

SECTION 3

ELECTRICAL DESIGN CONSIDERATIONS

3.1 CLASS OF OPERATION

Most power grid tubes used in AF or RF amplifiers can be operated over a wide range of grid bias voltage (or in the case of grounded grid configuration, cathode bias voltage) as determined by specific performance requirements such as gain, linearity and efficiency. Changes in the bias voltage will vary the *conduction angle* (that being the portion of the 360° cycle of varying anode voltage during which anode current flows.)

A useful system has been developed that identifies several common conditions of bias voltage (and resulting anode current conduction angle). The classifications thus assigned allow one to easily differentiate between the various operating conditions.

Class A is generally considered to define a conduction angle of 360° , class B is a conduction angle of 180° , with class C less than 180° conduction angle.

Class AB defines operation in the range between 180° and 360° of conduction. This class is further defined by using subscripts 1 and 2. Class AB₁ has no grid current flow and class AB₂ has some grid current flow during the anode conduction angle.

Example Class AB₂ operation - denotes an anode current conduction angle of 180° to 360° degrees and that grid current is flowing.

The class of operation has nothing to do with whether a tube is grid-driven or cathode-driven. The magnitude of the grid bias voltage establishes the class of operation; the amount of drive voltage applied to the tube determines the actual conduction angle.

The anode current conduction angle will determine to a great extent the overall anode efficiency. Anode efficiency in an amplifier is defined as the conversion of dc power into useful output power. Any power not delivered to the load is given off as heat, which must be removed from the anode to prevent damage to the tube.

System efficiency is often an important factor, particularly if prime power is obtained from a non-stationary source such as a diesel generator used in mobile systems. In applications where a power tube is operated continuously over a long period of time, improving the anode efficiency a few percent will result in considerable savings. Conversely, a reduction in anode efficiency not only raises operating costs for anode power to the tube but may also increase the cooling requirements to the extent that a larger blower motor (or water pump in the case of water-cooled tubes) is required, further raising operating costs.

Heat exchangers are needed to cool most power grid tubes, thermal energy from the anode must be removed by means of an air mover (blower), vapor-phase cooling system, water cooling, thermal link (i.e. conduction cooled) or by radiation (normally employed only in tubes with glass envelopes). Figure 23 illustrates the theoretical efficiency attainable with a tuned or resistive load, assuming the peak ac anode voltage is equal to the dc anode supply voltage.

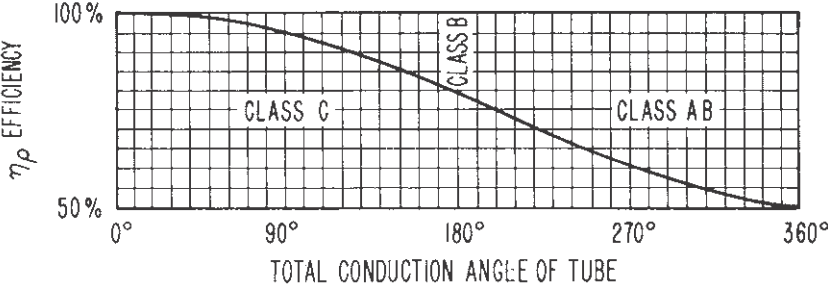


Figure 23. Anode efficiency vs. Conduction angle for an amplifier with tuned load. (From Thomas L. Martin, Jr., "Electronic Circuits," p. 452, Prentice-Hall, Inc., Englewood Cliffs, N.J., 1955).

A Class A power amplifier is used in applications requiring large amounts of low harmonic distortion power output. Class A amplifiers can be operated with very low intermodulation distortion in linear RF amplifier service (see section 4). Typical anode efficiency for a Class A amplifier is about 30 percent. The power gain is quite high due to the very low drive power required. Gains as high as 30 dB are typical. Class A service is widely used in audio amplifier and regulator service.

A Class AB power amplifier is capable of generating more power, with the same tube, than the Class A amplifier, but more

intermodulation distortion is generated at the same time. A Class B RF linear amplifier will generate still more intermodulation distortion, but is acceptable in certain applications. The anode efficiency is typically 66 percent, and stage gain is about 20 to 25 dB. Class B is used widely in audio amplifier applications, such as modulators for high level amplitude modulation.

Class C power amplifiers are used where large amounts of r-f power are to be generated with high efficiency. A Class C amplifier operates much like a limiter; therefore, it cannot amplify a modulated driver without serious distortion. Class C amplifiers are used for high level amplitude modulation wherein the anode voltage, or anode and screen voltage for tetrodes, is varied at an audio rate. Class C amplifiers must be used with tuned circuits or with a commutating output circuit with filtering. Class C cannot be used in a normal audio amplifier circuit. Class C amplifiers are also typically used in FM transmitters and in industrial applications, where large amounts of RF power are needed, but linearity is not an important factor

Class D⁴ describes a non-linear switch, where a square wave is applied to the grid and the anode alternates between non-conduction and saturation. Rise and fall times should be as fast as possible to ensure lowest losses. Practical anode efficiencies of an amplifier operating in Class D may exceed 85%, but significant harmonic energy is generated, thereby, raising filter requirements.

This class of operation requires the lowest possible saturation voltage. Therefore, significant grid current may be required while the tube is in conduction. A triode would be preferred over a tetrode because of its lower $e_{b_{min}}$.

3.2 TUBE PERFORMANCE COMPUTER FOR R-F AMPLIFIERS

It is quite easy to closely estimate the performance of a vacuum tube in radio-frequency power amplifier service, or an approximation in the case of harmonic amplifier service. Such estimates will give RF output power, dc input power, grid driving power, and all dc current values.

⁴ F.H. Raab, et al, "Power Amplifiers and Transmitters for RF and Microwave", IEEE Trans. Microwave Theory Tech., vol. 50, no. 3, pp 814-826, March 2002.

These estimates can be easily made by using the EIMAC Tube Performance Computer. This can be obtained at no cost by writing to: Application Engineering Department, CPI, EIMAC Division, 301 Industrial Road, San Carlos, CA 94070 USA. The computer is used with the characteristic curves of the tube, as plotted on the anode voltage/grid voltage coordinates (constant current curves).

By graphically laying out the trace of the anode and grid voltages as they rise and fall about the applied dc anode voltage and dc grid bias, a clear understanding of the action taking place within a tube is possible. With such an understanding, the operating conditions can be readily altered to suit individual requirements.

3.2.1 Simple Action in Class C RF Amplifiers

In an amplifier varying voltage is applied to the control grid of the tube. Simultaneously the anode voltage will vary in a similar manner, due to the action of the amplified current flowing in the anode circuit. In radio-frequency applications with resonant circuits, these voltage variations are smooth sine-wave variations, 180° out of phase (as the grid voltage rises and becomes **more** positive, the anode voltage falls and becomes **less** positive), as indicated in Figure 24. Note how these variations center about the dc anode voltage and the dc control grid bias.

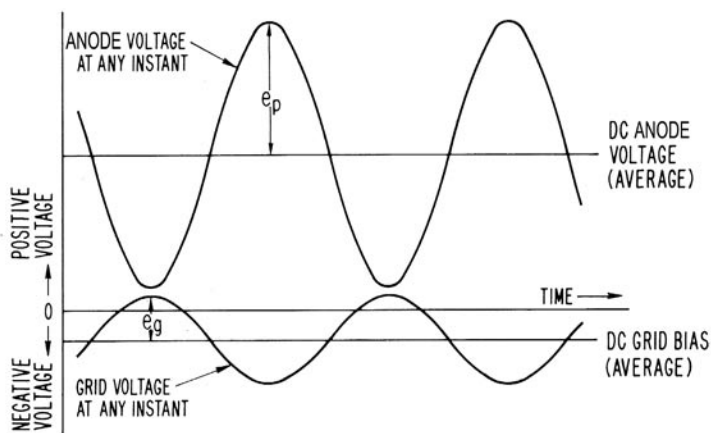


Figure 24. Variation of anode and grid voltage

Now let us see how such variations of the anode and grid voltages of a tube appear on the constant-current curves of a tube. In Figure 25 these variations have been indicated next to the anode voltage and grid voltage scales of a typical constant current curve. At some instant of time, shown as "t" on the time scales, the grid voltage has a value which is the point marked "e_g" on the grid-voltage sine wave. If one finds the point on the tube curves corresponding to these values (where a line drawn from "e_g" and a line drawn from "e_p" cross) he will be at Point A in Figure 25. As the values of the grid voltage "e_g" and anode voltage "e_p" vary over the RF cycle, Point A moves up and down a line, which in the case of the normal RF power amplifier is a straight line. This line is called the "Operating Line."

Any point on the operating line (when drawn on constant current curves as illustrated in Figures 25 or 27) tells the instantaneous values of anode current, screen current, and grid current which must flow when these particular values of grid and anode voltage are applied to the tube. Thus, by reading off the values of the currents and plotting them against time "t", one can obtain a curve of instantaneous values of anode and grid current (Figure 26).

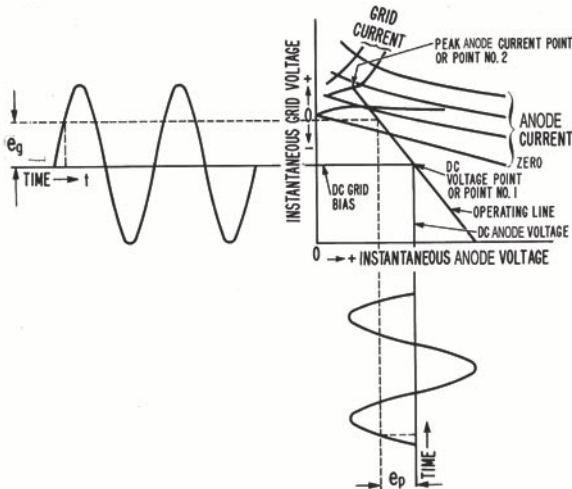


Figure 25. Anode voltage and grid voltage peak scales

If we analyze the anode and grid current values shown, we can predict that they will cause a dc ammeter to show a particular reading. This is called the dc component of the current. Also, we can predict that if the anode current flows through a properly-loaded resonant RF circuit, a certain amount of radio-frequency power

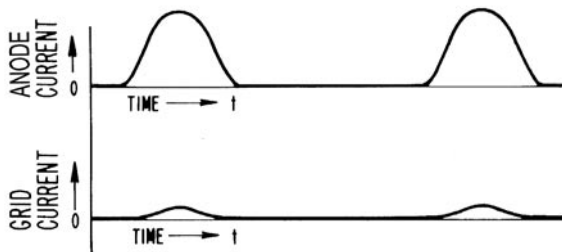


Figure 26. Instantaneous values of anode and grid current.

will be delivered to that circuit. If the circuit is tuned to the fundamental frequency (same frequency as the RF grid voltage), the power delivered will be due to the fundamental, or principal, radio-frequency component of anode current. If the circuit is tuned to a harmonic of the grid-voltage frequency, e.g., two or three times the frequency, the power delivered will be due to a harmonic component of the anode current.

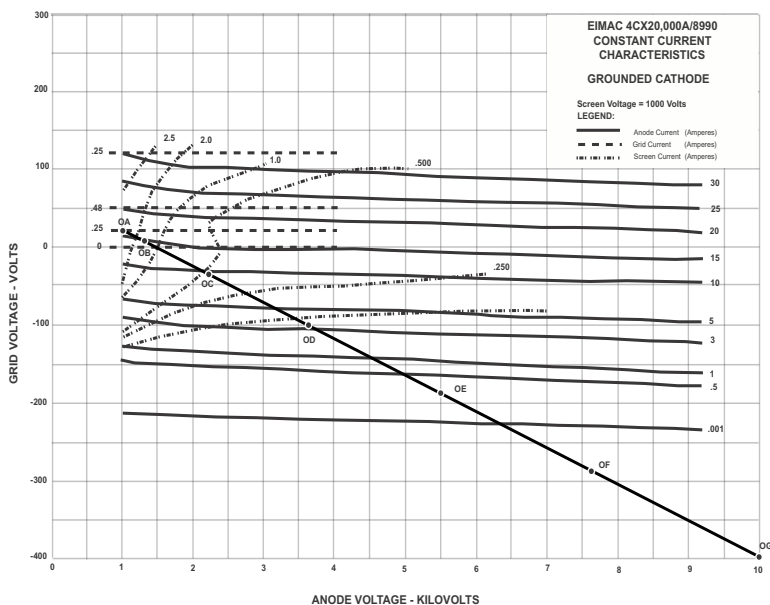


Figure 27. Constant current characteristics for 4CX20,000A/8990 tetrode.

