for so long – even the most complex_ music signal can be represented by a Fourier analysis.

This mathematical equation lists separately each frequency making up the signal, (together with its phase and amplitude). However, a Fourier analysis is only complete in the case of simple waveforms, with more complex waveforms it becomes only a convenient approximation.

To make a Fourier analysis of a signal the components of that signal have to be analysed over a period of time such that complete cycles of the lowest frequency can occur. Thus we take consideration of the time domain.

Where steady-state signals are concerned, the time domain is not normally considered, as the signal is of a continuous unchanging nature between any two periods. If the "time window", during which the signal is Fourier analysed, is reduced progressively it becomes apparent that an accurate spectral analysis becomes less possible. It can then be seen that the important characteristics of the signal are amplitude and rate of change. In other words its envelope.

WHAT DO WE WANT

What is required is the amplification of an audio waveform in such a way that the ear can detect no degradation.

Let us consider ways in which such degradation can occur. The waveform envelope can be distorted by amplitude changes of any component or by changes in the phase relationship of the component harmonics.

Experimental work has established that changes in the relative amplitudes of the harmonic structure of the waveform are readily detectable.

Other work has shown that the qualitative. characteristics of a complex sound depend upon the phase relationships of the component harmonics. It would seem that as a phase difference must be interpreted as a time delay between the component parts of the signal, then a sufficient phase shift in a system must eventually become audible as these component parts are moved in respect to each other in time. In practice large phase shifts are very audible and indeed telephone lines are often phase and delay corrected to render speech intelligible. However, establishing an acceptable degree of phase shift is extremely difficult.

Following the arrival of "linear phase" loudspeakers great controversy has raged over whether phase shifts affect sound quality. A study of the experimental work performed to date shows that

1. It seems to be very difficult to replicate someone else's experiment.
2. It seems, on balance, that where recurrent waveforms (steady state) such as sine-waves (and instruments producing a "continuous" although decaying tone) are concerned; then quite large phase shifts, between the extremes of the frequency band, have no identifiable effect on sound quality. However, a phase non-linearity on the leading edge of a true transient appears to be audibly more perceptible, Particularly on speech and percussive sounds.

BANDWIDTH AND TID

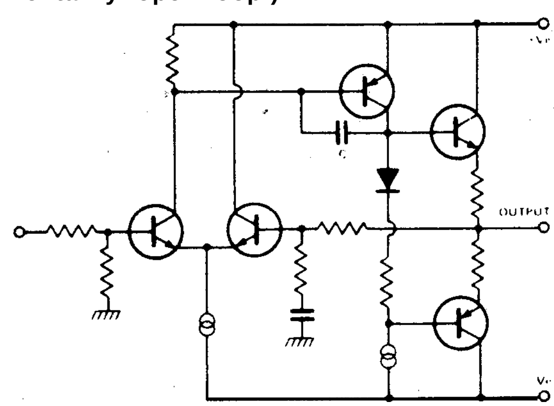
Transient signals cause many problems of which phase linearity is but one. Other problems include; instability and ringing, clipping,' slew-rate limiting, and transient intermodulation distortion.

Transient intermodulation distortion (TID or TIM) is much in vogue but is often misunderstood. TID most commonly occurs when an amplifier, with overall negative feedback over several stages, is driven by a large enough signal whose frequency (or equivalent rise time) is above the open loop bandwidth of that amplifier.

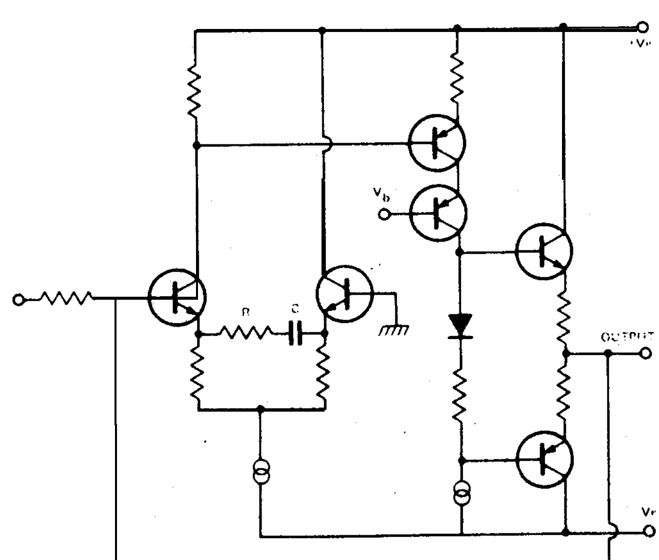
Because the feedback loop is fed from the output of the amplifier, there is no effective feedback until signal current flows at the output, i.e. during the open-loop rise time of the amplifier.

