Designer's Notebook

Power Amp Design
Cutting through the superstition and hype of **power amplifier design.**

By John Linsley Hood

A CONSIDERABLE amount of mystique surrounds the whole field of audio amplifier design, particularly power amplifiers, and a vast amount of time is spent in labs and listening rooms sorting the good from the not-so-good. How much of this mystique is justified is a speculative question.

My own feelings in this matter, particularly in respect to hifi devotees, are highly ambivalent. On one hand, I am convinced that much of the stock in trade of hifi journals, reviewers, and manufacturers who follow in the train of their approval, is built on the identification and exaggeration of differences which are, in reality, fairly small. On the other hand, I know that there are sound differences between differing designs. Contributing factors undoubtedly reside in a gray area of technology not yet clearly defined.

Output Power

Because of the nature of the sensitivity of the human ear, doubling the power fed to a speaker does not make the resultant sound twice as loud. In reality, this is a logarithmic relationship, in which apparent increases in sound loudness are ted to power by by the equation W = k.log P2/Pl. While precision in this field is difficult because the loudness/power relationship varies with level and frequency, one can say roughly that ten times the power doubles the loudness; in other words, 3W, 30W, and 300W are increments corresponding to doublings of loudness. One can see that very big steps in amplifier power are required to get significant gains in sound level. The converse is also true, that microscopic amounts of sound power, like the buzzing of a fly's wings or amplifier noise voltages, are still quite audible.

Power is related to the square of the voltage or current; if an output power of 100W is required into 8 ohms, the RMS output voltage must be 28.3V. This corresponds to an 80V peak-to-peak swing. Allowing for voltage drops in the output transistors, emitter resistors, etc., the supply should be closer to 100V. Now let us suppose that the loudspeaker impedance drops to 2 ohms at 15KHz. 100W will then correspond to 10A peak current. Our amplifier, to meet this spec, would require 28.3V RMS and 10A peak current capability. Fortunately, it usually doesn't require them simultaneously.

Power Bandwidth

The audio spectrum is assumed to lie between 20Hz and 20KHz, but with average listening rooms, average listeners, and average program material, 45Hz to 15KHz is more realistic. Happily, there isn't much program energy above 10KHz, so we don't need to cater for maximum power in this region unless the equipment is going to be reviewed. There is also not much very low frequency content from program material or loudspeakers. Unfortunately, if the amplifier cannot operate well below 30-40Hz it may sound thin and will probably overload on record rumble and the like.

Some listeners with acute hearing can undoubtedly hear the difference when HF response is curtailed even if this is above their ear's frequency response, because of the absence of beat-note effects due to the interaction of HF sounds within the non-linearities of their own ears. A better I-IF response may not make things sound better, but it can make them sound different.

This is where the first of the needs to compromise occurs. With typical power junction transistors, which are fairly sluggish devices, increased HF power bandwidth can only be obtained a the expense of loop stability in a negative feedback amplifier. If loop stability is poor, the amplifier transient response is bad, and this can introduce some pretty drastic distortions into pulse type signals such as drum beats or cymbal clashes. An amplifier with good loop stability is usually much more pleasant to listen to, and will certainly be less critical about the speaker load characteristics.

Feedback and Stability

Negative feedback is the comparison of the input signal with the output and the generation of a corrective adjustment to the input signal to make sure input and output are closely identical. It is the major tool of circuit designers, but there are snags.

We need to make sure that our use of feedback does not make the whole system unstable, and this is particularly the case for amplifiers which have to drive speakers, since these are notoriously com-

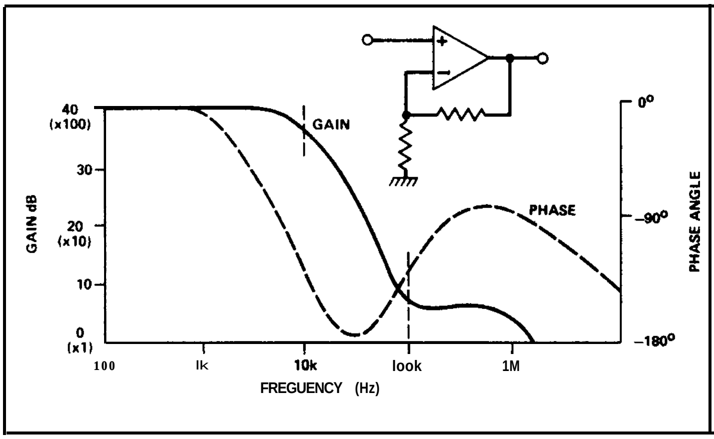


Fig. I Gain/phase or Bode diagram for conditionally stable feed-back amplifier (a change of load might make thb amplifier oscillate at 30-50 kHz).

