

KT88 – TRIODE

Parallel Push-Pull 60Watt

Mono-Block

Music Amplifier



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1.0 Acknowledgments.

The author wishes to express grateful appreciation to Kevin Kennedy (of Kennedy Tube Audio) and Steven Spicer (who built and maintains the Australian Kiewa Valley Stereo web site) for their mentorship. Both Kevin and Steven were extremely forthcoming and helpful during my efforts to advance my education and experience in this fascinating subject. Many pitfalls have been avoided through the willingness of both these gentlemen to share their expertise with me.

2.0 Introduction.

The design and development of this amplifier was inspired by dissatisfaction with semi-conductor equipment and the prohibitive cost of high-end vacuum tube hi-fi. The first iteration of the design (having a modified Hedge differential cascode drive topology¹) was completed in 1998; the design presented herein was completed in 1999.

This text attempts to explain how the design decisions were made and describe the design.

The subject amplifier is a medium power mono-block design, based on triode connected KT88 power beam tetrodes (also known as kinkless tetrodes) operating in Class AB1. Two cascaded differential voltage gain stages drive the parallel push-pull output stage without the use of cathode followers. The amplifier may be driven by either unbalanced or balanced sources. A separate precision, wide bandwidth voltage regulator is provided for each supply rail. The regulated power supplies confer not only enhanced resolution but also greater consistency in operation since the high-tension (HT) and bias voltages will remain on the design point regardless of output level or normal variations in the AC supply voltage.

Experimentation with global negative feedback (up to 12dB) on the prototype design led me to conclude that it detracts from the sonic performance of a good audio amplifier and thus I resolved to use no global negative feedback. Accordingly, great care was taken during the further development of the differential drive circuit to minimize any tendency to generate odd order harmonics. (Odd order harmonics do not cancel in a push-pull output stage.) The measured performance of this amplifier design reveals not only low total harmonic distortion but more significantly, low intermodulation distortion for a topology having no global or output stage negative feedback.

3.0 Design Approach.

It is well known that odd order harmonics are more readily detected by the human ear than even order harmonics. To gain an appreciation of this, try the following experiment:

Sit at a piano (preferably in good tune and without false beats and other defects) and play A below middle C (220Hz) and then play both A220 and A440 simultaneously (one octave thus the fundamental and 2nd harmonic) a richening of the tone will be apparent. It is possible to envisage that if you were seated out of sight of the keyboard while another strikes the octave, it would not be obvious whether a single note or an octave has been struck. However, if we were now to repeat the exercise using A220 and E above A440 (660Hz thus the third harmonic on A220) there would be no doubt at-all as to whether a single tone has been struck or the fundamental and 3rd harmonic struck simultaneously.² Further, Hamm observed that odd order harmonics cause the sound to be perceived as “stopped” or “covered” whereas even order harmonics are experienced as lending body to the sound making it “fuller” or more “choral” in aspect.³

Initially, the intention was to design an ultra-linear (also known as partial triode) amplifier but I soon came to correlate the behavior of triodes with what I had read and experienced regarding

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relative the sensitivity of human hearing to odd and even order harmonics. When viewed in the light of this information, the predisposition of triodes to produce predominantly second order harmonic distortion may partly explain why the sound of triodes is preferable to that of most, if not all, other amplification devices.

The predisposition of a triode to produce predominantly second harmonic distortion is due to the power law characteristic of the plate voltage-current relationship exhibited by triode devices as elucidated by Child's law viz; $I \propto V^{3/2}$

This gives rise to the characteristic "to the right and up" plate bend of the grid lines on the plate curves. If we plot a realistic load line on a set of typical triode plate curves, the right hand end will be in a region of asymmetrical compression of the grid lines caused by the plate curvature. This non-linear transfer characteristic will result in a waveform that is asymmetrically distorted about the x-axis, having higher negative going peaks than positive going. If we imagine now that the transfer characteristic is linear (uncompressed toward cut-off) we will obtain an undistorted sine wave of the same frequency. Algebraic subtraction of the distorted resultant from the undistorted waveform results in a new frequency double that of the fundamental and of lesser amplitude, a second harmonic.