Very large signals occurring in the intermediate stages of the amplifier cause those stages to distort or even to clip. With some amplifiers this clipping can cause the stage to latch-up for a time until the operating conditions restabilise. Thus not only is the leading edge of the signal severely distorted – in some cases it is removed completely.

TID is therefore a form of overloading that is dependent upon both amplitude and time. It is audibly (but at a higher signal level) similar to cross-over distortion, as both effects cause phase and amplitude modulation of the signal due to momentary change in gain. (Remember that at the cross-over point zero, there is no current flow in the output stage and hence no feedback current and so the amplifier is momentarily open-loop.)



Circuit diagram of a typical amplifier circuit which employs lag compensation techniques – provided by C.



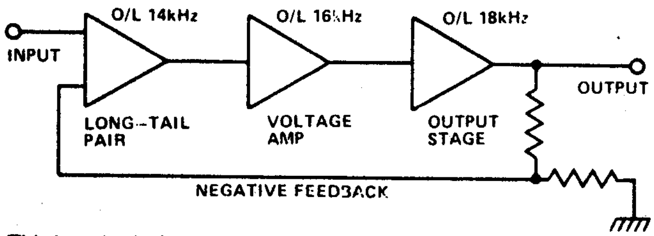
Lead compensation: components R end C provide the time constant.

MAKING BIG BANDS

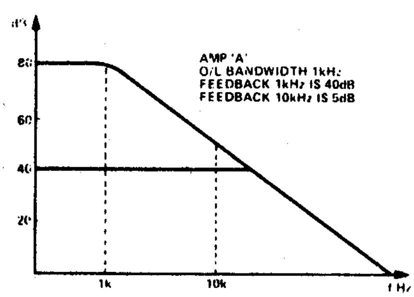
TID can be avoided by designing an amplifier whose open-loop bandwidth is greater than the highest frequency of the input signal. The maximum bandwidth can then be defined at the input by a passive RC filter. Thus if we decide upon a maximum signal bandwidth of 20 kHz than our filter will limit the signal waveform rise-time to T = 0.35.

$$T = \frac{0.35}{20\ \text{kHz}}$$

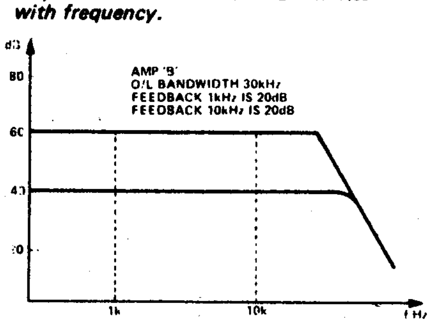
i.e. $17.5\ \mu s$.



Third method of avoiding TID. Each stage in the design has a wider bandwidth than the preceding one.



This amplifier design has a limited open loop bandwidth and the THD will rise with frequency.



Contrast this with the graph above. The bandwidth here is much wider, resulting in a more linear THD response.

Our amplifier's open-loop bandwidth should be designed to be, say, 23 kHz, giving it an open-loop rise-time of 15 μs and freedom from TID. If however, in the interests of a good specification, and possibly better reproduction, we decide upon a close-loop bandwidth of 100 kHz (i.e. a rise time of 3.5 μs) then our amplifier will need an open-loop bandwidth of greater than 100 kHz to maintain freedom from TID. In a power amplifier such performance is not easy to obtain. Fast power transistors are notoriously easy to blow-up and are expensive. The common form of lag compensation (used where the open-loop bandwidth is restricted) has to be replaced by lead compensation:—

Another technique is an extension of the first in that the preceeding stage of the power-amplifier is designed to have a lower open-loop band width than the next.

