RF Note 108

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K1200 RF Driver Circuit Analysis and Design

Contents.

1. Introduction.

The K1200 C transmitter has been retrofitted to use a Thomson TH555 tetrode in lieu of the RCA 4648 tetrode. This move was necessary for four reasons.

The RCA tetrode was:

- increasing rapidly in price.
- not readily available.
- proving to be unreliable.
- being phased out of production.

A great deal more information on these problems as well as the tube selection process is described in RF Note 107.

The initial step towards accomplishing this tube change was to mechanically modify the transmitter to accept the new tetrode. The number of mechanical changes necessary for the modification was minimized in order to proceed to the test phase quickly.

The first modification was aimed at neutralizing the tetrode feedback capacitance which caused 'tuned grid - tuned plate' self oscillation near the operating frequency. Since the transmitter design was totally chassis- referenced this was not a conventional problem. Many different types of neutralizers were designed and tested with different degrees of success; however, most of these were narrow-band devices which had to be adjusted for each operating frequency. Furthermore, due to the high Q of the load (cyclotron cavities), these tuned type neutralizers could not maintain stability without a damping load on the output of the transmitter. This lack of stability is due to the rapidly changing impedance seen by the transmitter during initial RF excitation of the cavities and also due to the cavity input impedance being a strong function of frequency. These combined effects led to tuned plate - tuned grid oscillations at a frequency slightly different from the driven signal.

2. Transformer Neutralizer

The most stable performance was achieved by a variable capacitor connected to the anode and then fed through a phase inversion transformer to the final grid. This method is theoretically broadband. The original tests that verified the stable and broadband performance were completed with a small transformer which could not handle the large amount of energy transfer necessary for operation. Therefore, while the design was underway for a suitable transformer to handle the large amounts of energy, one of the tuned variety we had tested was put into service. This was done to let the experimental program limp along until a more stable, robust, and reliable device could be designed and fabricated. The first transformer designed was cooled by ambient air and had relatively long connections, while the final version was oil and water cooled with much shorter connections. This allowed the cyclotron to work up to about 21 Mhz without difficulties though above this we had problems due only in part to poor neutralization.

Above 21 Mhz the amplifier would self excite at moderate drive levels in the VHF range. This was found to be due to two problems; the chassis design of the driver still largely resembled that used for the 4648 tetrode which was not optimum for this new tube, and the screen bypass capacitor was not large enough in addition to having stray undesirable reactances due to its geometry. A new screen by-pass capacitor was designed that had about 6 times more capacity while simultaneously reducing stray effects to a minimum. At this point the transmitter could be operated to a frequency of about 23 MHz without any problems, above which its operation was not stable or reliable. This problem is attributed to both the driver and the coupling chassis for the final tube grid. We have now embarked on a total redesign of the driver chassis.

Since it was determined that large geometric changes to the driver amplifier were necessary, we also decided to associate with this change a new, more conventional type of neutralization. This particular method is termed 'Single Ended Grid Neutralization' and will eliminate the need for an elaborate transformer while maintaining the needed broadband characteristics. A description of this type of neutralization is given in Appendix B. The new design goals are to perfect the electrical characteristics rather than to simplify the mechanical implementation. We feel we are now able to do this, having identified the fundamental electrical problems which must be minimized or eliminated. Since we now intend to pursue a larger scale development, a very thorough general analysis of the transmitter is being performed instead of the cursory specific problem analysis which had been performed previously.

3. Harmonic Content and Modes.

RF amplifier resonant tanks require the most careful and thorough analysis if the desired results are to be achieved. This is in sharp contrast with tank circuits or cavities to which power is coupled through a matching circuit. This is due to the fact that RF amplifier tank circuits contain the non-linear active driving element in shunt whereas coupled cavities do not. In many cases the driving element is driven in class AB to class C to gain efficiencies of 60 to 80 %, which is a major requirement for high power amplifiers. This means the device only conducts current for about 180 degrees of the cycle. Resolving this periodic current pulse into its Fourier series representation shows not only a component at the desired frequency, but many harmonic components of various strengths. It should be obvious that if the shunt tank circuit has higher order modes that cross one of these harmonics, than this harmonic will appear in the amplifier output. Designers of amplifiers designed to drive a matched cavity cannot assume that the harmonic content of the output will be directly proportional to the harmonic content of the driving element. This is because each harmonic component will see a different size load. For example, the TH555 amplifier has an equivalent loaded shunt impedance of about 580 Ohms at the fundamental driven frequency, whereas at any of the higher harmonics this resistance may be 20 times as great. Therefore, if the tank circuit has a mode which crosses any harmonic, it may show up many times worse than would be expected if we were driving a broadband load such as a antenna. Since a vacuum tube is basically a voltage controlled current source, the output voltage swing is a direct function of the real load seen by the plate. The output voltage of a higher harmonic with the conditions for the load described here may be significantly higher than the fundamental swing, leading to premature screen and grid currents. This limits the output of the amplifier at the desired operating frequency, demonstrating the need for a thorough harmonic analysis of such a tank circuit. In addition to power loss concerns, the coupling between the output tank and input tank circuit may lead to the separate problem of uncontrolled self excitation at these higher frequencies. There is a great deal more that could be said on this subject; however from what has been stated there is sufficient reason to require that the shunt circuits for any amplifier be investigated for frequencies many times greater than the maximum driven frequency. Any mode found by this analysis should be carefully investigated and preferably eliminated or at least reduced to an acceptable level. It should also be clear that the analysis will have to begin by determining the TH555 voltage gain into both a 580 Ohm and 10k load. We will investigate the higher order modes into a 10k load versus a 580 Ohm load.

An overwhelming fact is that for even the most trivial resonant tanks, higher order modes will be present. This is because inductive and capacitive elements contain stray capacitance and inductance, respectively. Also, associated with each of these elements may be distributed reactances. All these elements appear in our circuit description as stray inductors, capacitors, or transmission lines. Simple passive elements such as interconnects also have reactances attributed to them so they appear as transmission lines or stray inductances in the circuit model. The designer must recognize all of these strays, determine or approximate their value, and include them in the tank model. Furthermore, these modes will normally not vary across the tuning range in the same way that the driven modes do. For these reasons, it is normally not possible to eliminate higher order modes or to force them to occur between harmonics of the fundamental where they will not be excited. The question to be resolved is, what is the acceptable level for the different modes ? The four general mode types are:

1. Self-excitation at the driven frequency (Oscillator).
2. Harmonic distortion which does not set up in such a way to lead to self-excitation (Harmonic Distortion).
3. Harmonic distortion which sets up in such a way that it leads to self-excitation (Harmonic Oscillator).
4. Non-Harmonic modes which lead to uncontrolled self- excitation (Parasite).

Harmonic distortion always exists but can be minimized, usually to at least 15 dB down from the fundamental. Harmonic oscillations, parasitic modes, and oscillator modes must be eliminated altogether. The tank resonant modes which lead to harmonic oscillations, however, cannot be completely eliminated. Instead, the magnitude of the effect causing self-generation must be reduced until this is no longer possible. The capacitive feedback which causes self-oscillation is typically eliminated by neutralization. Finally, it should be noted that actions which eliminate one mode can either create or enhance another mode!

4. Circuit Model.

The fundamental skill necessary for effective RF design and analysis is to model the system as an equivalent circuit consisting of lumped and distributed elements. With a model that accurately predicts the total response of the RF device to many times the desired operating frequency, an analysis may be performed to identify trouble spots and attempt to eliminate any problems. Figure 1 is a model of the driver amplifier and final tube. We have no problems with the final resonator, so it is not included. However, the significant parts of the final tube are modeled to include the transmission line like structure of the grid and screen electrodes and feedback capacitance. How each element was identified and arrived at will not be described, however a list of which elements are fundamental, stray, or device models is listed below.

Driver Tubes Elements:

CG2, CFB2, GA, CA2

Final Tube Elements:

TGK1, TGK2, TGS1, TGS2, CTGK, CTGS, CFB5, EFA

Fundamental Tank Elements:

CBB1, CAG, RA, CA, TA1, TA2, TDL1, TDL2, TDLG, CN, RN, CG

Stray Elements:

LA2, LKG, LAG, LCG, TN2, TN1, LN, CD1

Stray Elements are things we would rather not have in the circuit, but can do little to remove or minimize anymore. The rest of the categories of elements should be relatively obvious.

The large numbers scattered around figure 1 denote the nodes of the circuit. From here on, voltages will be with regards to those nodes to common and referred to as V(1), V(2), ... , etc. Common is defined to be the main tube cathode plane. Phases will be the phase of the voltage at any given node with respect to the driving source which is set to 0 and will be denoted as VP(1), VP(2), ... , etc.

5. Analysis.

The first step in the analysis is to crudely determine the main tank values as a function of frequency. This is done by simplifying the circuit enormously and computing numbers by hand. With this data we can develop a feeling for what the element values should be as a function of frequency to insure the computer model is relatively correct. This is known as a 'back of the envelope analysis', although the envelope is typically several pages long. Appendix 1 contains this crude analysis.

