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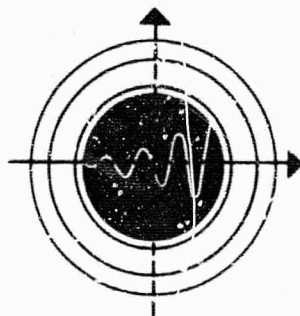
RESEARCH AND DEVELOPMENT
ON
L-BAND CROSSED-FIELD AMPLIFIER CHAIN

by
A. Wilczek

Report No. 9
Final Report

Contract Nr. AF 30(602)-2533
ARPA Order No. 136.61

Rome Air Development Center
Research and Technology Division
Air Force Systems Command
United States Air Force
Griffiss Air Force Base, New York



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ABSTRACT

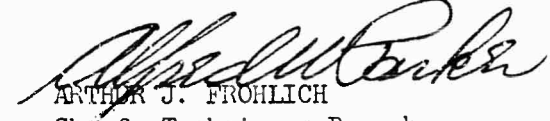
A crossed-field amplifier for use in an L-band phased array radar has been developed. The SFD-209 is designed to be stacked on half-wavelength centers in a permanent magnet array. The tube will generate peak power up to 150 kw with 16 db gain at an efficiency of 50% over a 100 Mc band centered on 1200 Mc, with exceptional phase characteristics. Average power of 1.5 kilowatts and operating life in excess of 5000 hours have been demonstrated.

The SFD-209 operates from a dc power supply with very low modulator power. Cold cathode operation with RF turn on and control electrode turn off is used. The duty cycle may be changed instantaneously by changing repetition rate and/or pulse width with no change in modulator power on a per pulse basis. Pulse widths from 0.1 μ sec to 600 μ sec and repetition rates as low as 1 pulse per minute have been demonstrated. The ability to pulse code a CW input signal and the ability to amplify a complex signal have been demonstrated.

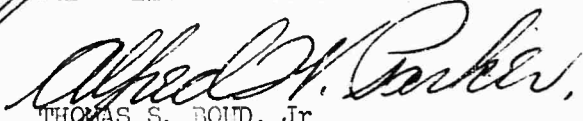
PUBLICATION REVIEW

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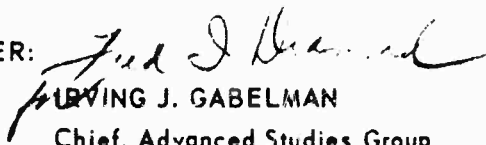

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1.0 INTRODUCTION

The purpose of this program was to develop a crossed-field amplifier suitable for use in an L-band phased array radar. There is no existing array radar for which these particular tubes were destined, although L-band experimental array radars have been built. However, a tube development of this sort must proceed well in advance of system development so that the system designers will have detailed advance knowledge of the amplifier properties. At the outset of the program, the amplifier was to be operated with a gain of 35 db with an efficiency in excess of 40% over a 10% bandwidth. The peak output power was to be 1 Mw for a 2 μ sec pulse with an average power output of 2 kw.

The intended application in an array radar placed some difficult requirements on the overall design. Some of the more stringent requirements were the following: the small cross-sectional area, which was to be sufficiently small so that the tubes could be stacked in a matrix with half-wavelength spacings; a relatively high degree of linearity and reproducibility of phase characteristics; and low cost per hour of operation for a large quantity of tubes. In the development program, emphasis was placed on those aspects which would lead to small diameter, reproducibility, and long life. In addition, consultations with array designers led to the conclusion that substantial contributions could be made by considering the effect of the amplifier design on the power supplies, modulators, etc., required to operate the amplifier. The features which lead to important simplification in system equipment were incorporated in the tube design during the course of the development program.

At the start of the program, it was decided that a forward wave circuit would best fulfill the many necessary requirements for application in a phased array. The forward wave circuit used consists of half-wavelength bars coupled by alternate capacitive strapping. A

description of this circuit is given in Appendix I. The use of a forward wave circuit in the amplifier makes it possible to operate with a constant anode voltage over the required band. This is possible with a forward wave circuit but not a backward wave circuit for the reasons indicated in Figure 1. Once the possibility of constant voltage operation was established, it became worthwhile to consider operating the amplifier from a dc voltage source with the space charge controlled by a low power control electrode. The sequence of events is illustrated in Figure 2. The input RF signal turns the amplifier on and it amplifies for the duration of the input RF signal pulse. At the end of the RF signal pulse, a control pulse is applied to an electrode located near the cathode and between the RF input and output coupling sections. The control pulse interrupts the circulating electron cloud, collects the circulating electrons, and thus causes the amplifier to turn off. The control electrode consumes very little power and requires an applied pulse amplitude less than the anode dc voltage, a pulse length of a few tenths of a microsecond, and a pulse shape which is not critical. Thus, the high power modulator usually required with a crossed-field amplifier is replaced by a dc power supply and a comparatively low power control electrode modulator, which represents a great savings in power supply equipment.

The mechanical design of the tube was affected by the limitations on the cross-sectional area available in a phased array with half-wavelength spacings of tubes. These limitations forced the use of straight vertical bars; i.e., the bars could not be bent into a radial plane in the inductive section as is sometimes done. This in turn meant that great care had to be taken in the design of the magnetic circuit to insure field uniformity over the relatively long interaction area. A further consideration was the thermal expansion of the bars in the axial direction as they are heated during operation at high average power levels. If the bars were firmly fixed at

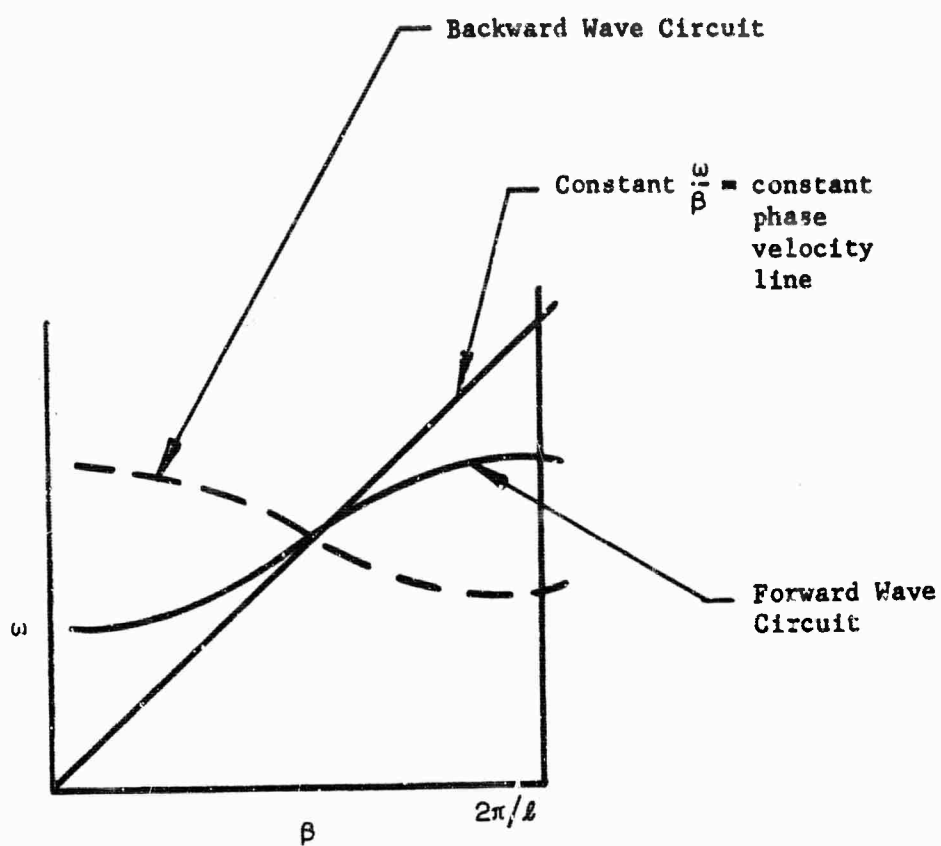


FIGURE 1 COMPARISON OF THE ω - β PLOTS FOR FORWARD AND BACKWARD WAVE CIRCUITS

It is seen that synchronism is maintained over a much wider band of frequencies with a forward wave circuit than with a backward wave circuit.

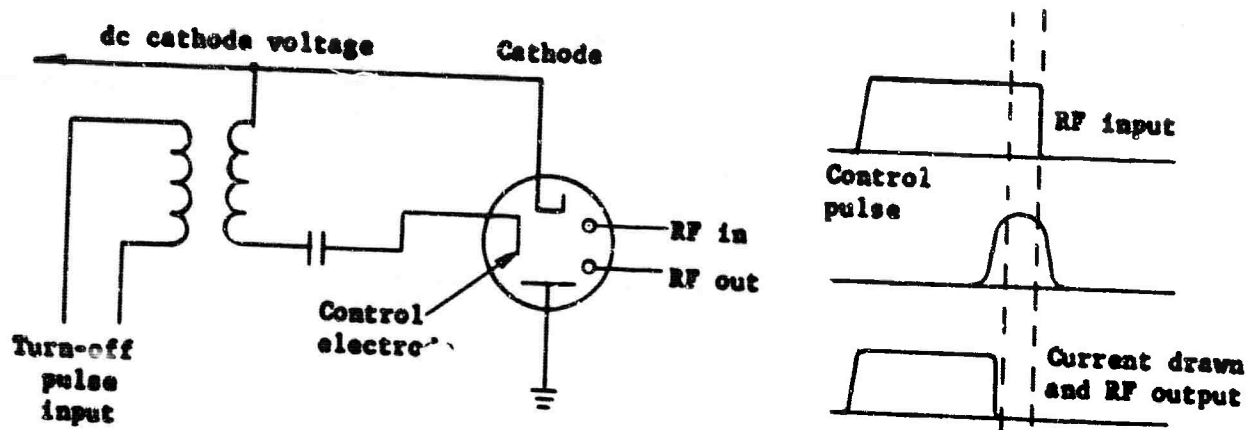


FIGURE 2 APPLICATION OF THE CONTROL PULSE FROM AN EXTERNAL SOURCE

The control pulse must last for at least 30 nsec after RF input has terminated. This arrangement has been shown to work using the SFD-202

both ends so that thermal expansion could not take place, the bars would bow. A little simple trigonometry shows that the amount of bowing is more than ten times the axial expansion.

A third mechanical requirement imposed by the strict reproducibility requirements called for in array applications is that the design be one in which critical dimensions may be accurately reproduced. Throughout the design stage, special precautions were taken to insure compliance with these requirements.

The initial decision was to meet the requirements of this program with two crossed-field amplifiers in tandem - a 20 db gain driver tube, giving an output of 100 kw, and a 10 db gain power tube, giving an output of 1 Mw. This division was made necessary by restrictions imposed by the electronics of the tubes and the desire to optimize efficiency (by low gain) in the output stage. As it turned out, the driver output level is close to that being considered for use in the next generation of phased array systems. The results of the driver program will be useful in the near future, while the complete package development may be desired at a later date. Accordingly, more emphasis was placed on the driver tube development program, while the power tube development program was carried along with reduced emphasis.

In this development program, emphasis was placed on those aspects of the tube design which will lead to long life and low unit cost for tubes produced in large quantities. Value considerations were thus an important part of the basic design approach chosen for the tubes. Figure 3 shows a layout drawing of the driver tube and Figure 4 shows a layout drawing of the power tube. Some of the areas studied as part of the value engineering program are indicated on these drawings. Not all of the desired modifications could be incorporated within the time and scope of the program; thus, our studies have also led to a list of recommendations for future changes which should reduce tube cost in quantity production.

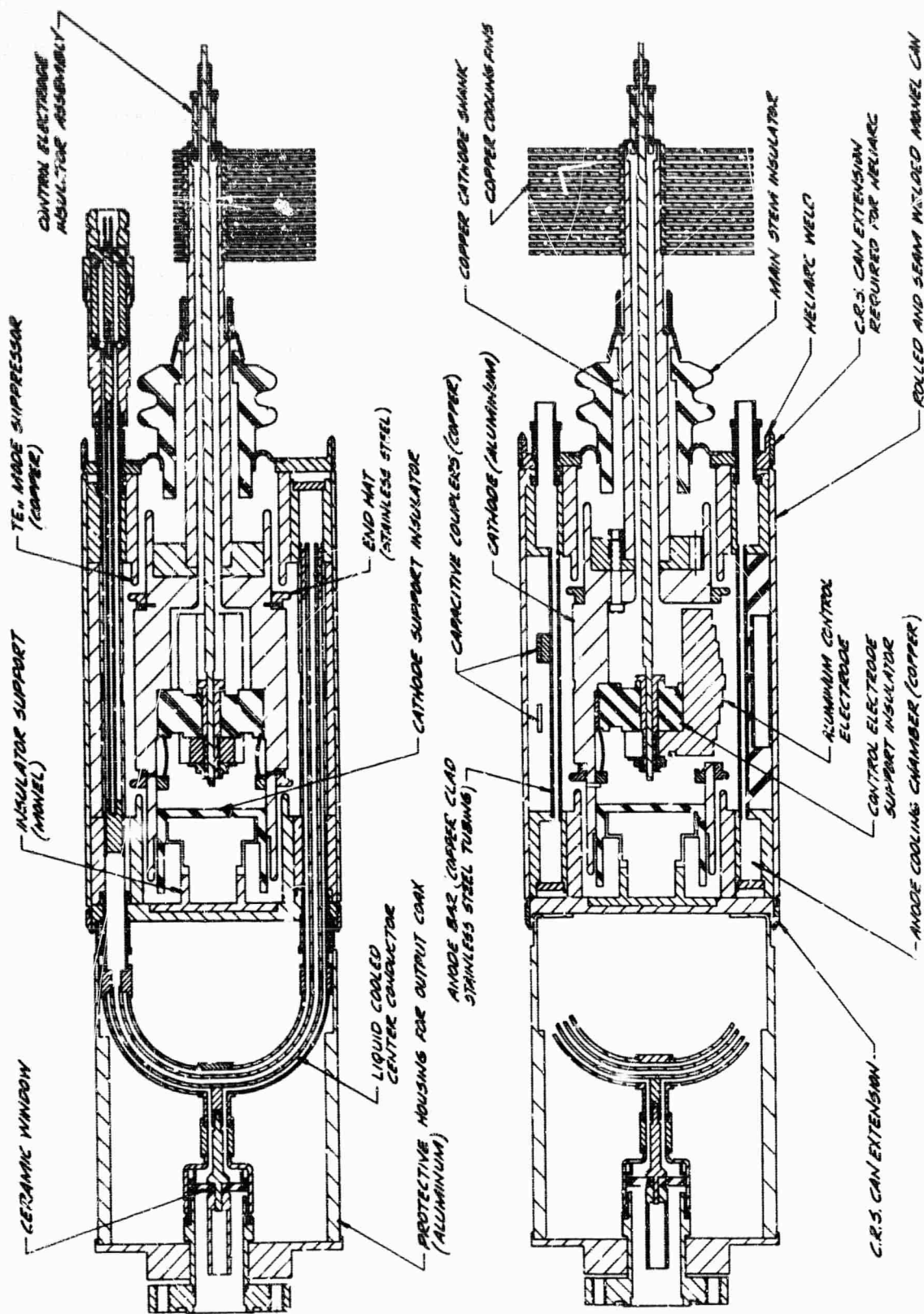


FIGURE 3 LAYOUT DRAWING FOR DRIVER TUBE

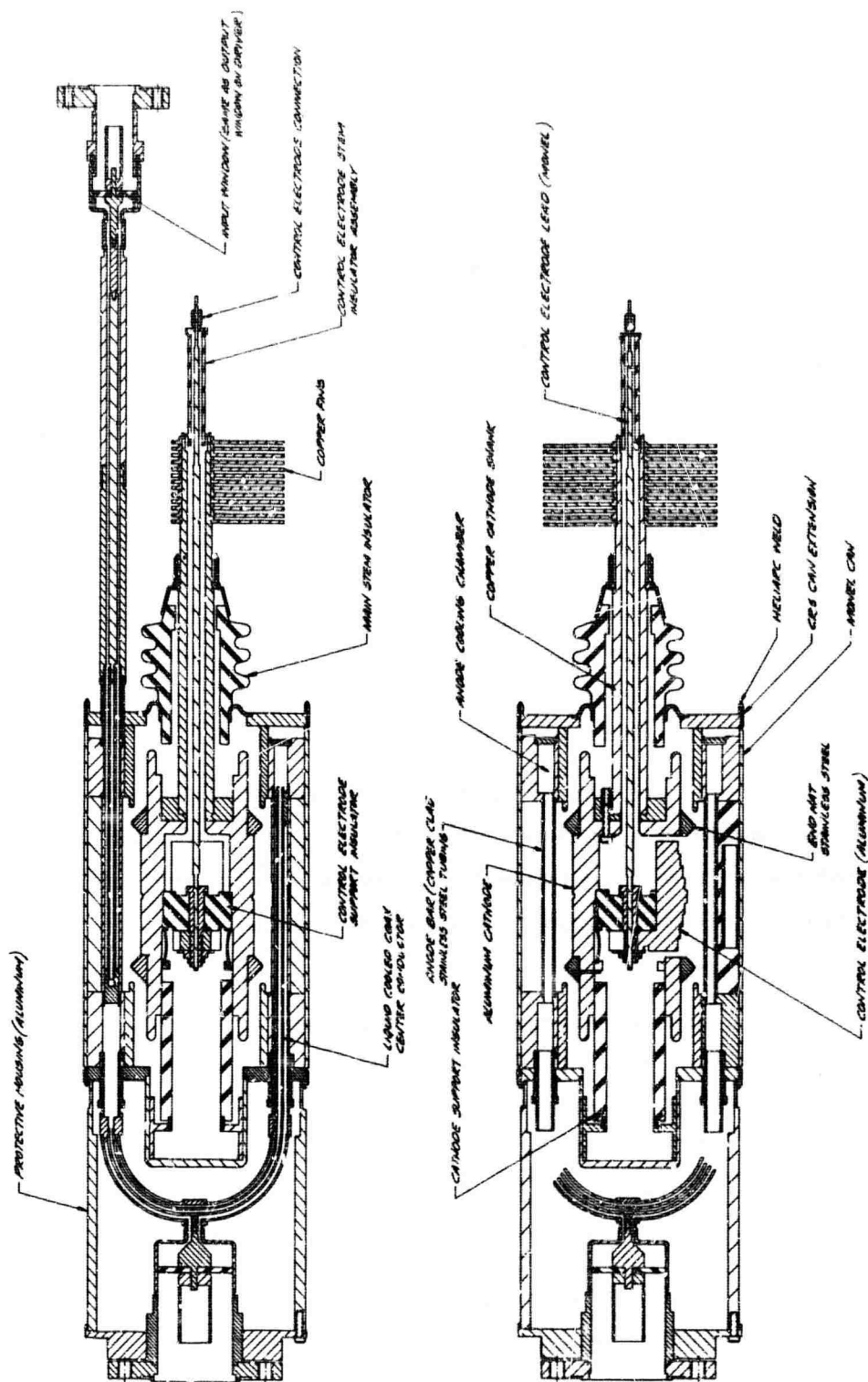


FIGURE 4 LAYOUT DRAWING OF POWER TUBE

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Some of the more significant achievements in terms of maximum performance capabilities that have been demonstrated are the following:

- (a) dc cold cathode operation with low power control
- (b) operating voltage less than 10 kv
- (c) gains of approximately 20 db
- (d) 10% bandwidth
- (e) efficiencies of approximately 60%
- (f) peak power in excess of 200 kw
- (g) average power of 1.5 kw
- (h) pulse widths from 0.1 μ sec to 600 μ sec
- (i) operating life in excess of 5000 hours
- (j) phase sensitivity of 2° for 1% change in voltage
- (k) 12° maximum phase deviation from linearity with input and output VSWR's of 1.5:1
- (m) ability to handle complex signals

Not all of the above mentioned performance characteristics have been achieved simultaneously. However, many of these are well beyond the initial requirements. A typical driver amplifier has demonstrated the following capabilities simultaneously.

- (a) dc cold cathode operation at constant voltage
- (b) operating voltage less than 10 kv
- (c) gain of approximately 15 db
- (d) 10% bandwidth
- (e) efficiencies of 50% over the 10% bandwidth
- (f) peak powers of 100 kw to 150 kw over the 10% band
- (g) average power in excess of 200 watts with a circuit and cathode capable of 1500 watts average
- (h) operating life well in excess of 2500 hours

It must be pointed out that not all of the improvements had been incorporated in the tubes at the time the life tests were begun. Consequently information obtained later in the program could not be incorporated in these tubes.

The results obtained in the power tube program were the following.

- (a) dc operation with low power control
- (b) operating voltage under 30 kv
- (c) gain of 10 db
- (d) 100 Mc bandwidth centered on driver band
- (e) peak power of 750 kw
- (f) average power over 1 kw

The above-mentioned performance represents the best achieved but not necessarily obtained simultaneously. Several tubes were constructed and tested with each one indicating a problem area. Most of the problems observed in the power tube were also observed in the driver tube. At this point, it was decided to suspend effort to insure the construction of a group of final design driver tubes in which the observed problem areas had been successfully corrected. The measurements on the power tubes indicated that a 750 kw output at 60% efficiency and 10 db gain would be obtained with dc operation when the problem areas are corrected.

The use of a forward wave circuit, cold cathode, dc operation with RF turn on, and relatively low power control has demonstrated versatility that is not usually found in other types of crossed-field amplifiers. The control electrode technique offers the capability of pulse coding and operation at continuously variable pulse lengths over a very wide range of duty cycles. It should be noted that the control pulse interrupts the recirculation of electrons and turns the tube off even in the presence of an RF input signal. With the electron recirculation interrupted by the control electrode, there is insufficient circuit length for space charge to build up and the resulting current

is negligible. Control electrode pulsing can thus be used to chop up a long input pulse into a coded pulse. The control electrode pulse amplitude in the 150 kw tube is about 5 kv with a peak current of about 9 amps drawn during the tube turn off. This current reaches a maximum for a control pulse voltage of 2-3 kv and then drops to zero even though the control voltage is maintained on the tube. The time required for turn off is in the order of 0.1 μ sec, depending upon the characteristic rise time of the control pulse. The volt-ampere characteristic of the control electrode is shown in Figure 5.

The length of the amplified pulse is determined by the delay between the start of the RF signal pulse amplification and the application of the control pulse. The pulse length can be varied, at low levels, by delaying the synchronizing pulse to the control modulator. The only qualification is that the control voltage must be maintained until the RF drive signal is removed from the amplifier. The pulse width has been varied from approximately 0.1 μ sec to 600 μ sec. The upper limit on pulse length in this case was imposed by the available storage capacity in the dc supply for the amplifier.

Aside from the obvious advantage of eliminating the heater power supply and the ability to operate from a dc supply, cold cathode operation offers several other attractive features. Cold cathode operation with RF turn on offers instant "warm-up" time in that no preheating or standby power is required. Cold cathode operation offers inherently long life; indeed, operating life in excess of 5000 hours has been demonstrated with no sign of imminent failure. Inasmuch as the cathode is operated "cold," there is no heater reduction schedule to be observed so that essentially instantaneous changes in duty cycle are easily accomplished by varying pulse width and/or repetition frequency. Repetition rates as low as approximately one pulse per minute have been demonstrated. The upper limit on the repetition rate is the maximum duty cycle. The maximum duty obtained during this period was in excess of 1%. The duty limit was imposed by the average power capability of the coaxial line components used in the test set.

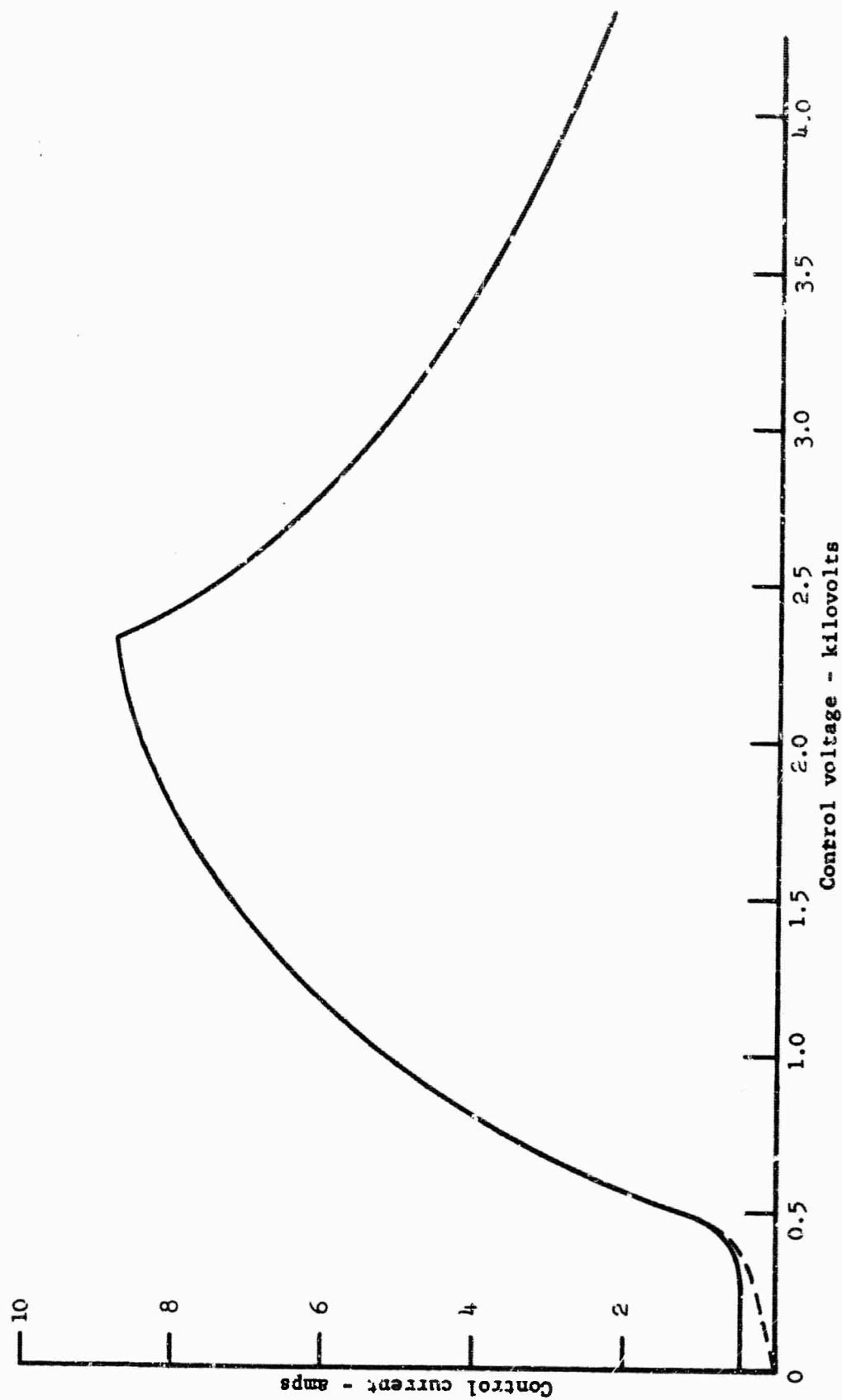


FIGURE 5 THE V-I CHARACTERISTICS OF THE CONTROL ELECTRODE

Since we do not have a long pulse L-band driver available, it was necessary to conduct the simultaneous tests of high average power and long pulse capability using a feedback loop technique. Figure 6 shows a block diagram of the system employed. The magnetron driver, which is also used for most of our RF characterization, is employed as a trigger signal in this long pulse test. The pulse length from this magnetron is 2.5 μsec . During this initial trigger interval, the crossed-field amplifier is driven directly by the magnetron. At the end of this trigger interval, the amplifier continues to be self-excited by the feedback loop. At the end of the desired long pulse, the control electrode pulse is applied to the amplifier to turn it off. Once turn off has occurred, the amplifier waits for the application of the next magnetron trigger pulse without drawing current. In this manner, the output pulse length is controlled by the time delay introduced between the magnetron trigger pulse and the control electrode turn-off pulse. Figure 7 shows waveforms obtained for a 25 μsec pulse length. The initial amplification period of approximately 2.5 μsec may be clearly discerned. The total attenuation in the feedback loop was comparatively high; i.e., approximately 17 db. Therefore, during the self-excited interval, the amplifier was operated at 17 db gain. This resulted in an effective input power between 1 kw and 2 kw peak (17 db below measured output) - a comparatively low value for this tube. This in turn resulted in a lower output power during the self-excited period as compared with the output when driven by the magnetron during the initial trigger interval. Figure 8 shows the pulse shapes obtained with a 200 μsec pulse length. In this case, the initial 2.5 μsec trigger interval is barely visible. The pulse droop is between 0.5 db and 1 db during the 200 μsec period. These tests were run with a 24 microfarad energy storage capacitor. The observed droop is approximately what would be predicted from a calculation of the voltage droop on the energy storage capacitor and a

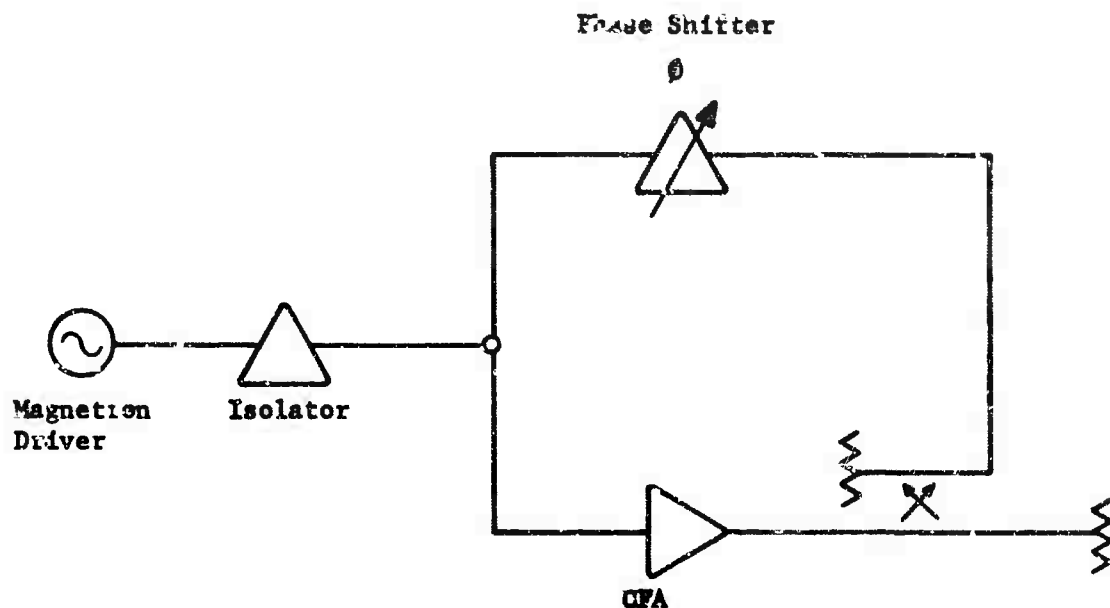


FIGURE 6 FEEDBACK LOOP USED IN LONG PULSE AND HIGH AVERAGE POWER TESTING

The magnetron triggers the crossed-field amplifier on. When the magnetron pulse ceases, the crossed-field amplifier continues to oscillate at a frequency determined by the feedback loop until the control pulse turns the crossed-field amplifier off.

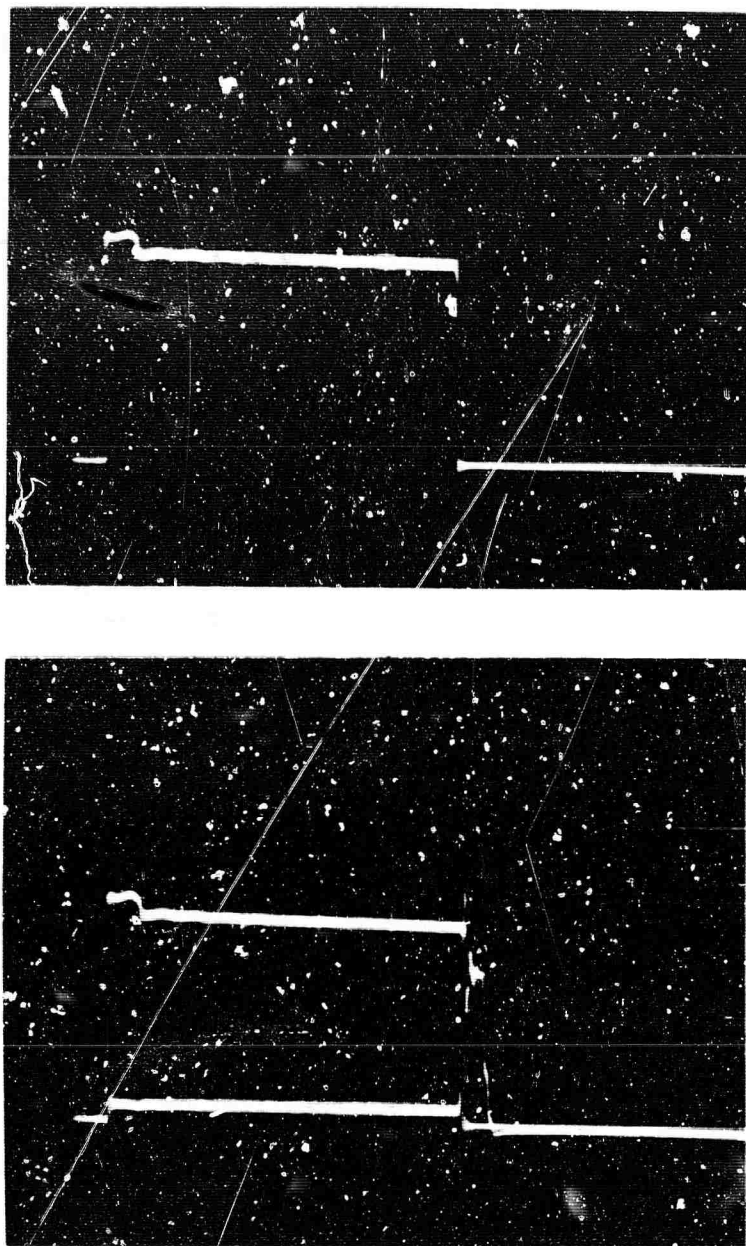


FIGURE 7 AMPLIFIED RF PULSE AND CONTROL PULSE (top) AND RF PULSE ALONE (bottom)

Tube triggered by 2.5 μ sec magnetron pulse and allowed to oscillate after cessation of drive pulse under influence of feedback loop until control pulse is applied. Effective RF drive from feedback loop is only ≈ 2 kw peak. This causes sudden drop in RF output when magnetron turns off. Total amplifier pulse length is 25 μ sec.

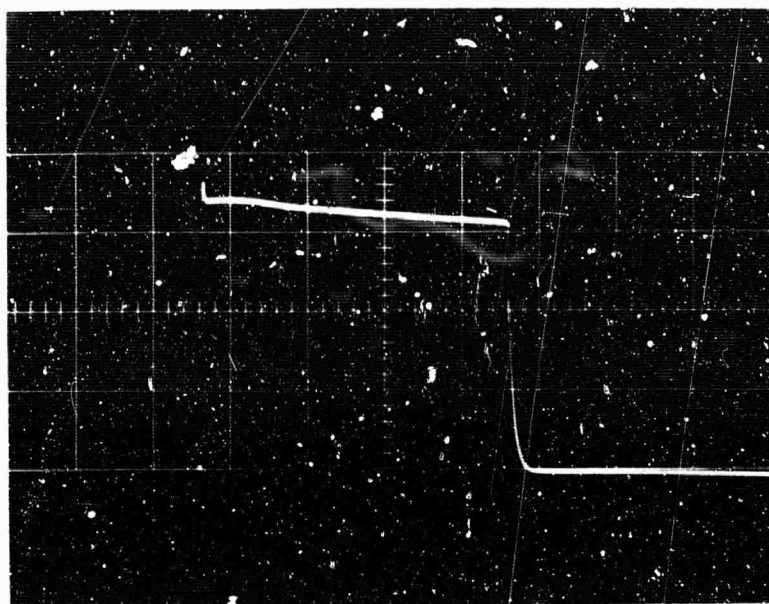


FIGURE 8 RF PULSE SHAPE FOR 200 MICROSECOND PULSE

Crossed-field amplifier operated on feedback loop.
Spike on leading edge represents the 2.5 μ sec
magnetron drive interval

knowledge of the dependence of output power on operating voltage as determined from the performance charts.

In these tests, the phase length of the feedback loop was sufficiently long so that self-excitation at several different frequencies within the operating bandwidth of the crossed-field amplifier would have been possible. The adjustable phase shifter was positioned so that one of these regenerative frequencies coincided with the frequency of the magnetron trigger signal. When this was done, the tube continued to operate at the trigger frequency after the termination of the magnetron drive signal. This arrangement hints at the possibility of a sort of "digitalized frequency operation." In such an arrangement, an electrically long feedback loop would be employed so that regeneration would be possible at a number of frequencies within the operating amplifier bandwidth. The frequency at which oscillation actually occurred could then be chosen by the proper frequency of the magnetron trigger signal. At the termination of this trigger signal, the amplifier will continue to oscillate at the regenerative frequency closest to that of the applied trigger signal.

Using this feedback loop technique, average power capability tests were run at a pulse length of 200 μ sec up to an average power of 1500 watts. The power limit reached was imposed by the average power capability of the coaxial line components used in the feedback loop. For example, the directional coupler used as the output power sampler had a rating of 200 watts. On one occasion, heating of this coupler was sufficient to unsolder a connection. The tests at 1500 watts of average power were run for a comparatively short time (less than half an hour) because of the limitations of the coaxial components. However, tests at approximately the 1100 watt average power level were run for several hours on several successive days. On the basis of these tests, we are confident that the present tube design may be rated for at least 1 kw of average power, and we suspect

that a rating of over 2 kw could be justified by further average power testing. During all of these tests, the tube operated cold to the touch and a very moderate air flow over the cathode cooling fins was required to maintain them at a temperature of approximately 60°C.

The achievements of this program are not only the development of a tube to meet a general set of specifications, but that several new and significant concepts, which have advanced the state of the art of microwave tubes, were demonstrated. The basic accomplishments of this program are

1. Forward wave circuit with all the necessary properties for practical tube use
2. Cold cathodes capable of starting and sustaining emission required for amplification with the drive power levels available
3. Technique of operating a pulsed amplifier from a dc power source to eliminate high power modulation and increase the versatility of the amplifier
4. Amplifier comprising the previous three properties and also capable of array matrix spacings on $\lambda/2$ centers
5. Techniques which are clearly extrapolable to other frequency bands, circuit structures, and power levels.

2.0 DESCRIPTION OF THE SFD-209

The tube developed during this program is a 100 kw to 150 kw crossed-field amplifier intended for phased array applications. Figure 9 shows a photograph of this tube. It has a diameter of slightly under 4 inches, was designed for stacking on approximate half-wavelength spacings, and is intended for use with a fixed permanent magnet matrix into which the individual tubes will be inserted.

The amplifier operates over the band of 1150 Mc to 1250 Mc, develops a peak power of 100 kw to 150 kw, controllable by anode voltage, and requires a 5 kw RF input drive signal. This tube was developed for a 0.002 duty requirement; however, because it was known that the eventual requirements would call for higher duty factors, a liquid-cooled anode capable of dissipating several kilowatts of average power was used. The tube has been tested to 1500 watts average power output at a 200 μ sec pulse length.

Life tests have been conducted on the SFD-209 amplifier at the 200 watt average power level. One tube operated for over 5500 hours and another for 2900 hours with no sign of failure. Several other tubes were life tested for over 1000 hours. A total of 13,023 hours of life testing has been accumulated on this program.

2.1 Principles of Operation

Figure 10 shows a schematic diagram of the internal structure of the SFD-209 crossed-field amplifier. Listed on the diagram are some of the principal features which, in combination, distinguish this type of crossed-field amplifier. The tube uses a secondary emitting continuous cathode. The electron cloud is reentrant; i.e., electrons are allowed to circulate around the tube without interruption. This means that the electron focusing mechanism, the electron trajectories, and the energy interchange mechanism are similar to those in a magnetron.

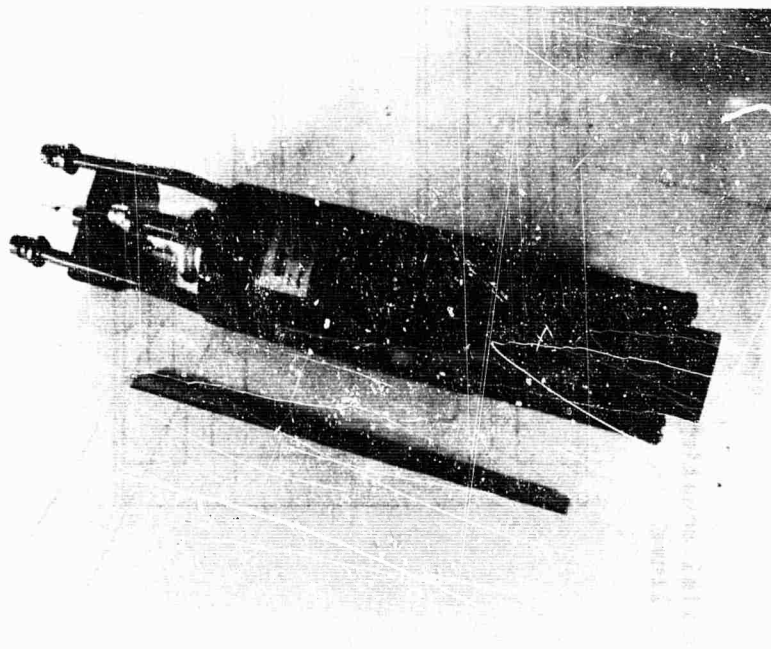


FIGURE 9 THE SFD-209 CROSSED-FIELD AMPLIFIER

Connections for dc voltage, control voltage, coolant and RF input are on the left end. RF output connector is centered on the right end of the tube

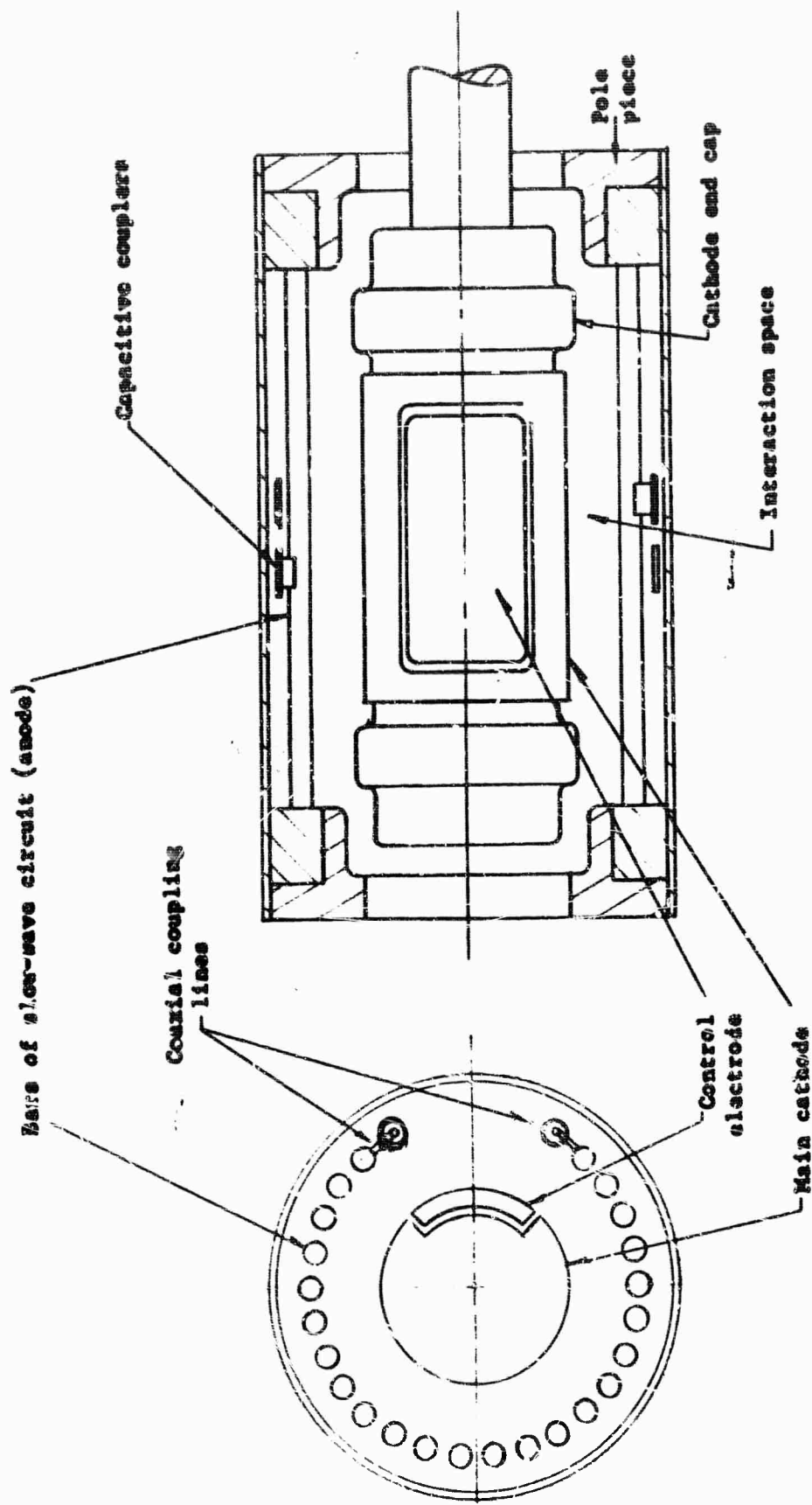


FIGURE 10 SCHEMATIC DRAWING OF MAJOR PARTS OF A TUBE OF THE SFD-209 TYPE

As a consequence, high magnetron-like efficiencies are obtained. The anode circuit is, however, a propagating structure instead of a resonant circuit as is the case in a magnetron. Amplification is thus obtained by a crossed-field traveling wave interaction between circuit wave and space charge cloud.

The anode slow wave circuit is constructed from half-wavelength bars which are capacitively coupled to form the propagating circuit; it is described more fully in Appendix I. This type of circuit has a forward wave characteristic. By this we mean that its phase and group velocities are in the same direction. This is also the direction in which the electrons rotate in the tube. Crossed-field amplifiers may also be constructed with backward wave circuits in which the phase and group velocities are in opposite directions. Such backward wave circuits have seen considerable application in crossed-field amplifiers. A significant difference between forward wave and backward wave circuits is that a forward wave circuit can have a constant phase velocity, whereas a backward wave circuit inherently has a variable phase velocity as a function of frequency. Each of these characteristics has its own advantages. For an array radar tube, the constant phase velocity characteristic of the forward wave circuit has the important advantage of permitting operation at a fixed voltage over the full bandwidth of the tube. This in turn makes operation directly from a dc power supply with control electrode modulation possible.

Figure 11 shows a photograph of the type of circuit used. In our present tubes, the anode bars are hollow and a liquid coolant is passed through them. Two rows of capacitive couplers may be seen near the center of the circuit. If these couplers were solid rings, they would represent alternate strapping such as is used in magnetron circuits. Such circuits would be backward wave. By cutting the straps between connection points, we obtain a forward wave characteristic. The anode bar diameter for the 150 kw tube is 0.135 inch and 0.250 inch for the 750 kw tube.

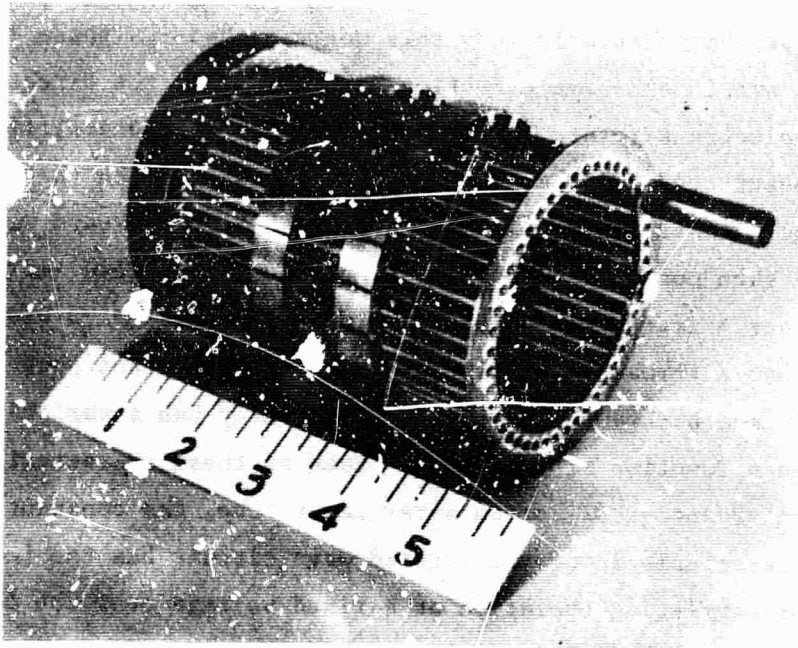


FIGURE 11 PHOTOGRAPH OF THE ANODE CIRCUIT

This version of the circuit used solid anode bars, but current versions use hollow fluid cooled bars. The coaxial input line protrudes to the right. The coaxial output line is seen beneath it. The three bars between the coaxial lines form the drift space anode. These bars are without capacitive couplers so that RF energy does not propagate across the drift space

The liquid cooling of these anode bars gives the anode circuit a power dissipation capability of many kilowatts. The ruggedness of this anode structure is evident from the photograph. Crowbar protection of this tube operated from a dc power supply will not be necessary.

A drift space is incorporated between the input and output coaxial lines in Figure 10. Heavily bunched electron spokes enter the drift space from the output end of the tube. Once in the drift space, they debunch under the influence of their own space charge since there are no RF circuit fields there to maintain the bunching. Electrons are thus returned from the output to the input with the RF modulation removed from them so that we do not incur RF feedback from the output to input. The incorporation of the drift space and the consequent debunching of the returning space charge are important features of this crossed-field amplifier. If RF feedback were present, it would be necessary to design the tube to operate as a regenerative amplifier, by limiting the electrical length of the circuit so that the feedback remained positive over the 10% bandwidth. This would limit the number of circuit sections to 17. By incorporating the drift space and eliminating feedback, this limitation for a given percentage bandwidth is removed. Thus 36 resonators were chosen, this being the maximum number which can be squeezed within the diameter restriction. In tubes with no stringent restriction on diameter, a considerably greater number of resonators could be used. Increasing the number of resonators is advantageous because it increases the area available for power generation. The removal of bandwidth limitations imposed by RF feedback opens the door to future increases in bandwidth to 15% or more by relatively minor modifications in the circuit structure.

Figure 12 shows a photograph of the cathode and control electrode sub-assembly. The control electrode, shown as the recessed

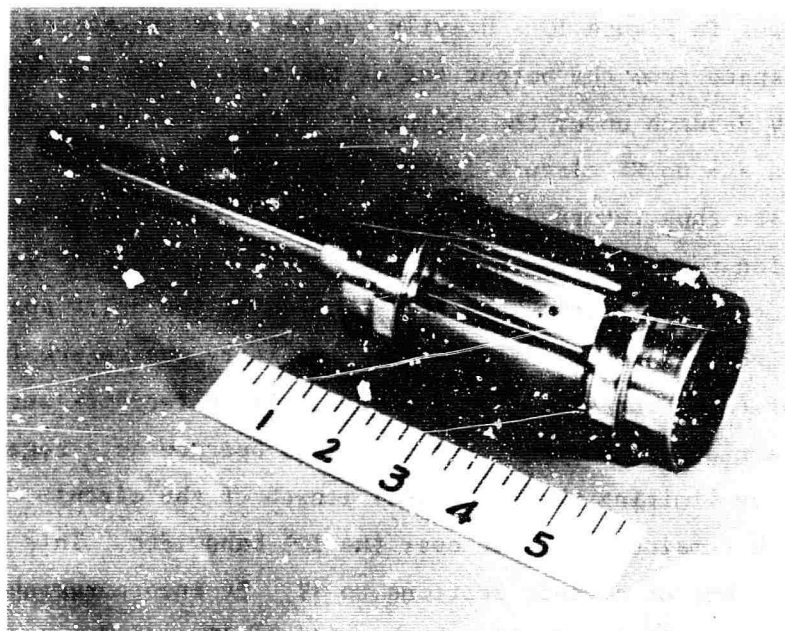


FIGURE 12 PHOTOGRAPH OF THE CATHODE ASSEMBLY

The control electrode is shown facing out from the paper. The main cathode operates entirely on secondary emission. Cathodes of beryllium copper and aluminum have been used. These cathodes may be operated cold and the tubes have no heaters

notch in the photograph, is located in the drift space and permits operation of the tube from a dc power supply as will be described in the next section. The cathode uses a metallic secondary emitter which is operated cold. No heater is required. The resulting structure is thus both rugged and inexpensive.

2.2 Operation from a dc Power Supply

An important feature of this tube is its ability to be operated from a dc power supply with low power modulation. This is obtained through the use of a purely secondary emitting cathode and the incorporation of a control electrode, which is shown in the drift space in Figure 10. Since there is no thermionic emission from the cathode, a dc voltage may be placed directly across the tube and, in the absence of an input signal no current will be drawn. The output from the tube, under these conditions, is essentially room temperature thermal noise. A few electrons will always be present in the interaction space as a consequence of ionization of the residual gas or other causes, but their number will be far too small to generate a measurable output in the interpulse period. However, when an RF input signal is applied to the tube, they will interact with it. Some of these electrons will gain energy from the RF wave and will be accelerated toward the cathode. This is the same mechanism by which magnetrons generate cathode back bombardment power. The electrons reaching the cathode give rise to secondary emission and thereby increase the number of electrons in the interaction space. Through a repetition of this process, space charge is built up and the tube turns on. Turn-on times are typically of the order of a few electron transits around the tube or a few tens of RF cycles. At L-band this means turn-on times of a few tens of nanoseconds. This time would scale down with wavelength in higher frequency tubes.

Once the tube has turned on, it becomes space charge limited so that the current drawn is not dependent on the secondary emission ratio of the cathode. To turn the tube off, it is necessary to sweep the space charge out of the interaction space. This is done by pulsing the control electrode positive with respect to the cathode. It then becomes a collector of electrons and stops their recirculation from output back to input. This prevents the build up of space charge and turns the tube off. The pulsing sequence used is shown in Figure 13. This figure also shows a schematic diagram of the control electrode circuitry employed. During the interpulse interval and the major portion of the amplification interval, the control electrode remains at cathode potential. To turn the tube off, the control electrode is pulsed positive at the end of the input drive pulse. It is held positive until the input drive has been removed. Once the drive pulse has been removed, the control pulse may also be removed without the tube going back into conduction. The tube will then draw no current until application of the next input RF drive pulse. It should be noted that the control pulse interrupts the recirculation of electrons and turns the tube off even in the presence of an RF input signal. With the electron reentrancy interrupted by the control electrode, there is insufficient circuit length for space charge build up and the resulting current is negligible. Control electrode pulsing can thus be used to chop up a long RF input pulse into a coded pulse. The control electrode pulse amplitude required to turn the 150 kw L-band tube off is about 5 kv. A peak control electrode current of about 9 amps is drawn during turn off. This current reaches a maximum as the control pulse voltage reaches 2-3 kv and then drops rapidly to zero even though the control voltage is maintained on the tube. The power required for the control modulator is thus very small. Lower control electrode voltages and currents have been demonstrated with a different control electrode configuration used in a higher frequency tube. Oscillograms of actual pulse shapes are shown in Figure 14.

