UNRAVELING TRANSMITTER RF POWER AMPLIFIER MYSTERIES

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Earlier transmitters are always tuned the same. Peak the grid, dip the plate, adjust loading for correct output as indicated by plate current, power output, dissipation, and efficiency.

Now it's different as each transmitter has a different method of tuning. Some dip the DC current, some dip the DC voltage, others peak the output power, or a few compute a predetermined power amplifier efficiency.

The reason for these changes in tuning methods can be understood by studying the available categories of active devices that are currently being used in transmitters.

There are four general categories of active power devices used as final amplifiers for todays transmitters. They are grouped here in terms of the effects of the devices output voltage on their output current.

The four devices are:

Triodes

Tetrodes with poorly regulated screen grid voltages

Tetrodes with well regulated screen grid voltages

Solid state devices (junction transistors, FET's, etc.)

These four types can be broken into two headings:

(1.) Devices in which the output voltage has much effect on output current. They are:

(A.) The triode, due to its inherent nature.

(B.) The tetrode, due to a poorly regulated screen grid voltage.

(2.) Devices in which the output voltage has minimum effect on output current. They are:

(A.) The tetrode, due to a well regulated screen supply.

(B.) Solid state devices, due to their inherent nature.

A block diagram of a typical RF power amplifier is shown in Figure 1. As the output resonant circuit of the transmitter's PA is tuned, the RF voltage level of the resonant circuit reaches a maximum when the resonant frequency (F_R) of the network equals the frequency of the active device's current pulses (the transmitters' resonant frequency or F_0). See Figure 2.

In Figure 3, the RF voltage is riding upon the DC supply voltage that is applied to the output port of the RF power amplifiers' active device. This combination is the voltage that appears at the output port of the active device. The DC voltage is fed to the active device from the DC power supply, and the RF voltage is the result of the output current pulses flowing through the resonant output network.

Review of Class C

To explain the effect of the plate voltage upon plate current in the various devices, a review of Class C is necessary. In Class C, plate current flows during the positive peak of the input cycle. This is the conduction angle, which is less than 180 degrees in Class C. In Figure 4, the transfer curve for a triode or tetrode is shown. The DC bias is far below cutoff and the RF drive voltage drives the grid positive into grid conduction. Thus the plate current is in the form of narrow pulses.

Figure 5 shows the grid voltage, plate current, and plate voltage waveforms. The resonant output network converts the plate voltage variation into a sine wave of the same frequency as the grid input signal.

Notice that instantaneous plate voltage is at its minimum when the plate current flows, and when the plate resonate circuit is tuned the amplitude of the RF plate voltage and thus the minimum instantaneous plate voltage value changes. See Figure 2 and 3.

This low value of instantaneous plate voltage can have an effect on plate current since this is the only time plate current flows in Class C operation. This forms the bases for plate voltage effecting plate current.

The Triode

For a triode Class C amplifier, the plate voltage (E_p) is constant, being set by the H.V. supply. See schematic Figure 6 and wave forms Figure 5. The DC plate current (I_p) is the average of the pulses of Class C plate current (I_p maximum). The level of Class C plate current pulses (I_p maximum) depends upon the grid drive, and the instantaneous plate voltage (E_p minimum) which is present when the plate current flows.

If the plate resonate tuning is at optimum value, its resonate frequency (F_R) equals the amplifiers drive frequency (F_0), maximum RF plate voltage (e_p) will occur causing E_p minimum to dip very close to zero volts. This minimum instantaneous plate voltage that occurs when plate current flows reduces the level of I_p maximum and thus the level of I_p.

Analysis:

The increased e_p increases RF output power

$$P_O = \frac{e_p^2}{2Z_p}$$

The reduced I_p decreases DC input power

$$Pin = E_p I_p$$

Plate efficiency is therefore increased

$$Eff = \frac{P_{out}}{P_{in}}$$

and plate dissipation is reduced

$$P_{diss} = P_{in} - P_{out}$$

When the output resonate circuit is detuned e_p is reduced and thus the plate voltage does not swing as close to zero volts. This higher value of E_p minimum causes greater I_p maximum to flow increasing I_p. This results in:

(1.) A decrease in P_{out}

(2.) An increase in i_p, I_p, and P_{in}

(3.) A decrease of efficiency

(4.) An increase of dissipation

NOTE: i_p is the peak value of the equivalent RF plate current sine wave.

The Triode Plate Family of Curves

Figure 7, the plate family of curves for a triode, shows the reason why E_p effects I_p. Notice the rising curves on the graph, they represent a fixed grid voltage. If the grid voltage is held constant and the plate voltage changes, a change of plate current occurs.

Example:

In the 3CX2500F3, the grid is held constant. The table shows the variation of plate current as plate voltage is changed.

If E_g = 0V, then:

I_p = 0.5A when E_p = 1000V

and I_p = 8.0A when E_p = 5000V

If E_g = -50V, then:

I_p = 1A when E_p = 2KV

and I_p = 6A when E_p = 5KV

From the above discussion, it should be obvious how the correct tuning of a triode RF power amplifier is indicated by a dip of DC plate current.

The Tetrode

If a tetrode is used as an RF amplifier, the block diagram is the same as Figure 1. The RF voltage versus current, and the combination of DC and RF voltage shown in Figures 2 and 3 are the same as for the triode. The transfer curve shown in Figure 4 and the waveforms on Figure 5 are also the same.

Figure 8 shows a family of E_p I_p curves for a typical tetrode. Notice that the plate voltage has minimum effect on plate current if the grid voltage is held constant.

Example:

Tube.....4CX3000A

E_g = -50V, then:

I_p = 4A when E_p = 1KV

and I_p = 6A when E_p = 7KV

If the tetrode plate current is relatively independent of the plate voltage, a question arises as to how the proper tuning of a tetrode RF power amplifier can be indicated by a dip of plate current as it can in a triode. The key is the relationship of the instantaneous plate voltage to positive screen grid current, and the relationship of the positive DC screen current to the value of the DC screen grid voltage.

Screen Current

The screen grid is normally operated positive and is physically located in the shadow of the control grid to minimize positive Isq. See Figure 9. Each electron emitted from the cathode is repelled by the negative charge on the control grid, attracted by the positive charge on the screen grid, and once it passes the control grid by it is attracted to the higher positive plate voltage. See Section A of Figure 9. When the plate voltage swings close to the value of the DC screen grid voltage (Esg) the electrons that pass the control grid experience a stronger attraction for the screen grid since it is physically closer. This causes an increase of positive screen grid current since this grid now intercepts more electrons. The lower the plate voltage swing the greater will be the positive screen grid current, and if the plate swings very far below the value of Esg there is a danger that an excessively large screen current will overdissipate the screen grid.

Power screen grid = Esg x Isg

where Isg = Positive DC screen current

The screen grid can emit electrons (negative screen current) when the plate voltage is at its maximum positive excursion, but this is usually a smaller current flow than the positive screen current mentioned above. The DC screen current (Isg) is an average between the positive and negative screen grid currents.

NOTE: The screen and control grids can emit electrons because of their high temperature, being physically close to the white hot cathode.

Relationship Between Screen Voltage and Plate Current

Figure 10 shows a set of tetrode transfer curves that change their horizontal position due to a change in screen voltage. As the screen voltage increases the space charge around the cathode decreases, increasing the amplification factor of the tube. The transfer curve shifts to the left changing the relationship of the control grid voltage (DC and RF) to plate current.

As the screen voltage is increased, the same bias voltage and RF drive level cause more plate current to flow.

In Figure 11A and 11B, two plate families of curves are drawn for the same tube. The difference in the two sets of curves is caused by the change of screen grid voltage. The graphs also show the increased sensitivity, more plate current for the same control grid voltage, as the transfer curves of Figure 10. Notice that although the screen voltage causes the plate family of curves to change their position on the graph, a change of plate voltage still has minimum effect on plate current when the control grid voltage is held constant.

Summary of the Tetrode

Plate voltage has minimum effect on plate current. Plate current is controlled by the value of DC screen grid voltage, DC control grid bias voltage, and the level of RF drive voltage presented to the grid or cathode.

Tetrode RF Amplifier with a Poorly Regulated Screen Grid

Figure 12 is a schematic of a tetrode RF amplifier. The resistor Rsg shown between the screen grid and its supply causes the screen grid voltage to be poorly regulated, thus it changes when Isg changes.

From our earlier discussion, it was noted that the level of plate voltage swing (E_p minimum) changed as the output resonant circuit was tuned, and that as E_p minimum approached the level of Esg, positive screen grid current (Isg) flowed.

As the output network is tuned closer to F_0, the E_p minimum dips lower, and the positive Isg increases. The increase of Isg causes a larger voltage drop across Rsg reducing Esg. As Esg is reduced, the amplitude of I_p maximum and I_p are reduced, thus correct plate tuning is indicated by minimum I_p which is caused by the poorly regulated Esg.

The resistance Rsg prevents screen grid overdissipation by limiting input power to it.

Since loading of this amplifier will effect e_p, which effects Isg, which effects Esg, which effects I_p maximum and I_p, there is one value of loading that will produce maximum output power.

Loading analysis:

Loading too heavy lowers Z_p which reduces e_p,

Lowering P_{out}

$$P_{out} = \frac{e_p^2}{2Z_p}$$

Loading too light increases Z_p and e_p which ultimately

causes a large reduction of I_p maximum and i_p (the equivalent RF

plate current) which lowers P_{out}

$$P_{out} = \frac{i_p^2 Z_p}{2}$$

where i_p = peak value of the equivalent RF plate current

Tetrode RF Amplifier with a Well Regulated Screen Grid

If the Rsg on Figure 12 is replaced with a piece of wire, the screen grid will be well regulated. The screen voltage will not change when the DC screen current varies, such as with tuning or loading.

If tuning or loading causes E_p minimum to dip closer to Esg, Isg will increase and Esg will remain constant. A danger now exists that excessive Isg will cause overdissipation of the screen grid so that a screen current overload circuit must be added to protect the tube.

Since Esg is constant, I_p does not change with tuning or loading so both are adjusted for maximum output power.

Analysis:

Constant E_p and I_p means fixed input power to the plate circuit. Thus proper tuning and loading will produce maximum efficiency and power output possible for the amplifier.

Light Loading

Light loading increases Z_p and e_p which increases positive Isg. If loading is too light, there is danger that a screen grid current overload trip might occur, but light loading and maximum Isg usually provide better efficiency.

Heavier Loading

Heavier loading produces higher I_p, lower e_p, and lower Isg, but can reduce efficiency because the higher I_p will increase input power while lower e_p will reduce output power.

NOTE: The higher I_p is due to increased grid drive required to make power and not directly due to loading changes.

When used as RF power amplifier, vacuum tubes have two disadvantages.

(1.) Required heater power for the cathode. This power does not appear in the RF output, it adds to the cooling load of the transmitter and lowers the overall efficiency.

(2.) The negative swing plate voltage must be kept above a minimum value so that the tube will continue to draw plate current and produce output power. This also wastes plate power and lowers efficiency.

Introduction to Class D Tube Circuits

Plate dissipation depends upon the value of plate voltage that is across the tube when plate current flows.

$P_{diss} = E_{pk} I_p$

where: E_{pk} = plate to cathode voltage

and I_p =plate current.

Figure 13 shows the E_p and I_p waveforms for tube type Class A, B, C, and D operation. The shaded area under the plate waveform, which represents the duration of plate current flow, is proportional to plate dissipation.

Observation confirms the known fact that Class A is the least efficient and Class D is the most efficient of the four classes.

Class D

The form of Class D referred to here is used in vacuum tube amplifiers, triode or tetrode, and is sometimes referred to as Class F or the Tyler high efficiency amplifier.

This amplifier, as shown in Figure 14, features a third harmonic resonator in the plate and cathode circuit. Some amplifiers have the third harmonic resonator in the grid circuit instead of the cathode but the same results are achieved.

The third harmonic resonator in the input and output circuits add a third harmonic component to the fundamental RF voltage, this is the first approximation of a square wave. In Figure 15 there are graphs of the various waveforms of Figure 14. The DC plate voltage is 5000 volts, the amplitude of the fundamental RF voltage component is 5300 volts, and the third harmonic component is 1060 volts. When the fundamental and third harmonic waveforms are added, the resultant waveform has a maximum value of 4590 volts at the peak and 4240 volts at the dip. When added to the DC plate voltage of 5000 volts, the minimum plate swing is 410 volts with 760 volts at the center. With a conduction angle of 120^{o} and an I_p maximum value of 9.1 amps, the DC plate current is 2 amps and the equivalent sinewave RF plate current (i_p) has a peak value of 3.6 amps.

The analysis of this amplifier is:

$E_p = 5KV \qquad I_p = 2A \qquad I_p$ maximum $= 9.1A$

$i_p = 3.6A \qquad e_p = 5300V$

$$P_{in} = E_p I_p = 5000V \times 2A = 10KW$$

$$P_{out} = \frac{e_p i_p}{2} = \frac{5300V \times 3.6A}{2} = 9.54KW$$

$$\text{Efficiency} = \frac{P_{out}}{P_{in}} \times 100 = \frac{9.54KW}{10KW} \times 100 = 95.4\%$$

$$P_{diss} = P_{in} - P_{out} = 10KW - 9.54KW = 460W$$

The reason Class D has a high efficiency is that when the plate current is flowing the plate voltage is an approximation of a square wave. It is caused by the addition of a third harmonic component to the fundamental RF plate voltage sine wave. The lower plate dissipation occurs because the plate voltage waveform swings closer to zero volts when plate current flows.

The addition of the third harmonic sine wave to the fundamental plate voltage sine wave also serves to increase output power. Notice in Figure 15 that this composite RF and DC plate waveform never drives the plate voltage to zero volts, but always allows enough positive plate voltage to keep the tube in conduction. If the plate voltage is dropped to zero volts or below, such as if e_p exceeded E_p as shown in Figure 16, the tube would draw no plate current, but in Class D this is exactly what happens. The extreme dip of plate voltage below zero volts is disguised by the addition of the third harmonic component.

At test point A in Figure 14, the third harmonic component is removed by the third harmonic resonator revealing the fundamental component of the plate voltage and plate current.

The third harmonic resonator in the input circuit changes the grid-cathode RF waveform to an approximation of a square wave so that the plate current has a component of third harmonic current of the proper amplitude. The third harmonic resonator in the cathode and plate have high impedances at that frequency.

The following is a table of measured impedances from a 5KW RF Class D RF power amplifier.

Z_{pfo} = 3100 ohms

Z_{p3f0} = 28K ohms

Z_{k3fo} = 2600 ohms

Where: fo = fundamental frequency

3fo = third harmonic

Zk = cathode impedance

Due to the high third harmonic impedances in the plate and cathode circuit, only a small percentage of third harmonic current exists in the plate current waveform. The plate current waveform can be treated as a pure Class C current waveform and the third harmonic component ignored. It will not cause significant errors in the appearance of the waveform or calculations involving it.

The large voltage and small current components of the third harmonic in the plate and cathode circuit can be understood if ohms law is applied to the impedances given above.

e_{pfo} = 5872 volts and Z_{pfo} = 3100 ohms

$$\text{Yielding:}\quad i_{pfo} = \frac{e_{pfo}}{Z_{pfo}} = \frac{5872V}{3100 \text{ ohms}}$$

$$i_{pfo} = 1.9 \text{ amps}$$

e_{p3fo} = 1200V and Z_{p3fo} = 28K ohms

$$i_{p3fo} = \frac{e_{p3fo}}{Z_{p3fo}} = \frac{1200V}{28K \text{ ohms}}$$

$$i_{p3fo} = 42.9 \text{ ma}$$

$$e_{k3fo} = i_{p3fo} \times Z_{k3fo}$$

$$= 42.9 \text{ ma} \times 2600 \text{ ohm}$$

$$e_{k3fo} = 111 \text{ volts}$$

Summary of Tube Type Class D

Class D efficiency is high because plate voltage is low when plate current flows, and this holds plate dissipation to a minimum. Good output is obtained because of the large component of fundamental frequency RF plate voltage. The Harris MW-5, MW-10, and MW-50, as well as the VP-100, use these high efficiency circuitry.

SOLID STATE CLASS D R.F. AMPLFIERS

SOLID STATE DEVICES SUCH AS JUNCTION TRANSISTORS OR THE VARIOUS TYPES OF FET'S HAVE SEVERAL PROPERTIES THAT LEND THEMSELVES FAVORABLY TO HIGH EFFICIENCY CLASS D USE. FOR THIS DISCUSSION THREE PROPERTIES ARE OF PARTICULAR INTEREST. THEY ARE:

(1.) OUTPUT D.C. VOLTAGE HAS MINIMUM EFFECT ON THE OUTPUT D.C. CURRENT.

COLLECTOR VOLTAGE AND CURRENT (V_C AND I_C) WITH JUNCTION TRANSISTOR, DRAIN VOLTAGE AND CURRENT (V_D AND I_D) WITH FIELD EFFECT TRANSISTORS.

(2.) LOW COLLECTOR OR DRAIN RESISTANCE.

(3.) LOW COLLECTOR OR DRAIN SATURATION VOLTAGE.

A GRAPH OF THE COLLECTOR FAMILY OF CURVES FOR A TYPICAL JUNCTION TRANSISTOR IS SHOWN IN FIGURE 17. ABOVE SATURATION A CHANGE OF COLLECTOR VOLTAGE HAS MINIMUM EFFECT ON THE COLLECT CURRENT. THIS ALLOWS A STEADY VALUE OF COLLECTOR CURRENT TO FLOW DOWN TO SATURATION. A GRAPH OF THE DRAIN VOLTAGE AND CURRENT FAMILY OF CURVES FOR THE IRF 350 MOS FET POWER TRANSISTOR, SHOWN IN FIGURE 18, SHOWS THE SAME INDEPENDANCE OF DRAIN CURRENT FROM DRAIN VOLTAGE AS THE JUNCTION TRANSISTOR.

THE LOW OUTPUT RESISTANCE OF THE SOLID STATE DEVICES REQUIRE THAT RELATIVELY LOW IMPEDANCES BE USED FOR THE R.F. CIRCUITS, AND THAT D.C. SUPPLY VOLTAGES OF A FEW HUNDRED VOLTS OR LESS BE USED. COLLECTOR OR DRAIN CURRENT IS USUALLY ON THE ORDER OF A FEW AMPS, RARELY OVER

TWENTY AMPS PER TRANSISTOR. THE LOW RESISTANCE OF THESE DEVICES MAKES OPERATION AT LOW VOLTAGE POSSIBLE.

SATURATION IS ON THE ORDER OF 0.1 VOLT TO 1.0 VOLT FOR A JUNCTION POWER TRANSISTOR AND USUALLY LESS THAN 4 VOLTS FOR A POWER FET. THE FET CURVES IN FIGURE 18 SHOWS 6 AMPS FLOWING WITH A COLLECTOR TO DRAIN VOLTAGE DROP OF 2 VOLTS.
THE LOW SATURATION VOLTAGE AND RESISTANCE MAKE THESE DEVICES ADAPTABLE FOR SWITCHING CIRCUITS.

WHEN A TRANSISTOR OF FET IS AT SATURATION, THE DEVICE IS CAPABLE OF PASSING SEVERAL AMPS OF CURRENT WITH A SMALL VOLTAGE DROP. IN FIGURE 19, THE COLLECTOR TO EMITTER VOLTAGE IS 0.5 VOLT AT 4 AMPS SO THAT THE TRANSISTOR IS DISSIPATING ABOVE 2 WATTS OF POWER. THIS LEAVES 99.5 VOLTS ACROSS THE 25 OHM LOAD FOR A LOAD POWER OF 398 WATTS. WITH 400 WATTS INPUT, THE DEVICE IS 99.5% EFFICIENT AS A CLOSED SWITCH. WHEN THE SWITCH IS OPEN, BECAUSE THE TRANSISTOR IS REVERSE BIASED, THE DEVICE DROPS ALL 100 VOLTS, SEE FIGURE 20, BUT DISSIPATES NO POWER SINCE THE CURRENT IS ZERO.

A SOLID STATE DEVICE IS AN EXCELLENT SWITCH UNDER STEADY STATE CONDITION, BUT IT WILL DISSIPATE POWER DURING SWITCHING TRANSIENTS, WHEN IT HAS BOTH A SIZABLE CURRENT FLOW AND VOLTAGE DROP.

IN FIGURE 21, DURING THE SWITCHING TRANSIENT OF THE TRANSISTOR THE FOLLOWING PARAMETERS OCCUR AT MID POINT OF THE SWITCHING ACTION:

$V_{CE} = 50$ VOLTS $I_C = 1$ AMP

$P_{DISS.} = V_{CE} \times I_C = 50$ VOLTS $\times$ 1 AMP $=$ 50 WATTS

$V_{CC} = 100$ VOLTS $I_C = 1$ AMP

$P_{IN} = V_{CC} \times I_C = 100$ VOLTS $\times$ 1 AMP $=$ 100 WATTS

$V_{RL} = 50$ VOLTS $I_C = 1$ AMP

$P_{RL} = V_{RL} \times I_C = 50$ VOLTS $\times$ 1 AMP $=$ 50 WATTS

DURING THE SWITCHING TIME THE EFFICIENCY OF THE CIRCUIT DROPS TO A MINIMUM OF 50%.

$$\text{EFFICIENCY} = \frac{P_{OUT}}{P_{IN}} \times 100 = \frac{50 \text{ VOLTS}}{100 \text{ VOLTS}} \times 100 = 50\%$$

A COMMON TYPE OF SWITCH IS PRODUCED BY DRIVING A TRANSISTOR, OR FET, WITH A SINE WAVE. IF THE AMPLIFIER IS ONLY SLIGHTLY OVERDRIVEN BY THE SINEWAVE, AS IN FIGURE 21, THE SWITCHING TRANSIENT TIME WILL BE LONG AND THE AVERAGE POWER DISSIPATION WILL BE HIGH.

IF HIGH SWITCHING RATES ARE ENCOUNTERED, SUCH AS WITH CLASS D R.F. AMPLIFIERS, TRANSIENT TIME POWER DISSIPATION WILL BECOME A SUBSTANTIAL PORTION OF THE TOTAL AMPLIFIER POWER DISSIPATION. THIS PROBLEM IS MINIMIZED BY KEEPING THE TRANSIENT TIME AS SHORT AS POSSIBLE. THUS THE HIGH TRANSIENT POWER DISSIPATION OCCURS FOR A SMALL PORTION OF THE SWITCHERS DUTY CYCLE KEEPING THE AVERAGE POWER DISSIPATION LOW. THIS IS ACCOMPLISHED IN FIGURE 22 BY GREATLY OVERDRIVING THE

SWITCHING TRANSISTORS WITH A LARGE AMPLITUDE SINE WAVE.

THE CLASS D R.F. POWER SWITCH

FIGURE 23 SHOWS THE SCHEMATIC AND WAVEFORMS FOR A CLASS D PUSH PULL R.F. POWER AMLIFIER. THE CONDUCTION ANGLE FOR EACH TRANSISTOR IS 180° AND THE BASE OF EACH AMPLIFIER IS OVERDRIVEN TO THE EXTENT THAT THE TRANSISTORS ARE ALTERNATELY DRIVEN FROM CUTOFF TO SATURATION AS EACH ONE CONDUCTS ON ITS HALF CYCLE.

Q_1 CONDUCTS ON THE POSITIVE HALF CYCLE OF THE OUTPUT SIGNAL AND Q_2 CONDUCTS FOR THE NEGATIVE HALF CYCLE. WHEN Q_1 CONDUCTS, THE VOLTAGE AT THE JUNCTION OF THE EMITTER OF Q_1 AND THE COLLECTOR OF Q_2 IS VERY CLOSE TO V_{CC}, BEING LOWER BY THE 0.5 VOLT TO 1.0 VOLT TRANSISTOR SATURATION VOLTAGE. THIS CONDUCTION CHARGES C_C TO ONE-HALF THE SUPPLY VOLTAGE (V_{CC}). WHEN Q_2 CONDUCTS, C_C DISCHARGE RETURNING SOME OF ITS STORED ENERGY TO THE CIRCUIT. THE COLLECTOR OF Q_2 HAS A SATURATION VOLTAGE OF 0.5 VOLT TO 1.0 VOLT WHILE THE REMAINDER OF THE C_C VOLTAGE APPEARS AT THE INPUT OF THE RESONANT CIRCUIT - THE JUNCTION OF L_1 AND C_C. CAPACITOR C_C HAS A LARGE ENOUGH VALUE SO THAT ITS D.C. CHARGE ($\frac{1}{2} V_{CC}$) REMAINS CONSTANT OVER THE PERIOD OF THE R.F. CYCLE.

THE OUTPUT VOLTAGE AT THE JUNCTION OF Q_1, Q_2, AND C_C IS A SQUARE WAVE WITH ITS:

$$V_{MAX} = V_{CC} - V_{Q1\ SAT.}$$

AND $V_{MIN} = V_{C2\ SAT.}$

On the other side C_C at the junction C_C and L_1 the voltage is a square wave with values of:

$$V_{MAX\ POS} = \frac{1}{2} V_{CC} - V_{Q1\ SAT.}$$

$$V_{MAX\ NEG} = V_{Q2\ SAT} - \frac{1}{2} V_{CC}$$

SEE WAVEFORM A OF FIGURE 23

RESONANT CIRCUIT $L_1 - C_1$ MATCHES THE HIGHER LOAD IMPEDANCE TO THE LOWER AMPLIFIER OUTPUT IMPEDANCE FOUND AT THE D.C. BLOCKING COUPLING CAPACITOR C_C.

THE INPUT CURRENT TO THE RESONANT CIRCUIT THROUGH C_C IN FIGURE 23, IS A SINE WAVE OF PEAK AMPLITUDE i_p, SEE WAVEFORM E OF FIGURE 23.

THE VOLTAGE WAVEFORM TO THE INPUT OF THE RESONANT CIRCUIT IS A SQUARE WAVE, CAUSED BY THE TRANSISTOR'S SATURATION, BUT THE SINE WAVE OF CURRENT IS DUE TO THE IMPEDANCE OF THE RESONANT CIRCUIT, THE VOLTAGE DROP ACROSS IT, AND THE FACT THAT AT SATURATION THE COLLECTOR CURRENT CAN BE QUITE HIGH BUT IS USUALLY LOWER DEPENDING UPON THIS FORMULA

$$I_C = \frac{V_{LOAD}}{Z_{LOAD}}$$

WHERE: Z_{LOAD} = THE AMPLIFIER'S OUTPUT LOAD IMPEDANCE AND V_{LOAD} = ITS VOLTAGE DROP

SINCE THE SIGNAL CURRENT THROUGH C_C IS A SINE WAVE OF PEAK VALUE i_p, EACH TRANSISTOR'S COLLECTOR CURRENT IS A HALF CYCLE OF CURRENT OF A VALUE OF I_C MAXIMUM $= i_p$, SEE FIGURE 23 WAVEFORM C AND D.

THE D.C. CURRENT DRAWN BY THE COLLECTOR OF Q_1 IS THE AVERAGE OF THE HALF CYCLE SINEWAVE PULSES AND IS:

$$I_C = \frac{i_p}{\pi}$$

THE INPUT POWER TO THE AMPLIFIER IS:

$$P_{IN} = V_{CC} \times I_C$$

THE SQUARE WAVE VOLTAGE AT THE RESONANT CIRCUIT INPUT, JUNCTION OF C_C AN L_1 HAS A VALUE OF:

$$V_{max} = \frac{V_{CC} - 2\,V_{SAT}}{2}$$

WHERE: V_{max} = PEAK VALUE OF THE SQUARE WAVE

V_{CC} = THE D.C. SUPPLY VOLTAGE

V_{SAT} = V_{CE} AT SATURATION FOR EACH OUTPUT TRANSISTOR, AND IT IS ASSUMED THAT $V_{Q1\,SAT} = V_{Q2\,SAT} = V_{SAT}$.

THE RESONANT CIRCUIT WILL ALLOW ONLY THE FUNDAMENTAL COMPONENT OF THE SQUARE WAVE VOLTAGE TO PASS, OFFERING AN EXTREMELY HIGH IMPEDANCE TO ALL OF ITS HARMONICS. THE PEAK VALUE OF THE FUNDAMENTAL SINEWAVE COMPONENT OF THE SQUARE WAVE VOLTAGE (e_p) IS LARGER THAN THE SQUARE WAVE VOLTAGE (V_{MAX}).

$$e_p = \frac{4}{\pi} \cdot V_{max}$$

THE R.F. OUTPUT POWER OF THE AMPLIFIER BECOMES:

$$P_{OUT} = \frac{e_p\, i_p}{2}$$

In the following example the output power, efficiency, power dissipation, etc. are calculated for the amplifier and waveforms given in Figure 23.

The high efficiency that results is due to:

(1.) The low V_{SAT} voltage

(2.) The extremely fast switching time of the transistors

(3.) The fact that the fundamental component of the amplifier's output voltage (e_p) is larger than the square wave voltage (V_{MAX}).

$V_{CC} = 25$ volts

$I_C = 4$ amps

$P_{IN} = E_p I_p = 100$ watts

$V_{MAX} = \dfrac{V_{CC} - (2V_{SAT})}{2} = 11.75$ volts where $V_{SAT} = V_{CE}$ at saturation

$e_p = \dfrac{4}{\pi} V_{max} = 14.96$ volts

$i_p = \pi I_c = 12.56$ amps

$$P_{OUT} = \frac{e_p i_p}{2} = \frac{14.96 \text{ Volts} \times 12.56 \text{ Amps}}{2} = 93.9 \text{ watts}$$

$$EFF = \frac{P_{OUT}}{P_{IN}} \times 100 = \frac{93.9 \text{ watts}}{100 \text{ watts}} \times 100 = 93.9\ \%$$

$P_{DISS} = P_{IN} - P_{OUT} = 100$ watts $- 93.9$ watts $= 6.1$ watts

$$Z_p = \frac{e_p}{i_p} = \frac{14.96 \text{ volts}}{12.56 \text{ amps}} = 1.19 \text{ ohms}$$

$$R_{DC} = \frac{V_{CC}}{I_c} = \frac{25 \text{ volts}}{4 \text{ amps}} = 6.25 \text{ ohms}$$

IT CAN BE PROVED THAT THIS TYPE OF CLASS D CIRCUIT HAS AN IDEAL EFFICIENCY OF 100%, IF VSAT EQUALS ZERO VOLTS BY THE FOLLOWING MATHEMATICAL SEQUENCE.

(1.) $P_{OUT} = \frac{e_p i_p}{2}$

(2.) $i_p = \pi \cdot I_C$

(3.) $e_p = \frac{2 V_{CC}}{\pi}$ (From $e_p = \frac{4}{\pi} V_{max}$ and $V_{max} = \frac{V_{CC} - 2 V_{SAT}}{2}$)

(4.) $P_{OUT} = \frac{2 V_{CC}}{\pi} \cdot \frac{\pi I_C}{2}$ (SUBSTITUTING IN 1 FROM 2 AND 3.)

(5.) $P_{OUT} = V_{CC} \cdot I_C$ (SIMPLY 4)

(6.) $P_{IN} = V_{CC} I_C$ (GIVEN)

(7.) $P_{OUT} = P_{IN}$ (SUBSTITUTING IN 5 FROM 6)

SOLID STATE PUSH PULL CLASS D AMPLIFIERS HAVE PRACTICAL EFFICIENCIES OF OVER 90%, BEING LIMITED ONLY BY THE FOLLOWING

(1.) TRANSIENT TIME NOT BEING INSTANTANEOUS.

(2.) THE SATURATION VOLTAGE OF THE ACTIVE DEVICES.

(3.) THE IR VOLTAGE DROP OF THE CONDUCTORS IN THE CIRCUIT.

(4.) THE VARIOUS CAPACITOR AND INDUCTOR LOSSES ASSOCIATED WITH THE OUTPUT NETWORK.

THE MOS FET AMPLIFIER SHOWN IN FIGURE 24 USES TWO COMPLEMENTARY PUSH PULL AMPLIFIERS. Q11 AND Q12 FORM ONE COMPLEMENTARY PAIR AND Q14 AND Q13 FORM THE OTHER PAIR. EACH INDIVIDUAL COMPLEMENTARY PUSH PULL AMPLIFIER PAIR WORKS LIKE THE ONE JUST DESCRIBED, PRODUCING A SQUARE WAVE OUTPUT, BUT THE TWO AMPLIFIER SQUARE WAVES ARE 180° OUT OF PHASE.

Q_{11} AND Q_{13} CONDUCT ON ONE HALF CYCLE OF THE INPUT AND Q_{14} AND Q_{12} CONDUCT ON THE OTHER HALF. ASSUMING, AS IN THE OTHER CIRCUIT JUST DESCRIBED, THAT EACH AMPLIFIER HAS A $V_{CC} = 25$ VOLTS AND $I_C = 4$ AMPS, THEN EACH AMPLIFIER HAS 100 WATTS INPUT FOR A COMBINED INPUT OF 200 WATTS.

ALLOWING FOR THE SAME V_{SAT} AS THE PREVIOUS EXAMPLE ($V_{DC} = 0.25$ VOLTS), THE SQUARE WAVE VOLTAGE ACROSS THE SERIES RESONANT OUTPUT NETWORK OF C_1, L_1, AND T_1 WOULD HAVE A V_{MAX} OF 23.5 VOLTS. THIS YIELDS A FUNDAMENTAL SINE WAVE VOLTAGE COMPONENT OF $e_p = 29.4$ VOLTS WITH THE SAME R.F. CURRENT OF $i_p = 12.56$ AMPS.

THE SERIES RESONANT CIRCUIT ALLOWS ONLY THE FUNDAMENTAL SINE WAVE COMPONENT OF THE OUTPUT SQUARE WAVE VOLTAGE TO PASS, OFFERING A HIGH IMPEDANCE TO ALL OTHER HARMONICS. THE OUTPUT CURRENT OF THIS AMPLIFIER LIKE THE PREVIOUS ONE, IS A SINE WAVE.

THE FOLLOWING TABLE OF CALCULATIONS SHOW THE AMPLIFIER CHARACTERISTICS BASED ON $V_{CC} = 25$ VOLTS AND I_C OF EACH AMPLIFIER $= 4$ AMPS.

$$P_{IN} = 2\,V_{CC}\,I_C = 2 \times 25 \text{ VOLTS} \times 4 \text{ AMPS} = 200 \text{ WATTS}$$

$$V_{MAX} = V_{CC} - 2\,V_{SAT} = 25 \text{ VOLTS} - 1.5 \text{ VOLTS} = 23.5 \text{ VOLTS}$$

$$e_p = \frac{4}{\pi} \cdot V_{MAX} = \frac{4}{\pi} \times 23.5 \text{ VOLTS} = 29.9 \text{ VOLTS}$$

$$i_p = \pi\, I_C = \pi \times 4 \text{ AMPS} = 12.56 \text{ AMPS}$$

$$P_{OUT} = \frac{e_p\, i_p}{2} = \frac{29.4 \text{ VOLTS} \times 12.56 \text{ AMPS}}{2} = 188 \text{ WATTS}$$

$$EFF = \frac{P_{OUT}}{P_{IN}} \times 100 = \frac{188 \text{ WATTS}}{200 \text{ WATTS}} = 94\%$$

$$Z_p = \frac{e_p}{i_p} = \frac{29.9 \text{ VOLTS}}{12.56 \text{ AMPS}} = 2.38 \text{ OHMS}$$

$$R_{DC} = \frac{V_{CC}}{I_C} = \frac{25 \text{ VOLTS}}{8 \text{ AMPS}} = 3.13 \text{ OHMS}$$

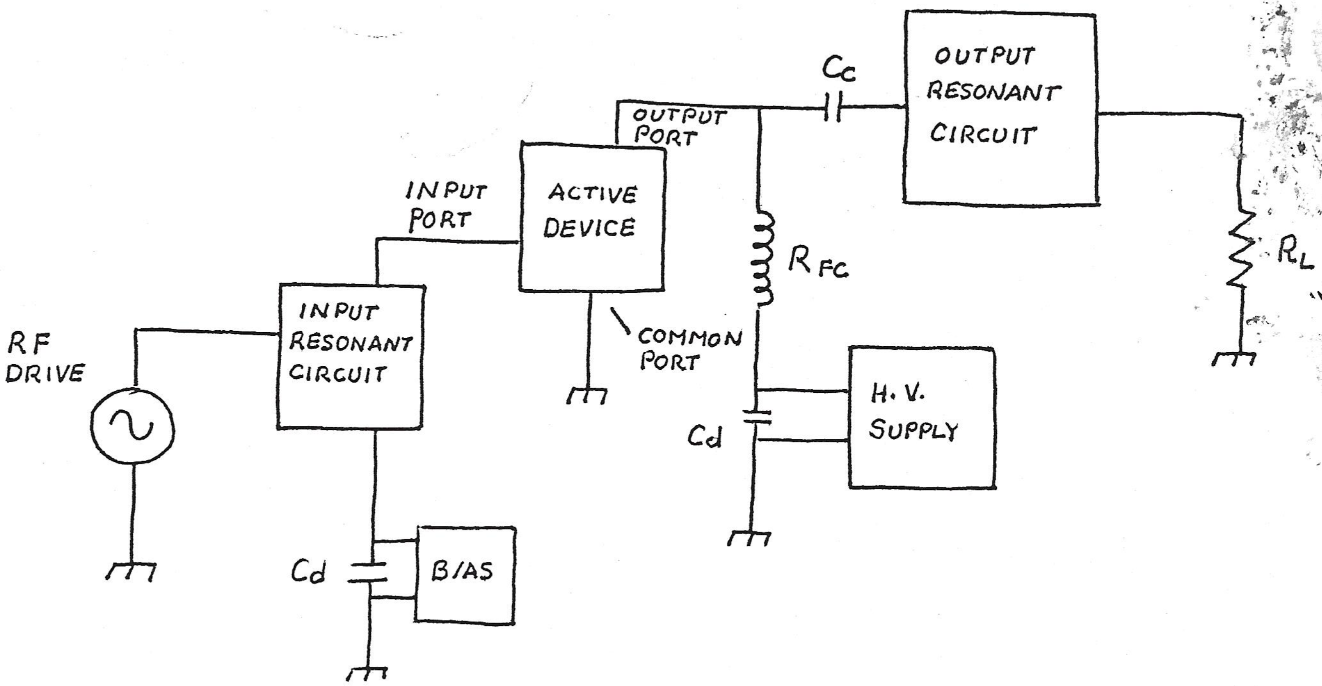


FIGURE 1 BLOCK DIAGRAM OF AN RF POWER AMPLIFIER

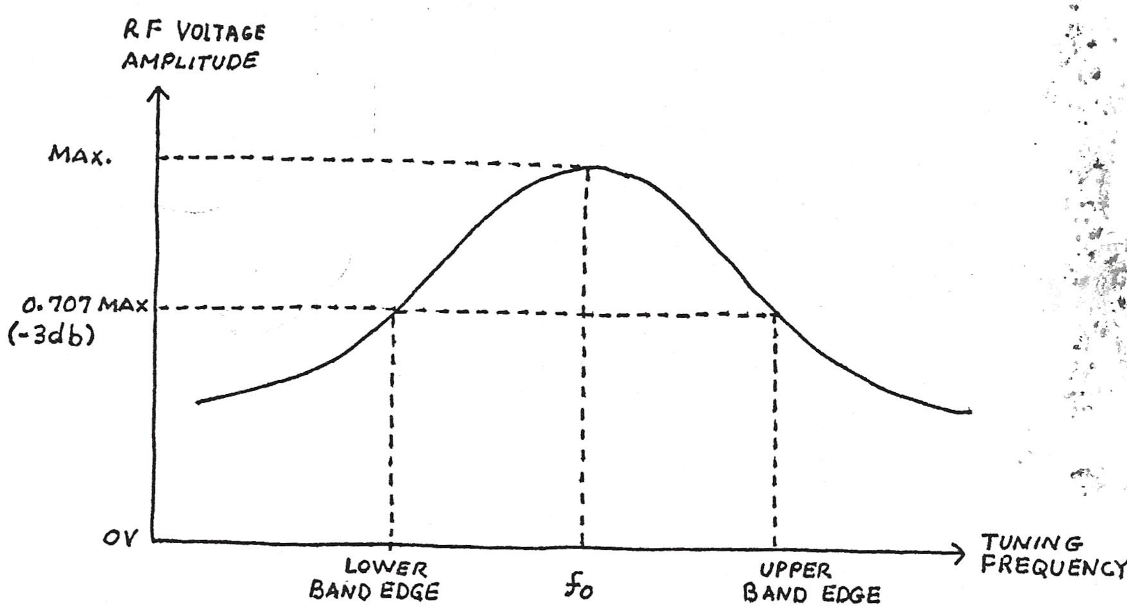


FIGURE 2 GRAPH SHOWING THE OUTPUT R.F. VOLTAGE AS THE OUTPUT PORT TUNING CONTROL, AND THUS THE RESONANT FREQUENCY, IS VARIED ABOVE AND BELOW THE FREQUENCY OF THE CURRENT PULSES DELIVERED BY THE ACTIVE DEVICE.

