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FINAL REPORT

MASTER

**MFTF ELECTRON CYCLOTRON RESONANCE
HEATING CONCEPTUAL DESIGN STUDY**

PREPARED FOR
**UNIVERSITY OF CALIFORNIA
LAWRENCE LIVERMORE LABORATORY
LIVERMORE, CALIFORNIA 94550**

UNDER PURCHASE ORDER NO. 6631709

NOVEMBER 26, 1979

**GENERAL ELECTRIC COMPANY
ENERGY SYSTEMS PROGRAMS DEPARTMENT
SCHENECTADY, NEW YORK**

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FOREWORD

This is the Final Report for the General Electric Company study on a conceptual design for electron cyclotron resonance heating for the Tandem Mirror Fusion Test Facility (MTTF-B). This report was prepared by the Energy Systems Programs Department, General Electric Company covering work performed during the period August 24, 1979 to November 26, 1979 under Purchase Order 6631709 of the University of California, Lawrence Livermore Laboratory, Livermore, California.

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SECTION I

INTRODUCTION

This report presents the results of a General Electric Company conceptual design study of an electron cyclotron resonance heating (ECRH) system for MFTF-B. The study was done under contract to the University of California, Lawrence Livermore Laboratory, Livermore, California.

The RF heating system in this study is based on the principles contained in the Rough Draft Proposal for the ECRH System prepared by LLL. Two MW of RF heating at 60 GHz are to be supplied to the plasma on MFTF-B for auxiliary heating. This study addresses the conversion of a single Sustaining Neutral Beam Power Supply System (SNBPSS) Accel Power Supply to provide dc power for eight 60 GHz gyrotrons, each capable of providing 250 kW. The RF pulse width is expected to be adjustable from 10 ms to 30s with a maximum duty factor of 10%. Additionally, the study addresses the waveguide distribution system which would be required to efficiently distribute the energy from the power source to the MFTF-B device. The power output from four tubes would be fed into each of the two end plugs.

The equipment necessary to provide pulsing and protective functions for the multiple channel load connected to the converted SNBPSS power supply are considered. Also, the internal interaction between the eight gyrotrons operated from the common power supply, and external interaction between the ECRH power system and the overall SNBPSS are analyzed.

Significant further development of gyrotrons and waveguide distribution components are required to enable production of the ECRH system described herein. Present DOE programs in place approach the problem of the development of high power millimeter wave gyrotrons. A program will have to be developed to address the specific tube characteristics desired for MFTF-B. In addition, an extensive development program on waveguide components must be undertaken.

This report presents conceptual designs, discusses research and development requirements, and provides schedule requirements and rough order of magnitude cost

estimates for the ECRH system. Requirements for the basic equipment needed to implement the ECRH power generators and distribute the power have been developed. Conceptual approaches to the development and fabrication of such a system have been generated.

Much of the information provided in this report is based on trade-off and design analysis using good engineering judgement with only some areas subjected to more detailed analysis. Schedules and costs have been generated from engineering estimates based on past experience, and should be considered as subject to considerable variation.

The study reported herein was performed under a fixed price, level of effort contract of fourteen weeks duration for Lawrence Livermore Laboratory (LLL). Following the direction provided at a meeting at LLL, substantially more time was spent on the gyrotron transmitter system than on the waveguide distribution system, and the latter did not address the problem of coupling the RF energy to the plasma.

SECTION 2

STUDY OBJECTIVES

The overall objective of this study is to develop a conceptual design of an RF Heating System for MFTF-B in accord with the proposal submitted by ESPD to Lawrence Livermore Lab and Technical Discussion of the problem forwarded by LLL as enclosure C to the RFQ. The study will consider the applicability of an MFTF sustaining neutral beam power supply to power eight gyrotron type mm wave tubes and will cover the neutral beam power supply modifications, the gyrotron transmitter and the interfaces to the distribution system.

Completion of this task will involve the following subtasks:

- Have liaison with tube manufacturer of gyrotron devices
- Review the status of applicable gyrotron tube development programs and obtain characteristics of operation
- Generate a conceptual equipment configuration that provides operation of eight gyrotrons from a single MFTF power supply
- Develop a block diagram for the conceptual system
- Prepare a description of the equipment represented by the diagram
- List important design parameters and requirements for the equipment represented by the diagram
- Provide equipment generation breakdown
- Perform a transient analysis of the MCRH system, to the extent possible, to determine the degree of interaction between gyrotron systems when driven from a common power supply
- Estimate cost
- Prepare a draft Final Report
- Prepare a final report incorporating LLL comments.

SECTION 3

ECRH DESIGN DESCRIPTION

3.1 GYROTRON TRANSMITTER SYSTEM

3.1.1 INTRODUCTION

MFTF-B will require an RF plasma/heating system for maintaining a high plasma temperature (high potential) at the center of each yin-yang set located at the ends of the machine. Since the RF power will be used for bulk heating and the field intensity in the center of a yin-yang set is continuous from 2 to 4 T, a single frequency of approximately 60 GHz is considered adequate and need not be adjustable. A total system output of 2 MW, exclusive of waveguide losses, is adequate for MFTF-B. The RF system will be comprised of two sets of four RF sources, each source rated at 250 kW maximum. All the RF sources will be operated at the same pulse length and power level to ensure symmetrical heating of the end plugs.

The ECRH system will be operated remotely through the MFTF supervisory control and diagnostics system during both the tune-up mode and the normal operating mode. During the tune-up mode, each of the eight RF sources will be operated individually into a water load at a maximum repetition rate of 1 pulse per second and a maximum duty factor of 10%. During the normal operating mode, all the RF sources, or a subset thereof provided symmetry is maintained, will be run under the same conditions and, in the event one RF source fails, the ECRH system may be turned off. Radio-frequency power produced by each RF source will be continuously monitored during normal operation. The RF pulse length will be adjustable from 10 ms to 30 s with a maximum duty factor of 10%.

Local operation will be provided for purposes of acceptance testing the equipment into water loads and maintenance. A local control and monitoring system will be integral with the equipment and compatible with the MFTF standard Local Control and Instrumentation System.

The purpose of this study is to develop a conceptual design for an ECRH system for use with MFTF-B. This heating system is based on the principles set forth in the Rough Draft Proposal for the ECRH System prepared by LLL. The major thrust of the study will center on the conversion of a single sustaining neutral beam power supply system (SNBPSS) ac/dc power supply to provide 8 power for eight 60 GHz gyrotron oscillators. Additionally, the equipment necessary to provide pulsing and protective functions for the multiple channel load connected to the converted SNBPSS power supply will be explored. Finally, internal interaction between the eight gyrotrons operated from the common supply, and external interaction between the ECRH power system and the overall SNBPSS power system will both be considered.

3.1.2 SUMMARY

The reference design ECRH is an eight channel gyrotron RF generator system which will be used for maintaining high plasma temperature at the ends of the MFTF-B machine. Figures 1 and 2 show plan views of the overall MFTF area involved. Specifically, several important equipment areas and interfaces are shown in Figure 1. One of the twenty-four 90 kV @ 90 ampere dc power supplies (e.g., the circled one denoted as A and shown in the upper left hand corner of the outdoor field) interfaces with the 230 kV/13.8 kV three phase 60 Hz substation via the dashed line from the right and interfaces with the MFTF-B via the high voltage dc cable running to the left. It is necessary to make modifications to this supply (e.g., change polarity of output voltage from plus to minus) to adapt it for ECRH use. Figure 2 shows that the 90 kV dc cable enters Bldg. 431 and proceeds to an area denoted as 2, which will house the high voltage dc regulation, pulse modulation, and fault diverter (crowbar) equipment. This pulsed output splits two ways and goes to the areas denoted as C1 and C2 which each house four 60 GHz gyrotron oscillator assemblies and their associated switching networks. Finally, the gyrotron RF outputs will be transmitted via oversize circular waveguides to the end plugs (denoted as D1 and D2) of the MFTF-B machine to provide the required RF plasma heating.

Table 1 gives a summary of the ECRH parameters used for the reference design.

Table 1

LIST OF ECRH PARAMETERS FOR REFERENCE DESIGN

A. Prime Power (at 230/13.8 kV substation)			
1. Voltage			13.8 kV, line to line
2. Frequency			60 Hz $\pm$ 0.25 Hz
3. Rated Capacity			250 MVA Pulses
4. Static Voltage Variation			$\pm$ 5.0 percent
5. Short Term Voltage Variations			130 sec $\pm$ 1.0 percent
6. Substation Equivalent Impedance at 250 MVA			24% Reactance 2% Resistance
7. 13.8 kV Connection			Wye with grounded Neutral
8. Pulse loads only			10% duty cycle
B. DC Power Supply			
1. Input Voltage			13.8 kV, line to line
2. Input Voltage Adjustment			50-100%
3. Rectifier Transformer			Primary = 1 Secondary #1 = 1 Secondary #2 = 7
4. Transformer Equivalent Impedance			20% Reactance 2% Resistance
5. Power Factor Correction			6 lines (secondary)
6. Rectification Mode			12 pulse
7. Output Voltage			90 kV
8. Output Current			90 A
9. Duty Factor			10%
10. DCR Limiting Resistor			20 ohms
11. Filter Capacitor			5 μ F
12. Capacitor Current Limit Resistor			50 ohms
C. Modulator/Regulator			
1. Pulse Duration			10 ms - 30 Seconds
2. Output Pulse Top Voltage Regulation			$\pm$ 0.5% of preset value
3. Pulse Rise Time			20 μ Seconds Max.
4. Load Fault Removal Time			20 μ Seconds Max.
5. Duty Factor			10% max.
6. Switchable Output Channels			9
7. Crowbar Delay (including Sensing)			5 μ Seconds
D. R-F Generator			
1. Electrical Operating Parameters	Minimum	Maximum	Typical
a. Beam Voltage	80	100	90 kV
b. Beam Current	7	11	9 A
c. Gun Control Anode (with respect to Cathode)	10	30	20 kV
d. Heater Voltage (AC)	6	12	9 V
e. Heater Current	4	3	6 A
f. Body Current		80 mA	40 mA
g. Gun Anode Current		80 mA	10 mA
h. Power Output	175	300	250 kW
i. Efficiency			30 %
2. Power Output Sensitivities			
a. Gun Control Anode Voltage		0.6dB/%	
b. Main Magnet Current		1.5dB/%	
c. Gun Magnet Current		0.3dB/%	
d. Heater Voltage		0.6dB/%	
e. Beam Voltage		0.6dB/%

Figure 1. SFTF Neutral Beam Sustaining System
Outdoor Equipment Layout

3.1.3 SYSTEM BLOCK DIAGRAM/MAJOR ITEM IDENTIFICATION/FUNCTIONAL DESCRIPTION

The ECRH system will be comprised of the following major elements (see Figure 4):

- A dc power supply rated at 90 kV (negative polarity), 90 A, 30 s, 10% duty factor
- A modulator/regulator with switching network
- 3 gyrotron systems
- 3 waveguide systems (not shown)
- Utilities, a 2-MVA unit substation, and an LCW cooling system (not shown)

The above items, with the exception of the waveguide systems and the utilities, etc, have received consideration from the point of view of an overall electrical system which starts at the public utility's prime line input and ends with the gyrotron RF generators. Major items within each block will be described with reference to a simplified basic system schematic.

