by Richard Lloyd Measures, AG6K.

as of 7/11/98

Preface

The following manuscript was originally written under contract for the Amplifier and Projects Chapter (13) of the 1995 ARRL Handbook. About a month before the due date on the contract, I received a phone call from the ARRL Handbook Editor. He told me to stop writing.

Introduction

Amplification can be defined as: The process of increasing the magnitude of a variable quantity--especially the magnitude of a voltage and/or current, without substantially altering any other quality.

The (pure) sine wave is the only periodic waveform that contains no harmonic energy. All of the energy in a sine wave is contained in the fundamental frequency. In other words, a pure sine wave is coherent. Maintaining the quality of sine waves during amplification is a major concern in the design and operation of amplifiers.

Even though the following discussion pertains to the operation of gridded electron tubes, some of the concepts apply to power FETs because both are voltage-driven amplifying devices.

In the discourse that follows: if a voltage is said to be negative-going, that does not necessarily mean that the voltage is negative. It could be a positive voltage that is moving in the negative direction. If a voltage is positive-going, or moving in the positive direction, it could be a negative voltage or a positive voltage that is moving in the positive direction.

In an electron tube, the cathode emits a cloud of electrons. Since electrons carry a negative charge, and unlike charges attract, electrons are strongly attracted by the positive voltage that is applied to the anode. Unless something is placed in the way, much current flows between the cathode and the anode.

The term "grid" describes appearance--not function. A grid is made from a number of closely spaced wires or bars--like a bird cage. The grid is placed close to the cathode. As a result, the grid has more influence over the cathode's electrons than does the more distant anode. Thus, a small change in grid-voltage produces a large change in the flow of electrons. Because the voltage applied to the grid controls the flow of electrons from the cathode to the anode, in function, the grid acts like a valve or gate. The grid requires virtually zero current to perform its job. As a result, the power gain of a gridded electron tube is theoretically high.

Since like charges repel, a sufficiently negative grid can stop electrons from traveling to the anode. As the grid is made less-negative, the flow of electrons to the anode steadily increases. In other words, a positive-going, albeit negative polarity, grid-voltage causes an increasing flow of electrons between the cathode and the anode. The reverse is also true: a negative-going grid-voltage causes a decreasing flow of electrons from the cathode to the anode. As long as the grid remains negative with respect to the cathode, the relationship between grid-to-cathode voltage and anode-current is fairly linear.

When an appropriate load resistor is connected between the anode and its positive voltage source, the changes in anode-current produced by changes in grid-voltage create a proportional, typically much larger, voltage change across the resistor. The ratio of changes in voltage across the load resistor to changes in grid-voltage is the voltage amplification factor. It is designated by the Greek letter Mu. Since Mu is higher at high anode-voltages than it is at low anode-voltages, average Mu is a more meaningful number than maximum Mu. Average Mu ratings vary from about 2 to 240. Mu is partly determined by the spacing between the grid wires and the distance between the grid and the cathode..

Classes of Operation

The duration of anode-current conduction per cycle determines the class of amplifier operation. A conduction angle of 360 degrees means that the anode is conducting current during 100% of the input sine wave cycle. A conduction angle of 90 degrees means that the anode is conducting current during 25% of the input sine wave cycle. Long conduction angles produce a more linear representation of the input sine wave. Short conduction angles produce more efficiency--and less linearity.

Class A is defined as a conduction angle of 360 degrees. Class B is defined as a conduction angle of 180 degrees. Class C is defined as a conduction angle of less than 180 degrees. The subscript 1 indicates that no grid-current flows. The subscript 2 indicates that grid-current flows--the result of driving the grid into the positive voltage region.

When the conduction angle is less than 360 degrees, the missing part of the sine wave must somehow be filled in. One way of filling in the missing part of the sine wave is by utilizing the flywheel effect of an output tank circuit. Another way to produce a smooth sine wave is to use a push-pull configuration. If each device in a push-pull circuit conducts for at least 180 degrees, a smooth sine wave can be produced.

Class A Operation

Class A is the most linear class of amplifier operation. Class A amplifiers produce only about 1/100,000 part, or minus 50dB, distortion. The theoretical efficiency of a Class A amplifier is 50%. The practical efficiency is slightly lower. Class A is used mainly in low level amplifiers--where efficiency is not much of an issue. Since Class A operates with continuous (360 degrees) conduction, no tank circuit is needed to complete the sine wave. Class A is ideal for wide band amplification.

The zero-signal anode-current [ZSAC] in Class A is set to roughly half of the electron tube's maximum anode-current rating. Although the meter-indicated anode-current remains constant from zero signal to maximum signal, the instantaneous anode-current typically varies from just above zero to many times the meter-indicated anode-current.

The maximum available power in Class A is roughly equal to the anode-dissipation rating of the electron tube.

The Class A amplifier can be compared with a gas turbine engine. Both have a smooth, continuous power stroke--and neither one is very efficient.

Class AB1 Operation

Class AB1 amplifiers are roughly 60% efficient. The trade-off for increased efficiency is slightly more distortion--roughly 1/10,000 part, or minus 40db. Since most transceivers produce more than minus 36db of IMD, such an amplifier would not add significant distortion.

The anode-current in Class AB1 varies in proportion to the grid-voltage for about 60% of the input sine wave cycle. Thus, the anode-current is off for about 40% of each input sine wave cycle. The missing 40% in the output sine wave is filled in by the flywheel effect of the output tank circuit.

In Class A or Class AB1 a somewhat unusual relationship exists between the work being performed by the tube and the grid voltage. The grid is operated in the zero to negative voltage region. Maximum instantaneous anode-current, maximum anode voltage swing and maximum peak power output coincide with an instantaneous grid-voltage of zero--i.e., maximum stoke equals zero grid volts.

The grid must not be allowed to become positive. If the grid became positive, electrons from the cathode would begin flowing into the grid. Whenever grid-current flows, the linear relationship between grid-voltage and anode-current deteriorates.

Since there is zero grid-current in Class A or Class AB1 operation--and any voltage multiplied by zero amperes is zero watts--the driving power is usually stated as zero on the tube manufacturer's technical information sheet. However, because charging and discharging a capacitor requires current flow, in the real world of conductor RF resistance, charging/discharging the grid capacitance at an RF rate consumes some power. In a typical HF amplifier, the drive power required for Class AB1 MF/HF operation is roughly 1% to 2% of the output power. Thus, the typical power gain is roughly 50 to 100. As frequency increases, conductor resistance increases due to skin-effect. More R causes more I^2 R loss. As frequency increases, the amount of current needed to charge and discharge the grid capacitance also increases--causing even more I^2 R loss. These losses can only be compensated for by adding more drive power.

Drivers (usually a transceiver) require a resistive load--which the capacitive grid does not provide--so a suitable resistance must be connected from the grid to RF-ground..

The zero-signal anode-current [ZSAC] in a Class AB1 amplifier is normally set to about 20% of the maximum-signal single-tone anode-current.

Tubes that are designed for Class A and Class AB1 service produce high peak anode-current when the instantaneous grid-voltage is zero. Typically, the peak anode-current is about three times the maximum rated (average) anode-current. Most of the tubes that are used in Class A and Class AB1 RF amplifier service are tetrodes and pentodes--devices that have the advantage of grid-to-screen amplification. Triodes are seldom used because the only grid designs that can produce high anode-current with zero grid volts are those that have a Mu of 2 to 5. Low Mu triodes require much more driving voltage than a comparable tube with a screen requires. Since imposing a high RF voltage across the capacitive grid is difficult, low Mu triode Class A and Class AB1 power amplifiers are only practical up to a few hundred kHz.

Class AB1 Cathode-Driven Operation

The most common configuration for Class AB1 operation is grid-driven. Since grid amplification as well as grid-to-screen amplification takes place, the resulting power gain is high. Class AB1 amplifiers can also be cathode-driven if the tube is a tetrode or pentode. The grid is tied to the cathode. Thus, the grid-voltage is always 0V--so no grid-current flows. The screen is grounded. The input signal is applied to the cathode/grid. Because input signal voltage is applied between the grid and the screen, grid-to-screen amplification takes place. However, since the grid is tied to the cathode, no grid amplification takes place. Although the power gain is relatively low, linearity is excellent. The Collins 30S-1 is an example of a cathode-driven Class AB1 amplifier.

The Class AB1 amplifier is like a 2-cycle, single-cylinder engine. The power stroke is roughly half of each crankshaft revolution--it has a flywheel (the tank circuit) that supplies power between power strokes--and it is more efficient than a gas turbine (Class A).

Class AB2 Operation

Class AB2 is similar to Class AB1 except that the grid is driven into the positive voltage region during a part of the anode conduction period. A Class AB2 amplifier can be grid-driven or cathode-driven.

Grid-Driven Class AB2 and Negative-feedback

When the grid is driven positive, it attracts and accelerates the cathode's electrons. Some of the electrons stick to the grid, resulting in grid-current. Electrons that miss the grid travel to the anode. The accelerated head start causes a sharp increase in instantaneous anode-current--and a sharp decrease in linearity. The distortion products from a single-ended, Class AB2 grid-driven amplifier are roughly 1/100 part [minus 20db]. In SSB service, this level of distortion is virtually certain to cause interference to other stations using adjacent frequencies. However, by limiting grid-current and by adding an unbypassed, low-L cathode feedback resistor (to develop an out-of-phase {negative} feedback voltage) it is possible to achieve acceptable linearity in grid-driven Class AB2 operation--but only if the grid current produced is small.

An unbypassed cathode resistor is also useful for improving linearity in Class AB1. For example, the 4CX250B has a somewhat objectionable distortion level in Class AB1, SSB service. Adding a 25 Ohm resistor between the unbypassed cathode and chassis ground improves linearity. The trade-off is that slightly more grid drive voltage is needed to achieve the same output level. Cathode negative-feedback is also useful with TV sweep tubes--devices that were originally designed for switching--the opposite of linear amplification. When an appropriate cathode resistor is used with a sweep tube, reasonable linearity can be achieved.

Class AB2 Grounded-Grid Operation

Even though Class AB2 cathode-driven/grounded-grid operation produces grid-current, it is never the less fairly linear due to the laundering effect of negative-feedback. This is the result of the input and output signals being in series with each other and out of phase. Due to the negative-feedback, the distortion level in Class AB2 grounded-grid service is low--typically about 40db below PEP.

High-Mu triodes work well in Class AB2 grounded-grid operation. Medium-Mu triodes can be used, but they have less power gain. Tetrodes and pentodes usually work well in grounded-grid operation. Since tetrodes and pentodes typically have a grid-to-screen amplification factor of about 5, its easy to assume that they offer an advantage over triodes in Class AB2 grounded-grid operation. However, RF-grounding the grid and the screen stops grid-to-screen amplification. Applying DC screen-voltage does NOT increase gain because grid-to-screen amplification can not take place unless input signal voltage is applied between the grid and the screen.

The maximum available power in Class AB2 is roughly double the anode-dissipation rating.

Class B Operation

Class B is defined as a conduction angle of 180 degrees. Class B RF amplifiers produce unacceptable distortion in SSB operation.

Class C Operation

Class C is defined as an anode conduction angle of less than 180 degrees. In Class C, the amplifying device is deliberately not operated linearly. Instead, it is operated as a switch in order to reduce resistance loss. The anode conduction angle in Class C operation is usually made as short as is possible. In effect, the tank circuit makes the RF output sine wave--like a bell that is struck at a constant rate by a hammer. This is similar to the principal behind the spark transmitter.

The efficiency of a typical Class C amplifier is high. When compared to a Class AB1 or Class AB2 amplifier operating at the same power input, a Class C amplifier will deliver a received signal increase of about 1db--in other words, 1/6 of 1 S-unit. However, significant trade-offs are required to achieve that 1/6 of 1 S-unit. As is the case with Class B operation, the distortion from Class C operation is so high that SSB operation is precluded. Only CW, FM or FSK operation is practical. The harmonic output level from a Class C amplifier is substantial. Extra filtering is usually needed to control harmonic radiation.

The maximum available power in Class C operation is roughly three to four times the anode-dissipation rating of the electron tube.

Class D Operation

Class D is used up to about 1.6MHz--mostly in AM broadcast service. In Class D operation, the amplifying device rapidly switches on and off at a fixed rate--like a switching power supply--except that the output voltage varies at an RF rate. The amplitude of the RF is controlled by varying the on period of the switch. Smoothing is accomplished by a complicated filter that converts some of the odd-harmonic energy from the rectangular waves back to fundamental frequency energy. Class D is highly efficient--but it is limited in frequency capability and frequency agility.

Amplifier Design Considerations

Arriving at an amplifier design that will give years of surprise-free service involves many considerations. Merely copying circuits from published amplifier designs or commercial amplifiers is not necessarily the best approach. Doing so may result in copying someone's mistakes. The best approach is to learn what you can about each section of an amplifier, discuss it with others--then reach your own conclusions.

Basic prerequisites for getting a handle on what's going on inside an amplifier are an understanding of Ohm's Law, inductive and capacitive reactance, impedance, resonance, how gridded electron tubes function, and some knowledge of L and pi networks.

A useful book on amplifier design is Eimac®'s ***Care and Feeding of Power Grid Tubes***. I will minimize the discussion of topics that are covered adequately in this book.

Tubes vs. FETs

When the first RF power FETs were introduced, it was commonly thought that FETs would eventually replace bipolar transistors and gridded electron tubes in HF power amplifiers. Since RF power FETs work better at 50V than they do at 12V, FETs have not replaced bipolar transistors in 12V mobile applications. Another difficulty with FETs is cost. A pair of FETs that can produce 1200W PEP at 29MHz cost about six times more than an electron tube, or tubes, that can do the same job. The FETs' input power requirements are 50V at 50A, i.e., 2500W--so there's considerable heat to dispose of. Meeting the cooling requirement is not nearly as easy as it is with tubes because tubes operate quite happily at surface temperatures that destroy silicon devices.

In low power applications at room-temperature, solid-state devices can last 100 years. However, at the junction temperatures encountered in high power applications, the P and N doped layers slowly diffuse into each other--thereby steadily eroding the device's amplifying ability. A relatively-large rigorous cooling system is needed to achieve a reasonable operating life from high power solid-state RF amplifying devices.

Another difficulty with solid-state high power RF amplifiers is their power supply requirements. Tubes are quite tolerant of moderate variations in their anode supply voltage. Transistors, however, are fatally sensitive to over-voltage. It is much easier to build a 3000V, 0.8A unregulated supply for a tube than to build a regulated 50V, 50A power supply with over-voltage, over-current and over-temperature protection circuitry for high-power solid-state devices.

The bottom-line is that 1500W, HF, gridded electron tube amplifiers are more efficient, more forgiving, easier to cool, more compact, weigh less, are more tolerant of high SWR and are less costly than 1500W HF semiconductor amplifiers. For instance, a pair of legal-limit FETs cost about $800 from Motorola®. The efficiency is about 10% less that what one can achieve with gridded electron tubes.

Grounded-Grid Versus Grid-Driven

For at least the last three decades, the vast majority of amateur radio amplifier designs have been Class AB2 cathode-driven--a.k.a. 'grounded-grid'. One reason for this is simplicity--or at least the appearance of simplicity. Ground the grid(s), drive the cathode. Only three supplies are needed--the T-R switching supply, the filament supply and the anode supply. Neutralization is theoretically not needed because the grounded grid(s) shield the output element, the anode, from the input element, the cathode. This theory works **almost** perfectly. Grounded-grid amplifiers are virtually always stable at the operating frequency because the reactance of the feedback C is too high at HF to allow regeneration. This is fortunate because there is no way of neutralizing a single ended grounded-grid amplifier. Another advantage is flexibility. Almost any tetrode, pentode, or high-Mu triode from the junk box will work. Linearity is usually good and the typical power gain--10db to 14db--is acceptable. So far, so good. Now for the trade-offs.

What goes on inside a grounded-grid amplifier is not as simple as it looks. The AC component of the anode-current and the grid-current, i.e., the RF cathode current, passes entirely through the cathode coupling capacitor and the tuned-input circuit--so the input circuit is in series with (and out of phase with) the output circuit. The components in the tuned circuit must be able to handle a substantial amount of RF current. Manufacturers of tubes that are designed for grounded-grid operation typically recommend using a tuned input pi-network with a Q of 2 to 5. To maintain an acceptable SWR and Q when the operating frequency changes appreciably, all three reactances in the tuned input must change proportionally. However, if Q is allowed to change, L can be left as is providing that C1 and C2 are retuned. [For more information on this problem, see the section titled "Tuned Input Circuits."]

Even though HF grounded-grid amplifiers are stable at their operating frequency, at VHF the grid looses its ability to shield the input from the output. HF grounded-grid amplifiers have a less-than-pristine reputation for VHF stability.

For wide frequency coverage, the Class AB1 grid-driven amplifier requires a much simpler tuned input than a grounded-grid amplifier requires. Typically, grid-driven amplifiers have more power gain than grounded-grid amplifiers. One Class AB1 grid-driven amplifier has about as much gain as two Class AB2 grounded-grid amplifiers in series. The trade-off is two additional DC supplies--a grid bias supply and a screen supply. Both of these supplies need to be adjustable. HV power FETs make this task easy.

Cathodes

There are two types of cathodes--directly-heated and indirectly-heated. In a directly-heated cathode, a ditungsten carbide layer on the hot (c.1800 degrees K) tungsten, alloyed with about 1.5% thorium--a.k.a. 'thoriated-tungsten', filament wire emits electrons. In an indirectly-heated cathode, the filament (a.k.a. heater) heats a metal cylinder that is coated with strontium oxide and barium oxide. This coating is relatively frangible--but highly emissive.

Ditungsten carbide is commonly formed by heating tungsten in an atmosphere of acetylene (C_2H_2) gas. Carbon atoms in the gas break their electron bonds with hydrogen atoms and bond with tungsten atoms to form ditungsten carbide on the surface of the filament wire. Since it is atomically linked to the underlying tungsten, the ditungsten carbide layer is very durable. During use, the process reverses. Ditungsten carbide gradually looses carbon and changes back to tungsten. Extra heat exponentially accelerates this process. A cathode is worn out when the carbon is mostly used up.

After their cathodes grow tired of emitting electrons, large external-anode amplifier tubes are commonly "recarburized" with acetylene, vacuum-pumped and resealed. This restores full emission. Although it is possible to recarburize a 3-500Z, doing so is not economically feasible. The smallest tube that is currently being recarburized is the 3CX1000A7.

Each type of cathode has advantages and disadvantages. Indirectly-heated cylinder [8877] and planar [3CX100A5] cathodes have much less inductance than a directly-heated cathode made from wires. Thus, indirectly-heated cathodes are more frequency-capable. Some indirectly-heated cathode tubes can perform satisfactorily at 2500MHz. The 3CX100A5 is an example.

Directly-heated/thoriated-tungsten cathodes are more resistant to damage from electrons that bounce off the anode. It's possible to use up to 22kV with the larger thoriated-tungsten cathode tubes. Electrons that have been accelerated by such voltages move at very high velocities. When they strike the anode, they produce X-rays.

A thoriated-tungsten cathode typically warms up in one second, while few indirectly-heated cathodes can warm up safely in one minute--and three to five minutes is not uncommon. For HF operation, indirectly-heated cathode tubes have a much higher cost to watt ratio than thoriated-tungsten cathode tubes. For VHF and especially for UHF operation, indirectly-heated cathode tubes are often the only choice. For super-power HF operation, thoriated-tungsten cathode tubes are the only choice.

Cathodes deserve respect. Filament-voltage and filament inrush current are the prime areas for concern.

Filament / Heater Considerations

For optimum life from a thoriated-tungsten cathode, the filament-voltage should be just above the voltage where PEP output begins to decrease. As a thoriated-tungsten cathode ages, filament-voltage needs to be increased incrementally to restore full PEP. By using this technique, commercial broadcasters typically achieve an operating life of more than 20,000 hours from thoriated-tungsten cathode tubes.

According to Eimac®'s *Care and Feeding of Power Grid Tubes*, "every 3% rise in thoriated-tungsten cathode filament-voltage results in a 50% decrease in life due to carbon loss." Each additional 3% rise in filament-voltage decreases the life by half. Thus, cathode life is proportional to [E1/E2]^23.4 where E1 is the lowest filament-voltage at which normal PEP output is realized--and E2 is the increased filament-voltage. However, for heater-type oxide cathodes, if the heater potential is allowed to fall below the specified level, the emissive material may flake off of the cathode, and cause a cathode-grid short. On the other hand, excessive heater potential causes barium migration to the grid - which results in primary grid emission.

Controlling Filament-Voltage

It's simple to make the filament voltage adjustable when the filament is powered by its own transformer. All that's needed is a small rheostat in series with the primary. For dual voltage, dual primary transformers, a dual ganged rheostat is required. However, when the filament is powered by a winding on the HV transformer, making the filament-voltage adjustable is more difficult since a dual ganged, very low resistance, high current rheostat must be connected to the low-voltage high-current secondary winding. The typical value needed for a pair of 3-500Zs would be (2) adjustable 0.01 Ohm @ 30A--most definitely not a common rheostat. A reasonable substitute can be made from a double pole, c.10 position, 30A rotary switch and short lengths of resistance wire or ribbon bridging the fixed contacts.

An indirectly-heated cathode can be ruined by operating it below the rated minimum filament-voltage. When operated above its maximum filament-voltage rating, an indirectly-heated cathode boils off emissive material (principally barium) onto the grid and other parts. This results in decreased cathode life and undesirable grid-emission when the grid warms up during transmit. This condition is indicated when the output power steadily drops off in AØ (max. signal, key-down, a.k.a. NØN) operation. The decrease in power normally begins within two seconds.

For maximum cathode life in HF communications service, an indirectly-heated cathode should be operated at the rated minimum filament-voltage. This can be accomplished best with a regulated DC supply.

Controlling Filament-Voltage During Receive

In a typical amateur radio amplifier, the filament-voltage rises about 5% during receive due to decreased load on the electric-mains. Not only is a 5% increase in filament-voltage during receive useless, it uses up thoriated-tungsten cathode emission life about three times as fast as during transmit. However, if an appropriate resistor is placed in series with the filament transformer primary, and shorted out by a relay during transmit, the filament-voltage will not change appreciably between transmit and receive. An appropriate type of relay for shorting the resistor on transmit is a 'power' reed-relay.

Filament Inrush Current

Thoriated-tungsten filaments commonly consist of two vertical intermeshing helices (coils) of tungsten wire that are suspended by their ends. (see Sept. 1990 QST, p.15) The conductance of tungsten at room temperature is about 8.33 times the conductance at the normal operating temperature. Thus, the start-up current for a 15-ampere filament can exceed 100 amperes. Needless to say, 100 amperes makes for a dandy electromagnet.

In a high amplification triode such as the 3-500Z, the filament helices clear the grid cage by a matter of thousandths of an inch. If the position of the filament changes, a grid-to-filament short may result. Therefore, it is prudent to limit filament inrush current in order to minimize thermal and magnetic stress.

Since the grid-to-cathode clearance in an indirectly-heated cathode is not affected by movement of the heater inside the rigid cathode cylinder, indirectly-heated cathodes are not affected by inrush current.

For many of its smaller thoriated-tungsten cathode amplifier tubes--such as the 3-400Z and 3-500Z--Eimac® recommends that filament inrush current be limited to no more than double the normal current. This rating is easily exceeded unless a special current-limiting filament transformer or a step-start circuit is used.

END PART 1

Grid Protection in G-G Service

Transmitting AØ for long periods on 10m with the load capacitor set for maximum C would result in very high grid-current and almost no RF output. Under such a condition it might be possible to overheat a grid. However, since most people tune an amplifier for maximum output--and maximum output virtually coincides with normal grid-current--very few people are likely to overheat a grid. Thus, complex electronic grid-protection circuits are seemingly unnecessary.

A disadvantage of electronic grid-protection circuits is that they are not effective against the most common source of grid damage--intermittent VHF parasitic oscillation.

Glitch Protection

During a major problem, the anode (plate) current meter and other amplifier components can be subjected to a large current surge as the HV filter capacitors discharge. The peak discharge current can exceed 1000a if a series resistor is not used to limit the short circuit current that can be delivered by the HV filter capacitors. The current limiting resistor is placed in series with the positive output lead from the filter capacitors. A wire wound resistor with a high length to diameter ratio works best. A 10 ohm, 10W wire wound resistor is adequate for up to about 3kV & 1A. For higher voltages, additional 10 ohm, 10W resistors can be added in series to share the voltage drop during a glitch. Wire wound resistors with a high length-to-diameter ratio are best for this type of service. Since about 1985, Eimac® has recommended the use of a glitch protection resistor in the anode supply circuit. Svetlana® typically recommends using a 10 to 25 ohm glitch resistor.