IMPORTANT OR NOT

Many people now consider that TID is unimportant or even that it doesn't exist. This is partly because it is very difficult to measure and only readily visible (in the laboratory) in the "clipping" state. To reach this stage with most amplifiers (but not TID — free designs) there is a requirement for either fast rise-time or higher signal levels or both, — conditions that are unlikely to occur in practice. However, a large degree of non-linearity and hence bad intermodulation will still occur with more realisable input signals. Although this cannot be measured yet (how do you measure say, 5% IM over a period of 5 milliseconds?) it can be predicted mathematically and, just as important, heard. Amplifiers free of TID have a very "open" quality with accuracy of depth.

An amplifier designed with a wide open-loop bandwidth, for low TID, often has other more tangible benefits. The high frequency THD is usually no higher than at the mid-point; in stark contrast to more traditional designs. This is because gain is still available at high frequencies for negative feedback. Such amplifiers also usually have much higher slew-rate.

SLEW

Slew-rate defines the speed with which the amplifier can deliver output voltage to the load. For example, if an amplifier has a maximum output of 100 volts p/p and a rise-time of 100 μs, then the amplifier, if it were perfect, should have an output of about 80 volts after 10 μs in response to a suitable square wave input. In other words the output voltage would have risen at the rate of 8 V/μs. However, amplifiers do not generally respond to large changes as fast as their small signal characteristics predict, for circuit and transistor capacitances can be charged only as fast as their driving circuits allow.

In its simplest form the slew-rate of an amplifier defines how fast the output voltage can change for large signal conditions, and it is normally quoted in volts per micro second. The maximum slew-rate of an amplifier is usually limited by the slowest stage in its circuit.

That stage will have an operating current T (as set in the design) and a capacitance C (usually a frequency compensation capacitor)

$$\text{Slew-Rate} = \frac{T}{C}$$

Thus if a transistor stage has a standing current of 100 μA and is compensated by a 43 pF capacitor then its slew-rate will be

$$\frac{100}{33}$$

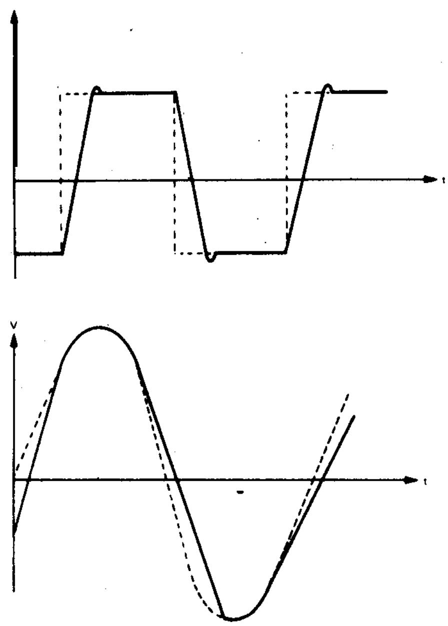
i.e. 3 V/μs

Depending upon the design some circuits have a different slew-rate depending upon whether their output is negative-going or positive-going. Slew limiting also defines the full-power bandwidth; a figure more commonly quoted by manufacturers.

$$fp = \frac{SR\ (10^6)}{2\pi\ E\ op}$$

E op = peak output swing in volts
fp = Full power bandwidth in hertz.

Thus in a 100 watt (into 8 ohms) amplifier having full-power bandwidth of 20 kHz the required minimum slew-



The effects of slew-rate on a signal passing through an amplifier prone to this fault. Top: a squarewave, note the slight overshoot. Below that, a sineweve. In both cases the dotted line represents the *input.*

rate would be about 5 V/µs. This is, however, the absolute minimum figure and experience suggests that such an amplifier would have a hard, gritty high-frequency sound. Such an amplifier should have a slew-rate greater than 20 V/µs to be certain of avoiding the increase in distortion caused by the gradual onset of slew-limiting.

Unfortunately the higher the power output of the amplifier the greater the required slew-rate as more volts swing at the output in the same period of time and so as our 100 W amp needs 20 V/µs an otherwise identical 50 W amp needs 14 V/µs and a 20 W amp needs only 9 V/µs. But these forms of distortion tend to give subtle audible effects compared to the most common amplifier problem – that of clipping.