Pentodes behave exponentially, giving rise to odd order harmonics. (A succession of exponential curves is a square wave with the higher order harmonics absent⁴ and a square wave expands as a Fourier series of odd order harmonics.)

It follows then, that the use of triodes in a push-pull circuit so as to minimize the residual characteristic triode coloration of the sound through cancellation of even order harmonics in the output stage is one approach to hi-fi reproduction.

I embarked upon the development of a drive circuit having wide bandwidth, low output impedance, low distortion - preferably predominantly even order - and near perfect balance.

I was determined to develop a fully balanced design, that is a topology which will produce balanced phase and inverse-phase signals from an unbalanced source at the first stage and from there to summation in the output transformer. Many permutations of **Long-Tailed Pair**⁵ (hereafter referred to as LTP) were analyzed, built and tested. It was not difficult to achieve near perfect balance, excellent bandwidth, low distortion and adequate gain in one stage. However, the distortion residual (isolated from the fundamental using a HP339a distortion analyzer) always showed evidence of both odd and even order harmonics. Further work revealed that suitably designed **Shunt Regulated Push-Pull** (hereafter referred to as SRPP) active loads for the 6SN7 LTP can provide sufficiently low output impedance to drive parallel pairs of triode connected KT88s with more than adequate swing at very low level of **Total Harmonic Distortion** (hereafter referred to as THD). Further, to the best that I could observe when viewing the THD residual against the fundamental on a dual trace CRO, the THD residual from this topology appeared to be primarily second harmonic. Since a spectrum analyzer was not available, this assertion has not been confirmed.

Subsequently, "Glass Audio" published an article by Perugini in which he discusses Pspice analysis of a 6SN7 SRPP having an applied B+ of 500V and for which the cathode resistor of the active load is varied over the range 1k Ω to 20k Ω .⁶ He uses Pspice simulation to demonstrate that as the cathode resistance for the active load is varied the THD dips sharply at a value around 10k Ω . (It is not quite clear to me, but I infer that the following stage grid leak resistance was maintained at 100k Ω for this portion of the analysis. An aspect of the SRPP, which is not often referred to is: To obtain the best performance, the design should be optimised for the load.) He

explains and demonstrates that the THD dip is due to reversal of the relative phase of the signal current in the active load to that delivered to the subsequent grid leak. This causes the proportions of the 2nd, 3rd and 4th harmonics to change. Germane to the present discussion is that with a cathode resistance less than 2k Ω at an input of 6Vp-p (Gain \approx 20 \therefore output \approx 120Vp-p, close to that required to drive the output stage of this design to full power), the 3rd harmonic is some 21dB down on the 2nd harmonic. (To those who are familiar with the spectral properties of the 6SN7, this result may not seem especially good but bear in mind that published spectral properties often pertain to small signal analysis; the data I have applies to an output of 2Vrms.⁷) Perugini suggests that the spectral flexibility of the SRPP could be applied in “performing power amplifiers” (his words) to apply harmonic cancellation techniques. This concept was the philosophical basis of the subject design viz; *“Minimize the tendency of the drive stage to generate odd order harmonics and ensure “perfect” balance of the phases so that the push-pull output stage is best employed to minimize the remaining even order harmonics through cancellation in the output transformer.”*

The balance of the 6SN7 SRPP LTP was not commensurate with the intention of maximizing even order harmonic cancellation in the output stage, also a further 10dB or so of gain was required. The obvious solution was to add a LTP input stage in front of the drive stage, both to increase the gain and to improve the balance. After some iterative fumbling around with curves, I established that a 6SL7 would mesh very nicely with an operating point as a function of the 6SN7 grid voltage, thereby permitting DC coupling. However, the gain would be too high. In addition, the HT voltage needed to be reduced to a more suitable level for this first stage. At this point, I hit upon the concept of connecting the 6SL7 plate loads to their respective SRPP cathodes. This circuit node is at an appropriate DC potential and permits reduction of the gain to more-or-less exactly what the required level by degeneration of the swing in the first stage plate loads. This of course, is negative feedback. I felt able to live with this approach since the feedback is applied around a simple DC coupled loop for which at audio frequencies, phase shift would not be problematic. Also, unlike applying global feedback from the output transformer secondary the fed-back signal would be entirely due to the internal voltage gain process and not influenced by the back e.m.f. from the loudspeakers(s). It has been argued that in the back e.m.f. entering the amplifier’s feedback loop via the output terminals, the amplifier is attempting to correct not only for internal non-linearity, but also for an erroneous back e.m.f. signal which will have phase anomalies due to the inductance of the voice coil, the crossover network(s) and the loudspeaker cable.⁸ This argument rings true for me and bolstered my determination not to employ global negative feedback.

