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National Laboratory

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## RF System Preliminary Design

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Collider Accelerator Department  
**Brookhaven National Laboratory**

**U.S. Department of Energy**

USDOE Office of Science (SC)

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RHIC-6

RHIC Technical Note No. 6

RF SYSTEM PRELIMINARY DESIGN

M. Puglisi

September 14, 1984

RF System ParametersRF Frequency  $\nu_0$  = 26.743 MHzHarmonic Number  $h$  = 342RF total Voltage  $V_t$  = 1.2 MVRF phase angle  $\phi$  = ~ 2.3°Number of bunches  $n_b$  = 57RF cycles per bunch  $\xi$  = 6Average stored current  $I_a$  = 65 mABeam revolution frequency  $\nu$  = 78.195906 KHz.

For a preliminary design of the RF system we should estimate the number of cavities required per ring and the maximum value for the stored current.

A reasonable guess for the cavity number is  $\sim 6$ ; consequently  $V_c = 200$  kV per cavity is required. Later the case of 12 cavities will be examined.

Following the criterion that the beam-induced voltage should be always much lower than the amplifier-induced voltage we could set an upper limit to the stored current.

An average output impedance of about  $100 \Omega$  can be obtained with a Class A feedback linear amplifier. With a Class C operation the average output impedance can easily reach more than  $\sim 400 \Omega$ . Because the typical dc plate voltage of a power amplifier is in the park of  $\sim 10$  kV then it follows that the transforming ratio should be near 20 and the transferred output impedance becomes  $>160$  k $\Omega$  per cavity.

For generating 200 kV on  $160$  k $\Omega$  the amplitude of the synchronous beam harmonic should be equal to 1.25 amp. that in turn demands an average current near to 0.625 A.

With a somewhat narrow safety margin we could assume for the average current  $I_m$  a value of 0.5 amp which would accelerate  $\sim 7 \cdot 10^{11}$  particles per bunch.

A value for  $n$  equal to  $5 \cdot 10^{11}$  particles per bunch seem then very reasonable and we obtain the following limits:

$n = 5 \cdot 10^{11}$	Particles per bunch	
$q = 8 \cdot 10^{-8}$	Charge per bunch	(Coulomb)
$Q = 4.56 \cdot 10^{-6}$	Total charge	(Coulomb)
$I_m = 0.356$	Average current	(Amp)
$I_1 = 0.713$	Synchronous harmonic	(Amp)
$RF = 2.80 \cdot 10^5$	Beam eq. Impedance	(Ohm)
$I_p \approx 24.$	Maximum beam peak current	*(Amp)

---

\* The beam is assumed to have parabolic distribution as follows:

### 3 - Parameters

$$I_b(t) = I_p - bt^2 \quad -\tau/2 \leq t \leq \tau/2$$

Where  $\tau$  is the beam time duration. Setting the boundary conditions we obtain:

1.5 m

$$\int_{-\tau/2}^{+\tau/2} (I_p - bt^2) dt = I_p \tau - b \frac{\tau^3}{12} = q$$

$$I_p - b \left(\frac{\tau}{2}\right)^2 = 0$$

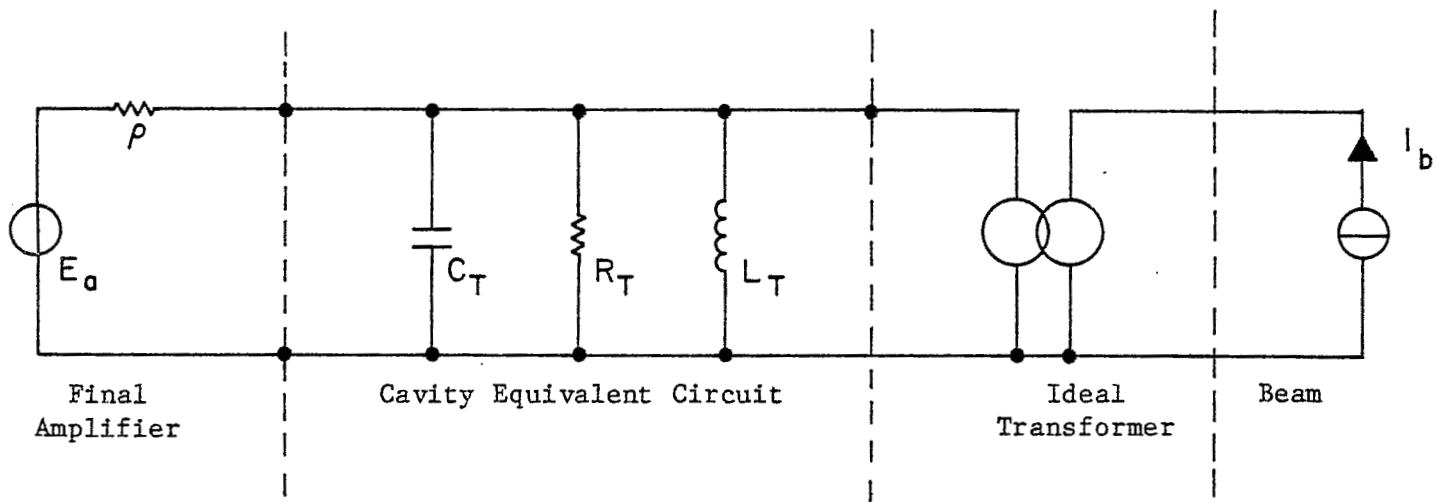
$$\text{Solving for } I_p \text{ we obtain: } I_p = \frac{3q}{2\tau}; \quad b = \frac{4I_p}{\tau^2}$$

$$\text{For } q = 8 \cdot 10^{-8}; \quad \tau = 0.5 \cdot 10^{-8} \quad I_p = 24 \text{ Amps.}$$

$$b = 16 \cdot 10^{16} \text{ A/sec}^2$$

## The Beam Loading

The following scheme could be used for describing the phenomenon as it is seen by the final amplifier.



If  $V$  is the voltage across the cavity eq. circuit (that is  $V = V_{\text{cavity}}/N$ ) then we can write:

$$\left\{ G + j \left( \omega C_t + \frac{1}{\omega L_t} \right) \right\} V = \frac{E_a}{\rho} e^{j\alpha} + N I l e^{j\beta} \quad (1)$$

$$G = \frac{1}{\rho} + \frac{1}{R_t}; \quad N = \text{Transforming ratio}$$

Where:  $\alpha$  and  $\beta$  are phase angles and  $I l$  is the beam current component that is synchronous with the RF voltage.

Equation (1) can be divided into two equations:

$$GV = \frac{E_a}{\rho} \cos \alpha + N I l \cos \beta$$

$$\left( \omega C_t - \frac{1}{\omega L_t} \right) V = \frac{E_a}{\rho} \sin \alpha + N I l \sin \beta$$

Now we want:  $\alpha = 0$  (that means that the final amplifier sees a real load) and  $\beta = \pm 90^\circ$  because the stable phase is assumed nearly equal to zero. Consequently the system reduces to:

$$GV = \frac{Ea}{\rho} \\ (\omega C_t - \frac{1}{\omega L_t}) V = \pm N I_l$$

Solving we obtain:

$$Ea = \left( 1 + \frac{\rho}{R} \right) V \\ C_t = \frac{1}{\omega^2 L_t} \pm \frac{N I_l}{\omega V} = C_0 \pm \Delta C$$

where, obviously,  $V$  is the voltage that appears at the terminals of the cavity equivalent circuit as is seen by the amplifier.

#### Numerical Example

The real cavity equivalent parameters are:

$$R = 0.8 \text{ M}\Omega, C = 50 \cdot 10^{-12} \text{ F}, L = 6.949 \cdot 10^{-7} \text{ H}, V_c = 200 \text{ kV.}$$

Upon transformation ( $N = 20$ ) we obtain:

$$R_t = 2000 \Omega; C_t = 2 \cdot 10^{-8} \text{ F}; L_t = 1.737 \cdot 10^{-9} \text{ H} \\ V = 200 \text{ kV}/20 = 10 \text{ kV}; I_{l_t} = (2 \cdot 0.356) \cdot 20 = 14.24 \text{ A}$$

(Where the subscript  $t$  indicates "Transformed Parameters"). Assuming that the

amplifier output impedance is equal to  $100\Omega$  we obtain:

$$E_a = \left(1 + \frac{100}{2000}\right) 10 \cdot 10^3 = 10.500 \cdot 10^3$$

$$\Delta C = \frac{14.24}{1.696 \cdot 10^3} = \pm 8.39 \cdot 10^{-12} \text{ F.}$$

It is easy, now, to calculate the total current  $I_t$  that the tube should supply:

If  $E_a = \left(1 + \frac{\rho}{R_t}\right)V$  then:

$$I_t = \frac{E_a - V}{\rho} = \frac{1}{\rho} \frac{\rho}{R_t} V = \frac{V}{R_t} = \frac{10 \cdot 10^3}{2 \cdot 10^3} = 5 \text{ A.}$$

where  $I_t$  indicates the total current that should be provided by the tube. (We note that because  $\beta = \pi/2$  then  $I_t$  do not depend upon  $I_1$ ).

## The Cavity

In Fig. 1 is represented an axial section of a cavity that could work with a gap voltage of 200. kV.

From Super-Fish calculation we have  $\nu_0 = 26.75$  MHz;  $Q_0 = 9030$ ;  $R_s = 0.916$  M $\Omega$ . and it follows that at 200. kV the power needed should be near to 22 kw.

The cavity eq. scheme is shown in Fig. 1a.

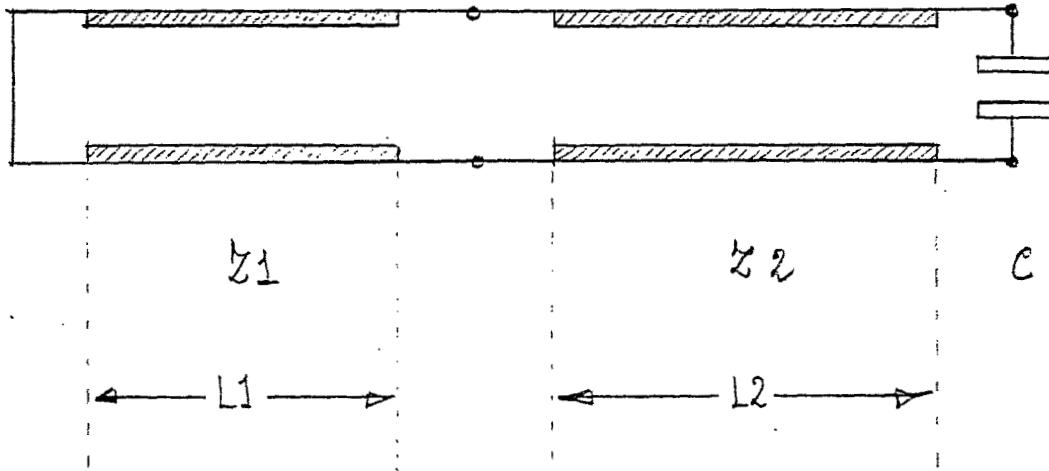


Fig. 1a. The cavity is schematized with a foreshortened coaxial line that contains a discontinuity in the characteristic impedance.

From the resonant frequency and the geometry of the coaxial line the value of the capacity C can be calculated as will be shown later on.

The cavity shape allows to have the driving tube directly connected to the low impedance side of the cavity\* and we assume that the cavity with the tube can be schematized as shown in Fig. 2.

\*The short stub can be in air so the tube can be easily connected to the cavity and adjusted for the wanted transforming ratio. Obviously the values for x and y are to be determined taking into account the total tube output capacity  $C_T$ .

