The aim of this article is to expand on some power supply issues related to valve amps that imho can benefit from additional detailed description.

The current rating of 1N4007 & UF4007 diodes are assessed for both continuous amplifier operation and amplifier turn-on. Comments are also made on SiC and Zener diodes.

The peak inverse voltage (PIV) capability of 1N4007 & UF4007 are assessed, including series connections.

The influence of power transformer secondary winding leakage inductance is assessed, and three practical means to alleviate commutation transients are discussed.

The practical use of the computer application PSUD2 to assessing valve diodes is presented. Datasheet details of GZ34, GZ37, GZ32, 5R4GY valve rectifier diodes are discussed, as an example of how to interpret datasheet limiting values, and how to use PSUD2 to assess diode applications in 'non-standard' circuits.

To simplify these discussions, only the commonly used capacitor-input filter power supply is discussed.

Power supply start-up sequencing issues are described.

The surge voltage capability of electrolytic capacitors is assessed.

Operational issues related to dual diode valves are described, including heater connections, emission and PIV degradation, PIV testing, and fitting 1N4007 in series.

DC powering of heaters using available 5V and 6.3Vac windings is discussed, including the use of small switchmode boost modules to generate 12.6Vdc. DC power from output stage cathode biasing is also described.

Fusing protection for valve amps is discussed in [13].

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Contents

1 1N4007 / UF4007 diode current rating

The 1N4007 is commonly used in valve amp power supplies, and the UF4007 is gaining favour fast. Few DIY'ers ever 'design' in that diode, but rather just choose it because others have used it, and it seems obvious that a '1A diode' is going to be appropriate for a 100-200mA power supply load.

If the diode datasheet was viewed, it doesn't present ratings that simply relate to an actual power supply rectifier application. So the aim of this section is to identify what can be gleaned from the datasheet, and how can that be applied to power supply design.

The 1N4007/UF4007 datasheets have two notable absolute max (ie. limit) current ratings - $I_{F(av)}$ of 1A, and I_{FSM} of 30A - and a derating chart for ambient temperature. These limits are now explored further, with specific relation to amplifier continuous operation and amplifier turn-on.

Power supply design is made eminently simple using the simulation application PSUD2 [2], which is used in this article, although some are using the more powerful simulator LTSpice for tube amplifier design.

1.1 Amplifier continuous operation

Unfortunately, there is no simple way to define a diode's continuous current capability in a tube amp because that rating depends on a variety of power system part values, including the type of filter circuit and loading, and the operating junction temperature Tj of the diode.

A continuous operating current capability is determined by the maximum allowed junction temperature Tj, which should be some reasonable value below the absolute maximum rating of 175^{o}C. Many designers will settle on a maximum design Tj value (such as 125^{o}C), and then derate further for known considerations – which can be many and varied. Junction temperature is related back to the local ambient temperature by the average power dissipation in the diode device, and the device's thermal resistance – these two characteristics are now discussed in more detail.

Power dissipation due to peak current

The 1N4007 datasheet parameter $I_{F(av)}$ refers to an absolute max 1A average current through a leaded diode whose leads are thermally clamped 9.5mm from the diode, and with the diode in a 75^{o}C ambient free-air environment (55^{o}C for UF4007). But this 1A average current level is based on a basic half-wave rectifier circuit with a resistive load where the diode current waveform is a 60Hz half-sine-wave with a peak of 3.14 times the average (ie. 3.14Apk, and 1.57Arms), and for Tj=150°C, close to its operating max of 175^{o}C. The PSUD2 [2] screengrab in Figure 1 shows this simple circuit, where the simulated diode current I(D1) has a 3.14A max (peak), 1.57Arms, and 1.0A mean (average). However, this half-wave rectifier circuit is not used in valve amplifiers, so we need to pursue further performance aspects to assist design.

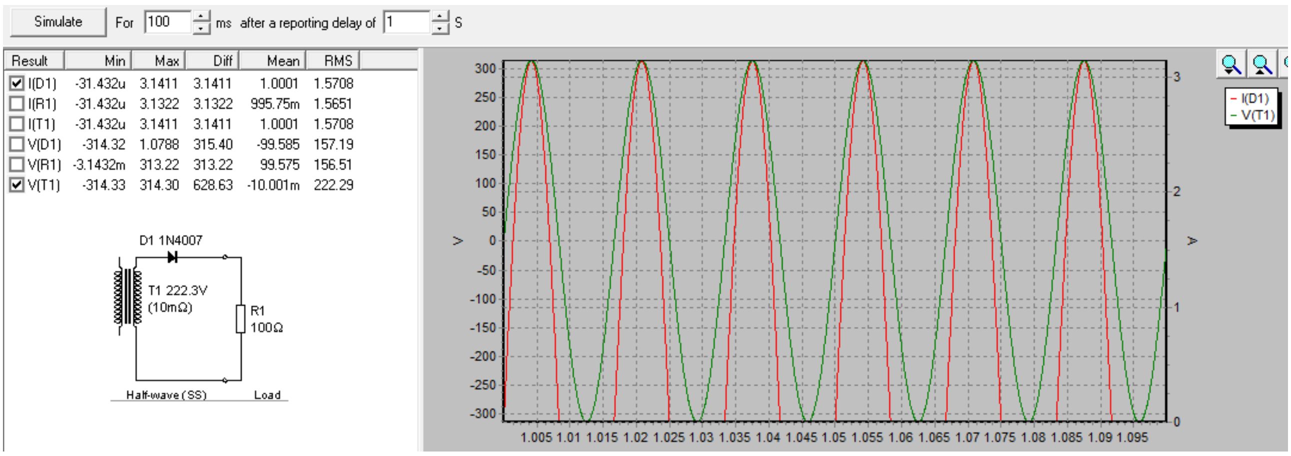


Figure 1. IN4007 simulated current waveform (red) for datasheet continuous current rating.

The current waveform in a capacitor input filter is very peaky (ratio of peak to average current >>3.14). The 'peakier' the waveform, the more energy is dissipated in the die per cycle (ie. the higher the average power), even though the same average DC current is rectified by the diode.

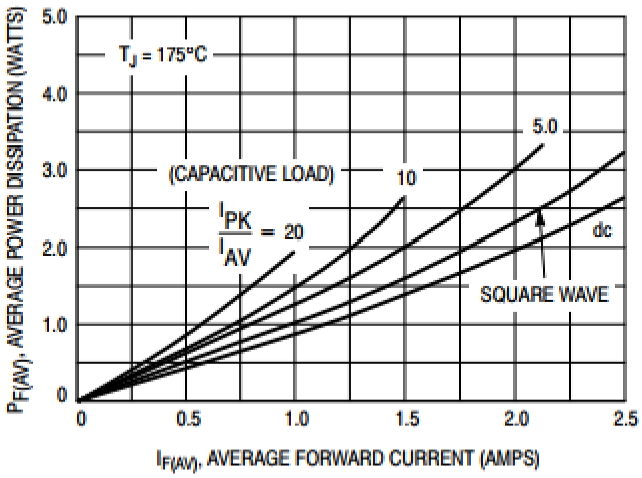


Figure 2. MUR160 power dissipation versus forward current, from [4].

The MUR120 series 1A diode datasheet [4] has additional design information to that of the 1N4007 datasheet, and allows a better awareness of the influence of peaky current waveforms on power dissipation, and on how mounting a 1A diode influences the die junction temperature rise.

For the MUR160, Figure 2 indicates that a 1A average current in a half-wave circuit will dissipate Pd ~ 1.0W (about the same as for a square-wave current), whereas if the diode had a peak current of 20A (Ipk/Iav=20), then the diode dissipation increases to about Pd ~ 2.0W. Note that this is when Tj=175^{o}C, which will cause the lowest dissipation (from forward voltage drop), but is too high to be a practical design aim.

Peak to average current ratio

The peak-to-average current ratio in a particular rectifier circuit can be estimated using Schade's design charts [5], as shown in Figure 3. Schade related filter capacitance and load resistance to the effective source resistance of the power transformer. PSUD2 can also calculate the peak and the mean currents for any particular circuit, allowing a simpler way to estimate peak-to-average current ratio.

Figure 3 shows two operating points from two power supplies that use the same rectifier, capacitive filter and load values, where the n.ω.C.R_L value is held constant at 75 (for n=2, ω=377, C=100uF, R_L=1kΩ).

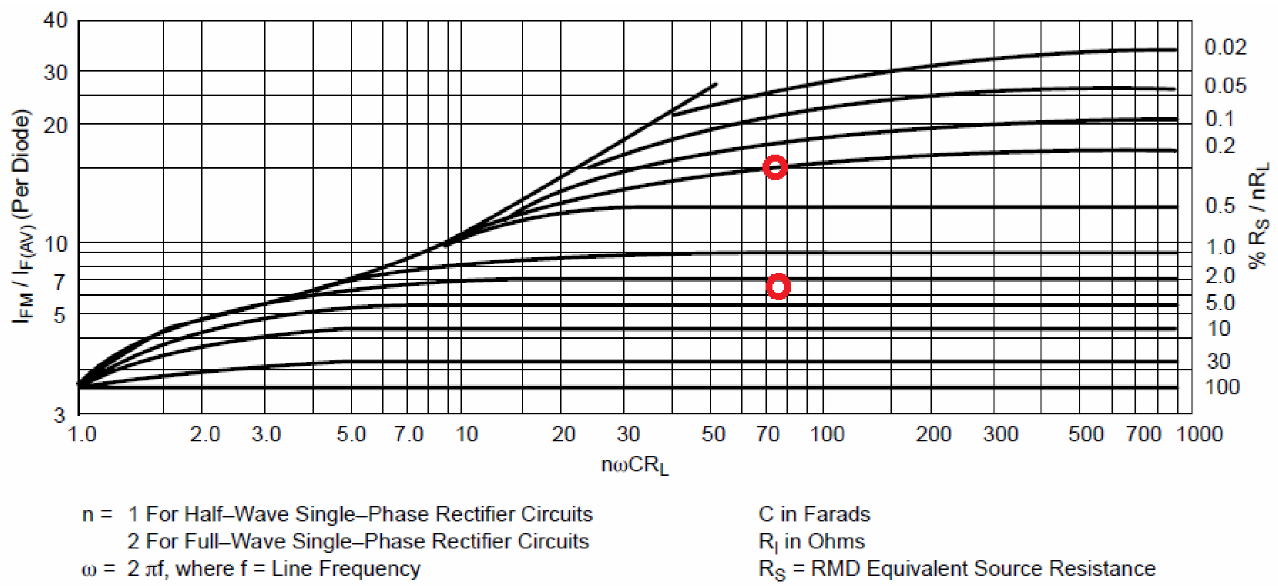


Figure 3. Schade's curve for peak current to average current ratio from [1], with 2 power supply operating points identified.

PSUD2 screengrabs from the two power supplies are shown in Figure 4, with the first screengrab showing diode Ipk=1.47A, and Iav=0.182mA, giving a peak-to-average ratio of 8, for an Rs/nR_L ratio of 2.5%. The second screengrab shows diode Ipk=3.14A, and Iav=0.182mA, giving a peak-to-average ratio of 17, for an Rs/nR_L ratio of 0.2%. The output DC voltage (~364V), and hence output DC current (~364mA) into R_L is effectively the same, although the power transformer voltages and effective source resistances are significantly different. When comparing Figure 2 for Ipk/Iav = 20 and 10 for the same average current level, the diode in the Rs/nR_L = 0.2% circuit is likely to dissipate noticeably more power.

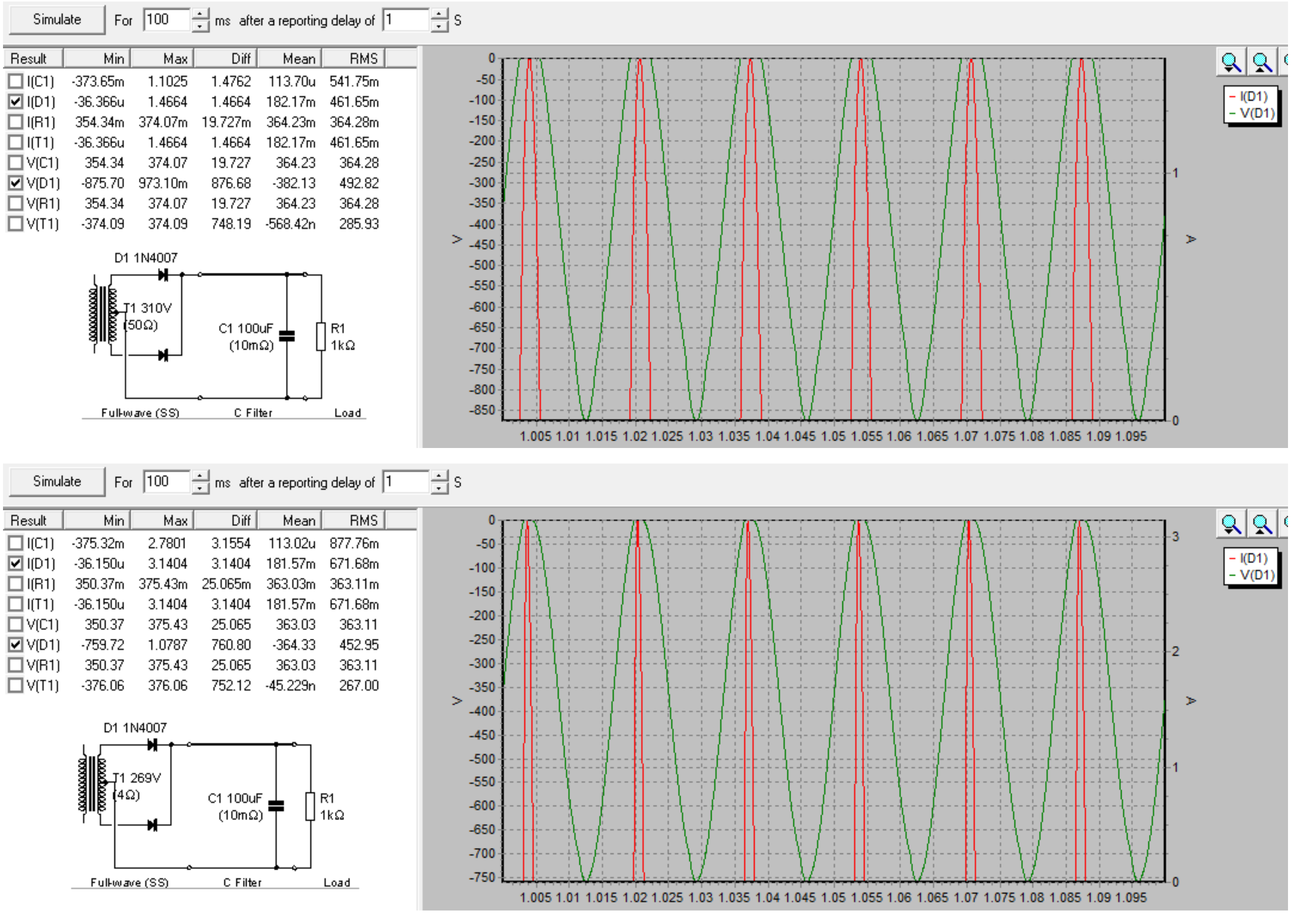


Figure 4. IN4007 simulated current waveform (red) for capacitor input circuit examples.

Comparison of diode forward voltage (Vf) is somewhat limited by comparable datasheet information. At Tj=25^{o}C, Table 1 indicates forward voltage and effective resistance for the 1N4007, UF4007 and MUR160 diodes. As Tj increases for any diode, its Vf characteristic curve for instantaneous current lowers, and it is expected that the three diodes would show a similar shift in Vf with Tj (although only the MUR160 datasheet [4] shows that shift).

Diode (Tj=25^{o}C)	1N4007	UF4007	MUR160
Vf at 1A (instantaneous)	0.94V	1.48V	1.02V
Vf at 3A (instantaneous)	1.06V	1.80V	1.20V
Resistance (average between 1A to 2A)	~0.07Ω	~0.13Ω	~0.11Ω
Resistance (average between 5A to 10A)	~0.022Ω	~0.8Ω (estimate)	~0.044Ω

Table 1. Nominal diode values from datasheets.

Power Dissipation & Forward Current Derating Curve

The 1N4007 forward current derating curve for ambient temperature is shown in Figure 5, and indicates that

Tj reaches 150^oC when ambient is 75^oC and Iav is 1A with resistive load. For the UF4007, the derating curves starts at 55^oC, and finishes at 150^oC. The derating curve is for a part in free air with 3/8" leads.

Figure 6 presents the MUR120 series datasheet thermal resistance levels for 3 lengths of lead in a simple mounting scheme. This data is assumed to be for Tj=175^oC, and the ambient Tamb=25^oC. The 1N4007, UF4007 and MUR160 all have the same JEDEC case and lead style and dimensions, so should have similar thermal resistance performance to ambient. These diodes are typically available in two case styles, DO-41 and A-405, however no datasheets indicate any change in thermal or current ratings for A-405 parts with the thinner leads, which typically have a /L appendix to the model number.

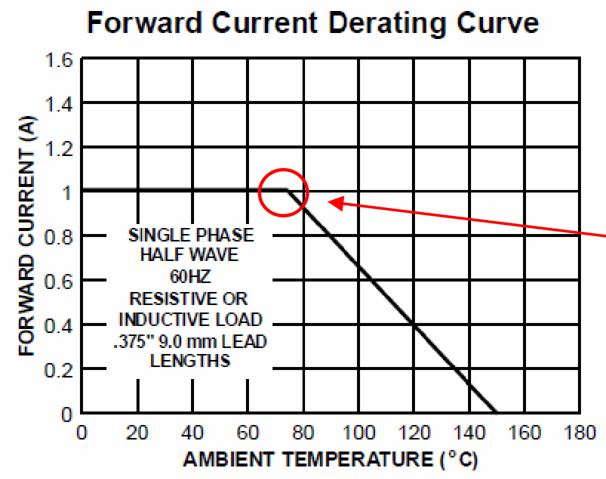


Figure 5. 1N4007 forward current derating curve, from datasheet.

Infer power dissipation at this operating point.

So the 1N4007 power dissipation Pd in Figure 5 can be inferred by noting the expected thermal resistance Rth (from MUR160), and calculating Pd = (150^oC – 75^oC)/Rth. For 0.375" lead lengths, the MUR120 indicates Rth ~ 84^oC/W (although this could be increased somewhat given a higher Tamb and a lower Tj for the 1N4007 operating conditions). Hence from Figure 5, 1N4007 Pd ~ 0.9W for 1A Iav, Tamb=75^oC and Tj=150^oC.

TYPICAL VALUES FOR $R_{\theta JA}$ IN STILL AIR

Mounting Method		Lead Length, L (in.)			Units
		1/8	1/4	1/2	
2	$R_{\theta JA}$	67	80	87	°C/W

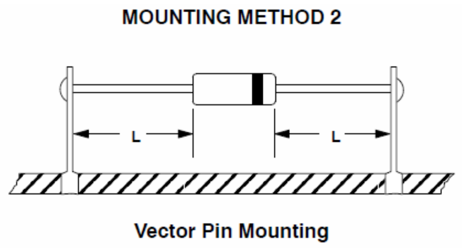


Figure 6. MUR120 series thermal resistance versus mounting scheme [4].

For the UF4007, Pd = (150 – 55)/Rth ~ 1.12W, for 1A Iav and nominal Tj=150^oC. The increased dissipation in the UF4007 is consistent with the higher datasheet Vf values in Table 1.

Previously, Figure 2 for MUR160 had indicated Pd~1.0W for 1A Iav, but at Tj=175C. This dissipation level is also consistent with estimated Pd from the 1N4007 and UF4007 and the datasheet Vf values in Table 1.

For design, Tj would aim for a lower value than 150^oC.

Design Outcomes

A design for continuous operation would likely consider the following factors:

- Maximum allowable diode power dissipation Pd (W), based on:
    - Maximum ambient temperature (perhaps design for at least 50°C, and possibly up to 70^oC).
    - Maximum junction temperature (perhaps don't aim to exceed 125^oC).

- Thermal resistance (perhaps anticipate up to 90ºC/W).
- Pd = 0.6W for Tamb=70ºC, Tj=125ºC, Rth=90ºC/W, as a likely design maximum.

- Ipk/Iav level (resulting from PSUD2 or Schade chart).
  - A combination of Ipk/Iav, and Pd, can be cross-checked for Iav in Figure 2.
    - The chart in Figure 2 is considered appropriate for 1N4007, as lower Tj would counter-balance the lower Pd.
    - The chart in Figure 2 would need to derate Iav by about 50% for UF4007, due to lower Tj and higher Pd.
  - An example combination of Ipk/Iav = 20 and Pd = 0.6W indicates a max Iav ~ 0.3A for a 1N4007, and max Iav ~ 0.2A for a UF4007.
  - An example combination of Ipk/Iav = 10 and Pd = 0.6W indicates a max Iav ~ 0.45A for a 1N4007, and max Iav ~ 0.3A for a UF4007.

Series connection of diodes

Diodes may be connected in series to allow for a higher effective PIV capability. If the series diodes are connected closely together (ie. leads soldered together with diode packages within say ½") then the combined power dissipation from the diodes will raise operating Tj, as thermal resistance is unlikely to reduce by much. The outcome is likely a 50% derating of Iav from the one diode situation, as Pd would double, but Rth may not significantly reduce.

Bridge connection of diodes

Modern bridge connected diodes in a single package, such as the Vishay W10G, have a different mix of thermal deratings. The datasheet already integrates the effect of Ipk/Iav in a derating curve. For example, the Ipk/Iav=20 curve indicates Tj=150ºC for Iav=0.98 and Ta=25ºC, with Rja ~ 36ºC/W, and hence about Pd = (150-25)/36 = 3.5W, or 0.87W per diode. Derating for Tamb=70ºC, and a Tj=125ºC as a maximum design capability, the max Iav ~ 44% of 0.98 = 0.43 A.

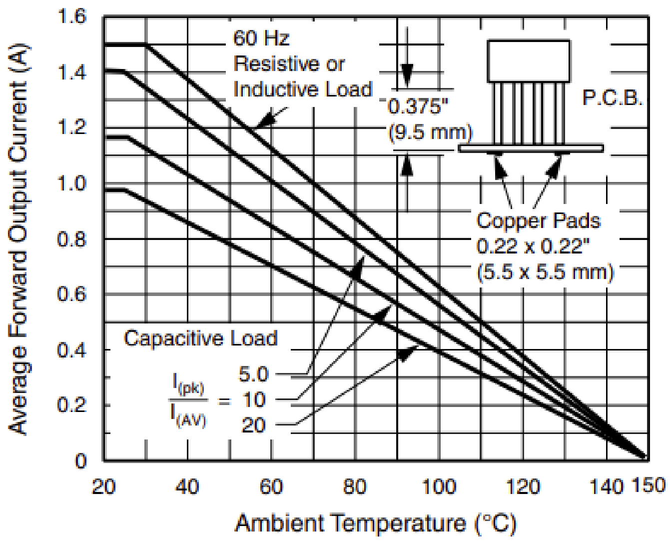


Figure 7. Vishay W10G forward current derating curve, from datasheet.

Reverse leakage current

A higher operating junction temperature has the disadvantage of increasing reverse leakage current as the diode withstands reverse voltage. The reverse current in a 1N4007 is shown in Figure 8, and for a capacitor input filter power supply the reverse current could reach 300uA at 1kV PIV at max operating Tj=150°C, equivalent to an additional 0.3Wpk dissipation during the off-state period. Due to this dissipation resulting from reverse voltage multiplied by reverse current, the peak dissipation level has a small duty-cycle, and so the average dissipation is quite small (ie. <0.05W). The UF4007 has approximately 4x higher reverse current than the 1N4007. Keeping Tj below 125°C is likely to make this issue insignificant.

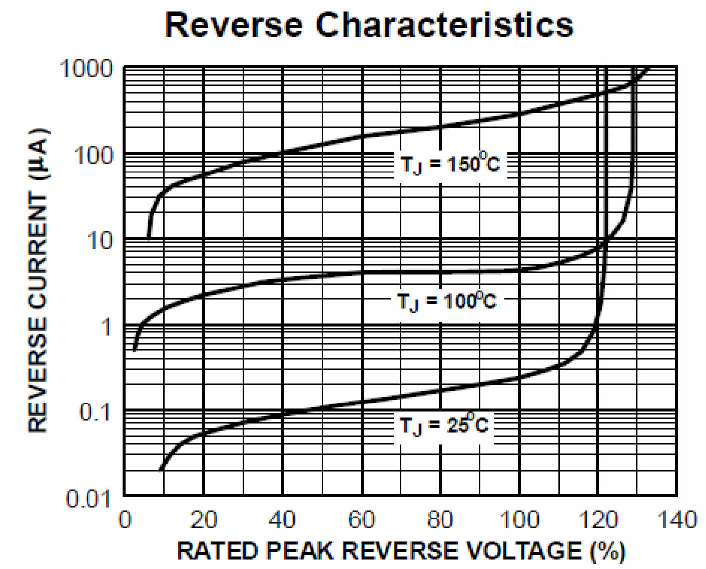


Figure 8. 1N4007 reverse current versus inverse voltage, from datasheet.