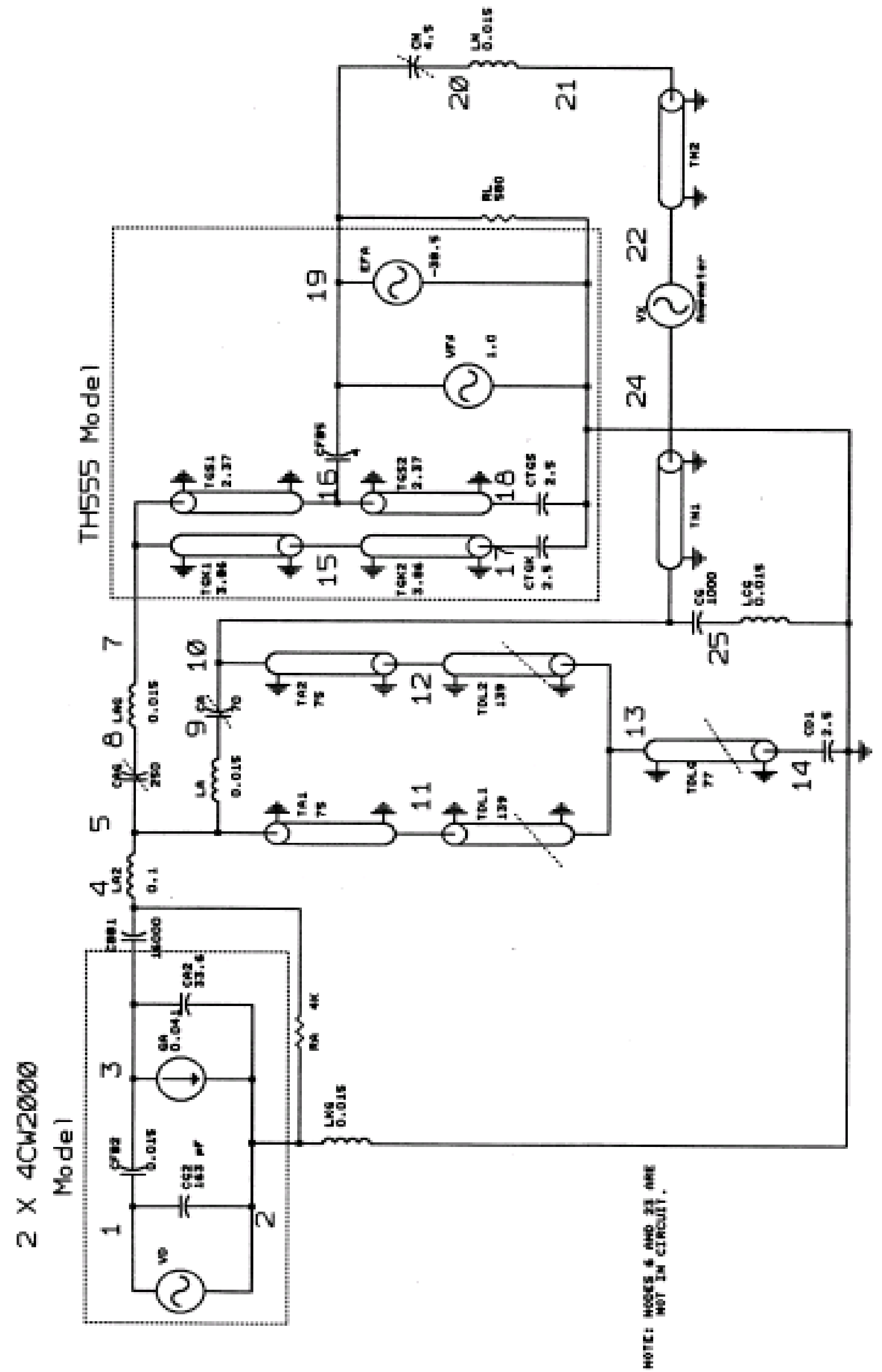


Fig. 1 K1200 RF TRANSMITTER SPICE MODEL

The next step is to enter the actual model in all its glory into a computer and determine the actual values for three points in the frequency band; Fmin, Fmid, Fmax. It is important that the final values arrived at are near enough to the cursory analysis to insure that all previous calculations and assumptions were valid. If problems exist here, it may be only a simple typographical mistake, or could be a gross modelling error. The signals of interest in figure 1 are:

V(1), V(3), V(14), V(15), V(19), I(VN)
VP(1), VP(3), VP(14), VP(15), VP(19)
Length of TDL1 = TDL2, Value of CN, RA, RN

The main tuning elements are TDL1, TDL2, and TDLG. All other values are to remain as shown on Figure 1. The analysis will proceed as follows.

1. Determine the voltage gain of the TH555 tetrode for both a 580 and 10k load.

2. Determine the values for the tuning elements for 9, 19, and 27 Mhz +/- .05 Mhz. At each of these points determine the value of CN to neutralize out the plate to grid capacitance CFB5. Find the necessary value of V(1) such that V(15) = 450. Do a frequency scan from 9 to 150Mhz step 1 Mhz to locate the approximate frequencies of any higher order modes.

3. If higher order modes were found in step 2, then we must find ways to eliminate any adverse effects due to these modes. Remove the driver drive voltage source and drive the final with a 1 volt, 0 degree source. Look at the response at V(15) and V(3) as well as other places of interest. If it appears that these modes could lead to self oscillation or would cause large amounts of harmonic distortion the circuit must be modified or adjusted to remove or reduce these effects. Once these changes are found than the analysis may proceed with this new circuit arrangement.

4. Iterate through steps 2 and 3 to insure that the circuit still functions properly, may still be neutralized, and no new modes have been created.

5. Go back to driving the driver circuit. Acquire tuning information and neutralization data for the final circuit configuration arrived at above for 9 through 27 Mhz step 2 Mhz. Also include for each point the drive requirements and location of the higher order modes.

6. Enter this data into a spread sheet. Plot and graph the above information to determine the intercepts of harmonics to higher order modes.

7. Go back to driving the final with a 1 volt 0 phase source. Look at the response at 9 through 27 Mhz step 2 Mhz plus the crossovers found in step 6. We are interested in the phase and magnitude information at the final grid V(15) and VP(15).

8. For each higher order mode enter the following data into a spreadsheet:

1. Tuning Frequency
2. Higher order mode frequency
3. Grid Magnitude
4. Grid Phase
5. Magnitude of grid (V(15)) voltage which is 180 degrees out of phase with the plate (V(19)) times the calculated voltage gain for the tube found in step 1 for a 10k Ohm load.
6. At each tune point, determine what harmonic of the fundamental the mode represents.
7. Enter this harmonic number found for each tune point into the spreadsheet along with the tube harmonic gain factor calculated found in step 1
8. Compute the adjusted feedback response of the final grid-plate circuit by multiplying the grid magnitude found in step 5 by the harmonic gain factor found in step 7.
9. Plot all of these effects to determine the circuit response across the entire band and ensure that self oscillation should not occur.

9. If any problems are remaining then some form of iteration of the above steps is necessary until all problems seem to be resolved.

The next two plots and data are the results of the analysis of the TH555 analysis for both a 580 Ohm and 10 kOhm load.

TH555 Operating into a 580 Ohm Load

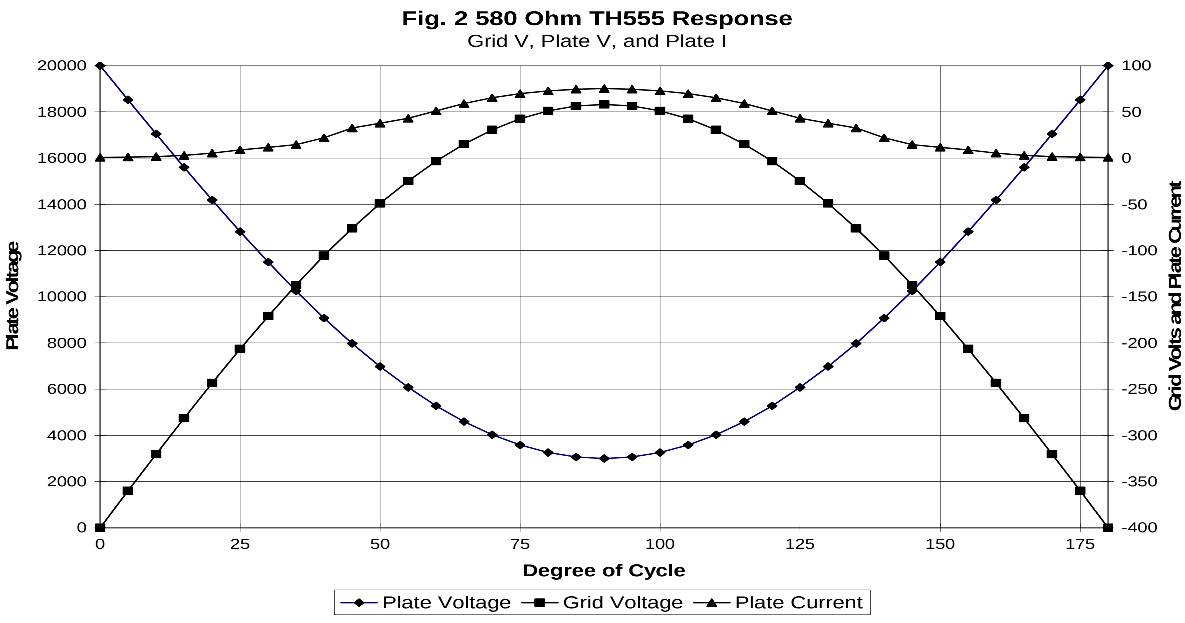
Dc Grid Bias	=	-400.00 Volts
Dc Screen Bias	=	1000.00 Volts
Dc Plate Bias	=	20.00 kV
Peak Grid Drive	=	455.00 Volts
Peak Anode Swing	=	17.00 kV
Anode Bias Current	=	0.45 Amperes