Crossed-Field Amplifier

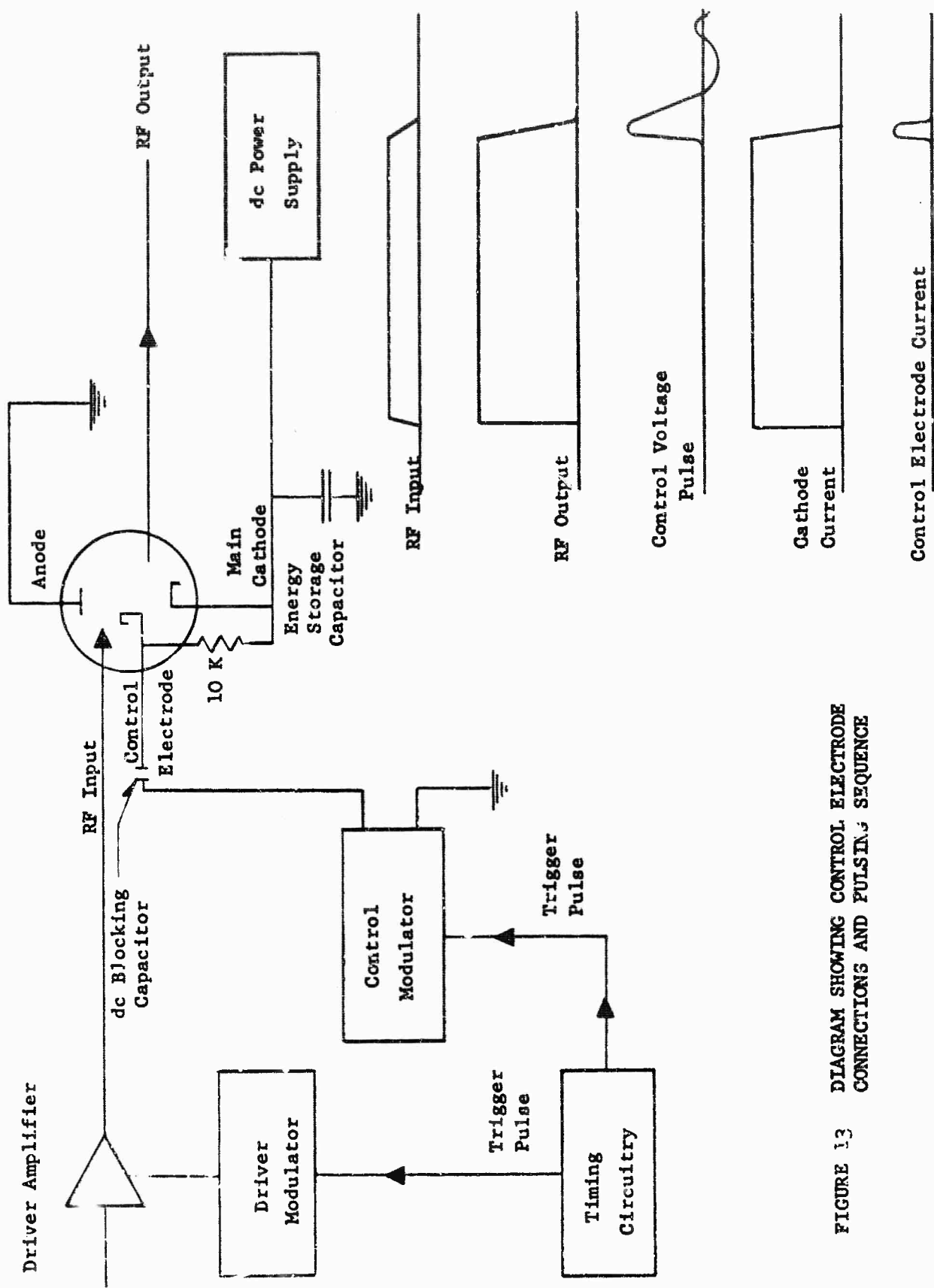


FIGURE 13 DIAGRAM SHOWING CONTROL ELECTRODE CONNECTIONS AND PULSING SEQUENCE

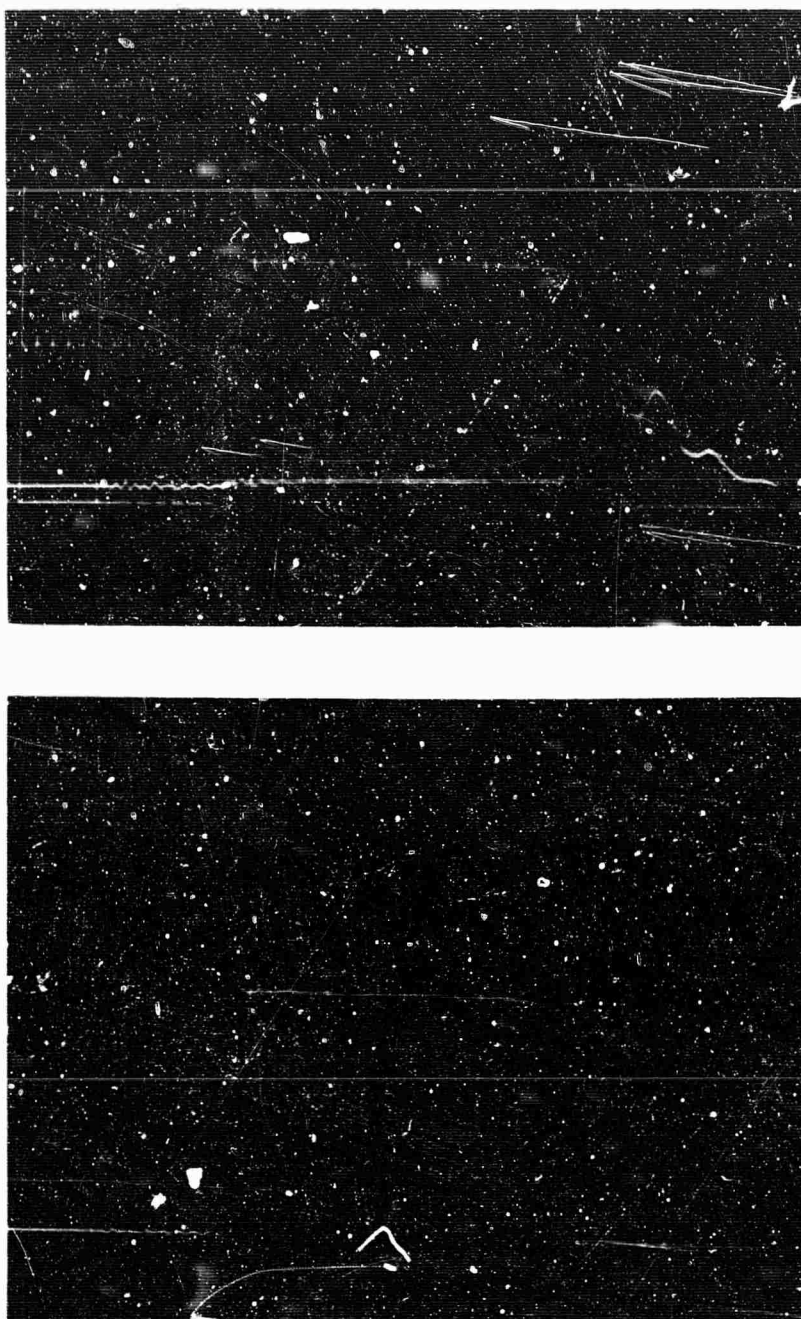


FIGURE 14 AMPLIFIED RF PULSE AND CONTROL ELECTRODE PULSE DISPLAYED
SIMULTANEOUSLY (top) AND RF PULSE ALONE (bottom).

Pulse length is 2.5 microseconds

The use of this modulation technique eliminates the need for a high power modulator, which should prove to be of considerable advantage in a radar transmitter application. It not only eliminates the modulator equipment required but also increases efficiency through the elimination of modulator power losses. It should be noted that, in order to take advantage of this technique, a forward wave slow wave circuit must be used so that operation may be obtained over the full bandwidth at a fixed value of dc voltage.

2.3 RF Performance Data

One way of representing performance data on crossed-field amplifiers is in the form of performance charts. Figures 15 and 16 show typical performance charts for an input drive level of 5 kw peak for two different tubes (L7F and D37F). On such charts, voltage is shown as the ordinate and current as the abscissa. V-I curves are plotted for different values of applied magnetic field. Contours of constant power output and constant efficiency are also shown. The curves are somewhat reminiscent of those for a magnetron except that the V-I curves are somewhat steeper.

Tube L7F was constructed with the end hats extending one-half the distance from cathode to anode. Based on magnetron experience, previous tubes had end hats extending one-third the distance. This change to larger diameter was made because results of an X-band crossed-field amplifier program indicated an improvement in efficiency of as much as 5 percentage points when the end hat extension was increased from one-third to one-half of the anode-cathode spacing. Results measured on tube L7F are shown in Figure 16. The performance chart of Figure 16 shows a significant increase in efficiency and a difference in the shape of the efficiency contours for the larger diameter end hats when compared to the results for the smaller end hat design shown in Figure 15. The increased end hat diameter

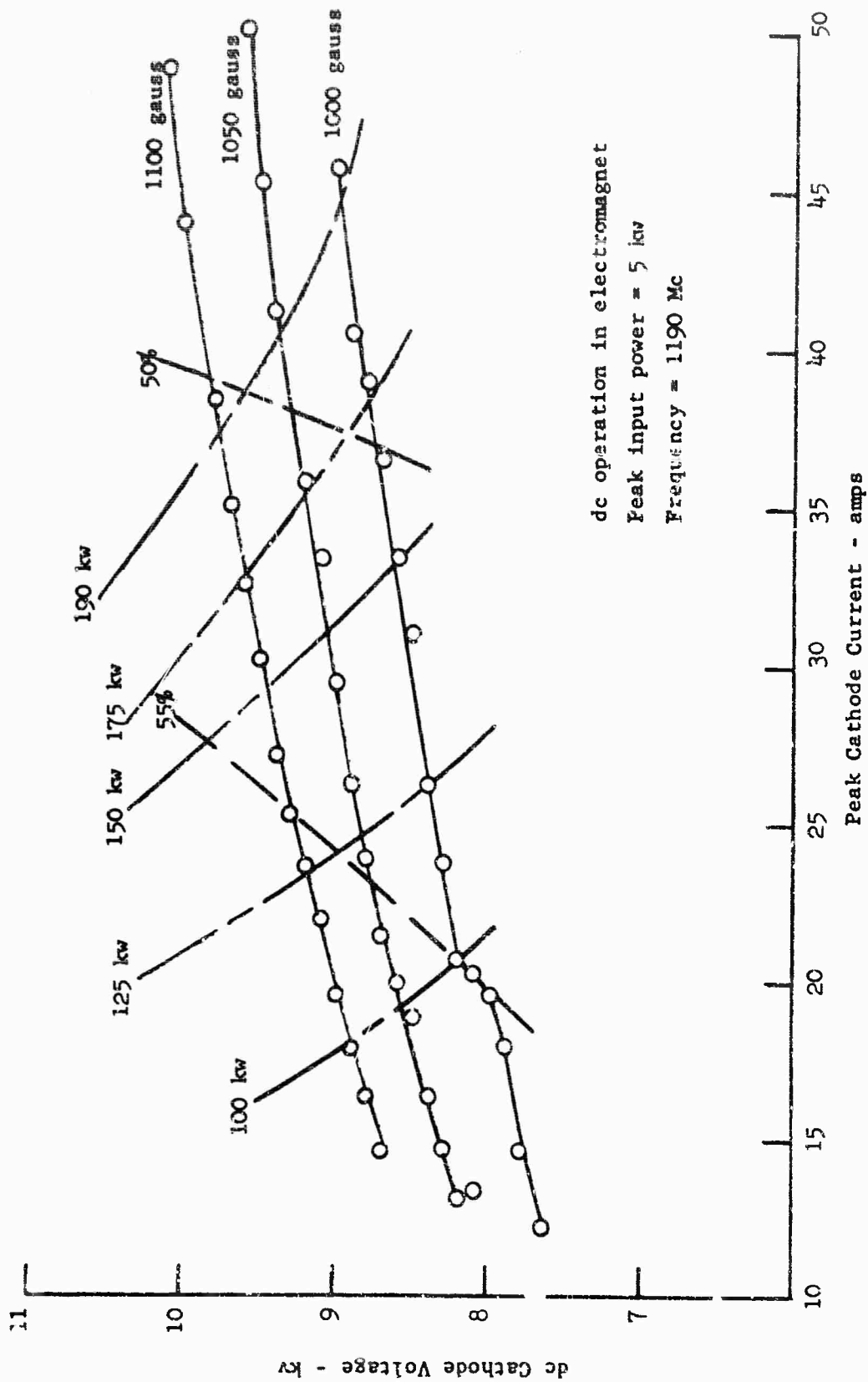


FIGURE 15 PERFORMANCE CHART FOR SFD-209, TUBE D37F

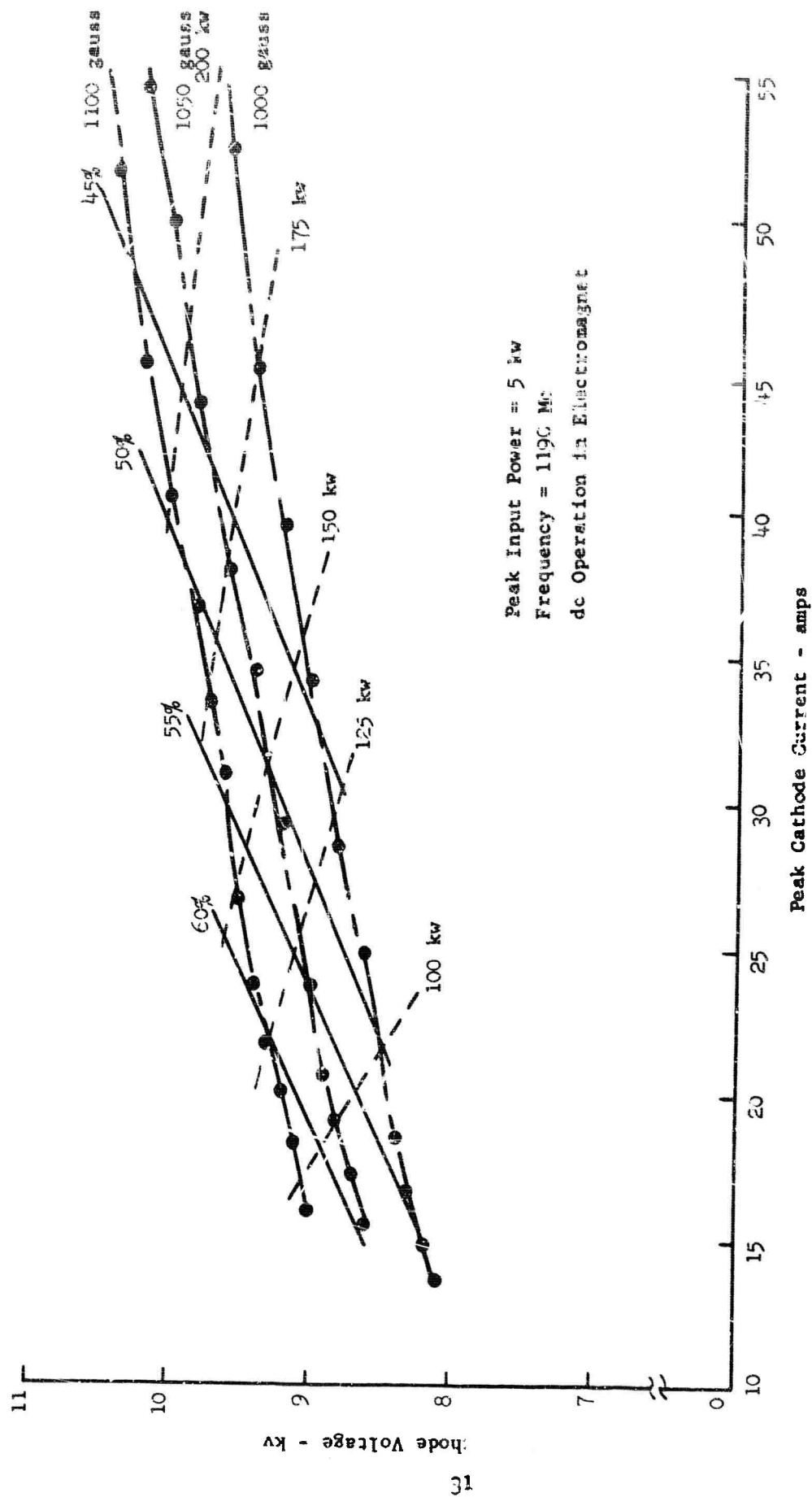


FIGURE 16 PERFORMANCE CHART FOR SFD-209, TUBE L7F

resulted in flatter efficiency contours which maintained high efficiency to higher values of operating current than had previously been the case. These flatter efficiency contours represent a very marked change as compared with the shape of efficiency contours seen in a number of previous tubes.

A direct cause-and-effect relationship between the higher efficiency and the changed end hat diameter is apparent. The design criteria previously employed called for the end hats to extend one-third of the cathode-anode spacing. This was based on previous magnetron experience. The results on our X-band and L-band crossed-field amplifier programs now indicate that these magnetron criteria are not adequate for crossed-field amplifiers and that additional space charge confinement is required. This probably results from electrons in the amplifier requiring a comparatively longer time to travel from the cathode to the anode because of the lower RF field strengths. This would be especially true near the input of the tube. During this longer transit interval, the axial space charge forces can be more effective in forcing the electrons axially out of the interaction region. It is therefore obvious that a higher degree of axial confinement is required in a crossed-field amplifier than in a magnetron.

Typical operation of this tube occurs at about 1100 gauss magnetic field. By proper choice of the operating point, we may maximize either efficiency or power output. Figures 17 and 18 show typical frequency response curves taken at fixed magnetic field and fixed anode voltage. It is seen that the tubes operate at close to 50% efficiency over a 100 Mc band and that the power variation with fixed operating voltage is 1 db or less. Operation at different peak power levels through adjustment of anode voltage is illustrated by Figures 17 and 18. Figure 19 shows operation at 1 kw peak drive over a 70 Mc bandwidth. Figure 20 shows a typical power output versus

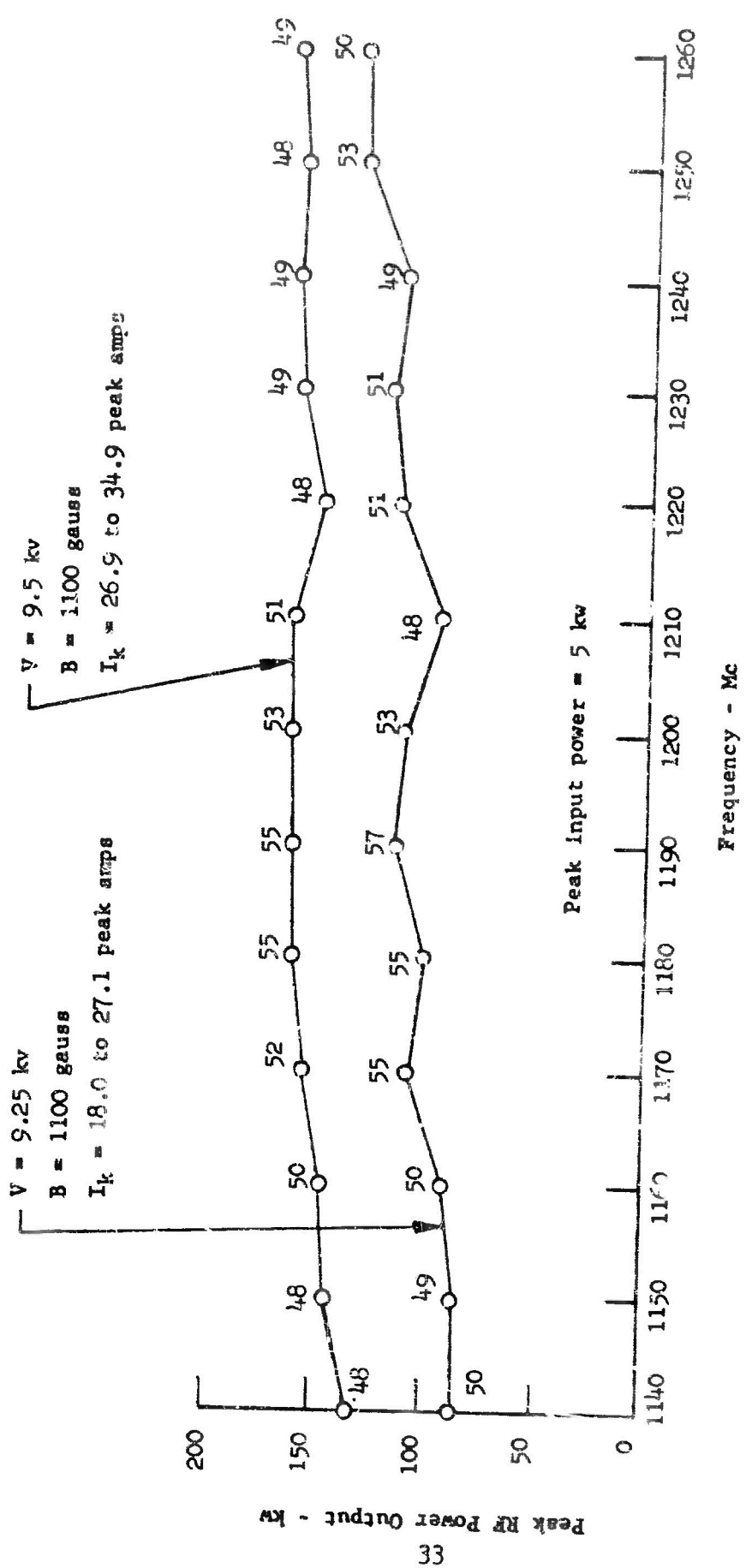


FIGURE 17 OUTPUT POWER VERSUS FREQUENCY AT FIXED ANODE VOLTAGE FOR SFD-209, TUBE D37F

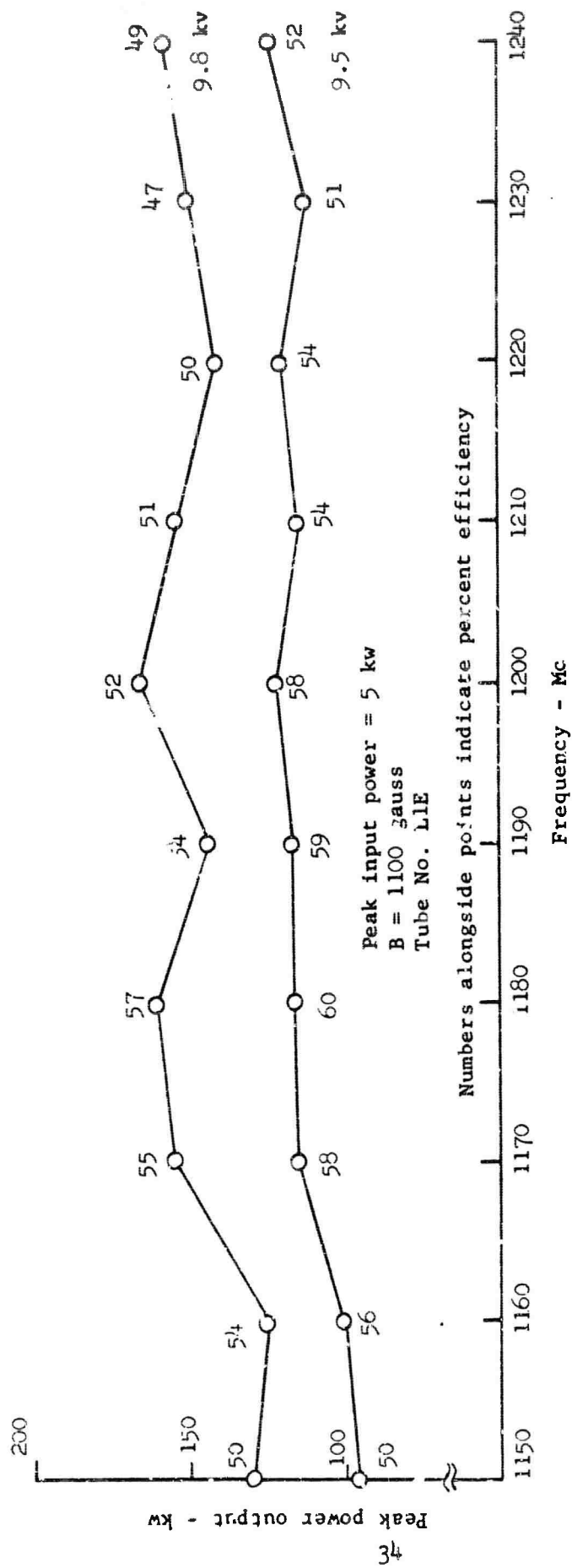


FIGURE 18 PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR THE SFD-209
 (L-BAND CROSSED-FIELD AMPLIFIER) WITH dc OPERATION

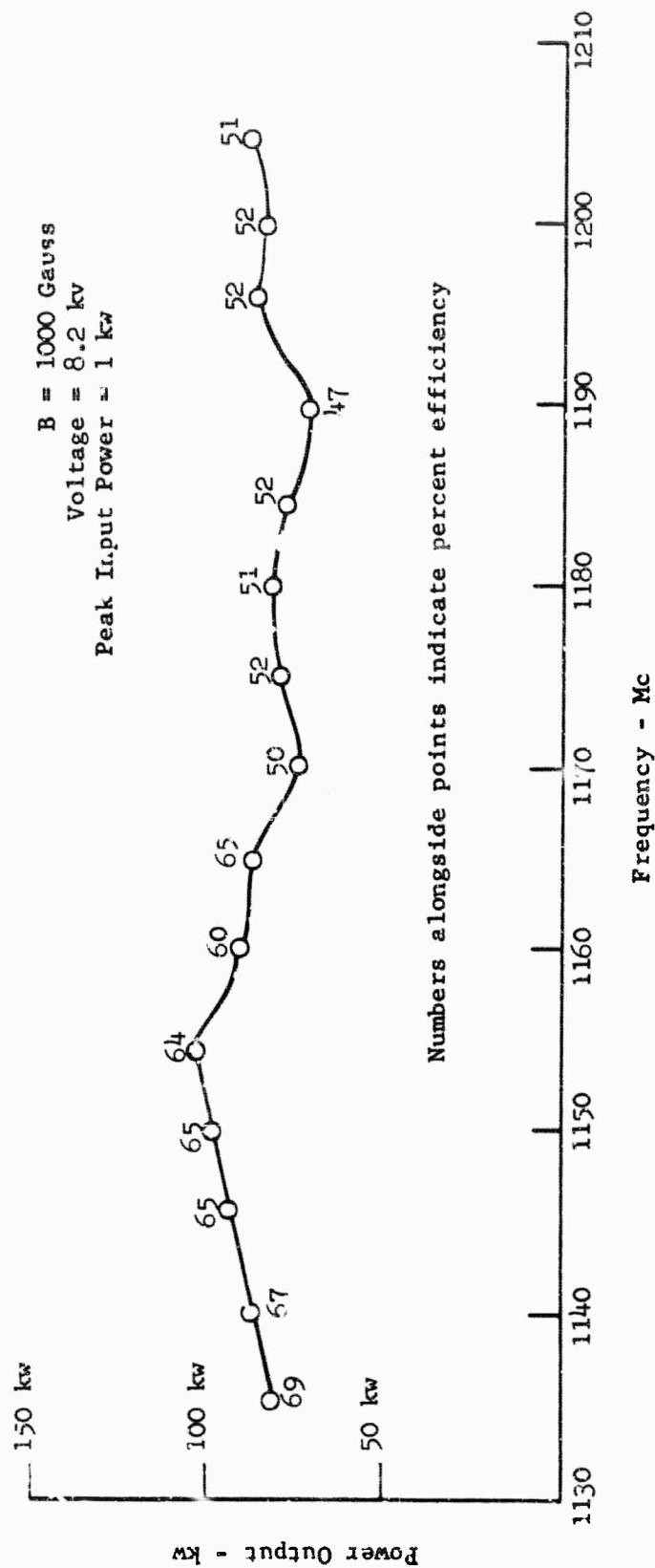


FIGURE 19 POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR THE SPD-2C9 TUBE WITH 1 kw DRIVE

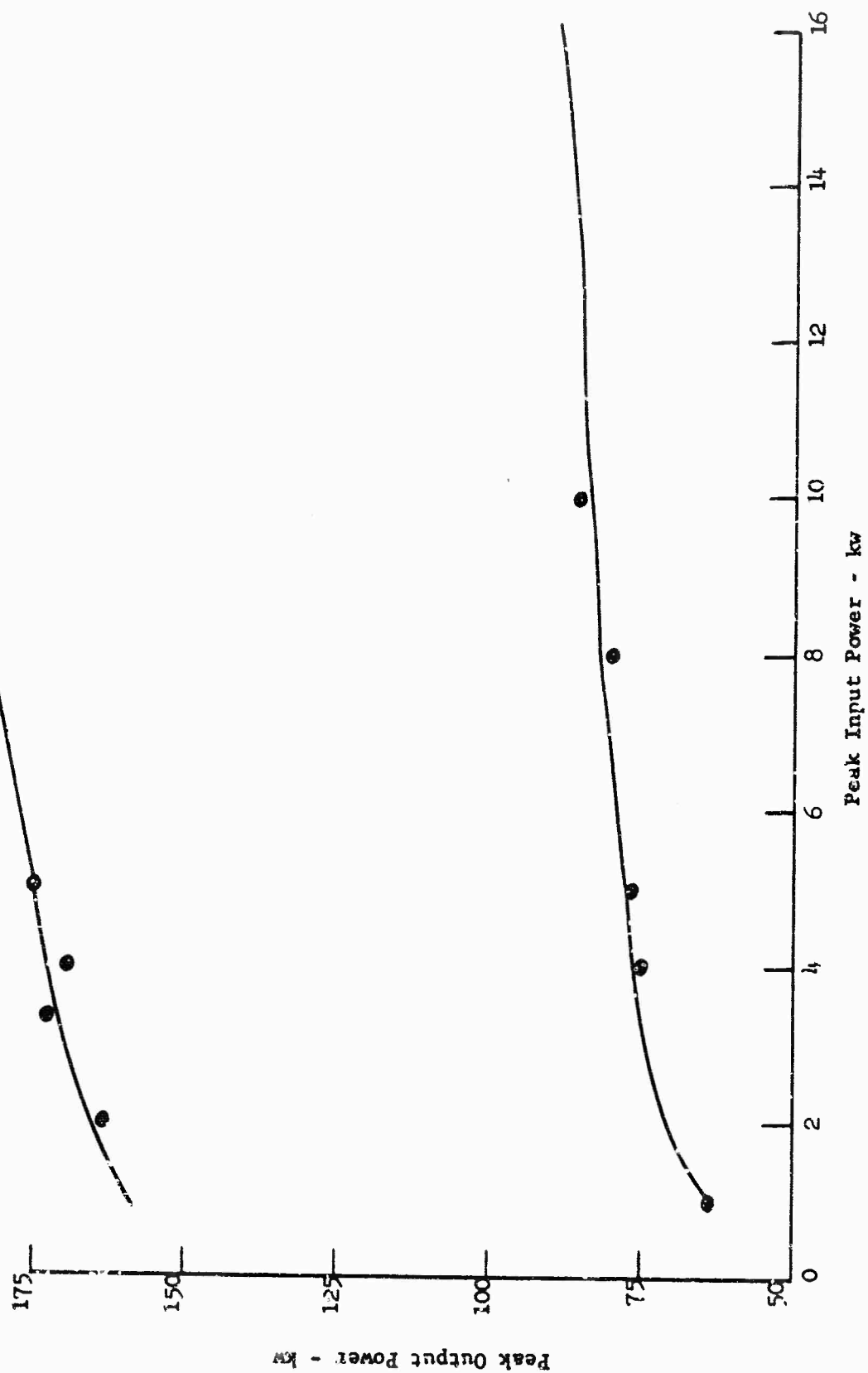


FIGURE 20 PEAK OUTPUT POWER AS A FUNCTION OF PEAK INPUT POWER
FOR TWO DIFFERENT CONDITIONS OF OPERATION OF THE SFD-209

power input characteristic taken at constant anode voltage. It is seen that the tube is heavily saturated; i.e., power output varies little with power input. This characteristic is advantageous in a transparent tube, giving a measure of immunity to the effects of multiple reflections between source and load.

2.4 Average Power and Pulse Length Capabilities

Most of the peak power characterization of the SFD-209 has been performed using a 2.5 μ sec pulse length which is available from our magnetron source. However, to evaluate the pulse length and average power capabilities, we have used the feedback loop arrangement shown in Figure 6. Here the crossed-field amplifier is triggered on by a 2.5 μ sec pulse from the driver. At the cessation of the drive pulse, the feedback loop takes over and the tube continues to oscillate at a frequency determined by the feedback loop until the control pulse is applied to turn the tube off. Since the feedback loop is external to it, the tube sees the same conditions it would when driven by a long pulse traveling-wave tube driver. The pulse length seen by the crossed-field amplifier may be controlled by the delay introduced between magnetron trigger pulse and control electrode turn-off pulse.

Figure 7 shows oscillograms of the pulse shapes for a 25 μ sec pulse and Figure 8 shows pulse shapes for a 200 μ sec pulse. The loss in the feedback loop is comparatively high so that the effective drive power under the feedback loop condition is less than 2 kw peak. This causes a drop in output power at the cessation of the drive pulse. The droop in the pulse shape subsequent to this is caused by the gradual discharge of the energy storage capacitor.

The 200 μ sec pulse tests were conducted with a 24 μ f energy storage capacitor. The droop was under 1 db as calculated from the measured V-I characteristics with an average output power of approximately 1100 watts. Average power was then raised to 1500 watts for a

period of about a half hour. At this point, we damaged components in the output coaxial line system which were not rated for this average power level. The tube, however, was operating satisfactorily at this level.

In other tests, we have operated the crossed-field amplifier using the feedback loop technique for 600 μ sec at about 100 watts average power output. We believe the tube is capable of millisecond pulse lengths.

2.5 Life Tests

Life tests were conducted on a group of SFD-209 tubes. One operated for over 5500 hours, another for 2900 hours and two more for over 1000 hours. The life tests were conducted with three tubes operated from a common dc power supply and fed by a common RF drive source. Figure 21 shows a photograph of the life test equipment. The tubes were tested in large laboratory permanent magnets which simulate the conditions they would see in an array. For a single tube application, there are much more efficient ways of building the magnetic circuit as will be discussed in a later section. Figure 22 shows a group of SFD-209 tubes awaiting life test. Figure 23 shows the RF test position where the tubes are characterized before being placed on life test.

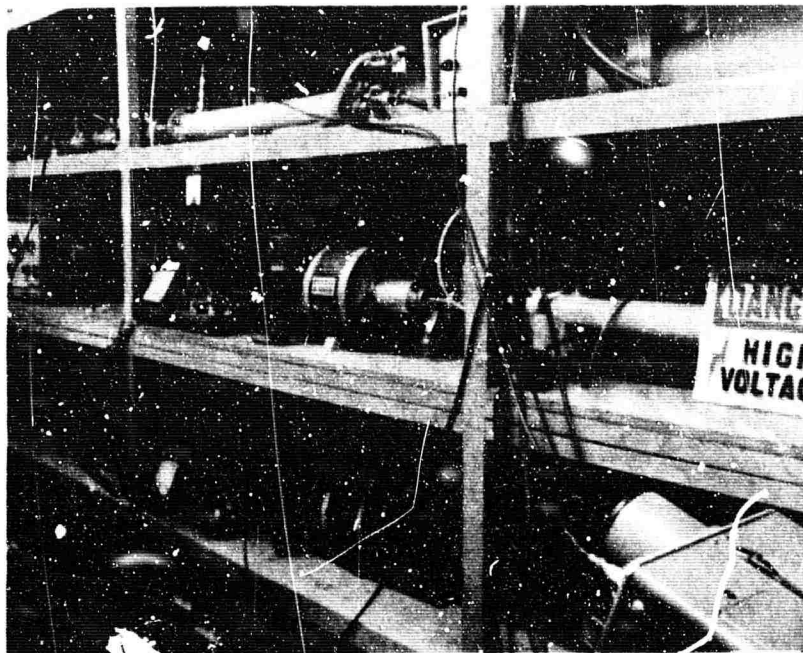


FIGURE 21 PHOTOGRAPH OF L-BAND LIFE TEST AREA

The two tubes on test are mounted in the cylindrical permanent magnets in the middle and lower center of the photograph. These magnets simulate the magnetic field configuration that would be used in a phased array magnet matrix and are larger than would be required for a package designed specifically for a single tube application

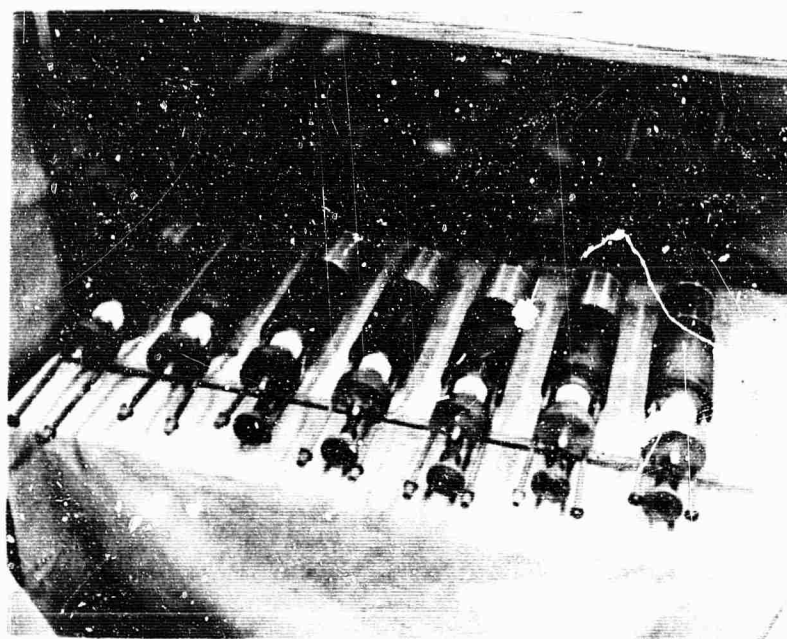


FIGURE 22 SFD-209 CROSSED-FIELD AMPLIFIERS WAITING
FOR LIFE TEST POSITIONS TO BECOME FREE

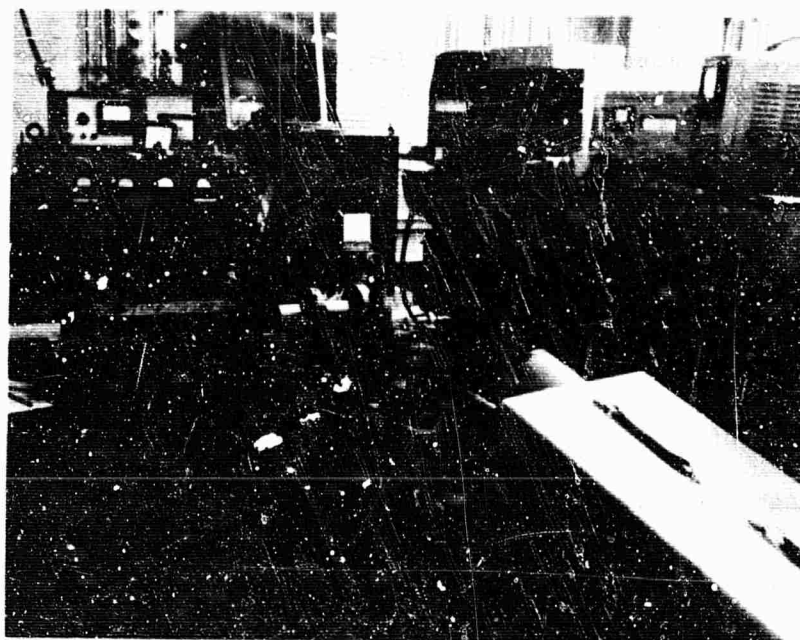


FIGURE 23 THE SFD-209 HOT TEST SET

**3.0 STUDY OF MAGNETIC DESIGN AND PROBLEMS RELATED TO MOUNTING
TUBES IN AN ARRAY**

3.1 The Requirements and Different Approaches to satisfying Them

A study was made of the possible magnet configurations for tubes in a phased array. Features of an ideal arrangement would include:

1. magnetic field uniformity in the interaction space.
2. minimum interaction of neighboring tubes and minimum magnetic fields external to the array
3. tube and magnet package to fit within a half-wavelength center-to-center spacing
4. minimum package weight
5. reproducibility of magnetic field from tube to tube. This is required so that all tubes may be operated from the same dc power supply.

The various magnet arrangements considered were:

1. shielded package in which the magnets were mounted on the ends of the tube and connected by a soft iron yoke
2. magnet located inside the cathode
3. magnet surrounding the tube and packaged with the tube
4. magnets fixed in the array with tubes plugged into them.

Of these arrangements, numbers 2 and 3 proved to be impractical. The magnet inside the cathode could give adequate magnetic field but the field uniformity in the interaction space was far short of what was desired. External magnets packaged with the tubes would result in excessive nearest-neighbor effects when the tube packages were changed. A detailed discussion of the different approaches was presented in the Third Semi-annual Report (November 1963, pp 54-75, AD 426 729).

A shielded package would be very desirable for minimizing nearest-neighbor effects and external magnetic fields. It would also permit adjustment of individual magnets to the same magnetic field within a close tolerance. The method of constructing a shielded magnetic package would be to mount magnets on either end of the tube and connect them with a soft iron yoke. Unfortunately, the presence of this yoke tends to make the magnetic field in the interaction space non-uniform unless the diameter of the yoke is fairly large. Thus the magnetically-shielded package tends to violate either the field uniformity or the maximum diameter requirement. It appears possible to compromise both of these requirements a bit and obtain a workable solution with this approach but requires some significant complication in the tube design.

Magnets fixed in the array and surrounding the tubes seems to represent the easiest means of obtaining a close approach to the desired objectives without compromising the tube design. The magnet weights are no more than the magnet plus yoke weight for the shielded package, and further this weight does not have to be handled each time a tube package is replaced. Replacement of tubes will have little effect on the magnets which remain fixed in place. The spacing requirements can be met without introducing field uniformity problems. The tube-to-tube reproducibility of magnetic field and the external field of the array are more serious problems than with the shielded package. However, as will be discussed below, there are ways of surmounting these problems.

Consideration of the relative merits of the shielded package and the fixed external magnets led to the conclusion that the latter approach would be preferable. Its disadvantages are minor and it does not sacrifice tube performance. Discussions were held with magnet suppliers about the problems of the fixed magnet array. From these discussions, we came to an agreement with the Arnold Engineering Company by which they would construct a nine-position demonstration magnet array module.

3.2 Magnets in a Fixed Array

Figure 24 shows the configuration for magnets fixed in an array worked out with the help of P. P. Cioffi of Arnold Engineering Company. Equality of magnetic field from cell to cell is assured by using common pole pieces for all the magnets. These pole pieces may be thin since they need only carry the flux for a single cell plus the small amount of flux flowing between cells because of inequality in the magnets from cell to cell. To further minimize the field differences from cell to cell, a soft iron flux guide will surround each tube. The flux guide will in turn be surrounded by a non-magnetic spacer and then by the magnet. The magnets will be 4.5" square and the cells 4.75" square so that there will be some space between magnets for additional supports, cooling liquid pipes, etc. The magnets will be made of Alnico 8 of improved quality; i.e., of higher energy product than usually shown in the literature for this material. With the means of equalizing flux between cells used here, cell to cell variation of less than 1% is predicted. A number of dummy cells will be placed around the edges of the array. These will provide a border of magnetic material to supply the external flux from the edges of the array. Thus, the first row of tube cells will see the same conditions surrounding them as do the cells in the center of the array.

Some thought has been given to assembly of the array. The best procedure appears to be to construct it in modules of nine or more tubes. These modules would be magnetized at the factory and then shunted with soft iron keepers around the outsides or in the tube positions. The modules could then be shipped and the array assembled. Once the array is assembled, keepers in the tube positions would be removed and the array allowed to come up to full strength. The pole pieces of the modules would mate in lap joints to provide a low impedance magnetic path.

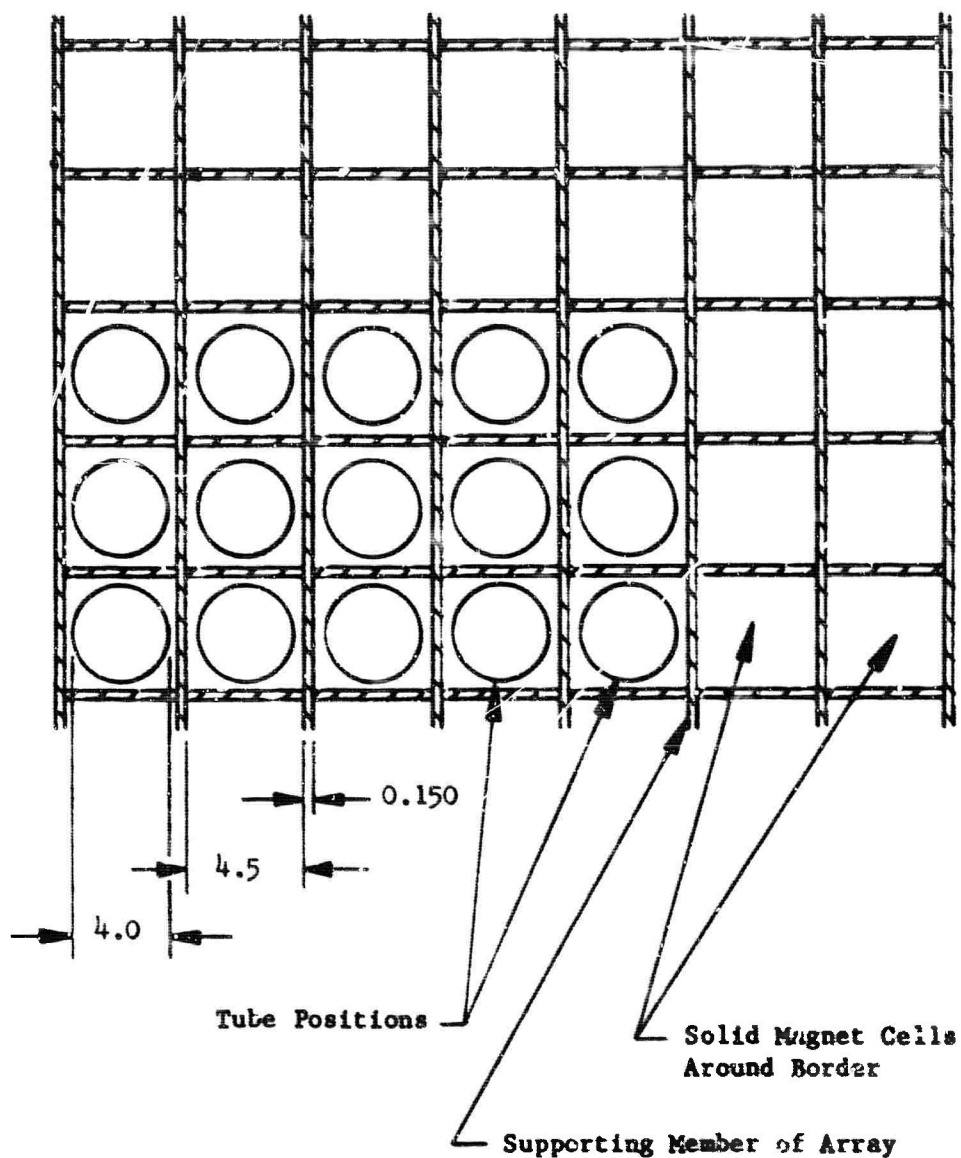


FIGURE 24 SKETCH OF AN ARRAY OF FIXED PERMANENT MAGNETS INTO WHICH TUBES MAY BE INSERTED

If the external magnetic field of an array of this sort is excessive, it can severely interfere with other electronic equipment nearby. If an array of 30 or 40 tubes on a side could be approximated by an ellipsoid, its ratio of thickness to diameter would be very small. As a consequence, the magnetic material would be operating very near its coercive force and the external magnetic flux density (and, therefore, magnetic field) on either side of the ellipsoid would be very small. (The field in the material approaches the coercive force and the external flux approaches zero as the ratio of thickness to diameter approaches zero.) Near the center our actual array looks like an ellipsoid with a large number of holes punched through it. The magnetic material now supplies the flux which floods these holes. The ratio of B to H in the magnetic material assumes a value equal to the ratio of the cross-sectional area of a cell filled with non-magnetic material to that of a cell filled with permanent magnet material. (The saturated flux guide would be approximated by an equivalent area of unity permeability carrying the same total flux.) Away from the array a distance equal to one or two cell spacings, the field would again look like the ellipsoid field and would be very low.

What we have said so far holds for the field near the center of a large array a short distance from its face. A practical array has a constant thickness rather than one which tapers toward zero near the edges as for an ellipsoid. Therefore, we can expect the field near the edges of the array and a short distance from its face to be larger than that for an ellipsoid. To investigate this effect, we measured the field near the top plate of the configuration shown in Figure 25 with the results shown. We know by analogy with the ellipsoid problem that the results will be a function of the ratio of transverse dimension (s) to thickness (t). No difference in field distribution beyond a couple of cell spacings from the array results from using solid plates to represent the array pole faces and providing

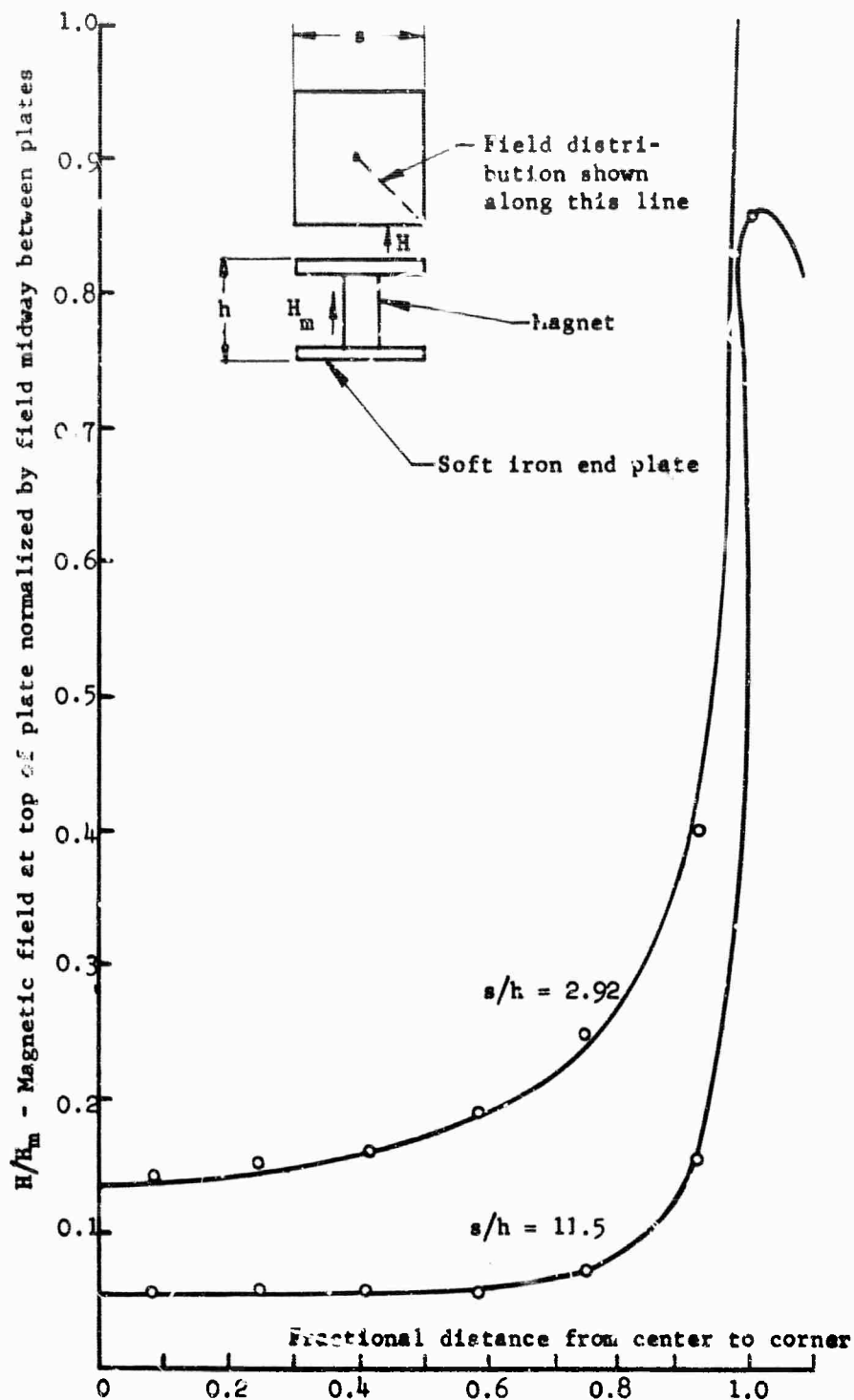


FIGURE 25 DISTRIBUTION OF MAGNETIC FIELD FROM CENTER TO CORNER OF POLE FACES FOR A MAGNET BETWEEN LARGE SOFT IRON POLE FACES

The pole faces simulate the end pole faces of an array and the magnet simulates the MMF between these pole faces. This experiment shows how the magnetic field near the pole faces drops to a low value near the center. For an actual array, the s/h ratio would be of the order of 100.

magnetomotive force between them by a small magnet in the center. Thus, we can get information of interest about a large array from this small scale model. (The plates were about 1 foot square.)

Figure 25 shows the distribution of magnetic field over the face of the square plates for different ratios of side length to thickness. These results show that the magnetic field drops very rapidly as we move in from the edge of the plate. Figure 26 shows the magnetic field a distance of 0.1 times the side of the plate in from the edge. This field decreases rapidly as we approach the thickness to side ratio which would be encountered in an actual array.

From Figure 26, we conclude that if we move in 0.1 of the way from the edge of an array operating with a magnetic field of 1000 oersteds in the tube positions, the field near the array face will be less than 10 oersteds. Thus we arrive at a recipe for insuring that electronic equipment behind the array will not be disturbed by stray magnetic fields. We simply plan to make the active part of our array occupy only 80% of the side of the full magnet array. Fields behind this active part of the array where we want to place driver tubes, control modulators, etc., will then be very small. Large leakage fields will be confined to the region around the edge of the array which we can conveniently avoid. This means a 12 foot by 12 foot active array surrounded by an 18 inch border of permanent magnet material. (To provide adequate field uniformity to the first row of tube positions, this border may need to be only 8" wide.) The 18" border could be tapered in thickness to further reduce the external leakage flux. Thus the only region of high leakage flux from the array would be this border - perhaps 18" wide by 2' thick around the edge of the array.

This approach to the magnetic design of an array meets all our objectives. Field uniformity is excellent. Total weight is no greater than by other approaches. Much of the weight is fixed in the

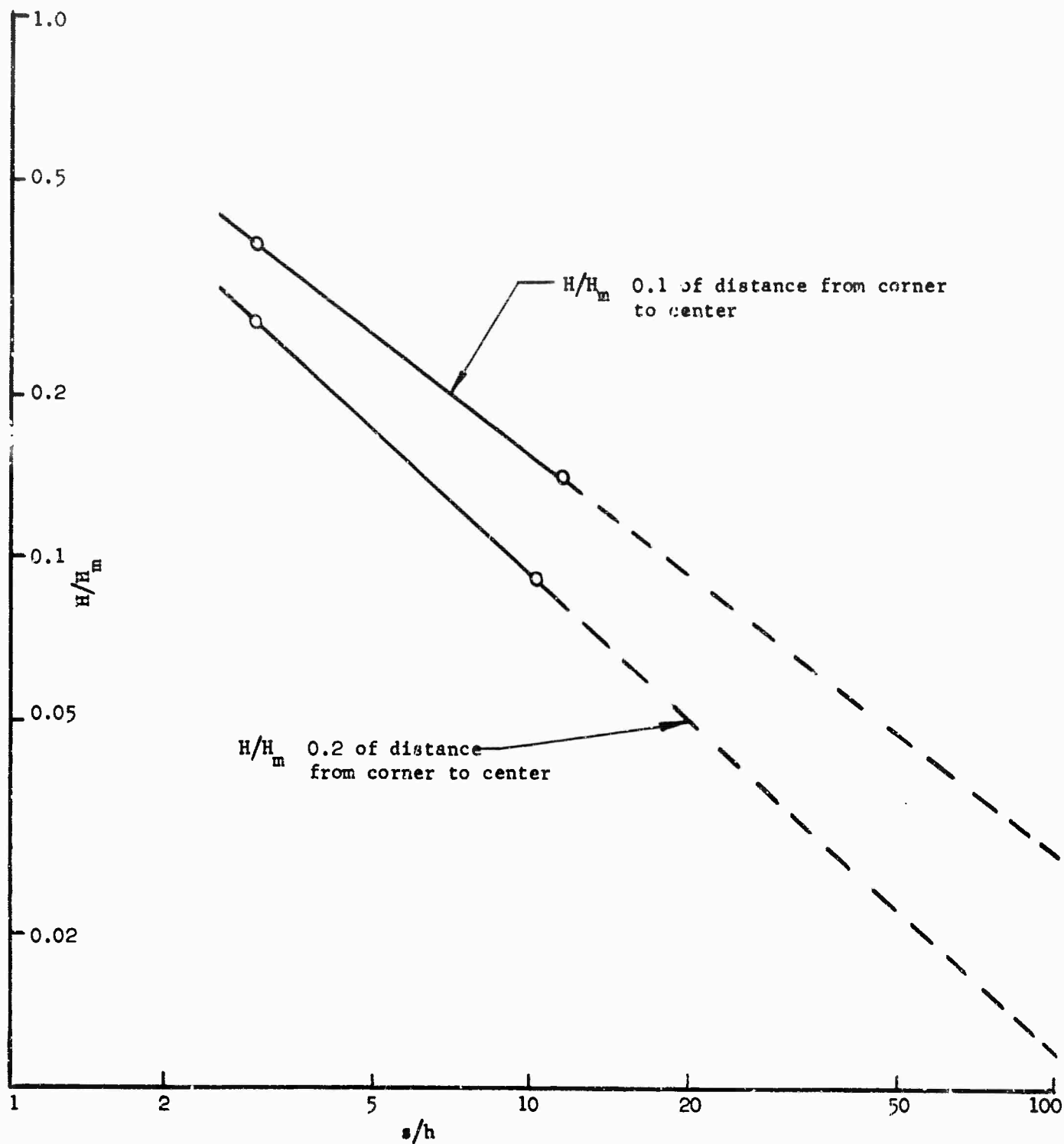


FIGURE 25 RATIO H/H_m FROM FIGURE 25 AS A FUNCTION OF THE RATIO s/h

Dotted portions of the curve show extrapolation to large s/h or thin flat structures such as would be used in a phased array.

array and does not have to be handled when a tube is replaced. The tubes fit within the desired spacing and external field is low, at least in the region where it is of significance. We have thus assumed this type of array construction in our tube design. The Arnold Engineering Company undertook the construction of a nine-position module plus border magnets to demonstrate the principles involved in this approach. The dimensions of this module were chosen to be compatible with the SFD-209 when used as a final transmitter tube in an array. Figure 27 is a photograph of the nine-cell array built by Arnold Engineering Company.

3.3 Magnets for an Isolated Tube

The tubes being developed on this program may at some point find application in systems in which only a single tube is required. The magnetic considerations in such cases are entirely different from what they are when the tube is used in an array. In the array application, an individual cell magnet is considerably smaller than the external magnet required if a tube is to be used in an isolated tube application. The magnet for an isolated tube must supply leakage flux to a large volume of space that in an array would be occupied by other cells and other magnets. The required increase in magnet cross-section causes an attendant increase in the outside perimeter of the magnet and hence in the leakage flux. For the case of the SFD-209, the array cell magnet is about 4.5" square whereas a comparable magnet for an isolated tube application is a cylinder with an outside diameter of 10". The cross-sectional area of this magnet is very large compared to that of the array cell magnet. Less than 10% of its total flux appears on the inside of the hole, the remainder is external leakage. Such magnets have been ordered for laboratory use and life testing in spite to their size because they will accurately reproduce the magnetic conditions a tube will see in an array.

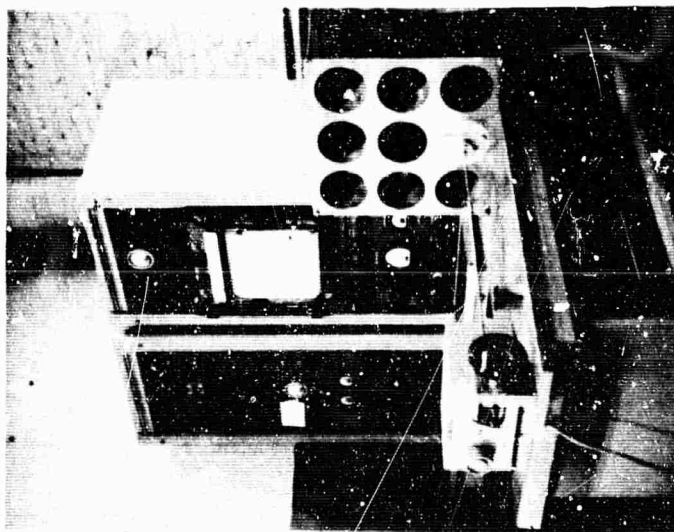
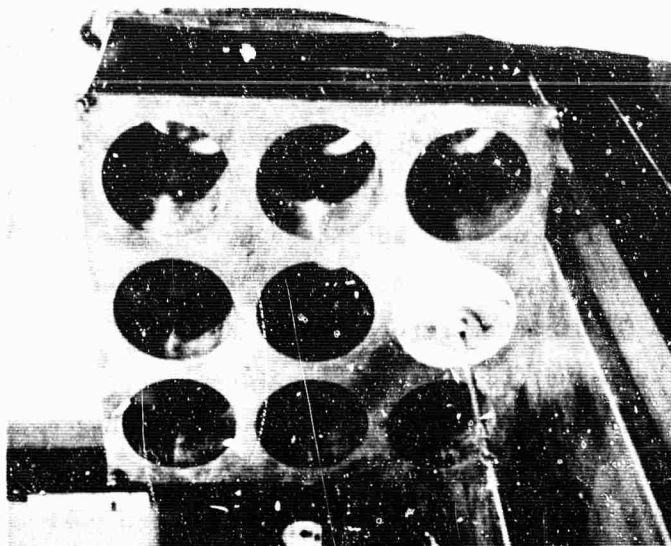


FIGURE 27 PERMANENT MAGNET MODULE WITH NINE TUBE POSITIONS
BUILT BY ARNOLD ENGINEERING COMPANY

For an isolated tube application, however, another approach is desirable. One possible alternative is to use a bowl or horse-shoe magnet of Alnico V-7, a higher energy product but lower coercive force material than the Alnico 8 required for the array cell magnet. An even more interesting result is obtained by returning to the magnetically-shielded package. By placing the magnets on the ends of the tube, their diameter is reduced by a considerable factor. Further, the Alnico V-7 magnets can be used to advantage here. The result is a reduction in weight from that required for the 10" diameter Alnico 8 magnet. Putting on a 9" diameter return yoke adds another 15 lbs. The total weight of the magnetically-shielded package, including the tube, thus becomes about 60 lbs. The consideration that has made the magnetically-shielded package practical in this case is the relaxation of the diameter requirement. Because of this, the return yoke can be moved far enough from the interaction space so that it does not interfere with the magnetic field uniformity there.

