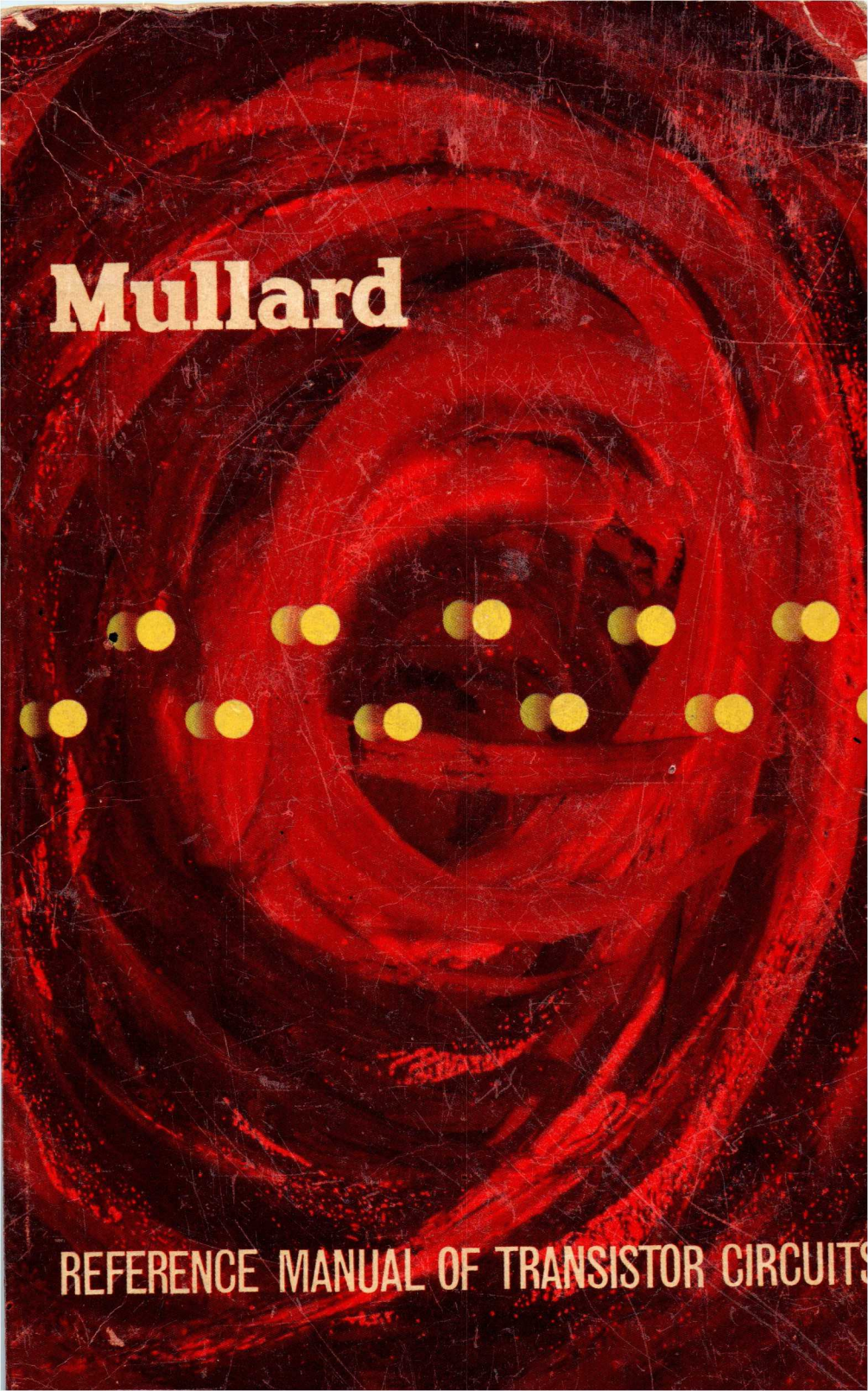


The background of the cover is a dark, textured red with black, swirling, concentric circular patterns that create a sense of depth and movement. Scattered across the middle of the cover are several pairs of small, bright yellow dots, arranged in a roughly horizontal line. The overall aesthetic is gritty and industrial.

Mullard

REFERENCE MANUAL OF TRANSISTOR CIRCUITS

Philippe
1. 3. 62

MULLARD TRANSISTOR CIRCUITS

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MULLARD REFERENCE MANUAL
OF **TRANSISTOR CIRCUITS**

Second edition 1961

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PREFACE TO THE FIRST EDITION

NO TECHNICAL BOOK EVER MADE LIGHT BEDTIME reading. Nor is it possible, in just over three-hundred pages of legible type, to transform the novice into a fully fledged circuit designer. What can be done, however, is to present the most important information to the largest number of users, in such a way that it can be absorbed as readily as possible. The result is the present *Reference Manual of Transistor Circuits*.

Not only is this manual not intended for the senior designer: he will probably prefer not to consult it—at least when anyone else is looking. The aim is therefore to help all the other people who have some professional (or spare-time) interest in radio and electronics to realise the possibilities of the transistor; and to show them how to exploit these possibilities intelligently. The approach is not a new one, though it is rarely carried out on the present scale. The most useful circuit diagrams are chosen and revised where necessary; as much practical information is given as possible; and the principles of operation are described in such a way that the user can make at least any minor modifications which he may require.

The method is not without its drawbacks. It can leave a false impression of the state of development and availability of the latest devices. There is always a time-lag between the introduction of a new device with its technical data, which enables the circuit designer to go ahead; and the production of a circuit for some particular application, which can be put in the hands of the general user. Broadly speaking the circuits have been restricted to devices which are available now. However, the breakthrough, which the alloy-diffusion process shows

every sign of being, means that all users must be acquainted with this technique, even if circuits cannot be included yet.

Other publications exist for reporting the latest developments. In particular, those who subscribe to the *Mullard Technical Handbook Service* are automatically supplied with the latest technical data sheets as these become available.

An important subsidiary aim has been to increase the ability of the average reader to appreciate technical data. The first nine chapters provide a background, with the main emphasis on small signals and audio frequencies. Subsequent chapters are more practical, though the necessary bits of theory are introduced where appropriate to extend the treatment of large signals and high frequencies.

Short chapters on semiconductor diodes and on the OCP71 photo-transistor have been included because of the close relationship that exists between the various semiconductor devices. A few topics—such as some switching applications and electronic computers—have been judged too specialised for inclusion in a book having a wide appeal.

The manual will be suitable for junior designers, graduate entrants into electronics, students of electrical engineering and physics, science teachers, radio service engineers, laboratory and works technicians, and keen hobbyists. The reader must use his judgement in choosing those parts which are most suitable to him. The apprentice service engineer may need to memorise the electrode connections, and the graduate entrant be more interested in equivalent circuits.

But whether on the bench or the bookshelf, this manual is primarily meant as a reference source of circuits.

PREFACE TO THE SECOND EDITION

In this second edition, which follows within a few months of the first, the opportunity has been taken to make a few minor alterations. The two editions have essentially the same content.

CHAPTER ONE

THE JUNCTION TRANSISTOR

Most junction transistors are, essentially, a sandwich of three layers of crystalline semiconductor. The manufacturing techniques and theory of operation are not so elementary as this simple picture would suggest, but a detailed knowledge of these topics is the concern of specialists, rather than of those who wish simply to make up transistor circuits.

Fortunately, the operation of the transistor as a circuit element can be understood from a few basic ideas about its construction and interior mechanism. The present chapter will provide this necessary basis, but it must be realised that the approach has been very much simplified.

MECHANICAL FEATURES

In the normal amplifying valve, the bulb is pumped out as thoroughly as possible, and a getter is provided to adsorb most of the residual gases. The flow of electrons from cathode to anode therefore takes place through an almost ideal vacuum. One or more grids placed between anode and cathode exert a controlling influence on the electron stream.

The transistor is, in essence, a crystal. It differs radically from the radio valve; in the valve the current flows through a near vacuum, necessitating an evacuated envelope and a more-or-less resilient assembly.

The current through the transistor, on the contrary, flows through a solid. The envelope does not have to be evacuated, and the device is mechanically stronger and non-microphonic.

A ready supply of current carriers is available at room temperature. The transistor does not require an auxiliary battery corresponding to the heater or filament supply of a valve. The transistor is therefore free from hum and less subject to deterioration and failure.

CONDUCTION THROUGH A SEMICONDUCTOR

In transistors and other semiconductor devices, current is carried by positive carriers as well as by negative electrons.

The positive carriers are referred to as holes. The hole has a positive charge exactly equal to the negative charge of the electron.

An electron and a hole, being of opposite sign, will attract each other, and the electron may eventually fill the hole. Neither can then take any further part in carrying current.

This process of recombination is fundamental to the operation of the transistor.

P- AND N-TYPE MATERIAL

Holes and electrons normally co-exist, but in varying proportions depending on the impurities present. In order to control the impurity content and hence the conductivity with the required precision, the semiconductor material is first very carefully purified. Then, by the addition of a small amount of selected additive, holes can be made much more numerous than electrons, or conversely.

Material in which holes predominate (Fig. 1) is known as p-type, where *p* stands for positive, the holes being positively charged. Where



Fig. 1—Semiconductor p- and n-type material

the majority carriers are electrons, the material is n-type, *n* standing for negative.

If holes are the majority carriers, electrons will also be present as minority carriers, and conversely.

THREE-ELECTRODE JUNCTION TRANSISTORS

Most transistors are in the three-electrode junction construction. Three layers of semiconductor merge into one another to form a

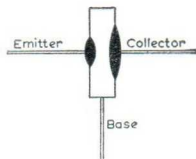


Fig. 2—Construction of three-layer (or triode) transistor

sandwich (Fig. 2). An ohmic, that is to say non-rectifying, contact is made to each of the three layers.

The central layer of the sandwich is very thin and is known as the

base, and the two outer layers are the emitter and collector. The junctions between (a) base and emitter and (b) base and collector are known as the emitter and collector junctions.

The transistors can be p-n-p or n-p-n, according to the additives used for emitter, base and collector (Fig. 3). The majority of transistors



Fig. 3—Sandwich arrangement of p-n-p and n-p-n layers

available at the moment are p-n-p. It will be simpler, however, to consider the n-p-n type first.

N-P-N Transistor

The emitter, as its name suggests, corresponds roughly to the cathode of the electronic valve. In the n-p-n transistor, the majority carriers in the n-type emitter are electrons. The emitter acts as a source of electrons, which flow into the base when the base is biased positively with respect to the emitter (Fig. 4).

The collector corresponds roughly to the anode of the valve. The collector of the n-p-n transistor, provided it is biased positively with respect to the base, absorbs electrons from the base.

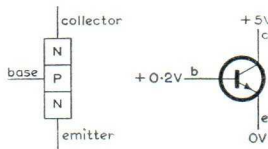


Fig. 4—Representation of n-p-n transistor. Conventional symbol on right shows polarities and typical voltages. Arrow on emitter points in conventional direction + to -.

Because the base is p-type, electrons only exist in the base as minority carriers. The electrons arriving at the collector are derived almost entirely from the emitter by diffusion through the base.

Not all the electrons flowing from the emitter into the base will be removed from the base by the collector, as a small proportion will recombine with holes in the p-type base layer.

This loss of charge in the base layer is made good by a flow of base current. Varying the base current varies the voltage across the emitter junction, and so controls the collector-emitter current.

P-N-P Transistor

The functions of emitter, base and collector are similar to those just described for the n-p-n transistor. This time, however, the emitter is p-type, and its majority carriers are holes. Holes flow from the emitter into the base when the base is biased negatively with respect to the

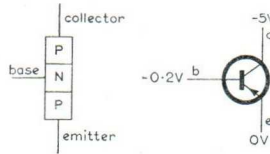


Fig. 5—Representation of p-n-p transistor

emitter (Fig. 5). The holes from the emitter diffuse through the base and are accepted by the collector. The collector has to be biased negatively with respect to the base, in order to absorb the positive holes.

A small proportion of the holes leaving the emitter recombine with electrons in the n-type base.

Emitter and Collector Junctions

The current carriers diffuse sideways through the base as well as forwards. The collector junction is therefore generally made considerably larger than the emitter junction (Fig. 2), to prevent excessive recombination of holes and electrons in the base.

Some types of transistor are sufficiently symmetrical to justify their use bi-directionally, that is, with what are normally the emitter and collector interchanged.

MANUFACTURE

The first process in transistor manufacture is the refining of germanium or silicon to a degree of purity beyond that attainable by normal

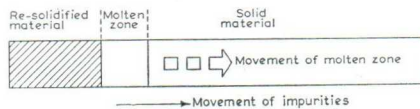


Fig. 6—Zone refining

chemical methods. The remaining impurity must not exceed about one part in ten thousand million.

The process is called zone refining (Fig. 6). An ingot of the metal is drawn slowly through a tube, which is surrounded at intervals by r.f. heating coils. Only a few zones of the ingot are molten at any one

time. The impurities become concentrated in the molten zones rather than in the solid portions. By drawing the ingot several times through the tube, the impurities are swept to one end, which is discarded.

The purified metal is then grown as a single crystal. This process may be carried out using the apparatus shown in Fig 7. The operation is performed in an atmosphere of nitrogen and hydrogen to prevent oxidation. The metal is kept just molten by an r.f. heating coil, the selected additive having been added in the proportion of

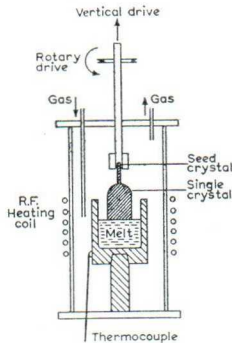


Fig. 7—Apparatus for growing single crystals

about one part in a hundred million. A seed crystal, held in a rotating chuck, is lowered into the molten metal and then slowly raised. A single crystal is withdrawn from the melt with the same crystal structure as the seed.

There are a number of ways of proceeding with the manufacture, but as most transistors are made by the alloy-junction method, only this method will be described in this chapter.

To make a transistor such as the OC71, a grown germanium crystal is turned into slices about $4 \times 2 \times 0.12$ mm. Each of these slices forms the base region of a transistor. Since the OC71 is a p-n-p transistor, the additive added before growing the crystal is n-type.

A pellet of a p-type additive is placed on each side of the slice, the one which will form the collector being about three times the size of the one used for the emitter. The assembly is heated in a hydrogen atmosphere, until the pellets melt and dissolve some of the germanium from the slice. On being cooled, the pellets with the dissolved germanium start to solidify, and a crystal of p-type germanium grows at the solid-liquid interface. At a lower temperature, the rest of the pellet solidifies.

Fig. 8 shows a cross-section of the resulting three-layer structure. Leads for the emitter and collector are soldered to the surplus material

in the pellets, thus making non-rectifying contacts. A nickel tab is soldered to the slice to make the connection to the base.

The assembly is etched to remove surface contamination.

Finally, the assembly is covered in moistureproof grease, and hermetically sealed into a small glass envelope, with the leads passing

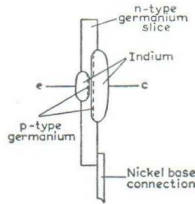


Fig. 8—Cross-section of p-n-p transistor

through the glass foot. The sealing process is performed so as to avoid overheating the transistor. A coating of opaque paint is applied to the outside of the glass bulb to exclude light.

CONCLUSION

Transistors depend for their operation upon the flow of current carriers (positive holes and negative electrons) within a semiconductor, that is, within a solid crystalline material. The presence of the two types of current carrier in differing concentrations, as majority and minority carriers, determines whether the material is p-type (excess holes) or n-type (excess electrons).

Junction transistors are commonly made in the form of a sandwich of three layers of p-n-p, or less frequently n-p-n, material. The central layer, or base, is always very thin. The outer layers form the emitter and collector, and generally speaking are not interchangeable. The collector circuit is connected to the negative side of the supply for a p-n-p transistor, and to the positive side for an n-p-n transistor. The emitter circuit is connected with the opposite polarity.

The characteristic part of the transistor is the junction, which is a region of transition between p- and n-type material. A feature of the junction is that a current will flow readily across it at room temperature, when a voltage of suitable polarity is applied; preliminary heating is not required.

APPENDIX**INSTALLATION AND GENERAL PROPERTIES****POLARITY**

Particular attention should be given to the circuit diagram. Connecting the transistor with the incorrect polarity may change its characteristics permanently for the worse, and could lead to its destruction.

ELECTRODE CONNECTIONS

The electrode connections are given in the technical data in the form of diagrams such as those shown in Fig. 9. The series of diagrams in the left-hand half of the figure, with the appropriate shape and dimensions, apply to most small-signal and low-power transistors, and to some small-signal high-frequency transistors.

The central lead is for the base, and there is a greater separation between the base and collector leads than between the base and emitter leads. The collector is further distinguished from the emitter by a spot on the adjacent part of the body.

The diagrams in the right-hand side of Fig. 9 are given as an example of a high-power output transistor. Two stiff leads project through the mounting base, and these can be identified as the base and emitter connections by the letters B and E stamped on

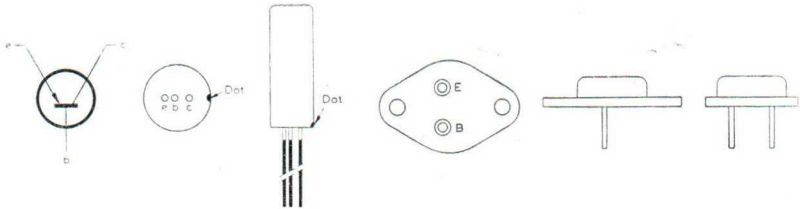


Fig. 9—Examples of electrode connections of transistors (not to scale)

the underside of the mounting base. The collector is electrically connected to the mounting base; and the collector connection is made to a tag which is held in contact with the top of the mounting base by one of the mounting nuts. The mounting base can be insulated from the chassis or heat sink as described in Chapter 9 (pp. 95 to 99).

RISK OF SHORT CIRCUIT

Disconnect the supply while installing transistors and, if possible, while making circuit adjustments. An accidental short circuit from the base to the collector supply line may cause sufficient current to flow to damage the transistor. Also, in some circuits, the supply voltage is higher than the rating of the transistor, the excess voltage being dropped by series resistance in the collector and/or emitter circuits. Short-circuiting this resistance may result in an excessive voltage being applied to the transistor.

Associated components and the devices themselves should not be inserted or

replaced with the power supplies connected, because of the surges which can occur, for instance, from the discharge of capacitors through the devices.

Reasonable care is called for, bearing in mind the proximity of flexible leads.

SOLDERING

The leads are tinned to facilitate soldering. Flexible leads should not be bent nearer than 1.5mm to the seal, otherwise the seal may be affected and the transistor no longer be moistureproof.

Soldering should be completed reasonably quickly. On no account should the transistor be allowed to heat up during the soldering operation. The leads may be held in a cool pair of long-jawed pliers. The pliers act as a thermal shunt which virtually prevents heat from being conducted to the transistor. Where the pliers would require a 'third hand', the transistor may be placed inside a crocodile clip, the jaws of which have been filed flat to grip the leads.

The electrical insulation between the heating element and the bit of some electrical soldering irons is sufficiently poor to cause a dangerously high current to flow through the transistor. Such irons should be disconnected during soldering. The plug may be arranged so that the iron is disconnected when it is picked up.

Transistors may be dip soldered, usually at a solder temperature of 240°C for a maximum of 10 seconds up to a point 2mm from the seal, but otherwise according to the specification.

MOUNTING POSITION

Transistors may be mounted in any position. The only restriction on the location of the transistor may arise from the need to provide adequate cooling and ventilation.

LIGHT AND OTHER RADIATION

Germanium and silicon are inherently sensitive to light. This light sensitivity is turned to useful account in the OCP71 phototransistor.

The ordinary transistor has to exclude light. For some transistors in a glass construction—for example, the OC71—the envelope is coated with an opaque paint. The paint resists normal handling, but should not be damaged. Some transistors are lightproofed internally. Power transistors, which are in a metal construction, require no special lightproofing.

High-energy radiations (for example, X rays, γ rays and neutrons) affect the behaviour of a junction, usually adversely and permanently.

STORAGE TEMPERATURE

Maximum and minimum limits are set to the temperatures at which transistors may be stored. For most transistors, the permissible range is from -55 to $+75^{\circ}\text{C}$. But the maximum storage temperature is lower for some types.

The minimum storage temperature is only likely to be encountered in specialised scientific, industrial and military applications.

Apart from the possibility of permanent mechanical damage from storage at temperatures below the minimum, the performance may be adversely affected by such low temperatures.

CHAPTER TWO

NEW TECHNIQUES

The OC71 process, which has served as a starting point for new constructions and techniques, is by no means typical of the latest designs. The new techniques have made possible radical extensions in performance in the directions of higher powers, higher frequencies, and higher voltage and current ratings.

POWER TRANSISTORS

The construction of a power transistor is represented in Fig. 1.

An important factor in the design of a power transistor is the emitter efficiency. In a p-n-p transistor, this is the ratio of the hole current to the total current across the emitter p-n junction, when it is positively biased in the forward direction. The emitter efficiency should be as high as possible.

The current amplification factor α' normally falls off as the operating current is increased, and the resulting non-linearity may be sufficient to limit the useful working range of the transistor. This decrease in α' depends on the emitter efficiency, which in turn depends on the solid solubility in germanium of the additive used for the emitter. A good emitter efficiency will be obtained, provided the conductivity of the p-type region is much higher than that of the n-type. A high solid solubility gives a high conductivity in the p-type region, and therefore

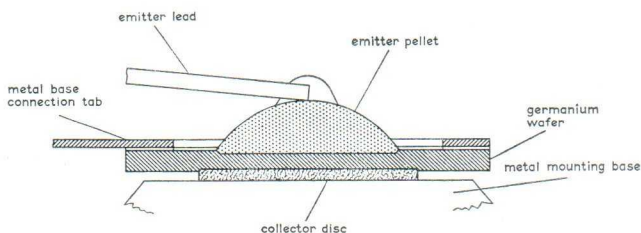


Fig. 1—Construction of power transistor

a high hole current. This in turn leads to a much more constant value of α' as the operating current is increased.

The solid solubilities of p-type additives in germanium increase in the order indium, gallium, aluminium and boron. At first, p-n-p transistors were made with emitter pellets of pure indium, but with the

demand for a higher α' at a higher operating current, the emitter efficiency was increased by the addition of a small quantity of gallium.

Even better emitter efficiency (and hence linearity) can be obtained by adding a small amount of aluminium to the emitter.

OC23 CONSTRUCTION

The base of the OC23 is very narrow and very uniform over a large area. The ratio of the diameter of the base to its width is about 80:1, as compared with about 25:1 for other germanium alloy-junction transistors.

To obtain this geometry, the penetration of the emitter into the base wafer has been kept small, so that the curvature of the edges of the emitter junction is much smaller than in transistors with larger emitter penetrations.

A special forming process has been developed to give the high-quality junctions necessary in this transistor.

The OC23 is designed and specially tested for driving square-loop ferrite computing elements and storage matrices. It provides one-amp pulses with a rise time of less than $0.8\mu\text{sec}$.

SILICON TRANSISTORS

Silicon transistors at present are made by the p-n-p alloy-junction process. Aluminium discs which form the emitter and collector are alloyed to opposite faces of a silicon disc (Fig. 2).

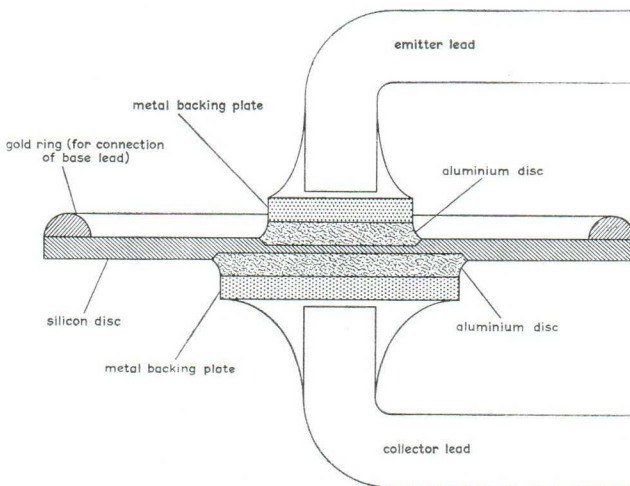


Fig. 2—Construction of a.f. silicon transistor

Two major advantages result from following what is fundamentally the same conception as that used for the OC71. First, basically the

same considerations apply when designing equipment for silicon as for germanium alloy-junction transistors, and equipment designers can derive maximum benefit from their experience gained with the germanium types. Second, the manufacturing experience obtained with millions of germanium devices is being used in the large-scale production of the silicon types.

Silicon transistors combine a number of distinct features:

- (a) low leakage current
- (b) relative freedom from thermal runaway
- (c) higher permissible junction temperature
- (d) higher collector-voltage ratings
- (e) low bottoming voltage
- (f) gain maintained at very low temperatures.

ALLOY-DIFFUSED TRANSISTORS

In alloy-junction transistors, the current is transferred by the diffusion of charged particles from the emitter to the collector junction within the field-free region of the base. Diffusion is a slow process, which severely limits the frequency performance of the transistor. In fact the

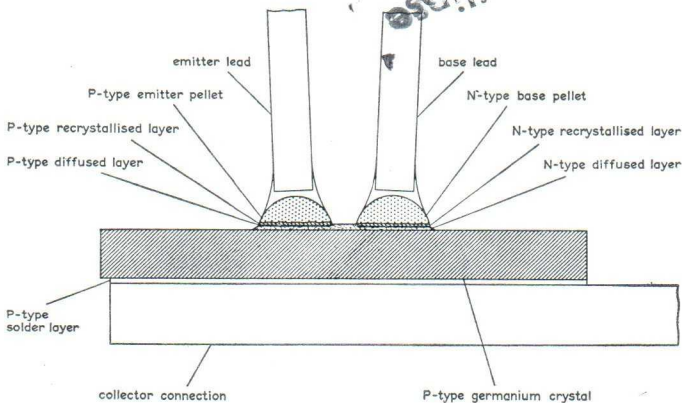


Fig. 3—Construction of alloy-diffused transistor

OC44 and OC42 are made with the narrowest base attainable in the alloy-junction technique, and give the best h.f. performance possible with this technique.

In the alloy-diffused construction (Fig. 3), the width of the base is reduced to a few thousandths of a millimetre, and an accelerating or 'drift' field is introduced between the emitter and collector junctions.

The accelerating field is created by doping the emitter pellet with both p- and n-type additives, and heating to a high temperature for a carefully controlled time. The n-type additive penetrates the crystal more deeply than the p-type, and forms a graded base layer which accelerates the holes towards the collector.

CHAPTER THREE

STATIC CHARACTERISTIC CURVES

The transistor will now be regarded as a three-electrode device, and the relationships between the currents and voltages at the electrodes examined. Comparisons will be made with the thermionic valve to bring out similarities and dissimilarities.

The transistor has four main parameters, namely, input voltage and current and output voltage and current, and its d.c. performance is described completely by a set of four graphs.

There is more than one way in which the set can be selected, but the graphs shown in this chapter are the most convenient. These graphs are typical of a small-signal transistor.

The temperature dependence and the techniques for dealing with it will not be considered until later chapters.

CURRENT AMPLIFIER

The transistor is usually regarded as a current amplifier, and its characteristics are often described in terms of current.

The base current, which is only a few percent of the emitter current, is important because it controls the current in the emitter-collector circuit. A similar controlling function is exercised in the valve by the control-grid voltage.

The direct or alternating currents at collector, emitter and base can be added up in the usual way. If the direct emitter current is 5mA and the collector current 4.9mA, the base current is 100 μ A.

A definite relationship exists between the currents at the transistor electrodes. This relationship is expressed by two ratios, α and α' , which are characteristic of the transistor.

α is the general symbol for collector current divided by emitter current, and α' for collector current divided by base current. One can be obtained from the other, since $\alpha' = \alpha/(1-\alpha)$ and $\alpha = \alpha'/(1+\alpha')$. If the current at one electrode is known, it suffices to know either α or α' to find the currents at the other electrodes.

In practice the values of α and α' to be used may depend on the frequency and current. In the present chapter only the simplest case will be considered, that of small changes of current at zero frequency (d.c.).

INPUT CHARACTERISTIC

The input characteristic shows the variation of input current with input voltage. When the input is applied to the base, the input characteristic shows the base current I_b plotted against the base voltage V_b (Fig. 1). The base voltage is measured relative to the emitter, and corresponds to the voltage V_{be} between base and emitter in a practical circuit.

When the input electrode is the emitter, the input characteristic shows a plot of emitter current I_e against emitter voltage V_e (Fig. 2). It is usually necessary to interpret V_e as V_{eb} in practice.

Input Resistance

The reciprocal of the gradient of the input characteristic is the input resistance. This is typically of the order of 50 to 100 Ω with the input

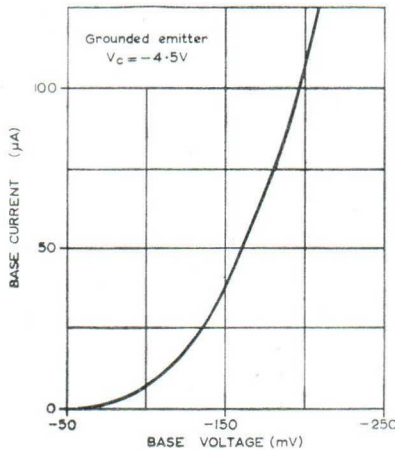


Fig. 1—Input characteristic in grounded emitter

applied to the emitter, and 500 Ω to 1k Ω with base input, and is much lower than for a valve.

The input characteristic derives from the characteristic of the forward-biased emitter junction. The input resistance is therefore low. The characteristic is approximately exponential in shape, as may be predicted theoretically for the semiconductor diode.

The input characteristic is thus quite non-linear. The input resistance depends very much on the current at which it is measured.

The collector voltage is kept constant during the plotting of the input characteristic; in a practical circuit, constant collector voltage could only arise with the output short-circuited to a.c. (zero load).

Current Drive

Since the resistance of the emitter diode changes with the current flowing through it, as shown by the non-linearity of the input characteristic, the transistor is normally current biased and driven from a

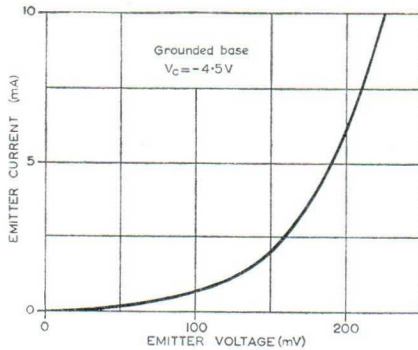


Fig. 2—Input characteristic in grounded base

current rather than a voltage source. Current drive is achieved by using an effective source resistance which is large in comparison with the input resistance.

Input-circuit Distortion

If the constant impedance of the source is not sufficiently high to swamp the varying impedance of the transistor under drive, high input-circuit distortion will result.

When using an oscilloscope to examine the input waveform for distortion, the current waveform should be monitored, as often the voltage waveform is misleading.

TRANSFER CHARACTERISTIC

The transfer characteristic of a valve shows the variation of anode current with control-grid voltage, that is, the dependence of the output current on the input voltage. The mutual conductance or slope g_m of the valve corresponds to the gradient at some point of this characteristic, and is normally specified in milliamperes per volt. A similar quantity is sometimes expressed for the transistor; this is the gradient of the I_c/V_b characteristic, which is of the order of tens or hundreds of milliamperes per volt.

However, if the collector current is plotted against the base current, the resulting curve is much more linear than the I_c/V_b characteristic. This characteristic (Fig. 3) is the one normally referred to as the

transfer characteristic of the transistor in grounded emitter (input to base).

The transfer characteristic in grounded base (input to emitter) shows I_c plotted against I_e (Fig. 4).

Non-linearity Distortion

Any non-linearity in the transfer characteristic gives rise to non-linearity distortion in the output. This distortion is low for the OC71 and similar transistors operated at collector currents in the region of

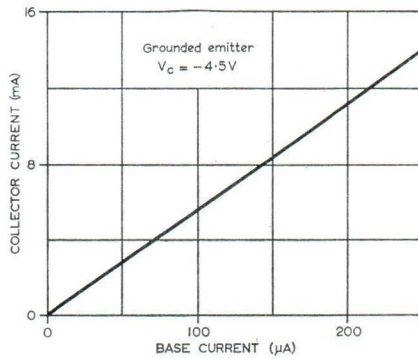


Fig. 3—Transfer characteristic in grounded emitter

a milliamp or so; however, operation below 0.3mA or above 25mA is inadvisable when low distortion is required.

Current Amplification Factor* with Input to Base (Grounded Emitter)

The gradient at any point of the I_c/I_b transfer characteristic is a ratio and has the dimensions of a pure number. This is the current amplification factor of the transistor with the input applied to the base. It is represented by α'_0 (alpha nought dash).

The subscript $_0$ indicates that α'_0 applies to zero frequency (d.c.).

From Fig. 3, it can be seen that the base current is $250\mu A$ at a collector current of 14mA, the collector voltage being kept constant at $-4.5V$. The corresponding value of α'_0 is therefore $1400/25 = 56$, the characteristic being linear in the range shown.

*The term current amplification factor may be applied to transistors, by analogy with the voltage amplification factor of a valve. The more usual expression is current gain; sometimes this term is qualified as the transistor current gain, to distinguish it from the current gain of the stage.

α'_o is not greatly affected by collector voltage. However, the I_c/I_b characteristic is measured at some constant value of collector voltage which, for the sake of completeness, is shown on the curve. Constant

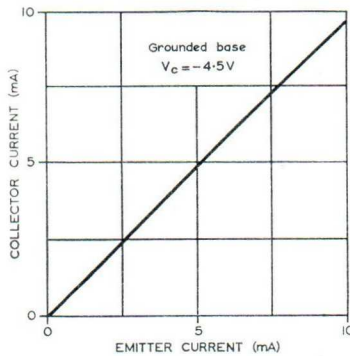


Fig. 4—Transfer characteristic in grounded base

collector voltage could only arise in a practical circuit with the output short-circuited to a.c.

The definition of α'_o is given mathematically by

$$\alpha'_o = \left| \frac{\partial I_c}{\partial I_b} \right|_{V_c}$$

This equation restates in mathematical terms that α'_o is a numerical quantity, which corresponds to the slope of the I_c/I_b curve, and is measured at constant collector voltage.

Current Amplification Factor with Input to Emitter (Grounded Base)

The gradient of the I_c/I_e curve (Fig. 4) is the current amplification factor in grounded base and is given the symbol α_o . The relationship is linear in the range shown, and α_o is $9.8/10 = 0.98$. α_o is defined mathematically as

$$\alpha_o = \left| \frac{\partial I_c}{\partial I_e} \right|_{V_c}$$

OUTPUT CHARACTERISTIC

The output characteristic of a valve consists of a plot of anode current against anode voltage for various control-grid voltages.

The output characteristic of the transistor, which is normally current driven, has input current as parameter instead of input voltage. Such a graph is shown in Fig. 5 for a transistor in grounded emitter (the

input current being the base current). The base current is normally negative, according to the usual sign convention.

In a practical circuit, which may contain collector and emitter resistors, for example, the collector voltage V_c is the collector-emitter voltage V_{ce} across the transistor.

Fig. 6 shows the output characteristic in grounded base. The input current is now the emitter current which is plotted as parameter.

Output Resistance

At voltages above the knee voltage, which is only about 0.2V in Fig. 5, a comparatively large change in collector voltage produces a relatively

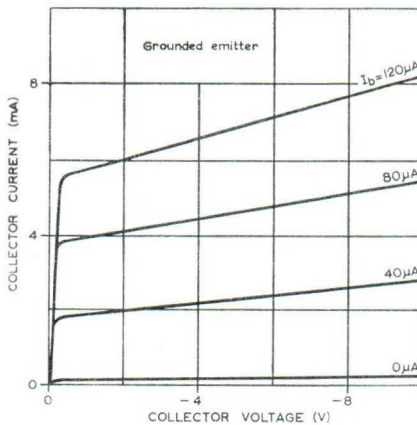


Fig. 5—Output characteristic in grounded emitter

small change in collector current. The transistor therefore has a high output resistance. This can be explained by regarding the collector-base junction as a diode biased in the reverse direction.

The gradient of the output characteristic, because of the way it is plotted, has the dimensions of a conductance (current/voltage) which is low. The reciprocal of this is the output resistance, which is high. The curves apply to constant base current, which would only be obtained with a high source resistance in the base circuit.

Pentode-like Characteristic

The transistor, which is basically a three-electrode or 'triode' device, has an output characteristic like that of a pentode valve in having a well-defined knee and a region of high output resistance. The knee

voltage of the transistor (about 0.2V to 3V) is much lower than that of the pentode valve (which may be about 30V). Hence low battery voltages can be used for transistors, while still obtaining high efficiency.

Output Characteristic related to Transfer Characteristic

If a straight line be drawn across the output characteristic for some particular value of collector voltage, and the readings of I_c and I_b or I_e replotted, the result is the transfer characteristic for the collector

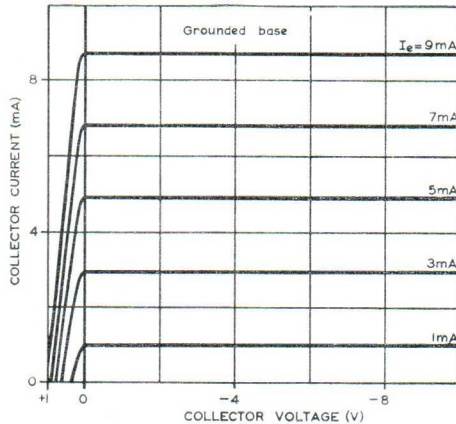


Fig. 6—Output characteristic in grounded base

voltage in question. The non-linearity in the transfer characteristic therefore appears as an uneven spacing of the curves on the output characteristic, for equal changes in input current.

Collector Leakage Current

GROUNDING EMITTER

A closer inspection of the output characteristic in grounded emitter (e.g. Fig. 5) leads to the conclusion that a finite collector current continues to flow, even when the base is open circuit ($I_b = 0$). This current is called the collector leakage current in grounded emitter, and is commonly given the symbol I'_{co} . Alternatively it may be represented by I_{ceo} , where the subscript ce designates a collector-emitter current, and the subscript o shows that the current in the remaining electrode (the base, b) is zero (Fig. 7).

As is shown by Fig. 5, I'_{co} increases slightly with collector voltage.

The value of I'_{co} is typically $150\mu A$ for an OC71 ($V_c = -4.5V$), at an ambient temperature of $25^\circ C$.

GROUNDING BASE

The collector leakage current in grounded base is the base-collector current with the emitter open circuit (Fig. 8). The corresponding symbol is I_{co} or I_{cbo} .

I_{co} is typically $4.5\mu A$ for an OC71 at $V_c = -4.5V$ and at an ambient temperature of $25^\circ C$. This quantity is too small to be shown on the output characteristic in grounded base (Fig. 6).

I_{co} is the current flowing in the collector-base diode biased in its reverse direction (Fig. 8). The term 'leakage current' arises because no current would flow through an ideal diode under similar conditions.

The full value of leakage current in grounded-emitter circuits ($= I'_{co}$) is only obtained when the base is open circuit. When the

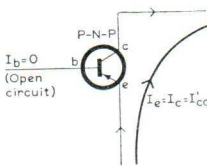


Fig. 7—Collector leakage current in grounded emitter

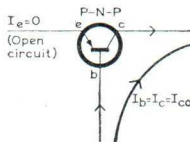


Fig. 8—Collector leakage current in grounded base

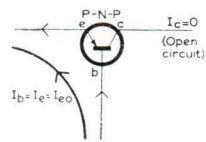


Fig. 9—Emitter leakage current

base is connected by a resistance to ground (i.e. to emitter), a leakage current lying between I'_{co} and a value somewhat higher than I_{co} is obtained. The leakage current does not drop to I_{co} when the base is short-circuited to ground, because of the small forward bias produced across the emitter-base junction by the flow of leakage current through the internal base resistance $r_{bb'}$.

Emitter Leakage Current

If the collector is open circuit as in Fig. 9 ($I_c = 0$), the transistor reduces to a base-emitter diode. The current which flows when this diode is biased in the reverse direction, with the emitter negative for a p-n-p transistor, is the emitter leakage current, I_{eo} or I_{ebo} .

The emitter leakage current is usually of the same order of magnitude as the collector leakage current in grounded base. Thus for the OC71, I_{eo} is typically $3.5\mu A$ at $V_e = -4.5V$ and at an ambient temperature of $25^\circ C$.

FEEDBACK CHARACTERISTIC

The variation of base voltage (base-emitter voltage) with collector voltage may be shown by means of a feedback characteristic. This curve

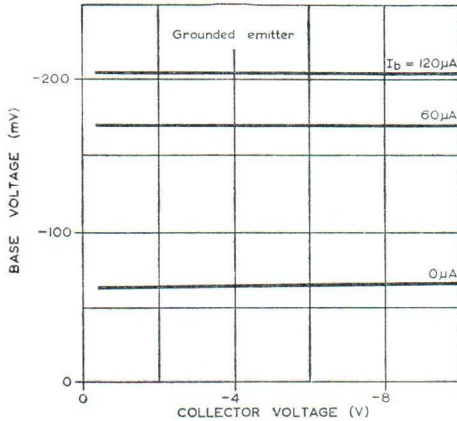


Fig. 10—Feedback characteristic in grounded emitter

(Fig. 10) is not of any great practical value, and is no longer given in the published data.

The effect of feedback within the transistor is extremely important, and not directly comparable with anything that occurs in a valve. This point will be taken up again when discussing small-signal characteristics and equivalent circuits.

UNIVERSALITY OF STATIC CHARACTERISTICS

The static characteristic curves are normally given for

- (a) emitter grounded (input to base, output from collector)
- (b) base grounded (input to emitter, output from collector).

The curves are not quoted with reference to grounded collector (input to base, output from emitter).

The information contained in the curves applies whichever electrodes are used for the input and output.

For example, the output characteristic in grounded emitter expresses a relationship between I_c , V_c and I_b , which is still valid in the other configurations, provided V_c is interpreted as V_{ce} . Similarly, the I_c/I_b curve is true for all configurations, although the term transfer characteristic is appropriate to this curve only when the transistor is in grounded emitter.

CONCLUSION

- (a) The thermionic valve has a comparatively high input impedance and is voltage driven. The transistor, on the other hand, has a comparatively low input resistance (impedance), and normally must be current driven from a high source impedance.
- (b) The characteristics of the transistor are often expressed in terms of input current rather than input voltage.
- (c) The output characteristic of the 'triode' transistor resembles that of a pentode valve in having a definite knee and a region of high output resistance. The knee voltage of the transistor is much lower than that of the pentode valve, and low supply voltages can be used while retaining high efficiency.
- (d) When one of the electrodes is open circuit, the transistor does not reduce to an ideal diode, because leakage current flows through the reverse-biased diode.
- (e) Transistor characteristics are normally given for grounded emitter and grounded base. This method of presentation is more extensive than that adopted in valve data.
- (f) One of the peculiarities of the transistor is its internal feedback.

CHAPTER FOUR

SMALL-SIGNAL CHARACTERISTICS

The static characteristic curves described in the previous chapter are typical of those normally given. Such curves are useful primarily for choosing a working point, and in particular for the design of a.f. output stages. The curves are less useful for the design of small-signal a.f. stages.

BLACK BOX

The most profitable way of approaching the a.c. performance of the transistor is to treat the device as a four-terminal (or four-pole) 'black box' (Fig. 1). The box has a pair of input and a pair of output terminals.

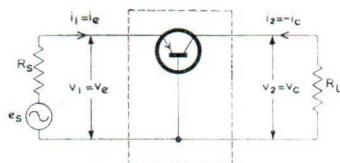


Fig. 1—A.C. quantities of 'black box' in grounded base

Relationships between the signal voltages and currents measured at the input or output of the black box are called four-pole characteristics or parameters.

The 'black box' treatment is not peculiar to transistors. It is a general method which can be applied to any electrical network.

Small-signal characteristics are published for a.f. transistors in the form of 'h' and 'modified z' systems. These parameters are suitable for calculating performance at audio frequencies or, more precisely, at frequencies low in comparison with the 'cut-off frequency'.

For some h.f. transistors the four-pole characteristics are quoted as a set of four admittances (y parameters).

HYBRID OR h PARAMETERS

One set of small-signal a.f. characteristics is formed by the slopes of the static characteristic curves at the working point. The slopes cannot

be found with sufficient accuracy from the graphs, and so they are given for one or two nominal working points.

The symbol for these characteristics is an h which is modified by numbers 1 and 2 in subscript. A subscript 1 refers to a voltage or

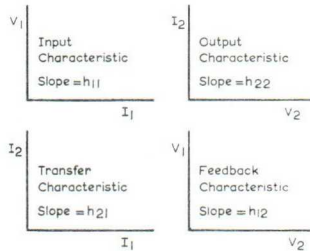


Fig. 2—Notation for h -parameter subscripts

current measured at the input terminals, while a subscript 2 denotes a voltage or current at the output terminals (Fig. 2).

Capital letters represent direct voltage and current at the input electrode (V_1 , I_1) and at the output electrode (V_2 , I_2). Small letters will represent a.c. quantities at the input electrode (v_1 , i_1) and output electrode (v_2 , i_2). In practice v_1 , i_1 and v_2 , i_2 represent small amplitudes.

Characteristics in the h system are defined as follows:

- $h_{11} = v_1/i_1$ = slope of input characteristic
= input impedance for constant output voltage
- $h_{21} = i_2/i_1$ = slope of transfer characteristic
= current amplification factor for constant output voltage
- $h_{22} = i_2/v_2$ = slope of output characteristic
= output admittance for constant input current
(= reciprocal of output impedance)
- $h_{12} = v_1/v_2$ = slope of feedback characteristic
= voltage-feedback ratio for constant input current.

Constant output voltage means that $v_2 = 0$ and that the output (load) is short-circuited to a.c. (zero load); for example, a large capacitance may be connected across the output terminals. Constant input current means that $i_1 = 0$ and that the input is open-circuited to a.c., as for example by including a large series resistance or inductance. Thus it is usual to see the characteristics expressed as:

- h_{11} = input impedance with output short-circuited to a.c.
- h_{21} = forward current transfer with output short-circuited to a.c.
- h_{22} = output admittance with input open-circuited to a.c.
- h_{12} = reverse voltage transfer with input open-circuited to a.c.

h_{11} is measured in ohms or kilohms, h_{21} is a ratio (that is, a pure number), h_{22} is measured in reciprocal ohms (mhos or micro-mhos) and h_{12} again is a ratio. Because of their different dimensions, these parameters are referred to as hybrid, whence the symbol h .

Similar quantities can be defined for grounded emitter from the appropriate set of characteristic curves, the symbols then being primed, thus: h'_{11} , h'_{21} , h'_{22} and h'_{12} . The values of the primed and unprimed quantities are normally different, corresponding to the different values measured in grounded emitter and grounded base.

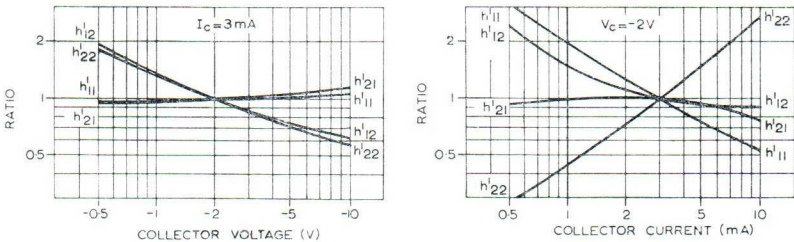


Fig. 3—Variation of h -parameters with working point, for OC71 in grounded emitter

The values of the h parameters as given by the slopes of the static characteristic curves would only apply to zero frequency. In practice, the parameters are measured at a representative audio frequency (1000c/s), using special equipment.

The values of the characteristics also depend on the working point. Fig. 3 for the OC71 in grounded emitter enables the characteristics for other working points to be derived from the values at $-2V$, $3mA$.

The h parameters can also be defined from the equations:

$$v_1 = h_{11}i_1 + h_{12}v_2 \quad \dots(1)$$

and

$$i_2 = h_{21}i_1 + h_{22}v_2. \quad \dots(2)$$

Performance equations based on the h -parameters are given in Table 1, at the end of the chapter (p. 31).

This system of characteristics shows to particular advantage when it is required to calculate the performance of circuits containing several stages, feedback networks, etc. The characteristics of the transistor can be expressed in the form of an h matrix, a matrix merely being a special notation for writing down coefficients. Thus Eqs. 1 and 2 can be re-written as

$$\begin{bmatrix} v_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ v_2 \end{bmatrix}.$$

The characteristics of any network, such as a feedback path for instance, can also be expressed in matrix form. The performance of a circuit containing transistors and other networks can be calculated by manipulating the matrices according to certain rules (matrix algebra)*.

GENERAL NOTATION

The subscripts ₁ and ₂ constitute a general notation for distinguishing quantities measured at the input and output terminals. The voltages and currents can be combined in the appropriate ways to give resistance and impedance (*r* and *z*), admittance (*y*) and conductance (*g*).

z PARAMETERS

A set of four parameters which all have the dimensions of impedance can be written out using the general notation, thus:

$z_{11} = v_1/i_1$ = input impedance with output open-circuited to a.c.

$z_{21} = v_2/i_1$ = forward transfer impedance with output open-circuited to a.c.

$z_{22} = v_2/i_2$ = output impedance with input open-circuited to a.c.

$z_{12} = v_1/i_2$ = reverse transfer impedance with input open-circuited to a.c.

MODIFIED z PARAMETERS

Equations based on the *h* system (Table 1) do not indicate clearly the characteristic values and the trends in performance as the quantities are varied. A further set of a.f. characteristics has therefore been produced. These are more suitable for elementary circuits which do not include a.c. feedback and other networks.

The system consists of five characteristics which are again defined from the 'black box'. They are (for grounded base):

forward current transfer α at constant collector voltage (output short-circuited to a.c.)

input impedance z_{in} at constant collector voltage (output short-circuited to a.c.)

input impedance z_{11} at constant collector current (output open-circuited to a.c.)

output impedance z_{out} for constant input voltage (input short-circuited to a.c.)

output impedance z_{22} for constant input current (input open-circuited to a.c.)

* Matrix methods for circuit calculations are described in *Principles of Transistor Circuits*, edited by R. F. SHEA. Wiley (New York). Chapman and Hall (London).

From these definitions, three of the five characteristics will be seen to be related directly to the h characteristics, thus:

$$\alpha = |h_{21}|$$

$$z_{in} = h_{11}$$

and

$$z_{22} = \frac{1}{h_{22}}.$$

The other two, z_{11} and z_{out} , are related to the h system as follows:

$$z_{11} = h_{11} - h_{21} \cdot \frac{h_{12}}{h_{22}} = \frac{\Delta}{h_{22}}$$

and

$$z_{out} = \frac{h_{11}}{h_{11}h_{22} - h_{21}h_{12}} = \frac{h_{11}}{\Delta},$$

where

$$\Delta = h_{11}h_{22} - h_{21}h_{12}.$$

A set of five characteristics is defined in a similar way for grounded emitter, the symbols being primed, thus: α' , z'_{in} , z'_{11} , z'_{out} and z'_{22} .

The equations using this system (Table 2) consist of a simple fixed term, dependent on the transistor only, multiplied by a term which varies in a fairly simple manner with the circuit values.

One of the merits of this system is that it shows the interdependence of the input and output impedances. In an actual circuit, the input impedance lies between the values with open-circuit and short-circuit output. The output impedance lies between the two extreme values which correspond to open-circuit and short-circuit input. The interdependence of the input and output impedances is the effect of internal feedback, which is allowed for in the h system by the voltage-feedback factor h_{12} .

As can be seen from the definitions,

$$z'_{11} = z_{11} \quad \text{and} \quad z'_{out} = z_{out}.$$

Other useful relationships are:

$$z'_{in} = z_{in}(1 + \alpha') \simeq \alpha' z'_{in};$$

$$z'_{22} = \frac{z_{22}}{1 + \alpha'} \simeq \frac{z_{22}}{\alpha'};$$

$$\frac{z'_{in}}{z'_{11}} = \frac{z_{out}}{z_{22}}; \quad \text{and} \quad \frac{z'_{in}}{z'_{11}} = \frac{z'_{out}}{z'_{22}}.$$

Dynamic Performance

The modified z characteristics are particularly useful for calculating the dynamic performance of individual transistor stages.

The equations to be applied depend on the relative magnitude of the load resistance R_L . Certain basic equations can be derived for an R_L which is of the order of the output impedance (say, $30k\Omega$). When R_L is low, only $1k\Omega$ or so, the equations can be simplified.

In fact the choice of equations is decided by the coupling. The value of R_L which will give maximum gain is high, and can normally only be provided by transformer coupling.

When RC coupling is favoured, the load is virtually formed by the input impedance of the following stage, which is low and effectively shorts the collector resistor to a.c.

Hence the calculations are simpler to perform for RC coupling.

TRANSFORMER COUPLING

Power gain is a maximum when the load resistance is $R_L = \sqrt{z'_{22}z'_{out}}$ and the source resistance is $R_s = \sqrt{z'_{11}z'_{in}}$.

Maximum gain occurs with these optimum values and is given by

$$\text{Matched Power Gain} = \left\{ \frac{\alpha'}{\sqrt{z'_{11}} + \sqrt{z'_{in}}} \right\}^2 \cdot z'_{22}.$$

(Similar equations apply in grounded base, using unprimed quantities.)

As an example, consider a typical OC75 in grounded-emitter connection. The following values of the parameters are for a working point of $-2V$, $3mA$. With $z'_{22} = 7.8k\Omega$ and $z'_{out} = 14k\Omega$, the optimum load resistance is $\sqrt{(7.8 \times 14)} = 10.5k\Omega$. With $z'_{11} = 720\Omega$ and $z'_{in} = 1.3k\Omega$, the optimum source resistance is

$$\sqrt{(720 \times 1300)} = 970\Omega.$$

The matched power gain is

$$\left\{ \frac{90}{\sqrt{720} + \sqrt{1300}} \right\}^2 \times 7800 = 15,900 = 42dB,$$

where the power gain in dB is ten times the logarithm to the base ten of the numerical power gain; thus $10 \log_{10} 15,900 = 10 \times 4.2014 = 42dB$. Similar calculations for grounded base, using the unprimed quantities α , z_{11} , z_{in} and z_{22} , show that the maximum gain available is about 29dB in this configuration.

For multistage amplifiers, matching can easily be achieved with interstage coupling transformers. A stepdown ratio is required.

RC COUPLING

With RC coupling, the collector resistor is normally several $k\Omega$'s, and the input resistance of the following stage only about 700Ω . Thus the output is more or less shorted to a.c. The current gain A_i and the input impedance of the grounded-emitter stage are approximately equal to the small-signal parameters, α' and z'_{in} .

The voltage gain A_v of the stage is

$$A_v = \alpha' \frac{R_L}{z'_{in}}$$

where R_L , the load resistance, is substantially equal to the input impedance of the following stage.

The power gain A_w is

$$\begin{aligned} A_w &= A_i \times A_v \\ &= (\alpha')^2 \frac{R_L}{z'_{in}} \end{aligned}$$

For identical stages in cascade, $R_L = z'_{in}$ and $A_w = (\alpha')^2$, so that for a typical α' of 41 (for the OC71),

$$A_w = (41)^2 = 1680 = 32\text{dB}.$$

In a practical RC-coupled stage, the power gain will always be less than $(\alpha')^2$ because:

- (a) there is considerable loss in the coupling network (collector resistor, coupling capacitor, and biasing components);
- (b) the factor R_L/z'_{in} is less than one, because succeeding stages are operated at higher currents to accommodate the increased signal swing, and the input impedances of the stages become progressively lower.

The power gain of an OC71 stage with RC coupling is typically about 26 to 30dB.

LOW-FREQUENCY T NETWORK

The low-frequency T network is described at greater length in a later chapter. Performance equations based on the T-network parameters are given in Table 3 for grounded-emitter operation. For the present purpose, the notation of these parameters is sufficiently explained by Fig. 4, where one version of the equivalent circuit includes a voltage generator $i_{e_r m}$, and the other a current generator αi_e . The equations can be considerably simplified by neglecting r_m in comparison with r_e and r_c .

The chief disadvantage of the T-network parameters, as compared with those defined in the h and modified z systems, is that they lead to

cumbersome expressions. Furthermore, a different equation is required for each of the three circuit configurations. The parameters themselves, however, have the same value for each configuration.

The parameters, which depend to some extent on the operating

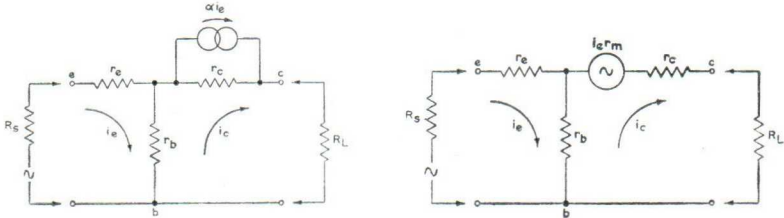


Fig. 4—Low-frequency T network

conditions, are useful up to a point for understanding large-signal performance, if average values are employed.

Tables 4 and 5 show how to evaluate the h and modified z parameters from the T-network parameters.

y PARAMETERS

The y parameters form another set of four-pole characteristics. The subscripts are used as for the h parameters to denote quantities measured at the input and output terminals. Similarly, a dash or prime

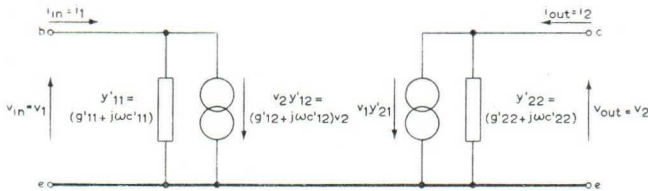


Fig. 5— y -parameters and their components

indicates values for grounded-emitter operation. Unlike the h parameters, however, the y parameters all have the same dimensions, namely, admittance. Thus in grounded emitter:

- $y'_{11} = i_1/v_1$ = input admittance with output short-circuited to a.c.
- $y'_{21} = i_2/v_1$ = forward transfer admittance with output short-circuited to a.c.
- $y'_{22} = i_2/v_2$ = output admittance with input short-circuited to a.c.
- $y'_{12} = i_1/v_2$ = reverse transfer admittance with input short-circuited to a.c.

Values of the y parameters are given in the published data for r.f. transistors such as the OC170. The input, output and reverse transfer admittances are resolved into a conductance and a capacitance (Fig. 5). The three conductances increase with frequency and the capacitances decrease.

The phase shift within the transistor is given as ϕ'_{21} , the phase angle of the forward transfer admittance.

TABLE 1
PERFORMANCE EQUATIONS BASED ON h PARAMETERS
Grounded Base*

$G_L = 1/R_L$	$\Delta = h_{11}h_{22} - h_{12}h_{21}$
Input Resistance	$h_{11} - \frac{h_{12}h_{21}}{h_{22} + G_L}$
Output Resistance	$\frac{1}{h_{22} - \frac{h_{12}h_{21}}{h_{11} + R_s}}$
Current Gain	$h_{21} \cdot \frac{G_L}{h_{22} + G_L}$
Voltage Gain	$-\frac{h_{21}}{\Delta + h_{11}G_L}$
Power Gain	$\frac{(h_{21})^2 G_L}{(h_{22} + G_L)(\Delta + h_{11}G_L)}$

* For grounded emitter use the same equations with primed quantities.

TABLE 2
PERFORMANCE EQUATIONS BASED ON MODIFIED z PARAMETERS
Grounded Base*

Input Resistance	$z_{11} \cdot \frac{R_L + z_{out}}{R_L + z_{22}}$
Output Resistance	$z_{22} \cdot \frac{R_s + z_{in}}{R_s + z_{11}}$
Current Gain	$\alpha \cdot \frac{z_{22}}{R_L + z_{22}}$
Voltage Gain	$\frac{\alpha z_{22}}{z_{11}} \cdot \frac{R_L}{(R_L + z_{out})}$
Power Gain	$\frac{(\alpha z_{22})^2}{z_{11}} \cdot \frac{R_L}{(R_L + z_{22})(R_L + z_{out})}$

* For grounded emitter use the same equations with primed quantities.

TABLE 3**PERFORMANCE EQUATIONS* BASED ON T-NETWORK PARAMETERS**
Grounded Emitter

Input Resistance	$r_b + r_e + \frac{r_e(r_m - r_e)}{R_L + r_c + r_e - r_m}$
Output Resistance	$r_e + r_c - r_m + \frac{r_e(r_m - r_e)}{R_s + r_b + r_e}$
Voltage Gain	$\frac{(r_e - r_m)R_L}{(R_s + r_b + r_e)(R_L + r_e + r_c - r_m) + r_e(r_m - r_e)}$
Power Gain†	$\frac{4R_LR_s(r_m - r_e)^2}{[(R_s + r_b + r_e)(R_L + r_e + r_c - r_m) + r_e(r_m - r_e)]^2}$

*These equations can be simplified by neglecting r_m in comparison with r_e and r_e

†Power output divided by max. power obtainable from generator

TABLE 4**CONVERSION FROM T-NETWORK PARAMETERS TO h PARAMETERS**

<i>Grounded Base</i>	<i>Grounded Emitter</i>	<i>Grounded Collector</i>
$h_{11} = r_e + (1 - \alpha)r_b$	$h'_{11} = (1 + \alpha')h_{11}$	$h''_{11} = h'_{11}$
$-h_{21} = \alpha$	$h'_{21} = \alpha'$	$-h''_{21} = 1 + h'_{21}$
$h_{22} = 1/r_c$	$h'_{22} = (1 + \alpha')h_{22}$	$h''_{22} = h'_{22}$
$h_{12} = r_b/r_c$	$h'_{12} = h'_{22}r_e$	$h''_{12} = 1/(1 + h'_{12})$

TABLE 5**CONVERSION FROM T-NETWORK PARAMETERS TO MODIFIED z PARAMETERS**

<i>Grounded Base</i>	<i>Grounded Emitter</i>
$\alpha = r_m/r_c$	$\alpha' = r_m/(r_c - r_m)$
$z_{11} = r_e + r_b$	$z'_{11} = r_e + r_b$
$z_{1n} = r_e + (1 - \alpha)r_b$	$z'_{1n} = r_b + r_e(1 + \alpha')$
$z_{22} = r_c$	$z'_{22} = r_c/(1 + \alpha')$
$z_{out} = r_c \left\{ 1 - \frac{\alpha r_b}{r_e + r_b} \right\}$	$z'_{out} = \frac{r_c}{1 + \alpha'} \left\{ 1 + \frac{\alpha' r_e}{r_e + r_b} \right\}$

CHAPTER FIVE

BASIC A.C. CIRCUITS

The three basic circuit arrangements of the transistor will be considered in this chapter from the a.c. point of view. The modified z parameters, described in the previous chapter, will be used to compare the performance, as the equations based on these parameters show most clearly the effects of the load and source resistances.

For the sake of comparison, the basic circuits of the triode valve are given in Figs. 1, 2 and 3. The first is the grounded-grid circuit, in which

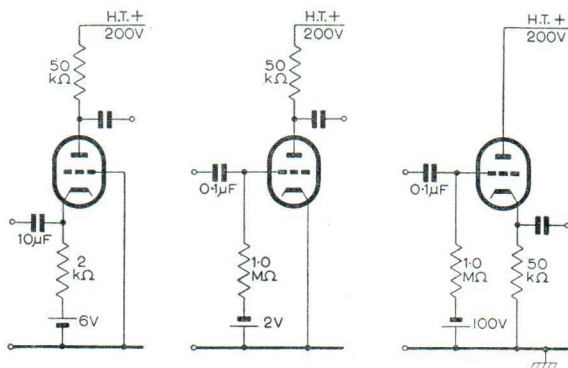


Fig. 1—Grounded grid
Fig. 2—Grounded cathode
Fig. 3—Cathode follower

the input is applied to the cathode. In the second, more common arrangement, the input is applied to the grid and the cathode usually grounded to a.c. Finally there is the cathode follower.

There are similarly three a.c. circuit configurations for the transistor (Figs. 4, 5 and 6). They are usually known as grounded base, grounded emitter and grounded collector, according to which electrode is common to the input and output circuits.

Table 1, at the end of the chapter (p. 43), summarises the characteristics of the three transistor configurations, and is useful when choosing the one required for any particular purpose.

GROUNDING BASE

The grounded-base or common-base circuit (Fig. 4) gives the best illustration of how a transistor works, though it is not the most frequently used. The input is applied to the emitter and the output taken

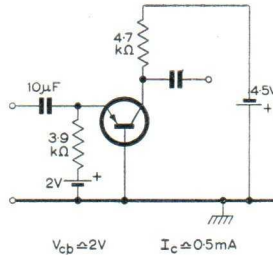


Fig. 4—Grounded base

from the collector. The base is common to the input and output circuits, and is normally grounded to a.c.

Input Impedance

The input impedance of the transistor in grounded base is

$$Z_{in} = z_{11} \cdot \frac{R_L + z_{out}}{R_L + z_{22}}$$

From the definitions of the modified z parameters, the input impedance lies between the extreme values of z_{11} (output open circuit) and z_{in} (output shorted). Thus if in the above expression $R_L \rightarrow \infty$, then

$$Z_{in} \rightarrow z_{11}$$

and if $R_L = 0$,

$$Z_{in} = \frac{z_{11}z_{out}}{z_{22}} = z_{in}$$

since

$$z_{out}/z_{22} = z_{in}/z_{11}.$$

For a typical OC71 at a working point of $-2V$, $1mA$,

$$z_{in} = 35\Omega \text{ and } z_{11} = 720\Omega.$$

The input impedance therefore varies widely with the load resistance and in this particular example, by as much as 20:1.

Output Impedance

The output impedance in grounded base is

$$Z_{out} = z_{22} \cdot \frac{R_s + z_{in}}{R_s + z_{11}}$$

The extreme value of output impedance is z_{22} for $R_s \rightarrow \infty$ (input open circuit). For $R_s = 0$ (input shorted), the extreme value is

$$\frac{z_{22}z_{1n}}{z_{11}} = z_{out}.$$

The variation of output impedance with source resistance is the same as the variation of input impedance with load resistance, since

$$\frac{z_{1n}}{z_{11}} = \frac{z_{out}}{z_{22}}.$$

For the typical OC71 and a working point of $-2V$, $1mA$,

$$z_{22} = 1.0M\Omega \text{ and } z_{out} = 50k\Omega.$$

The variation is therefore 20:1, as expected.

The ratio of output to input impedance is high, thus

$$\frac{z_{out}}{z_{in}} = \frac{z_{22}}{z_{11}} \simeq 1400.$$

Current Gain

The current gain in grounded base, in terms of the modified z parameters, is:

$$A_i = \alpha \cdot \frac{z_{22}}{R_L + z_{22}}.$$

If the load resistance R_L is made equal to zero (short-circuited output), the current gain becomes equal to α .

As R_L is increased, the current gain becomes progressively smaller. When R_L is infinite (open-circuited output), A_i is zero.

For an alloy-junction transistor, α is slightly less than one. The current gain in grounded base is therefore always less than one.

Current Amplification Factor

α is the current amplification factor of the transistor in grounded base. This is the maximum theoretical current gain.

The condition $A_i = \alpha$ when $R_L = 0$ (output shorted to a.c., $v_2 = 0$) agrees with the definition of α given in the previous chapter.

Voltage Gain

The voltage gain in grounded base is

$$A_v = \frac{\alpha z_{22}}{z_{11}} \cdot \frac{R_L}{R_L + z_{out}}.$$

The factor $R_L/(R_L + z_{out})$ approaches one as R_L is increased; the voltage gain increases as R_L increases, and reaches a theoretical maximum when R_L is infinite. A_v is zero when R_L is zero.

Although α is typically about 0.98, the ratio of the output impedance to input impedance (z_{22}/z_{11}) is high, and so is the possible voltage gain.

Voltage Amplification Factor

The maximum voltage gain, with infinite load ($R_L \rightarrow \infty$) is

$$\mu_{21} = \left(\frac{v_2}{v_1} \right)_{i_2 = 0} = \frac{\alpha z_{22}}{z_{11}}.$$

μ_{21} is the voltage amplification factor, and is a function of the transistor only.

The definition of μ_{21} is similar to that of the voltage amplification factor μ of the triode valve. However, for transistors, a μ without subscript is usually reserved for μ_{12} , the voltage feedback factor ($\approx h_{12}$).

For a typical OC71 operating at $-2V$, $1mA$:

$$\alpha = 0.976, \quad z_{22} = 1.0M\Omega, \quad \text{and} \quad z_{11} = 720\Omega.$$

Hence the voltage amplification factor (or maximum voltage gain) is

$$\mu_{21} = \frac{\alpha z_{22}}{z_{11}} = \frac{0.976 \times 10^6}{720} \simeq 1400.$$

Power Gain

The power gain A_w is the product of the current gain and the voltage gain, so that in grounded base

$$\begin{aligned} A_w &= A_i \times A_v \\ &= \frac{(\alpha z_{22})^2}{z_{11}} \cdot \frac{R_L}{(R_L + z_{22})(R_L + z_{out})}. \end{aligned}$$

The grounded-base circuit derives its power gain from its high voltage gain.

The power gain is a maximum when

$$R_L = \text{optimum } Z_{out} = \sqrt{(z_{22} z_{out})},$$

as can be shown by differentiating the expression for A_w with respect to R_L .

The source resistance which gives the optimum output impedance is found from

$$R_s = \text{optimum } Z_{in} = \sqrt{(z_{11} z_{in})}.$$

With the above values of R_L and R_s , the matched power gain is

$$A_{w\max} = z_{22} \left(\frac{\alpha}{\sqrt{z_{11}} + \sqrt{z_{in}}} \right)^2.$$

For a typical OC71 at a working point of $-2V$, $1mA$, the modified z parameters in grounded base are:

$\alpha = 0.976$; $z_{in} = 35\Omega$; $z_{11} = 720\Omega$; $z_{out} = 50k\Omega$; and $z_{22} = 1.0M\Omega$. Consequently the matched power gain at this working point is

$$10^6 \times \left(\frac{0.976}{\sqrt{720} + \sqrt{35}} \right)^2 = 890 = 29\text{dB}.$$

This power gain is obtained with

$$R_L = \sqrt{(z_{22}z_{out})} = \sqrt{(5 \times 10^{10})} = 224k\Omega$$

and

$$R_s = \sqrt{(z_{11}z_{in})} = \sqrt{(720 \times 35)} = 160\Omega.$$

The maximum power gain is only obtainable with transformer coupling, a step-down ratio being required of approximately

$$\sqrt{\frac{R_L}{R_s}} = \sqrt{\frac{224}{0.16}} = \sqrt{1400} = 37,$$

when feeding into an identically similar stage.

GROUNDING EMITTER

Grounded-emitter connection (Fig. 5) is the most commonly used of the three configurations. The input is applied to the base, the output is taken from the collector, and the emitter is common to the input and

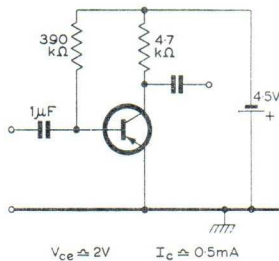


Fig. 5—Grounded emitter

output. It is sometimes called common-emitter connection, because the emitter is not necessarily grounded to a.c.

Using the modified z parameters, the performance equations are exactly the same as for grounded base, except that the quantities are dashed or primed to show that they are measured in grounded emitter. The equations will be repeated in their primed form, so that they can be associated with the values of the parameters in grounded emitter.

Input Impedance

The input impedance in grounded emitter is given by:

$$Z'_{in} = z'_{11} \cdot \frac{R_L + z'_{out}}{R_L + z'_{22}}$$

Putting $R_L \rightarrow \infty$,

$$Z'_{in} \rightarrow z'_{11}$$

and putting $R_L = 0$,

$$Z'_{in} = \frac{z'_{11} z'_{out}}{z'_{22}} = z'_{in},$$

as required by the definitions.

For a typical OC71 at $-2V$, $1mA$:

$$\alpha' = 41, \quad z'_{11} = 720\Omega, \quad \text{and} \quad z'_{in} = 1.45k\Omega.$$

There is therefore comparatively little variation in input impedance with load resistance in grounded emitter ($z'_{in}/z'_{11} = 2$).

From the above values, it can be confirmed that $z'_{11} = z_{11}$ and, within the errors of measurement, $z'_{in} = z_{in}(1 + \alpha')$.

Output Impedance

The output impedance in grounded emitter is

$$Z'_{out} = z'_{22} \cdot \frac{R_s + z'_{in}}{R_s + z'_{11}},$$

which lies between the extreme values z'_{22} ($R_s \rightarrow \infty$) and

$$\frac{z'_{22} z'_{in}}{z'_{11}} = z'_{out}$$

($R_s = 0$), as required by the definitions.

For the typical OC71 as before, z'_{22} is $25k\Omega$ and z'_{out} is $50k\Omega$. Again the variation is 2:1.

From the above values, it can be confirmed that $z'_{out} = z_{out}$ and $z'_{22} = z_{22}/(1 + \alpha')$.

The ratio of output to input impedance is medium, thus

$$\frac{z'_{out}}{z'_{in}} = \frac{z'_{22}}{z'_{11}} \simeq 35.$$

Current Gain and Current Amplification Factor α'

The current gain in grounded emitter is

$$A_I = \alpha' \cdot \frac{z'_{22}}{R_L + z'_{22}}.$$

As before, the current gain is a maximum when R_L is zero, and decreases to zero as R_L is increased to infinity.

The maximum theoretical current gain α' , which is typically 40 to 50 for the OC71, is much higher than in grounded base. α' is the

current amplification factor of the transistor in grounded emitter.

The current-gain equation agrees with the earlier definitions of α' , in that $A_1 = \alpha'$ only when $R_L = 0$.