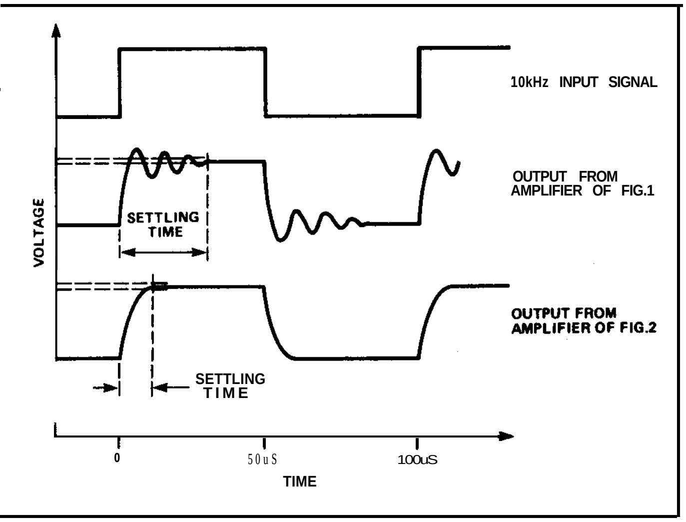


Fig. 2 Influence of gain and phase margins of NFB amplifiers and settling time.

plex in their impedance and delayed response characteristics. It is also essential to remember that a feedback path is just what its name suggests: a means by which signal components can be fed back from the speaker to the input. Since speakers can generate signals of their own because of internal cabinet echoes and inadequately damped reflections, we have to watch this point.

With regard to loop stability, this field was investigated by Bode and Nyquist many years ago in respect to closed-loop servomechanisms. I find the Bode diagram of **Fig. 1** the easiest to follow and explain. In this, the gain and phase shift are shown as a function of frequency. If the amplifier has a gain of 1 or more at a frequency where the feedback is in phase with the input (feedback shift of 180 deg.) it will oscillate. The reason for this is simple: the feedback path is providing an input signal of the right size and phase to generate the actual output without the need for any other signal at all (the circuit starts by amplifying its own inherent noise). If the gain is more than unity, the output will continue to increase until some other effect such as clipping reduces it to unity.

It isn't sufficient merely to ensure that the amplifier doesn't oscillate on load. There must be an adequate margin of gain or phase at the unity gain point to make sure that the amplifier is not triggered into misbehaviour during transients in the input signal. In particular, the settling time, or time required following an input voltage excursion for the circuit to settle to the new value, depends solely on the system's speed and stability margin, as shown in **Fig. 2. I** would very much much like reviewers of audio amps to measure this value for a step input with a real live speaker load, since this is one of the areas where the pursuit of very low THD figures at the top end of the audio spectrum can lead to circuit design characteristics which are bad for the transient handling qualities of the amplifier and make it fussy about the speaker with which it is used. It seems pointless to try to reduce .1 percent distortion at 20KHz to .01 percent if the price you pay is 20 to 50 percent distortion on transient signals.

In a typical audio amplifier, the major factor which dominates the gain and HF phase shift characteristics is the relative slowness of the output transistors. The faster the response of these, the easier it is to design a good, stable amp. The catch is that fast junction output transistors are also more fragile and require more restrictive protection circuitry. This makes the amp less good at driving low impedance loads. The answer, and a virtually complete one too, is to use power MOSFETs. Some of the recent ones have almost instantaneous response and are more linear than either tubes or junction transistors. They too have drawbacks, of which the main one is that they are particular about the phase characteristics of their loads, but there's a simple design answer to this. It is possible to design MOSFET amplifiers that are ten times better than their forerunners.

Distortion

Not all distortions are equal in their unpleasing effects, and the characteristic of distortion can be greatly influenced by the relative phases of the signal components. This effect has relevance to the behaviour of multi-driver speakers, which can jumble up the phases of a signal and thereby alter the nature of the amplifier nasties. Our efforts in the design exercise should be aimed at removing components of distortion which can be so transformed.

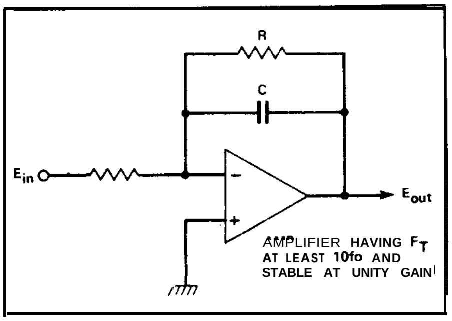


Fig. **3** ***Feedback amplifier having good gain and phase margins.***

Obviously it is helpful if we can keep the distortion of the amplifier as low as possible before feedback is applied, in that low feedback leeds to better loop stability. A useful design yardstick, cribbed from servo theory, is to determine the time constants of the bits of the circuit which lead to HF rolloff, and then make one of these ten times the size of the rest. This isn't as arbitrary as it seems; if we wish to end up with an ideal Bode plot, we must remember that the phase shift due to an RC element begins one decade below the -3dB rolloff point. In this way, the system will behave as though it had only a single phase-shifting component.

Another useful design approach is to limit the system to no more than two stages within the loop, adding more loops if necessary. This isn't always possible with power amps, so other approaches, such as the phase-lead generating step network, may be necessary. One can always put in a an internal loop operating only at HF where transistor trouble is likely to arise, and include only two gain stages.

The aim of the designer should be to produce an amplifier in which the harmonic distortion is as low as possible, and the gain bandwidth is as high as possible, before the application of feedback. The feedback should be used mainly to control the gain and output impedance characteristics rather than as a way of lowering distortion.

