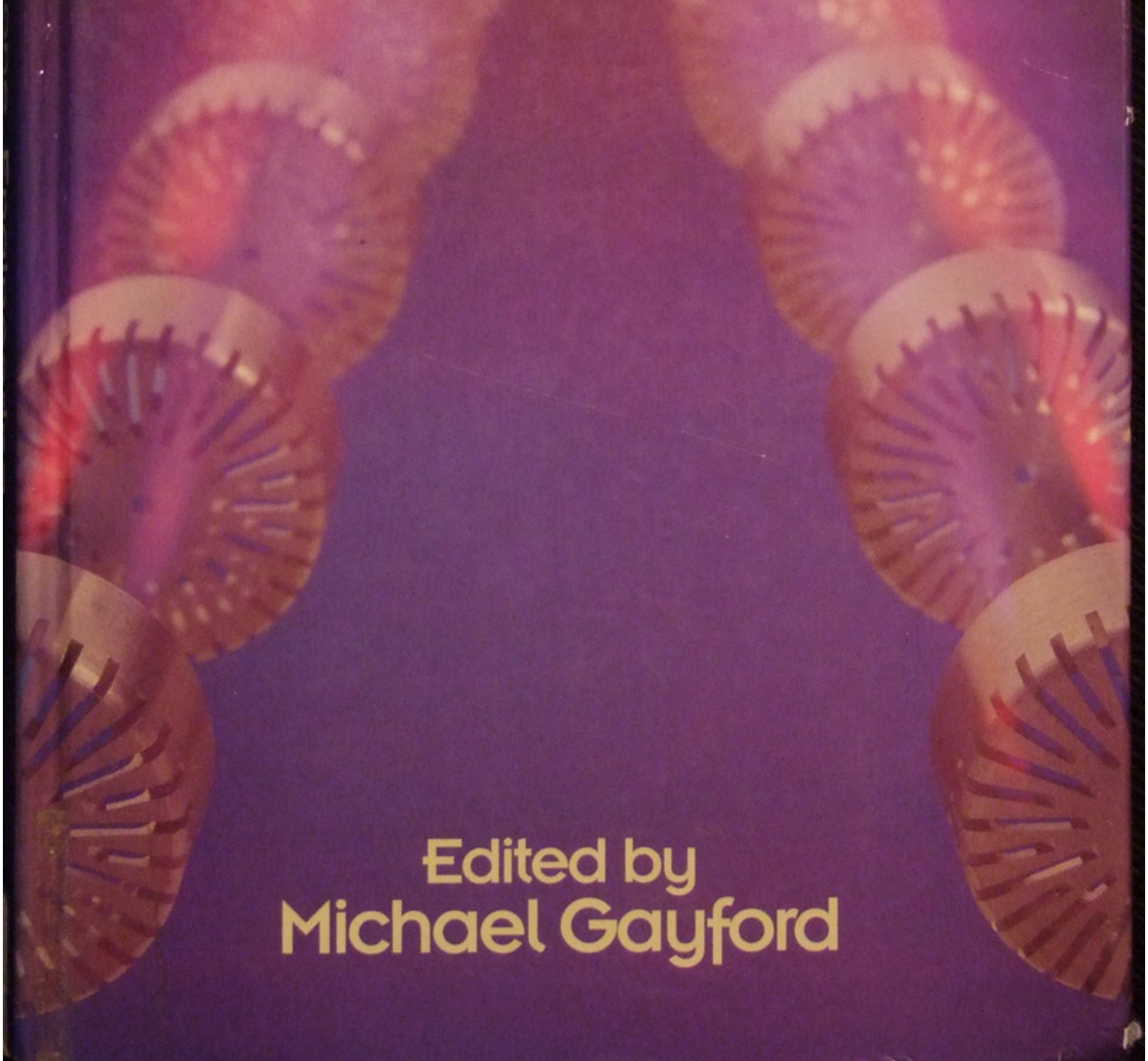
Microphone Engineering Handbook



Edited by
Michael Gayford

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For complete information the frequency dependence of R_{N_v} and R_{N_i} should ideally be included, though the variation in R_{N_v} is usually negligible. In practice, however, single broad-band or spot-frequency values are normally quite adequate.

