

Trimodal Audio power



part 2

The same components that dominate amplifier noise performance also determine the output dc offset; if R_9 is reduced to minimise the source resistance seen by Tr_3 , then the value of R_8 is scaled to preserve the same closed-loop gain, and this reduces the voltage drops caused by input transistor base currents.

My previous amplifier designs assumed that a $\pm 50\text{mV}$ output dc offset is acceptable. This allowed dc trimming, offset servos, etc to be gratefully dispensed with. However, it is not in my nature to leave well enough alone, and it could be argued that $\pm 50\text{mV}$ is on the high side for a top-flight amplifier. For this reason, I have reduced this range as much as possible without resorting to a servo; the required changes were already made when impedance of the feedback network was reduced to minimise Johnson noise. There were details on this in last month's issue.

With the usual range of component values, the dc offset is determined not so much by input transistor V_{be} mismatch, which tends to be only 5mV or so, but more by a second mechanism—imbalance in beta. This causes imbalance of base currents, I_b , drawn through input bias resistor R_1 and feedback resistor R_8 . Cancellation of the voltage-drops across these components is therefore compromised.

A third source of dc offset is non-ideal matching of input degeneration resistors $R_{2,3}$. Here they are 100Ω , with 300mV dropped across each, so two 1% components at opposite ends of their tolerance bands could give a maximum offset of 6mV . In practice, it is unlikely that the error from this source will exceed 2mV .

There are several ways to reduce dc offset. Firstly, a Class-A amplifier with a single output pair must be run from modest ht rails, so the requirement for high- V_{ce} input transistors is relaxed. This allows higher beta devices to be used, directly reducing I_b . The 2SA970 devices used in this design have a beta range of 350 to 700, compared with 100 or less for MP5A06/56. Note the pinout is *not* the same.

In the first half of this article, we reduced the impedance of

the feedback network by a factor of 4.5, and the offset component due to I_b imbalance is reduced by the same ratio. We might therefore hope to keep the dc output offset for the improved amplifier to within $\pm 15\text{mV}$ without trimming or servos. Using high-beta input devices, the I_b errors did not exceed $\pm 15\text{mV}$ for ten sample pairs — *not* all from the same batch — and only three pairs exceeded $\pm 10\text{mV}$. Errors in I_b are now reduced to the same order of magnitude as V_{be} mismatches, and so no great improvement can be expected from further reduction of circuit resistances. Drift over time was measured at less than 1mV , and this seems to be entirely a function of temperature equality in the input pair.

Figure 1 shows the ideal dc conditions in a perfectly-balanced input stage, assuming a β of 400, compared with a set of real voltages and currents from the prototype amplifier. In the latter case, there is a typical partial cancellation of offsets from the three different mechanisms, resulting in a creditable output offset of -2.6mV .

Biasing for three modes

Figure 2 shows a simplified rendering of the Trimodal biasing system; the full version appears in Fig. 3. The voltage between points A and B is determined by one of two controller systems, only one of which can be in command at a time. Since both are basically shunt voltage regulators sitting between A and B, the result is that the lowest voltage wins. The novel Class-A current-controller introduced in the original article¹ is used here adapted for 0.1Ω emitter resistors, mainly by reducing the reference voltage to 300mV , which gives a quiescent current (I_q) of 1.5A when established across the total emitter resistance of 0.2Ω .

In parallel with the current-controller is the V_{be} -multiplier Tr_{13} . In Class-B mode, the current-controller is disabled, and critical biasing for minimal crossover distortion is provided in the usual way by adjusting preset Pr_1 to set the voltage across Tr_{13} . In Class-A/AB mode, the voltage Tr_{13} attempts to establish is increased (by shorting out Pr_1) to a value greater than that required for Class-A. The current-controller therefore takes charge of the voltage between X and Y, and unless it fails Tr_{13} does not conduct. Points A B X Y are the same circuit nodes as in reference 1.

Class A/AB mode

In Class-A/AB mode, the current-controller, comprising $Tr_{14,15,16}$ in Fig. 2, is active and Tr_{13} is off, as Tr_{20} has short-

Douglas Self has already shown how the low-impedance negative feedback network integrated into his trimodal amp reduces noise. Here he demonstrates how low-Z nfb improves output dc offset performance too. He also details the amplifier's mode-switching and bias control systems and looks at overall performance.

ed out Pr_1 . Transistors $Tr_{15,16}$ form a simple differential amplifier that compares the reference voltage across R_{31} with the V_{bias} voltage across output emitter resistors R_{16} and R_{17} ; as explained in reference 1, for Class-A this voltage remains constant despite delivery of current into the load. If the voltage across $R_{16,17}$ tends to rise, then Tr_{16} conducts more, turning Tr_{14} more on and reducing the voltage between A and B. $Tr_{14,15,16}$ all move up and down with the amplifier output, and so a tail current-source Tr_{17} is used.

I am aware that the current-controller is more complex than the simple V_{be} -multiplier used in most Class-B designs. There is an obvious risk that an assembly error could cause a massive current that would prompt the output devices to lay down their lives to save the rail fuses. The tail-source Tr_{17} is particularly vulnerable because any fault that extinguishes the tail current removes the drive to Tr_{14} , the controller is disabled, and the current in the output stage will be very large. In Fig. 2 the V_{be} -multiplier Tr_{13} acts as a safety-circuit which limits V_{bias} to about 600mV rather than the normal 300mV, even if the current-controller is completely non-functional and Tr_{14} fully off. This gives a 'quiescent' of 3A, and I can testify this is a survivable experience for the output devices in the short-term; however they may eventually fail from overheating if the condition is allowed to persist.

