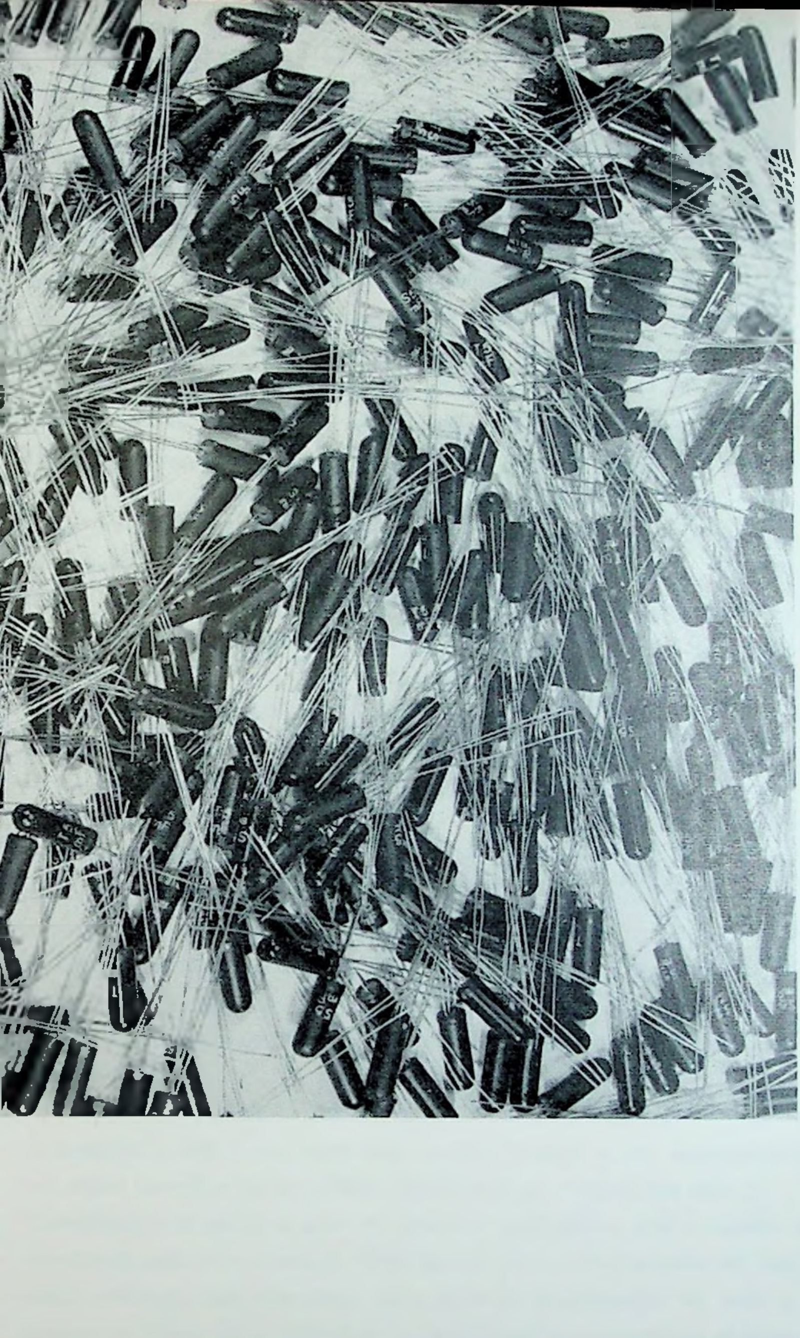
PHILIPS

THE TRANSISTOR

theory and applications

PHILIPS ELECTRON TUBE DIVISION

THE TRANSISTOR



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FOREWORD

Less than ten years ago, Bardeen and Brittain announced the invention of a new electronic component which they called the "transistor", thus combining the terms "transformer" and "resistor". Like electron tubes, transistors amplify electrical energy, but they operate on an entirely different principle.

In fact transistors and electron tubes differ so much and in so many respects that the former might almost be labelled "any resemblance to existing amplifying devices is purely coincidental".

The development of the transistor, in essence an improvement on the crystal detector, aroused intense interest in many quarters, not only amongst scientists, but also amongst manufacturers in all branches of the electronics industry, who began to wonder whether this discovery would revolutionize the design of their products.

Indeed, many people at present using electron tubes are of the opinion that if transistors do not actually supersede thermionic tubes, they will at any rate make possible a whole range of new applications.

That the development of the transistor made, and continues to make, such rapid progress is mainly due to the thoroughness of investigations carried out under the direction of Shockley in the Bell laboratories.

Experience has already shown that in many applications modern transistors can be used instead of electron tubes, and may even provide the answers to many switching problems which cannot be solved satisfactorily with tubes. At the same time, it does not follow that transistors can now be used as a matter of course instead of electron tubes; for one thing the only common function is amplification (usually the gain per stage is smaller in the case of the transistor) and, what is more important, the number of types of transistor is still strictly limited. Moreover, some transistors, notably those for use at high frequencies, are still in the comparatively early stages of development.

This book is an attempt to strike a happy medium by keeping the theoretical material as simple as possible *) and at the same time precise enough to be appreciated by scientific students, for whose benefit a list of suitable literature is provided at the end.

Considerable attention is given to practical applications, and a number of diagrams and constructional notes are provided. They include one or two circuits for high-frequency transistors, since, although such transistors have been in development for only a short time, sufficient progress is being made with them to ensure steady improvement of types and circuits.

As the circuits provided have been specially chosen to enable them to be constructed by those without technical training, many of them are not suitable for mass-production methods of manufacture and do not satisfy all the requirements imposed on such equipment by the Electronics Industry.

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*) To understand the different calculations fully, it is necessary to have some knowledge of algebra.

INTRODUCTION

Transistors are crystal amplifiers used, like certain electron tubes, to amplify electrical energy. In their manufacture semi-conducting materials such as germanium and silicon in the form of single crystals are employed.

Transistors fall into two categories, namely point-contact transistors (no longer very important from the technical point of view) and junction (or layer) transistors.

The theory of transistors is more complex than that of electron tubes. To understand how transistors work it is necessary to know something about the structure of atoms. Accordingly, the physical and chemical aspects of germanium crystals are discussed briefly in the first chapter.

PART 1 – THEORY

CHAPTER I

SEMI-CONDUCTORS

I. 1. ATOMIC STRUCTURE

According to the universally adopted conception, all atoms comprise a nucleus around which electrons revolve like the planets around the sun.
Bohr has shown that the electrons move in orbits or "shells" around the nucleus of the atom and that the shells, conventionally indicated by the letters K, L, M, N, etc., can accommodate only certain numbers of electrons, different for each shell (fig. 1b).

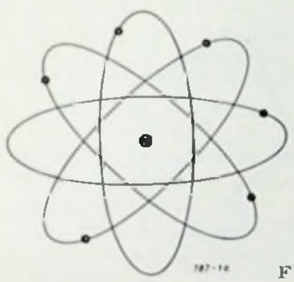

Fig. 1a

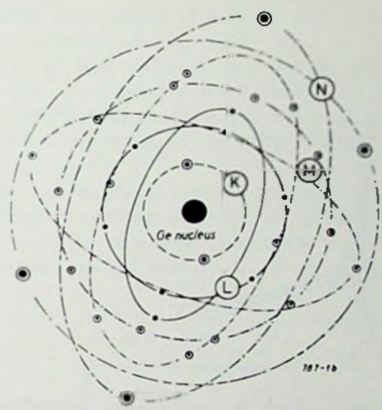

Fig. 1b

It is found that shells K, L, M, N . . . are fully occupied by 2, 8, 18, 32 . . . electrons respectively. From the theory of quantum mechanics it can be shown that shells fully occupied by electrons constitute perfectly symmetrical, and therefore also very stable, arrangements of electrons around the nucleus. Moreover, eight electrons in the outer shell likewise constitute a stable distribution, even when this is not the full quota for the particular shell.
Such stable shells are sometimes referred to as "rare gas shells", since the outer shells of the rare gas atoms, e.g. helium, neon and argon, contain this number of electrons.

I. 2. CHEMICAL BONDS

It also follows from the theory of quantum mechanics that the stable rare gas shells are associated with the minimum energy state; hence all atoms try to acquire such shells. So strong is the compulsion towards this state that atoms "borrow" electrons from their immediate neighbours in order to reach it.
Accordingly, chemical bonds should be regarded as an exchange of electrons between individual atoms whereby each atom tries to acquire a stable rare gas shell.

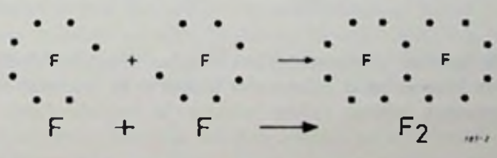

Fig. 2

Although this "borrowing" of electrons can take place in several ways, we are concerned with only one of them, producing what is known as the covalent bond between atoms each of which lacks one or more electrons in its outer shell.
A typical example of the covalent bond is that formed between fluorine atoms (fig. 2). Individual atoms of fluorine have seven electrons in the outer shell, but manage to change the effective number to eight, consistent with the stable state, by sharing one electron with a neighbouring

atom. Hence the bond is formed by the pairing of atoms. Fluorine, then, is univalent, since each atom is capable of binding one other atom. Of course, this is only a simplified explanation of the chemical bond. In reality the electrons are not stationary, but revolve about the nucleus at tremendous speed. Another point to be borne in mind is that of the seven electrons in the outer shell of fluorine atoms, two revolve in elliptical, and five in circular orbits (see fig. 1b, germanium atom). Accordingly, covalent bonds involve distortion of the atomic structure, particularly in complex compounds. Owing to the pairing of electrons, the valency forces (binding forces) act upon them at strictly defined angles; hence the atoms fall into characteristic three-dimensional patterns. For example, in the case of carbon atoms, with four electrons in the L-shell (and therefore with a strong compulsion to capture 4 more) the valency forces assume angles of 109° 28′ with each other, that is to say precisely the angles formed by the median axes of a tetrahedron (solid figure enclosed by four triangles) (fig. 3a). The arrangement of carbon atoms can also be shown as a cube, with the adjacent atoms placed diagonally at the corners (fig. 3b). This is what is known as a cubic lattice.

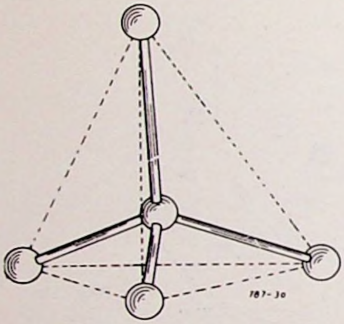

Fig. 3a

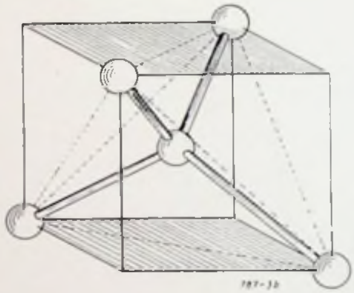

Fig. 3b

In diamond, say, which is really pure crystallized carbon, the cubes are stacked with their edges contiguous (fig. 4). To present a clearer picture, the distance between centres of the atoms is very much exaggerated in all these diagrams. It will be evident from what has been

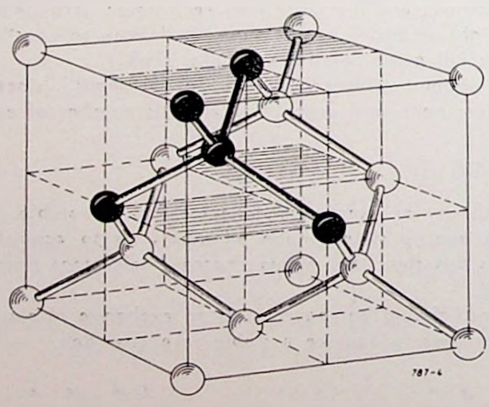

Fig. 4

said so far that in reality partial interpenetration of the electron shells occurs. The structure shown in fig. 4 is what is known as a "diamond lattice". In general terms, "lattice" means a regular pattern or arrangement.

I. 3. CURRENT CONDUCTION IN SEMI-CONDUCTORS

Germanium, No. 4 in Group IV of the periodic system of elements, is closely related to carbon and silicon. The nucleus of the germanium atom is surrounded by 32 electrons, four of them in the outer shell (N-shell). Hence it has four electrons fewer than the rare-gas structure of krypton, which has 36 electrons, eight of them in the N-shell.
Like carbon, then, germanium is tetravalent and crystallizes with a diamond lattice (fig. 4).

At absolute zero (– 273° C) all the valency electrons are in bonds; hence there are no excess electrons to move freely through the crystal and thus to produce an electric current. In other words, the material is, in effect, a perfect insulator. It is interesting to compare germanium with actual diamond, an insulator despite being simply pure carbon. Graphite, on the other hand, which has a different (hexagonal) lattice structure, is a good conductor. Accordingly, it will be evident that the electrical properties of materials depend very much on the arrangement of atoms in the particular crystal lattice.
The conductivity of most materials is affected by temperature. Heat, actually a form of energy, produces molecular vibration, which tends to disturb the arrangement of the atoms. The result is that some of the bonds are broken, and electrons are thus released (higher energy state; see I. 2). Such electrons leave gaps – or positively charged "holes" as they are called – in the crystal lattice ("weakened bonds" is really a more apt description). At the first opportunity, however, the ejected electrons fall back into the nearest hole (recombination); in this way, a random migration of free electrons and holes takes place (fig. 5).
When as many electrons are ejected as are recombined in a given time, a state of "thermal equilibrium" exists; this means that the number of free electrons is constant at a given temperature, but increases with the ambient temperature. Since such electrons are current-carriers, the electrical conductivity of the particular material depends on the temperature. Accordingly, the temperature coefficient of semi-conductors is negative, or in other words, their resistance decreases as the temperature rises.
As we have seen, the positions vacated by electrons become positively charged holes. In effect, since such holes appear as soon as the occupying electron is ejected from the bond, holes and electrons migrate in opposite directions, in very much the same way as with, say, the sliding pieces of certain puzzles (fig. 6). When one of the pieces is moved to the left, the hole shifts to the right.

Fig. 5

Fig. 6

Holes behave like electrons, i.e. like freely moving particles, but are oppositely charged. From the theory of quantum mechanics, the mass of a hole is slightly different from that of an electron; also they do not move quite as readily. This is because electrons are freely moving particles, whereas holes shift only when they happen to capture an electron, which involves a certain amount of delay.
The conductivity of pure germanium is low, and depends on the number of electrons released at room temperature. However, germanium can be rendered very much more conductive by adding certain impurities to it. Such impurities are of two kinds, the one promoting the production of free electrons and the other the production of holes.
The first group includes substances with five valency electrons, e.g. arsenic and vanadium, both of which are pentavalent. It will be evident that, when such atoms are introduced into the diamond lattice of germanium, they each leave one valency electron unable to enter a bond. Accordingly, a very small amount of arsenic added to the germanium is enough to produce a surplus of electrons; arsenic, then, is what is known as a "donor". The mixture of the two is called n-germanium. Donor atoms giving up an electron acquire a positive charge.
In the other group, such trivalent impurities as indium and aluminium create a deficiency of electrons; in other words, mixtures made with them contain more holes than electrons, because electrons released from the germanium are used by the indium (or aluminium) to form the 4th bond with the germanium. Because it captures free electrons, indium is called an "acceptor"; mixed with pure germanium it forms what is known as p-germanium. Through capturing an electron, acceptors acquire a negative charge. The amount of impurity required

is only very small, i.e. of the order of one part in 10^8 (or 1 in 100,000,000). Larger amounts only reduce the desired effect.

In bulk, p-germanium and n-germanium are both electrically neutral, like all metal alloys. This is because donors and acceptors alike are charged oppositely to the electrons and holes produced by them.

CHAPTER II

THE OPERATION OF TRANSISTORS

II. 1. P-N JUNCTIONS *)

Let us now consider what happens when a piece of p-germanium and a piece of n-germanium merge in a continuous crystal lattice, thus producing a p-n joint, or "junction".

The n-germanium yields a portion of its free-electron surplus to the p-germanium, which is short of electrons; similarly, the p-germanium supplies some of its excess holes to the n-germanium, which is short of them. This exchange bears some resemblance to what happens when two chambers containing different gases are in communication with each other; the molecules of the one gas are displaced by those of the other gas, and vice versa, so that after a time the mixture of the two becomes homogeneous throughout both chambers.

The situation produced by joining p-germanium and n-germanium is more complex, however, because the particles involved are electrically charged. Since both pieces of germanium are initially electrically neutral, electrons passing to the p-zone from the n-zone make the former negative with respect to the latter. The potential difference between the two is further increased by the migration of holes from the p-zone to the n-zone, making this positive with respect to the p-zone.

Being now negatively charged, the p-zone then repels electrons arriving from the n-zone, and at the same time the positively charged n-zone repels the holes coming from the p-zone. The result is that, when the two different types of germanium are brought together to form a junction, the exchange of electrons and holes is quickly reduced, and eventually ceases as soon as the potential difference between the two attains a certain value. The electrostatic field produced by this potential difference in the transition zone repels the charge carriers when they approach the junction (fig. 7), and with all free charge carriers thus excluded from the very narrow zone on both sides of the junction, this zone can be considered, more or less, as an insulator (I. 3).

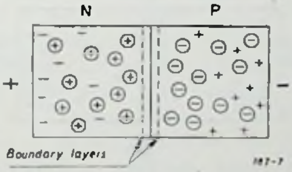

Fig. 7

The transition zone is roughly 1 micron (= 0.001 mm) wide. Such zones can be widened, however, by applying a direct voltage from an outside source to the p-germanium in such a sense that the p-germanium is rendered still more negative with respect to the n-germanium. Such a bias voltage strengthens the electrostatic field in the transition zone, thus driving the free charge carriers further back and therefore widening the insulating layer or "barrier" layer as it is called, and increasing the (barrier)resistance. The barrier resistance increases with the voltage only up to a point, since with very high voltage other effects occur, which ultimately cause breakdown. If the polarity of the applied direct voltage is reversed, making the p-germanium positive with respect to the n-germanium, the p-zone attracts electrons from the n-zone, and the n-zone, then negatively charged, attracts holes from the p-zone.

A very small potential applied in this direction, which is known as the "forward direction", is sufficient to produce a heavy current of electrons and holes. So far, however, we have not taken into account the electrons and holes released by thermal energy — the "intrinsic" charge-carriers as they are called. Their contribution to the total forward current, injected into the germanium through the terminals, is negligible. At the same time, they have the effect of making the insulation of the barrier layer in the reverse direction less than perfect, since thermal energy ejects electrons from bonds near the limits of the transition zone on both sides of the junction, thus producing not only free electrons, but also holes.

*) J. J. Ploos van Amstel, Electronic Application Bulletin No. 8/9, '54.

According to their respective charges, the intrinsic holes in the p-germanium are repelled, and the intrinsic electrons attracted, by the positively charged n-germanium. In this way a leakage current is produced, corresponding to the number of intrinsic electrons and holes, which in turn depends on the temperature. Hence, the barrier resistance is also temperature-dependent.
We have seen, then, that a crystal diode — or junction diode as it is called — is formed by bringing a piece of p-germanium and a piece of n-germanium together. This operation is not however quite as simple as it sounds, since the forming of junctions is not merely a matter of "joining" p and n germanium, but an intricate process requiring much technical knowledge and special equipment.

II. 2. JUNCTION TRANSISTORS

Junction - or layer - transistors comprise three layers of impure germanium in the form of a sandwich. The combination may be either p-n-p or n-p-n.
Other possible combinations, say p-n-p-n, having no equivalents amongst existing electron tubes, offer very interesting possibilities, but it is not within the scope of this book to describe them.
We shall consider mainly p-n-p transistors; n-p-n transistors operate in a similar way.
Fig. 8 shows the basic construction of p-n-p transistors. The different layers are referred to as the emitter, the base and the collector, respectively. Although the emitter and the collector are of similar material, there is a constructional difference between them, which will be defined later. Separate feed-wires, marked e, b and c respectively, are attached to each layer. With the base, the collector and the emitter each form an individual junction diode. The emitter/base diode is biased in the forward direction, and the collector/base diode in the reverse direction, by giving the emitter a positive potential of some tenths of a volt, and the collector a negative potential of some volts, with respect to the base.

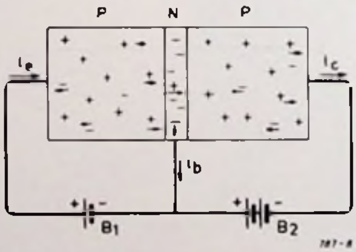

Fig. 8

Briefly, the transistor operates as follows: in the first instance the collector/base diode raises a potential barrier in the same way as the junction diode (see II. 1). The electrostatic charges of collector and base produce another barrier at the junction of the two by causing the collector to repel the excess electrons arriving from the base (of n-germanium), and the base to repel excess holes arriving from the collector. Being negatively charged, the collector tries to attract positive holes from the n-germanium (the base), but fails, of course, because no holes are being formed. As we have seen, however, the emitter/base diode is biased in the forward direction. As in the case of the ordinary junction diode, this means that the positive p-region (the emitter) attracts electrons from the n-region (the base) and the negatively charged base attracts holes from the emitter.
The result is a current of holes to the base, and this is precisely what is wanted by the collector/base diode.
Now, because the base is very thin, most of the incoming holes soon reach the collector/base transition zone, where the negative charge of the collector exerts a strong pull on them, thus producing a current of holes from the emitter to the collector; this current is sustained by two voltages, i.e. the positive emitter voltage and the negative collector voltage obtained from sources outside the transistor. The former governs the size of the hole current and the latter "absorbs" this current as fast as it is supplied by the emitter. Given a sufficiently high collector/

base voltage to dispose of all the holes, the hole current will be governed solely by the emitter/base potential.
Another factor, which we have not previously taken into account, is the recombination of holes and electrons (I. 3). A few of the holes reaching the base from the emitter recombine with the free electrons there, so that fewer holes reach the collector than enter the base from the emitter. Although the number of holes thus eliminated can be cut down considerably by employing a very thin base, constructional considerations naturally impose a minimum limit of thickness.
Apart from the hole current, what is known as "minority current" is produced owing to the presence of free electrons in the base. This portion of the overall emitter current does not form part of the actual collector current, but instead is supplied by the base and sustained by the emitter battery. Hence the collector current is really slightly smaller than the emitter current. The ratio of collector current (i_c) to emitter current (i_c) is expressed in terms of the current amplification factor α, ($\alpha = \dfrac{\Delta i_c}{\Delta i_e}$) which is therefore slightly less than unity. In well-made transistors the value of α is roughly 0.95 to 0.99.
Certain special transistors (e.g. point-contact transistors or p-n-p-n transistors) are so designed as to make the amplification factor greater than unity. Since it is beyond the scope of this book to discuss them in detail, however, those interested should consult the excellent text books on transistors by Shea and Shockley (see bibliography), in which full particulars will be found.
As we have seen, then, almost the whole of the emitter current, actually the emitter current multiplied by the factor α), is absorbed by the collector. To enable all the holes to be absorbed, the collector voltage must be at least $0.3 - 0.5$ V; provided that it satisfies this condition, the collector voltage does not affect the size of the emitter-to-collector hole current very much. In effect, then, the collector, like the conventional pentode valve, has a very high internal resistance (R_i). In transistors, R_i is of the order of 0.25 MΩ to 2 MΩ, so that in principle it is possible to employ a high external load resistance in the collector circuit without affecting the value of I_c appreciably. (In practice, with RC coupling, the resistance should not be taken beyond a certain limit, because it is also necessary to take into account the intrinsic leakage current of the collector/base diode, which may reduce the collector voltage unduly).
Because electrons are steadily withdrawn from the emitter, it is able to inject holes into the base continuously. The electrons pass into, and through, battery B_1 (fig. 8). Accordingly, (positive) current passes through the emitter/base diode in the forward direction. The forward resistance of this diode is very low — in the region of 10—50 Ω — hence a low voltage (that supplied by B_1) is sufficient to produce a relatively heavy current. Since the input circuit of the transistor is also a diode, the input resistance is not constant, but depends on potential V_{bc}. This will be evident from, say, the characteristics of the OC 71, in which $I_b = f(V_{be})$ is a curved line, indicating that input resistance depends on the working point and therefore that the input impedance is low. Slight variations in the voltage between emitter and base produce proportional variations in the emitter current. The collector current, being α times the emitter current, follows such variations exactly. Hence, substantial voltage variations occur in the high collector resistance, without there being any appreciable difference between the collector and emitter currents. In essence, then, we have what may be described as "resistance transformation" (as implicit in the term "transistor"), whereby voltage amplification is obtained notwithstanding that the emitter and collector currents are virtually equal.
Given proper matching of input and output, power amplification of 40 db or more can be procured with ordinary-grade transistors.
As junction transistors actually comprise two junction diodes, they are highly sensitive to temperature variations. Hence transistorized circuits must include some means of counteracting the effects of this temperature dependence (e.g. increased leakage current, displacement of working point).

II. 3. POINT-CONTACT TRANSISTORS

All modern junction transistors were developed from their predecessor, the point-contact transistor. That the point-contact transistor was not given pride of place at the beginning of the book is because the principles of transistor action can be explained more readily with reference to the junction transistor. Point-contact transistors also contain transition zones similar, in effect,

to junctions. In its simplest form, the point-contact transistor is a piece of germanium with two very fine whiskers of tungsten resting on it (fig. 9); the space between the two is approximately 50 microns.
During a process known as forming (involving, amongst other things, passing current pulses from whisker to crystal), impurities are introduced into the germanium, whereby what may be described as junction zones are formed. Improved methods of manufacturing junction transistors have resulted in point-contact transistors taking second place. Nevertheless, they are still used for special purposes, amongst other things in circuits involving negative resistance-characteristics. The reason is that with point-contact transistors in grounded-base circuits, high external resistance

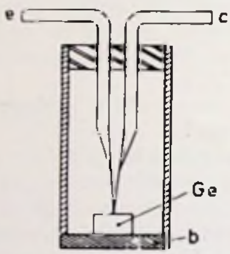

Fig. 9

may cause instability. This is turned to account by including a tuned network in the base circuit and so converting the device into a very simple oscillator. Because the frequency-limit of point-contact transistors was higher than that of the early junction transistors, they were originally employed for R.F. applications. Nowadays, however, they have been superseded, for all practical purposes, by R.F. junction transistors, e.g. types OC 44 and OC 45, which are more robust.

II. 4. SIMPLIFIED EXPLANATION OF TRANSISTOR ACTION

A much simplified explanation of transistor action, with reference to point-contact diodes and point-contact transistors, will now be given.
The point-contact diode (fig. 10) has o n e whisker contact on a germanium crystal, with low resistance in the direction from whisker to crystal, and high resistance in the direction from crystal to whisker.

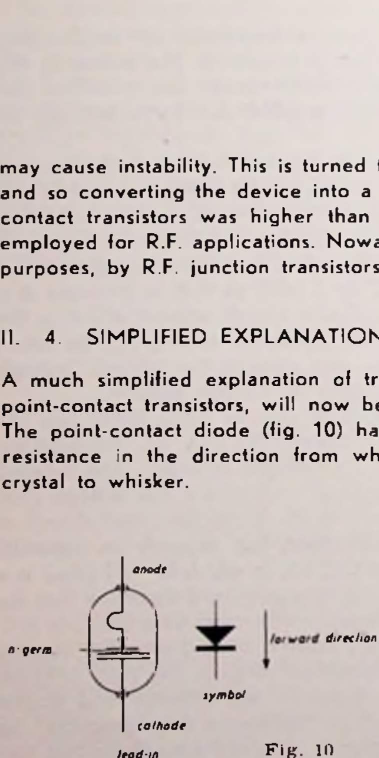

Fig. 10

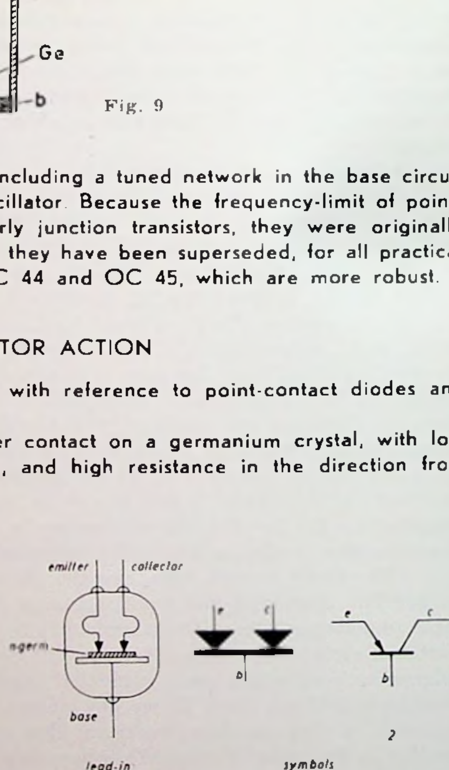

Fig. 11

The transistor (fig. 11) has two whiskers on the crystal, each contact constituting a separate diode.
Now, ideal diodes are, in effect, perfect "valves" (or contacts); in other words, the resistance is zero in the forward direction (contact closed) and infinite in the reverse direction (contact open).
However, neither germanium, nor thermionic [1]) diodes are in fact ideal in this respect. Both exhibit a certain amount of resistance in the forward direction (albeit lower in the crystal diode than in the valve), and the reverse resistance, although for all practical purposes infinite in thermionic diodes, does not greatly exceed 100 kΩ tot 2 MΩ in germanium diodes, where,

1) Vacuum diode with heated cathode.

moreover, it is subject to variation owing to external factors. For example, raising the temperature of the crystal or exposing it to strong light reduces the reverse resistance considerably [1]). Again, this resistance can also be decreased by passing an electric current through the crystal, very near the whisker. Whereas the first two of these factors are often inconvenient and inadvertent, the third is turned to good account in the manufacture of transistors.
Referring to fig. 11-1, if current be passed through the left-hand diode in the forward direction, the reverse resistance of the right-hand diode, some microns away, decreases. Now the left-hand whisker is the "emitter", the right-hand whisker the "collector" and the crystal the "base". In diagrams they are indicated by the symbols shown in fig. 11-2.

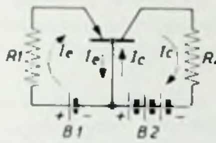

Fig. 12a

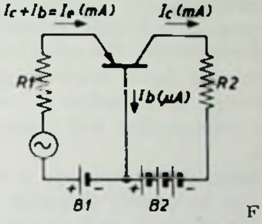

Fig. 12b

Referring to fig. 12a, we see that in the first instance emitter current (I_e) is produced. This reduces the reverse resistance of the collector/base diode, thus producing current I_c. As the two currents flow in opposite directions in the base, they partly compensate each other. The difference, I_e-I_c, is what is known as the base current, I_b [2]).
It is thus seen that we have a current I_c passing from $+B_1$ through the transistor etc. to $-B_2$, to which, on the emitter side, we must add the base current I_b (see fig. 12b). Current I_e cannot equal I_c, since if they were equal, the base current required, as we have seen, to reduce the reverse resistance of the collector/base diode, would simply not exist. The smaller I_b (the difference between current I_e and I_c), the larger the current amplification factor α; hence α is invariably less than unity.
In well-made transistors, α is between 0.95 and 0.99, or, to put it differently, the collector current is between 0.95 and 0.99 times the emitter current.
The collector load resistance R_2 is several times larger than R_1 (representing the internal resistance of the generator (or signal source) in fig. 12b). B_1 maintains a steady current, having very much the same function as negative grid bias in thermionic tubes. Let us suppose that the generator impresses on this bias current (not bias voltage) an alternating current such as to produce a variation of 0.1 V in the potential difference across R_1. Ignoring the base current for the moment, or, in other words, assuming that $I_e = I_c$, the potential difference across R_2 will also vary, and the ratio of this variation to the variation of the potential difference across R_1 will be the same as the ratio of R_2 to R_1 (e.g. if $R_2 = 30 \times R_1$, the variation in the potential across R_2 will be $30 \times 0.1 = 3$ V).
Of course, B_2 must then be capable of supplying a voltage at least equal to the variation in potential; it should be, say, a 6 V accumulator. Higher amplification can be obtained by making R_2 larger, provided that the voltage of B_2 is likewise increased so as to avoid saturation. At the same time, the battery voltage must not be made so high that it causes breakdown of the collector-base diode. Furthermore, high values of R_2 are associated with such problems as noise, dependence on temperature, and so on.

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[1]) So also with selenium cells. Rising temperature affects barrier resistance.
[2]) To simplify the argument, collector leakage-current I_{co} is left out of account.

CHAPTER III

TRANSISTOR CHARACTERISTICS

III. 1. TRANSISTORS AS COMPARED WITH ELECTRON TUBES

As pointed out at the beginning of the book, transistors bear only a very superficial resemblance to electron tubes. Nevertheless, it is useful to compare the properties of the two.
By so doing, the following advantages of transistors are brought to light:

1. They are usually very much smaller than average electron tubes. Hence, transistorized sets can be made relatively smaller, provided that the size of the other components can be likewise reduced.
2. The absence of a heated cathode means that no energy has to be supplied for cathode heating. Furthermore, the electrical properties are such as to enable transistors to be operated more efficiently (with larger loads) than electron tubes. By virtue of these properties, the overall energy consumption is very much lower, enabling portable transistorized sets to be equipped with smaller and cheaper batteries. They, in turn, permit still smaller sets to be produced.
   Another associated advantage is that transistors generate only a very small amount of heat, and this can be readily disposed of in a manner consistent with very compact construction. The amount of space and energy that can be saved in this way is considerable, particularly in the case of elaborate equipment such as electronic equipment which, if operated with tubes, are very bulky. Again, the voltages required for transistorized apparatus are very much smaller, so that there is no risk of shock or other personal hazards.
3. Very long life, which is also particularly important in the case of large equipment such as computors, which often incorporate thousands of tubes.
4. Not subject to microphony. In fact, the resistance to shock in general is very high, being limited only by the fact that glass envelopes are employed for many transistors.

On the other hand, the comparison also reveals the following drawbacks of transistors:

1. Properties depend very much on the temperature; hence special measures are necessary to prevent this dependence from affecting the circuits, or at any rate to lessen such effects.
2. The frequency limit of most of the existing transistors is low. However, low-power transistors for R.F. purposes are already beyond the laboratory stage of development; they are being used in miniature receivers.
3. Transistors do not give as much A.F. amplification as electron tubes. The ratio of the gains obtainable with electron tubes and with transistors is roughly between 2 to 1 and 3 to 2. For a given gain, therefore, more transistors than tubes are required.

III. 2. TRIODE ELECTRON TUBES

The characteristic currents and voltages of triodes are as follows:

1. Grid voltage V_g (between grid and cathode).
2. Anode voltage V_a (between anode and cathode).
3. Cathode current I_k.
4. Grid current I_g.
5. Anode current I_a.

With the grid maintained negative, $I_g = 0$, and therefore $I_k = I_a$.
Other characteristic quantities are:
with V_a constant, $\Delta I_a = S \cdot \Delta V_g$, where S is the mutual conductance (or slope) of the tube, and:

with V_g constant, $R_i = \dfrac{\Delta V_a}{\Delta I_a}$, where R_i is the internal resistance of the tube.

In triodes, R_i varies from a few kilo-ohms to some tens of megohms.
The amplification factor (μ) corresponds to:

$$\mu = R_i \cdot S = \frac{\Delta V_a}{\Delta V_g}.$$

Therefore: $\Delta V_a = \mu \cdot \Delta V_g$

and $$R_i = \frac{\Delta V_a}{\Delta i_a} = \frac{\mu}{S}.$$

III. 3. ELECTRON TUBE CHARACTERISTICS

Although triode amplification can be computed by means of the above formulae, in practice it is more often determined from the actual tube characteristics. These graphs indicate the relationship between I_a and V_g, with V_a as parameter, or the relationship between I_a and V_a, with V_g as parameter. The general form of $I_a = f(V_g)$ family of characteristics, with V_a constant, is reproduced in fig. 13, and that of the $I_a = f(V_a)$ family, with V_g constant, in fig. 14.

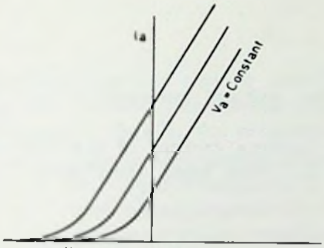


Fig. 13

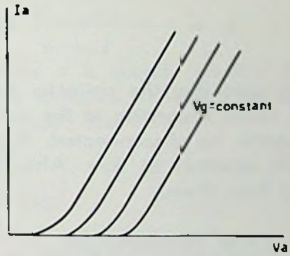


Fig. 14

III. 4. TRANSISTOR FORMULAE

As we have seen in II. 2, the emitter injects holes into the base and extracts electrons from the base. Almost all the holes are absorbed by the collector, and the electrons are carried away as minority current through the base connection.
From Kirchoff's Law:

$$I_b = I_e - I_c,$$

where I_c is the collector current, I_e the emitter current and I_b the base current. Since $I_c = \alpha I_e$, we have:

$$I_b = I_e - \alpha I_e = (1 - \alpha) I_e.$$

It should be noted that this does not take the leakage current of the collector/base diode into account. As we know, this leakage current is produced by the intrinsic electrons and holes liberated by thermal vibration, including those in the barrier layer. The result is that the intrinsic electrons pass from collector to base and the intrinsic holes from base to collector. Together, then, holes and electrons constitute the leakage current I_{co}, which, as we have seen (II. 1), depends very much on the temperature. Provided that the collector voltage remains above a certain critical value, I_{co} is invariably saturation current, or, in other words, it is independent of the voltage and governed exclusively by the temperature.

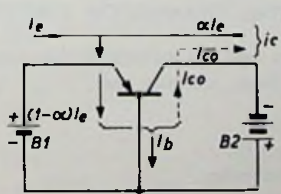


Fig. 15

Consequently, not only the holes injected by the emitter, but also the intrinsic holes migrate to the collector, thus increasing the overall collector current by an amount I_{co}. On the other hand, this migration cuts down the base current, I_b, by a similar amount, so that the resultant base current is in effect the algebraic sum of two opposed quantities (see fig. 15). The actual base current (I_b), then, is smaller than $(1 - \alpha) I_e$ by the amount I_{co}, and the actual collector current is larger than αI_e by the same amount. Thus: $I_b = (1-\alpha) I_e - I_{co}$, and $I_c = \alpha I_e + I_{co}$.

To sum up, the base current is in effect the algebraic sum of two opposed quantities, i.e. the difference between the emitter and collector currents (owing to recombination in the base crystal, see II. 2) on the one hand, and the leakage current of the collector/base diode on the other hand.
Again referring to fig. 15, we also see what happens when the base connection is cut. This produces what is known as a "floating base", whereby the base current is in effect the algebraic sum of two equal, but opposite, quantities. Thus: $I_b = (1-\alpha)\ I_e - I_{co} = 0$.
Also from Kirchoff's Law, in this case:

$$I_e = I_c.$$

Combining the two formulae, we find that the collector current produced depends on I_{co} and the current-amplification factor α.

$$I_e = I_c = \left(\frac{1}{1-\alpha}\right).\ I_{co} = I_{co}'.$$

It is thus seen that the collector current depends on the ambient temperature and the base potential varies according to the current.
If the emitter be disconnected, then, only I_{co}, i.e. the leakage current of the collector/base diode will continue to flow. Also, when once I_{co} and I_{co}' have been established, α can be computed from them.

III. 5. N-P-N TRANSISTORS

So far, only p-n-p transistors have been discussed. N-p-n transistors operate in very much the same way, the only difference being that all the currents and voltages are of opposite sign. In the basic form (II. 4), n-p-n transistors also comprise two diodes on a common base crystal, but in this case the directions of bias are reversed, as shown in fig. 16a. Fig. 16b is the conventional symbol.
The polarities of the required supply voltages are indicated in brackets; it will be seen that in this respect n-p-n transistors correspond to triode electron-tubes.

Fig. 16a

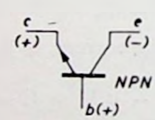


Fig. 16b

The collector (anode) is connected to the positive pole, and the emitter (cathode) to the negative pole of the battery. The transistor (tube) becomes non-conductive with negative base (grid) potential (cut-off voltage), and conductive with positive base (grid) potential (grid-current range).
By using n-p-n transistors in combination with p-n-p transistors, many interesting circuits, not possible with electron tubes, can be obtained.

III. 6. THE THREE FUNDAMENTAL TRANSISTOR CIRCUITS

Having three terminals, transistors can, in principle, be connected in three different ways to produce amplification. The three basic circuits are:

a) with grounded base,
b) with grounded emitter,
c) with grounded collector.

The characteristic feature is that in each of these circuits one electrode is common to both input and output. Since the particular electrode is usually earthed (for alternating current), the circuits are classified accordingly. They will now be discussed separately.

a) Grounded-base circuit. $I_c = f\ (I_e)$.
Fig. 15 is the fundamental grounded-base circuit, or "common-base circuit", as it is sometimes called. This diagram is shown in another form in fig. 17.

As we have seen in III. 4, the following formulae are applicable:

$$I_b = (1 - \alpha)\, I_e - I_{co} \text{ and}$$
$$I_c = \alpha\, I_e + I_{co}.$$

Since I_{co} is constant for a given temperature, it does not occur in the alternating current formulae:

$$I_{bw} = (1 - \alpha)\, I_{ew}$$
$$I_{cw} = \alpha\, I_{ew}.$$

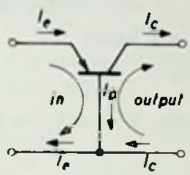


Fig. 17

This indicates that the current amplification equals α, where α is slightly smaller than unity. Although the internal impedances also do not occur in the formulae, they can be determined from the graphs. Briefly, the properties of grounded-base circuits are:

low current amplification (α = 0.92—0.99);
low input impedance, r_i in the order of 50 Ω;
high output impedance, r_o' in the order of 500 kΩ;
maximum power amplification, about 40 db.

The leakage current of the collector/base diode, which is, of course, temperature-dependent does not pass through the amplifying portion of the transistor. The circuit is therefore quite stable so far as temperature changes are concerned.
The equivalent electron tube is what is known as the grounded-grid amplifier, or grid-base circuit, in which the input and output signals are in phase.

b) Grounded-emitter circuit ($I_c = f\,(I_b)$).
This is also referred to as the "common-emitter circuit" (fig. 18).

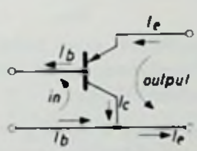


Fig. 18

It will be seen from the diagram that the emitter current is in effect the sum of the base and collector currents. Thus:

$$I_e = I_b + I_c \qquad (1)$$

Also, from III. 4:

$$I_c = \alpha\, I_e + I_{co} \qquad (2)$$

Substituting (2) for I_e in (1), we have:

$$\begin{aligned} I_c &= \alpha\,(I_b + I_c) + I_{co} \\ &= \alpha\, I_b + \alpha\, I_c + I_{co} \\ I_c - \alpha\, I_c &= \alpha\, I_b + I_{co} \\ I_c\,(1 - \alpha) &= \alpha\, I_b + I_{co} \end{aligned}$$

Therefore:

$$I_c = \frac{\alpha}{1 - \alpha}\, I_b + \frac{1}{1 - \alpha}\, I_{co} \qquad (3)$$

which indicates that the collector current is a function of the base current.

The factor $\frac{\alpha}{1 - \alpha}$ is usually indicated by the symbol α', that is, the current-amplification factor in the grounded-emitter circuit:

$$\frac{\alpha}{1-\alpha} = \alpha' \qquad (4)$$

thus indicating the ratio of I_b to I_c.
By means of a simple manipulation, it can be shown that:

$$\frac{1}{1-\alpha} = \frac{1-\alpha+\alpha}{1-\alpha} = \frac{1-\alpha}{1-\alpha} + \frac{\alpha}{1-\alpha} = 1 + \alpha' \qquad (5)$$

so that, substituting (4) and (5) in (3), we have:

$$I_c = \alpha' \, I_b + (\alpha' + 1) \, I_{co}. \qquad (6)$$

Now, α being only slightly less than unity, $1 - \alpha$ is very small and $\frac{\alpha}{1-\alpha}$, on the other hand, is very large (as also $\alpha' + 1$).
As compared with α', then, unity itself is small enough to be safely ignored; hence (6) can be simplified slightly, as follows:

$$I_c \approx \alpha' \, . \, I_b + \alpha' \, I_{co},$$

showing that in grounded-emitter circuits the leakage current I_{co} passes through the amplifying portion of the transistor and therefore emerges in the collector circuit amplified.
The amplified leakage current is designated I_{co}':

$$I_{co}' = (\alpha' + 1). \; I_{co} \approx \alpha' \, I_{co}.$$

Because I_{co} and I_{co}' are both constant at a given temperature, they do not occur in the alternating current formulae:

$$I_{ew} = I_{bw} + I_{cw} \text{ and}$$
$$I_{cw} = \alpha' \, . \, I_{bw}$$

Briefly, the characteristic features of grounded-emitter circuits are:

high current-amplification ($\alpha' = \frac{I_{cw}}{I_{bw}} = 35$ to 70)
relatively high input impedance, r_i = between 400 and 2000 Ω
relatively low output impedance, r_o = about 40 to 100 kΩ
maximum power gain about 50 db.

As we have seen, then, the leakage current of the collector/base diode passes through the amplifying portion of the transistor and therefore emerges, amplified, as I_{co}' in the collector circuit. Some means of stabilization is therefore necessary to prevent displacement of the working point owing to variation of the ambient temperature. Despite this drawback, however, earthed emitter circuits are often used because they provide greater gain than the grounded-base circuit previously described.
The electron tube equivalent in this case is the ordinary cathode base circuit, in which the input and output signals are in anti-phase.

c) Grounded-collector circuit. $I_e = f\,(I_b)$
For circuit diagram, see fig. 19. In grounded-collector circuits the collector current is in effect the difference between the emitter and base currents.

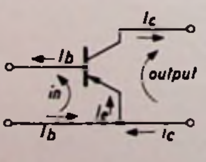


Fig. 19

In very much the same way as with the grounded-emitter circuit, it can be shown that:

$$I_e = \frac{1}{1-\alpha} \, . \, I_b + \frac{1}{1-\alpha} \, . \, I_{co}.$$

Substituting $(\alpha' + 1)$ for $\dfrac{1}{1 - \alpha}$, we have:

$$I_c = (\alpha' + 1) \cdot I_b + (\alpha' + 1) \cdot I_{co}.$$

Elimination of unity, negligible as compared with α', produces:

$$I_c \approx \alpha' (I_b + I_{co}).$$

With alternating current, I_{co} is also eliminated; therefore:

$I_c = (\alpha' + 1) \cdot I_b$, or, simplified:

$I_c \approx \alpha' \cdot I_b$.

To sum up, the characteristic features of grounded-collector circuits are:

voltage gain slightly less than unity;

substantial current gain $(= \dfrac{\alpha'}{\alpha})$;

very high input impedance, $r_i = > 20$ kΩ (depending on emitter resistance);

low output impedance, r_o = about 1 kΩ;

maximum power gain about 16 db.

Leakage current I_{co} emerges in the emitter amplified.

Equivalent electron tube is the cathode follower. Accordingly, grounded-collector circuits are sometimes called "emitter followers". As in grounded-base circuits, then, the input and output signals are in phase.

III. 7. TRANSISTOR CHARACTERISTICS

As we have seen, transistors, like triode electron tubes, are three-electrode systems. They operate with the following characteristic currents and voltages:

1. base potential V_{be} (between base and emitter);
2. collector potential V_{ce} (between collector and emitter);
3. base/collector potential V_{bc} (between base and collector);
4. emitter current I_e;
5. base current I_b;
6. collector current I_c.

Certain relationships, not linear, but depending on the action of the particular transistor, exist between these "transistor parameters" as they are called.

Given two of the six parameters within the operating range of the transistor, the others can be determined from the characteristics. For example, given I_b and V_{ce}, it is possible to compute I_c, I_e, V_{be} and V_{bc} from them. In this way, say, $I_c = f(V_{ce})$ can be computed for various fixed values of I_b, which is then the parameter of the function. This is the most important characteristic from the point of view of the practical application of transistors; it resembles the I_a/V_a characteristic, with V_g as parameter, of electron tubes. Fig. 20a shows the family of characteristics obtained by plotting the function $I_c = f(V_{ce})$, with I_b constant, for the circuit shown in fig. 18.

It will be seen that they resemble typical pentode electron-tube characteristics (e.g. the I_a/V_a characteristic of the EL 84; see fig. 20b). The straight portions are almost horizontal, indicating that the internal resistance (R_i) of the collector is very high, i.e. in the range from 10 kΩ to 100 kΩ; R_i decreases with the collector current.

Again, the lines are spaced quite regularly over a large region, an indication of linear amplification covering a considerable portion of the operating range.

The knee of the characteristic (i.e. saturation point, at which virtually all the holes injected by the emitter are absorbed by the collector) occurs at a very low collector potential, in the order of some tens of volts. This indicates that the transistor has a high load-rating, consistent with high efficiency. In the characteristics of what are known as power transistors, the knee is slightly higher (at approximately 1 V).

It will also be seen from the diagram that a certain amount of collector current, I_{co}', occurs when $I_b = 0$. As explained in section 4, this is produced by I_{co}.

Now an exponential relationship exists between I_{co} and the temperature; in germanium transistors, this current doubles itself for every 10° C temperature rise. If the junction temperature rises, say, owing to the load, the overall family of characteristics shifts upwards, thus moving

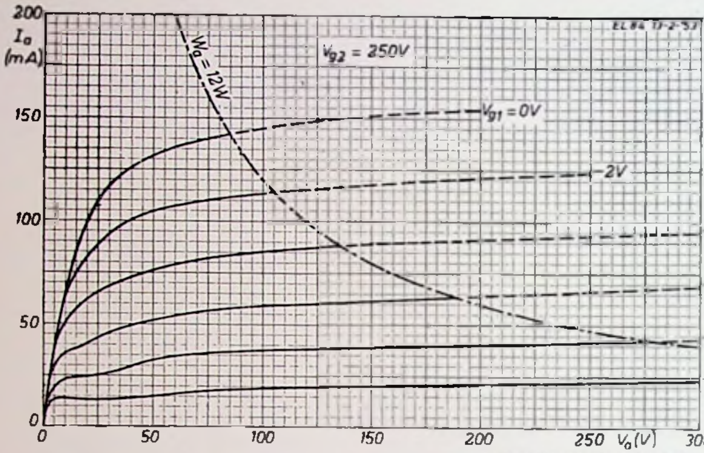


Fig. 20b

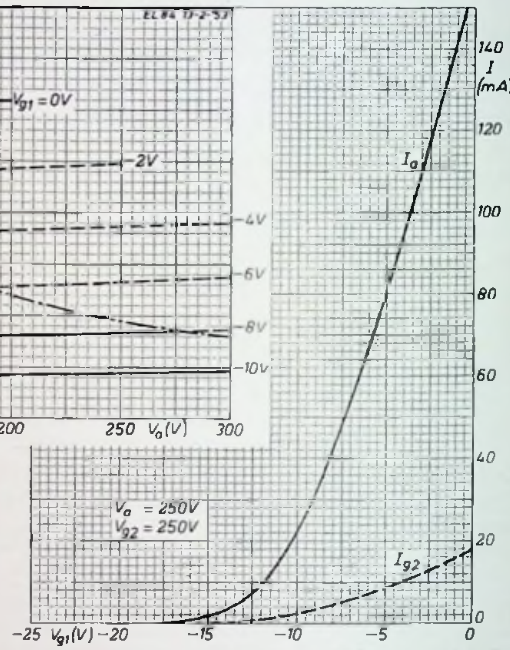


Fig. 21b

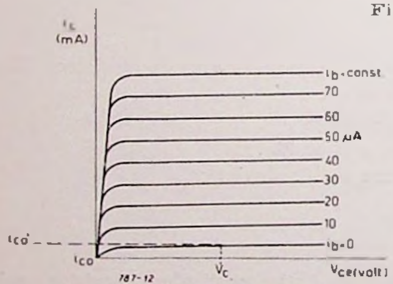


Fig. 20a

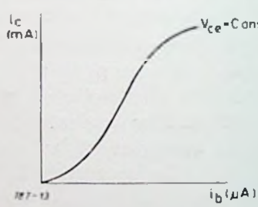


Fig. 21a

the working point away from the original operating conditions. It is therefore necessary to employ some means of preventing, or at any rate compensating, such temperature variations. Another important transistor characteristic is $I_c = f(I_b)$, with V_{ce} constant, representing the current amplification associated with a given collector voltage (fig. 21a). This graph is comparable to the I_a/V_g characteristic of electron tubes (see fig. 21b, I_a/V_g characteristic of pentode EL 84, where I_a corresponds to I_c and V_g to I_b).
Experience has shown that the amplification falls off at very low, and also towards very high, values of I_c. In modern transistors, maximum current amplification usually occurs in the region of 1 mA collector current. Transistor manufacturers try to keep the slope as far as possible constant. Whereas this may reach 100 in well-made modern transistors, it does not greatly exceed 20 in many older types.
The $I_c = f(I_b)$ characteristic can be derived from the $I_c = f(V_{ce})$ characteristic by reading the value of I_c corresponding to each value of I_b on a vertical line erected at a certain value of V_c (see fig. 22).

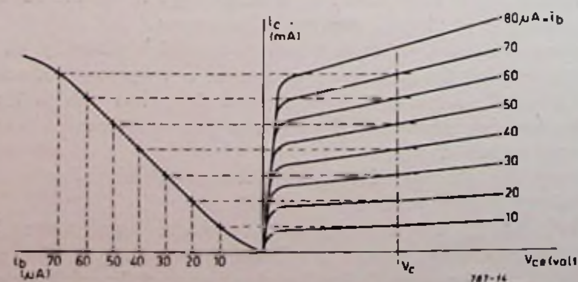


Fig. 22

Further examination of the curves in fig. 22 reveals that the A.C. parameters of the transistor can be deduced from the slope of the characteristics.
The amplitude of the output signal can be read from a load line drawn in the same way as with electron tube characteristics (for example, see fig. 23). Amplification is generally linear for small amplitudes; also for stronger signals, provided that the characteristics are straight enough.

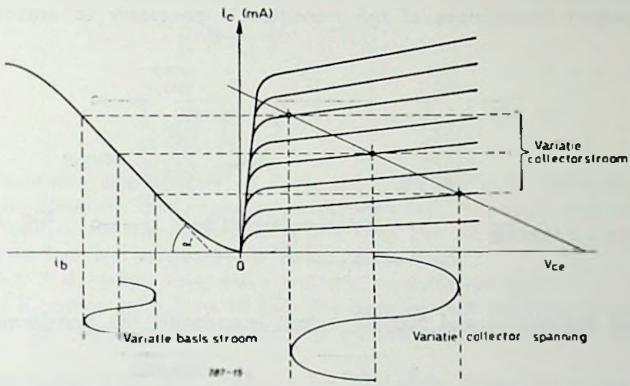


Fig. 23

Another feature of transistors, which we have not considered so far, is the internal feedback. In transistors, all the electrodes are, in effect, coupled by (non-linear) resistances. Any variation in one of the quantities (current, voltage) produces greater or smaller variations in the other. Another equally important characteristic is that of the input, $V_{be} = f\ (I_b)$, indicating the variation of the input current I_b as a function of the input voltage V_{be}, with the output voltage constant (fig. 24a).
As the input circuit of a junction transistor is actually a diode receiving current in the reverse direction, the impedance varies according to the current and voltage. Current amplification in transistors is linear; to minimize distortion, then, transistors must be operated with a current source, or, to put it in another way, the (constant) impedance of the current source must be high enough to ensure that, by comparison, the varying impedance of the transistor does not greatly affect the total resistance of the control circuit.

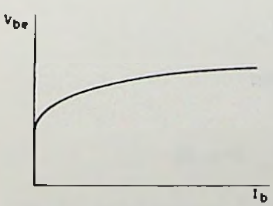


Fig. 24a

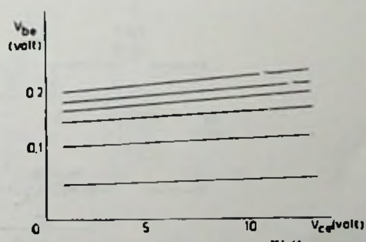


Fig. 24b

It should therefore be borne in mind that non-linearity of the base voltage does not necessarily mean non-linearity of the base current also. This is very important in connection with the observation of wave forms in oscilloscopes. The oscilloscope should be connected directly across a measuring resistor introduced into the circuit on test (i.e. oscilloscope input at the one end, and oscilloscope earth at the other end, of the measuring resistor).
The base voltage also depends on the collector voltage. By plotting the relationship between the two, V_{be}/V_{ce} characteristics, or feedback characteristics as they are called, are obtained (fig. 24b).

CHAPTER IV

TRANSISTOR AMPLIFIER DESIGN

IV. 1. CIRCUITS WITHOUT STABILIZATION

a) Grounded-base circuit.

Fig. 25 is the circuit diagram. As the input current is for all practical purposes the same as the output current, amplification is governed exclusively by the transformation ratios. Matching to the input and output impedances of the transistor is necessary to ensure maximum power amplification.

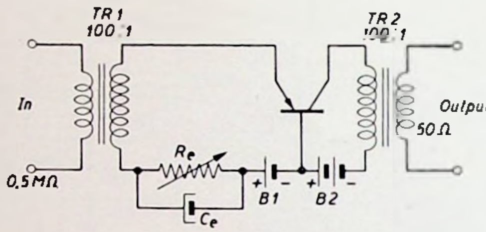


Fig. 25

Given 50 Ω input impedance and 500 kΩ output impedance, the transformation ratio will be:

$$n = \sqrt{\frac{Z_o}{Z_i}} = \sqrt{\frac{500{,}000}{50}} = 100.$$

A separate battery B_1 is required to regulate the emitter current. A tapping on the original battery, or a voltage divider with the base connected to it, can be used instead, but the divider must then have a fairly low D.C. resistance and therefore consumes more energy than is strictly necessary.

To enable the direct current to be regulated independently of the (varying) transistor impedance, the emitter series resistance must be at least ten times the input impedance of the transistor. Failing this, the variation will be amplified and may then cause distortion.