4.0 HISTORY OF DEVELOPMENT

The limitations on cross-section for this tube seriously affected the mechanical design of the tube. These limitations forced the use of straight vertical bars; i.e., the bars could not be bent into a radial plane in the inductive section as is sometimes done to decrease the axial length. Because of the axial thermal expansion as the bars heat up when the tubes run at full average power, the ends of the bars could not be firmly tied down. A further mechanical requirement imposed by the phase reproducibility called for in a phased array application was that the design be one in which a high degree of reproducibility could be obtained.

The higher gain requirements for the driver tube dictated a somewhat more complex circuit construction. The higher gain requirements necessitated the use of a relatively large number of bars, thereby increasing the difficulty of obtaining reproducibility. Several approaches to the problem were investigated. The first was to machine the entire anode circuit out of a solid block. Figure 28 is a photograph of the first machined anode. This procedure proved to be rather difficult and costly and therefore was not pursued further.

A somewhat different approach, combining the advantages of the simplicity of individual bars with the simple coupler arrangement and simultaneously providing maximum strength characteristics, was investigated. In this instance, molybdenum bars and a premachined molybdenum ring were assembled and brazed together. A subsequent machining operation simultaneously and accurately provides the correct capacitive gap and overall capacitive coupler diameter. Figure 29 is a photograph of the circuit prior to the final machining operation. This technique was used throughout the program with some slight modifications and proved to be the most economical method for insuring reproducibility of mechanical dimensions.



FIGURE 28 A VERSION OF THE SFD-209 ANODE IN WHICH THE CIRCUIT BARS ARE MACHINED BY SLOTTING A CYLINDER

The coupler rings are brazed to the circuit and slotted to form the capacitive couplers

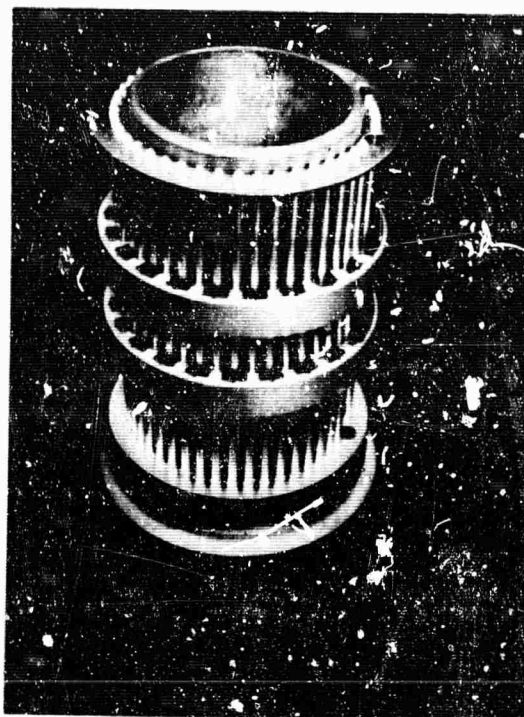


FIGURE 29 PHOTOGRAPH OF A VERSION OF THE SFD-209 ANODE
ASSEMBLED FROM MOLYBDENUM BARS

The anode is brazed together and the coupler rings added at the same time. Later the coupler rings are slotted. This assembly technique will also be used with bars made of copper and zirconium alloy

The tubes being made at this stage were demountable and designed for use in an electromagnet, so that maximum information could be obtained in minimum time. One end of the tube envelope had a flange for a demountable vacuum seal as shown in Figure 30. The demountable assembly allowed rapid testing of cathode geometries and materials. Utilizing this technique, the first significant RF tests were performed early in January 1962. The design and cold test work of the actual tubes was started in July 1961. Results on the tube were quite heartening and proved that the basic design was quite sound. These results were reported in the First Semi-annual Report (January 1962, AD 339 238) of this contract.

During the second period, control electrode operation from a dc power supply was successfully introduced. This feature had not been called for in the original contract, but it was evident that it would be of considerable advantage in an array radar because of the reduction in modulator equipment. With the agreement of Rome Air Development Center, effort was directed toward its inclusion.

By late June 1962, the driver tube development had resulted in a tube which could be characterized as fairly representative. This tube gave an output power of 100 kilowatts peak with a drive power of 3-5 kilowatts peak over slightly less than a 10% band. With an input of 1 kilowatt, an output power of 100 kilowatts could be obtained over a 5% bandwidth. The tube worked from a dc power supply with a control electrode μ of 3. The μ is defined as the ratio of the voltage necessary to turn the tube off to the operating voltage of the tube.

At this point, effort was concentrated on increasing the bandwidth, redesigning the tube in a sealed off version, and packaging it in a permanent magnet. Figures 31 and 32 are photographs of the redesigned tube and the tube in its permanent magnet package. A technical paper, summarizing the design philosophy and the results up to 1 August 1962, was presented at the Fourth International Congress

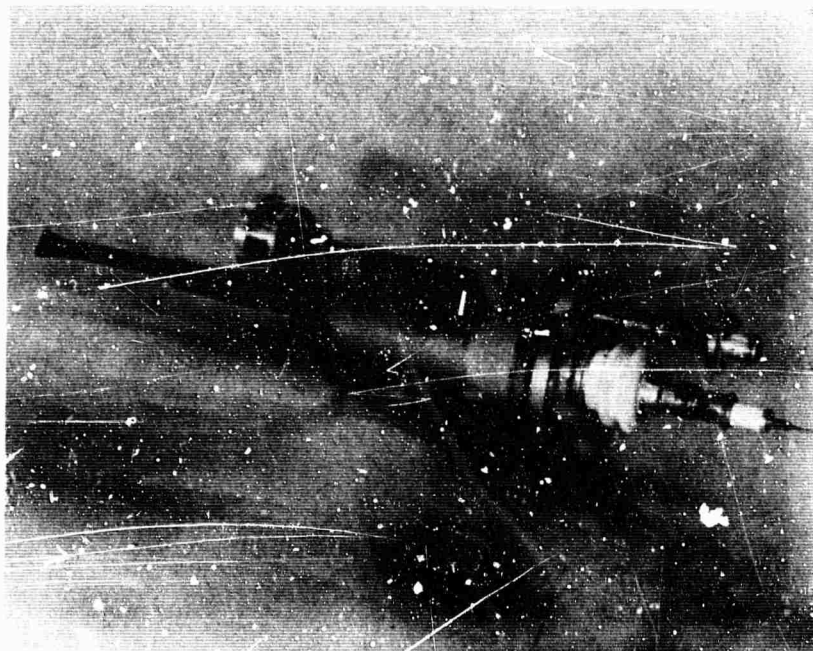


FIGURE 30 DEMOUNTABLE VERSION OF THE SFD-209 DRIVER TUBE
The demountable feature has permitted rapid evaluation of
minor changes

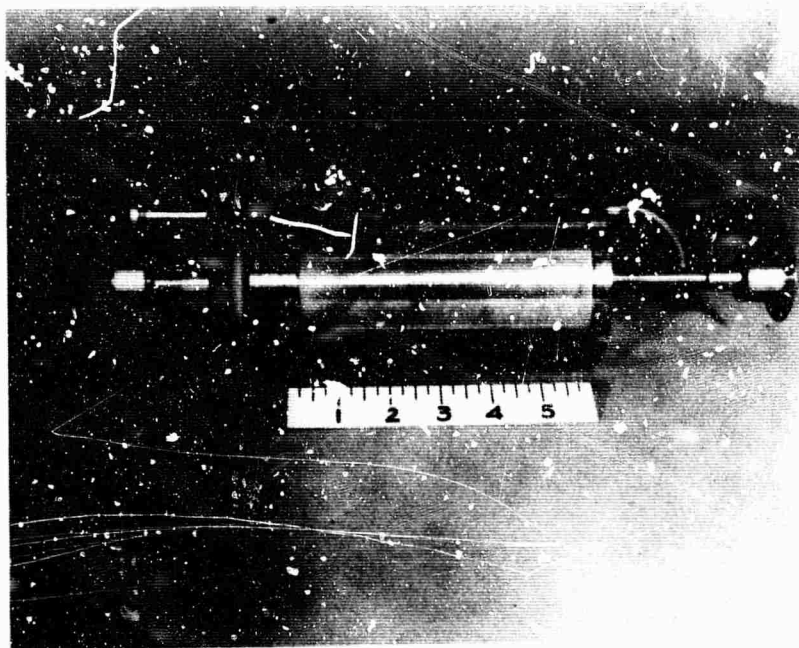


FIGURE 31 SEALED OFF VERSION OF THE SFD-209 DRIVER

A heliarc seal has replaced the demountable flange and the output window has been brought to the center of the tube so that the whole tube now fits within the required dimensions

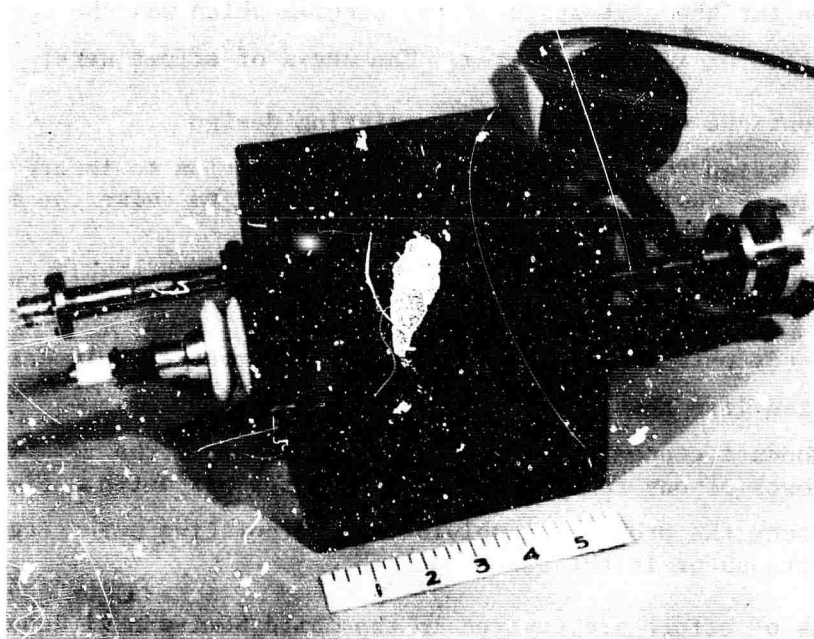


FIGURE 32 THE SFD-209 DRIVER TUBE PACKAGED IN A PERMANENT
MAGNET ADEQUATE FOR OPERATING THE TUBE AS AN
ISOLATED ENTITY

In an array, adjacent magnets would help one another out
so that a smaller magnet could then be used

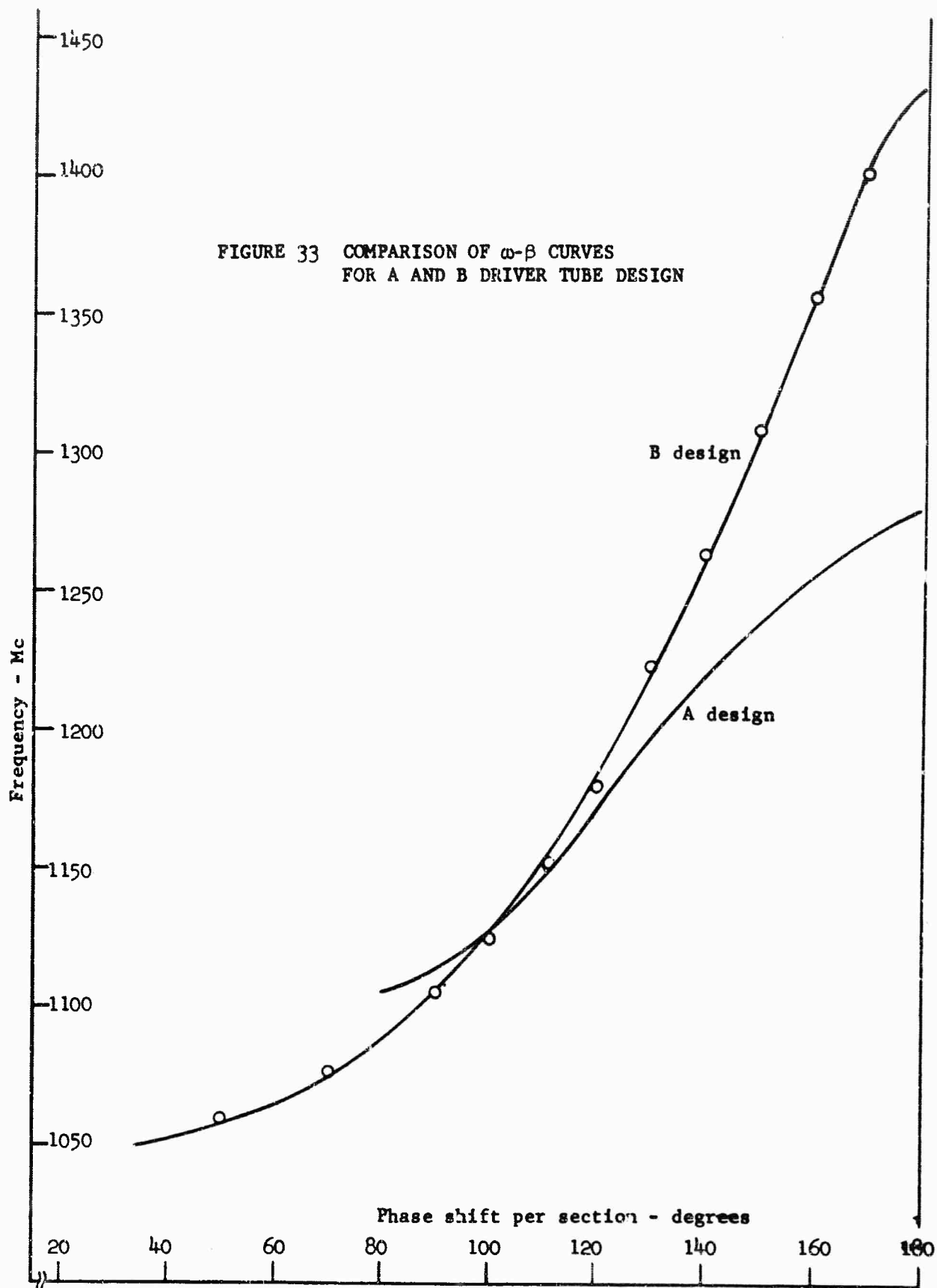
on Microwave Tubes in the Hague, Holland, on 6 September 1962. The results were presented in the Second Semi-annual Report (RADC-TDR-62-589, November 1962, AD 294 766) of this program.

During the next period from September 1962 to February 1963, considerable effort was expended along a number of different lines in preparation for the next stage of the program which was the construction and evaluation of identical tubes. The areas of effort were:

1. Continue evaluation of the existing design
2. Redesign the tubes based on this evaluation to incorporate liquid cooling through the circuit bars to raise the average power capability
3. Study and evaluate the magnetic problems for tubes mounted in an array
4. Study phase measuring techniques and procure equipment
5. Plan life test set up and procure equipment
6. Conduct economic study requested by ARPA of the relative costs of one tube versus several tubes per antenna element
7. Study the problems involved in incorporating a ferrite attenuator in the tubes.

These areas of effort are covered in the Third Semi-annual Report from September 1962 to February 1963 (RADC-TDR-63-492, November 1963, AD 426 729).

During the period from February 1963 to September 1963, attempts were made to increase the bandwidth of the driver tube by increasing the capacitive coupling between bars of the slow wave circuit. Figure 33 shows a comparison of the dispersion curves between the first design (Group A) and the second design (Group B). Figure 34 shows a comparison of the transmission bands for both designs. The cold bandwidth increases as predicted, but the gain of the tube dropped by much more than expected. This was found to be a consequence of the closer approach of the π mode to synchronism for the broad band tube. As a result, much higher RF drive was required to lock out the π



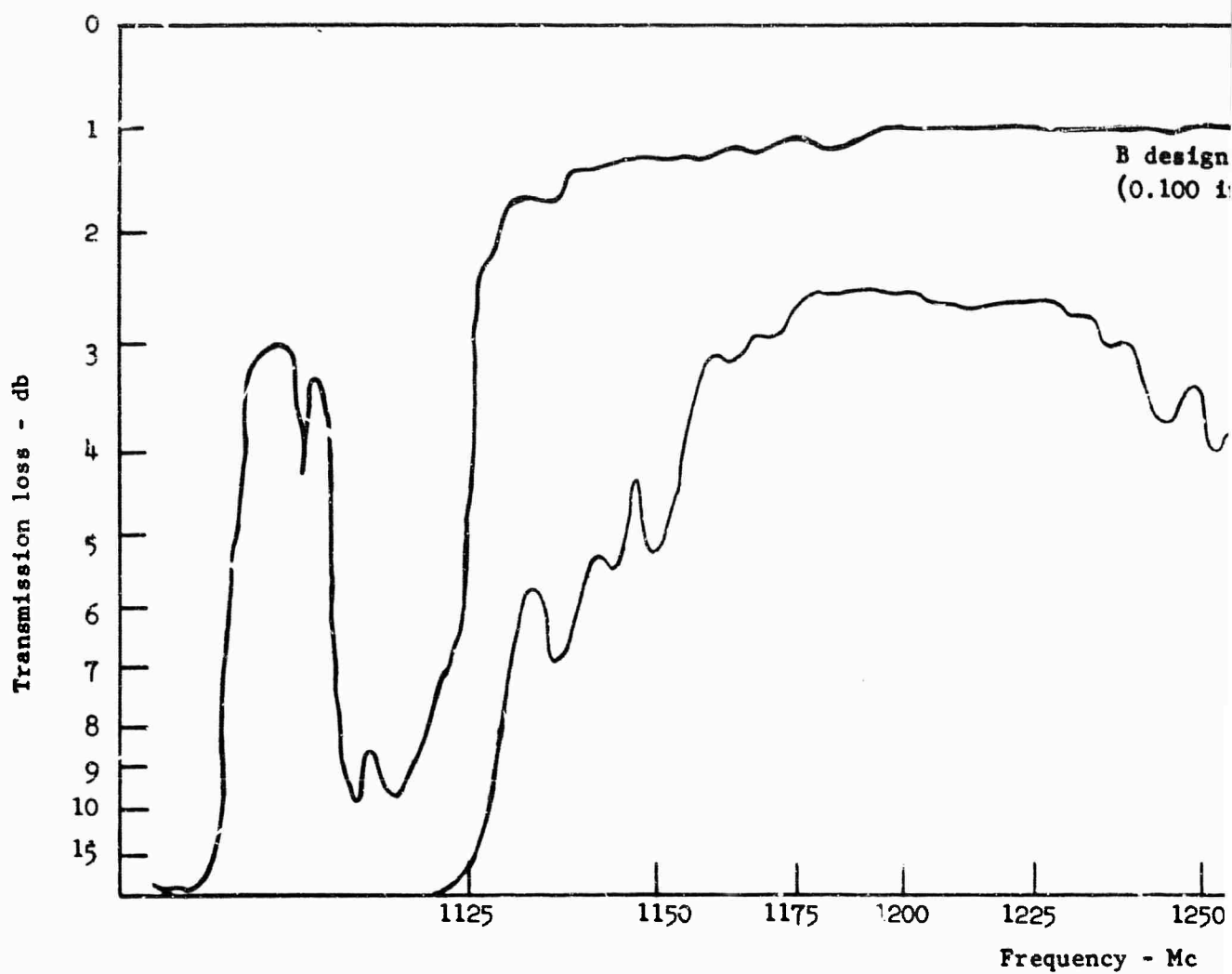
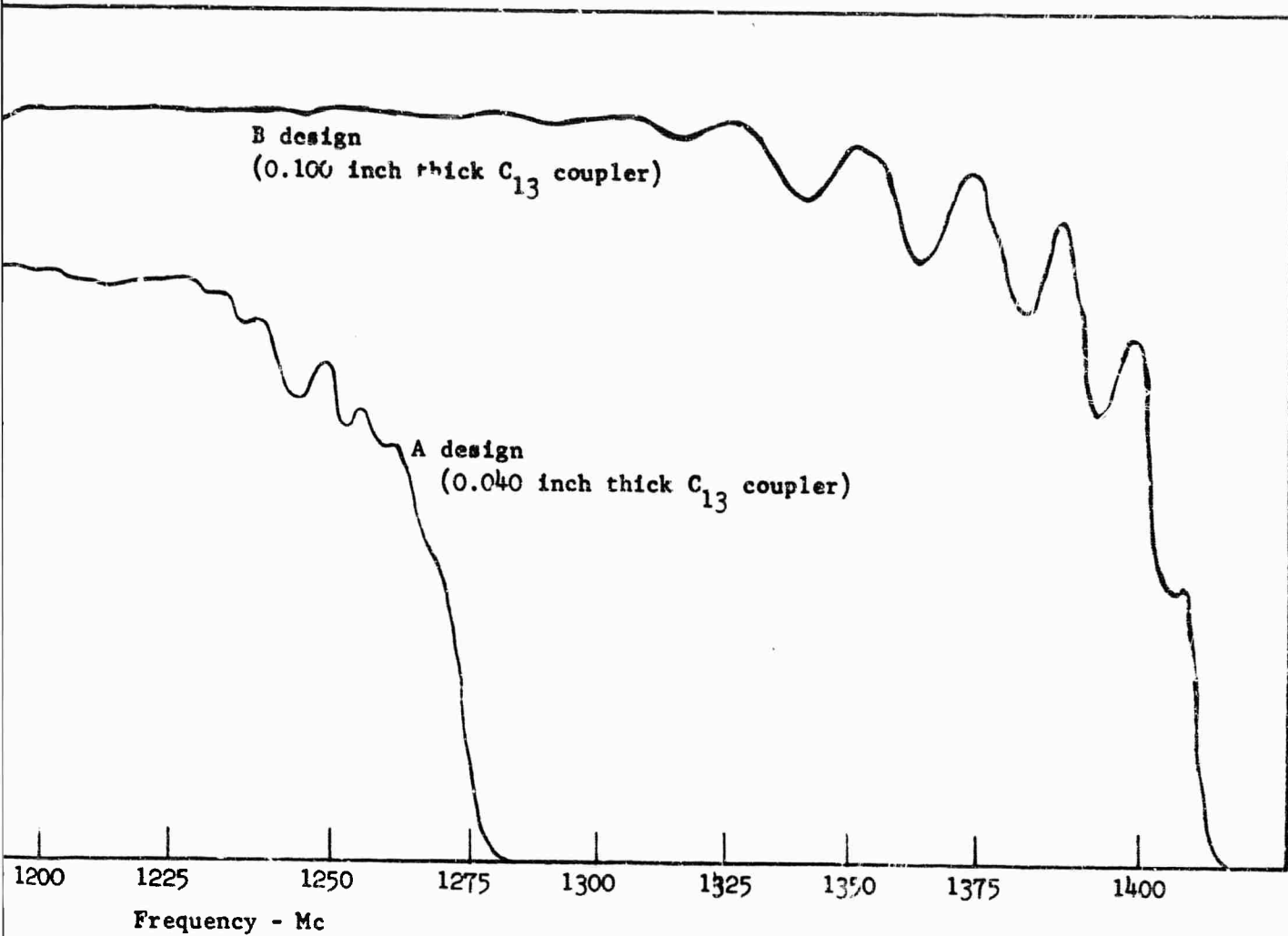


FIGURE 34 COMPARISON OF TRANSMISSION LOSS AS A FUNCTION OF FREQUENCY FOR THE A AND B DESIGNS - CATHODE WAS NOT IN POSITION FOR THESE TESTS



A FUNCTION OF FREQUENCY
AS NOT IN POSITION

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mode. From these experiments it was learned that, when no preferential damping of the π mode is incorporated into a forward wave tube, the circuit dispersion must be sufficient to make the voltage for a π mode oscillation different from that for amplification. Higher gains and broader bandwidths may be obtained by selective attenuation of the π mode frequency. A technique for doing this on the existing slow wave circuits was demonstrated. Because of time limitations and interest in obtaining life test data, it was decided not to attack the practical problems involved in incorporating the selective π mode attenuation in hot tubes. Instead it was decided to return to a narrower band design for the Group C tubes. These tubes would have a gain of about 16 db, a bandwidth of about 100 Mc, an output power of 100-150 kw peak and would operate from a dc voltage with control electrode turn off. They would thus represent quite an advance in the state of the art. Figure 9 is a photograph of the SFD-209 at this stage of development.

The appearance of an absorption resonance near band center of the new driver tubes led to a study of cathode structure resonances. The offending resonances were found to be coaxial modes between cathode and circuit as illustrated in Figure 35. Means for tuning these modes out of the band were found and applied, as shown in Figure 36. For the future, it was learned how to avoid such modes by proper interaction space design.

Improved performance of the driver tube was achieved through further investigation. A cathode study led to the conclusion that aluminum was superior to beryllium copper as a secondary emitting cathode material. Accordingly, it was decided to use aluminum cathodes in the final group of driver tubes. Studies of interaction space tapering showed that this method of increasing gain will be ineffective in a forward wave tube until selective π mode attenuation is introduced. The existence of π mode oscillations at low RF input drives presently

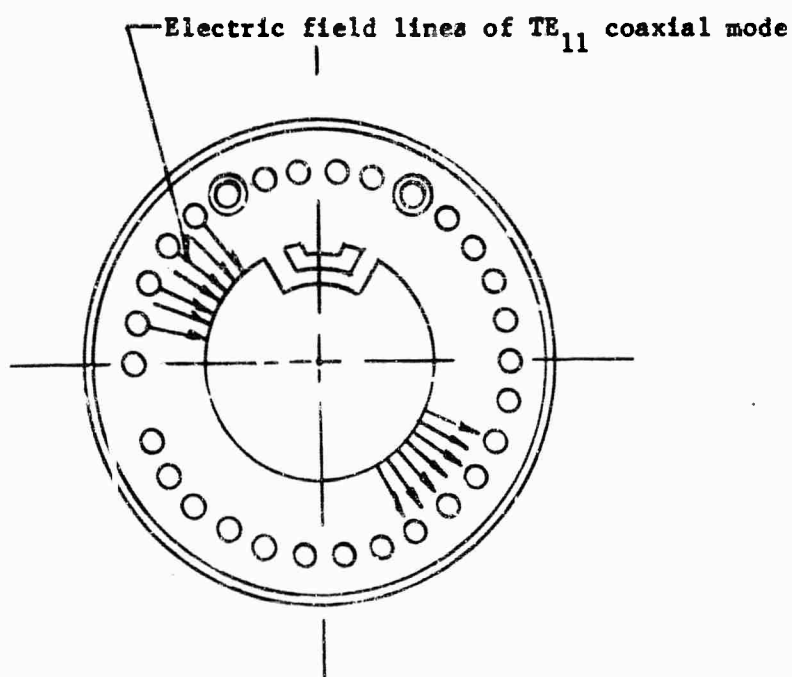
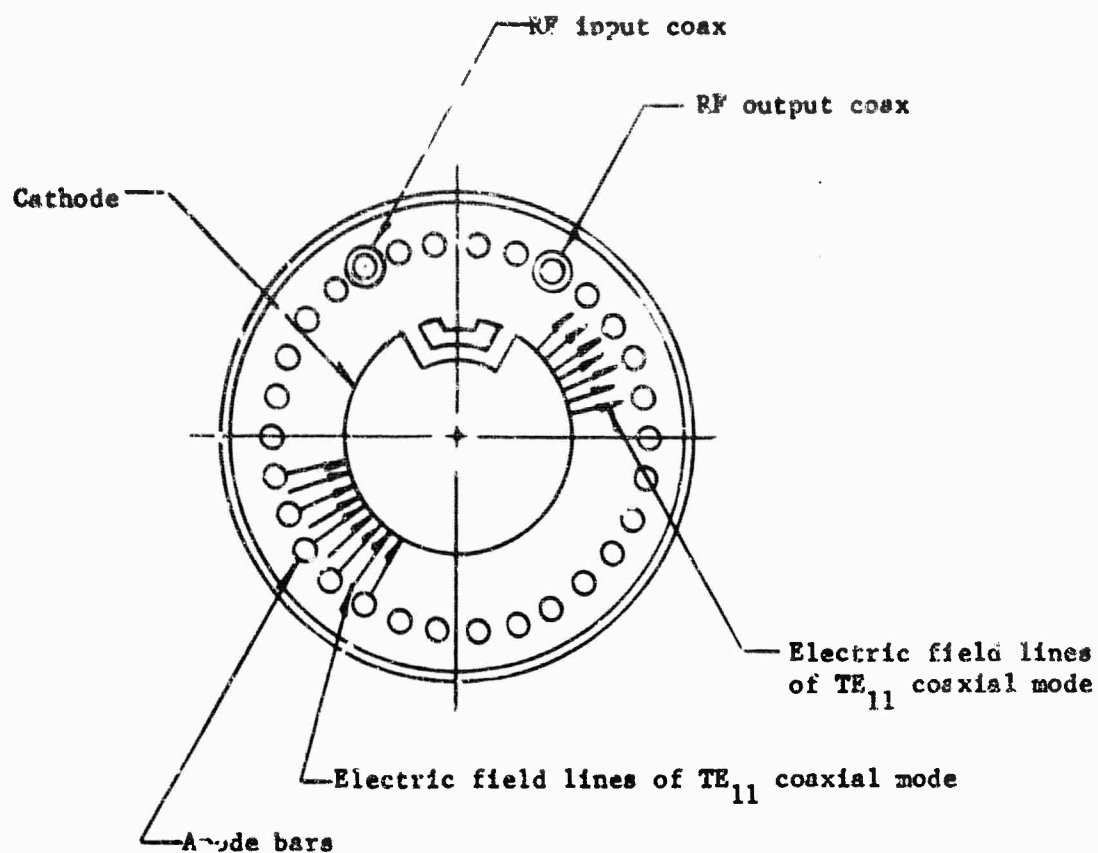


FIGURE 35 ELECTRIC FIELD PATTERNS OF THE TE_{11} MODES IN THE INTERACTION SPACE. THE FIELDS OF THE MODE DID NOT VARY APPRECIABLY IN THE AXIAL DIRECTION

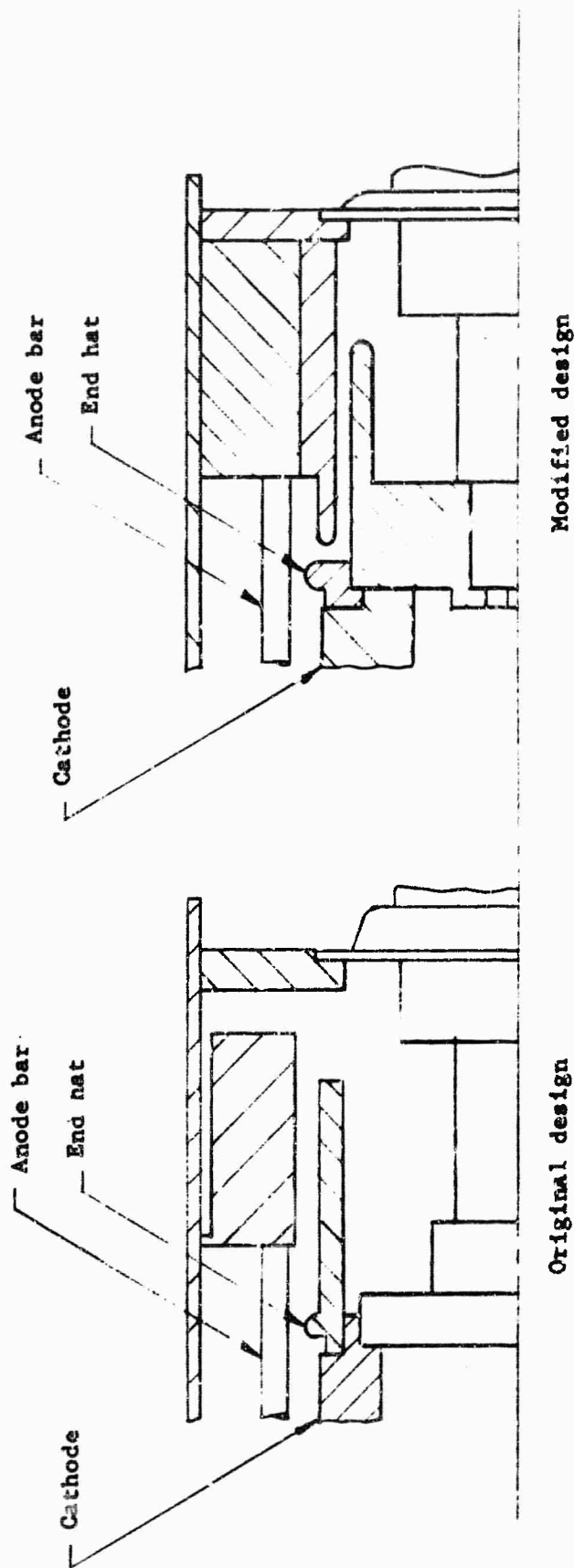


FIGURE 36 ORIGINAL AND MODIFIED DESIGN OF CATHODE END SPACES USED IN B DESIGN TUBES.
 THE MODIFIED DESIGN TUNED THE TE_{11} MODE RESONANCES TO A FREQUENCY ABOVE
 THE PASS BAND OF THE CIRCUIT

limits the gain and interaction space tapering does not weaken these oscillations. Studies showed the energy dissipated by the control electrode during turn off is a small fraction of the total energy per pulse drawn from the dc supply even for RF output pulse lengths of the order of tenths of microseconds. Thus the control electrode pulsing procedure which was originally designed for use with long pulse lengths was shown to be applicable to short pulse lengths.

The first of the C design anode circuits became available in September 1963. The bandwidth was as expected, but the π mode frequency was slightly higher than the design values. The required correction was made by placing an insert in the vacuum envelope to effectively load the circuit so as to lower the π mode frequency.

When the first C design tube (K19E) was tested, it showed difficulty in starting. The tube could be made to start only over a very limited range of operating parameters, specifically with high RF drive and high magnetic field at the low frequency end of the band. Even when starting did occur, there was often a delay of up to a microsecond before the tube turned on and, in some cases, resulted in large amounts of jitter in the turn-on time. Once the tube had turned on, it appeared to be operating normally. Similar results were obtained with the second C design tube (J10E).

To improve starting performance, the control electrode was redesigned. The new control electrode shape resulted in a decrease in the electronic feedback while at the same time permitting a smoother flow of electrons through the drift space during the starting interval.

The third Group C driver tube (K6E) was constructed with the new control electrode design. Its starting performance was markedly improved.

Life tests were started with the Group C tubes during November 1963. The first tube (J10E) operated for a total of 750 hours at the 200 watt average power level. This tube did not have the final stepped control electrode geometry and consequently it showed a marked starting delay and a slight leading edge jitter when it was placed on life test. The tube was removed when the leading edge jitter became more severe. The second tube (K6E) operated for 233 hours when a fault in the control electrode circuitry caused the tube to go into continuous conduction for a relatively long time. Consequently the tube became gassy to the point where the VacIon[®] pump became inoperable. This tube was subsequently recovered by reprocessing. The third tube (L1E) operated for 1288 hours when it, too, was damaged by a failure in the life test equipment. The fourth tube (L18E) placed on life test had accumulated over 5000 hours of operation and was still operating at the termination of the program. The fifth (D37F) and sixth (L30E) tubes, added so that three tubes were always operating in parallel, had accumulated approximately 2400 hours and 1030 hours, respectively, by the end of the program. The parallel operation of these last three tubes covered 1030 hours. A total of 11,881 life test hours were accumulated during the program.

Seven tubes incorporating some of the above-mentioned design changes were completed and designated as Group D tubes. Tubes in this group were tested and evaluated against the specification. The results are discussed in detail in section 6. Increased anode-cathode separation was studied and showed that improved operation in the region of lower power could be obtained without sacrificing performance in the higher 150 kilowatt operating region. Starting delay was measured with the increased separation and was found to be less than 20 nanoseconds. High average power tests up to 1500 watts with a 200 μ sec pulse length were conducted. Phase linearity and

phase sensitivity measurements were made. Evaluation of this group of tubes concluded the program.

5.0 GROUP C TEST RESULTS

5.1 Driver Tube Construction

A group of eight C design driver tubes were constructed and tested. Problems occurring in the initial tubes were corrected and are discussed in Section 5.3. The result was the first tube (H19E) being rebuilt twice and the second tube (J10E) being rebuilt once. No further construction difficulties or design faults showed up in the tube construction phase of this group.

5.2 Cold Test Results

When the first of the C design driver tube anode circuits became available for cold testing, the bandwidth was as expected but the π mode frequency was slightly higher than the design value. The required correction was obtained by placing an 0.080" insert in the vacuum envelope to effectively thicken it and thereby decrease the spacing between the C_{13} couplers and the envelope wall. This increased the capacitive loading between the centers of the anode bars and the ground plane, reducing the π mode frequency. In future tube designs, this insert may be eliminated and the vacuum envelope diameter decreased by a total of 0.160" to a diameter of 3.8".

Figure 37 shows an ω/β curve of the final version of the circuit. Propagation between 90° and 180° phase shift per section is a consequence of capacitive coupling from the bar centers to the vacuum envelope (ground plane) via the C_{13} couplers. This type of propagation is analagous to that obtained in a Karp circuit.

Figure 38 shows the transmission through a typical tube and the reflection looking into the input window with a termination placed on the output window of the tube. The reflections recorded are the net consequence of reflections from input and output windows and from input and output coaxial line to slow wave circuit transducers.

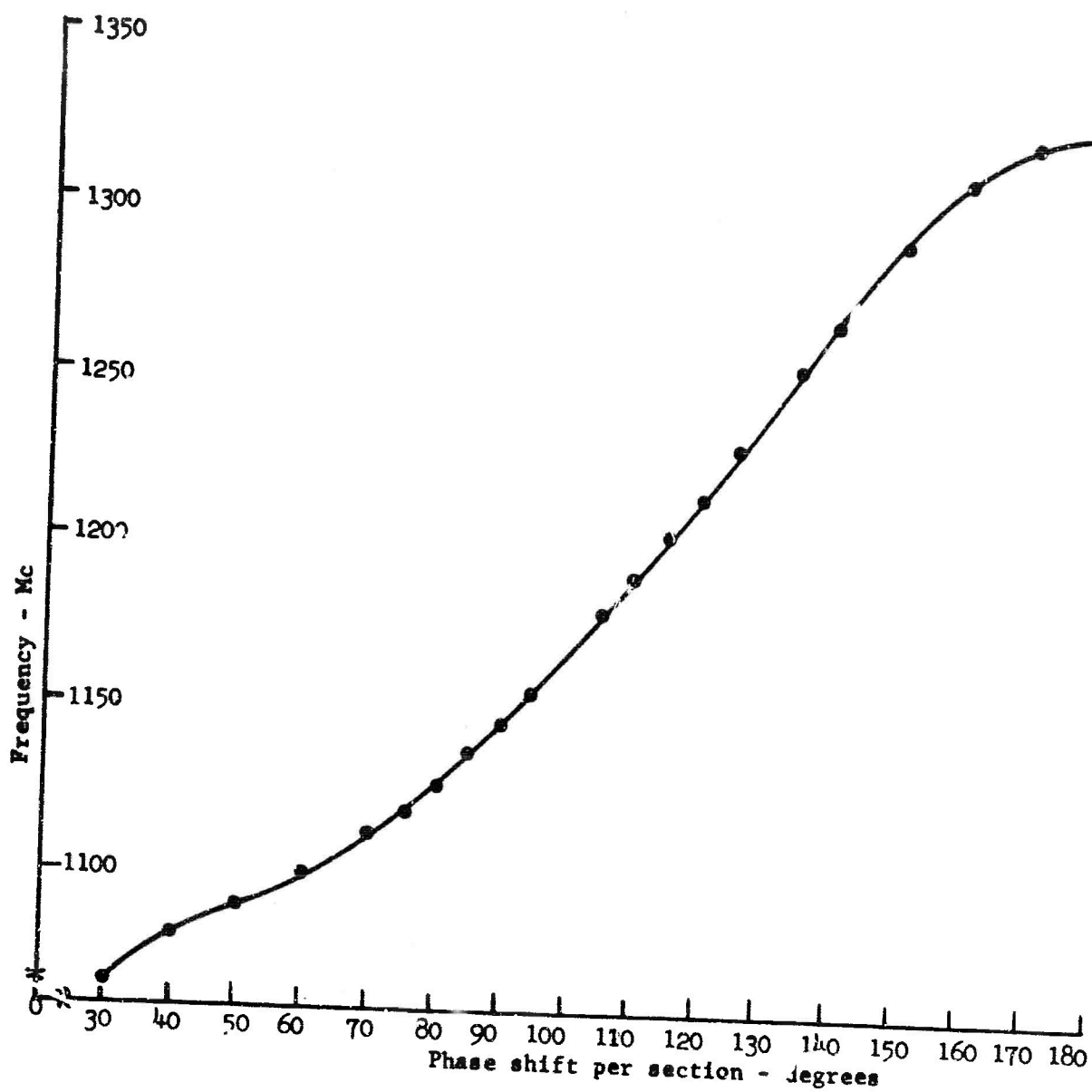


FIGURE 37 ω - β CURVE FOR THE C DESIGN DRIVER TUBES

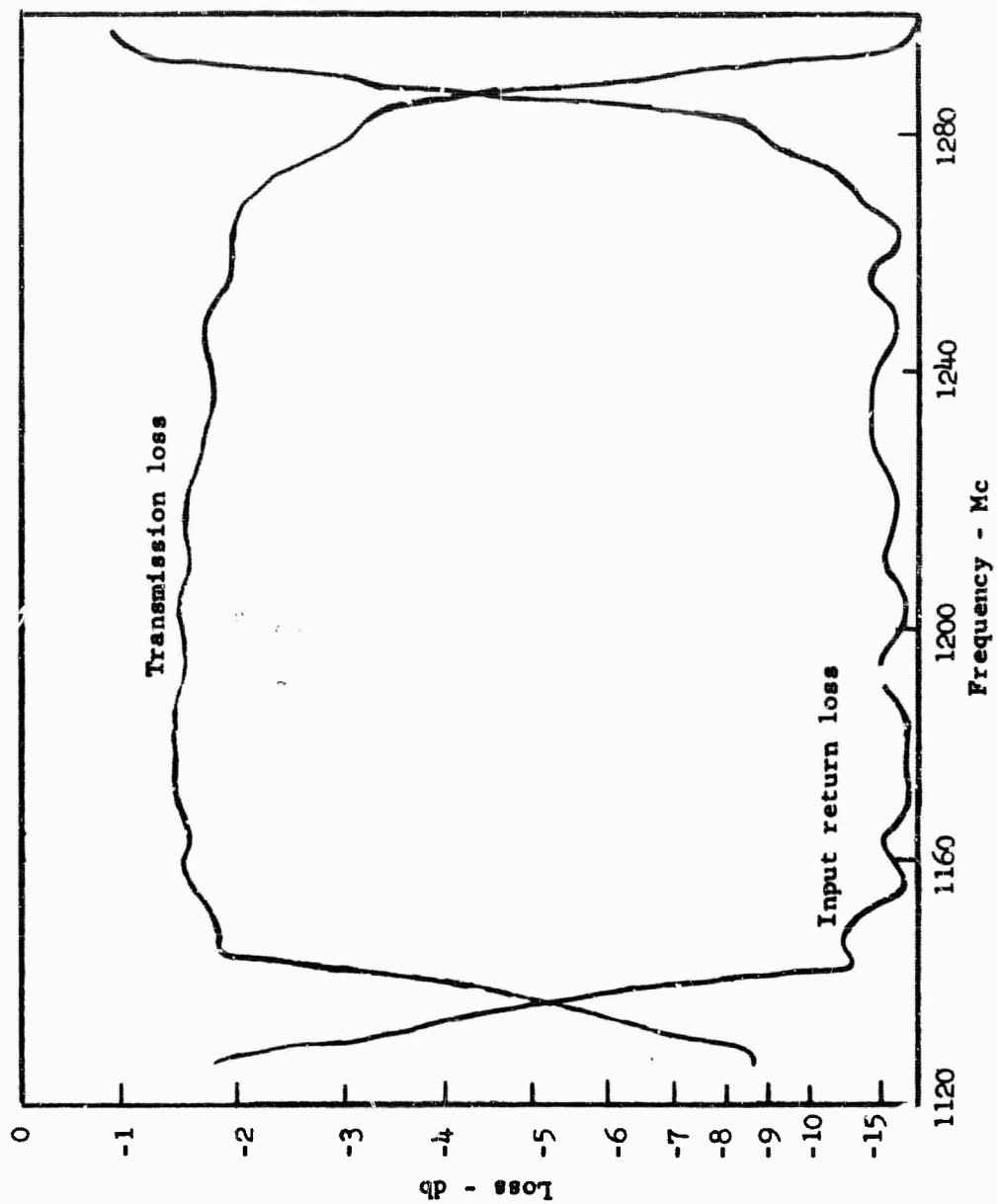


FIGURE 38 TRANSMISSION LOSS AND INPUT RETURN LOSS AS FUNCTIONS OF FREQUENCY FOR C DESIGN DRIVER TUBE

Reflections from the transducers have a return loss of approximately 17 db. Reflections from the window assemblies (including the bend and stub support of the output window) have a return loss of approximately 30 db over the 100 Mc operating bandwidth of the tube. Figure 39 shows the result of return loss measurements on a typical output window and stub support assembly. There is some tube-to-tube variation in the fine structure of the transmission and reflection curves, but no clear pattern of cause and effect has shown up. While both window and transducer matches are already quite good, it is believed that even better results could be obtained with further effort.

5.3 Hot Test Results on the C Design Driver Tubes

The first of the C design driver tubes (H19E) reached hot test during the end of September 1963. The tube operated from a uc voltage for only a few minutes and then an arc occurred. Subsequently, it was not possible to apply more than 2 kv to the tube. This tube was opened to determine the nature of the fault. It was found that the internal cathode support insulator had cracked and that voltage breakdown was occurring through the crack in this insulator. The location of the insulator may be seen by referring to Figure 3. From this figure, it may be noted that, if the support insulator cracks, there is a short path across the crack between metal parts separated by the full anode-to-cathode potential. In the present tube design, the metal cylinder inside the ceramic insulator was a close fit to the ceramic instead of a loose fit as shown in the figure. The ceramic had cracked during bake out as a consequence of differential expansions between the internal metal cylinder and the ceramic insulator. The problem was solved by increasing the clearance between these two parts and it has subsequently presented no further trouble.

The first driver tube was then rebuilt as H19E-1 with the increased clearance between cathode support insulator and its

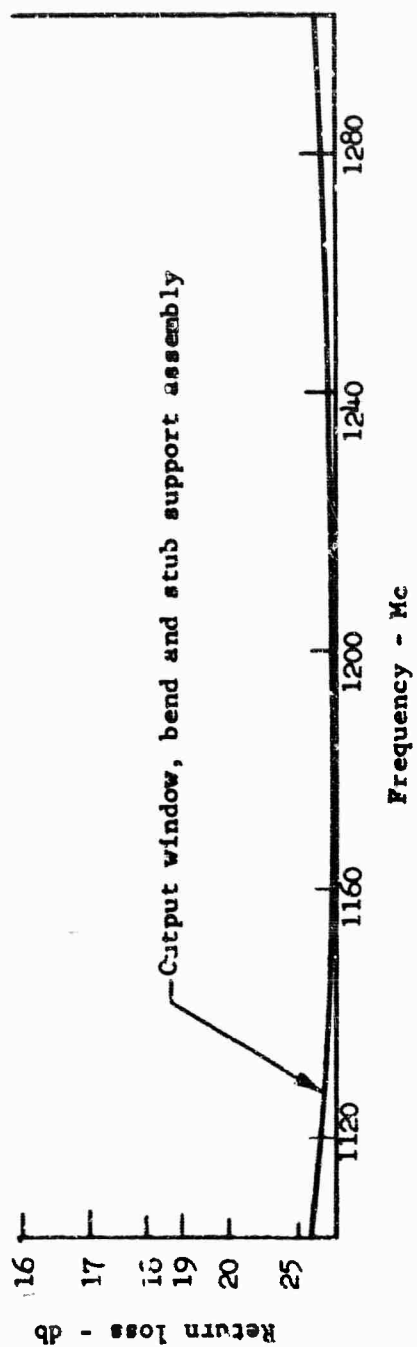


FIGURE 39 MATCH OF OUTPUT WINDOW, BEND AND STUB SUPPORT ASSEMBLY

concentric metal parts. When tested the tube could be made to start only over a very limited range of operating parameters, specifically with high RF drive and high magnetic field and at the low frequency end of the band. There was often a delay of up to a microsecond before the tube turned on and, in some cases, a large amount of jitter was present during turn-on time. Once the tube had turned on, it appeared to be operating normally. Further experimentation with this tube showed that starting could be effected by changes in the control circuit and by holding a magnetic shunt on the outside of the tube in the region of the drift space. When the magnetic shunt was held in just the right manner, the starting delay was decreased and starting improved. Other locations of the shunt resulted in degradation of the starting performance. Similar results were obtained with the second C design tube (J10E).

The results with the magnetic shunt appeared to indicate that the right type of distortion in the magnetic field in the drift space could assist electrons in getting through this region during the starting interval. The drift space in the C design driver tube is longer than it had been in the earlier A design. These two designs are compared in Figure 40. Typical trajectories of electrons traversing the drift space during the starting interval are sketched in the figure. These trajectories are for electrons so phased as to gain energy from the RF wave at the tube output. During starting, the majority of electrons will be bunched in this phase since only these electrons be continuously returned to the cathode and thus multiplied by secondary emission. The first time these electrons strike the control electrode, they do so with appreciable energy and produce many secondary electrons. In the drift space, however, there is no mechanism by which these newly generated secondary electrons can gain energy from the RF fields. They thus return to the cathode with low velocities and produce only a small number of secondaries. Thus for

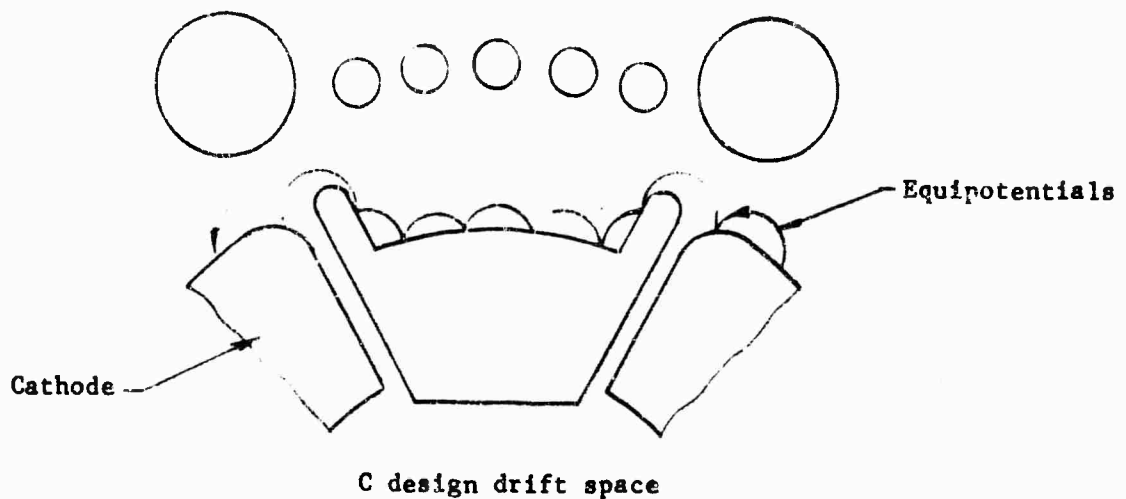
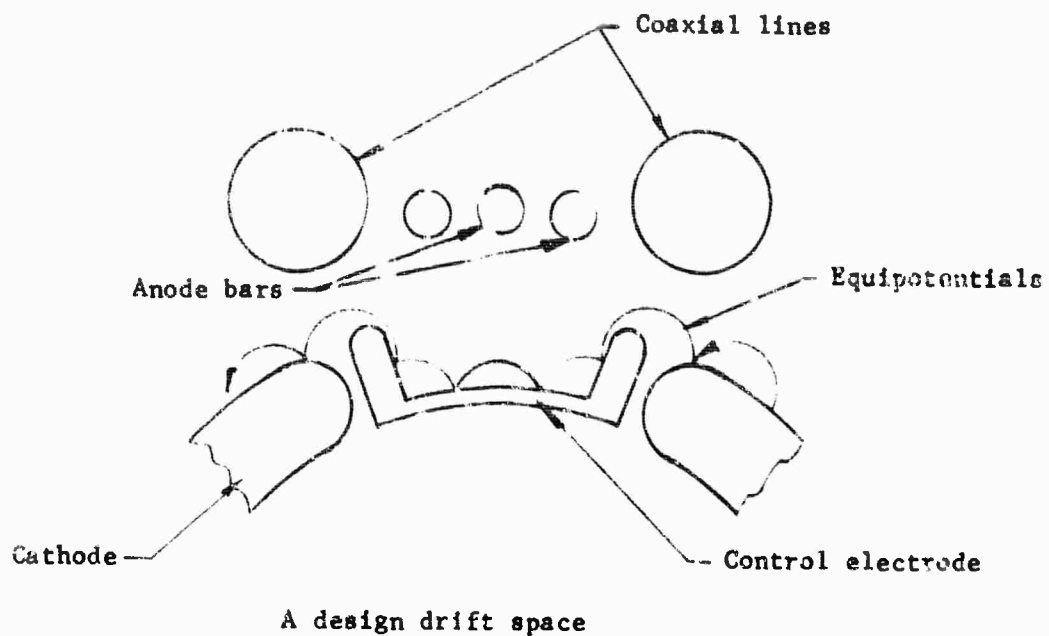


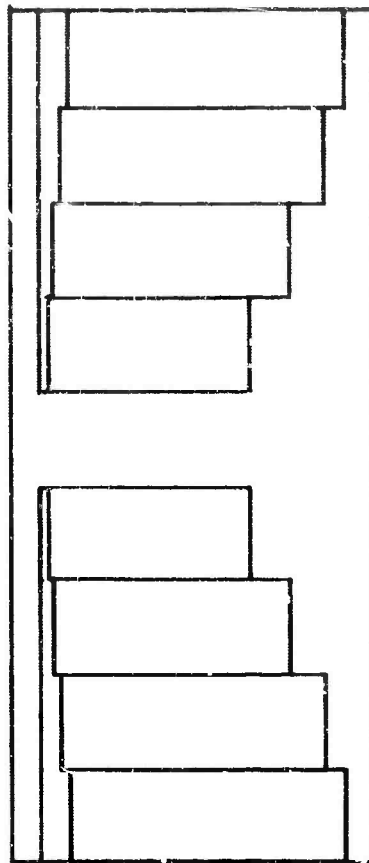
FIGURE 40 COMPARISON OF A AND INITIAL C DESIGN DRIFT SPACES

The longer control electrode recess in the C design permitted more strikes of electrons on the control electrode during the starting interval and thus permitted fewer electrons to pass through the drift space.

each cycloid in the drift space, there is a large decrease in the number of electrons present. The longer drift space in the new design makes this effect more pronounced and would explain the greater starting difficulties of the new design.

The control electrode in the first Group C SFD-209 (H19E) was made of titanium, which has a low secondary emission ratio, in order to minimize the control pulse voltage required for turn off. It was reasoned that improving the control electrode secondary emission ratio should increase the number of electrons traversing the drift region during the starting interval. With this line of reasoning, the second driver was rebuilt as J10E-1 with an aluminum control electrode in place of the titanium control electrode. Since aluminum has a good secondary emission ratio, less loss of electrons during the starting interval was expected as a consequence of electrons striking the control electrode. The starting performance of this tube was much improved, indicating that this reasoning was correct. The starting, however, was still not as good as it had been with the A design and there was still excessive starting delay over much of the range of parameters investigated. RF performance data were taken on this tube and it was placed on life test.

To further improve starting performance, the control electrode was redesigned into the complex shape shown in Figure 41. This control electrode provides a smooth bridge across its center so that electrons in the center of the interaction space will strike the control electrode a minimum number of times during the starting interval. The remainder of the control electrode recess was changed to a series of parallel ramps. It was reasoned that such a ramp geometry would produce minimum loss of electrons in the drift space. At the same time, through a variation in the anode-to-cathode spacing in the drift space as a function of axial height, a variation was introduced in the drift space transit time of electrons at different



Top view of control electrode

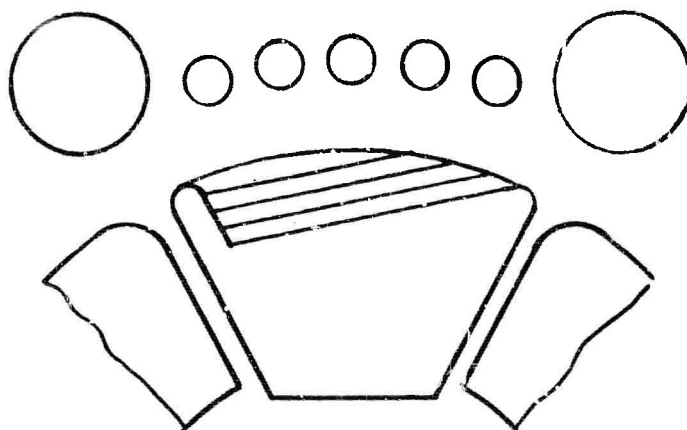


FIGURE 41 SKETCH OF REVISED C DESIGN DRIFT SPACE

Ramp type construction minimizes loss of electrons in the drift space. Differing slopes of ramps cause variation in drift transmit angle as a function of axial position. This results in cancellation of residual electronic feedback.

axial positions. The variation was large enough so that electrons at different axial positions had a transit angle through the drift space which varied by several times 2π . Consequently electronic feedback introduced by electrons at different axial positions was prevented. Introducing this means of eliminating electronic feedback permitted the reduction in the drift space cut back previous used to break up electronic feedback. The new control electrode shape actually resulted in decreased electronic feedback while simultaneously assuring a smoother flow of electrons through the drift space during the starting interval.

The third Group C driver (K6E) was constructed with the new control electrode design. Its starting performance was markedly improved and was judged to be equivalent to that obtained with the shorter drift space in the Group A tubes. Starting jitter was no greater than that of the driver magnetron. With a peak input power of 2.5 kw, an output power of over 100 kw was obtained over a 100 Mc bandwidth. Efficiencies varied from 32% to 37%. These results are shown in Figure 42 and are comparable to those obtained with earlier A design driver tubes. They represented the attainment of the goals which had been set for this C design driver. An attempt was then made to obtain higher efficiency by operating at higher magnetic field. This required an increase in drive power to 5 kw to obtain locking. However, output power also increased so that an output power of about 160 kw at efficiencies of between 44% and 49% were obtained. These results are shown in Figure 43. Gain in this high power region of operation was about 15 db, compared to 16 db gain in the lower power region of operation. Operation in the higher power region appeared to represent a more optimum use of the tube, and it was therefore decided to perform future characterizations and life tests of this design in the higher power region.

