

(a) the minimum gain is unity, not zero, and
(b) a floating signal source is required.
Disadvantage (b) is of little consequence when an input transformer is used, and (a) may be overcome by taking the signal output from the pot. slider. The latter change, of course, sacrifices the virtue of very low output impedance possessed by the Fig. 4 version.

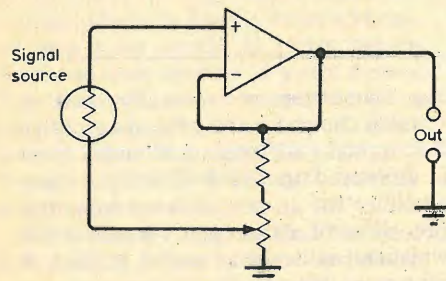


Fig. 4. Simple feedback gain control.

In assessing the pros and cons of various circuits, it is very helpful to appreciate the relationships between the circuits in the most vivid possible way, rather than relying purely on formal analysis. Very often the differences between circuits are much smaller than they appear to be, involving merely the choice of earthing point and/or the way of drawing the circuit diagram, rather than differences of more fundamental significance. Sometimes, in redrawing circuits employing op. amps. to facilitate better understanding of them, it is helpful to replace the op. amp. symbols by ordinary single-transistor symbols — an unfamiliar-looking circuit may then suddenly be recognised as an old friend! At other times, replacing a detailed transistor circuit by the op. amp. equivalent may reveal its true nature in the best way.

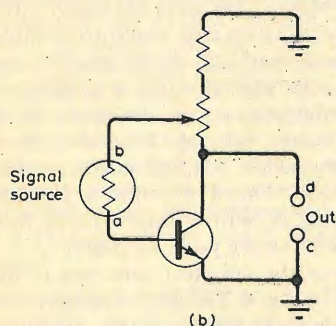
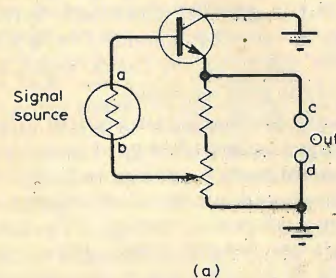


Fig. 5. Single transistor equivalent to Fig. 4, neglecting d.c. conditions. Rearrangement in (b) shows circuit to be easily recognizable.

On replacing the op. amp. in Fig. 4 by a transistor, the circuit of Fig. 5(a) is obtained. Though the collector would in practice be taken to a positive supply line, it is here shown as earthed, for in the present context we are concerned only with a.c. aspects and it is best to omit irrelevant details.

Shifting the earthing point to the emitter of the transistor, but making no other changes, leads to Fig. 5(b), which is a simple common-emitter amplifying stage with adjustable feedback.

If the output in the Fig. 4 circuit is taken from the pot. slider instead of from the point shown, then the circuit, redrawn with a transistor in place of the op. amp., is as in Fig. 6(a). Merely shifting the earthing point to the pot. slider then yields the circuit of Fig. 6(b). It is now evident that moving the slider to the right has two separate effects — it increases the amount of resistance in the emitter lead, thereby increasing the amount of negative feedback, and it reduces the collector load resistance. Both these effects contribute to reducing the gain, which becomes zero with the slider fully to the right.

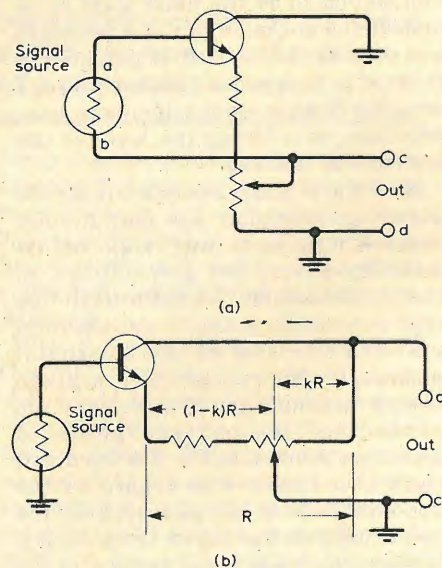


Fig. 6. Fig. 4. Circuit with output taken from pot. slider and rearranged at (b) to show dual function — varying emitter resistance and varying feedback.

Employing just a single transistor, as in Fig. 6(b), will give a noise performance which is inferior to that achievable with more elaborate arrangements. This is largely because the resistance inserted in the emitter lead is itself a generator of Johnson noise, which is effectively added in series with that generated in the internal resistance of the signal source. The transistor d.c. operating current must be chosen in relation to the source impedance, for good noise performance, and it will then be found that to obtain a substantial reduction of gain by inserting emitter resistance, the amount of resistance needed will give considerable degradation of the noise figure.