So, the drive stage had matured; fully balanced with a swing capability exceeding 160% of that required to fully drive the KT88s, low output impedance, excellent bandwidth, the correct amount of gain which generates substantively even order THD. An additional benefit of this drive topology is that not only is the balance “perfect” (i.e. better than 0.1dB) but also the THD residual measures **and** looks the same at both output terminals. I have not observed these three properties coalesce with any of the alternate topologies and permutations thereof that were investigated as a part of the design process. (Note, at this juncture I had not evaluated a transformer topology.) Thus arose the hope of developing a push-pull amplifier with zero global negative feedback that would have low distortion, lots of power and good bandwidth. It also became very clear to me that the design of the power supply is **at least as important** as that of the audio circuit. An amplifier is “merely” a device which increases the energy of a signal to a

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level at which it may be re-transduced into sound. The amplifier does this by modulating current from a voltage source with the musical signal. It thus follows that the voltage source must be as constant and clean as possible. A well-designed voltage regulator can provide a source impedance of less than 1Ω from DC to frequencies beyond the audible range together with extremely high levels of ripple and noise rejection. This design embodies three regulated voltage power supplies. Heater/cathode insulation limitations mandate that several individual heater supplies are needed. While it would be possible to use a multiplicity of off-the-shelf transformers, a better solution is to use custom wound power transformers. Data for T1 & T2 is given on the schematics. Potential vendors are suggested in section 6.0.

4.0 Measured Performance of the KT88 Parallel Push-Pull Amplifier.

Measurements using 8Ω output into 8Ω load at 550V / 65mA with JJ KT88s unless otherwise stated.

Intermodulation Distortion (SMPTE 60Hz/7kHz mixed 4:1)

using unbalanced input:	520V(Svet)	550V(JJ)	590V(KT90)
5W	0.9%	0.65%	0.84%
55W	4.0%	N/M	N/M
60W	N/A	2.5%	1.7%
70W	N/A	N/A	2.5%

(The IMD measurements were taken using a Heathkit IM-5248 analyzer)

THD @ 1kHz using unbalanced input:	520V(Svet)	550V(JJ)	590V(KT90)
5W	0.32%	0.20%	0.25%
55W	1.45%	N/M	N/M
60W	N/A	1.20%	0.95%
70W	N/A	N/A	1.25%

(The THD measurements were taken using an HP339a distortion analyzer.)

Damping factor @1kHz:	3.0
Power Bandwidth (-3dB referred to 60W):	10Hz to 50kHz
Rise time:	6 μ sec (10 to 90%), 1kHz square wave
Slew Rate (60V top to top):	11V/ μ sec, (20 to 80%) 10kHz square wave
Sinusoidal sensitivity for maximum continuous output:	1.6Vrms, unbalanced; 0.8Vrms balanced
Voltage gain:	23dB unbalanced; 29dB balanced
Input impedance:	100k Ω , unbalanced; 200k Ω , balanced
Signal to noise ratio:	90dB unweighted (referred to 60W)
Power consumption: Standby	160W
Idle	350W (@ 65mA)

5.0 Design Description.

5.1 Output Stage – Refer to Figure 1.

The amplifier operates in Class A up to approximately 15Watts at the quiescent current of 65mA, which is set by adjustable bias (fixed rather than self-biasing). Above 15W the amplifier operates in Class AB1. The quiescent output tube dissipation is 35.1Watts, the plate dissipation being 33.5Watts, 20% below the rated maximum of 42Watts. To improve reliability, 330 Ω stopper resistors are used to limit the screen grid screen grid dissipation to approximately 1.6Watts, 80% below the 8Watt maximum⁹.