Voltage Gain and Voltage Amplification Factor

The voltage gain in grounded emitter is

$$A_v = \frac{\alpha' z'_{22}}{z'_{11}} \cdot \frac{R_L}{R_L + z'_{out}}$$

The maximum voltage gain, when R_L approaches infinity, is

$$\frac{\alpha' z'_{22}}{z'_{11}},$$

which is the voltage amplification factor in grounded emitter. This value is exactly the same as in grounded base, since

$$z'_{11} = z_{in} \text{ and } \alpha' z'_{22} = \alpha z_{22}.$$

Thus the maximum voltage gain ($= \mu_{21}$) in grounded emitter, as in grounded base, is approximately equal to 1400 for the example quoted.

Power Gain

The power gain in grounded emitter is much higher than in grounded base because, although the voltage gain is the same, the current gain is increased. The power gain is calculated from

$$A_w = \frac{(\alpha' z'_{22})^2}{z'_{11}} \cdot \frac{R_L}{(R_L + z'_{22})(R_L + z'_{out})}$$

For an OC71 at a working point of $-2V$, $1mA$ as before, the parameters are:

$$\alpha' = 41; \quad z'_{in} = 1.45k\Omega; \quad z'_{11} = 720\Omega; \quad z_{out} = 50k\Omega; \\ \text{and } z'_{22} = 25k\Omega.$$

The maximum power gain is

$$z'_{22} \left(\frac{\alpha'}{\sqrt{z'_{11}} + \sqrt{z'_{in}}} \right)^2 = 25,000 \left(\frac{41}{\sqrt{720} + \sqrt{1450}} \right)^2 \\ = 40dB.$$

This gain is obtained with:

$$R_L = \sqrt{z'_{22} z'_{out}} = \sqrt{(25 \times 50)} = 35k\Omega$$

and

$$R_s = \sqrt{z'_{11} z'_{in}} = \sqrt{(720 \times 1450)} = 1.02k\Omega.$$

The transformer turns ratio is approximately

$$\sqrt{\frac{R_L}{R_s}} = \sqrt{\frac{35}{1.02}} = \sqrt{34.3} = 5.9,$$

when feeding into an identically similar stage.

GROUNDING COLLECTOR

Grounded-collector connection (Fig. 6) corresponds to the 'cathode follower' valve circuit. It is sometimes for this reason referred to as the 'emitter follower'. It is also referred to as common collector.

The input is applied to the base, the output is taken from across the load in the emitter, and the collector is common to the input and output.

The current amplification factor for small signals, which is given the symbol α'' , is approximately equal to α' . The exact relationship is

$$\alpha'' = 1 + \alpha'.$$

In this configuration, the voltage gain cannot exceed one and the power gain is lowest.

There is high input and low output impedance, and these impedances are more dependent on the load and source resistances than in the

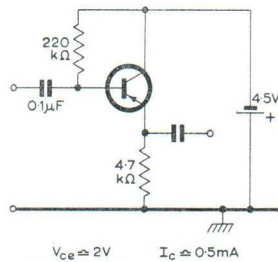


Fig. 6—Grounded collector

other configurations. Over a wide range of values the grounded-collector circuit can be considered as an impedance changer, the input impedance being approximately $\alpha' R_L$ and the output impedance R_s/α' .

The performance will not be calculated for grounded collector, as this circuit is little used for small-signal stages.

COMPARISON OF CONFIGURATIONS

The performance of the various configurations can be compared by taking the OC71 as an example. Table 2, at the end of the chapter (p. 43), summarises the typical performance of the OC71 for a working point of $-2V$, $1mA$. It is another version of Table 1 into which values have been inserted.

Maximum Current Gain

The current amplification factor in grounded emitter, α' , is 41 and the value in grounded base, α , is calculated as 0.976 from

$$\alpha' = \frac{\alpha}{1 - \alpha}$$

or explicitly

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

For grounded collector $\alpha'' = 1 + \alpha' = 42$.

The typical value of α' becomes 47 when the collector current is increased to 3mA.

Matched Input and Output Impedances

The input and output impedances are reduced appreciably when the collector current is increased from 1mA to, say, 3mA. Thus at 3mA, $z'_{in} = 800\Omega$, $z'_{11} = 500\Omega$, $z'_{out} = 21k\Omega$ and $z'_{22} = 12.5k\Omega$. The matched input impedance is then 630 Ω and the matched output impedance 16.2k Ω in grounded emitter.

Matched Power Gain

The power gain is highest in grounded emitter; useful power gain is also available in grounded base, but in grounded collector the power gain is rather low.

For most purposes the power gain is of prime importance when choosing the circuit configuration, and the grounded-emitter configuration is therefore most frequently used, and grounded collector least.

In none of the configurations is the power gain critically dependent on load resistance. In grounded emitter the reduction from 40dB to 37dB takes place at about 8k Ω at one extreme and 150k Ω at the other. The gain in grounded base is reduced from 29dB to 26dB at about 50k Ω and 1M Ω . In grounded collector the 16dB gain becomes 13dB at about 30 Ω and 20k Ω .

The matched power gain at 3mA is still 40dB in grounded emitter, as the increase in α' and the fall in input impedance compensate for the fall in output impedance.

Cut-off Frequency

The current amplification factor of the transistor falls as the frequency is increased. The point at which the fall-off becomes pronounced depends mainly on the type of transistor and also on the circuit configuration.

The cut-off frequency is the point at which the current amplification factor falls to 3dB below the low-frequency or 'zero frequency' value; 3dB corresponds to $1/\sqrt{2}$, that is, 0.707. The symbol for the cut-off frequency is f_α for grounded base and f'_α for grounded emitter.

Grounded-base connection gives the best h.f. performance, having a higher cut-off frequency (f_α) than in grounded emitter (f'_α) and

grounded collector ($\sim f'_\alpha$). The OC71 illustrated in Table 2 is only an a.f. transistor. For the OC45, f_α is typically 6Mc/s, and for the OC44, 15Mc/s.

The maximum frequency at which the transistor will continue to give satisfactory performance in any circuit is not necessarily limited to the cut-off frequency. Apart from spread in transistor characteristics, which means that any given transistor may have a better or worse performance than the typical, the choice of a reduction of 3dB to define the cut-off frequency is arbitrary. Nevertheless, the cut-off frequency usually gives a good indication of the suitability or otherwise of a transistor for h.f. performance, and is useful for the purpose of comparison.

RC AND TRANSFORMER COUPLING

A transistor in grounded base will not give useful voltage gain, if it is RC coupled into another grounded-base stage. Since the current gain is less than one, there is no useful power gain.

The low input impedance of the following stage virtually shorts the collector coupling resistor to a.c. Hence in the expression for voltage gain we can write $R_L = z_{in}$. Then since $\alpha \sim 1$,

$$\begin{aligned} A_v &= \frac{z_{22}}{z_{11}} \cdot \frac{z_{in}}{z_{in} + z_{out}} \\ &= \frac{z_{out}}{z_{in} + z_{out}} \\ &< 1. \end{aligned}$$

For multistage amplification using RC coupling, the grounded-base circuit must be combined with grounded-emitter or grounded-collector stages.

The above objection does not apply to grounded emitter, because with an α' of 40 or more, both the current and voltage gain are greater than one.

With transformer coupling there is no restriction on the cascading of grounded-base stages. The impedance ratio in grounded emitter, however, leads to more convenient turns ratios, apart from the higher matched power gain in this configuration.

APPLICATIONS

The grounded-emitter configuration gives the highest power gain and is normally used in a.f. amplifiers.

Grounded-base stages are not often used at a.f., but sometimes the low input and high output impedances are of value. The low input impedance is useful for pre-amplifiers for use with moving-coil micro-

phones, and the high output impedance for feeding into valve amplifiers.

The grounded-collector configuration is useful where high input and/or low output impedances are important. It is used in buffer stages and can sometimes replace a transformer.

Despite the higher cut-off frequency in grounded base, this mode of connection is not necessarily preferred at high frequencies. Practical i.f. amplifiers, for example, can be designed for medium and long waves using the OC45 in grounded emitter.

Grounded-base connection is nevertheless frequently exploited in oscillator and switching circuits.

TABLE 1
COMPARISON OF CIRCUIT CONFIGURATIONS

	<i>Grounded Base</i>	<i>Grounded Emitter</i>	<i>Grounded Collector</i>
Current gain	~ 1	High	High
Voltage gain	High	High	~ 1
Input impedance	Low	Medium	High
Output impedance	High	Medium	Low
Power gain	Medium	High	Low
Cut-off frequency	High	Low	Depends on R_L
Voltage phase shift at low frequencies	$\sim$ Zero	$\sim 180^\circ$	$\sim$ Zero

TABLE 2
CIRCUIT CONFIGURATIONS USING OC71
Collector Voltage $-2V$, Collector Current $1mA$

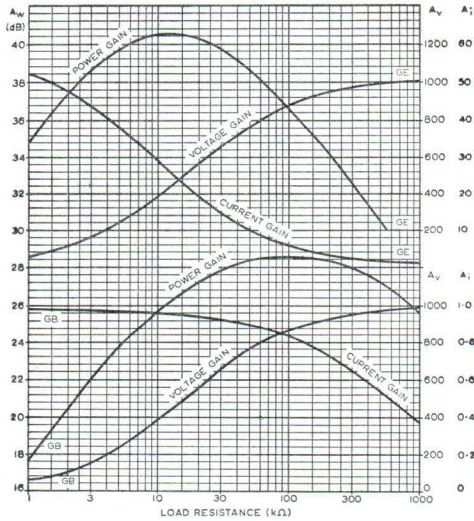
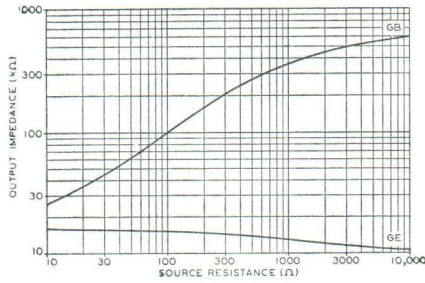
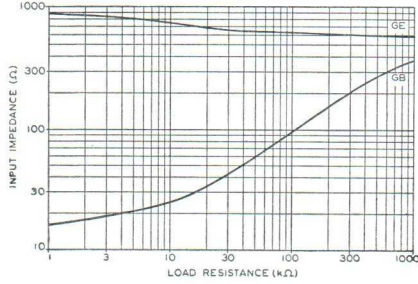
	<i>Grounded Base</i>	<i>Grounded Emitter</i>	<i>Grounded Collector</i>
Max. current gain ($R_L = 0$)	0.976	41	42
Max. voltage gain ($R_L \rightarrow \infty$)	1400	1400	1
Matched input impedance ($k\Omega$) ..	0.16	1.02	38
Matched output impedance ($k\Omega$) ..	224	35	0.93
Matched power gain (dB)	29	40	16
Cut-off frequency (kc/s)	600	11	11

PERFORMANCE OF PARTICULAR OC71 IN GROUND BASE (GB) AND GROUND Emitter (GE)

$$V_e = -2V, I_e = 3mA; z_{11} = z'_{11} = 590\Omega,$$

$$z_{out} = z'_{out} = 16k\Omega, \alpha = 0.982, \alpha' = 56;$$

$$z_{1n} = 16\Omega, z_{22} = 590k\Omega; z'_{1n} = 910\Omega, z'_{22} = 10.4k\Omega$$



Power gain A_w (dB), voltage gain A_v and current gain A_i

CHAPTER SIX

TRANSISTOR AS A CURRENT NETWORK

In Chapters 4 and 5 the transistor has been treated on a four-pole basis. Even the static characteristic curves described in Chapter 3 can be regarded as being derived from a black box.

This treatment has the advantage of allowing the performance of the transistor in circuit to be calculated, without making any assumptions about the interior mechanism of the device.

For a more complete grasp of transistor operation, and in particular in order to understand the biasing arrangements to be described in Chapter 7, it is necessary to regard the transistor as something more than a set of input and output terminals. Fortunately all the extra information required can be derived by treating the transistor as a simple current network.

CONSERVATION OF CURRENT

Use has already been made of the example of a transistor operating at an emitter current of 1mA; if the collector current is 0.98mA, then the base current must be equal to the difference, 0.02mA. Similar calculations are made on thermionic valves, the screen-grid current of a pentode, for example, normally being calculated from the known currents at the other electrodes.

Several important relationships can be derived from this principle, provided it is expressed in a form which is easier to manipulate.

The direct current flowing into the transistor is equal to that flowing out, or in symbols

$$I_e + I_b + I_c = 0. \quad \dots (1)$$

This equation is a restatement of Kirchhoff's law, which gives, for a three-terminal network:

$$I_1 + I_2 + I_3 = 0.$$

A similar equation holds for alternating currents, namely:

$$i_e + i_b + i_c = 0. \quad \dots (1a)$$

The usual sign convention is followed, according to which current flowing from + to - is considered positive. In practice it is usually only necessary to consider the sign (direction) of the base current.

RELATION OF α TO α'

The relationship between the various currents remains the same irrespective of the transistor configuration. The difference between α and α' arises solely from the choice of input electrode, and a definite numerical relationship exists between them.

From the definitions of α and α' , where no regard is paid to the direction of current flow,

$$\frac{\alpha}{\alpha'} = \frac{\delta i_b}{\delta i_e}.$$

Also, Eq. 1a can be written numerically as

$$i_b = i_e - i_c,$$

so that

$$\delta i_b = \delta i_e - \delta i_c.$$

Hence

$$\frac{\alpha}{\alpha'} = \frac{\delta i_b}{\delta i_e} = 1 - \frac{\delta i_c}{\delta i_e} = 1 - \alpha.$$

Rearranging this equation in its more usual form gives

$$\alpha' = \frac{\alpha}{1 - \alpha}.$$

Note that

$$1 + \alpha' = \frac{1}{1 - \alpha} \quad \text{and} \quad \alpha = \frac{\alpha'}{1 + \alpha'}.$$

REVERSAL OF BASE CURRENT

Consider the example given earlier, where the emitter current is 1mA. The collector current is about 0.98mA, and the base current is equal numerically to the difference between these, namely, 0.02mA.

The signs of these currents will be as follows, for a p-n-p transistor: I_e will be positive because it flows from + to - into the transistor; I_c will be negative because it flows from + to - out of the transistor; and I_b will therefore be negative because, from Eq. 1,

$$1.0 - 0.98 - 0.02 = 0.$$

Thus at normal operating currents and temperatures I_b is negative, and flows from + to -, out of the transistor.

To understand the operation of the transistor, it is necessary to consider what happens at very low currents. In fact, to take the extreme case, consider the emitter open-circuited ($I_e = 0$). A leakage current I_{co} then flows from base to collector. The base current in this condition

flows from + to - into the transistor, and is positive.

As the emitter current is increased from zero, the base current therefore reverses at some point from positive to negative.

This effect can be explained by means of Fig. 1, which shows a plot of collector current against emitter current. The full line shows the behaviour of a normal transistor. When $I_e = 0$, $I_c = I_{co}$, as required. As the emitter current is increased by an amount ΔI_e , the

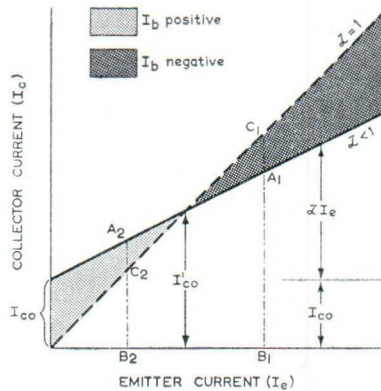


Fig. 1—Relation of transfer characteristic in grounded base to transistor parameters.

collector current increases by the fraction $\Delta I_c = \bar{\alpha} \times \Delta I_e$. The full line has a slope of $\bar{\alpha} < 1$. ($\bar{\alpha}$ is the value of α for large current changes.)

The broken line is drawn at an angle of 45° to the two axes and represents the condition in which the collector and emitter currents are equal. This line would correspond to a transistor in which $\bar{\alpha} = 1$ and $I_{co} = 0$.

At the point where the two lines intersect, $I_c = I_e$ and $I_b = 0$. When $I_b = 0$, the base is open circuit, and the collector and emitter currents are both equal to I'_{co} , the collector leakage current in grounded emitter.

Now examine the point A_1 on the full line ($\bar{\alpha} < 1$) at a current $I_c > I'_{co}$. The collector current at this point is A_1B_1 and the emitter current is C_1B_1 . The base current is equal to the difference A_1C_1 . In this region the emitter current is numerically greater than the collector current, and as I_e is positive, the base current is of the same sign as I_e , that is, negative (Eq. 1).

The point A_2 lies in the region where $I_c < I'_{co}$. The collector current is A_2B_2 , the emitter current C_2B_2 , and the base current equal to the difference A_2C_2 . Here the collector current is numerically greater than the emitter current, and the base current is positive in sign, like I_e .

The transistor can only be operated below I'_{co} by reversing the base current from its normal negative direction to the positive direction.

The current reversal can also be regarded from the point of view of the current network shown in Fig. 2. If we put $I_e = 0$, the leakage current I_{co} flows from base to collector. Normally, the current $\bar{\alpha}I_e$

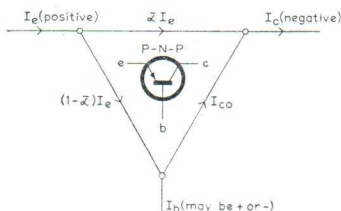


Fig. 2—Transistor currents in terms of $\bar{\alpha}$, I_e and I_{co} .

flows from emitter to collector. The current flowing from emitter to base is the difference between the currents flowing in the other two branches, namely

$$I_e - \bar{\alpha}I_e = (1 - \bar{\alpha})I_e.$$

The base current will therefore be positive, zero or negative depending on whether I_{co} is greater than, equal to, or less than $(1 - \bar{\alpha})I_e$.

The sign of the base current is sometimes important, because of the need to reverse this current. The signs of the collector and emitter currents do not need to be distinguished for practical purposes. The collector and base currents are no longer shown as negative in the published data.

LARGE-SIGNAL CURRENT AMPLIFICATION FACTOR

In audio output stages, driver stages, and switching circuits, the transistors are required to handle large changes of current. The current amplification factors for large signals are distinguished by adding a bar to α and α' , thus $\bar{\alpha}$ (alpha bar) and $\bar{\alpha}'$ (alpha dash bar).

An alternative symbol for $\bar{\alpha}'$ is h_{FE} , the subscript F denoting forward transfer and the subscript E showing that the emitter is the common electrode (grounded-emitter connection). Capital letters are used in the subscript to indicate that the currents under consideration are direct and can be read from the static characteristic curves. $\bar{\alpha}'$ may also be represented by β .

The alternative symbol for $\bar{\alpha}$ is h_{FB} .

The large-signal current amplification factor is defined with particular reference to switching applications. The transistor is considered to be OFF when the emitter current is zero. In this condition, the leakage current or cut-off current I_{co} continues to flow as a positive base current from base to collector. Suppose now that the transistor is switched ON to some collector current I_c . The change in collector current is numerically equal to $I_c - I_{co}$.

The base current I_b in the ON condition flows in the normal negative direction, out of the transistor. The change in base current is numeri-

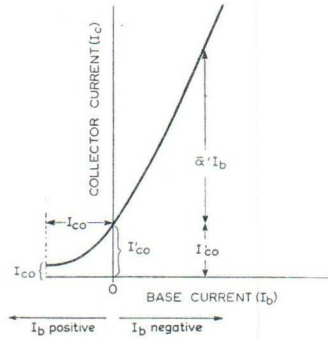


Fig. 3—Relation of transfer characteristic in grounded emitter to transistor parameters.

cally equal to $I_b + I_{co}$. Hence the large-signal current amplification factor in grounded-emitter is

$$\begin{aligned}\bar{\alpha}' &= \frac{\Delta I_c}{\Delta I_b} \\ &= \frac{I_c - I_{co}}{I_b + I_{co}},\end{aligned}$$

where Δ represents a large change. This definition is illustrated graphically by Fig. 3.

It is not usual to work in terms of $\bar{\alpha}$, but this quantity may be defined as

$$\begin{aligned}\bar{\alpha} &= \frac{\Delta I_c}{\Delta I_e} = \frac{I_c - I_{co}}{I_e} \\ &= \frac{I_c - I_{co}}{I_c + I_b},\end{aligned}$$

where the transistor is OFF, as before, when the emitter current is zero.

RELATION OF I'_{co} TO I_{co}

When the base is open circuit and $I_b = 0$, we have

$$I_c = I_e = I'_{co}$$

and, from Fig. 2,

$$(1 - \bar{\alpha})I_e = I_{co}.$$

Hence

$$\begin{aligned} I'_{co} &= \frac{I_{co}}{1 - \bar{\alpha}} \\ &= (1 + \bar{\alpha}')I_{co} \\ &\simeq \bar{\alpha}'I_{co}. \end{aligned}$$

This relationship is only true provided the value of $\bar{\alpha}'$ inserted in the

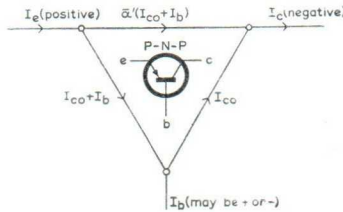


Fig. 4—Transistor currents in terms of $\bar{\alpha}'$, I_b and I'_{co}

equation is correct for very low collector currents. This value is considerably lower than at normal operating currents.

Fig. 1 is not to scale because, in order to make the stippled areas of appreciable size, I_{co} had to be represented as a much larger fraction of I'_{co} than it is in practice. Values are likewise not given on the scales of Fig. 3, and this figure too is not to scale.

FUNDAMENTAL TRANSISTOR EQUATION

From Fig. 2 it follows directly that, numerically,

$$I_c = I_{co} + \bar{\alpha}I_e. \quad \dots (2)$$

This is the fundamental equation for the transistor, and is true in all configurations. Eq. 2 is the equation to the straight line of Fig. 1.

Sometimes it is more convenient to express the collector current in terms of I'_{co} , $\bar{\alpha}'$ and I_b ; thus, numerically:

$$\begin{aligned} I_c &= I_{co} + \bar{\alpha}'(I_{co} + I_b) \\ &= I'_{co} + \bar{\alpha}'I_b. \end{aligned} \quad \dots (2a)$$

This equation corresponds to the diagrams shown in Figs. 3 and 4.

TEMPERATURE EFFECTS

The collector leakage currents I_{co} and I'_{co} , the current amplification factors α and α' , and the base-emitter voltage V_{be} are all temperature dependent to a greater or lesser extent.

Temperature dependence is greatest for the leakage currents. For a rise in junction temperature from 25°C to 45°C, I_{co} increases by a

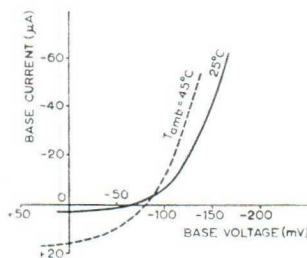
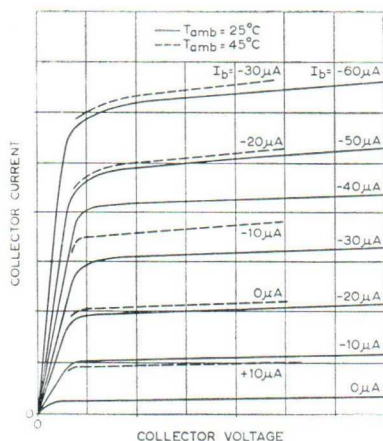


Fig. 5—Effect of temperature on output characteristic in grounded emitter

Fig. 6—Effect of temperature on input characteristic in grounded emitter

factor of five, and I'_{co} by a factor of eight, still considering the OC71. α' increases by about 1% per degree Centigrade.

V_{be} decreases by roughly 2mV per degree Centigrade. This rate of change is almost constant for all transistors (germanium and silicon).

The effect of temperature on the output and input characteristics is shown in Figs. 5 and 6.

In Fig. 5, the full lines apply to an ambient temperature of 25°C. The collector current is equal to the leakage current I'_{co} (at $I_b = 0$) plus the current produced by biasing the transistor from $I_b = 0$ to, say, $I_b = -30\mu A$. When the ambient temperature is increased from 25°C to 45°C, the increase in leakage current pushes the curves upwards relative to the I_c scale.

Consequently the curve at 45°C for $I_b = -30\mu A$ (broken line) lies above that at 25°C for $I_b = -60\mu A$ (full line).

Fig. 6, the input characteristic, shows the possibility of reversing the base current. The effect of this reversal is shown on the output characteristic by the broken curve for $I_b = +10\mu\text{A}$. Reversing the base current allows the grounded-emitter transistor to operate at low collector currents, even at high temperatures where I'_{co} reaches a high value.

SPREADS

Transistor characteristics, like those of the thermionic valve, exhibit production spreads. Every endeavour is made to keep the spreads as small as possible.

Spreads are of most importance when considering grounded-emitter operation. For the OC71, the collector leakage current I'_{co} , which is

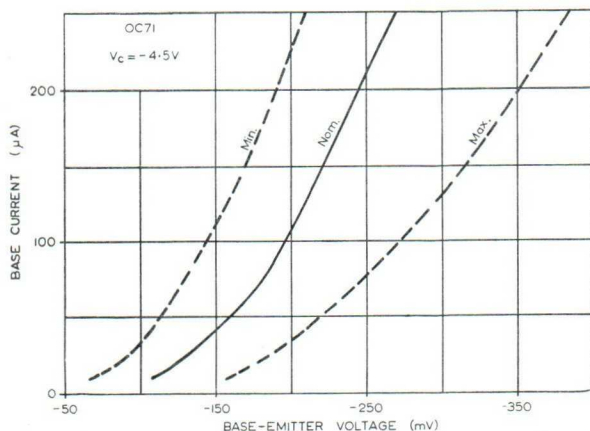


Fig. 7—Spread in base-emitter voltage of OC71

nominally $150\mu\text{A}$, may attain an extreme upper limit of $325\mu\text{A}$. These figures apply at an ambient temperature of 25°C . At 45°C , the leakage current in the extreme case may be as much as $8 \times 0.325 = 2.5\text{mA}$.

A fairly wide spread also occurs in α' and $\bar{\alpha}'$. A spread of only a few per cent in α introduces a much wider spread into the factor $1-\alpha$ and hence into α' , since

$$\alpha' = \frac{\alpha}{1-\alpha}.$$

A nominal α' of 47 for the OC71 (at a collector current of 3mA) corresponds to a maximum of 75, the spread $\Delta\alpha'/\alpha'$ therefore being $(75-47)/47 \simeq 0.5$. A similar spread applies to $\bar{\alpha}'$.

The spread in base-emitter voltage for the OC71 is shown in Fig. 7.

CHAPTER SEVEN

BIAS AND STABILISATION

In thermionic-valve circuits, the normal cathode resistor reduces the influence of valve spreads on the position of the working point. Although there are a number of ways of introducing d.c. feedback into transistor circuits, the best of these are a refinement from the idea of using an emitter resistor. For transistors, d.c. stabilisation is particularly important, because temperature effects are capable of introducing a much wider variation in collector current than that due to spreads alone.

The transistor should be biased by a method which prevents excessive shift of the d.c. working point. Insufficient d.c. stabilisation can give rise to the following effects:

- (a) Wide spread in input and output impedances.
- (b) Risk of overloading ('bottoming') at high ambient temperatures.
- (c) Possibility of thermal runaway. (This effect is normally important only in high-voltage and/or high-power stages.)

The biasing circuits which follow are described as grounded base, grounded emitter or grounded collector, but this description applies to

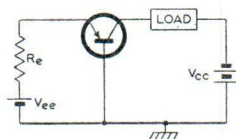


Fig.1—Grounded-base circuit

the d.c. conditions only, and does not prevent the use of the circuit in a different a.c. configuration. Any electrode can be grounded to a.c. by means of a large capacitance, or open-circuited to a.c. by means of inductance, without affecting the d.c. conditions.

GROUNDLED-BASE CIRCUIT

In the grounded-base configuration (Fig. 1), the base is biased with constant emitter current, by making the biasing voltage V_{ee} large in comparison with the transistor input voltage V_{eb} .

This is the most stable arrangement. It gives the least change in

collector current when transistors are replaced, and the least increase in current with temperature.

The two equations governing the performance are:

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

and

$$I_e = \frac{V_{ee} - V_{eb}}{R_e}$$

Strictly speaking, the large-signal value $\bar{\alpha}$ should be inserted in the first equation, but in practice, with small-signal transistors, little error is likely to arise in substituting the zero-frequency value α_0 .

Assuming constant emitter current I_e , the spread in collector current I_c will arise from the small spread in α_0 , which is normally less than 4%. At higher temperatures, the increase in the leakage current I_{co} may become apparent, depending on the relative magnitude of I_c and I_{co} . A transistor operating at a relatively high collector current will be less affected by temperature.

In practice, it may not be possible to satisfy the condition in which the bias voltage V_{ee} is very much greater than the input voltage V_{eb} , and the spread in V_{eb} and its change with temperature may influence I_e .

SIMPLE GROUND-EMITTER CIRCUIT

In the simple grounded-emitter circuit (Fig. 2), the emitter is common to the input and output circuits. The transistor is biased with constant base current, as the battery voltage V_{cc} is large in comparison with the input voltage V_{be} .

This is the simplest arrangement in that it requires only one battery and one resistor. Unfortunately the d.c. stability is poor. There is a

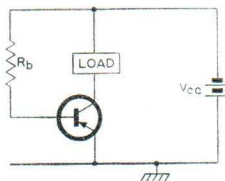


Fig. 2—Simple grounded-emitter circuit

large change in collector current on replacing transistors and a large increase in collector current with temperature, and there may be thermal runaway in high-voltage or high-power stages. Unless the intended application is extremely uncritical of these effects, this method of biasing is generally unsuitable.

The performance equations are:

$$I_c = I'_{co} + \alpha' I_b$$

and

$$I_b = \frac{V_{cc} - V_{be}}{R_b}$$

The disadvantages of this method of biasing are, firstly, the collector current must always lie above I'_{co} , which may become prohibitively high at high temperatures. Secondly, there is a wide spread in α' over a batch of transistors, and α' itself increases with temperature. Since V_{be} is normally small in comparison with V_{cc} , the spread in V_{be} and its change with temperature are usually of only secondary importance.

GROUND-EMITTER CIRCUIT WITH FEEDBACK RESISTOR

A collector-base feedback resistor (Fig. 3) is the simplest method of including some d.c. stabilisation in a grounded-emitter circuit. The

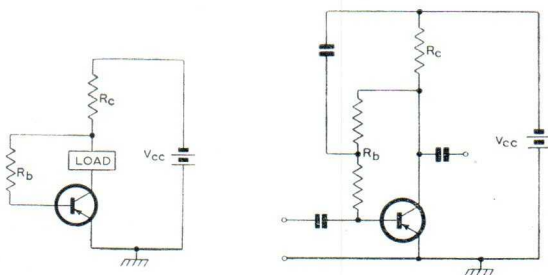


Fig. 3—Grounded-emitter circuit with feedback resistor

Fig. 4—Bypassing of feedback resistor to prevent a.c. feedback

base resistor R_b is returned to the collector end of R_c , instead of to the battery. The value of R_b is roughly equal to V_c/I_b . Any increase in collector current causes a drop in collector voltage, and hence reduces the current flowing through R_b to the base, so compensating partly for the original change.

With RC coupling, the collector resistor itself supplies the d.c. feedback, and no additional components are required. There will also be a.c. feedback, unless decoupling is used as shown in Fig. 4, where R_b is made up of two approximately equal resistances.

This arrangement provides some d.c. stabilisation, but again the collector current must always lie above I'_{co} . In the limiting case, where the increase in ambient temperature causes the stage to bottom, the voltage across R_b will be zero, the base current zero, and the collector current equal to I'_{co} .

If the collector current in the circuit tends to change by ΔI_c , the d.c. feedback will reduce this change to $K \times \Delta I_c$ ($K < 1$), where

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

K is the 'factor of stability' for the circuit.

A small value of K gives the best stability, and may be achieved by making R_c large or R_b small, the former implying a high battery voltage, and the latter a high collector current.

By rearranging the circuit, it may be used for grounded-collector

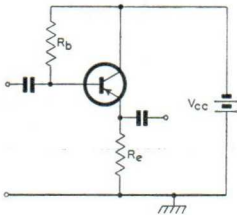


Fig. 5—Grounded-emitter circuit for d.c. arranged in grounded collector to a.c.

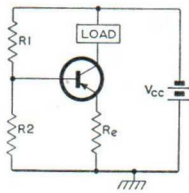


Fig. 6—Emitter-resistor and potential-divider circuit

operation, as shown in Fig. 5. Exactly the same equations apply, except that R_c must be replaced by R_e .

In practice this circuit gives a K of about 0.7 or 0.8, so the degree of stabilisation is not great. Nevertheless it may be sufficient for some applications.

EMITTER-RESISTOR AND POTENTIAL-DIVIDER CIRCUIT

In the emitter-resistor and potential-divider circuit (Fig. 6), the emitter resistor introduces negative d.c. feedback. The input voltage V_{be} is determined by the emitter resistor, in conjunction with the potential divider R_1 – R_2 connected across the battery.

This arrangement is by far the most commonly used, as the designer has most control over the stabilisation, whilst only one battery is required.

Any increase in emitter current causes a large voltage drop across the emitter resistor, and reduces the base-emitter voltage. The base current is reduced and, because of the exponential shape of the input characteristic, there is a large degree of compensation for the original change.

The feedback depends on R_e , a high value giving better stabilisation. The feedback also depends on how constant the base potential can be maintained during changes in base current; low values of R_1 and R_2 improve the stabilisation.

At one extreme, if R_e were large and R_1 and R_2 small enough to give a base potential of effectively zero resistance, the circuit would be indistinguishable from the grounded-base arrangement, with its inherently good stability. At the other extreme, if R_e were zero and R_1 and R_2 very large, the circuit would become equivalent to the simple grounded-emitter arrangement, with its very poor stability.

The circuit is very flexible, and a wide range of stability can be obtained between these two extremes. The collector current is no longer limited to the region above I'_{co} , and can be reduced almost to I_{co} .

In practice the relationship between the values of R_e , R_1 and R_2 will depend on the particular requirements of the circuit, but certain

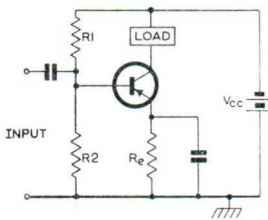


Fig. 7—RC-coupled circuit in grounded emitter to a.c.

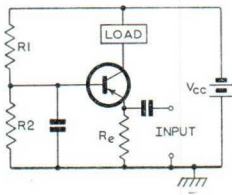


Fig. 8—RC-coupled circuit in grounded base to a.c.

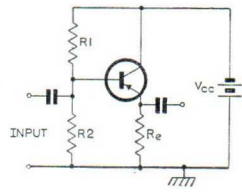


Fig. 9—RC-coupled circuit in grounded collector to a.c.

limitations exist. The maximum value of R_e depends on how much of the battery voltage can be dropped across it. The minimum values of R_1 and R_2 are dictated by the current which R_1 – R_2 can be allowed to bleed from the battery, and/or by the shunting of the incoming signal when RC coupling is used.

The factor of stability for this circuit is

$$K = \frac{1}{1 + \frac{\alpha'(R_e + r_e)}{R_b + r_{bb'} + R_e + r_e}}$$

where R_b is the effective base resistance formed by R_1 and R_2 in parallel, and is given by

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

and $r_{bb'}$ and r_e are internal resistances in series with base and emitter.

K will be low, and there will be good stability, if R_e is high and R_b low.

A low value of R_b gives a low value of K , but accentuates the effects of changes in V_{be} if R_e is also low.

The emitter-resistor and potential-divider circuit is suitable for RC or transformer coupling using any of the three a.c. configurations (Figs. 7 to 12).

N.T.C. Thermistor

The circuit may be modified by replacing R_2 by an n.t.c. thermistor (resistor having a negative temperature coefficient). However, a parallel combination of a normal and an n.t.c. thermistor is more usual (Fig. 13).

As the temperature rises, R_2 falls. The base voltage V_{be} decreases, offsetting the rise in collector current. Variations in α' from transistor

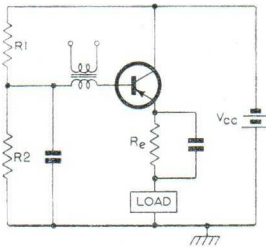


Fig. 10—Transformer-coupled circuit in grounded collector to a.c.

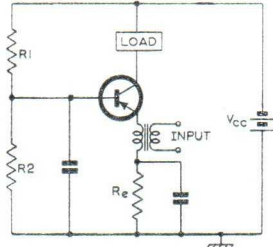


Fig. 11—Transformer-coupled circuit in grounded base to a.c.

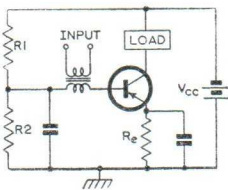


Fig. 12—Transformer-coupled circuit in grounded emitter to a.c.

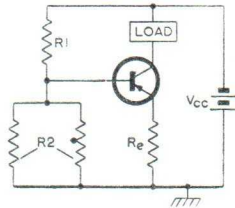


Fig. 13—N.T.C. thermistor in potential-divider bias circuit

to transistor are taken up as before by feedback through the emitter resistor, whilst the n.t.c. thermistor copes with the change in V_{be} with temperature. Better stabilisation is therefore obtained.

Overcompensation for changes in V_{be} counteracts changes in leakage current. Complete compensation over the whole temperature range cannot be provided, as the law of an n.t.c. thermistor is approximately

linear, whilst that of I'_{co} is exponential. The same collector current only occurs at two fixed temperatures.

P.T.C. Thermistor

A positive-temperature-coefficient (p.t.c.) thermistor is sometimes used.

The change in collector current with temperature is caused mainly by the temperature dependence of the leakage current and of the base-emitter voltage. Under certain conditions, and particularly in power-output stages and for silicon transistors, the change in V_{be} may become the more important of the two.

The change in V_{be} with temperature can be compensated by using emitter resistors of copper or nickel or other pure metal, which have a small positive temperature coefficient, instead of conventional resistors made from a zero-temperature-coefficient alloy.

Overcompensation for changes in V_{be} again helps to counteract changes in leakage current.

The p.t.c. thermistor may be particularly important in power-transistor circuits with low-impedance bias supplies.

TWO-BATTERY CIRCUIT WITH EMITTER RESISTOR

This circuit (Fig. 14) may be regarded as a special case of the preceding one, a base supply voltage V_{bb} providing the bias voltage instead of

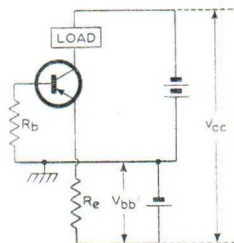


Fig. 14—Two-battery and emitter-resistor circuit

the potential divider R_1 – R_2 . Alternatively it may be regarded as a special case of the grounded-base circuit, in which a resistor R_b has been inserted in the base lead.

The collector current flows through the top part of the battery and the emitter current ($= I_c + I_b$) flows through the lower part. The base current is small, and a tapped battery can be used, as both parts will require replacement at about the same time.

The same equation for the factor of stability applies to this circuit as the preceding one.

The circuit saves one resistor with RC coupling, and two with trans-

former coupling. There is no potential divider to increase the current drain.

Since the voltage V_{bb} of the extra battery will often be higher than that required for biasing the base, the resulting higher value of R_e will

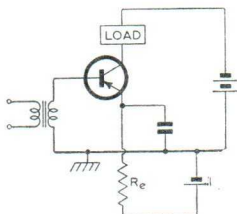


Fig. 15—Transformer-coupled circuit in grounded emitter to a.c.

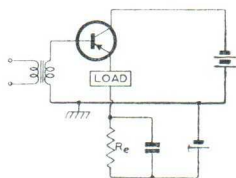


Fig. 16—Transformer-coupled circuit in grounded collector to a.c.

give better stabilisation. A higher combined battery voltage is implied, which to some extent offsets the saving of the current drain through R_1 - R_2 .

The design of the circuit is comparatively simple, as the current is almost entirely determined by V_{bb} and R_e and little affected by R_b .

This circuit is particularly suited to transformer coupling. With the transistor in grounded emitter or grounded collector to a.c. (Figs. 15

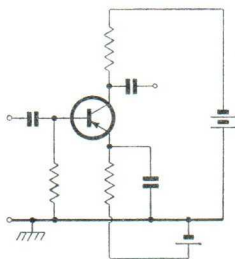


Fig. 17—RC-coupled circuit in grounded emitter to a.c.

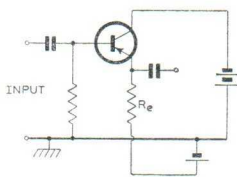


Fig. 18—RC-coupled circuit in grounded collector to a.c.

and 16), R_b is only the resistance of the transformer winding, and excellent stabilisation is achieved.

The circuit does not show to the same advantage for RC coupling, because R_b must then be several $k\Omega$'s, to avoid shunting of the input. Figs. 17 and 18 show the RC-coupled circuits, with the transistor in grounded emitter and grounded collector to a.c.

THERMAL RUNAWAY

Thermal runaway is a condition in which the collector current continues to rise until limited by some external means, or until the transistor is destroyed. The prevention of this effect is one of the chief purposes of d.c. stabilisation.

Thermal runaway is easily avoided in RC-coupled stages.

For a transistor to attain a state of thermal equilibrium, the junction temperature must reach some steady temperature T_j above the ambient temperature T_{amb} . These temperatures are related by the equation

$$T_j = T_{amb} + \theta p_{tot}$$

where p_{tot} is the collector dissipation plus base dissipation, and θ is the rise in junction temperature per unit collector dissipation. In most small-signal applications, the base dissipation may be neglected, and $p_{tot} \simeq p_c$.

On switching on the power supply, the following sequence of events takes place. The collector dissipation p_c makes the transistor start to warm up, thus initiating a rise in junction temperature ΔT_j . Because of the temperature dependence of I'_{co} , α' and V_{be} , the collector current will rise above its nominal quiescent value by an amount ΔI_c . In its turn, ΔI_c will lead to an increase Δp_c in collector dissipation. Finally, Δp_c leads to a further increase in junction temperature, equal to $\theta \Delta p_c$.

Positive thermal feedback exists in the system, the loop gain being

$$G = \frac{\theta \Delta p_c}{\Delta T_j}.$$

If the loop gain is greater than or equal to one, the transistor will be thermally unstable, and the collector current will 'run away'.

A simple expression can be derived for the loop gain. The collector dissipation is given by

$$p_c = I_c V_{ce}$$

and if R represents the total d.c. resistance in the collector and emitter circuits,

$$p_c = I_c(V_{ce} - I_c R).$$

Hence

$$p_c + \Delta p_c = (I_c + \Delta I_c)V_{ce} - R(I_c + \Delta I_c)^2$$

and by subtraction, neglecting $(\Delta I_c)^2$,

$$\Delta p_c = \Delta I_c(V_{ce} - 2I_c R) = \Delta I_c(2V_{ce} - V_{ce}).$$

Thermal runaway cannot take place provided the loop gain G is less than one, but as the final collector dissipation is $1/(1-G)$ times the initial switching-on dissipation, it is advisable to design for a G of about 0.5 for a limit transistor.

Thus the necessary condition for thermal stability becomes

$$G = \theta \cdot \frac{dI_c}{dT_j} (2V_{ce} - V_{ce}) < 1 \approx 0.5. \quad \dots(1)$$

If θ is given in $^{\circ}\text{C}$ per milliwatt, dI_c/dT_j should be expressed in milliamps per $^{\circ}\text{C}$, V_{ce} and V_{ce} being in volts. If θ is in $^{\circ}\text{C}/\text{W}$, dI_c/dT_j should be in amps per $^{\circ}\text{C}$.

Half-supply-voltage Principle

From Eq. 1, it follows that if the voltage across the transistor is equal to or less than half the supply voltage, the factor $(2V_{ce} - V_{ce})$ will be zero or negative. G will also be zero or negative. Such circuits are inherently thermally stable. This is the half-supply-voltage principle.

Most RC-coupled stages satisfy the half-supply-voltage condition without modification.

RECOMMENDED STABILISED CIRCUITS

Tables 1 and 2, at the end of the chapter (p. 72), give recommended circuits for the OC71. These circuits are suitable for operation up to a maximum operating ambient temperature of 45°C .

Examples have not been given in Table 1 for RC-coupled circuits operating at collector supply voltages of less than 4.5V , because of the reduction in gain which occurs at lower battery voltages. For special applications, such as hearing aids, the use of RC coupling at low battery voltages may still be desirable, however.

CLIPPING AND BOTTOMING

The working point must remain sufficiently outside the knee of the output characteristic to allow the required swing under drive, otherwise clipping will occur. Under extreme conditions, the increase in the direct collector current will 'bottom' the transistor.

As the collector current rises, because of either a rise in temperature or transistor spreads, the working point moves up along a 'd.c. load line' (Fig. 19). This line is drawn across the output characteristics to pass through the collector supply voltage V_{ce} , with a slope equal to the total d.c. resistance $R_c + R_e$ in the collector and emitter circuits. The working point (I_c , V_c) is given by

$$V_c = V_{ce} - I_c(R_c + R_e). \quad \dots(2)$$

The transistor bottoms when all the available voltage is dropped across the d.c. resistance, that is, when the collector current rises to a value

$$I_c = \frac{V_{ce} - V_{knee}}{R_c + R_e}, \quad \dots(3)$$

where V_{knee} is the knee voltage.

The knee voltage depends on the collector current. When designing amplifier stages for the typical small-signal germanium and silicon transistors, it may be assumed that V_{knee} will not exceed 0.2V.

It follows from Eq. 3 that stages in which V_{cc} is increased or $R_{\text{c}} + R_{\text{e}}$ decreased will not clip until a higher collector current is reached, that is, such stages can be operated up to higher ambient temperatures.

DESIGN PROCEDURE

Wherever possible, circuits for the OC71 should be selected from those given in Tables 1 and 2. A design procedure will now be described which is suitable for special requirements, such as higher supply voltages, and for transistors other than the OC71.

There are many ways of proceeding with the design, but this method has been chosen as being sufficiently rigorous and simple. Circuits designed along these lines will have if anything a larger margin of

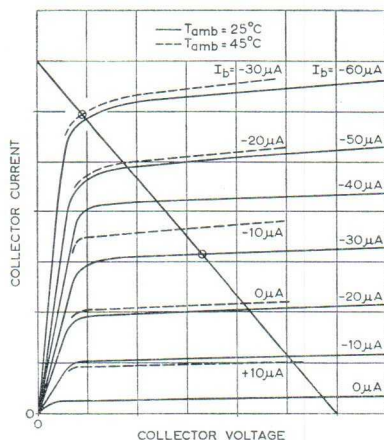


Fig. 19—Output characteristics with superimposed d.c. load line

safety than is really necessary, at the expense of some loss of gain and slightly increased battery drain. For 'one-off' quantities, however, the ultimate in performance is usually not essential. The circuits may be incorporated in permanently installed equipment, and need not be restricted to 'breadboards' or 'lash ups'.

The method is based on three rules of thumb:

- (a) The circuit should be designed for a collector current of 1mA, unless the signal level requires a higher current.

- (b) About $\frac{1}{2}$ to 1V should be dropped across the emitter resistor R_e . For a collector current of 1mA, the emitter resistor will be in the region of 470Ω to $1k\Omega$.
- (c) The external base resistance R_b normally should be chosen to be of the order of $10 \times R_e$. The circuit will then generally cope with junction temperatures of up to 45 to 50°C.

More flexibility is possible in the circuit design than these rules of thumb might suggest. The collector current may be increased above 1mA if the signal swing demands it. Collector currents below 1mA should be avoided as far as possible, because better stabilisation

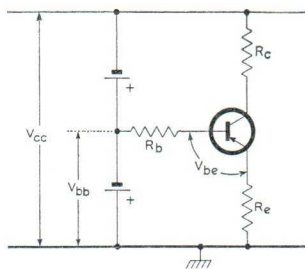


Fig. 20—Equivalent circuit for one-battery and two-battery circuits

becomes necessary. The stabilisation may also have to be improved if the transistor will be expected to operate at high ambient temperatures, and/or if the supply voltage is low. The improvement is effected by increasing the voltage dropped across the resistance in the emitter, and by decreasing the ratio of R_b/R_e from 10 to perhaps 5 or less. If R_b is very low, then it may be necessary to include $r_{bb'}$ in R_b , because $r_{bb'}$ will no longer be negligible in comparison with R_b .

The one-battery and two-battery circuits can be analysed by means of the same equivalent circuit (Fig. 20). From the equivalent circuit, it follows that

$$V_{bb} = I_b R_b + V_{be} + I_e R_e. \quad \dots (4)$$

Since the emitter and collector currents are approximately equal, and since the base current is approximately given by I_e/α' ,

$$V_{bb} \simeq \left(R_e + \frac{R_b}{\alpha'} \right) I_e + V_{be}. \quad \dots (4a)$$

For small-signal transistors, at collector currents of a few milliamps, α' may be taken as being equal to α'_0 .

In Fig. 21, V_{bb} is plotted against R_b with R_e as parameter for a

collector current of 1mA, the transistor being the OC71, for which $\alpha'_0 \simeq 40$ and $V_{be} \simeq 0.1V$.

The value of R_b is chosen with the aid of Fig. 21 or Eq. 4a. In the two-battery circuit, V_{bb} is fixed in advance, and there is some restriction on the choice of component values. In the one-battery circuit, the base can be regarded as being fed from an artificial tap on the battery, the

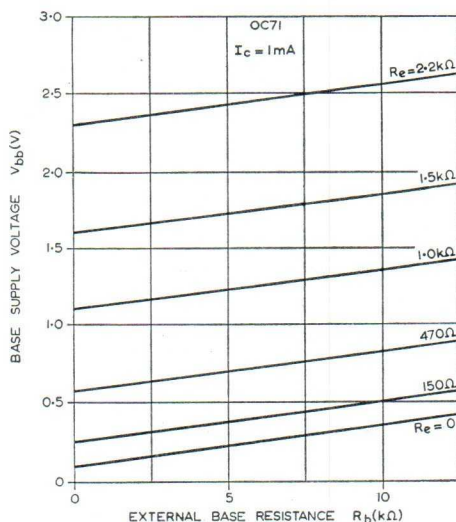


Fig. 21—Base supply voltage related to external base resistance and to resistance in emitter circuit.

proportion of the voltage tapped off being completely under the control of the designer.

From the values of R_b , R_e and α' , the factor of stability K can be calculated, using

$$K = \frac{(R_b + R_e)}{(R_b + R_e) + \alpha' R_e} = \frac{1}{1 + \frac{\alpha' R_e}{R_b + R_e}} \quad \dots (5)$$

If the factor $(R_b + R_e)$ is particularly small, $r_{bb'}$ must be added to R_b . K also can be found from a graph, such as that given in Fig. 22 for an α' of 75.

The changes in the collector current caused by (a) the spread in α' , (b) the increase in leakage current with temperature, and (c) the decrease in V_{be} with temperature and the spread in V_{be} , are additive and are given by:

$$\Delta I_{c(1)} = K \left(\frac{\Delta \alpha'}{\alpha'} \right) I_c, \quad \dots (6)$$

$$\Delta I_{c(2)} \simeq K(1 + \alpha') \Delta I_{c0}, \quad \dots (7)$$

and
$$\Delta I_{c(3)} = - \frac{K \alpha'}{R_b + R_e} \cdot \Delta V_{be} \quad \dots (8)$$

or
$$\simeq - \frac{\Delta V_{be}}{R_e}. \quad \dots (8a)$$

The total change in collector current

$$\Delta I_c = \Delta I_{c(1)} + \Delta I_{c(2)} + \Delta I_{c(3)} \quad \dots (9)$$

gives the maximum collector current as

$$I_{c \max} = I_{c \text{ nom}} + \Delta I_c. \quad \dots (10)$$

Eq. 6 allows for the effect of inserting a transistor having the maximum α' in a circuit designed for the nominal α' . K is calculated for

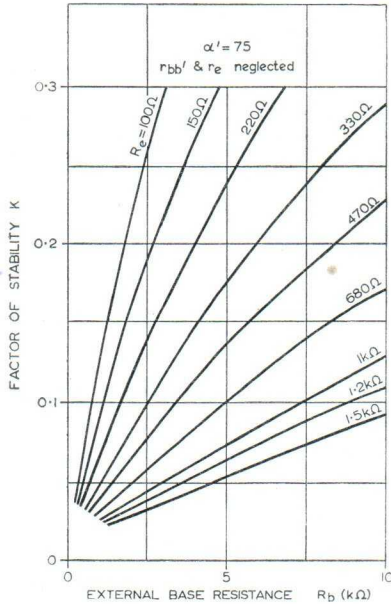


Fig. 22—Evaluation of K for an α' of 75

the maximum α' ; $\Delta \alpha'$ is the difference between the maximum and nominal α' ; and the nominal value of α' is inserted in the denominator. Usually $\Delta \alpha' / \alpha' \simeq 0.5$.

Eq. 6 may also be used for calculating the effect of inserting a

transistor having the minimum α' . K should then be calculated for the minimum value of α' . $\Delta\alpha'$ will now be negative and $\Delta I_{c(1)}$ will be negative.

Eq. 7 primarily allows for the increase in leakage current up to the maximum junction temperature occurring in the circuit. The increase can be found by means of a curve such as that shown in Fig. 23, which is for the OC71. The calculation should be made for a transistor having the maximum I_{co} and—although this combination is rather unlikely in practice—the maximum α' . For a transistor operating under

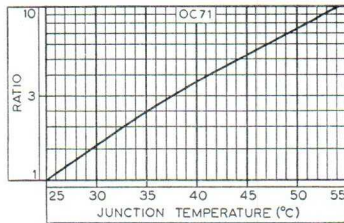


Fig. 23—Collector leakage current in grounded base, I_{co} , at a given junction temperature divided by its value at 25°C.

true small-signal conditions, the maximum junction temperature may be taken as being approximately equal to the maximum ambient temperature.

With reference to Eq. 8, V_{be} for both silicon and germanium transistors decreases by roughly 2mV for every °C rise in temperature. The sign of $\Delta I_{c(3)}$ is positive for a decrease in V_{be} and negative for an increase in V_{be} . Eq. 8 should be used for the one-battery case, but 8(a) is sufficiently exact for two-battery circuits.

The minimum acceptable collector-emitter voltage is

$$V_{cmin} = (V_{knee} + e_c) \times \frac{\text{nominal } V_{cc}}{\text{end-of-life } V_{cc}} \quad \dots(11)$$

where e_c is the peak signal required. No end-of-life correction is needed for a mercury cell, which has a constant-voltage characteristic. Nor is this correction required if the stage drives another whose handling capacity is also reduced as V_{cc} falls.

The resistance required in the collector is found from:

$$R_c = \frac{V_{cc} - V_{cmin}}{I_{cmax}} - R_e \quad \dots(12)$$

If the value of R_c is unacceptably low, then the circuit must be re-designed for a higher battery voltage, or for lower values of R_b and K .

The junction temperature at the maximum collector current is:

$$\max. T_j = T_{amb\max} + \theta(I_{c\max})(V_{c\min}) \quad \dots(13)$$

where θ is expressed in $^{\circ}\text{C}/\text{mW}$ and is 0.4 for the OC70, OC71 and OC75. If the maximum junction temperature exceeds the maximum ambient temperature by more than 2 to 3°C , the circuit must be re-designed using the true junction temperature when calculating the change in leakage current.

For the one-battery circuit, the values of the resistors in the potential divider are found from:

$$R_1 = \frac{V_{cc}R_b}{V_{bb}} \quad \dots(14)$$

$$\text{and} \quad R_2 = \frac{V_{cc}R_b}{V_{cc} - V_{bb}} \quad \dots(15)$$

If R_1 and R_2 are fairly similar in value, the tolerances should be $\pm 5\%$. Otherwise $\pm 10\%$ is sufficient.

In the following examples, no allowance has been made for the spread in V_{be} , but only for the change with temperature. There is little error, however. The nearest lower preferred value is chosen for R_2 and R_e .

Example 1

Tapped-battery RC-coupled small-signal circuit for OC71: $V_{cc} = -9V$, $V_{bb} = -1.5V$, max. operating $T_{amb} = 45^{\circ}\text{C}$.

Choose $I_c = 1\text{mA}$ and $R_e = 1.2\text{k}\Omega$.

From Fig. 21, $R_b = 10\text{k}\Omega$.

$$\begin{aligned} \text{Maximum } \alpha' &\simeq \text{nominal } \alpha' + 50\% \\ &= 40 + 20 = 60. \end{aligned}$$

The value of K for an α' of 60 is, from Eq. 5,

$$K = \frac{(R_b + R_e)}{(R_b + R_e) + \alpha' R_e} = \frac{11.2}{11.2 + 60 \times 1.2} = 0.13.$$

From Eq. 6,

$$\Delta I_{c(1)} = K \left(\frac{\Delta \alpha'}{\alpha'} \right) I_c = 0.13 \times 0.5 \times 1.0 = 0.07\text{mA}.$$

From Eq. 7 and Fig. 23, where $I_{c0\max} = 13\mu\text{A}$,

$$\Delta I_{c(2)} \simeq K(1 + \alpha') \Delta I_{c0} = 0.13 \times 61 \times (4 \times 13 \times 10^{-3}) = 0.4\text{mA}.$$

From Eq. 8(a),

$$\Delta I_{c(3)} \simeq \Delta T_{amb} \times 2\text{mV}/R_e = (45 - 25) \times 2/1.2 = 0.03\text{mA}.$$

From Eqs. 9 and 10,

$$I_{c\max} = 1.0 + 0.07 + 0.4 + 0.03 = 1.5\text{mA}.$$

From Eq. 11, for no end-of-life correction,

$$V_{c\min} = V_{knee} + e_c = 0.2 + 0.1 = 0.3V.$$

From Eq. 12,

$$R_c = \frac{V_{cc} - V_{c\min}}{I_{c\max}} - R_e = \frac{9 - 0.3}{1.5} - 1.2 \simeq 3.9k\Omega \pm 10\%.$$

The voltage dropped across the d.c. resistance is

$$V_R = I_c(R_c + R_e) = 1.0(3.9 + 1.2) = 5V,$$

and as this is greater than $4.5V (= \frac{1}{2}V_{cc})$ the circuit is necessarily thermally stable.

Example 2

One-battery RC-coupled small-signal circuit for OC201: $V_{cc} = -12V$, max. operating $T_{amb} = 100^\circ C$.

The OC201 is a silicon transistor having an I_{co} of only $10\mu A$ at $T_j = 100^\circ C$, and of only $100m\mu A$ at $25^\circ C$.

$$\theta = 0.5^\circ C/mW.$$

Stabilisation is now primarily for the spread in α' and change in V_{be} .

Choose $I_c = 1mA$ and $R_e = 470\Omega$.

Choose $R_b = 5k\Omega$.

At $1mA$, the nominal $\alpha' = 30$ and the maximum $\alpha' = 80$.

$$K = \frac{5.5}{5.5 + 80 \times 0.5} = 0.12.$$

$$\Delta I_{c(1)} = 0.12 \times \frac{80 - 30}{30} \times 1.0 = 0.2mA.$$

$$\Delta I_{c(2)} = 0.12 \times 81 \times (10 \times 10^{-3}) = 0.1mA.$$

$$\Delta I_{c(3)} = \frac{0.12 \times 81}{5.5} \cdot (100 - 25) \times 2 \times 10^{-3} = 0.26mA.$$

$$I_{c\max} = 1.0 + 0.2 + 0.1 + 0.26 = 1.56mA.$$

$$V_{c\min} = 0.2 + 0.1 = 0.3V.$$

$$R_c = \frac{12 - 0.3}{1.6} - 0.5 \simeq 5.6k\Omega \pm 10\%.$$

$$V_R = (5.6 + 0.5)1.0 = 6.1V > \frac{1}{2}V_{cc}.$$

For a one-battery circuit, V_{bb} , R_1 and R_2 have to be calculated using Eqs. 4(a), 14 and 15.

$$V_{bb} \simeq \left(R_e + \frac{R_b}{\alpha'} \right) I_c + V_{be}$$

$$= \left(0.5 + \frac{5}{30}\right) 1.0 + 0.1 = 0.77V.$$

$$R_1 = \frac{V_{cc}R_b}{V_{bb}} = \frac{12 \times 5}{0.77} \approx 68k\Omega \pm 10\%.$$

$$R_2 = \frac{V_{cc}R_b}{V_{cc} - V_{bb}} = \frac{12 \times 5}{12 - 0.77} = 4.7k\Omega \pm 10\%.$$

Example 3

One-battery RC-coupled large-signal circuit for OC71: $V_{cc} = -14V$, max. operating $T_{amb} = 45^\circ C$.

Choose $I_c = 3mA$.

In view of the high supply voltage, R_e may be chosen as high as 470Ω .

R_b may be chosen as $5k\Omega$ ($\approx 10 \times R_e$).

At $3mA$, the maximum α' is 75. Assume max. $T_j = 47^\circ C$.

From Fig. 22,

$$K = 0.14.$$

$$\Delta I_{c(1)} = 0.14 \times 0.5 \times 3 = 0.21mA.$$

$$\Delta I_{c(2)} = 0.14 \times 76 \times (5 \times 13 \times 10^{-3}) = 0.69mA.$$

$$\Delta I_{c(3)} = \frac{0.14 \times 75}{5.5} \cdot (47 - 25) \times 2 \times 10^{-3} = 0.08mA.$$

$$I_{cmax} = 3 + 0.21 + 0.69 + 0.08 = 4mA.$$

$$V_{cmin} = 0.2 + 0.5 = 0.7V \text{ (for } 1.0V \text{ peak-to-peak)}$$

$$R_c = \frac{14 - 0.7}{4} - 0.47 \approx 2.7k\Omega \pm 10\%.$$

The maximum junction temperature, from Eq. 13, is

$$\begin{aligned} \text{max. } T_j &= T_{ambmax} + \theta(I_{cmax})(V_{cmin}) \\ &= 45 + 0.4 \times 4 \times 0.7 = 46^\circ C. \end{aligned}$$

The calculation does not need to be repeated, since a max. T_j of $47^\circ C$ was assumed.

$$V_R = (2.7 + 0.5)3 = 9.6V > \frac{1}{2}V_{cc}.$$

As the design is satisfactory in the above respects, V_{bb} , R_1 and R_2 may be calculated.

$$V_{bb} \approx \left(0.5 + \frac{5}{50}\right) 3.0 + 0.1 = 1.9V.$$

$$R_1 = \frac{14 \times 5}{1.9} \approx 33k\Omega \pm 10\%.$$

$$R_2 = \frac{14 \times 5}{14 - 1.9} \approx 5.6k\Omega \pm 10\%.$$

Example 4

One-battery RC-coupled large-signal circuit for OC75: $V_{ce} = -18V$, max. operating $T_{amb} = 45^\circ C$.

The OC75 has a maximum α' of 130 at 3mA and a maximum I_{co} of $14\mu A$ at $25^\circ C$.

Choose $I_c = 3mA$.

R_e may be chosen as 680Ω , as there is plenty of supply voltage available.

Choose $R_b = 7k\Omega$ ($\simeq 10 \times R_e$).

Assume max. $T_j = 47^\circ C$.

$$K = \frac{7.7}{7.7 + 130 \times 0.7} = 0.08.$$

$$\Delta I_{c(1)} = 0.08 \times 0.5 \times 3 = 0.12mA.$$

$$\Delta I_{c(2)} = 0.08 \times 130 \times (5 \times 14 \times 10^{-8}) = 0.73mA.$$

$$\Delta I_{c(3)} = \frac{0.08 \times 130}{7.7} \cdot (47 - 25) \times 2 \times 10^{-3} = 0.06mA.$$

$$I_{cmax} = 3 + 0.12 + 0.73 + 0.06 = 3.9mA.$$

$$V_{emin} = 0.2 + 0.8 = 1.0V \text{ (for } 1.6V \text{ peak-to-peak)}$$

$$R_c = \frac{18 - 1}{3.9} - 0.7 = 3.3k\Omega \pm 10\%.$$

$$\text{max. } T_j = 45 + 0.4(3.9)(1.0) = 46.6^\circ C.$$

$$V_R = 3(3.3 + 0.7) = 12V > \frac{1}{2}V_{cc}.$$

$$V_{bb} = \left(0.7 + \frac{7}{130}\right) 3.0 + 0.1 = 2.36V.$$

$$R_1 = \frac{18 \times 7}{2.36} \simeq 47k\Omega \pm 10\%.$$

$$R_2 = \frac{18 \times 7}{18 - 2.36} \simeq 6.8k\Omega \pm 10\%.$$

TABLE 1

PREFERRED ONE-BATTERY CIRCUITS FOR OC71

max. operating $T_{amb} = 45^{\circ}\text{C}$

V_{cc} (V)	I_c (mA)	R_e ($k\Omega$)	R_1 ($k\Omega$)	R_2 ($k\Omega$)	R_c ($k\Omega$)
<i>RC Coupling</i>					
$4\frac{1}{2}$	0.5	1.0	18	2.7	3.3
6	0.5	1.0	33	3.9	3.3
6	1.0	1.0	39	10	2.2
6	1.5	1.0	22	10	1.5
9	1.0	1.0	62	10	3.9
9	1.5	1.0	39	10	2.7
12	1.0	1.0	82	10	5.6
12	1.5	1.0	56	10	4.7
<i>Transformer Coupling</i>					
$1\frac{1}{2}$	0.5	1.0	4.7	3.3	0.2*
3	0.5	1.0	10	2.7	0.2
$4\frac{1}{2}$	3.0	0.47	10	6.8	0.2
6	3.0	0.47	12	4.7	0.2

*Resistance of primary winding of output transformer

TABLE 2

PREFERRED TWO-BATTERY CIRCUITS FOR OC71

 $V_{cc} = -6\text{V}$, $V_{bb} = -1.5\text{V}$, max. operating $T_{amb} = 45^{\circ}\text{C}$

I_c (mA)	R_e ($k\Omega$)	R_b ($k\Omega$)	R_c ($k\Omega$)
<i>RC Coupling</i>			
0.5	2.7	10	3.9
0.5	2.7	6.8	4.7
1.0	1.2	10	2.2
<i>Transformer Coupling</i>			
0.5	2.7	0	0.2*
1.1	1.2	0	0.2
2.85	0.47	0	0.2

*Resistance of primary winding of output transformer

CHAPTER EIGHT

EQUIVALENT CIRCUITS

The major processes occurring in a junction transistor have already been outlined in Chapter 1. These processes will now be represented by electrical circuit elements, which will be assembled to form a complete equivalent circuit.

By making some approximations, the complete equivalent circuit can be simplified into a number of more familiar forms. The relation-

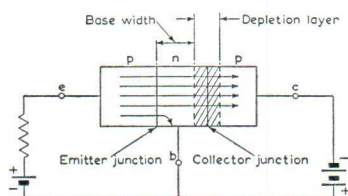


Fig. 1—Theoretical model of junction transistor

ship between the most important of these circuits is shown, and some of the approximations made in their derivation are pointed out.

A family of six circuits is described, each member of which covers the full useful range of operating frequency. Three circuits will also be given which apply to restricted frequency ranges.

From the values of not more than six transistor parameters, any of the equivalent circuits can be written out in full. For some of the circuits, fewer than six parameters need be known.

For reference, the six basic parameters are listed here in the order in which they will be introduced: r_e , α_0 , μ , $r_{bb'}$, C_{dep} and C_e .

Three equivalent circuits are used as a basis for drawing up the published data. They are the low-frequency T circuit, the hybrid π grounded-emitter circuit, and the complete grounded-base T circuit.

The approximate equivalent circuits described in this chapter apply to alloy-junction transistors only.

TRANSISTOR MODEL

The transistor may be represented by the theoretical model of Fig. 1.

Emitter Action: Carrier Injection

As all the current carriers leaving the emitter flow into the base, no

components are required to represent carrier injection. The emitter is therefore shown connected directly (Fig. 2) to the components which represent the diffusion of carriers through the base.

The carriers leaving the emitter will be holes in a p-n-p transistor, or electrons in an n-p-n one. In what follows a p-n-p transistor will be assumed, for the sake of definiteness.

Diffusion through Base

Holes (in a p-n-p transistor) cross the emitter junction and diffuse within the field-free region of the base, until they reach the collector

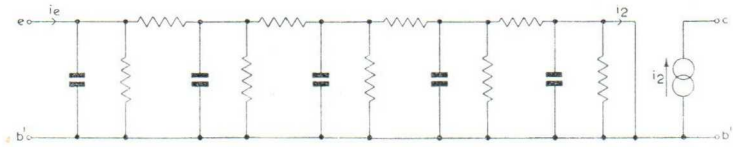


Fig. 2—Representation of diffusion and collector action

depletion layer. This process is represented by the symmetrical RC transmission line shown in Fig. 2.

The holes exist long enough in the base for a proportion of them to recombine with current carriers of opposite sign. Recombination in the base region is represented by the distributed shunt resistance along the transmission line.

The length of the transmission line corresponds to the width of the base. Recombination is less in a narrower base.

Collector Action

All the holes reaching the collector depletion layer, during their random movement of diffusion in the base, will be swept out of the base into the collector by the field existing across the depletion layer. This process corresponds to a short-circuit on the end of the transmission line, as shown in Fig. 2.

The flow of current carriers must appear finally as a current at the collector terminal flowing out of the reverse-biased collector junction. Since in short-circuiting the line the effect of this current has been removed from the equivalent circuit, an infinite-impedance current generator is connected between the collector and base terminals in Fig. 2, to restore it. The current flowing from the current generator equals that flowing in the short-circuited end of the transmission line.

Fig. 2, which still does not give a complete representation, may

be compared with the first low-frequency equivalent circuit proposed for the transistor (Fig. 3).

Emitter Resistance r_e

The input resistance of the short-circuited transmission line is known as the emitter resistance r_e , and is given by the formula

$$r_e = \frac{0.0863T}{I_e} \text{ ohms,}$$

where T is the junction temperature measured in $^{\circ}\text{K}$ and I_e is the

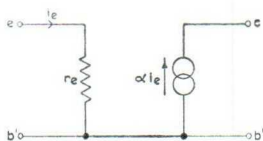


Fig. 3—First low-frequency equivalent circuit

emitter current in milliamps. An approximate but more useful expression, applicable at room temperatures (18 to 20°C), is

$$r_e \simeq \frac{25}{I_e} \text{ ohms.}$$

I_e may be 1mA and then r_e is 25Ω . r_e is proportional to $1/I_e$ and to T ($^{\circ}\text{K}$). The temperature in degrees Kelvin is $^{\circ}\text{C} + 273$.

Current Amplification Factor

The forward current transfer in grounded base with the output short-

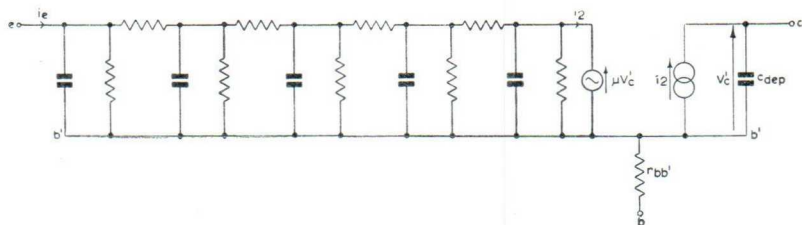


Fig. 4—Complete equivalent circuit with transmission line

circuited to a.c. is the current amplification factor $\alpha = i_2/i_e$. The value of α at low frequencies is denoted by α_0 and is usually about 0.98.

A number of components will now be added to Fig. 2 in order to derive the complete equivalent circuit shown in Fig. 4.

Width of Collector Depletion Layer: Feedback Factor μ

The width of the space-charge layer or depletion layer shown in Fig. 1 depends on the collector voltage (Early effect). If the collector voltage is increased, the width of the depletion layer increases, and so the effective width of the base decreases. Therefore the base width, and the length of the transmission line, vary with collector voltage.

The modulation of the base width by the collector voltage is represented by a zero-impedance voltage generator in the short-circuited end of the transmission line (Fig. 4). The voltage of the generator is a fraction μ of the voltage V'_c across the collector junction.

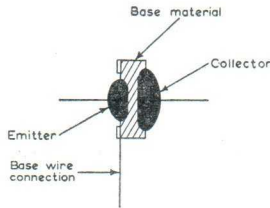


Fig. 5—Diagram of cross-section of practical junction transistor

The significance of μ is that it represents one form of feedback within the transistor.

If μ is put equal to zero, the internal feedback at low frequencies becomes zero, and the equivalent circuit becomes that much nearer to the simple one shown in Fig. 3.

μ is proportional to $\frac{1}{\sqrt{V_c}}$ and is independent of frequency.

Capacitance of Collector Depletion Layer c_{dep}

The collector depletion layer may be regarded as a parallel-plate capacitor c_{dep} , which is included in Fig. 4. A typical value for c_{dep} is 10pF. This capacitance causes feedback within the transistor at high frequencies.

c_{dep} is proportional to $\frac{1}{\sqrt{V_c}}$.

Ohmic Base Resistance $r_{bb'}$

In a practical transistor, shown diagrammatically in Fig. 5, it is necessary to take into account the resistance of the base material between the active part of the transistor (between emitter and collector) and the base connecting wire. The external base connection ideally would

EQUIVALENT CIRCUITS

be made to an imaginary point b' , in the active part of the base material. The resistance between b' and the actual base connection b is the internal base resistance $r_{bb'}$ included in Fig. 4. $r_{bb'}$ depends on the construction of the transistor, and a typical value (e.g. for the OC45) is 75Ω .

The complete equivalent circuit has now been assembled and all the components shown in Fig. 4 have been identified. This circuit is

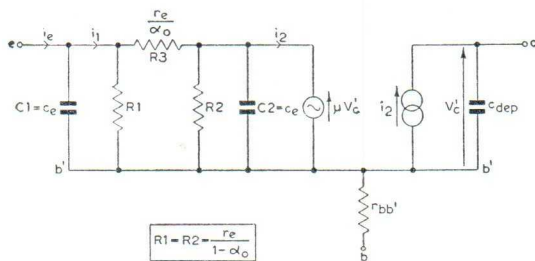


Fig. 6—Simplified equivalent circuit with collapsed transmission line

too complicated for most practical purposes. It will now be simplified and transformed into the more usual arrangements. The T equivalent circuits will be considered first, then the π circuits.

GROUND-BASE T EQUIVALENT CIRCUIT

First an approximation will be made to the transmission line, and then part of the circuit will be converted into an alternative network having identical properties.

