Distortion in low-noise amplifiers

1 — Distortion analysis

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The principles of low-noise circuit design are now well established and have been the subject of several articles in this journal, refs 1 & 2. In comparison the design of low distortion circuits has received relatively little attention. In this article distortion in feedback amplifiers is considered in detail with special reference to the distortion produced by the common-mode input signal in series feedback amplifiers. Distortion resulting from the exponential dependence of the collector current of a transistor on base-mitter voltage is also considered in detail, both theoretically and experimentally, and the analysis can be used to predict the effect of this non-linearity on the distortion performance of an amplifier.

In the second part of the article a preamplifier design will be described which embodies the design guidelines developed. Harmonic distortion, measured with magnetic pickup equalization, is less than 0.005% at all frequencies up to 20kHz and all overload levels up to 30 dB.

The inequality derived in the panel on page 31 expresses mathematically the requirement that a series feedback amplifier should have good common mode performance to minimize distortion. Unfortunately, design for good common mode rejection conflicts with the low-noise design requirements of operating the input transistors at low collector-emitter voltages.

Non-linearity due to common mode input

Operation of a transistor with a low collector-emitter voltage minimizes the noise due to leakage currents[1] but the transistor is obviously more sensitive to changes in the collector-base voltage (which occur as a direct result of a common mode input signal) than if the transistor were operated at a higher collector-base voltage. Changes in the collector-base voltage of a transistor manifests itself as a variation in the input base current and a common mode input voltage to a transistor amplifier therefore results in a common mode input current. The common mode input voltage and input current are related by common mode input admittance and it is the non-linearity of this which is primarily responsible for the distortion which arises from a common mode input signal.

The common mode input current would not be important if the source impedances seen by the inverting and non-inverting input of the amplifier were low or equal. However, in a series feedback amplifier designed for example for use with a magnetic impedance seen at the non-inverting input is predominantly inductive whereas the impedance presented by the feedback network to the inverting input is normally kept low so that the equivalent noise voltage generator of the feedback network is small. At the higher audio frequencies therefore there is a serious mismatch in source impedances. Under these conditions the common mode input current can produce a significant differential mode input which is indistinguishable from the input signal. A common mode input voltage is also capable of producing a differential mode input current but with a serious mismatch of source impedances the effect due to the common mode input current will be dominant.

The variation of base current of a transistor with collector-base voltage has been investigated with the circuit shown in Fig. 1 in which, for convenience the collector base voltage is modulated by a transformer in series with the collector d.c. supply. Figure 2 shows the waveform observed at the base of the transistor due to a 20 kHz, 1.0 V r.m.s. sine wave modulation of the collector-base voltage, a modulation level which might well be achieved in a series feedback amplifier when driven by a magnetic pickup at high overload. The waveform was obtained with a quiescent collector-base voltage of 2.0V and a G800E magnetic cartridge used for Z_b to simulate the source conditions of a practical amplifier. Notice that the base voltage waveform contains a high proportion of distortion products and harmonic analysis shows that the total harmonic distortion (t. h. d.) referred to the 1.0V r.m.s. sine wave is 0.17%. If used as an input stage of a series feedback amplifier these distortion products would be indistinguishable from the input signal and no amount of feedback would reduce the t.h.d. of the amplifier to less than 0.17%.

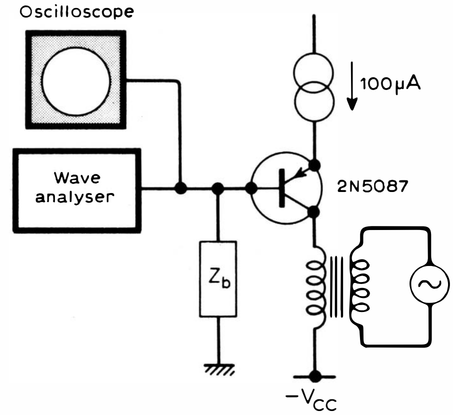
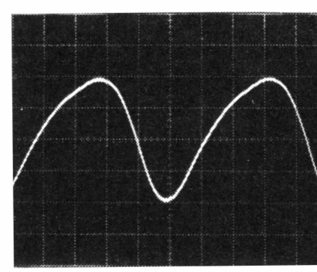


Fig. 1. Arrangement to investigate variation of base current of a transistor with collector-base voltage.



Vertical scale 5mV/div
Horizontal scale 10μs/div

Fig. 2. Voltage developed at the transistor base with a G800E magnetic pick-up cartridge used for Z_b. (Collector modulation 20kHz, 1.0V r.m.s. sinewave, V_{CB} 2.0V.)

The mechanism primarily responsible for the variation of the base current of a transistor with collector-base voltage is base-width modulation, otherwise known as the Early effect. Base-width modulation occurs because of changes in the width of the depletion layer of the collector-base junction as the collector-base potential is varied. Thus an increase in reverse bias causes the depletion layer to extend further into the base region of the transistor which reduces the effective base width and results in an increase in β because of increased base transport efficiency. The increase in width of the depletion layer is also accompanied by a decrease in the collector-base junction capacitance which varies according to

$$C \propto V^{-x}$$

where V is the reverse bias on the junction and x normally has a value between ½ and ⅓ according to the impurity profile across the junction.

The relative contributions of these two effects to the base current modulation have been investigated with the circuit shown in Fig 1 and the results are presented in Fig. 3 in which the fundamental and distortion products of the base current are plotted as a function of frequency for various values of I_c, and constant V_{cE} of 5.0V. At low frequencies base current modulation is independent of frequency but varies with collector current and it is reasonable to attribute this behaviour to variations in β of the transistor. At higher frequencies however base current modulation is independent of one collector current and approximately proportional to frequency which indicates that the collector-base capacitance is the dominant mechanism.

The break point in the characteristics at which the effects of the collector base capacitance starts to dominate over the effect of variations in β shifts to higher frequencies as the collector current is increased as would be expected if the mechanism described above are responsible for base current modulation. At the collector current levels normally encountered in the first stages of low noise audio amplifiers (10 to 100μA) and for frequencies greater than 500 Hz, the variation of the collector-base capacitance is primarily responsible for the distortion products present in the modulated base current.

Base current modulation has been plotted in Fig. 4 as a function of the quiescent collector-emitter voltage modulated by a 10 kHz sinewave. At this frequency and a collector current of 100 μA the collector base capacitance is the dominant base current modulation mechanism. Qualitatively the results agree with the prediction that base current modulation decreases with increasing V_{cE} and although a power-law dependence is indicated it has not been possible to obtain quantitative agreement with the distortion that would be expected from the non-linearity of the collector-base junction capacitance.

