Negative Feedback in Transistor Amplifiers

Principles of single-stage and two-stage circuits

by S. W. Amos, B.Sc., M.I.E.E.*

There is little doubt that transistors will soon have replaced valves in all low-power applications—hardly surprising because in many respects they are superior. Transistors are, however, inferior to valves in that properties such as h_{fe} are subject to wide manufacturing spreads and to large variations with temperature. Moreover transistors have a leakage current which is highly dependent on temperature although at normal temperatures it is negligible in silicon devices. To manufacture transistor equipments with a consistent performance, the effects of differences in the value of h_{fe} must be minimized and the normal technique is to employ negative feedback. Provided sufficient feedback is used many significant properties of an amplifying stage can be made substantially independent of the parameters of the transistors, being determined instead by the components of the passive network used to apply feedback. Thus it is often possible to state the gain, transfer resistance, mutual conductance, etc., from an inspection of the component values. The use of negative feedback reduces gain, of course, but has advantages such as an improvement in linearity in amplifying stages.

In this article various circuits which can be used to apply feedback are discussed and typical calculations are given.

Derivation and injection of feedback signal

In general negative feedback is applied to an amplifying stage by taking a signal from the output and returning it to the input in such a way that the returned signal is in antiphase to the input signal. The input to the amplifier is now the difference between the source and feedback signals. The larger source signal required for a given amplifier output expresses the effective reduction in gain due to feedback. The link which feedback establishes between the input and the output of the amplifier enables a number of deficiencies in the output to be reduced. For example, if the gain of the amplifier falls at high frequencies, so also does the feedback signal amplitude. If the source signal is of constant amplitude, the amplifier input is thus greater at high than at low frequencies. This tends to offset the reduced gain and maintain a constant output signal. In this way negative feedback improves frequency response: it also brings about the other improvements in performance mentioned in the first paragraph.

The single-line diagram of Fig. 1 illustrates the principle of negative feedback

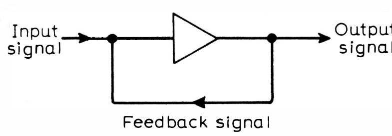

Fig. 1. Fundamental representation of negative feedback.

but does not show how the feedback signal is derived from the output or how it is injected into the input of the amplifier. The effect of feedback on the input resistance and output resistance of the amplifier is primarily determined by the way in which the feedback connections are made.

Two ways in which a feedback signal may be taken from the output of an amplifier are illustrated in Fig. 2. At (a) feedback is obtained directly from the output terminals

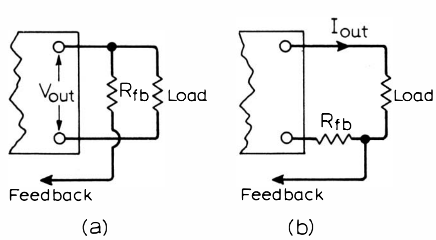

Fig. 2. General circuits for (a) parallel-derived and (b) series-derived feedback.

(or it can be taken from a potential divider connected across the output terminals). The significant feature is that the feedback circuit and the output circuit are in parallel and that the feedback signal is proportional to the output voltage. Any increase in the load resistor tends to increase the output voltage and hence the feedback signal. As a result the gain of the amplifier is reduced and the rise in output voltage minimized. Parallel-derived feedback thus tends to maintain a constant output voltage: in other words it effectively reduces the output resistance of the amplifier.

In Fig. 2(b) feedback is obtained from a resistor connected in series with the output load of the amplifier. The feedback signal is thus proportional to the output current of the amplifier. Any increase in the load resistor tends to reduce the output current and the feedback signal. As a result the gain of the amplifier is increased and the fall in output current is minimized. Series-derived feedback thus tends to maintain a constant output current: in other words it effectively increases the output resistance of the amplifier.

Two corresponding circuits for injecting the feedback signal into an amplifier are

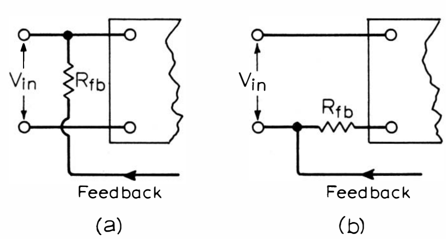

Fig. 3. General circuits for (a) parallel-injected and (b) series-injected feedback.

shown in Fig. 3. At (a) feedback is injected directly into the input terminals via a series resistor so that the feedback circuit and the input circuit of the amplifier are in parallel. The input signal has to offset the current from the feedback circuit and to supply the input current for the amplifier. For a given input voltage therefore, a larger input current is required as a result of the addition of feedback: this is equivalent to a reduction

Table 1

Type of feedback connection	Effect on input resistance	Effect on output resistance
series-derived		increased ($= g_m R_e r_c$)
parallel-derived		decreased ($= R_b/h_{fe}$)
series-injected	increased ($= h_{fe} R_e$)	
parallel-injected	decreased ($= R_b/A$)	

The formulae given in the table are derived in the article

*Head of Technical Publications Section, B.B.C.

in input resistance. Thus parallel-injected feedback effectively reduces the input resistance.