R_e must be decoupled with respect to alternating current by means of a fairly large electrolytic capacitor (C_e, at least 100 μF).

Temperature stabilization is not necessary in this circuit.

b) Grounded-emitter circuit (without stabilization).

Fig. 26 shows the circuit in its simplest form. As in a), transformer coupling is employed and matching is necessary to ensure maximum power gain.

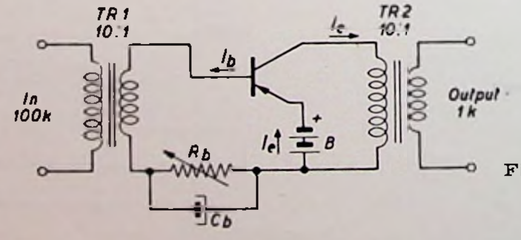


Fig. 26

Given 1 kΩ input impedance and 100 kΩ output impedance, the transformation ratio will be:

$$n = \sqrt{\frac{Z_o}{Z_i}} = \sqrt{\frac{100}{1}} = 10.$$

Because the direct current required for the base is only a few microamperes, the base resistance R_b is very high, enabling a very much smaller decoupling capacitor (one or two μF) to be employed. Hence only one battery is required. As the circuit is not stabilized with respect to temperature variations, it can be used only where the temperature is more or less constant.

RC-coupling can be used instead of transformer coupling (fig. 27). Usually, input and output matching is not possible in circuits thus coupled, and it is certainly out of the question in cascade circuits. The variation of collector current ($\triangle$ I_c) in resistor R_c produces a voltage variation $\triangle I_c \cdot R_c = \triangle V_o$, a factor α' times the variation of the input current ($\triangle$ I_b), in this resistor.

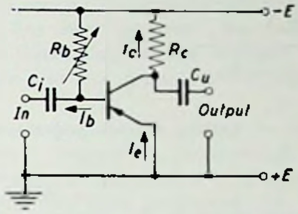


Fig. 27

In practice a compromise is necessary; if R_c is too high, the A.C. collector voltage drops to zero as soon as the (smaller) load impedance is connected. Although such voltage suppression is not serious in itself, since transistors are in any case current-operated (see II. 6), it involves a risk that, owing to I_{co}, the transistor will cease to function.
On the other hand, if R_c is too low, there will not be sufficient gain.
Practical values of R_c range from 5 to 10 kΩ. The base current (and therefore also the working point) is regulated by means of R_b. Depending on the collector current, R_b should be from 1 MΩ to 150 kΩ.
Because the circuit is not stabilized, it should be used only where the temperature is more or less constant. Although there are exceptions to this rule for sets operating with only very small amplitudes, the apparatus should not be exposed to the direct rays of the sun, or the transistor may nevertheless cease to function.
Apart from these exceptions, then, D.C. stabilization, or in essence D.C. feedback, is essential. Apart from its stabilizing effect, D.C. feedback tends to compensate for the spread of characteristics of individual transistors (see IV, 2), in much the same way as individual cathode resistances compensate for the spread of tubes in push-pull amplifiers, one of the stages being dispensed with in order to offset the difference between characteristics and thus ensure a symmetrical output signal.

c) Grounded-collector circuit.

In grounded-collector circuits, or emitter followers as they are also called (fig. 28), the transistor is employed as an impedance transformer. The input impedance (Z_i) is computed in the following manner:

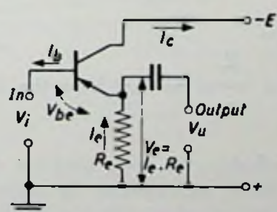


Fig. 28

Let the input voltage (V_i) equal the sum of the voltage drop (V_e) across the emitter resistance and the voltage drop (V_{be}) across the internal resistance (R_i) of the emitter/base diode:

$$V_i = V_e + V_{be} = V_e + I_b R_i. \qquad (1)$$

As we have seen, the emitter current is:

$$I_e = (\alpha' + 1) I_b + (\alpha' + 1) I_{co}, \qquad (2)$$

enabling the base current (I_b) to be expressed in terms of I_e:

$$I_b = \frac{I_e}{\alpha' + 1} - I_{co} = \frac{I_e - (\alpha' + 1) I_{co}}{\alpha' + 1}. \qquad (3)$$

The input impedance (Z_i) corresponds to:

$$Z_i = \frac{V_i}{I_b}, \qquad (4)$$

so that, by substituting (3) in (4), we obtain:

$$Z_i = V_i \frac{\alpha' + 1}{I_e - (\alpha' + 1) I_{co}}. \qquad (5)$$

The emitter current is then $I_e = \frac{V_e}{R_e}$, and by substituting this in (5), we find that:

$$Z_i = V_i \frac{\alpha' + 1}{V_e/R_e - (\alpha' + 1) I_{co}} = R_e \frac{(\alpha' + 1) V_i}{V_e - (\alpha' + 1) R_e I_{co}}. \qquad (6)$$

From formula (1) we have:

$$V_e = V_i - V_{be} \qquad (7)$$

so that formula (6) becomes:

$$Z_i = R_e \frac{(\alpha' + 1) . V_i}{(V_i - V_{be}) - (\alpha' + 1) . I_{co} . R_e}. \qquad (8)$$

Since I_b and R_e are both small, V_i is usually $\gg V_{be}$; hence V_{be} can be eliminated.

$$Z \approx R_e \frac{\alpha' + 1}{1 - (\alpha' + 1) I_{co} . R_e/V_i}. \qquad (9)$$

Assuming that I_{co} is also small, $(\alpha' + 1) . I_{co} \frac{R_e}{V_i}$ can also be eliminated, making the input impedance, by approximation:

$$Z_i \approx R_e (\alpha' + 1) R_e . \alpha'. \qquad (10)$$

From the point of view of the input, then, emitter resistance R_e is multiplied by a factor of α'.

IV. 2. DESIGN AND STABILIZATION OF GROUNDED-EMITTER TYPE TRANSISTOR AMPLIFIERS

In grounded emitter circuits:

$$I_c = \alpha' . I_b + \alpha' . I_{co}.$$

In the circuit shown in fig. 27, $I_b = \frac{E}{R_b}$, so that here the direct collector current I_c is:

$$I_c = \alpha' . \frac{E}{R_b} + \alpha' . I_{co}.$$

With a variation of ΔI_{co} in I_{co} owing to temperature variations, we have:

$$\Delta I_c (t) = \alpha' . \Delta I_{co}.$$

If the transistor in the circuit be replaced by another whose α' is a factor of $\Delta \alpha'$ different owing to manufacturing tolerances, the collector current will vary as follows:

$$\Delta I_c (\alpha') = \Delta \alpha' \left(\frac{E}{R_b} + I_{co} \right)$$

It will be seen, then, that I_c varies by a factor ΔI_{co} owing to temperature variations, and by a factor $\Delta \alpha'$ owing to spread in the transistors.

As $\Delta \alpha'$ may be as much as a factor of 2, considerable variations in gain may occur, which in some cases are not acceptable. However, such variations usually do not matter very much to amateurs.

Temperature variations of 25 to 45 °C may cause I_{co} and I_{co}' to vary by factors of from 5 to 8, but this is not serious, provided that the peak value of I_{co}' is only a fraction of the overall collector current I_c. At 25 °C the I_{co} of, say, the OC 71 is roughly 8 μA, and the associated I_{co}' is in the region of 150 μA; the values vary, however, as between individual transistors. Provided that I_c is considerably greater than the peak I_{co}' (8 x 150 μA), then, the variation of I_{co}' does not affect matters very much. To provide sufficient I_c, a relatively heavy current is required in the base (I_b), to that this also "swamps" the leakage current I_{co}, thus minimizing the relative variations in I_b due to fluctuations in the ambient temperature. On the other hand, there is a risk that the temperature will rise owing to the increase in dissipation, thereby producing an increase in I_{co}.
Designing a transistor amplifier according to the circuit in fig. 27 is a simple matter. When once the working point is chosen — depending, in practice, upon the maximum permissible, or available, supply voltage and the load impedance — the required base current can be read direct from the static characteristics. The base resistance can then be computed in accordance with Ohm's Law: $R_b = \dfrac{E}{I_b}$, bearing in mind that: $\dfrac{V}{\mu A} = M\Omega$.

The actual method is as follows:
Having established the collector (load) resistance R_c and the battery voltage E_b, draw the load line R_c in the diagram (see fig. 29), as in the case of tube characteristics (with $I_c = 0$, then, $V_{ce} = 0$, $I_c \dfrac{E_b}{R_c}$).
With heavier collector current, the knee voltage V_k is roughly 0.5 V, so that to ensure linear amplification it is necessary to shift the working point an amount $\dfrac{E_b - V_k}{2}$ from E_b (for convenience, $I_{co}' \times R_c$ not taken into account).

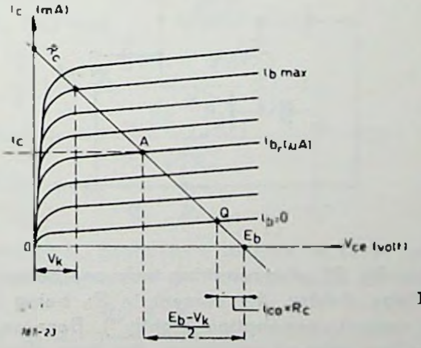


Fig. 29

Next, draw a vertical line from this working point through the load line, and read the required quiescent base current from the point of intersection. With regard to the maximum voltage permissible between emitter and collector, it should be borne in mind that: WITH TRANSFORMER COUPLING, A POTENTIAL DIFFERENCE EQUAL TO TWICE THE BATTERY VOLTAGE MAY OCCUR ACROSS THE TRANSISTOR, OWING TO SELF-INDUCTANCE.
With $V_{ce\ max.} = 25$ V, then, the battery voltage should not exceed 12.5 V. This applies especially to output transistors.
On the other hand, the simple circuit referred to cannot be employed in, say, pre-amplifiers and others operating with small collector current. Here, temperature stabilization is indispensable. What is known as the self-biasing method of connecting the base series resistance to the collector, as illustrated in fig. 30, provides a simple solution. The base resistance R_b then corresponds to:

$$R_b = \frac{V_{ce}}{I_b} = \frac{E - I_c R_c}{I_b}.$$

The stabilizing effect can be explained as follows: Suppose that the collector current I_c tends to increase owing to a rise in ambient temperature (or to the transistor replaced by another having a higher I_{co}'); the voltage drop $I_c \cdot R_c$ then increases, thus cutting down the collector voltage $V_{ce} = E - I_cR_c$, and therefore also the base current $I_b = E - \frac{I_cR_c}{R_b}$.

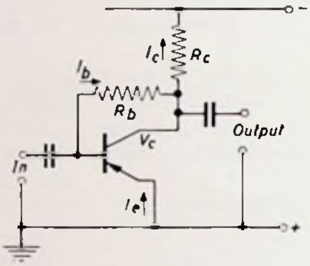


Fig. 30

In addition to I_b, the current $I_c = \alpha' (I_b + I_{co})$ also decreases. It will be evident that in this way the original increase in the collector current is compensated by feedback affecting not only the direct, but also the alternating current and thus reducing the overall amplification (to roughly 25 db). This feedback also compensates to some extent for variation between individual characteristics. The A.C. feedback can be eliminated by dividing Rb into two sections and decoupling at the mid-point (fig. 31).

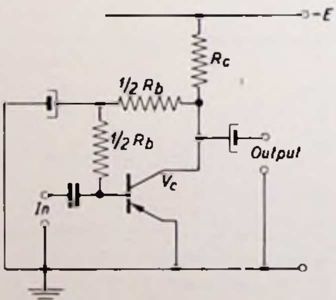


Fig. 31

The arrangement illustrated in fig. 32, also requiring only one battery, is more effective, however. R_b and R_s constitute a voltage divider, the current in R_b being $I_b + I_s$. Greater stability is obtained by increasing the current-consumption slightly *). Because the base-current is supplied by a source with relatively low internal resistance, the base potential is more constant (from Thevenin's theory, the voltage divider is in effect a voltage source with $R_i = \frac{R_b \cdot R_s}{R_b + R_s}$, whose E.M.F. is $\frac{R_s}{R_b + R_s} \times E$).

Provided R_e is high compared with the internal resistance of the emitter, and that the overall resistance in the base circuit is not too high, the emitter current will be fairly constant. Again, in junction transistors, the collector current (within the operating range) is for all practical purposes independent of the collector voltage (pentode characteristic). Whereas the circuit in fig. 27 satisfies the second of the two conditions, it has one or two unsatisfactory features from the point of view of the first condition. Because the base current is small, R_b is very high. Hence a separate battery supplying the base with lower voltage and thus enabling the same base current to be obtained with a lower R_b would be an improvement. Because extra batteries

*) From this point of view the circuit is then for all practical purposes in grounded base.

are usually inconvenient, however, it is better to provide a voltage divider ($R_b + R_s$ in fig. 32); the extra battery can then be dispensed with, although on the other hand the power consumption increases by a few milliwatts.

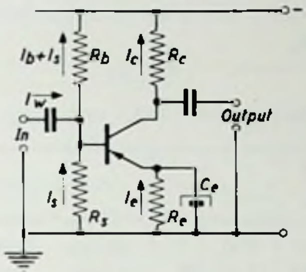


Fig. 32

With the addition of the voltage divider the circuit becomes, in effect, a grounded-base circuit and therefore less sensitive to temperature variations.
With the above-mentioned conditions satisfied, then:

$$I_b = \frac{\dfrac{R_s}{R_b + R_s} V_B - I_e R_e}{\dfrac{R_b\,R_s}{R_b + R_s}}. \quad (1)$$

on the assumption that V_{be} is negligible compared with $I_e R_e$ (although this is not strictly permissible, since dependence on temperature governs stability, the more so with class B output stages). Also:

$$I_c = I_e - I_b = \alpha I_e + I_{co} \quad (2)$$

Resolving these equations, we find that:

$$I_c = \frac{\alpha \dfrac{E}{R_b} + \left\{ 1 + R_e \dfrac{R_b + R_s}{R_b \,.\, R_s} \right\} I_{co}}{(1 - \alpha) + R_e \dfrac{R_b + R_s}{R_b \,.\, R_s}}. \quad (3)$$

Differentiation of I_c with respect to I_{co} produces a stability factor S, defining the amplified version of the collector/base diode leakage-current I_{co} appearing in the overall collector current I_c:

$$S = \frac{\delta I_c}{\delta I_{co}} = \frac{1 + R_e \dfrac{R_b + R_s}{R_b \,.\, R_s}}{(1 - \alpha) + R_e \dfrac{R_b + R_s}{R_b \,.\, R_s}}. \quad (4)$$

Converting formula (4) and substituting $\dfrac{R_b \,.\, R_s}{R_b + R_s} = R_v$ for the parallel circuit, we have:

$$S = \frac{1 + R_e \,.\, \dfrac{1}{R_v}}{(1 - \alpha) + R_e \dfrac{1}{R_v}} = \frac{R_v + R_e}{(1 - \alpha) R_v + R_e}. \quad (5)$$

Again, since $\alpha' = \dfrac{\alpha}{1 - \alpha}$, α can be expressed in terms of α',

so that formula (5) becomes:

$$S = \frac{R_s + R_e}{\dfrac{1}{1 + \alpha'} R_s + R_e}. \qquad (6)$$

Assuming that $R_s = R_e$:

$$S = \frac{2\ R_e}{\left\{ \dfrac{1}{1 + \alpha'} + 1 \right\} R_e} = \frac{2 + 2\alpha'}{2 + \alpha'}. \qquad (7)$$

Substituting (5) and (7) in (3) and developing the equation, we find that:

$$I_c = \frac{E}{R_b} (S - 1) + S\ .\ I_{co}$$

$$= \frac{E}{R_b}\ \frac{\alpha'}{2 + \alpha'} + \frac{2 + 2\alpha'}{2 + \alpha'}\ .\ I_{co}. \qquad (8)$$

With α' high, formula (8) becomes, by approximation:

$$I_c \approx \frac{E}{R_b} + 2\ I_{co}.$$

It will now be seen that the variation in the collector current, owing to a variation $\triangle\ I_{co}$ produced in I_{co} by change in temperature, is only:

$$\triangle\ I_c\ (t) = 2\ \triangle\ I_{co},$$

an improvement of $\alpha'/2$ as compared with the condition without stabilization.

Again, it can be shown, as regards alternating current:

$$I_{ca} \approx \alpha'\ I_{ba},$$

indicating that this method of stabilization does not affect amplification.

On the other hand, a certain loss of amplification does occur in base voltage dividers in grounded-emitter circuits (base injection). For this reason, a slightly higher S (roughly 8) is employed in such circuits, enabling what is in most cases adequate stabilization to be obtained at the sacrifice of only 3 or 4 db.

In circuits not of the grounded-emitter type, $S = 2$ is chosen as in the above case, where $R_s = R_e = R_c$. At the same time, almost half the D.C. power supplied to the collector is then dissipated in R_e.

Usually, R_e is roughly $^1/_3$ R_c, and S is from 6 to 25.

From the above equations, given α, E, V_c, I_c, I_{co} and R_c:

$$R_e = \frac{\alpha\ (E - V_c - I_c\ .\ R_c)}{I_c - I_{co}}$$

$$R_s = \frac{S - 1}{\dfrac{[\ 1 - S\ (1 - \alpha)\ ]\ (I_c - I_{co})}{\alpha\ (E - V_c - R_c\ .\ I_c)} - \dfrac{1}{E}\ (I_c - S\ .\ I_{co})}$$

$$R_b = \frac{E\ (S - 1)}{I_c - S\ .\ I_{co}},$$

and given α, E, I_{co}, R_e, R_b, R_s and R_c: $I_c = \dfrac{E}{R_b} (S - 1) + S\ I_{co}$

$$V_c = E \left\{ 1 - \frac{S-1}{R_b} \left(\frac{\alpha}{R_e} + R_i \right) \right\} - I_{co} \left\{ \frac{R_e (S-1)}{\alpha} + S \cdot R_e \right\}$$

The data for any amplifier can be computed by means of these formulae.

TABLE 1

STANDARD STABILIZED CIRCUITS FOR TRANSISTOR TYPE OC 71

Method of coupling	Application	Nr.	V_B (volt)	R_e (Ω)	R_b (kΩ)	R_s (kΩ)	R_c (kΩ)	i_c nom. (mA)
RC-coupling	weak signals	1	4,5	1000	18	2,7	3,3	0,5
	weak signals	2	4,5	1800	15	10	3,3	0,4
	stronger signals	3	6,0	1000	33	3,9	3,3	0,5
	stronger signals	4	6,0	1800	82	15	5,6	0,4
Transformer coupling	weak signals	5	1,5	1000	4,7	3,3	0,2*	0,5
	weak signals	6	3,0	1000	10	2,7	0,2*	0,5
	stronger signals	7	4,5	470	10	6,8	0,2*	3,0
	stronger signals	8	6,0	470	12	4,7	0,2*	3,0
	stronger signals	9	6,0	470	39	18	0,3*	2,6
Transformer coupling with common divider R_b/R_s (max. 3 stages)	weak signals	10	6,0	1000	2,2	0,56	0,2*	0,45
	stronger signals	11	6,0	2200	2,2	0,56	0,2*	1,0

* D.C. resistance of transformer primary.

The collector impedance is $Z_c = \dfrac{V_{ce\ max}}{I_c} = \dfrac{V_B - I_e R_e}{I_{c\ max}}$.

As the formulae are rather complex and take time to work out, however, it is simpler to employ the values specified in Table 1.

Another method of approximation suitable for circuits other than those discussed so far will now be described; it should be borne in mind, however, that this method ONLY HOLDS GOOD FOR EXPERIMENTAL CIRCUITS, which, although requiring a certain amount of stabilization, are nevertheless not unduly critical.

The first step is to choose a nominal collector current. Next, a value of R_e is selected, which ensures that the voltage drop $I_e R_e$ is between $1/2$ and $1/4$ of the battery voltage (see fig. 32), and R_s is made approximately $10 \times R_e$.

The final value of resistance R_b is then determined experimentally by increasing it gradually from the value of R_s until the milliammeter in the collector circuit indicates the required current. The greater the voltage drop in R_e and the smaller the ratio R_s/R_e, the better the stabilization. At the same time, this system involves the loss of much of the available battery voltage and of the impressed A.C. input power.

IV. 3. OUTPUT AMPLIFIERS

The prime object of an output amplifier is to produce as much output power as possible. Maximum power gain is obtained with the input and output impedances matched and the amplifier fully loaded. At the same time, of course, the maximum limit of collector dissipation must not be exceeded, since this dissipation affects transistors more than electron tubes.

It sometimes happens that maximum output power is obtained with less than maximum power gain.

Output amplifiers may be divided into two groups:

a) single-stage;
b) push-pull.

a) Single-stage output amplifiers

With transistor output amplifiers the maximum limit of collector dissipation, associated with high output power, is a problem.
The "heaviest" transistor at present made is the OC 16, whose output power in class A is of the order of 4 W.
In smaller transistors, say, the OC 70, 71 and 72, maximum collector dissipation is 100 mW. However, this falls by 2 mW per °C rise in temperature as from 25 °C. Hence it is necessary to cool the transistors effectively.
Transistors to operate at relatively high power levels are therefore provided with cooling fins, screwed to a heat-conducting chassis.
Referring to the I_c/V_{ce} characteristic, we see that by joining all the I_c and V_{ce} points whose product indicates the maximum permissible collector dissipation, we produce a hyperbola ($W_{c\ max}$). The load line may then be drawn anywhere below this dissipation line. Whether we should draw a tangent to $W_{c\ max}$ from point $V_{ce\ max}$ or from point $I_{c\ max}$ (maximum permissible collector current) depends entirely on the permissible collector voltage.

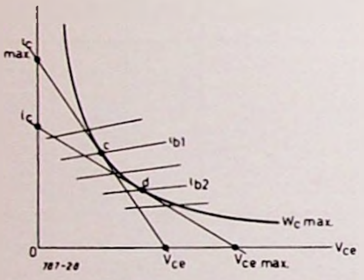


Fig. 33

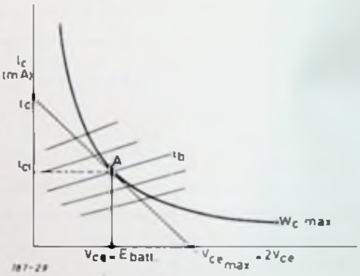


Fig. 34

As we have seen, with transformer coupling, 12.5 V from the battery produces 2 × 12.5 V = 25 V A.C. between collector and emitter, owing to the self-inductance of the transformer. With transformer coupling, then, the D.C. voltage loss in the transformer primary is only small; hence $V_{ce} = E_{batt.}$, and $E_{batt.}$ should not be more than half the maximum voltage permissible between collector and emitter. To return to the characteristic, from point $V_{ce} = E_{batt.}$ we erect a vertical line cutting the dissipation line W_c, draw another line from the point of intersection to point $2\ V_{ce} = V_{ce\ max}$, and extend it to cut the I_c co-ordinate. The required collector impedance Z_c can then be computed by means of the formula:

$$Z_c \frac{V_{ce\ max}}{I_c} \text{ or } \frac{V_{ce}}{I_{c\ max}} \text{ (fig. 33 and fig. 34).}$$

It will be seen from the diagram that the working point (A) is then at the centre of load line Z_c (at any rate, without taking the knee voltage and I_{co}' into account). The output power (W_o) to Z_c under full-load conditions is $I_c \cdot V_{ce}$. The power supplied by the battery is: $W_b = I_c \cdot V_{ce\ max} = I_c \cdot 2\ V_{ce}$ (half of this is supplied indirectly by the magnetic energy stored in the transformer core).

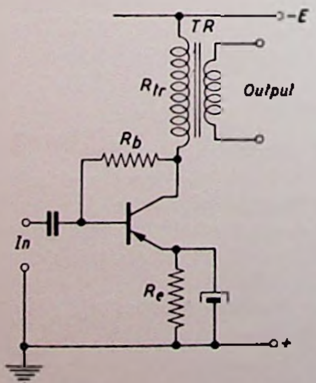


Fig. 35

Accordingly, the efficiency of the system is $\frac{W_o}{W_b} \cdot 100\,\% = 50\,\%$.

Fig. 35 illustrates the principle of stabilized output amplifiers. Note that emitter resistance R_e is decoupled by means of a large capacitor. The capacitance varies from 100—1000 μF, depending on the direct emitter current.

b) Push-pull output amplifiers

Very much more output power can be obtained by employing two transistors in push-pull. Like electron tubes, they can be operated in class A, class AB or class B.
In class A the total output power is twice that obtained with single-stage amplifiers. Other advantages are compensation of the D.C. magnetization of the output transformer and elimination of even harmonics. Because conditions for the individual transistors are the same as in single output stages, however, the efficiency is still only 50 %.

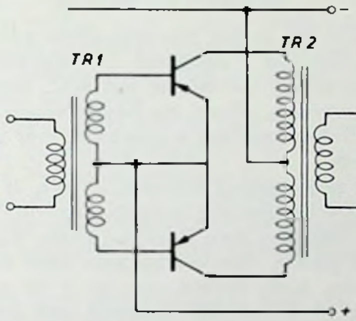


Fig. 36

For portable sets etc., it is best to operate both output transistors in class B (fig. 36). This ensures very low no-signal current and very high efficiency at full load. Biasing takes place with very weak collector current, theoretically $I_c = I_{co}$, and $I_b = 0$ (point Q in fig. 29). In theory, at any rate, the output power per transistor, fully loaded, is:

$$W_o = {}^1\!/_4\, E\, I_{c\ max.} \qquad (1)$$

and the collector dissipation per transistor:

$$W_c = \left(\frac{1}{\pi} - \frac{1}{4}\right) E\, I_{c\ max.} \qquad (2)$$

The power supplied by the battery is therefore:

$$W_b = \frac{1}{\pi} E \cdot I_{c\ max.} \qquad (3)$$

making the efficiency:

$$\eta = \frac{\pi}{4} = 78\,\%. \qquad (4)$$

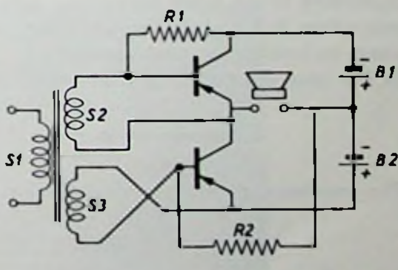


Fig. 37

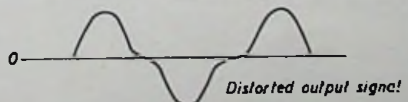


Fig. 38

In reality, however, the efficiency is slightly lower, firstly because transistors cannot be operated with 100 % load owing to the knee voltage and I_{co}', and secondly because a certain amount of power is invariably wasted in the output transformer and in possible base resistances. Nevertheless, transistors attain very close to theoretical efficiency.

A drawback of class B operation is that it necessitates a tapped output transformer, i.e., in effect, a transformer with two primary windings, which cannot be accommodated in a small space.
On the other hand, it eliminates D.C. magnetization of the transformer core, although in any case this does not affect class B output stages very much because the steady current is very small.
It is a moot point whether a centre tapping should be provided on the speech coil, as this is not easily accomplished. The output transformer can be simplified considerably, and in some cases dispensed with altogether, by employing what is known as a single-ended push-pull circuit. Because transistors operate with low voltage and comparatively heavy current, it is possible, with suitable transistors (e.g. power transistors type OC 16), to incorporate the normal speech-coil impedance in the collector circuit. Fig. 37 shows how this is done.
Resistors R_1 and R_2, with the D.C. resistance of windings S_2 and S_3 of the driver transformer, constitute voltage dividers imparting a slight negative bias to the bases. The reason for this arrangement is that transistors for class B operation are not reliable at $I_b = 0$, since this is associated with distortion of the output signal owing to non-linearity of the input characteristic (see fig. 38).
By virtue of the very low voltages in the circuit, this arrangement is very much simpler in many respects (insulation etc.) than it would be if equipped with electron tubes. The emitter of the top transistor (cathode in electron tubes) carries the full A.C. output voltage with respect to earth. With tubes, however, this invariably affects the heater supply. Although the two separate secondary windings on the driver transformer are a drawback, they can be avoided by arranging the circuit in a slightly different way.
Another drawback, common to all single-ended push-pull circuits, is that they require either two identical batteries, B_1 and B_2, or one battery with a centre-tapping.
It should be borne in mind when designing class B output stages that the battery voltage should be consistent with optimum output power. This will be seen from the following calculation.
Suppose that two power transistors, both limited, say, to 2 W collector dissipation, roughly 3 amp. peak collector-current and 24 V collector potential, are connected in push-pull, in the manner indicated in fig. 37. Given a 12 V accumulator with centre-tapping (i.e. 2 × 6 V), the peak collector current ($I_{c\,max}$), as computed by means of formula (2), will be:

$$I_{c\,max} = \frac{W_c}{\left(\frac{1}{\pi} - \frac{1}{4}\right) . E} = \frac{2}{\left(\frac{1}{\pi} - \frac{1}{4}\right) . 6} = 4.87 \text{ A.}$$

According to the latest provisional data, however, the collector current should not be very much more than 3 A.; hence the transistors should not be loaded to maximum dissipation, or they will cease to function at peak current.
In this case the matching impedance should be:

$$Z_c = \frac{E}{I_{c\,max}} = \frac{6}{3} = 2\ \Omega.$$

The maximum output power of the push-pull circuit is then:

$$W_{o\,tot} = 2 \times \frac{1}{4} E . I_{c\,max} = 2 . \frac{1}{4} . 6 . 3 = 9 \text{ watts.}$$

Accordingly, the collector dissipation is:

$$W_c = \left\{\frac{1}{\pi} - \frac{1}{4}\right\} E . I_{c\,max} = \left\{\frac{1}{\pi} - \frac{1}{4}\right\} . 6 . 3 = \text{roughly } 1.5 \text{ W (per transistor).}$$

On the other hand, given 2 × 12 V from the battery the peak current with the transistors loaded to maximum dissipation, will be:

$$I_{c\,max} = \frac{2}{\left(\frac{1}{\pi} - \frac{1}{4}\right) . 12} = 2.44 \text{ A,}$$

that is, below the prescribed limit.

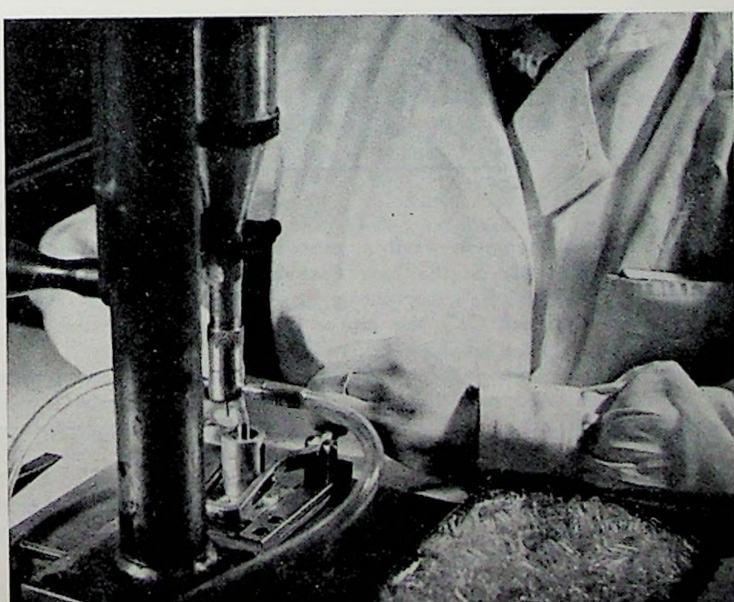
The matching impedance is then:

$$Z_c = \frac{12}{2.44} = \text{roughly } 5\ \Omega,$$

which is normal for speech coils.
The associated maximum output power is:

$$W_{o\ tot} = 2 \cdot \frac{1}{4} \cdot 12 \cdot 2.44 = \text{roughly } 14\ \text{W}.$$

So much for the theory.
In practice, however, the power level is very much lower, partly because the transistors cannot be operated with 100 % load and partly because of the rise in temperature caused by dissipation. With high power output, then, stabilization is essential.
Clearly, the most economical method is to employ NTC-resistors in the base circuit. Also, it is very difficult to maintain stable operating conditions at all temperatures from, say, 0° to 50° C. Displacement of the working point may cause the transistors to be destroyed. Another drawback is that NTC resistors are rather expensive.
In this connection, what is known as the stability limit of transistors should be mentioned.
As explained in section 11.2, germanium transistors are temperature-dependent. Almost all the transistor parameters affected by the temperature (I_{co}, α, V_{be}) tend to increase the collector current as the junction temperature rises. It is therefore important to provide an outlet (metal housing, vanes, etc.) for as much as possible of the heat generated by dissipation at the collector-base junction. If the outlet is inadequate, or if the ambient temperature rises unduly, the working point shifts.
This affects class B amplifiers most of all, because they are loaded to the maximum limit of dissipation. Owing to the displacement of the working point, the original class-B setting changes to class A, at which still greater dissipation takes place. The result is a further rise in the junction temperature, producing another change in the operating conditions, which raises the temperature still further, and so on until the transistor finally burns out, or "runs away" as it is called.



CHAPTER V

PROPERTIES AND CONSTRUCTION

V. 1. FREQUENCY LIMIT AND TRANSISTOR NOISE

As explained in section II. 2, the emitter of a p-n-p junction transistor injects holes into the base crystal owing to the fact that the applied voltage ejects electrons from their bonds in the base. The holes, attracted by the negative field of the collector, drift towards the collector. On the way, however, they encounter free electrons, which may deflect them or eliminate them altogether (by recombination). When the concentration of electrons and holes is not very high, their mutual action is not considerable. With high concentrations, however, there is more probability of recombination, so that not all the holes produced are able to reach the collector. What is known as the average life of the charge carriers is evaluated accordingly.
Again, the field between emitter and base retards the holes, so that it is only their initial velocity that enables them to pass through the field to the base. The result is that the holes have lost much of their initial velocity by the time they enter the base, and are therefore all the more readily deflected by the free electrons. Re-acceleration takes place only when the holes reach the field near the junction of base and collector; hence their path through the base crystal is rather tortuous. If all the charge carriers travelled the same distance, the only effect of deflection would be a phase difference between the input and output signals. As it is, however, the distance is purely a matter of chance and therefore varies for individual carriers; hence the difference in transit time may be such that one hole reaches the collector half a period before the next. It will be evident that the two then contribute nothing to the overall amplification. The result is a decrease in the amplitude of the output current. The probability of two holes being in anti-phase is greater at high frequencies, which means that the gain decreases as the frequency increases.
Transistors are therefore restricted to a certain frequency limit, namely, the frequency at which power amplification falls by 3db (to roughly 0.71 of the input voltage).
The frequency limit can be raised, and the phase difference thereby reduced, by employing suitable materials and shapes. Germanium of low conductivity is employed to reduce recombination. Again, the base layer should be as thin as possible, although it must not be too thin in view of the difficulty of manufacture and the possibility that the current density near the base connection will increase unduly, thus raising the base resistance. Moreover, if the base is very thin, the internal capacitances of the transistor are high. Fig. 39 is the equivalent circuit of the input of a junction transistor with grounded emitter.

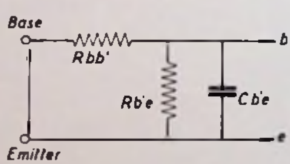


Fig. 39

In this diagram $R_{bb'}$ represents the non-reactive resistance of the base material and $C_{b'e}$ the capacitance, which depends mainly on the base thickness and the emitter current (recombination). The resistance in parallel, $R_{b'e}$, represents losses due to leakage current etc. It will be evident that the lower $R_{bb'}$ and $C_{b'e}$, the higher the frequency limit and the greater the proportion of V_w contributing to the amplification.
Transistors with frequency limits up to 600 Mc/s in grounded-base circuits are already obtainable. Another peculiar effect associated with the transit time in the base crystal is what is known as "hole storage", occurring as a result of high current density in transistors employed as switches (see V. 2).
With the transistor controlled by means of a square-wave voltage applied to the base, output voltage cut-off lags slightly (roughly one or two microseconds) behind control voltage cut-off. The result is that the square-wave pulse is wider at the output than at the input. Special methods of switching enable this effect to be reduced, however.
Noise in transistors differs from the ordinary kinds of noise, i.e. thermal, shot, and so on. The noise power produced in transistors varies in such a way as to be for all practical purposes

inversely proportional to the frequency. In effect, then, every "octave" of the overall frequency range includes more or less the same amount of noise power.
This noise may be defined as the ratio of the total noise power at the input (neglecting noise produced by the load impedance) to that portion of the thermal noise present at the output which is produced by the generator resistance (R_g) at the input.
Usual values, as measured at 1000 c/s, are 40—60 db for point-contact, and 10—15 db for junction transistors.
Transistor noise depends on the working point. Whereas noise produced by the emitter is mainly independent of the collector voltage V_{ce}, collector noise depends very much on this voltage. To reduce noise as much as possible, transistors generally should be operated with low collector voltage, i.e. of the order of 1 V.

V. 2. TRANSISTORS AS SWITCHES

Junction transistors have what are for all practical purposes ideal pentode characteristics. In other words, they can be operated with very low collector voltages. This is a useful feature of transistors employed as switches, as will now be explained.
With the collector resistance constant, increasing the base current causes the collector voltage to fall to some tenths of a volt as soon as the bend in the I_c/V_c characteristic is passed. Hence the transistors can carry heavy collector current without exceeding the maximum limit of collector dissipation. With, say, 0.15 V and 200 mA in the collector, collector dissipation is only $0.15 \times 200 = 30$ mW. The transistor is then equivalent to a resistor of only 3 Ω.
At the same time, the overall collector dissipation is augmented by the dissipation of the emitter/base diode, particularly when the current is heavy.
Supposing that in the present example the α' of the transistor is 25, the base current will be

$$\frac{I_c}{\alpha'} = \frac{200}{25} = 8 \text{ mA (approx.)}.$$

From the I_b/V_{bc} characteristic of, say, the OC 76, the required base/emitter potential is found to be roughly 1.5 V. In this case, then, the base dissipation is $1.5 \times 8 = 12$ mW. With heavier current it may happen that the base dissipation exceeds the collector dissipation.
This point should be kept in mind when designing transistorized switching circuits.
The OC 76 is especially suitable for switching applications by virtue of its very low input resistance, consistent with low V_{bc} ($= I_b \times R_{b'b}$) and therefore also low base dissipation.

V. 3. PHOTO-TRANSISTORS

As we saw in section II. 2, the current through p-n junctions depends very much on the amount of light falling on them. For this reason normal transistors, and the majority of modern diodes, are sealed into capsules which exclude light for example by means of a coating of black lacquer on a glass envelope or a metal cap.
This light sensitivity is, however, turned to good purpose in photo-diodes and photo-transistors (fig. 40). Photo-transistors also amplify the current produced by light.

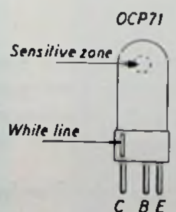


Fig. 40

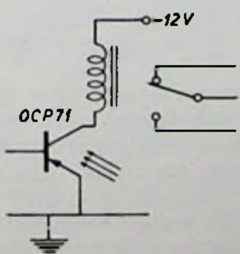


Fig. 41. Photo-transistor OCP 71 as photo-electric relay.

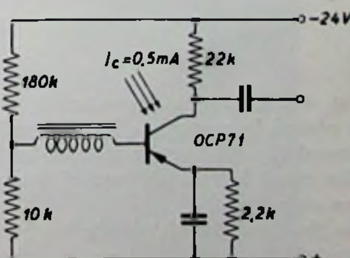


Fig. 42. Stabilised circuit for modulated light.

It was shown in III. 4, that with the base of the transistor "floating", an amplified leakage current $I_{co}' = \alpha' I_{co}$ is produced. Since I_{co} depends partly on the amount of incident light, phototransistors, operated with floating base for switching applications, amplify the initial photo-electric current. The region most sensitive to light is close to the emitter/base junction. Because the transistor capsule is filled with a substance resembling vaseline, which diffuses light, the sensitivity is not noticeably directional. In the case of type OCP 71, maximum current is obtained with the controlling luminous flux falling at right angles to the connecting wires, on the side of which the type number is printed. Roughly 30 % of the optimum sensitivity is attained with light falling vertically on the end, and roughly 50 % with light falling on the side, of the transistor.
The operating range of the OCP 71 covers the visible spectrum as far as the infra-red, e.g. the radiation from an Infraphil lamp, or the glow of a cigarette.
A disadvantage of this transistor is that it is also highly sensitive to temperature variations; as with ordinary transistors, therefore, stabilizing circuits are employed to compensate this as far as possible.

V. 4. MANUFACTURE AND CONSTRUCTION OF TRANSISTORS

As explained at the beginning, semi-conductors are made of exceptionally pure materials, to which controlled amounts of certain impurities are added. The proportion of impurities introduced into germanium as employed for transistors is of the order of 10^{-5} to 10^{-8} %.
To enable the amount of impurity added to be accurately controlled, the original germanium must first be made as pure as possible.
First, germanium obtained from the mineral germanite, or as a by-product of zinc manufacture, is purified by chemical means (i.e. by converting it into germanium tetrachloride, distilling this several times and then converting the distillate into germanium dioxide by hydrolysis).
Next, the dioxide is reduced in hydrogen to germanium powder, which is melted at approximately 950° C and cast into a rod. At this stage the germanium is already so pure that ordinary methods of chemical analysis do not reveal any trace of impurity in it. Nevertheless, it is not pure enough to be used in the manufacture of transistors. Further purification is carried out by other, less conventional, methods.
When molten metal sets, it invariably crystallizes with fewer impurities than were present in the original melt *). Advantage is taken of this in the process known as zone melting.

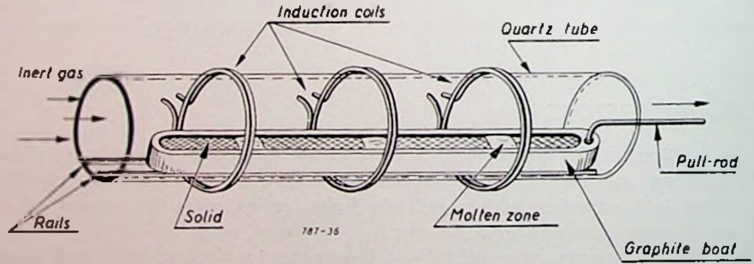


Fig. 43a

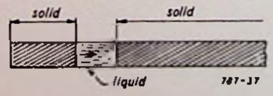


Fig. 43b

Fig. 43a illustrates the principle of the process as applied to germanium. The germanium rod is placed in a graphite boat travelling on rails through a quartz tube, which contains inert gas. Round the tube are a number of coils carrying heavy alternating current at radio frequency. The result is that portions of the germanium are melted by eddy current induced in them by the coils.
As the boat travels slowly to the right, the molten zones shift gradually to the left, thereby crystallizing purer and purer germanium at the right-hand sides of the zones and increasing the impurity concentration in the left-hand portion of the melt (fig. 43b).

*) Not strictly correct, but full explanation is beyond the scope of this book.

This has what may be described as a "sweeping" effect, in that the purity of the right-hand portion increases with repeated sweeping until all the impurities are finally concentrated in the left-hand portion of the rod. This is then cut off.

A very accurately computed amount of a given impurity is now introduced into the germanium, to convert it into either p-type or n-type germanium, depending on the nature of the impurity. What happens after this determines whether the final product will be an "alloy junction", or a "grown junction", transistor, and whether the composition will be p-n-p or n-p-n.

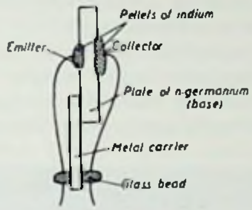


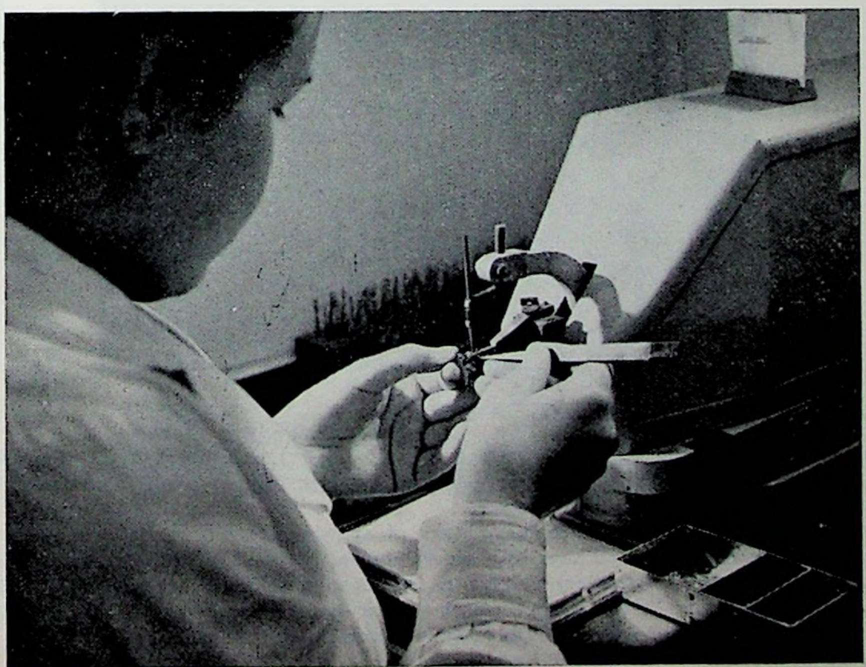
Fig. 44

Alloy junctions are formed by applying a single pellet of metal to each side of a germanium plate and heating them until they dissolve the germanium.

The manufacture of, say, p-n-p transistors begins with a plate of n-germanium, roughly 0.2 mm thick, welded to a metal support.

The first step is to clean the germanium carefully. Single drops of indium are then applied to each side of the plate, and this is heated at a temperature just above the melting point of indium [indium melts at 155° C; germanium at 958° C]. The molten indium dissolves germanium until saturated (amalgamation). The art is to leave a very thin layer of n-germanium unaffected. As the plate cools, most of the dissolved germanium separates again (segregation) just where it is needed, that is, next to the unaffected n-germanium. The layers of separated germanium still contain a certain amount of indium, in other words they consist of p-germanium.

Usually, the heating is carried out in stages, to enable the thickness of the residual n-layer to be controlled. The thinner the layer, the greater the amplifying power of the transistor. As the emitter is slightly smaller than the collector, the latter for all practical purposes encloses it. Fig. 44 shows a cross section of an alloy-junction p-n-p transistor. Grown junction transistors are produced as follows.

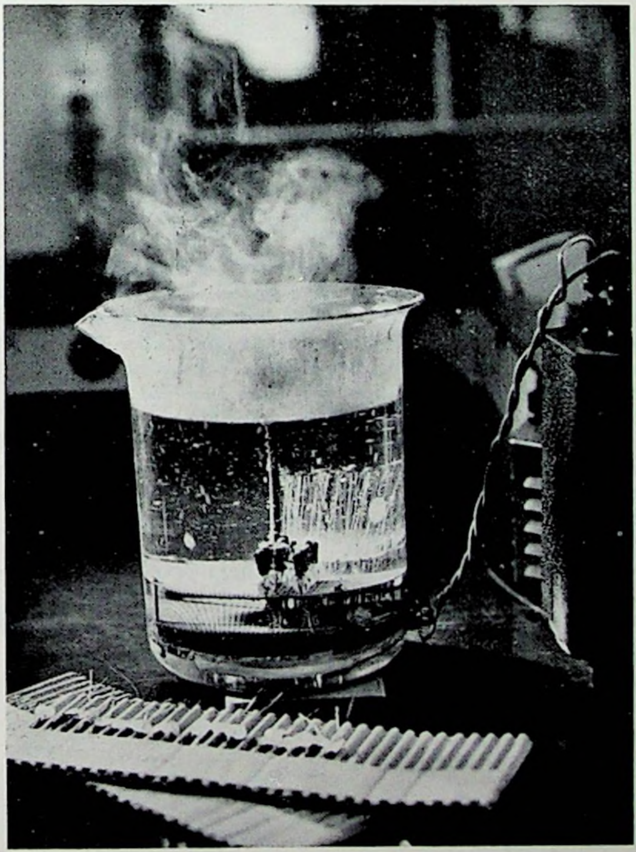


What is known as a seed crystal, say, a tiny rod of pure germanium is dipped into a graphite crucible containing pure molten germanium. The temperature is then lowered, with the result that the crystal "grows". Next, sufficient arsenic or antimony to produce n-germanium with the desired volume resistivity is introduced into the melt. When the crystal has grown a certain amount, a measured quantity of gallium or indium is added to over-compensate the n-conductance and thus produce p-germanium. The crystal is then allowed to grow 50 microns larger, after which enough arsenic or antimony to make the germanium predominantly n-conductive is again added to the melt. In this way an n-p-n single crystal is formed. It is sawn into many rods, each of which will make an n-p-n transistor.
The rods are etched to remove the superficial portion of the crystal lattice damaged by the saw. Attaching the connecting wires, especially to the p-layer, not visible to the naked eye, is a task for specialists. It is done by clamping the rod in a special measuring device which draws the wire along the rod and automatically welds them together at the proper position.

V. 5. NOTES ON THE CARE OF TRANSISTORS

Although fairly robust mechanically, transistors must be treated carefully from the electrical point of view, since their characteristics are very different from those of the more familiar and the voltages involved are very much lower.
Nowadays, no one is likely to make the mistake of applying the full anode voltage of a tube to the control grid. What is perhaps less widely known, however, is that it would be equally disastrous to apply the full battery voltage to the base of a transistor without a resistor in series. Since transistors "die" without warning, that is, noiselessly and without preliminary sparking, it is necessary to become thoroughly accustomed to the entirely different values associated with them.

It is particularly important to avoid overheating, whether during soldering or as a result of overloading. Flat pliers should be used to hold the leads and act as a "thermal shunt" during soldering-in. Overloading results in heating of the transistor, thus increasing the leakage current. The working point then shifts, thereby increasing the collector current, and also the dissipation, and this is cumulative, the transistor finally running away. With overloading, the junction temperature rises so quickly that transistors thus affected often run away before the heat generated becomes noticeable from the outside.
Other effects likely to damage, if not destroy, transistors are heavy capacitor discharges, switching-on phenomena affecting transistors inserted in circuits already under tension, etc. Accordingly, it is advisable to switch the voltage off before making any changes in the circuit. For the same reason, plug-in type transistor bases are unsafe in view of the temptation to withdraw transistors from sets under tension and then reinsert them a moment later.
The polarity of the applied voltages is another important factor. Bear in mind that with p-n-p transistors the "H.T." line is negative. On the other hand, if the particular transistor happens to be n-p-n, the collector should be positive. Reversing the polarity may destroy the transistor. A point to remember in connection with transistors in switching circuits is that in certain cases the dissipation of the emitter/base diode may equal the collector dissipation.



PART 2 — PRACTICE

CHAPTER VI

AMPLIFIER CIRCUITS

VI. 1. SIMPLE MICROPHONE AMPLIFIER FOR DYNAMIC MICROPHONES

Fig. 45 shows a circuit suitable for low-resistance dynamic microphones. Note that the base of the OC 70 is earthed, so that signal injection is into the emitter. Instead of a separate battery giving positive bias to the emitter, a voltage divider giving the base a slight negative bias with respect to the emitter is provided.
The current consumption of the amplifier is very low, enabling it to operate almost indefinitely with a 1½ V dry cell.

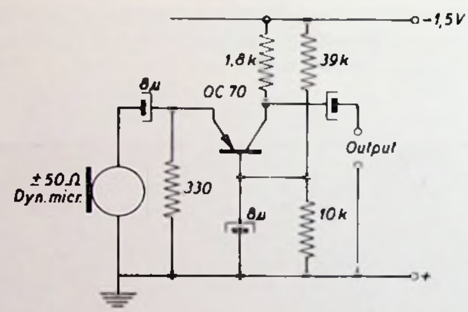


Fig. 45

VI. 2. AN INTERESTING PRE-AMPLIFIER STAGE

Generally speaking, transistors are employed as circuit elements for low-frequency amplifier purposes mainly where the supply voltage has to be taken from a small battery. Applications may arise, however, in which it is desired to drive the transistor from a supply voltage already available, in order to exploit other advantages which they offer. For example, pre-amplifier elements occupying very little space, providing a large voltage gain at a high impedance input and having low noise, microphony and hum, are often required for use with stationary thermionic amplifiers.
The transistor provides a convenient solution to this problem, as will be clear from the following description. The stage differs slightly from the customary power-gain transistor circuits, but shows that a transistor can also give a voltage gain such as would normally require two stages when electron tubes are utilized. To save using a separate battery, the transistor should be driven from the anode voltage of the main amplifier.
Diagram 46 shows the circuit of the pre-amplifier. The supply is 250 V. The transistor OC70 functions in a grounded-emitter circuit, since this is capable of providing the maximum voltage amplification. The source is assumed to be a microphone with a low impedance of from 30—50 ohms.

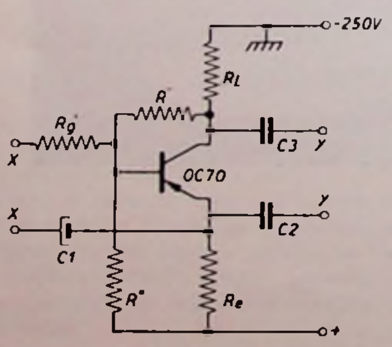


Fig. 46

R_g'	=	100 Ω	(5 %)
R'	=	100 kΩ	(5 %)
R''	=	470 kΩ	(5 %)
R_e	=	5.6 kΩ	(5 %)
R_L	=	330 kΩ	(5 %)
C_1	=	100 μF	
$C_{2\text{-}3}$	=	0,1 μF	

Resistances		$\pm$ 5 %
Supply voltage	$-V_o$	$\pm$ 10 %
Ambient temperature	T_{amb}	$\leq$ 45 °C

The d.c. stabilization also functions as a dynamic negative feed-back, the resistor R' being the most effective component. It actually provides a parallel negative feed-back, the chief effect of which is to keep the current gain constant. Another requirement is a constant voltage gain, because the low impedance of the microphone is at the input. This would call for a series negative feed-back in which a d.c. stabilizing circuit with a potential divider would have to be used for the base voltage, a practice familiar from conventional amplifier stages for small signals. Such a device, however, does not permit the desired stabilizing effect to be attained. The stabilizing effect achieved with the resistor R' in the circuit as suggested, on the other hand, is advantageous because the very large load resistance contributes almost in its entirety to the stabilizing effect.
The working point of the transistor lies nominally at

$$-I_c = 0.7 \text{ mA}$$
$$-V_{ce} = 4 \text{ V}$$

A microphone with an impedance of from 30 to 50 Ω is assumed to be the signal source, and the high-impedance input of a thermionic amplifier to be the load.
If the input potential is 5.5 mV (at the terminals), the stage supplies an output potential of 1.8 V (at the terminals yy), thus the voltage gain is

$$g_u \approx 330 \text{ (25.5 db)}.$$

The frequency response is very good, the $-$3db points being at 15 c/s and 20 kc/s. The input impedance at terminals xx amounts to 200 Ω, and the output impedance at terminals yy is approximately 4 kΩ. A disadvantage of the circuit illustrated in fig. 46 is that the output is "floating" since, normally, the output voltage will be applied between earth and the grid of the first thermionic amplifier stage.
The modified circuit as at fig. 47 may then be used. This circuit avoids risks for the microphone in case of short circuit to earth.

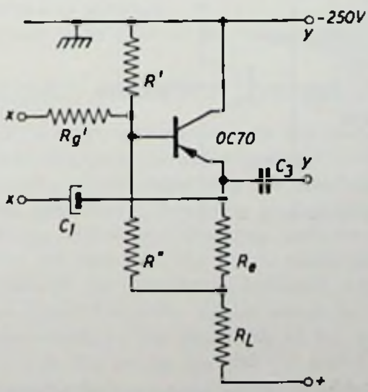


Fig. 47

R_g' = 100 Ω
R' = 100 kΩ
R" = 470 kΩ
R_e = 5.6 kΩ
R_L = 330 kΩ
C_1 = 100 μF
C_3 = 0.1 μF
(tolerances $\pm$ 5 %)

VI. 3. SIMPLE SIGNAL TRACER

This is really a two-stage transistor amplifier, very useful as a means of tracing faults in A.F. amplifiers.
Variable resistor R_1 is adjusted for roughly 1 mA collector current. Operating conditions depend on the D.C. resistance of the headphones employed. If necessary, the circuit will drive a small

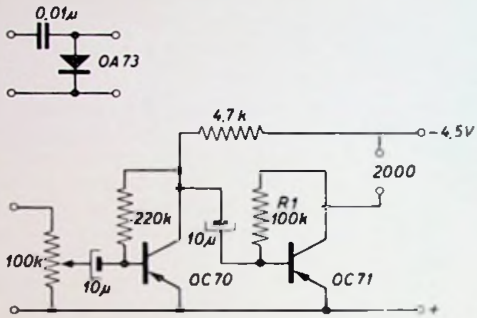


Fig. 48

loudspeaker instead. A plug-in diode is provided, enabling the tracer to be used also for testing R.F. and I.F. amplifiers (fig. 48).

VI. 4. TELEPHONE-CONVERSATION AMPLIFIER

This makes use of stray flux from the line transformer included in most telephones. As with certain hearing aids, there is a pick-up coil, to be held near the telephone. The alternating voltages generated in this coil are amplified, enabling others to "listen-in" to the telephone conversation.
Transformer coupling is employed in this amplifier, because it cuts down the required supply voltage and enables more efficient amplification to be obtained (fig. 49).

