DESIGN OF A STABLE 150 KW 23 MHz AMPLIFIER FOR THE TRIUMF CRM

by

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This thesis discusses the design of a stable 150 KW 23MHz RF system for the TRIUMF CRM cyclotron. The required characteristics of this system are presented with emphasis on the amplitude and phase modulation constraints. The composition of an amplifier system satisfying the power, bandwidth and noise requirements is discussed. Both the initial and present PA designs are presented, as is the design of the driver amplifier. Also included is a discussion of the choice of tubes and RF circuits.

The usefulness of feedback control in satisfying the RF stability requirements is shown. The conditions a stable feedback system must satisfy are also given. Two amplitude control systems (drive control and screen control) are designed using Bode techniques. A digital simulation of these systems using an electrical analogue is presented. Implementation and results are discussed.
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CHAPTER 1. SYSTEM DESCRIPTION

1.1 General Requirements

This thesis will describe the design of the RF system built for the TRIUMF CRM cyclotron. An exceptional feature of this system is the extreme stability required of the RF waveform if the energy resolution desired of the cyclotron \( \frac{\Delta KE}{KE} \leq 2 \times 10^{-5} \) is to be achieved. The steady state RF requirements for the TRIUMF cyclotron have been calculated; those applying to the CRM RF system may be briefly listed as follows:

(a) a frequency of 23 kHz,
(b) a power output of 150 KW,
(c) a peak cavity voltage of 100 KV,
(d) a frequency stability of \( \pm 1.25 \) parts in \( 10^6 \) for maximum duty factor and \( \pm 7.5 \) parts in \( 10^8 \) for single turn extraction, and
(e) a voltage amplitude stability of \( \pm 2 \) parts in \( 10^4 \) for maximum duty factor and \( \pm 2.5 \) parts in \( 10^5 \) for single turn extraction.

In addition, the delicate nature of the electronic instrumentation at TRIUMF demands that the level of RF radiated be extremely low. It has been found that spurious radiation at the TRIUMF site must be kept below a level of \(.5 \mu V/m\), a level 20 times lower than that stipulated by the Department of Transport.

1.2 RF Modulation Constraints

The stability requirements listed in the above section gave the maximum permissible static deviation of a waveform parameter from its desired value. However, disturbances to the system are not static in nature, so an estimate must be made of the permissible amplitude of time varying deviations in frequency, phase and amplitude. Since the frequency synthesizer has a short term frequency stability of less than 5 parts in \( 10^8 \), only amplitude and phase variations were considered when examining the CRM requirements. That is, if \( \frac{\Delta f}{f} = R \) then for

(a) maximum duty factor: \( \frac{R}{R \text{ synthesizer}} \geq 25 \)
(b) single turn extraction: \( \frac{R}{R \text{ synthesizer}} \geq 1.5 \)
if there is no amplitude or phase modulation present. Figure 1 shows how both the amplitude of the resonator RF (V) and its phase ($\theta_m$) with respect to the instant at which a beam pulse crosses the accelerating gap will affect the accelerating voltage (VA). Note that the effect of phase errors introduced into the RF system is twofold:

(a) the actual change in phase will introduce a phase error
(b) the rate of change of phase will introduce a frequency deviation which will, due to the resonant nature of the system, lower the resonator voltage.

Ideally, the resonator voltage should be maintained constant at the desired level ($V_0=100$ kV) with the beam pulses always crossing the accelerating gap at the peak of the RF waveform ($\theta_m=0$).

![Fig. 1. RF Accelerating Voltage; RF out of Phase with Beam](image)

1.2.1 RF Amplitude Modulation Constraints
(No Frequency or Phase Modulation Present)

A. Sinusoidal Modulation

Assume that the gap voltage is modulated sinusoidally, with the voltage remaining constant as the particle passes through the gap.
That is, let the energy gained per pass through the acceleration gap be 

\[ (q \times V_{\text{res}}) \]

where \( q \) = particle charge 
and \( V_{\text{res}} \) = gap voltage 

Hence, when amplitude modulation is present, the total energy gained by a particle will be

\[
KE_T = \sum_{n=0}^{N} 2qV_{o} (1 + m \cos (w_m n \Delta t + \phi)) = \sum_{n=0}^{N} q V_{\text{res}}
\]

(1.1)

\( V_o \) = magnitude of fundamental RF 
\( m \) = ratio of modulating freq. amplitude to fundamental amplitude \(< 1\) 
\( w_m \) = freq. of modulating signal rad. 
sec. 
\( \Delta t \) = time between impulses 
\( = \frac{1}{2} \) time for 1 rev. \( \approx 1 \times 10^{-7} \) sec. 
\( \phi \) = phase of modulating signal at instant of particle injection 
\( N \) = twice the total no. of rev. one particle makes 

The factor (2) is present because the resonators are to work in the push-pull mode.

The value of \( KE_T \) may be approximated by substituting an integral for the sum of the sequence. Thus, if \( \Delta t \) is sufficiently small, one may write:

\[
KE_T = 2NqV_{o} + \frac{2q m V_{o}}{\Delta t} \int_{0}^{t_f} \cos (w_m t + \phi) \, dt
\]

\[
= 2NqV_{o} + \frac{2q m V_{o}}{w_m \Delta t} \sin (w_m t_f + \phi) - \sin \phi
\]

(1.2)

\( t_f \) = total acceln. time \( \approx 250 \mu \) sec. 

Assuming that reasonable accuracy requires that the period of the sinusoid be at least 100 times as long as the time (\( \Delta t \)) between terms of the sequence for (\( KE_T \)), then the above approximation is adequate for modulating frequencies below 100 KHz. (That is, we require \( fm \frac{10^{-2}}{\Delta t} = 10^5 \) Hz.) 

It is not anticipated that higher modulating frequencies will be encountered. The energy resolution desired now requires that

\[
\frac{|\Delta KE|}{KE} = \frac{m}{w_m t_f} \left| \sin (w_m t_f + \phi) - \sin \phi \right| < \varepsilon
\]

(1.3)
Fig. 2. Relationship between $m/\epsilon$ and Modulating Frequency
Hence substitute: \( \theta = \text{wmt}_f \)
\[ \delta = \frac{\epsilon}{m} \]
and solve for \( \theta \)

\[ \frac{1}{\theta} \left| \sin(\theta + \phi) - \sin\phi \right| < \delta \]  
(1.4)

The solutions vary between extremes for which \( \phi = 0 \) and \( \phi = \pi/2 \). The first places the more stringent requirements on the amplitude of modulation at low frequencies.

Thus solving i) \( |\sin \theta| < \delta \theta \) (\( \phi = 0 \))

ii) \( |\cos \theta - 1| < \delta \theta \) (\( \phi = \pi/2 \))

for values of \( \theta = 2\pi f_m t \) and \( \delta = \frac{\epsilon}{m} \) gives the relationship between \( f_m \) and \( m \).

Figure 2 gives the resulting relationships between \( f_m \) and \( m \) for \( \phi = 0 \) and \( \phi = \pi/2 \).

Noting that below about 1 KHz the permissible upper bound on \( \frac{m}{\epsilon} \) is given for \( \phi = 0 \), while that above this frequency it is defined by the lowest points on the curves for \( \phi = \pi/2 \), one may conservatively define the overall bounds on \( \frac{m}{\epsilon} \) as:

\[ \frac{m}{\epsilon} \sim \begin{cases} 1 & f_m \leq 1.3 \text{ KHz} \\ (7.8 \times 10^4 f_m & f_m > 1.3 \text{ KHz} \end{cases} \]  
(1.7)

A plot of \( m = \text{amp. modulating signal in decibels against modulating amp. fundamental} \) frequency is given in Figure 3.

B. Rectangular Pulse Amplitude Modulation

An estimate was made of the restrictions on the amplitude and duration of a single rectangular amplitude pulse superimposed on the steady-state resonator RF amplitude. Assuming the worst case—that is, that the start of the pulse coincides with entry into the cyclotron of a particle to be accelerated—then the total energy gained by that particle will be

\[ KE_T = 2qV_o (N + mn) \]

\( V_o = \text{steady state amplitude of the accelerating RF voltage} \)
\( N = \text{total number of accelerating impulses per particle} = 2500 \)
\( n = \text{number of accelerating impulses for which the amplitude pulse persists} \)
Fig. 3. Plot of $m = \text{Max. allowable Mod. Amp.}/\text{RF Amp.}$ against Modulating Frequency ($fm$)
\[ m = \frac{V_p}{V_0} \text{ where } V_p = \text{pulse height} \]
\[ \epsilon = \text{energy resolution required} = 2 \times 10^{-5} \]
\[ t_p = \text{pulse length} \]

Then the restriction on the energy gained by the particle is

\[ \left| \frac{\Delta KE}{KE} \right| < \frac{mn}{N} \epsilon \tag{1.8} \]

Letting \( n = \frac{tp}{t} \) gives the restriction

\[ mt_p < \epsilon N \Delta t = 5 \times 10^{-9} \tag{1.9} \]

Figure 4 shows this restraint.

### 1.2.2 RF Phase Modulation Constraints
(No Amplitude or Frequency Modulation Present)

Calculations similar to those of part A of 1.2.1 have been done for sinusoidal phase modulation. That is, it was assumed that the phase error (\( \theta_m \)) was a function

\[ \theta_m = A_\theta \sin 2\pi f_m t \tag{1.10} \]

The resultant restrictions on the magnitude of \( A_\theta \) are shown in Figure 5, where it is apparent that for low frequencies \( (f_m \leq 10^2 \text{ Hz}) \) the restrictions on the amplitude of RF phase variations become increasingly relaxed.

This is because each new beam pulse is injected into the cyclotron with zero initial phase error by means of a control system that monitors the resonator RF and feeds particle bunches into the beam line at the proper times.

### 1.3 RF System Description

The RF system built to satisfy the CRM requirements is shown in Figure 6. It is apparent that the system may be divided into two functionally different—though not independent—subsystems: a system to generate the required RF, and a system to regulate its amplitude and phase. The following is a brief description of the main RF function blocks.
Fig. 4. Plot of $m$ (= Max. allowable Pulse Height/RF Amp.) against Pulse Length, $tp$
FIG. 5. PLOT OF MAXIMUM PERMISSIBLE AMPLITUDE OF SINUSOIDAL PHASE MODULATION

VS
MODULATING FREQUENCY
(NO AMP. MOD. PRESENT)
Fig. 6. Block Diagram of the TRIUMF CRM RF System
RF Synthesizer
(Rohde and Schwarz type ND 30 M)

This device is the primary signal source of the RF system. The manufacturer claims a frequency variation $\leq 5 \times 10^{-9}/1^\circ C$ for ambient temperatures ranging from $+10^\circ C$ to $+30^\circ C$. and a long term drift due to aging of less than $5 \times 10^{-8}/$month. Its maximum output voltage is .5 Vrms into a 50 \Ohm load. Hence, to satisfy the CRM voltage and power requirements, the RF amplifier chain must provide voltage and power gains of at least $1.4 \times 10^5$ and $3 \times 10^7$ respectively.