3.1.3.1 230/13.8 kV Substation/Other Substation Loads

The characteristics of the substation will influence, to some extent, the performance obtainable from the ECRH system. Conversely, the ECRH system can also influence, to some extent, the performance of the sustaining neutral beam (SNB) systems whose power is supplied from the same substation. This is a consequence of the fact that the substation's equivalent impedance (primarily associated with the 230kV/13.8kV distribution transformers) is common to both the SNB and ECRH systems. Since it was not obvious that the interaction between these two systems would be insignificant, this potential problem was set forth as an area of investigation for the study.

3.1.3.2 HVDC Power Supply

The 90 kV dc power supply will be one of the MFTF sustaining neutral beam power supplies (SNBPSS) with appropriate modifications as shown on the simplified schematic of Figure 4. The required modifications include a negative polarity rectifier assembly, fencing to separate the dc power supply from the sustaining neutral beam power supplies, and a modified control and monitoring system and safety system to make the power supply compatible with the ECRH system.

3.1.3.3 Modulator/Regulator

The modulator/regulator for ECRH will be new because of the requirement for negative polarity (the MFTF sustaining neutral beams require positive polarity) as

Figure 3. GCRH System Block Diagram

shown in Figure 4. The circuitry and components required are very similar to those used in the SNBPSS modulator. The modulator/regulator will regulate the output voltage to within $\pm 0.5\%$ of a preset value, apply and disconnect power from the loads, and crowbar the voltage off the loads in the event of a fault. A remotely controlled 3-port switching network at the output of the modulator, which was not required for SNBPSS, will select or deselect the gyrotron systems to be run. Integral with the equipment will be a local control and monitoring system similar to that required for SNBPSS.

3.1.3.4 Gyrotron Systems

Eight gyrotron systems will be required to convert electrical power into RF power, each rated at 250 kW output. Each gyrotron system will be comprised of the following elements as shown in Figure 4.

- Gyrotron tube with magnet assembly
- Set of magnet power supplies
- Isolation network
- R-C compensated network
- Control anode power supply
- Filament power supply
- LCW cooling system
- Local control and monitoring system

The gyrotron is a state of the art device that will be developed for this application. The tube will have the following general characteristics: $f_0 = 60$ GHz, 250kW output, 30 s pulse length, and 10% duty factor. If the conceptual design assumes this tube will have excellent voltage holding characteristics, it will allow all eight tubes to be driven from a common power supply and a common series modulator. Each tube will require typically -90kV at 9A with a voltage regulation of at least $\pm 1\%$. Since a gyrotron is relatively insensitive to input voltage, all eight tubes may, based on the above assumptions, be run at the same voltage. The tube requires a set of four conventional, pulsed, copper magnets to create a 2T magnetic field in the RF cavity. In addition to the main magnets, two smaller magnets are required to ensure proper focusing of the electrons in the anode cathode region of the tube. By adjusting the currents in these magnets the tube may be tuned for optimum performance.

Each magnet requires a remotely programmable, current regulated ($\pm 0.1\%$) power supply. The main magnet power supplies are rated at 1000A and 50V with a 10% duty

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Figur

Figure 4. Simplified Basic Schematic

factor. These power supplies are rated at 10A, 50V and continuous duty. Such supplies are commercially available and pose no unique design problems.

When several tubes are to be driven from a common power supply and a common modulator, an isolation network will be required for each tube. An air core inductor and resistor, rated at 150kV, 16A, and a 10% duty factor, could make up this network. The values of the elements of this network must be chosen to prevent large transients from being reflected back into the modulator/regulator in the event one of the tubes arcs down. The isolation network could be located in the supporting tank for the gyrotron tube and immersed in high voltage liquid/gaseous insulation.

An RC compensated voltage divider will be required to bias the control anode of the gyrotron tube at approximately 25kV with respect to the cathode. In addition to providing biasing, this network will ensure proper timing of the control anode voltage with respect to the 90kV beam voltage. The voltage holding capability of the tube during turn-on may be contingent upon delaying the application of the full accelerating potential with respect to application of the anode control voltage. The compensated divider could be housed in the supporting tank of the gyrotron.

To ensure sufficient control over the focusing characteristics of the tube, a separate, regulated ($\pm 0.1\%$) adjustable, low-current, 35kV power source is required. This source will float at -90kV and be directly tied to the gyrotron cathode. The supply will have a remotely settable voltage output from 20kV to 30kV, a current rating of 100mA, and will be located in the supporting tank of the gyrotron.

The beam current and thus the output power of a gyrotron is determined by the heater power supplied to a dispenser cathode that is indirectly heated and operated thermoionically limited. A 50W unregulated power supply with a primary to a secondary hold off capability of the full accelerating potential of 90kV is required. A special power supply will be needed for this application and will be located in the supporting tank of the gyrotron.

To ensure that the dimensions of the gyrotron's RF cavity are maintained during the 30 s pulse, the cavity should not experience a temperature change greater than 8°F. Consequently, a sophisticated cooling system is required similar to systems used for high power klytrons and tetrodes. The water in the system will have a dissolved CO₂ and O₂ level of less than 1 ppm, a resistivity of approximately 5M Ω - cm, and, when the magnet cooling requirements are included, a flow rate of 500 gal/min. Specific cooling requirements are shown in Table 2.

Table 2
COOLING REQUIREMENTS

Deionized water cooling is required as follows:

	FLOW RATE gpm	PRESSURE DROP psi
Tube Collector	200-400	120-150
Tube Body	20-30	75-140
Room Temperature Magnet Coils	5	100
Output Waveguide	10-20	75-100
Output Water Load	50-100	3-6

The cathode end of the tube is designed to be operated in oil for both voltage hold-off and cooling. An oil flow of 0.3 gpm or more is required to cool the gun. A small submersible oil pump is recommended to supply this flow.

Each gyrotron system will require a dedicated local control and monitoring system that is plug compatible with the standard MFTF local control and instrumentation systems. The system will have sufficient local controls to allow acceptance testing and maintenance independent from the MFTF computer control system. Normal operation and system tuneup will be performed with the use of the MFTF supervisory control and diagnostics system.

3.1.3.5 Design Variations

A number of assumptions have been noted above, particularly with respect to the gyrotron's capabilities. As a result of this study, several alternate approaches will be set forth in subsequent sections of this report.

3.1.4 SYSTEM ANALYSIS

3.1.4.1 Introduction

Several computer models have been developed for the purposes of, first, analyzing the inherent performance capabilities of the MFTF-B ECRH system, and second, examining the interaction that exists when both the SNB and ECRH systems operate from a common prime power source.

In order to accomplish the necessary computer modeling work in the limited period of time available for this study, recourse was made to some computer programming developed several years ago in connection with MFTF SNBPSS studies performed at GE. This previous work was primarily concerned with modeling the prime

power substation and the multiple accel power supply characteristics. While these programs may not reflect the exact parameters which are currently in use for MFTF, they are believed to be representative. If desired the programs could be readily modified to reflect the current system parameters.

The following will describe the models used and the results obtained from computer runs executed to explore several areas of concern in the overall MFTF-B ECRH system.

3.1.4.2 AC/DC Power System Model

Figure 5 shows the computer model used to simulate the substation/main power supply complex. The element rationale is as follows:

- SV1, SV2, SV3 - These represent a zero impedance three phase generator/substation source with no voltage or phase unbalance
- LG1, LG2, LG3, RG1, RG2, RG3 - It is assumed that the substation is rated 250MVA and that two 230kV/13.8kV distribution transformers are used. This simulation represents the load on one of the 125MVA transformers. LG1, LG2 and LG3 represent the leakage inductance values equivalent to those of a transformer with 28% reactance on a 125MVA base. Similarly RG1, RG2 and RG3 represent the resistance values for 2% resistance on a 125MVA base.
- RI1, RI2, RI3, LI1, LI2, LI3 - In this simulation two load groups are assumed. The group of major interest is a single accel (or gyrotron) power supply. A lumped load representing 12 additional supplies comprises the second load group. LI1, LI2 and LI3 represent the leakage inductance for the variable voltage transformer (in the single channel) and the values are equivalent to those of a unit with 2% reactance on a 7.37MVA base. RI1, RI2 and RI3 are equivalent to 2% resistance on a 7.87MVA base.
- LP1 to LP6, LS1 to LS6, K1 to K6, RS1 to RS6, RT1 to RT6 - These elements simulate the Δ -Y/ Δ - Δ transformers used for the 12 pulse rectifier of the single channel power supply. LP1 to LP6 represent the exciting inductances of the primary windings of the transformer. Their values were selected to produce a value of 2% exciting current. Secondary exciting inductances LS1 to LS6 then had their values established by the required turn ratios of the Δ -Y and Δ - Δ windings. K1 to K6 are the coupling coefficients between the 6 pairs of windings, and their values were selected so that a leakage inductance per phase--equivalent to a 17% transformer reactance on a 7.66MVA base--would result. RS1 to RS6 represent resistances equivalent to 3% transformer winding resistances on a 7.66MVA base. Finally, RT1 to RT6 simulate core losses of 1% on 7.66MVA base.
- CP1 to CP6 - These elements are the power factor correction capacitors. Their values were selected to cancel out the leakage inductances associated with the rectifier transformer, the variable voltage transformer and an appropriate portion of the generator inductance.