An alternative and widely used method for presenting the noise performance of a microphone amplifier, or amplifying device, is based on the notion that with some specific value of passive source resistance R_S connected to the input, the total noise output voltage magnitude may be accounted for in terms of a single equivalent input noise (EIN) voltage of white spectrum acting in series with the source.

A change in the value of R_S alters the magnitude of the EIN voltage, both because of the different Johnson noise voltage generated by R_S and also because the value of R_S affects the magnitude of the contribution from the amplifier's current-noise generator – see Figure 8.7. However, the total output noise may still be regarded as being due solely to a single EIN voltage, now of modified value, acting in series with R_S.

The EIN voltage itself may be expressed in several different ways.

It may be given in $nV/\sqrt{Hz}$, or as a noise voltage in a stated bandwidth. Figure 8.14(a) and (b), based on the data sheets for the NE5534 op. amp., embody these two methods, the descriptions above them being those used by the manufacturer.

Figure 8.14(c) is taken from the data sheets for the Trans-Amp balanced-input microphone amplifier module made by Valley People Inc., the EIN voltage here being expressed in dBu, though labelled 'dBv (dB re 0.775 V)' on the American data sheet.

It is evident from the definition of EIN voltage that if the amplifier itself was noiseless, then the EIN voltage would be just the Johnson noise voltage in the source resistance R_S, and it is helpful to include graphs, shown in broken-line, depicting this ideal performance. The vertical spacing between the device curve and the corresponding broken-line graph, at any given value of R_S, then gives the noise figure for that R_S value.* The excellent noise performance of the Trans-Amp when fed directly from a 200 Ω moving-coil microphone will be noticed, this unit having a considerable number of transistors in parallel in each half of its input stage.

8.4 Electronically balanced input circuits and noise

Though microphone amplifier circuits with transformerless balanced inputs differ considerably in design details, the circuit shown in Figure 8.15(a) is representative of good contemporary practice.[25] The value of R_G is varied to provide gain adjustment, sometimes in 10 dB steps. Additional components, here omitted for clarity, are often included to reduce the sensitivity to r.f. interference and to control the rate of attenuation of loop gain so as to give good stability margins.

The resistors labelled R' serve to hold the input circuit at the right d.c. level, but are normally made of sufficiently high value to avoid significant shunting of the signal source.

The negative feedback from each op. amp. output, via R_{fb_1} or R_{fb_2} to the emitter of the associated transistor, ensures that the two collector voltages are at all times

*For (a) and (b) in Figure 8.14, NF = 20 log (noise voltage ratio), whereas in (c) the spacing may be related directly to the decibel scaling.