Dismally, experience with the current production KT88 reveals a tendency to screen grid arc-over. The genuine KT88 specification calls for a maximum screen grid design rating of 600V. Current KT88s I have tried will not sustain voltages approaching 600. (This appears to be due to poor evacuation processing. The tubes exhibit excessive blue electrostatic haze, characteristic of gas.) I have used the JJ and Svetlana types, of which the JJ type seems to be somewhat less susceptible to failure. (At 550V/50mA, the blue haze appears to fill the envelope of all the Svetlana KT88s I have. I was reluctant to press them to 65mA at 550V.) The original intention was to operate at 585/590V but I reduced to 545/550V for peace of mind. Since then I have learned that tube hi-fi industry gurus seem to concur that it is safest to operate below 520V. (Note if you set the supply at 520/525V, the plate voltage will then be less than 520V due to the DC resistance of the output transformer primary.) Performance data at three B+ levels is given. This data in conjunction with the adjustable B+ facility, is intended to facilitate your own choice of operating point. The KT90 appears to be altogether more robust. It is possible that the best choice for both performance and reliability is to use KT90s at 550V.

It is essential to source the tubes from a dealer who conducts proper burn-in and matching.

Magic Parts seem to do a reliable job with their brand, “Ruby Tubes” while Jim McShane (jimmcshane@prodigy.net) or Upscale Audio (www.upscaleaudio.com) are excellent sources. Have them to match at your desired operating voltage and current in triode connection.

Bias set-up is facilitated by the provision of individual 30.9 Ω sensing resistors in each of the four cathode circuits. The 30.9 Ω bias sensing resistors are intentionally large to allow each tube a degree of auto-bias thereby minimizing the impact of tube aging on current sharing between each phase-pair of KT88s. These resistors are ½W metal film components, which are further intended to provide additional protection to the output transformer by fusing open circuit in the event that a catastrophic failure (such as a tube arcing over while unattended) occurs. The cathode end of each of phase-pair of sensing resistors is connected together by 10K Ω resistors to form a resistive center-tap at which the averaged bias sensing voltage for each phase-pair may be measured. One potentiometer is provided for each phase-pair. The wipers of the bias potentiometers are bypassed to ground with 0.22 μ F capacitors to prevent contact rectification effects⁹. Additionally, the wiper of each potentiometer is strapped to the negative end with a 100k Ω resistor to prevent disaster in the event of wiper contact failure. The grids each have a 1k Ω stopper resistor attached directly to pin 5 of each tube socket. The bias voltage is applied to the grids of each phase-pair using a single 100k Ω resistor. The coupling capacitors are 0.22 μ F/650V etched silver alloy foil in oil types. (Note that the set-up as described is only suitable for use with matched tubes, otherwise separate test points, potentiometers and coupling capacitors need to be provided for each tube.) While it is possible to use matched phase-pairs, it is far better to obtain matched quartets of output tubes. The output tube plate impedance is matched to 4, 8 or 16 ohms using a high quality double C core output transformer. The Bartolucci Model 106 transformer chosen for

this application has a primary inductance of 300H that combined with a reflected primary impedance of 3800 Ω confers tightly controlled low distortion bass performance. The reflected primary impedance could be lower (say 2500 Ω minimum) to obtain higher power, but at the expense of increased distortion.

5.2 Drive Stage – Refer to Figure 1.

The drive circuit is a classical topology in as much as it employs two DC coupled cascaded long-tailed pairs. However, while the first LTP uses 221k Ω resistive plate loads (made up of two resistors in series to distribute the voltage drop, \approx 220V), the second LTP is provided with SRPP active loads. The SRPP active load implementation, in addition to squewing the harmonic distribution of the stage towards even order harmonics confers an output impedance (3k Ω) that is capable of driving the Miller capacitance of the triode connected parallel KT88s to an HF corner of 104kHz (drive stage bandwidth).