As far as the feedback path is concerned, it is best not to use a parallel capacitor across the feedback resistor as this can make the amplifier sound less good on some speaker units. It is also helpful if a small resistor in the range of 0.15 to 0.33 ohms is added in series with the output to act as one element of an attenuator. Against the much lower output impedance of the amplifier, this will assist in attenuating signals originating from the speaker itself. ***(Editor's note: this goes against thestandard dogma of having the lowest possible output impedance to raise the damping factor. Comments from designers are invited.)***

A final thought with regard to amplifier sound. The human ear is not a particularly good judge of distortion in that a number of randomly chosen people, all reasonable aware of sound quality, either preferred the addition of 0.3 percent third harmonic distortion, or had no specific preferences. Similar work has been done by other investigators. This tends to cast doubt on the value of such judgements, where the listener may actually prefer inferior equipment because it adds a wanted colouration to the sound.

A Practical Design

My preference for power MOSFETs as the output devices is definite, but not just simple source followers where they require an output inductor in the speaker lead to

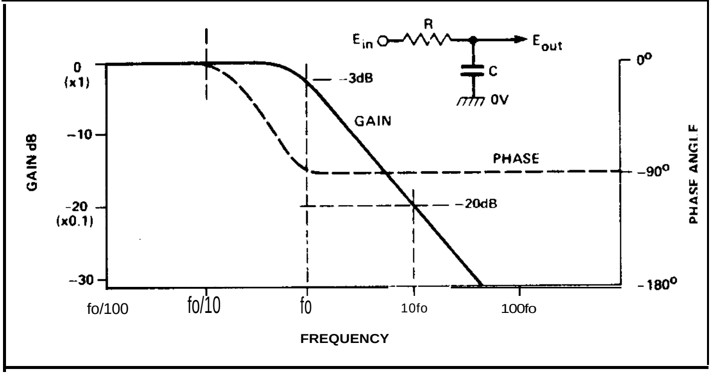
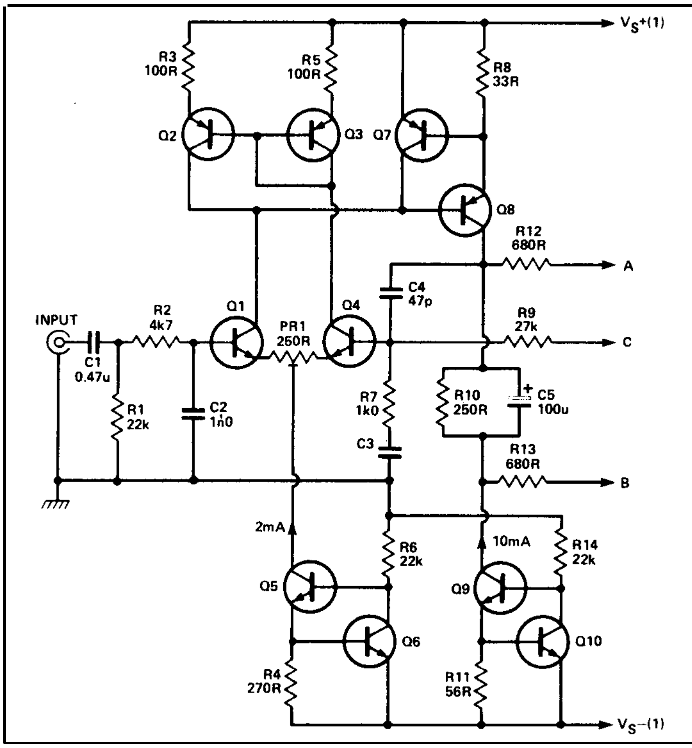


Fig. 4 Gain/phase diagram for a simple RC attenuator; the characteristics of the amplifier in Fig. 2 would be similar to this.



Fig, 5 Power amp. driver stage.

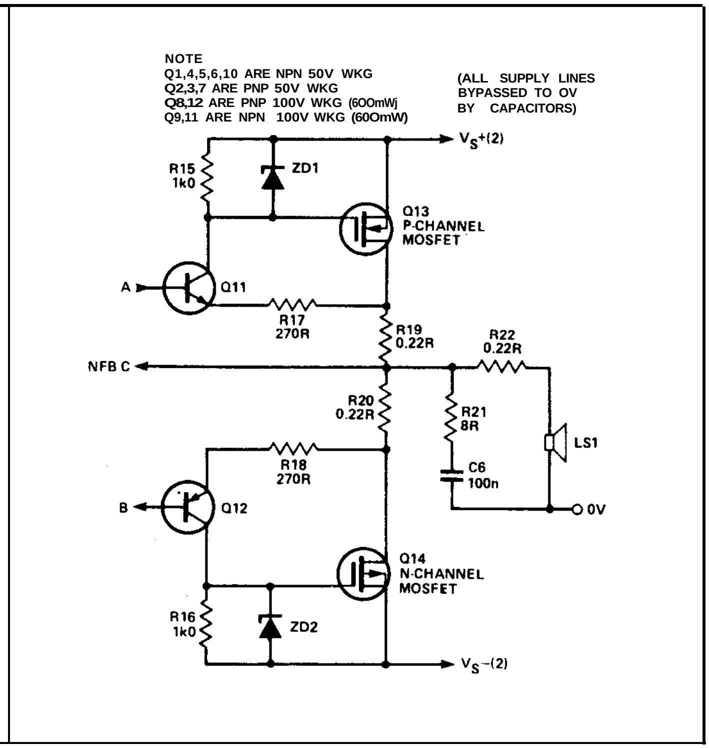


Fig. 6 MOSFET output stage.

prevent VHF oscillation due to the device's own small lead inductance. If the power MOSFET is used in combination with a bipolar small-signal transistor as shown, this small inelegance can be avoided without degradation of the fast response of the MOSFET. The only necessary device protection can then be an output fuse or a current-limited DC supply.