Tube Response to the above Bias Point

Dc Grid Current	=	0.25 Amperes
Dc Screen Current	=	0.17 Amperes
Dc Plate Current	=	16.86 Amperes
Anode Bias Current	=	0.45 Amperes
Grid Dissipation	=	99.43 Watts
Screen Dissipation	=	169.31 Watts
Plate Dissipation	=	86.70 kW
Output Power	=	250.50 kW
Necessary Load Resistance	=	576.84 Ohms
Plate Efficiency	=	74.29 %

Linear Harmonic Analysis of Anode Current Follows

Harmonic#	#dB
1	0.00
2	-12.62
3	-14.64
4	-19.65
5	-32.10



TH555 Operating into a 10000 Ohm Load

Dc Grid Bias	=	-400.00 Volts
Dc Screen Bias	=	1000.00 Volts
Dc Plate Bias	=	20.00 kV
Peak Grid Drive	=	225.00 Volts
Peak Anode Swing	=	17.00 kV
Anode Bias Current	=	0.45 Amperes

Tube Response to the above Bias Point

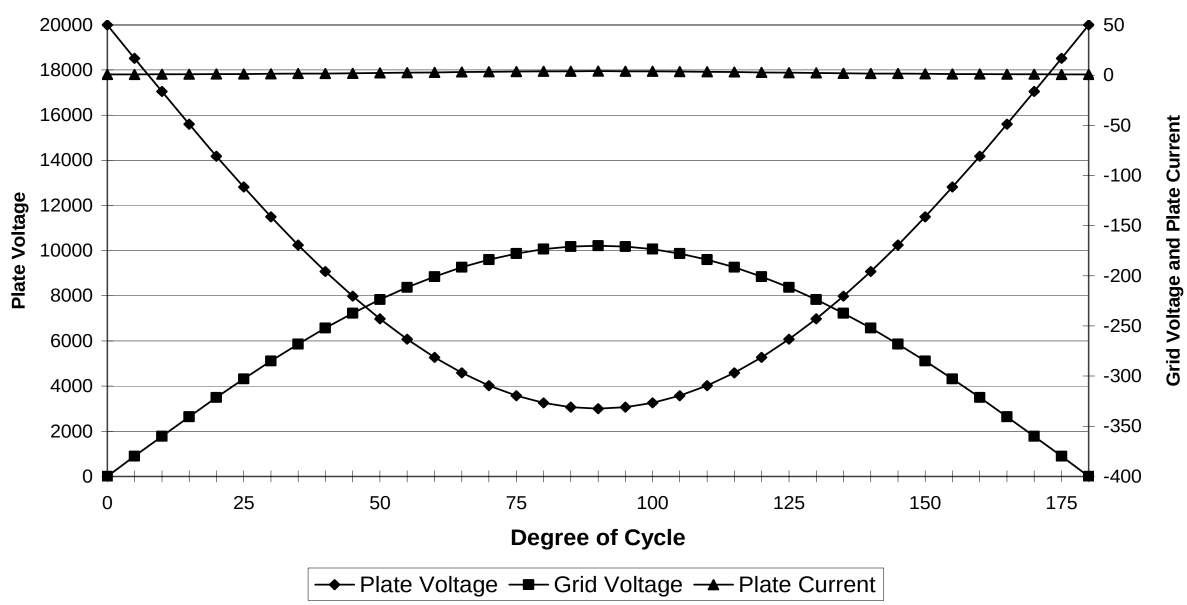
Dc Grid Current	=	0.00 Amperes
Dc Screen Current	=	0.00 Amperes
Dc Plate Current	=	1.01 Amperes
Anode Bias Current	=	0.45 Amperes
Grid Dissipation	=	0.00 Watts
Screen Dissipation	=	0.00 Watts
Plate Dissipation	=	6.60 kW
Output Power	=	13.67 kW
Necessary Load Resistance	=	10572.20 Ohms
Plate Efficiency	=	67.44 %

Linear Harmonic Analysis of Anode Current Follows

Harmonic#	#dB
1	0.00
2	-3.39
3	-4.23
4	-9.32
5	-29.16

Fig. 3 10,000 Ohm TH555 Response

Grid V, Plate V, and Plate I



Figures 2, 3, and the accompanying data indicate that the voltage gain for the TH555 into a 580 Ohm load is 17000/455 = 37.4 while into a 10000 Ohm load it is 17000/225 = 75.6. It is also worth noting the harmonic content of these two cases on the above charts and data. It is easily seen that the 10K Ohm load case, which is representative of the higher order modes, has more harmonic content then the 580 Ohm load case representing the driven mode. These facts will be used to determine the expected harmonic distortion and stability of the amplifier with respect to the driver circuit.

A simple hand analysis was performed and the results are in Appendix A. This analysis found the approximate value for the neutralizing capacitor and an equivalent inductance between nodes 5 and 10 to achieve resonance at 9, 19, and 27 Mhz. The computer model was run with these initial elements to verify the model and to fine tune their values. Once this was done, the actual elements were placed in the model with initial computed values. For example, the tuning inductor between nodes 5 and 10 was replaced by TA1, TA2, TDL1, TDL2, and TDLG. These elements were than adjusted to fine tune the resonances at 9, 19, and 27 Mhz. At each of these points in the band, the frequency was swept to find and quantify the higher order modes. Two modes were found between 80 and 100 Mhz. Both of these resonances would have caused problems in the original configuration, making it necessary to identify the factors which could dampen them. Since stray elements had already been minimized and all other elements in the system were necessary, we could only vary the location and magnitude of the resistive elements to eliminate the higher order modes. The ground rules for this manipulation were:

- Use 2 resistors in the circuit.
- One must either be between node (4, 2) or node (9, 10).
- The other resistor, if used, may be placed anywhere.
- The total power dissipated by all the resistors must be less than or equal to 2 kW.

We found that the neutralizer system was responsible for enhancement of the most troublesome mode, so a resistor would be used as a damping element in that subcircuit. The final arrangement was a resistor of 3.9 kOhms from node 4 to 2, and a resistor of 40 Ohms from node 10 to common. This arrangement led to what appears to be a very stable and harmonic free driver circuit with respect to both itself and the final.

Figures 4-7 describe the behavior of the final driver circuit. As indicated in these graphs the troublesome modes occurred at approximately 81 and 96 MHz. The lower frequency resonance was singled out for further study because it possessed a stronger ability to go into oscillation.