Figure 5. MFTF High Voltage DC Power Supply System Model

- SG1 to SG6 - These elements comprise a subprogram which senses when the voltage across their associated value of CP- exceeds a predetermined value. The gap then breaks down and inserts a predetermined value of resistance (i.e., 5.4 ohm) across the capacitor. When the voltage across the capacitor goes through zero--later in a cycle--the gap extinguishes.
- D11 to D112 - These are the 12 pulse rectifier diodes and are simulated in a subprogram to present a predetermined low resistance (i.e., 1 ohm) when their anodes are positive and a very high resistance (i.e., 1 megohm) when their anodes are negative.
- RC1, RC2 - These resistors represent elements provided for current limiting under fault conditions. Their combined value was 95 ohms for some of the early runs and 20 ohms in later runs.
- RS7, CS1, RS8, CS2 - These values represent the filter capacitor bank values. The combined value of series capacitance was 44 microfarads on some of the early runs and 5 microfarads on some of the later runs. The combined value of series resistance had values of 170 ohms and 50 ohms for the early and late runs respectively.
- R4 - For simple power supply analyses this resistor is programmed to represent power supply dc loading (e.g., to 965 ohms for 82kV @ 85 amp operation or to some low value for load fault operation.) For pulsed operation a more detailed load model is used.
- L1L1 to L1L3, R1L1 to R1L3, C1L1 to C1L3, R2L1 to R2L3 - These elements are used to represent the lumped load used to approximately simulate the effects of 12 additional power supplies of the type which has been described above in single channel format. L1L1 to L1L3 represent the reactive portion of the lumped power supply series inductance, R1L1 to R1L3 represent the series resistive component, C1L1 to C1L3 represent the power factor correction component, and R2L1 to R2L3 represent the load resistance portions of the model. Since rectification is not included as a part of the lumped model, this simulation is obviously not as good as the single channel model from the point of view of waveform distortion. It is, however, quite satisfactory for a number of power system investigations.

3.1.4.3 Common Power System Interactions

With the power supply model established as outlined in Section 2, four areas of technical concern were subjected to computer analysis. These areas involved, first, the power factor correction technique used in the system; second, the effects of adverse operation of a single power supply (e.g., crowbar operation) on the common power supply interface bus voltage; third, the relative magnitude of fault currents on a crowbarred power supply; and fourth, the power supply's dc fault build up current characteristics.

It should be noted that the power supply model described in Section 2 was programmed in CIRCUS 2 format. This permitted continuous transient solutions of system performance to be obtained--in both tubular and line printer plotted output format--as functions of time.

A representative CIRCUS 2 run showing several line printer outputs is shown in Figure 6. This reference run was actually for the SNBPSS circuit configuration (i.e., rectifiers connected for positive voltage output). The output load on the HVPS was 965 ohms and the average output was 82kV at 85 amperes. The run starts with initial currents in all circuit elements established from a previous run.

Figure 6A shows one of the line to line voltage waveforms at the output of the source (i.e., at the substation distribution transformer secondary) in the upper plot, and it shows one of the line currents to the power supply's variable voltage transformer as the lower waveform. The characteristic current distortion associated with 12 pulse rectification is clearly visible. Figure 6B shows a second pair of line voltage and line current waveforms. In both cases the approximate 30° displacement of waveforms is what one would expect with a nearly 100% power factor load. Hence it can be seen that the power factor correction circuit is effectively performing its function.

Figure 6C (upper plot) shows the power supply's dc voltage ripple across the load. The run indicates two things, first, the system has not established itself at a completely steady-state condition, and second, the peak to peak ripple output is typically 630 volts (n.b., these values were obtained from a tubular printout and are those associated with the fourth ripple peak). The lower waveform shows the dc output voltage across the lower power supply only. It is of course at only about 1/2 the full output voltage level, and its basic ripple frequency is at 360Hz rather than 720Hz.

Figure 6D (upper plot) shows the primary power input to the power supply. The plot is the summation of three volt ampere products. Each product is the instantaneous value of rectifier transformer primary leg current times leg voltage. The value of average power can be seen to be approximately 8 megawatts.

Figure 6D is a somewhat startling waveform at first glance. It is the summed power (i.e., obtained as described above) for the total load on the generator. It appears to have some much higher frequency components present than the basic 720Hz ripple associated with the power supply. These components are probably the result of high frequency ringing oscillations associated with phase to phase rectifier

Figure 6A. Common Interface Voltage #1 and Line Current #1

Figure 6B. Common Interface Voltage #2 and Line Current #2

Figure 60. Single Channel Power and 12 Channel Line Power

Figure 60. Single Channel AA + AY Rectified Output and Single channel AA Rectified Output

commutation. In any event their amplitude is quite low--approximately 500 watts on a 32 megawatt average power level base.

With the basic model capabilities established, several runs were performed to investigate the areas of prime concern noted previously. These will be described in the following paragraphs.

Power Factor

The power factor correction technique applied to the MFTF power supply was investigated for two reasons: first, to determine how effectively power factor could be corrected under normal operating conditions and second, to determine how a crowbar fault on a single power supply would affect the power factor of the total load applied to one of the two substation transformers.

Based on the power system parameters set forth in Section 2 above, the value of power factor correction capacitor for each rectifier transformer secondary phase was 23.04 microfarads. This appears to be about 20% higher than the 19.3 microfarad value we understand LLL has selected for this function. The difference is no doubt due to the assumptions made herein on power system impedances. In any event, the constant dc load computer run noted above showed that the active single channel power supply had power factor readings of .9994, .9963, .9978, and .9981 on the first four zero crossings (i.e., power factor was determined by comparing the common interface line to line voltage crossing with the primary leg current crossing). In all cases the current lagged the voltage, hence not quite complete compensation resulted from the capacitors selected.

The peak voltage appearing across the power factor capacitors was found to be between 10.3 and 10.4 kilovolts. Based on this value it was arbitrarily decided to set the protective gap across the capacitors at 15 kilovolts peak.

Since there was some uncertainty regarding the stability of the overall system when the spark gaps functioned (e.g., do objectionable oscillations occur when the gaps break down at 15kV or when they open when crossing zero?) a computer run was executed for the case where the power supply crowbarred at 2 milliseconds into a run. Figure 7 shows the waveforms of voltage across the six power factor correction capacitors. It will be noted first, that the first gap broke down approximately 1.5 milliseconds following the fault; second, that all gaps had broken down within 5.5 milliseconds after the fault; third, that the gaps cleared when crossing the zero following a breakdown; and fourth, that no objectionable oscillations occurred after

Figure 7. Spark Gap Voltages Under Load Fault Conditions

either the opening or closing of a gap. Based on this run no shortcomings could be seen in the technique.

A summary of the power factor analysis is shown in Figure 8. Three situations are examined here with regard to the power factor at the source (i.e., the common interface of the 13 loads). The dashed curve shows that for the constant dc load case the power factor remains constant at about .9998 lagging. (n.b., data is plotted in degrees lead or lag with power factor in parenthesis.) The solid curve shows how the power factor fluctuates at successive zero crossings - in a system without spark gaps - following a fault which occurs at 2 milliseconds. It is interesting to note that a lagging power factor is followed by a leading power factor (n.b., since only discrete points were calculated the artistry in connecting the points may be questioned, however, definite fluctuations do appear to occur). Finally, the dash-dot curve shows how the power factor would behave if 15kV breakdown level spark gaps are employed. Clearly, in this case, a lagging power factor prevails.

Common Interface Voltage Variation

The cause of concern in this area centers around the effect on performance of the nonfaulted groups when a fault occurs on one of the 13 power supplies connected to a single prime power distribution transformer. In this analysis the peak values of the line to line voltage waveforms at the common interface were used as a comparison reference.

Figure 9 summarizes the results of three computer runs. The first run was one with constant power supply loading and as such the voltage peaks stayed constant (i.e., zero percent voltage variation).

The second run faulted one power supply - which was not equipped with capacitor gap - at time equal to 2 milliseconds. This is the solid line plot of Figure 9. It can be seen that an initial dip in voltage of approximately 4 percent is followed by rise of approximately 5 percent at the interface. As in the case of the power factor plot, issue may be taken with the between point artistry--however--nonlinear variations appear to occur. The general pattern of this waveform bears a close family resemblance to Figure 8.

The dash-dot curve of Figure 9 shows how the interface voltage fluctuates for the case where 15 kV spark gaps are used across the capacitors. It appears that the interface voltage is reduced by about 3-1/2% in this case.

Figure 8. Power Factor Analysis

Figure 9. Common Interface Voltage Variation

From these limited investigations it can be concluded that a fault in a single power supply has a quite pronounced effect on the interface voltage when power factor correction capacitors are used in the supplies. However, no serious system instability appears to exist based on the situations analyzed.

Magnitude of Fault Currents/Voltages

In this area the points of concern regard potential transient damage to power supply components under dc load fault conditions. Figure 10 provides one yardstick for comparison of normal operation with fault operation as obtained from the three computer runs noted previously. The yardstick in this case is primary current to the rectifier transformer.

The dashed curve shows normal rectifier transformer primary current waveforms for all three phases--and it can be seen to have a nominal value of approximately 200 amperes.

As before, following the application of a dc fault at 2 milliseconds, the solid curve shows the manner in which the three phase currents build up in the absence of power factor correction capacitor spark gaps. The somewhat modulated waveform appears to be heading for a nominal value of 4.5 to 5 times normal current.

The dot-dash curve shows the same situation when the spark gaps are added across the power factor correction capacitors. This appears to have ironed out the superimposed fluctuations to a large extent, however, the two curves appear to be approaching approximately the same asymptote. This in itself is not too surprising when one considers that the net effective impedance in the power system is on the order of 5% resistance and 20% reactance, hence one might reasonably expect steady-state fault currents of 4 to 5 times normal. It is reassuring, nevertheless, to see that the currents are of an apparently well behaved nature and that no unusual oscillatory effects are present.