RF Driver

The RF driver takes the output from the synthesizer (or the amplitude modulator when the drive control system is in operation) and boosts this input signal to the level required to drive the main power amplifier. The power required (400 watts) is considerably less than the 2500 watts this stage is capable of. Figure 21 indicates in detail the makeup of this device.

Main RF Power Amplifier

This stage, designed and built by Continental Electronics, \(^3\) provides the 150 KW of RF power required by the CRM cyclotron. The single tetrode tube (EIMAC 4CW250000) employed requires a drive power of only 150 watts\(^4\) for an output of 162 KW. This is a power gain of $\approx 10^3$, and it is obtained with an efficiency of about 75 per cent. For circuit details see Figure 18.

RF Transmission Line

A resonant coaxial transmission line (Figure 8) is employed at TRIUMF to transfer the output of the final power amplifier to the CRM resonators. A resonant line (VSWR $\approx 30$) was chosen primarily because it reduces the sensitivity of the tube's load to changes in the resonator characteristics. As indicated in Figure 8, this line has 3 tuning capacitors which (1) reduce the physical length of the line (for a specified electrical length) and (2) facilitate the matching of the power tube to the resonator load. The first effect is illustrated in Figure 7, where A and B illustrate the voltage standing wave before and after the introduction of a capacitor (C2) in a part of the line where the impedance is inductive. The value of C2 is chosen so that the electrical length
ADD CAPACITOR SUCH THAT \( c || Z_b \rightarrow Z_c \)

ELECTRICAL LENGTH = \( 1 \sqrt{9} \lambda \)

Fig. 7. Effect of Lumped Capacitance on the Voltage Standing Wave of a Transmission Line
of the line is increased by an amount that results in an inductive input impedance \((Z_{\text{in}})\) with a parallel resistive component equal to the load impedance required by the tube. The capacitor \(C_3\) is then chosen so that, in parallel with the output capacitance of the tube, it will cancel the inductive component of \(Z_{\text{in}}\), thereby presenting the tube with its required load impedance. Capacitor \(C_1\), by modifying the impedance terminating the transmission line, permits greater freedom in the choice of \(C_2\) and \(C_3\). Coupling to the tube is achieved through a series (D.C. blocking) capacitor \((C_4)\); coupling to the resonators is achieved by means of a loop. During normal operation, the power loss between these 2 points is about 2000 watts; since this power is dissipated primarily at the current nodes, an excessive temperature rise at these points is avoided by blowing cooling air through the region between the conductors.

**Resonators**

The accelerating structure (RF load) used at TRIUMF is unusual because it has dees which are in the form of \(\lambda/4\) resonant cavities (see Figure 9). To use excitation at the ion rotation frequency would have required these structures to be prohibitively large; instead, the frequency chosen (23 MHz) is the 5th harmonic of the rotation frequency, thus reducing the size by a factor of 5.

The Q of these resonators is about 6000; hence to get the required 200 KV\(^5\) dee-to-dee, approximately 150 KW of RF power is needed.
Fig. 9. Coaxial Structure of Cyclotron Dees
CHAPTER 2. RF AMPLIFIER DESIGN

2.1 General Design Considerations

2.1.1 Achieving the Required Voltage and Power

The techniques required to achieve the power and voltage gains needed by the CRM RF system are well known—a brief summary of the important ideas will be given here. Since one is working at a high frequency \( f = 23 \text{ MHz} \), tuned amplifiers must, of course, be used. Also, since a high output power is desired, it is imperative that the input DC power be efficiently converted to usable RF output power. This is important because it leads to:

(a) lower costs for power
(b) relaxed cooling requirements
and (c) lower plate dissipation requirements.

As shown in Figure 10, high efficiency is obtained by operating the tube such that the total plate conduction angle is less than about \( 180^\circ \) (Class C).

![Figure 10](image)

Fig. 10. Plot of Asymptotic Efficiency \( (E_a) \) against Total Conduction Angle \( (\Theta) \)
Table I indicates that to obtain the required output power there must be at least 3 or 4 stages in the CRM RF amplifier chain. The actual number chosen depends largely on the capabilities of the final power tube used. That is, once an output tube has been chosen, the output power requirements dictate the driving power requirements; these in turn determine the choice of driving tube. This process is repeated until the driving requirements are met by the signal source.

<table>
<thead>
<tr>
<th>n</th>
<th>Gp</th>
<th>Gv</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$7.5 \times 10^6$</td>
<td>$7 \times 10^4$</td>
</tr>
<tr>
<td>2</td>
<td>$2.74 \times 10^3$</td>
<td>$2.65 \times 10^2$</td>
</tr>
<tr>
<td>3</td>
<td>$1.9 \times 10^2$</td>
<td>41.2</td>
</tr>
<tr>
<td>4</td>
<td>52.3</td>
<td>16.3</td>
</tr>
<tr>
<td>5</td>
<td>23.7</td>
<td>9.3</td>
</tr>
</tbody>
</table>

Table I. Average Power and Voltage Gain per Amplifier Stage

2.1.2 Bandwidth Requirements

To act as a guide in the design of the coupling circuits, an estimate was made of the bandwidth requirements of the CRM driver amplifier. Since this design work was done at a time when it was expected that under certain conditions the operating frequency would be increased by 3 percent, it was considered highly desirable to design the driver amplifier with a bandwidth sufficient to accommodate this change without retuning. Only the driver amplifier was to be broadbanded since the load on the final stage (resonators) would be retuned. A driver with a suitable bandwidth may be achieved most simply through the use of synchronously single tuned stages, each having a sufficiently large bandwidth. As shown in Figure 11, the minimum overall driver bandwidth ($B_0$) may be taken to be 3 percent.
of the normal operating frequency (fo).

This leads to an estimate of the minimum stage bandwidth required (Bn) by means of the relationship

$$Bn = \frac{Bo}{(2^{1/n} - 1)^{1/2}}$$  \hspace{1cm} (2.1)

where

$Bo = \text{overall bandwidth} = .03 \text{ fo}$

$Bn = \text{stage bandwidth}$

$n = \text{number of stages}.$

Since 4 tuned circuits are used in the driver, it is required that

$$Bn = 2.29 \text{ Bo} = 2.29 (.03 \text{ fo}).$$ \hspace{1cm} (2.2)

Hence the Q per stage must be less than

$$\frac{fc}{Bn} = 14.$$ \hspace{1cm} (2.3)

Note that with synchronous single tuned stages the overall bandwidth is always less than the stage bandwidth. If the above restraint (2.3)
on $Q$ is too severe, one may then stagger tune the amplifier. In this case, the overall bandwidth is larger than the stage bandwidth; however, to obtain this desirable result more effort must be expended in tuning the amplifier. For a maximally flat response, the circuits of the driver must be tuned\(^7\) as shown in Table II.

<table>
<thead>
<tr>
<th>No. of Circuits</th>
<th>Circuit Center Freq.</th>
<th>Circuit Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>$f_c$ .46 $Bo$</td>
<td>.38 $Bo$</td>
</tr>
<tr>
<td>2</td>
<td>$f_c$ .19 $Bo$</td>
<td>.92 $Bo$</td>
</tr>
</tbody>
</table>

Table II. Design Data for Maximally Flat Staggered Quadruple

The restraints on the stage $Q$'s are now considerably reduced: instead of Equation 2.3, the stage $Q$'s must be

\[ Q_{1,2} \leq 37 \]  
\[ Q_{3,4} \leq 89 \]  

where the stages having these $Q$'s are not required to be in any particular order. Maximal flatness is not very important in this case; hence $Bo$ could be further increased by overstaggering.

Some time after the driver was built, it was decided to drop the requirement for a 3 percent frequency shift, thus reducing the bandwidth requirements of the RF amplifiers. The bandwidth requirements are now set by the maximum drift in the natural frequency of the resonators during warmup. This is because the RF system is first run in a self-oscillatory mode when the system is turned on. Since the natural frequency of the "cold" resonators will differ by less than .5 percent from the frequency of the system during normal operation, the overall bandwidth ($Bo$) of the driver may be conservatively set to .01 $fo$. For a driver consisting of 4 synchronous single tuned circuits, the restraint on the $Q$ per circuit is now reduced to

\[ Q \leq \frac{fo}{2.29 Bo} = 44. \]
Since this restraint is easily met, the present driver amplifier is not
stagger tuned.

2.1.3 Achieving Waveform Purity

The attitude adopted in the design of the CRM RF amplifier system
has been to first eliminate as far as possible all sources which will
introduce noise into the RF chain and then to reduce the remaining noise
using feedback systems. It must be noted that the effect of noise
introduced into the RF chain is cumulative: noise introduced into the
amplifier's initial stages is especially troublesome since it is amplified
by the subsequent stages. With this effect in mind, it was decided that
during normal operation the driver amplifier would be driven well into
saturation. This reduces significantly the output due to noise signals
introduced in stages previous to the main power tube since, when saturated,
the plate current of the driver (4-1000A) is virtually independent of its
grid voltage. Under these conditions the amplitude variations of the RF
output are due almost entirely to sources of noise in the main power
amplifier (4CW250,000). Frequency and phase variations will, of course,
remain unreduced by this method of operation. In order of importance,
the sources of modulation in the CRM RF amplifier are

(a) detuning of resonant circuits
(b) vibration of tube elements (tube microphonics)
(c) power supply ripple
and (d) thermal noise.

These will be now discussed in turn.

A. Detuning of Resonant Circuits

Since the CRM RF chain will be driven at a fixed frequency, a shift
in the resonant frequency of any of the tuned stages will introduce both
phase and amplitude errors into the output. This may be illustrated most
simply by considering a single simply tuned stage for which the gain may
by written in the form

\[ A(j\omega) = \frac{Ar}{1 + jQ(\omega/\omega_0 - \omega_0/\omega)} \]  

(2.6)
where

\( A_r \) = gain at resonance
\( Q \) = stage Q
\( \omega_0 \) = resonant frequency
\( \omega \) = driving frequency.

For small changes in resonant frequency, a fractional change in gain is obtained

\[
\frac{|A(j\omega)|}{A_r} \approx \frac{1}{1 + 2(\delta/B)^2}
\]

(2.7)

where

\( \delta = \omega - \omega_0 \) (change in resonant frequency)
\( B \) = stage bandwidth.

The actual change in gain is

\[
\Delta |A(j\omega)| = -2A_r(\delta/B)^2.
\]

(2.8)

Since the change in gain is proportional to

\[
(\delta/B)^2 = (\delta/\omega)^2 Q^2
\]

a high Q circuit will introduce large deviations in amplitude for small deviations in resonant frequency. The resonators with their very high Q will therefore seriously affect the RF amplitude as their resonant frequency changes. Hence, if no other corrective measures were taken, the fractional change in the resonant frequency of the resonators would have to be controlled so that

\[
\frac{\Delta \omega_r}{\omega_r} < 1.2 \times 10^{-4} \varepsilon
\]

(2.9)

where \( \varepsilon \) is the desired amplitude stability.