2 - Cavity

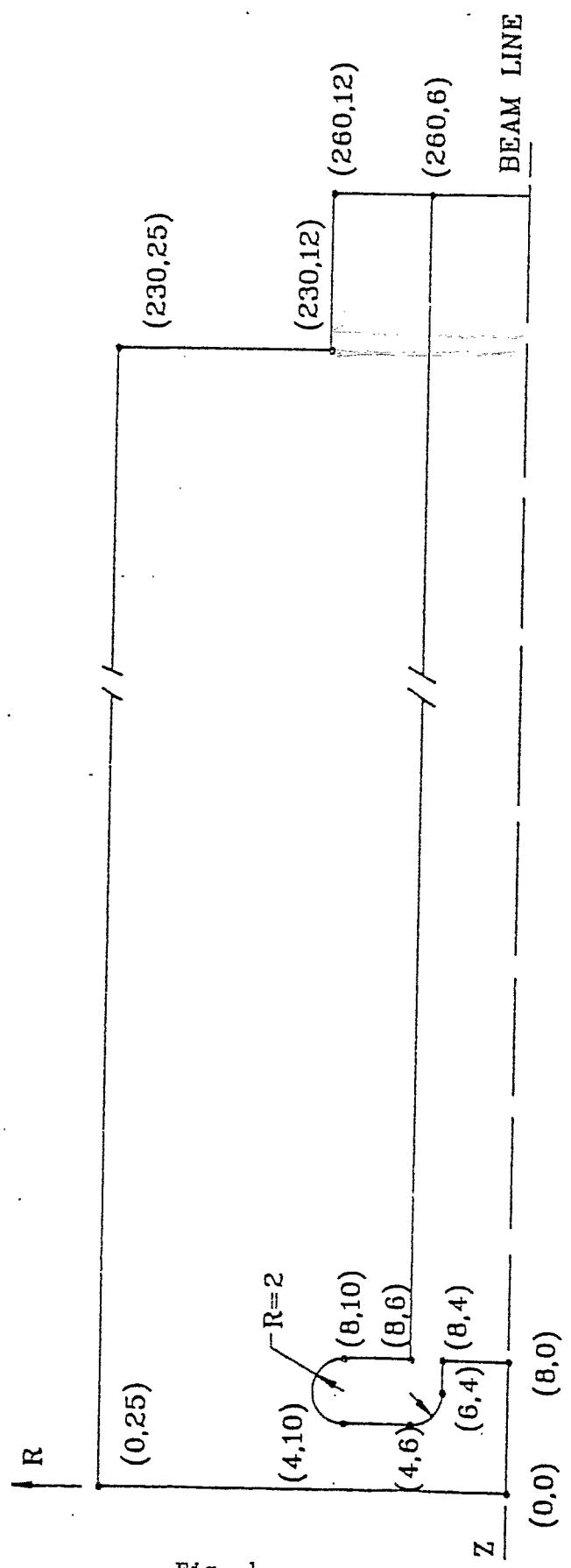


Fig. 1

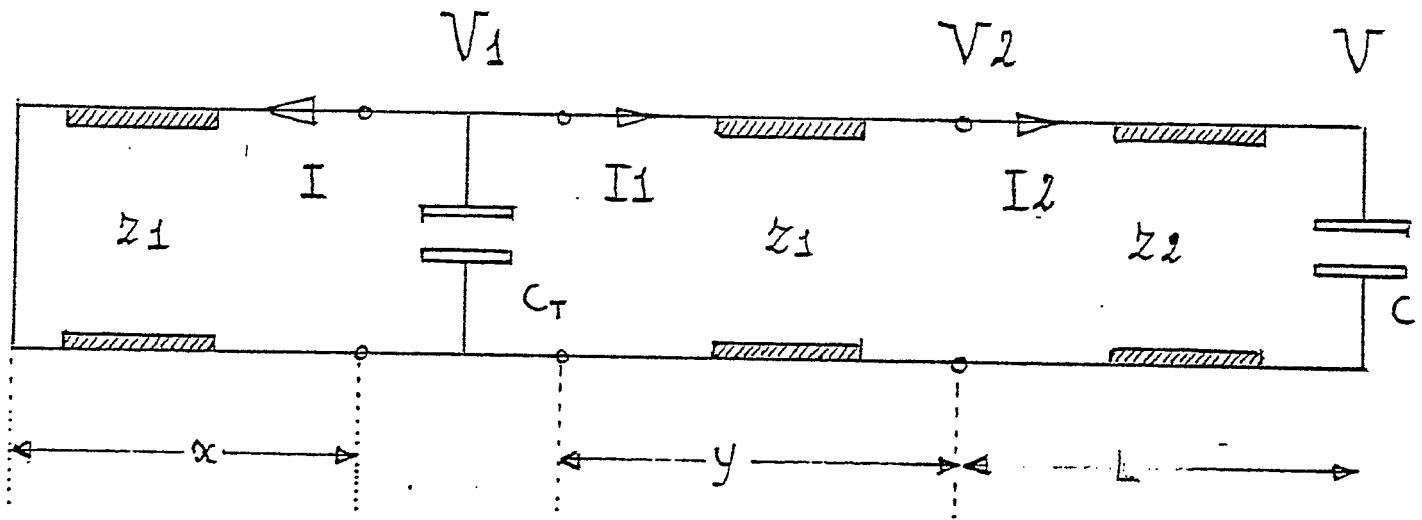


Fig. 2.  $C$  and  $C_T$  are the gap and the tube eq. capacities.  
 $Z_2$  and  $L$  are the characteristic impedance and the electrical length of that portion of the cavity that is under vacuum.  
 $Z_1$  and  $L_1 = x + y$  define the portion of the cavity that is in air.

From the lines theory we have:

$$V_2 = V (\cos \beta L - \omega c Z_2 \sin \beta L) = AV$$

$$I_2 = jV (\omega c \cos \beta L + \frac{1}{Z_2} \sin \beta L) = jBV$$

where  $\beta = \frac{2\pi}{\lambda} = \omega/\epsilon_0 \mu_0$  is the phase constant.

Transferring the voltage and the current at the junction with the tube we obtain:

$$V_1 = V(A \cos \beta y - B Z_1 \sin \beta y)$$

$$I_1 = jV(B \cos \beta y + \frac{A}{Z_1} \sin \beta y)$$

#### 4 - Cavity

From the first eq. because  $N = V/V_1$  is already known then we determine the value of  $y$ .

At this point all the quantitites are known and we should find the value of  $x$  for tuning the whole system.

For this purpose we set equal to zero the total admittance  $Y$  at the junction as follows:

If  $\epsilon = \frac{A}{BZ_1}$  then we can write the R.H.S. component of the total admittance as:

$$Y(R) = I_1/V_1 = \frac{j}{Z_1} \frac{1 + \epsilon \tan \beta_y}{\epsilon - \tan \beta_y}$$

The L.H.S. of the total admittance is:

$$Y(L) = \omega C_T - \frac{1}{Z_1 \tan \beta_x}$$

from the tuning condition:

$$Y = Y(R) + Y(L) = 0$$

we obtain:

$$\cot \beta_x = \omega C_T \beta_x = \frac{1 + \epsilon \tan \beta_y}{\epsilon - \tan \beta_y}$$

and the value of  $x$  is determined.

With the dimension indicated in Fig. 1 we find that the resonant frequency is equal to 26.75 MHz. (Super-Fish)

The output impedance of the coaxial structure is as follows:

$$Z = jZ_2 \frac{\frac{Z_1}{Z_2} \tan(\beta L_1) + \tan(\beta L_2)}{1 - \frac{Z_1}{Z_2} \tan(\beta L_1) * \tan(\beta L_2)}$$

Because:  $Z_1 = 41.58 \Omega$ ;  $Z_2 = 85.62$ ; and  $\beta = 0.5604$  for  $\nu_0 = 26.75$  MHz  
 then  $Z \approx j423 \Omega$ . Consequently the total gap capacitance ( $C_T$ ) that tunes the cavity should be:  $C_T = 1/\omega_0 Z \approx 14$  PF.

It follows that:

$$A \approx 0.084; B = 0.0118$$

and the eq.  $A \cos \beta y - B Z_1 \sin \beta y = 0.05$  (solved numerically) gives  $y \approx 0.12$  m.  
 Now  $\epsilon = 0.171$  and  $C_T = \sim 70$  PF.

$$\text{Consequently: } \frac{1}{\tan(\beta x)} = 0.489 + \frac{1 + 0.0118}{0.171 - 0.069} = 10.40$$

Solving we obtain:  $x = \sim 0.17$  m.

#### The Simplified Scheme

The above model is too complicated for a first design of the final amplifier and we consider the circuit given in Fig. 3.

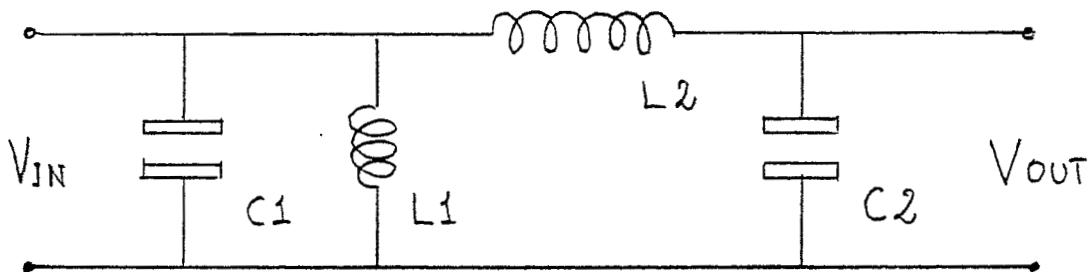


Fig. 3. Eq. circuit for the cavity loaded by the driving amplifier.  $C_1$  and  $C_2$  are given because they both depend upon the tube and the cavity.

## 6 - Cavity

Now the transforming ratio (no load) is as follows:

$$N = \frac{V_{out}}{V_{in}} = \frac{1}{1 - \omega^2 L_2 C_2}$$

and we obtain:

$$L_2 = \frac{1 - 1/N}{\omega^2 C_2}$$

The circuit should be tuned by the inductor L1. The input admittance can be written as follows:

$$Y(i) = \frac{1}{j\omega L_1} + j\omega \frac{C_1 (1 - \omega^2 L_2 C_2) + C_2}{1 - \omega^2 L_2 C_2}$$

setting  $Y(i) = 0$  (that means "tuning"), solving for L1 and recalling the condition:  $1 - \omega^2 L_2 C_2 = \frac{1}{N}$  we obtain:

$$L_1 = \frac{1}{\omega^2 C_2} \frac{1/N}{1 + \frac{C_1}{C_2} \frac{1}{N}}$$

The calculations already shown are correct only for "no load conditions". It can be demonstrated that for quality factors of the cavity as low as  $\sim 100$  the above formula are still valid for design purposes.

Simple checks made with the ECAP program shows both the validity of the previous calculation and the fact that the tube ignores the beam current if the beam is in quadrature with the gap voltage and an appropriate compensating capacity is added. The scheme prepared for ECAP calculations is shown in Fig. 4.