Fig.5 Neutralizing Capacitance
versus Frequency

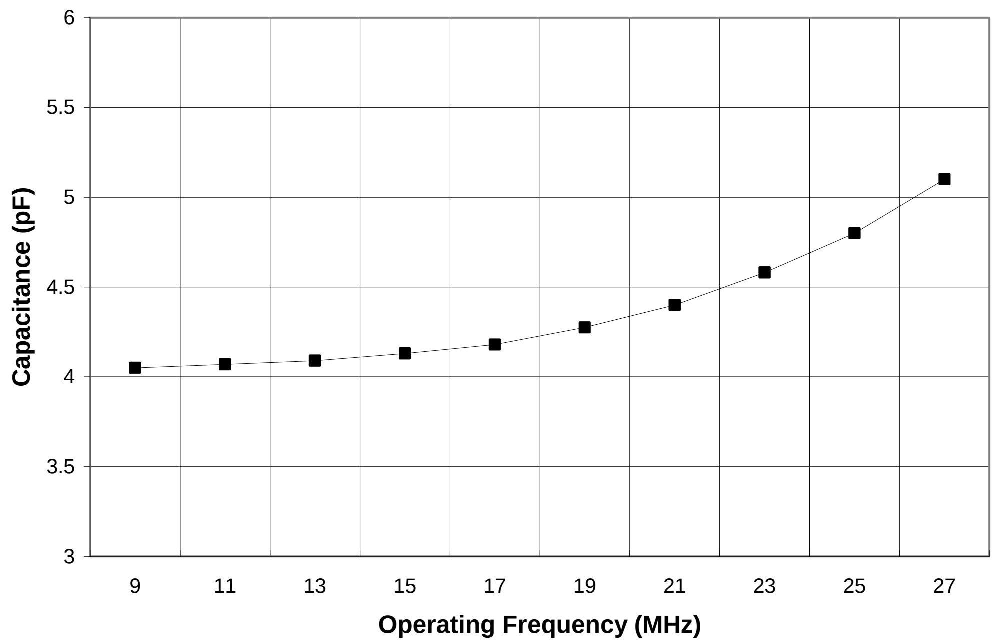
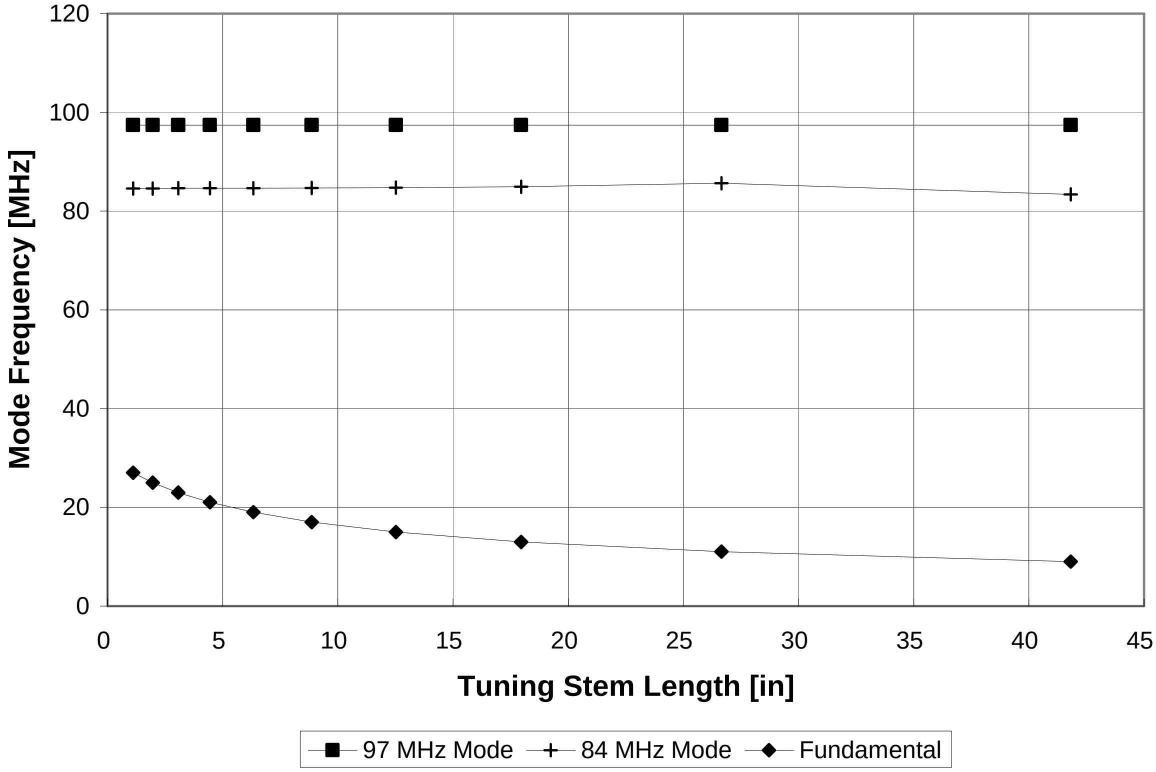
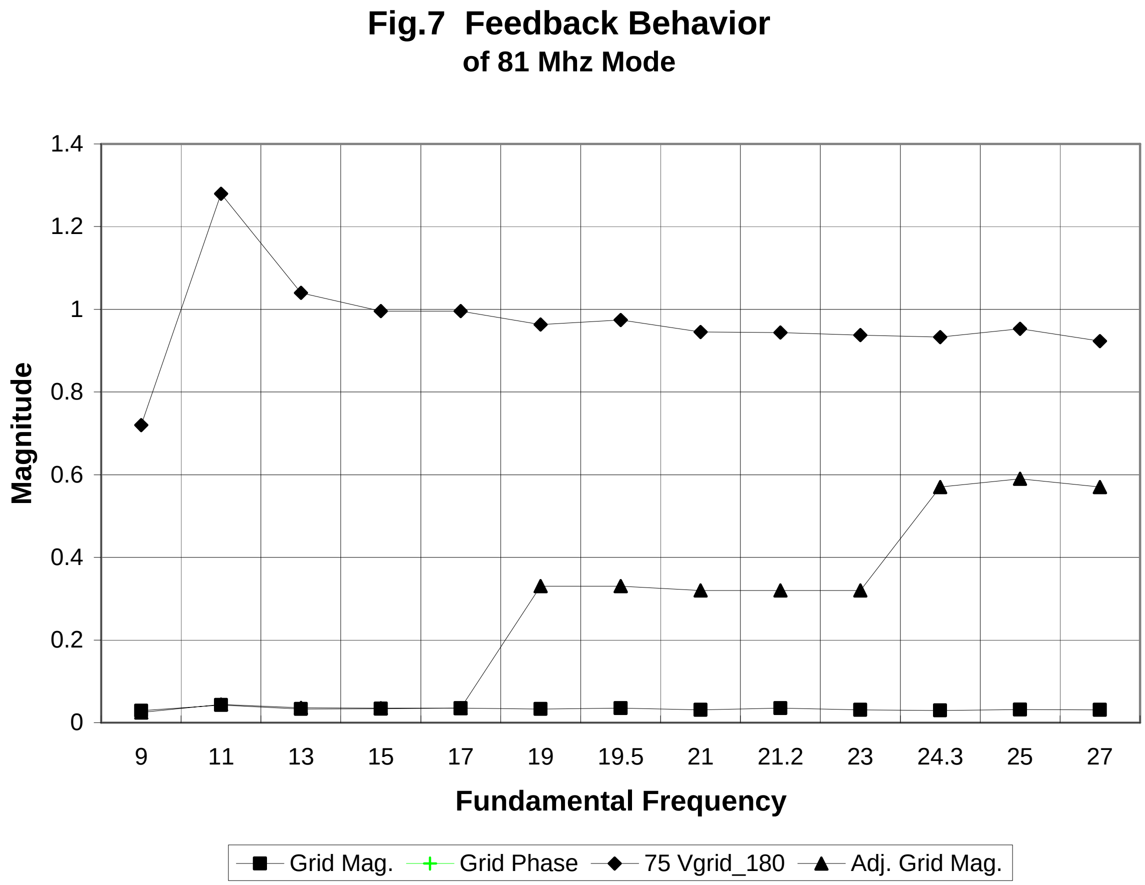


Fig.6 Frequency of Principal Modes
Over Tuning Range





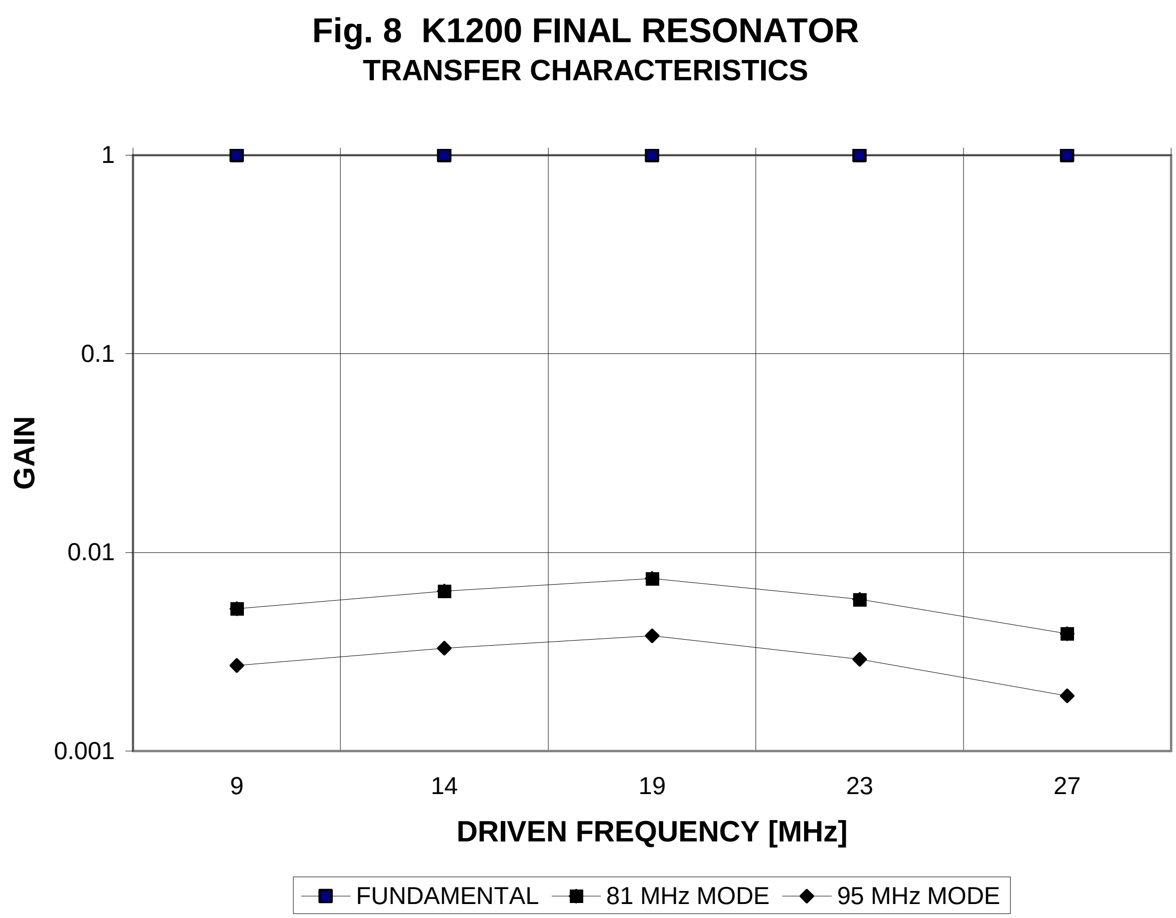
A brief explanation of the plots of Figure 7 amay be helpful. The curve marked 'Grid Mag.' is the magnitude of the voltage on the final grid with a signal of one volt placed on the final plate. Note that the tube is completely passive for this test since its voltage controlled voltage source is disconnected (EFA). The plot marked Grid Phase is just the phase of this same signal.