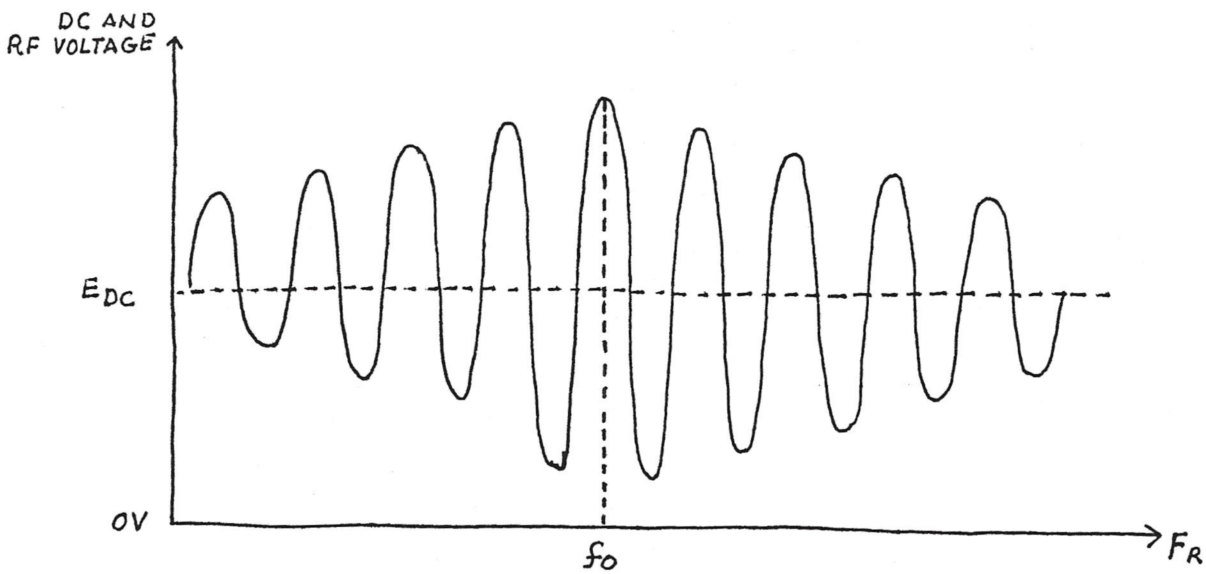


FIGURE 3 RF OUTPUT VOLTAGE IS RIDING ON THE D.C. SUPPLY VOLTAGE. NOTICE THE D.C. VOLTAGE IS CONSTANT AND THE R.F. VOLTAGE VARIES ABOUT IT. THE R.F. VOLTAGE AMPLITUDE DETERMINES THE INSTANTANEOUS VOLTAGE SWING FROM THE D.C. LEVEL.

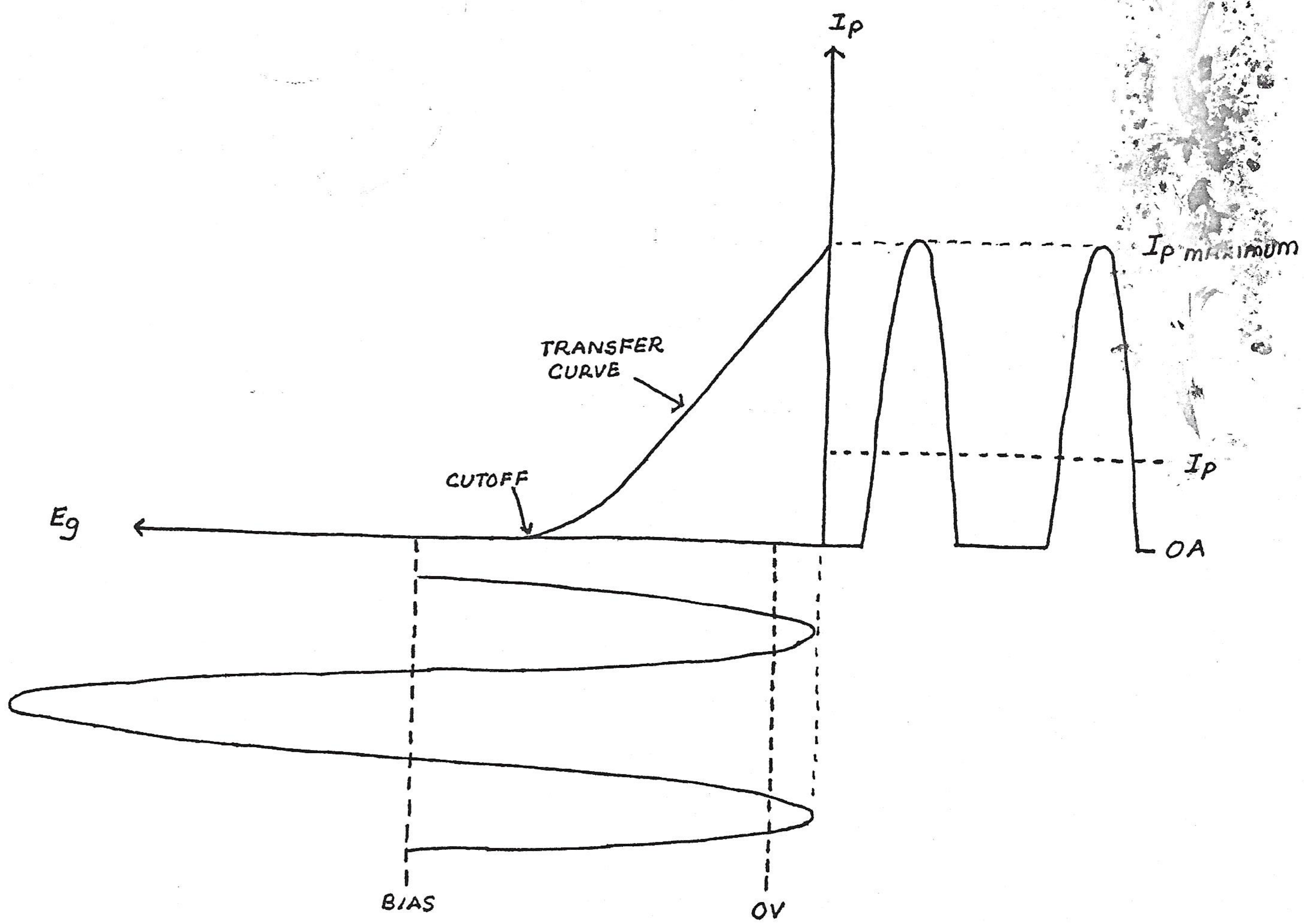


FIGURE 4 TRANSFER CURVE SHOWING BIAS, INPUT DRIVE, AND THE PLATE CURRENT FOR A CLASS C RF AMPLIFIER

THE CONDUCTION ANGLE IS LESS THAN 180°. THE D.C. PLATE CURRENT IS THE AVERAGE OF THE R.F. CURRENT PULSE.

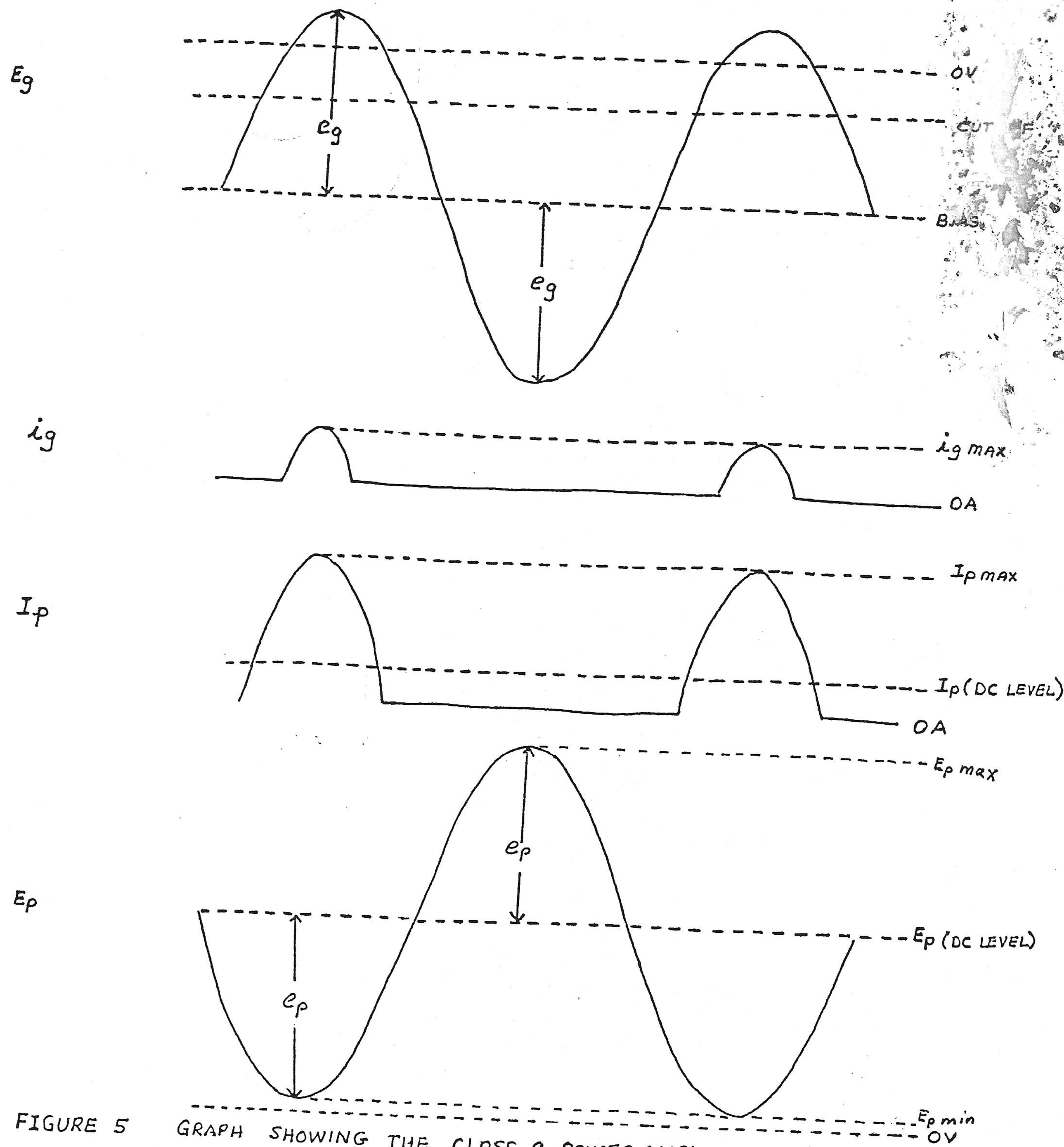


FIGURE 5 GRAPH SHOWING THE CLASS C POWER AMPLIFIER

E_g = DC BIAS, AND e_g = PEAK RF GRID DRIVE VOLTAGE; I_g max = THE INSTANTANEOUS POSITIVE GRID CURRENT, WHICH FLOWS WHEN THE GRID VOLTAGE IS POSITIVE; THE PLATE CURRENT, I_p MAXIMUM AND THE D.C. LEVEL(I_p); AND THE PLATE VOLTAGE, E_p = DC LEVEL AND e_p = THE PEAK RF LEVEL.

NOTE THAT THE PLATE VOLTAGE SWINGS CLOSE TO ZERO VOLTS WHEN PLATE CURRENT FLOWS.

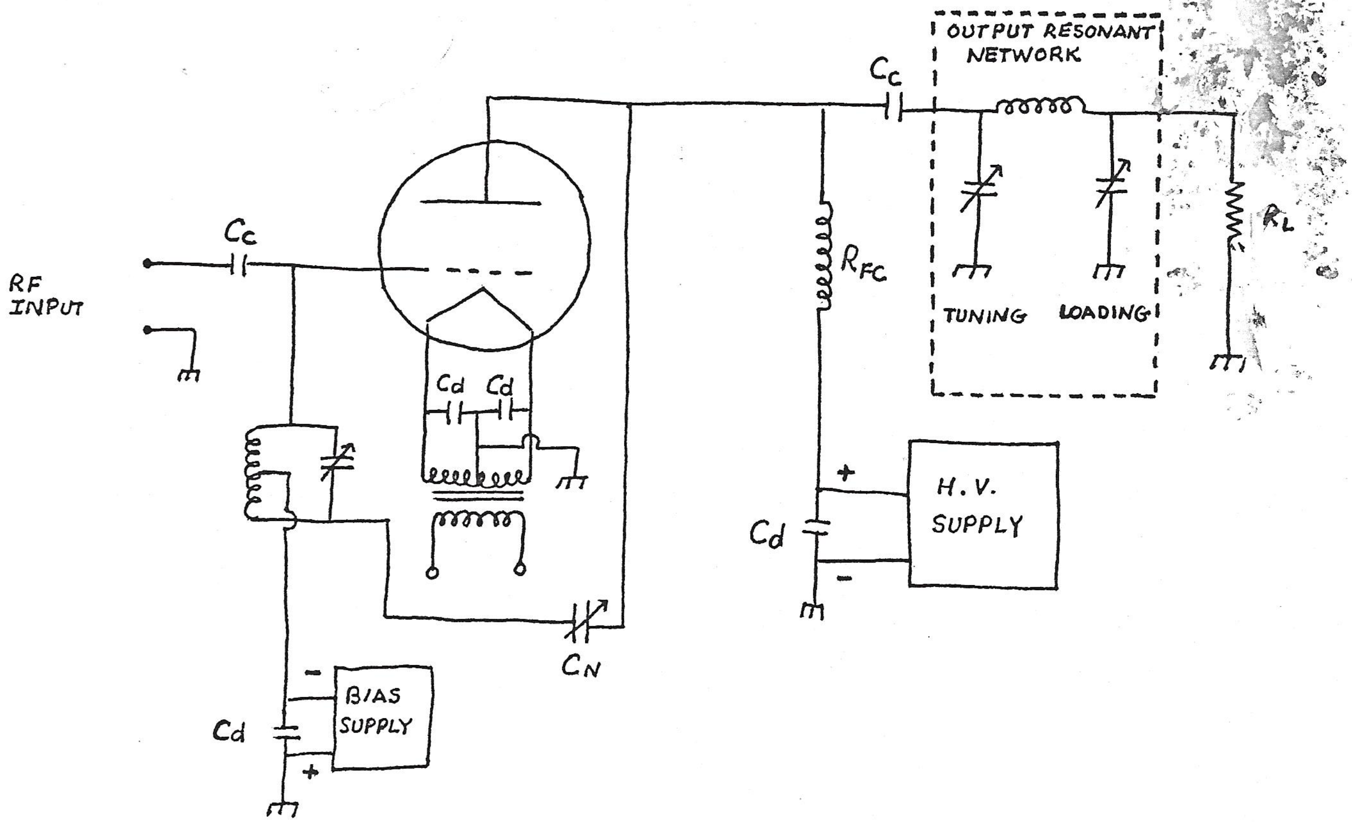


FIGURE 6 A TRIODE R.F. AMPLIFIER WITH NEUTRALIZATION. THIS IS A SHUNT FED OUTPUT SINCE THE D.C. AND R.F. OUTPUT CURRENTS HAVE SEPARATE PATHS OUTSIDE OF THE PLATE OF THE TUBE.

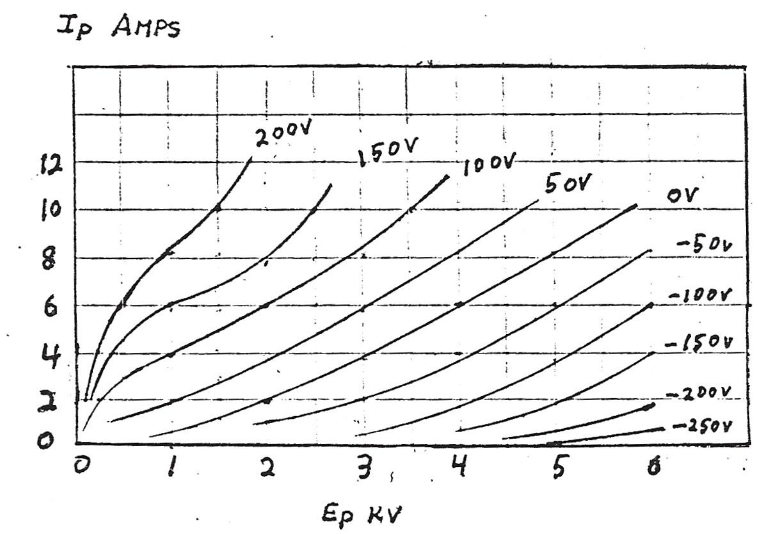


FIGURE 7 PLATE FAMILY OF CURVES FOR A TRIODE.
IF Eg IS HELD CONSTANT Ep AFFECTS Ip

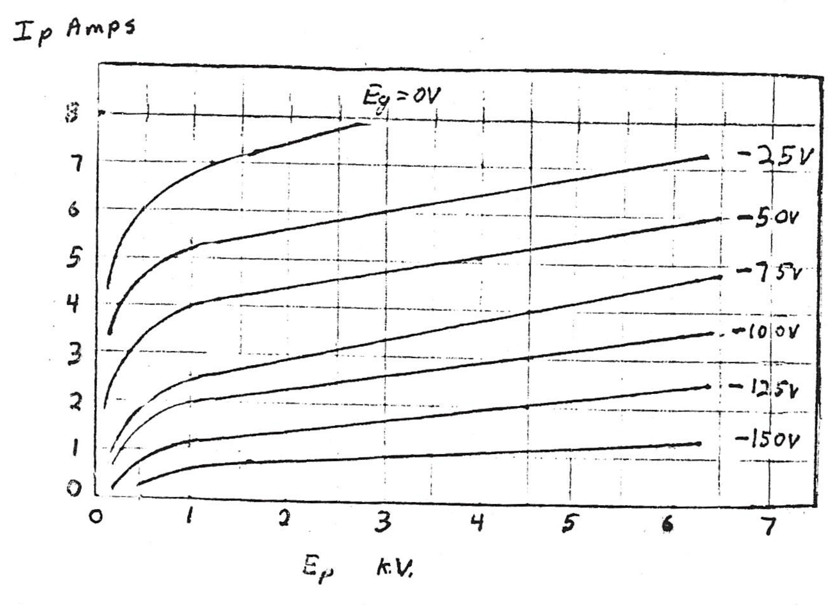


FIGURE 8 PLATE FAMILY OF CURVES FOR A TETRODE. IF Eg IS HELD CONSTANT, Ep HAS MINIMUM AFFECT ON Ip

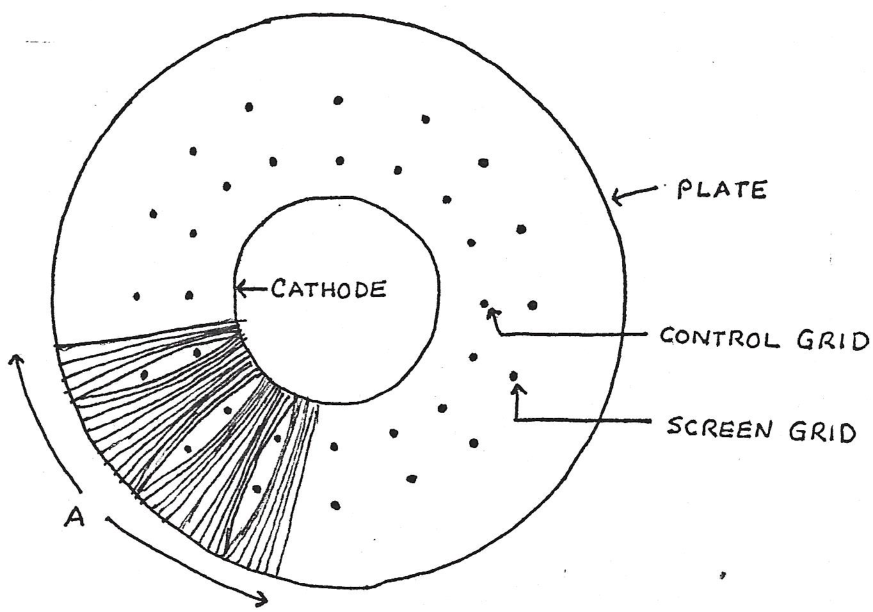


FIGURE 9 TOP VIEW OF A POWER GRID TETRODE SHOWING THE SHADOW WOUND GRIDS. SINCE THE SCREEN GRID IS IN LINE WITH THE CONTROL GRID, MOST OF THE ELECTRONS THAT PASS THE CONTROL GRID WILL MISS THE SCREEN GRID AND TRAVEL TO THE PLATE.

SECTION A SHOWS THE SHADOW EFFECT THAT THE NEGATIVE CONTROL GRID HAS ON THE POSITIVE SCREEN GRID WHEN PLATE CURRENT FLOWS. IN THIS DRAWING, THE INSTANTANEOUS PLATE VOLTAGE IS MUCH GREATER THAN THE VALUE OF E_{sg}.

ONLY A PORTION OF THE TUBE IS SHADED SO THAT THE CONSTRUCTION DETAILS OF THE TUBE CAN BE MORE CLEARLY OBSERVED.

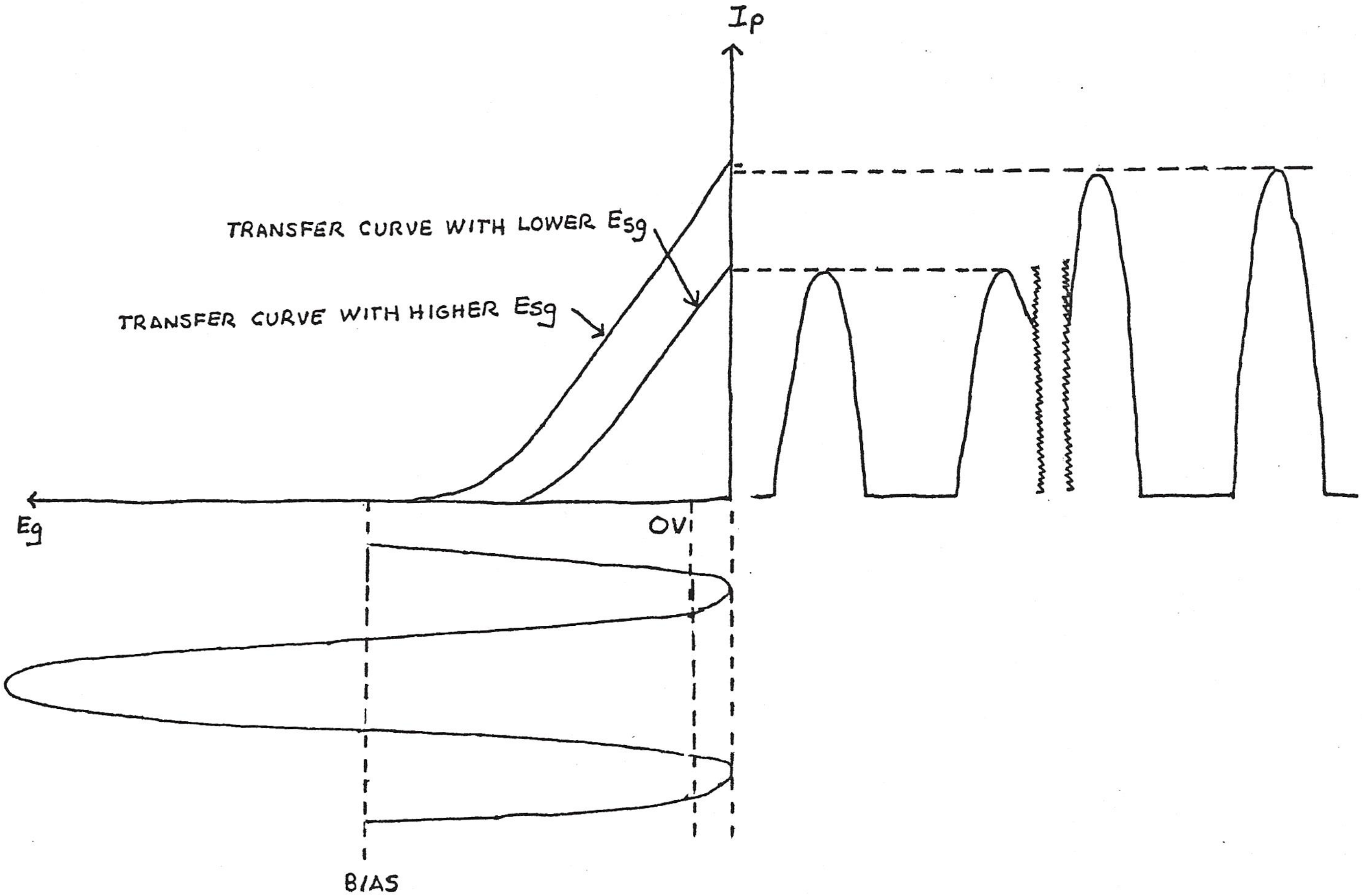


FIGURE 10 . HORIZONTAL SHIFT OF THE TRANSFER CURVES DUE TO A CHANGE OF SCREEN GRID VOLTAGE. THE BIAS AND R.F. DRIVE ARE KEPT CONSTANT BUT THE OUTPUT CURRENT PULSE AMPLITUDE CHANGES DUE TO THE SCREEN GRID VOLTAGE VARIATION. BY INCREASING THE SCREEN VOLTAGE THE SPACE CHARGE AROUND THE CATHODE IS REDUCED AND THE AMPLIFICATION FACTOR (GAIN) OF THE TETRODE IS INCREASED.

IT IS ASSUMED THAT THE CATHODE TEMPERATURE IS HIGH ENOUGH TO CREATE A SPACE CHARGE.

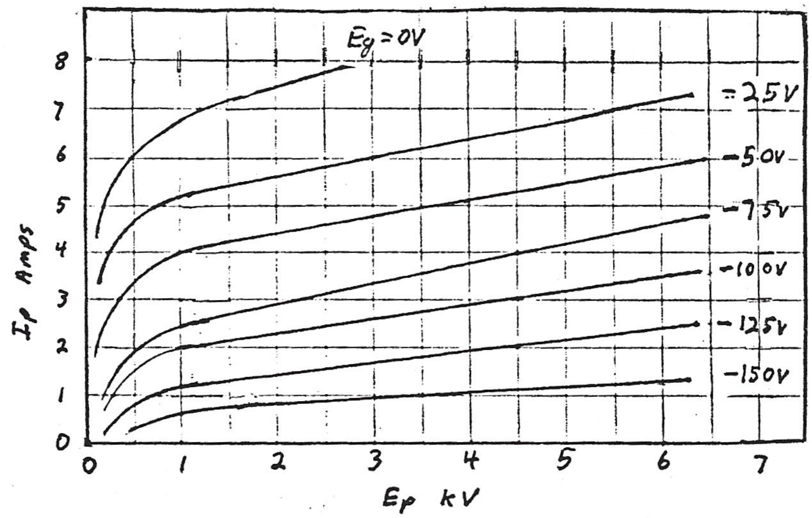


FIGURE 11 A GRAPH OF THE PLATE FAMILY OF CURVES FOR THE 4CX3000A TETRODE WITH Esg = 850V

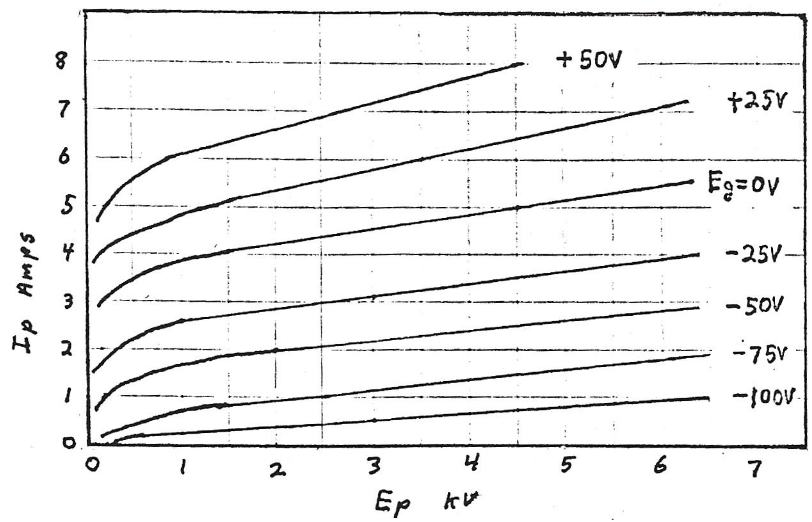


FIGURE 11 B GRAPH OF THE PLATE FAMILY OF CURVES FOR THE 4CX3000A TETRODE WITH Esg = 500V

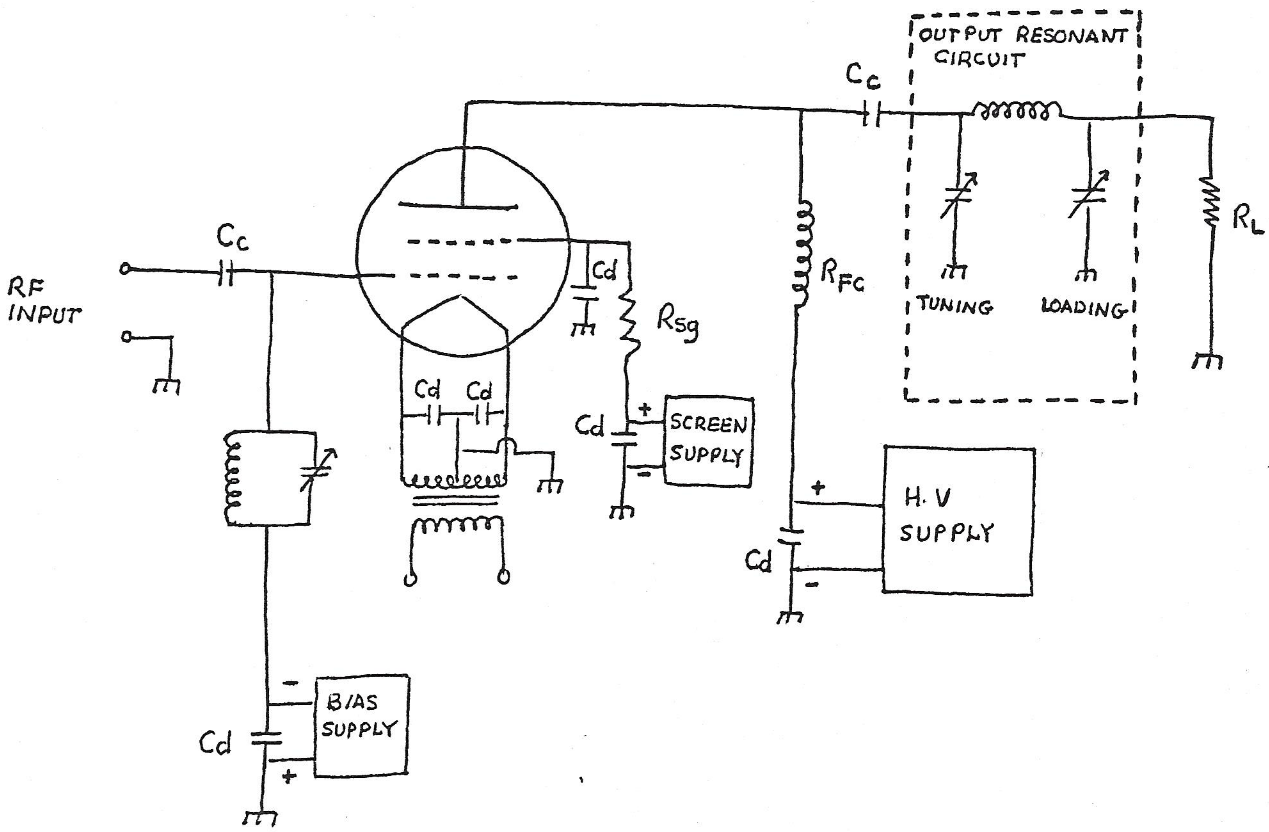
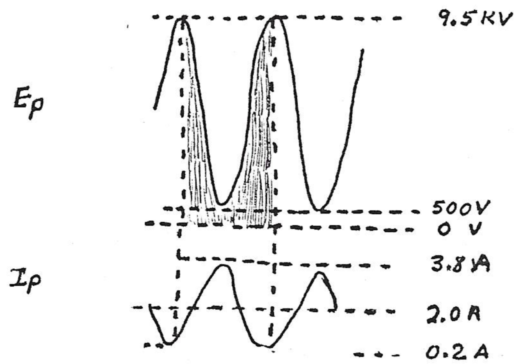


FIGURE 12 TETRODE RF AMPLIFIER

THE RESISTOR R_{sg} IN SERIES WITH THE SCREEN CAUSES SCREEN GRID VOLTAGE TO VARY AS THE SCREEN GRID CURRENT CHANGES. R_{sg} CAUSES THE SCREEN TO HAVE A POORLY REGULATED SCREEN GRID VOLTAGE.

IF R_{sg} IS **SHORTED** THE SCREEN GRID VOLTAGE WILL BECOME WELL REGULATED CREATING A DANGER OF SCREEN GRID OVERDISSIPATION.