3.2.2 Use of the EIMAC Tube Performance Computer

The EIMAC Tube Performance Computer provides the means to make these simple calculations. It is a means to determine the dc component, the fundamental RF component, or the approximate harmonic component of the current flowing in a tube when the tube is operating as a radio-frequency amplifier. It also enables one to determine what all the meter readings will be, and to predict the RF output power and the required drive power. With these factors known, we are able to forecast what will happen if any of the operating conditions are changed.

The EIMAC Tube Performance Computer

The EIMAC Tube Performance Computer consists of a transparent, graduated "Cosine Scale" overlay, which is used with the Constant-Current Characteristic curves for the tube under consideration. The transparent overlay is used to facilitate the determination of instantaneous current values of the tube at intervals of 15° along a 90° segment of the operating line of the tube, as plotted on the Constant-Current Characteristic curves. Simple arithmetic operations then yield the desired dc and peak fundamental values of the tube currents.

The EIMAC Tube Performance Computer is based on graphical methods of Fourier analysis as applied to the operation of power amplifiers and oscillators.

The EIMAC Tube Performance Computer is a simple aid to enable the selection of suitable values from the characteristic curves of a tube, and, by means of simple calculations, to forecast the performance of the tube in radio-frequency power amplifier applications.

The basic steps are outlined under "Instructions" on the Performance Computer. This requires selecting dc anode and grid-bias voltages, being guided by the typical operating values given on the technical data sheet for the tube type under investigation. Next, a suitable "operating line" must be chosen on the constant-current curves for the tube type (plotted on the grid-voltage/anode-voltage coordinates).

The computer, when properly placed over this operating line, permits one to obtain instantaneous values of the currents flowing at every 15° of the electrical cycle. The formulas given on the computer were derived by Chaffee⁵ to give the various average

5 E. L. Chaffee "A Simplified Harmonic Analysis," Review - Sci. Inst. 7, October 1936, p. 384.

and harmonic components of the resulting currents. Knowing these current component values and the radio-frequency voltage values which are determined through use of the computer, one can readily calculate the complete performance of the tube.

The fundamental methods of making such computations, the considerations necessary to stay within the ratings of the tube types while accomplishing various forms of modulation have been covered in literature.^{6, 7, 8, 9, 10} The method used for harmonic amplifier service is approximate and should be used only for tetrode and pentode tubes, where the anode voltage has little effect on the amount of anode current flowing. A more exact method, showing that for harmonic operation, the operating line is a simple Lissajou Figure, has been described by Brown.

The results obtained by using the Tube Performance Computer are a good first approximation for power amplifier design and will give good accuracy for most applications. The resulting accuracy is well within the normal tolerances of tube characteristics due to the normal variation in manufacturing dimensions of a tube. Since the published curves are only typical of the characteristics to be expected of a particular tube type, the calculated performance will be well within the values obtained when different tubes of a given tube type are operated under the assumed conditions.

3.2.3 Example Showing Detailed Use of the Computer in Class C RF Power Amplifiers

Assume that we have an EIMAC 4CX20,000A/8990 tetrode and want to make it work efficiently. Also assume that we have a 10 kilovolt dc power supply available.

Within frequency limits, we know that a tube should be able to run in Class C amplifier service with about 75% efficiency or higher, or, in other words, to convert 75% of the dc anode input power

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- 6 H. P. Thomas, "Determination of Grid Driving Power in Radio Frequency Power Amplifiers," Proc. IRE, Vol. 21, Aug. 1933, pp. 1134-1141.
 - 7 W. G. Wagener, "Simplified Methods for Computing Performance of Transmitting Tubes," Proc. IRE, January 1937.
 - 8 R. 1. Sarbacher, "Graphical Determination of PA Performance," Electronics, December 1942.
 - 9 R. 1. Sarbacher, "Performance of Self Biased Modulated Amplifier," Electronics, April 1943.
 - 10 Robert H. Brown, "Harmonic Amplifier Design," Proc. IRE, Vol. 35, August 1947, pp. 771 - 777.

to RF output power. The difference, or 25% of the input power, is dissipated and lost as heat on the anode of the tube. The dc anode input power is then about four times the power dissipated on the anode.

The 4CX20,000A/8990 tetrode has a maximum rated anode dissipation of 20 kilowatts. To illustrate performance near the maximum rating, we will choose an input power about 1.5 to 2 times the dissipation rating, or about 30 - 40 kilowatts. At 10 kilovolts the dc anode current must then be 3 to 4 A. We will use 3.2 Amperes of anode current for our example, striving for 80% efficiency. We hope to obtain at least 25 kilowatts of output power. It is normal practice with tetrodes and medium or low μ triodes in Class C amplifier service for the dc grid bias voltage to be roughly two to three times the grid voltage necessary to cut off the flow of anode current. By referring to the curves for the 4CX20,000A/8990, it is decided to use a dc grid bias voltage of -400 volts. Higher bias voltage may be employed in applications where highest possible efficiency is desired (and additional drive power is available). In such a case, points will have to be established below the bottom-most line in the data sheet, which is -400 Volts, by scaling a new grid voltage line horizontally below the existing line. Scaling for higher anode voltage can also be done in this way (within the absolute maximum ratings established for a given tube type).

Now locate the operating line on the constant-current curves for the 4CX20,000A/8990. (See Figure 27). First, mark the point where the dc grid bias and dc anode voltage cross. The operating line must go through this point, which can be referred to as "Point No. 1." Next, the peak value of the anode current and how low we can let the instantaneous value of anode voltage ($e_{b_{min}}$) be when the tube is passing this much current must be decided. This is necessary in order to locate the other end of the operating line, e.g., "Point No. 2".

The peak value of anode current in Class C amplifiers can range from three to five times the value of average dc anode current, the actual value being determined by the conduction angle (narrower conduction angles result in a higher peak-to-dc ratio). Less than 120 degrees is a typical conduction angle and a peak-to-average ratio of 5 is typical in high efficiency class-C service. The minimum value of $e_{b_{min}}$ is usually limited by the fact that if it is too low, the grid current in triodes or the screen current in tetrodes will be needlessly high, resulting in high grid or screen dissipation. Little will be gained as far as output power is concerned if the tube is

driven “harder.” The minimum value of anode voltage is usually in the region where the anode constant-current curves bend upward (entering saturation, see Figure 27.) In the case of triodes, this is near the “diode line,” where the instantaneous anode and grid voltages are equal. The practical procedure in calculating tube performance is to arbitrarily choose Point No. 2, complete the calculations, and compare the data to target values and repeat the process until the desired results are obtained.

In the case of the 4CX20,000A/8990, let us choose a peak value of anode current about five times that of the dc anode current of 3.2 Amperes, or approx. 16 A. Let us choose a minimum instantaneous anode voltage of 1 kilovolt, the same value of screen voltage used when the curves were taken. This defines the upper end of the *operating line*. Locate this point on the tube curves, and refer to it as Point No. 2 in Figure 27. (The anode currents which flow at various combinations of anode and grid voltages are shown by the anode-current lines.) The value of current for each line is noted. In-between values can be estimated closely enough for our purposes. Draw a straight line between Points 1 and 2. This line is the *operating line*, and shows the current and voltage values at each instant in the RF cycle when the current is being taken from the tube. The nonconducting half of the RF cycle would be shown by extending the line an equal distance on the opposite side of Point No. 1. There is little to be gained by this line extension, since no current flows during this half of the cycle.

The EIMAC Tube Performance Computer can now be used to obtain the average currents (what will become meter readings in actual operating conditions) and anticipated power values from this *operating line*. Overlay the computer on the constant-current curve so that the “guide lines” of the computer are parallel with the *operating line*. Slide the computer (without turning it) until the line **OG** passes through the dc voltage point No. 1, and line **OA** passes through the peak-current Point No. 2. Be sure the guide lines are still parallel to the *operating line*.

Note that the lines OB, OC, OD, OE, and OF all cross over the *operating line*.

At each point where the lines OA, OB, etc. cross the *operating line*, we must determine the instantaneous values of anode current and grid current (and screen current if a tetrode or pentode is used) which are flowing at that particular moment in the RF cycle.

Later, from these key values of current, values of anode current, grid current, screen current, and the RF components of the anode current may be calculated.

These current values should be listed for each point where the lines OA, OB, etc. cross the *operating line* so that they can be combined later to calculate the various tube currents. At points where OE and OF cross, the current values are insignificant or zero in Class C operation.

In the examples chosen, read off the instantaneous anode current values where these lines cross the operating line. At the point where the line OA crosses the operating line, the anode current is 16 A. Where OB crosses the operating line, the anode current is 15 A. At OC the anode current is about 9 A, OD is 3 A, OE is 0.25 A, and OF and OG are 0 mA. Similarly, the instantaneous screen current at the crossing of OA and the operating line is 2.25 A, and the instantaneous grid current is 0.25 A. Values are read for the other crossover points and written down. These values are put into columns for calculation:

Instantaneous current values				
Crossing of Line	Simplified name in Formulas	Anode	Screen	Control Grid
OA	A	16A	2.25A	0.25A
OB	B	15	1.5	0.14
OC	C	9	0.5	0
OD	D	3	0.1	0
OE	E	0.25	0	0
OF	F	0	0	0

In order to obtain the dc value of anode, screen, and control grid currents the formula (see computer) indicated that the above values should be added, using only one-half the A values (giving 8 A for anode, 1.125 A for screen, and 125 mA for grid), and then divided by 12 (or multiplied by .0833), as follows

D-C meter reading = 1/12 times (0.5A+B+C+D+E+F)

Anode	Screen	Control Grid
8A	1.125A	125 mA
15	1.5	140
9	0.5	0
3	0.1	0
0.25	0	0
0	0	0

D-C current = 1/12 totals = (approximately)

2.9A	269mA	22mA
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To calculate the RF output power it is necessary to use the formula for the peak RF current which is present in the tube anode current. Since the tube is being used as a straight RF power amplifier, use the formula for "Peak Fundamental RF" as shown on the computer. If estimating the performance of a doubler or tripler, use the formula for "Peak 2nd Harmonic RF" or "Peak 3rd Harmonic RF."

From the computer, it may be seen that the formula for the peak fundamental RF current is:

$$\begin{array}{rcl}
 & 1/12 (A+1.93B+1.73C+1.41D+E+0.52F) & \\
 A & = & 8 & = & 16 \text{ A} \\
 1.93B & = & 1.93 \times 15 & = & 28.95 \\
 1.73C & = & 1.73 \times 9 & = & 15.57 \\
 1.41D & = & 1.41 \times 3 & = & 4.23 \\
 E & = & 0.25 & = & 0.25 \\
 \text{Total} & & & & 65.0 \text{ A}
 \end{array}$$

Peak fundamental current = 1/12 total or 65/12 = 5.4 A

The various current values are now at hand. In order to calculate the powers involved, it is necessary to know not only the dc voltage values but also the greatest amount each value swings away from the dc value. This is known as the peak value of the RF voltage. Because the anode voltage swings from 10 kilovolts down to 1 kV, the peak RF anode voltage is the difference between these two Figures, or 9 kilovolts. Similarly, the grid voltage must rise and fall between the operating points No. 1 and No. 2, or from -400 volts to +20 volts. This is a peak swing of 420 volts, and the peak RF grid voltage is +20 volts. Now use the formula for output power and driving power: Output power = one-half RF anode current x peak RF anode voltage.

The peak RF anode current is found to be 5.4 ampere, and the peak RF anode voltage is found to be 9 kilovolts.