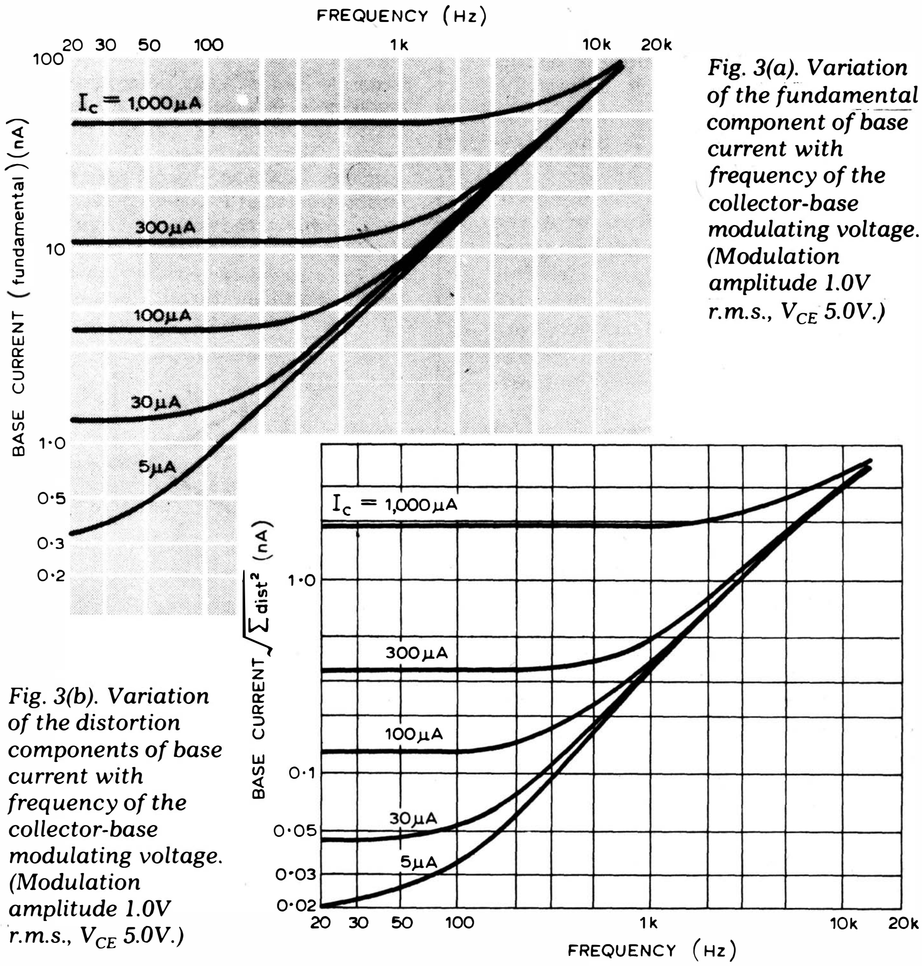


Fig. 3(a). Variation of the fundamental component of base current with frequency of the collector-base modulating voltage. (Modulation amplitude 1.0V r.m.s., V_{CE} 5.0V.)

Fig. 3(b). Variation of the distortion components of base current with frequency of the collector-base modulating voltage. (Modulation amplitude 1.0V r.m.s., V_{CE} 5.0V.)

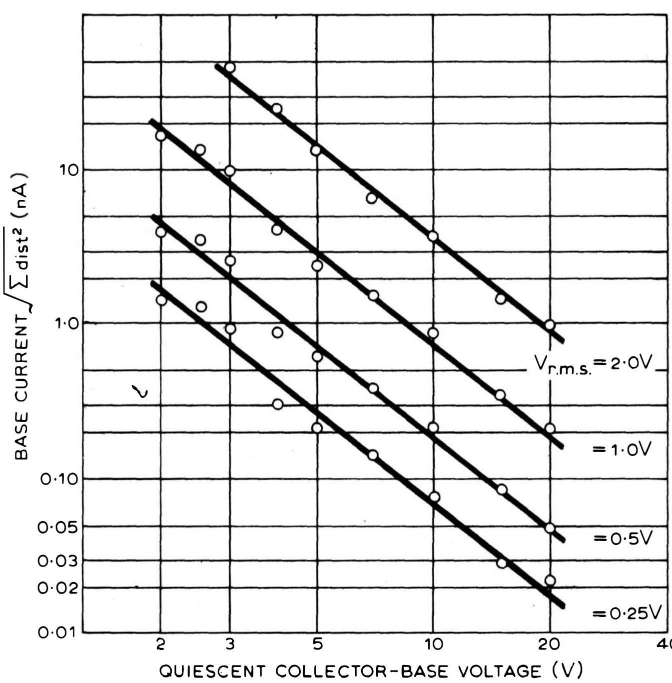


Fig. 4. Variation of the distortion components of base current with quiescent collector-base voltage. (Collector-base voltage modulation frequency 10kHz.)

Reduction of the common mode input signal

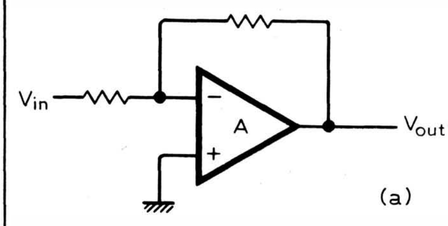
The common mode input signal present in a series feedback amplifier can produce distortion by generating harmonic components at the input which are indistinguishable from the input signal. Differential negative feedback can do nothing to reduce this type of distortion but common mode feedback can give an improvement. As the name implies common mode feedback uses the common mode output signal to reduce the common mode signal at the amplifier input. The application and advantages of common mode feedback, which is fully treated elsewhere,[4] will not be pursued in this article as a very simple technique for reducing the common mode signal which is more relevant to audio applications is to use the feedback connection shown in Fig. 5. In this connection the input signal is introduced in the feedback path of the amplifier so that the differential negative feedback subtraction process is performed external to the amplifier and the common mode signal at the amplifier input becomes identical with the common mode signal which occurs in the shunt feedback configuration. This circuit therefore has the overload capability of the shunt feedback connection but retains the noise performance of the series feedback connection.

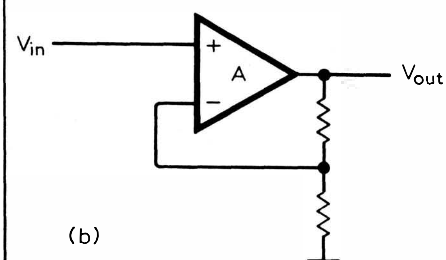
This type of connection does of course require that the signal source is floating. Fortunately this is normally the case in audio applications as the use of series feedback can only be justified in pre-amplifier stages for use with low-level signal sources, e.g. magnetic pickup or tape head.

The pre-amplifier design which is presented in the second part of this article utilises series feedback and the input can be connected conventionally as shown in Fig. 6 or in the feedback path as shown in Fig. 5. With the amplifier equalized for a magnetic cartridge, the last-mentioned connection gives a reduction in t.h.d. by a factor of 40 at high frequencies and high overload levels.

Use of feedback

Negative feedback can be applied to an amplifier by feeding back to the input an antiphase current or voltage which is derived from the output. The inverting amplifier shown in Fig. (a) uses current feedback in what is generally referred to as a shunt feedback configuration, whereas the non-inverting amplifier in Fig. (b) uses voltage feedback in a series feedback configuration.

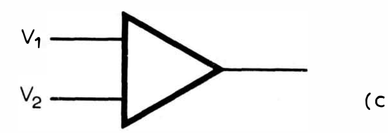


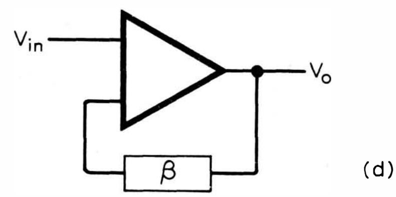


The relative merits of shunt and series feedback in low-noise pre-amplifiers has been the subject of many letters to this Journal.[3] Walker has shown conclusively[1] that with the source impedances associated with a magnetic cartridge, the thermally limited signal-to-noise ratio of the series feedback connection is 13.5dB better than that of the shunt feedback connection. It is generally agreed, however, that the shunt feedback connection has a better overload capability, i.e. lower distortion at high signal levels.

The inferior overload capability of the series feedback connection is a result of the large common mode signal which appears at the amplifier input terminals with voltage feedback but which is not present in the shunt feedback connection. To understand the effect of this common mode signal on the amplifier performance it is necessary to characterise the amplifier by a differential gain A_d and a common mode gain A_c. Thus for the basic amplifier shown in Fig. (c) the output voltage is

$$V_0 = A_d(V_1 - V_2) + A_c(V_1 + V_2)$$





If series negative feedback is now applied to the amplifier as shown in Fig. (d) this equation becomes

$$V_0 = A_d(V_{in} - \beta V_0) + A_c(V_{in} + \beta V_0)$$

$$\therefore A_f = \frac{V_o}{V_{in}} = \frac{A_d + A_c}{1 + \beta(A_d - A_c)}$$

Non-linearity of the differential mode gain

A voltage-driven transistor is an inherently non-linear device because of the exponential relation between collector current and base-emitter voltage. A more linear mode of operation results if the transistor is current driven, but as most audio signal sources approximate to voltage sources the distortion arising from the exponential relation of the input transistor of an amplifier can be significant. Large signal levels can also produce distortion because of the dependence of many transistor parameters on collector current and collector-emitter voltage but these problems can, with suitable design, be confined to the output stage of the amplifier.