CLASS A

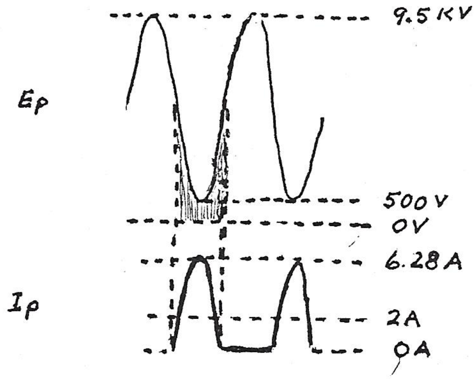
$\angle$ CONDUCTION $= 360°$

$P_{IN} = E_p \cdot I_p = 5000 \times 2 = 10\ KW$

$P_{OUT} = \frac{e_p i_p}{2} = \frac{4500 \times 1.8}{2} = 4.05\ KW$

$EFF = \frac{P_{OUT}}{P_{IN}} \times 100 = \frac{4.05}{10} \times 100 = 40.5\%$

$P_{DISS} = P_{IN} - P_{OUT} = 10KW - 4.05KW = 5.95\ KW$



CLASS B

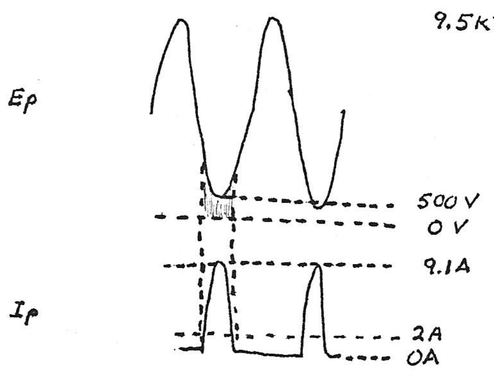
$\angle$ CONDUCTION $= 180°$

$P_{IN} = E_p \cdot I_p = 5000 \times 2 = 10\ KW$

$P_{OUT} = \frac{e_p i_p}{4} = \frac{4500 \times 6.28}{4} = 7.06\ KW$

$EFF = \frac{P_{OUT}}{P_{IN}} \times 100 = \frac{7.06}{10} \times 100 = 70.6\%$

$P_{DISS} = P_{IN} - P_{OUT} = 10KW - 7.06KW = 2.94\ KW$



CLASS C

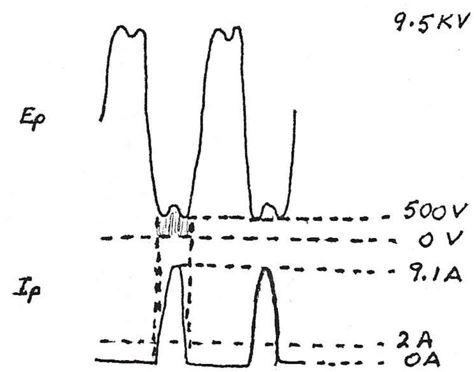
$\angle$ CONDUCTION $= 120°$

$P_{IN} = E_p I_p = 5000 \times 2 = 10\ KW$

$P_{OUT} = \frac{e_p i_p}{2} = \frac{4500 \times 3.6}{2} = 8.1\ KW$

$EFF = \frac{P_{OUT}}{P_{IN}} \times 100 = \frac{8.1}{10} \times 100 = 81\%$

$DISS = P_{IN} - P_{OUT} = 10KW - 8.1KW = 1.9\ KW$



CLASS D

$\angle$ CONDUCTION $= 120°$

$P_{IN} = E_p I_p = 5000 \times 2 = 10\ KW$

$P_{OUT} = \frac{e_p i_p}{2} = \frac{5200 \times 3.6}{2} = 9.360\ KW$

$EFF = \frac{P_{OUT}}{P_{IN}} = \frac{9.36}{10} \times 100 = 93.6\%$

$PISS = P_{IN} - P_{OUT} = 10KW - 9.36KW = 0.64\ KW$

FIGURE 13 GRAPH OF PLATE VOLTAGE AND CURRENT FOR CLASS A THROUGH D. SHADED PORTION SHOWS DISSIPATION

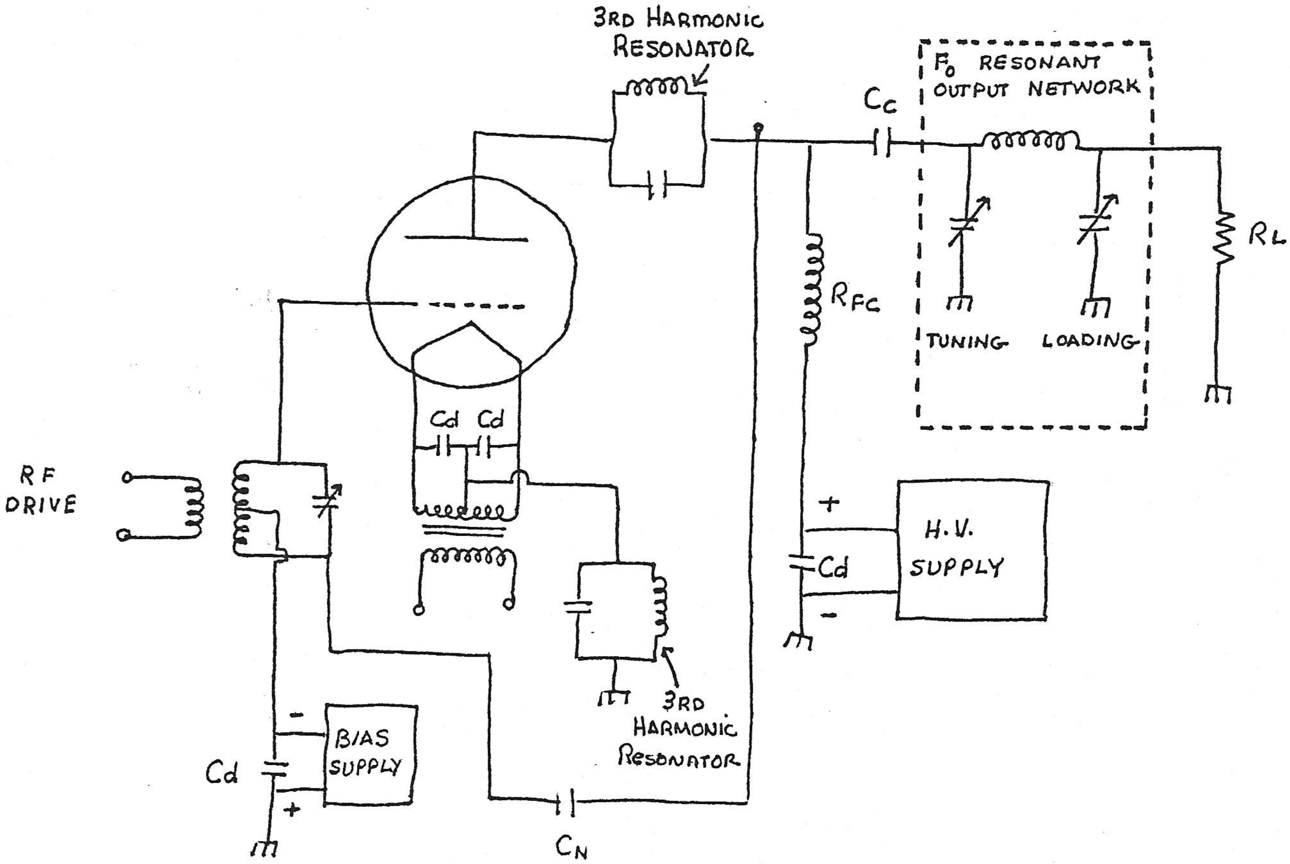
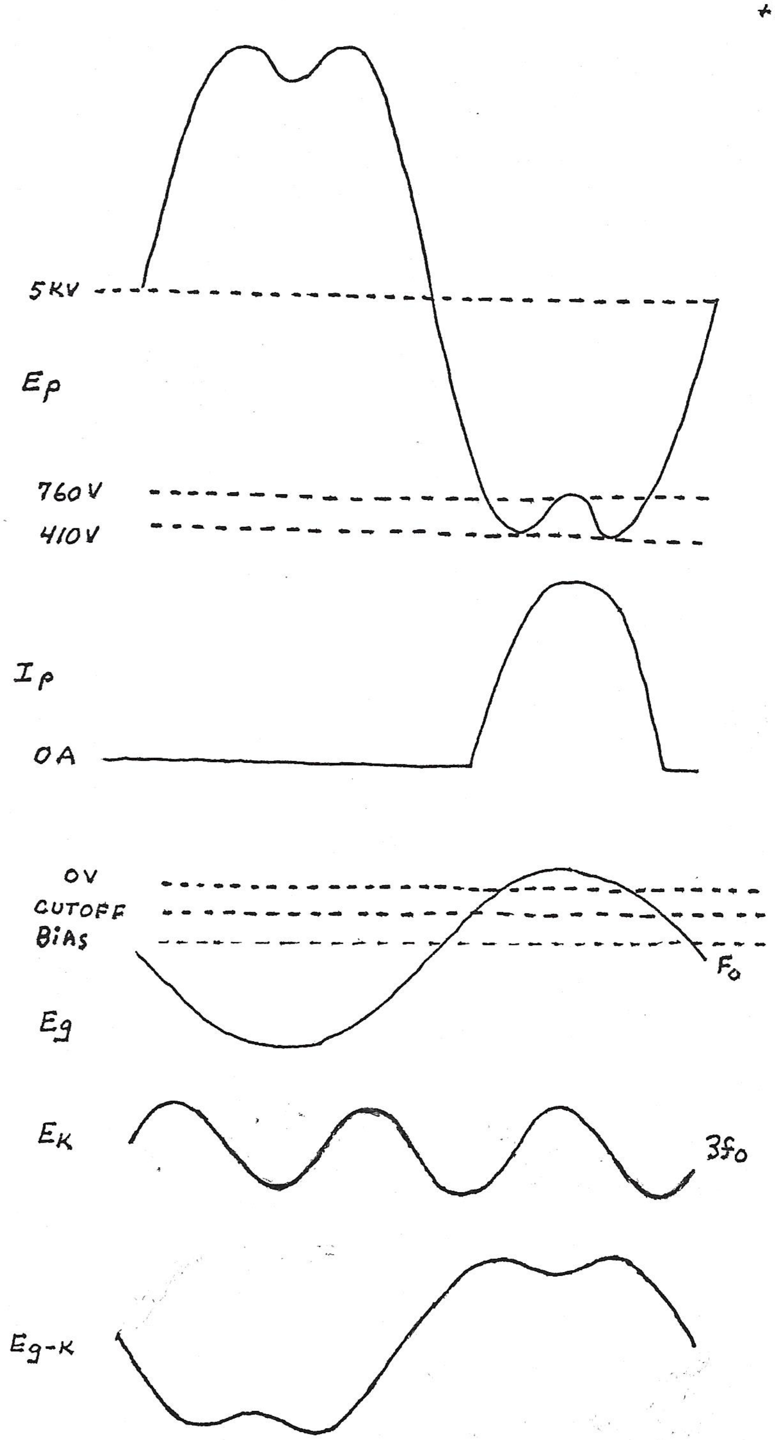


FIGURE 14 A CLASS D HIGH EFFICIENCY TRIODE RF POWER AMPLIFIER. THIRD HARMONIC RESONATORS IN THE PLATE AND CATHODE CIRCUIT CONVERT THE CLASS C AMPLIFIER INTO CLASS D.



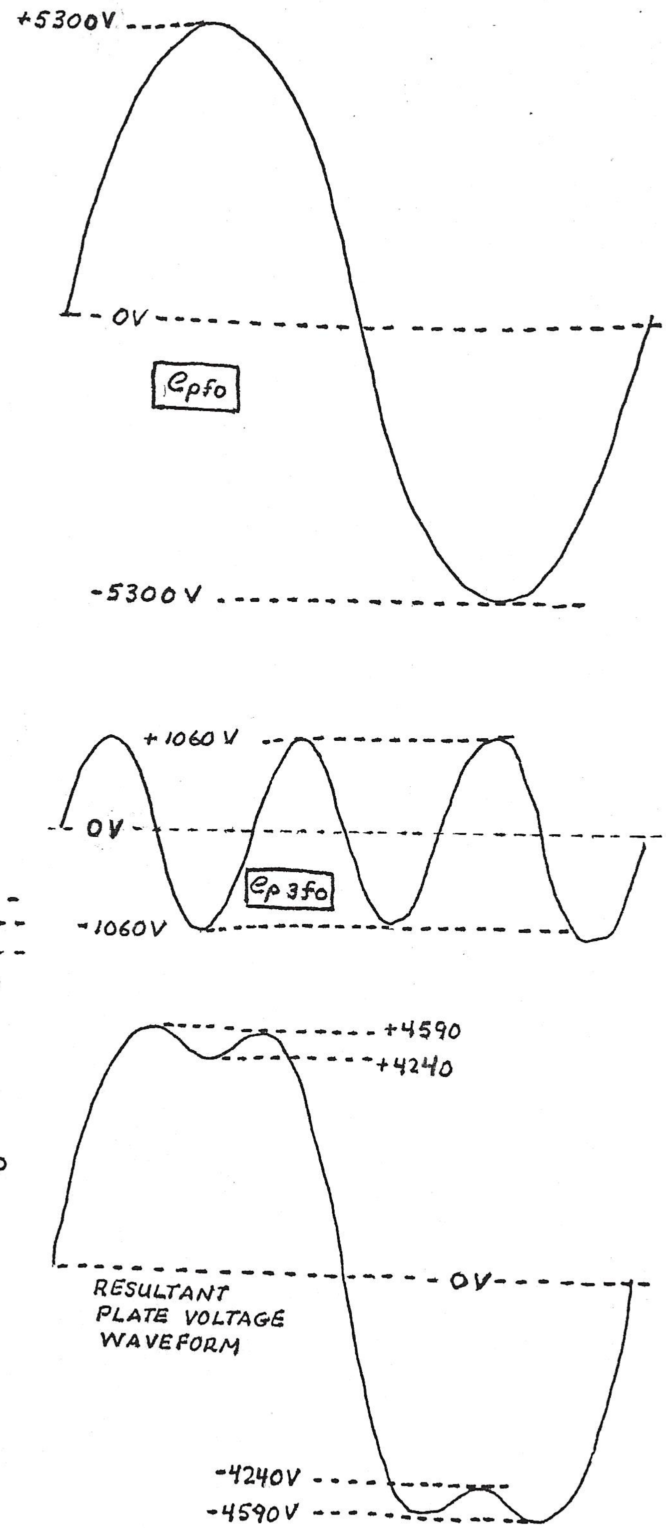


FIGURE 15

CLASS D AMPLIFIER WAVEFORM
SHOWN ABOVE

SHOWN ABOVE, ADDING THE FUNDAMENTAL AND THE THIRD HARMONIC TO CREATE AN APPROXIMATION OF A SQUARE WAVE

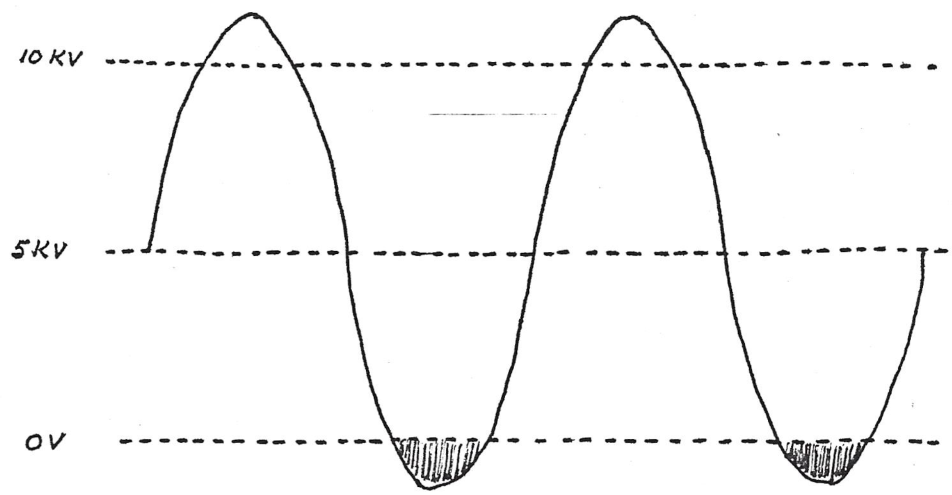


FIGURE 16 IF e_p IS GREATER THAN E_p, THE PLATE VOLTAGE WILL DROP BELOW ZERO VOLTS CAUSING PLATE CURRENT TO STOP AT THAT TIME.

THE SHADED AREA SHOWS WHERE PLATE CURRENT CAN NOT FLOW.

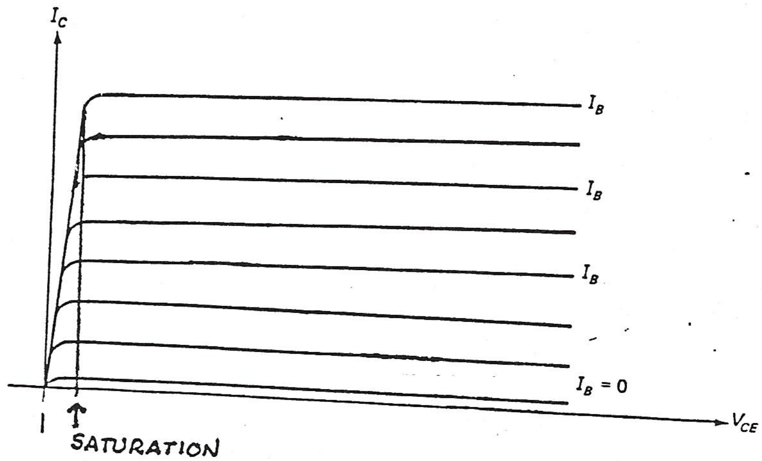


FIGURE 17: JUNCTION TRANSISTOR COLLECTOR FAMILY OF CURVES.

ABOVE SATURATION, WHICH IS LESS THAN 2 VOLTS COLLECTOR TO EMITTER, V_{CE} HAS MINIMUM AFFECT ON I_C.

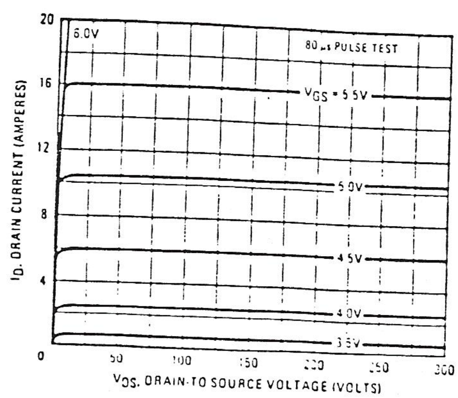


Fig. 3 – Typical Output Characteristics

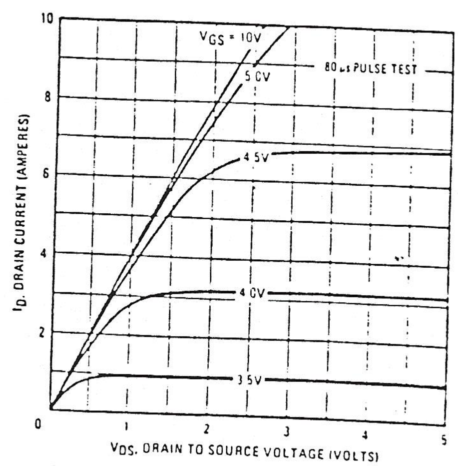
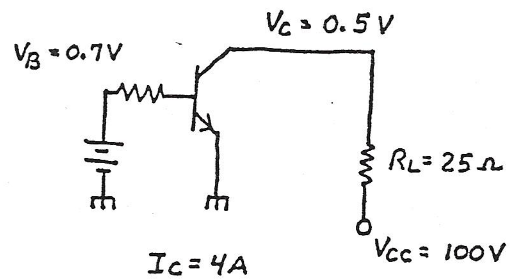


Fig. 5 – Typical Saturation Characteristics (IRF350, IRF351)

FIGURE 18: FET FAMILY OF CURVES. ABOVE SATURATION, WHICH IS LESS THAN $V_{DS} = 5V$, V_{DS} HAS MINIMUM AFFECT ON I_D

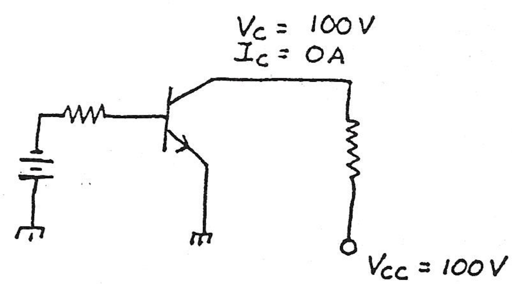


$P_{IN} = V_{CC} \cdot I_C$
$= 100 \times 4$
$P_{IN} = 400W$

$P_{DISS} = V_C \cdot I_C$
$= 0.5 \times 4$
$P_{DISS} = 2W$

$P_{RL} = V_{RL} \cdot I_C$
$= 99.5 \times 4$
$P_{RL} = 398\ W$

FIGURE 19: AT SATURATION MOST OF THE TRANSISTORS INPUT POWER IS DELIVERED TO THE LOAD, VERY LITTLE IS DISSIPATED IN THE COLLECTOR.



$P_{IN} = 0\ W$

FIGURE 20: AT CUTOFF NO POWER IS DELIVERED TO THE TRANSISTOR

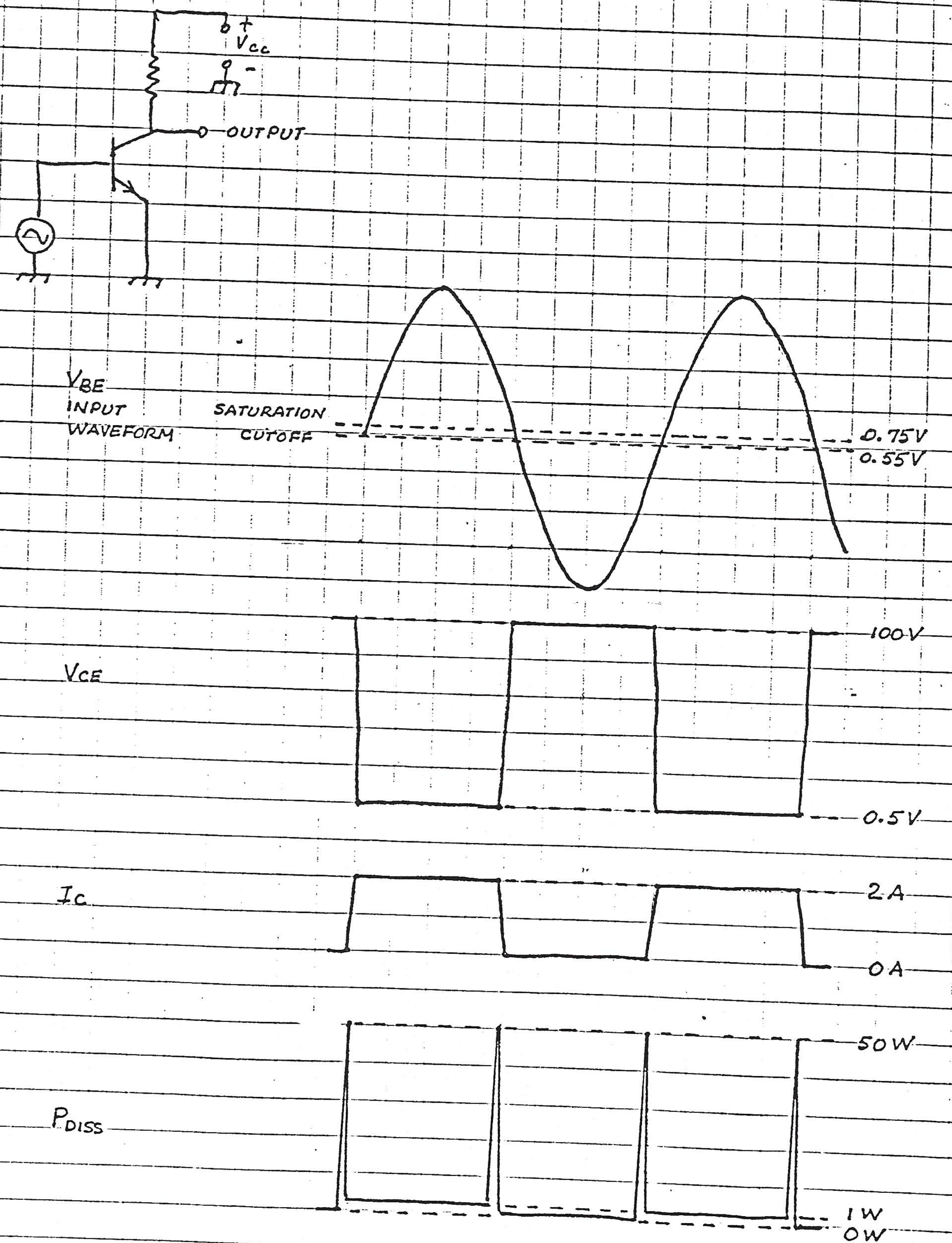


FIGURE 22

SWITCHING TRANSISTOR SHOWING LOW POWER DISSIPATION DUE TO SHORT SWITCHING TRANSIENTS. THIS IS ACCOMPLISHED BY A LARGE DRIVE SIGNAL.

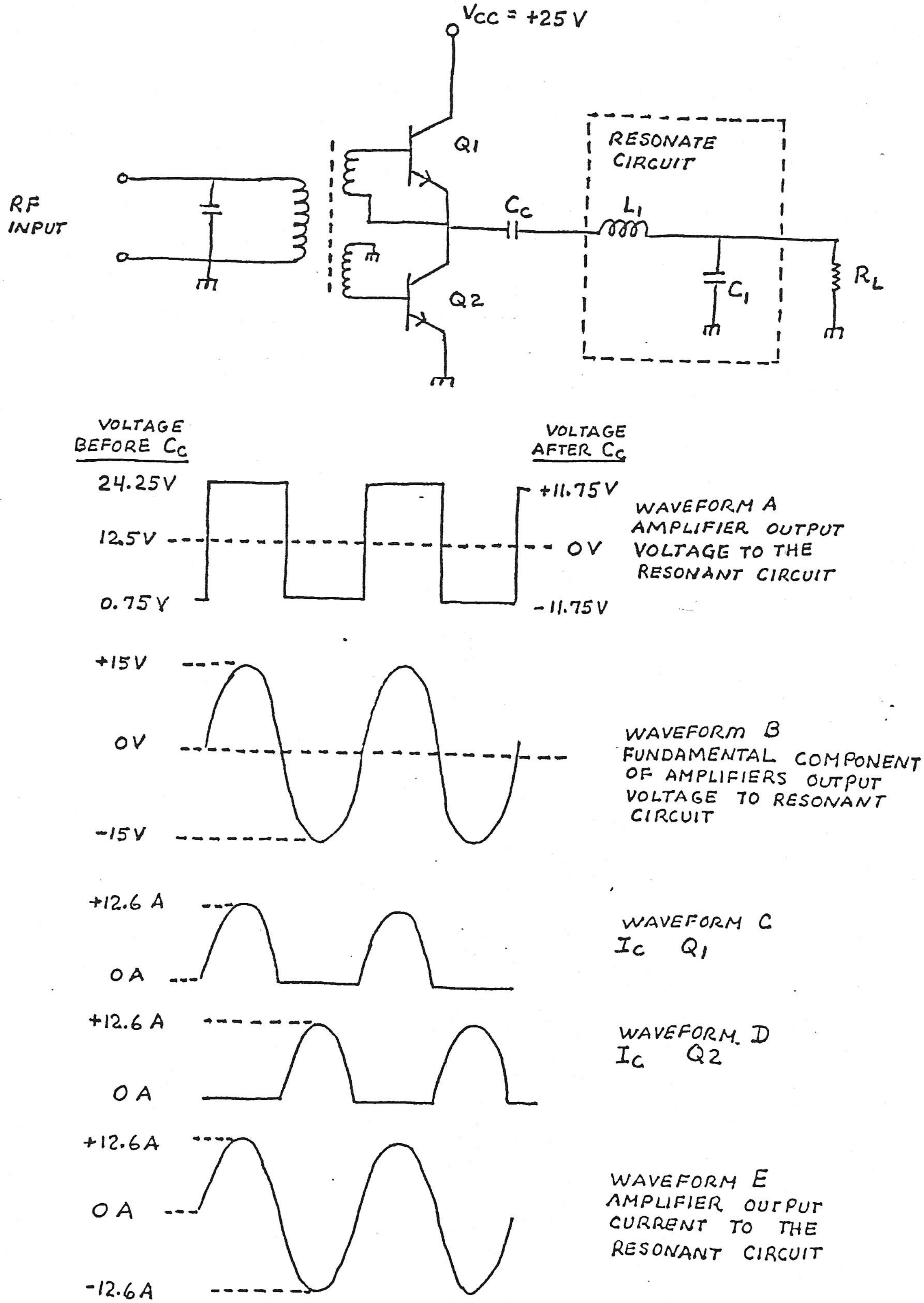


FIGURE 23: A CLASS D PUSH PULL R.F. POWER AMPLIFIER WITH ITS WAVEFORMS

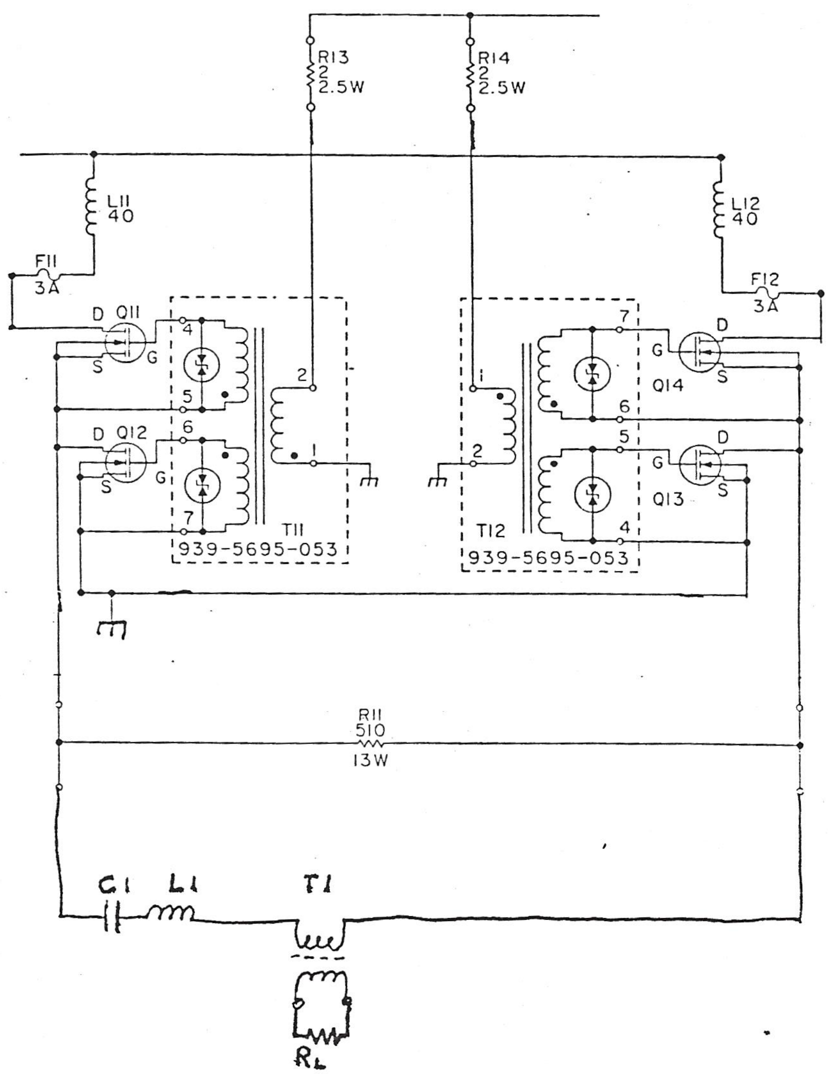


FIGURE 24: FET CLASS D R.F. POWER AMPLIFIER

STANDING WAVES

IF A TRANSMISSION LINE IS TERMINATED SUCH THAT $R_L = Z_0$ A UNIFORM TRAVELING WAVE WILL EXIST (THE INCIDENT WAVE), AND THE VOLTAGE ALL ALONG THE TRANSMISSION LINE WILL BE UNIFORM. IF THE TRANSMISSION LINE HAS LOSSES, THE VOLTAGE WILL DECREASE AS THE DISTANCE FROM THE SOURCE INCREASES.

IF THE LINE IS NOT PROPERLY TERMINATED ($R_L \neq Z_0$) THE LINE WILL HAVE AN INCIDENT WAVE TRAVELING FROM THE SOURCE TO THE LOAD AND A REFLECTED WAVE TRAVELING FROM THE LOAD TO THE SOURCE. SINCE THESE TWO WAVES ARE TRAVELING IN THE OPPOSITE DIRECTIONS THEY WILL ADD AND CANCEL AT POINTS ALONG THE LINE.

WHERE THE WAVES ADD A MAXIMUM (ANTINODE) WILL EXIST AND THE CANCELATION PRODUCES A MINIMUM (NODE).

THE VOLTAGE STANDING WAVE RATIO (VSWR OR ρ) IS A RATIO OF THE MAXIMUM VOLTAGE ON THE LINE TO THE MINIMUM VOLTAGE ON THE LINE.

THE MAXIMUM VOLTAGE (ANTINODE) CAN BE NO GREATER THAN TWICE THE INCIDENT VOLTAGE AND THE MINIMUM VOLTAGE (NODE) CAN BE DOWN TO ZERO VOLTS.

THE VOLTAGE NODES AND ANTINODES ARE ONE QUARTER WAVELENGTH APART. A VOLTAGE NODE WILL BE FORMED AT THE SAME LOCATION AS A CURRENT ANTINODE AND VISA VERSA.

A VOLTAGE NODE (MINIMUM) WILL BE FOUND ONE QUARTER WAVELENGTH TOWARD THE GENERATOR FROM AN OPEN AND A CURRENT NODE (MINIMUM) WILL BE FOUND ONE QUARTER WAVELENGTH TOWARD THE GENERATOR FROM A SHORTED LINE.

IF A STANDING WAVE OCCURS ON A LINE THE VSWR WILL DECREASE AS THE DISTANCE FROM THE VSWR CAUSING DISTURBANCE INCREASES. THIS IS DUE TO THE TRANSMISSION LINE ATTENUATION (α).

THE FOLLOWING FORMULI APPLY:

$$\Gamma = \frac{R_L - Z_0}{R_L + Z_0} \qquad \Gamma = \frac{VSWR - 1}{VSWR + 1}$$

$$VSWR = \frac{E_{MAX}}{E_{MIN}} = \frac{E_i + E_R}{E_i - E_R} = \frac{1 + |\Gamma|}{1 - |\Gamma|}$$

IN THE ATTACHED CHART:

COLUMN 1 GIVES THE VSWR

COLUMN 2 GIVES THE VSWR IN db's WHERE:

$$VSWR\ db = 20 \log_{10} VSWR$$

COLUMN 3 GIVES THE ABSOLUTE VALUE OF THE REFLECTION COEFFICIENT (K)

WHERE $K = \Gamma = \frac{VSWR - 1}{VSWR + 1}$ AND $VSWR = \frac{1 + K}{1 - K}$

ALSO $K = \frac{R_L - Z_0}{R_L + Z_0}$.

COLUMN 4 IS THE RETURN LOSS IN db WHERE:

$$K\ db = 20 \log_{10} \frac{1}{K}$$

THIS TELLS HOW FAR THE REFLECTION LEVEL IS IN db's BELOW THE TRANSMITTERS OUTPUT POWER LEVEL IN db's.

EXAMPLE: A TRANSMITTER OUTPUT LEVEL IS 20 db K (WHERE 0 dbK = 1 KW). THE RETURN LOSS IS 30 db. WHAT IS THE LEVEL OF POWER BEING REFLECTED BACK TO THE TRANSMITTER?

REFLECTED POWER LEVEL = XMITR OUTPUT LEVEL POWER (−) RETURN LOSS

= 20 dbK −30 db

REFLECTED POWER LEVEL = −10 dbK

IN THE EXAMPLE −10 dbK (THE POWER RETURNING TO THE TRANSMITTER) IS EQUAL TO 100 WATTS.

COLUMN 5 K^2 REPRESENTS THE FRACTION OF THE TRANSMITTERS OUTPUT POWER THAT IS REFLECTED BACK TO THE TRANSMITTER.

EXAMPLE: A 100 KW TRANSMITTER HAS A K = 0.0316. HOW MUCH POWER IS BEING REFLECTED BACK TO THE TRANSMITTER?

REFLECTED $= P_{TRANSMITTER} \cdot K^2$

$= 100 \times 10^3 \, (0.0316)^2$

= 100 WATTS

COLUMN 6 $1-K^2$ IS THE PART OF THE POWER THAT IS TRANSMITTED.

EXAMPLE: A 100 KW TRANSMITTER HAS A $K = 0.0316$. HOW MUCH POWER IS TRANSMITTED?

$$P_{TRANSMITTED} = P_{TRANSMITTER}\ (1-K^2)$$

$$= 100 \times 10^3\ (1 - 0.0316^2)$$

$$P_{TRANSMITTED} = 99{,}900 \text{ WATTS}$$

COLUMN 7 $1-K^2 db = -10 \log_{10} (1-K^2)$ IS THE LEVEL OF THE TRANSMITTED PORTION OF THE POWER IN db's BELOW THE TRANSMITTER'S OUTPUT POWER.

EXAMPLE: A 100 KW TRANSMITTER HAS AN OUTPUT LEVEL OF 20 db K. WHAT IS THE LEVEL OF THE TRANSMITTED POWER IN dbK IF $K = 0.0316$

$$\text{TRANSMITTED LEVEL} = 20\,dbK - \left[-10 \log (1-K^2)\right]$$

$$= 20\,dbK - 4.345 \times 10^{-3}$$

$$\text{TRANSMITTED LEVEL} = 19.996\ dbK$$

Reflection Measurements

The table of reflection measurements is made up of seven columns, computed as follows:

1. VSWR is voltage standing wave ratio tabulated from 1.000 to 4.000 in steps of .010.
2. VSWR (dB) = 20 $\log_{10}$ (VSWR).
3. K is the absolute value of the reflection coefficient:

$$K = \Gamma \qquad K = \frac{VSWR-1}{VSWR+1} \qquad k = \frac{R_L - Z_0}{R_L + Z_0} \qquad VSWR = \frac{1+k}{1-k}$$

4. K (dB) = 20 $\log_{10}$ (1/K). Return loss.
5. K^2 is the power reflection coefficient, or the fraction of power reflected.
6. $1-K^2$ is the fraction of power transmitted.
7. $1-K^2$ (dB) = $-10 \log_{10} (1-K^2)$.

VSWR	VSWR (dB)	K	K (dB)	K^2	$1-K^2$	$1-K^2$ (dB)
1.000	0.000	0.0000	∞	0.00000	1.00000	0.0000
1.001	.008361	.000490	66.20	.0000002	.9999998	.000089
1.002	.017077	.000999	60.01	.0000010	.9999990	.000089
1.003	.025793	.001498	56.49	.0000022	.9999978	.000089
1.004	.034332	.001996	54.00	.0000040	.9999960	.000089
1.005	.043049	.002494	52.06	.0000062	.9999938	.000089
1.006	.051765	.002991	50.48	.0000090	.9999910	.000089
1.007	.060304	.003488	49.15	.0000122	.9999872	.000089
1.008	.069020	.003984	47.99	.0000159	.9999841	.000089
1.009	.077559	.004480	46.97	.0000201	.9999799	.000089
1.010	.086	.0050	46.02	.00003	.99997	.0001
1.020	.173	.0099	40.09	.00010	.99990	.0004
1.030	.256	.0148	36.60	.00022	.99978	.001
1.040	.340	.0196	34.15	.00038	.99962	.002
1.050	.424	.0244	32.25	.00060	.99940	.003
1.060	.506	.0291	30.72	.00085	.99915	.004
1.070	.588	.0338	29.42	.00114	.99886	.005
1.080	.668	.0385	28.29	.00148	.99852	.006
1.090	.749	.0431	27.31	.00186	.99814	.008
1.100	.828	.0476	26.45	.00227	.99773	.010
1.110	.906	.0521	25.68	.00271	.99729	.012
1.120	.984	.0566	24.94	.00320	.99680	.014
1.130	1.062	.0610	24.29	.00372	.99628	.017
1.140	1.138	.0654	23.69	.00428	.99572	.019
1.150	1.214	.0698	23.12	.00487	.99513	.021
1.160	1.289	.0741	22.60	.00549	.99451	.024
1.170	1.364	.0783	22.12	.00613	.99387	.027
1.180	1.438	.0826	21.66	.00682	.99318	.030
1.190	1.511	.0868	21.23	.00753	.99247	.033
1.200	1.584	.0909	20.83	.00826	.99174	.036
1.210	1.656	.0950	20.44	.00903	.99097	.039
1.220	1.727	.0991	20.08	.00982	.99018	.043
1.230	1.798	.1031	19.73	.01063	.98937	.046
1.240	1.868	.1071	19.40	.01147	.98853	.050
1.250	1.938	.1111	19.09	.01234	.98766	.054
1.260	2.007	.1150	18.78	.01323	.98677	.058
1.270	2.076	.1189	18.49	.01414	.98586	.062
1.280	2.144	.1228	18.22	.01508	.98492	.066
1.290	2.212	.1266	17.51	.01603	.98397	.070
1.300	2.278	.1304	17.70	.01700	.98300	.074
1.310	2.345	.1342	17.44	.01801	.98199	.079
1.320	2.411	.1379	17.21	.01902	.98098	.083
1.330	2.477	.1416	16.98	.02005	.97995	.088
1.340	2.542	.1453	16.76	.02111	.97889	.093
1.350	2.607	.1489	16.54	.02217	.97783	.097
2.000	6.0206	0.3333	9.54	0.1111	0.88889	0.512
3.000	9.54	0.5	6.02	0.25	0.75	1.25

CAVITIES

QUARTER AND HALF WAVELENGTH
TUBE TYPE CAVITIES FOR FM
AND TV USE

BY: CLARENCE "DOC" DAUGHERTY

F.M. TUBETYPE CAVITY RF POWER AMPLIFIER CONSIDERATIONS

BACKGROUND:

When amplifiers were first built they were able to handle audio frequencies only. As the operating frequency increased the interelectrode capacity of the vacuum tube and the distributed capacity of the circuit (lumped together they can be referred to as stray capacity, C_s) tends to shunt the signal to ground. (See Figure 1). This limits the high frequency response of the amplifier.

One method used to overcome the shunting effects of stray capacity (C_s) is to lower the resistive load impedance of the amplifier.