A single 6SL7 is used for the first LTP and two 6SN7s are used for the long-tailed SRPP pair. The 6SL7 is biased at 1mA. The SRPP active loads are biased at 7.6mA, while the second 6SN7 differential pair is biased at 6.6mA. The difference in bias current between the 6SN7 LTP and the SRPP is due to the first LTP, the plate loads of which are connected to their respective SRPP cathodes thereby providing degeneration of the voltage swing in each plate load of the first LTP. The resulting gain of the first LTP is 13dB while the gain of the second LTP is 16dB (loaded by output stage) for a total drive circuit gain of 29dB. This will drive the KT88s to full output with a single ended sinusoidal input of 1.6V_{rms} (0.8V_{rms} with balanced input). The 127V peak-peak swing required to fully drive the output stage in Class AB1 is developed at a THD of 0.29% at each of the SRPP cathodes, measured with the KT88 grid leaks connected. Swing limitation with the KT88s energized occurs during the positive excursion of the cycle as grid current comes on. This sets the limit of the output of a Class AB1 amplifier at a given operating point. To assist DC balance within the limits of typical tube side-to-side matching, 2.0k Ω individual cathode resistors are provided for the first and second LTPs.⁹ The first stage tail resistor is 127k Ω while the second stage tail is 35.75k Ω , both are returned to the negative supply rail. Both tails are made up of a four resistor series-parallel combination to handle the dissipation and voltage stress with high reliability. The use of large tail resistors in combination with cascading two LTPs compensates for the slight AC balance degradation caused by the use of individual cathode resistors in a LTP. The circuit exhibits AC balance better than 0.1dB from sub-sonic to above 100kHz, using an unbalanced input.

5.3 High Tension Power Supplies – Refer to Figures 2 & 3.

The AC is rectified using Fast Recovery Epitaxial Diodes. To delay the HT during power up, a single slow turn-on 6D22S rectifier tube is placed between the diodes and the HT regulators. (I did this to save a little space: A better scheme would be to omit the diodes and use two 6D22S or 6CG3 or 6CJ3 ½ wave rectifiers.) A 51 μ F polypropylene film reservoir capacitor is located after the 6D22S, at the input to the HT regulators. The primary requirements for an audio amplifier power supply are stability, low noise with good ripple rejection and stiff regulation across the bandwidth of the amplifier. Accordingly, dedicated regulators are provided each for the output stage (550V), the output stage bias (-100V), the drive stage (650V) and the drive stage negative (-260V) supplies. Each error amplifier is a cascode connected twin triode.⁹ This approach provides high gain (good ripple rejection, good supply stiffness) and wide bandwidth (low Miller capacitance). In this application, variations of the AC line can cause the raw DC supply to vary by +/- 50V or more due to the transformer step-up. Because the 650V and 550V regulators are

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supplied from a common source, the voltage drop across the 650V regulator was deliberately limited to avoid excessive dissipation of the 550V series pass tubes which have to handle 260mA. To avoid drop-out of the 650V regulator under low line conditions, I decided to apply negative feed-forward from the unregulated HT input to the grid of the top triode of the cascode error amplifier (analogous to a screen grid). I test for drop-out (which is indicated by a sudden 'peaking up' of ripple as the input voltage is reduced) using a variac. The 650V drive stage supply error amplifier is implemented using a 12AX7 which has a high heater/cathode insulation rating, the bias supply uses a 12AX7 also. Each regulator uses a pentode cathode follower for the series regulation element. Proper decoupling of the cathode follower screen grid supply results in superior ripple rejection when compared to a triode series regulator. A pair of matched KT90s handle the output stage HT series regulation. KT88s are also suitable. 49.9 Ω /2W plate resistors are used to assist current balance between the KT90s. To obtain HF drive bandwidth of the combined grid capacitance of the two KT90s, this regulator uses the higher transconductance 6922 for the error amplifier, thereby allowing a lower value of plate resistor (compared with a 12AX7) while maintaining the error amplifier gain. The supply to the KT90 screen grids and output stage error amplifier is filtered and smoothed using a 5H choke combined with a 20 μ F film capacitor. This also serves as the take-off point to the drive stage regulator. The resulting supply stiffness is such that the supply drops less than 1 volt from zero signal to the full output of the (Class AB1) amplifier. The drive stage HT series regulator is a 6L6GC. The drive stage 6L6GC screen grid and error amplifier supply is further smoothed and filtered using a 2.2k Ω resistor and 20 μ F film capacitor. Zener diodes are used for the voltage references. A disadvantage of zener diodes is due to the temperature coefficient which causes the regulated voltage to slowly drift upward until thermal equilibrium is achieved. This effect is accounted for in the set-up procedure. It should be noted that if an error amplifier heater should fail, the respective regulated voltage would rise towards the input voltage. Therefore, if the 0.5ampere HT supply fuse fails, look to see that the output stage regulator 6922 heaters are glowing before replacing the fuse.