So: Output Power = $5.4 \times 9000 \times 0.5 = 24.3$ kilowatts.
 and Input Power = dc Anode Current x dc Anode Voltage = $2.9 \times 10,000 = 29$ kilowatts
 Anode Dissipation = dc Input Power - RF Output Power = $29 - 24.3 = 4.7$ kilowatts
 Efficiency = r - f Output Power divided by d - c Input Power = $24.3/29 = 83\%$
 Driving Power = dc Grid Current x Peak RF Grid Voltage = $0.0265 \times 420 = 111$ Watts

The power consumed by the bias source is the product of the dc grid current and the dc grid voltage, or $0.022 \times 400 = 88$ Watts.

The difference between the driving power and the power consumed by the bias source is the power dissipated on the control grid, or $111 - 88 = 23$ Watts.

The power dissipated on the screen grid is the product of the dc screen current and the dc screen voltage, because the screen grid has no impedance between it and the dc screen supply. Thus it is $0.269 \times 1000 = 269$ Watts.

The performance of the tube can be summarized:

dc Anode Voltage	10,000	Volts	Driving Power	111	Watts
dc Screen Voltage	1000	Volts	Grid Dissipation	23	Watts
dc Grid Voltage	-400	Volts	Screen Dissipation	269	Watts
dc Anode Current	2.9	A	Anode Input Power	29	kW
dc Screen Current	269	mA	Anode Output Power	24.3	kW
Peak RF Grid Voltage	420	Volts	Anode Dissipation	4.7	kW
calculated efficiency	83%				

We see that the resulting calculated value of dc anode current (2.9 A) is lower than our initially chosen value of 3.2 by approx. ten percent, evidence that calculated efficiency of 83% is higher than the target value of 80% we initially anticipated. In the "real world" actual transmitter output power will be somewhat lower than calculated values because of RF losses in the output matching network. One should also be aware that there may be variations in output from one tube to another due to varying electrical characteristics and that a tube that has been in service for an appreciable length of time may deliver less output than it did when it was new. Varying the dc screen voltage on a tetrode is an excellent method for

varying output power and is employed for maintaining a fixed power requirement; as a given tube ages and loses emission from the filament its output power will decline. Compensating for losses over time is easily accomplished by using a voltage source for the screen that is adjustable by means of a closed-loop regulator system. If one wishes to calculate tube performance of a tetrode at screen voltages other than those at which the constant current curves were taken, a compensating factor using the three-halves power law can be employed (see sec 6.4). To account for loss of output over time a good designer will choose a tube that has output power capabilities well above the specified requirement, using as much conservatism as practicable. In the case of the 4CX20,000A/8990 in FM broadcast service actual tube life of 10,000 to 20,000 hours is common, indicating this tube is a good choice for supplying RF power from approx. 20 to 30 kilowatts in continuous commercial service.

3.2.4 Use of EIMAC Tube Performance Computer for Class A, AB, and B Service

While the EIMAC Tube Performance Computer is primarily designed for use in Class C service, it may be used for Class A, AB, and B service where the idling (quiescent) anode current is not zero. To calculate performance for operating conditions having a large order of idling anode current, the anode current flow during the positive half of the anode voltage swing becomes appreciable and cannot be ignored. When the tube functions over 180 degrees or more of the operating cycle, a full set of ordinate points must be employed for the computations. The computer is therefore used in a two-step process.

First, determine the operating line. The computer is used in the normal fashion to derive the instantaneous values of anode, screen, and grid current during the negative half of the anode voltage swing. These current points are logged as explained under the Class C service description.

Now, determine the instantaneous current points over the positive portion of the anode voltage cycle. Combine these with the points taken for the negative half of the cycle. This is done in the following fashion: a line is penciled on the computer over the operating line, and of equivalent length. The computer is now inverted and rotated 180° and again aligned with the chosen operating line on the constant current curve, so that inverted point G falls on the idling current value (Point No. 1) and inverted point A passes through a minimum peak current point representing maximum positive anode

voltage swing. The penciled line on the computer now represents an extension of the operating line into the area of positive peak voltage swing. The extended operating line is a straight line, twice as long as the original operating line. Instantaneous values of anode, screen, and grid current (if any) are those observed where the reversed ordinate lines on the computer cross the extended operating line.

The current points measured with the computer inverted are logged and these, together with the points obtained in the first operation (computer right side up) provide a full set of instantaneous peak current values necessary for the calculation of Class A, AB, or B parameters. With the first set of computer readings termed A, B, C, D, E, F and G, the second (inverted) set of Figures may be termed A', B', C', D', E', and F', with the sets used in the following formulae, which are modifications of Chaffee's permitting the use of current values directly from the constant current curves:

$$(1) \text{ D. C. Current (Meter Reading)} = \frac{1}{12} \left[\frac{A + A'}{2} + (B + B') + (C + C') + (D + D') + (E + E') + (F + F') + G \right]$$

$$(2) \text{ Peak Fundamental RF Current} = \frac{1}{12} [(A - A') + 1.93 (B - B') + 1.73 (C - C') + 1.41 (D - D') + (E - E') + 0.52 (F - F')]$$

$$(3) \text{ Approx. 2d Harmonic RF current (tetrodes or pentodes only)} = \frac{1}{12} [(A + A') + 1.73 (B + B') + (C + C') - (E + E') - 1.73 (F + F') - 2G]$$

$$(4) \text{ Approx. 3d Harmonic RF current (tetrodes or pentodes only)} = \frac{1}{12} [(A - A') + 1.41 (B - B') - 1.41 (D - D') - 2(E - E') - 1.41 (F - F')]$$

$$(5) \text{ Power Output} = \frac{\text{Peak Fundamental RF current} \times \text{Peak RF voltage}}{2}$$

$$(6) \text{ Resonant Load Impedance} = \frac{\text{Peak RF Voltage}}{\text{Peak Fund. RF Current}}$$

3.2.5 An Example: Using the Computer for Class AB₁ Service

Operating data is to be derived for an EIMAC 4CW100,000D tetrode operating at a anode potential of 10 kV with a screen potential of 1.5 kV. Assume that power output design goal for this particular application is about 60 kW. Grid current is zero; that is, the tube is operated in the Class AB₁ mode, with the grid never being driven positive.

Within frequency limits, an anode circuit efficiency of about 55-60% may be assumed for Class AB₁ operation. Maximum dc anode input is therefore 2.2 to 2.75 times the anode dissipation. A maximum power input of 2.2 x 50,000 = .110 kW is chosen. At 10 kV, the maximum dc anode current is then 11.0 A. This is within the maximum rated anode current of 15 A for Class AB₁ service as specified on the data sheet.

For Class AB₁ service the tube does not operate in a cut-off condition, but rather a certain value of idling anode current exists. This must be taken into account when choosing Point 1 on the constant current curve. Generally speaking, high levels of resting anode current provide reduced levels of intermodulation distortion products and somewhat lower anode efficiency. Idling anode current is usually chosen so that anode dissipation under quiescent conditions is about 50-70% of the maximum dissipation rating. In the following example, idling anode current is chosen to be 4.5 amperes. From Figure 28, it may be determined that a grid potential of about -295 volts is required to produce the desired anode current at the chosen screen and anode potentials. The intersection of the -295 volt bias line and the 10 kV anode line determines the idling point on the operating line (Point No. 1).

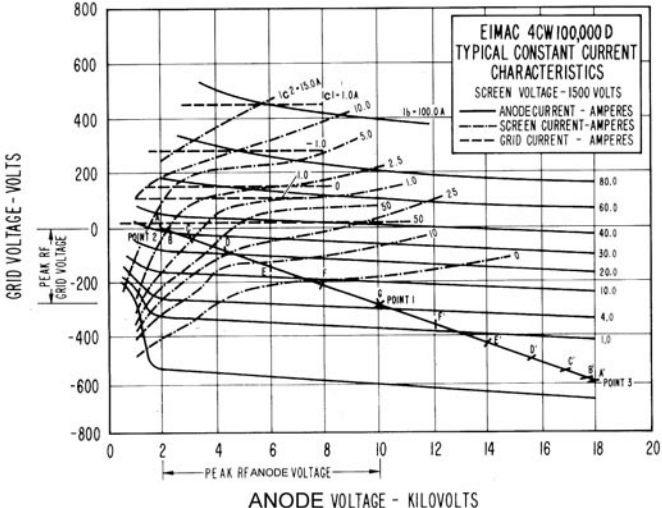


Figure 28. Constant current characteristics for 4CW100,000D tetrode.

Next, the peak value of anode current must be determined, and the minimum amount of instantaneous anode voltage chosen to pass this amount of current. Determination of these values will locate Point No. 2 and will thus define the operating line.

Class AB₁ service limits grid voltage excursions to negative grid regions on the constant current graph. Point No. 2 therefore may never be located above the zero grid voltage line. In addition, the minimum instantaneous anode voltage is usually not allowed to swing as low as the dc screen potential, since screen dissipation tends to become abnormally high. The location of Point No. 2 thus has certain restrictive limits defined by screen dissipation

and the maximum positive grid signal voltage. In this case, for the 4CW100,000D, minimum instantaneous anode voltage is about 2 kV. Peak RF voltage is thus 10,000 - 2,000 = 8,000 volts.

Peak RF anode current in a Class AB₁ amplifier usually runs about 2.5 to 3.0 times the average dc anode current. In this case, a maximum peak anode current of about 2.9 times the maximum signal dc anode current of 11.0 A, or 32 amperes, is chosen. This defines Point No. 2, which is at the intersection of the 2 kV minimum anode voltage line and thus falls within the limits discussed in the preceding paragraph. A straight line is drawn between Point No. 1 and Point No. 2 which is the negative anode cycle portion of the operating line.

When the operating line is extended to the right of Point No. 1, it can be observed that the tube conducts over the rest of the cycle where (by virtue of the "flywheel" effect of the resonant tank circuit) the instantaneous anode voltage swings as far above the normal dc value as it swings below. It is important to note that operation with less than cutoff bias requires that the EIMAC Tube Performance Computer employ points on the operating line falling to the right of Point No. 1. The operating line is accordingly extended and the computer is used in a two-part operation, as shown in the following example:

4CW100,000D COMPUTATIONS

d-c Plate Voltage = 10 kV	Power Input	= 110 kW
d-c Screen Voltage = 1.5 kV	Max. d-c Plate Current =	11 amp.
	Zero-Signal Plate Current =	4.5 amp.
	d - c Grid Voltage =	-295 volts
Constant Current Graph and	Peak Plate Current =	32 amp.
Operating Line determine:	Peak RF Voltage =	8,000

Step One

EIMAC Computer Readings		
Instantaneous Peak Current (amps)		
Ordinate Crossing	Plate	Screen
A	32	3
B	31	2
C	28	1.2
D	22	0.25
E	15	0.07
F	9	—
G	4.5	—

Step Two

Inverted Computer Readings		
Instantaneous Peak Current (amps)		
Ordinate Crossing	Plate	Screen
A'	0.20	—
B'	0.25	—
C'	0.30	—
D'	0.50	—
E'	0.80	—
F'	1.50	—

$$\begin{aligned}
 \text{d-c Plate Current} &= \frac{1}{12} \left[\frac{32^2}{2} + 31.25 \cdot 28.3 + 22.5 \cdot 15.8 + 10.5 \cdot 4.5 \right] = 10.75 \text{ amp.} \\
 \text{Plate Power Input} &= 10 \text{ kV} \times 10.75 \text{ A} = 107,500 \text{ watts} \\
 \text{Peak Fundamental RF Current} &= \frac{1}{12} \left[31.5 \cdot 59.4 + 47.7 \cdot 30.3 + 14.2 \cdot 3.9 \right] = 15.6 \text{ amp.} \\
 \text{Power Output} &= \frac{15.6 \times 8000}{2} = 62,500 \text{ watts} \\
 \text{Plate Dissipation} &= 45,000 \text{ watts} \\
 \text{Efficiency} &= \frac{62,500}{107,500} \times 100 = 58.2\% \\
 \text{Resonant Load Impedance} &= \frac{8000}{15.6} = 512 \text{ ohms} \\
 \text{d-c Screen Current} &= \frac{1}{12} \left[\frac{3}{2} + 2 + 1.2 + 0.25 + 0.07 \right] = .417 \text{ mA.}
 \end{aligned}$$

3.3 TYPICAL RF AMPLIFIER CIRCUIT DESIGNS

In the previous discussion of tube performance calculations, an example was worked out using the 4CX20,000A in class C service. Using the obtained operating parameters, it is now possible to demonstrate the next step in the circuit design. For the benefit of discussion, it will be assumed that the output circuit is a shunt-fed pi-network; it will also be assumed that the grid circuit is to be tuned by a conventional parallel tuned circuit. The circuit will be as shown in Figure 29.

