# Design Example: A High Power Vacuum Tube HF Amplifier Using the 8877 Triode

*Editor's note: Section and figure references in this article are from the 2013 edition of the ARRL Handbook. This material was originally contributed to the Handbook by John Stanley, K4ERO.*

Most popular HF transceivers produce approximately 100 W output. The EIMAC 8877 can deliver 1500 W output for approximately 60 W of drive when used in a grounded grid circuit. Grounded-grid operation is usually the easiest tube amplifier circuit to implement. Its input impedance is relatively low, often close to 50 Ω. Input/output shielding provided by the grid and negative feedback inherent in the grounded-grid circuit configuration reduce the likelihood of amplifier instability and provide excellent linearity without critical adjustments. Fewer supply voltages are needed in this configuration compared to others, often just high-voltage dc for the plate and low-voltage ac for the filament.

## 17.9.1 Tube Capabilities

The first step in the amplifier design process is to verify that the tube is actually capable of producing the desired results while remaining within manufacturer's ratings. The plate dissipation expected during normal operation of the amplifier is computed first. Since the amplifier will be used for SSB, a class of operation producing linear amplification must be used. Class AB2 provides a very good compromise between linearity and good efficiency, with effective efficiency typically exceeding 60%. Given that efficiency, an input power of 2500 W is needed to produce the desired 1500 W output. Operated under these conditions, the tube will dissipate about 1000 W — well within the manufacturer's specifications, provided adequate cooling airflow is supplied.

The grid in modern high-mu triodes is a relatively delicate structure, closely spaced to the cathode and carefully aligned to achieve high gain and excellent linearity. To avoid shortening tube life or even destruction of the tube, the specified maximum grid dissipation must not be exceeded for more than a few milliseconds under any conditions. For a given power output, the use of higher plate voltages tends to result in lower grid dissipation. It is important to use a plate voltage that is high enough to result in safe grid current levels at maximum output. In addition to maximum ratings, manufacturers' data sheets often provide one or more sets of "typical operation" parameters. This makes it even easier for the builder to achieve optimum results.

According to typical operation data, the 8877, operating at 3500 V, can produce 2075 W of RF output with excellent linearity and 64 W of drive. Operating at 2700 V it can deliver 1085 W with 40 W of drive. To some extent, the ease and cost of constructing a high-power amplifier, as well as its ultimate reliability, are enhanced by using the lowest plate voltage that will yield completely satisfactory performance. Working with various load lines on the characteristic curves shows that the 8877 can comfortably deliver 1.5 kW output with a 3100 V plate supply and 50 to 55 W of drive. Achieving 1640 W output power at this plate voltage requires 800 mA of plate current — well within the 8877's maximum rating of 1.0 A.

| Parameter | Result |
|---|---|
| Grid Current (mA) | 34.2 |
| Screen Current (mA) | 0.00 |
| Plate Current (A) | 0.800 |
| Input Power (Watts) | 2470 |
| Output Power (Watts) | 1640 |
| Plate Dissipation (W) | 831 |
| Plate Load (Ohms) | 2210 |
| Efficiency (%) | 66.3 |
| Grid Swing (Volts) | 80.0 |
| Resting Dissipation (W) | 620 |
| Input Resistance (Ohms) | 62.5 |
| Power Passed (Watts) | 48.8 |
| Total drive power (W) | 51.6 |