The phase shift due to the detuning of a single tuned circuit may be derived from Equation 2.6 as

\[
\Delta \theta \approx -2Q(\Delta \omega_r/\omega_r) \text{ radians}
\]

(2.10)

where \( (\Delta \omega_r/\omega_r) \) is a small fractional change in resonant frequency. Since the resonators have a Q of about 6000, they will be the principal source of phase errors.

Because the magnitude of the RF amplitude and phase errors due to resonator detuning is large, a resonator tuning system is being installed. The rate of change of amplitude and phase will, however, be small, since the errors are mechanical and thermal in origin. This means that the
feedback systems employed (resonator tuning, amplitude and phase feedback) to correct these errors will be able to reduce them below the levels specified in Chapter 1.

B. Tube Microphonics

This form of output distortion (consisting of low frequency amplitude and phase modulation) is caused by the relative motion of the elements of a vacuum tube. In a large power tube such as the 4CW250,000, many complex modes of vibration are possible; in this tube filament motion is most important, being the principal cause of microphonic noise. Since an actual mechanical vibration is involved, anything which jars or vibrates the tube will cause this effect. Typical causes are mechanical vibrations from nearby equipment transmitted to the tube via the floor, the vibration of cooling fans, vibration due to coolant (water) flow, and the vibration of power transformers installed in the cabinet. Filament motion may also be caused by the interaction of their own, or with external, magnetic fields. Typical noise values for the 4CW250,000 are -70 to -82 db.

Since the noise levels quoted for the 4CW250,000 are considerably higher than the total allowable noise level of -94 db, great care must be taken to eliminate the sources of microphonics. That is, the tube must be isolated from sources of mechanical vibration and from external magnetic fields. This requires that the main DC power transformers must not be mounted in the same cabinet as the power tube; coolant flow must not be turbulent and fans must be vibration free. DC filament supplies give further noise reductions, as would mesh filaments, if they were available for the 4CW250,000.

C. Power Supply Ripple

Any variation in the level of plate and bias supply voltages (Figure 12) will cause a change in the amplitude of the output RF voltage; the ripple on the output of these DC supplies therefore must be restricted to very low levels. This is not a serious problem for the driver stages of the RF amplifier chain where the power and voltage requirements are quite low. That is, the DC supplies for these stages may be easily regulated. Saturating the driver further reduces the problem, leaving the PA as the principal
source of ripple-related amplitude modulation.

From Figure 13 it is seen that
\[ \frac{\Delta E_p}{\Delta E_{bb}} \approx 0.75 \]  \hspace{1cm} (2.11)

where
\[ \Delta E_p = \text{change in the output RF amplitude} \]
\[ \Delta E_{bb} = \text{change in the DC plate voltage}. \]

Hence, if an amplitude stability of -100 dB is demanded, then the ripple on the plate supply must be restricted to less than -99 dB. That is, for
\[ \frac{\Delta E_p}{E_p} < \frac{1}{10^5} = -100 \text{ dB} \]  \hspace{1cm} (2.12)

it is required to have
\[ \frac{\Delta E_{bb}}{E_{bb}} < \frac{1.1}{10^5} = -99 \text{ dB} \]  \hspace{1cm} (2.13)

if the plate supply is the only source of modulation. In the CRM amplifier, this criterion is met by a series regulated DC supply.\(^9\) (If regulation was not available, the ripple on the output RF could be reduced by introducing a suitable fixed fraction of the ripple to the grid.) The
Fig. 13. Change in Plate Swing (Ep) for a Change in Plate Voltage (Ebb)
change in output amplitude for a change in screen voltage is shown in Figure 24, which indicates that
\[ \frac{\Delta E_p}{\Delta E_{c2}} \approx 8 \] (2.14)

where \( \Delta E_{c2} \) is the change in screen bias. The stability required of the screen bias supply therefore may be given as
\[ \frac{\Delta E_{c2}}{E_{c2}} \left( \frac{E_p}{E_{c2}} \right) \leq \frac{\varepsilon}{8} \] (2.15)

where \( \varepsilon \) is the allowable fractional change in plate swing. Taking the plate swing and DC screen voltage as 15 KV and .8 KV respectively requires that for
\[ \varepsilon = -100 \text{ db}, \]
\( \frac{\Delta E_{c2}}{E_{c2}} \) must be less than -92.6 db in the absence of other sources of modulation.

A similar calculation for the control grid indicates that the ripple on its DC bias must also be restricted to a level of less than -93 db. The frequencies of the modulation introduced will be the ripple frequencies associated with the DC supplies. These frequencies are rather low; the highest (that of the 6 phase plate supply) is 360 Hz. Since the effectiveness of the amplitude feedback system increases with decreasing frequency (see Figure 34), this system is able to significantly reduce the amplitude modulation caused by ripple on the plate and bias supplies.

An AC filament supply is another source of amplitude modulation. This modulation is the result of
(a) the change in potential along the filament caused by the IR drop of the filament current
(b) the action on the plate current by the magnetic field associated with the filament current
and (c) the change in filament temperature due to a change in the magnitude of the filament current.

The first cause is the most important. In this case, the voltage drop along the filament (cathode) causes this electrode to be biased with respect to ground; if one end of the filament is grounded, then the other will oscillate about ground by an amount equal to the filament supply voltage. This may be partially remedied by placing the ground near the midpoint of the filament. This is most easily done by connecting the
plate return leads to a center-tap on the output of the filament transformer (see Figure 17). The voltage drop in the filament now will cause the negative half of the filament to supply more electrons than the positive half. However, since the current drawn is proportional to the 3/2 power of the plate to cathode voltage, the reduction in current from the positive half of the filament is not equal to the increase in current from the negative half. Consequently, the plate current increases whenever filament current flows; that is, the plate current is modulated at twice the filament supply frequency.

The magnetic field produced by the filament current will deflect the electrons flowing to the plate; consequently, the plate current is decreased slightly whenever filament current flows. Again, the plate current is modulated at twice the filament supply frequency. The effect of the filament magnetic field will tend to cancel the effect of the filament voltage drop; by properly choosing the filament geometry and voltage, the modulation produced by an AC filament will be reduced to a very low level.

The third way in which an AC filament supply may modulate the plate current—through cyclic variations in filament temperature—is far less significant than the first two. This is because the thermal capacity of the filaments in the tubes used is so high that the variation in filament temperature during a cycle is not large enough to significantly affect the plate current. Since this effect depends on the power dissipated in the filament, the consequent modulation need not have a sinusoidal waveform; its frequency will be twice that of the filament supply.

The most effective way to eliminate the filament as a source of modulation is to use DC filament supplies—this will also eliminate the problem of microphonics caused by the interaction of the magnetic fields of the filament strands. DC filament supplies may not be used with some tubes because the voltage drop along the filament would significantly affect the tube's performance (that is, alter the $g_m$ of the tube). In our case, this is not a problem because the filament voltage is less than .1 percent of the plate supply voltage.
D. **Thermal Noise**

At the noise levels of interest at TRIUMF, the most familiar source of random noise—thermal noise—is of no concern. This is illustrated in the following brief calculation in which it is assumed the amplifier consists of two stages (Figure 14).

\[ G_v = \text{voltage gain} \]

\[ \overline{V_o} = \text{RMS thermal noise voltage on resonators} \]

**Fig. 14. Amplifier considered for Thermal Noise Calculation**

The RMS noise voltages generated at the inputs may be calculated using the formula

\[ E = \sqrt{\frac{2kT\beta R}{\pi}} \]  \hspace{1cm} (2.16)

where

- \( E \) = RMS noise voltage generated in the resistance \( R \) (ohms)
- \( k \) = Boltmann's constant = \( 1.38 \times 10^{-23} \) joule/°K
- \( T \) = temperature, °K
- \( R \) = resistance, ohms
- \( \beta \) = noise bandwidth.

For a synchronous single tuned amplifier, the noise bandwidth may be calculated from the formula

\[ \beta = B_n \int_{0}^{\infty} \frac{dx}{(1 + x^2)^n} \]  \hspace{1cm} (2.17)
where
\[ B_n = \text{bandwidth of a single stage}\]
\[ n = \text{number of stages}. \]

This integral may be evaluated by letting \( Z(n) = \frac{\Phi}{B_n} \) and first evaluating \( Z(1) \). (\( Z(1) \) was found to be \( \pi/2 \).) Then for a cascade of \( n \) stages, \( Z(n) \) may be found by recursively evaluating
\[
Z(n) = \left(1 - \frac{1}{2(n-1)}\right)Z(n-1).
\]

Since the overall bandwidth (\( B_0 \)) is given by
\[
B_0 = B_n(2^{1/n} - 1)^{1/2}
\]
the following Table (III) was constructed.

<table>
<thead>
<tr>
<th>( n )</th>
<th>( Z(n) )</th>
<th>( B_n/B_0 )</th>
<th>( \Phi/B_0 = m )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.57</td>
<td>1.0</td>
<td>1.571</td>
</tr>
<tr>
<td>2</td>
<td>0.785</td>
<td>1.55</td>
<td>1.220</td>
</tr>
<tr>
<td>3</td>
<td>0.589</td>
<td>1.96</td>
<td>1.155</td>
</tr>
<tr>
<td>4</td>
<td>0.491</td>
<td>2.30</td>
<td>1.129</td>
</tr>
<tr>
<td>5</td>
<td>0.423</td>
<td>2.59</td>
<td>1.097</td>
</tr>
</tbody>
</table>

Table III. Chart for Noise Bandwidth Calculations

The output noise voltage may now be simply evaluated for noise introduced by \( R_1 \) and \( R_2 \). Since the resonator \( Q (=6000) \) is much larger than all other system \( Q \)'s, this value may be used for both cases. Hence
\[
E = 2f_0 \sqrt{\frac{kTf_{\perp}m}{Q}} = 4.6 \times 10^{-10} \sqrt{TRm}
\]
where
\[ f_0 \approx 2.3 \times 10^7 \text{ Hz} \]
and \( m \) is taken from Table III.
The thermal noise introduced by $R_1$, will be the dominant thermal noise source if the driver is not saturated. In this case, there will be 4 stages in the amplifier—therefore $m = 1.129$. Taking $T = 100\degree \text{C}$
gives
\[ E_1 = 6.68 \times 10^{-8} \text{v.} \quad (2.21) \]
Since there is a voltage gain of about $2 \times 10^5$, the corresponding RMS thermal noise component at the output will have a level of about
\[ V_{o1} = 1.3 \times 10^{-2} \text{v.} \quad (2.22) \]
If the driver is saturated, the principal source of thermal noise will be the $2.5 \text{K} \Omega$ resistance at the input to the main power amplifier. In this case, there is only 1 stage of amplification—therefore $m = 1.571$. Taking $T = 100\degree \text{C}$
gives
\[ E_2 = 5.56 \times 10^{-7} \text{v.} \quad (2.23) \]
Then, due to the PA voltage gain of 166, the RMS thermal noise in the output will have a value of
\[ V_{o2} = 9.2 \times 10^{-5} \text{v.} \quad (2.24) \]
Both $V_{o1}$ and $V_{o2}$ are RMS values; however, even though the instantaneous values of thermally produced noise at the output may reach considerably higher levels, the sources previously discussed still contribute far more to the output noise. Shot and partition noise, the two other sources of random noise in a vacuum tube amplifier, may similarly be disregarded.