CLIPPING

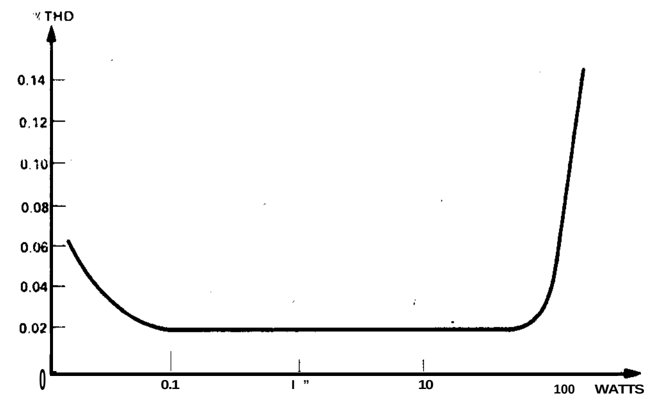
Clipping occurs when an amplifier is overloaded by high level signal peaks. Such peaks occur frequently in much music material and so the manner in which the amplifier clips determines its audibility. A soft, clipping effect where the distortion rises gradually (typical of valve amplifier circuits) is audibly preferable to the hard clipping typical of transistor circuits.

Worse still, some amplifiers tend to suffer saturation effects on clipping and take a time to recover; thus artificially extending the length of tirna the signal is clipped. The use of overall negative feedback to reduce distortion unfortunately makes things worse. Overall feedback effectively linearises the clipping – the distortion changes from 0.01% (say) to 10% and quite suddenly too.

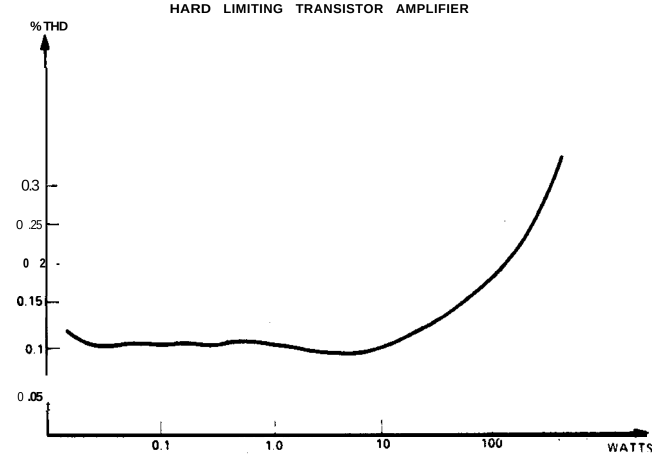
DESIGN PROCEDURE

We have covered just ‘a few of the requirements a designer must consider when working upon the design of power-amplifiers. There are many more to be considered to even rough out a design specification before the circuit hardware is considered. The following sequence is mandatory:

1. What parameters are important to prevent audible degradation of the signal?
2. Detail a performance specification that meets the requirementsof (1).
3. Decide upon the circuit technology necessary; Bipolar; MDSFET; ‘Tube; Class A; Class B; Switching; etc; etc.
4. Undertake a development programme to produce a prototype.



HARD LIMITING TRANSISTOR AMPLIFIER



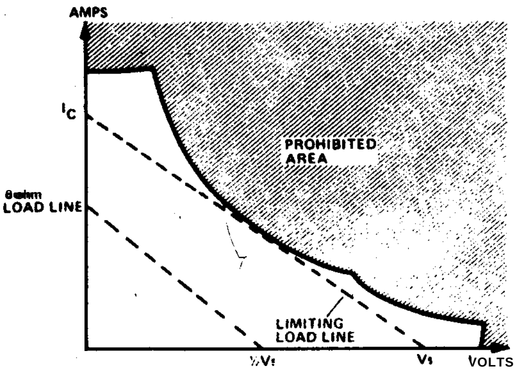
SOFT LIMITING VALVE AMPLIFIER

A comparison of the limiting characteristics – in general – of both transistor and valve amplifier types. There is a body of opinion which holds these curves to be the whole truth as to why valve amplifiers are preferred by many musicians.

At this point the designer has to accept that it’s a real world and that his performance specification cannot be achieved in a way that is acceptable to accountants, salesmen, customers, customer’s wives or whoever else is around. Tradeoffs are necessary and much of the “art” is in deciding which defects and degradations are more acceptable than others.