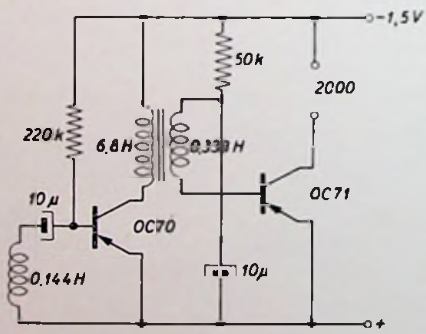


Fig. 49

VI. 5. TRANSISTOR VOLTMETER

Inexpensive voltmeters can be rendered highly sensitive by means of a simple D.C. transistor amplifier, as follows. The transistor is included in one of the arms of a bridge circuit whose other arm comprises two resistors R_1 and R_2 in series. Temperature compensation is dispensed with; therefore the meter must be zeroized before each test. To do so, short the input terminals (test prods) and adjust R_2 so as to bring the meter reading to zero.
If the ambient temperature is on the high side, it may happen that I_{co}' of the transistor is too high to enable the meter to be zeroized simply by means of R_2. In that case, fixed resistor R_1 should be shorted through S_3. Calibrate the meter to known voltages by means of R_3, controlling the sensitivity. R_2 and R_3 are both wire-wound. Adjust the range by means of S_1. The size of the range-resistors to be included in the circuit depends on the meter employed.

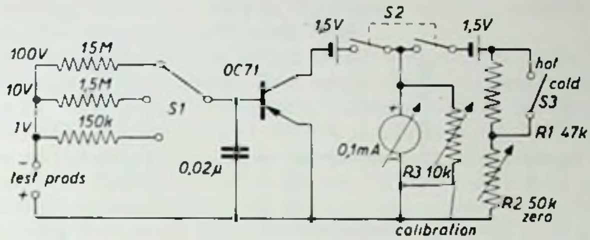


Fig. 50

The values indicated in the diagram (fig. 50) are as computed for a milliammeter reading 0.1 mA and a transistor type OC 71. For a meter reading to 1.0 mA, the range resistance should be divided by 10 (to 1.5 MΩ, 0.15 MΩ and 15 kΩ).
The accuracy of the transistor voltmeter is of the order of 5—10 % at room temperature (15° C—25° C); hence it is quite suitable for use by amateurs.

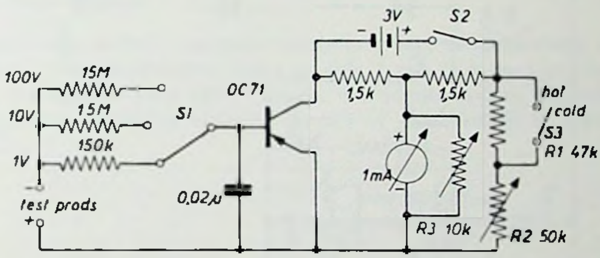


Fig. 51

Although in theory D.C. amplifiers with two or more stages are also possible, in practice they are very difficult to build owing to the temperature dependence of transistors generally and the spread amongst individual transistors.
Fig. 51 shows a slightly different circuit with only one battery, enabling a 3 V dry cell to be used complete instead of divided.
The 0.02 μF capacitor is to protect the meter when S_1 is changed over.

VI. 6. A FREQUENCY METER FOR LOW FREQUENCIES

Familiar devices for measuring frequency in the low-frequency ranges use circuits in which the frequency-dependent charging current of intermittently charged capacitors is utilized to indicate the frequency. Circuits of this type may be employed for measuring both sinusoidal voltages and periodic pulses and hence they may be considered for application in industrial revolution counters. When these circuits are designed with transistors they are so reliable that, as a rule, the indicating errors are less than the errors made when a reading is taken from the instrument.
The complete circuit of the frequency indicator and the corresponding values are shown in fig. 52. The initial three transistor stages serve to convert the sinusoidal input signal into a virtually square-wave voltage. The sensitivity of the equipment can be matched to the amplitude of the input signal with the aid of the 100 kΩ and 1 mΩ resistors at the input. The diodes that follow, which are connected in anti-parallel, and the 10 kΩ resistor protect the input transistor against test voltages that are too high. The virtually square-wave voltage appearing at the output of the third transistor controls the output transistor OC72, which functions as a switch. That the positive level of the controlling square-wave voltage always remains at zero potential is ensured by the diode in front of the base.
The collector voltage of the transistor OC72 changes periodically between its two switching values, i.e. the very small knee voltage (approx. 0.3 V) and the battery voltage V_o (4.5 V), less the small voltage drop across the load resistor R_L due to the residual current.
Basically, the frequency measurement is effected as follows. A capacitor is periodically charged on the battery voltage and discharged on the knee voltage. If the charge (or discharge) current is

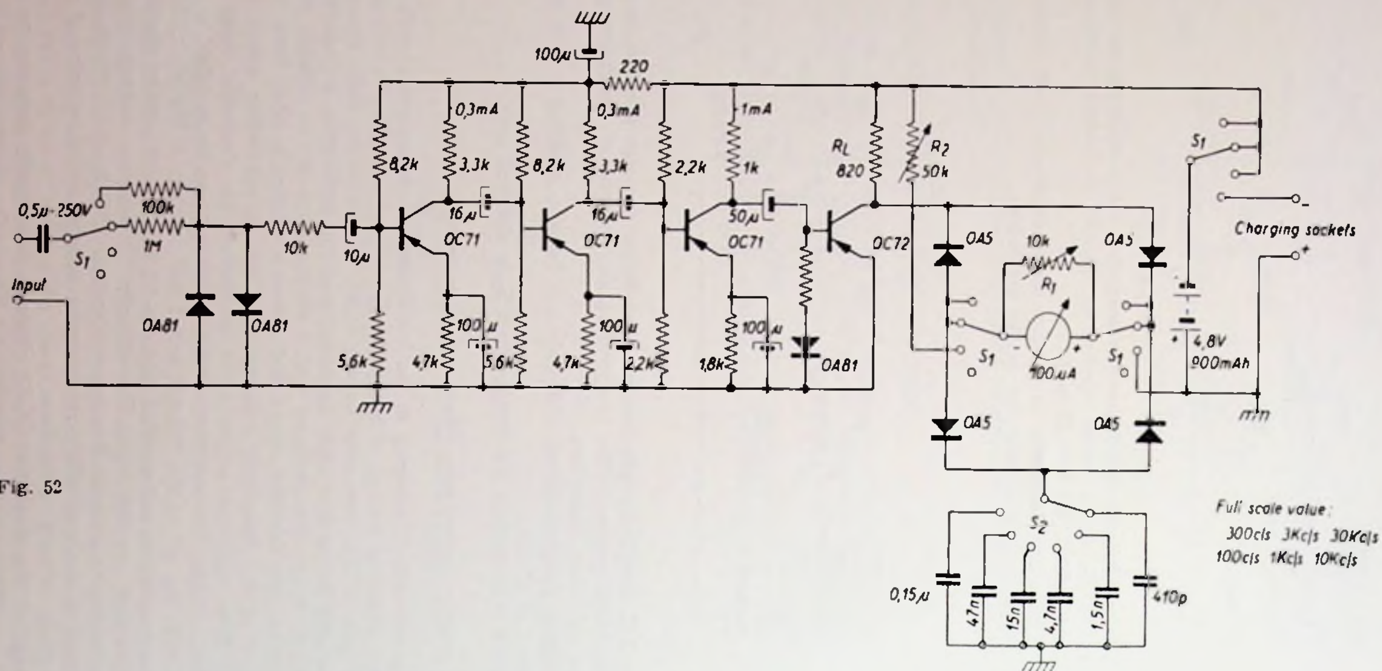

100µ
220
0,3mA
0,3mA
1mA
8,2k
3,3k
8,2k
3,3k
2,2k
1k
RL
820
R2
50k
S1
0,5µ-250V
100k
1M
S1
10k
10µ
16µ
OC71
16µ
OC71
50µ
OC71
OC72
Input
OA81
OA81
100µ
100µ
100µ
5,6k
4,7k
5,6k
4,7k
2,2k
1,8k
OA81
OA5
10k
R1
OA5
Charging sockets
S1
S1
100µA
4,8V
900mAh
OA5
OA5
S2
0,15µ
47n
15n
4,7n
1,5n
410p
Full scale value:
300c/s 3Kc/s 30Kc/s
100c/s 1Kc/s 10Kc/s

Fig. 52

measured by an integrating network, a reading is obtained which is proportional to frequency, provided the charging time-constant is sufficiently small.
The measuring instrument is included in a bridge circuit consisting of diodes so connected that both charge and discharge currents always flow across the instrument in the same direction. On the highest range (30 kc/s), when V_1—V_2 ≈ 4 V (battery almost discharged), C = 410 pF. R_i of the instrument = 2500 Ω a current i of 100 μA is obtained at full deflection.
This particular bridge circuit is equipped with the germanium gold-bonded diodes OA5, and hence their resistances in the forward direction do not influence the results.
The instrument is practically independent of the ambient temperature, as the knee voltage (which would be the only quantity that could affect the measurement) varies only slightly with temperature.
The entire circuit, including accumulator, is contained in quite a small space, and thus the equipment may be used for carrying out measurements directly at the place of measurement. The switch S_1 when set to the "off" position, enables the accumulator to be recharged at the charging sockets.
To cover the measuring ranges accurately, the charging capacitors need, in view of spread, to be selected as carefully as possible.

VI. 7. SENSITIVE D.C. RELAY

With a slightly different type of D.C. transistor amplifier a very sensitive relay can be constructed. Given, say, a telephone relay modified to make at only 1 mA (e.g. by removing all the contacts but one and giving the relay a very light trigger), the sensitivity can be increased to roughly 20 μA by including the relay in the collector circuit of a transistor. In view of possible self-

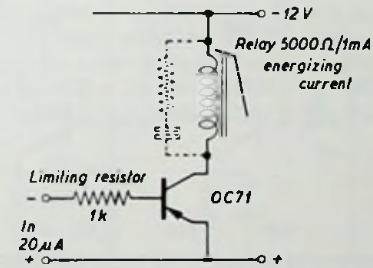


Fig. 53

induced pulses from the relay, it is advisable to limit the battery-voltage to 15 V at any rate, unless the induced voltage is limited by means of a resistance-capacitance shunt as shown in dotted lines in fig. 53, in which case a 22½ V hearing-aid battery may be used instead.

VI. 8. SENSITIVE R.F. RELAY

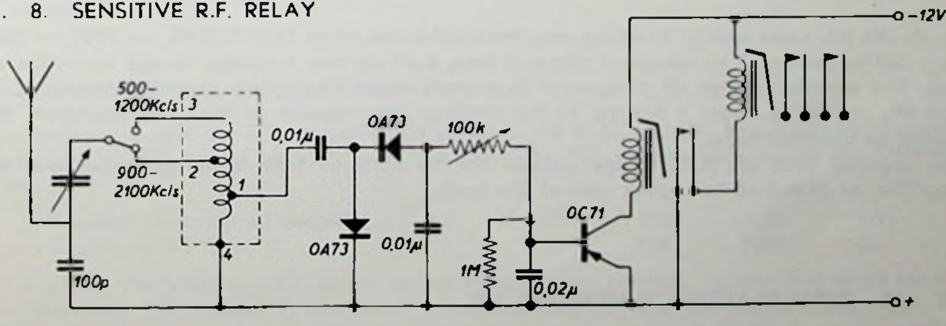


Fig. 54

COIL DATA
The tuning coil is wound on a (plastic) former with flanges at distances of 2 mm, an outer diameter of 20 mm and a breadth of 12 mm at the bottom. A powdered iron core should be inserted. The total number of turns is 360. Use 0,1 E (SWG no. 42 — AWG no. 38 — BWG no. 36) magnet wire.
Start winding at connection no. 4; make tap no. 1 at 24 turns, no. 2 at 220 turns.

If a germanium diode be added to the above-mentioned circuit, as indicated in fig. 54, the relay will also respond to A.F. and R.F. alternating current up to about 200 Mc/s. It could then be used, say, as a means of switching on the radio first thing in the morning. For this purpose, however, it should be used in combination with a heavier-duty relay, actuated by the sensitive relay so as to switch on the voltage for the (transistorized) receiver.
The sensitive relay must not be operated with heavy current, or the contacts will burn out.

VI. 9. TRANSISTORIZED MEASURING BRIDGE

This is really a tuned amplifier operating at 1000 c/s (fig. 55).
The transformers required can be wound on discarded output-transformer cores. To enable the self-inductance to be regulated it is best to make the air-gap variable. This can be done by placing a thin sheet of rubber under one side of the I-core and pressing this in by means of a screw (with nut soldered to mounting bracket). When once the I-core is properly positioned, pour spraying lacquer between all the laminations. When dry, this fixes the assembly permanently.

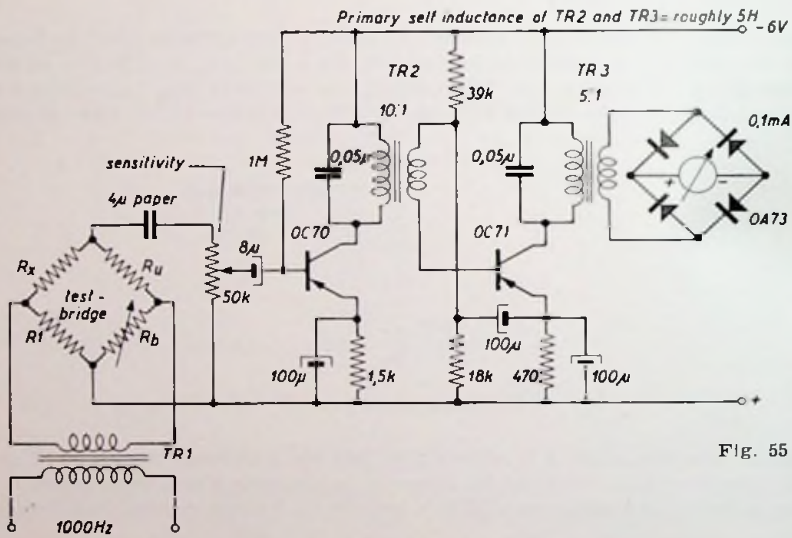


Fig. 55

To do the job more neatly, however, use ferroxcube pot cores (e.g. D25/16 — 1000 — 3B2); they can be regulated by moving a piece of tape, built up into a wedge, to and fro in the air-gap. This provides a high Q component in a small space. The primary self-inductance of TR_2 and TR_3 should be about 5 henrys; as the winding data depend on the core employed, they cannot be specified here.
The required 1000 c/s A.C. bridge voltage can be obtained from a transistor oscillator with amplifier, as described in another part of this book.

VI. 10. THREE-TRANSISTOR HEARING AID

CIRCUIT DESCRIPTION

Fig. 56 shows the circuit diagram of a transformer-coupled hearing aid equipped with two transistors OC 70 and one transistor OC 71. The hearing aid has been designed for supply by a battery with a nominal voltage of 1.2 V. The current drain is 4.6 mA, so that the total energy that must be supplied by the battery is approximately one tenth of that required by a

conventional hearing aid equipped with tubes. The nominal output power is 1.2 mW at a distortion of 5 %.
The coupling between the different stages of the amplifier can be achieved either by means of a resistance-capacitance network or by means of a transformer. (An RC-coupled hearing aid is described in the next Section). Transformer coupling offers the possibility of obtaining a nominal gain of 81 db with three transistors in common emitter connection notwithstanding the loss of gain caused by the negative feedback. The transformer coupling moreover offers the possibility of applying an efficient d.c. stabilisation even at the low supply voltage of 1.2 V. An autotransformer is used for the coupling between the driver and the output stage, a higher gain thus being possible than could be obtained by means of a transformer with separate windings.

D.C. STABILISATION OF THE WORKING POINTS

The working point of the transistors in the input and in the driver stage is effectively stabilised, and the influence of temperature fluctuations on the properties of these transistors is compensated. Negative feedback is, moreover, applied to the input stage and to the driver plus output stages for stabilising the gain.
The nominal electrical power gain is 81 db, which amounts to an overall acoustical gain of 60 db. Due to the negative feedback and the d.c. stabilisation the spread in overall power gain is reduced to ± 2 db for various conditions of ambient temperature, battery voltage and production spread in transistor parameters.
The working point of the input transistor (nominal $+ I_e = 0.5$ mA) and that of the driver transistor (nominal $+ I_e = 0.35$ mA) are stabilised by feeding the base of these transistors by voltage dividers shunted across the battery and including a resistor in the emitter circuit. Temperature fluctuations are likely to influence the direct current due to variation of the voltage between the base and emitter. The latter variation is compensated by shunting across part of the voltage dividers a resistor with a negative temperature coefficient.
The working point of the output transistor (nominal $-I_c = 3$ mA) is adjusted by means of a variable resistor of 8 to 30 kΩ. Applying d.c. stabilisation in the output stage would result in a rather high loss of output power.

NEGATIVE FEEDBACK

The negative feedback of the input transistor is obtained by including an un-bypassed resistor in the emitter circuit: The effect of this resistor is analogous to that of the familiar un-bypassed cathode resistor in a tube circuit.
Common negative feedback is applied to the driver and output transistors by feeding back part of the output current to an additional winding S_3 of the first transformer T_1. To avoid the number of turns of this winding becoming impracticably small, only part of the output current is fed back.

ADJUSTMENT OF THE OUTPUT STAGE

The maximum output power can be limited by adequate choice of the bypassed clip-in resistor R_{15} included in the supply lead of the output transistor. This resistor should have the following value:

Required max. output power	1.2	0.5	0.2	0.05	mW
Clip-in resistor R	0	120	390	1000	Ω

For adjusting the working point of the output transistor the variable resistor of 8 to 30 kΩ has been provided. This resistor is connected in such a way between the clip-in resistor and the telephone that sharp limiting of the output power is ensured at any value of the clip-in resistor without need of readjusting this variable resistor. The circuit has been designed for an electromagnetic type of telephone with an impedance of 270 Ω at 1000 c/s (d.c. resistance 90 Ω).
The frequency characteristic of the hearing aid can be modified by means of the capacitor in the collector circuit of the OC 71. This method of tone control offers the possibility of using a very simple combined tone control and on-off switch.

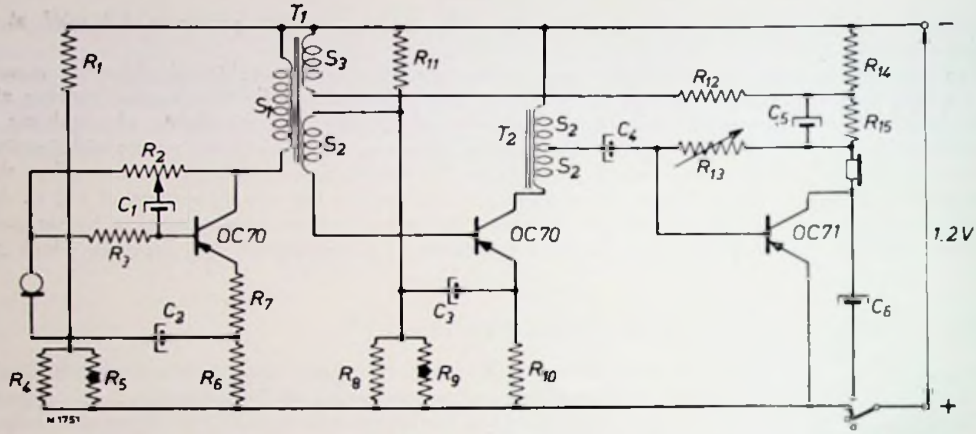


Fig. 56. Circuit of a transformer-coupled hearing aid with three transistors.

R_1 = 220 Ω
R_2 = 0.3—500 kΩ hyper log.
R_3 = 680 Ω
R_4 = 1500 Ω
R_5 = N.T.C. resistor
Resistance at 25 °C 2200 Ω
Temp. coefficient —3.7 %/°C at 25 °C
Tolerance at 25 °C 10 %
R_6 = 220 Ω
R_7 = 180 Ω
R_8 = 1800 Ω
R_9 = N.T.C. resistor
Resistance at 25 °C 1500 Ω
Temp. coefficient —3.4 %/°C at 25 °C
Tolerance at 25 °C 10 %

R_{10} = 270 Ω
R_{11} = 3900 Ω
R_{12} = 120 Ω
R_{13} = 8—30 kΩ
R_{14} = 10 Ω
R_{15} see below
C_1 = 10 μF
C_2 = 10 μF
C_3 = 10 μF
C_4 = 10 μF
C_5 = 10 μF
C_6 = 0.8 μF
All resistors are of the 5 % tolerance type

Transformer T_1

Primary S_1	2700 turns
Inductance of S_1	7.2 H at 0.5 mA
D.C. resistance of S_1	860 Ω
Tolerance	20 %
Secondary S_2	600 turns
D.C. resistance of S_2	300 Ω
Tolerance	20 %

Transformer T_2

Autotransformer

Winding S_1	2178 turns
Winding S_2	622 turns
Inductance of $S_1 + S_2$	8.6 H at 0.5 mA
D.C. resistance of S_1	650 Ω
D.C. resistance of S_2	130 Ω
Tolerance	20 %

Transformer wire — Enamelled copper

T_1S_1 = 0.045 mm
T_1S_3 = 0.100 mm
T_2S_2 = 0.060 mm
T_1S_2 = 0.040 mm
T_2S_1 = 0.045 mm

Microphone

Magnetic

D.C. resistance	200 W
Tolerance	20 %
Impedance	1000 Ω at 1 kc/s
Sensitivity	0.2 mV/μbar at 1 kc/s on 1 kΩ

Telephone

Magnetic

D.C. resistance	90 Ω
Tolerance	20 %
Impedance	270 Ω at 1 kc/s
Sensitivity	125 phone for 0.6 mW at 1 kc/s

Clip-in resistor R_{15} for decreasing the output power W_0

R (Ω)	$W_{0\,max}$ (mW)
0	1.2
120	0.5
390	0.2
1000	0.05

VOLUME CONTROL

The volume control by means of the hyper logarithmic potentiometer of 0.3 kΩ to 0.5 MΩ is based on the use of an electromagnetic type of microphone with an impedance of 1000 Ω at 1000 c/s, which ensures a good compromise between optimum noise match and optimum gain match without requiring the use of a matching transformer between microphone and input transistor. This method of volume control does not affect the signal-to-noise ratio at high signal levels, as is often experienced with other methods of volume control. When the amplification is reduced by means of the volume control the input impedance of the first transistor OC 70 is also decreased and since the microphone impedance is inductive, this would result in the frequency characteristic falling off towards the high frequencies. This effect is almost entirely suppressed by including a resistor of 680 Ω in the base lead of this transistor.

PERFORMANCE

The influence of the spread in the transistors, of battery voltage variations between 1.1 V and 1.3 V and of temperature fluctuations between 15 °C and 35 °C are tabulated below.
The table below shows the power gain spread figures that will be found when the power gain is calculated by combining the most extreme values of transistor parameters within their spread figures and component values within their tolerances, the battery voltage ranging from 1.1 V to 1.3 V and the ambient temperature ranging from 15 to 35 °C. In practice it is hardly likely that these combinations of extreme conditions occur.

Table A. Combination of extremes in transistor parameters, circuit tolerances, battery voltage and temperature *).

Transistors	Circuit	Battery voltage	Temperature	Power gain
minimum	minimum	1.1 V	15 °C	85.7 db
maximum	maximum	1.3 V	35 °C	74.9 db

Table B. Combination of extremes in transistor parameters and circuit tolerances, at the nominal battery voltage and the nominal temperature *).

Transistors	Circuit	Battery voltage	Temperature	Power gain
minimum	minimum	1.2 V	25 °C	75.7 db
nominal	nominal	1.2 V	25 °C	81.1 db
maximum	maximum	1.2 V	25 °C	85.3 db

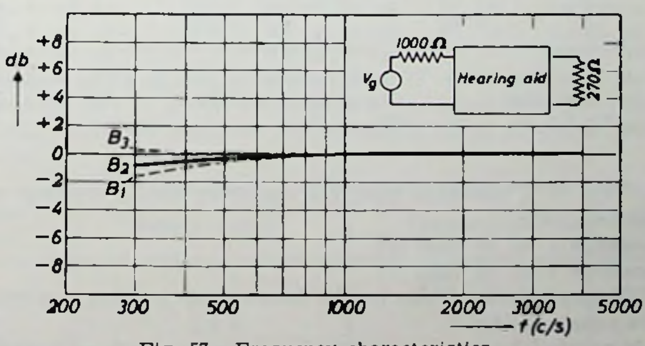


Fig. 57. Frequency characteristics.

*) A minimum (maximum) transistor is understood to be a transistor whose parameters within their spread figures are combined in such a way that minimum (maximum) gain results.
A minimum (maximum) circuit is understood to be a circuit whose component values are so combined within their tolerances that minimum (maximum) gain results.
The battery voltages and temperatures have been chosen either for minimum, for nominal or for maximum gain.

The frequency characteristics of the hearing aids according to table B have been plotted in Fig. 57, curve B_1 applying to minimum transistors and a minimum circuit, curve B_2 to nominal transistors and a nominal circuit, and curve B_3 to maximum transistors and a maximum circuit. For plotting these curves the microphone was replaced by a signal generator with an internal resistance of 1000 Ω, and the telephone was replaced by an impedance of 270 Ω.

VI. 11. FOUR-TRANSISTOR HEARING AID

In contrast to the transformer coupled hearing aid described before, this hearing aid is resistance-capacitance coupled. The lower stage gain obtained with this type of coupling necessitates four stages of amplification instead of the three stages required for the transformer coupled circuit.
Due to the application of d.c. stabilisation and negative feedback the spread in overall power gain is limited to ± 3.5 db for diverging conditions of ambient temperature, battery voltage, and production spread in transistor parameters.

CIRCUIT DESCRIPTION

The hearing aid (see Fig. 58) is equipped with three OC 70 transistors and one OC 71 transistor in common emitter connection. It has been designed for supply by a battery with a nominal voltage of 2.4 V.

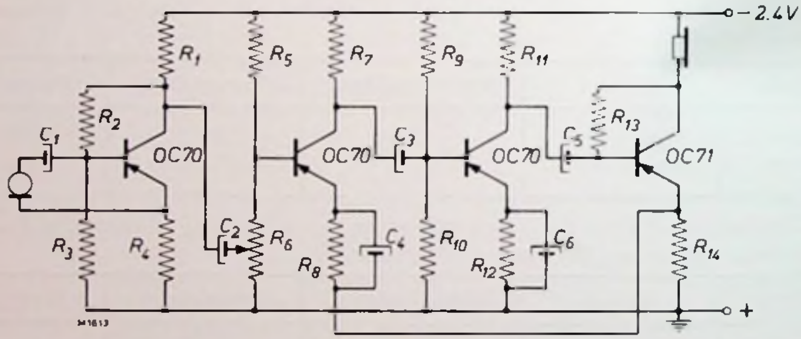


Fig. 58. Circuit of an RC-coupled hearing aid with four transistors.

R_1 = 56 kΩ	R_8 = 1 kΩ	C_1 = 8 μF
R_2 = 2.7 kΩ	R_9 = 22 kΩ	C_2 = 8 μF
R_3 = 33 kΩ	R_{10} = 10 kΩ	C_3 = 8 μF
R_4 = 1 kΩ	R_{11} = 1.8 kΩ	C_4 = 8 μF
R_5 = 18 kΩ	R_{12} = 1 kΩ	C_5 = 8 μF
R_6 = 5 kΩ log.	R_{13} = see text below	C_6 = 8 μF
R_7 = 3.9 kΩ	R_{14} = 2 Ω	

All resistors are of the 10 % tolerance type; the capacitors have a working voltage of 6 V⎓.
Microphone: Electromagnetic
Impedance 1 kΩ at 1000 c/s
Telephone: Impedance 1 kΩ at 1000 c/s
In the three OC 70 stages forward bias for the base-emitter junction is provided by potential dividers across the supply battery.
The input signal is supplied by an electromagnetic microphone with an impedance of 1 kΩ at 1000 c/s.
The volume control is obtained by feeding the output from the first stage to the second stage via the potentiometer R_6 (5 kΩ, logarithmic).
A low collector load resistance is required in the third stage to provide adequate current and voltage swing up to full drive without clipping.

The collector voltage at optimum operating conditions in the OC 71 output stage is 2 V. The collector load is the earphone with an impedance of 1 kΩ at 1000 c/s (resistance 250 Ω), so that the required collector current is 2 mA; this current is obtained by choosing a value for the resistor R_{13}, connected from collector to base, which is α' times the load impedance ($R_{13} = \alpha' R_L = \alpha'$ kΩ), where α' is the current gain of the OC 71 with the emitter grounded. The forward bias of the OC 71 is obtained by means of the resistor R_{13} mentioned above.

D.C. STABILISATION

In the three OC 70 stages the voltage dividers across the battery together with the resistors included in the emitter leads stabilise the working points. In the output stage the resistor R_{13} will tend to stabilise the working point of the OC 71.

NEGATIVE FEEDBACK

In the input stage some a.c. negative feedback from collector to base is provided by R_2, and with the values chosen for R_2 and R_3 the input impedance is approximately 1 kΩ, which matches the impedance of an electromagnetic microphone (1 kΩ at 1000 c/s). This feedback also improves the frequency response and lessens the spread of gain due to variations in temperature and transistor parameters. The un-bypassed resistor R_4 in the emitter circuit does not contribute to the a.c. negative feedback, since the input is fed between the base and emitter.

Negative feedback in the output stage is provided by the resistor R_{13} between the collector and base.

Apart from the stabilisation by feedback provided by R_2 and R_{13} in the input and output stages, gain stabilisation is also achieved by taking 12 db of negative feedback over the last three stages. A feedback voltage proportional to the output current is taken from the top of the 2 Ω-resistor R_{14} in the emitter circuit and applied to the second stage in series with the bypassed resistor R_8. The 2 Ω resistor consists of approximately 18 cm of Eureka wire 0.23 mm diameter, 11.7 Ω/m.

GAIN

The circuit provides a mean electrical power gain of 73 db. The following table shows how the power gain depends on the spread in α' at the nominal ambient temperature of 25 °C:

Specimen set of transistors	Battery voltage	Temperature	Power gain
high α'	2.4 V	25 °C	75 db
mean α'	2.4 V	25 °C	73 db
low α'	2.4 V	25 °C	70 db

The mean power gain of 73 db can be increased to 77 db by reducing the feedback resistor R_{14} in the output stage to 1 Ω, but at the expense of increased spread in gain.

The following table shows the maximum and minimum values of power gain obtained under extreme conditions:

Specimen set of transistors	Battery voltage	Temperature	Power gain
high α'	2.4 V	40 °C	76 db
low α'	2.4 V	20 °C	69 db

The 12 db feedback taken from the 2 Ω resistor in the OC 71 emitter circuit to the second stage keeps the spread in gain under extreme conditions down to ± 3.5 db. As the ambient temperature increases, the gain increases slightly up to a temperature of 40 °C and falls off with any further rise of temperature.

In these tables the term low α' here means that the three OC 70 transistors and the OC 71 transistors were selected from the lower end of the production spread in α'. Similar definitions apply to mean α' and high α'.
The following table finally shows the collector current and average power gain (without allowing for feedback) in each stage:

Stage	Power gain	Collector current
first	17 db	300 μA
second	21 db	300 μA
driver	18 db	500 μA
output	30 db	2 mA

FREQUENCY RESPONSE

Fig. 59 shows the frequency response of the circuit when fed from a source impedance of 1 kΩ, with an OC 71 collector load of 1 kΩ shunted by a choke (inductance 20 H, resistance 250 Ω).

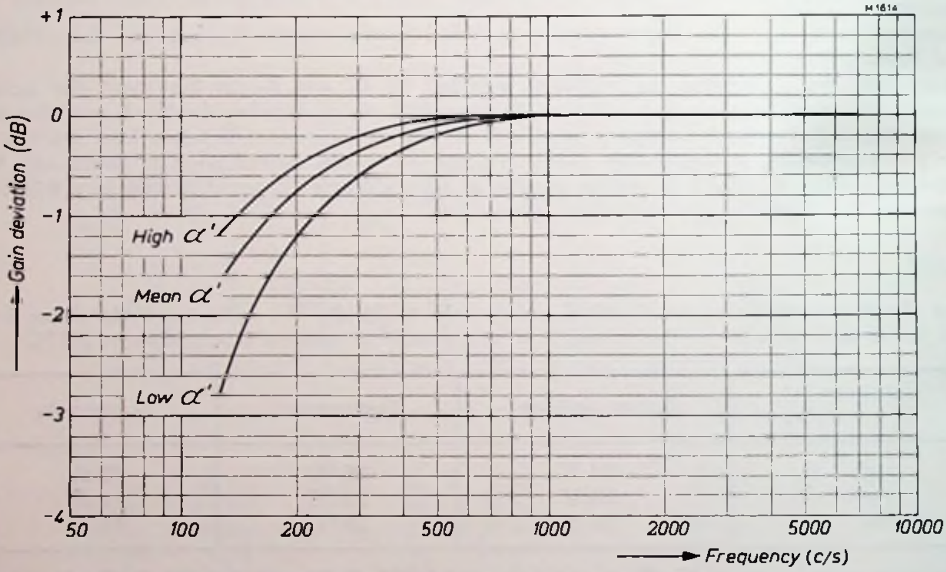


Fig. 59. Frequency response curves.

The reduction in gain is less than 2 db at 200 c/s and the response is flat up to 7 kc/s. The response of the complete hearing aid will therefore depend primarily on the microphone and earphone.

OUTPUT AND DISTORTION

The total harmonic distortion for various output power levels and with a 400 c/s input signal is given in Fig. 60. At 5 % total harmonic distortion the output is 1.9 mW for a circuit employing transistors with low current gain α'. This figure becomes 2.0 mW for transistors having a mean or high value of α'.
From Fig. 61 to 63 it appears that the second and fifth harmonic distortion is much less than the third harmonic distortion. When the signal amplitude is large enough to overload the OC 71, the limitation on the two halves of the waveform is the same, and the distortion therefore consists mainly of the third harmonic.

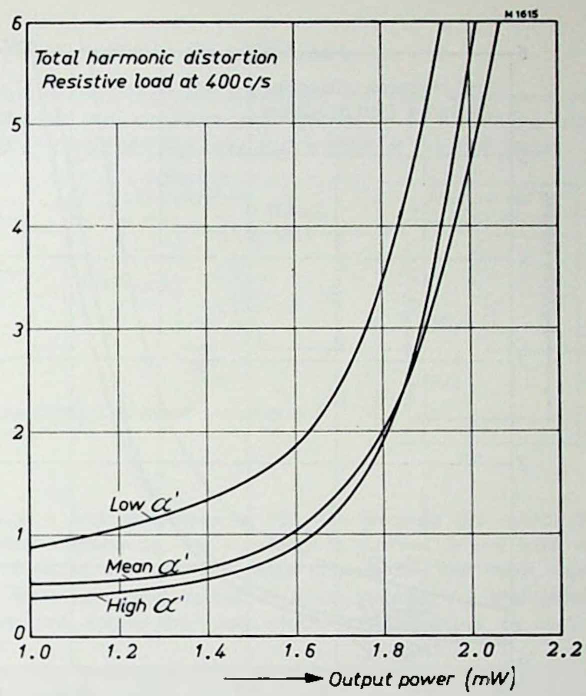


Fig. 60. Total harmonic distortion as a function of the output power.

FIXED VALUE FOR R_{13}

If a fixed value of 39 kΩ is chosen for R_{13} the output stage will operate under optimum conditions only when the OC 71 transistor has a low value of current gain α'. For transistors of mean or high α' values the clipping in the output stage under overload conditions becomes asymmetrical, and second harmonic distortion is increased.
From the following table it follows that with the mean or high values of α' the output power at 5 % distortion is not appreciably affected by the fixed value of R_{13}, but the current drain increases considerably (up to 30 % increase). With the correct matching values for R_{13} based on the collector current in the output stage of 2.0 to 2.1 mA, the total current drain is held to 3.5 mA, as shown by the second section of the table.

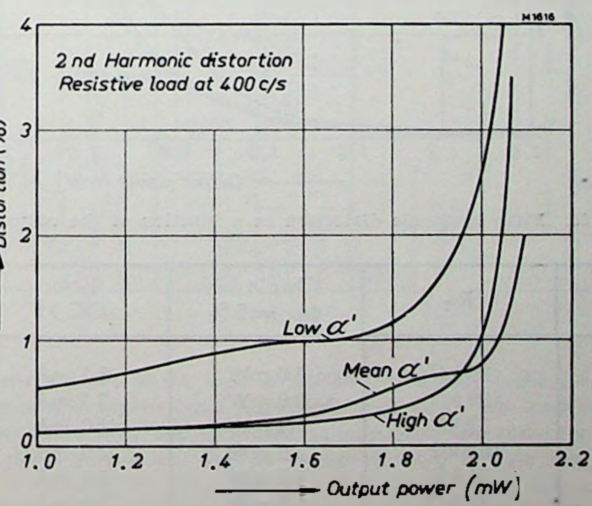


Fig. 61. Second harmonic distortion as a function of the output power.

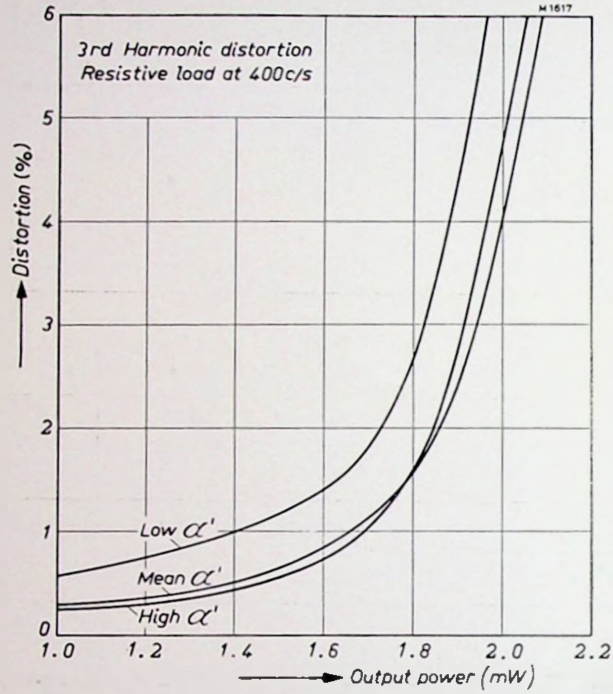


Fig. 62. Third harmonic distortion as a function of the output power.

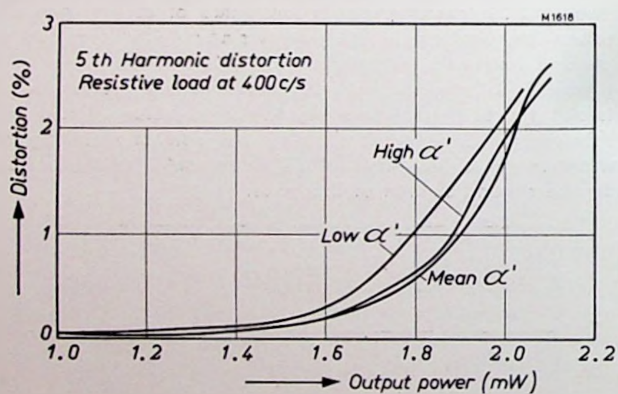


Fig. 63. Fifth harmonic distortion as a function of the output power.

Specimen set of transistors	R_{13}	Output for $d_{tot} = 5\,\%$	i_c for OC 71	Current drain
high α′	39 kΩ	∾ 2.0 mW	3.1 mA	4.6 mA
mean α′	39 kΩ	∾ 2.0 mW	2.7 mA	4.1 mA
low α′	39 kΩ	1.9 mW	2.0 mA	3.25 mA
high α′	68 kΩ	2.0 mW	2.0 mA	3.5 mA
mean α′	56 kΩ	2.0 mW	2.1 mA	3.5 mA
low α′	39 kΩ	1.9 mW	2.0 mA	3.25 mA

MEASUREMENT OF INPUT IMPEDANCE AND GAIN

Fig. 64 shows a suitable arrangement for measuring the input impedance and power gain. If the sensitive tube voltmeter and oscillator are earthed, a 1 : 1 isolating transformer must be used, since the input is fed between the base and emitter of the first stage.

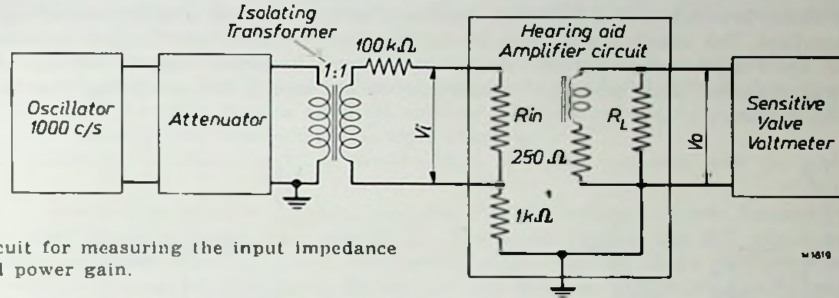


Fig. 64. Circuit for measuring the input impedance and power gain.

The input is fed through a high resistance of 100 kΩ. Because the source impedance is high compared with the input impedance, the transistor is current driven from a source which is unaffected by any variation in the transistor input impedance. The input impedance and input power are calculated from measured values of input voltage V_{in} and input current I_{in}. The output voltage is measured across the load which is represented by a 1 kΩ resistor from V_o^2/R_L.

VI. 12. 0.2 W GRAMOPHONE-AMPLIFIER

This comprises a preliminary stage, a driver stage and a class-B push-pull output amplifier with two OC 72 "matched pair" output transistors (that is to say transistors whose characteristics are sufficiently well-matched to ensure a symmetrical output signal).
The input impedance of the amplifier is about 7 kΩ, rather low to be matched to a crystal pickup; hence a series resistance is provided at the input. On the other hand, the resistor absorbs much of the energy supplied by the pick-up (about 6 db). Nevertheless, the overall amplification is enough to enable the output stage to be fully loaded. If the amplifier is to follow a crystal receiver, another emitter-follower stage should be connected in series with it (V_1 in fig. 65).

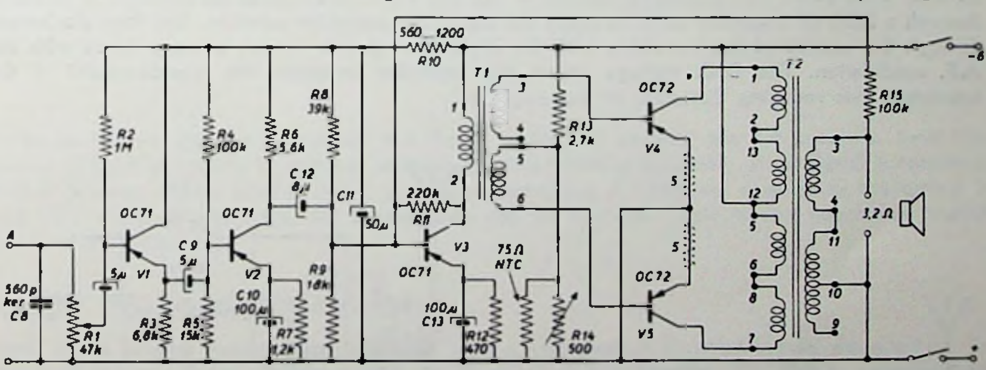


Fig. 65

This enables the series-resistance at the input to be dispensed with, making the full input power available. Also the relatively higher input impedance cuts down the load on the circuit, thus giving scope for greater selectivity and increased voltage gain, and reducing modulation distortion associated with deep modulation. This is explained more fully in the chapter "Receiver circuits".
With two OC72 transistors of current design, output power is in the region of 200 mW; transistors recently developed should enable this to be increased to 350 mW.

CHAPTER VII

RECEIVER CIRCUITS

VII. 1. SINGLE-TRANSISTOR RECEIVERS

A circuit in which one transistor is employed as detector and amplifier combined will now be described. This circuit is analogous to that in which a tube is used for grid detection, employing the bend of the characteristic. At the same time, transistors require nothing resembling the "grid capacitor" and "grid leak", which constitute a kind of R.F. smoothing filter in grid-detector tube circuits.

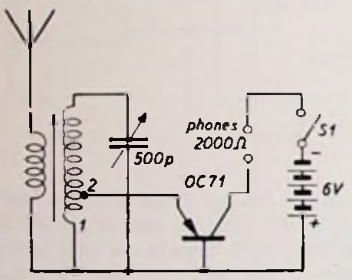


Fig. 66

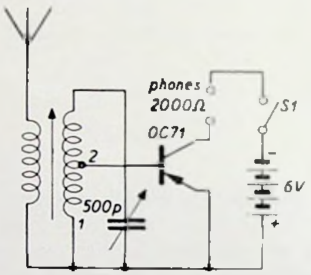


Fig. 67

Moreover, a capacitor in series with the base would cause reactive peak detection, thus reducing the efficiency of the receiver (already rather low) considerably. The transistor should therefore be connected direct to the tuning circuit. It is then a matter of choosing between grounded-base (fig. 66) and grounded-emitter connection (fig. 67). With the former, the coil should be tapped as near as possible to the bottom, as the input impedance is then low, and with the latter the tapping may be taken slightly higher.

Better is to split up the functions of detector and amplifier. Adding a germanium diode gives the circuit of fig. 68. To ensure that direct current from the transistor does not reach the diode, an isolating capacitor should be inserted between base and diode. This gives what is known as a floating-base circuit, with the transistor adjusted to the amplified leakage current I_{co}' (see III. 7b). Such circuits are satisfactory when operated with weak signals and sufficient I_{co}.

A very neat piece of circuitry is shown in fig. 70. Here, the diode rectifies the R.F. signal through a 2000 pF capacitor so as to make the top of the capacitor positive. This then discharges through the emitter of the transistor, with the discharge current varying in accordance with the A.F. modulation. The D.C. voltage across the capacitor improves the working-point of the transistor, thus reducing distortion in the amplifier.

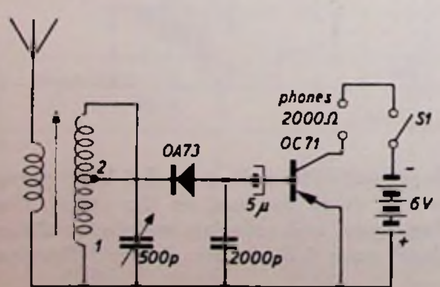


Fig. 68

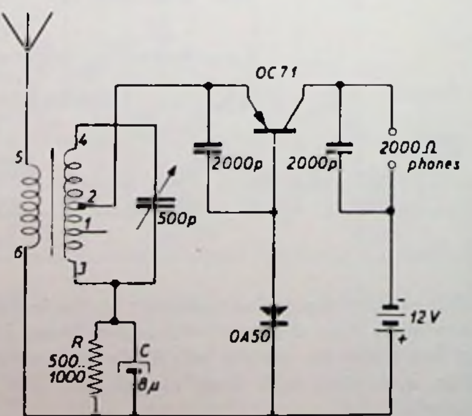


Fig. 69

A useful feature of this circuit is that its current consumption depends on signal strength. The collector current is 50 to 150 μA without signal, and from 1 to 2.5 mA with a signal of suitable strength.
Provided that the user happens to live fairly near the broadcasting station and has installed a good aerial, this circuit will even drive a speaker. The sound-level is then ample for small rooms such as bedrooms where there is not much ambient noise. Of course, the more sensitive the speaker, the better the reception.
A convenient value for the matching impedance of the output transformer is about 2000 Ω.
Another advantage from the point of view of users living near the transmitter is that power from the transmitter itself can be used to drive the receiver, no battery being required (fig. 70). In effect, this is a two-stage receiver with fairly tight coupling.
Winding 5—6 constitutes a "filter circuit" in series with a two-gang capacitor with the gangs in parallel.
A switch (S_1) is provided to bring in an extra capacitor (C_1) for tuning to lower frequencies. The actual receiver circuit is tuned by means of C_4. The diode rectifies the R.F. signal and supplies collector current to the OC71. It is important to ensure that the diode rectifies in the proper direction; if it is connected wrongly, the set will not function. The residual A.F. ripple across C_5 is amplified by the transistor.

COIL DATA to the figures 66, 67, 68, 70 and 72

The tuning coils are wound on a former with flanges at distances of 2 mm, an outer diameter of 20 mm and a breadth of 12 mm at the bottom.
A powdered iron core should be inserted.
Total number of turns (3—4) 80 Litz wire 16 × 0.04.
Start winding at connection no. 3, make tap 1 at 22 turns, no. 2 at 29 turns.
Number of turns (5—6) 23.

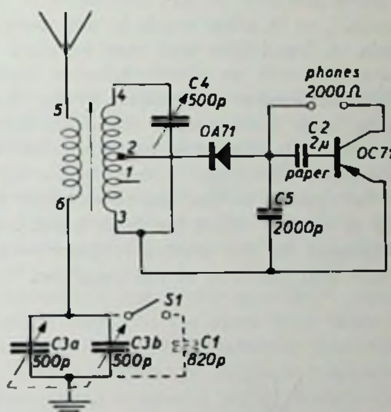


Fig. 70

Tune as follows: Having connected the (long) aerial and earthed the set properly, twist the knobs of C_{3a}/C_{3b} rapidly to and fro, at the same time rotating C_4 slowly. If no signal is received, switch S_1 over. When once a signal is obtained, trimming is relatively easy, since the tuning is not very sharp. If necessary, the tuning dials can be marked. Triple tuning capacitors cannot be used in this circuit.

VII. 2. TWO-TRANSISTOR RECEIVER

Very much greater sensitivity is obtained with a two-stage A.F. amplifier so designed as to increase the voltage gain in the tuning circuit as far as possible (fig. 71). This is accomplished by cutting down the circuit current, thereby combining limited tuning capacitance with high self-inductance.
The tuning capacitor comprises a 100 pF variable, and a 100 pF fixed capacitor in series. As the limited tuning capacitance covers only a narrow frequency band, a switch (S_2) is provided, enabling an overall range from 500 to 2100 kc/s to be covered in two steps, with a certain amount of overlap.
Using a whip aerial, headphone-reception is quite good, at any rate within reasonable distance of the transmitter. As the receiver has very few components, it can be accommodated, with the $1\frac{1}{2}$ V dry cell, in a very small box.

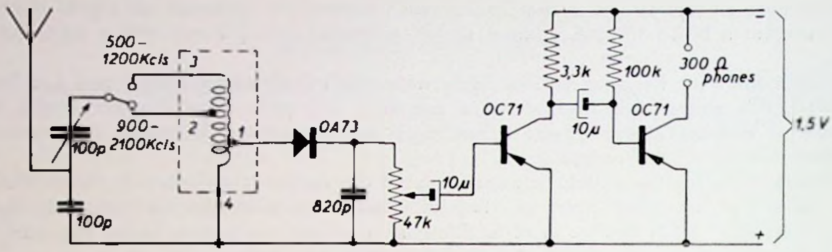


Fig. 71. The tuning coil is wound on a former with flanges at distances of 2 mm an outer diameter of 20 mm and a breadth of 12 mm at the bottom. A powdered iron core should be inserted. The total number of turns is 360. Litz wire 16 × 0,04 mm.
Start winding at connection no. 4; make tap no. 1 at 24 turns, no. 2 at 220 turns.

VII. 3. THREE-TRANSISTOR RECEIVER

Failing the use of R.F. transistors it is all the more necessary to "get the most out of the aerial", or in other words to pay more attention to the tuning circuits so as to ensure maximum gain. A band-filter unit may be used as tuner. To minimize circuit damping, the amplifier is provided with an emitter-follower input (fig. 72).
This has several advantages. Firstly, the input impedance is relatively high, which, as we have seen, cuts down circuit damping, thus enabling signals to be received from stations of low power. Secondly, high input impedance prevents distortion.
The third advantage is that dispensing with the series resistance enables the full input power to be applied to the amplifier, which results in a gain of roughly 3 db.
So as not to affect the high input impedance thus obtained, a floating base arrangement is employed for the emitter follower — in other words, an isolating capacitor is placed between diode and the base of the transistor.

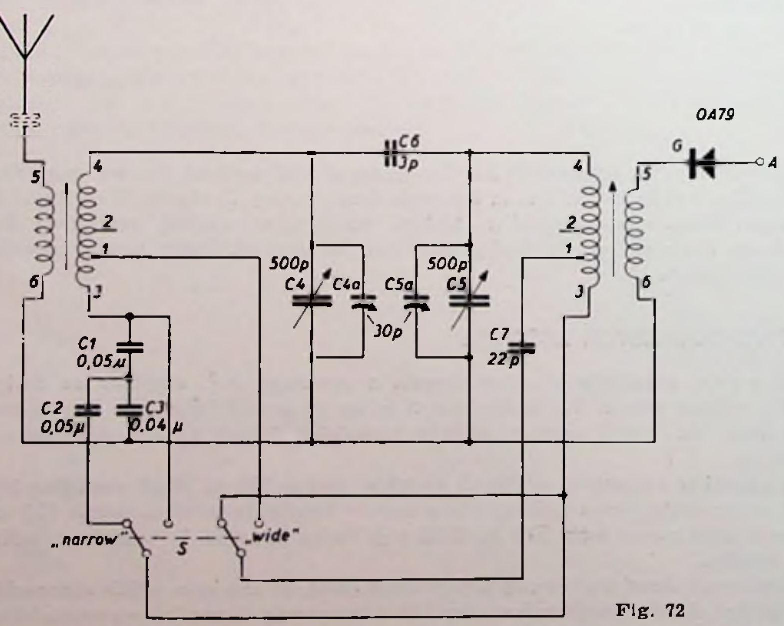
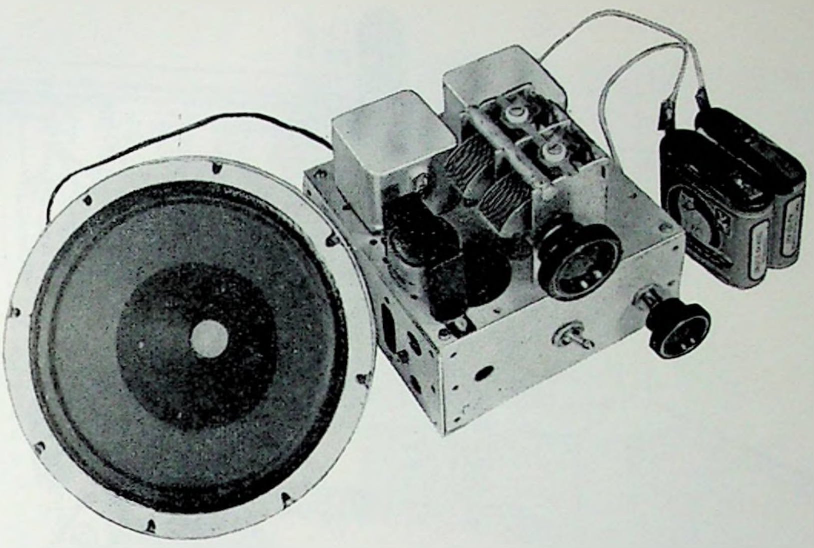


Fig. 72



The transistor is thus set to its leakage current I_{co}'. At the same time, it may happen under unfavourable conditions (low ambient temperature, strong signal) that the circuit produces distortion. To prevent this, a 1 MΩ resistor should be placed between —6 V and the base. It will be evident that the overall A.F. gain would be greater with V_1 as a voltage amplifier. On the other hand, the low input impedance of such amplifiers would cause circuit damping to become so heavy as to affect the overall performance of the set, firstly by reducing the selectivity and secondly by wasting much of the A.C. alternating voltage so carefully conserved at the input. Hence it is very much better to employ an emitter follower.

The main purpose of gain control R_1 is to give overload protection, not to the speaker — since the output power is only about 10 mW — but to the output stage. At the same time there is enough output power to develop a surprising volume of sound in sensitive speakers, and with a good-sized baffle, at any rate, bass-response is more than adequate. A baffle is necessary to ensure a proper ratio of bass and treble, since compensating networks would absorb too much energy.