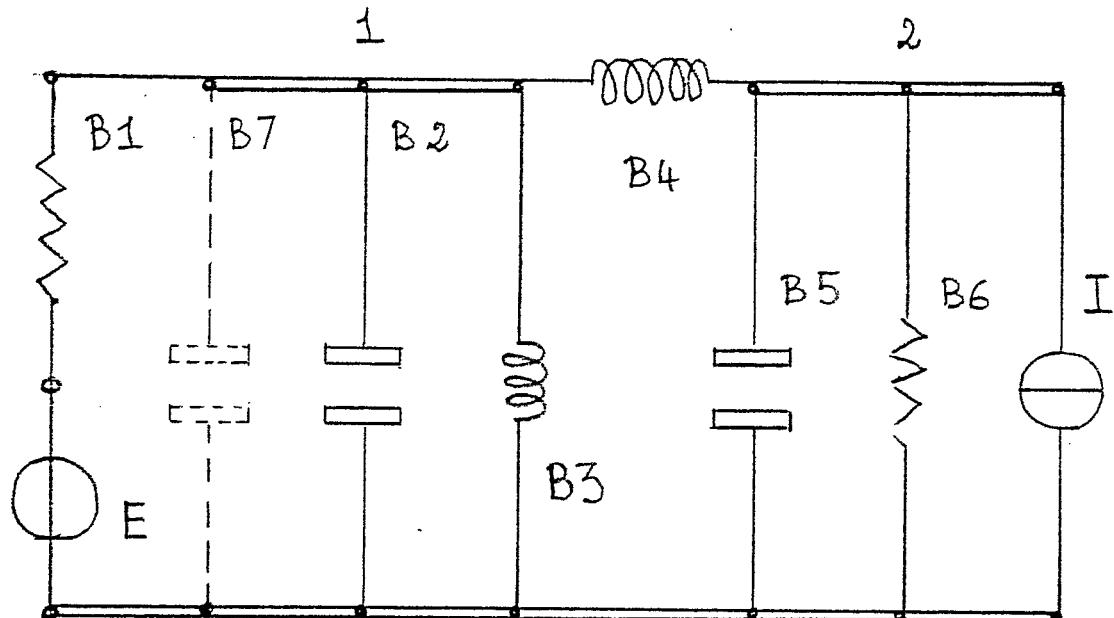


Fig. 4

ECAP. AC ANALYSIS

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PAGE NO. 1

```

1.      AC ANALYSIS
2.  B1    N(0,1),R=100,E=11.912E3
3.  B2    N(1,0),C=70,E-12
4.  B3    N(1,0),L=3.2473E-8
5.  B4    N(1,2),L=6.6018E-7
6.  B5    N(2,0),C=50,E-12
7.  B6    N(2,0),R=0.7BE6
8.          FREQUENCY=27E6
9.          EXECUTE

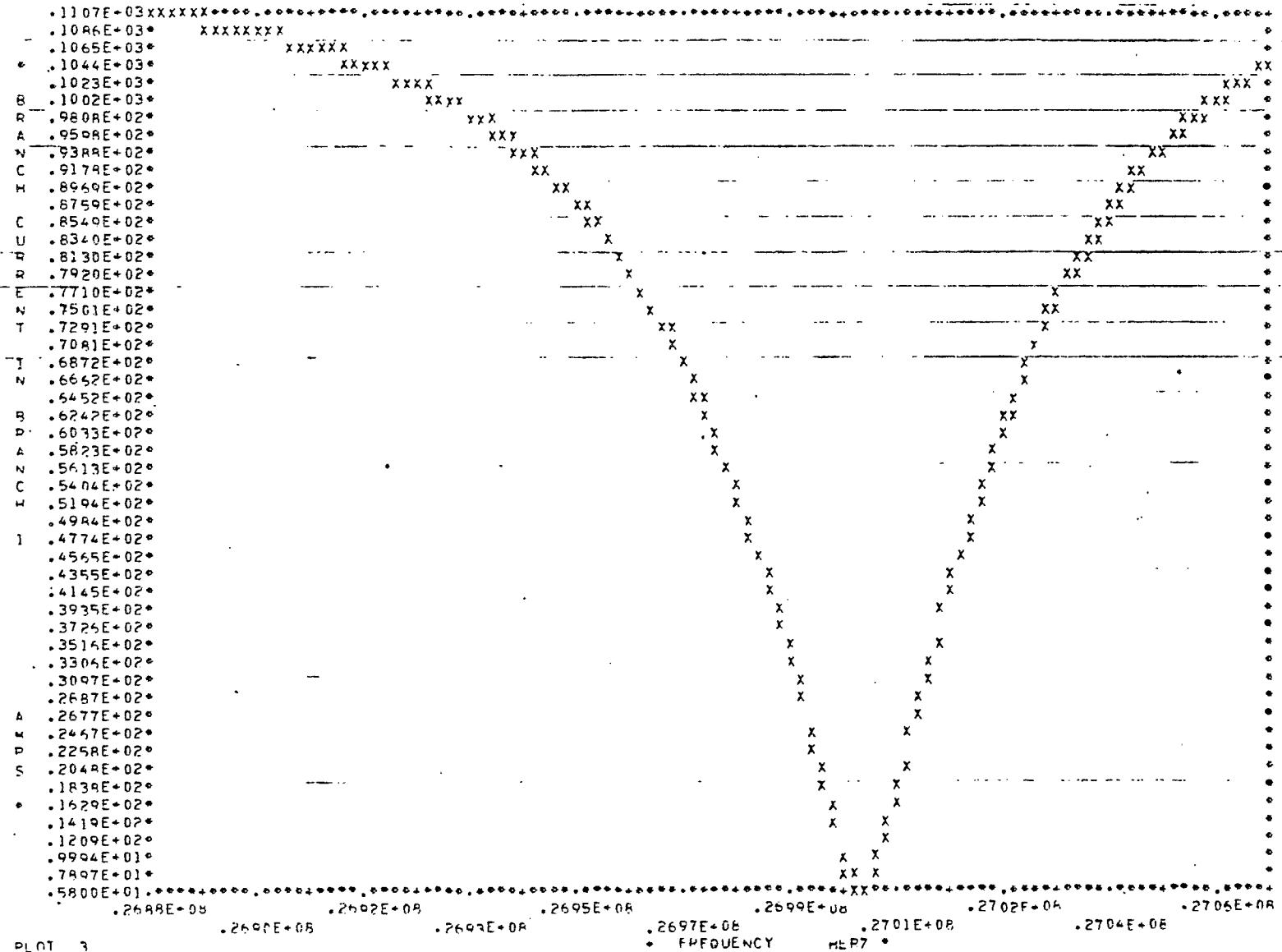
```

NO FATAL INPUT ERRORS DETECTED. EXECUTION INITIATED.

ECAP. AC ANALYSIS

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PAGE NO. 10



ECAP. AC ANALYSIS

09/09/84 19.02.71.

PAGE NO. 1

```

1.      AC ANALYSIS
2.      B1  N(0,1),R=100,E=11.912E3
3.      B2  N(1,0),C=70.E-12
4.      R3  N(1,0),L=3.2473E-8
5.      R4  N(1,2),L=6.6018E-7
6.      R5  N(2,0),C=50.E-12
7.      R6  N(2,0),R=0.6E6,I=-0.7
8.      FREQUENCY=27E6
9.      EXECUTE

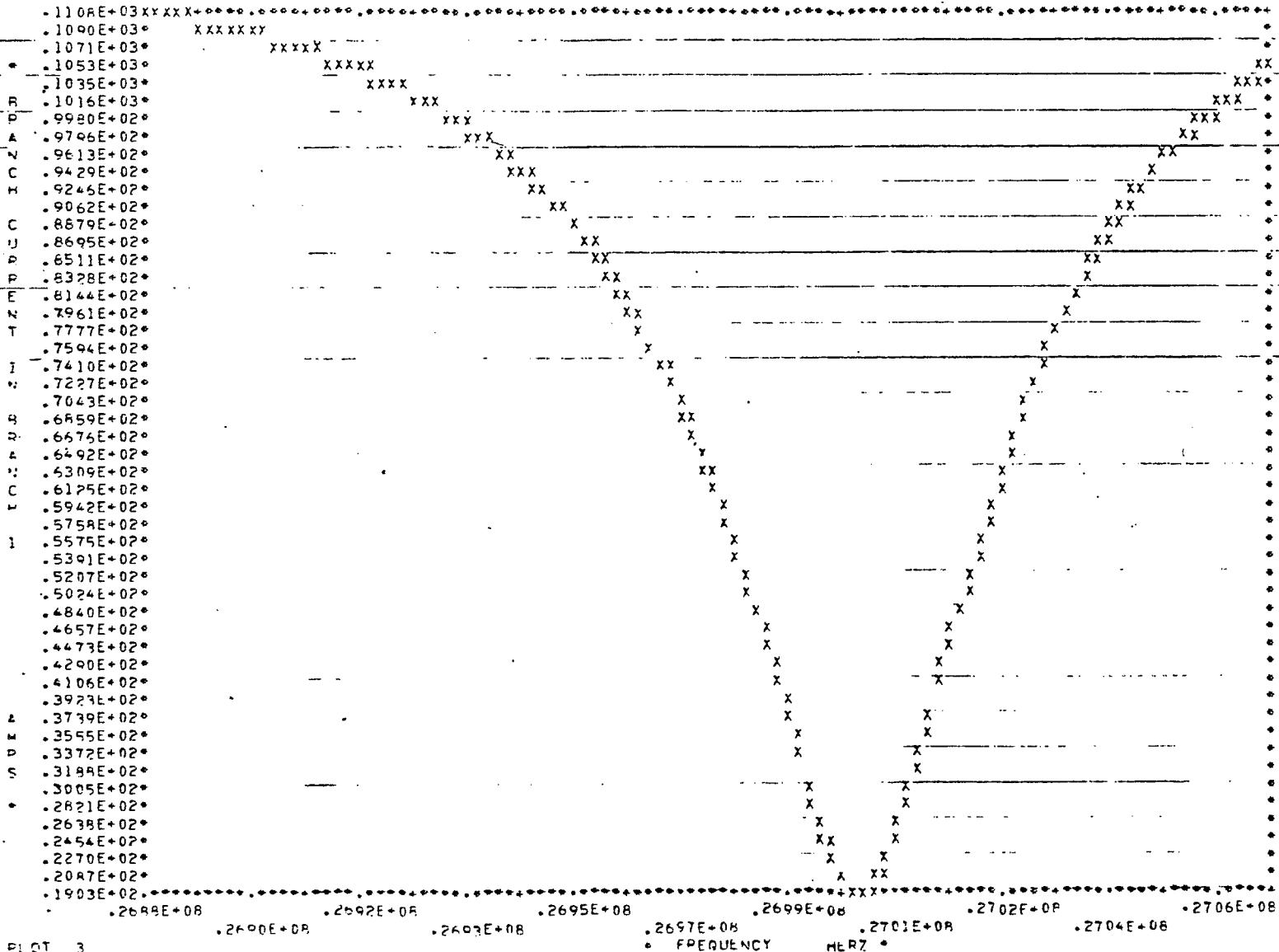
```

NO FATAL INPUT ERRORS DETECTED. EXECUTION INITIATED.

ECAP. AC ANALYSIS

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PAGE NO. 10



PLOT 3

## 10 - Cavity

EC4P. AC ANALYSIS

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PAGE NO. 1

```

1.      AC ANALYSIS
2.      R1  N(0,1),R=100,E=11.912E3
3.      R2  N(1,0),C=70.E-12
4.      R3  N(1,0),L=3.2473E-6
5.      R4  N(1,2),L=6.6018E-7
6.      R5  N(2,0),C=50.E-12
7.      R6  N(1,0),C=7.28E-12
8.      R7  N(2,0),R=0,RE6,I=0.7/90
9.      FREQUENCY=27E6
10.     EXECUTE

```

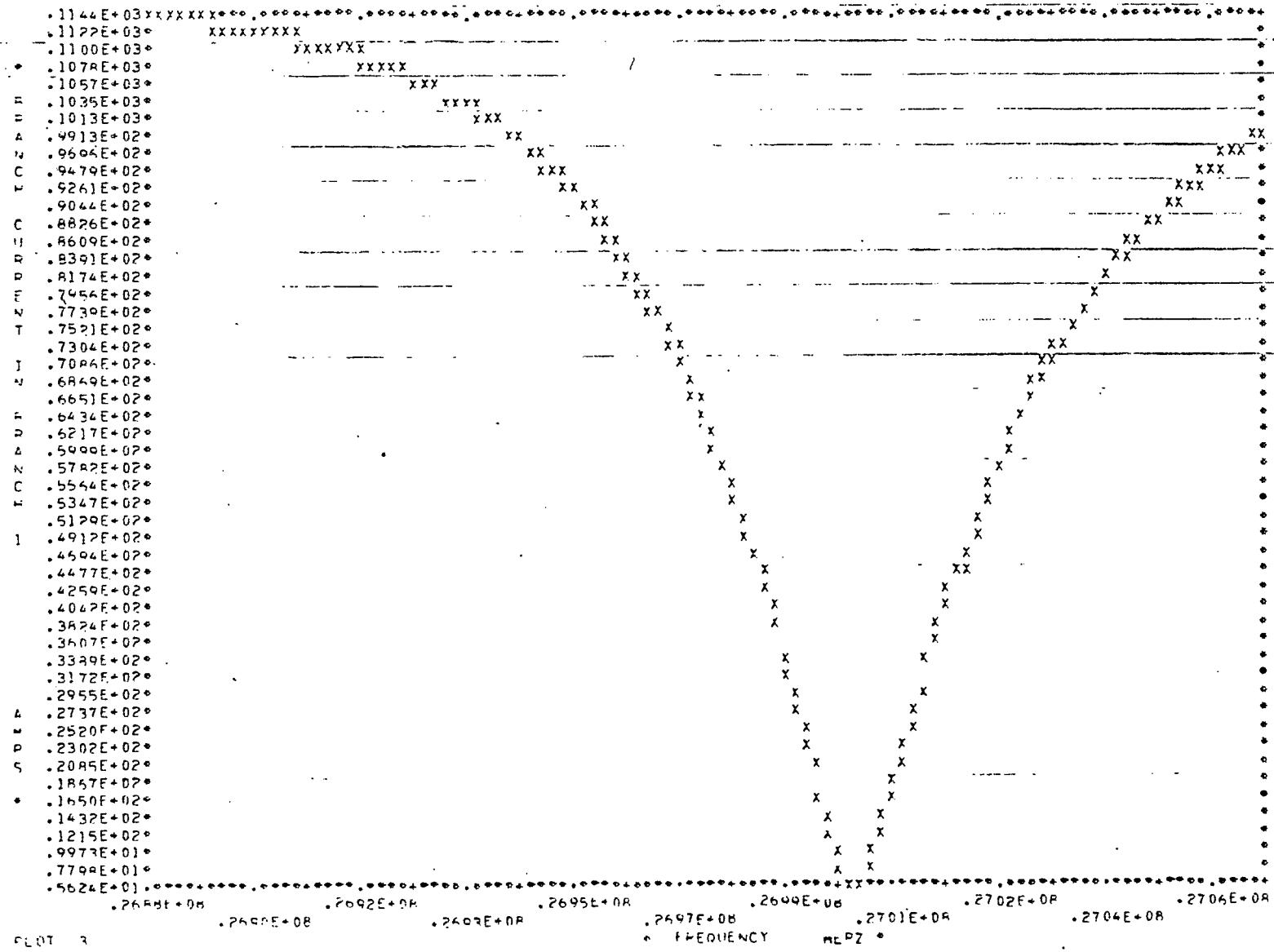
COMPENSATING CAPACITOR

NO FATAL INPUT ERRORS DETECTED. EXECUTION INITIATED.