There are important points about the current-controller. The entire tail-current for the error-amplifier, determined by Tr_{17} , is syphoned off from the voltage amplifier stage current source Tr_5 . This must be taken into account when ensuring that the upper output half gets enough drive current.

There must be enough tail current available to turn on Tr_{14} , remembering that most of Tr_{16} collector-current flows through R_{15} , to keep the pair roughly balanced. If you feel moved to alter the voltage-amplifier stage current, remember also that the base current for driver Tr_6 is higher in Class-A than Class-B, so the positive slew-rate is slightly reduced in going from Class-B to A.

I must admit that the details of the voltage reference were rather glossed over in reference 1, because space was running out fast. The original amplifier shown last month used a National LM385/1.2, its output voltage fixed at 1.223V nominal; this was reduced to approx 0.6V by a 1k Ω /1k Ω divider.

The circuit also worked well with V_{ref} provided by a silicon diode, 0.6V being an appropriate bias voltage drop across two 0.22 Ω output emitter resistors. This is simple, and retains the immunity of I_q to heatsink and output device temperatures, but it does sacrifice the total immunity to ambient temperature that a band-gap reference gives.

The LM385/1.2 is the lowest voltage band-gap reference commonly available; however, the voltages shown in Fig. 2 reveal a difficulty with the new lower V_{bias} value and the complementary feedback pair stage; points A & Y are now only 960mV apart, which does not give the reference room to work in if powered from node A, as in the original circuit.

The solution is to power the reference from the positive rail, via $R_{42,43}$. The midpoint of these two resistors is bootstrapped from the amplifier output rail by C_5 , keeping the voltage across R_{43} effectively constant. Alternatively, a current-source could be used, but this might reduce positive headroom. Since there is no longer a strict upper limit on the reference voltage, a more easily obtainable 2.56V device could be used providing R_{30} is suitably increased to 5k Ω to maintain V_{ref} at 300mV across R_{31} .

In practice, stability of I_q is very good, staying within 1% for long periods. The most obvious limitation on stability is differential heating of $Tr_{15,16}$ due to the main heatsink. Transistor Tr_{14} should also be sited with this in mind, as heating it will increase its beta and slightly imbalance $Tr_{15,16}$.

Class-B mode

In Class-B mode, the current-controller is disabled, by turn-

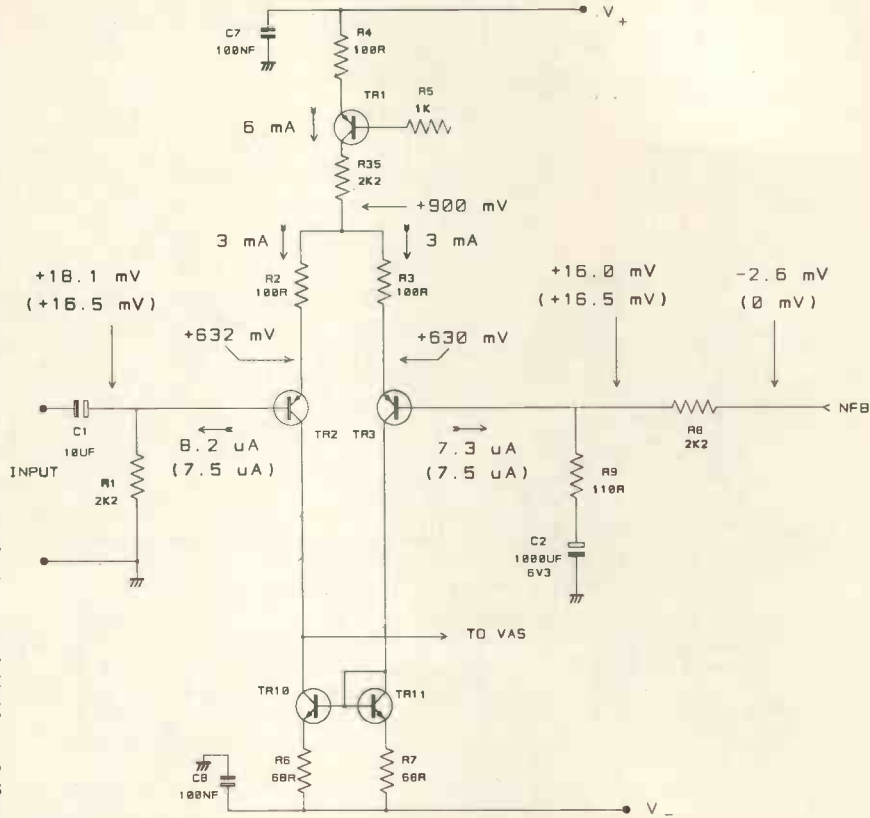


Fig. 1. A close look at input stage balance. Circuit conditions shown here are a real example. Ideal conditions for $\beta = 400$ are shown in brackets. All voltages measured to ground.