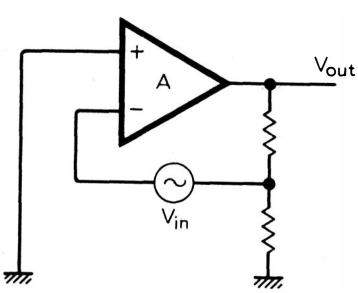


Fig. 5. Series feedback connection with reduced common-mode input signal.

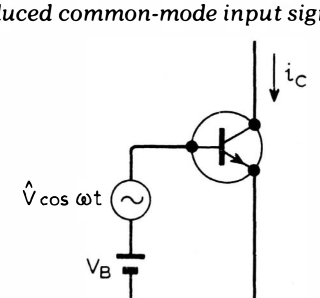


Fig. 6. Equivalent circuit used for distortion analysis of a common-emitter stage – see Fig. 7, curve (g).

Local negative feedback can be used to linearize the output stage of a pre-amplifier but this same technique cannot be used on the input stage without compromising the noise performance. Distortion due to the input stage is therefore a limiting factor in the gain linearity of a low noise pre-amplifier because in theory, if not in practice, the output stage can be made as linear as required simply by increasing the feedback. Information concerning the distortion resulting from the exponential i_c-v_{BE} characteristic of a transistor is therefore necessary to allow the ultimate distortion performance of a pre-amplifier to be predicted.

The distortion of a transistor can be found by expressing the collector current as a function of the input signal and then expanding the expression in a Fourier series which enables the distortion terms to be identified. Thus for the common-emitter stage shown in Fig. 6.

$$i_c = i_S \left| \exp\frac{e}{kT}(V_B + V\cos\omega t) - 1 \right|$$

$$\approx I_c \exp\frac{e}{kT}$$

where i_S is the reverse saturation current of b-e junction, e electron charge, k Boltzmann's constant, T temperature in Kelvins, and I_c quiescent collector current.

This equation now has to be expanded as a Fourier series by writing

$$\exp\frac{e}{kT}(V\cos\omega t)$$

$$= a_0 + a_1 \cos\omega t + \quad \cos 2\omega t \; + \ldots$$

Unfortunately this expression cannot be solved analytically and it is necessary to resort to numerical methods.

The method adopted takes the first ten terms of the Fourier series and gives cos ωt ten equally spaced values between 0 and 1.0 thus enabling a set of ten simultaneous equations with ten where A_f is the closed loop gain. The equation for V_0 can be rearranged in the form

$$V_o = A_d V_{in}\left[\frac{1-2\beta A_c}{1+\beta(A_d-A_c)}\right]$$

$$+A_c V_{in}\left[\frac{1+2\beta A_d}{1+\beta(A_d-A_c)}\right]$$

which allows the differential mode signal V_d and the common mode signal V_c at the amplifier input to be identified in terms of the signal input voltage V_{in}. Thus

$$V_d = \frac{(1-2\beta A_c)V_{in}}{1+\beta(A_d-A_c)} \approx \frac{V_{in}}{1+A_d\beta}$$

$$V_c = \frac{V_{in}}{2}\left[\frac{1+2\beta A_d}{1+\beta(A_d-A_c)}\right] \approx V_{in}$$

The approximations in these two equations make the assumptions $A_d\beta \gg 1$, $A_d \gg A_c$ and $2A_c\beta \ll 1$. Comparison of the two equations shows that in an amplifier with series negative feedback the common mode signal is approximately equal to the input signal and is greater than the differential mode signal by a factor $(1 + A_d\beta)$. In an amplifier with a high differential gain and a large amount of negative feedback the common mode signal can therefore be very much greater than the differential mode signal and the effect of the common mode gain on the amplifier performance may not be insignificant despite an apparently high common-mode rejection ratio.

The effects of non-linearities in the differential and common mode gains on the closed-loop gain can be found by partial differentation of the equation for A_f which gives

$$\frac{\partial A_f}{\partial A_d} \approx \frac{1}{1+A_d\beta}\cdot\frac{A_f}{A_d} \text{ and } \frac{\partial A_f}{\partial A_c} \approx 2\frac{A_f}{A_d}.$$

The approximations make the same assumptions as before. Using the relation

$$\delta A_f = \frac{\partial A_f}{\partial A_d}.\,\delta A_d + \frac{\partial A_f}{\partial A_c}\delta A_c \text{ gives}$$

$$\frac{\delta A_f}{A_f} = \frac{\delta A_d}{A_d}\cdot\frac{1}{1+A_d\beta} + \frac{2A_c}{A_d}\cdot\frac{\delta A_c}{A_c}$$

This equation gives the well-known result that differential negative feedback reduces the effect of changes in differential gain on the closed-loop gain by a factor $(1 + A_d\beta)$. However, differential negative feedback has no effect on the non-linearity of the closed-loop gain due to changes in the common mode gain and the resulting distortion ultimately limits the closed-loop performance of the amplifier. Thus, if the non-linearity of the common mode gain is of the same order as the non-linearity of the differential mode gain, any increase in differential negative feedback is only worthwhile in reducing distortion provided

$$1 + A_d\beta < \frac{A_d}{A_c}$$

In a practical amplifier design the useful limit of negative feedback will probably be reached well before this as some consideration will have been given to obtaining a linear differential gain characteristic.

unknowns to be generated. The solution of these equations is relatively painless with a digital computer and the Fourier coefficients have been evaluated for values of the peak input signal amplitude, $\hat{V}$, incremented in 1.0mV steps up to a maximum of 25mV. The t.h.d. is then readily calculated from the Fourier coefficients and the results of this analysis are presented graphically in Fig. 7(g). Experimental points plotted on the computed curve were determined from measurements made with a Marconi Instruments wave analyser type TF2330A on a 2N5087 transistor operating at a collector current of 100μA. There is excellent agreement between the theory and the experimental results.

Fig. 7(g) clearly confirms that the transistor is an inherently non-linear device; even with input signal amplitudes as low as 1.0mV the t.h.d. is 1% whereas at 10mV the t.h.d. has risen to 10%. The application of this distortion characteristic to the prediction of the distortion performance of an amplifier is perhaps best explained by an example. Consider an amplifier with a common-emitter input stage designed for a maximum output level of 2V peak with an open-loop gain of 2000 and a closed-loop gain (with feedback) of 200. Under these conditions the differential input signal to the amplifier is 1.0mV and the distortion generated in the input stage is, from Fig. 7(g), 1%. The amplifier has a loop gain of 10 and as feedback reduces the distortion by a factor $(1+A\beta)$, the distortion of the amplifier with feedback will be approximately 0.1%.

If better distortion performance is required the simplest design change is to increase the open-loop gain which, in addition to increasing the amount of feedback available to correct the overall non-linearity of the amplifier, reduces the input differential signal with a corresponding reduction in input stage distortion. (The effect of increasing the open-loop gain on the amplifier distortion is analysed in more detail in Appendix I). Ultimately, however, the maximum open-loop gain is limited by stability requirements and the distortion cannot be reduced indefinitely. In any case if very low distortion is the primary specification of an amplifier a better approach is to design for low inherent distortion rather than to try and straighten everything out with negative feedback.[5]

An alternative to the single transistor input stage is the two transistor long-tailed pair input stage. This type of transistor configuration has the advantage of being symmetrical so that

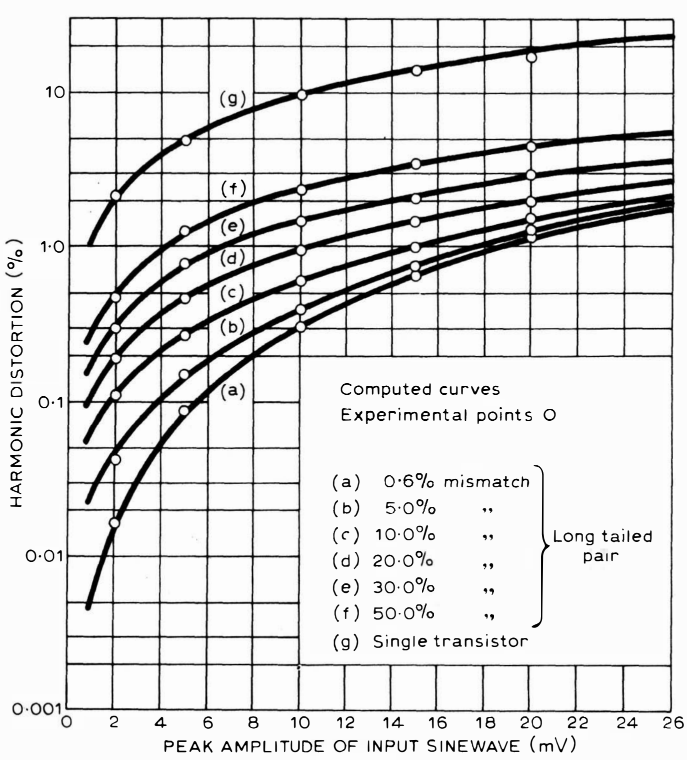


Fig. 7. Distortion curves calculated from coefficients in Fourier expansion of collector current as a function of input signal. Experimental points were measured on 2N5087 transistors with circuits of Figs. 6 and 8.

even-order harmonics are not generated and therefore second harmonic distortion, which is the predominant distortion component in the case of a single transistor, is eliminated.

