Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Feedback amplifiers

Following the interesting and informed article by Mr Walker on low noise amplifiers (*Wireless World*, May 1972) there has been a protracted and inconclusive series of letters discussing the various merits of shunt and series feedback connections with regard to noise and distortion.

I would almost certainly have been happy to let this die out in its own way had not the discussion gone completely off the rails in John Linsley Hood's letter "Feedback Amplifiers" in the May 1973 issue.

Mr Linsley Hood suggests that the difference between the series and shunt feedback connections in the circuit given arises because in the series feedback case the signal is not normally attenuated much between source and amplifier, whereas in the shunt feedback case it will be attenuated 4–6dB depending on suitable operating parameters.

The effect of a finite input impedance in a feedback amplifier can be considered as a reduction in loop gain, and for the two connections, see Fig. 1, the effect of input impedance are as below.

Series feedback:

$$\frac{E_0}{E_1} = A(s)\left[\frac{R_e+R_{fb}}{(R_e+R_{fb})\left(1+\dfrac{R_e+R_s}{R_{in}}\right)+A(s)R_e}\right]$$

When $R_e = R_s$ and $R_{in} \rightarrow \infty$, $A(s) \gg 1$

$$= \frac{R_e+R_{fb}}{R_e}$$

Shunt feedback:

$$\frac{E_0}{E_1} = -A(s)\frac{R_{fb}}{(A(s)+1)R_s+R_{fb}(1+R_s/R_{in})}$$

In the limit $R_{in} \rightarrow \infty$, $A(s) \gg 1$

$$= -\frac{R_{fb}}{R_s}$$

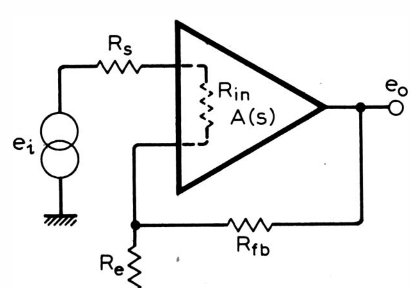
It can be seen that the sensitivity of the two circuits to finite input impedance is similar, with suitable values, e.g. a loop gain of 500 $R_{in} = 15\text{k}\Omega$, $R_s = 50\text{k}\Omega$, $R_{fb} = 500\text{k}\Omega$, the reduction in gain in each case by considering R_{in} finite is: series 1.3dB, shunt 0.8dB.

It is not correct to assert that the intrinsic problem with a shunt feedback amplifier is that its input impedance attenuates the signal by 4–6dB. It is readily seen from the equation that the input impedance becomes insignificant anyway when $R_{in} \rightarrow R_s$ and completely insignificant for $A(s) \gg 1$. It is therefore simply a problem of good design to assure that R_{in} is a suitable value, not a drawback of a feedback connection.

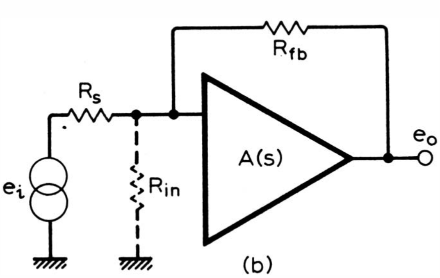
Mr Linsley Hood goes on to say that (in the shunt feedback case) the noise impedance seen by the input is not the input resistor circuit value but the value of the "virtual earth impedance", and suggests that this impedance is 600–1200 ohms. This comment is quite amazing. What is a virtual earth impedance? One can only assume that it is a phantom idea to describe how "earthy" the virtual earth point is.

It is quite misguided to use this idea. The virtual earth is a phenomenon resulting from the feedback connection but it does not have an impedance as such that can generate noise.

A shunt feedback amplifier is a current amplifier, and the low noise condition is with the input open circuited, i.e. in Fig. 1(b) the generator E_1 is open-circuited. The noise of the amplifier here is determined by the thermal noise current generator in R_{fb} and the noise factor of the amplifier with a source resistance of R_{fb}.



(a)



(b)

In its mode of use (E_1 short circuited) the source resistance is $R_s//R_{fb}$ and the noise current of R_s is significant. Certainly a 47kΩ source resistor will generate a noise voltage of 3.9μV and provided the input is short circuited this will be shown in the amplifier noise performance.

In the case of the pickup amplifier R_s = 47kΩ, $Z_{fb}(s) \gg R_s$ it can be shown by calculations that the maximum s/n ratio with a cartridge connected is 58dB ref. 2mV.

Experiment and theory clearly show a marked increase in the noise of such an amplifier when the input is short circuited. Perhaps Mr Linsley Hood could explain how connecting a 47kΩ resistor in parallel with a 1000Ω virtual earth impedance can give a 10dB rise in noise?

Finally, on the subject of distortion, good circuit design can easily permit a series feedback amplifier to have a repeatable performance of s/n better than 70dB s/c ref. 2mV and distortion less than 0.01% in the audio range. The fact that this cannot be achieved with a 741 should be considered irrelevant by any engineer concerned with these and any other important design parameters not covered in the arguments to date.