5.4 Negative Power Supplies – Refer to Figure 2.

5.4.1 Drive Stage.

The DC input to the drive stage negative regulator is provided by a floating full wave supply. The negative regulator uses a 6BQ5 for the series regulator, with the cathode connected to ground. The supply to the 6BQ5 screen grid and the error amplifier is further smoothed and filtered using a 3.32k Ω resistor and 40 μ F film capacitor thereby rendering the use of a π DC input filter superfluous.

5.4.2 Output Stage.

The output stage bias is not served from the drive stage negative regulator to avoid the consequences of failure of this regulator to the output tubes or transformer. The output stage bias supply voltage of minus 100V is developed across a zener diode that forms part of a potential divider placed across the floating supply for the drive stage regulator. The cathode of the zener diode is connected to ground, the anode being connected with a 38.3k Ω resistor to the negative end of the input supply. The zener bias current is developed by completing the circuit with a 49.9k Ω resistor connected between the positive end of the supply and ground. Thus, the vital output circuit bias supply relies only upon reliable and conservatively specified silicon diodes.

5.5 Power Supply Bypass.

Each supply is bypassed by a non-inductive polypropylene film capacitor, at the point of power application to the associated circuit.

5.6 DC Heater Supplies – Refer to Figure 3.

Separate regulated DC current supplies energize the 6SL7 and 6SN7 LTP tube heaters. The SRPP heater is provided with a separate AC current supply. The AC from separate 10 volt windings is rectified using bridge connected Schottky diodes. The resulting DC is smoothed using a 4700 μ F Nichicon capacitor. The voltage is regulated to 6.3V using an LM317K mounted with a mica insulator to the chassis. To maximize the ripple rejection available from the LM317K, the reference point is decoupled using a 22 μ F tantalum capacitor. A 2.7ohm / 1 μ F Zobel network is connected across the output to improve RF rejection and counter the resonance of the LM317 device.¹⁰ The 6SL7 supply is fitted with a pseudo center tap comprised of two 1k Ω resistors, which is grounded at the signal ground point adjacent to the 6SL7 socket. As the amplifier warms up, the bias voltage develops before the B+ which will stress the 6SL7 heater / cathode insulation since little current will be flowing in the 127k tail resistor. To prevent this stress, a miniature neon lamp (NE-2H) is connected between the 6SL7 common cathode node and ground.⁹ This acts to limit the negative voltage excursion at the cathodes during warm-up to approximately -85V. Once the amplifier has warmed up, the current flow in the first LTP will cause the common cathode node potential to approximate ground level and the neon will thus extinguish. The neon has negligible capacitance and generates no noise when extinguished.

5.7 Power Supply Management – Refer to Figure 3.

Protection is provided by a 0.5ampere fuse in the HT supply center tap in addition to a 5ampere fuse in the incoming AC supply circuit.

The power supply features standby and operate modes. If the standby / operate switch is set to “Operate”, the amplifier will automatically cycle through standby during power-up.

Approximately 55-60 seconds after switch-on, the relays are energized to bring the amplifier on-line. The status lights switch from amber to green as the changeover takes place. No bumps or pops are emitted from the loudspeakers as these amplifiers power up and down. Note, if the switch is left at “Standby”, the amplifier will remain in that mode until set to “Operate”.

In standby mode, the power consumption is reduced to approximately 160Watts from 350Watts. This is accomplished through reduction of the output stage and drive stage HT voltages by approximately 150 to 175 volts (depending on voltage setting). A 6 Ω / 50W power resistor is switched in series with the AC line reducing the AC heater voltages to approximately 6 volts. This resistance is sufficient to somewhat mitigate the cold start stress on the heaters without risking cathode poisoning during standby. The amplifiers thus may be left warm and ready for use in standby for a long period without significantly compromising the useable life of the tubes.