In Fig. 6, the circuit of Fig. 4 has been simplified by lumping the total resistance and capacitance shunted across the transmission line into two equal parts, and connecting half the total at the start of the line and half at the finish. Thus $R_1 = R_2$ and $C_1 = C_2$. Pictorially this process is equivalent to collapsing or telescoping the line, R_3 being the total series resistance of the line.

R_1 , R_2 and R_3 are all simply related to r_e by factors of α_o and $1 - \alpha_o$, as shown in Fig. 6. R_1 and R_2 are very much greater than R_3 , since

$$\frac{R_1}{R_3} = \frac{\alpha_o}{1 - \alpha_o} = \alpha'_o.$$

Emitter Capacitance c_e

C_1 in Fig. 6 is known as the emitter capacitance c_e . C_2 also equals c_e .

All the six basic parameters listed at the beginning of the chapter have now been introduced.

c_e is proportional to I_e and to $1/Tf_\alpha$, where T is in $^{\circ}\text{K}$ and f_α is the α cut-off frequency.

Network Conversion

Interchanging R_2 and C_2 in Fig. 6 does not affect the electrical properties of the circuit or in any way disturb the electrical symmetry. The central part of the network shown in Fig. 7 may be extracted and considered separately.

The central part of Fig. 7 can be converted into a different network, shown in Fig. 8, without in any way altering its electrical character-

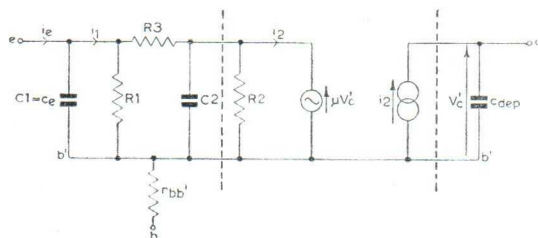


Fig. 7—Previous figure redrawn and divided into three

istics. The characteristics of the whole network will therefore also be unchanged.

The proof that the central part of Fig. 7 is identical with Fig. 8 runs as follows. The performance of this type of network may be described completely by means of four parameters. If the four parameters describing one network are identical with those describing another network, then all the characteristics of the two networks are identical. It is convenient to choose the input and output impedances for open- and short-circuit terminations as the four parameters. Then the two networks are identical since for both:

Input impedance with output terminals short-circuited = 0

Input impedance with output terminals open-circuited = R_2

Output impedance with input terminals short-circuited = 0

Output impedance with input terminals open-circuited = R_2/μ .

A typical value for R_2 is 1250Ω and μ could equal $1/2000$, making the value of R_2/μ in Figs. 8 and 9 equal to $2.5\text{M}\Omega$.

C_2 in Fig. 6 can be transferred to the right-hand side of the generator by the same method as used for R_2 . A similar change in impedance is made by multiplying the capacitance by μ , to give μC_e ($= \mu C_2$) in Fig. 9.

Collector Capacitance and Collector Resistance

The combined capacitance of μc_e and c_{dep} in parallel is known as the

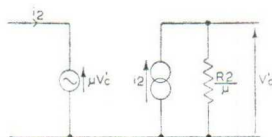


Fig. 8—Alternative form of centre network

collector capacitance c_c . For most purposes μc_e is negligibly small in comparison with c_{dep} , and $c_c \approx c_{dep}$.

Fig. 9 may now be simplified to Fig. 10. With $V'_e = 0$, the input resistance between b' and e is equal to R_1 and R_3 in parallel, that is, r_e . When V'_e is not zero, the voltage $\mu V'_e$ appears through R_3 at the

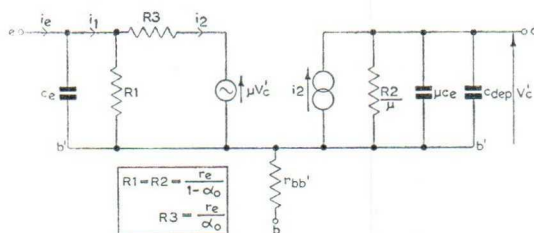


Fig. 9— R_2 and C_2 transferred to collector

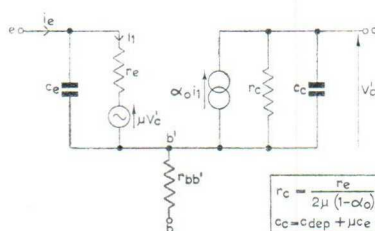


Fig. 10—Two-generator grounded-base T equivalent circuit

input terminals. Because $R_3 = r_e/\alpha_0 \approx r_e$, the $\mu V'_e$ generator may be placed in series with r_e as shown in Fig. 10.

The effect of R_1 on the output resistance has now been removed, so it must be restored by connecting an additional resistance across b' and c , as was previously done for R_2 . Because R_3 is very much smaller than R_1 , the effect of R_1 on the output resistance is approximately the same as that calculated for R_2 . Since two equal resistances

in parallel give a combined resistance equal to half either one of them, the total resistance between b' and c in Fig. 10 becomes half that in

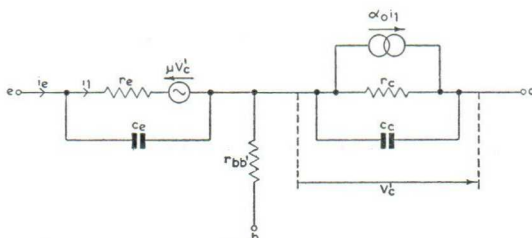


Fig. 11—Usual form of two-generator grounded-base T equivalent circuit

Fig. 9. This resistance is known as the collector resistance $r_c (= R_2/2\mu)$.

α is equal to i_2/i_e , and at low frequencies the effect of c_e is negligible, so that $i_1 = i_e$ and $\alpha_0 = i_2/i_1$. The current generator in Fig. 10 is therefore labelled $\alpha_0 i_1$.

The complete grounded-base T equivalent circuit shown in Fig. 10 is redrawn in its more usual form in Fig. 11.

A number of equivalent circuits are related directly to the grounded-base T circuit, and these will now be described, but without giving the full derivation.

DOUBLE-BASE-RESISTANCE EQUIVALENT CIRCUIT

The double-base-resistance equivalent circuit of Fig. 13 is equivalent

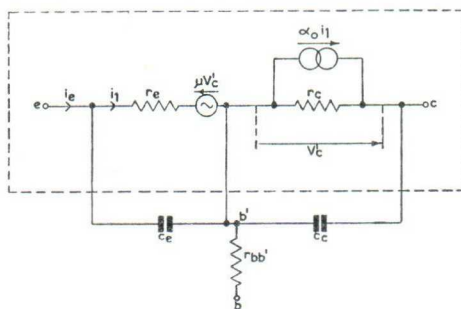


Fig. 12—Redrawn version of grounded-base T equivalent circuit divided into two

electrically to the grounded-base circuit of Fig. 11. To show the relationship between the two, Fig. 11 has been redrawn in Fig. 12.

The double-base-resistance circuit (Fig. 13) is obtained by substituting the network within the box for the corresponding network in

Fig. 12. The conditions for the two networks to be identical are found, as before, by equating the values of four parameters for each network. These conditions are included in Fig. 13.

The double-base-resistance circuit contains only one generator (the current generator). The effect of the form of internal feedback,

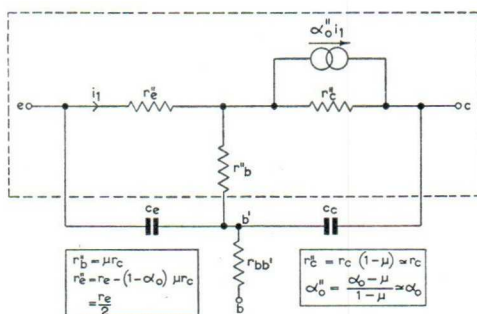


Fig. 13—Double-base-resistance grounded-base equivalent circuit

previously accounted for by the $\mu V'_c$ generator in the emitter circuit, is now taken care of by the factor μ , which appears in each of the expressions for r_b'' , r_e'' , r_c'' and α_o'' .

LOW-FREQUENCY T EQUIVALENT CIRCUIT

If the capacitors are omitted from the double-base-resistance equivalent

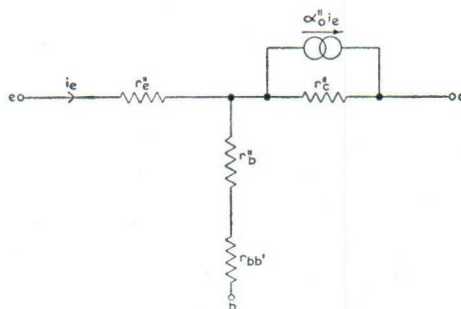


Fig. 14—Conventional low-frequency T equivalent circuit for grounded base

circuit (Fig. 13), the result is the grounded-base low-frequency T circuit shown in Fig. 14.

In Fig. 15 the grounded-base circuit has been redrawn with its emitter grounded. This is the grounded-emitter low-frequency T circuit. The quantities r_e'' , r_c'' and $(r_b'' + r_{bb'})$ are now represented by capital

letters as R_e , R_c and R_b . Often they are quoted simply with small letters as r_e , r_c and r_b , but this practice can be confusing since r'_e equals $r_e/2$.

R_b (or r_b) is $r_{bb'} + \mu r_c$ and $R_c \simeq r_c$.

This equivalent circuit is given in the published data for the OC70, OC71 and OC75 low-frequency transistors.

An alternative form of the circuit with the current generator expressed in terms of α'_o and i_b is shown in Fig. 16. To keep the output impedance

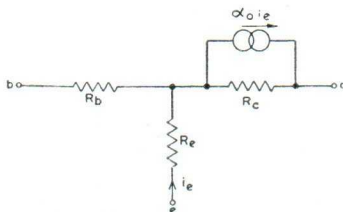


Fig. 15—Low-frequency T equivalent circuit for grounded emitter

the same, R_c has been replaced by $R_c(1 - \alpha_o)$, which is approximately equal to R_c/α'_o .

SIMPLIFIED HIGH-FREQUENCY T EQUIVALENT CIRCUIT

In the equivalent circuit of Fig. 11, the voltage V'_e that can be developed across the collector circuit at high frequencies is usually severely limited by the low reactance of c_c , and so the voltage $\mu V'_e$

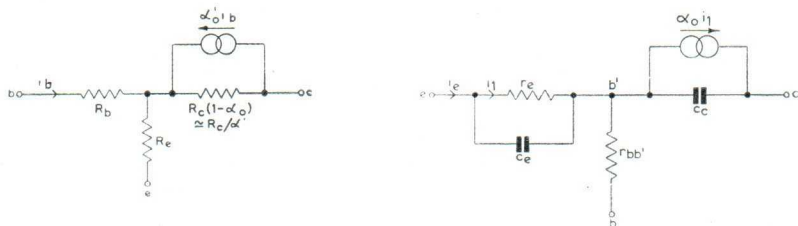


Fig. 16—Alternative low-frequency T equivalent circuit for grounded emitter

Fig. 17—Simplified high-frequency T equivalent circuit

in the emitter circuit may be neglected. Similarly, since c_c is a low reactance, little current flows through r_c , and this also may be neglected. The high-frequency circuit therefore reduces to that of Fig. 17.

GROUND-BASE π EQUIVALENT CIRCUIT

The grounded-base T circuit shown in Fig. 10 can be transformed into

the π circuit shown in Fig. 18. The circuit values have been calculated by the same methods as for the other circuits, by comparing alternative portions of the circuits and equating parameters. The resistance in parallel with c_e is equal to $2 \times r_e (= R_2/\mu)$.

g_m is the mutual conductance of the transistor, and its value is

$$g_m = \frac{\alpha_0}{r_e}$$

HYBRID π GROUNDED-EMITTER EQUIVALENT CIRCUIT

The hybrid π grounded-emitter circuit is perhaps the most generally

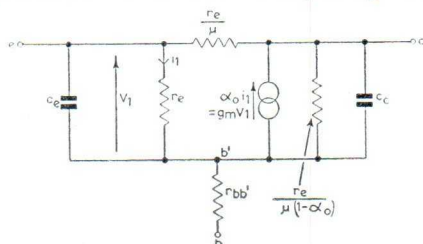


Fig. 18—Complete grounded-base π equivalent circuit

useful, and will therefore be dealt with in slightly more detail than some of the preceding circuits.

Derivation from Grounded-base π Circuit

The hybrid π grounded-emitter equivalent circuit can be derived from the grounded-base π circuit. In Fig. 19 the circuit of Fig. 18 is redrawn

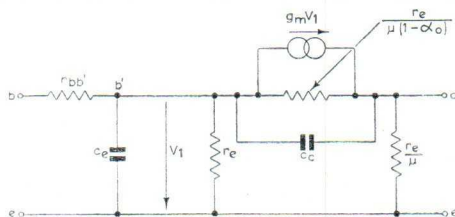


Fig. 19—Hybrid π grounded-emitter equivalent circuit (rearrangement of previous figure).

with the emitter connection common to the input and output; otherwise the two circuits are identical.

Fig. 19 may be transformed further into Fig. 20 by replacing the current generator by two identical generators in series, with their junction connected to the emitter.

The current leaving the point b' and flowing into the generator is the same in both circuits. Similarly the current entering the point c is the same. The current entering the common line e - e from the left-hand generator is extracted by the right-hand generator, and since both generators are by definition infinite impedances, the conditions at e - e are unchanged by connecting the generators. Consequently Figs. 19 and 20 are identical.

In Fig. 20 the left-hand generator is a function of the voltage across its terminals. Therefore it can be replaced by a resistance, which is

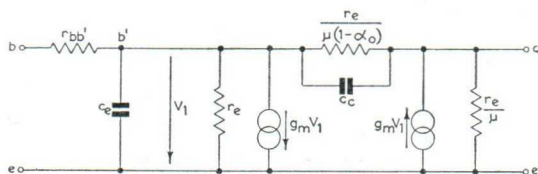


Fig. 20—Hybrid π grounded-emitter equivalent circuit with duplicate current generators.

given a negative sign because of the direction of current flow through the generator. The resistance can be shown to be $-1/g_m$, or $-r_e/\alpha_0$ ohms. This resistance is in parallel with r_e , so the total resistance $r_{b'e}$ between the points b' and e can be worked out from

$$\frac{1}{r_{b'e}} = \frac{1}{r_e} - \frac{\alpha_0}{r_e} = \frac{1 - \alpha_0}{r_e}$$

whence

$$r_{b'e} = \frac{r_e}{1 - \alpha_0}$$

Complete Hybrid π Grounded-emitter Circuit

The hybrid π equivalent circuit, which is used to describe the properties of high-frequency transistors in published data, is shown in Fig. 21, with its values stated in terms of the six basic parameters. In Fig. 22, the hybrid π circuit is shown with the circuit elements labelled with their usual symbols.

VALIDITY OF EQUIVALENT CIRCUITS

With the exception of the low-frequency and high-frequency circuits given in Figs. 3, 14, 15, 16 and 17, all the circuits give useful results up to the α cut-off frequency. Errors are greater, however, at the higher frequencies. The circuits are intended for small-signal operation under normal d.c. operating conditions, for example, as a class A amplifier.

No account has been taken of surface effects, collector-junction

leakage, emitter efficiency and so on. In practice these factors may entail some modification to the equivalent-circuit values.

CHOICE OF EQUIVALENT CIRCUIT

The first complete equivalent circuit (Fig. 4), which includes a transmission line, is the most accurate. It is closely related to the physical operation of the transistor. This circuit and its first approximate version (Fig. 6) are most useful for studying the transistor in its own right, as distinct from using it for particular applications.

For application work any of the following four conventional circuits may be used:

- two-generator grounded-base T circuit (Figs. 10 and 11);
- double-base-resistance grounded-base circuit (Fig. 13);
- grounded-base π circuit (Fig. 18); and
- hybrid π grounded-emitter circuit (Fig. 21).

All these circuits convey the same information with equal accuracy.

For grounded-base operation, the choice between the three grounded-base equivalent circuits will depend on the particular application, and to some extent on the personal preference of the circuit designer.

For grounded-emitter operation, the hybrid π circuit is probably the most generally useful, because the feedback within the transistor

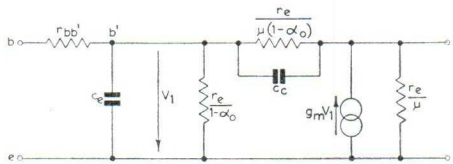


Fig. 21—Values of hybrid π grounded-emitter circuit related to grounded-base circuit of Fig. 10.

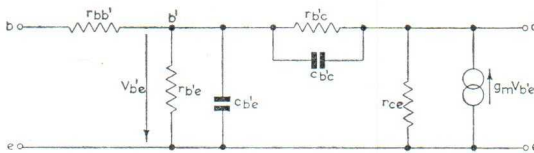


Fig. 22—Usual symbols for hybrid π grounded-emitter equivalent circuit

is expressed in its simplest form, so allowing the performance of the transistor to be estimated from an inspection of the circuit.

The various approximate versions of these circuits can also be extremely useful in circuit design, provided their limitations are assessed

EQUIVALENT CIRCUITS

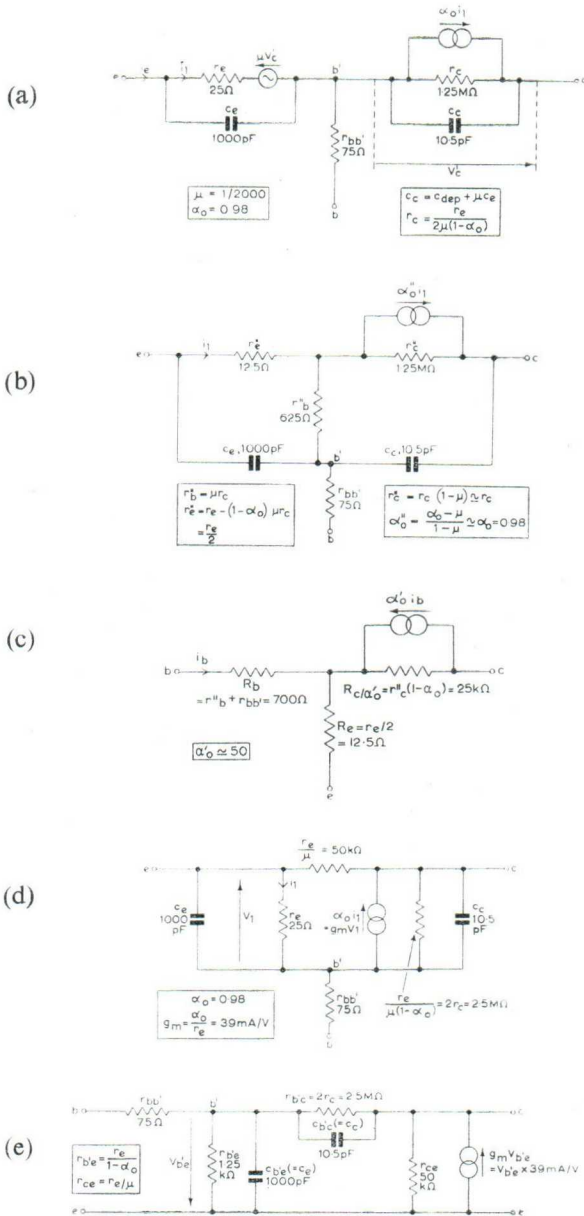


Fig. 23—Representative values inserted on five equivalent circuits

for the particular application. For example, in the design of a low-frequency amplifier having a gain of only a few times, the first approximate equivalent circuit shown in Fig. 3 would normally be quite adequate. At the other extreme, where no simplification is possible, such as in the design of some moderately high-frequency amplifiers, the equivalent circuit merely forms the basis for the calculation of a less complex but frequency-dependent network, which represents the transistor at one frequency only.

NUMERICAL EXAMPLE

As a numerical example, the following values will be assumed for a hypothetical transistor:

$$\alpha_o = 0.98 \quad (\alpha'_o \simeq 50)$$

$$r_e = 25\Omega$$

$$\mu = 1/2000$$

$$r_{bb'} = 75\Omega$$

$$c_{dep} = 10\text{pF}$$

and

$$c_e = 1000\text{pF}.$$

These values have been chosen to give round numbers, and while they do not apply to any particular type, they are sufficiently representative of an r.f. transistor.

Five equivalent circuits will now be written out using the above values (Fig. 23).

(a) The *two-generator grounded-base T circuit* is shown in Fig. 23(a). The values of r_e , c_e , μ , $r_{bb'}$ and α_o can be written directly on to the circuit. The only quantities to calculate are c_e and r_c , thus:

$$\begin{aligned} c_e &= c_{dep} + \mu c_e \\ &= 10 + \frac{1000}{2000} \\ &= 10.5\text{pF}; \end{aligned}$$

and

$$\begin{aligned} r_c &= \frac{r_e}{2\mu(1-\alpha_o)} \\ &= \frac{1}{2} \times 25 \times 2000 \times 50 \\ &= 1.25\text{M}\Omega. \end{aligned}$$

(b) The *double-base-resistance circuit* is shown in Fig. 23(b). On to this circuit we can write immediately:

$$c_e = 1000\text{pF} \quad \text{and} \quad r_{bb'} = 75\Omega.$$

Also, c_e has already been calculated as 10.5pF . The other quantities (using $r_e = 1.25\text{M}\Omega$) are:

$$\begin{aligned} r''_e &= r_e/2 = 12.5\Omega; \\ r''_b &= \mu r_e \\ &= 1.25 \times 10^6 / 2000 = 625\Omega; \\ r''_e &= r_e(1-\mu) \\ &\simeq r_e = 1.25\text{M}\Omega; \end{aligned}$$

and

$$\begin{aligned} \alpha''_o &= \frac{\alpha_o - \mu}{1 - \mu} \\ &\simeq \alpha_o = 0.98. \end{aligned}$$

(c) The *low-frequency T circuit* arranged for grounded emitter is shown in Fig. 23(c). Here the quantities are calculated as:

$$\begin{aligned} R_b &= r''_b + r_{bb'} \\ &= 625 + 75 = 700\Omega; \\ R_e &= r_e/2 = 12.5\Omega; \\ \alpha'_o &= \alpha_o / (1 - \alpha_o) \\ &= 0.98 / 0.02 = 49; \end{aligned}$$

and

$$\begin{aligned} R_c / \alpha'_o &\simeq r''_e(1 - \alpha_o) \\ &= 1.25 \times 10^6 / 50 = 25\text{k}\Omega. \end{aligned}$$

(d) The *complete grounded-base π circuit* is shown in Fig. 23(d). The following values can be filled in straight away from the six basic parameters:

$$c_e = 1000\text{pF}; \quad r_e = 25\Omega; \quad r_{bb'} = 75\Omega; \quad \text{and } \alpha_o = 0.98.$$

Also, $c_e = 10.5\text{pF}$ as calculated previously. Then

$$\begin{aligned} r_e/\mu &= 25 \times 2000 = 50\text{k}\Omega; \\ g_m &= \alpha_o/r_e = 0.98/25 = 39\text{mA/V}; \end{aligned}$$

and

$$\frac{r_e}{\mu(1-\alpha_o)} = 2r_e = 2.5\text{M}\Omega.$$

(e) The *hybrid π grounded-emitter circuit* is shown in Fig. 23(e). The basic parameters on this circuit are:

$$r_{bb'} = 75\Omega; \quad \text{and } c_{b'e} = c_e = 1000\text{pF}.$$

From previous calculations:

$$c_{b'e} = c_e = 10.5\text{pF}; \quad r_{b'e} = 2r_e = 2.5\text{M}\Omega; \quad \text{and } g_m = 39\text{mA/V}.$$

It only remains to calculate:

$$r_{b'e} = \frac{r_e}{1-\alpha_o} = 25 \times 50 = 1.25\text{k}\Omega;$$

and

$$r_{ce} = r_e/\mu = 25 \times 2000 = 50\text{k}\Omega.$$

CUT-OFF FREQUENCIES f_α AND f_1

Introduction: the Idealised Transistor

The complete equivalent circuit shown in Fig. 4 consists of

- (a) a transmission line and a current generator, and
- (b) c_{dep} , μ , and $r_{bb'}$.

The second group of quantities may be classed loosely as parasitic elements, since they are in no way fundamental to the operation of the transistor. They will be neglected, therefore, in this section, and only an ideal transistor consisting of an emitter-base transmission line and a collector-current generator will be considered (Fig. 24).

If r , g and c are respectively the resistance, conductance and capacitance per unit length of the line, and l is the length of the line, then by

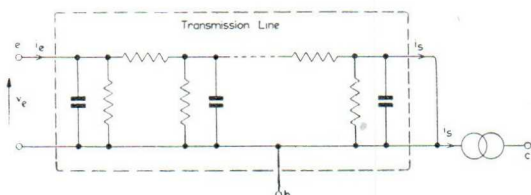


Fig. 24—Transmission line and output-current generator representing idealised transistor.

using the normal methods of network analysis, the general equations for the line may be written:

$$0 = v_e \cosh Pl - i_e Z_0 \sinh Pl$$

$$i_s = i_e \cosh Pl - \frac{v_e}{Z_0} \sinh Pl$$

where

$$P = \sqrt{\{r(g + j\omega c)\}}$$

$$Z_0 = \sqrt{\left\{\frac{r}{g + j\omega c}\right\}}.$$

Several facts may be established from these equations. For example, r_e and c_e may be related to r , g , c and l , and hence, by means of the physical theory of the transistor, to its physical dimensions and constants. This method of attack leads to the equation

$$r_e = \frac{kT_j}{qI_e},$$

on which the approximate expressions for r_e given earlier are based, where k is Boltzmann's constant, T_j is the junction temperature in $^{\circ}\text{K}$, and q is the electronic charge.

Alternatively, r , g , etc. may be dispensed with by relating them to

measurable transistor parameters, such as the current amplification factor and cut-off frequency. It is the latter approach which will be followed here, in order to discuss the current amplification factor in grounded base. The treatment is limited to an idealised transistor, and in practice the parasitic elements, particularly c_{dep} and $r_{bb'}$, necessitate some minor modifications.

Classical Approach to Current Amplification Factor

The classical approach to the effect of frequency on the current amplification factor will now be summarised.

The fundamental equations for the transmission line give the current amplification factor as

$$\alpha = \frac{1}{\cosh Pl}$$

This expression gives rise to the diagram of Fig. 25, in which, for a series of frequencies, the imaginary part of $\alpha \{Im(\alpha)\}$ has been plotted against the real part of $\alpha \{Re(\alpha)\}$, and the points joined up to form a locus. The point at which the magnitude of α (proportional to the length of the vector from the origin to the curve) has fallen by 3dB from its low-frequency value is marked on the locus. The frequency corresponding to this point is called the cut-off frequency f_α , and this can be related clearly to r , g etc. and thence to the physical dimensions and properties of the transistor under consideration. The expression for α may then be written as

$$\alpha \simeq \frac{\alpha_0}{\cosh \sqrt{(2 \cdot 44 j f / f_\alpha)}}, \quad \dots (1)$$

and c_e also may be expressed in terms of f_α .

DIFFICULTIES

The classical approach just outlined, which culminates in the definition of an arbitrary frequency characteristic f_α , leads to several major difficulties:

The expression for the α of an alloy-junction transistor (Eq. 1) is not particularly simple. For drift-field transistors, the transmission-line equations are more complicated and the corresponding expression for α is completely unmanageable.

Drift-field transistors have different loci for α according to the strength of the drift field. The direct significance of f_α , illustrated by Fig. 25, is therefore lost.

At relatively high frequencies, f_α is extremely difficult to measure, and almost impossible to measure with any accuracy.

The relation between f_α and other components of the equivalent circuits, particularly

$$C_{b'e} = \frac{1.22}{2\pi f_\alpha r_e},$$

is reasonably simple for alloy-junction transistors, but considerably more complicated for drift-field transistors.

The relation between f_α and the grounded-emitter cut-off frequency

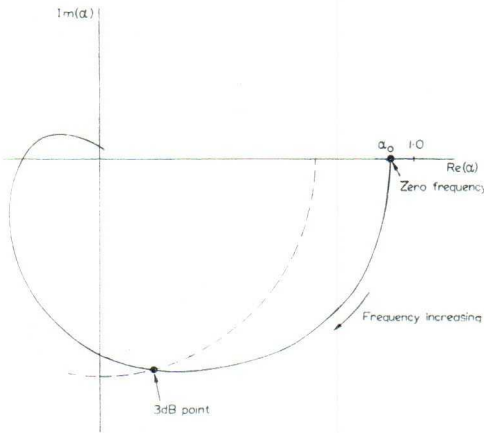


Fig. 25—Locus of α for increasing frequency

f'_α is not straightforward, particularly for drift-field transistors.

These difficulties have become more apparent over the past few years, and a number of proposals have been put forward for overcoming them.

High-frequency Parameter f_1

The present situation will now be outlined by discussing the difficulties in turn.

A relatively simple approximate expression has been found for α , which holds for all practical values of drift field. This is

$$\alpha \simeq \frac{\alpha_0 \exp(-j\phi f/f_\alpha)}{1 + jf/f_\alpha},$$

where ϕ is a constant, the value of which depends on the drift field.

A frequency having more direct significance than f_α is that at which the real part of α is 0.5.

A frequency which is much easier to measure than f_α is that at

which the magnitude of α' has fallen to one. This is called f_1 .

The two frequencies at which $\text{Re}(\alpha) = 0.5$ and $|\alpha'| = 1.0$ are for practical purposes identical. That is, when f_1 is measured, at which the magnitude of α' is one, what is in fact obtained is the frequency at which the real part of α is one-half. From Eq. 1, it can then be shown that

$$f_1 = \frac{f_\alpha}{1 + \phi}.$$

For alloy-junction transistors (which have zero drift field) $\phi = 0.22$, and thus $f_\alpha = 1.22f_1$. For alloy-diffused or other drift-field transistors, $0.22 \leq \phi \leq 1.0$.

The relation between $c_{b'e}$ and the frequency f_1 is now simple and, for all transistors, is independent of the value of the drift field:

$$c_{b'e} = \frac{1}{2\pi f_1 r_e}.$$

A simple relation also holds between f_1 and f'_α , namely

$$f'_\alpha = \frac{f_1}{1 + \alpha'_0}.$$

From Eq. 1, it can be shown that $|\alpha'|$ falls at a rate of 6dB/octave (20dB/decade) at frequencies several times greater than f'_α .

Relative Importance of f_1 and f_α

The alpha cut-off frequency f_α has served well in the past to characterise the high-frequency behaviour of transistors, but it was chosen rather arbitrarily and suffers from some drawbacks. The introduction of f_1 , the frequency at which $|\alpha'| = 1$, has eased the situation considerably.

For certain special purposes, such as the discussion of the detailed behaviour of α , or the consideration of high-frequency noise, it is advantageous to retain f_α , supplemented by f_1 . For many purposes, however, f_1 can be used by itself in place of f_α . In particular, f_1 will come into greater prominence in connection with alloy-diffused and other drift-field transistors.

CHAPTER NINE

LIMITING VALUES

Limits are set to the operating conditions in the interests of the user. The limits provide the circuit designer, whose experience is limited to comparatively small quantities of transistors and short operating periods, with ratings based on systematic life-testing of samples from the whole of the production.

If the limits are exceeded the transistor may fail immediately; alternatively there may be a gradual loss of performance, and the life, normally very long, may be shortened appreciably.

ABSOLUTE MAXIMUM RATINGS

Transistor ratings are absolute. The ratings must never be exceeded.

The circuit designer has to ensure that no transistor is operated outside the limits under any condition of operation and for any period of time. In arriving at the operating conditions, he must take into account: transistor spreads; variations in supply voltage; component tolerances; ambient temperature; and any other relevant conditions.

It is not safe to assume that one rating may be exceeded provided a reduction is made in some other rating. If the collector voltage is low, the collector current cannot necessarily be increased up to the maximum permitted by the rated collector dissipation; in so doing, the collector-current rating itself might be exceeded.

APPROACHING THE LIMIT

In the past, rather large safety factors were allowed in drawing up transistor data, because the users' and manufacturers' experience of transistors could not compare with that built up with valves over a great many years. Safe limits were adopted, and these were increased as more experience became available.

This phase is now largely over. As a result of extensive life-testing, the ratings of the first transistors to be introduced have been increased. Newer transistors are rated as realistically as possible from the start.

The existence of a limiting value may mean no more than that, to

date, the region beyond the limit has not been investigated. But as time has gone on, more and more of the ratings have come to represent real limits of operation, beyond which performance and life are known definitely to be impaired.

DISSIPATION AND TEMPERATURE RATINGS

Most of the current flowing during the normal operation of the transistor consists of an emitter-collector current.

During its passage through the transistor, the emitter-collector current flows through regions of differing resistance, in which it exerts a normal heating effect. Little heat is generated at the emitter junction, which is of low resistance, and most of the heat appears in the region of the collector junction, where the voltage gradient is high.

Provided the base power is small, the total dissipation in the transistor can be taken as being virtually equal to the collector dissipation $V_{ce} \times I_c$. When the base power is not small,

$$P_{tot} = V_{ce}I_c + V_{be}I_b.$$

The heat developed at the collector junction increases the junction temperature, the rise in temperature generally being greater for a smaller than for a larger junction.

If the junction temperature is allowed to rise indefinitely, the transistor will fail catastrophically.

At junction temperatures not sufficiently high to overheat the transistor, there may be long-term effects on life and reliability.

The junction temperature is therefore rated at a certain maximum value, T_{jmax} .

Effect of Ambient Temperature

The heat generated inside the transistor has to be conducted away to the casing and absorbed by the surroundings. In continuous operation the transistor must reach a thermal equilibrium with its immediate surroundings, in which the collector junction remains at some steady temperature above the ambient value.

The junction temperature of the transistor therefore depends on the ambient temperature T_{amb} as well as on the collector dissipation. If the ambient temperature is high, the dissipation must be reduced, in order that the maximum junction temperature shall not be exceeded.

The total permissible dissipation is calculated from

$$P_{tot} \simeq P_c = \frac{T_{jmax} - T_{amb}}{\theta},$$

where

p_{tot} = total dissipation ($\approx$ collector dissipation p_c)

$T_{j\text{max}}$ = maximum junction temperature

T_{amb} = ambient temperature

θ = thermal resistance, expressed as the rise in junction temperature above T_{amb} for every unit of power dissipation.

In this equation, the maximum junction temperature $T_{j\text{max}}$ is a limiting value and θ is a constant.

The permissible dissipation is therefore found from the maximum ambient temperature at which the equipment will be required to operate. For the OC71, $T_{j\text{max}}$ is 75°C for continuous operation,

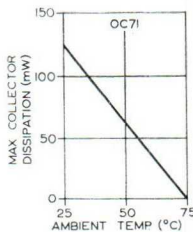


Fig. 1—Permissible collector dissipation of OC71 plotted against ambient temperature.

and θ is 0.4°C/mW . If the equipment is required to operate up to temperatures where the air surrounding the transistor reaches 45°C ,

$$p_{c\text{max}} = \frac{75 - 45}{0.4} = 75\text{mW}.$$

This information is most conveniently given by a straight-line graph, such as Fig. 1, which is for the OC71.

For a power transistor, more efficient means are provided for removing heat. For example, the effective or total θ (from junction to ambient) of an OC23 when mounted on a suitable heat sink may be only 7.5°C/W . For a maximum junction temperature of 90°C ,

$$p_{\text{tot}} = \frac{90 - 45}{7.5} = 6\text{W}$$

at a maximum ambient temperature of 45°C .

Efficient Removal of Heat: the Heat Sink

Heat is removed from the transistor by a combination of conduction, convection and radiation. The transistor manufacturer arranges for

the collector junction to be in the best possible thermal contact with the casing.

A further improvement can be obtained by making thermal contact between the casing and a metal plate, from whose surface heat is removed by convection and radiation. This plate may take the form of a heat sink or a clip-on cooling fin. Often the chassis forms the heat sink, though the effectiveness of this arrangement depends on the chassis temperature.

Unfortunately, substances which are good conductors of heat are also good electrical conductors. In high-power transistors the requirements of good heat conduction and electrical insulation are reconciled

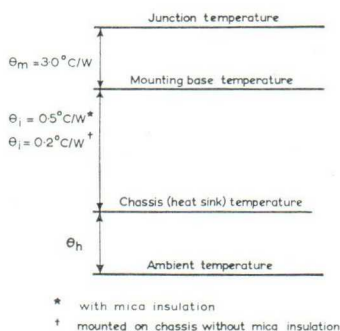


Fig. 2—Thermal resistances of OC22, OC23 and OC24 power transistors

as follows. The collector junction is connected electrically as well as thermally to a metal casing. Where no potentials exist on the heat sink which might affect the collector voltage, the transistor is bolted directly to the heat sink by means of two holes in the mounting base. Otherwise the transistor is insulated electrically from the heat sink (e.g., the chassis) by a mica washer and two bushes. Because the thermal insulation is as thin and as large in area as possible, this method of mounting affords the minimum resistance to the flow of heat.

For a power transistor, θ must be regarded as the total thermal resistance which is composed of three terms (Fig. 2). θ_m is the thermal resistance between the junction and the mounting base. θ_i exists between the mounting base and the chassis (heat sink). θ_h is the thermal resistance between the chassis (heat sink) and the surroundings. So

$$P_{\text{tot}} = \frac{T_j - T_{\text{amb}}}{\theta_m + \theta_i + \theta_h}$$

This equation expresses mathematically the information given by the derating characteristic (Fig. 3).

θ_m is a function of the construction of the transistor; the manufacturer keeps this as low as possible, and the circuit designer or transistor user can do nothing further to reduce it. The value of θ_m is 3.0°C per watt for the OC23. θ_i depends on the electrical insulation provided

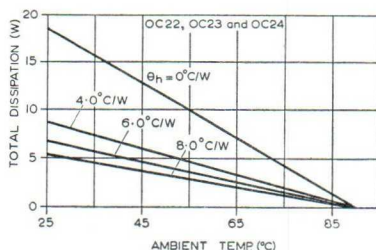


Fig. 3—Permissible total dissipation of OC22, OC23 and OC24 plotted against ambient temperature for various θ_h .

between the transistor and the chassis. With the transistor insulated from the chassis, θ_i is not more than 0.5°C per watt for the OC23.

θ_h depends on the heat sink provided by the user. It is determined mainly by the dimensions, position, material and finish of the heat sink. For example, an air-cooled heat sink may give a θ_h of 4°C/W .

θ_h can be determined for any heat sink from the temperature T_{mb} of the mounting base as measured by a thermocouple. Then

$$\theta_h = \frac{T_{mb} - T_{amb}}{P_{tot}} - \theta_i.$$

The following example illustrates the temperatures at various points of an OC23 power transistor with mica insulation for $p_c = 4\text{W}$, $\theta_h = 4^\circ\text{C/W}$, $\theta_m = 3.0^\circ\text{C/W}$ and $\theta_i = 0.5^\circ\text{C/W}$.

Junction temperature = 90°C

Mounting-base temperature = $90 - (4 \times 3.0) = 78^\circ\text{C}$

Chassis (heat sink) temperature = $78 - (4 \times 0.5) = 76^\circ\text{C}$

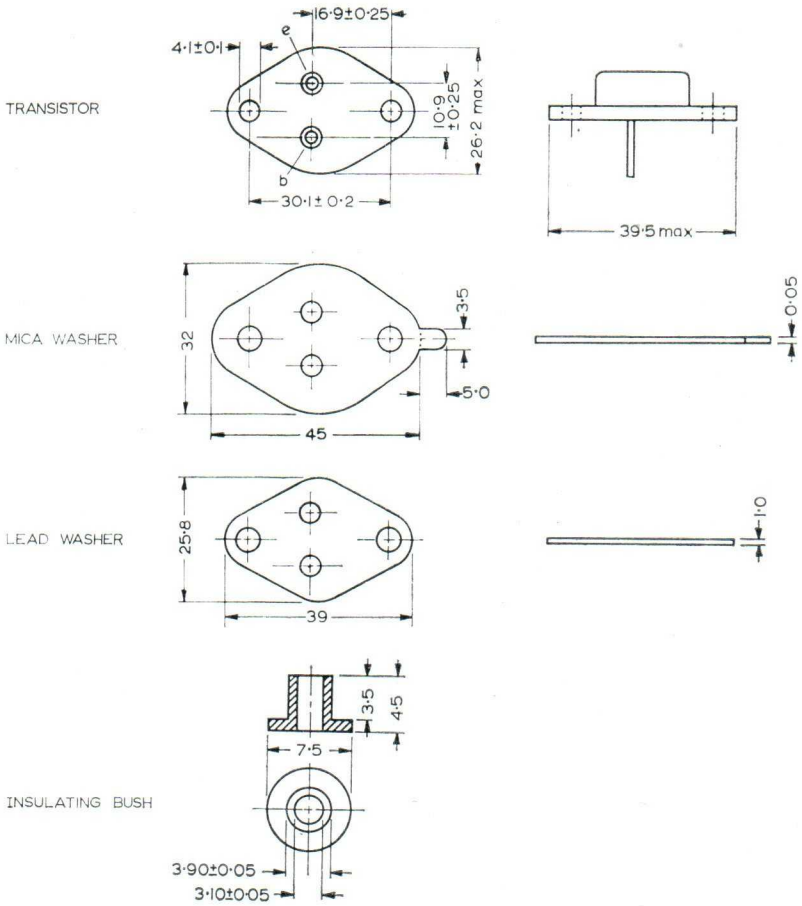
Permissible ambient temperature = $76 - (4 \times 4) = 60^\circ\text{C}$.

The suitability of any design can be checked by measuring the mounting-base temperature with a thermocouple, with the equipment operating at the required total dissipation and maximum ambient temperature. The result is compared with a graph of maximum total dissipation versus mounting-base temperature (e.g. Fig. 5, p. 99). If the point lies above the line, the design is unsatisfactory, and the dissipation must be reduced or the heat sink improved. The total dissipation is interpreted as the maximum reached by any transistor of the type in question.

The effectiveness of the heat sink is partly determined by its position

LIMITING VALUES

relative to other objects. Blackening assists cooling by radiation if there are no other hot surfaces in the vicinity. The effect of obstacles to cooling by convection should be taken into account.



All dimensions in mm

Fig. 4—Dimensions of mounting base, lead and mica washers, and insulating bushes for OC28, OC29, OC35 and OC36.

Good thermal contact is required between transistor and heat sink. The transistor should be bolted down evenly and the heat sink should be flat.

For minimum thermal resistance between the case and the heat

sink, the washer should be smeared with silicone grease (insulating quality). The edges of the mounting holes in the chassis should be free from burrs or thickening, which could puncture the washer or cause uneven contact.

Nothing should be done to impair existing arrangements for removing heat from the transistors or from neighbouring parts of the equipment, when servicing transistorised equipment.

An example of a dimensioned drawing of the washers and insulating bushes is given in Fig. 4.

COLLECTOR-CURRENT RATINGS

In the past, the heating effect of the collector current has been a useful starting point when drawing up the collector-current ratings.

If the usual working voltage of the transistor as an amplifier is x volts, and the maximum collector dissipation y watts, most requirements are met by a rating of y/x amps for the continuous collector

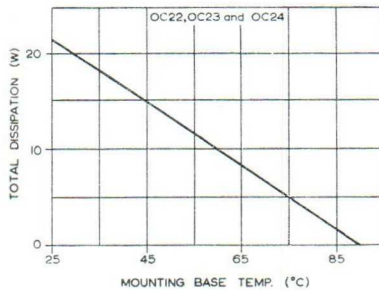


Fig. 5—Permissible total dissipation of OC22, OC23 and OC24 plotted against mounting-base temperature.

current. This reasoning may be confirmed by systematic life-testing of transistors at the current in question. Such a rating is satisfactory for small-signal circuits with the transistor biased into the linear portion of its transfer characteristic.

However, when the transistor is operated in a bottomed condition, as in a flip-flop or other switching circuit, the collector current that can be passed before the dissipation rating is exceeded may be very large. Such a high current can be damaging in itself, apart from its heating effect. To meet the requirements of such circuits, the rating is extended upwards by means of life-testing. Finally a real limit is found, or a rating established which covers all the proposed applications.

In practice the designer may have to restrict the collector current, whether peak or continuous, to less than the rated value, in order to remain within the permissible dissipation or limit distortion.

VOLTAGE RATINGS

A large amount of laboratory and field experience goes into the drawing up of the voltage ratings. The ultimate limit to the voltage ratings, however, is set by the avalanche effect. Avalanche multiplication of the current carriers takes place in the collector depletion layer if the voltage across the layer exceeds some critical value. The critical voltage depends on some operating conditions and not on others. It is determined by the physics of the transistor, and depends on such factors as current density; but it has the same value whether the applied voltage is peak or direct.

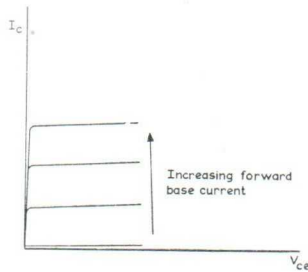


Fig. 6—Usual form of output characteristic

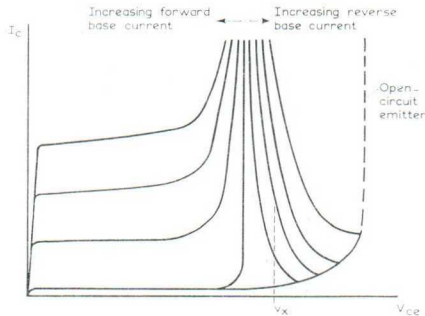


Fig. 7—Output characteristic plotted to higher voltages

Avalanche Characteristics

The output characteristic of a transistor in grounded-emitter connection is usually given in the form shown in Fig. 6. These curves are plotted for increasing forward base current.

If the characteristics for forward base current are plotted for higher collector voltages, the curves shown in the left-hand part of Fig. 7 are obtained. As the collector voltage is increased, the slope resistance of the characteristic falls, until a region is reached where only a slight increase in collector voltage is observed up to very high values of

LIMITING VALUES

collector current. This is called the avalanche region, the voltage across the transistor being the avalanche voltage.

Fig. 7 also shows the characteristics for increasing reverse base current. As the collector current is increased, the characteristic turns back on itself. The voltage across the transistor falls rapidly until the avalanche region is reached.

These curves are for current drive to the base (each of the curves being plotted for some constant value of base current), but the characteristics for voltage drive are similar.

The avalanche characteristics are of particular relevance to switching circuits. In these circuits the transistor is normally switched between: the ON or bottomed condition, in which the forward base current is high; and the OFF condition, in which a pulse is applied in a direction which tends to cut the transistor off, and reverse base current flows.

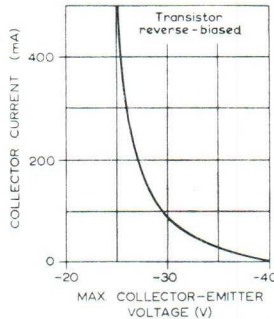


Fig. 8—Collector current plotted against absolute maximum collector-emitter voltage.

A voltage rating curve for a switching transistor is given in Fig. 8. This is for the transistor cut-off. If the collector-emitter voltage is allowed to exceed the values shown, catastrophic failure may occur. The precautions required to prevent failure depend on whether the load is resistive or inductive.

Resistive Load

There is no great difficulty in keeping the working point within the permissible area of operation with a purely resistive load. The resistive load line should be drawn in such a way that it does not intersect the avalanche characteristic, as shown in Fig. 9.

Inductive Load

The behaviour of the circuit is more complicated when the load is inductive, as in relay-switching circuits, d.c. converters and audio

output stages. On switching the transistor off, the rapid decrease in collector current induces a voltage

$$v = L \frac{di}{dt}$$

across the inductance L . Since the current will still be very high during this period, a condition of high current, high voltage exists. If the voltage is not limited by some circuit device (e.g. a catching diode), the working point may intersect the static characteristic for the value of reverse base current used (Fig. 7). If this condition is allowed to occur, the rate of decrease of collector current is given by

$$\frac{di}{dt} = \frac{v_x}{L}$$

rather than by the normal transistor switching time. The time taken to switch off may then be comparatively long (perhaps several milliseconds, depending on the L/R time-constant); during this period the transistor is operating in a high-dissipation region and may be damaged or even destroyed.

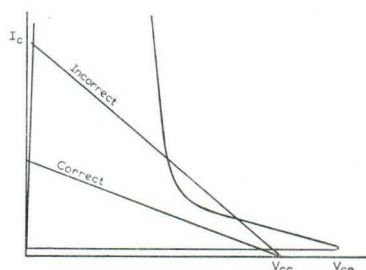


Fig. 9—Correct and incorrect resistive load lines and breakdown characteristic

This condition may be avoided by choosing a suitable transistor for the application and by correct circuit design.

Softening Voltage

The fall in slope resistance before the avalanche region is reached is referred to as softening. The softening voltage is defined as the voltage at which the slope resistance falls to some arbitrary value, and is a function of the external base-emitter impedance. While it is permissible to operate in the soft region of the characteristics—provided precautions are taken to see that avalanche cannot occur—the attendant non-linearity usually leads to excessive distortion. This effect is taken care of in circuit design by means of a characteristic which shows the permissible peak collector voltage plotted against the base-emitter impedance. The permissible voltage falls as the impedance is increased.

CHAPTER TEN

SEMICONDUCTOR DIODES

After the characteristics of the semiconductor diode have been outlined, the various types, such as point-contact, junction etc. will be described.

ADVANTAGES

Semiconductor diodes offer combinations of characteristics not available with thermionic diodes, and they have a number of other advantages.

Life will be long, if the ratings are observed, as there is no heater to fail, no cathode coating to lose emission, and no vacuum to soften.

Hum cannot occur. The saving of heater power may be significant in series-operated equipment. The absence of a heater favours the semiconductor diode in equipment where only diodes are required, or where a heater supply would require long leads.

Small size and weight, and the fact that no holder is required to connect it in circuit, allow the semiconductor diode to be combined with other components. Detector diodes which do not dissipate appreciable amounts of heat can be incorporated in the coil unit.

Interelectrode capacitance is low, and this is an advantage if the load also has a low capacitance.

These advantages bring with them some drawbacks which, in certain applications, make the semiconductor diode unsuitable.

Whereas ideally a thermionic diode passes no current in the reverse direction, appreciable reverse current flows through the semiconductor diode, particularly at higher voltages and temperatures.

The characteristic of the semiconductor diode is noticeably temperature dependent in both the forward and reverse directions.

Semiconductor diodes are hermetically sealed to protect them from atmospheric moisture and soldering contamination. The envelope may be metal, glass, or glass encased in metal.

POLARITY

The semiconductor diode is represented by the symbol shown on the

right of Fig. 1. The 'bar' is connected with the polarity of a cathode and the 'arrow head' with that of an anode.

The thermionic diode on the left of Fig. 1 shows this polarity.

The cathode is the positive output electrode in a rectifying circuit.

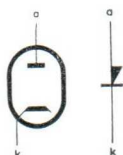


Fig. 1—Polarity of semiconductor diode

The smaller semiconductor diodes are made in either a double-ended or a single-ended construction (Fig. 2). In the double-ended construction, a coloured band is placed on the end of the body nearer to the cathode; alternatively the whole of the tip may be coloured. In the single-ended construction, a coloured spot is placed on the same side of the body as the cathode.

DIODE CHARACTERISTIC

The form of the characteristic of the semiconductor diode can be seen from Fig. 3. The dotted curve shows what is predicted from theory. The characteristics of the various types of semiconductor diode will

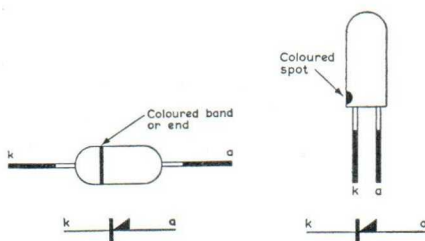


Fig. 2—Electrode connections of small double-ended and single-ended diodes

diverge to a greater or lesser extent from the theoretical, but the full-line curve can be taken as showing their general shape.

Forward Characteristic

The forward characteristic at first glance resembles that of a thermionic diode, but there are several important differences. The characteristic for the semiconductor diode rises comparatively fast, corresponding

to a lower forward resistance. The forward current for a voltage drop of $\frac{1}{2}$ V to 1 V ranges from a few milliamps for certain types to tens or hundreds of milliamps for others.

At low currents the characteristic follows an exponential rather than a three-halves power law. At high positive voltages, well beyond the working range, the characteristic approaches a straight line; there is no saturation effect.

Reverse Characteristic

A negative voltage produces a small negative current. The reverse characteristic is plotted in the opposite quadrant to the forward

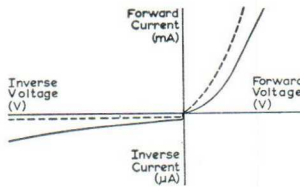


Fig. 3—General form of forward and reverse characteristics. Dotted line theoretical, full line practical.

characteristic, the scale of the reverse current being in microamps, while the forward current is usually plotted in milliamps. The reverse current is sometimes referred to as the leakage current, and is similar to the leakage current through the collector junction of a transistor.

The inverse current increases rapidly at first until the low-voltage 'knee' is reached. Then the current increases gradually until, at a point called the turnover voltage, there is a rapid increase and the characteristic may turn back on itself. If the current is not limited in the turnover region, and the maximum dissipation is exceeded, the diode overheats and is permanently damaged.

The rating for the peak inverse voltage is of fundamental significance and should be observed strictly. Excess voltages lasting even a few microseconds may be disastrous.

Zero Region

The characteristic passes through a true zero (no voltage, no current). In measuring circuits, compensation for the starting current, which is often necessary with a thermionic diode, is not required.

In the zero region (± 10 mV) the characteristic is approximately linear and, as the forward and reverse resistances are nearly equal,

the detection efficiency at this level is low. The detection (or rectification) efficiency in a specified circuit at a given peak input voltage is

$$\eta = \frac{\text{d.c. output voltage}}{\text{peak input voltage}} \times 100\%$$

DERATINGS

All measurements and ratings only apply to the temperatures at which they were made or specified. At higher temperatures, it is necessary to derate the permissible forward current and sometimes the maximum inverse voltage.

A further derating to the forward current is required if appreciable inverse voltages occur during any part of the cycle. Although the

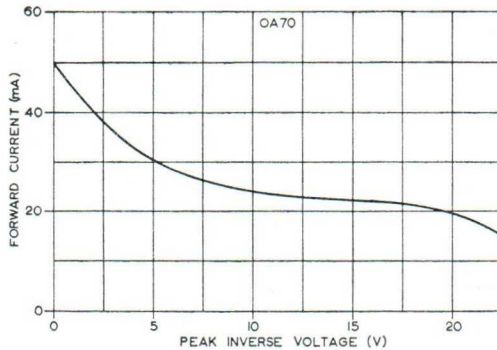


Fig. 4—Mean forward current (averaged over any 50 millisecc period) plotted against peak inverse voltage for the OA70 at ambient temperatures up to 25°C.

reverse current is much lower than the forward current, the high inverse voltage may give appreciable dissipation in the reverse direction.

Derating information for the OA70 is given in Fig. 4. At ambient temperatures above 25°C (up to the permitted maximum of 75°C), the maximum mean forward current should be multiplied by $25/T_{amb}$. Similar information is given for other types in the published data.

Disregarding the rating and derating information may cause the diode to run away and break down.

It is not to be assumed that the semiconductor diode is excessively sensitive to temperature and inverse voltage. The dangers have been mentioned to draw attention to the deratings required. This aspect of circuit design should be given as much attention as is given to correct h.t. and l.t. voltages in ordinary valve practice.

CARRIER STORAGE

One extremely important aspect of diode performance does not appear on the d.c. characteristic curve.

The semiconductor diode usually consists of a strongly p-type anode and a slightly n-type cathode. When the diode is conducting in the forward direction, holes are injected into the cathode. On reversing the polarity, the holes remaining in the 'cathode' flow back into the 'anode'. Thus the current through the diode does not fall to the leakage

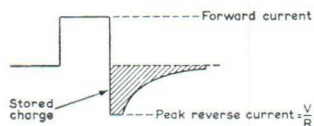


Fig. 5—Stored charge and reverse current

value as soon as the voltage across it is reversed, but a pulse of current flows for a short time in the reverse direction, until the current settles down to the leakage value (Fig. 5).

This effect results from the storage of current carriers. The effect is normally referred to as hole storage.

The peak inverse current is equal to the applied inverse voltage divided by the resistance of the external circuit.

The diode behaves as if an additional capacitance were connected across the junction of its two elements. This is called the storage capacitance, to distinguish it from the interelectrode or reverse capacitance of the diode biased in its reverse direction. The concept of storage capacitance is, however, not very helpful, as the capacitance is non-linear.

Carrier storage may be expressed quantitatively in terms of the inverse current flowing in the circuit a certain time after the forward current has been removed. For a typical OA86, when a forward current of 30mA is removed and an inverse voltage of -35V applied to the diode through a resistance of $2.5\text{k}\Omega$, the inverse current decays to $380\mu\text{A}$ after 0.5 microseconds and to $36\mu\text{A}$ after 3.5 microseconds. The OA86 is suitable where low carrier storage is essential, as in some computer applications.

Alternatively the carrier-storage characteristic may be quoted as the time required for the transient current to fall to a certain value. This time is the 'recovery time'. For the OA10, when a forward current of 10mA is removed, and an inverse voltage of 7V applied to the diode through 300Ω , the time taken for the inverse current to decay to 0.5mA is typically $0.18\mu\text{sec}$.

Two recovery times may be observed in practice. Most of the stored charge decays within a short time, but a longer term effect can be detected at low levels which is sometimes important.

The recovery time is meaningful only if the circuit, the forward current, and the switching time of the pulse applied to the diode are quoted. It is misleading to compare recovery times without at the same time considering the conditions of measurement.

There is a good deal of discussion at present as to the best method of measuring carrier storage. It is not really possible to infer the performance in every circuit from one measurement of carrier storage in

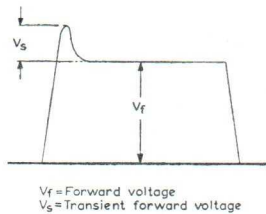


Fig. 6—Turn-on transient forward current

one standard circuit, but the information in the published data represents the practical requirements as closely as possible.

The recovered charge may be quoted in future. This is equal to the area under the reverse peak, the shape of the peak being determined by the circuit resistance.

Turn-on Transient Forward Voltage

During the build-up of stored charge in the base region, the resistance of the base decreases.

When a pulse of forward current is passed through the diode, the forward voltage is initially higher by an amount V_s than the steady forward voltage V_f (Fig. 6). This effect is known as the forward recovery of the diode. For the OA10, the transient forward voltage V_s is less than 0.2V for a forward current of 400mA and a rise time of 0.16 μ sec.

POINT-CONTACT DIODES

The point-contact diode was the first of the semiconductor diodes to be developed. It is still of great importance in circuit design, and is firmly entrenched in a number of applications, where it offers a performance not available with newer forms of construction. It is unlikely to suffer the fate of the point-contact transistor, which has been rendered obsolete by the junction technique.

The point-contact diode is the modern form of the crystal detector (cat's whisker) used in the simplest radio receivers, and was developed originally for operation at radar frequencies, where the low inter-electrode capacitance and short 'transit time' provided a great improvement over the thermionic diode.

Structurally the point-contact diode consists of a springy tungsten wire which presses against a crystal of germanium. During manufacture a current is passed between the whisker and the crystal to 'form' the diode. A small region of p-type material is formed beneath the point of the whisker. Rectification takes place between this region and the remaining material which is n-type.

Although the point-contact diode is effectively a junction diode of small interface area, there are considerable differences. At low reverse voltages the current in the point-contact diode approaches the theoretical value, but the slope of the characteristic thereafter rises up to the turnover voltage. Generally speaking the forward voltage across the point-contact diode is comparatively high. On the other hand, the reverse capacitance and usually the carrier storage are lower than for the junction diode. The temperature sensitivity of the two types is similar.

The point-contact diode is generally the most suitable of the semiconductor diodes for r.f. detectors and mixers.

High and Low Reverse Voltages

Point-contact diodes can be produced with a variety of characteristics. Heavily doping the germanium reduces the maximum reverse voltage. Lightly doping the germanium makes for a higher reverse-voltage rating.

A heavily doped diode can operate efficiently as a detector up to very high frequencies but has a low reverse-voltage rating. At the other extreme, a p.i.v. rating of over 100V can be provided in this technique.

The OA70 is suitable for operation as a detector up to 100Mc/s; the OA81 is a high-voltage diode for general applications requiring the higher reverse-voltage rating and is suitable, for example, for shunting relay contacts (either to suppress sparking or to protect a transistor from voltage surges).

The OA85 has a lower voltage drop than the OA81. The OA86, for which the maximum p.i.v. is 90V, has better hole-storage characteristics than the OA85 and is more suitable for computer circuits.

What are essentially miniature versions of the OA81 and OA85 are provided by the OA91 and OA95.

A reverse-voltage rating intermediate between the two extremes

(45V) is provided by the OA79. This diode is suitable for sound detection, and is available in matched pairs (under the type number 2-OA79) for use in the discriminator or ratio detector in f.m. reception (Fig. 7).

Applications of Point-contact Diodes

VIDEO DETECTOR

Fig. 8 shows a typical video-detector circuit for operation at 30Mc/s. The OA70 should be used in this circuit; the recharging time for the

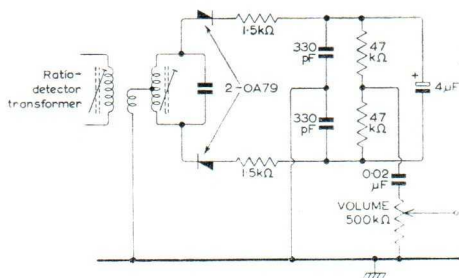


Fig. 7—Ratio detector using 2-OA79

10pF capacitor is short when operating at 30Mc/s, and this diode has a low forward resistance (about 10mA will flow for 1 volt drop) which will allow a substantial charging current to flow. The 100kΩ reverse resistance of the OA70 is more than sufficient to prevent the capacitor

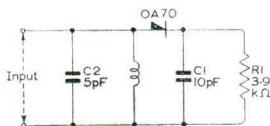


Fig. 8—OA70 as video detector at 30Mc/s. Peak input voltage 5V, damping resistance 3kΩ.

from discharging back through the diode, instead of through the 3.9kΩ load.

The rectification efficiency in this circuit is at least 54% for an OA70 at 30Mc/s, with a peak signal of 5V.

SOUND DETECTOR

An extended reverse characteristic is essential in some applications. For example, in a sound-detector circuit the 3.9kΩ load resistor of

Fig. 8 would be replaced by, say, $47\text{k}\Omega$, and the choice of diode lies between the OA70, OA79 and OA81, depending on the peak inverse voltage which will be encountered and on the value of the load resistor.

Detection in a transistor receiver is preferably performed by an

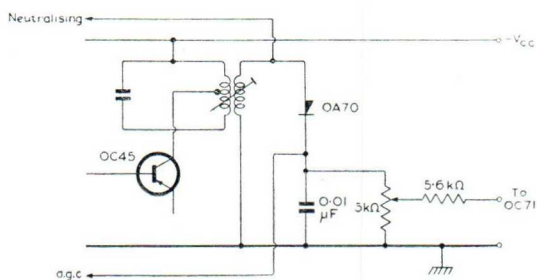


Fig. 9—OA70 as sound detector in transistor receiver

OA70 and not by a transistor (Fig. 9). The final i.f. amplifier is an OC45 and the first a.f. amplifier is an OC71.

TELEVISION NOISE LIMITER

The noise limiter shown in Fig. 10 requires a diode with a high reverse resistance, such as the OA81. A small current flows through the chain of

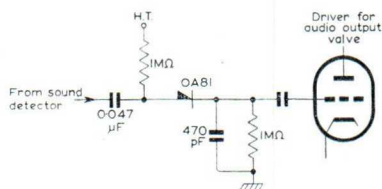


Fig. 10—OA81 as television noise limiter

$1\text{M}\Omega$ resistors and holds the diode in its conducting region. The diode therefore provides a path for normal audio-frequency signals. Interference, however, drives the diode into its reverse-current region, where the high reverse resistance virtually open-circuits the signal path.

GOLD-BONDED DIODES

The gold-bonded diode is structurally a point-contact diode in which the usual tungsten-wire contact is replaced by one made of gold. The result is effectively a small-area junction diode. The characteristics are a cross between those of the point-contact and junction diodes, the reverse capacitance being low and the forward current high. The great

improvement in front-back impedance ratio, and the low voltage at which forward conduction starts, make the device of particular interest.

The OA5 is in a single-ended construction. The forward-voltage drop is 0.4V at 10mA, while the capacitance of 1pF is very little more than that of a conventional point-contact diode. The maximum inverse voltage is 100V at 25°C and 50V at 75°C. Some of the possible applications are: in ring modulators for telephony; as a catching diode; and in transistor computing circuits.

The single-ended OA7 and the double-ended miniature OA47 have essentially the same electrical characteristics as one another. They are for use with transistors operating at peak currents of up to 50mA in the 'logic' circuits of computers operating at pulse-repetition rates of up to 1Mc/s.

JUNCTION DIODES

A junction transistor consists of two diodes formed on a base wafer, whilst a junction diode consists essentially of one such diode.

Comparison of Silicon and Germanium Junction Diodes

The properties of the junction diode are determined largely by whether it is made from germanium or silicon.

Silicon gives a considerably higher reverse-voltage rating, and is preferable for power rectifiers for medium and high voltages. Germanium gives a lower forward-voltage drop, and higher efficiency in a power rectifier for low voltages.

A higher junction-temperature rating can be realised with silicon than with germanium, so that this, too, makes silicon more suitable for high-power rectifiers.

As a rule the silicon diode has the lower reverse current, and the germanium diode the better high-frequency performance.

Characteristics of the Junction Diode

The forward volt drop of a junction diode is low—the forward current being about 100mA at 0.5V for an OA10. The reverse current remains low up to quite high voltages, but increases sharply if the maximum rated voltage is exceeded. The characteristics of most junction diodes approach the theoretical more closely than do those of point-contact diodes.

In addition to the storage capacitance, the capacitance of the diode when biased in the reverse direction may be important when the diode

is used with tuned circuits operating over a range of signal levels. The interelectrode or reverse capacitance usually varies inversely as the square root of the voltage.

Both forward and reverse currents increase substantially with temperature. The reverse-voltage rating usually decreases with increasing temperature.

Junction diodes are more robust than those manufactured by the point-contact technique and are generally able to stand higher levels of shock and vibration.

The OA10 is designed to have sufficiently low hole storage for square-loop-ferrite coupling circuits, and in consequence has a com-

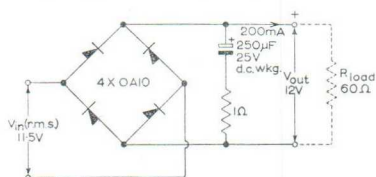


Fig. 11—Four OA10 in low-voltage bridge rectifier. Peak current 700mA

paratively low reverse-voltage rating (30V). This diode is suitable for low-voltage rectification (Fig. 11).

The OA200 and OA202 are miniature general-purpose silicon diodes.

Silicon junction rectifiers are available with very high reverse voltage ratings for rectification at currents of several hundred milliamps. Large metal-case rectifiers designed for bolting down to a heat sink commonly have rather lower voltage ratings but can rectify currents of 15A and upwards.

ZENER DIODES

A silicon rectifying junction has a more sharply defined 'knee' in its reverse characteristic than a germanium junction. Beyond a certain point the current increases sharply with voltage, the dynamic impedance in this region often being only a few ohms.

The silicon Zener diode exploits this part of the characteristic and provides a nearly constant voltage over a reverse current range of tens of milliamps. It is used as a voltage stabiliser or voltage limiter.

Two factors determine the usable current range. The lower limit is set by the fact that at low currents the slope resistance increases and the voltage falls. The upper limit is set either by the dissipation or by the maximum-current rating.

Zener diodes possess a normal forward characteristic.

Zener diodes can be specified as a series of types covering a range of nominal Zener voltages, the individual types having a narrow or wide tolerance on the particular values. Thus Zener diodes are available under the type numbers OAZ200 to OAZ207 and OAZ208 to OAZ213. The OAZ200-207 cover progressively higher nominal voltages from 4.7V (OAZ200) to 9.1V (OAZ207) with a tolerance on the nominal voltage of approximately $\pm 5\%$. The OAZ208-213 give a coverage from 4.2V (OAZ208) to 12.2V (OAZ213) with a tolerance of approximately $\pm 15\%$.

CHAPTER ELEVEN

THE PHOTOTRANSISTOR

The phototransistor is a junction transistor, manufactured in such a way that the inherent photoelectric properties are fully exploited. It is essentially a photodiode, in which the light current is amplified by normal transistor action.

PHOTODIODE

The mobile current carriers on the two sides of an isolated p-n junction will diffuse across the junction, until a counter-e.m.f. is built up which limits the further exchange of carriers. When a reverse voltage is applied to the diode, the barrier potential is increased correspondingly until equilibrium is again reached. Under these conditions, some carriers receive sufficient random energy to cross the barrier, and constitute the normal leakage current. This is the dark current.

If light is allowed to fall on the junction, hole-electron pairs are created on both sides of the junction. The barrier potential sweeps the holes one way and the electrons the other. The current now flowing is the light current, equal to the dark current plus the photoelectric current.

PHOTOTRANSISTOR

The phototransistor operates like a normal transistor into which light is allowed to enter. The hole-electron pairs generated by the incident light should preferably be created as near to the collector junction as possible, in order to minimise recombination. Hole-electron pairs stand very little chance of contributing to the light current, if they are generated more than one diffusion length (usually about 0.5mm) from the collector junction.

With the base open-circuit (Fig. 1), the leakage current I_{co} , which is only a few micro-amps, is transformed by transistor action into I'_{co} , which is equal to $(1 + \bar{\alpha}')I_{co}$. The dark current with the base open circuit is substantially equal to I'_{co} .

When light falls on the phototransistor, the photoelectric current I_{ph} contributes an internal base current which, if the base is open circuit, is amplified by the factor $(1 + \bar{\alpha}')$ into a much larger current at the collector. The light current in this circuit is approximately

$(1 + \bar{\alpha}')I_{ph} + I'_{co} = (1 + \bar{\alpha}')(I_{ph} + I_{co})$. Provided the photoelectric current is large in comparison with I_{co} , the light- to dark-current ratio is high.

Normally, the phototransistor is operated with a resistance between base and emitter, to ensure thermal stability at higher temperatures. The dark current (I'_{co}) is then reduced considerably below its value

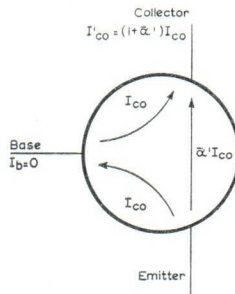


Fig. 1—Relation of I'_{co} to I_{co}

with open-circuit base, resulting in a greatly improved light- to dark-current ratio.

OCP71 PHOTOTRANSISTOR

The OCP71, which operates on the principles just described, is a

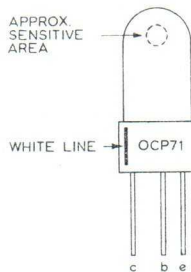


Fig. 2—Base connections of OCP71

general-purpose device of p-n-p alloy-junction construction. The base connections are as shown in Fig. 2.

Sensitivity and Response

Maximum spectral response occurs at a wavelength of 1.55μ , and 50% response is obtained at about 0.8μ and 1.6μ . (The micron [μ] is 10^{-6} of a metre, and is equal to $10,000\text{\AA}$.) The peak response is in the near

infra-red, but continues through the visible spectrum to the near ultra-violet.

Light falling on the emitter side of the crystal creates hole-electron pairs in the parts of the base not in the shadow of the emitter (Fig. 3). The chief contribution to the total light current is made by hole-electron pairs which have only a short distance to diffuse before reaching one of the junctions. The emitter junction of the OCP71 is smaller than the collector junction. The hole-electron pairs generated in an annulus ('doughnut'), the inner and outer boundaries of which are formed by the emitter and collector, are therefore most effective.

If the light is incident on the collector side of the crystal, the collector shades the annulus bounded by the emitter and collector. Hence fewer

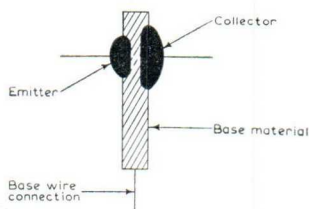


Fig. 3—Diagrammatic cross-section of OCP71

hole-electron pairs are created, and recombination is greater, and the total light current is very much smaller. However, the moistureproof grease in which the assembly is coated diffuses the light, and the device is not critically sensitive to its direction.

Maximum current is obtained when the light is allowed to fall on the side of the bulb bearing the type number (Fig. 2), in a direction perpendicular to the plane of the leads. The response is about 50% of the maximum, when the light is incident at 90° to this direction in the horizontal plane. If the light falls vertically downwards, the response is still 30% of the maximum.

Output Current

The output (I_e/V_e) characteristic of the OCP71 is plotted for a number of intensities of incident light. The intensity, which is specified in foot-candles (= lumens/ft²), has to be calculated in order to interpret this characteristic.

The amount of light (in lumens) falling on any surface is given by

$$\frac{\text{Candlepower of source} \times \text{Area of surface}}{(\text{Distance of surface from source})^2}.$$

For example, if a 40W tungsten-filament electric lamp (which gives

approximately 30 candlepower) is placed 120mm from a surface 7mm² in area, the flux at the surface is

$$\frac{30 \times 7}{120^2} = 0.015 \text{ lumen.}$$

An OCP71 phototransistor has an effective sensitive area of about 7mm². If used with a 40W unfocused lamp, as in this example, the OCP71 will pass a current of 4.5mA, since the sensitivity is about 300mA/lumen.

The above example using unfocused light does not give a very clear idea of the high sensitivity of the OCP71. A more vivid illustration is provided by a 2½V pea-lamp, under-run from a supply of 1½V. If the barely glowing filament is several centimetres away from the OCP71, and the light is focused by a simple glass lens on to the sensitive area, the current is at least 5mA.