**Fig 17.35 — Typical operating parameters for the 8877 triode used in the design example detailed in the text.**

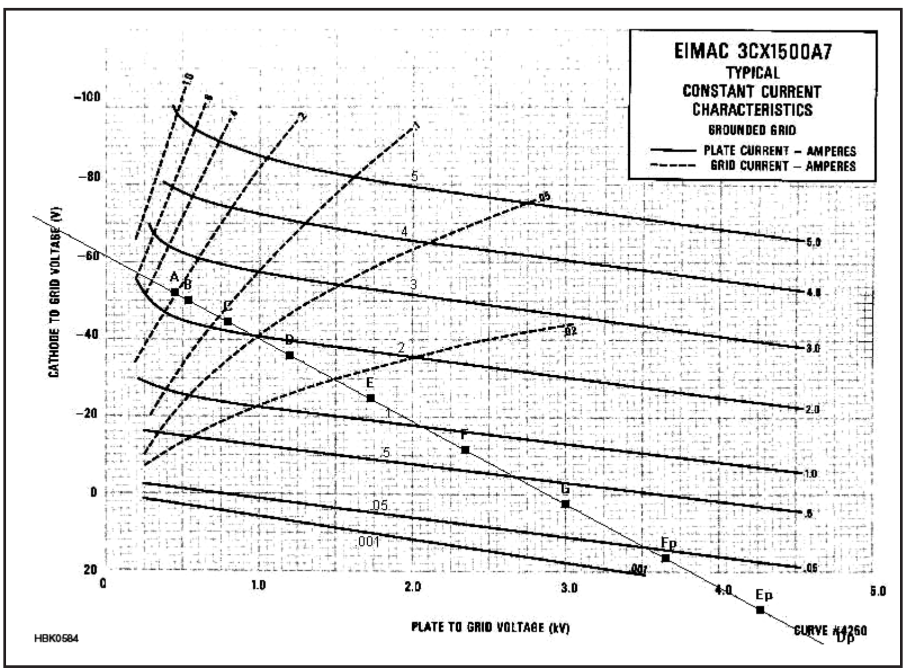


**Fig 17.34 — Characteristic curves for the 8877 triode used in the design example detailed in the text.**

## 17.9.2 Input and Output Circuits

The next step in the design process is to calculate the optimum plate load resistance at this plate voltage and current for Class AB2 operation and design an appropriate output-matching network. Using the operating line shown in **Fig 17.34**, the *TubeCalculator* program gives these values (**Fig 17.35**). The simple formula for load resistance shown earlier in this chapter (equation 1) gives a similar value, about 2200 Ω.

Several different output networks might be used to transform the nominal 50-Ω resistance of the actual load to the 2200-Ω load resistance required by the 8877, but experience shows that pi and pi-L networks are most practical. Each can provide reasonable harmonic attenuation, is relatively easy to build mechanically and uses readily available components. The pi-L gives significantly greater harmonic attenuation than the pi and usually is the better choice — at least in areas where there is any potential for TVI or crossband interference. In a multiband amplifier, the extra cost of using a pi-L network is the "L" inductor and its associated band switch section.

The input impedance of a grounded-grid 8877 is typically on the order of 50 to 55 Ω, shunted by input capacitance of about 38 pF. While this average impedance is close enough to 50 Ω to provide negligible input SWR, the instantaneous value varies greatly over the drive cycle — that is, it is nonlinear. This nonlinear impedance is reflected back as a nonlinear load impedance at the exciter output, resulting in increased intermodulation distortion, reduced output power, and often meaningless exciter SWR meter indications. In addition, the tube's parallel input capacitance, as well as parasitic circuit reactances, often are significant enough at 28 MHz to create significant SWR.

A tank circuit at the amplifier input can solve both of these problems by tuning out the stray reactances and stabilizing (linearizing) the tube input impedance through its flywheel effect. The input tank should have a loaded Q of at least two for good results. A Q of five results in a further small improvement in linearity and distortion, but at the cost of a narrower operating bandwidth. Using the *PI-EL Design* software, one can quickly determine values for C1, L1 and C2 for the various bands as well as the bandwidth that various values of Q will provide. Since the 3.5 to 4 MHz band is the widest in terms of percentage bandwidth, using a lower Q for that band seems wise. If we wish to cover two bands with the same switch position, for example the 24 and 28 MHz bands, that will also call for a lower Q. For the 40, 30 and 20 meter bands alone, a rather high Q would work fine. Remember to subtract the input capacitance of the tube from the calculated value of C2.