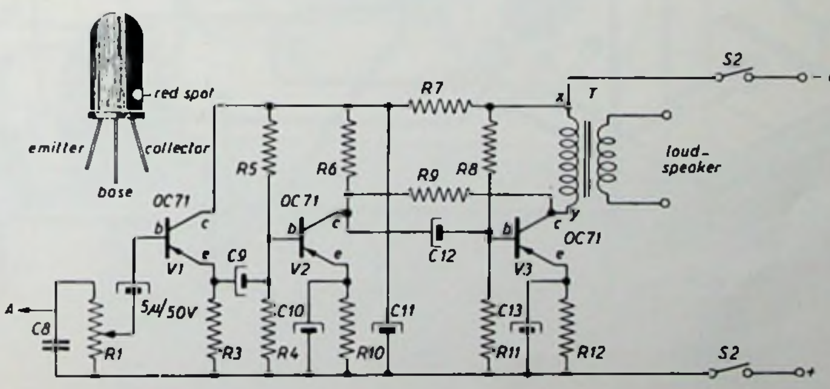


Fig. 73

R_1	= 100 kΩ	R_{7-10}	= 1.2 kΩ	C_8	= 470 pF
R_3	= 6.8 kΩ	R_8	= 39 kΩ	C_9	= 5 μF/50 V
R_4	= 15 kΩ	R_9	= 220 kΩ	C_{10-13}	= 100 μF/12 V
R_5	= 100 kΩ	R_{11}	= 18 kΩ	C_{11}	= 50 μF/12 V
R_6	= 5.6 kΩ	R_{12}	= 470 Ω	C_{12}	= 8 μF/12 V

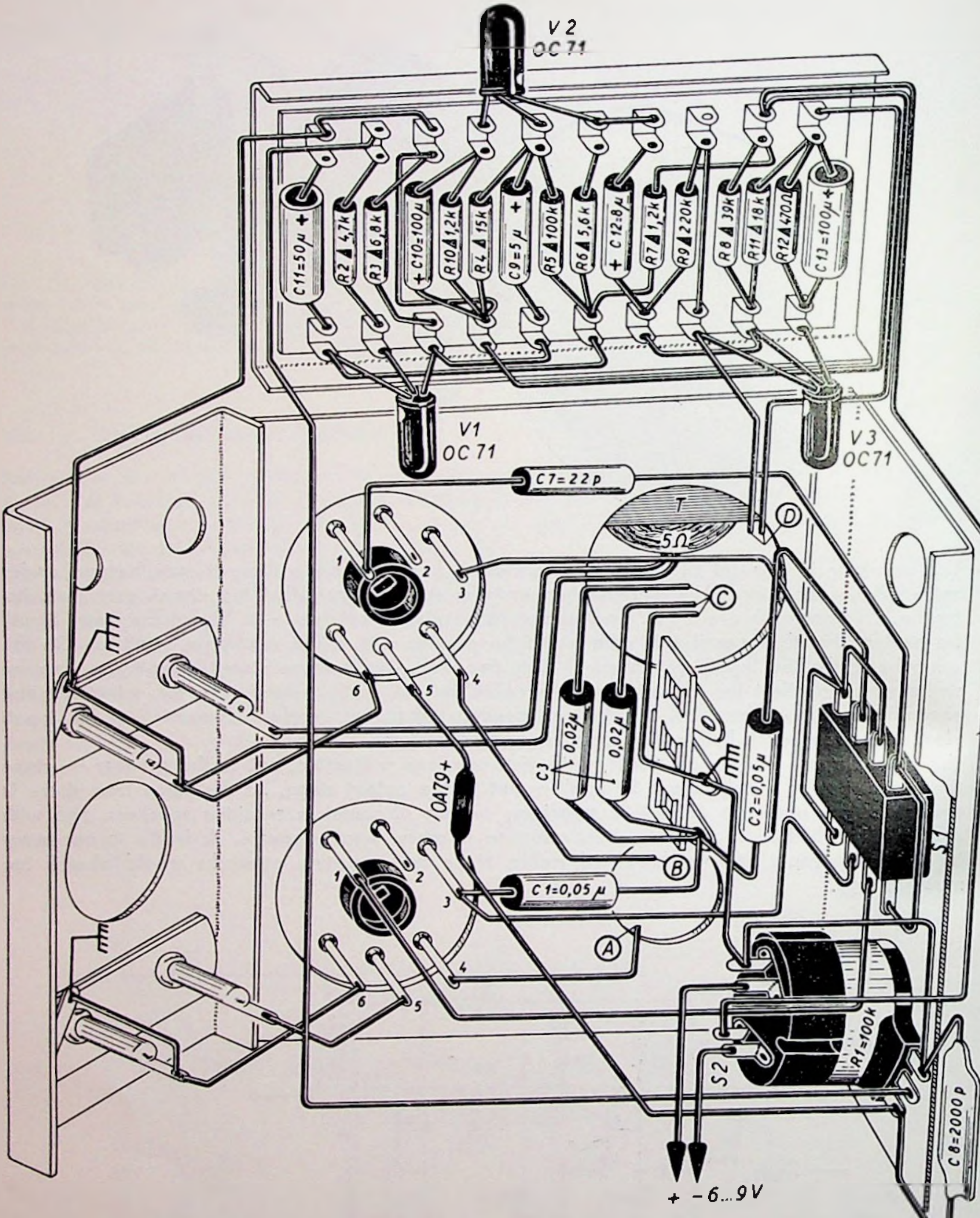
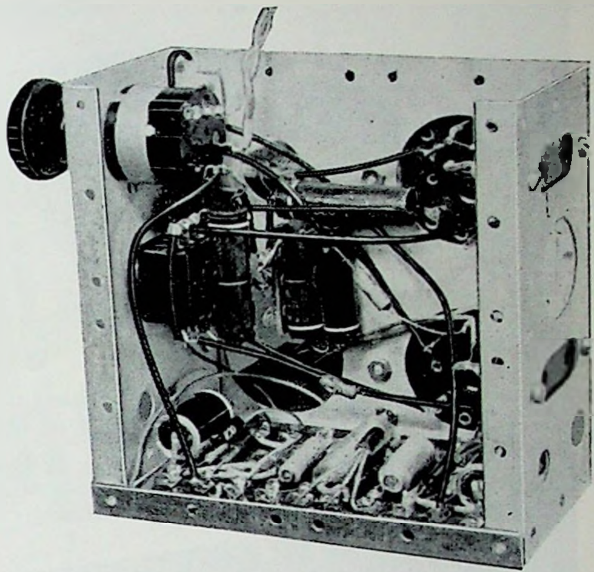

V 2
OC 71
C 11=50 µ +
R2 4,7k
R3 6,8k
+ C10=100 µ
R10 1,2k
R4 15k
C9=5 µ +
R5 100k
R6 5,6k
+ C 12=8 µ
R7 1,2k
R9 220k
R8 39k
R11 18k
R12 470Ω
C 13=100 µ +
V1
OC 71
V 3
OC71
C7=22 p
T
5Ω
D
C
1
2
3
4
5
6
C3
0,02 µ
0,02 µ
C2=0,05 µ
OA79 +
C 1=0,05 µ
B
A
S2
R1=100k
C 8=2000 p
+ -6...9 V

Fig. 74



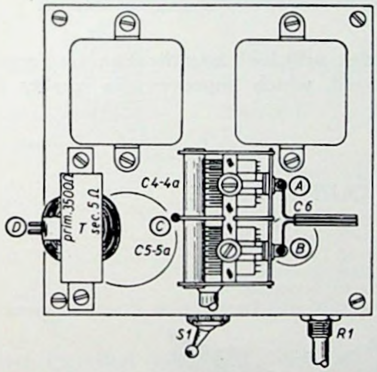
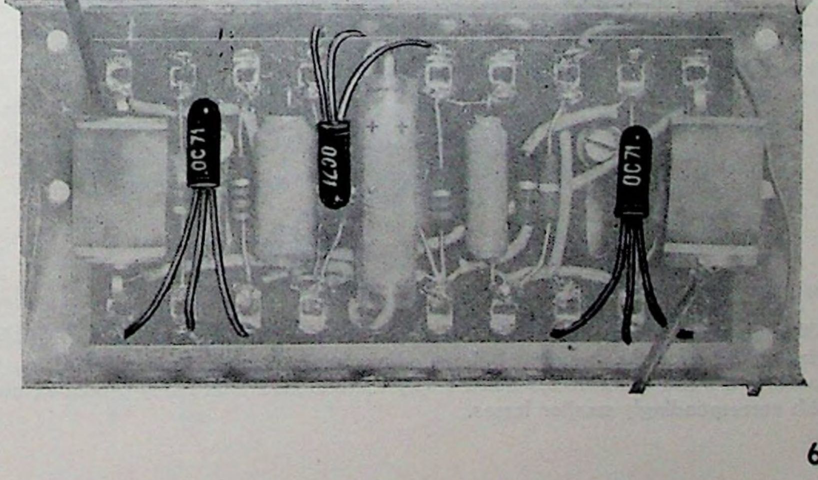
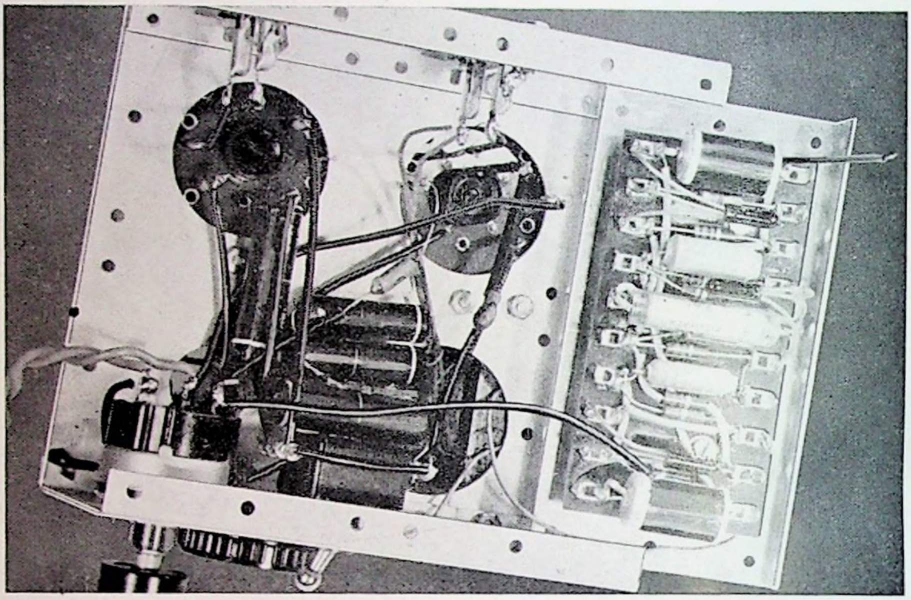


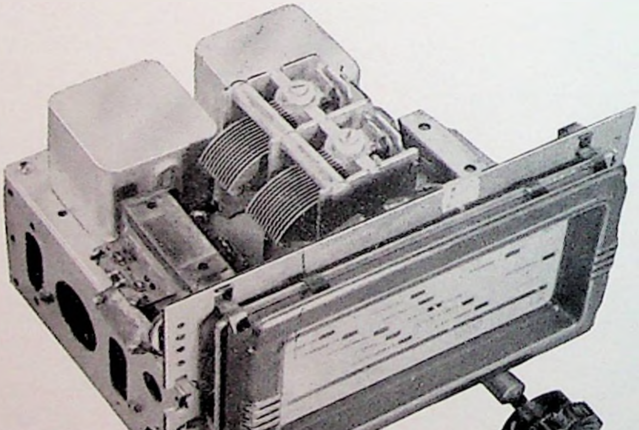
Fig. 75





Nevertheless, experience has shown that the circuit provides sufficient amplification to permit a certain amount of negative feedback (R_9) to be employed, which improves the quality of reproduction.
The maximum battery voltage is 9 V.

VII. 4. FIVE-TRANSISTOR RECEIVER WITH CLASS-B OUTPUT STAGE



Although suitable transformers are on the market, it is worth while to provide the necessary winding data for the benefit of those who prefer to build their own. Using these data it is possible to make transformers more amply dimensioned than many produced commercially, and with correspondingly smaller losses.

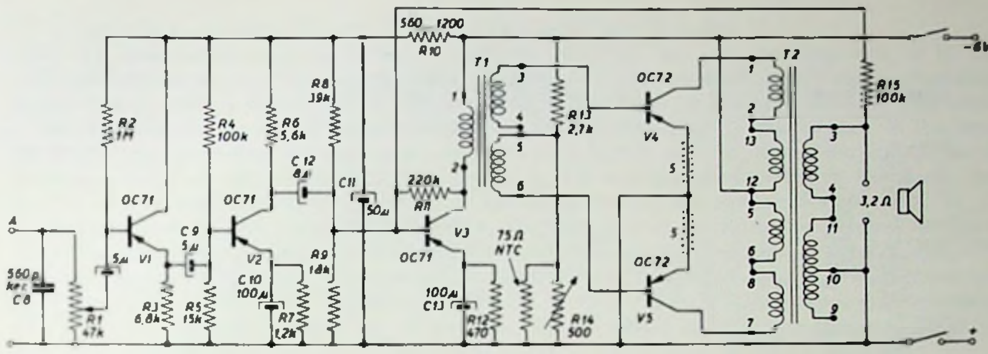


Fig. 76. Low-frequency part of 5-transistor receiver, to be combined with the band-filter unit of Fig. 72.

THE DRIVING TRANSFORMER

Mu-metal core taken from a microphone transformer. For dimensions see fig. 77. Windings are wound simultaneously to ensure sufficiently tight coupling between them.
Take care to number the soldering tabs on the coil bobbin so as to be able to check the connections after winding. Each arrow in fig. 77 indicates the start of a winding.

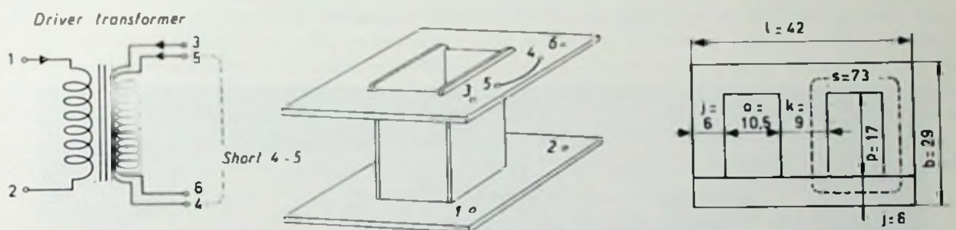


Fig. 77

primary 1—2	=	2500 turns, ø 0.09 enamelled
secondary 3—4, 5—6	=	715 turns, ø 0.16 enamelled, bifilarly wound
stacking height	=	9 mm
diameter of core	=	9 × 9 = 81 mm²
average length of lines of electric force	=	2 (b—j) + l — (k + j) = 73 mm

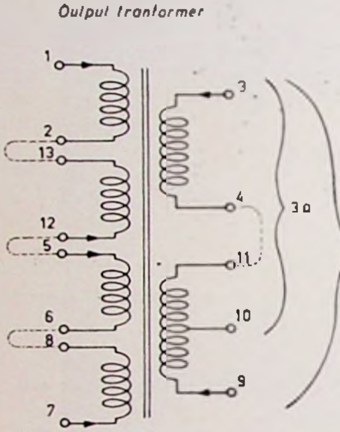
THE OUTPUT TRANSFORMER

Core consists of ordinary silicon iron from a discarded output transformer. Window should be on the large side in view of the thickness of the wire employed. If necessary, two sets of E-laminations can be used instead of the usual combination of E and I laminations. The legs of one set are shortened to permit of alternate stacking. In the example given in fig. 78, the original lamination height is halved.
To ensure good coupling, the principle of "equal average turn diameter" is employed, i.e. each coil is divided in two and so distributed that the average turn diameter is the same for any two halves in series. To offset parasitic capacitances, the winding direction is changed after the first two half-coils. As far as possible, the coil should fill the window in order to minimize stray flux. The speaker winding is provided with a tapping enabling either a 3 Ω speaker or a 5 Ω speaker to be employed.
The conversion formula for other core dimensions is:

$$n_x = n \sqrt{\frac{s_x}{s} \cdot \frac{A}{A_x}},$$

where n_x is the unknown, and n the specified, number of turns, s_x the mean flux path and A_x the cross-sectional area of the new core. Hence we have a conversion factor F:

$$F = \sqrt{\frac{s_x}{s} \cdot \frac{A}{A_x}}.$$



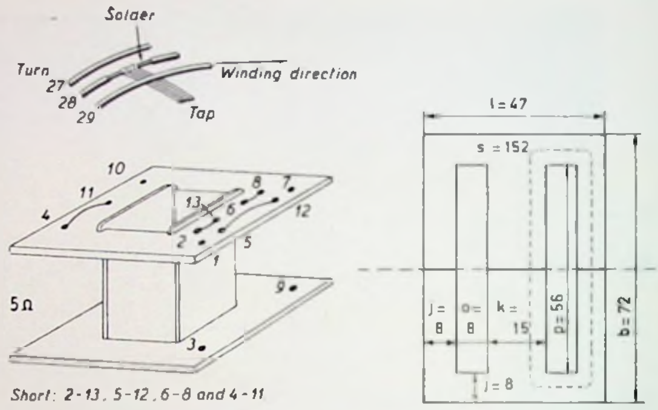
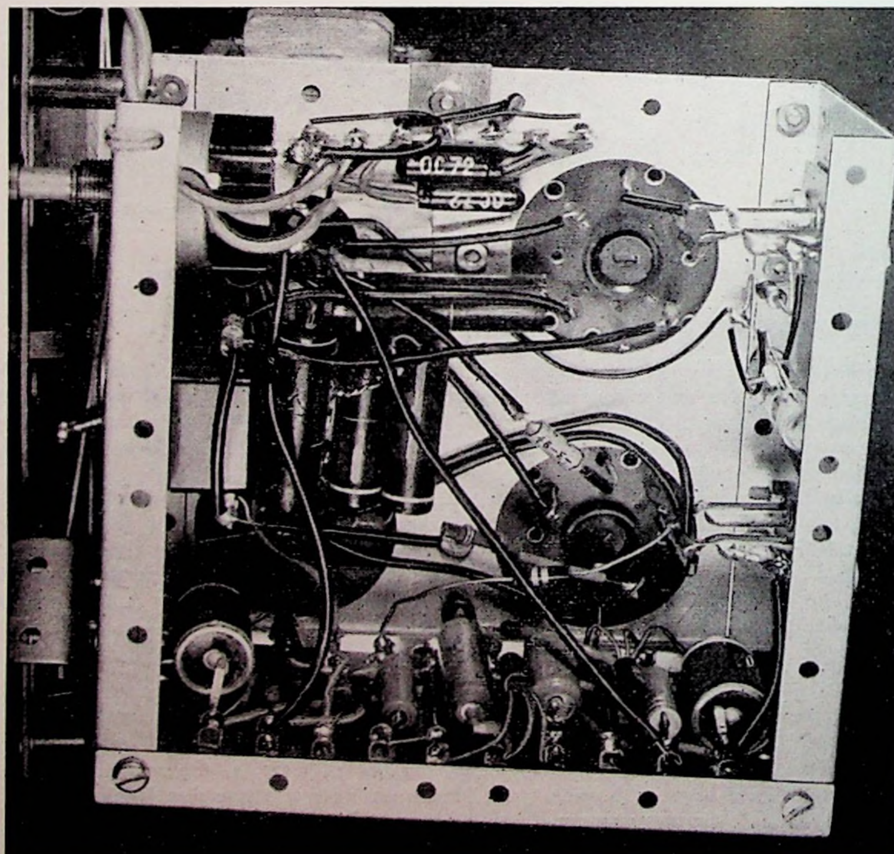


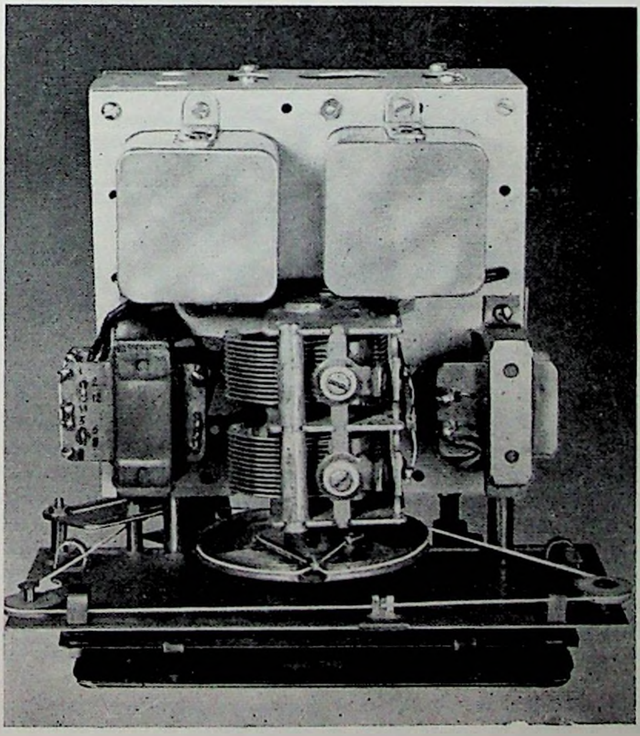
Fig. 78. Order of winding
w = turns; em = enamelled
reverse direction of winding tapped after 34 turns
stacking height = 7 mm
diameter of core = k × h = 15 × 7 = 105 mm²
average length of lines of force = 2 (b—j) + l — (k + j) = 152 mm



All the specified numbers of turns should be multiplied by this factor.
After this calculation it is necessary to ascertain whether the new winding will fit in the available window $o_x \cdot p_x$. Find the diameter of the particular enamelled wire from a wire-table, square it and multiply the square by the total number of turns of this type of wire. This gives a total area, which should not be more than half the effective area of the window in the core lamination. Provided that the turns are wound in neat layers, not too loose, the calculation is accurate enough. If this shows that the window in the available laminations is too small, the number of turns may be reduced — provided that the difference is not more than about 10 %, — by an equal amount. This actually increases the efficiency, but at the same time curtails the bass-response. Transformers wound to the specified data reproduce peaks up to roughly 450 mW without distortion. Even with the number of turns halved, maximum power without distortion is in the region of 140 mW. The turn-over point f_0 is then about 350 c/s. Hence it is advisable to adhere as strictly as possible to the winding data, in order to keep something in hand for peak power-levels.
To obtain very good bass response, increase the coupling and decoupling capacitances to, say, 100—200 μF. As this may cause "hiccup", however, the filter capacitance should also be increased.

CONSTRUCTION

Building the balanced stage is fairly simple. Care must be taken that the driver and output transformers are connected properly. Unless they are in anti-phase, the set will not work.
Now a word or two on NTC-resistors. "Thermistors", as employed in D.C./A.C. sets, are not suitable in the present case; the resistance is too high and the temperature coefficent too low. As a temporary expedient an ordinary 100 Ω carbon resistor can be used instead of the NTC, although it will then be necessary to check the working-point from time to time. Alternatively a germanium diode — e.g. type OA5 — operated in the forward direction, can be used instead (anode to earth). This provides sufficient voltage stabilization to prevent the class-B working-point from varying unduly with varying battery voltage. The earth connections must be carefully



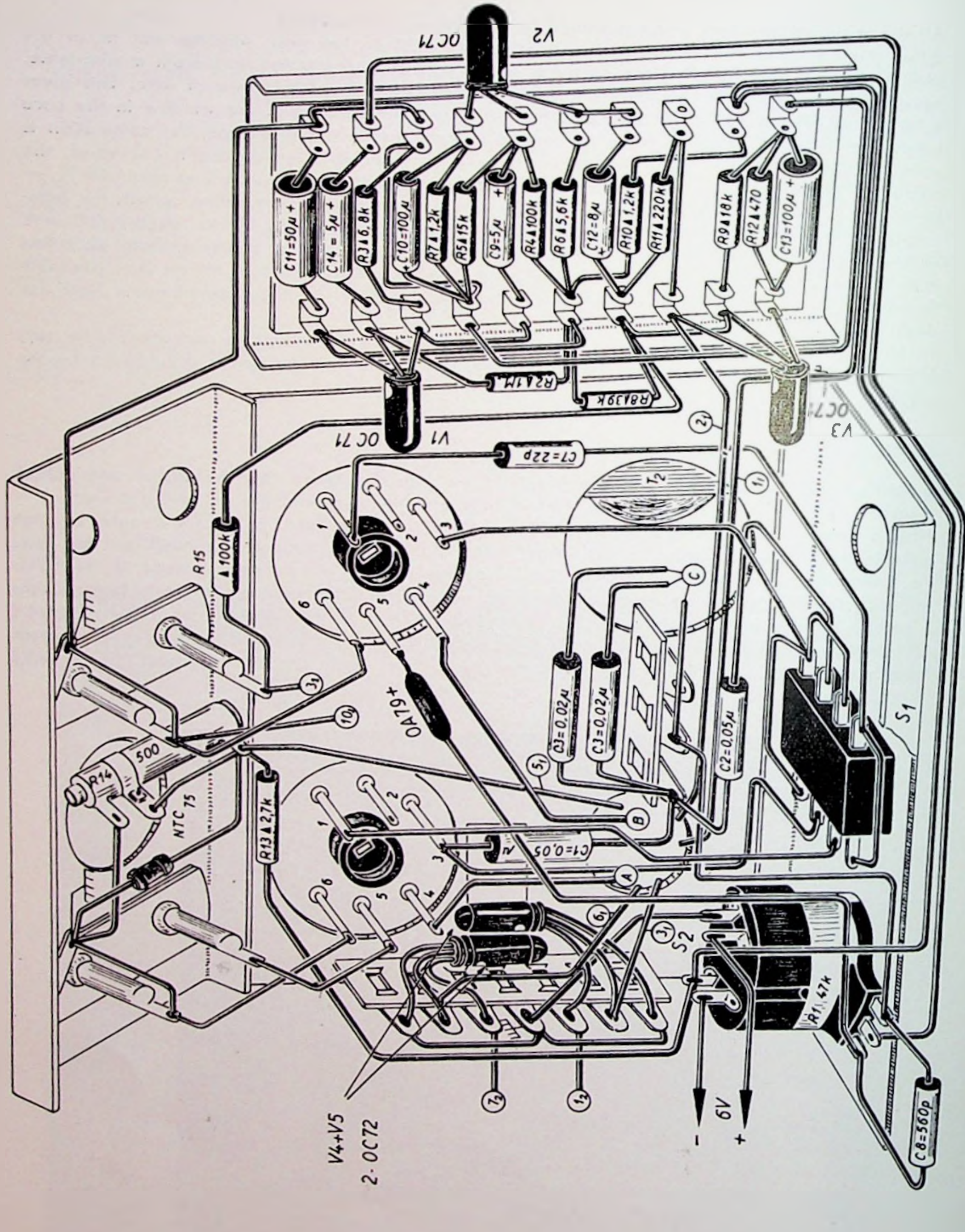

OC71
V2
C11=50μ +
C14 = 5μ +
R3 6,8k
C10=100μ
R7 1,2k
R5 15k
C9=5μ
R4 100k
R6 5,6k
C12=8μ
R10 1,2k
R11 220k
R9 18k
R12 470
C13=100μ +
R2 1M
R8 39k
V1
OC71
V3
C7=22p
R15
100k
OA79+
C3=0,02μ
C2=0,05μ
S1
500
R14
NTC 75
R13 2,7k
C1=0,05
S2
R1 47k
C8=560p
V4+V5
2·OC72
6V

Fig. 79

checked. The switch and the gain control should be earthed at different points, as a precaution against possible A.F. oscillation.

ADJUSTMENT

With a milliammeter connected between the centre-tapping of the output transformer and the —6 V supply, so adjust R_{14} that the overall collector current of the two transistors is 3 mA. Do not forget to connect both ends of R_{14}, as the meter may burn out if the circuit is broken when the slider is shifted. R_{13} is not variable, since the resistances specified provide sufficient compensation for normal temperature variations. If a variable resistor is preferred, however, it is advisable to place a fixed 1.5 Ω resistor in series with R_{13}, to protect the bases from possible full negative bias, which would destroy them.
Next, connect an output meter (or an A.C. voltmeter with 3 Ω shunt) to the speaker terminals. Disconnect the diode and connect a signal generator to the input. With 200 mW output, the milliammeter (without thermocouple) should indicate roughly 40 mA. Wait five minutes or so, then turn gain-control down and check steady-current setting. If necessary, trim R_{14} to bring current back to 3 mA.
For those who do not possess a signal generator, a straight receiver in oscillation will do just as well. Going slightly off-tune from the carrier of a fairly powerful station enables a strong test-signal to be taken from the speaker terminals.
The sensitivity of the amplifier is from 2 to 5 mV — depending on the spread in α' between the transistors employed.
Finally, if all is well, connect aerial and earth.

PERFORMANCE

This depends very much on the aerial and earth. Unduly long earth leads from upper stories of houses etc. may function as aerials; aerial and earth are then for all practical purposes in phase. Often, a separate piece of wire along the ground, as a counterpoise, improves matters.

DIMENSIONS

The set is not designed as a miniature receiver, mainly because suitable miniature components are not yet plentiful enough and because a good deal of experience and knowledge are necessary to make miniature sets (although well-made miniature components do not cause losses). Hence the rather ample dimensioning of the output transformer. In general it is not advisable to use nickalloy as the core, since slight asymmetry of the output transistors would then be enough to cause core saturation. On the other hand, since the permeability of nickalloy is about 4 to 5 times that of ordinary silicon iron, it enables the number of turns, and therefore also the size of the transformer, to be almost halved. With two identical resistors of about 5 Ω in the emitter circuits of the output transistors, the set is less dependent on possible differences in characteristics. If potentiometers are employed, very accurate balancing is possible. Nevertheless, the adjustment is rather critical.

VII. 5. 7-TRANSISTOR SINGLE-ENDED PUSH-PULL RECEIVER WITH AN OUTPUT OF 5 W

As explained in the section on amplifiers, it is possible to increase the A.F. output by employing a push-pull output stage. The trouble is that to drive a push-pull stage comprising two OC16 transistors requires a not inconsiderable amount of power, which must be supplied by the preceding stage. Suppose that the maximum driving power is 30 mW. To employ an OC16 in class-A for the driver stage would be practicable, but not very economical (30 mW can also be obtained with one OC72). By choosing a class-AB push-pull stage as the driver, we not only achieve a considerable reduction in current consumption, but also enable the present circuit to be designed as plug-in "extension" to the preceding one. The result is shown in figs. 80a and 80b. To enable a readily obtainable speaker-impedance to be employed, the circuit is designed to operate with a 3-Ω speaker and 2 $\times$ 6 V from the battery (ordinary 12-V car battery). The theoretical output power is then 6 W. In practice, however, certain limiting factors — amongst other things driving power — reduce the output to 4—5 W; nevertheless, this output is quite sufficient for, say, a closed car. The preamplifier + driver is for all practical purposes the same

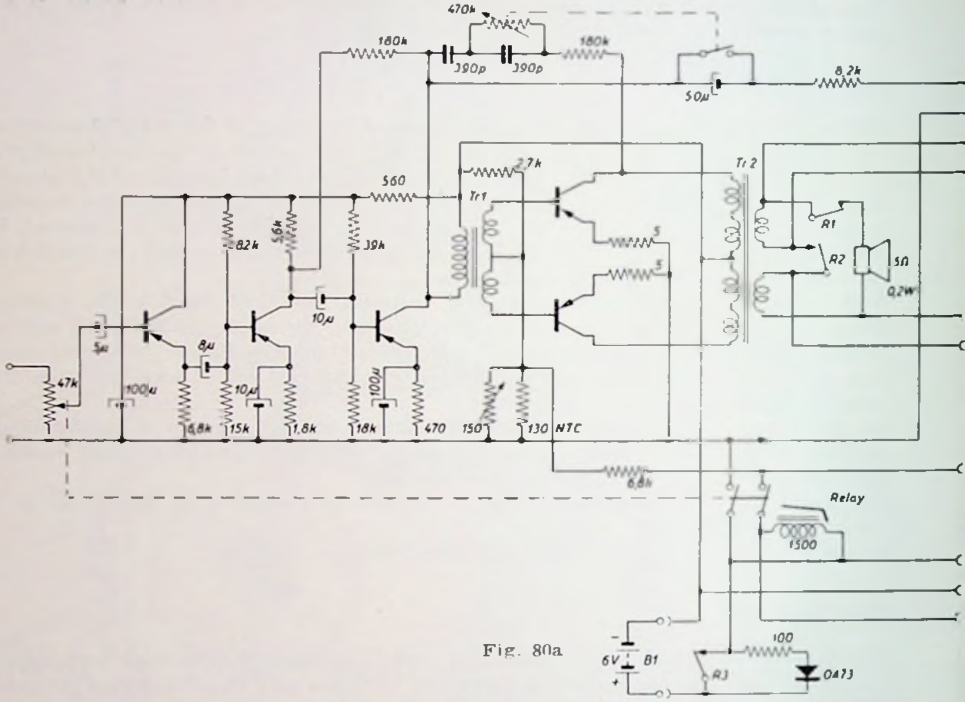


Fig. 80a

as in the preceding circuit. The only difference is that variable, frequency-governed negative feedback is employed to ensure effective tone control.
To obtain an alternative low-power output of about 0—2 W, a small 5-Ω speaker may be connected to TR_2. Both halves of the secondary of TR_2 are then in series with a relay contact (break contact) R_2. The two OC72 transistors are then in class B, and are supplied from a built-in miniature accumulator B_1 across relay stand-by contact R_3.
To increase the output power, insert the plug-in unit in a 10-point socket, thus closing the relay across the 12 V battery of the unit and opening all the relay contacts. Next, switch off the 5-Ω speaker and separate the halves of TR_2 secondary. TR_2 is than the driver transformer for the pair of OC16's. Closing on/off switch S_1 (on the gain control) brings the 6.8 kΩ resistor at the centre-tapping of TR_1 into series with the negative pole of the 12-V battery, thus bringing the two OC72's into class AB; this reduces distortion on weak signals.
Also, contact R_3 breaks the short-circuit across a germanium diode, thus enabling the built-in 6-V accumulator to be charged. The diode becomes non-conductive as soon as the voltage equals that of half the 12-V accumulator, thus preventing the built-in accumulator from discharging across the external battery as this runs down. A 100-Ω resistor in series prevents overloading of the diode by heavy charging-current.

VII. 6. POCKET RADIO WITH 4 TRANSISTORS

Pocket radios are always in demand. There is every temptation to employ transistors for this purpose, because they are small and require only one inexpensive battery. At the same time, transistors such as the OC70, 71, 72 and 73 do not always permit of good R.F. amplification, since their cut-off frequency is relatively low. The result is that feedback must be dispensed with; hence, really good circuit quality and a certain amount of extra A.F. amplification are essential. Reasonable head-phone reception is possible with a whip aerial of suitable size and a good earth.
However, the sensitivity is then on the low side for a "real" pocket radio; also, long whip-aerials are inconvenient. In this respect a ferrite rod is better, but since the amplification is

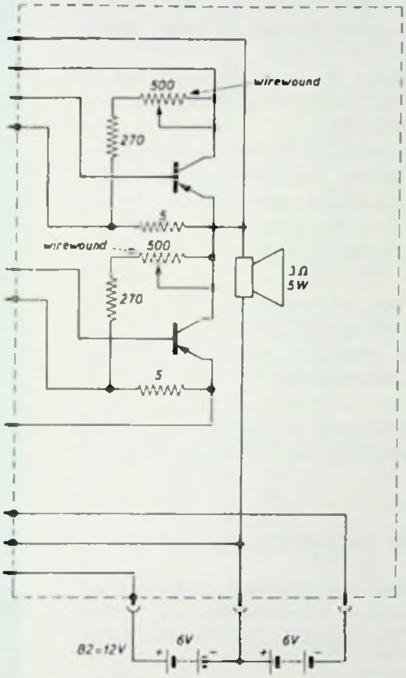


Fig. 80b

Circuit of a plug-in unit with power transistors OC 16.

still on the low side, an extra A.F. stage must be added, say, by repeating the second stage V_2 in the circuit of fig. 81.
To obtain maximum sensitivity with a ferrite rod aerial, it is necessary to use as many turns as possible. With many turns the inductance is high, so that a relatively small tuning capacitor must be employed and the tuning-range is limited.
On the other hand, since the small tuning capacitance cuts down the circuit current, thereby increasing the voltage gain in the tuning circuit, the D.C. losses in the aerial coil are smaller. This does not necessarily ensure optimum power matching, however, since the coil must then be tapped lower down. The Q of the coil without load should be as high as possible.
It is best to wind ordinary enamelled wire on the ferrite rod, first in order to determine the correct number of turns (which varies according to the size and material of the rod). When once the number of turns enabling a suitable range to be covered by means of the 100 pF capacitor is established, the rod should be rewound with litz wire. From 80 tot 120 turns are required, depending on the thickness of the rod.

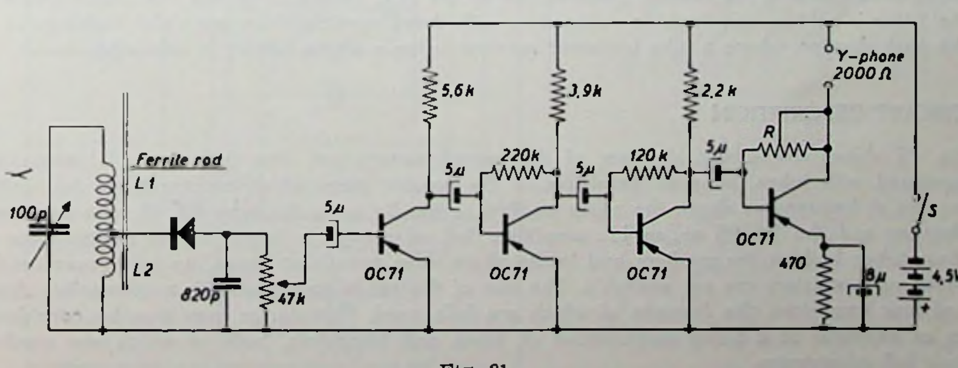


Fig. 81

The diode is connected to a tapping about 1/10 up from the earthy end and should detect even very weak R.F. signals. It is best to use an OA79 diode, which has a greater internal resistance, and therefore gives less attenuation than most diodes; on the other hand, the A.F. output voltage is slightly below normal, and therefore necessitates a certain amount of extra amplification.
Since the diode is on a tapping low enough on the coil, it enables the input transistor to be employed as a voltage amplifier. In general, it is not advisable to employ transistor detectors, as they require a relatively high R.F. voltage and are not very efficient.
With transformer coupling between the different stages, proper matching enables enough signal strength to be obtained with only 3 transistors (powered by a 3-V dry cell). With RC coupling an additional transistor is required. On the other hand, this arrangement makes the set lighter and easier to build, the more so as it avoids the necessity of winding the coupling transformers.
In order to achieve maximum economy as regards space and material, a simple circuit was chosen (fig. 81).
Stabilization is effected by feeding the base from the collector. Moreover, in this way a certain amount of A.F. negative feedback is produced, which improves the quality of the sound. Because stabilizing resistors in the emitter circuits are dispensed with, it is possible to employ a lower supply voltage (4½ V battery) than with the three-transistor receiver (fig. 73).
It is advisable to employ sensitive head-phones (2000 Ω).
The quality of reproduction thus obtained is remarkably high. If preferred, a change-over switch and one or two trimmers (with or without a fixed capacitor in parallel) can be used instead of the 100 pF tuning capacitor.

VII. 7. BATTERY RECEIVER WITH TWO TUBES AND FOUR TRANSISTORS

The introduction of the 96-range of battery tubes already marked an important milestone on the route to small portable battery receivers with very low power consumption. A further very drastic reduction of power consumption becomes possible by replacing part of the tubes by transistors.
This may be illustrated by the following description of an experimental battery receiver equipped with two 25 mA tubes of the 96-range, one germanium diode OA 79, and four transistors in the A.F. part. Two transistors OC 71 are used as A.F. amplifiers whilst the matched pair of transistors 2-OC 72 operates in the output stage of the receiver.
This receiver requires three separate batteries, namely one of 1.4 V, one of 6 V and one of 67.5 V. A 200 mW battery receiver equipped with the tubes DK 96, DF 96, DAF 96 and DL 96 has a heater current drain of 125 mA at 1.4 V and an anode current drain of 10.6 mA at 90 V. The mixed tube-transistor battery set described below has a current drain of 30 mA at 6 V together with 50 mA at 1.4 V and an anode current drain of 3.2 mA at 67.5 V. In the first case the total power consumption is therefore 190 mW + 950 mW, and in the second case 250 mW + 216 mW (L.T. plus H.T.). On the one hand there is a small increase in low-tension power consumption, but on the other hand there is a very substantial saving in high-tension power consumption, which is the more interesting since the cost of high-tension batteries compares unfavourably with that of low-tension batteries. The disadvantage of three batteries being needed in this circuit instead of two in an all-tube battery set is not of great importance because the current consumption of the H.T. battery is so low that replacement of the latter is seldom required. In connection with these considerations we refer furthermore to the next chapter, where a tube transistor receiver using a single battery is discussed.

CIRCUIT DESCRIPTION

Fig. 82 shows the circuit diagram of the mixed battery set. The A.F. and I.F. stages are equipped with tubes, because transistors in the present stage of development are less suited for use at frequencies above the super audible range. By using the tube DK 96 as a frequency changer and the DF 96 as an I.F. amplifier, full advantage is taken of the performance of these tubes both in the medium and in the short wave broadcast bands, up to frequencies for which no transistors are yet available. The rest of the set is equipped with a germanium diode and four transistors, the features of which are fully used. This design may thus be considered as an example of a lucky combination of tubes and transistors, both of which are used to their full advantages.

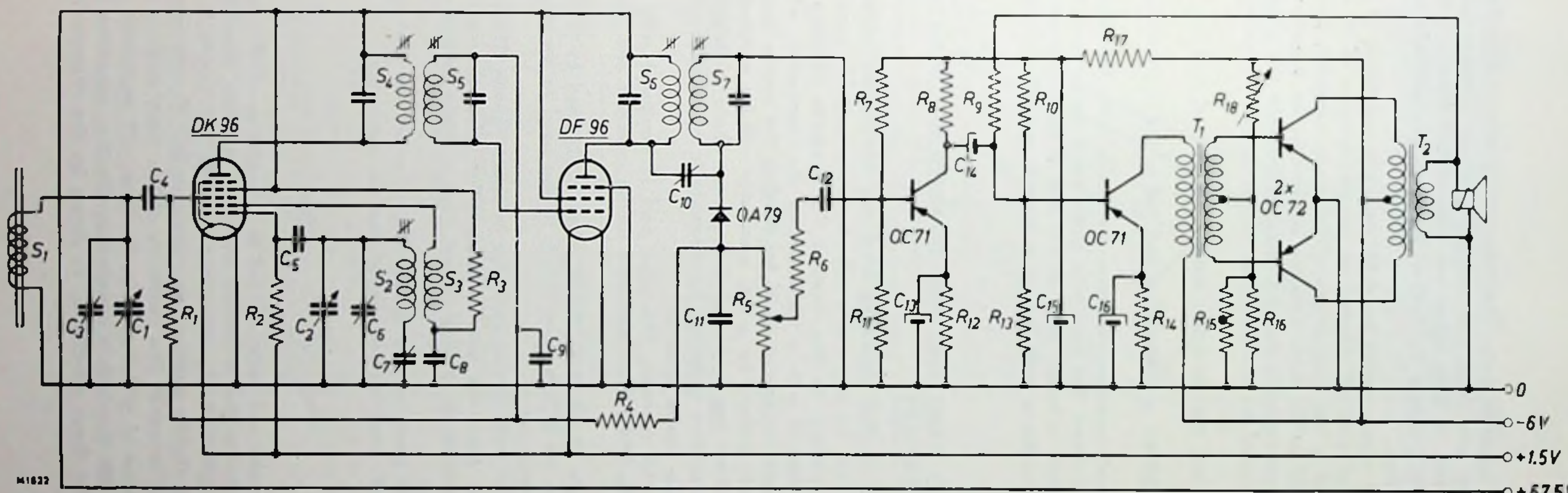


Fig. 82. Battery receiver equipped with the DK 96, DF 96, OA 79, OC 71 and 2-OC 72.

R_1 = 1 MΩ	R_8 = 5.6 kΩ	R_{15} = resistance at 25 °C 130 Ω (B = 45 °K)	C_4 = 120 pF ceramic	C_{11} = 1000 pF ceramic
R_2 = 27 kΩ	R_9 = 100 kΩ	R_{16} = 82 Ω	C_5 = 47 pF ceramic	C_{12} = 10 μF 3 V
R_3 = 18 kΩ	R_{10} = 39 kΩ	R_{17} = 150 Ω	C_6 = 6—30 pF trimmer	C_{13} = 32 μF 3 V
R_4 = 2.2 MΩ	R_{11} = 15 kΩ	R_{18} = 3 kΩ variable	C_7 = 200—500 pF wire trimmer	C_{14} = 32 μF 3 V
R_5 = 50 kΩ variable	R_{12} = 1.8 kΩ	$C_{1,2}$ = 2 × 500 pF variable	C_8 = 0.017 μF paper	C_{15} = 100 μF 12.5 V
R_6 = 8.2 kΩ	R_{13} = 18 kΩ	C_3 = 6—30 pF trimmer	C_9 = 0.1 μF paper	C_{16} = 100 μF 3 V
R_7 = 82 kΩ	R_{14} = 470 Ω		C_{10} = 1.5 pF ceramic trimmer	

S_1	Ferroxcube rod-aerial; coil 42 turns litz wire 24 × 0.04 closely wound on former of 12 mm ø; ferroxcube rod type C-9.5/203-IVB. The inductance is approximately 200 μH; adjustment takes place in the receiver.
S_2, S_3	Oscillator coil A3 125 730.
S_4, S_5 and S_6, S_7	I.F. transformers type AP 1001/52.
	Transformers T_1 and T_2 are identical to those of the amplifier for record players (page 65).

The circuit of Fig. 82 consists in fact of two separate parts. The tube part is quite conventional and no detail in the circuitry differs in any way from that published in our documentation "D 96-Series of Battery Tubes". The filaments of both tubes are parallel fed from a 1.4 V battery, and a 67.5 V battery is used for the H.T. supply. For further details of this part reference is made to the above-mentioned documentation.
The second part of the receiver begins with a new circuit for A.M. detection consisting of the germanium diode OA 79 with a capacitor of 1000 pF shunted by a potentiometer of 50 kΩ, and with a series resistor of 8.2 kΩ included in the base circuit of the input transistor. In this circuit the performance of the diode OA 79 as an A.M. detector appears to be superior to that of a tube diode. The only additional requirement is that the coupling between the primary and the secondary of the I.F. transformer should be increased. In this circuit this is achieved by means of the coupling capacitor C_{10}.
The automatic volume control is derived from the detection diode OA 79 and acts on the control grid of the DK 96 and the DF 96.
The rectified A.F. signal is taken from the 50 kΩ potentiometer and fed to the first transistor OC 71 of the A.F. part of the receiver.
The overall performance of this experimental receiver compares favourably with the battery receiver equipped with four tubes mentioned above. The sensitivity of the set is the same as that of the all-tube set. The frequency characteristic is better, both at low and at high frequencies, and the output power of 200 mW is in accordance with that of a normal portable broadcast receiver.

VII. 8. BATTERY RECEIVER WITH TWO TUBES AND FIVE TRANSISTORS OPERATING FROM A SINGLE 6 V SUPPLY BATTERY

In addition to the discussed tube-transistor receiver operating from two low-tension batteries and a 67.5 V anode battery, a description will be given of a tube-transistor portable receiver which operates from a single 6 V supply battery. In this receiver the necessity of using an H.T. anode battery is eliminated by providing it with a transistorized d.c. convertor, which converts the 6 V supply voltage into a direct voltage of 45 V for feeding the anode circuits of the tubes. This replacement of the anode battery by a d.c. convertor will be particularly interesting and economical in receivers which are to be used in areas where high-tension batteries are expensive compared with low-tension batteries, for example in tropical countries where much trouble is experienced in storing anode batteries.
Fig. 85 shows the circuit of an experimental receiver. The tube part is equipped with the battery tubes DK 96, DF 96 and DM 70. The DM 70 operates as a tuning indicator and, moreover, as an on-off pilot light.
An attractive solution for the filament supply of the three 1.4 V tubes is obtained by feeding their filaments in series with a resistor from the 6 V battery, which is also used as the supply battery for the transistor and the d.c. convertor.
The total power consumption of this receiver with its d.c. convertor is approximately 515 mW, and although this value is slightly higher than the power consumption of the previously described tube-transistor receiver, the advantages of feeding the set from a single battery very well outbalances this increased consumption.
In the circuit diagram shown in Fig. 85 four parts can be distinguished, namely the tube part, the detector with the A.G.C. circuit, the A.F. amplifier and, finally, the d.c. convertor.

THE TUBE PART

The input signal, originating from a ferroxcube rod aerial, is fed to the grid g_3 of the mixer DK 96, the circuit of which is identical to that in the previously described receiver, apart from the positive base pin of the filament of the DK 96 being connected to earth. This results in the potential of the grid g_3 of the DK 96 being slightly positive (approximately 0.7 V) with respect to the cathode, since this grid is connected to earth. The sensitivity of the input stage is thus increased; the damping due to the occurrence of grid current is, however, still negligible.
The tubes operate at an anode voltage of 45 V, which is supplied by the d.c. convertor. The performance of the tubes is still good under these conditions, as shown by the characteristics of the DK 96 and DF 96 on the next page.

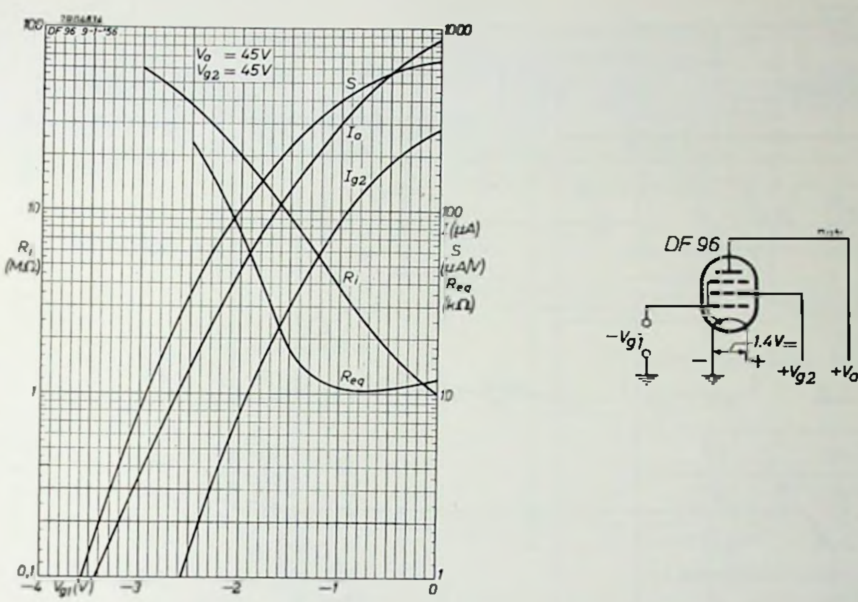


Fig. 83. Anode and screen-grid currents (I_a and I_{g2}), mutual conductance (S), the internal resistance (R_i) and the equivalent noise resistance (R_{eq}) of the tube DF 96, as functions of the control-grid voltage (V_{g1}) at a supply voltage of 45 V.

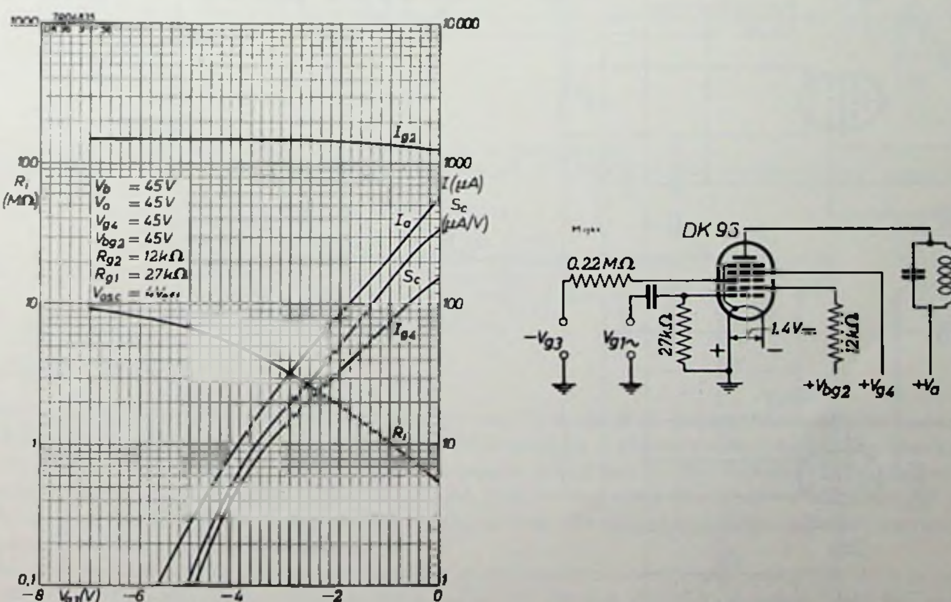


Fig. 84. Performance of the tube DK 96 as a function of the control-grid voltage (V_{g3}), at a supply voltage of 45 V, in a circuit as indicated. The oscillator voltage (V_{g1}) is 4 V_{eff}.

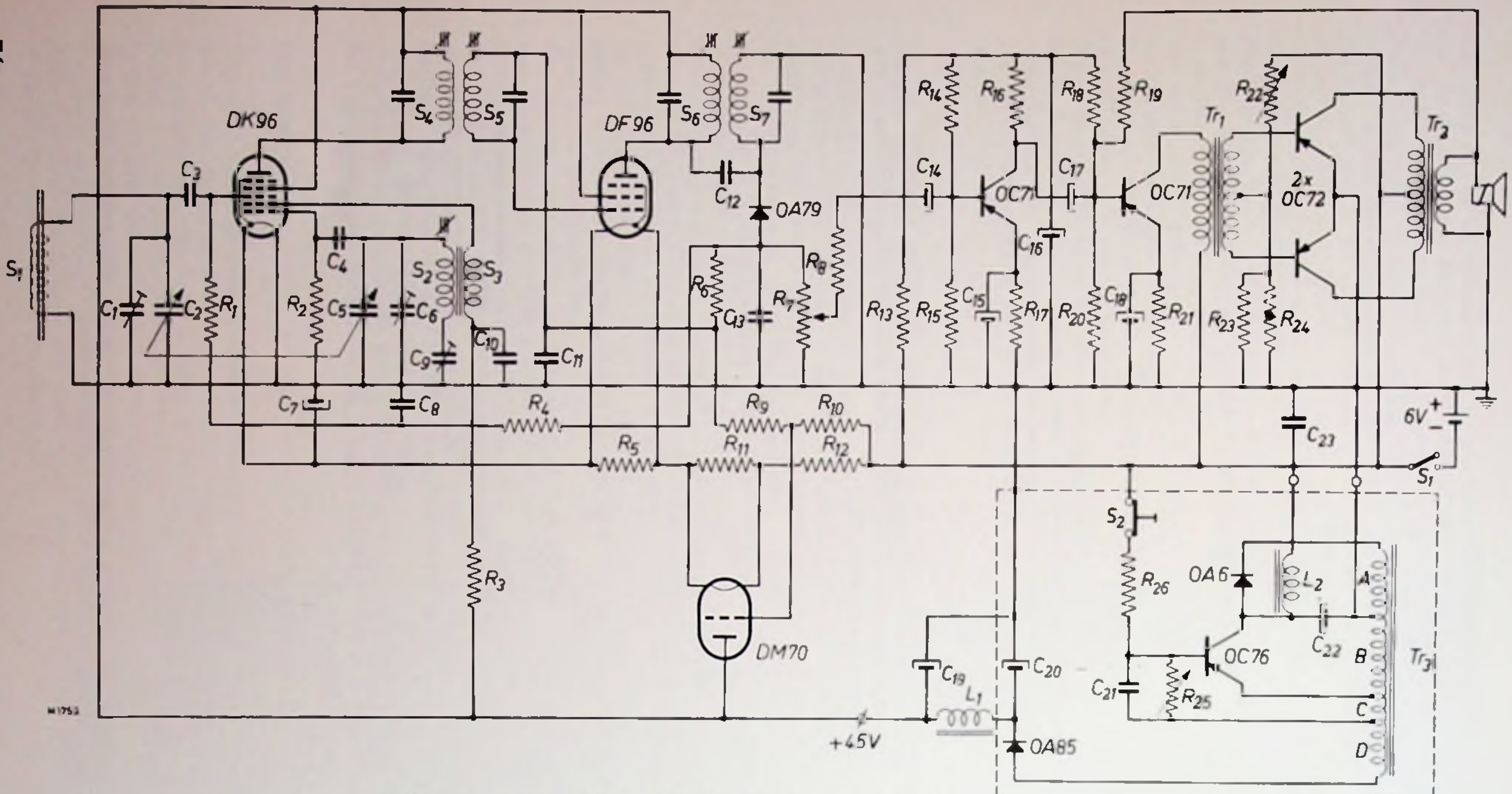


Fig. 85. Tube-transistor battery receiver with d.c. convertor.

R_1 = 1 MΩ	R_{11} = 420 Ω	R_{21} = 470 Ω	C_4 = 47 pF ceramic	C_{15} = 32 μF 6 V
R_2 = 27 kΩ	R_{12} = 68 Ω	R_{22} = 1—3 kΩ variable	C_6 = 6—30 pF ceramic	C_{16} = 100 μF 12.5 V
R_3 = 12 kΩ	R_{13} = 150 Ω	R_{23} = 82 Ω	C_7 = 8 μF 6 V	C_{17} = 32 μF 3 V
R_4 = 1 MΩ	R_{14} = 82 kΩ	R_{24} = 130 Ω at 25 °C. B = 4500 °K code nr. B8 320 00A/130E	C_8 = 0.047 μF paper	C_{18} = 100 μF 12.5 V
R_5 = 820 Ω	R_{15} = 15 kΩ	R_{25} = 1 kΩ variable	C_9 = 470 pF trimmer	C_{19} = 0.1 μF paper
R_6 = 3.9 MΩ	R_{16} = 5.6 kΩ	R_{26} = 2.2 kΩ	C_{10} = 0.047 μF paper	C_{20} = 3.2 μF 70 V
R_7 = 50 kΩ variable	R_{17} = 1.8 kΩ	C_1 = 6—30 pF trimmer	C_{11} = 0.047 μF paper	C_{21} = 0.05 μF
R_8 = 8.2 kΩ	R_{18} = 39 kΩ	$C_{2,5}$ = 2 × 500 pF variable	C_{12} = 1.8 pF ceramic	C_{22} = 100 μF 12.5 V
R_9 = 2.7 MΩ	R_{19} = 100 kΩ	C_3 = 120 pF ceramic	C_{13} = 0.001 μF ceramic	C_{23} = 0.068 μF
R_{10} = 1.8 MΩ	R_{20} = 18 kΩ		C_{14} = 10 μF 6 V	

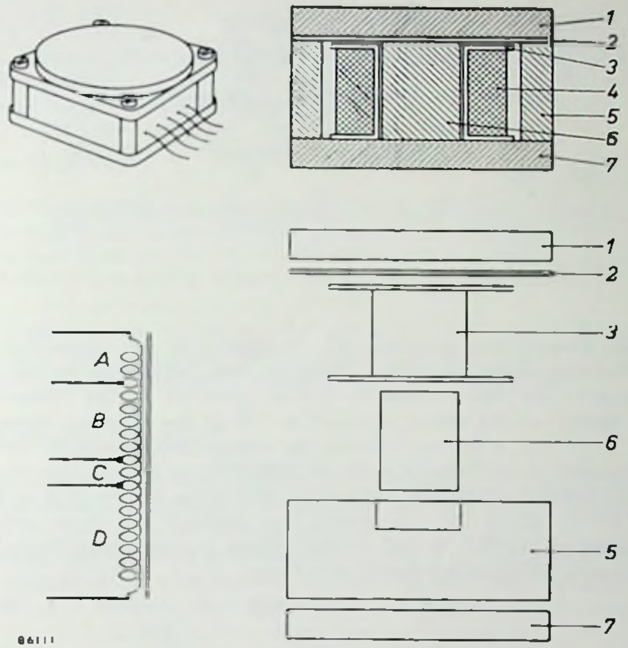


Fig. 86. Data of the transformers and of the coils of Fig. 85.

Transformers Tr_1:

Winding A silk covered copper wire 0.25 mm, 31 turns
Winding B silk covered copper wire 0.25 mm, 84 turns
Winding C silk covered copper wire 0.25 mm, 15 turns
Winding D silk covered copper wire 0.25 mm, 131 turns
The transformer is wound as an auto-transformer, without insulation between the windings. The air gap is 60 μ.