1.2 Amplifier turn-on

At amplifier turn-on, the diodes in the commonly used capacitor-filtered power supply experience a short duration series of high peak current waveforms, prior to the power supply and amplifier transitioning to a more steady-state condition when peak current levels are lower. PSUD2 allows a good estimate of this turn-on current characteristic.

A rectifier circuit (bridge or full-wave or half-wave or doubler) forces any diode in the rectifier to experience one current pulse each mains cycle (ie. each 16.6 or 20ms). The type of filter and load circuit dictates the short duration behaviour of the current pulses, ranging from a rapidly declining short sequence of pulse magnitudes to a combination of initial peaks followed by oscillatory behaviour.

Datasheet rating

The datasheet parameter I_{FSM} refers to a 30A peak single half-sine-wave current 'pulse' of 8.3ms duration that is non-repetitive. The key points to note are:

- As a non-repetitive pulse, any subsequent current pulses of sufficient magnitude may raise the die temperature beyond the rated max level and hence over-stress the diode. As such, a sequence of current pulses, as can be experienced during amplifier turn-on, would require sufficient derating of the initial peak level so that the combined sequence of pulses do not exceed die temperature max limit.
- The datasheet expands on this non-repetitive rating with a chart showing how many successive pulses of a set peak current value would similarly raise Tj to its max limit, based on those pulses occuring every 16.6ms. Some estimation is then required to translate this data to a set of successive pulses that don't all have a constant set peak value.
- Those of us on 50Hz power have a 10ms waveform width that increases pulse power dissipation, and need to apply some derating for turn-on current surge rating.

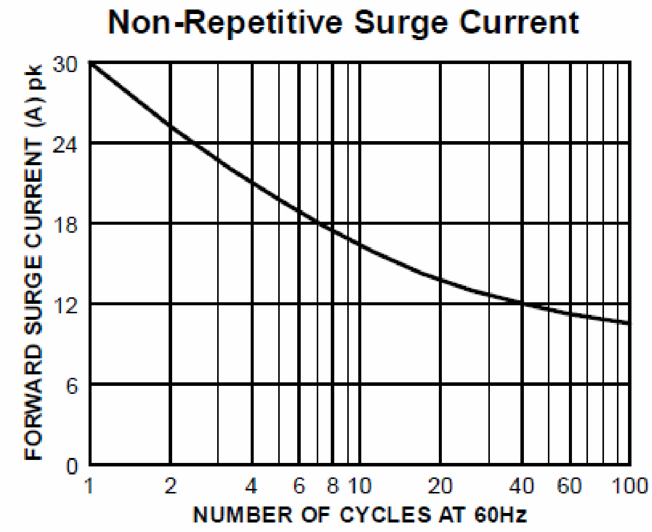


Figure 9. IN4007 surge current capability, from datasheet.

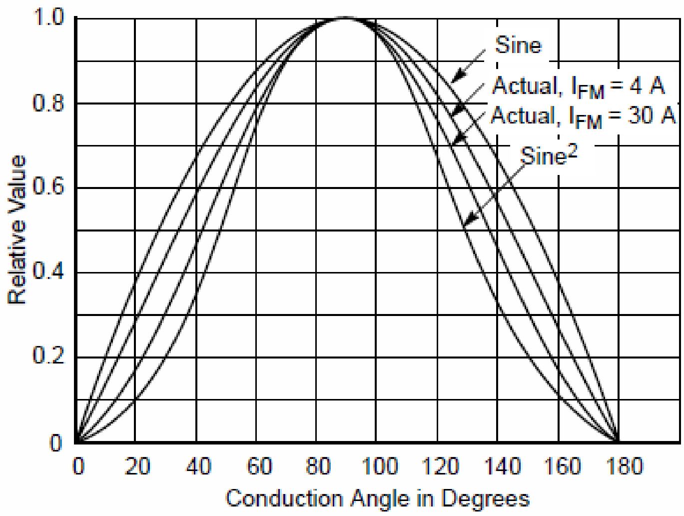


Figure 10. Example measured diode power waveform for half-sine-wave current, normalised to 100% (figure 73 from [1]).

Of note is the actual diode power waveform as illustrated in Figure 10. A relatively low current level causes a somewhat constant on-voltage and hence sine shaped power dissipation curve, whereas a relatively high current level causes a more resistance dominated voltage and hence a 'thinner' shape towards a sine-squared shape. For the 30A limit condition, the power waveform will peak to quite a high level over a duration of a few milliseconds. The 30A is a JEDEC peak surge rating, which is stressful to the part, and the part may only handle up to 100 such surge events before damage. 30Apk would need to be derated if it was used for turn-on assessment, as service life would likely extend beyond 100 turn-on events.

Simulation of turn-on peak current

Any simulation result of Ipk at amplifier turn-on must be regarded with some caution.

Simulation is based on a simplified representation of parts in a circuit. Even though the power transformer is modelled by a voltage transfer ratio, and winding resistances, other complex influences are typically ignored such as the mains voltage waveform point at turn-on, or the transient inductance of the primary due to in-rush [3], or leakage inductance of the secondary winding, or practical considerations such as switch contact bounce.

The simulated load is also problematic, as a worst-case condition may be a hot turn-on event where the output stage valves instantly conduct (due to a hot cathode) depending on bias voltage, and the power supply filter capacitors may have almost fully discharged in the short interval between turning the amp off and then on again. The output stage bias may have similarly discharged (eg. for cathode bias), and so the loading at hot turn-on may be the max zero-bias current drawn from both sides of a push-pull circuit (ie. more than the expected current drawn from the output stage during over-drive conditions) – although this would be subject to L/R rise time from the output transformer (which may only be a few millisecond such as for 10H half-primary inductance, and 50V/150mA = 300Ω conduction resistance).

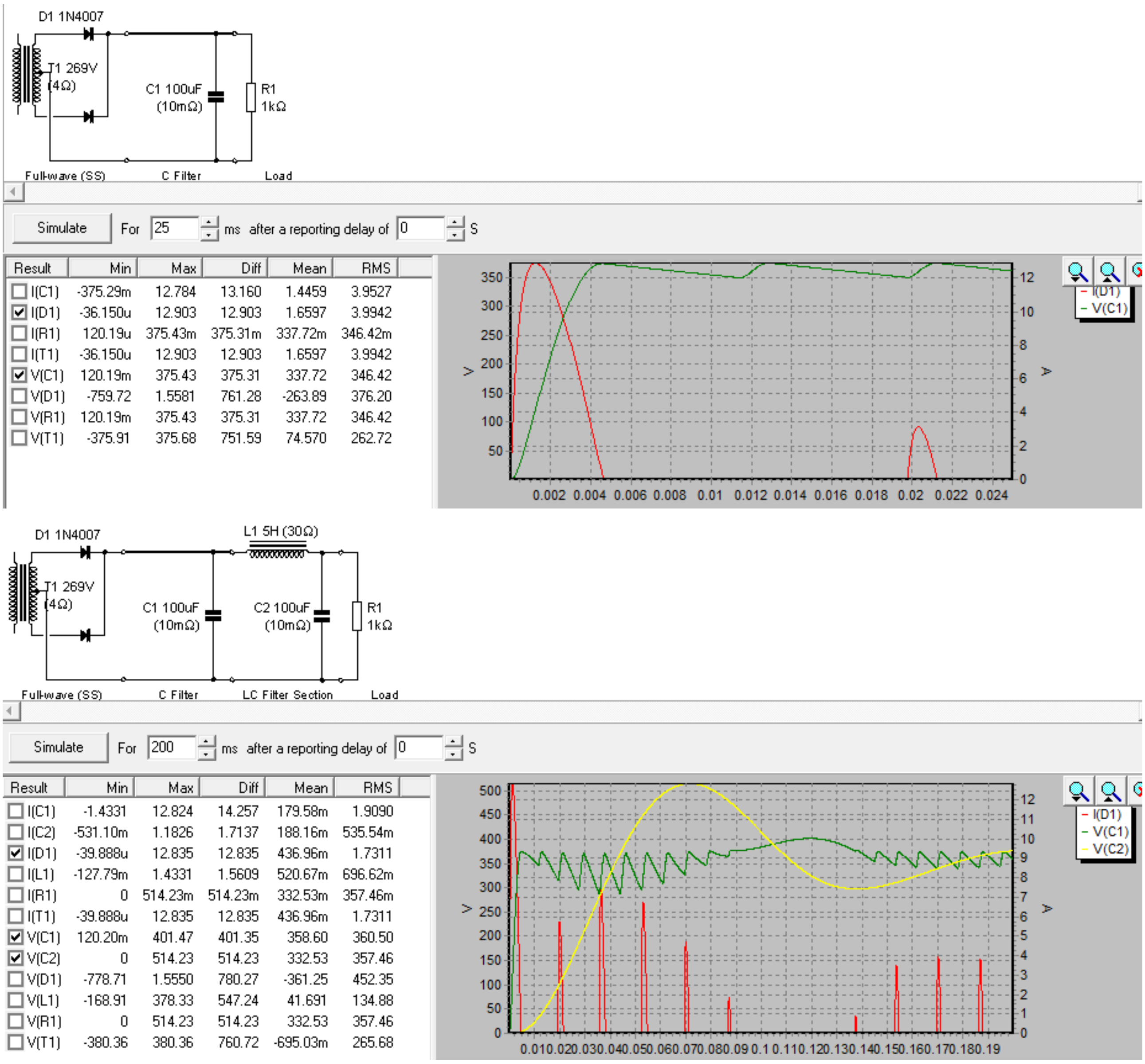


Figure 11. IN4007 peak current waveform (red) at amplifier turn-on for two forms of filter circuits.

The simulated filter parts may also show non-ideal behaviour, such as an inductor's inductance that can vary with DC current and may enter a saturation region during part of the AC voltage waveform across the inductor (choke).

Any simulation introduces boundary conditions that may change over time in the real world, such as mains voltage. An increase in mains voltage can be simulated to indicate sensitivity for diode current levels.

To indicate the peak current that could plausibly be expected for a valve amp power supply, Figure 11 presents 2 simulations for the same power transformer, rectifier, and first filter capacitor, but the second result is for a CLC filter. As simulated for continuous operation in Figure 4(b), the power transformer has a very low effective winding resistance (below anything I've come across) with an Rs/nR_L ratio of 0.2%.

For the simple C-only filter simulation, the initial current surge is 13Apk, and the duration of the pulse is about 5ms, so the energy absorbed by the diode is significantly less than the max rated 30Apk with an 8.3ms pulse duration.

For the C-L-C filter simulation, the initial current surge is still 13Apk, but the diode experiences subsequent pulses at significant levels due to the charging characteristic of the second LC filter circuit. From Figure 9, the diode could survive 20 pulses with 13Apk, although that capability should be reduced due to all but one of the factors described in this section.

Design Outcome

From a design perspective, peak current derating is required, but is difficult to quantify as only simulation is likely available to calculate the anticipated peak current level, and simulation results need some caution.

Just the concern about the damage capability of the 30Apk rating for 1N4007 / UF4007 would indicate that a derating from 1 to 2 cycles in Figure 9 is a minimum (ie. to 26Apk), and derating by another cycle to 3 would be reasonable for rapid turn-off turn-on switching (ie. to 22Apk).

Furthermore, derating by another cycle for 50Hz mains power (ie. to 21Apk), and derating by another cycle for circuits that cause multiple pulses (ie. to 20Apk), would seem appropriate.

Finally, some caution on how the peak value is determined may also suggest a need for further derating.

Comment on use of SiC diodes

Silicon carbide Schottky diodes have become mainstream for use in switchmode power supply applications over recent years. They have particular characteristics that suit switchmode converter operation.

However, they are not easily suited for mains frequency rectification applications such as in valve amps, as a typical datasheet indicates they have a relatively low repetitive peak forward surge current rating when compared to the main current rating displayed for a part, and their current ratings are predicated on heatsinking the part. As such, there is a need to cross compare datasheet details to clarify what ratings are appropriate for a particular valve amplifier's power supply, especially if the part is mounted in air without heatsinking.

For example, the Cree C4D02120A is the lowest current rating 1.2kV PIV device listed, and has a nominal device selection current rating of 2A, although the datasheet shows the rating is for continuous forward current at Tc=165°C, and the datasheet banner identifies a 5A rating (which is seen in the details as the continuous current rating for Tc=135°C operation). This is a large TO-220 package device, and in switchmode applications would be mounted on a heatsink.

However, it is likely that many DIYers would not use a heatsink, as the case is the cathode. So this discussion is about a free-standing part which is likely to be located under a chassis where local ambient temperature could easily be Tamb=50°C (and possibly up to the 70°C as used in earlier discussion on 1N4007). The thermal resistance Tj-c is given as 2.5°C/W, however no nominal Tc-a value is given for a free-standing device, so TO-220 case-to-ambient thermal resistance is likely to be circa 70°C/W. As such, 1W of dissipation would lead to Tc = 50°C + 1x70°C/W = 120°C (the junction temperature is pretty much the same as the case temperature). For some designers, Tj=125°C could be considered a maximum safe level given that Tamb could well be higher. A 1W dissipation capability is not much more than the 0.6W dissipation outcome for a 1N4007 as discussed earlier.

In alignment with a 1N4007 type datasheet, the SiC datasheet does provide a 10ms half-sine pulse repetitive peak current rating of 8.4A for Tc=135°C, but there is no confirmation if that relates to a 50Hz repetition (20ms period). The datasheet current derating graph shows the peak forward current rating versus Tc for a variety of duty-cycle operational curves – where it appears that the waveform would be a squarewave, as for Tc=135°C the peak current is shown as about 7A for a 50% duty-cycle. The diode on-voltage is somewhat comparable to a 'fast recovery' pn power diode, but increases with junction temperature for conduction level above about 1A, as compared to the pn diode whose on-voltage decreases with junction temperature. There is no datasheet clarification of what power dissipation would occur for particular duty-cycle operational curves.

With only 1W dissipation, and an on-voltage that is likely to be significantly higher than say a UF4007, the rectification current capability of the SiC diode may be no better than a 1N4007 or UF4007 for capacitor input filters, and may be not much greater than 1A for choke input filters.

Comment on use of Zener diodes

A not uncommon practise for DIYers is to use Zener diodes to reduce B+ rail voltage. Similar concerns are raised with Zener diode current capability as with rectifier diodes like 1N4007 and UF4007, including power dissipation due to peak current, forward current derating, and series connection of Zener diodes.

The same process of derating needs to be applied to a Zener diode, given the likely datasheet max junction operating temperature is 150°C and the default datasheet derating curve goes to 0W at 200°C. The thermal resistance of series connected parts can be quite high and determined by convection rather than lead conduction, and the maximum continuous current in the application likely needs to be measured.

For Zener diodes connected closely together (ie. leads soldered together with diode packages within say ½") then the combined power dissipation from the diodes will raise operating Tj, as thermal resistance is unlikely to reduce by much. The outcome is likely a 50% derating of power dissipation from the one diode situation, as Pd would double, but Rth may not significantly reduce.

A reasonable derating for temperature would be 40%, based on Tj=200°C reduced to 125°C, and Tamb=75°C. So two Zener diodes in series may need to be derated to 20% (ie. 1W for a 5W part).

The situation is more onerous when the Zener is subject to rectifier pulse currents, such as when located in the CT link of a power transformer secondary, where the diode is subject to peaky current pulses rather than a continuous current. In this situation the Zener is not as robust as a typical rectifier diode to handling the peak level of current, and may need to be further derated based on the following:

- Inspection of a typical 15V 5W part datasheet indicates the maximum peak non-recurring current (6.3A) is about 20x the nominal max current rating (0.32A). Given a turn-on event can have multiple current peaks that are higher than the allowed continuous peak level, the 6.3A peak rating needs to be derated for multiple peaks, as well as generally derated to provide margin from a worst-case limit. Simulation like PSUD2 is one method to estimate the turn-on peak in an amp's power supply.
- A meter should be used to measure the true-rms current through the Zener (for the purpose of calculating power dissipation). Note that the measurement may be significantly inaccurate due to high crest-factor (ie. check the meter's crest factor rating – it should be at least 10:1).
- Although the Zener will constrain its voltage rise at the peak of a current pulse, the effective Zener voltage (for the purpose of calculating power dissipation) is likely 5-10% higher than the rated Zener voltage.

For some applications of a Zener in the CT link, there may be a circuit configuration where the Zener doesn't pass rectifier pulse currents – see aikenamp's last circuit on backbiasing.

2 1N4007 / UF4007 diode PIV capability

The 1N4007 and UF4007 have a rated peak inverse voltage (PIV) rating of 1kV at 25°C. The reverse current versus reverse voltage characteristics shown in Figure 8 covers the likely range of Tj experienced in an amplifier application, with the 1N4007 having a rated 100% PIV = 1,000V. The manufacturer determines a PIV rating based on a variety of considerations, including batch tolerances, and continuous cyclic operation at maximum Tj (resulting from some level of operating Iav, Tamb and thermal resistance), followed by application of rated PIV.

The reverse characteristics indicate that reverse current has a typically constant low level up to 100% PIV rating, and a typical over-voltage margin of at least 120%, beyond which reverse current starts to avalanche increase to whatever current can be sustained in the circuit with the voltage being maintained.

A common recommendation is for diodes to include some level of PIV derating to account for a possible increase beyond the expected maximum AC mains voltage for the equipment, for example from a mains voltage swell event.

Mains borne transients could couple to the secondary winding and stress diode PIV, however this may be unlikely especially when the amplifier has a MOV on the primary winding, or the transformer has an electrostatic screen, or the secondary winding has a snubber network across it, or the secondary winding has substantial shunt capacitance, or the transient occurs away from the sinewave peak.

In Australia, the nominal rated mains voltage is 230V, although most domestic residences operate at around 240V, and it is not too uncommon to see up to 260V (due to solar PV generation or grid feeder tap setting). As such, a power transformer with a 280V-0-280V secondary for 230VAC mains may generate about 790V PIV on each diode in a full-wave rectifier (circuit in Figure 4). If the mains increased to 260VAC, then PIV would increase to about 900V. A +10% swell, or surge, in mains voltage could then expose a 1N4007 to its max rated PIV. Imho, at least a 10% derating (based on max applied mains VAC) such as exampled is worthwhile, and so some amplifier power supplies would need to assess how to increase their diode PIV capability if it exceeds 1kV and common 1kV rated diodes are being used.

Software such as PSUD2 can indicate the PIV reached in a circuit at the maximum expected AC mains voltage, and for some level of overvoltage, by adjusting the secondary winding voltage.

2.1 Series connected diodes

A common means to increase diode PIV is to connect two (or more) 1N4007 in series. The voltage across each diode in the series string will depend on:

- two static parameters
  - reverse leakage
  - reverse breakdown on-set
- three dynamic parameters
  - turn-on time
  - reverse-recovery charge Qrr
  - capacitance

The static parameters dictate sharing for constant or relatively slow changing voltage waveforms such as from mains AC frequency. The dynamic parameters become significant when there is a large transient change in voltage across the total diode string, where dV/dt is significant.

Operating above PIV rating, a normal junction power diode runs the risk of localised avalanche current conduction causing semiconductor die damage, compared to an avalanche rated diode that is specifically designed to spread avalanche current evenly over the die. Unfortunately, there is no known literature discussing this risk, especially where reverse current is constrained to no more than typical max level at rated PIV and max Tj, such as experienced when approaching the on-set of avalanche when Tj < 150°C.

Static balancing

Figure 12 is a graphical representation of dispersion in device leakage current, where three devices support the total PIV (ie. PIV = Vn + V2 + V1). At some increasing level of total PIV, D1 will approach the on-set of avalanche and its voltage won't increase at the same rate as the others (ie. the total PIV is increasingly supported by the other 2 diodes) as indicated by the red dashed line in Figure 12. As total PIV continues to increase, the string current (and hence each device's current) is still constrained to be within the normal range of reverse current. As such, the total PIV would have to extend beyond 3x the PIV (when all series diodes start to exhibit avalanche current) for diode avalanche current to increase significantly. Device leakage current dispersion is minimised by using only new parts from the same batch, by ensuring periodic maintenance to keep the diodes clean so that no build-up of surface pollution became significant (ie. increasing leakage current dispersion), and by keeping diode junction temperatures similar.

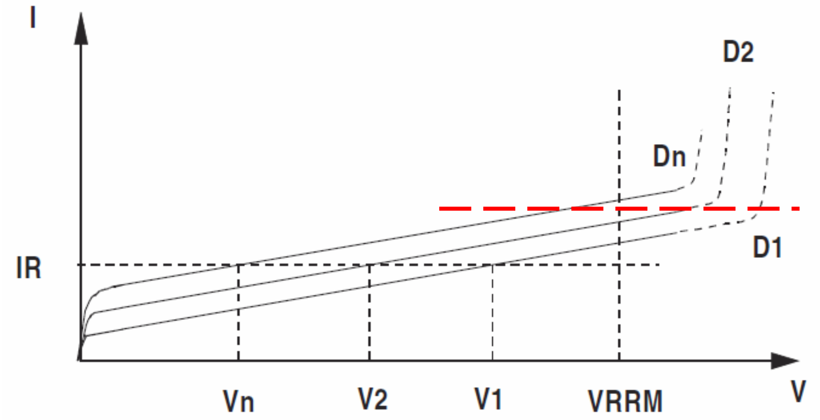


Figure 12. Graphical representation from [6] of diode leakage current dispersion.

Connecting diodes in close proximity (ie. leads soldered together within about ½") is the simplest means to ensure Tj is effectively the same for all series diodes, although it has the disadvantage of reducing forward current capability (see section 0). Reducing the variation in Tj is a form of balancing the leakage current dispersion.

Confirming how well diodes in a string have a balanced voltage (ie. minimal dispersion of leakage current) is not an easy measurement, as attaching any measurement probe would likely divert a substantial current, especially as leakage current is specified as 50uA at rated PIV and 100^{o}C, and Figure 8 indicates typically 5uA. Any test preferably has all diodes connected in series to retain the same Tj, and a typical multi-meter with 10MΩ input impedance will draw 50uA at 500VDC so won't be appropriate.

A more sensitive, but still common meter is a digital megohm-meter. However, even with 2000MΩ resolution and 1kVDC supply (ie. 0.5uA), diodes like 1N4004 (400V PIV) don't exceed 0.5uA, even when the diode is heated by a quartz-halogen bulb to likely above 100^{o}C. This reflects on improvements in modern manufacture, as the 1N4007 has been available since at least 1965, and the reverse current datasheet rating of 50uA hasn't changed since then. Also, manufacturers nowadays label the same diode as 1N4007, as well as 1N4004, 5 etc. With the digital megohm meter, the leakage current resolution can be simply increased by using a modern cheap 4-digit meter on mV range with 0.001mV resolution, and measuring across a 1kΩ sense resistor in series with the diode. Making measurements above 1kV would require a special jig or a specialist megohm-meter to characterise the level of dispersion in modern diodes, and some batch testing of 1N4004 has shown leakage current slowly increasing above 1kV linearly ramping up to about 0.1uA around 1.4kV and then breakdown starting between 1.5 and 1.6kV (similar observations made in [21]). A batch of UF4007 shows a markedly higher leakage current that is measurable down at a low voltage, rising consistently to nearly 0.2uA at 1kV and beyond, and avalanche onset starting at about 1.2kV.

If static balancing parts were to be added across each 1N4007 (perhaps for peace of mind), then they would need to be at least 1000VDC and 350VAC rated. For resistors, that is likely to require at least two PRO2 resistors [7] or four 0.6W metal-film resistors in series placed across each 1N4007. 1N4007 datasheets specify rated reverse leakage at 100^{o}C and 1kV as from 50uA to 500uA. With the lower level of 50uA, the suggested parallel resistance from [6] for 3 diodes blocking 70% of combined 3kV rating, is:

R = 0.3*3kV/[(3-1)*50uA] = 9MΩ (ie. 4x 2.2MΩ 0.6W resistors in series).

