Low-noise Audio Amplifiers

by H. P. Walker, M.Sc

Minimizing transistor noise-figure does not necessarily lead to an optimum low-noise design. In this article the effect of circuit configuration is included with a discussion of optimization with a complex source impedance, and a design procedure is given.

Much has been written on the optimization of transistor noise figure and while with very low and very high source impedances the transistor may be a severe limitation, in many audio amplifiers using modern low-noise transistors the circuit configuration and associated resistors can be the main factor determining signal-to-noise ratio. In this article both the effect of configuration and the limitations imposed by transistor noise are considered in achieving an optimum design.

Optimum configuration

High-quality audio amplifiers usually incorporate overall negative feedback to obtain gain stability and reduce distortion. Assuming that a low output impedance is required, there are two input configurations: series feedback, Fig. 1(a), and shunt feedback (virtual earth), Fig. 1(b), the former being voltage sensing and the latter current sensing.

The majority of audio input transducers are designed as voltage sources feeding an input resistance greater than the source impedance: for example, a magnetic pickup or tape head presents an inductive impedance which is less than the recommended 50-kΩ load up to frequencies of 10 to 15kHz. An exception would be a ceramic pickup driving a load resistance of say 200kΩ in a transistor pre-amplifier[1,2] where conditions approximate to current drive at low frequencies, the turnover occurring at around 1 or 2kHz.

When the source impedance is purely resistive, maximum signal-to-noise ratio is obtained when the input resistance is greater than the source resistance. This is a simple application of the well-known rule[3] 'no resistive attenuation before amplification' and it is clear that if the input resistance equals the source resistance, then Johnson noise* is reduced by 3dB and the signal attenuated by 6dB resulting in a 3-dB loss of signal-to-noise ratio.

Consider now the signal-to-noise ratio obtained for the two cases of Fig. 1. Identical frequency-dependent sources are loaded with an input resistance of R_{in} in both configurations. To examine the effect of configuration alone, assume that the amplifier A is noiseless, has infinite gain and high input impedance, thus creating a virtual earth at node 'E' in Fig. 1(b), and for Fig. 1(a), causing negligible loading at the input and therefore sensing the noise and signal voltage across R_{in}.

*The thermal noise generated by a resistive impedance in thermal equilibrium with the surroundings, and is $V_n = \sqrt{4kTR\Delta f}$ where $k = 1.38\times10^{-23}$ joules/deg C, T = absolute temperature in K, R = resistance in ohms, Δf is the noise bandwidth and V_n is the r.m.s. noise voltage.

The signal-to-noise ratio for the series circuit, derived in Appendix 1, at a frequency f for a bandwidth δf, is

$$\left|\frac{V_o}{V_n}\right| = \frac{V(f)}{\sqrt{4kT\delta f\cdot\left(\dfrac{|Z(jf)|^2}{R_{in}}+R(f)\right)}} \qquad (1)$$

assuming R_e is sufficiently small to contribute a negligible thermal noise voltage. Impedance $Z(jf)$ is the complex source impedance $R(f)+jX(f)$, where the resistive part $R(f)$ alone generates Johnson noise $\sqrt{4kT\cdot R(f)\cdot\delta f}$

The signal-to-noise ratio for the shunt circuit, derived in Appendix 1, is obtained by summing signal and noise currents at node 'E'

$$\left|\frac{V_o}{V_n}\right| = \frac{V(f)}{\sqrt{4kT\delta f[R_{in}+R(f)]}} \qquad (2)$$

assuming that the feedback resistor R_f is sufficiently high to contribute a negligible noise current $\sqrt{4kT\cdot\delta f/R_f}$ For practical purposes R_f should be made at least three times the impedance of the input arm.*

Comparing these two equations the only difference is in the bracketed terms in the denominator of each expression. The signal-to-noise ratio of the series feedback input will be greater than that of the shunt feedback input when

$$\frac{|Z(jf)|^2}{R_{in}} < R_{in} \quad \text{or when} \quad Z(jf) < R_{in}$$

That is, at a spot frequency f, the signal-to-noise ratio of the series circuit will be superior to the shunt feedback circuit if the source impedance at that frequency is less than the required input resistance. With a magnetic pickup or tape head, the source impedance is predominantly inductive and less than the input resistance at frequencies