**Class C Amplifier Operation (Effect on Waveform)**

Efficiency considerations dictate the use of class C operation in an RF power amplifier. Since in this case the plate current flows in pulses, the output signal will consist of the fundamental (input) frequency plus a high percentage of harmonics as shown in Figure 15. Although the restrictions on the harmonic content of the output are not as severe as those on its modulation (the second harmonic must be reduced to less than .5 percent) the percentage of harmonic content shown in Figure 15 is intolerable.
An interesting feature illustrated by this graph is that for a total conduction angle near 180° the third harmonic content in the output may be almost entirely eliminated, while the second harmonic is reduced to about 45 percent of the fundamental. A more significant reduction in harmonic content is obtained, however, because tuned circuits are used. Consider the simplest case: the single tuned stage. The reduction in gain for the second harmonic may be obtained by substituting \( \omega = 2\omega_0 \) in Equation 2.6:

\[
\left| \frac{A(2\omega_0)}{A(\omega_0)} \right| = \frac{1}{1.5Q} \quad (2.25)
\]

where

\[
Q \geq 10;
\]

\[
\left| \frac{A(2\omega_0)}{A(\omega_0)} \right| = 1.3\%
\]

if

\[
Q = 50.
\]
This equation indicates that sufficiently high attenuation of harmonics may be obtained by using a simple tank circuit with a sufficiently high Q. However, arbitrarily large Q's are not allowed since these would violate the bandwidth requirements given in Section 2.1.2. If greater harmonic attenuation is required, the low pass Pi network shown in Figure 15 may be used. (This network also has good impedance matching abilities.)

![Pi Network Diagram](image)

**Fig. 16. Pi Network Output Circuit**

No problem with harmonic content has been experienced during the operation of the CRM RF system—the transmission line (Figure 8) coupling the 4CW250,000 to the resonators provides sufficient harmonic attenuation.

2.2 Main Power Amplifier

2.2.1 Original Design

A circuit diagram of the first RF power amplifier built for the TRIUMF CRM is shown in Figure 17. This amplifier used a Machlett triode\(^1\) (ML-7560) in a self-biased, grounded cathode configuration. A combination of grid leak and cathode bias was used; the latter provided protection...
Fig. 17. Schematic of Triode (ML-7560) Power Amplifier
should the drive have been accidentally removed. The output circuit was the resonant transmission line discussed in Section 1.3. Neutralization, very important in a triode amplifier, was provided by the coil L₃ connected in parallel with the capacitor C₆. The parallel combination of L₃, C₆ and the plate to grid capacitance (75pF) was to be resonated at the operating frequency to eliminate the plate to grid feedback path. This form of neutralization (coil neutralization) is effective at only one frequency; hence, if the frequency of operation is changed, the neutralizing circuit must be readjusted. Since feedback at other frequencies is only slightly affected, the tendency of an amplifier to act as a tuned plate tuned grid oscillator is not appreciably reduced.

Typical operating conditions of the ML-7560 triode are given in Table IV. These were never fully realized because severe problems with parasitic oscillations caused this design to be abandoned.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC Plate Voltage</td>
<td>20KV</td>
</tr>
<tr>
<td>DC Grid Voltage</td>
<td>-700V</td>
</tr>
<tr>
<td>Plate Swing</td>
<td>17.6KV</td>
</tr>
<tr>
<td>Grid Swing</td>
<td>940V</td>
</tr>
<tr>
<td>DC Plate Current</td>
<td>8.75A</td>
</tr>
<tr>
<td>DC Grid Current</td>
<td>.51A</td>
</tr>
<tr>
<td>Fundamental Plate Current</td>
<td>15.9A (peak)</td>
</tr>
<tr>
<td>Fundamental Grid Current</td>
<td>.94A (peak)</td>
</tr>
<tr>
<td>Input Power</td>
<td>175KW</td>
</tr>
<tr>
<td>Output Power</td>
<td>139KW</td>
</tr>
<tr>
<td>Drive Power</td>
<td>517W</td>
</tr>
<tr>
<td>Efficiency</td>
<td>79%</td>
</tr>
<tr>
<td>Required Load Resistance</td>
<td>1100 Ohms</td>
</tr>
</tbody>
</table>

Table IV. Typical Operating Conditions of the ML-7560
2.2.2 Present Design

The present CRM RF amplifier, designed and built by Continental Electronics, is shown in Figure 18. Although a grounded cathode configuration is used, this amplifier differs considerably from the previous one.

The tube used is a tetrode (Eimac 4CW250,000); the screen grid of this tube reduces substantially the plate to grid feedback capacitance ($C_{pg} \approx 1.2\text{pF}$). The isolation of plate and grid circuits is further increased by placing them in separate compartments with the tube socket mounted in the wall dividing these two compartments. The screen and filament by-pass capacitors are built into this socket.

Even though this amplifier is less susceptible to parasitic oscillations than the previous one, neutralization must still be used; in this case single-ended grid neutralization is employed. To use this method, the input resonant circuit is taken slightly off ground by means of the by-pass capacitor $C$. The voltage across this capacitor, which is out of phase with the grid voltage, is then fed back to the plate via $C_n$ to provide neutralization. Since this is effectively a capacitive bridge circuit (see Figure 19), this form of neutralization is effective for a wide band of frequencies.

![Neutralization Diagram](image)

Fig. 19. Single-ended Grid Neutralization Drawn as a Bridge Circuit
Fig. 18. Schematic of Main CRM Power Amplifier
Separate DC supplies are used to provide the fixed bias shown in Figure 18. Although more expensive, this design has advantages: with fixed bias there is no danger of tube overload should the drive be removed; greater flexibility in setting the DC conditions also results. In addition, the amplitude of the output RF may be varied by adjusting the screen bias. This option, not present in a triode amplifier, was used to provide an amplitude control system utilizing feedback control of the screen bias voltage.

Since the restrictions on the amplifier Q given in Section 2.1.2 were found difficult to meet, a grid swamping resistor ($R_g$) is included in the circuit. To lower the Q of the input circuit to an acceptable level, this resistor must dissipate up to 250 watts; consequently, it is water cooled.

Table V gives the electrical characteristics of the 4CW250,000. Note that the maximum plate dissipation allowed is 250 KW. At 75 percent efficiency and an output of 150 KW, the dissipation will be only 50 KW, well within the tube's rating. Table VI gives typical operating conditions of this tube (in the CRM PA).

<table>
<thead>
<tr>
<th>Filament: Thoriated Tungsten</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum Filament Voltage</td>
</tr>
<tr>
<td>Maximum Filament Current</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Direct Interelectrode Capacitance, Grounded Cathode:</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input</td>
</tr>
<tr>
<td>Output</td>
</tr>
<tr>
<td>Feedback</td>
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<table>
<thead>
<tr>
<th>Maximum Ratings:</th>
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<tbody>
<tr>
<td>DC Plate Voltage</td>
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<tr>
<td>DC Screen Voltage</td>
</tr>
<tr>
<td>DC Plate Current</td>
</tr>
<tr>
<td>Plate Dissipation</td>
</tr>
<tr>
<td>Screen Dissipation</td>
</tr>
<tr>
<td>Grid Dissipation</td>
</tr>
</tbody>
</table>

Table V. Electrical Characteristics of the Eimac 4CW250,000
Table VI. Typical Operating Conditions of the 4CW250,000 Tetrode in the CRM Amplifier

2.3 Driver Amplifier Design

2.3.1 Specifications

In Section 2.1, the RF system requirements regarding power gain, bandwidth and noise reduction were discussed; Table VII summarizes the results applicable to the 3 stage driver amplifier. Note that the power output quoted is the minimum drive required to obtain an output of about 150 KW from the PA tube; "losses" refers to power lost in the coupling circuit and PA grid swamping resistor.
Table VII. Specifications for 3 Stage Driver Amplifier

2.3.2 Tube Selection

The tubes for the driver amplifier were chosen, in order, from the output to the input of the amplifier. Each was chosen on the basis of gain, input-output isolation, and output power. At the time the driver was designed, the PA power tube being used was the Machlett ML-7560; as indicated in Table VII, this tube required considerably more drive power than the present tube being used (Eimac 4CW250,000). On the basis of the drive requirements of the ML-7560, 3 Eimac power tetrodes were considered for the output stage of the driver: 4-250A, 4-400A, and 4-1000A. Tetrodes, rather than triodes, were considered because the former have much lower internal feedback coupling and lower driving power requirements (usually < 1 percent of output power). Assuming a typical efficiency of 75 percent, Table VIII indicates the maximum output power which could be expected of these tubes.
Table VIII. Maximum Output Power of the Tubes considered for the Output Stage of the Driver Amplifier

To help in the choice of the driver amplifier (DA) output tube, an estimate was made of the minimum output necessary to satisfy the Q requirements of this stage. To this end, the circuits in Figure 19 were considered; since the minimum capacitance that may be used in these circuits is the output capacitance of the final DA tube and the input capacitance of the PA, an estimate may be made of the minimum power which must be fed into the coupling circuit to obtain a given Q. The voltage (V) in both cases was taken to be the required peak RF voltage on the grid of the PA.

(a) Tank

(b) Pi

Fig. 20. Power Input to Tank and Pi Circuits
Table IX gives the power that must be fed into the tank circuit if the DA tube is the 4-1000A and the PA tube is either the ML-7560 or the 4CW250,000. (The circuit used is shown in Figure 20 (a), where C is the sum of the tube capacitances.)

<table>
<thead>
<tr>
<th>Q</th>
<th>Required Power (KW)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>ML-7560</td>
</tr>
<tr>
<td>10</td>
<td>1.22</td>
</tr>
<tr>
<td>20</td>
<td>.61</td>
</tr>
<tr>
<td>30</td>
<td>.41</td>
</tr>
<tr>
<td>40</td>
<td>.31</td>
</tr>
<tr>
<td>50</td>
<td>.24</td>
</tr>
<tr>
<td>60</td>
<td>.20</td>
</tr>
<tr>
<td>70</td>
<td>.17</td>
</tr>
<tr>
<td>80</td>
<td>.15</td>
</tr>
</tbody>
</table>

Table IX. Minimum Power Input to DA-PA Coupling Circuit for a given value of Q

If the Pi circuit is used (Figure 20 (b)) with $C_1 = C_o$ and $C_2 = C_i$, then one may write

$$P_{DT} = \frac{C_i}{C_o} P_{DT}.$$  \hspace{1cm} (2.26)

If the ML-7560 is used, $P_D = 25 P_{DT}$; if the 4CW250,000 is used, $P_{DT} = 100 P_{DT}$. Since additional reactance must be added to properly tune the circuits, the actual power required will be at least 1.5 times the figures given in Table IX, but lower than the $P_{DT}$ quote above. If a 3 percent shift in drive frequency must be handled by a synchronous single tuned driver, only the 4-1000A will adequately meet the power requirements. Even though the 3 percent frequency shift was discarded, the 4-1000A is still a good choice because it can easily satisfy the present power requirements without being run near its maximum ratings; also, if more power is
required, there is a considerable reserve available.