Dynamic balancing

Although the literature on series connected diodes will always describe methods to balance voltage across diodes to meet dynamic changes, in practise for valve amps, dynamic changes of the order needed to be significant are only likely as spike events – and those spikes are likely to be constrained by other means.

The influence of Qrr (see 3.2) and junction capacitance (see Figure 18) on PIV performance are effectively non-existent for series connected diodes in a valve amp power supply application, as dI/dt through the diode and dV/dt across the diode are negligible for the diode reverse voltage near PIV rating (ie. at the negative peak of secondary voltage waveform). Qrr is further discounted for fast-recovery diodes like the UF4007. This applies for either a capacitor-input or choke input filter power supply configuration.

Of note is that a choke input filter power supply has a near constant current level through the diode, which then steps to zero for the diode turning off, and steps from zero up to a constant level in the diode that turns on. A step change in voltage across each diode also occurs in practise due to transformer leakage inductance. However, diode commutation occurs when the transformer winding voltage is about zero, and so as long as the diode-choke junction is protected from over-voltage transients (ie. by some small level of shunt bypass capacitance, or a MOV across the choke winding) then the diodes would not normally be exposed to PIV stress conditions.

If dynamic balancing were to be added across each 1N4007 (perhaps for peace of mind), then they would need to be at least 1000VDC and 350VAC rated. The typical diode junction capacitance is shown in Figure 18 as being well below 10pF, and reducing as voltage approaches PIV. Diodes from the same batch will have a low spread in capacitance. The added capacitance has the aim of forcing a defined capacitance across each diode rather than relying on the inherent junction capacitance variation across parts. As such, a capacitance of just 100pF will swamp the diode's junction capacitance, and the capacitor would need to have a tight capacitance tolerance with respect to applied voltage

Practical use

Practical use of series connected ss diodes without added balancing parts requires diodes:

- to be from the same manufacturers batch
- to be soldered in close proximity so as to maintain a similar junction temperature
- to be adequately de-rated

Continuing the design example from the previous section, where using 2 series connected 1N4007 would be appropriate for full-wave rectifying power transformer secondary voltages in excess of 280-0-280VAC, it would be reasonable to include additional margin to determine a secondary voltage where 3 series connected 1N4007 would be appropriate. If the derating was increased to say 30% to cover imbalance, then 2x700V = 1400V PIV would cover up to 440-0-440VAC at nominal mains voltage (eg. 230VAC mains which could be expected to experience 260VAC max). And similarly, 3 series diodes would cover up to 660-0-660VAC when similarly derated.

Apart from using just 1N4007 solid-state (ss) diodes for power supply rectification, another common application in valve amplifiers is to prevent arcing in valve diodes – where ss diodes are inserted in series with the anode of each valve diode in a dual-diode rectifier. This type of application will typically need at least two 1N4007 in series, as valve diodes typically have PIV rating in excess of 1kV, and the ss diodes need to support the full PIV when a valve diode has degraded and would normally start to arc (and has been known to damage the power transformer without adding protection).

It is also worth noting that some valve amplifier manufacturers have used a series string of ss diodes to act as free-wheeling diodes that conduct to short the output transformer push-pull windings (valve anodes) to 0V ground if the anode voltage is forced below ground level. This protective measure exposes the diodes to fast transient reverse voltages arising from output transformer primary winding leakage inductance. It was not uncommon for these diode strings to fail after years of use, and although there could be a few plausible causes of failure it is likely that PIV stress was a cause for failure of those vintage parts.

2.2 UF4007 in series with 1N4007

An interesting adjunct to using the same two ss diodes in the manner previously described in section 2.1 is to use a combination of 1N4007 and UF4007.

Due to significantly different leakage current characteristics, the 1N4007 would increasingly support inverse voltage during each alternate mains half-cycle up till when its leakage current increases sufficient to cause the UF4007 to start sustaining inverse voltage. It may be that the 1N4007 inverse voltage reaches circa 1.5kV (the voltage typically reported as the onset of its avalanche region) before the UF4007 supports substantial inverse voltage. Even when the UF4007 supports a substantial inverse voltage, the 1N4007 doesn't enter its breakdown region as the series reverse current is still being limited by the UF4007.

Due to the faster turn-off characteristic of the UF4007, it will stop rectifier current through the 1N4007 before the 1N4007 itself would have turned off.

The current capability of the series pair would be at most the capability of the UF4007 (as it is lower than the 1N4007's capability), and need to be derated due to two diodes in close proximity (as discussed in 1.1). The 1N4007 would have a lower die temperature to some extent from lower IR conduction loss, although if the two diodes are soldered in close proximity, then that would tend to force a similar die temperature.

The characteristics pertaining to this combination of ss diodes is not too dissimilar to when a 1N4007 is used in series with a valve diode - as discussed in section 7.4.

3 The influence of secondary winding leakage inductance

One outstanding concern when rectifying AC power to DC power for audio amplifiers is to avoid 'noise' related to the diodes commutating (especially when they turn off).

Decades ago when only valve diodes were available, the main concern was ripple voltage and how to attenuate that with available parts, including chokes, paper-oil can style capacitors and early generation electrolytic capacitors. But the prevalent use of solid state diodes from the 1960's on, and the ease in using electrolytic capacitors, has meant that diode-related noise has become a concern. This situation is even more exacerbated for solid-state amplifiers that use lower voltage power supplies, with output current requirement that are orders of magnitude higher than for high-voltage (HV) power supplies for valve amps.

Although the diodes are often considered the culprit, and seem to receive the bulk of attention, it is the leakage inductance in the power transformer winding feeding the rectifier diodes that is the real culprit. The noise generation process has four main phases to consider:

- The filter capacitance and load, along with the transformer effective winding resistance, causes a peaky current pulse through the transformer secondary winding. The current pulses typically start and stop well within half a mains cycle.
- The rectifier diodes stop the winding current when it passes through zero, such that the leakage inductance in the winding experiences a stepped voltage across it, due to the step in dI/dt at the time when I=0.
- The voltage step generated by the leakage inductance is energy that dissipates by current flow through the winding into any available circuit loop, including looping through other power transformer windings through magnetic induction and capacitive coupling. This includes when the diode junction exhibits a capacitance (typically a non-linear value varying with junction voltage), such that current in the reverse direction can flow due to the voltage disturbance across the junction capacitance, until the energy in the winding is dissipated.
- If the diode junction exhibits a stored charge Q, current in the reverse direction can flow until the diode finally blocks current flow, and the step in dI/dt of that final blocking action can increase the observed voltage step.

3.1 Winding current waveform

The winding current equals the diode current when the diode is conducting and charging the filter capacitance. The current waveform is quite peaky, as discussed in section 1.1.

The higher the dI/dt at the time of current stopping at 0A, then the more abrupt the change from a non-zero dI/dt to when dI/dt=0. This is the basic aspect that determines the extent to which winding leakage inductance is a problem.

The conduction period reduces as the Ipk/Iav ratio increases, as indicated in Figure 13.

The dI/dt of the diode pulse when diode current reaches zero can be estimated from peak current Ipk and conduction duration Tc, where dI/dt = 2.π.Ipk/(2.Tc) , assuming conduction waveform is half of a sinusoid of frequency = 1/(2.Tc). For an onerous power supply where Ipk/Iav ~ 20, and Iav ~ 0.3A, and a short conduction time of about 28 degrees of a 180 deg 60Hz half-sine (from Figure 13) where Tc~1.3ms, then dI/dt ~ 2.π.6A/2.6ms ~ 0.015 A/us when current reaches zero. This dI/dt level is proportional to the Ipk magnitude, and inversely proportional to the conduction duration of the current pulse.

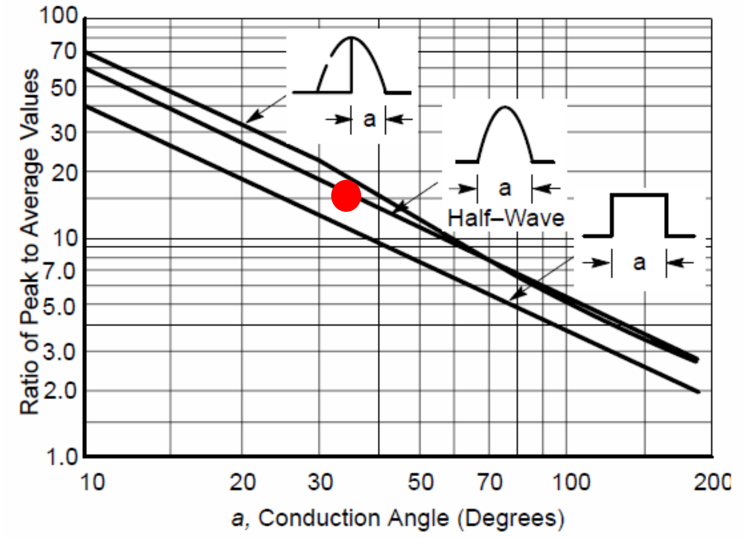


Figure 13. Graphical representation from [1] of diode conduction angle for different peak to average current ratios. For single phase rectifiers the maximum conduction angle is 180 degrees.

Red dot relates to example in Figure 3 and Figure 4.

The abruptness of the stop in current when it approaches zero is a characteristic of the diode. A valve vacuum diode has an incremental resistance (r_p) that increases smoothly and significantly as the plate current level approaches zero, such as shown in the plot for a 6H6 diode [8]. The incremental resistance is shown to increase from 278Ω at 28V, to 527Ω at 8V, and ~ 2kΩ at 1V. Common valve rectifier diodes used in power amplifiers have higher current capability, and lower incremental resistance (eg. 10%), although the characteristic is similar.

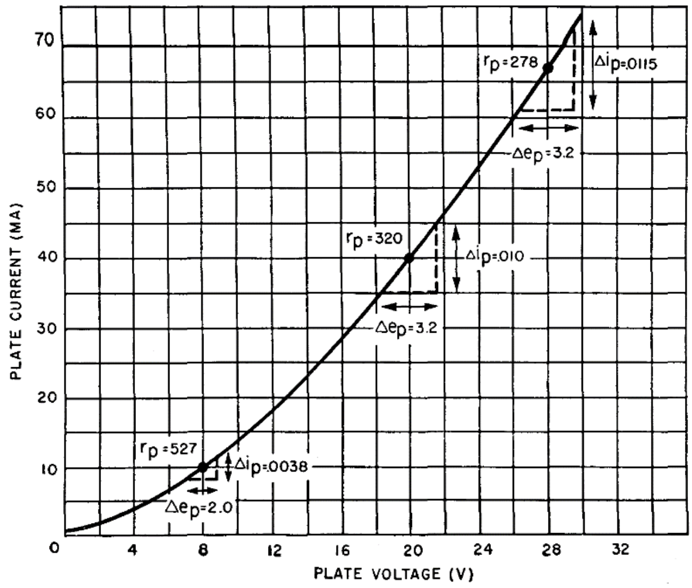


Figure 14. 6H6 valve diode V-I curve from [8].

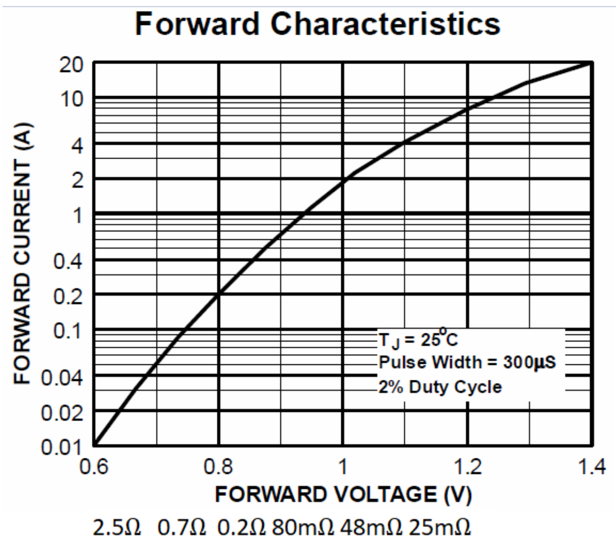


Figure 15. 1N4007 diode forward V-I curve, with incremental resistance levels added at bottom.

For the 1N4007, the incremental resistance also increases as the forward current level approaches zero, but resistance values are much lower, as indicated by the values added to the bottom of Figure 15. In addition, the diode voltage only reduces by about 1V from when it is conducting peak current to when it is effectively not conducting current.

It is illustrative to calculate the likely time it takes for the voltage across the diode to change by a level that commutates the diode.

The induced secondary winding voltage Vsec, due to the primary winding voltage, is a sinusoid with a dV/dt that varies throughout the mains frequency cycle. At the peak of the secondary voltage sinewave, when the diode is conducting its peak level Ipk, the dV/dt is zero. As the Vsec waveform then falls, the dV/dt increases in magnitude. For small conduction angles (Figure 13), diode turn-off typically occurs close to the peak of Vsec and well before Vsec passes through zero crossing. For a 60Hz mains frequency, and Vsec = 240Vrms, the maximum dV/dt at zero crossing is 2.π.f.Vsec.√2 = 0.13V/us. Given that dV/dt of Vsec at diode turn-off would be significantly lower, perhaps ~0.03V/us, then Vsec would take about 10us to reduce 0.3V, which is equivalent to 1N4007 diode current falling from 600mA to 10mA. During that time, diode cathode voltage will also be falling – dependant on the ripple voltage characteristic. This indicates the diode is 'forced' off over a duration of many microseconds, but so far excludes any effect that winding leakage inductance may have at that time.

3.2 Diode Turn-off Transient

A turn-off transient occurs when winding (diode) current reaches zero due to a mix of issues.

The pre-existing dI/dt magnitude, just prior to diode turn-off, is generating a voltage across the secondary winding's leakage inductance, as illustrated by the equivalent circuit in Figure 16.

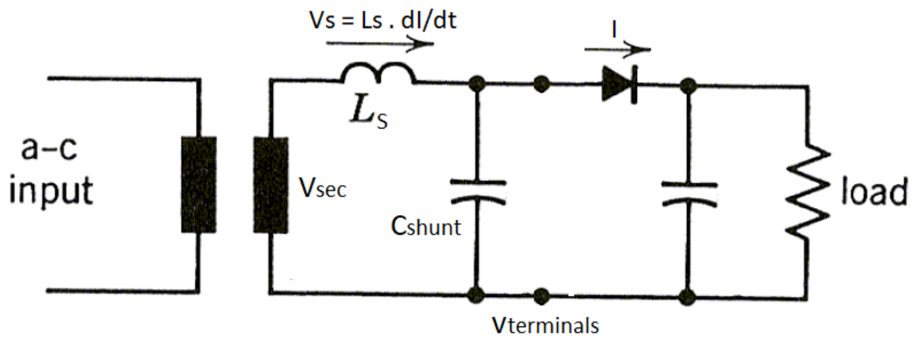


Figure 16. Equivalent circuit showing secondary winding with transformer voltage Vsec, leakage inductance (Ls) and shunt capacitance (Cshunt).

One key issue is how abruptly the diode stops conducting (ie. its dynamic change in resistance over time). A solid-state diode like a 1N400x presents a low dynamic resistance (<<1 ohm) when conducting significant current (due to the low on-voltage), and dynamic resistance only increases above 1Ω as on-voltage falls through the nominal threshold voltage (ie. 0.5-0.7V).

At the time of diode turn-off, the dV/dt of the secondary winding voltage Vsec is pretty slow, as described in the previous section, and is likely to take some microseconds for Vsec to fall even 0.1V. On the cathode side of the diode, the filter capacitor voltage is also falling, as the capacitor current reverses from charging (I > load), to discharging (I < load), which would tend to slow the dV/dt across the diode.

Figure 17 Illustrates a rectified current pulse (I), and the related voltage Vs developed across leakage inductance Ls within the secondary winding (due to Vs = Ls . dI/dt).

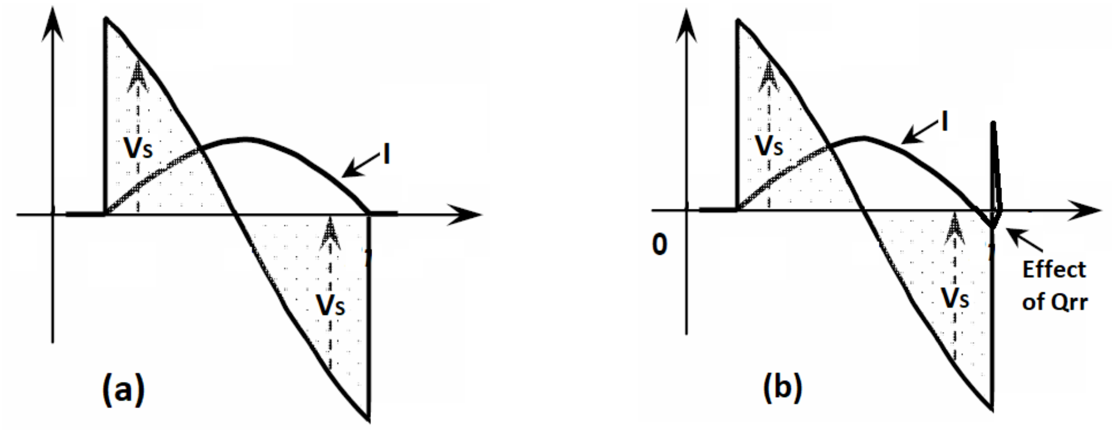


Figure 17. Voltage across secondary winding leakage inductance Ls due to changing winding current I with ideal diode (a), and when diode Qrr is significant (b).

When the diode stops conducting, and diode voltage falls through its threshold level (~0.5-0.6V) causing the diode dynamic resistance to rise quickly, the dI/dt steps from a negative level to zero as shown in Figure 17(a), and the step in voltage across the leakage inductance abruptly causes the winding's terminal voltage to step down.

As the secondary winding has internal stray shunt capacitance (Cshunt), it is the path of least resistance for current flow due to the step in Vs. Shunt capacitance (Cshunt) shown in Figure 16 is located at the winding terminals, but will be distributed throughout the secondary winding. Similarly, leakage inductance (Ls) is distributed throughout the secondary winding.

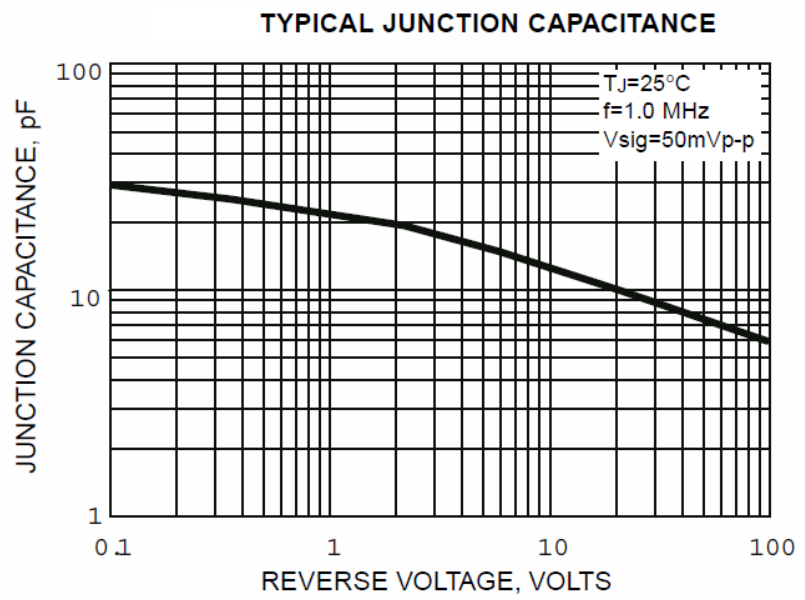


Figure 18. UF4007 diode junction capacitance for reverse voltage.

The diode has a junction capacitance, as shown in Figure 18 for the UF4007. Some 1N4007 datasheets provide that curve, and show it is effectively the same as the UF4007. However, when the junction is at zero or slightly forward biased, any amount of diode stored charge (Qrr in the junction) allows the diode current to continue to reduce through zero and go negative until Qrr is reduced to zero and the diode blocks any further junction current. Qrr is a representation of the diffusion capacitance presented by the junction – that capacitance has been measured by a few people and shown to significantly increase when the junction enters in to forward bias.

The relatively large effective capacitance of the diode junction when junction voltage is reducing through the on-voltage level, but still forward biased, allows a resonant current path with the winding leakage inductance. Resonant conditions are lost above a frequency where the winding shunt capacitance overcomes the leakage inductance. Morgan Jones investigated this topic [12].

From the previous section 3.1, the worst-case level of dI/dt ~ 15 mA/us is at least 100x slower than normal characterisation dI/dt levels of diode Qrr, as indicated by Figure 19 where dI/dt plot goes only down to 1A/us, and Qrr reduces significantly as dI/dt reduces. Measurement of dI/dt in example valve amplifiers indicates dI/dt is typically under 1mA/us.

The Qrr of the 1N4007 is much more than the UF4007 (which is a fast recovery diode). Test results indicate that Qrr is at least an order of magnitude less for the UF4007 [10].

The 1N4007 anode voltage can step down further than the expected Vs, due to an additional step in Vs as the diode finally blocks a negative I value and dI/dt again steps in level as illustrated in Figure 17(b). Example measurements of the voltage disturbance at diode turn-off have been presented by Merlin [11].

Any step change in winding terminal voltage is going to cause a transient current through the diode junction (diffusion) capacitance, which will loop around through the filter capacitor and load circuit paths. Mark Johnson [22] made comparison measurements of 48 different diode types in a special test rig that appears to force a high peak current level due to a n.ω.C.R_L value of ~ 916 (for n=1, ω=377, C=16,200uF, R_L=150Ω), although the transformer effective resistance was not identified. The special test conditions generate a substantial transient voltage on the sole secondary winding, which being only half-wave rectified, forces a measurable ringing current through the diode junction, to an extent influenced mainly by the diode's junction capacitance.

The simplified rectifier circuit in Figure 16 does not include the additional series diode used in a full-bridge, or the two reverse biased diodes in a full-bridge. At the diode bridge turn-off time the reverse biased diodes have about their peak reverse voltage across each diode, and so their junction capacitance is insignificant.

Any other winding on the power transformer will also experience a transient voltage step, due to transformer voltage coupling, and experience a transient current step due to parasitic winding capacitance coupling.

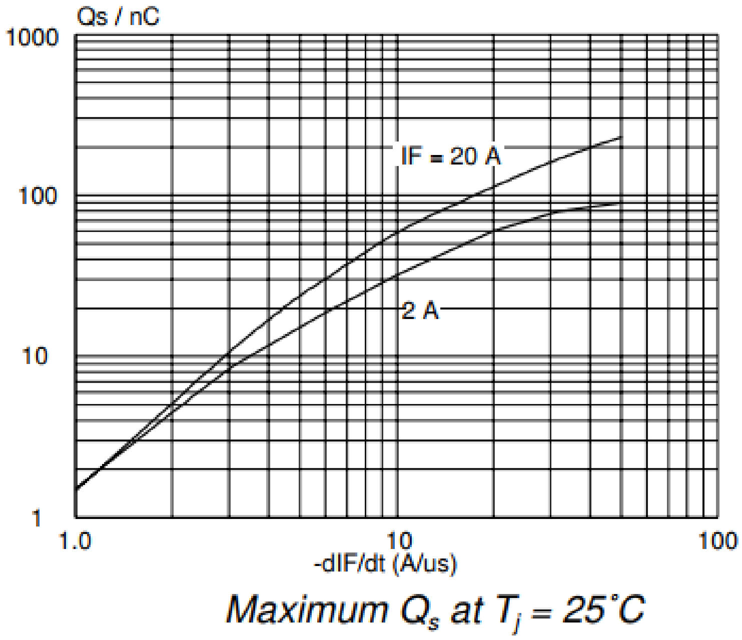


Figure 19. Typical reduction of Qrr with dI/dt from BYT79 14A 500V fast-recovery diode.

3.3 Alleviating the influence

Three means are practically available to alleviate the influence of winding leakage inductance:

- reduce dI/dt at on-set of diode turn-off arising from the charging pulse.
- add a snubber with low stray inductance as close to the winding terminals as possible.
- use a fast recovery diode with as small a junction capacitance, and as low a residual Qrr as possible.

It is uncommon to have a choice between commercial power transformers based on secondary winding leakage inductance, especially as many DIY projects are based on having a power transformer to hand. Leakage inductance is typically only of concern for output transformers, where much effort is expended to minimise leakage inductance, but balanced with also minimising shunt capacitance – whereas power transformers typically have simpler construction, dictated by safety isolation between primary and secondary windings. In general, a toroidal transformer will likely present a lower leakage inductance, and also a lower effective winding resistance.

