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Integrated linear basic circuits

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Monolithic integrated circuits, which have developed in a matter of ten years from a laboratory experiment into a mass-produced product, can be divided into two main classes, digital and linear. The linear devices can fulfil many functions, but they always contain the same basic circuits. The authors present some elegant solutions for these linear basic circuits, making use of the special capabilities of integrated-circuit technology, in what amounts to a new departure in electronics.

Introduction

The unsuspecting layman viewing an integrated circuit for the first time under a microscope might well believe that he is looking at a piece of modern art. The abstract play of patches and lines, often beautifully coloured, does not suggest a deliberate pattern of shapes designed to replace a whole board packed with resistors and transistors.

It is not surprising that many people believe integration technology to be capable of making the impossible possible. Admiration for an imperfectly understood technology may lead people to overestimate its capabilities. It should be remembered, however, that integration technology is only one of the methods of manufacturing electronic circuits. Like any other method, it has its advantages and disadvantages, and a finished product of high quality will not be produced unless the electronic engineer's design is able to exploit the one and avoid the other.

In this article we shall describe some examples of linear basic circuits which take advantage of the particular capabilities offered by integrated-circuit technology. To appreciate the beauty of these circuits we must first, however, take a closer look at the features of the new technology.

Integration techniques

In recent decades constant efforts have been made to produce electronic circuits more efficiently than can be done by making wire connections between individual components [1].

The first step was the introduction of printed circuits, the printed wiring pattern being applied to an insulating base, usually a resin-bonded paper board, by a photo-etching process. A second step was the advent of the "hybrid" circuit [2], in which not only the wiring but also the resistors and smaller capacitors are evaporated on to a glass or ceramic substrate. The active devices and other components are soldered on later. The circuits are called "hybrid" because of the combination of individual components with vacuum-evaporated ones; there is integration to the extent that some of the components are fabricated as a whole.

Both of these techniques obviously allow complete freedom in the choice of the active devices. There is no reason why *NPN*, *PNP*, field-effect and MOS transistors should not be included side by side in the same circuit. The introduction of these technologies therefore did not radically alter the work of the circuit designer.

A complete change was brought about, however, by the monolithic or solid circuit [3], which is usually what is intended by the term "integrated circuit". Here all the circuit elements, both active and passive, are

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[1] P. W. Haaijman, Integration of electronic circuits, Philips tech. Rev. 27, 180-181, 1966.
[2] E. C. Munk and A. Rademakers, Integrated circuits with evaporated thin films, Philips tech. Rev. 27, 182-191, 1966.
[3] A. Schmitz, Solid circuits, Philips tech. Rev. 27, 192-199, 1966.

formed at the same time in a thin layer of a silicon wafer by the "planar" technique, in a succession of oxidizing, photo-etching and diffusion processes. Finally the elements are interconnected by means of an evaporated pattern of conductor. The fact that all the elements are formed in the same steps of the process makes them interdependent. For example, the starting material and the individual steps could be chosen so as to produce optimum *NPN* transistors. But generally speaking this choice would then not be optimum for the other elements. The introduction of the monolithic integration technique therefore demanded a different approach from the circuit designer; the planar technique has its own special advantages, but it also has its limitations.

Capabilities and limitations of the planar technique; a new electronics

Advantages of the planar technique

In every circuit the external contacts are possible sources of undesirable effects, and the reliability of a circuit generally decreases with an increase in the number of contacts. Complicated circuits are more reliable in integrated form than when they are built up from separate transistors, since the total number of contacts is then much smaller than the total number of contacts of all the individual transistors. This greater reliability is the important factor that has led to growing interest in the building of systems from standard integrated-circuit units.

Provided the masks for the photo-etching operations in the planar technique are accurately drawn and the various manufacturing steps are carried out with scrupulous care, it is possible to produce almost identical transistors in an integrated circuit. The variation of base-emitter voltage V_{BE} with collector current I_C can be reproduced fairly easily ($\Delta V_{BE}/V_{BE} < 1\%$), but less success is achieved with the current-gain factor h_{FE} ($\Delta h_{FE}/h_{FE} < 10\%$).

The reason for this is that, as can be seen from the relation

$$I_C = I_{C0}\,(e^{eV_{BE}/kT} - 1) \qquad (1)$$

(where e is the electronic charge, k Boltzmann's constant and T the absolute temperature), the collector current I_C at a given base-emitter voltage V_{BE} is proportional to the leakage current I_{C0}, and I_{C0} is in turn proportional to the surface area of the emitter. This area is critically determined by the masks used. The current gain factor h_{FE}, on the other hand, is dependent on the thickness d of the base layer, i.e. on the depth of the base diffusion less the depth of the emitter diffusion (*fig. 1*), and of course the difference between these two diffusion depths is much more difficult to make identical than the emitter areas.

The d.c. operating point of a transistor is affected by the temperature (see equation 1). Now in integrated circuits the distances between the elements are so small, and the thermal conductivity of the silicon is so high, that two closely adjacent transistors vary with temperature in practically the same way, provided that dissipating elements are kept sufficiently far apart.

This can be established by measuring the small difference in drift of the base-emitter voltage V_{BE} between two transistors in the same circuit that carry identical currents I_C. A value of 1 μV/°C is quite feasible. Now it can be demonstrated that at a V_{BE} of 0.6 V — approximately the value of V_{BE} at 100 μA — a temperature difference of 1 °C would cause a difference of 2 mV in V_{BE}. This indicates that the temperature difference between the transistors varies by only about 0.0005 °C for the same temperature change of 1 °C. By using balanced circuits, such as differential amplifiers, the temperature effects of the individual transistors can be made to compensate each other almost completely. Any slight temperature drift remaining is not so much due to temperature differences as to slight physical differences between the devices.

The price of monolithic circuits is determined by the initial costs, such as the costs of design and drawing the masks, and by the production costs, which in turn depend on chip size and on the number of contacts per circuit. Where large quantities are produced the initial costs are usually negligible; if the designer can succeed in keeping the chip size down and minimizing the number of contacts required, then the planar technique is very suitable for the inexpensive manufacture of reliable electronic circuits of high quality.