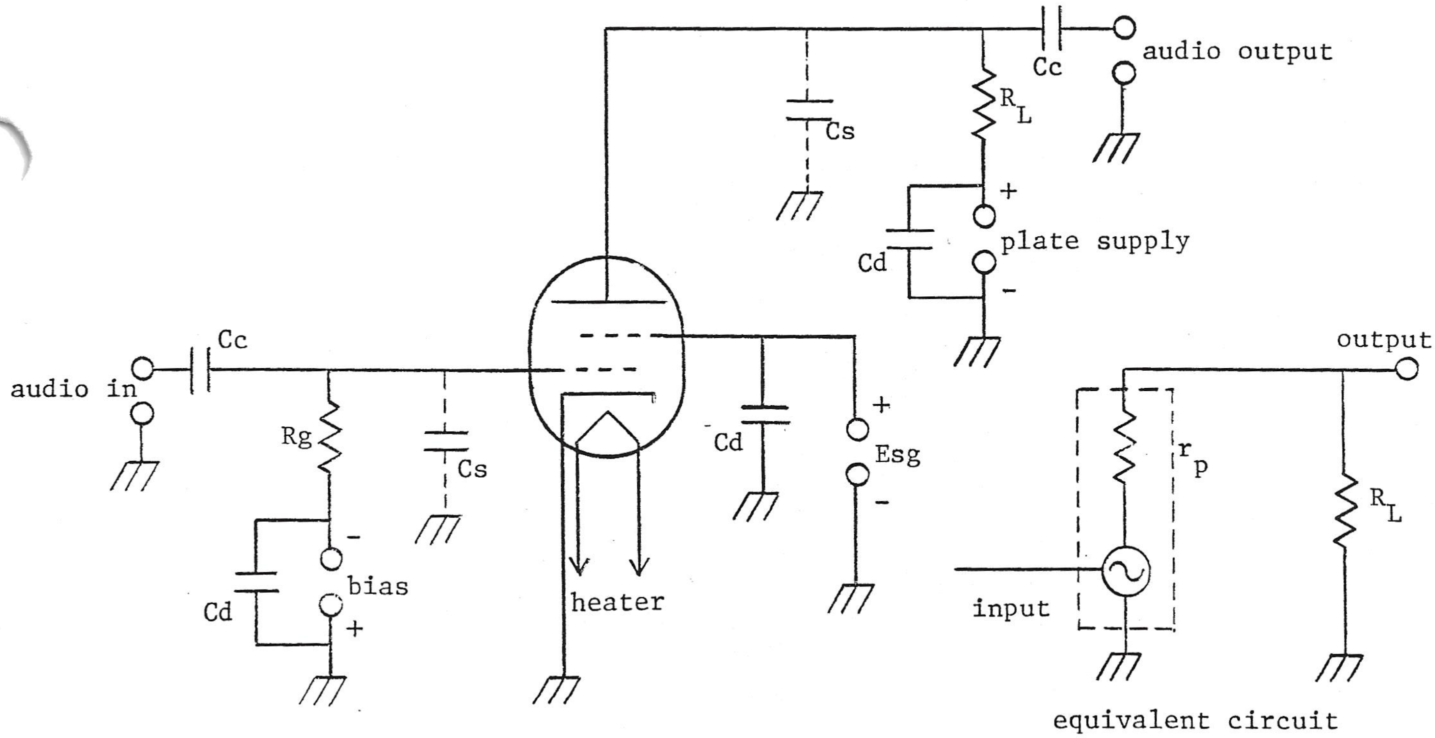


FIGURE 1 - A tube type amplifier and the equivalent circuit

High frequency rolloff is determined by the formula:

$$f_{max} = \frac{1}{2 \pi R C_s}$$

Where: f_{max} = Maximum frequency output. At this point the voltage gain is 0.7 of maximum and the power gain is half of maximum (The -3db point).

R = The plate load resistance or the grid load resistance

C_s = The plate circuit stray capacity if R_L is used, or the input circuit stray capacity if R_g is used.

The voltage gain of the amplifier is determined by the formula:

$$A_v = \frac{-R_L}{R_L + r_p}$$

Where: A_v = Amplifier voltage gain

r_p = The internal plate resistance of the tube

R_L = The plate load resistance of the tube.

From the above formuli it is apparent that as the resistance of R_L decreases, the maximum operating frequency increases, but at the expense of voltage gain. (See circuit in Figure 1).

Another method of eliminating the shunting effects of stray capacitance is to make it a part of a resonant circuit. To do this R_L is replaced by an inductance. The inductance is resonated by the stray capacity (C_s) and an additional capacity (C_1 and C_2 shown in Figure 2 A).

The resonant circuit offers a high impedance for the tube's input and output circuits at the operating frequency. This produces a high amplifier gain. Notice that C_s is now part of the resonant circuit of the amplifier and no longer bypasses the desired signal.

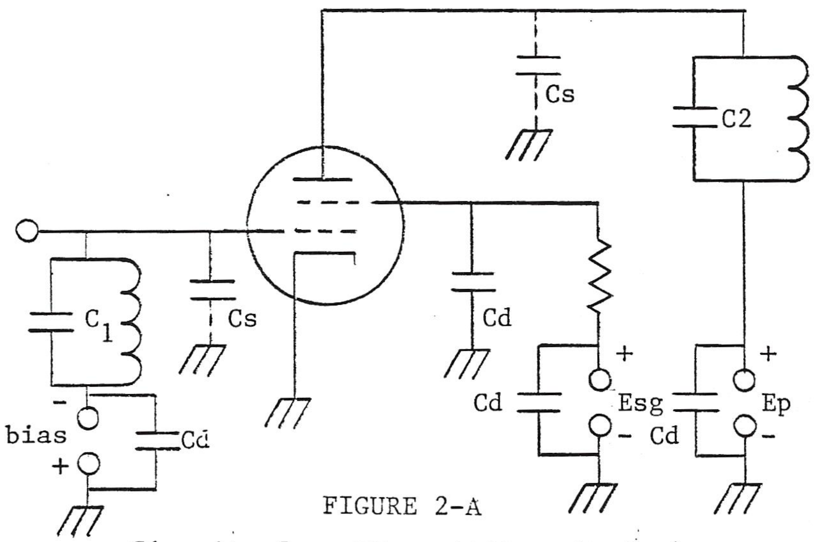


FIGURE 2-A

Circuit of an RF amplifier showing the input and output stray capacity and tuned circuits

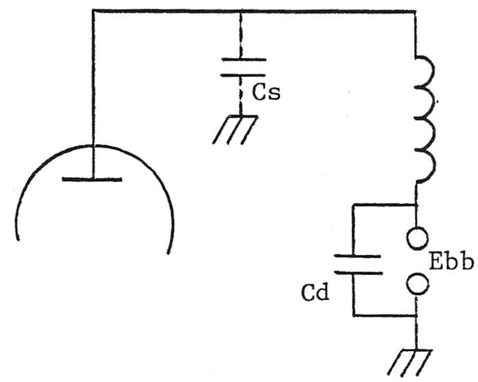


FIGURE 2-B

Plate circuit of higher frequency amplifier where C_s is the resonating capacity.

The scheme worked fine when 10 MH_z was the highest usable frequency. As higher frequencies were used several new problems became apparent. They were: skin effect, decreasing values of L and C required to resonante the circuit, and stray inductance.

REDUCING THE VALUE OF INDUCTANCE AND CAPACITANCE

The formula for resonant frequency is:

$$f_R = \frac{1}{2 \pi \sqrt{L C}}$$

Where: f_R = Resonant frequency in Hertz

L = Inductance in Henries

C = Capacitance in Farads

To increase the resonant frequency, L and/or C must be reduced. In the RF amplifier the value of the added C can be reduced until C_s is the only resonating capacity in the circuit. (See Figure 2-B).

The value of inductance is reduced by winding fewer turns on the coil and keeping the coil's diameter small. Eventually as the frequency increases the coil might have less than one turn or even be a straight conductor. The inductance of the straight conductor is determined by its length and diameter.

The longer the conductor, the greater its inductance will be.

The greater the area of the conductor, the smaller the inductance will be.

SOLVING THE SKIN EFFECT PROBLEM

Skin effect is a condition that causes RF currents to flow only on the surface of the conductor at higher frequencies. As the frequency of the RF increases, the RF current tends to flow closer to the surface of the conductor. At VHF TV and FM frequencies, the conducting layer is only a few thousandths of an inch thick. Thus the RF resistance of a conductor is much greater than the DC resistance. This necessitates the use of larger conductors with greater surface area for high power use at VHF frequencies. This will keep skin effect losses low. Silver plating is often used on VHF RF components to lower the RF resistance.

STRAY INDUCTANCE

We now have the stray capacity of the circuit resonating an inductor that consists of a large area conductor with few or no turns. This gives rise to the problem of stray inductance. The leads that connect the various elements of the tube to the amplifier circuitry now contain much inductance compared to the rest of the amplifier's circuitry. This inductance can act as:

(1) An RF choke which will reduce the RF output.

(2) Part of another, unplanned, resonant circuit with the stray capacity of the tube. This can cause the tube to have parasitic oscillations.

PARASITIC OSCILLATIONS

Parasitic oscillation can occur within the tube or can occur because of the tube and its external circuitry. They are stopped by several methods. The most common method is the losser resistance (also called the parasitic suppressor). It is a small value of resistance (usually less than 100 ohms) found in series with the plate, screen, or grid circuits of the tube. It lowers the Q of circuit that causes the parasitic oscillations and eliminates them.

A material called Eccosorb is also used in cavity type RF amplifiers to prevent parasitic oscillations. It offers resistance to the parasitic oscillations and can be in the form of hard blocks or various molded shapes, or thin flexible sheets. It is used in selected places in the RF path of the amplifier to lower the Q of the parasitic oscillation circuits and eliminate them.

Another consideration of operation at VHF frequencies is that unlike operation at lower frequencies, relatively pure lumped components of resistance, capacitance, and inductance are extremely difficult to produce. All the components will exhibit considerable values of resistance, inductance and capacitance.

Thus as frequency increases, lumped component resonant circuits get smaller and smaller (to reduce L and C); larger in diameter (to reduce skin effect); closer to the tube to reduce the effects of stray L; and there is great difficulty in predicting exactly what values of R, L, and C a component or circuit may have.

These problems can be managed in low power circuits but with high power circuits, arcs and shorts due to high DC and RF voltages become a problem. Larger size and spacing of components is a good start towards arc and short prevention, but this is in opposition to the smaller size and spacing dictated by the high frequency operation. Also in high power circuits, the unpredictability of the circuit values of R, L, and C make it difficult to control the vitally important parameters of dissipation, efficiency, and reliability of operation.

THE TRANSMISSION LINE CAVITY

One solution to the above problems is the resonant transmission line cavity amplifier. In this type of amplifier the tube becomes part of a resonant transmission line. The elements of these tubes are arranged to look like concentric coaxial transmission lines. The design of these tubes stresses low interelectrode capacity and low distributed inductance. The stray (interelectrode and distributed) capacity and inductance of the tube becomes part of the resonant transmission line. The resonant transmission line is physically larger than the equivalent lumped constant L-C resonant circuit operating in the same frequency. This larger physical size aids in solving the high power operation problems of skin effect losses, prevention of arcs and shorts, and reliable and predictable operation. Before we can study the transmission line cavity, we must review the basics of resonant transmission lines.

REVIEW OF RESONANT TRANSMISSION LINES

Any transmission line (parallel wire, coax, or the microstrip type used on printed circuits) has a characteristic impedance. If a transmission line is terminated in its characteristic impedance by a resistive load (Example: a 50 ohm resister terminating a 50 ohm line), a signal traveling down the line will be totally absorbed by the load and no reflection will result. If the impedance of the termination (Z_L) does not equal the impedance of the line (Z_o), or if Z_L does equal Z_o but it is not totally resistive, a signal sent down the line will not be totally absorbed by the termination. Some or all of it will be reflected back down the line to the source.

FOUR SPECIAL CASES OF RESONANT LINES

There exists four special cases of improper termination of a transmission line. They are:

(1) A quarter wavelength section with termination open

(2) A quarter wavelength section with termination shorted

(3) A half wavelength section with termination open

(4) A half wavelength section with termination shorted.

The symbol for wavelength is λ .

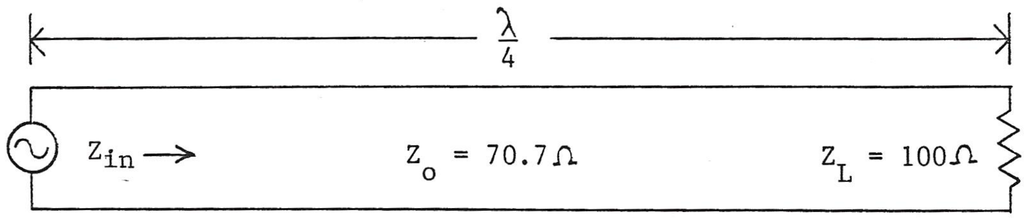
GENERAL RULES FOR RESONANT TRANSMISSION LINES

Two general rules exist for these four cases. They are:

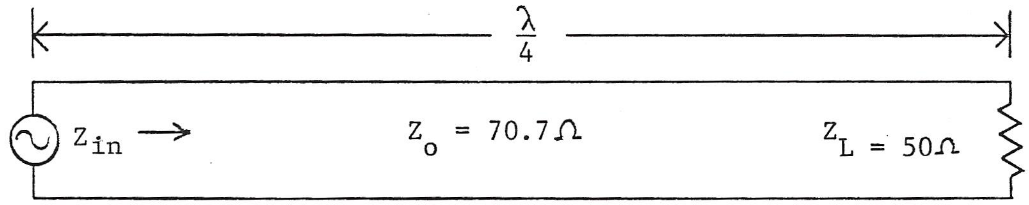
(1) A quarter wavelength of transmission line inverts impedance.

(a) If the Z_L is less than Z_o, then the input impedance of the line(Z_{in}) will be greater than Z_o. (See Figure 3)

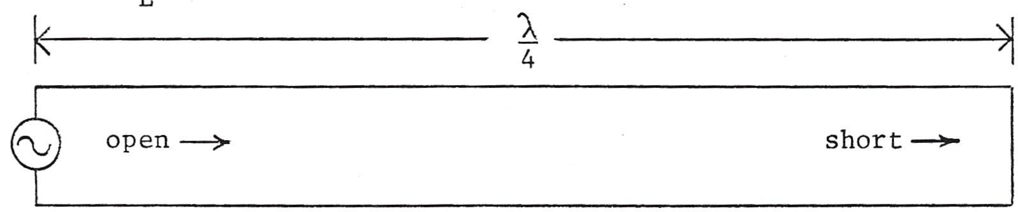
(b) If Z_L is greater than Z_o, then Z_{in} will be less than Z_o. (See Figure 3)



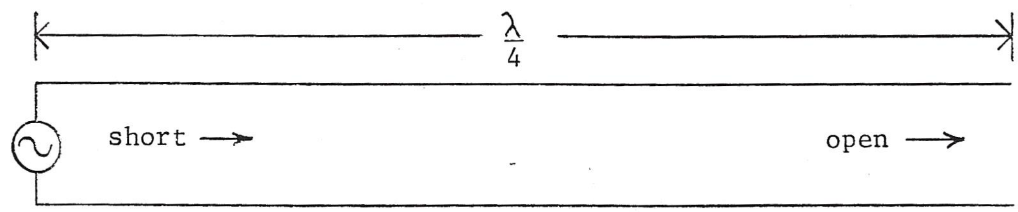
A. $Z_{in} = \dfrac{Z_o^{\,2}}{Z_L} = 50\Omega$



B. $Z_{in} = \dfrac{Z_o^{\,2}}{Z_L} = 100\Omega$



C. A shorted termination yields an open source.
The source is equivalent to a parallel resonant circuit.



D. An open termination yields a shorted source.
The source is equivalent to a series resonant circuit.

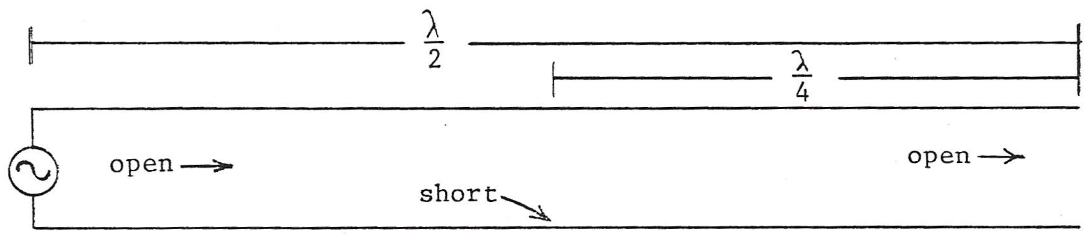
FIGURE 3 - Four examples of improper termination (Z_L) that shows the inverting properties of a quarter wavelength of a transmission line. The line used in Examples (A) and (B) had a characteristic impedance (Z_{in}) of 70.7 ohm, but the line impedance (Z_o) of A through D could be any value and still yield correct results using the formula:

$$Z_{in} = \frac{Z_o^{\,2}}{Z_L}$$

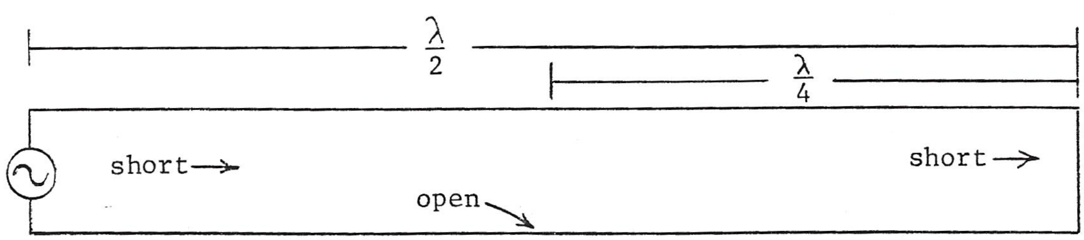
(2) A half wavelength section of transmission line repeats impedance.

(a) Think of it as two quarter wavelength sections in series. (See Figure 4 A and B).

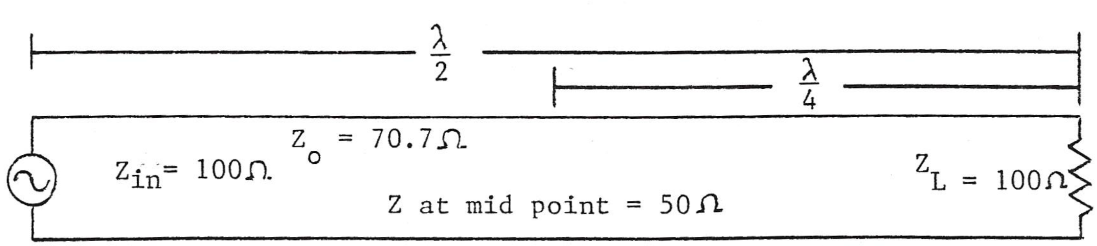
(b) The input impedance (Z_{in}) always equals the load impedance. (See Figure 4 C).



A. An open termination of a half wavelength section yields an open source. The source is equivalent to a parallel resonant circuit.



B. A shorted termination of a half wavelength section yields a shorted source. This is equivalent to a series resonant circuit.



C. A half wavelength line repeats impedance. At $\lambda/4$ from the termination $Z_{mid} = \frac{Z_o^2}{Z_L} = 50\Omega$ $\lambda/4$ from this point (the source) $Z_{in} = \frac{Z_o^2}{Z_{mid}} = 100\Omega$

FIGURE 4 - Conditions of a half wavelength of transmission line. The Z_o of the lines in Figures A, B, C can be any value, but Figure C is shown in this example at Z_o = 70.7 ohms.

THE SHORTED QUARTER WAVELENGTH LINE

A shorted quarter wavelength transmission line has a high (almost open), purely resistive input impedance. Electrically it looks like a parallel resonant circuit. If the applied frequency is changed slightly so that the shorted line is no longer one quarter wavelength long, the input impedance drops and no longer remains purely resistive. (See Figure 6).

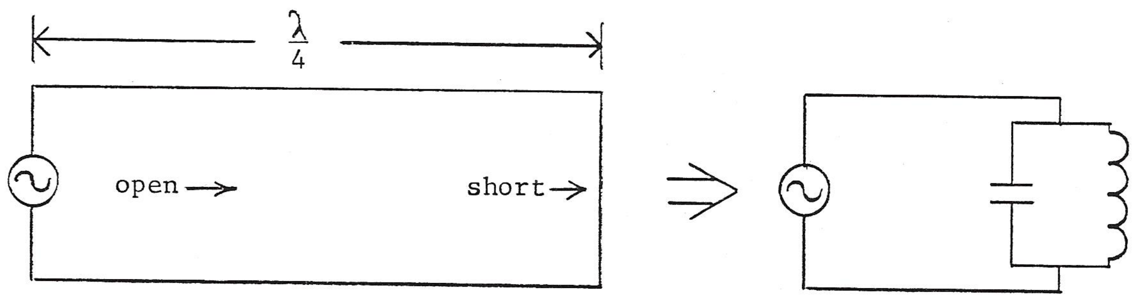
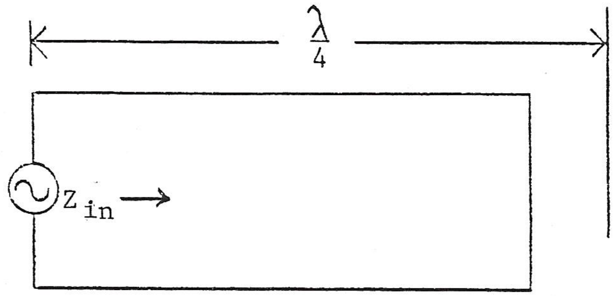
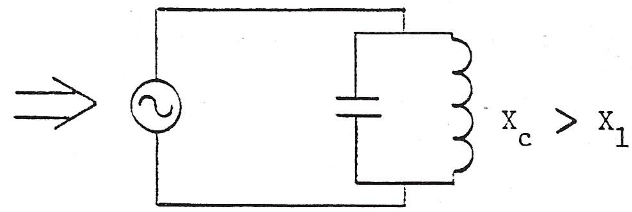


FIGURE 5 - A shorted, quarter wavelength line looks like a parallel resonant circuit.

THE SHORTED TRANSMISSION LINE LESS THAN A QUARTER WAVELENGTH LONG

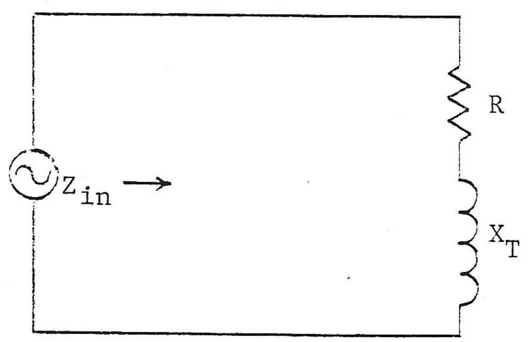
When operated at a frequency below that for which the shorted line is one quarter wavelength long, the physical length of the line at the new lower frequency will be less than one quarter wavelength. The impendance will be lower and the line will look inductive. Actually, the impedance will be a combination of resistance and inductance. As the applied frequency is lowered further, the resistance will become lower, the inductance will become greater, and the impedance will become smaller. (See Figure 6 A and C). This same effect is seen on a parallel resonant circuit when it is operated below resonance (See the Example in Figure 6).





A. The applied frequency is lower than the frequency for which the line is $\lambda/4$. The Z_{in} is inductive and resistive.

B. The equivalent circuit of A. A parallel resonant circuit is operated below resonance, X_c is greater than X_L and i_L is greater than i_c.



X_T is the total equivalent reactance. I_T can be inductive or capacitive. In this case it is inductive.

C. The equivalent circuit of A and B showing how Z_{in} appears.

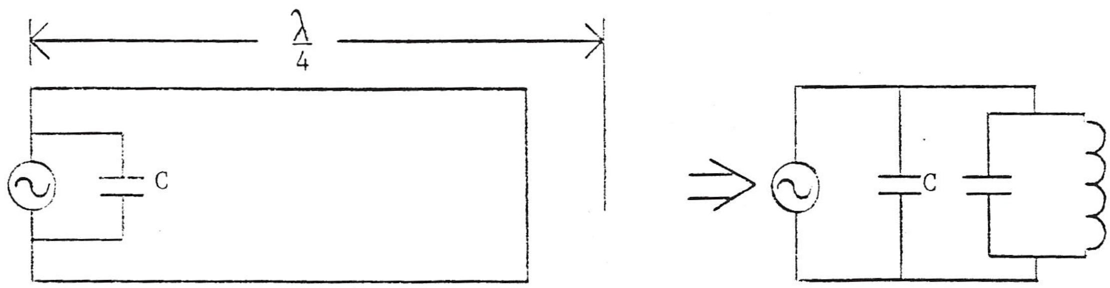
FIGURE 6 - The shorted line less than $\frac{\lambda}{4}$ and its eqivalent circuits.

In Figure 6-C, the equivalent circuit of the shorted line less than $\frac{\lambda}{4}$ and the parallel resonant circuit operated below resonance are both inductive. In Figure 6-B, if the capacitive reactance (X_c) is greater than the inductive reactance (X_L), than the current flow through the inductance (i_L) will be greater than the current through the capacitance (i_c).

NOTE: The current i_L and i_c are 180° out of phase and would cancel at resonance, where $X_c = X_L$ and $i_c = i_L$. The generator supplies a small current to make up for circuit losses.

Below resonance where i_L is greater than i_c, the currents do not cancel and the generator's current is equal to that value of i_L not cancelled by i_c. To solve this problem and reresonate the circuit, we need only to make X_L equal to X_c.

In Figure 7, the circuits shown in A and B (in both cases the applied frequency is lower than the resonant frequency) are resonated by adding parallel capacity. In Figure 7-B, this lowers X_c to make it equal to X_L. In Figure 7-A, the shorted line is physically less than $\frac{\lambda}{4}$, but it has been electrically lengthened to $\frac{\lambda}{4}$ by the parallel capacity and is again resonant.

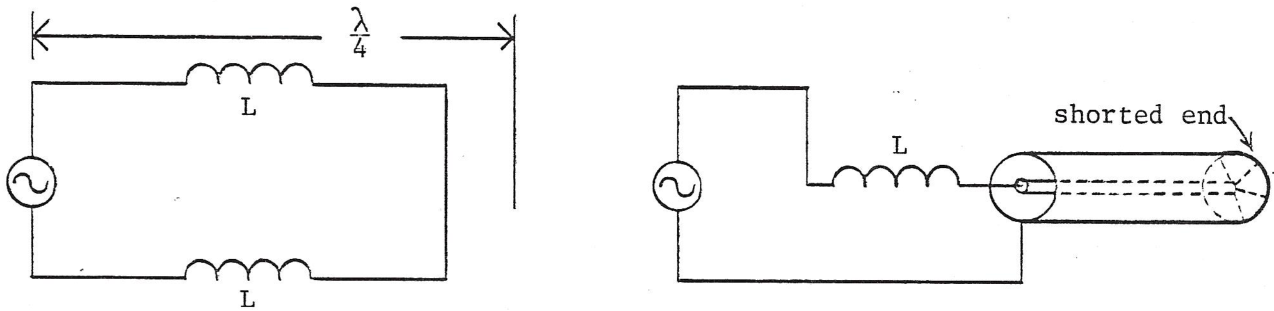


A. Capacitance (C) is added to electrically lengthen the line to $\lambda/4$ (resonate it).

B. Capacitance (C) is added to lower X_c and make it equal to X_L and resonate the circuit.

FIGURE 7 - Applied frequency is lower than resonant frequency and capacitance is added to resonate the circuit.

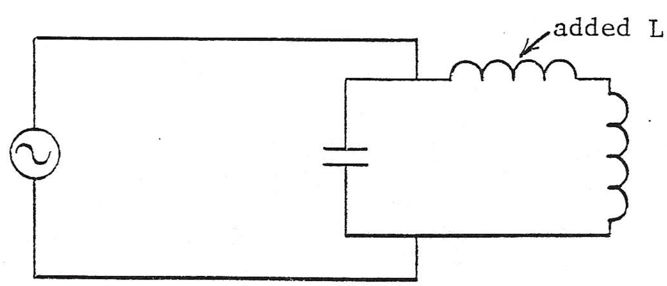
Another approach to resonating a circuit where the applied frequency is lower than the resonant frequency is to add a series inductance. In Figure 8-B, a small amount of series inductance is added to increase the X_L and make it equal to X_c. In Figure 8-A, the series inductance (or inductances in the case of a balanced line) electrically lengthen the physically short line to $\frac{\lambda}{4}$ and resonate it.



parrallel line

coaxial line

A. Adding series inductance somewhere in the physcially short line to electrically lengthen it and cause it to resonate.



B. Inductance is added to increase X_L and make it equal to X_c.

FIGURE 8 - When applied frequency is lower than resonant frequency, series inductance can be used to reresonate the circuit to this new lower frequency.

This concept is not new to many of you. How do you electrically lengthen an antenna that is physically too short? In an antenna system, series inductance (a loading coil) is used to lengthen an antenna whose length is not equal to $\frac{\lambda}{4}$ (vertical) or $\frac{\lambda}{2}$ (horizontal). See Figure 9.

Parallel, or shunt, capacity can also electrically lengthen an antenna that is physcially too short. In Figure 10, capacitive top hat loading electrically lengthens a vertical antenna. Shunt capacity can be used to lengthen a horizontal antenna, but it is less common.

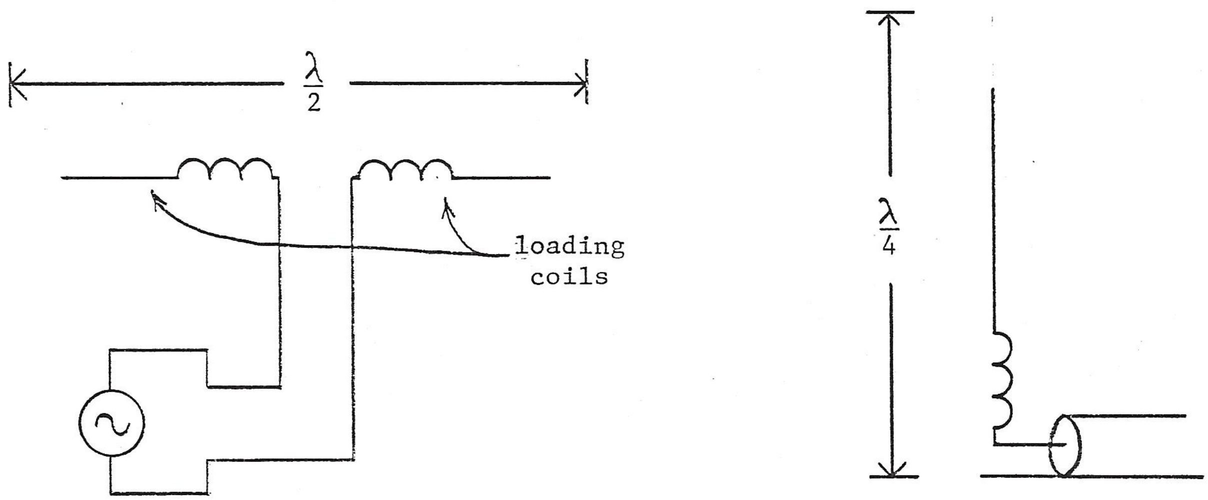


FIGURE 9 - Inductive loading electrically lengthens (and resonates) a physically short vertical and horizontal antenna.

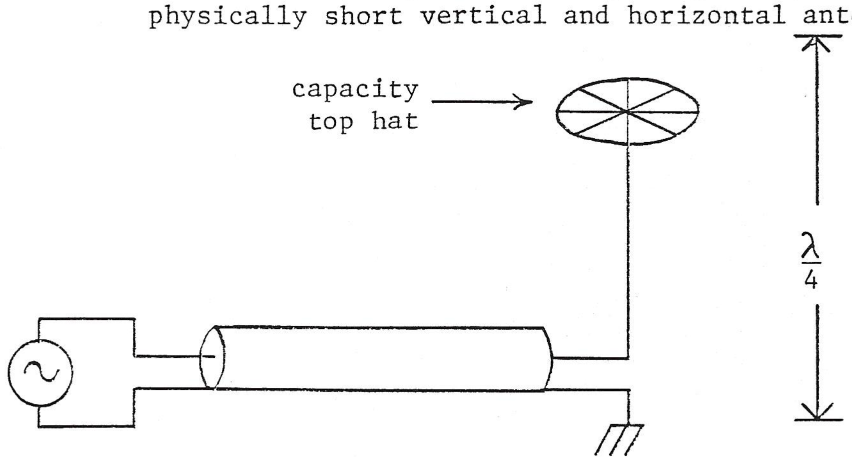


FIGURE 10 - A capacity top hat loading disc adds the shunt capacity that electrically lengthens the physically short antenna and resonates it.

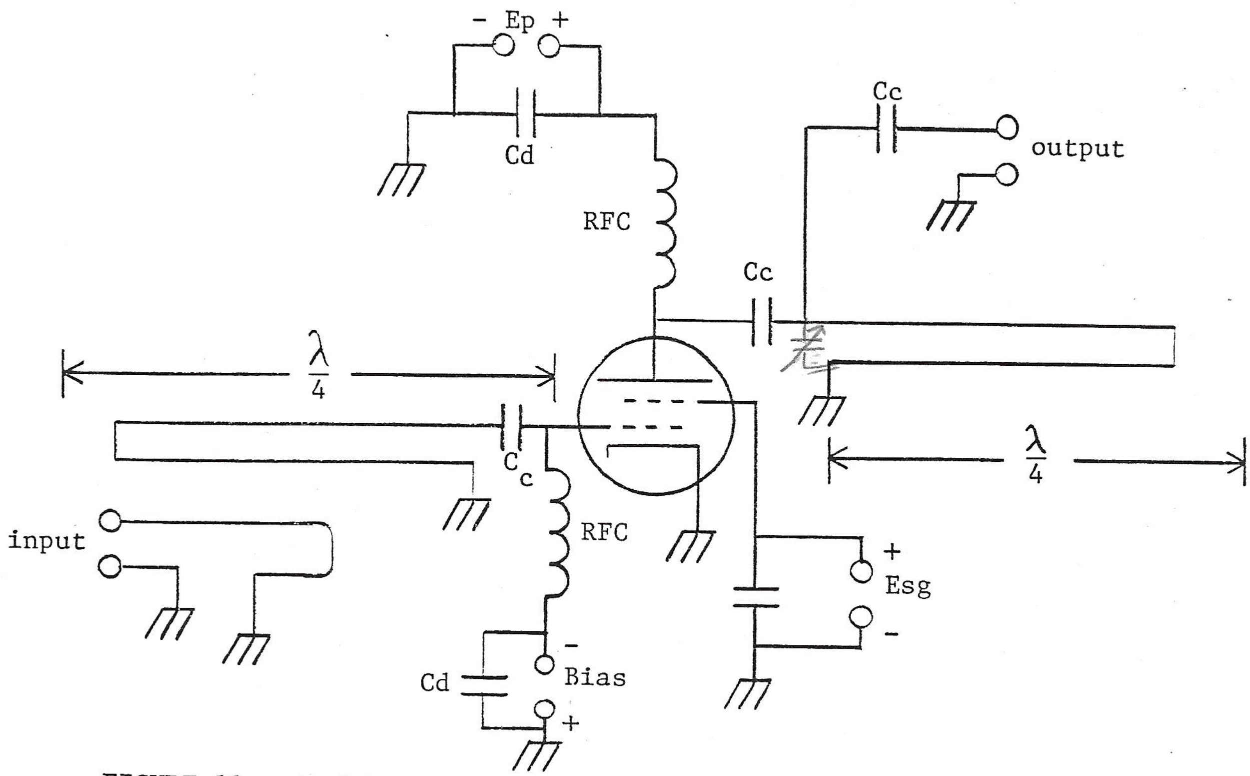


FIGURE 11 - A shorted, quarter wavelength transmission line amplifier.

C_c = Coupling capacitor

C_d = Decoupling capacitor

RFC = Radio frequency choke

In Figure 11, shorted transmission lines are used to resonate the inputs and outputs of this amplifier. Notice that the length of the lines are less than $\frac{\lambda}{4}$ but the tubes shunt input and output capacity and its series lead inductance will electrically lengthen and resonate the transmission lines. The input is shown inductively coupled, but it could just as easily have been capacitively coupled to the grid. The input could also have a lumped constant resonant circuit or a transmission line resonant circuit since its power level is low. The output coupling is capacitive, but it also could been inductive.

THE OPEN HALF WAVELENGTH TRANSMISSION LINE

The open ended half wavelength transmission line displays the same input characteristics as the shorted quarter wavelength transmission line. At $\frac{\lambda}{2}$ it acts like a parallel resonant circuit. It acts like an inductive/resistive circuit when it is shorter than $\frac{\lambda}{2}$, but greater than $\frac{\lambda}{4}$. As its wavelength progresses from $\frac{\lambda}{2}$ to just greater than $\frac{\lambda}{4}$, its impedance decreases and it gets more inductive. An open ended transmission line shorter than $\frac{\lambda}{2}$ can be electrically lengthened (brought to resonance at a lower frequency) by adding shunt capacity and/or series inductance.

THE OPEN HALF WAVELENGTH TRANSMSSION LINE AMPLIFIER

When incorporated into an amplifier circuit, the input may consist of a transmission line type resonant circuit or a lumped constant type resonant circuit. The lumped constant resonant circuit can be used at the input because of its lower power requirement.

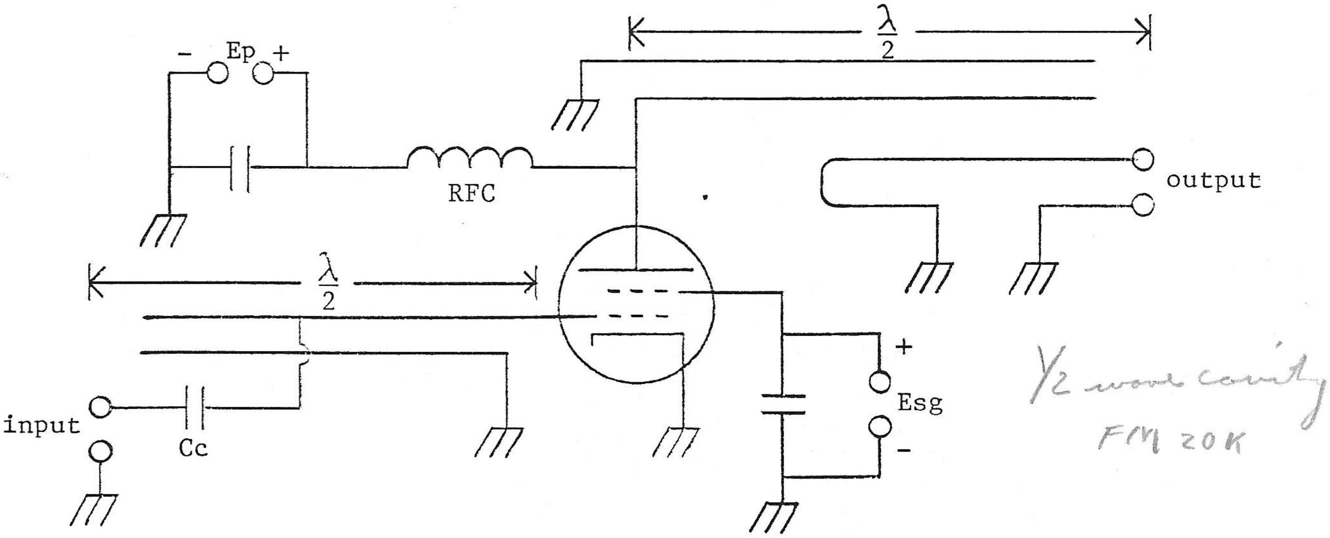


FIGURE 12 - A half wave open ended transmission line amplifier. The input and output can be inductively or capactively coupled.

The input and output can be either inductive loop coupled or capacitively coupled. The transmission lines used in this amplifier are shorter than $\frac{\lambda}{2}$. The stray inductance and capacitance of the tube becomes part of the resonant circuit and electrically lengthens (lower the resonant frequency) of the input and output transmission lines. (See Figure 12).