Diode current waveform

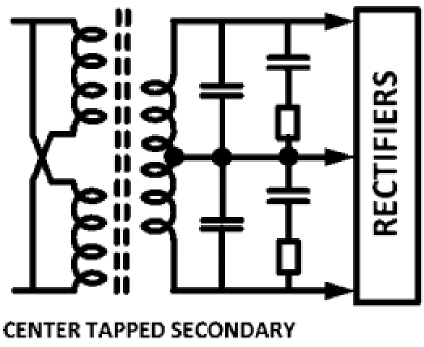
The dI/dt at the time when diode current returns to zero depends on the peak current Ipk of the waveform, and the conduction duration of the waveform – as discussed in section 2.1, and relating to capacitor input filter design as discussed in section 1.1.

Unfortunately, many power supplies are designed for minimum output DC voltage ripple, and better transformer regulation. The resulting choice of higher filter capacitance C, and lower effective transformer winding resistance, pushes the diode current waveform to a higher Ipk, and a smaller conduction duration, and so causes higher dI/dt as diode current falls to zero.

Snubber discussion

It is always better to mitigate a noise source at its origin, and that is the preference here too. As the noise source is the leakage inductance, then the closest available access nodes are the transformer winding terminals. Whether or not any such noise is significant or noticeable can depend on many factors, and is not easy to confirm even with sensitive measurement equipment, let alone human ears, so be aware that there may be no necessity for adding additional snubber circuitry and no change in performance may be detectable.

A C-RC snubber across the winding terminals allows transient current to be constrained to a local circuit loop. The simplest method to determine a suitable snubber is the ‘bell-ringer’ technique [9], which aims to load the windings as they would be loaded in normal operation, apply a step voltage to the relevant winding, and then observe the resulting transient waveform whilst adjusting snubber network values to obtain a suitably damped response.



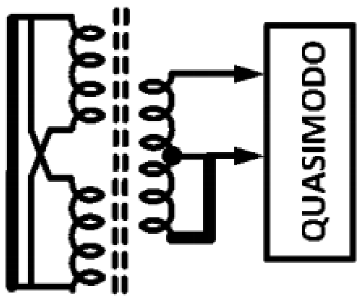


Figure 20. Snubber arrangement for full-wave rectified power supply as commonly found in tube amps, along with Mark Johnson’s recommended ‘Quasimodo’ test-jig setup [9].

The power transformer primary windings are shorted for the test, as the AC mains presents a low impedance (eg. <100Ω) supply. Any valve heater windings on the power transformer would also be shorted, as the heaters effectively short those windings in operation.

The non-conducting secondary winding is also shown as being shorted for the test (eg. for a full-wave rectifier configuration with a CT), although in operation that winding is effectively open-circuited (the non-conducting diode has close to max PIV across it). However, the aim is to then load each secondary half-winding with a snubber, which can then be approximated by a short circuit, or preferably a snubber is installed for the test.

In addition, the windings not being tested and the transformer core should be shorted to the secondary winding CT, and any bell-ends fitted, so that standard stray capacitances are accounted for in the test, and the mains AC distribution may by design short the neutral to the protective earth (which likely connects to the amp chassis and to the transformer core for common E-I cored transformers).

The snubber parts are best located at the power transformer, and connected in a manner to minimise loop area. The capacitors should be X rated, especially for valve B+ supplies, so that they are safely rated for continuous exposure to high AC voltage. Similarly the snubber resistor is preferably rated for the peak AC voltage across the winding (eg. 350V rated for a 250Vac winding).

Tests on one full-wave 400-0-400V power supply for a valve amplifier indicated that:

- a very low-level hump in the frequency spectrum of the winding voltage could be attributed to the resonance of winding leakage inductance and winding shunt capacitance when using 1N4007 diodes.
- adding the RC part only of a C-RC snubber noticeably suppressed the hump and noise in that region. There was no benefit in adding the first C part of a C-RC snubber.
- changing diodes to UF4007 noticeably reduced the noise level in the region of the hump, and in the hump level.
- no related noise was seen on the frequency spectrum of the rectified and filtered voltage (after the diodes on the first filter cap).
- the observed noise hump was around 60kHz, and worse case (1N4007) was at a level -90db below the fundamental frequency voltage level of the winding, and only measurable using a soundcard with a low noise floor.
- the resonance characteristic of the secondary winding was identified by measuring the impedance of each 400-0-400V half winding using a soundcard and REW software technique [24].

Diode selection

The third means to alleviate the impact of a given winding leakage inductance, is to use a fast recovery diode such as a UF4007. Compared to a standard silicon diode such as a 1N4007, the UF4007 has shown in tests to have negligible Qrr.

The preference is also to choose a diode with the lowest current rating, as that is likely to also provide a lower Qrr level and lower reverse bias capacitance, both of which minimise excess diode junction reverse current flow. For example, the UF5408 is a 3A 1kV fast recovery diode often used when the UF4007 is considered marginal for current rating. Unfortunately, the UF4007 and UF5408 datasheets don’t provide Qrr data, and only indicate that junction capacitance is about 3x larger for the UF5408.

There may also be benefit in using multiple series ss diodes to effectively slow down the abruptness of the turn-off instant. In this situation, as the turn-off voltage is increased then the time taken to transition from on to off is also increased.

Note that valve diodes do not exhibit any reverse current from a minority carrier Qrr characteristic, and along with their much slower change from on to off, they show negligible turn-off transient disturbance.

4 Using PSUD2 and assessing valve diode operating limits

4.1 PSUD2 simulation

The PSU Designer computer application (PSUD2 is the latest version [2]) provides an easy to use tool to design, assess and compare power supply related issues by simulating their performance.

Valve diodes are not as forgiving as solid-state diodes, with many failure modes, and a far greater cost to purchase and implement. PSUD2 can help determine diode related issues like:

- if a particular diode is suitable for an application and how different diodes compare.
- what transformer and filter values can be used with a particular valve diode, including if extra resistance in series with a power transformer winding is needed, and what maximum filter capacitance can be used.

PSUD2 also helps to determine many other related issues like:

- filter capacitor ripple current level (to help choose a capacitor part).
- filter capacitor maximum voltage requirement (under all load and power-on conditions).
- minimum loading for choke input filters.
- the influence and selection of an NTC part added to primary or secondary side of transformer.
- fuse rating for secondary winding protection (see [13]), and fault current levels from faulty parts or short circuits.
- DC and ripple voltage levels for RC and LC ladder power distribution schemes.
- filter resonances from step load changes (when inductors are used).

PSUD2 models a power circuit schematic and provides a simple process to insert or change sections of the schematic, and then modify specific part values and attributes by double clicking a part. The schematic can then be simulated, and a results table provides key operating data for the parts in the schematic, and a graph can plot the waveform of voltage or current through selected parts.

The simulated operation of any part requires caution as no part is ideal – see discussion in 1.2 above.

PSUD2 circuit schematic setup

Confirm that PSUD2 options are appropriately set, in particular the mains frequency, and that soft-start is de-selected (only select this if you have a specific need).

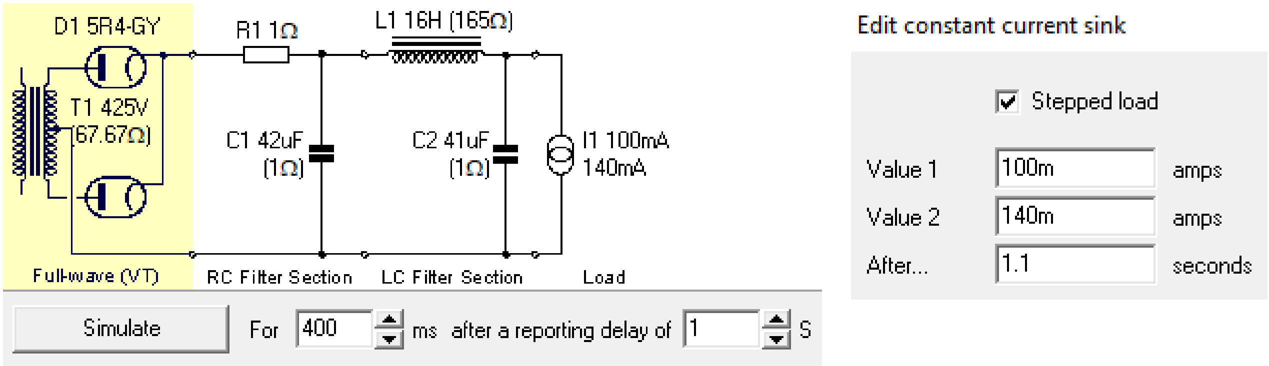
It is often convenient to use a simple single load resistance to represent an amplifier, even though the amplifier may have many circuit stages and a complex power distribution arrangement, especially for diode performance assessment. Typically, the output stage draws much more current than all the other stages combined, as all other stages may only represent an additional 5-10% of load current. Sometimes the output stage screen supply and preamp stage supply are taken from a different winding or a winding tap, which cannot be modelled in the same PSUD2 circuit, although that typically has no consequence to PSUD2 assessing the main B+ supply performance.

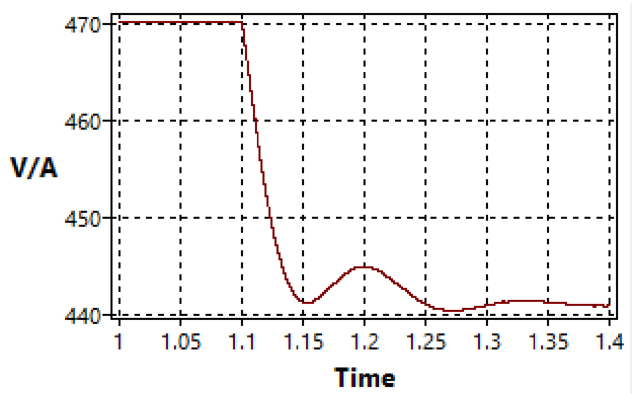
PSUD2 can include a variety of sequential filter stages, but is limited in only being able to resistively load the supply at the end of the filter stages. Current taps can be inserted between filter stages, and to replace a load resistance, however caution is required as a current tap may not adequately represent dynamic loading conditions, and so may not be appropriate to use for simulating transient turn-on events. On the other hand, current taps allow a step change in loading to be applied, and that allows dynamic load changes to be simulated to check for ringing type responses (for example see [15]).

For the transformer parameters I prefer to use the unloaded winding voltage and the effective series resistance configuration (see 4.4 below), noting that transformer ratings and catalog data often simply states the loaded secondary winding voltages (eg. 6.3V or 380-0-380V) which can be 6 to 9% below an unloaded voltage.

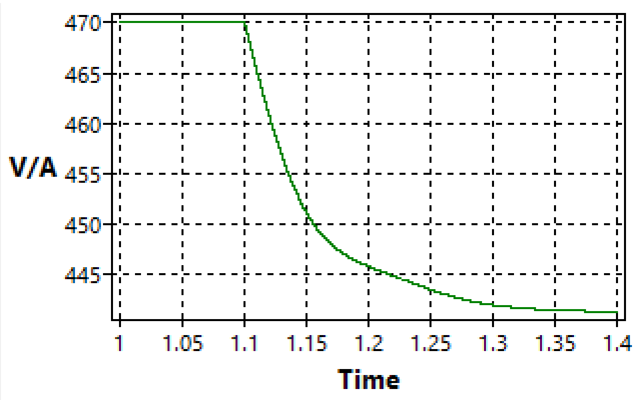
Filter resonance assessment

A C-L-C power supply filter could exhibit a damped oscillatory response to a step load change, depending on part values and inductor ESR, which PSUD2 can illustrate by configuring a load tap to instantaneously change current level during steady-state operating conditions. For example, use a step current change at 1.1s after a 1s delay (assuming the load has achieved steady-state conditions by 1s), and then view the plot waveform for say 400ms and look at voltage waveform across each C and the current through L, as per the following example. For a power amp where the first filter C handles rectifier ripple, and the second filter C handles output stage and amp load, then if the plots show a damped oscillation it is often better to increase (eg. double) the second C capacitance until no oscillation is visible in the plot. The frequency of oscillation is estimated by 1/T, where T is the time between oscillation peaks (or dips). In the following example, a damped oscillation of about T~0.12s (8Hz) occurs, which is then avoided by doubling C2 to 82uF.





C2=41uF



C2=82uF

In general, it is not valid to assess oscillatory response at turn-on, as that is a one-off event with abnormal initial conditions (eg. part voltages and currents are zero) and the part models likely do no represent actual circuit operation.

4.2 *Power turn-on and steady state conditions*

Two time-related regions of operation are well worth simulating and assessing when using valve diodes in a power supply: the initial power-on region when the mains AC is applied by a switch; and the steady-state operating conditions. Valve diodes and valve amplifying stages do not instantly operate from a cold state, and so an awareness is needed as to how that complicates the assessment of a valve amplifier. Many valve amplifiers have semiconductor diode power supplies, with the diode operating instantly, and although this situation is simpler to appreciate, it still requires care to appreciate how to simulate and assess.

For a semiconductor diode rectifier, current passes through the rectifier and into the filter and load immediately that AC mains power is turned on, which results in a dynamic change in voltage and current levels in the power supply. Three operating conditions are worthwhile appreciating, as they can be simulated by PSUD2:

a) Power turn-on with cold valves.
   - Semiconductor diodes start conducting immediately, and well before amplifier valves, and so the power supply and DC distribution parts charge up, but with no valve loading.

b) Power turn-on with hot valves.
   - This relates to turning an amplifier off, then on with only a short pause.
   - Semiconductor diodes start conducting immediately, and the amplifier valves start conducting as the power supply and DC distribution voltages rise.

c) Steady state operation.
   - This can relate to an idle condition (ie. no signal, or class A operation), or to a substantial signal (ie. where loading is much heavier and power supply voltage may sag).

For a valve diode rectifier when AC mains power is turned on, there is a delay while cold diodes heat up, and then current slowly increases. Three operating conditions are worthwhile appreciating, as they can be simulated by PSUD2:

d) Power turn-on with cold valves.
   - There is an initial delay as no valves conduct immediately.
   - Valve diodes typically start conducting before amplifier valves, even for indirectly heated diode valves, and so the power supply and DC distribution parts charge up, but with no amplifier stage valve loading.

e) Power turn-on with hot valves.
   - This relates to turning an amplifier off, then on with only a short pause.
   - Valve diodes start conducting immediately – this is the same as (b) above.

f) Steady state operation.
   - This is the same as (c) above.

4.3 Valve diode design parameters

A valve diode for power supply rectifier applications is typically selected by its:

- heater voltage and current (such as 5V 2A, 5V 3A, 6.3V 1A, etc).
- valve base style (eg. octal 8-pin, noval 9-pin).
- diode configuration (eg. single or dual diode with common cathode).
- direct or indirectly heated cathode (this impacts the diode turn-on time).

From a design perspective, the key diode operating limits are:

- peak inverse plate voltage, **PIV**.
- steady state peak plate current (per plate), $\mathbf{I_{ap}}$.
- transient peak plate current (per plate), for a maximum duration (typically 200ms), $\mathbf{I_{apt}}$.

These key diode limits are typically presented in a datasheet as 'design centre maximum ratings', and indicate maximum operating levels for which acceptable diode service life should still be obtainable. The rating system changed from 1957 to 'design maximum' levels, although they are pretty similar.

There are many vintage reference books and magazine articles describing power supply and rectifier diode operation - all are worthwhile reading and typically accessible from the internet. Some books focus specifically on power supplies for valve amps, or have major sections on that topic, and so are even more worthwhile reading to provide a close link between your amp of interest and its power supply.

Many valve diode datasheets try to assist designers by presenting the key operating limits along with some example transformer and filter values that operate a diode within its limits. But many amps don't use datasheet circuit values, and few people have the ability to make measurements to confirm operation is within the key limits, as that requires specialist equipment and time and an amp that is basically built.

Not all diode datasheets present all the key operating limits, so it can be constructive to look at some common valve diode datasheets. The two valve diodes GZ34 and GZ37 are compared to work out how they differ, as the GZ37 datasheet has quite limited information and those two devices are often considered equivalent. The GZ34 has very similar ratings to the 5AR4, and some datasheets combine them. The GZ32 and 5R4GY also have limited or no peak current information, and a variety of plate voltage capabilities, and are also discussed.

4.4 GZ34, GZ37, GZ32, 5R4GY datasheet interpretations

GZ34

The earliest GZ34 datasheet I have is Philips dated Feb 1958 (which contains some 1954 data). The datasheet provides operating limit values for PIV = 1500V and I_{ap} = 750mA, but no I_{apt} rating, along with graphs for a range of transformer secondary voltages, based on 60uF filter capacitor and a given transformer effective resistance for each voltage level, and the expected load output voltage for a target load current.

PSUD2 can be used to assess the diode's operation in the typical rectifier circuit used by the datasheet – a full-wave rectifier with capacitor input filter and a resistive load, as shown in Figure 21. PSUD2 simulates operation over a preset time duration (eg. 100ms) after a preset reporting delay (eg. 2 secs) to show a waveform plot of a particular result parameter (eg. diode peak current I(D1)). The transformer has the 450V and 150Ω values shown in datasheet graph 7R05950, and the results table shows output voltage V(R1) and output current I(R1) values the same as in the datasheet.

PSUD2 can be used to show that the datasheet page 7R05950 (dated 1958) presents curves where I_{ap} =0.75A for a 200mA load, and I_{ap} is higher for a load higher than 200mA. PSUD2 shows that datasheet page 7R04099 (dated 1954) indicates that I_{ap} =0.75A is not associated with any specific parameter like output current. It could be that the 1958 datasheet update attempts to normalise the required Rt level for a rated I_{ap}.

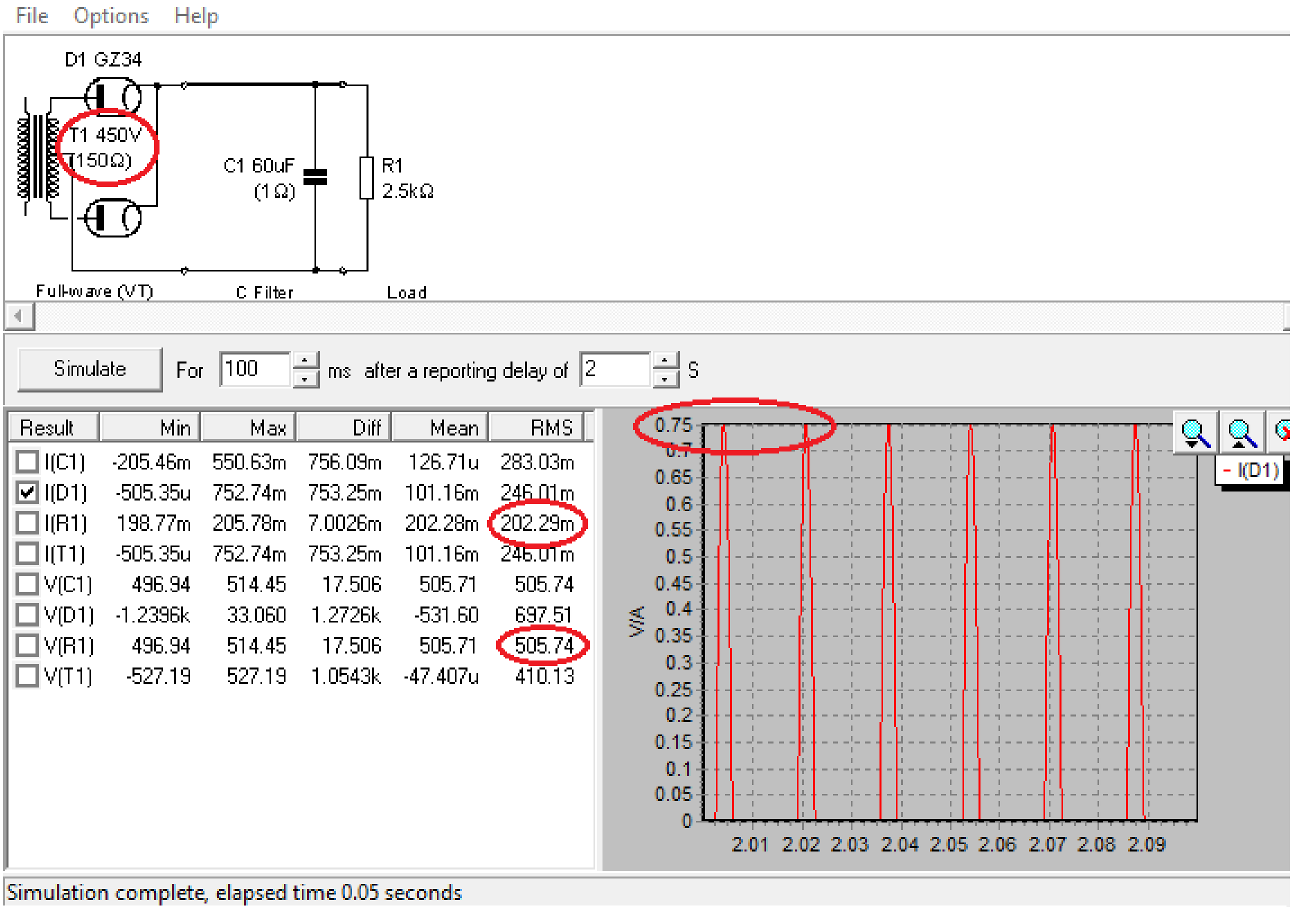


Figure 21. PSUD2 simulation of GZ34 datasheet limit for 450V secondary with 200mA load.

Power transformer winding resistance is inserted in the PSUD2 transformer model. Double mouse clicking on the transformer symbol brings up the 'Edit transformer properties' box, as shown in Figure 22. For a full-wave vacuum tube source, the RMS V value is the off load (or no-load) winding voltage from the CT to an anode terminal, and the 'Source res' value is the effective resistance (Rt) in series with that winding voltage. Clicking on the '…' Ohms button brings up the 'Source Impedance Calculator' box, where winding resistance measurements for both the primary and secondary windings can be input to calculate the effective resistance value Rt for any transformer. PSUD2 can be set up to simulate any point on the GZ34 datasheet curves by setting the appropriate transformer Vtr and Rt values, and varying R1 to show the target output current in the rms column for I(R1).

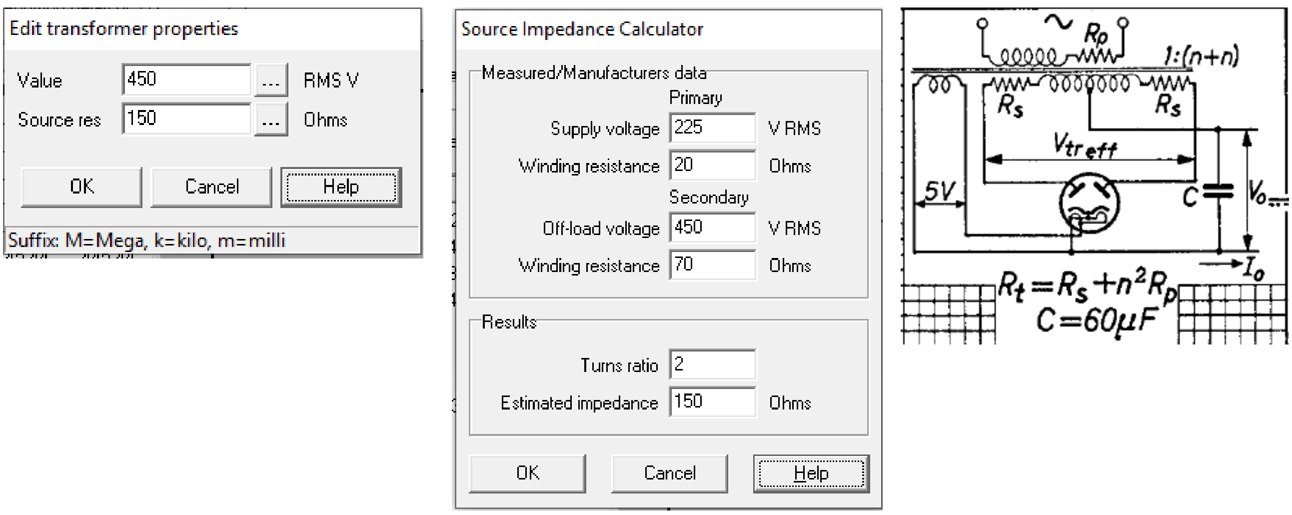


Figure 22. PSUD2 transformer parameter input screens.