Analysis of the long-tailed pair stage shown in Fig. 8 is given in Appendix II and the relation between collector current of Tr_1 and input signal has been Fourier analysed using a similar technique to that used for the single transistor stage and the results are presented in curves (a) to (f) of Fig. 7. If the collector currents of Tr_1 and Tr_2 are equal, i.e. $\lambda = 1$, second harmonic distortion is virtually eliminated and for input levels of less than 3mV the distortion is two orders of magnitude lower than that of a single transistor. Thus if a balanced long-tailed pair stage were substituted for the single transistor input stage in the design example previously described the t.h.d. would now be 0.0004%, a very respectable performance considering the small amount of feedback employed.

An interesting point which emerges from the analysis is that distortion is independent of the V_{BE} match between the transistors and this is confirmed by the close agreement between the computed curves and experimental points which were obtained using two transistors deliberately selected from a batch for the largest V_{BE} mismatch, the mismatch being 24mV at I_c of 100μA and V_{CE} of 5.0V. Matching of the collector currents however is essential to obtain the lowest distortion. Examination of the harmonic content of the collector current shows that the increase in distortion as the collector currents are progressively mismatched is due, almost exclusively, to increased second harmonic generation.

The experimental points plotted on the computed curves of Fig. 7 were obtained from measurements performed at 10kHz but further experiments have verified that the results are valid over the whole audio frequency range.

To be concluded

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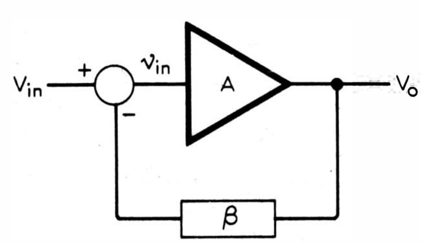
Appendix I — Effect of differential negative feedback on amplifier distortion

Consider an amplifier with a non-linear gain A which can be expressed in terms of the input voltage V_{in} by the Maclaurin series.

$$A = A_0 + v_{in}\frac{dA}{dv_{in}} + \frac{v_{in}^2 d^2A}{2!dv_{in}^2} + \ldots \qquad (1)$$

If this amplifier is now incorporated in the feedback configuration shown in Fig. A1 the



closed-loop gain A_f can similarly be expressed as a Maclaurin series of the form

$$A_f = A_{f0} + v_{in}\frac{dA_f}{dv_{in}} + \frac{v_{in}^2 d^2A_f}{2!dv_{in}^2} + \ldots \qquad (2)$$

$$\text{Now } A_f = \frac{A}{1+A\beta} \quad \therefore \frac{dA_f}{dA} = \frac{1}{(1+A\beta)^2}$$

$$\text{So that } \frac{dA_f}{dv_{in}} = \frac{dA_f}{dA}\cdot\frac{dA}{dv_{in}}$$

$$= \frac{dA}{dv_{in}(1+A\beta)^2} \qquad (3)$$

$$\text{Also } \frac{d^2A_f}{dv_{in}^2} = \frac{d}{dv_{in}}\left(\frac{dA_f}{dv_{in}}\right) = \frac{1}{(1+A\beta)^2}\cdot\frac{d^2A}{dv_{in}^2}$$

$$-\frac{2\beta(dA/dv_{in})^2}{(1+A\beta)^3} \approx \frac{1}{(1+A\beta)^2}\cdot\frac{d^2A}{dv_{in}^2} \qquad (4)$$

Substituting equations 3 and 4 in 2 gives

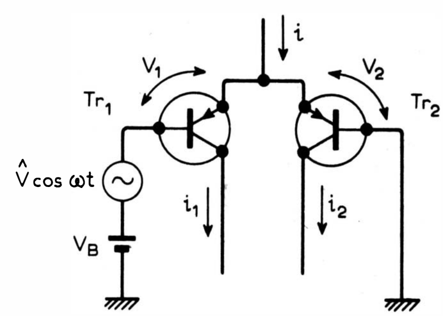
$$A_f = \frac{1}{1+A\beta}\left| A_0 + \frac{v_{in}}{1+A\beta}\cdot\frac{dA}{dv_{in}} + \frac{v_{in}^2}{2(1+A\beta)}\cdot\frac{d^2A}{dv_{in}^2} + \ldots \right|$$

Comparison of this with equation 1 shows that the effect of negative feedback has been to reduce the coefficients of the terms of the power series representing the non-linearity by a factor $(1+A\beta)$ compared with the open-loop configuration.

Increasing the open-loop gain of a feedback amplifier is therefore doubly beneficial in the case of distortion which is dependent on the amplitude of the differential input signal e.g. distortion associated with the exponential $I_c V_{CB}$ characteristic of the input transistor(s); not only is the input differential signal reduced but the amount of feed-back available to correct for non-linearity is increased.

Appendix 2 — Analysis of the long-tailed pair

The collector currents of a long-tailed pair (Fig. A2) are



$$i_1 = i_{S1}\left| \exp\frac{eV_1}{kT} - 1 \right| \approx i_{S1}\exp\frac{eV_1}{kT}$$

$$i_2 = i_{S2}\left| \exp\frac{eV_2}{kT} - 1 \right| \approx i_{S2}\exp\frac{eV_2}{kT}$$

$$\therefore \frac{i_1}{i_2} = \frac{i_{S1}}{i_{S2}}\exp\left| \frac{e}{kT}(V_1 - V_2) \right|$$

$$= \frac{i_{S1}}{i_{S2}}\exp\left| \frac{e}{kT}(\hat{V}\cos\omega t + V_B) \right|$$

When $\hat{V} = 0$, i.e. in the absence of any signal input, let $i_1/i_2 = \lambda$. Then

$$\frac{i_1}{i_2} = \lambda \exp\left| \frac{e\hat{V}\cos\omega t}{kT} \right|$$

But $i_1 + i_2 = i$

$$\therefore i_1 = \frac{i\lambda}{\lambda + \exp - \left| \frac{e\hat{V}\cos\omega t}{kT} \right|}$$

Distortion in low-noise amplifiers

Low-noise, low-distortion preamplifier design with RIAA equalization

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The first part of this article considered the effects of transistor non-linearities on the distortion performance of feedback amplifiers. This concluding part illustrates the practical application of some of the low distortion design principles established, by the design of a low-noise, low-distortion, audio preamplifier equalized for use with a magnetic pickup. With a nominal output of 100mV for 5mV input at 1kHz, it has 30dB overload capability and an harmonic distortion of 0.005% at all frequencies and all overload levels.

The primary function of an audio preamplifier is to raise the input signal above the system noise level whilst meeting certain specifications regarding distortion and overload. Nominal output level should be high enough to prevent the design of subsequent stages being compromised by noise considerations but should not be so high as to severely restrict the overload capability of the amplifier. A nominal output level of 100mV is a reasonable compromise but even so an overload capability of 30dB demands a peak-to-peak output swing of approximately 9V.