Collector Dissipation

The junction temperature of the OCP71 must not be allowed to exceed 65°C. Junction temperature, ambient temperature and collector dissipation are related in the usual way. The thermal resistance is 0.4°C/mW, and the permissible collector dissipation for any maximum ambient temperature is found from

$$p_{c\max}(\text{mW}) = \frac{T_{j\max} - T_{\text{amb}\max}}{0.4}.$$

To take a specific example, if the circuit is intended to operate up to an ambient temperature of 45°C,

$$p_{c\max} = \frac{65 - 45}{0.4} = 50\text{mW}.$$

Precautions should be taken where necessary to prevent thermal runaway, and d.c. stabilisation should be provided as required (Chap. 7). Without some form of d.c. stabilisation or an external base resistance, the ambient temperature must be restricted to 25°C, at the maximum permissible collector voltage for the OCP71 of 25V.

Circuit Design

The circuits which may be employed are extremely simple, and may amount to nothing more than the phototransistor connected in series with a relay coil and a d.c. supply of 12 to 18V, with the base left unconnected. If it is required to operate the phototransistor over a wider range of temperature or from a high voltage, as in most industrial applications, a resistor should be connected between base and emitter, to reduce the dark current and hence improve the thermal stability. A suitable value may be of the order of 5kΩ.

Fig. 4 shows a basic circuit for d.c. (unmodulated light). This circuit

is adequate for 'on-off' applications. With a base-emitter resistor of $5\text{k}\Omega$, the light- to dark-current ratio may be, for example, 480 at 25°C and 20 at 45°C . Without this resistor, the corresponding values may be 90 and 9.5. The resistor should preferably be an n.t.c. type; the exact value depends on the maximum ambient temperature and the light level.

Fig. 5 includes an OC201, which is used as a simple d.c. amplifier following the OCP71 to give extra sensitivity. This circuit has temperature compensation.

Maximum dissipation in the transistor occurs at half of full drive. Maximum power is available to operate the relay when the transistor

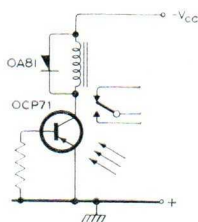


Fig. 4—Basic circuit for unmodulated light

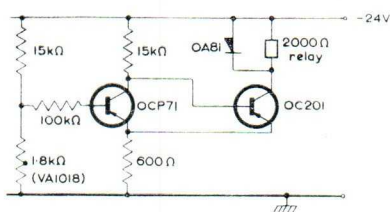


Fig. 5—Temperature-compensated circuit with d.c. amplifier

is bottomed, under full drive. For example, consider a phototransistor operating from a 20V supply. A relay of $5\text{k}\Omega$ is in the collector. Sufficient light is available to bottom the phototransistor and 4mA flows in the collector. Thus:

$$\text{Power in the load} \simeq 4\text{mA} \times 20\text{V} = 80\text{mW};$$

Max. power in the phototransistor on switching

$$\simeq \frac{4}{2}\text{mA} \times \frac{20}{2}\text{V} = 20\text{mW};$$

Phototransistor dissipation when 'bottomed'

$$\simeq 0.15\text{V} \times 4\text{mA} = 0.6\text{mW}.$$

To ensure reliable operation, it may be necessary to choose a relay which will pull in at a power rather lower than four times the maximum power in the transistor.

Where temperature stability is important and a 'chopped' light source is available, the circuit of Fig. 6 can serve as a basis for design. The base is returned to the emitter through an inductance, which is preferably parallel-tuned to the light-modulation frequency. For the modulation frequency, the base is essentially open-circuit, and maximum

THE PHOTOTRANSISTOR

amplification of the light signal is obtained. For d.c. (dark current, or current due to unwanted background illumination), the base impedance is so low that the amplification of this current is reduced to a minimum. Stabilisation is provided by the potential divider and by the bypassed resistance in series with the emitter.

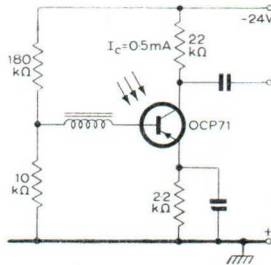


Fig. 6—Basic circuit for modulated light

The cut-off frequency for a phototransistor is the modulation frequency at which the gain is reduced 3dB below the value with unmodulated light. The typical value is 3kc/s for the OCP71 in the circuit shown in Fig. 6.

If the emitter resistor is unbypassed, a.c. feedback will be introduced. The overall gain will be considerably lower, but the frequency response will be considerably extended.

Applications

Special features of the OCP71 are its small size, low operating voltage, high sensitivity, relatively large sensitive area, robust construction, ability to respond to infra-red radiation, and quick response.

The OCP71 is suitable for a variety of industrial-control and other applications such as:

- Photoelectric counters
- Speed measurement
- Liquid-level controls
- Edge detection (in paper making, textiles, and belting)
- Burglar alarms
- Door opening
- Curve followers
- Smoke detection
- Industrial on/off controllers.

CHAPTER TWELVE

AUDIO AMPLIFIER STAGES

In designing a transistor amplifier, it is more convenient to consider the early stages as current amplifiers rather than as voltage or power amplifiers. First the output stage is designed and the drive requirements are determined. The driver stage is then designed, and a sufficient number of amplifier (or preamplifier) stages added to give the required sensitivity.

The stages of the amplifier fall roughly into three groups:

- (a) Low signal (1st stage and possibly 2nd)
- (b) Medium signal (middle stage or stages)
- (c) Large signal (output stage, and possibly the driver).

Case (b) is the simplest, and (a) and (c) are extreme cases of (b).

The power output and driver stages are covered chiefly by the two following chapters. This chapter will be devoted primarily to the amplifier stages, but will include some general remarks which apply to audio amplifiers as a whole.

The emphasis will be on RC coupling, as this is almost always used in practice for audio amplifier stages, and is most convenient for those who want to design their own circuits.

CIRCUIT CONFIGURATION

Common-emitter connection is almost always used for low-level a.f. amplifiers, because of the high power gain obtainable in this configuration. Either transformer or RC coupling is suitable.

RC coupling is preferable for a multistage amplifier, since otherwise instability may be introduced by magnetic coupling through the transformers. Also, the design is simpler for RC coupling, the components are readily available, and the final equipment is lighter and less bulky.

Even when an additional stage is required to achieve the same gain with RC as with transformer coupling, the cost of the equipment is not likely to be increased.

D.C. STABILISATION

By means of d.c. stabilisation, satisfactory performance can be assured for any transistor of a given type over a specified temperature range. Reference should be made to Chapter 7, where the stability of various

circuits and the influence of each component are described. The present section gives a practical method for calculating the maximum and minimum collector currents over the range of operating temperature.

As was shown in the earlier chapter, the preferred biasing circuit for common-emitter connection is that of Fig. 1. The supply is assumed

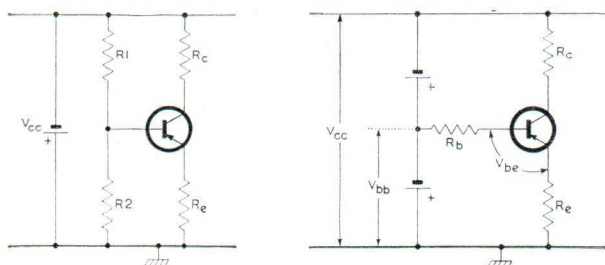


Fig. 1—Conventional potential-divider and emitter-resistor circuit

Fig. 2—Two-battery circuit equivalent to preceding circuit

to be a single battery, as is normal practice. The two-battery circuit, which is a more basic biasing arrangement, is shown in Fig. 2. Both these circuits are reducible to the same equivalent circuit.

From Fig. 2,

$$V_{bb} = I_b R_b + V_{be} + I_e R_e \quad \dots (1)$$

where R_b is the resistance in series with the base. For the one-battery circuit,

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

and

$$V_{bb} = \frac{V_{cc} R_2}{R_1 + R_2}$$

The collector current is given by

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

and since numerically

$$I_c = I_e - I_b,$$

$$I_b = I_e(1 - \alpha) - I_{co}.$$

Substituting for I_b in Eq. 1 and rearranging gives

$$I_e = \frac{V_{bb} - V_{be} + I_{co} R_b}{R_e + R_b(1 - \alpha)} \simeq \frac{V_{bb} - V_{be} + I_{co} R_b}{R_e + R_b/\alpha'} \quad \dots (2)$$

V_{bb} , R_b and R_e are under the control of the circuit designer, but variations in battery voltage, and resistor tolerances, necessarily

introduce a spread in emitter current. V_{be} , α' and I_{co} are properties of the transistor, and of these I_{co} and to a lesser extent V_{be} are temperature dependent. Hence transistor production spreads and changes in temperature introduce a further spread in emitter current.

Maximum and minimum values of emitter current can be found by substituting the appropriate extreme values in Eq. 2.

The changes in I_{co} and V_{be} both increase the emitter current with temperature. Maximum emitter current occurs at maximum ambient temperature, and conversely.

V_{be} decreases with temperature at the rate of roughly $2\text{mV}/^\circ\text{C}$, so that the value given by the characteristic curves, which normally are plotted for an ambient temperature of 25°C , must be corrected for other temperatures.

I_{co} increases exponentially with temperature, and its value can be found from a graph, such as Fig. 3 for the OC71.

Operating Current and Voltage

The maximum and minimum emitter currents must be such that sufficient peak voltage and current is available to drive the next stage, and the values of R_b and R_e have to be chosen accordingly. The circuit

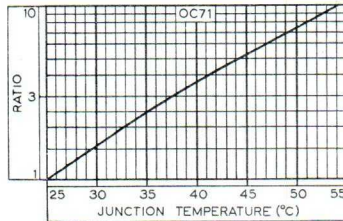


Fig. 3—Ratio of I_{co} at a given junction temperature to its value at 25°C

should not be designed for peak currents or voltages greater than the required values, or the gain will be unnecessarily low.

The OC70, OC71 and OC75 should not be operated in the region below 0.3mA , where the non-linearity of the α'/I_e characteristic produces excessive distortion. The current drive available is

$$i_{c(pk)} = I_c - 0.3 \text{ [mA]}.$$

There is no advantage in providing a higher current than required.

The voltage drive available is given by

$$V_{c(pk)} = V_{cc} - V_{knee} - I_e R_e - I_e R_c.$$

V_{knee} , the knee voltage, is the minimum collector-emitter voltage

for satisfactory transistor action. In class A amplifier circuits, the transistor is 'bottomed' when the base current is such that the collector-emitter voltage is equal to the knee voltage. Any further increase in base current does not influence the collector current.

MAXIMUM DISSIPATION AND THERMAL RUNAWAY

The collector current increases with junction temperature. Whether or not the increase in collector current causes an increase in dissipation depends on the circuit conditions.

If the collector current of the transistor is varied, while keeping the values of the collector and emitter resistors fixed, the collector dissipation is a maximum when the voltage across the transistor, V_{ce} , is equal to half the supply voltage. The collector dissipation is then

$$\left[\frac{V^2}{R} \right] = \frac{V_{cc}^2}{4(R_c + R_e)}.$$

If the operating conditions are such that V_{ce} is greater than $\frac{1}{2}V_{cc}$, then an increase in I_c causes an increase in the collector dissipation. If V_{ce} is less than $\frac{1}{2}V_{cc}$, then an increase in I_c causes a decrease in dissipation.

Circuits with $V_{ce} < \frac{1}{2}V_{cc}$

Most RC-coupled and some transformer-coupled stages come into this class. Such circuits are thermally stable because the dissipation decreases with increase in temperature. This property can be turned to advantage to obtain higher dissipations at normal ambient temperatures.

To evaluate the maximum junction temperature, the lowest value of collector current should be calculated for the maximum ambient temperature.

Circuits with $V_{ce} > \frac{1}{2}V_{cc}$

Many transformer-coupled stages and output stages come into this class. For such circuits two calculations are necessary.

MAXIMUM JUNCTION TEMPERATURE

The maximum emitter current ($\simeq I_c \text{ max}$) is determined at the maximum ambient temperature, as described under *D.C. Stabilisation*.

The power dissipated in the transistor may be taken as

$$p_c = I_c V_{ce},$$

and the maximum junction temperature is found by inserting the maximum values of ambient temperature and collector dissipation in the equation

$$T_j = T_{amb} + \theta p_c,$$

where θ is the thermal resistance from the junction to the surroundings. T_j must not exceed the maximum permissible value.

THERMAL STABILITY

The circuit is thermally unstable if

$$\frac{dp_c}{dT_j} \cdot \theta > 1.0$$

and the dissipation then increases to the limit, $V_{cc}^2/4(R_e + R_c)$, imposed by the circuit. If the increased dissipation results in a temperature greater than the maximum permissible junction temperature, the transistor may be permanently damaged and possibly destroyed.

dp_c/dT_j may be calculated from the following equations:

$$\frac{dp_c}{dT_j} = [V_{cc} - 2I_c(R_e + R_c)] \cdot \frac{dI_c}{dT_j}$$

and

$$\frac{dI_c}{dT_j} = \frac{(R_e + R_b) \left(\frac{dI_{co}}{dT_j} \right) - \frac{dV_{be}}{dT_j}}{R_e + R_b/\alpha'}$$

In the above equation,

$$\frac{dI_{co}}{dT_j} = 0.08I_{co}(T_j)$$

($I_{co}(T_j)$ being the value of I_{co} at the junction temperature T_j), and

$$\frac{dV_{be}}{dT_j} \sim -2\text{mV}/^\circ\text{C}.$$

GAIN

The a.c. load R_L on the transistor is formed by the following in parallel:

- (a) the output resistance r_{ce} of the transistor
- (b) the load resistance R_c in the collector circuit
- (c) the bias resistance R_b
- (d) the input resistance R_{in} of the following stage.

Hence

$$\frac{1}{R_L} = \frac{1}{r_{ce}} + \frac{1}{R_c} + \frac{1}{R_b} + \frac{1}{R_{in}}.$$

This equation can be evaluated more conveniently by substituting conductances for the reciprocals, thus

$$G_L = g_{ce} + G_c + G_b + G_{in}.$$

The output current from the transistor, which is equal to $\alpha' i_b$, will not all flow into G_{in} , some of it being shunted off by the other conductances. The current gain of the stage is therefore

$$A_i = \alpha' \cdot \frac{G_{in}}{g_{ce} + G_c + G_b + G_{in}}.$$

The gain will be increased as G_c and G_b are reduced, that is, as R_c and R_b are increased.

The effect of the coupling capacitance is discussed below.

COUPLING CAPACITANCE

The coupling capacitance is considered to be connected between the input resistance R_{in} of the following stage and the source resistance R_s of the preceding stage. R_s will be formed by the collector load resistor R_c in parallel with the high output resistance of the transistor, and is approximately equal to R_c .

A 3dB reduction in gain occurs at the frequency at which the reactance of the coupling capacitance is equal to $R_c + R_{in}R_b/(R_{in} + R_b)$, that is, when

$$\frac{1}{2\pi fC} = R_c + \frac{R_{in}R_b}{R_{in} + R_b}.$$

The input resistance is typically $1k\Omega$, the collector resistor about $5k\Omega$, and R_b may be only $2k\Omega$. For a 3dB reduction in response at $100c/s$, and using these values, the coupling capacitance would be

$$C = \frac{1}{2\pi \cdot 100 \cdot 5700} \text{ F} \\ \sim 0.3\mu\text{F}.$$

Normally the coupling capacitance will be an order of magnitude larger than this to ensure good bass response, and values of 4, 8 and $10\mu\text{F}$ are commonly used.

The polarity of the electrolytic coupling capacitor must be observed. The negative side of an interstage coupling capacitor is connected to the collector of the preceding transistor, and the positive side to the base of the following transistor (Fig. 4, p. 130), assuming that the transistors are p-n-p types.

In an input stage, where the capacitor forms part of a base-emitter network, the negative side of the capacitor is connected to the base.

EMITTER DECOUPLING CAPACITANCE

The resistance R to be decoupled at the emitter is not that of the emitter resistor alone, but is given by

$$\frac{1}{R} = \frac{\alpha'}{R_s} + \frac{1}{R_e}$$

where R_s is the total source resistance feeding the base.

As an example, α' can be taken as 50 and R_e as $1\text{k}\Omega$. The output resistance of the transistor is in the region of 20 to $50\text{k}\Omega$, hence the source resistance is substantially equal to the collector load resistance of, say, $5\text{k}\Omega$. R_e is large in comparison with R_s/α' , and the resistance

$$\frac{R_s}{\alpha'} = \frac{5\text{k}\Omega}{50} = 100\Omega.$$

Again a large capacity is normally used to ensure good bass response. The decoupling capacitance is usually $100\mu\text{F}$, the reactance of which is about 16Ω at 100c/s , 32Ω at 50c/s , etc. Decoupling capacitors of $250\mu\text{F}$ are not uncommon. For a p-n-p transistor, the negative side of the decoupling capacitor will be connected to the emitter and the positive side to the positive supply line (Fig. 4).

FREQUENCY RESPONSE

The h.f. response of the circuit is determined by the transistors, the cut-off frequency in grounded emitter, f'_α or f_β , giving a fairly good guide to the frequency at which a 3dB loss of gain occurs in practical grounded-emitter circuits.

The l.f. response is determined by the coupling capacitances and the emitter decoupling capacitances.

If the interstage coupling capacitors and emitter decoupling capacitor are added to the stabilised circuit of Fig. 1, the result is the basic RC-coupled amplifier stage of Fig. 4.

In practice an amplifier consisting of a number of RC-coupled stages is designed initially to have as wide a frequency response as possible, using coupling capacitances much higher than those indicated by simple theory. Then the response is limited as required by giving a lower value to one or two of the capacitors, care being taken to provide sufficient voltage drive in the preceding stage to allow for the increased reactance of the capacitance at low frequencies.

DISTORTION

The source resistance should be as high as possible to overcome the effect of the non-linear input resistance, otherwise there will be significant distortion. The source resistance is determined mainly by the collector load resistance of the preceding stage.

The distortion produced by the variation of α' with collector current

is of lesser importance, provided the collector-current excursion does not extend below 0.3mA. This distortion is a minimum for the OC71 when the current is in the region of 1 to 2mA.

Peak collector currents of up to 50mA are permissible for the OC70, OC71 and OC75, but it is inadvisable to design for peak currents of greater than 25mA where low distortion is required.

NOISE

In many a.f. stages, the signal level is such that noise need not be considered. Provided the volume control is suitably sited, only the input transistor need be operated under minimum-noise conditions. Good results are obtained with a collector current of 0.3 to 0.5mA and a source impedance of 500 to 2000Ω. To avoid introducing noise from the resistors, high-stability types may have to be used in the first stage.

All stages prior to the volume control should have a dynamic range of at least 10:1.

If the source impedance is high and no transformer is used, the transistor should be operated in grounded emitter at as low a current as possible (0.3 to 0.5mA). Silicon transistors, because of their low leakage currents, are probably more suitable than germanium transistors in such circumstances.

INPUT IMPEDANCE

A high input impedance may be obtained with a grounded-emitter stage by adding resistance in series with the input. Alternatively an unbypassed resistance R_e may be inserted in series with the emitter, the input resistance then being approximately

$$r_b + (r_e + R_e)(1 + \alpha')$$

The first method is preferable as the circuit can be stabilised against temperature changes without appreciable loss of gain.

Grounded-base stages have a very low input impedance, and give a high voltage gain when feeding into a high impedance.

NEGATIVE FEEDBACK

Single-stage Feedback

Two methods are possible for applying negative feedback to a single stage: (a) emitter feedback from an unbypassed emitter resistance;

(b) shunt feedback from collector to base.

Distortion and noise are decreased, and the cut-off frequency increased, by the feedback factor. Thus 6dB of negative feedback (feedback factor of two) halves the distortion and noise and doubles the cut-off frequency. The signal-noise ratio, however, is not improved.

Multistage Feedback

For a specified overall gain it is preferable, in principle, to include as many stages as possible in the feedback loop(s). Because of the wide spreads and low cut-off frequencies of most audio transistors, a feedback loop containing more than two stages tends to be unstable, and the suppression of the instability decreases the bandwidth. It has been found that a two-stage amplifier using current feedback is 'some sort of optimum'.

If the phase shift in the coupling circuits is neglected, a two-stage RC-coupled amplifier is inherently stable with any degree of negative feedback. If the transistors are directly coupled, a variable negative-feedback resistor may be used as the gain control, without any risk of instability. With transformer coupling, the transformers should have low leakage inductances if appreciable negative feedback is to be applied.

Amount of Feedback

The circuit has to be designed for the required minimum of feedback to be present with the lowest-gain combination of transistors. A check then has to be made that the circuit will remain stable with the highest-gain combination of transistors, and where appropriate with a loud-speaker load.

In a two-stage amplifier designed for a minimum of 6dB of feedback, the maximum feedback may be about 20dB.

In output stages, a design for a minimum of 6dB of feedback gives sufficient improvement in performance, without making the transformers too expensive.

Individual amplifiers can be designed with greater amounts of feedback. Any particular amplifier can be checked for stability by verifying that the application of feedback does not produce undue peaks in the very low and very high frequency regions.

Frequency Correction

A non-level response, such as is required for pick-up equalisation, for example, may be obtained by including frequency-sensitive components in the feedback networks. The feedback networks should be between

well-defined impedances.

SUPPLY DECOUPLING

The stability of a multistage amplifier is sometimes impaired by coupling between input and output circuits through the supply. Instability is avoided by decoupling the earlier stages by an RC network. The decoupling capacitance is usually $100\mu\text{F}$.

A dry battery shows a marked increase in internal impedance near the end of its life. This effect has to be borne in mind when considering the amount of decoupling required.

TYPICAL AMPLIFIER STAGE

Table 1 gives operating conditions and performance for an OC71 in the basic amplifier stage of Fig. 4. Gain and distortion are for a source

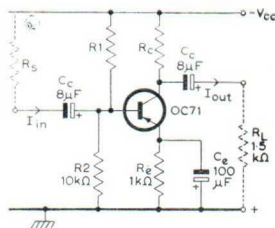


Fig. 4—Typical RC-coupled stage for OC71

TABLE 1

Operating Conditions and Performance for Basic OC71 Stage

Operating $T_{\text{amb max}} = 45^\circ\text{C}$

V_{cc} (V)	I_c (mA)	R_1 (k Ω)	R_e (k Ω)	$\frac{I_{out}}{I_{in}}$	I_{out} for $D_{tot} = 5\%$ ($\mu\text{A}_{r.m.s.}$)
6	1.0	39	2.2	23	200
6	1.5	22	1.5	18.5	290
9	1.0	62	3.9	28	260
9	1.5	39	2.7	24	430
12	1.0	82	5.6	31	270
12	1.5	56	4.7	30	535

impedance equal to the collector load resistance R_c of an identical preceding stage, and for a load equal to the input impedance of an identical following stage. The current flowing in the $1.5\text{k}\Omega$ load resistance is the output current considered in the distortion and gain measurements. The performance therefore applies to one OC71 in a series of

identical stages in cascade, but is not unrepresentative of what happens when a number of fairly similar stages are cascaded.

STAGES IN CASCADE

Simple amplifiers can be constructed by connecting a number of the stages given in Table 1 in cascade.

Apart from simplicity, there is no advantage in cascading identical stages. The output is limited to that of the final stage. In the earlier stages there is a sacrifice of gain, because the unnecessarily high operating current implies a lower collector load resistance.

In a correctly designed amplifier, therefore, each stage should be designed to supply the drive required by the following stage.

The overall distortion will be less than that obtained by adding together the distortions in the individual stages, because there will be partial cancellation of second-harmonic distortion.

PREAMPLIFIER FOR 250V SUPPLY

A high-gain preamplifier for operation from a 250V supply is shown in Fig. 5. The circuit can be rearranged as in Fig. 6 to accord with the positive polarity of the supplies in a succeeding valve amplifier. There is no need for a separate supply. In Fig. 6 one side of the output is earthed, whereas in the more familiar arrangement shown in Fig. 5 the output is floating. The performance of the circuit is given in Table 2. The preamplifier is suitable for low-impedance microphones and pick-ups.

The supply voltage is reduced to a safe level across the transistor by the flow of the quiescent current ($\sim 0.7\text{mA}$) through the $330\text{k}\Omega$ load resistor and $5.6\text{k}\Omega$ emitter resistor. High voltage gain arises from the high resistance in the collector-emitter circuit and the high input impedance (about $1\text{M}\Omega$) of the following valve stage, the shunting effect of which is small.

The circuit will work safely with a supply voltage of up to 275V. It has been operated successfully from a 100V supply taken from a valve amplifier.

There is no hum or microphony. Noise is not as low as in a valve circuit, but is low enough to make the preamplifier acceptable for most applications.

R2 is returned to the collector instead of to the supply and provides a.c. as well as d.c. feedback, the a.c. feedback path being through R2, R1 and the source. R1 ensures that a certain amount of a.c. feedback is present even with a very low source impedance.

AMPLIFIER STAGES

R5, the emitter resistor, does not contribute a.c. feedback, even though it is not bypassed. The input is applied between base and emitter and not through R5, which forms part of the d.c. load.

The a.c. feedback, apart from improving the frequency response and reducing distortion, reduces the input and output impedances. If the feedback is undesirable, R2 can consist of two equal or nearly

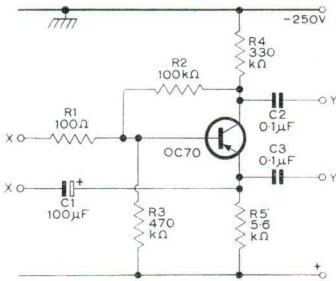


Fig. 5—Basic preamplifier for 250V supply

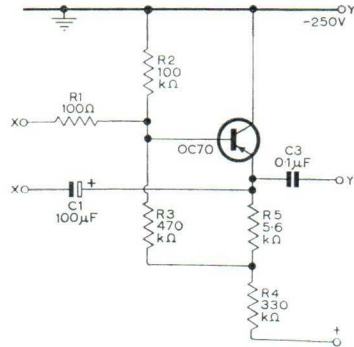


Fig. 6—Rearrangement of 250V preamplifier for feeding valve amplifier

TABLE 2
Performance of Preamplifier for 250V Supply

Operating T_{amb} max = 45°C

Output voltage	1.8V
Input voltage	5.5mV
Voltage gain	330
Output impedance	~5kΩ
Input impedance	200Ω
Freq. resp. (50Ω source) rel. to 1kc/s	
3dB reduction in gain at 15c/s and 13kc/s	
Total-harmonic distortion	0.4% at 0.5V
Current drain	~0.7mA

equal resistances with their common point bypassed to the emitter by a capacitance of about 0.5μF.

Although in both arrangements of the circuit the input terminals are floating, no hum should be introduced provided the preamplifier is mounted reasonably close to the microphone or pick-up. With the circuit of Fig. 5, the input terminals must never be shorted to earth. With that of Fig. 6, the current through the microphone or pick-up will not be excessive if the input is accidentally shorted to earth. High-stability 5% resistors are recommended. The OC71 could be used instead of the OC70, but it would be necessary to re-design the circuit.

CHAPTER THIRTEEN

CLASS A OUTPUT STAGES

Class A operation will be discussed only for single-ended output stages, since class B is normally preferred for push-pull circuits.

BASIC CONSIDERATIONS

Fig. 1 represents a transistor in any configuration operating in class A from a supply voltage V_{cc} . First of all it will be convenient to consider an ideal transistor, for which the knee voltage is zero; an ideal transformer will also be assumed. Similar relationships to those usually given for a valve will then hold.

The working point is the midpoint of the dynamic collector load line. The quiescent current I_q has superimposed upon it a signal

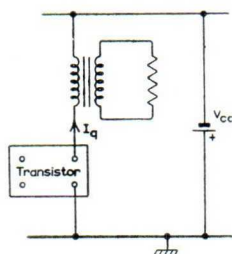


Fig. 1—Transistor in any configuration operating in class A

current which, with transformer coupling, swings the collector current between $2 \times I_q$ and zero, the signal-current amplitude being

$$i_{c(pk)} = I_q.$$

At the maximum collector current, the voltage across the transistor is zero, the voltage across the transformer being equal and in the opposite direction to the supply voltage V_{cc} . On the reverse half cycle, the voltage across the transformer is again equal to V_{cc} , but in such a direction as to reinforce the supply voltage. The collector voltage therefore swings between zero and $2 \times V_{cc}$, the signal-voltage amplitude being

$$V_{c(pk)} = V_{cc}. \quad \dots (1)$$

If the reflected load presented by the transformer is R_L , the

signal-voltage amplitude is

$$V_{e(pk)} = I_q \times R_L. \quad \dots (2)$$

From Eqs. 1 and 2, the required load resistance is

$$R_L = \frac{V_{cc}}{I_q}.$$

The load resistance may be represented in the usual way by a load line drawn across the output characteristic (Fig. 2). The load line will

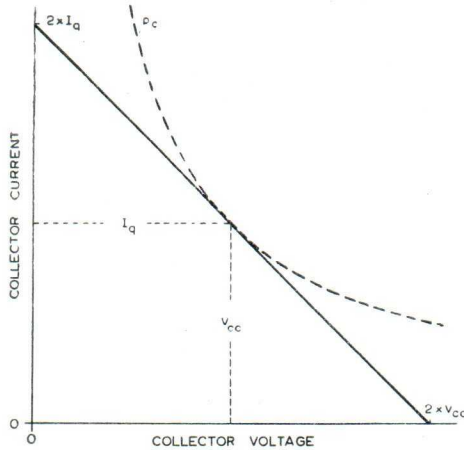


Fig. 2—Idealised load line

join the points $(2 \times V_{cc}, 0)$ and $(0, 2 \times I_q)$, the midpoint of the line being the working point (V_{cc}, I_q) .

The maximum output power is the r.m.s. value of the voltage excursion times the r.m.s. value of the current excursion, thus:

$$P_{outmax} = \frac{V_{cc}}{\sqrt{2}} \cdot \frac{I_q}{\sqrt{2}} = \frac{1}{2} V_{cc} I_q \quad (= \frac{1}{2} I_q^2 R_L).$$

The quiescent power which has to be supplied to the stage, irrespective of the output, is equal to the dissipation at the working point:

$$P_q = V_{cc} I_q.$$

The maximum theoretical efficiency is therefore $P_{outmax}/P_q = 50\%$.

The collector dissipation has its maximum value of $V_{cc} \times I_q$ under zero-signal conditions.

Losses

In practice there is always some series resistance $R_{d.e.}$ in the output

circuit. For example, in Fig. 3, $R_{d.c.}$ consists of the resistance of the output-transformer primary (R_p) and the unbypassed resistance in the emitter. Also, the minimum allowable collector voltage is not quite zero, but equal to or slightly greater than the knee voltage V_{knee} . Then

$$R_L = \frac{V_{cc} - V_{knee}}{I_q} - R_{d.c.}$$

and

$$P_{out \max} = \frac{(V_{cc} - V_{knee})I_q}{2} - I_q^2 R_{d.c.}$$

Drive Requirements

For a collector-current excursion of I_q , the required base-current excursion is

$$i_b = \frac{I_q}{\alpha'}$$

The preceding stage must be able to provide this peak drive current.

The drive voltage required is given by

$$v_b = I_q R_e + V_{be}$$

The drive requirements should be calculated for a transistor having the minimum α' and maximum V_{be} . The input impedance is v_b/i_b and the input power required is

$$\frac{v_b i_b}{2}$$

The operating conditions of the driver stage and the ratio of the driver transformer must be chosen with these values in mind. In

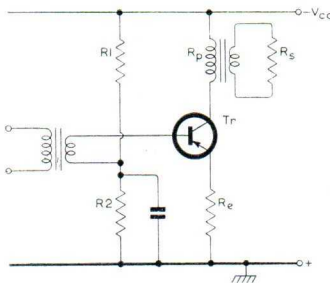


Fig. 3—Grounded-emitter output stage with d.c. resistance $R_p + R_e$

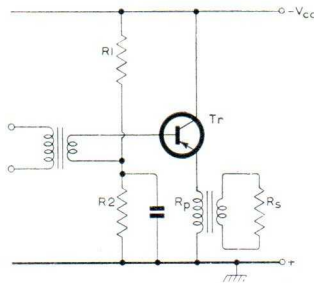


Fig. 4—Grounded-collector output stage

principle, the impedance at the transformer secondary should be fairly high for low distortion; in fact, however, the transformer usually has

a step-down ratio in order to reduce the current swing required from the driver stage.

Grounded Collector

In the grounded-collector output stage shown in Fig. 4, the emitter stabilising resistance can be the resistance of the primary of the output transformer.

The drive currents are the same as for grounded emitter, but the peak drive voltage is approximately equal to the supply voltage V_{cc} . Thus more input power is required. This circuit can be driven from a low source impedance.

Grounded Base

A transistor in grounded base requires a very high load resistance to give useful power gain. This circuit is, therefore, usually not suitable for output stages in which the load resistances are low.

HALF-SUPPLY-VOLTAGE PRINCIPLE

Simple output stages can be designed using the half-supply-voltage principle (pp. 62 and 124-5). A single-transistor amplifier based on

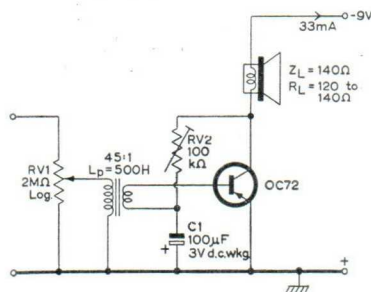


Fig. 5—60mW output stage using OC72 operating on half-supply-voltage principle

this principle is given in Fig. 5. The load is formed directly by a high-impedance speaker, the resistance of which drops half the supply voltage. As half the d.c. power is dissipated in the load, the maximum efficiency (a.c. power/d.c. power) is 25%; however, there is no output transformer to introduce further losses.

The collector current is preset to 33mA at 25°C by means of RV2. The output power is about 60mW at room temperature, but is considerably reduced at higher temperatures. A high-output crystal pick-up, giving about 3V at 1.2cm/sec, provides sufficient drive through the

45:1 transformer for full output from the OC72. Although the output is meagre, an appreciable volume of sound is generated in a speaker having a high flux density ($> 10,000$ gauss).

The cost of the complete amplifier is low, even with a high-efficiency speaker, as there is no output transformer and only one transistor is required.

CAR-RADIO OUTPUT STAGES

The output from the OC26 power transistor, which can be about three watts in a conventional class A circuit, is sufficient for most car-radio receivers. Such receivers may be all-transistorised or 'hybrid' (containing both transistors and valves). The valves in the hybrid receivers are specially designed to operate from a 14V supply.

Low-voltage Operation

A receiver using conventional valves and operating from a low-voltage d.c. source, such as a car battery, must necessarily include some device (usually a vibrator) to provide an h.t. line of, say, 200V. Although valves have been designed to give perfectly satisfactory performance in the early stages of a receiver, using low anode and screen-grid voltages, it is not practicable to obtain sufficient power from a valve output stage at low voltages. The introduction of power transistors has therefore made possible the design of complete low-voltage receivers, since the transistor requires only a low operating voltage.

'Hybrid' Receiver

The hybrid car-radio receiver has the following advantages:

- (a) the vibrator and its associated components are not required
- (b) filtering is simplified
- (c) the cost is (at present) less than that of an all-transistor design
- (d) the current drain is considerably reduced.

The performance is at least equal to that of a conventional all-valve receiver with a vibrator supply.

Two car-radio output stages will now be described, one for an OC26 output transistor driven by an EF98 valve, the other for an OC26 driven by an OC71.

Supply Voltage

Both circuits (Figs. 6 and 7) have been designed for a nominal line voltage of 14V. Occasionally the terminal voltage of the car battery may rise to 15V or more. Another complication is that a filter is usually included in the supply line to suppress interference, and this filter may

cause a voltage drop of about 0.5 or 1.0V. The actual line voltage therefore may only be determined with reference to a practical design. It should be noted that a lower line voltage will reduce the maximum available output power. Also, the circuit should be changed if necessary to ensure (i) a screen-grid voltage of at least 12.6V and an anode voltage of at least 12V for the EF98 in the valve-driver circuit, or (ii) a line voltage of 12.6V for the ECH83 and the OC71 in the transistor-driver circuit.

OUTPUT STAGE WITH EF98 DRIVER

The circuit for an OC26 output stage driven by an EF98 valve is given in Fig. 6. The gain of the OC26 is sufficient for it to be driven by the EF98 via a 25:1 driver transformer.

Output Stage

The quiescent current of the OC26 is preset to 600mA by adjusting RV2. With a collector load of 21Ω , the maximum peak collector current is 550mA. This gives a maximum output of 3.6W from the transistor and 3.3W into the primary of the output transformer, at about 10% total-harmonic distortion. With an output transformer as specified, the output power into the loudspeaker is about 2.8W.

OUTPUT TRANSFORMER

Turns ratio	2.5:1 (3Ω speaker)
Primary resistance	$<0.75\Omega$
Secondary resistance	$<0.25\Omega$
Primary inductance	$>30\text{mH}$ at 600mA d.c.

THERMAL STABILITY

With a 1Ω emitter resistor, the total base resistance is

$$6.8 + 0.75 = 7.55\Omega,$$

where the resistance of the driver-transformer secondary is 0.75Ω . The maximum junction temperature is 90°C at an ambient temperature of 55°C , provided the thermal resistance between mounting base and ambient does not exceed 2.6°C/W . The transistor under these conditions is always thermally stable.

If this thermal resistance is increased to 3.3°C/W , the maximum junction temperature becomes 100°C . (The transistor is rated for a maximum of 200 hours intermittent operation at this temperature.) The thermal stability, however, is much impaired, as the stability factor

$$\theta \cdot \frac{dI_c}{dT_j} \cdot V_{ce}$$

is now 0.8. This is dangerously near 1.0, the limit for stability. It is

recommended, therefore, that the thermal resistance should not exceed $2.8^{\circ}\text{C}/\text{W}$.

LOCAL NEGATIVE FEEDBACK

A minimum of 3dB of local negative feedback is applied to the OC26 through the collector-base resistor R4. Apart from reducing distortion, the feedback also safeguards the transistor against excessive collector voltage under overdrive conditions. The feedback of 3dB is sufficient for this purpose, because of the small spread in the drive voltage of the OC26.

Driver Stage

The maximum base current required by an output transistor having the minimum $\bar{\alpha}'$ of 20 is $600/20 = 30\text{mA}$ peak. Local negative feedback

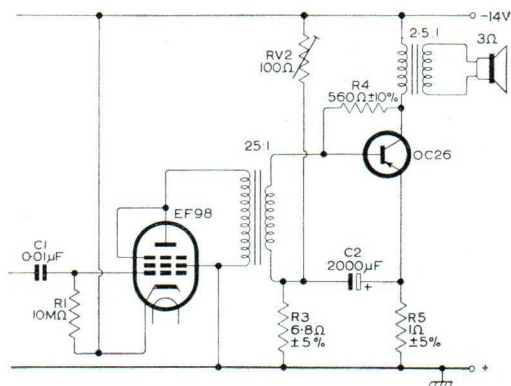


Fig. 6—OC26 output stage with EF98 valve driver

accounts for a further 14mA peak, so that the total drive current required is 44mA peak. A drive voltage of 330mV should also be available at the secondary of the driver transformer, allowance having been made for the flow of base current through the secondary.

With a 25:1 driver transformer, the EF98 has to provide an output-current amplitude of $44/25 = 1.76\text{mA}$ peak and a voltage of $330 \times 25 = 8.25\text{V}$ peak. This drive can be obtained from the driver valve at less than 5% total-harmonic distortion, provided the voltage V_{ak} across the valve is equal to or greater than 12V, and provided V_{g2k} is equal to or greater than 12.6V.

The operating conditions of the EF98 are:

$$V_{a_k} = 12V, \quad V_{g2k} = 14V, \quad R_{g1k} = 10M\Omega, \quad R_a = 4.4k\Omega.$$

DRIVER TRANSFORMER

Turns ratio	25:1
Primary resistance	$<300\Omega$
Secondary resistance	$<0.75\Omega$
Primary inductance	$>6\text{H}$ at 6mA d.c.

The driver transformer should be phased such that an increase in the collector current of the OC26 corresponds to a decrease in the anode current of the EF98. Although the second-harmonic distortion in the output will not be reduced, because the distortion components of the transistor and valve do not cancel, the transformer must be phased in this way to keep overdrive within safe limits.

Reduction in Size and Weight

The circuit shown in Fig. 6 does not represent the most compact design which is possible with a valve driver. A thermistor may be incorporated in the bias network, and the quiescent current of the output stage reduced. It would then be possible to use a smaller heat sink for the output transistor, and to reduce the value, and therefore bulk, of the bypass capacitor C2. However, where saving of weight and space are important, a better solution is to use a transistor driver.

OUTPUT STAGE WITH OC71 DRIVER

The second circuit for a car-radio output stage (Fig. 7) is novel in that it incorporates a transistor driver. This circuit has been produced primarily with the valve-cum-transistor ('hybrid') receiver in mind. The EF98 valve driver is replaced to considerable advantage by an OC71 transistor, the rest of the receiver being equipped with valves. However, the circuit will be equally suitable for all-transistor receivers, provided the preceding stages develop the input required by the OC71.

The usual saving in space and weight is effected when a transistor replaces a valve. To follow up this advantage, the heat sink and bypass capacitor C7 have also been reduced in size. However, the main advantage in adopting an OC71 transistor as the driver is an increase in gain of approximately 6dB. This increase in gain enables overall negative feedback to be applied. The sensitivity has been kept the same as in the valve-driver circuit, but the total-harmonic distortion is only about 3% instead of 10%.

Also, with a transistor driver, overdrive conditions are easier to avoid, because the knee voltage of the driver transistor, and thus the maximum available drive voltage, is much more accurately defined than for a valve.

Smaller Heat Sink and Bypass Capacitor

The area of the heat sink can be reduced by approximately 30% by accepting a slight reduction in output power, and by improving the thermal stabilisation at the higher ambient temperatures by means of a thermistor.

The quiescent current in the output stage has been reduced from 600mA to 500mA, thus reducing the output power from the transistor from 3.6W to 3.2W. In a practical class A stage, the efficiency is always less than 50%, so that the transistor dissipations at normal room temperature are 7.8W and 6.8W respectively.

As the ambient temperature increases, the resistance of the thermistor falls and prevents the collector current from rising. Whereas at the maximum ambient temperature, the collector dissipation rises to about 9W in the valve-driver circuit, in the transistor-driver circuit, it remains virtually constant at 6.8W.

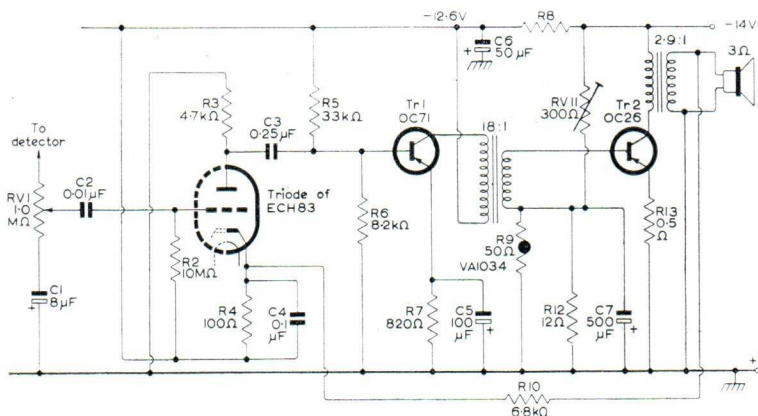


Fig. 7—OC26 output stage with OC71 driver

The resistances in the potential divider across the base of the output transistor can be increased by a factor of about two, with a similar decrease in bypass capacitance. In fact it is possible to decrease the bypass capacitance by a factor of four, from 2000 μ F to 500 μ F. With the extra drive available from the transistor, there is no appreciable reduction in output or frequency response.

A similar reduction in the size of the heat sink, and some reduction in the value of the capacitor, could be obtained with the valve-driver circuit. However, as the sensitivity is not so great, and less feedback is available to improve the frequency response, the capacitance could not be reduced by a factor of four.

Output Stage

The OC26 has a 0.5Ω emitter resistor as part of the d.c. stabilisation network. This resistor is not bypassed to a.c., so that some local negative feedback is applied to the stage.

The quiescent current is preset to 500mA by means of RV11, and the maximum peak collector current is approximately 450mA. With a collector load of 29Ω , the maximum output from the transistor is 3.15W, giving 3W into the primary of the output transformer, at about 3% total-harmonic distortion (with overall negative feedback). Using an output transformer as specified, the output power into the loudspeaker is about 2.6W.

These figures are derived below by way of a worked example.

OUTPUT TRANSFORMER

Turns ratio	2.9:1 (3 Ω speaker)
Primary resistance	<1.0 Ω
Secondary resistance	<0.3 Ω
Primary inductance	>45mH at 500mA d.c.

COLLECTOR LOADING

The variation of current gain with collector current causes the load line of a transistor class A output stage to shift with changes in the signal conditions. The output also changes and the optimum load may be slightly different from that obtained by the simple theoretical determination given below. The optimum loading is usually found by experiment.

A maximum output power of 3W is required into the primary of the output transformer. (On account of the uncoupled 0.5Ω emitter resistor, a slightly higher output of 3.15W is required from the transistor.) The maximum signal-voltage amplitude available is the collector supply voltage of 14V minus the knee voltage of 0.4V, thus $v_{c(pk)} = 14 - 0.4 = 13.6V$. The total load R_L required for the transistor is given by

$$R_L = \frac{(13.6/\sqrt{2})^2}{3.15} \simeq 29.5\Omega,$$

and the maximum collector-current excursion is therefore

$$i_{c(pk)} = \frac{13.6}{29.5} \simeq 450mA.$$

As the emitter resistor is part of the total load on the transistor, the load required across the primary of the output transformer is $29.5 - 0.5 = 29\Omega$.

The output power into the primary of the transformer is

$$\left(\frac{0.45}{\sqrt{2}}\right)^2 \cdot 29 = 3\text{W}.$$

OUTPUT-TRANSFORMER TURNS RATIO

The collector loading of 29Ω consists of the resistance of the primary winding of the output transformer plus the reflected load impedance. For a really efficient transformer, the primary resistance should not exceed 1Ω . The reflected load is then $29 - 1 = 28\Omega$. For a 3Ω loud-speaker, the resistance of the secondary winding should not exceed 0.3Ω for reasonable efficiency. The total secondary impedance is thus 3.3Ω , and this should be reflected as 28Ω into the primary. The transformer turns ratio is

$$\sqrt{\frac{28}{3.3}} = 2.9:1.$$

POWER INTO LOUDSPEAKER

The power into the primary of the output transformer is 3W . The power into the load with the winding resistances as specified above is

$$3.0 \times \frac{28}{28+1} \times \frac{3}{3+0.3} = 2.6\text{W}.$$

THERMAL STABILITY

The output transistor is stabilised by means of a thermistor placed across the lower resistor, R_{12} , of the potential divider which provides the base-bias voltage. When the ambient temperature rises, the fall in the resistance of the thermistor effectively prevents the collector current from increasing. At an ambient temperature of 55°C , the collector current is 510mA , which is only 10mA greater than at 25°C , and the maximum dissipation is then 6.8W at the nominal battery voltage of 14V . The transistor is stable against thermal runaway under the above conditions, provided the thermal resistance between the mounting base of the transistor and the surroundings is 3.3°C/W . The transistor is also stable for a battery voltage of 15V at an ambient temperature of 50°C .

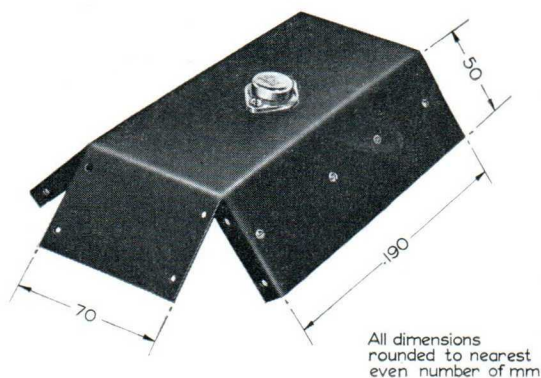
The maximum junction temperature with the specified thermal resistance is approximately 85°C , and this is well within the maximum of 90°C for continuous operation. Furthermore, the maximum is 100°C for 200 hours intermittent operation. An amplifier could be designed to take advantage of these maximum junction-temperature limits. However, the sensitivity would be reduced, as better d.c. stabilisation would be required to prevent thermal runaway. Also, a loss in output power would occur, unless a high-value and bulky capacitor were

added to decouple the emitter resistor of the OC26. The circuit of Fig. 7 is considered to give the best compromise between considerations of size and sensitivity.

Negative Feedback

LOCAL FEEDBACK

The local negative feedback from the unbypassed emitter resistor tends to linearise the input impedance of the OC26. This feedback reduces the possibility of overdriving output transistors which are characterised



Exploded view of heat sink/chassis for OC26 in laboratory model of transistor-driver circuit. Material 2mm blackened-aluminium sheet metal.

by a low base-emitter voltage. The feedback increases the drive voltage, and not the drive current, required by the output stage.

OVERALL FEEDBACK

The maximum input voltage required at the grid of the triode of the ECH83 is basically 80mV r.m.s. for full output. The maximum input voltage required if an EF98 is used instead of an OC71 is 150mV r.m.s. Nearly 6dB more gain is thus obtained from the transistor driver, and this increase in gain is used to add overall negative feedback.

The feedback loop is taken from the secondary of the output transformer to the cathode of the ECH83 triode. The feedback has to be taken to the input of the valve, in order that a higher output current will not be required. The grid of the triode is not a suitable point at which to apply the feedback, as the input impedance of the stage, and thus the load on the preceding diode, would be reduced. This effect is particularly important when the gain control RV1 is set to give

minimum resistance. The 100Ω feedback resistor in the cathode circuit does not affect the working of the heptode section of the valve, since it is decoupled to r.f. by the $0.1\mu\text{F}$ capacitor C4.

The feedback loop, although it includes two transformers and three stages of amplification, can be made stable for considerable amounts of feedback.

Driver Stage

The maximum base current required by the OC26 for the maximum peak collector current is $500/20 = 25\text{mA}$ peak. A total drive voltage of 550mV should be available at the secondary of the driver transformer; this includes the voltage drop across the secondary winding and also the voltage across the unbypassed emitter resistor R13. If the voltage available across the primary of the driver transformer is 10V , a driver-transformer ratio of $10/0.55 = 18:1$ is suitable. The current swing in the primary is then $25/18 = 1.4\text{mA}$ peak, and the minimum direct collector current of the driver transistor must be 1.9mA , to avoid swinging below 0.5mA , where the characteristic is more non-linear and there is excessive distortion.

The triode section of the ECH83 gives $50\mu\text{A}$ peak into the $1\text{k}\Omega$ load presented by the input impedance of the OC71. Total-harmonic distortion is less than 5%, although more current is obtainable with an increase in distortion. With a base current of $50\mu\text{A}$, the driver transistor must have a current gain of $1.4 \times 10^3 / 50 = 28$ in order to drive an OC26 having the minimum α' to the full output of 3.2W . For the great majority of OC71 transistors, less than $50\mu\text{A}$ drive will be required from the valve for full output. Occasionally, if both the OC71 and OC26 have the minimum α' , up to $67\mu\text{A}$ peak will be taken from the triode. However, as the valve is within the feedback loop, the increase in distortion at the output will be small.

The operating conditions for the ECH83 triode are:

$$V_b = 12.6\text{V}; \quad R_a = 4.7\text{k}\Omega; \quad R_{g1} = 10\text{M}\Omega;$$

$$D_{\text{tot}}(\text{for } I_{\text{out}} = 50\mu\text{A}) = 5\%.$$

SUPPLY VOLTAGE TO DRIVER STAGE

The resistance of R8 in the negative supply line should be such that the supply to the driver stage at the nominal battery voltage (14V) is not greater than -12.6V . The collector-emitter voltage is then $12.6 - 1.9 \times 0.82 = 11\text{V}$.

The d.c. voltage rating V_{cmax} of the OC71, with the $6.5\text{k}\Omega$ source impedance existing in the circuit, is 12.7V . The collector-emitter voltage of the transistor will not exceed this rating, even if the battery

voltage rises to 16V, provided the collector-emitter voltage is 11V with a 14V battery.

The a.c. source impedance of the OC71 is $3\text{k}\Omega$ and the $V_{c(\text{pk})}$ max rating of the OC71 under this condition is greater than 25V. Provided the supply voltage to the driver stage is not greater than -12.6V , the a.c. rating will not be exceeded.

DRIVER TRANSFORMER

Turns ratio	18:1
Primary resistance	$<150\Omega$
Secondary resistance	$<1.0\Omega$
Primary inductance	$>10\text{H}$ at 2mA d.c.

As in the valve-driver circuit, in order to avoid overdriving of the OC26, the phasing of the driver transformer should be such that an increase in the collector current of the OC26 corresponds to a decrease in the collector current of the OC71.

Frequency Response

The gain of the complete audio amplifier (triode valve plus OC71 and OC26) falls to 3dB below the mid-frequency value at 70c/s and 8kc/s. This response is for nominal transistors, with the overall feedback applied, and with a $200\text{k}\Omega$ source impedance.

Sensitivity

The voltage inputs required at the grid of the triode of the ECH83, with the overall negative feedback applied, are given in the following table:

<i>Output from OC26</i>	<i>Minimum-α' transistors</i>	<i>Nominal transistors</i>
Full output (3.0W)	150mV	116mV
50mW	22mV	17mV

Although a drive current of $67\mu\text{A}$ peak is required at the base of the OC71 when both the OC71 and OC26 have the minimum α' , for nominal transistors, the typical figure is $30\mu\text{A}$ peak (without overall feedback).

CHAPTER FOURTEEN

CLASS B PUSH-PULL OUTPUT STAGES

In class B push-pull operation, one transistor conducts while the other is cut off. Thus the two transistors amplify alternate halves of the waveform, and their inputs are 180° out of phase.

In principle the two transistors should be biased to cut off. Strict adherence to this condition, however, results in 'crossover' distortion

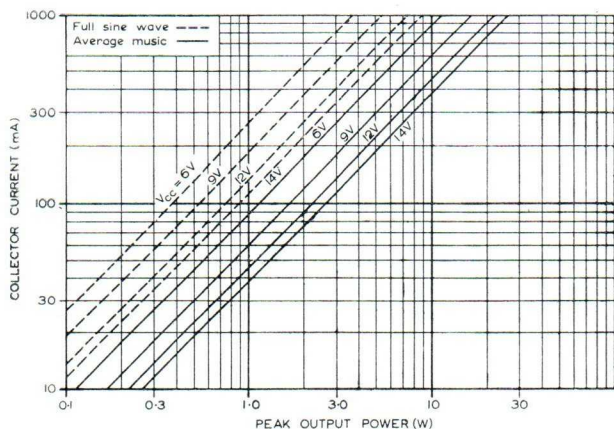


Fig. 1—Direct current plotted against nominal peak output power, for average music and for full sine-wave drive.

which is unpleasant to the listener. This distortion is overcome by applying a small forward bias to each transistor.

The main advantages of class B push-pull as compared with class A operation are

- (a) low quiescent current, and
- (b) high efficiency at full output.

Because the current drain is low, class B operation is favoured for the output stage for equipment operating from dry batteries.

The battery drain depends on the signal being handled. The average current consumption on music is about a third of that on maximum sine-wave output.

In Fig. 1, the full lines show the direct current with average music

plotted against the nominal peak output power, for various battery voltages. The broken lines are for full sine-wave drive.

The output from the class B push-pull pair is about five times the collector dissipation of a single transistor. The maximum theoretical efficiency (a.c. power/d.c. power) is 78.5%, and usually in practice an efficiency of 70 to 75% can be realised.

The transistors could be connected in common base, common emitter or common collector. Common-base operation is not suitable

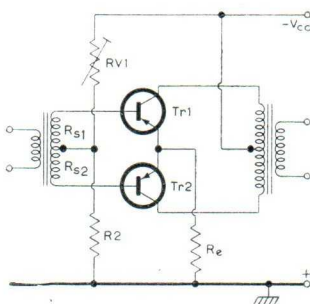


Fig. 2—Basic common-emitter circuit

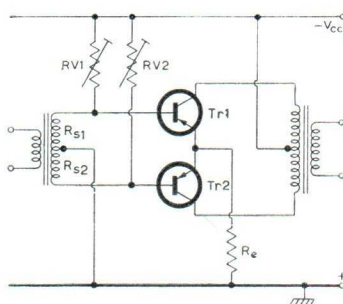


Fig. 3—Common-emitter circuit with integral secondary resistances

for practical circuits because of its low power gain. Common-emitter and common-collector circuits, and also a split-load arrangement which lies between the two, are described in this chapter.

COMMON-EMITTER STAGES

The basic common-emitter circuit is shown in Fig. 2. The potential divider RV_1 , R_2 provides the quiescent bias necessary to eliminate crossover distortion. The sum of the quiescent currents of the two transistors is preset to the design value by means of RV_1 . An emitter resistor R_e is required for thermal stability. R_2 and the resistances of the transformer secondaries also influence thermal stability. In general the secondaries should have as low a resistance as possible.

Occasionally it will be cheaper to use transformers having high secondary resistances. These resistances can form the lower part of the bias potential dividers, as shown in Fig. 3. The quiescent current of each transistor is preset to the design value by means of RV_1 and RV_2 .

In Fig. 4, the centre-tap of the output transformer is connected to the supply. The transistors still operate in the common-emitter mode,

even though the collectors are taken directly to the negative supply. This circuit has the advantage, for power transistors, that both transistors can be mounted directly (without mica washers) on a common heat sink, which is connected to the negative supply terminal.

Whenever it is possible to adopt a centre-tapped supply, the single-ended push-pull circuit of Fig. 5 offers the least costly solution. A high-impedance speaker provides the load, instead of a conventional low-impedance speaker and output transformer. The performance is

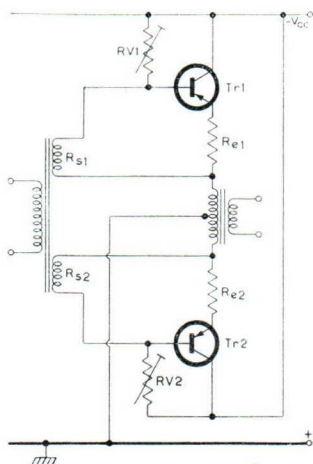


Fig. 4—Common-emitter circuit with collectors connected to negative side of supply

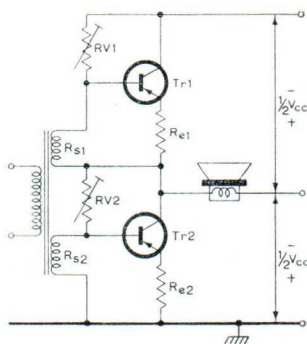


Fig. 5—Transformerless output stage for centre-tapped supply

the same as for the other circuits, if each half of the centre-tapped battery has the same voltage as that of the supply in the other circuits.

$RV1$ and $RV2$ in all the above circuits can be fixed instead of variable for transistors with close V_{be} spreads. The bias conditions given in the published data should be adopted.

Thermal Stability

None of the operating conditions should be changed without investigating the thermal stability. The quiescent currents must be set to the design values, the stability being impaired at higher settings.

The maximum dissipation and the thermal stability depend on the total thermal resistance of the transistor from junction to ambient. With power transistors, the thermal resistance largely depends on the

heat sink and the manner in which the transistor is mounted. The thermal resistance (junction to ambient) must always be that required by the design or lower.

Normally, users of this book will be concerned with operating or servicing existing equipment which may include a heat sink, or with building up designs which include a specification for a suitable heat sink. Provided the function of the heat sink is understood, and nothing is done to impair its efficiency, there should be no trouble from thermal instability. Further information on heat sinks will be found in Chapter 9.

Transformer Turns Ratios

The turns ratios of the output and driver transformers can be calculated simply from the information supplied. The usual equation applies to the output transformer, namely

$$R_c = R_p + m^2 R_s$$

where R_c is the load impedance per transistor, R_p is the resistance of the primary, R_s is the total resistance of the secondary and speaker, and the turns ratio is $m+1$. In an efficient transformer, R_p will be not more than 5% of R_c , and the resistance of the secondary winding will be not more than 5% of the speaker resistance.

The published data show the peak drive voltage required at the secondary of the driver transformers. Let this be v_2 . Let the peak voltage available at the transformer primary be v_1 . Then the turns ratio $n:1$ is given by $n = v_2/v_1$.

SPLIT-LOAD STAGES

In the split-load circuit (Fig. 6) part of the load is in the emitter. This arrangement reduces crossover distortion at lower battery voltages and lower ambient temperatures.

In conventional class B common-emitter output stages, a quiescent bias voltage is applied, which results in a linear transfer characteristic at nominal battery voltage. As the battery voltage decreases, the quiescent bias decreases, and the transfer characteristic becomes non-linear, resulting in crossover distortion.

In the split-load circuit, the presence of part of the load in the emitter means that a higher drive voltage v_b is required at the base for the same peak collector current $i_{c(pk)}$. The non-linear region of the characteristic forms a smaller proportion of the peak drive than before, and the distortion is reduced.

Best results would be obtained with the entire load in the emitter (that is, as in common-collector amplifiers) but the loss in gain may be too high. The fraction of the load in the emitter circuit is thus a

compromise between gain and performance. As a rough guide, the tolerable variation in battery voltage increases by the same factor as the increase in base input voltage.

The base input voltage is greater than that of an equivalent common-emitter amplifier by the voltage developed across the emitter load, and the driver and earlier stages must be designed accordingly.

Crossover distortion at low battery voltages (or low ambient temperatures) is lower than for a common-emitter amplifier. In all other respects the performance of the two types of circuit is similar.

COMMON-COLLECTOR STAGES

The transistors can also be used in common-collector class B push-pull. The drive current is the same as for a similar common-emitter circuit, but the drive voltage is approximately equal to the sum of the output

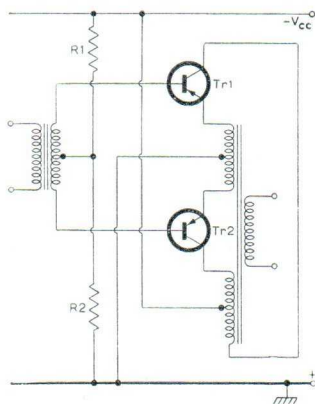


Fig. 6—Split-load circuit

voltage and the drive voltage for common-emitter operation. Therefore much greater drive power is required. The driver transformer must have a step-up ratio, and the inductances of the driver and output transformers must be higher than for a common-emitter output stage.

Crossover distortion and non-linearity distortion are less than in common-emitter or split-load circuits, but the gain is considerably reduced. Common-collector stages are therefore adopted only in special circumstances.

DESIGN PROCEDURE

The design procedure of common-emitter class B stages will now be

described. A complete specimen design will not be given, but some of the points will be illustrated by figures based on a 15W output stage. The procedure can be modified to cover the design of various types of output stage, as indicated in the preceding sections.

In designing an output stage, it is necessary to determine: (a) the maximum dissipation of the transistor; (b) the peak currents and voltages; (c) the optimum load; (d) the output power; and (e) the input current and input voltage. The distortion at various outputs can also be evaluated.

Allowance has to be made for the full spread of the transistor characteristics. All of (a) to (d) above are very nearly independent of transistor characteristics, provided that when calculating (e), the input current and voltage requirements, the extreme-limit characteristics are used.

Calculation for Zero Knee Voltage

The design equations are best illustrated by giving them for transistors having zero knee voltage, and then making adjustments for the departure from this ideal. In practice the power output is about 85 to 95% of the ideal. The following relations can be proved for a pair of transistors having zero knee voltage:

Maximum peak voltage = supply voltage V_{ce}

Maximum peak current = $\frac{V_{ce}}{R_{ce}}$,

where R_{ce} is the collector-emitter load per transistor

Maximum output power of the pair ($P_{out\ max}$) = $\frac{(V_{ce})^2}{2R_{ce}}$

Efficiency at maximum power = 78.5%

Maximum dissipation per transistor = $\frac{2}{\pi^2} \cdot \frac{(V_{ce})^2}{2R_{ce}}$
 $\simeq \frac{1}{5} P_{out\ max}.$

A graph showing the dissipation and efficiency related to peak signal is given in Fig. 7 for sine-wave drive. Maximum dissipation occurs when the actual maximum collector-current amplitude is 0.638 of the ideal, and the efficiency at this current is 50%. Consider the load line shown in Fig. 8. Then:

Maximum output voltage (ideal) = 14V peak

Maximum output current (ideal) = 3.3A peak

Load per transistor $R_{ce} = 14/3.3 = 4.25\Omega$

Maximum output power from the pair (ideal) = $\frac{1}{2}(14)(3.3) = 23.1W$

Maximum dissipation per transistor (ideal) = $\frac{1}{5}(23.1) = 4.6W.$

Effect of Knee Voltage

In the above expressions, the minimum collector-emitter voltage is assumed to be zero. In fact, the minimum collector-emitter voltage at a particular collector current is the knee voltage given in the published data. Because of the non-linearity in the vicinity of the knee voltage, a slightly higher voltage is taken as the minimum for the purpose of design. This minimum voltage is proportional to the collector current and can be represented by a straight line on the output characteristic

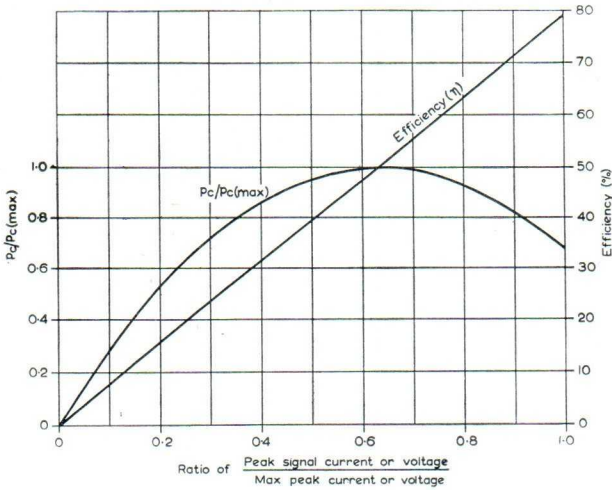


Fig. 7—Efficiency and relative dissipation plotted against relative signal

(Fig. 8). $\tan \phi$ is the minimum voltage for a particular current I_c divided by the current I_c , and has the dimensions of a resistance. This method of dealing with the minimum collector-emitter voltage is convenient, as $\tan \phi$ is the same for all collector currents.

The effect of the minimum voltage is to decrease the maximum peak output voltage. Therefore the maximum peak current, the maximum output power, and the efficiency at maximum output are all slightly decreased. For example, in Fig. 8, $\tan \phi = 0.4\Omega$. With $V_{ce} = -14V$ and $R_{ce} = 4.25\Omega$ as before;

$$\text{Maximum output voltage} = V_{ce} \cdot \frac{R_{ce}}{R_{ce} + \tan \phi} = 12.8V \text{ peak}$$

$$\text{Maximum output current} = \frac{V_{ce}}{R_{ce} + \tan \phi} = 3.0A \text{ peak}$$

$$\text{Maximum output power} = \frac{1}{2}(12.8)(3.0) = 19.2W.$$

In Figs. 7 and 8,

$$\begin{aligned} \frac{i_{c\max. \text{ actual}}}{i_{c\max. \text{ ideal}}} &= \frac{v_{c\max. \text{ actual}}}{v_{c\max. \text{ ideal}}} = \frac{R_{ce}}{R_{ce} + \tan \phi} \\ &= \frac{4.25}{4.25 + 0.4} = 0.915, \end{aligned}$$

so that

Efficiency at maximum output power = 72.5%.

Since the maximum dissipation occurs at 0.638 of $I_{c\max}$, the currents and voltages at which maximum dissipation occurs are unaffected by

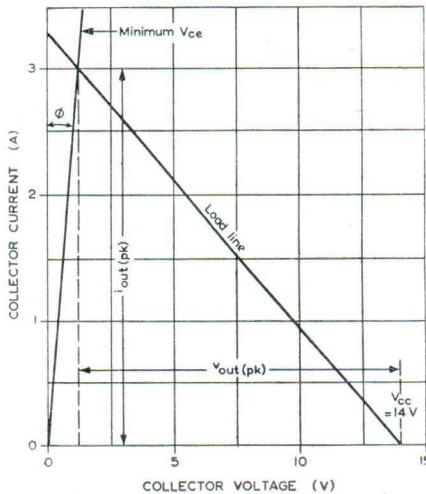


Fig. 8—Output characteristic and load line

the knee voltage. Therefore the maximum dissipation is the same as for zero knee voltage.

Emitter Resistor

The emitter resistor is in the collector-emitter loop and forms part of the load on the collector. As the emitter resistor cannot be decoupled, this resistor introduces a loss of output power.

If R_c is the reflected useful load in the collector, and R_{ce} is the total load in the collector-emitter loop,

$$R_{ce} = R_e + R_c.$$

Then

$$\text{Useful output power} = P_{\text{out}} \max. \frac{R_c}{R_{ce}}.$$

For the same example as before:

$$\text{Reflected useful load} = 4.25 - 0.5 = 3.75\Omega$$

$$\begin{aligned} \text{Useful output power} &= 19.2 \cdot \frac{3.75}{4.25} \\ &= 16.9\text{W}. \end{aligned}$$

Quiescent Current and Power Dissipation

In general the maximum value of the quiescent current is small in comparison with the maximum collector current. It can be shown that the effect of the quiescent current on the maximum collector dissipation is to increase the dissipation by $\frac{1}{5}(I_q V_{ce})$.

In an output stage giving 15W output, the maximum quiescent current might be 130mA at an ambient temperature of 55°C. The increase in dissipation would then be

$$0.2 \times 0.13 \times 14 = 0.36\text{W}.$$

Hence the maximum collector dissipation is

$$4.6 + 0.36 = 5\text{W}.$$

The quiescent current does not affect the maximum output power.

Drive Current

The maximum drive current can be found from $i_c/\bar{\alpha}'$, where i_c is the maximum collector current required (e.g. 3.0A) and the value of $\bar{\alpha}'$ is the minimum which may occur for the type of transistor in question.

Drive Voltage

The drive voltage required to produce a collector current i_c consists of:

- (a) the change in the voltage across the emitter resistor, equal to $i_c R_e - I_q R_e$
- (b) the change required in the base-emitter voltage, allowance being made for transistor spreads. The required change in base-emitter voltage is equal to the base-emitter voltage at the collector current i_c ($= V_{be(i_c)}$) minus the base-emitter voltage at the collector current I_q ($= V_{be(I_q)}$)
- (c) the additional voltage required for the base current i_b to flow in the resistance of the base circuit, that is, the circuit consisting of the transformer secondary and the equivalent bias source resistance.

The voltage required in the base circuit is $i_b R_b$, where R_b is the total base-circuit resistance.

Therefore the drive voltage required is

$$(i_e - I_Q)R_e + V_{be(i_e)} - V_{be(I_Q)} + i_b R_b.$$

(The above considerations of drive current and voltage include the assumption that maximum V_{be} occurs with transistors on the low-gain limit. The actual combination of characteristics will not always be as unfavourable as this, and the maximum drive required is then somewhat less.)

The drive power required (r.m.s.) is given by

$$\frac{1}{2}(i_{\text{drive peak}} \times v_{\text{drive peak}}).$$

This value of drive power includes the loss in the driver-transformer secondary.

Distortion

The main causes of distortion are non-linearity of the I_c/I_b (transfer) characteristic, crossover distortion, and mismatch of the two transistors.

NON-LINEARITY

The $\bar{\alpha}'$ of a transistor decreases with increasing current over most of the current range. The $\bar{\alpha}'/I_c$ characteristic is, however, controlled so that the maximum variation in gain is restricted to within acceptable limits. The effect of this type of non-linearity is to produce predominantly third-harmonic distortion.

The distortion can be calculated for any given or derived I_c/I_b characteristic. The following method is simple and sufficiently accurate. Let i_{c1} be the peak collector current at which the distortion is to be calculated. Then:

- from the characteristics obtain the base current (represented by i_{b1}) which corresponds to the collector current i_{c1}
- determine the collector current (represented by i_{c2}) which corresponds to $\frac{2}{3}i_{b1}$
- calculate the third-harmonic distortion from

$$D_3 = \frac{1}{2} \cdot \frac{i_{c1}/i_{c2} - 2}{i_{c1}/i_{c2} + 1} \cdot 100\%.$$

MISMATCH IN $\bar{\alpha}'$

The difference in $\bar{\alpha}'$ of the two transistors causes the two halves of a sine wave to be unequal (assuming current drive). The distortion at any output current can be calculated for any two specified $\bar{\alpha}'$. The

second-harmonic component can be obtained readily from the following table, where D_2 is correlated with the ratio of the two $\bar{\alpha}'$.

Ratio of $\bar{\alpha}'$	1.1	1.2	1.3	1.4	1.5	1.6	1.7	1.8	1.9	2.0
$D_2\%$	2.4	4.6	6.5	8.3	9.8	11.5	13.0	14.5	15.5	16.5

If the maximum ratio of the two $\bar{\alpha}'$ is 1.35:1, the corresponding value of D_2 is 7.5%.

The total-harmonic distortion (neglecting harmonics higher than the third) is given by

$$D_{\text{tot}} = \sqrt{\{(D_2)^2 + (D_3)^2\}}.$$

MISMATCH IN INPUT IMPEDANCES

Any difference in the input impedances of the two halves of the circuit causes second-harmonic distortion. The amount of distortion is dependent also on the source impedance, high source impedances giving less distortion. In general, with a transistor driver stage, the source impedance is high, and mismatch of the input impedances is not of any serious consequence.

MISMATCH IN CUT-OFF FREQUENCIES

If the cut-off frequencies of the transistors are not equal, the phase shifts of the two halves of a sine wave are unequal, the difference becoming larger as the frequency is increased. The resulting distortion is somewhat similar to crossover distortion and is unpleasant, the intermodulation distortion of frequencies higher than the cut-off frequencies being fairly high. In designing complete amplifiers, it is therefore preferable to limit the upper cut-off frequency of the amplifier to below the cut-off frequency of the output stage, the frequency-restricting circuit being inserted prior to the output stage.