The above noise difficulty may be

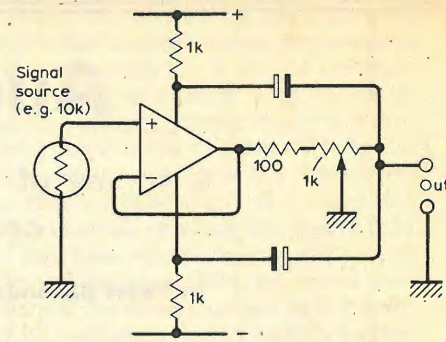


Fig. 7. Fig. 6(b) using an op. amp.

solved by replacing the single transistor by a suitable pair or triple, having a much higher mutual conductance than the single transistor but whose input stage operates at a similarly low current. The increased mutual conductance and output current permit the resistance values associated with the gain-control pot. to be made much lower, with a correspondingly reduced effect on the noise performance at low gain settings. The well-known configurations for pairs and triples as used in audio class 'B' output stages may be adapted to the present application, but an interesting alternative is that shown in Fig. 7. Here the supply connexions to the op. amp. are used as the equivalent of the transistor collector in the Fig. 6(b) circuit — a way of using an op. amp. which perhaps deserves to be more widely borne in mind.

Assuming infinite mutual conductance, the voltage gain of the Fig. 6(b) idealized circuit is simply $k/(1-k)$. Expressing this in decibels gives the graph of Fig. 8(a). The Fig. 8(b) graph is a measured one for the circuit of Fig. 7.

With the idealized circuit of Fig. 6(b), unity gain occurs when the pot. is set for $k=0.5$, and the curve is quite symmetrical about this centre point. With the Fig. 7 circuit, however, the curve is not symmetrical about the unity-gain point. This is because the right-hand part of the pot. is shunted by the parallel value of the two $1k\Omega$ resistors going to the supply lines.

Another very simple feedback gain-control circuit is shown in Fig. 9. With high forward gain in the op. amp. itself, this circuit gives a gain, between the input and output terminal pairs shown, accurately equal to $k/(1-k)$. (This formula, as for the Fig. 6(b) case, may be prefixed by a minus sign if it is desired to allow for the fact that phase inversion occurs.)

The Fig. 9 circuit, unlike those previously discussed, has the feature that the current in the gain-control resistance chain is supplied by the signal source. This makes it impossible to achieve a good noise figure over a wide range of gain adjustment, no matter how the resistance values are chosen in relation to the signal-source impedance. That this must be so can best be understood as follows. Negative feedback as such never has any effect on the signal-

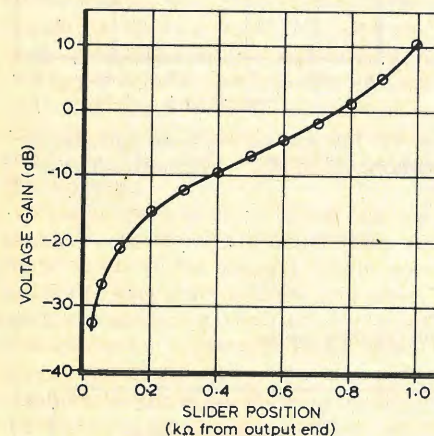
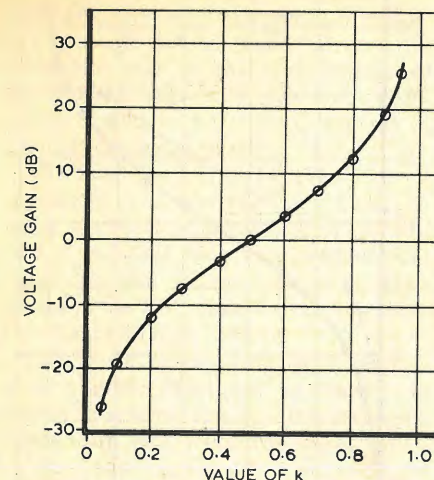


Fig. 8. Gain variation of Fig. 6(b) circuit is at (a). Measured performance of equivalent op. amp. circuit of Fig. 7 is shown at (b).

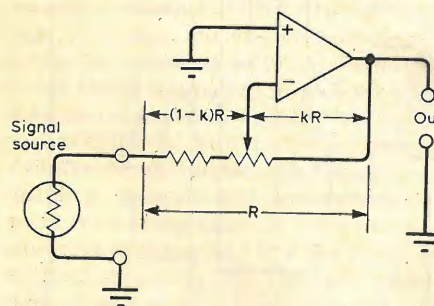


Fig. 9. Feedback gain-control circuit, which has disadvantage of source-fed resistor chain, giving poor noise figure over wide range.

to-noise ratio, at a given frequency, of an amplifier circuit to which it is applied, though the resistors introduced for the purpose of providing the feedback may do so. Thus the output signal-to-noise ratio of the Fig. 9 circuit is the same as that of the circuit shown in Fig. 10. If R is made low, say equal to the internal resistance of the signal source, it will degrade the signal-to-noise ratio at the source terminals*, whereas if R is made much higher, a

*When a resistive source of internal resistance R is shunted by a load resistance equal to R , the signal voltage is halved, but the Johnson noise voltage is reduced by a factor of only 2. The signal-to-Johnson-noise ratio is therefore worsened by 3dB.