Limitations of the planar technique

It is not possible, as we have said, to choose the starting material and processes of the planar technique in such a way that optimum *NPN* and *PNP* transistors can be produced at the same time, perhaps with MOS or other field-effect transistors as well. *NPN* transistors

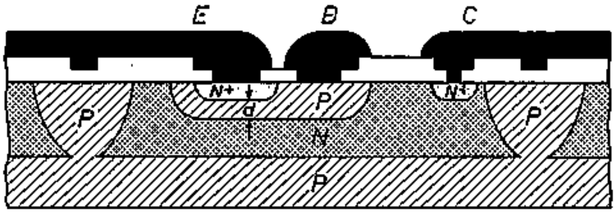

Fig. 1. Cross-section of an *NPN* transistor in an integrated circuit. The integrated circuit is made in a layer of *N*-type silicon applied epitaxially to a *P*-type silicon substrate. Part of the epitaxial *N*-type layer, separated from the rest by a *P*-type diffusion, serves as the collector. A *P*-type diffusion in the collector island forms the base of the transistor, an N^+ diffusion in the base forms the emitter. A second N^+ diffusion forms the contact with the *N*-type collector. *E*, *B* and *C* are metal conductors for the connections to emitter, base and collector. *d* thickness of base layer.

and resistors are usually regarded as the main product and *PNP* transistors as secondary products. Consequently the starting material is an epitaxially grown layer of *N*-type silicon, which serves as the collector material, into which a *P*-type region is subsequently diffused as the base, followed by the diffusion into this *P*-type zone of an *N*-type zone as the emitter (fig. 1). *PNP* transistors are made by diffusing two closely spaced *P*-type areas into the *N*-type layer. The characteristics of these lateral *PNP* transistors are not so good as those of the *NPN* transistors. The quality of the *PNP* transistors can be increased by means of a number of extra operations, but these of course make the circuit more expensive.

The resistors in a monolithic circuit are usually formed by channels of *P*-type material which are produced at the same time as the bases of the transistors in the epitaxial *N*-type layer. As might be expected, the accuracy of these resistors is not very high ($\Delta R/R \approx 10\%$), since the value depends not only on the surface area, which is determined by the masks, but also on the concentration of the *P*-type doping. The relative values of two resistors are maintained much more accurately (about 3%, and as good as 1% for resistors of the same value).

The same applies to the temperature coefficient. This is fairly high and, depending on the sheet resistance of the *P*-channel, varies between 0.1 and 0.3% per °C (the sheet resistance or surface resistivity is measured at the surface between two opposite sides of a square; the value is independent of the size of the square). Because the temperature difference is so small, the temperature coefficient of the ratio of two resistors may be much smaller.

Integration technique imposes a certain limitation on the size of the resistors. The area of one 10 kΩ resistor, for example, is equal to that of six transistors. In the circuits to be described transistors have deliberately been used instead of resistors wherever possible.

In cases where high resistances are indispensable, "buried resistors" sometimes provide the answer. These are resistors of *P*-type material covered by an *N*-type layer that is applied at the same time as the emitter diffusion. This has the effect of increasing the resistance value about ten times. Such a resistor has a field-effect-transistor configuration, and consequently the resistance depends on the voltage and there is some spread in the value.

In an integrated circuit a reverse-biased *PN* junction can be used as a capacitor. Its capacitance depends strongly on the reverse voltage. Another possibility is to apply an aluminium layer above an N^+ layer with the protective layer of silicon dioxide in the circuit acting as the dielectric.

The capacitance of both types of capacitor is proportional to the area that they occupy on the chip. A 200 pF capacitor occupies an area of the order of 0.1 mm². Only capacitors of very small values are therefore eligible for integration in a monolithic circuit.

Inductors cannot be made by the monolithic technique.

An integrated circuit always contains parasitic elements; the resistors and transistors, for example, always have parasitic capacitance to the *P*-type substrate (fig. 1). The parasitic *PNP* transistor formed in every *NPN* transistor by the base diffusion (*P*), the epitaxial layer (*N*) and the substrate (*P*) can often be particularly disturbing. This starts to conduct as soon as there is a forward voltage across the collector junction of the *NPN* transistor, which happens when it is driven into saturation. It may also happen, however, when the *NPN* transistor is operated as a diode by connecting the collector and base together (*fig. 2*), owing to the effect of the parasitic collector series resistance $r_{cc'}$ formed by the relatively poorly conducting epitaxial *N*-type layer. If the current through the transistor in a diode configuration becomes high enough for the voltage across the collector resistance $r_{cc'}$ to make the parasitic *PNP* transistor conduct, the current I will not

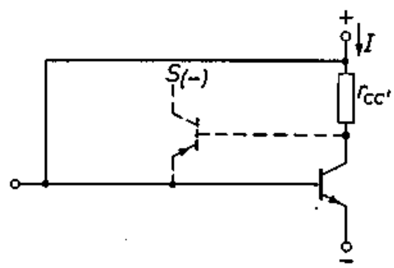

Fig. 2. Every integrated *NPN* transistor incorporates a parasitic *PNP* transistor, whose collector is formed by the substrate. If the *NPN* transistor is connected as a diode, the voltage across the internal collector resistance $r_{cc'}$ can make the parasitic *PNP* transistor conduct, so that part of the current flows to the substrate *S*.

flow entirely through the diode but partly through the *PNP* transistor to the substrate, which is at a negative potential to maintain the reverse-biased junction between substrate and epitaxial *N*-type layer. The collector series resistance can be reduced by means of a buried N^+ layer under the *N*-type silicon of the collector [3]. The collector contact diffusion is sometimes made so deep that it joins up with the buried layer, forming a "collector wall".

A new electronics

The capabilities and limitations of integration technology make it necessary to rewrite or add new material to our textbooks on electronics.

In the chapter on *Basic Circuits*, for example, the ordinary amplifier stage (*fig. 3a*) is not suitable for integration because of its many resistors and the large decoupling capacitor across the emitter resistor. The differential amplifier (fig. 3*b*) can fulfil the same function [4] [5], and is very suitable for integration, particularly since the load resistors can be replaced by a controlled current source — a new basic circuit that consists entirely of transistors. The resistors needed for limiting thermal drift in the circuit shown in fig. 3*a* are superfluous here since the drift in two identical transistors operating at the same temperature is exactly the same and does not give rise to any voltage between the output terminals, and therefore produces no output signal. Nor are these resistors needed for the d.c. bias, since this is also supplied by a current source. Under the heading of *Basic Circuits* a considerable amount of space will therefore have to be devoted to current sources.

The use of differential amplifiers and current sources also offers a wide variety of possible ways of coupling amplifier stages; these would have to be included in the chapter on *Amplifier Circuits*.

In the following we shall discuss in turn a number of circuits (current sources, input amplifiers, output amplifiers) which have been designed on the principles of this new integration electronics. Combination of these component circuits on a single chip of silicon gives complete integrated circuits, such as the operational amplifiers that are used in instrument electronics.

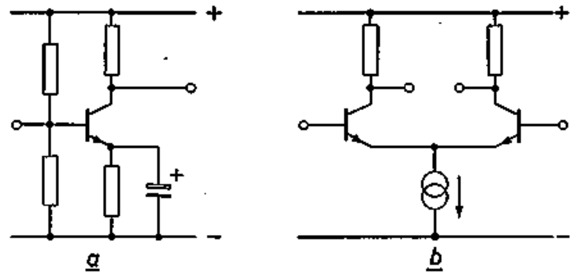

Fig. 3. *a*) Conventional transistor amplifier stage, not suitable for integration. *b*) Differential amplifier stage, very suitable for integration, particularly when the load resistors are replaced by transistor circuits.