In Fig. 3(b) the feedback signal is connected in series with the amplifier input circuit. The input signal thus has to offset the voltage from the feedback circuit and to supply the input voltage for the amplifier. For a given amplifier input current, therefore, a larger input voltage is required as a result of adding feedback; this is equivalent to an increase in input resistance. Series-injected feedback effectively increases the input resistance of the amplifier.

The effects of the two types of feedback derivation and injection are summarized in Table 1.

Current and voltage amplification

A knowledge of the input resistance of an amplifier is necessary if it is to be matched to an external source, e.g. a microphone, to obtain maximum input signal. Similarly a knowledge of the output resistance is important if the amplifier is required to feed a line which must be accurately terminated.

However, the input and output resistance of the individual stages of a multi-stage amplifier are also important. When transistor stages are connected in cascade the performance of each stage should not be seriously affected by the coupling to the previous or the following stage. There are two ways in which this isolation can be achieved. One method is to ensure that a stage with a low output resistance is followed by one with a high input resistance, the high resistance being large compared with the low. The output voltage of the first stage is then also the input voltage of the second and this common voltage is little affected by variations in input or output resistance. In such inter-transistor coupling circuits the signal is clearly most conveniently regarded as a voltage and the design of such circuits is best carried out in terms of this voltage.

Alternatively isolation can be achieved by arranging for a transistor with a high output resistance to be followed by one with a low input resistance, the high resistance again being large compared with the low. In a connection of this type the output current of the first transistor is the input current for the second and this current is little affected by variations in the two resistances. For transistor couplings of this type the signal is most conveniently regarded as a current and the design of the circuit is best carried out in terms of this current.

The input and output circuits of transistors can thus be classified as suitable for voltage or current operation depending on the magnitude of the resistance as indicated Table 2. However it is not the absolute magnitude of an input or output resistance which is of importance, but their ratio. For example a transistor stage with an input resistance of 2000 Ω is best regarded as a current amplifier if the source resistance is 100,000 Ω but as a voltage amplifier if the source is only 50 Ω.

The input and output resistances of common-emitter and common-base transistor stages are.such that a cascade of them is best regarded as a current amplifier.

Table 2

Signal best considered as	Input resistance	Output resistance
voltage	high	low
current	low	high

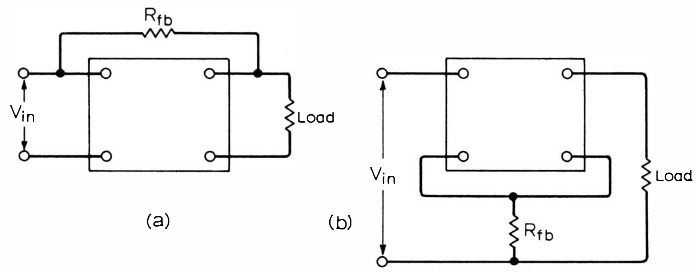

Fig. 4. General circuits for (a) parallel-derived, parallel-injected feedback and (b) series-derived, series-injected feedback.

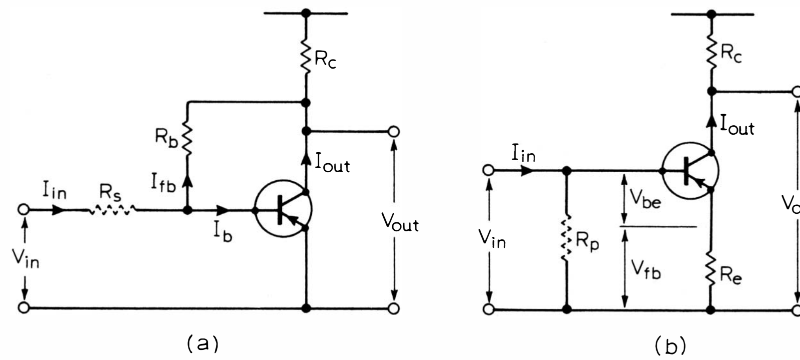

Fig. 5. A single-stage amplifier with (a) parallel-derived, parallel-injected feedback and (b) series-derived, series-injected feedback.