PSUD2 uses a GZ34 model with DRes = 0, Vlaw = 1.5, Vfac = 0.00396, Vpiv = 1500, Ipks = 99, Ipkr = 0.75. The simulated diode drop is 30V at 650mA, which aligns with the datasheet curve.

Although the GZ34 datasheet does not present a transient peak plate current I_{apt} rating, PSUD2 can estimate that value for the presented operating curves, as shown in Figure 23 when the reporting delay is reduced to 0 secs. The datasheet page 7R05950 (dated 1958) curves with 200mA output current all indicate I_{apt} = 3.1A at 60Hz, or 3.0A at 50Hz.

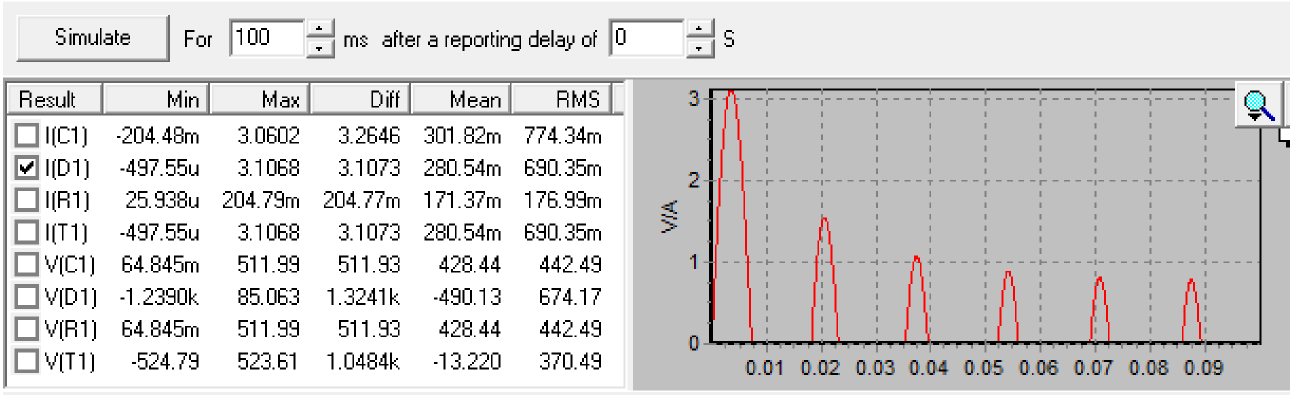


Figure 23. PSUD2 results showing hot turn-on I_{apt} level of 3.1A for GZ34.

GZ37

The Mullard GZ37 datasheet (dated Oct 1958) provides operating limit values for PIV = 1600V and I_{ap} = 750mA, but no I_{apt} rating and no graphs, and just presents a table with transformer secondary voltages based on 4uF filter capacitor and a given transformer effective resistance of 75 ohm, and the expected load output voltage for the max rated 250mA load current.

A Chelmer datasheet indicates the anode on-voltage is significantly more than the GZ34 (30V at 130mA), and PSUD2 uses a GZ37 model with DRes = 0, Vlaw = 1.35, Vfac = 0.001409, Vpiv = 1600, Ipks = 99, Ipkr = 0.75. PSUD2 shows I_{ap} =0.71A and 485V output for 250mA load with a transformer secondary of 500V, which aligns with the datasheet. PSUD2 shows that I_{ap} reduces for lower transformer secondary voltages, and that I_{apt} = 0.93A for 60Hz and 500V secondary is the highest level.

Morgan Jones 3rd Ed book (p.294) makes a note that as the continuous peak of 0.75A is the same for GZ34 and GZ37, then the 60uF max cap capability for the GZ37 can be assumed to be the same for a 300V output (although I presume that was intended to be 300V secondary voltage). Using PSUD2 with 300V secondary, Rt = 75 ohm, 250mA output DC, 60Hz, the GZ34 provides Vo=330Vdc, I_{ap} = 0.91A and I_{apt} =3.2A, and the GZ37 provides Vo=277Vdc, I_{ap} = 0.74A and I_{apt}=2.0A. For that 300V secondary scenario, the GZ37 on-resistance causes a significantly lower output voltage, as well as less stressful peak diode current levels.

Although the maximum Vac limits are close (they can be considered the same if series 1N4007's were inserted in series with each valve anode), the on-resistance of the GZ37 is certainly higher and the transient peak plate current rating is uncertain and likely to be lower than for the GZ34.

GEC introduced the U54 in 1955, with similar datasheet specs, but notably I_{ap} = 1500mA.

GZ32

GZ32 datasheets from 1949 do not specify either an I_{ap} or I_{apt} rating. The plate voltage drop characteristic is very close to that of 5V4.

PSUD2 can estimate I_{ap} and I_{apt} for the presented 16uF, 32uF and 60uF capacitor input filter operating conditions (with 50Ω, 100Ω and 150Ω effective impedance limits respectively) for the 300V, 350V and 500Vac plate winding conditions (with 300mA, 250mA and 125mA DC output limit conditions respectively). Table 2 shows that I_{ap} or I_{apt} capability can vary significantly, so caution is advised when interpolating for conditions other than given in the datasheet.

	Vtr (Vac) / Idc (mAdc)	I_{ap} (A)	I_{apt} (A)	PSUD2 load (Ω)
C=16uF Rt=50Ω	300 / 300	0.88	1.35	935
	350 / 250	0.82	1.56	1440
	500 / 125	0.58	2.2	4870
C=32uF Rt=100Ω	300 / 300	0.81	1.55	880
	350 / 250	0.77	1.82	1370
	500 / 125	0.54	2.6	4750
C=60uF Rt=150Ω	300 / 300	0.78	1.55	800
	350 / 250	0.71	1.8	1285
	500 / 125	0.50	2.6	4640

Table 2 GZ32 PSUD2 results

5R4GY

Tungsol datasheets from 1945 and 1957, and RCA 1948, indicate I_{ap} = 650mA, but provide no I_{apt} rating for the 5R4GY. Although the heater is 5V 2A, the plate voltage drop characteristic is very close to that of the 5U4, 5X4G and 5Z3 with 3A heaters and I_{ap} = 675mA.

The datasheet indicates up to 700V with same effective impedance and DC loading is allowed, but states that the heater should be on for about 10 seconds before applying anode voltage. Without any pre-heat, PSUD2 can estimate I_{apt} for the presented 4uF capacitor input filter operating condition of 550V plate winding with 125Ω effective impedance and 250mA DC output, and indicates I_{ap} = 700mA, and I_{apt} = 0.85A. With pre-heat and 700Vrms per plate, I_{ap} rises to 800mA, and I_{apt} to 1.05A.

When the effective impedance is increased from 125Ω to 575Ω, the datasheet allows plate voltage to increase from 500V to 900V whilst still allowing 250mA DC output. PSUD2 indicates that first filter capacitance could similarly be raised to 50uF and still maintain I_{apt} at 1.0A, with I_{ap} down at 560mA.

As such, care is needed when using the 5R4GY due to its apparent low I_{apt} capability.

4.5 Hot turn-on conditions

The hot turn-on current rating I_{apt} of many valve diodes is given as a peak level in the first 200ms of operation.

In the context of mains waveform cycles, 200ms is a long time, and no common rectifier and capacitor input filter scenario typically imposes an initial peak current for more than 1 to 2 cycles of mains (ie. 20-40 ms), so there seems to be quite some margin for valve diode cathodes to cope with peak surge current levels for a capacitor input filter.

Aspects of a hot turn-on event include:

- B+ supply input filter capacitor has discharged.
    - 100uF 500V cap discharges in 0.5 sec with 100mA loading (eg. from output stage).
- Cathode temperature of diode is still high

    - Heater warm up time is typically at least 10 secs, however heater temperature fall profile is expected to be slower due to the temperature and thermal mass of the plate structure surrounding the cathode/heater.
    - The resistance of all valve heaters are still high, so reapplying mains AC again has more capability to cause a high peak diode plate current as the transformer has less loading on it from heaters.

In addition, as output stage valve cathode temperature(s) would similarly be still high, they can immediately conduct if cathode bias or fixed bias voltage is zero or a low level, or if output stage grid coupling caps are discharged.

- both output valves in a PP stage could conduct, and could conduct much more than idle current level, so output stage loading could be heavier than worst-case loading in normal operation.
- In-rush protection devices like NTC thermistors or surgistor relays may not have had time to cool or reset, and so peak diode current may not be adequately protected by those devices.
    - NTC devices typically have a thermal cooling time-constant in excess of 30 seconds.

Every amplifier circuit will be different and is worth assessing for whether a substantial hot turn-on event could arise. Of note is that hot turn-on conditions are not 'typical', and so very few such events may occur over a long period.

Hot turn-on assessment may need to use a different load resistance (than for idle) if the load is considered to have high surge current characteristic, or where the amplifier immediately operates with a high level signal (ie. loading is heavier than just under idle conditions).

Output stage coupling cap time constant

An example hi-fi amp with a driver stage supply feed resistor of 100k, a 470nF coupling cap, and a 500k grid leak to the output stage valve grid, has a coupling capacitor charge time constant of 300ms. A long CR time constant may hold an output stage valve's Vg1 close to 0V for a noticeable duration, and hence allow output stage current to transiently surge above idle current level. This may be alleviated if the driver stage supply voltage takes more time to reach operating level than B+ to the output stage due to power supply filtering or regulation.

Output stage transformer inductive time constant

Output stage current typically passes through an output transformer primary winding, whose inductance could range from under 10H for a single-ended stage, to above 100H for a hi-fi OPT. The effective series resistance to that inductance is around 50-200Ω from the transformer winding, in series with valve conduction which could fall to around 100Ω for a large valve at Vg1=0V condition once current has built up. The L/R time constant of that current path could increase into the range of 0.01 to 0.5 sec. A short L/R level may not significantly delay any rise of current through the output transformer, whereas a higher L/R may noticeably delay a rise in current (and hence protect the amp)

4.6 Practical considerations

Adding extra series resistance

Often the required value of Rt exceeds the secondary winding resistance in the chosen power transformer, and so an additional series resistor has to be added between the winding and valve anode to bring up the total resistance to at least the required value. PSUD2 can calculate the expected rms current in that added resistor (it is the diode rms current), and so the nominal dissipated power in the resistor can be calculated by $P = I^2R$. It is recommended to use a resistor wattage rating at least 2x the nominal dissipated power, and preferably 3x.

Secondary winding resistance is likely to be a bit different when measuring each winding arm, so preferably use the lower value as the default Rt contribution from the transformer.

Parameter spread

Parameter values used in simulation will have some spread, such as from measurement error or using a bogey value. For example, measured winding resistance at ambient temperature could be 30-40% below what exists at internal winding operating temperature. Stated capacitor part value often has a wide tolerance on actual value. A valve diode's on-resistance may change significantly with age, as could the capability to operate at the bogey peak current level.

Extra protection

Rectifier diodes age and go gassy with time, as they operate at anode temperatures similar to output stage valves, in part due to the high heater power level. Apart from relying on a mains side fuse for protection it is recommended that secondary side fuse protection be included, as well as adding solid-state diodes in series with each diode anode as discussed in 2.1, and [13], [14].

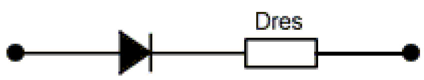
Damper Diodes

Damper diodes are used in TV line deflection circuits, and usually carry a note that they are not recommended for power rectification service. The datasheet doesn't include an I_{apt} rating, and the I_{ap} rating must be taken with caution as it is not designed for the same conduction duty cycle. That said, many have deployed damper diodes for rectification.

4.7 PSUD2 diode modelling.

The PSU Designer computer application (PSUD2 is the latest version [2]) provides a text file library of diode models named 'Rectifiers.txt' that is typically located in C:\Program Files (x86)\PSU Designer II on a WIN10 operating system. That txt file can be updated with new diode models by keeping to the same model syntax. The Help index shows a rectifier section about 'adding your own rectifiers', with the following description about the diode model.

The model



The models is based on an almost perfect rectifier in series with a resistance. The model for the rectifier for situations where Va > Vk is:

$$\mathrm{Vfac}\left(\mathrm{Va}^{\mathrm{Vlaw}}\right)$$

This is the internal rectifier model to which consideration should be made for the series resistance Dres, if used.

For example, if the 5AR4 example above is used, and Va is 20V, then Ia = (20 ^ 1.5) x 0.00396 = 0.354A.

Adding your own models

This is best achieved by using a spreadsheet. Make up a simple calculation with three variables Va, Vlaw and Vfac. The result (Ia) should be calculated in accordance with the formula above. Select values which correspond to a number of points chosen on the data sheet.

Using more modern spreadsheets, it becomes easier as the information from the data sheet can be entered in for a series of points and use goal-seeking functions to obtain the parameters for Vlaw and Vfac.

Table 3 shows parameters for 2 diodes, and the diode voltage-current characteristic can be plotted using the webpage www.duncanamps.com/psud2/show_rect .

Name	Type	DRes	Vlaw	Vfac	Vpiv	Ipks	Ipkr
1N4007	SS	0.042	20	9.353	1000	30	6
GZ34	VT	0.0	1.5	0.00396	1500	3.0	0.75

Table 3. Diode model parameters used in PSUD2

Figure 24 overlays the PSUD2 model points on a 1N4007 datasheet, but note that datasheets can show a variety of test conditions, such as different junction temperatures, and different duty-cycles.

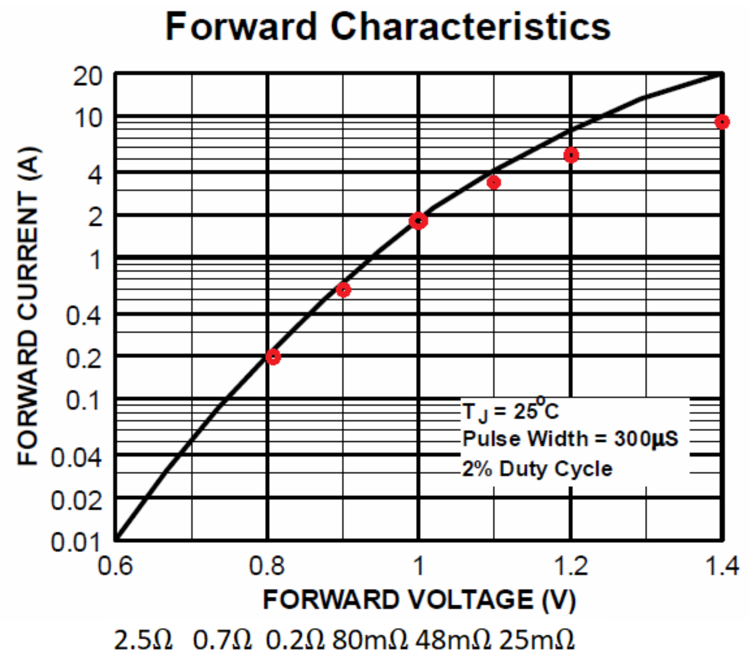


Figure 24. 1N4007 diode characteristic with PSUD2 model values.

5 Power supply rails – sequencing & options

The power supply of a common valve amplifier generates B+ prior to the signal valves starting to conduct. This is especially the situation when power supply diodes are solid-state, or have directly heated cathodes, and typically causes all signal stage bypass capacitors and coupling capacitors to charge to the un-loaded B+ level. Even with indirectly heated valve diodes, B+ can reach the un-loaded B+ level for a second or two depending on the output tubes used. Some mods like The Port Arthur Rectifier in a 2005 book by Gerald Weber, or the Weber Copper Cap replacement, inadvertently disable the slow start provided by a GZ34. Two issues arise from this start-up sequence – peak diode current and peak capacitor voltage.

The typical sequence of firstly charging power supply capacitors, and subsequently suppling load current, alleviates the peak current level experienced by power supply diodes.

The issue of over-voltage stress on coupling and bypass capacitors can be managed by using capacitors with voltage ratings that meet or exceed the anticipated turn-on stress levels. However, this can be onerous for choke-input type power supplies or amps with a high B+. And it is not uncommon to see modern amps use capacitors with marginal voltage ratings – perhaps from a perception that bypass caps for pre-amp circuitry will be protected by power supply dropping resistors, or that modern capacitors can survive short duration over-voltages. How well an electrolytic capacitor can manage over-voltage is assessed in Section 6.

5.1 Suppressing over-voltage stress at turn-on

A few methods are in common use to avoid over-voltage stress at turn-on, based on slowing down or delaying the power supply turn-on action.

Delay relays that use a thermo-mechanical (bi-metallic) switch can be found in vintage equipment for some specialist applications like mercury arc rectification, transmitters and electrometer low leakage valves that specifically require a delay between heater application and HV application.

Nowadays a standard electro-mechanical relay can be used for delay applications by allowing a delay time circuit to activate the relay coil. As with a vintage delay relay, the relay contact can be used in a few ways, such as:

- to turn on the primary voltage of a B+ supply transformer – requires separate filament transformer.
- to turn on the secondary HV AC supply to the rectifier circuitry – this may require the contact to have an AC voltage rating that exceeds the capability of common relays.

Semiconductor devices such as triacs, SCRs, FETs and solid-state relays can be an alternative to a relay contact [20].

Slowing down the rise in B+ voltage can be achieved in a few ways, including the use of resistance added in series with the power supply circuitry such as:

- adding a negative temperature coefficient (NTC) thermistor in series with either the primary or secondary current flow can provide a short duration delay, although in practise a practical NTC part is typically not sufficient to achieve more than a few seconds of delay.
- adding a resistor or NTC in series with either the primary or secondary current flow, and including a relay contact in parallel with the resistor/NTC to short it out of operation after a delayed period of time. The delay period can be from a timer circuit, or by the rise of load current itself (if the coil is part of the load), or by voltage rise across the relay coil. A practical design can suppress initial B+ for longer but will depend on circuit design.
- adding an NTC or PTC in series with a valve diode's heater – this technique delays the time it takes for the power supply diodes to conduct, and its use is not common so is discussed in detail below.

Combining both a delay time and a slowing of rise time, as sequential actions, can also be practically achieved.

5.2 *Power start-up influence on valve diodes*

The only known valve diode part with a start-up condition is the Tung Sol 5R4GY, whose 1945 datasheet shows a characteristic graph with output current versus AC voltage, for two situations:

a) normal start up (heater and plate voltages applied at same time)

b) heater applied separately for 10 secs first

The datasheet graph implies that diode continuous peak current level must be derated until the heater is at normal operating level. Note that this diode is an early type of directly heated cathode with no specification for transient peak current, as is typically provided for latter generation valve diodes. PSUD2 can be used to show that the transient peak current allowed for this diode appears to be quite low – about 2x the continuous peak level.

5.3 *Power start-up influence on output stage valves*

Many audio amplifiers have the output stage exposed to B+ before the output stage valve(s) start to conduct. Valve conduction starts as cathode temperature increases sufficiently, with the locus of conduction starting from zero current, and increasing to idle point at the final B+ level.

If the output stage valve cathodes were hot already then load current and voltage could rise as indicated in Figure 25, being dependent on the bias circuitry, and possibly being influenced by other circuit time-constants such as from the driver coupling capacitor and output stage grid leak, or filtering on the output stage screen supply. For example, a coupling capacitor to an output stage valve may have a time constant exceeding 100ms for a hi-fi type circuit, and may force a heavy load on the power supply from both output stage PP valves if B+ rises too quickly.

The Standby switch in an amplifier has been described as having multiple uses – one use was to introduce a delay in applying B+, to avoid cathode degradation in output stage valves. This issue has divided views.

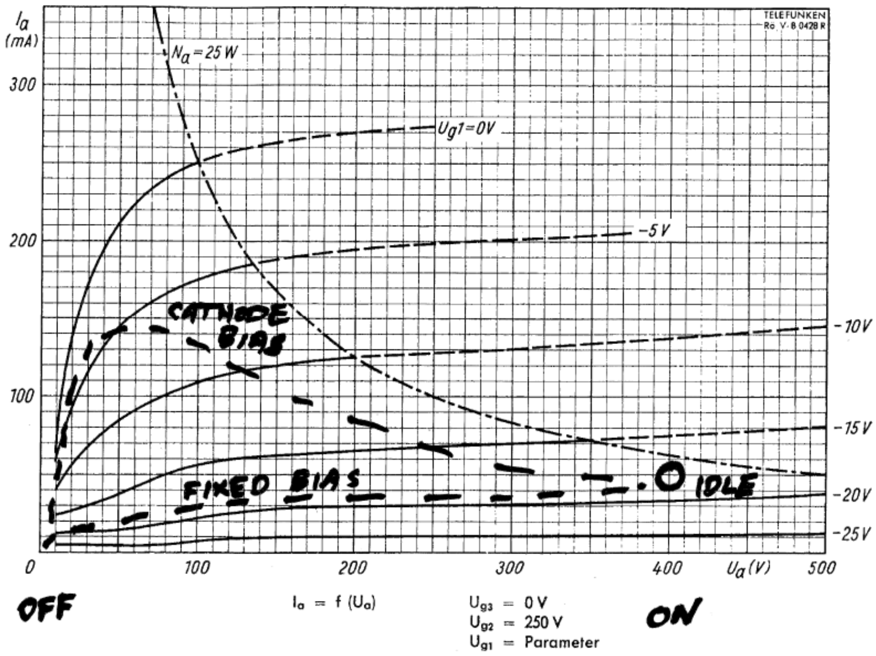


Figure 25. Example turn-on loci for hot output stage valves.

5.4 *Valve diode heater in-rush and delay*

A valve diode can cause a large heater current in-rush, as many large diodes have a 2 or 3A rated heater, and the in-rush could initially peak to about 5x the normal operating current. Just for this diode valve, the in-rush VA requirement could be up to about 75VA, and could take at least a second to substantially reduce. Time to conduct of a directly heated cathode diode like a 5U4 can be about 2 seconds, compared to an indirectly heated cathode diode like a 5V4 taking about 10 seconds to start conducting.

An NTC thermistor can be added in series with a diode valve heater, although this is rarely seen [1]. An NTC with a current rating in excess of 2-3A, and a low hot resistance appropriate for this heater application, is not readily available, as the hot resistance would need to be no more than about 100 mΩ in order to reduce voltage drop to below 0.3V - a 6% reduction in applied heater voltage. NTC parts with hot resistance of at most 100 mΩ typically have max current ratings of 6-8A, and won't reach a sufficiently high operating temperature at 3A to sufficiently reduce operating resistance.

As an example, a 3D-15 NTC part has a hot resistance below 100 mΩ with its 7A max rating. At 3A, the part temperature is ~136^{o}C in free air. Covering the part with a 10mm thick aerogel sheet (a cheap and simple form of thermal insulation) reduced the part resistance to ~130mΩ [2] at 3A. Even if the operating resistance could be reduced enough by adding thermal insulation, the cool-down time of the part would be extended, and the part could take a long time for its resistance to reset when mains AC is turned off.

One option is to use a relay contact to apply a short across the NTC part after sufficient time dely. Diode heater current will then go through the relay contact, effectively bypassing the NTC thermistor and allowing it to thermally reset during amp operation, and the heater operates at nominal AC voltage. The relay coil can be conveniently powered from the voltage across the heater itself, as that heater voltage is initially suppressed by the NTC being in circuit, and the heater voltage increases when the NTC is bypassed and so acts to latch the relay coil on. The NTC means the relay contact only has to switch and conduct the nominal hot heater 2-3A, and so avoids switching a cold heater current of circa 10-15A.

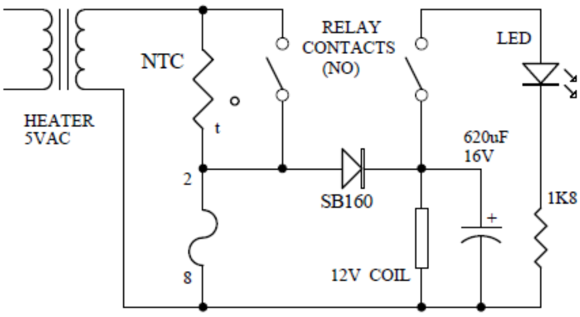


Figure 26. Diode heater delayed using NTC shorted by relay contact

AC coil relays are not commonly available, whereas a relay with DC coil is very common. The AC voltage across the heater can be rectified and filtered to provide DC, with a range of design choices including using a half-wave, full bridge, or doubler rectifier circuit, using common diode choices of PN or Schottky, and the filter capacitance value. Just using a rising DC supply voltage to energise the coil may cause the contact to chatter and then pull in and latch, but once the contact pulls in then contact resistance is low and is not affected by coil voltage (which can be significantly below the rated coil voltage). Once the contact is pulled in, then it is effectively latched, as the Vac available for the coil increases (as the NTC is shorted out), and the coil will only drop out when DC volts has fallen to typically 50% of the pull-in voltage. Each relay type is likely to need a custom configuration, but once determined the operation is consistent. The circuit in Figure 26 uses a 10.5V DC coil relay with 10A DPDT contacts (Carlo Gavazi MZPA002 44 05) and pulls in after 20 seconds at about 4.5Vdc using a half-wave Schottky rectifier that latches on without chatter for a 5V heater with 5U4 rectifier and 5D-15 NTC. In that circuit, the 3D-15 is not appropriate as it allows B+ to rise before most output stage valves start conducting, and an 8D-15 has a long pull-in time of about 45 secs.