1—7 discs ferroxcube
2 paper 60 μ
3 coil former, 88 488
4 coil
5 ring ferroxcube
6 core ferroxcube

The pot core is available under type number D 25/16-10.00-3B2.
Transformers Tr_1 and Tr_2 are identical to those of the amplifier for record players (page 65).

L_1 = 1 mH; L_2 = 1 mH
S_1: Litz wire, 43 turns, closely wound; coil former; 40 × 12 × 10 mm; core: ferroxcube type no. 56 681 62/4B.
S_2: 108 μH, enamelled copper wire 0.07 mm, 109.5 turns;
S_3: enamelled copper wire 0,07 mm, 70.5 turns coil former 7 mm diam.; coil length 2 mm; core powdered iron: $t = S_3/S_2 = 1/2.1$.
S_4, S_5 and S_6, S_7: I.F. transformer type AP 1001/52.

FILAMENT SUPPLY

The series-connected filaments of the tube are fed from the 6 V battery; the emission currents, which also flow through the filaments, are compensated by resistors shunted across the filaments. Full details on this type of filament current supply are given in the Bulletin "D 96 Series of Battery Tubes". It should be noted, however, that in this case the positive base pin of the filament of the DK 96 is connected to earth, so that the direction of the filament current is opposite to the customary direction.

The filament current is 25 mA, the voltage across each filament then being 1.4 V. It is true that the normal maximum ratings of these tubes are 24 mA and 1.3 V respectively for series filament supply, but in this special case, where the supply consists of a low-voltage battery and the tubes moreover operate at a low anode voltage, a filament current of 25 mA is permissible.

At an anode voltage of 45 V the emission currents of the DK 96, DF 96 and DM 70 are

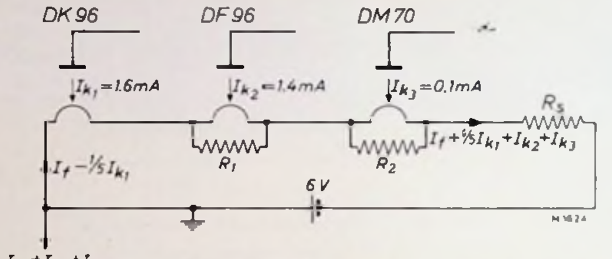


Fig. 87. Filament supply circuit of the tubes.

1.6, 1.4 and 0.1 mA respectively (see Fig. 87). According to the above-mentioned Bulletin it may be assumed that 1/5 of the emission current of each tube flows to the positive base pin and 4/5 to the negative pin. The resistor R_s is thus traversed by the filament current and by the total emission current of the tubes, reduced by 1/5 of the emission current of the DK 96, which gives 25 + 3.1 — 0.3 = 27.8 mA. Since the voltage drop across R_s should be 6 — 3 × 1.4 = 1.8 V, this resistor should have a value of 1800/27.8 = 64.8 Ω. The current entering the filament of the DM 70 will be 27.8 mA reduced by 4/5 of I_{k3}, hence 27.8 — 0.08 = 27.7 mA. This means that 27.7 — 25 = 2.7 mA should flow through the shunt resistor R_2, which is therefore given a value of 1400/2.7 = 520 Ω, the voltage across R_2 thus being 1.4 V.
It can be calculated in a similar way, that R_1 should be:

$$\frac{1400}{2.7 - 4/5\, I_{k2}} = \frac{1400}{2.7 - 1.1} = 880\ \Omega$$

It will be convenient to choose a 5 % standard resistor of 68 Ω for R_s. The filament current and the voltage distribution across the filaments will then deviate slightly from the values calculated. The deviation is, however, so small that the results of the calculation remain valid.

DETECTOR PART AND A.G.C. CIRCUIT

The detector part of the receiver is identical to the corresponding part of the tube-transistor receiver with anode battery (chapter VII. 7.). The A.G.C. voltage is also taken from the detection diode OA 79 and acts on the control grids of the DK 96 and DF 96.
The voltage divider formed by R_6, R_9, and R_{10} is so dimensioned that the control grid g_1 of the DF 96 has the same potential as its negative filament terminal when no signal is applied to the top of the potentiometer R_7. The same measure has been taken for the tuning indicator DM 70.
The A.G.C. characteristic of the complete receiver has been plotted in Fig. 88.

Fig. 88. A.G.C. curve of the receiver.

A.F. AMPLIFIER

The rectified A.F. signal is taken from the 50 kΩ potentiometer R_7 and fed to the first transistor OC71 of the A.F. amplifier.
For a description of the D.C. convertor see page 91.

SUPPLY OF THE RECEIVER

A nickel-cadmium accumulator was used for the supply of the receiver according to Fig. 85. Although a dry battery could be used for this purpose, it is preferable to employ an accumulator because it usually has a longer life, its internal resistance is smaller, and it can be recharged. The advantages of only one small 6 V battery being required to operate the receiver described above are obvious and have already been mentioned.
However, depending on local conditions, the use of H.T. batteries may sometimes not be so objectionable, and it may then very well be possible to use a 45 V H.T. battery for feeding the anodes of the tubes, thus rendering the d.c. convertor superfluous.

PERFORMANCE OF THE RECEIVER

The overall performance of this tube-transistor receiver with d.c. convertor compares very well with that of the previously described tube-transistor receiver. The frequency characteristic and the output power are the same for both receivers.
The sensitivity of the receiver with d.c. convertor is 50 μV at 1 Mc/s for an output of 50 mW. Owing to the application of transistors in the A.F. part of the receiver and the use of a germanium diode as a detector, the distortion of this experimental receiver could be kept fairly small.

VII. 9. I.F. AMPLIFIER WITH OC 45 P-N-P TRANSISTORS

Because the output capacitance of the OC45 is very low by transistor standards (collector capacitance $C_{b'c}$ = 12.5) and the internal base-resistance $R_{bb'}$ is also low ($R_{bb'}$ = 75 Ω), the internal reaction in it is small.
The I.F. frequency is 455 kc/s and power amplification 59 db; the input and output impedances are 70 kΩ and 1 kΩ, respectively.
The circuit does not include AGC. All supply points are decoupled by means of RC-filters, and the internal reaction is neutralized by negative feedback from collector to base. At the same time it is difficult to prevent transistorized I.F. amplifiers from oscillating.

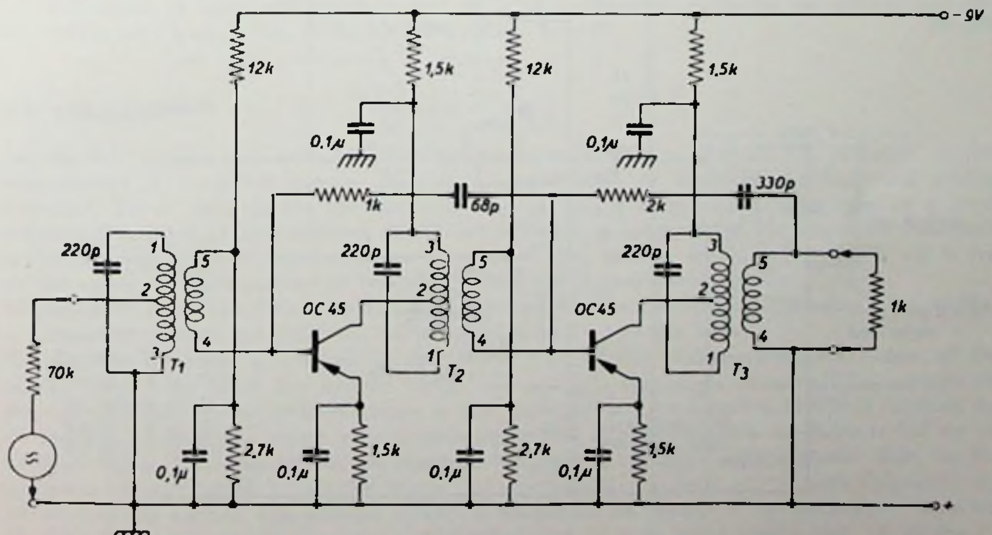


Fig. 89

The winding ratios of the I.F. transformers employed (cores D14/8/04) are as follows:

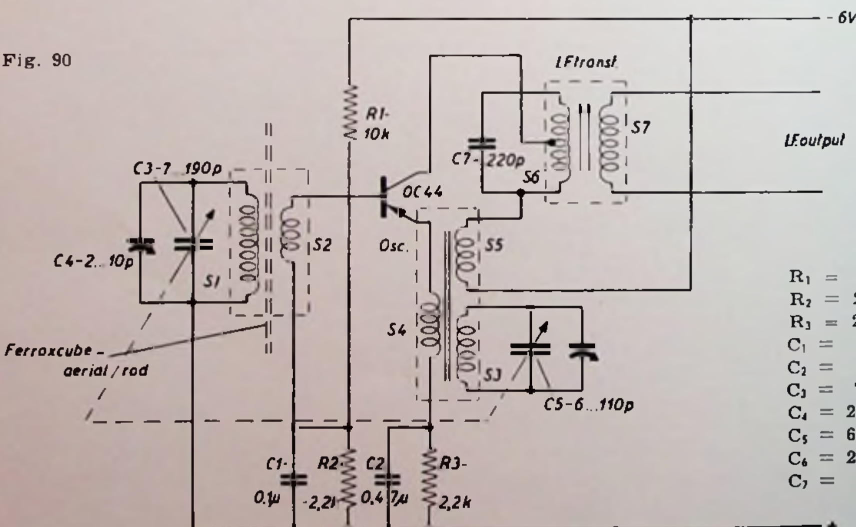
	TR_1	TR_2	TR_3
$\frac{2-3}{1-3}$	0,8	0,315	0,26
$\frac{4-5}{1-3}$	0,0458	0,0554	0,0955
Q_u	110	70	110
Q_l	35	35	35

VII. 10. SELF-OSCILLATING MIXER STAGE WITH OC44

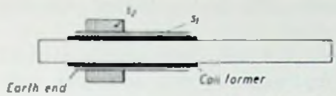
Fig. 90 shows the circuit diagram of a transistor mixer stage with an OC44 transistor. This has a frequency limit of 7—15 Mc/s, and is therefore very suitable for oscillating and mixer circuits operating in the medium-wave range. Although the mixer stage could be provided with separate oscillator and mixer transistors, it is more economical to employ a self-oscillating mixer transistor, as in that case only one OC44 is required.

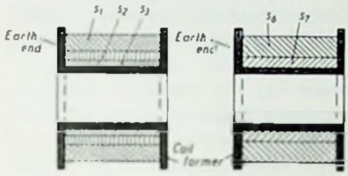
Conversion gain is in the region of 28 db for the single transistor, and for the 2-transistor models. This gain is defined as the ratio of the I.F. power developed in a 680 Ω load resistor, connected to the secondary of the I.F. transformer, to the R.F. energy available in the aerial circuit; 680 Ω is the average input resistance of OC45 transistors, as employed in the I.F. amplifier following the mixer.

Fig. 90



R_1 = 10 kΩ
R_2 = 2.2 kΩ
R_3 = 2.2 kΩ
C_1 = 0.1 μF
C_2 = 47000 pF
C_3 = 7—190 pF
C_4 = 2— 10 pF
C_5 = 6—110 pF
C_6 = 2— 10 pF
C_7 = 220 pF





Antenna coil

S_1 77 turns of 32 × 0.04 silk insulated Litz wire, closely wound on a former, diameter 12 mm.
Rod: ferroxcube 4 B, dimensions: 10 × 200 mm.
Unloaded Q at 1 Mc/s: 150 (mounted in chassis)
L: 480 μH

S_2 5 turns of 0.3 mm enamelled copper wire, wound at the earth side of S_1.

Oscillator coil

The oscillator coil is mounted in a potcore D 18/12, ferroxcube 3B3, airgap 1 mm.

S_3 51 turns of 32 × 0.04 silk insulated Litz wire.
Unloaded Q: 55 at 1.5 Mc/s.

S_4 2 turns of 0.3 mm enamelled copper wire.

S_5 5 turns of 0.3 mm enamelled copper wire.

I.F. transformer

The I.F. transformer is mounted in a potcore D 18/12 ferroxcube 3B3, airgap 0.3 mm.

S_6 65 turns of 16 × 0.04 silk insulated Litz wire, collector tap at 52 turns from earth side.
Unloaded Q: 110.

S_7 3 turns of 0.3 mm enamelled copper wire.

VII. 11. AN EXPERIMENTAL TRANSISTOR SUPERHET

INTRODUCTION

Present-day r.f. transistors are not very suitable for use in straight receivers on account of frequency-dependent feedback and low gain in the M.W. range. This produces circuits that are difficult to neutralize. It is, moreover, necessary to use a multiple-stage r.f. amplifier, which results in still greater instability, not to mention the synchronization problems which arise with the indispensable triple tuning capacitor. A high cut-off frequency of the transistors is necessary in order to obtain, in addition to a greater gain, a good signal/noise ratio.

With the superheterodyne circuit a better state of affairs in this respect exists, because here we are only concerned with one fixed frequency. A difficulty is still encountered in the mixing stage. The transistor(s) used in this stage must have a high cut-off frequency, because otherwise noise will be excessive. A high cut-off frequency ensures at the same time uniform oscillation over the whole band. The most satisfactory results are given here by the recently introduced OC44.

The i.f. amplifier, which, as is customary with transistor superhets, consists of two stages, is equipped with OC45 transistors. The first OC45 stage incorporates an AGC circuit, which gives excellent results in practice. The a.f. amplifier is the well-known 5-transistor amplifier with three OC71's and a matched pair of OC72's in class B.

Its construction is more difficult than that of a valve superhet, because the chassis and the i.f. transformers have to be made by the amateur himself.

THE DIAGRAM

For the first stage a self-oscillating mixing transistor OC44 was chosen. The radiation in this arrangement is somewhat greater than in a combination of separate oscillator and mixing transistor, but in view of the low oscillator energy this is hardly noticeable even at a small distance. Regarded as an oscillator, the OC44 is in the grounded-base configuration. Feedback is consequently effected from collector to emitter; the base is connected to earth via a tap on the aerial rod. This method of feedback is best for a transistor oscillator.

Regarded as an r.f. amplifier, the OC44 is in the grounded-emitter configuration. The emitter is connected to earth via L_5C_4; the aerial signal is applied to the base.

Because the transistor is adjusted to the curved part of the characteristic, the output of the mixing stage from which the desired 467.5 kc/s signal is filtered by means of T_1, appears on the collector side. In the collector chain is also incorporated the negative feedback winding L_4. To prevent the oscillator signal from penetrating to the i.f. amplifier, the oscillator is fed via an r.f. filter $R_{21}C_5$. The base derives the required bias from a common voltage divider R_1R_2 for the complete r.f. part, which brings the "earth" to a potential of approx. —1 V with respect to the terminal of the battery. The voltage divider is decoupled by means of an electrolytic capacitor C_3. The impedance of the divider is sufficiently low to ensure good stabilization of all the i.f.

transistors. A further advantage of a common divider is the considerable saving in material and space.

The resistor R_3 adjusts the emitter current of V_1 to approximately 0.4 mA, with which a good conversion gain ratio is obtained. Decoupling of R_3 is effected by means of C_4, which is connected directly to the stator of the double capacitor C_1C_2. C_p is the oscillator padder, which can be omitted if a special miniature double capacitor with reduced oscillator section is used. Since this proved to be unobtainable, a normal 2 × 500 pF double capacitor was used with a padder of 470 pF (approx. 5 %).

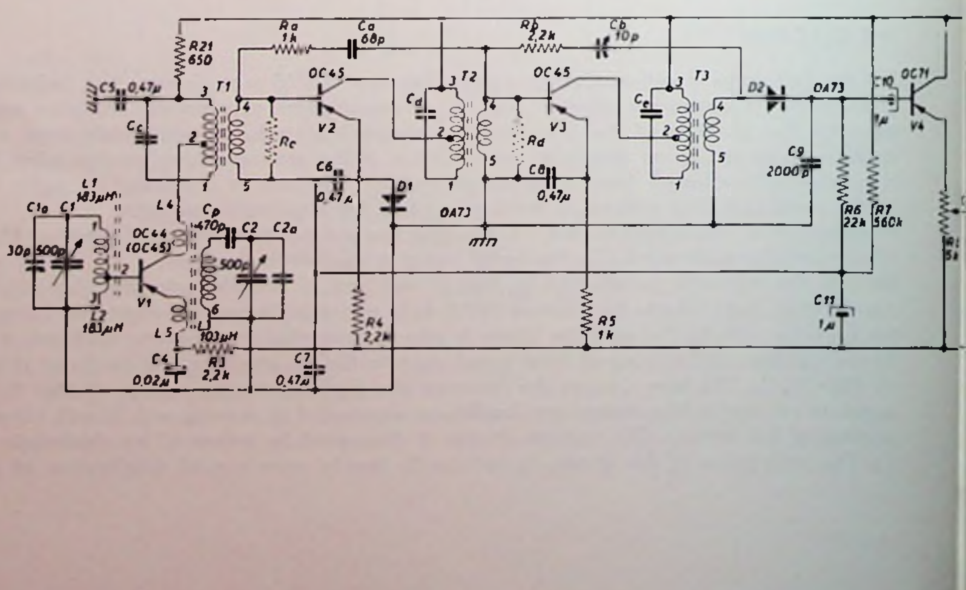
V_2 is included in the AGC circuit, which receives its base current via the detector diode D_2, and R_6 and R_7. Owing to this, this transistor draws an emitter current (approximately 0.5 mA) such that the voltage drop across R_4 is just slightly greater than the potential across R_2. The result is that the delay diode D_1 normally blocks owing to its anode being negative with respect to the cathode. As a result of the (positive) discharge current of C_9, the i.f. signal detected by D_2 influences the potential of C_{11}-C_7, which thus becomes slightly less negative. Since the base potential of V_2 falls at the same time, its emitter current will decrease. As long as the i.f. signal is weak, the potential across R_3 will change but little and D_1 will thus continue non-conductive. Since the emitter potential "follows" the base potential, the amplification changes but little. A strong i.f. signal, on the other hand, will cause the base potential of V_2 to drop to such an extent that the emitter current is greatly decreased, and consequently the voltage drop across R_3 is reduced and diode D_1 becomes conductive. If this happens, however, the emitter of V_2 will be maintained at "earth" potential; the transistor thus becomes conductive and the amplification decreases. In this manner a delayed AGC is obtained.

The next i.f. stage is adjusted to a collector current of approximately 1 mA, at which value it gives maximum amplification.

R_AC_A and R_BC_B serve to neutralize the internal feedback of the transistors V_2 and V_3, thus reducing the tendency to oscillation.

If during trimming it appears that this tendency cannot be eliminated with C_AC_B, then resistors R_c and R_d, whose values have to be determined experimentally, must be fitted. Their order of magnitude lies between a few hundreds and a few tens of ohms, but is in any case such that the Q_1 of the circuit under load becomes about 30. Especially the Q_1 of T_2 must not become too high.

Apart from the complication in respect of the AGC circuit, the detector circuit is of a fairly conventional type. C_{11} determines, in conjunction with R_7, the AGC time constant, and therefore must not be too large. However, since C_{11} serves at the same time as a.f. decoupling of R_6, the reactance would become excessive if V_5 were directly connected via C_{13} to D_2. This would result in a serious attenuation of the lower audio frequencies. V_4 has therefore been inserted, and is connected as an emitter follower, thus raising the input impedance of the amplifier to such a high level that no low-frequency attenuation occurs. An additional advantage is the



smaller detection distortion at large modulation depths and the absence of the otherwise necessary matching series resistor in the base circuit of the input transistor, which absorbs half of the power supplied.
The volume control is incorporated in the emitter circuit of V_4. This control should preferably be a wire-wound component, as otherwise crackling will occur when it is being adjusted. A carbon potentiometer must be connected, via a separating electrolytic capacitor of 8 μF, across the fixed 560 Ω emitter resistor which must be fitted in this case [1]).
The a.f. amplifier has already been described previously in this paper and therefore needs no further explanation.

THE 6-V SUPPLY

For the 6-V supply two 3-V cells are fitted one on top of the other in a space reserved next to the tuning capacitor. The contacts intended for it are partly fixed to the pertinax chassis and partly to the small partition; the battery panel is fitted with an interconnecting strip.

CONSTRUCTION

Since very low impedances are used in transistor circuits, there is no need to pay much attention to capacitive screening. One should however guard against inductive and galvanic couplings. As is shown in figs. 95 and 96, the "chassis" consists of a 2 or 3 mm thick pertinax plate, on which all components are mounted. On it is riveted also an aluminium front plate flanged on one side, to which are fitted the volume control, the on/off switch, the output transistors and the scale. The fixing of the resistors, capacitors and transistors is effected by means of the hollow insert bushes riveted in the pertinax. The wire ends of the coils are likewise finished by having insert bushes fitted to them.
This method of mounting, which greatly resembles that of printed circuits, makes possible a very compact construction. The size of the chassis depends on the available components. All this requires a thorough preparation, in which a cardboard trial chassis renders useful service. Due account must be taken of the dimensions of the electrolytic capacitors to be used. In some radio shops sub-miniature types are already obtainable.
(The most economical mounting is vertically on the chassis. To this end one wire is inserted into the relevant insert bush, the other wire being bent downward and fixed in a second bush by means of a piece of strong extension wire. It is advisable to use soldering springs in case of multiple-wire joints.)
Sufficient space must, of course, be left for the loudspeaker and for the rotating plates of the tuning capacitor.

[1]) In the test model this decoupling is omitted.

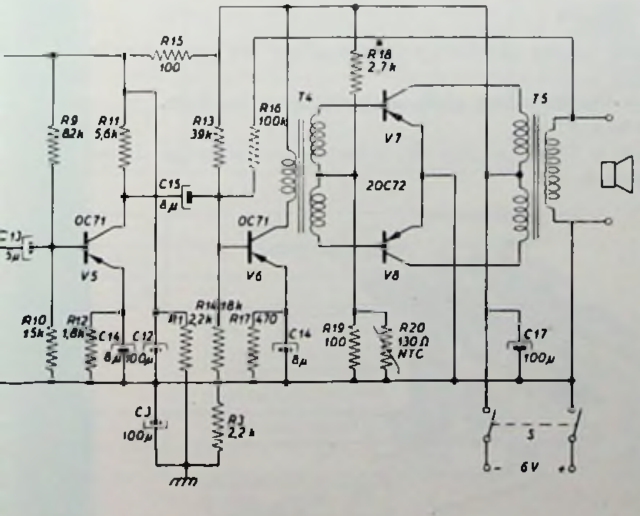


Fig. 91

After all holes and insert bushes have been carefully marked off on the cardboard test chassis, they can be transferred to the pertinax plate. After this the insert bushes are riveted in and the front plate is fixed; next the partition for the battery space and the contacts for the battery are fixed, likewise by means of insert bushes.
When riveting the insert bushes into the pertinax plate, care must be taken not to tear or break the plate. Hence the riveting support should be sufficiently rigid and the rivets should not be driven in too tightly. First use a blunt centre punch or ball for stretching the rim, then further flange it carefully by means of a flat hammer. (Practise a few times on a piece of spare material.)
For fixing the resistors etc., fluted pins can be used instead of hollow soldering rivets. These are steel pins having a longitudinal groove, which enables them to be fitted tightly into the pertinax. They have the drawback, however, of being difficult to solder and of oxidizing in the course of time. It might be possible for the amateur to make them himself from phosphor-bronze aerial wire. The groove is then pinched in with cutting pliers. The vertical mounting of the capacitor by means of fluted pins is a bit more difficult, since they are not readily accessible for soldering.
Before mounting the larger components, the necessary connecting wires are laid between the various insert bushes. The ends are not yet soldered, but are inserted through the holes and bent. Then the i.f. transformers are mounted. The previously cleaned and tinned litz ends are soldered in the appropriate bushes via the lead-ins. Next the screening bush is fitted and secured with M_2 fixing bolts to the coil body.
After mounting the transformers, tuning capacitor, volume control, etc., the fitting of the capacitors can be undertaken. From time to time a check should be made that sufficient space is left for the loudspeaker. Where more than one wire meet in a bush, soldering should not be

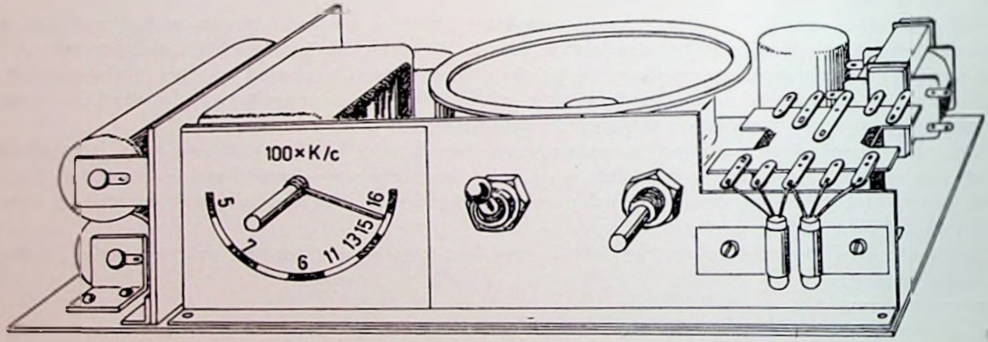


Fig. 92

done until all wires are fitted. The wire ends of the resistors and capacitors are provisionally bent in the manner suggested above.
The double capacitor is directly screwed on the mounting plate without fixing brackets.
For the necessary a.f. push-pull transformer see VII. 4.
On the driver-transformer a mounting plate is fitted, on which output transistors, the driver transistor and a few other parts can be mounted.

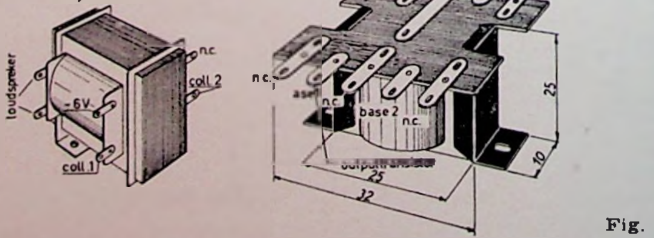
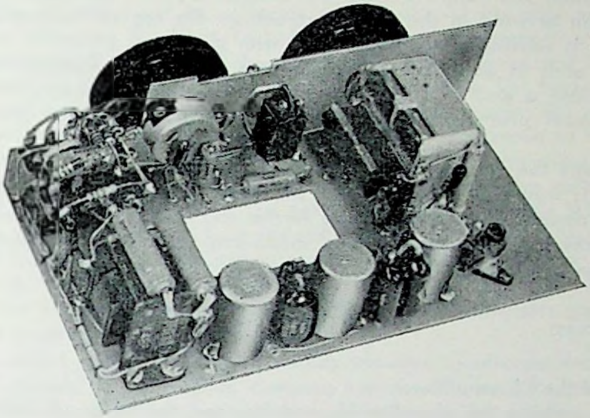


Fig. 93

THE I.F. TRANSFORMERS AND THE OSCILLATOR COIL

The i.f. transformers can be fairly easily made by the amateur himself. They are wound on to the well-known pin or dumb-bell cores which are placed in screening bushes. The latter are made from small aluminium tubes.



(The use of brass or iron bushes is not advisable on account of the greater losses involved. In view of the small diameter of the bush, the quality factor of the non-screened coil must be as large as possible (Q_0 = approx. 170). In this way a Q of approx. 75 can be obtained when the screen is in position.)
Appropriate tappings cause the Q to drop during operation to approx. 30, which is the value required for a good bandwidth (6 kc/s for 3 db).

TABLE 1
I.F. TRANSFORMERS

	T_1	T_2	T_3
1—2	46 turns	157 turns	170 turns
2—3	184 ,,	73 ,,	60 ,,
4—5	11 ,,	13 ,,	67 ,,
Q_0*)	75 ,,	75 ,,	75 ,,
C_0 to C_0	220 pF	220 pF	220 pF

OSCILLATOR COIL **)

1—2	3 turns 0.2 enamelled
3—4	8 ,, 0.2 enamelled
5—6	85 ,, 12 × 0.04 enamelled

L_{osc} = 103 μH
Padder 472 pF

AERIAL COIL ***)

1—2	90 turns 12× 0.04 mm
2—3	5 ,, 12 × 0.04 mm

L_{ant} = 183 μH

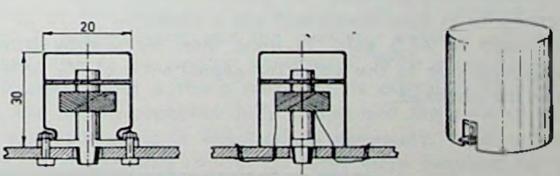


Fig. 94. Construction of the I.F. transformer.

*) In aluminium tubes 20 mm ø, 30 mm high. Litz wire 12 × 0.04 mm. Dumb-bell cores. I.F.467.5 kc/s.

**) Wound on dumb-bell cores (not screening).

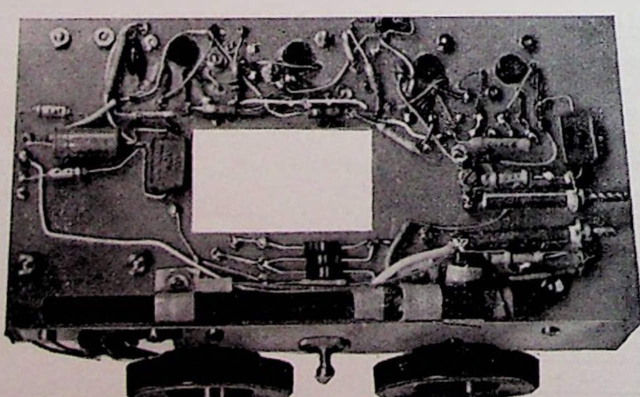
***) Directly wound on a ferrite rod ø 8 mm

The litz ends must be clearly identified; to this end small pieces of coloured insulating sleeving or small beads should be fitted as close as possible to the coil. Wave-winding gives of course the best appearance, but ordinary hand-winding also gives satisfactory results. The stated Q applies to the latter winding method.
After the start is marked, the wire is wound as far as the tapping. A loop is made in the wire at this point, twisted into a ball and marked. Winding is continued until the required number of windings is reached. The "ball" is fixed with a drop of pure beeswax (in no case use stearin). Care must be taken to wind in the correct direction. On top of the primary the secondary is wound. The whole is subsequently impregnated with beeswax.
Cleaning the litz ends is done by rapidly dipping them when red-hot into spirit. The ends should be twisted into a kind of small wad, so that the heat capacity increases somewhat and the ends do not cool prematurely. If necessary, first practise a few times with a piece of waste wire.
Fig. 94 clearly shows the construction of the i.f. transformer.
The oscillator coil is wound in a similar way. Screening is not required and is even detrimental. The aerial coil is directly wound on to the ferrite rod and is likewise secured with beeswax. The number of turns has intentionally been made too high. During trimming the correct self-induction is obtained by unwinding.

We shall now discuss:

a) Trimming;
b) Neutralizing of the i.f. amplifier;
c) The a.f. part contains the amplifier V_5 and V_6, and the push-pull output stage V_7 and V_8 (see diagram fig. 91).

Because the coils are home-made, the trimming of the all-transistor superhet requires a somewhat greater dexterity than in a valve superhet with ready-made coils. An R.F. generator is, in fact, indispensable for this purpose, as well as an output meter and a valve voltmeter.
As already stated above, the transistors V_1, V_2 and V_3 are only soldered in during trimming.
First T_3 is set to 467.5 kc/s by means of the R.F. generator. When doing this, the loudspeaker is removed and replaced by the output meter (or alternating voltage meter with 5 Ω shunt, set to the most sensitive measuring range). The R.F. generator is coupled, via a 10 pF capacitor, to tapping 2 on T_3 (i.e. collector connection V_3). The R.F. generator is set to 467.5 kc/s, and the core of T_3 is adjusted till the maximum deflection of the output meter is obtained. If necessary, wind on more turns or unwind a few turns if no clear maximum is obtained; alternatively, slightly increase or reduce C_c (diagram 91). Next transistor V_3 is soldered in. Its collector current should be approximately 1 mA. If necessary, correct by shunting R_5. The R.F. generator is then connected via a capacitor of 10 pF to tapping 2 on T_2, after which T_2 is also adjusted to 47.5 kc/s. T_3 must then be slightly trimmed again in order to compensate the detuning due to the collector capacitance of V_3. When this has been done, the neutralization can be checked.



To do this, a valve voltmeter with 20 mV range (if necessary a transistor voltmeter) is connected between the frame of the tuning capacitor (= "earth") and the base of V_3. Next the R.F. generator is connected via a 0.1 μF capacitor to the collector of V_3, and the R.F. generator signal is adjusted to approximately 400 mV/467.5 kc/s (earth terminal to "earth"). Next should be removed C_b, and a small trimmer of 15 pF maximum capacitance inserted in its place. R_b is now temporarily short-circuited, and the trimmer is adjusted in such a way that the valve voltmeter indicates a minimum. When this has been done, the short-circuit from R_b is removed and V_2 is soldered in. At no signal its collector current should be approximately 0.5 mA. If necessary, correct by changing R_7. Next T_1 is trimmed. The R.F. generator is coupled, via a 10 pF capacitor, to tapping 2 on T_1, and the output meter is reconnected to the loudspeaker terminals. The control of the R.F. generator is then turned so far back that at maximum sensitivity of the output meter (and maximum setting of volume control R_8) a distinct maximum is just obtained, in order to prevent blurring of the maximum as a result of the AGC.
As soon as T_1 is trimmed, and T_2 is corrected again to remove the detuning resulting from the collector capacitance of V_2, this stage, too, can be neutralized. The valve voltmeter is connected between the base of V_2 and the frame of the tuning capacitor, and the R.F. generator signal of 200 mV/467.5 kc/s is applied, via a 0.1 μF capacitor, to the collector of V_2. Next R_a is temporarily short-circuited, C_a is removed and capacitors of 39 pF, 47 pF, 56 pF or 82 pF are successively fitted into place, corresponding deflections of the tube voltmeter being moved. The capacitor that gives the lowest reading is finally selected. If two capacitors give almost the same meter indication, the smaller one should be selected.
Finally C_1 can be mounted. Again T_1 must be slightly trimmed to remedy the detuning.
If i.f. oscillation still occurs, the additional damping resistors R_c and R_d must be fitted. Their values should be so chosen that they just prevent oscillation. If the winding data, type of wire and dimensions of the screening bush for the i.f. transformers, as specified above, are adhered to, there is but slight risk of oscillating. The possibility remains, however, that the Q_o of the coils turns out to be somewhat greater than expected, so that instability occurs. A damping resistor will then restore stable operation of the amplifier. The trimming of the aerial and oscillator circuits is effected as follows: First a simple scale is fitted consisting of a piece of white cardboard, fixed to the front plate extended for this purpose, and a pointer directly clamped on the capacitor spindle (helical spring with the end drawn out). On the cardboard an arc of a circle is drawn, on which the two extreme positions of the variable capacitor are indicated. The position at which the capacitor is fully opened, is marked 1600 kc/s, and that at which the plates are fully closed is marked 500 kc/s. The arc is then divided into 11 equal parts, and at each point the relevant frequency (500, 600, 700 kc/s, etc. up to 1600 kc/s) is marked.
We can now proceed to check the frequency range. The output meter is reconnected to the loudspeaker terminals and the R.F. generator coupled to the ferrite rod by bringing the R.F. generator cable near to the rod or winding it a few times around the rod. The R.F. generator is now set to 1600 kc/s, the tuning capacitor is fully opened and the oscillator trimmer adjusted until the output meter indicates a maximum. This is the coarse adjustment. With a temporarily fitted aerial trimmer, further adjustment is made until a sharp maximum is obtained. The R.F. generator is subsequently set to 508 kc/s, the tuning capacitor fully closed and the core of the oscillator coil adjusted until a maximum output reading is obtained. If necessary, wind on or unwind the aerial rod until a sharp maximum is obtained. About 50 turns were required for a 8 mm rod; the 95 turns originally specified are, as previously suggested, rather excessive. The set is now retuned to 1600 kc/s, and the oscillator and aerial trimmer readjusted, after which the set is again adjusted at 508 kc/s. Slight readjustments at both these frequencies are continued until no further improvement can be noticed.
When the required frequency range has thus been determined, an attempt should be made to replace the temporary aerial trimmer by a fixed capacitor of a very approximate value. This, for instance, was done in the original design to save space. From a slightly over-sized ceramic capacitor small bits were nipped off on one side by means of tweezers until the required value was obtained. It is advisable, however, to maintain a trimmer here, too. This must be fixed so that it does not come into contact with the oscillator trimmer. If necessary, it may be wrapped in cellophane, gummed tape, or inserted in a piece of wide insulating sleeving.
The trimming process proper can now be attempted, and is solely carried out with the oscillator controls. Nothing in the aerial circuit must be changed.
The R.F. generator is set to the first synchronous frequency of 1535 kc/s and the tuning capacitor is slowly opened from fully closed position until the output meter deflects. (A headphone in parallel with the output meter prevents confusion with other transmitters.) The indication of the

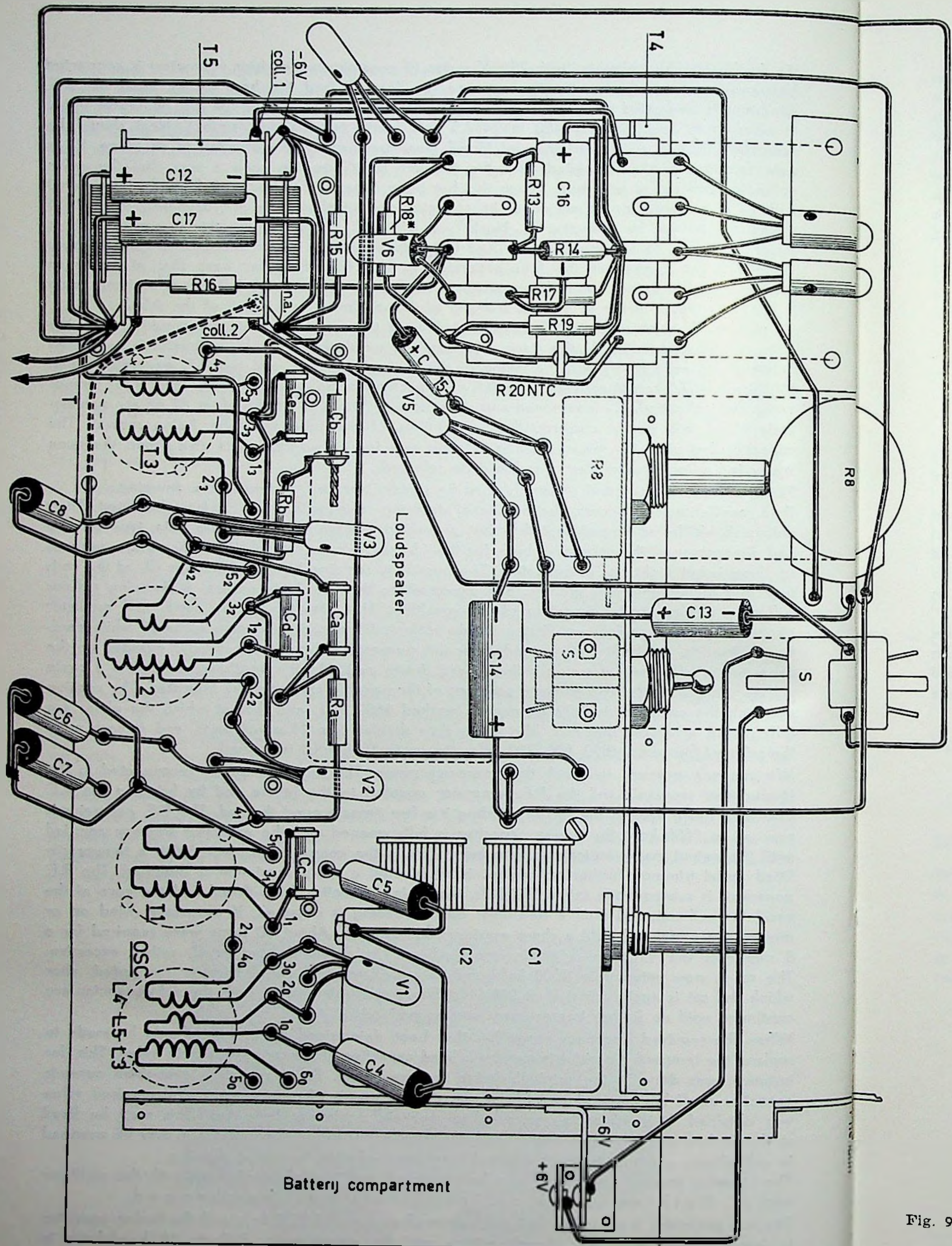

to loudspeaker
Loudspeaker
Batterij compartment
R 20 NTC

Fig. 95

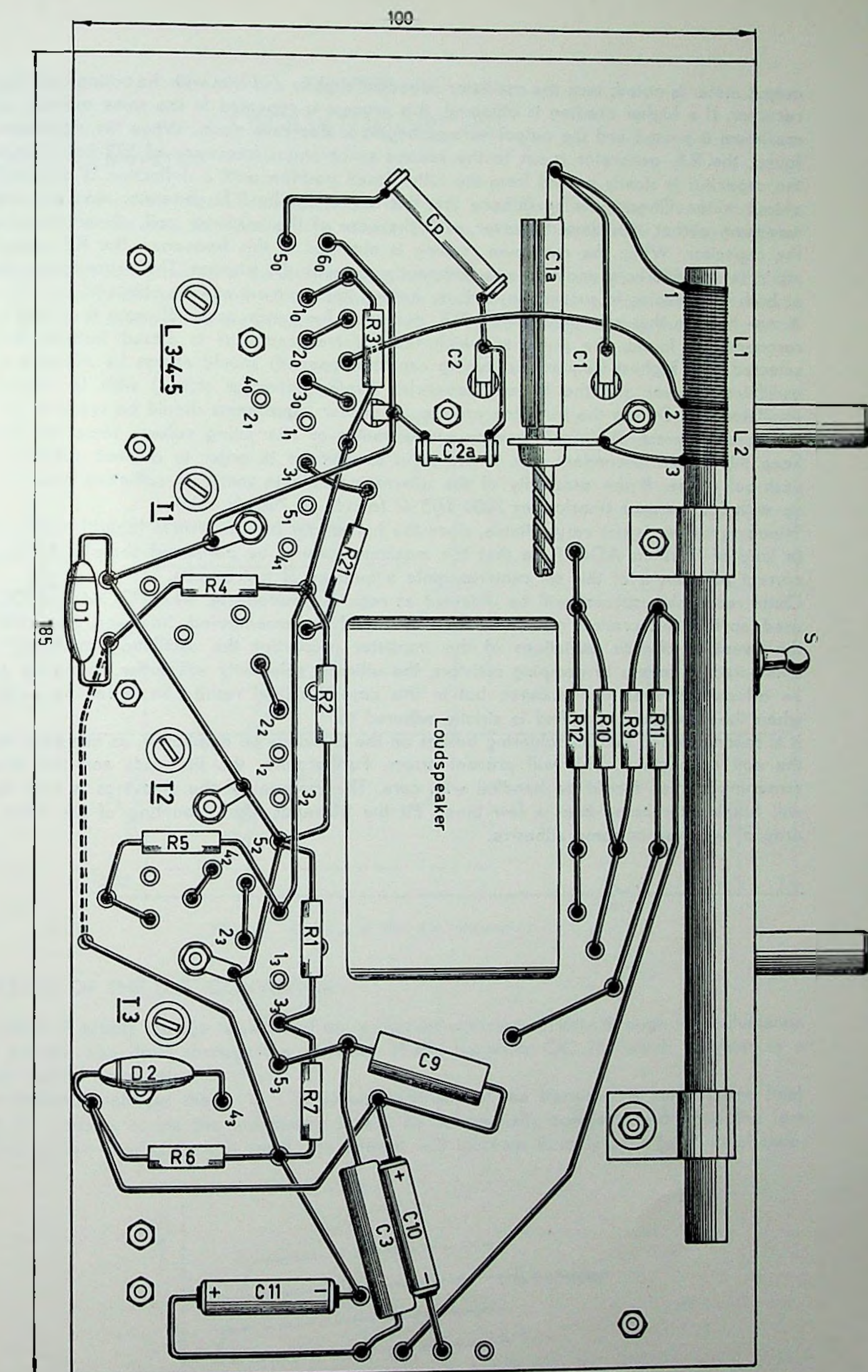

100
185
Loudspeaker

Fig. 96

output meter is noted; turn the oscillator capacitor slightly and trim with the tuning capacitor of the receiver. If a higher reading is obtained, this process is repeated in the same manner, until the maximum is passed and the output voltage begins to decrease again. When the maximum is thus found, the R.F. generator is set to the second synchronous frequency of 572 kc/s. Once again the capacitor is slowly moved from the fully closed position until a deflection is obtained in the output meter. Check with headphone that this is actually the R.F. generator tone and adjust for maximum output, this time, however, with the core of the oscillator coil, always trimming with the capacitor. When the maximum reading is obtained at this frequency, the R.F. generator is again set to 1535 kc/s, and further adjustment made with the trimmer. This procedure is repeated at both frequencies in succession until no further improvement can be obtained.
It may happen that harmonics of the R.F. generator frequency occur. If more than one point of resonance is found, the point at which the receiver capacitor is closed furthest should be selected. The highest frequencies (tuning capacitor opened) should always be adjusted with the oscillator trimmer and the lower frequencies (tuning capacitor closed) with the core of the oscillator coil. During the trimming process no further adjustments should be made to the aerial circuit already set. Preferably use an output meter or alternating voltage meter for trimming; keep this on the secondary side of the input transformer in order to prevent instability of the push-pull stage. If the sensitivity of the alternating voltage meter is sufficient, then transform up with loudspeaker transformer 7000 Ω/5 Ω (meter on 7000 Ω side).
Trimming by ear is not very reliable, since the human ear has an inverse logarithmical sensitivity (a kind of "built-in AGC"!) so that the maximum cannot be perceived sharply. All in all the correct adjustment of this set requires quite a number of instruments.
Quite reasonable success will be obtained as regards neutralizing, when OC44 and OC45 are used, so that this process can be omitted. It should be remembered, however, that although in the event of chance deviations of the transistor properties the oscillating tendency can be eliminated by means of damping resistors, the ultimate selectivity will suffer. Trimming can also be effected in a simpler manner, but in this case the final result can never be as good as when the procedure specified is strictly adhered to.
It is recommended that all soldering bushes on the drawings be numbered, as has been done for the coil connections. This will prevent errors. Furthermore, the litz ends and the aluminium screening bushes should be handled with care. The material of the sleevings is very thin and will break after being bent a few times. Fit the litz ends after mounting of the coils with a drop of beeswax or some adhesive.

CHAPTER VIII

OSCILLATOR CIRCUITS

VIII. 1. D.C. CONVERTOR

Fig. 97 shows the circuit of the d.c. convertor which is used for converting the 6 V supply voltage to 45 V required for feeding the anode circuits of the tubes in this tube-transistor receiver.

The basic part of the circuit is a transformer-coupled pulse oscillator using a transistor OC 76. Energy is stored in the inductance of the transformer during the "on" period of the transistor, and this energy is delivered to the output circuit at an increased voltage during the "off" period.

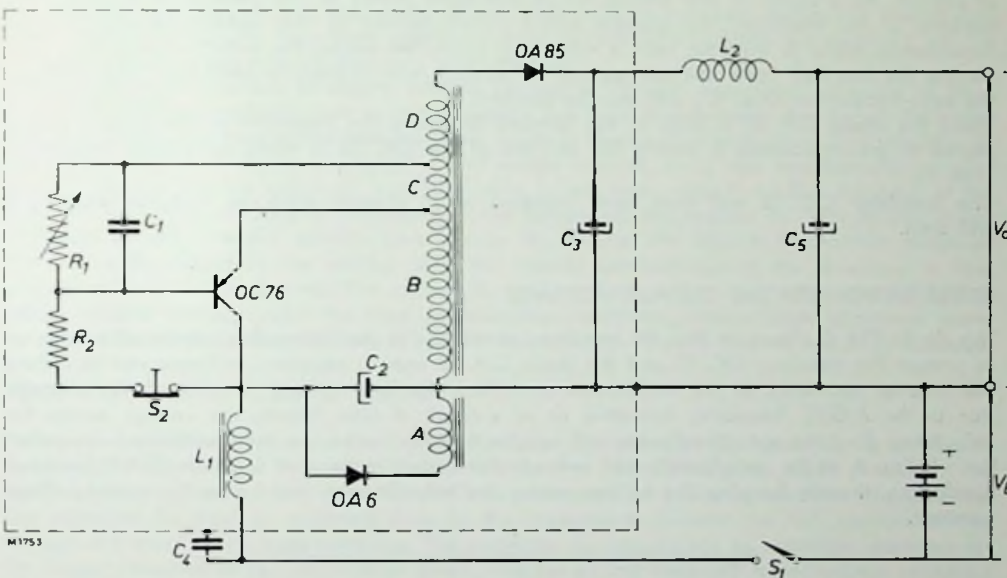


Fig. 97. Circuit of the d.c. convertor.

OPERATION OF THE D.C. CONVERTOR

When the 6 V supply voltage is switched on, a current will start to flow through the inductance B; this current also flows through the collector of the transistor OC 76, which operates in a common emitter connection.

At low collector voltages the $I_c = f(V_{ce})$ characteristic of the transistor is steep (knee line) so that the voltage across the inductance B will be substantially constant, and since the loss resistance of this conductance is small, the current will increase linearly as a function of time.

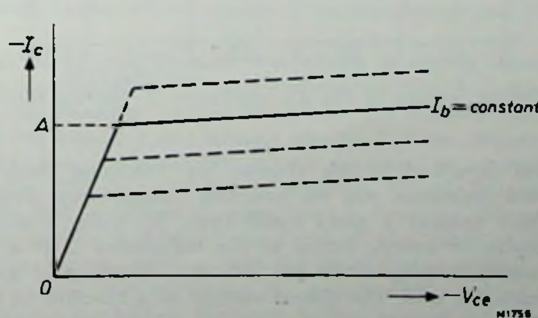


Fig. 98.

A constant voltage will thus be produced across the coil C of the auto-transformer, which is so wound that the base of the transistor becomes negative with respect to the emitter. An almost constant base current will thus flow into the base. The $I_c = f(V_c')$ characteristics of the transistor shown in Fig. 98 reveal that two regions can be distinguished, namely the region below point A having a very high slope and the region beyond point A having a very low slope. It will thus be clear that at the constant base current mentioned above, the collector current will assume a determined value (point A, Fig. 98) that cannot increase any further. When the collector current has reached this determined value the collector current, and hence the current flowing through the inductance B, can therefore not increase any further, which means that the voltages across the inductances B and C will start to decrease. This causes the base current and hence also the collector current to decrease. The polarity of the voltages across B and C is thus reversed because the sign of L di/dt is also reversed, and the transistor is cut off. As a result of the energy which has been stored in the inductance B during the on-time of the transistor, the parallel tuned circuit formed by this inductance and its stray capacitance starts to oscillate, and a voltage is produced across the inductance.
During the first half cycle this voltage will be positive with respect to earth, so that the top of the auto-transformer (Fig. 97) will also be positive.
Since the diode OA 85 is then in the forward direction, the magnetic energy that has been stored in the inductance B during the on-time of the OC 76 is taken up by the load capacitor C_3.
The transistor OC 76 will then have returned to its original state, so that the next cycle will start.

STABILISATION OF THE OUTPUT VOLTAGE

The diode OA 6 shown in Fig. 97 has been provided for stabilising the output voltage so as to protect the transistor OC 76 and the diode OA 85 against excessive voltages and to reduce the internal resistance of the d.c. convertor. When the load current is reduced, for example due to the A.G.C. becoming operative or as a result of tube failure, the voltage across the inductance A of the auto-transformer will assume a higher value. As soon as the voltage across the winding A of the auto-transformer exceeds the battery voltage of 6 V, the OA 6 becomes conducting, thereby keeping the voltage across the inductance A and hence the output voltage constant.

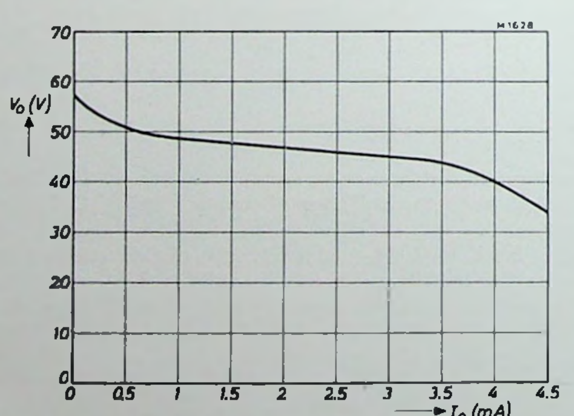


Fig. 99. Output voltage of the d.c. convertor at varying load current.

In the experimental set described in this chapter the stabilising diode is the gold bonded type OA 6. The forward resistance and the resulting voltage drop across this type of diode is extremely small, which ensures a good stabilisation. By using a diode for stabilising the output voltage, the surplus of energy stored in the inductance B at a small load current is returned to the 6 V supply battery, which prevents the efficiency of the d.c. convertor from becoming low at a decreasing load. The output voltage as a function of the output current has been plotted in Fig. 99.

STARTING OF THE D.C. CONVERTOR

When the battery circuit of the d.c. convertor shown in Fig. 97 is closed, the base-emitter voltage V_b' of the transistor will initially be zero. Since the mutual conductance of the transistor is proportional to the emitter current I_e' and $I_e = 0$ at $V_{be} = 0$, the mutual conductance will then also be zero, so that the d.c. convertor will not start to oscillate.
Starting may be initiated by the disturbance which is produced by the transient occurring when the battery switch is closed; a voltage pulse is then induced by which the base of the transistor is temporarily rendered negative, and the mutual conductance assumes a finite value. Starting is, however, hampered because the system is heavily damped initially by the output circuit. This is due to the fact that the load capacitor is uncharged when the d.c. convertor has been switched off, so that the output circuit is shunted across the H.T. winding of the transformer with only the relatively small forward resistance of the H.T. diode in series. It is true that once the d.c. convertor has started, the anode of the diode will be negative during the critical part of the pulses, so that its high back resistance then considerably reduces the damping on the H.T. winding; this, however, is obviously cold comfort.
There are several solutions to ensure reliable starting of the d.c. convertor, but most of them have their specific drawback, such as loss of energy and radio interference. A detailed discussion of a great number of starting methods has been given in Electronic Applications (Vol. 16, No. 2). The circuit of Fig. 97 shows an attractive and simple method, using two mechanically coupled switches S_1 and S_2. The switch S_2 is opened after S_1 has been closed, so that the base of the transistor OC 76 is temporarily connected to the battery via the resistor R_2. This resistor forms a voltage divider with the variable base resistor R_1, so that the base is temporarily rendered negative with respect to the emitter, and the mutual conductance of the transistor is thus prevented from becoming zero. The resistor R_2 can be given a very low value, say 2.7 kΩ, to ensure reliable starting under the most unfavourable conditions, without high additional losses being introduced, since the switch S_2 is opened immediately after the battery switch S_1 is closed.

DECOUPLING OF THE SUPPLY BATTERY

To suppress A.F. interference via the battery, the latter is shunted by the capacitor C_2 of 100 μF. To decouple the battery lead for H.F. interference the filter $L_1 - C_4$ has been provided. The choke L_1 is wound on a ferroxcube 4B core with a diameter of 6 mm to minimize its resistance. The capacitor C_4 must be mounted close to the grommet to prevent the H.F. current flowing through this supply lead from radiating. The capacitor C_2 also serves as a storage capacitor for the energy returned by the stabilisation diode. The use of this capacitor is particularly necessary when the apparatus is fed from a dry battery, because this has a fairly high internal resistance and the recharging of dry batteries takes place at a very poor efficiency.
To prevent magnetic radiation, the entire d.c. convertor is placed in an iron box, which should, moreover, be provided with a copper coating.
The operating frequency of the d.c. convertor is approximately 5.5 kc/s; the duty cycle of the transistor OC 76 is 0.75. The overall efficiency of the convertor is approximately 80 %.

VIII. 2 PUSH-PULL D.C. CONVERTOR

This unit, shown in fig. 100, can be used to supply anode voltage in hybrid (tube and transistor) receivers, and is in effect an "inductive mutivibrator" or square-wave oscillator, since the time-constant is governed partly by the self-inductance of the transformer and partly by the capacitors employed.
The converter operates as follows: Because the circuit is slightly asymmetrical, when S_1 is closed one of the two transistors produces more collector current (I_{co}') than the other. Assuming that this applies to the top transistor, current flows from the positive pole of the battery to the emitter-collector of the top transistor and through winding I of the negative pole of the battery. In accordance with Lenz's law, a counter-EMF is then generated in winding I, which opposes the applied voltage. A similar EMF is generated in windings III and IV, which are on the same core. The polarity of this EMF is such as to make the top transistor still more conductive and at the same time drive the bottom transistor beyond cut-off (the starts of the windings, indicated by arrows in the diagram, become positive, and the other ends negative). The result is that the current in winding I increases, thus increasing the counter-EMF, making the top transistor still more conductive and driving the bottom transistor further beyond cut-off.