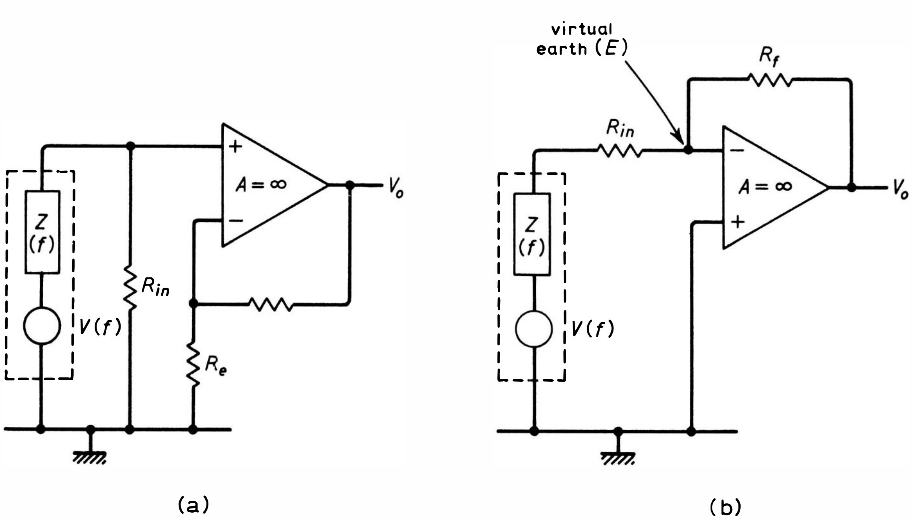
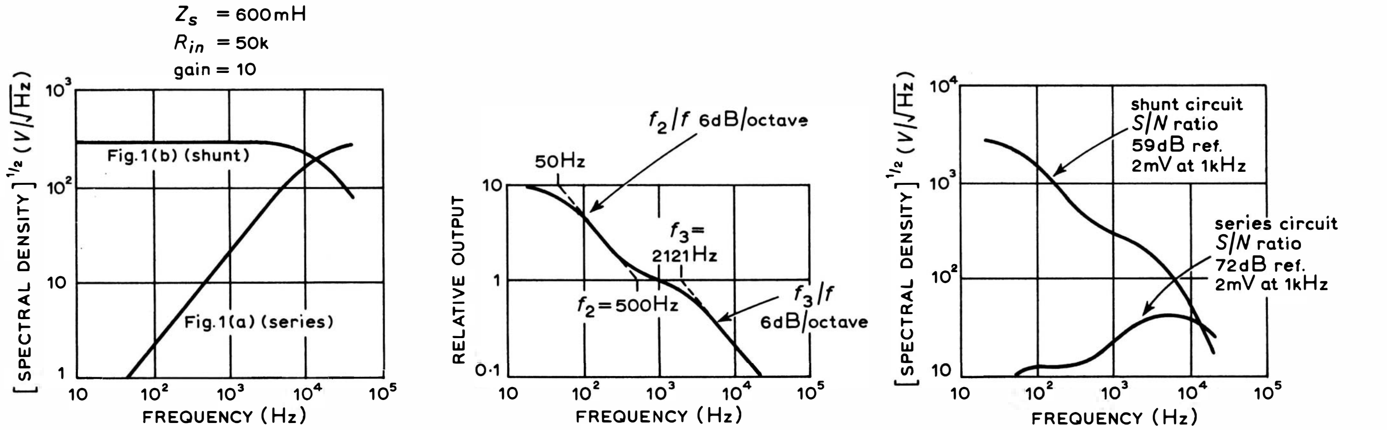


Fig. 1. Signal-to-noise ratio for series (a) and shunt (b) feedback arrangements, as derived in Appendix 1, is shown in graphical form in Fig. 2.

*This same condition applies to biasing resistors connected across the input of amplifier A. Although the connection of a low resistance from the summing junction 'E' to ground will have a negligible effect on the closed-loop gain, particularly with a high-gain amplifier, the signal-to-noise ratio will be seriously reduced because of the high noise current being injected into the virtual earth.



(a) (b) (c)

Fig. 2. Signal-to-noise ratio of series circuit is better over most of the band–illustrated in (a) where noise voltage per unit bandwidth falls at 6dB/octave for the series circuit. Curves shown at (c) apply after equalization curve at (b) has been applied. (Signal-to-noise ratio figures are derived in Appendix 2.)

below the turnover in the region 10 to 15kHz. This implies that the series circuit will provide a better signal-to-noise ratio over most of the audio band.

To illustrate this, the r.m.s. noise voltage per unit bandwidth at the outputs of the two circuits is plotted against frequency to obtain the spectral density* functions shown in Fig. 2(a) using typical values of R_{in} as 50kΩ and transducer inductance of 600mH. The graphs indicate that for the series circuit the spectral density falls at 6dB/octave below the turnover frequency of 13kHz due to the inductive source impedance, whereas for the shunt feedback circuit the spectral density is nearly constant up to 13kHz and falls at 6dB/octave beyond this frequency. This latter result is as expected for at low frequencies when $|Z(jf)| < R_{in}$, the noise current in R_{in} entering the virtual earth is nearly constant with frequency; in other words signal-to-noise ratio is independent of source impedance when $|Z(jf)| \ll R_{in}$.

Now pass the outputs of these two amplifiers through an R.I.A.A. equalization network with unity gain at 1kHz and with the frequency response shown in Fig. 2(b). The resulting spectral density functions are shown in Fig. 2(c). Notice that apart from a higher average noise level than that of the series circuit, the shunt feedback configuration generates a large portion of the total noise in the band below 1kHz where the subjective effect is the more disturbing. To calculate the overall signal-to-noise ratio we must find the total noise power by integrating the square of the functions plotted in Fig. 2(c), see Appendix 2. The signal-to-noise ratio, with full R.I.A.A. equalization in the band 100Hz to 20kHz, obtained for the series and shunt feedback circuits respectively are 72dB and 58.5dB referred to 2mV at 1kHz, a typical signal from a tape head or low-output magnetic pickup cartridge. Practical measurements on these two circuits give results in good agreement with the theoretical (Appendix 2).