Current sources

A current source has to supply a current that does not vary with the voltage across it; the ideal current source therefore has an infinitely high output impedance. In many cases it is desirable to be able to control the magnitude of the output current; in the circuits described here this is done by means of a reference current.

Controlled current source using two transistors

The simplest controlled current source consists of two identical transistors, one of which is connected as a diode (*fig. 4*). The two transistors have the same emitter area and therefore the same leakage current I_{C0}. Since they have the same base-emitter voltage V_{BE}, their collector currents I_{C1} and I_o are also equal:

$$I_{C1} = I_o = I_{C0}\,(e^{eV_{BE}/kT} - 1).$$

The base currents are thus $I_{C1}/h_{FE} = I_o/h_{FE}$ and are supplied by the reference-current source I_{ref}, so that $I_o = I_{ref} - 2\,I_o/h_{FE}$ or

$$I_o = I_{ref}\left(1 - \frac{2}{h_{FE}+2}\right), \qquad (2)$$

where the difference term expresses the two base currents. Since h_{FE} is of the order of magnitude of 100, I_o is approximately equal to I_{ref}. The circuit gives gain because I_o is delivered across a high output impedance, the collector impedance of transistor *2*, while the reference-current source (see page 7) has a conducting diode as its load and therefore does not need a high output impedance. Because of the symmetric structure of the circuit it is relatively insensitive to variations in temperature (eV_{BE}/kT has the same value for both transistors) and to voltage fluctuations.

With transistors of unequal emitter areas the ratio of the reference and output currents will be the same as the ratio of the emitter areas. Since these areas are fixed when the openings in the masks are drawn, their ratio can be fairly well controlled. To ensure accuracy in this ratio each of the two emitters is sometimes built up from a number of diffusions of equal magnitude.

As equation (2) shows, the emitter areas only determine the currents if the current gain h_{FE} of the transistors is sufficiently high.

An investigation of the behaviour of the circuit as a function of frequency involves the quantity h_{fe}, the current gain for the a.c. component in the base current of a transistor. At relatively low frequencies, h_{fe} is independent of frequency but at high frequencies h_{fe} decreases with rising frequency. The behaviour of h_{fe} as

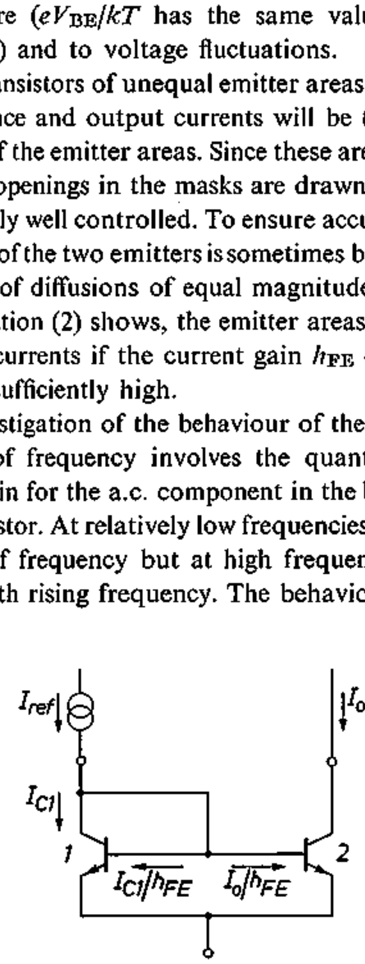

Fig. 4. Controlled current source with two transistors. Transistor *1* is connected as a diode. The output current I_o is independent of the voltage across the output and is approximately equal to the reference current I_{ref}.

a function of frequency f is approximated by the expression

$$h_{fe} = \frac{h_{fe0}}{1 + jh_{fe0}\, f/f_T}, \qquad (3)$$

where h_{fe0} is the value of $|h_{fe}|$ at low frequencies, and f_T is the frequency at which $|h_{fe}|$ has decreased to 1. Substitution of this expression in (2) gives the following equation for the a.c. components in output and reference current:

$$I_o = \frac{I_{ref}}{1 + 2/h_{fe0} + 2j\, f/f_T} \approx \frac{I_{ref}}{1 + 2j\, f/f_T}. \qquad (4)$$

We see from this that the equality of I_o and I_{ref} is no longer adequate when $f > \frac{1}{2} f_T$; the frequency characteristic of this current source is given by curve *a* in *fig. 5*.

The collector-emitter breakdown voltage $V_{(BR)CE0}$ of transistor *2* in fig. 4 is two or three times greater in this circuit than that of the transistor itself.

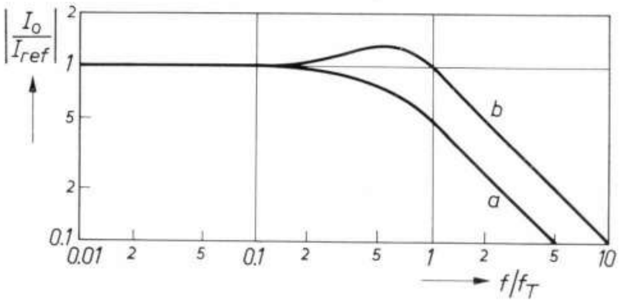

Fig. 5. Frequency characteristic of the controlled current source using two transistors (*curve a*) and of the controlled current source using three transistors (*curve b*).