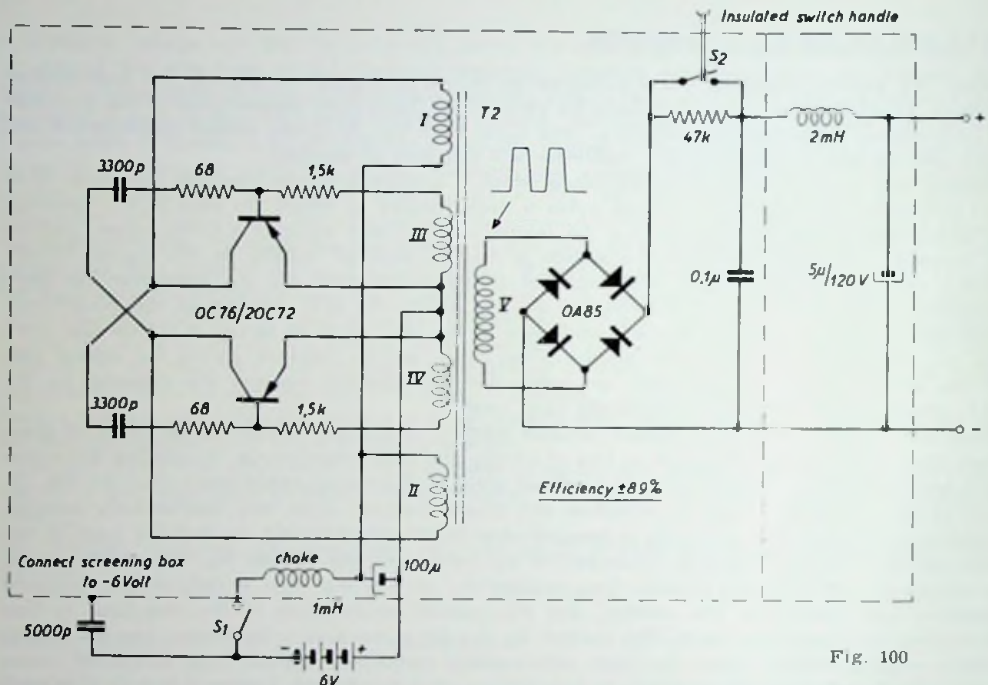


Fig. 100

Now, the transformer has a certain self-inductance L, which, combined with the circuit resistance R (i.e. resistance of winding + impedance of transistor), produces a time constant:

$\tau = \frac{L}{R}$. Hence the current increases as a power of e.

This continues until the transformer core (ferroxcube or mu-metal) reaches magnetic saturation. For all practical purposes the current is then governed only by D.C. resistance, since magnetic saturation causes the self-inductance to fall almost to zero. Also, the cessation of flux-variation in the core cuts off the counter-EMF abruptly, thus driving the top transistor beyond cut-off. As the current is then interrupted, the field in the core dies away, but, in so doing, generates another EMF in the opposite direction. The result is that the bottom transistor suddenly becomes conductive, sending current through winding II to magnetize the core in the opposite direction. Again a counter-EMF is induced in winding II, and also in windings III and IV, by reason of which the top transistor is driven beyond cut-off, and the bottom transistor becomes still more conductive, and so on. The frequency of the two cycles can be computed, approximately, as follows:

$$V = nL \frac{di}{dt}$$

$$V\,dt = nLdi$$

$$\int_0^t V\,dt = nL \int_{-i_{max}}^{+i_{max}} di$$

$$\int_0^{\frac{1}{2f}} V\,dt = nL \int_{-\Phi_{max}}^{+\Phi_{max}} d\Phi$$

$$V \frac{1}{2f} = 2\,nL\,\Phi_{max}$$

$$f = \frac{V}{5\,nL\,\Phi_{max}}$$

showing that the frequency is proportional to the battery voltage and inversely proportional to the number of turns, the self-induction (depending on the load) and the maximum flux Φ_{max}. This last factor depends on the material and cross-sectional area of the core.

To keep the frequency within reasonable limits, and thus to avoid absurdly heavy losses, it is necessary to employ many turns and to ensure that the core material combines suitable cross-sectional area with high B_{max}. Although ferroxcube is the most suitable material, mu-metal is satisfactory as an alternative. At any rate its hysteresis loop is small, which is the essential requirement for minimizing hysteresis losses. On the other hand, eddy currents are heavier in mu-metal, so that the frequency should be kept as low as possible. It is also important to use relatively thick wire for windings I and II; 0.4—0.6 mm enamelled wire is suitable. With proper dimensioning, the efficiency may be as much as 80 %. To facilitate starting the converter, a resistor in series with the smoothing capacitance is provided. With the converter switches on and the capacitors charged, the resistor can be shorted. The converter should be suitably screened to prevent stray radiation. S_2 is operated by a rod of insulating material. Winding ratio $\frac{\mathrm{I}}{\mathrm{III}} = \frac{\mathrm{II}}{\mathrm{IV}}$ is 2. Numbers of turns depend on the particular core; on a core of 1 sq. cm cross-sectional area, I = II has one hundred, and III = IV fifty turns. Winding V contains roughly 200 turns of 0.2 mm enamelled wire.

VIII. 3. A D.C. CONVERTOR FOR 500 V OUTPUT POTENTIAL

The following is a description of the circuit of a d.c. convertor which generates a high d.c. voltage at relatively large power, viz. 500 V at 25 mW from a 3 V battery. The efficiency is 52 %. Convertors of this type are frequently used in portable apparatus.

Diagram 101 shows the complete circuit with the corresponding component values. Resistor R_2 (of 18 kΩ) ensures "safe" starting of the convertor.

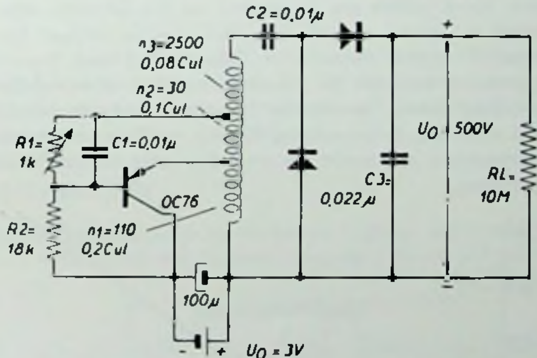


Fig. 101

With a d.c. convertor type, care must be taken that the load is not removed, as otherwise the voltages across the transistor assume excessively high values. Rectification is effected by means of a voltage doubler circuit containing two selenium rectifiers.

The peak value of the collector current is 50 mA and that of the collector voltage $-V_{cb}$ about 28 V. The interruption frequency is about 2 kc/s, and at that frequency the filtering is easy. Without additional filtering components a hum voltage of 3 V was produced in the 500 V output potential.

The peak collector current may be regulated by means of the resistor R_1 connected in series with the base. The smaller the value of the resistor, the larger will be the collector current and also the output voltage. An increase of R_1, however, slightly reduces the efficiency of the apparatus. Reliable functioning of the convertor does not call for a critical adjustment of the battery voltage. Even with a battery voltage of 2 V it is still possible to maintain the output at 450 V by adjusting R1.

No difficulties are presented in adapting the component values for higher voltages, and, of course, a tripler circuit may also be used in the secondary circuit.

VIII. 4. A SIMPLE REVOLUTION COUNTER FOR PETROL ENGINES

In the foregoing section an indicator for low frequencies has been described. Using the same principle, it is also possible to construct revolution counters for industrial machines. This is particularly simple when electrical pulses are already at hand, as is the case with the ignition

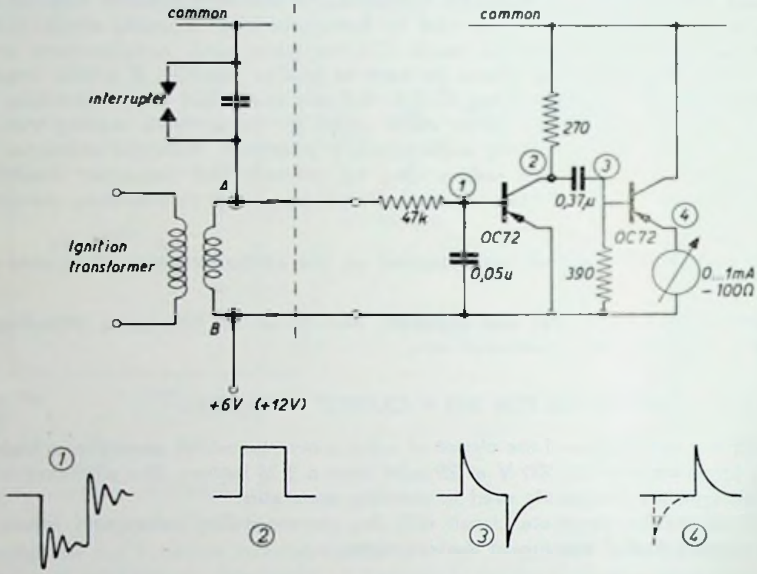


Fig. 102

circuit of a petrol engine. Here pulses are generated on the primary side of the ignition transformer each time the spark is passed, and these pulses can be utilized for a frequency reading. Depending upon the type of engine (two-stroke, four-stroke) and the number of cylinders, a particular ignition sequence corresponds to a certain number of revolutions per minute.
With the instrument described below, an attempt has been made to combine optimum compactness in design with good accuracy in measuring the r.p.m.. This counter may be used for other machines, provided arrangements are made for an interrupter in combination with a choke coil to supply the necessary pulses.

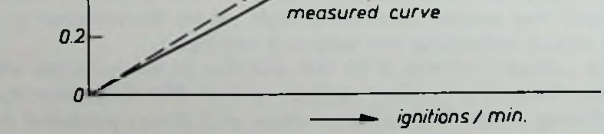


Fig. 103

In fig. 102 the ignition transformer with the interrupter is shown. On the right the circuit of the revolution counter is reproduced. From points A and B the primary voltage of the ignition transformer is passed to the input of the instrument. The signal is filtered and is used to drive a transistor OC72. In the collector circuit a square-wave voltage is generated, which is differentiated and applied to the base of the second transistor. This latter functions as a switch and in such a way that it is actuated only by negative pulses, since the positive pulses occur in the cut-off range of the transistor. The reading is thus independent of the duration of the square-wave pulses. The negative pulses set up current pulses in the emitter circuit.

If the time constant is sufficiently small and the instrument is connected via an integrating network, a reading is obtained which is proportional to frequency and hence to the r.p.m.
The voltages arising at the individual points are indicated in diagram 102 for the sake of clarity. Depending upon the choice of the RC member and the magnitude of the resistance of the instrument, the instrument will indicate the actual r.p.m. of the engine.
The counter can operate from a 6 or 12 V battery. The circuit has been designed as compactly as possible. Compared with a mechanical revolution counter, this instrument offers numerous advantages, viz. only two connecting wires are required, and as a rule the negative pole may be connected direct to the earth of the engine.

VIII. 5. BLOCKING OSCILLATOR

This is a one-shot multivibrator operating on the same principle as the previously described circuits. The circuit is shown in fig. 104. Unlike the previous units, however, it produces only one square pulse. The blocking oscillator is employed as a pulse regenerator (or shaper), say, to control flip-flop circuits, as described later. It is sensitive enough for, say, spark-discharge operation with an aerial only a few metres long. This would enable the oscillator to be employed in combination with a flip-flop circuit and a relay controlled thereby through an amplifier, as a radio-control unit (e.g. for models). See VIII. 12.

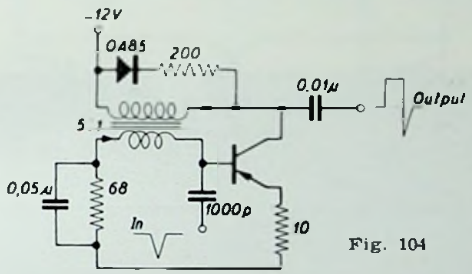


Fig. 104

The emitter resistance must not be decoupled, or the oscillator will become, in effect, a multivibrator controlled by the leakage current I_{CO}'. As the frequency may be very high, excessive dissipation occurs, with the result that the transistor soon breaks down. Hence it is better to connect a voltage divider to the emitter (see VIII. 12).

VIII. 6. 1 KC/S-MULTIVIBRATOR

Fig. 105 is the circuit of a transistorized multivibrator, whose signal can be taken from one of the collectors, or from a tapping on the collector resistor. The 5 kΩ resistor provides a certain

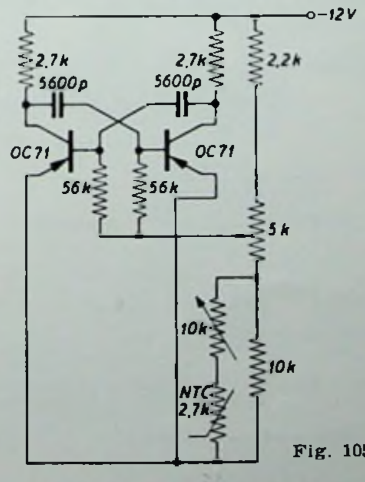


Fig. 105

amount of frequency control. Temperature compensation can be varied by means of the 10 kΩ resistor so as to minimize frequency drift owing to variations in temperature.

VIII. 7. 1 KC/S RC GENERATOR

A sinusoidal signal is obtained by means of a suitable combination of positive and negative feedback (fig. 106). As in the preceding circuit, temperature compensation and frequency are both adjustable.

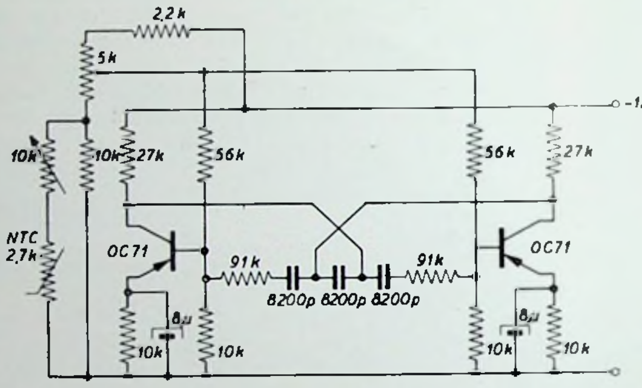


Fig. 106

VIII. 8. COLPITTS OSCILLATOR

Transistor in grounded-base circuit. As the input and output signals are in phase, positive feedback is obtained by returning the collector signal to the emitter (fig. 107). This is connected to a "capacitive branch" of the circuit. Ratio C_2/C_1 should be 10 : 1. Output signal can be taken from the emitter, or from a coupling coil on L.

Max. frequency is roughly 250 kc/s with an average, and 500 kc/s with an above-average, OC71, enabling a simple I.F. trimming oscillator to be constructed with L as the coil of an I.F. aerial filter (series capacitor shown in the diagram should then be removed or short-circuited).

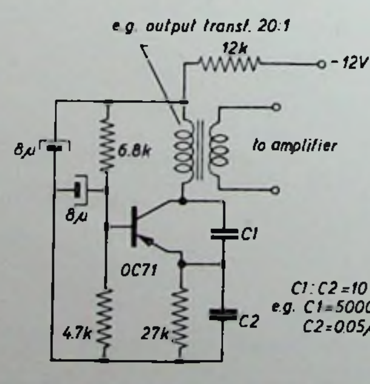


Fig. 107

VIII. 9. SOUNDER UNIT

Very simple morse-sounder comprising magnetic headphones and a transistor. Actually it comprises the above-mentioned Colpitts circuit with the self-inductance of the headphone coils taking the place of L.
With 2000 Ω headphones, the frequency is about 1000 c/s (fig. 108).

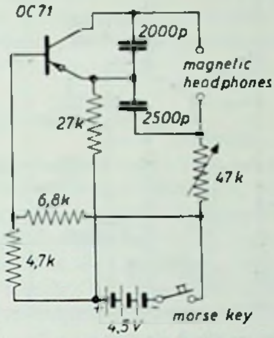


Fig. 108

VIII. 10. MODULATED LONG-WAVE AUXILIARY OSCILLATOR FOR SETS WITHOUT PICK-UP SOCKET

With the Colpitts oscillator modulated by means of a pick-up signal, any receiver can be readily converted to operate with a pick-up (see fig. 109). The 2 mH short-wave choke provides the necessary self-inductance.

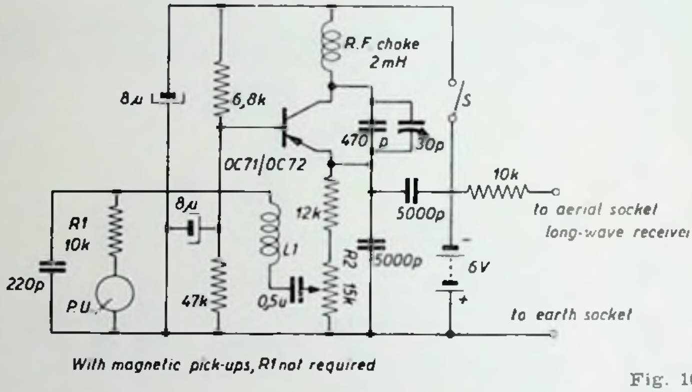


Fig. 109

The frequency can be varied slightly by means of a 30 pF trimmer so as to avoid possible heterodyning with long-wave transmitters. Receiver should be tuned to roughly 160 kc/s (about 1875 metres). Disconnect ordinary aerial and plug oscillator output into aerial socket.
Given a low-resistance magnetic pick-up with low output voltage, R_1 can be removed or shorted.
Percentage modulation is adjusted semi-permanently by means of R_2.

VIII. 11. FLIP-FLOP CIRCUIT

Transistors can be employed in very much the same way as double triodes in circuits with two stable settings. The base potential of the one transistor is then independent of the collector potential of the other, so that the one opens and the other closes whenever the setting changes. On-off switching is more positive than with tubes. With a saturation flip-flop as indicated in

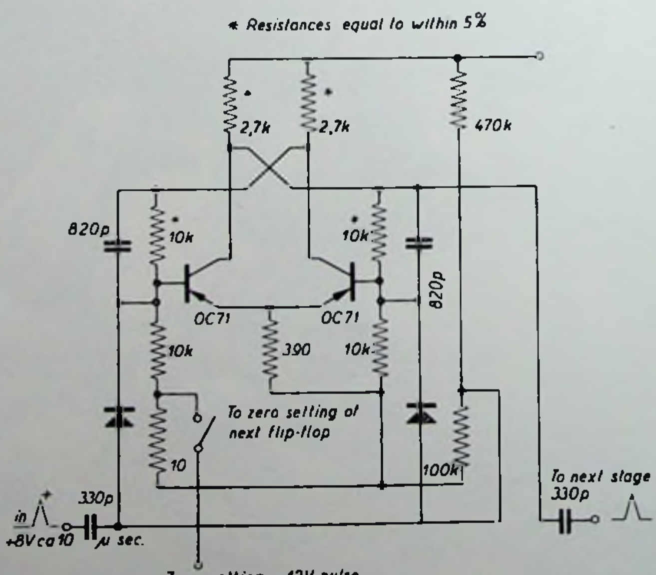


Fig. 110

fig. 110, the difference between the collector and emitter voltages is only some tenths of a volt. Hence the flip-flop makes a good switch or relay. One of the possible uses is described in VIII. 12. On the other hand, owing to hole-storage, the response is somewhat slower than in flip-flops with electron tubes (about 4 μS as compared with 1 μS). The difference can be reduced by cutting the load on the transistor, say, to the normal working point; in the circuit of fig. 110 this means replacing the 10 kΩ resistors in the lower part of the circuit with 1 kΩ resistors. At the same time, this also reduces the swing of the collector potential. Several flip-flops can be arranged to form a simple frequency divider, each stage halving the frequency of the preceding stage; the first stage can be controlled, say, by a multivibrator and pulse-former.

VIII. 12. RADIO-CONTROL UNIT

Fig. 111 is the circuit diagram of the radio-control unit referred to in section VIII. 5 with an induction coil, producing a spark when switched on or off, as the "transmitter".
To avoid possible swamping of the flip-flop by pulses arriving one on top of the other (oscillation), a blocking oscillator with a suitable time constant (wide pulses) is connected in series with it. The necessary time constant is obtained by combining high transformer self-inductance with low D.C. resistance, or in other words by winding 1000 turns of 0.15 mm enamelled wire on the mu-metal core of a discarded microphone transformer. The secondary should contain 1/5 as many turns, i.e. 200. The control unit operates as follows: when the induction coil is switched on or off, the transmitter provides a "click-pulse". The receiving aerial — a whip roughly ½ metre long — passes this pulse to transformer TR, where it is stepped down to match the low base impedance of V_1.

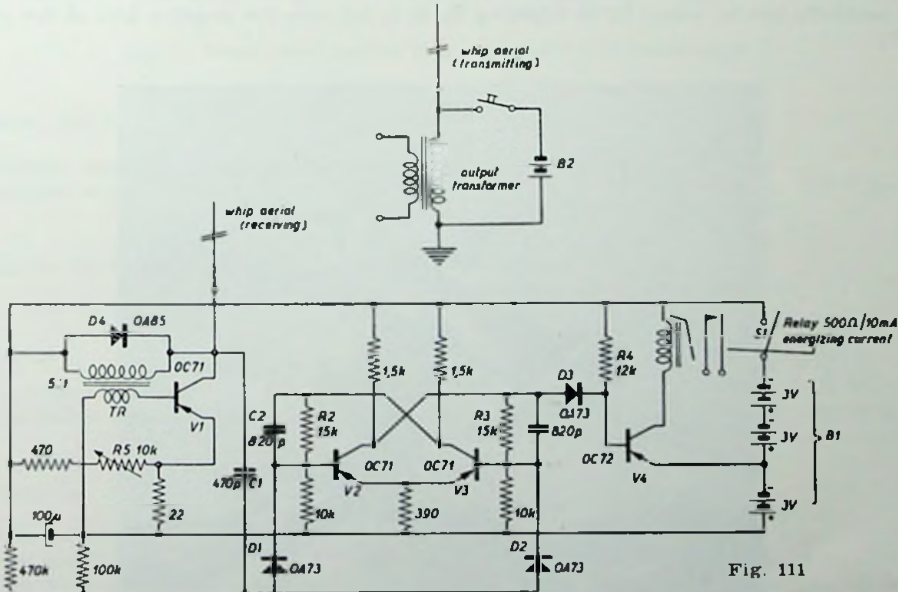
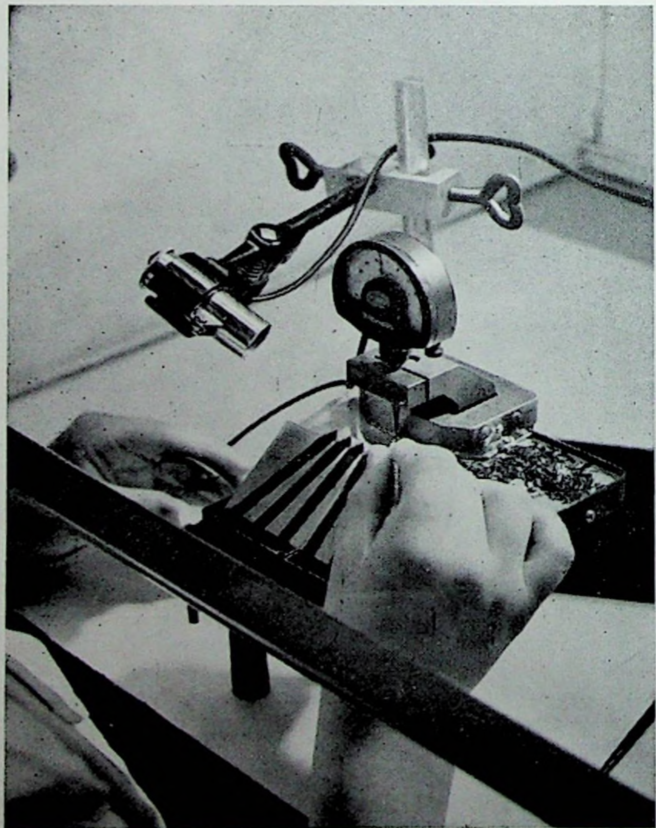


Fig. 111

As the field variation in the transformer builds up, the transistor, only slightly conductive at first, is driven fully conductive (avalanche effect), and remains so until the field in the core reaches saturation point, or until the current reaches the maximum limit imposed by the transformer resistance (in the latter case it may happen that the time interval is too short; hence the resistance should be low).
As we have seen, as soon as the field in the core ceases to vary, the induced EMF disappears from the base winding and the transistor becomes non-conductive, thus cutting off the transformer current. With the disappearance of this current, the field in the core dies away, but in so doing induces another EMF in the coil, in the opposite direction. This gives the base positive bias, thus driving the transistor beyond cut-off. In this way, the indefinite signal in the aerial is converted into a true square-wave voltage.

At the same time, the blocking oscillator does not remain open long enough to operate the relay. This is where the electronic switch, that is the saturation flip-flop, comes in.
Let us suppose that V_2 is open and V_3 closed. In saturation flip-flops this means that the collector of V_2 is for all practical purposes at earth potential; hence the anode of D_3 is positive with respect to its cathode.
Because the diode is conductive, its cathode is also at earth potential. The emitter potential of V_4 is —3 V; hence it is non-conductive and the relay is open. When the square pulse arrives from the oscillator it is differentiated, that is narrowed, by means of C_1.
The positive transient passes to the anodes of D_1 and D_2, making them conductive. V_2, originally conductive, is then driven beyond cut-off and V_3, already non-conductive, remains so. In becoming non-conductive, however, V_2 produces in its base a negative transient, which passes through C_3/R_3 to the base of V_3. With the termination of the positive pulse in D_1/D_2, V_3 conducts again. Thus the flip-flop changes over.
The anode of diode D_3 then becomes negative, so that this diode is non-conductive. R_3 then imparts negative bias (with respect to the emitter) to the base of V_4, enabling this to become fully conductive and, since V_4 functions as a switch, reducing the potential difference between collector and emitter almost to zero. Negative emitter bias to cut off V_4 completely in the steady state is obtained from a tapping on the 9-V battery (three 3-V dry cells). Without this bias, dissipation in the (semi) cut-off state would be sufficient to damage the transistor. Successive pulses from the "transmitter" drive the flip-flop to and fro, thereby opening and closing the relay.
The distance range of the unit is in the region of 10 metres, but can be extended by employing a longer, matched, aerial. It should be borne in mind that the circuit also responds to interference from other sources than the transmitter (thunder, switch-clicks, car-ignitions and so on).
The sensitivity can be varied by so adjusting R_5 as to increase the negative bias of the emitter of V_1.



CHAPTER IX

TRANSISTOR DATA

ALL-GLASS GERMANIUM TRANSISTOR OC 70

The OC 70 is a junction transistor of the P-N-P type in all-glass construction. It is particularly recommended for medium gain low-power audio frequency applications and especially designed for use in the first stage of hearing-aid circuits. Moreover the OC 70 can be used as a general-purpose amplifier at frequencies up to ∾ 0.3 Mc/s. Further applications are in switching and oscillator circuits where large signals are involved.
The all-glass construction ensures absolute moisture resistance and long life. The transistor is shock-proof and insensitive to ambient illumination. It may be soldered in or clipped for plug-in connection.

MECHANICAL DATA

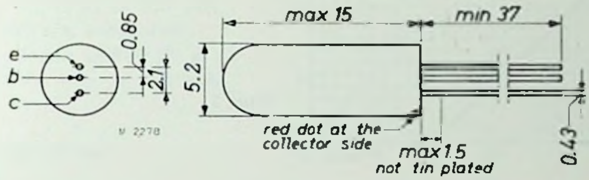


Fig. 1. Dimensional outline in mm and electrode connections.

THERMAL DATA

Junction temperature rise to ambient temperature, transistor in free air	K	=	max.	0.4 °C/mW

ABSOLUTE MAXIMUM RATINGS

Junction temperature

storage	T_s	=	min. / max.	−55 °C / +75 °C
continuous operation	T_j	=	max.	75 °C
intermittent operation, total duration max. 200 h	T_j	=	max.	90 °C [1]

Collector

Voltage (emitter reference)			
Peak and D.C.	see Fig. 16.		
Peak, base at least 0.1 V positive	$-V_{CEM}$	=	max. 30 V
Average, base at least 0.1 V positive	$-V_{CE}$	=	max. 30 V
Current			
Peak	$-I_{CM}$	=	max. 50 mA
Average	$-I_C$	=	max. 10 mA
Dissipation (in free air)	see figure 2.		

[1] Likelihood of full performance of a circuit at this temperature is also dependent on the type of application.

Emitter

Current

Peak	I_{EM}	= max. 55 mA
Average	I_E	= max. 12 mA

Base

Current

Peak	$-I_{BM}$	= max. 5 mA
Average	$-I_B$	= max. 2 mA

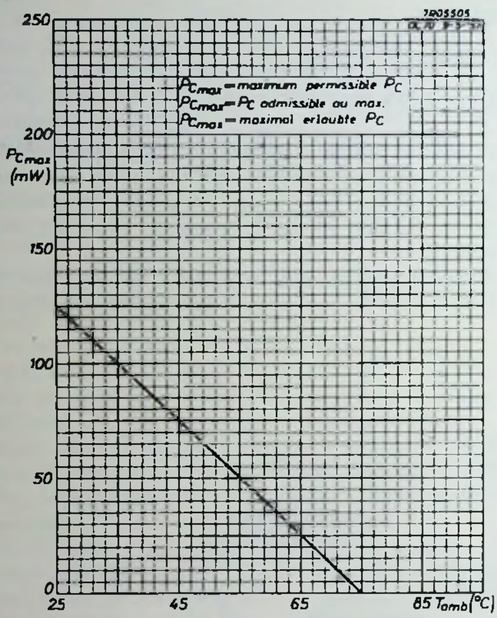


Fig. 2

EQUIVALENT CIRCUIT PARAMETERS AT AN AMBIENT TEMPERATURE OF 25 °C

Common base circuit

at $-V_{CB}$ = 2 V
$-I_C$ = 0.5 mA
f = 1000 c/s

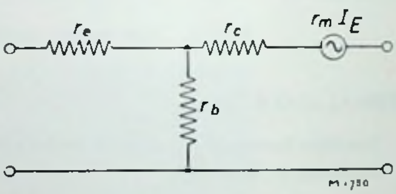


Fig. 3

Emitter resistance	r_e	= 39 Ω
Base resistance	r_b	= 1000 Ω
Collector resistance	r_c	= 1.43 MΩ
Transfer resistance	r_m	= 1.38 MΩ

		min. [1]	average	max. [1]
Input impedance (output short-circuited)	h_{11b} =	58	71	88 Ω
Current ampl. factor (output short-circuited)	$-h_{21b}$ =	0.952	0.968	0.976
Output admittance (input open)	h_{22b} =		0.7	1.3 μA/V
Voltage feedback ratio (input open)	h_{12b} =		7.10^{-4}	

Common emitter circuit

at $-V_{CE}$ = 2 V
$-I_C$ = 0.5 mA
f = 1000 c/s

		min. [1]	average	max. [1]
Input impedance (output short-circuited)	h_{11e} =	1.2	2.2	3.6 kΩ
Current ampl. factor (output short-circuited)	h_{21e} =	20	30	40
Output admittance (input open)	h_{22e} =		23	53 μA/V
Voltage feedback ratio (input open)	h_{12e} =		9.10^{-4}	27.10^{-4}

[1] Typical production spread.

CHARACTERISTICS AT AN AMBIENT TEMPERATURE OF 25 °C

Common base circuit

			min.	average	max.	
Collector current at						
$-V_{CB}$ = 4.5 V; I_E = 0	$-I_{CBO}$	=		5	12	μA

Common emitter circuit

			min.	average	max.	
Collector current at						
$-V_{CE}$ = 4.5 V; I_B = 0	$-I_{CEO}$	=		110	225	μA
$-V_{CE}$ = 4.5 V; $-I_B$ = 10 μA	$-I_C$	=	0.21	0.4	0.65	mA
$-V_{CE}$ = 4.5 V; $-I_B$ = 250 μA	$-I_C$	=	4.6	10	13.2	mA
Base voltage at						
$-V_{CE}$ = 4.5 V; $-I_B$ = 10 μA	$-V_{BE}$	=	75	110	150	mV
$-V_{CE}$ = 4.5 V; $-I_B$ = 250 μA	$-V_{BE}$	=	200	275	385	mV
Current amplification cut-off frequency at						
$-V_{CE}$ = 2 V; $-I_C$ = 0.5 mA	f_{α_e}	=		15		kc/s
Noise factor at 1000 c/s, measured with a source impedance of 500 Ω, at						
$-V_{CE}$ = 2 V; $-I_C$ = 0.5 mA	F	=		10	15 [1]	dB

[1] Typical production spread.

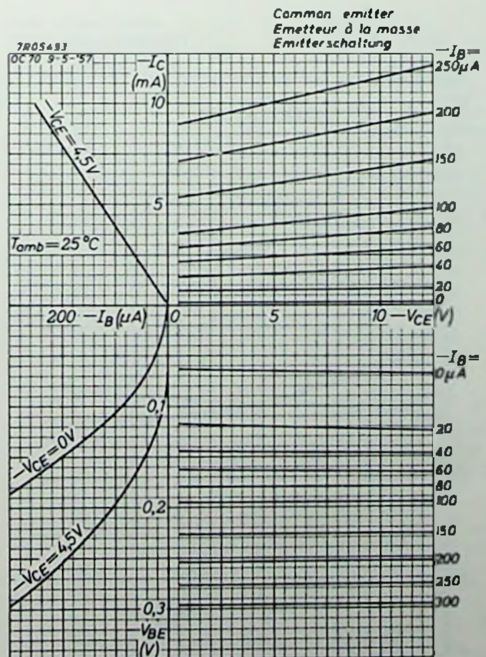


Fig. 4. Typical characteristics in common emitter connection at an ambient temperature of 25 °C.

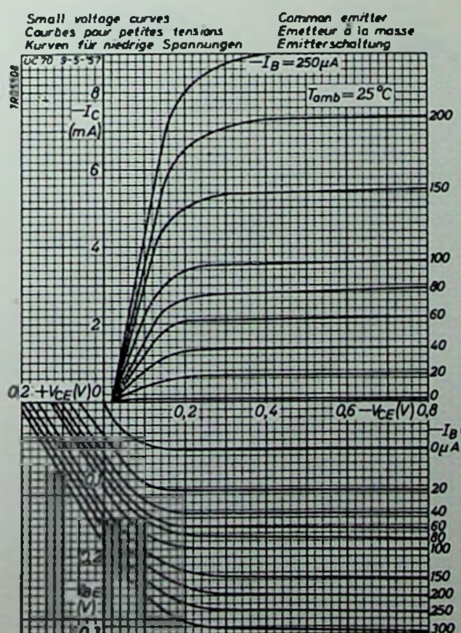


Fig. 5. Typical low-voltage characteristics in common emitter connection at an ambient temperature of 25 °C.

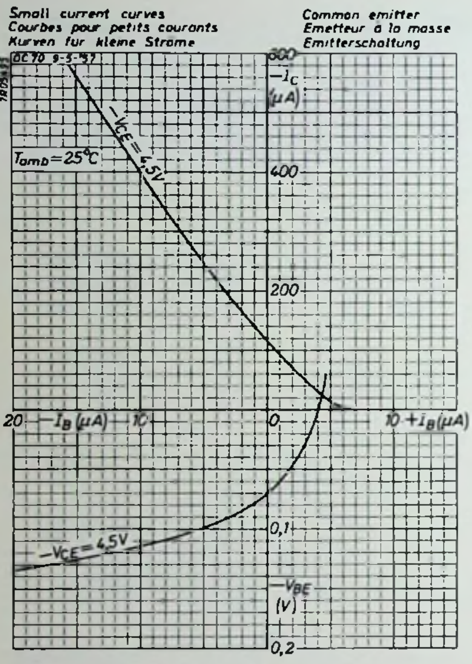


Fig. 6. Typical low-current characteristics in common emitter connection at an ambient temperature of 25 °C.

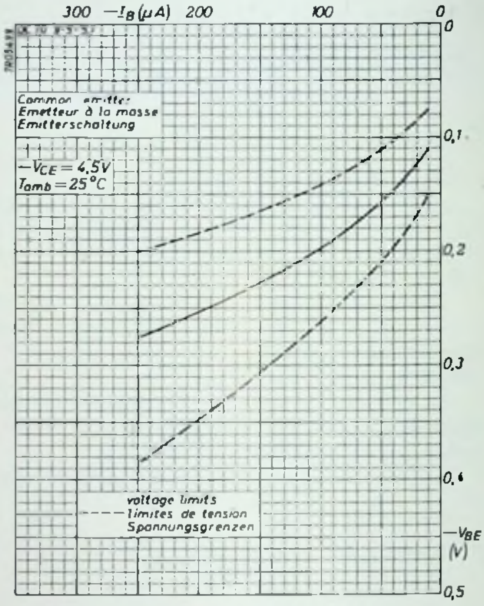


Fig. 7. Input characteristics in common emitter connection at an ambient temperature of 25 °C.

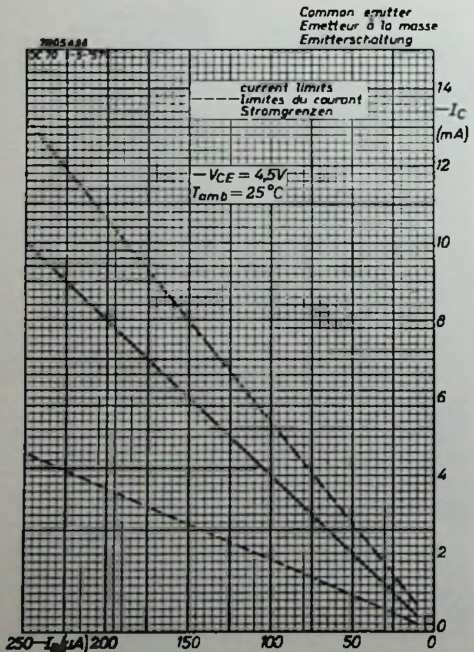


Fig. 8. Current amplification characteristics in common emitter connection at an ambient temperature of 25 °C.

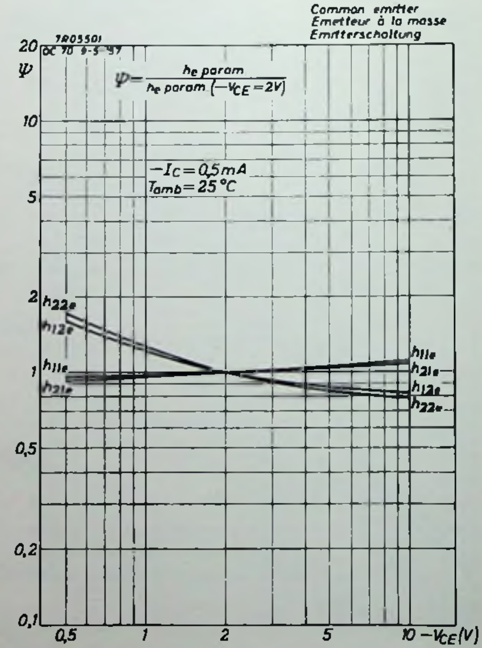


Fig. 9. h-parameters versus collector voltage in common emitter connection at an ambient temperature of 25 °C.

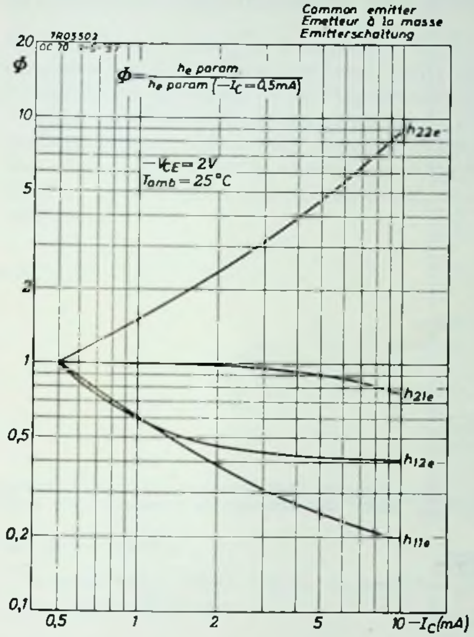


Fig. 10. h-parameters versus collector current in common emitter connection at an ambient temperature of 25 °C.

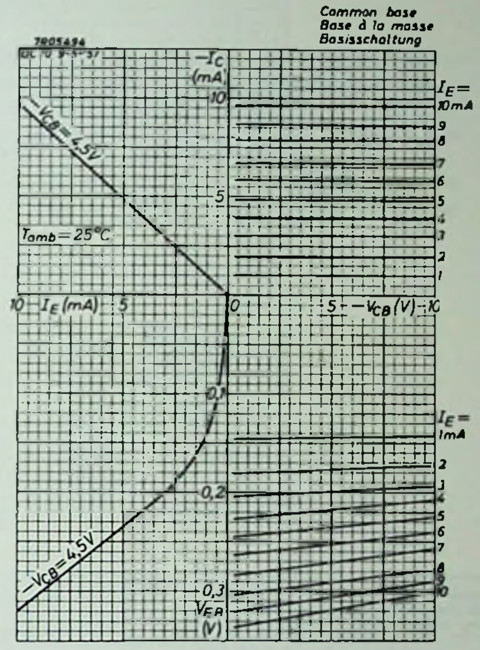


Fig. 11. Typical characteristics in common base connection at an ambient temperature of 25 °C.

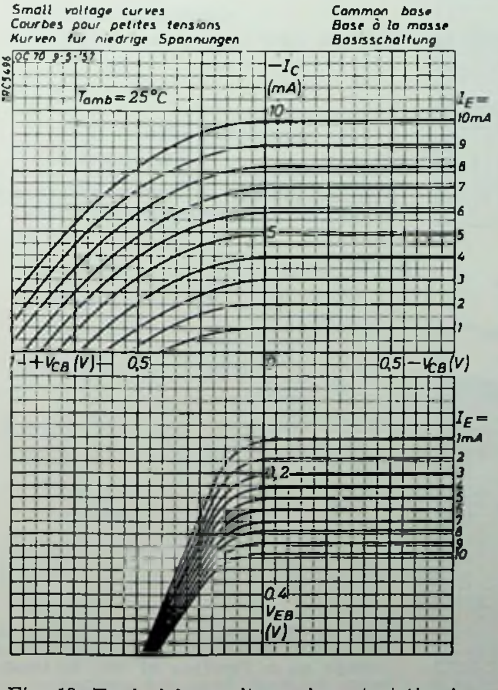


Fig. 12. Typical low-voltage characteristics in common base connection at an ambient temperature of 25 °C.

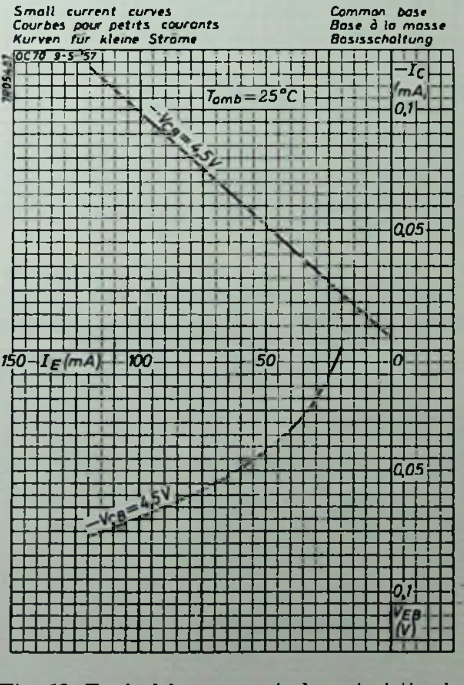


Fig. 13. Typical low-current characteristics in common base connection at an ambient temperature of 25 °C.

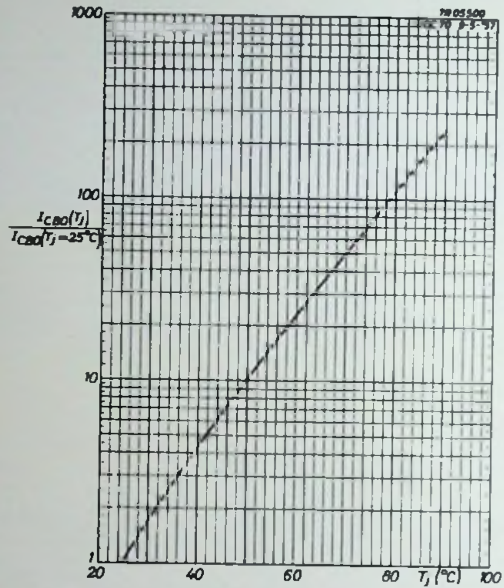


Fig. 14. Collector cut-off current versus junction temperature in common base connection

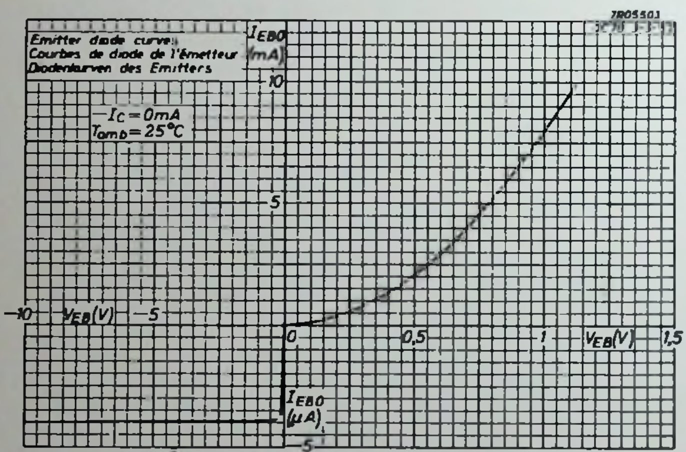


Fig. 15. Typical emitter diode characteristic at an ambient temperature of 25 °C.

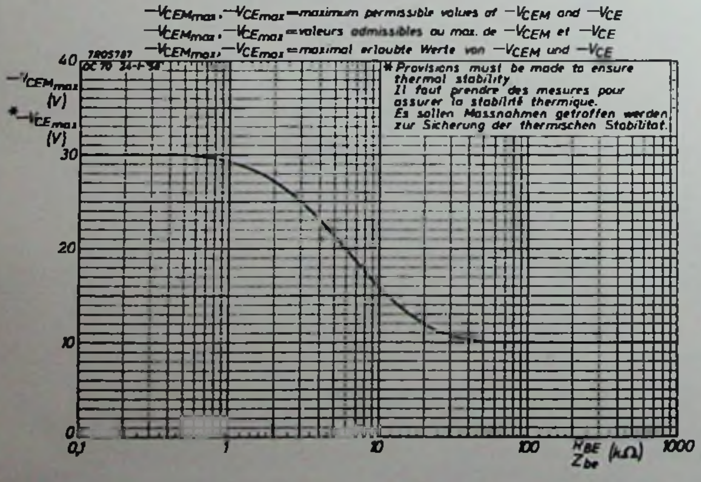


Fig. 16. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the external impedance between base and emitter.

ALL-GLASS GERMANIUM TRANSISTOR OC 71

The OC 71 is a junction transistor of the P-N-P type in all-glass construction. It is particularly recommended for medium gain lower-power audio frequency applications and especially designed for use in the first stage of hearing-aid circuits. Moreover the OC 71 can be used as a general-purpose amplifier at frequencies up to ∾ 0.3 Mc/s. Further applications are in switching and oscillator circuits where large signals are involved.
The all-glass construction ensures absolute moisture resistance and long life. The transistor is shock-proof and insensitive to ambient illumination. It may be soldered in or clipped for plug-in connection.

MECHANICAL DATA

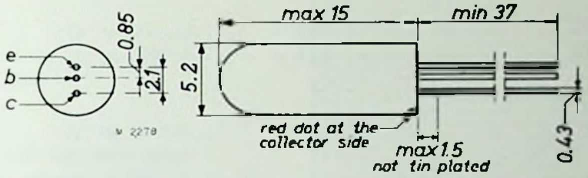


Fig. 1. Dimensional outline in mm and electrode connections.

THERMAL DATA

Junction temperature rise to ambient temperature, transistor in free air	K = max. 0.4 °C/mW

ABSOLUTE MAXIMUM RATINGS

Junction temperature

storage	T_s = min. −55 °C, max. +75 °C
continuous operation	T_j = max. 75 °C
intermittent operation, total duration max. 200 h	T_j = max. 90 °C [1]

Collector

Voltage (emitter reference)	
Peak and D.C.	see Fig. 16.
Peak, base at least 0.1 V positive	$-V_{CEM}$ = max. 30 V
Average, base at least 0.1 V positive	$-V_{CE}$ = max. 30 V
Current	
Peak	$-I_{CM}$ = max. 50 mA
Average	$-I_C$ = max. 10 mA
Dissipation (in free air)	see figure 2.

Emitter

Current	
Peak	I_{EM} = max. 55 mA
Average	I_E = max. 12 mA

Base

Current	
Peak	$-I_{BM}$ = max. 5 mA
Average	$-I_B$ = max. 2 mA

[1] Likelihood of full performance of a circuit at this temperature is also dependent on the type of application.

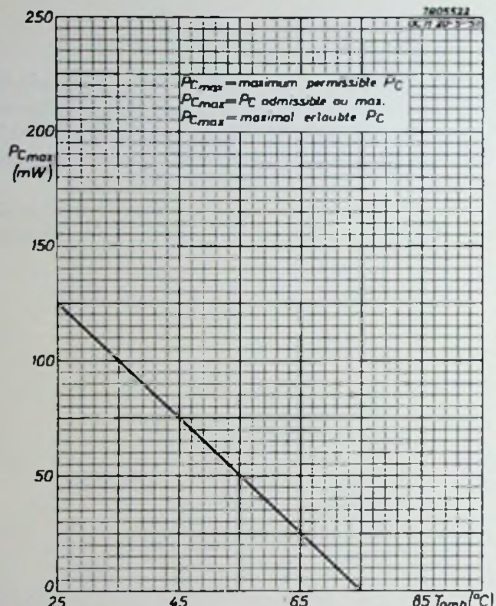


Fig. 2. Maximum collector dissipation as a function of the ambient temperature.

EQUIVALENT CIRCUIT PARAMETERS AT AN AMBIENT TEMPERATURE OF 25 °C

Common base circuit

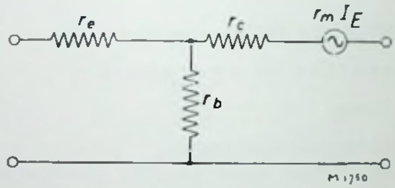


Fig. 3.

at $-V_{CB}$ = 2 V
$-I_C$ = 3 mA
f = 1000 c/s

Emitter resistance	r_e	= 6.5 Ω
Base resistance	r_b	= 500 Ω
Collector resistance	r_c	= 625 kΩ
Transfer resistance	r_m	= 611 kΩ

		min. [1])	average	max. [1])
Input impedance (output short-circuited)	h_{11b} =	10	17	25 Ω
Current ampl. factor (output short-circuited)	$-h_{21b}$ =	0.968	0.979	0.987
Output admittance (input open)	h_{22b} =		1.6	2.7 μA/V
Voltage feedback ratio (input open)	h_{12b} =		8.10^{-4}	

Common emitter circuit

at $-V_{CE}$ = 2 V
$-I_C$ = 3 mA
f = 1000 c/s

		min. [1])	average	max. [1])
Input impedance (output short-circuited)	h_{11e} =	0.4	0.8	1.5 kΩ
Current ampl. factor (output short-circuited)	h_{21e} =	30	47	75
Output admittance (input open)	h_{22e} =		80	200 μA/V
Voltage feedback ratio (input open)	h_{12e} =		$5.4.10^{-4}$	17.10^{-4}

---
[1]) Typical production spread.

CHARACTERISTICS AT AN AMBIENT TEMPERATURE OF 25 °C

Common base circuit

			min.	average	max.
Collector current at					
$-V_{CB}$ = 4.5 V; I_E = 0	$-I_{CBO}$	=		4.5	12 μA

Common emitter circuit

			min.	average	max.
Collector current at					
$-V_{CE}$ = 4.5 V, I_B = 0	$-I_{CEO}$	=		150	325 μA
$-V_{CE}$ = 4.5 V; $-I_B$ = 10 μA	$-I_C$	=	0.33	0.7	1.2 mA
$-V_{CE}$ = 4.5 V; $-I_B$ = 250 μA	$-I_C$	=	7.2	14	21 mA
Base voltage at					
$-V_{CE}$ = 4.5 V; $-I_B$ = 10 μA	$-V_{BE}$	=	80	110	155 mV
$-V_{CE}$ = 4.5 V; $-I_B$ = 250 μA	$-V_{BE}$	=	210	270	385 mV
Current amplification cut-off frequency at					
$-V_{CE}$ = 2 V; $-I_C$ = 3 mA	f_{α_e}	=		10	kc/s
Noise factor at 1000 c/s, measured with a source impedance of 500 Ω at					
$-V_{CE}$ = 2 V; $-I_C$ = 0.5 mA	F	=		10	15 [1]) dB

---

[1]) Typical production spread.

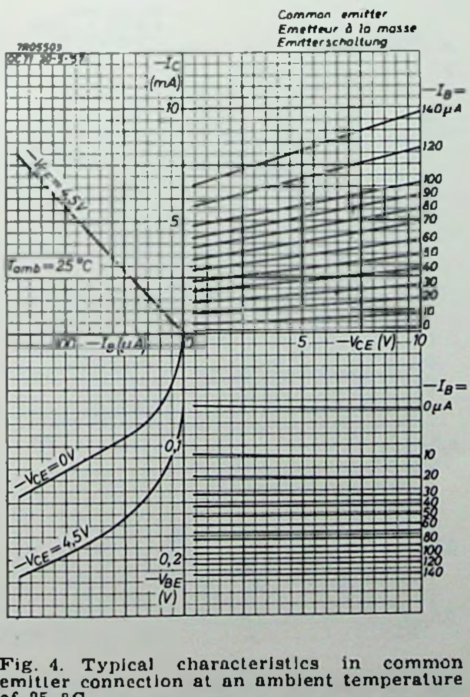


Fig. 4. Typical characteristics in common emitter connection at an ambient temperature of 25 °C.

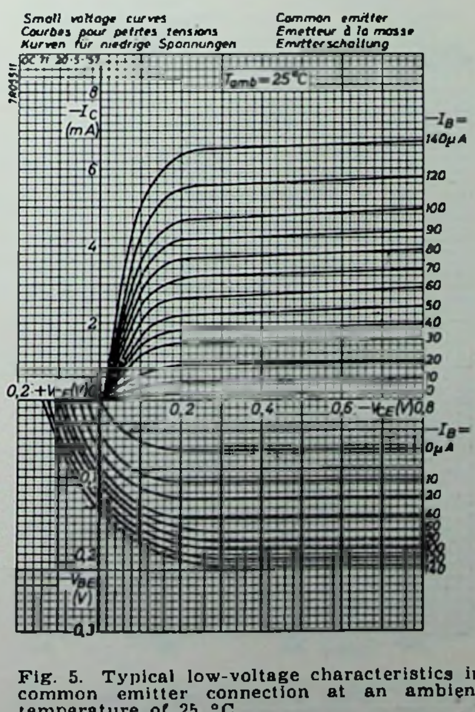


Fig. 5. Typical low-voltage characteristics in common emitter connection at an ambient temperature of 25 °C.

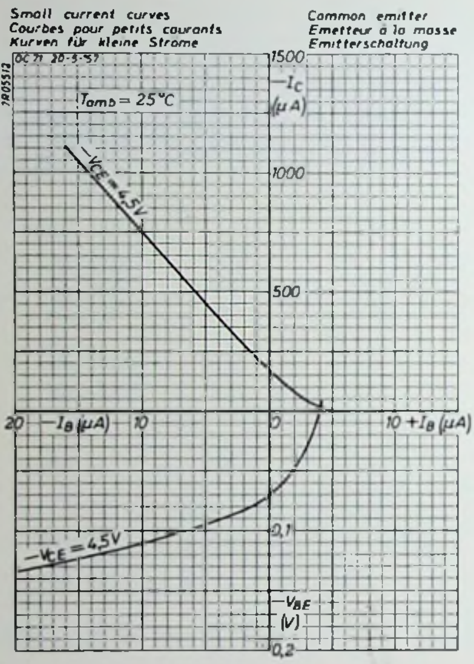


Fig. 6. Typical low-current characteristics in common emitter connection at an ambient temperature of 25 °C.

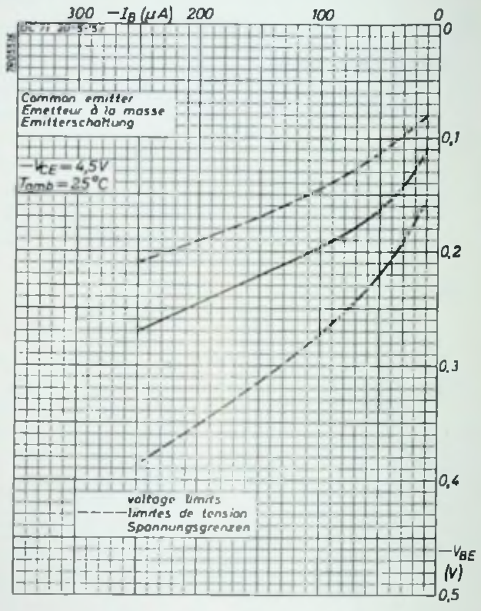


Fig. 7. Input characteristics in common emitter connection at an ambient temperature of 25 °C.

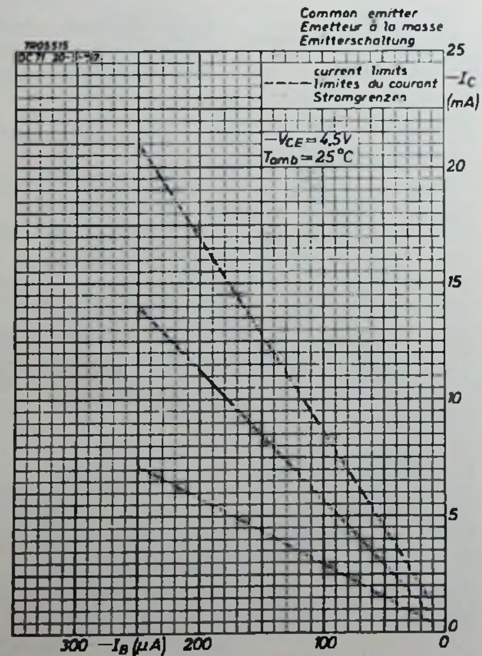


Fig. 8. Current amplification characteristics in common emitter connection at an ambient temperature of 25 °C.

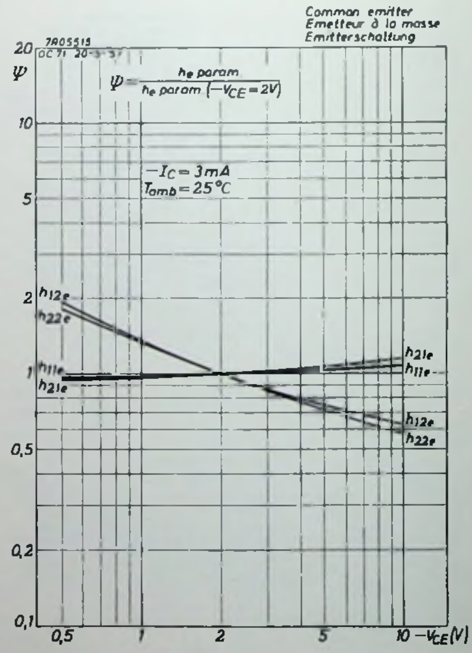


Fig. 9. h-parameters versus collector voltage in common emitter connection at an ambient temperature of 25 °C.