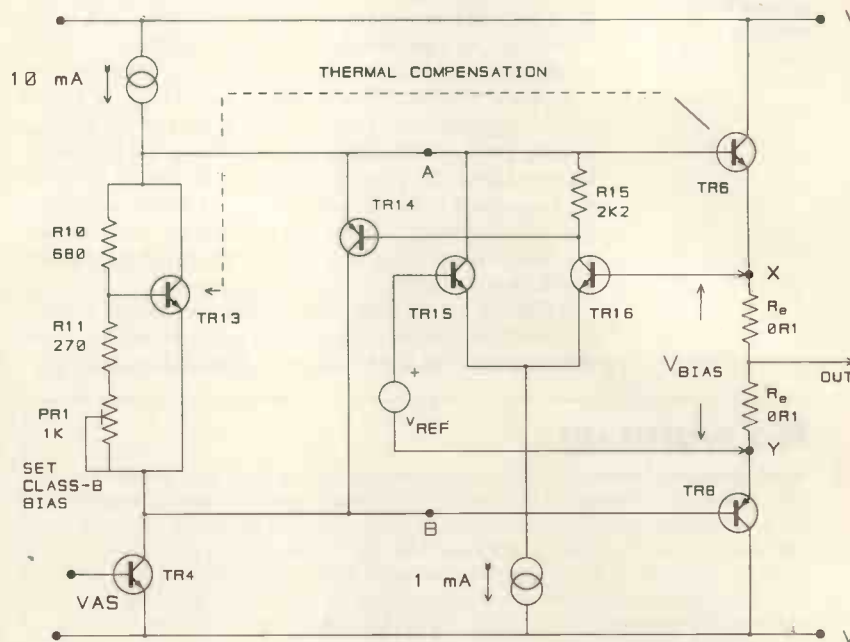


Fig. 2. Simplified current-controller in action, showing typical dc voltages in class-A. Points A, B, X and Y are the same as in the original class-A article. The grey panel on the left is the V_{be} multiplier, Class-B biasing and Class-A safety circuit. Panel in the middle is the Class-A current regulator. Voltage over points A,B is 1.5V while over X,Y, i.e. V_{bias} , there is 300mV.

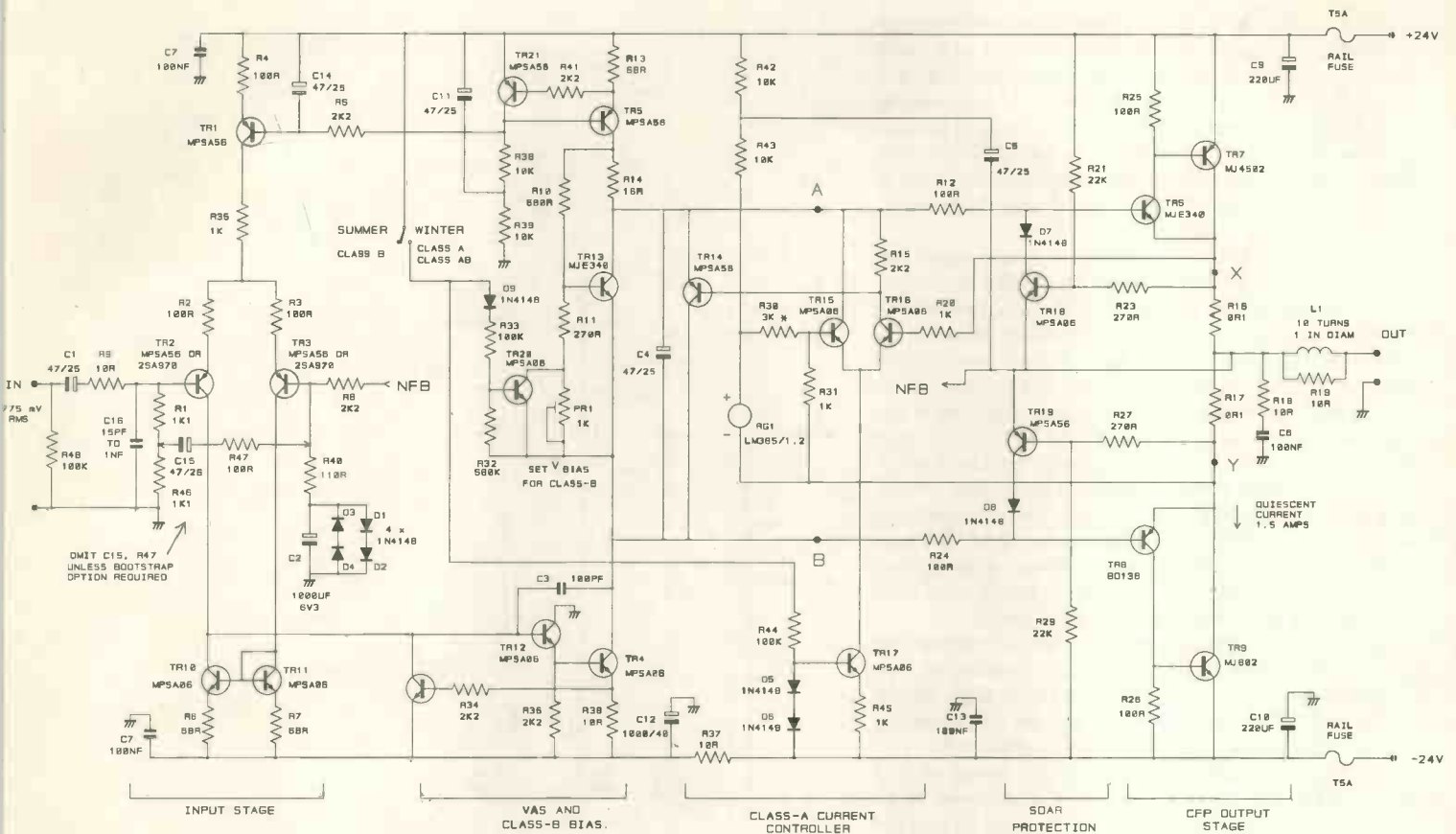


Fig. 3. Complete circuit diagram of class-A amplifier, including the optional bootstrapping components, R47 and C15.

ing off tail-source Tr_{17} so Tr_{14} is firmly off, and critical biasing for minimal crossover distortion is provided as usual by V_{be} -multiplier Tr_{13} . With 0.1Ω emitter resistors V_{bias} (between X and Y) is approx 10 mV. I would emphasise that in Class-B this design, if constructed correctly, will be as 'blameless' as a purpose-built Class-B amplifier. No compromises have been made in adding the mode-switching.