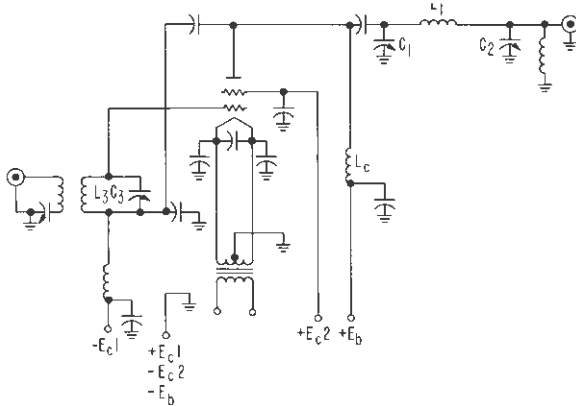


Figure 29. A typical circuit for an RF amplifier.

The first step in designing the output circuit is to specify the resonant load impedance of the tube, the loaded Q of the circuit and the desired output impedance of the network. The resonant load impedance for the 4CX20,000A is determined by dividing the anode peak RF voltage swing by the anode peak fundamental RF current.

Resonant Load Impedance = Peak RF Voltage/Peak Fundamental RF Current

$$RL = 9000 \text{ volts}/5.4 \text{ amperes}$$

$$= 1667 \text{ Ohms}$$

If it is assumed that the output impedance of the network is to be 50 Ohms and the loaded Q is to be 15, the output tuned circuit may now be designed.

The output impedance of 50 Ohms will match into a properly terminated 50-Ohm transmission line. The loaded Q of 15 is a compromise between circuit efficiency and harmonic attenuation

(see Fig. 30). Figures 31, 32, and 33 are design graphs for matching typical tube load impedances into 50 Ohms with a loaded Q of 15. The general equations are given for other load

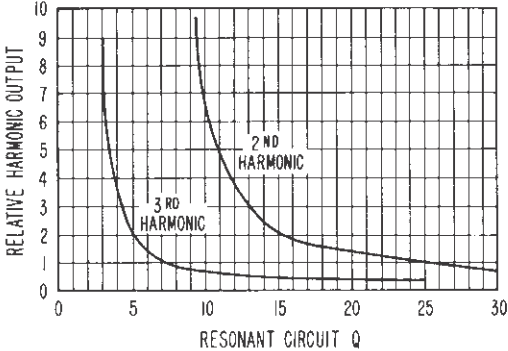


Figure 30: Relative harmonic vs. Resonant circuit Q.

impedances and other Q's. For the purpose of this example, the parameters for 2 MHz will be determined. At this frequency the value of C1 (Figure 31) will be 700 pF, C2 will be 4000 pF and the value of L1 will be 9.5 microhenries. The C1 value includes the tube output capacitance, the stray capacitance from tube to ground

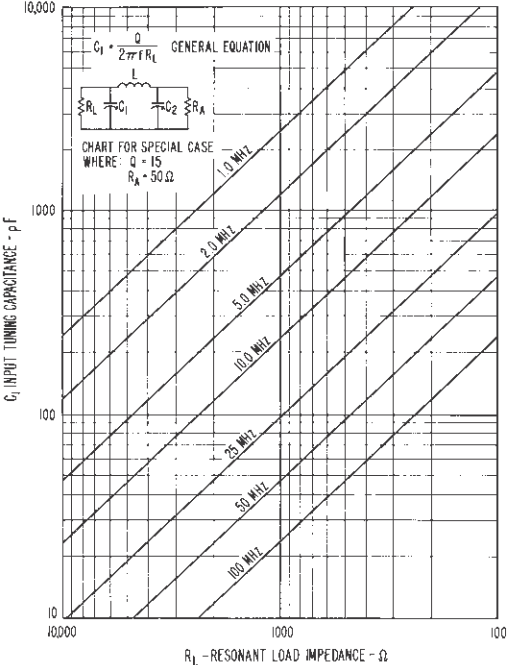


Figure 31: Determination of input capacitor C1.

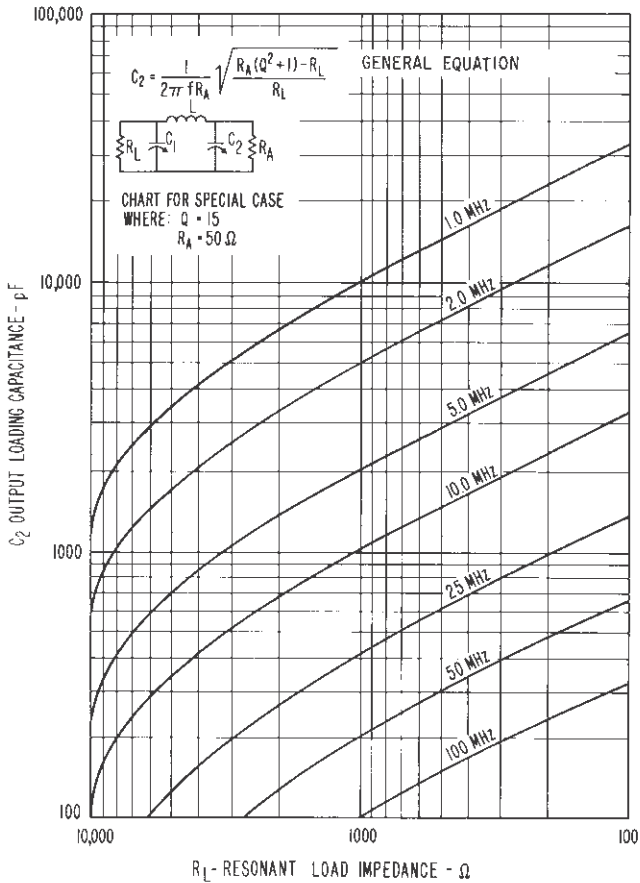


Figure 32: Determination of loading Capacitor C₂.

and the lumped capacitance added to the circuit. The output capacitance of the tube is given on the Eimac technical data sheet. Tube capacitance is measured in a shielded fixture (using a 4-wire bridge at 100 kHz), therefore the value that is obtained is one that is confined to the tube alone. Stray capacitance becomes increasingly important at higher frequencies; at 2 MHz it may be neglected.

The inductance of the anode choke (L_C in Figure 29) may be determined by noting that it is in shunt with the tuned circuit. If the RF current through the choke is to be limited to one percent of the tuned circuit circulating current, the inductance of L_C must then

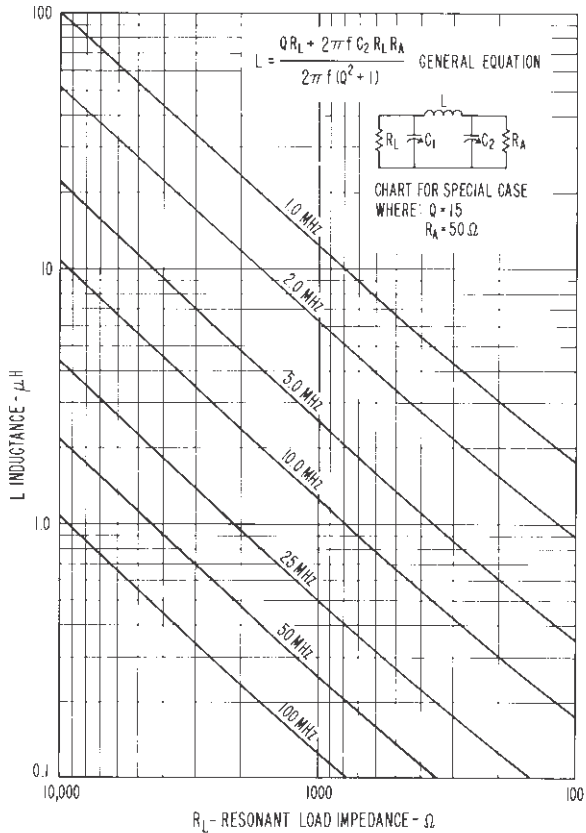


Figure 33A. Determination of Inductor L

be 100 times that of coil L. The inductance of L_c would then be 950 microhenries. The series resonant frequency of the anode choke must be determined to assure that it is outside the normal operating frequency range of the amplifier. This can be checked by shorting the ends of the choke together using a low inductance strap and measuring the resonant frequency by means of a grid-dip meter or impedance bridge. Building an amplifier that covers a wide range of frequencies (for example 2 to 30 MHz) may present such a demand on the anode choke that the easiest and most effective solution is to use two RF chokes in series, shorting out one section on all but the lowest frequency band. Varying the pitch of a portion of the solenoid windings is another technique for avoiding a series resonance at an undesired frequency. Pi-wound chokes are often riddled with resonances and are generally not used in high power amplifiers.

$$R = \frac{P}{I^2}$$

R = input resistance
P = power delivered to grid
I = d-c grid current

$$R = \frac{1.4 \text{ watts}}{(.014 \text{ amps})^2} = 7140 \text{ ohms}$$

$$X_C = X_L = \frac{R}{Q}$$

X_C = capacitive reactance of input circuit
 X_L = inductive reactance of input circuit
Q = loaded Q of grid circuit
 $X_C = X_L = \frac{7140}{15} = 476 \text{ ohms}$

It is also important to determine the parallel-resonant frequency of the anode choke due to the stray capacitance involved. A choke will exhibit inductive reactance below the parallel resonance and capacitive reactance above the resonance. The parallel resonant frequency must therefore be above the operating frequency of the amplifier by a small amount.

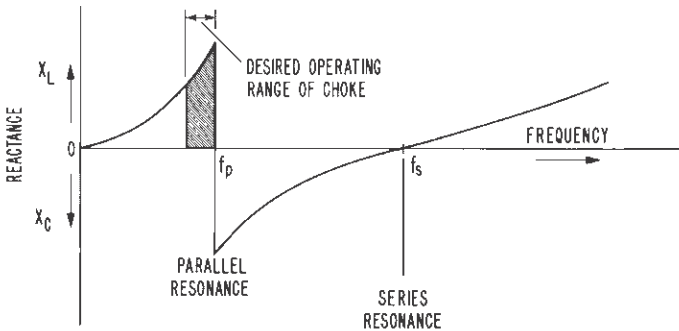


Figure 33B. Reactance of an RF choke vs. Frequency.

The design of a typical grid-tuned circuit is quite similar to the design of the anode resonant circuit. For class C operation, or any class where grid current is flowing, the input circuit must have sufficient Q to maintain a sinusoidal grid wave shape. The tube's input resistance loads the input circuit. It is considered good engineering practice to have a circuit Q of between 12 and 15. In any class of operation with no grid current, the grid circuit requirements are not as stringent, provided broad bandwidth is not a factor. Applications where broad frequency coverage is desired may better utilize a ferrite loaded impedance matching transformer terminated into a non-inductive resistor (typically 400 Ohms to 1K Ohms); this is

then coupled to a parallel resonant circuit at the control grid. The net power gain resulting from such a circuit is somewhat less than that of a conventional high-impedance grid-tuned circuit but this circuit generally does not require as much neutralization, due to the swamping effect of the low grid impedance.

For the 4CX20,000A example using a parallel tuned matching circuit (Fig. 27), the tube's input resistance is approx. the power delivered to the grid of the tube divided by the square of the dc grid current.

The input circuit capacitance C3, for 2 MHz operation will therefore be approx. 1250 pF. The input circuit inductance L3 will be approx. 5 microhenries.