Before a tube to drive the 4-1000A can be chosen, the operating conditions of the 4-1000A must be calculated. This information is also necessary for the design of the bias and coupling circuitry. The analysis of a class C amplifier tube is semigraphical, utilizing the fact that the operating line of the constant current curves of the given tube is a straight line. This line is straight because a tuned load is used. That is, if the voltage on the grid of the tube is

\[ E_{gg} = E_{cl} + E_g \cos \omega t, \]  

(2.27)

then the voltage on the plate of the tube will be

\[ E_{pp} = E_b - E_p \cos \omega t \]  

(2.28)

where

\[ E_{gg}, E_{bb} = \text{instantaneous grid and plate voltages} \]
\[ E_{cl} = \text{grid bias} \]
\[ E_b = \text{DC plate voltage} \]
\[ E_g = \text{grid swing} \]
\[ E_p = \text{plate swing}. \]

On the constant current characteristics of the tube (see Figure 20), the relationship between \( E_{gg} \) and \( E_{pp} \) then appears as a straight line:

\[ E_{gg} = \left( - \frac{E_g}{E_p} \right) E_{pp} + \left( E_{cl} + \frac{E_g E_b}{E_p} \right). \]  

(2.29)

The actual analysis consists of specifying an operating line and then calculating the desired quantities from data points on this line. To specify the operating line for a given screen bias, the grid and plate bias plus grid and plate swing must be fixed. Points on the operating line are then found that are 15° (conduction angle) apart. Using the values of voltage and current obtained from these points, an approximate harmonic analysis is made of the plate, control grid, and screen grid currents; that is, the magnitude of the DC and fundamental (driving frequency) components are found. The operating line given for the 4-1000A in Figure 21 shows this tube operating near maximum output; to find the DC and fundamental current the equations

\[ I_{DC} = \frac{(.5A + B + C + D + E + F)}{12} \]  

(2.30)
A, B, C, D, E, F, G are 15° apart.

Fig. 21. Operating Line of the 4-1000A
and

\[ I_{pp} = \frac{(A + 1.93B + 1.73C + 1.41D + E + .52F)}{12} \]

are used, where

- \( I_{DC} \) = DC component of the plate current pulse
- \( I_{pp} \) = peak fundamental component of the plate current pulse

and \( A, B, C, D, E, F \) = the instantaneous values of current taken from the operating line.

This analysis is done for the plate current to find the power output and plate dissipation; for the grid current to find the drive power and grid dissipation, and for the screen current to find the screen dissipation. The process is repeated until an operating line is found that gives the desired output power with a high efficiency (> 75 percent) and without exceeding the maximum ratings of the tube. Then, in order for the tube to operate as calculated, the given bias and drive must be provided, as well as the proper load impedance for the tube, where the load impedance is given by

\[ R_L = \frac{\text{Peak fundamental voltage}}{\text{Peak fundamental current}} \]

(2.31)

The operating conditions calculated for the 4-1000A based on the operating line in Figure 21 are given in Table X; Table XI gives the electrical ratings for this tube.
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC Plate Voltage</td>
<td>5 KV</td>
</tr>
<tr>
<td>DC Grid Voltage</td>
<td>-200 V</td>
</tr>
<tr>
<td>DC Screen Voltage</td>
<td>500 V</td>
</tr>
<tr>
<td>Plate Swing</td>
<td>4.4 KV</td>
</tr>
<tr>
<td>Grid Swing</td>
<td>350 V</td>
</tr>
<tr>
<td>DC Plate Current</td>
<td>.71 A</td>
</tr>
<tr>
<td>DC Grid Current</td>
<td>.04 A</td>
</tr>
<tr>
<td>DC Screen Current</td>
<td>.12 A</td>
</tr>
<tr>
<td>Fundamental Plate Current</td>
<td>1.24 A (peak)</td>
</tr>
<tr>
<td>Fundamental Grid Current</td>
<td>.08 A (peak)</td>
</tr>
<tr>
<td>Input Power</td>
<td>3.57 KW</td>
</tr>
<tr>
<td>Output Power</td>
<td>2.73 KW</td>
</tr>
<tr>
<td>Drive Power</td>
<td>13.1 W</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>.84 KW</td>
</tr>
<tr>
<td>Grid Dissipation</td>
<td>5.1 W</td>
</tr>
<tr>
<td>Screen Dissipation</td>
<td>61.5 W</td>
</tr>
<tr>
<td>Efficiency</td>
<td>76%</td>
</tr>
<tr>
<td>Required Load Resistance</td>
<td>3550 Ohms</td>
</tr>
</tbody>
</table>

Table X. Operating Conditions for the Eimac 4-1000A near maximum Output
Filament: Thoriated Tungsten

- Maximum Filament Voltage: 7.5 V
- Maximum Filament Current: 22.7 A

Direct Interelectrode Capacitance, Grounded Cathode:

- Input: 32.4 pF
- Output: 9.4 pF
- Feedback: 0.35 pF

Maximum Ratings:

- DC Plate Voltage: 6.0 KV
- DC Screen Voltage: 1.0 KV
- DC Plate Current: 0.7 A
- Plate Dissipation: 1.0 KW
- Screen Dissipation: 75 W
- Grid Dissipation: 25 W

Table XI. Electrical Characteristics of the Eimac 4-1000A

The tube chosen to drive the 4-1000A is an RCA 6146A (see Table XII). This is a beam power pentode with an indirectly heated cathode; consequently, this tube provides high stage isolation and will be free of filament hum. As Table XII shows, this tube is capable of an output power of 48 watts which is considerably larger than the maximum drive requirement of the 4-1000A (≈ 13 W). However, even if the drive requirement of the 4-1000A is very small, a significant amount of power must be dissipated in the coupling circuitry to satisfy the driver Q requirements. For example, if only the tube capacities are used, the tank circuit of Figure 20 (a) must dissipate about 18 watts if a stage Q of 10 is to be achieved. Since the 4-1000A requires less than 18 watts of drive under normal operating conditions, a resistor must be added to lower the Q by dissipating the excess power.

Typical operating conditions for the 6146A are shown in Table XIII.
Filament Voltage (AC/DC) 6.3 V
Filament Current 1.25 A
Transconductance 7000 $\mu$ mhos

Direct Interelectrode Capacitances:
- Input 13 pF
- Output 8.5 pF
- Feedback .24 pF

Maximum Ratings (CCS):
- DC Plate Voltage 600 V
- DC Screen Voltage 250 V
- DC Grid Voltage -150 V
- DC Plate Current 140 mA
- DC Grid Current 3.5 mA
- Plate Dissipation 67.5 W

Table XII. Electrical characteristics of the RCA 6146A Beam Power Pentode

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC Plate Voltage</td>
<td>500 V</td>
</tr>
<tr>
<td>DC Screen Voltage</td>
<td>170 V</td>
</tr>
<tr>
<td>DC Grid Voltage</td>
<td>-66 V</td>
</tr>
<tr>
<td>Grid Swing</td>
<td>84 V (peak)</td>
</tr>
<tr>
<td>DC Plate Current</td>
<td>135 mA</td>
</tr>
<tr>
<td>DC Screen Current</td>
<td>9 mA</td>
</tr>
<tr>
<td>DC Grid Current</td>
<td>2.5 mA</td>
</tr>
<tr>
<td>Driving Power</td>
<td>.2 W</td>
</tr>
<tr>
<td>Output Power</td>
<td>48 W</td>
</tr>
<tr>
<td>Efficiency</td>
<td>70%</td>
</tr>
<tr>
<td>Required Load Resistance</td>
<td>1500 Ohms</td>
</tr>
</tbody>
</table>

Table XIII. Typical Operating Conditions of the RCA 6146A
The maximum drive power required by the 6146A (.2 W) is much larger than the available power output from the frequency synthesizer (5 mW); however, under normal operating conditions (output about 18 W), the 6146A will require only a few milliwatts of drive. This would seem to indicate that the synthesizer could normally be used to drive the 6146A. However, this cannot be done if the stage Q requirements are to be met; that is, if the input circuit of the 6146A is a tank circuit using only the input capacitance of this tube it is found that

\[ Q \text{ Pin} = 2.3 \]  \hspace{1cm} (2.32)

where

\[ \text{Pin} = \text{power dissipated in grid circuit of the 6146A} \]
\[ C_{in} = 13 \text{ pF} \]
and \[ V_g = 50 \text{ V (peak)} = \text{RF drive voltage required}. \]

Hence, for a \( Q = 10 \), the synthesizer would have to supply 230 mW; for a \( Q = 70 \), the synthesizer would have to supply 14 mW. Since the synthesizer is capable of only 5 mW, another stage must be added.

An RCA 6CL6 power pentode (Table XIV) was chosen to drive the 6146A. Although this tube has an input capacitance (11 pF) that is almost as large as that of the 6146A, the 6CL6 requires an RF grid voltage of less than 1.5 volts (peak). Because the energy storage in the input capacitance depends on this voltage squared, it is found that

\[ Q \text{ Pin} < 1.8 \times 10^{-3} \]  \hspace{1cm} (2.33)

where

\[ \text{Pin} = \text{minimum power to be dissipated in the grid circuit of the 6CL6} \]
and \( Q = \text{stage Q required} \).

Consequently, the drive requirements of this stage may easily be met by the synthesizer: a \( Q \) of 10 requires a power dissipation of only .2 mW.
Filament Voltage (AC/DC) 6.3 V
Filament Current .65 A
Direct Interelectrode Capacitance (without external shield):
  Input 11 pF
  Output 5.5 pF
  Feedback .12 pF
Maximum Ratings:
  DC Plate Voltage 300 V
  DC Suppressor Grid Voltage 0 V
  DC Screen Voltage 300 V
  DC Grid Voltage -50 V
  DC Plate Current 30 mA
  DC Screen Current 7.2 mA
  Plate Dissipation 7.5 W
Transconductance $1.1 \times 10^{-2}$ mhos

Table XIV. Electrical Characteristics of the 6CL6 Power Pentode

2.3.3 Driver RF Circuits

A circuit diagram of the present driver amplifier is shown in Figure 22; Plates 1 to 3 show the top, front and bottom of the driver with covers removed.

The circuitry of the driver must provide proper interstage coupling; that is, it must
(a) present the proper load impedance to each tube
(b) provide the required RF drive voltages
and (c) transfer power efficiently. (This is imperative in the output stages where the power generated is large.)

This circuitry must also sufficiently attenuate the harmonics of the drive frequency that are generated by Class C operation of the tubes. Any tendency of the amplifier to break into spontaneous oscillation is, of course, intolerable. To prevent this, neutralizing circuits are used to reduce plate to grid coupling, while parasitic suppressors are added to
Fig. 22. Schematic of CRM Driver Amplifier
load the grid and plate circuits for frequencies much higher than the operating frequency.

Since the ability to function with a 3 percent shift in driving frequency is no longer required of the CRM RF system, the driver consists of synchronous single tuned stages with circuit Q's less than 44. Since the stage requirements are not all the same, several different types of single tuned circuits are utilized.