A brief examination of potentially destructive voltage transients in the power supply revealed the following: first, the rectifier transformer secondary line to line voltage--on a normal load basis--had transient peak values of approximately 34.7kV, and the peak voltage across the power factor capacitors was typically 10.4kV, second, after a dc fault was executed on a system not equipped with power factor spark gaps the transformer secondary voltage rose to repetitive peak transient values as high as 81.6kV and peak voltages across the power factor correction capacitors as high as 45.8kV, and third, when the spark gaps were added the transformer voltages (after the fault was established at 2 milliseconds) appeared to go

Figure 10. Primary Current Peaks of Successive 3 Phases

no higher than 32.5kV while the capacitors were of course limited to values no greater than 15kV by the action of the gaps.

Based on these observations it would appear that if reasonable precautions are exercised in the power supply design destructive voltage and current transient conditions should not occur. It is important to note, however, that if a highly dependable type of spark gap is not used across the power factor correction capacitors, a substantial additional level of transient voltage withstand capability should be incorporated in both the power factor capacitors and the rectifier transformers.

DC Fault Current Build Up Characteristic

This characteristic is of interest for a number of reasons, for example, in evaluating the coulombs of charge passing through crowbar devices prior to line circuit breaker interruption.

The same three conditions--no load and dc fault with and without power factor correction capacitors spark gaps--have been plotted in Figure 11. A high degree of similarity to the plots of Figure 10, where the absolute values of line current peaks were plotted, can be observed.

It should be noted that these runs have all assumed a dc output current limiting resistor of 95 ohms. Somewhat higher currents would probably result if this value is lowered to 20 ohms. The predominant factor in fault current determination would appear to be the ac prime line current reactance, however. In any event, it would appear that if a 1.5 cycle interruption capability were available in the transformer primary circuit (i.e., following fault occurrence) the coulomb charge passing through the crowbar device could be limited to values within the state of the art of available devices.

3.1.4.4 Modulator/Regulator Performance

In this section the power supply system just described will be extended to include the modulator, regulator and gyrotron loads as parts of the overall cyclotron resonant heating system model. Figure 12A shows the generic system circuitry concept used for this portion of the study. It consists of first, a coaxial cable to deliver the dc power from the high voltage power supply to the modulator/regulator; second, a shunt limiter/regulator tube and its associated low voltage power/control/isolation circuits; third, a crowbar fault diverter and its associated firing/control circuits; fourth, a series regulator tube(s) and associated low voltage power/control/isolation circuits; fifth, a pulse distribution network consisting of multiple coaxial cables, multiple load isolation switches, multiple load fault

Figure 11. DC Current From Rectifier Bank Connected to Y Winding

Figure 12A. Basic Regulator/Modulator and Gun Anode H.V. Circuitry

Figure 12B. Beam Regulator/Modulator and Gun Regulator H.V. Circuitry

isolation elements, and multiple gun anode voltage application elements, and sixth, the gyrotron loading placed on each of the multiple channels.

The operating philosophy for a multiple load system such as this can follow any of a number of points of view. For example, if it is assumed that the gyrotrons present very stable load characteristics (e.g., that they seldom if ever have internal arcs) and that the tubes are very similar in their characteristics (e.g., their cathode voltages, gun voltages, etc. are essentially identical from tube to tube) then a high voltage current configuration such as that shown in Figure 12A would probably be adequate for the system.

If one makes a somewhat more realistic assessment of the probable tube-to-tube characteristics resulting from a fundamentally new tube development program, some variation in the basic circuit configuration would probably result. For example, rather than relying on a simple voltage divider for the gun anodes of the gyrotrons, a shunt regulator for each tube, as shown in Figure 12B, presents an attractive alternative. This provides the dual advantages of first providing an independent, and remote means of voltage adjustment for each tube, and second, providing additional ripple filtering for the gun voltage.

If a still less optimistic assessment of the overall system is considered, the arrangement shown in Figure 13 might be advisable. In this configuration the point of view is that rather frequent arcs will occur in the gyrotrons, hence each tube will have its own series switch tube. This insures maintenance of the highest degree of system RF output continuity since a faulted gyrotron may be almost instantaneously removed from the power system without affecting the output from the other tubes. Additionally, it provides the highest degree of protection to the faulted tube. This latter feature could easily justify the expense of multiple switch channels, since the expense of frequent gyrotron replacement would far outweigh the expense of the switches. This approach, when combined with the gun anode shunt regulator tubes shown in Figure 12B, would appear to represent the most desirable modulator/regulator configuration for this system.

Before examining the details of this circuit configuration, some of the fundamental transient voltage and current conditions which could exist will be examined. Included are transients associated with system snap on, system pulsing, and load faults. These three situations will be examined in the following paragraphs.

Figure 13. Individually Switched Gyrotron System

System Snap-On/Pulsing

There are a number of ways the system may be turned on. For example, the variable voltage transformer could be run down to its minimum voltage output position (e.g., 50% voltage) before closing in the three phase 13.8kV lines to the power supply. This method is perhaps more appropriate for turning on the system the first time each day, rather than after a crowbar firing sequence. This rationale is based on the assumption that with < 30 second pulse duration conditions an excessively long period of time would be required to run down and then run up the variable voltage transformer, whereas, if snap-on at full voltage were possible very little lost time would be required to return the system to operation following a crowbar. In any event, the transient situation for a full primary voltage snap-on was analyzed.

Figure 14 shows the modification made to the basic power supply computer model for this study. Current generator J, battery V, and diode DI1 simulate the high voltage shunt limiter/regulator, while the switch and resistor R simulate the pulser and gyrotron load respectively. A more elaborate model was not required since only first order (i.e., relatively long time base) effects were desired from this run.

The first objective of the model was to have the limiter/regulator operate as a 95kV based diode limiter (i.e., $V = 95\text{kV}$) which has a 2 ampere rating current (i.e., $J = 2$ amperes). This would permit an assessment of the current transients the limiter/regulator would have to absorb for this situation (i.e., the need for this form of limiting was evident when a trial run--without voltage limiting--was executed and the output voltage of the power supply was found to overshoot to an excess of 140kV under full voltage snap on conditions).

The second objective of the model was to ascertain how the transient output voltage of the supply would behave when zero resistance switch SW is closed to insert resistor R into the circuit (i.e., R was programmed to a 1 megohm open circuit for the first 24 milliseconds of the run and 1000 ohms--approximately equivalent to 8 gyrotrons in parallel--for the second 24 milliseconds of the run).

Figure 15 summarizes the key outputs. The upper waveform indicates that the output voltage of the supply reaches 95kV at approximately 7 milliseconds after the 13.8kV lines are closed in. The 95kV clamping action is evident from this point to the 24 millisecond time marker.

The lower waveform shows the current which would pass through the limiter/regulator during this period of time. A peak value of approximately 124 amperes is indicated.

Figure 14. MFTF High Voltage DC Power Supply System Model with Shunt Regulator and Pulsed Load

Figure 15. Snap-On DC Voltage (Upper Trace) and Current (Lower Trace)

At time equals 24 milliseconds, switch SW was arbitrarily closed to simulate the start of pulsing action. It should be noted that the power supply was still delivering about 50 amperes at this time. Nevertheless, the power supply and energy storage capacitor were incapable of supporting the load current demand and a slow oscillation of output voltage ensued--undershooting to as low as 77.6kV, then overshooting to 91.3kV etc.--in the form of a damped wave superimposed on the dc voltage.

With this output in hand a more appropriate operating sequence is immediately evident. Rather than allowing the current waveform to decay so far, the switch should have been closed at this waveform crossed approximately 90 amperes on the way down (i.e., at about 12 milliseconds). The system will then more closely stabilize to the output voltage value required by the load since both the power supply current and the storage capacitor voltage would be at more nearly their steady-state values. A computer run using these conditions is shown in Figure 16. Unless an approach of this general nature is used, it would appear that severe pulse top fluctuations at the beginning of each pulse would result.

Another area of possible concern is associated with the line currents which take place during a full voltage snap-on. A check of this run revealed that successive rectifier transformer primary current peaks were 403, 361, 190, 359, 313 and 197 over the first 12 milliseconds of the run. If these values are compared with the data of Figure 10 it can be seen that these peaks are less than 2 times normal primary current, albeit the waveforms have a rather ragged characteristic. In any event, such full voltage snap-on operation would appear to introduce a negligible impact on overall power system operation.

Gyrotron Faults

Arcs within the gyrotron tubes create power supply faults which must be interrupted rapidly if the ECRH system is to effectively provide RF output to the MFTF-B machine. Four aspects of the gyrotron fault situation are significant; first, some form of current limiting is highly desirable in the fault path; second, the fault path should be interrupted as rapidly as is practical in order to reduce the destructive effects of the arc on the gyrotrons internal structure; third, following interruption, the reapplication of voltage to the faulted tube should be carefully programmed as a function of time, and fourth, the fault interruption sequence of events should produce the least possible effect on the operation of the remaining gyrotrons.

Figure 16. DC Voltage Output Waveform if Load Pulsing Starts at 12.2 Milliseconds

High voltage tetrode vacuum tubes have several features which make them well suited to the role of a high voltage high current interrupter. They have no moving parts, no arc quenching problems, adjustable current limiting in the case of a shorted load, and the ability to dissipate power. A major feature which recommends a tetrode for this application is the fact that its plate current is virtually independent of the voltage drop across the tube. Hence, in the case of a load arc, the fault current will increase typically by only about ten percent. Additionally, tetrodes have the capability to interrupt such faults within a matter of nanoseconds if necessary.

From the above it can be seen that a single tetrode interrupter would, ideally, be capable of providing all of the capabilities necessary for combatting the load fault situation problems except for one. That one problem concerns interruption of the RF output from the other gyrotrons during load fault and reapplication of voltage to the fault channel.

Multiple switch channels can overcome the system RF interruption problem. Two additional advantages accrue from the multiple switch approach, first, the switches may be operated more conservatively, hence, if desired they could also provide a voltage regulator/level adjustment function and allow the gyrotrons to operate at different cathode voltages, and second, a switch tube failure would not necessitate shutting down all channels, rather, the faulted channel could be isolated and the remaining seven channels could continue to operate.

