SUPPLEMENT TO THE LINEAR AUDIO CUBE AMP VOL 8 ARTICLE

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Scope

This brief covers areas mentioned briefly in the Cube-law power amplifier article published in Linear Audio Volume 8.

1. Output stages with gain - the CFP and its derivatives

Why are common emitter output stages not so common?

The 'HC Lin' topology is used in most audio power amplifiers and uses the common-collector output stage where the emitters drive the load, also known as a Voltage Follower. Normally bipolar power transistors are used; either the Darlington, or the Sziklai compound pair are used to provide a high current gain; so the earlier stages can operate a small fraction of the power handled by the power transistors.

Since the Darlington or Sziklai compound output stage operates quite close to unity voltage gain it means that at least one of these earlier low power stages need to operate at the *full supply rail voltage*, and this is usually the VAS (Voltage Amplifying Stage) coined by Douglas Self.

If you want to hear how to pronounce "Sziklai" then listen to an interview of 'Jimmy' Lin here.

Incidentally, the VAS can be understood as part of a CFP-with-gain when a single input transistor is used (as in early variants of the HC Lin amp). The VAS can be either a transimpedance (I-to-V) stage *or* a voltage gain stage depending on whether a low value base-emitter resistor is used with the VAS bipolar – it depends on the frequency since C_{dom} creates a virtual earth at the base at high frequencies.

The following covers output stages with voltage gain? There are many ways it can be done. Some ways are better than others, some sound better than others.

Douglas Self's book has a section on output stages with voltage gain (here or faster Google docs) and similar by Rod Elliot here and neither recommend output stages with voltage gain. R Tobey & J Dinsdale's 'Lin' variant in Wireless World November 1961 showed a "grounded emitter with 100% voltage feedback" here (or CFP as we now know it) and it was shown as one of 3 possible output stage types, but they opted for the Lin type quasi-complementary output stage, and this choice seemed to cement thinking of most designers into this single path.

Bob Cordell's in his book gives a good overview of the problems of the standard CFP. Bob Cordell's 1st Edition does not mention anything about output stages with gain. One of the advantages not mentioned by Douglas Self or Rod Elliot is the ease of scaling up amplifiers to higher voltage rails such as ±80V. With output stage gain the earlier stages (VAS predriver and input stages) can run at a lower voltage which allows smaller lower dissipation transistors, ones that offer higher F_T's and lower cost. With output stage gain we could drive the power stage with one of the many low distortion fast opamps. An example (view here) is covered below.

Power amplifier design books probably do not cover output stages with gain design books due to page limits. I have explored some of the options in my own designs and am fortunate to be able to present some output stages with gain that are already 'out there' without stringent page limits.

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2. Modifying the CFP output stage for gain

Modifying the CFP output stage for gain was demonstrated briefly by Douglas Self and Rod Elliot (links above) but they left out a lot of useful stuff. For example, the so called 'grounded output stage' **Figure 2.1** that provide us voltage gain in the power stage. I call it a 'floating power supply amp'. A 'grounded output stage' implies shorting the output to ground. But of course the other side of the load which connects to the centre rail of the power supply is *now floating*, Figure 2.1. So the whole power supply goes up and down with the output signal and with one simple step we get an output stage with voltage gain. Of course the output stage distortion increases when the gain is increased, but this is usually irrelevant because the gain becomes part of some other feedback loop – either local or global or a combination of these.

A floating power supply is not a problem as long as your auxiliary supplies come from a separate transformer secondary, and you use separate power supplies for each channel, and shield the secondaries from capacitive coupling from the primary.

Some commercial examples of a common emitter (CE) or common source (CS) output stages using a floating power supply are given in Ben Duncan's Audio Power Amplifier book (eg p112). The "Trans•nova" (TRANSconductance NOdal Voltage Amplifier) is a less known example of Error Correction. James Strickland never wrote a technical articles describing it (apart from his US Patent disclosure US4,467,288 21 Aug 1984, patent Fig.2.2). Instead he designed car amps for Rockford-Forsgate and later for David Hafler. BTW David Hafler worked with Herbert Keroes on the *Ultralinear* Williamson kit and founded Acrosound and Dynaco (more).

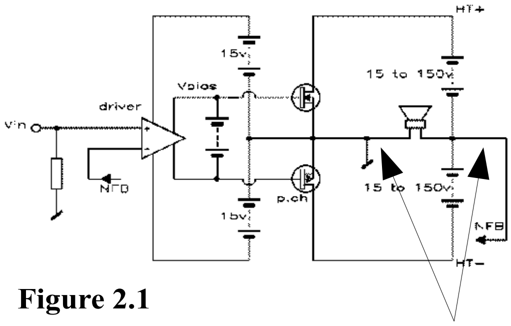

Figure 2.1
Output stage with gain.
Floating power supply.
=> LV opamp drive

Move the common.
Simple!

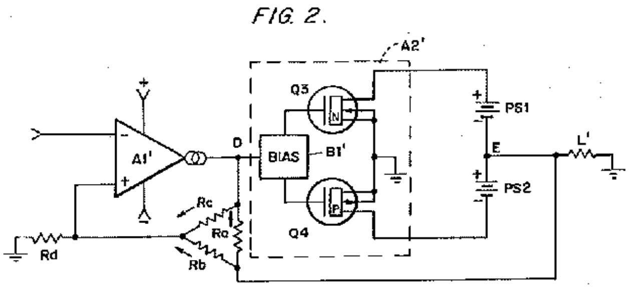

Figure 2.2
'Trans•nova' Strickland 1984 (US4467288)
– Floating PS, LV opamp driven
– Error correction

Now a look at some variants for the bipolar CFP, then variants with gain, then variants with FET's.

Figure 2.3 is the standard Sziklai compound pair (CFP). The power transistors feed boosted driver collector current into the emitter resistors. A CFP is more linear without base-emitter resistors. In the CFP the power transistor ideally operates as a current booster – current-in-current-out or a CCCS. When a resistor is added the current transfer ratio (or effective Beta) becomes much more nonlinear than the raw transistor's Beta. It is not a big concern in Class-A but when used for Class-B the crossover region becomes more jagged which means more high order harmonics. The base-emitter resistors are nearly always added to Class-B stages to increase the turn-off speed and allow more overall feedback but if weighted distortion measurements were used then they show it creates high order distortion. When the power transistors operate as current-in-current-out devices (without base-emitter resistors) then the Class-B crossover region is widened and less high order crossover distortion is generated.

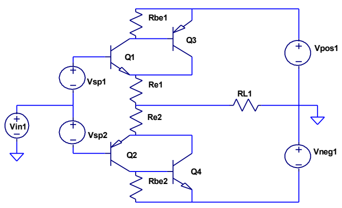

Figure 2.3 Standard CFP
– 100% VFB
– base-emitter R's

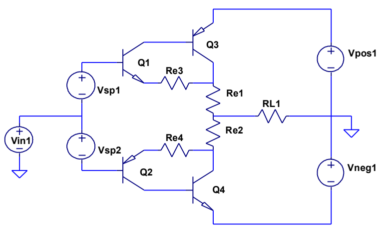

Figure 2.4 CFP Variant-1
– Extra Re's, no Rbe's
– still 100% VFB

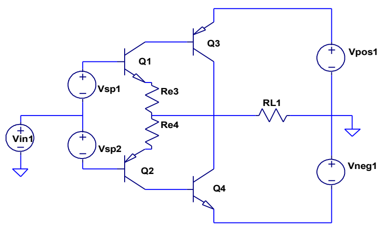

Figure 2.5 CFP Variant-2
– Only driver Re's
– still 100% VFB

Figure 2.4 shows a CFP Variant-1 where the driver transistors are given extra resistors. With extra driver emitter resistance the other resistors can be reduced. This reduces the relative amount of feedback around the top and bottom halves (so less feedback to linearise Beta variations with current and temperature). But this variant does not alter the overall feedback for the combined currents through the load, so the overall linearity not significantly affected. With less feedback around each half the Class-AB crossover region is wider and less high order crossover distortion is generated by this variant.

As a general principle it is better for weighted distortion (what we actually can hear) not to apply local feedback to each half of a push-pull output stage, but to save the available gain for feedback for the combined currents through the load as local or even global feedback (or some combination of these). Effectively what happens when the available gain is used after the currents are combined is the Class-A crossover region of optimally biased Class-B becomes wider which means less high order harmonics and lower weighted distortion (even though standard THD reading appear to be the same). Widening the Class-A crossover region in Class-B necessarily increases the idle current for optimal bias. (Green-conscious citizens prefer as low idle current as possible but making the idle current 2 or 3 times higher does not represent much extra threat to global warming IMHO).

The lack of harmonic weighting for comparing alternative power amplifier approaches has led to 5 decades of misleading indications of supposed 'improvements' in topologies – but in reality very few designs have resulted in a better listening experience. What a disaster! If harmonic weighting was carefully applied then most of these 'improvements' would have been picked up before others test them and give yet another thumb's down review, it's “back to the drawing board”.

Most of my amplifier circuits (eg see LACAv8 Cube-amp circuits for LTspice) now include harmonic weighting using a subcircuit in the amplifier simulation so the ratio of weighted harmonic distortion to unweighted can be checked. If the ratio of the weighted THD is more than about 3 times the unweighted THD then alarm bells should ring because there are significant high order distortion and we know the THD readings will not give a true indication of what we can expect from listening tests. Conversely, an amp that gives relatively high unweighted THD readings (eg 0.01% to 0.1% range) can still sound very good if we also find the ratio of weighted to unweighted harmonics is low (eg 3 or less).

Figure 2.5 takes CFP Variant-1 to the limit where the high current resistors Re1 and Re2 are omitted. The Class-AB crossover region is widened so less high order crossover distortion are generated.

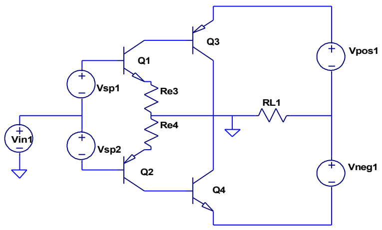

Figure 2.6 CP
– with gain, no VFB
– floating PS

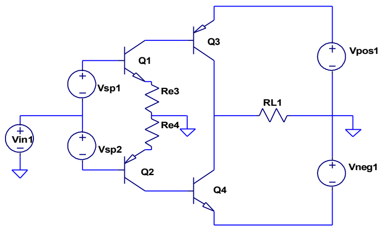

Figure 2.7 CP
– with gain, no VFB
– non-floating PS

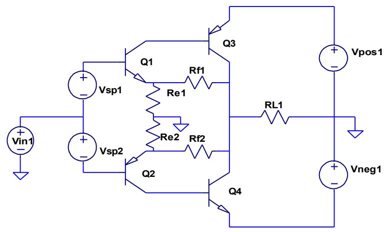

Figure 2.8 CP 'PE Texan'
– with gain
– variable VFB

In the CFP Variant-2 the power transistors are now operating with no emitter resistors and no base-emitter 'pull-out' resistors. This should ring alarm bells, “Warning! DANGER! Thermal runaway!” But NO, Surprise, the power transistors do not runaway … provided the drivers are thermally isolated from the power transistors, the *driver* transistors are thermally linked to the bias voltage generators, and the power transistors have a high enough Vce_{BEO} rating, say half the Vce_{R} (Vce_{max} with a specified base-emitter resistor). Omitting the base-emitter 'pull-out' resistors improves the thermal stability (since the Beta temp. co. is less than the gm temp. co). Charge removal is possible by adding a current source or high resistance to auxiliary rails.

Figure 2.9 shows a practical circuit using MJ15004/5's (here). Notice the absence of emitter resistors and the base is loaded by a diode plus resistor 'mirror' with trimpots to trim symmetry. It was later learned that it is better to use a transdiode (eg BC337 b-c shorted) rather than a 1N4148 since the SPICE model shows the 1N4148 has an N of 2 effectively two diodes in series but a transdiode gives an N of 1 to more closely mirror the power transistor). Another improvement is to add current sources to keep the power transistors slightly on (so called non-switching mode) either a current source to bases of Tr5 & Tr6 (about 100uA for 10mA Ic min) or two resistors (eg 150k) to common with an electrolytic (eg 10uF) to the supply to filter most of the ripple.

Notice the driver's are BiCMOS composite pairs (see ref. 4 at this site); this combination BiCMOS can generate square-law driver currents, giving what I call Square-law Class-AB, and this gives a wider Class-A power region and a lot less high-order crossover distortion than standard bipolar optimum biased Class-B.

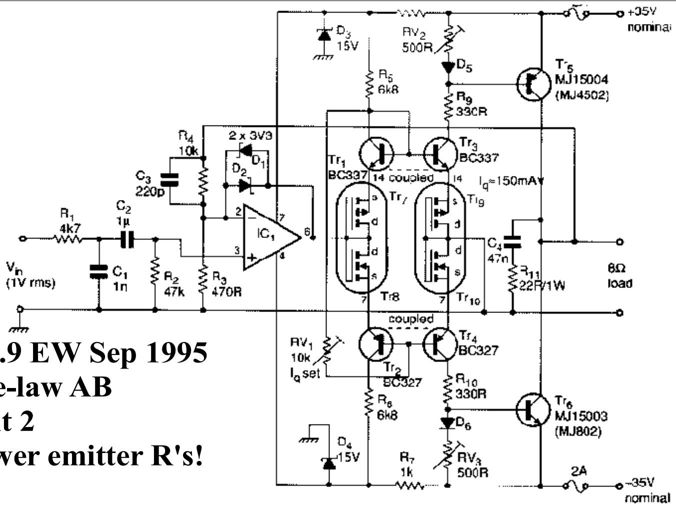

Figure 2.9 EW Sep 1995
– Square-law AB
– Variant 2
– No power emitter R's!

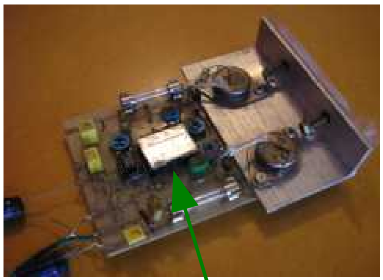

Figure 2.10
CMOS inverters
coupling under plate

The output stage distortion was 0.1% at half the full output swing and for frequencies up to 2kHz in *open loop with a voltage gain of 13* into 8 ohms. Compared to optimally biased bipolar Class-B in CE the unweighted distortion is a factor of 10 lower using Fig.2.9 (better than 10 when weighted THD's are compared) [w-THD update is needed].

Figure 2.9 design shows that output transistors in current gain mode (current boosters) and do not need the careful thermal link that Voltage Follower output stages need. Notice no emitter resistors are used – they are not needed. Only the driver stage needs care with thermal linking to the bias voltage generator. The power transistor operating as a current boosters have a temperature coefficient of +0.3%/°C and this does not cause thermal runaway as long as the base sees a high incremental resistance to the emitter. Also the Beta falls at high currents so the net electrothermal Beta temperature coefficient is reduced. Are you wondering what the diodes soldered to the power transistor cases in Fig.2.10 are for? They were added in case thermal runaway was a problem, mainly from collector-base leakage (see the EW article for details). But this did not prove to be necessary and it can be safely omitted.

Back up to **Figure 2.7.** This is the same as Fig.2.5 but with the output grounded and the power supply floated. This removes all the voltage feedback so we now call it a “CP” for Compound Pair with gain. Think of a “CP” as a CFP without the “F” – a Complementary Feedback Pair without the Feedback.

Interestingly, if we now add a small value current sensing resistor in the output *common* we get load current feedback and the load see a higher output resistance, making a better current source for driving the load and with lower the distortion. As mentioned, there is an advantage placing the feedback after the currents are combined (rather than before they combine such as shown in Fig.3.1 and Fig.3.2). It makes *better* use of the power transistors gain and there's less high-order distortion. In most power amplifiers we don't want a high output resistance, so a better feedback scheme is to use voltage feedback and reduce the voltage gain.

Square-law Class-A distortion is quite sensitive to source degeneration local negative feedback. For example, my square-law Class-A article in Linear Audio Vol.1 showed local feedback was 8 times more with normal value source resistors compared to no source resistors (L|A vol.1 p41).

This is a good example where negative feedback makes output stage distortion sound worse. Class-A and Class-B output stages also sound worse when emitter (source) degeneration is applied to each half of push-pull rather than applying the negative feedback after the two halves currents are combined. It does not mean negative feedback is inherently bad, nor is it related to how much negative feedback is used (see the *Footnote below on misunderstanding Baxandall's feedback analysis).

In bipolar output stages the unweighted distortion (measured as THD) does not change much when local feedback is applied before the two halves currents are summed. But it does change the idle current for optimum bias for minimum THD. The more local feedback used before the two halves currents are summed, the lower the idle current for optimum bias. Lowering the idle current for optimum bias increases the levels of the higher-order harmonics because the crossover region represents a small duration of each cycle and Fourier analysis can show a shorter duration of a variation within a cycle pushes more distortion components into the higher frequency range. When the harmonics are weighted we are better off with less local negative feedback in each half and instead use the gain for negative feedback after the currents are combined.

It means we are better off using as high as possible *optimum* Class-B idle current to give as wide as possible Class-A region, stopping when the heatsink idle temperature reaches the worst-case for Class-B, at about 40% of $Pout_{max}$ (if this level is reached then the heatsink runs at a constant temperature independent of volume level). But the standard bipolar Voltage Follower output stage cannot be operated to such a high optimum bias current because it becomes thermally unstable (Bob Codell's book provides the calculation for thermal stability in this mode on p299-301). To get a higher *optimum* Class-B idle current in a practical bipolar power output stage then the power transistors need to be operated in their Beta current boosting mode, as in amplifier in Fig.2.9.

One way to get the idle current up to where the heatsink runs at a constant temperature no matter what the volume setting is to use Cube-law Class-A.

*Footnote: Peter Baxandall 'Audio Power Amplifier Design, Pt. 6' Wireless World Jan 1994 (view) follows earlier analysis by J. Frommer, Wireless Engineer (Wireless World) Jan 1938 p20-22 and referenced by Fritz Langford Smith in the Radiotron Handbook (p64 free download here) concluding that intermodulation from using a small amount of negative feedback is not a concern with typical amplifiers because the open loop linearity is not bad enough and music seldom reaches peak levels where the intermodulation is a problem. Peter Baxandall's last sentence in his article states that the artefacts from a small amount of negative feedback are not significant for audio because it decreases so rapidly with reducing power [due to music's low crest factor] and Douglas Self and other's following missed this important fact when commenting on using small amounts of negative feedback (eg Douglas Self view and Bruno Putzeys Linear Audio Vol.1 p112-132, Fig.14) and take note that Douglas said, “a careful reading of the Baxandall series is absolutely indispensable” and Douglas' comments on weighting of re-entrant harmonics generated by negative feedback intermodulation are not supported by any measurements and I suggest that they should not be accepted as fact until they are supported by measurements at a range various power levels likely in audio power amplifiers.

Douglas Self has looked at the effect of using low value emitter degeneration resistors in the EF and CFP output stages (EW and book p148-151; EF Table 5.3, CFP Table 5.4). He found lower-value emitter resistors 'wingspread' gain plots had a wider Class-A crossover region and higher optimal idle currents (eg book Fig.5.43 & Fig.5.44). Douglas concluded that the lowest possible emitter resistance should be used where the thermal stability is still adequate. End of story.

Back up to **Figure 2.8** the CP with gain, aka, the PE Texan (Practical Electronics May 1972 view) shown as **Figure 2.11**. This is one way to provide some voltage feedback around the driver transistors. Some designs add a capacitor across the feedback resistors to improve the phase margin when overall feedback is applied to the input stage. For example the Australian ETI-480 power amplifier **Figure 2.12**, which incidentally sold multi-10,000's over 25 years from 1975 to 2000, and it was upgraded by Silicon Chip magazine to a blameless one called the SC-480. I made 4 amplifiers using the ETI-480 including a bridged version for a 200W guitar amp. It was low cost, easy to build and to get running, and none of my amps 'popped' even though it never used SOA load-line protection nor an inductor in the speaker line.

Is the inductor need for this amp? Paul Cambie advises it but has he checked that it is necessary? (BTW, please let me know if you found that an output inductor was actually needed on the ETI-480).

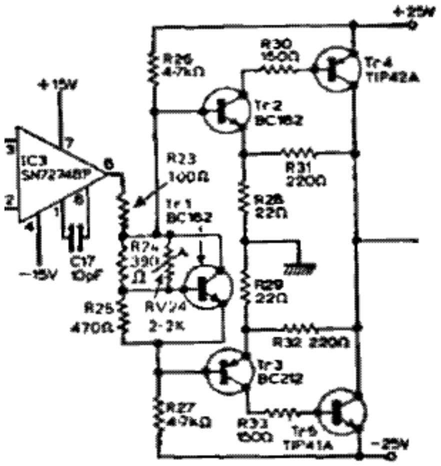

Figure 2.11 PE Texan 1972
– Gain of 9
– No power emitter R's!
– No power base-emiter R's!

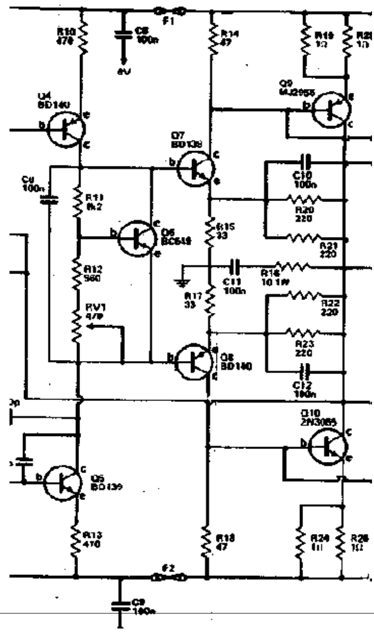

Figure 2.12
– ETI-480 1975
– Gain of 3

The only time I encountered instability was when I bench tested a module with the heatsink floating rather than earthed in the case. With the heatsink floating the capacitance of the thermal washers coupled the output to the driver transistors mounted on the flange causing ultrasonic oscillation. The power transistors survived because I noticed it before they got too hot.