A HV current limiting/glitch resistor may disintegrate during a major glitch--so it should be given a wide berth with plenty of chassis clearance. If the chassis clearance is minimal, its a good idea to cover the chassis with electrical insulating tape. Glass-coated (a.k.a. vitreous) wire wound resistors are the most suitable type of resistor for this application. If a glass-coated resistor comes apart during a major glitch, it won't be throwing chunks of shrapnel around--like a less-expensive rectangular ceramic-cased resistor often does. Metal-case power resistors should not be used in this application. If a glass-coated glitch resistor is damaged during a glitch, it should be replaced with two such resistors in series to reduce the peak V-gradient per unit of length during a problem.

If the positive HV arcs to chassis ground--due to lint, a hapless insect, a VHF parasitic oscillation, or moisture--the negative HV circuit will try to spike to several kilovolts negative in the typical 1500W amplifier. In the real world, this type of glitch is not an uncommon occurrence. Anything that gets in the way of the negative spike may be damaged. Since the grid-current meter is normally connected between chassis ground and the negative HV circuit, the meter can be exposed to kilovolts at hundreds of amperes.

The easiest way to protect a current meter is to connect a silicon rectifier diode across it, or across its shunt resistor. Usually, only one diode {cathode band to meter negative} is needed in parallel with a DC meter. In some circuits, it is best to use two diodes in parallel [anode to cathode] with the meter movement to protect against positive and negative surges.

It may take more than one diode to protect a meter shunt resistor. A silicon diode begins to conduct at a forward voltage of about 0.5V. To avoid affecting meter accuracy, the operating voltage per glitch protection diode should not exceed 0.5V. For example, a 1 ohm shunt, at a reading of 1A full-scale, has 1V across it. Thus, two protection diodes in series would be needed to preserve meter accuracy. Similarly, if the shunt resistor for a 1A full-scale meter is 1.5 ohm, the maximum shunt voltage is 1.5V--so three diodes are needed.

Glitch protection diodes should not be petite. Big, ugly diodes with a peak current rating of 200a or more are best. Smaller diodes--and the meter they were supposed to be protecting--can be destroyed during a glitch. Suitable glitch protection diodes are 1N5400 (50PIV) to 1N5408 (1000PIV). In this application, PIV is not important. The 1N5400 family of diodes is rated at 200a for 8.3mS.

During an extremely high current surge, a glitch protection diode may short out--and by so doing protect the precious parts. Replacing a shorted protection diode instead of a kaput meter is almost fun.

To reduce the chance of the negative HV circuit spiking to several kilovolts, connect a string of glitch protection diodes from the negative terminal on the HV filter capacitor to chassis. At 200a, each diode will limit the surge voltage across it to about 1.5v. Typically, three diodes are needed--thusly limiting the negative spike to about 4.5 volts. Diode polarity is: cathode band toward the negative HV. With one simple wiring change, the same string of diodes can also protect the grid I meter and the anode I meter. This dual protection technique is incorporated into the Adjustable Electronic Cathode Bias Switch on Figure 7.

Design Considerations for Indirectly-Heated Cathode Tubes

A HV arc can destroy an indirectly-heated cathode tube. Here's how it happens: In some amplifiers, one side of the filament/heater is grounded. The cathode is connected to the negative HV circuit. If the negative HV spikes to several kilovolts, the cathode will often arc to the grounded heater. At a minimum, this breaks down the insulation between the heater and the cathode. Sometimes the heater wire burns out--and sometimes the cathode arcs to the grounded grid. Either way the tube is kaput.

Grounding one side of the heater is an invitation for cathode-to-filament breakdown. Instead, let both heater wires float. If the heater is fed through an c.40micro H bifilar RFC, one side of the heater can be wired to the cathode. Even though this arrangement can not protect against cathode-to-grid breakdown, it assures that the voltage between the filament and the cathode is unlikely to rise to dangerous levels.

Safety Devices

HV-Shorting

Manufactured amplifiers typically use a safety device to automatically short the +HV supply to chassis ground when the output section cover is removed. If the cover is removed before the HV filter capacitors have discharged, the resulting positive HV to ground short can damage the amplifier. In most g-g amplifiers, the only DC current path between the negative HV circuit and chassis ground is the grid-current meter and its shunt resistor. Even if the remaining charge in the HV filter capacitors is only 200V when the short from positive to ground occurs, without glitch protection diodes, the entire 200V appears across the grid-current meter shunt and the grid-current meter. Many potentially-fatal amperes can flow into the grid meter as the HV filter capacitors finish discharging. If the amplifier is accidentally switched on with the cover removed, rectifier diodes are a common casualty.

Automatic-shorting safety devices are not only dangerous to amplifier components, they can be dangerous to operators. It is dangerous to assume that an amplifier is safe to work on because it contains a safety device. Even though an amplifier's HV supply is shorted, if the amplifier is plugged in, its electric-mains circuitry is still alive and potentially fatal. Amplifiers are inherently dangerous. They should not be worked on casually--even if they have so-called safety devices.

The safest quick method of discharging HV filter capacitors is through a paralleled pair of wire wound resistors. The resistors limit the discharge current to a safe amount. In the unlikely event that one resistor opens, the remaining resistor will do the job. For the average 1500W amplifier, a paralleled pair of non-adjustable 1k ohm to 5k ohm, 50W resistors will do the job. **Its always a good idea to check the anode supply voltmeter before putting your hands inside an amplifier.**

Fuses

Fuses have current and voltage ratings. For a fuse, the real test is opening safely--not operating without opening during normal operation. A fuse's maximum voltage rating is important. In some amplifiers, ordinary 250V 3AG fuses are casually used in circuits where they may be required to interrupt several kilovolts. Examples are the anode or cathode circuit. All's well until a problem occurs. When a 250V fuse attempts to interrupt a potentially-damaging flow of current in such circuits, the frangible link inside the fuse parts as it should. However, due to the available voltage, when the link melts, a metal vapour arc forms in its place. Metal vapour arcs typically have a voltage drop of around 20V--so the unsafe current will continue to flow in the circuit. At some point, the fuse will eventually explode--usually after serious damage has been done to other components.

During a glitch, circuits which normally carry low voltages can spike to several kilovolts. For example, cathode circuits normally see a maximum voltage of 30V to 100V. Thus, it might seem appropriate to use a 250V fuse to protect the cathode from excessive current. However, when a glitch occurs, several kilovolts can appear across anything that attempts to interrupt the flow of cathode current. The safest place to use ordinary 250V fuses is in the primaries of transformers.

Power Supplies

Ripple Filters

There are basically two types of DC filters: inductor-input / capacitor-output, and capacitor. Each type of filter has advantages and trade-offs.

Capacitor filters have good transient response. Since no inductor and resonating capacitor are used, the capacitor filter is simple to build, compact, cost-effective, requires no tuning and it is lightweight. The main disadvantage of a capacitor filter is that the capacitor is charged only during a small fraction of the waveform supplied by the transformer. No charging current flows until the instantaneous output voltage from the rectifiers exceeds the instantaneous voltage on the filter capacitor. This means that the transformer is either not loaded or severely loaded at different times during each cycle.

For example, with a electric-mains frequency of 60Hz, the duration of a half cycle is 8.333mS. Under load, using a capacitor filter, the capacitor charging time per half-cycle is typically only about 1mS out of the 8.333mS. This means that the ratio between output current and peak charging current can be 8 to 1. To combat I^2 R loss in a capacitor filter power supply, the transformer, all circuitry in the primary (including the electric-mains) and the filter capacitor should have low resistance. Capacitor filters are not appropriate for use with older-design transformers that were intended for use with inductor filters. Typically, such transformers have high winding resistance.

Inductor-input / capacitor filters can be of the resonated type or the non-resonated type. A non-resonated inductor tries to maintain a constant DC-current despite changes in the load current. This is the nature of any inductor. It always tries to maintain constant current by temporarily increasing the output voltage when the load current decreases suddenly, or by decreasing the output voltage when the load current increases suddenly.

When a conventional voltmeter is used to monitor the output voltage from a non-resonated inductor / capacitor filter power supply, the transient unregulation characteristic will usually not be detected because of the damped response in the meter movement. If a DC oscilloscope is used to monitor the output voltage while a string of caround 5 WPM CW dashes are sent, the instantaneous output voltage swings can be easily observed. On make, the output voltage spikes downward. On break, the output voltage spikes upward. The amplitude and width of the spike depends on how much filter capacitance is used after the inductor and on the change in current. Upward and downward voltage spikes of more than ±50% are possible during a sudden load change on a non-resonated inductor / capacitor filter power supply.

- Transient unregulation is probably not an important consideration unless SSB operation is used. With SSB, the PEP output and the linearity of the amplifier would be adversely affected by a non-resonant inductor DC filter.

The resonant DC filter maintains a fairly constant output voltage during rapid or slow changes in current demand--provided that a minimum current passes through the inductor. This minimum current can be the zero-signal anode-current [a.k.a. 'idling current'] of the tube itself. The inductor is resonated with a parallel capacitor. In actual practice, the value of capacitance used--as well as the bleeder resistance--is that which produces satisfactory voltage regulation. The resulting resonant frequency is usually slightly higher than double the frequency of the electric-mains. Resonant L/C pairs are available from Peter W. Dahl, Inc.

DC filter inductors come in two types, fixed-inductance and swinging-inductance. A swinging-inductor changes its inductance according to the current that is passing through it. Obviously, a swinging inductor can not stay tuned correctly with changes in current. Therefore, resonated-inductor filters can **only** use a **fixed** inductor.

The disadvantages of a resonant inductor filter are:

- The resonating capacitor must have a DC-working voltage rating of about three times the DC output voltage of the supply Typical values are 0.1 to 0.15 micro F @7.5kV to 15kV.
- To maintain voltage regulation during standby, a minimum 'bleeder' current must flow in the inductor. Typically, the bleeder current is 10%. Considerable heat is dissipated by the bleeder resistor(s). However, if the filter capacitor can withstand the approximately. 50% increase in voltage during receive, the 10% standby bleeder current requirement can be reduced to 0.5%.
- The inductor is heavy and costly.
- A resonant inductor filter usually makes an audible noise--unless the inductor has been potted in plastic resin.

The advantages of a resonant-inductor DC filter are:

- Excellent voltage regulation.
- Greatly reduced peak current demand on the transformer and the electric-mains. Most importantly, this reduces transformer heating.
- Transformers have about double the output current capability when a resonant inductor filter is used instead of a capacitor filter. However, the output V from a capacitor filter is higher.

The resonant filter is used extensively by commercial and military amplifier manufacturers. Since a resonant filter demands much less peak power from the electric-mains than a capacitor filter demands, for 120V operation, where available power is typically much more limited than with 240V operation, a resonant filter is clearly the best choice. The resonant filter is also the best choice for high duty-cycle modes such as RTTY, FM or AM.

Rectifier Circuits

- Half wave. ----**Advantages**: may be used where one side of the AC input is grounded. **Disadvantages**: requires highest filter C; causes DC current to flow in the transformer; poor voltage regulation; transient-voltage protection is needed to protect the rectifier from reverse voltage spikes.
- **Fullwave-centertap----** The fullwave-centertap rectifier circuit was used in ancient times when

tube-type rectifiers were the only game in town. Only one rectifier filament-winding was needed to produce full wave rectification, so the center-tapped secondary winding was the norm in older transformers. If a more efficient fullwave-bridge rectifier circuit had been used, three rectifier filament windings would have been needed. **Advantage**: if needed, reduces output voltage to one-half or to approximately. one-fourth of the DC voltage that would be obtained with a full wave bridge or full wave voltage-doubler. **Disadvantage**: inefficiently utilizes only half of the secondary winding at any instant--resulting in less than optimum transformer efficiency.

- **Full wave bridge---- Advantages**: full utilization of the transformer's capability; may be used with a resonant filter. **Disadvantage**: requires twice as many transformer secondary turns as the full wave voltage-doubler requires. This means more layers of insulating paper--and that takes up winding space--so smaller wire must be used.
- **Full wave voltage doubler---- Advantages**: full utilization of the transformer's capability; to achieve a specific DC voltage, only half as many transformer secondary turns are needed compared to a fullwave-bridge circuit. This means that the power transformer secondary will have a higher ratio of copper to paper. If switched secondary taps are used to control the DC output-voltage, the voltage stress on the switch is only half of what it would be with full wave bridge rectification. Full wave doubler supplies typically have a remarkably low ripple content. This is because one half of the filter capacitance is being charged at the same time the other half is being discharged. Since the charging sawtooth waveform is similar and opposite to the discharging sawtooth waveform, the result is a fairly smooth DC output. **Disadvantages**: To achieve acceptable voltage regulation, the full wave doubler requires twice as much filter capacitance as a fullwave-bridge. This is the case because each of the two filter capacitor sections is charged once per cycle versus being charged twice per cycle with the fullwave-bridge circuit. Thus, in a full wave doubler, the filter-capacitors must be able to hold their charge twice as long--so twice as much filter capacitance is needed. The capacitance requirement is easily met with modern aluminum electrolytic capacitors. They provide a large amount of capacitance in a small space at a reasonable cost. Full wave voltage doublers are not practical for use with a resonant inductor filter

Transformers

Transformers are available in two basic types: E-I (conventional) core and toroidal core. The E-I core is made from a stack of thin E-shaped and I-shaped iron plates. When placed together they form a rectangle with two windows for the windings. A stack of E-I rectangles make the completed core. The toroidal core is made from a continuous tape of grain-oriented material that contains iron and silicon plus other elements that increase the permeability of the core and decrease loss. This core material is known as Hipersil. Westinghouse Corp. was the original patent and copyright holder. Their patent expired decades ago. There are different grades of Hipersil tape. Grade 5 has the highest performance. Grade 22 has the lowest performance.

Higher permeability means that fewer turns are needed to achieve the required inductance in each winding. This means that larger diameter wire can be used. The end result is a transformer with low resistance and high efficiency. The Hipersil core is so efficient that the principal loss factor is the resistive loss in the copper wire. Hipersil core transformers are capable of producing extremely high peak currents. Thus, the Hipersil core transformer is ideally suited for capacitor filter power supplies.

It is difficult and time-consuming to thread a continuous tape core through the completed transformer windings--so someone came up with a faster way of uniting the core with the windings. Here's how it's done: The tape is wound on a form of the appropriate dimensions. The tape is spot welded together, removed from the form, and annealed at about 700 degrees C to relieve internal stresses. After cooling, the core is varnished and dried. Then the core is cut in half with a machine that makes a precise square cut. The faces of the cut are then polished flat. Thus, the halves of the core can slip into the completed windings, contact each other closely--restoring nearly perfect magnetic coupling between the halves of the core. The matched halves of the core are marked so that they can not be inadvertently mixed up with other core halves. The reunited halves of the core are held together tightly by steel bands like those used for binding heavy cartons and crates. If future access to the primary and secondary windings is needed, a Hipersil transformer can be disassembled by cutting the steel bands and removing the core halves.

Hipersil is no longer the most efficient type of core material. The new amorphous core transformer is starting to come into use by electric utilities. An amorphous core transformer is so efficient that if the secondary is unloaded and the primary is disconnected from the electric-mains, the collapsing magnetic field generates a voltage spike that can destroy the transformer. To avoid this problem, one winding is paralleled with a suitable voltage surge absorber.

Transformer Power Ratings

Transformers are commonly rated in maximum "volt-amperes" [VA]. Maximum VA are roughly equal to maximum RMS watts when the rated RMS current is flowing in each transformer winding and the transformer is operated from the rated input voltage at the design frequency. If the electric-mains voltage is reduced, the VA capability of the transformer decreases.

For SSB and CW operation, a lighter transformer may do the job just as well as a much heavier and more costly transformer. Manufactured 1500W amplifiers typically use a HV transformer with a continuous capability of roughly 600W--or VA. Such transformers are completely satisfactory for normal SSB operation. Such transformers are also capable of handling brief FM and RTTY transmissions--provided that the lower voltage tap is used.

If a power supply's DC output voltage drops more than about 10% under modulation, it's a fairly safe assumption that a more capable transformer is needed. Of course, not using enough filter capacitance or excessive electric-mains resistance can also cause poor regulation.

Transformer Current Ratings

Increasing the current in any conductor causes a square-law increase in the amount of power dissipated in the conductor. Since P=I^2 x R, doubling the current causes a 4 times increase in dissipation. This is an especially important consideration with transformers because they have considerable difficulty dissipating the heat that is generated deep inside their windings. This problem is compounded because copper has a positive, resistance versus temperature, coefficient. Thus, as the copper heats up, its resistance increases--which increases the dissipation--which increases the resistance, et cetera. This can lead to thermal runaway and transformer failure.

If a transformer has a secondary rating of 1A RMS, it means 1A with a resistive load. If connected to a rectifier and DC filter, the 1A rating does not necessarily apply. For example, if a fullwave-bridge rectifier, resonant filter circuit is used, the RMS current rating can be multiplied by at least 1.2. The DC output voltage will be about 0.85 times the RMS voltage. If a fullwave-bridge rectifier, capacitor-filter circuit is used, the loaded DC output voltage will be about 1.3 times the RMS voltage. A 30% increase in voltage sounds good, but obviously you don't get something for nothing. The trade-off for the increase in voltage is a decrease in current capability. The high peak current demanded by the capacitor filter translates into a substantive current capability decrease. .

Any formula for converting a transformer's RMS current rating to a DC output current rating is bound to be problematic due to the large number of variables. Here's a rule-of-thumb that is fairly accurate. If, after about an hour of typical operation, the outside of the transformer is uncomfortably hot for one's thumb, the internal parts of the transformer are probably deteriorating. Reducing the average load current slightly will greatly reduce transformer heating because of the square-law relationship between current and power dissipation. For example, reducing the current by 30% will reduce winding dissipation by about 50%.

There is a simple, reasonably accurate, 2-step approximation for determining the safe SSB, maximum current rating for a specific transformer for use with a capacitor filter and a full wave bridge rectifier. A slightly different approximation is used for a full wave voltage-doubler. These approximations are based on the DC resistance and the AC-voltage of the transformer's secondary winding. These approximations are useful when shopping around surplus stores or swap meets. All that's needed is an ohm-meter and a clip-lead. The clip-lead is used to short the primary of the transformer. This dampens the inductive voltage spike that occurs when the ohm-meter is disconnected.

The fullwave-bridge, capacitor filter approximations are: Multiply the secondary winding resistance by 70 to find the minimum intermittent load resistance that can be placed on the power supply. To find the DC output voltage under load, multiply the secondary RMS-voltage by 1.3. To find the safe intermittent current rating for SSB service, use Ohm's law and divide the output voltage by the minimum load resistance. For a more accurate evaluation, use the appropriate graphs in this book.

For example, a 2000V RMS secondary winding has a DC-resistance of 60 ohms. A full wave bridge rectifier, capacitor filter, circuit will be used. The safe, minimum, intermittent load resistance is approximately 70 x 60 ohm = 4200 ohms. The approximate voltage delivered under load would be 1.3 x 2000V = 2600V DC. Thus, the maximum intermittent load current is 2600V ÷ 4200 ohms = 0.62a.

Another approximation can be used to find the amount of filter capacitance needed. The approximation is 50,000 divided by the minimum load resistance. In the above example this is 50,000 ÷ 4200 = 12micro F.

For a full wave voltage-doubler, capacitor filter, power supply, the SSB-service approximations are: Minimum intermittent DC-load resistance equals 300 times the winding resistance; DC-output voltage, under load, equals 2.5 times the secondary RMS voltage.

For example: A 1000VRMS transformer has a winding resistance of 10 ohms, the minimum load resistance for full wave doubler operation would be 300 x 10 ohms=3000 ohms and the output voltage would be 2.5 x 1000V=2500V. The maximum intermittent load current is 2500V ÷ 3000 ohms = 0.83A.

The amount of filter capacitance needed for each half of the full wave voltage-doubler circuit is approximately 200,000 divided by the minimum load resistance. In the above example each of the two capacitors should have a minimum of 200,000 ÷ 3000 = 67micro F.

There is more to transformer performance than secondary resistance. If a Hipersil® core is used, core loss is minimal and the maximum intermittent power capability increases. Primary resistance is another factor to consider since it is effectively in series with the electric-mains resistance. Electric-mains resistance can cause a voltage drop problem if the amplifier is a fair distance from the service entrance box and you are using a capacitor filter power supply. One solution is to use larger diameter wires than the electric code requires. Another solution is to install the power supply near the service box and bring the HV DC to the amplifier.

END OF PART 2

Changing Voltage

It's nice to have the ability to reduce the output power from an amplifier. One way to do this is to reduce the anode-voltage and anode-current simultaneously so that the output load-R of the amplifier tube or tubes does not change appreciably. This allows the tank circuit to function at its design Q for both high and low power.

Switching primary taps is not an efficient method of reducing output voltage because in order to do so extra turns must added to the primary. To make room for the extra turns, the primary's wire diameter must be decreased--and that increases R. An efficient method of reducing the DC output voltage in a HV power supply is by switching secondary taps on the transformer. If a fairly ordinary ceramic rotary switch is insulated from the chassis, it can easily perform this job. The taps should not be switched under load.

If no secondary tap is provided on a transformer, it is possible to lower the output voltage 50% by switching from fullwave doubler to fullwave bridge rectification. All that's needed is a suitable SPST vacuum relay, or well-insulated ceramic switch, two filter capacitors and four strings of rectifiers. For example, a power supply that produces 4000V for SSB could be operated at 2000V for RTTY, CW, or FM. The DC output current capability doubles when the output voltage is halved during fullwave bridge operation--just what's needed for FM's and RTTY's much higher duty-cycle. When switching the voltage output, it is best to temporarily switch the power supply off and then restart. The output voltage may be switched down without switching the supply off--provided that the amplifier is in standby.

Variable Electric-Mains Transformers

On paper, the variable auto-transformer, a.k.a. Variac® or Powerstat®, looks good. Variacs/Powerstats are intended to be used with resistive loads. When a Variac is set at or near 100% of the input voltage, it adds only a small amount of series R. However, when a Variac is set to produce a fraction of the input voltage it adds more series R. This is of little consequence with resistive loads. However, when the load is a capacitor filter DC supply, due the demand for high peak current, additional series R is most unwelcome. Although Variacs perform acceptably with resonant choke filter power supplies, using a Variac to control the output voltage from a capacitor filter supply is not good engineering practice.

A Variac can be used in place of a step-start relay. Provided that the operator always remembers to set the Variac to near-zero before switching the amplifier on, all will be well. A step-start relay offers some advantages: it is cheaper, mistake-proof, saves many kilograms, and it adds substantially less series R.

There is, however, an appropriate step-start application for a Variac. Eimac® recommends using a motor-driven Variac, feeding the filament transformer primary, to bring up and bring down the filament-voltage (over a period of two minutes each) on its tubes which incorporate water-cooled filament supports. An example is the 8973 tetrode--just what you need if you are building a 600kW linear amplifier.

Transformer Insulation

Most transformers use paper to separate and insulate each layer of windings. Paper is hygroscopic--i.e., it absorbs water vapour from the air. The presence of water reduces the insulating ability of the paper. In time, insulation breakdown is likely. The solution is to pot the windings. Plastic resins are best. Petroleum tar is next best. Since potting fills up the air spaces in the windings--and air is a poor heat conductor--potting also improves heat transfer--thereby reducing internal temperature and increasing MTBF. Potting adds very little to the initial cost of a transformer and subtracts substantially from the long-term cost. Some custom transformer manufacturers offer potting as an extra-cost option. Peter Dahl Co. has a potting option.

Transformer Potting

Commercial transformer potting is normally done in a vacuum chamber to facilitate the evacuation of air bubbles. However, with a little patience, it is possible to pot transformers satisfactorily without special equipment. Bake the transformer in an oven at a temperature of about 175 degrees F/80 degrees C. Bake for two to three hours per pound. Baking drives out internal moisture. After baking, place the transformer on a table covered with a thick layer of newspapers. Position the transformer so that the leads or lugs are down. Using masking tape, seal the end of the transformer windings opposite the leads/ lugs so that liquid can not escape easily when the transformer is inverted.

Polyester fiberglass laminating resin is designed to flow into small spaces and expel air bubbles. It can be used for potting transformers.