3.1.5 CANDIDATE HIGH VOLTAGE CIRCUIT FOR EACH GYROTRON

The general area of investigation addressed in the preceding analysis sections is defined as 1.5 Gyrotron System in the MFTF Work Breakdown Structure. More specifically, the areas of concern are the following:

- 1.5.3 Isolation Network
- 1.5.4 Disconnect Switch
- 1.5.5 Control Anode Network
- 1.5.6 Tube Arc Protection

As a result of the points of view established while conducting the study, a candidate hardware configuration for the conceptual design in the above areas is shown in Figure 17. The rationale of the configuration will be discussed briefly.

The fundamental premise set forth is that the gyrotrons should be provided with a degree of isolation/disconnectability from the common power supply. Additionally,

Figure 17. Basic High Voltage Circuit for One Gyrotron

the gun control anode voltage should be remotely adjustable from ground level while the gyrotron is in operation.

The first requirement -- isolation -- is provided by utilizing a series switch tube for each gyrotron channel. This approach will ensure that a faulted gyrotron will not result in a load current increase of more than about 10%. The tentative tube selection is a Varian X2062J. It will operate with a cut-off bias of -1000 volts and a screen voltage of 1000 volts. Based on an assumed gyrotron cathode plus control anode current of 10 amperes, the 2062J plate dissipation during the pulse would be approximately 16 kW and the anode would absorb approximately 430 kilowatts during a 30 second pulse. The tube's screen dissipation would be approximately 500 W. These ratings are all well below the 2062J's maximum ratings, hence extremely long life and reliable service should be observed.

The grid driving circuitry for this tube is patterned after that set forth by Sterling McNees.¹ The arrangement shown is a direct bootstrap circuit using two stages of power amplification. Pulse control signals, generated at ground level, are transported to the floating switch tube at the 90 kV level by means of optical isolators or other appropriate voltage isolation techniques. A positive going 10 to 15 volt pulse excursion is adequate to fully drive the grid of the first amplifier, a Varian 8875 zero bias triode. This stage, which has gain of about 10, is direct-coupled to a Varian 8990 tetrode. This tube is in its full conduction stage during the modulator's interpulse period and is driven toward cut-off by the negative going pulse from the 8875 during the pulse output period. This then causes the -1000 volt interpulse negative cut-off bias voltage drop across the 100 ohm grid-to-cathode resistor of the 2062J to assume the required grid cathode voltage during the pulse (e.g., approximately -100 volts).

The low-impedance grid drive circuitry shown has as its purpose the achievement of stability in the X2062J. While the tube's characteristic curves may imply a relatively high impedance circuit would be adequate to drive the grid of the X2062J, the bootstrap driver has been shown to effectively suppress the parasitic oscillations and also to minimize the pulse ringing effects typical of all high power tetrodes.

1. S.G. McNees, "A 100 kV, 90 Amp Long Pulse Switch Tube," IEEE Conference Record of 1978 15th Modulator Symposium, pp. 180-182.

Similar circuitry is shown for the gun anode voltage source. In this case however, the gun voltage is derived from the gyrotron cathode voltage, and a Machlett 8495 high voltage triode (i.e., rated for 160 kV anode voltage), connected in series with a 750K resistor, provides the voltage adjustment mechanism. This tube is controlled by bootstrap circuitry virtually identical to that discussed in connection with the X2062J.

The 60 Hz auxiliary power requirement for the cathode and gun anode equipment is also summarized in Figure 17. Although not shown, it has been assumed that two more X2062J tubes will be used as shunt limiters/regulators across the output of the dc power supply (i.e., as discussed in Section V-4), thus making a total of 10 circuits similar to the one shown on the left half of Figure 17. The estimated auxiliary load is 385 kVA which must be obtained from an appropriate isolation transformer bank (i.e., 100 kV primary to secondary transformer voltage differential). In the case of the gun anode circuits, 8 separate isolation transformers of approximately 5 kVA would be required.

One final area of Figure 17 concerns limiting resistors RL1 and RL2. The purpose of these resistors is to provide additional internal arc protection to the gyrotrons. These resistors--located as close as possible to the gyrotron's cathode and gun electrodes--absorb the major portion of the energy stored in stray circuit capacitance when internal arcs occur within the tube. Neither the series switches nor the crowbar can provide this form of protection to the gyrotron, hence the resistors are a valuable asset to the overall circuit.

3.1.6 CONCLUSIONS

The significant findings of this study are summarized below for the several areas investigated.

3.1.6.1 AC/DC Power System (WBS 1.2 and 1.3)

From the power system modeling study using 12 SNBPSS power supplies and one ECRH power supply:

- The HVPS power factor correction technique used is quite effective under constant system loading conditions
- The power factor at the public utility interface becomes substantially lagging (e.g., $\bar{.96}$) when a single power supply crowbar fires.
- The common 13.8kV-3 ϕ interface voltage drops approximately 3.5% when a single HVPS crowbar fires
- Individual HVPS voltage and current internal transients under crowbar conditions are acceptably low provided the power factor capacitor spark gaps function properly
- The dc current build up from a crowbarred power supply is at an initial rate of approximately 125 amperes per millisecond. Its steady-state fault value is approximately 400 amperes.

3.1.6.2 High Voltage Activation (WBS 1.2, 1.3, 1.4)

For the case of a single HVPS turn-on (i.e., the other 12 supplies are assumed to be already on-line)--with the supply's variable voltage transformer set for full dc output from the supply--the power supply/modulator modeling showed that:

- The nonlimited dc output voltage will exceed 140kV during the turn-on transient
- A shunt limiter/regulator with 95kV limiting capability will pass a peak current of approximately 125 amperes during the turn on transient
- Unless the modulator turn-on point is properly timed with respect to the power supply turn-on transient a highly oscillatory modulator output pulse top will result
- Reasonably close timing of these two events will limit the oscillatory peaks to 2% or less of the steady-state pulse top value
- The primary current to the "snapped-on" power supply does not exceed approximately 2 times normal.

3.1.6.3 Modulator/Regulator/Gyrotron (WBS 1.4 and 1.5)

A design which incorporates individual modulator/regulator tubes for the cathode and gun electrodes of each gyrotron has the following advantages:

- It provides current limiting (i.e., approximately 110% of normal current) to the fault within the gyrotron
- It allows the faulted gyrotron to be rapidly removed from the common power supply (i.e., insures minimum internal damage will be inflicted to the gyrotron).
- It reduces any effect on the performance of the other 7 gyrotrons to negligible proportions
- It allows the faulted gyrotron to be placed back on line after any predetermined delay period

- It provides the capability of individual adjustment of the cathode/gun voltages to the gyrotrons (i.e., assures optimum gyrotron performance) by programming the drive signal to the modulator.
- It can provide a series tube voltage regulation function to each channel if desired.
- It provides the capability of time stepped turn-on and turn-off programming to the gyrotrons.

It should be noted that the following disadvantages can be attributed to this approach:

- With eight series tubes the potential for switch tube arcs is increased
- The cost of the system is increased.

The counterargument to these potential problems are, first, by operating the series tubes conservatively (i.e., as has been indicated in the candidate design) this problem should be reduced to negligible proportions, and second, the additional cost of the multiple tube system should be easily justified by the longer life and hence reduced replacement cost for the gyrotron tubes (i.e., the major objective of this design approach is to provide the gyrotrons with an "easy ride").

3.1.7 SIZE AND WEIGHT ESTIMATES

The candidate high voltage circuit described in Section 3.1.5 was briefly examined to determine its size and weight impact. Table 3 summarizes the estimates for the revised switching networks (i.e., there are two involved and they are located in areas C1 and C2 of Figure 2) and Table 4 provides an estimate for each of the revised gyrotron assemblies (i.e., there are 8 assemblies, 4 in area C1 and 4 in area C2).

3.2 WAVEGUIDE DISTRIBUTION SYSTEM

3.2.1 INTRODUCTION

The objective of this study is to explore the feasibility of developing the critical microwave components required for the practical realization of the NFTF "B" RF heating system. This system will employ gyrotron tubes operating at 60 GHz with a CW power output of 250 kW from each tube. A conversion efficiency of 40 percent is expected. Four tubes will be fed to each of the two NFTF "B" end plugs to provide a total heating microwave power of 2 MW. The actual experiments will be conducted with long pulse widths ranging between 10 MS to 30 sec with a maximum duty factor of 10 percent.

Table J
SWITCH NETWORK CABINET SIZE AND WEIGHT ESTIMATE
(2 required)

	KW	L. In.	W In.	H In.	Dia. In.	Vol. Cu Ft	Wt. No.
• Switch Tube W/Shield				30	24	7.85	150
- Fil XFMR	8	12	12	16		1.33	114
- Screen PS 1.0 kV 1A	1.0	20	12	10		1.39	40
• Driver Tube				10	9	.37	1
- Grid Resistor	10			20	6	1.31	25
- Plate PS	25	28	16	19		4.93	500
- Screen PS	.8	20	12	10		1.33	114
- Fil. XFMR	1.4	7	6	9		.22	20
• Pre Driver Tube				3	3		
- Plate PS	.4	12	18	.6		.75	20
• Cooling Ass'y		48	24	8		5.33	150
• Total-four Switch Ass'ys						245	4536
• Isolation XFMR	150	36	46	46		44	4550
• Cabinet Shell		180	72	96		720	3550
• Loaded Cabinet						720	12,586

Table 4
GYROTRON ASS'Y SIZE AND WEIGHT ESTIMATE
(8 required)

	KW	L In.	W In.	H In.	Dia. In.	Vol. Cu Ft	Wt. No.
• Gyrotron		100			14	8.91	330
- Fil. XFMR	.1	4	4	3		.03	5
- Fil. Reg. Chassis	.1	15	15	10		1.30	20
• Mod. Anode. Reg. Tube				20	8	.58	20
- Fil. XFMR	.4	5	5	6		.09	17
• Driver Tube				10	9	.37	1
- Fil. XFMR	1.4	7	6	9		.22	20
- Plate Supply	2.5	12	12	9		.8	80
- Screen Supply	.1	10	12	6		.4	40
• Pre Driver Tube				3	3	--	--
- Plate Supply	.05	10	12	6		.4	40
• Isolation XFMR	5	19	27	28		8.3	700
• Gyrotron Magnet (Main)	200	14			16	8.2	800
• Magnet PS	200	36	36	108		81.0	3700
• Cooling Ass'y		48	30	12		10.0	300
• Cabinet Shell and Lead		72	60	108		270	3000
Loaded Cabinet						270	9073

Each of the eight gyrotrons and its associated microwave output transmission components can be considered a separate and essentially identical subsystem.