As in the previous Class-B design, the addition of R_{14} to the V_{be} -multiplier compensates against drift of the voltage amplifier stage current-source Tr_5 . To make an old but much-neglected point, the preset potentiometer should always be in the bottom arm of the V_{be} divider $R_{10,11}$, because when presets fail it is usually by the wiper going open; in the bottom arm this gives minimum bias voltage, but in the upper arm it would give maximum.

In Class-B, temperature compensation for changes in driver dissipation remains vital. Thermal runaway with the complementary feedback pair is most unlikely, but accurate qui-

escent setting is the only way to minimise cross-over distortion. Tr_{13} is therefore mounted on the same small heatsink as driver Tr_6 . This is often called thermal feedback, but it is no such thing as Tr_{13} in no way controls the temperature of Tr_6 ; 'thermal feedforward' would be a more accurate term.

Switching modes

The dual nature of the biasing system means Class-A/Class-B switching is easily implemented, as in Fig. 3. A Class-A amplifier is an uneasy companion in hot weather, and so I was unable to resist the temptation to sub-title the mode switch 'Summer/Winter', by analogy with a car air intake.

Switchover is dc-controlled, as it is not desirable to have more signal than necessary running around inside the box, possibly compromising interchannel crosstalk. In Class-A/AB mode, S_1 is closed, so Tr_{17} is biased normally by $D_{5,6}$, and Tr_{20} is held on via R_{33} , shorting out preset Pr_1 and setting Tr_{13} to safety mode, maintaining a maximum V_{bias} limit of 600mV. For Class-B, S_1 is opened, turning off Tr_{17} and therefore $Tr_{15,16}$ and Tr_{14} . Transistor Tr_{20} also ceases to conduct, protected against reverse-bias by D_9 , and reduces the voltage set by Tr_{13} to a suitable level for Class-B. The two control pins of a stereo amplifier can be connected together, and the switching performed with a single-pole switch, without interaction or increased crosstalk.

Mode-switching affects the current flowing in the output devices, but the output voltage is controlled by the global feedback loop, and switching is completely silent in operation. The mode is switchable while the amplifier is handling audio, allowing some interesting 'A/B' listening tests.

It may be questioned why it is necessary to explicitly disable the current-controller in Class-B; Tr_{13} is establishing a lower voltage than the current-controller which latter subsystem will therefore turn Tr_{14} off as it strives futilely to increase V_{bias} . This is true for 8Ω loads, but 4Ω impedances increase the currents flowing in $R_{16,17}$ so they are transient-

No warm up

Audio magazines often state that semiconductor amplifiers sound better after hours of warm-up. If this is true – in most cases it almost certainly isn't – the admission represents truly spectacular design incompetence. Accusations of this type are applied with particular venom to class-A designs, because it is obvious that the large heat sinks required take time to reach final temperature. So it is important to record that in class-A operation this design stabilises its electrical operating conditions in less than a second, giving the full intended performance.

No "warm-up time" beyond this is required.

Obviously the heat sinks take time to reach thermal equilibrium. But as already described, measures have been taken to ensure that component temperature has no significant effect on operating conditions or performance.

Supplying power

Regulated supplies are quite unnecessary, and are virtually certain to do more harm than a good unregulated power supply (Fig. 4).

The supply must be designed for continuous operation at maximum current, so the bridge rectifier should be properly heat-sunk, and careful

consideration given to the ripple-current ratings of the reservoirs. This is one reason why reservoir capacitance has been doubled to 20,000µF per rail: the ripple voltage is halved, improving voltage efficiency as it is the ripple troughs that determine clipping onset. But the ripple current, although unchanged in total value, is now split between two components. (The

capacitance was *not* increased to reduce ripple injection. This is dealt with far more efficiently and economically by making amplifier psrr high³.)

Do not omit the secondary fuses. Even in these modern times rectifiers do fail, and transformers are horribly expensive...