However, within limits we can make the input and output resistances what value we please by suitable choice of feedback circuit. Thus, by suitable design, we can make a transistor stage suitable for an input voltage or current and an output voltage or current. The performance of the stage can thus be measured by the values of I_{out}/I_{in}, V_{out}/V_{in}, I_{out}/V_{in} or V_{out}/I_{in} depending on the type of feedback applied to the stage.

Single transistor stage with base-collector resistor

By combining the circuits of Figs. 2(a) and 3(a) we can produce the circuit shown in Fig. 4(a) in which a single resistor bridges the input and output terminals. The resulting parallel-derived and parallel-injected feedback gives the amplifier a low input and a low output resistance.

Feedback of this type can be obtained by connecting a resistor R_b between base and collector of a common-emitter stage as shown in Fig. 5(a). www.keith-snook.info

This resistor returns a current V_{out}/R_b (proportional to output voltage) to the base of the transistor and this gives rise to a collector current I_c where

$$I_c = h_{fe} V_{out}/R_b$$

from which

$$\text{output resistance} = \frac{V_{out}}{I_c} = \frac{R_b}{h_{fe}}$$

In the absence of feedback the output resistance is the collector a.c. resistance of the transistor which can be of the order of 200 kΩ for silicon transistors. In a practical circuit R_b might be 50 kΩ and h_{fe} 150, giving an output resistance of 330 Ω. This illustrates the effective reduction of output resistance brought about by parallel-derived feedback.

At the input circuit we have

$$I_{in} = I_b + I_{fb}$$

and if R_b is low enough, I_{fb} is large compared with I_b and we can say

$$I_{in} \approx I_{fb}$$

Now $I_{fb} = V_{out}/R_b$ and $V_{out} = AV_{in}$ where A is the voltage gain of the transistor from base to collector. Thus

$$\text{input resistance} = \frac{V_{in}}{I_{in}} = \frac{R_b}{A}$$

which is less than in the absence of feedback.

This may be written $R_b/g_m R_c$ where R_c is the external collector load resistance. In an amplifier in which $R_b = 50$ kΩ, $R_c = 5$ kΩ and $g_m = 40$ mA/V, the input resistance is 250 Ω. In a practical circuit this may be effectively reduced by other resistors connected to the base for bias purposes.

By virtue of these low values of input and output resistance, the circuit is well suited for use with a current input and a voltage output and we are thus particularly interested in the value of V_{out}/I_{in}, the transfer resistance. We have already shown that if the degree of feedback is large I_{in} is approximately equal to I_{fb}. But $I_{fb} = V_{out}/R_b$.

$$\therefore \frac{V_{out}}{I_{in}} = R_b \qquad (1)$$

For the chosen numerical values

$$\frac{V_{out}}{I_{in}} = 50\ \text{k}\Omega$$

The circuit can also be used as a current amplifier. Because $V_{out} = I_{out}R_c$ we have

$$\frac{I_{out}}{I_{in}} = \frac{R_b}{R_c} \qquad (2)$$

and this has the value 10 for the numerical values used earlier.

Expressions (1) and (2) assume a current input but the circuit can be used with a voltage input provided a series resistor R_s is included as shown dotted in Fig. 5(a). R_s must be large compared with the input resistance of the amplifier so that the input current is given approximately by V_{in}/R_s. R_s could be the resistance of the signal source itself provided this is high enough and there is then no need to add a resistor to the circuit to provide the required high resistance. Substituting for I_{in} in (1) and (2) we have

$$\frac{V_{out}}{V_{in}} = \frac{R_b}{R_s} \qquad (3)$$

$$\frac{I_{out}}{V_{in}} = \frac{R_b}{R_c R_s} \qquad (4)$$

In the numerical example the input resistance was 250 Ω. R_s could then be 2·5 kΩ. From (3) the voltage gain is 20. From (4) the effective mutual conductance is 4 mA/V.

Circuit with emitter resistor

By combining the circuits of Figs. 2(b) and 3(b) we can produce the circuit shown in Fig. 4(b) in which a common resistor is included in the input and output circuits. The resulting series-derived and series-injected feedback gives the amplifier a high input and high output resistance.

Feedback of the type of Fig. 4(b) can be obtained by including a resistor R_e in the emitter circuit of a common-emitter stage as shown in Fig. 5(b). The voltage I_cR_e (proportional to output current) developed across R_e is applied between base and emitter of the transistor. This voltage is amplified by a factor $g_m r_c$ where r_c is the collector a.c. resistance of the transistor if, as assumed in calculations of output resistance, the output load is infinite.