In a clean tin-can, pour in a quantity of laminating resin that will fill up the air spaces in the bottom 5% of the transformer's windings. Using pliers, bend the rim to facilitate pouring the resin. Mix in about 5 drops of catalyst per ounce of resin. Depending on the ambient temperature and humidity, this amount of catalyst will result in a moderately fast gel time. Pour the resin slowly into the windings. Resin pouring should be done steadily and from only one area of the windings to avoid trapping air bubbles. Any leaks from the bottom can be patched by forcing raw silicone rubber into the area of the leak. When the resin gels, it forms a thin bottom plug. The bottom plug need not be more than about 5mm thick.

Pour an amount of resin into the can that will fill the remainder of the windings. For a several-kVA transformer, use about 1.5 drops of catalyst per ounce of resin. For smaller transformers, slightly more catalyst is needed. The resin must not gel before the air bubbles have had a chance to escape--so it is better to err on the light side for the amount of catalyst.

Heat increases the fluidity of the resin--hastening the exit of bubbles. However, heat tends to decrease gel-time. Internal transformer heating is accomplished by forcing current through the windings with a Variac. Connect the Variac to the highest voltage winding. Short the highest current winding with an AC ampmeter. Increase the voltage until the ampmeter indicates the rated winding current. At this level fairly normal internal heating results. As soon as the resin begins to gel, stop the current and direct a cooling fan at the transformer. Resin-gelling is an exothermic reaction.

Rectifiers

The most frequent failure mechanism for HV power supply rectifiers is too much reverse current. This problem can be virtually eliminated in 50Hz/60Hz, fullwave bridge and fullwave doubler, capacitor filter circuits if the total PIV in each string of diodes exceeds the no-load DC output voltage by at least 50%. For operation in high-temperature environments, a 100% factor may be needed.

Modern solid-state rectifiers are made differently than they were 30 years ago. In those ancient times, same-type rectifiers did not have uniform reverse characteristics. Rectifier failure was common. In an attempt to compensate for the inherent weaknesses in early solid-state devices, rectifier protection schemes were used. Resistors and capacitors were paralleled with series rectifiers--probably a take-off on the practice of using equalizer resistors on electrolytic filter capacitors. However, in any series circuit, the currents in all of the elements are exactly equal. Thus, when rectifiers are in series, the reverse current burden is exactly the same for each rectifier--provided that no parallel resistors are used. Manufacturers of series rectifier units long ago abandoned the practice of using parallel resistors and capacitors. The 1995/6/7 Radio Amateur's Handbook explains why rectifier 'equalization' is prone to cause premature rectifier failure.[page 11-9, middle column, top]

Series-connected rectifiers should be of the same type. Mixing rectifiers types in the same series string could cause a problem during the reverse half-cycle.

When a rectifier has been conducting, it takes a finite amount of recovery time for the rectifier to stop conducting after the source of forward current reaches zero. It is important that a rectifier not be conducting when the reverse voltage arrives. This can be a problem when rectifying high frequency AC or when rectifying square waves. Paralleling a capacitor with each rectifier may help the rectifiers to stop conducting sooner. If you need to rectify high frequency AC, one solution is to use fast-recovery epitaxial rectifiers. 1000PIV, 1A, 70 nanosecond recovery time units are currently priced at about 50¢ each in quantities of 100.

Packaged HV Rectifiers

Rectifiers that have a rating above 1kV PIV are typically made from a series of individual rectifiers that are entombed in an epoxy package. This arrangement makes for a neat-appearing installation--but there is a trade-off. Epoxy is a poor conductor of heat. Individual series-connected diodes mounted on perfboard and exposed to open air dispose of heat much more efficiently than do multi-diode packages.

Filter Capacitors

Filter capacitors usually have a ripple-current rating. The ripple-current rating should be at least equal to the maximum DC output current capability of the supply. Quality filter capacitors are designed to minimize equivalent series resistance [ESR]. Low ESR ohms translates into a high ripple-current rating.

Oil-filled capacitors are available in two types: filter service, for use in power supplies, and flash service for use in photography or pulsed laser applications. The flash capacitor is designed for maximum capacitance per unit volume. To reduce volume, very thin metal foil is used to make the plates of a flash capacitor. Thin plates have more ESR--so they dissipate more I-squared times R power when they are subjected to ripple-current.

For longest life in high duty-cycle applications, cool air should be allowed to circulate freely around filter capacitors.

According to some manufacturers, flash capacitors can be used in filter service if they are operated at 60% of their rated peak volts. In intermittent duty applications, it may be possible to use flash capacitors at more than 60% of their rated peak-voltage. To discover how a flash-capacitor is faring in ripple-current service, after about an hour of contesting, if the capacitor is warm to the touch, an internal heating problem is indicated. Internal heating causes expansion and stresses the capacitor's case--which may eventually come apart at the seams and begin to leak dielectric oil.

If not plainly stated on its label, there is a way of determining the intended type of service for an oil-filled capacitor. Flash-capacitors usually have a peak voltage [PV or VP] rating. Filter-service capacitors are usually rated in DC working volts [DCWV]. Capacitors can also be rated in AC working volts. To convert AC-working volts to DC-working volts, multiply the AC voltage by three.

There have been instances where surplus flash capacitors were offered for sale with altered or counterfeit markings. For example, a capacitor that was originally marked "3.5kVP" was changed to "5kV" by erasing characters. Thus, a capacitor that should have been de-rated to 0.6 x 3.5kV = 2100V for filter service would appear to be good for 5kV. A practical way of determining whether an oil-filled capacitor can withstand ripple-current is to connect it in series with an AC-ampere meter and an AC voltage source. The voltage is adjusted until the AC current is equal to the expected maximum output current of the power supply. If, after an hour, the capacitor shows little or no internal heating, you have a winner.

There is also a flash service type of aluminum electrolytic capacitor, that is not designed to handle ripple-current.

Electrolytic filter capacitors are intolerant of reverse current and heat. Electrolytic capacitor working voltage [WV] ratings should be treated with respect. The WV rating is virtually the maximum voltage rating. Despite their more delicate nature, electrolytic filter capacitors offer substantial advantages over oil-filled filter capacitors. The main advantages are more joules of energy storage per dollar, reduced weight and reduced volume.

When electrolytic capacitors are operated in series, they should share the voltage equally. In order to do this, a voltage equalizer resistor is connected across each capacitor. Equalizing resistors must have fairly equal resistance--and their resistance should not change appreciably during aging. If an equalizer resistor changes value appreciably, domino-effect destruction of an entire section of filter capacitors may result.

There is no formula for determining the optimum resistance for an equalizer resistor. Less resistance equals less bleed-down time. However, less resistance produces more heat. A compromise is in order.

Carbon-composition resistors change resistance with age. This characteristic is unacceptable for equalizer resistor service. High resistance, wire wound resistors are wound with extremely fine resistance wire. They are not remarkably reliable. Metal oxide film [MOF] resistors are more reliable. The initial resistance of a MOF resistor is typically much closer to the labeled value--and it will stay that way for many years. A Matsushita/Panasonic® 3W, 100k ohm MOF resistor makes a good equalizer resistor for 450V capacitors. It produces a reasonable bleed-down time and a reasonable amount of heat. These resistors are available from Digi-Key.

Electrolytic filter capacitors are ruined quickly by reverse current. Reverse current often occurs when a rectifier fails. To protect electrolytic capacitors from reverse current, connect a >600PIV diode across each capacitor. The cathode band of the diode connects to the capacitor's positive terminal.

Biasing

When a grounded-grid amplifier's operating bias is obtained from a single Zener diode, there is no way to compensate for tube variation. One solution is to obtain the operating bias from a series string of forward biased rectifier diodes. By switching the number of diodes in and out with a rotary switch, the bias can be changed in approximately 0.7V increments.

Traditionally, a mechanical relay has been used to switch amplifier bias between receive and transmit. An optoisolator coupled to a transistor switch, i.e., an electronic bias switch, can do this job faster, more reliably, sans-noise, and cheaper.

There are principally two means of actuating electronic bias switches--RF-actuation and coil-current actuation. Although it sounds hip, RF-actuation creates two problems. The amplifier rapidly switches between linear bias and non-linear bias during softly-spoken syllables of speech. This causes choppy-sounding audio and splatter. When the electronic bias switch is controlled by the current that passes through the RF relays' coils, it is not possible to intermittently switch the amplifier into non-linear bias during transmit. Coil current actuated bias switching can be accomplished with an optoisolator. The optoisolator's input LED is driven by the coil-current. The output of the optoisolator drives the bias switch transistor.

Class AB1 Grid Bias

Since the grid draws virtually zero current, it is easy to make the bias continuously adjustable in Class AB1 operation. Typically, the cutoff bias voltage during receive will be about 50% higher than the transmit bias voltage. An optoisolator driving a HV FET can be used to switch the bias between transmit and receive. A circuit is provided.

High Speed RF Switching

A conventional relay switches in roughly 25mS. Such relays have traditionally been used for RF and bias switching in RF amplifiers. This was acceptable when transceivers also used conventional relays. Currently manufactured transceivers are designed for AMTOR, QSK telegraphy, and unobtrusive SSB VOX operation. Modern transceivers T/R and R/T switch quietly, and do so in as little as 5mS. Such radios typically use a high-power SPDT reed relay to switch the antenna between transmit and receive. Similar relays can be used for amplifier input RF switching. Jennings and Kilovac manufacture high speed, SPDT vacuum-relays that have a continuous rating of 7A at 32MHz [2450W into 50 ohms]. The Jennings relay is the RJ-1A. Kilovac's relay is the HC-1. When used with a speedup-circuit, either relay can switch in under 2mS. Although both manufacturers make DPDT RF vacuum-relays, none are as speedy as their fastest single pole models. Thus, separate input and output relays are usually faster than a single DPDT relay.

PIN Diode Switching

Another device that can be used for high-speed RF switching is the PIN [P-Intrinsic-N] diode. PIN diodes are similar to 1000PIV rectifier diodes--i.e., they have a wide intrinsic region. PIN diodes are utilized extensively in radars as transmit-receive switches.

A PIN diode is switched off by applying DC reverse voltage to widen its intrinsic region. The PIN diode is switched on by passing DC current in the forward direction to fill its intrinsic region with current carriers. PIN diodes are extremely fast switches. Their lifetime is virtually unlimited as long as the allowable PIV is not exceeded.

The typical reverse breakdown voltage rating for a PIN diode is around 1000V. A legal-limit amateur radio amplifier produces an output voltage of about 800 volts peak-to-peak [p-p] into a 50 ohm load--so a 1000PIV PIN diode is more than adequate. When the load Z is higher, due to a somewhat less than wonderful SWR, the switching device may be exposed to more than 1000Vp-p. This poses no problem for a typical high-speed vacuum-relay. Even if a vacuum-relay's breakdown voltage is temporarily exceeded, there is little likelihood that permanent damage to the vacuum-relay's contacts will result. However, solid-state devices are not so forgiving. A single voltage-transient can destroy a PIN diode.

For 100 WPM computer-CW, the PIN diode is clearly the only choice. For 30 WPM CW, AMTOR and high-speed VOX, a vacuum-relay has advantages.

Solid-state Component Ratings

Different types of solid-state components are rated somewhat differently. Some ratings are realistic. Some ratings are not realistic. The maximum ratings of large transistors and large Zener diodes can not be realized unless drastic, extreme measures are used to keep the case temperature below the maximum allowable 25 degrees C at full ratings. In the real world, operation at 30% of a published dissipation or current rating is usually safe. Additionally, bipolar power transistors suffer from a generic weakness called secondary-breakdown phenomenon. For example, a "1500V, 8A, 150W" power transistor may only be able to safely dissipate 15W at moderate collector-to-emitter voltages. T-MOS power FETs are much more resistant to secondary-breakdown.

Wire-lead rectifier current ratings are fairly realistic when they are mounted on perfboard and cooling air is allowed to circulate freely around individual rectifiers.

Measuring PIV

There is some variation in the inverse breakdown voltage of solid-state rectifiers of the same type. Measuring the breakdown voltage of each rectifier diode is a good precaution. When a diode's reverse current reaches 1 to 2 uA, the voltage across the diode is the breakdown voltage. Exceeding this voltage is likely to be fatal. As operating temperature increases, breakdown voltage decreases.

Breakdown Voltage Testers

A breakdown voltage tester (a.k.a. high-pot) could be described as a variable HV Ohmmeter that does not read directly in ohms. It is a useful tool. Breakdown testers are essential for testing vacuum relays, vacuum capacitors, blocking capacitors, air-variable capacitors, rectifiers, and for finding problems with insulation. Building or troubleshooting a tube-type RF amplifier without a breakdown tester is like crossing an ocean without a navigation instrument. For most amateur radio applications, the highest voltage component rating commonly encountered [with vacuum-capacitors and vacuum-relays] is 15kVDC/9kV RF peak, so a 0 to 15kV breakdown tester should suffice.

Although commercial breakdown testers are available, they are not inexpensive. A suitable breakdown tester can easily be constructed from mostly-surplus parts. The main parts are a 50/60Hz low-current HV transformer, a >1A, 120V variable transformer, a 120V incandescent bulb, some diodes, resistors, a sensitive uA meter and two HV filter capacitors.

Commercial, low-current HV DC supplies may also be used provided that they are connected to a Variac in series with a 120V incandescent light bulb to limit current. The bulb limits the short-circuit current to a safe value--obviating the need for a fuse. The wattage of the bulb is roughly proportional to the wattage of the supply. The rated [I=P/E] bulb current should be similar to the appropriate fuse rating for the HV supply primary. A multi megohm resistor is used to limit the current flow into the device under test. The uA meter should be protected with back to back 1A diodes. A circuit is provided.

Step-Start

Most power supplies benefit from something to soften the shock of start-up. A 10A DPST-NO or 10A DPDT relay and two approx..25 ohm 10W resistors are just about all that's needed to add a step-start circuit to the average 1500W amplifier. The step-start circuit goes in series with the mains fuses or circuit breakers. With this arrangement the filaments, the HV supply and the LV supplies enjoy the benefit of a kinder and gentler start-up.

VHF Stability

Every HF amplifier has at least two resonant circuits in its output circuitry. The more obvious one is the HF-resonant tank circuit. A less obvious one is the VHF-resonant circuit that is principally formed by the anode capacitance and the inductance of the conductors between the tank circuit and the anode. In 1500W amplifiers, anode resonance typically occurs around 100MHz--well within manufacturers' ratings for "Amplifier and Oscillator Service" for the tubes that are commonly used in such amplifiers.

The equivalent resistance of a high Q parallel resonant circuit is virtually infinite. A low Q parallel resonant circuit has a relatively low equivalent resistance.

The voltage gain of an amplifier tube is roughly proportional to the load resistance. High load resistance produces more gain. Low load resistance produces less gain.

If the conductors in the anode resonant circuit have a high VHF Q, the equivalent load resistance presented to the anode will be high and the tube will exhibit increased voltage gain at the VHF resonance. If the conductors in the anode resonant circuit have a low VHF Q, the load resistance presented to the anode will be low and the voltage gain at the VHF anode resonant frequency will be reduced. Of course, if no VHF energy were present, it would make no difference how much VHF gain an HF amplifier had.

When a transient current passes through a resonant circuit, the resonant circuit rings like a bell--producing a damped sine wave signal. This is how ancient spark transmitters produce RF--and the larger ones produced many kilowatts of it.

Whenever the anode-current in an HF amplifier changes, a small VHF damped sine wave signal is produced in the anode's VHF resonant circuit. This signal can be observed with a VHF oscilloscope or a spectrum analyzer. The amplitude of the RF voltage produced is proportional to the Q of the anode resonant circuit. If none of this damped wave signal were fed back to the input, there would be no problem.

In a grounded-grid amplifier, the grid appears to shield the input from the output. In a grid-driven Class AB1 amplifier, The RF-grounded screen appears to shield the input from the output. However, no grid and no screen is perfect--so some of the damped wave VHF signal at the anode is capacitively fed back to the input--and amplified.

Although it's unlikely, if the phase and amplitude of the damped wave signal happens to be just right, oscillation at the anode's VHF resonance can occur. If the VHF energy that is produced could find its way to a load, no danger would be posed by a VHF oscillation. However, the HF tank circuit is a low pass filter that effectively blocks VHF energy. Thus, the oscillator is unloaded and the resulting grid-current is very high. The unloaded condition can cause VHF voltage transients in the anode circuit. These transients may cause tune-capacitor arcing and band switch arcing across open contacts. Since they are closest to the anode resonant circuit, open tune-capacitor padder contacts, as well as open 10m contacts are most vulnerable to parasitic-instigated arcing. Band switch contacts can be melted and/or vapourized by such occurrences.

Parasitic Suppression

On page 72, the 1926 Edition of ***The Radio Amateur's Handbook*** tells us how to build an improved VHF parasitic suppressor The logic was elementary. A suppressor is supposed to dampen the anode circuit. Since low Q is synonymous with high dampening, why not decrease Q by using resistance-wire? Quoting from page 72:........ *"The combination of both resistance and inductance is very effective in limiting parasitic oscillations to a negligible value of current."*

After 1929, someone forgot to include this information in the Handbook. In those days, the oversight probably didn't matter very much. Large amplifier tubes generally had low VHF amplification, so VHF instability was not a major issue. During the ensuing decades, people got into the habit of using parasitic suppressors made from copper or silver-plated copper. This was an easy habit to get into since copper and silver can be soldered more easily and cheaply than nickel/chromium [nichrome] resistance wire. Meanwhile, the performance of amplifier tubes kept improving. Because of these improvements, modern high-amplification tubes appear to benefit more from 1926-vintage low VHF-Q parasitic suppressors than 1926-vintage tubes. NOTE: In 1926, a 'high-mu' triode had a mu of around 40.

Anti-parasitic Techniques

Low VHF-Q conductor material can be used to increase VHF loading in the anode resonant circuit. Nickel-chromium-iron alloys are best, Nickel-chromium (nichrome) and some types of stainless-steel are almost as good. The use of copper, aluminum, and silver should be kept to a minimum. However, good conductors are desirable beyond the tune capacitor, which marks the end of the anode VHF resonant circuit and the beginning of the HF tank circuit.

Output Z: The output impedance of most tubes is a matter of kilo-ohms--not ohms. There is no scientific reason to use "heavy duty" conductors between the anode and the tune capacitor. If good VHF stability is a design goal, it's best to use conductors that are no larger than is necessary to carry the highest current present, i.e., the 10m RF circulating current between the anode (output) capacitance and the tune capacitor. Round conductors have a lower VHF Q than flat conductors. To increase current handling ability, or to reduce inductance, two paralleled round conductors, separated by a wide air-gap, are better than a flat conductor of the same overall width.

- When designing the layout for an HF amplifier, locate the tune capacitor fairly close to the anode. This reduces the inductance in the anode resonant circuit--increasing the VHF resonant frequency. If the distance from the anode to the tune capacitor is grossly excessive, the 3/4 wave anode resonance may cause instability problems--especially with tubes that are UHF-rated like the 8877, 8874, and 3CX800A7.
- In order to prevent a conductor from sharing the anode VHF-resonant circuit with the HF tank circuit, connect the tank inductor directly to the tune capacitor. It is best not to connect the tank inductor directly to the blocking capacitor.
- The output enclosure in an HF amplifier is a high Q VHF-resonant cavity. The output enclosure can become a player in parasitic oscillations. Cavity dampening may be needed. This is a

common practice in commercial high-power amplifiers. Closed loops of resistance-wire can be used to dampen cavity resonances.

- In some cases, the HV-RFC contains a VHF resonance that abets parasitic oscillations. This problem is indicated by several burned turns appearing on the HV-RFC after a glitch. To isolate a problematic HV-RFC VHF resonance, place at least one VHF attenuator-rated ferrite bead on the lead between the top of the HV-RFC and the anode circuit.

Designing VHF Parasitic Suppressors

The simplest type of parasitic suppressor is a resistor. It reduces Q by adding R. This technique is effective. However, it is mostly limited to low power applications. The traditional staggered-resonance parasitic suppressor provides two advantages over a resistor suppressor--it can handle more current, and it causes the VHF resonance to work against itself.

A staggered-resonance parasitic suppressor typically consists of a coil inductor paralleled with a low-inductance resistor. The axis of the coil inductor is parallel to the resistor. Here's how it works: The magnetic field from current flowing in the resistor is at a right angle to the direction of current flow. The magnetic field from the inductor is parallel to the direction of current flow. Because the two magnetic fields are 90 degrees apart, the inductances act independently instead of mutually. The two independent-inductances connect to one fixed-capacitance--i.e., the anode. Since the coil has more inductance than the resistor, it creates a second VHF resonance that is slightly lower in frequency than that produced by the resistor. The conflicting resonances work against each other. This technique broadbands the anode's VHF resonance, i.e., it reduces the Q--very much like stagger-tuning IF transformers to widen a receiver's band pass. Reducing VHF Q lowers the parallel equivalent VHF load resistance on the anode. This reduces the VHF voltage gain--and that reduces an amplifier's ability to oscillate at VHF.

Choosing the optimum amount of inductance for a suppressor inductor [Ls] is best determined experimentally by operating the amplifier on 10m. Since 10m is almost VHF, a device that suppresses VHF energy should get hot from 10m RF. If Ls is too small, the suppressor resistor [Rs] will not exhibit visible signs of heating during operation on 10m. If Ls is too large, there will be too much voltage drop on 10m and Rs will burn out.

Every straight conductor has inductance. The amount of inductance is fairly proportional to length. Manufactured high-power "non-inductive" resistors are long--and therefore somewhat inductive. They are too inductive for use in VHF suppressors. However, it's easy to make a suppressor resistor of sufficiently low inductance by paralleling straight nichrome wires that are separated by air-gaps.

Staggered-resonance suppressors can be built without a resistor by paralleling two unequal-inductance nichrome wires. For example, a silver plated strap in the anode circuit can be changed from a potential source of grief into a Q-reducing asset by replacing it with two, parallel, nichrome wire conductors. One of the conductors is made about 25% longer than is necessary to span the distance. Its length is shortened by winding a small 1 to 2 turn coil. The axis of the coil is parallel to the shorter wire. This arrangement decouples the magnetic field in the coil from the magnetic field in the straight conductor.

For large amplifiers, 100% nichrome staggered-resonance suppressors solve the problem of not being able to find a high-power resistor of sufficiently low inductance for use in parasitic suppression service. For very large HF amplifiers, all-nichrome staggered-resonance suppressors should be made from flat nichrome conductors in order to carry the large RF circulating current between the tune capacitor and the anode capacitance.

In two-tube amplifiers, if the two suppressors are allowed to magnetically couple to each other, a VHF parasitic oscillation may occur. In a two-tube amplifier, the suppressor coils should be positioned at a right angle. If the suppressor coils are parallel to each other, the coils should be wound in opposite directions, and separated as much as is practical.

Evaluating VHF Parasitic Suppressors

Some amateur radio operators--and some electronic engineers--do not believe that VHF oscillations can take place in an HF amplifier. This is understandable because the most common and most destructive type of VHF-parasitic-oscillation, the push-push variety, lasts only a matter of microseconds. Push-pull VHF parasitic oscillation is obviously only possible in multi-tube amplifiers. A steady oscillation between anodes is the result. Push-pull parasitic oscillation is characterized by extremely high anode dissipation, moderate grid and anode currents with zero drive, and no arcing. Push-pull parasitic oscillations can be stopped by switching the amplifier to standby. This is not possible with a push-push parasitic oscillation---wherein the event is probably over before the sound of the parasitic arc reaches the operator's ears.

VHF parasitic oscillations are not cooperative. It may take a particular sequence of anode-current transients to initiate a parasitic oscillation. Even though there is no concrete scientific evidence to prove it, the phrase "CQ contest" may have a propensity to produce the key sequence of anode-current transients--especially if the contest is one you've been waiting for, and the local radio parts emporium has just closed for the weekend.

A major factor with parasitics is the VHF-gain of the particular amplifier-tube or tubes that happen to be installed in the amplifier. Even among new amplifier-tubes from the same production lot, there is some variation in VHF gain. Tubes with below average VHF gain may never have a parasitic-oscillation--no matter how poorly their parasitic-suppressors perform. Thus, when below average gain tubes happen to be installed in an amplifier, it's easy to assume that the amplifier design is perfectly stable.

Since catching a parasitic oscillation in the act is virtually impossible with ham-type test gear, a different analytical approach must be used. It's a reasonable assumption that a resonant circuit which supports parasitic oscillation can be found and evaluated with a dipmeter.