**Fig 17.36** illustrates these input and output networks applied in the amplifier circuit. The schematic shows the major components in the amplifier RF section, but with band-switching and cathode dc return circuits omitted for clarity. C1 and C2 and L1 form the input pi network. C3 is a blocking capacitor to isolate the exciter from the cathode dc potential. Note that when the tube's average input resistance is close to 50 Ω, as in the case of the 8877, a simple parallel-resonant tank often can successfully perform the tuning and flywheel functions, since no impedance transformation is necessary. In this case, it is important to minimize stray lead inductance between the tank and tube to avoid undesired impedance transformation.

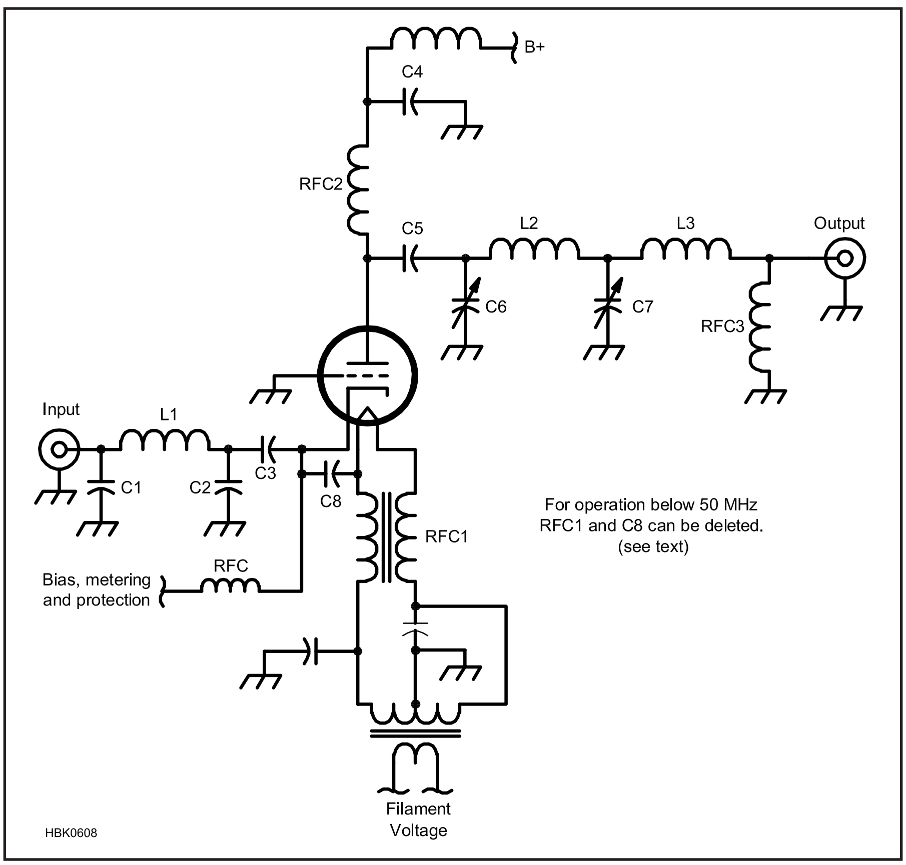


**Fig 17.36 — A simplified schematic of a grounded-grid amplifier using a pi network input and pi-L network output.**

## 17.9.3 Filament Supply

The filament or "heater" in indirectly heated tubes such as the 8877 must be very close to the cathode to heat the cathode efficiently. A capacitance of several picofarads exists between the two. Particularly at very high frequencies, where these few picofarads represent a relatively low reactance, RF drive intended for the cathode can be capacitively coupled to the lossy filament and dissipated as heat. To avoid this, above about 50 MHz, the filament must be kept at a high RF impedance above ground. The high impedance (represented by choke RFC1 in Fig 17.36) minimizes RF current flow in the filament circuit so that RF dissipated in the filament becomes negligible. The choke's low-frequency resistance should be kept to a minimum to lessen voltage drops in the high-current filament circuit.

The choke most commonly used in this application is a pair of heavy-gauge insulated wires, bifilar-wound over a ferrite rod. The ferrite core raises the inductive reactance throughout the HF region so that a minimum of wire is needed, keeping filament-circuit voltage drops low. The bifilar winding technique assures that both filament terminals are at the same RF potential.