A further point of interest is the magnitude of the current necessary to charge the tube's interelectrode capacitances; these must be charged and discharged during each RF cycle. The input capacitance of the 4CX20,000A is approx. 190 pF and the output capacitance is approx. 24 pF. With these values of capacitance, the input capacitive reactance at 2 MHz is 418 Ohms, and the output reactance is about 3300 Ohms. In the 4CX20,000A example, the grid voltage swing is 420 Volts; therefore, the grid charging current will be $420/418$ or about 1 Ampere. The anode voltage swing is 9000 Volts; therefore, the anode charging current will be $9000/3300$ or 2.72 Amperes. Current values for this tube are quite low in this case because the frequency is relatively low. Operating this tube under the same conditions at fifty times this frequency (100 MHz) the charging currents will be fifty times the 2 MHz example. The greater the charging current the greater the tube seal heating, electrode heating, circuit losses and the greater the current in the screen bypass capacitor. At higher frequencies it is better to limit the anode voltage swing and raise the anode current to supply the same power output. This technique reduces the magnitude of the anode charging current with generally only a slight loss of anode efficiency.

3.4 COMPONENT PARTS

If one is to maintain good isolation of the output and input circuits, some thought must be given to the location of the component parts of the amplifier. All component parts of the grid or input circuit and any earlier stages must be kept out of the anode circuit compartment. Similarly, anode circuit parts must be kept out of the input compartment. A discussion of compartmental enclosures is presented in sec. 6.7.3.

Proper choice of materials for interconnecting the tube and circuit is also important where long life and good performance is concerned. Copper is the preferred material for connection straps although brass and aluminum are also suitable in most cases; surface plating is optional in low power applications and not generally necessary, except for cosmetic reasons or corrosion prevention. Silver plating is preferred in components used for VHF and UHF, especially where high power (over approx. 10 KW) is anticipated; oxidized or tarnished silver does not exhibit significantly more loss than its untarnished state, due to the fact that it is still a good conductor. Nickel plating is often employed where cost may be a factor; its greater resistance to chemicals is another advantage. Components used with water-cooled tubes or tubes designed to be immersed in insulating oil or gas are often nickel plated. Gold plating is preferred in microwave cavities to reduce losses due to skin effect and because gold does not oxidize. Steel and stainless steel alloys should be avoided in RF power amplifiers because of the high RF losses associated with those materials in the presence of RF fields. If steel hardware is necessary for the mechanical strength, it can be copper plated; at HF and above, RF current will flow primarily through the plating. At LF iron or steel should be avoided due to increased skin depth and losses associated with its ferromagnetic properties.

3.5 LEAD LENGTHS

Most of the connecting leads should be designed with extremely low inductance to minimize the possible VHF parasitic circuits. Filament and screen bypass leads, suppressor bypass leads, leads to the grid and the anode tuning capacitors, and interconnections between tubes in both push-pull and parallel circuits are in this category. An exception is a parasitic suppressor (if used), which must have an inductance selected according to the operating frequency. For a lead to have low inductance, it must have a relatively large surface area and be short in length, as in a strap or ribbon shape. This consideration also applies to that portion of a lead inside a bypass capacitor or connecting from the tube to a bypass capacitor. If two or more tubes are used, they should be placed reasonably close together to help provide short lead lengths. At HF the length of leads connecting RF tuned circuits to the tube(s) can usually be much longer than bypass capacitor lead lengths. The dc supply, keying, modulating and control circuit wires can be quite long if properly filtered or shielded and arranged out of any RF fields.

It should be noted that the lead connecting to the screen grid for tetrodes and pentodes is common to both the output and input

resonant circuits. Due to the anode-to-screen capacitance of a tetrode or pentode, the RF anode voltage (developed in the output circuit) causes an RF current to flow through the screen bypass capacitor to the chassis. In the case of a push-pull stage this current may flow from the screen terminal of one tube to the screen terminal of the other tube. Similarly, due to the grid-to-screen capacitance of the tube, an RF voltage in the input circuit will cause an RF current to flow in this same screen lead to the chassis, or to the opposite tube in the push-pull circuit.

The inductance of the lead common to both the output and input circuits has the desirable feature of providing voltage of opposite polarities to neutralize the feedback voltage of the residual anode-to-control grid capacitance in the tube (this is discussed under "Neutralization" in section 5). It should be noted, however, that the mutual coupling from the screen lead to the input resonant circuit may be a possible source of trouble if the inductance is excessive.

With common-grid (cathode-driven) triode circuits, the control grid and associated leads are common to the output and input circuits. The inductance of the control grid lead can help or hinder the stabilization of a grounded grid amplifier (this is discussed under "Neutralization" in section 5).

In common-cathode (grid-driven) circuits, excessive cathode lead inductance can be detrimental to good performance. RF current through any inductance in the cathode (or in the case of thoriated tungsten tubes the filament) lead to ground will cause an RF voltage to appear across the grid-to-cathode region and, due to phase relationships, will decrease the input conductance. This will result in increased drive power requirements as the frequency is raised. The net anode (output) efficiency can also be reduced by this effect. A mathematical formula for power loss vs. frequency due to cathode lead inductance is presented in sec. 6.8. In large power grid tubes, this loss of efficiency and low gain may begin to occur at frequencies above 20 to 30 MHz, depending on the physical size of the filament. Some Eimac triodes using filament leads (also called flying leads) are provided with threaded holes for connecting the input circuit directly to the tube's filament ring; this is also the correct point for a filament bypass capacitor, not at the tube lug ends. At VHF and UHF frequencies it often becomes necessary to utilize an input tuned circuit that is a direct extension of the tube, i.e., there is no discrete socket. This so-called "cavity" structure eliminates series inductance, which is so detrimental at these frequencies.

3.6 *FILAMENT BY-PASSING*

Tubes having thoriated tungsten filaments generally use AC voltage supplied from a filament transformer with the center tap used as the “cathode” dc return. DC is also used in some applications to prevent hum or modulation of the rf. In either case, with grid-driven (common cathode) circuits, RF bypassing the filament to ground is necessary to ensure that stability is maintained and that power gain is not lost as the operating frequency increases (see Section 3.5). This path must be of sufficiently low inductance to prevent the filament from becoming above ground with respect to the RF. It may be preferable to ground one filament terminal directly to the chassis and bypass the other (hot) terminal. Ceramic and multi-layer mica capacitors are suitable for bypassing the filaments up to and above HF frequencies. With tubes having concentric isolated cone terminals (for an example, see Figure. 19), the preferred VHF circuit design will avoid straps altogether. In this case, an effective method is to bypass the filaments directly to ground through low inductance film capacitors mounted against the same chassis deck or cavity wall to which the screen grid is bypassed. The Eimac SK-350 is an example of a VHF socket with integral film capacitors used for the filament bypass.

In cathode-driven (grounded grid) circuits, the filament terminals are isolated from ground in order to apply RF drive to that element. Any stray capacitance here can present a problem, because it is in shunt with applied RF voltage as well as being part of the input tuned circuit. The Eimac SK-375 is an example of a low capacitance socket available for use with certain tetrode tubes.

Inductance in the thoriated-tungsten filament structure itself may be resonant at the fundamental or some harmonic frequency. To prevent this resonance from absorbing a portion of the applied drive power, a bypass capacitor directly across the filament terminals is used. This applies to both grid-driven and cathode-driven circuits, because RF fields exist within the tube between the control grid and filament in both cases. In some VHF fixed-frequency applications using thoriated tungsten tubes, a variable capacitor is connected directly across the filament terminals to effect a series resonance, which allows for optimizing the anode efficiency. This is believed to peak filament emission at the crest of the applied RF sine wave. RF chokes are often used in the filament leads even with the filament bypassed to ground. This will help reduce RF from appearing at the filament supply, because filament bypassing tends to be less than perfect even in the best designs, due to finite series internal inductances, which exists in nearly all capacitors.

Indirectly heated oxide-cathode tubes may exhibit a broad gain variation vs. frequency in some VHF and UHF applications. This effect is caused by absorption of some of the applied drive power by the low impedance heater coil at its resonant frequency. The distributed capacitance from the heater coil to the cathode can, in effect become a helical resonator. The frequency of this resonance is typically in the high VHF to UHF range and is of relatively low Q , because of the fairly low resistance of the heater at operating temperature. A solution for this is to use RF chokes in both heater leads, with a capacitor connected across the heater to de-tune the resonance. The exact value of capacitance may need to be determined empirically. Looking for minute changes in the heater current with and without RF drive is a test for determining whether heater resonance is occurring.

In the grounded grid configuration at HF with either a triode or tetrode having a cathode that is electrically separated from the heater (for example 4CX250B and 3CX1500A7 tubes), ferrite-loaded chokes in the heater supply leads are sometimes employed, although not absolutely necessary. Even though the heater-to-cathode capacitance is relatively low, there is a small reactance that can lead to stray RF if the leads are not well shielded. The use of heater chokes in this configuration becomes more necessary at VHF and UHF. Some designers choose to “common mode” the heater and the cathode by bypassing these elements together, assuring that the RF potential between them will be insignificant. This requires using RF chokes in all three leads; a trifilar choke could also be employed. The maximum dc voltage rating between heater and cathode is in the range of only a few hundred volts in small tubes.

In this context, one scenario should be examined; namely, if anode voltage goes to ground in a fault condition. Referring to Figure 34, one would note that if E_{b+} is grounded, the full potential E_b appears across the 20 Ohm resistor and, if this resistor fails (opens), the cathode voltage approaches the same potential as the anode supply, but negative in polarity. Stored energy in the E_b supply will be applied to the cathode end of the tube (although a filament type tube is illustrated) and cathode bypass capacitors, the cathode and grid current meter as well as the secondary windings of the heater transformer will receive considerable energy. Heater (and cathode) bypass capacitors should be protected from fault conditions of this nature by clamping the B- side of the anode supply to ground using diodes capable of handling high peak current. Devices such as metal oxide varistors (MOV's) are available and worth investigating. The center tap of the heater supply transformer (if one is present) employed with oxide cathode tubes should not be grounded. See further discussion on fault protection in sec. 3.9.

3.7 SCREEN AND SUPPRESSOR GRID BY-PASSING AND SCREEN TUNING

Low inductance leads are generally advisable for screen and suppressor grid terminal connections. For all frequencies, it is a good practice to route the screen and suppressor bypass capacitors directly from one filament terminal. The suppressor grid is bypassed in the same manner when the suppressor is operated at other than cathode potential. With the suppressor operating at cathode potential, the suppressor should be grounded to the chassis directly.

Above the self-neutralization frequency (see "Neutralization" Section 5) of the tetrode or pentode, the screen bypass capacitors are sometimes variable. By proper adjustment of this capacitance the amount and phase of the screen RF voltage can be made to cancel the effects of the feedback capacitance within the tube. Thus, neutralization is accomplished. It is worth noting that the screen lead inductance and the variable capacitor are not series resonant, the variable capacitor is adjusted so that a net inductive reactance remains to provide the proper voltage and phase for neutralization.

The preceding paragraphs apply directly to tubes having the screen and suppressor grids mounted on supporting rods, notably tubes with glass envelopes. Ceramic tubes having concentric isolated cone terminals use a series of collets to connect to the tube. The screen collet may be bypassed with ceramic "doorknob" capacitors at HF. At VHF frequencies, the preferred bypass capacitor is made from a film-type dielectric mounted directly against the chassis deck, with the collet connected directly to a plate, which rests against the adjacent side of the dielectric. Dielectric materials can be Teflon®, Mica, or Mylar®. A modern material named Kapton® is also desirable, because opposing surfaces can be metallized for increasing the effective capacitance per unit area. The Eimac SK-840 is an example of a socket for the 5CX1500B pentode having the screen bypass capacitor built into mounting plate. This allows excellent stability up to several hundred MHz.

Many FM broadcast transmitters use a simple, effective scheme in which the length of several straps between the screen collet and chassis ground plane are adjusted according to operating frequency, establishing the proper neutralization. The theory of this is discussed in sec. 5.5.4

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Another method of assuring that the screen grid is at RF ground is to use a quarter-wavelength coaxial stub, shorted at one end and resonant at the operating frequency, connected from the screen to ground. A blocking capacitor is required to prevent shorting the screen dc supply voltage to ground. This is generally used in an amplifier operating at a single fixed frequency at VHF. The length of the stub becomes physically too long at lower operating frequencies.