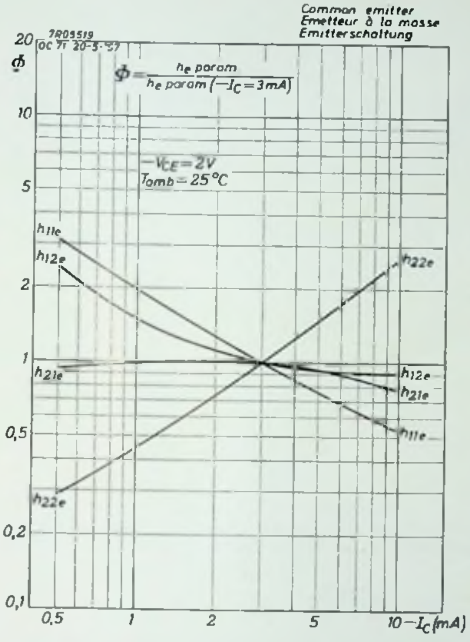


Fig. 10. h-parameters versus collector current in common emitter connection at an ambient temperature of 25 °C.

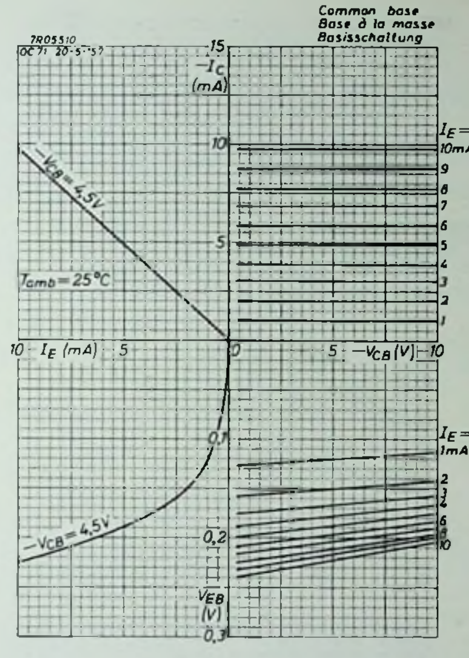


Fig. 11. Typical characteristics in common base connection at an ambient temperature of 25 °C.

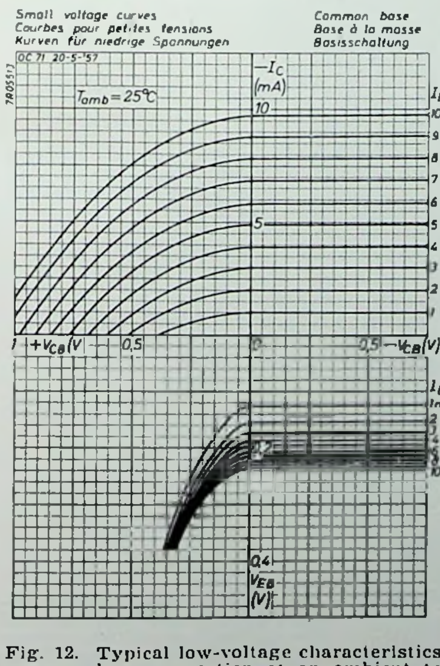


Fig. 12. Typical low-voltage characteristics in common base connection at an ambient temperature of 25 °C.

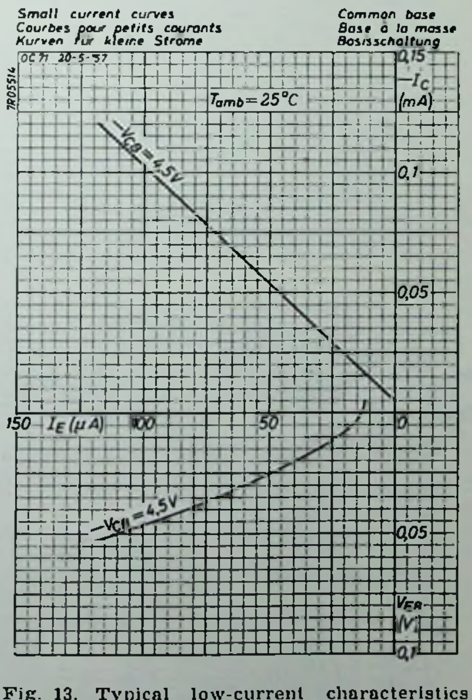


Fig. 13. Typical low-current characteristics in common base connection at an ambient temperature of 25 °C.

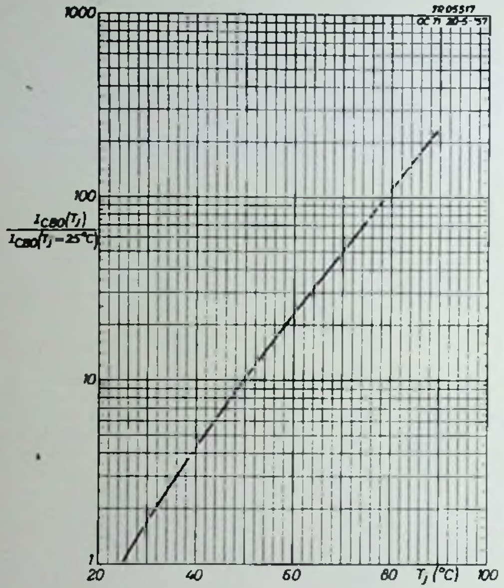


Fig. 14. Collector cut-off current versus junction temperature in common base connection.

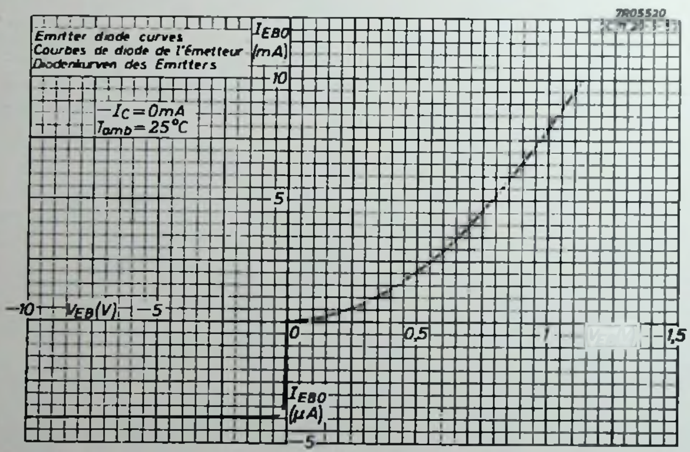


Fig. 15. Typical emitter diode characteristic at an ambient temperature of 25 °C.

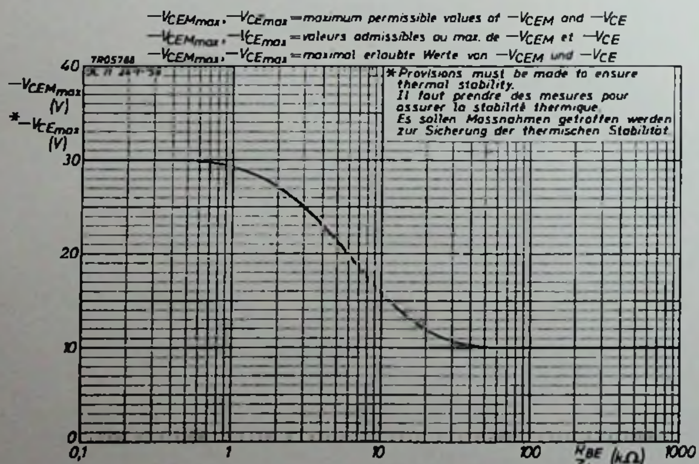


Fig. 16. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the external impedance between base and emitter.

P-N-P GERMANIUM TRANSISTOR OC 72 AND 2-OC 72

The OC 72 is an alloy junction transistor of the germanium P-N-P type of hermetically sealed construction with metal envelope.
This type has been designed for use in audio frequency output and driver stages and is also suitable for low-frequency switching and pulse-oscillator circuits, such as d.c. convertors.
Owing to the design, the transistors show a remarkable constancy of the current amplification factor up to high values of the collector current.
The OC 72 is capable of dissipating 75 mW at an ambient temperature of 45 °C. By using a copper cooling fin such as represented in Fig. 2, and connecting the fin to a heatsink, the maximum permissible collector dissipation may be increased up to 100 mW at an ambient temperature of 45 °C.
For use in class B output stages, a matched pair of transistors OC 72, selected to operate with low distortion and small spread in quiescent currents is available under type number 2-OC 72.

MECHANICAL DATA:

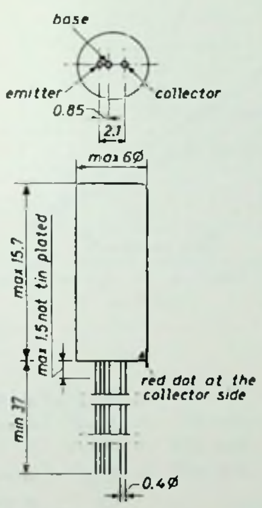


Fig. 1. Dimensional outline in mm and electrode connections.

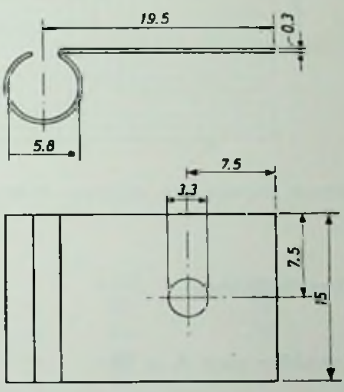


Fig. 2. Dimensions of the cooling fin, which is separately available under type number 56200.

THERMAL DATA:

Junction temperature rise to ambient temperature:

without cooling fin, transistor in 0 – 45 °C free air	K = max. 0.4 °C/mW
with cooling fin, mounted on a heat sink of 12.5 cm² in 0 – 45 °C free air	K = max. 0.3 °C/mW

ABSOLUTE MAXIMUM RATINGS:

T e m p e r a t u r e

storage	T_s = min. −55 °C, max. 75 °C
junction, continuous operation	T_j = max. 75 °C
junction, incidental operation (total duration max. 200 h)	T_j = max. 90 °C

C o l l e c t o r

Voltage (base reference)	
Peak	$-V_{CBM}$ = max. 32 V
D.C.	$-V_{CB}$ = max. 32 V
Voltage (emitter reference)	see Fig. 13

Current (amplifier class A or B)		
Peak	$-I_{CM}$	= 125 mA [1]
Average (averaging time max. 20 msec.)	$-I_C$	= 50 mA [1]
Current (switching or oscillating purposes)		
Peak (in both directions)	$\lvert I_{CM} \rvert$	= max. 250 mA
Average (averaging time max. 20 msec)	$-I_C$	= max. 125 mA
Dissipation (in free air)		

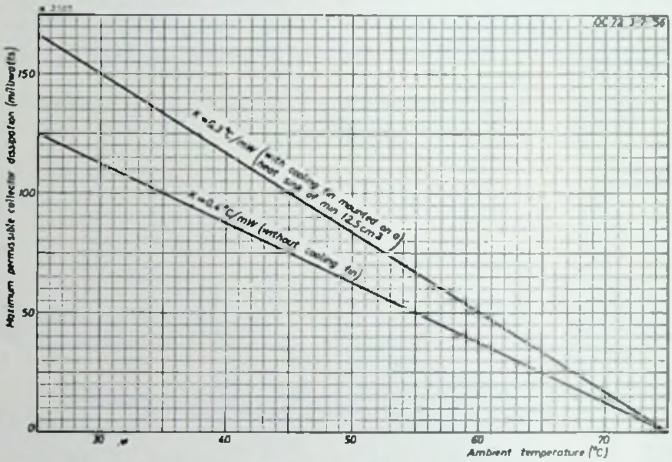


Fig. 3. Maximum permissible collector dissipation as a function of the ambient temperature.

Emitter

Voltage (base reference)		
Peak	$-V_{EBM}$	= max. 10 V
D.C.	$-V_{EB}$	= max. 10 V
Current (amplifier class A or B)		
Peak	I_{EM}	= max. 130 mA [1]
Average (averaging time max. 20 msec)	I_E	= max. 50 mA [1]
Current (switching or oscillating purposes)		
Peak (in both directions)	$\lvert I_{EM} \rvert$	= max. 250 mA
Average (averaging time max. 20 msec)	I_E	= max. 125 mA

Base

Current (switching or oscillating purposes)		
Peak (in both directions)	$\lvert I_{BM} \rvert$	= max. 125 mA
Average (averaging time max. 20 msec)	$-I_B$	= max. 20 mA

CHARACTERISTICS of the SINGLE TRANSISTOR at AMBIENT TEMPERATURE of 25 °C

Common Base Circuit (base reference)

			min.	average	max.
Collector current at					
$-V_{CB}$ = 10 V; I_E = 0	$-I_{CBO}$	=		4.5	10 μA
Emitter current at					
$-V_{EB}$ = 10 V; I_C = 0	$-I_{EBO}$	=		4.5	10 μA
Current multiplication frequency cut-off at					
$-V_{CB}$ = 6 V; I_E = 10 mA	$f_{\alpha b}$	=	350		kc/s

For footnotes see page 121.

Common Emitter Circuit (emitter reference)

		min.	average	max.
Collector current at				
$-V_{CE}$ = 6 V; I_B = 0	$-I_{CEO}$ =	50	125	300 μA 2)
$-V_{CE}$ = 30 V; V_{BE} > 0.5 V	$-I_C$ =	3	7.5	15 μA 2)
Collector knee voltage at $-I_C$ = 125 mA and for the value of $-I_B$ that occurs at $-V_{CE}$ = 1 V and $-I_C$ = 135 mA (see Fig. 4)	$-V_{CEK}$ =			0.4 V

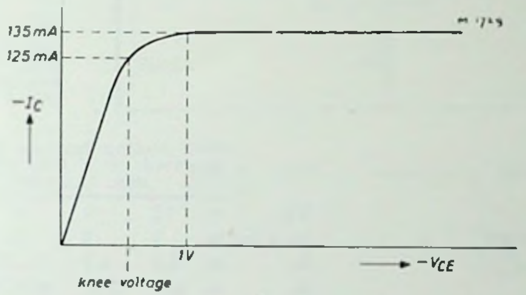


Fig. 4

Large signal d.c. current amplification factor

$$\alpha_{FE} = \frac{I_C - I_{CEO}}{I_B}$$

		min.	average	max.
at $-V_{CE}$ = 5.4 V; I_E = 10 mA	α_{FE} =	45	70	120 2)
at $-V_{CE}$ = 0.7 V; I_E = 80 mA	α_{FE} =	30	50	90 2)
at $-V_{CE}$ = 0.7 V; I_E = 125 mA	α_{FE} =	25		
at $-V_{CE}$ = 1 V; I_E = 250 mA	α_{FE} =	15		
Base input voltage				
at $-V_{CE}$ = 6 V; I_E = 1.5 mA	$-V_{BE}$ =	0.13		0.17 V 2)
at $-V_{CE}$ = 0.7 V; I_E = 80 mA	$-V_{BE}$ =			0.45 V
at $-V_{CE}$ = 0.7 V; I_E = 125 mA	$-V_{BE}$ =			0.70 V
Collector current				
at $-V_{CE}$ = 6 V; $-V_{BE}$ = 0.15 V	$-I_C$ =	0.70		3.20 mA

			min.	max.
Noise factor at 1000 c/s, measured with a source impedance of 500 Ω				
at $-V_{CE}$ = 2 V; I_E = 0.5 mA	F =			15 dB
Current multiplication frequency cut-off				
at $-V_{CE}$ = 6 V; I_E = 10 mA	f_{α_e} =		8	kc/s

CHARACTERISTICS of the MATCHED PAIR 2-OC 72 at an AMBIENT TEMPERATURE OF 25 °C

The two transistors of a matched pair are selected to operate in a class B circuit with low distortion at small and large signals and to have a low spread in quiescent currents.

	min.	average	max.
Ratio of the large signal d.c. current amplification factors of the two transistors at I_E = 80 and I_E = 10 mA	1.0	1.15	1.3 2)

For footnotes see page 121.

OPERATING CHARACTERISTICS as CLASS B AMPLIFIER in the OUTPUT STAGE of FIG. 5 at an AMBIENT TEMPERATURE of 25 °C

The circuits below are designed for stable operation up to an ambient temperature of 45 °C.

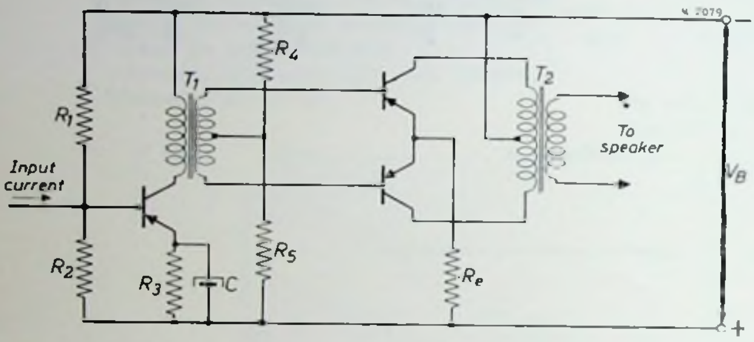


Fig. 5. Output stage with a matched pair of transistors 2-OC 72 in a class B amplifier and a transistor OC 71 in the driver stage.

		with cooling fin (see fig. 2) mounted on a heat sink of min. 12.5 cm²				without cooling fin	
Supply voltage	V_B =	12	9	6	6	6	4.5 V
Zero signal emitter current	$I_{E1}+I_{E2}$ =	3	3	3	3	3	3 mA
Bias resistor	R_4 =	4.7	4.7	1 – 3[3])	3.3	3.3	2.7 kΩ
Bias resistor	R_5 =	100	100	NTC[4])	100	100	100 Ω
Emitter resistor	R_E [5]) =	30	14		5	10	5 Ω
Maximum transistor output power, two transistors	$P_{C\,max}$ =	500	420	240	350	350	260 mW
Maximum power delivered to the primary of the output transformer, two transistors	$P_{O\,max}$ =	390	355	240	310	275	220 mW
Load impedance collector to collector	R_{CC} =	430	305	280	160	140	115 Ω
Load impedance ($R_{CE} = \frac{R_{CC}}{4} + R_E$) per transistor	R_{CE} =	138	90	70	45	45	34 Ω
At maximum power output:							
peak collector current	$-I_{CM}$ =	85	100	85	125	125	125 mA
D.C. collector current	$-I_C$ =	27	32	27	40	40	40 mA
maximum required peak driving voltage per transistor (see notes 6 and 7)	V_{bm} =	3.4	2.4	0.6	2.1	2.8	2.1 V
maximum required peak driving current per transistor (see note 7)	I_{bm} =	2.8	3.2	2.8	4.9	4.9	4.9 mA
total harmonic distortion	d_{tot} =	8.5	8.5	8.5	9.5	9.5	9 %
At an output power of 50 mW in the primary of the output transformer:							
peak driving voltage per transistor [6])	V_{bm} =	1.0	0.66	0.20	0.53	0.80	0.63 V
peak driving current per transistor	I_{bm} =	0.42	0.49	0.56	0.70	0.74	1.0 mA
total harmonic distortion	d_{tot} =	4.5	4.5	5.5	5	5	5.5 %
Data of the driver stage:							
D.C. collector to emitter voltage	$-V_{CE}$ =	10.5	4.1	4.2	4.5	4.5	3.0 V
D.C. emitter current	I_E =	1.3	3.0	2.3	4.0	4.8	6.5 mA
Bias resistor	R_1 =	68	12	39	15	8.2	6.8 kΩ
Bias resistor	R_2 =	8.2	15	15	4.7	2.7	2.2 kΩ
Emitter resistor	R_3 =	820	1500	470	270	220	120 Ω
Emitter capacitor	C =	100	100	100	100	100	100 μF
At an output power of 50 mW in the primary of the output transformer:							
peak base current	$-I_{bm}$ =	7	10.5	3.6	11	14	23 μA
peak input current (see Fig. 5)	I_{im} =	8.4	12	4.0	13.5	17.5	31 μA
Data of the driver transformer:							
primary to secondary turns ratio		$\frac{3.0}{1+1}$	$\frac{1.4}{1+1}$	$\frac{3.5}{1+1}$	$\frac{1.7}{1+1}$	$\frac{1.35}{1+1}$	$\frac{1.0}{1+1}$

For footnotes see page 121.

OPERATING CHARACTERISTICS of the CLASS A AMPLIFIER STAGE of FIG. 6 at an AMBIENT TEMPERATURE of 25 °C

The values of the emitter resistance in the circuits below are based upon full interchangeability of the transistors and such a stabilisation of the currents that the maximum junction temperature is not exceeded up to an ambient temperature of 45 °C.

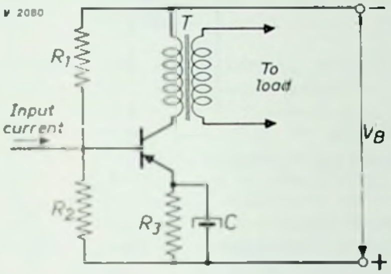


Fig. 6. Class A amplifying stage with a transistor OC 72.

			With cooling fin (see fig. 2) mounted on a heat sink of min. 12.5 cm²		
Supply voltage	V_B	=	6	9	12 V
D.C. collector current	$-I_C$	=	16.3	10.6	8.2 mA
Bias resistor	R_1	=	3.3	8.2	18 kΩ
Bias resistor	R_2	=	1.0	2.2	4.7 kΩ
Emitter resistor	R_E	=	62	140	280 Ω
Emitter capacitor	C	=	250	250	250 μF
Maximum power output	$P_{O\,max}$	=	38	38	38 mW
Collector load impedance	R_L	=	300	680	1150 Ω
At maximum power output:					
base current	I_B [8])	=	0.16	0.11	0.09 mA (rms)
input current (see Fig. 6)	I_i	=	0.22	0.13	0.09 mA (rms)
total harmonic distortion	d_{tot}	=	3.6	3.8	3.6 %

OPERATING CHARACTERISTICS as SWITCHING TRANSISTOR at an AMBIENT TEMPERATURE of 25 °C

Circuit			Fig. 7	Fig. 8
Supply voltage	V_B	=	6	6 V
Battery current	I_B	=	154	28 mA
Input power	P_I	=	924	168 mW
Output voltage	V_O	=	75.5	45 V
Output current	I_O	=	9.4	3 mA
Output power	P_O	=	710	135 mW
Efficiency	η	=	77	81 %
Total transistor dissipation	P_T	=	86	11.7 mW
Total diode losses			39	6.1 mW
Total transformer losses			35	14.3 mW
Total resistor losses			54	0.9 mW
Output resistance			< 1.4	2 kΩ

For footnotes see page 121.

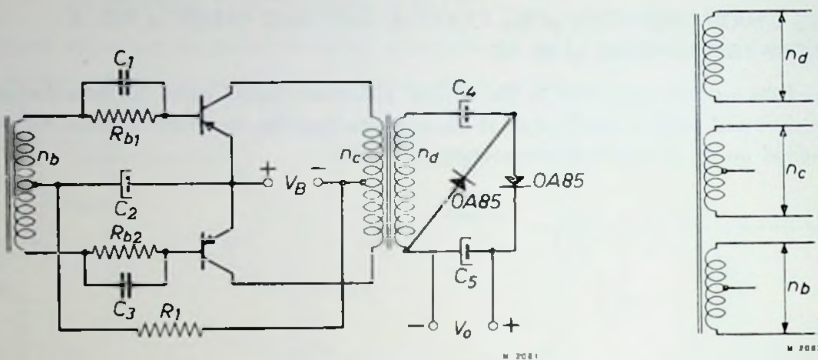


Fig. 7. Push-pull D.C. convertor with two transistors OC 72 without cooling fin.

R_{b1} = 270 Ω
R_{b2} = 270 Ω
R_1 = 820 Ω
C_1 = 47000 pF
C_2 = 16 μF
C_3 = 47000 pF
C_4 = 8 μF
C_5 = 8 μF

transformer turns ratio:

$$\frac{n_d}{n_c} = \frac{10}{1.37 + 1.37}$$

$$\frac{n_c}{n_b} = 2.7$$

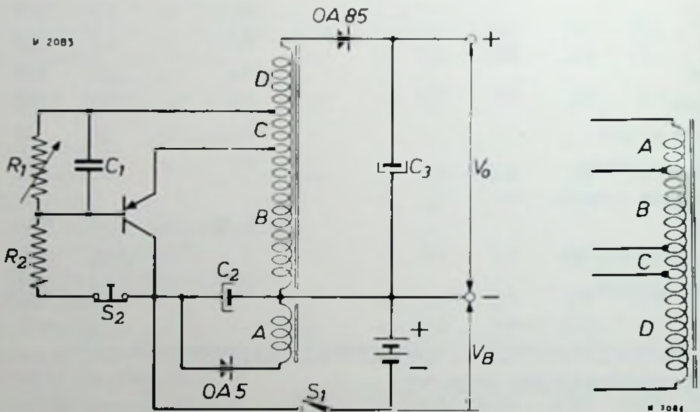


Fig. 8. D.C. convertor with a single transistor OC 72 without cooling fin. By means of the switches S_1 and S_2 which are mechanically coupled so that S_2 is opened after S_1 has been closed, the oscillation is initiated.

R_1 = 1 kΩ
R_2 = 2.7 kΩ
C_1 = 0.03 μF
C_2 = 100 μF
C_3 = 3.2 μF

transformer turns ratio:

$$\frac{n_A}{n_t} = 0.12 \qquad \frac{n_B}{n_t} = 0.32$$

$$\frac{n_C}{n_t} = 0.058 \qquad \frac{n_D}{n_t} = 0.5$$

$$(n_t = n_A + n_B + n_C + n_D)$$

NOTES

[1]) These ratings are based on the low-distortion requirements in amplifiers.

[2]) Typical production spread.

[3]) In this case R_4 is a variable resistor of 1—3 kΩ.

[4]) Here R_5 consists of an 85 Ω resistor in parallel with an NTC-resistor of 130 Ω at 25 °C, B = 4500 °K.

[5]) Instability of the working point of transistors due to thermal run-away can be prevented by using either a temperature-dependent base bias (e.g. with an NTC resistor) or a resistor R_e in the emitter lead. Both types of stabilisation, ensuring stable operation of the circuits up to an ambient temperature of 45 °C, are given; it can be seen from the operating characteristics that R_e decreases the useful power output and the sensitivity compared with the NTC resistor method. However, stabilisation by means of an emitter resistor has the important advantage of full interchangeability of the transistors (no adjustable bias required).
In the circuits shown under operating characteristics, the two transistors OC 72 have a common emitter resistor, which causes a sort of cross-over distortion, only noticeable at very small signals. This effect can be eliminated by using separate emitter resistors. To ensure the same thermal stability of the circuits the values of each resistor should be a factor 1.15 higher than the values of the common emitter resistor; consequently the circuits should be adapted to these higher values.

[6]) The losses in the bias resistor R_5 and in the resistances of the driver transformer are included (total resistance transformed to secondary is 100 Ω).

[7]) For the determination of these values the most unfavourable transistor properties with respect to maximum required driver voltage and current have been combined. These values enable the designer to dimension the driver stage.

[8]) The maximum required input currents to be taken into account in the design of the preceding stage, are 50 % higher (maximum values based upon the most unfavourable combination of the transistor properties).

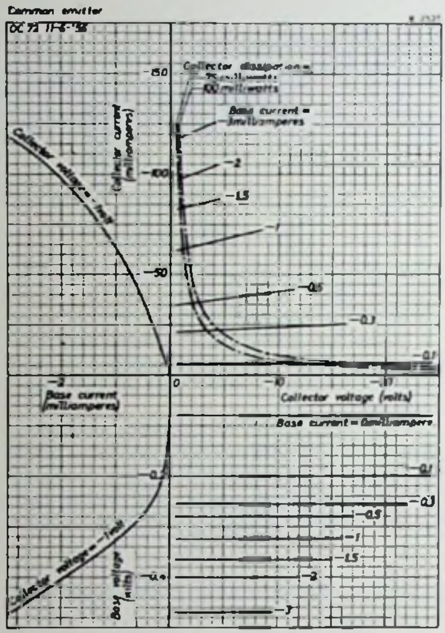


Fig. 9. Typical characteristics of the OC 72 in common emitter connection at constant junction temperature.

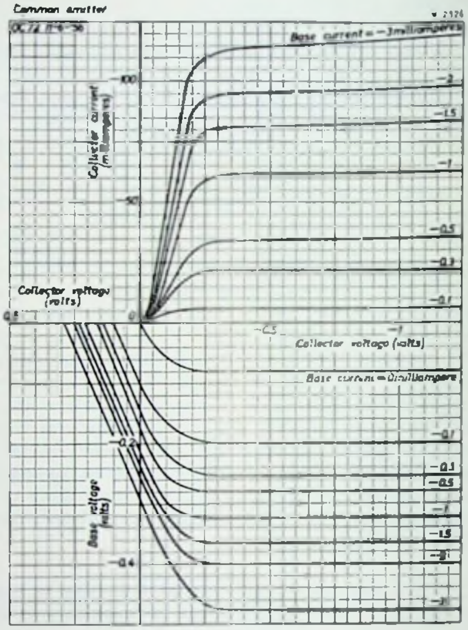


Fig. 10. Typical low-voltage characteristics of the OC 72 in common emitter connection at constant junction temperature.

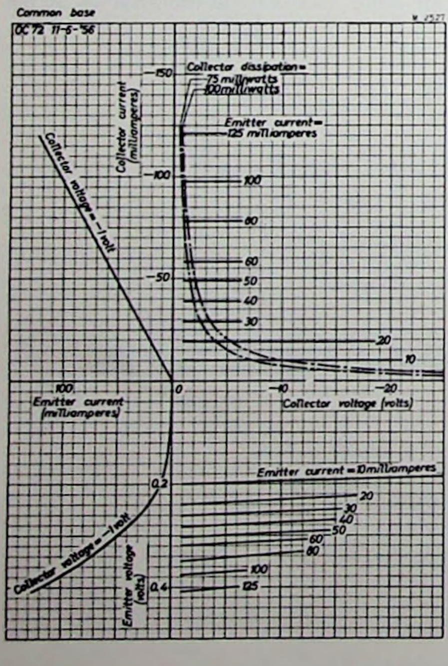


Fig. 11. Typical characteristics of the OC 72 in common base connection at constant junction temperature.

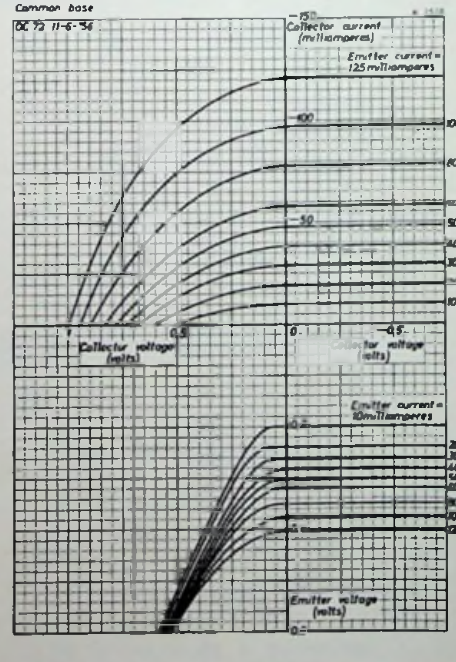


Fig. 12. Typical low-voltage characteristics of the OC 72 in common base connection at constant junction temperature.

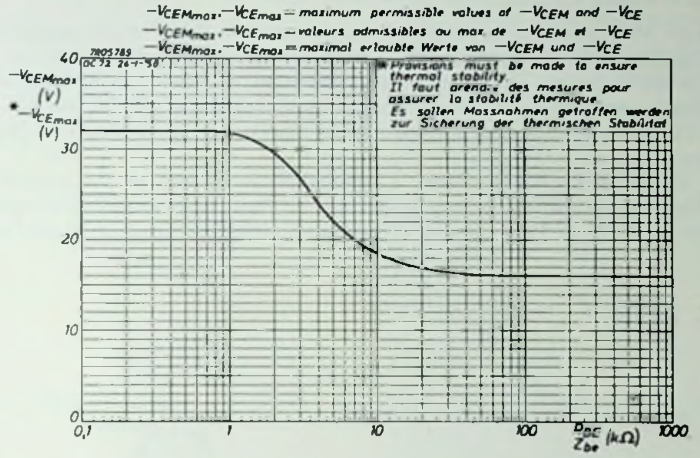


Fig 13. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the external impedance between base and emitter. Provisions must be made to ensure thermal stability.

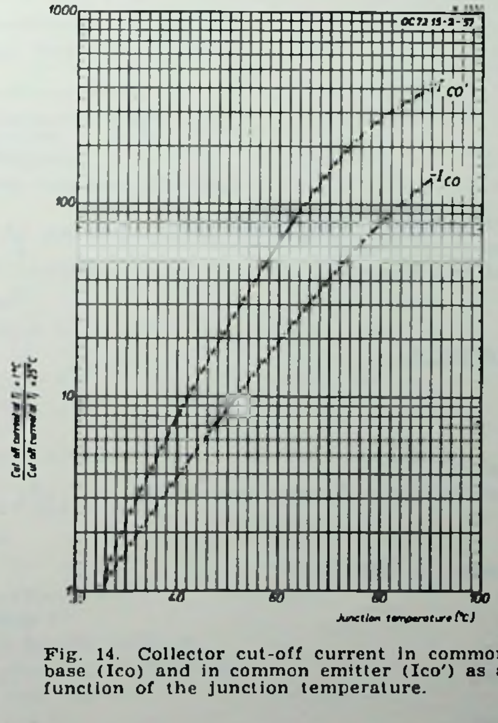


Fig. 14. Collector cut-off current in common base (Ico) and in common emitter (Ico') as a function of the junction temperature.

P-N-P GERMANIUM TRANSISTOR OC 76

The OC 76 is an alloy junction transistor of the germanium P-N-P type, especially suitable for switching and pulse-oscillating circuits. This transistor is of hermetically sealed all-glass construction with a metal cover. The all-glass encapsulation ensures high stability and makes the transistor absolutely moisture proof. At an ambient temperature of 45 °C the OC 76 is capable of dissipating 75 mW. By using a cooling fin as shown in Fig. 2 this may be increased to 100 mW.
Owing to the design, the OC 76 shows a remarkable constancy of the current amplification factor up to high values of the collector current, so that in push-pull d.c. converter circuits an output power of 1 W at 12 V supply voltage and of 0.75 W at 6 V supply voltage can be obtained.

MECHANICAL DATA

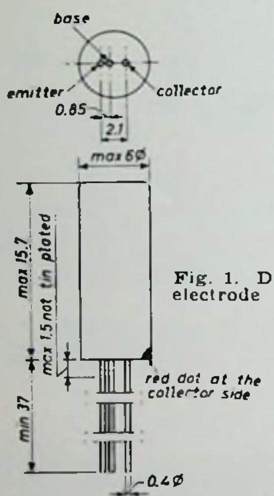


Fig. 1. Dimensional outline in mm and electrode connections of the transistor OC 76

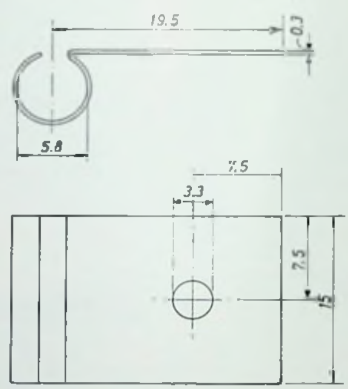


Fig 2. Dimensions of the cooling fin, which is separately available under type number 56200

THERMAL DATA:

Junction temperature rise to ambient temperature:

without cooling fin, transistor in free air	K = max. 0.4 °C/mW
with cooling fin, mounted on a heat sink of 12.5 cm², in free air	K = max. 0.3 °C/mW

ABSOLUTE MAXIMUM RATINGS:

Temperature

storage	T_s	=	max.	−55 °C
		=	max.	75 °C
junction, continous operation	T_j	=	max.	75 °C
junction, intermittent operation (total duration max. 200 h)	T_j	=	max.	90 °C [1]

Collector

Voltage (see derating curves, p. 129)				
peak (base reference)	$-V_{CBM}$	=	max.	32 V
d.c. (base reference)	$-V_{CB}$	=	max.	32 V
peak (emitter reference)	$-V_{CEM}$	=	max.	32 V
d.c. (emitter reference)	$-V_{CE}$	=	max.	32 V
Current				
peak (in both directions)	$\lvert I_{CM} \rvert$	=	max.	250 mA
average (averaging time max. 20 msec)	$-I_C$	=	max.	125 mA

[1] Likelihood of full performance of a circuit at this temperature is also dependent on the type of application.

Dissipation (in free air)

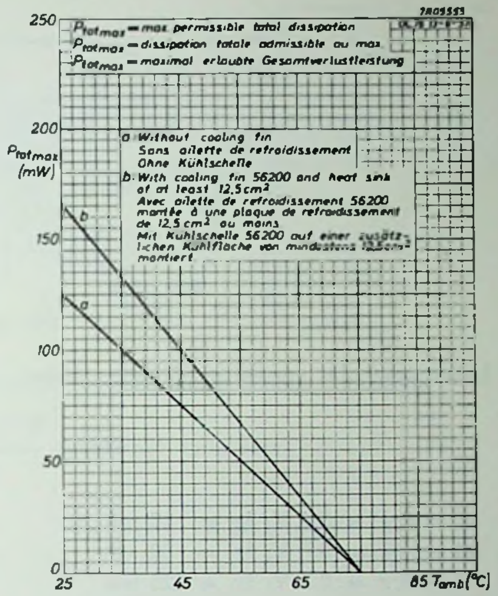


Fig. 3. Maximum collector dissipation as a function of the ambient temperature.

Emitter

Voltage (base reference)			
peak	$-V_{EBM}$	= max.	10 V
Current			
peak (in both directions)	$\lvert I_{EM} \rvert$	= max.	250 mA
average (averaging time max. 20 msec)	I_E	= max.	125 mA

Base

Current			
peak (in both directions)	$\lvert I_{BM} \rvert$	= max.	125 mA
average (averaging time max. 20 msec)	$-I_B$	= max.	20 mA

CHARACTERISTICS AT AN AMBIENT TEMPERATURE OF 25 °C

Common Base Circuit

			min.	average	max.
Collector current at					
$-V_{CB}$ = 10 V; I_E = 0	$-I_{CBO}$	=		4.5	10 μA
Emitter current at					
$-V_{EB}$ = 10 V; I_C = 0	$-I_{EBO}$	=		4.5	8 μA
Current amplification frequency cut-off at					
$-V_{CB}$ = 6 V; I_E = 10 mA	f_{α_b}	=	350 [1])	900	kc/s

Common Emitter Circuit

Collector current at				
$-V_{CE}$ = 6 V; I_B = 0	$-I_{CEO}$	=	200	600 μA [1])
$-V_{CE}$ = 30 V; V_{BE} > 0.5 V	$-I_C$	=	7.5	15 μA
Collector knee voltage at $-I_C$ 125 mA and for the value of $-I_B$ that occurs at $-V_{CE}$ = 1 V and $-I_C$ = 135 mA (see Fig. 4)	$-V_{CEK}$	=		0.4 V [1])

[1]) Typical production spread.

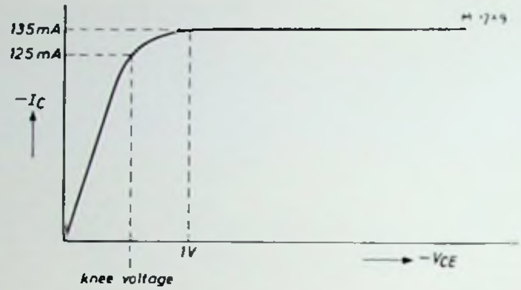


Fig. 4.

Large signal d.c. current amplification factor

$$\alpha_{FE} = \frac{I_C - I_{CEO}}{I_B}$$

at $-V_{CE}$ = 5.4 V; I_E = 10 mA	α_{FE} = min. 45 [1]
at $-V_{CE}$ = 0.7 V; I_E = 80 mA	α_{FE} = min. 30 [1]
at $-V_{CE}$ = 0.7 V; I_E = 125 mA	α_{FE} = min. 25 [1]
at $-V_{CE}$ = 1 V; I_E = 250 mA	α_{FE} = min. 15 [1]
Base input voltage	
at $-V_{CE}$ = 0.7 V; I_E = 80 mA	$-V_{BE}$ = max. 0.45 V [1]
at $-V_{CE}$ = 0.7 V; I_E = 125 mA	$-V_{BE}$ = max. 0.70 V [1]
Noise factor at 1000 c/s, measured with a source impedance of 500 Ω	
at $-V_{CE}$ = 2 V; I_E = 0.5 mA	F = max. 15 dB [1]

OPERATING CHARACTERISTICS as SWITCHING TRANSISTOR at an AMBIENT TEMPERATURE of 25 °C

Circuit		Fig. 5	Fig. 6	
Supply voltage	V_B =	6	6	V
Battery current	I_B =	154	28	mA
Input power	P_I =	924	168	mW
Output voltage	V_O =	75.5	45	V
Output current	I_O =	9.4	3	mA
Output power	P_O =	710	135	mW
Efficiency	η =	77	81	%
Total transistor dissipation	P_T =	86	11.7	mW
Total diode losses		39	6.1	mW
Total transformer losses		35	14.3	mW
Total resistor losses		54	0.9	mW
Output resistance		1.4	2	kΩ

---

[1]) Typical production spread.

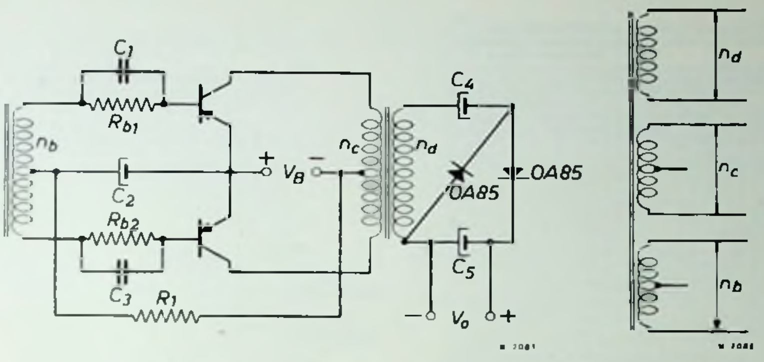


Fig. 5. Push-pull D.C. converter with two transistors OC 76 without cooling fin.

R_{b1} = 270 Ω
R_{b2} = 270 Ω
R_1 = 820 Ω
C_1 = 47000 pF
C_2 = 16 μF
C_3 = 47000 pF
C_4 = 8 μF
C_5 = 8 μF

transformer turns ratio:

$$\frac{n_d}{n_c} = \frac{10}{1.37 + 1.37}$$

$$\frac{n_c}{n_b} = 2.7$$

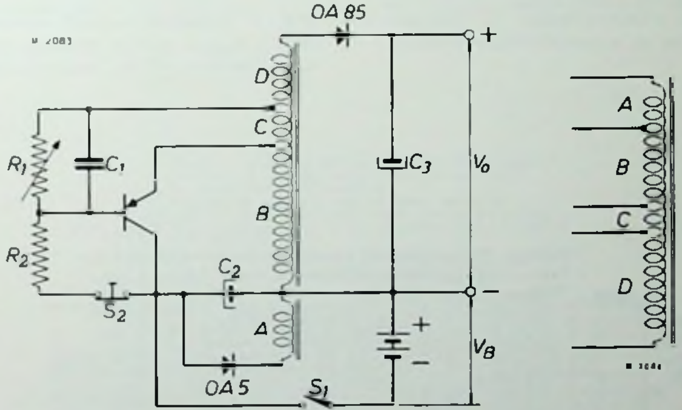


Fig. 6. D.C. converter with a transistor OC 76 without cooling fin. By means of the switches S_1 and S_2 which are mechanically coupled so that S_2 is opened after S_1 has been closed, the oscillation is initiated.

R_1 = 1 kΩ
R_2 = 2.7 kΩ
C_1 = 0.03 μF
C_2 = 100 μF
C_3 = 3.2 μF

transformer turns ratio:

$$\frac{n_A}{n_t} = 0.12 \quad \frac{n_B}{n_t} = 0.32 \quad \frac{n_C}{n_t} = 0.058 \quad \frac{n_D}{n_t} = 0.5$$

$(n_t = n_A + n_B + n_C + n_D)$

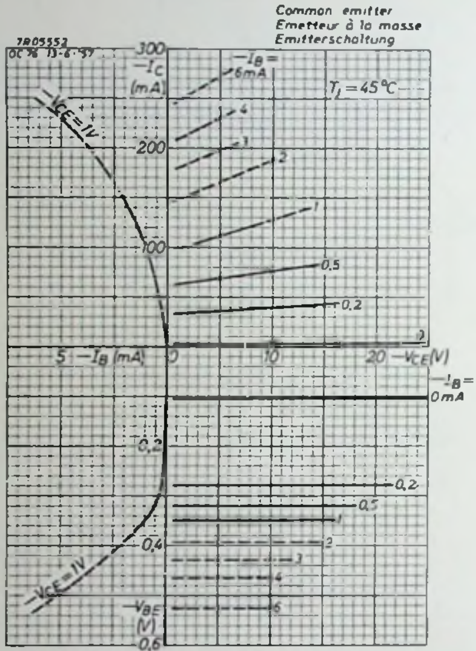


Fig. 7. Typical characteristics in common emitter connection at a junction temperature of 45 °C.

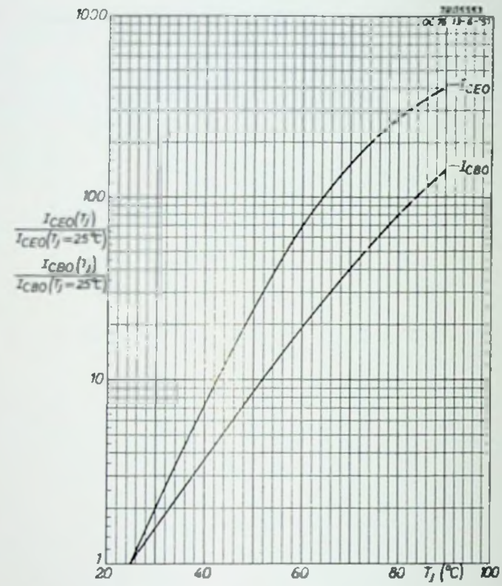


Fig. 8. Collector cut-off current in common base ($-I_{CBO}$) and in common emitter ($-I_{CEO}$) as a function of the junction temperature.

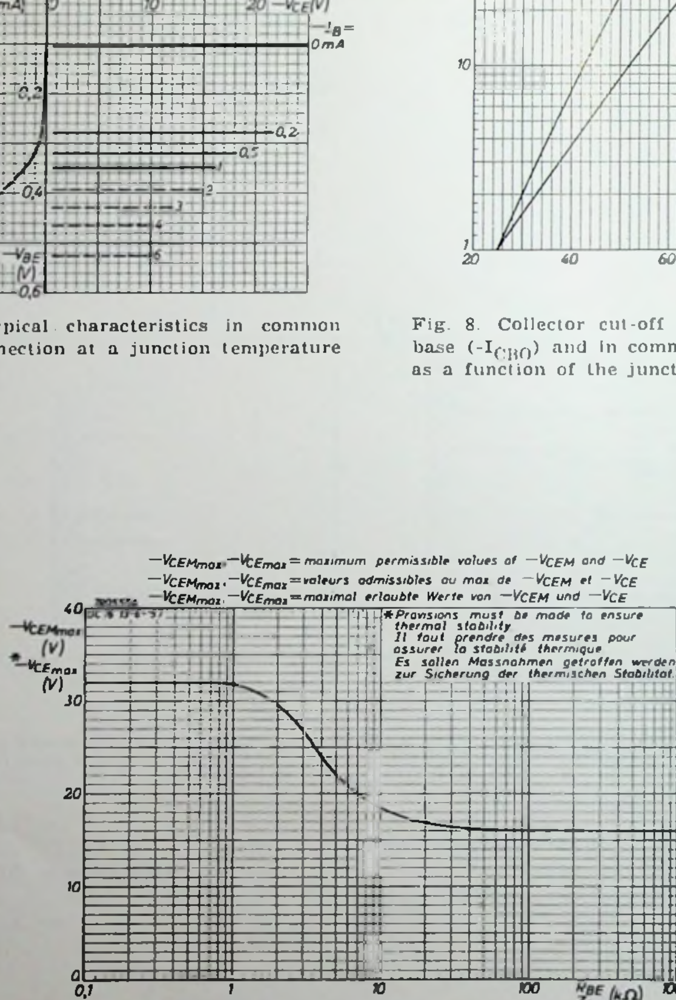


Fig. 9. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the external impedance between base and emitter.

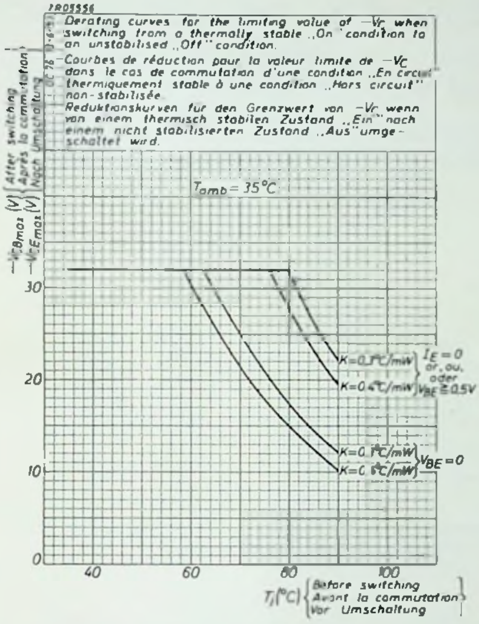


Fig. 10. Collector voltage derating curves at an ambient temperature of 35 °C.

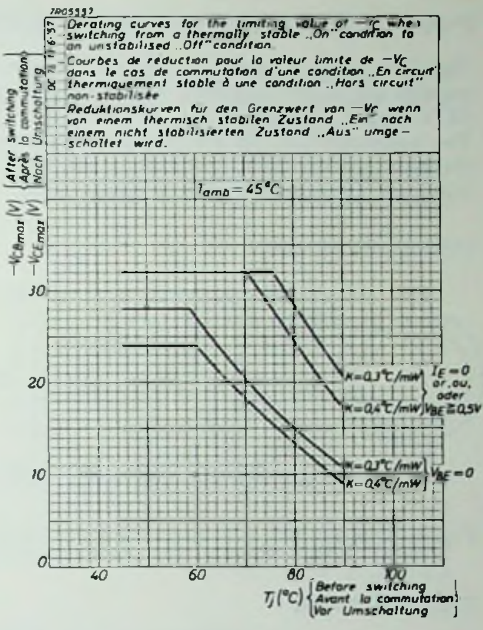


Fig. 11. Collector voltage derating curves at an ambient temperature of 45 °C.

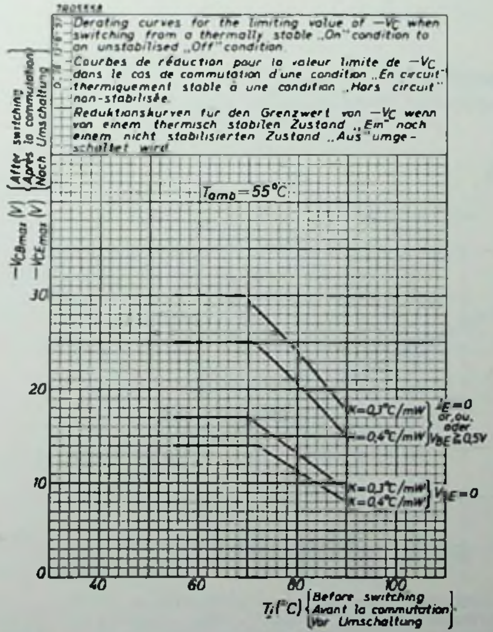


Fig. 12. Collector voltage derating curves at an ambient temperature of 55 °C.

P-N-P GERMANIUM POWER TRANSISTOR OC 16 AND 2-OC 16

The OC 16 is a germanium junction power transistor of the P-N-P alloy type, suitable for general use in low-frequency amplifying, switching and pulse-oscillating applications with 6 V or 12 V supply batteries.
The transistors OC 16 are supplied either in single units or in matched pairs; the latter have the type number 2-OC 16 and are selected in order to have low distortion and low spread in quiescent currents in a class B output stage.

MECHANICAL DATA

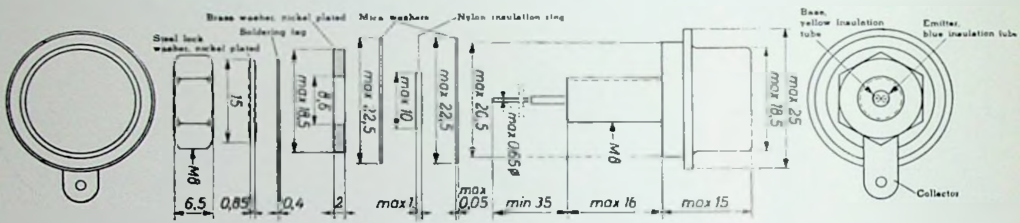


Fig. 1. Dimensional outline (in mm) and electrode connections. The collector is connected to the case; mica washers to insulate the case from the chassis are provided.

THERMAL DATA

Junction temperature rise to mounting base temperature (see application considerations p. 134)	K_m = max. 1.0 °C/W

ABSOLUTE MAXIMUM RATINGS

T e m p e r a t u r e

storage	T_s	= min. −55 °C = max. 75 °C
junction, continuous operation	T_j	= max. 75 °C
junction, intermittent operation (total duration max. 200 h)	T_j	= nax. 90 °C [1]

C o l l e c t o r

Dissipation (see Fig. 2)

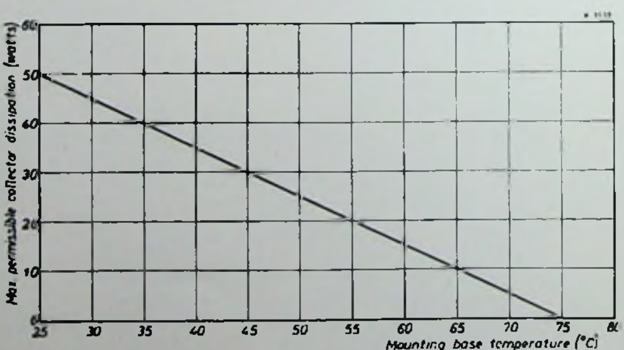


Fig. 2. The permissible collector dissipation is determined by the mounting base temperature. The mounting base temperature depends on the cooling capacity of the chassis or some other heat dissipating medium to which the transistor is attached.

---

[1]) Likelihood of full performance of a circuit at this temperature is also dependent on the type of application.

C o l l e c t o r

Voltage, base reference		
peak	$-V_{CBM}$	= max. 32 V
d.c.	$-V_{CB}$	= max. 32 V
Voltage, emitter reference	see Fig. 11	
Current		
peak	$-I_{CM}$	= max. 3 A
average (averaging time max. 20 msec)	$-I_C$	= max. 1.5 A

E m i t t e r

Voltage, base reference		
peak	$-V_{EBM}$	= max. 10 V
d.c.	$-V_{EB}$	= max. 10 V
Current		
peak	I_{EM}	= max. 3.3 A
average (averaging time max. 20 msec)	I_E	= max. 1.6 A

B a s e

Current		
peak	$-I_{BM}$	= max. 0.5 A
average (averaging time max. 20 msec)	$-I_B$	= max. 0.2 A

CHARACTERISTICS OF THE SINGLE TRANSISTOR AT AN AMBIENT TEMPERATURE OF 25 °C

C o m m o n b a s e c i r c u i t			average	max.
Collector current at $-V_{CB}$ = 14 V; I_E = 0	$-I_{CBO}$	=	0.02	0.1 mA
Emitter current at $-V_{EB}$ = 7 V; I_C = 0	$-I_{EBO}$	=	0.01	0.05 mA
Current amplification frequency cut-off at $-V_{CB}$ = 7 V; I_E = 0.3 A	f_{α_b}	=	200	kc/s

C o m m o n e m i t t e r c i r c u i t				
Collector current at $-V_{CE}$ = 14 V; I_B = 0	$-I_{CEO}$	=	0.6	2.5 mA
Collector knee voltage at $-I_C$ = 3 A; and for the value of $-I_B$ that occurs at $-V_{CE}$ = 1 V and $-I_C$ = 3.3 A (see Fig. 3)	$-V_{CEK}$	=	0.4	0.8 V

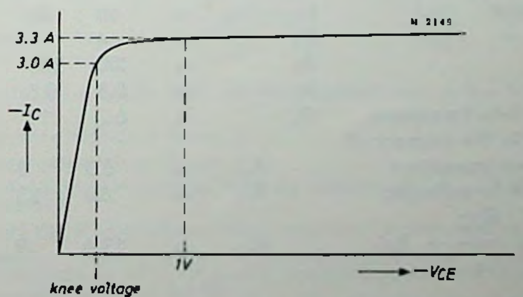


Fig. 3

Large signal characteristics:
d.c. current amplification factor

$$\alpha_{FE} = \frac{I_C - I_{CEO}}{I_B}$$

		average
at $-V_{CE}$ = 14 V; I_E = 0.03 A	α_{FE} =	40
at $-V_{CE}$ = 7 V; I_E = 0.3 A	α_{FE} =	45
at $-V_{CE}$ = 1 V; I_E = 2.0 A	α_{FE} =	22
at $-V_{CE}$ = 1 V; I_E = 3.0 A	α_{FE} =	18

base input voltage

at $-V_{CE}$ = 14 V; I_E = 0.03 A	$-V_{BE}$ =	0.19 V
at $-V_{CE}$ = 7 V; I_E = 0.3 A	$-V_{BE}$ =	0.32 V
at $-V_{CE}$ = 1 V; I_E = 2.0 A	$-V_{BE}$ =	0.8 V
at $-V_{CE}$ = 1 V; I_E = 3.0 A	$-V_{BE}$ =	1.0 V

CHARACTERISTICS OF THE MATCHED PAIR 2-OC 16 AT AN AMBIENT TEMPERATURE OF 25 °C

Ratio of the d.c. collector currents of the two transistors at
$-V_{CE}$ = 14 V; $-V_{BE}$ = 0.20 V max. 3

Ratio of the large signal d.c. current amplification factors of the two transistors
at I_E = 0.3 A; $-V_{CE}$ = 7 V and
I_E = 2.0 A; $-V_{CE}$ = 1 V max. 1.37

OPERATING CHARACTERISTICS AS CLASS B AMPLIFIER IN THE OUTPUT STAGE OF FIG. 4 AT AN AMBIENT TEMPERATURE OF 25 °C

The circuits below are designed for stable operation up to an ambient temperature of 55 °C. The total thermal resistance from collector junction to ambient temperature is taken to be 7 °C/W.