CROSSOVER DISTORTION

Crossover distortion occurs if the change-over in current from one transistor to the other is not smooth. If the composite transfer characteristic is not a straight line, but shows discontinuities or changes in slope, then crossover distortion results. This distortion results in a large amount of intermodulation distortion, and is the most objectionable form of distortion occurring in class B output stages.

To eliminate crossover distortion, it is necessary to bias the transistors such that the resultant transfer characteristic is a straight line. The optimum bias conditions, as given in the published data, ensure that the performance is satisfactory under all normal operating conditions. Crossover distortion will re-appear, however, under extremes of temperature or battery voltage.

Temperature-sensitive elements, such as n.t.c. thermistors or junction

diodes, can be used to compensate for variations in ambient temperature; and non-linear resistive elements, such as copper-oxide rectifiers, or lamps of suitable resistance, can give compensation for voltage variations.

Emitter feedback offers one way of minimising crossover distortion at low battery voltages. This method was discussed in the section on split-load stages (p. 150).

HOLE STORAGE

Hole storage in the base region produces small pulses of current at the instants when the current in the two transistors changes over. These pulses may make the driver transformer ring, if its leakage inductance is high. The use of bifilar windings or a CR damping circuit across the primary will decrease this effect, where it is troublesome.

PRACTICAL CIRCUITS

Some practical class B push-pull output stages are given in Figs. 9 to 13. The values of the dropper resistors are not given, as these will depend on the current drawn by the preceding stages. Complete amplifiers based on the circuitry given in Figs. 11 and 13 are included in Chapter 16 (pp. 171-2).

The driver and output transistors for the circuits illustrated are supplied together in a plastic packet. The LCR3 packet contains an OC82D driver transistor and a matched pair of OC26 output transistors. The LFH3 packet consists of an OC81D driver transistor and a matched pair of OC81 output transistors.

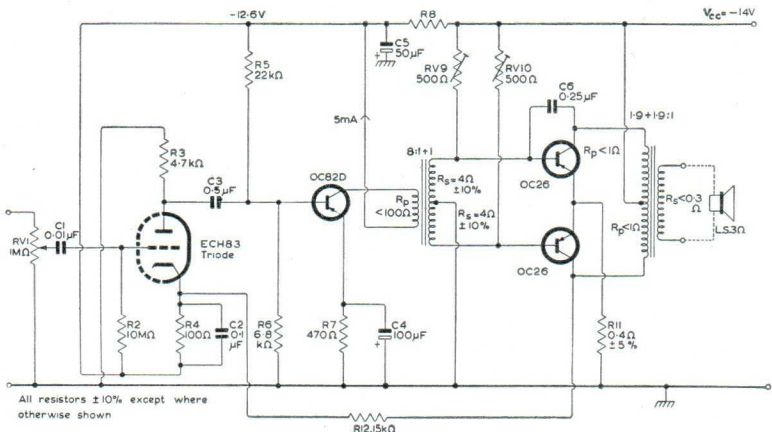


Fig. 9—Common-emitter 7W push-pull amplifier for hybrid car-radio receiver. The maximum drive (r.m.s.) required at the base of the driver transistor is $70\mu\text{A}$ and 40mV for full output. The output transistors must be mounted on heat sinks giving a thermal resistance of 7°C/W per transistor. Ambient temperature not to exceed 55°C .

CLASS B PUSH-PULL OUTPUT STAGES

The advantage of these packets is that the possible spread in performance is reduced and circuit design is simplified. Transistors from one packet should not be interchanged with those from another, or loss of performance may result.

Transformers for all the LCR3 and LFH3 circuits can be supplied by R. F. Gilson Ltd. For the 25 Ω speaker for Fig. 12, application may be made to TSL.

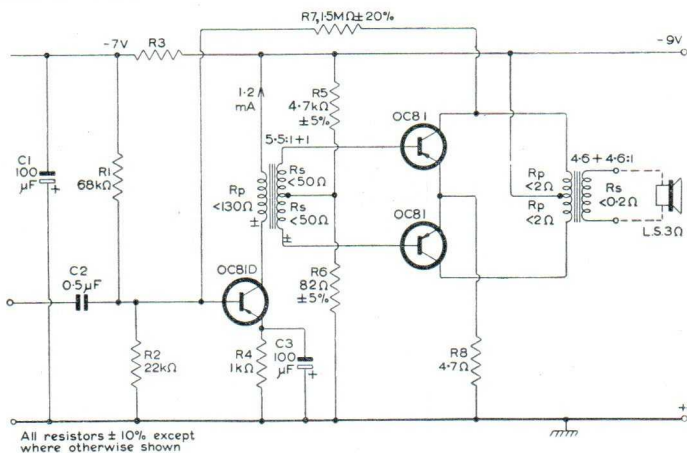


Fig. 10—Common-emitter 500mW push-pull amplifier. The input current (r.m.s.) for full output is 6.4 to 10 μ A. For operation at ambient temperatures not exceeding 45°C, the transistors may be mounted in free air, without cooling clips.

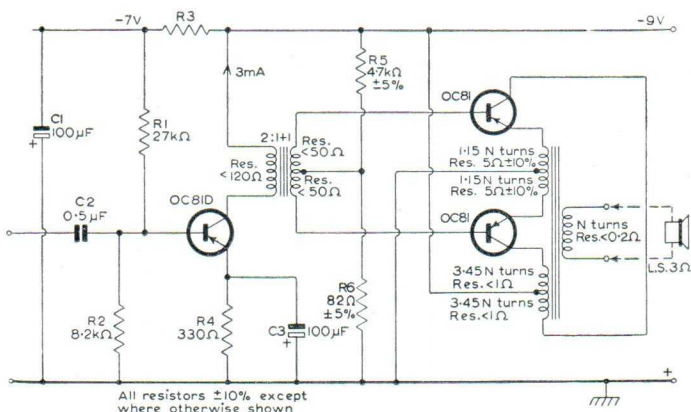


Fig. 11—Split-load 540mW push-pull amplifier. The input current (r.m.s.) for full output is 5.5 to 17.5 μ A. For operation at ambient temperatures not exceeding 45°C, the transistors may be mounted in free air, without cooling clips.

CLASS B PUSH-PULL OUTPUT STAGES

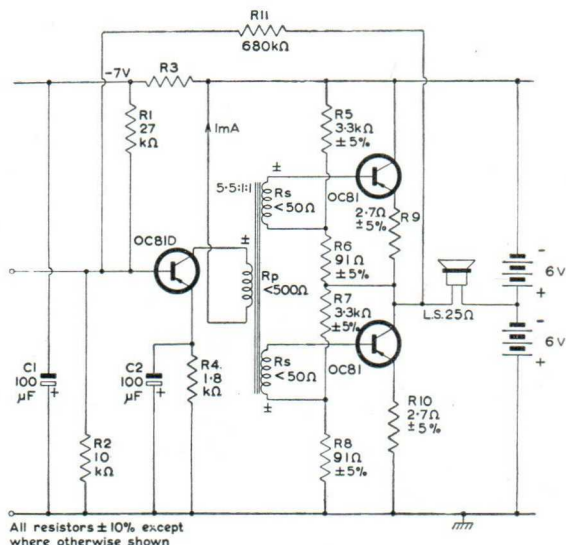


Fig. 12—Single-ended 500mW push-pull amplifier. The input current (r.m.s.) for full output is 10 to 15.5μA. For operation at ambient temperatures not exceeding 45°C, the transistors may be mounted in free air, without cooling clips.

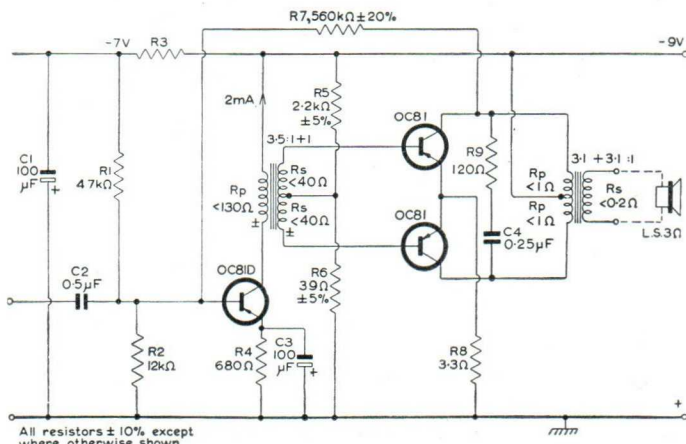


Fig. 13—Common-emitter 1W push-pull amplifier. The input current (r.m.s.) for full output is 16.5 to 26.5μA. The transistors must be mounted on heat sinks of 5×7cm of 16 s.w.g. aluminium. The transistor is bolted down to its heat sink by means of a close-fitting cooling clip (obtainable from Kimber - Allen Ltd. or distributors). $T_{amb} \leq 45^{\circ}\text{C}$.

CHAPTER FIFTEEN

HEARING AIDS

The OC57, OC58, OC59 and OC60 form a range of transistors specially developed for hearing aids. These transistors are cylindrical in shape and are only 4mm long and 3mm in diameter. They are already available to the industry and will be generally available in due course. These types are in keeping with the general trend towards miniaturisation of the components of the hearing aid.

Using these transistors, a complete hearing aid can be mounted in spectacle frames, or as a clip-on unit concealed behind the ear or in the hair or clothing.

Three hearing-aid circuits will now be described as an illustration of the general design principles of audio amplifiers, and in particular for their bearing on the operation of transistors at low levels and from low supply voltages. These circuits are not intended to be typical of modern commercial practice. The first two circuits are RC coupled; they represent a line of development which has led to extremely compact designs, such as the third circuit, which is directly coupled.

FOUR-TRANSISTOR RC-COUPLED CIRCUIT

Before special transistors were introduced for the purpose, hearing aids were designed using the standard small-signal a.f. transistors, that is, the OC70 and OC71. Fig. 1 shows a circuit of this type, in which the supply voltage is 2.4V.

The usual considerations of d.c. stability show that better stabilisation is required at lower collector supply voltages. Good stability is therefore provided in the circuit of Fig. 1, but this necessarily means using a large number of resistors and capacitors. Also, the OC70 and OC71, which are approximately cylindrical in shape, are about 5mm in diameter and 15mm long. Consequently the complete hearing aid is bulky by modern standards, although of course a tremendous advance over designs equipped with subminiature valves, as regards both size and battery consumption.

Later on, some reduction in size was made possible by substituting, in this circuit, the OC65 for the OC70 and the OC66 for the OC71. The OC65 and OC66 were in a metal construction, which was about

3mm × 4mm in cross-section and 7mm long. These two transistors were the forerunners of the OC57 series.

Demoded though this circuit is, it illustrates many of the techniques discussed in earlier chapters, and therefore will be discussed more fully than its present commercial position would warrant.

Noise

The first two stages are operated at currents of only 0.3mA to minimise noise. The gain control RV6 is sited between the first and second stages

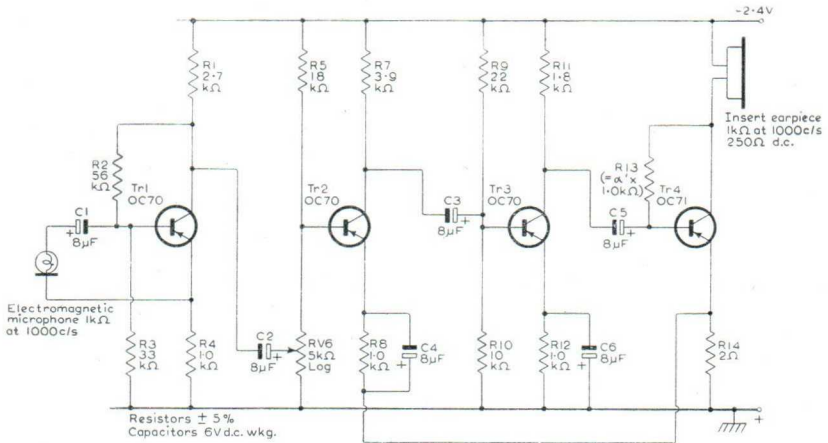


Fig. 1—Four-transistor RC-coupled hearing aid

to ensure a low noise level, while keeping this control clear of the feedback loop.

Input Stage

The input impedance of the amplifier is about 1kΩ. This matches the impedance of the electromagnetic microphone, which is specified as 1kΩ at 1000c/s. An arrangement similar to that described previously for the 250V high-gain preamplifier provides a.c. and d.c. feedback, except that extra resistance is not required in series with the microphone. The unbypassed emitter resistor R4 does not contribute a.c. feedback, since the input is applied between base and emitter.

Driver Stage

The collector current in the driver transistor Tr3 is 0.5mA. A low collector load resistance is required in this stage to ensure adequate swing up to full drive without clipping.

Output Stage

The output stage is designed round an OC71 to provide an output of nearly 2mW to the earpiece. Since in class A the maximum theoretical efficiency is 50%, a collector dissipation of about 4mW is required. This is provided by a collector current of 2mA and a collector-emitter voltage of 2V, the load impedance being $2V/2mA = 1k\Omega$ (at 1000c/s). The d.c. resistance of the earpiece is 250Ω , which at 2mA drops $2 \times 250 = 0.5V$, and leaves nearly 2V across the transistor.

The collector current is set to 2mA by the choice of the feedback resistor R13, the value of which should be α' times the $1k\Omega$ load impedance. Thus for a transistor having an α' of 47, R13 should be $47k\Omega$.

R13, in addition to biasing the transistor, provides a.c. and d.c. feedback, the d.c. stability being satisfactory with this method of stabilisation because of the higher collector current of 2mA, and because R13 is chosen to match individual transistors.

Feedback Loop

In addition to the a.c. feedback provided by R2 in the input stage and R13 in the output stage, the gain is stabilised by 12dB of negative feedback taken over the last three stages. A feedback voltage proportional to the output current is taken from across R14 in the emitter of the output stage, and injected into the second stage in series with the bypassed emitter resistor R8. The 2Ω resistance for R14 can be made from approximately 7in. of Eureka wire 0.0092in. in diameter and having a resistance of $10.6\Omega/\text{yard}$ (1 in. = 2.54cm, 1 yd. = 36 in.).

Performance

The power gains in the four stages are 17, 21, 18 and 30dB, giving a basic total of 86dB. As there is 12dB of loop feedback, the overall power gain reduces to 74dB.

The frequency response of the amplifier itself is 2dB down at 150c/s relative to the response at middle frequencies, the electrical response being virtually flat from 1000c/s to 7kc/s. The overall acoustical response of a complete hearing aid would depend primarily on the microphone and earphone.

The performance is satisfactory up to an ambient temperature of 40°C (104°F).

Total-harmonic distortion, measured at a test frequency of 400c/s,

is 5% at full output.

Current drain is 3.5mA with the correctly matched value for R13.

THREE-TRANSISTOR RC-COUPLED CIRCUIT

The use of the latest hearing-aid transistors in an RC-coupled circuit is illustrated by Fig. 2.

In this circuit the electrical power gain is about 85dB using only three transistors. The earpiece should have an impedance of 650Ω and a d.c. resistance of 200Ω , and the microphone should be a magnetic type of $2k\Omega$ impedance. The loss in the microphone and earpiece is about 35dB, so that the acoustical or air-to-air gain is $85 - 35 = 50\text{dB}$.

In the OC58 output stage, the collector current is about 2mA, and at full supply voltage the output power is then about 0.5mW, which is

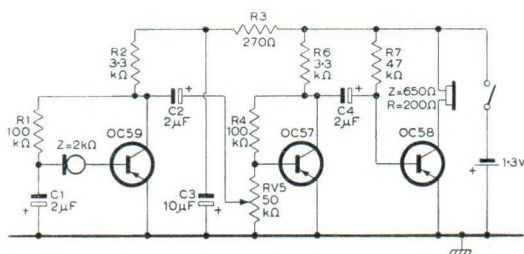


Fig. 2—Three-transistor RC-coupled hearing aid

sufficient for this application. R7 does not have to be set for individual transistors.

In the OC57 driver stage and OC59 input stage the collector current is 0.25mA. R1 and R4 provide sufficiently good d.c. stability to permit a wide range of ambient temperature. R4 introduces some a.c. feedback, but there is little loss in gain, because the input impedance falls and reduces the loss in the coupling network. C1 prevents a.c. feedback in the input stage.

By siting the volume control between the input and driver stages, the amplification of contact noise is reduced.

With a mercury cell (such as the Mallory RM625) the decoupling components for the first stage, R3 and C3, become unnecessary.

Average current drain is 2.7mA, and a life of about 100 hours can be obtained from a Mallory cell RM625. Total-harmonic distortion is 5% at 0.4mW, and the electrical frequency response is flat to within 0.25dB from 100c/s to 4kc/s.

THREE-TRANSISTOR DIRECTLY COUPLED CIRCUIT

The previous circuit, if it is used with a mercury cell so that R_3 and C_3 may be omitted, still contains five fixed resistors and three capacitors. The need for a subminiature hearing aid which can be concealed in a pair of spectacles or a hair slide has led to the consideration of direct coupling for the amplifier. Such a circuit is shown in Fig. 3, and apart from the usual microphone, earpiece, battery and volume control,

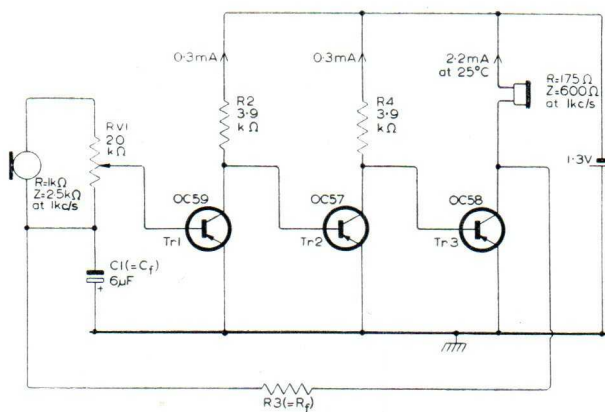


Fig. 3—Three-transistor directly coupled hearing aid

the only components required are the three transistors, three resistors and one capacitor.

The circuit is experimental rather than a production prototype. A pre-requisite of any large-scale production of a hearing aid utilising this circuit would be an examination of the effects of the production spreads of the transistors.

Choice of Circuit

The only directly coupled circuit which will give a useful power gain with three transistors consists of three grounded-emitter stages in cascade. Stabilisation is provided by a d.c. negative-feedback loop. This feedback loop governs the choice of circuit. Overall feedback is applied from output to input, so that the d.c. gain of the whole amplifier is used to compensate the temperature-dependence of the transistor characteristics.

The volume control is sited so as not appreciably to affect the d.c. working conditions.

Operating Conditions

To achieve low power consumption, the first two stages are operated at 0.3mA, which is the lowest practicable collector current. The collector voltage in each of these stages is limited to the base-emitter voltage of the following transistor. The collector voltage, which will be below the knee of the typical output characteristic, is therefore about 120 and 170mV for Tr1 and Tr2.

From the supply voltage (1.3V) and the resistance of the ear-piece, a working point of $V_c = 1.02V$, $I_c = 1.6mA$ can be derived for the output stage. This is the ideal working point at which maximum output can be obtained without clipping. In practice, to secure a useful performance over a wide range of ambient temperature, the collector current of the output stage should be set to 2.2mA at an ambient temperature of 25°C. The collector voltage is then 910mV. The ideal working point is only realised at lower ambient temperatures.

Decoupling Resistance

The value of the decoupling capacitor $C_f (= C1)$ is determined by the loss of gain which can be tolerated in the working-frequency range. A loss of 6dB at 1kc/s is considered permissible. A time-constant $C_f \times R_f$ of about one second is required, and as R_f , which is chosen to suit the characteristics of the output transistor, may be as low as 160k Ω , a nominal capacitance of 6 μF is specified for C_f .

Performance

The nominal air-to-air gain is estimated to be about 48dB. The performance should be acceptable over a range of ambient temperature from 0 to 39°C (32°F to 102°F). The predicted electrical response is within 1.5dB of the response level at 1kc/s for the range 300 to 5000c/s.

The current drain is 2.8mA at 25°C, giving a life of about 90 hours from a Mallory cell, type RM625.

CHAPTER SIXTEEN

LOW-POWER AUDIO AMPLIFIERS

Output transistors fall into two classes: the larger, higher-power ones in a construction similar to that of the OC26; and the smaller, lower-power types in a construction similar to that of the OC72. Output stages for both groups of transistor are designed along the lines given in Chapter 14. In the second group, matched pairs of transistors operating in class B give output powers at various levels up to the region of 1.0W. Such transistors are already a familiar feature of portable radios and record players.

Complete amplifiers using matched pairs of OC72 are not untypical of low-power audio amplifiers, and two 200mW circuits based on the OC72 are described in some detail in this chapter. Two circuits are then given for 540mW and 1W amplifiers using the LFH3 (OC81D driver plus a matched pair of OC81 output transistors). An experimental high-quality amplifier which provides an output of about 5W and uses OC22 and OC42 transistors is included at the end of the chapter.

6V 200mW AMPLIFIER

The operating conditions quoted in the published data for the OC72 show that a matched pair of transistors can give an output of 390mW when the transistors are fitted with the specified cooling fin, and 275mW without the cooling fins.

The circuits about to be described are the familiar ones for 200mW; these have two advantages for the general user: (a) cooling fins and special stabilisation arrangements are not required; (b) the transformers are available from a number of manufacturers.

Two versions of the 200mW audio amplifier are described, one for a 4.5V and one for a 6V supply. The 6V circuit is treated as the basic version. The performance of the two circuits is compared in Table 1.

Fig. 1 shows the 6V version of the amplifier, which gives the full 200mW output for an input of about 400mV. This input may be provided by a crystal pick-up connected to the high-impedance input terminals XX.

A low-impedance input may be connected to YY, an input of 3mV then being sufficient for full output.

The amplifier is suitable for a portable record player or, with a front-end, for a portable receiver. The same 6V supply may be used to feed the amplifier and the turntable motor.

Circuit Description

If the output stage is compared with the recommended biasing conditions given in the published data, it will be seen that some compromise of d.c. stability has been accepted. The emitter resistor has been discarded and, because of the expense, a thermistor has not been adopted in the biasing network. The stability is sufficient, however, to permit operation up to an ambient temperature of 45°C (113°F), without an undue increase in crossover distortion.

RV11 is a preset control which should be adjusted to give a combined quiescent current of 1.2mA in the output stage.

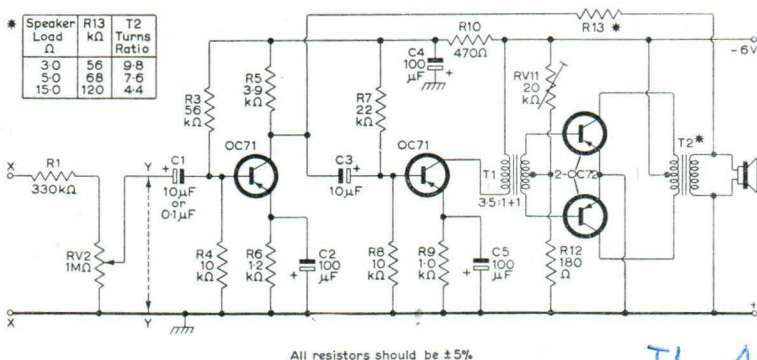


Fig. 1—200mW audio amplifier for supply voltage of 6V

Feedback from the secondary of the output transformer is applied directly to the collector of the input stage via R13. The value of R13 must be chosen to suit the impedance of the speaker, using the table in the top left-hand corner of the circuit diagram.

The OC71 driver stage is designed round a collector current of 1.5mA. Thermal stability is ensured by the 1kΩ emitter resistor.

The supply line is decoupled by R10-C4, thus reducing the distortion which would otherwise occur with the increase in battery impedance during life. The collector of the driver being much less sensitive to the feedback than the base, it has been possible to incorporate the filter at a low-current point, between the collector and base connections to the supply.

The low collector current in the input stage gives a comparatively high input impedance of several kΩ's at the terminals YY. An input impedance of about 500kΩ is needed, however, to match a crystal pick-up, and this is achieved at the terminals XX by inserting a 330kΩ

resistor R1 in series with the $1\text{M}\Omega$ gain control RV2. The load on the pick-up will be greater than $1\text{M}\Omega$ when the control is turned well down, so that maximum bass output is obtained at low volumes. When only the high-impedance input XX is to be used, C1 may be reduced from 10 to $0.1\mu\text{F}$.

Resistors of 5% tolerance are recommended, and coupling capacitors of $10\mu\text{F}$ prevent any serious reduction in bass response. The emitter resistors in the input and driver stages are bypassed by $100\mu\text{F}$ capacitors. In addition, the filter capacitor C4, to be effective at all audio frequencies, has to be $100\mu\text{F}$. The capacitors should be 6V d.c. working.

OUTPUT TRANSFORMER

With a matched pair of transistors operating in push-pull, there is no d.c. component of magnetic field, and each half of the primary must be capable of carrying the peak current of 83mA without saturation. As it is possible for the α' of the two transistors to be mismatched by up to 35%, the output transformer should be constructed to allow for unbalanced direct currents.

The whole of the primary should have an inductance of 0.5H , and its resistance should be as low as possible ($< 3\Omega$). The secondary resistance should be less than 5% of the load resistance. Leakage inductance should be as low as possible. Turns ratios (whole primary to secondary) for various speaker loads are given in the table on the circuit diagram. The output transformer is rated at 250mW.

DRIVER TRANSFORMER

The primary inductance should be 10H at a primary direct current of 1.5mA. Lower primary inductances give poorer bass response. The turns ratio is 3.5:1+1. The d.c. resistance of the primary should be less than 200Ω and that of each half of the secondary less than 50Ω , and the d.c. resistances of the two halves of the secondary should be equal. Leakage inductance should be as low as possible. The secondary should preferably be bifilar wound to give closer coupling between the two halves. The driver transformer is rated at 2mW.

4.5V 200mW AMPLIFIER

A version of the 200mW amplifier for operation from a 4.5V supply is shown in Fig. 2. Like the 6V version, this amplifier can be driven to full output by a crystal pick-up connected to the terminals XX. The two circuits are similar in principle, but there are many differences in detail. Attention will be drawn only to the most important of these.

The output stage is biased by presetting RV11 to give a total quiescent current in the stage of $1.3\text{mA} \pm 10\%$ at 20°C (68°F) or $1.6\text{mA} \pm 10\%$

LOW-POWER AUDIO AMPLIFIERS

at 25°C (77°F). With this adjustment, the circuit will operate satisfactorily at ambient temperatures from 15°C (59°F) to 45°C (113°F). The

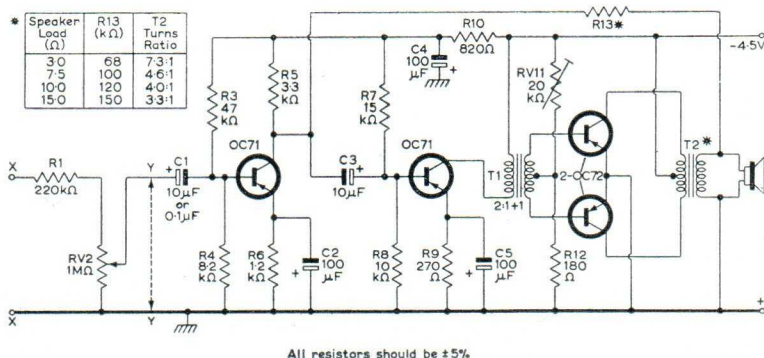


Fig. 2—200mW audio amplifier for supply voltage of 4.5V

TABLE 1

Performance of 6V and 4.5V 200mW Amplifiers

					6V	4.5V
Output power	..	..	..	..	215mW	215mW
Total-harmonic distortion	..	..	..	..	10%	10%
Sensitivity						
at YY	..	..	..	..	3.0mV	6.0mV
at XX	..	..	..	..	400mV	500mV
Frequency response						
3dB fall relative to 1kc/s at:	..	..	..	..	50c/s & 8kc/s	35c/s & 8kc/s
Current drain						
zero drive	..	..	..	..	5 to 6mA	
average music	..	..	..	..	12 to 15mA	
sine-wave drive, max output	..	..	..	..	55mA	75mA

range of operating temperature may be extended down to 10°C (50°F), by substituting for R12 (= 180Ω) a resistor of 270Ω shunted by a Varite thermistor VA1039.

Different values for the feedback resistance and turns ratio (whole primary to secondary) of the output transformer are specified, as shown in the table in the top left-hand corner of the circuit diagram. Each half of the primary should have a d.c. resistance of less than 2.5Ω. The driver stage is designed for a collector current of 3mA and the turns ratio of the driver transformer is 2:1+1. The primary inductance of this transformer should be 10H at a primary direct current of 3mA, and the d.c. resistance of the primary should be less than 150Ω. The resistance of each half of the secondary should be less than 75Ω.

OTHER OPERATING CONDITIONS FOR 2-OC72

Other possible operating conditions will be found in the published data for the 2-OC72. These conditions apply to stages which differ from those just described in that they contain an emitter stabilising resistor (Fig. 3) and have a higher quiescent current. Most of the conditions permit

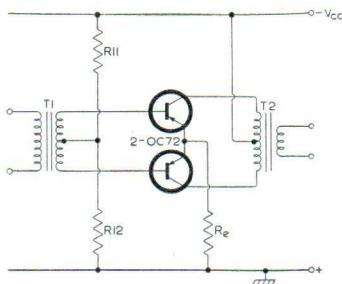


Fig. 3—Modified output stage with shared emitter resistor

higher output power, all of them give lower crossover distortion. However, if these conditions are used as the basis of design, the sensitivity will be worse and the driver stage and driver transformer will have to be redesigned, in addition to using a different output transformer.

540mW AND 1W AMPLIFIERS

The 1W amplifier (Fig. 4) uses the LFH3 transistors in the circuit of Fig. 13, Chapter 14 (p.160). The response, relative to that at 1.5kc/s,

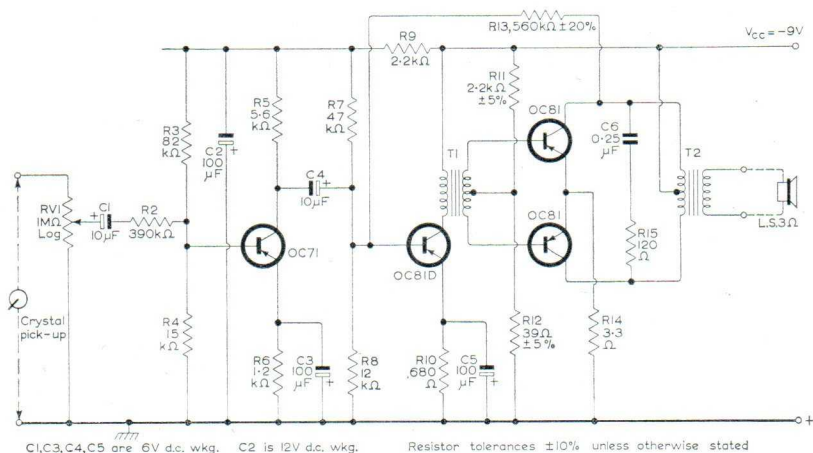


Fig. 4—1W amplifier. Colne transformers 06005 (T1) and 06006 (T2)

is 3dB down at 85c/s and 6kc/s. The OC81D and OC81 transistors must be mounted on heat sinks of 5×7 cm of 16 s.w.g. aluminium for operation at ambient temperatures up to but not exceeding $45^\circ C$. The tran-

sistor is bolted down to the heat sink by means of a close-fitting cooling clip (obtainable from Kimber-Allen Ltd. or distributors).

The 540mW amplifier (Fig. 5) uses LFH3 transistors in a split-load circuit similar to that of Fig. 11, Chapter 14 (p. 159). The bias network has been slightly modified to simplify the construction of the driver

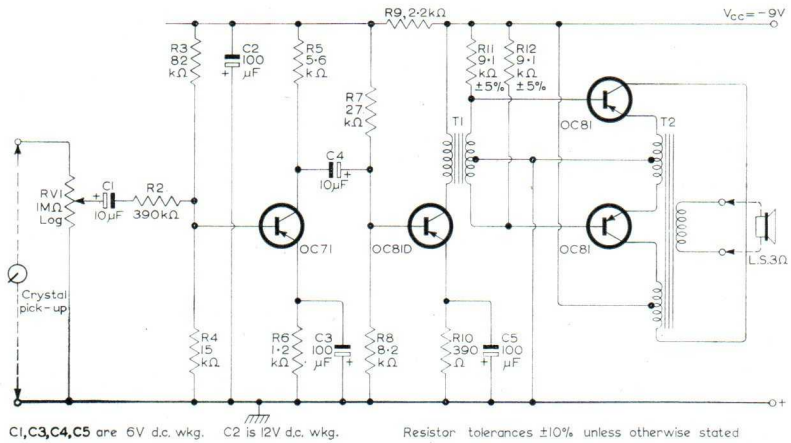


Fig. 5—540mW amplifier. Colne transformers 06003 (T1) and 06004 (T2)

transformer. The response, relative to that at 1.5kc/s, is 3dB down at 110c/s and 4.3kc/s. If the LFH3 transistors are mounted in free air without cooling clips, ambient temperatures of up to 45°C are permissible.

The sensitivity in both circuits is 250 to 350mV for full output at 10% total-harmonic distortion and for a 9V supply.

EXPERIMENTAL HIGH-QUALITY 5W AMPLIFIER

Circuit Description

An experimental high-quality amplifier is illustrated in Fig. 6. The output power is 5W into a resistive load, the total power gain being 63dB at 1kc/s. To extend the frequency response at high and low audio frequencies, high-frequency transistors and direct interstage coupling are used throughout. A single negative-feedback loop further extends the frequency response and reduces internal non-linearity distortion. The amplifier will deliver its rated output power at ambient temperatures up to 45°C.

The output transistors are operated in class A push-pull, because low distortion is a requirement. OC22 transistors have been chosen

because of their high cut-off frequency and the good linearity of the V_b/I_c characteristic. The transistors do not need to be exactly matched, but they should not be grossly mismatched. The maximum permissible collector-junction temperature of 90°C allows a quiescent collector current of 600mA at an ambient temperature of 45°C with a 14V supply. Temperature stabilisation is effected by the emitter resistance R20.

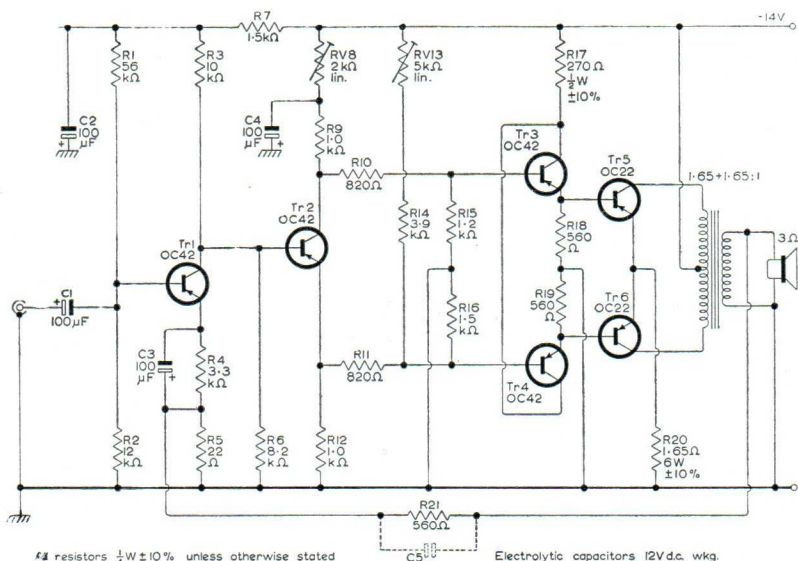


Fig. 6—Experimental 5W high-quality audio amplifier

which, being common to both emitters, introduces no degenerative feedback in a class A stage. This resistor may be made up from two 3.3Ω , 3W wirewound resistors in parallel.

Each OC22 is driven by an OC42 'emitter follower', the base voltage of the driver stage, and therefore the base bias of the output stage, being derived from a resistive potential divider. Two potential dividers couple the collector and emitter of a conventional phase splitter (Tr2) to the bases of the two driver transistors. Equal a.c. (signal) loads are presented to the collector and emitter of Tr2.

A conventional grounded-emitter amplifier (Tr1) drives the phase splitter. Negative feedback from the output transformer is applied to the emitter circuit of this stage.

The quiescent current in each OC22, and therefore the balance, is determined by the settings of RV8 and RV13. The following sequence of adjustments for the initial setting-up of these controls is recommended:

- (1) with suitable meters in the collector circuits of Tr5 and Tr6, RV8 is adjusted until the current in Tr5 is 600mA
- (2) RV13 is then adjusted to make the current in Tr6 equal to 600mA
- (3) RV8 is re-adjusted to return the current in Tr5 to 600mA
- (4) repeat (2) and (3) in that order as necessary.

PERFORMANCE

The performance of an experimental model of the amplifier, with 17dB of negative feedback, was:

Max. power output at 1kc/s 5W (into 3Ω load)

Frequency response

(a) 1W = 0dB -3dB at 3c/s & 50kc/s*

(b) 5W = 0dB -3dB at 7c/s & 40kc/s

Total-harmonic distortion

(a) 1W at 1kc/s 0.3%

(b) 5W at 1kc/s 0.6%

Input impedance 8.3kΩ

Sensitivity 16.8μA for 5W at 1kc/s

*A peak of +2.5dB occurs at 40kc/s following a continuous increase from 10kc/s (0dB)

OUTPUT TRANSFORMER

The output transformer should have a turns ratio of 1.65+1.65:1 (for a 3Ω load). The primary resistance should be less than 0.5Ω (each half) and the secondary resistance less than 0.1Ω, and the primary inductance should be at least 1H. C5 may be inserted if necessary, depending on the output transformer used. This capacitor was not incorporated in the experimental model.

Heat Sink

Each OC22 must be provided with a heat sink having a thermal resistance of 2°C/W.

The heat sink for each OC22 may consist, as in the experimental model, of a 9½-in. strip (≈ 24cm) of extruded aluminium, type Noral 6182. The strip as supplied is 4×1¼ in. (≈ 10×3cm). The finish may be bright and the strip may be mounted in any position.

An equivalent heat sink has been calculated for 14 s.w.g. Duralumin sheet metal. Each heat sink should be 8×8in. (20.5×20.5cm). The finish again may be bright, but the heat sink must be mounted vertically, although it may be folded along the vertical axis if required.

Whichever of the two types of heat sink is adopted, each OC22 should be insulated by a mica washer and insulating bushes. A thin smear of silicone grease should be provided between the washer and the heat sink. The two heat sinks allow a large margin of safety.

CHAPTER SEVENTEEN

PUBLIC-ADDRESS AMPLIFIERS

In this chapter two versions of a public-address amplifier are described, one for a 14V and one for a 28V supply. The output power is 15W at less than 4% total-harmonic distortion. The sensitivity is sufficient for the amplifiers to be driven fully by low-impedance microphones, and the performance is more than adequate for the intended application.

The 14V version of the circuit can be regarded as the basic amplifier and will be described first; it is then only necessary to describe the more important differences in the 28V circuit.

The performance of the two versions of the circuit is compared in Table 1. The amplifiers are suitable for operation at ambient temperatures which normally do not exceed 45°C, but occasional rises up to 55°C are permissible.

14V CIRCUIT

The basic amplifier (Fig. 1) is designed for a supply of 14V, which is the average voltage of a fully charged 12V accumulator.

An output stage of the type shown in Fig. 4 of Chapter 14 has been adopted. The output is provided by two matched OC26 operating in the common-emitter configuration in symmetrical class B push-pull.

The output stage is preceded by an OC26 driver and by OC72 and OC71 amplifier stages.

Output Stage

Although in the circuit of Fig. 1 the collectors are connected to the negative line, the transistors operate as common-emitter amplifiers. The advantage of this arrangement, it will be remembered, is that the output transistors can be mounted directly on a common heat sink connected to the negative line.

Emitter resistances of 0.5Ω are necessary for thermal stability at the higher ambient temperatures. To minimise crossover distortion, the quiescent currents are set individually to 30mA at normal ambient temperatures (~ 25°C) by means of RV19 and RV20. Each of these variable resistors forms the upper half of a potential divider biasing the appropriate transistor. The lower half is formed by the resistance of half the secondary.

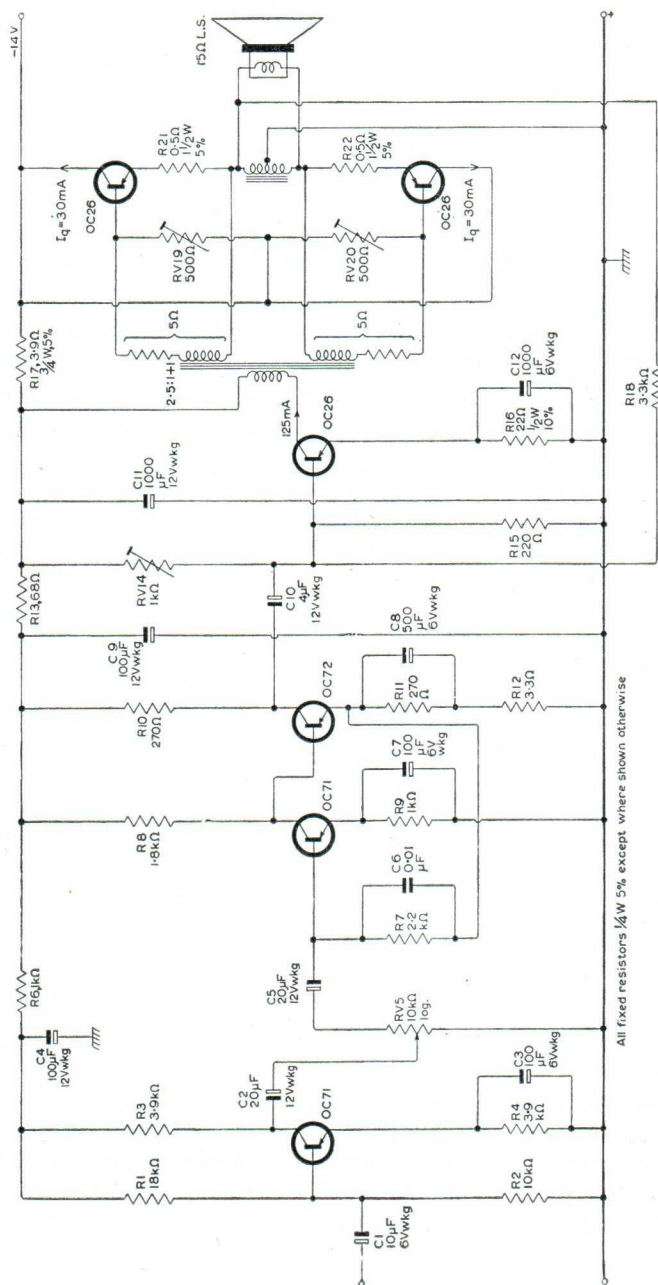


Fig. 1—15W public-address audio amplifier for supply voltage of 14V

The optimum load per transistor is 4.25Ω , of which 0.5Ω is provided by the emitter resistor and 3.75Ω has to be matched to the speaker. A centre-tapped choke, which effectively acts as a 2:1 auto transformer, provides the 4:1 impedance ratio for matching the 15Ω speaker. This choke is cheaper and more efficient than a conventional transformer.

The peak current on full drive is 3.0A , and the maximum transistor dissipation is 5W at an ambient temperature of 55°C . The maximum

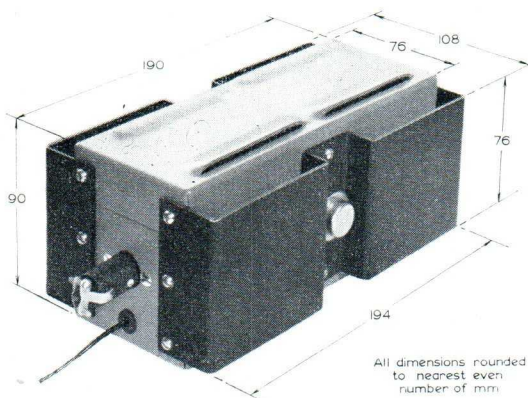


Fig. 2—Approximate dimensions of heat sinks for OC26 output transistors made from 2mm blackened-aluminium sheet metal.

junction temperature is 85°C , with the transistors mounted on a heat sink giving a total thermal resistance from junction to ambient of 6°C/W . The transistors are thermally stable at this junction temperature. Compact and cheap heat sinks can easily be made to give the required total thermal resistance (Fig. 2).

Negative feedback is applied from the output to the base of the driver to minimise distortion.

OUTPUT TRANSFORMER (14V)

For a 15Ω speaker a centre-tapped choke is suitable.

Total d.c. resistance	$< 0.2\Omega$
Total inductance	$> 100\text{mH}$

Driver Stage

The driver stage consists of a single OC26 in a conventional class A

circuit. The collector current is adjusted to 125mA by means of RV14. The collector dissipation is considerably less than in the output stage, and the total thermal resistance, from junction to mounting base, only has to be less than 15°C/W. This value is achieved very easily by mounting the transistor with a mica washer on the chassis.

DRIVER TRANSFORMER (14V)

Turns ratio	2.5:1+1 (bifilar secondary)
Primary inductance	> 500mH at 125mA d.c.
Primary d.c. resistance	< 6Ω
Secondary d.c. resistance	5Ω+5Ω

The resistance of each half of the secondary should be $5\Omega \pm 10\%$. If the resistance of the windings is less than this, external resistances must be added to make up the required value.

Amplifying Stages

The first stage is equipped with an OC71 and amplifies the signal from the microphone input. This stage is followed by OC71 and OC72 current-amplifying stages. The circuit is conventional except that the second OC71 is directly coupled to the OC72, and a.c. and d.c. negative feedback is applied over these two stages. This method of coupling requires fewer components and provides better temperature stability than two conventional RC-coupled stages.

Crossover distortion in the output stage increases at higher frequencies. It is therefore preferable to limit the upper cut-off frequency of the amplifier to about 7kc/s. C6 in the feedback loop provides the necessary limiting.

Reproduction of frequencies below 150c/s is not desirable in public-address systems. The smaller value used for C10 (4μF instead of 10μF) provides a convenient method of limiting the low-frequency response.

28V CIRCUIT

Fig. 3 shows the version of the circuit for operation from a 28V supply (usually two fully charged 12V accumulators in series). This circuit is of the transformerless push-pull type shown in Fig. 5 of Chapter 14. The 3.75Ω load impedance is provided directly by the speaker. There are also some differences in component values and ratings between the two versions of the circuit.

If a centre-tapped 28V supply is not available, an artificial centre-tap can be provided. A possible method is to connect two 24Ω, 10W resistors in series across the supply, with the centre-tap decoupled by a 1000μF capacitor of 50V d.c. wkg.

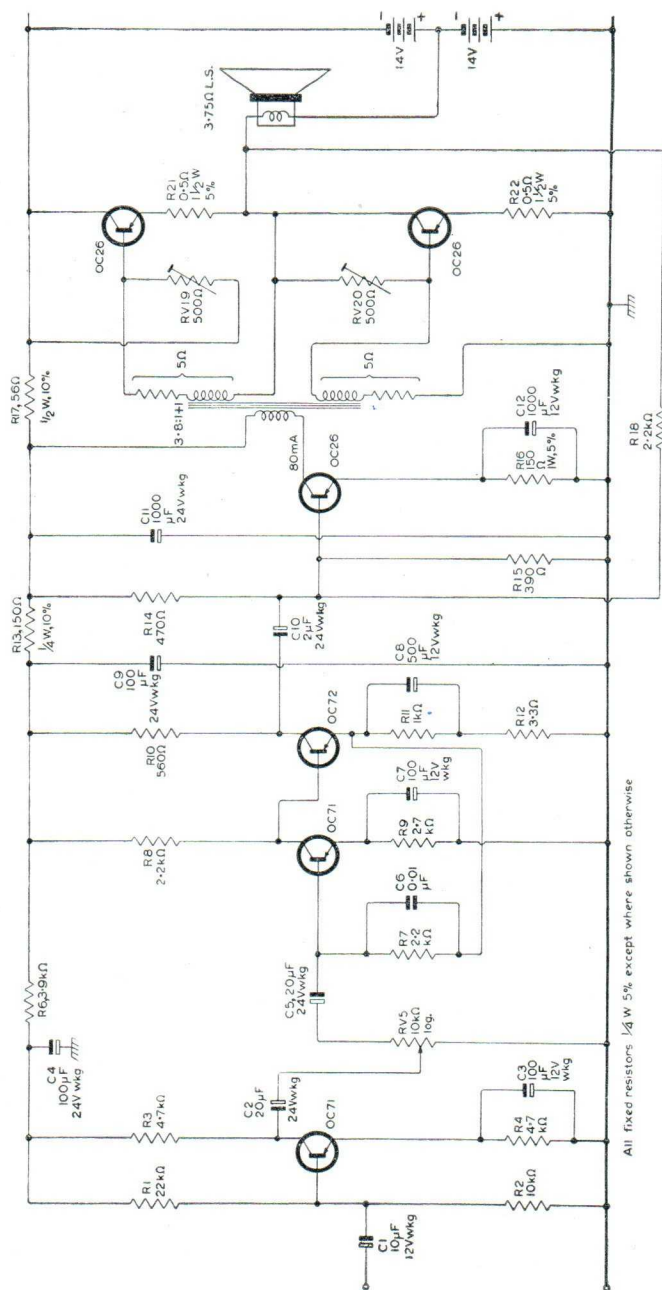


Fig. 3—15W public-address audio amplifier for supply voltage of 28V

DRIVER TRANSFORMER (28V)

Turns ratio	3·8:1+1 (bifilar secondary)
Primary inductance	> 600mH at 80mA d.c.
Primary d.c. resistance	< 8Ω
Secondary d.c. resistance	5Ω+5Ω

The resistance of each half winding of the secondary should be made up if necessary to $5\Omega \pm 10\%$.

TABLE 1
Performance of 14V and 28V Circuits

	14V	28V
Current Consumption I_q	220mA	av. 150mA
speech and music	$\simeq 800\text{mA}$	$\simeq 400\text{mA}$
Sensitivity (for full output)	0·2μA	0·1μA
impedance 1kΩ	0·2mV	0·1mV
Distortion (at full output)	< 4%	< 4%
Frequency Response	Flat within 3dB from 150c/s to 7kc/s	

HIGHER OUTPUT POWER

A higher output power, of perhaps 20W, will be available from OC26 circuits, although the sensitivity will necessarily be reduced. While such circuits will be similar in principle to those just described, the transformers and almost all the component values will be different.

CHAPTER EIGHTEEN

I.F. AMPLIFIER FOR 470kc/s

An i.f. amplifier can be designed for a.m. reception whose performance compares favourably with that obtainable with thermionic valves. This is perhaps the most interesting example of the method to be followed when using transistors as h.f. amplifiers. Equations for calculating the component values are given in fuller treatments of the subject.

INTERNAL FEEDBACK

Internal feedback is of major importance at high frequencies. It can produce instability in much the same way as feedback through the anode-grid capacitance of a triode valve. Even if oscillation does not occur, the bandpass characteristic may be highly asymmetrical.

Internal feedback can be neutralised by means of external feedback. The design of the amplifier then becomes relatively straightforward.

When both real and imaginary parts of the feedback are cancelled, the process is known as *unilateralisation*.

GAIN AND CUT-OFF FREQUENCY

A fall in gain with frequency is experienced with all transistors. The cut-off frequency is the point at which the current amplification factor falls to 3dB below its low-frequency value, and for an h.f. transistor is at least several Mc/s in grounded base. This loss of gain sets the limit to the usefulness of a transistor at high frequencies.

The gain in a grounded-emitter circuit falls off more rapidly with frequency than in grounded base, so the cut-off frequency needs to be well above the frequency of operation.

For an OC45 the maximum theoretical gain is about 38dB at 470kc/s. The stage gain in a practical narrow-band amplifier would normally be from 2 to 12dB lower than this, because of the insertion loss of the coupling elements, or intentional mismatching losses.

CHOICE OF TRANSISTOR

Two requirements therefore emerge for a transistor required to operate at h.f. First, the internal feedback should be small and not subject to too great a spread. Second, the transistor will be suitable if the cut-off frequency f_{α} in grounded base is about ten times the operating frequency.

The OC45 has been designed specially for these requirements, the average value of f_{α} being 6Mc/s.

CIRCUIT OF I.F. AMPLIFIER

Fig. 1 shows the circuit of a 470kc/s amplifier, the first stage of which will be taken as a design example. The d.c. conditions are set up in the normal way. The emitter current is 1mA, and with an emitter resistor R4 of 680 Ω the voltage at the emitter is about -0.68V . The voltage between base and emitter (across the emitter junction) is small,

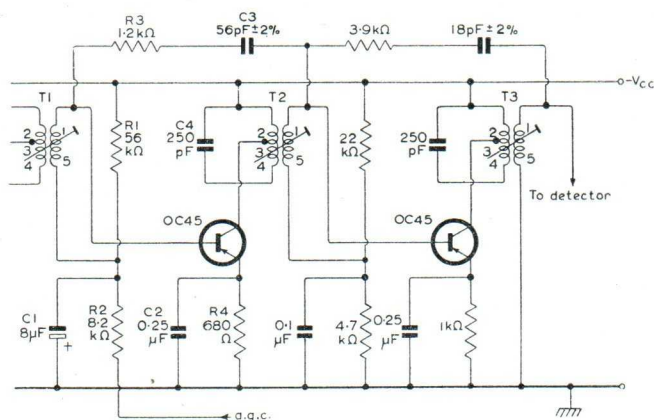


Fig. 1—Complete circuit of i.f. amplifier for 470kc/s

the voltage at the base being about -0.85V . The base voltage is fixed by the potential divider formed by R1 ($= 56\text{k}\Omega$) and R2 ($= 8.2\text{k}\Omega$), and by the voltage on the a.g.c. line.

C1 and C2 are bypass capacitors and C4 is the tuning capacitor for the coil. Unilateralisation is provided by R3 and C3 which form a feedback path from output to input. The tap on the coil allows a more convenient value for C4, as will be explained later.

T1 and T2 can be identical and are designed for a transistor output impedance of 28k Ω and an input impedance of 800 Ω .

EQUIVALENT CIRCUIT

Setting up the a.c. conditions requires a detailed knowledge of the characteristics of the transistor, and these are best expressed by means of an equivalent circuit.

There are a number of equivalent circuits which give an accurate representation of transistor characteristics. Any one of these could be

used as a basis for designing an h.f. amplifier without affecting the final result, and in fact the various equivalent circuits are only rearrangements of each other which can be obtained by normal circuit theory.

The most convenient equivalent circuit for grounded-emitter i.f. stages, however, is that shown in Fig. 2. The amplification of the

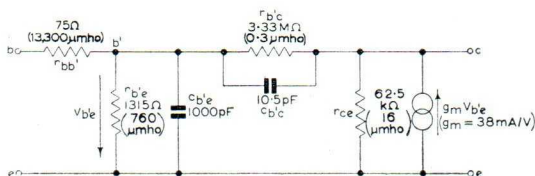


Fig. 2—Equivalent circuit of OC45 at any frequency

transistor is represented by a current generator acting directly across the output terminals. The value of the current generator (in milliamps) depends on the signal voltage $v_{b'e}$ between the points b' and e . The current generator is therefore conveniently designated by a mutual conductance, g_m , in mA/V. For most purposes, where the operating frequency is well below f_α , the values of all the circuit elements (including the generator) can be regarded as independent of frequency.

This circuit is called the hybrid π equivalent circuit. The values on the diagram are for an OC45 operating at $V_{ce} = -6V$, $I_e = 1mA$.

Provided only one frequency is being considered, the much simpler form shown in Fig. 3 can be used. (This simplified arrangement is the normal π equivalent circuit.) All the circuit elements except C_3 depend

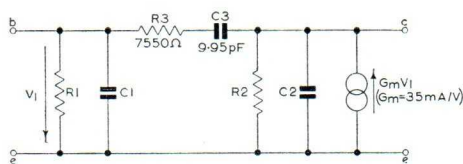


Fig. 3—Equivalent circuit of OC45 at 470kc/s

on the operating frequency. The only connection between the input and output is the feedthrough path of $R_3 = 7.55k\Omega$ in series with $C_3 = 9.95pF$.

The value of the current generator has to be changed, since it must now be defined in terms of the true input voltage v_{be} instead of the voltage $v_{b'e}$ between b' and e .

The difference between these voltages is caused by the resistance $r_{bb'}$ between the points b and b'; this resistance is an internal resistance (hence the small r) in series with the base, and is contributed by the base material lying between the active region of the base layer and the base contact.

Because of the voltage drop in $r_{bb'}$, the mutual conductance is reduced to 35mA/V, this value being denoted by G_m to distinguish it from the $g_m = 38\text{mA/V}$ of the previous circuit.

UNILATERALISATION

The effect of the internal feedback is that a voltage produced at the output terminals by the current generator will produce an unwanted voltage at the input terminals across the source impedance.

Neutralisation is effected by connecting a phase-changing transformer across the output (Fig. 4) and incorporating a suitable impedance in the feedback path. The current fed back to the input through

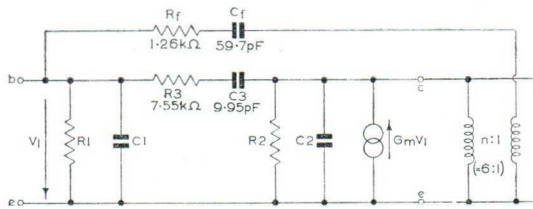


Fig. 4— π equivalent circuit with unilaterising components added

the external feedback path is equal in amplitude, but opposite in phase, to that fed back through the internal feedback path. So the total feedback is zero.

The external feedback is most effective when the transistor is unilateralised, that is, when both the real and imaginary parts of the internal feedback are cancelled. Thus if the transformer ratio were 1:1, R_f would be made equal to R_3 ($= 7.55\text{k}\Omega$) and C_f equal to C_3 ($= 9.95\text{pF}$). In fact a step-down ratio of $n:1$ is used to match the high output impedance into the low input impedance of the following stage, and $C_f = nC_3$ and $R_f = R_3/n$. For a transformer ratio of 6:1, the values will be $R_f = 1.26\text{k}\Omega$ and $C_f = 59.7\text{pF}$.

When the transistor is unilateralised, changes on the output side cannot affect the input circuit, and conversely. The transistor reduces to a box having independent input and output circuits (Fig. 5). In this condition the external feedback elements R_f and C_f merely exert a shunting effect on the input and output impedances. This will be

taken into account shortly. In practice, the power lost in R_f is often negligible and C_f may have only a slight influence on tuning.

INPUT AND OUTPUT IMPEDANCES

The input and output impedances of the unilateralised transistor can be calculated using equations derived from the equivalent circuit. The

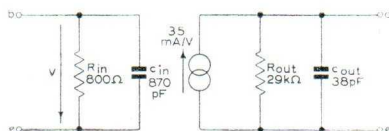


Fig. 5—Equivalent circuit of unilateralised OC45 at 470kc/s excluding effects of neutralising components.

effect of the elements in the external feedback path will not be taken into account yet.

The input resistance R_{in} is calculated as 800Ω and this is in parallel with an input capacitance c_{in} of 870pF .

To obtain the output impedance the assumption is made in the first instance that the transformer is perfectly lossless. The output resistance R_{out} is then $29\text{k}\Omega$ and the output capacitance c_{out} is 38pF .

EFFECT OF FEEDBACK ELEMENTS

The condition for the power lost in R_f to be a minimum is that

$$n = \sqrt{\frac{R_{out}}{R_{in}}} = \sqrt{\frac{29,000}{800}} = 6,$$

where R_{in} is the input resistance of the following stage.

The impedance in the external feedback path must therefore be one-sixth of that originally calculated, that is,

$$R_f = \frac{7.55}{6} = 1.26\text{k}\Omega$$

and

$$C_f = 6 \times 9.95 = 59.7\text{pF}.$$

The shunting effect of the neutralising components on the input and output impedances can now be calculated; first R_f and C_f are converted into an equivalent parallel combination of resistance and capacitance ($R_p = 26.8\text{k}\Omega$ and $C_p = 57\text{pF}$) which shunts the input impedance (Fig. 6).

Because the turns ratio of the transformer is 6:1, R_p and C_p appear on the output side of the transformer as $n^2 R_p = 36 \times 26.8 = 965\text{k}\Omega$ and $C_p/n^2 = 57/36 = 1.6\text{pF}$.

The combined input and output impedances are shown in Fig. 7, R_{out} becoming about 28k Ω .

In the final design the nearest preferred values are used for R_f ($= 1.2k\Omega$) and C_f ($= 56pF$).

MAXIMUM THEORETICAL GAIN

After unilateralisation the transistor can be considered as a box with

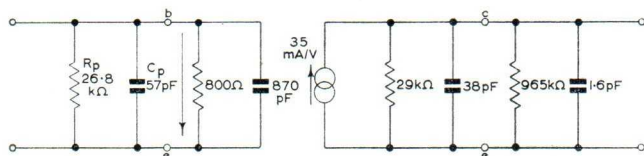


Fig. 6—Equivalent circuit of unilateralised OC45 at 470kc/s including effects of neutralising components.

independent input and output impedances, and the design of the i.f. amplifier is straightforward.

The possible stage gain can now be calculated. To obtain maximum power from the box, the load resistance R_L is made equal to the output resistance R_{out} ($= 28k\Omega$) as in Fig. 8. An inductance must also be connected across the output terminals to tune out the output

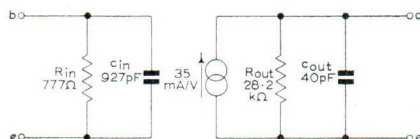


Fig. 7—Simplified equivalent circuit of unilateralised OC45 at 470kc/s including neutralising components.

capacitance. For the time being the inductance can be regarded as lossless, that is, of infinite Q .

Since the current flowing from the current generator will be divided equally between the output and load resistances, the power in the load, P_{out} , is given by:

$$\begin{aligned} P_{out} &= [I^2 R] \\ &= (\frac{1}{2} G_m V_{in})^2 R_L \\ &= \frac{1}{4} (G_m V_{in})^2 R_{out} . \end{aligned}$$

The input power is given by

$$P_{in} = \left[\frac{V^2}{R} \right] = \frac{(V_{in})^2}{R_{in}} .$$

Thus the maximum theoretical power gain is:

$$\begin{aligned}\frac{P_{out}}{P_{in}} &= \frac{1}{4}(G_m)^2 R_{in} R_{out} \\ &= \frac{1}{4} \cdot \left(\frac{35}{1000}\right)^2 \cdot 777 \cdot 28,000 = 6700 \\ &= 38\text{dB}.\end{aligned}$$

COIL DESIGN

The output capacitance of the transistor ($= 40\text{pF}$) and the input

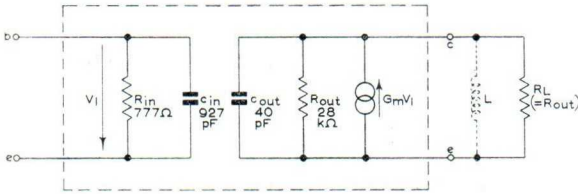


Fig. 8—Unilateralised transistor with load resistance and tuning inductance

capacitance of the following stage reflected back through the transformer ($= 927/n^2 = 26\text{pF}$) contribute a total of 66pF to the tuning capacitance. However, using 66pF as the sole tuning capacitance would give far too wide a bandwidth. The bandwidth can be reduced without

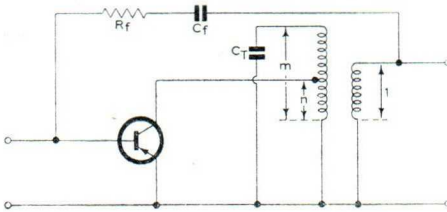


Fig. 9—Practical arrangement based on circuit shown in Fig. 8: d.c. components not shown.

loss of gain by adding extra capacitance and reducing the inductance to tune again to 470kc/s .

To avoid inconveniently large values for the extra capacitance C a tapped primary is used (Fig. 9). The capacitance C_T appears as a much larger capacitance at the transistor collector, the positioning of the tap being determined from

$$\frac{m}{n} = \sqrt{\frac{C}{C_T}}.$$

At the same time the ratio $n:1$ is maintained between primary and secondary.

Stability

With an unloaded Q of 100 a sufficient margin of stability will normally be assured in a two-stage i.f. amplifier.

Coil Loss

The loss in a practical coil depends on the initial Q and is given by the following expression:

$$\frac{P_{\text{out with practical coil}}}{P_{\text{out with perfect coil}}} = \left(\frac{Q_0 - Q_w}{Q_0} \right)^2$$

where Q_0 is the unloaded Q and Q_w is the working Q .

The insertion loss in decibels is

$$20 \log \frac{Q_0}{Q_0 - Q_w} \text{ dB}$$

and using practical values of $Q_0 = 100$ and $Q_w = 56$ the loss is 7dB.

ACTUAL STAGE GAIN

The actual stage gain is the maximum theoretical gain minus the coil loss, that is, $38 - 7 = 31\text{dB}$.

The amplifier gain is limited to 31dB by allowing a coil loss of 7dB. By not using the maximum gain of the amplifier, the circuit can be made stable, allowance being made for the spread in transistor gain and other circuit tolerances, such as those introduced by the neutralisation components and the Q 's of the coils.

CHAPTER NINETEEN

RADIO RECEIVERS

The six-transistor receiver described in this chapter is intended primarily for portable equipment, and has a performance comparable with that of a conventional four-valve portable receiver. However, the performance is also acceptable for normal indoor listening.

The receiver can be constructed to be carried as easily as a handbag. Only slight variations in the design are needed to adapt the circuit for either a miniature personal receiver or else a larger transportable 'set about the house'.

The set is designed for medium- and long-wave reception. It compares favourably as regards sensitivity and output power with typical four-valve battery receivers, and has a much lower battery consumption. The receiver gives 200mW at full output, and has a sensitivity of approximately $500\mu\text{V}/\text{metre}$ (or $20\mu\text{V}$ at the base of the mixer) for an output of 50mW.

The circuit in this as in portable valve equipment is necessarily a compromise between performance and cost. Component and transistor tolerances have been examined, so that the permitted spreads should not create any difficulty.

Special precautions in the layout are required only to ensure that the aerial is not heavily damped by adjacent components, and that appreciable feedback does not occur between the i.f. or a.f. stages and the aerial.

GENERAL DESCRIPTION

Six transistors and one germanium diode are used in the basic form of the receiver shown in Fig. 1. The OC44 is the frequency changer and the two OC45 form the i.f. amplifier. The detector is the OA70 germanium diode. Three transistors make up the audio stages, an OC71 being used to drive a matched pair of OC72 (the 2-OC72) in the transformerless push-pull output stage.

A superhet circuit is chosen for the same reasons that apply to valve receivers. The aerial is a Ferroxcube rod, which gives the required selectivity and sensitivity and can be coupled conveniently to the frequency changer.

The standard i.f. of 470kc/s is used, and the local-oscillator frequency is above the signal frequency, in accordance with normal practice.



Fig. 1—Original version of portable radio receiver

The OC44 operates as a self-oscillating mixer. The r.f. signals from the aerial coupling coil are fed into the base of the OC44, which produces its own local oscillation by means of feedback from the collector to the emitter. Tracking of the aerial and oscillator coil is obtained in a conventional way by means of a tuning capacitor with shaped oscillator vanes. The i.f. is selected at the collector of the OC44 by the first i.f. transformer T3.

The i.f. amplifier consists of two OC45 operating in unilateralised grounded-emitter circuits. The choice of bandwidth is a compromise between quality and selectivity. To obtain satisfactory adjacent-channel rejection, the i.f. bandwidth has been reduced. The resulting treble attenuation will not generally be noticeable during normal listening. In miniature receivers, in which the bass response is also reduced, the compromise treble response is completely adequate.

Double-tuned i.f. transformers could be used to improve either or both the frequency response and adjacent-channel rejection.

The third i.f. transformer T5 is connected to an OA70 detector diode which provides an audio output and a d.c. output. The d.c. output is fed back to control the operating current of the first i.f. transistor, so providing automatic gain control.

The a.f. output from the OA70 is taken to an OC71 driver stage. The OC71 is transformer-coupled to a class B output stage consisting of the 2-OC72. The output stage is of the transformerless or 'single-ended' type, in which no output transformer is required and the loudspeaker forms a direct load for the output transistors. A loudspeaker with a 35Ω speech coil provides the correct load for an output of 200mW. Negative feedback is applied to the emitter of the OC71 from the loudspeaker terminal.

A battery voltage of 9V is selected and this voltage is allocated in the following way:

- (a) the h.f. transistors work at the collector-emitter voltage of 6V which gives maximum gain
- (b) a voltage drop of about 1V is allowed across the emitter resistors of the h.f. transistors for stabilisation of the working point
- (c) a drop of 2V is allowed across the decoupling resistor from the audio output stage, so that the decoupling resistance can be high enough to make a very high decoupling capacitance unnecessary.

AERIAL CIRCUIT

In general, in order to obtain the maximum possible power from the Ferroxcube rod aerial, the rod should be as long as is practicable in a given cabinet. Increasing the diameter of the rod increases the Q and the power output of the aerial, but the choice of diameter is limited by the increased weight and cost of the thicker rods. The FX1268 Ferroxcube rod chosen for this receiver is 7in. long and has a diameter of approximately $\frac{5}{8}$ in.

The aerial is illustrated in Fig. 2. There are separate windings for the medium- and long-wave bands. The coils are placed at opposite ends of the rod, the centre of each being approximately $1\frac{3}{8}$ in. from the end. This spacing minimises interaction between the coils. Small adjustments

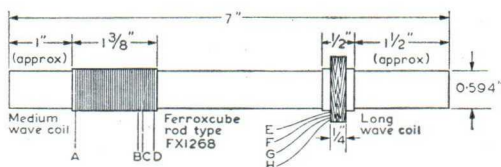


Fig. 2—Dimensions of aerial components (1 in. = 2.54cm)

Coil Details

Medium-wave coil (single-layer windings)

Coil AB: 64 turns, 19/0-0028 bunched conductors

Coil CD: 6 turns, 19/0-0028 bunched conductors

Long-wave coil (wave-wound)

Coil EF: 41 turns 0.0076 in. rayon-covered, enamelled wire

Coil GH: 175 turns 0.0076 in. rayon-covered, enamelled wire

to the coil inductance can be made when aligning the receiver by sliding the coils along the rod.

If the long-wave aerial coil is left open-circuited during medium-wave operation, the coil can resonate with its self-capacitance at a frequency in the medium-wave band. The resonance causes heavy damping of the medium-wave coil at that frequency. The long-wave coil is therefore short-circuited by SA1 during operation in the medium-wave band. The medium-wave coil is left open-circuited during long-wave operation.

The aerial is coupled to the frequency changer by low-impedance coils placed adjacent to the aerial coils. The number of turns on the medium-wave coupling coil has been adjusted to reduce the Q of the aerial from an initial unloaded value of 200 (measured at 1Mc/s) to a working value of 110 when the aerial coil is loaded by the input resistance of the frequency changer. The ratio of unloaded to loaded Q of almost 2:1 gives approximately maximum transfer of power from the aerial to the frequency changer, and also effects a reasonable compromise between bandwidth (9kc/s) and second-channel rejection. For

long-wave operation, the Q of the aerial is reduced from an unloaded value of 80 to a working value of 22, again to obtain a bandwidth of 9kc/s.

FREQUENCY CHANGER

The frequency changer in a transistor receiver may be one of two types. It may consist either of a separately excited mixer requiring a second transistor as a local oscillator, or of a self-oscillating mixer in which one transistor combines the functions of mixer and oscillator. There is little difference in performance with either type, but the self-oscillating mixer has two particular advantages, lower cost and better frequency stability. Both circuits can be designed for satisfactory frequency stability, however, so that cost is the main consideration.

In this receiver the mixer is of the self-oscillating type and is designed round an OC44. The input signal is taken from the aerial coupling coil by way of the wave-band selector switch to the base of the transistor, the base being the more sensitive input electrode. Oscillator feedback is taken from the collector to the emitter through low-impedance coupling windings on the oscillator coil.

The oscillator tuned circuit is similar to that used in valve receivers, since the capacitance reflected from the transistor is very small, only about 1pF. The Q of the oscillator coil is somewhat higher than in valve receivers to allow for transistor damping.

To ensure easy starting for the oscillator, the transistor is biased initially in class A by the normal d.c. stabilisation circuit. As the amplitude of oscillation increases, rectification of the oscillator voltage at the emitter causes a steady negative voltage to be developed across the emitter resistor and bypass capacitor. This voltage tends to drive the transistor into class B and also stabilises the amplitude of oscillation. At the same time the quiescent current increases slightly.

Operating Current

The direct emitter current in the OC44 was chosen to be 0.25mA (it rises to about 0.3mA when the oscillator is functioning) for two reasons:

- (a) the noise level is a minimum when the current is in the region of 0.25mA
- (b) the cut-off frequency of the mutual conductance at this current is approximately equal to the alpha cut-off frequency f_α , which is 15Mc/s.

To maintain a high working Q and good frequency stability, the

oscillator coil is only lightly loaded by the input resistance of the transistor at its emitter. The emitter is thus voltage- rather than current-driven at the frequency of oscillation, and so it is the mutual conductance, rather than the current amplification factor α , that determines the high-frequency performance of the oscillator. At high direct currents, the cut-off frequency of the mutual conductance may be considerably lower than f_α . However, if the current is so chosen that the internal emitter resistance becomes equal to the internal base resistance, the two cut-off frequencies become approximately equal.

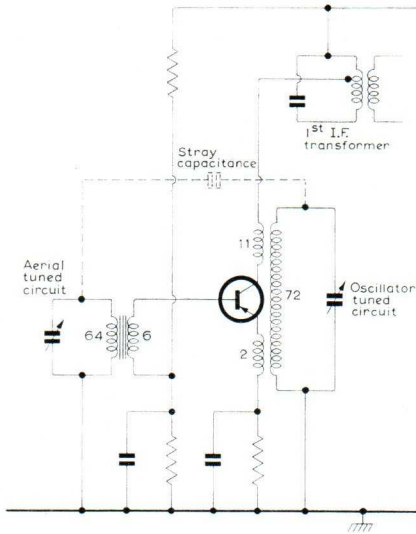


Fig. 3—Self-oscillating mixer with stray capacitance in tuning capacitor

A cut-off frequency of 15Mc/s ensures that the internal phase shift and fall in gain are small up to the maximum oscillator frequency of 2.07Mc/s. Furthermore, since the nominal effect of the drop in the h.f. performance of the transistor is small, the effect of spreads on the h.f. performance is also small.