The plot marked as ' 75 Vgrid_180 ' is the magnitude of the grid voltage vector that is 180 degrees out of phase with the driving signal at the plate. This component corresponds to a signal which would begin to oscillate if its magnitude exceeded one volt. The multiplication factor of 75 is the voltage gain of the tube into a 10 KOhm load as determined in Figure 3.

The last plot indicated as 'Adj. Grid Mag.' is the magnitude of the previous signal multiplied by another gain factor. This gain factor is the attenuation due to the tube's non-linear response for the harmonics of the fundamental. Since the 81 MHz mode can be represented as a harmonic of the fundamental frequency (i.e. 81 MHz is the 9th harmonic of 9 MHz), and we have determined the attenuation for the harmonics of the tube (see figs. 2, 3), we can specify how much the signal on the grid is reduced by this gain factor.

6. Final Resonator.

An additional simulation was performed to determine the transfer characteristics of the final resonator to insure that it would not support any unwanted modes. The gain characteristics of each of the higher order modes detected in the earlier stages plus the fundamental are shown in Figure 8. From this graph we can see that the problematic mode at 81 MHz, which came closest to self-oscillation previously, is again dampened by the final resonator. The 95 MHz mode is also dampened considerably in comparison to the fundamental though it was never a significant threat to self-oscillate. The graph of Figure 7 could be adjusted again by taking into account the gain factor represented by figure 8. Part of the reason for the severe damping of these modes by the resonator is due to the advantageous location of a zero at 90 MHz in the final stage's transfer characteristic. It should be noted that in this simulation the output coupling capacitor was held constant, though it is varied when actually tuning the transmitter. Another important factor in the simulation is that the final load is 1200 Ohms, not the actual matched case of 50 Ohms.



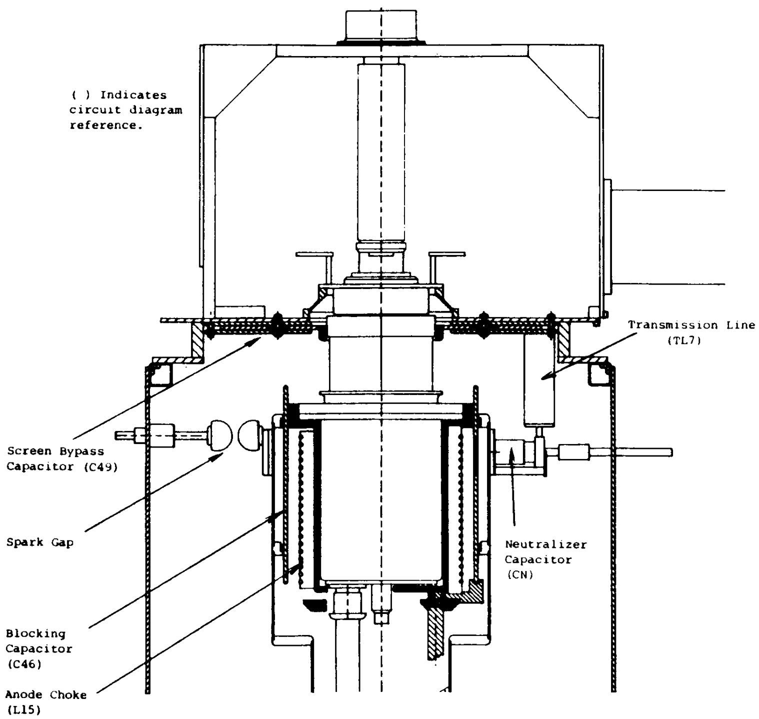


Fig. 11 New Transmitter.

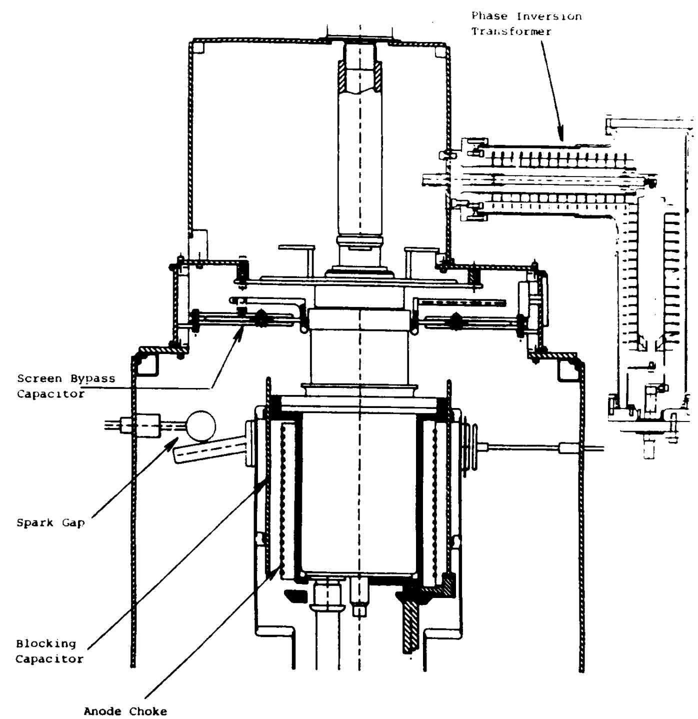


Fig. 12 Previous Transmitter.

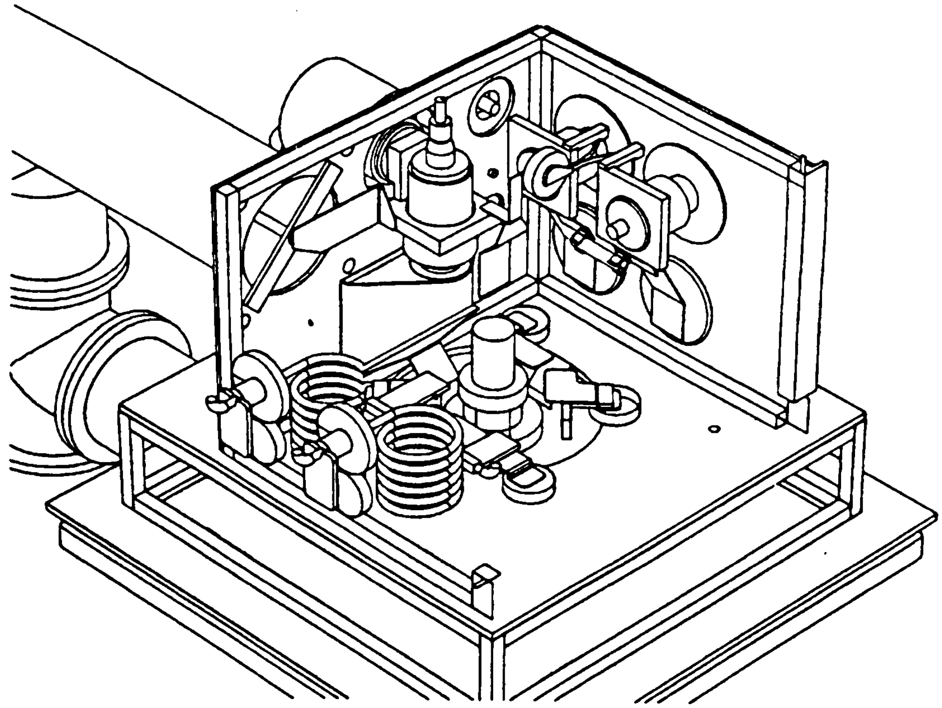


FIG. 9: Original driver amplifier.