A number of performance charts were taken at different frequencies for 3.6 kw and 5 kw input drive. The results are shown

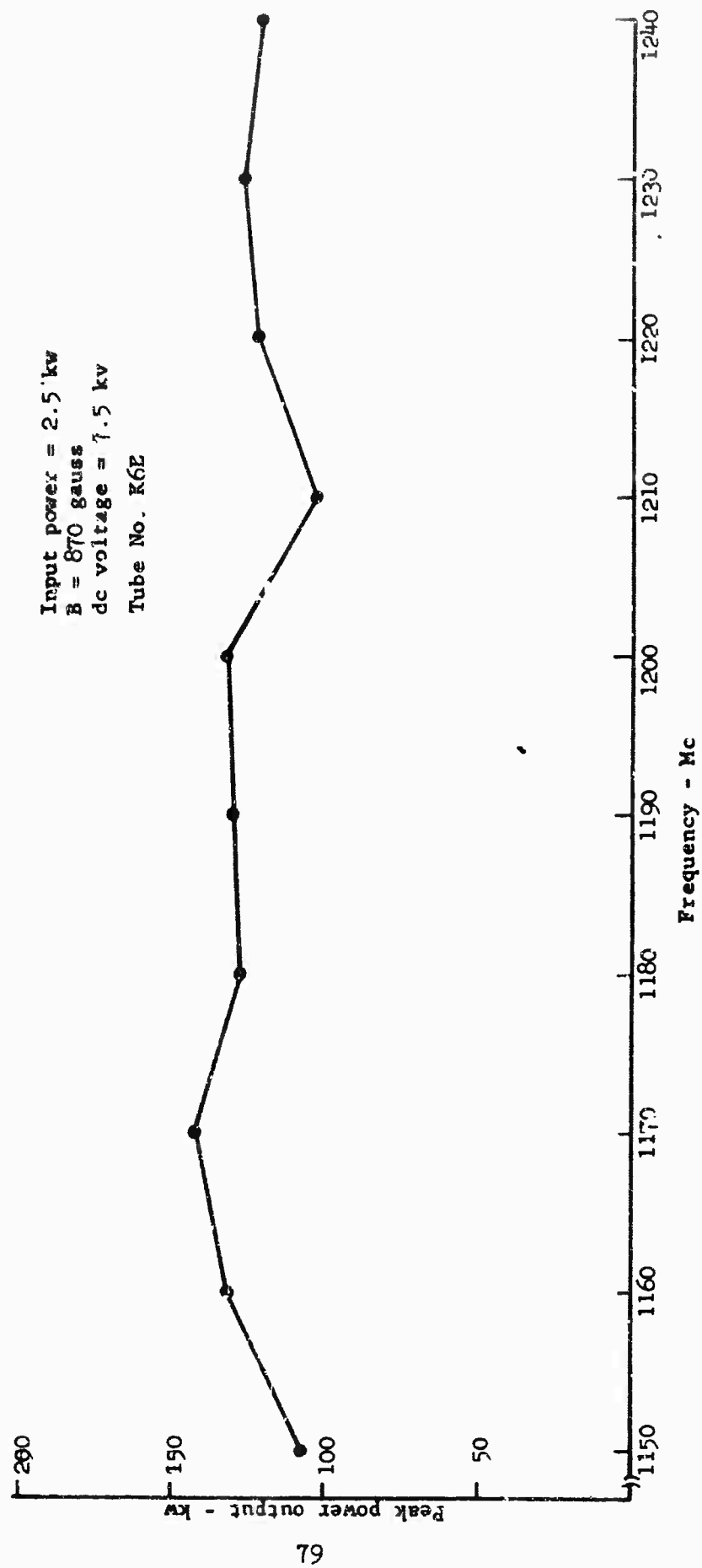


FIGURE 42 PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR SFD-209
WITH 2.5 kw INPUT POWER

FIGURE 43 · PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR SFD-209
WITH 5 kw INPUT POWER

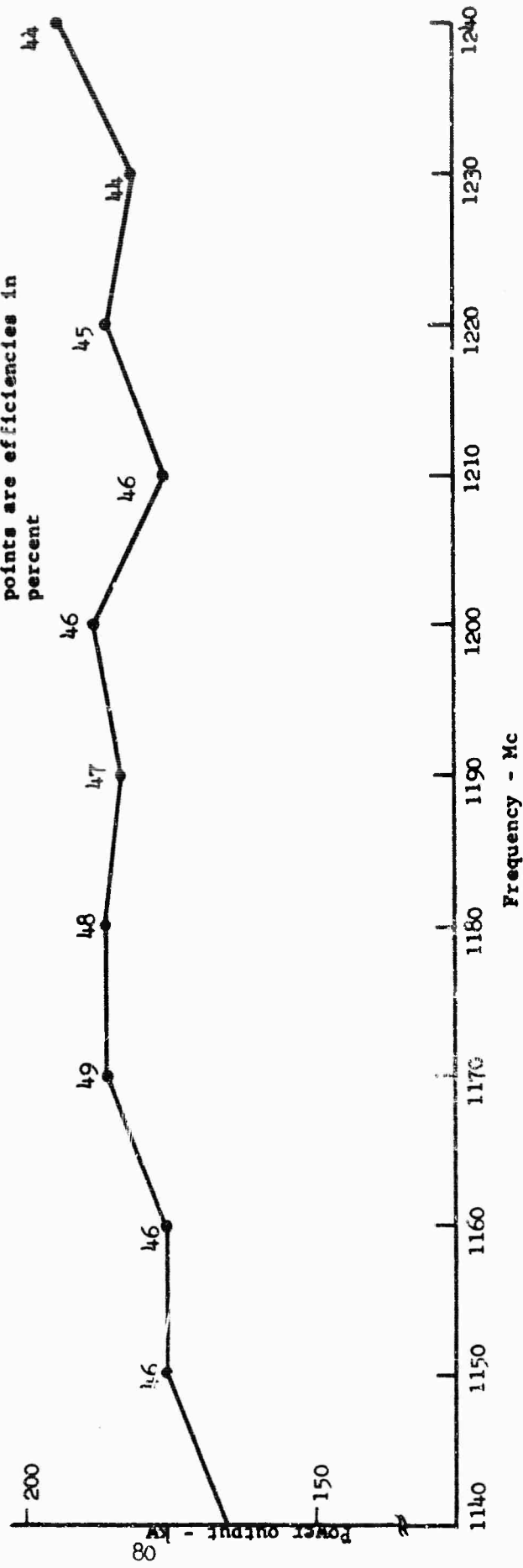
Tube No. K6Z

Input power = 5 kw

B = 1100 gauss

dc voltage = 9.2 kv

Numbers beside experimental
points are efficiencies in
percent



in Figures 44 through 49. Performance changed much less when going from a frequency corresponding to positive electronic feedback to a frequency corresponding to negative electronic feedback than had been the case in the earlier A design tubes. (Performance charts showing the difference between positive and negative electronic feedback frequencies in the A design are shown in Figures 50 and 51.) The design of the drift space and control electrode in these tubes thus appears to be superior to that used in the Group A tubes. Starting is as good as it was in the Group A tubes while the effects of electronic feedback are reduced.

Abbreviated RF tests were conducted on the remaining Group C driver tubes before they were placed on life test. Results obtained are shown in Figures 52 through 55. Tube J10E-1 was the second of the C design driver tubes and, as explained above, did not have the final control electrode design which was first incorporated in the third tube of the group (K6E). Consequently, its performance was not quite as good as that of tube K6E. Tubes L30E and L48E were capable of the same performance as tube K6E but required a somewhat higher voltage for the same output. A decision was made to characterize these tubes at the same voltage as tube K6E with the corresponding lower output shown. The voltage variation problem was later traced to a magnetic field calibration problem, but by that time tube J48E had been placed on life test and it was decided not to interrupt the life tests to rerun the RF measurements. However, tubes L30E and D37F were tested at a slightly higher voltage corresponding to the corrected magnetic field and the results are shown in Figures 56 through 59.

An experiment was then performed to determine if improved confinement of the electrons in the interaction space would improve the efficiency. A change was made in the electromagnet-pole piece configuration so as to provide a more barreling magnetic field thereby

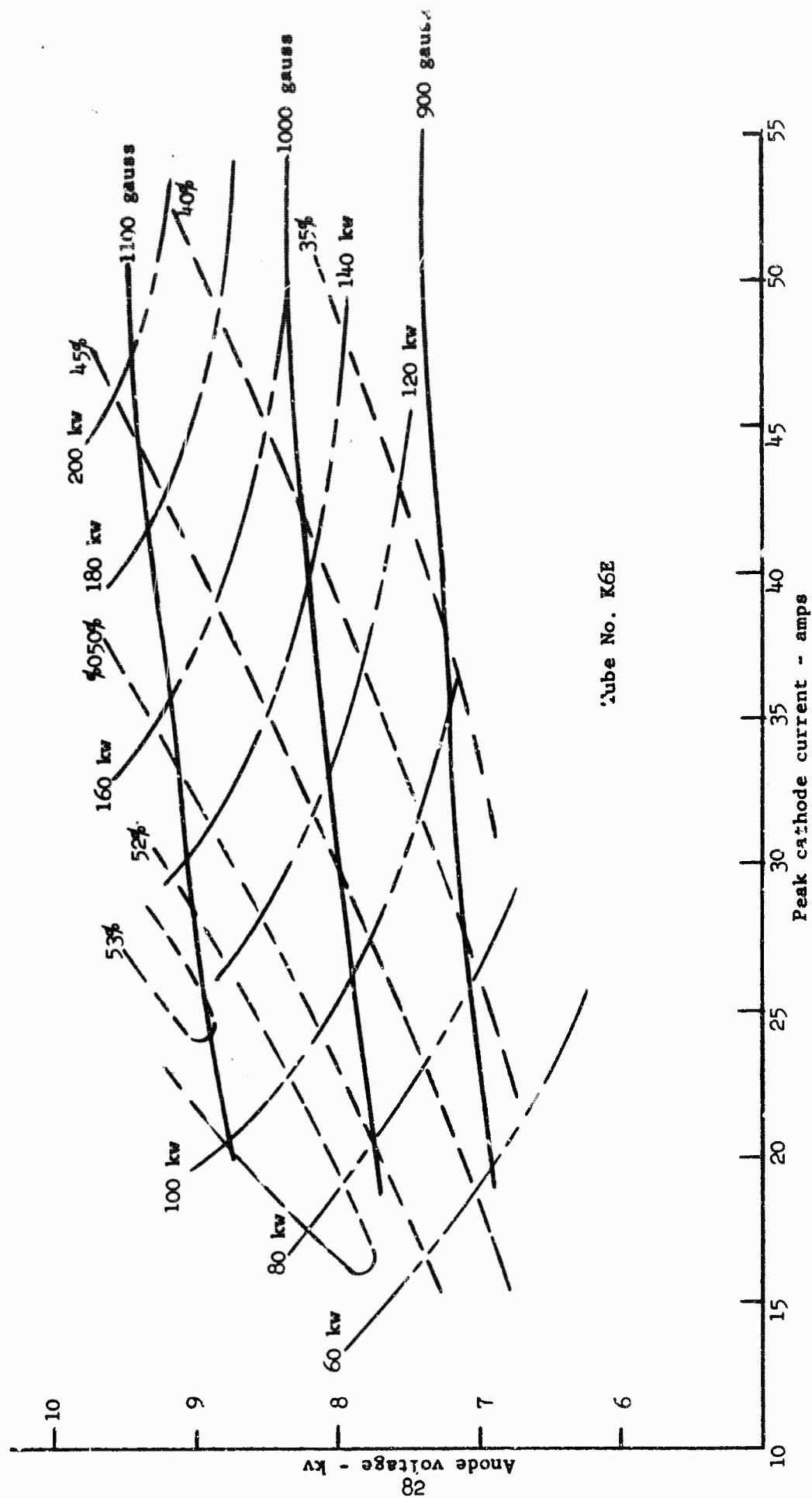


FIGURE 14 PERFORMANCE CHART FOR THE SFD-209 WITH 3.6 kw INPUT DRIVE AT 1140 Mc

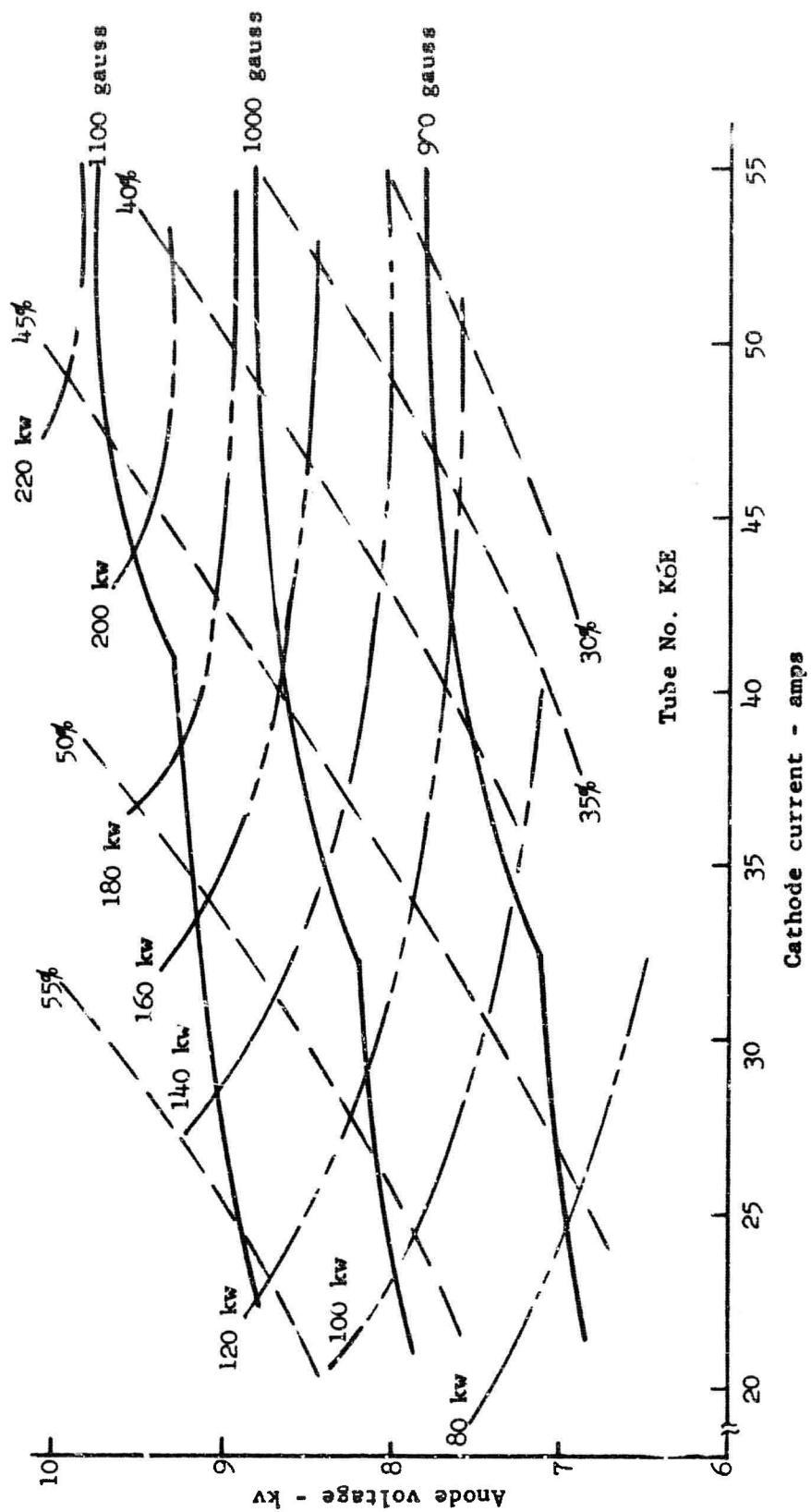


FIGURE 45 PERFORMANCE CHART FOR THE SFD-209 WITH 5 kw INPUT DRIVE AT 1150 Mc

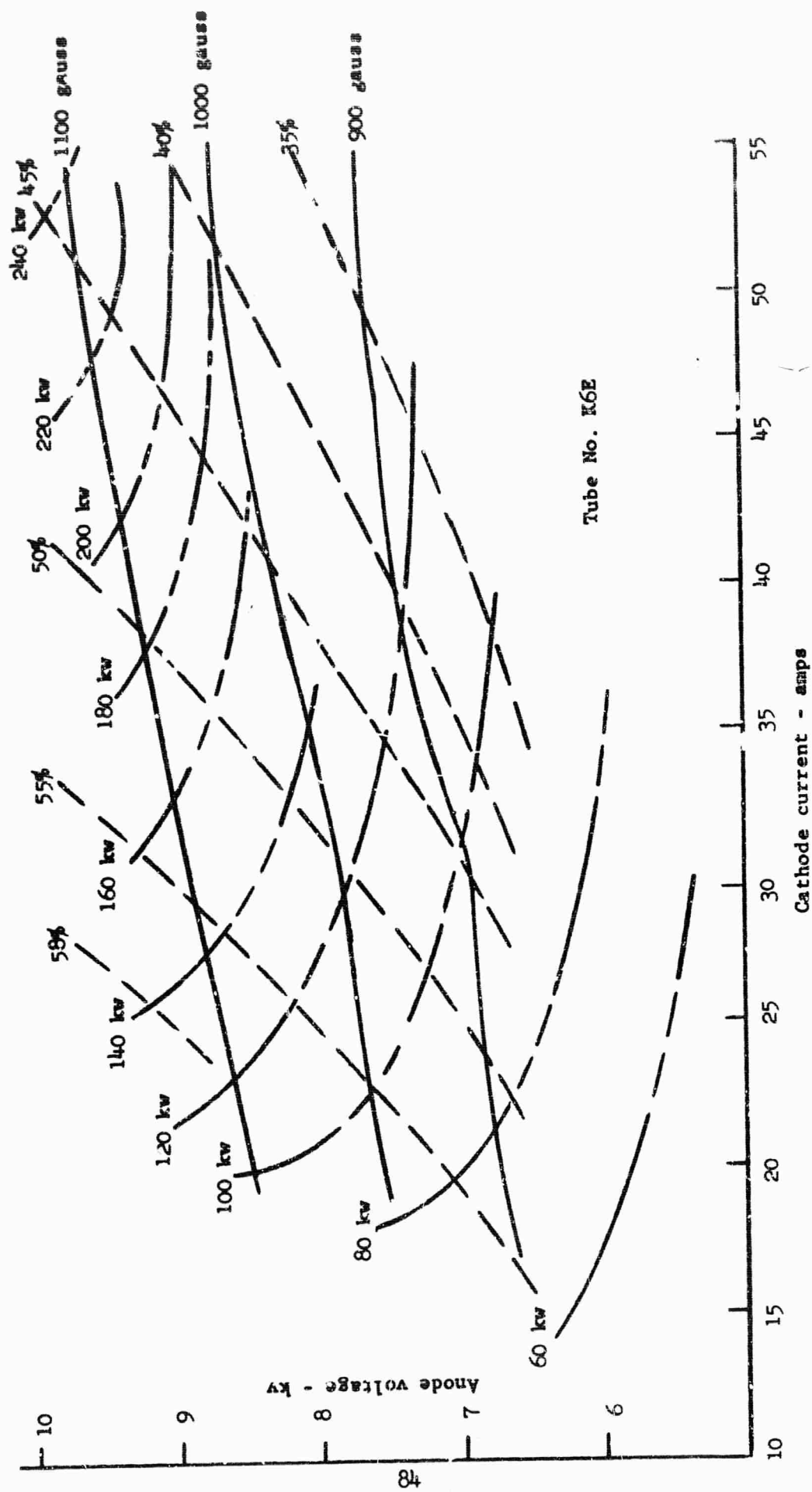


FIGURE 46 PERFORMANCE CHART FOR THE SFD-209 WITH 5 kw INPUT DRIVE AT 1180 Mc

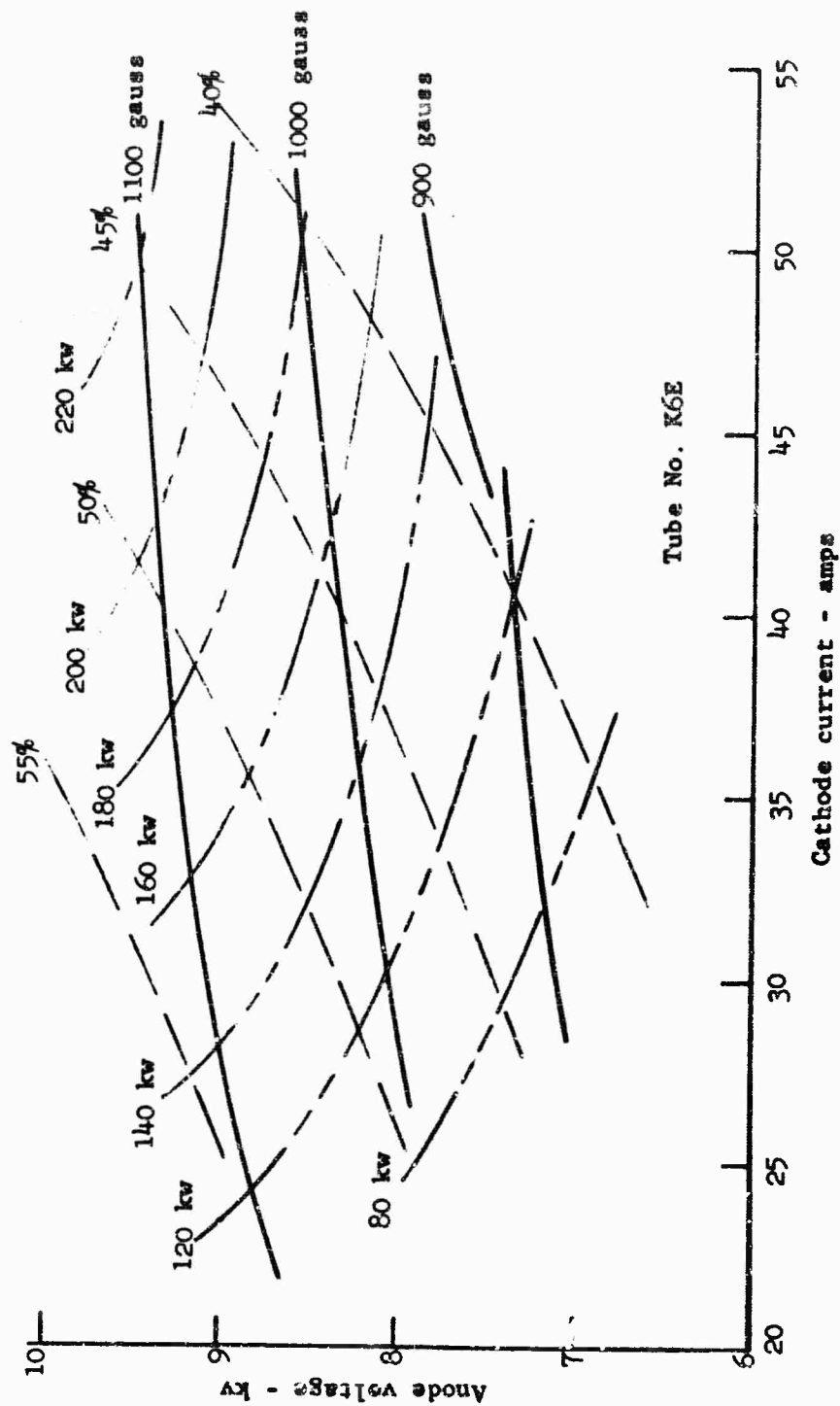


FIGURE 4 7 PERFORMANCE CHART FOR THE SFD-209 WITH 5 kw INPUT DRIVE AT 1190 Mc

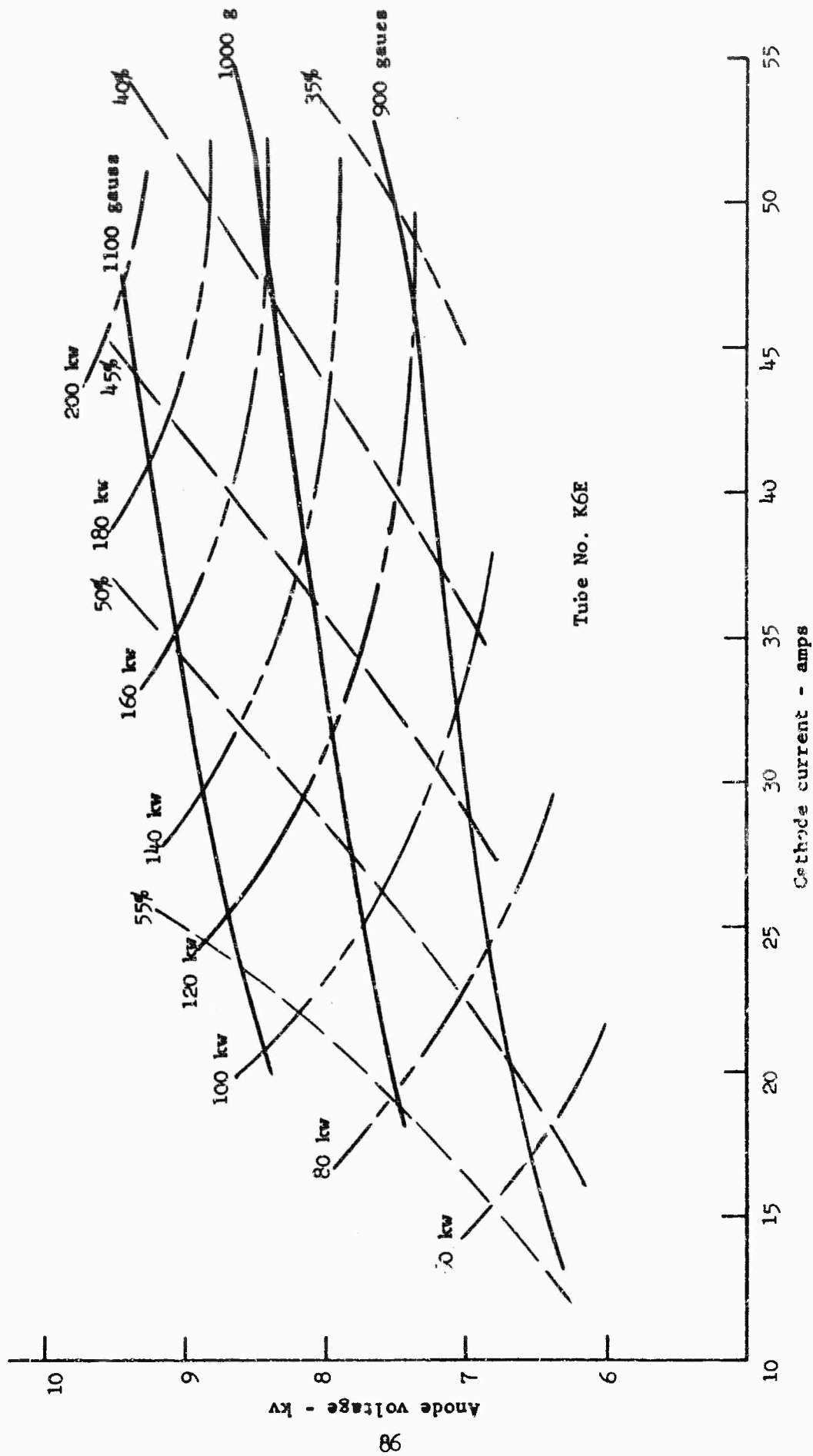


FIGURE 48 PERFORMANCE CHART FOR THE SFD-209 WITH 5 kw INPUT DRIVE AT 1230 Mc

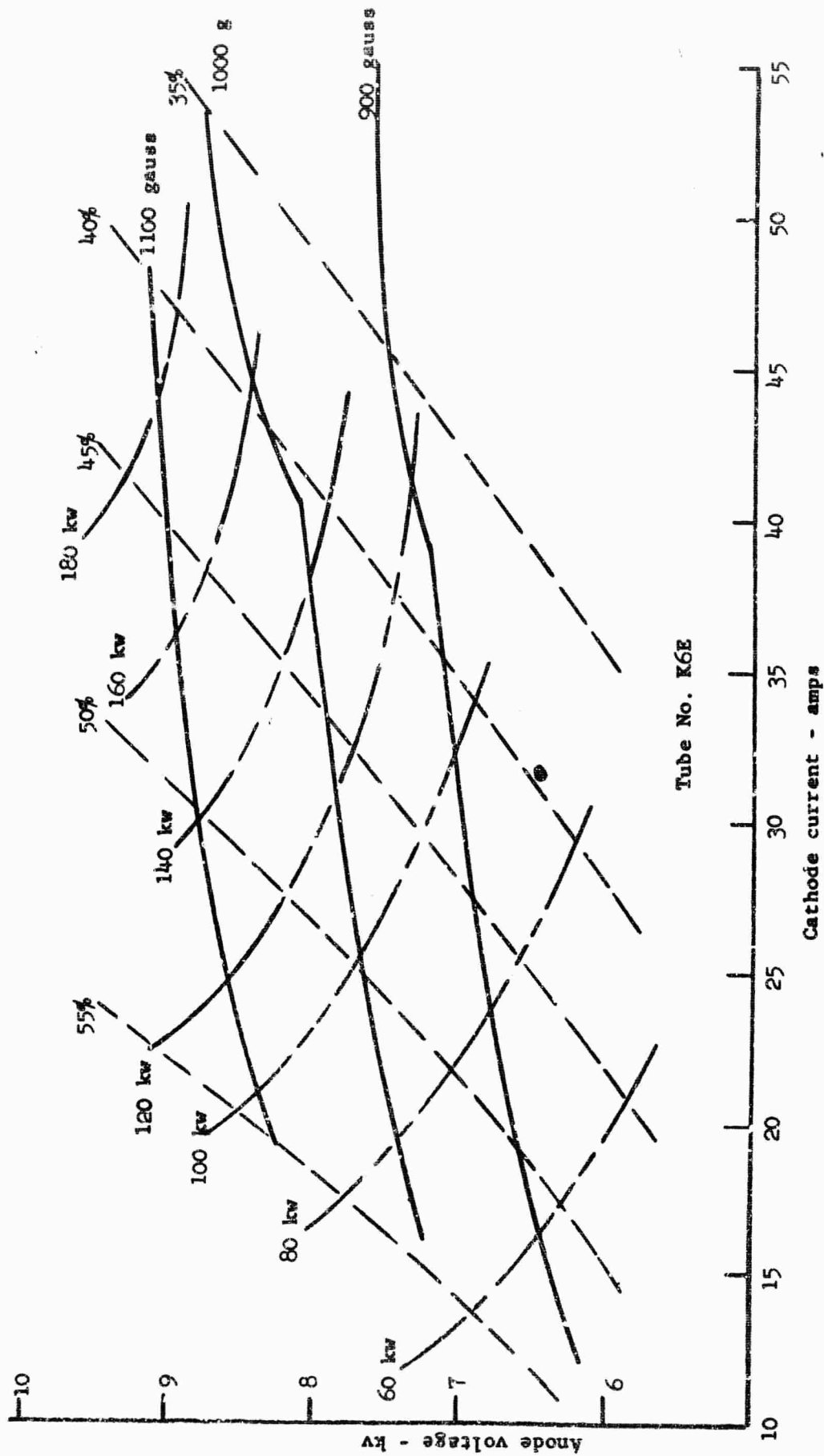


FIGURE 49 PERFORMANCE CHART FOR THE SFD-209 WITH 5 kw INPUT DRIVE AT 1240 Mc

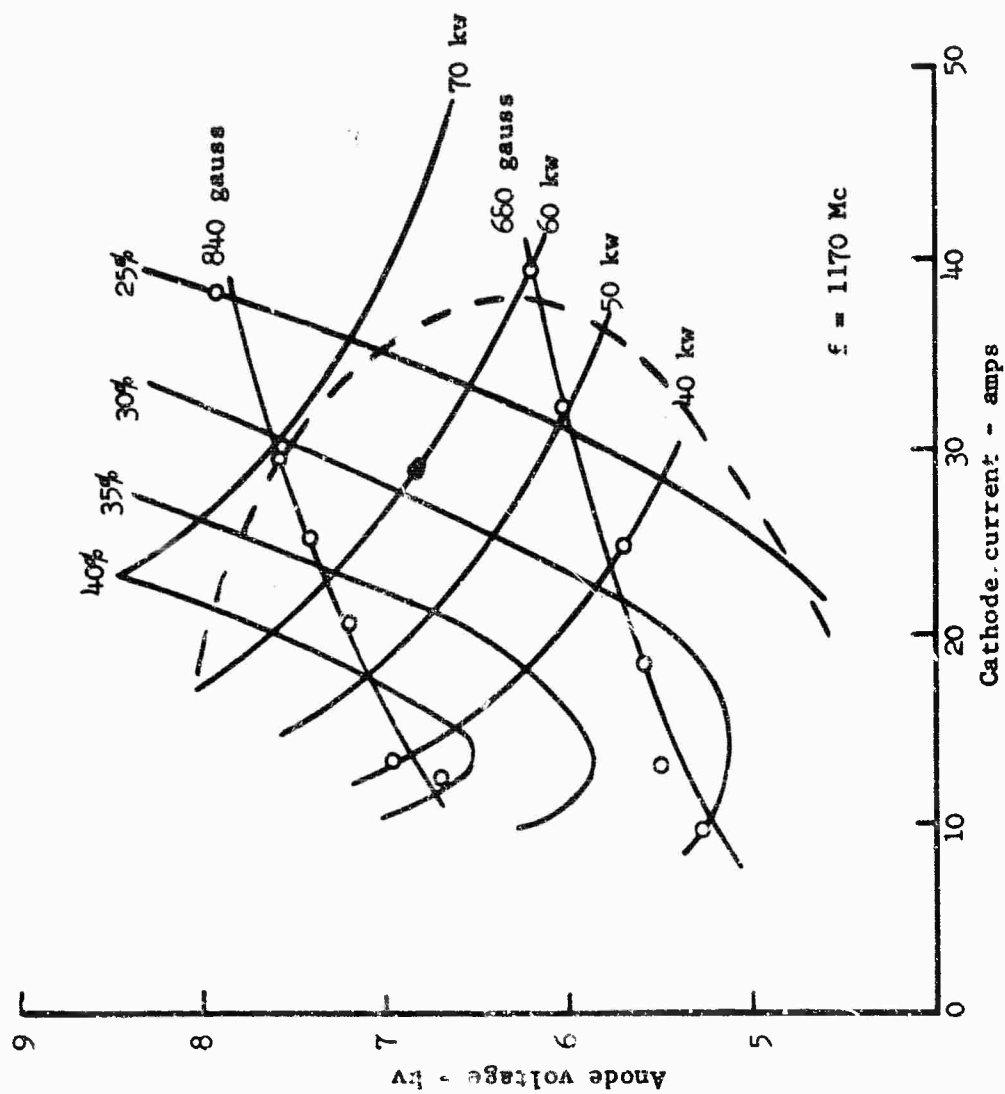


FIGURE 50 PERFORMANCE CHART FOR 1 kw INPUT FOR A FREQUENCY FOR WHICH THE FEEDBACK IS DEGENERATIVE

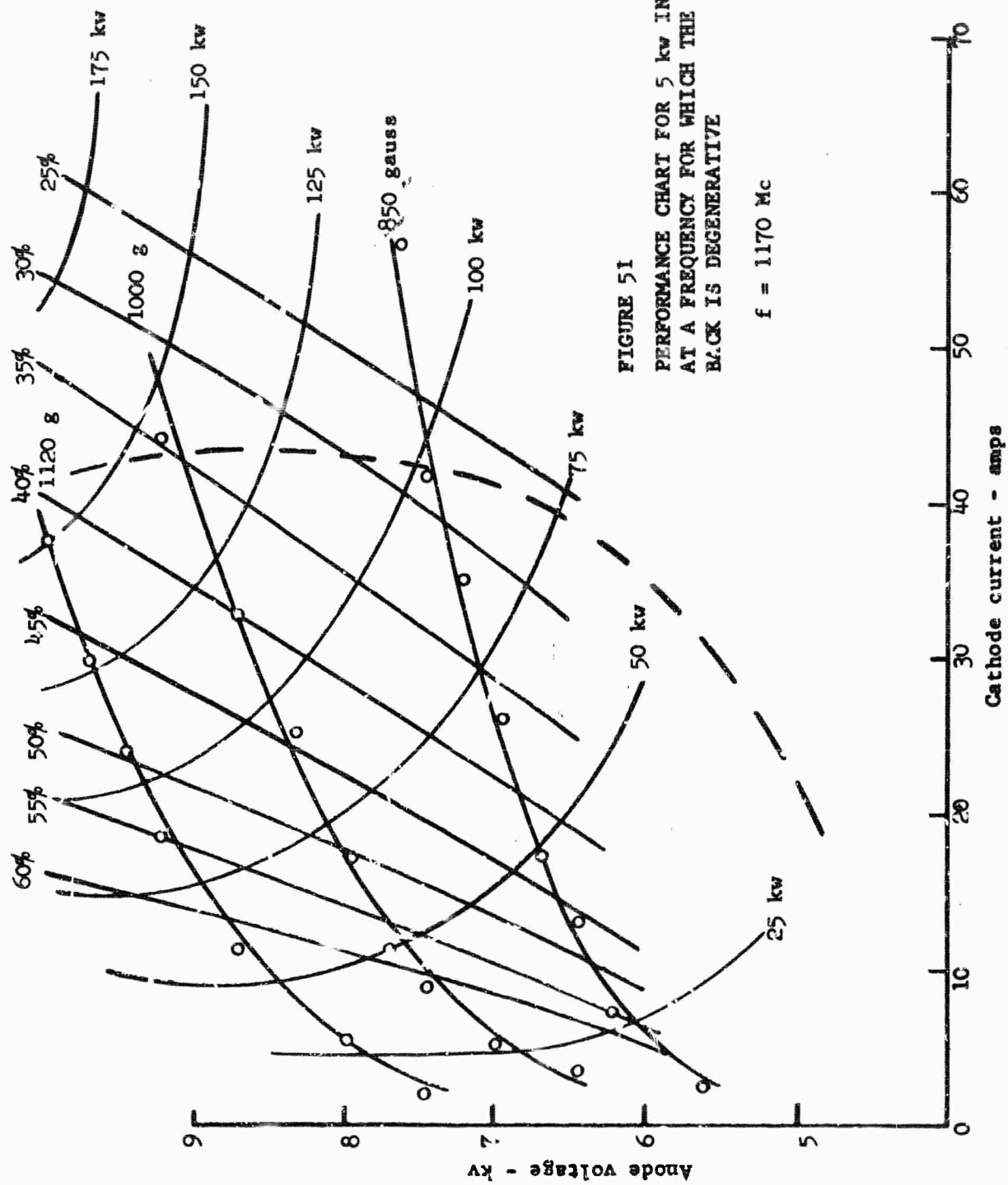


FIGURE 51
PERFORMANCE CHART FOR 5 kw INPUT
AT A FREQUENCY FOR WHICH THE FEED-
BACK IS DEGENERATIVE

$f = 1170 \text{ Mc}$

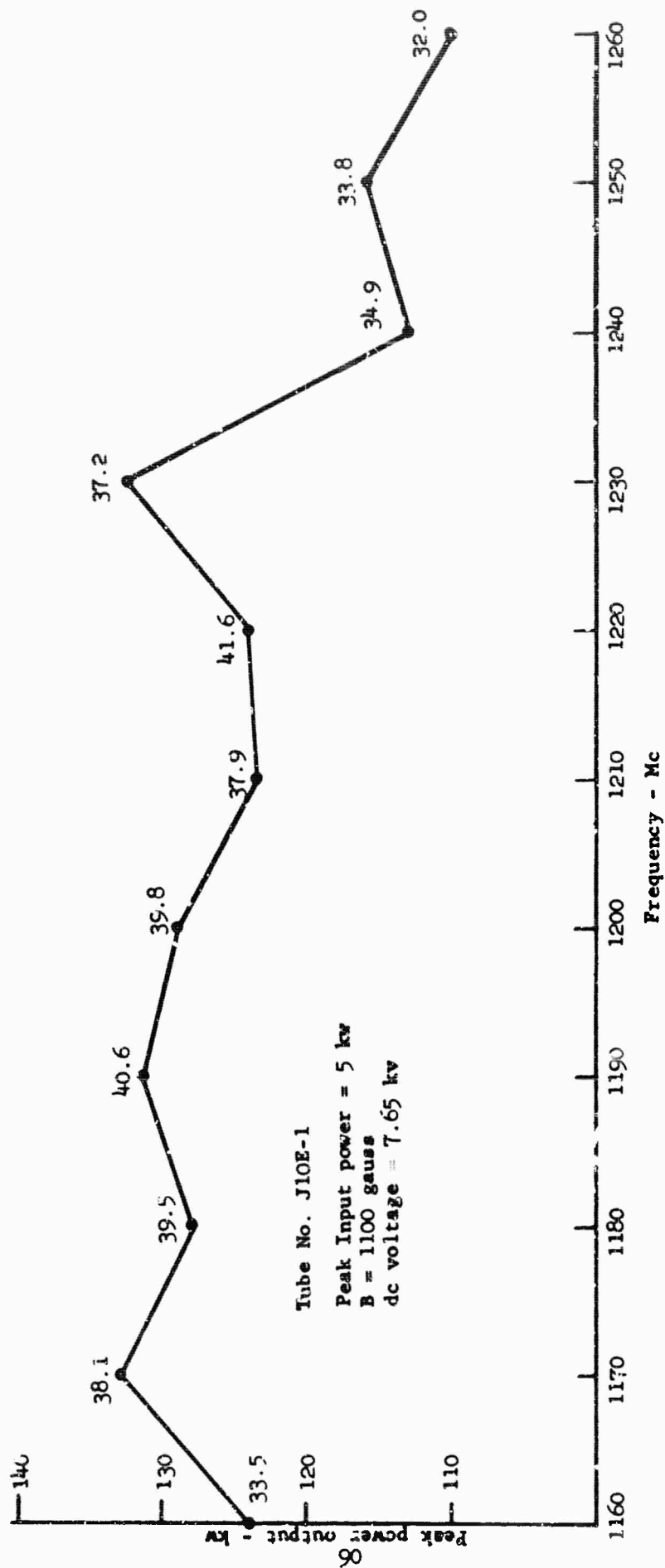


FIGURE 52 PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR THE SFD-209 WITH dc OPERATION

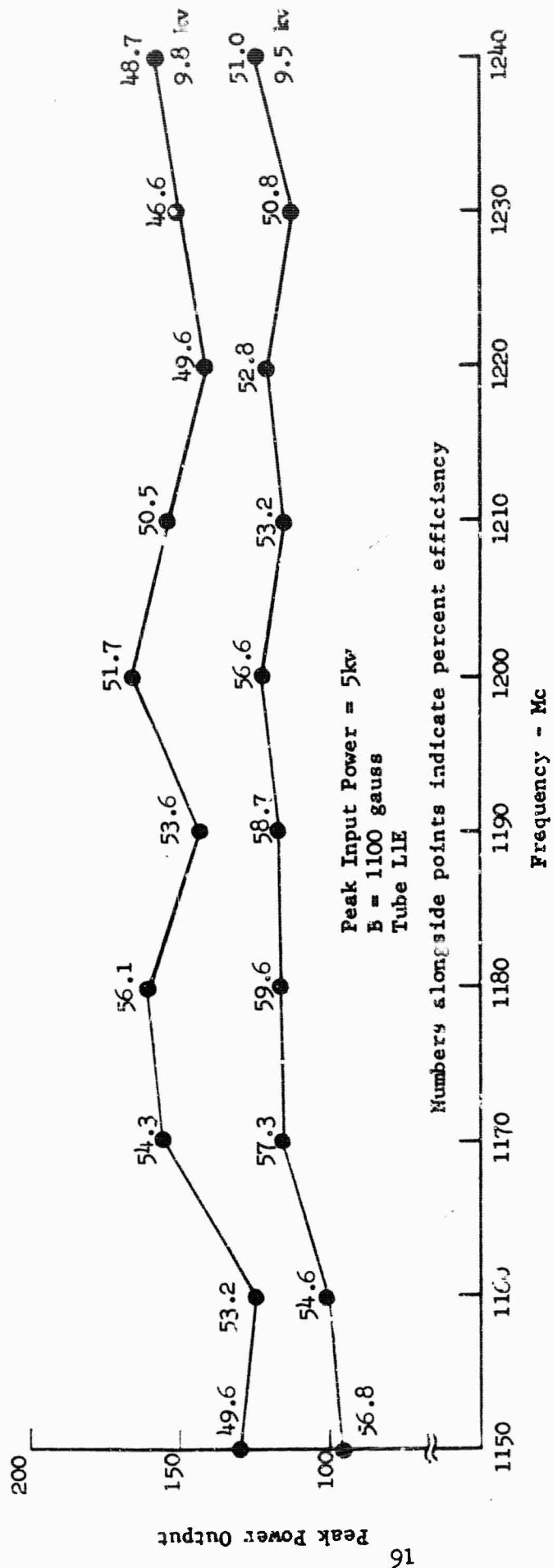


FIGURE 53 PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR THE SFD-209 WITH dc OPERATION

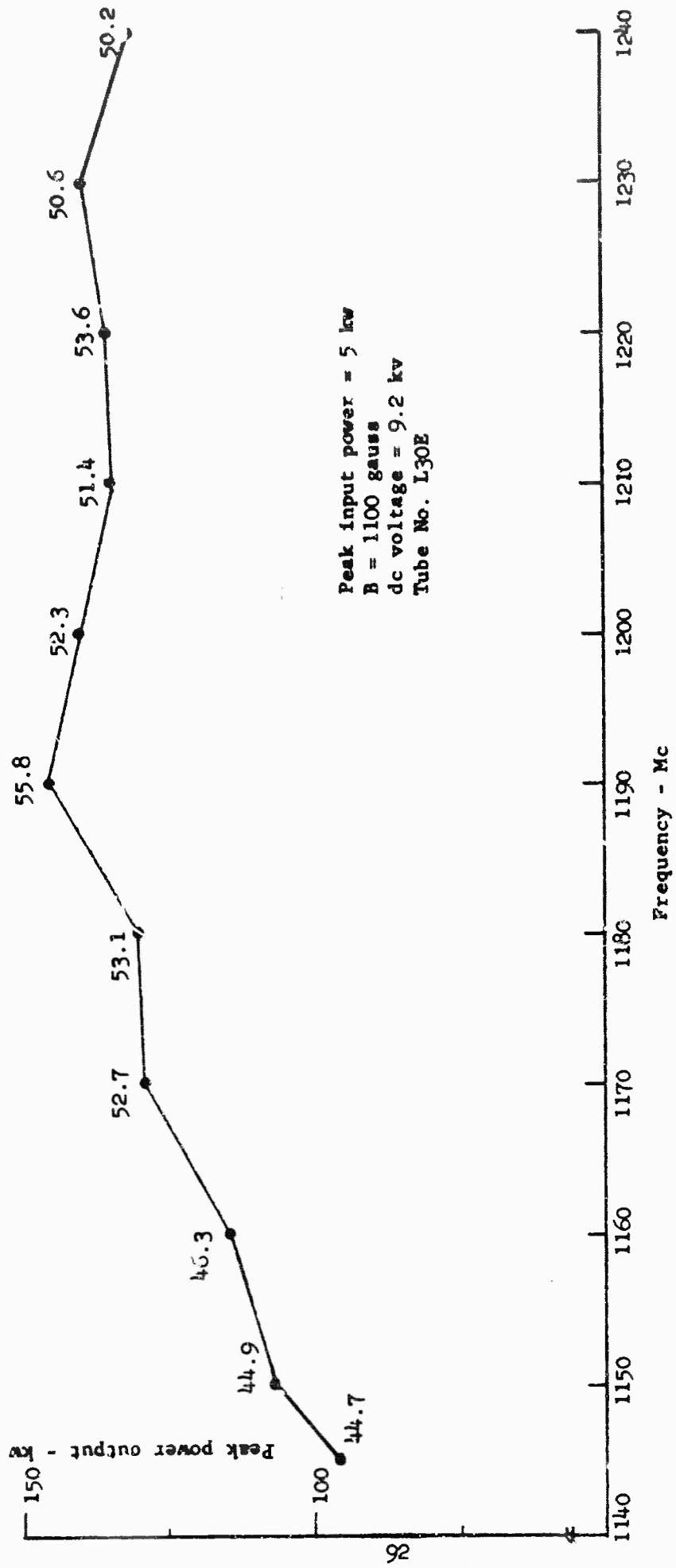


FIGURE 54 PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR THE SFD-209 WITH dc OPERATION

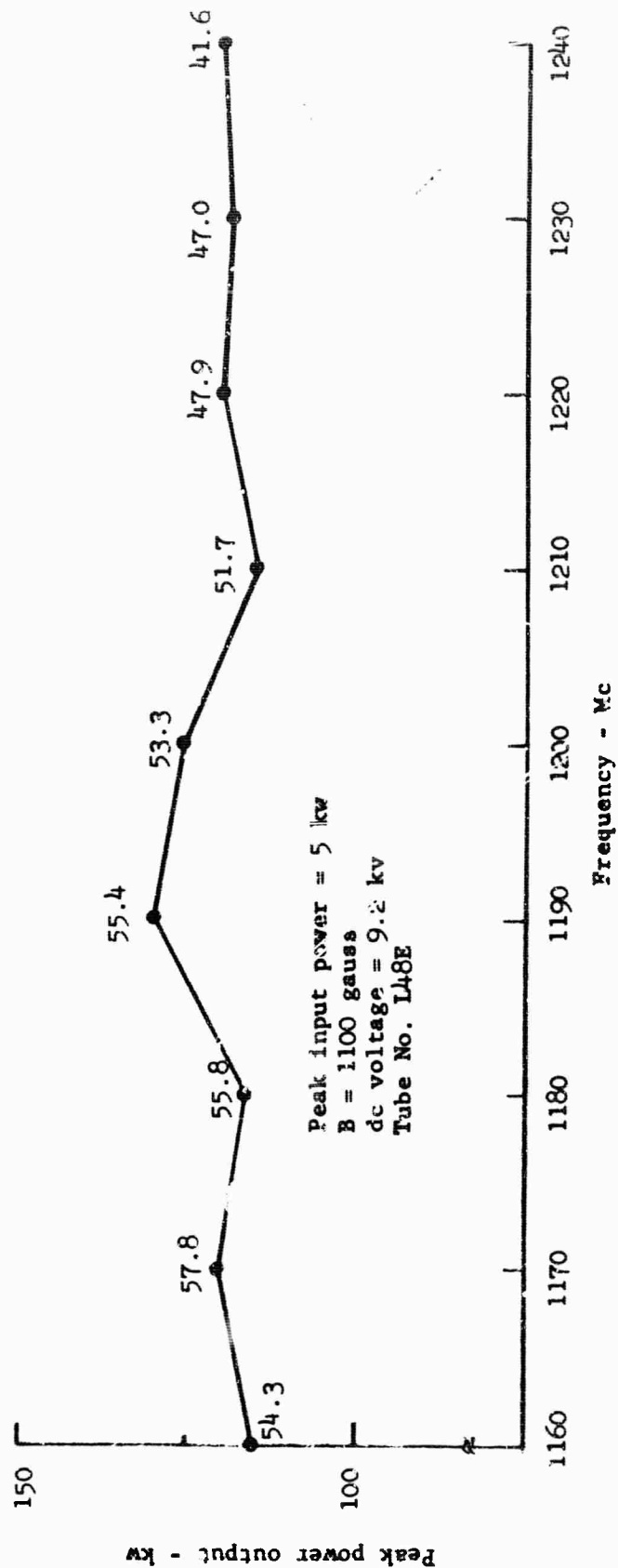


FIGURE 55 PEAK POWER OUTPUT AS A FUNCTION OF FREQUENCY FOR THE SFD-209 WITH dc OPERATION

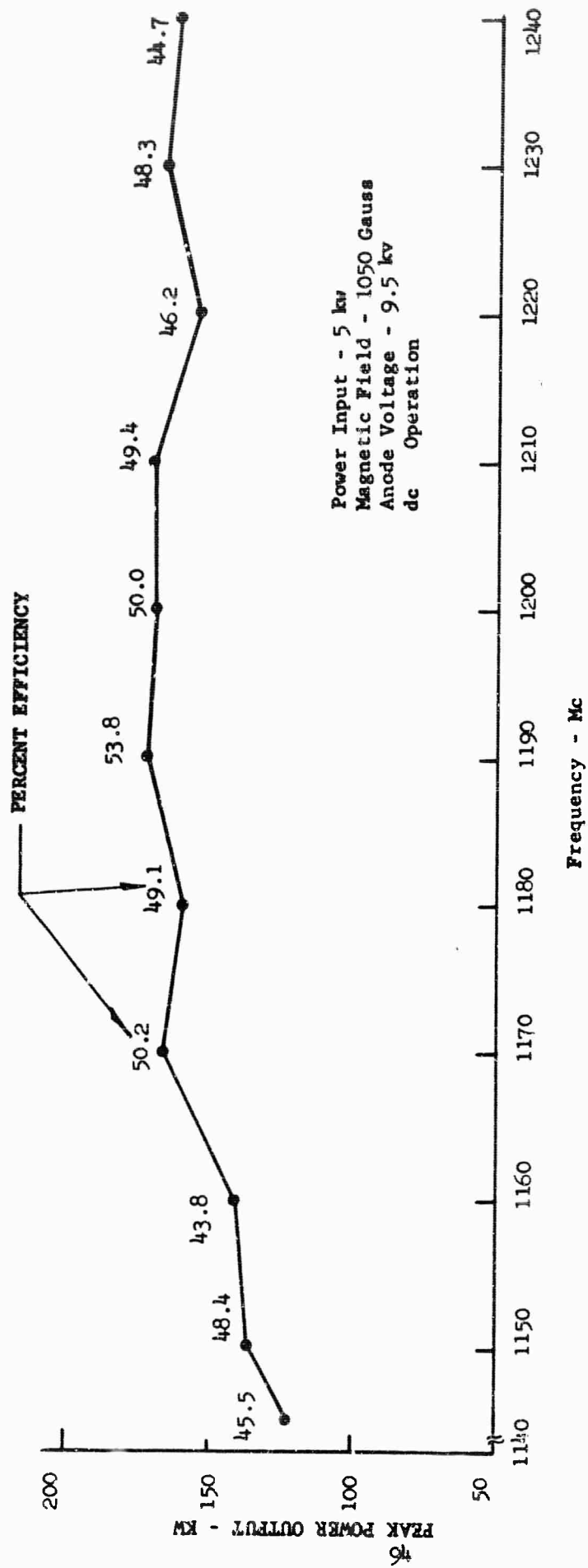


FIGURE 56 SFD-209 TUBE L3OE FREQUENCY RESPONSE

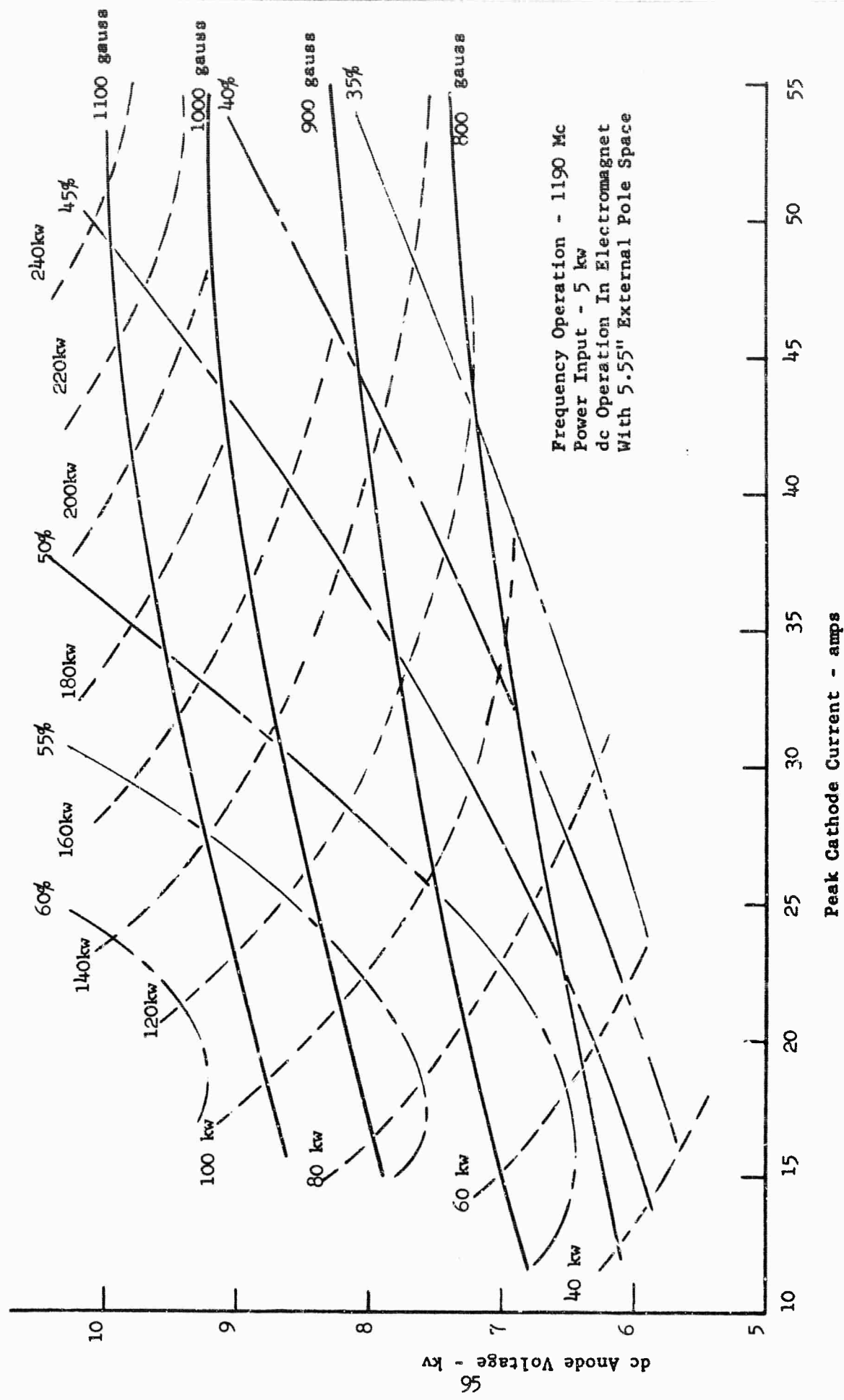


FIGURE 57 PERFORMANCE CHART SFD-209 TUBE L30E

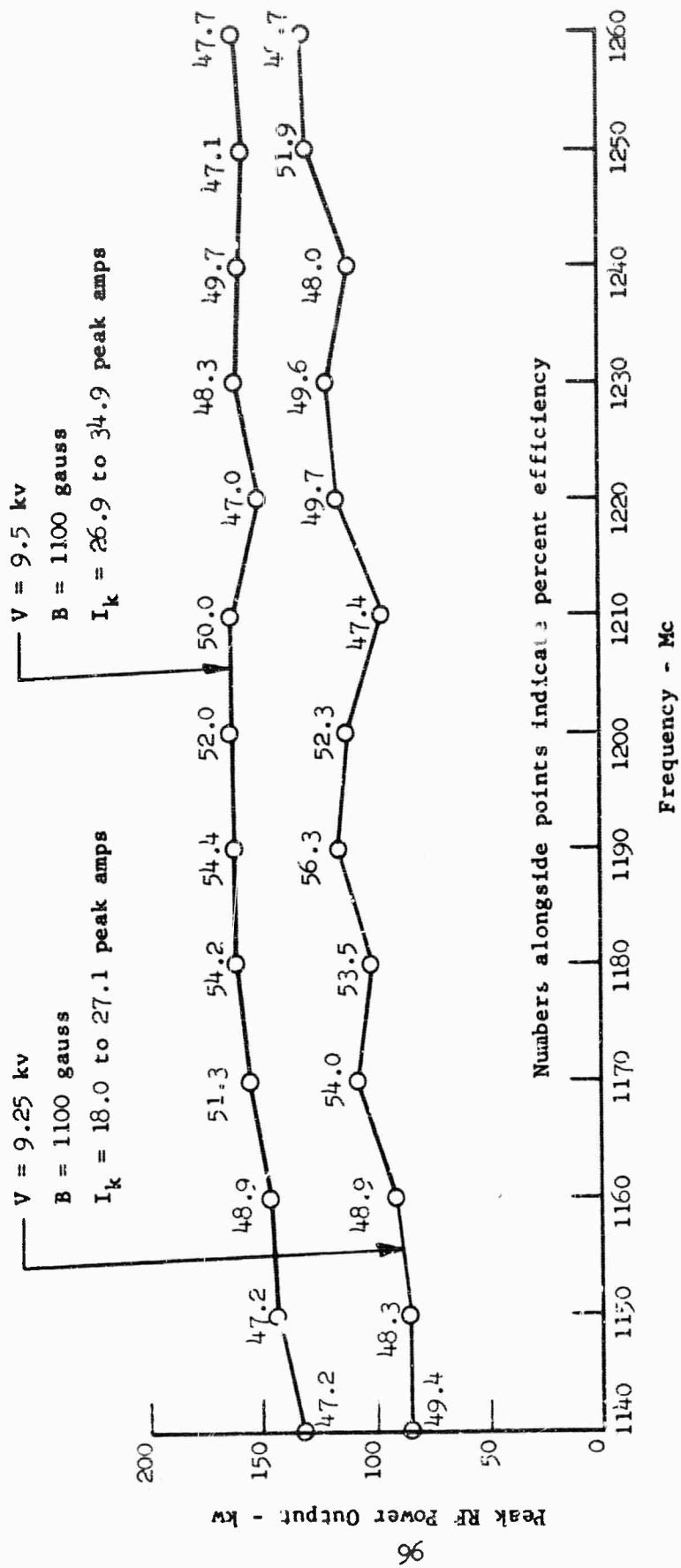


FIGURE 58 OUTPUT POWER VERSUS FREQUENCY AT FIXED ANODE VOLTAGE FOR SFD-209, TUBE D37F

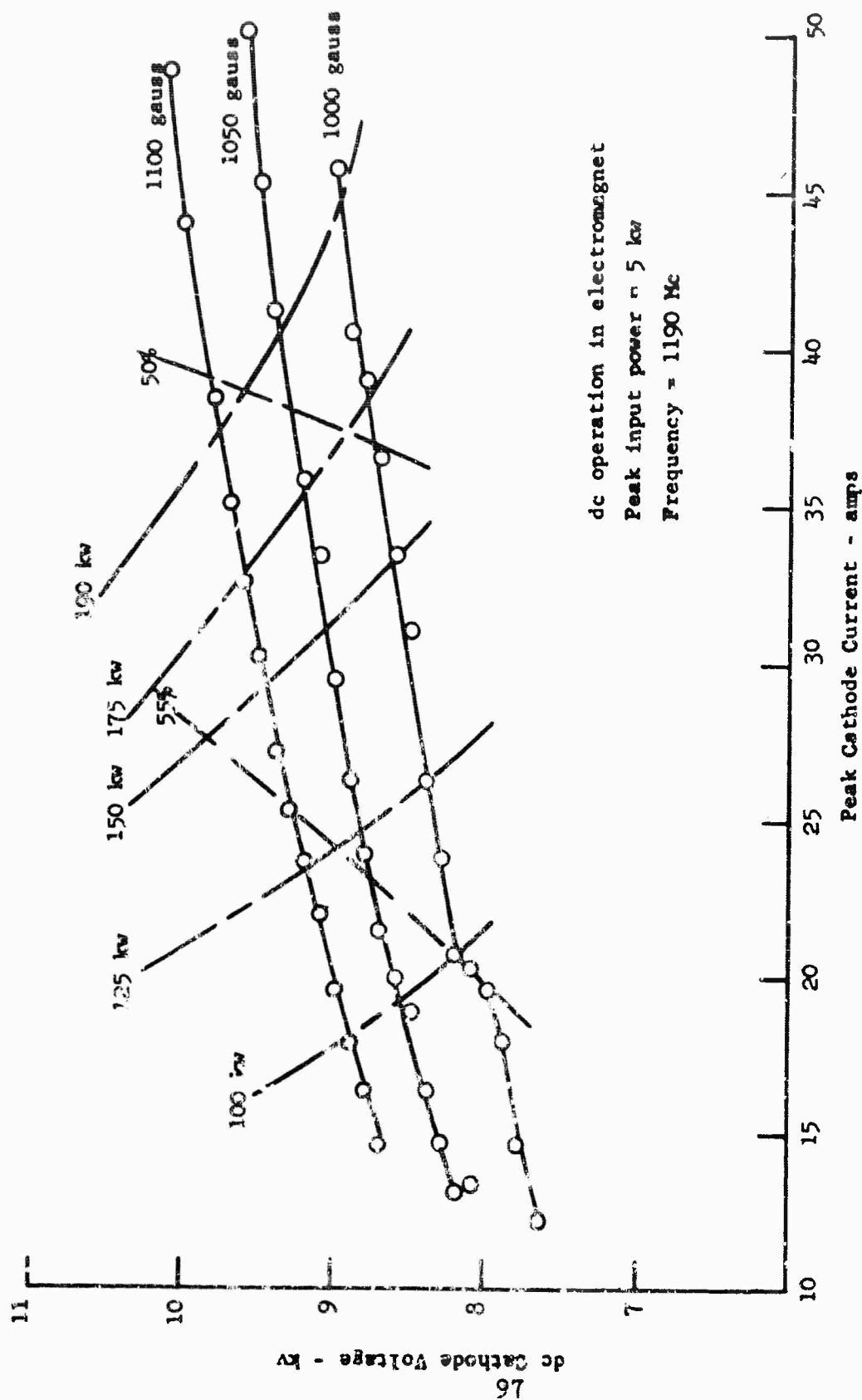


FIGURE 59 PERFORMANCE CHART FOR SFD-209, TUBE D37F

providing more magnetic confinement of electrons near the ends of the cathode. The magnetic field configurations for the original measurements of tube L30E shown in Figure 57 and for the changed pole piece configuration are shown in Figures 60 and 61, respectively. Figure 62 shows a performance chart taken under the same conditions as Figure 57, but with the new magnetic field configuration. It is seen that the results are essentially the same as those obtained in Figure 57. We therefore concluded that leakage of electrons from the interaction space is not significantly affected by a moderate barreling of the field and that a more severe change must be made to improve the efficiency.