Below 30 MHz, the use of such a choke seldom is necessary or beneficial, but actually can introduce another potential problem. Common values of cathode-to-heater capacitance and heater-choke inductance often are series resonant in the 1.8 to 29.7 MHz HF range. A capacitance of 5 pF and an inductance of 50 µH, for example, resonate at 10.0 MHz; the actual components are just as likely to resonate near 7 or 14 MHz. At resonance, the circuit constitutes a relatively low impedance shunt from cathode to ground, which affects input impedance and sucks out drive signal. An unintended resonance like this near any operating frequency usually increases input SWR and decreases gain on that one particular band. While aggravating, the problem rarely completely disables or damages the amplifier, and so is seldom pursued or identified.

Fortunately, the entire problem is easily avoided — below 30 MHz the heater choke can be deleted. At VHF-UHF, or wherever a heater isolation choke is used for any reason, the resonance can be moved below the lowest operating frequency by connecting a sufficiently large capacitance (about 1000 pF) between the tube cathode and one side of the heater. It is good practice also to connect a similar capacitor between the heater terminals. It also would be good practice in designing other VHF/UHF amplifiers, such as those using 3CX800A7 tubes, unless the builder can ensure that the actual series resonance is well outside of the operating frequency range.

### 17.9.4 Plate Choke and DC Blocking

Plate voltage is supplied to the tube through RFC2. C5 is the plate blocking capacitor. The output pi-L network consists of tuning capacitor C6, loading capacitor C7, pi coil L2, and output L coil L3. RFC3 is a high-inductance RF choke placed at the output for safety purposes. Its value, usually 100 µH to 2 mH, is high enough so that it appears as an open circuit across the output connector for RF. However, should the plate blocking capacitor fail and allow high voltage onto the output matching network, RFC3 would short the dc to ground and blow the power-supply fuse or breaker. This prevents dangerous high voltage from appearing on the feed line or antenna. It also prevents electrostatic charge — from the antenna or from blocking capacitor leakage — from building up on the tank capacitors and causing periodic dc discharge arcs to ground. If such a dc discharge occurs while the amplifier is transmitting, it can trigger a potentially damaging RF arc.

### 17.9.5 Tank Circuit Design

The output pi-L network must transform the nominal 50-Ω amplifier load to a pure resistance of 2200 Ω. We previously calculated that the 8877 tube's plate must see 2200 Ω for optimum performance. In practice, real antenna loads are seldom purely resistive or exactly 50 Ω; they often exhibit SWRs of 2:1 or greater on some frequencies. It's desirable that the amplifier output network be able to transform any complex load impedance corresponding to an SWR up to about 2:1 into a resistance of 2200 Ω. The network also must compensate for tube $C_{OUT}$ and other stray plate-circuit reactances, such as those of interconnecting leads and the plate RF choke. These reactances, shown in **Fig 17.37**, must be taken into account when designing the matching networks. Because the values of most stray reactances are not accurately known, the most satisfactory approach is to estimate them, and then allow sufficient flexibility in the matching network to accommodate modest errors.