The output circuit of the 4-1000A is a modified Pi\textsuperscript{14,15} circuit. Initially, a tank circuit with loop coupling was tried to match this tube to the 50 \R transmission line. It was found that when the loop was coupled loosely enough to provide the proper load impedance, the Q became excessively high. A Pi circuit was then built. When properly matched, it had a Q of about 30. This Q satisfied the bandwidth requirements, but the circulating current (equal to Q times the input current) heated the tank circuit excessively. The present output circuit is a modification of the Pi. By tapping the inductor as shown, inductance was added in series with the plate, hence lowering the input shunt capacitance of the Pi. This lowered the Q and reduced the heating of the inductor. To reduce the chance of parasitic oscillations, the parasitic suppressor (PS2) was added.

The circuit coupling the 6146A to the 4-1000A was originally a Pi circuit. It was to have been used in conjunction with the 4-1000A in a grounded screen configuration which it was hoped would eliminate the need for neutralization. Because long leads and floating supplies were required, the isolation of the 4-1000A's plate and grid was less than in a conventional grounded cathode configuration. Since the floating supplies presented a safety problem, and since the need for neutralization had not been eliminated, the output stage was changed to its present grounded cathode configuration. In addition, the grid circuit of the 4-1000A was changed to a transformer-coupled (tuned secondary) configuration\textsuperscript{16} in order to simplify the neutralization of this tube. This was accomplished by raising the secondary of the transformer off ground by means of a 1000 pF capacitor (Cng) and then coupling this point to the plate using capacitor Cn. (Cn is a section of foil placed next to the air chimney surrounding the 4-1000A.) The 5.6 K \Omega resistor placed across the
secondary of the transformer lowers the Q of this stage to the required level.

The coupling between the 6CL6 and the 6146A consists of a Pi circuit with resistive loading at the grid of the 6146A. A Pi circuit is used not because a wide impedance transformation is required, but because these two tubes are relatively far apart. By using a Pi circuit and placing the required capacitors near the tubes, the inductance of the lead is incorporated in the inductor of the Pi.

Transformer coupling utilizing a tuned secondary is used at the input to the 6CL6. To minimize the reactance in this circuit and therefore minimize the power dissipation required to attain a given Q, no capacitance is added to the input circuit (input capacitance of the 6CL6 $\approx 11$ pF). Resonance is obtained by slug tuning the transformer.
Plate I. Top View of Driver Amplifier (Cover Removed) showing the 4-1000A and its Output Circuit; the Input to the Transmission Line is on the Right-hand Side
Plate II. Front View of Driver Amplifier (Cover Removed) showing the 6CL6 and its Associated Circuitry
Plate III. Bottom View of Driver Amplifier (Cover Removed) showing the Filament Chokes and RF Transformer Coil; Filament Transformer on the Right
CHAPTER 3. AMPLITUDE CONTROL SYSTEM DESIGN

Two amplitude control systems were designed and built for the TRIUMF CRM RF system. For use when the driver amplifier is not saturated, a drive control system is used which controls the amplitude of the output RF by appropriately modulating the synthesizer signal. When the driver is saturated, this form of control is not possible—in this case, the resonator RF amplitude is controlled by adjusting the screen grid voltage on the 4CW250,000.

3.1 System Element Transfer Functions

Before the amplitude control systems may be analyzed as a whole, the input-output relationships of all system components must be found. Briefly, these are

(a) transmission line-resonator system
(b) RF chain
(c) resonator voltage detector
(d) input RF amplitude modulator
(e) Hewlett-Packard programmable power supply
(f) PA plate amplitude response to PA screen voltage modulation
and (g) RF by-pass circuit on the screen of the PA.

It was assumed that all other components to be added would be broadband enough so that their frequency response could be ignored.

A. Transmission Line-Resonator System

The input-output amplitude relationship of the transmission line-resonator system (1 dee) was found using the lumped parameter circuit of Figure 23 to model the actual distributed system assuming a fixed source frequency, \( \omega_0 \).
To find the output amplitude response to a small amplitude modulation at the input to the transmission line, the input was taken to be

\[ V_g(t) = \text{Im} \left( V (1 + ku(t))e^{j\omega_0 t} \right) \]  

(3.1)

where

- \( V \) = initial steady state amplitude
- \( k \ll 1 \) is the fractional increase in input amplitude as a result of step amplitude modulation.

The output is then of the form

\[ \frac{V_{\text{res}}}{V_g} = \text{Im} \left( (a(t)e^{j\phi(t)})e^{j\omega_0 t} \right) \]  

(3.2)

where \( a(t)e^{j\phi(t)} \) is the complex resonator voltage amplitude transfer function. It was found

\[ a(t)e^{j\phi(t)} = \gamma R_o C_o + \delta (k'_1 C'_1 + k'_2 C'_2) + j(\gamma R_o S_o + \delta (R'_1 S'_1 + R'_2 S'_2)) \]  

(3.3)

where

- \( \delta = ku(t) \)
- \( \gamma = 1 + \delta \)
- \( C_o = \cos \theta_o \)
- \( S_o = \sin \theta_o \)
- \( C_1 = \cos ((b_1 - \omega_0)t + \theta_1) \)
- \( S_1 = \sin ((b_1 - \omega_0)t + \theta_1) \)
- \( C_2 = \cos ((b_2 - \omega_0)t + \theta_2) \)
- \( S_2 = \sin ((b_2 - \omega_0)t + \theta_2) \)
The constants $\theta_0$, $\theta_1$, $\theta_2$, $a_1$, $a_2$, $R_0$, $R_1$, $R_2$ cannot be given explicitly in terms of the equivalent circuit parameters, since their determination involved a computer solution of the circuit's fourth order characteristic polynomial.

From the computer solution, it was found that $\theta_0$, $\theta_1$, $\theta_2 \ll 1$ and $b_1 \sim \omega_0$, $b_2 \sim \omega_0$; consequently, a very good approximation would be to take

$$C_0, C_1, C_2 \sim 1$$
$$S_0, S_1, S_2 \sim 0.$$  

It was found by comparing the actual and approximate solutions for the amplitude response that the error varied from about $1.5 \times 10^{-4}\%$ at $t = 0$ to about $2 \times 10^{-7}\%$ at $t = 50$ $\mu$sec. when the input was a step. Hence, one may take the amplitude response to be

$$\Delta a_{RES}(t) = kV\rho_0(1 + A_1e^{a_1t} + A_2e^{a_2t}) u(t) \quad (3.4)$$

where

$$k \ll 1 = \text{input step amplitude}$$
$$a_1 = -2.00 \times 10^4 \text{ sec}^{-1}$$
$$a_2 = -5.23 \times 10^5 \text{ sec}^{-1}$$

Time constants from computer solution.

Taking the Laplace transform this becomes

$$\Delta a_{RES}(S) = kV\rho_0\left(\frac{1}{S} + \frac{A_1}{S-a_1} + \frac{A_2}{S-a_2}\right) \quad (3.5)$$

$$R_0 = \text{Steady State Resonator Amp.}$$
$$\text{Corresponding Steady State Input Amp.}$$

$A_1, A_2$ are the relative amplitudes of the two exponential decay terms.

To satisfy the initial conditions (that is, $\Delta a(t) = 0$ at $t = 0$) one must have

$$1 + A_1 + A_2 = 0 \quad (3.6)$$
$$A_1 = -1.0417 \text{ (from computer solution).}$$

The small signal input to output amplitude response function $T(S)$ may now be found for the transmission line-resonator system.

$$T(S) = \frac{\Delta a_{RES}(S)}{\Delta a_{in}(S)} = \frac{R_0(1 + S\kappa_2)}{(1 + S\kappa_0)(1 + S\kappa_1)} \quad (3.7)$$
where

\[ \omega_0 = 1.966 \times 10^4 \approx 2 \times 10^4 \text{ sec}^{-1} \]
\[ \omega_1 = 5.227 \times 10^5 \approx 5.2 \times 10^5 \text{ sec}^{-1} \]
\[ \omega_2 = 7.0 \times 10^6 \text{ sec}^{-1}. \]

B. RF Chain

The amplitude response of the RF chain was estimated by examining the step response of the system when the PA was loaded by the transmission line which was terminated by a 50 ohm dummy load. It was found that the output, measured at the dummy load, had a rise time of about \( 3 \times 10^{-6} \) seconds. This then corresponds to a \( Q \) of about 200—a reasonable figure for the overall RF system. That is,

\[ Q = \frac{\omega_0 \gamma}{2} \approx \frac{2\pi \times 22 \times 10^6 \times 3 \times 10^{-6}}{2} \approx 200. \]  \hspace{1cm} (3.8)

Consequently, the transfer function for the RF chain was taken to be

\[ \frac{\Delta a_{\text{out}}}{\Delta a_{\text{in}}} = \frac{K}{1 + \gamma S} = \frac{K}{1 + S/3.3 \times 10^5}. \]  \hspace{1cm} (3.9)

where

\[ K = \text{ratio of output to input steady state amp.} \approx 2 \times 10^4 \]
\[ \gamma = \text{time constant} \approx 3 \times 10^{-6} \text{ sec}. \]

C. Resonator Voltage Detector

The dominant time constant of this device is determined by the R-C combination in the cathode circuit of the detecting tube. We have a time constant

\[ \gamma = RC = 180 \text{ pF} \times 100 \text{ K} = 1.8 \times 10^{-5} \text{ sec}. \]

Therefore, the detector transfer function may be taken as:

\[ T(S) = \frac{KD}{1 + 1.8 \times 10^{-5} S} \approx \frac{KD}{1 + S/6 \times 10^4}. \]  \hspace{1cm} (3.10)

where

\[ KD = 10^{-3} = \frac{\text{Steady State Detector Output}}{\text{Steady State Resonator Voltage}}. \]
D. Input RF Amplitude Modulator

This function is performed by a Hewlett-Packard double balanced mixer (Model 10514A). Since this device has a bandwidth in excess of 50 MHz, its frequency characteristic will be disregarded in the following calculations.