As an illustration of the changes in design approach over the years we will briefly illustrate three designs for which the author has been responsible:

1. Cambridge Audio P60 (P80)
2. Lecson AP3 Mk II
3. Mission Electronics Voltage Amplifier



Illustrating the load line conditions for output stages

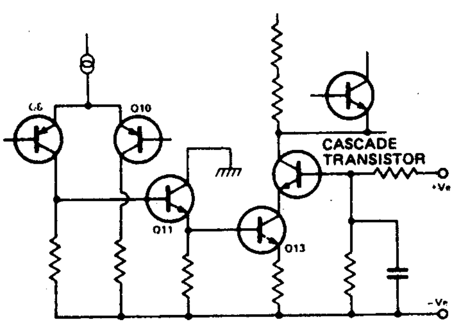
The P60 is capable of good mid-band performance (THD 0.01% at 1 kHz is 30 W) but its high frequency distortion is poor because of the limited open-loop bandwidth. Generally this amplifier performs well at low and moderate levels but at high levels its sound quality becomes hard and aggressive. Some improvements to this circuit can be quite simply made as follows:

1. A resistor is inserted between. Q10 collector and the negative rail to give better balance between Q8 and Q10.
2. A cascade transistor is fitted to Q13 collector to reduce "early effect" distortion due to the collector-base capacitance of Q13.
3. An emitter resistor is fitted to Q13 to provide local negative feedback.

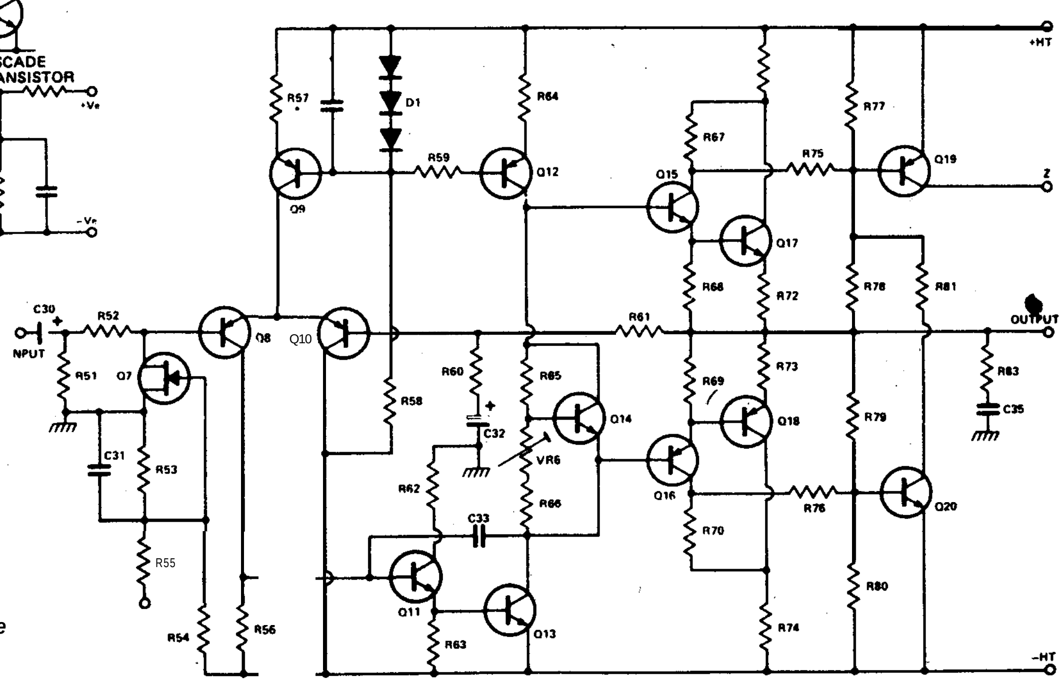
The Lecson AP3 Mk II incorporates much of the thinking in this article and is representative of the latest types of high performance amplifiers. It is a directly-coupled Class 8 design using a fully complementary output stage of series connected transistors and gives a power output of around 150 watts per channel.

The New Mission Voltage Amplifier represents an attempt to produce an amplifier that performs well irrespective of load. The circuits cannot be described at this stage as they are the subject of patent applications. However, a brief description will illustrate the philosophy behind the design.