In Part 1 of this article attention to the non-linearity of the differential gain of a low-noise amplifier was confined to the non-linearity of the input stage on the ground that the output stage could be made as linear as required by local feedback. Adopting a similar approach and assuming that all distortion is produced by the exponential $I_C V_{BE}$ characteristic of the transistors in the input stage, allows the minimum open-loop gain necessary to meet the distortion specification to be determined as follows.

The peak output amplitude V_o is determined for the specified overload capability; in the present design it is equal to 4.47V for 30dB overload referred to 100 mV. For a given value of open-loop amplifier gain A the differential input voltage to the amplifier is then V_O/A and the harmonic distortion can then be found either from the graph of Fig. 7 (Part 1) or more conveniently from the table given in Appendix 3. Thus if for example the gain A was equal to 1000, the differential input signal for 30dB overload would be 4.47mV and the distortion generated by a single common-emitter stage would be 4.3%.

It is now necessary to determine the feedback factor of the amplifier, $(1 + A\beta)$, as distortion in the open loop gain is reduced by this factor in the closed-loop configuration.* The feedback factor is readily determined from the expression for the closed-loop gain A_f.

$$A_f = \frac{A}{(1+A\beta)} \qquad (1+A\beta) = \frac{A}{A_f}$$

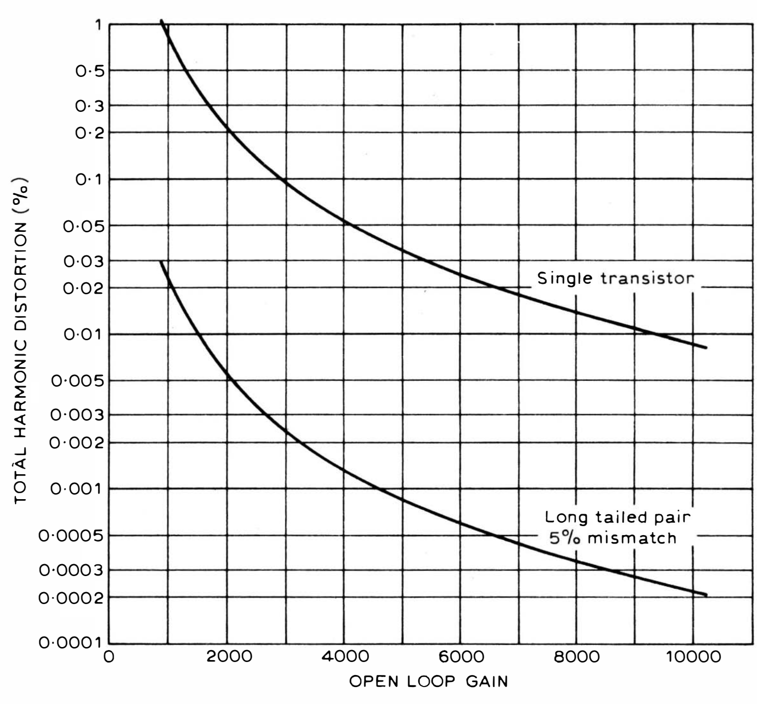
With RIAA equalization the feedback factor should be determined for frequencies below 50Hz as the amount of feedback reaches a minimum at these frequencies. In the present design the sensitivity is specified as 100mV output for a 5mV input at 1kHz and therefore at frequencies below 50Hz the closed-loop gain of the amplifier will be 200. From this equation the feedback factor is therefore equal to 5 and the closed-loop distortion will be $4.3/5 = 0.86\%$.

Repeating these calculations enables the distortion to be plotted as a function of the open-loop gain and this has been done in Fig 8 for a single transistor stage and a two transistor long-tailed pair stage in which the collector currents are matched to within 5%. With the single transistor input stage an open-loop gain of at least 9500 is required to meet the 0.01% distortion specification whereas with the two transistor long-tailed pair input stage the open-loop gain need only be 1500.

The open-loop gain also needs to be sufficient for the closed-loop gain to be

*This is not strictly correct because with frequency-dependent feedback all harmonic components are not subject to the same attenuation. With equalization which has a falling gain-frequency characteristic the distortion will therefore be less than the calculated value.

Fig. 8. Calculated distortion due to input stage of preamplifier as a function of open-loop gain.



closely defined as a function of frequency according to the RIAA equalization characteristic. At frequencies below 50Hz a closed-loop gain of 200 is required and an open-loop gain of 2000 would give an acceptable 20dB of negative feedback.

With a long-tailed pair input stage the minimum open-loop gain is therefore dictated by feedback requirements and should be approximately 2000, whereas with a single transistor input stage the open-loop gain is dictated by the distortion specification and should exceed 9500.

The input stage

The superior distortion performance of the long-tailed pair input stage compared to the single transistor input stage has been established beyond question. The signal-to-noise ratio of a long-tailed pair input stage is of course inferior to that of a single transistor input stage, but as shown in Appendix 4 the deterioration in the signal-to-noise ratio of an amplifier designed for use with a magnetic pickup is only 0.22 dB.* There seems to be little reason therefore for not using the long-tailed pair input stage unless the ultimate in noise performance is required.

Figure 9 shows the complete circuit diagram of the preamplifier. The long-tailed pair input transistors each operate at a collector current of approximately 90μA for optimum noise performance with a magnetic cartridge input and the tail current is derived from a current source to give a good positive supply rejection ratio and improve the common-mode performance of the amplifier. A single-ended output is taken from the input stage via a current mirror, the advantages of this arrangement being

—the useful gain of the input stage is doubled
—a good negative supply rejection ratio is achieved
—the current mirror can be used to balance the collector currents of the long-tailed pair.

The importance of balancing the long-tailed pair stage to obtain optimum distortion performance was emphasised in the first part of this article. With 10kΩ, 1% resistors in the current mirror overall negative feedback around the preamplifier maintains the collector currents of Tr2 and Tr3 to within 5% for up to 25mV mismatch in V_{BE} of Tr4 and Tr5.

The output stage

The noise contribution of the output stage of a preamplifier cannot be ignored but the design is primarily influenced by the overload capability, and therefore output voltage swing, that is required.

*Experimentally it has been found that with 2N5087 transistors the signal-to-noise ratio of the long-tailed pair input stage is approximately 0.6 dB worse than that of the common-emitter stage.

Low noise of series feedback + high overload of shunt

A low-noise, low-distortion audio pre-amplifier, equalized for use with a magnetic pick-up cartridge, has been developed using low cost, readily available components. The basic amplifier can however be considered as a high performance, 7.5 MHz unit-gain bandwidth operational amplifier which can easily be adapted for other purposes, e.g. different sensitivities and/or equalization.

Distortion measurements on the preamplifier have verified much of the theoretical treatment and have clearly shown that the distortion performance of a series feedback amplifier with a standard input is limited at high audio frequencies by distortion resulting from the common-mode input signal and the non-linearity of the common-mode input impedance. The common-mode input signal can however, be virtually eliminated by using an unconventional feedback connection in which the input signal is introduced directly in the feedback path of the amplifier. With this connection it is possible to achieve the low-noise performance of the series feedback connection with the high overload capability of the shunt feedback connection.