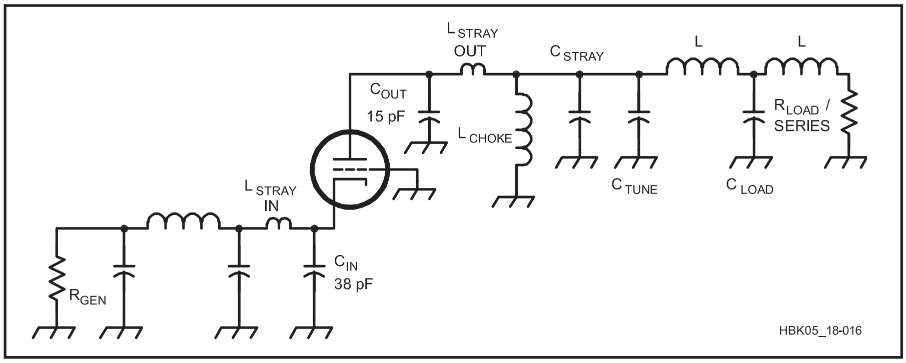


**Fig 17.37 — The effective reactances for the amplifier in Fig 17.36.**

Fig 17.37 shows the principal reactances in the amplifier circuit. $C_{OUT}$ is the actual tube output capacitance of 10 pF plus the stray capacitance between its anode and the enclosure metalwork. This stray C varies with layout; we will approximate it as 5 pF, so $C_{OUT}$ is roughly 15 pF. $L_{OUT}$ is the stray inductance of leads from the tube plate to the tuning capacitor (internal to the tube as well as external circuit wiring.) External-anode tubes like the 8877 have essentially no internal plate leads, so $L_{OUT}$ is almost entirely external. It seldom exceeds about 0.3 µH and is not very significant below 30 MHz. $L_{CHOKE}$ is the reactance presented by the plate choke, which usually is significant only below 7 MHz. $C_{STRAY}$ represents the combined stray capacitances to ground of the tuning capacitor stator and of interconnecting RF plate circuit leads. In a well-constructed, carefully thought out power amplifier, $C_{STRAY}$ can be estimated to be approximately 10 pF. Remaining components $C_{TUNE}$, $C_{LOAD}$, and the two tuning inductors, form the pi-L network proper.

The values for the output network components can be calculated using the *PI-EL Design* software, taken from the graphical charts in Figs 17.17 to 17.19, or from tables included on the *Handbook* CD. For pi networks, a Q of 12 is a good compromise between harmonic suppression and circuit losses. In practice, it often is most realistic and practical with both pi and pi-L output networks to accept somewhat higher Q values on the highest HF frequencies — perhaps as large as 18 or even 20 at 28 MHz. When using a pi-L on the 1.8 and 3.5 MHz bands, it often is desirable to choose a moderately lower Q , perhaps 8 to 10, to permit using a more reasonably-sized plate tuning capacitor.

#### CIRCUIT REACTANCES

The calculated output network values must be adjusted to allow for circuit reactances outside the pi-L proper. First, low-frequency component values should be examined. At 3.5 MHz, assuming that total tuning capacitance C1 is 140 pF, we know that three other stray reactances are directly in parallel with $C_{TUNE}$ (assuming that $L_{OUT}$ is negligible at the operating frequency as it should be). The tube's internal and external plate capacitance to ground, $C_{OUT}$, is about 15 pF. Strays in the RF circuit, $C_{STRAY}$, are roughly 10 pF.

The impedance of the plate choke, $X_{CHOKE}$, is also in parallel with $C_{TUNE}$. Plate chokes with self-resonance characteristics suitable for use in amateur HF amplifiers typically have inductances of about 90 µH. At 3.5 MHz this is an inductive reactance of +1979 Ω. This appears in parallel with the tuning capacitance, effectively canceling an equal value of capacitive reactance. At 3.5 MHz, an $X_C$ of 1979 Ω corresponds to 23 pF of capacitance — the amount by which tuning capacitor $C_{TUNE}$ must be increased at 3.5 MHz to compensate for the effect of the plate choke.

If the pi-L network requires an effective capacitance of 140 pF at its input at 3.5 MHz, subtracting the 25 pF provided by $C_{OUT}$ and $C_{STRAY}$ and adding the 23 pF canceled by $X_{CHOKE}$, the actual value of $C_{TUNE}$ must be 140 – 25 + 23 = 138 pF. It is good practice to provide at least 10% extra capacitance range to allow matching loads having SWRs up to 2:1. So, if 3.5 MHz is the lower frequency limit of the amplifier, a variable tuning capacitor with a maximum value of at least 150 to 160 pF should be used.