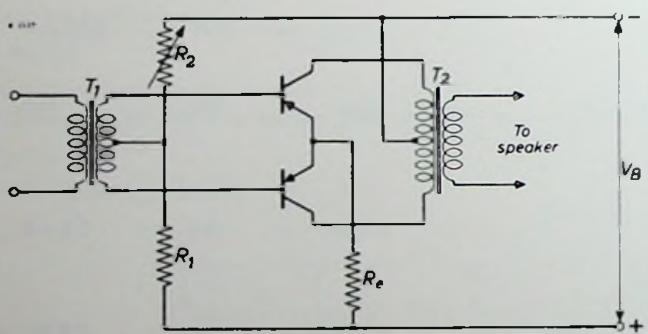


Fig. 4. Output stage with a matched pair of transistors 2-OC 16 in a class B amplifier

Supply voltage (see note 5, p. 133)	V_B =	14	14	7	7	V
Zero signal emitter current	$I_{E1} + I_{E2}$ =	60	60	60	60	mA
Bias resistor [1])	R_1 =	4	4	4	4	Ω
Bias resistor (variable)	R_2 =	300	300	200	200	Ω
Emitter resistor [1])	R_E =	0.8	0.8	0	0	Ω
Transistor output power, two transistors	P_c =	6.7	10	3.2	6.4	W
Output power delivered to the primary of the output transformer, two transistors	P_o =	6.3	9	3.2	6.4	W
Load impedance collector to collector	R_{cc} =	50	33	26	13	Ω
Load impedance per transistor $\left(R_{CE} = \frac{R_{CC}}{4} + R_E\right)$	R_{ce} =	13.4	9	6.4	3.2	Ω

[1]) The d.c. resistance of the driver transformer secondary is assumed to be about 2 × 4 Ω.

At maximum power output:							
peak collector current	I_{cm}	=	1	1.5	1	2	A
d.c. collector current	$-I_C$	=	320	480	320	640	mA
peak driving voltage per transistor	nom. V_{im}	=	1.3	2.0	0.56	1.17	V
	max.[2]) V_{im}	=	1.8	2.8	1.1	3	V
peak driving current per transistor	nom. I_{bm}	=	35	58	35	90	mA
	max.[2]) I_{bm}	=	70	120	70	270	mA
total harmonic distortion	d_{tot}	=	<10	<10[3])	<10	<10[4])	%
At an output power of 50 mW in the primary of the output transformer							
peak driving current per transistor	I_{bm}	=	2.2	2.8	3.2	4.6	mA
total harmonic distortion	d_{tot}	=	2.0	1.5[3])	3.0	2.0[4])	%

OPERATING CHARACTERISTICS OF THE CLASS A AMPLIFYING STAGE OF FIG. 5 AT AN AMBIENT TEMPERATURE OF 25 °C

The circuits below are designed for stable operation up to an ambient temperature of 55 °C. The total thermal resistance from collector junction to ambient temperature is taken to be 4.5 °C/W.

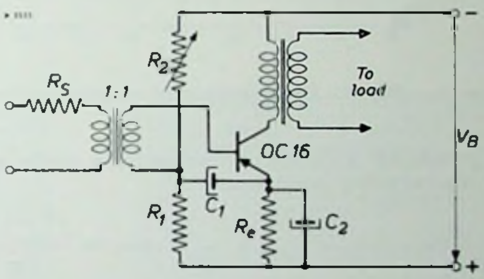


Fig. 5. Class A amplifying stage with a transistor OC16

Supply voltage	V_B[5])	=	14	7	V
D.C. collector current	$-I_C$[6])	=	440	950	mA
Bias resistor	R^1)	=	12	6	Ω
Bias resistor (variable)	R^2)	=	100	50	Ω
Emitter resistor	R_E[6])	=	3	0.8	Ω
Emitter capacitor	C_1	=	500	500	μF
Emitter capacitor	C_2	=	200	1000	μF
Output power in R_L	P_o	=	2.5	2.2[7])	W
Collector load impedance	R_C	=	26	5.5	Ω
At maximum power output:					
peak base current	I_{bm}	=	16	44.5	mA
total harmonic distortion	d_{tot}[8])	=	7	10	%
At an output power of 50 mW in R_L					
peak base current	I_{bm}	=	2.5	5.8	mA
total harmonic distortion	d_{tot}[8])	=	1	2	%

---

[2]) These figures indicate the maxima that can be expected due to the spread in transistor parameters.

[3]) With feedback of approximately a factor 2.

[4]) With feedback of approximately a factor 3.

[5]) These figures are based on the nominal working voltage of 7 V and 14 V of a car accumulator of 6 V and 12 V respectively.

[6]) The d.c. resistance of the driver transformer secondary is assumed to be about 4 Ω.

[7]) Due to the non-linearity of the α' curve, a maximum output power of 2.2 W can be obtained without exceeding a distortion of 10 %. If power feedback is applied the maximum output power for 10 % distortion increases to about 2.5 W.

[8]) Distortion measurements are made with R_s = 30 Ω.

APPLICATION CONSIDERATIONS

1. The mounting base (bottom) temperature of the transistor, which is difficult to measure when the transistor is mounted on a chassis, can be determined by measuring the stud temperature by attaching a thermocouple to the stud and taking into account a thermal resistance between mounting base and stud of 0.2 °C/W (transistor mounted with the two mica washers and the brass washer of Fig. 1).
   In Fig. 6 the maximum permissible stud temperature is shown as a function of the collector dissipation.

Fig. 6. Permissible stud temperature versus collector dissipation

2. The permissible collector dissipation at a given ambient temperature can be derived from the following expression:

$$W_{c\,max} = \frac{75 - t_{amb}}{K}$$

The thermal resistance K can be divided into several parts, which is illustrated in Fig. 7.

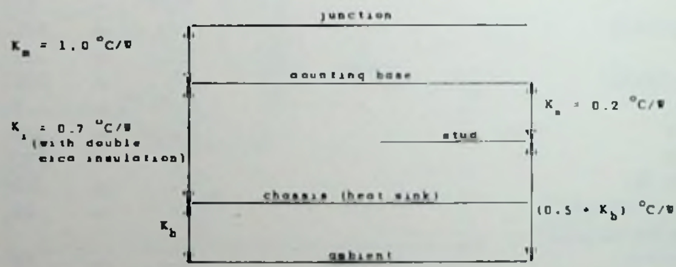


Fig. 7. The various thermal resistances. It is assumed that the transistor is mounted with the two mica washers and the brass washers of Fig. 1.
$K = K_m + K_i + K_h$

The value of K_h depends on the cooling conditions under which the transistors are used (dimensions and position of the heat sink, etc.); K_h can easily be controlled by measuring the stud temperature at a certain ambient temperature and collector dissipation.

$$t_{stud} = t_{amb} + (K_h + 0.5)\, W_c$$

In the following example the temperatures which occur at the various points at a given ambient temperature of e.g. 50 °C and a collector dissipation of e.g. 4 W are calculated with the aid of the figures in Fig. 7; K_h is assumed to be 4.55 °C/W.

71 + 1.0 × 4 = 75 °C — junction

68.2 + 0.7 × 4 = 71 °C — mounting base

stud — 50 + (4.55 + 0.5) 4 = 70.2 °C

50 + 4.55 × 4 = 68.2 °C — chassis (heat sink)

50 °C — ambient — 50 °C

Fig. 8. Temperature scale for W_c = 4 W. t_{amb} = 50 °C.

In Fig. 9 the maximum permissible collector dissipation is shown as a function of the ambient temperature for different values of K_h.

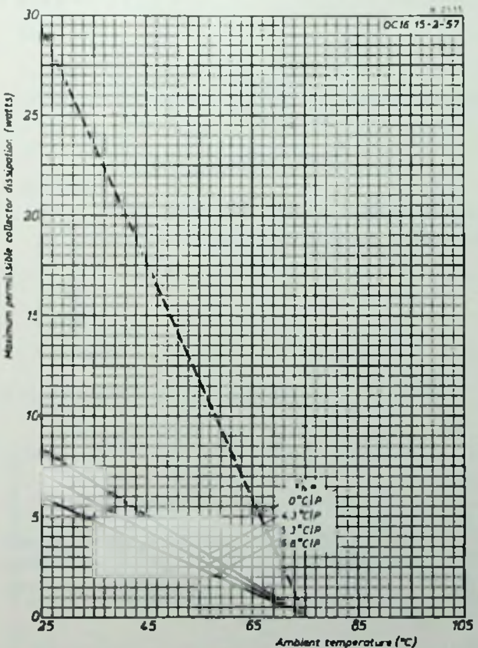


Fig. 9. $W_{c\ max}$ as a function of t_{amb} and K_h.

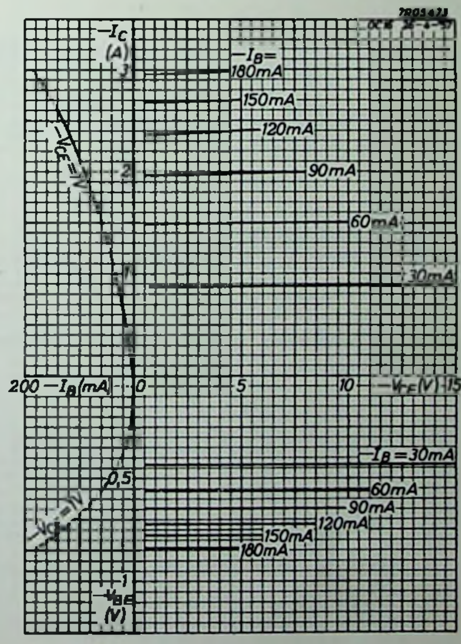


Fig. 10. Typical characteristics of the OC 16 in common emitter connection at a junction temperature of 45° C.

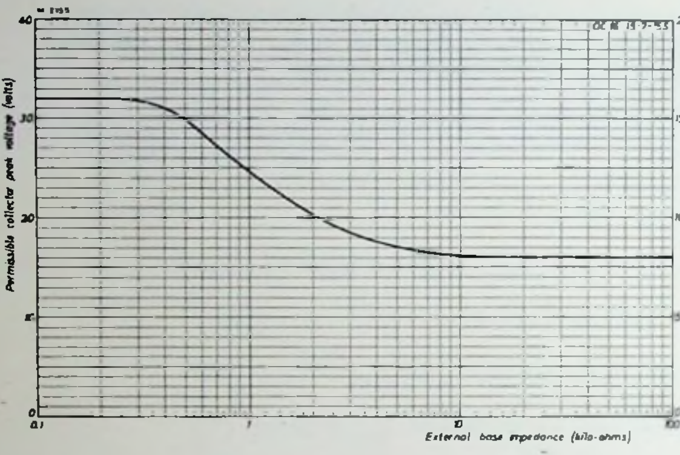


Fig. 11. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the external impedance between base and emitter. Provisions must be made to ensure thermal stability.

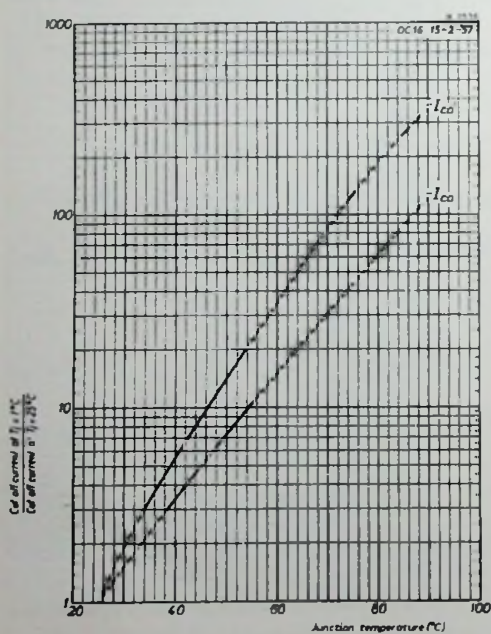


Fig. 12. Collector cut-off current in common base (I_{co}) and in common emitter (I_{co}') as a function of the junction temperature.

P-N-P GERMANIUM TRANSISTOR OC 44

The OC 44 is a high-frequency germanium junction transistor of the P-N-P alloy type, in all-glass construction, especially suitable for use in convertors and mixer-oscillator circuits.
The transistors are hermetically sealed and absolutely moisture-proof. They may be soldered in or clipped for plug-in construction.

MECHANICAL DATA

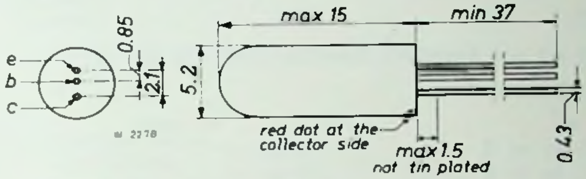


Fig. 1. Dimensional outline in mm and electrode connections.

THERMAL DATA

Junction temperature rise to ambient temperature, transistor in free air	K	=	max. 0.6 °C/mW

ABSOLUTE MAXIMUM RATINGS

Temperature

Storage	T_s	=	min. −55 °C max. 75 °C
Junction, continuous operation	T_j	=	max. 75 °C
intermittent operation, total duration max. 200 h	T_j	=	max. 90 °C [1]

Collector

Voltage

Peak, base reference (emitter reference see Fig. 8)	$-V_{CBM}$	= max.	15 V
Average, base reference (emitter reference see Fig. 8)	$-V_{CB}$	= max.	15 V

Current

Peak	$-I_{CM}$	= max.	10 mA
Average	$-I_C$	= max.	5 mA

Dissipation (in free air)

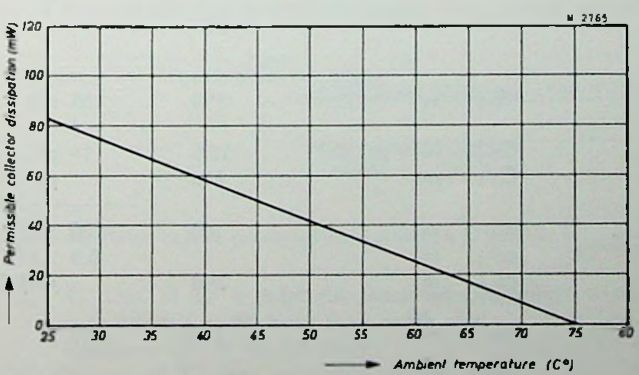


Fig. 2. Maximum continuous collector dissipation as a function of the ambient temperature.

[1]) Likelihood of full performance at this temperature is also dependent on the type of application.

E m i t t e r

Voltage

Peak, base reference	$-V_{EBM}$	= max.	12 V
Average, base reference	$-V_{EB}$	= max.	12 V

Current

Peak	I_{EM}	= max.	10 mA
Average	I_E	= max.	5 mA

CHARACTERISTICS AT AN AMBIENT TEMPERATURE OF 25 °C

C o m m o n B a s e C i r c u i t

			min.	average	max.
Collector current at					
$-V_{CB}$ = 2 V; I_E = 0	$-I_{CBO}$	=		0.5	2.0 μA
$-V_{CB}$ = 15 V; I_E = 0	$-I_{CBO}$	=			10 μA
Emitter current at					
$-V_{EB}$ = 2 V; I_C = 0	$-I_{EBO}$	=		0.4	2.0 μA
$-V_{EB}$ = 12 V; I_C = 0	$-I_{EBO}$	=			40 μA
Current amplification frequency cut-off measured at					
$-V_{CB}$ = 6 V; I_E = 1 mA	$f_{\alpha b}$	=	7.5	15	30 Mc/s

C o m m o n E m i t t e r C i r c u i t

			min.	average	max.
Collector current at					
$-V_{CE}$ = 2 V; I_B = 0	$-I_{CEO}$	=		25	75 μA
Current amplification factor, output short-circuited, measured at					
$-V_{CE}$ = 6 V; I_E = 1 mA f = 1000 c/s	α_{FE}	=	45	100	225

Small-signal hybrid π parameters:

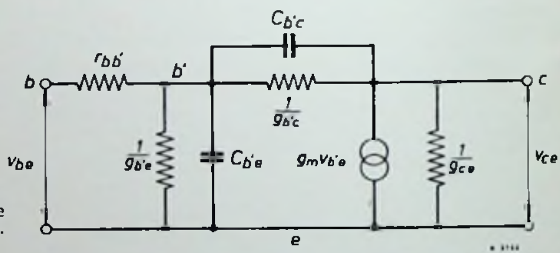


Fig. 3. Hybrid π equivalent circuit. The internal base connection is indicated by b'.

Measured at $-V_{CE}$ = 6 V; I_E = 1 mA

			min.	average	max.
Base input voltage	$-V_{BE}$	=	125	150	185 mV
Capacitance					
between points b' and c	$C_{b'c}$	=	7	10.5	14 pF
between points b' and e	$C_{b'e}$	=		410	pF
Conductance					
between points c and e	g_{ce}	=		40	100 μA/V
between points b' and c	$g_{b'c}$	=			0.5 μA/V
between points b' and e	$g_{b'e}$	=		390	μA/V
Resistance					
between points b and b'	$r_{bb'}$	=		110	250 Ω
quotient of r_{bb} and f_a	$\frac{r_{bb'}}{f_{\alpha b}}$	=	3.5	7.3	20 Ωs/Mc
Intrinsic transconductance	g_m	=		39	μA/V

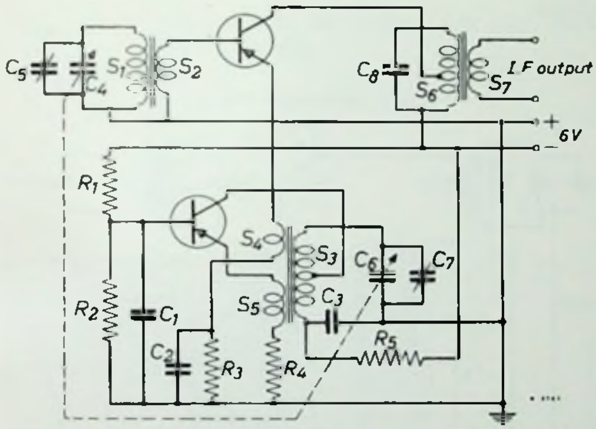
TYPICAL OPERATION AT AN INPUT SIGNAL FREQUENCY OF 1 Mc/s IN THE MIXER- AND OSCILLATOR CIRCUIT OF FIG. 4 AT AN AMBIENT TEMPERATURE OF 25 °C

Mixing transistor

D.C. collector-to-emitter voltage	$-V_{CE}$ =	5.8 V
D.C. emitter current	I_E =	0.4 mA
Oscillator injection voltage between emitter and earth	V_{osc} =	0.3 V (rms)
H.F. input damping (approx.)		0.3 mA/V
I.F. output damping (approx.)		17 μA/V
Conversion gain [1]) (approx.)	P_o/P_i =	28 dB

Oscillator transistor

D.C. collector-to-emitter voltage	$-V_{CE}$ =	4.8 V
D.C. emitter current	I_E =	0.3 mA
Oscillator voltage across tuned circuit		5.7 V (rms)
Oscillator voltage collector-to-base		3.5 V (rms)
Oscillator voltage emitter-to-base		0.3 V (rms)



R_1 =	22 kΩ
R_2 =	330 Ω
R_3 =	560 Ω
R_4 =	1.8 kΩ
R_5 =	2.2 kΩ
C_1 =	22000 pF
C_2 =	0.1 μF
C_3 =	0.1 μF
C_4 =	7 − 190 pF
C_5 =	2 − 10 pF
C_6 =	6 − 110 pF
C_7 =	2 − 10 pF
C_8 =	220 pF

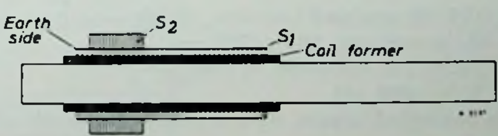
Fig. 4. Mixer-oscillator circuit with two transistors OC 44.

Circuit data

Antenna coil

S_1 77 turns of 32 × 0.04 silk insulated Litz wire, closely wound on a former, diameter 12 mm.
Rod: ferroxcube 4B; dimensions: 10 × 200 mm.
Unloaded Q at 1 Mc/s: 150 (mounted in chassis)
L: 480 μH

S_2 7 turns of 0.3 mm enamelled copper wire, closely wound at the earth side of S_1.



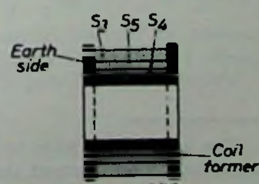
Oscillator coil

The oscillator coil is mounted in a potcore D 18/12, ferroxcube 3B3, air gap 0.5 mm.

S_3 43 turns of 32 × 0.04 silk insulated Litz wire, collector tap at 26 turns from earth side. Unloaded Q: 45 at 1.5 Mc/s

S_4 2 turns of 0.3 mm enamelled copper wire.

S_5 8 turns of 0.3 mm enamelled copper wire.



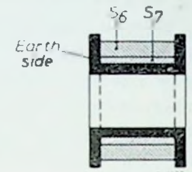
See footnote on page 140.

I.F. transformer

The I.F. transformer is mounted in a potcore D 18/12, ferroxcube 3B3, air gap 0.3 mm.

S_6 65 turns of 16 × 0.04 silk insulated Litz wire, collector tap at 52 turns from earth side.
Unloaded Q: 110.

S_7 3 turns of 0.3 mm enamelled copper wire.



TYPICAL OPERATION AT AN INPUT SIGNAL FREQUENCY OF 1 Mc/s IN THE SELF-EXCITING CONVERTOR OF FIG. 5 AT AN AMBIENT TEMPERATURE OF 25 °C

D.C. collector-to-emitter voltage	$-V_{CE}$	=	5.1 V
D.C. emitter current	I_E	=	0.4 mA
Oscillator voltage emitter-to-earth	V_{osc}	=	50 mV
Oscillator voltage collector-to-earth			0.12 V
Oscillator voltage across the tuned circuit			2 V
H.F. input damping (approx.)	g_i	=	0.5 mA/V
I.F. output damping (approx.)	g_o	=	17 μA/V
Conversion gain [1] (approx.)	P_o/P_i	=	28 dB

R_1 = 10 kΩ
R_2 = 2.2 kΩ
R_3 = 2.2 kΩ
C_1 = 0.1 μF
C_2 = 47000 pF
C_3 = 7 − 190 pF
C_4 = 2 − 10 pF
C_5 = 6 − 110 pF
C_6 = 2 − 10 pF
C_7 = 220 pF

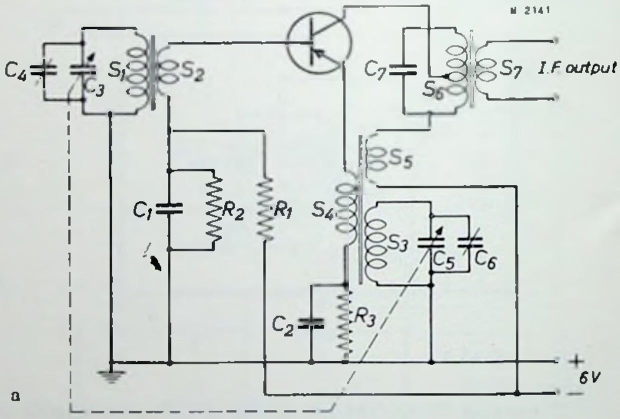


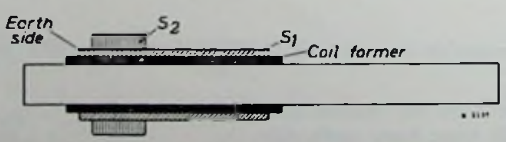
Fig. 5. Self-exciting convertor with a transistor OC 44.

Circuit data

Antenna coil

S_1 77 turns of 32 × 0.04 silk insulated Litz wire, closely wound on a former, diameter 12 mm.
Rod: ferroxcube 4B, dimensions: 10 × 200 mm.
Unloaded Q at 1 Mc/s: 150 (mounted in chassis)
L : 480 μH

S_2 5 turns of 0.3 mm enamelled copper wire, wound at the earth side of S_1.

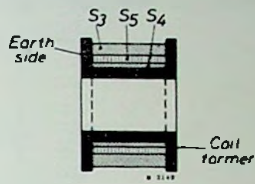


[1] The conversion gain is defined at the ratio between the I.F. power in a 680 Ω load resistor connected to the output terminals of the I.F. filter (680 Ω being the average value of the input resistance of an OC 45) and the available H.F. power in the antenna circuit.

Oscillator coil

The oscillator coil is mounted in a potcore D 18/12, ferroxcube 3B3, airgap 1 mm.

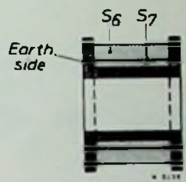
S_3 54 turns of 32 × 0.04 silk insulated Litz wire.
Unloaded Q: 55 at 1.5 Mc/s
S_4 2 turns of 0.3 mm enamelled copper wire.
S_5 5 turns of 0.3 mm enamelled copper wire.



I.F. transformer

The I.F. transformer is mounted in a potcore D 18/12, ferroxcube 3B3, airgap 0.3 mm.

S_6 65 turns of 16 × 0.04 silk insulated Litz wire, collector tap at 52 turns from earth side.
Unloaded Q: 110.
S_7 3 turns of 0.3 mm enamelled copper wire.



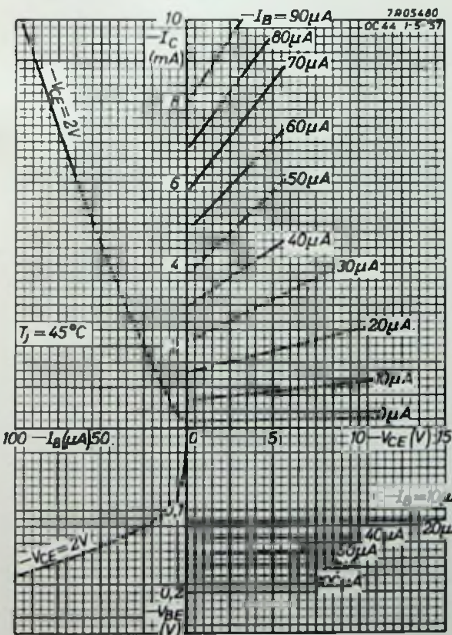


Fig. 6. Typical characteristics of the OC 44 in common emitter connection at a junction temperature of 45 °C.

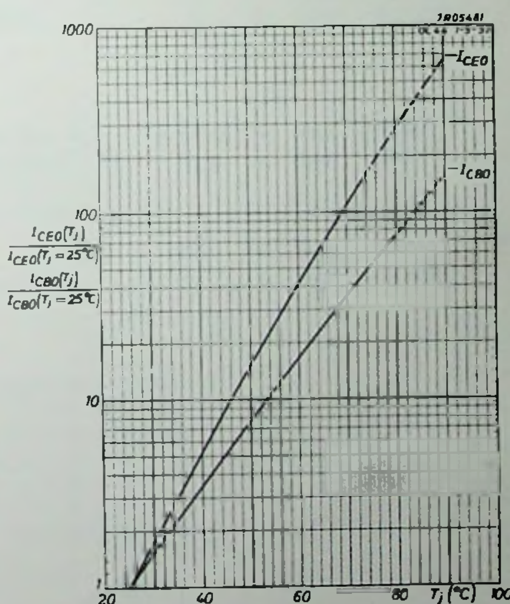


Fig. 7. Collector cut-off current in common base (I_{CBO}) and in common emitter (I_{CEO}) as a function of the junction temperature.

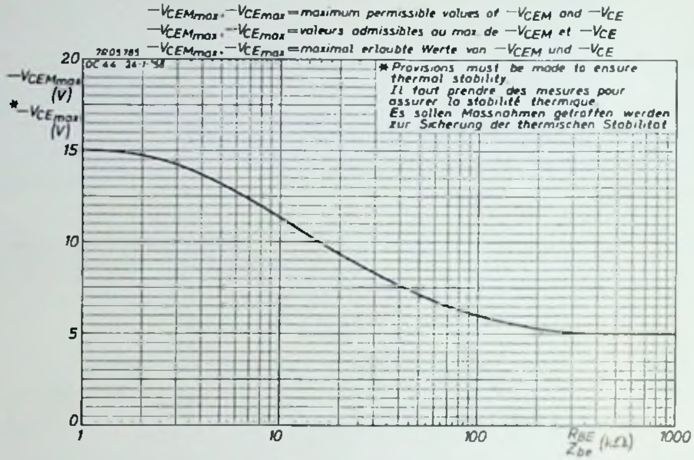


Fig. 8. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the impedance between base and emitter. Provisions must be made to ensure thermal stability.

P-N-P GERMANIUM TRANSISTOR OC 45

The OC 45 is a high-frequency germanium junction transistor of the P-N-P alloy type, in all-glass construction, especially suitable for use at intermediate frequencies.
The transistors are hermetically sealed and absolutely moisture-proof. They may be soldered in or clipped for plug-in connection.

MECHANICAL DATA

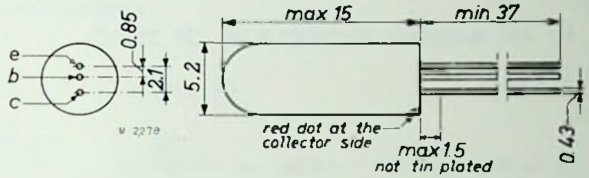


Fig. 1. Dimensional outline in mm and electrode connections.

THERMAL DATA

Junction temperature rise to ambient temperature, transistor in free air,	K	=	max.	0.6 °C/mW

ABSOLUTE MAXIMUM RATINGS

Temperature

Storage temperature	T_s	=	min.	−55 °C
			max.	75 °C
Junction, continuous operation	T_j	=	max.	75 °C
Intermittent operation, total duration max. 200 h	T_j	=	max.	90 °C [1]

Collector

Voltage

peak, base reference (emitter reference see Fig. 6)	$-V_{CBM}$	=	max.	15 V
average, base reference (emitter reference see Fig. 6)	$-V_{CB}$	=	max.	10 V

Current

peak	$-I_{CM}$	=	max.	10 mA
average	$-I_C$	=	max.	5 mA

Dissipation (in free air)

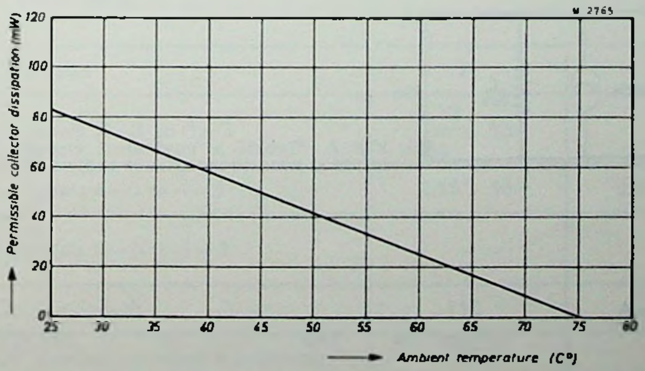


Fig. 2. Maximum continuous collector dissipation as a function of the ambient temperature.

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[1]) Likelihood of full performance at this temperature is also dependent on the type of application.

Emitter

Voltage

peak, base reference	$-V_{EBM}$	=	max.	12 V
average, base reference	$-V_{EB}$	=	max.	8 V

Current

peak	I_{EM}	=	max.	10 mA
average	I_E	=	max.	5 mA

CHARACTERISTICS AT AN AMBIENT TEMPERATURE OF 25 °C

Common Base Circuit

			min.	average	max.
Collector current at					
$-V_{CB}$ = 2 V; I_E = 0	$-I_{CBO}$	=		0.5	2.0 μA
$-V_{CE}$ = 15 V; I_E = 0	$-I_{CBO}$	=			10 μA
Emitter current at					
$-V_{EB}$ = 2 V; I_C = 0	$-I_{EO}$	=		0.4	2.0 μA
$-V_{EB}$ = 12 V; I_C = 0	$-I_{EO}$	=			40 μA
Current amplification frequency cut-off measured at $-V_{CB}$ = 6 V; I_E = 1 mA	f_{α_b}	=	3	6	12 Mc/s

Common Emitter Circuit

			min.	average	max.
Collector current at $-V_{CE}$ = 2 V; I_B = 0	$-I_{CEO}$	=		12	40 μA
Current amplification factor, output short-circuited, measured at $-V_{CE}$ = 6 V; I_E = 1 mA; f = 1000 c/s	α_{FE}	=	25	50	125

Small-signal hybrid π parameters (see Fig. 3)

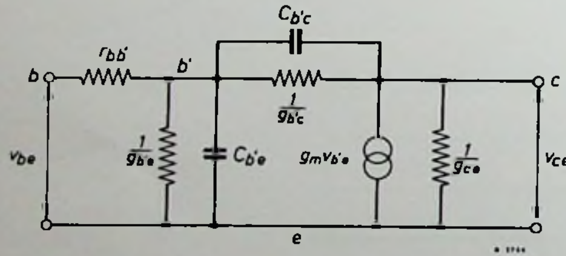

Fig. 3. Hybrid π equivalent circuit. The internal base connection is indicated by b'.

Measured at

$-V_{CE}$ = 6 V; I_E = 1 mA

			min.	average	max.
Base input voltage	$-V_{BE}$	=	145	170	195 mV
Capacitance					
between points b' and c	$C_{b'c}$	=	7	10.5	14 pF
between points b' and e	$C_{b'e}$	=		1000	pF

			min.	average	max.
Conductance					
between points c and e	g_{ce}	=		15	40 μA/V
between points b′ and c	$g_{b'c}$	=			0.5 μA/V
between points b′ and e	$g_{b'e}$	=		760	μA/V
Resistance					
between points b and b′	$r_{bb'}$	=		75	200 Ω
quotient of $r_{bb'}$ and $f_{\alpha b}$	$\dfrac{r_{bb'}}{f_{\alpha b}}$	=	5	12.5	30 Ωs/Mc
Intrinsic transconductance	g_m	=		39	μA/V

CHARACTERISTICS AS I.F. AMPLIFIER

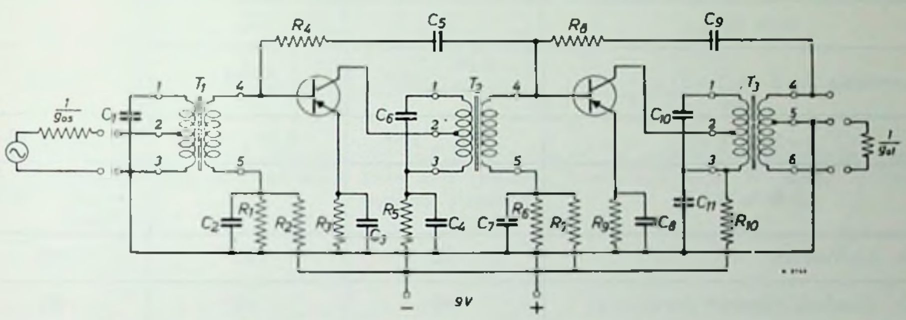

Fig. 4. Basic circuit of a two-stage I.F. amplifier. The design of this circuit provides stable operation and interchangeability of the transistors.

CIRCUIT DATA

A. For a selectivity factor of 5.0 at a detuning of 9 kc/s

R_1, R_6	=	2.7 kΩ
R_2, R_7	=	12 kΩ
R_3, R_5, R_9, R_{10}	=	1.5 kΩ
R_{14}	=	1 kΩ (5 %)
R_8	=	1 kΩ (5 %)
g_{os}	=	source admittance
g_{ol}	=	load admittance
C_1, C_6, C_{10}	=	220 pF
C_2, C_3, C_4, C_7, C_8, C_{11}	=	0.1 μF
C_5, C_9	=	56 pF (5 %)

Transformers

Turn ratio	T_1	T_2	T_3
Terminals 2–3 to 1–3	$\dfrac{3.12}{\sqrt{g_{os}}} \cdot 10^{-3}$	0.35	0.256
Terminals 4–5 to 1–3	$4.34 \cdot 10^{-2}$	$5.92 \cdot 10^{-2}$	$4.34 \cdot 10^{-2}$
Terminals 5–6 to 1–3	–	–	$\dfrac{3.12}{\sqrt{g_{ol}}} \cdot 10^{-3}$
Q_0 (unloaded)	110	70	110
Q_L (loaded, nominal transistors)	35	35	35
Core material	ferroxcube	ferroxcube	ferroxcube

Performance

Frequency	f	=	450 kc/s
Collector bias voltage	$-V_{CE}$	=	6 V
Collector bias current	$-I_C$	=	1 mA
Ambient temperature	T_{amb}	=	25 °C
Power gain	P_o/P_i	=	60 dB
Selectivity factor	S_9	=	5.0

B. **For a selectivity of 40 at a detuning of 9 kc/s**

Resistor and capacitor as under A, except $C_1 = C_6 = C_{10} = 390$ pF

Transformers

turn ratio	T_1	T_2	T_3
Terminals 2 – 3 to 1 – 3	$\frac{2.25}{\sqrt{g_{os}}} \cdot 10^{-3}$	0.307	0.225
4 – 5 to 1 – 3	$3.82 \cdot 10^{-2}$	$5.21 \cdot 10^{-2}$	$3.82 \cdot 10^{-2}$
5 – 6 to 1 – 3			$\frac{2.25}{\sqrt{g_{ol}}} \cdot 10^{-3}$
Q_o (unloaded)	160	160	160
Q_l (loaded, nominal transistors)	80	80	80
Core material	ferroxcube	ferroxcube	ferroxcube

Performance

Frequency	f	=	450 kc/s
Collector bias voltage	$-V_{CE}$	=	6 V
Collector bias current	$-I_C$	=	1 mA
Ambient temperature	T_{amb}	=	25 °C
Power gain	P_o/P_i	=	57 dB
Selectivity factor	S_9	=	40

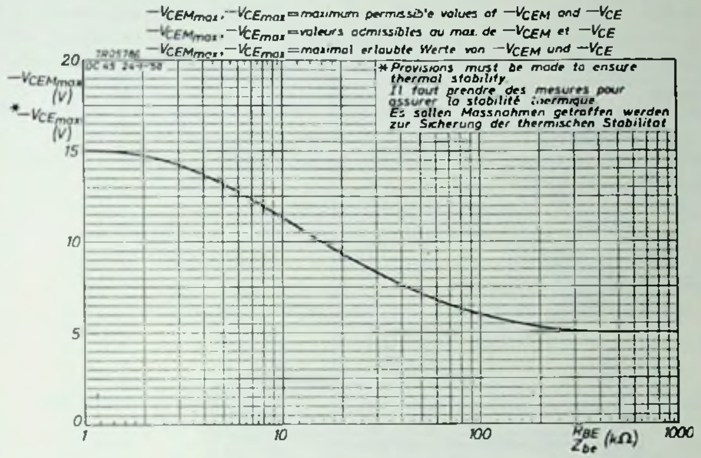

Fig. 6. The maximum permissible average and peak value of the collector-to-emitter voltage as a function of the resistance between base and emitter. Provisions must be made to ensure thermal stability.

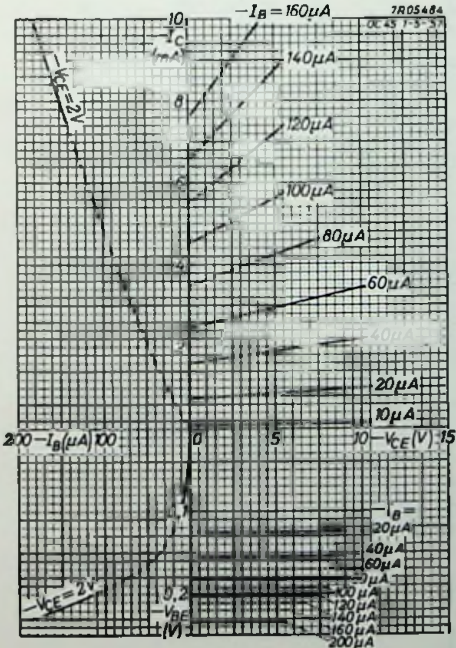

Fig. 5. Typical characteristics of the OC 45, in common emitter connection at a junction temperature of 45 °C.

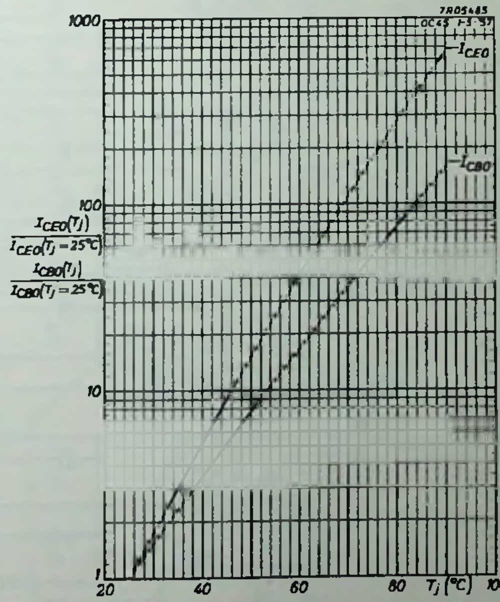

Fig. 7. Collector cut-off current in common base (I_{CBO}) and in common emitter (I_{CEO}) as a function of the junction temperature.

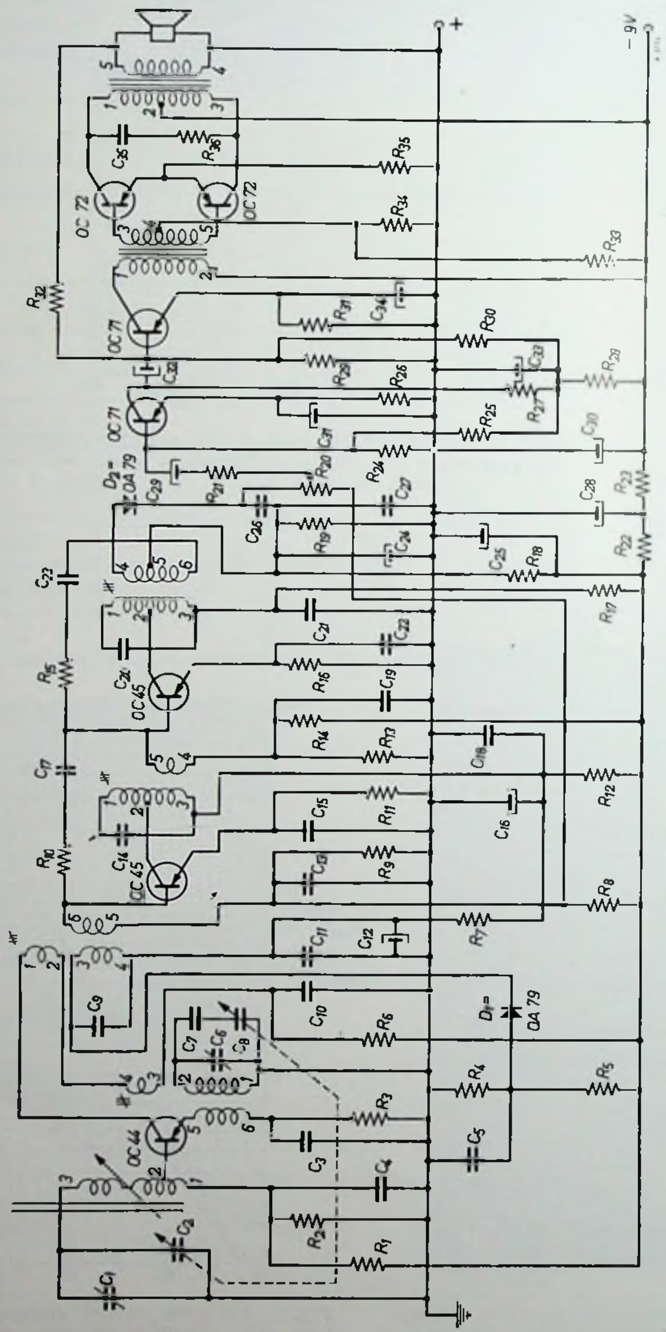

Fig. 8. Circuit diagram of an experimental all-transistor superhet receiver.

To illustrate the possibilities offered by the RF transistors OC44 and OC 45, Fig. 8 shows the circuit of a superheterodyne receiver for the medium-wave range, in which the typical circuits given in the Tentative Data of these transistors have been combined to a complete set. This receiver, which is fully stabilised, contains a new A.G.C. circuit and is fed from a 9 V supply battery.
In the RF part an OC 44 transistor operates as a self-oscillating mixer. The IF amplifier, equipped with two OC 45 transistors, operates at 450 Mc/s. The detector D_2 is an OA 79 diode. In this receiver the following AGC circuit is used. The direct current of the diode D_2, being amplified by the first OC 45, gives a voltage variation at the collector resistor of this transistor. This voltage variation is fed to the diode D_1, (OA 79), thus affecting the inverse bias voltage of D_1, which is in parallel with the first IF tuned circuit. In this way an increase of the RF signal will result in an increase of the inverse conductance of D_1 (damping on the first IF-tuned circuit). The AF part is quite conventional, and is equipped with a pre-amplifier and a driver stage with an OC 71 transistor and a push-pull class-B output stage with a matched pair of transistors 2-OC 72.

LIST OF COMPONENTS (Fig. 8)

R_1 = 10 kΩ	R_{25} = 56 kΩ	C_{12} = 3 μF
R_2 = 2.2 kΩ	R_{26} = 1 kΩ	C_{13} = 100 kpF
R_3 = 2.2 kΩ	R_{27} = 5.6 kΩ	C_{14} = 390 pF
R_4 = 12 kΩ ± 5 %	R_{28} = 150 Ω	C_{15} = 47 kpF
R_5 = 3.9 kΩ ± 5 %	R_{29} = 15 kΩ	C_{16} = 3 μF
R_6 = 6.8 kΩ	R_{30} = 12 kΩ	C_{17} = 56 pF
R_7 = 470 Ω	R_{31} = 1.5 kΩ	C_{18} = 47 kpF
R_8 = 27 kΩ ± 5 %	R_{32} = 33 kΩ	C_{19} = 47 kpF
R_9 = 2.2 kΩ ± 5 %	R_{33} = 4.7 kΩ	C_{20} = 330 pF
R_{10} = 1 kΩ	R_{34} = 100 Ω	C_{21} = 47 kpF
R_{11} = 470 Ω ± 5 %	R_{35} = 15 Ω	C_{22} = 47 kpF
R_{12} = 2.7 kΩ ± 5 %	R_{36} = 220 Ω	C_{23} = 56 pF
R_{13} = 2.7 kΩ		C_{24} = 25 μF
R_{14} = 12 kΩ	C_1 = 27 pF	C_{25} = 100 μF
R_{15} = 1 kΩ	C_2 = 499 pF	C_{26} = 1500 pF
R_{16} = 1.5 kΩ	C_3 = 47 kpF	C_{27} = 47 kpF
R_{17} = 1.5 kΩ	C_4 = 47 kpF	C_{28} = 100 μF
R_{18} = 39 kΩ	C_5 = 47 kpF	C_{29} = 10 μF
R_{19} = 3.3 kΩ	C_6 = 27 pF	C_{30} = 100 μF
R_{20} = 50 kΩ p. meter	C_7 = 470 pF	C_{31} = 100 μF
R_{21} = 8.2 kΩ	C_8 = 499 pF	C_{32} = 10 μF
R_{22} = 47 Ω	C_9 = 220 pF	C_{33} = 100 μF
R_{23} = 47 Ω	C_{10} = 47 kpF	C_{34} = 100 μF
R_{24} = 12 kΩ	C_{11} = 47 kpF	C_{35} = 56 kpF

Antenna Coil

terminals 1 – 3 .... 2 × 24 turns
terminals 1 – 2 .... 5 turns

Oscillator transformer

Selfinductance L_{1-2} = 103 μH
Turns ratios:
terminals 3 – 4 to 1 – 2 .... 0.0714
5 – 6 to 1 – 2 .... 0.0429

First I.F. transformer

Selfinductance L_{3-4} = 561 μH
Turns ratios:
terminals 1 – 2 to 3 – 4 .... 0.8064
5 – 6 to 3 – 4 .... 0.0387

Second I.F. transformer

Selfinductance L_{1-3} = 317 μH

Turns ratios:

terminals 2 – 3 to 1 – 3	0.3095
4 – 5 to 1 – 3	0.05

Third I.F. transformer

Selfinductance L_{1-3} = 374 μH

Turns ratios:

terminals 2 – 3 to 1 – 3	0.225
4 – 5 to 1 – 3	0.05
5 – 6 to 1 – 3	0.45

The design of the AF amplifier, including the transformers, is equivalent to the circuit given in the data of the OC 72.

GERMANIUM PHOTO TRANSISTOR OCP 71

The OCP 71 is a general-purpose germanium phototransistor of the P-N-P alloy type, in all-glass construction. This type has an extremely high sensitivity, and it will find immediate use in those applications where its rapid response and its ability to operate relays direct offer great advantages. The maximum permissible collector current of 10 mA is more than sufficient to operate a semi-robust relay.

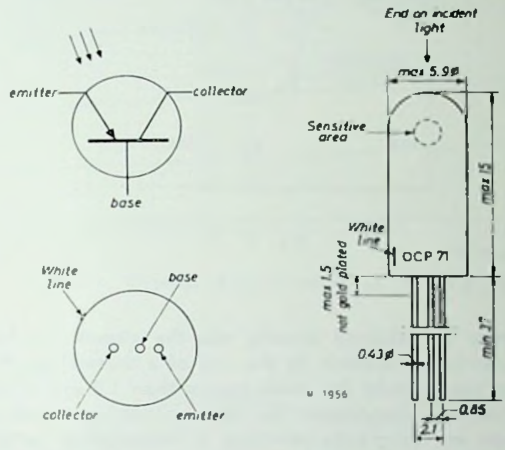

Fig. 1. Dimensional drawing (dimensions in mm) and electrode connections of the OCP 71. The preferred direction of incident light is perpendicular to the plane of the leads and is on the side of the bulb bearing the type number.

ABSOLUTE MAXIMUM RATINGS AT AMBIENT TEMPERATURE OF 45 °C

Collector		
Voltage		
Peak	$-V_{CEM}$	= max. 25 V
Average	$-V_{CE}$	= max. 25 V
Current		
Peak	$-I_{CM}$	= max. 10 mA
Average	$-I_C$	= max. 10 mA
Dissipation	P_C	= max. 25 mW
Temperature		
Junction temperature rise	K	= 0.4 °C/mW
Storage	T_s	= max. 55 °C

CHARACTERISTICS AT AMBIENT TEMPERATURE OF 25 °C

Dark current measured at $-V_{CE} = 10$ V; $I_B = 0$	$-I_{CO}'$	= max. 300 μA
Cut-off frequency (modulated light, Fig. 2)	f_{CO}	= 3 kc/s

SENSITIVITY (Colour temperature of the light source 2700 °K)

Measured at $-V_{CE} = 2$ V	
and uniform illumination of 75 ft. candle (= 807 lux)	
with preferred direction of incident light (see outline drawing)	1.5 to 4 mA
with end-on (see outline drawing)	0.5 to 1.3 mA
With sensitive area of 7 mm²	0.3 A/lumen

SPECTRAL RESPONSE

Peak	1.55 μ
Cut-off	2 μ

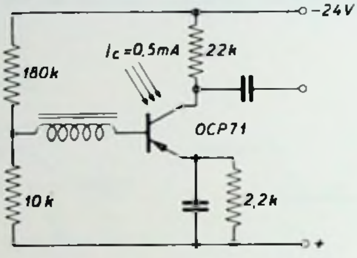

Fig. 2

OPERATING NOTES

1. The phototransistor may be soldered directly into the circuit, but heat conducted to the junction should be kept to a minimum by the use of a thermal shunt.
2. Care should be taken not to bend the leads nearer than 1.5 mm to the seal.
3. When used with a constant illumination the base connection is left open circuit. In such service phototransistors are inherently sensitive to temperature variations which result in variations of the output current which cannot be distinguished from the light signal. Designers must take this factor into consideration and must be conservative with regard to dissipations and operating temperature.

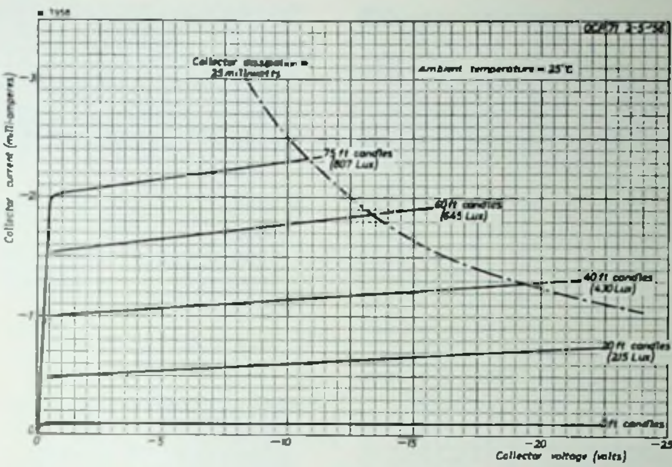

Fig. 3. Typical characteristics of the OCP 71.

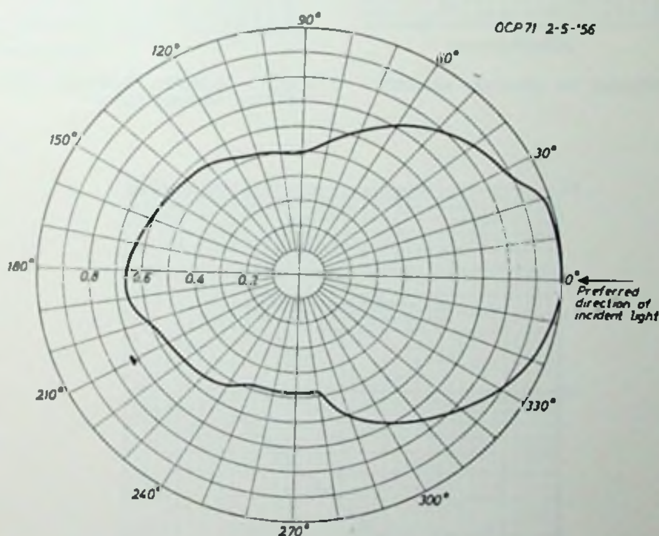

Fig. 4. Relation between the sensitivity and direction of incident light.

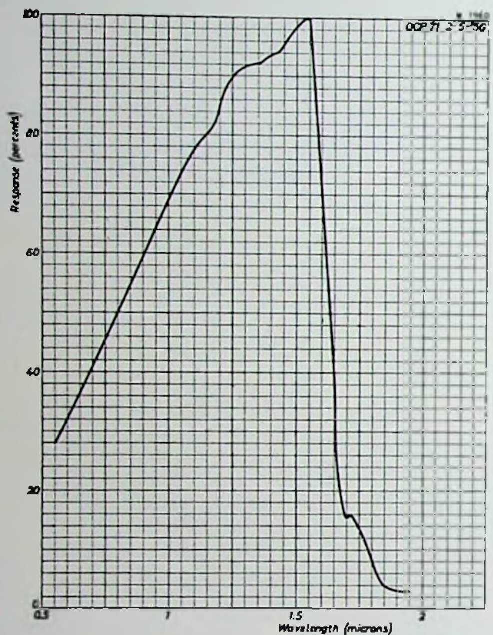

Fig. 5. Spectral response to constant energy spectrum.

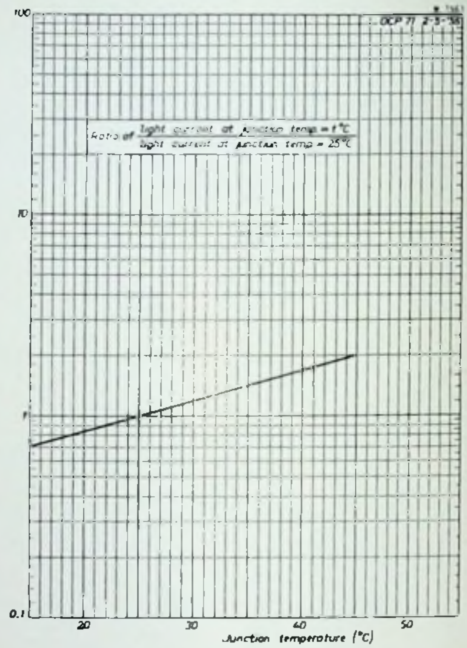

Fig. 6. Relation between light current and junction temperature.

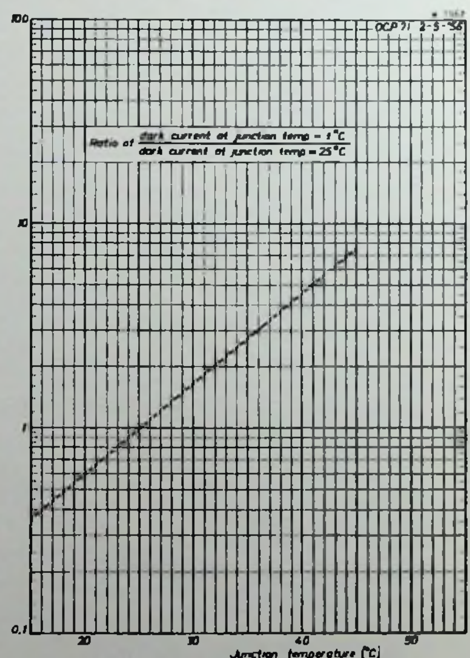

Fig. 7. Relation between dark current and junction temperature.