Thus

$$V_{out} = g_m r_c R_e I_c$$

and

$$\text{output resistance} = \frac{V_{out}}{I_c} = g_m r_c R_e$$

$g_m R_e$ is normally greater than unity, confirming the effective increase in output resistance. As r_c is commonly of the order of 200 kΩ for a silicon transistor, this type of feedback can give very high output resistances.

At the input circuit

$$V_{in} = V_{fb} + V_{be}$$

and if R_e is large enough V_{fb} is large compared with V_{be}

$$\therefore\ V_{in} \approx V_{fb}$$

Now $V_{fb} = I_eR_e = (h_{fe}+1)I_bR_e$. Thus the input resistance is given by

$$\therefore\ \frac{V_{in}}{I_b} = (h_{fe}+1)R_e \approx h_{fe}R_e$$

The input resistance is given by $h_{fe}R_e$ approximately and, for a transistor with $h_{fe} = 150$ and $R_e = 1$ kΩ, is equal to 150 kΩ. In a practical circuit this may be effectively reduced by resistors connected to the base e.g. for bias purposes.

By virtue of the high input and output resistances the amplifier is well suited for use with an input voltage and an output current and we are particularly interested in the value of I_{out}/V_{in} i.e. the mutual conductance g_m.

We have already shown that if the degree of feedback is large V_{in} is approximately equal to V_{fb}. But $V_{fb} = I_eR_e \approx I_cR_e \approx I_{out}R_e$.

$$\therefore\ \frac{I_{out}}{V_{in}} = \frac{1}{R_e} \qquad (5)$$

If, as assumed earlier, R_e is 1 kΩ the effective mutual conductance is 1 mA/V.

If the amplifier is required to give a voltage output, the low output resistance necessary is provided by R_c which can be given a suitably low value such as the 5 kΩ assumed earlier. If, however, the amplifier is required to give a current output, R_c should be removed: the high value of the internal collector a.c. resistance then provides the required high value of output resistance. Alternatively when the amplifier is required to give a current output, R_c can be taken as representing the low input resistance of the following stage.

When R_c is present $V_{out} = I_{out}R_c$ and thus, as a voltage amplifier, the gain of the circuit is given by

$$\frac{V_{out}}{V_{in}} = \frac{R_c}{R_e} \qquad (6)$$

For the numerical values quoted earlier the voltage gain is 5.

Expressions (5) and (6) are in terms of a voltage input but the amplifier can be used with a current input provided a low-value resistor R_p is connected across the input terminals as shown dotted in Fig. 5(b). Resistor R_p should be small compared with the input resistance of the amplifier so that $V_{in} = I_{in}R_p$. Resistor R_p could be the resistance of the signal source itself if this is small enough, and there is then no need to add a resistor to provide the required low resistance.

Substituting $I_{in}R_p$ for V_{in} in (5) and (6) we have

$$\frac{I_{out}}{I_{in}} = \frac{R_p}{R_e} \qquad (7)$$

$$\frac{V_{out}}{I_{in}} = \frac{R_pR_c}{R_e} \qquad (8)$$

If R_p is 5 kΩ, the current gain is 5 and the transfer resistance 25 kΩ.

Single-transistor circuit

The circuits so far discussed have been simplified by omission of all components except those essential for amplification or for feedback. In a practical circuit provision must be made for biasing the base to give the required value of mean collector current and for the stabilization of this current. For a single-stage amplifier a satisfactory means of meeting these requirements is that shown in Fig. 6 in which the base is returned to a potential divider across the supply. This, together with the emitter resistor (which

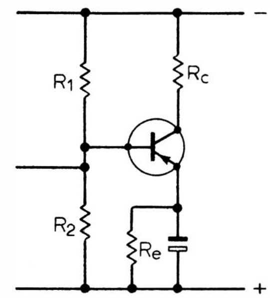

Fig. 6. Mean collector current stabilization by means of a potential divider and emitter resistor.

can be used for signal-frequency feedback also if desired) ensures reasonable stability of collector current and thus minimizes variations in performance due to variations in leakage current or in the value of h_{fe}. It also minimizes the effects of spreads in h_{fe} so permitting the construction of a number of circuits with consistent performance.

The circuit operates by impressing a constant potential on the base of the transistor. Any tendency of the emitter current to increase causes the voltage across R_e to rise and reduces the base-emitter voltage thus reducing the increases in emitter current. The circuit is an example of zero-frequency feedback. The effectiveness of the circuit in stabilizing collector current is increased by increasing the value of R_e and by decreasing the resistance of R_1 and R_2 in parallel (which is the effective internal resistance of the source of base voltage).