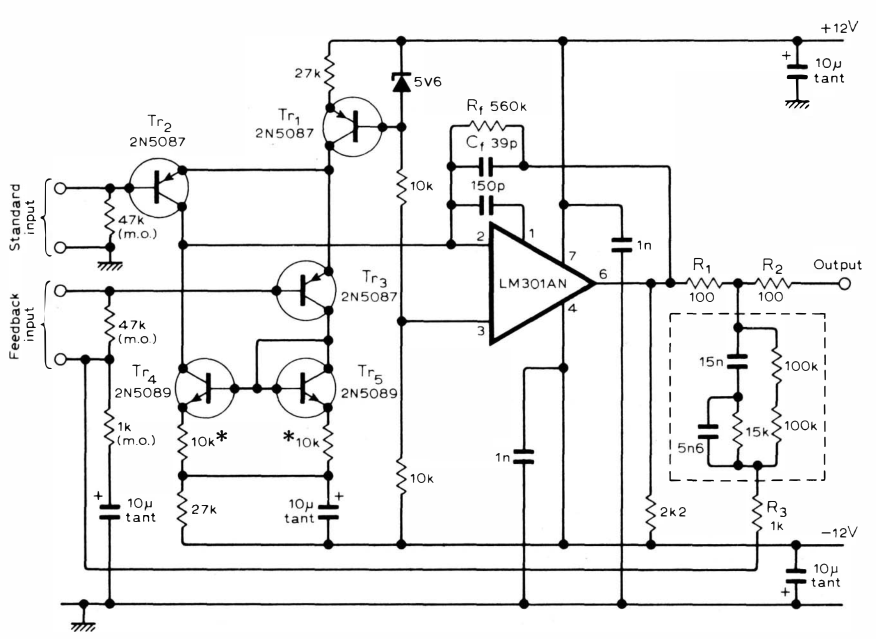
At low frequencies the distortion of a low-noise audio amplifier is dominated by the non-linearity of the differential-mode gain and ultimate performance is limited by the exponential relation between collector current and base-emitter voltage of the input stage transistor or transistors. The two-transistor long-tailed pair has a much more linear transfer characteristic than a single common-emitter input stage and enables a significant improvement in distortion performance to be achieved with only a slight deterioration in signal-to-noise ratio.

The design example shows how low-distortion design can be treated quantatively and that it is not difficult, at least in an audio preamplifier, to achieve an harmonic distortion of less than 0.005%. It may be argued that this level of performance is academic when other imperfections in an audio system are considered, but if it has been achieved at low cost then such an argument can only conform that progress at least has been made towards the ideal preamplifier.

Large voltage swings in any transistor circuit inevitably lead to distortion because of the effects of base-width modulation. Even the popular current-driven common-emitter stage is subject to this type of distortion, because variations of β with V_{CE} are not insignificant. A current-driven common-base configuration would probably be the most linear single-transistor output stage because the current gain α is relatively independent of V_{CE}. However the high output impedance of both the common-emitter and the common-base stage make them unsuitable for use as an output stage in a feedback amplifier unless the output is buffered to prevent instability with capacitive loads.

An output stage consisting of at least two transistors is therefore indicated

Fig. 9. Complete circuit diagram of RIAA equalized preamplifier. Unused input must be shorted. Resistors marked 10kΩ are matched to within 2%. Three of the input resistors should be metal oxide types.*



and at this point the use of an operational amplifier becomes attractive in terms of cost and performance. An integrated circuit operational amplifier with shunt feedback and the output stage operating in class A is used in the present design, the advantages of this arrangement being

—large output swing capability
—low distortion due to local feedback and class A output
—low output impedance
—virtual earth input minimizes voltage changes and therefore distortion of the preamplifier input stage
—optimum feedback configuration for low-noise amplification of the signal from the input stage
—the open-loop gain of the pre-amplifier is well defined.

The operational amplifier used in the output stage of the preamplifier has to meet certain large signal voltage swing and slew rate specifications to operate satisfactorily under overload conditions. The preamplifier is designed to give a nominal 100mV r.m.s. output with a 30dB overload capability which demands a maximum peak-to-peak output of approximately 9V. The maximum slew rate under these conditions for a sine wave output is calculated as follows

$$V_{out} = V_o \sin 2\pi ft$$

$$\frac{dV_{out}}{dt} = 2\pi f V_o \cos 2\pi ft$$

$$\left.\frac{dV_{out}}{dt}\right|_{max} = 2\pi f V_o$$

Evaluated at f = 20kHz for V_o = 4.47V (30dB overload) this indicates a maximum slew rate requirement of 0.56 V/μs.

The ubiquitous 741 operational amplifier is just capable of meeting the voltage swing and slew rate requirements but the LM301 is a much better alternative at little extra cost. With feedforward compensation[1] the LM 301 has a limiting slew rate of 10V/μs and a peak-to-peak voltage swing in excess of 24V at 20kHz. In addition whereas the 741 has a unity-gain bandwidth of 1 MHz, feedforward compensation extends the unity-gain bandwidth of the LM301 to 10MHz, a significant improvement as the loop roll-off frequency of the preamplifier is a function of the unity-gain bandwidth.

Little information is available concerning the distortion performance of general purpose integrated circuit operational amplifiers. However, Linsley Hood[2] has obtained figures of less than 0.02% harmonic distortion at 1V r.m.s. output with a 741 in a shunt feedback configuration and measurements by Walker[3] show that intermodulation distortion in an LM 301 under similar conditions is less than 0.03%. As the output stage of the preamplifier is contained within the overall negative feedback loop, it would appear that both of these amplifiers would enable the 0.01% distortion specification to be achieved.

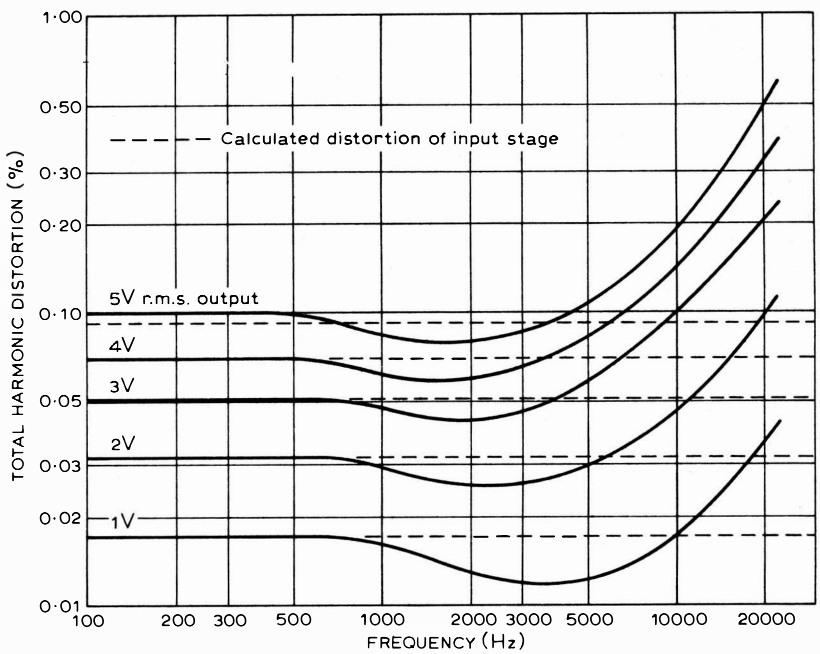


Fig. 10. Open-loop distortion of the preamplifier as a function of frequency and output amplitude.

Frequency compensation

The low-frequency open-loop gain of the amplifier is

$$A_o = -g_m R_f$$

where the mutual conductance of the input transistors g_m, is equal to 3.6mA/V with the transistors operating at a collector current of 90μA. The high-frequency break point of the input stage is calculated to be 12.0MHz and the h.f. break point of the output stage is 10MHz. Compensating the amplifier for unity loop gain at 7.5MHz gives a reasonable stability margin.

It is not necessary for the amplifier to be compensated for unconditional closed-loop stability as the feedback network which defines the equalization characteristic can be used to attenuate the loop gain. Thus the resistor R_3 in the equalization network (Fig. 9) usefully extends the frequency at which the loop gain must be rolled off by the compensation network to ensure stability by a factor of two.

The amplifier is compensated by the capacitor C_f in the output stage which gives a dominant pole in the open-loop response. The required value of C_f is given by

$$\frac{1}{2\pi C_f R_f} = \frac{7.5\times10^6}{A_o/2} = \frac{2\times7.5\times10^6}{g_m R_f}$$

which gives 38pF. For an open-loop gain of 2000 R_f needs to be 560kΩ (A_o/g_m) and the loop gain then rolls off at 7.5kHz.

It is interesting to note that the value of C_f necessary for stability is a function only of the input stage transconductance and the high frequency attenuation of the loop gain by the feedback network. If the high frequency attenuation of the feedback network can be increased, as may be possible for example in a high-gain equalized preamplifier, then the value of C_f may be reduced proportionately to maintain the 7.5 kHz break frequency in the loop response. It is not recommended that C_f is reduced below 10pF however as the operational amplifier output stage may become unstable within its own local feedback loop.