With a low-impedance resistive source such as the output of a negative feedback amplifier, it is usual to specify a high input resistance so that moderately-sized coupling capacitors can be used and so that several inputs can be fed simultaneously. Equations 1 and 2 indicate that in these circumstances the series input will provide the better signal-to-noise ratio, though usually signal levels are high and noise is not a problem.

As an example the reader may like to compare the performance of the two line amplifiers of my stereo mixer[4] design; the series input of Fig. 16 gives a residual noise level (i.e. with $R_s = 0$) nearly 15dB lower than that of Fig. 15 which is a shunt feedback arrangement. When the source resistance equals the input resistance (e.g. 600-Ω line matching), it follows from the equations that both configurations will give the same signal-to-noise ratio.*

Finally an interesting property of the shunt feedback circuit is that, contrary to common experience, it will generate minimum noise when the input is open circuit as no noise current flows in the input arm and only the feedback and biasing resistors remain. In practice this effect may be masked or even reversed by deterioration of transistor noise figure when operating with such a high source impedance. This is discussed in the next section.

To summarize: the series feedback configuration gives the better signal-to-noise ratio when the source approximates to voltage drive, while the shunt feedback circuit is superior for current drive conditions ($Z_s > R_{in}$). The designer must also ensure that the feedback resistors, R_e for the series circuit and R_f for the shunt circuit, do not introduce an unnecessary source of noise as implied in the derivation of the equations.

Noise in transistors

The equivalent noise generators of Fig. 3 are a universal representation of any noisy amplifier[3,5]. These generators may be thought of as equivalent noise resistances[3,6] R_{nv} and R_{ni}, which in the case of bipolar transistors are a function of h_{FE} and the collector current. There exists, for any value of these parameters, an optimum source resistance[5], $R_{s\,opt} = \sqrt{R_{nv} \cdot R_{ni}}$ which minimizes the noise contributed by the amplifying device.

For silicon bipolar transistors*[3,7,8]

$$V_n = \sqrt{4kT(r_b + 1/2g_m) \cdot \delta f}$$

$$\text{hence} \quad R_{nv} = (r_b + 1/2g_m) \qquad (3)$$

$$I_n = \sqrt{\frac{4kT\delta f}{2h_{FE}/g_m}}$$

$$\text{hence} \quad R_{ni} = 2h_{FE}/g_m \qquad (4)$$

assuming no correlation between I_n and V_n, and $g_m = qI_e/kT$. In the case of field effect transistors, as with valves, the value of R_{ni} is extremely high and the circuit can often be reduced to an equivalent noise resistance (typically several kilohms at low audio frequencies) in series with the input[6].

It can be shown[6] that noise figure

$$= 10\log_{10}\left(1 + \frac{R_{nv}}{R_s} + \frac{R_s}{R_{ni}}\right)\text{dB} \qquad (5)$$

and it follows that the larger the ratio R_{ni}/R_{nv}, the lower the optimum noise figure becomes and the greater the range of R_s over which a good noise figure (<2dB) can be obtained.

However, the expression for R_{nv}, given in equation 3, includes one term which is inversely proportional to collector current

*If the matching input resistance is established by a combination of shunt and series feedback (see footnote on page 236), no loss of signal-to-noise ratio occurs, except that due to the feedback resistors, as the amplifier is both voltage and current sensing. See for example the microphone amplifier described in B.B.C. Monograph No. 46, Feb. 1963, 'Application of Transistors to Sound Broadcasting'.

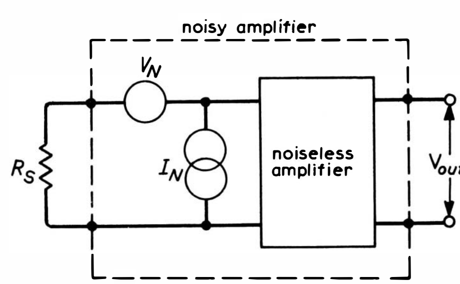


Fig. 3. Equivalent noise generators shown are usually thought of as equivalent noise resistances.

*Spectral density is usually defined as mean-square noise voltage (i.e. power) per unit bandwidth, but as the plot is of r.m.s. noise voltage per unit bandwidth, the vertical scales in Fig. 2 are in $V_{r.m.s.}/\sqrt{f(\text{Hz})}$.

*We assume that $h_{FE} \gg 1$; also neglect leakage current effects or for practical purposes assume a collector-emitter voltage of less than 5V. The 'effective' base resistance, r_b, is obtained by noise measurement, and is not the h.f. value quoted by manufacturers.

(and therefore under the control of the designer) and a fixed term, r_b, the 'effective' base resistance[9], which is a characteristic of a particular transistor and relatively independent of collector current.