To determine the parasitic frequency and evaluate the parasitic suppressor, unplug the HF amplifier from the electric-mains and measure the anode-resonance with a dipmeter. The best place to do this is on either side of the HV blocking capacitor. The resonant frequency typically varies inversely with the amplifier's power capability. 700W single-tube amplifiers typically resonate from 100MHz to 150MHz. 1500W amplifiers typically resonate from 80MHz to 140MHz. 100kW amplifiers typically resonate from 35MHz to 45MHz. You should be able to tune the resonance a few MHz by adjusting the tune capacitor. The VHF dip in some amplifiers' anode-circuits is so sharp that it will "suck-out" the oscillator in the dipmeter. If that is the case, the dipmeter must be backed away (decoupled) from the conductor to accurately observe the dip. A broad, smooth dip is good. A sharp dip indicates that the anode-circuit has a high VHF-Q. This is not good news unless you happen to need a VHF oscillator.

If the suppressor design is changed to (hopefully) lower the VHF Q, check the dip again. The frequency will usually not change appreciably but the dip should now be smoother, more broad, and it should be necessary to couple the dipmeter coil closer to the anode-circuit to achieve the same degree of dip.

If you make an experimental change to a parasitic suppressor, and you want to evaluate the change more precisely, use a plastic ruler to measure the distance from the anode circuit to the tip of the dipmeter coil that will result in a 20% dip on the dipmeter. If the coupling distance required for the 20% dip decreases after a change to a suppressor, a lower VHF-Q is indicated and the change was obviously an improvement. If the dip distance increases, the VHF-Q went up and the change in the suppressor was a step backward.

Amplifiers that exhibit a tuning vagary in the area of resonance on one or two bands are probably in need of better parasitic suppression. A stable amplifier usually exhibits smooth, symmetrical tuning.

Additional information on parasitic oscillation was published in the September and October 1990 issues of *QST*.

END OF PART 3

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ALC

Today, the typical transceiver output is 100W to 200W. There are amplifier tubes that can be destroyed by 100W of drive. A good example is the 3CX800A7. Driving a 3CX800A7 with 100W PEP will eventually strip flakes off of the cathode. The flakes lodge between the cathode and the grid cage--creating a fatal short. Even a pair of 3CX800A7s are clearly over-driven by 100W. The fix is: connect a approx. 40 ohm cathode negative-feedback resistor in series with each 3CX800A7 cathode. As a result, the 3CX800A7s won't be driven above their maximum ratings--and into non-linearity--by a 100W transceiver. Naturally, when cathode negative-feedback resistors are added, the cathode driving impedance increases. The driving impedance for a pair of 3CX800A7s is about 25 ohms. With 40 ohm cathode resistors, the driving impedance is roughly 50 ohms.

Cathode negative-feedback resistors are better than having a matched pair of 3CX800A7s. The cathode currents automatically equalize themselves---and unlike ALC circuits, cathode feedback resistors work instantaneously--eliminating ALC's generic flaw--leading edge splatter on SSB. Amplifier-to-transceiver ALC works properly only on constant signal level modes such as RTTY and FM.

The 3-500Z is rated at approx. 60 watts drive. When a single 3-500Z is driven by 100W, it "flat-tops" and produces distortion. A 25 ohm low-L cathode feedback resistor will make a 3-500Z linear with 100W of drive. The resistor is placed in series with the cathode RF coupling capacitor.

Capacitors and RF Current

Capacitors that carry RF current are subject to two types of internal heating. Like ripple filter capacitors, the ESR in the capacitor's conductors generates heat that is proportional to I^2 x R. Due to skin-effect, R goes up with frequency. Another source of heat is RF dielectric loss. Since dielectric loss usually varies with frequency, the current carrying ability of capacitors changes with frequency. Typically, transmitting capacitors are current rated at three widely-spaced frequencies. It's a good idea to check the manufacturer's current ratings before using a transmitting capacitor in a specific application. Just because a capacitor is a transmitting-type does not mean that it will work reliably in all RF applications.

Tuned Circuit Q

Most of the tuned input and tuned output circuits in HF amplifiers are pi-networks. There are a number of ways to define the Q of a pi-network. In what follows, Q is defined as the input impedance of the pi-network divided by the reactance of the input, shunt element--typically a capacitor. This definition of Q is the one used by Eimac® in *Care and Feeding of Power Grid Tubes*.

Tuned Input Circuits for Class AB2 Cathode-Driven [grounded-grid]

Even though grounded-grid amplifier circuits look simple, they are not. The grounded-grid amplifier's tuned input circuit is in series with and out of phase with the anode current pulses. The RF cathode current's approx. half sine wave pulses are the sum of the anode and grid currents. Since the driver is connected to the other end of the tuned input, some of the RF cathode current finds its way back to the driver. Consequently the driver interacts with the amplifier. The Q of the amplifier's tuned input affects this interaction.

Modern solid-state output MF/HF transceivers use a broadband push-pull RF output stage. In order to meet FCC requirements, Butterworth and/or Chebyshev pass band filters are used to suppress spurious emissions. Such filters introduce inductive reactance or capacitive reactance within their pass bands. In other words, the output impedance of a modern transceiver is seldom 50 ±j0 ohms. When driving a tuned input in a grounded-grid amplifier, filter reactance interacts with the input reactance in the tuned input. The length of the coax between the driver and the tuned input affects the interaction.

When tube manufacturers state the cathode driving impedance in grounded-grid operation, they are talking about an average value. The instantaneous driving impedance fluctuates wildly during the sine wave input signal. During most of the positive half of the input cycle, the grounded-grid looks negative with respect to the cathode--so the flow of current is cut-off. Since virtually no current flows, the driving impedance is extremely high.

During the negative swing in the input cycle, the grounded-grid is relatively positive. A positive grid accelerates electrons away from the cathode, producing high anode-current and grid-current. Due to the large flow of current, the input-impedance is low during the negative half of the input cycle.

Consider a pair of 3-500Zs. When the driving voltage is peaking at negative 117v, the anode-current is at its peak, and the instantaneous anode-voltage is at its lowest point--about +250v. At this instant, the total, peak cathode-current is 3.4a. Thus, the instantaneous cathode driving impedance is 117v/3.4a = 34.5 ohm--and the peak driving power = 117v x 3.4a = 397W.

In other words, the instantaneous driving impedance swing is from near-infinite all the way down to 34.5 ohms. The instantaneous drive power requirement varies from 0w at the positive peak to 397w at the negative peak of the input sine wave. Thus, the input pi-network's job is to act as a flywheel/energy storage system and a matching transformer. That's why a simple broadband transformer can not adequately do the job of matching the driver impedance to the cathode impedance in a grounded-grid amplifier.

The Q of a tuned circuit is like the mass of a flywheel. More Q makes for a better flywheel--which does a better job of averaging the wild swings in input-Z--thereby producing a lower input-SWR. The trade-off is that more Q means less bandwidth. With a high Q, the input SWR may be near-perfect at the center of the band, but unacceptable at the band edges. Thus, a compromise is in order. Eimac® typically recommends using a pi input network Q of 2 for Class AB2 grounded-grid operation. To arrive at a Q of 2, the reactance [X] of the input capacitor, C1, is minus j50 ohm÷2=minus j25 ohm. Using C=1÷[25(2f)], approximately 220pF of input capacitance is needed for a Q of 2 on the 10m band. In actual practice, however, 220pF may be far from the value that produces a satisfactory SWR with a particular model transceiver and a particular length of coax. It may be possible to find a length of coax that would ameliorate this problem on 10m--but there are eight other bands to contend with below

30MHz. Since band switching different lengths of coax is hardly practicable, it would be useful if the input capacitors were adjustable in a grounded-grid amplifier's tuned input circuits. Adjustable coils are also useful.

Tank Circuits

When the Q of the output pi-network tank circuit is low, two problems can occur. The harmonic attenuation may be inadequate to meet FCC requirements--and the load impedance matching range decreases. In other words, when Q is low, the tank circuit may be incapable of matching even a 50 ohm load. When the Q of the tank is too high, efficiency decreases due to the increase in I^2 R circulating current losses. A compromise is in order. A Q of 10 is about minimum. A Q of 20 may cause excessive tank component heating due to high circulating current. A Q of 12 to 15 is a fair compromise.

Better tank performance can be achieved by using a pi-L tank circuit. When compared to a simple pi, the pi-L has roughly 15db better harmonic attenuation and it typically has a wider matching range. The trade-offs are that the pi-L requires an extra switch section and a tapped inductor.

Skin Effect and Current Capability

As frequency increases, progressively less current flows inside a wire--so current progressively concentrates on the surface. Since a steadily decreasing part of the conductor is being used, resistance increases as frequency increases. For example, a 12 gauge (copper) wire will carry 20A at 60Hz with very little heating. At 30MHz, the RF current carrying ability of 12 gauge wire is about 5A. Band switch contact current ratings need to be similarly de-rated as frequency increases. Paralleling contacts is a good way of increasing the current handling ability of a band switch. Directing a portion of an amplifier's cooling air flow at the band switch improves the RF current handling ability of band switch contacts.

HF tank inductors can become quite lossy unless the conductor surface area varies in proportion to frequency. Inadequate tank conductor size is the main reason for decreasing amplifier efficiency at the higher frequencies. A tank inductor made from 14 gauge wire is usually more than adequate for efficient 1.8MHz operation at 1500W PEP. For efficient operation at 29MHz, approx.10 mm o.d. copper tubing (or copper strap with an equivalent surface area) is appropriate. However, due to normal QSB--at the receiving end, even a one-third decrease in transmit power is virtually undetectable. Thus, squeezing out the last percentage of efficiency on 10m is not very important.

Calculating the RF circulating current in a tank inductor is fairly complex. A quick approximation is to multiply the maximum anode-current by Q. For example, if the anode-current is 1.2A and the tank Q is 15, the RF circulating current in the tank will be 1.2*15 =18A. At 29MHz, 18A is a formidable amount of current.

Silver

Compared to copper, silver [Ag] is cosmetically more attractive and more immune to oxidation. However, silver does not make an amplifier measurably more efficient at frequencies below about 100MHz. Copper oxidation can be prevented by polishing copper with extra fine steel wool and applying clear, gloss, polyurethane varnish.

Silver is useful as a component of solder. 95% tin [Sn], 5% silver, solder has a melting temperature of 221 degrees-C/430 degrees-F. Compared to tin-lead electronics solder, 95/5 Sn/Ag solder is about 3.5 times stronger and it has better wetability--especially on hard-to-solder materials. 95/5 Sn/Ag solder is ideal for soldering tank components, band switches, surface-mount solid state devices, loose vacuum tube pins, and low Q parasitic suppressors. When resoldering a tin-lead solder joint with tin-silver solder, first remove as much of the tin-lead solder as possible.

Anode HV RF Chokes

The basic requirements are: 1. The choke must have ample reactance at the lowest operating frequency to limit the RF current through the choke to a reasonable amount. 2. The choke can not be self-resonant near an operating frequency. 3. The wire gauge used must be able to carry the DC anode current plus the RF current at the lowest operating frequency without excessive heating.

If the HV-RFC has a self-resonance on or near an operating frequency, potentials of many times the anode supply voltage can appear on the choke. When this occurs, a choke arc and fire is likely. Choke fires can destroy more than just the choke because the rising plume of ionized gasses from the choke fire often creates a conduction path to the ceiling of the RF output compartment. If an arc occurs, pervasive damage is likely if no glitch protection resistor was used in the HV positive circuit.

Choke Design

Materials:
There are two types of wire insulation materials that are suitable for use in HV-RFCs--silicone varnish and Teflon. Modern, high-temperature electric motor wire is insulated with a tough, silicone varnish that can handle high DC voltage and high RF voltage. At room temperature, a twisted pair of #20 silicone varnished wires can withstand more than 5000VDC or 1500W in a 50 ohm circuit at 29MHz. This type of wire is sold by the pound in electric motor rewinding shops. If you want to buy some, bring your own empty spools and winding device--such as a variable-speed electric drill, with a homemade adapter to hold the spool. Due to its toughness, silicone varnish insulation requires a special method of stripping. An open flame from a butane lighter causes the silicone varnish to decompose and combust. The remaining ash residue can be removed from the copper with steel wool.

Teflon insulated magnet wire is not common. Although ordinary Teflon insulated hookup wire may be used, the extra insulation thickness requires that a longer coil form be used. One potential trade-off with Teflon insulated wire is phosgene. When Teflon burns, deadly phosgene [$COCl_2$] gas is produced.

Due to contact with air, the current carrying ability of either type of wire is much higher in an HV-RFC than it would be in a transformer. #28 wire will easily carry 1A in a HV-RFC. #24 will carry several amperes with acceptable heating.

G10 or G11 epoxy-fiberglass tubing is RF-resistant, strong, and easy to work with. It is an ideal material for building HV RF chokes. It can be obtained from plastic supply houses. 1mm wall thickness is more than adequate. Diameters of 16 to 25 mm are typically used for building HV-RFCs. G10 tubing can be cemented to a G10 base plate with silicone rubber adhesive or epoxy. A source of G10 tubing: Plastifab,1425 Palomares, La Verne, CA 91750 818 967 9376.

It is probably a good idea to limit RF current in the HV-RFC to no more than 1 ampere. To calculate current in the choke, take roughtly 2/3 of the anode supply volts and divide it by the reactance in ohms at the lowest operating frequency -- a.k.a. Ohm's Law.

Determining Bypass C

Power supply components can be damaged by RF. Electrolytic filter capacitors are especially at risk. Thus, adequate RF bypassing on the power supply side of the HV-RFC is needed. Probably no more than 10V of RF should be allowed to appear on the +HV supply at the lowest operating frequency. Determining just how much bypass C is needed basically involves using ohm's Law. The amount of RF current flowing through the choke and the amount of bypass C need to be evaluated for the lowest operating frequency--usually 1.8MHz. For example, if the reactance of the choke is +j2000 ohms, and the AC anode voltage is 2000Vrms, then I=2000V/2000 ohm=1A of RF flows through the choke. In order to limit the RF voltage to 10V maximum at 1.8MHz, 10V/1A=10 ohm of capacitive reactance is needed for an adequate bypass. Using C=1/(Xc * 2pi * f), this equates to a HV bypass capacitance of 8842pF. Obviously, a typical 1000pF bypass C [minus j88 ohm] is not going to do the job because it would allow approx. 88V of RF to appear across the HV supply if 1A were flowing through the choke.

500pF 20kV TV-type doorknob capacitors are NOT designed to handle RF current--so they do not make satisfactory HV bypass capacitors. Disk ceramic capacitors may be used for HV bypassing. Disk ceramic capacitors are somewhat limited in the amount of RF current they can safely handle. Manufacturers typically don't publish RF current ratings for them. To find out how different capacitors react to RF current, you must test them yourself. Even a 7500WVDC, 2500pF disk ceramic capacitor becomes warm from 1A at 1.8MHz. Thus, it is often best to parallel a number of individual bypass capacitors so that the RF current will be shared among them.

Determining L

At the lowest operating frequency, the HV-RFC should have enough reactance to limit the RF circulating current through the choke to a reasonable amount. Allowing a RF current of 1A RMS through the choke usually does not create problems for the wire-lead disc-ceramic capacitors that are typically used to bypass RF on the power supply side of the HV-RFC. To minimize RF current through the choke, it would seem that more inductance is the answer. However, more inductance means more choke resonances and a greater likelihood of choke fires. A compromise is indicated.

Over the years, various schemes have been used to minimize choke resonances. Adding gaps at presumably esoteric positions in the winding was represented as a means of decoupling parts of the choke winding--allegedly ameliorating the self-resonance problem. However, when the resonances of gapped chokes are compared to similar chokes without gaps, no real improvement is observed on a dipmeter. This should not be surprising. Optimum decoupling between two coils occurs when they are mounted at a right angle. Adding end-to-end spacing with gaps is the least effective decoupling method possible. To minimize resonance problems, instead of using a single large choke, use two smaller chokes mounted at right angles.

The highest-L choke that can built that is free of self-resonances in the HF spectrum is roughly 60µH. At 1.8MHz, 60µH has a reactance of about +j679 ohm.

The RMS voltage that appears across an amplifier's HV-RFC is approximately two-thirds of the anode supply voltage. For example, an amplifier that is powered by a 3000V supply subjects its HV-RFC to about 2000V RMS. If a 60µH inductor was used in this amplifier, at 1.8MHz the RF current through the choke would be 2000V/679 ohm=2.95A RMS. Adequately bypassing approx. 3A of current on the power supply side of the choke is difficult. A typical HV disk ceramic bypass capacitor can handle only about 1A. Another problem is that at 1.8MHz 130pF [minus j679 ohm] of extra capacitance is required from the tune capacitor to cancel the +679 ohms of reactance in the choke. Adequately bypassing 3A at 1.8MHz requires a substantial amount of capacitance. To hold the voltage across the bypass capacitors to less than 10V at 1.8MHz, roughly 0.026µF [minus j3.3 ohm] is indicated. To handle this amount of current, four approx. 0.0075µF HV disc ceramic capacitors would probably be needed. All things considered, using more inductance is indicated. Limiting the HV-RFC's RF current to a maximum of !A would make the task of bypassing a lot easier. However, increasing the inductance above 60µH is virtually certain to move choke resonances into the HF range. Unless these resonances are prudently parked between operating frequencies, a choke fire may result.

To realistically evaluate the self-resonance situation, HV-RFCs should be checked with a dipmeter after they are installed and wired in the amplifier. If a self-resonance is within about 5% of an operating frequency, there may be a problem. When re-parking resonances, it is usually best to remove turns from the choke. This will move the resonances up in frequency--and only slightly increase the maximum RF current through the choke.

In continuous coverage amplifiers, there are obviously no safe parking places for choke resonances. The only solution is to switch HV-RFCs with one or more HV vacuum relays.

HV-RFCs should be single-layer solenoid wound. To minimize wire vibration during operation, the wire should be under constant tension when winding and soldering the ends to the solder lugs. When silicone varnish insulated wire is used to wind a HV-RFC, the finished winding should be given a coat of gloss urethane varnish to hold the wire in place. Since varnish will not adhere to Teflon wire, a different method is needed to keep a Teflon winding taught. Small tensioning springs are soldered to the ends of the wire. The springs provide constant pull to minimize wire vibration during modulation. An S-shaped copper foil jumper should be connected across each tensioning spring.

DC Blocking Capacitors

Blocking high voltage DC is the least difficult part of the blocking capacitor's job. During operation on 10m, the DC blocking capacitor must be able to carry most of the RF circulating current in the tank. Here's why: The amplifier tube's anode capacitance normally provides most of the tune capacitance during 10m operation. Thus, a major portion of the tank circulating current passes through the anode capacitance and therefore through the DC blocking capacitor. In an amateur radio amplifier, blocking capacitor currents of 5 to 10 A RMS are not uncommon during operation on the 10m band.

Selection of a blocking capacitor should not be guesswork. It is advisable to select a capacitor or capacitors that is rated to carry the calculated maximum RF current present. Merely selecting an RF-type (transmitting) capacitor is not good enough. Some RF-type capacitors have rather unspectacular current capabilities. The capacitance of the DC blocking capacitor is not very critical. 1000pF seems to be more than adequate for operation at 1.8MHz. 88 ohms of Xc is relatively insigificant in comparison to the typical 1000 to 2000 ohm anode output Z.

Vacuum Components

Vacuum capacitors and vacuum relays are ideal for use in high power RF amplifiers because they can withstand high RF voltages. vacuum capacitors are able to handle more RF current than any other type of capacitor. There are some trade-offs. Vacuum components depend on their glass-to-metal or ceramic-to-metal seals to maintain their near-perfect vacuum. If a seal leaks, air molecules enter and the vacuum component is kaput. Vacuum component seals should not be subjected to unnecessary mechanical stress.

Vacuum Capacitors

Although vacuum capacitors can be mounted in any position, vertical mounting places the least stress on the soft copper plates. Vertical mounting also makes the most efficient use of chassis space. With vertical mounting, a right-angle drive is used to bring the 1/4" diameter tuning shaft to the front panel. Cardwell-Multronics® makes a compact right-angle drive mechanism that is ideal for this application. It is designed to replace the shaft-cap on a vacuum capacitor's tuning shaft. The vacuum capacitor should be set for minimum C before the drive shaft cap's setscrews are loosened.

A vacuum capacitor should not be used as a standoff-insulator to support heavy components. High G force can be fatal to a vacuum capacitor. The danger is not necessarily breakage or damage to the seals. The plates in a vacuum capacitor consist of a series of concentric, intermeshing, soft copper cylinders that almost touch each other. A vacuum capacitor can be shorted by an inertia force that is capable of bending the soft copper plates.

Vacuum Relays

To avoid stressing the seals, connections to the contact terminals of vacuum relays should be made with soft copper ribbon.

The molded-in coil terminals on vacuum relays are easily broken. Connections to the coil terminals should be made with approx. 24 gauge stranded hookup wire.

Vacuum relays generate sharp mechanical vibrations when they switch. If one is mounted securely to the chassis, the chassis acts like a speaker cone--coupling the vibrations more efficiently to the air. One way of overcoming this problem is to mount the vacuum relay on small beads of silicone rubber. To accomplish this, drill a approx. 3mm oversize mounting hole in the chassis. Use temporary L-shaped poster board spacers to prevent the relay from touching the chassis. After cleaning the surfaces with acetone, apply three small beads of silicone rubber between the relay mounting flange and the chassis. Allow the silicone rubber to cure for 2 days. Remove the spacers. The relay should float quietly on silicone rubber shock absorbers. The vacuum relay's body should be grounded to the chassis with thin copper ribbon. The ribbon may be soldered to the edge of the relay flange. To avoid overheating the seals, use a large soldering iron--and tarry ye not.

Controlling Make and Break Times

All relay coils have inductance. Since inductance delays a change in the flow of current, coil-inductance tends to increase the make-time of relays. Make-time is an important design consideration when using vacuum relays for RF switching. RF-rated vacuum relays use copper contacts to obtain high conductivity. However, copper is vulnerable to damage from hot-switching. For example, if an amplifier's RF output relay contacts are not closed and finished bouncing before the RF arrives, arcing and contact damage is likely.

Make-time can be decreased by supplying extra voltage to the coil during start-up with what is commonly called a speed-up circuit. Jennings® and Kilovac® recommend using them to accelerate relay closure. A speed-up circuit consists of a resistor in series with the relay's coil and a power supply that supplies two to three times the rated coil voltage. At turn-on, the extra voltage hastens the flow of current in the coil. The resistor limits the steady-state coil voltage to a safe value after the flow of current builds up in the coil.

DC relay coils are usually paralleled with a diode to absorb the reverse voltage spike that results when current stops flowing through the coil. If no reverse diode is used, the reverse voltage spike can exceed 20 times the rated coil voltage. The break-time of a DC relay can be controlled by adding a resistor in series with the diode. As R increases, the break-time decreases. R should probably not exceed three times the coil resistance.

Testing Vacuum Components

When a vacuum seal leaks air, the breakdown voltage decreases. This problem is easy to spot in a glass-body vacuum relay--because when electrons flow through air, blue-purple photons are emitted. With glass-body vacuum capacitors, this problem is not as obvious. In a leaky glass-body vacuum capacitor, internal ionization/arcing is often not visible since the problem usually occurs deep inside the meshed concentric plates.

It is a good idea to test all vacuum components, whether they be new or used,before constructing the amplifier.

When a vacuum component in an amplifier becomes gassy, arcing typically occurs near the crest of the RF sine wave--so a bad vacuum component typically reduces the peak power output. Since many amplifiers use more than one vacuum component, finding the bad one is difficult without individual evaluation using a breakdown tester.

Random replacement--a.k.a. "Easter-egging"--is not an efficient way to repair an amplifier that uses vacuum components. For instance--if an amplifier's RF output vacuum relay becomes gassy, it is virtually certain to divert high power RF into the (usually more delicate) RF input relay. If a thusly-damaged RF input relay is replaced, the new RF input relay may also be damaged by the gassy RF output relay. Thus, it is desirable to be able to individually test vacuum components with a breakdown tester.

Testing the quality of a vacuum is similar to testing the breakdown voltage of a diode. Connect a approx. 100M ohm HV resistor and a approx. 20 microampere meter in series with a breakdown tester. Increase the voltage until about 1 to 2 µA of leakage is detected. This voltage is the breakdown voltage. The peak RF working voltage of a vacuum component is roughly 60% of the DC breakdown voltage.