A typical small-signal voltage amplifier to drive this output stage is shown. Because small-signal transistors are cheap I have been lavish in using them to confer some practical benefit to the circuit. R2 and C2 roll off the response at 33KHz to prevent very fast transients from affecting the circuit. Rl defines the zero-volts DC level. Q5 and Q6 are a compound constant-current source, setting 1mA each through Ql and Q4, and minimizing noise pickup through the negative supply. PRl allows the output DC voltage (offset voltage) to be set to precisely zero.

I have chosen to use a current mirror, Q2 and Q3, as the load for the differential amp in order to combine the signals from the two input transistors and optimize the input stage gain. From this, the signal is fed to the second stage PNP amplifier Q8, again loaded by a constant current source to ensure high AC gain and low distortion. R1 1 is chosen to give a collector current of 1OmA in order to lessen the effect of variable drive current into the output stage and to push the maximum slewing rate possible with C4 up to a high value.

RlO, bypassed by C5, provides the 2.5V bias to set the output stage at lOOmA quiescent current.

Under overload conditions, Q9 and 10 limit the current which can be drawn from the -Ve line to lOmA. Q7 and R8 provide a similar protective function for Q8.

R9, R7 and C3 provide the negative feedback path from the output stage to stabilize the AC gain to 27 which allows maximum output (in this case, 50W) from 0 VU or .77V input. C4 provides HF stabilization by means of an internal HF feedback loop enclosing the two stages Q1/4 and Q8, which gives a good well-damped transient response, especially with reactive loads. Eight ohm speakers most definitely do not behave like resistances.

R12 and R13 serve the useful function of preventing temporary latch-up if the amplifier is driven into clipping; in their absence clicks and bangs are prolonged and sound louder.

The output stage is unusual in that it uses MOSFETs in a compound emitter follower configuration with Qll and Q12 as the input devices. Excess voltage across the gate/source junctions is prevented by ZDl and ZD2, while R15/16 and R16/18 limit the AC gain of the output to 5.

R22 is the resistor which attenuates signals returned from the speaker unit, and R21/C6 is the Zobel network which prevents the output from seeing an open circuit if the speaker is removed.

The power supplies are shown with two numbers; if two channels are used, it is advantageous to use two separate power supplies to prevent interaction, particularly to prevent power supply nasties from intruding into the sensitive input stages. You could always run the inputs from a regulator, since it draws only 12mA; you can then use a single large power supply with protective fuses in the speaker lines.

This article is meant as a design exercise; I am not going to fill in all the small details of power supplies, PCB, and so on. Designers may wish to do a bit of "fine-tuning" to make sure it as good as possible in all the various conflicting requirements of the system. ■

Transient intermodulation in amplifiers.

Simpler design procedure for t.i.m.-free amplifiers

by Bert Sundqvist

The usual way to avoid transient intermodulation distortion in an audio power amplifier is to use a very large open-loop bandwith and a high-frequency preamplifier roll-off.

In this article it is shown that this is not necessarily the only way; it is possible to reach the same goal by making the first stage inside the feedback loop determine the open-loop bandwith. This bandwidth can then be arbitrarily low, permitting the use of standard lag compensation stabilization.

During the last few years it has become more and more obvious that the traditional steady-state measurements of harmonic and intermodulation distortion in an audio system do not give the whole truth about the qualities of the system when handling complex signals like music. As a result, much work has been done in studying the, dynamic behaviour of different links of the audio reproducing chain.

The most interesting work in this field in recent years is probably Professor M. Otala's identification of the mechanisms producing transient intermodulation distortion. Work by Otala and others[1-5] show that negative feedback, when incorrectly used in an amplifier design, may make the amplifier sound worse than it did without feedback, while measurements of steady-state harmonic and intermodulation distotiion show an improvement in amplifier quality (Jan., pp. 41-3).

Transient intermodulation arises when heavy negative feedback is applied to an amplifier with low open-loop bandwith. It is basically an overload phenomenon, giving an audible result that resembles crossover distortion. Transient intermodulation can be avoided by careful design'-" and probably the best known of the design rules that have evolved is that the amplifier open-loop bandwidth should be greater than the bandwidth of the preceding preamplifier or transducer, which must therefore not be unnecessarily large. A preamplifier bandwidth of several hundred kilohertz might give power amplifier troubles and should be rolled off using a passive RC filter.

In a power amplifier, a large open-loop bandwidth is not easy to obtain. Firstly, fast power transistors are neither easily obtained nor cheap. Secondly, the simplest way to stabilize an amplifier is to use lag compensation. which requires a dominant low-frequency pole to be inserted in the open-loop frequency response of the amplifier. Whe pushing this pole above 20 or even 50kHz, the rest of the amplifier must be designed for a bandwidth of perhaps several megahertz. This method can, of course, be used and has been very successful[4, 6]. The first difficulty can be overcome by using the output transistors in the emitter-follower configuration, thus increasing their cut-off frequency. The second can be evaded by using lead compensation[3,6] instead.