J. R. Stuart,
Lecson Audio Ltd.,
St Ives,
Huntingdon.

May I offer the following points regarding recent correspondence on distortion and noise?

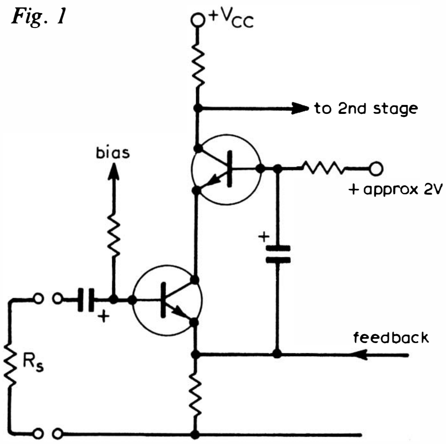
(1) "Common-mode distortion". In many cases of practical interest, it is the variation of C_{cb} of the first transistor with V_{cb} (Early effect) which dominates. Considering a BC214 input stage run at $V_{ce} = 5$ and handling an input of 1 volt r.m.s., the Texas data sheet indicates a capacitance swing of 3pF. This corresponds to a second harmonic distortion of 0.1% at 20kHz if the source impedance is 10kΩ. A considerable reduction in Early effect distortion, and almost complete elimination of the other distortions which are not amenable to reduction by feedback, may be obtained by using a bootstrapped cascode arrangement (Fig. 1).

There is now an Early effect from the upper transistor's C_{cb}, but it is much less than before since it injects distortion into the output, not the input. In fact if the quiescent current through the transistors is chosen for optimum noise figure from R_s, then the Early effect will be reduced by a factor $\sqrt{\beta}$. This circuit permits the lower transistor to be run at a very low V_{ce}, for optimum noise performance, without compromising the ability to handle large signals.

(2) Reduction of distortion by feedback. The statement by Mr Hood and quoted by Messrs Mornington-West and Vereker (May issue), that quadrature components of the feedback are ineffective in reducing the distortion, is absolutely without foundation, as is shown in the appendix to this letter.

To understand the poor high frequency performance of Mr Linsley Hood's *Hi-Fi News* design it is sufficient to consider how much feedback is applied round the output

Fig. 1

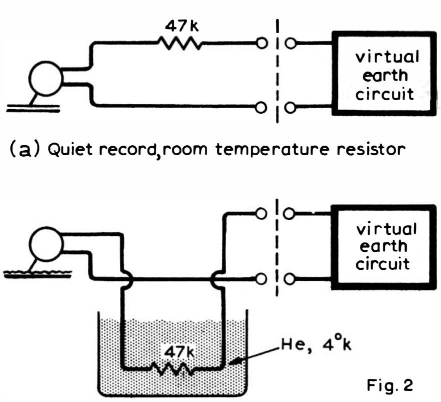


stages. Apart from the usual local feedback, it amounts to 5½dB at 20kHz.

It is not nowadays safe to assume that the effect of a "h.f. stabilising capacitor" is confined to high frequencies. In the design mentioned above, the dominant lag capacitor looks harmless enough at 220pF, yet it gives an open-loop break point of 10Hz.

(3) The measurements reported in Mr Linsley Hood's second letter in the May issue point to interesting possibilities in noise reduction.

Consider a notional dividing line between the 47kΩ resistor and the rest of Mr Linsley Hood's virtual earth circuit (Fig. 2 (a)). The combination on the left will, as he says, produce an open-circuit noise of 3.9μV.



(a) Quiet record, room temperature resistor

(b) Noisy record, cooled resistor

Fig. 2

Now let us take a gramophone record on which, by some mischance, tape hiss has been recorded, and let us choose a pickup of the right sensitivity so that this hiss appears as exactly 3.87μV on its output terminals. Pickup and resistor will then produce $\sqrt{3.9^2+3.87^2} = 5.5\mu$V of noise, but if we now immerse the resistor in a dewar of liquid helium (Fig. 2 (b)) the open circuit noise will once again be 3.9μV.

The impedance presented to the circuit on the right is of course exactly the same for Fig. 2 (b) as for 2 (a), so it should need only a little fiddling of the frequency spectra to convince the circuit that it is connected as in Fig. 2 (a), when the truth is 2 (b). If now the circuit can achieve the noise value of 0.6μV claimed by Mr Linsley Hood, then a noise reduction of $20\log_{10}\dfrac{3.87}{0.6} = 16.2$dB will have been obtained.

Perhaps some enterprising record company will consider this technique for revitalising its pre-Dolby LPs?
Peter G. Craven,
Oxford.

Appendix

Let the amplifier have perfect common mode rejection so that V_{out} is a function of V_1 only. Suppose that we are trying to reproduce a sine wave of unit amplitude and that it is possible to predistort V_1 so that V_{out} is a pure sinusoid. Let "X" be the assumption that the gain of the system to a small signal superimposed on the input is not greatly affected by the presence of the large signal. X will be false if the amplifier is near to clipping.