In the ETI-480 two pairs of 220 ohm 1 watt feedback resistors are used for Rf1 & Rf2 and they would get hot and discolour during full power sinewave bench tests, but in normal operation they did not get very hot. In the ETI-480 the output stage was set for a voltage gain of 3 and 100nF capacitors were added across these feedback resistors. In Douglas Self's section on output stages with gain (links above) he gave an example with a gain of 2 using 47 ohm feedback resistors and the power dissipation was 4 times higher than the ETI-480 and this amount of heat is not welcome on a PCB and it's power wasted that could go to the load (mainly to boost sales by getting as many watts output as possible from the parts). If a capacitor is used across the feedback resistors then a higher value resistor can be used without sacrificing bandwidth. So the high power dissipation of the feedback resistors in output stages with gain is not a sufficiently good reason to not use output stages with gain.

In the case of Fig.2.6 & Fig.2.7 (CP no voltage feedback) the voltage gain is given by $Av=gm_{Tot}\times R_L$ where gm with optimal bias is $gm_{Tot} \approx {}^1/_{Re1}\times k\times Beta_3$, for symmetry $gm_{Tot} \approx {}^1/_{Re2}\times k\times Beta_4$ for a balanced circuit, where k is a factor slightly less than 1 due to some current through Rbe_1 and Rbe_2 if they are used (if not then k=1). The base-emitter resistor creates the same beta fall at the low current end as the beta fall inside the transistor itself except the base-emitter resistor makes low current beta-fall much worse. BTW, this does not apply to the Locanthi arrangement where one base-emitter resistor is connected between two bases of a Darling follower output stage called the Locanthi-T, eg Fig's 6.9,17,18 (fast or slow) and Bob Cordell's book eg Fig 5.1b p98-9 www.cordell.com.

There is another variant of Fig.2.8 where only one feedback resistor is used (eg see Fig.3.3). Since the feedback resistors are usually in the 1W to 5W range it is better to use several smaller resistors in parallel to increase the dissipation and spread the heat over a larger area. Notice Fig.2.8 is more convenient with two feedback resistors rather than Fig.3.3 because two 1W carbon resistors are lower cost than one 2W wire-wound resistor.

Also, with Fig.2.8 it is possible to trim one feedback resistor by paralleling to give more feedback to one emitter than the other and this is useful since one power transistor or FET will have a slightly higher gain than the other (see **Figure 3.6**). This output stage gave 0.8% THD and a gain of 20 into 8 ohms in open loop and the class-A power band was 4W [w-THD update needed].

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3. Using FET's in the CFP and the Compound Pair with gain

In most cases FET's can be used as drop-in replacements for bipolar's provided there is always a gate to source resistor present (was labelled the base-emitter resistor above). Possible combinations are FET drivers and bipolar power transistors, or bipolar drivers and power FET's, or FET drivers and power FET's. Until recently matched complementary small signal FET's have been rare and parameter spreads we large requiring matching by hand. Also power FET's (usually MOSFET's) are more robust SOA wise since they do not suffer from secondary breakdown and do not require harsh de-rating to survive domestic and stage environments. Another advantage is FET's do not have storage turn-off delay so they can slew from rail-to rail up to several MHz without blowing up – bipolar's storage restricts full rail-to rail slewing to less than 1/10th of their F_T with good charge removal and 1/100th of their F_T (maybe less) with no base-charge pull-out (ball-park figures).

It is more common to use bipolar driver transistors and MOSFET power transistors but this is not a rule and you can make good output stage and good amplifiers with any of these combinations. As mentioned some of the options give less high-order distortion in Class-B.

Figure 3.1 shows the standard CFP with bipolar drivers and power MOSFET's.

Figure 3.2 shows a CP with no voltage feedback using bipolar drivers and power MOSFET's.

Figure 3.3 shows a CP with some voltage feedback with bipolar drivers and power MOSFET's. Notice in this case only one feedback resistor is used. Two resistors could have been used as in Fig.2.8.

Figure 3.6 uses bipolar drivers and power MOSFET's and two resistors to balance the gain of each half (as mentioned above). **Figure 4.1 and Figure 4.2** show plots of a circuit similar circuit to Figure 3.6 show the 'Batman' gain curves (Fig.4.2) arise when using MOSFET's in common source with no source degeneration resistors to get a high idle current and as wide as possible Class-A power band. These are discussed in the following section.

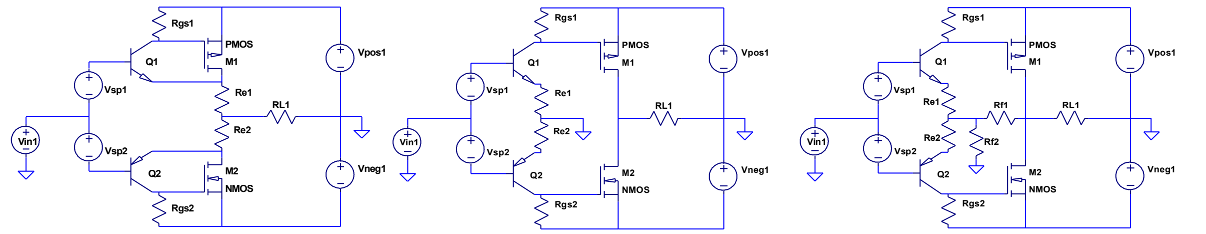

Figure 3.1 BJT-FET CFP – 100% FB | **Figure 3.2 BJT-FET CP – gain, no VFB** | **Figure 3.3 – single feedback resistor**

Examples of MOSFET drivers with MOSFET power transistors are Nelson Pass F5 Class-A (audioXpress May 2008) using j-FET drivers and d-MOSFET's for power FET's. An earlier example is Ivor Brown EW April 1990 (**Fig 3.5**) with a gain of 9.

The F5 is a complete amplifier, it is driven by a low source resistance line driver stage at about 2V rms and gives around 10 ppm mostly 3rd distortion at 1 watt and 25W/8R 1% THD. Idle current is 1.2A and idle dissipation is 62 watts and can provide ±10A peaks into low impedances (by operating in Class-AB mode with low impedance loads) and has an output resistance of 0.1 ohms. The -3dB bandwidth is 1MHz. This amazingly simple amplifier sound good but needs quiet power supply rails, such as the capacitance multiplier as described in Linear Audio Vol.3 p17 by 'Patrick K' earlier described in Andrew Ciuffoli's Class-A design here and here (EW May 2000).

Ivor Brown's output stage Fig 3.5 is similar to the Pass F5 but Ivor uses lower gain lateral power MOSFET's and is designed to operate in Class-AB with a lower idle current. The output stage bandwidth is 3MHz and the overall amplifier is 1MHz but reduced to 40kHz using the input filter and output inductor and could drive a 2uF load with little overshoot indicating a good phase margin.

The distortion was very low and hard to measure with the available equipment but reassuring that there were no measurable high order harmonics popping up like bipolar power amps (THD was around 0.001% and the 7^{th} 0.0001% at the limit of the equipment). This was a remarkable achievement and none of Douglas Self's subsequent bipolar designs claimed the low-order distortion that Ivor Brown in 1989. High-order distortion from standard Class-B bipolar amps generate distortion that when weighted is at least 10 times more than a good MOSFET designs such as Ivor Brown's 1990 one (L|A Vol.1).

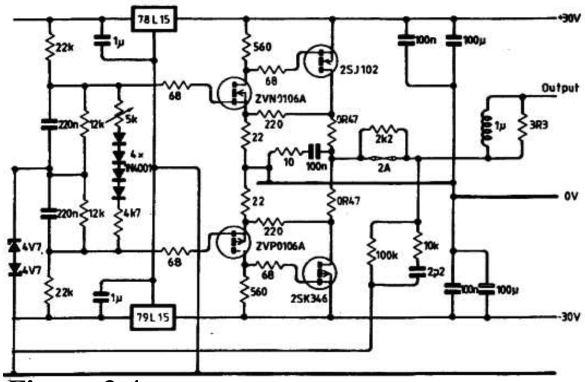

Figure 3.4
Ivor Brown, EW Apr 1990
CP Common Source, MOSFET with gain of 9

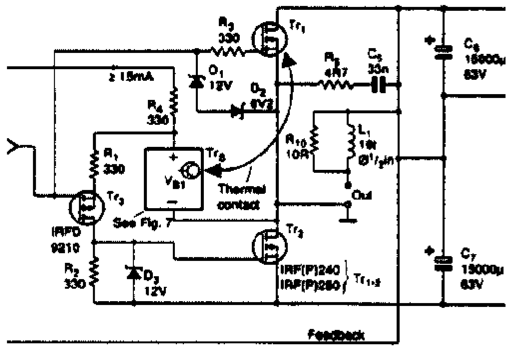

Figure 3.5
Bengt Olsson, EW Dec 1994
– Non-complementary power MOSFET's
– Common Source Sq-Class-A or Sq-Class-AB

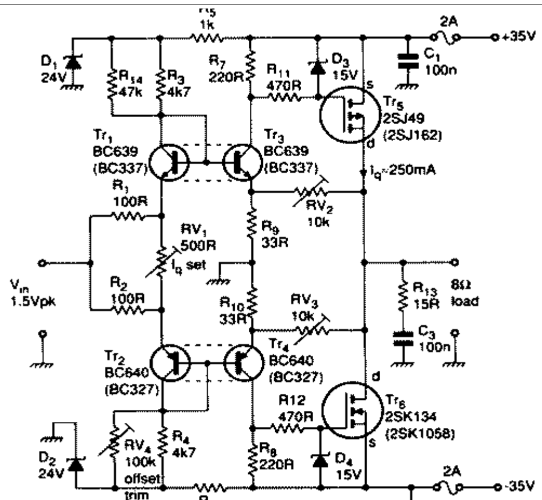

Figure 3.6 Sq-AB
EW Sep 1995

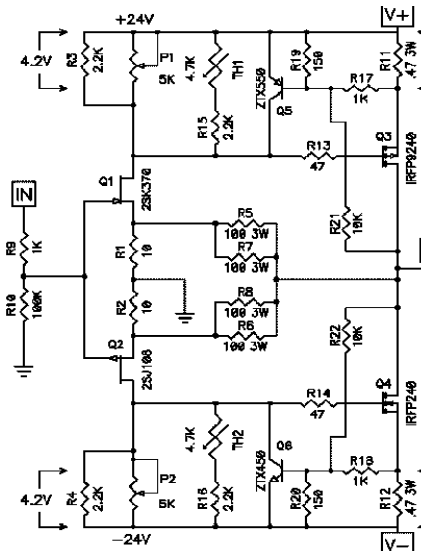

Figure 3.7
Pass F5

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4. MOSFET 'Batman' gain curve

Douglas Self questioned whether MOSFET's are more liner than bipolar's in power amps, and asked whether it was a case of Hi-Fi Press hype to sell the new MOSFET technology,
"*The worst drawback of FET's is that they are so depressingly nonlinear, despite what you sometimes read in the Hi-Fi press...*"
(commenting on Ivor Brown's Electronics World 1989 MOSFET design in Electronics World Letters February 1994).

This could have been a deliberate statement to provocative a vigorous debate. This debate ran for over a decade in Electronics World from 1990 to 2004 and stopped when a new editor banned more audio amplifier articles. Over this period there were around 70 articles on audio power amplifier design. Around 65 of them used the Lin/Self topology with the Voltage Follower output stage. Only 2 or 3 used an output stage *with voltage gain* like Ivor Brown.

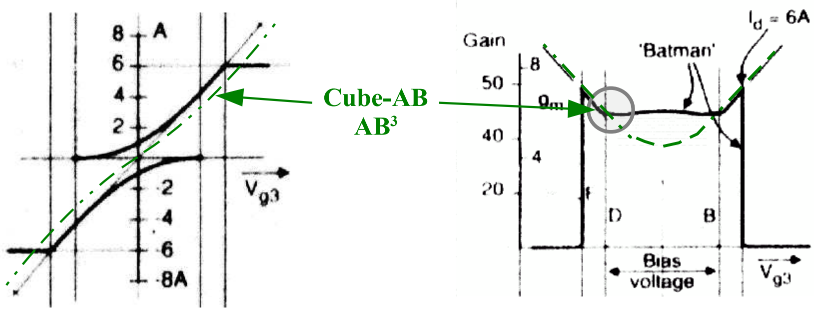
One MOSFET output stage *with voltage gain* was by Bengt Olsson's (Dec 1994) **Figure 3.5. A**nother with gain was my article Fig.2.9 & Fig.3.6. My article appeared within months of Bengt's article and all our development was done independently. Bengt Olsson's design was critiqued by Douglas Self (EW Sep 1995) but I was spared.

One key weakness of Douglas' arguments over linearity was that he was relying on standard unweighted THD measurements (just like everyone else) with no calculations for how much (or how little) high order harmonics from Class-B crossover distortion affected the listening experience for MOSFET and bipolar output stages. There was (and still is) no accepted standard mathematical formula for doing this. Even though Douglas mentioned Shorter's 1930's $n^{\hat{}2}/4$ harmonic weighting algorithm he never applied it to any of his measurements in his articles or books.

In a recent article in Linear Audio Vol.4 a way was given to weight the harmonics from an amplifier using a filter to emulates our hearing sensitivity so that it can be added to a standard distortion analyser. After publishing this I noticed someone had independently developed this approach for a PhD using add-on filers here. I am hoping another independent researcher will take up the cause and check whether the method works as stated.

Continuing on the debate in Electronics World. Douglas wrote the following shortly after Bengt's and my article,
"*It may be possible to partly cancel FET square-law distortion by push-pull operation. But this can only work in Class-A, when both upper and lower output devices are connected at the same time. Economic necessity and energy conservation mean that most amplifiers are Class-B, and to date there is no practicable compromise between these two modes. If FET's can only give acceptable linearity in Class-A, then this is not much of a recommendation for them*"
Douglas Self, 'Hazy linearity notions?' (Letters EW Apr 1996 p330) a reply to Mr Kiyokawa's Letters EW Jan 1996 p37-8.

Notice the good linearity of MOSFET's derives from their square-law nature. Douglas is right saying square-law's are not so good with lower idle currents, for Class-AB mode. But Douglas never used harmonic weighting to quantify this disadvantage (to see if it was really a disadvantage in practice). The next section covers this question in detail.



Cube-AB advantages:
- **smooth gain curve**
- **clean A to B crossover**
- **easy to linearise with NFB**
- **sounds like Class-A**
- **efficiency of Class-B**

Figure 4.1 **Figure 4.2**

Ref: Bengt Olsson, 'Super symmetry misunderstood?' EW+WW Letters Oct 1995 p886-7. Fig's.1,3

One question: Is distortion from MOSFET's in Class-AB is better (or worse or same) compared to bipolar Class-B at optimum bias? This is where the 'Batman' gain curves in **Figure 4.1** and **Figure 4.2** can tell us something.

By inspection, MOSFET's the idle current can be set a lot higher than bipolar's we omit the source resistors to give a wide Class-A power region that can extend to more than 1 watt which is more than enough to cover most of the audio signal range in a 50W or 100W power amplifier.

Similary, bipolar Class-B at optimum bias has a much lower idle current that only gives 25mW Class-A range and after that we get into a higher-order distortion range that is worst in the few watt region where most of the audio information is conveyed (Ben Duncan's PA book p136).

So there seems to be a significant advantage here for MOSFET's in Class-AB over bipolar Class-B at optimum bias.

BTW the Square-law Class-AB 50W amp Fig.2.9 gives 4W in Class-A with 250mA idle current which gives an idle dissipation of 16W for a 50W/8R amplifier. The current gain mode of the power transistors is the key to this design that makes it notably different to standard Voltage Follower bipolar Class-B. It is really a current boosted MOSFET amplifier operating in Class-AB. A niffty synergistic trick to get good linearity at high currents was to use the normally unwanted high current beta fall to partly flatten the 'Batman' wings (the unwanted gain rise from square-law FET's after leaving Class-A). As mentioned, open loop output stage distortion was 0.1% at half the full output swing and for frequencies up to 2kHz in *open loop with a voltage gain of 13* into 8 ohms – . so the unweighted distortion is a factor of 10 lower than bipolar Class-B at optimal biased in CE and the class-A region is 4W compared to 50mW for bipolar Class-B.

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5. Cube-law Class-AB advantages

In my latest article in Linear Audio Vol.8 shows that cube-law's offers an improvement over square-law's in Class-AB by doubling the Class-A width or 4 times the power with square-law's and the gain slope when crossing the Class-A boundary is smoother being a continuation of a parabolic curve in Class-A before crossover that changes to a ramp after crossover so there is not an abrupt change in the gain slope as shown in Fig.4.2 which is an open loop gain plot. Cube-law Class-AB gives a gain curve like the dotted **green** curve in Figure 4.2 and since this is the open loop gain when feedback is to be applied it squashes the curve to an almost flat line – and because there are no apparent high-order harmonics in there you can say it is *as good as Class-A*.

The Class-A boundary is smoother being a continuation of a parabolic curve in Class-A before crossover that changes to a ramp after crossover so there is still some remaining crossover distortion produced but the spectrum of harmonics falls away 20dB per decade faster than square-law Class-AB and 40dB per decade faster than standard high bias Class-AB.

In a nutshell this is the nub of AB^3 – the **green** curve in Figure 4.2 shows what, how and why AB^3 is *as good as Class-A* once you apply some feedback to suppress the mainly 3rd low order harmonics. Here's the main points:

- Circled is where square-law gain changes abruptly when leaving Class-A – hence crossover distortion from Square-law-AB.
- Notice the green curve glides through without a noticeable bump – hence very little crossover distortion from Cube-law-AB.
- Because the green curve is smooth it's distortion is just like Class-A – all low-order harmonics, very little high-order crossover distortion.
- We can now reduce the low order harmonics with a small amount of negative feedback.
- It sounds as good as Class-A with very little feedback
- It has the efficiency of Class-B

With cube-law's there now is a *practicable compromise between Class-A and Class-B modes*.

In mathematical terms square-law Class-AB gives a *discontinuous* second derivative at the Class-A to -B boundary whereas cube-law Class-AB has a *continuous* second derivative at the Class-A to -B boundary. Simulations show the harmonic spectrum falls off -20dB per decade *faster* with cube-law Class-AB than square-law Class-AB and this can be understood as being equivalent to integration of harmonics or a Low Pass Filter (LPF) of harmonics but not the fundamental.

Applying the same analogy to gm-doubling generated by high bias bipolar Class-AB we see the gain has an almost step change (doubling or halving) when crossing the Class-A boundary. Compare this to square-law Class-AB gain which is a ramp and the gain ramp can be seen as an integration of a gain step. Therefore the harmonic spectrum with square-law Class-AB falls off -20dB per decade faster than standard high biased Class-AB (linear-law Class-AB) and this is seen in simulations.

For cube-law Class-AB it does not need precise cube-law functions to be effective so it is not difficult to make a cube-law Class-AB power amplifier using off the shelf part (see my Linear Audio Vol.8 article). Furthermore, power MOSFET's have to be used in the output stage to achieve this so we could be applied to bipolar output stages (but nothing simulated or tested to date). Using small signal MOSFET's for generating square-law's in the driver stage may be better than all bipolar driver stages, but it is still early days for cube-law amps.

And there's no need to be overly concerned about relatively high distortion levels in open loop (around 10% for -AB) which arises from the parabolic gain curve (see Fig.4.2 green line) – because with cube-law's crossover distortion is effectively *all low order distortion* so we know that relatively high levels of low-order crossover distortion can be easily made inaudible simply by applying a bit of overall feedback. Only 20dB of feedback is adequate and this also lowers the output resistance for driving standard loudspeakers. We can add a safety margin of 10dB or 20dB more to cover all bases.

Cube-law Class-AB's crossover distortion effectively gives us an audio quality indistinguishable from Class-A for the same THD levels. So we only need to apply the same relatively small amount of feedback to make all distortion inaudible. This means simpler amplifiers with good sound quality and high efficiency as well.

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6. Simulated Cube-law-AB distortion roll-off

Figure 6.1 left spectrum plot shows simulated distortion for the 50W MOSFET Square-law Class-AB circuit (Fig.3.6 EW Sep 1995 here) at two power levels 50W and 25W into 8 ohms and no local or overall feedback. Unweighed THD is 4% at 50W and the idle current is 244mA. The first few harmonics (3rd to 11th) follow around -60dB/decade (light dotted line) then higher-order harmonics fall faster. The average rolloff including higher-order harmonics is -80dB/decade (dark dotted line).

The right side spectrum plot show the same MOSFET's configured for cube-law Class-AB at 50W and the same idle current and no local or overall feedback. The unweighted THD is 10% at 50W.

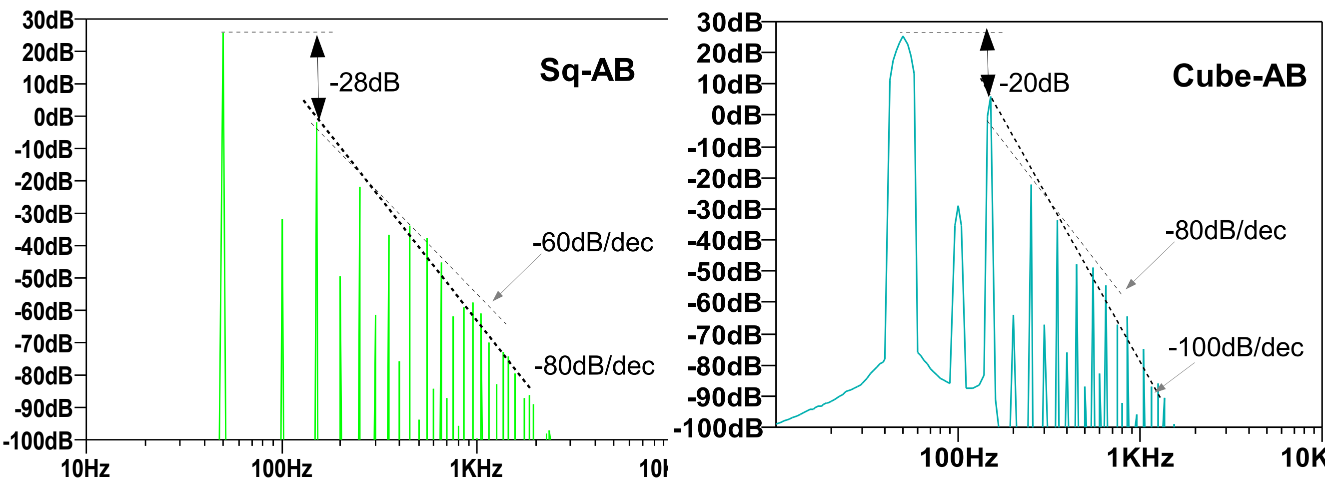

Figure 6.1 Left. Square-law-AB, 50W/8R THD 4% open loop. Rolloff -80dB/dec.
Figure 6.2 Right. Cube-law-AB at the same power and Iq(244mA). Rolloff is -100dB/dec.