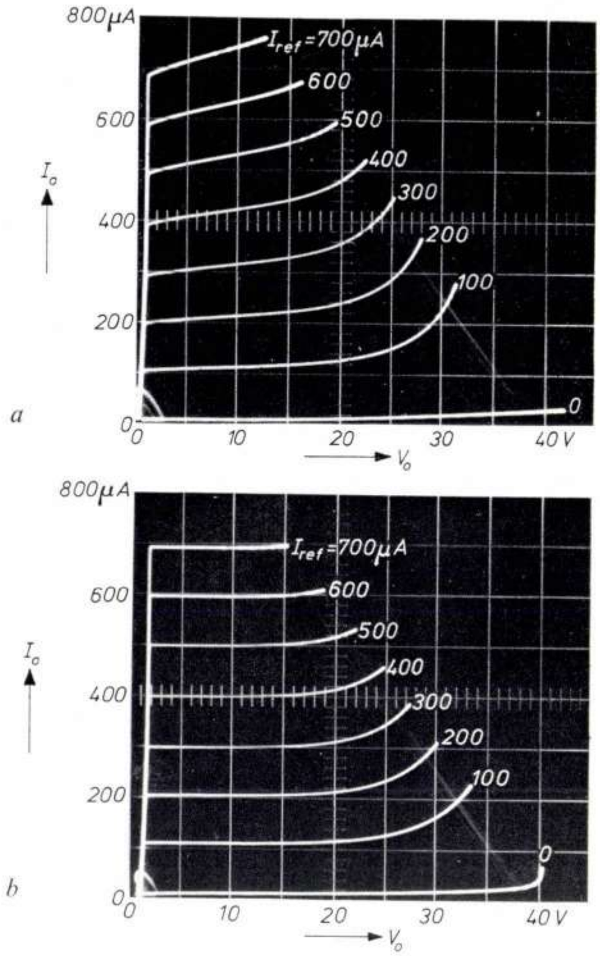

Fig. 6. Current-voltage characteristics of the controlled current source with two transistors (*a*) and of the controlled current source with three transistors (*b*). In both cases the breakdown voltage is about 0.8 times the collector-base breakdown voltage $V_{(BR)CB0}$ of the output transistor. The curves in (*b*) are flatter because of the higher output impedance of the circuit with three transistors.

This can be understood if we treat the whole circuit of fig. 4 as a single transistor with a current gain of $h_{FE} = 1$ and consider that in general:

$$V_{(BR)CE0} = V_{(BR)CB0}\,(1 + h_{FE})^{-1/N}$$

(where $V_{(BR)CB0}$ is the collector-base breakdown voltage and N has a value between 2 and 4). For the individual transistor h_{FE} is about 100, and therefore $V_{(BR)CE0}$ is two or three times lower than for our circuit as a whole, which has a breakdown voltage of $V_{(BR)CE0} = 0.7$ to $0.8\ V_{(BR)CB0}$. The breakdown effect is visible in the characteristics shown in *fig. 6a*.

Controlled current source using three transistors

By using a third transistor it is possible to make the output current I_o of the current source more accurately equal to the reference current I_{ref} (*fig. 7*). The circuit of fig. 7 operates by feeding back variations of the current through transistor *3* to the base of transistor *3* in the opposite sense by means of a current source like

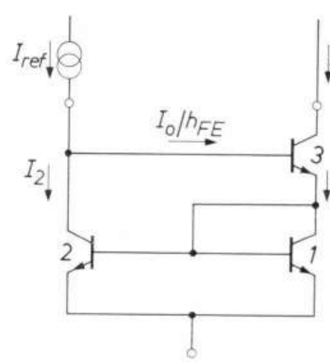

Fig. 7. Controlled current source with three transistors. I_{ref} reference current. I_o output current.

[4] G. Klein and J. J. Zaalberg van Zelst, General considerations on difference amplifiers, Philips tech. Rev. **22**, 345-351, 1960/61.

[5] G. Klein and J. J. Zaalberg van Zelst, Precision electronics, Philips Technical Library, Eindhoven 1967.

the one shown in fig. 4. For this current source eq. (2) shows that :

$$I_2 = I_1\left(1 - \frac{2}{h_{FE}+2}\right).$$

From the circuit it also follows that:

$$I_1 = I_o + I_o/h_{FE},$$

$$I_2 = I_{ref} - I_o/h_{FE}.$$

From these three equations we arrive at the output current:

$$I_o = I_{ref}\left(1 - \frac{2}{h_{FE}^2 + 2h_{FE} + 2}\right). \qquad (5)$$

A comparison with equation (2) shows that the difference term here is about h_{FE} times smaller than in the case of the circuit with two transistors. There is no difference term of the order of $1/h_{FE}$, since here I_{ref} and I_o each deliver a single base current.

The output impedance of the current source is equal to $\frac{1}{2}h_{FE}$ times the collector output impedance of a single transistor and therefore is $\frac{1}{2}h_{FE}$ times greater than that of the circuit in fig. 4. The reference current source has the impedance of two diodes connected in series as its load.

An idea of the high-frequency behaviour of this current source can be obtained by substituting equation (3) in (5). Since the value of h_{fe0} is high, some terms can be neglected, and we obtain the following expression for the a.c. components of the currents:

$$I_o \approx I_{ref}\left(1 - \frac{2}{h^2_{fe0} + 2\,h_{fe0} + 2}\right) \times$$

$$\times\,\frac{1 + 2\,\mathrm{j}\,f/f_T}{1 + 2\,\mathrm{j}\,f/f_T + 2(\mathrm{j}\,f/f_T)^2}. \qquad (6)$$

Curve *b* in fig. 5 shows the variation of $|I_o/I_{ref}|$ as a function of frequency. It can be seen from this that the current source with three transistors can be used up to higher frequencies than the one with two transistors.

The input and output impedances of the circuit can be calculated by determining the voltage variations at the base and collector of transistor *3* when I_{ref} and I_o are varied. If I_{ref} changes by an amount ΔI_{ref}, then I_o and I_1 change by the same amount and if S is the transconductance the current change ΔI_1 produces a voltage change $2\Delta I_1/S$ across the base-emitter junction of transistor *3* and across diode *1* [5]. The input impedance is therefore $2/S$, i.e. the impedance of two diodes in series.

At a variation of I_o the value of I_{ref} remains constant, and since I_2 follows the variations of I_1 a variation of $\frac{1}{2}\Delta I_o$ occurs in both the emitter current and the base current of transistor *3*. The base current variation $\frac{1}{2}\Delta I_o$ gives a variation in the base-emitter voltage of $\frac{1}{2}\Delta I_o h_{FE}/S$. Transistor *3* amplifies this voltage μ times (μ is the amplification factor and is equal to the product of the output resistance R_o of the transistor and the transconductance S), so that a voltage variation of $\frac{1}{2}\Delta I_o h_{FE}\mu/S = \frac{1}{2}\Delta I_o h_{FE} R_o$ appears at the collector. The output impedance of the current source is thus seen to be $\frac{1}{2}h_{FE}R_o$, i.e. $\frac{1}{2}h_{FE}$ times the output impedance of a single transistor.