It is anticipated that the tube output power will be predominately in the TE_{02} mode. Because of the high output power the output circular waveguide will have a diameter of 2.5 inches and will be backfilled with 1 atm of dry nitrogen to give a peak power handling capability of 4.0 MW.

Figure 13 shows a schematic diagram of the waveguide system. This diagram is intended only for discussion, the individual components not necessarily being arranged in their optimum position.

Arc detectors are placed at each end of the system to detect arcs and turn off the system before damage occurs to the windows. These detectors employ standard designs used in most high power microwave systems, and will not be discussed further here.

Two diagnostic couplers are employed to determine the total output power and modal content at a point near the output of the gyrotron and at a point near the input to the fusion vessel. These couplers are discussed in more detail in Section 3.2.2.1. Diagnostic coupler 1 permits the monitoring of output power and mode spectrum during the adjustment of the tube when power is switched to the calorimeter load of the load switch. Diagnostic coupler 2 is employed for monitoring power and modal spectrum close to the fusion vessel. This coupler is shown connected to the output of the mode converter, in which case the spectra of diagnostic couplers 1 and 2 would be very different. On the other hand, an external mode converter may not be employed, any desired mode conversion being accomplished inside the fusion vessel. In this case the modal spectra indicated by the two diagnostic couplers would also differ to some extent because of the effect of the mode absorber which removes some of the undesired modes produced by the tube and as a result of mode conversion by the waveguide components.

Three 90° bends are required in the system configuration being considered. These bends are considered a critical item and are discussed in Section 3.2.2.2.

The load switch is discussed in Section 3.2.2.3 and as already stated allows the power to be switched to a calorimeter load during tube adjustment.

The nitrogen pressurization equipment is also standard. With one atmosphere of dry nitrogen the peak power handling capability of 2.5" diameter waveguide is roughly 4.0 MW.

Figure 18. Waveguide System Schematic Diagram.

A discussion of undesired mode absorbers is given in Section 3.2.2.4. The undesired mode absorber can be designed to selectively absorb all modes except those of the TE_{on} mode family, and can serve 3 purposes. First, undesired modes generated by the tube and by mode conversion in the waveguide components can be removed before the high power enters the fusion vessel. Second, it is expected that a high percentage of the power reflected back from the fusion vessel will be contained in higher order modes. This power would also be absorbed by the mode absorber before it can return to the tube. Third, a mode absorber can dampen the resonances of higher order modes that can be trapped between system components, e.g., between two 90° corner bends. It may be important to dampen these resonances in order to prevent overheating and possibly arcing.

There are now strong indications that the modal content of the power entering the fusion vessel is not a critical consideration, total power being the primary factor. In this case, removal of undesired modes may not be important. Furthermore, convincing arguments can be made which indicate that the power reflected back from the fusion vessel will be small under most conditions with or without a plasma present. This follows from the fact that the 2.5" waveguide being many wavelengths in diameter has inherently only a small aperture reflection even when radiating directly into free space with no additional impedance matching elements. When this same waveguide is employed to radiate into the large fusion vessel, the reflected power will still be low, because the radiated power will be scattered into many directions with very little power finding its way back to the feed waveguide. On this basis, it appears that a mode absorber may be needed only to suppress the resonances of trapped modes. As indicated in Section 3.2.2.4, it may be possible to effectively dampen these resonances with a section of stainless steel waveguide.

Mode converters are discussed in Section 3.2.2.5. These may be needed primarily near the input to the fusion vessel to convert from the TE_{on} mode family to another mode which can radiate a pencil beam.

At this time, it is not clear whether a vacuum window will be needed at the input to the fusion vessel. These windows can have design similar to that employed for the tube window. For this application a pair of BeO plates are employed with Fluorocarbon coolant in between. It is generally believed that this design can be made to handle the required power levels with sufficient development effort.

3.2.2 DISCUSSION OF CRITICAL SYSTEM COMPONENTS

3.2.2.1 RF Power Diagnostics Coupler

It is desirable to be able to measure the total absolute power and modal spectrum generated by the gyrotron and transmitted into the fusion vessel. However, this task is made difficult because of the very large number of modes that can propagate in the oversize waveguide that must be employed in order to handle the very high peak and average power levels. It may also be desirable to measure the power and modal spectrum reflected from the fusion machine.

The following discussion is intended to be only preliminary and the approaches discussed should not be considered as being optimum proposed approaches.

Limitations of Simple Waveguide Wall Coupler Approach

Figure A1 shows one possible configuration based on this approach. In this configuration N longitudinal slots are arranged in a circumferential array to measure the total circumferential current produced by the TE_{mn} modes on the cylindrical wall of the oversize waveguide of diameter D . Since TM_{mn} modes have only longitudinal wall currents, these modes will not excite the longitudinal slots. Each of the N slots can be coupled to a separate standard size rectangular waveguide and the phase and magnitude of the output of each waveguide can be measured and the data analyzed to determine the characteristics of the modal spectrum. However, it will be clear from the following discussion that the data obtainable from the simple configuration of Figure 19 cannot yield a complete analysis of the mode spectrum.

The required number of slots N is determined by the number of modes contained in the spectrum. In general, it is necessary to measure at least one complex probe voltage for each mode present. Thus it is necessary to measure both magnitude and phase.

In the case of the configuration of Figure 19, phase measurements are necessary even if one wishes to measure only the total power in the TE_{on} mode family because the TE_{on} mode family is not the only set of modes which can have wall current magnitudes which are independent of circumferential angle ϕ . Thus, circularly polarized TE_{mn} mode pairs also have this characteristic. This can be understood by noting that a "vertically polarized" TE_{mn} mode varies as $\cos m\phi$ while its degenerate "horizontally polarized" partner varies as $\sin m\phi$. If these two modes are 90 degrees apart in time phase the result is $\cos m\phi + j \sin m\phi = \exp(jm\phi)$

which has unit magnitude for all θ , but a progressive phase as a function of θ . The progressive phase is the only way to distinguish this perfectly legitimate circularly polarized mode from the TE_{on} mode family based on a measurement of circumferential wall currents alone. Additional measurements with transverse slots would be needed to detect longitudinal currents of the TE_{11} mode.

The configuration of Figure 19 also has the serious limitation that the measured data cannot be employed to separate the members of the TE_{on} mode family. The problem is made even worse by the fact that the slot coupling increases with mode order. Thus, for example, the coupling to the TE_{02} mode exceeds the coupling to the TE_{01} mode by about 5.25 dB.

In order to separate the TE_{on} mode family, it would be necessary to employ several circumferential slot arrays of the type shown in Figure 19 at several points along the axis of the circular waveguide. The separation of the TE_{on} mode family would then be made from a knowledge of the wavelength differences among these modes. However, the accuracy would be poor because the wavelength differences are small in the oversize waveguide.

From the above discussion it is clear that the configuration of Figure 11 has serious limitations and cannot satisfy the requirements. It has been discussed in some detail only to record these limitations and justify a consideration of other more complicated configurations. It is clear that in order to obtain sufficient information on the modal spectrum, it is necessary to make measurements along a surface and not merely along a single line.

Diagnostic Coupler Comprising Slots on Reflecting Surface of 90 Degrees Bends

The diagnostic coupler configuration of Figure 20 employs coupling slots placed on the reflecting surface of a 90° bend to determine both the circumferential and radial variations of the waveguide fields. Furthermore, both circumferential and radial slots can be employed to measure both polarizations of the currents on the reflector surface.

This configuration clearly has the important advantage that the members of the TE_{on} mode family can be separated from one another. Furthermore, the coupling involves an interaction with the transverse waveguide fields and, therefore, is less dependent on mode index.

Figure 19. Simple Diagnostic Coupler Employing Circumferential Array of Slots on Waveguide Wall.

Figure 20. Diagnostic Coupler Employing Slot Array on Reflector Surface.

As discussed previously, the amount of information obtainable on the modal spectrum depends on the number of measurements one is willing to make [1]. However, it may be possible to gain meaningful knowledge if measurements are made only along a few radii and circumferences. Hopefully, a measurement approach can be developed for obtaining sufficient meaningful data on modal spectrum by means of magnitude measurements only.

Calorimeter Coupler

A calorimeter coupler which would allow continuous monitoring of total power also appears to be feasible. Figure 21 shows this concept. A metal plate with a coupling hole array is employed as the reflecting surface in a 90 degree oversize waveguide four-port junction. Incident high power enters port 1, reflects off the 45° perforated reflector with low loss and leaves by way of port 2. A small fraction of the incident power (perhaps - 40 dB) passes through the coupling holes to port 3 to which is connected a water load calorimeter for measuring the total heating power in the multimode spectrum. Port 4 is terminated in an absorbing load for absorbing the small power reflected by the calorimeter. Alternately, port 4 can be terminated in a second calorimeter for measuring power reflected by the fusion machine. Thus, the configuration of Figure 21 is a dual directional coupler which can be used as a power reflectometer.

Low loss coolant is passed over the coupling hole array in order to prevent warping by overheating. However, experience may show that this provision is not necessary. It should be noted that the overall structure of coupling plate and coolant assembly does not have to be impedance matched as in the case of a microwave window.

The design of the coupling hole array may not be critical. First, since the holes are small, they will have only a small effect on the output power and spectrum at port 2. Second, there is no requirement that the mode spectrum transmitted to port 3 be undistorted, provided excessive mode conversion does not occur to modes which are close to cutoff since these modes are not effectively absorbed by the calorimeter. One has the option of employing a uniform array of small holes with small spacing S designed to minimize conversion of energy from one mode to another; or, one can employ strategically placed holes to couple only to a specific set of modes incident from port 1.