Similar to the previous case, the first few harmonics (3rd to 11th) follow around -80dB/decade (light dotted line) then higher-order harmonics fall faster. The average rolloff including higher-order harmonics is -100dB/decade (dark dotted line).

There is a crossover point at the 5th harmonic where the open loop level is the same for is the same for both square-law and cube-law. Cube-law Class-AB harmonics then reduce faster than Square-law-AB harmonics starting from the 7th harmonic.

It show Cube-law Class-AB harmonics fall away -20dB/decade faster that Square-law Class-A.

The simulation was repeated at half full power (70% of full output swing) and the Square-law Class-AB distortion reduces to 1.7% and the harmonic roll-off rate increases slightly (x3). Notice Cube-law Class-AB harmonic roll-off rate is *significantly faster* (x100) at half power (**Table 6.1** right column).

Table 6.1 also shows weighted distortion with two RTU-468 weighting filters at 50 Hz. This shows how much the roll-off slopes are changed for each of the 2 output stages at the 2 power levels. The bottom row shows that in both cases weighting increases the slopes by 40dB as expected (Linear Audio article Vol.4).

Table 6.1. Harmonic roll-off rates for Square-law-AB and Cube-law-AB

	Unweighted slope (ave)		**Weighted slope (ave)**		**Power change**	
	SLR-AB	**Cube-AB**	**SLR-AB**	**Cube-AB**	**SLR**	**Cube-law-AB**
Full power	-80 dB/dec	-100 dB/dec	-40 dB/dec	-60 dB/dec	ref	ref
Half swing	-90 dB/dec	-140 dB/dec	-50 dB/dec	-100 dB/dec	-10dB	**-40dB**
Class change	ref	-10dB **-50dB**	ref	-20dB **-50dB**		
Weighting change	ref	ref	+40dB	+40dB	-	-

Summary: Harmonics from Cube-law Class-AB crossover distortion reduce as the harmonic number increases at the rate of -20dB/decade *faster* that Square-law Class-A. A -20dB/decade improvement in higher order harmonics is just like adding an integrator or LPF to attenuate the crossover distortion. This makes high-order harmonics from Cube-law Class-AB an insignificant problem to our hearing.

Consequently, Cube-law Class-AB can be accepted in the Class-A hall of fame. There is no a "practicable compromise between these two* modes" [*Class-A and Class-B]. Cube-law Class-AB offers the excellent sound quality of Class-A with an acceptably low idle power dissipation comparable with Class-B. For example, Cube-law Class-AB_{50} (half the idle dissipation of Cube-A) offers a quarter the idle dissipation of Square-A, or an eighth the idle dissipation of standard Class-A. Cube-law Class-AB is not difficult to make with standard parts to be stable with temperature changes due to power level variations and idle current setting is not critical like Class-B's optimum bias condition.

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7. Estimating the distortion spectrum from gain plots

A gain versus input signal plot is a versatile method for estimating the distortion spectrum. Whereas a distortion spectrum must be measured at one specific input level, the gain plot method gives a bigger picture of how the distortion spectrum changes over a wide range of input levels without the need for processing many individual distortion spectra. It can be done by inspection once you understand what causes what.

First a look at the distortion spectrum of various common signals like the square wave, triangle wave. A triangle wave can be obtained from a square-wave by integration as shown in **Figure 7.1**.

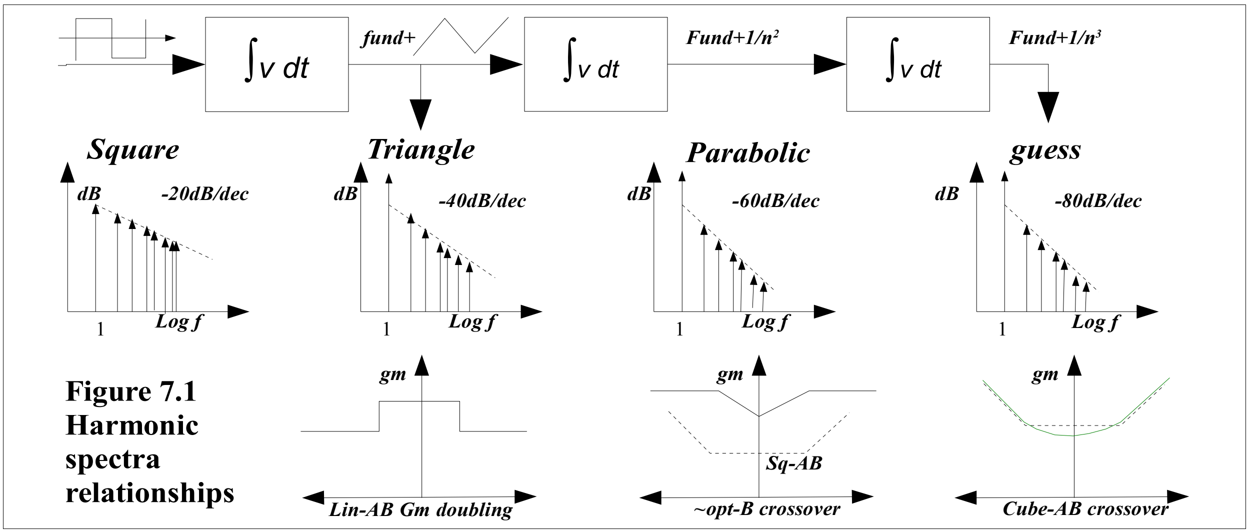

Figure 7.1 Harmonic spectra relationships

A square wave has a harmonic spectrum that reduces with 1/n or a slope of -20dB/decade on a Log-Log plot. When a square wave is integrated we get a triangle waveform and from Fourier analysis we know the harmonics fall as $1/n^2$ which means -40dB/decade rolloff as shown in Figure 7.1. The gain plot shows the fundamental is the main part of a composite waveform made up of the input sinewave and the triangle wave represents the harmonic distortion.

The next level is a ramping gain plot. This can be viewed as an integration of the previous gain step, so we can now expect a slope of -60dB/decade for the harmonics generated by this type of nonlinear distortion. The 3rd level with a gain plot that is a parabola we can apply another integration giving an expected -80dB/decade roll-off in harmonics.

This intuitive approach is supported by simulations in the previous section. Cube-law-AB rolled off initially at -80dB/decade, higher harmonics were slightly faster and this can be explained by extra gain rounding in practical circuits.

In Figure 7.1 the gain step is typical Linear-law Class-AB where gm doubling occurs causing the gain to double in the Class-A region. With some negative feedback the gain change in the Class-A region is compressed (as shown). The spectrum contains suppressed harmonics from the 3rd and rolling-off at -40dB/decade.

The middle gain plot has two cases: the *solid line* approximates optimum bias Class-B. The *dotted line* approximates square-law Class-AB where the gain ramps up after leaving the flat square-law Class-A region. For these the distortion rolls off at -60dB/decade spectrum. For Class-B this occurs at most power levels but with square-law Class-AB the spectrum only appears after leaving the Class-A region with mostly 3rd in the Class-A region.

The right side gain lot is cube-law Class-AB; the gain curve is parabolic in the Class-A region and beyond the Class-A region the curve continues to rise as part of a parabola. Within the Class-A region there is mainly 3rd harmonic distortion. Beyond the Class-A boundary harmonics rolloff at -80dB/decade. So there is a -20dB/decade harmonic rolloff advantage compared to Square-law-AB and this is seen in simulations (above).

The final step in this approach is to apply weighting to harmonics to give a idea of whether the harmonic roll-off is reversed by weighting to predict whether we are likely to hear the distortion even though the THD values seem very low. Based on hearing sensitivity curves a +50dB weighting covers the worst case. This is what we expect to hear:

- Linear-law Class-AB (gm doubling) will have increased weighted THD levels due to high-order harmonics. But this only applies once the power exceeds the Class-A limit. With music the audibility depends on a lot of factors including: amplifier Class-A limit, loudspeaker sensitivity, volume setting and music type.

- Optimum-biased Class-B is on the border line since roll-off is -60dB/dec and weighting boosts by +50dB/decade. But Class-B distortion is generated at all power levels since the Class-A band is usually only a few ten's of milliwatts. An optimum-bias is difficult to maintain in standard output stages so Class-B depends on using as much negative feedback as possible to ensure distortion is not audible.
- Square-law-AB is also border line but high-order harmonics only arise once the power is greater than the Class-A limit and in the Class-A band distortion is mainly 3rd harmonic. If the Class-A limit is more than the average power (usually only a few watts before clipping occurs) then Square-law-AB has an advantage over Class-B and Square-law-AB will not need to use as much negative feedback as possible to ensure distortion is not audible like Class-B.
- Cube-law-AB is safe from high-order distortion; the roll-off rate of -80dB/dec after weighting +50dB/dec is well below the level needed to make high-order harmonics a significant part of what we heard. Beyond the Class-A power level we will still hear only low-order harmonics (3rd and maybe 5th) from Cube-law-AB and anything else is masked. Since low-order harmonics from Cube-law-A are also 3rd and maybe 5th we can say that there is no significant deterioration in sound quality when operating into Cube-law-AB mode. This is unlike all the other modes where high-order harmonics can affect the sound quality when operating beyond the Class-A limit and into the Class-AB region.

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8. A flow chart for designing a Cube-law amp

Flow Chart 8.1 shows the steps for designing a cube-law Class-A amplifier. The aim is to get the end of Class-A at the rated full power with soft clipping. 'Rated full power' is where the soft clip slope is half the normal slope and the distortion is just audible with a sinewave and not audible with most music that stays just below this clip level.

For example the Linear Audio circuit simulation at the rated power measured where the slope is half the normal slope is 100.5W average and 3.3% THD. The clip indicator output is 6V peak. The Linear Audio circuit gives Class-A up to 66W peak which is 75W average due to the distortion and the THD at this power is 1.2% (clip indicator voltage is 2Vpk). The LED indicator begins to light at 1.5V peak which is at about 0.5% THD which is the level that the 3rd harmonic is just detectable. The difference in input level going from the end of the Class-A power to the rated power is 1.5dB. The dynamic range of music that spans at least 40dB and the top few dB are visited less frequently.

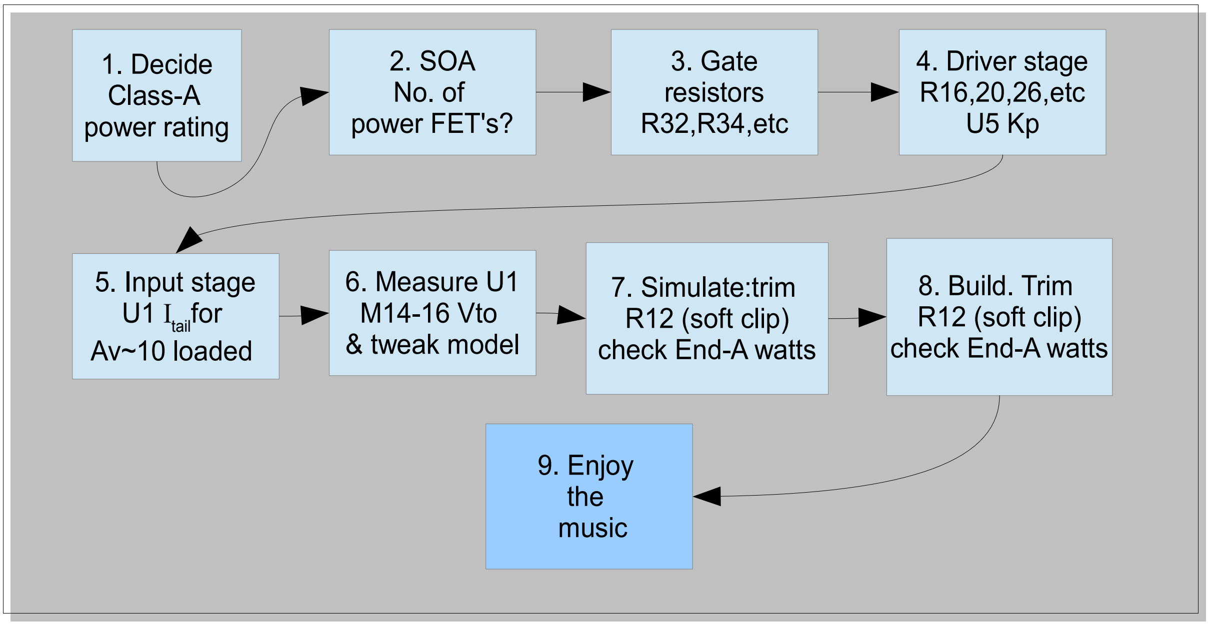

Chart 8.1. Flow diagram for designing Cube-law amps

The 2nd step is the SOA for the power MOSFET's. This is covered in **Section 9**.

The 3rd step is choosing gate resistors. The gate stopper resistors are the same but depend on whether singe or dual-die packages are used. The gate to source termination resistor R32 & R33 depend on how many MOSFET's are used in parallel, being around 600 ohms per single die and reducing in proportion to the number of die in parallel. For example the 100W version with two dual die packages in parallel uses 150 ohms for R32. This is covered in Section 9.

The 4th step is resistors for the idle current of the square-law stage R16 & R17, resistors for the idle current of the linear stage R20 & R21, and the gain of the linear stage R26 & R27, and the gain of the square-law stage which is determined by U5 CMOS model parameter Kp. This is also covered in **Section 10**.

The 5th step is getting enough gain from the input stage when driving the output stage.

The 6th step is to measure the CMOS inverters threshold voltages using a test jig and tweak the models to these values. CMOS inverters threshold voltages vary quite a lot between manufacturers and the threshold voltage determines the input voltage range and affects the feedback resistor R12. This is also covered in User Guide Chapter 2 Section 6.

The 7th step is to simulate it. The EKV lateral MOSFET model is used for simulations and the fitting of this model to your measured threshold voltages is covered in the User Guide Chapter 2 Section 7.

The last design step is to build and test it and measure how many watts you get at the end of Class-A. Why is that important? To see if you achieved your design goal.

If your design goal was to make an amplifier that sounds good then don't forget to do listening tests. Set up another amplifier and switch between with equalised volume levels, possibly with a common volume control, and keep a diary so you or others can repeat your auditions.

9. SOA calculations

The output current needs to be sufficient to drive nominal 8 ohms speakers with impedance dips down to 3 ohms at certain frequencies. Usually this requirement is bench tested with an intermittent 4 ohm resistive load. This means the amplifier needs to deliver twice the peak current for a nominal 8 ohm load. For a 100W RMS amplifier into nominal 8 ohms the peak current is 5 amps. So we need 10A peak into 4 ohms. The rail voltage needs to be around 50 volts to achieve this.

Next, we want the amplifier to survive accidentally shorted output terminals are when the amplifier is running. The power transistors carry the rail voltage and twice the nominal load peak current. So the peak dissipation is 500 watts and this is shared with an alternating signal so it averages at 250W per device. The SOA of the power MOSFET's is 125W for single-die types and 250W for the dual-die types. This level of power dissipation cannot be sustained for long with a heatsink chosen to handle the the heat dissipation for 100W RMS into 8 ohms, which is around 30W to 60W (depending on the signals crest factor and whether the Class-AB or Class-A mode is in use). A relatively fast circuit trip is therefore needed to stop the power transistors overheating when the output is shorted for more than a few seconds.

Dual-die lateral MOSFET's are rated at 250 watts when their case is at 25°C. They should not be pushed to their absolute maximum so some derating is needed. A 20% de-rating limits their junction temperature 250°C rather than the 300°C limit. De-rating 250W becomes 200W. Therefore 400W average requires 2 pair of dual die power MOSFET's (see the parts list for suitable parts).

A fuse in the output line is a simple and reliable way to break the output circuit for longer term output shorts. The fuse can be placed inside the global feedback path so fuse distortion becomes a non problem. To make a fuse open in say 2 seconds then the current needs to be at least twice the fuse rating. But we don't want the fuse to blow when bench testing the amplifier into a nominal 8 ohm load, not even when the amplifier is run continuously with lots of clipping which increases the average output current by up to 1.4 times to 5A for our 100W design. Therefore we need to use a 5A UL rated fuse. A similar IEC fuse could be 4A since the standards are slightly different [ref littlefuse URL]

Therefore a direct short will blow a 5A UL fast fuse in around 2 seconds with a 10A current limited output stage and when the output hard limits when shorted with negative voltage feedback. But a safety margin is needed to cover variations in fuses, circuit current limit variations and non-square waveforms. A safety factor of 50% should cover these where the peak current is increased to around 15A.

Also the power MOSFET's need to be increased from 250W to 375W. And to provide a safety margin for the MOSFET's we need two dual-die 250W devices in parallel per side, or alternatively 4 single-die 125W devices in parallel per side. This will make a robust 100W 8 ohms design.

A more economical version is possible with one dual-die (or two single-die) for 100W into 8 ohms by reducing the fuse to 4A UL (or 3A IEC) but the test into 8 ohms with lots of clipping will eventually blow the fuse. It will do 100W continuously into 8 ohms with a sinewave OK and it will blow the fuse when shorted. But I find it is worth adding the extra MOSFET's to make it robust.

Calculations for gate resistors

The termination resistors R31 and R32 form part of the gate stopper resistor path and with two MOSFET's in parallel each with gate resistors we can view the two gate resistors effectively appear in parallel and then are in series with a termination resistor.

A complication in the calculations is the p-channel input capacitance is higher than the n-channel. We want both p- and n- sides to have the same bandwidth and we can achieve this by using higher gate stopper resistances for the n-channel. But we also want the p- and n- termination resistances (R31 and R32) to be about the same values.

The p-channel's gate capacitance is about 1.5 times higher than the n-channel so we make the n-channel gate stopper twice the value of the p-channel so when the driver termination resistance is added the totals are in the desired ratio of 1.5 times. For dual-die MOSFET's this means 220 ohm p-gate resistors and 470 ohm n-gate resistors with termination resistors R31 and R32 of 150 ohms.

The termination resistance for the n-channel is padded down by R33 since the n-MOSFET's are on average 10% higher gain than the p-MOSFET's as shown on the data sheets gm curves. This difference is consistent across the various suppliers. R33 is added to trim out any other asymmetry from earlier stages such as the square-law FET driver.

10. Designing the driver stages

10.1 Peak current for the square-law driver

Using the lateral MOSFET data sheet curves at 75°C we need 2.5V on the gates to get 5A peak. The idle current needs 1.25V so we know the current through the 150 ohm termination resistor is 8mA and since the square-law stage doubles the peak current we find the peak current for the square-law MOSFET's is around 4mA at the peaks.

If we know how much current the driver needs to deliver then we can calculate the FET gain. Previously we selected R30,31 as 150 ohms and the lateral MOSFET's deliver 5A peak with an 8 ohm load so we can use the power MOSFET's gm at the peak output current to calculate the gate voltage needed to drive the power MOSFET's and that gives us the driver current.

A data sheet [ref Alfet] for a dual-die lateral MOSFET's carrying 2.5A peak at 75°C junction temperature needs a gate voltage of 2.5 volts. The bias voltage needed to set up the idle current of 630mA is 1.25V so FET's M11,12 need to provide around half the current to get from 1.25V to 2.5V driving into 150 ohms. This is 8.3mA total or 4.15mA from the square-law stage via M11,12.

10.2 Understanding the driver stage

The operation of the Linear Audio Cube-law amplifier circuit which uses two driver stages in conjunction with lateral power MOSFET's can be understood by plotting the gains in open loop.

Figure 10.1 shows plots for an earlier 50W Cube-law Class-A circuit simulation where the top blue curve d($\sqrt[3]{Id}$) shows a good cube-law is obtained over the full Class-A range and the driver gains and currents are plotted in the lower two panes. The d($\sqrt[3]{Id}$) plot is a good cube-law where it gives a flat line since $\sqrt[3]{Id}$ would be a linear ramp and the derivative gives the slope of the ramp.

The overall gain plot (red) shows the Class-A band gain doubles at end of of the Class-A band where the Class-A band ends when the gain of one side reaches zero. In practice the tail end is exponential so the current and gain never gets to zero but an extrapolation of the total driver gain back to zero is one way to define the end/start of the Class-A band (mid pane, red plot) at ±0.60V. The curved part of the knee indicates the exponential nature. This is also the same knee point as the d($\sqrt[3]{Id}$) plot and is a better indicator than the driver gains since d($\sqrt[3]{Id}$) includes the power MOSFET's.

In this simulation a voltage divider was used to scale the input voltage to U4/U5 as a convenient way to trim the tilt of the d($\sqrt[3]{Id}$) plot until it is flat over a wide range of input voltage since this divider ratio does not change the idle current.

There is a ramping up region within the Class-A band before a good cube-law relationship is established. This ramping to does not affect the operation very much because it involves very small currents that are added to very large currents on the other half of the push-pull output stage. What is important for achieving cube-law Class-A over the full power range of the amplifier is the d($\sqrt[3]{Id}$) plot remains flat to the high current end of the Class-A band until one power MOSFET finally turns fully on causing the gain to ramp down to zero (red plot). The rounding at the end is due to the power MOSFET operating in the ohmic region. Incidentally, simulations with the EKV model the change from the current saturation region to the ohmic region (known as pinchoff) is an abrupt point (hence the point of red curve) but in practice pinch-off is rounded and improved models do provide a rounded pinchoff (which helps convergence and simulation speed).

Similarly, there is a ramping up region for the square-law driver before a good square-law relationship is established. This is the subthreshold region of U5. This ramp-up to does not affect the operation very much because it also involves very small currents that are being added to very large currents on the other half of the push-pull output stage. Extrapolation of the square-law to the x-axis gives an intercept that is well inside the Class-A band and this does not cause much of a problem as long as it occurs not much later than where the linear stage reached about 70% of its final linear gain (marked 'X' in the middle pane).