So if we let $R_s = r_b \approx 300$ and make $I_C \geqslant 0.5$mA, the absolute minimum noise figure would be 3dB. Obviously there will be a very severe limitation on achievable noise figure when $R_s < r_b$ and so a matching transformer is normally used with a low-impedance microphone[2] (Z_s = 30 to 50Ω or 200 to 600Ω) to step-up the impedance to the 10 to 30kΩ region where transistor noise figure has an optimum value of less than 1dB. The alternative is to use the technique of parallelling n transistors[6,10] and so reducing the series noise resistance by a factor n; a practical example is the medium-impedance microphone amplifier (Fig. 7) of the stereo mixer[2]. It has also been found[8] that p-n-p transistors have lower effective base resistances than n-p-n types. Figs 4(a) and (b) show the noise performance of a suitable p-n-p transistor, Motorola 2N4126*, which for I_C = 0.5mA† gives a 3-dB noise figure at 1kHz when R_s = 100Ω, (hence R_{nv} < 100Ω), and with R_s = 200Ω a noise figure of 2dB at 1kHz, rising to 3dB at 100Hz.

Both bipolar and field-effect transistors suffer from flicker noise‡, which varies from one device to the next and is very hard to predict. With bipolar transistors it may be represented by an increase in I_n (or a decrease in R_{ni}), Fig. 3, below a certain frequency, and may be characterized by the extra noise generator[8] $|I_f|^2 = K.I_B^{\gamma}.f^{-\alpha}.\delta f$, where γ and α are approximately unity and K varies widely with different transistors. Since this generator is proportional to base current, its effect will be reduced by using a low collector current and a device with a high h_{FE} (ref. 12). Fig. 5 demonstrates the relation between collector current and low-frequency noise performance of the Texas TIS97. Flicker noise in j.f.e.ts appears in the voltage generator and can be represented by an increasing value of R_{nv} at low frequencies. Thus for a good noise figure (<4dB) it is preferable to operate bipolar transistors with source resistances less than say 200kΩ, unless selected devices are used, and with low collector currents of say 10μA or less. On the other hand, as the series noise resistance dominates at low frequencies in j.f.e.ts, best noise performance will be achieved with high source resistances of greater than 50kΩ; again this lower limit can be reduced with selected devices.

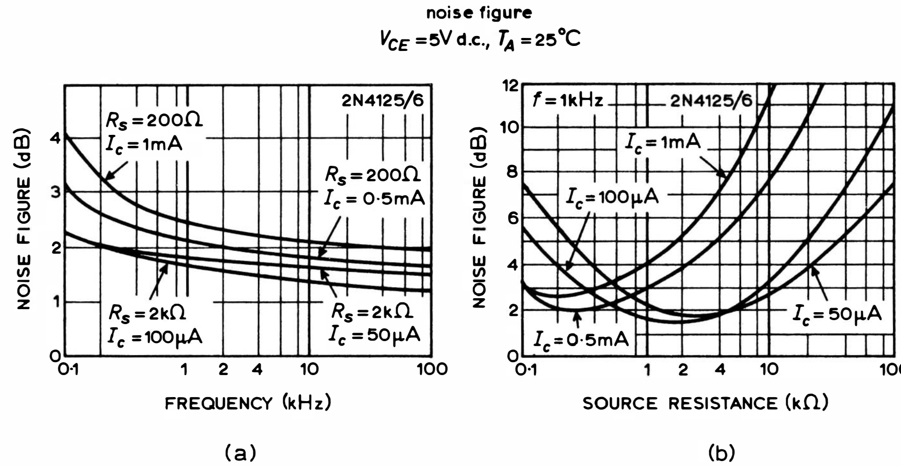


Fig. 4. Noise performance of a p–n–p transistor giving 2-dB noise figure at 1kHz with R_S = 200Ω.

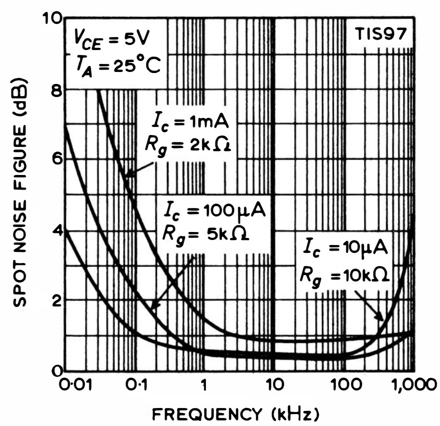


Fig. 5. Effect of flicker noise at low frequencies is reduced by using a low collector current.

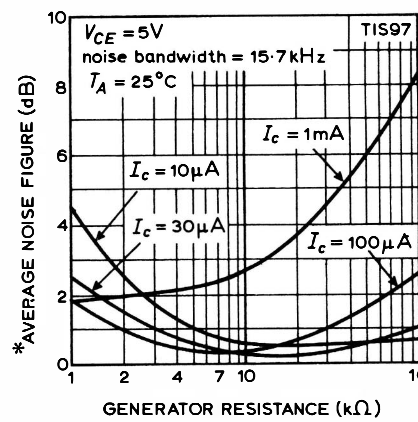


Fig. 6. Broadband noise figure is less than 2dB at I_C of 100μA for a mean source resistance of 5kΩ.