It can be shown that the frequency characteristic should have the shape of curve *b* in fig. 5 by using Bode diagrams [6]. Using the current symbols to represent a.c. components again, we can write for the currents in transistor *3*:

$$I_1 = (h_{fe} + 1)\,I_o/h_{fe}.$$

Making use of equation (4), we can also write:

$$I_o/h_{fe} = I_{ref} - I_2 = I_{ref} - \frac{I_1}{1 + 2\mathrm{j}f/f_T},$$

from which it follows that

$$I_1 = \frac{I_{ref}}{\dfrac{1}{h_{fe}+1} + \dfrac{1}{1+2\mathrm{j}f/f_T}}. \qquad (7)$$

The Bode diagrams of $h_{fe} + 1$ and of $1 + 2\mathrm{j}f/f_T$ are given in *fig. 8a* and *8b*. Below $\frac{1}{2}f_T$ the second term is dominant in the denominator of (7), since $h_{fe} + 1$ is still $\gg 1$. Between $\frac{1}{2}f_T$

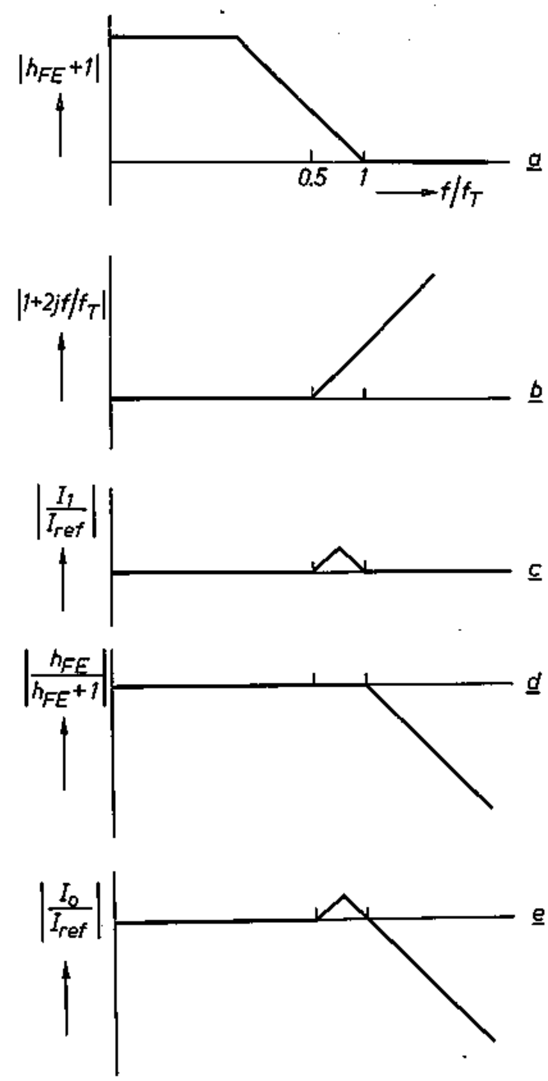

Fig. 8. Derivation of the frequency characteristic of the three-transistor current source represented by curve *b* in fig. 5, by means of Bode diagrams. Combining diagrams (*a*) to (*d*) gives diagram (*e*), which corresponds to the frequency characteristic.

and f_T both terms become equal to each other, and above this value the first term predominates. The Bode diagram of I_1/I_{ref} is therefore as shown in fig. 8*c*. What we are interested in, however, is not I_1/I_{ref} but I_o/I_{ref}, which is given by

$$I_o/I_{ref} = \frac{h_{fe}}{h_{fe}+1} I_1/I_{ref}.$$

The Bode diagram of $h_{fe}/(h_{fe}+1)$ is shown in fig. 8*d*; by adding fig. 8*c* and fig. 8*d* we get fig. 8*e*, which does indeed correspond to curve *b* in fig. 5.

The breakdown voltage of the current source with three transistors is about 0.8 $V_{(BR)CBO}$, just as in the case of the current source with two transistors. This can clearly be seen from the characteristics in fig. 6*b*, which, compared with fig. 6*a*, also show the higher output impedance of the circuit with three transistors.

Output current unequal to reference current

We have already seen that by giving reference diode *1* and transistor *2* of the current source in fig. 4 dissimilar emitter areas we are free to choose the ratio between reference and output current. This applies only within certain limits; ratios that are too high give large emitter areas and low cut-off frequencies.

A fixed ratio between output and reference current can also be obtained by incorporating a resistor in the emitter lead of diode or transistor. For the circuit shown in *fig. 9* it can be shown that

$$I_oR = (kT/e) \ln (I_{ref}/I_o). \qquad (8)$$

By adding relatively small resistances sources can be made that supply a very low current. If for example we have $I_{ref} = 100$ μA, then for an output current of $I_o = 10$ μA, (8) shows that R should have a value of 6 kΩ.

The use of emitters of different area does not upset the symmetry of the circuit, because the quantity eV_{BE}/kT is not dependent on the emitter area. The symmetry is however upset by the introduction of an emitter resistance, and the circuit no longer retains its basic insensitivity to temperature fluctuations. The temperature effect caused by the resistance is sometimes used for compensating other temperature effects.

It will be evident that the feedback current source of fig. 7 can also be designed with emitters of different area or with an emitter resistance. In that case, however, the expression for the output current again contains the difference term of the order $1/h_{FE}$ which did not occur in equation (5), since now all base currents no longer have the same magnitude.

Reference-current source

A reference-current source, which is an important element in the circuits dealt with here, is often obtained by deriving the current from the supply voltage via a large resistance.

A much more attractive current source for integration, which has only a small resistance and which is moreover independent of the supply voltage, can be obtained by combining two of the current sources described above (*fig. 10*). The resistor R serves for adjusting the output current.

The upper current source, consisting of *PNP* transistors, causes identical currents I_o to flow in both branches. To make the currents identical in the lower current source when there is a resistor R, transistor *2* is given a larger emitter area than transistor *1*. Equation (8) shows that R and the ratio p of the areas should then satisfy the condition:

$$I_oR = (kT/e) \ln p. \qquad (9)$$

For a given R and p the value of I_o is then fixed.

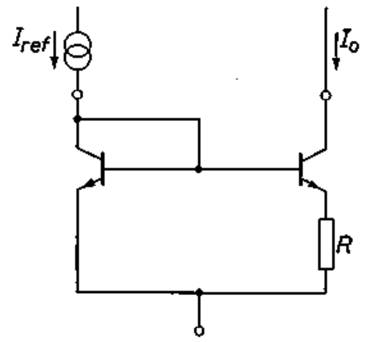

Fig. 9. Controlled current source in which the reference current and output current are unequal.

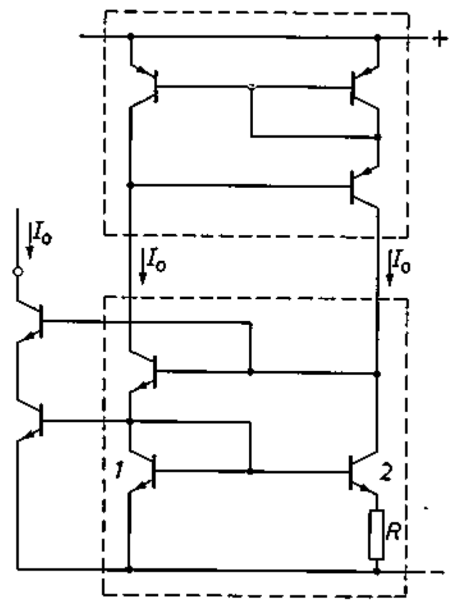

Fig. 10. Reference-current source which is independent of the supply voltage. A current source with three *PNP* transistors (*upper circuit inside dashed lines*) is connected with a current source using three *NPN* transistors (*lower circuit in dashed lines*); the output current of the one current source is the reference current for the other. An extra pair of transistors provides a constant current I_o.