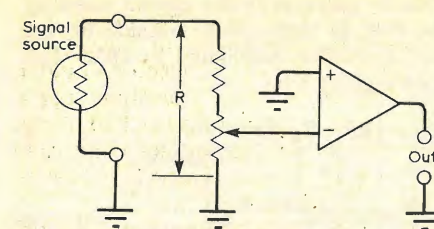


Fig. 10. Circuit of Fig. 9 gives same noise performance as circuit shown here.

large amount of resistance is introduced into the op. amp. input circuit at intermediate slider settings, with correspondingly large Johnson noise and maybe noise from the op. amp. equivalent current-noise generator.

Comparing Fig. 10 with Fig. 1(b) might suggest that the Fig. 9 circuit is no better than that of Fig. 1(b) as regards noise performance. This is not so, however, for to effect a given number of decibels reduction of gain below maximum, the slider in Fig. 9 has to be moved a smaller fraction of the way from the signal-source end of R than is necessary for the same gain reduction in the Fig. 1(b) circuit. The noise performance of Fig. 9 is better than that of Fig. 1(b), but is nevertheless not very good.

Another feature of the Fig. 9 circuit which makes it undesirable for some applications is that the loading of the signal source varies with the pot. setting. If the signal source has a complex internal impedance, the overall frequency response will vary with the gain setting.

This undesirable characteristic of the Fig. 9 circuit may, to a large extent, be overcome by inserting an emitter-follower (or op. amp. follower) between the signal-source and the left-hand end of the resistance chain. With a $50k\Omega$ signal-source, for example, R could be made about $5k\Omega$, giving reduced Johnson noise from R but nevertheless subjecting the signal-source to negligible loading.

As already mentioned, the Fig. 9 circuit as it stands produces the gain-control characteristic shown in Fig. 8(a), which is symmetrical about the unity-gain point. Over a range of about 30dB, and using an ordinary linear pot., the scale shape obtained approximates fairly reasonably to the desirable one having uniformly-spaced decibel divisions, though for many applications a gain of more than unity would be preferred at the centre of this control range. The modification shown in Fig. 11 provides an increased gain at the point of inflexion of the control characteristic, but has the weakness that the gain cannot be reduced right down to zero. Provided R_a and R_b are made much lower in value than the pot. resistance, however, the minimum gain may be made sufficiently low for many purposes.

If a stud type pot. is used, and assuming there is complete freedom in the choice of its law and total resistance

value, the Fig. 11 modification gives no advantage, the required performance being obtainable with better economy of components by adopting the Fig. 9 arrangement.

The circuit of Fig. 12 possesses a combination of several good features. It employs only one op. amp., has a high input impedance, the feedback network can be of low resistance for good noise performance; and the values of R_a and R_b can be chosen, in relation to R , to make the point of inflexion in the control characteristic occur at a gain of much greater than unity, as sometimes desired.

Analysis shows that the gain of the Fig. 12 circuit is given by:

$$\frac{V_{out}}{V_{in}} = \frac{R + \frac{R_b}{1-k}}{R + \frac{R_a}{k}} \quad 1.$$

$$\text{or } \frac{V_{out}}{V_{in}} = \frac{k}{1-k} \times \frac{R(1-k) + R_b}{R_k + R_a} \quad 2.$$

Thus, if R_a and R_b are each much greater than R , the gain is approximately proportional simply to $k/(1-k)$, and is approximately equal to R_b/R_a when $k=0.5$. Thus the control characteristic is fairly closely as in Fig. 8(a) but shifted upwards. For lower values of R_a and/or R_b the characteristic is of modified form, covering a smaller number of decibels with reasonable linearity.

The curve shown in Fig. 13 is the result of a measurement using the Fig. 12 circuit with the following values:

$$R = 1k\Omega, R_a = 330\Omega, R_b = 3.3k\Omega$$

Comparison of this curve with Fig. 8(a) shows that it gives a poorer

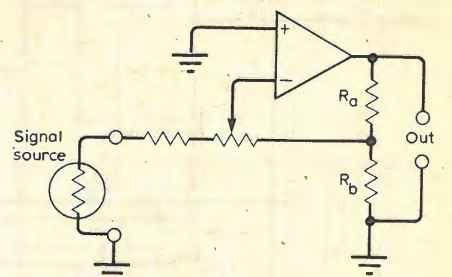


Fig. 11. Variation of Fig. 9, giving increased gain at halfway position of slider.

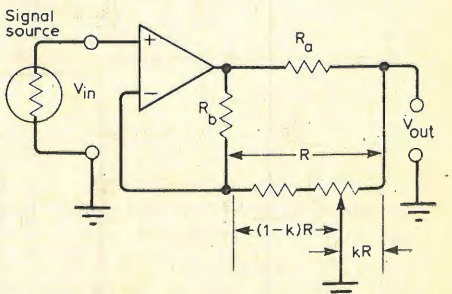


Fig. 12. Circuit featuring only one amplifier, high input impedance, low-resistance feedback chain for low noise and flexibility in choice of inflexion point.