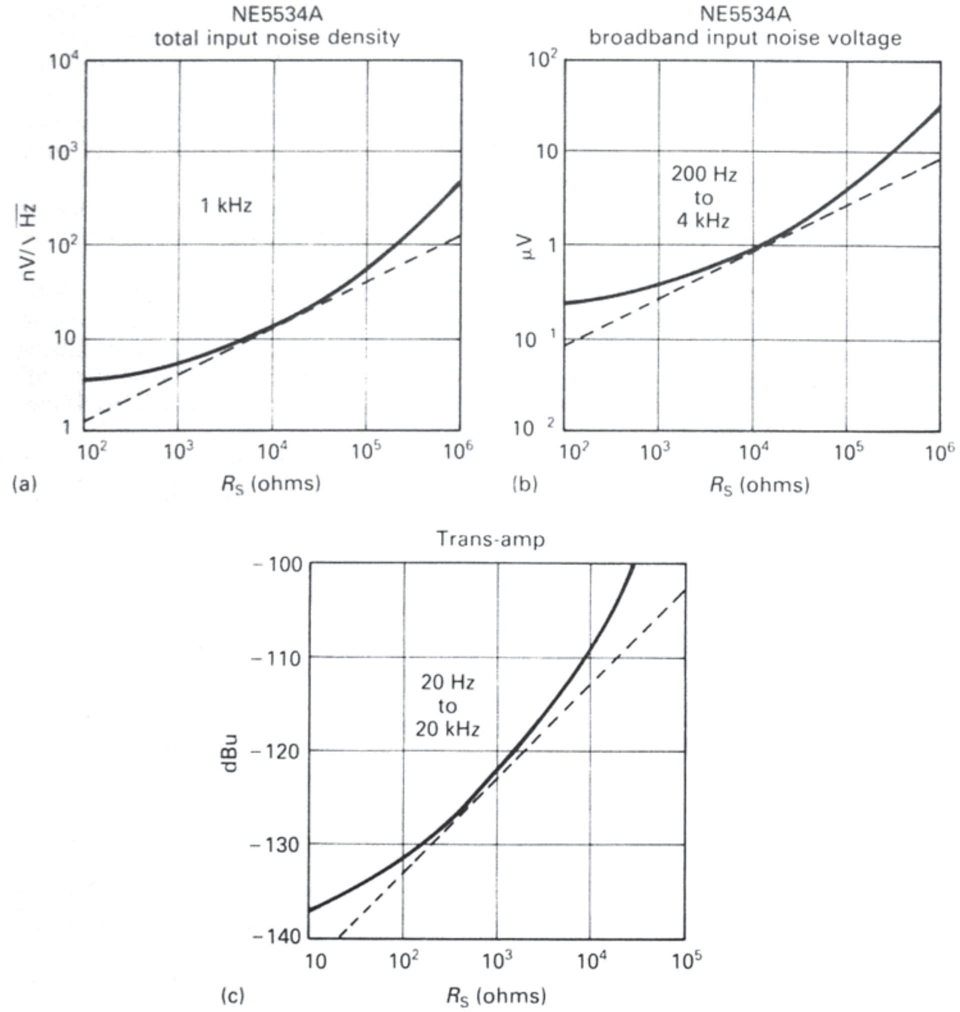

Figure 8.14 *Equivalent Input Noise (EIN) voltage data (20°C). The broken-line graphs represent the source Johnson noise alone.*

held very close to the d.c. bias voltage level $+V_B$, thus keeping the instantaneous collector currents very nearly constant.

When a common-mode input voltage V_{com} is applied to input terminals 1 and 4, voltage changes virtually equal to V_{com} are caused to appear also on points 3 and 6, thus keeping constant the voltages across R_{fb_1} and R_{fb_2} and hence the transistor currents.

The equal voltages on 3 and 6 are applied as a common-mode input to the balun stage involving op. amp. 3, giving zero output from it if the resistor values are accurately matched. Thus the complete Figure 8.15(a) circuit has ideally zero response to common-mode inputs. No voltage appears across R_G for such inputs.

When a balanced input voltage V_{bal} is applied between terminals 1 and 4, the feedback again holds the transistor currents virtually constant, and to achieve this it is now necessary for the feedback circuit to produce a voltage change between the emitters equal to V_{bal}. Hence

$$V_{3\text{-to-}6} \times \frac{R_G}{2R_{fb} + R_G} = V_{bal} \tag{8.17}$$

where $R_{fb_1} = R_{fb_2} = R_{fb}$.

The voltage between points 3 and 6 is subjected to an inverting gain of unity in the balun stage, the final output voltage therefore being given by

$$V_{out} = -V_{bal} \times \frac{2R_{fb} + R_G}{R_G} \tag{8.18}$$

The circuitry within each broken-line enclosure in Figure 8.15(a) may be regarded as constituting an op. amp. with modified characteristics, terminal 1 of the left-hand circuit being the non-inverting input, terminal 2 the inverting input and terminal 3 the output. Representing these modified op. amps. by simple broken-line triangles enables the Figure 8.15(a) circuit to be drawn as at (b).