Resistors R_1 and R_2 in series with the output are used to isolate the LM301 from any load capacitance and prevent high frequency instability.

Performance

The distortion performance of the amplifier is presented graphically in Figs 10 & 11. Figure 10 shows the open-loop distortion of the amplifier as a function of frequency for several values of output voltage. At low frequencies the distortion corresponds closely to that predicted for the input stage. As the frequency is increased above 1kHz there is a slight reduction in distortion, probably as a result of the 3.25kHz break frequency in the output stage (for these measurements the amplifier was compensated for unconditional closed-loop stability) which will attenuate the predominantly third-order harmonic distortion components generated in the long-tailed pair input stage. Above 5kHz the distortion increases rapidly with frequency and must be attributed to the output stage of the amplifier as distortion generated in the input stage is independent of frequency. At 3.0V r.m.s. output however, corresponding approximately to 30dB overload, the distortion has only risen to 0.2% at 20kHz.

The distortion of the amplifier with

RIAA equalization is shown in Fig. 11. These characteristics were obtained using the standard input configuration and a source impedance equivalent to that of a 600mH cartridge. At low frequencies the distortion decreases with increasing frequency as expected because of the increase in loop gain of the amplifier. The distortion reaches a minimum at 1.5kHz and with a 3V output (30dB overload) the distortion is less than 0.001%. Above 2kHz the distortion increases rapidly with frequency until at 20kHz the distortion with a 3V output has risen to 0.1%.

Measurement with the feedback input connection have shown that the distortion is less than 0.005% at all frequencies up to 20kHz and all overload levels up to 30dB. Unfortunately it has not been possible to plot any meaningful distortion characteristics for the feedback input connection because of the difficulty in making reliable distortion measurements below 0.001%.

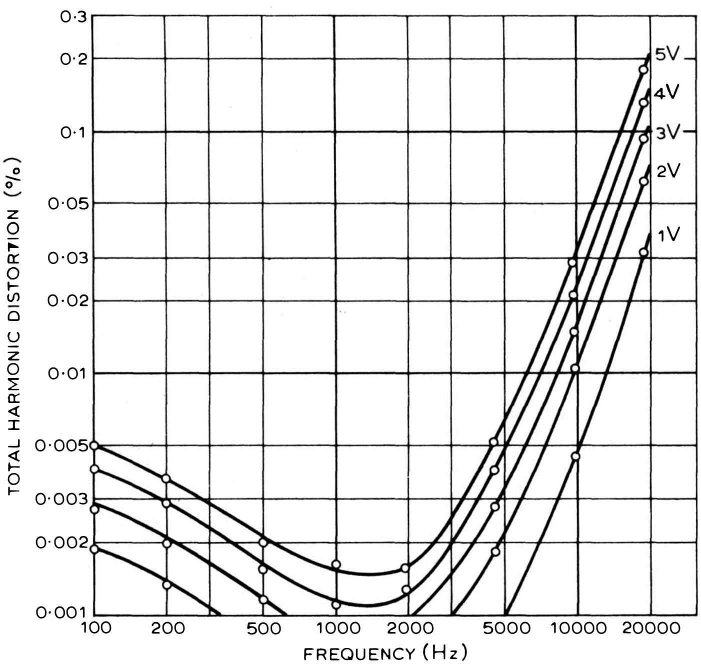
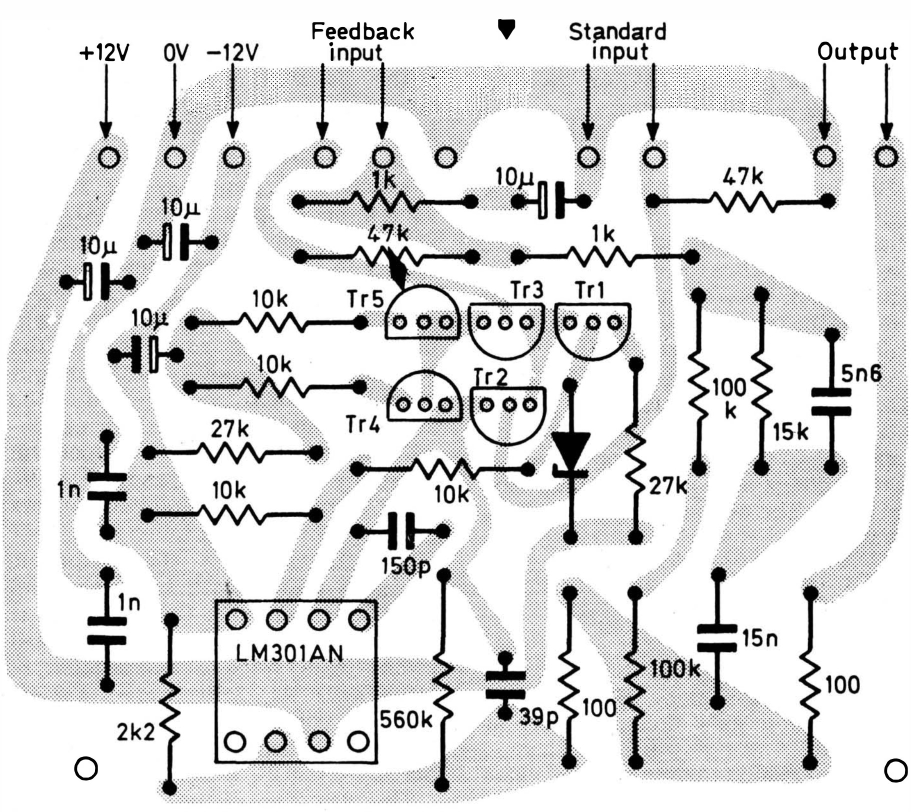


Fig. 11. Total harmonic distortion of the preamplifier, with RIAA equalization, as a function of frequency for various output amplitudes for standard input configuration.

Fig. 12. Printed circuit board layout viewed from component side. Ready-made and drilled boards will be available from M. R. Sagin, 23 Keynes Road, London, NW2.



The maximum output signal amplitude before clipping is 5.6V r.m.s. which gives a 35dB overload capability referred to 100mV.

Signal-to-noise ratio of the preamplifier is greater than 75dB ref. 5mV at 1kHz for both the standard and feedback input connection with a 600mH source inductance.

Construction

Figure 12 shows a printed circuit board layout of the preamplifier and two amplifiers for stereo operation can easily be mounted in an Eddystone 7134P die-cast box measuring 111×60×31mm. The printed circuit board allows for either the standard input or floating input connection. In my system the preamplifier is mounted directly adjacent to the pickup and no problems with hum or instability have been encountered with the floating input connection.

The power supply is not critical and the circuit operates satisfactorily with the positive and negative supplies derived from a simple half-wave rectifier with Zener stabilization. The positive and negative supplies should be capable of providing approximately 10mA.

Acknowledgements. The assistance of Dr D. A. Edwards with the computer programming and Mr D. H. Warne with the design of the printed circuit board is acknowledged.

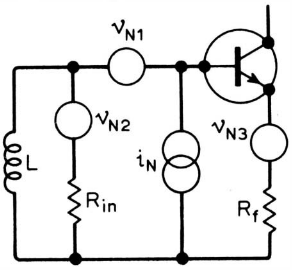
Appendix 3 —Total harmonic distortion (%) of a common emitter and long-tailed pair transistor stage due to the exponential relation between collector current and base-emitter voltage of a transistor.