Oscillation is maintained in this design at all frequencies under the most adverse conditions, that is, when the battery voltage has fallen by half, f_α is at the lower limit of its permitted tolerance range, and $r_{bb'}$ has its maximum permitted value. Also, the oscillator remains stable with a transistor which has the maximum permissible f_α and the minimum $r_{bb'}$. Squegging can only be made to occur by doubling the emitter bypass capacitance.

The design, therefore, will accept any OC44.

Tuning Capacitor

Correct tracking could be obtained with either a conventional paddler capacitor or a shaped oscillator section, as has been adopted in this design.

The value of tuning capacitance is not critical, but must be sufficient to provide the desired frequency coverage. The aerial section has a capacitance of 175pF, and the oscillator section a capacitance of 123pF.

If there is no screen between the two sections of the tuning capacitor, a stray capacitance exists between them. A self-oscillating mixer which includes stray capacitance is shown in Fig. 3. If the aerial is matched correctly to the input impedance of the OC44, the stray capacitance can cause spurious oscillations at the high-frequency end of the medium-wave band.

The circuit elements in Fig. 3 which control these unwanted oscillations are the aerial and oscillator tuned circuits and the stray capacitance. The circuit resembles that of a triode valve oscillator of the tuned-anode tuned-grid type. The unwanted oscillatory voltage appearing at the collector of the OC44 is stepped up in the oscillator transformer by a factor of 6.55 ($= 72/11$), and a corresponding feedback current flows through the stray capacitance. This feedback current is stepped up in the aerial transformer by a factor of 10.7 ($= 64/6$) to appear as a feedback current in the coupling coil. Consequently, feedback through the stray capacitance is approximately 70 times more important than feedback through the collector-base capacitance of the transistor. A stray capacitance of 1pF in the tuning capacitor alone will be equivalent to about 7 times the internal feedback capacitance of the OC44, which is about 10pF.

The effect of the feedback depends on the phasing of the oscillator and aerial transformers. If the phase of the feedback current is such that the amplitude of oscillation is increased, squegging may occur. Alternatively, if the phase is reversed, the amplitude of the wanted oscillation may be decreased. The unwanted feedback is a maximum when the receiver is tuned to its highest frequency, and the oscillator coil then loses control of the frequency of oscillation, which is then determined by the overall properties of the feedback path. The feedback is increased when the tuning capacitance has a low value, such as is common in miniature tuning capacitors, because of the higher transformer ratios.

The most practical way of reducing the undesired feedback in the circuit is to specify a tuning capacitor in which there is a screen between the oscillator and aerial sections, as in this design. Other forms of stray capacitance between the oscillator and aerial coils, for example, the

capacitance of the wiring of the wave-band selector switch, should be kept as low as possible.

Input Resistance of Mixer

The r.f. input resistance of the mixer transistor does not depend very much on whether the local oscillator is functioning. The change in input resistance does not exceed 10%, provided the steady emitter current is adjusted to be the same for both oscillating and non-oscillating conditions. It follows that the input resistance of an average OC44 may be calculated from values derived from its equivalent circuit.

The load impedance in the collector is low at radio frequencies. Thus it can be shown that the input resistance of a grounded-emitter transistor with its output short-circuited to a.c. is given approximately by

$$R_{in} = r_{bb'} + r_{b'e} \frac{r_{b'e} + r_{bb'}}{r_{b'e} + r_{bb'} + \omega^2 (c_{b'e} + c_{b'e})^2 (r_{b'e})^2 r_{bb'}}$$

The hybrid π equivalent circuit gives

$$r_{bb'} = 110\Omega,$$

$$g_{b'e} = 390\mu\text{mhos},$$

$$c_{b'e} = 410\text{pF},$$

and

$$c_{b'e} = 10.5\text{pF},$$

for an OC44 operating at an emitter current of 1mA and a collector-emitter voltage of -6V.

The values of $g_{b'e}$ and $c_{b'e}$ are directly proportional to the emitter current. Thus for operation at 0.32mA, which is the current when the receiver is tuned to 1Mc/s, $g_{b'e}$ is 125 μmhos and $c_{b'e}$ is 131pF. Hence $r_{b'e}$, which is the reciprocal of $g_{b'e}$, is 8k Ω . Substituting these values in the above equation gives an input impedance of 5k Ω at 1Mc/s.

At a frequency of 200kc/s, the input resistance of the mixer stage is approximately 10k Ω for an average transistor. This input resistance is controlled not only by the characteristics of the transistor as given by the above equation, but also by the bypass capacitance in the emitter. This capacitance has a greater effect at lower frequencies.

Output Resistance of Mixer

The output resistance at a frequency of 470kc/s varies widely and depends on the frequency to which the receiver is tuned. This variation results from feedback within the mixer. Without feedback, the output resistance would be about 35k Ω .

In practice, the output resistance reaches a minimum several times lower than 35k Ω in the long-wave band, and a maximum several times higher than 35k Ω in the medium-wave band. The mixer is therefore

coupled to the first i.f. transistor by an i.f. transformer similar to that used for coupling together the two i.f. stages. The output resistance with this arrangement is $28\text{k}\Omega$.

This method of coupling makes it possible to standardise the design of the first and second i.f. transformers. It also limits the load resistance in the mixer output to a suitably low value, so that feedback in the mixer is not serious.

Conversion Gain

At a signal frequency of 1Mc/s , the ratio between the r.f. power at the input of the mixer transistor and the r.f. power at the input of the i.f. transistor is approximately 27dB .

I.F. AMPLIFIER

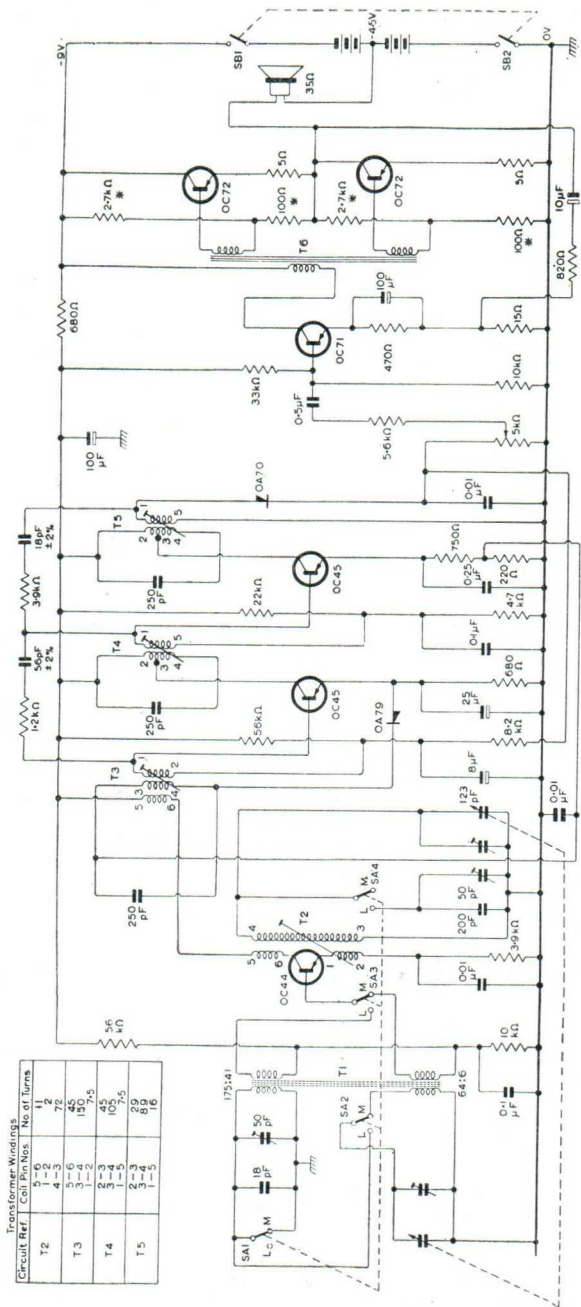
The basic design of the i.f. amplifier has been discussed in Chapter 18. Two OC45 are used in unilateralised grounded-emitter stages. The intrinsic power gain of an OC45 at 470kc/s is 38dB and the insertion loss of the second i.f. transformer 7dB . The gain of the first stage, without automatic gain control, is therefore 31dB .

The unloaded Q (that is, Q_0) of the second i.f. transformer is 100. The working Q (or Q_w) is 56, which gives a bandwidth of 8.4kc/s . These values of Q_0 and Q_w , giving an insertion loss of 7dB , ensure that a stability factor of four is maintained in a two-stage amplifier containing three similar coils. Thus the loop gain of either i.f. stage does not exceed a quarter of the gain needed for oscillation, even when the feedback capacitances of the transistor are at the extremes of the tolerance range.

The third i.f. transformer is designed to have an unloaded Q of 160. The load resistance on the secondary is arranged to maintain the stability factor of four, and also to give a bandwidth of 9kc/s and a stage gain of 34dB .

The identical first and second i.f. transformers are designed for the $28\text{k}\Omega$ output resistance in the preceding stage. However, the output resistance of the mixer stage may rise to many times this value, so that the stability factor of the first i.f. stage may be reduced undesirably from four to three. Although this reduction should not cause oscillation, even with the tolerances at their most unfavourable extremes, the bandwidth may be somewhat narrower. The required stability factor of four can be restored very simply, and at the expense of only about 2.3dB of gain, by connecting a 750Ω resistor across the secondary of the first i.f. transformer.

RADIO RECEIVERS



Tolerance of resistors marked with an asterisk (*) should be 5%. Tolerance of all other resistors should be 10%.

Fig. 4—Modified version of receiver with damping diode across first i.f. transformer

DETECTOR

The detector is an OA70 germanium diode in a conventional circuit. The steady voltage developed across the diode load resistance (which is also the volume control) is fed back to the first i.f. transistor to give automatic gain control.

The small forward bias applied to the diode by the a.g.c. circuit serves to keep the input resistance of the detector constant at all signal levels. This small bias gives constant loading on the i.f. amplifier, thus preserving the stability factor, and also improves the detection efficiency on weak signals.

AUTOMATIC GAIN CONTROL

Fig. 5 shows the performance of the simple system of a.g.c. incorporated in the circuit of Fig. 1. The range of control is 33dB, and the

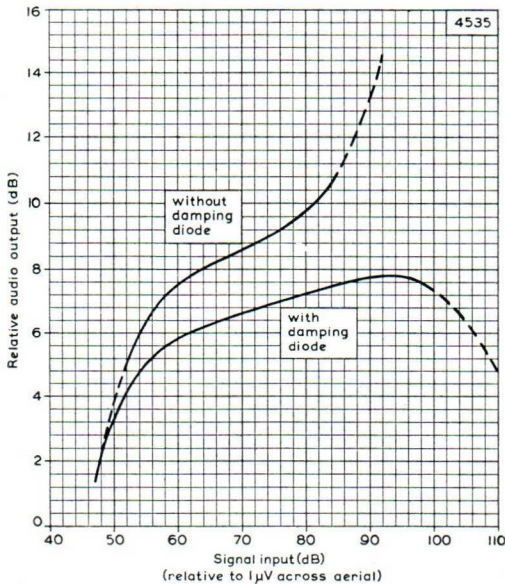


Fig. 5—Performance of a.g.c. arrangement with and without damping diode

corresponding variation in signal level at the detector is 6dB. (If the direct voltage across the diode load resistance is measured instead of the audio output, then a better figure for a.g.c. will be obtained.)

Fig. 4 shows a modification to the a.g.c. system used in Fig. 1 which increases the range of control to 52dB. The original feedback path is retained, but another diode (an OA79) is included to damp the first

RADIO RECEIVERS

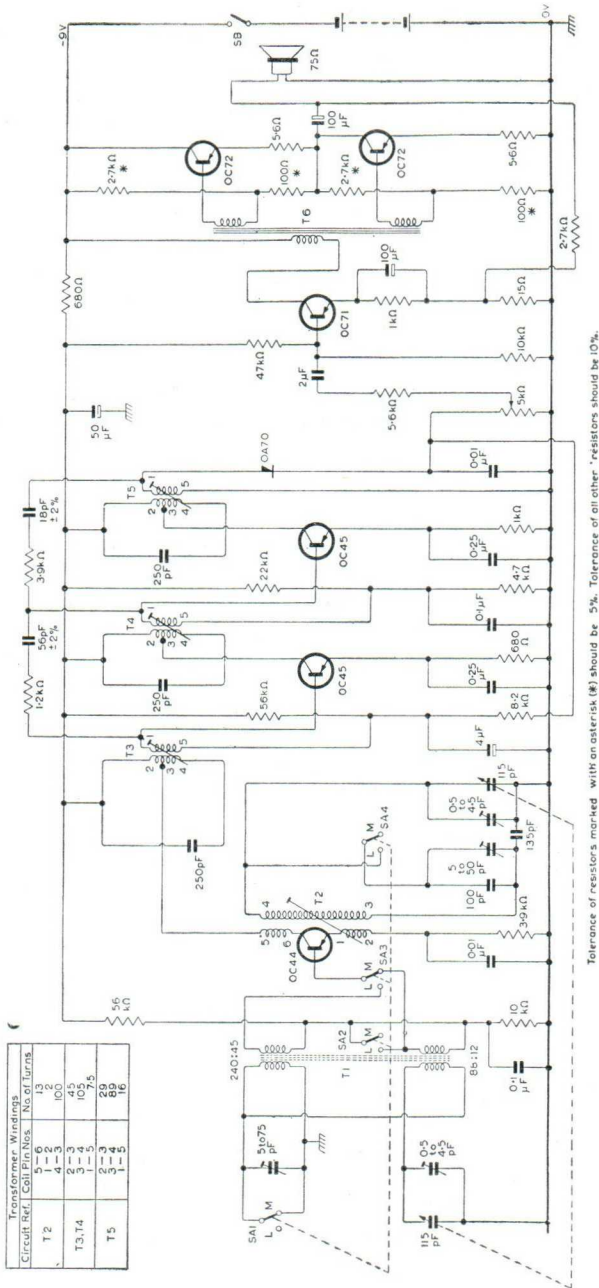


Fig. 6—Miniaturised version of receiver

i.f. transformer. This transformer is now constructed with a separate input winding, so that the tuned winding can be held at a steady potential of -0.22V by connecting it to a point in the emitter circuit of the second OC45.

When no signal is applied, the emitter of the first i.f. transistor has a nominal potential of 0.68V . Thus the reverse bias across the damping diode is 0.46V . The diode therefore has a high impedance and its effect on the gain of the receiver is negligible.

If an input signal of increasing amplitude is applied, the a.g.c. voltage on the base of the transistor causes the emitter voltage of the first i.f. transistor to fall until it becomes less than 0.22V . The damping diode is then biased in its forward direction, and its impedance decreases rapidly. The diode heavily loads the first i.f. transformer, thus widening the bandwidth and allowing a much larger input signal to be handled.

Fig. 5 also shows the variation in audio output with r.f. input voltage for the circuit using the modified a.g.c. system. The fall in output at very high signal levels results from clipping of the modulation envelope in the mixer.

For the modified form of a.g.c., the emitter bypass capacitance of the first i.f. transistor is increased to $25\mu\text{F}$. This capacitance prevents a.f. voltages, arising from unwanted detection, from being developed at the emitter. Such voltages would cause distortion of the modulation envelope at some signal levels.

AUDIO AMPLIFIER

The audio amplifier consists of a driver stage and a transformerless or 'single-ended' output stage. The driver stage incorporates an OC71 and the output stage a matched pair of OC72 (the 2-OC72). An inter-stage coupling (driver) transformer is required, but the loudspeaker forms the direct load for the output stage.

The audio amplifier is designed to have an input impedance of $9\text{k}\Omega$ at the volume control, so that a good a.c./d.c. load ratio is maintained for the diode detector. With nominal transistors, a drive voltage of 200mV is required for an output of 200mW . Negative feedback of 5dB is taken from the loudspeaker terminal to the emitter circuit of the OC71 driver.

Higher Output Powers

Audio amplifiers consisting of the LFH3 packet (an OC81D driver transistor and a matched pair of OC81 output transistors) may be incorporated if desired. These circuits were described at the end of Chapter 14 (Figs. 10 to 13). According to the form of circuit adopted,

output powers of up to 1W are obtainable. For all four circuits, the resistance in series with the connection from the volume control should be increased from $5.6\text{k}\Omega$ to about $8.2\text{k}\Omega$.

Modern commercial receivers normally provide a much higher output power than is obtainable from the 2-OC72.

MINIATURE RECEIVERS

The circuit of a semi-miniaturised version of the receiver is shown in Fig. 6. Basically this modified version and the original are very similar. However, the output power of the miniature version is reduced to 100mW (as compared with the performance given on p. 203) and smaller components are used. The current consumption is also reduced and is 7mA with zero signal and 13mA with average programme. The miniature receiver has been made with overall dimensions of $6\frac{1}{2} \times 4\frac{1}{4} \times 1\frac{1}{2}$ in. ($\approx 16.5 \times 11 \times 4\text{cm}$) and further reductions in these are possible.

The frequency range is modified to 550 to 1600kc/s in the medium-wave band and 150 to 250kc/s in the long-wave band. The lowest possible stray capacitance is essential in the wave-band selector switch in the oscillator section. To avoid switch capacitance in the aerial circuit, the selector switch has been transferred to the low end of the medium-wave coil, so that both coils are in series for long-wave reception.

In small personal receivers, the aerial is small and is almost always damped by adjacent metal parts. Thus two i.f. stages are still required if the receiver is to have sufficient sensitivity for use throughout Great Britain, for example.

Similarly, because the loudspeaker in a miniature receiver is small and relatively inefficient, no great reduction in output power is possible. The driver and class B stages are therefore still required, especially since battery economy is of major importance when battery size is reduced to a minimum. Nevertheless, it is advisable to reduce the output power to 100mW in order to limit vibration in the receiver case, and also to limit the battery consumption.

When low initial cost is more important than battery life, and a limited output is acceptable, a class A output stage could be used. Thus the number of transistors may be reduced to five. Another suggestion for reducing the number of transistors is to use a reflex arrangement in which the last i.f. transistor also acts as the first audio amplifier.

LARGER RECEIVERS

For enlarged receivers intended for permanent use in the home, the

RADIO RECEIVERS

existing design can easily be adapted, with very few modifications. A large Ferroxcube rod can be incorporated to give maximum sensitivity. Also a highly efficient loudspeaker can be used. With such a speaker, the volume obtained from the existing output of 200mW should be adequate for most listening conditions, while retaining a low battery consumption. The life of a moderately large battery could be expected to be about a year.

Higher output powers may be provided as just described where more frequent battery replacement is not a disadvantage.

PERFORMANCE OF ORIGINAL VERSION

Output power	200mW
Frequency range	
medium-wave	540 to 1640kc/s
long-wave	155 to 280kc/s
Frequency response	6dB below response level at 1kc/s at 70c/s and 3kc/s
Sensitivity 200 μ V signal (30% modulation at 400c/s) across aerial tuned circuit for audio output of 50mW	
Selectivity	
adjacent-channel rejection ratio at 9kc/s off-tune	29dB
second-channel rejection ratio	50dB
A.G.C. range	
change in input signal for 6dB change in audio output	
(a) without damping diode	33dB
(b) with damping diode.. .. .	52dB
Battery consumption	
zero signal	9mA
average programme	20mA

MAJOR COMPONENTS FOR ORIGINAL VERSION

Printed circuit: Weymouth Radio Manufacturing Co. Ltd. PCA1

Aerial

Mullard Ferroxcube rod FX1268

M.W. TRANSFORMER

Primary 64 turns of 19/0-0028 bunched conductors wound in single layer
($Q_0 = 200$ at 1Mc/s)

Secondary 6 turns of 19/0-0028 bunched conductors wound at low end of primary coil

L.W. TRANSFORMER

Primary 175 turns of 34 s.w.g. rayon-covered enamelled wire, wave wound
($Q_0 = 80$ at 200kc/s)

Secondary 41 turns (continuation of primary coil)

Complete aerial: Weymouth Radio Manufacturing Co. Ltd. RA2W/2

RADIO RECEIVERS

Tuning Capacitor

Aerial section	..	..	..	..	..	..	..	175pF max
Oscillator section	..	..	..	..	..	..	..	123pF max
Wingrove and Rogers. 196pF max (aerial section) and 110pF max (oscillator section). Types C74-02 with slow-motion drive and C74-22 with direct drive. Additional trimmers to order.								

Oscillator Coil (screened)

Inductance	..	..	..	..	..	..	..	173μH
Windings	..	..	..	..	..	..	..	Main = 72 turns
								Collector = 11 turns
								Emitter = 2 turns
Q ₀	..	..	..	..	130 at f _{osc} = 1Mc/s with tuning capacitance of 132pF			
Weymouth Radio Manufacturing Co. Ltd. P50/1AC								

I.F. Transformers

The leakage inductance in the coils in the i.f. transformers should be kept as low as possible. Enclosed pot cores are desirable.

FIRST* AND SECOND I.F. TRANSFORMERS

Tuning capacitance (referred to collector)	..	..	..	..	..	3000pF
Ratio of collector winding to secondary winding	..	..	..	..	..	6:1
Q ₀	..	..	..	..	..	100
Q _w (in circuit) nominal	..	..	..	..	..	56
Weymouth Radio Manufacturing Co. Ltd. P50/2CC						

*If a damping diode is used in the a.g.c. system, a separate collector winding on the first i.f. transformer is necessary. Otherwise, this transformer is identical with the second i.f. transformer. Weymouth Radio Manufacturing Co. Ltd. (type no. not available)

THIRD I.F. TRANSFORMER

Tuning capacitance (referred to collector)	..	..	..	..	..	4000pF
Ratio of collector winding to secondary winding	..	..	..	..	..	1.85:1
Q ₀	..	..	..	..	..	160
Q _w (in circuit)	..	..	..	..	..	52
Weymouth Radio Manufacturing Co. Ltd. P50/3CC						

Driver Transformer

Turns ratio	..	..	..	..	..	..	3.6:(1+1)
Primary inductance	..	..	..	..	..	..	5H at 2mA d.c.
Primary resistance	..	..	..	..	..	..	less than 200Ω
Secondary resistance	..	..	..	..	..	..	less than 60Ω per winding
Fortiphone A443							

Loudspeaker

Speech-coil impedance	..	..	..	..	..	..	35Ω
Elac 5in. 5D/211 ; 3in. P/319; 4 × 7in. 47D/108.							
TSL (type no. not available)							

RADIO RECEIVERS

Battery

Voltage (4.5+4.5)V
 Ever Ready Batrymax D

MAJOR COMPONENTS FOR MINIATURE VERSION

Aerial, oscillator coil, and aerial and i.f. transformers:

Weymouth Radio Manufacturing Co. Ltd. (type nos. not available)

Aerial

Mullard Ferroxcube rod FX1057

M.W. TRANSFORMER

Primary .. 88 turns of 19/0-0028 bunched conductors wound in single layer

Secondary 12 turns of 19/0-0028 bunched conductors wound at low end of primary,
 nearest centre of rod

L.W. TRANSFORMER

Primary .. 240 turns of 3/0-0024 bunched conductors wave-wound in three sections

Secondary 45 turns of 3/0-0024 bunched conductors wound as fourth-pie section,
 nearest centre of rod

The total length of the four l.w. windings (including three spaces of $\frac{3}{32}$ -in. each)
 is one inch (≈ 2.5 cm)

Tuning Capacitor

Aerial section 115pF swing

Oscillator section 115pF swing

Wingrove and Rogers C78-22 or C78-02 (with slow-motion drive). Both types
 to include an inter-section screen.

Oscillator Coil

Inductance 313 μ H

Windings Main = 100 turns

Collector = 13 turns

Emitter = 2 turns

Q_o 105 at 1Mc/s with tuning capacitance of 70pF

I.F. Transformers

As given for original version

Driver Transformer

Turns ratio 7:(1+1)

Primary inductance 5H at 1.5mA d.c.

Primary resistance less than 750 Ω

Secondary resistance less than 100 Ω per winding

Fortiphone L442

RADIO RECEIVERS

Loudspeaker

Speech-coil impedance		75Ω
TSL 2in. CMS50		

Battery

Voltage		9V
Ever Ready Batrymax PP4		

NOTE: The inclusion of a type number should not be taken to imply either that the list is exhaustive or that specific components are generally available.

CHAPTER TWENTY

4W 500kc/s TRANSMITTER

The transmitter described in this chapter is capable of delivering an output of 4W into a resistive load, and has been designed to operate at the international marine distress frequency of 500kc/s. This frequency is, of course, strictly reserved for the use of those in distress, but the design procedure is fully described and the circuit can easily be modified for neighbouring frequencies.

Modulation arrangements are described for manual keying and radio telephony.

Apart from the interest of the transmitter as a whole, it illustrates many of the applications described in other chapters, plus some new

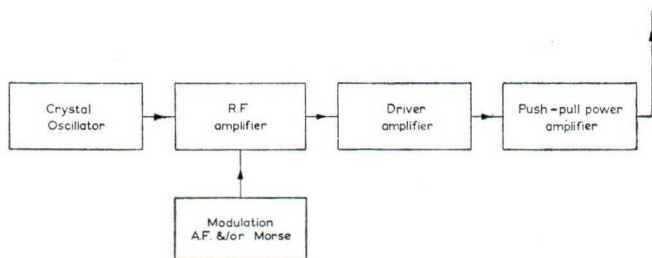


Fig. 1—Block schematic of transmitter

ones. The design includes a crystal-controlled oscillator, a silicon transistor used other than as a d.c. amplifier, and Ferroxcube pot cores.

GENERAL ARRANGEMENT OF TRANSMITTER

A block schematic of the 500kc/s transmitter is shown in Fig. 1. The transmitter comprises an oscillator, an r.f. amplifier, a driver and a power amplifier.

The oscillator is crystal controlled and uses an OC45. The resonant circuit in the collector lead is tuned to the series-resonant frequency of the crystal. The crystal is connected in series with the feedback winding to control the frequency of the oscillation.

The r.f. stage is an OC201 silicon transistor. The output from this stage is applied to a driver amplifier and then to a push-pull power

amplifier which feeds an aerial. Both the driver and the power amplifier use OC24 high-frequency power transistors.

The r.f., driver and power-amplifier stages are operated under class B conditions. Class C operation could be adopted, but the modulation would have to be applied to the output stage and, since a higher modulator power would be required, there would be little or no improvement in the efficiency of the transmitter as a whole. Class C operation of transistors is in any case not particularly attractive; hole-storage effects reduce the efficiency considerably below the theoretical value, and the

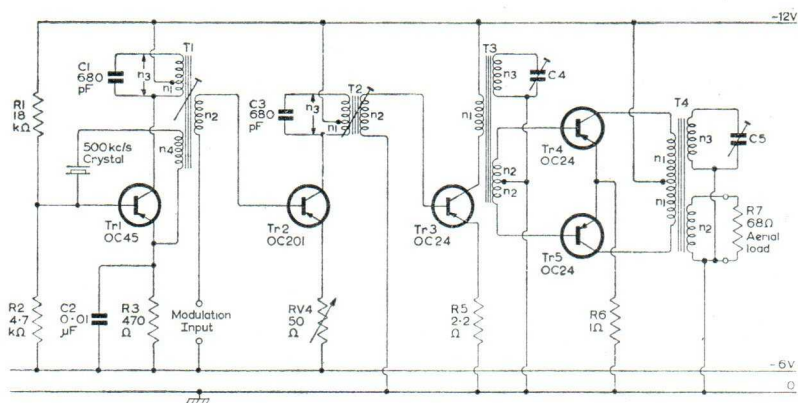


Fig. 2—4W transmitter for 500kc/s

variation in current gain with frequency and current give a further departure from the ideal. These effects would probably necessitate an empirical design.

The main circuit of the transmitter is given in Fig. 2.

PUSH-PULL AMPLIFIER

Collector Dissipation

The amplifier has been designed and tested for thermal stability at a maximum operating ambient temperature of 60°C (140°F) and at a maximum operating junction temperature of 75°C.

Suitable heat sinks for the two OC24 transistors in the push-pull amplifier can be made from $\frac{1}{16}$ -in. (≈ 16 s.w.g.) aluminium sheet. The surface should be blackened to assist cooling by radiation. Each transistor should be mounted centrally on its own heat sink of area 5×6 in. ($\approx 12.5 \times 15$ cm). The thermal resistance of the heat sinks should be about 4.4°C/W. The transistors should be mounted directly on the heat sinks, and the heat sinks insulated from the chassis.

In practice it may be more convenient to use the chassis as a common heat sink, the transistors being insulated from the chassis in the usual way. The mounting-base temperature of the transistors, for a maximum operating ambient temperature of 60°C, should then not exceed 72°C.

Collector Load

In a class B push-pull amplifier, the mean a.c. power P_{out} is given by

$$P_{out} = \frac{(V_{r.m.s.})^2}{R_{ce}}$$

where R_{ce} is the load resistance applied to each transistor in turn, and $V_{r.m.s.}$ is the r.m.s. value of the output voltage.

The design of the class B push-pull amplifier follows the same basic technique as that for an audio amplifier except that, since the collector circuit is tuned, no bias is required to eliminate crossover distortion. If a 12V battery is used to supply the collector circuit, and allowing 0.5V for bottoming, the output voltage at the collector is

$$\frac{12-0.5}{\sqrt{2}} = 8.1V_{r.m.s.}$$

The collector-emitter load resistance per transistor is

$$R_{ce} = (V_{r.m.s.})^2 / P_{out} = \frac{8.1^2}{4} = 16.4\Omega.$$

The use of a shared emitter resistor reduces the effect of variations in the base-emitter voltage V_{be} , and reduces the lengthening, caused by hole storage, of the collector-current pulse. The reduction of the hole-storage current gives a marked improvement in the efficiency of the amplifier. The resistance in the emitter lead must not be too high, because it reduces the gain of the stage. With an emitter resistor of 1Ω, the collector load becomes $R_c = 16.4 - 1 = 15.4\Omega$. The collector-collector load R_{c-c} is then $4R_c = 61.6\Omega$.

Output-transformer Design

To avoid loss in the tuned circuit, the working Q of the circuit should be as low as possible, though too low a Q results in greater harmonic content. A Q of 10 or 20 is recommended. In the design of the output transformer, using a Ferroxcube pot-core assembly LA6, a working Q (Q_w) of 15 is assumed with an unloaded Q (Q_o) of 230. Because $Q_o \gg Q_w$, the dynamic impedance of the circuit at resonance is effectively equal to R_c , the collector load. Thus

$$R_c = \omega L Q_w$$

and

$$L = R_c / \omega Q_w.$$

For $R_c = 15.4\Omega$ at 500kc/s,

$$L = \frac{15.4}{2\pi \times 0.5 \times 10^6 \times 15} = 0.327\mu\text{H}.$$

The number of turns required for $0.327\mu\text{H}$ using a Ferroxcube pot-core assembly LA6 is

$$n_1 = 92\sqrt{L} \quad [\text{mH}] = 92\sqrt{(0.327 \times 10^{-3})} = 1.66 \simeq 2 \text{ turns.}$$

Q_w cannot be increased because the winding would be less than 2 turns.

If $n_1 = 2$ turns, the new value for L is

$$L = \left(\frac{n_1}{92}\right)^2 \quad [\text{mH}] = \left(\frac{2}{92}\right)^2 = 0.472\mu\text{H}.$$

The capacitance to tune the collector to resonance at 500kc/s is

$$C = 1/\omega^2 L = \frac{1}{(2\pi \times 0.5 \times 10^6)^2 \times 0.472 \times 10^{-6}} = 0.215\mu\text{F}.$$

In practice, it is impossible to obtain a high value of Q_0 with only two turns; therefore a tertiary winding is used for tuning. If the number of turns across the transistor output is n_1 , and C_T is the tuning capacitance for the tertiary winding, then the number of turns n_3 required for the tuned circuit is

$$n_3 = n_1\sqrt{(C/C_T)}.$$

The highest practicable tuning capacitance is used to avoid high voltages across the circuit. If the tuning capacitance C_T is 820pF,

$$n_3 = 2\sqrt{\left(\frac{0.215 \times 10^{-6}}{820 \times 10^{-12}}\right)} \simeq 33 \text{ turns.}$$

To feed an aerial of impedance R_{ae} , the turns ratio n_1/n_2 between the collector and aerial is chosen to obtain a reflected load of 61.6Ω for R_{c-c} or $61.6\Omega/4$ for one transistor. Alternatively a turns ratio of 1:1 or 1:2, in conjunction with an impedance-matching network, may be used. This can take the form of either an L or a π section.

The π network shown in Fig. 3 is for matching the output impedance R of the transmitter to the aerial impedance R_{ae} . The values of X_A , X_B , and X_C can be found from

$$X_A = X_C = -\sqrt{(R \times R_{ae})}$$

and

$$X_B = \sqrt{(R \times R_{ae})}.$$

The values of L and C are given by

$$L = \frac{X_B}{2\pi f}$$

and

$$C = \frac{1}{2\pi f \cdot X_A}.$$

The advantage of such a network is that the reactance X_C can include the capacitive reactance of the aerial. The efficiency of the matching network is of the order of 90%.

A resistive load and a turns ratio of 1:2 may be used for checking the transmitter. If $n_1 = 2$ turns and $n_2 = 4$ turns, the preferred value of $68\Omega \pm 10\%$ may be taken for the resistive load of 61.5Ω .

DRIVER STAGE

The driver stage Tr3 is also operated in class B. This stage has to provide an output power of about 500mW, and operates at less than about 25% of the dissipation of each of the output transistors. The heat sink

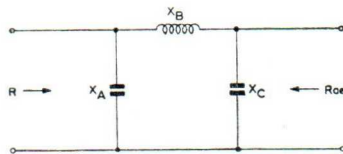


Fig. 3— π network for matching aerial

therefore needs to be only about 25% of the area of theirs, that is, about $2\frac{1}{2} \times 3$ in. ($\approx 6.5 \times 7.5$ cm). If the transistor is mounted on the chassis, with suitable insulation, the mounting-base temperature should not exceed 74°C , for a maximum operating ambient temperature of 60°C .

The input impedance Z_{in} of the output stage, at higher emitter currents, is very nearly $r_{bb'}$ (70Ω). Because the driver amplifier is working at a higher level of impedance than the output stage, Q_w may be increased. If $Q_w = 20$, the inductance required for the base winding is

$$L = Z_{in}/\omega \cdot Q_w = \frac{70}{2\pi \times 0.5 \times 10^6 \times 20} \approx 1.12\mu\text{H},$$

and the number of turns, using Ferroxcube pot-core assembly LA3, is

$$n_2 = 85\sqrt{L} \quad [\text{mH}] = 85\sqrt{(1.12 \times 10^{-3})} \approx 3 \text{ turns.}$$

The input power required to drive the push-pull amplifier using low-gain transistors was measured as 500mW. The input power is given by

$$P_{in} = \frac{(V_{in(r.m.s.)})^2}{Z_{in}}$$

so that

$$V_{in} = \sqrt{(P_{in}Z_{in})} = \sqrt{(0.5 \times 70)} = 5.9V_{r.m.s.}$$

If n_1 is the number of turns required at the collector of the driver amplifier,

$$n_1 = \frac{V_c}{V_{in}} \cdot n_2.$$

Allowing 0.5V for bottoming, V_c is equal to 8.1V r.m.s., and

$$n_1 = \frac{8.1 \times 3}{5.9} \simeq 4 \text{ turns.}$$

A slightly higher driving voltage is obtained at the input to the push-pull amplifier with the approximated winding of 4 turns. If required, the input power may be adjusted by means of the variable emitter resistor RV4 at the input to the driver amplifier. A tertiary winding is again used for tuning the collector current.

If 4 turns are used for n_1 , the new value of inductance is

$$L = \left(\frac{n_1}{85}\right)^2 [\text{mH}] = \left(\frac{4}{85}\right)^2 = 2.22 \mu\text{H}.$$

The tuning capacitance is:

$$C = 1/\omega^2 L = \frac{1}{(2\pi \times 0.5 \times 10^6)^2 \times 2.22 \times 10^{-6}} = 0.0457 \mu\text{F}.$$

If $C_T = 820 \text{ pF}$, the number of turns required for the tuned winding is given by

$$n_3 = n_1 \sqrt{(C/C_T)} = 4 \sqrt{\left(\frac{0.0457 \times 10^{-6}}{820 \times 10^{-12}}\right)} \simeq 30 \text{ turns.}$$

An emitter resistor of 2.2Ω is required to reduce hole-storage effects and increase the efficiency of the amplifier.

R.F. AMPLIFIER

The design procedure is similar to that in the previous section. The r.f. amplifier is also designed for class B operation. A 6V supply is used because of the voltage limitations of the oscillator transistor and because it is convenient for the modulator; it is also used because of the keying system adopted.

The input power required by the driver amplifier was 35mW at 2.3V r.m.s., measured with a low-gain transistor. The average input impedance is

$$Z_{in} = \frac{(V_{r.m.s.})^2}{P} = \frac{2.3^2}{35 \times 10^{-3}} = 151 \Omega.$$

As the stage is working at a higher level of impedance than the driver stage, the working Q can be further increased. A Q_w of 50 was chosen for good harmonic rejection.

The inductance required for the base winding of the r.f. amplifier is

$$L = Z_{in}/\omega \cdot Q_w = \frac{151}{2\pi \times 0.5 \times 10^6 \times 50} = 0.96 \mu\text{H}.$$

For this stage a Ferroxcube adjustable pot-core assembly LA35 is used. The number of turns required for the base winding is given by

$$n_2 = 135\sqrt{L} \quad [\text{mH}] = 135\sqrt{(0.96 \times 10^{-3})} \simeq 4 \text{ turns}.$$

The available voltage swing, allowing for the voltage across the transistor and emitter resistor, is approximately 3.25V r.m.s.

The number of turns n_1 required at the collector is found from

$$n_1 = n_2 \cdot V_c/V_{in} = \frac{4 \times 3.25}{2.3} \simeq 6 \text{ turns}.$$

The inductance of the collector winding is

$$L = \left(\frac{n_1}{135}\right)^2 [\text{mH}] = \left(\frac{6}{135}\right)^2 = 1.98 \mu\text{H},$$

and the tuning capacitance is

$$C = 1/\omega^2 L = \frac{1}{(2\pi \times 0.5 \times 10^6)^2 \times 1.98 \times 10^{-6}} = 0.0513 \mu\text{F}.$$

This high value of capacitance is rather inconvenient, so some extra turns are added to the collector winding and the tuning capacitance reduced to C_T . With C_T equal to 680pF, the total number of primary turns is

$$n_3 = n_1\sqrt{(C/C_T)} = 6\sqrt{\left(\frac{0.0513 \times 10^{-6}}{680 \times 10^{-12}}\right)} \simeq 52 \text{ turns}.$$

OSCILLATOR

A crystal-controlled oscillator with a tuned collector circuit is used. Feedback is obtained by a transformer winding which couples the parallel-tuned circuit in the collector to the base. The crystal is connected in series with the feedback winding, and the collector circuit is tuned to the series-resonant frequency of the crystal. The series resistance of the crystal should be sufficiently low for the feedback current to start oscillation. For the feedback to be of sufficient magnitude and the loop gain greater than unity, the turns ratio of the transformer should be 3:1.

The oscillator works under class A conditions. Base bias is provided by a potential divider R1-R2, in conjunction with an emitter resistor R_e , which is bypassed to r.f.

The oscillator is an OC45. The circuit is designed to drive the OC201 r.f. stage. This amplifier requires an input of 1.3V r.m.s. for a transistor having a minimum α' of 20.

The oscillator has a nominal quiescent collector current of 2mA. The available voltage swing, taking into account the voltage drop across the emitter resistor, is approximately equal to 3.4V r.m.s.

If n_2 , the base winding for the OC201, is 7 turns, the number of turns at the collector is

$$n_1 = n_2 \cdot V_c/V_b = \frac{7 \times 3.4}{1.3} \simeq 18 \text{ turns.}$$

Therefore 6 turns are required for the feedback winding if a 3:1 turns ratio is used.

With a Ferroxcube adjustable pot-core assembly LA35, the inductance of the collector winding is

$$L = \left(\frac{n_1}{135} \right)^2 [\text{mH}] = \left(\frac{18}{135} \right)^2 = 17.8 \mu\text{H.}$$

The capacitance required is

$$C = 1/\omega^2 L = \frac{1}{(2\pi \times 0.5 \times 10^6)^2 \times 17.8 \times 10^{-6}} = 5680 \text{ pF.}$$

As before, extra turns are added to the collector winding, so that a smaller tuning capacitance, C_T , of 680pF may be used.

The total number of collector turns now required is

$$n_3 = n_1 \sqrt{(C/C_T)} = 18 \sqrt{\left(\frac{5680 \times 10^{-12}}{680 \times 10^{-12}} \right)} \simeq 52 \text{ turns.}$$

TUNING PROCEDURE

The procedure adopted is the usual one of rough tuning at reduced power, followed by final adjustment at full power.

A resistor of 27Ω is connected in the emitter lead of the r.f. amplifier, Tr2, in order to reduce the available input to the driver amplifier. (RV4 may be set to the middle of its range for this purpose.) The 12V supply is then connected to the transmitter, and the oscillator tuned circuit is adjusted to the series-resonant frequency of the crystal, by adjusting the inductance of the pot core. The r.f. amplifier is tuned next, again by adjusting the pot core. The driver and the power-amplifier stages are then tuned for maximum output voltage across the 68Ω load, by adjusting the tuning capacitance. (Use a fixed capacitance of just below the calculated value and shunt this with a variable trimmer.) An output of about 2W should be obtained.

The emitter resistor of Tr2 should then be short-circuited. The complete transmitter should be readjusted, starting with the oscillator, followed by the r.f. amplifier, the driver stage, and finally the power

amplifier. The tuning of the power amplifier should be carried out carefully, as excessive collector current flows when the stage is off tune. The full output power of 4W should now be obtained.

RV4 is subsequently used to set the output power to the desired level.

PERFORMANCE

An experimental transmitter has been built using the circuit shown in Fig. 2. The performance has been tested over a range of ambient temperature from 0 to 60°C. The variation of output power, measured across a resistive load of 68Ω, is shown in Fig. 4. From the graph, it can be seen that the output is higher at lower temperatures, and falls

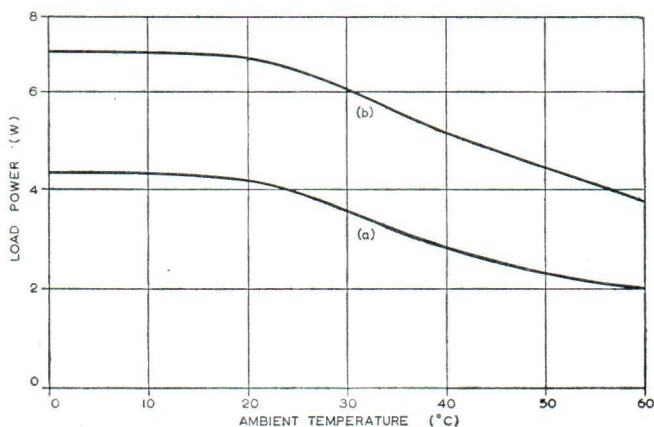


Fig. 4—Variation in output from transmitter with temperature

gradually at temperatures above 25°C. Quite reasonable output power is obtainable even at 60°C. The reduction of the output power arises from the fall in α cut-off frequency and the increase in hole-storage current with temperature. The latter can be seen from an expression for the hole-storage time-constant,

$$\tau_s = \frac{1/\omega_{1(N)} + 1/\omega_{1(I)}}{1 - \alpha_N \alpha_I},$$

(where N denotes normal and I inverse).

Because ω falls and α rises with temperature, there is a rise in the time-constant τ_s with temperature, which accounts for the increased hole-storage current.

Curve (a) in Fig. 4 is for low-limit transistors. With typical transistors, the output and driver stages may be overdriven at low temperatures, in order to obtain increased output at higher temperatures. Overdriving is achieved by selecting transistors with higher gain for the r.f. amplifier

and driver amplifier, or by adjusting the turns ratio. The variation in output power for high-gain transistors is shown in curve (b) in Fig. 4. The variation in frequency has been found to be less than one part in 10^6 per $^{\circ}\text{C}$ over the entire range of temperature.

Some field tests have been carried out at 520kc/s. The only modifications to the design meant using a 520kc/s crystal for the oscillator and retuning the transmitter to the new frequency. A transmitting aerial 30 ft. high ($\sim 9\text{m}$) and a receiving aerial of about 30 ft. of wire were used. Strong signals were received within 12 miles ($\sim 19\text{km}$) over land. This is not the maximum range over land, and a range of at least 50 miles ($\sim 80\text{km}$) may be expected at sea.

RADIO TELEPHONY

Low-level modulation is used for radio telephony. The r.f. amplifier Tr2 is biased into class C operation, using a parallel CR combination,

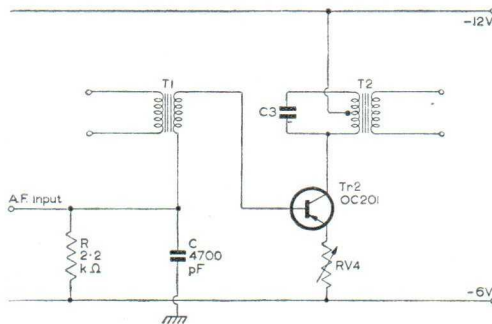


Fig. 5—Modulator for radio telephony

by the normal rectification of the emitter diode. The bias is adjusted to the point where the output of the transmitter falls to half its normal value, and an a.f. signal is applied across the bias resistor (Fig. 5).

RADIO TELEGRAPHY

Manual Keying

A circuit for on-off keying is given in Fig. 6. A morse key is connected in series with the base winding, and an r.f. choke and a bypass capacitor are added to prevent r.f. voltages from floating across the key.

The r.f. amplifier is thus d.c. keyed in its base circuit. This keying point is possible because, although with the base open circuit the leakage current assumes its full value of I'_{co} , the value of I'_{co} is very low for a silicon transistor, as used for this stage.

Automatic Keying

The morse key in Fig. 6 may be replaced by an automatic key which will give continuous unattended operation. The transistor multivibrator provides a simple switching element around which to design such a device.

Automatically-keyed transmitters are of interest to the whaling industry. A transmitter can be left aboard the harpooned whale, which is collected later when the kill is complete. The circuit is modified to

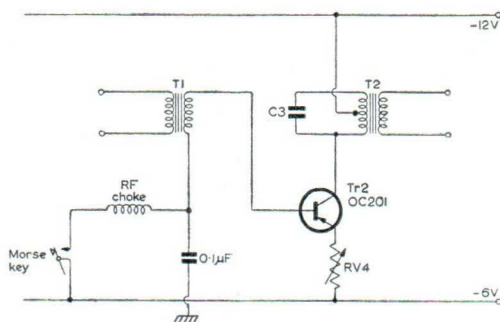


Fig. 6—Modulator for manual keying

operate at a fixed frequency other than 500kc/s, and is provided with an automatic-keying device which generates the simplest convenient signal. Such transmitters may be operated on continuous wave or modulated continuous wave.

Another application of automatic keying is to the generation of the S.O.S. distress signal. The signal to be generated consists of the group $\cdot \cdot \cdot - - - \cdot \cdot \cdot$, transmitted as a single signal with the dots sufficiently distinguished from the dashes. A minimum of four multivibrators and a gate would be needed to generate the signal. When intended for distress-signal transmissions, the transmitter may be designed for 500kc/s, but should be modified for modulated-continuous-wave operation, in order to comply with the international requirements.

Continuous Wave (C.W.) or Class A1 Emission

With the arrangement shown in Fig. 6, the transmission consists of the r.f. carrier interrupted by the on-off key. As there is no audio modulation, the result is continuous wave (c.w.) or class A1 emission.

Modulated Continuous Wave (M.C.W.) or Class A2 Emission

The modification for modulated-continuous-wave operation entails

providing an auxiliary audio oscillator to modulate the r.f. carrier. It is permissible to key either the audio modulation alone, or the r.f. carrier and the audio modulation together. One possible system for the latter method (Figs. 2 and 6) is to apply the keying signal in series with the audio modulation to the base winding of Tr2.

EMERGENCY EQUIPMENT

The transmitter is particularly suitable for lifeboats and life-rafts, but can be adapted for life-jackets, air-sea rescue equipment, helicopters, and radio beacons for coastal stations. The transmitter may also be of interest to climbing and scientific expeditions.

A 4 ampere-hour accumulator (the Exide 6MNA17, for example) will power the transmitter and an automatic-keying device for some 50 hours.

TRANSFORMER WINDINGS

Oscillator Transformer T1

Core:	Ferroxcube adjustable pot-core assembly LA35			
Wire:	London Electric Wire Co. and Smiths Ltd., 9/0024 bunched conductors			
Winding:	<i>Primary</i>	<i>Secondary</i>	<i>Tuned</i>	<i>Feedback</i>
	n_1	n_2	n_3	n_4
Turns:	18	7	52	6

Modulator Transformer T2

Core:	Ferroxcube adjustable pot-core assembly LA35		
Wire:	London Electric Wire Co. and Smiths Ltd., 9/0024 bunched conductors		
Winding:	<i>Primary</i>	<i>Secondary</i>	<i>Tuned</i>
	n_1	n_2	n_3
Turns:	6	4	52

Driver Transformer T3

Core:	Ferroxcube pot-core assembly LA3		
Wire:	Connollys Ltd., 3 strands of 19/0028 bunched conductors (n_1 , n_2) and 19/0028 bunched conductors (n_3)		
Winding:	<i>Primary</i>	<i>Secondary</i>	<i>Tuned</i>
	n_1	n_2	n_3
Turns:	4	3+3	30

Output Transformer T4

Core:	Ferroxcube pot-core assembly LA6		
Wire:	21 s.w.g. enamelled copper (n_1 , n_2) and Connollys Ltd., 19/0028 bunched conductors (n_3)		
Winding:	<i>Primary</i>	<i>Secondary</i>	<i>Tuned</i>
	n_1	n_2	n_3
Turns:	2+2	4	33

CHAPTER TWENTY-ONE

TUNED OSCILLATORS

A sinusoidal oscillator may be regarded as a tuned amplifier, with part of the output signal fed back to the input such that the output signal is maintained. Oscillation will occur if both the following conditions are satisfied: (a) the phase of the feedback signal is such that, after passing through the amplifier, the amplified feedback is exactly in phase with the output from which it is considered to be derived; and (b) the new output is equal to or greater than the original output. These two conditions mean that the loop gain of the circuit must be real and equal to or greater than one.

For oscillations to build up from zero, practical oscillators have a loop gain greater than one. On switching on the oscillator, the oscillations continue to build up until the effective gain is somehow reduced so that the loop gain becomes one.

The reduction in gain generally occurs in one of two ways: either the amplifier bottoms, thereby reducing the load resistance; or else part of the sinusoidal signal is converted to a bias, which is arranged to cut off the amplifier for part of each cycle, and thereby reduce the effective transfer conductance.

Bottomed oscillators have been rather out of favour in the past. The minimum gain of a modern transistor such as the OC72 or OC84, however, is sufficient for these oscillators to be designed to work with all transistors of the type in question, and at fairly low distortion even when a maximum-gain transistor is inserted in the circuit.

The bottomed and unbottomed oscillators described in this chapter are designed in a similar manner to small-signal transistor amplifiers. The oscillator performance can be predicted from the mean values of collector/emitter current, α' and input capacity. In particular, the output voltage can readily be calculated.

OUTPUT VOLTAGE

Fig. 1 shows the bare essentials of the output circuit of an oscillator. The transistor collector current, which generally consists of pulses, is set to a mean value I_{mean} by some suitable bias circuit. (For the purpose of this chapter, the mean emitter current can be taken to be equal

to the mean collector current. These will be referred to indifferently as I_{mean} .) Usually the bias arrangement takes the form of either a resistor connected between the base of the transistor and the negative side of the supply, or a base potential divider, which defines the voltage at the base and, in conjunction with a bypassed emitter resistor, defines the mean emitter current.

The peak output voltage of the oscillator, $v_{\text{out(pk)}}$, is related to the mean current by the feedback applied, and will have a maximum value of

$$2 I_{\text{mean}} |Z_L|$$

where $|Z_L|$ is the magnitude of the load impedance at the frequency of oscillation. In practice $v_{\text{out(pk)}}$ will never be more than 30% less than

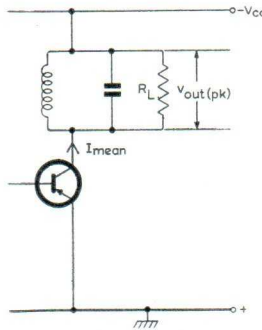


Fig. 1—Output circuit of oscillator

this value in any acceptable design. If, as is generally the case, the oscillator is working at the resonant frequency of the tuned circuit, the load impedance is simply R_L , and the peak output voltage will be approximately (within 30%) equal to $2 \times I_{\text{mean}} R_L$.

The value of R_L can be determined for the desired output power once the maximum permissible value of $v_{\text{out(pk)}}$ has been decided. From first principles

$$P_{\text{out}} = \left(\frac{v_{\text{out(pk)}}}{\sqrt{2}} \right)^2 \cdot \frac{1}{R_L}$$

$$R_L = \frac{(v_{\text{out(pk)}})^2}{2P_{\text{out}}}$$

If the oscillator is allowed to bottom, and assuming no emitter resistor, $v_{\text{out(pk)}}$ will be approximately equal to V_{cc} .

LOOP GAIN

The above approximation, that

$$V_{out(pk)} = 2I_{mean} R_L,$$

becomes more accurate as the collector-base feedback (and hence the loop gain) is increased, although the feedback cannot be increased indefinitely because of squegging. Again the d.c. bias circuit does not

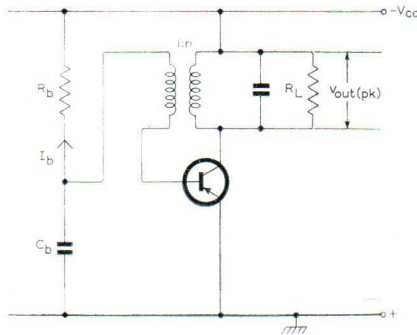


Fig. 2—Simplest bias circuit

affect the general principles of operation, and there are several ways of applying the bias. Fig. 2 shows the simplest bias circuit with a base resistor R_b .

An unbypassed resistor R_f may be included in the emitter to introduce a.c. feedback, but it should be remembered that in some circuits, $R_f = 0$.

The magnitude of the collector-base feedback can be assessed from the loop gain A_o , which is calculated on a small-signal basis using mean values of the transistor parameters. A_o is defined as the product of the forward and reverse voltage-transfer ratios.

The forward voltage transfer is $g_m R_L$, where g_m is the transfer (or mutual) conductance. The reverse voltage transfer is simply the transformer ratio $1/n$; the loading effect of the transistor input on the secondary is small and can usually be neglected.

Thus the loop gain is

$$A_o = g_m R_L \cdot \frac{1}{n} \quad \dots(1)$$

If the internal base resistance is ignored, then substituting

$$g_m = \frac{\alpha_o}{r_e} \simeq \frac{1}{r_e}$$

gives

$$A_o = \frac{R_L}{(r_e + R_f)} \times \frac{1}{n}, \quad \dots(1a)$$

where R_f is the external unbypassed resistance, if any, in the emitter circuit.

The peak output voltage, which has been given as $2I_{\text{mean}} R_L$, is strictly speaking equal to $R_L \times i_{\text{pk(fund)}}$, where $i_{\text{pk(fund)}}$ is the peak

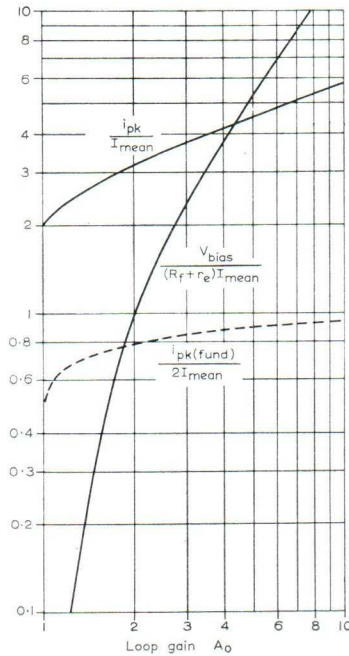


Fig. 3—Design curves for R_f greater than zero

collector current at the fundamental frequency of the oscillation contained in the collector-current pulses. The value of $i_{\text{pk(fund)}}$ can be found from Fig. 3 for circuits which include a finite feedback resistance R_f , or from Fig. 4 for circuits in which $R_f = 0$.

For instance, if the loop gain A_o is 3, then from Fig. 4 (for $R_f = 0$),

$$\frac{i_{\text{pk(fund)}}}{2I_{\text{mean}}} = 0.9,$$

instead of 1.0, as required by the approximation made above. For a mean current of 1.5mA, the peak collector current at the fundamental

frequency is

$$2 \times 1.5 \times 0.9 = 2.7 \text{ mA}$$

instead of approximately 3 mA, and if the load resistance R_L is $4 \text{ k}\Omega$, the peak output voltage will be 10.8 V , instead of 12 V .

Only the fundamental component of the peak collector current need

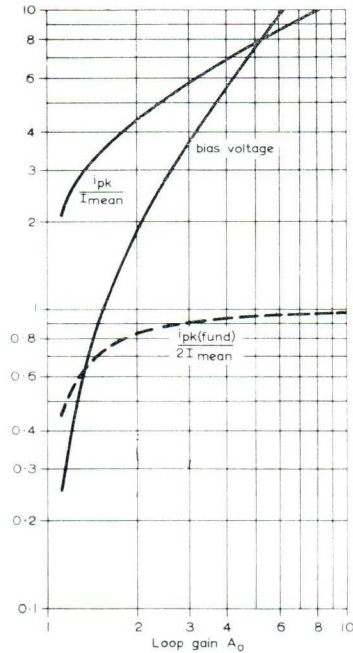


Fig. 4—Design curves for R_f equal to zero. Bias voltage in units of kT/q , where $kT/q \simeq 25 \text{ mV}$ at $T_{amb} = 298^\circ \text{K}$ (25°C).

be considered, because the second and higher harmonics are bypassed by the capacitor in the tuned circuit.

EFFECT OF INTERNAL BASE RESISTANCE

The difficulty presented by the internal base resistance $r_{bb'}$ is that, unless this is only a small fraction of the input impedance of the transistor, an appreciable amount of the feedback will be lost. Since $r_{bb'}$, and hence the loss, varies from transistor to transistor, so will the output from the oscillator.

Low Frequencies

At low frequencies, the loss arises because α'_{oe} is not infinite, and the

input voltage is tapped down in the ratio

$$\frac{v_{in}}{e_s} = \frac{\alpha'_o(r_e + R_f)}{r_{bb'} + R_f + \alpha'_o(r_e + R_f)}, \quad \dots (2)$$

as is represented by the equivalent circuit of Fig. 5. This attenuation factor can be reduced by operating with small mean currents and therefore large r_e , or else it can be controlled to some extent by means of the feedback resistance R_f . The value of R_f is typically made such that

$$\alpha'_o(r_e + R_f) = 3(r_{bb'} + R_f).$$

R_f raises the input impedance of the stage more than it increases the resistance effectively in series with the base.

The attenuation factor can be taken into account when calculating the reverse voltage transfer and the loop gain A_o . The forward voltage

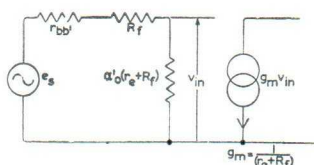


Fig. 5—Equivalent circuit of oscillator for low frequencies

transfer also will be affected by R_f , since as R_f is increased, g_m falls.

High Frequencies

At high frequencies the problem is a little more complicated, but may be treated in a similar manner. The shunting of the input by the emitter capacitance of the transistor, c_e , is shown in Fig. 6. The input impedance of $\alpha'_o(r_e + R_f)$ in parallel with the reactance of c_e must now be made three or more times greater than $(r_{bb'} + R_f)$. This condition may be satisfied by a suitable choice of R_f . However, since

$$r_e[\Omega] \simeq \frac{25}{I_e[\text{mA}]},$$

r_e will have a sufficiently high value at low mean currents for R_f to be zero. The exact calculation of the loop gain is more difficult at high frequencies because of the phase shift introduced by the input capacity.

The situation is simplified if the operating frequency is sufficiently high for only the reactance of c_e to be considered. The input capacity c_e causes a phase shift between the input voltage v_{in} and the voltage returned by the transformer, e_s . Because an oscillator always works with a real loop gain (zero phase shift), there will be an equal and opposite phase shift in the collector tuned circuit, and the load, equal

to $|Z_L|$, will be somewhat less than R_L . It happens that the same proportion of voltage is lost at the output as at the input. The reduced value of the loop gain A_0 can therefore be calculated using the attenuation factor

$$\frac{(X_{ce})^2}{(r_{bb'} + R_f)^2 + (X_{ce})^2}$$

From Fig. 3 or Fig. 4, the value of the peak collector current at the fundamental frequency can be found which corresponds to the reduced A_0 .

The poorest transistor is the one having the largest input capacity, that is, the lowest ω_1 , since

$$c_e = \frac{1}{\omega_1(r_e + R_f)}$$

The output voltage with this transistor is less than for a nominal

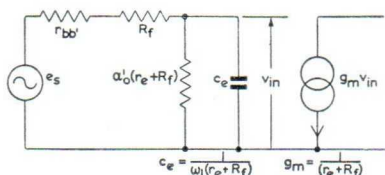


Fig. 6—Equivalent circuit of oscillator for high frequencies

transistor, and will have a value given by

$$\text{reduced } v_{\text{out(pk)}} = \text{reduced } R_L \times \text{reduced } i_{\text{pk(fund)}} ,$$

the reduced R_L being found from

$$\frac{R_L \cdot X_{ce}}{\sqrt{\{(r_{bb'} + R_f)^2 + (X_{ce})^2\}}}$$

Maximum Value of R_f

It can be seen that the value of R_f is decided by the internal base resistance $r_{bb'}$, by α' , and by the frequency of operation, but is independent of the required output power.

At the higher power levels, the load resistance becomes comparable with R_f , so that the maximum output power is limited, unless the supply voltage is increased. This limit can easily be illustrated by referring to the graphs. For instance, at a loop gain of 2, the peak collector/emitter current is about 3 times the mean (Fig. 3), and so when R_f is equal to one-third of the load, the same voltage, though of different waveform, is developed across both R_f and R_L . The circuit is then not capable of more than 50% efficiency.

BIAS CIRCUITS

Base-resistor Bias

With the simple bias arrangement shown in Fig. 7, the mean current is approximately equal to the mean $\bar{\alpha}'$ times the mean base current, that is, $I_{\text{mean}} \simeq \bar{\alpha}' \times I_{b(\text{mean})}$. Since $I_{b(\text{mean})} R_b \simeq V_{cc}$,

$$R_b \simeq \frac{\bar{\alpha}' V_{cc}}{I_{\text{mean}}},$$

where the leakage current is ignored. From the known value of $\bar{\alpha}'$, the mean current can be set to the desired value by choosing a suitable

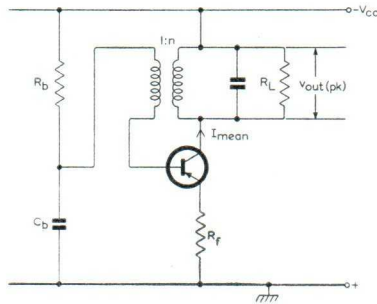


Fig. 7—Schematic oscillator with base-resistor bias

value for R_b . This arrangement is not very satisfactory as it stands, because $\bar{\alpha}'$ varies from transistor to transistor and so introduces a spread into the mean current for any fixed value of R_b .

However, the circuit works very well if the value of R_b is made equal to that which will just cause the lowest- $\bar{\alpha}'$ transistor to bottom ($V_{\text{out(pk)}} = V_{cc}$). Higher- $\bar{\alpha}'$ transistors are then more heavily bottomed, so that their mean $\bar{\alpha}'$ is almost reduced to that for which the circuit was designed, and there is little change in the mean current. For the same reason, the circuit is not affected by the increase in I_{co} with temperature, except at very low currents.

Circuits can be designed on this basis to give distortions of only 1% with the most heavily bottomed transistor.

The mean base current when the oscillator is working is higher than the direct current in the quiescent state, on account of the bias voltage developed across the decoupling capacitor C_b . This voltage is in a direction such that it tends to increase the mean base current, so that for complete accuracy the bias voltage should really be added to V_{cc} when calculating R_b . Also, the mean base-emitter voltage

$V_{be} + I_{\text{mean}} R_f$ should be deducted from V_{cc} . The design equation therefore becomes

$$R_b = \frac{(V_{cc} + V_{\text{bias}} - V_{be} - I_{\text{mean}} R_f) \times \bar{\alpha}'_{\text{min}}}{I_{\text{mean}}}$$

The bias voltage can be found by means of Fig. 3.

Bias by Potential Divider and Emitter Resistor

In oscillators which are to be modulated (for instance, self-oscillating

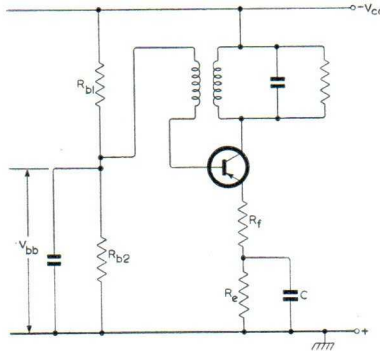


Fig. 8—Schematic oscillator for potential-divider and emitter-resistor bias

mixers in radio receivers), control of the amplitude by bottoming is usually objectionable.

Also, the bottomed oscillator is not capable of giving very low distortion, and may be unsuitable for some applications.

For such circuits, the mean current has to be well defined at a value less than that which causes bottoming. A suitable bias arrangement is the conventional potential-divider and emitter-resistor circuit shown in Fig. 8. The potential divider should normally pass a current which is about ten times the expected base current of the transistor.

For transistor mean currents up to about 5mA, there is little difference between the mean current and the direct or quiescent current when not oscillating, so that the values of R_{b1} , R_{b2} and R_e can be determined as for a small-signal amplifier. Also, the effect of R_f on the bias network can be neglected. Thus up to 5mA,

$$I_{\text{mean}} \simeq \frac{V_{bb} - V_{be}}{R_e}$$

Above 5mA, the oscillatory current is somewhat greater, and if complete accuracy is required, the bias voltage (Fig. 3) must be added to V_{bb} before calculating I_{mean} .

Hence above 5mA,

$$I_{\text{mean}} = \frac{V_{\text{bb}} - V_{\text{be}} + V_{\text{bias}}}{R_f + R_e}.$$

Furthermore, in the circuit of Fig. 8, if too small a value is chosen for V_{bb} , the change in the bias voltage with the variation in loop gain from transistor to transistor will cause a large change in I_{mean} and therefore in output voltage.

V_{bb} may be obtained from a tapped battery in the usual way, if desired.

TEMPERATURE STABILITY

The temperature stability of the mean current of this circuit is similar to that with a small-signal amplifier. For instance, the change in V_{be} ($\sim -2\text{mV}/^\circ\text{C}$) has the same effect upon the mean current as the same change in V_{bb} . A 10% change in V_{bb} , for example, will cause a 10% change in mean current. V_{bb} can be chosen sufficiently high to keep the variation in mean current within the limit required for any particular application.

DISTORTION

Since the above types of oscillator supply current pulses to the tuned circuit, the oscillators usually work in class C, and the distortion is greater than for class A operation.

The distortion can be limited as desired by a suitable choice of the working Q (Q_w), as the following example illustrates. For a loop gain A_o of 3 or 4, the second-harmonic current is typically equal to 0.6 times the fundamental. Then the collector voltage will contain 1% second-harmonic distortion when

$$0.6 \times \left(\frac{2}{3Q_w} \right) = \frac{1}{100}$$

whence

$$Q_w = 40.$$

Also, the third-harmonic current is typically 0.25 times the fundamental. The collector voltage will contain 1% third-harmonic distortion when

$$0.25 \times \left(\frac{3}{8Q_w} \right) = \frac{1}{100},$$

that is, when

$$Q_w \simeq 9.$$

The factors $2/3Q_w$ and $3/8Q_w$ are obtained from considerations of the impedance of the tuned circuit to the respective harmonics and therefore of the voltage that will be developed.

SQUEGGING

The value of C in the circuit of Fig. 8 has to be chosen carefully in order to avoid squegging. As a general rule, the reactance of C should be about $\frac{1}{10}$ to $\frac{1}{4}$ of the value of $(r_e + R_f)$. The situation is aggravated by large amounts of feedback, and A_o should be restricted to a maximum of about 4. The position is worst for low-power stages, where the high value of R_e gives a large value to the time-constant $C \times R_e$.

Squegging is much less of a problem with the circuit of Fig. 7, when this has an amount of bottoming, but it is good practice to use a capacitance which only just gives the necessary decoupling. A useful guide is

$$X_{Cb} \sim \frac{\bar{\alpha}'_{\min}(r_e + R_f)}{20}.$$

FREQUENCY STABILITY

Each circuit should be designed to cope with the variations in temperature and supply voltage present in the particular design.

The inductance of the coil used in the tuned circuit is a function of temperature and, for a magnetic core, it is also a function of the peak output voltage. The Ferroxcube pot core for which these effects are least troublesome is the LA6.

For the LA6,

$$\delta L \sim +6 \text{ parts per million}$$

for every 1% increase in peak output voltage. Also

$$\delta L < +60 \text{ parts per million}$$

for every 1°C rise in ambient temperature.

For an air-cored coil, a variation of

$$\delta L \sim +25 \text{ parts per million}$$

per °C rise in ambient temperature is typical.

In the transistor itself, the principal frequency drift is caused by the phase shift introduced by the internal base resistance. The phase shift ϕ of v_{in} referred to the feedback voltage e_s is

$$\phi = \tan^{-1} (r_{bb}'/X_{ce}) \quad \text{if} \quad X_{ce} > r_{bb}'.$$

Therefore

$$\phi = \tan^{-1} \frac{r_{bb}' f_r}{r_e f_1},$$

where f_r is the resonant frequency of the tuned circuit, and is approximately equal to the operating frequency, and f_1 is the high-frequency parameter. Provided ϕ is less than 0.5 radian,

$$\phi \simeq \frac{r_{bb'} f_r}{r_e f_1}.$$

The phase shift between voltage and current in the tuned circuit is given by

$$\phi' = 2Q_w \Delta$$

where

$$\Delta = \frac{f_r - f}{f_r}$$

and f is the operating frequency.

The phase shift ϕ is balanced out by ϕ' , so that $\phi' = -\phi$, whence

$$\Delta = -\frac{r_{bb'} f_r}{2Q_w r_e f_1}.$$

As an example, suppose $r_{bb'} = 100\Omega$, $f_r = 1000\text{c/s}$, $f_1 = 500\text{kc/s}$, $Q_w = 10$, and $r_e = 25\Omega$. Then $\Delta = -400 \times 10^{-6}$. For a 10% decrease in r_e , caused by an increase in collector current with temperature, Δ decreases by 40 parts per million.

20mW, 10kc/s OSCILLATOR

As an illustration of the design procedure, a requirement is postulated for a 10kc/s grounded-emitter oscillator. The supply voltage is to be 12V and the peak output power about 20mW (that is, 10mW mean) at 20°C, and this output must not increase by more than 10% if the ambient temperature rises to 45°C. Further, the oscillator must not bottom.

D.C. Stabilisation

First the stage is stabilised to have a quiescent current which does not increase by more than 10%. The stability calculations are essentially the same as those for a class A amplifier stage with a mean current I_{mean} .

Briefly, the 25°C increase in temperature causes a 50mV decrease in the base-emitter voltage V_{be} for the same quiescent current, so that the base supply voltage V_{bb} must be at least 0.5V, if the quiescent current is not to increase by more than 10%. V_{bb} may be obtained from a potential divider, provided the source resistance is sufficiently low for (a) I_{co} not to cause trouble, and for (b) the mean base current (which is dependent upon α'_o —a quantity which differs from transistor to transistor) not to introduce any appreciable change in V_{bb} . The second condition is satisfied if the potential divider has a standing current of about ten or more times the base current.

As high efficiency is not a requirement, a base supply voltage V_{bb} of 1.5V may be used. The temperature dependence of the oscillator bias

may be ignored, because it is insignificant compared with the change in V_{be} .

Load Resistance

If the collector voltage is allowed to swing within 2.5V of the emitter voltage, the peak oscillatory voltage will be

$$12 - 1.5 - 2.5 = 8V,$$

where the voltage dropped across the emitter resistor is approximately equal to V_{bb} ($= 1.5V$). The amplitude will not be sufficient to bottom the transistor when it increases by 10% at the higher temperature. The collector load required to dissipate a peak power of 20mW is

$$R_L = \frac{(V_{out(pk)})^2}{P_{out}} = \frac{8^2}{20 \times 10^{-3}} = 3.2k\Omega.$$

Mean Stage Current

Because the peak collector current at the fundamental frequency is approximately twice the mean current,

$$I_{mean} = \frac{V_{out(pk)}}{2 \times R_L} = \frac{8}{2 \times 3.2} = 1.25mA.$$

Emitter Resistor and Base Potential Divider

The value of R_e can now be calculated from

$$R_e = \frac{V_{bb} - V_{be} + V_{bias}}{I_{mean}}.$$

V_{bias} can be obtained from Fig. 4, and is 140mV for a loop gain of, say, 4. Thus

$$R_e = \frac{1.5 - 0.15 + 0.14}{1.25} = 1.19k\Omega \simeq 1.2k\Omega,$$

where 0.15V is a typical figure for the bias voltage of a low-power germanium transistor operating at an emitter current of 1.25mA, and 1.2k Ω is taken as the nearest standard resistance value.

