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, 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, deficiencies in the output to be reduced.

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

![Fig. 1. Fundamental representation of negative feedback.](image)

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-injected and (b) series-injected feedback.](image)

(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 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.

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

The input signal has to offset the current 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 formulae given in the table are derived in the article

<table>
<thead>
<tr>
<th>Type of feedback connection</th>
<th>Effect on input resistance</th>
<th>Effect on output resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>series-derived</td>
<td>increased (= $g_{m}R_{f}$)</td>
<td>decreased (= $R_{f}/h_{fe}$)</td>
</tr>
<tr>
<td>parallel-derived series-injected</td>
<td>increased (=$R_{f}/R_{s}$)</td>
<td>decreased (=$R_{s}/h_{fe}$)</td>
</tr>
<tr>
<td>parallel-injected</td>
<td>decreased (=$R_{f}/A$)</td>
<td></td>
</tr>
</tbody>
</table>

*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 of 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 in 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 kΩ is best regarded as a current amplifier if the source resistance is 100,000 kΩ but as a voltage amplifier if the source is only 50 kΩ.

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>
<thead>
<tr>
<th>Signal best considered as</th>
<th>Input resistance</th>
<th>Output resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>voltage</td>
<td>Input current</td>
<td>Output voltage</td>
</tr>
<tr>
<td>current</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td></td>
<td>low</td>
<td>high</td>
</tr>
</tbody>
</table>

Table 2

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 $I_{in}/R_{in}$, $V_{out}/V_{in}$, $I_{out}/I_{in}$ 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 transistor 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

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

and if $R_b$ is low enough, $I_b$ is large compared with $I_f$ and we can say

$$I_{in} \approx I_f$$

Now $I_{in} = V_{in}/R_b$ and $V_{out} = A V_{in}$ 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 kΩ. 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 $I_f$. But $I_f = V_{out}/R_b$.

$$V_{out} = R_b$$

(1)
For the chosen numerical values
\[
\frac{V_{\text{out}}}{I_{\text{in}}} = 50 \, \text{k}\Omega
\]
The circuit can also be used as a current amplifier. Because \(V_{\text{out}} = I_{\text{out}} R_e\) we have
\[
\frac{I_{\text{out}}}{I_{\text{in}}} = \frac{R_b}{R_e}
\]
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 \(V_{\text{in}}/R_s\). \(R_s\) 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 \(I_{\text{in}}\) in (1) and (2) we have
\[
\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{R_b}{R_e}
\]
(3)
\[
\frac{I_{\text{out}}}{V_{\text{in}}} = \frac{R_b}{R_e}
\]
In the numerical example the input resistance was 250 \(\text{k}\Omega\). \(R_s\) could then be 2.5 \(\text{k}\Omega\).

From (3) the voltage gain is 20. From (4) the effective mutual conductance is 4 \(\text{mA/V}\).

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 \(I_{\text{R}}\) (proportional to output current) developed across \(R_e\) is applied between base and emitter of the transistor. This voltage is amplified by a factor \(g_{mR_e}\) where \(r_e\) is the collector a.c. resistance of the transistor if, as assumed in calculations of output resistance, the output load is infinite. Thus
\[
V_{\text{out}} = g_{mR_e} I_{\text{R}} R_e
\]
and
\[
\text{output resistance} = \frac{V_{\text{out}}}{I_{\text{out}}} = g_{mR_e} R_e
\]
\(g_{mR_e}\) is normally greater than unity,confirming the effective increase in output resistance. As \(r_e\) is commonly of the order of 200 \(\text{k}\Omega\) for a silicon transistor, this type of feedback can give very high output resistances.

At the input circuit
\[
V_{\text{in}} = V_{\text{ib}} + V_a
\]
and if \(R_e\) is large enough \(V_{\text{ib}}\) is large compared with \(V_a\),
\[
V_{\text{in}} \approx V_{\text{ib}}
\]
Now \(V_{\text{ib}} = I_{\text{b}} R_e = (h_{fe} + 1) I_{\text{b}} R_e\). Thus the input resistance is given by
\[
\frac{V_{\text{in}}}{I_{\text{in}}} = (h_{fe} + 1) R_e \approx h_{fe} R_e
\]
The input resistance is given by \(h_{fe} R_e\) approximately and, for a transistor with \(h_{fe} = 150\) and \(R_e = 1 \, \text{\Omega}\), is equal to 150 \(\text{k}\Omega\).