E. Hewlett-Packard Programmable Power Supply

The frequency characteristics of this device were taken from its service manual. Its half-power frequency is given as 20 KHz with a voltage gain variable from 0 to -10. Therefore, its transfer function may be written as:

\[ T(S) = \Delta V_{\text{HP}} = -\frac{K_{\text{HP}}}{AV_{\text{in}}} \frac{1}{1 + S/a} \]  \hspace{1cm} (3.11)

\[ a = 2\pi \times 20 \times 10^3 \text{ rad. sec.} \approx 1.25 \times 10^5 \text{ rad. sec.} \]

\[ K_{\text{HP}} = \text{gain (taken as 10 in all following calculations)}. \]

F. PA Plate Amplitude Response to PA Screen Voltage Modulation

It is necessary to know \( \Delta V_p \) if a screen modulating system is to be analyzed. \( \Delta V_p \) = change in the RF amplitude at the plate of the 4CW250,000 in response to a change in the screen voltage of this tube. The time constant of such a change will be extremely short and may therefore be neglected. The magnitude of \( \Delta V_p \) was estimated from the constant \( \Delta V_{g_2} \) current characteristics of the Eimac 4CW250,000 tube by drawing load lines corresponding to the same operating conditions on curves for different screen voltages. That is, load lines were found for the different screen voltages of 400, 800, 1000 and 1200 volts (the characteristics available at that time). These load lines had the same load resistance (RL), DC plate voltage (Eb), control grid bias (Ec1) and grid drive (Eg).

\[ \text{RL} = 690 \ \Omega \]
\[ \text{Eb} = 18,000 \text{ volts} \]
\[ \text{Ec}_1 = 500 \text{ volts} \]
\[ \text{Eg} = 600 \text{ volts peak} \]
$R_L = 690 \Omega$
$E_g = 600 \text{ V (peak)}$
$E_b = 18000 \text{ V}$
$E_C = 500 \text{ V}$

{BIAS}

Fig. 24. Plot of the Output RF Amplitude for varying Screen Voltages (4CW250,000 Power Tube)
The results obtained are shown on Figure 24 where one sees that \( V_p \) is an approximately linear function of \( V_{g2} \) and that \( \frac{\Delta V_p}{\Delta V_{g2}} \approx 8 \).

**G. RF By-Pass Circuit on Screen of PA**

Figure 25 below shows the RF by-pass circuit which connects the modulated DC screen supply to the screen of the 4CW250,000 power amplifier tube. \( RD \) represents the dynamic resistance of the screen. That is,

\[
RD = \frac{\Delta V_{g2}}{\Delta I_{g2}}
\]

(3.12)

![RF By-pass Circuit Diagram](image)

Fig. 25. RF By-pass Circuit at the Screen of the 4CW250,000

Using the same load lines as those used for finding \( \Delta V_p \), it was possible to find \( \Delta V_{g2} \). However, from the nature of the curves it was possible to find \( I_{g2} \) for \( V_{g2} \) equal to only 1000 and 800 volts. Thus

\[
\begin{array}{c|c}
V_{g2} & I_{g2} \\
1000 & .95A \\
800 & .48A \\
\end{array}
\]

Therefore,

\[
RD = \frac{\Delta V_{g2}}{\Delta I_{g2}} \approx \frac{1000 - 800}{.95 - .48} = 425.5 \Omega
\]

(3.13)

For the purpose of these calculations, \( RD \) was consequently taken to be 425\( \Omega \). The frequency response (db magnitude and phase) of the screen circuit was then found using the Ecap computer program. A plot of the results is shown on Figure 26, from which one sees that if a control system with a phase crossover frequency below \( 5 \times 10^5 \text{ rad/sec} \) is adequate, then this circuit may be ignored in control calculations.
Fig. 26. Frequency Response of the RF By-pass Circuit at the Screen of the 4CW250,000
3.2 Drive Control System

A. Uncompensated Open Loop Response

Having found the transfer functions of the components of the proposed drive control system, one has a system as shown in Figure 27.

\[
\begin{align*}
\text{RF CHAIN} & : \frac{K_1}{1 + S/3.3 \times 10^5} \\
\text{MODULATOR} & : K_3 \\
\text{TL-RES SYSTEM} & : \frac{K_2(1 + S/7 \times 10^6)}{(1 + S/2 \times 10^4)(1 + S/5.2 \times 10^5)} \\
\text{VOLTAGE DETECTOR} & : \frac{10^{-4}}{1 + S/6 \times 10^4}
\end{align*}
\]

Fig. 27. Block Diagram of Uncompensated Drive Control System

Note that the output of the voltage detector has been decreased by a power of 10 to make its output compatible with the DAC voltage reference. From Figure 27, one may find the open loop transfer function of the system:

\[
GH = \frac{K(1 + S/7 \times 10^6)}{(1 + S/2 \times 10^4)(1 + S/6 \times 10^4)(1 + S/3.3 \times 10^5)(1 + S/5.2 \times 10^5)}
\]

(3.14)

where \((K)\) is the DC gain of the loop. The Bode plot of equation 3.14 is given in Figure 28. Since positive phase and gain margins in most cases ensure a stable closed loop system, we see that, based on this criterion, the maximum stable open loop gain of the system will be 40 db - 18 db = 22 db.

A check of this stability limit may be made by applying the Routh stability criterion using, however, an approximation of \((GH)\) to simplify calculations.
Fig. 28. Open Loop Frequency Response of the Uncompensated Drive Control System
Let
\[ GH = \frac{K}{(1 + S/2 \times 10^4)(1 + S/6 \times 10^4)(1 + S/3.3 \times 10^5)} \]  \hspace{1cm} (3.15)

Then the characteristic equation of the system, \( 1 + GH = 0 \) becomes
\[ S^3 + (4.1 \times 10^5)S^2 + (2.76 \times 10^{10})S + (3.69 \times 10^{14}) + K^1 = 0 \]
\[ K^1 = 3.96 \times 10^{14} K \]

### Table XV. Routh Table for the Approximate Characteristic Equation

<table>
<thead>
<tr>
<th>( S^3 )</th>
<th>1</th>
<th>2.76\times10^{10}</th>
</tr>
</thead>
<tbody>
<tr>
<td>( S^2 )</td>
<td>4.1\times10^5</td>
<td>3.69\times10^{14} + K^1</td>
</tr>
<tr>
<td>( S^1 )</td>
<td>2.76\times10^{10} - .09\times10^{10} - .24\times10^{-5}\times K^1</td>
<td>0</td>
</tr>
<tr>
<td>( S^0 )</td>
<td>3.69\times10^{14} + K^1</td>
<td>0</td>
</tr>
</tbody>
</table>

The characteristic equation represents a stable system (having roots with only negative real parts) if the signs of all entries of the first row of the Routh table (Table XV) are the same. Since \( K^1 > 0 \), this requires that
\[ 2.76 \times 10^{10} - .09 \times 10^{10} - .24 \times 10^{-5} K^1 > 0 \] \hspace{1cm} (3.16)
or
\[ K^1 < 28 = 28.65 \text{ db} \]

This confirms the value 22 db as the maximum stable gain of our system.

The final value theorem (Eqn 3.17) may now be used to estimate the reduction of resonator voltage errors by the uncompensated drive control system.
\[ f(\infty) = \lim_{t \to \infty} f(t) = \lim_{s \to 0} s F(s) \] \hspace{1cm} (3.17)

If a unit step error \( (1/S) \) should occur in the output, we may write
\[ e(\infty) = \lim_{s \to 0} s \left\{ \frac{G_1}{1 + G_1 H_1} \right\} = \lim_{s \to 0} \frac{G_1}{1 + G_1 H_1} = \frac{1}{1 + K} \] \hspace{1cm} (3.18)

where our actual system configuration is as shown in Figure 29.
Fig. 29. Block Diagram of Drive Control System with Step Error in the Output

\[ G_1 = 1 \]
\[ H_1 = GH \text{ (system open loop transfer function)} \]

Therefore, since \( K = 22 \text{ dB} \approx 12.6 \), step errors in the output will be reduced by a factor of \( \frac{1}{1 + 12.6} = 0.074 = -22.6 \text{ dB} \).
B. **Compensated Open Loop Response**

Since as large as possible a reduction of the amplitude errors is desired, an increase in the open loop gain is required. However, increasing the system gain above 22 db would lead to an unstable system unless:

(a) the time constants of the system are decreased at the same time or (b) compensating elements are added to stabilize the system.

An examination of the RF chain, transmission line-resonator system, and voltage detector shows that the system time constants can not be decreased significantly. Hence, compensation—either parallel or cascade—must be used in this case. Cascade compensation placed between the reference comparator and the input modulator is by far the most practical choice.

Two cascaded phase lag circuits having a transfer function

\[
T(S)_{\text{comp}} = \frac{Kc (1 + S/2 \times 10^4)(1 + S/6 \times 10^4)}{(1 + S/10^3)(1 + S/6 \times 10^3)}
\]  

were tried first as compensation. From bode plot considerations this limited the system gain to less than 42 db and gave a phase crossover frequency of \(3.0 \times 10^4 \text{ rad} \). This would reduce the response to a step error input by about 42 db, an improvement of about -20 db over the uncompensated case. However, the phase crossover frequency had been lowered—this was not desirable since it would reduce the effectiveness of this system when confronted with modulation in the kilohertz range.

A better compensator would be constructed using phase lag-lead compensation (note the pole cancellation as before) of the form:

\[
T(S)_{\text{comp}} = \frac{Kc (1 + S/2 \times 10^4)(1 + S/6 \times 10^4)}{(1 + S/a_0)(1 + S/a_1)}
\]  

letting

\[
a_0 = 10^3 \text{ rad/sec.}
\]

\[
a_1 = 6 \times 10^5.
\]

The frequency response of this compensator is shown on Figure 31; the open loop response of the drive control system using this compensation is given on Figure 32. The latter plot indicates that the maximum stable
Fig. 31. Frequency Response of the Lag-Lead Compensator
Fig. 32. Open Loop Response of the Compensated Drive Control System
system gain was then $64 \text{ db} - 15 \text{ db} = 49 \text{ db}$, a considerable increase over lag-lag compensation (simulation shows the limit is $64 - 11 = 53 \text{ db}$). In addition, the phase crossover frequency had been raised to $2.7 \times 10^5 \text{ Hz}$, an important gain.

It must be noted that the system gain in these two compensating schemes depended upon the compensator's lowest break point frequency which was arbitrarily chosen to be $10^3 \frac{\text{rad.}}{\text{sec.}}$. In both cases, the position error constant (DC error) could be improved indefinitely by lowering this frequency and increasing the gain. This would not, however, improve the system response to amplitude errors having a frequency larger than about $10^4 \frac{\text{rad.}}{\text{sec.}} \approx 1.6 \text{ KHz}$. The choice of $a_1$ as $(6 \times 10^5 \frac{\text{rad.}}{\text{sec.}})$ is approximately the maximum possible frequency in terms of ease of implementation and usefulness.

C. Simulation Results

It was deemed useful to examine the system closed loop frequency and time response. Since closed loop calculations with anything other than the simplest systems are extremely tedious, it was decided to digitally simulate an electrical analog of this control system using the Ecap circuit analysis program. Each system block was first modelled separately, its Ecap frequency response being checked against its Bode plot. Once all function blocks were correctly modelled individually, the system was then modelled by interconnecting, but not loading, the function blocks by means of the Ecap dependent current source ($T_{ij}$).
Fig. 33. Circuit used for the ECAP Simulation of the Drive Control System
i. **Frequency Response**

Using the circuit of Figure 33, the open loop response of the drive control system was found and compared to the equivalent Bode plot (see Figure 32), showing (as expected) that the Bode plot is an accurate approximation except at system break points. The system closed loop frequency response was then found for a loop gain of 44 db (10 db gain margin, 40° phase margin) for a unit error input to the plate of the PA. This was done to examine the effect of amplitude modulation introduced by this stage.