There are other drawbacks with extremely wide-band amplifiers; for example, such an amplifier must be very well shielded, as it is prone to pick up radio transmissions inside (and outside) its passband. High frequency noise could also be a problem, from the intermodulation point of view. However, there is no doubt that designing a t.i.m. - free amplifier is a rewarding task for the serious listener, as it is particularly annoying[4,5]; a t.i.m.-free amplifier sounds better than most traditional designs, especially on transient-rich musical material.

Is there, then, a way to design a t.i.m.-free amplifier without having to rely on a very high open-loo@ bandwidth? To answer this question we take a close look at the mechanisms producing t.i.m.

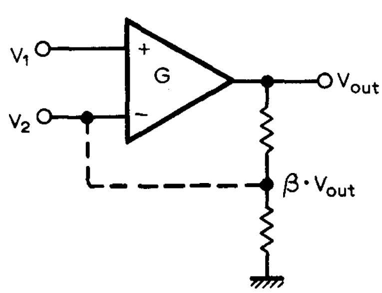


Fig. 1. Single stage amplifier.

Feedback in an amplifier

Suppose that we have a one-stage amplifier as in Fig. 1. The gain of this stage can be approximated by $V_{out} = G(V_2 - V_2)$ where G = $\boldsymbol{Aa/(a+s)}$, with $S = j\omega$; we have a low frequency gain of G= A and an upper cut-off frequency $2\pi f_a = \boldsymbol{a}$. If we now apply the input signal V_{in} to input 1 and a feedback signal βV_{out} to input 2 we get $V_{out} = V_{in} G/(1 + \beta G) = V_{in} Aa/(s + a(1 + \beta A))$.

From this equation the low-frequency gain with feedback is $A/(1 + \beta A) \approx \beta^{-1}$, and the upper cut-off frequency is now $2\pi f_c = a(1 + \beta A)$ Further analysis shows that low frequency distortion, rise time and output impedance have been reduced and input impedance has been increased by a large factor. Thus, on this single stage, negative feedback has nothing but beneficial effects.

If the two similar single-stage amplifiers of Fig. 2, with gains G, = Aa/(a + s) and G_2 =Bbl(b+s), are cascaded, total gain is $G = G_1 G_2$ = ABab/(a + s) ***(b + s), see*** Fig. 3. If we now apply feedback in the same way as before we obtain

$$V_{out} = V_{in} \frac{ABab}{(s+a)(s+b)\left[1 + \beta \frac{ABab}{(a+s)(b+s)}\right]}$$

$$= V_{in} \frac{ABab}{s^2 + s(a+b) + ab(1 + \beta AB)}$$

The non-inverting configuration has

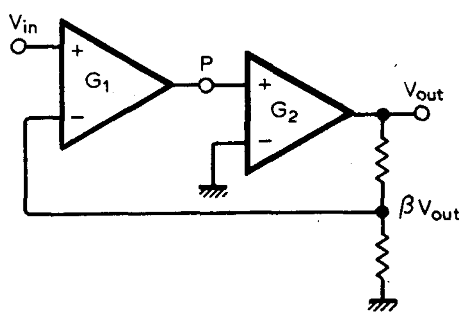


Fig.2. Two-stage amplifier with overall ***negative feedback.***

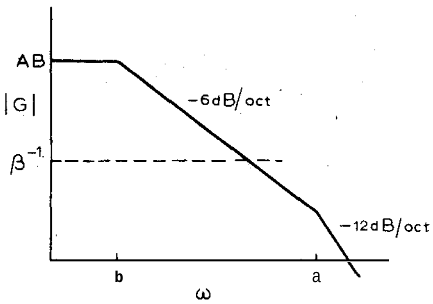


Fig.3. ***Gain vs frequency plot for the amplifier*** in Fig. 2; both axes logarithmic.

been chosen to avoid confusion in signs. The open-loop gain for this cascaded amplifier has two poles, at a and b. To obtain a stable amplifier it is necessary that the open-loop gain diminishes by less than 12dB/octave at the intersection of the open-loop gain curve and the desired closed-loop gain line (broken line in Fig. 3). Supposing A and B to be large we thus have, with feedback, a stable amplifier in which we probably have reduced harmonic and intermodulation distortion to very low values and which has a very large closed-loop bandwidth.