REVIEW OF COUPLED PARALLEL INDUCTORS

To understand one feature of the FM 2.5K through FM 20K FM transmitter RF power amplifiers, it is necessary to review inductively coupled, parallel inductors.

If two inductors are connected in parallel but not magnetically coupled, the total inductance will be equal to the formula:

$$L_T = \frac{L_1 \times L_2}{L_1 + L_2} \qquad \text{or} \qquad L_T = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2}}$$

Assume each inductor is 1 uh. The total inductance will be 0.5 uh. (See Figure 13-A).

If two inductors are positioned so that there is an aiding magnet coupling, the inductance of each inductor will increase. This increased inductance is due to the mutual inductance provided by the magnetic coupling.

$$L_1' = L_1 + L_m \qquad \text{and} \qquad L_2' = L_2 + L_m$$

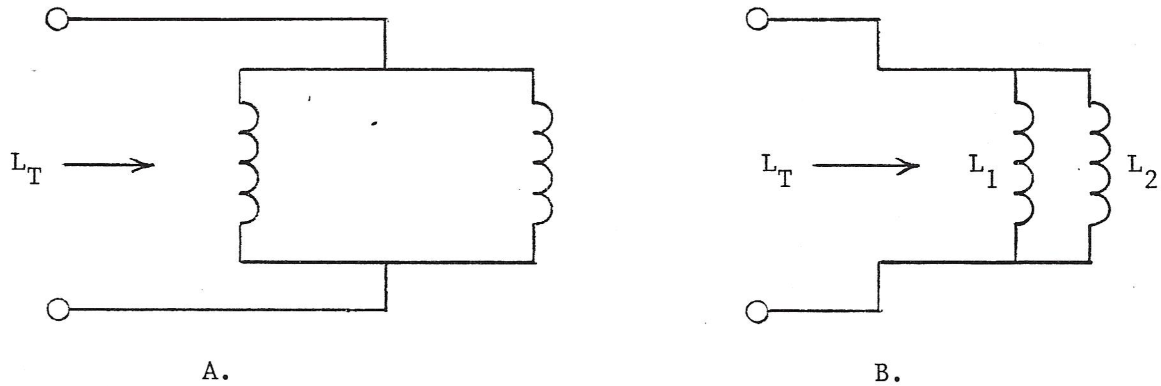
Where: L_m is the mutual inductance provided by the magnetic coupling.

If these two coupled inductors are placed in parallel (still magnetic aiding, the total inductance will be:

$$L_T = \frac{1}{\frac{1}{L_1 + L_m} + \frac{1}{L_2 + L_m}}$$

(See Figure 13-B)

FM 20K



FM 20 K

FIGURE 13 -(A) Two parallel 1 uh inductors without magnetic coupling.

$$L_T = 0.5 \text{ uh}$$

(B) Two parallel 1 uh inductors that have aiding magnetic coupling.

$$L_T = 0.55 \text{ uh}$$

If two parallel inductors are constructed so that their magnetic coupling can be changed, the total inductance can be varied over a small range.

INDUCTANCE OF A STRAIGHT CONDUCTOR

Every conductor has some value of inductance at VHF frequenceis. This inductance has a large effect on the circuit due to its high inductive reactance (X_L). $X_L = 2 \pi F L$. A small diameter conductor a few inches in length may act as an RF choke. A simple rule exists to help us control this value of inductance. Remembering that at VHF frequencies, the skin effect causes most, if not all of the RF current to flow on the surface of the conductor. The rule is this:

(1) As the surface area of a conductor of given length increases, its inductance decreases. A large area conductor may be thought of as many smaller area conductors in parallel.

The inductance of the large area conductor is less than the inductance of the small area conductor because of the law of parallel inductors.

$$L_T = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3}}$$

FM120K

(2) The inductance of a straight conductor is directly proportional to its length.

COAXIAL LINE FEATURES OF A TETRODE R.F. POWER AMPLIFIER TUBE

THE ANODE (PLATE)

The plate resembles a copper cup with half of the plate contact ring welded to the mouth and the cooling fins silver soldered or welded to the outside of the cup. (See Figure 14 and Figure 15).

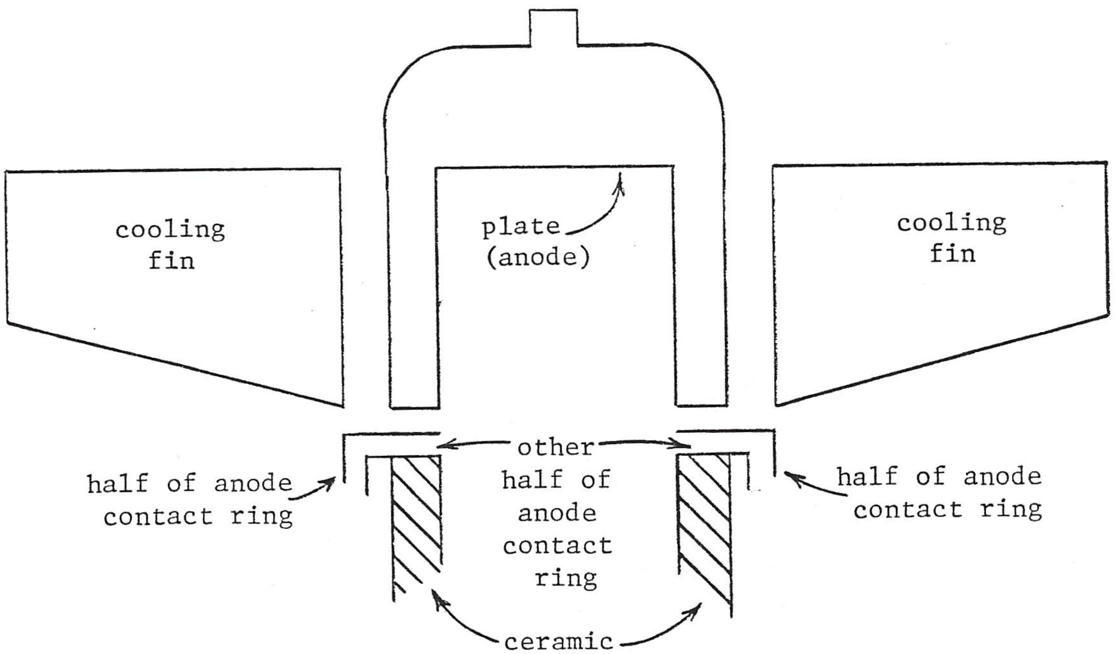


FIGURE 14 - Cut away view of the anode structure.

The other half of the anode contact ring is bonded to the base ceramic spacer. It fits into the half of the anode contact ring fastened to the plate structure, and the two halves are welded together. This ceramic spacer is the same ceramic that is shown above the screen contact ring in Figure 16.

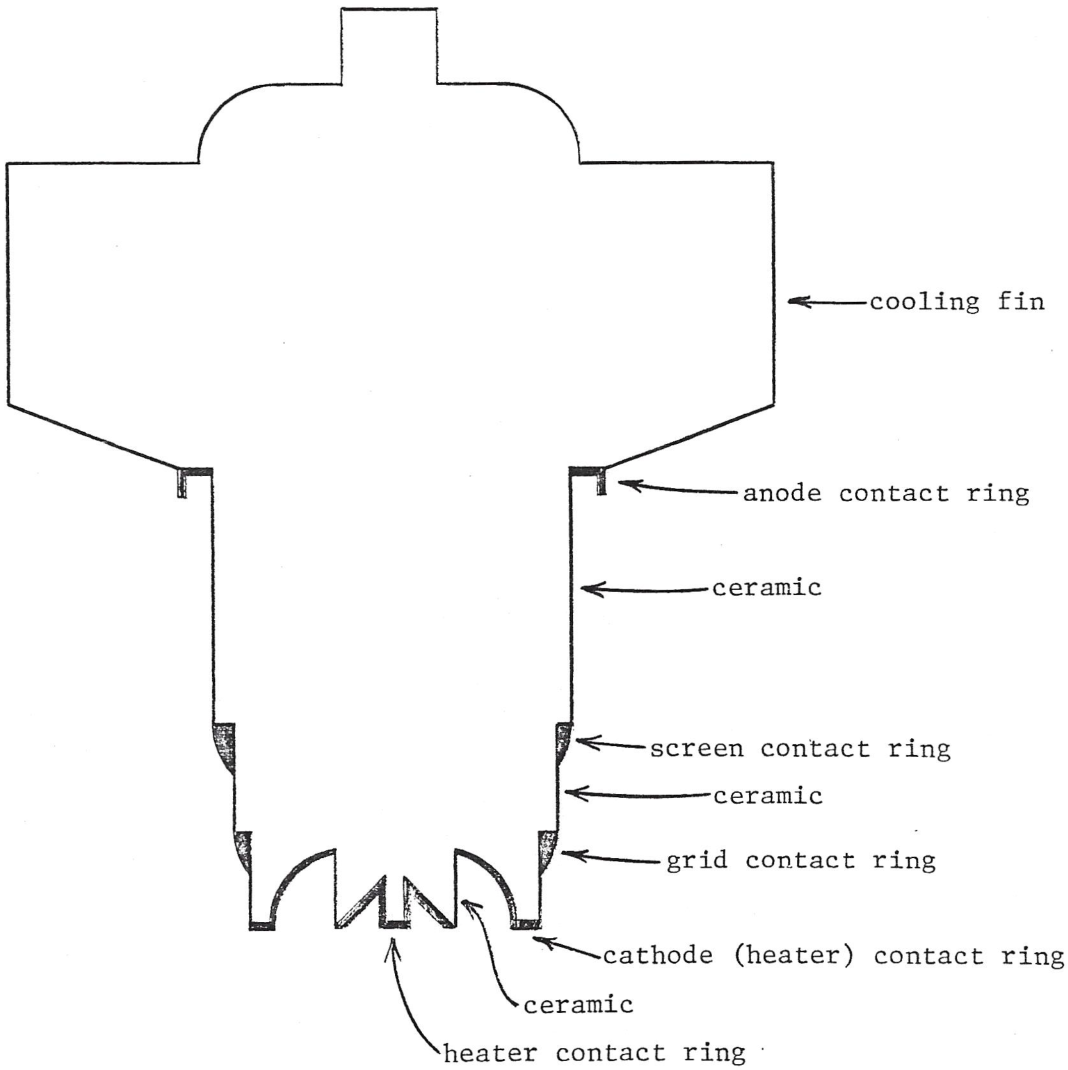


FIGURE 15 - A cutaway view of the exterior of a tetrode RF Power amplifier of the type used on F.M. transmitters.

THE SCREEN STRUCTURE

The screen grid consists of many vertical supports fastened to a metal base cone. The other end of the metal base cone fastens to the screen contact ring. The inductance of the individual vertical supports is reduced by building the screen grid of many of them in parallel. The vertical supports are held rigid by horizontal rings welded to them and a metal cap on the top of the assembly. The screen contact ring, metal base, and metal base cone also functions to reduce lead inductance and RF resistance due to skin effect. (See Figure 16).

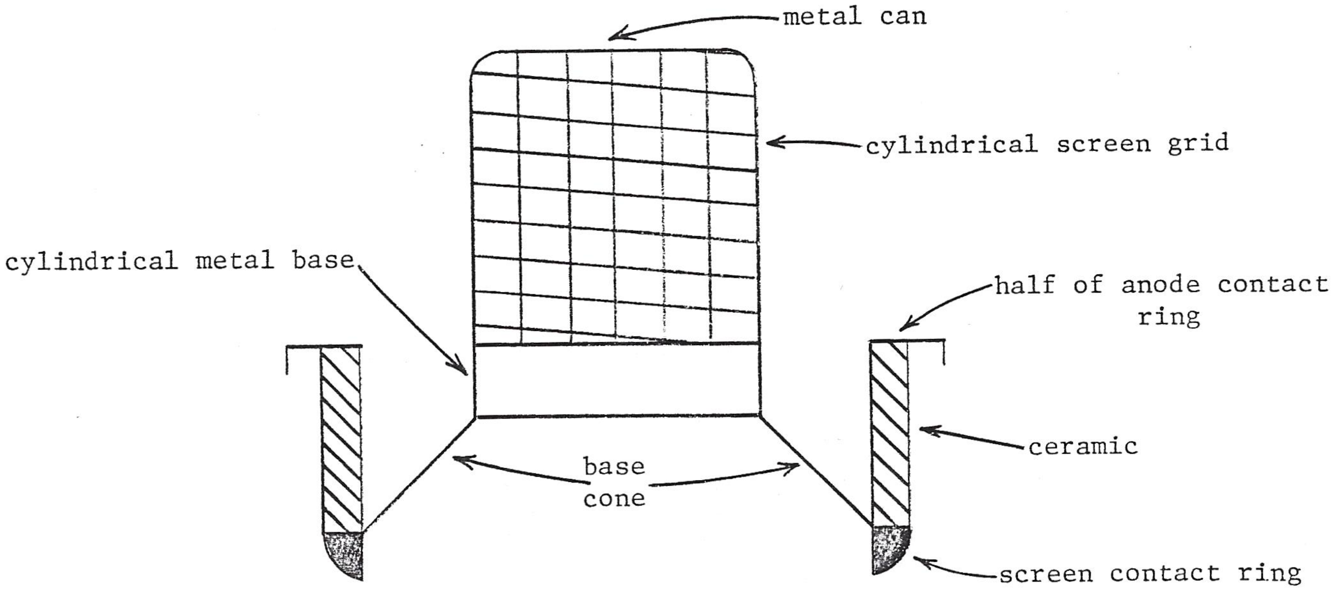


FIGURE 16 - The screen grid assembly

A cut away view of the plate circuit and the screen circuit in Figure 17 shows a concentric construction that resembles a coaxial transmission line.

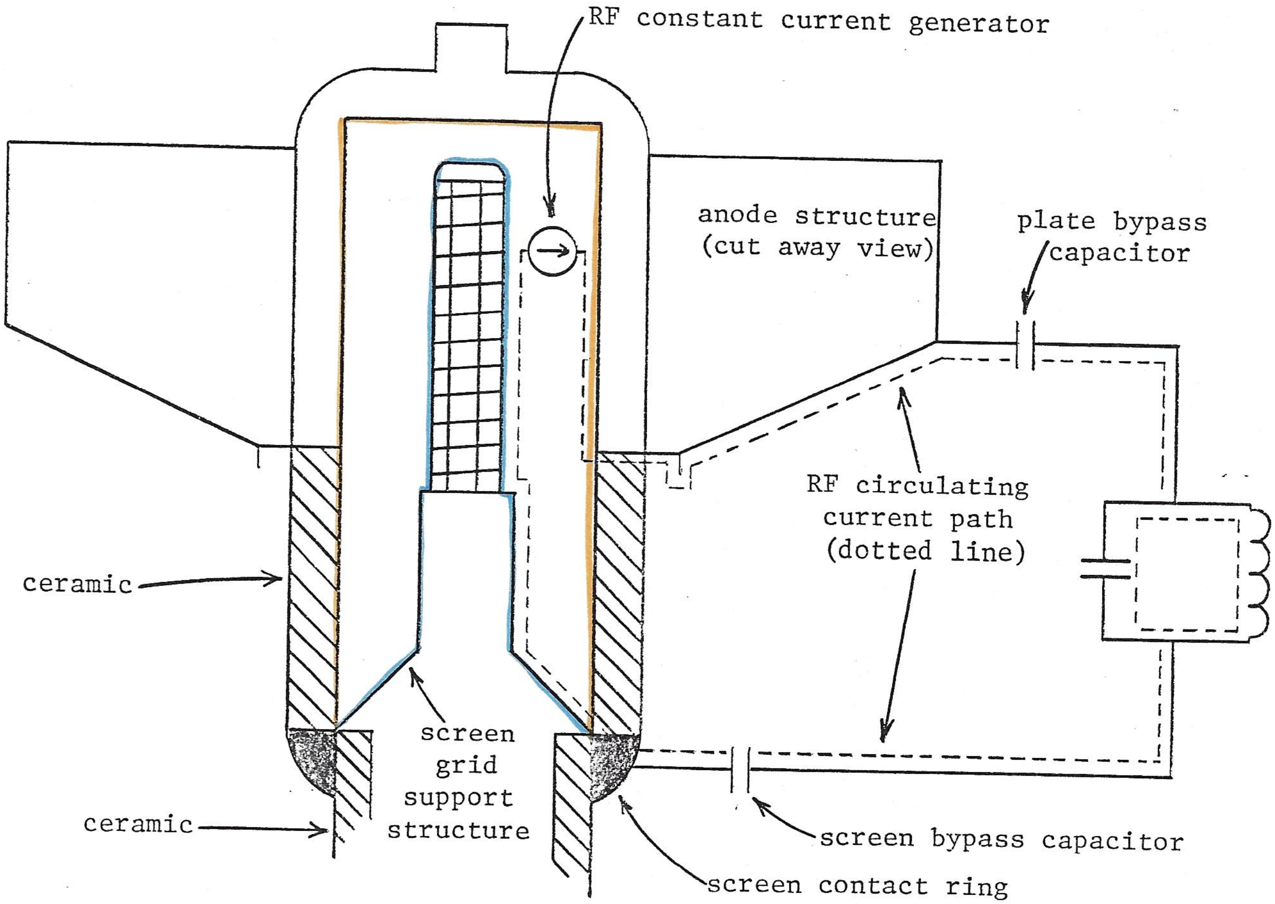


FIGURE 17 - Showing the plate and screen assembly and RF circulating current path (dotted line)

Consider that the output RF current is generated by an imaginary current generator between the plate and screen grid. The RF current travels along the inside of the plate structure on its surface (skin effect), through the ceramic at the bottom of the anode contact ring, around the anode contact ring, across the bottom of the fins, and to the band around the outside of the fins. From here it flows through the plate bypass capacitor to the RF tuned circuit and load, and returns to the screen grid. The return current travels through the screen bypass capacitor, then through the screen contact ring, up the cone, and up the screen grid to return to the imaginary generator. The screen grid has RF current returning to it but due to its low impedance, the screen grid is at RF ground potential. The RF current generator appears to be feeding an open ended transmission line consisting of the anode (plate) assembly, and the screen assembly. The RF voltage developed by the anode is due to the plate impedance (Z_p) presented to the anode by the resonant circuit and its load.

The control grid assembly and the cathode assembly are also cylindrically constructed and concentric. The control grid assembly is constructed similarly to the screen grid and is slightly smaller than it. Figure 18 shows the screen grid, control grid, and the cathode assemblies as they are placed in the tube.

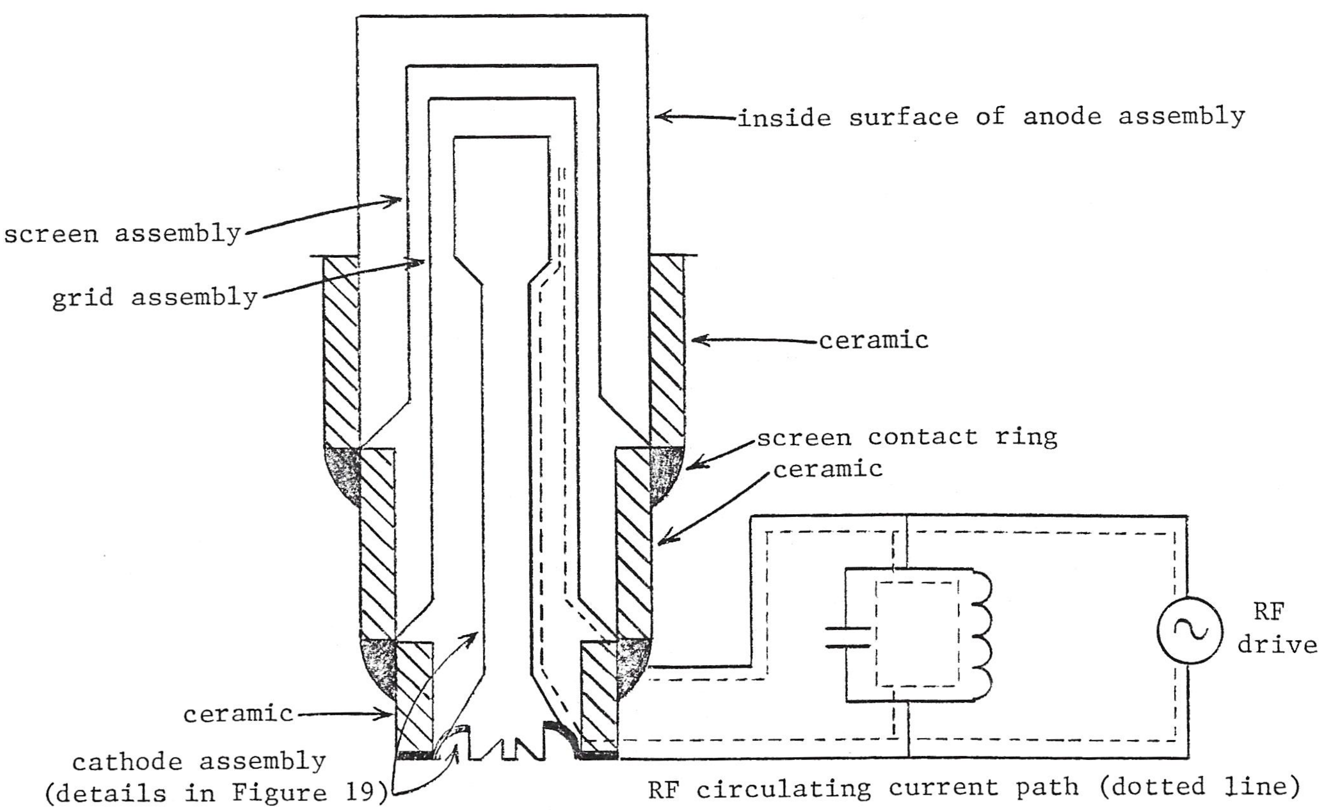


FIGURE 18 - Showing details of assembly of grids and cathode components, and the simplified RF input circuit (bias circuit not shown)

In Figure 18, an RF generator (The RF driver output) feeds a signal to the grid cathode circuit. The grid cathode assembly resembles a transmission line whose termination is the RF resistance of the electron stream within the tube.

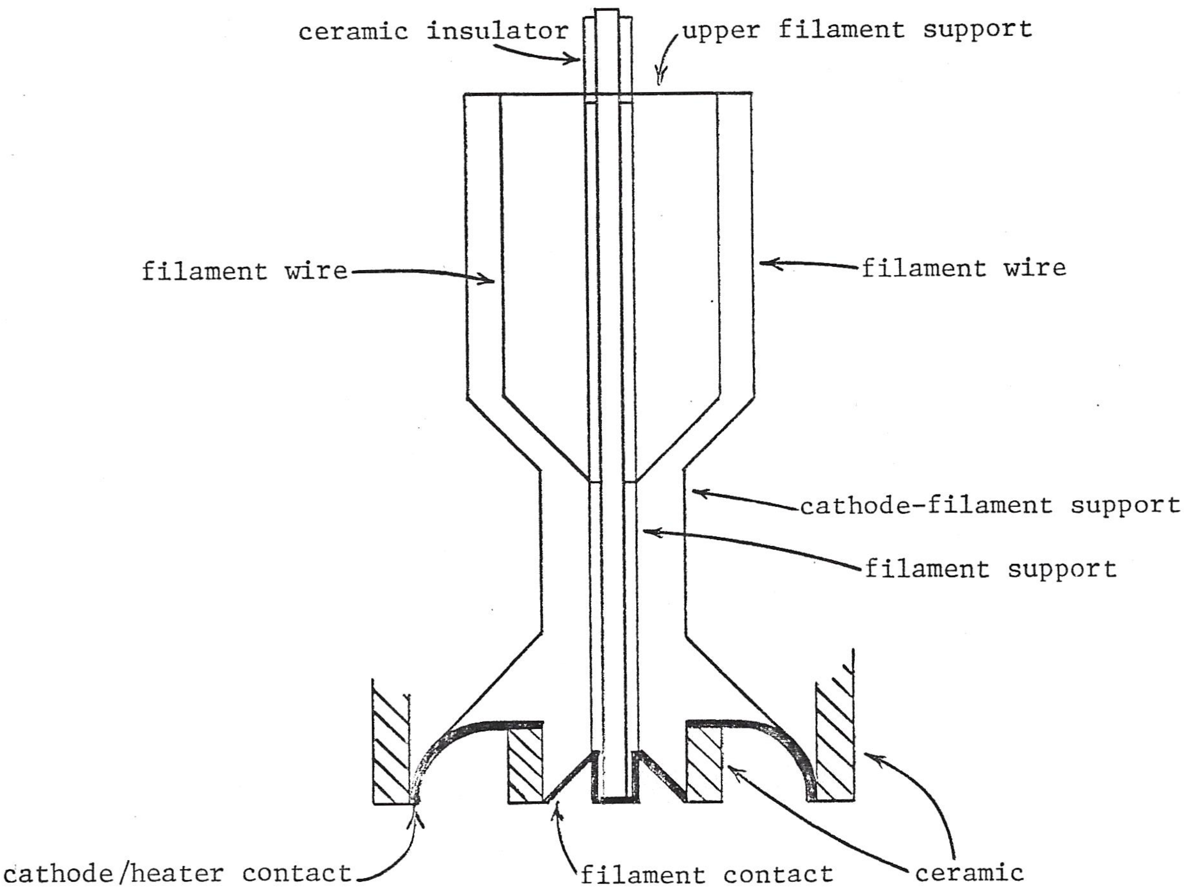
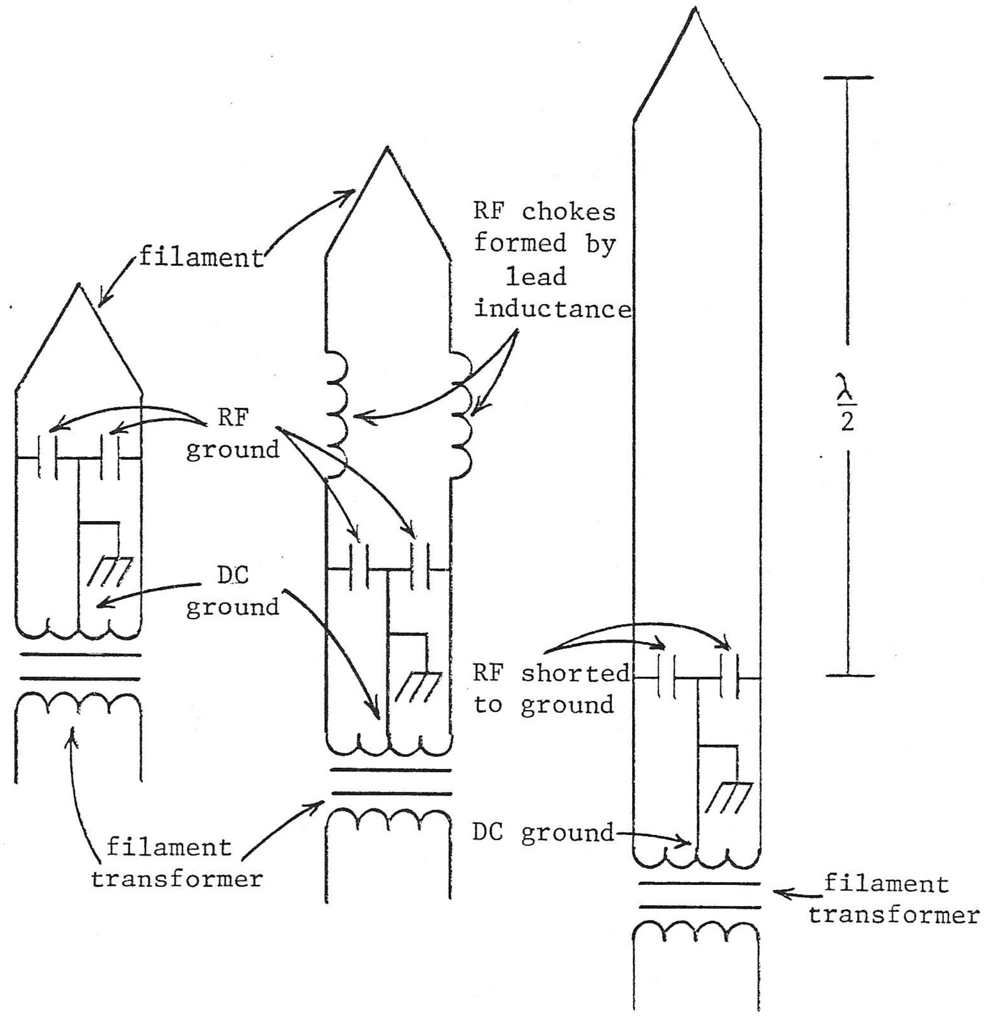


FIGURE 19 - The cathode assembly. The filament assembly consists of many parallel loops of wire supported at the top. Both sides of the filament supply feed the bottom of each loop.

The details of the cathode assembly are shown in Figure 19. The outer ring (the cathode heater contact) is the inner conductor of a coaxial transmission line formed by the cathode and control grid assemblies. The other side of the filaments is returned down the center of the cathode assembly.



A. Grounding the cathode below 30MHz

B. Grounding the cathode above 30MHz

C. Grounding the cathode via a λ/2 transmission line

FIGURE 20 - Bypassing the cathode

When operating an amplifier stage grounded cathode, feeding the RF into the grid, RF current flows into the cathode and grid circuits, but the cathode must have a low impedance (and thus low RF voltage).

Below 30 MH_z (Figure 20-A), the cathode can be grounded by simply bypassing the filament connections with capacitors.

Above 30 MH_z, the same technique does not work well because of the stray inductance of the filament leads. Notice in Figure 20-B the filament leads appear as RF chokes preventing the cathode from being placed at RF ground potential. This causes negative feedback and effects the efficiency of the input and output circuits.

In Figure 20-C, the cathode circuit is incorporated into a half wavelength transmission line. The line is shorted to ground by large values of capacitance one-half wavelength from the center of the filament (at the filament voltage feed point). This short is repeated one-half wavelength away at the cathode (heater assembly) and effectively places it at ground potential.

Since half wavelength bypassing is bulky and expensive, the selection of proper values of inductance and capacitance in the filament/cathode circuit can create an artifical transmission line that simulates the one-half wavelength shorted line shown in Figure 20-C. If Figure 20-B is observed, the inductance and capacitance can resemble an artifical transmission line of one-half wavelength, if the values of L and C are properly selected.

If this still does not make sense to you, remember that a half wavelength shorted transmission line appear to be a series resonant circuit. Now if proper values of inductance and capacitance are selected, does each side of Figure 20-B resembles a series resonant circuit?

REMEMBER: A series resonant circuit offers minimum impedance at resonances.

If you have a VHF tube type amplifier whose grid/cathode circuit is not the concentric transmission line type, you can remember selecting various lengths, widths (sizes) and numbers of conductors (inductors in this case) connecting the cathode to ground. You have bent, shaped, and changed those conductors (fine tuning values of L and C) until the amplifier has the proper efficiency of operation, grid current, RF input driver, and etc. You were really resonating the cathode circuit to place it at RF ground. To resonate the grid circuit much of these same types of bending, shaping, changing, and resonating adjustments took place. This is done to make the grid circuit operate at a high RF potential.

Most of these adjustment on your transmitter were performed by Factory Test and/or Field Service personnel. They need not be redone unless the tube manufacturer changes the tube internally (we'll notify you if this happens), or if you must change the transmitter's operating frequency. If the operating frequency must be changed, you should notify us and get Field Service assistance to perform this change.

THE HALF WAVELENGTH CAVITY (FM 2.5K THROUGH FM 20K)

½R

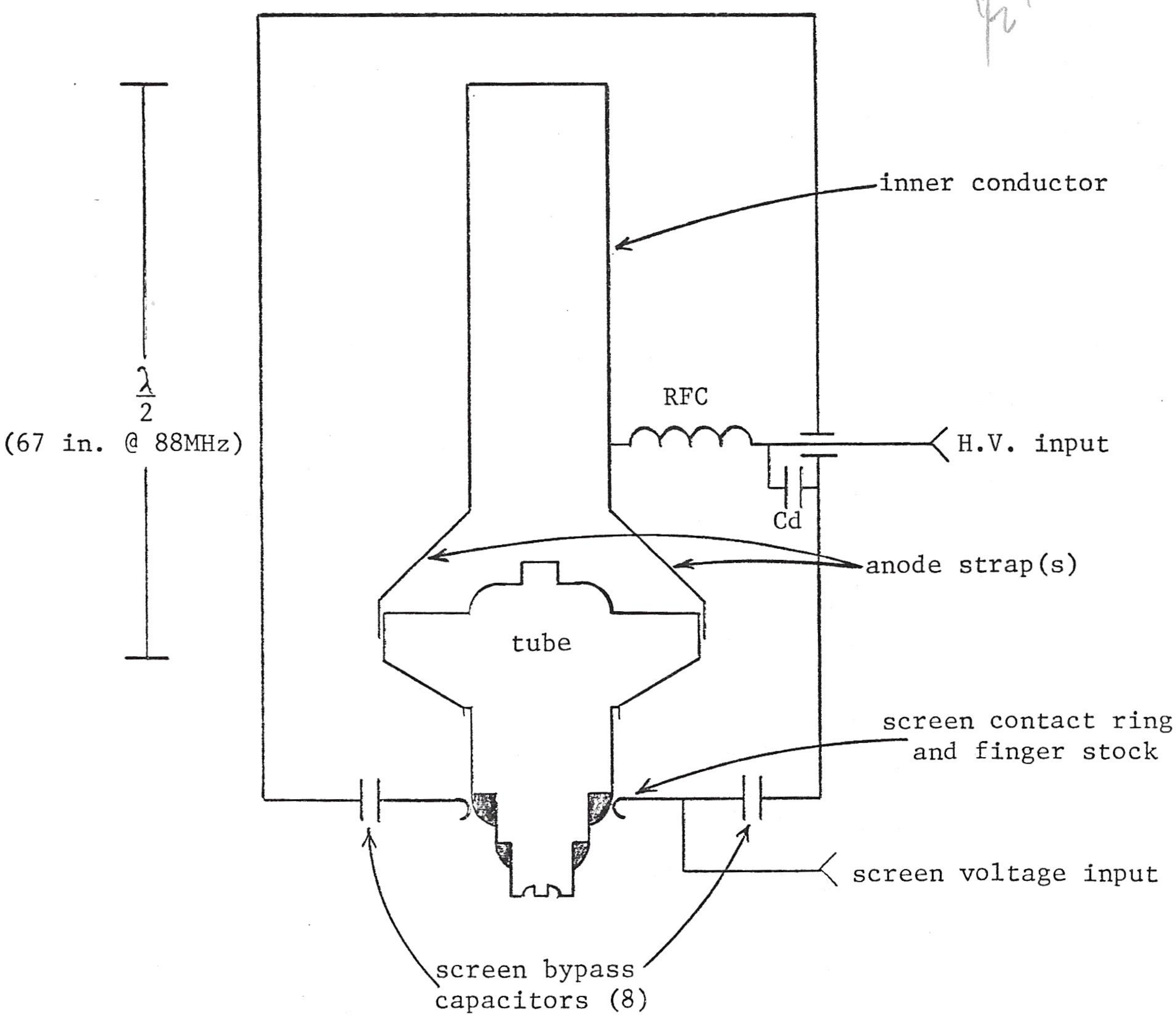


FIGURE 21 - A half wavelength cavity

This cavity (Figure 21) appears as a half wavelength circuit with the tube's anode and a silver plated brass pipe serving as the inner conductor, and the cavity box serving as the outer conductor. The transmission line is open at the far end and repeats this open condition at the tube. Remember that one half wavelength of transmission line repeats impedance. The line appears to be a parallel resonant circuit.

The circuit shown above was calculated for 88 MH_z. The inner conductor is 67 inches high. To allow room for the open condition at the top and the space for the input circuitry at the bottom, the cavity box would have to be almost eight feet tall. This size is too large for a practical transmitter and does not take tuning or operation at any other frequency between 88 MHz and 108 MHz into account.

If a graph of RF voltage and current and impedance were drawn for the inner conductor of the transmission line and the anode of the tube (See Figure 22), the plate inpedance of the tube would be many thousands of ohms. The plate's RF current would be extremely small and its RF voltage would be very large. Arcing would become a problem and the high plate impedance would make the amplifier operate inefficiently.

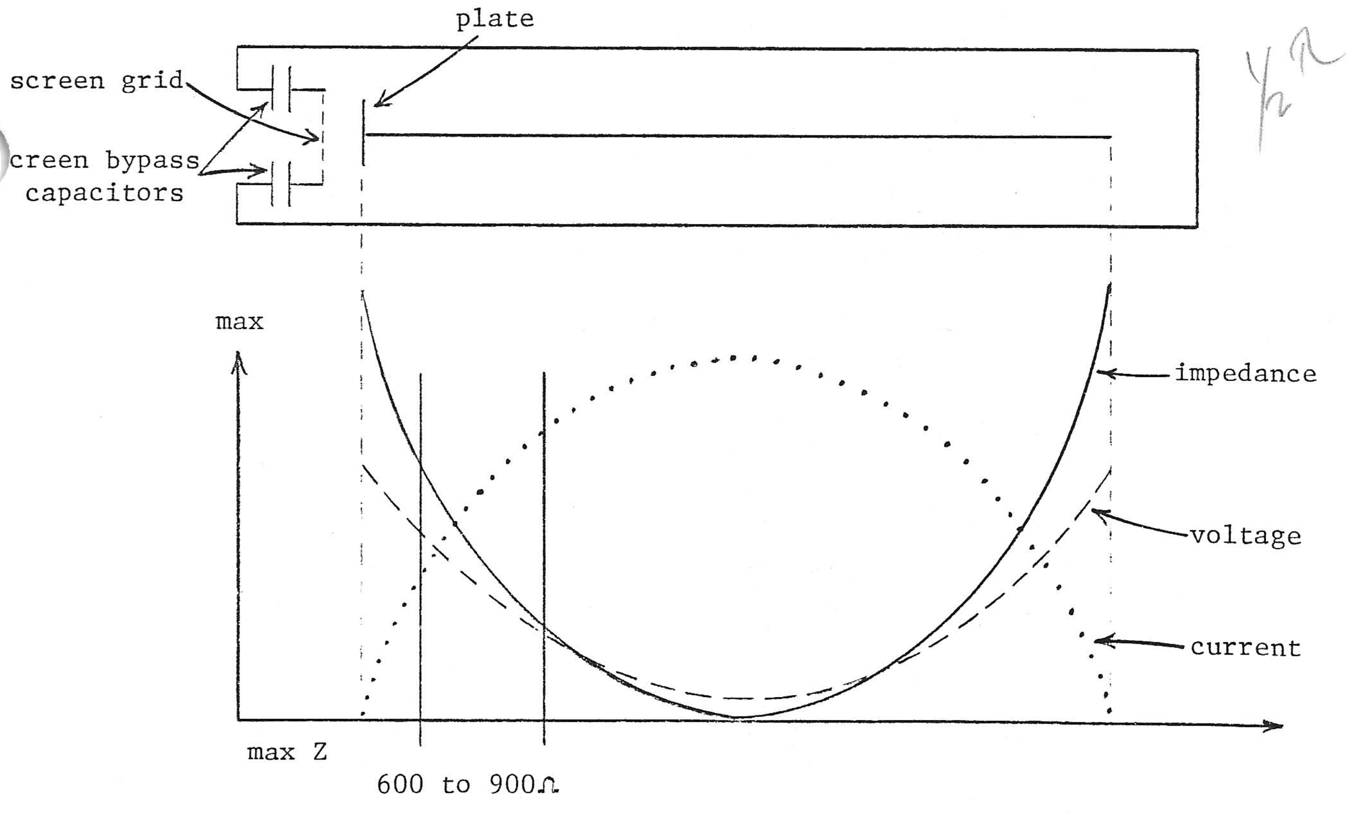


FIGURE 22 - Graph showing distribution of RF voltage, current, and impedance along the plate inner conductor.