A study was also made of possible reasons for variations in the V-I curves of the Group C tubes. Figure 63 shows an assembly of V-I curves on these tubes for a magnetic field strength of 1100 gauss. It is not surprising that tube L1E differs from the remainder of the group. It had been necessary to repair the shell of tube L1E and, as a consequence, the dimension from the C_{13} couplers to the back wall of the envelope was not within the required tolerances. A change in this dimension shifts the ω/β curve along the frequency axis thus changing the circuit synchronous voltage and consequently the operating voltage for a given magnetic field. A major part of the variation among the remainder of the group of tubes was traced to problems with calibration of the electromagnet. At least 300 volts spread along the voltage axis could be attributed to this. When errors had been corrected and a comparison made among the three tubes, the results shown in Figure 64 were obtained. The results appear to indicate that, when this source of variability is eliminated, the V-I curves are quite reproducible from tube to tube.

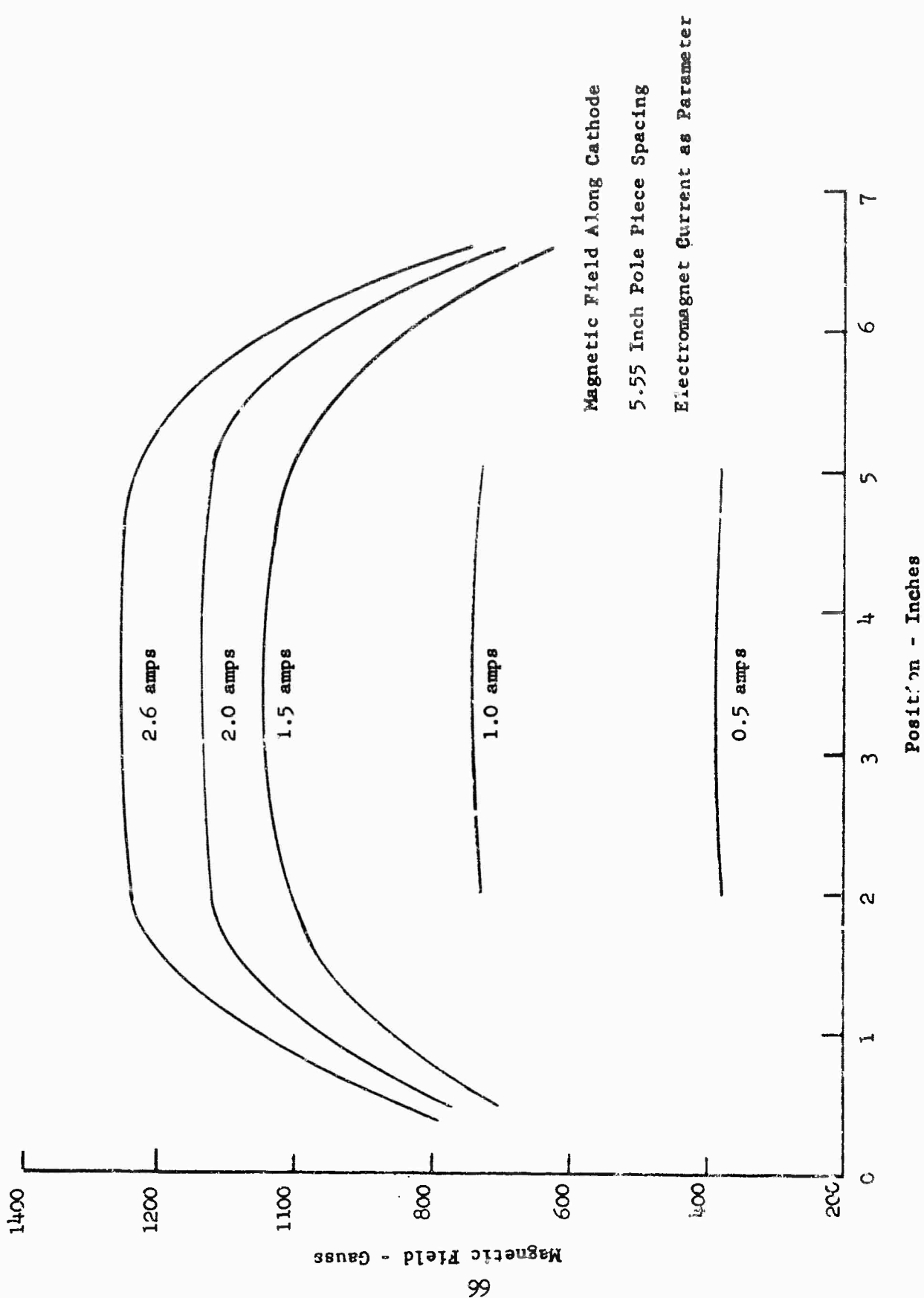


FIGURE 60 MEASUREMENTS ON HOT TEST ELECTROMAGNET

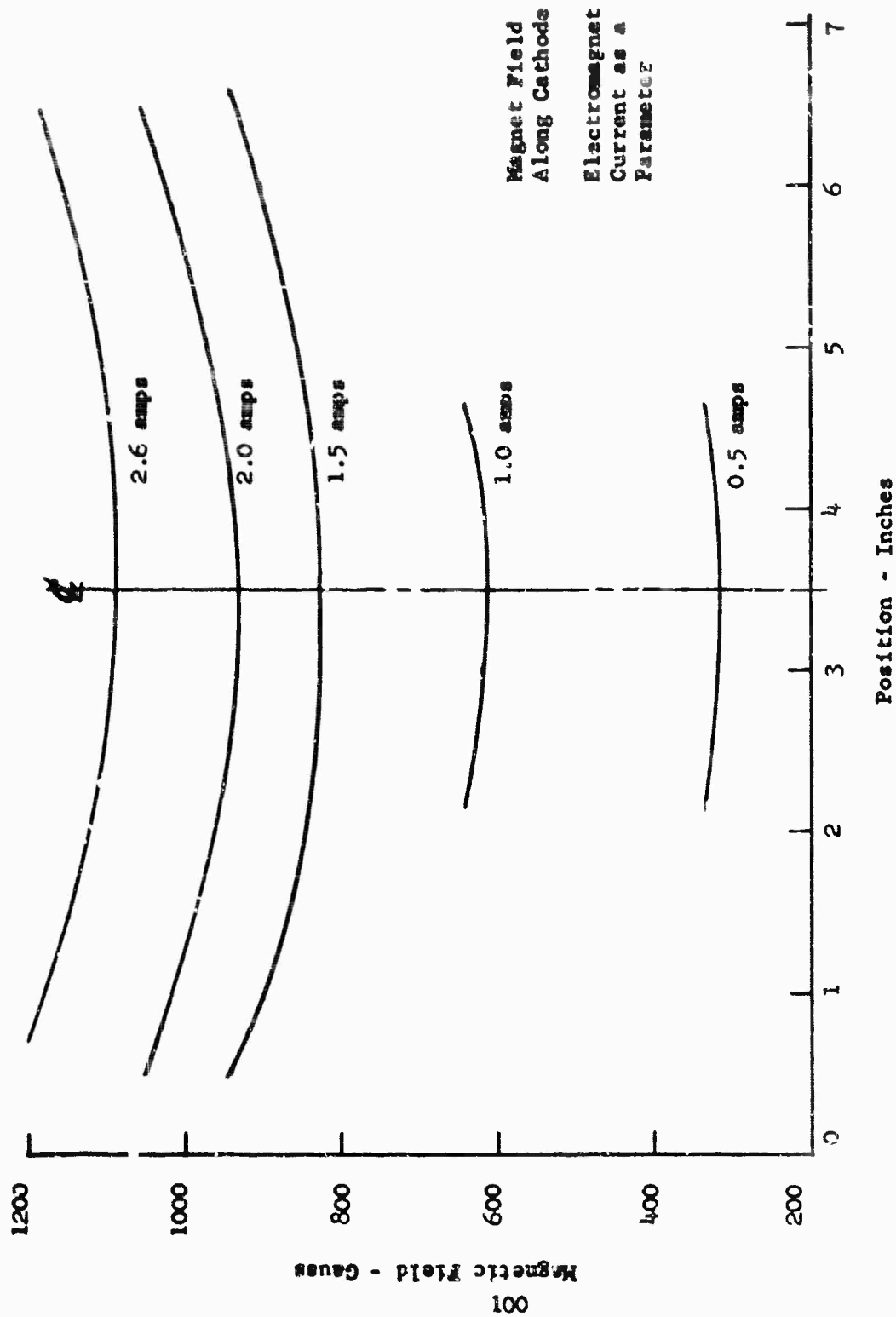
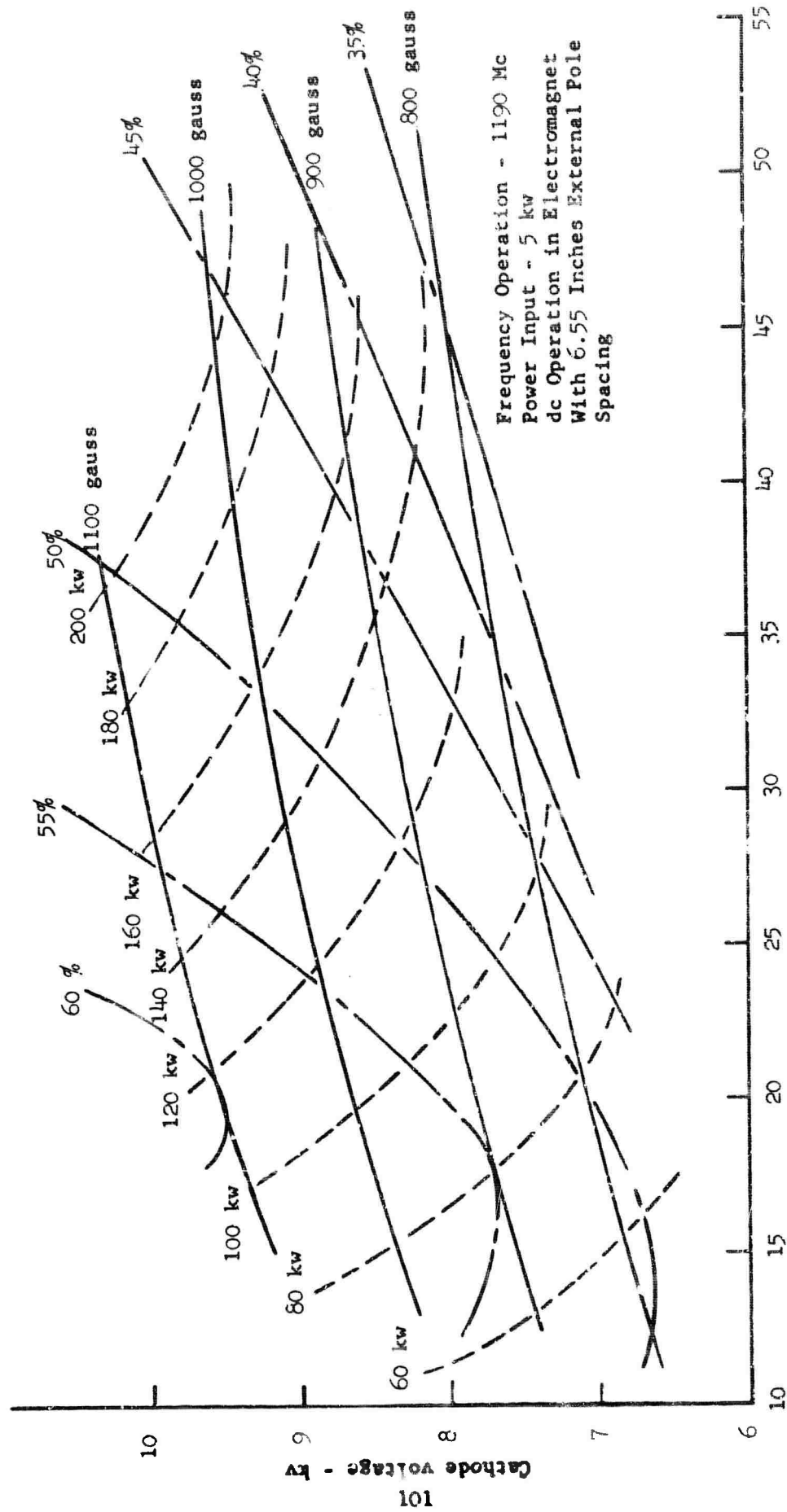


FIGURE 61 MEASUREMENTS ON HOT TEST ELECTROMAGNET



Peak Cathode Current - Amps

FIGURE 62 PERFORMANCE CHART FOR SFD-209, TUBE L302

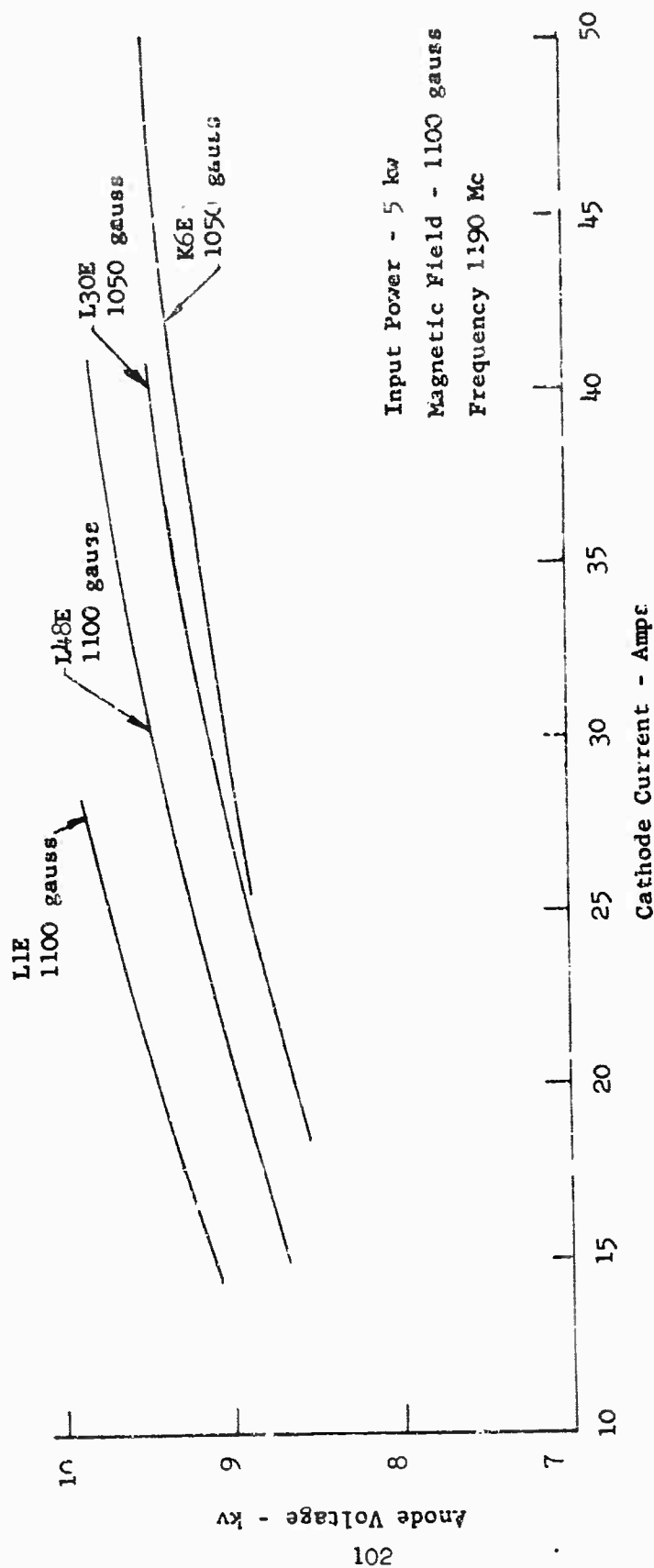


FIGURE 63 V-I CURVES FOR A GROUP OF SFD-209 TUBES

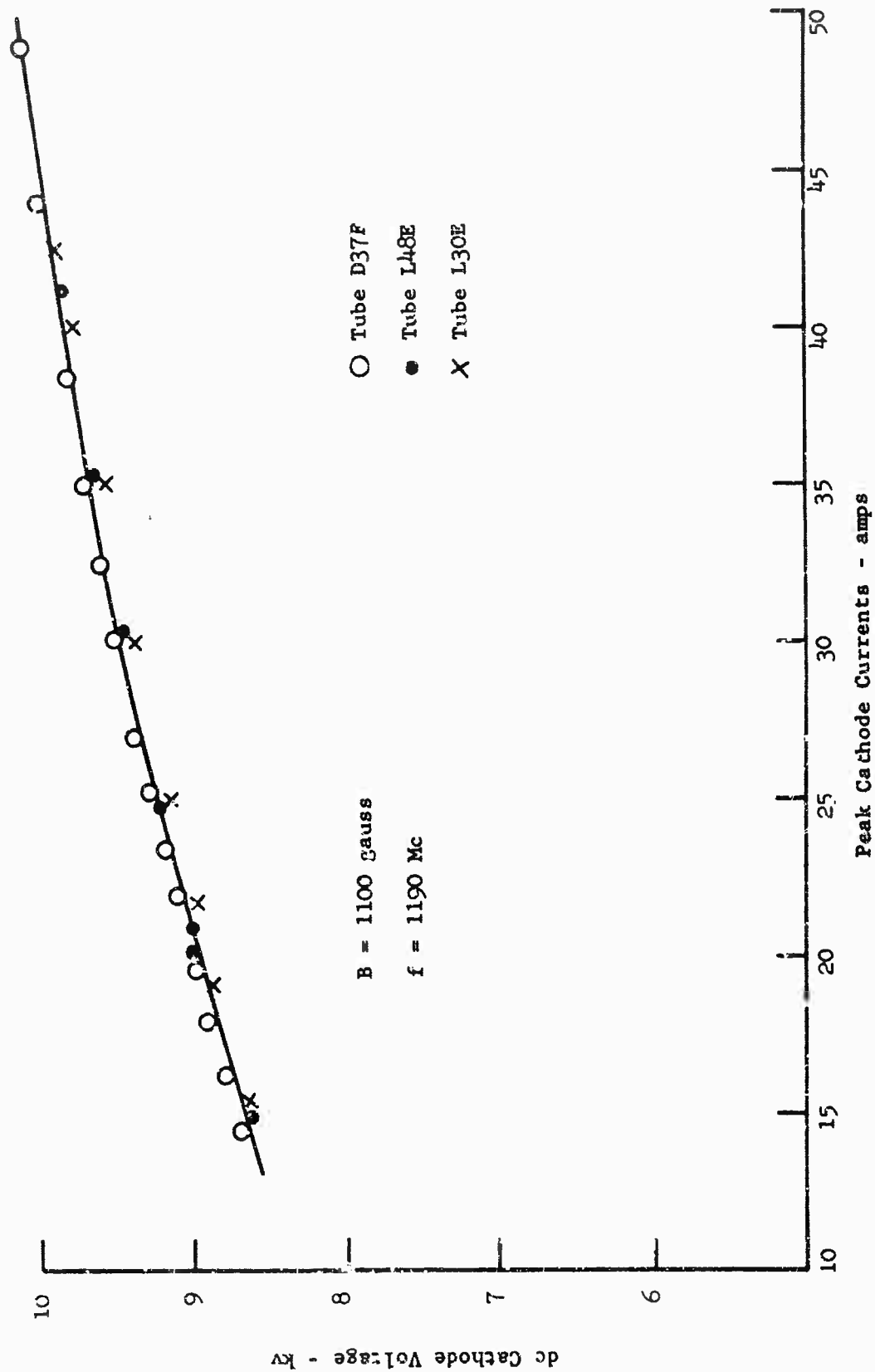


FIGURE 64 V-I CURVES OF TUBES AT START OF LIFE TEST

6.0 GROUP D TEST RESULTS

6.1 Construction

Performance tests and life tests on the Group C SFD-209 tubes indicated the desirability of several design changes. From the life tests, it was learned that the control electrode operated at excessive temperature. A first quick fix was made on the seventh (D34F) and eighth (D37F) Group C tubes by brazing copper heat shunts over the kovar support members for the control electrode insulator. However, it was evident that a redesign of the control electrode support was required. The supporting structure therefore was changed to the design shown in Figure 65. A beryllia insulator replaced the alumina insulator used in the previous design. The cross-section of the insulator was simultaneously increased. Finally, the method of mounting the insulator eliminated the thin kovar members of the previous design. The net result was considerably improved heat transfer from this new design control electrode support.

Another change shown to be necessary from the life tests involves the input window. Examination of the input window of the tubes removed from life test indicated considerable sparking between the center conductor and the outer conductor of the window on the atmospheric side. This resulted in a black deposit covering the input windows of tubes after several hundred hours of operation. The deposit does not seem to interfere with the cold transmission or reflection through the tube but is evidence of a marginal design and could very well cause difficulty as the average power level of the tube is increased. It was therefore decided to change the input window to 7/8 coaxial line size, making it the same as that used for the output window. This results in an input window somewhat larger than absolutely necessary.

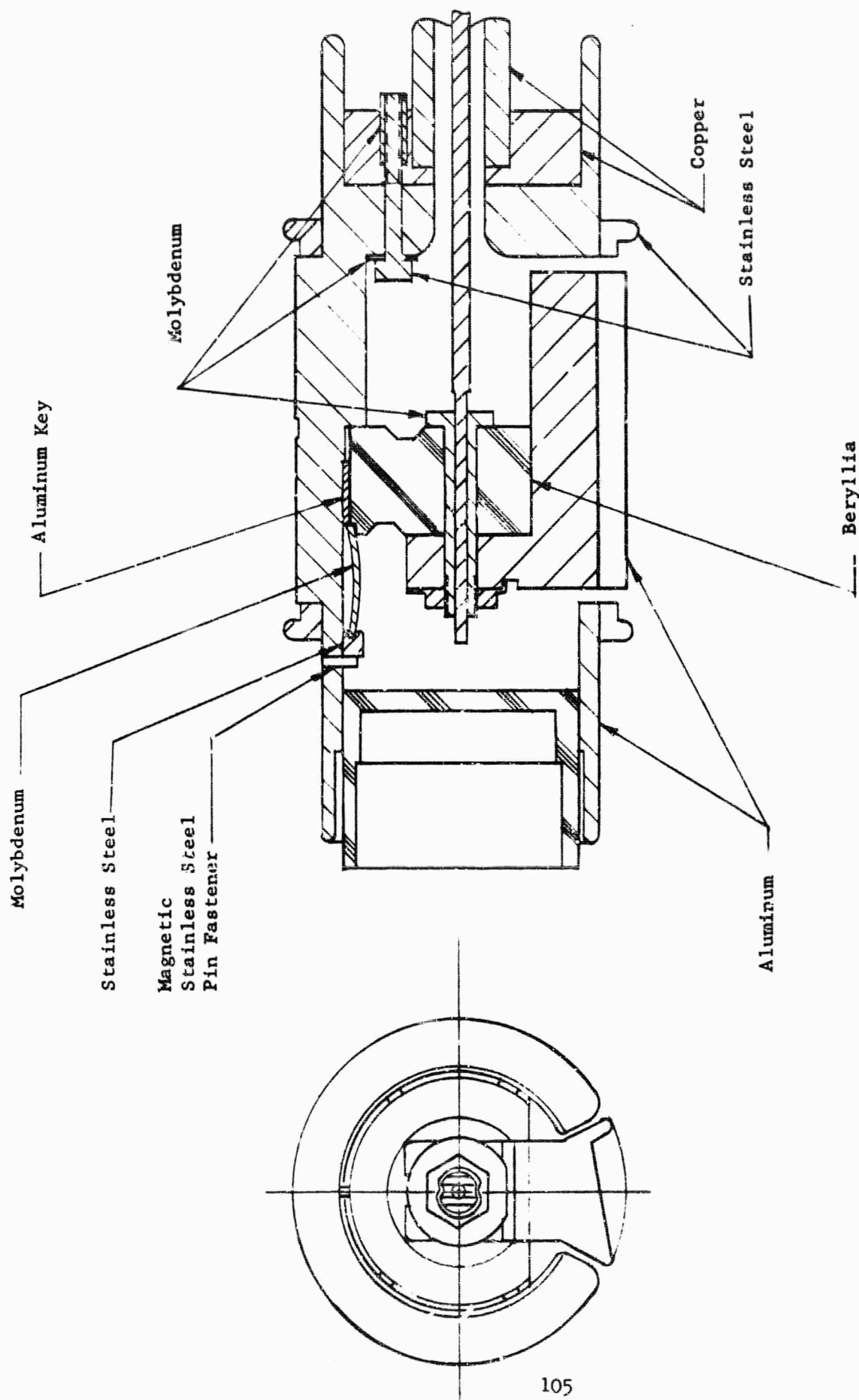


FIGURE 65 REVISED CONTROL ELECTRODE SUPPORT FOR SFD-209

6.2 Cold Tests

Detailed performance measurements have indicated that further suppression of π mode oscillations may be desirable to increase the dynamic range of the tube. Such oscillations do not exist in the normal operating ranges of this tube. However, we know that there is a region of π mode oscillation at lower drive power levels and at low operating voltages. If it were necessary to increase the gain of the tube by decreasing the input RF drive, further suppression of these oscillations would be required. If this suppression could be accomplished, the input drive power to the tube could be reduced with little change in output power. Even if the tube were to be operated with approximately 5 kw of input drive as we have done for the RF testing, it would be advantageous to decrease the required input power for satisfactory locking and thereby provide a greater degree of insensitivity to effects of load VSWR. (The effective reduction of input power by signals reflected from the antenna and re-reflected from the tube input is discussed in Appendix II.) Attenuation located between the C_{13} couplers and the vacuum envelope will introduce differential attenuation at the band edges. An example of such attenuation obtained on cold test was shown in Figure 9 of the Fourth Semi-annual Report.

As an experimental vehicle for differential attenuation studies, a linear version of the slow wave circuit for the SFD-209 was used. The linear circuit was assembled and matched. Figures 66 and 67 show the results of cold transmission and reflection measurements from the linear cold test circuit, with and without an Eccosorb absorber located between the C_{13} couplers and the back wall of the tube. The figures show that the absorber introduced differential attenuation at both band edges. Indications are that the attenuation at the π mode frequency may be quite large. It is believed that a smaller amount of attenuation will be satisfactory, with consequent reduction in the amount of attenuation introduced at band center.

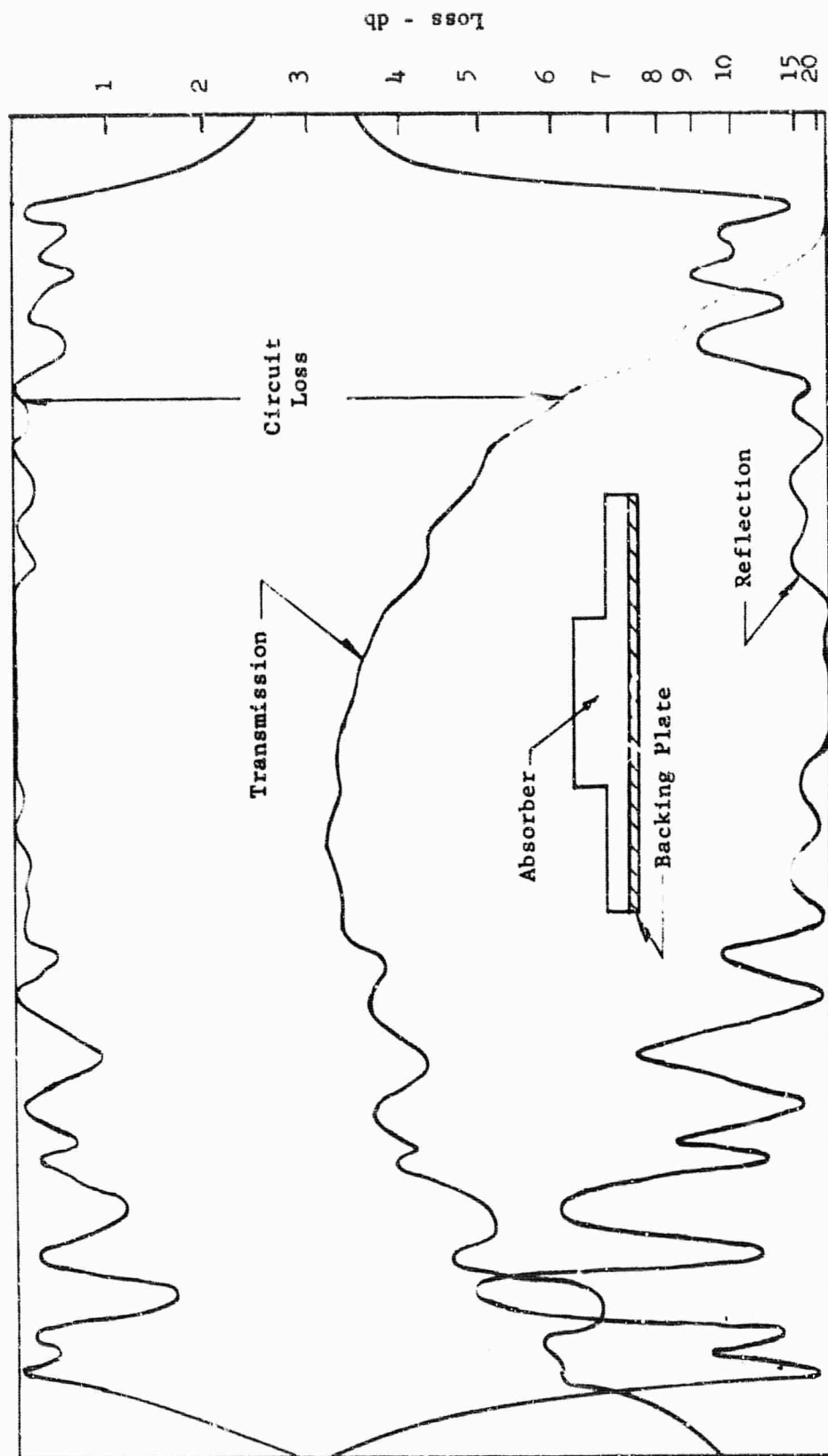


FIGURE 65 CIRCUIT LOSS INTRODUCED BY PLACING ABSORBER BETWEEN CIRCUIT AND ENVELOPE

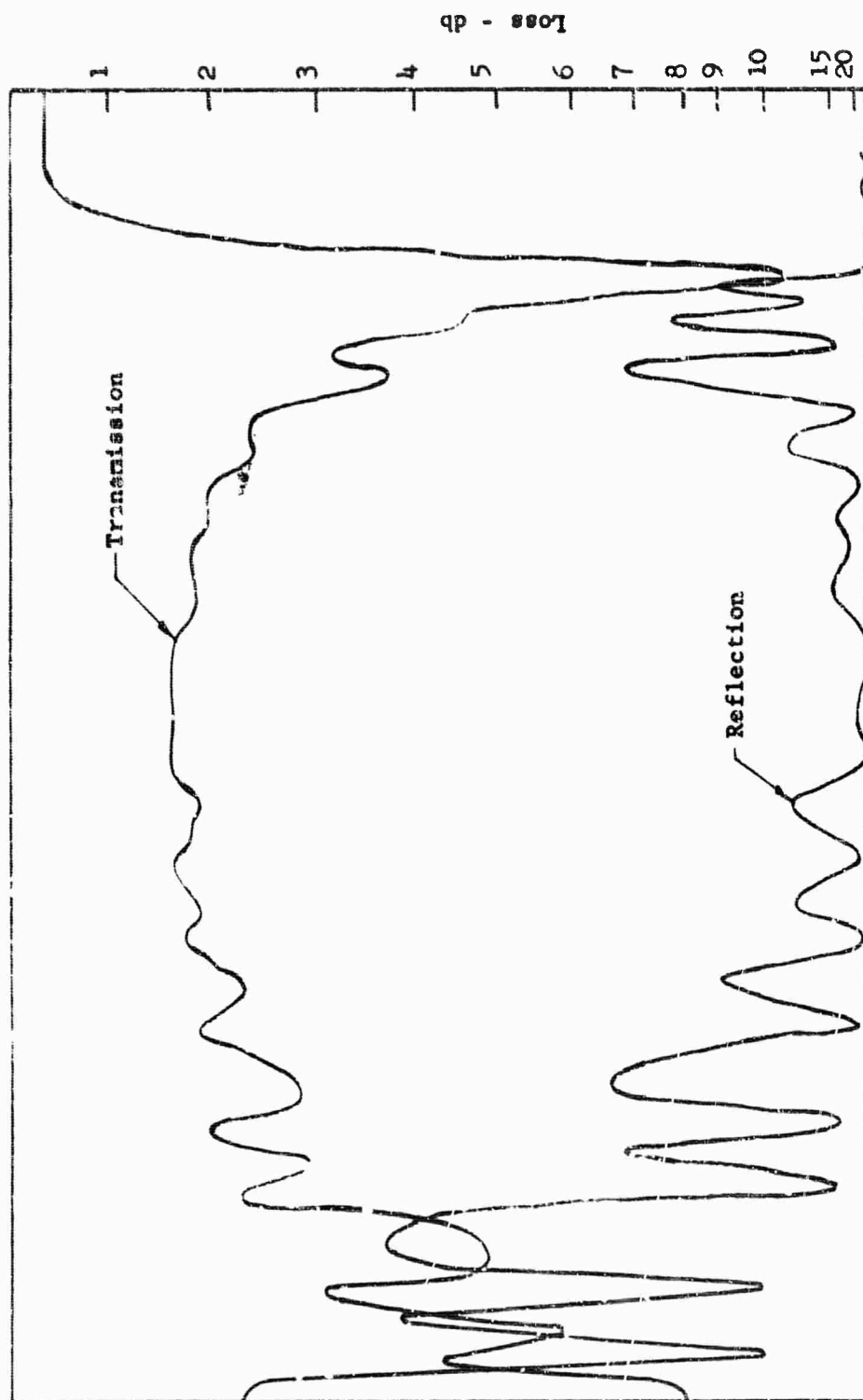


FIGURE 67 INSERTION LOSS OF CIRCUIT WITHOUT ABSORBER

A mounting procedure for locating a porous ceramic absorber in the actual tube to achieve results similar to those in Figure 66 was investigated. The absorbers were constructed from a relatively brittle, machinable ceramic because they could be obtained on a short delivery schedule. Unfortunately, they cracked during the tube bake-out procedure and fell against the capacitive couplers, resulting in high circuit loss and severe reflections.

Measurements on the linear cold test circuit produced some interesting results on the effect of back wall spacing. Figures 68 and 69 show dispersion curves as a function of spacing between the C_{13} coupler rings and a smooth back wall. In both cases, it can be seen that bringing a metal wall close to the C_{13} couplers depresses the entire dispersion curve and simultaneously depresses the π mode frequency. This indicates that more energy is stored in the region of the π mode frequency than at band center. Such selective π mode depression eliminates π mode synchronism in the Group C design tubes. Figure 70 shows the percentage depression of various points on the dispersion curve as the spacing of the C_{13} couplers to the back wall is varied. These data show how the back wall spacing may be used as a vernier adjustment on the location of the pass band.

6.3 Hot Test Measurements

Though tube F63F (the first Group D design tube) was mechanically different due to repeated brazing operations, measurements show that it may be classified as electrically similar to the Group C tubes. The V-1 characteristics and frequency response of tube F63F are shown in Figures 71 and 72. This tube was used to evaluate the high average power capability of this new design. Average power capability tests were run at a pulse length of 200 μ sec up to an average power of 1500 watts. This power limit was imposed by the average power capability of the coaxial components used in the test set.

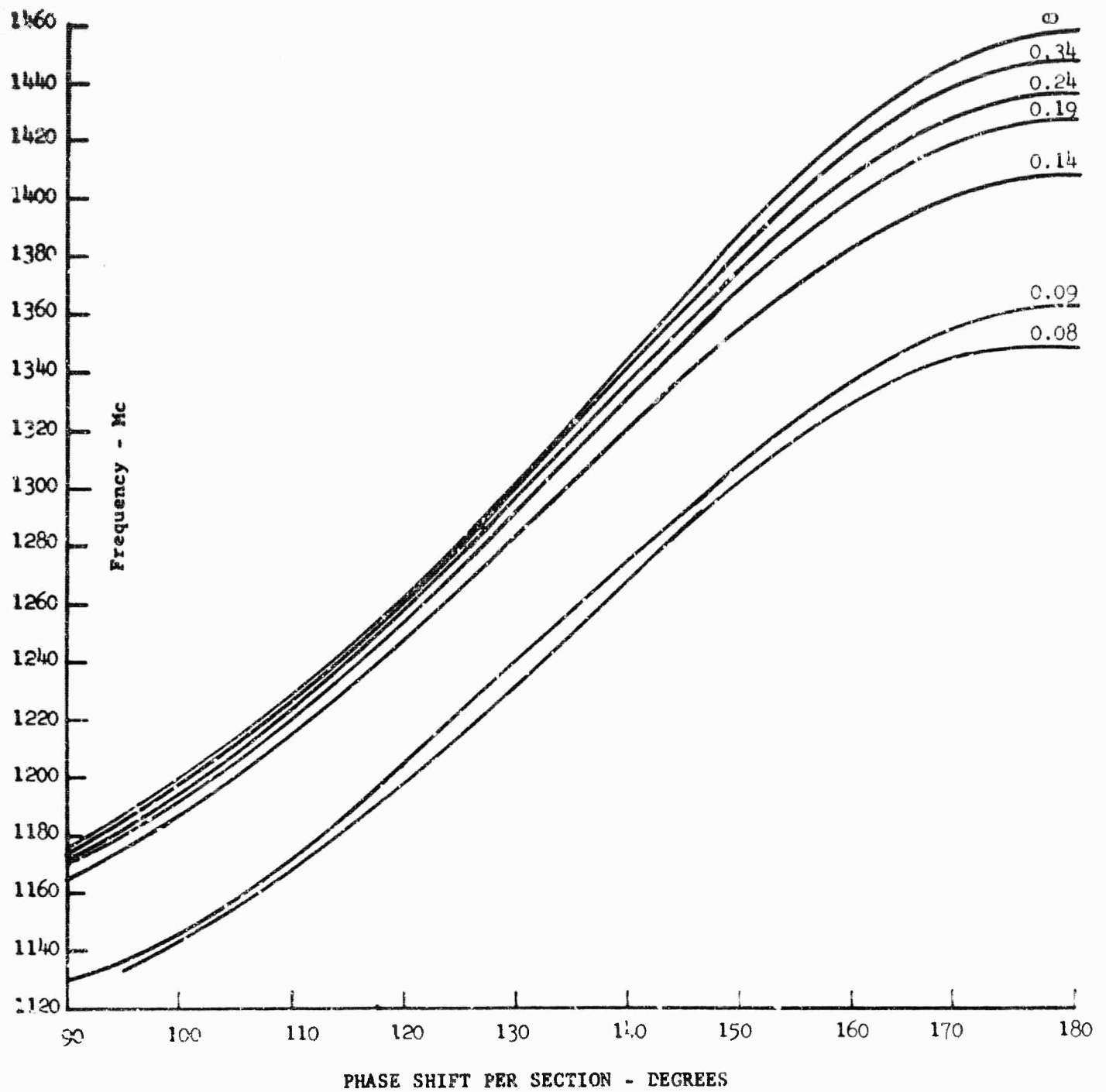


FIGURE 68 ω - β CURVES AS A FUNCTION OF SPACING BETWEEN E-PLANE WALL AND C_{13} COUPLER RINGS FOR L-BAND LINEAR CIRCUIT

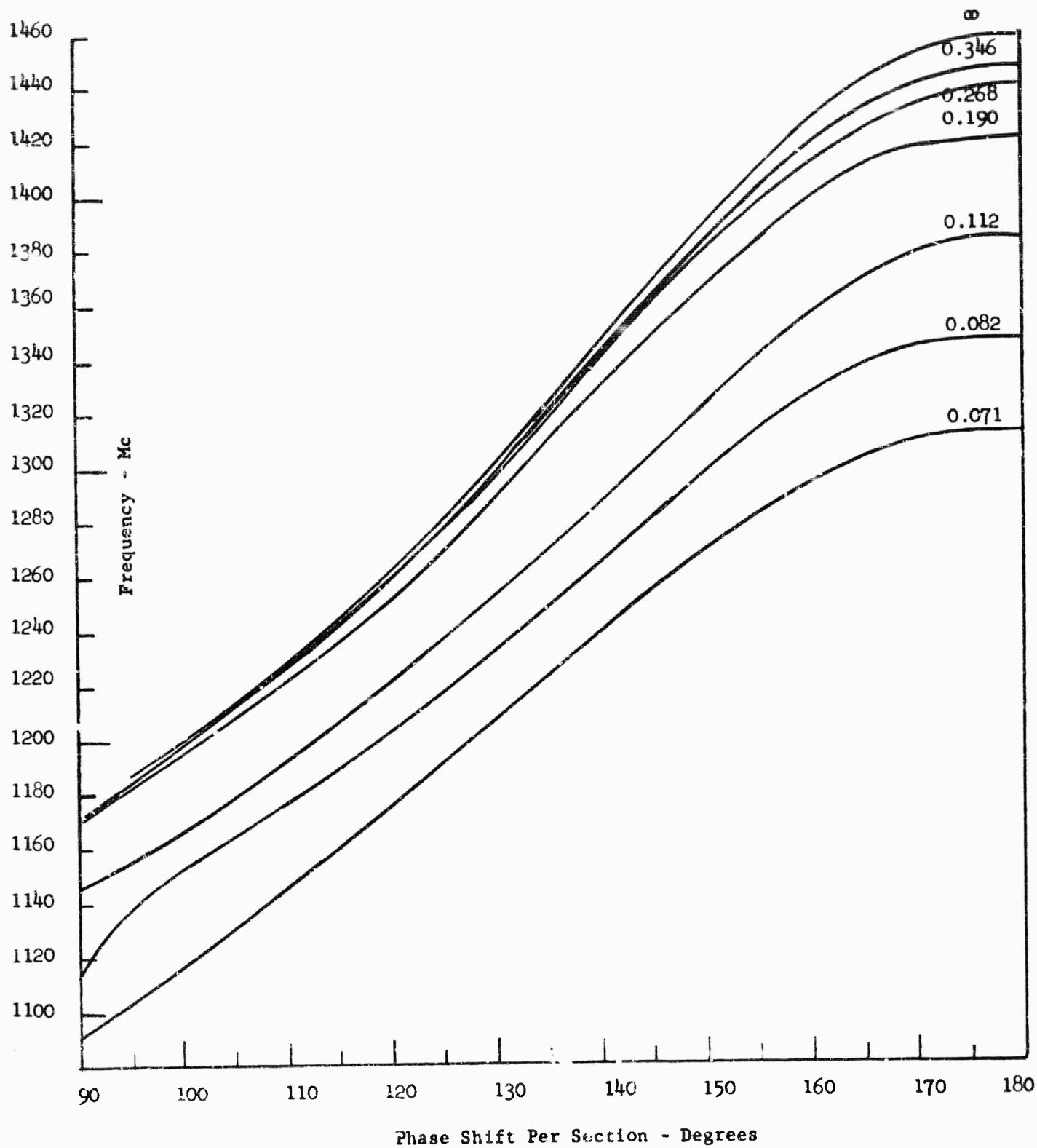


FIGURE 69 ω - β CURVES AS A FUNCTION OF SPACING BETWEEN 2 INCH RIDGE AND C_{13} COUPLER RING FOR L-BAND LINEAR CIRCUIT

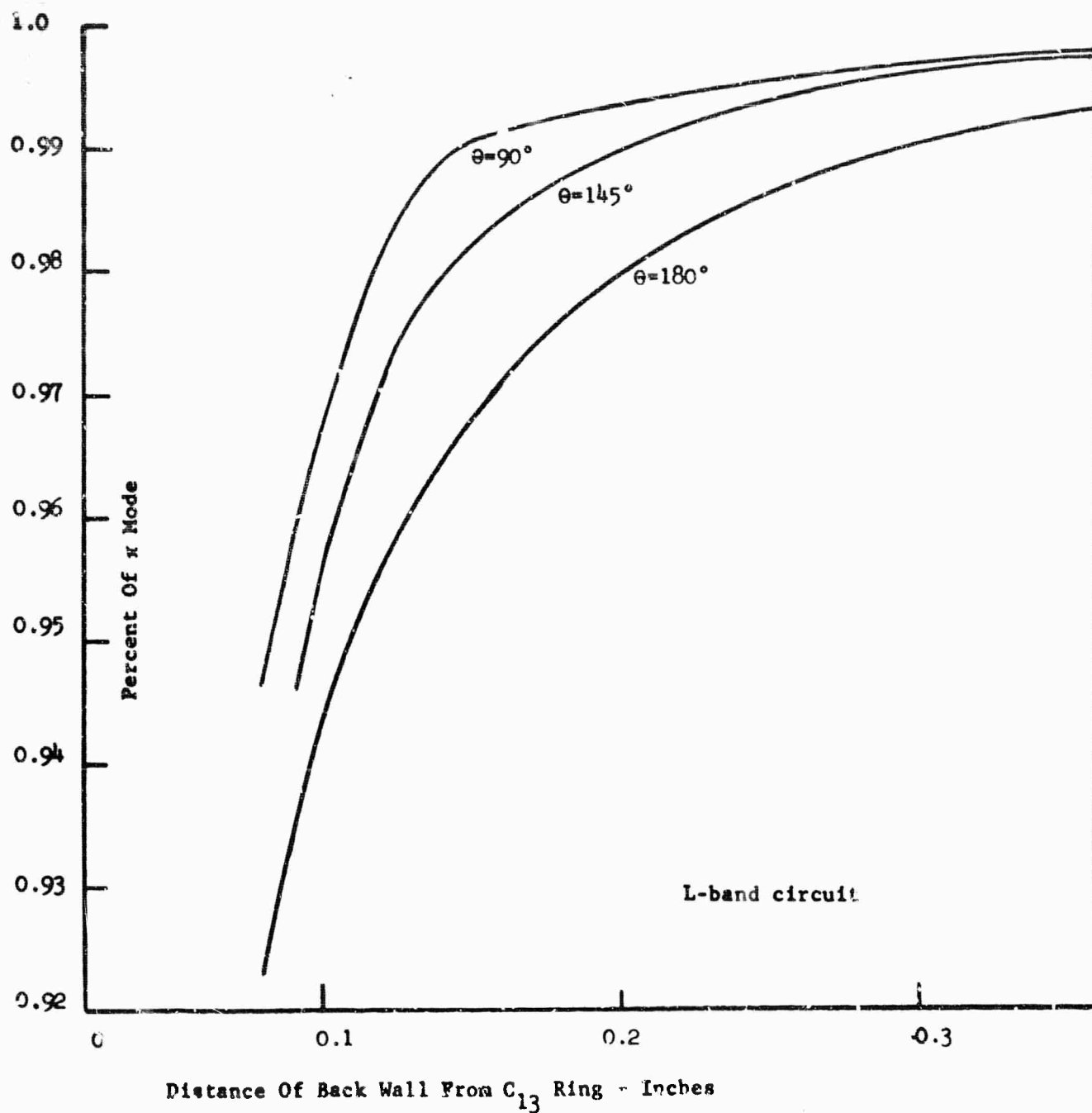


FIGURE 70 RELATIVE EFFECT OF BACK WALL SPACING ON FREQUENCY AT THREE DIFFERENT PHASE SHIFTS PER SECTION

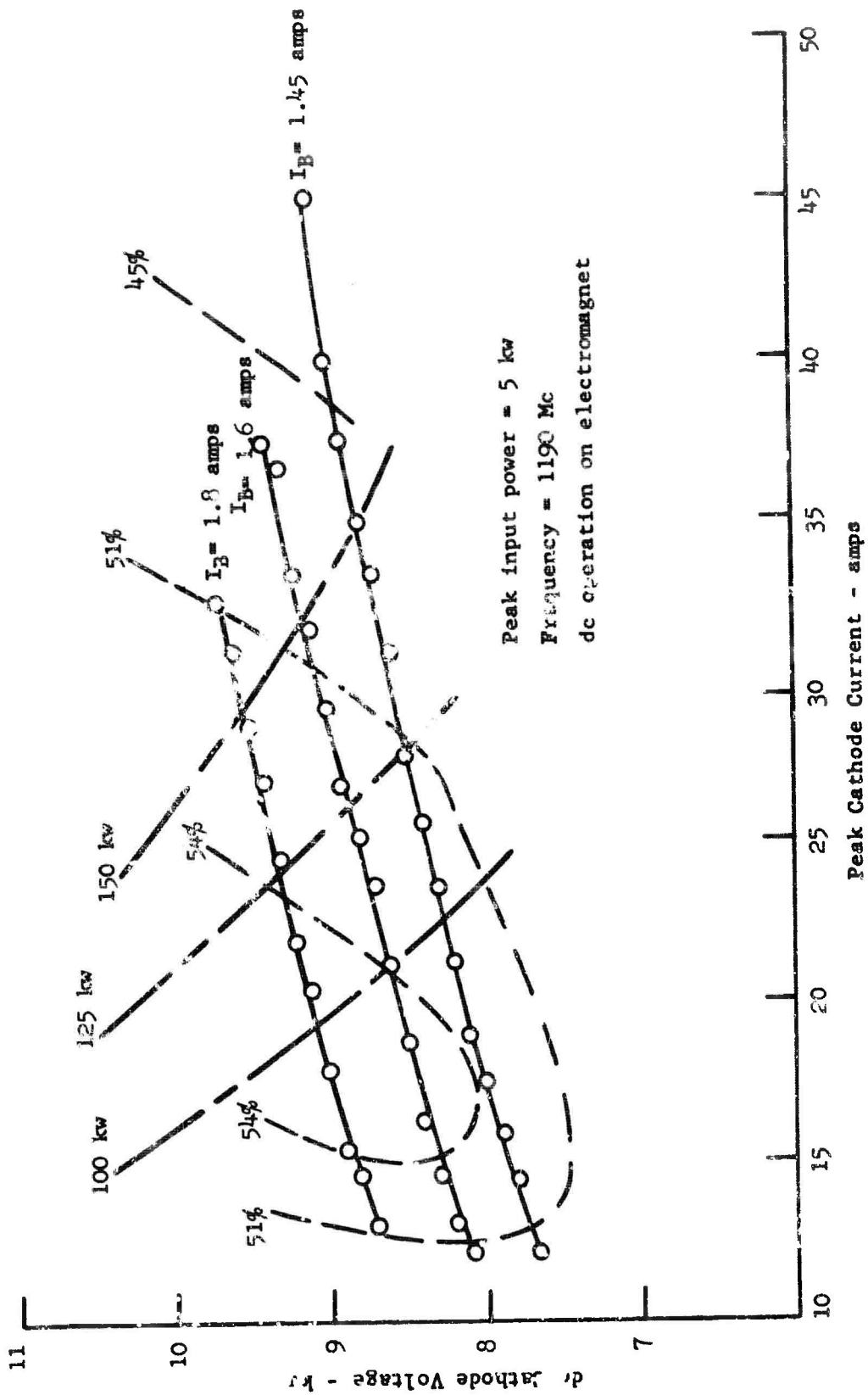


FIGURE 71 PERFORMANCE CHART FOR SFT-209, TUBE F63F

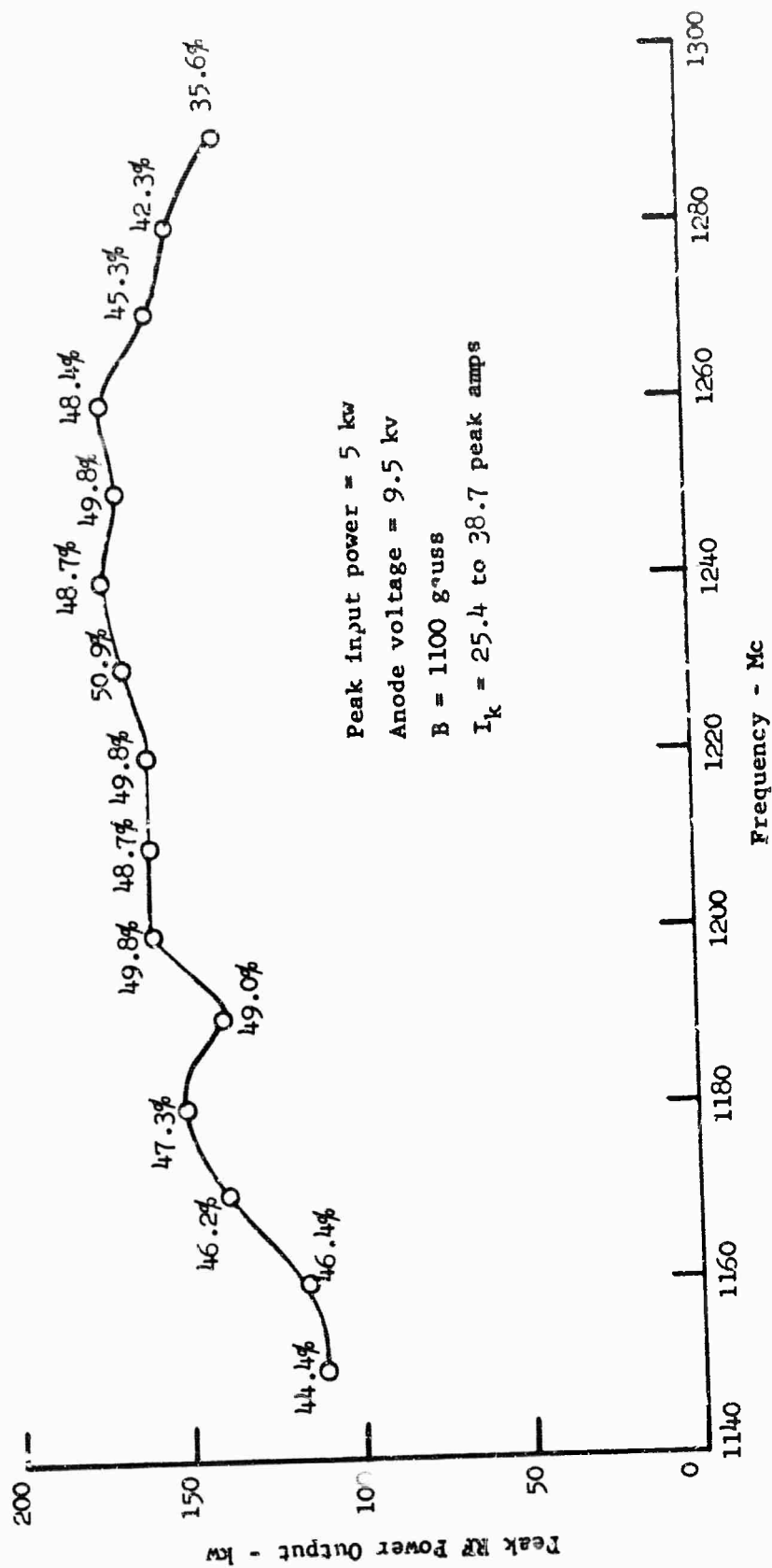


FIGURE 72 OUTPUT POWER VERSUS FREQUENCY AT FIXED ANODE VOLTAGE FOR SFD-209, TUBE #63F

In tube G35F, the second Group D tube, the anode-cathode spacing was increased 10% by decreasing the cathode diameter to observe the effect of the increased spacing on the dynamic operating range of the tube. During the initial test it was discovered that the control electrode diameter was not reduced proportionately to the cathode diameter, and therefore, the control electrode protruded radially from the cathode in the drift space. Upon completion of the test, the tube was opened, corrected and reprocessed as G35F-1. The subsequent measurements gave an evaluation of the effect produced by the protruding control electrode. Figure 73 shows the frequency response of the tube under both conditions. It is evident that the protruding control electrode causes a reduction in output power and efficiency. The frequency response of tube G35F-1 operating at a higher dc potential, is also shown in Figure 73. Figure 74 shows the V-I characteristics for this tube.