Dynamic considerations

To see how transient intermodulation arises and thus how it can be avoided consider the voltage at point P (Fig. 2). The voltage V_p at this point is

$$V_p = V_{out}/G_2 = G_1(V_{in} - \beta V_{out}) = V_{in}\frac{Aa(b+s)}{s^2+s(a+b)+ab(1+\beta AB)} \qquad (1)$$

As a suitable transient signal we can apply a unit step voltage to the input, that is

$$V_{in}(t) = \begin{cases} 0, & t<0 \\ 1, & t>0 \end{cases}$$

The voltages V_p and V_{out} can easily be found as functions of time by using standard Laplace transform techniques. First we solve the equation $s^2 + s(a+b) + ab(1+\beta AB) = 0$ to find the roots $p_{1,2} = -0.5(a+b) \pm 0.5[(a-b)^2 - 4ab\beta AB]^{1/2}$. We then find, for p_1 and p_2 both real and $t \geqslant 0$:

$$V_{out}(t) = \frac{AB}{1+\beta AB}\left[1 + \frac{p_2 e^{p_1 t}}{p_1-p_2} - \frac{p_1 e^{p_2 t}}{p_1-p_2}\right]$$

$$V_p(t) = \frac{A}{1+\beta AB}\left[1 + \frac{(b+p_2)p_2 e^{p_1 t}}{(p_1-p_2)b} - \frac{(b+p_1)p_1 e^{p_2 t}}{(p_1-p_2)b}\right]$$

By taking the time derivative of these two equations we find that V_{out} is always monotonically rising with no overshoot, and that the derivative of V_p with respect to time is zero for

$$t = t_0 = (p_1-p_2)^{-1}\log_e((b+p_2)/(b+p_1)) \qquad (2)$$

This means that for $t_0 \geqslant 0$ we must have a maximum in V_p at time $t = t_0$. This maximum value of V_p might be very large, and here is the mechanism that produces t.i.m. If the maximum value (V_{pmax}) of V_p is larger than the maximum voltage capability of the amplifier at point P, we get an overload situation in which the amplifier may be blocked for several milliseconds, thus causing severe intermodulation. Fig. 4 shows a plot of $V_{pmax}/V_p(t\to\infty)$ versus a for $b = 10^3$ and $b = 10^4$ and for different values of, βAB. The value of $V_{pmax}/V_p(t\to\infty)$ is approximately equal to βAB if a is large. To see why, let $a\to\infty$ in equation 1:

$$V_p = V_{in}\frac{A(b+s)}{s+b(1+\beta AB)}.$$

With V_{in} a unit step voltage as before this gives

$$V_p(t) = \frac{A}{(1+\beta AB)}\left[1 + \beta AB\, e^{-tb(1+\beta AB)}\right]$$

and $V_{pmax} = (1+\beta AB)V_p(t\to\infty)$, in agreement with Fig. 4 (cf also Fig.5,

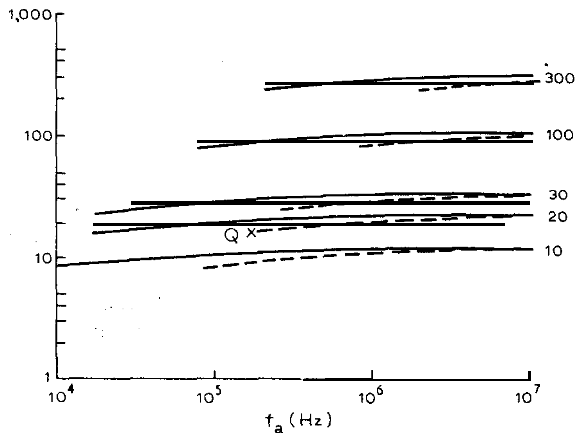


Fig. 4. Plots of V_{pmax}/V_p $(t\to\infty)$ vs $f_a = a/2\pi$ (in ***Hz***) ***for*** βAB = ***10, 20, 30,100, and*** 300. ***Broken lines:*** b = 10^4, solid lines: ***b*** = 16.

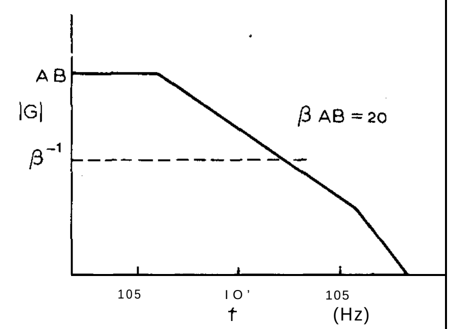


Fig. 5. Gain vs frequency plot for the amplifiers in the ***example (logarithmic axes).***

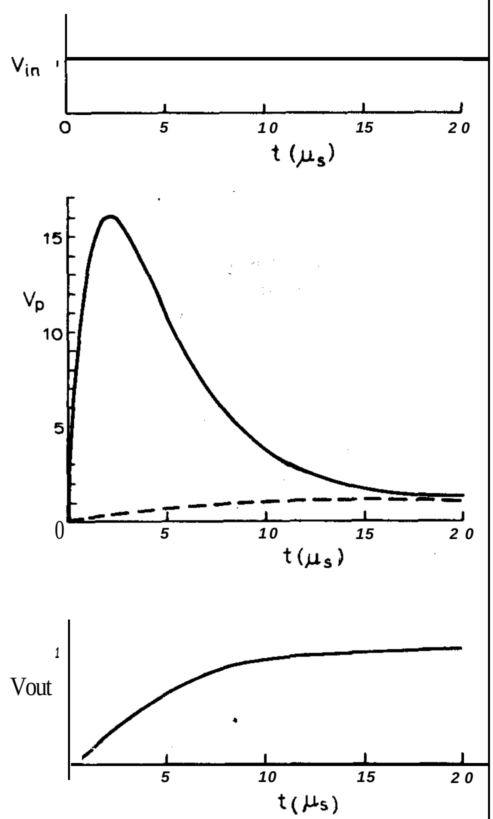


Fig. 6. Plot ***of*** $V_{in}(t)$, $V_p(t)$, ***and*** $V_{out}(t)$ ***for two amplifiers with a step*** voltage input signal. V_{in} and V_{out} ***are*** the same in ***both cases; for*** V_p ***we have: solid*** line: case 1, broken line: case 2 (see text).

ref 1 for $\gamma = 0$). Thus for a $\gg$ b we always **gei** for a step input, a maximum voltage V_p that is approqimately $(1 + \beta AB)$ times the steady state voltage at infinite time.