Measuring Relay Contact Resistance

There are basically two types of vacuum relays--those that are designed for hot-switching, and those that are not. Hot-switching-capable relays have tungsten contacts. Such relays are intended for use primarily in power supplies. Relays that are designed for RF have copper contacts. They should never be allowed to hot-switch. Copper-contact relays have approximately one-third the contact resistance that similar tungsten-contact relays have. For instance, the Jennings RJ-1A is the copper-contact version of the tungsten-contact RJ-1H. The rated contact resistance of the RJ-1H is 30 milli-ohms. The rated contact resistance of the RJ-1A is 10 milli-ohms. Tungsten contact relays are not rated for current RF current. However, they should work fine for RF service if they are operated at roughly two-thirds of the RF current rating for their copper-contact counterparts. Tungsten contacts are extremely hard. They are capable of more operations than copper contacts. For heavy , full break-in telegraphy use, tungsten contacts are preferable--even though they do not have the continuous RF current handling capability of copper contacts.

With vacuum relays, contact failure is not uncommon. Contacts suffer from contact erosion. This condition increases contact resistance. Eventually, an eroding contact will open completely. To test a vacuum relay, the resistance of normally open [NO] contacts and the resistance of normally closed [NC] contacts should be measured and compared with the manufacturer's specifications. Ordinary ohm-meters are not suitable for detecting contact problems other than an open circuit. The voltage drop across relay contacts should be measured with a substantial current flowing. 1A is a reasonable current to use. Measure the mV drop directly across the contact terminals using a DMM with test prod leads. Most of the vacuum relays that are designed to handle RF current have a rated contact resistance of less than15 milli ohm--so no more than 15 milli V should appear across the terminals with 1A flowing through the contacts.

Testing Vacuum Capacitors

Vacuum capacitors store energy efficiently because they have virtually zero ESR [equivalent series resistance] and internal L--thus, the peak discharge current can be astronomical. When the breakdown test voltage is high enough to create more than a few microamperes of leakage, a vacuum capacitor will normally self-discharge--producing a clearly audible tick due to the large peak discharge current and commensurately large electromagnetic force. After a vacuum capacitor self-discharges, it begins charging and the process repeats. A vacuum capacitor should not be allowed to self-discharge more than a few times unless the capacitor has been in storage for many years. During long-term storage, for some as yet unexplained reason, copper atoms tend to line up, forming whiskers on the surface of the plates. Copper whiskers initially reduce the breakdown voltage. Copper whiskers can be dislodged by self-discharge. If the breakdown voltage increases after a self-discharge, another self-discharge may be beneficial. Repeated self-discharge will cause a decrease in breakdown voltage.

Grounded-Grid Amplifier Tune-up

Linear amplifiers are like induction motors--they are designed to run fully-loaded. If your grounded-grid amplifier's instruction book says to reduce drive power during tune-up--and most of them do--it is not giving you correct information. In order to be linear, amplifiers must be tuned-up with the same peak drive-power level that they will be driven with during actual operation. Reducing drive power changes the output Z of the amplifying device to something other than the tank circuit was designed to match. Thus, the tune and load settings with low drive will be wrong when normal drive is applied.

Tune-up method #1: Set the amplifier's HV supply to the CW-Tune/low-V-tap. If you are not sure where to preset the Load control, set it to >70% of maximum loading [30% of C] to be safe. Apply the drive level that you intend to drive the amplifier with during actual use. Alternately adjust the amplifier's Tune and Load controls for maximum relative power output. The whole process should take less than 6 seconds. It may sound brutal, but this tune-up method results in good amplifier linearity and it won't damage the tubes if the maximum anode-current rating is not exceeded. If the anode-current is excessive, the resistance of the cathode, RF negative-feedback resistor needs to be increased slightly--or the PEP adjust control in the transceiver needs to be turned down.

Tune-up method #2: [not for FM, AØ, and RTTY operation] To reduce the stress on an amplifier during tune-up, use a reduced duty-cycle driving signal. This can be accomplished by keying the transceiver, on CW mode, with a CW keyer, set to send approx. 50wpm dits. CW dits have a 1/2-on, 1/2-off, or 50% duty-cycle. Using this method, the amplifier may be tuned-up, again for maximum power output, in its higher-voltage, SSB-mode. Keyers that have a weighting adjustment can be set to produce a light dit that has a duty cycle of about 30% instead of the normal 50%. Another device for reducing the duty-cycle during tune-up is a tuning-pulser.

If you want to operate with reduced power during good band conditions, first tune up your amplifier with normal drive power, then turn the microphone gain down to reduce power.

Class AB1 Design Considerations

Choosing the Optimum Value of Grid Terninating Resistance.

Tetrodes and pentodes require a peak RF drive voltage that semi-matches up to the grid bias voltage.

My strategy is to choose a value of grid termination resistance that roughly provides the needed peak RF drive V to the grid with exciters that develop 100v-p (100w rms) to 141v-p (200w rms) across 50 ohms. In other words, the goal is to match peak RF drive volts with the needed grid bias volts from the tetrode/ pentode manufacturer's technical specifications.

- -- For tubes that require 50v to 70v of grid bias, like the 4CX800A and 4CX1000A, a voltage-halving bifilar stepdown transformer driven 12.5 ohm grid termination is used.
- -- For tubes that require 100v to 140v of grid bias, like the 4CX1500A, a directly-driven 50 ohm grid termination would be used.
- -- For tubes that require 200v to 280v of grid bias, like the 4-1000A and 8169, a voltage-doubling bifilar stepup-transformer driven, 200 ohm grid termination is used.
- -- For tubes that require greater than 300v grid bias, like the 8171, a trifilar stepup-transformer driven, 450 ohm grid termination is used.

However, if the peak grid V with max. drive is still a bit much, a cathode RF negative feedback reisistor (Rk) can be added to make up the difference. However, a trade-off is that the peak V drop across Rk normally subtracts from the screen to cathode V at the critical anode current peak. A workaround is to use the circuit shown in Figure 10.

Tuned Input Circuits for Neutralized, Class AB1 Grid-Driven Operation

Class AB1 grid-driven amplifiers look more complex than Class AB2 grounded-grid amplifiers. However, the tuned input circuitry for multi band Class AB1 grid-driven operation is comparatively simple.

The grid capacitance of tubes that are commonly used in Class AB1 grid-driven amateur radio power amplifier service ranges from about 15pF to 130pF. Since the capacitance of the grid is in parallel with the input, as frequency increases, input SWR worsens. This problem can be corrected by connecting a variable inductor in parallel with the grid. The inductive reactance {+j ohms} of the inductor is adjusted to cancel the capacitive reactance {minus j ohms} of the grid--thereby resonating the grid at the operating frequency. When the input SWR is tuned to minimum, the grid circuit is resonant. A simplified diagram is provided.

If the other end of the variable inductor is connected to a properly-adjusted capacitive voltage divider (connected between the anode and chassis ground), the amplifier is neutralized at whatever frequency the grid is tuned to. Obviously, this type of Class AB1 input circuit is a natural for continuous HF and MF coverage--just what's needed for operation on the 9 amateur bands below 30MHz. The ratio of the capacitances in the capacitive voltage divider equals the ratio of the feedback capacitance (the anode to grid capacitance) divided by the grid input capacitance. Typical ratios are 150 to 1 ... Achieving wide frequency coverage is not as easy in Class AB2 grounded-grid operation. A pi-network tuned input with the recommended Q of 2 has a limited bandwidth--so many, switched, tuned input circuits are required for wide frequency coverage.

Screen and Grid Supplies

There are many tetrodes and pentodes to choose from that are satisfactory for Class AB1 grid-driven operation. The essential criteria is that, with zero grid volts, the tube is capable of a peak anode-current that is at least triple its maximum (average) current rating. In most cases, this condition can only be met if near-maximum screen-voltage is applied. Relatively high screen-voltage is important because peak anode-current is a function of the screen-voltage raised to the 1.5 power.

For the best linearity, screen voltage should be regulated. For smaller tetrodes and pentodes, a Zener diode shunt regulator offers a good solution. Typically, a series of 10v to 30v, 5W Zeners are used. Screen voltage is adjusted by shorting out Zener diodes with a rotary switch. For larger tubes, an adjustable series-regulator is the best way to supply voltage to the screen. Thanks to modern power FETs and the venerable 723 IC linear regulator, building a reliable, regulated supply of 2kV or less is fairly simple.

Since the grid does not pass current in Class AB1 operation, there is no necessity to regulate the bias voltage. However, the bias supply should not have an extremely high output impedance. A maximum grid circuit R of 1k to 100k ohms is typically recommended by tube manufacturers.

Work Space

'Work-space' and 'head room' are terms that describe the range in which instantaneous anode-voltage is free to move up and down--thereby performing work. In a tetrode, at the maximum peak anode-current, to avoid excessive screen-current and a decrease in linearity, the instantaneous anode-voltage should not dip much below the screen-voltage. For example, a tetrode with a 4kV anode supply and an 700V screen supply, the work-space is approximately 4000V minus 600V = 3400V peak

In a pentode, the instantaneous anode-voltage may dip close to the suppressor-voltage--which is typically zero volts. In the above example with a screen-voltage of 800V, if the tube happened to be a pentode, the work-space would be around 3750V peak. Thus, pentodes enjoy slightly more work-space than tetrodes. As a result, pentodes are slightly more efficient than tetrodes. However, pentodes are more expensive than tetrodes because they are more complex to build. Sockets with low-L suppressor and screen bypass capacitors are needed for stable operation. Pentode sockets are **not** inexpensive. Another trade-off is that there are relatively few types of pentodes to choose from. A (if not the) suitable pentode for amateur radio Class AB1 grid-driven service is the 5CX1500.

Pentode Caveats

Pentodes typically have less feedback capacitance than tetrodes. This advantage theoretically makes pentodes more stable. Some designers do not neutralize pentodes because they feel the relatively low feedback capacitance between the anode and the grid is insignificant. However, for optimum linearity and stability, plus low input SWR, a pentode should be neutralized. This can easily be accomplished with the grid input circuit diagram [Figure 5] for Class AB1 tetrodes. To use this circuit with a pentode, DC-connect the suppressor to the cathode with a l0 or so ohm resistor. However, the suppressor must always be RF-bypassed to chassis ground to decrease feedback from anode to grid.

Screen Protection

Every screen type tube has a maximum screen dissipation rating in watts. If screen current times screen voltage exceeds this rating, the tube could be destroyed. This can easily happen with a no load or light load condition--so various protection schemes are used. If the anode voltage disappears while screen voltage is present, screen current will be excessive unless a means of protection is provided. Another hazard is reverse screen current. Reverse screen current can easily become a runaway condition. It happens virtually instantaneously. Reverse screen current is commonly experienced in Class AB1 operation. Unless bled off into a resistor load or into a shunt Zener voltage-regulator, reverse screen current can quickly destroy a tube. For tubes with screen voltages in the 300V to 800V range, a shunt regulator using a Zener diode string is a good solution. The Zener regulator string is connected through a high-R resistor to the anode supply. A sample circuit is provided. A suitable tube would be the 4CX1500B, or similar types.

Advantages of Shunt Zener Screen Regulation:

- Limits the maximum current that can be drawn by the screen.
- Protects against reverse screen current.
- If the high voltage disappears, so does the screen voltage.

However, for larger tubes with higher screen current and screen voltage requirements, a Zener shunt regulator is somewhat impractical. A continuously-adjustable series-regulator screen supply is a better solution. To protect against reverse screen current, a shunt resistor/bleeder must be connected across the screen supply. A bleeder current flow of roughly 20% of the normal screen current seems to be adequate. 25% might be safer. To protect against excessive forward screen current, a fast acting fuse or magnetic-type circuit breaker is incorporated in the primary of the screen supply power transformer. An adjustable series regulator circuit is provided.

Grid-Driven Class AB1 Amplifier Neutralization and Tuneup

ßAdjusting a Class AB1 amplifier may look complicated at first, but after you have done it a few times, and you begin to understand the reason behind each step, it gets easier.

Neutralization: The goal of neutralization is to isolate the anode from the grid at the operating frequency. Neutralization discourages regeneration--oscillation. Neutralization usually needs to be adjusted only once.

1. Disconnect the amplifier from the electric-mains.

2. Temporarily disconnect the tank circuit from the HV blocking-capacitor.

3. Substitute a low-L film resistor, with the same R as the design anode-load [output] resistance, in place of the tank circuit. Typical values would be 1000 ohm to 4000 ohm, 2W. The resistor connects to the blocking-capacitor and to chassis-ground. Connect an RF-voltmeter or an oscilloscope equipped with a 10 to 1 hign impedance probe across the resistor.

4. Connect the amplifier to the electric-mains and turn on the transmit-receive relay power supply plus the grid and filament supplies. Do not turn on the screen or anode supplies.

5. Drive the amplifier with 20m or 15m RF. Tune the grid-circuit variable-inductor [L1] for minimum input SWR or minimum reflected power. If necessary, adjust the DC grid-voltage so that virtually no grid-current flows.

6. Adjust the neutralizing-capacitor (C3) for minimum RF-voltage at the anode-load resistor. If needed, readjust L1 for best input SWR followed by readjustment of C3. This completes the neutralizing procedure.

After C3 is nulled, the amplifier is neutralized for all bands. To confirm this, check the neutralization on another band. Readjust L1 for minimum SWR. The RF voltage across the output load resistor should not change appreciably. Typically, no further adjustment is necessary--even if the tube is replaced.

Remove the resistor and reconnect the tank circuit.

Tune-up.

1. Switch off the screen and HV anode supplies. Switch on the T/R relay supply, the filament supply and the grid supply.

2. Transmit on CW-mode into the amplifier and adjust L1, the grid roller-inductor, for minimum input reflected power. This tunes out the grid-reactance and simultaneously neutralizes the amplifier at the operating frequency. If you are using a transistor-output transceiver, to preclude SWR shutdown,

initially tune the grid with no more than 5W of signal.

3. Apply full drive power using an electronic keyer sending dits at about 50wpm, or a use a tuning-pulser. Adjust the DC grid-voltage so that <0.1mA of grid-current flows. The grid-voltage is adjusted so that the grid is on the threshold of current flow. The grid-voltage adjustment is not used to set the zero-signal anode-current [ZSAC]--also known as 'idling current' or 'resting current'. Although the grid-voltage adjustment can discretely be used to make a small adjustment in the ZSAC, in Class AB1 operation, the primary criteria for setting the grid-voltage is that virtually NO grid-current flow with maximum drive. ZSAC is set by adjusting the ***screen voltage***. Switch on the screen and HV supplies. Key the amplifier but do not apply drive power. Using the screen voltage adjustment, set the ZSAC as recommended by the tube manufacturer. For most tubes, the ZSAC should be about 20% of the rated anode current.

4. If a variable tank inductor and a variable tune capacitor is used, preset the tune capacitor and the tank inductor for the desired operating Q on the band in use. Preset the load capacitor and inductor by calculation. It is best to error on the side of too-little load C [heavy loading]. If too-light loading [too much load C] is used, excessive screen-current is likely. Remember that the tune C sets the operating Q. Most of the tuning should be done with the variable tank inductor. Fine tuning can be done with the tune C--but the final setting should no be very far from the setting for the correct operating Q.

5. When any amplifier is tuned-up, the anode-current must be driven to the maximum, peak, design value so that the tube's output load resistance will meet the design criteria for the pi output tank circuit. If a lesser current is used without proportionately decreasing the supply voltage, the output load resistance will be too-high and the subsequent adjustment of the tank will be for the incorrect output load resistance.

To be both linear and deliver good power output, an amplifier tube must be adjusted by loading it for the optimum peak anode-voltage swing. The indicated screen-current is an accurate way of tuning up a tetrode or pentode. If the anode-voltage swing is too great because of too-light loading, the screen-current [and distortion] will increase. This means that the instantaneous minimum anode-voltage is less than it should be--a situation which causes too many electrons to stick to the screen--thereby depriving the anode of electrons. If the screen-current is too low, the anode-voltage swing is inadequate--meaning that the loading is too heavy. This condition causes lower power-output. When the output tank circuit is tuned correctly, the screen-current meter peaks. This is done by adjusting the tune capacitor or by adjusting the tank inductor. Do not peak the screen current with the loading capacitor.

·· *Thus, by using only the screen-current meter, tuning and loading can be adjusted for good linearity and good power output.*

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6. Set the transceiver to CW-mode. Apply full drive power. To reduce stress during tune-up, use an electronic keyer to send dits at about 50wpm. Standard dits are a 50% duty-cycle waveform, so current

meter indications are roughly half of the actual value. A tuning pulser works even better. [Figure 9]

7. Peak the screen-current by tuning the tank inductor or the tune capacitor. If the screen-current begins to become excessive, stop short of the peak, increase loading and continue. If the screen-current is too low, lighter loading is needed.

--The last step is to re-peak the screen-current with the tank inductor or the tune capacitor.

Loading for slightly less screen-current increases linearity with the trade-off of slightly less power output.

·· It is useful to keep a log of the various final settings for different frequencies. This saves time during future tune-ups.

For thoriated-tungsten cathode tubes only: - - While sending dits at full power, gradually reduce the filament-voltage until the relative output just begins to decrease. Increase the filament-voltage about 2%. This is the optimum filament-voltage. This should be rechecked every few hundred operating hours. The same thing applies to grounded grid amplifiers. For indirectly heated cathode tubes, like the 8877, the ideal filament voltage for communications services is near the minimum filament voltage rating. Under no circumstance should such a tube be operated at less than the minimum filament rating.

Distortion

Perfectly linear amplification produces nothing except a larger representation of the input signal. Non-linear amplification produces mixing--and mixing creates distortion products.

Inter-modulation distortion [IMD] is the result of mixing between two or more input signals. The human voice produces many frequencies at any instant. When voice modulation is amplified non-linearly, many mixing products are produced. This is called "splatter" or, more descriptively, "rotten splatter." IMD is usually measured by simultaneously applying two equal-amplitude, not harmonically related modulation frequencies such as 2000Hz and 2200Hz. When two or more frequencies mix they produce spurious signals at their sum and their difference frequencies--in this case 4200Hz and 200Hz. The first level of mixing produces what are called "third order products." Additional products are produced by third order products mixing with the two fundamental frequencies. For instance, 2200Hz and 4200Hz mix to produce a signal at 6400Hz.

When distortion products are inside the fundamental pass band of an AM or SSB transmitter, audible distortion results. This gives voice modulation a rough, unpleasant characteristic that reduces intelligibility. Odd-order distortion products which lie outside the pass band can cause interference on adjacent frequencies.

There are two methods of referencing IMD measurements. In method A, the IMD power level is referenced to either one of two equal-amplitude input signals. The power ratio of PEP to either of two equal-amplitude sine waves is four to one [6db]. In method B, the IMD level is referenced to the PEP level. Thus, an IMD level of minus 34db using method A equals an IMD level of minus 40db using method B. Amateur radio operators tend to use method B because receiver S-meters respond to PEP. In commercial radio, the military, and the FCC--where distortion measurements are typically made with a spectrum analyzer--method A is used. When using a spectrum analyzer, distortion can be broken down further into third order products, fifth order products, seventh order products. However, total IMD referenced to PEP is a more significant number.

It is possible to measure IMD without expensive laboratory equipment. All that's needed is a receiver and some understanding of what's required to make a meaningful measurement.

By comparing the signal strength in the transmitter's fundamental pass band window with the signal strength in the adjacent pass band windows, IMD can be measured fairly accurately--even over the air. The amount of receive frequency offset is critical. If the receive pass band is too close to the transmitter's fundamental pass band, the receiver will not be able to separate the IMD energy from the fundamental energy. As a result of this overlap, the distortion measurement will be higher than the actual amount. If the receive frequency offset is too far from the fundamental pass band, the receiver's pass band will not receive all of the IMD--and the distortion measurement will be lower than the actual amount.

For a receiver with two, cascaded SSB filters, a receive offset of 3.6kHz is about right--provided that the receiver is set to the same sideband as the transmitter. For a receiver with one SSB filter, an offset of about 4.5kHz is needed. To measure the IMD level of a LSB signal, offset LSB-receive higher in frequency. For measuring the IMD from an USB signal, offset USB-receive lower in frequency.

Since very few S-meters are linear, a calibration chart of S-meter readings versus decibels is a prerequisite for making accurate measurements. A calibration chart can be made with a step-attenuator and a signal source, or with a signal generator/attenuator.

In order to measure IMD, at least two modulation frequencies are required. Human speech is a good signal source for measuring IMD because, at any instant, speech contains many fundamental frequencies and harmonics. As its name suggests, another harmonic-rich signal source is a harmonica. By simultaneously blowing into two or three adjacent holes at the low note end, a plethora of frequencies can be produced that are optimal for making distortion measurements.

Splatter Reports

Before reporting splatter, it is important to keep in mind that all SSB, DSB, and AM signals have IMD. In other words, everybody splatters. The obvious question is how many decibels down is the IMD? Minus 40db is excellent; minus 30db is objectionable; minus 20db is abundantly abominable. With one exception, FCC rules allow virtually any level of IMD inside the ham bands. The exception is when IMD causes harmful interference to emergency communications. Splattering on non-emergency communications is NOT considered to be harmful.

Before reporting a station's level of IMD, it is advisable to determine whether or not the station operator is interested in hearing your report. Although most amateur radio operators are interested in transmitting a high quality signal, some operators deliberately misadjust their equipment to maximize IMD.

Notes on Measuring Power

Since E-peak = E-rms x 2^0.5, and P = E^2 ÷ R, at its crest, the instantaneous peak power in a sine wave is double the RMS power. A common unit of measuring amplifier output is the PEP [peak envelope power] watt. Despite the name, peak envelope power watts are not peak watts--they are RMS watts at the crest of modulation. If an amplifier was powered by a regulated anode supply, there would be virtually no difference between PEP watts and AØ [NØN] watts. In a typical amplifier, the anode-voltage sags appreciably under AØ conditions--so PEP watts are typically about 20% higher than AØ watts. PEP need not be measured with voice modulation. PEP can also be measured by keying the driver at 30 pps with a steady string of pulses that approximates the duty-cycle of a human voice--roughly 30%.

Tube Ratings

Traditionally, amateur radio operators have taken a cavalier attitude toward tube manufacturer's ratings. While some ratings can be exceeded judiciously, exceeding other ratings can be costly. Examples of ratings which should not be exceeded for indirectly-heated cathode tubes are minimum filament-voltage and maximum anode-current. Violation of either rating can result in destruction of the delicate cathode. Directly-heated cathodes are more rugged. The maximum anode-current rating for directly-heated cathode tubes is a linearity issue--not a cathode destruction issue. One rating which should not be exceeded is maximum seal temperature. It has been said that the way to tell when the blower is too big is if it blows the tube out the socket.

END OF PART 4

Calculating Tank Q, Tune C (C1), and Optimal Anode Load Resistance (R_L)

Tank Q, the reactance of C1, and the optimal anode load resistance for linear operation (R_L) are inter-related. Tank Q is defined as the capacitive reactance of C1 (X_{C1}) at the frequency of operation, divided into R_L ---i.e., $Q = R_L/X_{C1}$......and $X_{C1} = R_L/Q$. Note: C1 includes the anode (output) capacitance (Ca) of the amplifier tube. At 29MHz, Ca may be a sizeable fraction of C1.

$R_L = E_{supply}/2*I_{An}$ where I_{An} is the average anode current in amperes.

(Note: There is some variation in the constant in the denominator of the R_L formula. For tubes with minimal anode-cathode potential at peak anode-current, like the 8877, a constant of 1.6 should give more accurate results. However, for tetrodes like the 8171, which use a high screen potential (reduces anode AC peak-V), a constant of 2 seems to be more accurate.

Thus, for a tube operating from 2500v @ 1A, whose anode capacitance (Ca) is 10pF:

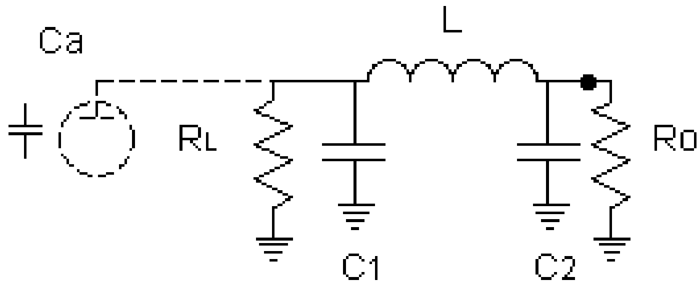
R_L = 2500v/2*1A = 1250 ohms.