3.8 GROUNDED-GRID CIRCUITS

Zero-bias, high μ triodes employed in RF power amplifiers for general purpose RF amplification, linear and even class-C FM service are typically used only in the grounded-grid configuration. The control grid is operated at RF ground, similar to the screen grid in a tetrode or pentode; therefore, good isolation is obtained between the input and output. The control grid may be bypassed directly at the socket or grounded directly to the chassis. The dc-grounded approach is favored, because no bypass capacitor is required and stability is optimum. Figures 34 and 35 illustrate the two circuit configurations.

Above 30 MHz the example in Figure. 35 should be avoided because of less than ideal RF grounding associated with this configuration. A case can also be made wherein Figure. 35 has less inherent reliability. In the event of an internal or external arc from anode to grid, the grid bypass capacitor and metering circuit could be destroyed, unless the available energy is limited to less than a few joules. Protection diodes connected across the meters are highly recommended in both circuits.

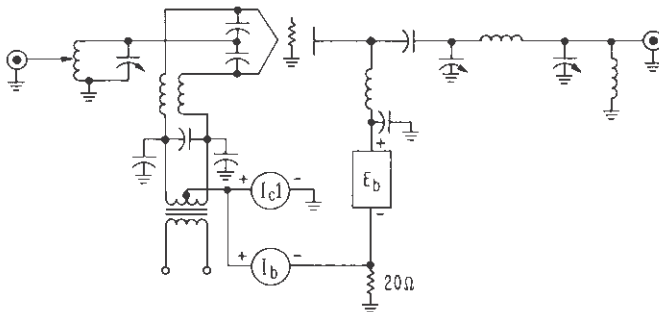


Figure 34. A typical circuit using “zero-bias” triodes showing metering circuits and method of grounding the grid to RF. The grid current is measured in the return lead from ground to filament.

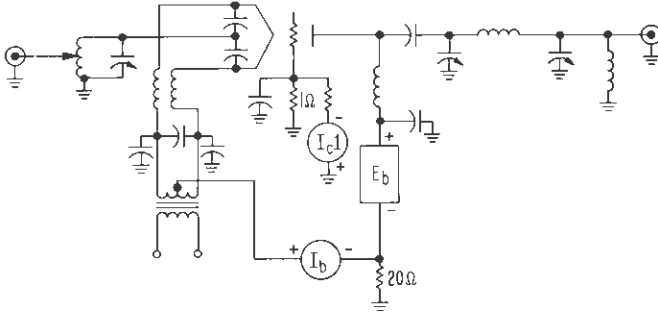


Figure 35. A typical circuit using “Zero-bias” triodes showing metering circuits. The grid is grounded to RF with a bypass capacitor. The grid is raised 1Ω above dc ground to allow the grid current to be measured.

Some Eimac triodes are designed with a grid flange that support the tube directly on the chassis and thereby provides the lowest inductance connection possible (see Figure. 36), while simultaneously eliminating the need for a socket.



Figure 36. 3CX5000A7 with integral grid flange

In grounded-grid circuits, the cathode lead inductance must be minimized to prevent loss of drive power before it reaches the grid-cathode region of the tube. The problem is similar to that seen in grid-driven circuits (discussed in Section 3.5.5). Several Eimac triodes have more than one cathode terminal (for example the 3CX800A7 has 6 of the 11 pins connected to the cathode). In

VHF circuits all of the available cathode pins should be utilized to assure that series inductance is as low as possible. The 3CX800A7 is useful to frequencies over 450 MHz with these pins connected directly to the cathode matching network.

3.9 PROTECTION

EIMAC Power Grid Tubes are designed to stand considerable abuse. For instance, the excess anode dissipation resulting from detuning the anode circuit of the tube will have no ill effects if not applied for periods of time sufficient to overheat the envelope and the seal structure.

Similarly, the control, screen and suppressor grids will stand some excess dissipation. The maximum dissipation for each grid indicated on the data sheet should not be exceeded except for time intervals of a few milliseconds. The maximum dissipation rating for each grid structure is usually considerably above typical values used for maximum output so that ample operating leeway is provided. The time of duration of overloads on a grid structure is necessarily short because of the small heat storage capacity of the wires. Furthermore, grid temperatures cannot be seen, so no visual warning of accidental overload is apparent.

The type and degree of protection required in an RF amplifier against circuit failure will vary with the type of screen and grid voltage supply. Figure 37 is a chart of tetrode and pentode protection as related to certain kinds of circuit failures.

For screen voltage taken through a dropping resistor from the anode supply, an anode relay provides almost universal protection. For a fixed screen supply a relay provides protection in most cases. For protection against excess antenna loading and consequent high anode dissipation, a screen undercurrent interlock may also be used in some services.

The anode, screen and bias voltages may be applied simultaneously to a tetrode. The same holds true for a pentode, plus the application of the suppressor voltage. In a grid driven amplifier, the grid bias and excitation can usually be applied alone to the tube, especially if a grid leak resistor is used for bias voltage. Anode voltage can be applied to the tetrode and pentode before the screen voltage with or without excitation to the control grid. **NEVER APPLY SCREEN VOLTAGE BEFORE ANODE VOLTAGE.** The only exception would be when the tube is cut off so that no space current (screen or anode current) will flow, or when the excitation and screen voltage are low. If screen voltage is applied before the anode voltage and screen current can flow,

CIRCUIT FAILURE	FIXED SCREEN SUPPLY		SCREEN VOLTAGE THROUGH DROPPING RESISTOR	
	FIXED GRID BIAS	RESISTOR GRID BIAS	FIXED GRID BIAS	RESISTOR GRID BIAS
Loss of Excitation	No Protection Required	Anode Current Relay	Anode Current Relay	Anode Current Relay or Screen Control Circuit
Loss of Antenna Loading	Screen Current Relay	Screen Current Relay	Grid Current Relay	Nothing Required
Excess Antenna Loading	Screen UnderCurrent Relay	Screen UnderCurrent Relay	Anode Current Relay	Anode Current Relay
Failure of Anode Supply	Screen Current Relay	Screen Current Relay	Grid Current Relay	Nothing Required
Failure of Screen Supply	Grid Current Relay	Nothing Required	—	—
Failure of Grid Bias Supply	Anode Current Relay or Screen Current Relay	—	Anode Current Relay & Grid Current Relay	—

Figure 37. Tetrode and pentode protection chart. This chart indicates the location of a suitable relay which should act to remove the principal supply voltage from the stage or transmitter to prevent damage to the tubes.

the maximum allowable screen dissipation will almost always be exceeded and tube damage will result.

Figure 38 is a chart for the protection of a triode tube. This chart covers the grid driven triode amplifier and the high- μ (zero bias) cathode driven triode amplifier. **DRIVE VOLTAGE MUST NEVER BE APPLIED TO A ZERO-BIAS TRIODE AMPLIFIER WITHOUT ANODE VOLTAGE BEING APPLIED.**

CIRCUIT FAILURE	TRIODE		ZERO-BIASED TRIODE
	FIXED GRID BIAS	RESISTOR GRID BIAS	
Loss of Excitation	No Protection Required	Anode Overcurrent Relay	No Protection Required
Loss of Antenna Loading	RF Output Detector & Relay	RF Output Detector & Relay	Grid Over-Current Relay
Excess Antenna Loading	RF Output Dctetro & Relay	RF Output Detector & Relay	RF Output Detector & Relay
Failure of Anode Supply	No Protection Required	No Protection Required	Grid Over-Current Relay
Failure of Grid Bias Supply	Anode Over-Current Relay	—	—

Figure 38. Triode protection chart. This chart indicates the location of a suitable relay which should act to remove the principal supply voltage from the stage or transmitter to prevent damage to the tube or transmitter.

3.9.1 Arc Protection

The damaging effects of electrical arcing is an important subject for any circuit designer.

Capacitors used as filters in high voltage power supplies can store substantial energy, more than enough to damage tubes and other circuit components. The level of energy available can be determined in the formula $Q=1/2CE^2$, where Q is the stored energy in Joules (or Watt-seconds), C is the power supply filter capacitance in Farads and E is the dc voltage across the capacitor. It is obvious from this formula that increasing the voltage increases the stored energy exponentially.

As an example, a 20 microfarad capacitor charged to 5 kilovolts represent 250 Joules of stored energy; enough to melt metal or ceramic-to-metal seals, which are commonly used as part of a vacuum envelope in power grid tubes. Catastrophic loss of vacuum during operation causes rapid oxidation of the filament or cathode, making the tube inoperable and usually beyond repair.

Arcing is a phenomena that occurs when the atmosphere between conductors breaks down in the presence of a high voltage field

and becomes ionized. Electrical charges build up and eventually cross the gap between the conductors to form a path of hot, ionized gas with an equivalent resistance of a few tenths of an ohm. The presence of metallic ions from vaporized metal that may exist at the terminus of an arc further reduces the impedance. Unpredictable by nature, arcing under fault conditions can be the result of power surges, parasitic oscillations, unloaded conditions in an RF amplifier, dirt build-up across a ceramic or a number of other causes.

Internal arcing in a power grid tube is the result of the same mechanism, but may be aggravated by residual gases within the vacuum space. Arcing can also be caused by exceeding the breakdown rating between tube elements.

Internal arcing at the cathode of oxide cathode tubes may eventually occur as a result of normal evaporation of oxide material. The remaining oxide must then supply considerably more current per unit area than for which it was designed. For instance, if 50% of the oxide becomes depleted, then the remaining oxide material must supply twice the normal current.

The process of an internal arc caused by sparking at the cathode can best be understood by first examining the electrical function of the cathode material. There is an Ohmic factor at the oxide barrier; that molecular region where the oxide material meets the base metal and forms an electrochemical junction. Oxide material has relatively high electrical loss. When electron current through the resistive oxide layer is increased substantially above normal limits, localized heating occurs due to I^2R losses. This heating takes place at microscopic points on the cathode surface. The additional heat, added to oxide material already at high working temperature, will cause a rapid decomposition of cathode material, which then releases gases that will ionize in the presence of high voltage, leading to an internal arc. Additional heat is generated by the arc energy, releasing more gas. A sustained arc can result if fuses, overload circuit or other protection devices are not present to interrupt the high voltage to the tube.

Although good design measures can be taken to prevent arcing, ionization and arcing will occur whenever high voltage exists. The task then, to prevent internal damage to the tube, is to remove or divert the stored energy away from the power grid tube if an arc should occur.

Since power grid tubes are made in various sizes, internal tube elements will have varying mass. Some tubes can safely withstand more arc energy than others. Two general types of fault protection are used for power grid tubes, depending upon whether they are operated at low energy or high energy.

Systems using power grid tubes with oxide cathodes and tubes with thoriated tungsten with small diameter wire grids generally operate at low power and are classified as low energy tubes. Anode dissipation ratings for low energy tubes typically range from a few Watts up to approximately 25 kW. Generally, any arcing associated with these tubes should be limited to less than 4 Joules of energy if catastrophic damage is to be prevented.

A simple technique for protecting low energy tubes is to use a protection resistor in the high voltage supply lead, coupled with a fuse or circuit breaker to quickly remove power from the high-voltage transformer. The resistor should have a fairly high power rating. Bulk-carbon Carborundum™ or large wire wound resistors are recommended. Typical values range from 10 to 50 ohms. Additionally, in a tetrode, a protection resistor should also be used in the screen power supply, especially if it contains considerable filter capacitance.

It is recommended that the largest resistance value that will still conduct the full average (dc) anode current without overheating be used. The fuse or fast acting circuit breaker completes the protection circuit. This prevents follow-on energy, which must be interrupted to prevent the resistor from failing.

Power grid tubes operating at substantially higher voltages are subjected to far greater arc energies than low power tubes. Large power grid tubes used for high energy operation are generally capable of withstanding arc energy of up to 50 Joules of total energy.

High energy power grid tubes are generally protected by a special circuit, the so-called “crowbar”, that shunts all the stored power supply energy to ground. A series protection resistor will not work for these applications, because anode currents used for these tubes are typically 20 to 200 A, which would cause excessive voltage drop across the resistor.

A typical crowbar uses a shunt device such as a hydrogen thyratron, an ignitron or a triggered spark gap. If an arc is detected, the crowbar device is activated, shorting the power supply filter

capacitor to ground. The crowbar must be fast acting and be able to absorb large levels of energy without failure.