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 resistance the amplifier is well suited for use with an input voltage and an output current and we are particularly interested in the value of \(l_{\text{out}}/l_{\text{in}}\). i.e. the mutual conductance \(g_{m}\).

We have already shown that if the degree of feedback is large \(V_{\text{in}}\) is approximately equal to \(V_{\text{ib}}\). But \(V_{\text{ib}} = I_{\text{b}} R_e \approx I_{\text{R}} R_e\).
\[
\frac{I_{\text{out}}}{V_{\text{in}}} = \frac{1}{R_e}
\]
(5)

If, as assumed earlier, \(R_e\) is 1 \(\text{\Omega}\), the effective mutual conductance is 1 \(\text{mA/V}\).

If the amplifier is required to give a voltage output, the low output resistance necessary is provided by \(R_e\), which can be given a suitably low value such as the 5 \(\text{k}\Omega\) assumed earlier. If, however, the amplifier is required to give a current output, \(R_e\) 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, \(R_e\) can be taken as representing the low input resistance of the following stage.

When \(R_e\) is present \(V_{\text{in}} = I_{\text{out}} R_e\) and thus, as a voltage amplifier, the gain of the circuit is given by
\[
\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{R_e}{R_e}
\]
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 \(R_e\) is connected across the input terminals as shown dotted in Fig. 5(b). Resistor \(R_e\) should be small compared with the input resistance of the amplifier so that \(V_{\text{ib}} = I_{\text{b}} R_e\). Resistor \(R_e\) 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 \(I_{\text{out}} R_e\) for \(V_{\text{in}}\) in (5) and (6) we have
\[
\frac{I_{\text{out}}}{I_{\text{in}}} = \frac{R_p}{R_e}
\]
(7)
\[
\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{R_p R_e}{R_e}
\]
(8)
If \(R_p\) is 5 \(\text{k}\Omega\), the current gain is 5 and the transfer resistance 25 \(\text{k}\Omega\).

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 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 \(R_e\) 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 V. For good stabilization, the voltage at $R_1 R_2$ junction should be large compared with changes in base-emitter voltage due to temperature and transistor tolerances. For a silicon transistor, $V_T$ is commonly 0.7 V and thus the voltage across $R_3$ is 0.7 V less than that at the potential-divider junction. A suitable voltage across $R_3$ is 7 V and this gives the value of $R_3$ as 7 kΩ. The voltage across $R_2$ and the transistor is thus 17 V and, to permit the greatest output voltage swing, this should be shared equally between them. Thus, the no-signal voltage across $R_1$ is 8.5 kΩ and $R_1$ should be 8.5 kΩ.

The voltage gain of the stage is given by $R_1 / R_2$, and to give the required value of 20, $R_1$ should be 420 Ω. Thus, for zero-frequency feedback $R_1$ should be 7 kΩ and for signal-frequency feedback it should be 420 Ω. This can be achieved as shown in Fig. 7 by constructing $R_1$ of 420 Ω and $6580 \Omega$ 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 $h_in R_1$, 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 $I_h / h_{pr}$, i.e. 7 mA and a convenient value for the potential divider current is 100 mA. $R_2$ thus carries a current of 0.1 mA and the voltage across it is 7.7 V: the resistance is therefore 77 kΩ. $R_1$ carries 0.017 mA and the voltage across it is 27.7 V, i.e. 1.5 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 $R_e$, 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 $h_{pr}$ 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 $\pm 50\% h_{pr}$ gives only a $\pm 2\%$ spread in collector currents.

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

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 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 $Tr_1$

$$V_{in} / I_{in} = R_{b1}$$

and for $Tr_2$ from (5)

$$V_{out} / I_{out} = R_{b2}$$

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

$$I_{in} = 1 / R_{e2}$$

If, as assumed earlier, $R_{b1} = 50$ kΩ and $R_{b2} = 1$ kΩ the current gain is 50.