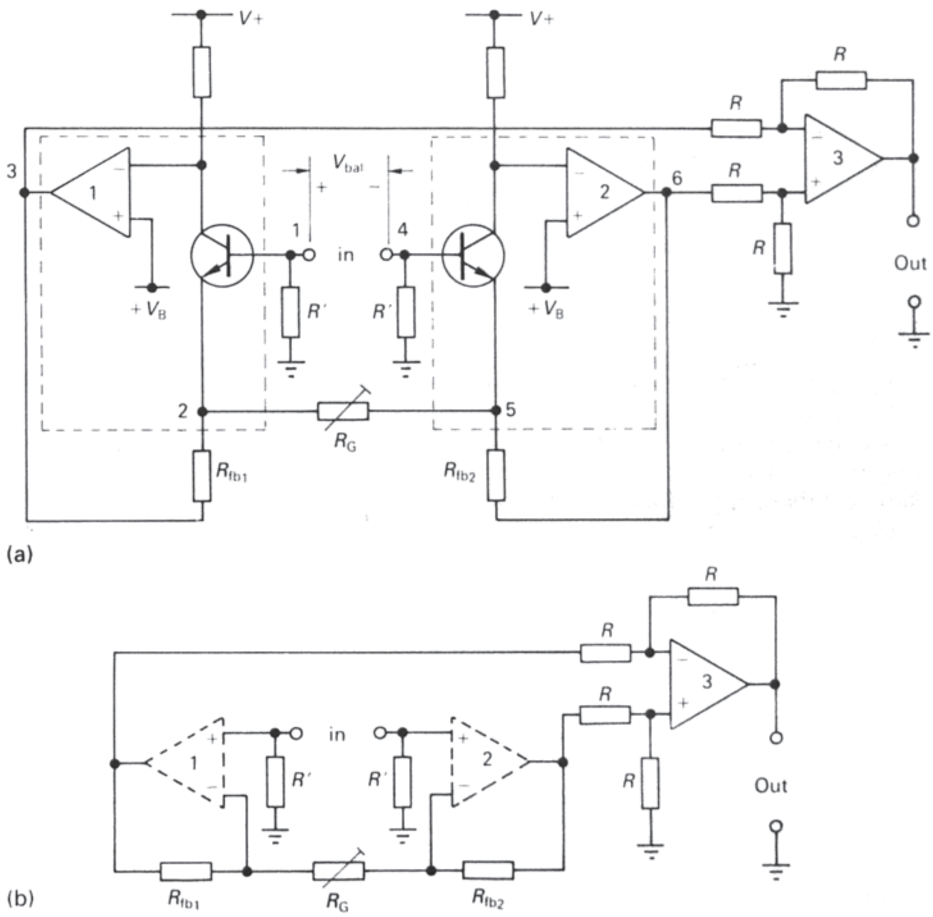

Figure 8.15 *(a) Basic circuit of a modern transformerless microphone amplifier with balanced input; (b) simplified version of (a).*

A satisfactory practical microphone amplifier, suitable for feeding straight from 200 Ω moving-coil and ribbon microphones, could indeed be directly based on Figure 8.15(b), op. amps. such as the LT1028 (see Table 1), which have low voltage-noise, being used in positions 1 and 2.

In recent times, however, a single op. amp., type SSM2015P, intended specifically as a balanced-input microphone amplifier, has become available and has internal circuitry in broad conformity with Figure 8.15(a). This product provides a neat, economical and generally highly satisfactory solution to the microphone amplifier problem, tending, indeed, to render most other solutions somewhat obsolescent.

The noise aspect of electronically balanced microphone amplifiers is interesting and demands careful consideration.

Referring again to Figure 8.15(a), it will be assumed that the input transistors provide sufficient gain to make it reasonable to ignore the noise produced by the op. amps. Thus the significant noise sources are the transistor voltage and current noise generators, Johnson noise in the feedback network, and source Johnson noise.