A specially designed high-voltage switch-mode power supply can be used as an alternative to crowbar shunt protection. Operating at a high switching rate, this type of supply requires very low values of filter capacitance. Therefore the stored energy is low. At the same time, the ability to quickly interrupt the oscillator or switching devices (within tens of microseconds) insures that follow-on energy is minimal.

Eimac Application Bulletin AB-17 "Fault Protection" is available on Eimac's web site (www.eimac.com) under Applications.

3.10 KEYING

Although "keying" in the traditional sense of turning an amplifier on and off to generate Morse code is no longer used, there are instances where it is useful to switch an amplifier on and off. Thus, this section is included for reference.

The tetrode and pentode power amplifier can be keyed using the same basic principles employed with any power amplifier. In addition, the screen electrode provides another low power circuit where keying can be introduced. Suitable filters must be used so that the make and break is slow enough to avoid high frequency sidebands known as "key clicks." The usual "key click" filter techniques apply.

There are several good methods of controlling the tetrode and pentode RF power amplifier when exciter keying is used. With the screen voltage fixed and with fixed bias greater than cutoff, the tube will pass no current when the excitation is removed. A low or medium- μ triode amplifier can be keyed in the same manner. With the high- μ (zero bias) triodes, it is even simpler. When the drive is removed, the anode current falls to the normal, safe quiescent anode current.

It is also possible to key the exciter stage when the screen voltage for a tetrode or pentode is obtained through a dropping resistor and grid leak bias is used (see Figure 39A). In this system a high transconductance, low- μ triode is connected between screen and cathode, and the controlling bias for the small triode is taken from

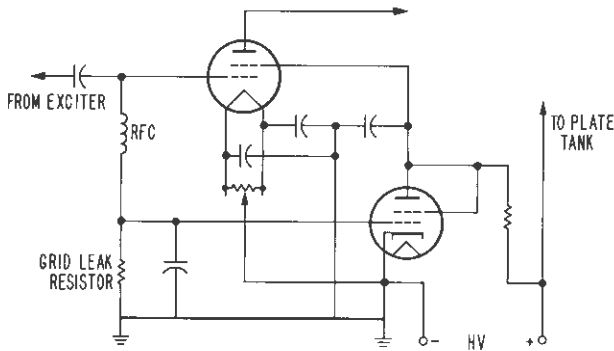


Figure 39A. Screen voltage control circuit for exciter keying or protection against loss of excitation when supplying screen from high voltage source.

the tetrode or pentode bias developed in the grid leak resistor. When normal excitation is present on the RF amplifier and grid bias is developed, the triode control tube is cut off and the screen voltage circuit operates normally. If excitation voltage is removed from the tetrode or pentode power amplifier, the bias voltage developed in the resistor drops to zero and the control triode becomes conducting. The current drawn by the triode control tube will increase the voltage drop across the series screen resistor and lower the screen voltage to a very low value.

There is still some voltage on the tetrode or pentode screen and a small static anode current flows. Usually this remaining current is not enough for the anode dissipation rating to be exceeded. This value can be reduced further by the use of a second control triode in parallel with the first (this redundancy is also insurance against failure of the single triode), by putting a gas regulator tube in series with the lead to the screen before the screen RF by-pass capacitor, or by introducing a small amount of fixed bias on the power amplifier between the grid resistor and the power tube grid.

Figure 39B shows a method of keying a tetrode or pentode RF power amplifier where the low voltage power supply for the screen of the power tube and for the anode and screen of the driver stage is keyed directly. This permits keying in a relatively low voltage, low current circuit. The key click filter capacitor, resistor, and choke are simple and assure positive control of the keying wave shape.

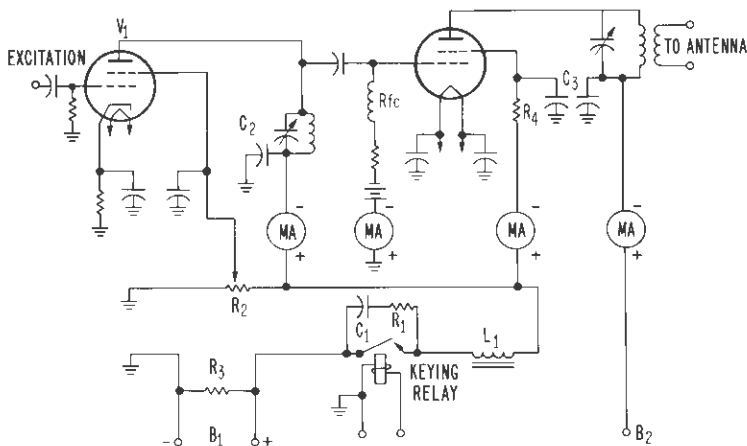


Figure 39B. A typical method of keying a tetrode or pentode amplifier.

One important point should be made regarding keying. Some circuit designers may be tempted to disconnect the cathode return to the anode supply, B-, to key-off a power grid tube. This action can have extremely serious consequences. If the cathode return (or the center tap of a filament transformer used with a thoriated tungsten filament tube) is open-circuited, the potential at the cathode will rise to nearly that of the anode supply. Internal arcing from the cathode or filament to the control grid would be likely. Under such conditions, a tube will saturate, or "run-away", and draw excessive anode and screen currents. External arcing in the socket would also be likely, because insulation ratings would be inadequate for fault conditions of this type. Cathode circuit components such as bypass capacitors and the secondary winding of the filament or heater transformer may also be subject to voltages well beyond normal design values. **To prevent such possibilities, the cathode return must always have a secure path to ground and, ultimately, the B- of the anode supply.**

Similar conditions could exist if a fuse is placed in the cathode return. If the fuse should open, the circuit has infinite resistance, causing the cathode voltage to raise to nearly the anode potential. To prevent this, one should shunt the fuse (if used) with a fairly high value resistor. The current through the resistor will develop a bias voltage that will act to cut-off the anode current, or hold it to a very

low value. A resistance of approximately 5,000 ohms is used in most applications to provide this protection. A better solution would be to use a high voltage fuse in the anode supply instead of fusing the cathode return.

3.11 AMPLITUDE MODULATION

A triode RF amplifier can be anode, grid or cathode modulated. Anode modulation is the type most extensively used. A triode may also be operated under a linear set of conditions for amplifying an amplitude modulated driver.

A tetrode or pentode RF amplifier can be anode and screen, screen grid, control grid or cathode modulated. Usually the system chosen is a combination of any of the previously listed

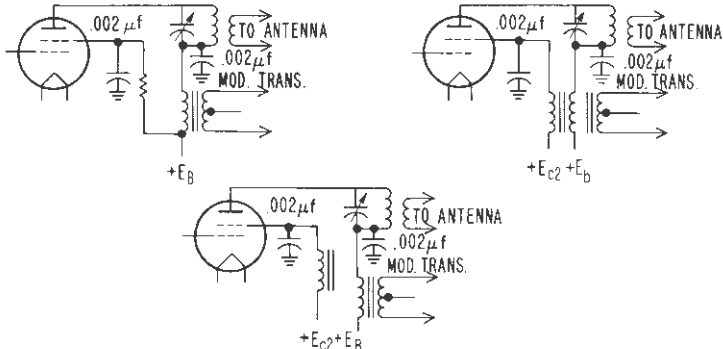


Figure 40. Basic screen and anode modulation circuits.

techniques. The most extensively used technique is a combination of anode and screen modulation. Often additional modulation must be provided on the control grid to reach 100 per cent modulation on the positive peaks. Figure 40 shows three of the basic anode and screen modulation circuits.

In anode and screen modulation it is necessary to introduce not only amplitude modulation of the anode voltage, but also to develop 70 to 100 per cent amplitude modulation of the screen voltage for 100 per cent carrier modulation. Modulation of the screen voltage can be developed in one of the following three ways:

- (a) By supplying the screen voltage through a dropping resistor connected to the unmodulated dc anode supply.
- (b) When a low voltage fixed screen supply is used, a modulation choke is placed in series with the supply. In the case of voice modulation, this is about a 5 to 10 henry choke.
- (c) A third winding on the modulation transformer designed to develop the required screen modulation voltage.

It is interesting to note that in all three cases the screen of the tetrode or pentode tube supplies the necessary audio power. During the portion of the modulation cycle, when the anode voltage is increased, the screen current decreases. If the screen is supplied through an impedance, such as the screen dropping resistor of a modulation choke, the voltage drop in this series impedance becomes less and the screen voltage rises in the desired manner. On the part of the modulation cycle when the anode voltage is decreased, the screen current increases causing a greater voltage drop in the screen series impedance and thus lowering the voltage on the screen of the tube. The screen by-pass capacitor value in the Class C stage is a compromise between good RF bypassing and the shunting effect of this capacitance on the screen modulation circuit.

Where 100 per cent modulation capabilities are desired, the tube efficiency under carrier conditions is about half that expected in the RF amplifier when anode and screen modulation is used. This efficiency is usually on the order of 35 to 40 per cent. Grid and screen modulation is used when there is a desire to save on physical size and cost of the modulation source.

When grid modulation is used, the screen voltage and grid bias must be taken from sources with good regulation. This usually means a separate low voltage power supply source. In the case of screen modulation, the grid bias should be taken from a grid leak bias resistor to provide some drive modulation.

The output of a pentode Class C amplifier can be controlled by applying to the suppressor grid a modulating voltage superimposed upon a suitable bias. As the suppressor grid in such an arrangement becomes more negative, the minimum instantaneous anode potential at which current can be drawn to the anode is increased. Thus, as the modulation varies the suppressor-grid potential, the output varies.

The suppressor-grid modulated amplifier has about the same anode efficiency as the grid modulated Class C amplifier. The

overall efficiency is somewhat less because of the screen-grid losses. The modulating power is about the same. The linearity of modulation is not particularly high.

The screen-grid losses are higher because as the anode potential decreases, the current to the screen increases. This tendency toward high screen losses is the factor that usually limits the output power obtainable from a suppressor-grid modulated amplifier.

3.12 POWER SUPPLY CONSIDERATION

The power supply requirements for a triode are straightforward. The degree of regulation and ripple depends upon the requirements of the system. In the case of a linear RF amplifier, it is important to have good anode power supply regulation. Without good regulation, the anode voltage will drop during the time the anode is conducting current heavily. This tendency for the voltage to drop will cause "flat topping" and will appear as distortion in the output. In a push-pull audio application where grid current flows, it is important to keep the grid circuit resistance to a minimum. If this is not done, positive peak clipping will occur. In the case of the tetrode and pentode, the need for screen voltage introduces some new considerations and provides some new possibilities.

Voltage for the screen grid of a low power tetrode or pentode can readily be taken from the power supply used for the anode of the tube. In this case, a series resistor, or potential dividing resistor, is chosen so that with the intended screen current flowing the voltage drop through the resistor is adequate to give the desired screen voltage. The potential dividing resistor is the preferred technique for those tubes with significant secondary screen emission (see Figure 6).

It is possible to take the screen voltage from a low voltage supply, frequently using an already available source in the equipment. There is considerable latitude so that an available voltage can be used. Sometimes a combination might be employed, where a dropping resistor is used in conjunction with a low voltage or intermediate voltage supply. Frequently a combination of series resistor and voltage source can be chosen so that the rated screen dissipation will not be exceeded regardless of the variations in screen current. With a fixed screen supply, there are advantages in using an appreciable amount of fixed grid bias so as to provide protection against loss of excitation, or for cases where the driver stage is being keyed.

If the screen voltage is taken through a dropping resistor from the anode supply, there is usually little point in using a fixed grid bias because an unreasonable amount of bias would be required to protect the tube if the excitation failed. When a screen dropping resistor is used, most of the bias is normally supplied through a grid resistor and other means are used for tube protection.

Under operating conditions with normal screen voltage, the cutoff bias is low (screen voltage divided by the screen μ). When a stage loses excitation and runs statically, the screen current falls close to zero. (See static curves of tube in question.) If the screen voltage is obtained through a simple dropping resistor from the anode supply, the screen voltage will then rise close to full anode voltage. Because the cutoff bias required is proportional to the screen voltage, the grid bias required will be much greater than the amount of bias desired under normal operating conditions.