Calculating C1

For a Q of 12.5, X_{C1}=1250 ohms/12.5 = 100 ohms.

The needed tune capacitance, C1 = 1/(2*Pi*f*X_{C1}). For 14MHz, C1 = 1/(6.28*14*10^6Hz*100 ohms) = 113.7pF. However, since part of C1 is comprised of Ca, the net tune C is 113.7pF -10pF = 103.7pF. At 28MHz, the tune C would be roughly: 57pF - 10pF = 47pF. At 280MHz, 10pF has about 100 ohms of X, so, for a Q of 12.5, Ca furnishes 100% of C1, so no tune C can be used.

Calculating the Load Capacitance (C2) and Tank Inductance, L



f = frequency in Hz
ω = 2π f (radians per second)
Ca = anode capacitance
C_1 = tune capacitance, including Ca
C_2 = load capacitance
R_L = calculated anode load R
R_O = output load R

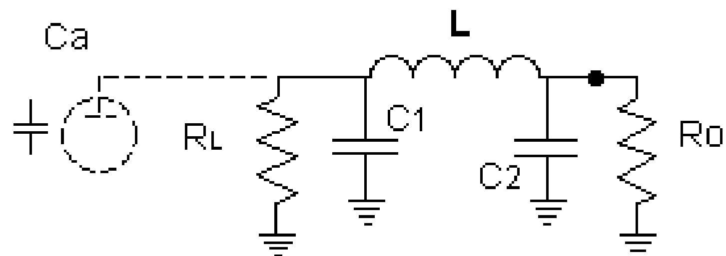
$$C_2 = \frac{1}{\omega R_O}\left[\frac{R_O (Q^2+1) - R_L}{R_L}\right]^{0.5}$$

Example: R_L=1250Ω, Ca=10pF, f=14 MHz, Q=12.5, Ro=50Ω
Solution: ω = 2π f = 6.28x14x10^6 = 88x10^6 radians per sec., (Q^2+1) = 12.5^2 +1 = 156 +1 = 157

$$C_2 = \frac{1}{88 \times 10^6 \times 50}\left[\frac{(50 \times 157) - 1250}{1250}\right]^{0.5}$$

C_2 = 521 × 10^-12 Farads $\boxed{C2 = 521\text{pF}}$

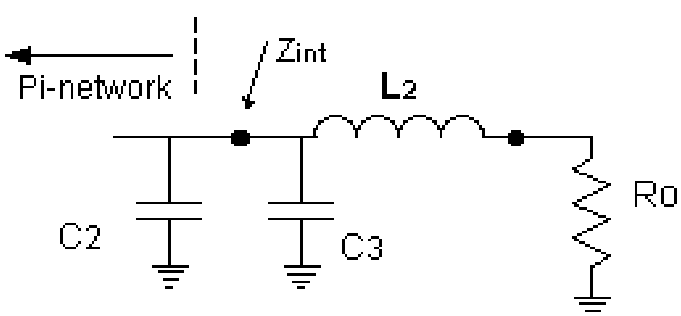
$$L = \frac{Q R_L + (\omega\, C2\, R_L\, R_O)}{\omega\, (Q^2+1)}$$



$$L = \frac{12.5 \times 1250 + (88 \times 10^6 \times 521 \times 10^{-12} \times 1250 \times 50)}{88 \times 10^6 \times 157}$$

$\boxed{L = 1.34\mu H}$

A Pi-L-network provides about 14db better harmonic suppression---and a wider load matching range---than a Pi-network. Typically, Z_{int} (the intermediate impedance) is 300Ω, and Ro is 50Ω. [••• Note: Z_{int} must be substituted for Ro in the Pi-network equations for C2 and L. The equation for C1 is the same for either a Pi-network tank or a Pi-L-network tank.]



$$X_{L2} = \left[(Z_{int}\ R_o) - R_o^2\right]^{0.5}$$

$$X_{L2} = [(300 \times 50) - (50 \times 50)]^{0.5} \quad \boxed{= 111.8\Omega}$$

$$X_{C3} = \frac{Z_{int} \times R_o}{X_{L2}} = \frac{300 \times 50}{111.8} \quad \boxed{= 134\Omega}$$

X_{L2}

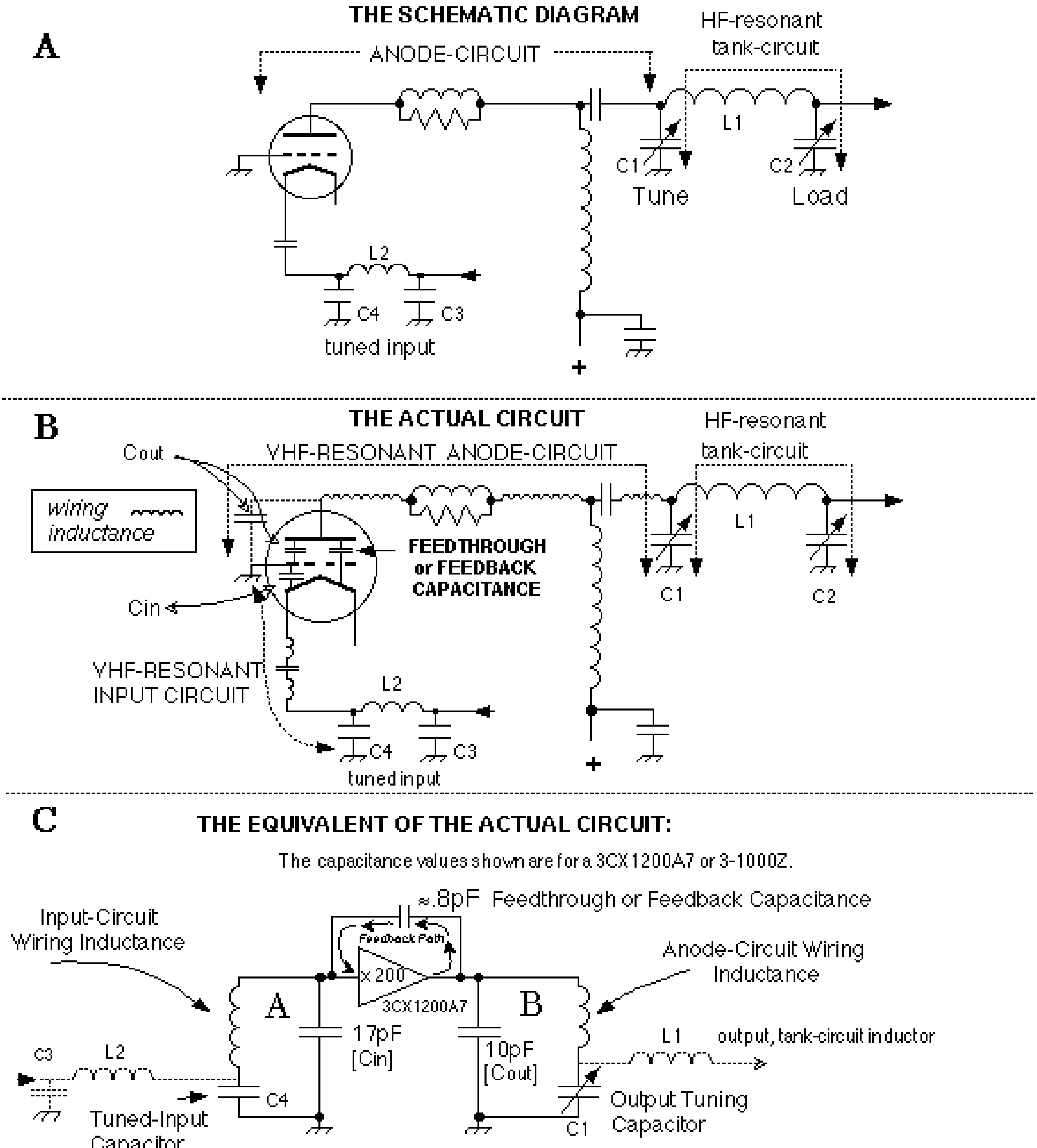
X_{L2} 111.8

$$L_2 = \frac{X_{L2}}{\omega} \qquad C_3 = \frac{1}{\omega \; X_{C3}}$$

$\omega = 2\,\pi\,f$

C2 and C3 are typically combined into one capacitor.

FIGURE 1 of 1 for: "Improved Anode-Circuit Parasitic Suppression For Modern Amplifier-Tubes"



The high-Q inductances and capacitances at A and B form high-Z, parallel-resonant, VHF circuits that cause the amplifier to exhibit a near-maximum VHF voltage gain. And so, the possibility of VHF regeneration and oscillation is enhanced. This problem can be corrected by reducing the VHF-Q at A and at B.

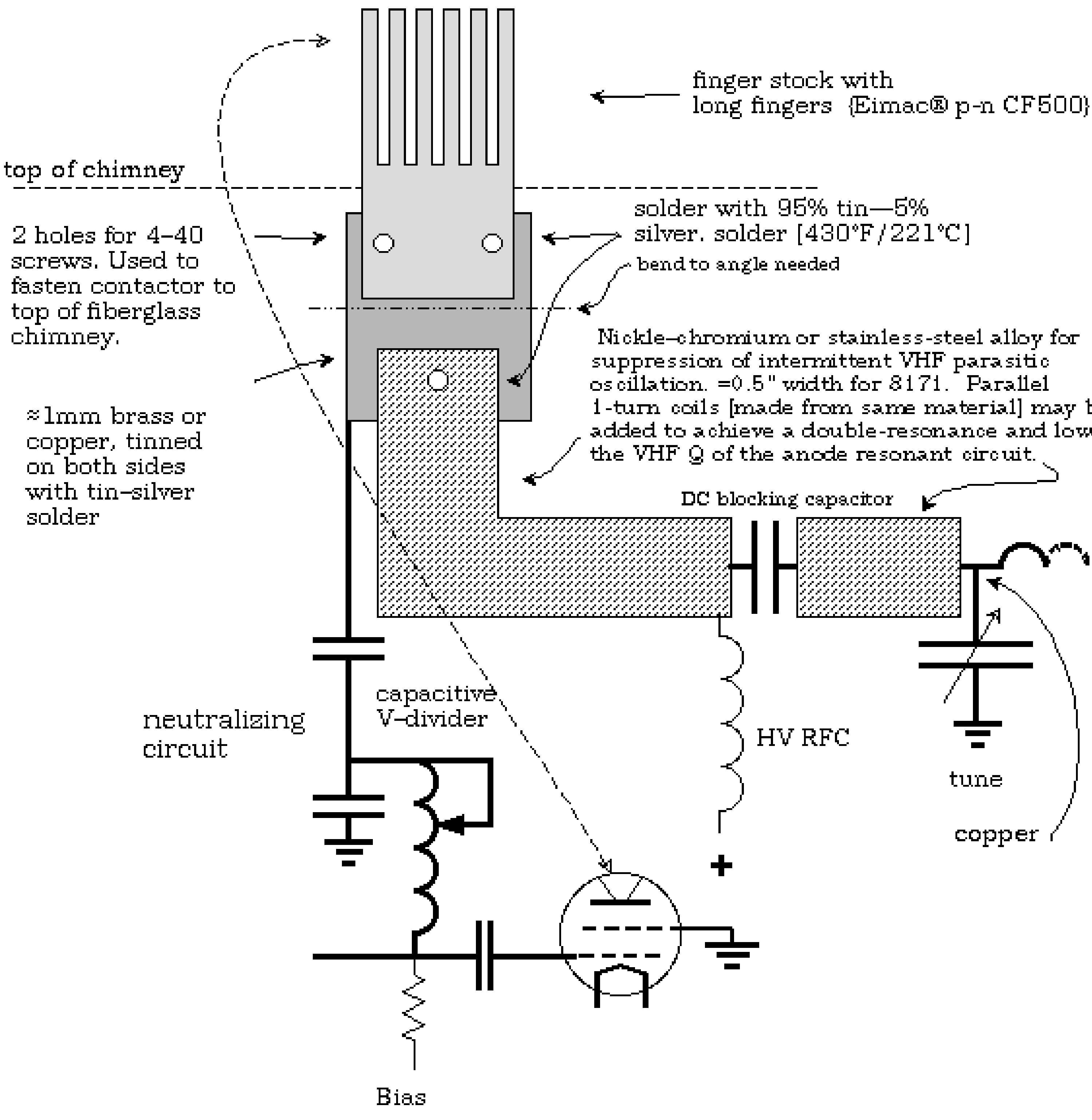
≤30MHz Anode Contactor for 8169, 8170, 8171, 8281, et cetera

This type of anode contactor:

1. Reduces the stray inductance in the Class AB1 amplifier's neutralizing circuit.
2. Makes the job of changing the amplifier–tube easier.
3. Reduces the inductance in the amplifier's anode circuit, thereby increasing the VHF self-resonant frequency of the anode–circuit and reducing the chance of an intermittent VHF parasitic oscillation

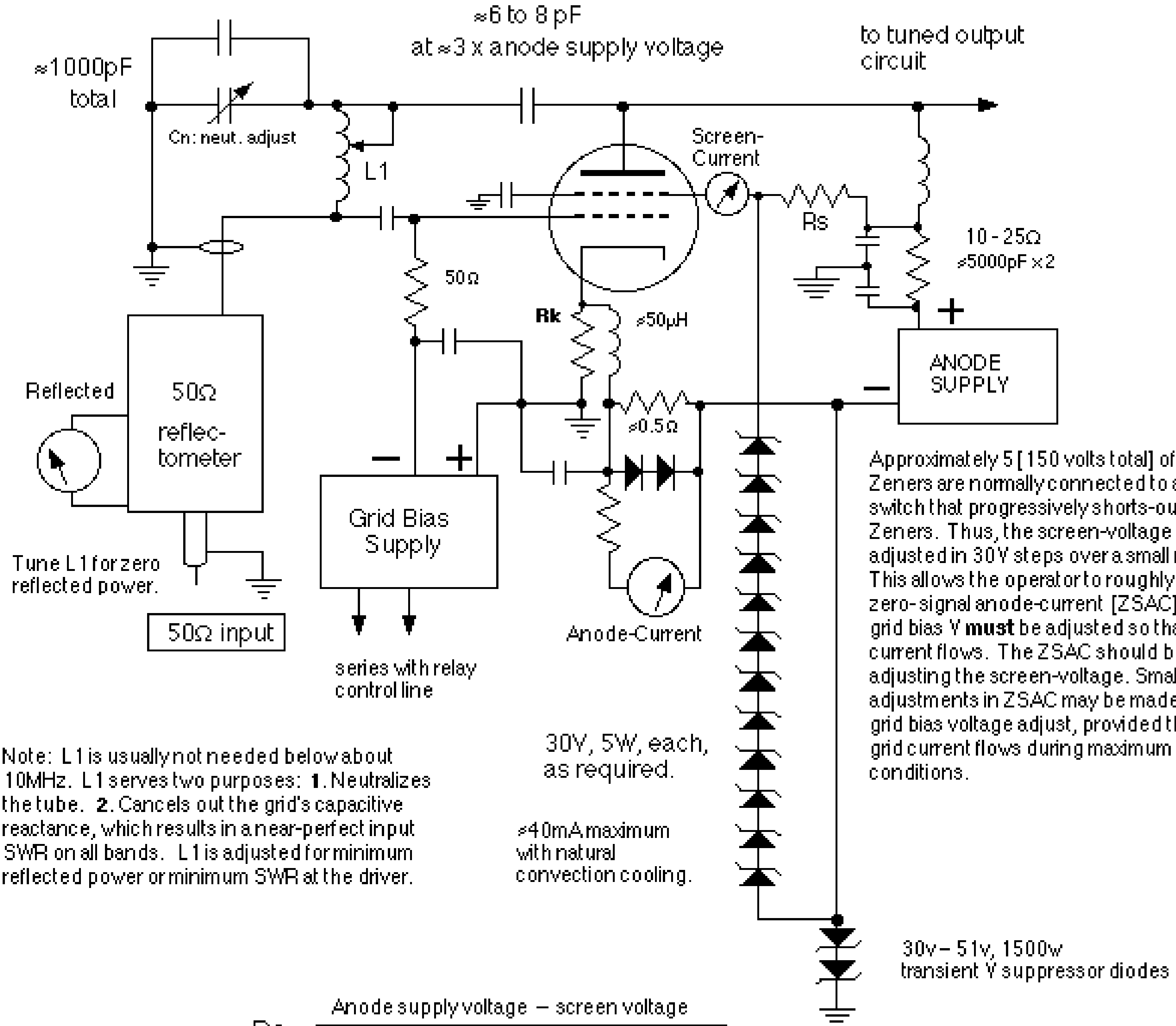
****Note**: If the low VHF Q [nichrome/stainless] conductor material overheats on the highest frequency used, increase the width of the conductor. Use of a wider conductor results in a higher VHF Q.

**If you replacing existing copper or silver strap strap in an amplifier, it might be interesting to do a before and after comparison with a dipmeter.



10. General Coverage, Class AB1, Grid-Driven Amplifier, Simplified Diagram, for Tubes with Less Than ≈1000 Screen Volts [see text]

This circuit is fail-safe. If the HV supply fails, the screen voltage goes away. If the amplifier is inadvertently operated without a load, the screen can not draw excessive current. When the screen current reverses during tuneup, the reverse current is safely shunted through the zener diodes.



Approximately 5 [150 volts total] of the Zeners are normally connected to a rotary switch that progressively shorts-out these Zeners. Thus, the screen-voltage can be adjusted in 30V steps over a small range. This allows the operator to roughly set the zero-signal anode-current [ZSAC]. The grid bias V **must** be adjusted so that no grid current flows. The ZSAC should be set by adjusting the screen-voltage. Small adjustments in ZSAC may be made with the grid bias voltage adjust, provided that no grid current flows during maximum signal conditions.

Note: L1 is usually not needed below about 10MHz. L1 serves two purposes: **1.** Neutralizes the tube. **2.** Cancels out the grid's capacitive reactance, which results in a near-perfect input SWR on all bands. L1 is adjusted for minimum reflected power or minimum SWR at the driver.

$$Rs = \frac{\text{Anode supply voltage} - \text{screen voltage}}{\text{maximum screen-current}}$$

Rk: cathode RF negative feedback resistor, typically 5Ω to 20Ω; resistance = excess grid drive peak volts divided by peak cathode current. Rk reduces intermodulation distortion and makes the tube harder to over-drive. Do not bypass the cathode. Use one cathode RFC and one Rk for each tube.

▪▪ Neutralization: When L1 is tuned for ≈zero reflected power, adjust Cn for minimum RF at the anode.{Use a suitable terminating resistance in place of the tuned output circuit} 15m is a good band for neutralization. This completes the neutralization adjustment. No further adjustment is required.

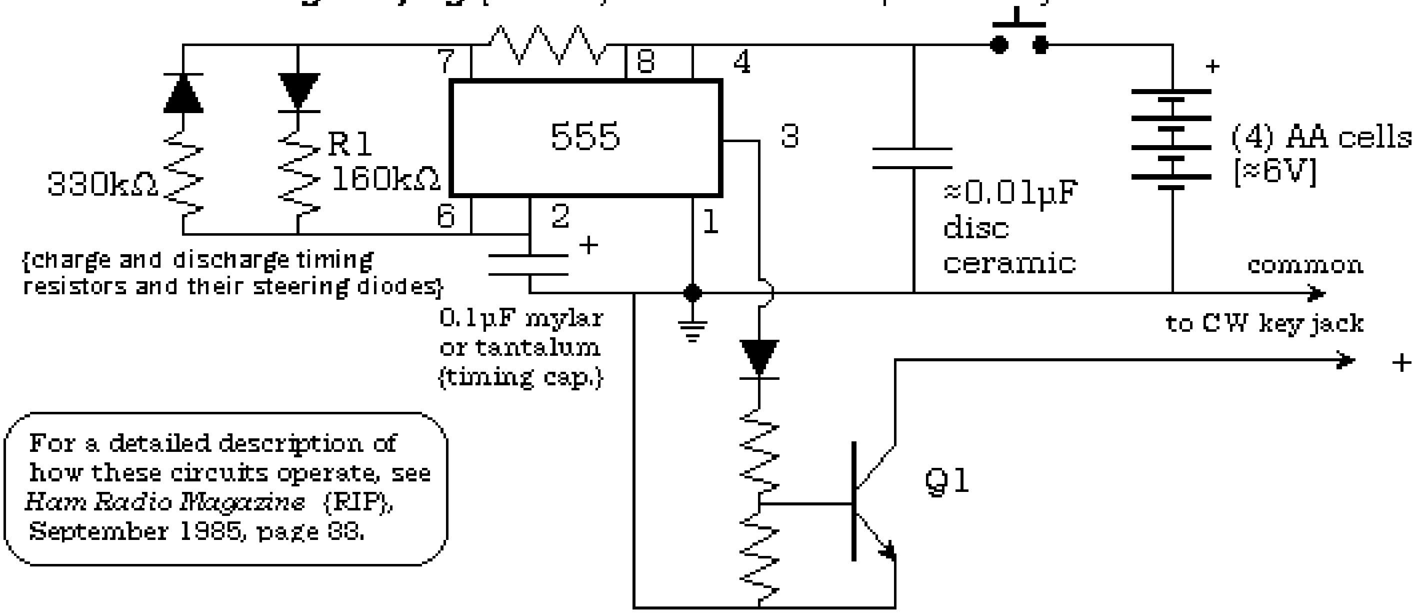
▪▪ To reduce the chances of a heater to cathode arc, connect one side of the heater to the cathode, feed the heater through a ≈40μH bifilar choke. Do not ground either side of the heater.

Rich, AG6K, [805] 386-3734

Tuning Pulser for SSB Amplifiers

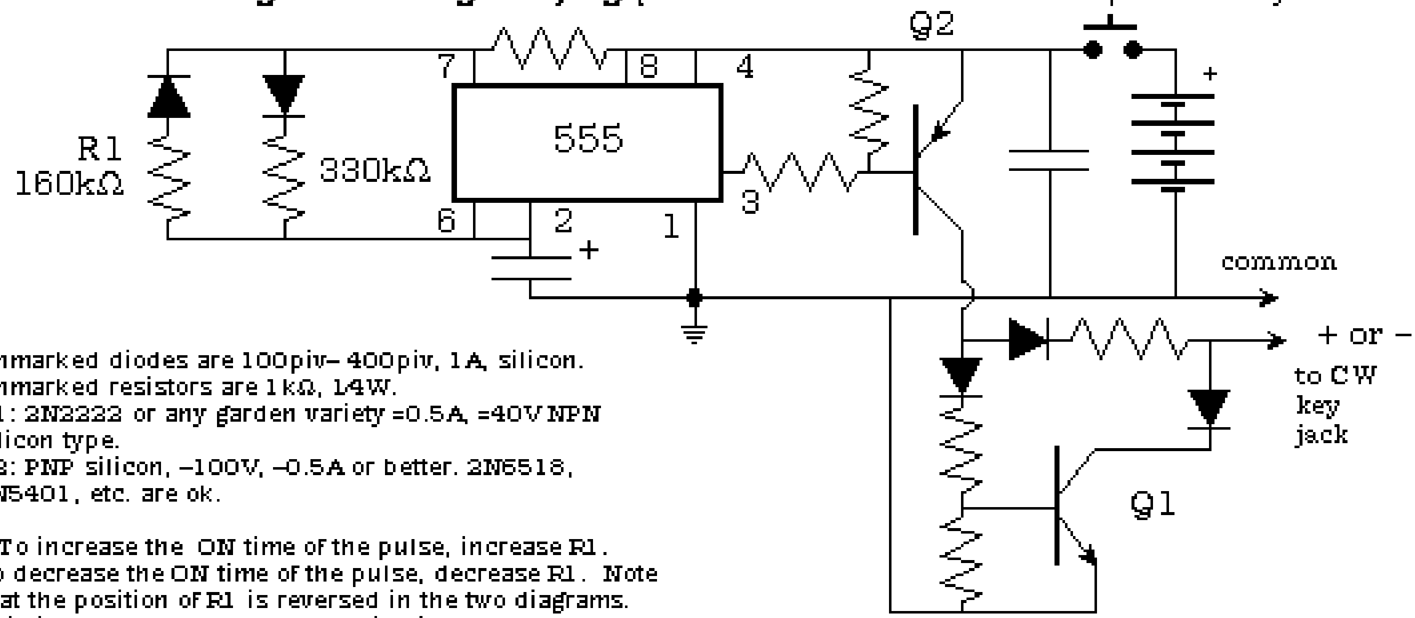
Richard L. Measures, AG6K

For Positive Voltage Keying *[virtually all transistor-output radios]:*



For a detailed description of how these circuits operate, see *Ham Radio Magazine* (RIP), September 1985, page 33.