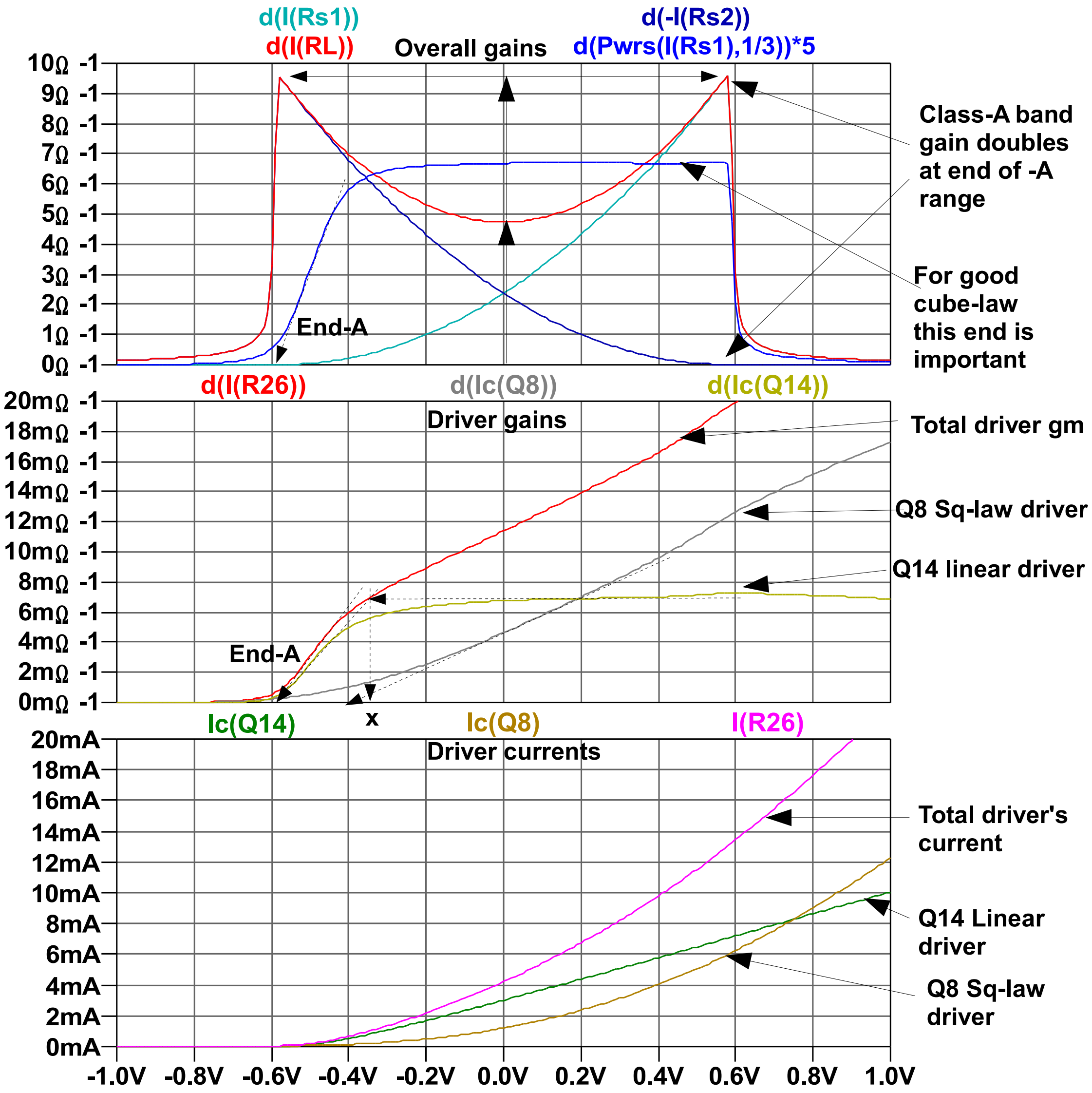

Figure 10.1. Plots for an earlier 50W Cube-law Class-A circuit simulation. It shows a good cube-law over the full Class-A range.

These plots show the main requirements to get cube-law's to give Class-A up to the full rated power of the amplifier. Not included in this plot is how the Early effect upsets things up when the load resistance changes. This is covered in the next plots for the Linear Audio 100W Class-A design where Class-A is achieved over the first 2/3rd of the rated power range.

Figure 10.2 shows similar plots for the published Vol.8 100W version but with the load stepped from 4 ohms to 8 ohms to show what causes the loss of cube-laws where Class-A is achieved over the first 2/3rd of the rated power range and not the full power range as above (Fig 10.1). The end of Class-A is identified from the intercept of the driver gain plot (mid pane) to give 75W (at 40mV input). This is close to the Class-A power for prototype (see Section 1v0 above).

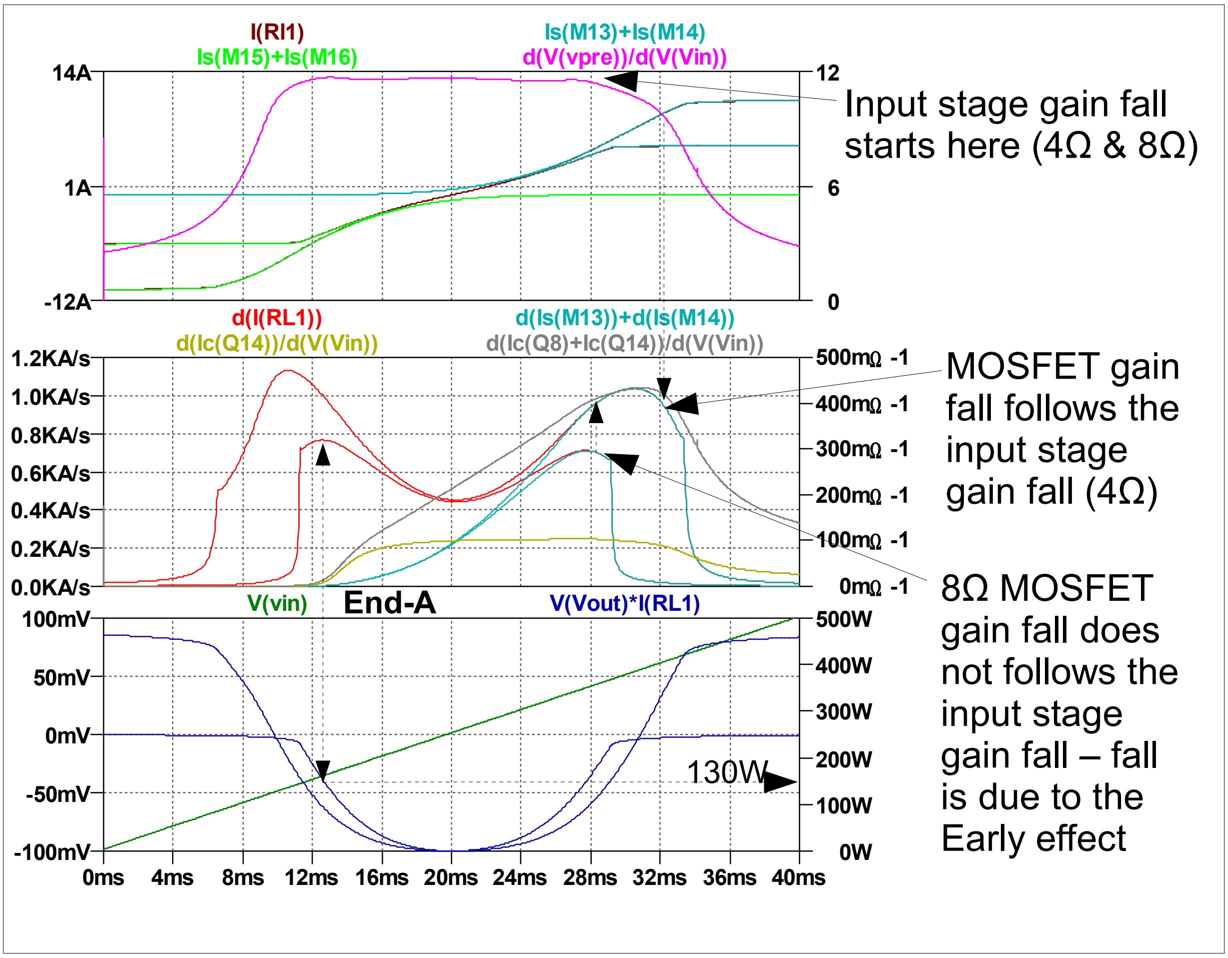

Figure 10.2. Plots of original 100W to see where the loss of cube-laws comes from. With 8 ohms it is due to the Early effect in the power MOSFET's and not the input gain stage. The end of Class-A is 130Wpk or 75W (from the driver gain intercept).

From the top pane we see where the input stage starts to reduce its gain. In the second pane we see that the 8 ohms gain plot increases more slowly than the 8 ohm plot and reaches a peak before the input stage gain starts to roll off. From this we can deduce that it is not the input stage but something else that is related to the load resistance and the power MOSFET's Early effect is the most likely cause. Setting the power MOSFET's Lambda parameter to zero (the MOSFET's Early effect parameter) and repeating the above plots (not shown here) makes the red lines gain rise merge for the 8 and 4 ohm cases thereby proving that the power MOSFET's Early effect is the cause of early gain peaking, departure from a cube-law and departure from Class-A before reaching full power.

Plots in **Figure 10.3** show $d(\sqrt[3]{Id})$ and $d(\sqrt[3.5]{Id})$ plots to see how close to a cube-law this version gives. For 8 ohms the $d(\sqrt[3]{Id})$ covers a slightly broader range for 90% of the top part and for 4 ohms the $d(\sqrt[3.5]{Id})$ covers the top range better. BTW, Figure 10.1 shows an excellent $d(\sqrt[3]{Id})$ alignment.

With the loss of cube-law's at the high current end it is not possible to achieve 1/8th the idle current for Class-A to 100W in the circuit. As was shown in above (Section 1v1) it is possible to stretch the Class-A range just enough to get cube-laws to higher currents to get 100W in Class-A. It was done by increasing the resistance of the emitter resistors (R28,R29) in the linear driver stage to 180 ohms (from 100Ω) and increasing R18,R19 to 1k8 to get enough bias voltage for the linear driver stage with higher value emitter resistors.

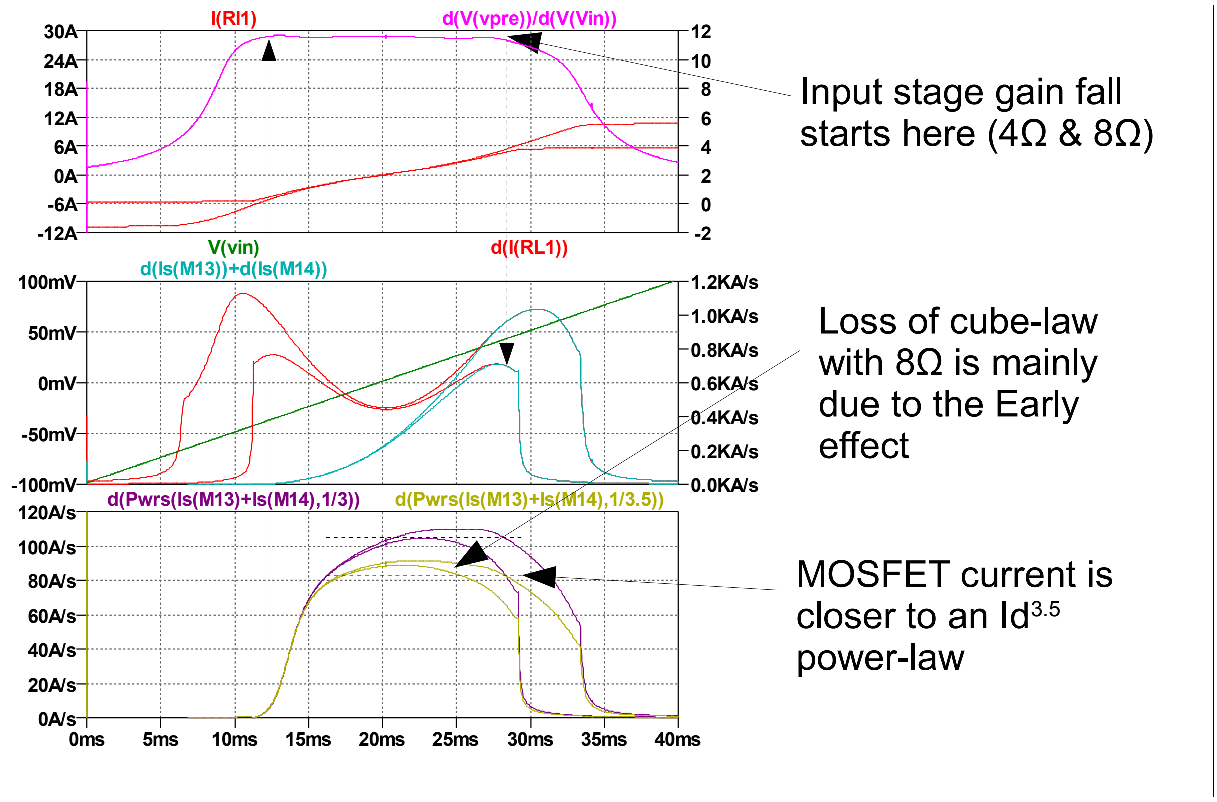

Figure 10.3. Plots of original 100W to check for a cube-law at high currents.

Figure 10.4 shows the same plots as in Figure 10.2 but for the modified 100W version for cube-laws to 100W by widening the linear drivers range with 180 ohm emitter resistors and adjusting the bias to 1/8th the peak current (625mA). The square-law driver settings remain the same.

The peak gain is 739A/s and gm_0 of 377A/s or 1.96 times (cf peak 767 and gm_0 of 443 1.73 in Fig.U9.2). The end of Class-A is at -43mV or 11ms (compared to Fig.U9.2 shows -36mV or 12.7ms).

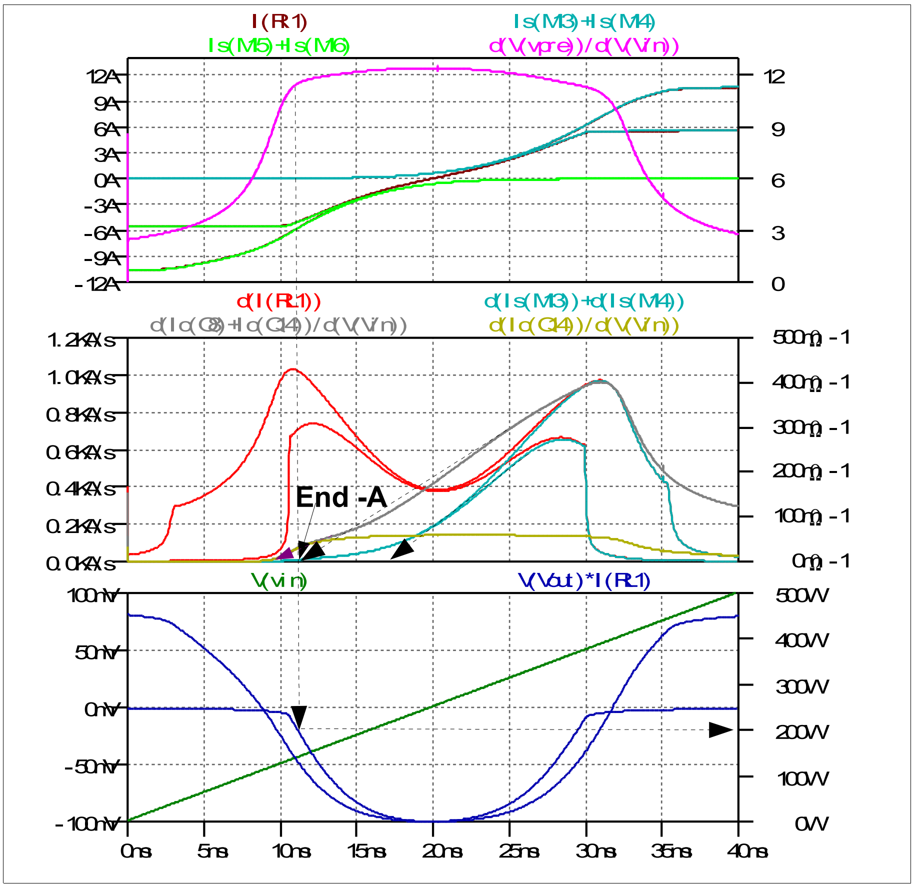

Figure 10.4. Plots of modified 100W for cube-laws to 100W by widening the linear drivers range while the square-law driver settings remain the same. The peak gain is now 1.96 gm_0.

Interestingly the X-axis intercept of the total driver gain does not follow the previous plots because the lead-in is stretched wider. The linear stage intercept is now more distant from the square-law driver x-axis intercept which has not changed. The total driver gain x-axis intercept was created from where the 200 watt peak level passes x-axis and drawn to the peak gain. It was found that the intercept needs to be slightly wider than the 100W-A (200Wpk) intercept since in closed loop feedback compresses the time scale used in this method. An extra 10% is required in these open loop plots to achieve the target of 100W-A in closed loop.

This variation shows that there are a wide range of alignments possible to get full rated power in cube-law Class-A, that is to achieve an idle current for Class-A 1/8th the peak current.

However, some alignments give less variation in THD when the load changes (see below). Optimisation of linearity for a range of loads from 4 ohms to 8 ohms is being studied. It is complicated by the need for good MOSFET models for both the PC stage and the power MOSFET's since the Early parameter plays such a large role in changing THD figures with load. The following demonstrates linearity changes with load with cube-law and square-law alinement's.

10.3 THD changes when driving 4 ohms and 8 ohms

Figure 10.5 shows the linearity (distortion) does not change much when driving a 4 ohm loads and can be almost as linear as an 8 ohm load. The middle pane, top red curve is 4 ohms. This is still cube-law.

This is interesting because the above open loop plots (10.2-10.4) were significantly affected by the Early effect and the PC correction is not upset nearly as much as expected. This was mentioned in the Linear Audio article without any plots or distortion versus power plots.

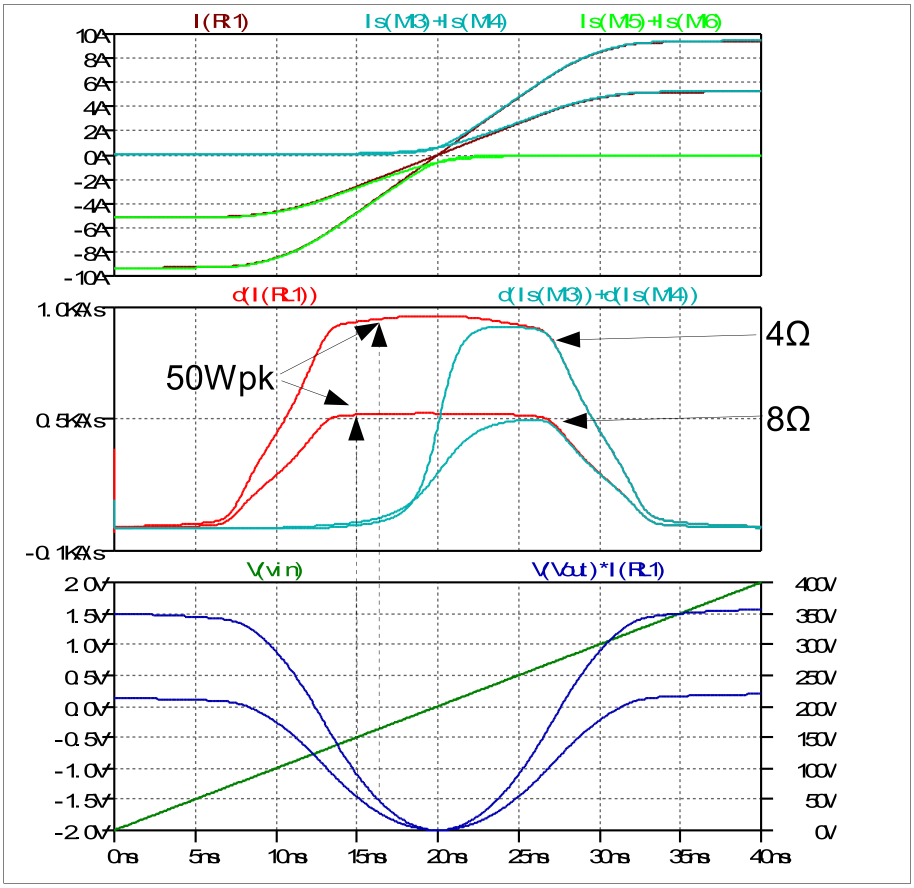

Figure 10.5. Cube-law closed loop modified 100W version showing the linearity when the load is changed from 8 ohms (mid pane lower red) is ΔG/G=1.1% and 4 ohms is ΔG/G=1.1% same.

Figure 10.5 shows points where 50W peak occurs and the gain deviation from the peak (and hence distortion) are similar within a factor of two. As mentioned it may be possible to choose an alignment for the drivers to minimise the gain deviations in the top plateau region. The ramping down gain regions at the ends is due to the input stage and this cannot be altered by the choice of driver alignment.

Figure 10.6 shows square-law closed loop. Linearity is good to 50Wpk (half swing). No PC needed with square-laws because square-law's are inherently linear. The mid pane lower red curve shows 8 ohms gain deviation is 0.01% and 4 ohms gain deviation is 0.04% which is less variation than the cube-law alignment (usually 10 times less linear with 4 ohms).