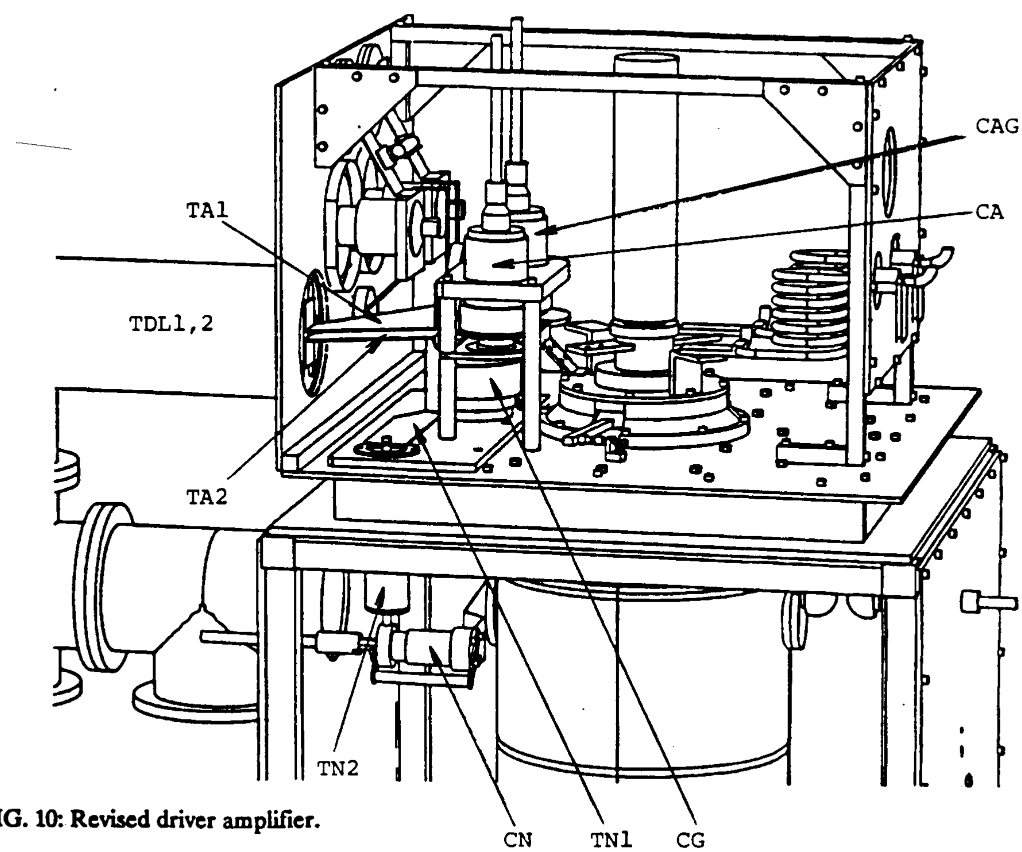


FIG. 10: Revised driver amplifier.

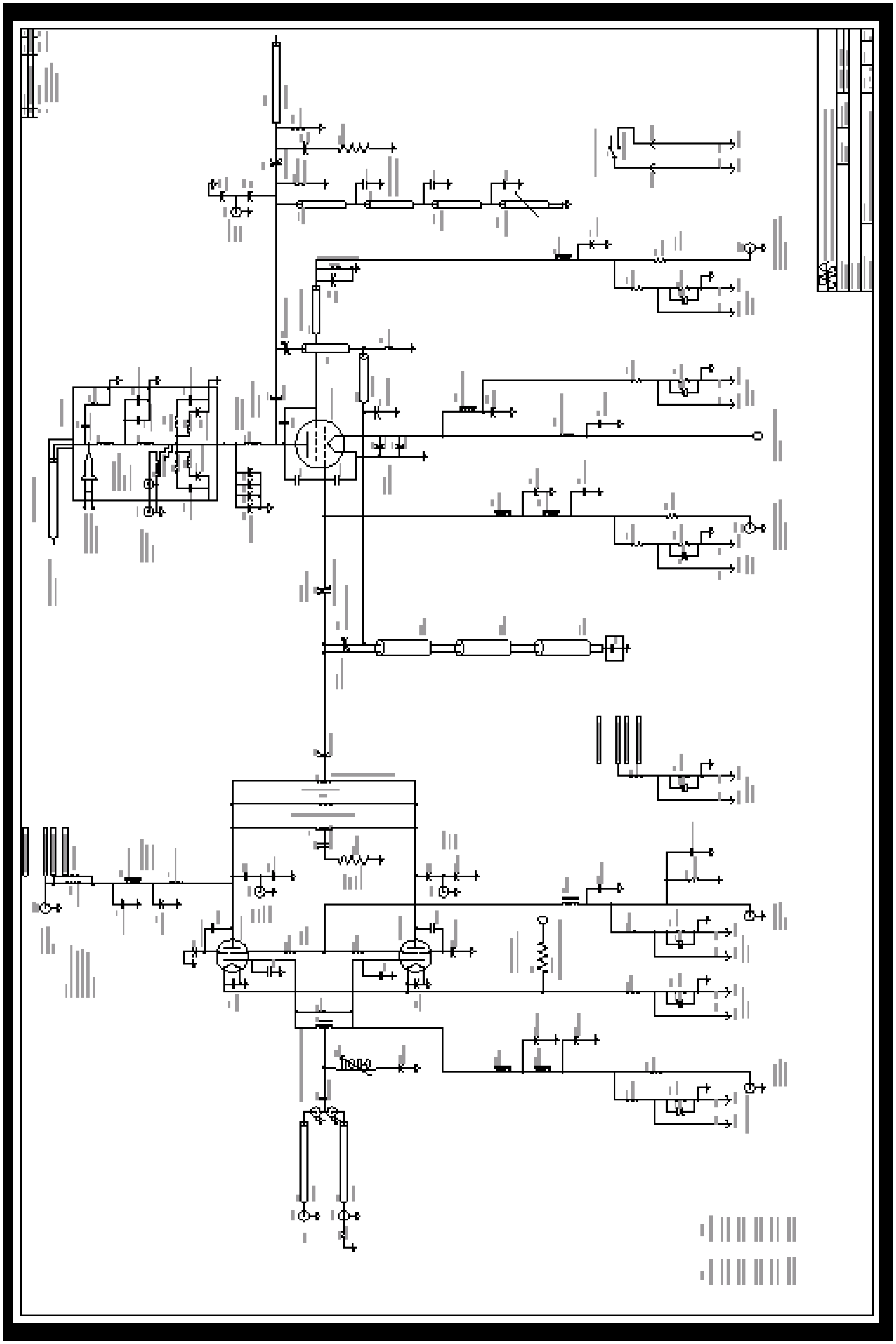


Fig. 13 K1200 RF Transmitter

7. Field Testing.

Field testing commenced with tuning the driver only. The driver was successfully tuned through its frequency range by setting the tank circuit capacitance CA to a constant value and varying the tank circuit inductor (transmission line). The next important analysis was to determine whether sufficient drive power existed to create the necessary voltage swing on the final grid. This was initially not possible because the 50 Ohm shunt impedance in parallel with CG dissipated too much power. This load was originally included in the circuit as a means of damping unwanted modes, though fortunately could be removed entirely with no loss of stability due to the final resonator.

Neutralization was achievable with this configuration, though the capacitance required was the minimum value of the variable capacitor, CN. Further tests of the initial configuration revealed that the transmitter was stable across the entire band when run into an open circuit, though a strong second harmonic component was visible at 23 MHz.

The next iteration on the transmitter circuit was to change the value of CG from 750 pF to 1000 pF. In addition, a shunt resistor of 4 kOhm was placed from the driver anode to ground. As mentioned above the final grid load was removed entirely. The capacitor change resulted in stable neutralization using a less than minimum value of the neutralizing capacitor, CN. The changes also dictated that the tank circuit capacitance (CA) must equal 70 pF in order to tune over the frequency range using only the variable transmission line.

Lastly, this configuration was run into a high power 75 Ohm load. The transmitter was stable across the tuning spectrum and successfully reached peak power output at 27 MHz (20 Amps @ 18 KV with 70% efficiency, giving 252KW). For comparison purposes three dimensional views of both the present and previous driver boxes are shown in Figures 9 and 10. Also included for reference are side views of the driver/final boxes showing the previous configuration using a transformer neutralizer (Fig. 11), and the present structure using the capacitive bridge neutralizer (Fig. 12).

8. Cyclotron Resonator Modifications

Before discussing the final tests into the cyclotron cavity, it would be useful to describe changes made in the cyclotron stem since they affected the performance of the RF system. Alterations were made to the vacuum seals of the C upper dee stem insulator which is shown in Figure 14. The previous insulator was vacuum sealed using an indium seal which developed leaks over time (Figure 15). These leaks were attributed to voids forming in the seal which would not re-fill due to the inelasticity of indium. The movement responsible for leaks developing over time is probably due to both RF heating of the Indium and pressure changes on the sealing surface due to thermal cycling. Because these leaks require 2 or more days of downtime to repair and occur every few months we are testing the feasibility of using O-rings to seal these locations. The C upper stem seals were refitted with silicone rubber O-rings, and a copper spacer which both held the O-ring in place and supported the stem. This configuration is shown in figure 16. Silicon rubber was chosen as the initial material to test because it is readily available and a good RF material, although it has a high helium permeabilty. A material which has both good RF characteristics and good vacuum characteristics is pure butyl rubber. O-rings made of butyl rubber are currently on order.