For Positive or Negative Voltage Keying *[TS–830S and other tube-output radios] :*



Unmarked diodes are 100piv– 400piv, 1A, silicon.
Unmarked resistors are 1kΩ, 1⁄4W.
Q1: 2N2222 or any garden variety ≈0.5A, ≈40V NPN silicon type.
Q2: PNP silicon, –100V, –0.5A or better. 2N6518, 2N5401, etc. are ok.

To increase the ON time of the pulse, increase R1. To decrease the ON time of the pulse, decrease R1. Note that the position of R1 is reversed in the two diagrams. This is due to the effect of Q2 which inverts the 555's output pulse in the circuit that is capable of switching positive or negative voltages.

Either one of these circuits produces a continuous pulse string with an ON time of ≈12mS and an OFF time of ≈24mS — for a duty cycle of 0.333. The resultant pulse rate is roughly 28 per second — fast enough to produce a fairly smooth meter indication during tune-up. This allows one to tune up a SSB amplifier safely and more linearly at the higher voltage–SSB tap.

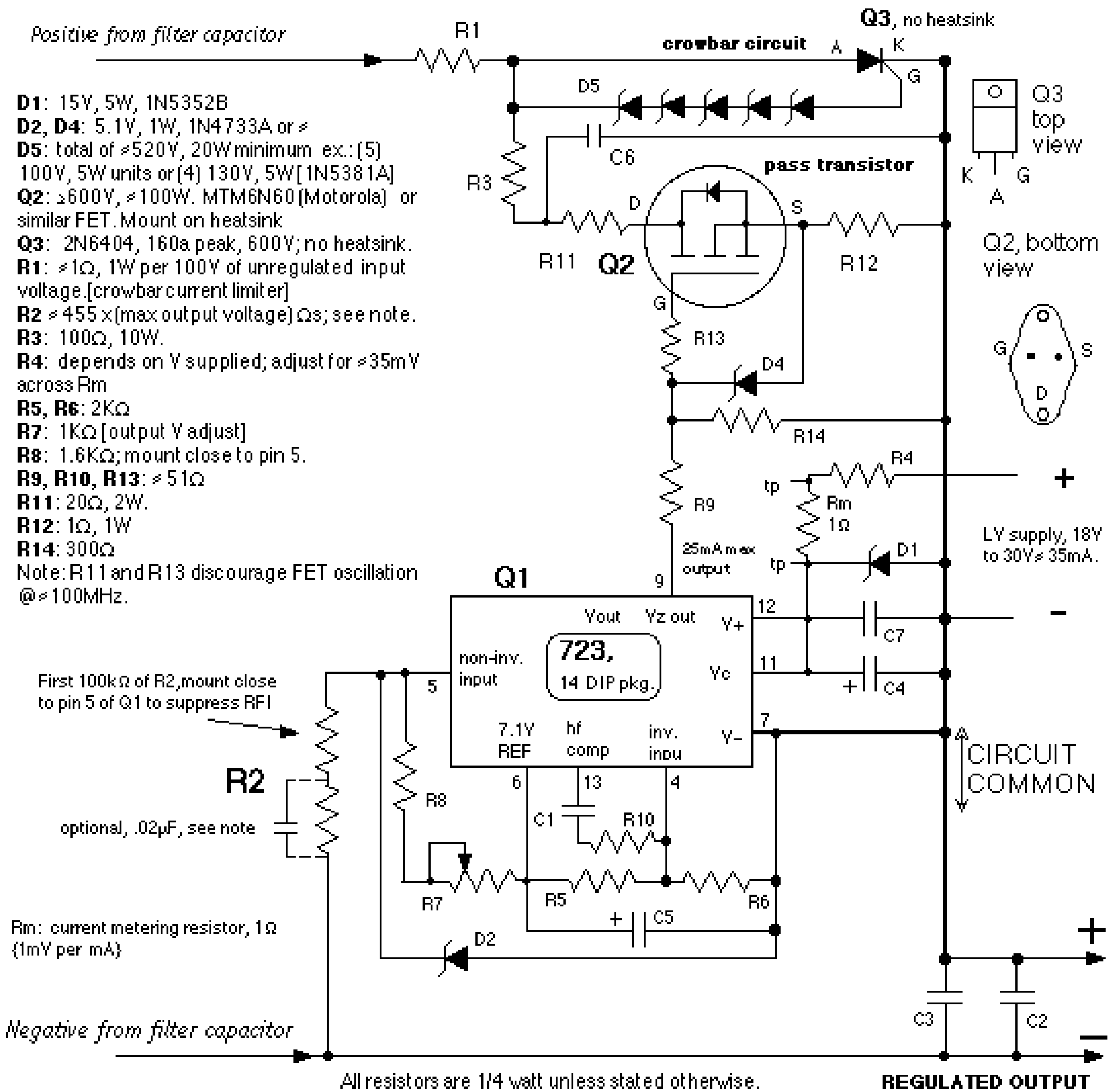
☛To tune-up any g–g amplifier:

1. The transceiver must be in CW mode, the amplifier is off.
2. Key the pulser and adjust the CW peak power for the *same peak power level that will be used on SSB.* Not all transceivers have adjustable peak power on SSB. For example, the TS–440S and TS–830S allow the operator to set the peak CW power but they do not have a means of setting the peak SSB power to anything other than maximum. If you own such a radio, adjust the CW power level until the ALC meter just begins to move. This indicates that the radio's maximum peak power capability has been reached.
3. To be on the safe side, unless you already know where to preset it, initially set the amplifier's Load control to about 25% of maximum capacitance [75% of maximum loading]. This assures that the grid–current will not run amuck during the initial tune-up. Switch the amplifier on.
4. Key the pulser. Alternately adjust the amplifier's Tune and Load controls for maximum power output. A relative power indicator is all that's needed.
5. When you are fairly sure that the maximum relative power output has been achieved, the linearity of the amplifier may be further improved if you adjust the Load capacitor for slightly more loading —enough to cause the power indicator to drop about 2%. More loading usually, *but not always*, means a higher number on the Load capacitor's panel scale. If you are not sure if this convention was followed in your amplifier, maximum loading is minimum C and minimum loading is maximum C.

Note: When the pulser is in use, the amplifier's I meters will indicate only about a third of the actual I. For example, if the grid–I meter indicates 50mA with the pulser on, the real grid–I is: 3 x 50mA=150mA.

Figure 8 723-based screen V regulator, 2kV maximum. This circuit is short-proof and RF-proof. Voltage regulation: ±0.1%.

D1: 15V, 5W, 1N5352B
D2, D4: 5.1V, 1W, 1N4733A or ≈
D5: total of ≈520V, 20W minimum ex.: (5) 100V, 5W units or (4) 130V, 5W [1N5381A]
Q2: ≥600V, ≈100W. MTM6N60 (Motorola) or similar FET. Mount on heatsink
Q3: 2N6404, 160a peak, 600V; no heatsink.
R1: ≈1Ω, 1W per 100V of unregulated input voltage. [crowbar current limiter]
R2 ≈ 455 x (max output voltage) Ωs; see note.
R3: 100Ω, 10W.
R4: depends on V supplied; adjust for ≈35mV across Rm
R5, R6: 2KΩ
R7: 1KΩ [output V adjust]
R8: 1.6KΩ; mount close to pin 5.
R9, R10, R13: ≈ 51Ω
R11: 20Ω, 2W.
R12: 1Ω, 1W
R14: 300Ω
Note: R11 and R13 discourage FET oscillation @≈100MHz.



All resistors are 1/4 watt unless stated otherwise.

C1: 1000pF 50V
C2: ≈2uF
C3: ≈.01µF
C4, C5: 1µF, 35V tantalum.
C6: .047µF, 1000V
C7: .01µF, ≥50V

About 90% of R2 can be bypassed as shown with 0.02µF, to improve the regulator's response to load transients and to reduce noise and ripple.
The **voltage** rating of R2 must be ≥ the output voltage.
☞*Caution*: If this supply will be used with a *grounded screen circuit* Class AB1 amplifier, the + circuit common is grounded, so the LV supply does not require special insulating ability. However, if this regulator will be used with an conventional, elevated screen circuit, the –output is grounded and the circuit common of this regulator is elevated with respect to chassis ground. In this case, the LV supply must be able to float at the full output voltage of the screen supply.
Note: The unregulated input should be filtered by a C filter or filtered by a tuned-choke/C filter. A non-resonant choke/C filter is not satisfactory.

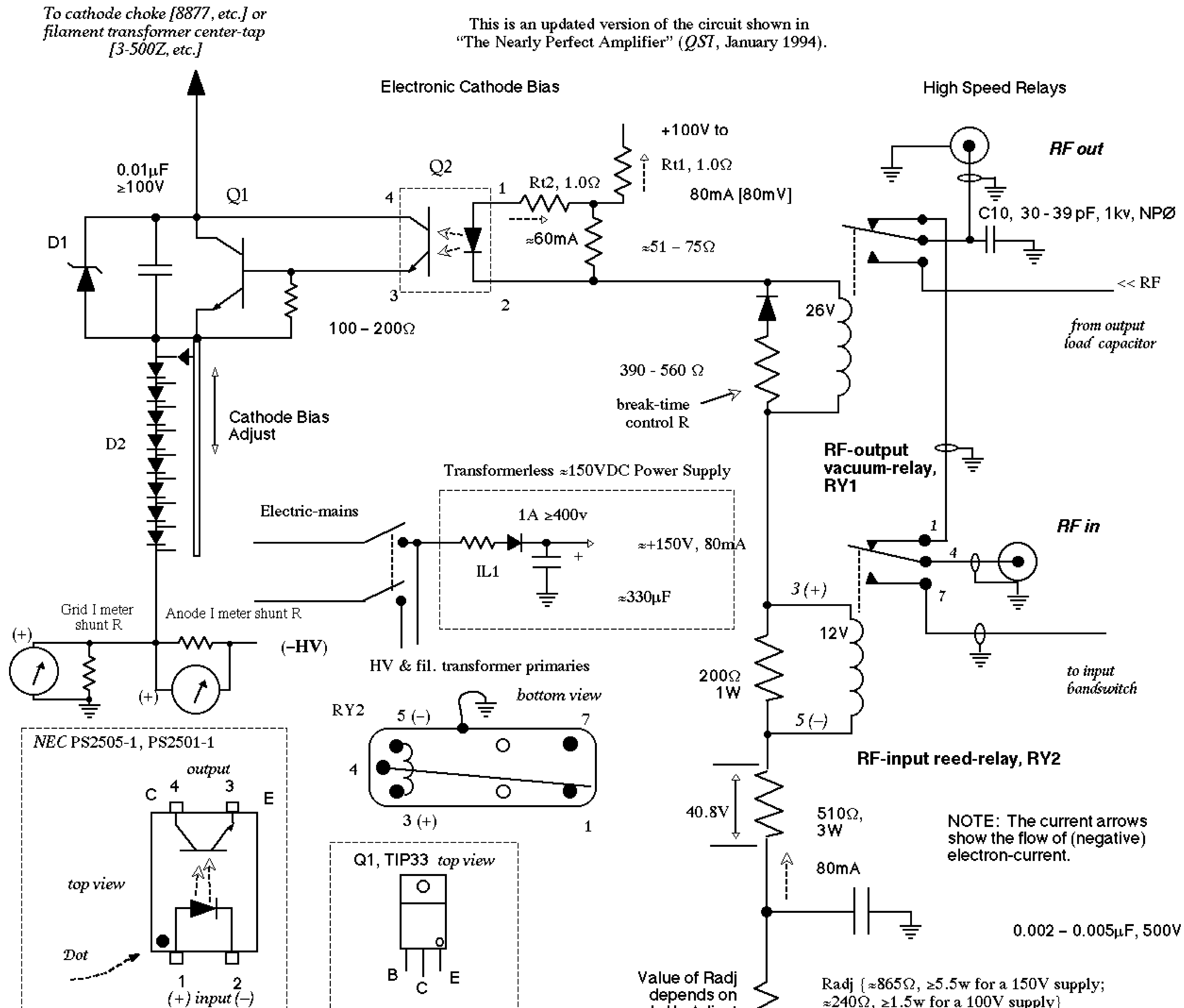
This circuit is based on National Semiconductor's® 'Positive Floating Regulator' for the LM723 in their *Linear Databook*.

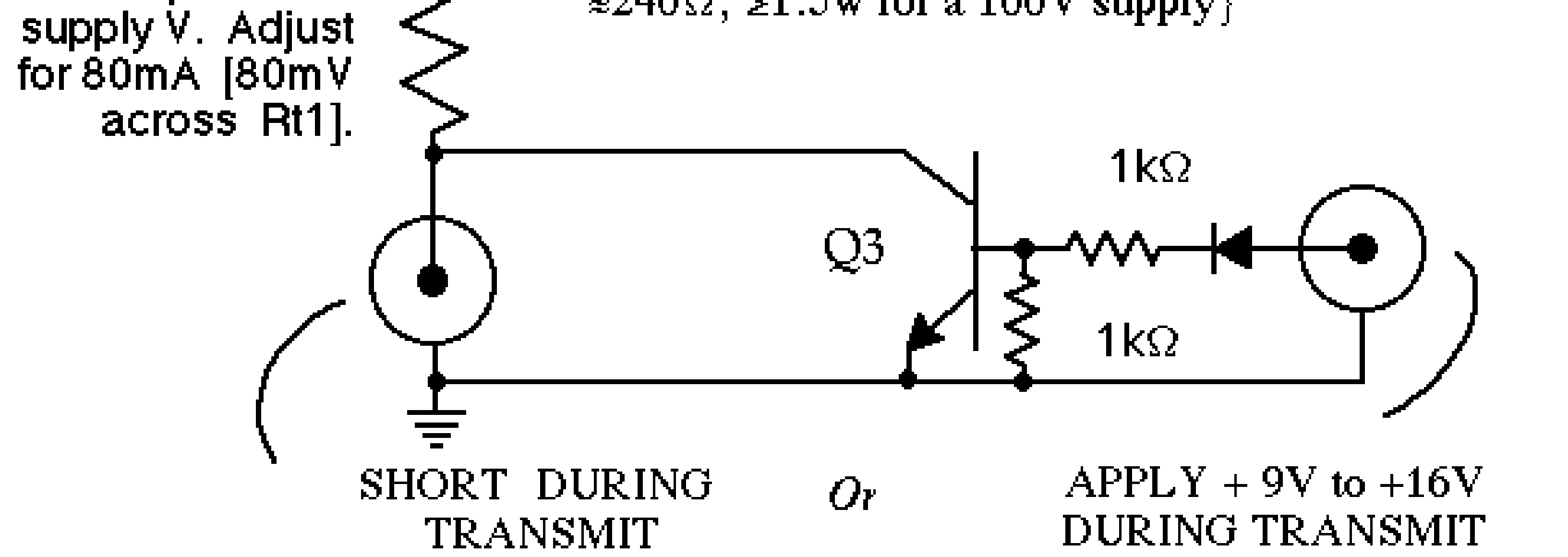
Rich, AG6K

Richard L. Measures, AG6K, ☎805 386 3734

High Speed Switching (≈2mS) for G-G Amplifiers

This is an updated version of the circuit shown in "The Nearly Perfect Amplifier" (*QST*, January 1994).



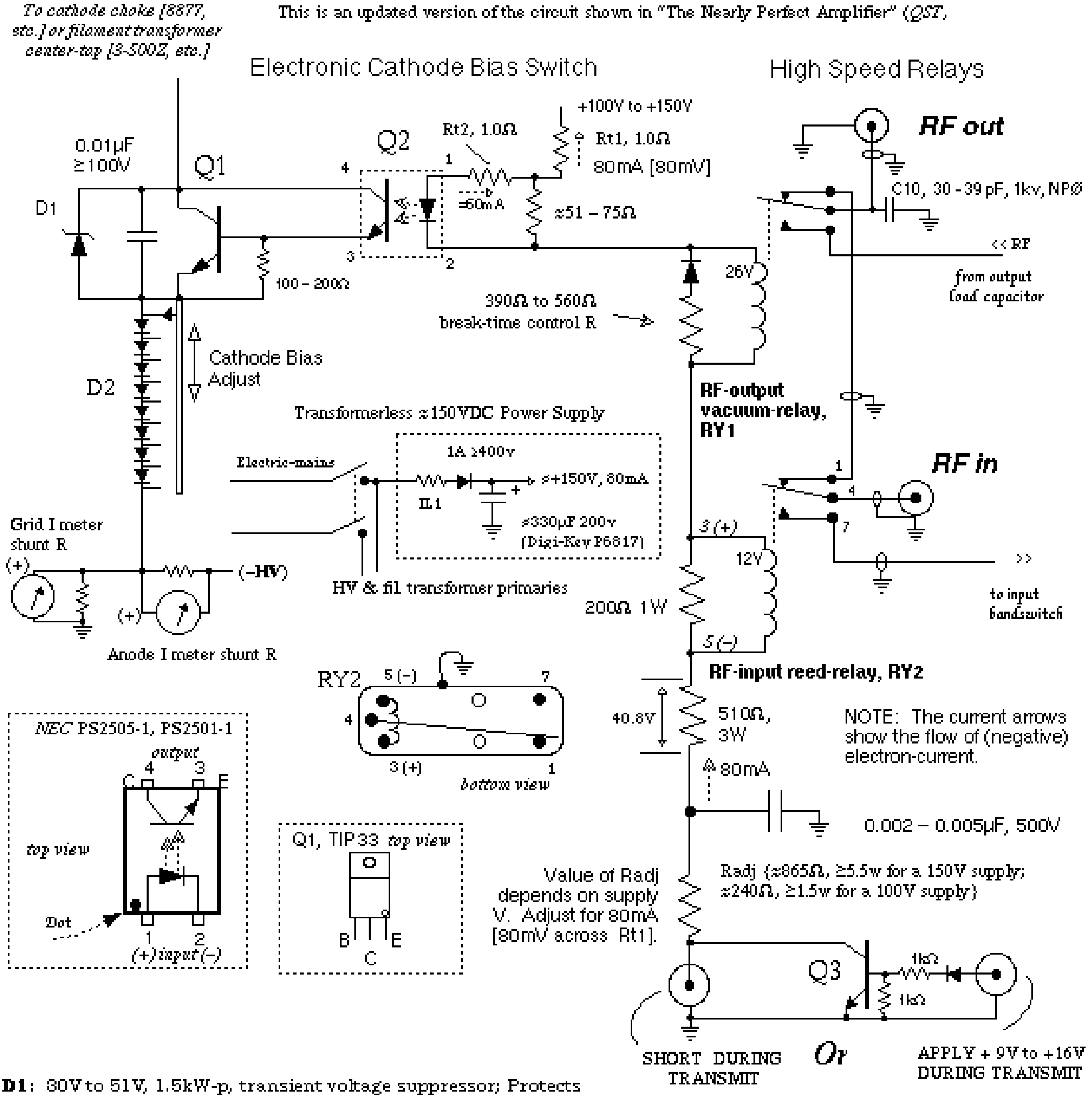


D1: 30V to 51V, 1.5kW-p, transient voltage suppressor; Protects Q1 and Q2; Digi-Key (✆1-800-344-4539) part #1.5KE51CA-ND or similar. D1 may short during a glitch – so it should be mounted on solder posts in an accessable place.
D2: Any piv, current rating ≥cathode-I 3A maximum. # of diodes depends on tube and anode supply V.
IL1: Keystone inrush I limiter, ≈10Ω cold, Digi-Key KC012L-nd or similar.
Q1: NPN, ≥80V, 10A, TIP33B, C, etc. mounted on small heatsink. This device can reliably switch cathode currents up to 3 amperes.
Q2: *NEC* PS2505-1 or PS2501-1 optoisolator, 80mA MAXIMUM input, 80VCE, gain ≥300%, 6µS or faster.
Q3: 300V, 0.5A, MPSA42 or =.
RY1: Kilovac HC-1 or Jennings RJ-1A {7A max @32MHz}. 26v coil. Shock-mount with silicone-rubber; do NOT use factory hardware for mounting; ground frame with copper foil ribbon. For extra contact life with heavy CW use, use Jennnings RJ-1H or Kilovac HC-3, which have tungsten contacts—and less power handling ability on 15m - 10m. The RJ-1A and the HC-1 use copper contacts.
RY2: RF reed-relay. Matsushita RSD-12v. Shock-mount with silicone-rubber; ground frame. Note: pin 3 [coil +] must be positive with respect to pin 5 [coil –].
NOTES: Unspecified diodes are ≥1A, ≥100PIV. ••The number of diodes in the Cathode Bias Adjust circuit, D2, depends on the bias requirement of the amplifier. ••The two 1.0Ω resistors, Rt1 and Rt2, are for measuring current with a DVM. 1mV=1mA. ••The loop current must be set to 80 – 84 mA with Radj. To be safe, jumper pins 1-2 of Q2 until the loop current has been adjusted to a safe value. ••For long life, the optoisolator input current should not exceed 65mA.
☞ ***Warning: ≥100V @ 0.080A can be lethal.***
I sell parts for this circuit. ☎805-386-3734.

NOTES, continued: ••The collector-to-emitter V drop [Vce] across Q1 is ≤1.4V on transmit. On receive, Vce depends on the supply V to the tubes and on the particular tubes used—typically Vce on receive is ≈22V. ••C10: Adjust for best 10m SWR with amplifier off using a 50Ω termination. ••The Transformerless 150VDC supply should not be used with a ground fault interrupter type circuit breaker. ••All RF connections to the relays are made with copper foil ribbons. Do not solder coax directly to the relays.
••**Vacuum Relays**, used RJ-1A or HC-1 26v: RF Parts, ☎ 619 744 0700; Fair Radio Sales ☎419-227-6573. New relays: ($120) Surcom ☎619 438 4420. All vacuum relays should be tested before installation for the presence of a vacuum. About a third of the used vacuum relays we test have bad vacuums. For these relays a leakage current of less than 2µA should flow with 3500VDC across open contacts. If the relay can not withstand at least 3500V, the relay has a bad vacuum and it may destroy RY2 when high power RF is present. Then you will need to replace two relays instead of just one.

High Speed Switching (≈2mS) for G-G Amplifiers

This is an updated version of the circuit shown in "The Nearly Perfect Amplifier" (*QST*,



D1: 30V to 51V, 1.5kW-p, transient voltage suppressor; Protects Q1 and Q2; Digi-Key (&1-800-344-4539) part #1.5KE51CA-ND or similar. D1 may short during a glitch, so it should be mounted on solder posts in an accessable place.

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Q1: NPN, ≥80V, 10A, TIP33B, C, etc. mounted on small heatsink. This device can reliably switch cathode currents up to 3 amperes.

Q2: *NEC* PS2505-1 or PS2501-1 optoisolator, 80mA MAXIMUM input, 80VCE, gain ≥300%, 6µS or faster.

Q3: 300V, 0.5A, MPSA42 or =.

RY1: Kilovac HC-1 or Jennings RJ-1A (7A max @32MHz), 26v coil. Shock-mount with silicone-rubber; do NOT use factory hardware for mounting; ground frame with copper foil ribbon. For extra contact life with heavy CW use, use Jennnings RJ-1H or Kilovac HC-3, which have tungsten contacts—and less power handling ability on 15m - 10m. The RJ-1A and the HC-1 use copper contacts.

RY2: RF reed-relay, Matsushita RSD-12v. Shock-mount with silicone-rubber; ground frame. Note: pin 3 [coil +] must be positive with respect to pin 5 [coil –].

NOTES: Unspecified diodes are ≥1A, ≥100PIV. ••The number of diodes in the Cathode Bias Adjust circuit, D2, depends on the bias requirement of the amplifier. ••The two 1.0Ω resistors, Rt1 and Rt2, are for measuring current with a DVM. 1mV=1mA. ••The loop current must be set to 80 – 84 mA with Radj. To be safe, jumper pins 1-2 of Q2 until the loop current has been adjusted to a safe value. ••For long life, the optoisolator input current should not exceed 65mA.