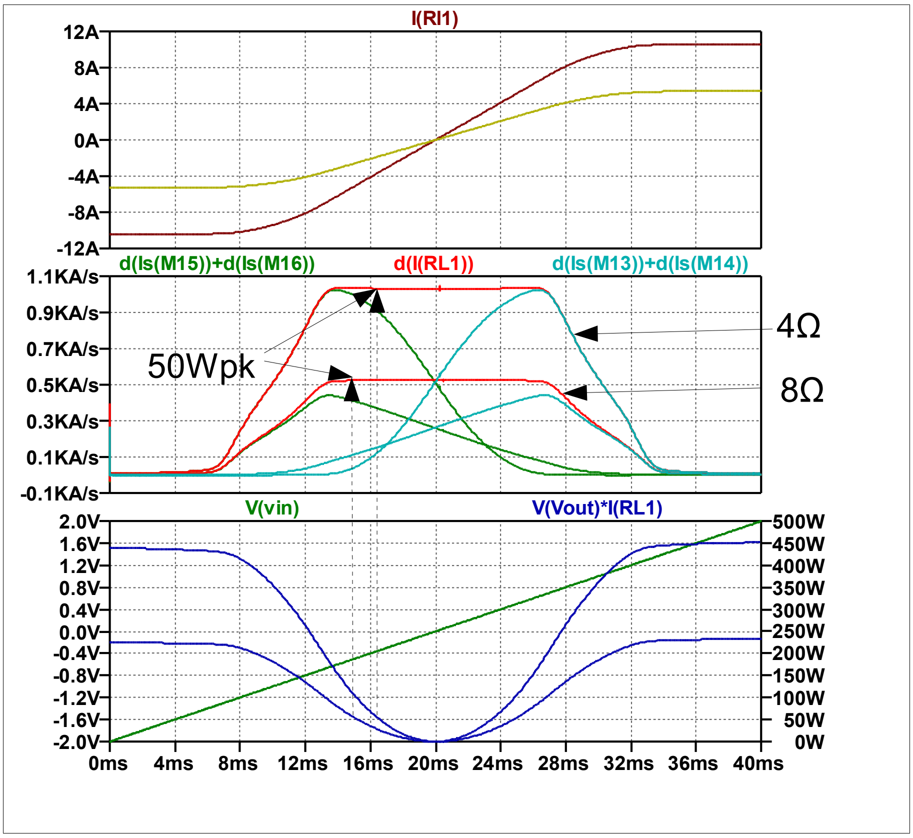

Figure 10.6. Square-law closed loop. Linearity is good to 50Wpk (half swing). No PC needed. Mid pane, lower red curve 8 ohms ΔG/G=0.01%, upper is 4 ohms ΔG/G=0.04%.

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11. Understanding PC Pre-Compensation as a current mirror

Figure 11.1 is the block diagram for a cube-law power amplifier. Error correction is used to linearise the open loop gain as much as possible. Notice the shape of the Pre-Compensation PC stage is the reverse of the output stage distortion.

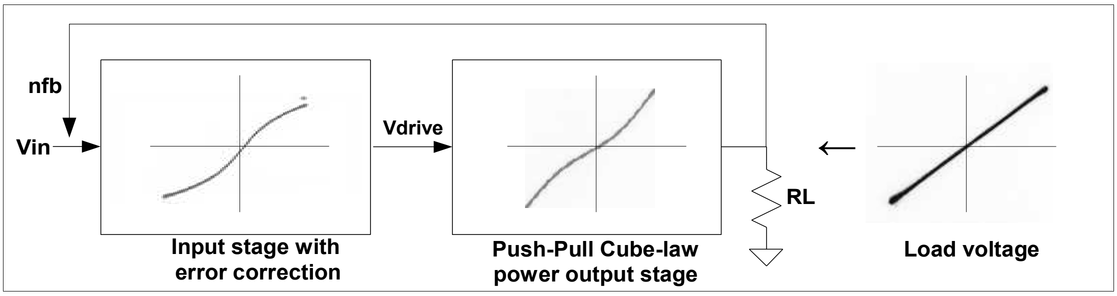


Figure 11.1. A Cube-law Class-A power output stage is nonlinear and error correction can remove the main nonlinearites. There are virtually no high-order harmonics to remove so Pre-Compensation error correction can be fairly effective.

Global feedback is used to further reduce distortion and to give an acceptable output resistance to drive loudspeakers.

The following circuits and plots show how the PC stage uses local feedback in a similar way to a basic current mirror generating a very linear output from two highly nonlinear gain devices and does not use any overall negative feedback. In the PC the mirror action takes place in complementary form, but the basic principle is the same.

Figure 11.1 shows a basic current mirror. An input current is fed to the input MOSFET and this generates a voltage proportional to the inverse of it's transconductance since the gate is tied to the

Back to Contents

drain, effectively giving a current to voltage conversion. This voltage drives the output FET and the output current is linear with the input signal (Iin) if the two MOSFET's are the same. Although there is local feedback around the input FET a current mirror is said to operate in open loop since no *overall* negative feedback is applied from the output of the second FET.

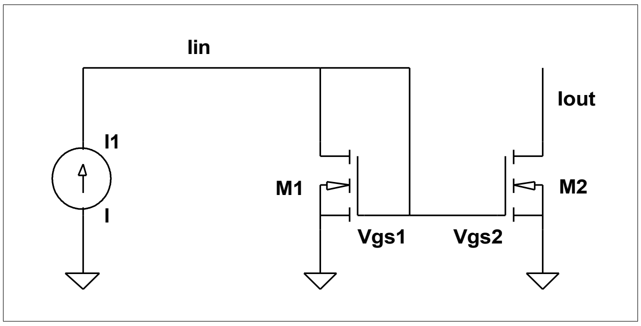


Figure 11.1. The current mirror principle with FET's. First M1's gate voltage rises until Id1 equals Iin, then Vgs2 generates Iout which is a linear version of Iin.

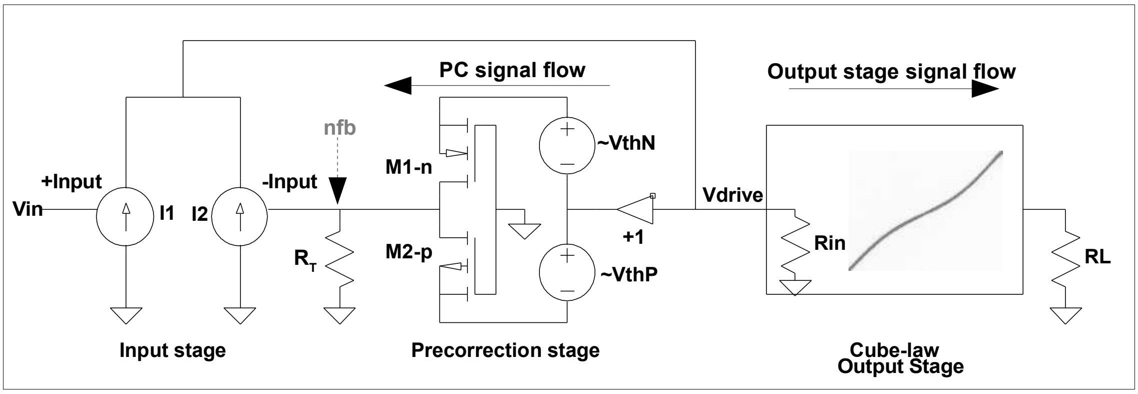
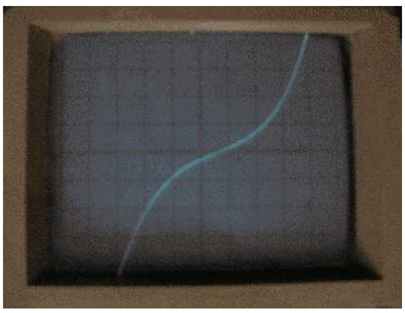


Figure 11.3. Current mirror principle applied to a cube-law output stage.
FET's M1 and M2 with I2 form a local loop (like M1 in Fig.B1).
The pre-compensated drive voltage gives a linear output current in the load
It gives linearity in open loop – no overall feedback is used.

Now to the PC stage. **Figure 11.3** shows the current mirror principle applied to the cube-law output stage. The PC stage is shown as a buffer driving MOSFET's that are biased at their threshold voltages and their output current drives into a termination resistor R_T (R5 in Figure 4). Then local feedback via the input stage inverting input inverts the current due to the input signal (I1). The resulting inverted voltage (Vdrive) is fed to the output stage and this makes the overall gain linear in open loop. At this stage no global feedback is applied from the load.

Operation of the PC stage is demonstrated using several scope screen photos using Figure 4 circuit. **Figure 11.4** shows the output voltage into R5 (330 ohm) and plotted against the input drive voltage to the input stage (Vin).



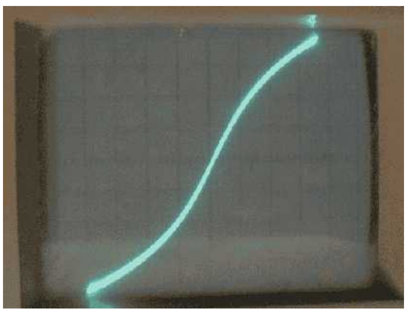


Figure 11.4. The PC stage voltage across R5 330 ohm against Vdrive. It looks similar to the output stage (Fig.B2). X is 100mV/div (same as Fig.4) and Y is 100mV/div.
Figure 11.5. Now with local feedback, plotting drive voltage against Vin. It gives a cube-root

like plot at the ends and fairly linear in the middle range. X is 10mV/div, Y is 100mV/div.

This trace was generated by breaking the inverting input connection to common and the X input is connected to Vdrive and with the input stage is driven with Vin. The input voltage swing used is the same as the swing needed to drive the output stage to the start of clip with an 8 ohm load which corresponds to the same 4 divisions of the x axis.

Figure 11.5 shows the drive voltage after local feedback is used to invert it. It gives a cube-root shape at high input levels which is what is required for a cube-law output stage.

The most convenient way to set up the PC stage to get the best overall linearity is using R12 or R13 to alter the gain of the PC stage rather than R_T (R5). The global feedback resistor (R9) can be chosen for the amount of global feedback to give soft clipping without upsetting the PC stage linearisation. It also conveniently allows the PC stage to be broken while the amplifier is running so the difference with and without the PC stage can be seen and heard.

Finally, **Figure 11.6** shows the voltage across load with PC applied and no global feedback. Compare this to Figure 4 for the cube-law output stage. Linearisation occurs over the first half of the output swing. Over the last half "S-curve" soft clipping occurs.

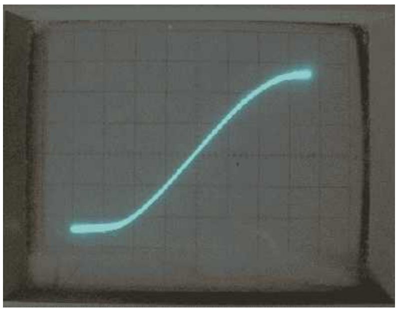


Figure 11.6. The load voltage in open loop with PC. A listening test at 50Hz reveals distortion is starts to be audible at around half full output swing or 12 watts.

This rounding can occur from two places in this circuit. The input stage M9,10 rounds the input current (I1 in Fig.11.3). Or soft clipping can came from the PC stage M11.22 where the PC generated curve (Fig.11.4) increases faster than it should for linearising the output stage. When the maximum amount of global feedback is used the input stage is used to control the clip level for normal loading. With low resistance loads the PC stage can dominate soft clip. It is a fairly complex process but which ever way causes the clipping we get about the same soft rounding shape. ... [more sometime?]

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12. On Error Correction

Malcolm Hawksford the developer of Hawksford Error Correction (HEC) said,
"You often see that when the distortion is driven down by feedback or EC, *it works for the first few harmonics at the expense of increasing higher harmonic components*. Again, just like what we observe with noise shaping in digital systems!" link p25 mid. col. end of 1st para. *Emphasis* added.

It is almost impossible to use a very simple error correction circuit to linearise complex linearities created by bipolar and MOSFET's in power output stages. Although bipolar's are modelled primarily as exponential with base-emitter voltage and FET's are modelled primarily as square-law in the gate to source voltage we find real devices are far more complicated because they all have peculiar second order effects added to these primary effects and they are all affected in different ways by temperature changes in the power devices. Error correction, including HEC, aims to correct the largest nonlinearities of an output stage (and usually at one operating temperature) and they leave the smaller parts of the nonlinearities either unaffected or made worse.

The vexing thing in audio amplifiers is that our hearing is much more sensitive to very small amounts of high-order nonlinearities. EC circuits appear to reduce the THD by a good amount such as 10 times or -20dB but when you listen the distortion is still audible. Oh HECk, Verflixt!

It annoys designers because what looked like a good tool for designing high quality amplifiers turns out to be either useless, or worse, a backward step with more parts and yet another added trimpot to tweak, and a possibility that the EC may misbehave when clipping, and worst of all no improvement in sound or even the dreaded “sterile sound” review that was given to the Halcro DM58 review by Martin Colloms (HFN June 2004 p60-63).

That review for the lowest distortion audio power amplifier in the world sent ripple through the design world. We are still asking, “where on earth did he get it wrong?” And “He” could refer to the designer Bruce Halcro Candy (BHC) or the reviewer Martin Colloms (MC). MC it was rated “a little below average” (p59) even though its distortion and most other specifications were the best ever.

Bruce Candy (see patent) uses most of the tricks in the power amplifier design books for all 3 stages: the input stage, the VAS, and output stage; with lots of local feedback and minimal global feedback;with lateral power MOSFET's (I think) using HEC correction similar to Bob Cordell's benchmark 1984 design (view here).

From the previous information we can surmise that HEC may be *one* reason why it did not get the good review expected after doing everything 'according to the book'. As mentioned, HEC lowers THD readings but leaves the high-order distortion either untouched or sometimes the high order harmonics are increased in level, and recall it is mainly the high-order harmonics that we hear. Although HEC can give outstanding THD improvements it can be counter productive in terms of reducing the distortion bits that we actually hear, so in general HEC does not make it sound better. Oh HECk!

The Halcro DM58 also uses sophisticated bootstrapped cascoding (with higher auxiliary rails and not electrolytics) and complexity brings the likelihood of additions problems, particularly with overload recover after clipping. Therefore auditioning will depend a lot on on speaker sensitivity because this affects the amount of clipping to show up normally unseen clip overload recovery problems. Reviewers must be aware of this they need to monitor whether clipping takes place or not. Otherwise an unscrupulous reviewer could choose a speaker efficiency to either give the best possible review or conversely to get a mediocre review.

There are other possible reasons why a good amplifier might be given a mediocre review. Outstanding lab test results like the DM58 raise the expectation of the reviewer and if their listening experience is about the same other amplifier with ordinary specifications then there is disappointment and this may cause the reviewer to give it a mediocre review below the others without realising their mistake. A reviewer may believe that the listening experience gets better and better no matter how small the improvement, eg, reducing THD from 1 ppm to 0.1 ppm like the DM58 should sound better, but this ignores physical limits of resolution of hearing where there is a level of distortion below which no amount of listening time can reveal an audible difference even though our lab equipment can 'hear' a difference. A reviewer who believes there is no hearing threshold will be disappointed when reviewing a DM58 because the lowest possible resolution for hearing distortion is 3 ppm and the DM58 and several other modern amps goes well below that, and knowing the test results before a listening review could create disappointment and a biased review. The 1978 listening tests by James Moir and friends [here] showed that most power amplifiers sounded the same when they never clipped with THD readings below 0.1% (1,000 ppm), but that was later revised down to 0.01% (100 ppm) to cover some transistor Class-B amps with higher than usual levels of high-order harmonics (and that would include HEC amps).

Yet another problem with reviewers is normalisation by only hearing distorted music with distortion above the hearing threshold. When played music with distortion below the hearing threshold it sounds unnatural. It is necessary to train (or retrain) our to an accurate reference to learn what to listen for.

Back to the HEC problem. Cube-law Class-A and Cube-law Class-A can help design. In Cube-law Class-A and Cube-law Class-AB we do not generate very much high-order wiggly gain bits anywhere in the output current even up to much higher currents than we need to drive the normal load impedance. If we don't generate high-order stuff like standard Class-B then we only have low order stuff to correct – and we *can* do that reasonably easily using simple circuits. In the Linear Audio Cube amp this is done using a simple Pre-Compensation stage. This stage is explained in a section in the "Upgrades" companion document (from same link as this document).

13. Simple two stage power amplifiers

Malcolm Hawksford in the interview quoted above also said,
"*...It's better to go for a simple system, with as few stages as possible because an additional stage cannot fully undo*."

John Linsley-Hood, Nelson Pass, Jean Hiraga, Erno Borbely, Michael Renardson, Bengt Olsson, myself and others have been designing power amplifier that are as simple as possible, with as few stages as possible. Many of these designers examples appear in this section above, eg Fig's 3.4 to Fig 3.7. Bengt Olsson in a EW Letter said with only two gain stages that compensation was the easiest he experienced in his 30 years of amplifier design. It was able to drive any capacitive load without the usual inductor in the output speaker line and that is very rare … except in two stage designs.

The technical reason for the easy of compensation (if it is needed at all) is two stages can only contribute a maximum phase shift of 180 degrees so the unity gain frequency will always have less than 180 degrees and so it can never oscillate and go unstable (and in bipolar designs that usually means the end of the power transistors and a good channel it will take the drivers and VAS stage as well). In practice two stages can contribute more than 180 degrees due to parasitics but when some voltage gain with simple two stages is used then there is less chance that there will be enough gain around the feedback path to oscillate at the 180 degree frequency. And if there is too much ringing then it is usually possible to add a capacitor somewhere for some simple compensation.

Michael Richardson has reverted his power amp topology to more like the original two-stages Lin topology. The original two-stages Lin uses one stage for what we now call the VAS and the other stage is quasi-complementary output stage.

Michael Richardson (new site) has left off the differential input stage that was added to the original HC Lin topology, eg, Michael Richardson's earlier info at www.angelfire.com/ab3/mjramp/ and see Footnote 3 for discussion on alternative compensation methods. Without the input stage we have a return to the early single rail power amps with an output capacitor in series with the loudspeaker. A bridge arrangement can eliminate the output capacitor but both speaker terminals must floated (that's not normally a problem but you need to remember never to connect the speaker leads to the amps common!) Using output capacitors may seem like a backward step but it has the advantage of being easier to compensate. A turn on thump is avoided by slowly increasing the output voltage at turn on.

Without the differential input stage the phase shift around the feedback loop is reduced making compensation easier and it now allows capacitive loads to be driven without sacrificing bandwidth to cater for highly capacitive loads like electrostatic loudspeakers. It gives remarkably stable and high bandwidth amplifiers that can easily drive any capacitive load you throw at it. The next section mentions an alternative compensation scheme for the 3 stage topology that can achieve a similar result without deleting the input stage of the standard 3 stage topology.

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14. Alternative compensation

The standard 3 stage Lin topology usually uses Miller compensation C_{dom} across the VAS. Simulations using the 3 stage Lin topology with Miller compensation reveals a problem driving capacitive loads. C_{dom} needs to be set to give adequate margin of stability into a the largest likely load capacitance and when driving ordinary loads the compensation is much too large. What is needed is compensation that adapts to load capacitance to give a wider bandwidth with ordinary loads but reduces the bandwidth when driving large capacitive loads.

An alternative compensation method used by John Linsley-Hood (examples Jul 1970 Fig 3, Letters Sep 1973, Aug 1982 Fig 14, Mar 1989, Nov 1989 Fig 17 view) and possibly Marshall Leach (view) allows large capacitive loads to be driven without oscillation and still retains a high bandwidth when driving more common low capacitive resistive loads of standard moving coil loudspeakers. In other words the amplifiers bandwidth reduces adaptively to load capacitance with the square-root of load capacitance without ever oscillating no matter how large the load capacitance. This is a better type of compensation for an audio power amplifier.

Discussion at DIYaudio forum 2010 here shows this type of compensation has not been thoroughly explored. Douglas Self who calls it "Inclusive compensation..." in Linear Audio Vol.0. Bob Cordell calls it "Miller input compensation" (Book p). Marshall Leach Jr. calls it “feedforward compensation”. John Linsley-Hood calls it “the loop stabilising capacitor across stages 1 and 2” (WW Sep 1973) and in a March 1989 amp (p263-4 view) with compensation by C10 says:
“C10 is all that is needed to provide an adequate gain and phase margin in the feedback loop; C10 is employed in a position which ***greatly*** lessens the tendency to slew-rate limiting, in contrast with the more conventional and less satisfactory technique in which C10 would be connected to the gate of Tr7 [VAS] to provide a 'dominant-lag' [Cdom Miller] type high frequency compensation.” [*emphasis* added].
John Linsley-Hood's November 1989 summary article (view) simply refers to it as “the preferred position for HF loop compensation capacitor.” We are still waiting for agreement on what to call this type of compensation. What about “the best compensation”?

Miller compensation is not a very good choice for the 3 stage Lin topology but it is common because it is simple and reasonably effective only for loads with little capacitance. Because Miller compensation does not encompass the input stage it very susceptible to additional phase shift from the output node which makes it hard to compensate for large capacitive loads like 1uF from electrostatic loudspeakers.

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15. Soft and hard clipping in power amps

A simulation of a 50W cube-law Class-A amplifier was done with a 50Hz sinewave input with soft clipping. To emulate hard clipping the circuit was modified by adding a small-signal diodes from the drain to gate of each power MOSFET. This clamps the gate voltage when the MOSFET's drain voltage falls below the gate voltage which occurs near the peak output voltage.

The power is shown in **Figure 15.1** show the normal soft clip where rounding starting from about 70 watts and reaching 100W peak. **Figure 15.2** shows the hard-clip which starts at 90 watts and limits at 92 watts. In both cases the average power is the same 50 watts.

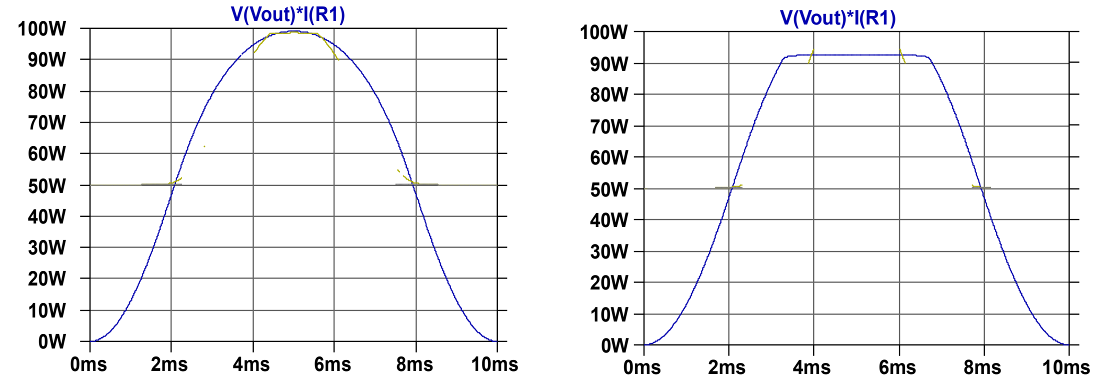


Figure 15.1 Left. Standard soft-clip.
Figure 15.2 Right. Hard-clip (right).