Now two points to consider when optimizing noise figure in a practical circuit. Firstly, equation 5 is true only for a fixed source resistance; practical sources loaded by an input resistance present a complex source impedance, $Z_s(jf) = R_s(f) + jX_s(f)$, of which only the real part $R_s(f)$ is responsible for thermal noise from the source. Analysis of Fig. 3 under these conditions gives the mean-square noise output voltage as

$$\bar{V}_{no}{}^2 = 4kT\delta f\left(R_s(f) + R_{nv} + \frac{|Z_s(jf)|^2}{R_{ni}}\right) \quad (6)$$

at a frequency f for a bandwidth δf, and the spot noise figure is

$$F(f) = 10\log_{10}\left(1 + \frac{R_{nv}}{R_s(f)} + \frac{|Z_s(jf)|^2}{R_s(f)\cdot R_{ni}}\right) \quad (7)$$

The broadband noise output voltage can be found by integration of equation 6. Optimization is now much more tedious particularly when there is excess noise in the noise current generator at low frequencies and when there is an equalization network as is usual in tape and disc pre-amplifiers. A brief discussion of this optimization technique is available from the editorial office.

However, for practical purposes a good approximation can be made by finding the range over which the source impedance varies and then choosing a collector current which gives a good noise figure for the same range of source resistances, either from manufacturers' data sheets or by calculation using equation 5. For our example of the magnetic pickup or tape head loaded by a 50kΩ resistor, the source impedance varies from less than 1kΩ at 100Hz to nearly 50kΩ at 20kHz, having a geometric mean of about 5kΩ.

The noise figure is optimized for this source resistance when the collector current is in the region 70 to 100μA. Fig. 6 shows that the Texas TIS97 gives a broadband noise figure of less than 2dB over the required range of source resistances when the collector current is 100μA. The computed signal-to-noise ratio referred to 2mV at 1kHz of an R.I.A.A.-equalized amplifier (similar to Fig. 3 in the stereo mixer[2]) is shown in Fig. 7, plotted against input stage collector current; the maximum is achieved for a current of 35 to 40μA.

The second point concerns the effect of feedback and also transistor configuration. The fact that applying negative feedback affects input impedance may lead to the erroneous assumption that it will also alter the noise figure. Since signal and noise are both reduced by the same factor[6], the open and closed-loop noise figures are the same provided the noise bandwidth remains constant. Thus noise figure must be optimized under open-loop conditions and negative

*A reader, Mr Curl, of San Francisco, has kindly pointed out the excellent noise performance of the Motorola 2N4401 and 2N4403.

†Operating the transistor at a current of several mA causes R_{nv} to increase as the noise current generator, I_n, begins to develop a significant noise voltage across r_b; the generators will then be slightly correlated. It is probably for this reason, particularly when excess noise is present, that higher current transistors do not give an improved equivalent series noise resistance although the ohmic base resistance is less.

‡Flicker noise (1/f noise) is known to arise in bipolar transistors from generation and recombination at defects (i.e. dislocations and impurities) in the emitter-base junction where there is a high concentration of minority carriers[11,12], while burst ('popcorn') noise results from surface states existing in the passivating silicon dioxide layer on the base region. As such, both effects are a function of the processing used in manufacture, and the improved noise performance of modern transistors results from fewer impurities and from low-defect processing. Excess noise in f.e.ts[7] is due to bulk generation/recombination in the vicinity of the channel and is relatively independent of operating conditions.

feedback used to increase or decrease input impedance*. For example, consider the virtual earth configuration of Fig. 1(b). In a low-noise amplifier, the input arm should have a lower impedance than either the feedback arm or biasing resistors, so the source impedance (open loop) seen by the first transistor is very nearly that of the input arm. As a magnetic-pickup amplifier, this impedance will tend towards R_{in} (50kΩ) at low frequencies, so the first transistor should be run at a low collector current of 10µA or less for a satisfactory flicker noise performance, though perhaps a j.f.e.t. would be a more suitable input device in this configuration.

The fact that the input becomes a virtual earth when the loop is closed does not alter the noise figure, as once the source resistance is defined the transistor noise can be represented by a single current generator across the input which injects the same current regardless of whether the loop is open or closed.

Likewise, despite the differing input impedance of the three transistor configurations, all give approximately the same noise figure with a given source resistance and collector current. For example a low source impedance (<200Ω say) may intuitively be thought to give best noise figure with the very low input impedance of a common-base amplifier. This is not so as the base resistance, which limits the noise figure, is still in series with the signal source and the base-emitter junction.

Design procedure

The important points in low-noise audio design can be summarized under the following headings:

Step 1: configuration. By using equations 1 and 2, determine which feedback arrangement will provide the better signal-to-noise ratio with the particular source impedance and input resistance.

Step 2: feedback components. Check that the feedback loop does not cause a deterioration in the signal-to-noise ratio. Keep series resistors in the input circuit low in value and shunt resistors high.

Step 3: first stage noise figure. The source impedance comprises the total impedance (including that of the source itself) seen by the input transistor between its base and emitter[6]. When this is purely resistive, the optimum first stage collector current is found directly from equations 3, 4 and 5. With a complex source impedance, the approximate method described in the previous section may be used, or a more accurate result obtained by using equation 6 as discussed in Appendix 3. The input transistor should be run with a V_{CE} of less than 5V to prevent excess noise due to leakage currents.

Step 4: minimize noise contribution of later stages. The first-stage gain should be high to minimize the noise contribution of the second stage. If the second stage is run at h_{FE} times the current in the first stage (as with a Darlington connection), the shot noise in the base current of the second transistor will equal the shot noise in the collector current of the first. Also the second stage may generate flicker noise, so the ratio of the collector currents should be much less than h_{FE}. E. A. Faulkner[6] recommends that the second stage be run at the same current as the first, though usually other requirements such as gain and harmonic distortion must be considered.