EC4P. AC ANALYSIS

09/11/84 10.57.21.

PAGE NO. 10



Full Beam - Compensated Circuit

## The Tube

Depending upon the shape of the accelerating cavity the power needed for 200 kV is always in the park of 20÷25 kW per cavity. The power delivered to the beam, even during the acceleration, is negligible. Consequently a tube capable of ~ 30 kW output power is surely enough.

A good candidate could be the Eimac triode 3CW30,000H3 (\$1500 per tube) and the data sheets are included.

From the firm we had two recommended sets of operating conditions:

## a) High Voltage Operation

DC and DE voltage	10 kV
DC grid bias	-450 V
Driving signal	630 V
Plate current (idle)	0.6 A
Plate current	3.16 A
Grid current	0.183 A
Driving power	120 W
Output power	21.5 kW
Load	1880 Ohms
Anode signal	~ 9000 Volts

## b) Low Voltage Operation

DC anode voltage	5.5 kV
DC grid bias	-200 V
Driving signal	825 V
Plate current (idle)	2.5 A
Plate current	8 A
Grid current	1.1 A
Driving power	690 W
Output power	29 kW
Load	400 Ohms
Anode signal	4.8 kV

In both the operating conditions the dynamic anode impedance can be considered as equal to  $330\Omega$ .

From the two sets it is evident that the tube can be operated as a Class A-B amplifier (very low output impedance). Nevertheless, another recommended typical Class C operation is as follows:

DC Anode Voltage	10 kV
DC grid bias	-800 V
Driving signal	1160 V
DC anode current (idle)	0.0 Amp.
DC anode current	6 Amp.
DC grid current	0.315 Amp.
Driving power	365 Watts
Output power	42 KW
Load impedance	750 Ohms
Anode signal	8000 V

In this condition the output impedance of the tube is no longer definable. Any model should be analyzed with the state variable method and the tube should be treated as a highly non-linear device. However, it can be shown that even a Class C operation could be satisfactory if an appropriate tuning is provided for.

#### Modeling the Tube

We assume that, to a first approximation, the following equation holds:

$$I_p = \frac{1}{\rho} (v_p + \mu v_g)^\alpha; I_p > 0. \quad (1)$$

Where  $v_p$  and  $v_g$  are respectively the anode and grid voltages with reference to the cathode and  $\alpha$  is normally near to 3/2. Eq. (1) can be rewritten as follows:

$$\sqrt[\alpha]{\rho} * \sqrt[\alpha]{I_p} - \mu v_g = v_p \quad (2)$$

Assuming  $\alpha = 1.5$  the constants  $\rho$  and  $\mu$  can be determined from two points wisely chosen on the tube data sheet.

Let  $\sqrt[\alpha]{\rho} = x$  then we have:

$$\begin{aligned} x \sqrt[\alpha]{I_{p1}} - \mu v_{g1} &= v_{p1} \\ x \sqrt[\alpha]{I_{p2}} - \mu v_{g2} &= v_{p2} \end{aligned} \quad (3)$$

where the subscript 1 and 2 refer to the points already chosen.

Solving with the Kramer rule we obtain:

$$x = \frac{\begin{vmatrix} V_{1p} & -V_{1g} \\ V_{2p} & -V_{2g} \end{vmatrix}}{\begin{vmatrix} V_{1p} & V_{2g} \\ V_{2p} & V_{1g} \end{vmatrix}} \quad (4)$$

$$\begin{matrix} \alpha\sqrt{I_{1p}} & -V_{1g} \\ \alpha\sqrt{I_{2p}} & -V_{2g} \end{matrix}$$

$$\mu = \frac{\begin{vmatrix} \alpha\sqrt{I_{1p}} & V_{1p} \\ \alpha\sqrt{I_{2p}} & V_{2p} \end{vmatrix}}{\begin{vmatrix} V_{1p} & \alpha\sqrt{I_{2p}} \\ V_{2p} & \alpha\sqrt{I_{1p}} \end{vmatrix}} \quad (5)$$

$$\begin{matrix} \alpha\sqrt{I_{1p}} & -V_{1g} \\ \alpha\sqrt{I_{2p}} & -V_{2g} \end{matrix}$$

When  $x$  is known then  $\rho = x^\alpha$ . Obviously the model is not very accurate. Nevertheless it is always very useful because it allows to take into account both the variation of the internal resistance and the phenomenon of the cut-off that, in many cases, could be extremely important. (For the indicated tube a possible choice could be  $\rho = 30 \cdot 10^3$ ,  $\mu = 20$  with the condition already seen  $\alpha = 1.5$ ).

*Eimac*®

E I M A C

Division of Varian

S A N C A R L O S

C A L I F O R N I A

3CW30,000H3

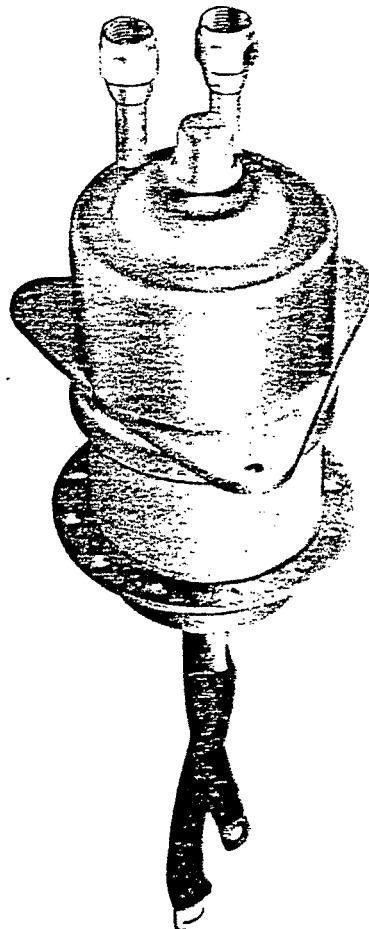
MEDIUM-MU

WATER-COOLED

POWER TRIODE

The Eimac 3CW30,000H3 is a water-cooled, ceramic-metal power triode designed primarily for use in industrial radio-frequency heating services. Its water-cooled anode is conservatively rated at 30 kilowatts of plate dissipation with low water flow and pressure drop.

Input of 60 kilowatts is permissible up to 90 megahertz. Plentiful reserve emission is available from its one kilowatt filament. The grid structure is rated at 500 watts making this tube an excellent choice for severe applications.



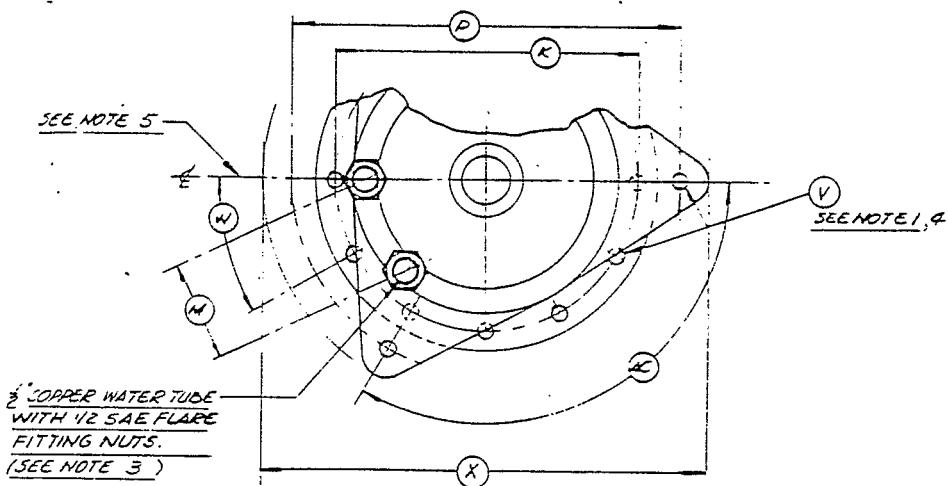
## GENERAL CHARACTERISTICS

### ELECTRICAL

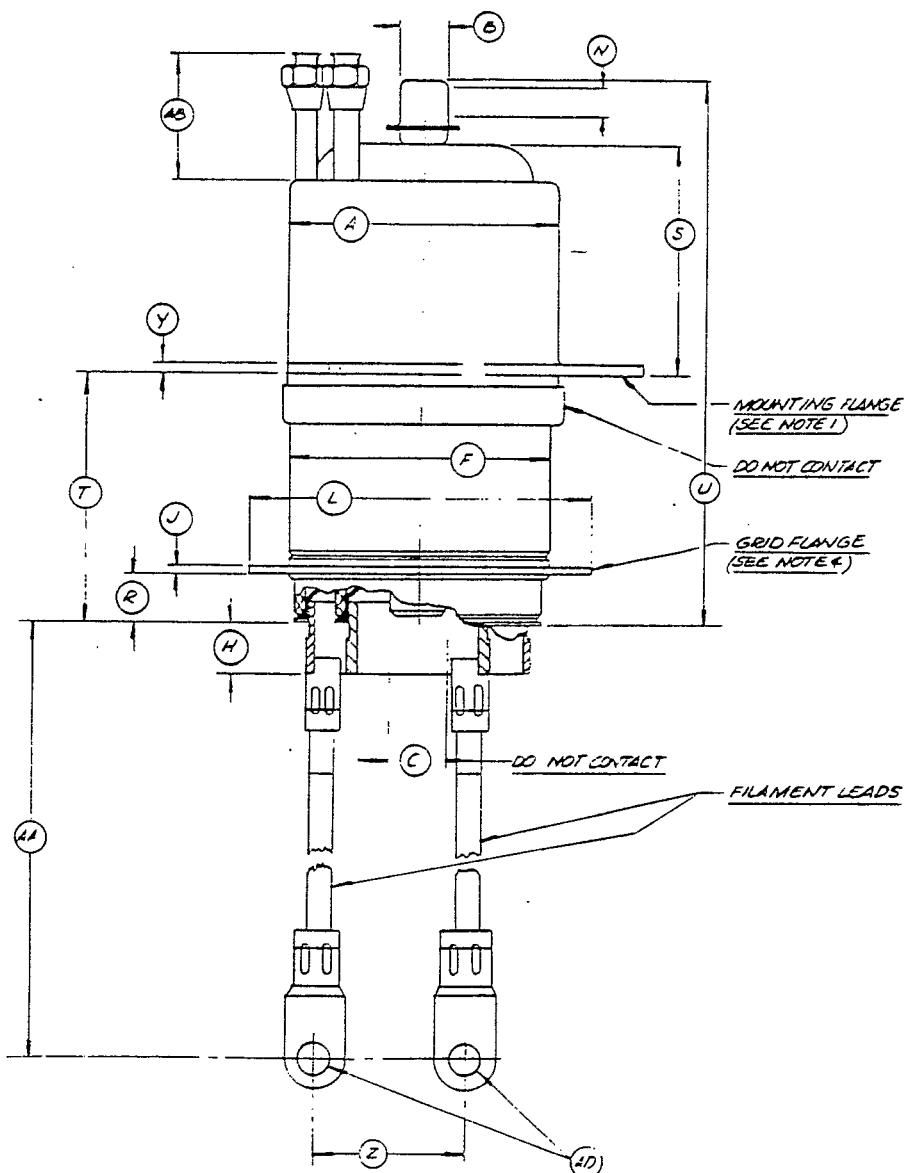
Filament:	Thoriated-Tungsten	Min.	Nom.	Max.
Voltage	- - - - -		6.3	V
Current	- - - - -	152		172 A
Amplification Factor	- - - - -		20	
Interelectrode Capacitances, Grounded Cathode:				
Grid-Filament	- - - - -	48		58 pF
Plate-Filament	- - - - -	1.2		1.5 pF
Grid-Plate	- - - - -	30		38 pF
Frequency for Maximum Ratings	- - - - -		90	MHz