The high-value resistors R' exert a minor shunting effect across the source, which only very slightly degrades the noise performance, and they also produce a common-mode noise input, to which, however, the system is non-responsive. Hence the influence of these resistors on the noise output can be neglected in practice.

Figure 8.16 represents the essential elements of Figure 8.15 from a noise point of view.

Considering voltage noise first, it is evident that there will be introduced in series with the input circuit, in a balanced manner, a total Johnson noise voltage corresponding to a resistance value of $R_{N_{v1}} + R_{N_{v2}} + R_S + R'_G$, where R'_G is the parallel value of R_G and $(R_{fb_1} + R_{fb_2})$.

R_{fb_1} and R_{fb_2} are made as low in value as is practicable, consistently with the ability of the op. amps. to drive them to a sufficient output level when R_G is set to a low value. Thus the Johnson noise contribution from the feedback network is kept minimal.

The total Johnson noise voltage from the above-mentioned four resistances in series is subjected to a gain of $(R_{fb_1} + R_{fb_2} + R_G)/R_G$.

The noise current generators I_{N_1} and I_{N_2} together produce both a common-mode component of current noise, to which the complete circuit is non-responsive, and also a balanced component flowing round the input circuit. This latter component produces voltages across R_S and R'_G which give rise to a noise

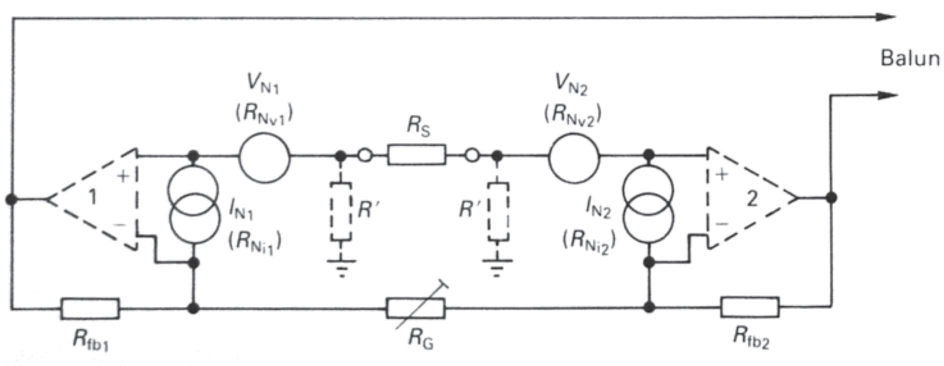

Figure 8.16 *Diagram for explanation of noise performance of Figure 8.15*

output. It is therefore necessary to consider the derivation of these separate noise current components.

Let the instantaneous values of I_{N_1} and I_{N_2} be i_{n_1} and i_{n_2} respectively, both being taken as positive when flowing upwards, say. Whatever the values of i_{n_1} and i_{n_2} may be, another pair of current components i_a and i_b can always be chosen so that, at the time instant concerned, the following relationships are satisfied:

$$i_{n_1} = i_a + i_b \tag{8.19}$$

$$i_{n_2} = i_a - i_b \tag{8.20}$$

These are two equations with two unknowns, so that given the i_{n_1} and i_{n_2} values at any instant, i_a and i_b can be determined.

It is evident that i_a constitutes the common-mode current component, flowing in the same direction in both current generators, whereas i_b constitutes the balanced component which flows round the circuit through R_S and R'_G.

Adding and subtracting equations (8.19) and (8.20) gives

$$i_a = \frac{i_{n_1} + i_{n_2}}{2} \tag{8.21}$$

$$i_b = \frac{i_{n_1} - i_{n_2}}{2} \tag{8.22}$$

Now the noise waveforms of which i_{n_1} and i_{n_2} in equation (8.22) are instantaneous values each have an r.m.s. value of I_N, it being assumed that $I_{N_1} = I_{N_2} = I_N$ in Figure 8.16. However, since I_{N_1} and I_{N_2} are uncorrelated, the minus sign in equation (8.22) is now of no consequence and the r.m.s. value of the right-hand side of the equation is $\sqrt{2}I_N/2$, i.e. the r.m.s. value of the balanced noise current flowing round the circuit due to the noise current generators is $I_N/\sqrt{2}$. The effective value of equivalent current-noise resistance for the complete circuit is therefore $2R_{N_i}$, where R_{N_i} is the value for a single transistor.