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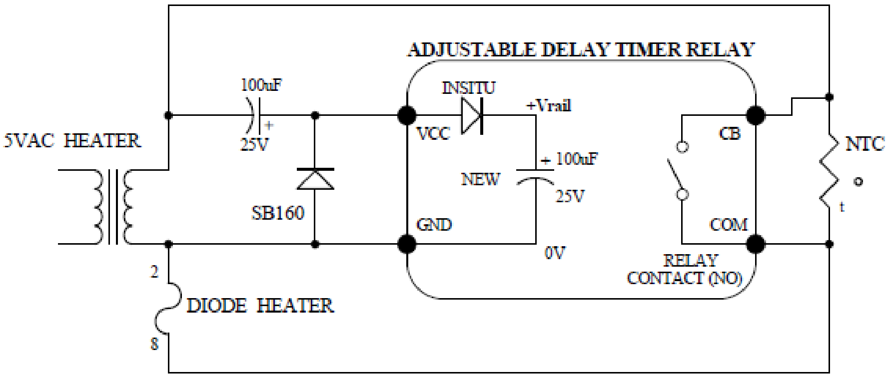
[1] The use of a thermistor to gradually apply heater voltage was used industrially in the 1950's on the first large computers to lower the failure rate of valves [16].

[2] Testing a 3D-15 in free air gave ~113mΩ for a surface temperature of ~178^{o}C at 5A. Part resistance reduced to ~94mΩ for a surface temperature of ~204^{o}C at 6A.



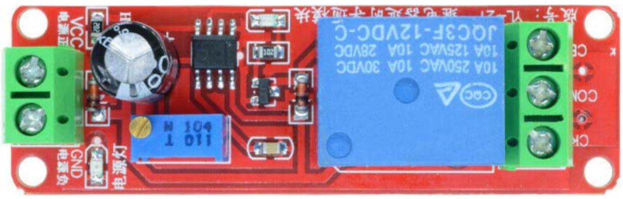


Figure 27. Common eBay adjustable delay timer relay switch module.

Another option is a timer relay module like that shown in Figure 27, and although rated as 12Vdc, can be powered from 5VAC using a doubler rectifier comprising two 100uF 25V caps, and two diodes (the module conveniently has one of those diodes in place for reverse polarity protection).

A PTC thermistor may also be an option for delaying conduction of a valve diode and hence delaying B+ voltage rise. The PTC could be sized for nominal conduction of a 2 or 3A heater supply, but could also dynamically raise its resistance during a power-on event, and hence suppress heater current for a short duration. The PTC could also have the advantage of acting as a resettable fuse for the heater circuit, and reducing heater supply voltage by about 0.6V if mains voltage was nominally high. This PTC technique has not yet been performance tested.

5.5 Sequencing options

An indirectly heated rectifier valve can be used with an NTC and delay relay contact bypass (as in Figure 26 and Figure 27), such that the NTC inhibits diode heater current inrush for about 1-2 secs, and then the delay contact bypasses the NTC such that the diodes start conducting at the same time as the signal valves start loading B+. Benefits are suppressed 5V heater in-rush, suppressed HV winding / diode surge current, and controlled sequencing of B+.

A directly heated rectifier valve can be used with an NTC and delay relay contact bypass (as in Figure 26 and Figure 27), such that the NTC suppresses diode heater current for about 10 secs, and then the delay contact bypasses the NTC to allow full heater current as the signal valves start loading B+. Benefits are suppressed 5V heater in-rush, suppressed HV winding/diode surge current, and controlled sequencing of B+. The NTC must have a high enough cold resistance to delay diode conduction for about 12 secs.

If there is not enough delay introduced by the NTC itself, then one alternative is to use two delay modules in step sequence as per Figure 28, where the first module delays the start of heater conduction, and the second module then restrains the heater current with an NTC.

An ss diode B+ rectifier can use two delay modules operating in step sequence to delay and slow the B+ rail, as in Figure 29. The first delay module is timed to allow the signal valve heaters to warm up to a point where they would start conducting. The second delay module uses an NTC with relay contact bypass, such that relay contacts do not carry surge currents, and also allows B+ to rise softly via the NTC and then to operate as per normal, with the contact/NTC located in the full-bridge CT or diode bridge 0V leg. Benefits are suppressed mains AC in-rush current, suppressed HV winding/diode surge current, and controlled sequencing of B+. This scheme places one end of the contact/NTC at 0V ground. The eBay module only has minimal trace separation between relay contacts and timing circuitry, so the relay contact and heater supply should be operated with minimal voltage difference.

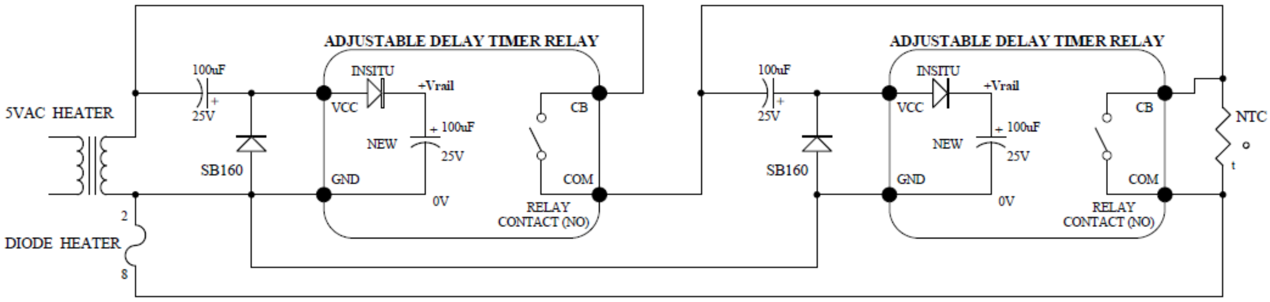


Figure 28. Adjustable delay timer relay switch modules for heater in step sequence.

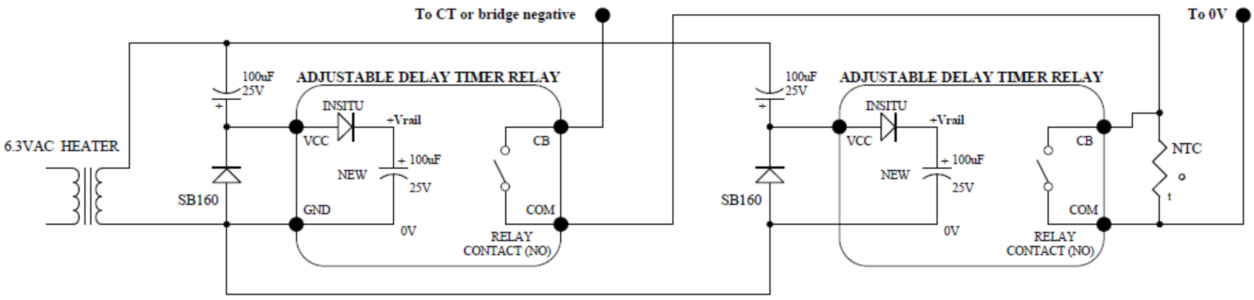


Figure 29. Adjustable delay timer relay switch modules for CT link in step sequence.

In addition to the above, an NTC and delay module could be used to soft-start the heater current to all signal valves for about 1-2 secs duration. Benefit is a suppressed AC mains in-rush.

Controlled sequencing of B+ aims to avoid any turn-on over-voltage surge to the amplifier's filter and bypass capacitors.

5.6 Rail options

Many tube amplifiers generate a high B+ rail voltage for the output stage, and then use R-C droppers to generate progressively lower local B+ rail voltages for earlier amplifier stages. Some power supplies allow a 50% rail voltage to be easily generated as part of a power transformer secondary winding configuration with a CT, which can be very convenient for output stage valves that need a significantly lower voltage for screen and preamp stages (such as for TV line output valve screens, or for valves operating pentode mode with say a 250-300V screen limit).

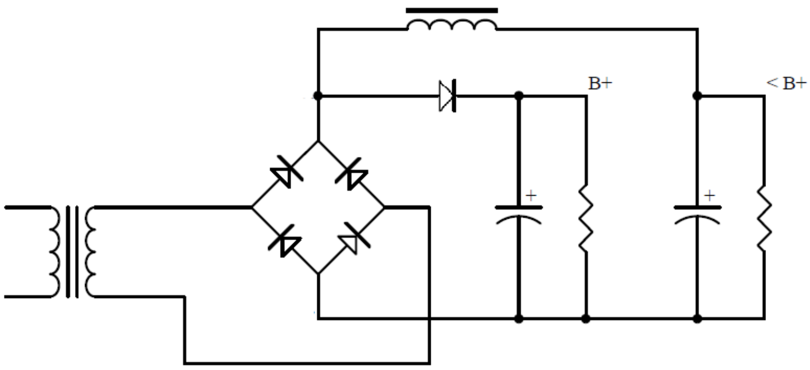


Figure 30. Capacitor and choke input filtered rails from same diode bridge.

An uncommon configuration uses a diode bridge with capacitor input filter to generate B+ for the output stage, and also uses the same diode bridge and a choke-input filter to generate a lower B+ rail for screen and preamp supply. An extra diode is used to generate the capacitor input filtered rail, in order to isolate the filter capacitor from the voltage waveform available at the diode bridge. Given the choke input filter, it is recommended to use fast diodes like UF4007, as per discussion in section 3.

The load current requirement for screen and preamp supply is typically much lower than for the output stage B+, and so a relatively small choke can be utilised (in lieu of say a FET based regulator, or a dropper resistor or Zener diode, all of which may dissipate substantial power). An advantage of the choke-input option is that the rail's ripple voltage can be quite low. PSUD2 is useful to estimate a choke inductance that is just sufficient to maintain critical inductance for the load current requirement, as a choke with less inductance will typically be physically smaller and have a lower series resistance (DCR). PSUD2 would estimate the lower rail voltage level for the given choke resistance, and if that voltage is lower than desired then adding a small value capacitance before the choke can raise the rail voltage.

6 Electrolytic filter capacitor over-voltage stress

When a typical valve amplifier is turned on, the power supply initially generates a B+ level that is higher than normal, as the signal valves haven't started conducting. This is especially the situation when power supply diodes are solid-state, or have directly heated cathodes. This section discusses the ability of electrolytic capacitors to survive a short period of over-voltage stress.

Vintage capacitor manufacturers often specified electrolytic capacitors with a rated (or working) voltage and a surge (or peak) voltage, such as on common 8uF and 16uF chassis mounted cans. The photo shows a Ducon can with 500V VW (max working voltage), and 600V VP (max peak voltage). Some Ducon cans just had a 600VP rating (ie. no VW rating). A 1960 UCC advert specified:

PEAK OR SURGE VOLTAGE:
Is the maximum voltage the capacitor will withstand without damage when connected in series with the resistor having a value of $\frac{20,000}{C}$ ohms for 30 seconds: the voltage is measured across the resistor/capacitor combination.

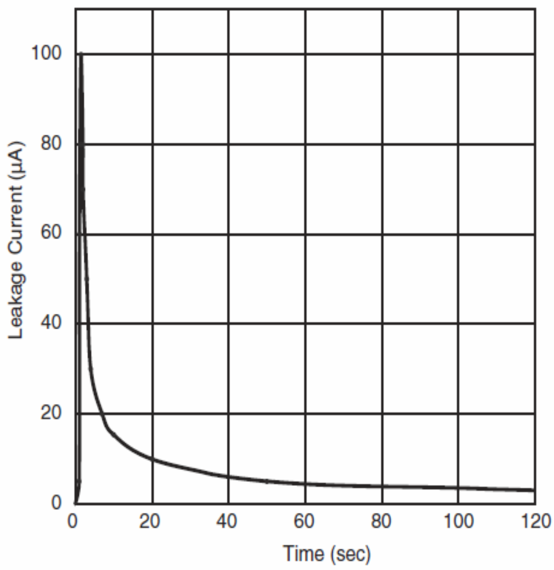
Vintage magazines from the 1950's often had commentary that these capacitors had a short service life, and likely when they were the first filter capacitor directly following the rectifier. The example UCC peak/surge requirement would need a 1k2Ω series resistor for a 16uF e-cap, which would likely only be applicable to filter capacitors located along the resistor dropper chain, and not for the first filter capacitor.

In the 1980's, datasheets from manufacturers often used a 110% surge rating for caps with >350V rating. Nowadays the datasheets from many manufacturers only provide a rated maximum voltage, and datasheets or application guides often state not to exceed that maximum operating voltage.

However, from circa 1977 the standard IEC 60384-4 that covers testing of fixed capacitors has included a surge voltage test (clause 4.14) that applies a surge voltage of 110% of rated voltage for cap rating > 315 V, through a source resistance where RC = 0.1 sec for 1000 cycles of 30 secs on and 330 secs off, and the outcome has to be the same leakage current spec and no more than 15% capacitance change. From [19] it is likely that the surge test is applicable in order to comply with the standard, although it is uncertain if the test is done at rated maximum operating temperature (as possibly implied by [18]), or just at 20^{o}C where leakage current is specified.

The IEC 60384-4 surge voltage test relates to an example 1kΩ resistance in series with a 100uF e-cap, which is likely not valid for the first filter capacitor location, but may exist for screen and preamp type filter capacitor locations that are buffered by feed resistance.

At power turn-on, a filter capacitor draws current from the rectifier as part of charging its voltage up to the B+ level, but also to form up the insulation layer with a current called leakage current. At turn-on the leakage current surges to a level that can easily exceed 10 to 100x the normal level, and can take many minutes to subside, as indicated by the generic plot.



For an example 24uF 500V vintage e-cap where leakage may be 50uA, the initial leakage may exceed many mA, which along with charging current can suppress preamp B+ levels and provide a form of safety pre-load. However, the first filter capacitor after the rectifier Is not afforded such protection, especially with a solid-state rectifier, so that filter capacitor requires sufficient voltage margin and not rely on its surge voltage rating.

A key concern of operating a capacitor marginally over its continuous max voltage rating is leakage current. Nichicon [17] provides manufacturer information related to over-voltage operation of their electrolytic capacitors, and the influence of leakage current. Modern cap manufacturers have effectively supressed the influence of impurities on capacitor leakage current and aging, and provide a package that supresses any external influence but still allows safety venting under high internal pressure. As such, the operation of a capacitor at marginal over-voltage is definable to a point that it could be relied upon from a practical design and operation viewpoint.

Figure 31 shows the leakage current versus voltage characteristic for a 35V cap that is likely to be generic for higher voltage caps. The plot shows that leakage current increases as applied voltage increases, and that during manufacture a forming voltage is applied that is above the rated surge voltage. The forming voltage appears to be at about the maximum level that can be applied whilst maintaining leakage current under a limit where internal power dissipation is managed. The sustainable level of voltage reduces as internal capacitor temperature increases.

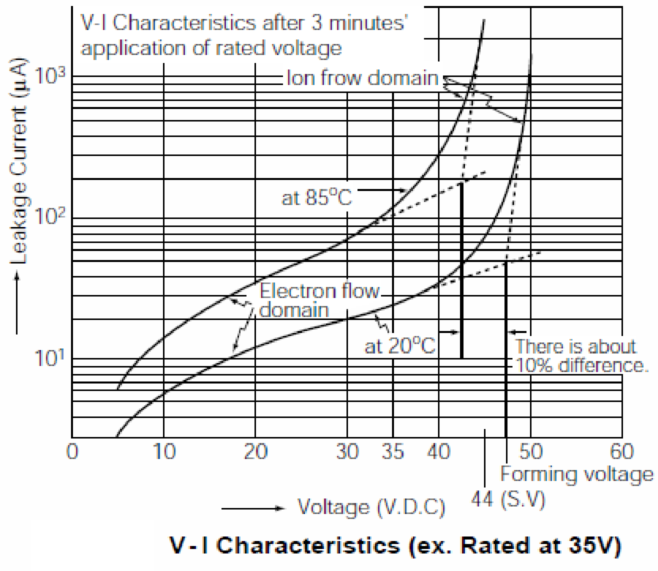


Figure 31. **L**eakage current characteristic versus voltage [17].

Evox Rifa [18] identify how, in practise, internal partial discharges start to occur as voltage exceeds rated voltage. The paper indicates how a capacitor's leakage current changes during the 30 second application of surge voltage for the IEC60384-4 test, and indicates that peak initial leakage can be substantially high compared to rated leakage current, especially for one-off applications of power to an amplifier where the next pulse may occur hours or days later.

The Evox Rifa paper indicates an inherent self-healing process for certain forms of partial discharge whereby capacitor performance is not degraded. Above some voltage level, the number and severity of partial discharges rise rapidly in an avalanche behaviour.

A 2011 paper [19] describes relevant test results that show the increase in leakage current as applied voltage is ramped up and beyond the rated capacitor voltage. Example plots show partial discharge events occurring with an increasing severity as voltage rises beyond the surge voltage rating (although the ramp rate is unfortunately not disclosed).

From the three referenced papers it appears practical to allow B+ filter capacitors to peak to 110% of rated voltage for the infrequent, one-off surge that occurs when a valve amplifier is turned on and where the surge typically lasts no longer than 10-12 seconds, even with solid-state rectifier diodes. This is especially the situation where the amplifier has had time to cool down before turning it back on again. The proviso is that the e-cap has not deteriorated (eg. significant loss of electrolyte although still within capacitance spec), or been non-operational for a long duration (eg. perhaps >5 years, such that a higher than nominal leakage current occurs as part of maintenance reforming).

It may be practical for an even higher voltage to be reliably applied, as an over-voltage margin may exist for a number of reasons, including:

- Low capacitor temperature at power on.
- Recent use of the amplifier (due to lowered surge level of initial leakage current).
- The capacitor has not degraded for its age (due to lowered default leakage current level).

- The effective series resistance of the supply is substantial when a partial discharge occurs in a capacitor (especially for screen and preamp filtering with buffer resistance).
    - A valve amp B+ power supply with solid state diodes could have an effective series resistance of well below 100Ω. In contrast, the effective series resistance of a valve diode may exceed 100Ω when there is practically no charging current (ie. after the filter capacitor has charged, and prior to any rising load current).
- The manufacturer has incorporated additional margin, whether by design, or material availability, or accounting for statistical variation between parts.

However, any daily or seasonal variation in mains AC supply voltage can equally counter any perceived margin, and a 5-10% swell in mains AC voltage could easily push peak capacitor voltage into a region where partial discharge events could start to avalanche, and the capacitor may be degraded.

After a long period in storage (eg. Nichicon recommend > 2 yrs), it is recommended to limit the in-rush current in to filter capacitors whilst they reform (and possibly experience an increased number of partial discharges) for at least 30min (some manufacturers recommend at least an hour). Storage without applied voltage degrades the oxide layer, causing increased leakage current when voltage is next applied, as the capacitor attempts to reform the oxide layer. As an example, KEMET recommend limiting reforming current to at most 5mA for their ESH range.

Mains voltage reduction is one means to reduce initial capacitor reforming current (eg. a variac, or a step-down transformer) but only the variac can raise the voltage in a timely manner up to normal operating level. Another simple technique is a light-bulb limiter with a low wattage bulb whereby the bulb is bright during initial reforming (and hence initial capacitor current is being significantly limited). One custom technique is to replace the power transformer secondary fuse in the CT lead with a resistor (eg. 1kΩ 5W) and remove the output valves, but that requires monitoring that amplifier voltages do not become excessive. Another custom technique is to temporarily insert a resistor in series with the first filter capacitor. Capacitors downstream of the first filter capacitor likely have sufficient series dropper resistance to avoid leakage current stress.

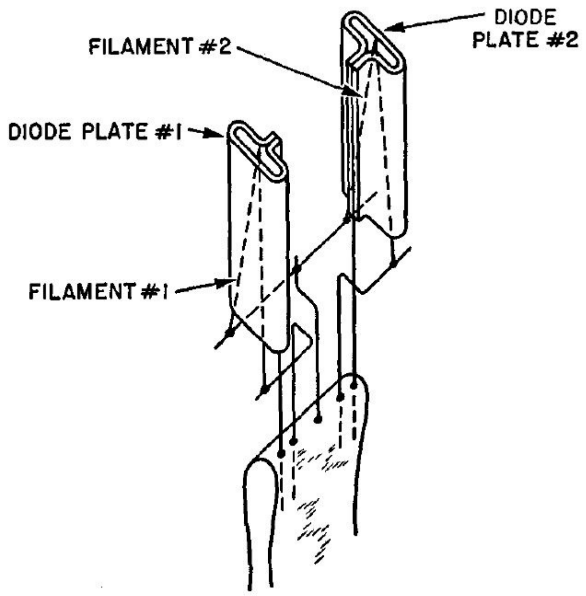
Capacitors that have not been used for decades require even more caution to avoid initial over-heating during the start of reforming (eg. using a series resistor >> 1kΩ). Many restorers just install new e-caps, but some like to reform vintage e-caps especially if the original e-cap seal appears to be in good condition and the visual aesthetic of the amplifier benefits from retaining the original capacitors. Many vintage e-caps in commercial equipment only had rudimentary seals and it is commonly found that the e-cap can't be reformed adequately, or that subsequent use of the cap in an elevated temperature environment within an amp would be quite a reliability risk. In contrast, some quality e-caps from the 1960's have excellent seals, and although some loss of liquid is anticipated, they may well reform nicely and be reused if their measured levels of capacitance, ESR, and leakage current at rated voltage are nominal.

Also note that e-caps operated for many years/decades typically only experience the actual operating voltage provided by the equipment, which can be significantly below the rated e-cap voltage. Hence the e-cap may have low leakage current at the nominal operating voltage but likely exhibits a relatively high reforming current if ever the supply voltage increases, which could cause noticeable effects. This indicates that maintenance on equipment that is older than say 10 years should consider reforming power supply e-caps up to their rated voltage even when that action is onerous.

7 Valve diode operational issues

7.1 Dual diode directly heated valve options

A dual diode rectifier valve has two separated cathode-anode structures. For the directly heated cathode types, like 5AS4, 5Y3, 5U4, 5X4, 4W4, 5T4 and 5R4 models, the heater wire inside the anode structure is the cathode (also referred to as the filament). Typically, the filament is a series connection of heater wires, as shown below in Figure 32, and with no external connection to the central joint.



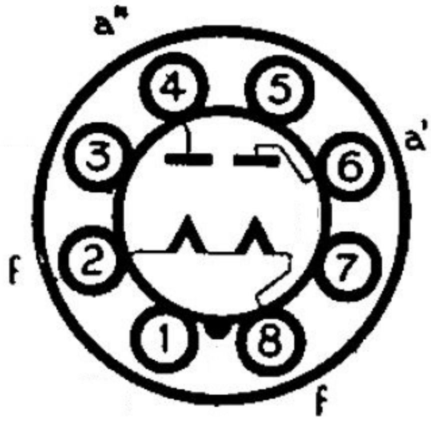


Figure 32. Directly heated filament configuration.

In quality vintage equipment, the 5V heater supply winding on a power transformer would come with a centre-tap, and the centre tap (CT) would become the output terminal of the rectifier that connects to the first filter capacitor or choke. This arrangement places the heater CT at effectively the mid-point of the diode heater, and means each diode is alternately exposed to the same magnitude of voltage waveform as the other diode. The schematic below indicates that B+ = V – Vd – 2.5V, or B+ = V – Vd + 2.5V, depending on the phasing of the heater winding to the HV winding.