Stabilization could be improved by returning the potential divider to the collector instead of to the collector supply voltage because increase in collector current is now minimized in two ways. The emitter potential is raised (as in the simple potential divider circuit) as a result of the increased voltage generated across the emitter resistor. In addition, however, the base potential is lowered as a result of the increased voltage generated across the collector resistor by the increased collector current.

If however R_1 is simply transferred to the collector, difficulties arise from the potential-divider bleed current which now flows through R_c and from the unwanted signal-frequency feedback introduced by the potential divider. Because of these difficulties, collector feed of the potential-divider circuit is unlikely to be employed in this simple form. It is, however, the basis of a very effective form of stabilization used in two-stage amplifiers described later.

We will now consider the design of a practical single-transistor stage required to operate with a voltage input, to give a voltage output, the voltage gain being 20. The basic circuit of Fig. 5(b) is suitable and we will assume that a silicon n-p-n transistor with $h_{fe} = 150$ (at a collector current of 1 mA) is to be used with a supply voltage of 24. For good stabilization the voltage at R_1R_2 junction should be large compared with changes in base-emitter voltage due to temperature and transistor tolerances. For a silicon transistor V_{be} is commonly 0·7 V and thus the voltage across R_e is 0·7 V less than that at the potential-divider junction. A suitable voltage across R_e is 7 and this gives the value of R_e as 7 kΩ. The voltage across R_c and the transistor is thus 17 and, to permit the greatest output voltage swing, this should be shared equally between them. Thus the no-signal voltage across R_c is 8·5 and R_c should be 8·5 kΩ.

The voltage gain of the stage is given by R_c/R_e and to give the required value of 20, R_e should be 420 Ω. Thus for zero-frequency feedback R_e should be 7 kΩ and for signal-frequency feedback it should be 420 Ω. This can be achieved as shown in Fig. 7 by

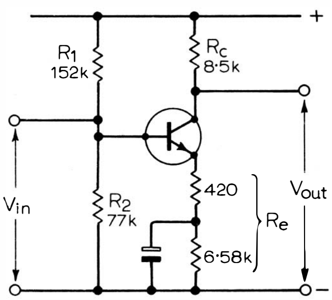

Fig. 7. Practical version of the circuit of Fig. 6, designed for a voltage gain of 20.

constructing R_e of 420 Ω and 6580 Ω in series and by decoupling the larger of the two resistors by a capacitor with a low reactance at the lowest signal frequency. The input resistance of the transistor at signal frequency is given by $h_{fe}R_e$ i.e. 73 kΩ but this is effectively reduced by R_1 and R_2.

For reasonable stability the current taken by the potential divider from the supply should be at least 10 times the standing base current of the transistor. I_b is approximately I_c/h_{fe}* i.e. 7 μA and a convenient value for the potential divider current is 100 μA. R_2 thus carries a current of 0·1 mA and the voltage across it is 7·7: the resistance is therefore 77 kΩ. R_1 carries 0·107 mA and the voltage across it is 27 − 7·7, i.e. 16·3 V: the resistance is therefore 152 kΩ. The input resistance of the amplifier is made up of 77 kΩ, 152 kΩ and 73 kΩ in parallel i.e. 30 kΩ. The output resistance is equal to R_c i.e. 8·5 kΩ.

The stability factor of this circuit, using the calculated component values, is 0·04,† that is to say the variations in collector current due to changes in h_{fe} or in leakage current are reduced to 4% of what they would be without the stabilizing circuit. We can also say that a spread of ±50% h_{fe} gives only a ±2% spread in collector currents.

*Strictly I_b is given by I_c/h_{FE} where h_{FE} is the static or d.c. current amplification factor. For simplicity we are assuming here that there is little difference between h_{fe} and h_{FE}.

†This is calculated from the expression

$$K = 1/[1 + h_{fe}R_e/(R_e + R_b')]$$

where $R_b' = R_1R_2/(R_1 + R_2)$.

Two-stage current amplifier

If we arrange for a Fig. 5(a) type of stage to feed into a stage of the type of Fig. 5(b) we obtain the two-stage amplifier shown in skeleton form in Fig. 8. The low input resistance and high output resistance

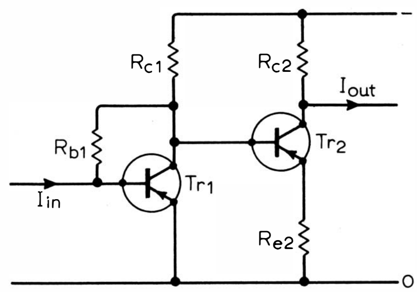

Fig. 8. Skeleton form of two-stage current amplifier.

of the two-stage circuit makes it suitable as a current amplifier. The input resistance of the second stage is high and it can be connected across the low output resistance of the first stage without mutual interaction. The output voltage of the first stage is the input voltage of the second.