6.0 Construction.

Most of the components used in the design are available from Michael Percy Audio www.bainbridge.net/percyaudio. T1 & T2 may be obtained custom wound through Kennedy Tube Audio (kennedyk@kta-hifi.com). Other possible sources include; Electra-Print (702-396-4909), Sowter (<http://homepages.tcp.co.uk/~sowter/>) or Bartolucci (audiomarketing@omniway.com). Note the maximum specified DCR of 45 Ω /phase for the 625V winding. The transformers should be rated for continuous commercial service (vs intermittent continuous amateur) to ensure reliability.

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The amplifiers are built on chassis measuring 18in by 9in by 3in deep. Having previously built the amplifiers on the same chassis with choke-input power supplies, these dimensions turned out to be marginal for the final design. I would suggest at least 19in by 10in. I made the chassis by cutting a 3in x 3in x 1/8th in aluminum box section to form channels having 1/2in flanges. These were attached using No6 screws to an aluminum top plate 1/8th in thick. The side rails have the flanges turned inwards while the end rails have the flanges turned outwards. The corners are locally TIG welded at the bottom to ensure rigidity. The chassis were finished by powder coating, hiding a multitude of sins.

Layout should follow the established practice of separating the power supply from the audio circuit. I bent this rule a little bit by putting the inputs on the same end as the outputs and power connection. By using a high-quality foil screened balanced signal lead and a screen around the input sockets, this did not compromise the signal to noise performance. I arranged the four output tubes to form a square around the drive tubes with the input tube off to one side. This arrangement means that one phase of the KT88 plate interconnections runs through the drive circuit. I prevented interaction between this and the drive circuit by triple insulating the wire with teflon sleeving and using grommets, routed it over the top of the chassis.

Through-flow convective cooling was accommodated by building the tube stages as modules on platforms of perforated stainless steel. These were then bolted into the chassis. Keep the output tubes at least 4 inches apart and not less than 1 1/2 inches from the output transformer. If you do not provide for through-flow convection, it would be prudent to increase these distances to 5 and 2 inches respectively.

Grounding calls for a little thought: My method is to create star ground nodes, one for each DC input filter, one for each regulator, one for the drive stages and one for the output stage. The regulator grounds are connected to each DC input ground separately, the audio stages being grounded to the appropriate regulator(s). The foil screen should be grounded at the input socket end only, and use the common wire in the lead to make the signal ground connection from the input sockets to the first audio stage. I actually isolated the electronic ground system from the chassis thereby allowing for the use of a 'ground lift' switch. This can be useful if there is a ground loop problem in your line supply arrangement. The connection between the signal ground and earth was made via the switch at the input sockets. Some people like to place a 10nF ceramic capacitor across the ground lift switch, presumably for RF purposes.

THE CHASSIS MUST REMAIN GROUNDED VIA THE LINE CORD.

The chassis ground connection should be made adjacent to the line IEC receptacle.

Grid stopper resistors should be attached directly and closely to the grid pin(s) of the tubes sockets. Similarly, the 10nF screen capacitors used in the power supplies should be attached directly to the screen grid pins.

For anyone wishing to build the design whom is wary of voltage regulators, consider starting with the negative regulator since this is slightly less complex yet similar to the high-voltage regulators.

SAFETY: Allow a minimum of three minutes for the polypropylene capacitors to bleed down via the bleed resistors. DO NOT WORK on the circuits without REMOVING THE LINE CORD FROM THE IEC RECEPTACLE. Use a voltmeter to check that the CAPACITORS HAVE DISCHARGED.

Having got the negative regulator working, complete and check out the power supply. (The regulators control without the load applied, so you can check their operation at this stage.)

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Initially, set each B+ regulator at the lowest voltage. Set the negative regulator to -255/-260V. Final adjustment is a part of the amplifier set-up procedure. Having successfully completed the power supply, move on to the audio stages.

The set-up procedure for the completed amplifier is given on the amplifier schematic.

The author cannot accept responsibility for your ability or otherwise to work safely with the high voltages present in this design.

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