### MECHANICAL

Base	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	See Outline
Operating Position	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	Vertical, base up or down
Cooling	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	Water and Forced Air
Maximum Operating Temperatures:							
Ceramic-to-Metal Seals	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	250°C
Maximum Dimensions:							
Height	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	See Outline
Diameter	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	See Outline
Net Weight	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	- - - - -	12 pounds

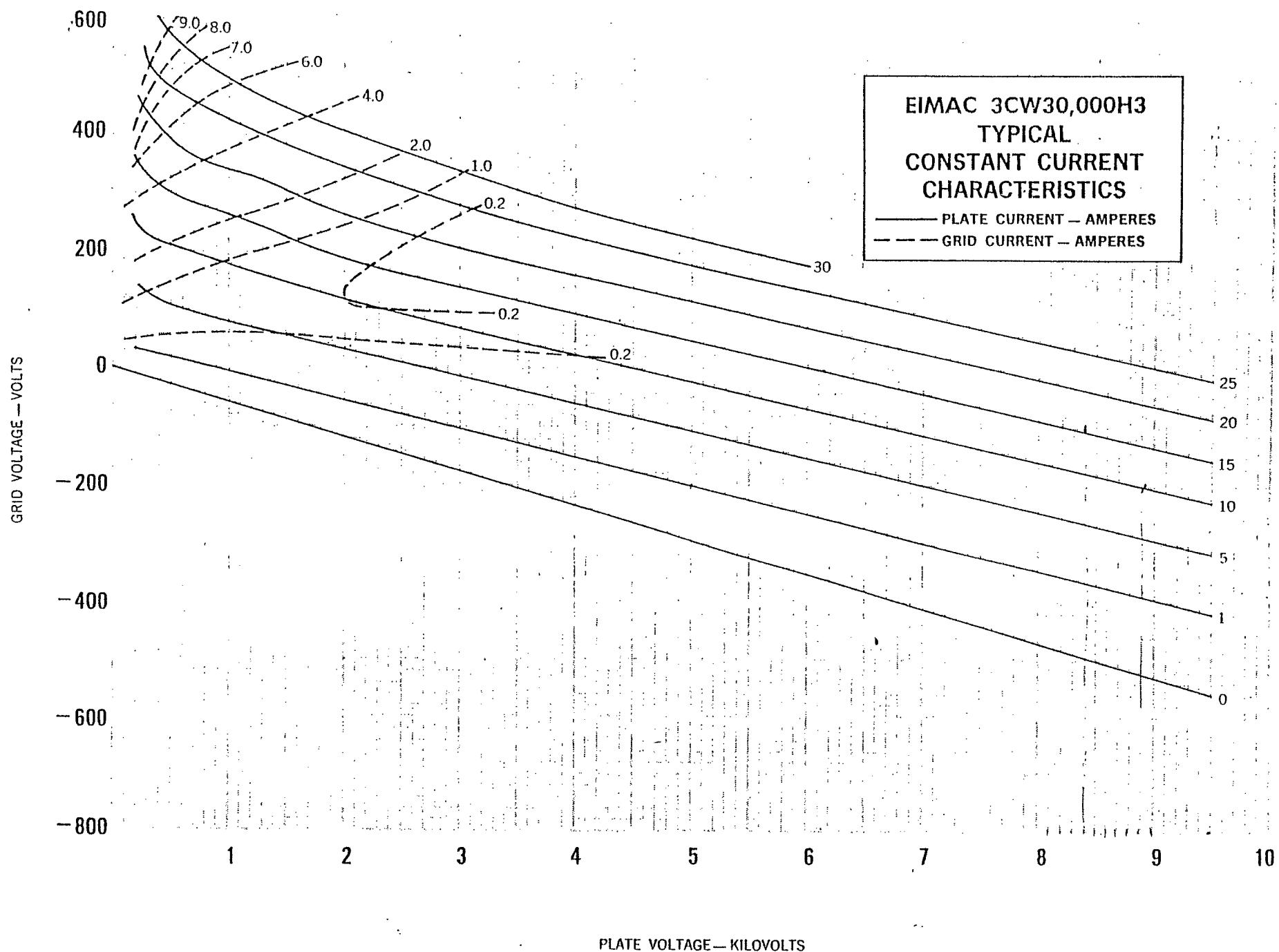


DIMENSIONAL DATA		
DIMENSIONS IN INCHES		
DIM.	MIN.	MAX. REF.
A	.050	.250
B	.860	.890
C	.720	.760
F	3.792	3.832
H	.530	.700
J		.125
K	.025	.445
L	5.030	5.090
M		1.500
N	.375	
P	5.990	6.010
R	.200	.660
S	3.300	3.500
T	3.950	4.100
U	8.250	8.750
V		.250
W	29°	31°
X		6.750
Y		.250
Z		2.000
AA	8.500	9.000
AB		2.000
AC	118°	122°
ID		.390



NOTES:

1. 3 MOUNTING HOLES IN MTG. FLANGE.
2. REF. DIMS. ARE FOR INFO. ONLY AND ARE NOT RECD FOR INSPECTION PURPOSES.
3. EITHER FITTING CAN BE USED AS INLET OR OUTLET.
4. 12 HOLES IN GRID FLANGE
5. MTG. FLANGE, FIL. LEADS & WATER FITTINGS ARE TO BE ORIENTED AS SHOWN.



**EIMAC 3CW30,000H3  
TYPICAL  
CONSTANT CURRENT  
CHARACTERISTICS**

PLATE CURRENT - AMPERES  
GRID CURRENT - AMPERES

~~3C~~ 3CW30,000Hz

### The Beam Loading Transient Analysis

The steady state sinusoidal linear analysis already made shows that if the beam is always in quadrature with the accelerating voltage then the beam current (transformed) do not affect the final tube provided the amplifier is properly tuned. Unfortunately the real situation is quite different because the beam passes throughout the cavity in a very short time.

Each beam occurs every 6 periods of the RF voltage, and the tube could be operated in Class C.

A simple model for simulating this situation is given in Fig. 4.

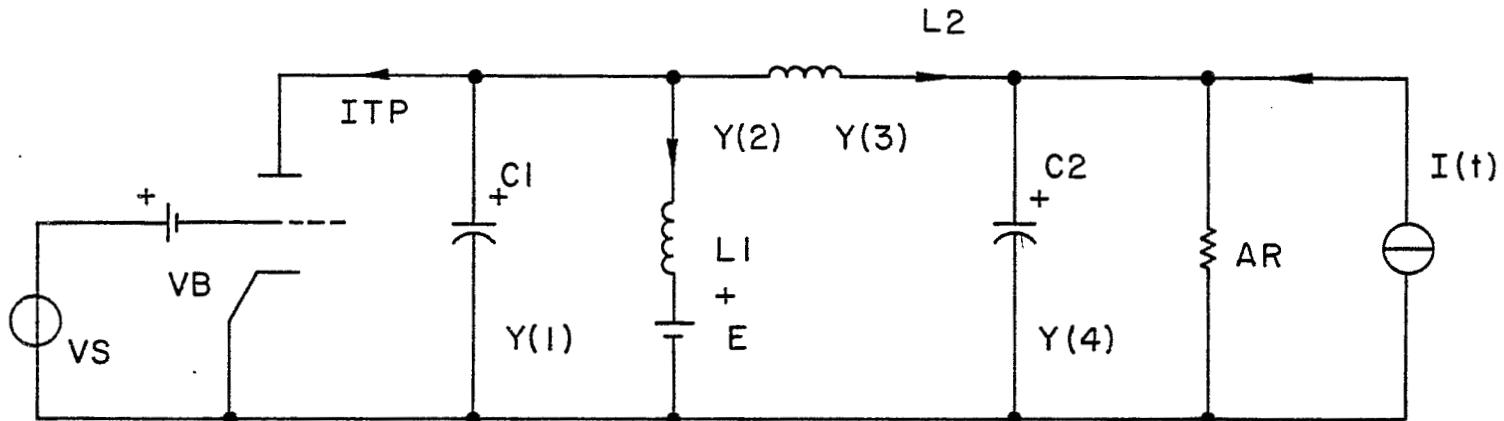


Fig. 4. Equivalent circuit for the transient non-linear analysis. The state variables are as follows:  $Y(1)$ ;  $Y(4)$  are voltages;  $Y(2)$ ;  $Y(3)$  are currents.

After some elementary algebra we obtain the following set of differential equations:

$$\begin{aligned}
 \dot{Y}(1) &= (-Y(2) - Y(3) - ITP)/C1 \\
 \dot{Y}(2) &= (Y(1) - E)/L1 \\
 \dot{Y}(3) &= (Y(1) - Y(4))/L2 \\
 \dot{Y}(4) &= (IT + Y(3) - Y(4)/AR)/C2
 \end{aligned}$$

The beam current could have parabolic distribution and must be periodic with period equal to  $6T$ . That is we have a bunch every 6 RF cycles. The first period is defined as follows:

$$I(t) = -I_p \cdot [1 - (\frac{2t - T}{\tau})^2]$$

$$I(t) = 0.0$$

$$I(t) = 0.0$$

$$\frac{T - \tau}{2} < t < \frac{T + \tau}{2}$$

$$0.0 < \frac{T - \tau}{2}$$

$$\frac{T + \tau}{2} < t < 6T$$

The model for the tube current is not critical. Obviously the cut-off and the dependency of the anode current upon the anode voltage are the most important factors.

From the data sheet already seen we could assume:

$$I_{pt} = 0.3 \cdot 10^{-5} \cdot (V_{pk} + 20 V_{gk})^{1.5}$$

where  $E$  = dc anode voltage = 11 kV;  $V_b$  = dc grid bias = -550;

$V_s$  = Peak driving voltage = 700.V;  $V_{pk}$  = Y(1)

$V_{gk}$  =  $V_s * \sin \omega t + V_b$

The system was integrated with the normal fourth order Runge-Kutta method and the results show that the tube does not see the beam peak current and that the gap voltage remains reasonably sinusoidal even if the cut-off occurs.

In Fig. 1 the beam current and the gap voltage are represented for two different values of the compensating capacity  $\Delta C_T \approx 11 \text{ pF}$  and  $\Delta C_E \approx 13 \text{ pF}$ . Obviously the slope changes (because the amplitudes change) but the waveform remains sinusoidal. In both cases the model was working on Class B condition.

We conclude that the standard linear analysis is applicable for the system "Power Amplifier - Acc. Cavity."