The gain of the amplifier is easily assessed. We know from (1) that for Tr_1

$$\frac{V_{out}}{I_{in}} = R_{b1}$$

and for Tr_2 from (5)

$$\frac{I_{out}}{V_{in}} = \frac{1}{R_{e2}}$$

But $V_{in} = V_{out}$

$$\therefore \quad \frac{I_{out}}{I_{in}} = \frac{R_{b1}}{R_{e2}}$$

If, as assumed earlier, R_{b1} = 50 kΩ and R_{e2} = 1 kΩ the current gain is 50.

Stability considerations

The signal at Tr_2 emitter is a copy of that at Tr_2 base which is directly connected to Tr_1 collector. Thus the performance of the circuit is unaffected if R_{b1} is transferred from Tr_1 collector to Tr_2 emitter as shown in Fig. 9. This modification is of considerable help in practical versions of this circuit because it makes possible a simple but very

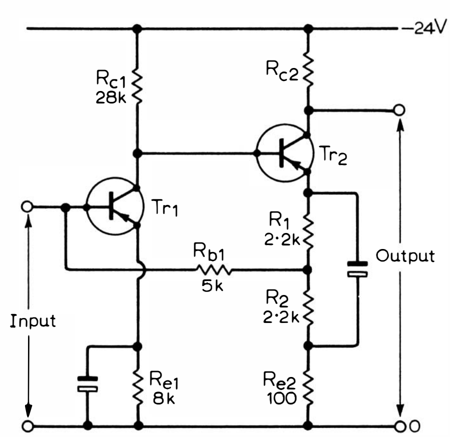

Fig. 9. More detailed version of the circuit in Fig. 8, designed for a current gain of 50.

effective means of stabilizing the mean collector current of both transistors.

If R_{b1} is returned to a potential divider R_1R_2 included, in addition to R_{e2}, in Tr_2 emitter circuit its value can be reduced whilst keeping the gain (R_{b1}/R_{e2}) constant. The circuit is now very similar to that of the potential divider method of stabilization (Fig. 6) but Tr_1 requires an emitter resistor (R_{e1}) to complete the circuit. This should be decoupled because it is not required to give signal-frequency feedback. This is a particularly good circuit because R_{b1}, R_1 and R_2 can be of low resistance and in addition the potential divider is returned in effect to the collector of Tr_1 by emitter-follower action in Tr_2. Both factors, as mentioned earlier, make for good stabilization. R_{b1} can be reduced to 5 kΩ and, for a current gain of 50, R_{e2} should be 100 Ω.

Estimation of component values

Suppose a 24-V supply is available. A suitable value for the no-signal voltage at Tr_1 collector is 10, giving the voltage across R_{c1} as 14. If we decide on a mean collector current for Tr_1 of 0·5 mA, R_{c1} is 28 kΩ. Tr_2 base voltage is 10 but because of the standing 0·7 V base-emitter voltage of silicon transistors, Tr_2 emitter voltage is 9·3. If Tr_2 is to take a mean emitter current of 2 mA, the total emitter resistance is 4·65 kΩ. Of this 100 Ω must provide signal-frequency feedback R_{b1} being taken as 5 kΩ, and the balance could consist of two 2·2 kΩ resistors in series, decoupled, the centre point providing bias for Tr_1 base. The voltage across R_{b1} caused by Tr_1 base current can be neglected and thus we can say that Tr_1 base voltage is approximately 4·7. Because of the voltage across Tr_1 base-emitter path, the emitter voltage can be taken as 4·0. The emitter resistance is thus 8 kΩ.

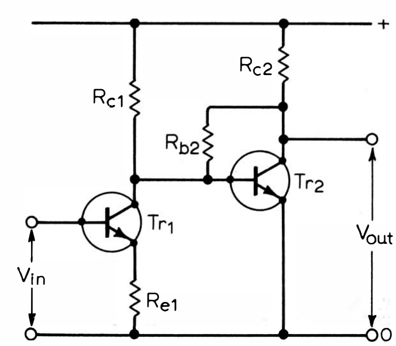

Fig. 10. Skeleton form of two-stage voltage amplifier.