#### PERFORMANCE AT HIGH FREQUENCIES

Component values for the high end of the amplifier frequency range also must be examined, for this is where the most losses will occur. At 29.7 MHz we can assume a minimum pi-L input capacitance of 35 pF. Since $C_{OUT}$ and $C_{STRAY}$ contribute 25 pF, $C_{TUNE}$ must have a minimum value no greater than 10 pF. A problem exists, because this value is not readily achievable with a 150 to 160-pF air variable capacitor suitable for operation with a 3100 V plate supply. Such a capacitor typically has a minimum capacitance of 25 to 30 pF. Usually, little or nothing can be done to reduce the tube's $C_{OUT}$ or the circuit $C_{STRAY}$, and in fact the estimates of these may even be a little low. If 1.8 MHz capability is desired, the maximum tuning capacitance will be at least 200 to 250 pF, making the minimum-capacitance problem at 29.7 MHz even more severe.

There are three potential solutions to this dilemma. We could accept the actual minimum value of pi-L input capacitance, around 50 to 55 pF, realizing that this will raise the pi-L network's loaded Q to about 32. This results in very large values of circulating tank current. To avoid damage to tank components — particularly the band switch and pi inductor — from heat due to $I^2R$ losses, it will be necessary to either use oversize components or reduce power on the highest-frequency bands. Neither option is appealing.

A second potential solution is to reduce the minimum capacitance provided by $C_{TUNE}$. We could use a vacuum variable capacitor with a 300-pF maximum and a 5-pF minimum capacitance. These are rated at 5 to 15 kV, and are readily available. This reduces the minimum effective circuit capacitance to 30 pF, allowing use of the pi-L values for a Q of 12 on all bands from 1.8 through 29.7 MHz. While brand new vacuum variables are quite expensive, suitable models are widely available in the surplus and used markets for prices not much higher than the cost of a new air variable. A most important caveat in purchasing a vacuum capacitor is to ensure that its vacuum seal is intact and that it is not damaged in any way. The best way to accomplish this is to "hi-pot" test the capacitor throughout its range, using a dc or peak ac test voltage of 1.5 to 2 times the amplifier plate supply voltage. For all-band amplifiers using plate voltages in excess of about 2500 V, the initial expense and effort of securing and using a vacuum-variable input tuning capacitor often is well repaid in efficient and reliable operation of the amplifier.

A third possibility is the use of an additional inductance connected in series between the tube and the tuning capacitor. In conjunction with $C_{OUT}$ of the tube, the added inductor acts as an L network to transform the impedance at the input of the pi-L network up to the 2200-Ω load resistance needed by the tube. This is shown in **Fig 17.38A**. Since the impedance at the input of the main pi-L matching network is reduced, the loaded Q for the total capacitance actually in the circuit is lower. With lower Q, the circulating RF currents are lower, and thus tank losses are lower.

$C_{OUT}$ in Fig 17.38 is the output capaci-tance of the tube, including stray C from the anode to metal enclosure. $X_L$ is the additional series inductance to be added. As determined previously, the impedance seen by the tube anode must be a 2200 Ω resistance for best linearity and efficiency, and we have estimated C+ of the tube as 15 pF. If the network consisting of $C_{OUT}$ and $X_L$ is terminated at A by 2200 Ω, we can calculate the equiva-lent impedance at point B, the input to the pi-L network, for various values of series $X_L$. The pi-L network must then transform the nominal 50-Ω load at the transmitter output to this equivalent impedance.

### IMPEDANCE TRANSFORMATIONS

We work backwards from the plate of the tube towards the $C_{TUNE}$ capacitor. First, calculate the series-equivalent impedance of the parallel combination of the desired 2200-Ω plate load and the tube $X_{OUT}$ (15 pF at 29.7 MHz = $-j\,357$ Ω). The series-equivalent impedance of this parallel combination is $56.5 - j\,348$ Ω, as shown in Fig 17.38B. Now suppose we use a 0.5 μH inductor, having an impedance of $+j\,93$ Ω at 29.7 MHz, as the series inductance $X_L$. The resulting series-equivalent impedance is $56.5 - j348 + j93$, or $56.5 - j255$ Ω. Converting back to the parallel equivalent gives the network of Fig 17.38C: 1205 Ω resistance in parallel with $-j267$ Ω, or 20 pF at 29.7 MHz. The pi-L tuning network must now transform the 50-Ω load to a resistive load of 1205 Ω at B, and absorb the shunt capacitance of 20 pF.