Accurate measurement of the total power can be made if the coupling is the same for all modes. Under these conditions, the modal spectrum of the transmitted

Figure 21. Diagnostic Coupler Calorimeter on 90° Reflector Surface.

energy would be essentially the same as that of the incident energy. However, coupling through the holes depends on polarization and angle of incidence, and cannot be made exactly the same for all modes. Therefore, the calibration factor for total power would depend on the modal spectrum. Analysis shows that under these conditions, significant errors can occur in the measurement of the total power. For example, if the TE_{0N} family and TE_{11} modes, respectively, have 1.0 and 2.0 units of power the error can range between 10 and 77 percent depending on the polarization of the TE_{11} mode.

Two approaches are under consideration for overcoming this problem. In the first, the use of a nonuniform distribution in the coupling may lead to smaller dependence of coupling calibration on the polarization of the incident mode. In the second approach, a pair of four-port junctions containing 45° reflectors are cascaded as shown in Figures 22 and 23 and are employed to measure separately the horizontally and vertically polarized field components. For simplicity, rectangular waveguides cross sections are shown; however, the same conclusions apply for circular cross sections.

Figure 22. Dual-Reflector Absolute Power Monitor

Figure 23. Coplanar Absolute Power Monitor

In the configurations of Figure 22, the two cascaded four-port junctions are not coplanar and result in a net 90° change in direction as well as a small offset. High power containing both horizontally and vertically polarized components represented by magnetic fields H_H and H_V , respectively, is incident on port 1a of junction 1. The 45° metallic reflectors contain regular periodic arrays of narrow slots having spacing S and length l which is large compared to the slot width W in order to provide polarization selectivity. The slot length l is made small compared to wavelength so as to limit the power coupled to port 1c to a very small value (e.g., -40 dB or less). The spacing S is also made small compared to wavelength to limit so-called grating waves. With l made small, high power incident on port 1a in either polarization reflects to port 1b with low loss. Since the slots in reflector 1 are vertically oriented, H_H couples extremely weakly (ideally zero) to port 1c. However, since H_V is parallel to these slots, a finite magnetic field coupling of value of k (e.g., -40 dB) results between ports 1a and 1c for the vertically polarized fields H_V .

If the modes coupled to port 1c are reasonably well above cutoff, the calorimeter will accurately measure the total power coupled to port 1c, and the power coupling coefficient will be k^2 for all modes incident from port 1a. Thus, the absolute power indicated by the calorimeter will not depend significantly on the modal spectrum of the high power incident on port 1a.

In the same way, the high power reflected from junction 1 will enter port 2a of junction 2 and will reflect from 45° reflector 2 to port 2b. Reflector 2 is oriented to provide the same conditions for H_H as experienced by H_V at reflector 1. Thus, in both cases the magnetic field that is being coupled lies parallel to the plane of the reflector and also parallel to the long dimension l of the slots. The angle of incidence is approximately 45° in both cases assuming all incident modes are well above cutoff. The reflection and transmission coefficients are therefore identical at reflectors 1 and 2 for H_V and H_H , respectively. Thus, the power calibration factors for the calorimeters at ports 1c and 2c are identical and equal to k^2 for the vertically and horizontally polarized power, respectively.

It should be noted that calorimeters placed in ports 2d and 1d function in the same way for power reflected from the load on port 2b as the incident power calorimeters on ports 2c and 1c, respectively.

Figure 23 shows an alternate configuration in which the pair of four-port junctions are coplanar, and result only in a small offset between input and output high power waveguides. Note that junctions 1 in Figure 22 and 23 are identical, only junctions 2 being different. In junction 2 of Figure 23, the magnetic field H_H which has a component parallel to the long dimension l of the slots, does not lie in the plane of the reflector. Under these different conditions, the coupling coefficients k_V and k_H at reflectors 1 and 2, respectively, cannot be expected to be equal. However, analysis shows that k_V and k_H each are essentially the same for all modes. Accurate measurement of the absolute power can therefore be accomplished by calibrating both k_V and k_H separately.

An estimate of the relative values of k_V and k_H can be obtained by the following arguments. First, with the narrow slots employed, the coupling results only from the magnetic polarizability of the slots. That is, coupled power depends only on the tangential magnetic field component parallel to the long dimension of the slots on the reflector surfaces and does not depend on the normal electric field. Second, note that the angle of incidence θ is essentially 45° at both reflectors for all incident modes reasonably well above cutoff. The tangential magnetic field at reflector 1 is equal to H_V and is independent of θ . In this case, k_V in Figure 23 is equal to k in Figure A4 and both are independent of θ and, therefore, the same for all modes.

On the other hand, the tangential magnetic at reflector number 2 in Figure 23 is $H_H \cos \theta$ and the value of k_H is, therefore, 6 dB less than that of k_V . It is also noted that k_V is approximately equal to the value of k obtained for normal incidence provided the slot spacing S is small enough to prevent grating lobes from occurring, the maximum value of S being $\lambda/(1+\sin\theta) = 0.586 \lambda$ where λ is the free space wavelength.

It may also be possible to combine the configurations of Figures 20 and 22 (or 23) to obtain a single diagnostic coupler which can measure total power reasonably accurately and also establish a reasonable estimate of the modal content.

The diagnostic couplers must be calibrated by some means. The absolute coupling through a small hole can be calculated with an accuracy of only 1 or 2 dB compared to actual experimental values. Therefore, the absolute power calibrations should be done by comparison with an accurate calorimeter. However, the calculated coupling can provide a reliable check on the measurements.

3.2.2.2 90 Degree Bends

Bends for waveguides having cross-sectional dimensions which are very large compared to the free space wavelength can be analyzed by resorting to quasi-optical concepts. For example, quasi optical concepts can be applied to obtain accurate results if the diameter D of a circular waveguide exceeds perhaps 3 to 5 free space lengths λ .

The losses in a bend comprise dissipation loss due to the finite conductivity and mode conversion loss. The dissipation loss is usually not significant, the major loss component being due to mode conversion which decreases as the cross-sectional dimensions or frequency is increased.

The mode conversion loss for the TE_{01} mode in a quasi optical circular waveguide bend is given by [2]

$$L(\text{dB}) = 2.4 \left(\frac{2\lambda}{D} \right)^{3/2}. \quad (1)$$

Equation (1) is plotted in Figure 24. These data show that the mode conversion loss for the TE_{01} mode transmitted around a 90 degree plane-mirror bend in 2.5 inch diameter circular waveguide is 0.15 dB at 60 GHz ($D/\lambda = 12.71$). This amount of loss would be tolerable for the present application. However, the gyrotron that will be employed will have output mostly in the TE_{02} mode. In response to this, Dr. Bernard Arfen* has recently extended the analyses of Katsenelenbaum to the TE_{02} mode. His result for a 90° bend is

$$L(\text{dB}) = 8.034 \left(\frac{2\lambda}{D} \right)^{3/2}. \quad (2)$$

Equation (2) which is also plotted in Figure 24 indicates that the mode conversion loss in decibels for the TE_{02} mode is $8.034/2.4 = 3.3475$ times that of the TE_{01} mode. For the present application at 60 GHz, the TE_{02} mode loss with $D = 2.5$ inches would be about 0.5 dB which may be considerably higher than desired.

The reduction in mode conversion that can be obtained by using curved mirrors instead of plane mirrors has been investigated [3], [4]. Bends in rectangular waveguide are discussed in [3]. It is indicated that the mode conversion loss in decibels for the TE_{10} mode in an H-plane bend using a curved mirror can be reduced

*Unpublished notes by Dr. Arfen who is with the Lawrence Livermore Laboratory.

by a factor of about 1.5 compared to the decibel loss in a plane mirror bend. Curved mirror bends for TE_{01} mode in circular waveguide are discussed in [4], and these results also indicate lower loss for curved mirrors than for plane mirrors. However, the data apply only for D/λ between 1.77 and 2.57. No data is available for the losses for the TE_{02} mode in waveguide bends employing curved mirrors. More work is clearly called for in this area.

Cullen [5] has studied mirror bends employing relatively large ellipsoidal reflectors as shown in Figure B2. Using circular waveguides having D/λ of 5.9 the calculated mode conversion loss is 0.25 dB for the TE_{01} mode. Reference to Figure B1 shows that the corresponding loss for the TE_{01} mode in a waveguide bend with plane mirror is about 0.475 dB. In an experiment at 35 GHz, the measured loss was 0.5 dB, but Cullen attributes some of this loss to dissipation and mode conversion in transducers and tapers employed in the experiment.

Figure 24. Mode Conversion Loss for 90° Plane Mirror Bend in Circular Waveguide.

Figure 25. Quasi Optical Bend Employing Ellipsoidal Reflector.

Cullen has also calculated the loss for the TE_{02} mode in the ellipsoidal reflector bend. One set of calculations shows about 1 dB loss. However, it is not clear that the reflector was optimized for TE_{02} mode transmission. More work is required to fully evaluate the potential of the ellipsoidal reflector. Furthermore, it is not clear that an ellipsoidal shape is optimum for the reflector.

Some thought should also be given to employing a reflector system (ellipsoidal or other shape) of Figure 25 as part of the output system of the gyrotron tube. In this case, the reflector surface could serve as the electron beam collector. Alternately, the electron beam could be deflected downward magnetically to be collected on an auxiliary metal plate placed below the ellipsoidal reflector well out of the way of the microwave beam. Because the reflector system is "open" the spurious mode resonance problem would be much less severe than in one current approach which employs back-to-back waveguide tapers and is therefore a "closed" system.

3.2.2.3 RF Load Switch

The load switch is required for the insertion of a matched water load calorimeter into the waveguide system during the time the gyrotron and associated equipment is being adjusted.

Figure 26 shows one possible configuration which employs a sliding switch element. In the position shown, the sliding element provides a straight section of circular waveguide to cause low loss transmission between ports 1 and 2. If the sliding element is moved to the left in Figure 26 a 45° reflecting surface will be provided to cause low loss transmissions between ports 1 and 3.

The design of the switch is not critical. In both switch positions back scatter to port 1 will be very small. Furthermore, very low loss can be achieved in the position for power transmission between ports 1 and 2. The small amount of mode conversion that can occur in the switch position for transmission between ports 1 and 3 is not important since the calorimeter will measure the net heating power.