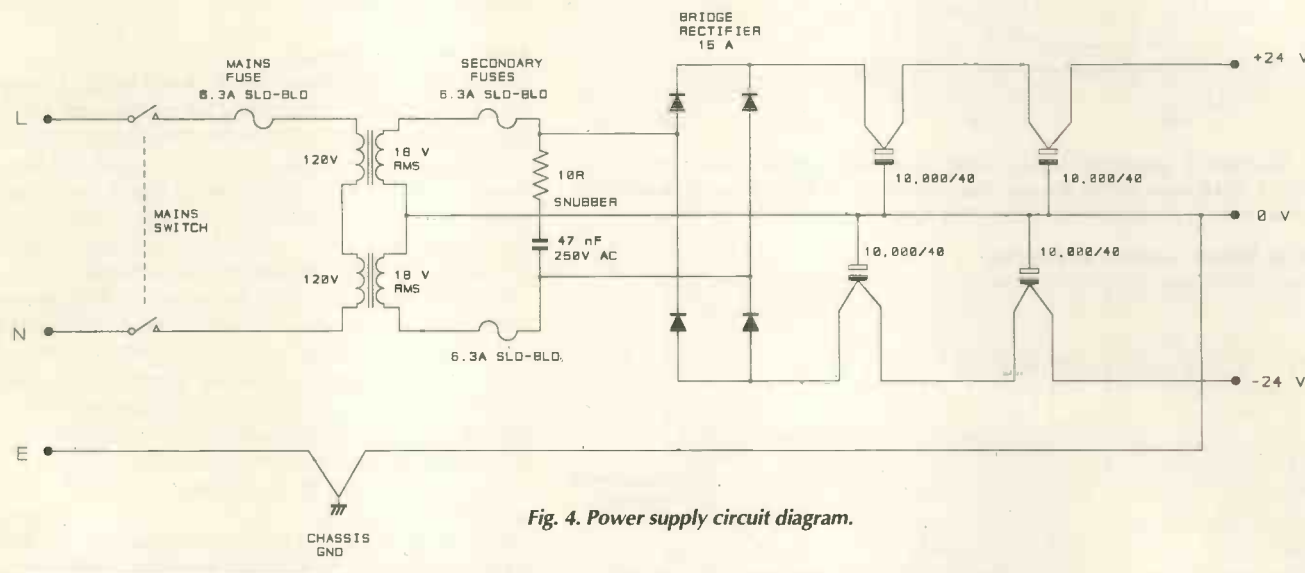


Fig. 4. Power supply circuit diagram.

ly greater than the Class-A I_q , and the controller will therefore intermittently take control in an attempt to reduce the average current to 1.5A. Disabling the controller by turning off T_{r17} via R_{44} prevents this.

Test mode

If the Class-A controller is enabled, but preset Pr_1 is left in circuit, (eg by shorting T_{r20} base-emitter) we have a test mode which allows suitably cautious testing; current I_q is zero with the preset fully down, as T_{r13} over-rides the current-controller, but increases steadily as Pr_1 is advanced, until it suddenly locks at the desired quiescent current. If the current-controller is faulty then I_q continues to increase to the defined maximum of 3A.

Thermal design

Class-A amplifiers are hot almost by definition, and careful thermal design is needed if they are to be reliable, and not take the varnish off the Sheraton. Since the internal dissipation of the amplifier is maximal with no signal, simply turning on the prototype and leaving it to idle for several hours will give an excellent idea of worst-case component temperatures. In Class-B the power dissipation is very programme-dependant, and estimates of actual device temperatures in realistic use are notoriously difficult.

Table 1 shows the output power available in the various modes, with typical transformer regulation, etc; the output mode diagram in Part 1, Fig. 1, showed exactly how the

Table 1. Power capability of the trimodal power amplifier.

Load resistance	W	W	W	Distortion
Class A	8Ω	6Ω	4Ω	low
Class AB	n/a	n/a	39	high
Class B	21	28	39	medium

AUDIO PRECISION APLASTS THD+N(%) vs FREQ(Hz)

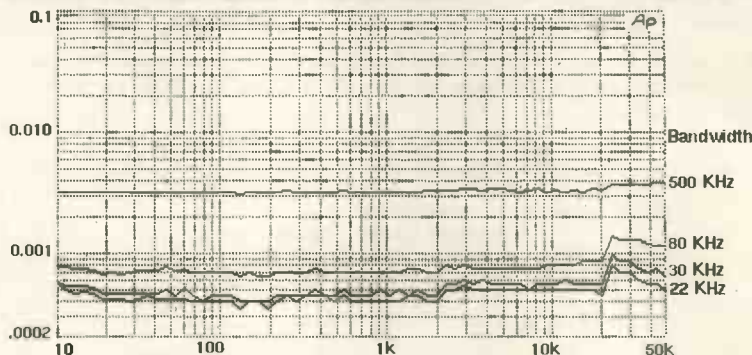


Fig. 5. Distortion plot of the Audio Precision oscillator/analyser combination alone, for measurement bandwidths of 500, 80, 30 and 22kHz. The saw-teeth below 1kHz are artefacts. The residual appears to be pure noise.

AUDIO PRECISION POWRAMP THD+N(%) vs FREQ(Hz)

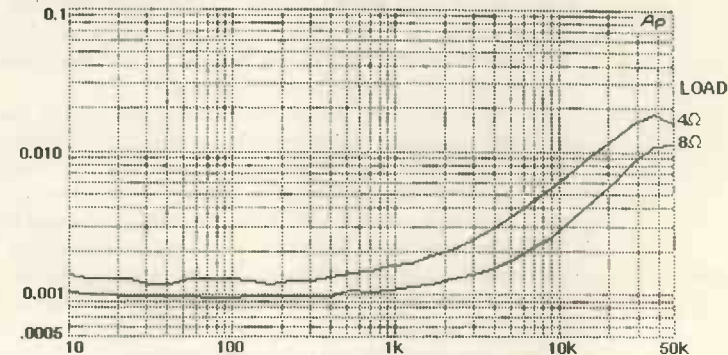


Fig. 6. Distortion in class-B (summer) mode. Distortion into 4Ω is always worse. Power was 20W in 8Ω and 40W in 4Ω, bandwidth 80kHz.