The casing contains two completely separate mono amplifiers, each with its own power supply. A separate module carries the dc-voltage offset protection circuits; the delay switched-on circuits; and the thermal protection



Showing how some of the improvements mentioned can be added to the P60 basic design.



Full circuit diagram of the Cambridge P60 power amplifier design.

HOW IT' WORKS–Cambridge P60

The P60 power amplifier is of a conventional design but with care being taken to optimise each stage. Q8 and Q10 form a long-tailed pair with Q9 as their emitter current source. Q8 and Q10 must be very closely matched for minimum DC offset and for maximum common-mode rejection to avoid H. T. ripple appearing at the output. The next stage is the Q13 voltage amplifier which is loaded by a current source (Q12) instead of the more common "bootstrapped" resistors. Note that Q13 is buffered from the long-tail pair by an emitter follower (Q11) to prevent any loading of that stage worsening the distortion characteristics.

Capacitor C33 gives lag compensation which defines the dominant pole of the amplifiers. The open-loop bandwidth is quite high (for this type of circuit) at 12 kHz but none the less this amplifier is prone to TID effects. The protection circuit is very unusual in that the output is limited by an FET (Q7). QI9 and Q20 each ferm conventional V-I summing circuits which monitor the loading on the output stage.

If either Q19 or Q20 turns-on, the gate of the FET Q7 (normally biased-off by R54 to the negative HT) is biased positive and it starts to turn-on. It then acts as a potential divider with R52 and thus attenuates the audio signal. This protection only turns on at the equivalent of 50 W into 2 Ohms load and when it turns on it only adds moderate distortion (0.2% typically) as distinct from clipping.

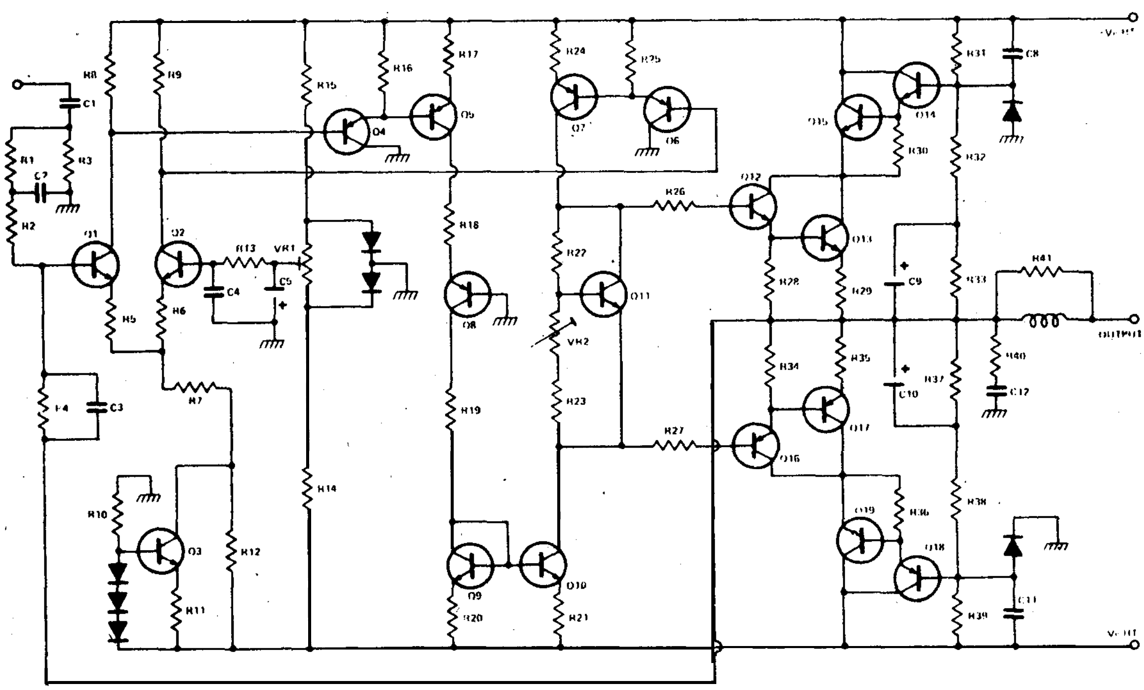
circuits. Particular attention has been paid in the design to achieving:

1. Low distortion with a very low order of overall feedback
2. Wide open-loop bandwidth with an excellent slewing rate
3. Minimum time and phase distortion
4. A high transient power capability with virtual freedom from clipping effects.