From the above, assuming equal R_{N_v} and R_{N_i} values for the two transistors and ignoring the small noise effect of the R' resistors, the noise performance of the Figure 8.16 balanced circuit is the same as that of the unbalanced circuit shown in Figure 8.17(a). Comparison with Figure 8.7 and reference to equations (8.9), (8.10) and (8.11) shows that, for the Figure 8.17(a) circuit

$$\mathrm{NF} = 10 \log_{10}\left[1 + \frac{2R_{N_v} + R'_G}{R_S} + \frac{R_S}{2R_{N_i}}\right] \tag{8.23}$$

$$R_{S_{opt}} = \sqrt{[(2R_{N_v} + R'_G)(2R_{N_i})]} \tag{8.24}$$

$$\mathrm{NF}_{opt} = 10 \log_{10}\left[1 + 2\sqrt{\frac{2R_{N_v} + R'_G}{2R_{N_i}}}\right] \tag{8.25}$$

When, as is often the case, the feedback network resistance R'_G is negligibly small, we then have

$$\mathrm{NF} \approx 10 \log_{10}\left[1 + \frac{2R_{N_v}}{R_S} + \frac{R_S}{2R_{N_i}}\right] \tag{8.26}$$

$$R_{S_{opt}} \approx 2\sqrt{(R_{N_v}R_{N_i})} \tag{8.27}$$

$$NF_{opt} \approx 10 \log_{10}\left[1 + 2\sqrt{\frac{R_{N_v}}{R_{N_i}}}\right] \tag{8.28}$$

It is seen from equations (8.27) and (8.28) that though $R_{S_{opt}}$ is twice as high as it would be for a single transistor of the same type, nevertheless use of this higher R_S value gives the same noise figure as for the single transistor.

It is of interest to determine what happens to the noise performance of the Figure 8.15 type of circuit if, instead of operating with a balanced signal source, one end of the source is earthed to give unbalanced operation.

For simplicity it will again be assumed that R'_G is negligibly small. In other words it is being assumed that the currents in the bottom leads of the two current generators in Figure 8.16 flowing into the feedback network produce negligible noise output, the significant effect in normal balanced operation being produced by the balanced component of current in their top leads flowing through R_S. However, when the left-hand end, say, of R_S is earthed, I_{N_1} is virtually prevented from contributing anything at all to the noise output, the observed current-generator-originated noise output now being due almost entirely to the voltage across R_S produced by the full value of I_{N_2} flowing through it.

Voltage noise from $R_{N_{v1}}$, R_S and $R_{N_{v2}}$ is of course still applied between the non-inverting inputs as before. The noise performance of the system is therefore approximately as for the circuit of Figure 8.17(b), and we now have

$$\text{NF} \approx 10 \log_{10}\left[1 + \frac{2R_{N_v}}{R_S} + \frac{R_S}{R_{N_i}}\right] \tag{8.29}$$

$$R_{S_{opt}} \approx \sqrt{(2R_{N_v}R_{N_i})} \tag{8.30}$$

$$NF_{opt} \approx 10 \log_{10}\left[1 + 2\sqrt{\frac{2R_{N_v}}{R_{N_i}}}\right] \tag{8.31}$$

Comparisons of equations (8.29), (8.30) and (8.31) with (8.26), (8.27) and (8.28) shows that earthing one end of R_S to give unbalanced operation has the following effects:

(1) If the value of R_S remains unchanged, the noise figure is worsened.
(2) The value of R_S required for optimum noise performance is reduced by a factor of $\sqrt{2}$.
(3) Even with R_S reduced as in (2), NF_{opt} is still not as good as with balanced operation.