In Figure 22 there exists an area between the anode and the quarter wavelength short location where the impedance would be ideal for the anode of the tube (typically 600 to 800 ohms). To achieve this ideal plate impedance, the inner conductor should be less than one-half wavelength in physical length (physically foreshortened) and electrically resonated (electrically lengthened to one-half wavelength).

> Some engineers prefer to talk about physically foreshortening the line so that the shorter length resonates at the desired (lower) frequency. I prefer to speak of electrically lengthening the physically short line to be electrically one-half wavelength at the desired (lower) frequency. I believe this approach helps make clear the process of resonating the line.

If the line length were changed to operate at different frequencies, the plate impedance would also change because of the new distribution of RF voltage and current on this new length of line. The problem of frequency change now becomes twofold: (1) Change the length of the line to resonate it, and (2) keep the plate impedance of the tube constant for good plate efficiency (plate efficiency explained later under electrical parameters).

To solve the problems of operation at different frequencies while keeping the plate impedance constant, two forms of coarse tuning and one fine tune (plate tune) provisions are built into the cavity.

Lets explore the actual configuration of the cavity and explain the methods used to solve the problems of physical size, operating frequency, plate impedance, and the coarse and fine tuning arrangement. Figure 23 shows the tube and its plate line (inner conductor). The inner conductor is bent into a "U" shape to reduce the cavity height.

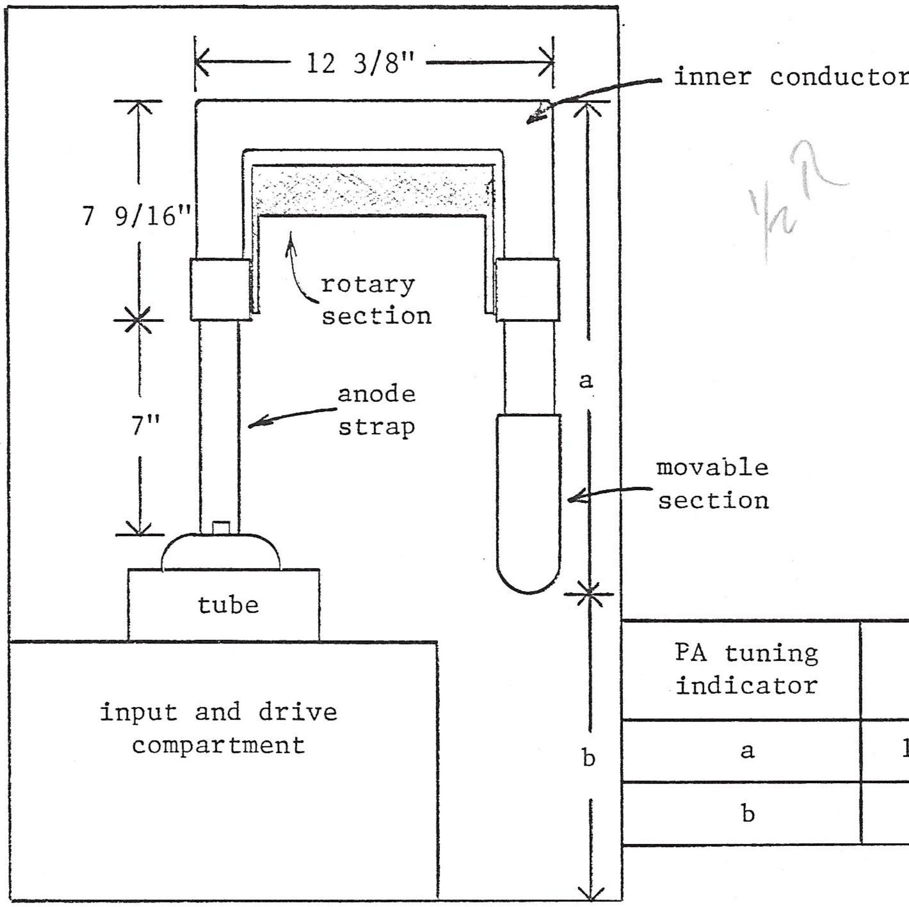


FIGURE 23 - The configuration of the halfwave cavity

PA tuning indicator	000	max
a	18 1/16	13 3/4
b	8 1/16	12 3/8

With the movable extension fully extended (plate tune indicator 000) the inner conductor measures 38 inches, and the anode strap measures 7 inches. The path from the anode strap to the inside surface of the tube's anode (along the surface due to skin effect) is estimated to be about 8 inches. This makes the inner conductor's maximum possible length about 53 inches. This is too short to be a physical half wavelength at any F.M. frequency. The length of a half-wave length line is 54.7 inches at 108 MH_z and 67.1 inches at 88 MH_z.

The two coarse tuning arrangements, the fine tuning arrangement, and the tube's output capacity resonate (electrically lengthen the physically foreshortened line) the plate line to the exact operating frequency. This process, along with proper loading, determines the proper plate impedance and therefore the efficiency.

ELECTRICALLY LENGTHENING THE PLATE LINE

THE STRAY CAPACITY

The output capacity of the tube is the first element that electrically lengthens the line. A halfwave transmission line that is too short offers a high impedance that is resistive and inductive. The tube's output capacity resonates this inductance and the detrimental effects of the output capacity are eliminated.

Another stray capacity that will electrically lengthen the line is the capacity between the movable section (the plate tune) and the cavity box (the outer conductor).

THE ANODE STRAP (COARSE TUNING PROVISION)

The anode strap has much less cross sectional area than the inner conductor of the transmission line. It therefore has more inductance than an equal length of the inner conductor. Thus the anode coupling strap acts as a series inductance and electrically lengthens the plate circuit.

At low frequencies, one narrow strap is used. This high inductance lengthens the plate circuit more. At the mid F.M. frequencies, one wider strap is used. This provides less inductance than the narrow strap and does not electrically lengthen the plate circuit as much. At the upper end of the F.M. band, two anode straps are used. This parallels the inductance of the anode straps (lower total inductance) and thus electrically lengthens the plate circuit even less.

This coarse adjustment gives three line lengths to choose from.

THE ROTARY SECTION AS A VARIABLE INDUCTOR

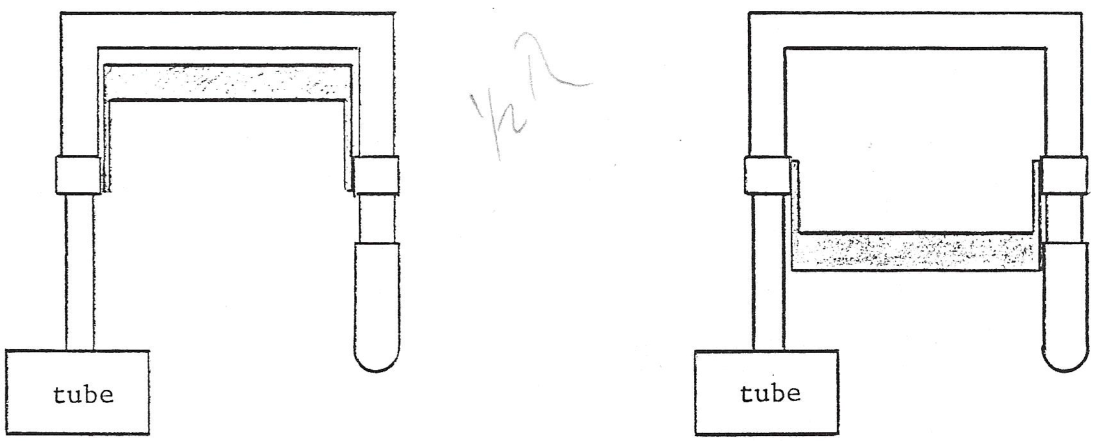
The main section and the rotary section of the plate resonant transmission line can be thought of as parallel inductance.

RF current flows in the same direction in the main transmission line and the rotary section, and thus their magnetic fields would aid.

When the rotary section is at maximum height, the magnetic coupling between the main sections of transmission line and the rotary section is at maximum. Due to the relatively large mutual inductance provided by this close coupling, the total inductance of these parallel inductors would increase. This electrically lengthens the transmission line and lowers its resonant frequency. (See Figure 24-A).

When the rotary section is at minimum height, the magnetic coupling between these two parts of the inner transmission line is minimum. This reduced coupling lowers the mutual inductance which lowers the total inductance of the parallel combination. The reduced inductance does not electrically lengthen the line as much and it therefore operates at a higher resonant frequency. (See Figure 24-B).

The rotary section provides an infinite number of coarse setting for the various operating frequencies.



(A) The rotary section at maximum height

(B) The rotary section at minimum height.

FIGURE 24 - The positions of the movable section.

THE PLATE TUNING ASSEMBLY

The movable plate tune assembly is at the end of the plate inner transmission line. It is moved up and down, changing the physical length of the inner conductor by about 4 11/16 inches. It is linked to the plate tuning knob and provides a fine tuning adjustment for the cavity.

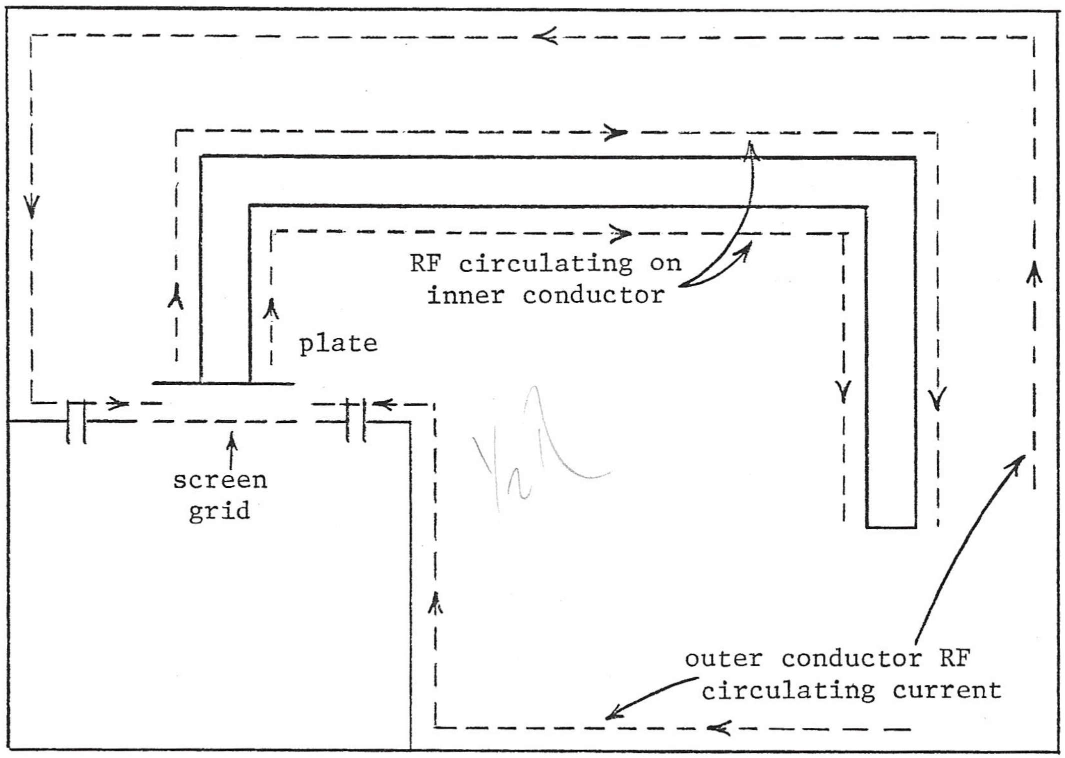


FIGURE 25 - RF circulating currents as a result of the imaginary RF current generator located between the plate and the screen grid. The direction shown for current flow is arbitrary (one-half cycle later it will reverse).

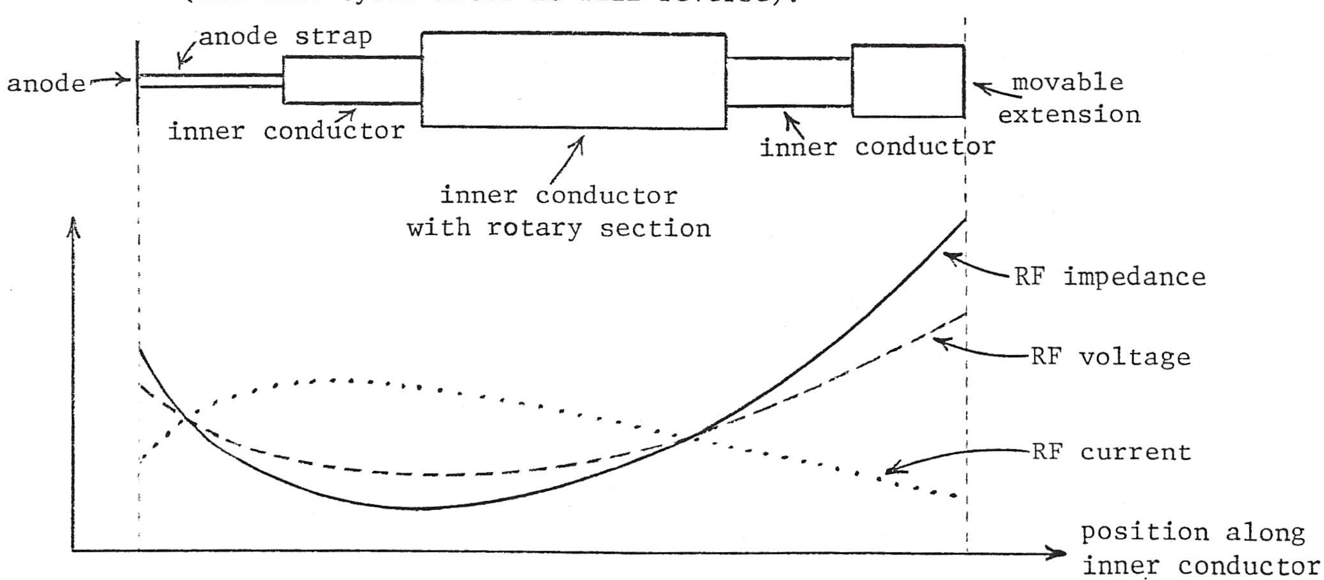


FIGURE 26 - Graph of RF current, voltage, and impedance for the cavities actual inner conductor.

THE RF CIRCULATING CURRENTS

CIRCULATING CURRENTS IN THE INNER CONDUCTOR (PLATE CIRCUIT)

Figure 25 shows the cavity RF circulating currents. The circuit impedance reflected back from the resonant circuit to the plate screen circuit of the tube is 600 to 800 ohms. RF current leaves the plate and flows down the electrical half wavelength plate inner conductor. When the current reaches the far end of the plate line it can go no further. The current flow is stopped and a high RF voltage is developed between the cavity box (outer conductor) and the end of the inner conductor. (See Figure 25). As with any other open terminated transmission line, the high RF voltage developed at the end of the line pushes (reflects) the RF current back down the line to the plate. This is the RF circulating current, the same as would be found in a conventional L-C tank circuit. If no load were placed on the resonant circuit it would have an extremely high Q. The circulating currents would gradually dampen out over several cycles if the plate-screen circuit were to receive only one pulse from the grid

NOTE: As with any other transmission line RF currents will flow in equal magnitude and opposite directions on the inner conductor and the outer conductor. We will study the RF circulating currents in the outer conductor shortly. The inner conductor is used to determine the resonant frequency and determines where the load is coupled into it.

COUPLING THE LOAD

To couple energy out of the cavity two methods can be used. They are inductive coupling and capacitive coupling.

Capacitive coupling must take place at a maximum RF voltage point, at the far end of the line in this case. Inductive coupling must take place at the maximum RF current point. It is approximately one-quarter wavelength from the end of the inner conductor. (See Figure 26). In this cavity the two coarse tuning adjustments are located just before and after the place on the line where the inductive output coupling occurs. (See Figure 27). By proper combination of these two coarse tuning controls, the maximum current point is placed exactly over the output coupling point.

At the far end of the plate line, RF current does not go quite to zero. RF voltage and impedance never get quite to maximum due to the capacity between the end of the line and the cavity box (outer conductor). This also has the effect of physically foreshortening (electrically lengthening) the line.

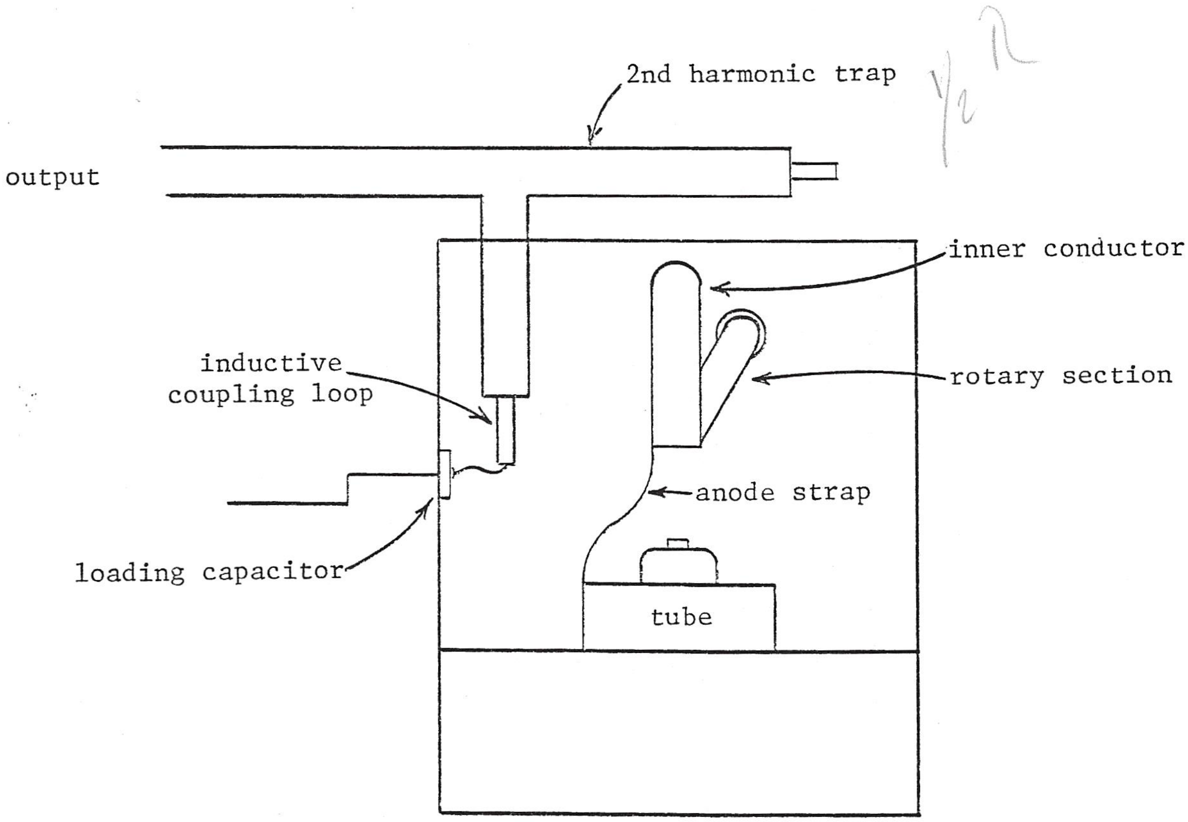


FIGURE 27 - Showing side view of cavity, and the location of the inductive coupling loop

THE CIRCULATING CURRENTS AND Q

The amount of cavity RF circulating current is directly dependent on the loaded Q of the cavity and is usually much higher than the RF output current or the RF plate current.

$$Q = \frac{Z_p}{X_L} \quad \text{and} \quad i_{circulating} = i_p \times Q$$

Where: $i_{circulating}$ = The cavity RF circulating current

i_p = The plate RF output current.

X_L = The inductive reactance required if the cavity resonant circuit had lumped constants of L and C.

This will be dicussed later under electrical parameters.

RF CIRCULATING CURRENTS IN THE OUTER CONDUCTOR

When current flows on one conductor of a transmission line an equal magnitude and current flows in the opposite direction on the other conductor. This means that a large value of RF circulating current is flowing in the cavity amplifier's outer conductor (the cavity box). All of the outer conductor's circulating currents start out at and return to the screen grid.

The back access panel (door) of the cavity is part of the cavity outer conductor and large values of circulating current flow through it, into it, and out of it. The amplifier must never be run with the back panel removed or any of the fasteners loose or damaged.

The mesh contact strap electrically connects the back panel to the rest of the cavity. If a fastener is loose or damaged, or the back panel is loose, or the mesh contact strap is damaged or defective, arcs will develope between the cavity box and that area of the back panel. Once an arc forms, the arced, pitted surface forms an insulator to the flow of RF currents. The arced surface can be cleaned but the surface must be flat to insure a good electrical contact. Any pit mark left by or under the mesh will cause a reoccurance of the arc.

THE SCREEN GRID'S RF PATH

The screen grid is connected to the screen contact ring on the tube base by a cone constructed inside the tube. The purpose of the cone is to greatly reduce the stray inductance and lower the RF resistance caused by skin effect. To take advantage of these parameters, the RF currents should flow evenly up all parts of the screen grid assembly. In Figure 28, one screen bypass capacitor is shown. This would cause all of the RF circulating current to flow at one point of the screen assembly and upset the field, increase skin effect losses, and value the apparent stray inductance would appear greater.

If the number of screen bypass capacitors were increased to two, the RF current distribution would improve. If eight bypass capacitors are used, see Figure 29, the RF current is evenly distributed throughout the screen assembly, less skin effect losses and lower stray inductance would result.

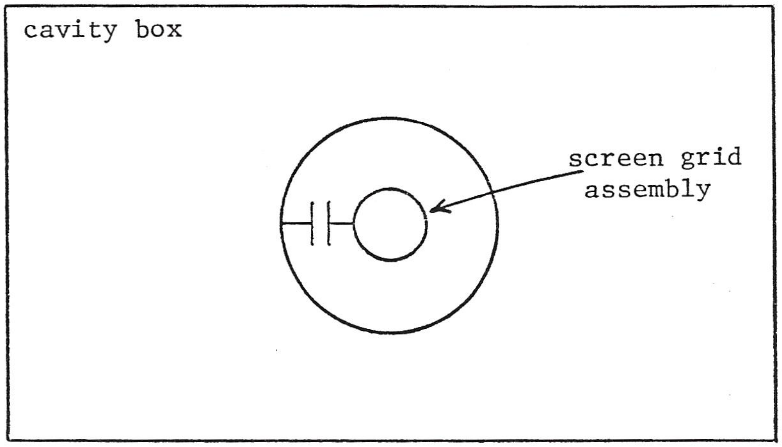


FIGURE 28 - If one screen bypass capcitor is used, an uneven distribution of RF current will result.

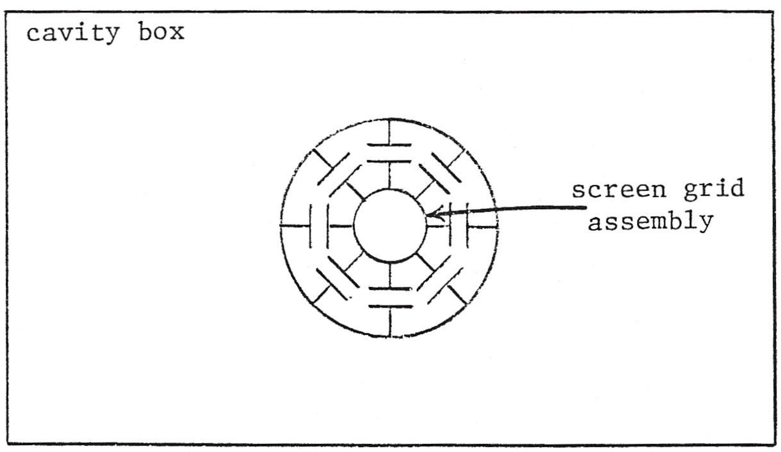
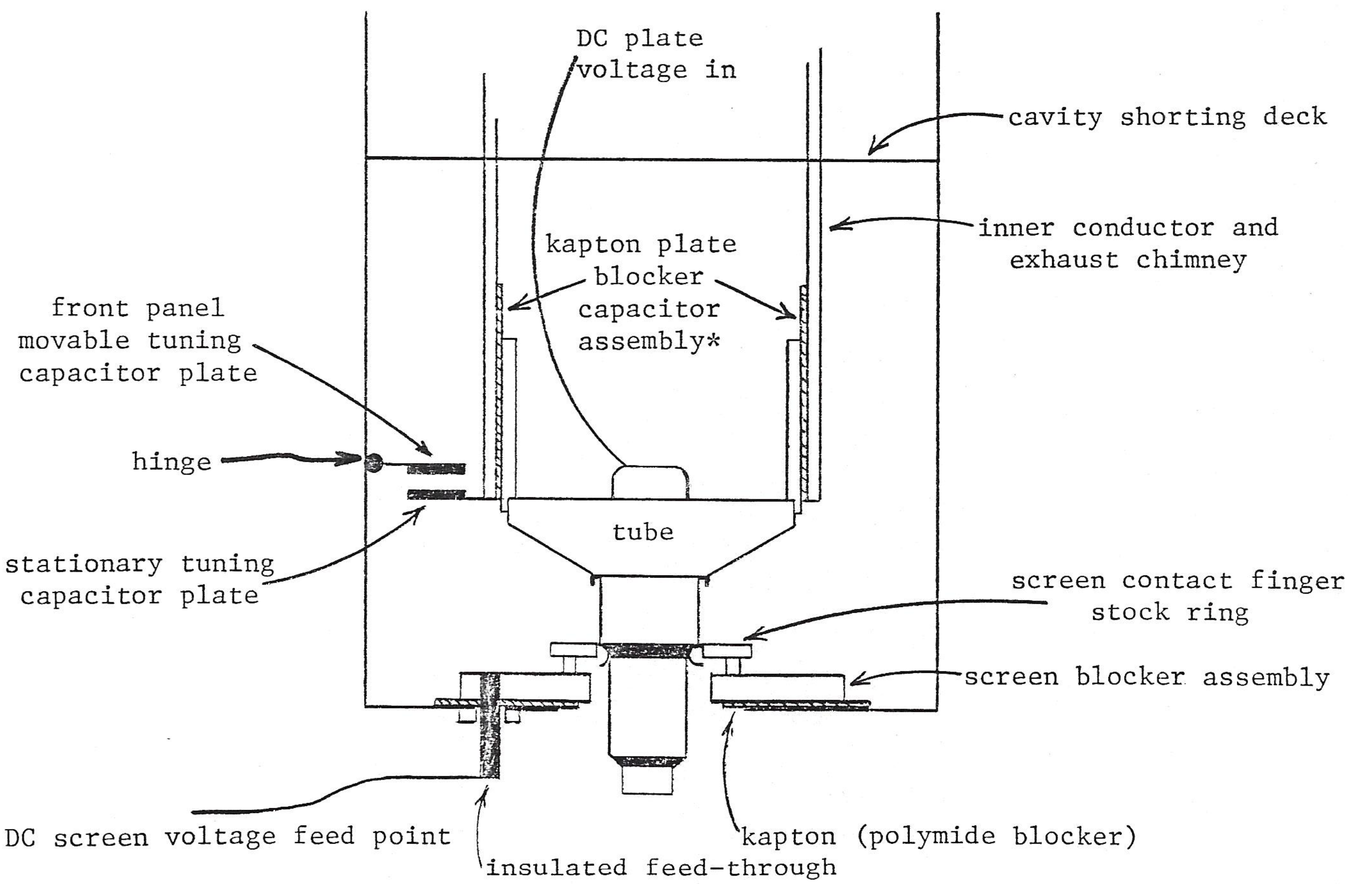


FIGURE 29 - Eight evenly spaced bypass capacitors cause the RF circulating current to divide evenly around the screen assembly.

THE F.M. 25 K QUARTER WAVELENGTH P.A. CAVITY

The quarter wavelength cavity is much simpler than the half wavelength cavity just discussed. It has one coarse tune and one fine tune control. Since the quarter wavelength cavity is shorted at the far end, a plate blocker capacitor must be used to isolate the D.C. plate voltage from ground.



*parts of plate blocker
1. inner tube
2. polymide film(kapton) 0.005" X 8" X 11ft 5 wraps
3. outer anode blocker shell

FIGURE 30 - The FM 25K Quarter Wavelength Cavity

Figure 30 shows a drawing of the quarter wavelength F.M. cavity. The plate of the tube connects directly to the inner half of the exhaust chimney (the inner tube of the plate blocker). The other part of the chimney (the outer anode blocker shell) is connected to the top of the cavity. The D.C. plate voltage is present on the inner tube of the chimney and is isolated from the grounded outer shell of the chimney by the plate blocker capacitor. The plate blocker is formed by wrapping the outside surface of the inner tube of the chimney with five wraps of eight inch wide 0.005 inch thick polymide (kapton) film. The screen contact fingerstock ring is mounted on a metal plate (the screen blocker assembly) which is isolated from the grounded cavity deck by a kapton (polymide film) blocker. The D.C. screen voltage is fed from underneath the cavity deck through an insulated feed through arrangement with one of the corner mounting screws.

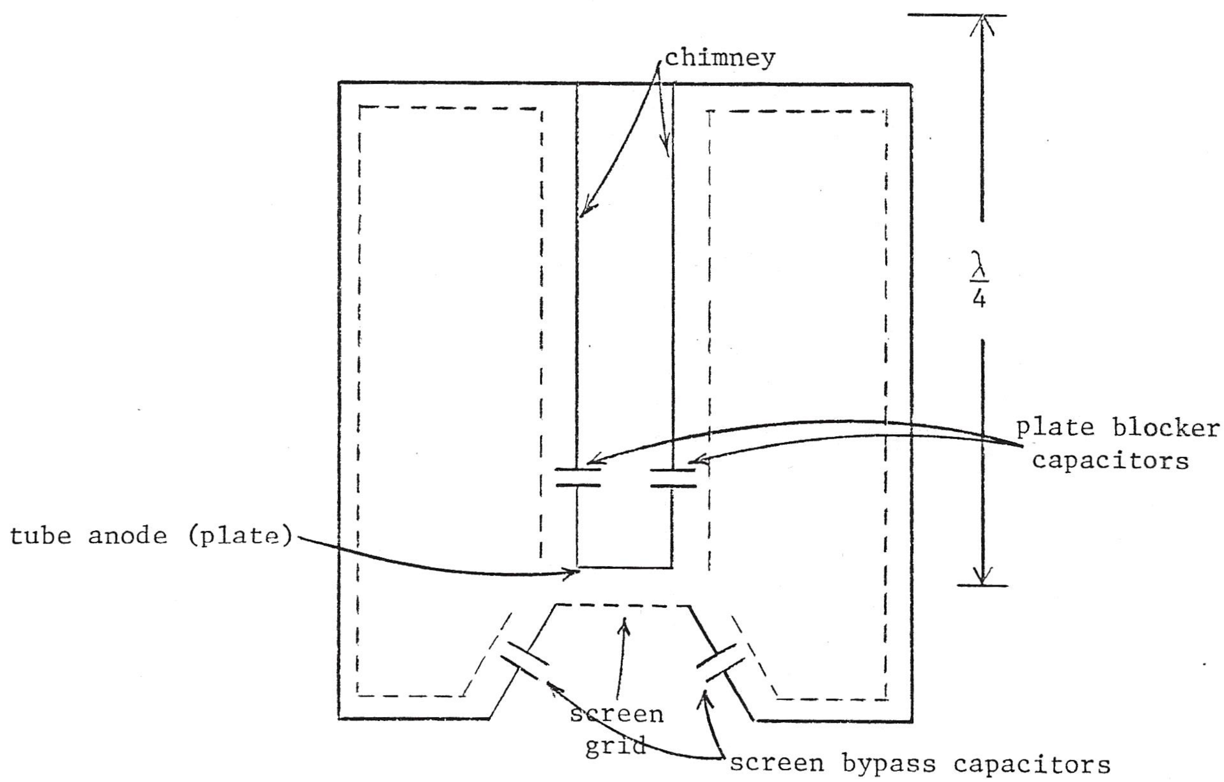


FIGURE 31 - The path for RF circulating currents in the quarter wavelength cavity.

The cavity is slightly shorter than a quarter wavelength. This makes the load inductive and it resonates the tube's output capacity. Thus, the physically foreshortened shorted transmission line is resonated and electrically lengthened to one quarter wavelength.

The RF circulating current flows from the plate, through the plate blocker capacity, and along the inside surface of the cavity (skin effect). It flows up the chimney (the inner conductor), across the top of the cavity, down the inside surface of the cavity box (the outer conductor), across the cavity deck, through the screen blocker, over the screen blocker plate, over the screen contact fingerstock, and into and up the screen grid.(See Figure 31)

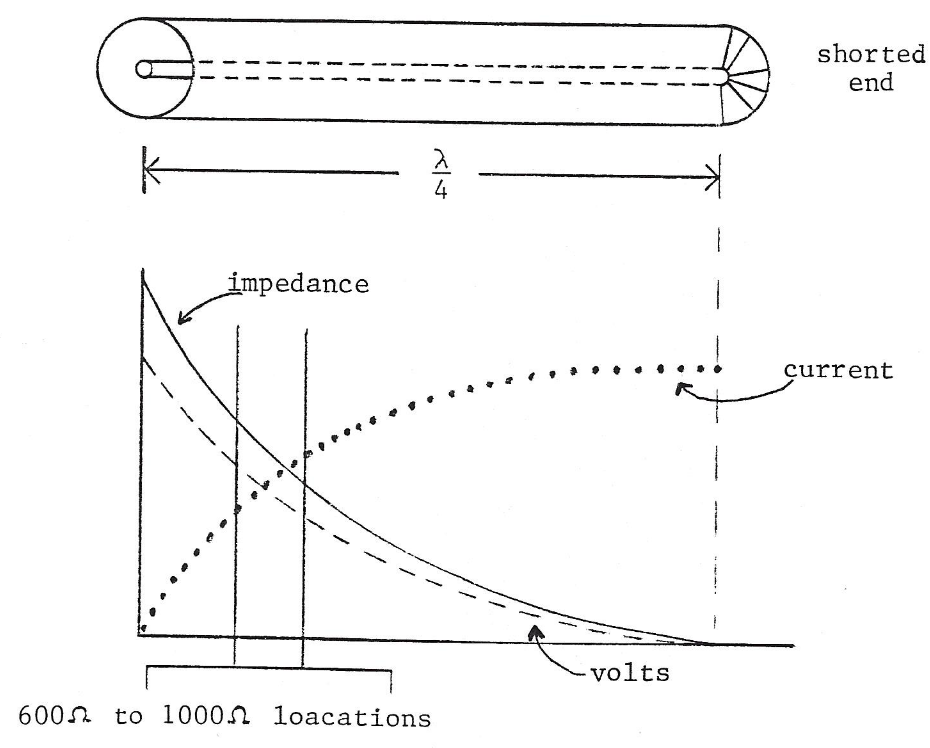


FIGURE 32 - Graph of RF current (.....), RF voltage (----) and RF impedance (———) for a quarter wavelength shorted transmission line. Notice that at the feed point RF current is zero, the RF voltage is maximum, and the RF impedance is infinite.

A graph of RF current, voltage, and impedance for a shorted, quarter wavelength coaxial transmission line shows infinte impedance, zero RF current, and maximum RF voltage at the feed point. This would not be suitable for a tube's plate impedance as the mismatch would cause arcing and poor efficiency. A point on the graph slightly less than $\frac{\lambda}{4}$ is marked. This length yields an impedance of 600 to 800 ohms and would be ideal for the plate. (See Figure 32)

The output capacity of the tube shunts the transmission line that forms the cavity and electrically lengthens it. It is now necessary to physically foreshorten the shorted coaxial transmission line (the cavity) to slightly less than $\frac{\lambda}{4}$. This shorter length is the required length from Figure 32 that will present the required plate impedance.

Figure 33 shows a graph of the RF current, voltage, and impedance presented to the plate of the tube as a result of the physically foreshortened line. This plate impedance now appears to be closer to the ideal 600 to 800 ohms required by the tube's anode.

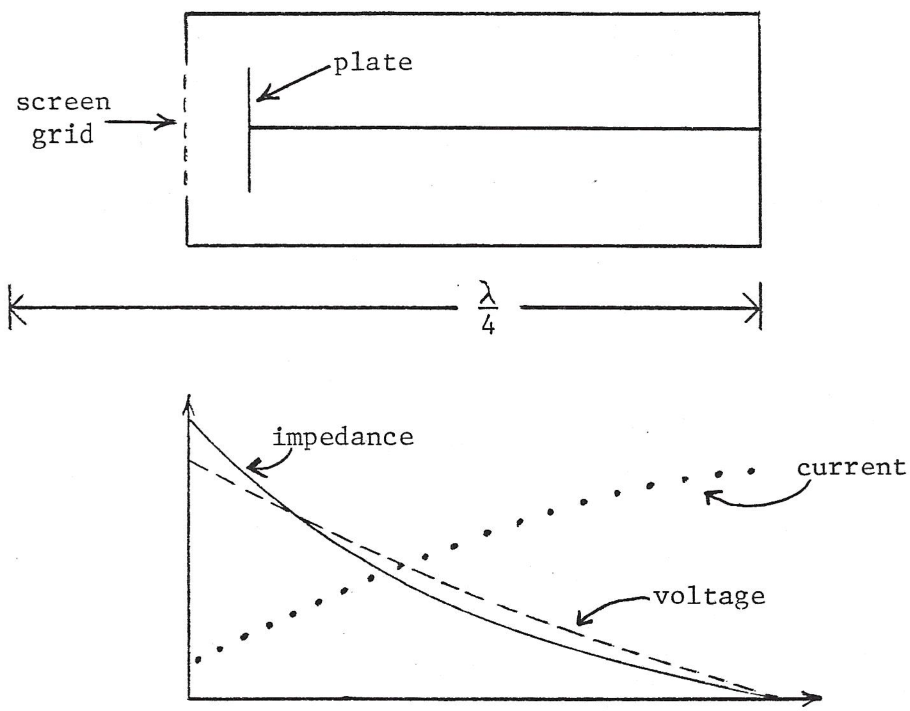


FIGURE 33 - Graph of RF current (......), RF voltage (------) and impedance (———) produced by the physically fore-shortened coaxial transmission line cavity.

COARSE TUNING

The cavity coarse tuning is accomplished by adjusting the cavity length. The top of the cavity (the cavity shorting deck) is fastened by screws and can be raised or lowered to set the length of the cavity for the operating frequency.