To eliminate t.i.m. we want to minimize or, still better, avoid this overshoot. One way to do this is to use a low value of ßAB. By limiting the bandwidth of the pre-amplifier we can then decrease the overshoot still further or even eliminate it by slowing down the rise of the input voltage', and in this way it is thus possible to design an amplifier with very low t.i.m.

There is, however, another way. If $t_0 < 0$ we see that V_p rises monotonically towards its final value and no overvoltage blocking is possible. From equation 2 we see that $t_o < 0$ is equivalent to $(b+p_2)/(b+p_1) < 1$, which is equivalent to $a\text{-}b < ((a\text{-}b)^2 - 4ab\beta AB)^{1/2}$. Thus, if $a r b$ no blocking can occur and no t.i.m. is generated, however small a is! This possibility seems to have been overlooked earlier.

Look at a simple example. Suppose that a two-stage amplifier has open-loop stage bandwidths a and b. We study two cases: case 1 a = 10^6, b = 10^4 (this resembles those studied in refs 1 and 4; it is shown as point Q in Fig. 4) and case 2 $a = 10^4$, b= 10^6. In both cases $1 + \beta AB = 21$. The gain vs frequency plot in Fig. 5 describes both amplifiers equally well, and shows two things. The amplifier is probably stable and has a closed-loop bandwidth of approximately 30kHz. And secondly, as the amplifier open-loop bandwidth is only $10^4/2\pi \approx 1.6$kHz this amplifier might give rise to appreciable values of t.i.m., even if preceded by a pre-amplifier with 2OkHz bandwidth'.

Fig.6 plots what happens if we apply a umit step voltage V_{in} to the amplifier input. (All voltages have been normalized to give $V_{in} = V_p = V_{out} = 1$ at infinite time). In both case 1 and case 2 we have the same $V_{in}(t)$ and V,,,(t), if the amplifiers have infinite voltage capabilities. V,(t), however, differs strongly between the two cases, and we see that while the amplifier in case 1 might produce severe t.i.m. with a transient input, this is not possible in case 2. It should be pointed out that if the amplifier in case 1 was designed with this situation in mind and the gain A before the "slow" stage 2 was kept low, an overshoot of this magnitude might be within the voltage capabilities of stage 1 and thus no harm, that is, t.i.m., would be done in either case. However, from Fig. 6 and the preceding discussion it seems wise to let the first stage in the feedback loop determine the overall open-loop bandwidth.

Conclusions

A good design procedure to obtain a t.i.m.-free amplifier is given in refs l-3,6. From the preceding discussion in this article, however, it seems that this procedure could be simplified. Simply stated: instead of designing the power amplifier for Bn open-loop bandwidth greater than that of the pre-amplifier, all that is needed to avoid t.i.m. is to let the first stage in the power amplifier determine the open-loop bandwidth. This bandwidth could then, theoretically, have any value; even with an open-loop bandwidth of 1Hz we would still have no t.i.m.! On the other hand, what should always be avoided is to let the last stage be the slowest, especially if this has a low gain.

Low first stage bandwidth could be obtained in several ways, for instance, by input lag compensation*, by using a very-high-impedance current source as collector or by using a very low collector current in the input stage. The low current technique has the advantage of giving at the same time very low input noise. One drawback is that the second stage in this case must have high input impedance and low input capacitance so as not to exceed the first stage output current capability and thus cause t.i.m. in this way instead*.

The stages following the first can be designed using accepted "rules"[2,3,6] Transistors should be run at high collector currents and voltages to give large overload margins and local feedback used to obtain a high bandwidth. Distortion can be reduced by using a symmetrical design. If the input stage bandwidth is not low enough to give a stable amplifier at the desired feedback lead compensation can be used to enhance stability.

By designing the power amplifier in this way it would also be possible to use larger amounts of feedback than in an amplifier relying only on a wide bandwidth to eliminate t.i.m., and thus very low harmonic and intermodulation distortion could easily be obtained. However, this possibility should be used with caution as there is always a possibility of current or voltage limiting at some stage in a real amplifier with heavy enough feedback.

No experimental work has been done on this subject yet because of lack of available time, but it would certainly be very interesting to see or listen to the result of some experiments along these lines!

References

1. M. Otala, Transient distortion in transistorized audio power amplifiers, IEEE Trans., vol.AU-18, 1970, p.234-9.

2. M. Otala, Circuit design modifications for minimizing transient intermodulation distortion in audio amplifiers, J. Audio Eng. Soc., vol.20, 1972, p.396-9.