The resistances in the potential divider are calculated in the usual way.

Transformer Ratio

The loop gain (or more precisely the initial loop gain) which has been chosen to be 4, is given by:

$$A_o = g_{mo} \times \frac{R_L}{n},$$

where g_{mo} is the value of g_m immediately before the onset of oscillation, and is the small-signal g_m with the direct current of the transistor equal to the mean current of the oscillator.

The value of g_{mo} in A/V is equal to $I_{mean}/25$ if I_{mean} is in mA, so that

$$A_o = 4 = \frac{1.25}{25} \times \frac{3200}{n},$$

from which $n = 40$. The source resistance looking into the secondary of the transformer is

$$R_L/n^2 = 3200/40^2 = 2\Omega.$$

This value justifies the assumption that the transistor may be considered as being driven from a voltage source.

Complete Circuit

The complete circuit is shown in Fig. 9. A working Q of 10 is used (the unloaded Q being very much greater than 10), so that the collector load can be specified as nearly as possible by an ordinary resistor.

If the output voltage is calculated more exactly, it is found that, with an A_o of 4, the approximation

$$\frac{i_{pk(fund)}}{2I_{mean}} = 1$$

gives an output voltage which is 7% too high (Fig. 4), so that the expected peak output voltage is changed from 8 to 7.44V. In practice,

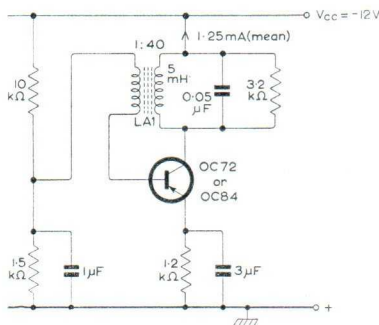


Fig. 9—10kc/s oscillator for output of 20mW

the mean current would be increased by 7%, but this adjustment has not been made to the circuit of Fig. 9.

A loop gain greater than 4 should not be used, because complete decoupling of the emitter is then not always possible. The partially unbypassed emitter resistance causes degeneration; this reduces the effective loop gain below the calculated value, and the output voltage may be more difficult to calculate because of the phase shift. Because of the high degree of feedback, the circuit will oscillate with a supply voltage of down to about 3V.

SQUEGGING

The $3\mu\text{F}$ capacitor decoupling the emitter has a reactance of 5Ω at 10kc/s and represents the largest practical value that can be used without squegging, which starts at $4\mu\text{F}$. This capacitance needs to be chosen carefully if squegging is to be avoided, but in the design described, in which the loop gain does not change appreciably from transistor to transistor, the problem is not a difficult one. This component should be chosen with care; if necessary three $1\mu\text{F}$ paper capacitors should be used in parallel. The resulting capacitance at the maximum of the tolerance range should not exceed $4\mu\text{F}$.

90mW, 50kc/s OSCILLATOR

The design procedure will not be given for this oscillator (Fig. 10), as the requirements for this circuit were deliberately made as difficult as

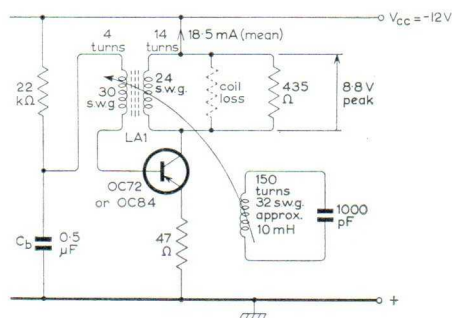


Fig. 10—50kc/s oscillator for output of 90mW

possible whereas, in general, the design problems would be much easier. The circuit has been designed to work with the 'worst' transistor.

The oscillator operates at 50kc/s from a 12V supply, and provides an output power of about 90mW . The circuit is arranged so that all transistors will be bottomed. When fifty OC72 were tried in the circuit, the output voltage showed a spread of only 2%.

Care should be taken with the base-resistor-bias circuit—such as that shown in Fig. 10—to ensure that the oscillator output is never short-circuited and that the polarity of the feedback winding is never reversed. Otherwise α' is fully effective with no oscillation, and the collector current might be destructively high with high- α' transistors.

Because of the stabilising action of the 47Ω unbypassed emitter resistor, the circuit will oscillate with a supply voltage of down to about 4V .

42kc/s OSCILLATOR FOR TAPE RECORDERS

The oscillator shown in Fig. 11 has been included to help those who are experimenting with transistor circuits for portable tape recorders.

A single-transistor oscillator is sufficient to provide the small amount of power required for a.c. biasing. Considerably more power would be needed to cover the requirements of a.c. erase, but this function can be performed by a permanent-magnet system.

The bias oscillator of Fig. 11 is based on the circuit given in Fig. 10. The frequency of oscillation of 42kc/s has been chosen as being five times the highest frequency to be reproduced in a tape recorder of reasonable quality. The circuit will operate satisfactorily with any OC72 and will continue to oscillate with a battery voltage of only about

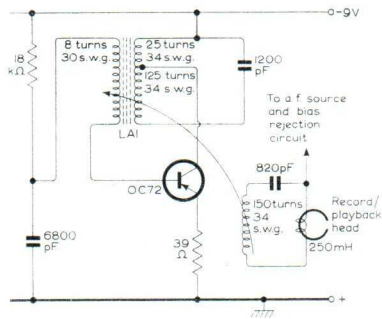


Fig. 11—42kc/s oscillator for feeding playback/record head of tape recorder

one-third of the nominal value. The current consumption is 5 to 10mA at 9V. The output voltage of between 32 and 35V peak is sufficient to provide a minimum of 1mA bias current in the 250mH inductance of the record/playback head shown in the circuit.

CHAPTER TWENTY-TWO

PHASE-SHIFT OSCILLATORS

The design of RC phase-shift oscillators is similar for transistors and thermionic valves, but three factors must be borne in mind:

- (a) the RC phase-shifting network is required to feed into the input of the transistor, which is of low impedance
- (b) the internal phase-shift of the transistor will be added to, or subtracted from, that of the network
- (c) when ladder networks are used, the current amplification factor of the transistor must be appreciably greater than the attenuation of the network.

ONE-TRANSISTOR OSCILLATORS

Where it is required to produce an oscillator with only one transistor, the grounded-emitter configuration will be chosen, because of its high gain. From the small-signal grounded-emitter equivalent circuit

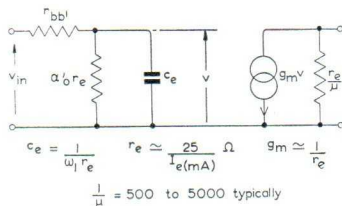


Fig. 1—Small-signal grounded-emitter equivalent circuit

shown in Fig. 1, it may be seen that the output-current generator $g_m v$ is in antiphase with the voltage v derived from the input voltage v_{in} . A network giving a phase shift of 180° between the output and input of the transistor is therefore wanted. Ladder networks are commonly used for this purpose.

Ladder Networks

Ladder networks may be built up by cascading a number of similar RC phase-shifting sections, using any one of the sections shown in Fig. 2 as a basic element.

Sections (a) and (b) give a phase shift between the input and output currents, so that the input of the transistor, which is connected across

the pair of terminals on the right-hand side, is fed with a current. The input impedance of the transistor should preferably be much less than the impedance of the network—which is R in (a) or the reactance of C in (b)—in order that the transistor impedance shall not disturb the operation of the circuit. This requirement is fairly easy to meet. However, an α'_o of at least 60 is needed for satisfactory operation, and

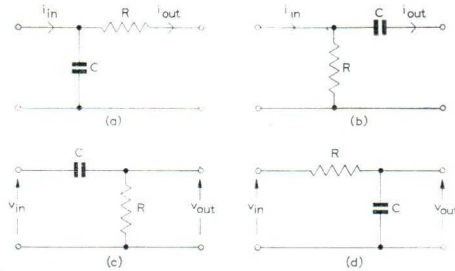


Fig. 2—Basic elements of ladder networks

since the input impedance of the transistor in grounded-emitter connection may be as high as $2.5k\Omega$ or more at $I_e = 1mA$, the resulting circuits are of high impedance, and a collector supply voltage of 12V or more is needed.

Sections (c) and (d) in Fig. 2 are voltage-transfer networks, and are normally employed with thermionic valves. For transistors, the input impedance of the transistor, which is connected across the right-hand side, needs to be large compared with the impedance of the network.

The voltage v_{in} across the left-hand terminals of (d) could be generated by allowing the output-current generator of the transistor (Fig. 1) to work into R , as may be seen from the equivalence in Fig. 3. No such

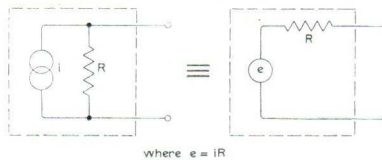


Fig. 3—Equivalence of i, R arrangement to e, R arrangement

simple method can be found for (c), and this circuit may be dismissed as unsuitable.

This way of generating v_{in} sets a lower limit to the value of R which may be employed while obtaining sufficient voltage for oscillation.

The condition is $R > Ar_e$, where A is the attenuation factor of the network, and is equal to 29 for a network of three equal sections. At $I_e = 1\text{mA}$, r_e is equal to $25/I_e = 25\Omega$, and R needs to be greater than $29 \times 25 = 725\Omega$ (say, $1.2\text{k}\Omega$).

To have an input impedance which is sufficiently high for satisfactory operation, the transistor needs to have an α'_o greater than 100. These networks therefore are better suited to thermionic valves than to transistors; although, if transistors of sufficiently high α' are available, the circuit can be made to work from lower supply voltages than when using sections (a) or (b).

Number of Sections

The most suitable RC phase-shift networks may be built up, therefore, from sections of the (a) or (b) type.

Neglecting the internal phase shift of the transistor for the moment, at least three sections are necessary, and the networks could be of

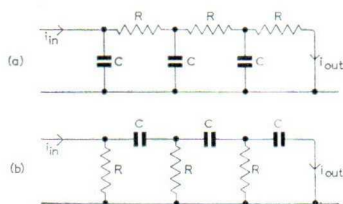


Fig. 4—Preferred ladder networks for transistors

the form of (a) or (b) in Fig. 4. If circuit (b) is used, the first R can be the load resistance of the transistor, and the circuit design is considerably simplified.

The current attenuation of these networks, at the frequency where the phase shift between the input and output currents is 180° , is 29. To allow for losses in the input and output impedances of the transistor, the α'_o should be greater than about 60. The OC75 is therefore suitable, though transistors at the top end of the OC71 production spread should also work in this type of circuit.

The current attenuation of 29 applies to a network having three equal resistances and three equal capacitances. These are the simplest to design and the most commonly used. Also, since the grounded-emitter transistor has an input impedance which is only an order or so different from its output impedance, tapered networks are of little use.

Operating Frequency

Single-transistor phase-shift oscillators are best restricted to low-

frequency operation, where the internal phase shift of the transistor need not be considered. At higher frequencies, both the phase shift and the reduction in α'_o cause design difficulties. For instance, with the network shown at (b) in Fig. 4, the phase shifts in the network and in the transistor vary in opposite senses, so that at high frequencies more phase shift has to be provided by the network. The phase shift in each section has to be increased, or a fourth section added. There is consequently more attenuation.

With network (a), the transistor and network phase shifts vary in the same sense, and a two-section oscillator can be constructed, though the operating frequency will be somewhat dependent upon the particular transistor. Also, since any shunt resistance across the input capacity of this network reduces the phase shift, a higher collector-load resistance, and possibly a higher transistor output impedance, will be demanded.

From Fig. 1, it may be seen that when the transistor is current fed, the phase shift in v , and hence in the output-current generator, will be 45° when the reactance of c_e is equal to $\alpha'_o r_e$. This frequency, which is above that which would normally be used with circuit (b), is called the grounded-emitter cut-off frequency, and is designated by f'_α . It is given by $f'_\alpha = f_1/\alpha'$, where $f_1 = f_\alpha/1.22$ for alloy-junction transistors.

800c/s Oscillator

A circuit using the (b) network and operating at 800c/s is shown in Fig. 5. Ideally the operating frequency is given by $f = 1/2\pi CR\sqrt{6}$,

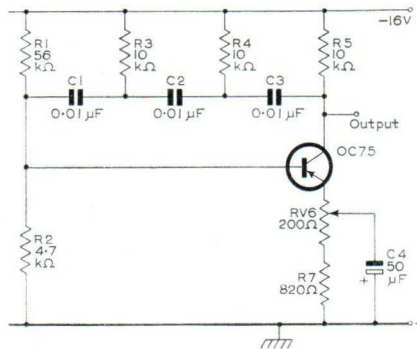


Fig. 5—800c/s oscillator using three-section ladder network

which for the circuit shown is 650c/s. The transistor input and output impedances modify this to 800c/s.

The value of the phase-shift resistors (10kΩ) is chosen to be a mean

between that which will be appreciably affected by the transistor output impedance, which is high, and that which will be appreciably affected by the transistor input impedance, which is low. With these networks, it is not easy to control the amplitude of oscillation without somewhat affecting the frequency of operation, and the amplitude control may change the frequency by 10% or so. The gain is controlled by changing the distribution of the feedback current between the base-bias resistors and the transistor input, the unbypassed resistance in the emitter increasing the transistor input impedance. The control should be adjusted so that the oscillation amplitude is smaller than that giving objectionable distortion.

WIEN-NETWORK OSCILLATORS

RC oscillators may also be constructed by arranging the phase-shift components to form a Wien network. The advantage of this network

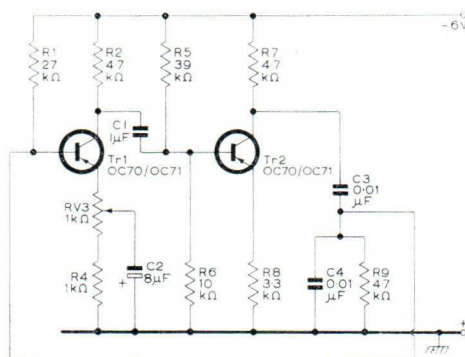


Fig. 6—Two-transistor Wien-network oscillator for about 3.4kc/s

is that the attenuation factor is only 3 at the frequency which gives zero phase shift although, since the output is in phase with the input, at least two amplifier stages are necessary.

Two-transistor Circuit

A simple two-stage oscillator of this type is given in Fig. 6, where

PHASE-SHIFT OSCILLATORS

the Wien network is R7, C3, C4, R9. Both stages are d.c. stabilised, the lower base-bias resistance of Tr1 also being part of the bridge

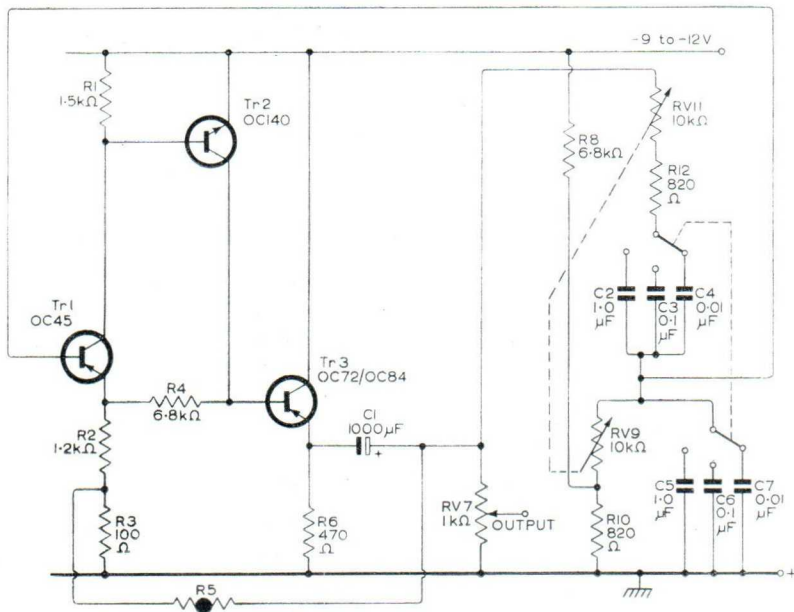


Fig. 7—Three-transistor Wien-network oscillator for 15c/s to 20kc/s

network (R9). The 3.3kΩ emitter resistor (R8) of Tr2 is left unbypassed, so that this stage has considerable a.c. negative feedback.

RV3 in the emitter of Tr1 provides a convenient means of adjusting the waveform for amplitude and distortion, and of compensating for changes in temperature from, say, one day to the next. The waveform is good, and the short-term temperature stability is fairly good. Supply voltages from at least -3V to -6V may be used, provided RV3 is suitably adjusted. The output may be taken from the collector or emitter of Tr2 (R7 or R8).

This circuit operates very nearly at the theoretical frequency of

$$\frac{1}{2\pi\sqrt{(C_3C_4R_7R_9)}} = \frac{1}{2\pi C_3R_7},$$

since in fact $C_3 = C_4$ and $R_7 = R_9$. R7 can be shown to be effectively in series with C3 by means of the equivalence given in Fig. 3.

With the capacitances shown, the circuit oscillates at a frequency of about 3.4kc/s. For operation at other frequencies, the capacitances should all be increased or decreased by the appropriate factor. (The resistances cannot be changed without altering the d.c. conditions.) The values of C3 and C4 should be chosen according to the frequency accuracy required, but for C1 and C2, the nearest standard values may be taken.

The circuit is sufficiently uncritical of gain to accept the OC70 or OC71, but if operation is required at higher frequencies, it will probably be better to use two OC45.

Three-transistor Circuit

A more professional Wien-bridge oscillator is shown in Fig. 7. This circuit incorporates a thermistor R5 as an amplitude-control device, and the output is essentially independent of small changes in supply voltage or ambient temperature. (A suitable component for R5 is the S.T.C. thermistor type R53.) Apart from the frequency-determining capacitors, only one capacitor is required. Consequently the unit is compact, and no difficulties arise from phase shift in the coupling capacitance.

The output voltage is 1V r.m.s., and the circuit operates with supply voltages between 7V and 12V and consumes about 10mA. In Fig. 7, a lower limit of 9V has been set to the supply voltage to ensure low distortion.

The frequency coverage is from 15c/s to 20kc/s in three ranges: 15 to 200c/s, 150 to 2000c/s and 1.5 to 20kc/s, the lower frequencies being associated with the larger capacitances. The ganged variable resistors, RV9 and RV11, allow the frequency to be adjusted within any given range. In an experimental model, the amplitude over the full range was constant to within better than 2%.

If it is desired to extend appreciably the upper frequency limit of the oscillator, Tr3 should be changed to an OC41. This modification will enable the oscillator to work satisfactorily at frequencies in excess of 100kc/s. For other transistors, R8, the 6.8k Ω bias resistor feeding the bridge, may need to be adjusted to ensure optimum working points for the transistors.

The OC140 (Tr2) is an n-p-n transistor, and its connections should therefore be made as in the circuit diagram, with the emitter connected to the negative supply line.

To avoid excessive distortion, the external load connected to the oscillator when the output is at its maximum should be not less than 1k Ω . With the addition of the load, the change in the maximum output voltage is less than 1%.

STABILITY

With a change of 3V in the supply voltage (from 9V to 12V), the change in frequency at 10kc/s and the change in output voltage are less than 1%.

CHAPTER TWENTY-THREE

TRANSISTOR AS A SWITCH

Many transistor pulse circuits are similar in form to classic thermionic-valve circuits, and perform similar operations. However, the transistor gives a closer approach to the ideal switch than the thermionic valve, and the limitations of the two types of device are different. The design of transistor circuits is based, therefore, not upon analogues, but on an analysis of the operating states of the transistor. Most of this chapter is consequently given over to the large-signal equivalent circuit.

In most pulse circuits, the transistor is used as a voltage or current switch, the timing operations being performed by combinations of resistance and capacitance ($C \times R$), or resistance and inductance (L/R). The limit to the pulse repetition frequency is set ultimately by the switching times of the transistor.

VOLTAGE SWITCH

The basic circuit of the transistor as a voltage switch is given in Fig. 1.

On Condition

The collector current that the transistor is capable of passing in the normal condition is $\bar{\alpha}I_e$ ($= \bar{\alpha}'I_b$). If the collector current is limited to some lower value—for example, by a resistive load—the emitter and base currents are higher than those required for normal transistor

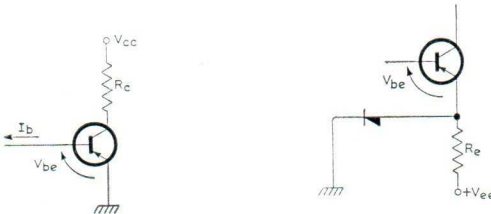


Fig. 1—Transistor as voltage switch

Fig. 2—Transistor as current switch

action. This is the bottomed (or current-saturated) condition. Both the collector and emitter diodes are forward biased.

Under normal operating conditions, the holes diffusing across the base of the transistor are swept rapidly across the collector depletion layer. As the collector current increases (Fig. 1), the voltage across the transistor is reduced. Consequently the field across the depletion

layer is diminished. To the first order of approximation, however, the flow of holes across the collector junction is not affected by the collector voltage. Even in the bottomed state, when many more holes leave the emitter than are required to maintain the maximum load current in the collector, the normal flow of holes across the collector junction is not affected. The excess holes that cross the junction bias it in the forward direction, so that these holes recross the junction into the base as a forward current through the collector diode.

The bottoming voltage in common (or grounded) emitter is therefore the difference between the voltages across the two forward-biased diodes. This voltage is quite small—about 100mV for alloy-junction transistors—and smaller than the knee voltage usually considered in the design of a.f. amplifiers.

Off Condition

If the base-emitter voltage is greater than $I_{co}r_{bb'}$, the emitter junction will be reverse biased. In this condition, therefore, both the diodes are reverse biased. The collector and emitter leakage currents are then given (as will be shown later) by:

$$I_{co(o)} = \frac{I_{co}(1-\alpha_I)}{1-\alpha_N\alpha_I} \quad \dots (1)$$

and

$$I_{eo(o)} = \frac{I_{eo}(1-\alpha_N)}{1-\alpha_N\alpha_I}, \quad \dots (2)$$

where α_N is the normal large-signal α of the transistor, and α_I is the large-signal α with the collector used as an emitter and the emitter used as a collector (inverse connection).

A suitable step of current on the base of the transistor will switch the collector voltage between the bottoming voltage and a voltage nearly equal to the supply voltage V_{cc} .

CURRENT SWITCH

The basic circuit of the transistor as a current switch is given in Fig. 2.

On Condition

Assuming that the emitter supply voltage V_{ee} is much greater than the switching voltage applied to the base, then a current

$$\simeq \frac{V_{ee}}{R_e}$$

is switched into the emitter of the transistor when the base is made more negative than the voltage dropped across the emitter junction.

(V_{be} is not more than 0.3V for the OC41 and OC42 at $I_e = 10\text{mA}$.) The collector current in this condition is

$$\approx \frac{\alpha V_{ee}}{R_e}.$$

When the transistor is used as a current switch, it must not be allowed to bottom; otherwise the output impedance would become very low, and the transistor would cease to act as a constant-current generator.

Off Condition

When the base is made positive to ground by a voltage greater than the voltage drop across the emitter junction, the emitter junction will be reverse biased, and a current

$$\approx \frac{V_{ee}}{R_e}$$

will be switched into the diode. (In practice the diode may be the emitter diode of another transistor.) The collector current in this condition is $I_{co(o)}$.

Thus a voltage step of about 1V on the base of the transistor will switch the collector current from $I_{co(o)}$ to approximately $\alpha V_{ee}/R_e$.

LARGE-SIGNAL EQUIVALENT CIRCUIT

The analysis of the operating states leads to the setting up of a large-signal equivalent circuit, from which the switching times of the transistor can be calculated. As this approach will be extensively employed in future, the derivation will be given in full.

The equivalent circuit for large-signal operation may be derived from a simple physical model. It is assumed that the transistor is a p-n-p type, but by reversing the sign where appropriate, the derivation will apply equally well to n-p-n transistors. Only alloy-junction transistors will be considered.

D.C. Equivalent Circuit

The large-signal d.c. equivalent circuit of an alloy-junction transistor is shown in Fig. 3. The diodes D_e and D_c represent the emitter and collector junctions of the transistor. As the emitter and collector junctions are very close together and form a single crystal with the base, part of the current flowing in at one junction will flow out at the other. These currents are represented on the equivalent circuit by the current generators $\alpha_{N_{e1}}$ and $\alpha_{I_{c1}}$ (where N and I denote normal and inverse connection). An appreciable resistance exists between the external base contact and the active area of the transistor. This resistance is represented in the equivalent circuit by $r_{bb'}$. The resistivity of the emitter and collector material is lower than that of the base, and

the currents in the emitter and collector flow through regions of much greater cross-sectional area. The internal resistances in series with the emitter and collector are therefore negligibly small, and are not included on the equivalent circuit.

Depletion Capacitance

When the voltage across either junction of the transistor changes, the

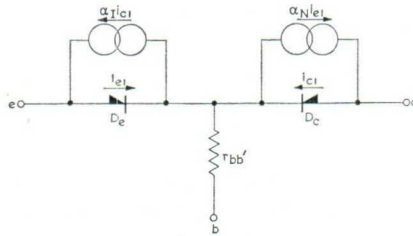


Fig. 3—D.C. equivalent circuit for large signals

width of the corresponding depletion layer also changes, and a charging/discharging current must flow. To take this effect into account, the depletion layers are represented by capacitances $c_{e(dep)}$ and $c_{c(dep)}$ on the full equivalent circuit (Fig. 4). The depletion capacitance of

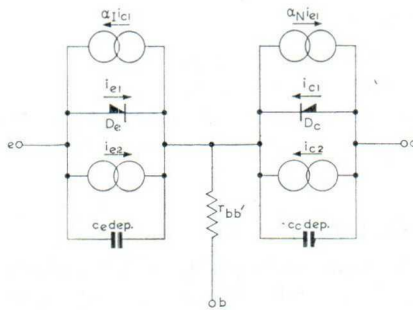


Fig. 4—Full equivalent circuit

alloy-junction transistors is inversely proportional to the square root of the voltage across the junction.

Diffusion Capacitance

The flow of current carriers through the base layer arises from the diffusion of holes within the field-free region of the base. There will

therefore be a concentration gradient of holes in the base, proportional to the flow of current through the base.

The hole-concentration pattern for a transistor operating in the normal active region is shown in Fig. 5(a). In the absence of a drift field, the concentration gradient is uniform and is represented by a straight line. The hole concentration at the collector junction is zero, because all holes which reach the junction are swept into the collector

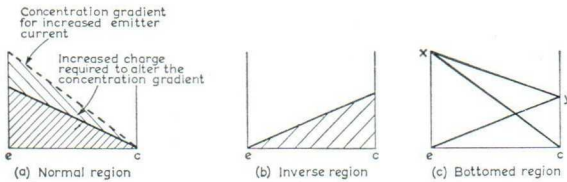


Fig. 5—Distribution of base charge for (a) normal region of characteristics (b) inverse region (c) bottomed region.

by the potential gradient which exists across the collector depletion layer. As can be seen from Fig. 5(a), when the emitter current changes, a capacitive current must flow to alter the hole-concentration gradient. The value of the capacitive current is given by

$$i_{e2} = \frac{1}{\omega_{1(N)}} \cdot \frac{di_{e1}}{dt}, \quad \dots(3)$$

where $\omega_{1(N)}$ is the angular frequency at which the normal α' becomes equal to one.

Also, it can be shown that the charge required in the base to provide a collector current of $\alpha_N I_e$ is given by

$$Q_b = \frac{I_e}{\omega_{1(N)}} \cdot \quad \dots(4)$$

When the transistor is operating in the inverse active region, the concentration pattern is as shown in Fig. 5(b). The only difference between this and the previous case is that, since current is flowing from collector to emitter, the concentration gradient is reversed. By similar reasoning to that used to derive Eq. 3, it can be shown that the capacitive current is given by

$$i_{c2} = \frac{1}{\omega_{1(I)}} \cdot \frac{di_{c1}}{dt}, \quad \dots(5)$$

where $\omega_{1(I)}$ is the angular frequency at which the inverse α' becomes equal to one.

The effect of the charges in the base can be taken into account by including two capacitive current generators i_{e2} and i_{c2} in the full

equivalent circuit (Fig. 4). The values of i_{e2} and i_{c2} are given by Eqs. 3 and 5.

When the transistor is operated in the saturation region, the concentration gradient is given by the line xy in Fig. 5(c). The concentrations given by this line can be considered to be the sum of the concentrations given by the lines xc and ye , which are represented on the equivalent circuit by the two capacitive generators i_{e2} and i_{c2} .

Simplification of Full Equivalent Circuit

The diodes D_e and D_c of Fig. 4 have characteristics of the form shown in Fig. 6(a). When the junction is reverse biased it can be considered

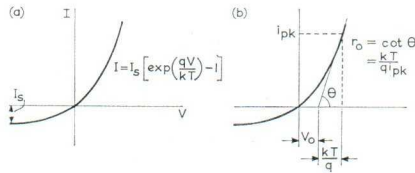


Fig. 6—Complete diode characteristic (a) and simplification of characteristic (b)

to be a current generator I_s . The equation of the curve is then given by

$$I = I_s \{ \exp (qV/kT) - 1 \}. \quad \dots (6)$$

A linear approximation may be made to this characteristic, in order to simplify the mathematical expressions obtained when using the equivalent circuit.

The linear approximation is shown in Fig. 6(b). When the junction is biased in the forward direction, the diode is replaced by a resistance r_o in series with a cell V_o . The value of r_o which correlates best with practical results is given by the slope of the diode characteristic at the peak current. The value can be shown, by differentiating Eq. 6, to be:

$$r_o = \frac{kT}{qI_{pk}}. \quad \dots (7)$$

The value of V_o is given by the intercept of the tangent at any particular value of i_{pk} on the voltage axis.

CUT-OFF STATE

The equivalent circuit for the cut-off state (Fig. 7) can be derived from the full equivalent circuit given in Fig. 4. Since D_e and D_c are reverse biased, they can be represented by current generators I_{es} and I_{cs} . The capacitive currents i_{e2} and i_{c2} are zero, because they are proportional

to the rates of change of I_{es} and I_{cs} , which are constant. The current generators $\alpha_N i_{e1}$ and $\alpha_I i_{c1}$ in Fig. 4 become $\alpha_N I_{es}$ and $\alpha_I I_{cs}$ in Fig. 7, since $i_{e1} = -I_{es}$ and $i_{c1} = -I_{cs}$ when the junctions are reverse biased.

I_{es} and I_{cs} are the saturation currents of the emitter and collector diodes with the opposite diode short-circuited. These currents can be

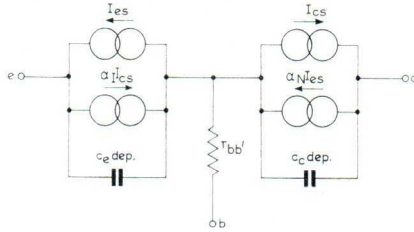


Fig. 7—Equivalent circuit for cut-off region

expressed in terms of I_{eo} and I_{co} . From Fig. 7, the value of the collector current with the emitter open-circuited is

$$I_{co} = I_{cs} - \alpha_N I_{es},$$

and since in this state

$$\begin{aligned} I_{es} &= \alpha_I I_{cs}, \\ I_{co} &= I_{cs}(1 - \alpha_N \alpha_I). \end{aligned}$$

Therefore

$$I_{cs} = \frac{I_{co}}{1 - \alpha_N \alpha_I} \quad \dots (8)$$

Similarly

$$I_{es} = \frac{I_{eo}}{1 - \alpha_N \alpha_I} \quad \dots (9)$$

The value of collector current when the emitter junction is reverse biased is obtained from Fig. 7, thus:

$$I_{co(o)} = I_{cs} - \alpha_N I_{es},$$

and since now

$$\begin{aligned} \alpha_N I_{es} &= \alpha_I I_{cs}, \\ I_{co(o)} &= I_{cs}(1 - \alpha_I) \\ &= \frac{I_{co}(1 - \alpha_I)}{1 - \alpha_N \alpha_I} \quad (\text{from Eq. 8}). \end{aligned}$$

Similarly

$$I_{eo(o)} = \frac{I_{eo}(1 - \alpha_N)}{1 - \alpha_N \alpha_I}.$$

NORMAL ACTIVE REGION

In the normal active region the emitter diode D_e (Fig. 4) is biased in the forward direction. It can therefore be represented by a resistance r_{oe} in series with a cell V_{oe} (Fig. 8). The capacitive current

$$i_{e2} = \frac{1}{\omega_1(N)} \cdot \frac{di_{e1}}{dt}$$

will then flow through a capacitance of

$$c_e = \frac{1}{\omega_1(N)r_{oe}}$$

in parallel with r_{oe} , since

$$\begin{aligned} i_{e2} &= c_e \cdot \frac{dV}{dt} = \frac{1}{\omega_1(N)r_{oe}} \cdot \frac{d(i_{e1}r_{oe})}{dt} \\ &= \frac{1}{\omega_1(N)} \cdot \frac{di_{e1}}{dt} \end{aligned} \quad \dots(10)$$

The depletion capacitance $c_{e(dep)}$ is not included in Fig. 8, because it

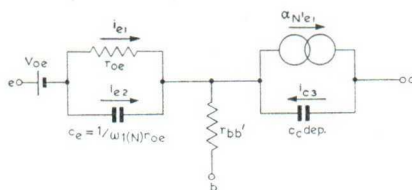


Fig. 8—Equivalent circuit for normal region

is negligibly small in comparison with the diffusion capacitance c_e .

This equivalent circuit is used to calculate switching times. The current generators (I_{cs} and $\alpha_1 I_{cs}$) have very little effect on switching times and have therefore not been included in Fig. 8.

INVERSE ACTIVE REGION

The considerations leading to the derivation of the equivalent circuit for the normal active region apply equally well to the inverse active region and result in the circuit of Fig. 9. Here, however, the inverse parameters α_1 and $\omega_1(I)$ are used instead of the normal parameters.

BOTTOMED REGION

Fig. 10 shows the equivalent circuit for the bottomed region. The diodes D_e and D_c (Fig. 4) are biased in the forward direction, and are represented by resistances r_{oe} and r_{oc} in series with the cells V_{oe} and V_{oc} . The capacitive currents i_{e2} and i_{c2} flow through the diffusion

capacitances c_e and c_c . The depletion capacitances $c_{e(\text{dep})}$ and $c_{c(\text{dep})}$ are omitted because they are negligibly small in comparison with c_e and c_c .

Use of Large-signal Equivalent Circuit

The equivalent circuit derived above can be used to calculate the behaviour of the transistor under large-signal conditions. In particular

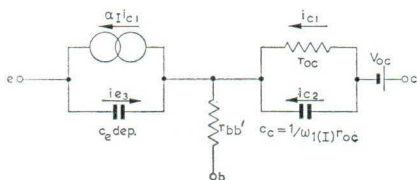


Fig. 9—Equivalent circuit for inverse region

the switching times of the transistor may be calculated. The switching times are evaluated by forming a differential equation for the equivalent circuit of the transistor and the external circuit, and then solving the equation. The theoretical values thus obtained agree reasonably well with those measured in practice.

SPECIAL CASE

The equations governing the performance for generalised conditions are quite complicated. As an example of the use of the large-signal equivalent circuit, it will therefore be convenient to consider only a

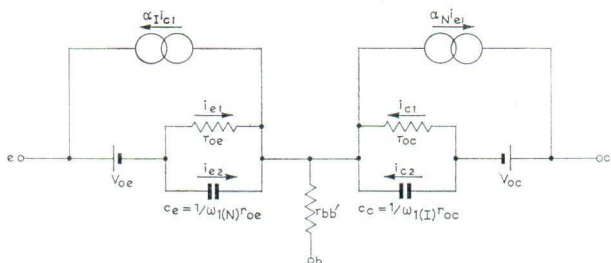


Fig. 10—Equivalent circuit for bottomed region

special case, that in which the transistors are overdriven (as is common practice) to obtain fast switching times.

Recombination can be assumed to give an approximately exponential decay of base charge, and a recombination time-constant may be defined from the exponential curve in the usual way. If the switching time is small in comparison with the recombination time-constant, the

effect of recombination may be ignored. The switching time can then be found by dividing the required base charge by the base drive current, since by definition $[Q] = [t] \times [I]$.

The recombination time-constant can be derived as follows:

From Eq. 4,

$$Q_b = \frac{I_e}{\omega_{1(N)}} .$$

The recombination current is given by

$$I_b = \frac{I_e}{1 + \bar{\alpha}'}$$

assuming an emitter efficiency of one. The recombination time-constant is therefore given by

$$\frac{Q_b}{I_b} = \frac{1 + \bar{\alpha}'}{\omega_{1(N)}} .$$

The base charge required to switch the transistor on to a collector current I_c can be found from the equivalent circuit (Fig. 8). If recombination is ignored, $\alpha_N = 1$ and all the drive current is available to charge c_e and $c_{c(dep)}$. The turn-on time is then given by

$$t_{on} = \frac{Q_{b(on)}}{I_{b(on)}} \quad \dots(11)$$

and the charge required to switch the transistor on is

$$\begin{aligned} Q_{b(on)} &= Q_{ce} + Q_{c_{c(dep)}} \\ &= \frac{1}{\omega_{1(N)} r_{oe}} \cdot I_c r_{oe} + c_{c(dep)} \Delta V_c \\ &= \frac{I_c}{\omega_{1(N)}} + c_{c(dep)} \Delta V_c \quad \dots(12) \end{aligned}$$

where ΔV_c is the change in collector voltage.

The charge required to switch the transistor off, if it is not bottomed, is equal to that required to switch it on. When the transistor is bottomed, additional charge is required to switch it off, in order to overcome the excess base current. The extra charge is

$$\left(I_b - \frac{I_c}{\bar{\alpha}'} \right) \tau_s$$

where τ_s is the hole-storage time-constant of the transistor and is given by

$$\tau_s = \frac{1/\omega_{1(N)} + 1/\omega_{1(I)}}{1 - \alpha_N \alpha_I} .$$

The total charge required to switch the transistor off is then

$$Q_{b(off)} = \frac{I_c}{\omega_{1(N)}} + c_{c(dep)} \Delta V_c + \left(I_b - \frac{I_c}{\beta'} \right) \tau_s. \quad \dots (13)$$

The turn-off time is given by

$$t_{off} = \frac{Q_{b(off)}}{I_{b(off)}}. \quad \dots (14)$$

POWER SWITCH

In relay-switching circuits, the transistor operates essentially as a power switch. A typical relay drive circuit is shown in Fig. 11. R_L is the resistance of the relay winding.

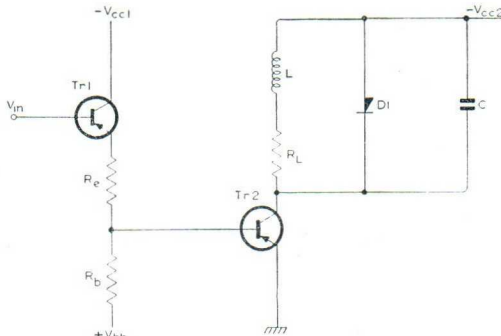


Fig. 11—Relay drive circuit

On Condition

When V_{in} is negative, the negative base current ($-I_{b1}$) flowing out of Tr2 is sufficient to bottom this transistor. The current in the load rises to V_{cc2}/R_L . The total dissipation in the transistor is given by

$$p_{tot} = \frac{V_{ce} V_{cc2}}{R_L} + V_{be} I_b.$$

The power in the load is $(V_{cc2})^2/R_L$ and is comparatively high.

Off Condition

When V_{in} goes positive, Tr1 is cut off. A positive base current of approximately $+I_{b2} = V_{bb}/(R_b + r_{bb'})$ flows into Tr2, bringing Tr2 out of bottoming and tending to switch it off.

The collector current in this condition (emitter and collector diodes reverse biased) is $I_{co(0)}$. The dissipation in the transistor is approximately $V_{ce2} I_{co(0)}$ and the power in the load is $(I_{co(0)})^2 R_L$.

Delayed Switch-off

A description of the delayed switch-off which occurs with an inductive load has already been given in Chapter 9 (pp. 101 and 102).

The induced voltage at switch-off is usually sufficient to make the collector potential rise almost immediately to the supply voltage. The relay should be shunted by a catching diode, such as D1 in Fig. 11, which will conduct when $V_c = V_{ce2}$ and prevent any further rise in collector voltage. The OA81, OA5 or OA10 should be used, according to the peak current and the voltage to be handled.

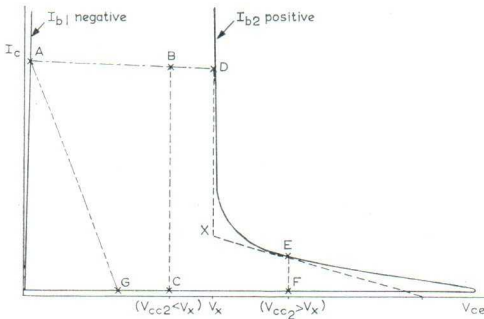


Fig. 12—Static output characteristic illustrating switch-off loci

During switch-off, the operating point (Fig. 12) follows some path such as ABC ($V_{ce2} < V_x$) or ADEF ($V_{ce2} > V_x$). The transistor is then in a high-dissipation region. The switch-off time may be only a few microseconds for the path ABC, but may be tens of milliseconds ($\sim L/R_L$) for ADEF.

Overheating of the transistor for the condition $V_{ce2} > V_x$ can be prevented by shunting the relay by a capacitance C (Fig. 11). By choosing a suitable value for C , the switch-off time can be shortened and the locus of the operating point modified to some path, such as AGF, where the dissipation is lower.

The value of C which ensures that the collector current reaches zero by the time the collector voltage reaches some value V_1 is given by $C = It/2V_1$, where I is the collector current immediately before switch off and t is the maximum switch-off time of collector current (assumed linear). C should not be given a larger value than necessary, since it increases the dissipation during switch on.

CHAPTER TWENTY-FOUR

PULSE CIRCUITS

Switching between two operating states may be achieved by means of regenerative feedback. The two states may both be unstable (multivibrator or astable circuit) or one may be unstable and the other stable (flip-flop or monostable circuit), or both may be stable (Eccles-Jordan or bistable circuit). The family resemblance between the three types of circuit is brought out by the basic circuits of Figs. 1, 2 and 3. The distinguishing features of the circuits are the cross-coupling components.

MULTIVIBRATOR (Astable Circuit)

The multivibrator (Fig. 1) will be considered first, as this is probably the best known and, in appearance at least, the simplest. The circuit in

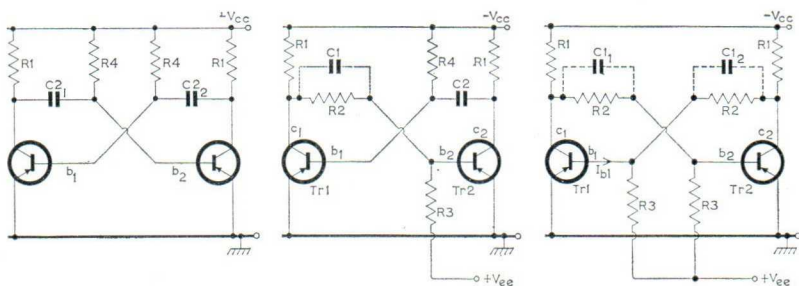


Fig. 1—Basic multivibrator (astable circuit)

Fig. 2—Basic flip-flop (monostable circuit)

Fig. 3—Basic Eccles-Jordan (bistable) circuit

its thermionic-valve form was described by Abraham and Bloch in 1918. It was first conceived as a square-wave generator, and since a square-wave is very rich in harmonics, the name multivibrator was coined.

Upon switching on the circuit, slight unbalance in the components, or random variations in the currents, will cause the transistors to go into one of the unstable states rather than the other. The circuit is therefore self-starting.

The circuit subsequently oscillates between the two unstable states. The multivibrator is the only one of the three circuits which is a free-

running oscillator, although if required pulses may be applied to synchronise the frequency.

Positive feedback is provided by the capacitive cross-coupling from the collector of each transistor to the base of the other transistor. The loop gain is greater than one. Assuming an initial condition where Tr1 is on and Tr2 off (Fig. 1), then as C_{21} charges, the base b_2 of Tr2 goes negative and brings Tr2 into conduction. Regenerative switching occurs, and Tr1 is now off and Tr2 on. C_{22} now charges, b_1 goes negative until Tr1 conducts, and the cycle starts again.

The duration t of the quasi-stable state — quasi-stable rather than unstable, since the circuit stays in this state for a certain length of time — is given approximately (Fig. 1) by

$$t \sim 0.7C_2R_4.$$

This expression neglects, among other things, the fact that in reality C2 has two discharge paths, one through R_4 and the other through the base of the transistor. When the transistor is cut off, its base current is constant and equal to the leakage current. The frequency of operation may therefore be somewhat temperature dependent. The output may be taken from either collector.

The multivibrator is sometimes used in frequency multiplication, the required harmonic of the fundamental frequency being selectively amplified and appearing as a sinusoidal output from the tuned circuit.

A representative multivibrator is given in Fig. 4, the values of C_1 , C_2 , R_2 and R_3 being chosen from the accompanying table accord-

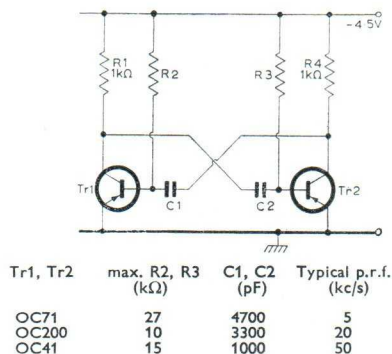


Fig. 4—Practical multivibrator

ing to the transistor which is to be used. With these particular values, the circuit will operate with all transistors of the type in question at the stated pulse repetition frequency.

The value of R_2 and R_3 is the maximum which ensures that the lowest-gain transistor will be bottomed when turned on. As the p.r.f. of the circuit is governed by the time-constant C_1R_2 ($= C_2R_3$), the only possible modification is to increase the value of C_1 and C_2 if a lower operating frequency is required.

In a symmetrical circuit, where $C_1R_2 = C_2R_3$, the square waveform has equal on and off times. Unequal on and off times can be achieved by making the two time-constants unequal, but the circuit is not suitable for use with very large mark/space ratios.

The limit to the p.r.f. is set by the switching times of the transistor, which are chiefly a function of cut-off frequency. It is only to be expected that, of the transistors considered in Fig. 4, the highest p.r.f. would be obtained with the OC41, but in some applications, the OC200 may have the advantage because of its very low leakage current.

FLIP-FLOP (Monostable Circuit)

The flip-flop circuit (Fig. 2) has one stable and one unstable or quasi-stable state. A trigger pulse flips the circuit into the unstable state, and the circuit subsequently flops back into the stable state.

The astable circuit can be made monostable by replacing one of the capacitive cross-couplings by a resistive coupling. (In fact the coupling resistor is shunted by a capacitance, but this should be ignored for the moment.) The stable state is with $Tr1$ bottomed and $Tr2$ cut off. The trigger pulse switches $Tr1$ off and $Tr2$ on. The collector c_2 of $Tr2$ goes positive, and takes the base b_1 of $Tr1$ positive with it, thus cutting off $Tr1$. C_2 now discharges through R_4 , and the circuit automatically switches back when the base-emitter voltage applied to $Tr1$ is approximately zero.

The input capacitance of $Tr2$ reduces the loop gain at high frequencies, thus increasing the switching time. In practice the attenuation caused by the input capacitance is compensated by shunting R_2 by $C1$.

The duration of the quasi-stable state is again given approximately (Fig. 2) by

$$t \sim 0.7C_2R_4.$$

The (monostable) flip-flop delivers one output pulse for every input pulse. It may be used for pulse amplification and pulse shaping, or simply to delay the trigger pulse applied to the input, triggering of the following stages being effected by the rear edge of the output pulse. The output may be taken from either collector.

A representative circuit for the (monostable) flip-flop is given in Fig. 5, with extra details in the table. The trigger pulse has to last long enough for the circuit to switch, that is, the trigger width has to be longer than the switching time. The length of the output pulse is determined mainly by the time-constant C_3R_3 . The value of C_2

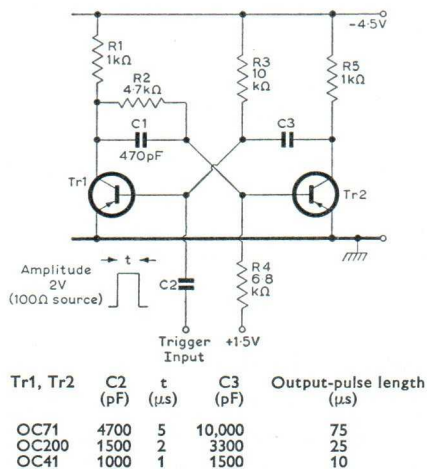


Fig. 5—Practical flip-flop

should be high enough to ensure that sufficient charge is extracted from the base of Tr1 when the transistor is switched off by the trigger pulse. Carrier storage is lower for transistors having higher cut-off frequencies, and a smaller trigger capacitance (C_2) is therefore satisfactory for the OC200 and, to an even greater extent, the OC41.

ECCLES-JORDAN CIRCUIT (Bistable Circuit)

The original thermionic-valve version of the bistable circuit was described by Eccles and Jordan in 1919. There are two stable states, and the circuit will only change from one to the other when a trigger pulse is applied.

The bistable circuit (Fig. 3) can be derived from the astable circuit by replacing both the capacitive cross-couplings by resistive couplings. The two resistors are in fact shunted by capacitors, for the same reason as in the monostable circuit.

The two stable states of the circuit are:

- (a) Tr1 bottomed and Tr2 cut off;
- (b) Tr1 cut off and Tr2 bottomed.

If Tr1 is bottomed and Tr2 cut off, a positive trigger pulse on the base

of Tr1 will tend to cut this transistor off. The collector c_1 of Tr1 will go negative, and the base b_2 of Tr2 will go negative in turn. Tr2 therefore comes into conduction, and c_2 and hence b_1 go more positive. Thus the loop can be made regenerative, the necessary condition being that the voltage gain from base to base be greater than one.

The choice of the various component values is always a compromise. A conventional bistable circuit with representative values is shown in

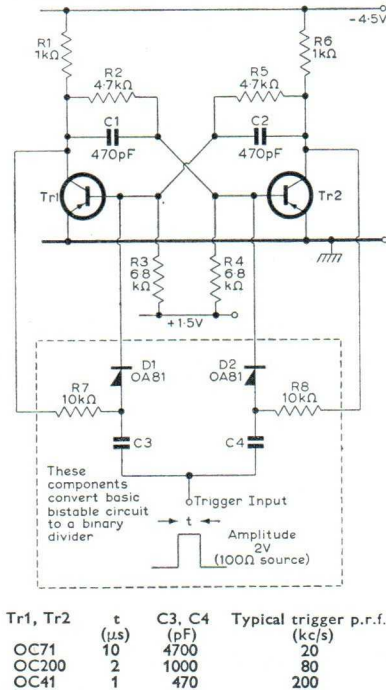


Fig. 6—Practical bistable circuit

the top part of Fig. 6; however, for some particular application, it may be necessary to change the values considerably.

Each trigger pulse causes the circuit to change from one stable state to the other. The bistable circuit carries out one complete cycle for every two input pulses. During a complete cycle, one pulse appears at each collector. The pulse repetition frequency of the output from either collector is therefore half that of the input. (This is the basis of binary counting.)

The simplest method of taking off the output is by capacitive coupling from either of the collectors. A higher switching speed may

be obtained, however, with a coupling transformer connected in series with one of the collector load resistors. With transformer coupling, repetition rates of up to approximately one-fifth of f_{α} are possible.

One sound method of converting the bistable circuit to a binary divider (binary counter) is illustrated by the components boxed in by the dashed line in the lower part of Fig. 6. This circuit has been designed for reliable operation with all transistors of the type in question up to the stated values of trigger-pulse repetition frequency, so that these values do not represent the ultimate in performance.

The diodes D1 and D2 are controlled by the collector voltages. When Tr1 is on, it is bottomed, so that D1 is just conducting. Tr2 is off, and D2 is biased off by the negative voltage at the collector. The next positive pulse is passed to the base of Tr1, switching Tr1 off and Tr2 on. The time-constant $C_3R_7 (= C_4R_8)$ is arranged such that the diode remains conducting until the end of the input pulse, thus ensuring that the circuit will always complete its switching before the 'gate' is opened to the next input pulse.

The time that must be allowed to elapse between trigger pulses is determined by the time required for C3 (or C4) to return to its initial voltage. Thus the input time-constant $C_3R_7 = C_4R_8$ sets an upper limit to the trigger-pulse repetition frequency. For effective triggering, this time-constant needs to be about five times the trigger width.

The typical p.r.f. is of course highest for the OC41 which, with its higher-frequency version the OC42, is recommended for this type of circuit.

The pulse repetition frequency may be divided by any power of two, by combining the requisite number of binary counters in cascade (binary system).

ASYMMETRICAL BISTABLE CIRCUIT

The Eccles-Jordan bistable circuit is completely symmetrical. For some purposes an asymmetrical circuit may be preferable. Such a circuit is shown in Fig. 7.

Tr1 is a grounded-base and Tr2 a grounded-collector stage. In the grounded-base configuration, the current amplification factor is almost one, the input impedance is low and the output impedance is high. At low frequencies, the transistor does not introduce any phase shift between the input and output signals. To obtain regenerative feedback from the output to the input, it is necessary to bring about a transformation in impedance without introducing any change in phase.

In the grounded-collector arrangement, a transistor has a high input and a low output impedance. It has a current amplification

factor of α' and at low frequencies there is again no phase shift within the transistor.

A combination of two such stages therefore gives regenerative feedback. The loop gain is greater than unity, since $R_2 > R_4$.

The two stable states of the circuit are: (a) Tr1 conducting, Tr2 almost cut-off; (b) Tr1 cut-off by a negative bias on its emitter, Tr2

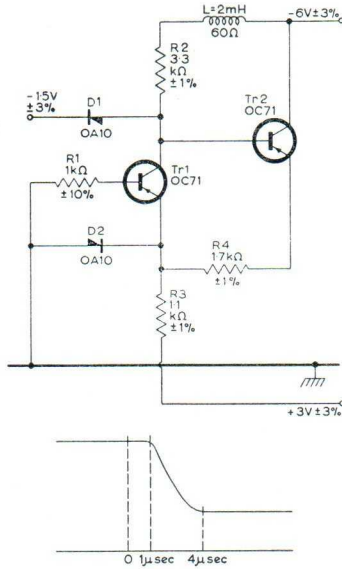


Fig. 7—Asymmetrical bistable circuit

conducting. The circuit may be triggered on either the base or emitter of Tr1, both points being sensitive to pulses of either polarity.

The two diodes D1 and D2 are added to the circuit to prevent excessive variation in trigger sensitivity from circuit to circuit. They also improve the switching time. D1 is a catching diode which prevents Tr1 from bottoming, thus greatly reducing the hole storage. A smaller trigger pulse is required, since it is no longer necessary to expel so many holes from the base, and at the same time the switch-off time is shortened. D2 limits the cut-off bias and thus reduces the negative trigger pulse required on the base to switch on the transistor. The 2mH choke is a further improvement which increases the high-frequency loop gain and improves the switching time by about 1μsec.

The asymmetrical bistable circuit is suitable for a repetition rate of about one-fifth of f_{α} .

P-N-P/N-P-N BISTABLE CIRCUIT

The OC139 and OC140 are n-p-n transistors and are approximately opposite-polarity versions of the OC41 and OC42 respectively. The output from the collector of a p-n-p transistor is of suitable polarity for direct coupling to the base of an n-p-n transistor. Fig. 8 shows

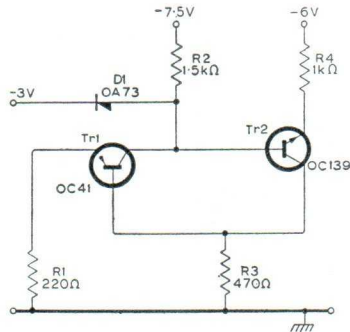


Fig. 8—Bistable circuit using p-n-p and n-p-n combination

such a bistable circuit using the OC41 and OC139. The current gain is greater than one and there is no phase reversal.

The base of the n-p-n transistor, Tr2, is connected by a series resistor to a voltage more negative than the collector supply voltage. This arrangement allows the base current to reverse, and improves the switch-off time.

D1 is a catching diode which prevents bottoming of Tr1.

SCHMITT TRIGGER ACTION

The asymmetrical bistable circuit (Fig. 7) may be used as a Schmitt trigger circuit. The input is applied to the base of Tr1 via the existing 1kΩ resistor R1. The circuit will then trigger in one direction when the input exceeds a certain level, and will trigger back again when the input returns to approximately the same level.

The backlash or hysteresis of the system depends on the loop gain and only disappears when the loop gain is just equal to one.

The p-n-p/n-p-n bistable circuit (Fig. 8) may be used in a similar manner, the input being applied to the emitter of Tr1.

BLOCKING OSCILLATORS

Bistable, monostable and astable circuits may be designed using transformer coupling to provide the positive feedback, instead of RC

coupling. The transformer increases the loop gain, and hence results in faster switching.

Blocking oscillators as such are transformer-coupled monostable or astable pulse generators. Whether or not the circuit is free-running (astable) depends on the d.c. conditions. If a d.c. state exists in which the loop gain of the circuit is less than one, then the circuit will be stable in that state and will require to be triggered out of it (monostable circuit).

The circuit consists essentially of a transistor with transformer-coupled feedback from the collector to either the base or the emitter, and with an RC timing network in either the emitter or the base circuit. Sufficient feedback is applied to drive the transistor into bottoming at the beginning of the first cycle of oscillation. The transistor then remains bottomed for a period determined by the circuit constants, after which regeneration causes a rapid switch-off. The output voltage takes the form of a rectangular pulse.

The advantages of the blocking oscillator are a low output impedance, short output pulses if required, and an output which can be adjusted to any required amplitude by a suitable choice of transformer winding.

Triggered Circuit

A practical circuit for a triggered blocking oscillator is given in Fig. 9. This circuit will operate at a pulse repetition frequency of 100kc/s at

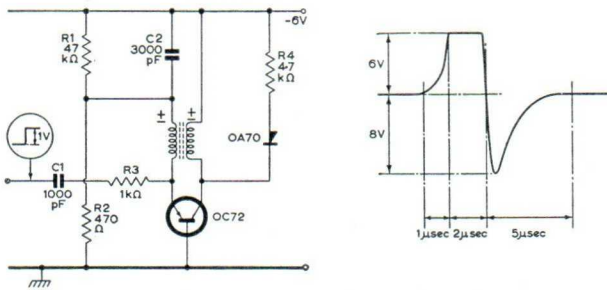


Fig. 9—Triggered blocking oscillator

ambient temperatures of up to 60°C.

The transformer details are: Ferroxcube core, FX1011; primary (collector) winding, 60 turns of 38 s.w.g. wire; secondary (emitter) winding, 10 turns of same wire.

A diode and a damping resistor R4 are connected across the primary of the transformer to prevent excessive overshoot. Bias is applied by R1 and R2, C1 is the coupling capacitor, and R3 ensures a certain minimum source impedance for the transistor.

During the bottomed period, when the collector voltage is constant, the collector current continues to increase at a constant rate through the inductive branch of the load. The emitter current however is virtually constant, and is determined by the transformer feedback voltage and the resistance in the emitter circuit. When the emitter current is no longer sufficient to maintain the rate of increase of collector current, the voltage across the transformer falls; the transistor comes out of bottoming, and regeneration causes the voltages across the transformer windings to reverse and the transistor to switch off rapidly.

C2 then discharges through R2, and the energy in the transformer is dissipated in the diode and damping resistor. The larger of these time-constants determines the period before the cycle can recommence.

Free-running Circuit

Fig. 10 shows a free-running blocking oscillator which can operate as a frequency divider. If the oscillator is synchronised by an input p.r.f. of up to 100kc/s, a p.r.f. of 10kc/s, or proportionately less, may

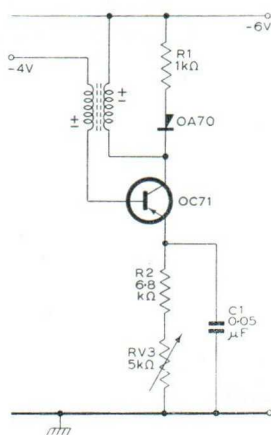


Fig. 10—Free-running blocking oscillator (frequency divider)

be obtained at the output. The division ratio may be as high as 10:1 and is stable for ambient temperatures of up to 60°C. The transformer core is again the FX1011 and the primary (collector) winding 60 turns of 38 s.w.g. wire, but the secondary (base) winding is now increased to 30 turns of the same wire.

Despite first appearances, this circuit is essentially a variation on the preceding one. The feedback is applied to the base instead of to the emitter, as the loop gain is then higher and the circuit oscillates

more freely. The timing circuit is in the emitter circuit, RV3 being the frequency control.

The prime requirement of a frequency divider is that the reset time be stable. The stability of this time determines the maximum division-ratio of the circuit.

The reset time of the circuit is the time required for the capacitor to return to a voltage at which the transistor can conduct again. In addition to the discharge through the associated resistance, the leakage current $I_{eo(o)}$ helps to discharge the capacitor. As the leakage current is temperature dependent, for a stable reset time it is essential to ensure that $I_{eo(o)}$ at its maximum value is a negligible part of the total discharge current.

If the RC timing network is in the base circuit, the leakage current which contributes to the discharge is the collector leakage current $I_{co(o)}$. With the timing network in the emitter circuit, only $I_{eo(o)}$ flows to the capacitor. As $I_{eo(o)}$ is only about one-fortieth of $I_{co(o)}$, the temperature dependence of the circuit is very much reduced.

SAWTOOTH-WAVEFORM GENERATION

Transistors may be used to generate sawtooth-waveforms in circuits which take their names, at least, from thermionic-valve counterparts.

A well-known technique in thermionic-valve circuits is the use of an anode-grid capacitor to provide Miller voltage feedback. A feed-

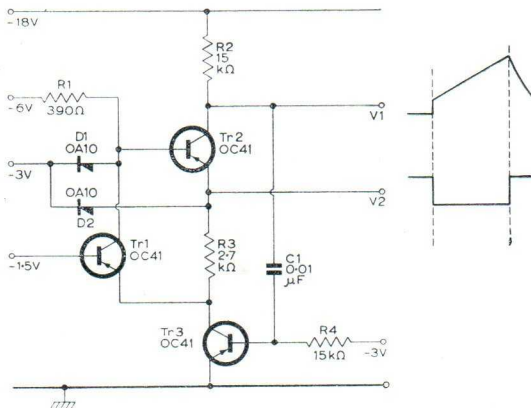


Fig. 11—Self-gating Miller circuit

back capacitor may likewise be used with transistors to provide current feedback.

The discharge of the capacitor can be made sufficiently linear to

provide a sweep voltage for a cathode-ray tube. For a fast flyback, it is necessary to interrupt the negative-feedback action of the circuit. One method is to connect two transistors in series, the feedback loop being open or closed according to whether the second transistor is cut-off or conducting. A further transistor, used instead of a catching diode, makes the circuit self-gating.

A practical self-gating Miller circuit is given in Fig. 11. Tr1 is the catching transistor; Tr2 and Tr3, which are connected in series by R3, provide two gating electrodes in the current path and simulate, to some extent, the gating action of a pentode.

In the rest condition, Tr1 and Tr3 are conducting, and the collector of Tr3 is at approximately -1.5V , the base potential of Tr1. Both the diodes D1 and D2 are conducting, so that the base and emitter of Tr2 are both at the same potential (-3V) and the transistor is just cut off. The collector of Tr2 is therefore at the line potential of -18V .

On applying a trigger pulse to cut off Tr1 (for example, a negative trigger pulse on the emitter), V_{e1} switches to -3V . D1 and D2 are both cut off; Tr2 conducts, feeding back a step of current to the base of Tr3 via C1. The capacitive feedback loop through Tr3, Tr2 and C1 is thus closed, and the current through these two transistors rises linearly with time. The positive-going flanks of the sawtooth occur at the collectors of Tr3 and Tr2. When V_{c3} reaches -1.5V , Tr1 starts to conduct; V_{e1} goes positive, cutting off Tr2, so that V_{c2} flies back to -18V . The circuit thus resets itself to the rest condition.

In addition to the sawtooth waveform at the collector of Tr2, a rectangular waveform, corresponding to the sweep time, is available at the emitter. The circuit may be triggered by a positive pulse applied to the collector of Tr2, or by a negative pulse at either the collector or emitter of Tr1.

The circuit illustrated is only one among several possible circuit combinations. A parallel instead of a series arrangement may be adopted for the gating transistor, and a directly coupled p-n-p/n-p-n combination (giving a faster flyback time) is suitable for the feedback loop. For a particularly linear sweep, a circuit may be developed from the thermionic-valve 'bootstrap' circuit. The need for suitable trigger points is another factor which influences the choice of circuit arrangement.

CHAPTER TWENTY-FIVE

D.C. AMPLIFIERS

Amplification of a d.c. input is frequently required for industrial and scientific purposes. Many applications will probably suggest themselves to the worker in any particular subject:

- (a) d.c. amplifiers may form part of temperature-measuring devices using thermocouples, platinum resistance thermometers or thermistors
- (b) in general, the d.c. amplifier permits low-range meters, such as micro-ammeters and high-impedance voltmeters, to be replaced by more robust types
- (c) once the d.c. level has been raised sufficiently, it can be used to operate a relay in an alarm or automatic-control system
- (d) other applications include analogue computers and electro-cardiographs.

Amplification of a d.c. signal is a simple requirement, but the choice of a suitable circuit can be a difficult problem, as many of the circuits which have been published deal with specific applications. This chapter gives a survey which will make it fairly simple to choose the type of d.c. amplifier required.

A comparison of the performance of the circuits is given at the end of the chapter. The circuits illustrate the principles of the various types of d.c. amplifier, and usually modifications will be necessary for any particular application.

Directly coupled and chopper types are discussed.

DIRECTLY COUPLED AMPLIFIERS

The main problems of d.c. amplification are drift of the zero-reading and variations in gain. The gain can be kept as constant as required by applying negative feedback. Feedback does not dispose of drift, however, as the gain and drift are reduced proportionally.

Temperature variations are mainly responsible for drift in transistor circuits, and their effect is particularly pronounced in directly coupled amplifiers. In these amplifiers, the temperature dependence of the leakage current and of the base-emitter voltage is most important. The temperature dependence of α' is not usually significant, though in

some circuits it has to be taken into account. These effects are additive, that is, they all cause an increase in collector current with temperature.

The effect of the base-emitter voltage V_{be} is less well known than that of the leakage current. V_{be} at constant emitter current falls by roughly 2mV per °C rise in temperature, for both silicon and germanium transistors made by the alloy-junction method. If the stage is fed from a low source impedance, the drift introduced by the change in V_{be} as the temperature varies may be larger than that introduced by the change in leakage current. In the simpler type of directly coupled amplifier, the minimum source resistance is specified at which the drift caused by V_{be} becomes comparable with leakage-current drift. In more elaborate arrangements, this drift is taken care of in other ways.

Amplifier using Germanium Transistors

Fig. 1 shows a straightforward experimental circuit using two germanium transistors (an OC72 and an OC35). Grounded-emitter connection is the obvious choice for direct coupling. The amplifier should be

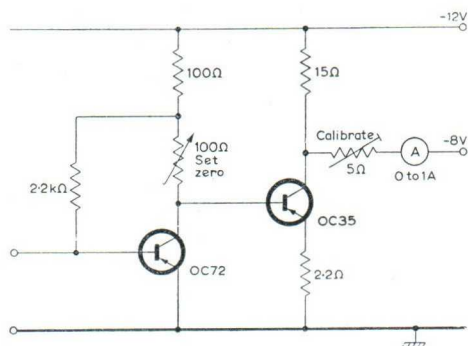


Fig. 1—Directly coupled amplifier using germanium transistors. Supply -12V tapped at -8V . Current drain 1.3A for full-scale output. OC35 passes constant 0.3A .

driven from a reasonably high source impedance ($> 3\text{k}\Omega$) to reduce the V_{be} drift to the same level as leakage-current drift. A set-zero potentiometer is included, and a tap on the supply facilitates a zero-current reading.

Amplifiers of this type might be used as the output stages following a more sensitive amplifier after the signal level has been raised sufficiently compared with the maximum drift.

Amplifier using Silicon Transistors

The leakage current is much lower for silicon than for germanium transistors.

Drift is therefore very low for the amplifier using silicon transistors (Fig. 2), and as the leakage current no longer masks the temperature dependence of V_{be} and α' , these become of greater significance.

A source resistance of at least $300\text{k}\Omega$ should be provided for the amplifier for the V_{be} drift to be negligible. It is likewise necessary to provide a high source resistance for the second stage. To use the requisite high load resistance ($= 33\text{k}\Omega$) in the first stage, a supply voltage of at least 24V is required.

The change of α' with temperature is no longer negligible. Negative feedback is incorporated which reduces the drift caused by variations in α' (and also the gain) by a factor of about 8. The feedback is provided over three stages, not merely the first stage, and is thus more effective because of the higher gain over the three stages. The feedback also stabilises the operating points of all three stages, not the first

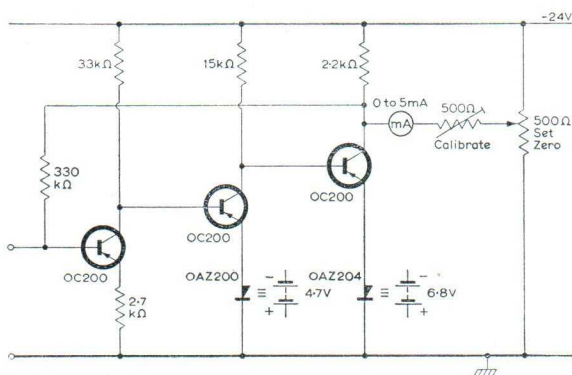


Fig. 2—Directly coupled amplifier using silicon transistors. Supply -24V at 60mA . Zener diodes equivalent to voltages shown.

stage alone. The feedback resistance should not be made so low that changes in V_{be} begin to introduce appreciable drift.

Direct coupling brings a further difficulty. The feedback resistance must establish the operating point and also provide the feedback. In this circuit, the resistance required for bias was considerably higher than that required for feedback. An emitter resistor is therefore included in the first stage.

The resistance in the emitter circuit raises the base potential to a few volts, so that a reverse current flows through the source. In some circuits this may not be permissible. Another arrangement would be to connect the source between base and emitter.

The effective gain of the amplifier is 1000 with feedback, but depends on the load resistance, which can be varied to adjust the gain.

Operating the transistor at very low voltages, comparable with V_{be} , would make the gain sensitive to changes in the V_{be} of the next stage. Hence Zener diodes are used in the emitters of the second and third stages to ensure that sufficient voltage is applied to the collectors of the first two stages, so avoiding the loss of gain which would occur if emitter resistors were used.

Long-tailed-pair Amplifier using Germanium Transistors

Both the amplifiers described so far require a high or relatively high source resistance. For amplifying signals from low-resistance sources, some method must be found of eliminating drift introduced by V_{be} ($\sim -2\text{mV}/^\circ\text{C}$). Because this drift is very nearly identical for any two

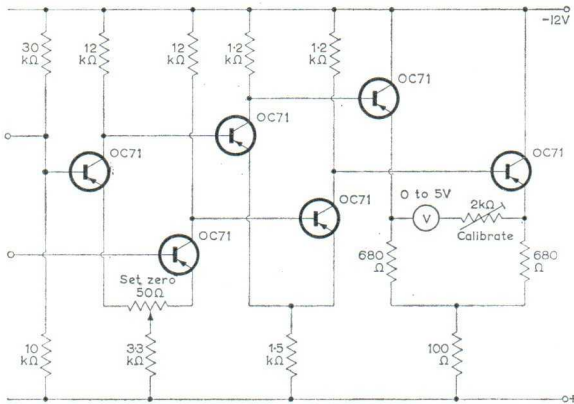


Fig. 3—Long-tailed-pair amplifier using germanium transistors. Supply -12V at 20mA .

transistors of any given type (say, two OC71), it can be cancelled by using another transistor of the same type to balance out the change.

The resulting circuit is the long-tailed-pair amplifier, an example of which is shown in Fig. 3. In this circuit, the error arising from differences in leakage current can be made small by using a low source resistance ($< 100\Omega$). The set-zero resistance in the emitter circuit introduces some unbalanced feedback and is therefore kept as low as possible. For reasonable stability, the voltage drop across the shared emitter resistors has to be much greater than the change in V_{be} .

The long-tailed-pair circuit is effectively a method of applying differential negative feedback so that in-phase changes produce no output, an output only occurring with push-pull signals.

Long-tailed-pair Amplifier using Silicon Transistors

The leakage current of silicon transistors is very small, and does not merit consideration over the usual ambient-temperature range up to 35°C (95°F). Differences in V_{be} and α' can introduce drift, however.

Any source resistance can be used, since it is not limited, as when using germanium transistors, by the need to eliminate differences in leakage current. The biasing arrangement is therefore changed to that shown in Fig. 4. The d.c. conditions in the first two transistors are

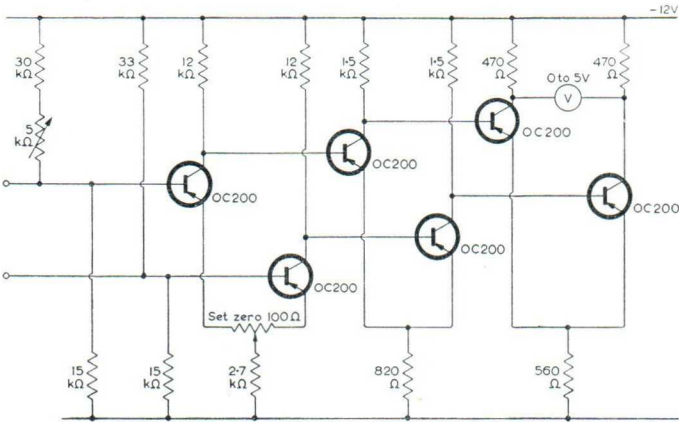


Fig. 4—Long-tailed-pair amplifier using silicon transistors. Supply $-12V$ at 20mA.

set by two potential dividers, which do not significantly shunt the input resistances, although they have the usual stabilising effect on the operating point. The zero under open-circuit conditions is adjusted by varying one of the bias resistors. The zero adjustment under short-circuit conditions is made as before by setting the potentiometer in the emitter circuit.