Another modification to the cyclotron is being fabricated. This modification will change the dee tips in such a way that the necessary peak voltage for all beams of interest will decrease by about 20%. Tests performed on the resonators showed that 150KV peak could be maintained over a 12mm gap. Over a 10mm gap the voltage decreases to 140KV. The central region now has a minimum gap of about 15mm with around 20mm gaps everywhere else. The new central region will have a minimum gap of 14mm with gaps of about 15mm everywhere else. To obtain the maximum energy from the cyclotron 150KV will be needed in lieu of 186KV. The maximum voltage necessary will be 163KV versus 204KV for beams requiring the highest voltages. This will reduce the necessary power form the RF system substantially and probably make the system more stable and reliable. However the tests described in the next section seem to indicate that we could, if necessary, develop the voltages necessary with the original central region.

9. Final Tests.

Initial tests at 27 MHz were successful even though the vacuum system was not running to its full capability. The lack of pumping speed was due to ongoing facility work which made it impossible to run the helium cryo-panels. The turbomolecular pumps and nitrogen cryo-panels, however, were run during the test to provide the minimum required vacuum of approximately 1 X 10-5 Torr. At 27 MHz we held steady at 140 kV and could run intermittently as high as 180 kV. At this point we were limited by continuous dee sparking and regular vacuum excursions.

The next test was run at a lower frequency to force a higher voltage on the insulator. The reason a lower frequency can achieve a higher voltage on the stem insulator is due to the resonator geometry (Fig. 14). The low voltage node of the resonator occurs at the shorted end, while the high voltage node occurs at the dee. Since the insulator is offset from the dee, it is located closer to the low voltage point at high frequencies (shorted end close to the dee). When we run at lower frequencies (shorted end extended away from the dee) the insulator is located proportionally closer to the high voltage node (the dee) and further from the low voltage point.

At 17 MHz it was difficult to go beyond 100 kV without creating sparks on the upper stem insulator. After inspection it was found that a small amount of debris had collected in the upper insulator area and was the main cause of the sparking. After cleaning this area, further tests revealed that we could run to 170 kV at 17 MHz without significant insulator sparking. This would correspond to a voltage at the insulator points of 103 kV peak. 17 MHz was chosen as the frequency for these tests because it is the frequency corresponding to the maximum insulator voltage required for cyclotron operation, i.e 160 kV on the dees and 97 kV on the insulator. When sparks did occur they were equally distributed between the upper and lower stems.

Although the seals now seemed to work admirably, an additional problem was found when running at high frequency and high voltages. Under these conditions, a large amount of heat is generated at the dees. These thermal changes are of large enough magnitude to shift the resonant frequency through a significant portion of the dee fine tuner's range. When sparks occur the dee suddenly cools enough to change the resonant frequency and the required position of the dee fine tuner. If the difference between the dee fine tuner position at the two extreme thermal conditions is large enough the RF will not restart after a spark, or requires constant re-tuning of the dee fine tuner during operation. We witnessed this behavior during a test at 27 MHz and voltages greater than 140 kV. Evidently there is not sufficient water cooling of the dees to allow for stable, long term runs under these extreme conditions. As pointed out in the previous section, the overall reduction of voltage requirements because of central region changes should also reduce this problem.

At this point we are satisfied with the performance of the RF system and the new seals. In the near future when the vacuum is sufficient, we will run long term high kV tests to determine the behavior of the seals under more rigorous conditions.

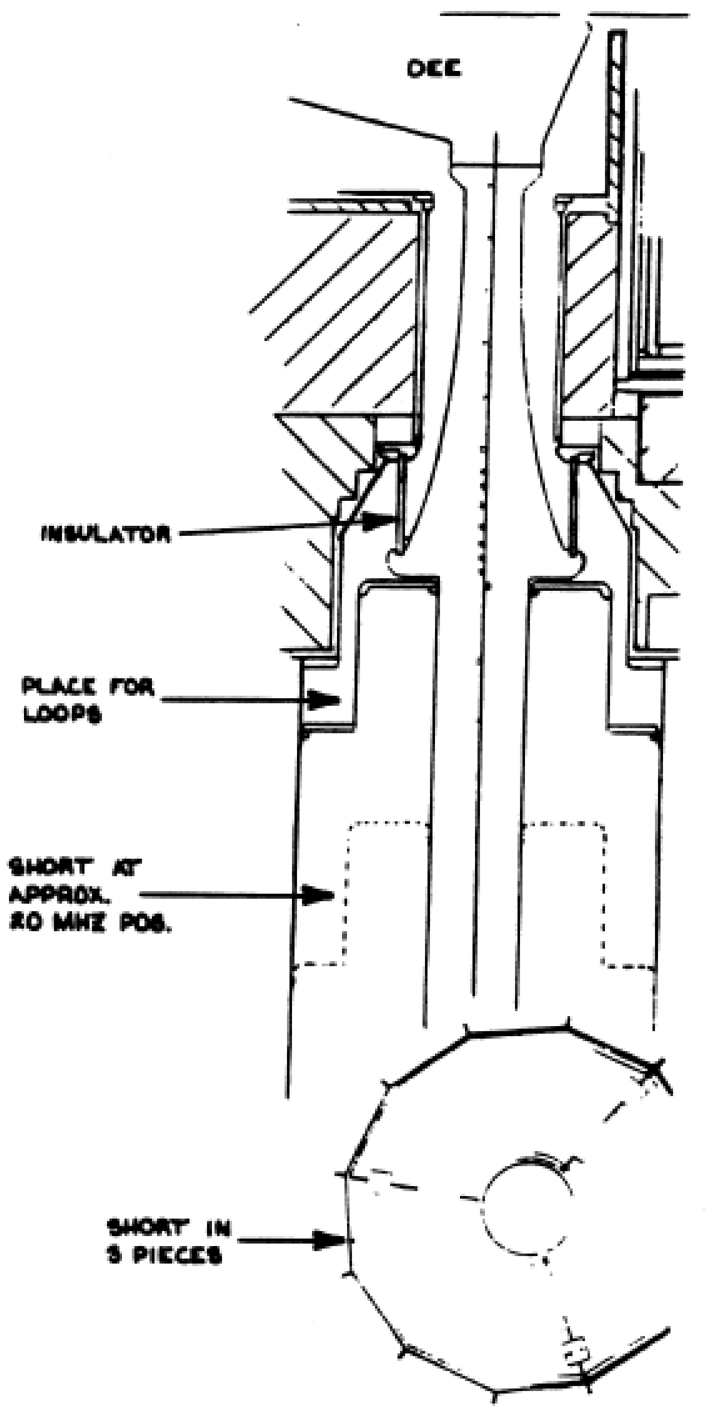


Fig. 14 Basic mechanical sketch of a lower quarter wave dee resonator.

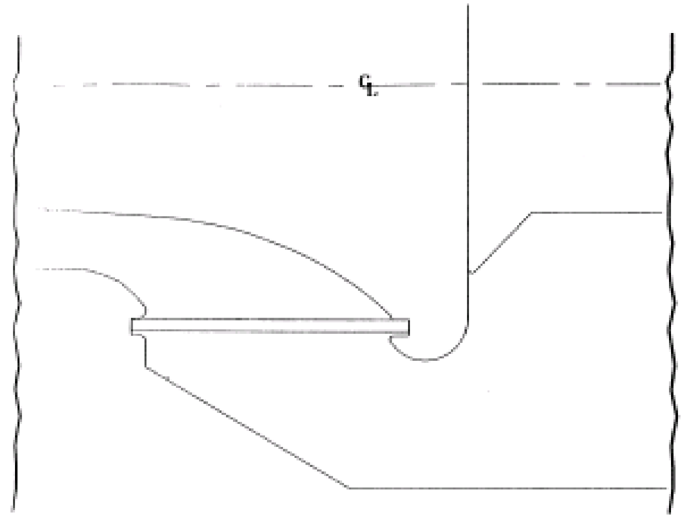


FIG. 15 INSULATOR SEAL WITH INDIUM

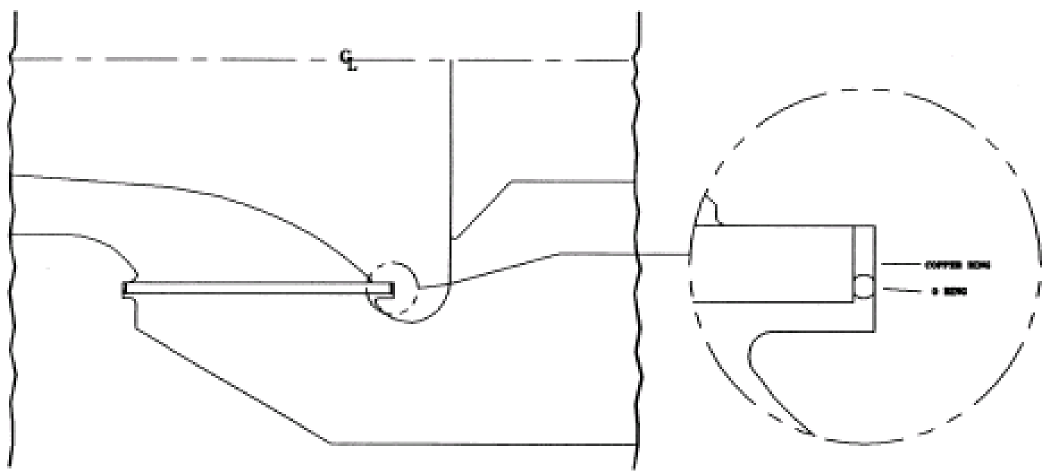


FIG. 16 INSULATOR SEAL WITH COPPER RING AND O RING

Appendix A. Hand Calculations.