The output for the control loop both open and closed is shown on Figure 34. Note that the F-B system ceases to provide any useful error reduction at about $10^6 \text{ rad.} = 160 \text{ KHz.}$

ii. **Time Domain Response**

The time domain response of the compensated drive control system to a unit step error input to the plate of the PA was found using the Ecap transient response program. The results are shown in Figures 35, 36, and 37. Note that without F-B, the output would be an exponential rise to $(6.6) \text{ Vin}$ with a time constant of $50 \mu \text{sec.}$

Figure 35 shows a stable system response corresponding to a 10 db gain margin, 40° phase margin, and a unit step error input. Because of the nature of our amplitude tolerances, a smooth response of this kind is less useful than the response shown on Figure 36, which is due to a system gain close to the stability limit. However, when this limit is exceeded, an oscillation with an exponentially increasing amplitude, such as shown on Figure 37, results. The choice of system gain must therefore be such as to place the system near the stability limit, but not so close that the amplifier drift could cause the system to become unstable.
Fig. 34. Frequency Response of Drive Control System with Error Input at the Plate of the PA
Fig. 35. Plot of Change in Resonator Voltage Amplitude after an Initial Unit Step Change in RF Amplitude at the PA Plate (Drive Amplitude Control used; FB Amp. has 24.1 db of Gain)
Fig. 36. Plot of Change in Resonator Voltage Amplitude after an Initial Unit Step Change in RF Amplitude at the PA Plate (Drive Amplitude Control used; FB Amp. has 32.1 db of Gain)

$\Delta A = \text{Change in Amplitude at the Resonator}$
Fig. 37. Plot of Change in Resonator Voltage Amplitude after an Initial Unit Step Change in RF Amplitude at the PA Plate (Drive Amplitude Control used; FB Amp. has 240 db of Gain)
3.3 Screen Modulating System

The proposed screen modulating system is shown in Figure 38 below.

![Block Diagram of Proposed Screen Modulating System](image)

Note that two components of this system—the resonator voltage detector circuit and the transmission line-resonator system—are common to both amplitude control schemes. This indicates that the compensating amplifier used in the drive control system might work equally well as the F-B amp. in the screen modulating system. This is apparent when it is noted that the proposed compensation was used to cancel the pole contributed by the voltage detector and also the lower pole due to the transmission line-resonator circuit. The use of a common F-B amplifier considerably simplifies the amplitude control system.

A. Screen Modulator

Screen voltage modulation is accomplished as shown in Figure 39, where the variac is used to set the DC screen voltage of the 4CW250,000. For purposes of control calculations, this arrangement is shown more simply as in Figure 40, where the DC screen voltage is considered constant and hence does not enter into control calculations. Furthermore, the RF by-pass circuit, whose transfer function is shown in Figure 26, may be ignored if an overall system phase crossover frequency lower
than $10^6$ rad. is considered adequate to meet the amplitude error
sec.
tolerances. This will be assumed to be the case unless measurements on
the actual system show otherwise.

![Block Diagram of Screen Modulating Circuit](image)

**Fig. 39.** Block Diagram of Screen Modulating Circuit

![Screen Modulator](image)

**Fig. 40.** Screen Modulator
The screen arrangement as shown in Figure 40 may be adequate; however, an improvement can be made by feedback compensating the Hewlett-Packard programmable power supply. This would provide advantages:

(a) the system phase crossover frequency would be increased, allowing more system gain
and (b) the DC screen supply would be regulated, removing—to a large degree—amplitude errors from this source.

Since the Hewlett-Packard supply has a simple pole, the effect of F-B compensation may easily be determined analytically. The proposed F-B arrangement is shown in Figure 41, where (H) is chosen to be \((10^{-2})\) so that the DC screen reference (R) may be taken as 8 volts.

---

![Diagram 41](image1)

**Fig. 41. Regulation of the Screen Supply**

![Diagram 42](image2)

**Fig. 42. Equivalent Block Diagram of Figure 41**
The equivalent block is shown in Figure 42. For purpose of calculation, take \( A = 10^4 \), which then gives

\[
\frac{V_{g2}}{e} \sim \frac{10^2}{1 + S/10^2a} = \frac{10^2}{1 + S/1.25 \times 10^7}
\]

(3.21)

This circuit would then give approximately a 40 db reduction in DC screen voltage errors, and move the modulating supply break frequency up to about

\[
10^2 \times 1.25 \times 10^5 \text{ rad. sec.} \sim 2 \times 10^6 \text{ Hz.}
\]

If the amplifier providing the required 60 db of feedback has a 20 db/decade drop-off in gain above a single break frequency, this feedback loop will be absolutely stable. However, to increase the break frequency of the equivalent block considerably above \( 1.25 \times 10^5 \text{ rad. sec.} \), the local F-B amplifier must have a break frequency several decades above \( 1.25 \times 10^5 \text{ rad. sec.} \).

This requirement would be very hard to meet; therefore, one must settle for a lower overall break frequency. In the remaining calculations, a F-B compensated screen modulator having an equivalent transfer function

\[
T(S) = \frac{10^2}{1 + S/1.25 \times 10^7}
\]

(3.22)

will be taken as the limit on what can be achieved in this direction.

B. Open Loop Response

Figure 43 gives the open loop frequency response of the screen modulating system using an uncompensated screen modulator. A compensated modulator is used for Figure 44. In both cases, the feedback amplifier providing the signal to the screen modulator provides the same compensation as was used in the drive control system; that is,

\[
T(S)_{\text{amp.}} = \frac{K_{\text{F-B}}}{(1 + S/2 \times 10^4)(1 + S/6 \times 10^4)}
\]

(1 + S/10^3)(1 + S/6 \times 10^5)

(3.23)

for the feedback amplifier. The Bode plots indicate that without screen compensation, the maximum system gain would be about 51 db, with a phase crossover frequency of

\[
1.8 \times 10^5 \text{ rad. sec.} = 29 \text{ KHz.}
\]
Fig. 43. Open Loop Frequency Response of the Screen Modulating Feedback System with Lag-Lead Compensation (Screen Modulator is Uncompensated)
Fig. 44. Open Loop Frequency Response of the Screen Modulating Feedback System with Lag-Lead Compensation (Screen Modulator is Compensated)
When the screen compensation of the previous section is used, allowable system gain is raised to about 62 db with a phase crossover frequency of
\[ 5.8 \times 10^5 \text{ rad.} = 92 \text{ KHz.} \]

C. Screen Modulation System Simulation

As for the drive control system, the Ecap program was used to analyze the behavior of an electrical analog of this system. Since the only essential difference between the screen modulating system and drive control system is that the first has a screen modulator instead of an RF chain, a function block for the first (Figure 45) was substituted for the latter in the program.

![Simulation of Screen Modulator Function Block](image)

\[ C = 8 \times 10^{-8} \quad \text{(Screen Modulator Uncompensated)} \]
\[ C = 8 \times 10^{-10} \quad \text{(Screen Modulator Compensated)} \]

Fig. 45. Simulation of Screen Modulator Function Block

1. Frequency Response

Again, the closed loop system frequency response with respect to a unit amplitude error signal introduced at the plate of the PA was found. Figure 46 shows the response for the screen with and without compensation. Also shown for comparison is a plot of the response with the F-B loop open. As is to be expected, the error reduction for low frequencies is in each case equal to
\[ \frac{1}{1 + Kp} \]
Fig. 46. Closed Loop Frequency Response of the Screen Modulating System (Error Input at the Plate of the PA)
where

\[ K_p = \text{position error constant} \]

\[ K_p = \text{DC gain in this case.} \]

If the modulating supply is uncompensated, the graphs show that below about

\[ 10^3 \text{ rad.} = 160 \text{ Hz}, \]
\[ \text{sec.} \]

the maximum error reduction will be about \(-51\) dB. Above this frequency, the amount of reduction decreases by about 20 dB per decade, the effect of F-B becoming minimal at about

\[ 10^5 \text{ rad.} = 16 \text{ KHz}. \]
\[ \text{sec.} \]

ii. Time Response

Using the Ecap transient response program and the electric analog circuits already described, the error output for an amplitude error at the PA plate was found. Figures 47 and 48 show the outputs for unit step error inputs to the PA plate for maximum F-B amplifier gain. In both cases, a significant decrease in the error as a result of increasing the feedback amplifier gain is seen.
Fig. 47. Plot of Change in Resonator Voltage Amplitude for a Unit Step Change in RF Amplitude at the PA Plate (Screen Modulation Control used; Screen Modulator Uncompensated)

- **NO F-B**
  - F-B AMP HAS 38.45 db
  - OF GAIN
  - FINAL VALUE = 1.21

- **F-B AMP HAS 76.9 db**
  - OF GAIN
  - FINAL VALUE = .018
Fig. 48. Plot of Change in Resonator Voltage Amplitude for a Unit Step Change in RF Amplitude at the PA Plate (Screen Modulation Control used; Screen Modulator Compensated)
4 CONCLUSIONS

4.1 RF Amplifier System

The RF amplifier system described in this thesis has satisfied the steady state voltage and power requirements given in Section 1.1. The power required by the CRM cyclotron (150 KW) is easily supplied by the final output tube which is capable of an output of up to 460 KW. Actual operation of the amplifier has indicated that it is stable: no parasitic oscillations have been observed. At present the longest period of continuous operation with a cavity voltage of 100 KV has been about 5 hours; longer periods of operation have not been possible due to problems with arcing in the resonators.

4.2 RF Amplitude Control System

The two amplitude control systems described in this thesis were built with a common compensating feedback amplifier differing from the one proposed only in that its first break frequency was 79.6 Hz instead of 159.0 Hz.

Initial tests on the operating systems showed results consistent with those calculated. For example, it was found that both control systems showed a maximum error reduction of about 50 db which is very near to the calculated values. It was also found that the drive control system oscillated at about 50 KHz for excessive feedback gain. This compares closely with the results shown in Figures 36 and 37.

The open loop error consisted mainly of a 120 Hz signal at about -30 db for the drive unsaturated and -35 db for the drive saturated. The 50 db reduction gained by closing the loop still left the system performance short of the -94 db error ratio desired. The installation of DC filament supplies should improve this performance. To accurately evaluate the system, more comprehensive noise measurements (magnitude and frequency) will have to be made in addition to checking the control system frequency response. The information provided by the latter measurements will be needed to optimize the present feedback system settings or to suggest, if necessary, improvements to the compensator.
FOOTNOTES


3. Continental Electronics Manufacturing Co. is a subsidiary of Resalab, Inc., Dallas, Texas.

4. The other 250 watts of drive required by this stage are dissipated in the grid swamping resistor and in the tank coil.

5. The peak voltage on each resonator is 100 KV; push-pull operation of the 2 dees brings the gap voltage to 200 KV.

6. The ability to function with the operating frequency increased by 3% was initially demanded since it was desired to have a cyclotron with an optional, higher output energy.


8. These noise figures were obtained from correspondence with Continental Electronics.

9. 20 KV series regulated DC supply, built by Cober-Electronics Inc., Stamford, Conn.

10. Martin, p. 446.


A SELECTED BIBLIOGRAPHY


Jameson, Robert A. Analysis of a Proton Linear Accelerator RF System and Application to RF Phase Control. Los Alamos, 1965.