Step 5: biasing. In the shunt circuit, biasing resistors connected to the input should be at least three times the input resistance to prevent avoidable noise current flowing into the virtual earth. In the series circuit the biasing resistors should be greater than or equal to the specified input resistance; never pad out the input impedance with series resistance. In addition to Johnson noise, resistors carrying a direct current give rise to excess noise (in proportion to their voltage drop) so avoid large voltage drops across biasing resistors, although low-noise resistors and the low voltages in transistor circuits make this a secondary consideration.

*In fact this is a useful advantage of feedback, as it allows impedance levels to be modified without altering noise performance; in an open-loop configuration, shunt or series buffering resistors would be required with a consequent deterioration in signal-to-noise ratio.

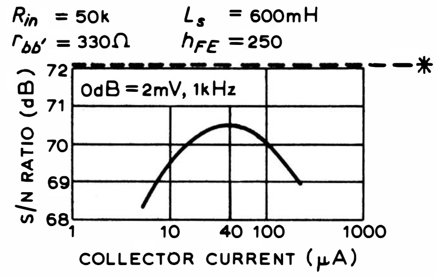


*Fig. 7. Calculated signal-to-noise ratio of an equalized amplifier as plotted against I_C of first stage. *Broken line represents theoretical maximum with noiseless transistor. (Graph neglects excess noise.)*

Example

The amplifier shown in Fig. 8 was primarily intended for a magnetic pickup[2], and as the source impedance is less than the required input resistance over most of the audio band, a series feedback arrangement is chosen (Step 1). The graph of Fig. 7 shows the optimum first-stage collector current to be 40µA (Step 3), whereas the approximate analysis gives 80µA. Note, however, that the difference in signal-to-noise is only 0.2dB. Resistors R_1 in parallel with R_4, the biasing resistor which carries only the very small base current to Tr_1, provides the required input resistance of 50kΩ. Resistor R_7 is kept as low as possible (Step 2) bearing in mind the loading effect of the feedback network on the emitter follower Tr_3. The voltage across Tr_1 is about 2.5V (Step 3), and the high collector load, R_6, allows Tr_2 to be current driven for low distortion. The optimum second-stage collector current for minimum distortion with the particular value of R_6 is about 0.5mA and this gives a satisfactory collector current ratio (Step 4) of 12. The open-loop gain is in the range 5000 to 7000 and harmonic distortion about 1% for $5V_{r.m.s}$ output. www.keith-snook.info

Appendix 1

Signal-to-noise ratio of series-feedback circuit

The signal voltage appearing across R_{in}, Fig. 1(a), at a frequency f is

$$|V_{in}| = \frac{V(f)\cdot R_{in}}{|R_{in}+Z(jf)|}$$

where $Z(jf) = R(f)+jX(f)$

Short out $V(f)$; then noise voltage across R_{in} at a frequency f for a bandwidth δf is

$$\overline{V_n^2} = 4kT\cdot R_{in}\cdot\delta f\cdot\left|\frac{Z(jf)}{R_{in}+Z(jf)}\right|^2$$

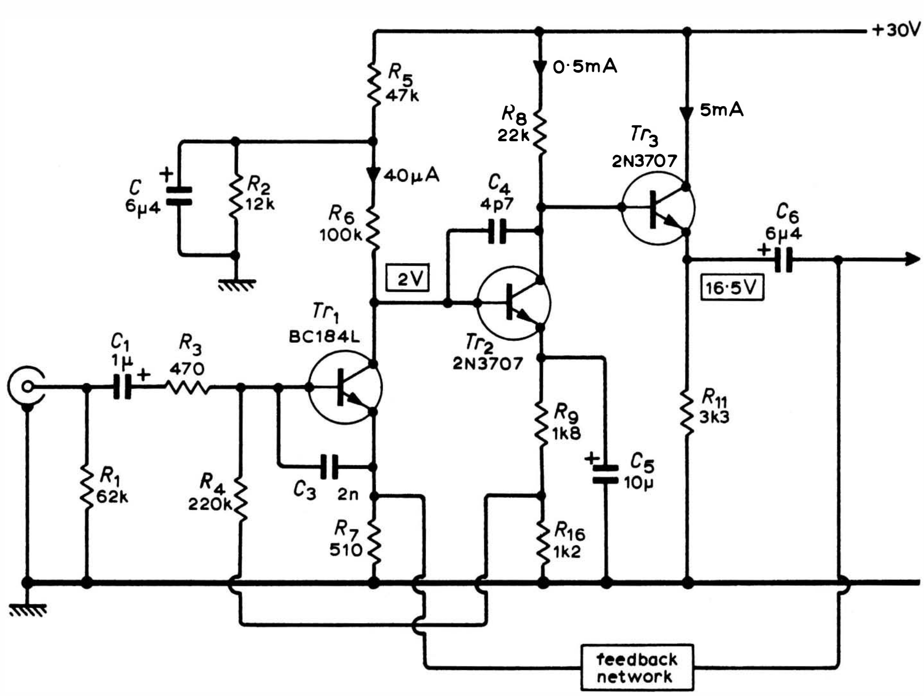


Fig. 8. Circuit of pre-amplifier for a magnetic pickup used to illustrate low-noise design procedure.