AUDIO PRECISION POWRAMP THD+N(%) vs FREQ(Hz)

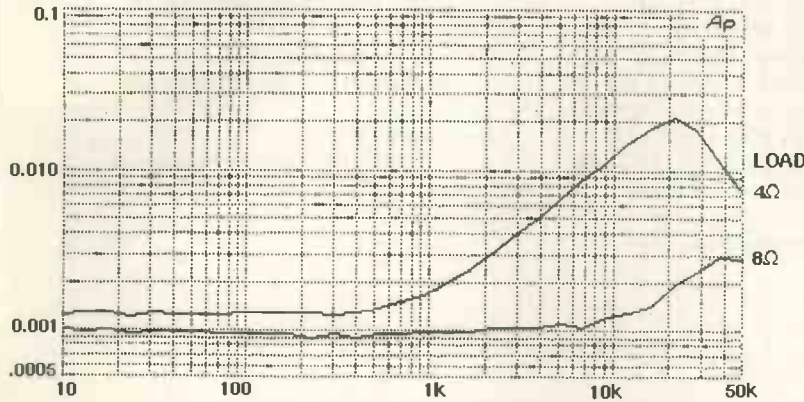


Fig. 7. Distortion in class-A/AB (winter, same power and bandwidth. The amplifier is in AB mode for the 4Ω case, and so distortion is higher than for class-B2. At 80kHz bandwidth, the class-A plot below 10kHz merely shows the noise floor.

AUDIO PRECISION POWRAMP THD+N(%) vs FREQ(Hz)

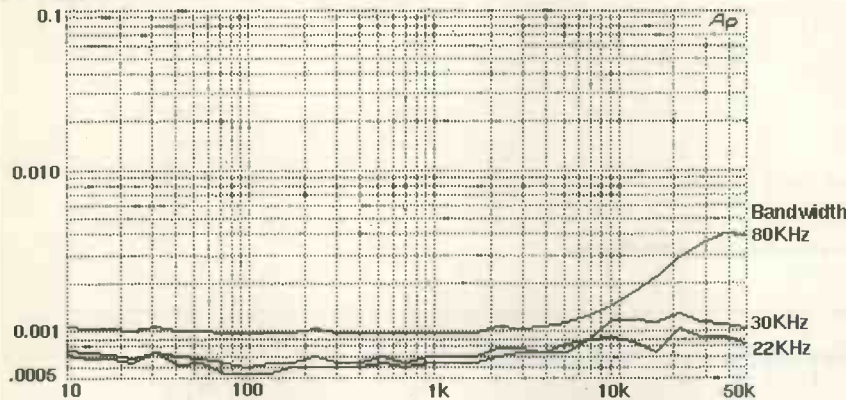


Fig. 8. Distortion in class-A only (20W/8Ω) for varying measurement bandwidths. The lower bandwidths ignore hf distortion, but give a much clearer view of the excellent linearity below 10kHz.

Table 2. Temperature considerations.

	Thermal resist °C/W	heat flow W	temp rise °C	temp °C
Juncn to to3 case	0.7	36	25	100 junction
Case to sink	0.23	36	8	75 TO3 case
Sink to air	0.65	72	47	67 heatsink
Total			80	20 ambient

amplifier changes mode from A to AB with decreasing load resistance. Remember that in this context 'high distortion' means 0.002% at 1kHz. This diagram was produced in the analysis section of PSpice simply by typing in equations, and without actually simulating anything at all.

The most important thermal decision is the size of the heatsink; it is going to be expensive, so there is a powerful incentive to make it no bigger than necessary. I have ruled out fan cooling as it tends to make concern for ultra-low electrical noise look rather foolish; let us rather spend the cost of the fan on extra cooling fins and convect in ghostly silence. The exact thermal design calculations are simple but tedious, with many parameters to enter; the perfect job for a spreadsheet. The final answer is the margin between the predicted junction temperatures and the rated maximum.

Once power output and impedance range is decided, the heatsink thermal resistance to ambient is the main variable to

manipulate; and this is a compromise between coolness and cost, for high junction temperatures always reduce semiconductor reliability, Table 2.

This shows that the transistor junctions will be 80°C above ambient, ie at around 100°C; the rated junction maximum is 200°C, but it isn't wise to get anywhere close to this very real limit. Note the Case-Sink thermal washers are made from high-efficiency material. Standard versions have a slightly higher thermal resistance.

The heatsinks used in the prototype had a thermal resistance of 0.65°C/W per channel. This is a substantial piece of metal, and is expensive.

The complete circuit

The complete Class-A amplifier is shown in Fig. 3, complete with optional input bootstrapping but omitting any balanced-line input amplifier or gain control. The circuitry may look a little complex at first, but we have only added four low-cost transistors to realise a high-accuracy Class-A quiescent controller, and one more to implement mode-switching. Since the biasing system has been described above, only the remaining amplifier subsystems are dealt with here.

The input stage follows my design methodology in running at a high tail current to maximise transconductance, and then linearising it by adding input degeneration resistors R_{2,3}. These reduce the final transconductance to a suitable level. Current-mirror Tr_{10,11} forces the collector currents of the two input devices Tr_{2,3} to be equal, balancing the input stage to prevent the generation of second-harmonic distortion. The mirror is degenerated by R_{6,7} to eliminate the effects of V_{be} mismatches in Tr_{10,11}.

With some misgivings I added the input network R₉, C₁₅, which is definitely not intended to define the system bandwidth, unless fed from a buffer stage; with practical values the hf rolloff could vary widely with the source impedance driving the amplifier. It is intended rather to give the possibility of dealing with rf interference without having to cut tracks. Resistor R₉ could be increased for bandwidth definition if the source impedance is known, fixed, and taken into account when choosing R₉; bear in mind that any value over 47Ω will measurably degrade the noise performance. The values given roll off above 150MHz to keep out uhf.