These effects are borne out experimentally, but if the value of R_S used is well below $R_{S_{opt}}$, as is frequently the case in practice, the differences become quite negligible since voltage noise then tends to become dominant and is the same for balanced and unbalanced operation.

An interesting sidelight on the above relates to the ordinary use of op. amps., for example in the virtual-earth manner shown in Figure 8.18(a). The long-tailed-pair input stage of the op. amp. is here being operated in the unbalanced manner referred to above, and the noise performance obtainable is commensurate with the device's R_{N_v} and R_{N_i} (or nV/$\sqrt{}$Hz and pA/$\sqrt{}$Hz) values as normally quoted. However, by changing to the (b) scheme, op. amp. 1 input stage is now operated under proper balanced conditions and a noise performance somewhat better than given by the data sheet figures is obtained.

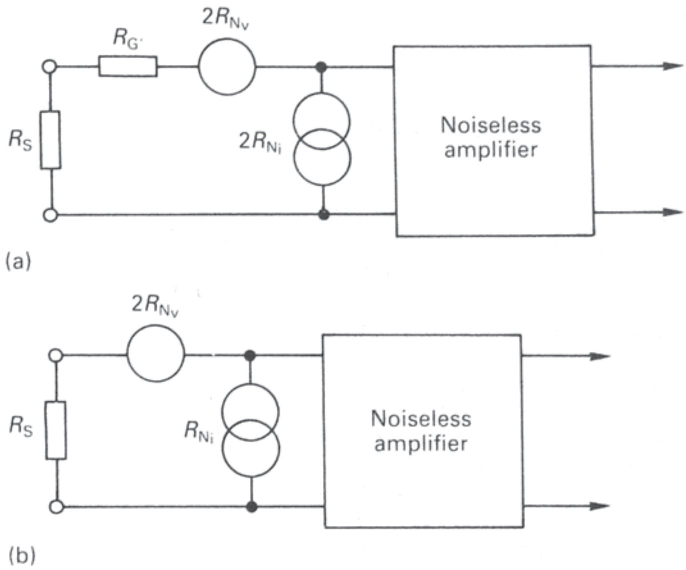

Figure 8.17 *(a) Circuit having the same noise performance as the circuit of Fig 8.16; (b) as for (a) but with one end of R_S in Figure 8.16 earthed. $R_{G'}$ in (a) is the small parallel value of R_G and (R_{fb1} + R_{fb2}), and was neglected in deriving (b).*

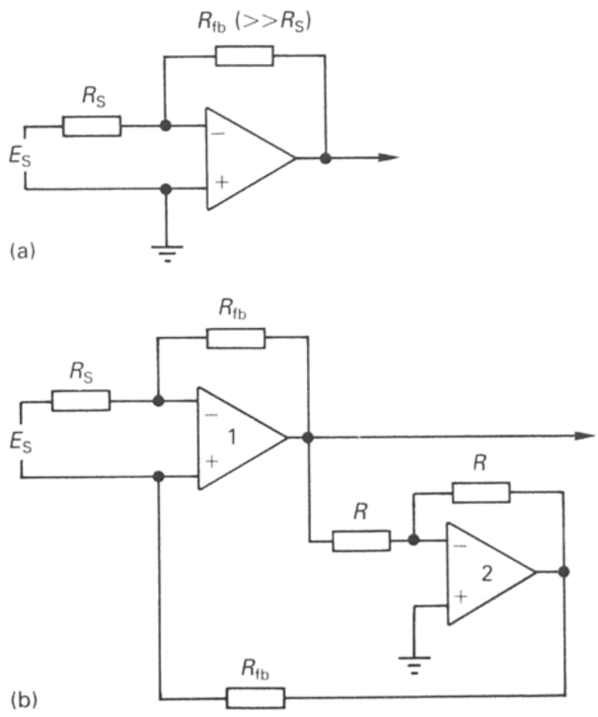

Figure 8.18 *Diagrams illustrating a point about op. amp. noise characteristics.*

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