due to R_{in}

$$+4kT\cdot R(f)\cdot\delta f\cdot\left|\frac{R_{in}}{R_{in}+Z(jf)}\right|^2$$

due to real part of $Z(jf)$

Therefore r.m.s. noise voltage

$$\overline{V_n}=\frac{R_{in}}{|R_{in}+Z(jf)|}\cdot\sqrt{4kT\cdot\left(\frac{|Z(jf)|^2}{R_{in}}+R(f)\right)\cdot\delta f}\quad\text{(A1)}$$

Hence signal-to-noise ratio is

$$\left|\frac{V_{in}}{\overline{V_n}}\right|=\frac{V(f)}{\sqrt{4kT\cdot\left(\frac{|Z(jf)|^2}{R_{in}}+R(f)\right)\cdot\delta f}}$$

Shunt-feedback circuit

The signal voltage at a frequency f appearing at the output, Fig. 1(b), is obtained by summing the currents at node 'E'.

$$\frac{V(f)}{|Z(jf)+R_{in}|}+\frac{V_o}{R_f}=0$$

$$\text{Therefore } V_o=-\frac{R_f\cdot V(f)}{|R_{in}+Z(jf)|}$$

Short out $V(f)$ and equate the mean square noise currents flowing into node 'E'. Assume R_f is sufficiently large that its noise current can be neglected.

$$\frac{4kT(R_{in}+R(f))\cdot\delta f}{|R_{in}+Z(jf)|^2}=\frac{\overline{V_n}^2}{R_f^2}$$

at a frequency f over a bandwidth δf.

$$\text{Therefore } \overline{V_n}=\frac{R_f}{|R_{in}+Z(jf)|}\cdot\sqrt{4kT(R_{in}+R(f))\cdot\delta f}\quad\text{(A2)}$$

$$\text{Thus s/n ratio is}\quad\frac{V(f)}{\sqrt{4kT(R_{in}+R(f))\cdot\delta f}}$$

Appendix 2

Signal-to-noise ratio of magnetic pickup amplifiers with R.I.A.A. equalization

Assumptions: Pickup cartridge purely inductive; piecewise linear approximation to R.I.A.A. curve, as in Fig. 2(b).

The subject of noise in frequency-dependent networks is treated in several texts[12]. When a noise voltage or current with a certain spectral density is passed through a frequency-dependent network the resulting output has a spectral density equal to the product of the input spectral density function and the square of the magnitude of the transfer function (equation A3).

$$S_o(f)=S_i(f)\cdot|H(jf)|^2\quad\text{(A3)}$$

Series circuit. From equation A1, the noise voltage at a frequency f for a bandwidth f is

$$\overline{V_n}(f)\Big|_{B=\delta f}=\left|\frac{R_{in}}{R_{in}+j2\pi fL}\right|\cdot\sqrt{4kT\cdot\left(\frac{|j2\pi fL|^2}{R_{in}}+0\right)\cdot\delta f}$$

Let $L/R_{in}=1/2\pi f_t$ where L is the inductance of the pickup.

$$\overline{V_n^2}(f)\Big|_{B=\delta f}=\left|\frac{1}{1+jf/f_t}\right|^2\cdot 4kTR_{in}\cdot(f/f_t)^2\cdot\delta f=S_i(f)\cdot\delta f$$

The R.I.A.A. network can be characterized by three regions

$$|A(jf)|^2=(f_2/f)^2 \text{ for } f_1<f<f_2$$
$$\text{,,}\quad =1 \text{ for } f_2<f<f_3$$
$$\text{,,}\quad =(f_3/f)^2 \text{ for } f_3<f<f_4$$

Output spectral density is

$$S_o(f)=S_i(f)\cdot|A(jf)|^2$$

Total mean square voltage over the band f_1 to f_4 is $\int_{f_1}^{f_4}S_o(f)\cdot df$ Thus

$$\overline{V_{no}}^2=4kTR_{in}\cdot\int_{f_1}^{f_4}\frac{(f/f_t)^2}{1+(f/f_t)^2}\cdot|A(jf)|^2\cdot df$$

This integral is evaluated in three parts corresponding to the three regions of the R.I.A.A. characteristic. With L = 600mH. R_{in} = 50kΩ, f_1 = 50Hz, f_2 = 500Hz, f_3 = 2120Hz and f_4 = 20kHz,

$$\overline{V_{no}}^2=4kTR_{in}\cdot(0.65+18.8+290)$$

Thus the r.m.s. noise voltage $\overline{V_n}=0.5\mu V$, which referred to 2mV at 1kHz gives a signal-to-noise ratio of 72dB.

Shunt circuit. Starting from equation A2 and assuming $R_f=R_{in}$, but neglecting the noise current due to the feedback resistor, the mean square noise voltage is

$$\overline{V_{no}}^2=4kTR_{in}\cdot(4000+1600+1600)$$

Hence the r.m.s. noise voltage is 2.5μV, which gives a signal-to-noise ratio of 58.5dB referred to 2mV at 1kHz.