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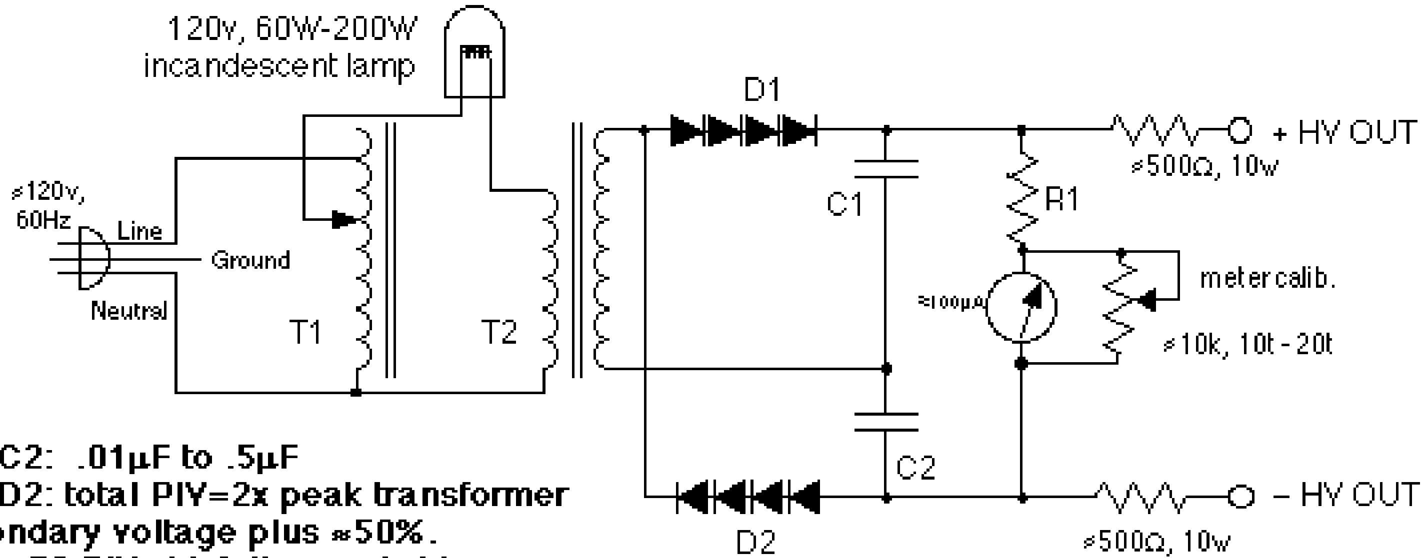
(–HV) on the TL-922 is the "IG" terminal on the X54-1300-10 circuit board. NOTE: An existing wire connects IG to the negative terminal of C26 / anode terminal of the original Zener cathode-bias diode, which is removed. The original Zener heatsink is adjacent to the filament transformer on the bottom of the chassis. Q1 is mounted to the vacated heatsink with an insulation kit.

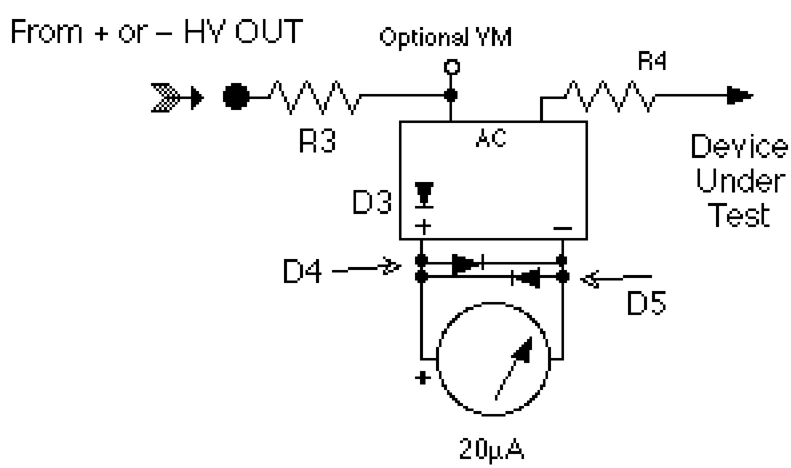
Figure 6

Breakdown Voltage Tester



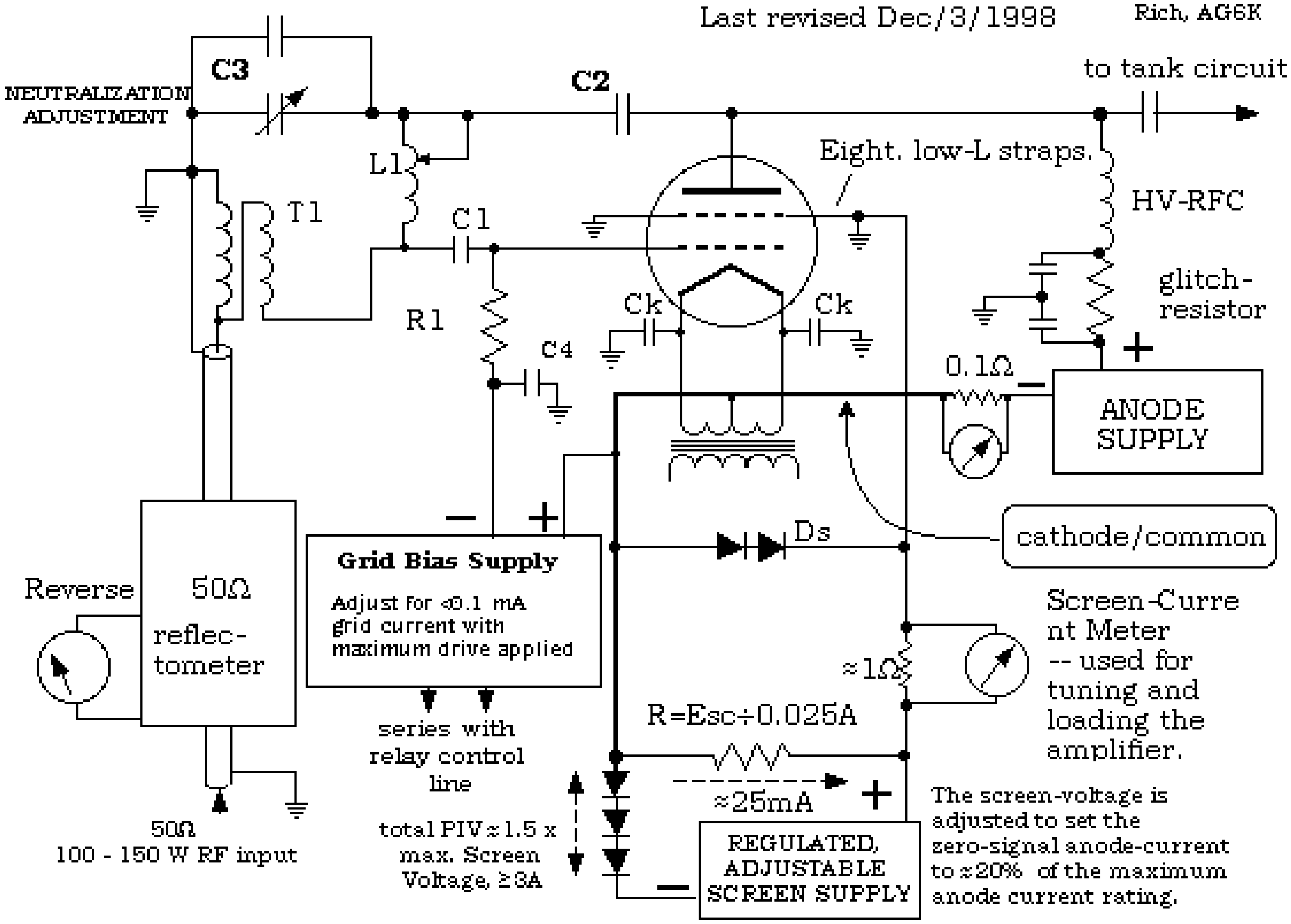
C1, C2: .01µF to .5µF
D1, D2: total PIV=2x peak transformer secondary voltage plus ≈50%.
D3: ≥50 PIV, 1A full-wave bridge.
D4, D5: ≥50 PIV, 3A.
R1: Resistance as needed - typically 10MΩ to 500MΩ, ≥5W, designed for HV [long, spiral-film type].
R3: ≈5MΩ to ≈50MΩ. Must be able to withstand the full output voltage.
R4: 100KΩ to 200KΩ, 2W.
T1: Variable transformer, ≥1A, 120V input, 0-132V output.
T2; 120V 50/60Hz primary, ≈5–10kVRMS secondary. A 5kV secondary will produce a DC output of about 15kV.
--Custom scale meters are available from Metermaster at [800]-962-8128 or [213]-685-4340. or from Metercraft at [714]-973-8378. The meters must be mounted on plexiglass or some other insulating surface. Metal chassis/panels are problematic in HVBTs.

CURRENT LIMITER-DETECTOR



5. General Coverage Grid-Driven Class AB1 Amplifier – Simplified Diagram

Last revised Dec/3/1998 Rich, AG6K



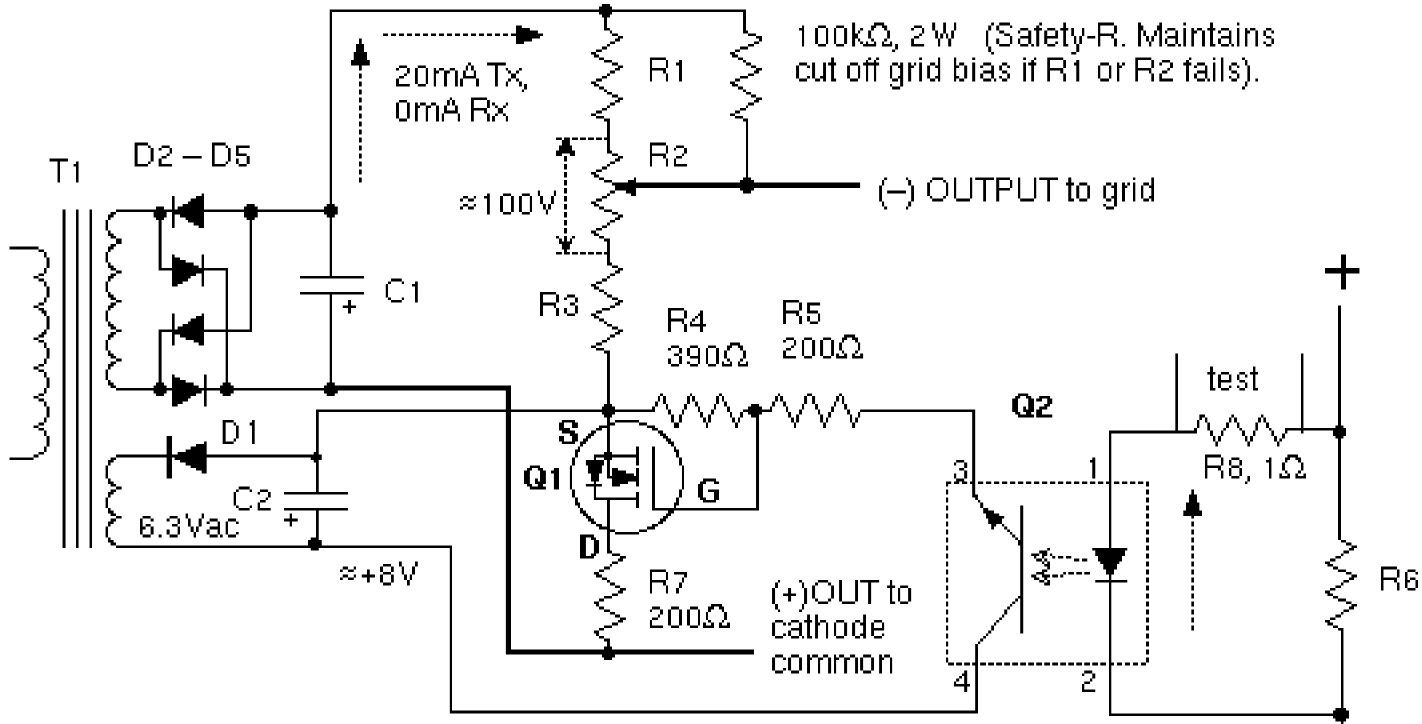
C1: ≈1nF [1000pF] @ 5KV for 80m to 10m. ≈2nF @5kV for 160m to 10m.
C2: ≈8 pF, rated at ≥3 x anode supply voltage/
C3: Neutralizing C. ≈1100 pF total. The variable C is ≈250pF, compression mica.
C4: ≈0.02uF in parallel with ≈2000pF.
R1: R1≈200Ω for bifilar T1; R1≈450Ω for trifilar T1. Must be able to dissipate the average power delivered by the driver.
L1: Sufficient inductance at the lowest operating frequency to resonate with the grid capacitance. Resonate at the operating frequency by tuning for minimum reflected power. This should be done with the anode and screen supplies OFF.
T1: Perm. = 40 [Amidon #67 material] 2.4 inch diameter. A typical bifilar wound T1 has about 12 turns of #20 – #22 teflon, stranded, hook-up wire for 1.8MHz to 29MHz operation. A bifilar [voltage-doubling] winding is used for amplifier tubes that require a grid voltage of approximately –200V [4CX1500A, 5CX1500A, 4-1000A and the 8169]. A trifilar [voltage-tripling] winding is used for tubes [8171] that require a grid voltage of approximately –300v. A typical trifilar winding would have 8-turns on the same type core. The bandwidth of trifilar windings is typically less than that of bifilar windings.

- In order to minimize inductance/phase-shift in the neutralization circuit's capacitive voltage-divider [C2 - C3], the anode connection to the amplifier-tube should be made with 2cm-3cm wide section of long-finger type, Eimac® finger-stock fastened to the upper rim of the fiberglass chimney. This point is also used for the output & DC connection. The "plate-cap" connection point is not used.
- The inductance of all leads in the neutralization circuit should be minimized. C2 should be placed between the finger-stock and a feedthrough insulator that is mounted next to the tube-socket. L1 and C3 should be mounted near the feedthrough insulator under the chassis. This is an important layout consideration for a neutralized AB1 circuit.

Ds: V-surge breakdown diodes: (hand picked from garden-variety rectifiers) Protects components which connect to cathode-common—such as the filament-transformer, grid bias supply, and the screen supply. Ds protects by short-circuiting during a +HV-to-chassis arc-over or an anode-to-screen arc-over. The measured PIV total for Ds should = the maximum screen-voltage plus about 500V. The current rating should be ≥3A [1N5400 family, etc.]. Since these diodes will short during a glitch, they should be mounted in an accessable place to facilitate replacement.
Ck: cathode bypass capacitors. Must be able to withstand the peak RF cathode current. Approximately 6 disc ceramic capacitors should be used for each Ck to share the RF-current burden and to mimimize L between the filament collets and ground. Typical values: 3kV - 6kV, 2000pF to 5100pF each. It's best to bypass with at least two different values of capacitors.

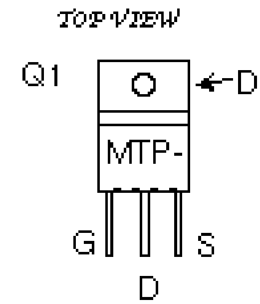
Figure 4

Solid-State Switched
GRID BIAS SUPPLY FOR CLASS AB1 AMPLIFIERS



C1: ≈20 µF, voltage rating ≥1.2 times the no load peak voltage of T1 secondary.
C2: ≈200µF, 16V
D1: ≥50piv, 1A, 1N4002 or similar.
D2, D3, D4, D5: PIV≥1.5 x peak secondary voltage of T1, ≥1A.
Q1: MOS FET power transistor, N channel, ≈1A, Vds≥ voltage rating of C1. The Motorola™ MTP3N80 is rated at 800V, 1.5A; no heatsink needed.
Q2: Optoisolator; *NEC* PS2505-1 or equivalent.
R1: Must be calculated: R1=[V on C1 @20mA load, minus (expected transmit grid bias V plus 50V)] ÷ .02A. Wattage≥ .02A x .02A x *Ωs*.
R2: 5KΩ, ≥2W.
R3: R3 is whatever resistance is needed to produce 100V across R2 (or a current flow of 20mA through R1 and R2).
R6: Shunts excess current around Q2. Resistance depends on relay coil current. [Q2 input current is measured at R8]. Adjust R6 for ≈20mA through R8/Q2. For a relay coil current of 80mA, R6≈1.5V÷(80mA – 20mA)≈25*Ωs*. The abs. maximum input current rating for Q2 is 80mA.
T1: Secondary RMS volts≈1.25 x the normal operating bias voltage needed. Example, for an operate bias requirement of –300V, the transformer should be 300 x 1.25 ≈ 375V RMS, with a current rating ≥50mA. In the above example the cutoff bias voltage is 375V x √2 ≈ –530V.
• Resistors are 0.25 watt unless specified otherwise.

How It Works: When Q1's gate is at zero V, Q1 is cut off (∞R). Q1 switches to a low resustance when a few V positive is applied to its gate. During receive, when no current passes through the amplifier's T/R relay coils, Q2 is off—so Q1 is off [∞R] and zero current flows through R1/R2/R3. Thus, the grid bias voltage is high no matter where R2 is set, cutting off anode current flow in the amplifier tube. On transmit, current passes through the T/R relay coils; Q2 turns on and applys +V to the gate of Q1. As a result, Q1 switches from ∞R to low R. This causes ≈20mA to flow through R1/R2/R3, creating a voltage drop in these resistors. Thus, during transmit, the grid bias voltage decreases and it is adjustable. The grid bias voltage should be adjusted so that virtually zero grid current flows during maximum drive.



Q2

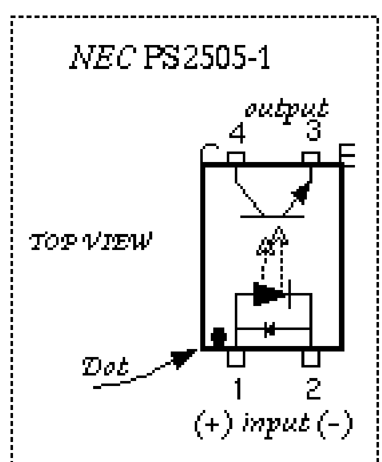
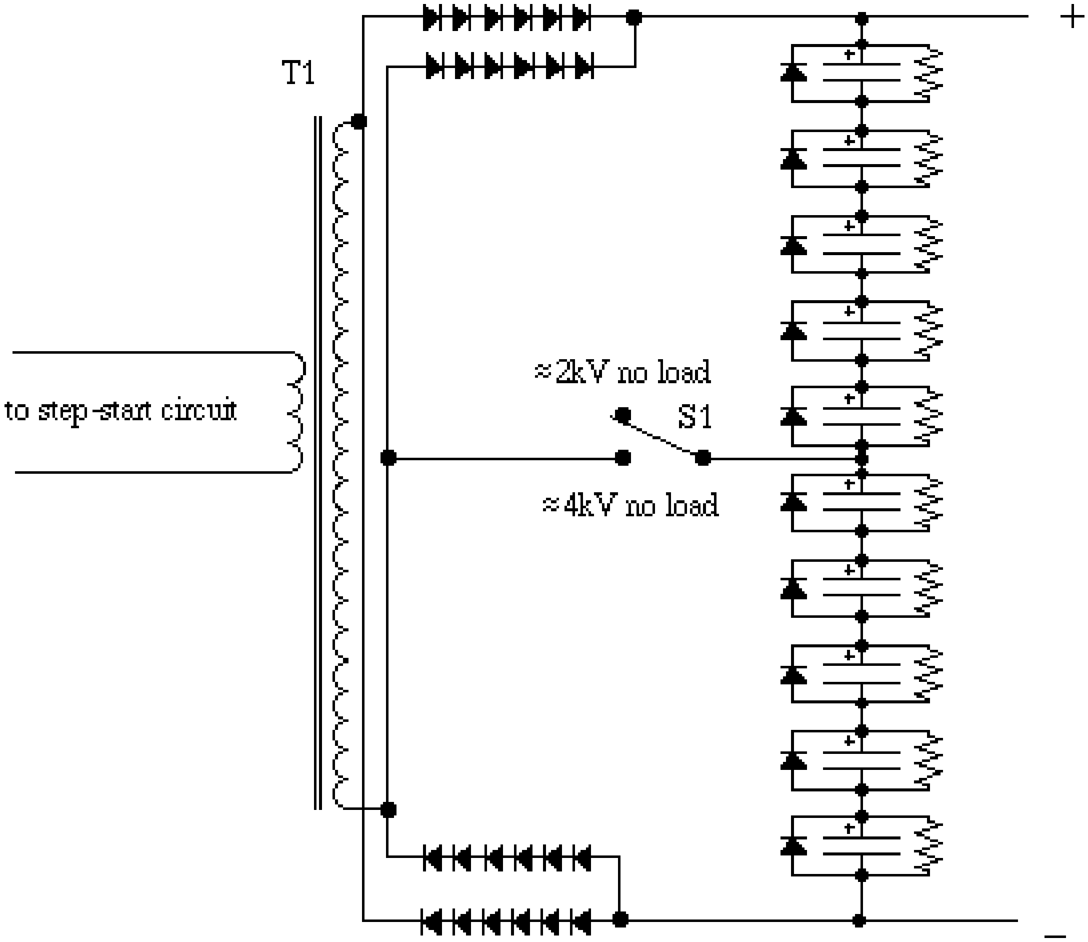


Figure 3

Example of a switchable
Dual Voltage Power Supply
using a transformer with no
secondary or primary taps.

Rich, AG6K, 3 Feb 1994



•All unmarked diodes are 1kV, 3A [1N5408 or =].
•All unmarked resistors are 100kΩ 3W MOF.
•All capacitors are ≈330µF 450VDC electrolytic

T1:unloaded secondary voltage is ≈1414VRMS [2kV peak].
S1: vacuum-relay or suitably insulated ceramic switch.

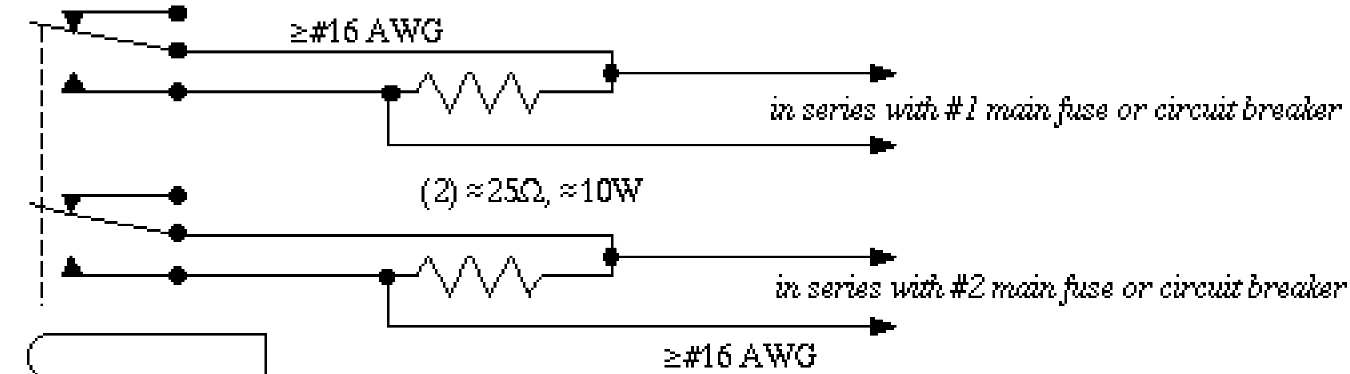
When S1 is open, the circuit operates as a fullwave bridge—producing about 2kV for CW, RTTY or FM operation. When S1 is closed, the circuit operates as a fullwave voltage doubler—producing about 4kV for SSB operation--but at a lower average current capability. Do not close S1 when the power is on.

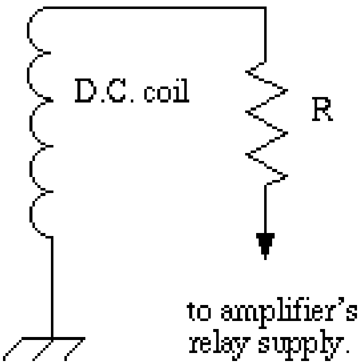
Figure 1

Step-Start

Richard L. Measures, AG6K. 6 Feb. 1994

Step-Start circuit for a typical 1500W dual-voltage operation [120V/240V] amplifier:





In effect, the timing capacitor for this step-start circuit is the HV filter capacitor.

The relay's coil voltage is approximately the same as the DC output voltage of the amplifier's relay supply. A contact rating of 10A is sufficient for ≈1500W amplifiers.

Adjust R for relay pull in when the HV reaches ≈2/3 of normal after turn-on. Typical pull in time is 1 second ±0.5 seconds. R is roughly half of the relay's coil resistance.