As a result of insights gained while studying the slewing behaviour of the generic/Lin configuration, I have increased the input-stage tail current from 4 to 6mA, and increased the voltage amplifier stage standing current from 6 to 10mA over the original circuit. This increases the maximum positive and negative slew rates from the basic +21, -48V/μs of reference 4 to +37, -52V/μs; as described elsewhere² this amplifier architecture is always assymetrical in slew rate. One reason is feedthrough in the voltage amplifier current source; in the original circuit an unexpected slew-rate limit was set by fast edges coupling through the current source c-b capacitance to reduce the bias voltage during positive slewing. This effect is minimised here by using the negative-feedback type of current source bias generator, with voltage amplifier collector current chosen as the controlled variable.

Transistor Tr₂₁ senses the voltage across R₁₃, and if it attempts to exceed V_{be}, turns on further to pull up the bases of Tr₁ and Tr₅. Capacitor C₁₁ filters the dc supply to this circuit and prevents ripple injection from the positive rail. Capacitor C₁₄, with R₅, provides decoupling. Increasing input tail-current also mildly improves input-stage linearity, as it raises the basic transistor g_m and allows R_{2,3} to apply more local feedback.

The voltage amplifier stage is linearised by beta-enhancing stage Tr₁₂, which increases the amount of local feedback through Miller dominant-pole capacitor C₃, often referred to as C_{dom}. Resistor R₃₆ has been increased to 2.2kΩ to minimise power dissipation, as there seems to be no significant

effect on linearity or slewing. Do not, however, attempt to omit it altogether, or linearity *will* be affected and slewing much compromised.

As described in reference 3, the simplest way to prevent ripple from entering the voltage amplifier via the negative rail is old-fashioned RC decoupling, with a small R and a big C. We have some 200mV in hand (see Part 1) in the negative direction, compared with the positive, and expending this as the voltage-drop through the RC decoupling will give symmetrical clipping. R_{37} and C_{12} perform this function; the low rail voltages in this design allow the 1000 μ F capacitor C_{12} to be a fairly compact component.

The output stage is of the complementary feedback pair, or CFP, type. As described in Part 1, this gives the best linearity and quiescent stability, due to the two local negative feedback loops around driver and output device. Quiescent stability is particularly important with $R_{16,17}$ as low as 0.1 Ω , and this low value would probably be rather dicey in a double emitter-follower output stage.

Voltage efficiency of the complementary feedback pair is also higher than the emitter follower version. Resistor $R_{25,26}$ define a suitable quiescent collector current for the drivers $Tr_{6,8}$, and pull charge carriers from the output device bases when they are turning off. The lower driver is now a BD136; this has a higher f_T than the MJE350, and seems to be more immune to odd parasitics at negative clipping.

The new lower values for the output emitter resistors $R_{16,17}$ halve the distortion in Class-AB. This is equally effective when in Class-A with too low a load impedance, or in Class-B but with I_Q maladjusted too high. It is now true in the latter case that too much I_Q really is better than too little – but not much better, and AB still comes a poor third in linearity to Classes A and B.

Safe operating area protection is given by the networks around $Tr_{18,19}$. This is a single-slope safe operating area system that is simpler than two-slope safe area, and therefore somewhat less efficient in terms of snuggling the limiting

An adaptive trimodal design?

One interesting extension of the ideas presented here is the adaptive trimodal amplifier. This would switch into class-B on detecting device or heat-sink over-temperature, and would be a unique example of an amplifier that changed mode to suit the operating conditions.

Thermal protection would need to be latching as flipping from class-A to class-B every few minutes would subject the output devices to unnecessary thermal cycling.

characteristic up to the true safe operating area of the output transistor. However, in this application, with low rail voltages, maximum utilisation of the transistor safe area is not really an issue; the important thing is to observe maximum junction temperatures in the A/AB mode.

AUDIO PRECISION APLAST55 THD+N(%) vs FREQ(Hz)

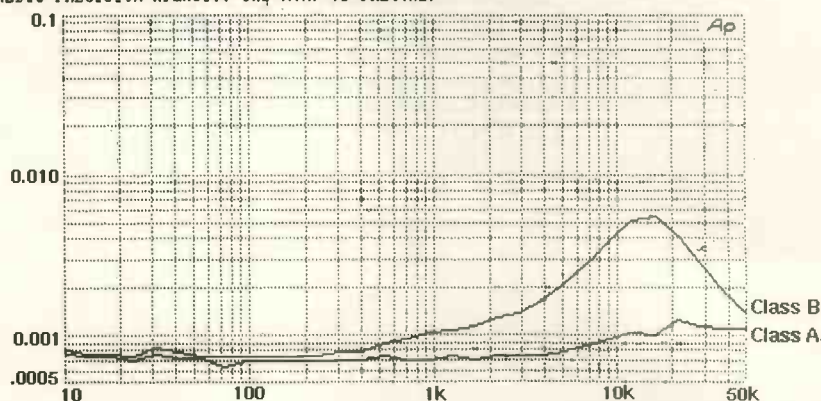


Fig. 9. Direct comparison of classes A and B (20W/8 Ω) at 30kHz bandwidth. The hf rise for B is due to the inability of negative feedback that falls with frequency to linearise the high-order crossover distortion in the output stage.