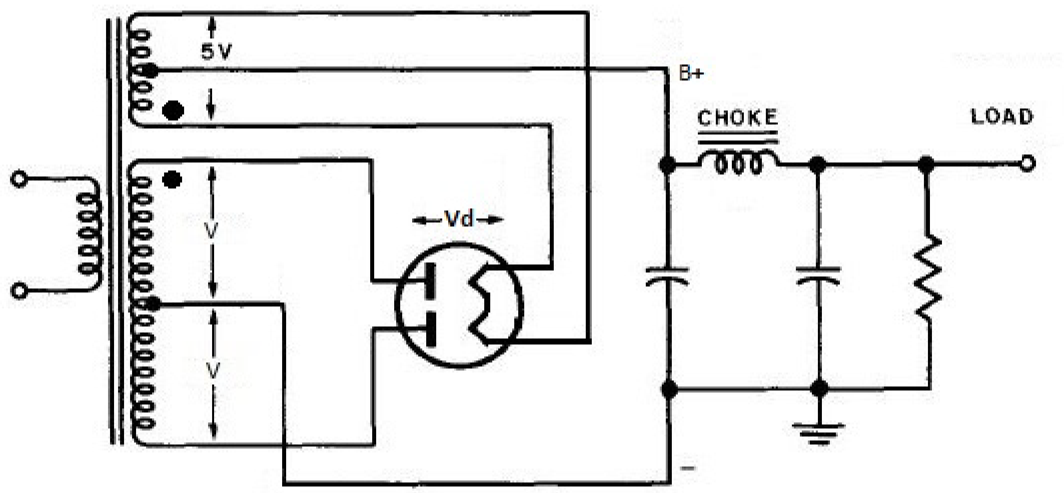


Figure 33. Typical power supply circuit with 5V centre-tapped heater winding.

When diodes are well matched (as well as HT winding DCR for each V winding), this configuration achieves matched rectification of current pulses, with minimum 2H and H ripple harmonics (2H is the typical 2x50=100Hz ripple, and H is fundamental mains 50Hz residual ripple due to asymmetry).

When a power transformer's heater winding does not have a CT, then the output terminal of the rectifier can either be pin 2 or pin 8 (although connecting to pin 8 is preferred to allow an indirectly heated rectifier to be used). The schematic below indicates (in simple terms) that B+ = V – Vd for one diode, and B+ = V – Vd + 5V or B+ = V – Vd - 5V, depending on the phasing of the heater winding to the HV winding. As such, this configuration will lead to some additional asymmetry in the magnitude of voltage waveforms to the two diodes, and so increase the level of H and 2H ripple harmonics if the diodes had matched on-voltages. If the diodes in a valve have un-matched on-voltages then swapping the heater phase (or swapping connection to pin 2/8) may reduce the H ripple harmonic.

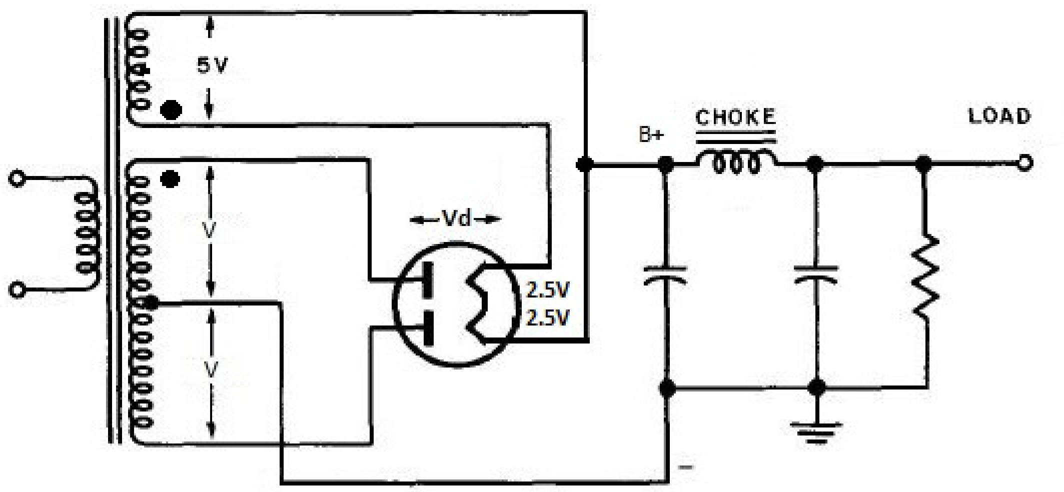
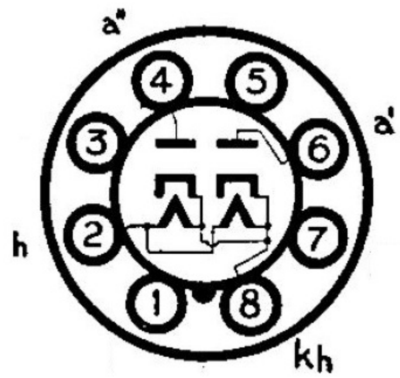


Figure 34. Typical power supply circuit with 5V heater winding (no centre tap available).

Although not substantial (2.8Vdc difference for 310Vdc B+ with 5Y3 rectifier in a small amp), the option to slightly increase or decrease B+ by means of swapping the phasing of the heater winding, may have use. This situation also relates to doubler rectifier configurations which use valves like the 5U4G where both anodes are paralleled for higher current. For the doubler situation, the HT winding voltage is about 50% of that used for a bridge rectifier, so the percentage increase or decrease of B+ due to rectifier heater phasing may be of more influence. However, a doubler has two separate rectifier heaters (the two 'diodes' in a doubler do not have a common cathode) and ripple is more, making it difficult to discern the effect of heater phasing of each diode.

In contrast, the indirectly heated cathode type dual diodes use a heater/filament that is fitted inside a tubular cathode housing, similar to the heater/cathode assemblies used by typical output stage valves. As such, each heater filament is 5V rated, and so the two diodes have their separate filaments connected in parallel to the terminal pins 2 and 8. Each heater filament only needs to be functionally insulated from the cathode.



The plate structure is also noticeably different, with the cathode/heater tube structure located inside a slightly larger cylindrical plate tube, which has cinched edge strips that extend outwards and act as two large thermal radiation fins.

With pin 8 (cathode) connected to B+, each diode is alternately exposed to the same magnitude of voltage waveform as the other diode. Only differences in cathode emissions, and power transformer winding resistances (typically about 5%, but can be up to 10%), cause noticeable difference in diode peak currents.

7.2 Dual diode valve emission difference

The difference in cathode emission between each diode in a dual-diode rectifier valve can become quite substantial from aging, and go unnoticed in an amplifier application until some form of gross performance change occurs. A 2015 forum thread showed this difference in a scope plot (Figure 35) of current waveforms on the power transformer secondary CT lead of an amplifier, with a random used 5AR4 valve.

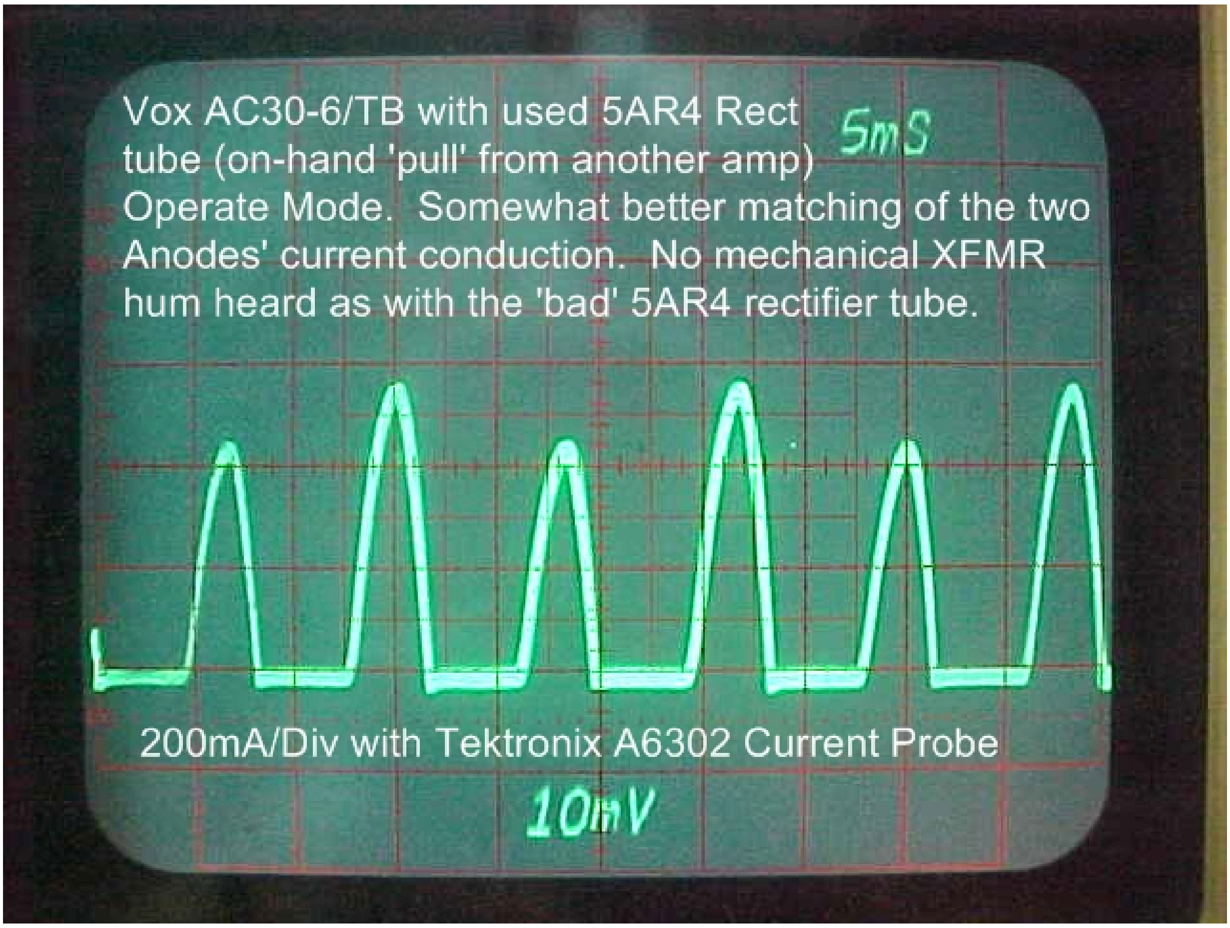


Figure 35. Rectifier current pulses on CT connection.

To illustrate the variation that may exist between the two diodes in a dual diode rectifier, a batch of 29 5U4G/5AS4 valves, of unknown history, were measured for their current difference between the two anodes when forward conducting with 24V, 36V and 49V drop across each diode. Most valves showed good balance between the two diodes, with only a few having one diode conduct less than 80% of the current of the other diode (for the same applied voltage). 40% of the sample showed at least 90% balanced conduction. Figure 36 is a plot of the 29 samples tested, showing the % balance in plate currents.

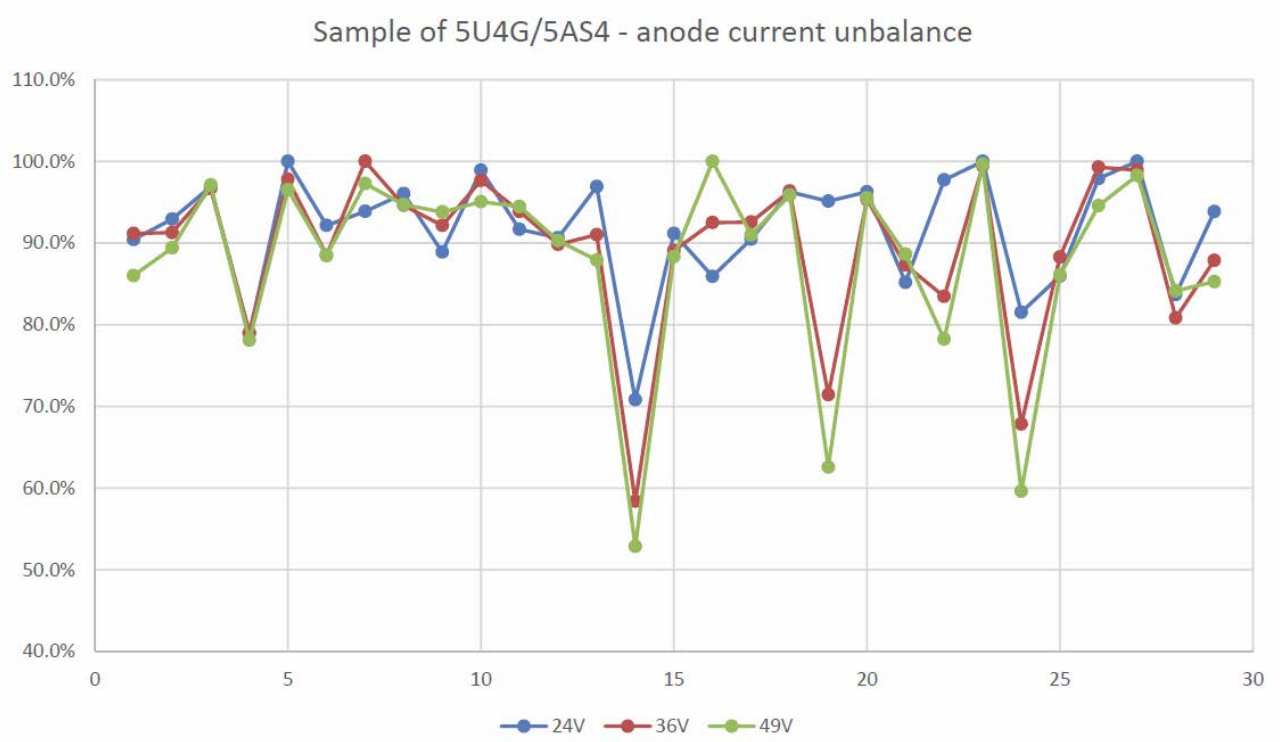


Figure 36. Anode current imbalance for a batch of 5U4G/5AS4 rectifier diodes

Across the batch of 29 valves, there was a wide spread of current level conducted at the three applied forward voltage levels, as shown in Figure 37. The spread in current level increases significantly with applied forward voltage. With reference to the 5U4 datasheet, plate voltage drops 49V for 180mA in a bogey valve, indicating that 47 of the batch of 58 anodes (ie. 81%) at least meet bogey performance. For a 36V drop, the number increase to 49 of 58, and for 24V drop the number is 54 (ie. 93%). This variation aligns with drooping cathode emission with lifetime.

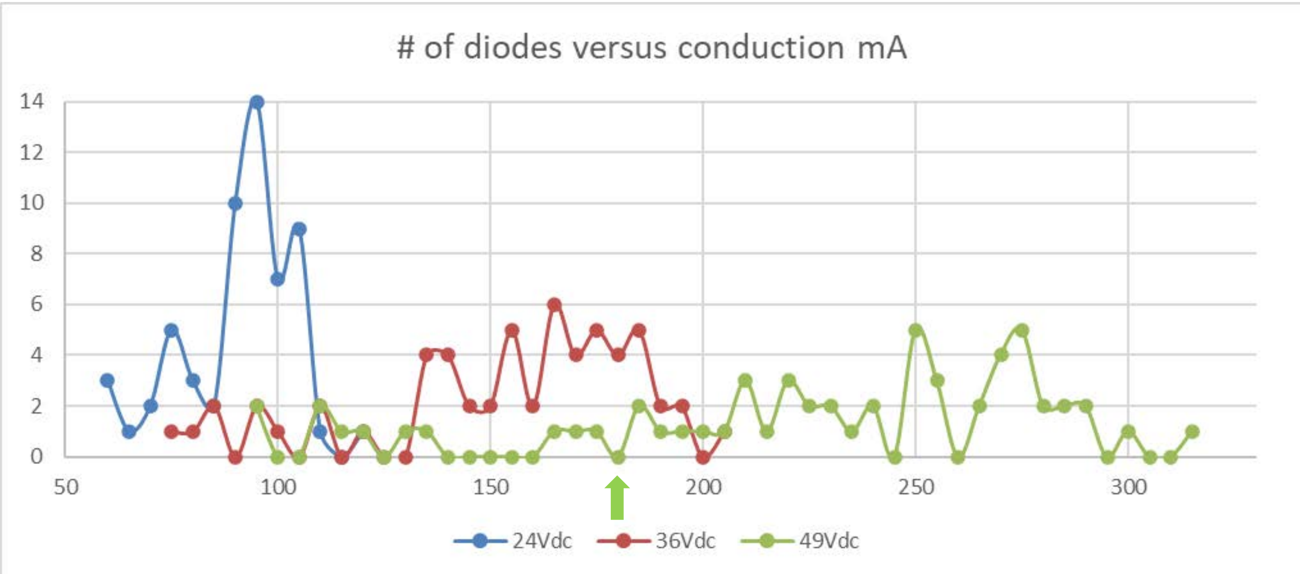


Figure 37. Number of diodes in batch with mA anode current levels (5mA span).

In service, a 5U4 plate is rated to continuously conduct up to 675mA peak per half mains cycle, whereby the anode drop is likely to be circa 76+60=136V by inspection of the datasheet and assuming a constant resistance curve. Although there is no transient peak plate current rating, the 5U4 is likely to have a capability in excess of 2A (eg. see discussion in 4.4), however the plate voltage drop likely becomes a very large percentage of the transformer winding voltage beyond 1A.

7.3 Valve diode PIV degradation

Internal arcing within valve rectifier diodes is not uncommon, and effectively presents a low-resistance across the entire V-0-V power transformer secondary HT winding (see Figure 38), as one anode is conducting to its cathode in a normal manner, and then the other anode arcs to its cathode rather than presenting an open-circuit that can withstand the presented PIV. The peak winding current is then likely similar to the first pulse in a hot turn-on event, and could occur during each mains half-cycle.

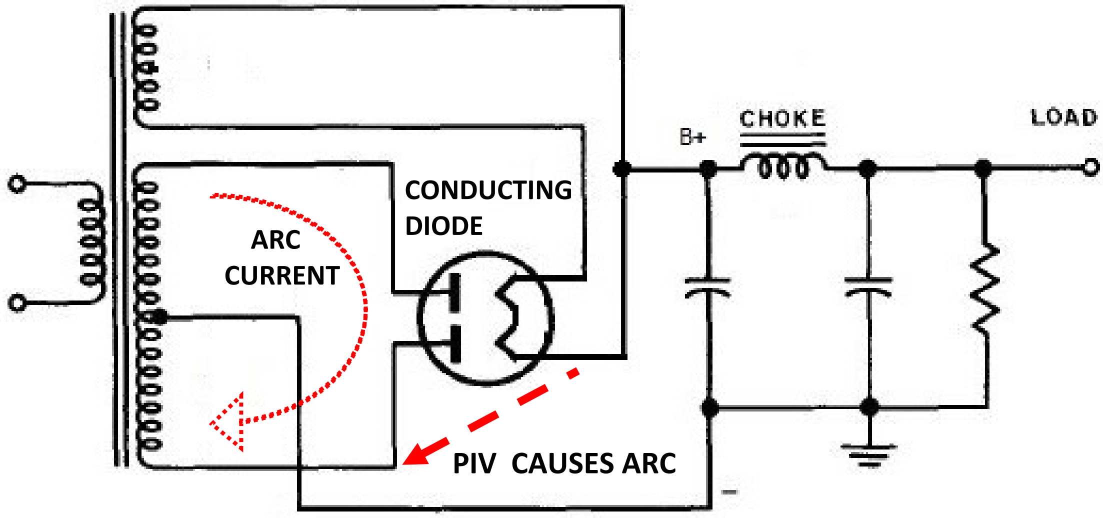


Figure 38. Illustration of arcing current in diode causing low resistance across transformer secondary winding.

The PIV performance of valve diodes can be tested using a modern Insulation Resistance (IR) meter (ie. megger). Many valve diodes have a PIV rating of typically 1.4 or 1.55kV, exceeding the 1kVdc limit of most IR meters. To illustrate the use of a generic IR meter (QM-1492 with 2GΩ 1kVdc capability), a batch of 28 5AS4/5U4G valves were measured for anode to cathode/filament resistance at 1kVdc. 14 showed IR > 2 Gigohm (ie. <0.5 microamp leakage) for both anodes, both initially when cold, and after the heater was on for a while. 5 valves showed both diodes with a relatively low IR (eg. < 100 Megohm). 3 valves initially showed some measurable leakage, but IR then recovered to > 2G when heaters were powered – and one valve showed the opposite effect. 5 valves showed one diode in the valve with IR > 2G, but the other diode with a measurable level of leakage.

13 valves that showed no leakage at 1kVdc PIV were re-tested at 1.5kVdc PIV continuous (using a special rig). 7 showed no leakage. 1 showed just a bit of leakage in one diode, 4 showed at least one diode with no more than 3uA leakage, and 1 showed more than 30uA leakage in one diode.

If a valve had gone gassy, then it is likely that each diode would exhibit a similar level of IR at the same PIV. Of the tested batch of 28 valves, 8 showed likely gassy symptoms, with 5 showing both diodes <100Megohm at 1kVdc, and 3 valves initially showed some leakage at 1.5kV but then recovered when powered.

Another likely leakage path is from filament (or cathode) vapour deposition on to mica, especially if only one diode has low IR. Of the tested batch of 28 valves, 7 valves showed leakage in just one of the 2 diodes. On a few tested samples that showed leakage, the IR value increased with lower test voltage (ie. leakage resistance was not constant).

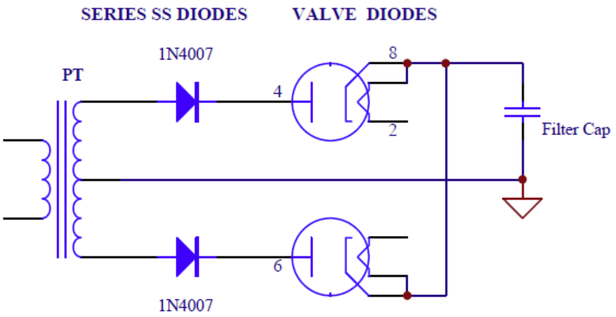
A PIV type leakage test does not appear to be standard for manufacturing acceptance, as indicated by the 6X4WA datasheet.

This IR testing example does indicate that a common 1kVdc IR meter can show up suspect valve diodes. A diode with low IR is likely to be more prone to initial arcing (ie. when PIV is max). A diode may need to show quite a low IR for it to exhibit a problem in a power supply, as indicated by PSUD2 with 10k Rectifier Leak option. When an ss diode is inserted in series with a valve diode for PIV protection, a leaky valve diode is not obvious as the ss diode leakage is low, and so an IR test may be the simplest means of maintenance checking.

7.4 Using 1N4007 in series with a valve diode anode

As indicated in Section 2.1, one or more series connected solid-state 1N4007 diodes can be fitted to each anode of a valve rectifier to suppress valve diode arcing, as shown below. The rectifier action is effectively the same during the on portion of the diode, as the valve diodes' voltage drop dominates the total voltage drop and determines the diode current waveform (as described in Section 3.1). There is no abrupt turn-off characteristic experienced due to the relatively soft turn-off characteristic of the valve diode.

When withstanding inverse voltage, the diode with lowest leakage current will support the most inverse voltage (see Figure 12 in Section 2.1). Initially that could be the valve diode, although over time a 1N4007 would end up supporting the majority of the inverse voltage. Even when a valve diode has significant leakage when inverse voltage is being supported, the series connection of a 1N4007 suppresses any leakage current to just that passed by the 1N4007.



When retrofitting a 1N4007 in with a valve diode the level of B+ voltage would only negligibly change as the 1N4007 voltage drop is likely negligible compared to the valve diode voltage drop, and reverse leakage current through the valve diode was likely not significantly loading the B+ voltage.

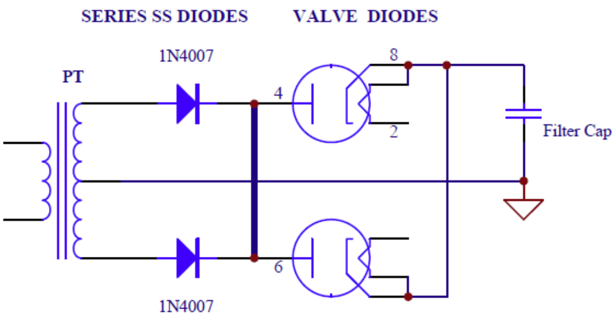
If a valve diode has started to show signs of arcing then it may be adequate to install ss diodes and still continue to use the 'faulty' valve diode with no performance degradation – this may not be appropriate if the valve diode has been damaged with respect to its on-voltage characteristic.

Bridging option

When 1N4007 diodes are fitted then an option is to bridge the valve diode anodes together as shown below. In this configuration the valve diodes operate in parallel (rather than separately), and conduct during every mains half-cycle (rather than every alternate half-cycle). This option provides an increase in B+ voltage of perhaps 10-20V depending on the valve type (if that was a benefit), and reduces the peak current level through each valve diode.

The 1N4007 diodes effectively take over the role of completely supporting inverse voltage, but don't modify the valve diode's dominant influence on the rectified current waveform and relatively soft turn-off characteristic.