FINE TUNING (PLATE TUNE)

The cavity fine tuning is accomplished by the variable capacity that is built into the cavity. One plate of this capacity (the stationary plate) fastens to the inner conductor just above the plate blocker. The front panel movable tuning plate is fastened to the cavity box (outer conductor) and is linked to the plate tuning control. This capacity shunts the inner conductor to the outer conductor and can vary the electrical length and the resonant frequency of the cavity.

THE SCREEN BLOCKER ASSEMBLY

The screen blocker assembly (bypass capacitor) is formed by a metal plate, the deck of the cavity, and a kapton (polymide film) insulating sheet. The RF circulating currents that enter and leave the screen grid follow the surface of the plate, and pass through the kapton blocker to the cavity tube socket deck at the edge of the plate.

THE CAVITY ACCESS DOOR

The cavity access door is part of the outer conductor of the coaxial transmission line. Large values of RF circulating current flow along the inner surface of the door, so it must be fastened securely to prevent arcing.

INDUCTIVE OUTPUT COUPLING

The output coupling circuit is the same for the quarter wavelength and half wavelength cavities just discussed. Both are inductively coupled to the output. In both cavities, the coupling is on the side opposite the cavity access door. The inductive pickup for the half wavelength cavity is a short length of transmission line inner conductor that is terminated by the loading capacitor (See Figure 27). For the quarter wavelength cavity, the inductive pickup loop is a half loop of flat copper bar stock that terminates in the loading capacitor at one end and feeds the output transmission line inner conductor at the other end. In both cavities, the inductive pickup is positioned at a maximum current point in the cavity. They are coupled lightly so that changes in the loading will have minimum effects on the plate tuning.

Adjustment of the loading capacitor matches the 50 ohm transmission line impedance to the impedance of the cavity. Heavy loading, clockwise rotation of the loading control (minimum capacity) lowers the plate impedance presented to the tube by the cavity. Light loading reflects a much higher load impedance to the amplifier's plate.

THE SECOND HARMONIC TRAP

Both cavities have the same type of second harmonic trap in their output transmission line (See Figure 34). The trap is connected to the output transmission line by a tee. It is $\frac{\lambda}{2}$ at the second harmonic frequency and the far end is shorted. The short is repeated at the tee and effectively shorts the cavities output transmission line at the second harmonic fequencies. Any energy at this frequency leaves the cavity, travels down the transmission line to the short, and is reflected back into the cavity and cancels the second harmonic energy present at the "T" due to the 180° phase reversal incurred in the reflection at the end of the filter. It does not get to the antenna.

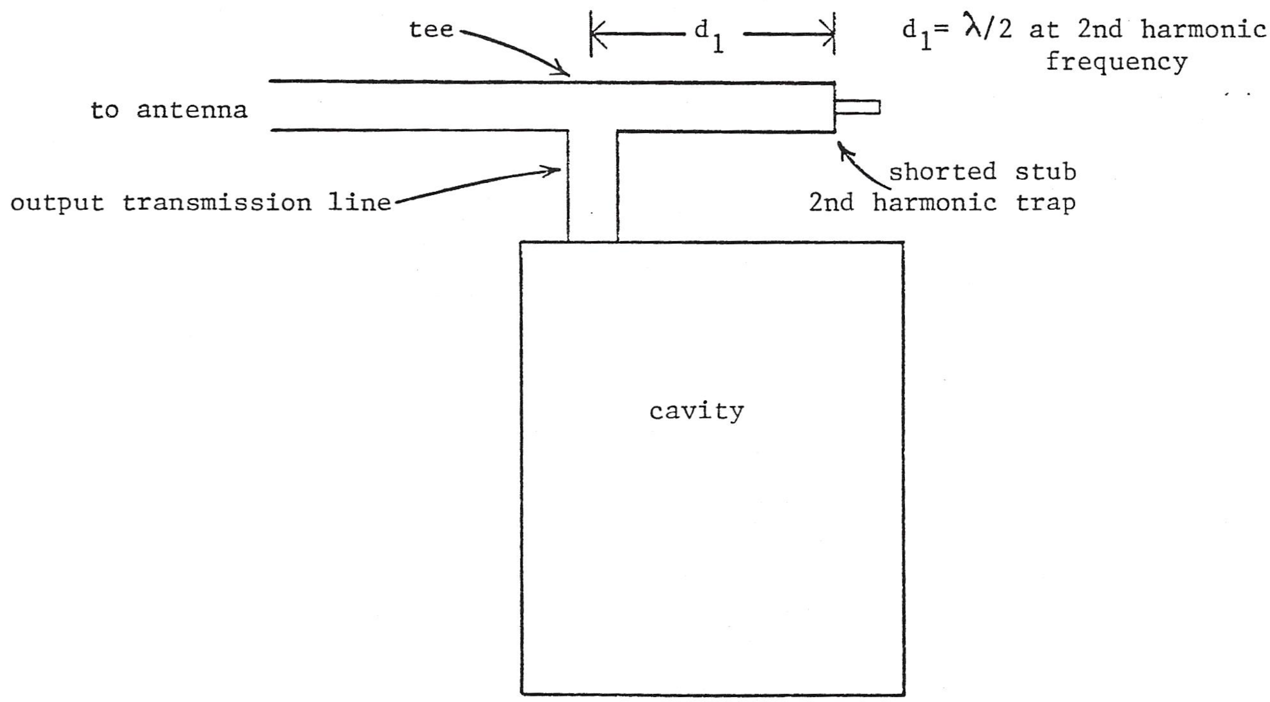


FIGURE 34 - The Second Harmonic Trap.

At the fundamental frequency of the amplifier, the trap is $\frac{\lambda}{4}$ long. The short at the end reflects as an open one-quarter wavelength away at the tee and does not interfere with the fundamental frequency energy traveling down the transmission line to the antenna.

COUPLING AND LOADING CONSIDERATIONS FOR QUARTER AND HALF WAVELENGTH TRANSMISSION LINE TYPE CAVITY AMPLIFIERS

BY: CLARENCE E. DAUGHERTY, JR.

Some additional notes on coupling will be useful. Coupling refers to the methods used to transfer the R.F. energy from the cavity that resonates the active power amplifying device (tube) to the output transmission line. Television R.F. power amplifiers use coupling to transfer energy from the primary cavity to the secondary cavity. Coupling in tube type power amplifiers usually matches a high (plate or cavity) impedance to a lower output (transmission line) impedance.

Light coupling produces light loading and results in a high plate impedance while heavy coupling results in heavier loading which lowers the plate impedance. Loading and plate impedance will be discussed in greater detail later in this report, but for now a few notes about plate impedance are in order.

Maximum output power coincident with good efficiency and acceptable dissipation dictates a definite plate impedance for a cavity of given design. This plate impedance is also dependent upon the values of DC plate voltage (Ep) and plate current (Ip).

Plate impedance dictates the cavity parameters of loaded Q, RF circulating current, and bandwidth.

1. Loaded Q is inversely proportional to the plate impedance and controls the other two cavity parameters listed.

$$\text{Loaded Q} = \frac{Zp}{X_L}$$

where X_L = cavity inductive reactance

Z_p = cavity plate impedance

2. Circulating current in the cavity is much greater (by a factor of the loaded Q) thus the RF current supplied by the tube.

$$\text{Circulating current} = Q\ (x)\ i_p$$

where: i_p = the RF current supplied to the cavity by the tube

3. The cavities bandwidth is dependent on the loaded Q and the operating frequency.

$$\text{Bandwidth} = \frac{F_r}{Q}$$

where: F_r = Cavity Resonate frequency.

Capacitive (electrostatic) and inductive (magnetic) coupling methods are used in RF amplifiers.

MAGNETIC COUPLING

Both the 1/4 and 1/2 wavelength cavities discussed use magnetic coupling.

Magnetic coupling employs the principal of transformer action and depends upon three conditions:

1. The cross sectional area under the coupling loop compared to the cross sectional area of the cavity, see Figure 1.
   - (A.) This compares to the turns ratio of the transformer.

2. The orientation of the coupling loop to the axis of the magnetic field, see Figure 2.
   - (A.) The coupling is proportional to the cosine of the angle which the coupling loop is rotated away from the axis of the magnetic field.

3. The amount of magnetic field that the coupling loop intercepts.
   - (A.) The greatest magnetic field will be found at the point of maximum RF current of the cavity. This is the place where maximum inductive coupling is attained.
   - (B.) Greater magnetic field strength is found closer to the center conductor of the cavity, thus coupling is inversely proportional to the distance of the coupling loop from the center conductor.

EXAMPLE: The effects metioned above are similar to the operation and adjustment of a variable directional coupler probe used to sample forward and reflected power in a transmission line.

THE 1/2 WAVELENGTH CAVITY

In the 1/2 wavelength cavity that was discussed previously, the coupling loop is located just above the tube, see Figure 3, and magnetically links the center conductor at that point to the coupling loop.

The two methods used to electrically lengthen the transmission line and influence the amount of coupling are:

1. The anode straps which are located before the coupling loop.

2. The rotary arm which is located after the coupling loop.

The proper adjustment of these two coarse tuning controls:

1. Places the maximum RF current point of the center conductor under the coupling loop.

2. Sets the electrical length of the plate line.

These in conjunction with the loading control, discussed later, sets the plate impedance of the P.A. tube.

THE 1/4 WAVELENGTH CAVITY

In the practical 1/4 wavelength cavity the position of the coupling loop is fixed. The short is movable as a coarse tuning adjustment.

The maximum RF current point is located at the short. If the coupling loop is located at this maximum current point, the magnetic coupling to the loop occurs from the inner conductor of the cavity and the short, see Figure 4. When the short is moved to change frequencies, the amount of the coupling also changes (due to changing distance between the short and the coupling loop). This effect is eliminated by moving the coupling loop away from the short so that the coupling occurs from the inner conductor only. This location is at a point of lower RF current, but sufficient coupling can be achieved.

If greater coupling is desired, for example, to lower the cavity Q and broaden the bandwidth, the coupling loop (as shown in Figure 5) can be made from a piece of wide, flat bar stock. This will still result in magnetic coupling, but the greater surface area will also capacitively couple energy from the cavities center conductor to the coupling loop.

By using both capacitive and inductive coupling, a large value of coupling is available at a location that is not at the maxiumum RF voltage or current point. This offers two advantages:

1. Good isolation of the amplifier from changes in the load.

2. The provision for changing the coupling, as dictated by different operating frequencies, by varying the width (capacity) of the coupling loop.

In both the 1/4 and 1/2 wavelength cavities, the coupling loop feeds a 50 ohm resistive transmission line (load). The loop is in series with the load and has considerable inductance. This inductance will reduce the RF current that flows into the load thus reducing the output. This effect is overcome by putting a variable capacitor in series with the coupling loop, see Figure 3 and 5. The load is connected to one end of the coupling loop and the variable capacitor connects the other end of the loop to ground.

The variable capacitor cancels some or all of the loops inductance. This is the amplifier's loading control.

Maximum load current and output power occurs when the loading capacitor cancels all of the inductance of the coupling loop. This lowers the plate impedance and results in heavier loading.

Light loading results if the loading capacitance does not cancel all of the coupling loop's inductance. The loop inductance that is not cancelled causes the load current and the power output to be reduced and the plate impedance to be raised.

CAPACITIVE COUPLING

Capacitive coupling, which physically appears straightforward, often baffles the technicians due to its unique characteristics.

Figure 6 shows a cavity amplifier with a capactive coupling plate positioned near its center conductor. This coupling plate is connected to the output load, which can be a transmission line or a secondary cavity. Figure 7 shows a cavity that is used as an 18 KW T.V. visual power amplifier when the secondary cavity is capactively coupled to the primary cavity, or a 10 KW aural cavity when the aural secondary capacitively couples the cavity amplifier directly to the transmission line.

Parameters that control the amount of capactive coupling are:

1. Area of the coupling capacitor plate
   (A.) Larger area = greater coupling

2. Distance of coupling plate to the center conductor.
   (A.) Greater distance = less coupling

3. Greater capactive coupling occurs when the coupling plate is at the maximum voltage point on the cavities center conductor.

To understand the effects of capactive coupling the equivalent circuit of the cavity must be observed. In the equivalent circuit shown in Figure 8, the cavity appears as a parallel resonate circuit with the tube connected in parallel with it. The plate blocker capacitor isolates the tube's DC voltage from the cavity. Physically the coupling capacitor and output load are in series (see Figure 8), but electrically they appear to be in parallel as shown in Figure 9. The attached series to parallel chart (Table 1) shows how the resistive component of the equivalent parallel circuit is increased by the coupling reactance. The equivalent parallel coupling reactance is absorbed into the parallel resonate circuit, thus explaining the necessity to retune after changing coupling (loading). The coupling (or loading) reactance can be a series capacitor or inductor.

NOTICE: In the double tuned cavity amplifier shown in Figure 7, coupling is controlled by a series capacitor (the coupling plate) and loading is controlled by a series loading inductor.

The series to parallel transformations are accomplished by the formuli listed below.

$$R_p = \frac{R_s^2 + X_s^2}{R_s} \quad \text{and} \quad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

Resistive Load Impedence	Series Coupling Capacitive or Inductive reactance	Equivalent Parallel Circuit Transformed Resistance (Plate Impedance)	Transformed Reactance
50 Ω	100 Ω	250 Ω	125 Ω
50 Ω	200 Ω	850 Ω	213 Ω
50 Ω	300 Ω	1850 Ω	308 Ω
50 Ω	400 Ω	3250 Ω	406 Ω

TABLE 1 The effects of the series coupling reactance on plate impedance.

HEAVY LOADING

Heavy loading, or large coupling, lowers the plate impedance and cavity Q.

The effects of lower plate impedance are:

1. Higher RF and DC plate currents.
   (A.) Remember: Lower Q reduces the cavities' RF circulating current.
   In some cavities high circulating currents can cause cavity heating and premature failure of plate and screen blocker capacitors.

2. Reduced RF plate voltage. This reduces the swing of DC plate voltage.
   (A.) The instantaneous plate voltage is the result of the RF plate voltage added to the DC plate voltage.

3. The reduced swing of plate voltage causes less positive DC screen current to flow.
    (A.) Positive screen current flows only when the plate voltage swings close to or below the value of positive screen grid voltage.

ANALYSIS OF HEAVY LOADING

The DC plate current increases but the DC plate voltage stays constant, therefore the DC imput power will increase.

Any change in RF output power is dependent upon the new RF plate impedance. There is one value of plate impedance that will yield optimum output power, efficiency, dissipation, and dependable operation. It is dictated by cavity design and the values of the various DC and RF voltages and currents supplied to it.

If the RF output power increases, efficiency may increase but it will 1
decrease if the RF output power stays the same or drops.

LIGHT LOADING

Light loading, or low coupling, raises the plate impedance and cavity Q.

The effects of the higher plate impedance are:

1. Reduced RF and DC plate current.
    (A.) Remember: Higher cavity Q will increase the cavity circulating currents which could damage some cavities.

2. The RF plate voltage increases, thus the DC plate voltage will swing further from its average value.
    (A.) The higher cavity RF or peak DC voltages may cause arcing.

3, The larger swing of DC plate voltage causes it to dip closer to or below the screen grid DC voltage. The result will be a greater DC screen current.

ANALYSIS OF LIGHT LOADING

The DC plate voltage is constant (set by the power supply) and the lower DC plate current reduces the tube's DC input power. The RF output power may or may not change depending on the plate impedance and cavity design while the efficiency will probably increase, or at worst remain constant. Caution must be exercised because the increased RF voltage and circulating current may cause cavity arcing, overheating, or other problems such as shorted capacitors. The manufacturers recommendation for the operation of the cavity and P.A. tube should be observed.

SCREEN GRID CURRENT

The screen grid of the P.A. tubes is operated at a positive DC potential. The instantaneous screen grid current will be positive when the plate voltage swings near to or below the DC screen grid voltage and can become negative (the screen grid emits electrons) at the most positive swing of the plate voltage. The screen current meter reads only the average (DC) screen current.

SCREEN GRID VOLTAGE

If the screen grid voltage is high greater plate current and output power can be produced for a given plate voltage, but the plate voltage will reach the DC screen voltage more quickly on its negative swing. This will cause a higher positive DC screen current to flow and necessitates operating with a lower plate impedance (heavier loading and/or more coupling).

A lower screen grid voltage allows the plate voltage to swing further but it lowers the tube gain and the amount of plate current that will flow. This requires a higher plate impedance for proper operation.

THE SCREEN GRID SUPPLY

The screen grid supply must have provisions to allow negative DC current to flow since the screen grid can operate under that condition. This can be accomplished by the bleeder resistor in the screen supply.

If the supply is well regulated the DC screen grid voltage will not drop when positive screen current flows. Two conditions result from this.

1. With light loading and proper tuning the flow of positive screen current can become quite large causing the screen grid to overdissipate and become damaged. This necessitates screen grid current metering and overload protection.

2. As the plate tuning is adjusted the plate current remains constant, due to the stiff screen grid voltage. The plate is tuned for maximum output power.

The screen grid supply can deliberately have poor regulation so that when excessive positive screen current flows the DC screen grid voltage drops. Under these conditions the screen grid can not be overdissipated, therefore the requirement for screen grid metering and overload protection is reduced.

When tuning and loading conditions cause a larger positive DC screen grid current to flow the screen voltage drops, due to the poor regulation, causing the plate current to reduce. Proper tuning is indicated by a dip in plate current and the desired output power is obtained only when proper tuning and the correct loading is achieved.

CONCLUSION

The mystery of cavity amplifiers vanishes when one discovers that all that is required to understand them are the basic principles of AC and DC theory, amplification , active devices, transmission line theory, and math that can be preformed on a simple calculator. The only mystery is the lack of material that explains to the technicians how all the above principles combine. I have tried to accomplish this in these papers.

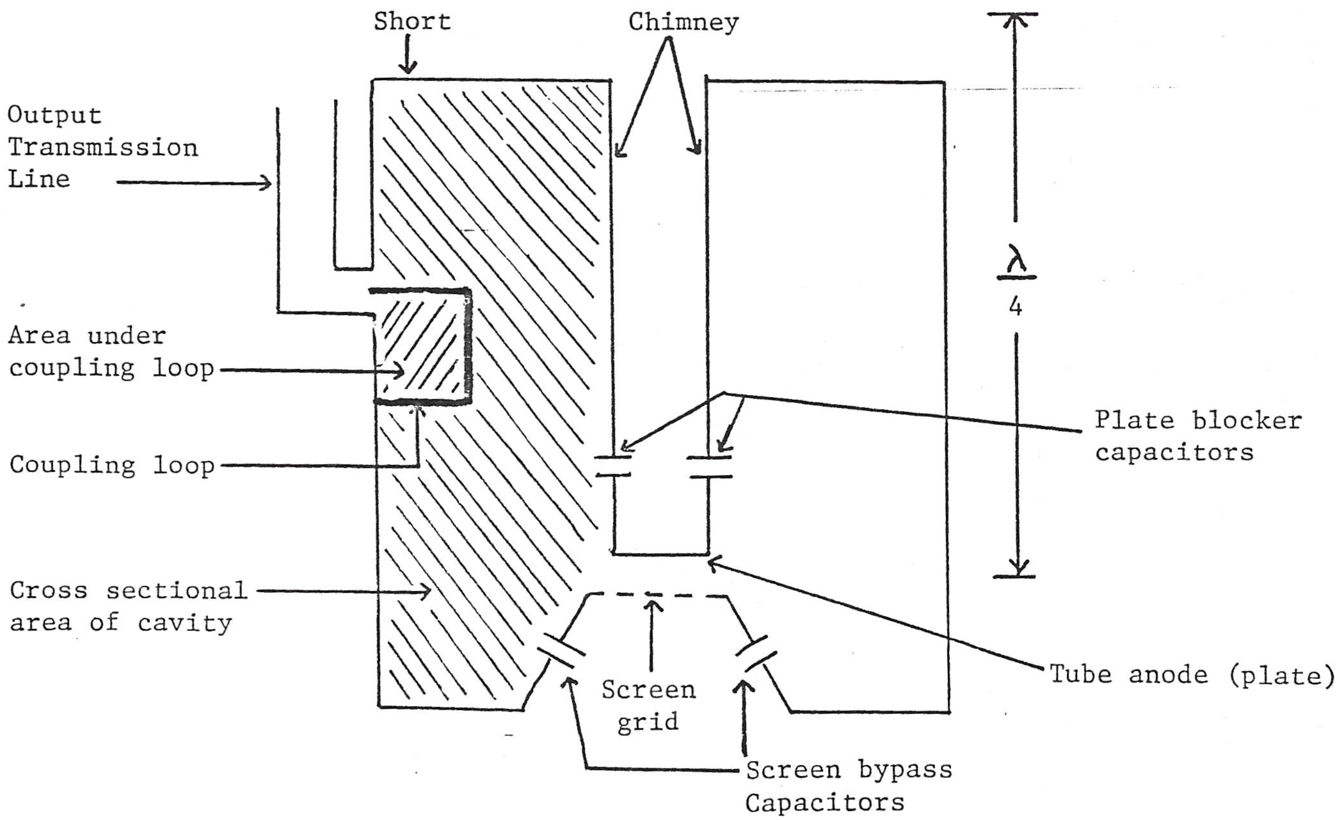


FIGURE 1 Shows one of the factors that determines the amount of coupling is the area under the coupling loop compared to the cross sectional area of the cavity.

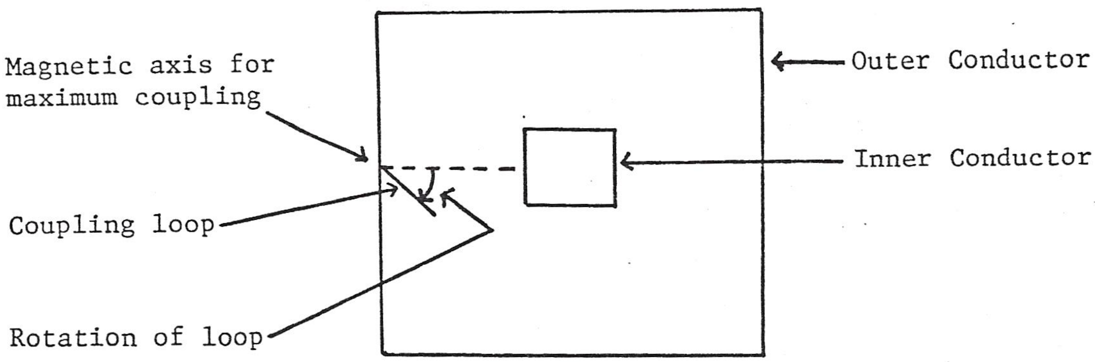


FIGURE 2A Top view of a cavity showing the coupling loop rotated away from the axis of the cavities magnetic field.

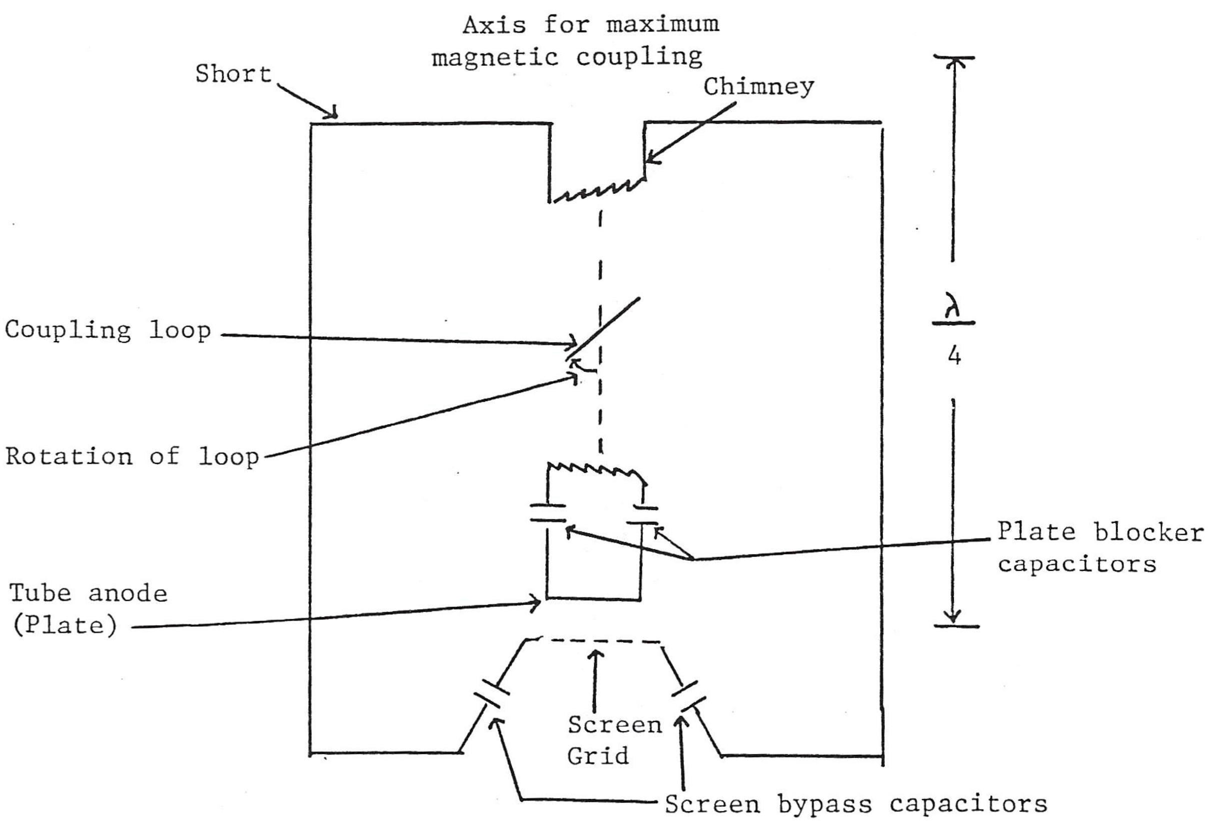


FIGURE 2B Cutaway view showing the coupling loop rotated away from the axis of the cavities magnetic field.

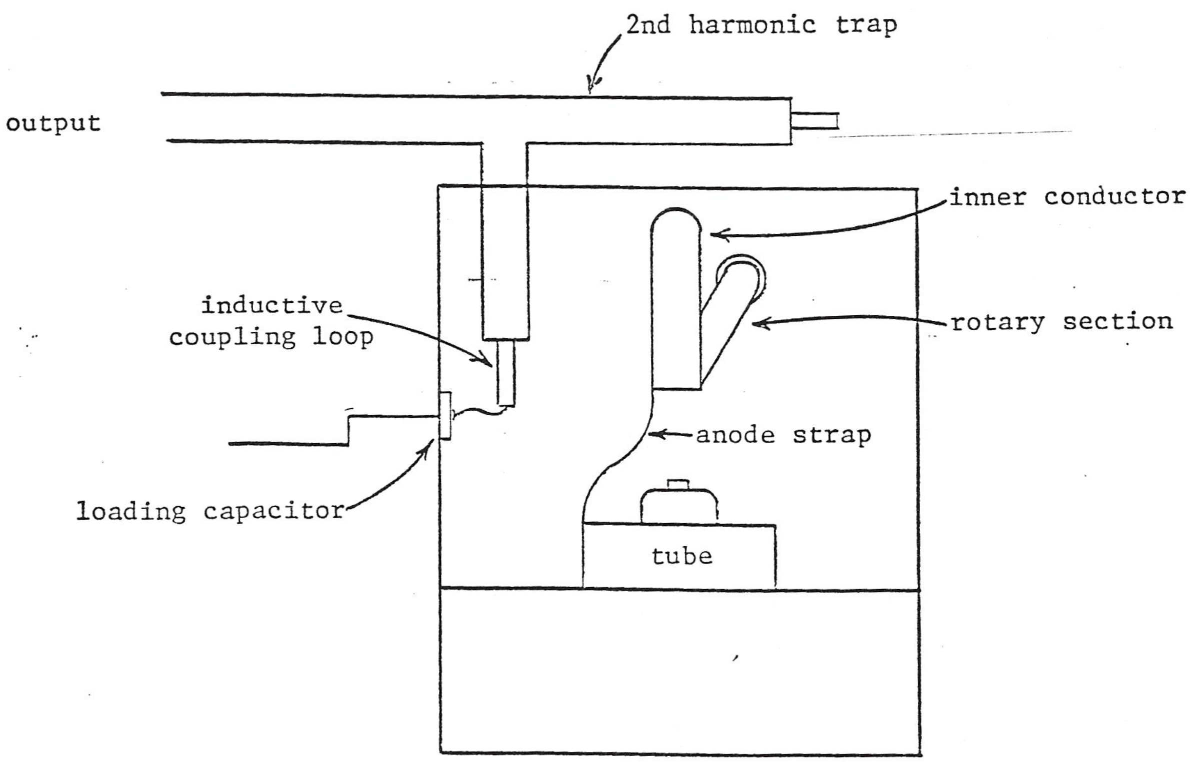


FIGURE 3 - Showing side view of cavity, and the location of the inductive coupling loop

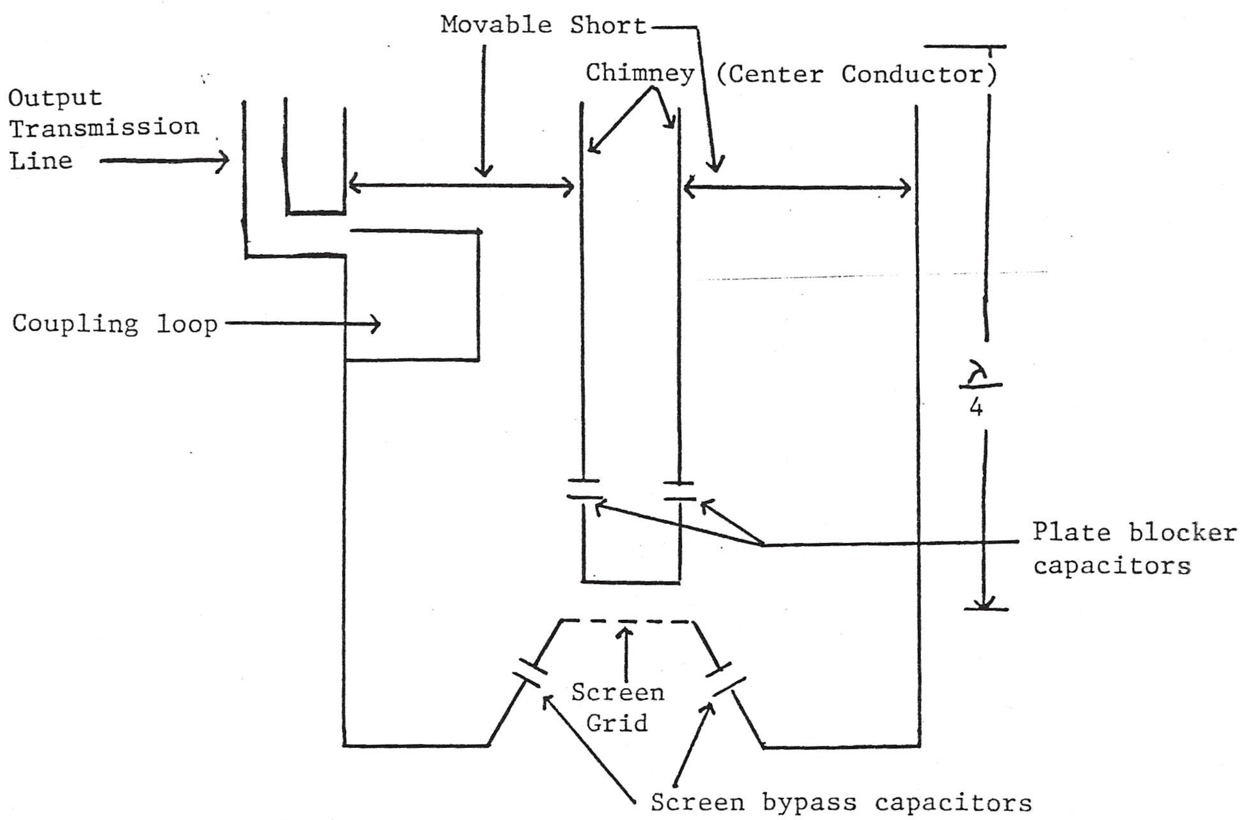
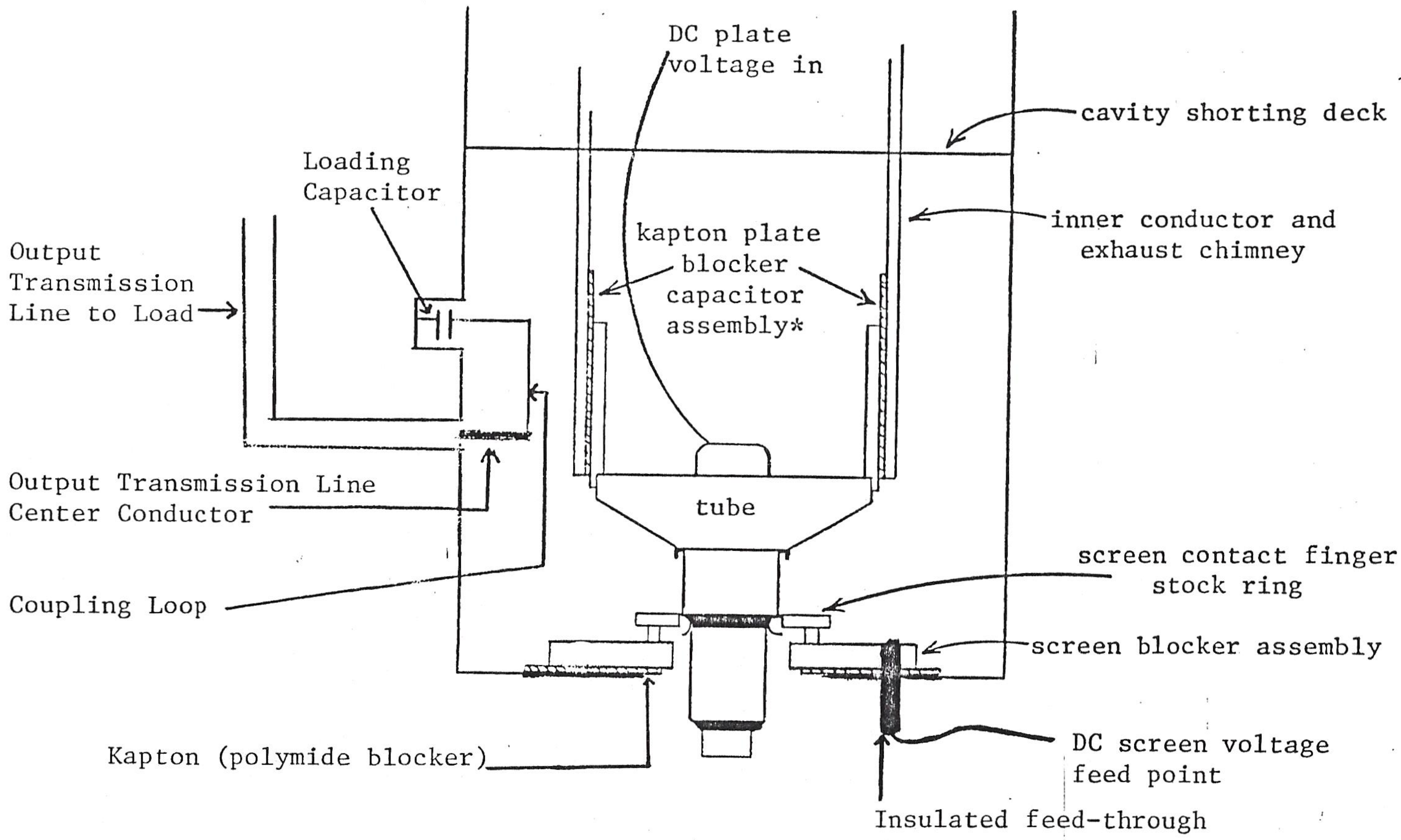


FIGURE 4 When the coupling loop is at the maximum current point of the ¼ wavelength cavity, coupling is obtained from the center conductor and the short.



*parts of plate blocker
1. inner tube
2. polymide film(kapton) 0.005" X 8" X 11ft 5 wraps
3. outer anode blocker shell

FIGURE 5 The FM 25K Quarter Wavelength Cavity showing the position of the coupling loop. It is at a position where the RF current and RF voltage are not at their maximum point.

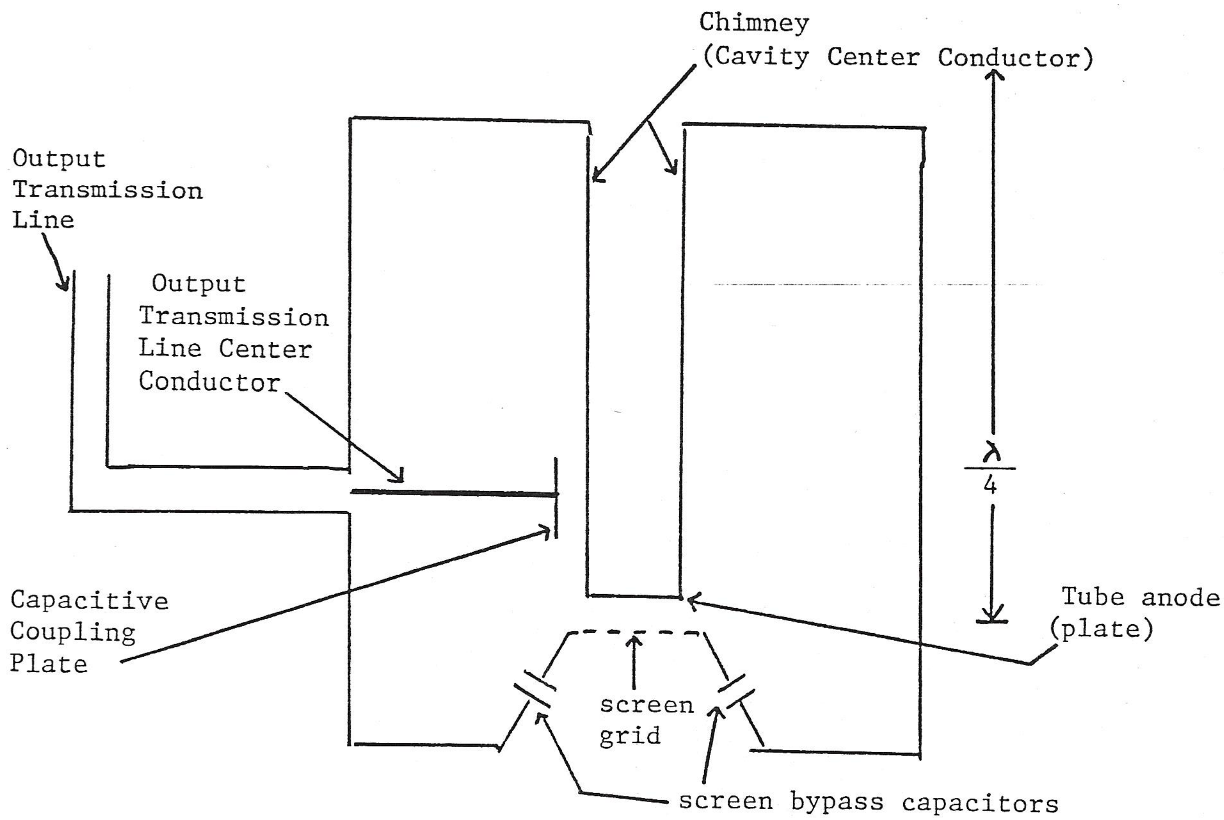


FIGURE 6 A quarter wavelength cavity with capacitive coupling to the output load.