Amplitude (mV)	Single transistor	Long-tailed pair 0% Mismatch	5% Mis match	10% Mis match
0.1	0.0967	0.0000312	0.00242	0.00484
0.2	0.193	0.000125	0.00484	0.00967
0.3	0.290	0.00218	0.00726	0.0145
0.4	0.387	0.000499	0.00968	0.0194
0.5	0.484	0.000780	0.0121	0.0242
0.6	0.580	0.00112	0.0146	0.0290
0.7	0.677	0.00153	0.0170	0.0339
0.8	0.774	0.00200	0.0194	0.0387
0.9	0.870	0.00253	0.0219	0.0435
1.0	0.967	0.00312	0.0244	0.0485
2.0	1.93	0.0125	0.0499	0.0975
3.0	2.90	0.0280	0.0777	0.148
4.0	3.87	0.0498	0.109	0.199
5.0	4.83	0.0778	0.143	0.253
6.0	5.79	0.112	0.182	0.309
7.0	6.76	0.152	0.226	0.368
8.0	7.72	0.198	0.276	0.431
9.0	8.68	0.251	0.330	0.497
10.0	9.63	0.309	0.390	0.566
11.0	10.6	0.373	0.455	0.640
12.0	11.5	0.443	0.526	0.718
13.0	12.5	0.519	0.602	0.800
14.0	13.4	0.600	0.683	0.887
15.0	14.4	0.687	0.770	0.978

Note. % mismatch for the long-tailed pair stage is defined by $2(I_{C1}-I_{C2}) / (I_{C1}+I_{C2})$, where I_{C1} and I_{C2} are the collector currents of the transistors.

Appendix 4 — Input stage noise

The noise generators of an amplifier with a single transistor common-emitter input stage and designed for use with a magnetic pick-up cartridge can be represented as



where v_{N1} is the equivalent noise voltage generator of the transistor, v_{N2} the equivalent noise voltage generator of the input resistance R_{in}, v_{N3} the equivalent noise voltage generator of the equivalent feedback network resistance R_f, i_N the equivalent noise current generator of the transistor, and L the inductance of the magnetic cartridge, assumed purely inductive.

The total mean square noise voltage at a frequency f for a bandwidth δf referred to the input can be shown to be

$$4kT\delta f\left\{R_{Nvl}+R_f+R_{in}\left[\frac{j\omega L}{R_{in}+j\omega L}\right]^2+\frac{1}{R_{Ni}}\left[\frac{R_{in}j\omega L}{R_{in}+j\omega L}\right]^2\right\}$$

$$=4kT\delta f\left\{R_{Nvl}+R_f+R_{in}\left[\frac{(\omega/\omega_o)^2}{1+(\omega/\omega_o)^2}\right]+\frac{R_{in}{}^2}{R_{Ni}}\left[\frac{(\omega/\omega_o)^2}{1+(\omega/\omega_o)^2}\right]\right\}$$

where the noise voltage and current generators have been replaced by equivalent noise resistors and $\omega_o = R_{in}/L$. If this noise is now passed through an RIAA equalizing network with a transfer function A(jf), the total mean square noise voltage over a band of frequencies is

$$\overline{V_N{}^2}=4kT\int\left\{R_{Nvl}+R_f+R_{in}\left[\frac{(f/f_o)^2}{1+(f/f_o)^2}\right]+\frac{R_{in}{}^2}{R_{Ni}}\left[\frac{(f/f_o)^2}{1+(f/f_o)^2}\right]\right\}|A(jf)|^2df \quad \ldots (5)$$

With L of 600mH and R_{in} of 50kΩ, it can be shown[3] that

$$\int_{50}^{20{,}000}\frac{(f/f_o)^2}{1+(f/f_o)^2}\,|A(jf)|^2df=298.4$$

and it is readily shown that

$$\int_{50}^{20{,}000}|A(jf)|^2df=8.015\times10^3$$

For a 2N5087 transistor operating at I_c of 100μA with a β of 250 and neglecting flicker noise the equivalent noise resistors are[4]

$$R_{Nvl}=(r_{bb'}+1/2g_m)\approx200\Omega$$
$$R_{Ni}=2\beta/g_m=1.25\times10^5\Omega$$

Putting R_f = 1000Ω, the value used in the design example, and substituting for all values in equation 5 gives

$$\overline{V_N{}^2}=2.655\times10^{14}+1.327\times10^{-13}+2.472\times10^{-13}+9.887\times10^{-14}$$

where the components are due to the noise voltage of the transistor, the noise voltage of the feedback network; the noise voltage of the input resistance and the noise current of the transistor respectively. Thus

$$V_N=\sqrt{5.053\times10^{-13}}=0.711\mu V$$

which corresponds to a signal-to-noise ratio of 76.94dB referred to 5mV.

With the long-tailed pair input stage two additional noise generators are introduced into the equivalent circuit as shown in Fig. A4. These noise generators are identical with the noise generators of the transistor in the common-emitter input stage (they are not correlated however) and the total mean square noise voltage is now

$$\overline{V_N{}^2}=5.053\times10^{-13}+2.655\times10^{-14}\times7.14\times10^{-16}$$

The first term of this expression is the noise present in the single transistor input stage and the last two terms represent the additional noise due to the noise voltage and noise current generators respectively of the second transistor. Thus

$$V_N=\sqrt{5.326\times10^{-13}}=0.730\mu V$$

which corresponds to a signal-to-noise ratio of 76.72dB referred to 5mV. The deterioration in signal-to-noise ratio of the long-tailed pair compared with the single transistor is therefore 0.22dB.

The reason for only a small deterioration in signal-to-noise ratio with the long-tailed pair is that the noise voltage generator associated with the additional transistor is small compared with the noise voltage associated with the 50kΩ input resistance and the noise voltage produced across the source impedance by the noise current generator of the original transistor. The noise current generator of the additional transistor produces a negligible noise voltage across the low impedance of the feedback network.

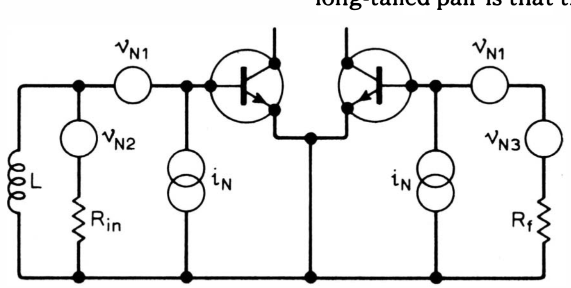


Fig. A4. Equivalent noise circuit of long-tailed pair stage.

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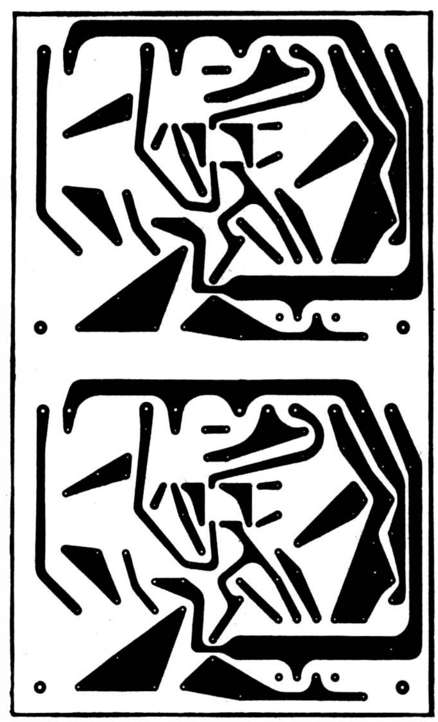

Drilled boards to this design, shown actual size, will be available for £1.65 inclusive from M. R. Sagin, 23 Keynes Road, London NW2.

Surround-sound decoders — correction

An error in the components list for the Sansui Variomatrix decoder circuit (September 1976 issue) was regretably perpetuated in the variable-matrix H decoder list on page 38 of the June issue. Values of C_{63} to C_{65} and of C_{87}, C_{90} and C_{91} should be ten times greater than shown. (In the original QS list this also applies to C_{55}, C_{56} and C_{73} to C_{75}. QS kit constructors will also have noticed values for R_{91} and R_{92} were transposed in the list with those of R_{125} and R_{126}, and that R_{107}, R_{108} are 6.8kΩ and not 68kΩ.) Input capacitors for the output phase shift circuits on page 35 are 4.7μF.

Should constructors of either circuit find that the voltages on pins 5-8 and 12-15 on the HA1327 i.cs do not reach their proper value of 5V, Sansui recommend a modification, which we understand is now applied to all Variomatrix circuits. Capacitors C_{58} to C_{61} and C_{79}, C_{80}, C_{85} and C_{86} should be taken to the +24V rail rather than 0V; this means capacitor polarity must be reversed.