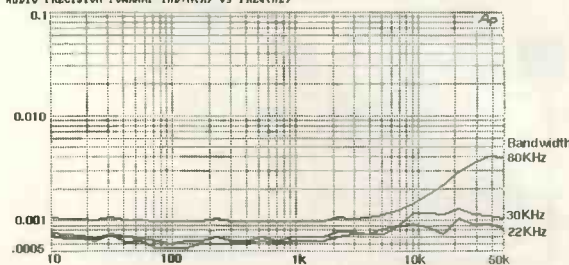
Trimodal power amplifier PCBs

"Performance of a properly-designed class-A amplifier challenges even the ability of an Audio Precision measurement system."

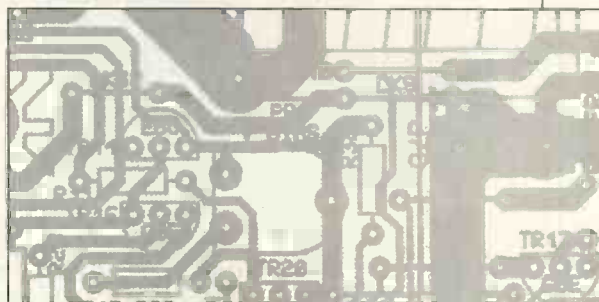
Printed circuit boards for Douglas Self's Trimodal audio power amplifier – detailed in the June and July issues of *EW+WW* – are available exclusively via *EW+WW*. This amplifier can be switched between Class A/AB and Class B to provide remarkable performance over a wide range of operating conditions. In Class A it delivers up to 27W with ultra-low distortion. But presented with a low impedance, the amplifier has recourse to an unusually linear AB configuration.

Designed by Gareth Connor and supplied with a 12 page manual, the silk-screened boards are supplied in pairs at £49.48 per pair, fully inclusive of VAT and UK or overseas postage. Send a postal order or cheque payable to Reed Business Publishing to Trimodal Power, *EW+WW*, room 1.333, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS, together with your address. Alternatively e-mail your address, credit-card number, credit-card type (i.e. Access/Visa) and the card's expiry date to jackie.lowe@rbp.co.uk or fax the same details on 0181 652 8956.

AUDIO PRECISION POWRAMP THD+N(%) vs FREQ(Hz)



Distortion of the Trimodal power amplifier in its class-A mode at 20W into 8 Ω .



The global negative-feedback factor is 32dB at 20kHz, and this should give a good margin of safety against Nyquist-type oscillation. Global negative feedback increases at 6dB/octave with decreasing frequency to a plateau of around 64dB, the corner being at a rather ill-defined 300Hz; this is then maintained down to 10Hz. It is fortunate that magnitude and frequency here are non-critical, as they depend on transistor beta and other doubtful parameters.

Performance

The performance of a properly-designed Class-A amplifier challenges the ability of even the Audio Precision measurement system. To give some perspective on this, Fig. 5 shows the distortion of the AP oscillator driving the analyser section directly for various bandwidths. There appear to be internal mode changes at 2kHz and 20kHz, causing step increases in oscillator distortion content; these are just visible in the thd plots for Class-A mode.

Figure 6 shows Class-B distortion for 20W into 8 and 4Ω, while Fig. 7 shows the same in Class-A/AB.

I would like to acknowledge the invaluable help and encouragement of Gareth Connor. Credit goes to him for the tricky task of pcb layout – and not me, as previous adverts have implied. ■

References

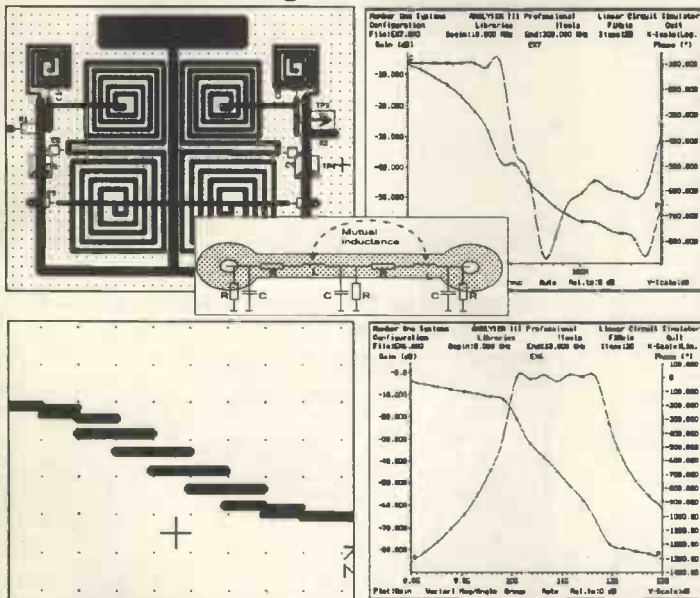
1. Self, D, "Distortion In Power Amplifiers; Part 8" *Electronics World & Wireless World*, March 94, p 225.
2. Self, D, "High Speed Audio Power", *Electronics World & Wireless World*, Sept 1994, p760.
3. Self, D, "Off the Rails" *Electronics World & Wireless World*, March 1995, p201.
4. Self, D, "Distortion In Power Amplifiers; Part 7." *Electronics World & Wireless World*, Feb 1994, p 139.

Notes on part 1

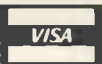
Regrettably, a couple of errors crept into the original article on Class-A¹. On page 229, second column: " $Tr_{15,16}$ then compares the reference voltage with that at point Y" should read "at point X". On page 229, third column: "This comes to the same thing as maintaining a constant V_{bias} across Tr_5 " should read "across Tr_{13} ". This is nobody's fault but mine, and I humbly apologise as it cannot have made understanding the current-controller action any easier. D.S.

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