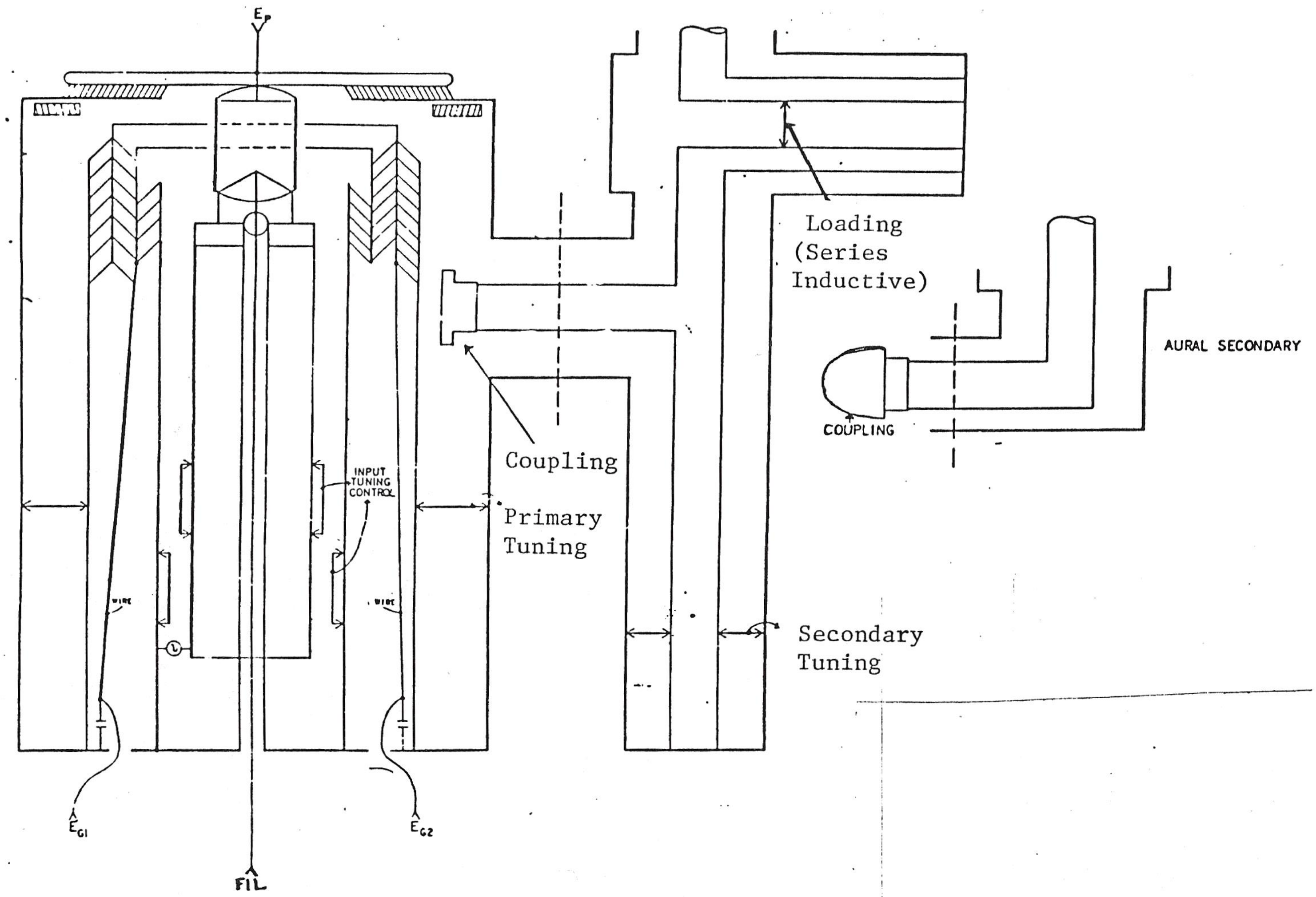


FIGURE 7 A double tuned 1/4 wavelength cavity used as a T.V. visual amplifier. It has capacitive coupling to a secondary cavity. If the aural secondary (shown to the right) replaces the secondary cavity, it becomes a single tuned capacitive coupled 1/4 wavelength cavity amplifier that is used in T.V. aural service.

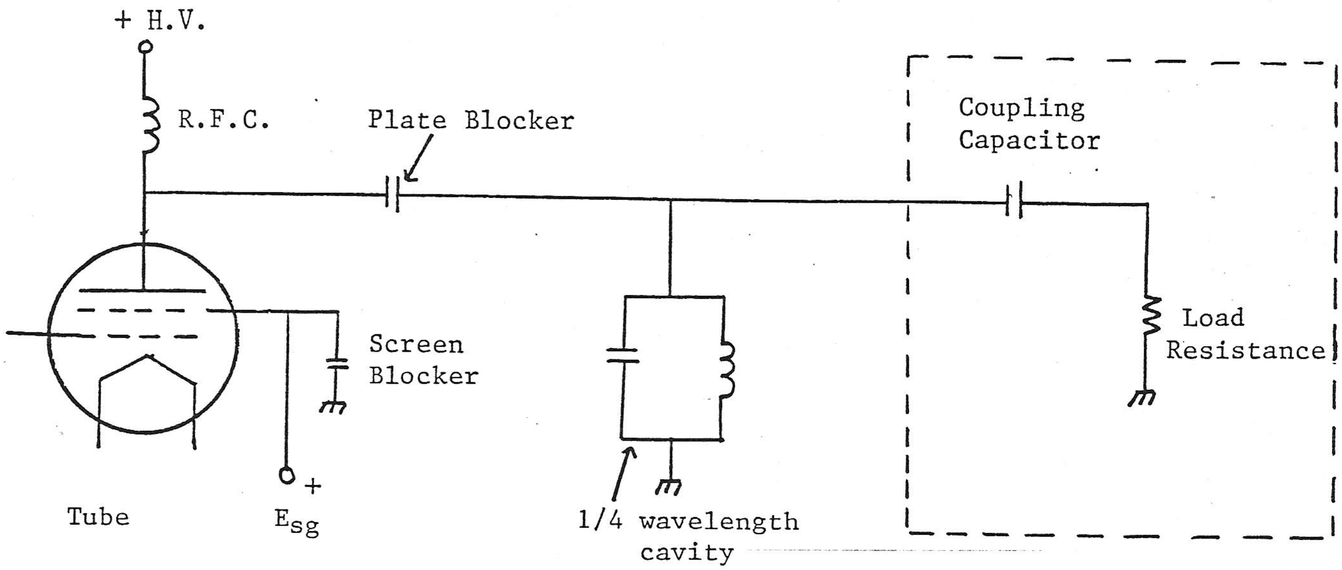


FIGURE 8 The equivalent circuit of a 1/4 wavelength cavity amplifier with capacitive coupling.

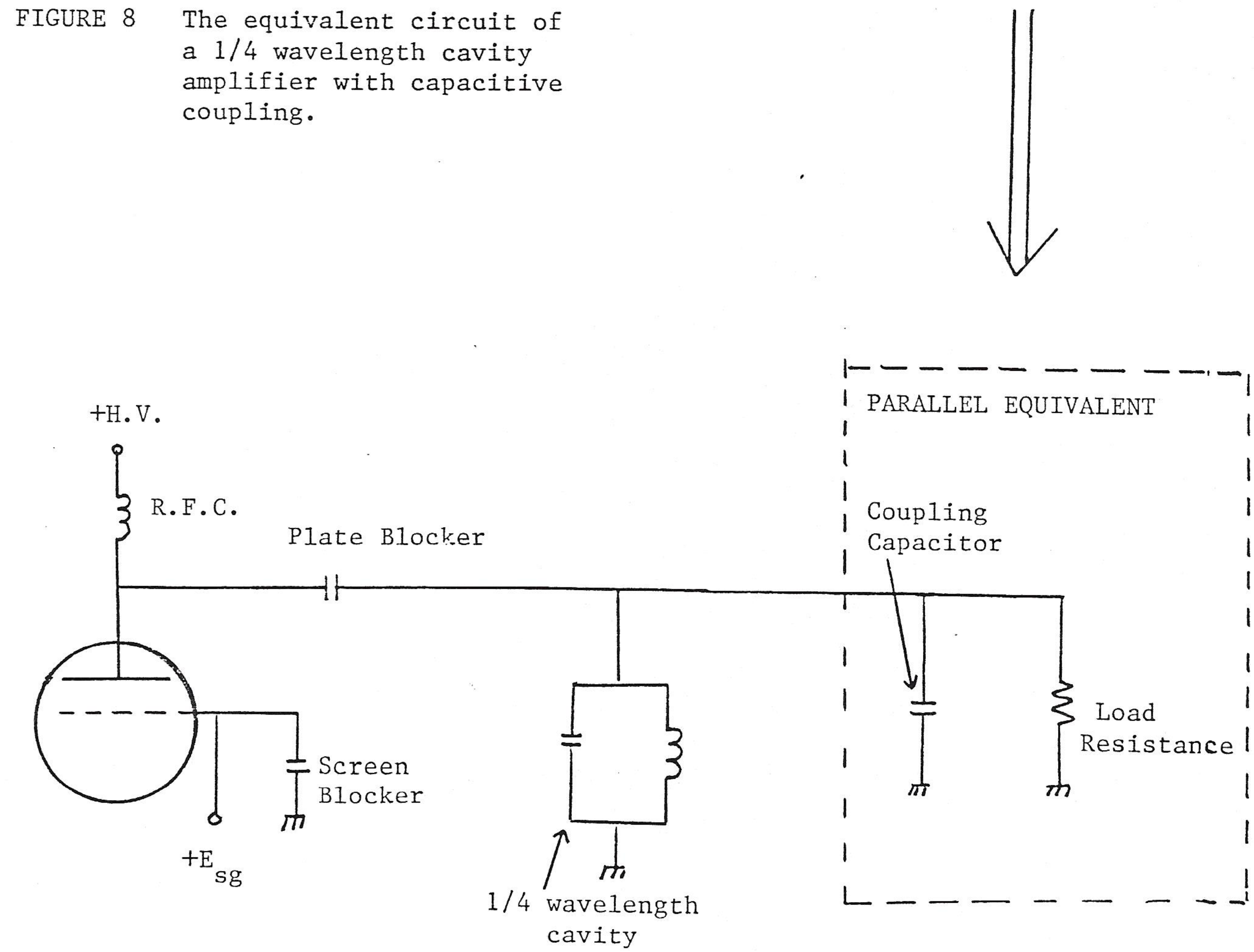


FIGURE 9 The equivalent circuit of a 1/4 wavelength cavity amplifier showing how the series capacitive coupling appears electrically to the tubes plate as a parallel circuit.

DSM THEORY

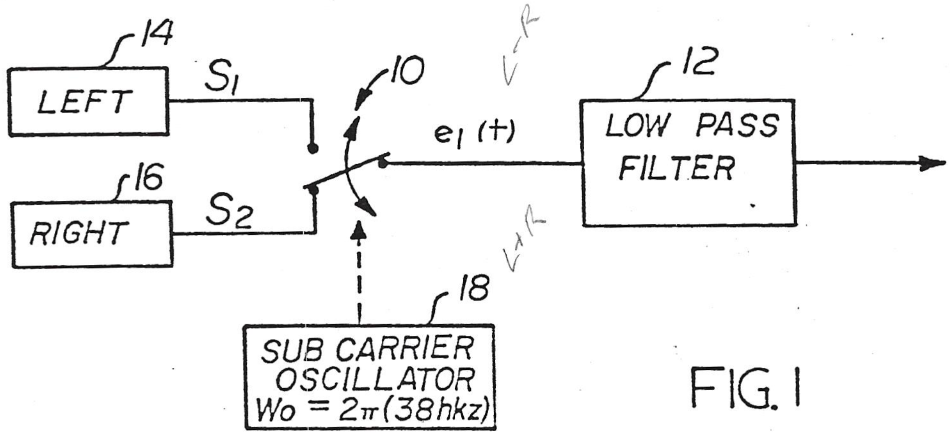


FIG. 1

(PRIOR ART)

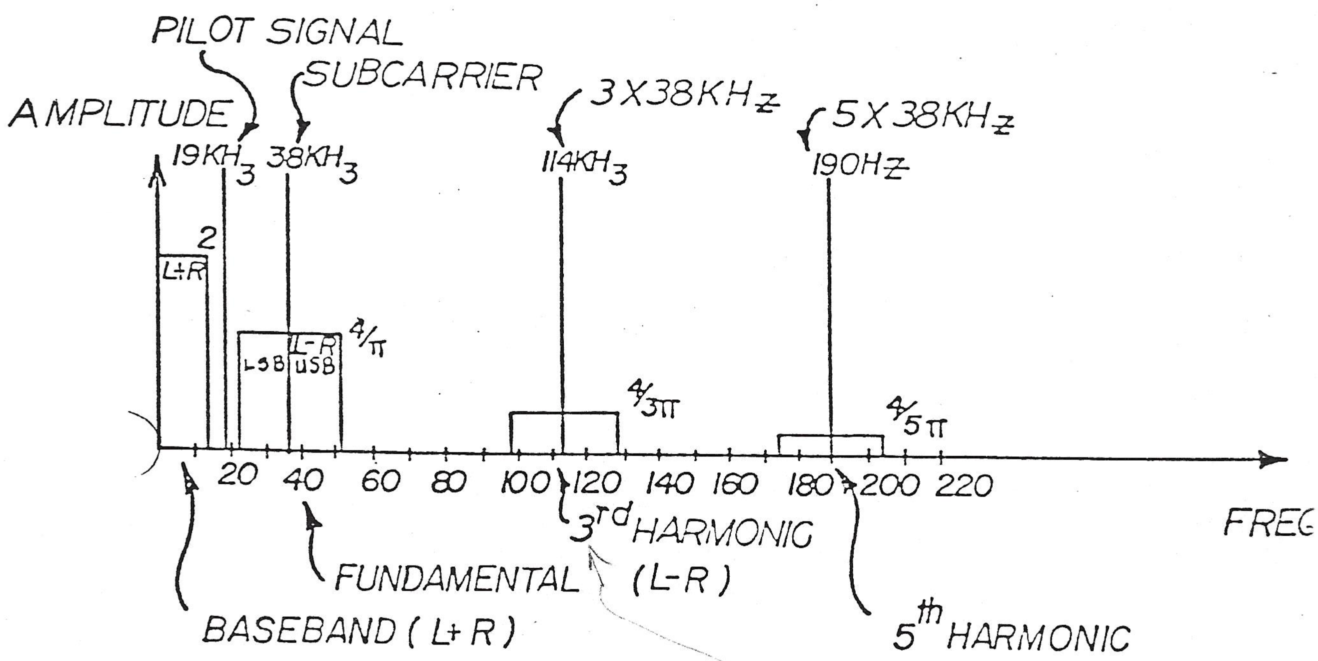


FIG. 2(A)

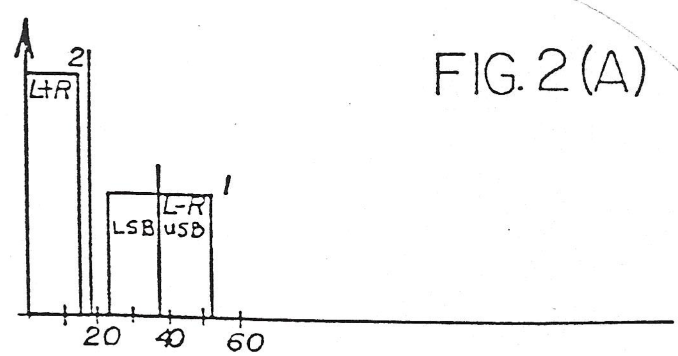


FIG. 2(B)

NOTE: NOT ONLY IS THE 114KHz HARMONIC undesirable it also generates (L – R) Sidebands.

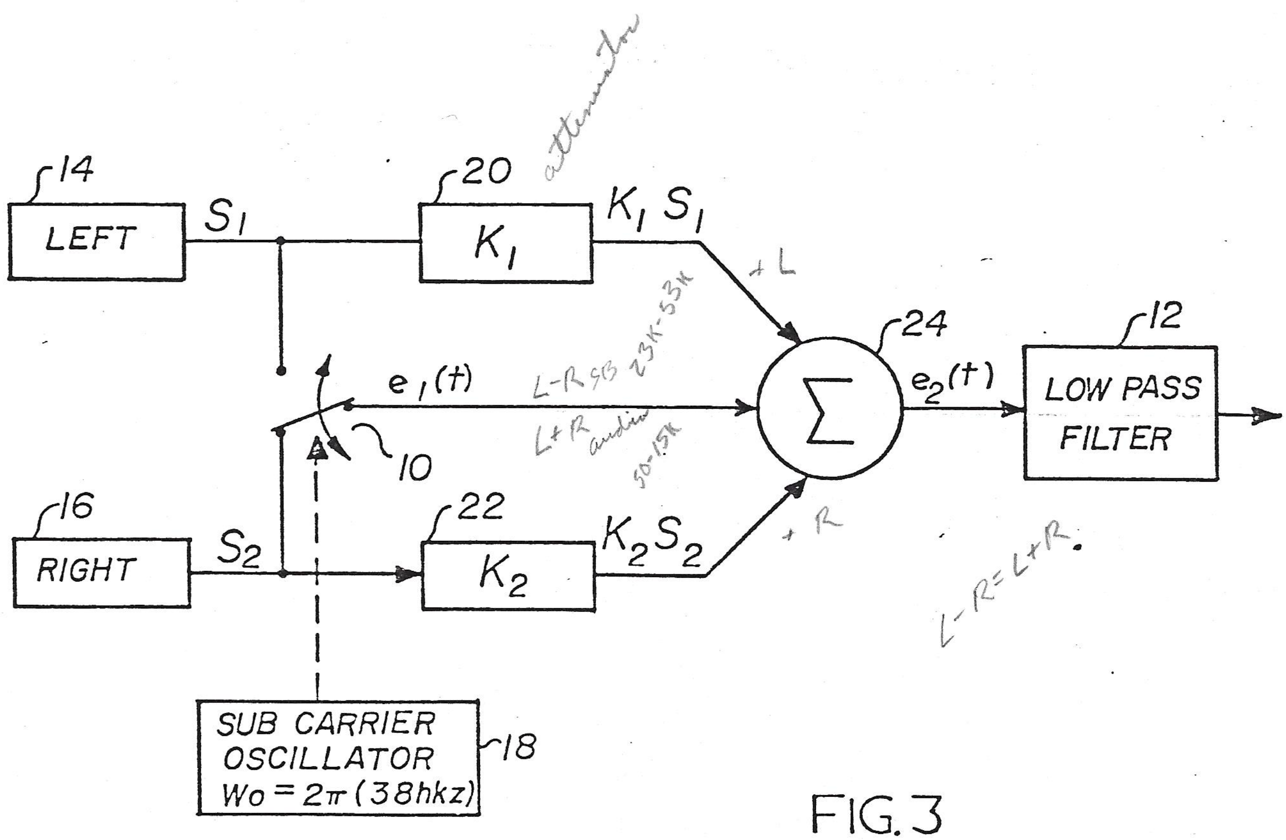

14
LEFT
S_1
20
attenuator
K_1
$K_1 S_1$
+L
24
Σ
12
LOW PASS
FILTER
$e_1(t)$
L-R SB 23K-53K
L+R audio 50-15K
10
$e_2(t)$
16
RIGHT
S_2
22
K_2
$K_2 S_2$
+R
L-R=L+R.
SUB CARRIER
OSCILLATOR
$W_o = 2\pi(38hkz)$
18

FIG. 3

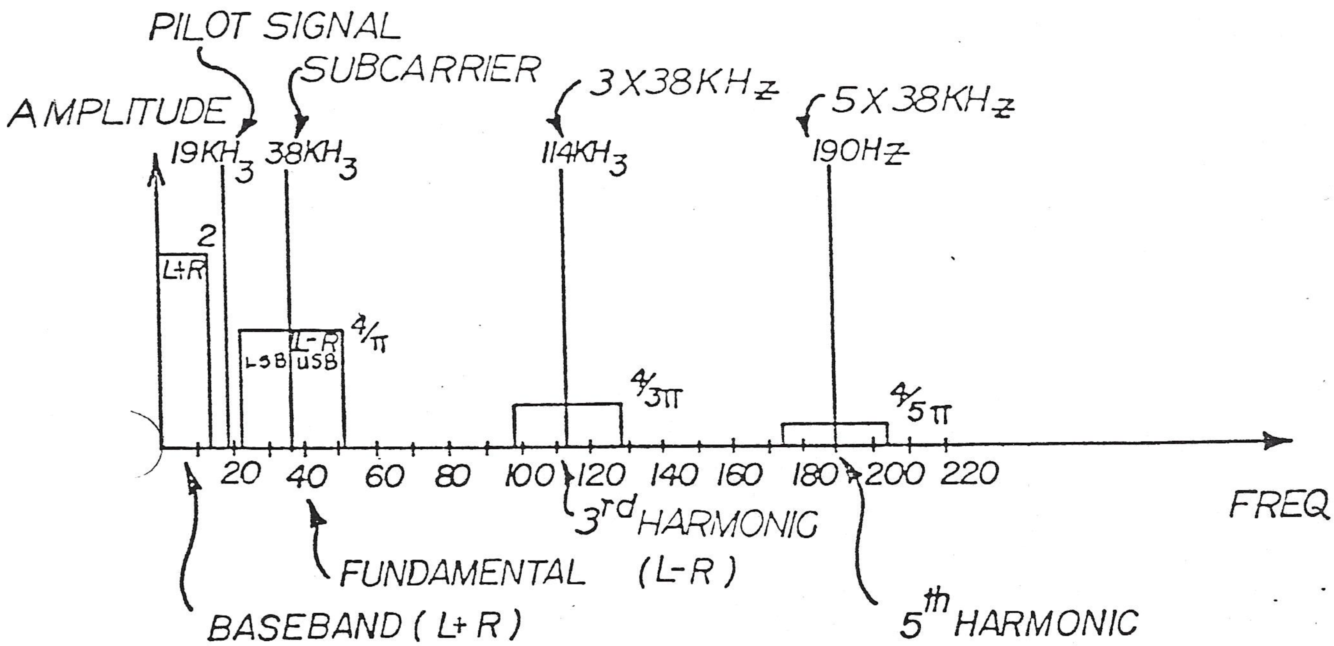

PILOT SIGNAL
SUBCARRIER
3X38KHz
5X38KHz
AMPLITUDE
19KH3
38KH3
114KH3
190HZ
2
L+R
L-R
LSB
USB
4/π
4/3π
4/5π
20
40
60
80
100
120
140
160
180
200
220
FREQ
3rd HARMONIC
(L-R)
FUNDAMENTAL
BASEBAND (L+ R)
5th HARMONIC

FIG. 2(A)

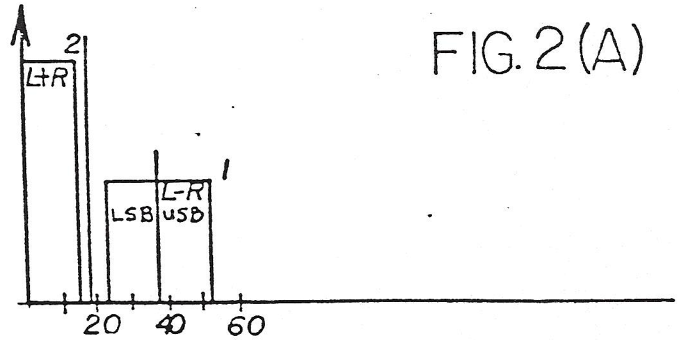
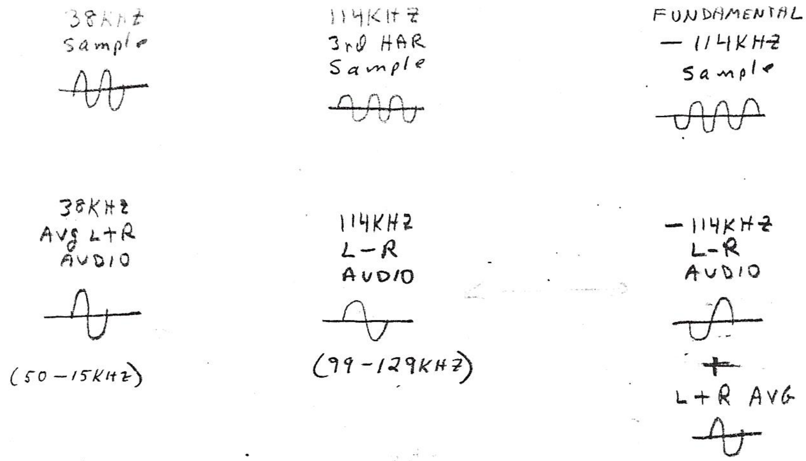

2
L+R
L-R
LSB
USB
1
20
40
60

FIG. 2(B)



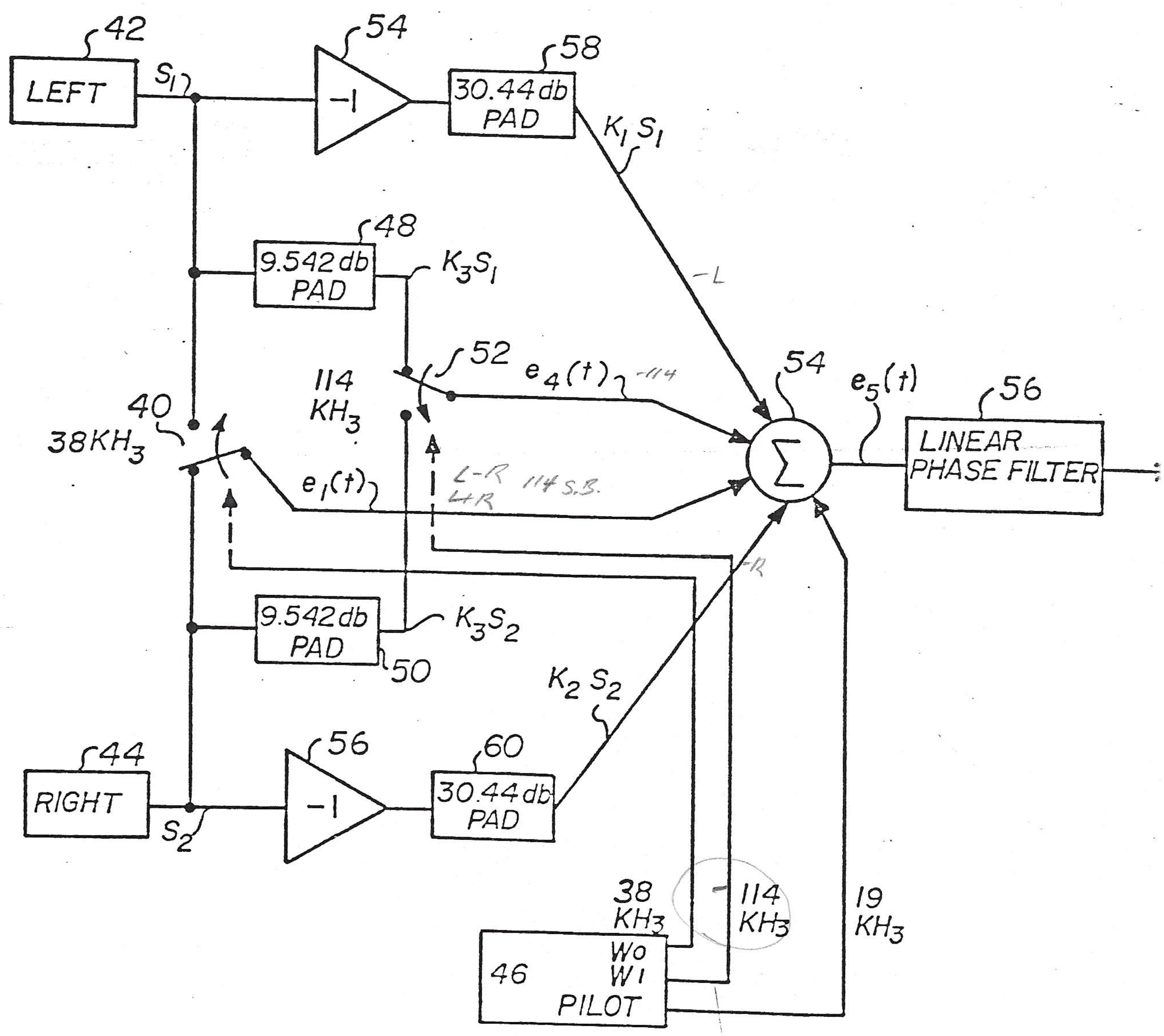
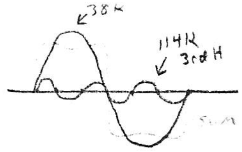


FIG.4

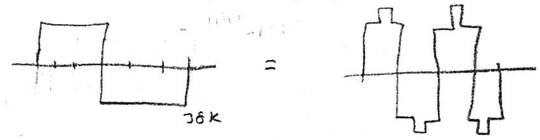
1. The 114KHZ signal can be used to cancel the 114KHZ 3rd Har of 38KHZ but what about the out of phase L – R audio sidebands that were generated by the 3rd HAR?

2. Sol. Sample the audio with 114KHZ SIGNAL that is out of phase with 3rd HAR of 38K. This sampled signal generaters (L–R) side bands that will be out of phase with the 3rd Har 114KHZ and will cancel the out completely when the amplitudes are ALSO. The L+R ...

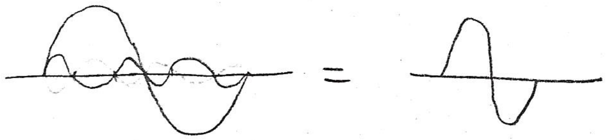
3. How do we balance it all out? Inject –L and –R! This decreases L+R until L+R = L–R.

1. 

+ 5^{th} + 7^{th} etc =

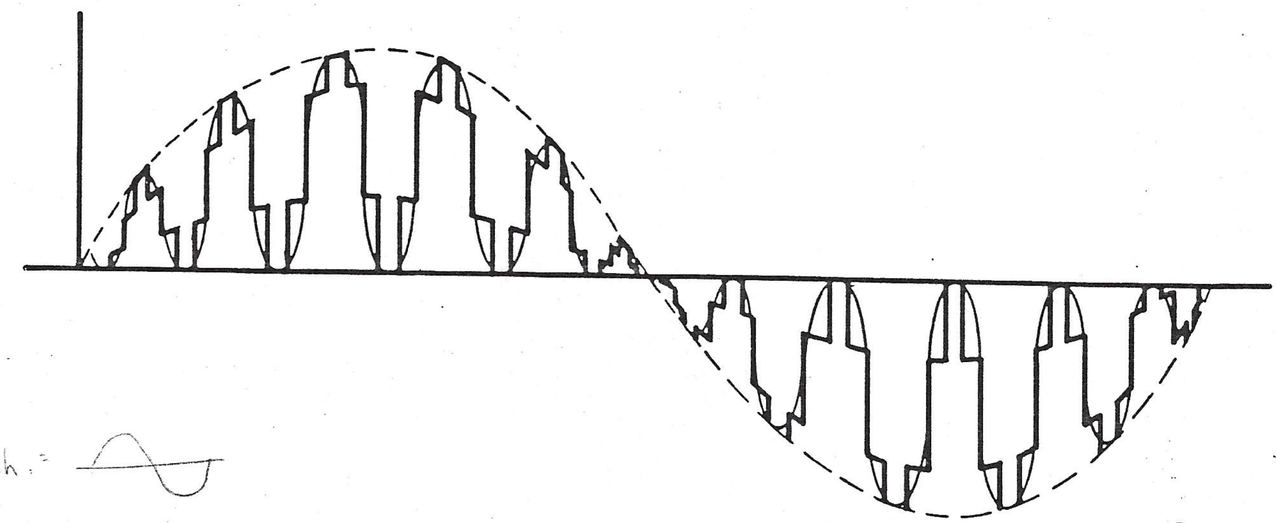
2. 

3. Since square waves are made up of sine waves, -114KHZ + 5^{Th} + 7^{th} etc cancel.



Fund. 38KHZ Sample sine-wave. This is what is seen on scope.

4-20. The output of the sum amplifier feeds the output amplifier through FL-1 which provides the required low-pass filtering. Output buffer U9 amplifies the signal level and provides a low impedance output. The COMP LEVEL control R27 adjusts the composite signal level to 1.0 VRMS for 100% modulation to drive the MOD OSC module circuitry. Several cycles of the digital sampling signal and the 38 kHz fundamental component as would appear at the output of sum amplifier (U8) (pilot off) are shown in figure 4-2.

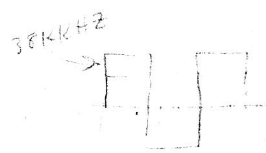


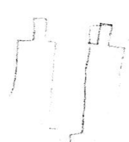
1742-31

Figure 4-2. DSM Waveform

Left ch =

Right ch =





when 3rd harmonics cancel

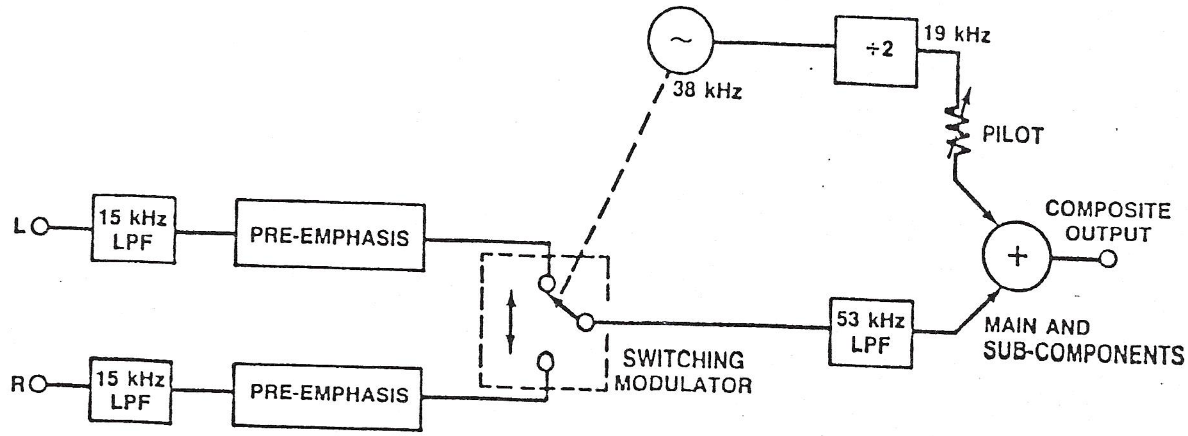


Fig. 5. Functional blocks of a time division multiplex stereo generator.

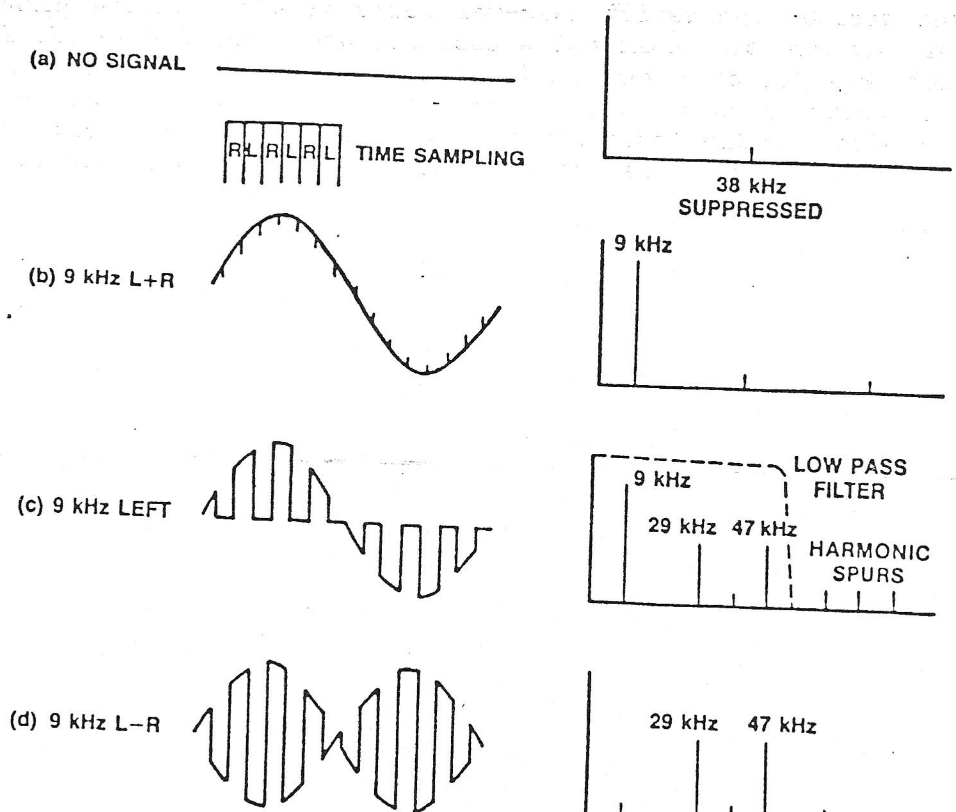


Fig. 7. Time domain and frequency domain diagrams of stereo baseband signals.

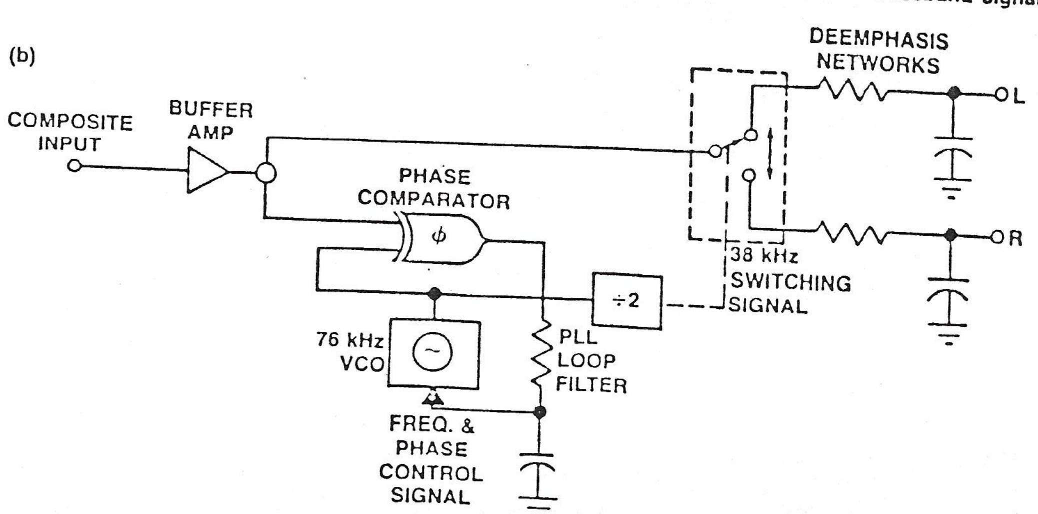


Fig. 8. Functional blocks of stereo decoders using (a) L+R and L−R matrixing and (b) phase-locked time-division multiplexing.