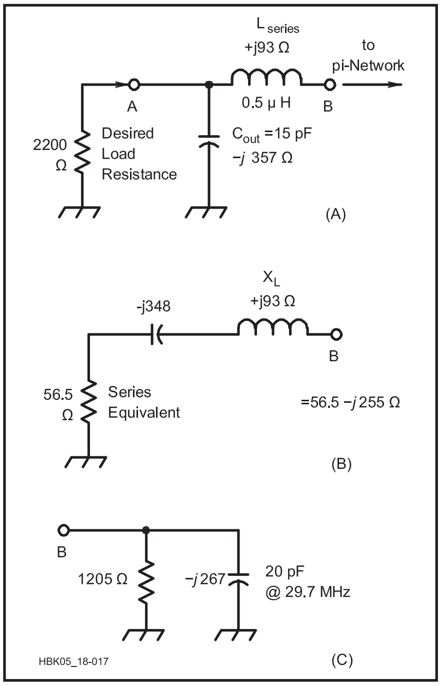


**Fig 17.38 — The effect of adding a small inductor in series with the tube plate to aid matching at high frequencies. See text for details.**

Using the *PI-EL Design* software or pi network formulas on the *Handbook* CD, R1 = 1205 Ω and Q = 15 at 29.7 MHz, yields a required total capacitance of about 67 pF at 29.7 MHz. Note that for the same loaded Q for a 2200-Ω load line without the series inductor, the capacitance was about 36 pF. When the 20 pF of transformed input capacitance is subtracted from the 67 pF total needed, the amount of capacitance is 47 pF. If the minimum capacitance in $C_{TUNE}$ is 25 pF and the stray capacitance is 10 pF, then there is a margin of 47– 35 = 12 pF beyond the minimum capacitance for handling SWRs greater than 1:1 at the load.

The series inductor should be a high-Q coil wound from copper tubing to keep losses low. This inductor has a decreasing, yet significant effect, on progressively lower frequencies. A similar calculation to the above should be made on each band to determine the transformed equivalent plate impedance, before calculating the network values. The impedance-transformation effect of the additional inductor decreases rapidly with decreasing frequency. Below 21 MHz, it usually may be ignored and pi-L network values calculated for R1 = 2200 Ω.

The nominal 90-μH plate choke remains in parallel with $C_{TUNE}$. It is rarely possible to calculate the impedance of a real HF plate choke at frequencies higher than about 5 MHz because of self-resonances. However, as mentioned previously, the choke's reactance should be sufficiently high that the calculations are not seriously affected if the choke's first series-resonance is at 23.2 MHz.

This amplifier is made operational on multiple bands by changing the values of inductance at L2 and L3 for different bands. The usual practice is to use inductors for the lowest operating frequency, and short out part of each inductor with a switch, as necessary, to provide the inductance needed for each individual band. Wiring to the switch and the switch itself add stray inductance and capacitance to the circuit. To minimize these effects at the higher frequencies, the unswitched 10-m L2 should be placed closest to the high-impedance end of the network at C6. Stray capacitance associated with the switch then is effectively in parallel with C7, where the impedance level is around 300 Ω. The effects of stray capacitance are relatively insignificant at this low impedance level. This configuration also minimizes the peak RF voltage that the switch insulation must withstand.

## 17.9.6 Checking Operation

After the input and output networks are designed, cold tuning as described earlier will confirm that all of the tuned circuits are working properly. These tests are well worth doing before any power is applied to the amplifier. The band switch itself will have significant inductance especially on the higher frequencies. To determine the proper taps for the various bands on the tank inductor, start with the highest frequency and verify that the calculated number of turns gives the frequency range desired, moving the tap as needed to allow for the inductance of the band switch. As each tap is located, it should be securely wired with strap or braid and the process repeated for successively lower bands. Once the cold tuning looks good, proceed to the tests for parasitics.

You should now be ready to apply full power to the amplifier and see how it performs.