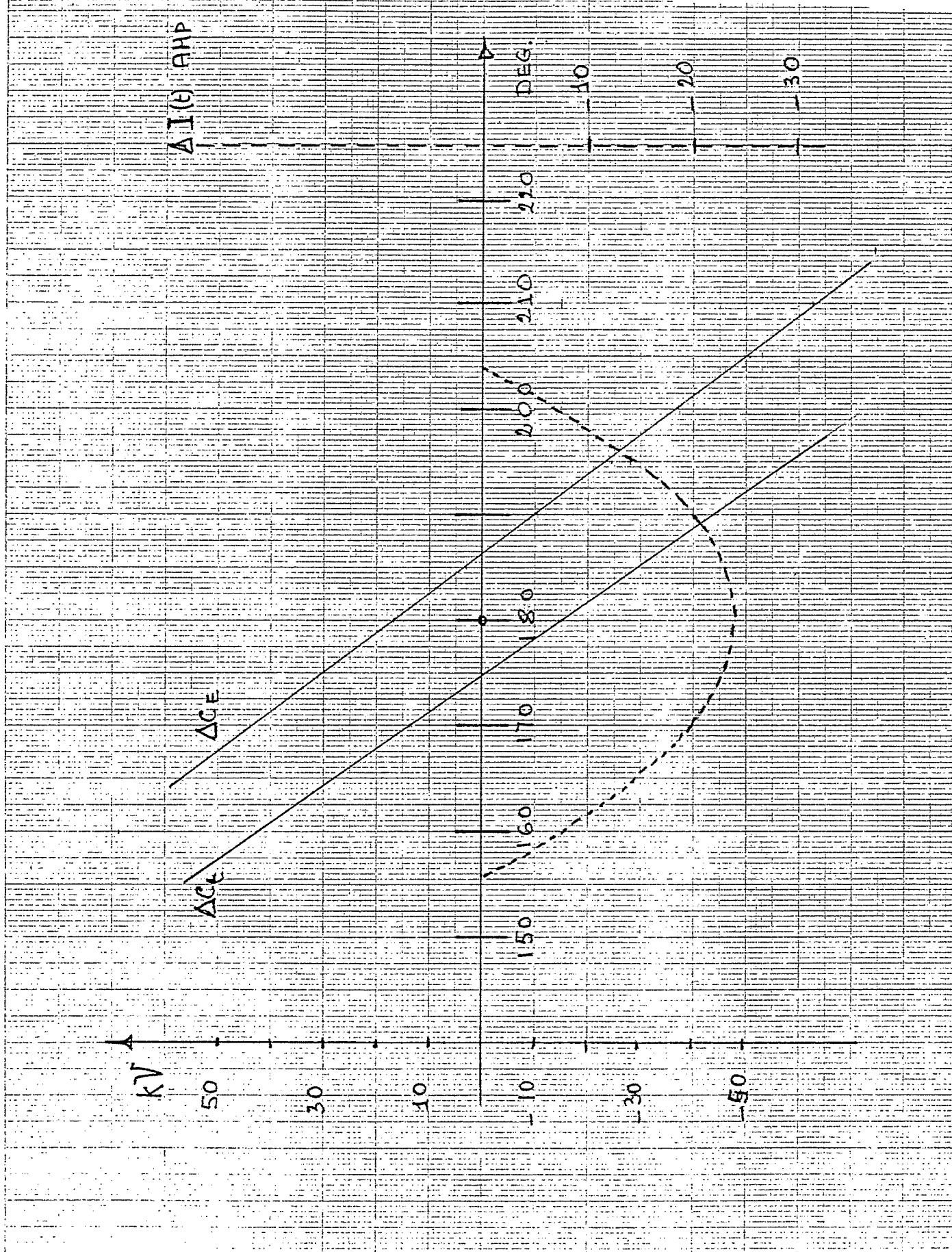


Fig. 1

## The Power Amplifier

1. General Considerations

The final amplifier should exhibit a very low output impedance mainly because this impedance appears across the accelerating cap, at best, multiplied by the square of the voltage transforming ratio ( $N \approx 20$ ).

The cathode follower amplifier could be used but at 27 MHz the stabilization becomes critical and, on the other hand, the achievable output impedance is really low only on a very narrow frequency range.

The grounded grid amplifier is always very stable but its output impedance is very high (~ 40 k $\Omega$  for Class A linear operation using the Eimac tube 3CW30,000H3). The grounded cathode amplifier offers a relatively low output impedance but must be neutralized in order to achieve the wanted stability.

The neutralizing techniques are well-known and for each one there are advantages and disadvantages.

Among the very many the one which could offer a high degree of feedback is the old and practically abandoned "split grid neutering circuit" that is shown in Fig. 1.\*

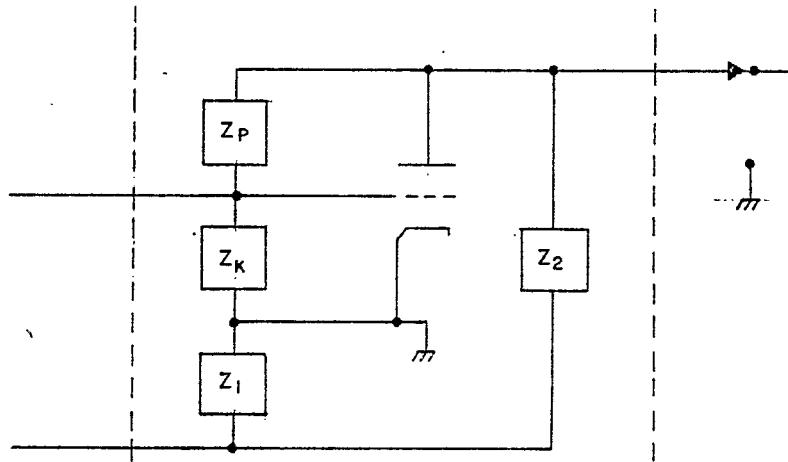


Fig. 1. Split grid neutralized amplifier.

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\*Fig. 1 is only a functional diagram where the input and output circuits (the so-called 'external circuits') are not shown. Normally the Z impedances are purely capacitive. When the grid current is not negligible then Z<sub>K</sub> and Z<sub>1</sub> becomes complex.

Where the impedances  $Z_1$  and  $Z_2$  take into account  $C_{pk}$  and  $C_{gp}$  that are the inherent capacitances of the tube.

It is well-known that over a theoretically infinite range of frequencies a perfect neutralization occurs if and only if:

$$\frac{Z_p}{Z_k} = \frac{Z_2}{Z_1} \quad (1)$$

If the neutralizing condition is fulfilled then the external grid and plate circuits are connected across the opposing vertices of a balanced bridge and, consequently, the circuits ignore each other. Nevertheless the tube is an active element and the impedances  $Z_p$ ;  $Z_k$ ;  $C_1$ ;  $C_2$  creates a negative feedback pattern that strongly reduces the amplifier output impedance.

The simplest possible equivalent scheme for the amplifier is given in Fig. 2.

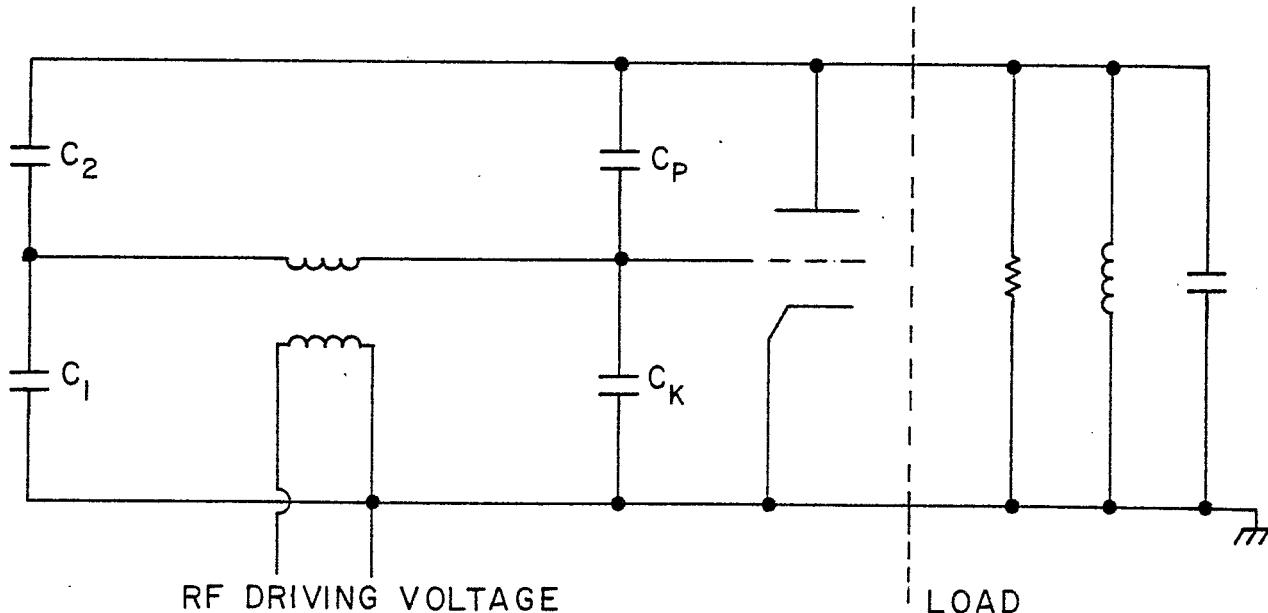


Fig. 2. The neutralized amplifier connected to the load and to the driver.

And on Fig. 3 is drawn the equivalent scheme for calculating the amplifier output impedance.

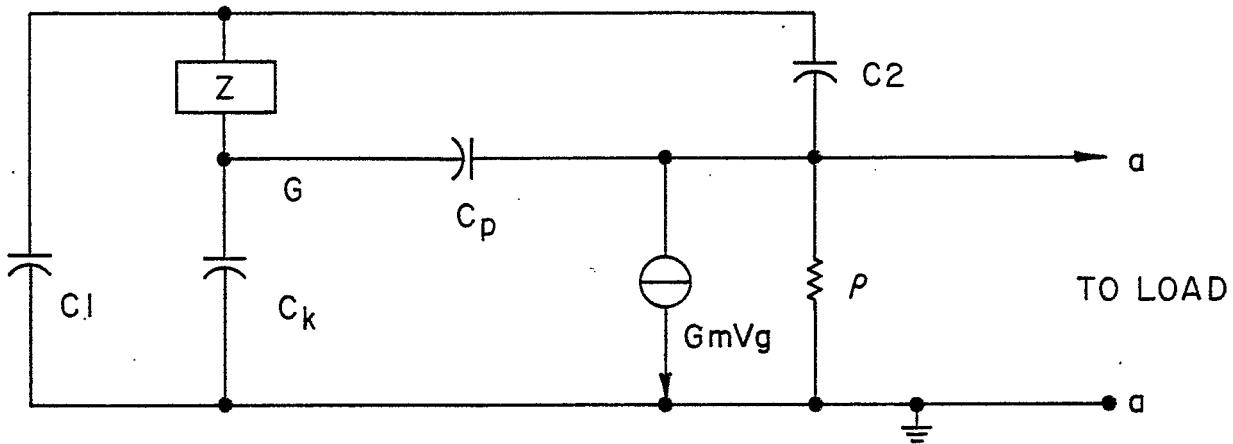


Fig. 3. Equivalent scheme for the neutralized amplifier.  
The impedance  $Z$  takes into account all the grid external circuitry.

In the worst case we could assume that  $\rho = \infty$ ; moreover if we assume

$$C_k = C_1; C_p = C_2$$

the neutralizing condition is satisfied and the calculations are somewhat simpler. Under the previous hypothesis the admittance seen from the terminals aa is as follows:

$$Y(aa) = \frac{G_m + 2j\omega C_2}{1 + \frac{C_1}{C_2}}$$

that does not depend neither upon  $Z$  nor upon the load. This means that the output impedance depend upon the parallel combination of a resistor  $R_{eq}$  and a capacitor  $C_{eq}$ .

$$R_{eq} = \frac{1 + C_1/C_2}{G_m}$$

$$C_{eq} = \frac{2}{\frac{1}{C_1} + \frac{1}{C_2}} \quad (2)$$

It should be observed the resistive component is effective only if the tube is "on". This is not the case for the capacitive component that is always present.

It can be shown that if the impedances  $Z_p$ ;  $Z_k$ ;  $Z_2$ ;  $Z_1$  are purely capacitive and the neutralizing condition is fulfilled then:

$$C_{pt} = \frac{C_1 + C_k}{1 + C_k/C_p}$$

$$C_{gt} = \frac{C_p + C_k}{1 + C_k/C_1}$$

where  $C_{pt}$  and  $C_{gt}$  are the total capacities that appear respectively on the plate and on the grid circuits while  $C_1$  is arbitrary and determines the degree of feedback we like to introduce. (Increasing  $C_1$  the feedback is reduced and the normal trade-off is to make  $C_1$  nearly equal to twice the value of  $C_k$ ).