The circuit of Figure 1 is reduced by resolving all stray reactances and transmission lines, giving the circuit of figure A1. Here we are ignoring all resistive elements.

The following notes are a summary of the calculations for circuit A1.

L is the equivalent inductance of the combination of TA1, TA2, TDL1, TDL2. TDL6 is ignored here as well as the neutralizing elements. All the stray, lumped inductive and capacitive elements are ignored. TA1 and TA2 both have a Z0 of 50 Ohms and are 6 inches long.

L ~ 2dZ0/c ~ .051 uH

TDL1 and TDL2 are each 139 Ohm lines with a variable length of 1.5 - 42 inches.

L2(9 MHz) ~ 2dZ0/c ~ .989 uH

L2(27 MHz) ~ 2dZ0/c ~ .035 uH

Adding L1 and L2 we get:

Lmax = L(9 MHz) ~ .989 + .051 = 1.04 uH

Lmin = L(27 MHz) ~ .035 + .051 = .086 uH

This tells us that the equivalent lumped tuning inductance may vary as :

L = 0.086 - 1.04 uH.

We want to reach 2000 Volts peak swing at the driver anode to correspond to 460 Volts peak at the final grid. A divider which achieves this is formed by CAG and the 800 pF equivalent capacitance of the transmission lines that represent the final tube electrodes.

Vg/Va = 460/2000 = 0.23

CAG = 800 (Vg/Va)/(1 - Vg/Va) ~ 240 pF

We then round this off to 250 pF for extra drive.

CAG = 250 pF

The circuit is now redrawn and further reduced in figures A2-A4.

In the reduced version we set CA so that the resonant frequency occurs at 9.25 MHz for L = 1.04 uH. Note that 9.25 MHZ was chosen as the absolute tuning minimum after several iterations of testing and simulation. That procedure is not indicated in this note since it was a rather lengthy and disjoint process.

CA + 183 pF = 1/(1.04 w2) which implies,

CA = 285pF - 183pF = 102pF ~ 100 pF

With this choice of CA we would like to determine whether a resonance of 27 MHz is achievable given the known range of L.

fmax = 1/(2 LC) = 1/ (0.086uH * 283 pF)

fmax = 32.26 MHz implying that 27 MHz is indeed within our tuning range.

In reality CA will be varied, both in the field and on the computer, in an effort to achieve resonance at the desired frequency. The value of CA will most likely be smaller due to ignored stray reactances, both known and unknown.

The circuit is redrawn in figure A5 with all the values that have been determined. The proper value(s) and placement(s) of resistive components will be found through computer and field analysis. These numbers will be determined to optimize stability and to adjust driver power requirements within acceptable limits.

In figure A6 the assumed neutralization bridge is shown. This is further simplified by ignoring L and CA, combining 250 pF and 34 pF in series and then combining this in parallel with 800 pF. The result is shown in Figure A7.

Assuming a drive signal between the plate and cathode, we require Vg = Vx so that the driver resonator elements between the grid and point X are unaffected.

1/(1 + 830/4) = 1/(1 + 1000/CN) which implies that

CN = 4.82 pF

Fig. A1

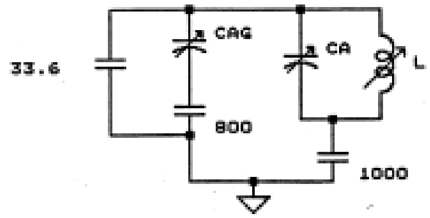


Fig. A2

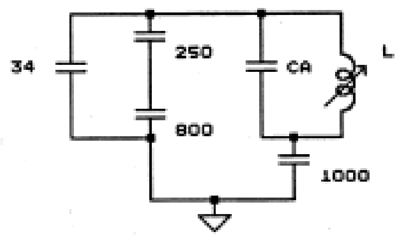


Fig. A3

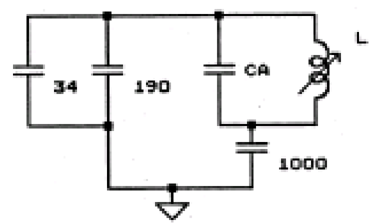


Fig. A4

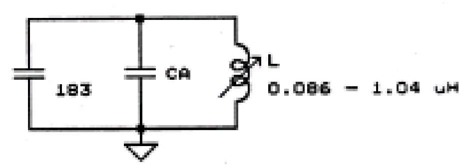


Fig. A5

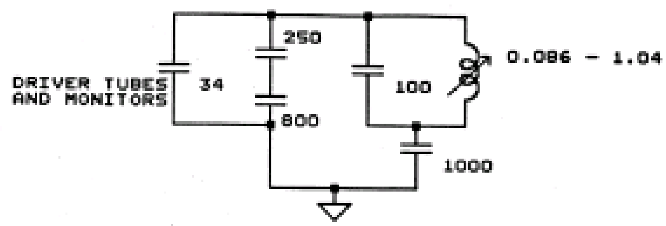


Fig. A6

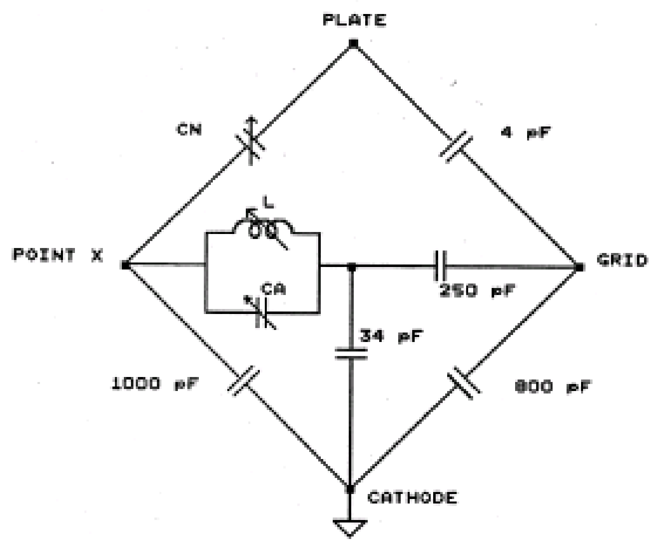
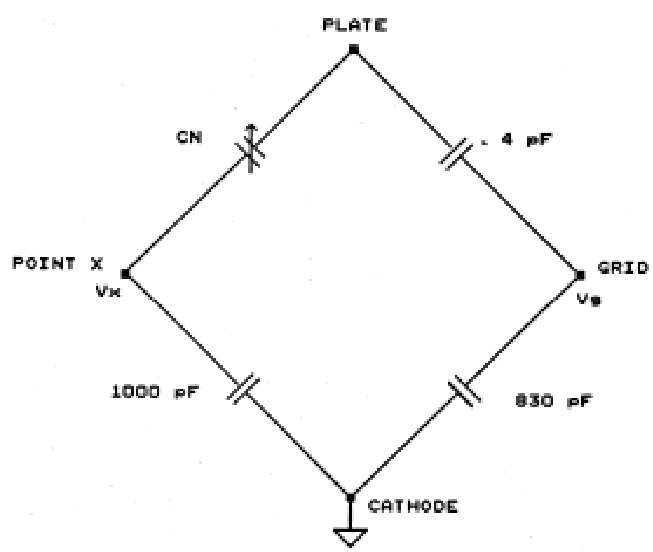


Fig. A7



Appendix B. Neutralizer Bridge

The following is a brief discussion of single-ended grid neutralization as it applies to the K1200 transmitter. The neutralizer can be effectively described in terms of an impedance bridge. Figure B1 shows this equivalent bridge circuit using the same element names as the circuit diagram of Figure 1. The elements are described here for clarity:

Cn Neutralizing capacitor (4.0 - 5.0 pF).
CFB5 Effective tube plate-grid capacitance (4 pF).
CGC Parallel combination of the grid-cathode and grid-screen capacitances.(800pF) This capacitance is represented in Fig. 1 by the transmission lines of the screen electrode and the grid electrode.
NOTE: The screen is grounded through a .05 uF capacitor.
CG Fixed bridge element (1000 pF).
CA Driver resonator capacitor (70 pF).
LDL Equivalent inductance of driver resonator transmission line.
CAG Coupling capacitor (250 pF).

Given a voltage applied across the plate - cathode, we want equal voltages appearing on the grid and point X. Equating the voltages on the grid and point X is done as in any impedance bridge; Cn is varied until the ratio Cn/(Cn + Cg) is equal to the corresponding ration on the other side of the bridge, CFB5/(CFB5 + CGC).

The power delivered from the grid to the plate will be positive for any frequency below the neutralization point, i.e. the point where the bridge is perfectly neutralized. This is referred to as a region of 'over-neutralization'. For frequencies above the neutralized point power will be delivered TO the grid from the plate, which corresponds to self-oscillation. This region is termed 'under-neutralized.' For the K1200 transmitter we set the perfect neutralization point at 27 MHz so that the entire operating range is over-neutralized.

Fig. B1 Grid Neutralization Bridge