The V-I characteristics of tubes G35F-1 and F63F are compared in Figure 75. The increased anode-cathode separation in tube G35F-1 allows operation in the region of lower power output where, under conditions of less separation, the output would contain some π mode oscillations. It also is seen that, at a given percentage of the characteristic current, the power output and efficiency are approximately the same and, at a given current, the tube with the larger anode-cathode separation operates at a higher voltage. Therefore, the net effect of increased anode-cathode separation, in a given magnetic field, is to move the V-I curve upward toward higher voltage and left toward lower current. This shift in operating conditions increases the dynamic operating range by extending the low power region without sacrificing too much of the higher power region.

An important scaling variable for crossed-field devices is known as the characteristic current. This current is given by

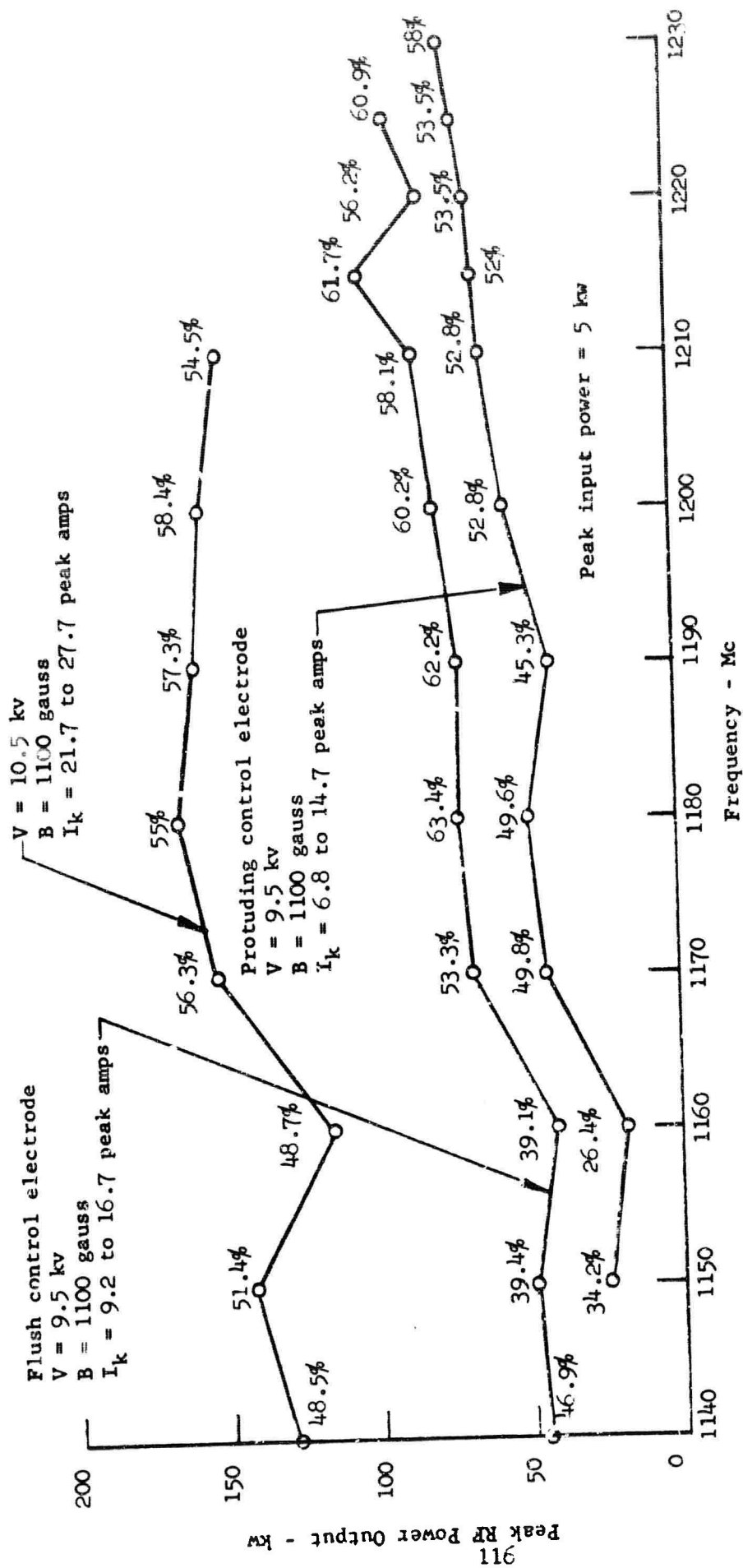


FIGURE 73 OUTPUT POWER VERSUS FREQUENCY AT CONSTANT ANODE VOLTAGE FOR SFD-203, TUBE G35F-1

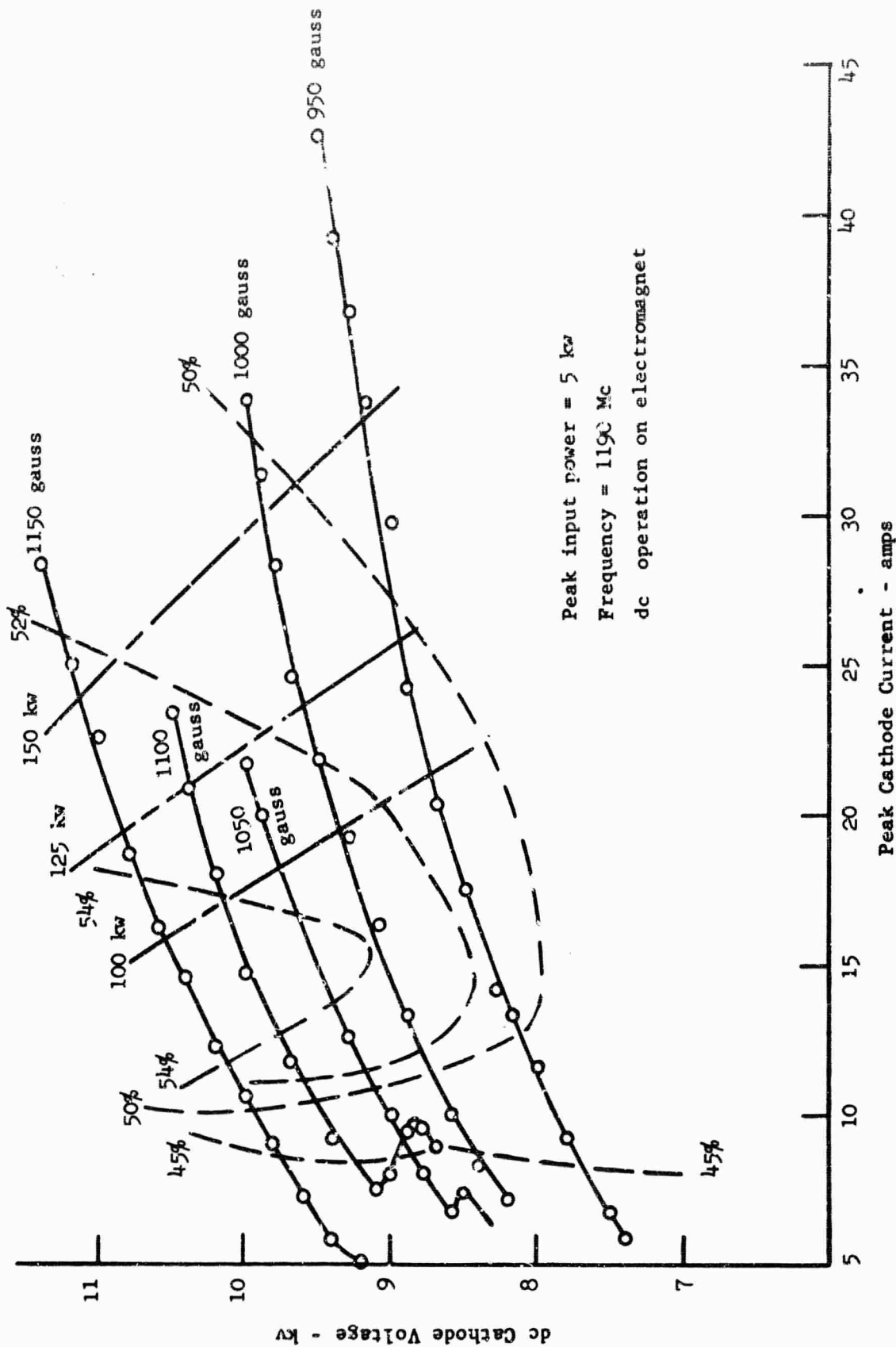


FIGURE 74 PERFORMANCE CHART FOR SFD-209, TUBE G35F-1

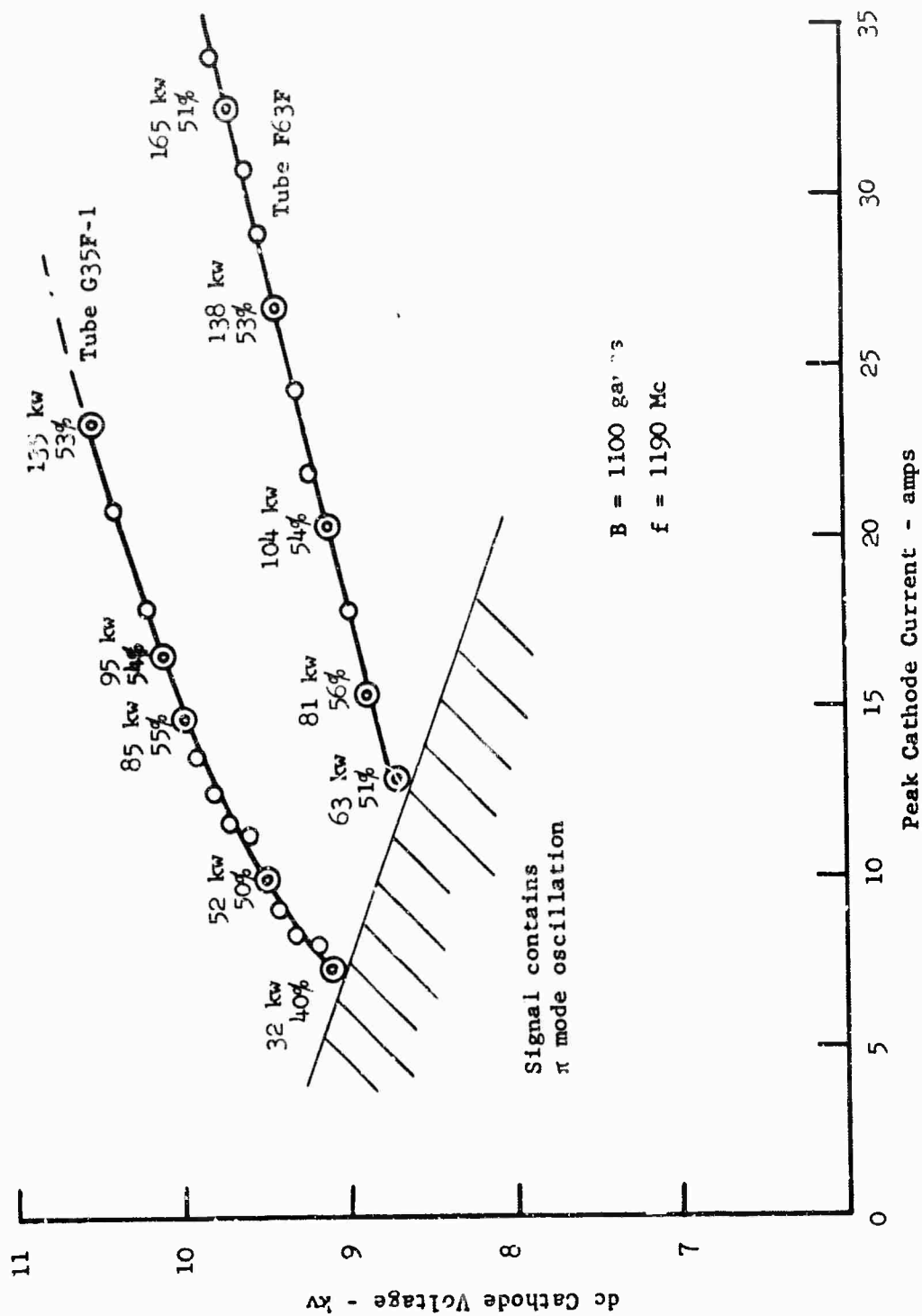


FIGURE 75 COMPARISON OF V-I CURVES OF TUBES F63F and G35F-1
AT 1100 GAUSS. ANODE-CATHODE SPACING IS INCREASED
BY 10% IN TUBE G35F-1

$$I_c = 8.96 \times 10^{-6} V_c^{3/2} \frac{h}{d} \frac{r_a}{d} R_1 \frac{\theta}{2\pi}$$

where $R_1 = 8r_a^2 r_c / (r_a + r_c)^3$
 h is the cathode axial height
 V_o is the synchronous voltage
 r_a is the anode radius
 r_c is the cathode radius
 $\theta/2\pi$ is the fraction of a full circle encompassed by the active circuit
 d is the anode-to-cathode spacing

It is seen that the characteristic current is inversely proportional to the square of the anode-cathode spacing. Characteristic currents calculated at band center for tubes F63F and G35F-1 are 24.7 amperes and 19.7 amperes, respectively. The peak output power from a crossed-field device is given by

$$P_o = \eta \frac{V}{V_o} V_o I_c$$

where P_o is the power output
 η is the efficiency factor (% efficiency/100)
 V is the operating voltage
 and the other symbols are as defined above.

The output power is thus a function of the characteristic current. For an output power of 150 kw peak, a tube having the normal anode-cathode spacing of tube F63F operates at about 1.5 times characteristic current. This is comparatively high for crossed-field devices with operation at characteristic current being more typical. The performance charts for the tubes, however, show that efficiency tends to maximize near the characteristic current. This is

also quite typical for magnetrons. To see if the scaling law for a characteristic can be properly applied to the SFD-209 type of tube, we have plotted the performance charts for tubes F63F and G35G-1 in terms of normalized parameters. (A number of charts of this sort are presented for magnetrons in Collins*.) On this chart, the abscissa is normalized by the characteristic current. The ordinate is normalized by the synchronous voltage which is given by

$$V_o = \frac{v_o^2}{2 \frac{e}{m}}$$

where v_o is the wave phase velocity
 e/m is the charge to mass ratio for an electron

and the magnetic field is normalized by

$$B_o = \frac{v_o}{\frac{e}{m} d} R_2$$

where $R_2 = 2r_a / (2r_a - d)$

and the other symbols are as defined above.

As the anode-cathode spacing is increased, the magnetic field must be decreased for the same V/V_o ratio. This results from the lowered dc electric field at the increased spacing. To maintain the same E/B electron drift velocity, the magnetic field must be correspondingly reduced. Figure 76 shows that the curves for F63F and G35F-1 practically coincide when plotted in normalized coordinates. Further, they show that increasing the anode-cathode spacing does not cause any significant degradation in efficiency. Thus the scaling law for anode-cathode spacing applies to this case. As was reported

*George B. Collins, MICROWAVE MAGNETRONS, McGraw-Hill Book Company, Inc., Vol. 6, MIT Radiation Laboratory Series (1948), Chapter 10

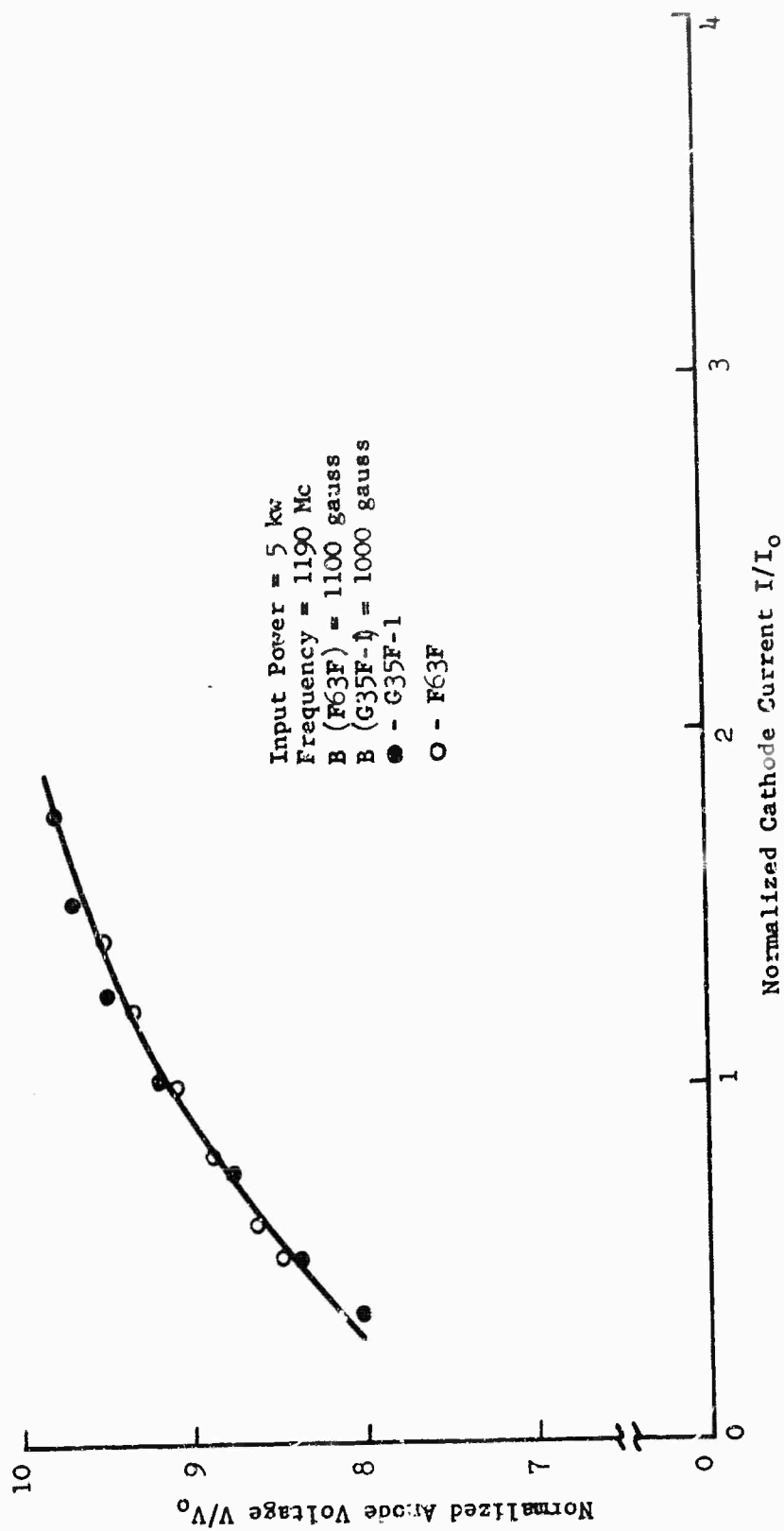


FIGURE 76 NORMALIZED PERFORMANCE FOR STD-209, TUBES F63F and G35F-1

in Report No. 7 (October 1964, RADC-TR-65-82, AD 613 995), increased spacing results in a significant decrease in the range of currents over which band edge oscillations occur. Figure 77 shows the location of the mode boundary for the spacings of tubes F63F and G35F-1. Stable operation occurs to the right of these mode boundaries and band edge oscillations occur to the left of them. The increased anode-cathode spacing has resulted in the mode boundary of tube G35F-1 being moved to approximately one-third of the current of tube F63F. Thus, it is seen that at some sacrifice in output power (approximately 1 db), the tendency toward π mode oscillation at low currents may be decreased. In some applications, this may be necessary where operation over a wide range of powers is required.

A two-signal experiment was made on another program using this amplifier. In this measurement, two signals of equal amplitude were fed into the crossed-field amplifier. When this is done, the instantaneous amplitude of the input to the crossed-field amplifier is given by

$$\sin \omega_1 t + \sin \omega_2 t = \frac{1}{2} \sin \left(\frac{\omega_1 - \omega_2}{2} \right) t \sin \left(\frac{\omega_1 + \omega_2}{2} \right) t$$

This signal is sketched in Figure 78. It consists of the product of a sinusoidal envelope function and a sine wave with frequency equal to the average of the two input frequencies. Between nulls of the envelope, the frequency appears to be constant; however, the phase undergoes a sudden 180° reversal at the time of the envelope null. Thus, the instantaneous input signal is a rather complex function containing amplitude nulls and simultaneously a sudden shift in phase. We expected that the manner in which the crossed-field amplifier would respond to this input signal might give a rough indication of its response to the switched frequency signals required in the present application.

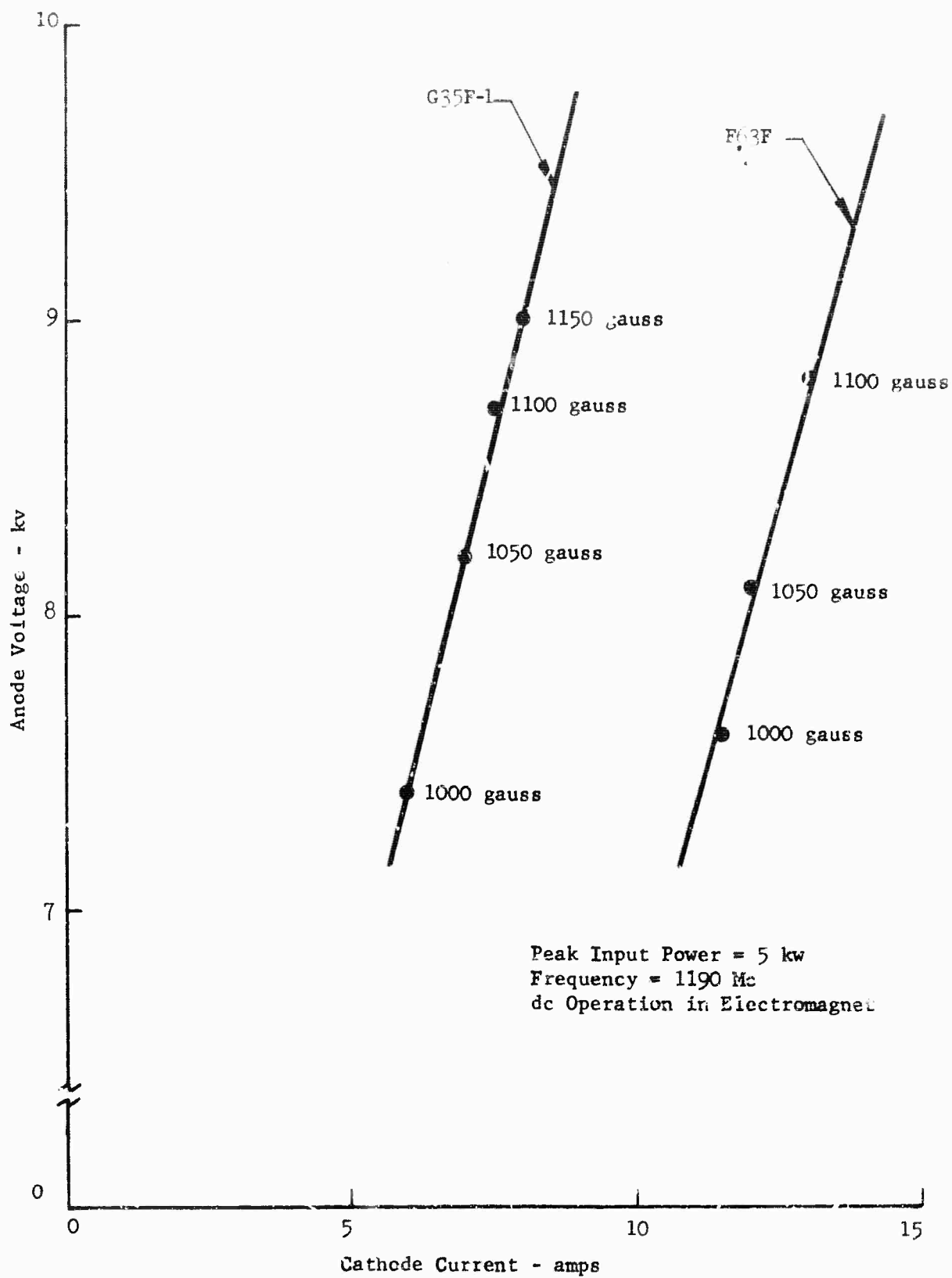
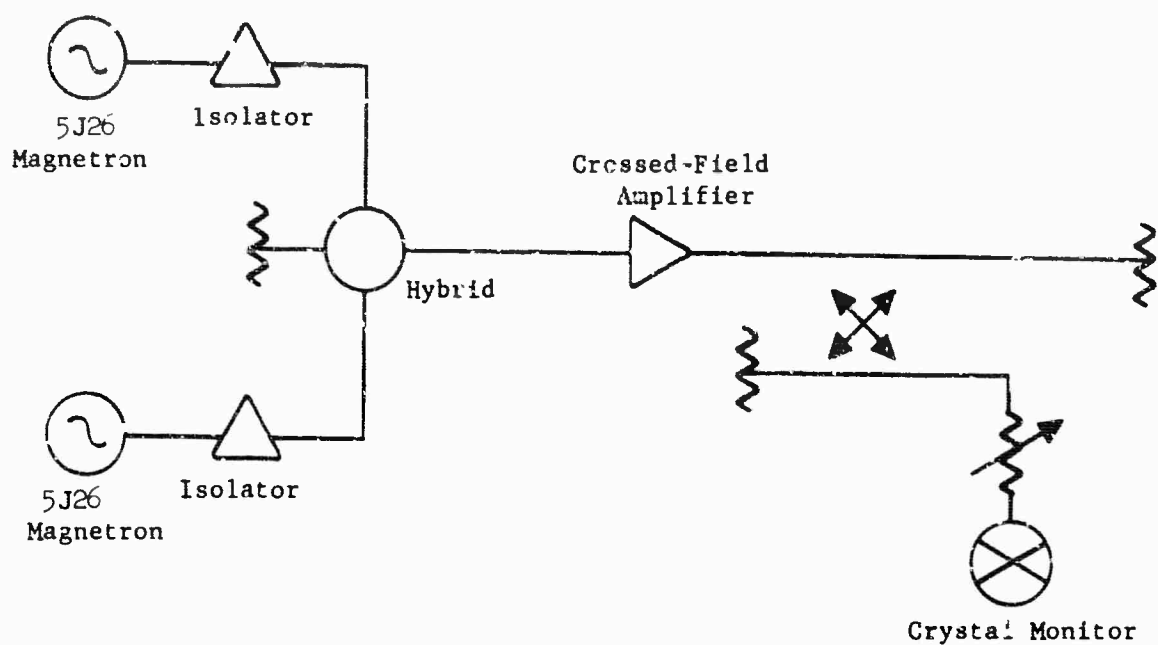


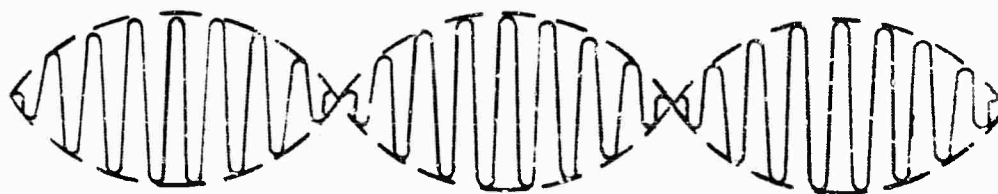
FIGURE 77 MODE BOUNDARY LOCATION FOR SFD-209, TUBES G35F-1 AND F63F

In the experiments conducted, the two input frequencies were derived from two magnetrons which were isolated from one another by about 30 db. These signals were combined and fed into the crossed-field amplifier. A sample of the output signal from the crossed-field amplifier was detected by a conventional crystal detector and the resulting output displayed on an oscilloscope. A block diagram of the set up is shown in Figure 78. Measurements were first made without voltages applied to the crossed-field amplifier. We therefore measured the input signal to the crossed-field amplifier as attenuated by its insertion loss. A photograph of the detected signal is shown in Figure 79a. It illustrates the expected beat pattern which is the rectified envelope of the waveform shown in Figure 78. This envelope appears in a fixed relation to the leading edge of the pulse because of phase locking of one magnetron by the other. One of the magnetron RF pulses is slightly longer than the other. The leakage signal from this magnetron, which starts first, into the second magnetron fixes the initial phase of this second magnetron; i.e., the leakage signal starts the second magnetron initially in phase with the first magnetron. If it were not for this effect, the detected envelope would appear in random relation to the leading edge of the pulse from pulse-to-pulse and the scope display would not show a fixed envelope pattern. The signal injected into the second magnetron is not strong enough to frequency lock this magnetron, so gradually the phases of the two magnetron outputs drift apart as each oscillates at its own frequency. A beat pattern with fixed envelope relationship to the pulse leading edge is thus established.

The operating voltages were then applied to the crossed-field amplifier. The resulting detected output waveform is shown in Figure 79b. (Additional attenuation was inserted in front of the crystal before taking this photograph to reduce its input to approximately the same level as in the first photograph, Figure 79a.) It is

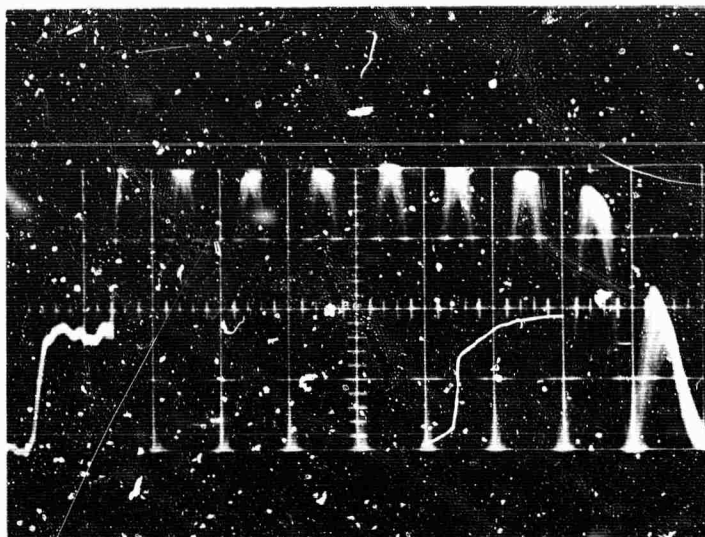


Block diagram of test set used in L-Band experiment with two equal amplitude input signals

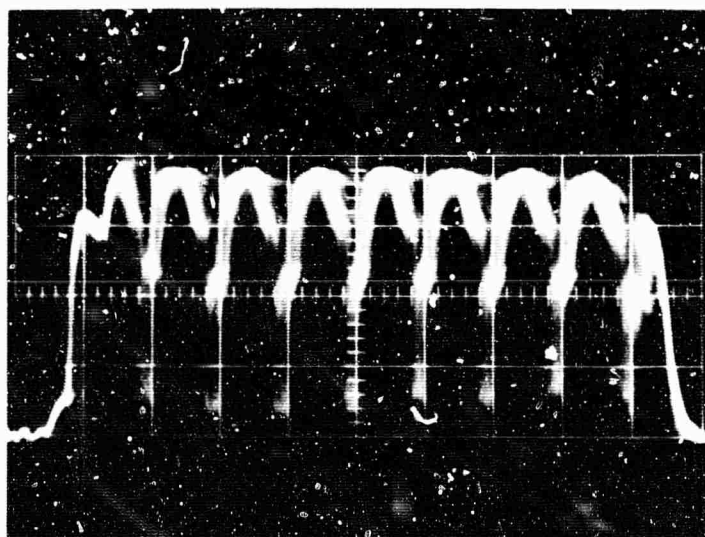


Instantaneous waveform at input of crossed-field amplifier

FIGURE 78 BLOCK DIAGRAM OF TEST SET USED TO OBTAIN INSTANTANEOUS WAVEFORMS SHOWN AT INPUT OF CROSSED-FIELD AMPLIFIER



a Photograph showing detected envelope of combined output of two magnetrons before amplification by crossed-field amplifier. Pulse length is 2 μ sec



b Photograph showing detected envelope of two signals after amplification by crossed-field amplifier

FIGURE 79 DETECTED SIGNAL BEFORE AND AFTER AMPLIFICATION BY CROSSED-FIELD AMPLIFIER

seen that the output signal maintains the null pattern of the input signal, although the compression in the crossed-field amplifier has broadened the tops of the cusps and sharpened the nulls. There was no indication of a tendency toward oscillation or noise generation during the null intervals. Although the exact significance of this experiment requires further evaluation, we believe that the ability of the crossed-field amplifier to reproduce the null pattern is at least a rough indication of its ability to reproduce the switched frequency input signal.

The experiments described above were made with the input signals a few megacycles apart at L-band. In similar experiments, the two equal amplitude signals were separated anywhere from 2 Mc to 100 Mc. In all cases, the crossed-field amplifier amplified both signals and split its output roughly equally between them, so that about half of the saturated output occurred at each of the input frequencies. The total output with two equal amplitude signals was slightly above that with only one of the signals present, indicating that the amplifier was not completely saturated. These measurements indicate that the crossed-field amplifier is indeed a true saturated amplifier and not simply a locked oscillator.

In an attempt to understand the starting delay and jitter observed in the Group C tubes on life test, the starting delay in tube G35F-1 (Group D) was studied. The delay time was measured as a function of anode voltage and input power at several magnetic fields. These measurements showed a starting delay of less than 20 nanoseconds for a 5 kw peak input over a 1000 volt range of cathode voltages encompassing the normal operating values. (The delays quoted here are in excess of the pulse envelope transit time through the tube and thus represent a shortening of the pulse as sketched in Figure 80.) For lower peak input powers, delays were greater and the range of operating voltage for minimum delay was reduced. At 1 kw input drive

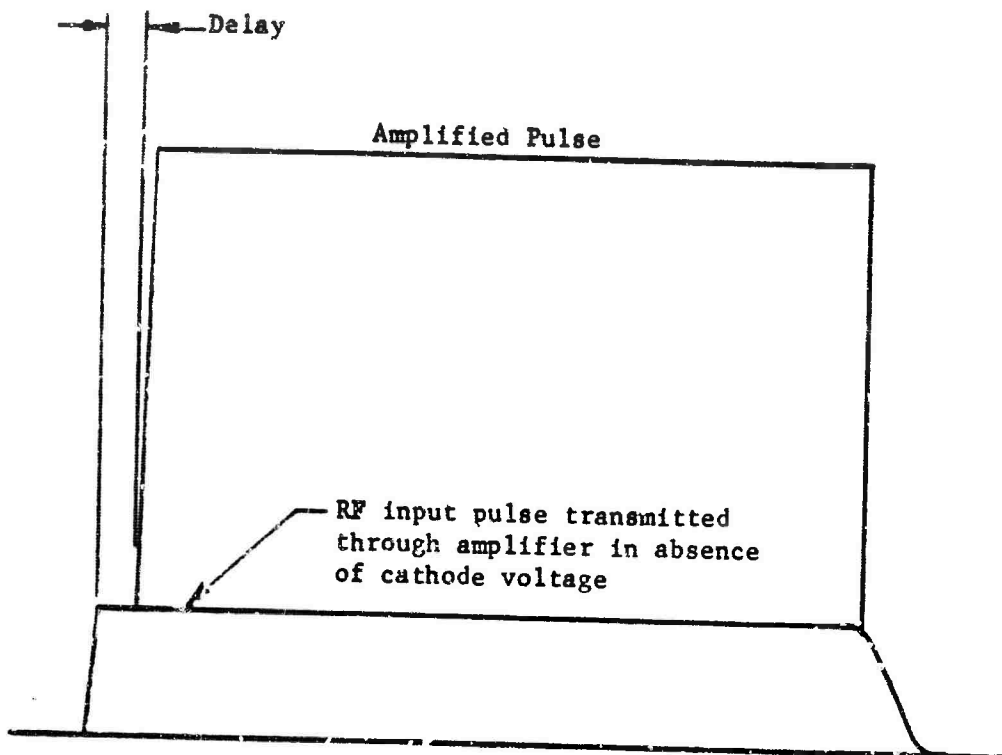


FIGURE 80 PULSE SHAPES OF RF INPUT PULSE AS TRANSMITTED THROUGH TUBE IN ABSENCE OF CATHODE VOLTAGE AND AMPLIFIED RF PULSE WHEN CATHODE VOLTAGE IS APPLIED

Starting delay is measured at approximately 85% points. The amount of delay is exaggerated in this sketch and is ordinarily less than 20 nanoseconds

the minimum delay was 40 nanoseconds and delays of several tenths of a microsecond were encountered when cathode voltages were raised to the values required for outputs over 100 kw peak.

The minimum delays are believed to occur when electrons are able to pick up enough energy on a single cycloid trajectory during the starting interval to make the secondary emission ratio greater than one. Delays are encountered when this ratio does not exceed unity. Under this condition, an initial interval is required during which space charge is accumulated by trapping electrons which give up energy to the RF fields. This stored space charge gradually lengthens the cycloid trajectories until the required energy for unity secondary emission ratio can be picked up along a single cycloid trajectory. Once this happens, the tube proceeds to turn on rapidly. This mechanism accounts for the delay interval (before starting) in which nothing appears to happen, followed by the very sudden turn on. It also accounts for the dependence of delay on peak input power. The dependence on voltage is a consequence of the necessity of the cycloid trajectories, along which electron multiplication occurs, maintaining synchronism with the circuit wave. At the higher cathode voltages, the conditions for synchronism are violated and multiplication can occur along only a few cycloids before the trajectory falls out of synchronism with the circuit wave. Thus, the rough outlines of a satisfactory theory of starting are on hand.

For our present purposes, it is sufficient to note that starting delays are small under normal operating conditions but can become appreciable (several tenths of a microsecond) at low input powers and high cathode voltages. Under such conditions, the amount of delay is very sensitive to voltage. The range of conditions over which delays are negligible can be increased by improving the secondary emission properties of the cathode so that less striking voltage

is needed for unity secondary emission ratio. Since they depend on transit times of electrons, the delays will scale down proportionally to wavelength in higher frequency tubes.

The measurements of starting delay in tube G35F-1 provided some insight to the behavior of the tubes being life tested. The starting delays observed can be explained in terms of changes in RF input power and operating voltage. Tube D37F showed no sign of either starting delay or jitter after 750 hours of operation, while the other tubes had both starting delay and jitter. The simultaneous occurrence of delay and jitter was also observed in other tubes life tested. Thus, there exists a strong possibility that starting jitter will only appear in a tube that has starting delay. A possible explanation for this lies in external circuit properties and in the sensitivity of delay to voltage under conditions of low RF input drive. Thus, if starting delay occurs on one pulse, less total energy may be drawn from the energy storage capacitor than if no delay had occurred. If external circuit time constants are long enough so that this higher voltage is still applied to the tube at the start of the next pulse, the delay for the next pulse may be even greater. An effect in the opposite direction is to draw greater peak current and greater total energy from the storage capacitor for higher cathode voltages. Between these two effects, there may be a tendency for a slight pulse-to-pulse jitter in applied voltage and therefore in starting delay. Proper design of the external circuitry may possibly eliminate jitter though it will not affect starting delay which is a property of the tube itself.

Tube I15F was constructed similar to the first two tubes of the Group D series but with an anode-cathode spacing between that of the first two tubes. Initial performance of this tube appeared to be very good. Unfortunately before tests could be run, the input window of the tube cracked as a result of mechanical strains imposed

by a revised coaxial line set up. The tube was repaired by replacing the input window without opening the main body of the tube and was then returned to test. Characterization of the tube after this repair showed significant degradation in its performance level, probably as a result of degradation of the cathode surface when the tube went down to air. This is the first indication we have had of an aluminum cathode surface degrading. This, however, was the result of the highly unusual processing history outlined above. Characteristics measured on tube I15F are shown in Figures 81 through 83.

Tube I28F was built similar in all respects to tube F63F. The results of measurements on this tube are shown in Figures 84 and 85. The performance was essentially similar to that of tube F63F.

Tube J4F was similar to tubes F63F and I28F in all respects, except that a matrix type impregnated cathode was employed in place of the usual aluminum cathode. This cathode was obtained from Semicon Associates. It was impregnated with a mixture usually employed in S-F-D laboratories' magnetrons and higher frequency crossed-field amplifiers. The cathode was outgassed to a temperature in excess of 800°C in a bell jar before incorporating it in the tube. However, because the tube has no heater, there was no means of activating the cathode once inside the tube. Semicon Associates believed that activation was necessary only to develop thermionic emission, not secondary emission. There is good reason to believe that this would be the case, since the impregnant includes a large amount of aluminum oxide which is known to be a good secondary emitter, even before cathode activation. This procedure has been used successfully at S-F-D laboratories on X-band crossed-field amplifiers employing liquid cooled cathodes. Experiments using a secondary emission tester had also demonstrated a secondary emission ratio at least equal to that of aluminum. Tests on tube J4F, however, showed very poor performance. It would appear that these results could only be attributed to a comparatively inactive

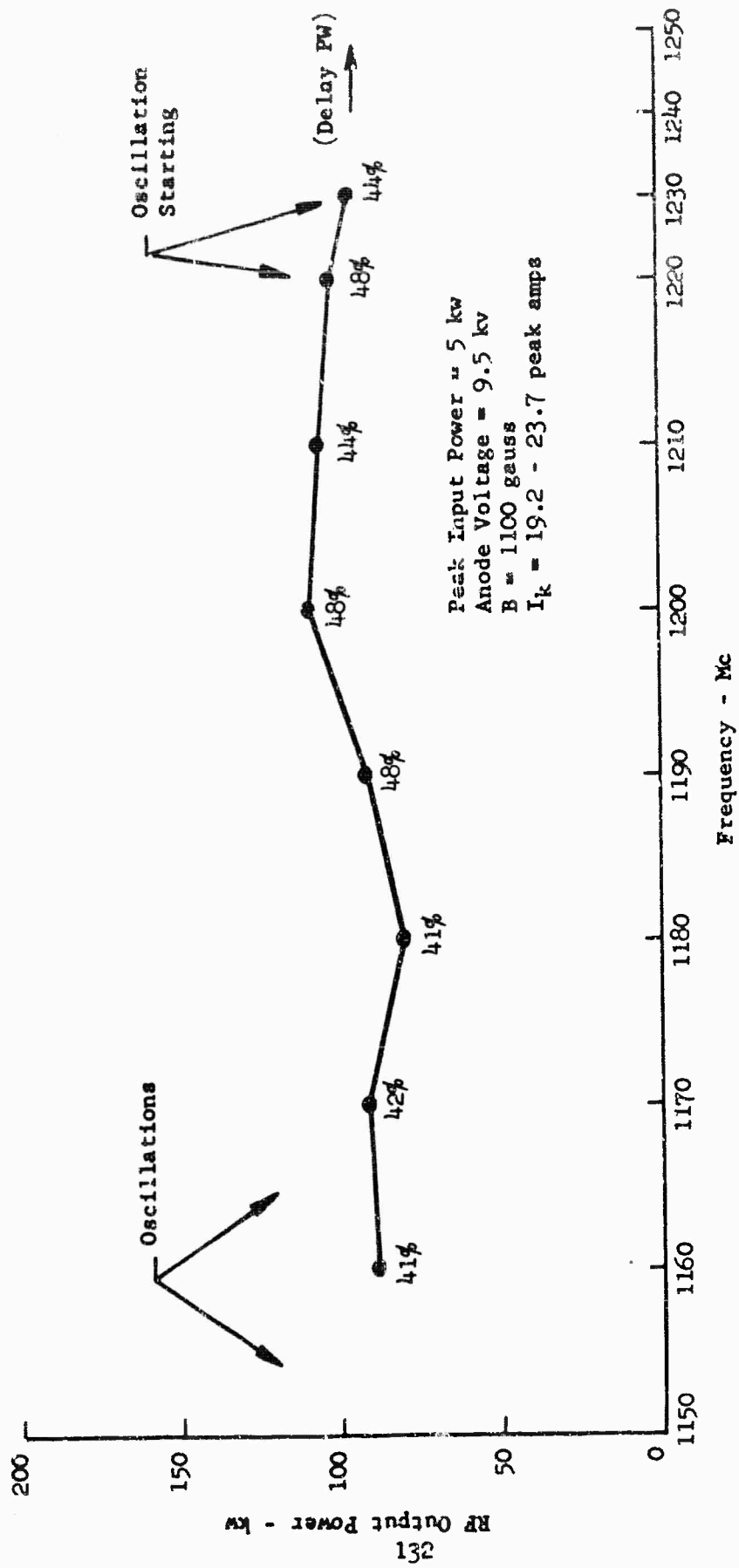


FIGURE 51 OUTPUT POWER VERSUS FREQUENCY AT FIXED ANODE VOLTAGE
 FOR SFD-209, TUBE 115F

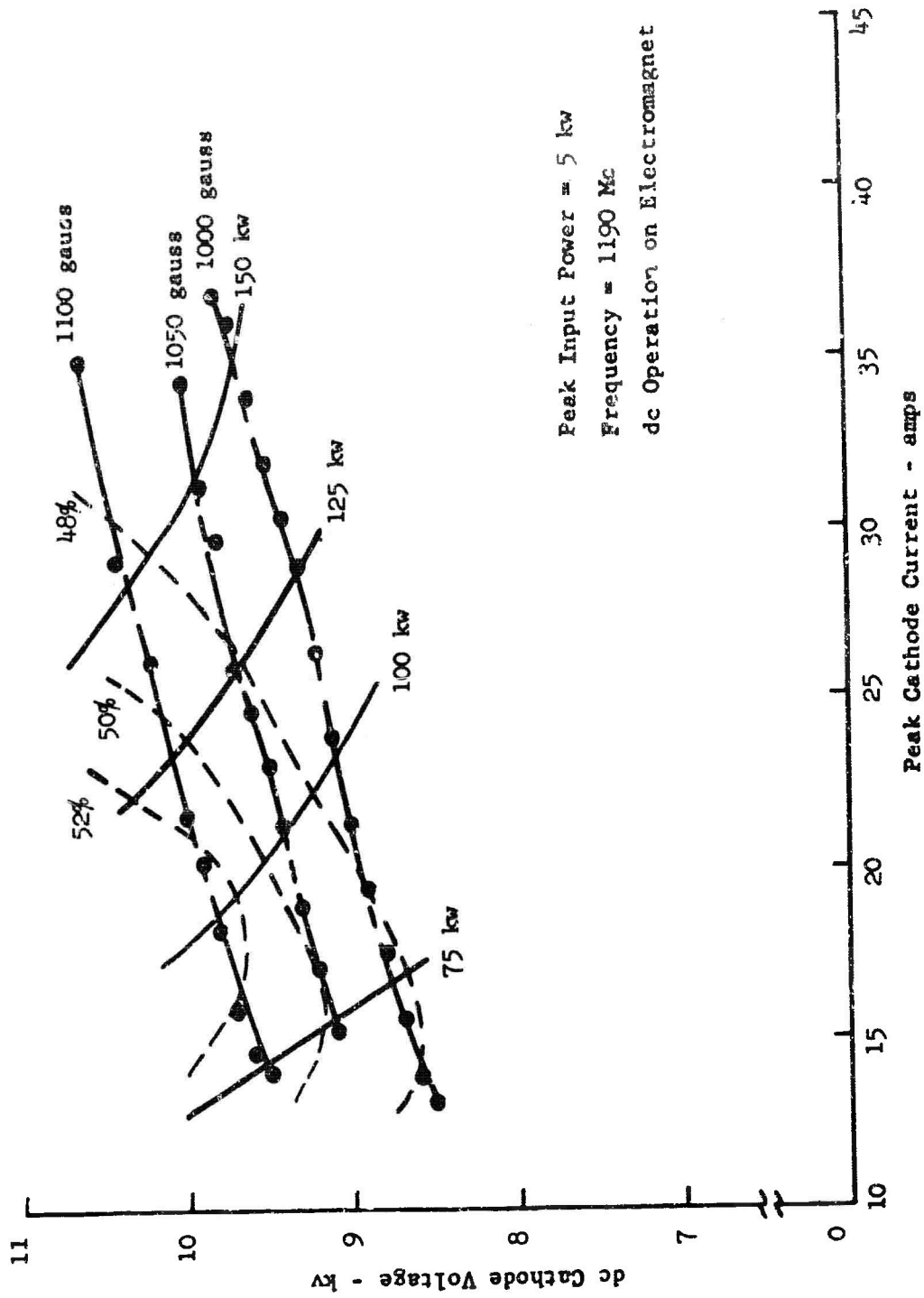


FIGURE 82 PERFORMANCE CHART FOR SFD-209, TUBE 115F

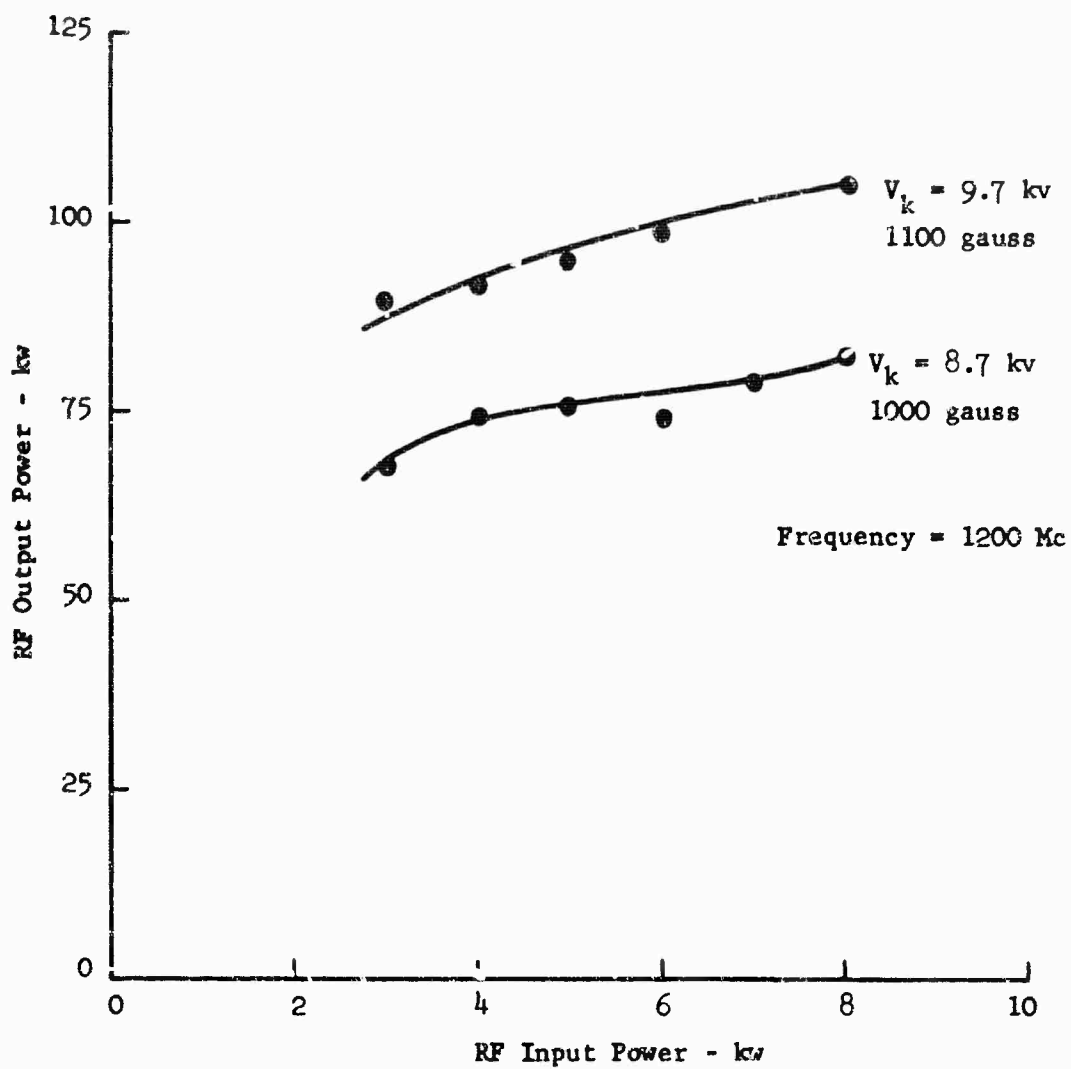


FIGURE 33 POWER OUTPUT VERSUS POWER INPUT AT FIXED VOLTAGE FOR SFD-209, TUBE 115F

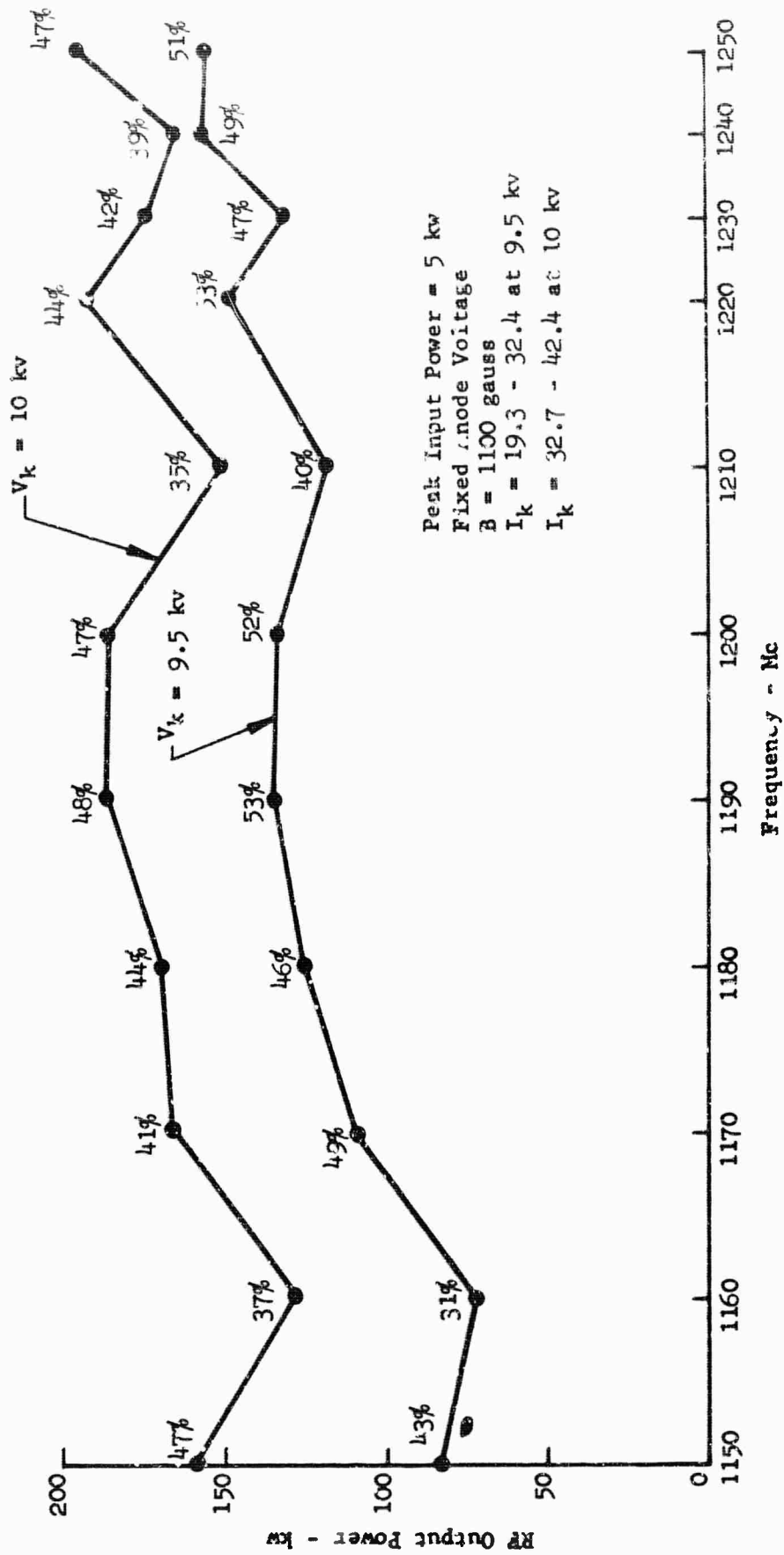


FIGURE 84 OUTPUT POWER VERSUS FREQUENCY AT FIXED ANODE VOLTAGE FOR SFD-209, TUBE 128F

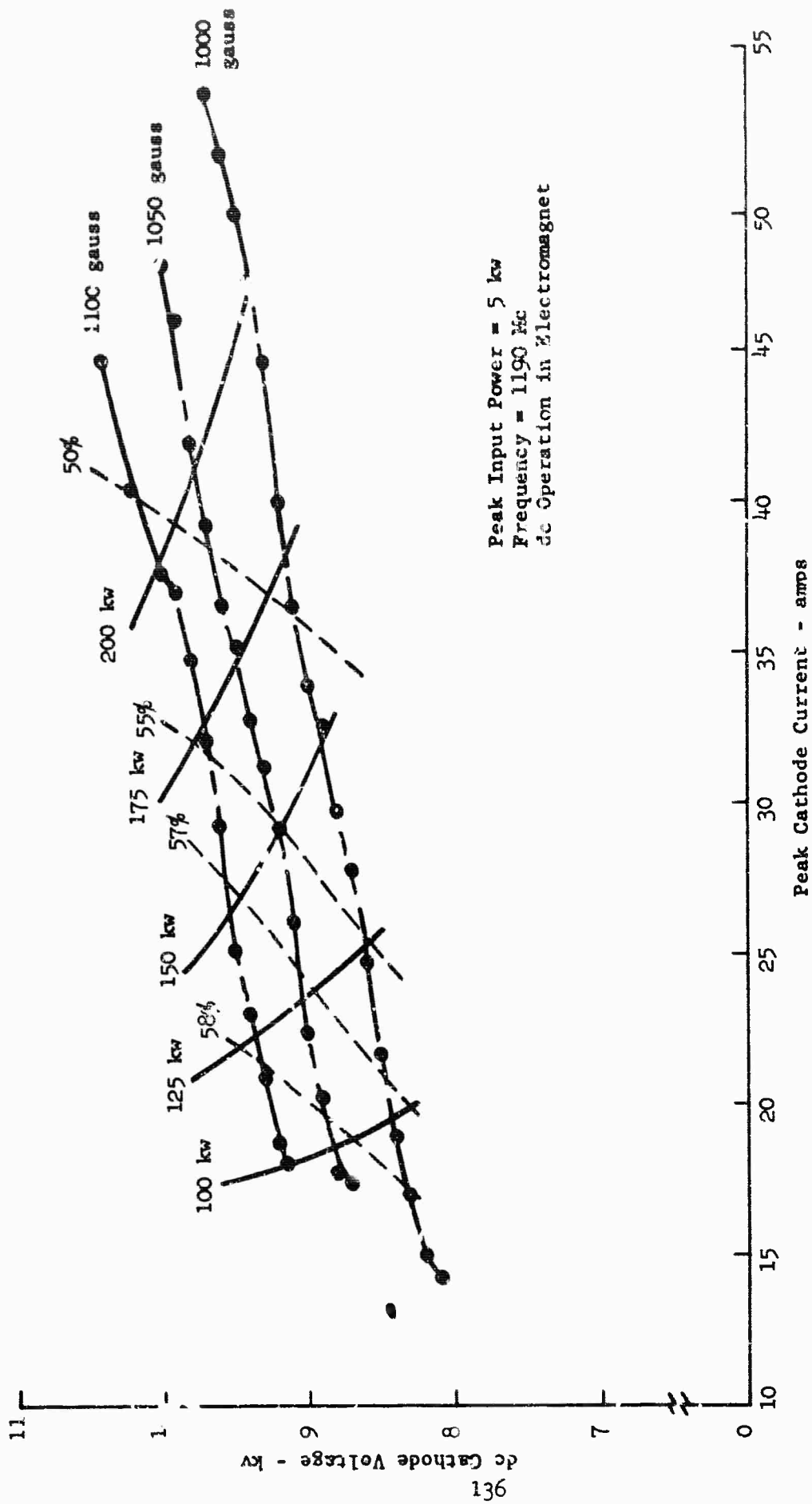


FIGURE 85 PERFORMANCE CHART FOR SFD-200, TUBE 128F

cathode surface. Some of these results are shown in Figure 86. We can only conclude that somehow we have not managed to obtain as high a secondary emission ratio in these experiments as we have at X-band. This could be connected with the difficulties encountered in heating the large L-band cathode in a bell jar. The 800°C temperature obtained in this step is significantly lower than the temperature to which the X-band cathodes are exposed.