Figure 15.3 shows the soft-clip weighted harmonic spectrum. Weighting shows high-order harmonics in the 1-5kHz range will not be audible and are masked by the lower order harmonics. The dotted line is the hearing threshold for weighted harmonics (for average hearing) with a sinewave.

Figure 15.4 shows the hard-clip weighted harmonic spectrum and high order harmonics will be audible making the sound harsh and unpleasant. With soft clipping the distortion is mellow.

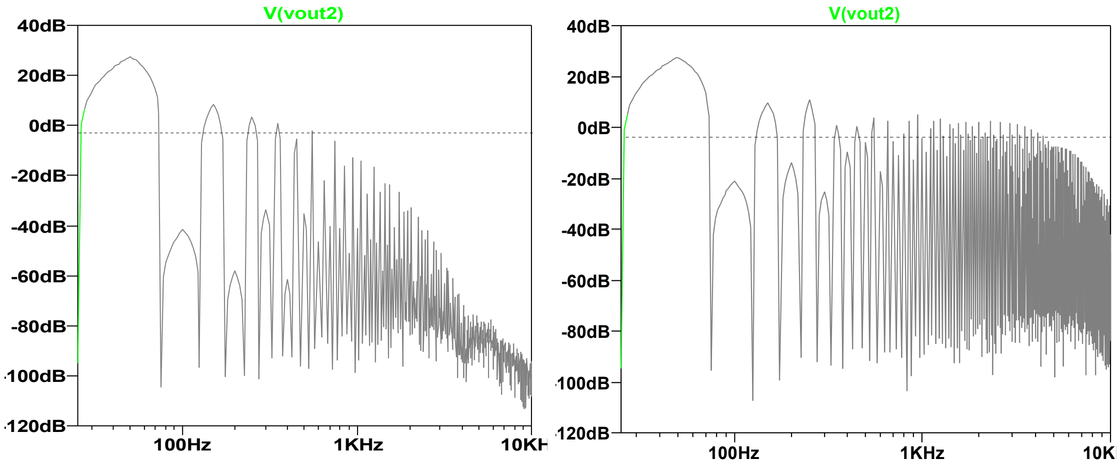


Figure 15.3 Left. Soft-clip weighted harmonic spectrum.
Figure 15.4 Right. Hard-clip weighted harmonic spectrum.
With hard-clip high order harmonics in the 1-5kHz range are audible.

Most transistor amplifiers clip harder than this example and with higher levels of high order harmonics and a harsher sound when clipping. This example looks close to how the Williamson valve amplifier clips [here] which shows a moderately hard-clip. But transistor amps clip even harder than this with more high order harmonics than shown above.

What sort of clipping and harmonics do we get from over driven valves (tubes)? **Figure 15.5** shows the unweighted spectrum for a common cathode 12AX7 triode over-driven (here Fig.12). Weighting boosts harmonics by about 50dB/decade so clipping here can be considered moderately hard-clipping like the previous simulation with diodes. It also shows the cube-law amp's soft clipping is significantly softer and mellower clip than used in typical tube guitar overdrive and the cube-amp is also a symmetrical clip with very little even harmonics.

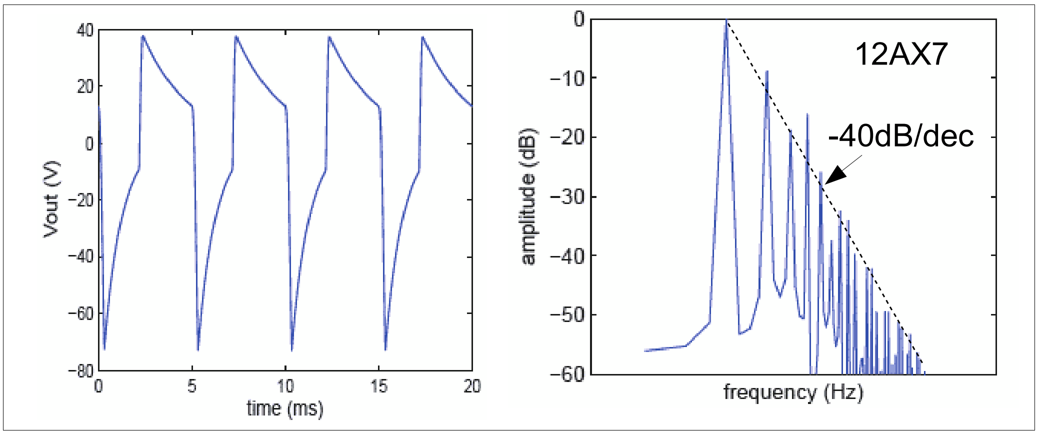


Figure 15.5. Output waveform and unweighted harmonics from an over-driven common cathode 12AX7 triode (Cohen AES-SF 2010).

Variable clip potentiometer

The soft clip level can be varied over a 100:1 power range by varying R8. The PCB includes a connector to allow a dual-gang 50k log potentiometer to be mounted on the front (or back) panel. One reason for making it available to adjust is for low mains voltages. Another reason is so the amount of clipping can be reduced until clipping distortion becomes just noticeable and this gives an idea of how much headroom is available for the speakers being used and the music used. It is also possible to work out how frequently the clip LED can flash at before clipping becomes audible. It is also useful when testing someone else's expensive speakers if your power amp is far too powerful. Placing a trimpot inside the amplifier can stop someone else from exceeding the limit.

The remote clip level potentiometer can also be used for guitar amplifier "power-scaling™" a trademark of London Power Ltd (here and here). It allows a powerful amplifier to be used at home without annoying others too much by limiting the maximum output power to 1 watt or less.

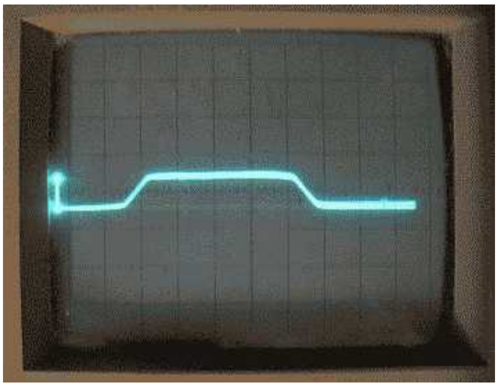


Figure 15.6. Soft-clip with R8 of 15k. Input 1Vpk, output 3Vpk (2Wrms). 5V/div.

Figure 15.6 Shows clip with R8 of 15k. The input voltage is 1Vpk sinewave which would give nearly full output power (50W) if the clip pot was set to zero but set at 15k we get an output of only 3Vpk or 2Wrms. The clip is not as soft as with full output and the waveform is more of a trapezoid than a rounded over sinewave.

Ripple free soft clipping

This clip level can be used to stop power supply ripple getting through to the output when clipping. It does not matter how much negative feedback is used in your amp because during clipping the loop gain is wiped out by hard saturation in the output transistors.

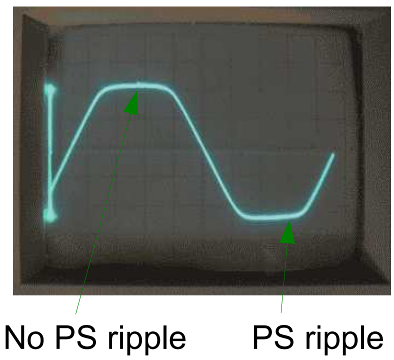
With hard clipping the power supply ripple substitutes for the signal that the power amplifier should be providing. A typical power rail might have 1V RMS of ripple which for a 50W tone would be a level of 5% at twice mains frequency. Tests were done to try to find out how audible this is (results are given later).

Some guitar power amplifiers include a circuit to keep out the ripple during clipping called a “sag circuit” [ref]. A simple way to keep power supply ripple out is to use a capacitance multiplier such as the MOSFET eg Andrea Ciuffoli (EW May 2000) here & here since it adapts to power supply voltage changes, unlike a true regulated supply which wastes lots of power during normal mains voltages.

Soft clipping must be applied before the power amps negative feedback loop and after the volume control. If it is done within the feedback loop then the output current is limited since the output stage is a transconductance stage. In this amplifier the input stage clip just before the output transistors saturate. This design does not adapt to supply variations for simplicity. It can be manually trimmed using the clip level pot (R8).

For my power supply using a 180VA toroidal transformer and 10,000uF capacitors. 30.2V-26.3V=3.9V and currents were 3.55A and 6.0A. The equivalent power rail resistance is $\Delta V/\Delta I$ ohms or 1.6 ohms. When the output resistance was reduced to 3 ohms (by changing the feedback resistor) until the ripple was only just locked out for 8R *and* 4R. This corresponds to an output resistance that is is twice the power supply resistance for one rail to common, or alternatively, the same as the rail-to-rail resistance. If the output resistance is less than this then the clip level needs to be set to lock out ripple with a 4 ohm load. The 100W version with a 500VA transformer had a rail-to-rail resistance of about 1 ohms – about the same as the amplifier's output resistance.

Figure 15.7 shows soft clipping with 20dB of NFB and R8 shorted. The top is ripple free but the bottom includes ripple due to clip asymmetry from the CMOS input stage. Clip asymmetry can be trimmed by RV2 to minimise the clip overhead and maximise output watts. RV2 adds a small dc offset to the common side of the clip level pot (around 50mV) to offset the difference in p-and n-threshold voltages in the CMOS input stage and this voltage can be derived from the 12V rail as with a voltage divider as shown in the Upgrade circuit (1v1).



[XY plot for this test]

Figure 15.7. Clipping with 20dB of NFB and R8 shorted.
The top is ripple free but the bottom includes ripple due to clip asymmetry from the CMOS input stage. Clip asymmetry can be trimmed to minimise the clip overhead and maximise output watts.

Test observations with soft clipping

One characteristic of valve (tube) amplifiers identified by Jean Hiraga (here) is the distortion (THD) is reasonably low at normal listening levels and rises steadily as the power increases on a Log-Log plot until clipping proper starts giving a rapid rise in distortion with power. Transistor power amplifiers give a different distortion profile with increasing power, that is, very low distortion (THD) at normal power listening levels and distortion (THD) usually remains very low until just before clipping and then it rises very rapidly with power.

If tube amplifier's give a more desirable sound and if it is mainly because of the way the distortion rises before clipping and not as fast during clipping due to soft clipping then the questions is how far up the power scale does soft clipping need to start to give a solid state amplifier the same desirable quality?.

An interesting observation is that *a slower broader clip curve* is better than a faster sharper one because the slower one generates mostly lower-order harmonics and less of the higher-order harmonics that are so easy to hear and objectionable.

One reason why designer's shy away from soft-clipping is the distortion starts to rise due to soft clipping from half full power or less and this can look bad on a THD versus linear power scale. Bob Cordell said, “When it comes to introducing soft clipping, some are reluctant because they have worked so hard to get THD vanishingly low right up to full power, and then with the addition of soft clipping distortion will begin to rise into the 0.01% or 0.1% range even at maybe half power.”

16. Open loop listening tests with 50Hz

During the development of this amplifier a dummy load and a small monitor loudspeaker was used with a 50Hz sinewave to listen for audible distortion in open loop then later in closed loop with 20dB of negative feedback. At the same time the linearity was monitored using an oscilloscope used in X-Y mode. For these tests the input sinewave did not need to be very low distortion because the X-Y are used. The PC stage was switched in and out while running with feedback to check how much change in distortion there was in open loop.

These tests were done with an 8 ohm 200W dummy load. The small speaker was an 8 ohm transistor radio 50mm 0.5W speaker with a dropper resistor of 220 ohm 2W and a 470 ohm 2W all in series with the speaker. The speaker sat on the box it came in. A bipolar 470uF/50V capacitor was also included to stop DC. (Adding two back-back 5V 1W zener's across across the speaker terminals would have been helpful since the speaker out the smell of burnt paper). This small speaker is rolls off from around 300Hz and this makes it more sensitive to the 5thand higher harmonics and gives a slight weighting filter effect. If the distortion was not able to be heard with this test rig then it was unlikely to be audible with music and full-range speakers.

A 50Hz sinewave was obtained from the transformer's secondary after passing it through a 2-stage RC filter, sometimes 3 stages for critical listening tests. With a 10k volume pot I used two 10k resistors and two 3u3 bipolar caps. The secondary waveform was heavily chopped off at the top and after filtering was about 2% THD, and 0.1%? with a 3 stage filter. The advantage of using a mains derived sinewave is that the clipping is phase locked unlike a signal generator that will drift and roll one way or the other and make it hard to take photos.

A spectrum analyser was not used until after the prototype was built on a PCB designed for me Jan Didden. The analyser was Christian Zietnitz PC 'oszilloscope' (here). X-Y plots were also possible with the PC scope with an external loop-back trimpot. X-Y plots were copied to the clipboard (using Print Screen) and the points captured for plotting in a spreadsheet by using Steven Benbow's free ‘Graph grabber’ software here. Some spectrum plots are included in the Linear Audio article Vol.8.

Christian Zietnitz PC 'oszilloscope' could resolve to around 0.03% THD and the amplifiers distortion fell below this below about 10 watts. The clip circuit output voltage “Vdist” can also be used as a distortion magnifier for higher resolution of distortion and X-Y transfer curves. No spectrum plots were done using the distortion magnifier but simulations suggest the distortion falls to around 0.001% (~10ppm) at 1 watt after the PC and symmetry are trimmed.

Test observations – to use PC or not PC?

One of the questions about this topology is whether the extra PC stage does enough good to use it. The bottom line is whether it improves the listening experience. It has the ability to add or remove most of the odd harmonics at normal listening levels and it can also reverse the phase of the odd harmonics by over or under correction. It provides a lot of scope for experimenting and that's the main reason it was included in the Linear Audio Vol.8 amp.

The Linear Audio Vol.8 amp has enough negative feedback to make the distortion inaudible at normal power levels without the PC enabled. The PC stage when engaged can further reduce distortion at moderate power such as 1 watt down to the 0.001% region which makes the amplifier look quite good in the eyes of those who want Ultra-Low Distortion readings.

The PC stage under some conditions when clipping can create a gain step that is easily heard. But it can be controlled by adding R7 (22 ohms) in the feedback circuit. The resistance was chosen using a trimpot while looking at the X-Y plot of the clip distortion output. With 22 ohms the changeover to soft clipping (at about half output swing) becomes a smooth knee er than a sharp knee [screen shot not taken].

Test observations – effect of 2nd and 3rd harmonics phase

The phase of the second and third harmonics creates a different sounding amplifier in designs where distortion is high enough to be audible such as low feedback amplifiers [Pass 2012 BAS 2013 PDF p12, audio 32 min].

Table 12.1 and **Table 12.2** show the phase relationships for the second and third harmonics for different gain transfer curves typical of Class-A amplifiers. A rising gain means the gain increases slightly with increasing input signal (and decreases with decreasing input signal) – this gives rise to mainly 2nd harmonics (so the 3rd harmonic is not listed in Table 12.1).

Table 12.1. Phase 2nd harmonic for different gain functions

	Gain Rise	Flat	Gain Fall
2nd	-90	0	+90

Table 12.2. Phase 3rd harmonic for different gain functions

	Bow Up	Flat	Bow Down
3rd	+180	+90	0

In Table 12.2 an *upward bowing* gain means the gain increase is for both rising and falling input voltage. A *downward bowing* gain means the gain decrease is for both rising and falling input voltage. Gain plots for these 4 cases are shown below as **Figure 12.2** to **Figure 12.5**. The first plot **Figure 12.1** is for very little gain change gives the "Flat" case.

In many of Nelson Pass' designs the distortion becomes audible at high power levels and he prefers the sound of distortion from an amplifier with the *decreasing gain function*.

Interestingly, the the 3rd harmonic in Cube-law Class-A is opposite phase to that of square-law Class-A causing the current to bend *upwards* near peaks, whereas in square-law Class-A the current bends downward near the peaks. This upward bowing can be reversed using PC. A test can be done where the PC is adjusted to reverse the bowing and then the PC link can be opened and closed while listening to music or a sinewave to hear the effect.

When I was doing closed loop tests on my Cube-law amp I was able to continuously alter the shape of the gain function from increasing to decreasing and I prefer the increasing gain function. But I referred it more when there was minimal gain change, a flat gain curve, with very little added distortion. To my ear the 3rd harmonic distortion from Cube-law Class-A in open loop sounds *like* the difference between a major and minor chord – major (better) for decreasing gain function and minor (worse) for an increasing. This is an analogy only because unlike major and minor chords this example both have major 5th's (here) – the only difference is the phase reversal of the 3rd harmonic.

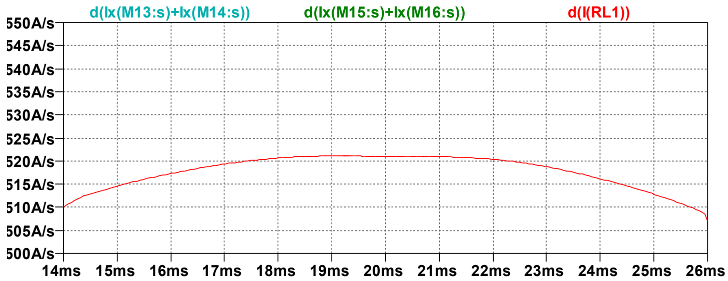


Figure 12.1 Reference Case 1 -- minimum distortion

1v1 x=512m, y=402m, bias=415m, Idle=624m Ios=6.6mA, gain slope -100.

Harmonic Number	Frequency [Hz]	Normalized Component	Normalized Phase[deg]
1	5.000e+01	1.000e+00	0.00°
2	**1.000e+02**	**4.660e-06**	**25.41°**
3	**1.500e+02**	**5.748e-06**	**98.75°**
4	2.000e+02	1.882e-06	88.03°
5	2.500e+02	1.553e-06	-178.82°
6	3.000e+02	1.113e-07	77.37°
7	3.500e+02	4.746e-09	171.31°
8	4.000e+02	3.586e-08	2.24°
9	4.500e+02	3.967e-08	1.84°

100mVin **Total Harmonic Distortion: 0.000779%**

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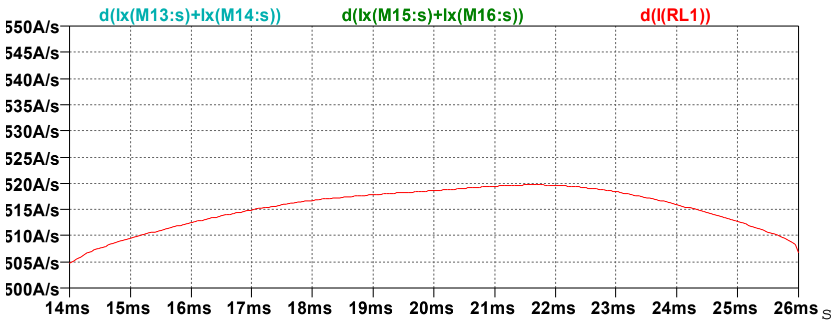

Figure 12.2. Case 2 – Positive gain slope

1v1 x=512m, **y=350m.**Gain slope=+880

Harmonic Number	Frequency[Hz]	Normalized Component	Normalized Phase [deg]
1	5.000e+01	1.000e+00	0.00°
2	1.000e+02	**4.611e-04**	**-88.86°**
3	1.500e+02	1.004e-05	140.02°
4	2.000e+02	1.378e-06	-89.86°
5	2.500e+02	2.003e-06	-179.15°
6	3.000e+02	5.691e-08	83.72°
7	3.500e+02	4.226e-08	175.58°
8	4.000e+02	1.729e-08	1.32°
9	4.500e+02	1.901e-08	18.52°

100mVin **Total Harmonic Distortion: 0.046120%**

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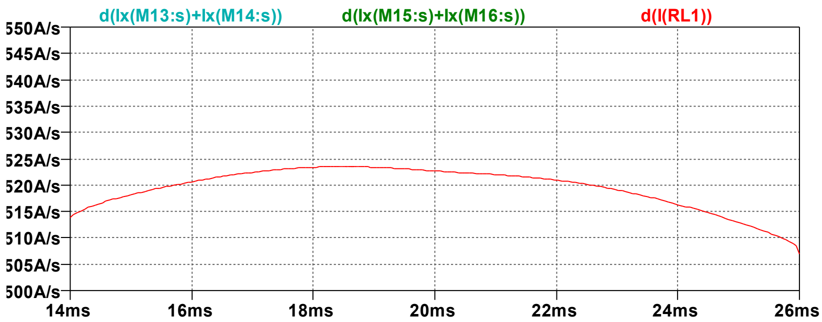

Figure 12.3. Case 3 – Negative gain slope

1v1 x=512m, **y=450m.** Gain slope=-703

Harmonic Number	Frequency[Hz]	Normalized Component	Normalized Phase [deg]
1	5.000e+01	1.00E+000	0.00°
2	1.000e+02	**3.03E-004**	**89.82°**
3	1.500e+02	7.20E-006	46.22°
4	2.000e+02	3.65E-006	88.49°
5	2.500e+02	1.25E-006	-178.99°
6	3.000e+02	1.24E-007	77.44°
7	3.500e+02	8.66E-009	10.62°
8	4.000e+02	3.83E-008	7.15°
9	4.500e+02	4.46E-008	1.75°

100mVin **Total Harmonic Distortion: 0.030337%**

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Case 3 = +slope: 1v1 **x=612m,** y=350m

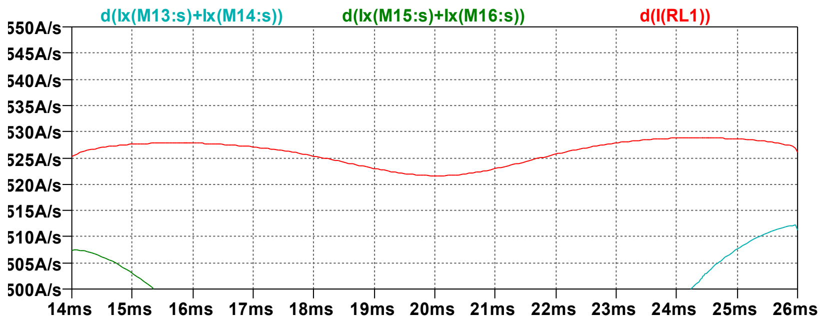

Figure 12.4. Case 4 – Upward bowing gain

slope=-1400(L),+1370(R)