The input resistance of the amplifier is given by R_{b1}/A where $A = g_{m1}R_{c1}$. The mutual inductance of a transistor is directly proportional to emitter current and for a current of 0·5 mA is approximately 20 mA/V. A is thus 560 and the input resistance is 5 kΩ/560 i.e. approximately 9 Ω. The output resistance is given by $g_{m2}R_{e2}r_{c2}$. If g_m is taken as 80 mA/V, $g_{m2}R_{e2}$ is 40 and the output resistance is thus several megohms, r_{c2} being say 200 kΩ.

Two-stage voltage amplifier

An amplifier consisting of a first stage of the type shown in Fig. 5(b) and a second stage of the type shown in Fig. 5(a) has a high input resistance and a low output resistance, making the amplifier (shown in skeleton form in Fig. 10) suitable for voltage amplification. At the inter-transistor coupling circuit the high output resistance of Tr_1 feeds into the low input resistance of Tr_2. There is no interaction provided the inter-transistor signal is taken as a current.

For Tr_1 from (5)

$$\frac{I_{out}}{V_{in}} = \frac{1}{R_{e1}}$$

For Tr_2 from (1)

$$\frac{V_{out}}{I_{in}} = R_{b2}$$

But $I_{in} = I_{out}$

$$\therefore \quad \frac{V_{out}}{V_{in}} = \frac{R_{b2}}{R_{e1}}$$

R_{b2} is normally connected to Tr_1 emitter instead of to Tr_2 base. It is not immediately obvious that such an alteration makes little difference to the performance of the circuit. In fact the input current for Tr_2 is (provided R_{c1} is large enough) the output current of Tr_1 and this is also the emitter current of Tr_1. Thus any current injected into Tr_1 emitter by R_{b2} is conveyed to Tr_2 base with little loss.

Suppose a voltage gain of 100 is required. Any values of R_{b2} and R_{e1}, provided their ratio is 100, will give this value of gain but high values of R_{b2} will give the amplifier an unnecessarily-high output resistance and low values of R_{e1} will give undesirably-low values of amplifier input resistance. A compromise such as $R_{e1} = 500\,\Omega$ and $R_{b2} = 50\,k\Omega$ is suitable.

Finally, means must be provided for stabilizing the collector currents of both transistors and here the same technique of zero-frequency feedback from Tr_2 emitter to Tr_1 base can be used as in the two-stage current amplifier and very similar component values can be used also. The emitter circuit of Tr_2 should be fully decoupled because no signal-frequency feedback is required here. In Tr_1 emitter circuit, however, a resistor of $8\,k\Omega$ is required for stability and $500\,\Omega$ for signal-frequency feedback. Both requirements can be met by using two resistors in series, the larger being decoupled as shown in Fig. 11. In the two-stage current amplifier R_{b1} provided the zero-frequency feedback necessary for stability and the signal-frequency feedback required to give the desired gain. In this voltage amplifier R_{b1} provides only zero-frequency feedback and R_{b2} provides signal-frequency feedback.

The input resistance of transistor Tr_1 is $50\,k\Omega$ if h_{fe} is taken as 100 and R_{e1} as $500\,\Omega$ but R_{b1} is in parallel with the base circuit. R_{b1} can be given any value within a wide range: high values degrade stability but give a high input resistance to the amplifier; low values give good stability but low input resistance. A compromise value such as $20\,k\Omega$ might be suitable and this gives an amplifier input resistance of $14\,k\Omega$.

The output resistance of the amplifier is given by R_{b2}/h_{fe} i.e. $500\,\Omega$ if $h_{fe} = 100$.

Some departures from the calculated resistor values indicated in Figs. 7, 9 and 11

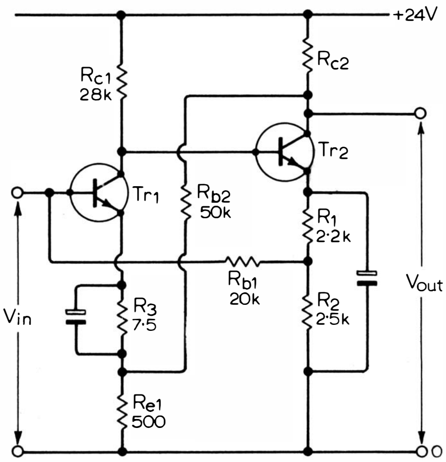

Fig. 11. More detailed version of the circuit in Fig. 10, designed for a voltage gain of 100.

may be desirable to permit the use of preferred-value resistors. Because of such departures and the spread of resistance likely to be encountered, it is advisable to use a preset component for one of the resistors in the amplifier and to adjust this to give the required working voltages. A suitable component to make preset is the decoupled part of Tr_1 emitter resistor.