The drift with open-circuit input for this amplifier is the same as for the one using germanium transistors with a restricted source resistance.

Temperature Control

Changes of temperature are mainly responsible for drift. Drift can be very much reduced by using a circuit such as that shown in Fig. 5 to maintain the temperature of the amplifier at a constant level. The amplifier is mounted in a metal block drilled to take the transistors. A closed-loop feedback system controls a power transistor, which supplies the current for heating a coil wound uniformly around the metal block. The sensing element used to provide the error signal is

an a.f. transistor (the first OC71) mounted in the block. The leakage current I'_{co} of this transistor changes markedly with temperature, and the changes, after further amplification, control the current through the heating coil.

CHOPPER-TYPE AMPLIFIERS

For applications where drift requirements are more stringent, direct voltages or currents can be converted to proportional a.c. signals: this technique is known as chopping. The first amplifiers built on this

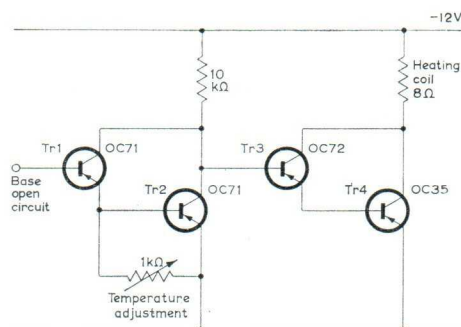


Fig. 5—Simple temperature-control circuit. Supply -12V at 0.3mA .

principle made use of mechanical switches in which a vibrating reed interrupted the d.c. signal periodically.

The problems associated with chopper-type d.c. amplifiers are centred on the design of the actual chopper, differing slightly for mechanical and semiconductor choppers.

Mechanical Choppers

The design of mechanical choppers will not be described here. The maximum practical speed of operation is at present about 400 to 500c/s.

Semiconductor Choppers

The speed of operation of semiconductor choppers is appreciably higher than for mechanical types, the limit being set by minority-carrier storage. Chopping frequencies of about 1000 to 1500c/s are commonly used at present. The higher speed makes possible a wider amplifier bandwidth.

Semiconductor choppers are still somewhat inferior to the best mechanical choppers as regards zero drift. The advantages of rugged-

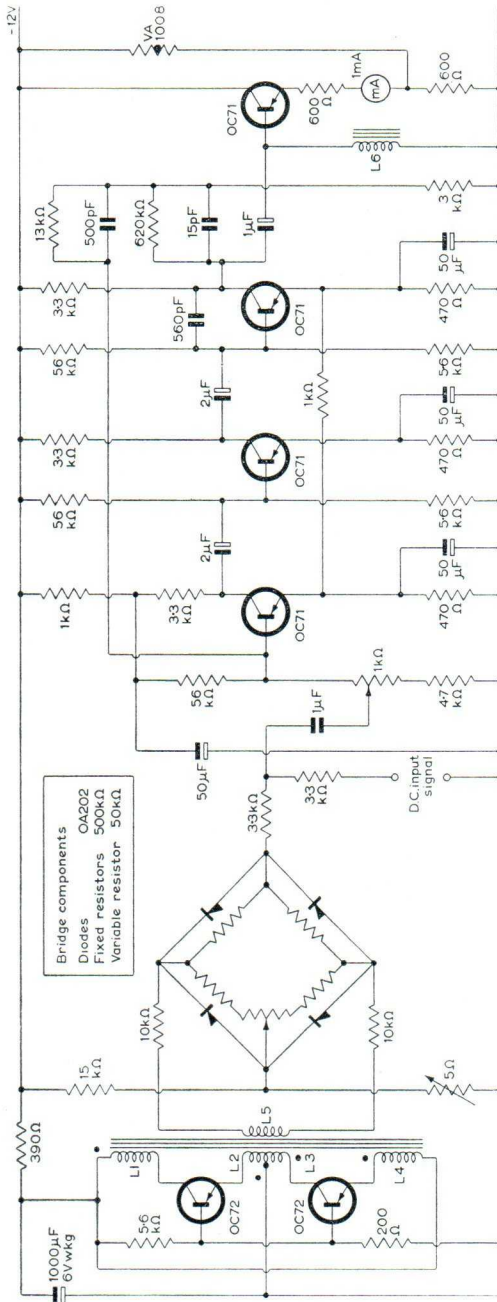
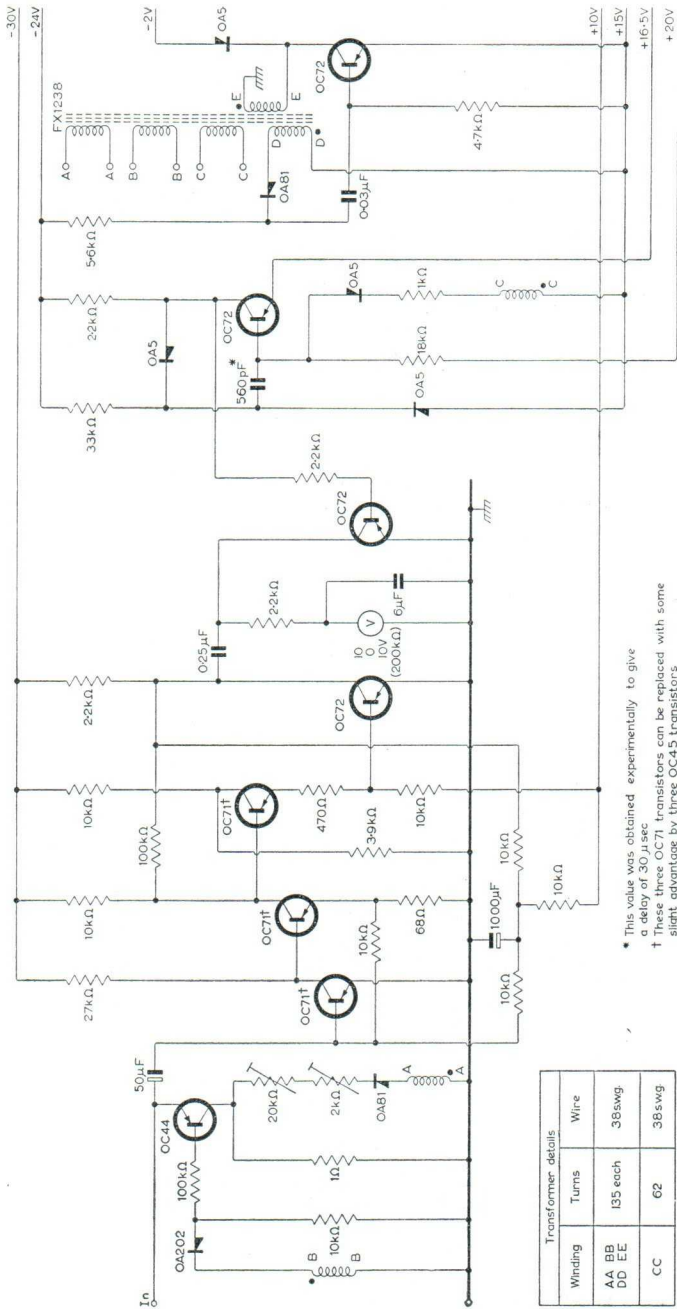


Fig. 6—Chopper-type amplifier with silicon-diode bridge. Supply —12V at 30mA.

D.C. AMPLIFIERS



* This value was obtained experimentally to give a delay of 30 μsec.
† These three OC71 transistors can be replaced with some slight advantage by three OC45 transistors.

Transformer details	
Winding	Turns
AA BB	135 each
DD EE	62
CC	62

Fig. 7—Single-transistor chopper and amplifier. Supplies are: -30V at 40mA; -2V; +15V at 30mA; +16.5V at 15mA; and +20V at 0.2mA. Use dropper resistors and capacitors to supply -24V from -30V and +10V from +20V.

ness and long life, however, make the former superior for continuous operation and for mobile equipment.

Silicon-diode Chopper

The amplifier shown in Fig. 6 incorporates a bridge chopper which uses silicon diodes. OA202 silicon diodes are completely satisfactory in this design.

L1 and L4 are 130 turns, and L2 and L3 are 30 turns, all of 35 s.w.g. Lewmex wire. L5 is 400 turns of 38 s.w.g. Lewmex wire. L6, the detector choke, has an inductance of 3H and a d.c. resistance of 90 Ω . All formers are Ekco type DP10857, and the laminations are Mu-metal 187.

Single-transistor Chopper

Fig. 7 shows an amplifier using a transistor chopper. The stability is significantly superior to that of other purely semiconductor amplifiers and approaches the performance of amplifiers using mechanical choppers. The amplifier was designed to amplify very low currents from high-impedance sources.

In its present form the amplifier is rather susceptible to changes in supply voltages to the driving-waveform circuits. Use of a slightly lower chopping frequency and omission of the delay circuit can overcome this drawback. The d.c. amplifier as a whole would then have reduced bandwidth, which is not always a disadvantage. It should be possible to make a significant improvement to this type of amplifier using silicon alloy-junction transistors such as the OC200 and OC201.

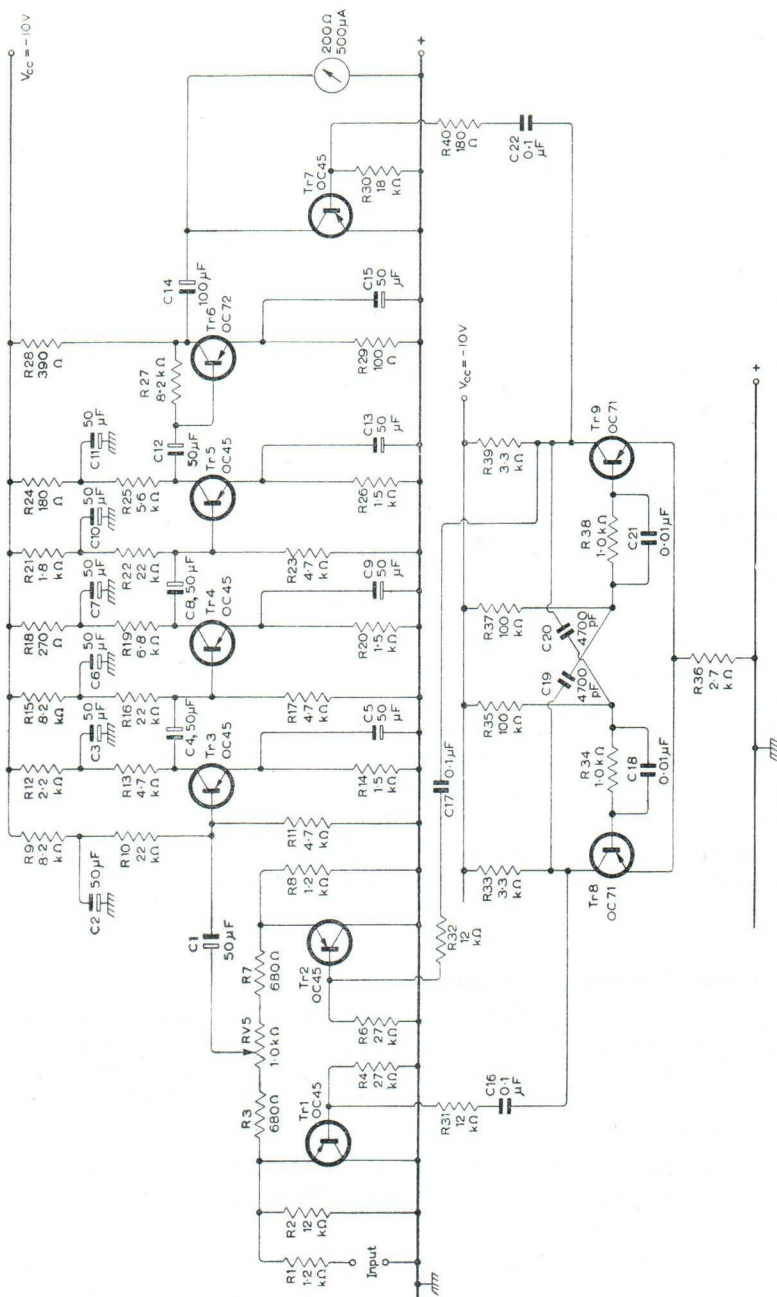
With OC71 transistors in the a.c. amplifier, the output is not identical for identical inputs of opposite polarities. This drawback can be remedied where necessary by using OC45 transistors in the first three stages.

Balanced Transistor Chopper

In the circuit shown in Fig. 8, two transistors (Tr1, Tr2) are used as a balanced chopper. The input is applied to only one of the transistors, while the error voltages of the two transistors appear in opposition. Balance is effected by setting the 1k Ω potentiometer RV5 to give a zero output reading for zero input.

The complete circuit shown in Fig. 8 is intended for amplifying voltages of less than $\pm 200\mu\text{V}$ from low-resistance sources (for example, thermocouples and strain gauges).

The a.c. amplifier consists of Tr3 to Tr6. A synchronous detector Tr7 gives an output in the same phase as the input. The chopper and

Fig. 8—Balanced transistor chopper and amplifier. Supply voltage $-10V$.

the detector are driven by a multivibrator (Tr8, Tr9) operating at a frequency of 1650c/s.

If the output signal is required as a current, the amplifier consists of three OC45 and one OC72 (Tr6), as shown in Fig. 8. The detector is then loaded by an ammeter which gives full-scale deflection at 0.5mA. With this arrangement the meter reads about one micro-amp per microvolt. If a voltage output is preferable, the a.c. amplifier consists

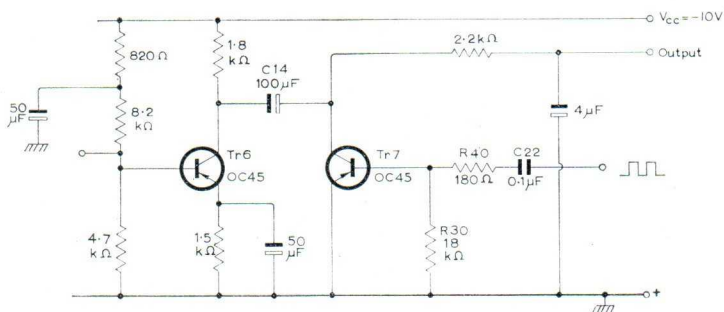


Fig. 9—Alternative final amplifier stage (Tr6) for voltage output from balanced-chopper amplifier.

of four OC45, and the final stage is arranged as shown in Fig. 9. This circuit allows an output of 750mV to be obtained for an input of $400\mu\text{V}$. This voltage may be fed into the input of a valve amplifier, or into an oscilloscope to display slowly varying low voltages.

Mechanical-chopper Amplifier

Fig. 10 is the circuit diagram of a chopper-type amplifier using a mechanical chopper which short-circuits the input periodically. The chopper is a Carpenter 3PK55 polarised relay supplied specially adapted for use as a chopper. It is driven by a simple multivibrator at a frequency of about 400c/s. The resulting square wave is amplified in a transformer-coupled grounded-base amplifier, to which two RC-coupled grounded-emitter stages are added to raise the signal to a suitable level for rectification and for driving a meter.

Very careful attention must be paid to the screening of the chopper and of the input to the amplifier.

COMPARISON OF PERFORMANCE

The performance of the amplifiers described in this chapter is compared in Tables 1 and 2. No claim is made that the amplifiers give the

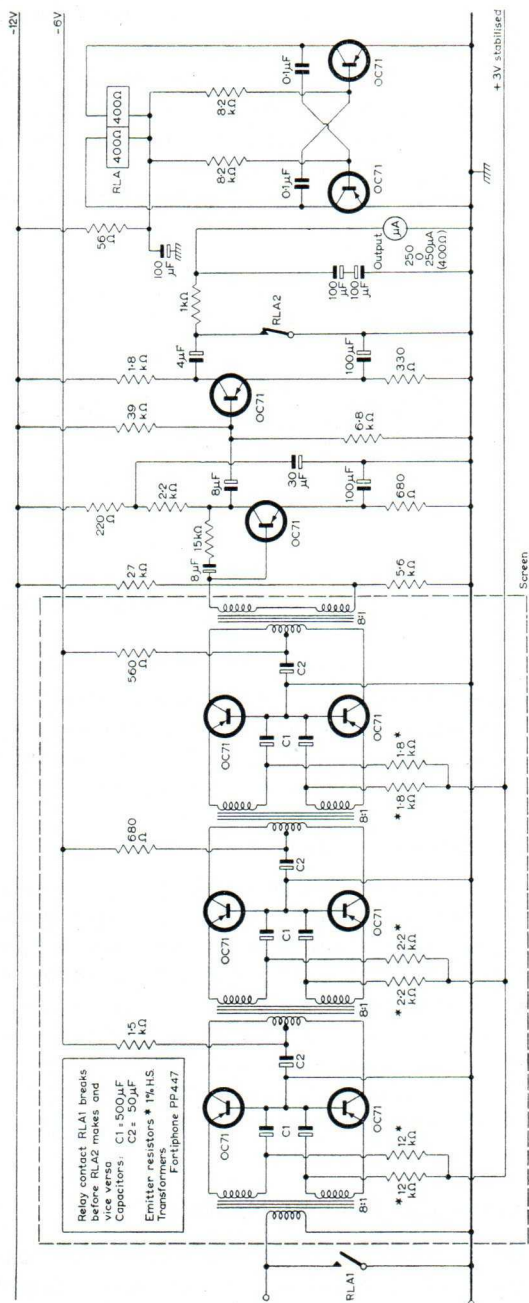


Fig. 10—Mechanical chopper and transformer-coupled amplifier. Supplies are: $-6V$ at $10mA$; $+3V$ at $10mA$; and $-12V$ at $18mA$ for chopper

best possible performance, and therefore component tolerances have not been considered. The figures quoted are typical of what can be achieved using average transistors in the circuits described.

Drift

Drift for the various semiconductor types is quoted in terms of changes in temperature. For the mechanical chopper, drift is essentially a question of mechanical design, and is therefore quoted in μV per hour.

Where chopper types are concerned, the mechanical chopper offers the best performance in terms of zero stability.

The source resistance of a directly coupled amplifier either has to be very low or very high if minimum drift is to be achieved. The choice of d.c. amplifier will be governed to some extent by the source resistance it is proposed to use. For example, for thermocouples, which typically have a source resistance of less than 100Ω , a long-tailed-pair amplifier is the obvious choice.

Silicon transistors offer a great improvement for current-fed amplifiers, where the source resistance is much greater than the transistor input resistance, as the figures in the first two columns of Table 1 show. Silicon types do not offer much advantage where a low-resistance source is specified, as can be seen by comparing the figures in the third and fourth columns for the long-tailed-pair circuit.

Drift for a long-tailed-pair amplifier can be reduced by a factor of about 20 by stabilising the temperature. A similar improvement would be obtained if any of the other directly coupled amplifiers were mounted in the temperature-controlled block. The supply must be correspondingly stabilised to obtain the full advantage of temperature control. Drift in a directly coupled amplifier with temperature control is comparable with that of chopper types.

The circuitry becomes correspondingly more complicated as the drift requirements are made more stringent, consequently the simplest circuit should be chosen which is adequate for the application in mind. If milliamps are to be amplified, it is not necessary to design a circuit with a drift of the order of millimicroamps.

TABLE 1
Performance of Directly Coupled Amplifiers

<i>Circuit</i>	<i>Basic</i>		<i>Long-tailed Pair</i>		
<i>Type of Transistor</i>					<i>Temperature Controlled</i>
	<i>Germanium</i>	<i>Silicon</i>	<i>Germanium</i>	<i>Silicon</i>	<i>Germanium</i>
Gain	500 current	1000 current	500 voltage	500 voltage	500 voltage
Zero drift	4 to 8 $\mu\text{A}/^\circ\text{C}$	0.1 $\mu\text{A}/^\circ\text{C}$	0.1 $\text{mV}/^\circ\text{C}$	0.1 $\text{mV}/^\circ\text{C}$	5 $\mu\text{V}/^\circ\text{C}$
Full-scale output ..	1A	5mA	5V	5V	5V
Specified source resistance	$>3\text{k}\Omega$	$>300\text{k}\Omega$	$<100\Omega$	any	$<100\Omega$
Input resistance	250Ω	$1\text{k}\Omega$	$5\text{k}\Omega$	$5\text{k}\Omega$	$5\text{k}\Omega$
Load resistance	0.1Ω	100Ω	$5\text{k}\Omega$	$5\text{k}\Omega$	$5\text{k}\Omega$

TABLE 2
Performance of Chopper-type Amplifiers

<i>Type of Chopper</i>	<i>Silicon-diode</i>	<i>Single-transistor*</i>	<i>Balanced transistor</i>	<i>Mechanical</i>
Gain	1000 current	10,000 voltage	52dB power	60,000 current
Zero drift	2.5 $\text{m}\mu\text{A}/^\circ\text{C}$	0.5 $\text{m}\mu\text{A}/^\circ\text{C}$	2 $\mu\text{V}/^\circ\text{C}$	2 to 3 $\mu\text{V}/\text{hour}$
Full-scale output	1mA	5V	$500\mu\text{A}$	$200\mu\text{A}$
Specified source resistance	any	any	$<2\text{k}\Omega$	any
Input resistance	$10\text{k}\Omega$	500Ω	$\sim 1\text{k}\Omega$	$15\text{k}\Omega$
Load resistance	150Ω	200Ω	200Ω	400Ω

*Measured for a change in temperature from 20 to 35°C

CHAPTER TWENTY-SIX

D.C./D.C. CONVERTERS

In many of the d.c./d.c. converters which have so far been published, the switching is controlled by increasing the collector current of the transistors. Most of the circuits described in this chapter belong to a family of possible arrangements for controlling the instant of switching by decreasing the base current. All these circuits use two transformers.

Details are given of the design and performance of a high-power push-pull converter capable of 100W output.

PRINCIPLE OF OPERATION

Transistor d.c./d.c. converters compare very favourably with rotary converters, vibrator-transformer-rectifier converters and mechanical methods generally for converting energy from one level to another. Without exception such methods are inefficient at low powers. At higher powers, where their efficiency is fairly good, the mechanical systems still suffer from a number of disadvantages, namely: high initial cost; bulk and weight; maintenance costs; and interference from arcing at the contacts. (Nevertheless, the transistor converter needs to be adequately screened.)

Transistor d.c./d.c. converters perform the same sequence of operations as the vibrator-transformer-rectifier circuit. First, the d.c. from the battery is chopped, using a square-wave oscillator. The chopped input is then stepped up to a higher level. Finally, the stepped-up signal is rectified to convert it back to d.c., and the output is smoothed and delivered to the load.

Usually some bias is provided, at least when first switching on, to start the oscillator. Further components may be incorporated to improve the regulation, and conventional voltage multipliers are used to step up the voltage on the output side.

Long leads to the supply possess appreciable inductance, and the supply will have to be smoothed by a capacitor sited near to the converter.

TWO-TRANSFORMER PUSH-PULL CIRCUIT

The two-transformer converter (as in Fig. 1) overcomes various disadvantages of the usual one-transformer type. The new circuit differs

from the conventional type of push-pull converter, in that a small saturating drive transformer is used to control the switching, and a larger transformer, working linearly, steps up the output to the required value. The essential improvement is that a higher proportion of the transistor peak-current rating can be used with all transistors, and therefore the output power can be increased, while tolerating the full production spreads in the characteristics of the transistors.

On connecting the supplies to the circuit in Fig. 1, one of the transistors (say Tr1) will conduct, because of the unbalance in the circuit, causing its collector voltage to swing (to zero) by very nearly the

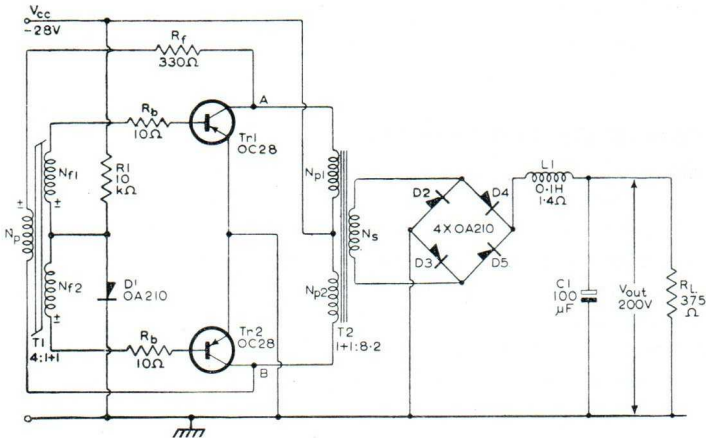


Fig. 1—Practical two-transformer converter

supply voltage. The voltage building up across the primary of the output transformer is applied across the primary of the drive transformer T1 in series with a feedback resistor R_f . The secondary windings are so arranged that Tr2 will be reverse biased and will remain cut off and Tr1 will be held in the bottomed condition.

As soon as the core of T1 reaches saturation, rapidly increasing primary current causes an additional voltage drop across the feedback resistor R_f . This drop reduces the drive; and the collector current of Tr1, which was bottomed, starts to decrease, causing the polarities of the voltages in all the windings to reverse. Tr1 is rapidly driven to cut off and Tr2 switched on. Tr2 continues in this state until the negative saturation of the transformer is reached.

The circuit switches back to the initial state and the cycle is repeated. The oscillation then continues at a frequency determined by the

design of the saturating transformer T1 and by the value of the feedback resistor.

For reliable starting, the transistors are initially biased into conduction by using a resistor and a diode (R1 and D1, Fig. 1). The external base resistors are added to reduce the effect of V_{be} on the operation of the circuit.

The collector current in either of the transistors rises to the load current, plus the magnetising current of the output transformer, plus the feedback current needed to produce the drive. Because the output

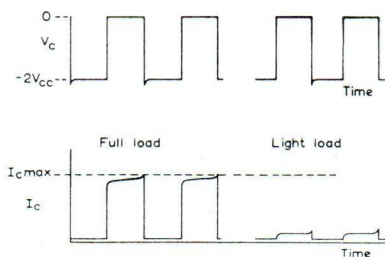


Fig. 2—Collector voltage and current waveforms for purely resistive load

transformer is not allowed to saturate, the magnetising current is only a small fraction of the load current.

The collector voltage and current waveforms, for a purely resistive load, are shown in Fig. 2.

DESIGN CONSIDERATIONS

The design of a converter is normally based on the available supply voltage, the required output voltage, and the output power.

The peak voltage at the collector of either transistor, when cut off, is approximately twice the supply voltage. The supply voltage should therefore always be less than half the collector breakdown voltage at the peak value of inductive current.

The design of the transformers is not critical, and a wide choice of operating frequency can be tolerated, depending on the required size and weight of the converter and its efficiency.

Drive Transformer

The primary of the drive transformer, in series with a feedback resistor R_f , is connected across the two collectors of the transistors of the converter (Fig. 1).

The peak voltage produced across the two collectors by the primaries

of the output transformer is approximately twice the supply voltage. The voltage applied across the primary of the input transformer, however, depends on the value of the feedback resistor and the required drive current. The value of the feedback resistor, in turn, is a compromise between the requirements of the saturation current of the transformer, the voltage applied across the primary, and the operating frequency of the converter.

The number of turns N_p required for the primary winding is determined as follows. The expression relating the operating frequency (f) and the various transformer and circuit parameters is:

$$f = \frac{V_{in} \cdot 10^8}{4N_p AB_s} \quad \dots(1)$$

where

V_{in} is the voltage applied across the primary

N_p is the number of turns on the primary

B_s is the flux density at saturation in gauss

A is the cross-sectional area in square centimetres.

Another condition for correct operation is that there should be sufficient current available to saturate the core. This condition is given by a commonly used transformer equation:

$$H = \frac{4\pi NI}{10l} \simeq \frac{1.26NI}{l};$$

from which, for saturation:

$$H_s = \frac{1.26N_p I_L}{l} > H_o, \quad \dots(2)$$

where

H_s is the strength of the magnetising field at saturation, expressed in oersteds

H_o is the intrinsic strength of the magnetising field of the material used for the core, expressed in oersteds

I_L is the inductive current in amperes

l is the length of the magnetic path in centimetres.

In Eq. 2, H_s is fixed by the material chosen for the core, and l by the size of the laminations. Therefore a value for N_p can be obtained for a particular inductive current I_L .

This value of N_p can now be substituted in Eq. 1 to obtain the operating frequency. If a different frequency is required, adjustment

of the cross-sectional area is necessary, which entails a change in the number of laminations.

The number of primary turns having been decided, the feedback winding N_f is designed from the expression:

$$N_f = N_p \cdot \frac{V_{in}}{V_f} \quad \dots(3)$$

where V_f , the required feedback voltage, is given by

$$V_f = V_{be} + I_B R_b + V_{D1}, \quad \dots(4)$$

and:

V_{be} is the base-emitter voltage for peak collector current

I_B is the base current required for peak collector current

R_b is the external base resistance

V_{D1} is the forward voltage drop across the starting diode D1.

Output Transformer

The output transformer, T2, is a normal linear transformer and is designed using conventional techniques. The primary windings must have a sufficiently high inductance to keep the required value of magnetising current low. Also, the leakage inductance must be made negligibly small, by using bifilar windings.

The inductance required for each half of the primary can be calculated from

$$L = V_{cc} \cdot \frac{dt}{di_m}, \quad \dots(5)$$

where

V_{cc} is the supply voltage

t is the time of half a cycle

i_m is the magnetising current.

Starting Circuit

The basic circuit arrangement will not necessarily start to oscillate, especially when heavily loaded, because both transistors are initially cut off. A permanent bias is therefore applied, by means of R1 and D1, so that the circuit has a loop gain greater than unity and will always start to oscillate.

To ensure a loop gain greater than unity, the base current I_b (in mA) must be greater than

$$\frac{25n}{(\alpha' R_L' - n R_{bb})}, \quad \dots(6)$$

where

R_L' is the resistive load appearing across the primary winding

R_{bb} is the total base resistance, both internal and external

n is the turns ratio of the feedback winding.

The value of R_1 can be calculated approximately from

$$R_1 = \frac{V_{cc}}{2I_b + I_d}, \quad \dots(7)$$

where I_d is the inverse current of the diode. For silicon diodes I_d is usually small enough to be neglected.

It is possible to use a resistor instead of the starting diode but, if the value of the two starting resistors is high, the drive power will need to be increased substantially. Small values for the resistors increase the current drain and so lower the efficiency.

If a resistor, R_2 , is used instead of the diode D_1 , the value of R_1 can be found from

$$R_1 \simeq \frac{V_{cc}}{2I_b + \frac{V_{be} + I_b R_{bb}}{R_2}}, \quad \dots(8)$$

where V_{be} is the base-emitter voltage for the required base current, I_b .

DESIGN OF PRACTICAL CIRCUIT

Several factors affect the design of a practical two-transformer converter; these factors and the performance of the circuit are now examined in detail.

Operating Frequency

The choice of operating frequency is not very critical and will depend on the efficiency and physical size of the converter.

Although losses in the transformer cores and transient losses of the transistors increase with operating frequency, the efficiency varies only a few per cent over the frequency range 300 to 1000c/s.

Feedback Resistor

The optimum value of the feedback resistor R_f is found, experimentally, to be that value which will drop about half the available voltage, at the drive current corresponding to the maximum load current.

Increasing R_f causes a greater drop in voltage across it, so that less

voltage is available across the primary of the drive transformer. As inferred from Eq. 1, the operating frequency will decrease.

Decreasing R_f will increase the operating frequency and increase the losses arising (a) from the circuit resistance and (b) in the transformer core, because of the higher magnetising current.

Drive Transformer

Since the required drive is less than one watt, only a small core is needed. A square stack of Telcon HCR alloy laminations (pattern 224) can be used. This material has the following characteristics:

$$B_s = 15,000 \text{ gauss}, \quad H_s = 2 \text{ oersteds}, \quad l = 5.72 \text{ cm.}$$

The cross-sectional area (A) of 50 laminations equals 0.331 square centimetres.

If the magnetising current I_L is 40mA then, from Eq. 2,

$$N_p = \frac{2 \times 5.72}{1.26 \times 40 \times 10^{-3}} = 227.$$

Thus there should be 227 turns on the primary winding.

To evaluate Eq. 1, the values of the primary voltage V_{in} and of the feedback voltage V_f must first be calculated.

The maximum base-emitter voltage $V_{be \text{ max}}$ required for lower-limit OC28 transistors is 1.6V, and the maximum base current $I_{B \text{ max}}$ is 375mA for the maximum peak collector current of 6A. With an external base resistor of 10 Ω , and allowing 1V across the starting diode, the feedback voltage required at 375mA is 6.35V (from Eq. 4).

If a turns ratio of 4:1 is chosen, 57 turns are required for each feedback winding, and the primary current I_p is about 94mA. The voltage developed across the primary under these conditions is

$$\begin{aligned} V_{in} &= n \cdot V_f \\ &= 4 \times 6.35 \\ &= 25.4V. \end{aligned}$$

From Eq. 1 the frequency is now given by

$$\begin{aligned} f &= \frac{25.4 \times 10^8}{4 \times 227 \times 0.331 \times 15 \times 10^3} \\ &= 564\text{c/s.} \end{aligned}$$

The value of the feedback resistor R_f is given by

$$R_f = \frac{2V_{cc} - V_{in}}{I_p} = \frac{(2 \times 28) - 25.4}{94 \times 10^{-3}} = 326\Omega.$$

The nearest preferred value of 330 Ω is used in the practical circuit.

Transistor Spreads

The drive transformer designed in the previous section is intended to drive a circuit containing low- $\bar{\alpha}'$, high- V_{be} transistors.

The performance of the converter with transistors having high $\bar{\alpha}'$ and low V_{be} will be modified very slightly, except that the operating frequency might decrease by a maximum of 14% from the calculated value. The frequency may be adjusted, if required, by extracting a few laminations from the core of the drive transformer. With the components shown in the circuit in Fig. 1, and for typical transistors, the operating frequency is about 500c/s.

For maximum spreads in transistor characteristics, the change in output voltage, output power and efficiency will be less than 3%.

Output Transformer

If the magnetising current of the output transformer is to rise to 400mA during the half-cycle time t (equal to 1msec), the inductance required for each half of the primary, as given by Eq. 5, is

$$L = 28 \times \frac{1 \times 10^{-3}}{400 \times 10^{-3}} = 70\text{mH}.$$

To avoid excessive loss of power, the resistance of each primary winding should be less than 0.2Ω .

The peak collector current is the sum of the magnetising, feedback and load currents. Therefore it would appear that the circuit can operate up to 5.5A load current, provided the two halves of the circuit and the transistors are identical. In practice, because of slight unbalance in the circuit and the fact that the transistors are not matched, the out-of-balance current through the output transformer causes some premagnetisation of the core. As a result the collector current of one of the transistors will rise to a higher peak value than the other. If the circuit has been designed for operation up to the maximum ratings of the transistors, the peak collector current can thus exceed the safe value, if the circuit is not modified.

The unbalance of the circuit can be reduced by using bifilar windings both for the primary of the output transformer and for the secondaries of the drive transformer.

Spreads in $\bar{\alpha}'$ are more difficult to deal with; the best method for obtaining balanced collector currents is to use a matched pair of transistors. It would then be possible to operate the circuit up to the full theoretical value of load current, 5.5A, with a consequent increase in output power of about 20%. An external base resistor can be used to reduce the effect of spreads in V_{be} .

If no precautions are taken to avoid the unbalance, the load current

must be limited to 4.5A; and the peak collector current must not exceed 5A, including the feedback current and the magnetising current of the output transformer. This allows for up to 1A of out-of-balance current plus surges from the smoothing system.

Even with these limitations, it is possible to obtain an output power of 100W with a 28V supply, with the additional advantage of using transistors with full spreads.

With matched transistors and a purely resistive load, 130W output can be obtained at about 90% efficiency.

Starting Circuit

On full load, the reflected load resistance is approximately 5.6Ω . With a feedback turns ratio of 2, a minimum low-current α' of 20, and R_{bb}

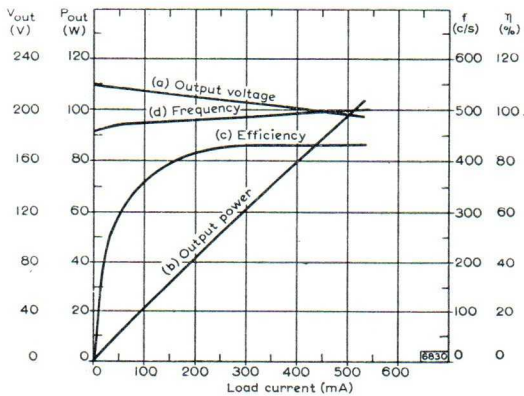


Fig. 3—Effect on operation of varying the load current

equal to 35.6Ω , the minimum base current required for oscillation is (Eq. 6) given by

$$I_b > \frac{25 \times 2}{20 \times 5.6 - 2 \times 35.6} \\ > 1.23 \text{ mA.}$$

DIODE STARTING

The value of R_1 , using a diode to initiate oscillation, is (from Eq. 7)

$$R_1 = \frac{28}{2 \times 1.23 \times 10^{-3} + 0.02 \times 10^{-3}} = 11.3 \text{ k}\Omega.$$

The nearest lower preferred value of $10 \text{ k}\Omega$ should be used.

RESISTOR STARTING

With R_2 equal to 3.3Ω , a value of $3.3k\Omega$ was found to be adequate for R_1 for reliable starting. These values were found to be satisfactory for both choke input and purely resistive loads. Higher values for R_1 might prevent starting with large capacitive loads, and lower values would reduce the efficiency.

Performance

The performance of the converter shown in Fig. 1 is as follows:

Supply voltage ..	..	..	..	..	..	..	28V
Supply current ..	..	..	..	..	..	..	4.3A
Input power ..	..	..	..	..	..	..	120W
Frequency ..	..	..	..	..	..	..	500c/s
Ripple voltage ..	..	..	..	..	..	..	220mV
Output voltage ..	..	..	..	..	..	..	195V
Output current ..	..	..	..	..	..	..	529mA
Output power ..	..	..	..	..	..	..	103W
Efficiency ..	..	..	..	..	..	..	86%

Over the range of temperature from -10°C to $+80^\circ\text{C}$ the performance is hardly affected. Reducing the copper losses in the output

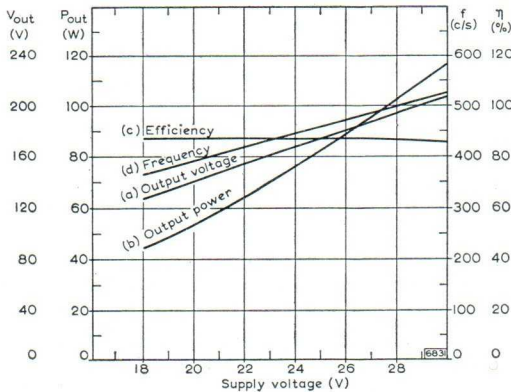


Fig. 4—Effect on operation of varying the supply voltage

transformer can lead to a higher output and an efficiency of about 90%.

The effects of varying the load current and supply voltage over a wide range are shown graphically in Figs. 3 and 4.

HEAT SINKS

The heat sinks for the practical circuit (Fig. 1) can be made of commercial copper $\frac{1}{8}$ in. thick of area $3\frac{3}{4} \times 14$ in. (conveniently folded), that

is, about 3.2mm thick by 95×355 mm. The surface should be blackened to assist cooling by radiation. An equivalent heat sink in aluminium could also be used. The thermal resistance of the heat sink, θ_h , should be about 2°C/W . The maximum ambient temperature at which the converter will operate satisfactorily is then approximately equal to 80°C for a dissipation of 3W per transistor.

CAPACITIVE LOAD

Fig. 5 shows the collector voltage and current waveforms of the two-transformer converter working under full load conditions. The collector current waveform for a purely resistive load is shown in Fig. 5(b).

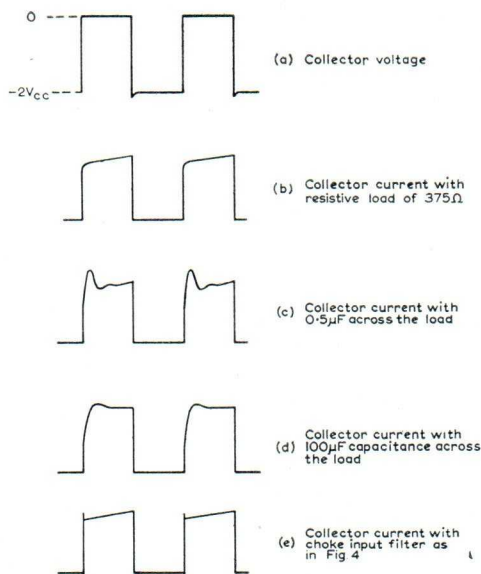


Fig. 5—Collector voltage and current waveforms for various load conditions

With a small capacitance across the load, the output transformer starts to ring. As a result the collector current (c) rises to a higher peak value. If the capacitance is much higher, the oscillation is damped and the collector current does not rise to such a high value (d). On no account must the peak-current rating of the transistors be exceeded.

The disadvantage of a large capacitive load is that it can affect starting when it initially short circuits the load. However, a surge limiting resistor can be connected in series with the load, being progressively short-circuited as the converter is switched on.

So that the converter can operate satisfactorily with a large capacitance across the output, it is necessary either to reduce the load current or, much the better solution, to use a resistive or choke input filter. The collector-current waveform when the latter is used is shown in Fig. 5(e). The spikes at the beginning of the waveform are caused by the inductance of the transformer and choke, and must not exceed the peak-current rating.

PERFORMANCE WITH RESISTIVE STARTING

Satisfactory operation of the converter (Fig. 1), with resistive starting, can be obtained by reducing the value of R_1 to $3.3\text{k}\Omega$ and replacing the diode D_1 by a resistor, R_2 , of 3.3Ω . Performance figures for this modified circuit are given below:

Supply voltage	28V
Supply current	4.36A
Input power	122W
Frequency	510c/s
Output voltage	193V
Output current	526mA
Output power	101W
Efficiency	83%

Voltage Doubler

A voltage doubler is often required instead of a bridge rectifier. A

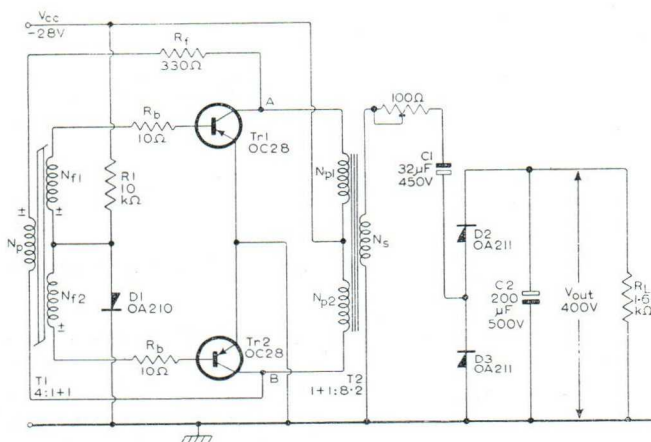


Fig. 6—Voltage-doubler output

suitable circuit is given in Fig. 6. Results of measurements carried out on this circuit are displayed graphically in Fig. 7.

The voltage doubler presents a large capacitive load, therefore a progressively short-circuited resistor is recommended to prevent large peaks of charging current from appearing immediately after switching

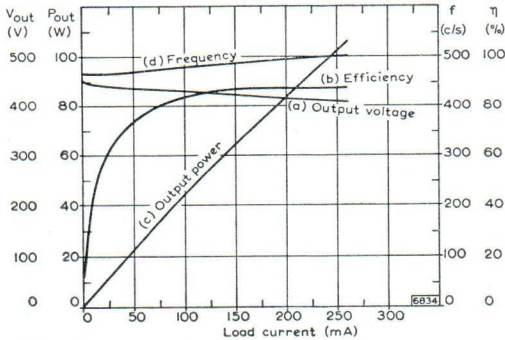


Fig. 7—Effect of varying the load current on operation of voltage-doubler circuit

on the supplies. This resistor can be in series either with the supply voltage or with the output-transformer connection and C1 (see Fig. 6).

Summary of Transformer Details

SATURATING DRIVE TRANSFORMER T1

Core material	H.C.R. alloy (Telegraph Construction and Maintenance Co. Ltd.) pattern 224, 50 laminations
Bobbin	Insulated Components and Materials Ltd. 187A
Primary winding	227 turns of 34 s.w.g. enamelled copper wire
Secondary winding	57+57 turns (bifilar winding) of 30 s.w.g. enamelled copper wire

LINEAR OUTPUT TRANSFORMER T2

Primary windings	Inductance = 70mH/winding Resistance < 0.2Ω/winding (bifilar winding)
Secondary winding	Resistance < 15Ω
Turns ratio	1+1:8.2

Further Modification

A modified circuit is shown in Fig. 8. Although the collectors can be connected to the same heat sink, or directly to chassis in equipment having the negative side earthed, the circuit operates as a push-pull common-emitter amplifier with the input applied between base and emitter. Fig. 8 is a redrawn version of Fig. 1, with a resistor R2 in place of D1. The main difference between the two circuits is that the collector and emitter connections are interchanged; a separate starting circuit is used for each transistor. In this arrangement a diode must not be used in place of R2, because there would be no means of

diverting base current and the transistor would never be cut off.

The performance of this modified circuit is almost identical to that of the circuit in Fig. 1, with resistive starting, except that the efficiency is one or two per cent lower because of the additional current drain

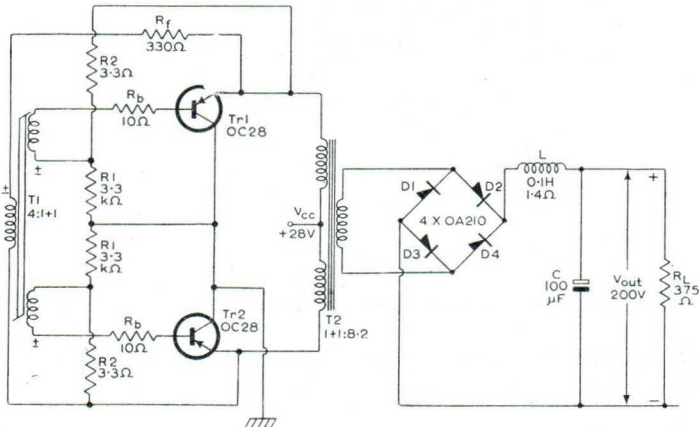


Fig. 8—Further modification of circuit shown in Fig. 1. The emitter and collector connections are reversed and separate starting resistors provided for each transistor.

arising from the separate biasing arrangements. The performance is as follows:

Supply voltage ..	..	..	..	..	..	..	28V
Supply current ..	..	..	..	..	..	..	4.4A
Input power ..	..	..	..	..	..	..	123W
Frequency ..	..	..	..	..	..	..	514c/s
Output voltage ..	..	..	..	..	..	..	193V
Output current ..	..	..	..	..	..	..	522mA
Output power ..	..	..	..	..	..	..	100W
Efficiency ..	..	..	..	..	..	..	82%

BRIDGE CIRCUIT

In push-pull d.c. converters, the peak voltage applied to the transistors in the cut-off state is twice the supply voltage. The supply voltage must be restricted to half the allowable peak collector voltage for the transistors.

It is often desirable to operate the converters from a higher supply voltage than the simple push-pull circuit will allow. A bridge circuit of four transistors can then be used.

A practical circuit with a suitable starting arrangement is shown in Fig. 9.

The required drive is applied by a small saturating transformer T1, in conjunction with a feedback resistor R_f . T2 has a linear characteristic and is used to step up the voltage to the value required for the load.

Diagonally opposite transistors (Tr1 and Tr3, or Tr2 and Tr4) conduct together. Thus, when Tr1 and Tr3 are 'bottomed', the supply voltage will appear across Tr2 and Tr4, which are cut off. Therefore the voltage across any transistor will never exceed the supply voltage V_{cc} , and this converter can be used with a supply voltage of twice the value allowed for any other push-pull arrangement.

This converter, like the two-transistor versions, is suitable for variable loads. This is because the collector current in any transistor does not rise to the peak value determined by the drive but, as in the push-pull

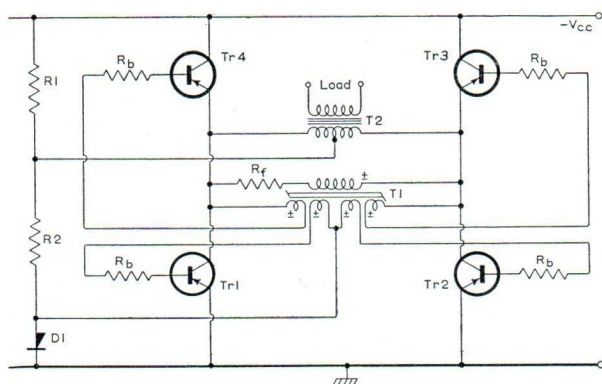


Fig. 9—Practical arrangement of bridge converter

circuit already described, to a value equal to the load current plus the magnetising current of the output transformer and the feedback current.

This converter, also, has good regulation. It is economical to construct, because it uses only a small saturating transformer with relatively inexpensive core material. The output transformer, being conventional, is relatively cheap for the output power obtained.

OC28 transistors can be used in the suggested circuit with a 56V supply, and it is possible to obtain output powers of up to 200W, with an overall efficiency greater than 80%. The same transformer-design procedure can be adopted as for the other circuits. However, the circuit designer must still ensure that the ratings of the transistors are never exceeded.

LOW-POWER CONVERTERS

A single-transformer converter with capacitive-resistive timing is

shown in Fig. 10. This should be particularly suitable for low powers. It operates on the same principle as the other circuits, the instant of switching being controlled by decreasing the base currents rather than by increasing the collector currents.

A normal linear output transformer can be used. The circuit is complete as shown, except that the usual biasing arrangements must be added. Each transistor should have separate bias, and this can be

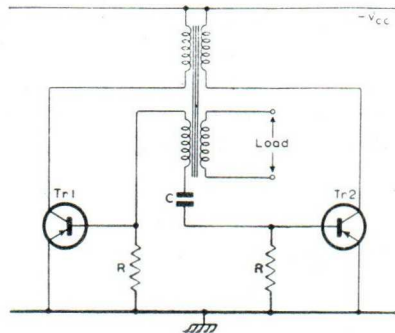


Fig. 10—Capacitive-resistive timing suitable for low-power converter

provided by feeding the base from a potential divider connected across the supply. As the timing resistors R form the lower branches of the potential dividers, only two extra resistors are required, connected between the negative supply line and the base of each transistor.

The drive is arranged such that the transistors operate in the bottomed or cut-off condition, so generating a square-wave output.

In a circuit as shown in Fig. 10, OC84 or OC83 could be expected to give an output of about 5W from a 12V supply. For even lower output powers a one-transistor circuit may be suitable, but here again switching will have to be controlled by the base current rather than by the collector current.

CONCLUSION

The circuits described have distinct advantages over the one-transformer arrangements using a saturating transformer. Briefly these advantages are:

- (a) improved performance under varying load conditions with reduced stress on the transistors, even though operation up to the maximum ratings of the transistors is possible
- (b) transformer design is less critical, so permitting the use of one small saturating transformer and a larger, linear output transformer. This is a cheaper solution than using one large saturating transformer.

CHAPTER TWENTY-SEVEN

PROTECTION CIRCUIT FOR STABILISED D.C. POWER SUPPLY

Where high-power transistors are used in non-mobile equipment, the supplies can most conveniently be taken from the mains. A mains-operated stabilised d.c. power unit is required, which is capable of providing low voltages at currents of the order of a few amps. This chapter is concerned only with the basic principles of stabilised d.c. power supplies, and more particularly with the provision of a protection circuit. Although practical values are given, the final circuit will depend very much on particular requirements.

PRINCIPLE OF OPERATION

The simplest voltage stabiliser is the emitter follower with the base connected to some reference potential, as in Fig. 1(a). In principle the

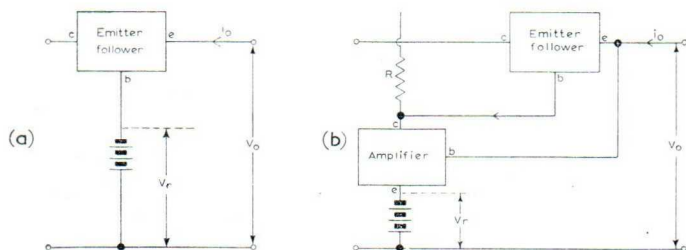


Fig. 1—Simple form of stabiliser (a) and application of d.c. feedback (b)

reference voltage may be taken from a standard cell or a Zener diode, or from a line stabilised by a (gas-discharge) voltage-stabiliser tube by means of a potential divider. The emitter follower divides the output impedance of the supply by approximately $1 + \alpha_o$. When a lower output impedance is required, for better regulation or to prevent feedback through the supply, an amplifier can be inserted before the emitter follower and d.c. feedback applied, as shown in Fig. 1(b).

The 'black boxes' of Fig. 1 may each represent one or more transistors. Referring to the one-amp supply unit of Fig. 2, the amplifier

POWER-SUPPLY PROTECTION CIRCUIT

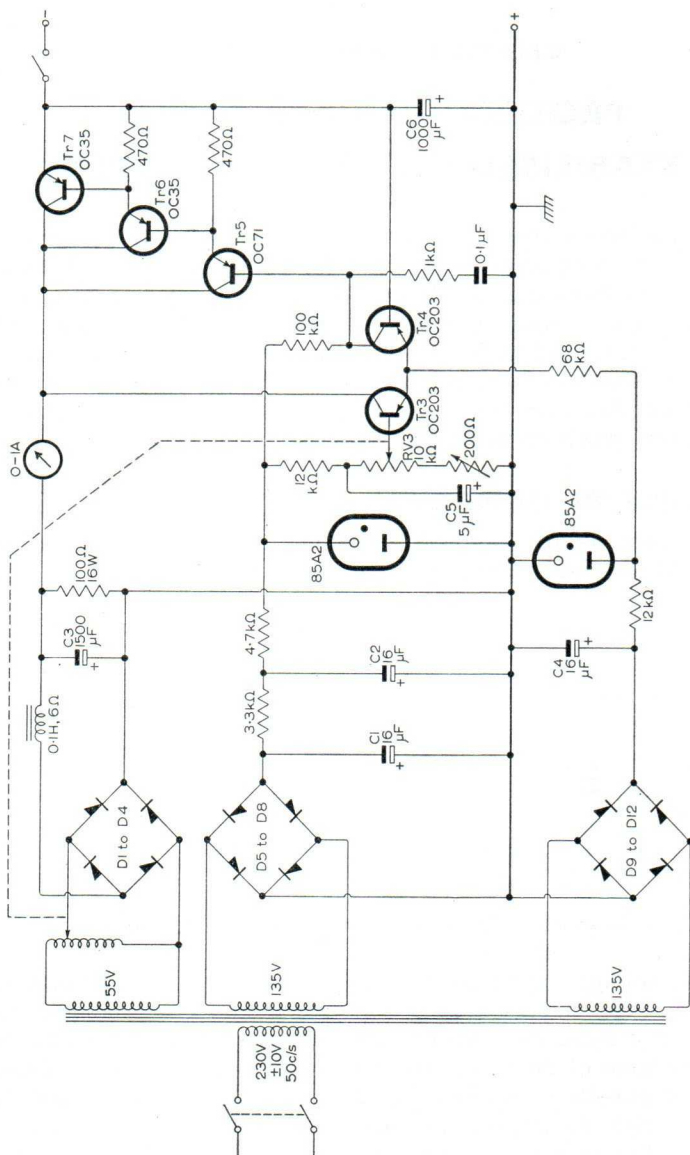


Fig. 2—Stabilised power supply for 1A

is the long-tailed pair (Tr3 and Tr4) and a compound emitter follower is used (Tr5 to Tr7), consisting of three transistors connected in cascade. The latter is no more than a device for effecting a greater reduction in output impedance.

The long-tailed-pair amplifier balances out the variations in base-emitter voltage with temperature. The effect of the variations in I_{co} is made very small by using silicon transistors. The effects of temperature on the germanium transistors in the compound emitter follower are taken care of by the d.c. feedback loop, which extends from the emitter circuits of Tr5, Tr6 and Tr7 to the base of Tr4 and thence from the collector of Tr4 to the base of Tr5.

The reference voltage is applied to the base of Tr3 and is derived from the stabilised $-85V$ line by a potential-divider chain. RV3 is ganged to the variable input transformer which supplies the emitter follower, to limit the power dissipation in the output transistor. A Zenith V544 is suitable. The coupled emitters of the long-tailed pair are run from a stabilised positive line rather than from the earth line, so that control can be maintained right down to zero output voltage. The positive voltage is stabilised by a second 85A2. For high gain, and to render the output voltage independent of the input, the $100k\Omega$ collector load is connected to the $-85V$ stabilised line.

The $1k\Omega$ resistor in series with the $0.1\mu F$ capacitor shunts the output of the long-tailed pair, and reduces the gain of the amplifier at the frequencies ($\sim 10kc/s$) at which the phase shift causes regeneration.

The series transistor Tr7 operates at a high dissipation and requires mounting on a substantial heat sink.

PROTECTION CIRCUIT

High-power transistors often need to be operated near their peak current and voltage ratings, and so can be destroyed by a sudden overload, such as may occur in an experimental circuit under fault conditions.

The overload protection circuit about to be described can switch off the supplies in less than $50\mu sec$. The transistors in the stabilised power supply itself are protected against overload and short-circuit conditions. In addition, the current at which the overload protection operates can be set well below the available output from the power supply, and can be varied over a wide range. Thus experimental circuits, powered by the stabilised supply, can also be protected against their own shortcomings. This feature is especially useful for laboratory power supplies where experimental circuits may need to be protected against such conditions as thermal runaway.

POWER-SUPPLY PROTECTION CIRCUIT

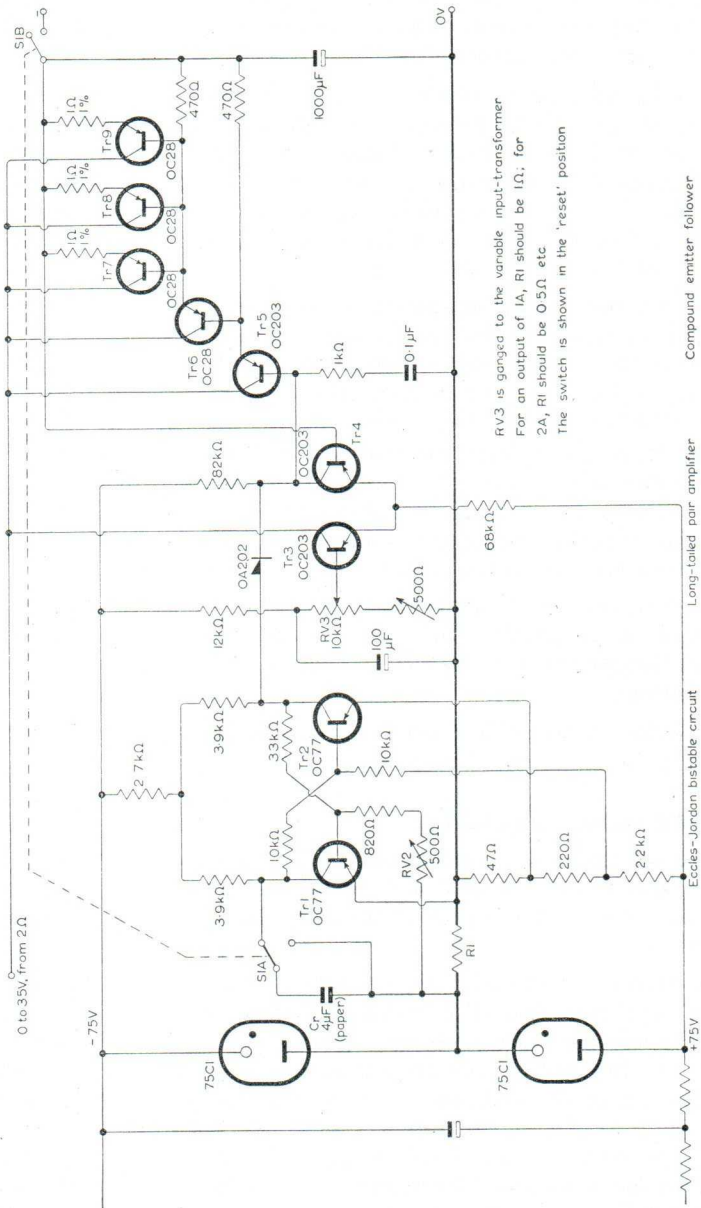


Fig. 3—Protection circuit inserted in stabilised power supply for 3A

Fig. 3 shows how such a protection circuit (Tr1 and Tr2) may be incorporated in a power supply of the type shown in Fig. 2. In fact Fig. 3 differs from the previous one in a number of other respects: thus the stabilised lines are $\pm 75\text{V}$; and in the compound emitter follower, the final step in the cascade consists of three OC28 in parallel, so that the unit can supply output currents of up to 3A.

The protection circuit is an Eccles-Jordan bistable circuit. Switch off is effected by cutting off the emitter follower by means of a small positive voltage applied to the base of Tr5. Tr2 is normally in the cut-off condition. The OA202 diode, which being a silicon diode has a low leakage current, ensures that changes in the leakage current of Tr2 do not affect the stability of the supply. In normal operation the anode of the diode is more negative than the maximum negative voltage on the base of Tr5.

In the emitter circuit of Tr1 is a low resistance (R1) through which the load current flows, and across which is developed a voltage proportional to the load current. The circuit is arranged so that, when the load current exceeds a certain value, the bistable circuit is triggered into its other state. Tr2 bottoms, causing the OA202 to conduct. The base of Tr5 is now at a small positive voltage with respect to the zero-voltage line, and the emitter follower is therefore cut off.

The emitter of Tr2 is connected to a positive voltage (+2V) to allow for the small voltage drop across Tr2 and the OA202. This arrangement ensures that the emitter follower can be definitely cut off. The base of Tr2 is connected to a voltage (+10V) more positive than that at its emitter (+2V), to maintain Tr2 in the cut-off state, when not triggered, over the whole range of ambient temperature (20°C to 45°C) for which the circuit is designed.

The current through the potential divider consisting of the 47 Ω , 220 Ω and 2.2k Ω resistors must be relatively large (40mA) because, when the circuit triggers to its other stable state, the current from the base of Tr5 will flow through the bottomed transistor Tr2 and through the 47 Ω resistor to the common line. If the latter current is comparable with that through the potential divider, it can inhibit the change-over action of the bistable circuit after that circuit has been triggered.

After a fault has been cleared, C_r can be connected momentarily to the collector of Tr1 by a spring-loaded switch, to reset the circuit. The charging current of C_r produces a positive triggering voltage at the base of Tr2, cutting off this transistor. The switch is shown in the reset position in the circuit diagram.

Practical Details

The complete circuit (Fig. 3) is given by way of example only; the effects of component tolerances have not been considered. The power

supply is designed to provide an output current of up to 3A at a stabilised voltage of 27V. The protection circuit operates at the 3A maximum output.

The protection circuit needs a triggering voltage, developed across R1, of approximately 1V. Different values of R1 can be switched in to vary the load current at which triggering occurs. For an output of 1A, R1 should be 1 Ω ; for 2A, R1 should be 0.5 Ω ; and so on. Approximate values can be used for R1, and RV2 adjusted so that triggering occurs at the exact value of required load current.

If there is a fault in the load, the discharge current of the output capacitor (limited only by the resistance of the fault) is added to the current flowing through the output terminals. The combined current pulse could damage the circuit in which the fault has occurred. For this reason, the output capacitor may need to be reduced to 100 μ F.

The protection circuit is inoperative until C_r has been charged. The value of C_r is dependent on the value of the output capacitor, and 4 μ F is the minimum for a 1000 μ F output capacitor. C_r may be reduced if a lower value is used for the output capacitor.

If the stabiliser circuit is triggered 'on' with a fault still present, the protection circuit is inoperative for a few milliseconds; and the dissipation in the series transistors could become excessive, because the full unstabilised voltage acts across the output transistors, causing a large current to flow through them. A double-pole switch is therefore used for the reset operation, so that one output terminal is disconnected at the same time as C_r is connected across Tr1.

Tr6, Tr7, Tr8 and Tr9 need to be adequately mounted to ensure that they are within the junction-temperature rating.

Expected Performance

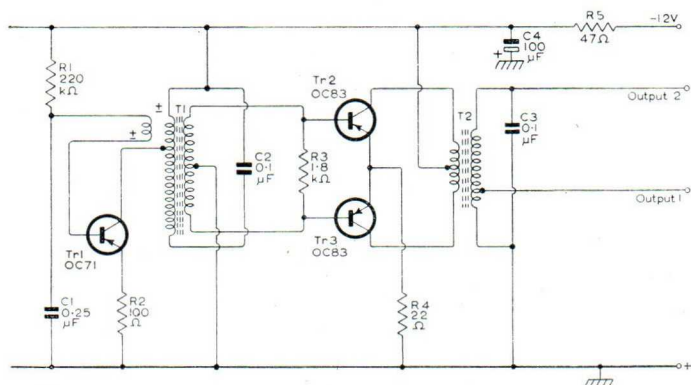
Basically, changing the value of R1 adjusts the triggering current between 20mA and 3A. With R1 equal to 1 Ω , the variable resistor can be adjusted, for example, until the circuit triggers at exactly 1A. The variation of a given triggering current over a temperature range of 20 to 45°C should be less than 10%. The value of triggering current should be nearly constant for all settings of output voltage down to less than 1V.

If the supply is short circuited when set to an output voltage of 25V, and the circuit set to trigger at 1A, the peak current is expected to be 2A and the time of the pulse about 25 μ sec.

CHAPTER TWENTY-EIGHT

1000c/s OSCILLATOR AND TUNED AMPLIFIER FOR H.F. MEASUREMENTS

A number of high-frequency measurements can be made using an r.f. input signal modulated at a low frequency (1000c/s). The r.f. signals at various points in the circuit can be measured by demodulating with a diode detector, and feeding the resultant low-frequency signal into a high-gain tuned amplifier. The output from the amplifier is rectified and the d.c. output fed to a moving-coil indicator meter. This



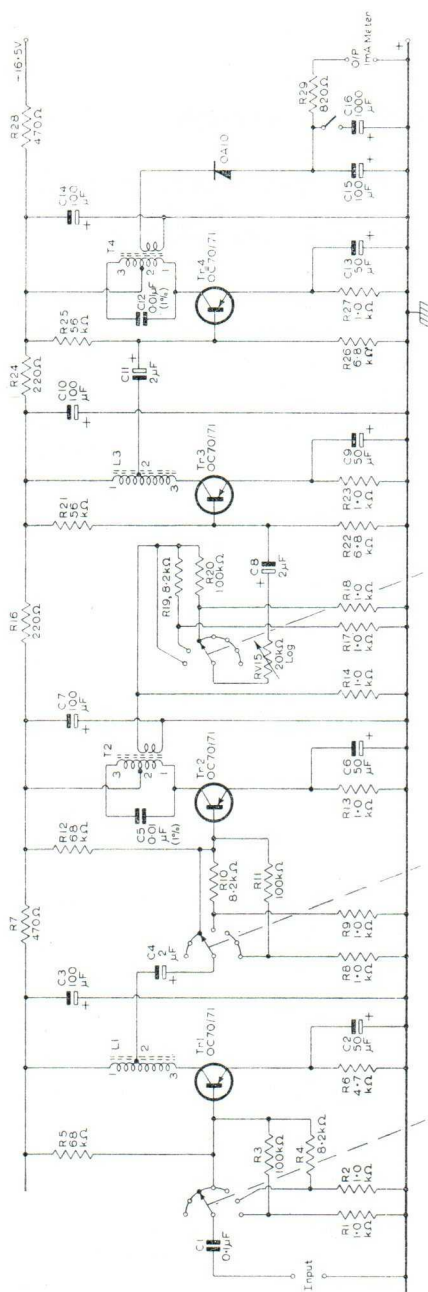
T1 wound on Vinkor former LA2303. Main winding 667 turns tapped at 230 turns from supply end, feedback winding 15 turns, output winding 136 turns centre-tapped, all of 38 s.w.g. enamelled copper wire. **T2** wound on Vinkor former LA2103. Primary 104 turns centre-tapped, secondary 538 turns tapped at 86 turns from chassis, both of 34 s.w.g. enamelled copper wire.

Fig. 1—1000c/s oscillator

method obviates the use of a high-gain r.f. amplifier, which would have to be either wideband or tunable to the r.f. signal to be detected.

1000c/s OSCILLATOR

Most signal generators either have a high input impedance ($\sim 1\text{M}\Omega$) requiring an input of 30 to 50V, or a low input impedance ($\sim 600\Omega$), requiring a relatively high input power ($\sim 100\text{mW}$). In the circuit shown in Fig. 1, these requirements are met by outputs (2) and (1).



L1 and L3 are Ferroxcube LA1 pot cores: 3680 turns of 45 s.w.g. Lewmex type F wire, tapped 726 turns; primary inductance 5H. L2 is Ferroxcube LA13 pot core: primary 1812 turns of 35 s.w.g. Lewmex type F wire tapped at 176 turns; primary inductance $2.54\text{H} \pm 0.5\%$ at 1000c/s for small signals; undamped $Q \approx 150$;

secondary 25 turns of same wire. T4 is Ferroxcube LA1 pot core: primary 2620 turns of 44 s.w.g. Lewmex type F wire tapped at 605 turns; primary inductance $2.54\text{H} \pm 5\%$ at 1000c/s for small signals; secondary 150 turns of same wire.

Fig. 2—1000c/s tuned amplifier

A low-power sinusoidal oscillator drives a class B push-pull amplifier. Class B operation is preferred in the interests of low battery drain. Negative feedback is applied to the push-pull output stage by R4, which restricts the mutual conductance of each transistor, during its conduction period, to 45mA/V.

The oscillator provides a voltage feed to the output transistors which, having a fixed mutual conductance, deliver an output current independent of the transistor input impedance. The input impedance, however, can vary over a very wide range, and to avoid the possibility of the oscillator squegging under certain conditions, the oscillator tuned circuit is damped by R3.

The maximum power dissipation in the output transistors occurs if the output is short-circuited. The transistors then pass peak currents of 110mA, resulting in a dissipation of 300mW per transistor if the maximum supply voltage is 12V. Using OC83 transistors mounted on a heat sink of 7×7cm of 16 s.w.g. aluminium per transistor, this dissipation allows a maximum ambient temperature of 55°C. The output transistors should be bolted down to their heat sinks by means of standard 20mm cooling clips (obtainable from Kimber-Allen Ltd. or distributors).

C3 tunes the secondary of the output transformer, thus reducing the higher-harmonic content, and permitting a lower inductance and hence lower losses.

The output stage can deliver a maximum output power of 120mW (into a resistance of 600Ω) at a nominal supply voltage of 10V. The total-harmonic distortion under all conditions is less than 1%. The output voltage will decrease slightly as the load increases, because of the effect of R4.

At a supply voltage of 10V, the remaining performance figures are as follows. The direct-current consumption is 14mA with no load, 30mA on full load, and 51mA with the output short-circuited. The frequency variation with temperature is less than 200 p.p.m. per °C. The frequency variation with supply voltage is 250 p.p.m. per volt. The output voltage from output (1) is 9.7V r.m.s. with no load, and 8.7V r.m.s. on full load. The output voltage from output (2) is 6.25 times that from output (1).

All resistors should be $\pm 10\%$. R1, R2, R3 and R4 should be $\frac{1}{4}W$, and R5 should be $\frac{1}{2}W$. C1 should be $\pm 20\%$, C2 and C3 should be $\pm 1\%$ with polystyrene dielectric, and C4 should be 12V d.c. wkg.

1000c/s TUNED AMPLIFIER

The 1000c/s tuned amplifier is shown in Fig. 2. To obtain a good signal-noise ratio, the first stage has an emitter current of 200μA. All

the other stages have emitter currents of about 1mA. The primaries of T2 and T4 are tuned to 1000c/s to reduce the bandwidth of the amplifier and thus reduce noise. The working Q of the tuned circuit of T2 is approximately 100. The working Q of the tuned circuit of T4 is about 10, but depends to some extent on the signal level supplied to the detector. The turns ratio of T4 is chosen so that, when the transistor overloads at full supply voltage, the meter current is about 2mA; thus the meter cannot be excessively overloaded. The time-constant of the detector circuit is normally 100 milliseconds, but can be increased, when working at very low levels, to one second, by switching in the 1000 μ F capacitor.

RV15 is a variable attenuator covering 0 to 20dB. Switched attenuation is provided in three sections. The first two 20dB steps of switched attenuation are placed between the second and third stages. In order to prevent overloading of the preceding stage, the second two steps of 20dB attenuation are placed between the first and second stages. The final two steps precede the first stage. The distribution of the attenuation in this way, rather than by putting it all at the input to the amplifier, ensures that the input voltages to the first and second stages are always the maximum possible, resulting in the best possible signal-noise ratio. Each stage is individually decoupled, to prevent oscillation and the feedback of signals through the supply. The stability margin round each possible feedback loop is about 20dB at 1000c/s.

Typical performance for OC71 transistors is as follows: The maximum power gain is 150dB, assuming the detector circuit to be replaced by a resistor. The maximum current gain—the direct output current divided by the r.m.s. input current—is 2×10^6 . The equivalent noise input current is 0.2m μ A. The bandwidth is 10c/s. The centre frequency drift with supply voltage is less than +500 p.p.m. per V. With temperature, the main source of drift is the tuned circuit coil which will cause a frequency shift of less than -300 p.p.m. per °C. The centre-frequency accuracy is $\pm 1\%$. Power consumption is approximately 4mA at full supply voltage (16.5V). The estimated life of an Ever Ready 16.5V grid-bias battery is several hundred hours when used for 4 hours per day, after which the voltage will have dropped to 10V.

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