[9] H. W. Bode, Network analysis and feedback amplifier design, Van Nostrand, Princeton, N.J., 1959.

If it is desired for example to bias the circuit to give a current of 100 μA, equation (9) indicates that for $p = 2$ we should choose $R = 180\ \Omega$.

The resistor does not have to be included in the current source with the unequal emitter areas. If it is not necessary to have identical currents in both branches, the resistor could be incorporated in the *PNP* current source.

Two extra transistors can be connected to the circuit to give a high-grade current source that could be used in a differential amplifier. Several such pairs can be connected to the circuit, enabling it to act as a common reference for a number of current sources, which can be necessary in a large amplifier circuit.

From equation (9) we see that I_0 is independent of the supply voltage but proportional to the absolute temperature, indicating that there might even be an application as a thermometer. This feature can be utilized for making the gain of a differential amplifier independent of temperature. The gain here is determined by the transconductance S of the transistors, which is given by $S = eI/kT$, where I is the current at the operating point. If I is obtained from the current source described here, we see from equation (9) that the transconductance and hence the gain is independent of temperature [7].

Amplifier circuits

The amplifier circuits we shall deal with are all based on the principle of the differential amplifier with a current source in the common emitter lead (fig. 3*b*). In the differential amplifier the input signal is the difference between the base voltages or currents of the two transistors. The difference between the collector currents is the output signal. Consequently, temperature and supply-voltage fluctuations, which cause the same variation in both collector currents, cancel out in the output signal, and the same is true for signals that appear in the same phase at the two bases [4] [5].

If a differential amplifier is followed by an output stage, the coupling is usually via a single-ended output; if it is followed by a second differential amplifier stage, then the coupling is balanced.

Differential amplifiers

When a controlled-current source is used as the collector load of a differential amplifier, the result is a circuit like the one shown in *fig. 11*. A voltage V_i between the input terminals gives rise to difference currents $\Delta I = \frac{1}{2}SV_i$. The controlled current source causes the difference currents to be added at the output, so that a current SV_i appears there across an impedance which is equal to the output impedances of the differential amplifier and of the current source in parallel.

Since there are no collector resistances there is very little decrease in the collector voltage at a steep increase in the current I. This enables the circuit to handle signals appearing in phase at the two inputs even when the signals are almost equal to the supply voltage. A low impedance, e.g. a transistor, should be used for taking off the difference current $2\Delta I$.

If a circuit with a higher output impedance than the circuit in fig. 11 is desired, the transistors in the differential amplifier can each be replaced by a cascode configuration, which increases the output impedance of the differential amplifier h_{FE} times. In this case a current source with three transistors must be used, which as we saw earlier gives an output impedance $\frac{1}{2}h_{FE}$ times higher than that of a single transistor. If all the transistors have the same parameters, the result is an output impedance $\frac{1}{4}h_{FE}$ times that of a single transistor. The amplification factor of the circuit thus obtained, which is the product of this output impedance and the transconductance S, may in practice be as high as 10^6. The same high amplification factor can also be achieved with two amplifier stages connected in series, but the circuit indicated here has the advantage that only a single time constant is significant at high frequencies, so that any negative feedback present will not give rise to instability.

An example of how a balanced coupling can be made between two differential amplifiers is shown in *fig. 12*. The load for the collectors of the input stage is a double

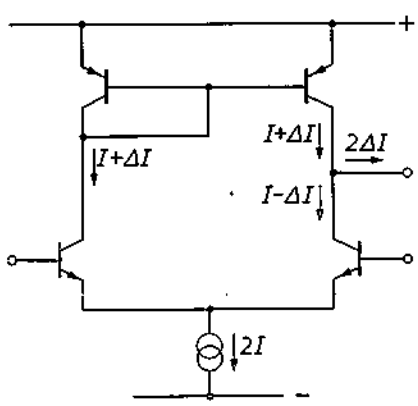

Fig. 11. Differential amplifier with single-ended output.

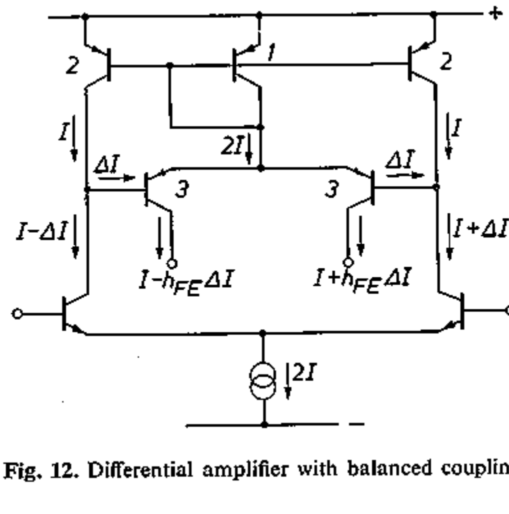

Fig. 12. Differential amplifier with balanced coupling.

controlled current source, which has two output transistors *2* and one combined reference diode *1* with an emitter area twice as large. The current in diode *1* is the sum current, which is independent of the drive and equal to $2I$, and since this diode keeps the base-emitter voltages of transistors *2* constant current I also flows through each transistor *2*. The difference currents $\pm \Delta I$ must therefore flow through the transistors *3*, and are therefore multiplied by the current gain h_{FE}.

The circuit can be extended similarly with a third amplifier stage by incorporating another such balanced circuit, now with *NPN* transistors, in the collector leads of the transistors *3*.

The collector voltage of the input transistors is equal to the supply voltage less two diode voltages. A positive in-phase signal at both inputs approximately equal to this voltage can be applied. The same applies to negative in-phase signals when the current source in fig. 7 is incorporated in the common emitter lead of the input transistors. This means that the differential amplifier with balanced coupling as described here is capable of handling exceptionally large in-phase signals.

Input circuits

After these general examples of differential amplifier stages we shall now examine the particular requirements which a differential amplifier must satisfy if it is to form the input stage of an integrated circuit.

When there is a d.c. coupling to an external circuit, both the input signal and the d.c. bias for the bases of both transistors have to be supplied from outside. In many applications it is desirable that the d.c. base currents should be small. This has led to the development of differential amplifiers with a low input current. A familiar example is the Darlington amplifier, a differential amplifier with series-connected emitter followers (*fig. 13*).