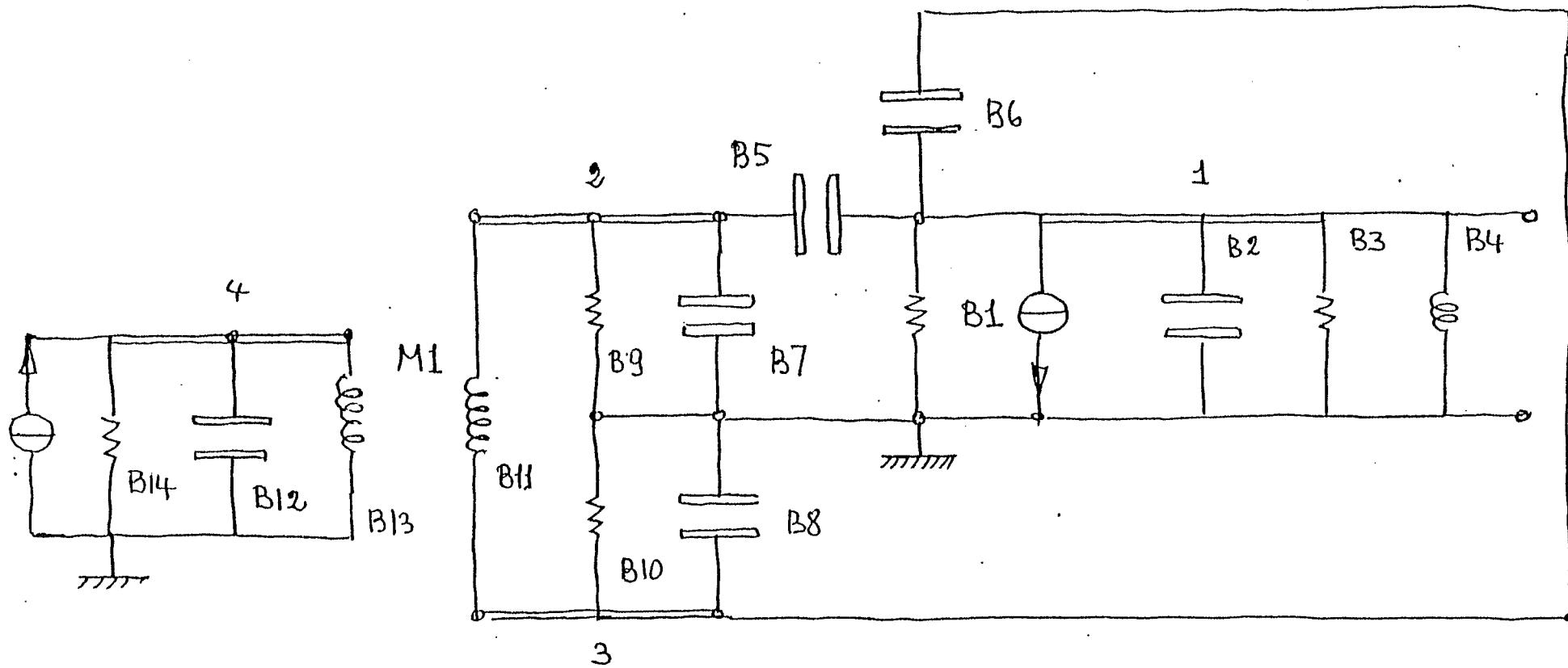
## 2. The Linear Analysis

The triode Eimac 3CW 30,000H3 can be linearized using the following parameters:

Anode dynamic resistance	$\rho = 400\Omega$
Transconductance	$G\mu = 0.05$ Siemens
Grid to cathode capacity	$C_k = 60$ pf
Grid to anode capacity	$C_p = 40$ pf

and in Fig. 3 is given the amplifier eq. scheme prepared for the ECAP ac analysis. It should be noted that we assumed a load of about  $700\Omega$ . That surely represents the worst case (cavity shunt impedance  $\sim 0.8 M\Omega$ ; anode ac voltage  $\sim 6000$  V;  $N = 33.8$ ).

FINAL AMPLIFIER EQ. MODEL  
FOR AC. ECAP ANALYSIS



$H_1 = B(13, 11)$ ,  $L = 0.1 \times 10^{-7}$

$B_1 = N(1, 0)$ ,  $R = 400$ .

$T_1 = B(9, 1)$ ,  $GM = 0.05$

$B_{15} = N(1, 0)$ ,  $R = 1.E6$ ,  $I = 1$ . IS USED ONLY FOR MEASURING THE OUTPUT IMPEDANCE.

## 6 - Amplifier

The first three plots give information for the transfer functions. Namely the ratio NV1/NV4 = 0.65 should be interpreted as follows.

The input impedance is equal to  $\sim 8.6 \text{ k}\Omega$  and if we need 6 kV on the output we need a driving current near to  $\sim 1.07 \text{ amp}$ . That means  $\sim 5. \text{ kW}$ , while the amplifier is generating 6 kV on  $700 \Omega$  which means an output power  $W_o$  equal to  $\sim 25.7 \text{ kW}$ .

The real voltage gain  $G_o$  should be measured between the driving node (NV(2)) and the driven node (NV(1)) and we find  $G_o = 12.29$ .

On the last plot we have the output impedance that, due to the feedback, is reduced to  $54 \Omega$ .

It should be observed that the situation we have simulated is really the worst case because we assumed that the grid circuit is loaded with  $2000 \Omega$  (see the listing for ECAP).

Normally much larger resistors are used and the input impedance can be easily raised by a factor of 3.

Moreover if an anode voltage of 10 kV is required then the anode load should be near to  $2000 \Omega$ .

Consequently we conclude that in any case the split-grid neutralized amplifier would be effective.

7 - Amplifier

ECAP. AC ANALYSIS

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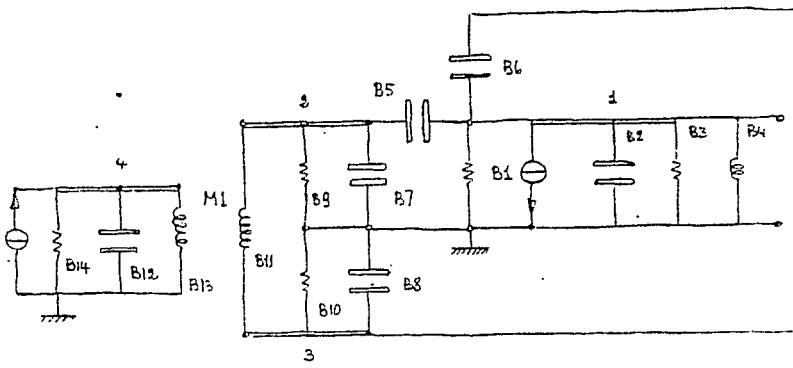
```

1.      AC ANALYSIS
2.      B1      N(1,0),R=400.
3.      B2      N(1,0),C=0.40E-10
4.      B3      N(1,0),R=700.
5.      B4      N(1,0),L=0.35768E-6
6.      B5      N(1,2),C=40.E-12
7.      B6      N(1,3),C=40.E-12
8.      B7      N(2,0),C=0.1E-9
9.      B8      N(3,0),C=0.1E-9
10.     B9      N(2,0),R=0.1E4
11.     B10     N(3,0),R=0.1E4
12.     B11     N(2,3),L=0.49638E-6
13.     B12     N(4,0),C=0.70E-10
14.     B13     N(4,0),L=0.49638E-6
15.     B14     N(4,0),R=1.E6,I=1.
16.     M1      B(13,11),L=0.1E-7
17.     T1      B(9,1),GM=0.05
18.     FREQUENCY=27.E6
19.     EXECUTE

```

NO FATAL INPUT ERRORS DETECTED. EXECUTION INITIATED.

FINAL AMPLIFIER EQ. MODEL  
FOR AC ECAP ANALYSIS



M1  $B(13,11)$ ,  $L = 0.1 E-7$       B1  $N(1,0)$ ,  $R = 400$ .