No mention has been made in this article of the frequency range of the circuits discussed. Modern silicon planar transistors, even those intended for a.f. applications, have transition frequencies of hundreds of MHz and if these are used, the passband of the amplifiers will probably extend to several MHz. If the amplifiers are used for a.f. applications such a response could be an embarrassment (e.g. because of amplification of any r.f. signals present) and should be curtailed by making the feedback increase above say 15 kHz. This can be done, for example, by shunting R_{b2} in Fig. 11 by a capacitor so chosen that its reactance at 30 kHz equals the resistor value, i.e. approximately 0·005 μF. For a.f. applications the decoupling capacitors should be large enough to perform adequately down to 30 Hz.

Announcements

Two six-week courses are to be held at **Norwood Technical College.** The first, commencing 2nd February, is entitled "Pulse Code Modulation Techniques". The second, "Single Standard Colour Television Receivers" commences 25th January. Further details are available from The Secretary, Norwood Technical College, Knight's Hill, London S.E.27. Fee 30s per course.

Japanese made EVR players may soon become available in Britain as a result of a licence agreement which the EVR Partnership has concluded with Mitsubishi Electric of Japan for the manufacture and distribution of EVR teleplayers internationally (with the present exception of the U.S.A. and Canada). A similar agreement has also been made with Hitachi.

By mutual agreement the arrangement whereby Siemens components and telecommunications test equipment was handled in the U.K. by Cole Electronics has been terminated. From January 1st Siemens (U.K.) Ltd, Great West House, Great West Road, Brentford, Middx, will handle these products.

MCP Electronics Ltd, Alperton, Wembley, HAO 4PE, Middx, have been appointed sole U.K. representatives and distributors for Telefunken semiconductors.

LST Components, 7 Coptfold Road, Brentwood, Essex, now distribute heat sinks manufactured by Marston Excelsior Ltd (Imperial Metal Industries Group) and the 20-W integrated circuit amplifier made by Toshiba.

Russian test and measuring equipment. Z & I Aero Services now market a range of measuring equipment. Maintenance facilities are available. Z & I Aero Services Ltd, 44a Westbourne Grove, London W.2.

GDS (Sales) Ltd, of Michaelmas House, Salt Hill, Bath Road, Slough, Bucks, have been appointed a franchised distributor by **Radiatron Components Ltd.** The agreement covers the Elma range of collet knobs and rotary stud switches and Jaquet stopwatches.

Semicomps Ltd, 5 Northfield Estate, Beresford Avenue, Wembley, Middx, have been appointed **sole U.K. distributors** for the EMC range of dual-in-line i.c. sockets by Teknis Ltd, of Guildford, Surrey.

An agreement has been signed which gives AB Sonab, of Sweden, the sole **marketing rights in Northern Europe** for Ultra Electronics' range of communications equipment.

Microsystems International, of Canada, specialists in **standard and custom-designed microcircuits,** have appointed Pinnacle Electronics Ltd, Achilles Street, New Cross, London S.E14, as their first U.K. distributor.

Jason Electronic Designs Ltd are no longer wholesaling stocks of **Dansette and Perdio spares.**

AB Electronic Components are to take over the entire European manufacturing and marketing operations of the American component manufacturer, **CTS Corporation,** in exchange for 10% of their equity.

Jasmin Electronics Ltd have moved from Hainault, Essex, to a new factory at Station Road, Quorn, Leics. LE12 8BP.

The sales and service divisions of **Carston Electronics** have moved from Chinnor to Shirley House, 27 Camden Road, London N.W.1 (Tel: 01-267 2748).

The Broadcast Division of **Rank Precision Industries Ltd** are moving from Welwyn Garden City to Watton Road, Ware, Herts. (Tel: Ware 3939).

Interscan Data Systems (UK) Ltd have moved from London to Hoechst House, Salisbury Road, Hounslow, Middx. Tel: 01-572 2871.

Spectra-Physics Ltd have moved to premises at 5 Wolsey Road, Hemel Hempstead, Herts.

The Avionics Division of Plessey Electronics Group has been awarded contracts valued at £227,000 for **automatic and manual test equipment,** from the Ministry of Aviation Supply.

Pye Telecommunications Ltd has received a £¾M contract to supply a **nationwide communications system** for the Sierra Leone police.

Racal-BCC Ltd, of Bracknell has received an order valued at over £1,300,000 from the Malaysian Ministry of Defence. The order is for over 1,000 **military radio telephones** with accessories and spares.