Stability considerations

The signal at $Tr_1$ collector is a copy of that at $Tr_2$ base which is directly connected to $Tr_1$ collector. Thus the performance of the transistor 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 effective means of stabilizing the mean collector current of both transistors.

If $R_{b1}$ is returned to a potential divider $R_1 R_2$ included in addition to $R_2$, in $Tr_2$ emitter circuit its value can be reduced whilst keeping the gain $(R_1 / R_2)$ 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 $(R_{b1})$ 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, $R_{b2}$ 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 V. We can then calculate the current across $R_1$ as 0.5 mA and the voltage across $R_2$ is 100 V. $Tr_1$ base current is 100 mA but because of the standing 0.7 V base-emitter voltage of silicon transistors, $Tr_2$ emitter voltage is 9.3 V. If $Tr_2$ is to take a mean emitter current of 2 mA, the total emitter resistance is 465 kΩ. Of this 100 Ω must provide signal-frequency feedback $R_{b1}$ being taken as 5 kΩ, and the balance could consist of two 22 kΩ resistors in series, decoupled, the centre point providing bias for $Tr_2$ base. The voltage across $R_{b1}$ caused by $Tr_1$ base current can be neglected and thus we can say that $Tr_2$ base voltage is approximately 4 V. Because of the voltage across $Tr_2$ base-emitter path, the emitter voltage can be taken as 40 V. The emitter resistance is thus 8 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 schematic form in Fig. 10) suitable for voltage amplification. At the inter-transistor coupling circuit the high output resistance of \( T_1 \) feeds into the low input resistance of \( T_2 \). There is no interaction provided the inter-transistor signal is taken as a current. For \( T_1 \) from (5)

\[
\frac{I_{\text{out}}}{V_{\text{in}}} = \frac{1}{R_{e1}}
\]

For \( T_2 \) from (1)

\[
\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{R_{b2}}{R_{e2}}
\]

But \( V_{\text{in}} = I_{\text{out}} \)

\[ R_{b2} \]

is normally connected to \( T_1 \) emitter instead of to \( T_2 \) 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 \( T_2 \) is (provided \( R_{e1} \) is large enough) the output current of \( T_1 \) and this is also the emitter current of \( T_2 \). Thus any current injected into \( T_1 \) emitter by \( R_{b2} \) is conveyed to \( T_2 \) 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 \( R_{b2} = 50 \Omega \) 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 \( T_2 \) emitter to \( T_1 \) base can be used as in the two-stage current amplifier and very similar component values can be used also. The emitter circuit of \( T_2 \) should be fully decoupled because no signal-frequency feedback is required here. In \( T_1 \) 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 \( R_{e1} \) provided the zero-frequency feedback necessary for stability and the signal-frequency feedback required to give the desired gain. In this voltage amplifier \( R_{e1} \) provides only zero-frequency feedback and \( R_{b2} \) provides signal-frequency feedback.

The input resistance of transistor \( T_1 \) is 50 k\( \Omega \) if \( \beta_f \) is taken as 100 and \( R_{e1} = 500 \Omega \) but \( R_{e1} \) is in parallel with the base circuit. \( R_{e1} \) 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 \( R_{b2}/\beta_f \) i.e. 500 \( \Omega \) if \( \beta_f = 100 \).

Some departures from the calculated resistor values indicated in Figs. 7, 9 and 11 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 \( T_1 \) 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 \( \mu F \). For a.f. applications the decoupling capacitors should be large enough to perform adequately down to 30 Hz.

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, HA0 4PE, Middx., have been appointed sole U.K. representatives and distributors for Telefunken semiconductors.

LST Components, 7 Copworth 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.

Semicoms Ltd, 5 Northfield Estate, Bresciafs Avenue, Wembley, Middx, have been appointed sole U.K. distributors for the IMC range of dual-in-line i.e. 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.

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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 Hockeath House, Salisbury Road, Hunslow, 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 Defence. The order is for over 1,000 military radio telephones with accessories and spares.

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.

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 3os per course.

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The Broadca