An extra benefit can be the ability to use a higher value of first filter capacitance, and/or support a higher load current, as each valve diode now experiences a lowered peak current level (ie. lowered by about 50%) allowing per diode current to be increased by √2 (ie. to 140%) for the same anode dissipation level. However, subtle differences in the non-linear resistance of a conducting valve diode, and the shape of the diode current waveform, may increase diode dissipation more than anticipated so caution is required.



Bridged anodes

If the load current is not significantly increased then this option likely improves service life of the valve due to a lower operating temperature, and a lower operating peak current level.

8 Valve heater DC powering & step-up B+ converters

Powering an amp's input stage valve(s) heater by DC is typically a topic related to suppressing AC mains related hum in the audio signal, although reduction of hum relies on many aspects (see [23]). The generation of a DC supply can induce other noise and hum signals through other pathways, so caution is required to achieve a net benefit for the added complexity.

Powering heaters from DC can also relate to the practical use of an external power supply module, such as a mains powered switchmode supply 'brick' that generates a regulated DC supply such as 12V 5A. For lower powered DIY amps, such a module avoids the need to connect AC mains to the amp, which eliminates mains AC safety risks. This topic is discussed further at the end of this section.

8.1 Power transformer 5 or 6.3Vac windings for 6.3Vdc

Power transformer windings for 6.3Vac can be rectified and filtered to provide 6.3Vdc for DC powering the heaters of preamp stages for low noise operation. However, the current rating of the winding along with the loading on all transformer windings, and the available mains voltage level, may have a significant effect on the achieved DC voltage. Given a valve's heater AC voltage specification is typically +/- 10%, and diyers often aim to tweak the voltage to +/-5%, then a target of between 6.0 and 6.6Vdc is a good outcome.

Diodes with higher on-voltage (eg. UF4007) may need a higher filter capacitance to achieve 6.3Vdc. Diodes with lower on-voltage may allow a low value resistor to be inserted in series with the diode bridge to achieve 6.3Vdc, which then reduces the diode peak current requirement and the filter capacitor ripple current. Diodes with lower on-voltage may also allow more heater current loading on the DC supply. Adding additional capacitance, above what may be needed, reduces the ripple voltage on the heater and may not noticeably alter the diode current waveform. It may be practical to generate 6.3Vdc from the 5Vac winding using appropriate Schottky diodes (eg. like the 1N5819, or even 1N5818 or 1N5817).

The example results below are from a 72VA filament transformer with the 6.3Vac test winding measured at 6.9Vrms with no loading on any of the 4 heater windings, and as such these results may show higher Vdc levels than observed in an amplifier with loaded transformer windings. The identified diode was used in a full bridge configuration, and the identified capacitance was used as the only ripple filter.

Example 5Vac 3A winding powering one 12AX7 heater (300mA)

Diode type (Ipk, Tc)	Filter cap (ripple A)	Heater voltage	Note
1N5819 (1.1A, 3.8ms)	4,700uF (0.51Arms)	6.2Vdc + 0.46Vpp	0.5Ω in series
1N5819 (1.1A, 3.8ms)	2x 4,700uF (0.69Arms)	6.3Vdc + 0.22Vpp	0.5Ω in series

Example 6.3Vac 3A winding powering one 12AX7 heater (300mA)

Diode type (Ipk, Tc)	Filter cap (ripple A)	Heater voltage	Note
1N4004 (1.4A, 4.8ms)	470uF (0.46Arms)	6.3Vdc + 3.6Vpp	
UF4007 (1.3A, 4.4ms)	1,000uF (0.44Arms)	6.4Vdc + 1.6Vpp	
1N5819 (0.82A, 5.4ms)	470uF (0.32Arms)	6.3Vdc + 3.0Vpp	2Ω in series
1N5819 (0.82A, 5.4ms)	4,700uF (0.35Arms)	6.3Vdc + 0.4Vpp	3Ω in series

Example 6.3Vac 3A winding powering two 12AX7 heaters (600mA)

Diode type (Ipk, Tc)	Filter cap (ripple A)	Heater voltage	Note
1N5819 (1.5A, 5.4ms)	4,700uF (0.68Arms)	6.2Vdc + 0.72Vpp	1.5Ω in series
1N5819 (1.5A, 5.4ms)	2x 4,700uF (0.69Arms)	6.3Vdc + 0.36Vpp	1.5Ω in series

The diode should have a current rating, and sufficient conductive cooling, to handle the heater loading. Mains AC voltage waveform distortion has a noticeable effect on diode conduction waveform. The filter capacitor should have sufficient ripple current rating to exceed the expected ripple current (that may be difficult for capacitor values of 470uF or lower).

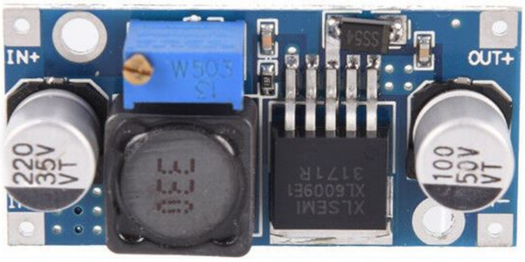
Heater related hum can ingress the audio signal via a few mechanisms [23]. The measured heater voltage Vpp level is an indication of the AC voltage waveform presented to the valve socket, compared to AC heater powering where an 18-20Vpp level would typically exist.

8.2 Power transformer 5 or 6.3Vac windings for 12.6Vdc

Power transformer windings for 5V or 6.3Vac can be rectified and filtered to provide an intermediate DC voltage that can then be converted to a regulated 12.6Vdc using a small pcb switchmode step-up (boost) module. Many preamp valves, and in particular the 12AX7, are designed to operate from 12.6V as well as 6.3V. The intermediate voltage can have substantial ripple voltage, as long as the minimum voltage level of the rectified waveform is above the under-voltage lockout level (UVLO) of the switchmode module, and the maximum voltage level of the rectified waveform is suitably below the regulated output DCV.

Two small pcb assemblies that provide a complete switchmode boost converter are the MT3608 [3] and the XTW-SY-8 (based on the XL6009 device). With a 12.6Vdc output setting, a rectified 6.3Vac winding has a peak rectified voltage (of circa 8.8V) that is sufficiently under 12.6V to allow boost duty-cycle control. Both the pcb modules allow the minimum rectified and filtered voltage to go down to below 4V, however, to avoid the controller from disabling operation due to under-voltage lockout setting, sufficient filter capacitance is needed for a particular heater loading (circa 470uF for each 150mA of heater load).

Pcb label: MT3608



PCB label: XTW-SY-8 (based on XL6009)

The default switching ripple voltage on the output terminal of each module was different in frequency and magnitude. The MT3608 exhibited <20mVpp level of 1.0MHz ripple, with 2 transient glitches each period of about 150-200mVpk. The XL6009 exhibited about 220mVpp of 400kHz ripple. If needed, those levels can be attenuated by adding a filter directly across the module output pads. For the XL6009, an LC filter using a 10uH leaded inductor (B82144A, 0.22Ω DCR; 1.4Adc; SRF~60MHz) followed by a 0.68uF leaded capacitor (63V MK10) attenuated ripple to <20mVpp. For the MT3608, an additional 10uF 16V smt capacitor soldered to the module's output pads reduced the glitch levels to about 30-40mVpp.

A suitable bridge rectifier diode for these modules is likely a 1A Schottky when loading is just 12.6V 150mA, but a higher current rated diode may be needed for higher loading.

During power turn-on, the rectified input voltage to a module increases in DC level over a few seconds as the heater resistance increases and the minimum input voltage rises up above the UVLO level. While the module's minimum input voltage dips below the UVLO level, the output voltage cycles between a constant 12.6V and a ramping down voltage (when the UVLO is active).

The MT3608 appears adequate for 12.6V loading up to about 600mA (the datasheet shows a performance plot up to 800mA). At 300mA load, the diode peak current increases to about 3.5A and pcb rear temp rise was 5^{o}C. At 450mA load with 1500uF filter, the diode peak current increases to about 5A, pcb rear temp rise

[3] The pcb assembly on eBay uses the TMI ST3508 device (with marking S35xx) which appears to be equivalent to the Aerosemi MT3608 except it operates at a slightly lower 1MHz switching frequency.

was 10°C, minimum input voltage was 4.8V, and output switching ripple increased to about 200mVpp with some mains frequency ripple of about 150mVpp related to control mode change when the input voltage waveform was near its minimum level. Adding additional input filter capacitance reduces the control mode related ripple by reducing the input voltage ripple, however the peak diode current increases.

The XL6009 module can also generate a negative bias voltage up to about 40Vdc max from a spare 5Vac heater. A single Schottky diode rectifier needs at least another 220uF 16V e-cap to augment the on-board 220uF 35V e-cap to generate about 6.1Vdc and suppress input mains frequency ripple sufficient to keep output ripple below about 30mVrms. The output ripple includes about 90mVpp 400kHz ripple spikes that can be attenuated by an LC filter (preferably an smt L part of a few hundred uH). The module may show regulation instability for certain levels of input and output voltage – one example showed instability for input between 8 to 14Vdc with no output load and output set for >28Vdc – the instability was an output saw-tooth ripple voltage of circa 100-150mVpp at circa 90-330Hz, with 400kHz ripple spikes on rising edge of saw-tooth.

8.3 Output stage cathode bias for dc heater powering

Preamp valves such as 12AX7 allow a 12.6V 150mA heater configuration. Some output stages with cathode bias can be configured to idle with a cathode resistance that presents a bias of 12.6V with 150mA common cathode current, and hence the 12AX7 heater can replace a fixed cathode bias resistor. Where two 12AX7 are required, then the output stage bias can be configured for 25Vdc at 150mA.

For a typical push-pull pair output stage, each valve would operate with a cathode current of about 75mA at idle. Alternatively, a quad could operate each valve at 38mA. The output stage anode and screen voltages, and the type of valve need to align with the available idle bias voltage and idle plate dissipation.

Known examples are these Australian PA amps:

- AWA PA872 20W amp uses a KT66 PP output stage, with two 12AX7 preamp valve heaters in series.
- AWA PA1003 25W amp uses a 7025A PP output stage, with two 12AX7 preamp valve heaters in series.
- AWA PA1005 12.5W amp uses a quad 6AQ5 PP output stage, with one 12AX7 preamp valve heater.

If needed, some additional resistor in parallel with the heater circuit, or in series with the heater circuit, can be used to modify the actual bias conditions to suit.

Care is required to avoid inadvertent signal coupling from the output stage cathode current to the input stage circuitry – for example the output stage cathodes should be decoupled with a capacitor that shunts most of the signal current with a low impedance compared to the heater resistance (eg. 25V 150mA resistance is 168Ω, which is the impedance of a 25uF cap at 40Hz). Similar to AC heater power wiring, the heater wiring should be kept away from input grids and have a small loop area. The heater wiring loop should also connect directly to the common cathode node and the output stage star 0V node, as with the decoupling capacitor, to restrain the signal current flow to just the output stage.

Some margin for an increased heater voltage arising from max continuous amp output is worthwhile, which indicates the idle bias should be somewhat below the rated heater voltage.

At power turn-on there is typically a few seconds of over-current stress on the output stage valves due to the cold resistance of the preamp heaters being much lower than their hot-state resistance.

- 150% current stress above idle level was measured on an AWA PA1003 amp.
- This condition likely won’t stress an output stage’s cathode, as cathode idle current level is typically far below the peak operational current level of the amp.
- This condition likely won’t stress an output stage’s plate, as plate temperature would initially be low and the plate current level is also likely well below the peak operational current level of the amp.
- This affect could be alleviated by adding an NTC resistor in series with the preamp heaters (which act as a PTC resistor). A 200D7 NTC inserts 200Ω into the heater circuit when cold, which reduces to about 20-30Ω when hot, and has an appropriate max current rating of 200mA which is above the

150mA target current by a suitable margin. In the PA1003, a 200D7 NTC restrained the over-current level to about 105% during power up.

Care is required to safely manage identifiable fault conditions:

- If a preamp valve is removed, or a preamp heater filament goes open, or the preamp heater connectivity goes open (eg. a poor socket contact to the valve) then the output stage cathodes stop conducting. That fault could cause an abrupt step in current and hence may cause an over-voltage surge on an output transformer primary winding.
- If an output stage fault causes fault current through the preamp heater circuit, then the preamp valve heaters may be stressed.

8.4 External DC supply and step-up B+ discussion

Perhaps the most commonly available external DC supply is a 12V regulated plugpack (aka smps 'brick') with a max current rating typically between about 1 to 8A. This form of supply can be used to generate a B+ supply rail, as discussed later in this section. Issues can arise if this path is taken, such as:

a) Can 6V valve heaters be powered from 12Vdc.

b) Is AC mains protective earth connected through the plugpack to the DC negative side.

c) Are all plugpack modules the same.

a) A 12Vdc regulated module likely supplies a bit over 12V to allow for feed cable, fusing and distribution voltage drop. If the voltage available to heaters is 12.0V (or 6.0V) then that is -5% of design centre nominal, which is not a concern for acceptable 12.6V (or 6.3V) valve operation, nor is it a risk to valve service life. However, if voltage at the socket heater terminals were to fall to -10% (11.4V or 5.7V) then that would be a concern and need attention.

   One simple method to arrange heater circuitry is to form two groups of 6.3V heaters that have the same, or similar, total group current, and connect those two groups in series. For example, a 12V 5A plugpack was used to power a Ferrograph 2A/N amplifier chassis that had no AC mains supply for heaters or B+. The 6.3V heater loads were 3x EF86, 1x 12AT7, 1x 6BQ5, and an 8V 1.6W bulb, which were grouped as 3x EF86 and 1x 12AT7 for 3x0.2+0.3=0.9A nominal, and 1x 6BQ5 and 8V 0.2A for 0.76+0.16=0.96A nominal. With the two groups connected in series, and a 12.3V supply connected, measurement showed that the 6BQ5//bulb had the lower resistance with a voltage of 5.7V, and the EF86/12AT7 group with 6.1V. Balance at 5.95V for that set of valves was obtained by loading the EF86/12AT7 group with a parallel 150Ω 1W. With this method, each group of heaters can be protected from over-voltage by placing a 6V2 5W Zener across each group, limiting any imbalance during power up, or if a valve is pulled for faultfinding. With this plugpack, the zeners had minimal 42mA idle current with all valves pulled but would experience significant heating if one group's heaters were removed (ie. 6.3x0.9=6W) so 2x Zener in parallel were used for the amplifier restoration. If convenient for grouping, valves like a 12AX7 can be configured separately for direct 12V heater operation.

b) Some plugpack models have a double-insulated rating between the mains AC primary and the secondary 12Vdc, so the 12Vdc is floating, but noise current can flow through the plugpack in a loop comprising the mains AC and other audio equipment due to stray capacitance between primary and secondary sides in the plugpack. In contrast, some plugpack models have a direct connection between the mains AC protective earth and the negative side of the 12VDC, which can similarly allow a noise earth loop to form. Either way, care is needed to manage and suppress the chance for mains related noise loops to couple into the audio signal.

c) In recent years, many smps plugpacks have become available for LED string powering, but that may not be explicitly stated by the seller, or it is ambiguous – and those plugpacks are likely to behave poorly as a 12Vdc regulated power supply for this type of application and should be avoided.

A step-up B+ supply can be configured from some cheap pcb-based switchmode inverters designed for 12Vdc input that use an isolation transformer with high voltage secondary windings, to generate convenient B+ DC levels when rectified and filtered. The inverter typically uses a FET push-pull primary, and a generic switchmode controller like 3525A or TL494 operating at 40-50kHz with a 20-25kHz secondary side ripple. Fast diodes like UF4007, or smt equivalents, are needed for secondary side rectification. The rectifier/filter circuity should have minimal current loop area using low value capacitors located on the inverter pcb (a main B+ bulk e-cap would be located with the amplifier's output stage circuitry). Even though the inverter has an unregulated output voltage, its level is defined and typically has little sag due to the regulated incoming 12Vdc from the plugpack.

The rectifier diodes could use a full-wave, full-bridge, or doubler configuration, depending on the target B+ voltage level, and the available secondary HV windings of the inverter. Also, the 12V supply level can be dropped by 1-2 volts (if needed) to lower the B+ level pro-rata (due to the fixed turns ratio transformer).

With the Ferrograph example, the original B+ was 270V, but 300-310V was the target. A recently available ebay inverter module was used, as shown in Figure 39. For this inverter, the target B+ voltage was met using a doubler across the 110V + 62V secondary taps (taps V0 to V3), and by inserting two 1N5404 diodes in series in the 12V feed to drop the inverter input voltage to 10.5V. The amplifier required about 12-15W of B+ supply, given the 6BQ5 SE class A output stage, so 12Vdc input current was a bit over 1A – indicating that any dropping diodes need to be selected and kept cool for the application.

Note that the power capability of a cheap module is likely significantly less than the advertised rating, and mainly due to the minimal or non-existent heatsinking of the switching FETs. This example '150W' module had no heatsinking, and was modified to bolt each FET to the pcb, along with thermal paste, but further heatsinking was not needed as the power output requirement was no more than 15W.

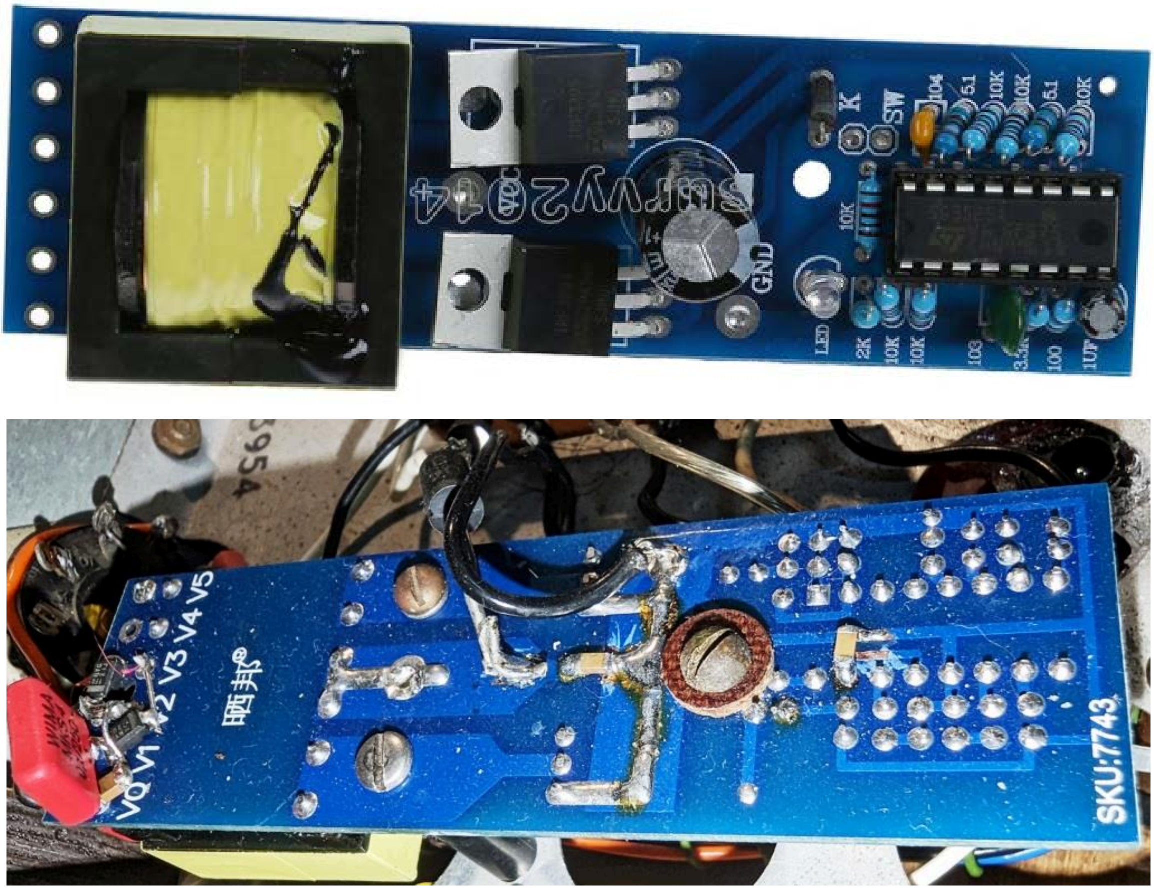


Figure 39. '150W' inverter module with added rectifier/filter.

Another type of small cheap dc-dc converter is being marketed as a 35W DC-AC Boost Inverter Module, with you-tube video's identifying it as a Royer inverter. The converter is a [Baxandall style converter](#) that Williams mis-represented as a Royer-based converter in Linear Technology's [AN65](#) and then as a Resonant Royer topology in AN118, and for brevity just used the term of Royer converter. A generic photo and schematic are shown in Figure 40.

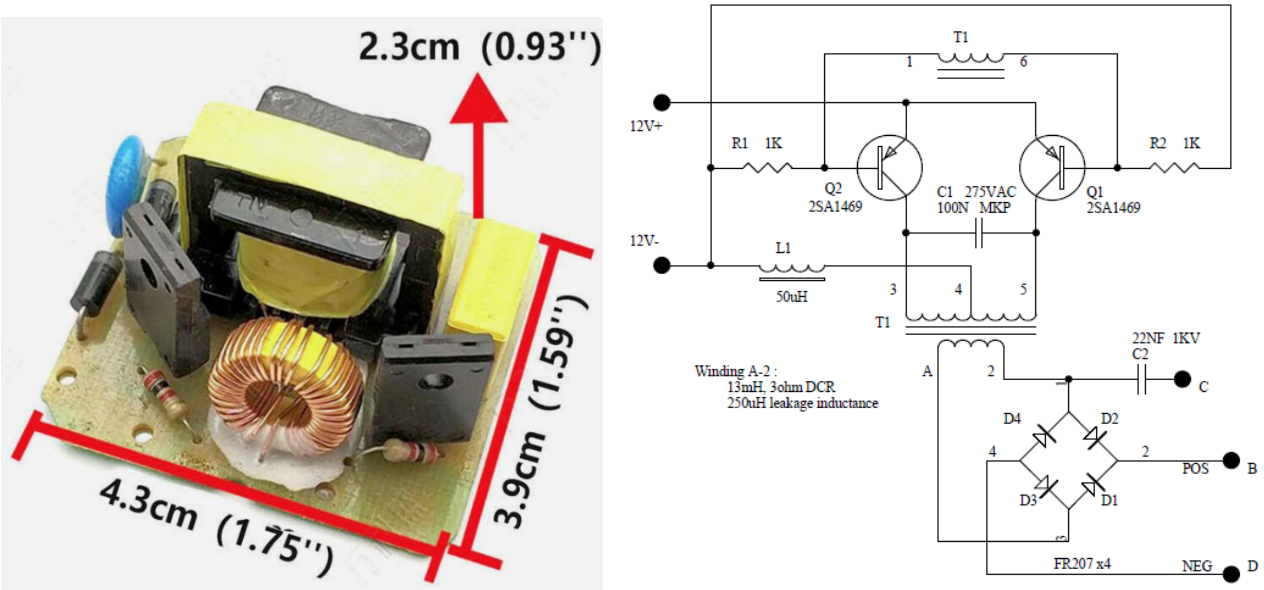


Figure 40. '35W' Baxandall inverter module with included bridge rectifier.

This converter uses a soft-switched pull-pull operation of the isolating ferrite-cored transformer primary due to self-excited near-resonant operation of the transistors Q1, Q2 at about 60kHz. As the converter uses no auxiliary control or drive circuitry, it can operate with input voltage down below 5V, and a nominal 12-15Vdc maximum (due to the 60Vceo rating of the transistors and some resonant Vce overshoot measured as 37V for 12Vin). This allows setting of the input Vdc to provide a nominal B+ level and loading, with 12Vdc generating 260Vdc 10W for the default bridge rectified configuration, although sag on B+ indicates a bleed resistor is necessary to constrain max B+ below 350-400V (no load). It's plausible to deploy a doubler rectifier/filter for a higher B+ level.

The module does not come with input or output rail capacitor filtering (C2 is meant for AC output applications), which are best added directly to the pcb to constrain ripple currents to the module. At least a 220uF 16V e-cap is recommended for the '12V' input, to cover capacitor ripple current rating above 10W. Similarly, a 4u7 400V e-cap is recommended for the output, although 2u2 may be ok (and certainly would be for a foil cap).

Initial testing to 10W output level indicates max power operation may be limited by temperature rise of the ferrite core, and caution would be needed for any continuous testing of >20W output.

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