3. W. M. Leach, Transient IM Distortion in power amplifiers, Audio, Feb.1975, p.34-41.

4. J. R. Stuart Approach to audio amplifier design. Wireless World. vol.79, 1973, pp.387-91,439-46,491-4.

5. H. Levitt et al., Perception of slope-overload distortion in delta-modulated speech signals, IEEE Trans. vol.AU-18. 1970, p.240-7.

6. J. Lohstroh and M. Otala, Audio power amplifier for ultimate quality requirements, IEEE Trans. vol. AU-21, 1973, p.545-51.

TRANSIENT INTERMODULATION DISTORTION

I would like to comment on the very informative article on transient intermodulation distortion by Bert Sundqvist published in your February 1977 issue:

He has shown by analysis that in order to prevent transient intermodulation distortion in an amplifier, the method proposed by Professor M. Otala (that of extended open-loop bandwidth in the power amplifier with subsequent passive band limiting in the preamplifier) need not be adhered to rigidly and the simpler method of band limiting the first stage of the amplifier achieves the same result. He suggested three methods for producing this band limiting: (1) input lag compensation, (2) use of a high-impedance current source as collector load, (3) operation of the first stage with a very low collector current. Of these, however. only the third seems to be new, as far as preventing transient intermodulation distortion is concerned.

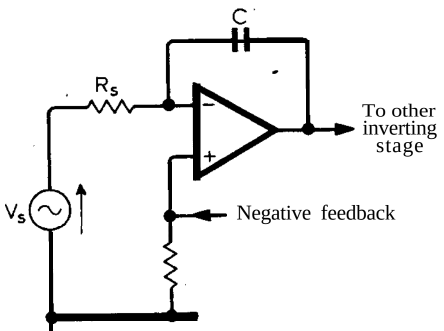


Fig. 1

To see why this is so, consider the frequency limiting mechanisms at work in the basic common emitter stage. There are mainly two. Firstly, the transfer mechanism, which is a physical motion of charge carriers, introduces dispersion and delay of the carriers and this results in the fall off of current gain (produces f_t). Secondly, existing between the various terminals of the transistor are frequency dependent impedances that are predominantly capacitive and these contribute to frequency limiting.

Considering Fig. 1, C represents the collector to base capacitance of the common emitter stage. Using Miller's theorem, this capacitance can be replaced by C_1 and C_2 as shown in Fig. 2, where A_v is the voltage gain between the inverting input and output

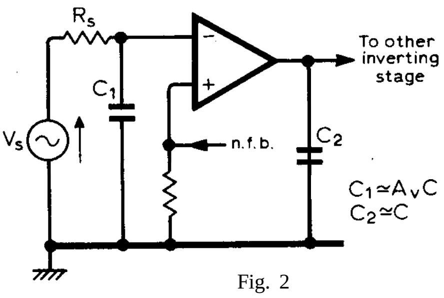


Fig. 2

(corresponding to the base and collector of the transistor). The time constant introduced by R_s and C_1 produces a dominant pole, and in general this is the mechanism that produces frequency roll-off in the common emitter stage. Input stage lag compensation increases C and a very high-impedance collector load increases A_v. Both result in a reduction of the bandwidth of the resulting input *RC* network. However, this RC network lies outside the loop of the feedback amplifier of which this stage is a part and indeed corresponds to the passive *RC* filter that Professor Otala recommends be placed before the input of power amplifiers in order to prevent the transmission of frequencies outside their open-loop bandwidths.

Thus, the only new technique which the results of Mr Sundqvist's analysis has uncovered is the reduction of the cut-off frequency of the input transistor which can be done by lowering the collector current, as he suggests. In fact, this method is more directly as a result of his analysis since the first pole within the loop encountered by an 'input signal is that due to fall-off of current gain of the input transistor.

Stephan Gift,
University of the West Indies,
St Augustine, Trinidad.

Reference

1. M. Otala. J. Lohstroh. An Audio Power Amplifier for Ultimate Quality Requirements. IEEE Transactions on Audio. vol. AU-21, no. 6, December 1973.

Mr Sundqvist replies:

I would like to thank Mr Gift for his clear explanation of the input stage frequency roll-off mechanisms. When I wrote my article I had not yet considered the details of how the frequency roll-off should be effected in practice. However, I would suggest that any band limiting procedure that gives a high input capacitance should be avoided, as this could give trouble when using a pre-amplifier with high output impedance, especially in combination with long connecting cables.

I have two other comments on my article which could be of interest to the readers. Using the original Otala design method, one ends up with a power amplifier with very wide bandwidth. However, the total audio system bandwidth is still limited by the pre-amplifier roll-off at 20-30 kHz. Although I do not think that an excessively large bandwidth is always desirable, this has always seemed to me to be a waste of good design work. Using the method outlined in my article the system bandwidth can be made as large or small as desired, as no frequency limits are involved in the design.

I would also like to point out that there are other methods to avoid t.i.m. without using Professor Otala's design method. My article was written in January, 1976, and since then Malmqvist' has published an interesting analysis of why t.i.m. is not produced by the Xelex range of amplifiers in spite of their relatively heavy feedback.

Bert Sundqvist,
Umeo,
Sweden.

Reference

1. M. Malmqvist, "Transient distortion", Musiktidningen, vol. 4. no. 4, p.53, Aug. 1976 (in Swedish); presented at the 56th AES Convention. Paris, March 1977.