Harmonic Number	Frequency[Hz]	Normalized Component	Normalized Phase [deg]
1	5.000e+01	1.00E+000	0.00°
2	1.000e+02	8.44E-006	59.31°
3	1.500e+02	**2.23E-004**	**179.38°**
4	2.000e+02	4.80E-006	89.83°
5	2.500e+02	4.39E-006	-179.70°
6	3.000e+02	1.82E-007	88.63°
7	3.500e+02	8.00E-008	-177.58°
8	4.000e+02	1.82E-008	13.66°
9	4.500e+02	1.93E-008	7.36°

100mVin **Total Harmonic Distortion: 0.022309%%**

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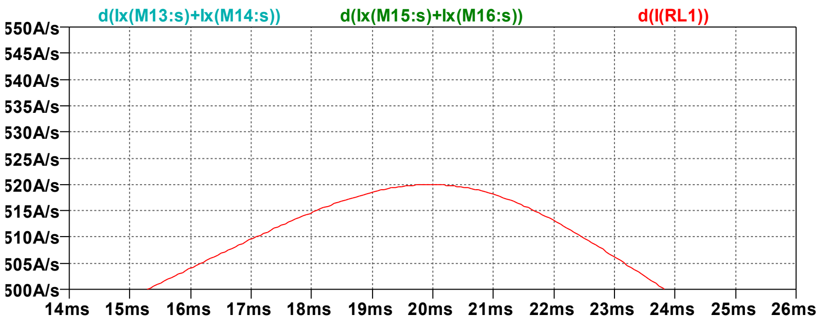

Figure 12.5. Case 5 – Downward bowing gain

1v1 **x=412m**, y=450m. Gain slope=+1520(L),-1920(R)

Harmonic Number	Frequency[Hz]	Normalized Component	Normalized Phase [deg]
1	5.000e+01	1.00E+000	0.00°
2	1.000e+02	9.29E-006	68.43°
3	1.500e+02	**2.75E-004**	**2.02°**
4	2.000e+02	3.93E-006	-89.15°
5	2.500e+02	1.32E-006	-3.86°
6	3.000e+02	1.21E-007	168.27°
7	3.500e+02	1.88E-007	-24.20°
8	4.000e+02	1.70E-007	137.82°
9	4.500e+02	2.31E-007	-37.40°

100mVin **Total Harmonic Distortion: 0.027494%**

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Misc References

A few references related to Linear Audio Vol.8 Cube-amp article on ultrasonic bandwidth effects, tube sound, Power Scaling™, soft clipping, and more:

1. David E. Blackmeyer, 'Life Beyond 20kHz', SVC (Sound & Video Contractor), Sept 1998, http://wilson-benesch.com/reviews/Life_Beyond_20kHz_Blackmer_SVC_Sep-1998.pdf
2. Dr. Hans van Maanen, 'Audibility of high-resolution digital audio,' http://www.temporalcoherence.nl/docs/HighReso.pdf . P13 "...the subjective sonic improvement is large..."
3. Dr. Hans van Maanen, Private correspondence 30 Dec 2012. Some some titles that show the effect of a higher bandwidth are:
   A. Santana Abraxas. Columbia/Legacy CS 65490 Sony Music Ent. Inc. Track 1(Singing Winds,Crying Beasts)
   B. Global Percussion Network "Rauk". Opus 3 CD 22011, track 5.
   C. FIM SACD 029 Audiophile Reference IV, e.g. tracks 3 and 16.
   D. Telarc SACD - 60011 "The Absolute Sound" sampler. Track 1.
4. Menno van der Veen, 'Subjective Sound Quality evaluation of SACD compared to CD,' http://www.temporalcoherence.nl/cms/images/docs/PAPER%20Subjective%20Sound%20Quality%20evaluation.pdf p1 "...SACD recordings compared to the same CD recordings, show significant quality differences between these two digital formats."
5. David Griesinger, 'Perception of mid frequency and high frequency intermodulation distortion in loudspeakers, and its relationship to high-definition audio' 30 May 2003, slide #23, "None of the popular music samples had anything at all above 23kHz" http://www.davidgriesinger.com/intermod.ppt
6. W. Steven Bussey, Robert M. Haigler, 'Tubes versus transistors in electric guitar amplifiers', Proceedings IEEE Ch 1610, p800-3, May-1981.

http://www.milbert.com/articles/tubes_vs_transistors_in_electric_guitar_amps

7. Eric Barbor, 'The Cool Sound of Tubes', IEEE Spectrum, August 1998 Vol 35 No 8. http://spectrum.ieee.org/consumer-electronics/audiovideo/the-cool-sound-of-tubes
Distortion spectra for SE triode, pentode, jFET, MOSFET, BJT – with 2V rms output for transistors and 20V rms for valves. Must-see sidebar AP distortion spectra comparison of SE triode, pentode, jFET, MOSFET, BJT by John Attwood http://spectrum.ieee.org/consumer-electronics/audiovideo/the-cool-sound-of-tubes/distortion
The comparison summary table can be found in text form with comments from David Berning at http://www.milbert.com/articles/tube_transistor_feature_comparison .

8. Peter Schüller, 'MST', A lecture at the High-End Fair, Munich 2012. http://burosch.de/audio-technik/509-high-end-2012-klang-2-english.html'

9. John Linsley Hood, WW Apr 1969 p_. Used a PS capacitance multiplier http://sound.westhost.com/jll_hood.htm http://sound.westhost.com/project15.htm
MOSFET version see Andrea Ciuoffoli, '30W-Class-A-Power-Follower' Electronics World May 2000 p382-3 http://www.audiodesignguide.com/my/Follower_99c.gif

10. Colloms (Martin), 'What has gone wrong with amplifier evaluation?', HFN/RR Oct 1977 p83-5. Summary of points held responsible for subjective sound quality and power amplifier improvements over the years: 1. Linearity and distortion; 2. Crossover [distortion] effects; 3. Bandwidth; 4. Power supplies; 5. Driver and input stages; 6. Transient and intermodulation distortion (TID); 7. Protection circuitry; 8. Solid-state versus valves; 9. Power output; 10. On evaluation techniques. "One still cannot predict from this how well the amplifier will audition". "This dramatic contradiction shows us quite clearly that at this elevated quality level there exists some important factor (or factors) affecting performance, that we have failed to recognise". Called factor 'X'. He hopes psycho-acoustic research will uncover the mechanism that will allow lab measurements to make a valid quality judgement. "**With only one or two exceptions we found that these [measured parameters] did not reflect subjective performance**". Letter: West (Ralph) HFN/RR Jan 1978 p79.

11. Comments on Klever-Klipper circuit
http://www.diyaudio.com/forums/solid-state/89023-bob-cordell-interview-error-correction-45.html
http://www.diyaudio.com/forums/solid-state/171159-bob-cordells-power-amplifier-book-51.html
#506 Sep 2006 As a solid-state centric designer for many years, my first reaction on hearing the tube amp was that it doesn't have a right to sound that good.
BTW, I always try to keep in mind the distinctions among:
▪amplifier is musical
▪amplifier sounds good
▪amplifier is neutral
Ideally, a well-behaved neutral amplifier always sounds good, but some program material and loudspeaker combinations sound better with an amplifier that is synergistically not neutral.
It seems that tube amplifiers have always had a reputation of being more musically tolerant to clipping than SS amps. We often hear something like "A 35-watt tube amp can play louder than a 35-watt SS amp".
I included optional dynamic soft clip (I called it the Klever Klipper) into my Super Gain Clone (SGC) amplifier and also described it in a chapter of my book. It does indeed give the SGC a more tube-like tolerance for signals that will clip it. It sounds good, **even though THD begins to rise significantly at power levels above about 1/3 power in the SGC,** which is rated at 40-50 watts. I show a curve of THD vs power for the SGC in the book with the Klever Klipper turned on and off.

12. Brian Santo, Volume cranked up in amp debate: Can solid-state sound really match that of tubes? Reprinted from Electronic Engineering Times, October 3, 1994. available at

http://www.trueaudio.com/at_eetjlm.htm . Interview with John Murphy of True Image Audio (Escondido), “When operated in a linear (or unclipped) mode, Murphy explained, tube amps sound the same as their solid-state counterparts, provided that their frequency response and group delay characteristics are well matched and their distortion levels are sufficiently low. The audible difference between tube and solid-state amps emerges *only when they are clipped*.” [emphasis added]

13. Jean Hiraga, 'Construction of a 20W class-A amplifier, General design', l’Audiophile No. 10 https://web.archive.org/web/20121010180230/http://www.tcaas.btinternet.co.uk/hiraga1.htm One of his design objectives is "Soft" distortion characteristic (rate of distortion going up steadily with the output power increase)”. Fig. 3 shows a 'natural or soft distortion' a progressive harmonic rise with power of valves.
14. [Yet to add more soft-clip refs] Nelson Pass said somewhere? “NFB takes something away” and my hunch is that it takes away soft clipping (and this can affects the amp sound even before harder clip starts *because* soft clipping is heard even before harder clip starts); Boyk; Norman Korean (Pt.2 2001); Fritz Langford Smith (referenced in Vol.8 article); Bob Carver's interview “misbehave differently when clip” (referenced in Vol.8 article), more?

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Appendix A: Cube-law Class-A expansion terms

In Linear Audio Vol.8 it was stated: “In Cubic-law Class-A $I_L={}^{15}/_{16}\,k(\sin wt - {}^{1}/_{15}\,k^{2}\sin 3wt)$, so there is 6.67% 3rd harmonic distortion at full output.” The following provides the derivation by expanding and collecting the terms for Cube-law Class-A. Also included is the derivation of the efficiency with the distortion. The efficiency without distortion is determined by simulation using a jig available from www.pak-project.org such as the file “PAK-ABM-Jig1-Cube-law-A”.

Derivation of output current for Cubic-law Class-A

$I_{out}=Ia - Ib$ where $Ia={}^{1}/_{8}I_m(1+V_{in}/V_b)^{3}$ and $Ib={}^{1}/_{8}I_m(1-V_{in}/V_b)^{3}$ so

$I_{out}= {}^{1}/_{8}I_m(1+V_{in}/V_b)^{3} - {}^{1}/_{8}I_m(1-V_{in}/V_b)^{3}$

$I_{out}= {}^{1}/_{8}I_m(1+V_{in}/V_b)(1+V_{in}/V_b)^{2} - {}^{1}/_{8}I_m(1-V_{in}/V_b)(1-V_{in}/V_b)^{2}$

$I_{out}= {}^{1}/_{8}I_m(1+V_{in}/V_b)^{2} + {}^{1}/_{8}I_m(V_{in}/V_b)(1+V_{in}/V_b)^{2} - {}^{1}/_{8}I_m(1-V_{in}/V_b)^{2} + {}^{1}/_{8}I_m(V_{in}/V_b)(1-V_{in}/V_b)^{2}$

$I_{out}= {}^{1}/_{8}I_m(1+V_{in}/V_b)^{2} - {}^{1}/_{8}I_m(1-V_{in}/V_b)^{2} + {}^{1}/_{8}I_m(V_{in}/V_b)[(1+V_{in}/V_b)^{2} + (1-V_{in}/V_b)^{2}]$

Notice first two terms are the Square-law difference type Class-A, so these are expanded separately:

$I_{sq} = {}^{1}/_{8}I_m(1+V_{in}/V_b)^{2} - {}^{1}/_{8}I_m(1-V_{in}/V_b)^{2}$

$I_{sq} = {}^{1}/_{8}I_m[1+2(V_{in}/V_b)+(V_{in}/V_b)^{2}] - {}^{1}/_{8}I_m[1 - 2(V_{in}/V_b)+(V_{in}/V_b)^{2}]$

$I_{sq} = {}^{1}/_{8}I_m\,4(V_{in}/V_b) = {}^{1}/_{2}\,I_m\,(V_{in}/V_b)$

and as expected this is purely linear.

The remaining terms involve cubic terms so are called Icu:

$I_{cube} = {}^{1}/_{8}I_m(V_{in}/V_b)[1+2(V_{in}/V_b)+(V_{in}/V_b)^{2} + 1 - 2(V_{in}/V_b)+(V_{in}/V_b)^{2}]$

$I_{cube} = {}^{1}/_{8}I_m(V_{in}/V_b)[2+2(V_{in}/V_b)^{2}] = {}^{1}/_{4}\,I_m(V_{in}/V_b) + {}^{1}/_{4}\,I_m(V_{in}/V_b)^{3}$

We can see that I_{cube} contains a linear part.

Now combining the two square-law and cubic parts we obtain:

$I_{out}= {}^{1}/_{2}\,I_m\,(V_{in}/V_b) + {}^{1}/_{4}\,I_m(V_{in}/V_b) + {}^{1}/_{4}\,I_m(V_{in}/V_b)^{3}$

$$\boxed{\mathbf{Iout}= {}^{3}/_{4}\,\mathbf{I_m}({}^{Vin}/_{Vb}) + {}^{1}/_{4}\,\mathbf{I_m}({}^{Vin}/_{Vb})^{3}}$$

Differentiating with respect to Vin gives the gm

$g_m = {}^3/_4 (I_m/V_b) + {}^3/_4 (I_m/V_b{}^{\wedge 3}) V_{in}{}^{\wedge 2})$ so

$Gm = {}^3/_4\ {}^{Im}/_{Vb}\ (1+({}^{Vin}/_{Vb})^{\wedge}2)$

This is the upward curving gain plot seen in Figure 2 of the above simulation. In Cubic-law Class-A the gain at the ends are twice that of the middle dip.

With a sinewave

With a sinewave, $V_{in}/V_b = k \sin \omega t$ where 0<=k<=1, gives

$I_{out} = ¾\ I_m\ k \sin \omega t + ¼\ I_m (k \sin \omega t)^{\wedge 3}$

$I_{out} = ¾\ I_m\ k \sin \omega t + ¼\ I_m\ k^{\wedge 3} (\sin^{\wedge 3} \omega t)$ where

$\text{Sin}^{\wedge 3}\ wt = (\sin \omega t)(½)(1\text{-}\cos 2\omega t) = (½)\sin \omega t – (½)[(½\sin 3\omega t)+½\sin \omega t] = ¾ \sin \omega t – ¼ \sin 3\omega t$

$I_{out} = ¾\ I_m\ k \sin \omega t + ¼\ I_m\ k^{\wedge 3} (¾\sin \omega t – ¼ \sin 3\omega t)$

$I_{out} = ({}^3/_4 + {}^3/_{16}\ k^{\wedge 2}) k I_m \sin \omega t – {}^1/_{16}\ I_m\ k^{\wedge 3} \sin 3\omega t$

when k=1 gives $I_{out} = ({}^{15}/_{16})\ I_m \sin \omega t – {}^1/_{16}\ I_m \sin 3\omega t$ or alternatively

$I_{out} = ({}^{15}/_{16})\ I_m (\sin \omega t – {}^1/_{15} \sin 3\omega t)$ showing there is ${}^1/_{15}$ or 6.67% 3rd harmonic distortion at full output due to the gm variation above.

Efficiency analysis (with 3rd harmonic distortion)

In Cubic-law Class-A $I_L = ({}^3/_4 + {}^3/_{16}\ k^{\wedge 2}) k\ I_m \sin \omega t – {}^1/_{16}\ I_m\ k^{\wedge 3} \sin 3\omega t$.

The output power is: $I^{\wedge 2}\ R_L = I^{\wedge 2} \times V_{dd}/Im$

$I^{\wedge 2}\ R_L = V_{dd}/Im\ [k(¾ + {}^3/_{16}\ k^{\wedge 2}) I_m \sin \omega t - {}^1/_{16}\ k^{\wedge 3}\ I_m \sin 3\omega t]^{\wedge 2}$
$I^{\wedge 2}\ R_L = {}^1/_{256} (V_{dd}/Im)\ Im^{\wedge 2}\ k^{\wedge 2} [256(¾ + {}^3/_{16} k^{\wedge 2})^{\wedge 2} \sin wt - k^{\wedge 2} \sin 3\omega t]^{\wedge 2}$
$I^{\wedge 2}\ R_L = {}^1/_{256}\ V_{dd}\ Im)\ k^{\wedge 2} [256(¾ + {}^3/_{16}\ k^{\wedge 2})^{\wedge 2} \sin^{\wedge 2}\omega t – 2\times 12(1 + {}^3/_{16}\ k^{\wedge 2}) k^{\wedge 2} \sin \omega t \times \sin 3\omega t + k^{\wedge 4} \sin 3\omega t]$

Since we only want the average output power and sin(nωt) terms average to zero and only $\sin^{k\wedge 2}$ wt terms have a dc component.

Average $I^{\wedge 2}\ R_L = {}^1/_{256}\ Im\ Vdd\ k^{\wedge 2} [256(¾ + {}^3/_{16} k^{\wedge 2})^{\wedge 2}(0.5)(1\text{-}\cos 2\omega t) – 0 + k^{\wedge 4}\ 0.5(1\text{-}\cos 6\omega t)]$

$Pout_{ave}$ **$= ½\ Im\ Vdd\ k^{\wedge 2}\ [(¾ + {}^3/_{16} k^{\wedge 2})^{\wedge 2} + {}^1/_{256} k^{\wedge 4}]$**

Since $Pout_{Sine} = ½\ Im\ Vdd$

$Pout_{ave\ distorted} / Pout_{sine} = {}^{225}/_{256}\ k^{\wedge 2} + {}^1/_{256}\ k^{\wedge 6}$

(For k = 1 $Pout_{ave\ distorted} / Pout_{sine} = {}^{226}/_{256} = 0.8828125$ or 88.25%).

Since $Pin = V_{dd} \times I1 + V_{dd} \times I2$ and due to symmetry $Pin_{ave} = 2V_{dd} \times I1_{\ ave}$

$Pin_{ave} = 2V_{dd} \times Im/8\ (1+k \sin \omega t)^{\wedge 3}$

$Pin_{ave} = 2V_{dd} \times Im/8\ (1+k \sin \omega t)(1+k \sin \omega t)^{\wedge 2}$

$Pin_{ave} = 2V_{dd} \times Im/8\ (1+k \sin \omega t)\ (1+2k \sin \omega t + k^{\wedge 2} \sin^{\wedge 2}\omega t)$

$Pin_{ave} = 2V_{dd} \times Im/8\ [1+3k \sin \omega t + 3k^{\wedge 2} \sin^{\wedge 2} \omega t + k^{\wedge 3} \times \sin^{\wedge 2}\omega t \times \sin \omega t]$

Since the average $\sin \omega t = 0$

$Pin_{ave} = 2V_{dd} \times Im/8\ [(1+0+ 3k^{\wedge 2} \times ½\ (1\text{-}\cos 2\omega t) + 0]$

$Pin_{ave} = ¼ V_{dd}\ Im\ (1 + {}^3/_2\ k^{\wedge 2})$

$\text{Efficiency}_{\ distorted} = Pout_{ave\ distorted} / Pin_{\ ave}$

$\text{Efficiency}_{\ distorted} =$ **$\{½\ V_{dd}\ Im\ k^{\wedge 2} [256(¾ + {}^3/_{16} k^{\wedge 2})^{\wedge 2} + k^{\wedge 4}]/256\}$** $/\ ¼\ V_{dd}\ Im\ (1 + {}^3/_2\ k^{\wedge 2})$

$\text{Efficiency}_{\ distorted} = k^{\wedge 2}\ \{(¾ + {}^3/_{16} k^{\wedge 2})^{\wedge 2} + {}^1/_{256}\ k^{\wedge 4}\} / ½\ (1 + {}^3/_2 k^{\wedge 2})$

Max Efficiency $_{\text{distorted}} = ({}^{225}/_{256} + {}^{1}/_{256}) / {}^{5}/_{4} = {}^{4}/_{5} \times {}^{226}/_{256} = {}^{226}/_{5x64} = {}^{226}/_{320} = 70.625\%$

(Check: simulated 6.667% 3rd gave $^{28.25W}/_{40W}$ = 70.625% ie same).

Efficiency analysis (with no 3rd harmonic distortion)

When NFB used to remove the distortion in the output signal 31.9797W average output and 42.8289W average input.

Maximum undistorted Class-A^3 sine efficiency = $^{31.9797W}/_{42.8289W}$ = 0.74668

Rounded to 2 dp as percentage is 74.67% (as in Linear AudioVol.8 article).

Maximum undistorted Class-A^3 sine efficiency 74.67%

A simulation jig is available from www.pak-project.org where the distortion in Class-A is removed using the cubic equation for Pre-Compensation (PC). This jig can also do PC for Class-AB (such as the file "PAK-ABM-Jig1-Cube-law-AB50"). For example AB50% is where the idle current is half the Cube-law Class-A level, or the peak current is 2x the Class-A current, or the peak current is 16×Iq and

Maximum undistorted Class-AB^3 sine efficiency $^{128.0W}/_{165.0W}$ = 0.7757 or 77.57%

BTW Class-B max sine efficiency is π/4 or 78.54% to 2 dp.