Tube L7F was constructed similar to tubes F63F and I28F in all respects, except that the end hat diameter was made larger so that the end hats extended half the distance from cathode to anode rather than the usual one-third. This change was made because results on an X-band crossed-field amplifier program indicated an improvement in efficiency by as much as 5 percentage points when the end hat extension was increased from one-third to one-half of the anode-cathode spacing. Results measured on tube L7F are shown in Figures 87 through 89. The performance chart of Figure 87 shows a significant increase in efficiency as a result of the larger diameter end hat. Note that the shape of the efficiency contours on this performance chart is different from that for the smaller diameter end hat design as shown in Figures 71 and 74. The increased end hat diameter has resulted in flatter efficiency contours thereby maintaining high efficiency to higher values of operating current than has previously been the case. These flatter efficiency contours represent a very marked change compared with the shape of efficiency contours seen in a number of previous tubes. Therefore, we are justified in drawing a direct cause-and-effect relationship between the higher efficiency and the changed end hat diameter.

The design criteria previously employed (based on magnetron experience) called for the end hats to extend one-third of the cathode-anode spacing. The results on our X-band and L-band crossed-field amplifier programs now indicate that these magnetron criteria are not

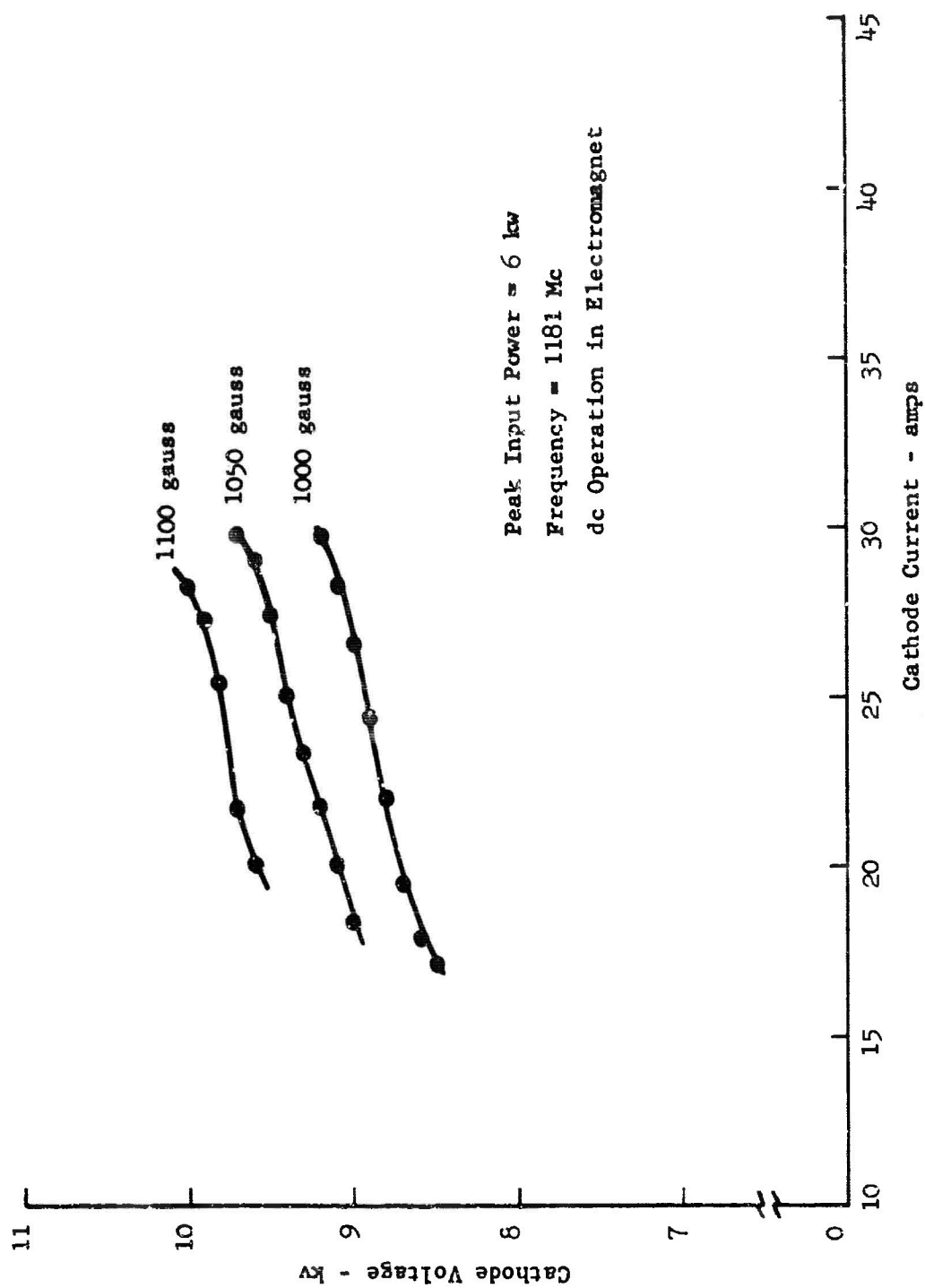


FIGURE 86 PERFORMANCE CHART FOR SFD-209, TUBE J4F

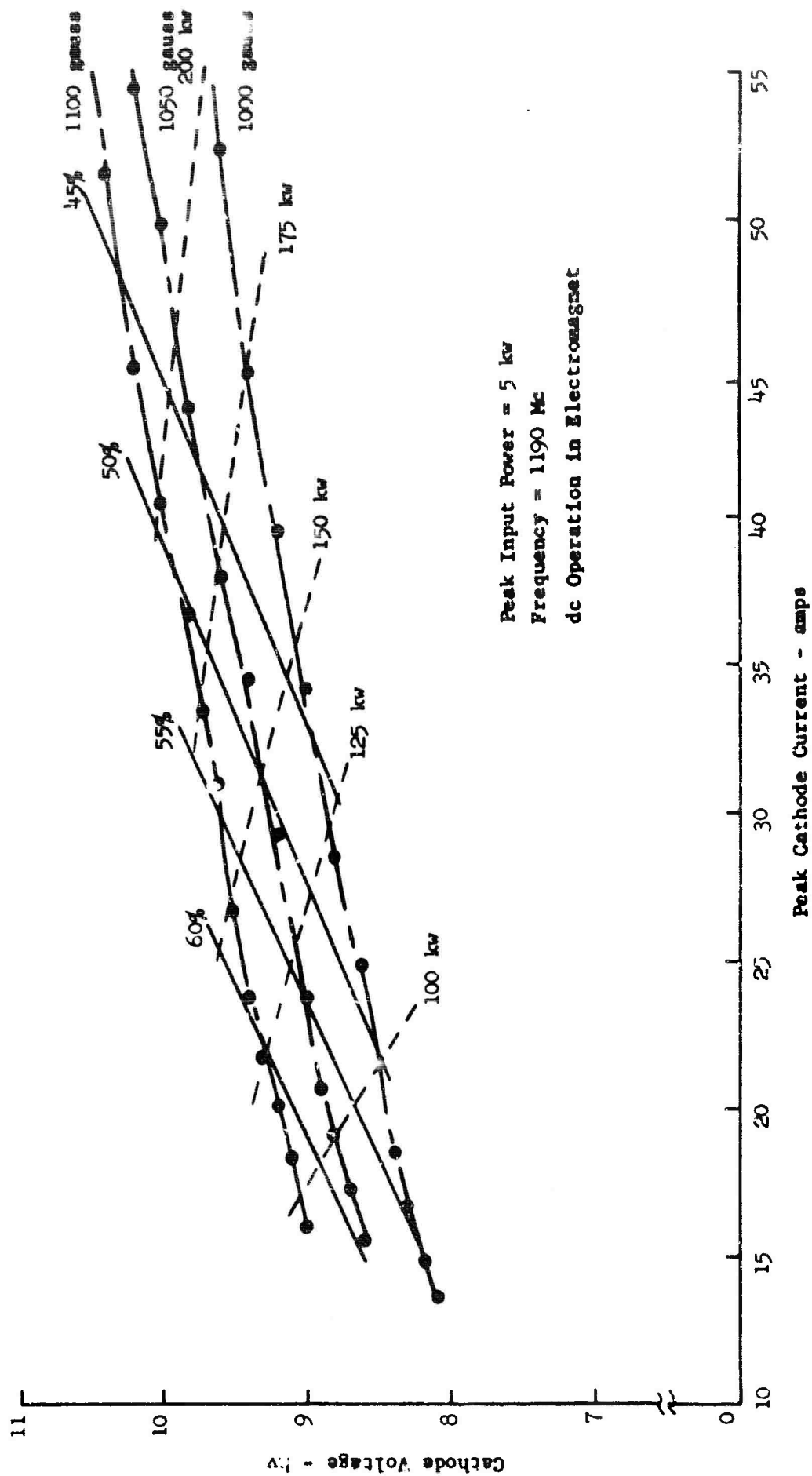


FIGURE 87 PERFORMANCE CHART FOR 5FD-209, TUBE L7F

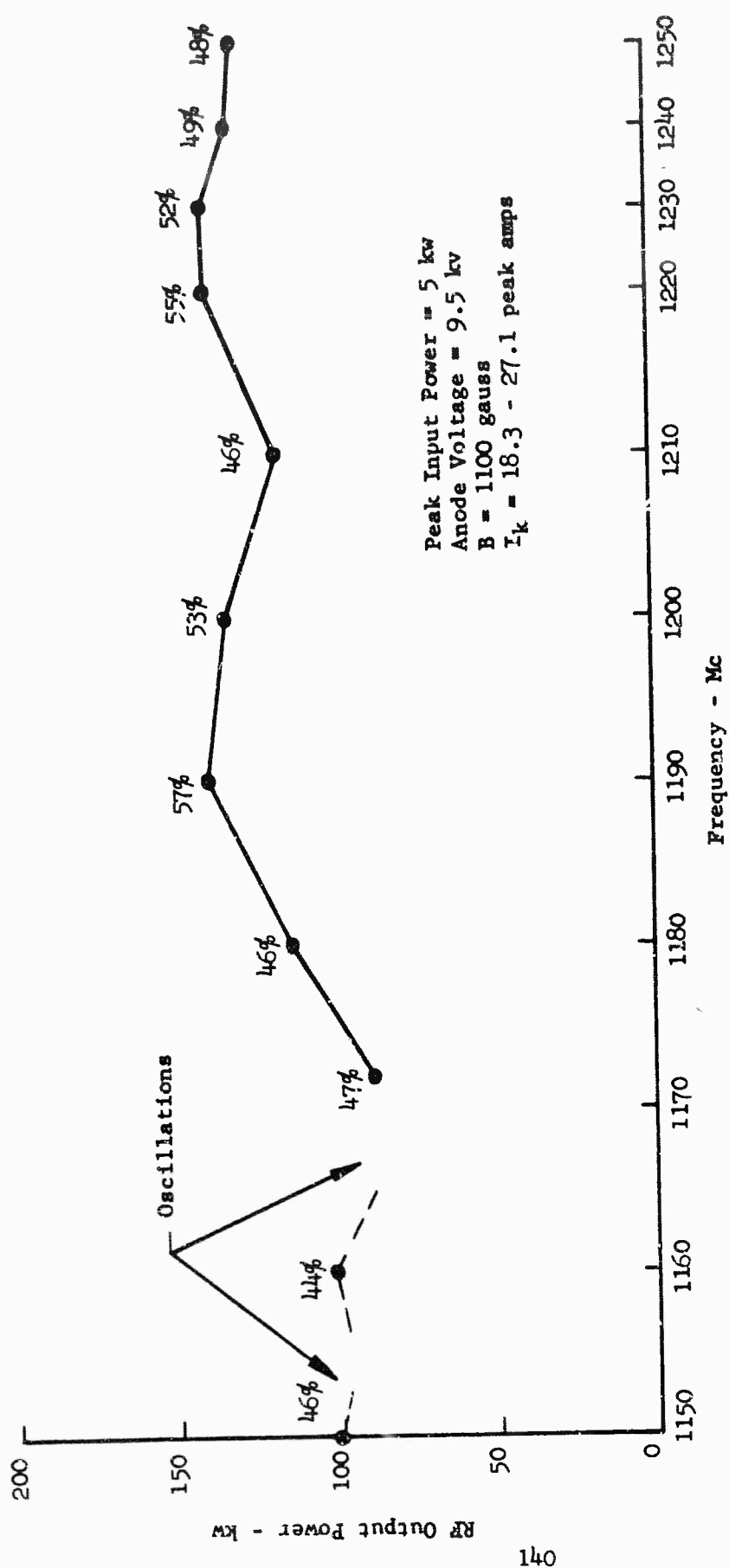


FIGURE 88 OUTPUT POWER VERSUS FREQUENCY AT FIXED ANODE VOLTAGE FOR SFD-200, TUBE L7F

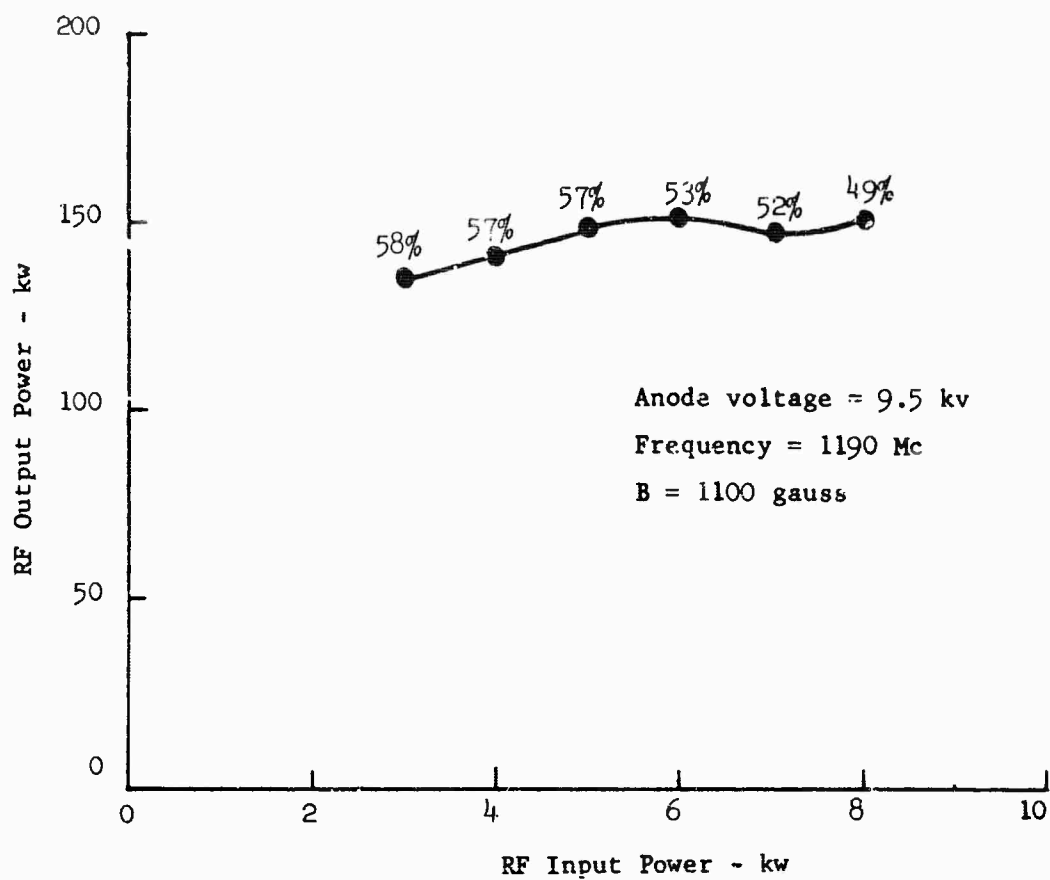


FIGURE 89 POWER OUTPUT VERSUS POWER INPUT AT FIXED VOLTAGE FOR SFD-209, TUBE L7F

adequate for crossed-field amplifiers and that additional space charge confinement is required. This probably results from electrons in the amplifier requiring a comparatively longer time to travel from the cathode to the anode because of the lower RF field strengths. This would be especially true at the input end of the tube. During the longer transit interval, the axial space charge forces can be more effective in forcing the electrons axially out of the interaction region. It is therefore reasonable to assume that a higher degree of axial confinement is required in a crossed-field amplifier than in a magnetron.

The frequency response curve for tube L7F (Figure 88) shows two holes near the lower frequency end of the band, which resulted from a poor match in these regions. As a result of the poor match, the tube did not amplify properly at these frequencies but instead oscillated at a band edge frequency. In general, the match of Group D tubes has not been as good as that of the good Group C tubes. Figures 90 and 91 compare the transmission and impedance matching characteristics of tube D37F of Group C and tube L7F of Group D. Tube L7F has the poorest match encountered among the Group D tubes. In both Figures 90 and 91, the match represents the reflection from a completed tube with a termination on the opposite port of the tube, and thus represents the reflections from both the tube input and the tube output transducers. A single one of these transducers would have a reflection equal to about half that of the maximum values shown by the interference pattern. Figures 90 and 91 show a much better match in tube D37F of Group C than in tube L7F of Group D. This better match is reflected in the tube performance. The characteristics measured on tube D37F, as shown in Figures 58 and 59, are the best we have obtained. Throughout the construction of the Group D tubes, we have looked for the significant difference that resulted in the poorer matches of the Group D tubes. On some of these tubes,

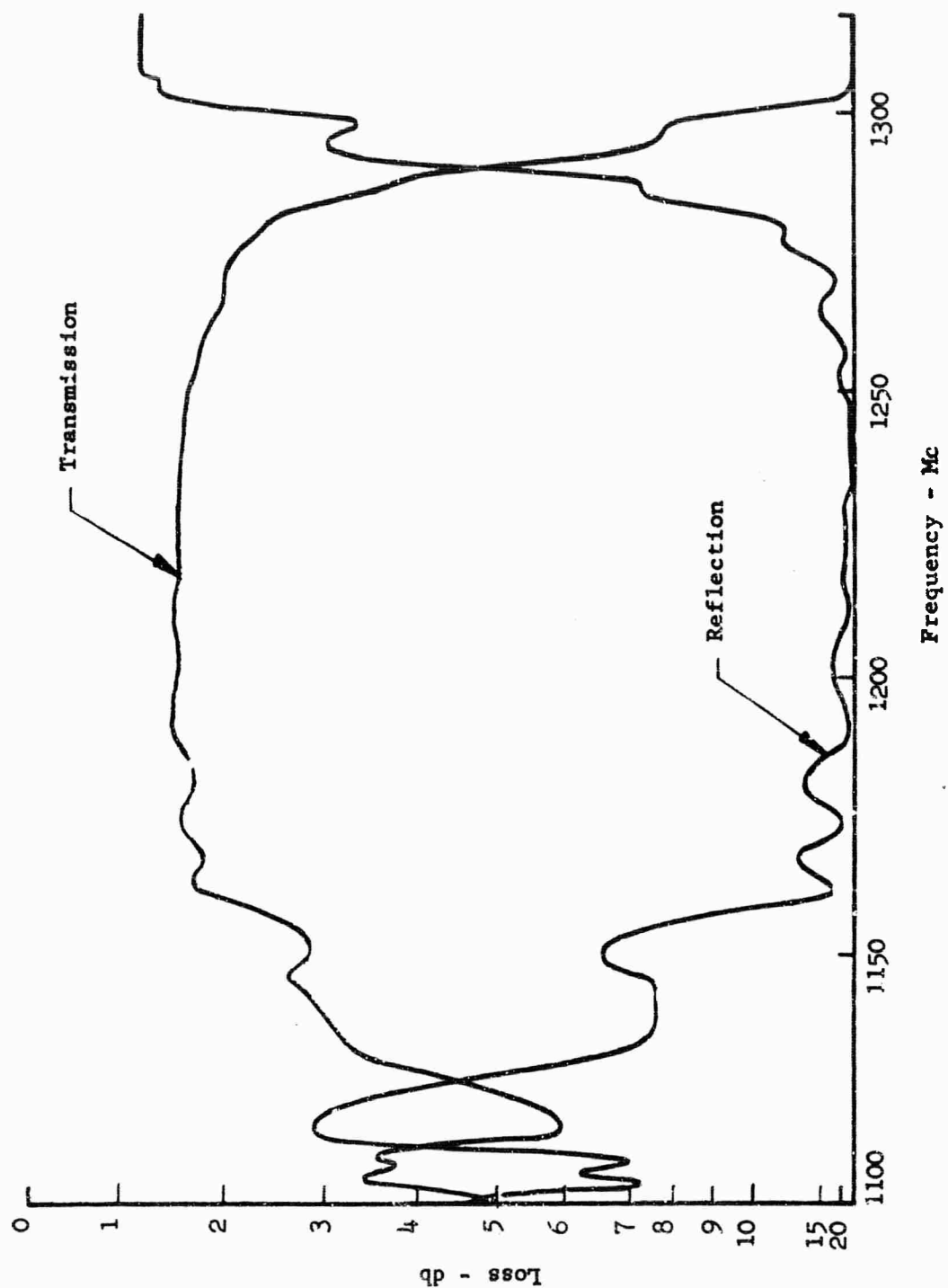


FIGURE 90 TRANSMISSION LOSS AND RETURN LOSS AS A FUNCTION OF FREQUENCY FOR STD-209, TUBE D37F

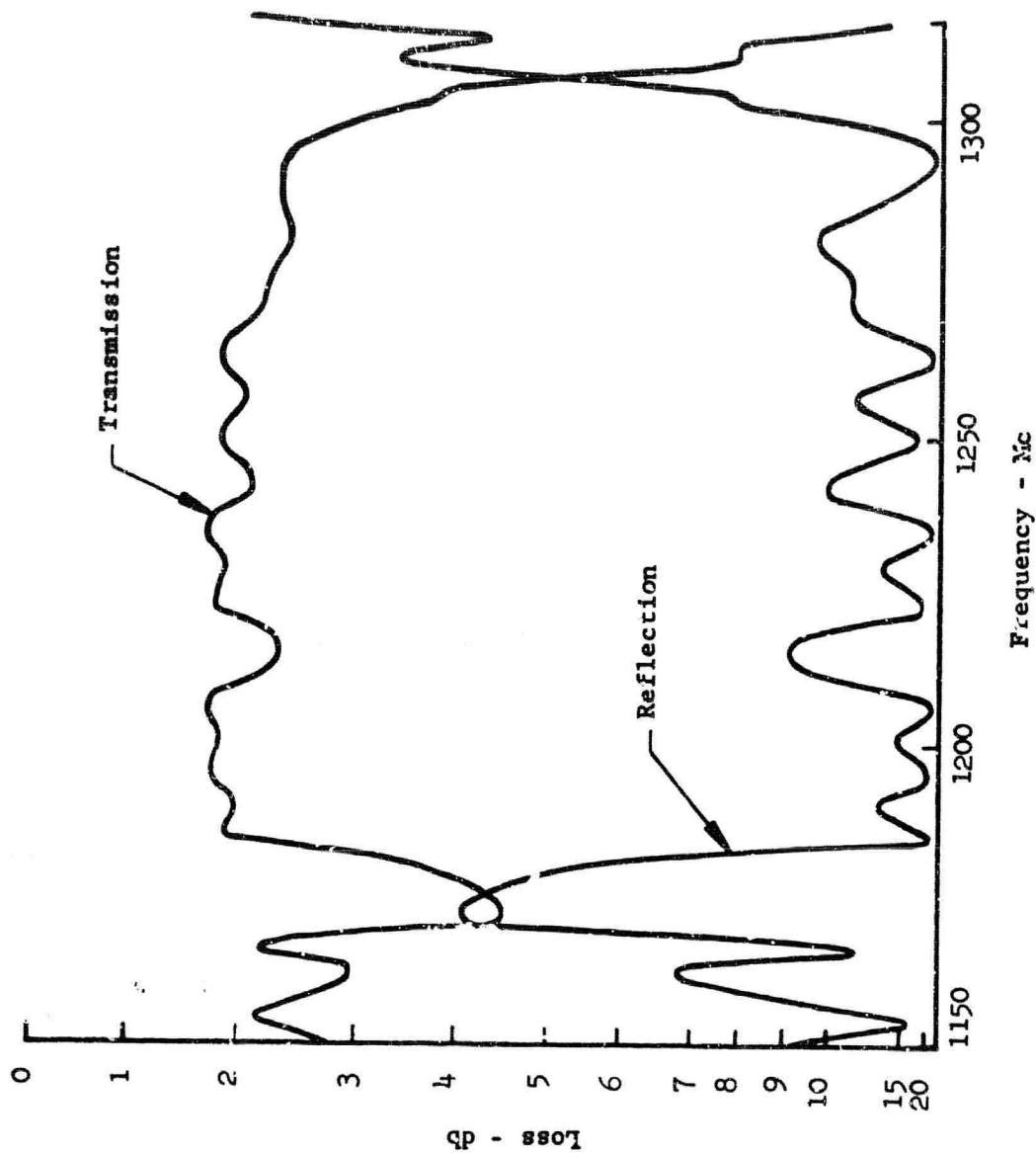


FIGURE 91 TRANSMISSION LOSS AND RETURN LOSS AS A FUNCTION OF FREQUENCY FOR SFD-209, TUBE L7F

we were never able to get as good a match as that on tube D37F. On others, we initially obtained a good match, but the match degraded on successive steps during assembly and bake out of the tube. We originally suspected problems with the output coaxial line and U-bend and still believe that some of the problems may be associated with this assembly. However, during the past quarter, two additional sources of variations of the impedance match have been discovered. Both of these are quite subtle. One results from the exact shape of the matching tab which is used to apply a capacitive susceptance in shunt across the first resonator. Depending on exactly how this tab is shaped, there is a different ratio of the capacity introduced from the coaxial line to the first and to the second anode circuit bar. A comparison of the tab shape in tube H29F of Group D, which was opened, with the tab shapes of Group C tubes, which were opened, showed that there was a significant change in this shape between the two groups of tubes. This tab is hand-formed during the assembly process and the change in shape apparently coincides with a change in assembly personnel assigned to the program. In the future, this problem may be avoided through a change in the matching tab design. The tab may thus be split into two sections - one section will introduce capacity only from the coaxial line to the anode circuit bar and will be so shaped that negligible capacity is introduced to the second bar; the second portion of the tab will be used to adjust capacity from the coaxial line to the second anode bar. This tab will overlap the capacitive coupler extending from the coaxial line to the second bar, since this adjustment is in parallel with this coupler. In this manner, the exact shape of the tabs used for this matching procedure will not be critical and essentially the lumped capacity introduced from coaxial line to the first and second anode bars will be the only parameter of significance.

Another source of variation in the impedance match is believed to be the motion of the first anode bar during the window and coaxial line assembly brazes or during take out. This first bar is rigidly connected to the center conductor of the coaxial line. Because of difficulties in induction brazing the output coaxial line assembly, we suspect that significant stresses are applied to this center conductor. These may cause the center conductor to move slightly. If this center conductor is off center by a moderate amount, there is still very little reflection introduced. However, the motion of this conductor can also push the first bar of the circuit, causing it to move underneath the matching tab. Since the spacing between the first anode bar and the matching tab is critical, this movement produces a significant change in the circuit match. A more flexible connection between the first anode bar and the center conductor of the coaxial line thus appears to be necessary. In addition, rearranging the coaxial line assembly procedure to eliminate stresses during brazing operations is required.

Phase sensitivity as a function of voltage and phase linearity as a function of frequency were measured on tube L7F. Figure 92 shows a block diagram of the test set used to make these measurements. This system has been used in phased array studies at Lincoln Laboratories. A simultaneous amplitude and phase balance is required and the system is very sensitive to the balance conditions. When used in a pulsed system, the widths of the pulses fed into the two arms of the bridge must be the same in order to prevent overloading the video amplifier. If the pulses do not have the same width, relatively large spikes will appear at the leading and trailing edges when the pulse centers are balanced. Fortunately control electrode operation from a dc power supply provides an extra degree of freedom in that the pulse widths of the two pulses may be closely matched.

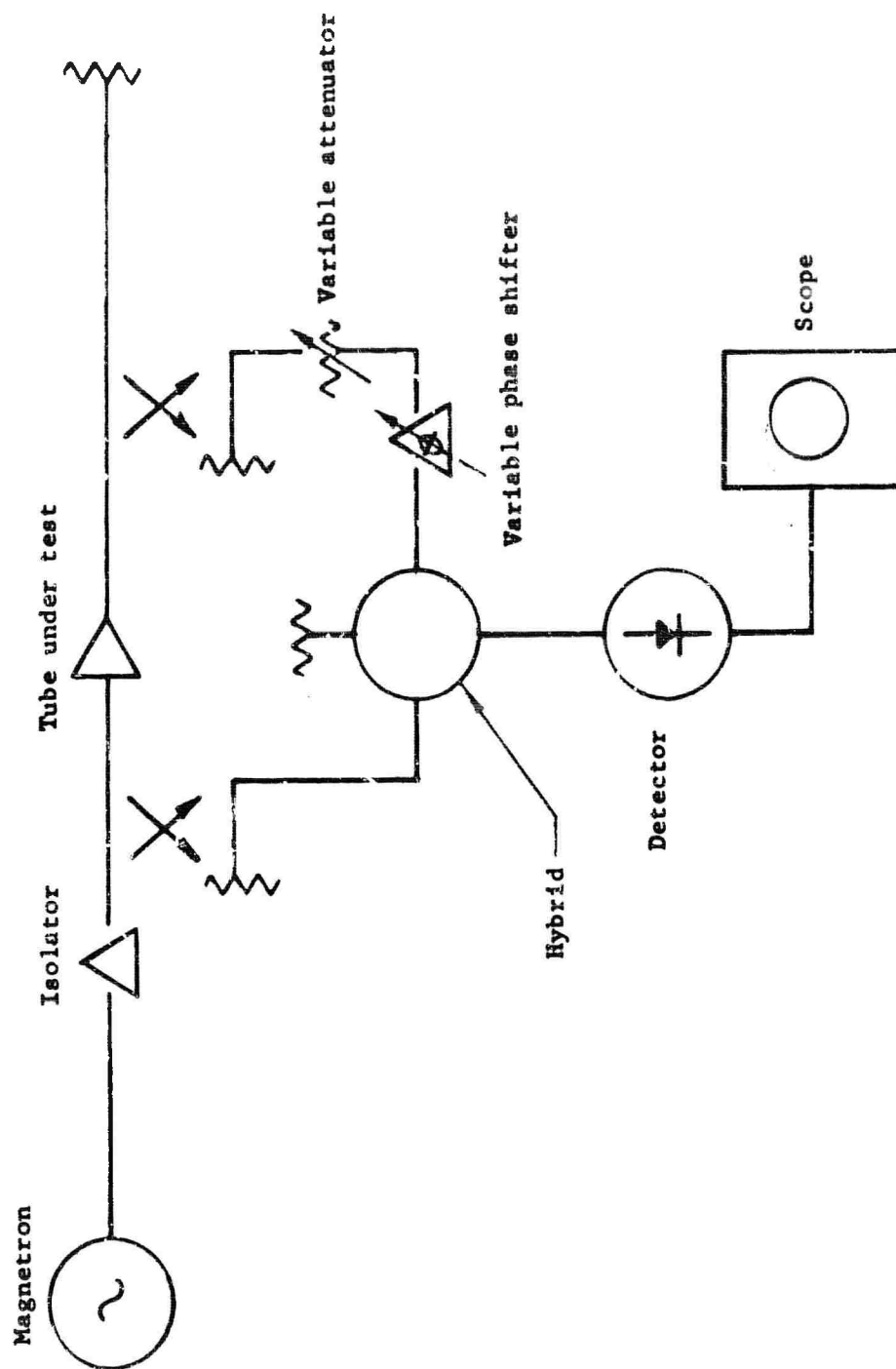


FIGURE 92 BLOCK DIAGRAM OF PHASE BRIDGE SET UP REQUIRING A SIMULTANEOUS AMPLITUDE AND PHASE BALANCE. A HYBRID JUNCTION IS USED AS THE PHASE COMPARATOR

Rigid coaxial lines were used throughout the test set so that, with proper padding, the measurement errors could be held to a minimum. The lines were adjusted so that the rate of phase change in each arm of the bridge was nearly equal in order to reduce the phase slope in the measurements. The load and source VSWR's as seen by the tube under test were approximately 1.2:1.

Figure 93 shows the phase linearity as a function of frequency with three different levels of input drive signal. Only the measurements from 1200 Mc to 1250 Mc have any significance since the match below this frequency range was extremely bad as shown in Figure 91. It should also be noted that the regions where deviation from linearity becomes significant coincide with the regions of abnormal match conditions in the tube. Even under these adverse conditions, the maximum peak-to-peak deviation is approximately 12° and is not significantly dependent upon drive power. The linearity slopes, however, are dependent upon drive power since varying drive power with constant voltage applied is analogous to varying the voltage with a fixed input drive which is a measure of phase sensitivity to voltage. Figure 94 shows the phase sensitivity as a function of applied dc voltage and power input. The measurements were made over a range of power output from approximately 100 kw to 150 kw.

□ 10 kw
 ● 5 kw
 ○ 3 kw
 /// Region of very poor match
 Anode Voltage = 9 kv
 B = 1100 gauss

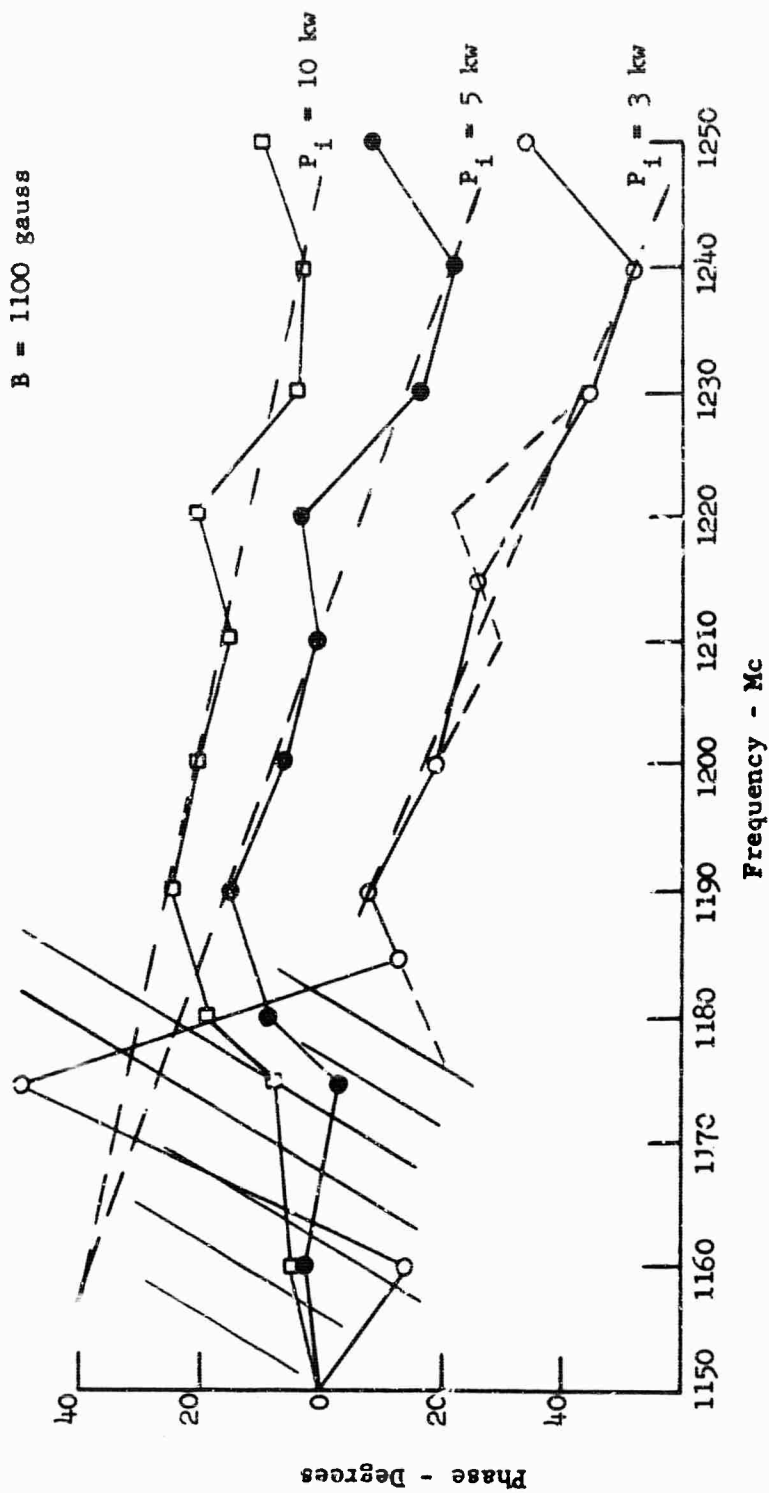


FIGURE 93 PHASE LINEARITY AS A FUNCTION OF FREQUENCY FOR SFD-209, TUBE L7F

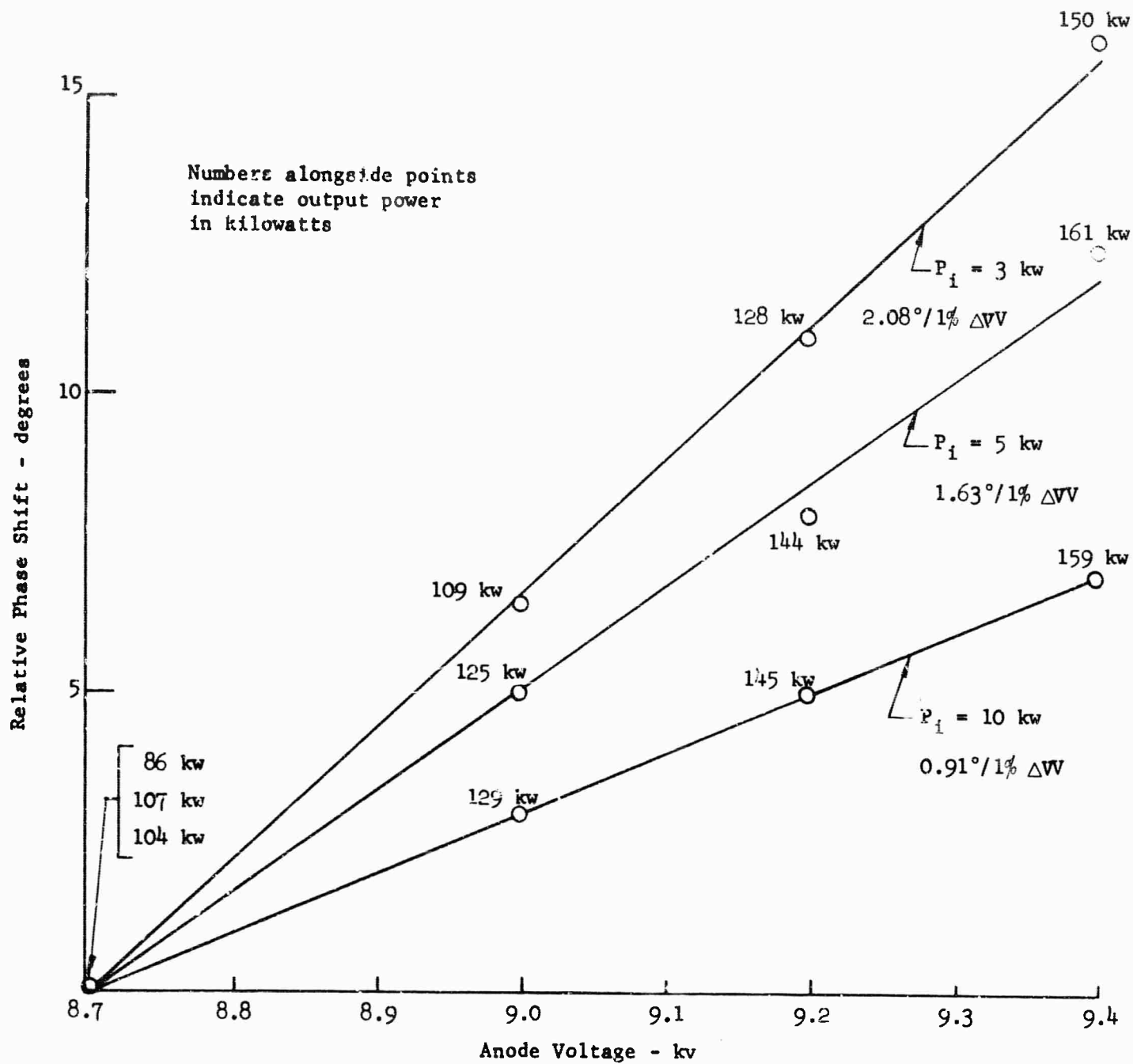


FIGURE 9¹ RELATIVE PHASE SHIFT VERSUS ANODE VOLTAGE FOR SFD-209, TUBE L7F

7.0 SUMMARY OF THE POWER TUBE DEVELOPMENT

The initial design of the power tube consisted of a cage of vertical bars with the cathode in the center. The circuit was made with 12 bars, 0.380 inch in diameter and approximately 4 inches long. The bars were cut out in the center, on the side away from the cathode, to provide room for the capacitive coupler rings. The coupling rings used an interdigital finger arrangement. The top flange of this circuit was made a flexible member to permit thermal expansion of the bars without distorting them. Figure 95 is a photograph of this circuit. This design was designated the SFD-208.

Matching of this circuit to a coaxial line was accomplished by attaching the center conductor of a coax line to an adjacent bar on the circuit. The outer conductor of the coax line served as a bar in the circuit. The tubes at this stage were demountable to facilitate changes in the cathode assembly and they were designed for use in an electromagnet. Figure 96 is a photograph of one of the first tubes of this type, and Figure 97 is a photograph of this tube mounted in the electromagnet in the hot test set.

The first version of the SFD-208 had a drift space and was intended to operate without electronic feedback. Tests on this model showed that proper spoke break up was not occurring in the drift space. Moreover, the characteristic of the tube was such that the residual feedback was degenerative at the band center and regenerative at the band edges. As a result, the tube would oscillate at the π mode frequency when driven at the band center and produce only a few db of gain at the band edges. Figure 98 shows the gain as a function of frequency for this design.

The second tube incorporated a slightly longer drift space but the improvement was not significant. It was then decided to try the opposite approach of introducing regenerative feedback. The tube was redesigned to have a one resonator drift space. This version was

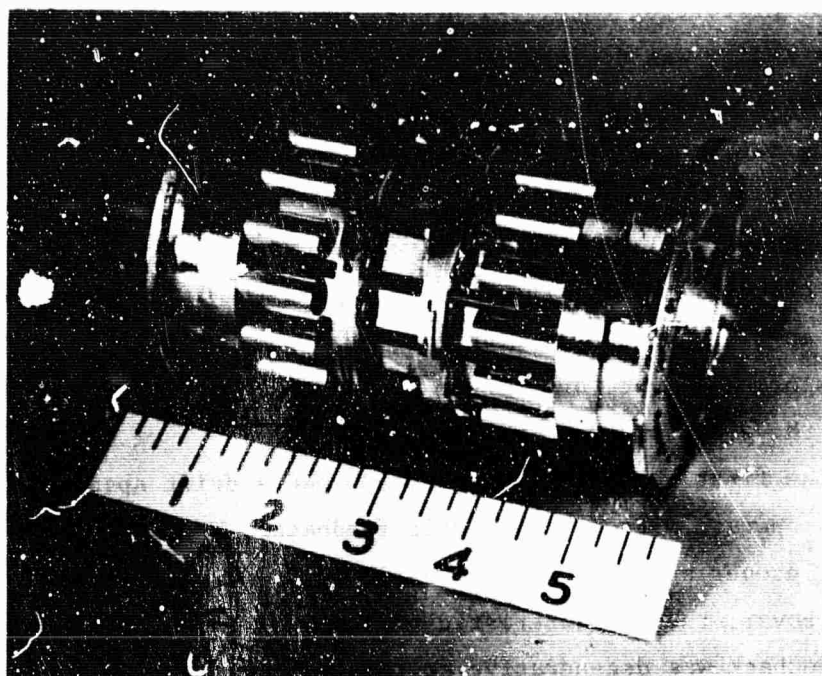


FIGURE 95 **PHOTOGRAPH OF THE SLOW WAVE CIRCUIT ASSEMBLY
FOR THE SFD-208 POWER TUBE**

The C_{13} couplers for this tube are of an interdigital type to minimize the incidental C_{12} and C_{10} capacities introduced by the couplers

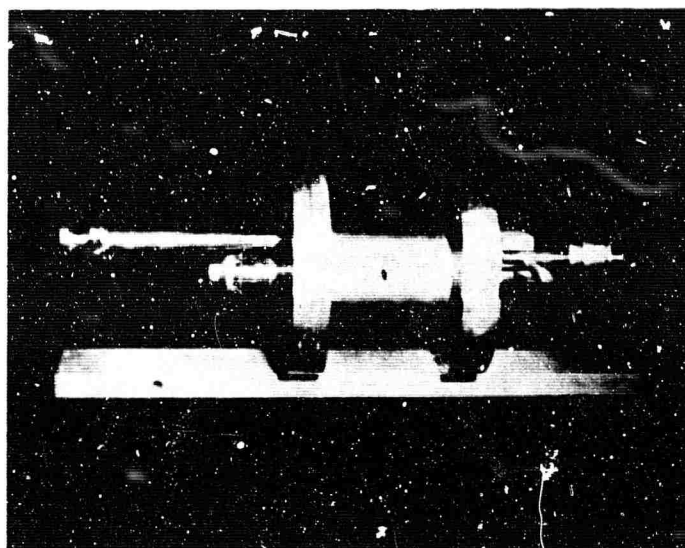


FIGURE 96 PHOTOGRAPH OF COMPLETED SFD-208 TUBE

An external pole piece for mounting the tube on an electromagnet used in early RF testing has been fixed to one end of the tube

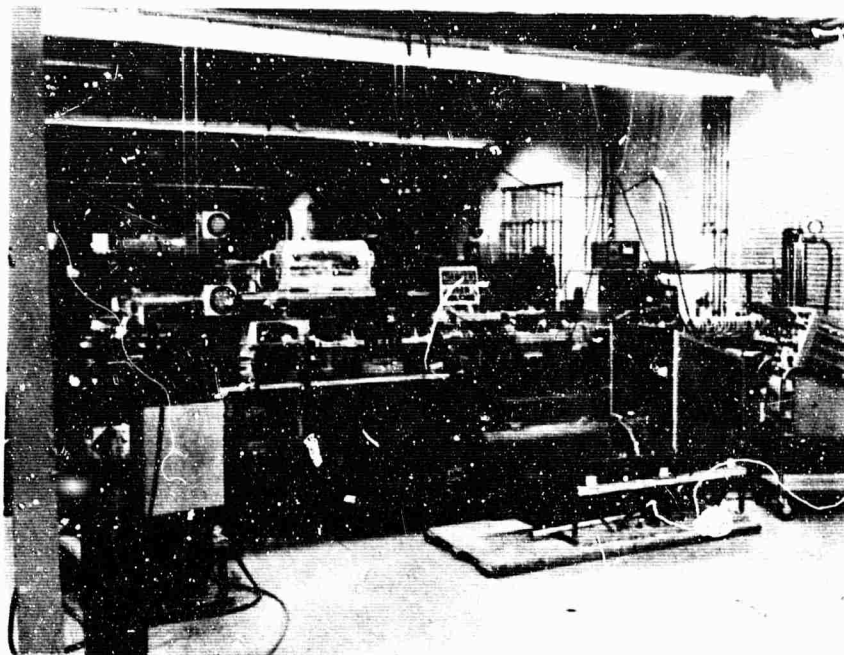


FIGURE 97 PHOTOGRAPH OF THE SFD-208 HOT TEST SET

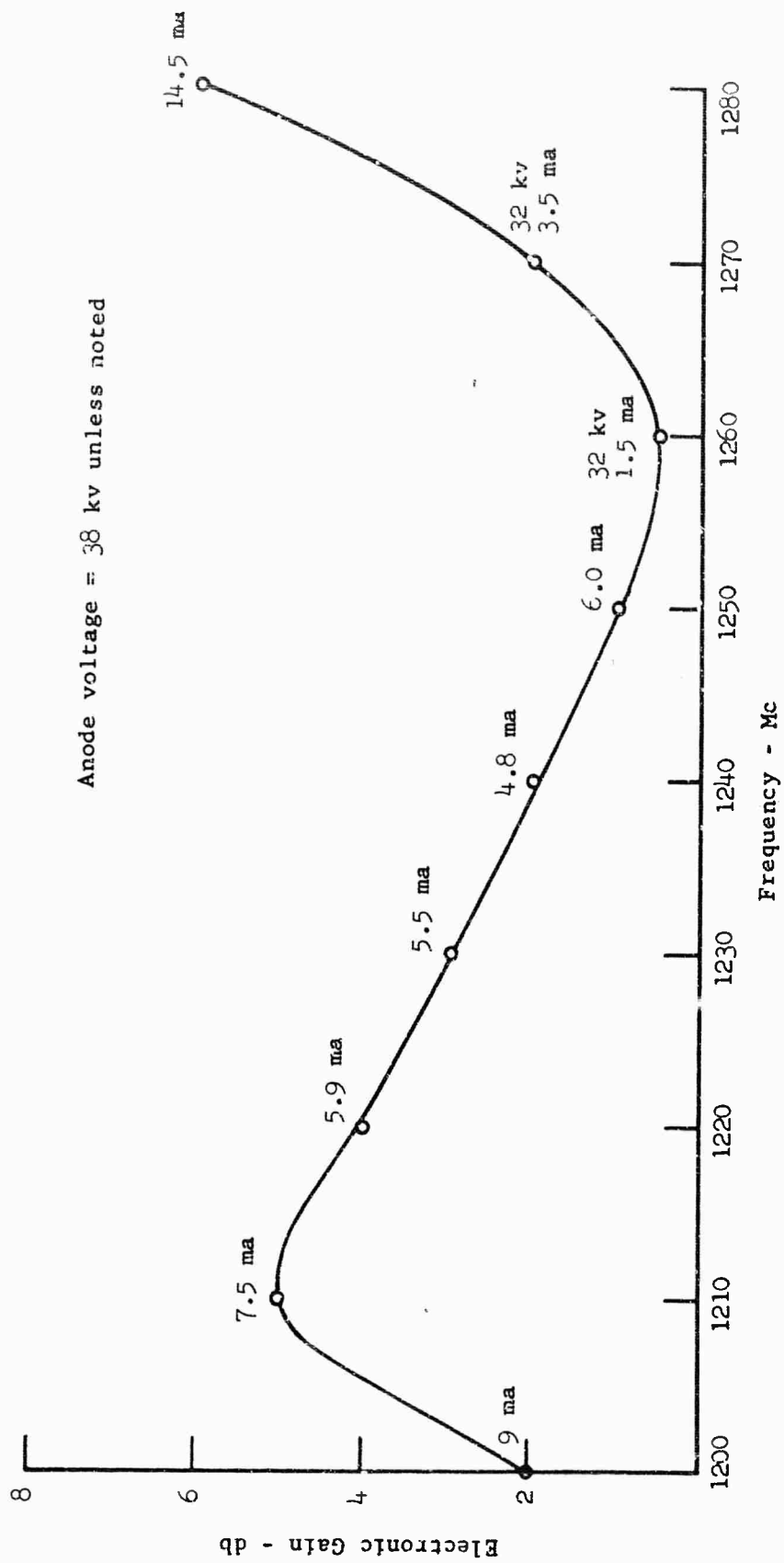


FIGURE 98 GAIN AS A FUNCTION OF FREQUENCY FOR THE SFD-208

operated successfully but only over a restricted bandwidth. A performance chart at band center is shown in Figure 99 and a frequency response curve is shown in Figure 100. Outputs of about 500 kilowatts over a 5% bandwidth were obtained. The tube was again redesigned to increase the bandwidth by increasing the capacitive coupling of the slow wave circuit bars. To obtain adequate bandwidth in the presence of positive feedback, a cold band pass of 40% was required. With this much coupling, we ran into difficulties with a second pass band located above our desired band. This upper pass band was associated with the inductance in the capacitive coupling rings. Attempts to eliminate this problem were not successful and the program was slowed down with this approach. At this point it was decided to redesign the power tube to resemble the more successful driver tube. This design was designated the SFD-216. The number of bars in the circuit was increased to 20 and the bar diameters decreased to 0.250 inch. Figure 101 is a photograph of an anode assembly for this design. Figure 102 is a summary of the important design dimensions for the SFD-209 (driver tube) and the SFD-208 and SFD-216 (power tubes).

The first tube of this design was constructed as a demountable tube and was used successfully to evaluate six different cathode geometries. The primary problems encountered with this design were π mode contamination and RF pulse break up on the tail end of the pulse. A change in cathode geometry removed the π mode contamination but did not prevent the pulse break up. The pulse break up did not occur sooner than 1 μ sec after the start of the pulse so that, if short pulses of less than 1 μ sec were used, relatively high power could be achieved. If longer pulse lengths were to be used, we had to settle for reduced power output. Figure 103 shows a frequency response for this tube with a 1.0 μ sec pulse width.