Let the gain (V_{out}/V_1) of the amplifier at the ith harmonic of the sinusoid be A_i and let β_i be the corresponding feedback factor (V_f/V_{out}). To take account of phase shifts, A_i and β_i will be complex. Suppose that we have succeeded in making V_{out} a pure sinewave and that d_i (also complex) is the amplitude of the ith harmonic of the predistorted signal V_1 necessary to achieve this. Since the feedback network is assumed linear, V_f will be a pure sine wave, and since $V_{in} = V_1 + V_f$, it follows that V_{in} must also have an ith harmonic of amplitude d_i.

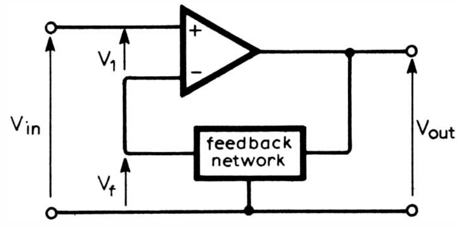
We wish to consider a pure V_{in}, and this we get from the predistorted V_{in} by adding $-d_i$ of the ith harmonic for $i = 1 \ldots \infty$. By assumption X there will appear harmonics at the output, the amplitude of the ith being $-G_i d_i$, where G_i is the gain of the system (including feedback if any) at the ith frequency.

Since d_i does not depend on the feedback, we have proved the well known fact that feedback reduces each distortion product in the same ratio as it reduces the gain at the frequency of the distortion product.

By elementary feedback theory, G_i is given by

$$G_i = \frac{A_i}{1+A_i\beta_i},$$

and comparing with the case $\beta = 0$, it is clear that introducing the feedback has reduced the gain, and hence the distortion



by a factor $|1+A_i\beta_i|$. This factor we now evaluate for an amplifier with a loop gain of 10 ($|A_i\beta_i| = 10$).

No phase shift $\rightarrow A_i\beta_i$ is real and positive
$\rightarrow |1+A_i\beta_i| = 11$
90° phase shift $\rightarrow A_i\beta_i$ is pure imaginary
$\rightarrow |1+A_i\beta_i| = 10.05$
180° phase shift $\rightarrow A_i\beta_i$ is real and negative
$\rightarrow |1+A_i\beta_i| = 9$

Mr Linsley Hood replies:
I am sorry that Mr Stuart feels that the debate on feedback amplifiers "has gone completely off the rails", but he has taken my letter somewhat out of its intended context.

To refer specifically to the main point of this—measurements suggest that the s/n ratio of an amplifying circuit with shunt f.b. is a few dB worse than in the case of the series f.b. circuit with the same value of input resistance.

I believe that this phenomenon is real, and that it is due to the fact that any real amplifying device will require some input energy—significant in a bipolar transistor—and that in the shunt circuit this is obtained from the input signal.

In the latter part of my letter I suggested an alternative method of considering the noise impedance seen at the input—which is a voltage node—in a shunt f.b. amplifier. If one considers the amplifying element, having a known open loop gain, as being detached from the feedback loop but amplifying the noise voltage seen at that point, the noise impedance of the "virtual earth" can be derived, if one is interested to do this.

This observation was not specifically related to the s/n ratio of a shunt feedback circuit, which is best approached by considering it as a current amplifier. In this case the input noise currents decrease as the root of the admittance ($1/Z$) of the input limb, whereas the signal current decreases linearly. Other things being equal the lower the input limb impedance, the better.

In the particular case of a pickup amplifier circuit with R.I.A.A. equalisation, it should be remembered that the effective noise bandwidth is only about 500Hz. Since this allows a s/n ratio with a 47kΩ input resistor and a shunt f.b. circuit to be −72dB ref. 5mV (−64dB ref. 2mV) I suspect that the "calculations" to which Mr Stuart refers assume a wider bandwidth than this. The relative advantage of the series circuit diminishes with frequency when used with an inductive element such as a magnetic p.u. cartridge, from about 11dB at 1kHz to some 3dB at 5kHz. (Assuming a 600mH cartridge inductance, and a series f.b. input d.c. resistance of 2kΩ).

In reply to Mr Craven, on the more important point of the extent of distortion reduction by feedback at phase angles other than 180°, the problem is that the predicted distortion reduction from the formula

$$\frac{D_F}{D} = \frac{A_F}{A}$$

gives unsound results under these conditions, whether the gain is calculated by the method Mr Craven shows or whether it is derived by the classical formula below.

$$A_F = \frac{A}{\sqrt{1+|\beta A|^2-2|\beta A|\cos\Phi}}$$

where Φ is the f.b. phase angle.

As an example, a non-linear amplifier element, having a gain of 100×, an input impedance of 4.7kΩ and a t.h.d. of approximately 4% at 1kHz and 1.5 volts r.m.s. output, was set up as shown in the figure, with an output lag circuit whose values were chosen to give a phase lag of 90° at 1kHz.