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Appendix B: Specifications for Cube-A, Cube-AB, Square-A & JLH

Table B1: Specifications for 3 Class-A designs

	Cube-law A Vol.8	**Cube-law AB$_{50}$** Vol.8	**Square-law (Vol.1 Fig 2)**	**JLH Class-A** (15W 1996)
Rated power 8R & Class	100W Class-A	100W Class-AB	50W Class-A	15W Class-A
Input sensitivity (8R load)	0.7V rms	0.7V rms	2V rms	~1V rms
Input impedance	500k // 47pF	500k // 47pF	~1k // ~1.5nF	47k // 220pF
Sinewave efficiency max P	65%	68%	59%	28% with reg PS
Peak output current	15A?	15A?	20A	-
Open loop gain into 8R	50dB	50dB	27dB	-
Open loop bandwidth	50kHz	50kHz	50kHz	-
Power bandwidth	100kHz	100kHz	100kHz	-
CL Load ripple (rms)	?	?	~1mV	3mV
CL s/n ratio w.r.t. max P	-100dB	-100dB	-83dB (80kHz bw)	-75dB†
Output resistance at 1kHz	1.2R	1.2R	6.5R	0.25R
CL Slew rate	70V/us?	70V/us?	100V/us	40V/us
Harmonic distortion (1kHz)	0.06%% 15W	0.1% 15W	0.15% 15W	<0.1% 15W
Typical THD at 1W	TBD<<0.01%	TBD<<0.02%	0.01%	0.03%
Quiescent current	0.625A	0.31A	0.9A	1.0A
Supply voltage at FL	±46V	±46V	±32V	±22V (27V unreg.)
Supply voltage at idle	±49V	±49V	±33V	-
Quiescent power dissipation	61	31	60W	64W (10W reg.)
Heatsinks	0.55K/W	0.5K/W	0.5K/W	2x 0.3K/W
P.S. Capacitors	2x15,000uF pc	2x10,000uF pc	2x10,000uF	2x33mF

Mains transformer	250VA pc	200VA pc	160VA	~180VA

† Estimated.

Appendix C: BoM for Linear Audio Vol.8 (revised Aug 2014)

...BoM

Parts BoM (for 1 channel 100W/8R)

Part	Description	Qty 1ch	Mouser Part #	sr Qty1C	Other suppliers	Notes	
C1	1uF 63V PET	1	871-B32529C105J	1	Farnell 975-0932		http://au.mouser.com/Pro
C2	1000uF 16V Electro	1	140-REA102M1CBK1016P	1			http://au.mouser.com/Pro
C3	47pF Ceramic NPO	1	810-FK28C0G1H470J	1			http://au.mouser.com/Pro
C4	470pF Ceramic NPO	1	810-FK28C0G1H471J	1			http://au.mouser.com/Pro
C5,C6	100nF 63V	2	80-R82DC3100AA50J	2			http://au.mouser.com/Pro
C7,C8	220pF 200V Ceramic NPO	2	80-C315C221J2G	2	Farnell 121-6415	Note 200V NPO	http://au.mouser.com/Pro
C9	22nF 63V	1	80-R82EC2220DQ50J	1			http://au.mouser.com/Pro
C10	100nF 200V	1	667-ECQ-E2104KF	1	Farnell 170-2665	Note 200V	http://au.mouser.com/Pro
C11,C12	100uF 16V electro	2	871-B41827A4107M000	2			http://au.mouser.com/Pro
ZD1 – ZD4	15V 1W zener	4	512-1N4744ATR	4		Z3,4 only needed for -Z laterals (no internal Z's)	http://au.mouser.com/Pro
ZD5, ZD6	24V 1W zener	2	512-BZX85C24	2			http://au.mouser.com/Se
J1 – J9	Jumper PCB pins	9	649-67996-406HLF	2		break into 6x2pins (Mouser)	http://au.mouser.com/Se
J1A – J9A	Jumper shunt	10	571-881545-5	10			http://au.mouser.com/Prod
RF1	5A fast 5x20mm fuse	1	576-0217005.HXP	1		Alt: 4A fast IEC	http://au.mouser.com/Pro
RF1a	20mm fuse holder	1	693-0031.8211	1	Element 14 116-2741		http://au.mouser.com/Pro
CON1	4-way PCB screw terminal 5mm	1	51-1706640	1		green	http://au.mouser.com/Prod
CON2	3-way PCB screw terminal 5mm	1	651-5442183	1		green	http://au.mouser.com/Pro
CON3	2-way PCB screw terminal 5mm	1	651-5442170	1		green	http://au.mouser.com/Pro
Socket	DIP-14 turned	6	575-11044314	6	Farnell 136-2806, RS 813-121	Extra to plug in some R's	http://au.mouser.com/Pro
Socket	DIP-8 turned	1	575-199308	1			
U1-U5	74HCU04 hex inverter	5	595-SN74HCU04N	5	Farnell 128-7555 (TI)	prefer NXP or TI	http://au.mouser.com/Pro
U6	7812 pos 12V 3 term reg	1	512-KA7812AETU	1	Match to neg 12V reg.	AETU +-2% or +-0.25V	http://au.mouser.com/Pro
U7	7912 neg 12V 3 term reg	1	512-KA7912ATU	1			http://au.mouser.com/Pro
U8	TL072 FET dual opamp	1	595-TL072IP	1		Alt LT1057	
D1,2	Green LED 3mm	2	859-LTL-2231AT	2			http://au.mouser.com/Pro
D3	BiColor Gr/Red 20mcd 20mA	1	645-521-9458F	1	Farnell 114-2488	prefer 3mm	http://au.mouser.com/Pro
M13, M14	p-lateral 16P16 or 20P20	2	-	0	Farnell 185-6764	Exicon or Magnatec	
M15, M16	n-lateral 16N16 or 20N20	2	-	0	Farnell 185-6762	Exicon or Magnatec	
Q1,Q3, Q5,Q11	BC547BP	4	512-BC547BBU	4			http://au.mouser.com/Pro
Q2,Q4, Q6,Q12	BC557BP	4	863-BC557BG	8			http://au.mouser.com/Pro
Q8,Q14, Q18	2SC5706 NPN 50V 5A 300MHz 300 Hfe	3	863-2SC5706-H	0	RS 792-4997	Alt dpak Mouser 863-2SC5706-TL-H	http://au.mouser.com/Pro
Q9,Q16, Q20	2SA2039 PNP 50V 5A 300MHz 300 Hfe	3	863-2SA2039-E	0	RS 774-0730	Alt dpak Mouser 863-2SA2039-TL-E	http://au.mouser.com/Pro
R13	50k trimpot	1	652-3296W-1-503LF	1		Multi-turn	http://au.mouser.com/Pro
R22A,R23A	10k trimpot	2	652-3296W-1-103LF	2		Multi-turn	http://au.mouser.com/Pro
R33	1k trimpot	1	652-3296W-1-102LF	1			
R48	500k trimpot	1	652-3296W-1-504LF	1		1. Alt: 1 Meg	http://au.mouser.com/Pro
R8A	50k dual logarithmic (*see note)	1*	313-1240F-50K	1	Farnell 182-2849 (metal) or Farnell 136-4	Alt Mouser metal shaft 313-1240F-50K. *	http://au.mouser.com/Se
R1,R2	1Meg 0.25W	3	660-MF1/4DC1004F	3			http://au.mouser.com/Pro
R3,R9	15k 0.25W 1%	1	660-MF1/4DC1502F	1			http://au.mouser.com/Pro
R4	-	-	-	0			
R5	330R 0.25W	1	660-MF1/4DC3300F	1			http://au.mouser.com/Pro
R6	1k2 0.25W	1	660-MF1/4DCT52R1201F	1			http://au.mouser.com/Pro
R7	22 0.25W	1	660-MF1/4DC22R0F	1			http://au.mouser.com/Pro
R8	Wire link	-	-	0			
R10,R11	10k 0.25W 1%	2	660-MF1/4DCT52A1002F	2			http://au.mouser.com/Pro
R14,R15, R51	2M2 0.25W 1%	3	660-MF1/4LCT52R225G	3			http://au.mouser.com/Pro
R16,R17	12k 0.25W 1%	2	660-MF1/4DC1202F	2			http://au.mouser.com/Pro
R12,R18,R19,R44-47	1k 0.25W 1%	6	660-MF1/4DCT52A1001F	6		R44-47 matched to 0.2%	http://au.mouser.com/Pro
R20,R21, R24,R25	22k 0.25W 1%	3	660-MF1/4DC2202F	3			http://au.mouser.com/Pro
R23	3k3 0.25W	1	660-MF1/4DC3301F	1			http://au.mouser.com/Pro
R26,R27	470R 0.25W	2	660-MF1/4DC4700F	2			http://au.mouser.com/Pro
R28,R29	100R 0.25W 1%	2	660-MF1/4DC1000F	2			http://au.mouser.com/Pro
R30	560R 0.25W	1	660-MF1/4DC5600F	1			http://au.mouser.com/Pro
R31	150R 0.25W	1	660-MF1/4DCT52A1500F	1			http://au.mouser.com/Pro
R32,R34, R36	220R 0.25W	3	660-MF1/4DC2200F	3			http://au.mouser.com/Pro
R32A	leave open	0	-	0			
R35,R37	470R 0.25W	2	660-MF1/4DC4700F	4			http://au.mouser.com/Pro
R38	50mR 2W 10% or 50mR 5W 1%	1	66-SPHR050JTR	1	Farnell 175-1806 2W/10% or Farnell 232-8368 5W/1%		http://au.mouser.com/Pro
R39	22R 1W	1	594-5083NW22R00J	1			http://au.mouser.com/Se
R40,R41	100k 0.25W 1%	2	660-MF1/4DC1003F	2		match to 0.2%	http://au.mouser.com/Pro
R42,R43	10k 0.25W 1%	2	660-MF1/4DC1002F	2		match to 0.2%	http://au.mouser.com/Pro
R48	Trimpots - see above	-	-	0			
R49	1M 0.25W	0	-	0		Was 1M2. Use 1M with 1M trimpot	
R50	33k 0.25W	1	660-MF1/4DC3302F	1			http://au.mouser.com/Pro
R52,R53, R80	2k2 0.25W	3	660-MF1/4DC2201F	3			http://au.mouser.com/Pro
R54, R55	1k 1W	2	594-5083NW1K000J	2			http://au.mouser.com/Pro
Heatsink	0.55K/W	1	-	0	Altronics H0536	As shown in Fig.1	http://www.altronics.com.
Thermal washers	TO-264 24mm x21mm BER 180-ND	4	951-SP900S-009-00114	4			http://au.mouser.com/Pro
Transformer	500VA 38-0-38	1*	546-1182R30 +mods	0	Farnell 167-5097 40V or RS 223-8285 40V	One transformer does 2 ch. Alt 40-0-40	http://au.mouser.com/Pro
PS caps	Triple 4700uF/63V	6	598-SLPX472M063E3P3	0	Farnell 945-2923 x3 (85C 2000h 40x26x10 very low	Triple C for lowest cost	http://au.mouser.com/Se
Bridges	25A 400V	2	512-GBPC2506	0		1 bridge per sec. is best	http://au.mouser.com/Pro
PA mains fuse	5AT Slow	1	693-0034.3124	0		Alt 4AF	http://au.mouser.com/Pro
PCB		1	-	0			

Alternative suppliers

M13, M14	p-lateral 20P20 using single die	4			Class-D UK for 'Alfet' lateral - lower cost	ACD100NSD/ALF08N20V =8A,200V,TO-2	http://www.class-d.com/l
M15, M16	n-lateral 20N20 using single die	4			Class-D UK for 'Alfet' lateral - lower cost	ACD102PSD/ALF08P20V =8A,200V,TO-2	http://www.class-d.com/l
M13, M14	p-lateral 20P20 using single die	4			Altronics (Australia & NZ) Z1432	ECX10P20 TO-247 plastic single die	http://www.altronics.com.
M15, M16	n-lateral 20N20 using single die	4			Altronics (Australia & NZ) Z1430	ECX10N20 TO-247 plastic single die	http://www.altronics.com.au
M13, M14	p-lateral 20P20 using TO-3's	2			Altronics (Australia & NZ) Z1476	ECF20N20 TO-3 metal	http://www.altronics.com.au
M15, M16	n-lateral 20N20 using TO-3's	2			Altronics (Australia & NZ) Z1474	ECF20N20 TO-3 metal	http://www.altronics.com.au
M13, M14	p-lateral 20P20 using TO-264	2			Lede Electronics	BUZ906CDP Pch-200V-16A TO-3PBL	http://www.ledeaudio.com
M15, M16	n-lateral 20N20 using TO-264	2			Lede Electronics	BUZ901DP Nch-200V-16A TO-3PBL	http://www.ledeaudio.com
M13, M14	p-lateral 20P20 using TO-264	2			Mouser need to call for quote	BUZ906DP Pch-200V-16A TO-3PBL	http://au.mouser.com/Pro
M15, M16	n-lateral 20N20 using TO-264	2			Mouser need to call for quote	BUZ901DP Nch-200V-16A TO-3PBL	http://au.mouser.com/Pro

Misc							
U1-U5	74HCU04 SMD package	0	771-74HCU04D-T	0	Mouser NXP but only SMD (need adapters)		
SO-14 pin adaper		0		0	RS 158-2890		http://au.rs-online.com/we
Thermal washers	TO-264 24mm x 21mm		-	0	Digikey BER 180-ND		http://www.digikey.com.a
Transformer	600VA 38-0-38			0	Antekinc AN-6438	USA suppier. Good price, LV sec's	http://www.antekinc.com/
Heatsink	0.55K/W	1	-	0	Alt Jaycar 2xHH8555	Jaycar HH8546 no longer stocked	http://www.jaycar.com.au
PS caps (long life)	Triple 4700uF/63V	6	647-LGY1J472MELB40	0	Alt 105C 5000h 30x40x10 (costs 25% more than 85C 3000h)		http://au.mouser.com/Pro
Overtemp switch		2	802-STO-160	0		Thermostats 68-74C (155-165F) N/C	http://au.mouser.com/Pro
Alt, PCB connectors	36 way plug			0	Jaycar HM3272 (RS? Farnell)	Cut to length (Aug 2014 out of stock)	http://www.jaycar.com.au
Alt, PCB connectors	36 way pcb socket			0	Jaycar HM3270 (RS? Farnell)	Cut to length (Aug 2014 out of stock)	http://www.jaycar.com.au
Alt 40 pin jumper strip				0	Jaycar HM3212	Cut to length	http://www.jaycar.com.au
Alt 2x20 jumper strip				0	Jaycar HM3250	Cut to length	http://www.jaycar.com.au
Jumper shunts links				0	Jaycar HM3240	Only black (pk 10)	http://search.jaycar.com.
Alt Bourns trimpots	1k 25T Spectrol (equiv 3296W)			0	Jaycar RT4644		http://search.jaycar.com.
Alt Bourns trimpots	10k 25T Spectrol (equiv 3296W)			0	Jaycar RT4650		
Alt Bourns trimpots	1Meg 25T Spectrol (equiv 3296W)			0	Jaycar RT4658		
Clip level pot	50k log dual metal shaft			0	Jaycar RP3760		
1Meg trimpot	R48 alt	1	652-3296W-1-105LF	1		R48 Alt. 1 Meg for 150W upgrade	http://au.mouser.com/Pro

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Extras for set up and testing:

1. For R12 extra values (2 off each for 2 ch): 1k2,1k5,5k6. For R44,47 extra values (4 off each): 470, 390, 330. For R30 extra values (2 off each): 470, 510, 620, 680.

2. Dummy load 8 ohms at least 100W. 100R 10W (series with speaker). Optional 2nd 100W 8 ohm dummy load (for testing at 4 ohms).
3. 2 trimpots for simple PC sound card interface (eg Christian Zeitnitz free PC scope): 2x 10k single turn. 100k ¼ watt 5% (drop output volts).

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Appendix D: Assembly details

PCB assembly (outline)

1. SMD d-pak transistors (Q8,9,14,16,18,20). If using through hole packs then thermally link Q5 & Q8 together, also Q6 & Q9 together (does not need thermal paste – too messy).
2. Insertion sockets for resistors R3, 6, 12, 18, 19, 22, 23, 27, 30, 44, 47 (Underlined are recommended, the others are optional)
3. Insertion sockets for C3, C4 (both are optional)
4. All other resistors except R38 (50mR).
5. Zeners (Z3,4 only needed for ...-Z MOSFET's, those without the internal zener)
6. IC sockets
7. TO-92 BJT's (Note Linear Audio PCB has TO-92 silk-screen's reversed, so place TO-92's opposite to what's shown).
8. Caps
9. Fuse holder
10. Connectors
11. 12V Reg's
12. R38 (50mR)
13. Mount MOSFET's on heatsink (drill to match PCB pitch), finger tighten, then...
14. Solder MOSFET's and tighten bolts.
15. Commence setup sequence (see below)

Set up [extracted from Vol-8 article p13,14]

Several links are provided to help get the amp running. If you use IC sockets for all the IC's then you can swap IC's to get the lowest output offset voltage. The clip indicator IC should be socketed since it can generate unwanted spikes during setup or for sensitive THD measurements. Some of the fixed resistors can be socketed using machined pins extracted from IC sockets. This avoids the difficulty of de-soldering components from plated through hole PCBs. After completing setup you can solder the wires into the pins. Resistors that you may want to change are R30 (for Vos), R22, 23 for idle current, R12 for clip shape, and R18, 19. If you would like to experiment with frequency compensation then provide pins for C3, R3, C4, and R6.

The first stage of setup is to test the linear driver section with the power MOSFETs with the square-law driver IC's and the input IC left off the board and no jumpers, R22A at minimum (shorts R22) and R33 at maximum resistance. You need a dummy load to do full bench tests, but for set up you can get away with a 100 ohm 10 watt resistor, or use an electric water cooker jug which is around 30 ohms for 230V elements or 8 ohms for 110V mains. Place a high current ammeter in series with the positive power rail. Temporarily short R31 and R32 and apply power. The ammeter should read less than 100mA. Check that the regulator voltages are ±12V and LEDs D1,2 light up. There should be less than

5mV across R28 and R29. Remove the shorts across R31 and R32 and check that the ammeter indication does not increase.

You can now increase trimpot R22A until the idle current increases and you should be able to get over 100mA. Adjust trimpot R33 and measure the output voltage across the dummy load, it should reduce to zero and then reverse. You can now apply an ac input signal to the R18, 19 node and check the output with an oscilloscope or an ac meter; about 1V should drive the output stage into clipping. Next you need to find a low offset voltage CMOS IC for U1. Plug U1 while the power is off then power up and short the input to output of U1b (you can place a screwdriver to short pins 3 and 4) and measure the voltage on these pins. Choose the one with the lowest voltage (my lowest was 3mV). If you are making two channels then find the two lowest ones. Group the others into matched pairs. You can now feed an input voltage to the input connector (via a volume control) and drive the output into clipping. This is open loop so less than 50mV rms is enough for some clipping. You can add link J2 to add some feedback. The amplifier in this form is standard Class-AB.

To complete the amplifier for cube-law Class-A add the remaining IC's still with links J3, J4 and J5 open. Reduce R22A to minimum. On power up short Vpre (eg R18, 19 node) to common and measure the current through J3 and J4. The current should be in the 1-3mA range and the average around 1.5mA to 2mA. Try swapping U4 and U5 to find which combination gives the highest average current through the links.

Now add links to J4 and J5 and also J2 and J9. With the power off adjust R33 until the resistance across R32 is around 130 ohms. Turn on and increase the idle current with no input signal to initially 300mA and check that the dc output voltage is less than 1 volt and check that this voltage can be trimmed to zero using R33. R32/R33 net resistance should be above 100 ohms and if it needs to be less than this then you need to change R30 by replacing it with a temporary 1k trimpot and then use a new fixed resistor.

Once the output voltage is zero the idle current can be increased to 620mA. After the heatsink warms up the idle current will increase to around 630mA. You can run the amp with an input signal – this is operating in cube-law Class-A without error correction. To add error correction you can close link J3 while running and adjust R13 for minimum distortion with a spectrum analyser.

Next open J9 link to reduce the output offset voltage by a factor of about 10 and it should reduce to under 10mV. If the Vos is still too high then the best way to trim Vos is by replacing R30 with with a temporary 1k trimpot rather than using R22 which alters the symmetry (2nd harmonic) as well as Vos. R30 was not made a permanent trimpot because R22 and R30 are interactive making setup overly complicated. Reducing the Vos below 100mV is not considered to be that important since 100mV into 8 ohms is only 1mW.

I used Christian Zeitnitz's soundcard software [**19**] with a 100k dropper resistor to a 10k trimpot to attenuate the output signal to the mic input (since my PC line input did not seem to work) as this input clipped at 500mV. I also looped back the soundcard 1V output to the other 0.5V input via another 10k trimpot for the X-Y plots. The X-Y plot mode is good for trimming R33 for symmetry and the FFT display is best for trimming out the 3rd harmonic at about 8 volts rms into 8 ohms.

Use an X-Y plot to set the clip level using a 10k trimpot in series with R8 to common and an 8 ohm dummy load. It is helpful to include a 100 ohm 10 watt resistor in series with a loudspeaker and connect these across the dummy load to hear the difference in clipping when the ripple is present and after it is removed by increasing the trimpot in series with R8. Use 300Hz to hear the harmonics more easily. Set the clip level so no power rail ripple is heard when over driving with an input voltage of around 1V rms. If you have a second 8 ohm dummy load then connect it in parallel with the other 8 ohm load (only for a few seconds) while it is clipping to check that no ripple is heard using 4 ohms, and adjust the clip level if necessary. If your power supply is shared with another amplifier then run them with the same input signal; one into 8 ohms and the other into 4 ohms (for a few seconds) and check that no ripple is heard in the channel with 4 ohms.

If you want to make the circuit as symmetrical as possible then R16, 17 and R20, 21 can be matched as close as possible from a pack of 10 using a DMM. Also the positive and negative 12V regulators can be matched before soldering them in. Alternatively, you can trim the lower voltage regulator to within 10mV of the other by adding a 100 ohm trimpot in series with the reference lead and then replace it with a fixed resistor.

As mentioned, the square-law stage is used to boost the peak output current at the end of the Class- A region to 8 times the idle current. Since the square-law driver section has a fixed gain which is fixed by the CMOS inverter FET gain, we can change the gain of the linear stage by changing R28, 29 while checking the peak output current at the end of the Class-A region. This can be easier observed at half the gain at the origin. In this design the slopes are closer to a ratio of 2:3 (as mentioned above and seen in Fig 3 due to soft clipping and the MOSFETs subthreshold region). Figure 3 was generated in open loop with links J2 and J3 open and applying a 100mV sinewave input voltage.

The Clip indicator

To set trimpot R48, run the amplifier at 15V rms (half full output swing) with an 8 ohm dummy load (50W will do). Reduce R48 until the LED's go out. Then count the turns until the other LED turns on. Then back up by half the number of turns. (If you use a single turn trimpot then visually note the two boundaries and set it to the middle of this range).

==================The-End===================