A further difficulty with this design was that the operating band was centered in the region of high phase shift per section. The

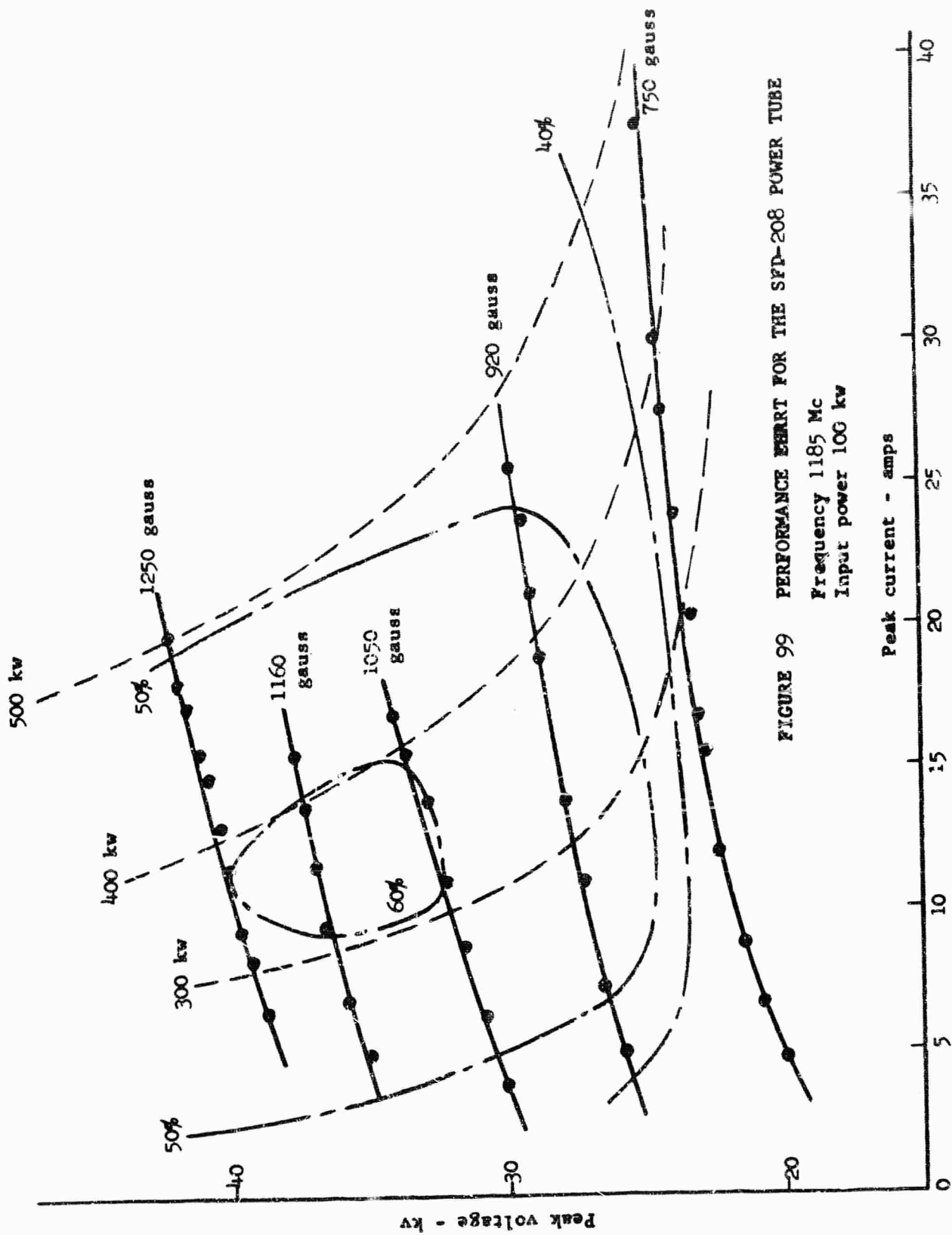


FIGURE 99 PERFORMANCE CHART FOR THE SFD-208 POWER TUBE
Frequency 1185 Mc
Input power 100 kw

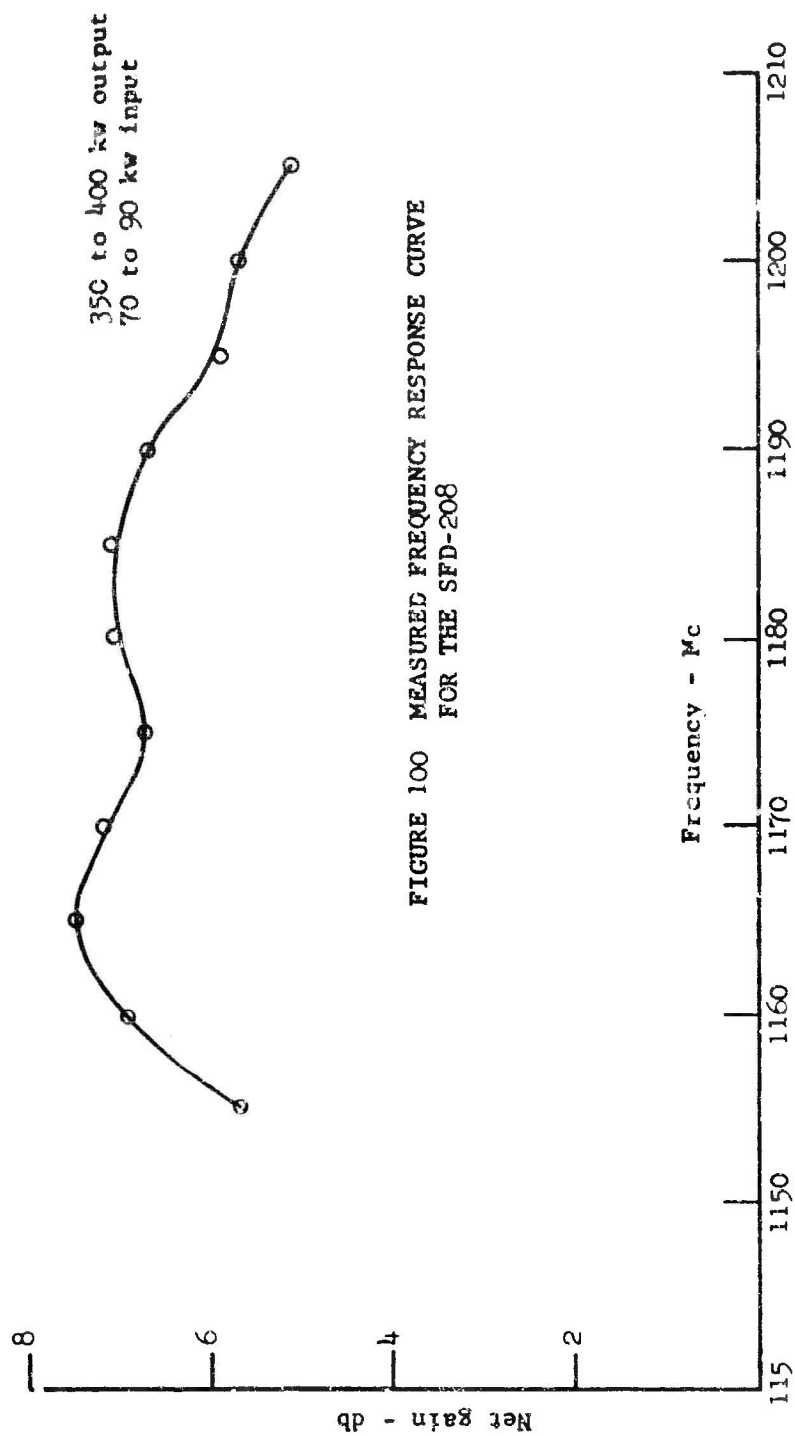


FIGURE 100 MEASURED FREQUENCY RESPONSE CURVE
FOR THE SFD-208

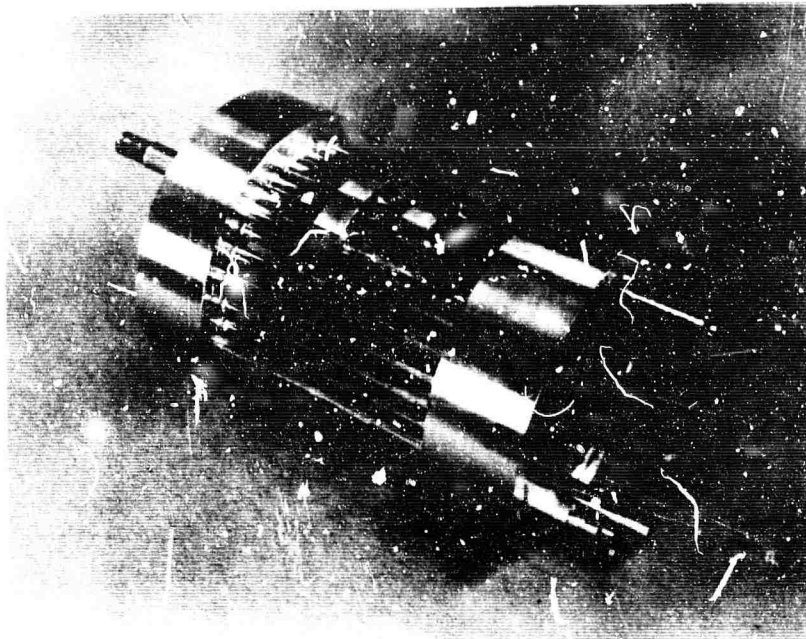


FIGURE 101 PHOTOGRAPH OF ANODE ASSEMBLY FOR SFD-216

	Driver tube <u>SFD-209</u>	Power Tubes	
		<u>SFD-208</u>	<u>SFD-216</u>
1. Anode bore diameter	2.750	2.250	3.00
2. Anode bar diameter	0.135	0.380	0.250
3. Anode bar length	3.800-4.000	3.350-4.000	4.000
4. Pitch angle of anode	8°	24°	13.3°
5. Cathode diameter	2.340-2.370	1.320	2.367
6. Cathode length	2.500	2.100-2.500	2.500
7. Can outside diameter	3.625	3.625	4.500
8. Can thickness	1/16	1/16	1/16

(All dimensions in inches)

FIGURE 102 PRINCIPLE DIMENSIONS OF THE TUBES

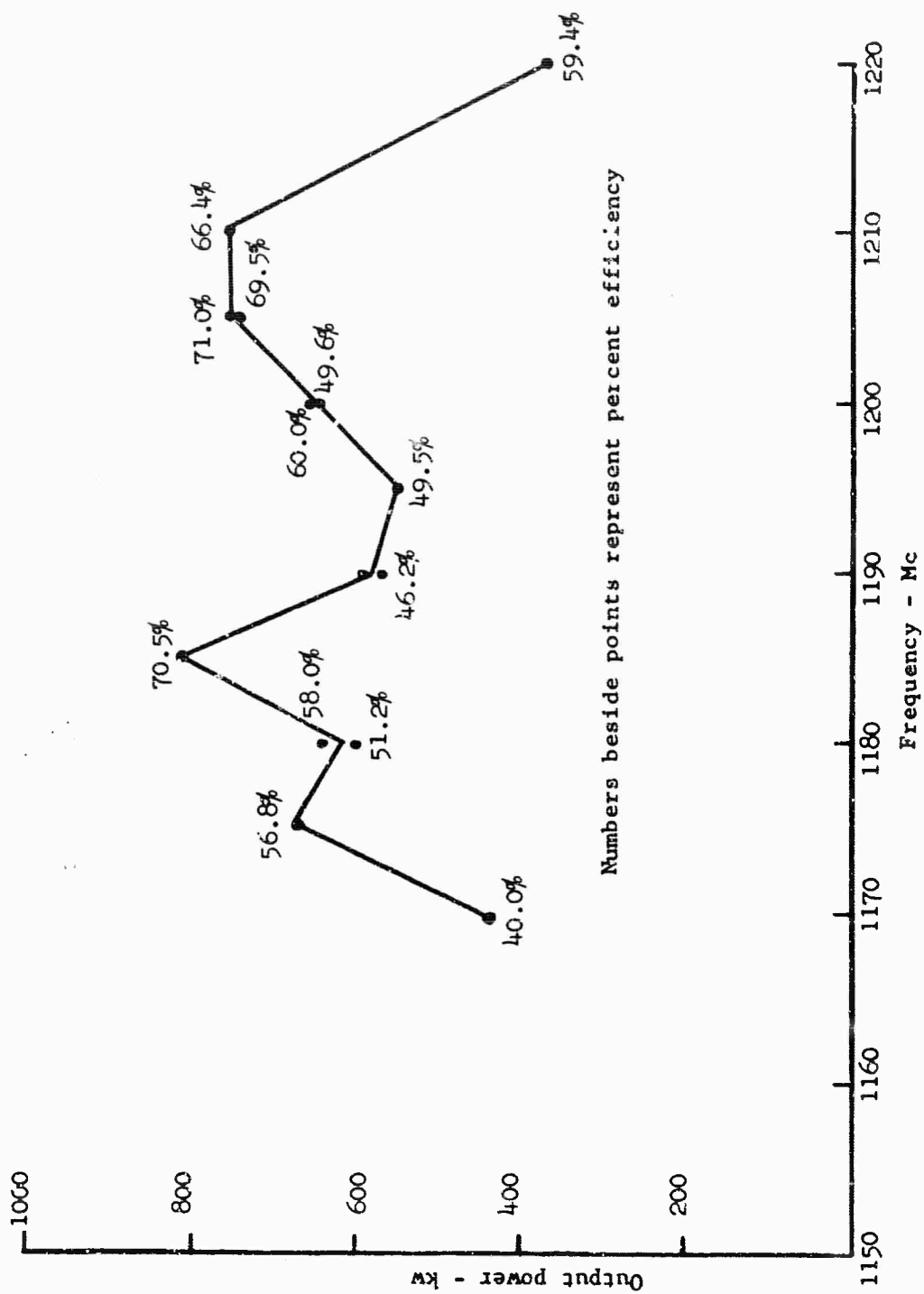


FIGURE 103 A FREQUENCY RESPONSE CURVE MEASURED ON THE SFD-216

lowest frequency range of 1150-1160 Mc which we could reach with our driver magnetron corresponded to phase shifts per section of over 130° as shown in Figure 104. We could not test the tube in the frequency range where phase shifts were lower and performance optimum.

With this background, it was decided to proceed to the next step which was to build a water cooled version for higher average power. Unfortunately the first tube of the water cooled version developed a crack in an insulating ceramic and, when voltage was applied, an arc through the ceramic punctured the vacuum envelope. The second water cooled tube (H20E) was successfully tested. Amplification was possible at low magnetic fields corresponding to operating voltages of 20 kv to 22 kv. Under these conditions, an output power of 800 kw was obtained for an RF input power of 80 kw. Efficiency was over 50%. Indications were that higher efficiency was obtainable with higher magnetic fields corresponding more nearly to the design value which would place the operating voltage between 25 kv and 30 kv. This tube was operated from a dc voltage with control electrode turn off. Under these conditions, a peak power output of 1 Mw with a peak RF input power of 150 kw was obtained at 22 kv. The average power in this test was slightly in excess of 1 kw.

A third water cooled tube (J26E) was built but, when it was hot tested, the cathode was found to be inactive. All the power tubes were constructed with beryllium copper cathodes and all were found to be inactive on their initial tests. We had not yet introduced the more successful aluminum cathode in the power tubes because of the higher powers required. Tests on the driver tube with aluminum cathodes indicated that we could obtain over 1 kw average power with aluminum. Before we could incorporate the aluminum cathodes in the power tube, it was decided to suspend effort on the power tube to insure construction of a group of final design driver tubes for life test and phase measurements. These portions of the program were considered of primary importance and it was felt that they should be emphasized at this time.

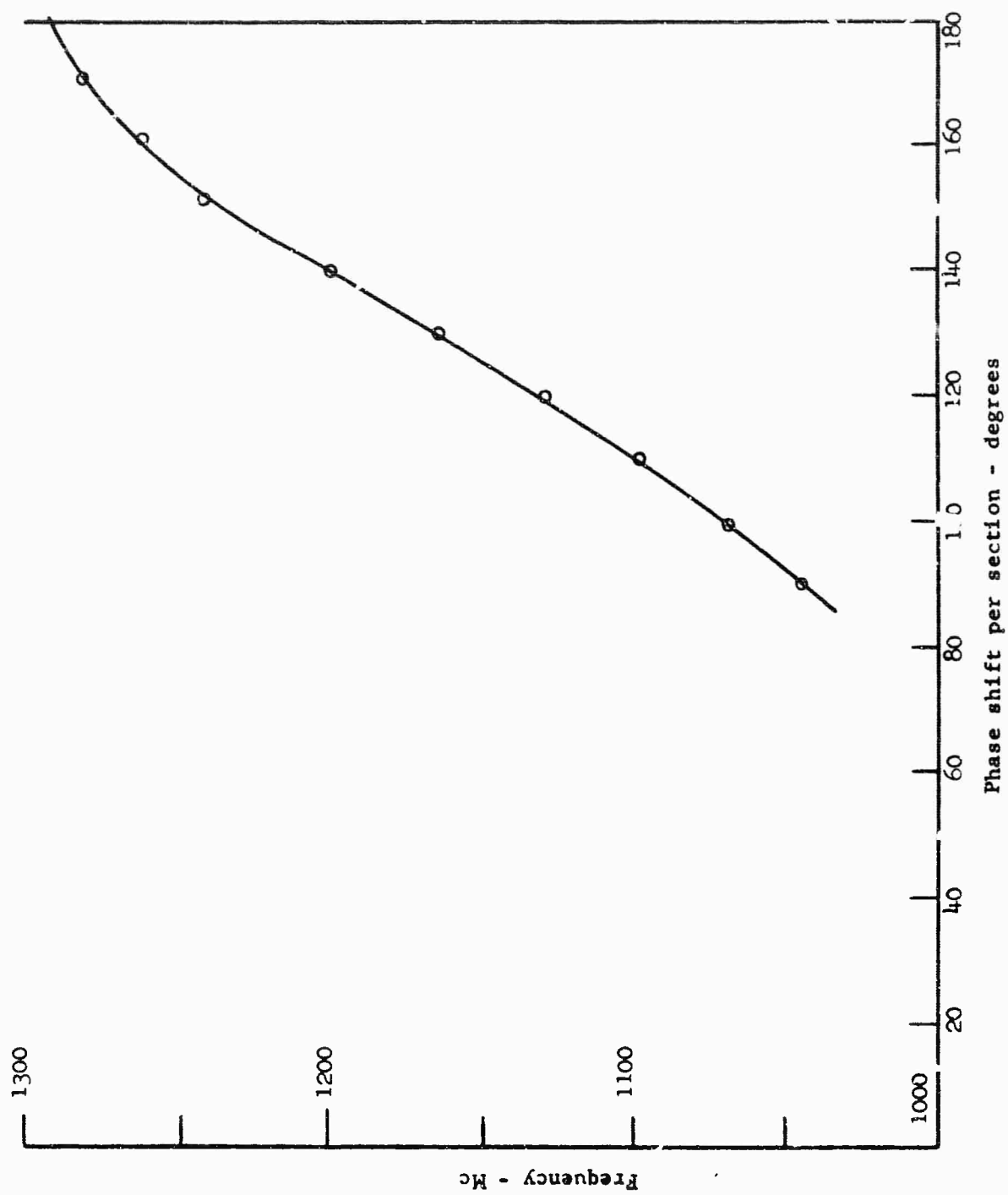


FIGURE 104 THE ω - β CURVE FOR THE SFD-216

S·F·D laboratories, inc.

The measurements obtained on the power tube program indicate that this tube would be very successful once some of the difficulties were surmounted. The resulting tube would have at least 10 db of gain, 750 kw output (minimum) and over 60% efficiency and would operate from a dc power supply with control electrode turn off.

8.0 CONCLUSIONS

The tube developed during this program is a crossed-field amplifier suitable for use in a phased array radar. The amplifier operates from a dc voltage supply through "cold" cathode operation with RF turn on and low level modulation. Operation over the frequency band from 1150 Mc to 1250 Mc has been demonstrated. Peak power output of 100 kw to 150 kw, controllable by an anode voltage of 10 kv or less, and average power output of up to 1500 watts has been demonstrated. The amplifier operates at an efficiency of 50% with a 5 kw input RF signal. Operating life in excess of 5 000 hours has been demonstrated.

An important feature of this tube is operation from a dc power supply with low power (control electrode) modulation, which should prove advantageous in a radar transmitter application. It not only simplifies the modulator equipment required but also increases the overall efficiency by eliminating the high power modulator losses. The low level modulation technique provides versatility not to be found in other devices of this type. Complex modulation such as pulse coding can be accomplished with relatively small amounts of power being expended in the control electrode modulator. RF output pulse widths may be continuously varied from 0.1 μ sec to milliseconds with essentially no change in the power per pulse required from the control electrode modulator.

Cold cathode operation from a dc power supply with RF turn on also offers versatility. With cold cathode operation, warm-up time is instantaneous and no heater reduction schedule is necessary. These features permit instantaneous changes in duty cycle by varying pulse width and/or repetition rate. Pulse widths from 0.1 μ sec to 600 μ sec and repetition rates as low as a pulse per minute have been demonstrated.

The mechanical design of the tube incorporates those features which would allow economical production in large quantities. The construction utilizes relatively simple parts built into sub-assemblies which can be checked for tolerances and tested electrically at various stages. This method was used so that uniformity could be assured before the tube is completed. The raw materials used were those that are readily available, economical and easily handled in accepted machining and brazing procedures. The overall package was designed to be stacked on approximate half-wavelength spacings by simply pushing the tube into its position in the magnetic matrix of the array.

9.0 RECOMMENDATIONS

The SFD-209, in its present state, is well suited for operation in a phased array radar and represents a significant advancement in the "state of the art" for high power amplifiers. However, it is felt that it can be further improved. It is possible to make additional advancements which could not be incorporated within the scope of this program. A very interesting area for future study is the possibility of introducing "self turn off" so that even the low level control modulator can be eliminated. This would result in a high power amplifier which would be turned on by the application of the RF signal and turned off by the removal of that signal. Included in this study would be the optimization of the control electrode geometry to provide further improvement in cold cathode starting.

A second area of study would be to increase the gain beyond the present 16-17 db range. It is felt that the gain can be increased by improvements in the matching techniques and incorporation of absorbers or unidirectional attenuators such as a ferrite isolator in the tube. Some preliminary experiments, which could not be pursued in this program, indicated that either of these methods could be successfully employed to suppress the π mode oscillations which in general limit the maximum usable gain.

A third area of study would be to examine the many secondary emitting materials available for use as cold cathodes to further improve starting performance and to increase the operating life of the amplifier. The present program has demonstrated operating life well in excess of 5000 hours. However, it is expected that 10,000 hours and more is a reasonable goal. The cathode studies should also include optimizing cathode geometry and electron confinement to the interaction area to raise the efficiency of the amplifier beyond its present 60% level.

The fourth area of study that should be included is improvement in the overall package design. The time and scope of this program did not allow optimizing the tube design. It is felt that the tube can be significantly improved. Increasing the anode bore slightly will move the frequency of the coaxial TE_{11} mode outside the operating band and shorten the axial length of the tube. Consequently, the end spaces would be shortened, eliminating the projections from the anode circuit and the extended cathode skirts, both of which presently serve to detune the TE_{11} coaxial mode. This would simplify the tube design and result in a corresponding cost reduction. The resulting shorter axial length would require a shorter, lighter permanent magnet in the array.

APPENDIX I

THE SLOW WAVE CIRCUIT

Principle of Operation

For tubes intended to operate from a dc power supply with control electrode pulsing, a forward wave circuit is required so that the dc operating voltage may remain constant over the frequency band. If a backward wave circuit were used, it would be necessary to program the voltage as frequency was changed, thereby considerably complicating the power supply arrangement. The circuit chosen is one derived from capacitive coupling of half-wavelength bars which are arranged in a circle around the cathode, parallel to the axis of the tube as was shown in Figure 10. The means of coupling the bars to form a propagating circuit are shown in Figure 105. The two rows of capacitive couplers may be visualized as being similar to the strapping used in magnetron circuits but with the straps cut between connection points to the half-wavelength bars. This makes the series capacitance introduced between alternate bars predominate over the series inductance introduced by the straps.

One means of visualizing the operation of the circuit is through an equivalent circuit approach as shown in Figure 106. Here the center of each bar is considered as a node and parallel L-C circuits are drawn between these nodes to represent the resonant spaces between bars. The C_{13} capacitors are then added as lumped elements between the nodes. Next each coupler is split into two halves in series. This permits dividing the resultant network into a series of four pole networks which will be recognized as high pass filter sections. The arrows beside the capacitors in the shunt arms in Figure 106 indicate the direction in which electrons traverse these capacitors. It is seen that, in the final equivalent circuit, the

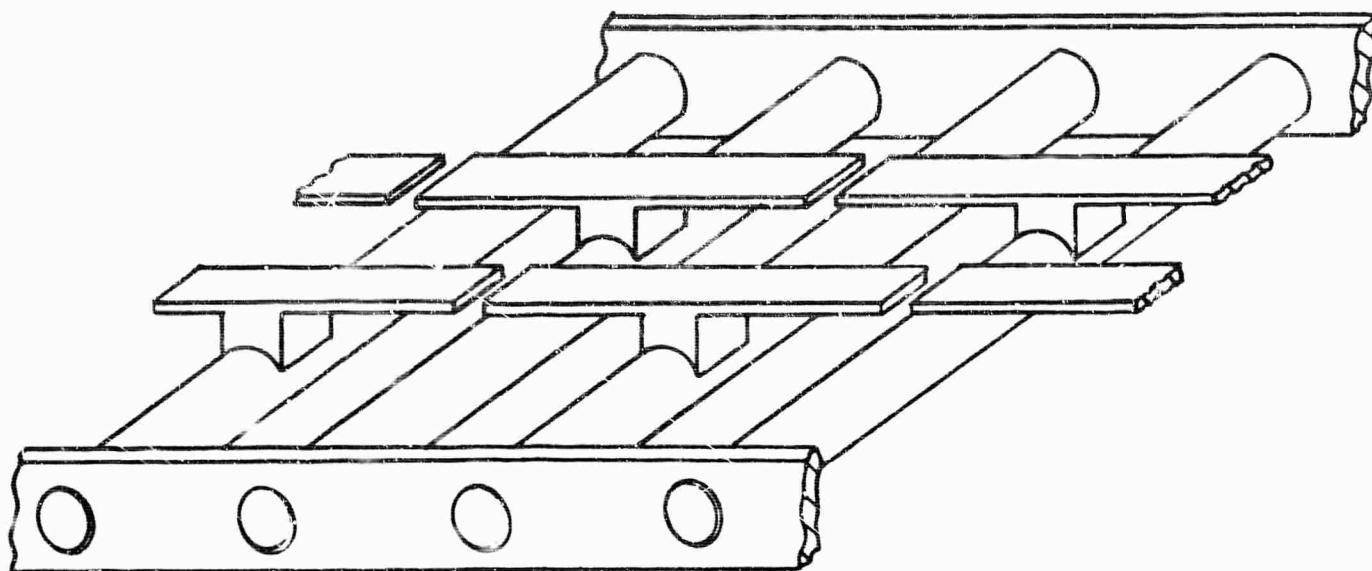
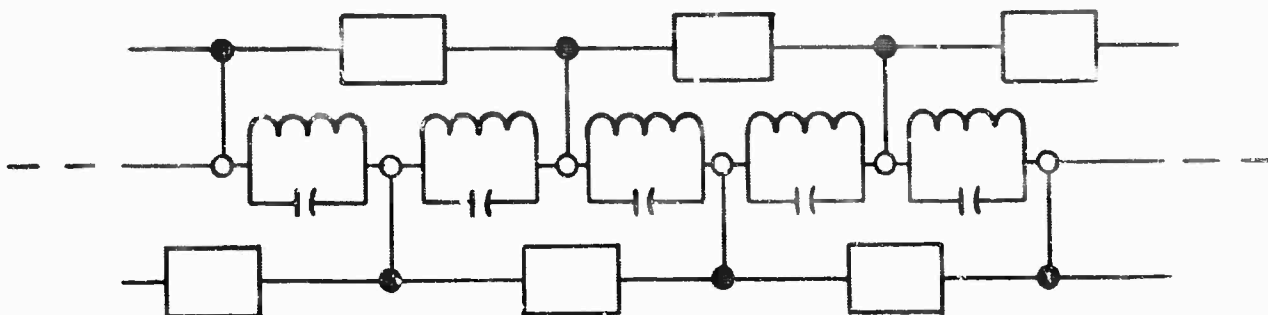


FIGURE 105 SKETCH OF THE TYPE OF SLOW WAVE STRUCTURE
USED IN THE TUBES

The bars are a half-wavelength long at the upper cut-off frequency. Alternate bars are coupled by rows of capacitive couplers. The circuit is forward wave and has its bandwidth between 90 degrees and 180 degrees phase shift per section.



Equivalent circuit of a strapped structure--points a represent the center of bars. The boxes represent the series impedance which may be either capacitive or inductive.

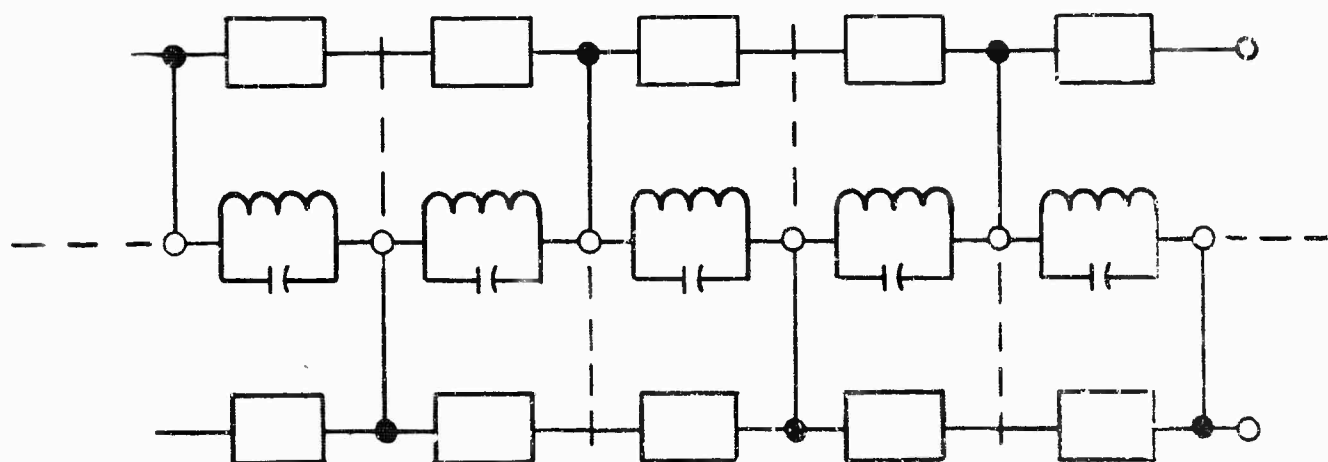
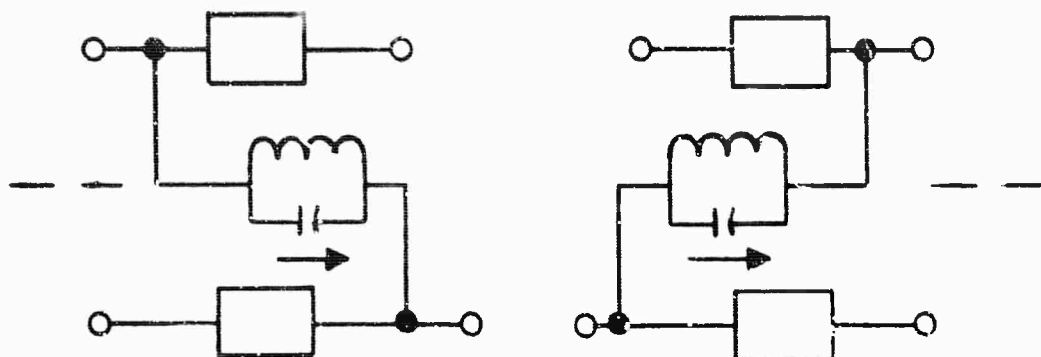
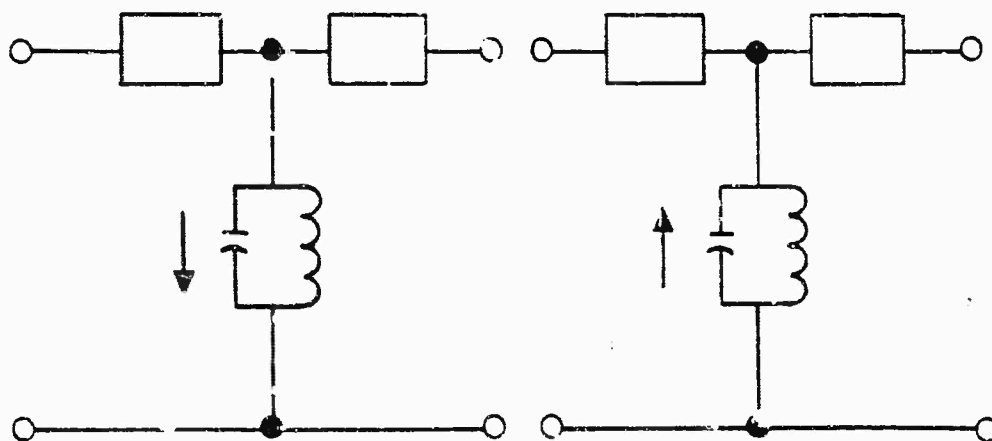


FIGURE 106 DEVELOPMENT OF EQUIVALENT CIRCUIT FOR STRAPPED BAR CIRCUITS

Here the series impedances are split into two impedances. This permits the circuit to be divided along the dotted lines into a cascade of four-poles like this:



The arrows have been added to indicate the direction in which electrons pass through the capacitors of the parallel resonant circuit. Straightening out the circuit and moving the series impedance from the bottom leg to the top since the two locations are electrically equivalent, we have:



This is a cascaded series of four-pole networks with an additional phase shift of π to account for the change in direction with which electrons traverse the center capacitor.

direction is reversed in every other resonator. Therefore, to obtain the ω - β curve, π phase shift must be added to the phase shift obtained from the equivalent lumped four pole networks.

The expression for the ω - β curve in terms of the parameters of the equivalent four pole circuit of Figure 106 is

$$\cos (\pi - \theta) = 1 + \frac{Z_1}{2Z_2} \quad (1)$$

where θ is the phase shift per section

$Z_1 = 1/j\omega C_1$ is the series arm impedance

$Z_2 = j\omega L_2 / [1 - (\omega/\omega_0)^2]$ is the shunt arm impedance

$\omega_0 = 1/\sqrt{L_2 C_2}$ is the π mode radian frequency

The angle $(\pi - \theta)$ is the phase shift of the equivalent circuit.

Because the direction with which electrons traverse the shunt capacitor in adjacent sections reverses, the effective angle for electrons is θ .

Expressing L_2 in terms of ω_0 and C_2 , we have

$$L_2 = \frac{1}{\omega_0^2 C_2} \quad (2)$$

Substituting into eq. (1) we have an expression for θ in terms of ω/ω_0 and C_1/C_2

$$\cos (\pi - \theta) = \frac{\left(\frac{\omega}{\omega_0}\right)^2 \left(\frac{C_1}{C_2} + 1\right) - 1}{\left(\frac{\omega}{\omega_0}\right) \left(\frac{C_1}{C_2}\right)} \quad (3)$$

Curves showing θ as a function of ω/ω_0 are plotted in Figure 107 for various values of C_1/C_2 . We see that increasing the C_1/C_2 ratio

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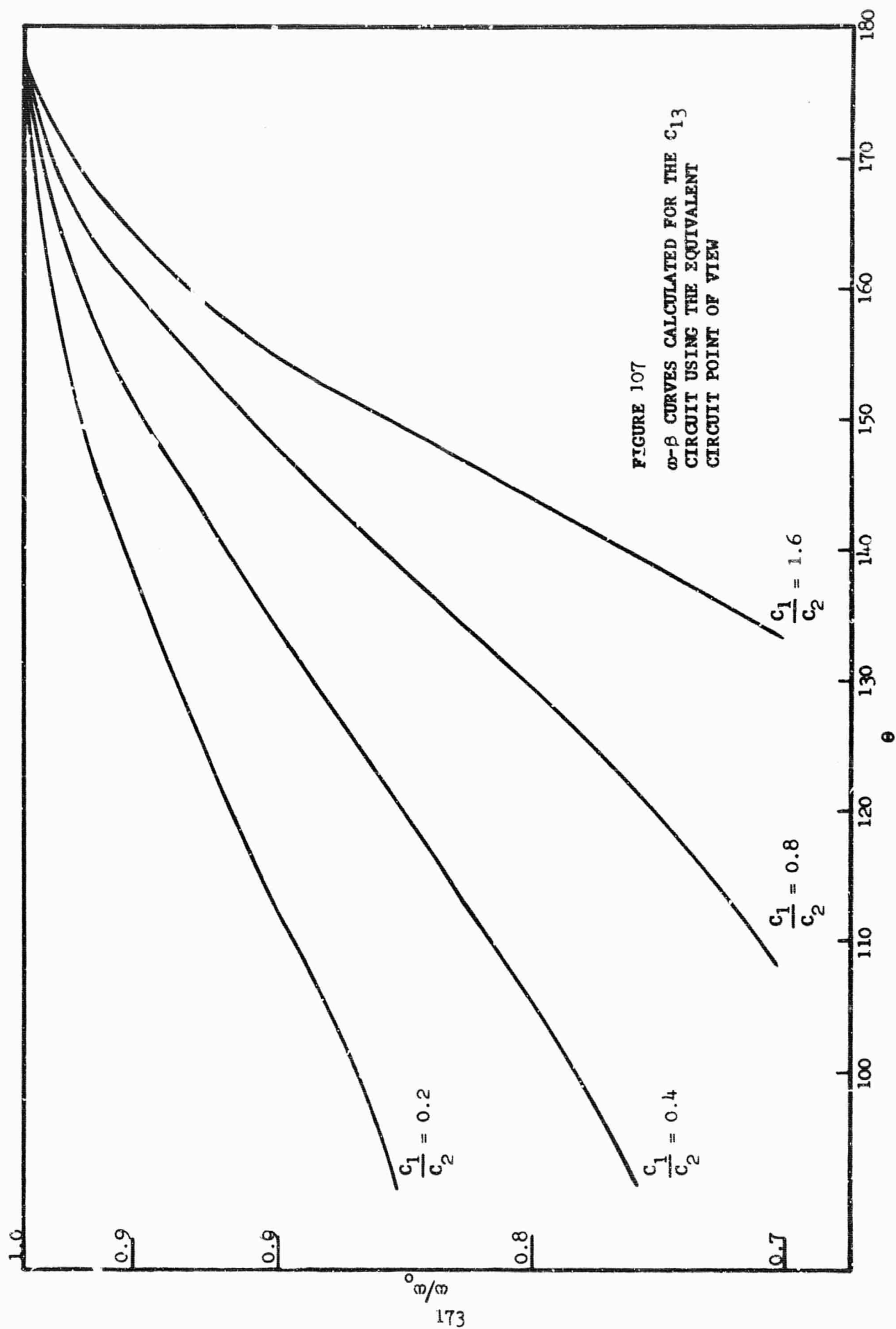


FIGURE 107
 ω - β CURVES CALCULATED FOR THE c_{13}
 CIRCUIT USING THE EQUIVALENT
 CIRCUIT POINT OF VIEW

increases the bandwidth. The π mode frequency is determined by the frequency of the shunt resonator which is in turn determined by the bar length.

An alternate method of analyzing the slow wave circuit is based on a theory by LeBlonde and Mourier^{*} which permits deriving an expression for the ω - β curve in terms of the capacity coefficients between various bars of the circuit. In this theory, the bars of the circuit and the adjacent ground plane are considered as a multi-wire transmission line which propagates a wave axially (i.e., at right angles to the direction of propagation of the slow wave circuit) along the bars at the velocity of light. An expression is then written for the impedance as a function of the capacity and inductance coefficients per unit length of the bars just as one would do for a two wire line. This expression also contains θ , the phase shift per section between bars, which now takes on any arbitrary value whereas for a two wire line it is inherently assumed to be π . Using the assumption that the wave is traveling at the velocity of light, the inductance coefficients may then be eliminated leaving an expression for impedance in terms of capacity coefficients and θ . If each bar is then considered to consist of a capacitive portion made up of the center half of the bar and two inductive portions made up of the two end quarters of the bars, we may impose the condition that the impedance be matched at the junction point between capacitive and inductive sections. This condition leads directly to a relation between the capacity coefficients, θ , and frequency. The resulting expression is

$$\tan^2 \frac{\pi \omega}{4 \omega_1} = \frac{C'_{10} + 4C'_{12} \sin^2 \frac{\theta}{2} + 4C'_{13} \sin^2 \theta}{C_{10} + 4C_{12} \sin^2 \frac{\theta}{2} + 4C_{13} \sin^2 \theta} \quad (4)$$

* A. LeBlonde and G. Mourier, "Study of Bar Lines with Periodic Structure for UHF Electron Tubes," (French) Ann. Radioelect., 9, 311-328 (1954)

- where ω is the radian frequency
- ω_1 is the radian frequency for which the bars are a half free space wavelength long (note that this is not necessarily the π mode frequency)
- C'_{10} is the capacity per unit length in the inductive region between a bar and the ground plane when the remainder of the bars are grounded
- C_{10} is a similar capacity in the capacitive region of the bars
- C'_{12} is the capacity per unit length in the inductive region between a bar and an adjacent bar when the remainder of the bars are grounded
- C_{12} is a similar capacity in the capacitive region
- C'_{13} is a capacity per unit length in the inductive region between a bar and an alternate bar
- C_{13} is a similar capacity in the capacitive region

These capacities are indicated in Figure 108. If desired the expression could be extended to include C_{14} , C_{15} , etc., terms but in our case these are negligible. LeBlonde and Mourier used expressions of this sort containing C_{10} and C_{12} terms to predict the behavior of Karp circuits, interdigital lines and meander lines. However, to our knowledge this is the first time advantage has been taken of the C_{13} term in the design of a slow wave circuit.

The derivation of expression (4) is based on some rather drastic assumptions. For example, it ignores completely the distortion of field patterns which occurs at the junction of inductive and capacitive sections of the bars. We make a further approximation by taking a lumped C_{13} capacity and expressing it as an equivalent distributed capacity. These assumptions, plus the difficulty of accurately evaluating the capacity coefficients, lead one to believe that the above expression cannot be used to obtain accurate predictions of the ω - β curve. However, as will be seen, it provides very useful information concerning the way in which various geometrical changes affect the ω - β curve.

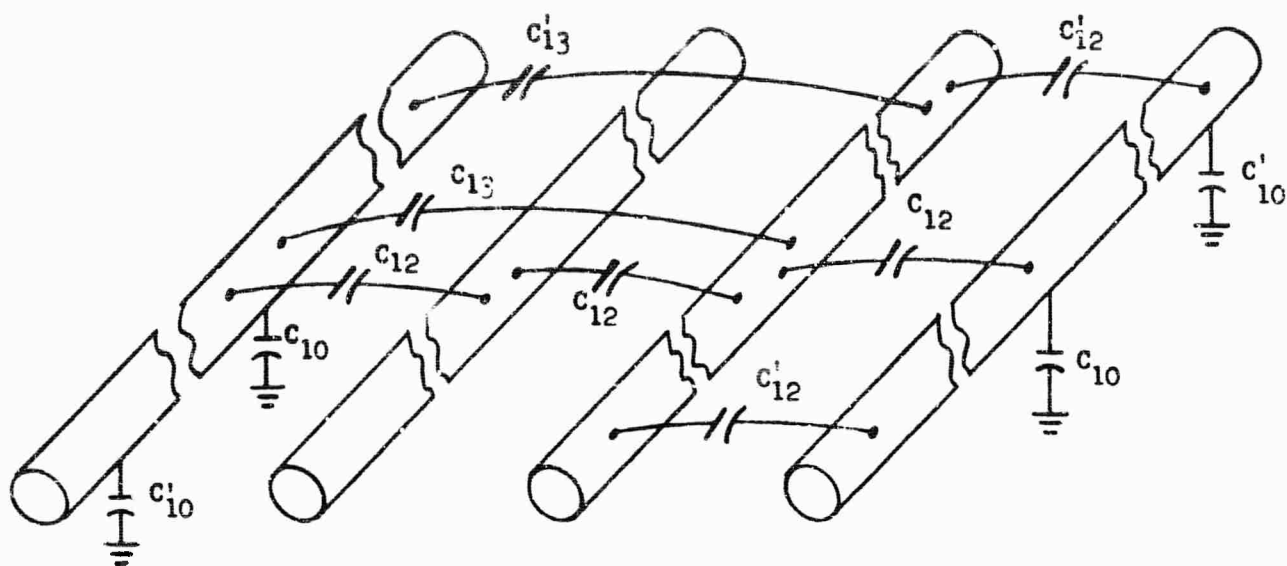


FIGURE 108 SKETCH SHOWING THE MEANING OF THE CAPACITY COEFFICIENTS IN THE Mourier-LeBlonde FORMULA.

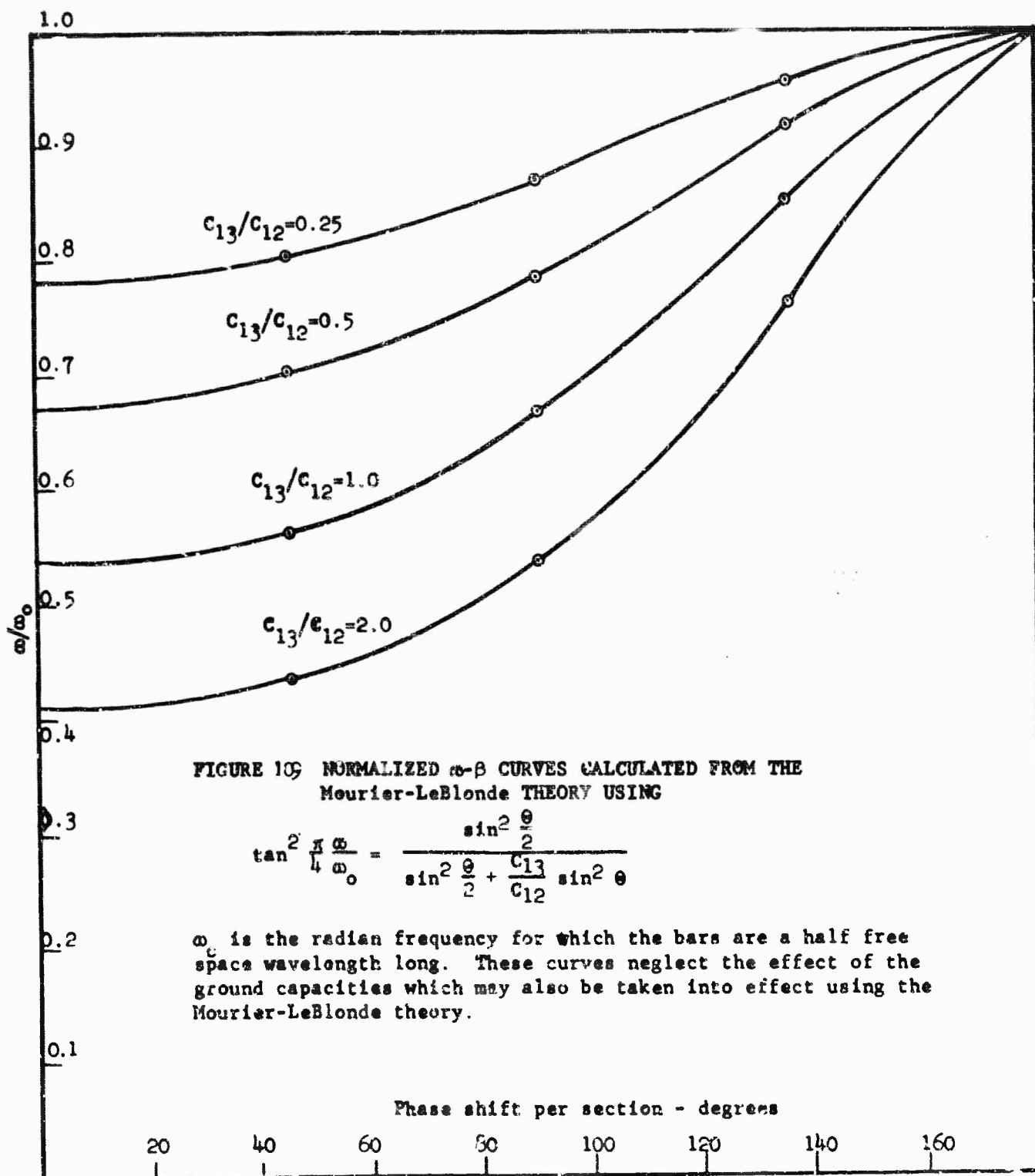
All capacities in the original analysis are per unit length. In deriving the relation for the C_{13} circuit the additional assumption of lumping the capacities is made. The center half of the bars is referred to as their capacitive region and the two end quarters as their inductive region.

We shall consider first the case for which $C_{10}' = C_{13}' = C_{10} = 0$ and $C_{12}' = C_{12}$. Expression (4) then reduces to

$$\tan^2 \frac{\pi}{4} \frac{\omega}{\omega_1} = \frac{C_{12} \sin^2 \frac{\theta}{2}}{C_{12} \sin^2 \frac{\theta}{2} + C_{13} \sin^2 \theta} \quad (5)$$

It is seen that when $\theta = \pi$, the righthand side of the expression reduces to unity and, therefore, $\omega = \omega_1$. Thus the π mode frequency becomes that for which the bars are a half-wavelength long. When $\theta = \pi/2$, the term involving C_{13} in the denominator becomes a maximum. This causes a reduction in the frequency for $\pi/2$ phase shift per section below the π mode frequency thereby giving rise to a forward wave circuit having its bandwidth between 90° and 180° phase shift per section. Some typical ω - β curves calculated for different C_{13}/C_{12} ratios are given in Figure 109. Again, it is seen that the bandwidth is determined by the amount of capacitive coupling while the π mode cut off is determined by the bar length - a result in agreement with the conclusion drawn from the equivalent circuit analysis. The analysis based on the capacity coefficients is the one originally used to invent the circuit. For this reason, we have given this type of slow wave circuit the name of " C_{13} circuit." Neither type of analysis has been used to calculate actual circuit properties. These have been obtained by an empirical procedure because of the difficulty of relating geometrical configurations to the parameters in the analyses and because of the approximate nature of the analyses. Each type of analysis has, however, added to our understanding of the circuit's operation.

In practice the π mode frequency of the circuits is not exactly the frequency for which the bars are a half-wavelength long. This effect can be best understood by reference to expression (4). If we set $\theta = \pi$, we obtain the following expression for the π mode frequency.



$$\tan^2 \frac{\pi}{4} \frac{\omega}{\omega_1} = \frac{C'_{10} + 4C'_{12}}{C_{10} + 4C_{12}} \quad (6)$$

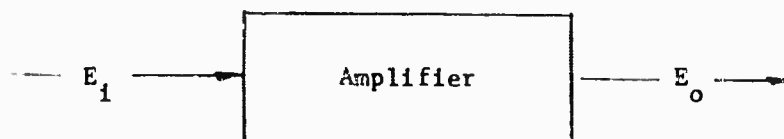
In practice, the C_{10} and C'_{10} terms represent capacities from the bars to the can and to the cathode. These are far from negligible in a practical tube, especially one in which the can closely surrounds the bars to minimize tube diameter. The C_{12} and C'_{12} terms are not identical either since the C_{13} couplers, in passing over the adjacent bars, also introduce some C_{12} capacity. The net consequence of these factors is that the π mode frequency is somewhat less than the frequency for which the bars are a half-wavelength long. The proper π mode frequency must be arrived at by a trial and error procedure. A first guess is taken, a circuit is built and the π mode frequency is measured. The bar length is then corrected in subsequent circuits until the desired frequency is obtained.

APPENDIX II

EFFECT OF CYCLICALLY VARYING ANTENNA VSWR

An antenna VSWR of 2:1 will affect tube performance through the effect it will have on phase linearity. It is not expected that this VSWR will in any way harm the tube or have any deleterious effect on life. A 2:1 VSWR means that about 10% of the output power will be reflected back to the tube. For a 150 kw peak output power, this would amount to a peak reflected power of 15 kw, which will be attenuated to about 10 kw in passing back through the tube to the input line. Provision must be made for dissipating this 10 kw peak power in an isolator or circulator or other component in the input line. The fact that the antenna VSWR is cycling at a four-second rate will have no particular consequence. The input isolator should be designed for the 10 kw peak return power which it will experience with the antenna VSWR at maximum value.

The effect of antenna VSWR on phase linearity may be determined from the following analysis:



Let $E_1 = A = Ae^{j\theta}$ be applied to the tube.

Let $\Delta\theta$ be the total linear phase variation through the tube for a specified Δf (the bandwidth required). Then

$$E_o = K_1 A e^{j\Delta\theta}$$

where K_1 is the voltage gain of the tube

Due to the mismatch at the output end, $K_2 E_o$ is reflected back to the input and, by the time it reaches the input, the reflected signal is

$$K_2 K_3 E_o e^{j\Delta\theta} = K_1 K_2 K_3 A e^{j2\Delta\theta}$$

where K_2 represents the fraction of amplified wave reflected back to input

K_3 represents the fraction of reflected wave not absorbed in circuit loss

If K_4 represents the fraction of reflected wave which is reflected back to the output, then the total input signal is

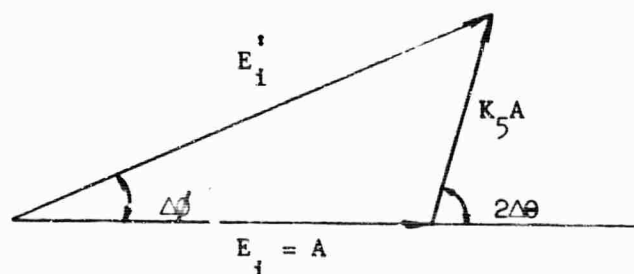
$$E_i' = A e^{j\theta} + K_5 A e^{j2\Delta\theta} \quad (\text{vector addition})$$

where $K_5 = K_1 K_2 K_3 K_4$

We can see that the effect of the two mismatches is to change the effective input signal from $A e^{j\theta}$ to $A e^{j\theta} + K_5 A e^{j2\Delta\theta}$

Graphically, we have the following situation at the input.

$\Delta\phi$ gives an indication of the degree of non-linearity.



The peak phase ripple occurs when $2\Delta\theta = 90^\circ$ and the frequency deviation (Δf) over which this occurs is one-quarter cycle of the phase ripple periodicity. Also, $2\Delta\theta$ is given by the expression

$$2\Delta\theta = 2 \frac{d\theta}{d\omega} N \Delta f$$

where $d\theta/d\omega$ = the reciprocal of the slope of the dispersion curve (deg/res)/Mc

N = number of resonators

Δf = frequency deviation

If we solve for Δf when $2\Delta\theta = 360^\circ$, we get one cycle of the phase deviation frequency. For this circuit

$$\frac{d\theta}{d\omega} = \frac{0.4^\circ/\text{res}}{\text{Mc}} \quad \text{and } \Delta f = 12.5 \text{ Mc for } N = 36$$

Thus the phase deviation periodicity will be eight cycles over the 100 Mc band for phase deviations caused by input and output mismatches.

The peak phase ripple is given by the expression

$$\Delta\phi = \tan^{-1} K_5$$

For the SFD-209 tube, the following numbers apply:

$$K_1 = 6.3 \text{ (16 db gain)}$$

$$K_2 = 0.14 \text{ (corresponds to 17 db return loss)}$$

$$K_3 = 0.8$$

$$K_4 = 0.14 \text{ (corresponds to 17 db return loss)}$$

$$K_5 = K_1 K_2 K_3 K_4 \approx 0.10$$

$$\Delta\phi = 5.7^\circ$$

This suggests that if the tube has a gain of 16 db, there might be a 5.7° ripple on the linear ω - β curve, assuming externally matched input and output lines.

Figure 110 shows the expected plot of phase versus frequency for the SFD-209 tube with 16 db gain.

The derivation so far is for externally matched lines. If there is a mismatch at the antenna, a second reflected component will be added to the component K_2 in the above derivation. This will introduce a second periodic ripple into the phase versus frequency characteristic. The periodicity of this component will be greater (i.e., more cycles over the bandwidth) than that due to the reflection from the tube output because the length of line between tube output and antenna will contribute an additional amount to the phase shift θ in the above derivation. These two components of phase shift will then add algebraically to give the net phase versus frequency characteristic. When the antenna VSWR reaches 2:1, the reflections from the antenna will predominate over those from the tube output. In this case, the peak amplitude of the phase deviation from linearity ($\Delta\phi$) will increase to about 13° and the periodicity will be increased by an amount depending on the distance between tube and antenna.

The effect of a cyclically varying VSWR will thus be deviations of phase from linearity with the deviations varying in amplitude and periodicity as the VSWR cycles. The minimum amplitude of the deviation will be about 5.7° and the maximum amplitude (for a 2:1 VSWR) will be about 13° .

If the antenna VSWR becomes significantly worse than 2:1, eventually the point will be reached at which the signal reflected from the antenna and again from the tube input (or from a point in the input line) will be large enough to produce significant variations in the effective input signal amplitude. For those frequencies at which the effective input amplitude is reduced, the amplitude may be too small to effectively lock the space charge. This could lead to spurious outputs from the tube. The antenna VSWR for which this occurs will depend on the input match. This problem should not appear at antenna VSWR of up to 2:1.

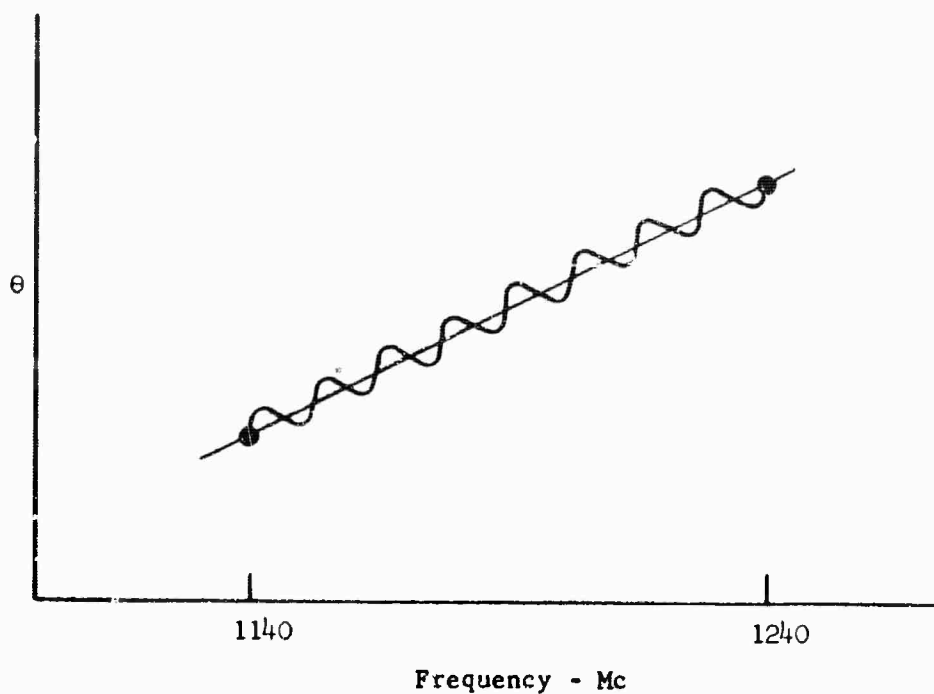


FIGURE 110 THE PHASE VERSUS FREQUENCY CHARACTERISTIC EXPECTED

The linear slope is taken from cold test measurements.
The amplitude of the sinusoidal ripples will depend
on tube and line mismatches.

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13. ABSTRACT A crossed-field amplifier for use in an L-band phased array radar has been developed. The SFD-209 is designed to be stacked on half-wavelength centers in a permanent magnet array. The tube will generate peak power up to 150 kw with 16 db gain at an efficiency of 50% over a 100 Mc band centered on 1200 Mc, with exceptional phase characteristics. Average power of 1.5 kilowatts and operating life in excess of 5000 hours have been demonstrated. The SFD-209 operates from a dc power supply with very low modulator power. Cold cathode operation with RF turn on and control electrode turn off is used. The duty cycle may be changed instantaneously by changing repetition rate and/or pulse width with no change in modulator power on a per pulse basis. Pulse widths from 0.1 μ sec to 600 μ sec and repetition rates as low as 1 pulse per minute have been demonstrated. The ability to pulse code a CW input signal and the ability to amplify a complex signal have been demonstrated.			

KEY WORDS

Crossed-field

Microwave

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