To check these results, noise measurements were made on the series feedback circuit shown in Fig. 8 and on the shunt feedback circuit described by Linsley-Hood (July 1969 issue). The inputs of the amplifiers were loaded with a 600-mH inductor to simulate the cartridge. The signal-to-noise ratios for the series and shunt circuits referred to 2mV input at 1kHz were 70dB and 58dB respectively. These results show good agreement with the theory. Measurement bandwidth was less than that used in the calculation and transistor noise is also making a contribution.

References

1. Burrows, B. J. C., Ceramic Pickup Equalization, *Wireless World*, July and August 1971, pp.321–4 & 379–82.
2. Walker, H. P., Stereo Mixer—1, *Wireless World*, May 1971, pp.221–5.
3. Baxandall, P. J., Noise in Transistor Circuits, *Wireless World*, vol. 74 1968, pp.388–92 & 454–9.
4. Walker, H. P. Stereo Mixer—2, *Wireless World*, June 1971, pp.295–300.
5. Haus, H. A. *et al*, Representation of noise in Linear Twoports, *Proc.I.R.E.*, vol. 48 1960, pp.69–74. www.keith-snook.info
6. Faulkner, E. A., Design of Low-noise Audio-frequency Amplifiers, *Radio and Electronic Engineer*, vol. 36, 1968, pp.17–30.
7. Mantena, N. R., Sources of Noise in Transistors, *Hewlett-Packard Journal*, vol. 21, 1969, No. 2. Swiontek, M. C. and Hassun, R., Influence of Transistor Parameters on Transistor Noise Performance—a simplified Presentation, *Hewlett-Packard Journal*, vol. 16, 1965. No. 7.
8. Chenette, E. R., Low-noise Transistor Amplifiers, *Solid-State Design*, February 1964, pp.27–30.
9. Faulkner, E. A. and Harding, D. W., Some Measurements on Low-noise Transistors for Audio-frequency Applications, *Radio and Electronic Engineer*, vol. 36, 1968, pp.31–3.
10. Faulkner, E. A., Optimum Design of Low-noise Amplifiers, *Electronics Letters* vol. 2, 1966, pp.426–7.
11. Jaeger, R. C. and Brodersen, A. J., Low-frequency Noise Sources in Bipolar Junction Transistors, *I.E.E.E.Trans.* vol. ED-17, 1970. pp.128–36.
Plumb, J. L. and Chenette, E. R., Flicker Noise in Transistors, *I.E.E.E.Trans.* vol. ED-10, 1963, pp. 304–8.
Conti, M., Surface and Bulk Effects in Low-Frequency Noise in N-P-N Planar Transistors, *Solid-State Electronics* vol. 13, 1970, pp. 1461–9.www.keith-snook.info
12. Hsu, S. T., Noise in High-gain Transistors and its Application to the Measurement of Certain Transistor Parameters, *I.E.E.E. Trans.* vol. ED-18, 1971, no. 7.

Meetings

LONDON

1st. IEE — Colloquium on "Digital frequency synthesis in communication systems" at 10.00 at Savoy Pl., W.C.2.

2nd. IERE/IEE — Colloquium on "Hybrid computing systems" at 14.30 at 9 Bedford Sq., W.C.1.

3rd. SERT — "The registration of technician engineers and technicians" by A. J. Kenward at 18.30 at Mullard House, Torrington Pl., W.C.1.

5th. IEE/I. Meas. Control — Colloquium on "Industrial applications of queueing theory" at 10.00 at Savoy Pl., W.C.2.

8th. IEE — Discussion on "Experiences with amorphous semiconductor devices" at 17.30 at Savoy Pl., W.C.2.

9th. AES — "Transformers and the audio engineer" by P. J. Baxandall at 19.15 at the Mechanical Engineering Dept., Imperial College, Exhibition Rd., S.W.7.

10th. I.Phys./IEE — Colloquium on "Semiconductor memories" at 10.00 at Savoy Pl., W.C.2.

10th. IEE — "The applications of electroluminescence and acoustic surface waves" by Dr. E. V. D. Glazier at 17.30 at Savoy Pl., W.C.2.

10th. SERT — "Digital processing of radio and television signals" by Dr. B. Moffatt at 19.00 at the I.T.A., 70 Brompton Rd, S.W.3.

11th. IERE/IEE — Colloquium on "Implantable cardiac pacemakers" at 14.30 at 9 Bedford Sq., W.C.1.

11th. IEE — "The future role of the technical journal" by M. G. Lowe at 17.30 at Savoy Pl., W.C.2.

12th. IEE — "Integrated navigation systems for aircraft and hovercraft" by D. C. Price at 17.30 at Savoy Pl., W.C.2.

18th. RTS — "Graphics in B.B.C. television" by Colin Cheesman at 19.00 at the ITA Conference Suite, 70 Brompton Rd., S.W.3.

GLASGOW

31st. SERT — "The registration of technician engineers and technicians" by A. J. Kenward at 19.30 at McClellan Galleries, Sauchiehall St.

IPSWICH

10th. SERT — "Video tape recorders" by C. H. W. Allvey at 19.30 at Ipswich Civic College, Rope Walk.