This configuration has a number of serious drawbacks. One is the considerable voltage drift, which is particularly undesirable in an input circuit. The current gain of the two inner transistors may differ appreciably, resulting in unequal currents through the outer transistors and thus causing an unbalance that leads to a marked voltage drift. Another drawback is that the output impedance of the outer transistors and the input capacitance of the inner transistors introduces an *RC* time constant which may be fairly high, since at the low emitter currents that flow the output impedance of the outer transistors is high.

The unbalance and the extra *RC* time constant can be reduced by taking an extra current from the outer pair (*fig. 14*). This minimizes the effect of the unequal base currents of the inner pair. The *RC* time constant becomes smaller because the emitter output impedance is reduced. The extra currents do not have to be high (e.g. 10 μA), but of course they cancel out to some extent the advantage of the Darlington amplifier.

The circuit in fig. 14 is not so attractive for integration because it contains fairly high resistances. In the version shown in *fig. 15* resistances ten times smaller can be used, and this circuit is therefore much more suitable for integration.

A small d.c. bias on the base is not the only requirement for the input stage of a differential amplifier. In some cases it may be necessary to stabilize the collector currents to keep the transconductance of the input transistors and the dissipation constant.

Fig. 16 shows a circuit that meets this requirement. The emitter leads of the differential amplifier in this circuit incorporate two *PNP* transistors *3* which perform three functions simultaneously. In the first place

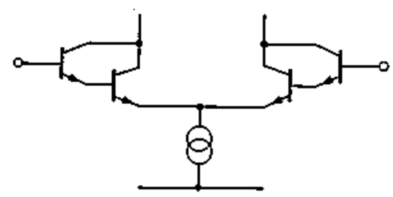

Fig. 13. Differential amplifiers with an emitter follower at the input to reduce the input currents (Darlington amplifier).

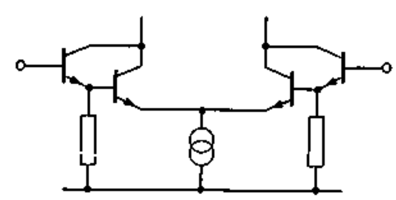

Fig. 14. Circuit as in fig. 13, in which extra currents are taken from the emitter followers to minimize the influence of the unequal current gain of the inner transistors.

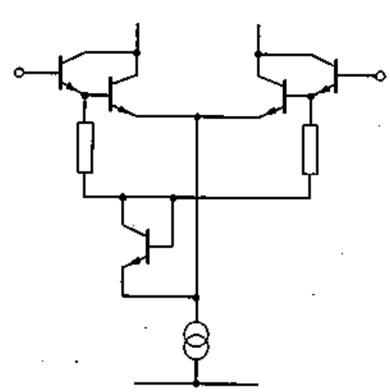

Fig. 15. Version of the circuit in fig. 14 with smaller resistors.

[7] A. J. W. M. van Overbeek and W. A. J. M. Zwijsen, Tunable integrated circuits, Philips tech. Rev. 27, 264, 1966.

they raise the breakdown voltage between the two input terminals to about 30 V; with the emitters connected the breakdown voltage would be equal to the Zener voltage of one of the base-emitter diodes, i.e. about 6 V. In the second place they give an output with a high internal impedance at the emitter end of the differential amplifier, i.e. at a d.c. voltage level close to that of the negative supply voltage. The advantage of this is that with an extra *NPN* transistor (shown dashed in fig. 16) an output is obtained at a d.c. voltage level midway between the supply voltages, i.e. at earth potential — a facility that is often required for the following stages. With an output at the collector end as in fig. 11 it is also possible to include an extra transistor to obtain an output at earth potential, but in this case the extra transistor must be of the *PNP* type, which has a lower cut-off frequency. Although the *PNP* transistors *3* in fig. 16 have a lower cut-off frequency, this does not matter so much since they are incorporated in a common-base configuration.

In the third place these same transistors help to stabilize the collector currents of the differential amplifier. This is because they form part of a controlled current source as in fig. 7, which also includes the transistors *1* and *2*. Unlike the configuration of fig. 7, there are two transistors *3*. The rather variable magnitude of the current gain h_{FE} of the *PNP* transistors is no drawback in this application, as eq. (5) shows, provided that h_{FE} is greater than about 5.

In this circuit, as in fig. 11, the difference signal is taken off by means of a current source consisting of two transistors. Here again, very large in-phase signals are permissible at both inputs.

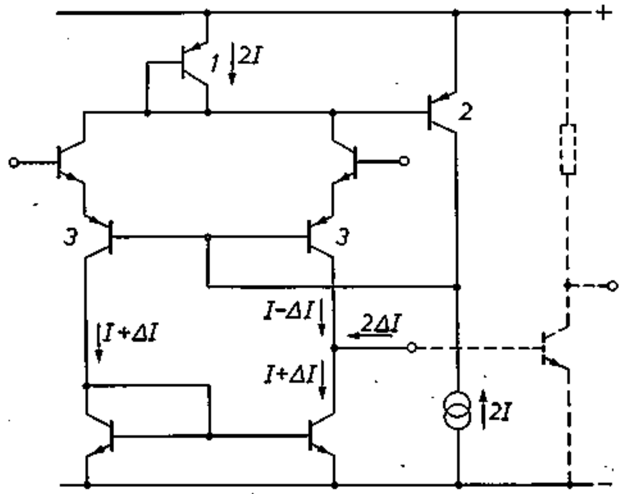

Fig. 16. Input differential amplifier in which the current source formed by transistors *1*, *2* and *3* keeps the collector currents constant. An *NPN* output transistor (*dashed lines*) is used to bring the d.c. level of the output signal to earth potential (midway between the positive and negative supply voltages).

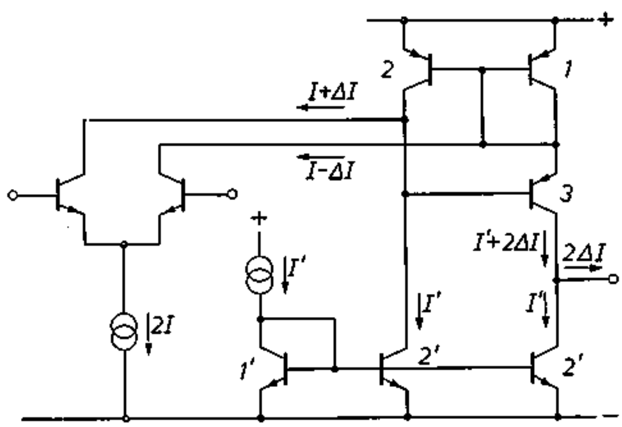

Fig. 17. Differential amplifier with d.c. level restorer, which brings the d.c. voltage level at the output midway between the positive and negative supply voltages, i.e. to earth potential.