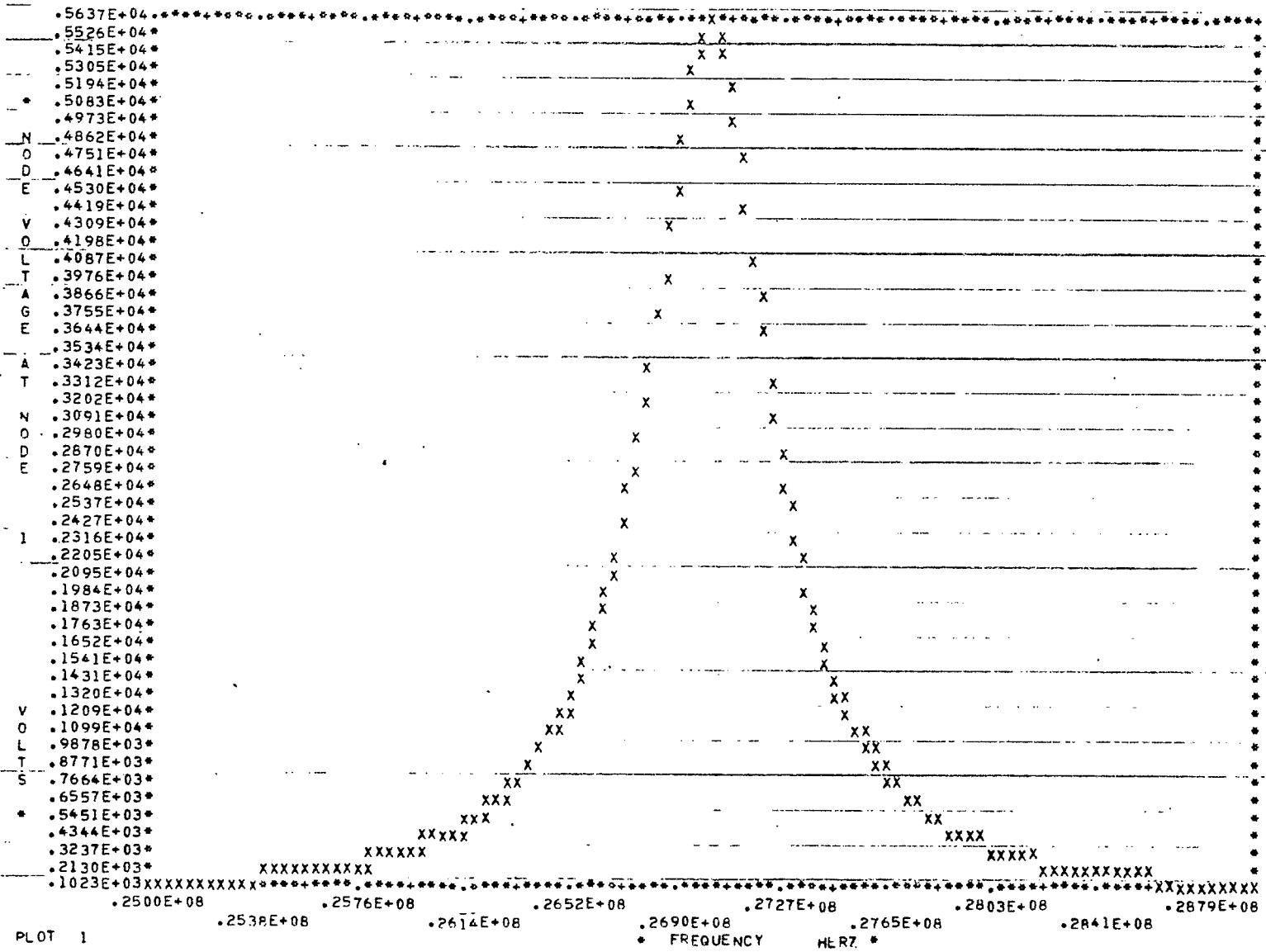
T1  $B(9,1)$ ,  $GM = 0.05$

## 8 - Amplifier

ECAP. AC ANALYSIS

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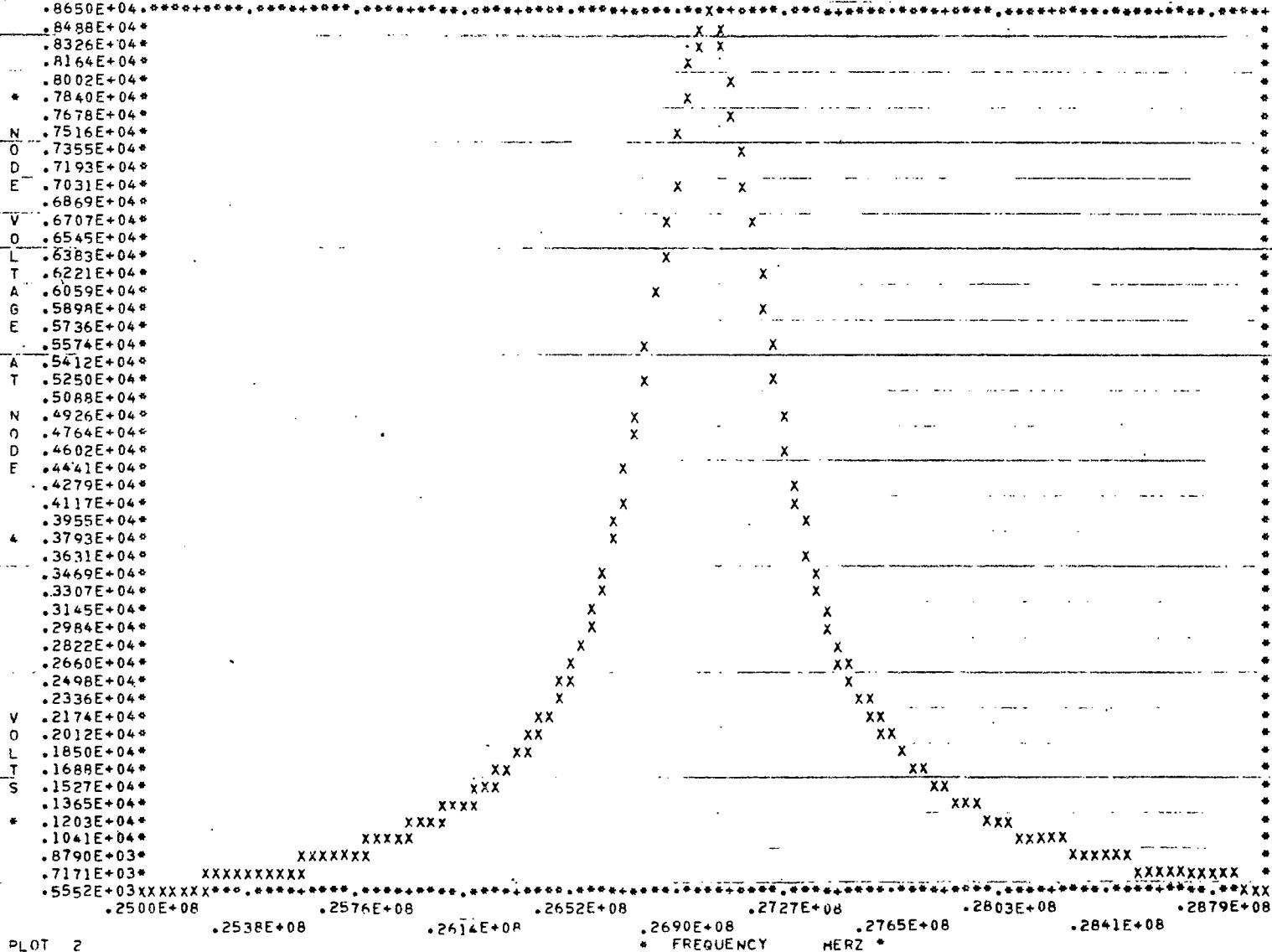


PLOT 1

ECAP. AC ANALYSIS

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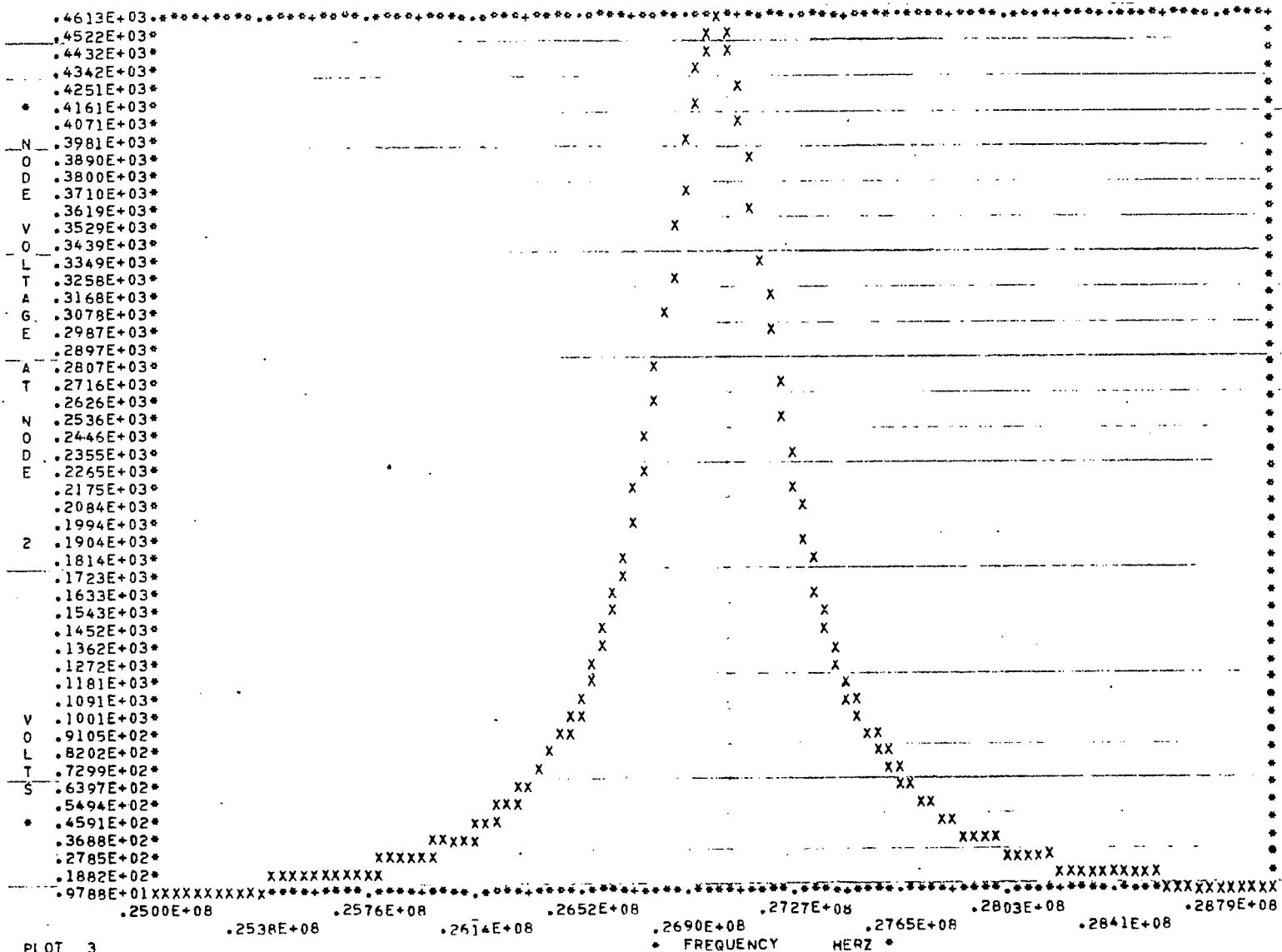


PLOT 2

## ECAP. AC ANALYSIS

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ECAP. AC ANALYSIS

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PAGE NO. 1

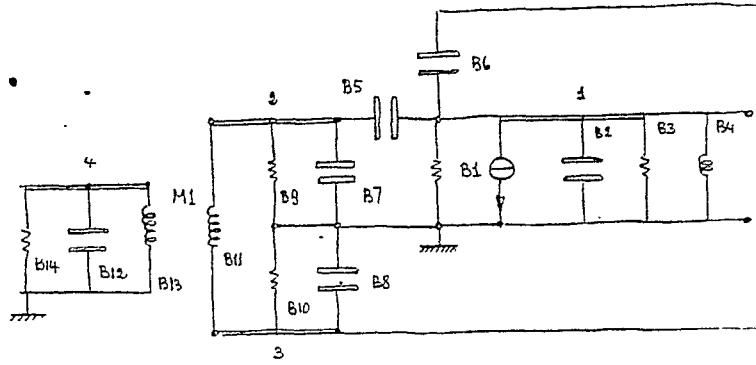
```

1.      AC ANALYSIS
2.      B1      N(1,0),R=400,
3.      B2      N(1,0),C=0.40E-10
4.      B3      N(1,0),R=700.
5.      B4      N(1,0),L=0.3576BE-6
6.      B5      N(1,2),C=40.E-12
7.      B6      N(1,3),C=40.E-12
8.      B7      N(2,0),C=0.1E-9
9.      B8      N(3,0),C=0.1E-9
10.     B9      N(2+0),R=0.1E4
11.     B10     N(3,0),R=0.1E4
12.     B11     N(2,3),L=0.49638E-6
13.     B12     N(4,0),C=0.70E-10
14.     B13     N(4,0),L=0.49638E-6
15.     B14     N(4,0),R=1.E6
16.     B15     N(1,0),R=1.E6,I=1.
17.     M1      B(13,11),L=0.1E-7
18.     T1      B(9,1),GM=0.05
19.          FREQUENCY=27.E6
20.          EXECUTE

```

NO FATAL INPUT ERRORS DETECTED. EXECUTION INITIATED.

FINAL AMPLIFIER EQ MODEL  
FOR AC ECAP ANALYSIS

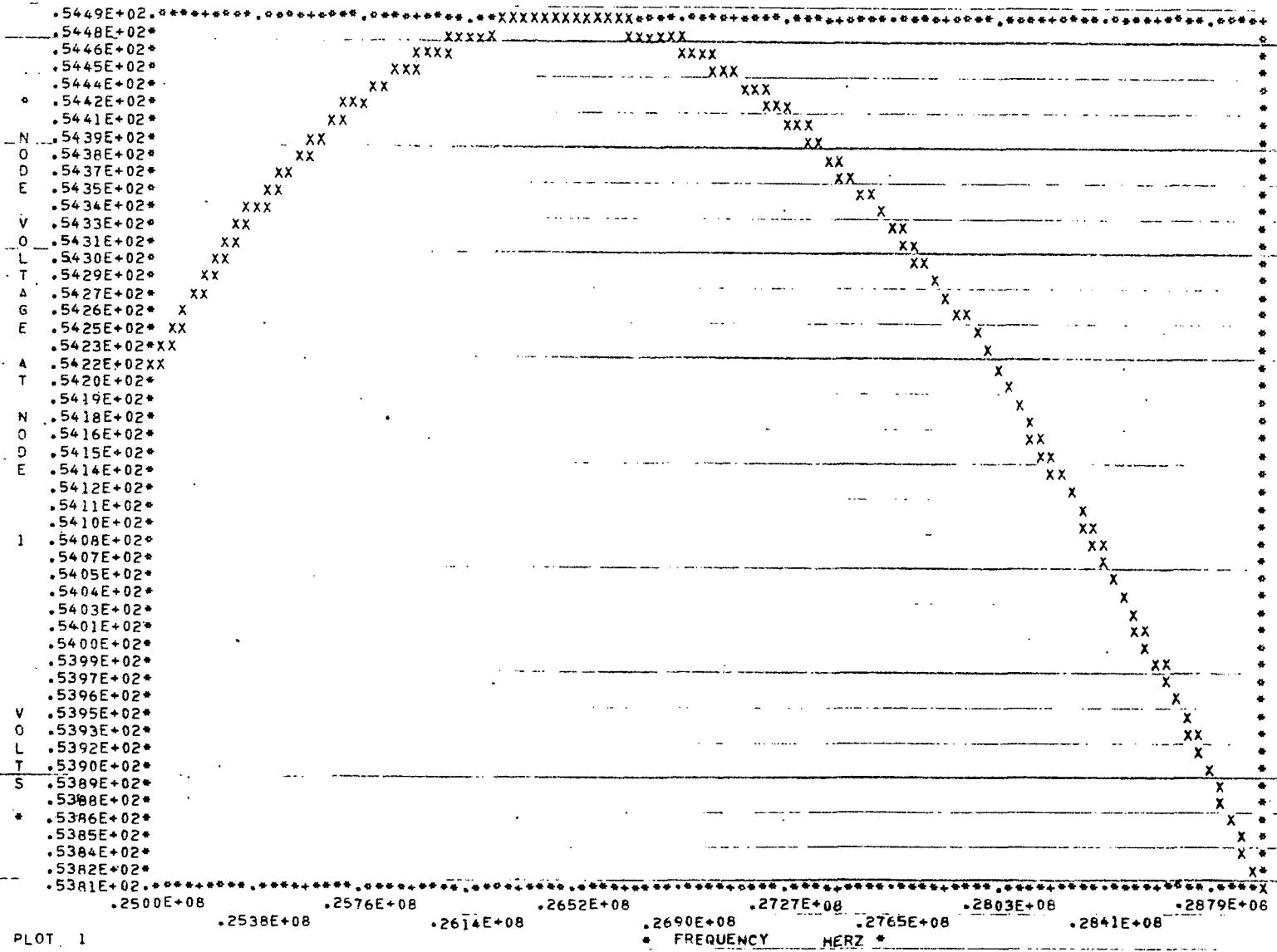
M1  $B(13,11)$ ,  $L = 0.1 E-7$       B1  $N(1,0)$ ,  $R = 400$ .T1  $B(9,1)$ ,  $GM = 0.05$ B15  $N(1,0)$ ,  $R = 1.E6$ ,  $I = 1$ . IS USED ONLY FOR MEASURING  
THE OUTPUT IMPEDANCE.

## 12 - Amplifier

ECAP. AC ANALYSIS

09/19/84 09.37.47.

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PLOT. 1