The output stages have a very high current capability but have no protection circuits, the output transistors being designed to sink the full energy of the power-supply into the load. A patented form of voltage feed to this stage gives the amplifier a short term power delivery capability of about 600 watts (compared to the rated 150 watts 8 ohms). This represents a 8 db increase in power availability over the rated figure. The voltage amplifing stages are designed to clip softly and this combined with the low-overall feedback gives overload characteristics similar to those of an equivalent tube amplifier.

CONCLUSION

This feature has discussed just some aspects of modern audio amplifier design. At present much attention is still given to whether an amplifiir is designed around bipolar transistors, FETs, valves, or switching transistors. However designers are beginning to appreciate that the major stumbling block is not designing a circuit using any of these technologies but in deciding upon what is the performance specification required that *will give faithful reproduction of the sound source.* Until this problem is solved there will continue to be an element of uncertainty in amplifier design.



Full circuit diagram for the Lecson AP3 power amplifier design, producing wound 150W.

HOW IT WORKS-Lecson AP3

Transistors Q1 and Q2 form a long-tailed pair differential amplifier with Q3 as the emitter current source. Local feedback is applied in the form of emitter resistors R5 and R6. The base of Q2. instead of being grounded, is connected to a potential divider RVI which permits the DC offset at the output to be set to zero. The input signal to Q1 is passed through a low-pass filter (R1, C2) which sets the bandwidth to 22 kHz (i.e. below the open loop bandwidth for no TID effects). The bi-phase outputs of the long-tail pair feed a second differential amplifier Q5 and Q7. Transistor Q5 has a constant current load (Q8) whilst is terminated by a current mirror (Q9 and Q10). Transistor Q10 will always deliver the same current as transistor Q9 hence the term "Current Mirror" and the excellent symmetry and balance this stage achieves. Functionally, however, Q10 can be considered as an active load whilst Q7 is a voltage amplifier from whose collector the drive to the output stage is taken. Note that Q5 and Q7 both have local emitter feedback (R17, R24) and that both are buffered from the long-tail pair (Q4 and Q6 emitter followers).

Transistors Q12, Q13, Q16 and Q17 each form conventional Darlington emitter follower stages. Each stage is series connected to a further power transistor (Q14, Q15 and Q18, Q19 respectively) which is permanently biased ON. Their emitter potentials are determined by the ratio of the base potential dividers. This ratio was chosen such that Q13 and Q15 each has half the supply rail across them.

The whole amplifier is in the inverting mode with overall shunt feedback through R4 and C3.

This amplifier is quite fast having an open-loop bandwidth of about 27 kHz. The circuit is stable without the usual compensation capacitors within the loop. THD is low being typically (at 100 W into 8 Ohms) 0.004% at 1 kHz and 0.02% at 10 kHz. The HF distortion can be further improved by selection of transistor Q7 foi a device with a low collector-base capacitance.

No conventional protection circuits are used as extremely high power transistors are fitted and these can survive a short-circuit condition in the time taken for the power supply to shut down.

Audio amplifier design

I have read the further letter from Mr Stuart of Lecson Audio in your August issue, and I am somewhat surprised that so much of conjecture or personal opinion should be stated by Mr Stuart as matters of established fact.

However, to take the main points on which Mr Stuart has thrown down the gauntlet:

1. Transient intennodulation distortion

It has been known by audio amplifier designers for very many years that unsatisfactory results were frequently obtained if the input bandwidth to the power amplifier was excessive, and more recently this has focused attention on the manner in which the amplifier responds to a transient input. A part of this problem was formerly analysed by Mr Otala in his paper presented in 1970, and he coined the above term for this problem.