D.c. level restorer

In the foregoing we have seen how the single-ended output of a differential amplifier is brought to a d.c voltage level between the positive and negative supply voltage, i.e. earth potential, by means of an extra transistor. This is desirable when this output has to be connected to an external load, whether or not via an output stage. In the circuit shown in fig. 12 we encountered a balanced output in which the amplifier stages shown, with possible extra ones, brought the difference signal to a level that lay alternately a few diode voltage levels above the negative or below the positive supply voltage. In this case a d.c. level shift is needed before the signal can be applied to an output or to an output circuit. *Fig. 17* shows a d.c. level restorer of this type. The signal is taken off this circuit by means of transistors in a common-base configuration. The voltage gain obtained with the circuit in fig. 16 is not obtained here, but on the other hand the bandwidth is greater.

The circuit consists of the two current sources *1*, *2*, *3* and *1'*, *2'*. The preceding differential amplifier is shown in simplified form on the left. The current through the source *1'*, *2'* with double output transistor *2'* is determined by the reference-current source I'; the reference current for transistor *2* of the upper current source is equal to the sum of I' and the current $I + \Delta I$ through the differential amplifier. The current through diode *1* is the same; the differential amplifier takes a fraction $I - \Delta I$ of this, and the remainder $I' + 2\,\Delta I$ flows through transistor *3*, and then divides between transistor *2'* which carries a current I', and the output.

The output impedance is equal to the parallel collector impedances of transistors *3* and *2'*. The circuit can be driven to within a few diode voltages of the supply voltage.

Class B output stage

Since a small load on the level shifter is sufficient to cause a loss of gain, it may be desirable in some cases to connect to the shifter an output stage that gives current gain only. A controlled current source can again successfully be used in such an output stage, as *fig. 18* shows. The current source here consists of the transistors *1* to *5*. A reference-current source I causes a constant current of about the magnitude of I to flow in the left-hand branch; this has the effect that the sum of the voltages across the base-emitter diodes of transistors *2* and *4* is constant.

This constant sum voltage also appears across the two base-emitter diodes of transistors *3* and *5*. This does not mean that the same current necessarily flows through these transistors. This is the case, though, when there is no output signal; a current approximately equal to I then flows in both transistors. However, if the voltage at the input of the circuit rises, the base-emitter voltage of transistor *5* rises with it, as does the current through this transistor. At the same time the base-emitter voltage of transistor *3* decreases by the same amount, so that a lower current flows in this transistor. Because of the exponential diode characteristic (1), an increase of the current in transistor *5* to γI ($\gamma > 1$) causes a decrease of the current in transistor *3* to I/γ. This is in fact a type of class B amplifier, but one in which the current in one branch never drops completely to zero.

If the voltage at the input of the circuit decreases, the current in transistor *5* also decreases and the current in transistors *3* and *1* rises. The base current for this is derived from the current source I.

The maximum output current during negative control is thus equal to the current I multiplied by the current gain of transistor *1*. If this output current is not sufficient, the circuit can be extended by adding a class C amplifier to it, as shown in *fig. 19*. The values of the resistors R are chosen so that transistors *7* and *8* do not conduct when the output current is small. When the output current is increased there comes a point at which the transistors start to conduct because the current supplied by transistors *3* and *5* increases the voltage across the resistors R. When this happens transistors *7* and *8* start to supply a large part of the output current.

Conclusion

We have seen from the treatment of basic and other circuits that in linear-circuits design today the trend is to adapt the circuit to the requirements of integrated-circuit technology in such a way as to obtain the optimum product. There are hardly any resistors. The circuits are as far as possible laid out symmetrically, thus minimizing temperature effects.

All this has been made possible through the successful application of the differential amplifier and a new element — a current source that can be controlled by a reference current. The introduction of this element is associated with a design approach in which current sources and current control are the dominant considerations. The voltages generated by the controlled currents are usually limited to diode voltages. A relatively low supply voltage is therefore sufficient, in spite of the stacking of transistors which characterizes linear integrated circuits today.

In this development the computer is an extremely useful tool. It is particularly useful in design calculations for determining high-frequency behaviour and the effect of spread in the elements and parasitic effects. The computer is also of great value in the drawing of masks and for circuit testing in production.

But it is the electronic engineer who will be responsible for the creation of new basic circuits. It is already

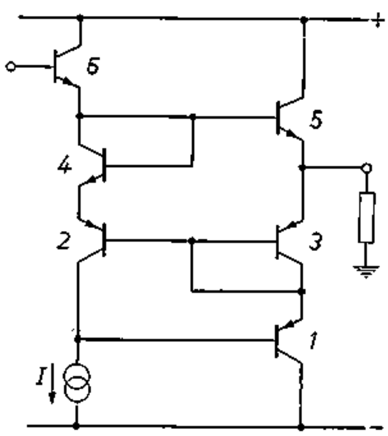

Fig. 18. Class B output stage.

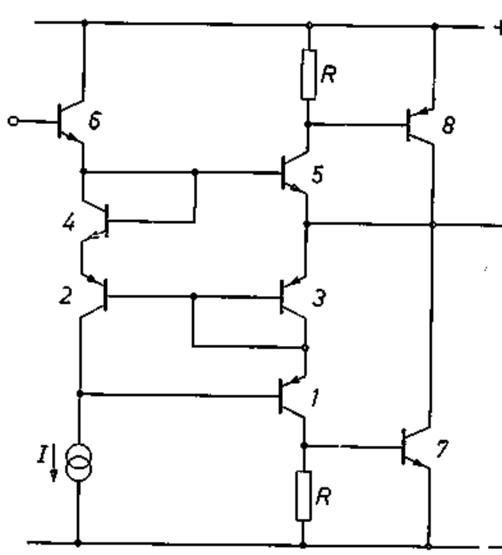

Fig. 19. Extension of the output stage in fig. 18 with a class C amplifier (transistor *7*, *8*) which only passes the peaks of the output current. This modification enables the circuit to supply higher output currents.

clear that circuit designers have accepted the challenge of integration with some enthusiasm and have found that there is much to be gained from the wealth of possibilities that it offers. Well may the layman look with some astonishment at integrated circuits, for in taking up the monolithic technique the electronic engineer — aided by the skill of the technologist — has produced circuits that are at least the equal of the traditional ones.

Summary. Integrated-circuit technology has both special capabilities and special difficulties for linear-circuit electronics. Transistors can be made that are identical and all operate at the same temperature; on the other hand, resistors and capacitors take up too much chip space and their use is to be avoided as far as possible. This is leading to a new electronics in which wide use is made of balanced circuits, e.g. differential amplifiers, and in which resistors are being replaced by current sources built up from transistors. Examples are discussed of current sources controlled by a reference current, differential amplifiers, and input and output stages. From these individual circuits complete integrated circuits such as operational amplifiers can be built up.