The power output from a tetrode or pentode is very sensitive to screen voltage. For this reason, any application requiring a high degree of linearity through the amplifier requires a well regulated screen power supply. A screen dropping resistor from the anode supply is not recommended in such applications.

The suppressor grid power supply requirements are quite similar to the control grid power supply. The suppressor grid intercepts very little current, and therefore a low power supply may be used. Any variation in suppressor voltage due to ripple or lack of regulation will appear in the output of the amplifier due to the suppressor grid modulation of the anode current.

3.13 STABILIZING THE AMPLIFIER

3.13.1 Testing for parasitic oscillations

In the case of an RF amplifier, it will be necessary to investigate not only the possibility of self-oscillation, but also lack of feedback on the fundamental frequency. The basic steps of checking for self oscillation are threefold:

- (a) The amplifier should be operated without RF excitation and without fixed grid bias, with light loading and with low voltages applied to the anode and to the screen of a tetrode or pentode.

The voltage should be high enough to develop full anode

dissipation. For this test, grid leak bias should be used. If the screen and anode voltage supplied cannot be adjusted directly to low voltages, suitable series resistance should be used, either in series with the rectifier output or transformer primary so that the voltages developed at the tube will be low.

Light bulbs of the correct size will serve as resistors in series with the primary of the rectifier transformers. The RF circuits should be tuned off-resonance to see if self-oscillation of the amplifier can be started. The indication of any current by the grid milliammeter means that self-oscillation is present.

- (b) The frequency of the self-oscillation must be determined. A simple neon bulb taped to a long rod (for safety reasons), will, indicate whether the frequency of oscillation is high or low when it is held near the anode of the tube. The lower the frequency the more orange will be the glow. A purple color indicates a VHF or UHF oscillation. With a little experience, it will become possible to guess the approximate frequency very closely. A wavemeter or an oscilloscope will then determine the exact frequency.
- (c) After the frequency of oscillation is measured, it remains to identify the circuit supporting the oscillation. The circuit must then be altered without disturbing the performance of the amplifier on the normal frequency of the amplifier.

3.13.2 Correction of Parasitic Oscillations

The usual self-oscillations in RF power amplifiers have been found to fall in the following three classes:

- (a) Oscillation at VHF from about 40 MHz to 200 MHz, regardless of the normal frequency of the amplifier.
- (b) Self-oscillation on the fundamental frequency of the amplifier.
- (c) Oscillation at a low radio frequency below the normal frequency of the amplifier.

The low frequency oscillation in an amplifier usually involves the RF chokes, especially when chokes are used in both the output and input circuits.

Oscillation near the fundamental frequency involves the normal resonant circuits, and brings up the question of neutralizing the RF power amplifier. This general subject is discussed under "Neutralization," Section (5).

When a parasitic self-oscillation is found on a very high frequency, the interconnecting leads of the tube, the tuning capacitor and the by-pass capacitors are involved. This type of oscillation does not usually occur when the power amplifier is designed for operation in the VHF region and where the RF circuits external to the tube have negligibly small tuning capacitors. Without tuning capacitors, the highest frequency oscillating circuit possible is then the fundamental, and there would be no higher frequency circuit available for the parasitic. The only exception would be where higher order modes of transmission line circuits might provide a parasitic circuit.

VHF parasitic oscillation may occur in amplifiers where the RF circuits consist of coils and capacitors. In Figure 41, the capacitors and associated grid and anode leads act as inductances for the parasitic oscillation circuit. The tube capacitances help form the tuned-anode tuned-grid oscillation circuits. The circuit is indicated by the heavy lines in Figure 41.

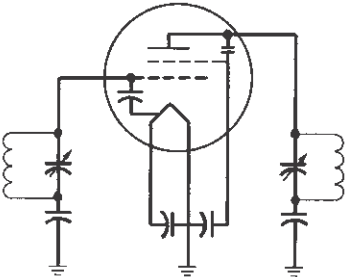


Figure 41. Usual circuit supporting VHF parasitic oscillation in HF RF amplifiers.

There are several straightforward ways to suppress VHF parasitic oscillation. In general, it will probably be more easily suppressed if the general layout and by-passing methods indicated earlier are followed.

It turns out that the frequency at which a VHF parasitic oscillation occurs is well above the self-neutralizing frequency of the tube (see Section 5). However, if the self-neutralizing frequency of the tube can be increased and the frequency of the parasitic lowered,

complete suppression of the parasitic will result, or its suppression by resistor-inductor (R-L) parasitic suppressors may be made easier.

It is also possible to predict fairly closely with a grid dip wavemeter the parasitic frequency to be expected in a given equipment. The circuit should be complete and with no voltages on the tube. Couple the meter to the anode or screen lead and determine the resonant frequency. The following two methods of eliminating the VHF parasitic oscillation have been used successfully:

- (a) Place a small R-L combination in the anode lead between the anode of the tube and the tank circuit (see Figure 42). The R-L combination is usually made up of a non-inductive resistor of about 25 to 100 ohms, shunted by three or four turns of wire or copper strap, frequently wound right around the resistor. In some cases (with a common cathode - grid driven circuit), it may be necessary to use such a suppressor in both the anode and grid leads. The R-L combination operates on the principle that the resistor

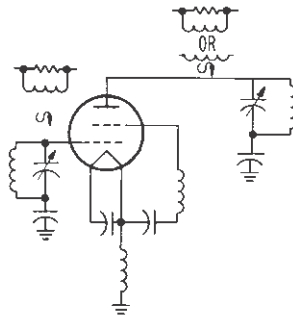


Figure 42: Placement of parasitic suppressors to eliminate VHF parasitic oscillations in HF r-f amplifiers.

loads the VHF circuit but is shunted by the coil for the lower fundamental frequency. IF it is found that the resistor runs too hot, the inductance in the coil may be excessive. Just enough inductance should be used to suppress the parasitic and no more; otherwise, excessive fundamental energy will cause the resistor to overheat. Once the parasitic has been suppressed there will be no parasitic voltage or current present. Therefore, there is no parasitic power to be dissipated.

- (b) Use small parasitic chokes in the anode lead (see Figure 42).

The size of this coil will vary considerably depending upon the tube and the circuit layout, and may run from about four to ten turns of about a one-half inch diameter. The presence of this choke in the frequency determining part of the circuit lowers the frequency of a possible VHF parasitic so that it falls near the self-neutralizing frequency of the tube and bypass leads. In addition to varying the size of the suppressor choke, the amount of inductance common to the screen and filament in the filament grounding strap may be a factor. This can be varied simultaneously with the suppressor choke.

Of the two methods indicated above for suppressing VHF parasitic oscillations, the first one is probably the simpler to use and has been widely employed. The popular belief that a "low Q VHF parasitic choke" is beneficial can be de-bunked by realizing that any parasitic choke has low Q, because of the loading caused by the low resistance across the choke (20-50 ohms). Special nichrome wire is unnecessary - regular copper wire or strap is recommended for the inductor in a parasitic choke.

In rf amplifiers, which operate between 25 MHz and 50 MHz, the selection of a proper parasitic suppressor is more difficult, especially with high output power. A parasitic inductor for these frequencies may consist of a simple loop or "U" shaped strap (approximately 0.5 inches in width and 6 to 8 inches in length, prior to bending) with a resistor connected across the open end. A bulk carbon rod (Carborundum) type of resistor (20 to 50 Ohms) is recommended.

VHF amplifiers operating above 50 MHz seldom use a parasitic suppressor, because excessive power would be induced in the "R" component and the "L" component would have an exceedingly low value. Fortunately, the excellent isolation provided by many modern ceramic-metal tubes allow stable operation at high VHF frequencies without a parasitic choke.

The procedure of checking for self-oscillation in an RF power amplifier described previously will normally show up most trouble and allow for its correction. If, however, the correction is marginal, it may sometimes happen that under operating conditions the self-oscillation will be triggered off. The oscillation may occur only on the peaks of amplitude modulation or on keying surges. By observing the RF envelope on a cathode ray oscilloscope, the oscillation can usually be seen. The trouble can be fully eliminated by pursuing further the outlined corrective procedure.

A more difficult self-oscillation to locate is one occurring on a harmonic of the fundamental frequency and occurring only when

the stage is operating. It will show up when testing for the presence of abnormal power in the harmonics under operating conditions.

In audio amplifiers employing tetrodes or pentodes, small non-inductive resistors of about 100 ohms resistance should be placed in series with the anode, and possibly the grid as well, in case self-oscillation of the amplifier occurs in the very high frequency portion of the RF spectrum. Should the audio or dc voltage drop in the resistor be objectionable, it can be shunted with a small coil.

Another form of commonly encountered self-oscillation is known as “dynatron” oscillation. Dynatron oscillation is caused when any electrode in a vacuum tube has negative resistance. Secondary emission characteristics of the screen grid in a tetrode were discussed in Section 2.2; it was pointed out that at times there may be more electrons leaving the screen grid than are arriving. If the screen voltage is allowed to increase under these conditions, even more electrons will leave the grid; the phenomenon implies a negative resistance characteristic. If there is high alternating current impedance in the circuit from the screen grid through the screen grid power supply, and from the anode power supply to the anode, dynatron oscillation may be sustained.

Dynatron oscillation typically occurs in the region of one to 20 Hz. This low-frequency oscillation is usually accompanied by another oscillation in the 1000-2000 kHz region. Suppression of these oscillations can be accomplished by placing a large bypass capacitor (1000 μ F) across the output of the screen grid power supply. The circuit supporting the oscillation can also be detuned by a large inductor. Increasing the circuit losses at the frequency of oscillation is also effective.

3.14 NOISE IN TUBE AMPLIFIERS

Power grid tube designers often come across descriptions of various types of noise associated with electron tubes. Phenomena such as a random noise from electron-thermal agitation or shot noise have been well described. These types of noise, which affect small signal, high voltage gain tubes, are usually of very little concern for large triode and tetrode power amplifier applications. Filament induced random noise in the output of power amplifiers is, typically, so far below the full output value that it is virtually non-measurable using common instrumentation

Power grid tubes with thoriated-tungsten filaments may, however, exhibit low levels of hum in the output if ac filament power is used.

This phenomena is well understood¹¹; basically, electrons at the crest of the negative voltage excursion are accelerated more at one end of the filament than the other end. This ac difference slightly modulates overall cathode emission and the net plate current will, therefore, contain energy at the second harmonic of the filament ac voltage. The use of center-tapped secondary winding on the filament transformer, with the center tap representing the cathode or common return point, will reduce hum to the point that it is generally insignificant.

However, when a complete absence of any ac-related component in the output is required, dc filament power is recommended. An example of this is analog (NTSC) television transmitters. Any line frequency AM component in the filament may beat with the precise sync pulses that are being amplified in the tube. Television transmitters, which use power grid tubes in the power amplifier stage, typically use a three-phase, rectified and filtered dc power supply for filament voltage.

Radar switch tubes often have very low noise requirements. Mechanical movement of the filament, grids or anode may induce artifacts in the anode current during conduction. Liquid cooled tetrodes, with specially designed, low-turbulence anode coolers that reduce mechanically induced noise, are available.

FM broadcast transmitter amplifiers sometimes encounter a form of noise called synchronous noise (also called "Incidental AM"). Synchronous noise occurs when a FM signal is applied to a narrow-band circuit, causing an AM component to be added to the signal. This noise is generally not noticeable on the main FM program, but it can cross-modulate any sub-carrier (SCA) program material.

Synchronous noise is generally the result of using a high-Q tuned circuit, typically in the grid input of the tetrode used in the IPA stage (although it may sometime be developed in the Power Amplifier input circuit). Saturating the grid by slightly over-driving the input, which, thereby, loads the circuit and reduces Q, minimizes this effect.

Proper input tuning is critical in most cases to reduce synchronous noise. Grounded-grid FM amplifiers generally never exhibit synchronous AM noise, because the bandwidth of the cathode input circuit is significantly greater than the applied FM signal bandwidth, and the circuit Q is quite low.

11 F. E. Terman, Radio Engineers Handbook 13th Edition, p. 617