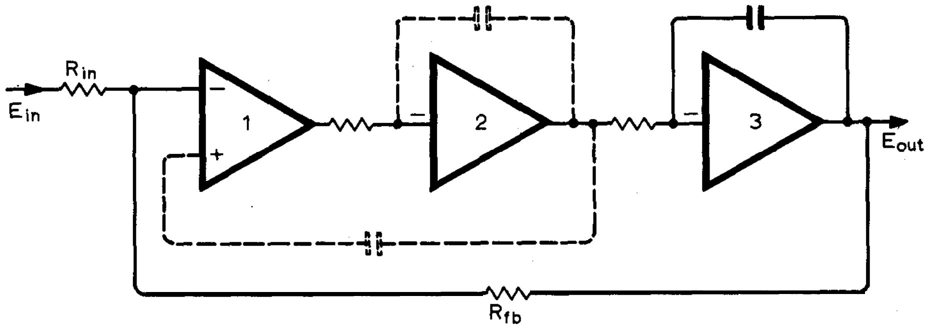
I will try to summarize Mr Otala's argument with reference to the accompanying diagram.

If we take a feedback amplifier which consists of a chain of separate op-amp elements (for convenience I have shown three) of which, say, amplifier 3 is arranged 'to have the slowest response, and we apply a step-function input of sufficiently rapid rise time and sufficient magnitude, amplifier 1 will overload and produce severe intermodulation distortion in the transmitted signal, even when the magnitude of the input signal is within the normal input range of the amplifier, because the propagation delay of the main feedback loop is too long for the feedback to be effective in diminishing the magnitude of the applied signal — as seen by the first amplifier stage — during the transient condition.

This is a matter of practical concern in the design of transistor audio amplifiers, since the output transistors, represented by stage 3, will generally have a more sluggish response than the small signal voltage amplifier stages 1 and 2. Moreover, designers sometimes exacerbate this problem by choosing to stabilize the feedback loop by connecting a capacitor between the collector and base of the second voltage amplifier transistor, represented in my diagram by op-amp 2.

There are (at least) three solutions to this problem.

(a) To ensure that the propagation delay



R_{in}
E_{in}
1
2
3
E_{out}
R_{fb}

through the whole system is very low, in comparison with the signal bandwidth. This is the approach favoured by Mr Otala, but is expensive.

(b) To apply the loop **stabilizing capacitor** across stages 1 and 2, thereby ensuring that stage 1 does not run out of negative feedback under transient conditions. (This was what I referred to as a divided feedback loop.)

(c) To interpose an input bandwidth limiting circuit, which could well be in the pre-amplifier, to ensure that the rise time of the input signal does not exceed the handling capability of the amplifier.

I favour a combination of the last two methods and, in particular, I believe that the dissociation of the h.f. feedback capacitor from the amplifier output point helps to ensure that the amplifier is tolerant of unexpected reactive loads, even though one pays the price of a t.h.d. curve which worsens somewhat at the h.f. end. Since I also believe that it is better to put predictable tonal quality with unknown loudspeakers before refinement of paper spec. I remain content with this decision.

2. ***Noise output from "virtual earth" amplifiers.***

Mr Stuart asks again how it should arise, if the "virtual earth" point appears to have a low noise impedance, that the noise output should rise when this is shunted by **47kΩ.** Since Mr Stuart must know very well that the gain of a shunt feedback amplifier is determined by the ratio of the input and feedback limb impedances, and that, other things being equal the higher the gain the larger the noise output voltage, I had treated this as a rhetorical question.

Obviously, the "virtual earth" impedance is a notional thing which arises because of the feedback connection. Equally obviously, the gain of the system increases when the input limb is reduced from an impedance of infinity (o/c) to any lower value one chooses.

3. ***Effective bandwidth of R.I.A.A. equalized stages.***

Mr Stuart says "of course the noise is calculated in a 20kHz bandwidth". May I suggest a simple experiment. If one takes a wideband feedback amplifier, of any input resistance value one likes, and measures the output noise with a 20kHz bandwidth, one will get **a** value which is in reasonable experimental agreement with the value predicted from

this bandwidth, the known gain of the circuit, and the thermal noise of the input elements. If now one connects across the feedback circuit the components necessary to provide the falling h.f. gain characteristic of the R.I.A.A. curve, the output noise will drop, simply because one has restricted the effective noise bandwidth. If this were not the case, the series feedback connection using an inductive input would be worse than it is.

J. L. Linsley Hood,
Taunton, Somerset.