The choke grooves shown in Figure 26 are not needed for the TE_{on} mode family, but may be used to prevent leakage of spurious modes that may be present in the waveguide system. If necessary, glossy material may be placed within the housing to prevent undesired high-Q absorption resonances from building up in the enclosure.

The switch can be designed to operate under the required conditions of nitrogen pressurization by enclosing the water load calorimeter in the pressurized region as indicated in Figure 26. Flexible hoses, not shown in the figure, will supply water to the water load calorimeter in a manner that will allow the water load to move freely with the sliding switch element.

Alternately the switch can be designed so that the water load is outside the nitrogen pressurized region. A flexible bellows would be employed to allow the necessary movement of the sliding element and to provide a pressure seal between the housing the flange at port 3.

A previously built oversize waveguide load switch of the sliding type employed linear bearings sliding on stainless steel rods to provide precision alignment and mechanical support for the sliding element and the water load. The switch was activated by an electrically controlled air cylinder. Positive stops were used to assure accurate alignment of the waveguides in the two switch positions and micro-switches were used to provide assurance that the switch was in the desired position.

There do not appear to be any fundamental problems in the development of a load switch of this type for application at millimeter wavelengths.

Figure 26. Sectional View of Load Switch.

3.2.2.4 Absorbers for Undesired Modes

As stated in Section 3.2.1, mode absorbers may be employed in order to remove modes other than the TE_{on} family generated by the tube and waveguide components, and in order to reduce the total power reflected back to the tube from the fusion vessel. Mode absorbers can also be used to dampen the effects of trapped mode resonance which can cause overheating and arcing.

Mode absorbers for circular waveguide have been developed in the past which employ circumferential slots to interrupt the longitudinal currents of undesired modes. 360° slots can be obtained by a stacked disc configuration. These slots can be terminated in matched loads to obtain a broad band wave impedance at the input to the slot.

For 360° circumferential slots with free space wave impedance, the attenuation for the TE_{11} mode is on the order of 8.686 g/s decibels (dB) in a length equal to the diameter D of the waveguide where g is the slot (or groove) width and S is the slot spacing. Thus, if g/S is 0.3 an attenuation of 2.6 dB is obtained in a length equal to D , corresponding roughly to 1 dB per inch in 2.5" diameter waveguide. These calculations assume free-space slots. Mismatched slots can be employed to reduce the attenuation per unit length. In this way, the absorbed power may be dissipated more easily over a longer length. Furthermore, the attenuation should be tapered along the slot array in order to equalize the power absorbed by each slot. Design considerations, therefore, are similar to those for a high power matched load.

An alternate approach for mode absorbers is to employ smooth metal wall waveguides made of material such as stainless steel. In this case relatively higher loss would be obtained for the undesired modes than for the TE_{on} mode family. The attenuation in decibels in one diameter for several modes in smooth circular waveguides is listed below. These expressions for attenuation have been simplified by assuming that all modes are well above cutoff.

MODE TYPE	ATTENUATION
TE_{on} Family	$1.732 R_s / \gamma (fc/c)^2$
All TM_{mn} Modes	$1.732 R_s / \eta$
TE_{11} Mode	$1.455 R_s / \eta$

R_s/η is the dimensionless ratio of surface resistance to free space wave impedance (120π) and is given by

$$\frac{R_s}{\eta} = 21.88 \times 10^{-6} \sqrt{F(GH_z)/\sigma r}$$

Where F is frequency and σr is conductivity relative to that of copper. Using this relation, the attenuation of TM_{mn} modes far above cutoff in stainless steel circular waveguide is 0.018 dB in a length equal to one diameter. Thus, the attenuation in 10 feet of 2.5 inch diameter waveguide is 0.864 dB. This is not sufficient loss for the undesired modes if a highly pure TE_{on} mode family spectrum is desired, but can be sufficient to dampen resonances caused by trapped modes.

3.2.2.5 Mode Converters

For the present purpose a mode converter can be defined as any component employed to transform from one mode or set of modes in a closed cylindrical waveguide or in an open beam waveguide. Any component employing a means for converging an antenna radiation pattern from one form to another is also considered a mode converter. Mode converters may be required to convert from a given gyrotron output mode to another mode more advantageous for low loss waveguide transmission. A mode converter may also be desirable at the input to the fusion vessel to obtain optimum heating conditions.

In the past, mode converters have been employed for waveguide systems having relatively small cross-sectional dimensions (i.e., $D/\lambda \leq 2.0$) [6]. These types of transducers cannot handle the required power levels and, furthermore, the basic designs cannot be extended to the large values of D (i.e., $D/\lambda > 10$) required in the present application. Mode converter design for the TE_{on} mode family for very large D/λ must recognize the fact that these modes are coupled only very weakly to the cylindrical waveguide walls. These designs must therefore be based on an interaction with the fields far removed from the cylindrical walls.

Wengenroth [7] has followed this approach in the design of what he called a "Mode Transducing Antenna." By stepping and grooving the surface of a radar reflector antenna, Wengenroth showed that a TE_{01} mode radiation pattern which has a hole on axis can be converted to a reasonable pattern having maximum radiation on axis. Circular polarization can also be obtained. As stated by Wengenroth the results presented in [7] are very preliminary, essentially representing proof of principle. Further work is clearly indicated along these lines.

During the course of the present short study the question has also been considered as to whether Wengenroth's approach can be employed if the incident mode is one of the higher order TE_{on} modes. A qualitative answer to this question can be obtained if one considers the limiting case of employing a very high order incident mode. In this case the use of phasing steps would result in narrow regions of "rectified" reflector currents having high periodicity corresponding to mode index n . The resulting "pencil" beam radiated would, therefore, have lower side lobes than would be obtained with small index n . On this basis alone, it seems that conditions may be slightly better with TE_{02} than with TE_{01} mode incident. More significant improvement might be expected with higher values of n , but in this case the problem of alignment of the incident mode fields with the reflector phasing elements becomes much worse.

3.2.3 REVIEW OF STATUS OF GYROTRON TUBE DEVELOPMENT

Discussions on the current status of gyrotron development were held with groups at the Naval Research Laboratory and at Varian Corporation. As a result of these discussions the following areas requiring further intensive study and development can be identified:

1. Peak and average output power capabilities
2. Separation of electron beam and microwave output power
3. Frequency and phase stability
4. Amplifier operation

The discussions revealed that the fundamental limitations on peak and average output power capability may be determined by the necked-down diameter in the microwave/electron beam region. The use of the necked-down interaction region reduces the problem of beam interception in the interaction region by eliminating the "reflectors" that would be required if a "longitudinal" standing - wave resonator were employed.

Current practice [8], [9], is to make the minimum diameter in the necked-down region equal to a value only slightly greater than the cutoff diameter of the desired TE_{mn} operating mode. Thus:

Mode	Cutoff Diameter (Wavelength)	Diameter for 60 GHz (Inches)
TE ₁₁	0.586	0.1153
TE ₀₁	1.219	0.2398
TE ₀₂	2.233	0.4393
TE ₀₃	3.238	0.6370

The TE modes develop very high electric fields and very low transverse magnetic fields close to cutoff, and therefore interact strongly with the electron beam.

The above data show that the higher order modes require a larger diameter, implying that the upper limit on power generation may be higher with these modes. Currently this is an important consideration in selecting the output mode. Apparently, the TE₀₁ and TE₀₂ modes are considered good compromises between power output limitation and usefulness of the output mode, in a practical system, modes of higher order having unacceptable mode pattern and the lower order TE₁₁ mode being too limited in output power.

Currently, the problem of separating the microwave power and collecting the electron beam can cause a limitation on the output power that can be achieved. However, it is believed that these problems are not fundamental in nature, and can be resolved by further engineering and development effort. The use of the ellipsoidal reflector configuration on Figure B2 may represent one possible approach.

There is general optimism concerning the frequency and phase stability of the basic gyrotron oscillation mechanism. Although extensive data on oscillator line widths and noise figures have not yet been recorded, available data indicate very acceptable characteristics. It has been said that in these respects a gyrotron is "better than magnetron."

The principal stability problem apparently originates from mode jumping, and this is a principal reason for employing the restricted diameter cutoff resonator for the interaction region.

There is not universal optimism concerning the ease with which stable amplifier operation will be obtained. Principal problems cited in this case result from small back scatter at the tube output.

Power levels achieved to date include 105 kW CW at 28 GHz with an efficiency of 17 percent. Under more optimum conditions which can be realized in practice, CW efficiencies of 40 percent are expected. Forty millisecond pulsed operation has also been demonstrated at 28 GHz with 200 kW peak, and 10 kW average power. Efficiency in this case was 37 percent. It is projected that 200 kW CW at 28 GHz will be obtained by September 1980 under a current program. Pulsed operation with 200 kW peak power at 60 GHz is projected for about 1982. During the 1983-84 time period it is hoped that 200 kW CW will be achieved at 110 GHz.

3.2.4 SUMMARY

The feasibility of the required microwave components for the RF heating system of the MFTF-B has been studied. The results of the study clearly show that further conceptual design and analyses are required in the following areas:

1. Diagnostic Couplers - Further study is required to develop practical methods for measurement of the modal spectrum. Work is also needed to develop accurate couplers for measurement of total power in a multi-mode spectrum.
2. 90° Bends - These are also a critical item especially for the TE_{02} mode. This problem clearly requires further study. The use of ellipsoidal reflectors for gyrotron outputs should also be explored.
3. Load Switch - This is an important component, but poses no difficult problems.
4. Mode Converters - The importance of these components depends on what kind of excitation will prove to be optimum for the fusion vessel.
5. Mode Absorbers - may be required principally to dampen trapped mode resonances which can cause overheating and perhaps arcing. Mode absorbers for this purpose are required to provide only relatively small attenuation of undesired modes, and can comprise a length of stainless steel waveguide. These can be placed between 90° bends where significant trapping can occur. The bends themselves could be made of stainless steel with extended straight stainless steel sections on one or both ports.

3.2.5 REFERENCES

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SECTION 4

PROGRAM COST ESTIMATES

Table 5 shows a work breakdown structure which has been generated for the ECRH system and Figure 27 presents a program plan for development, design, fabrication, test and installation of the equipment.

A schedule for the program has been developed and is shown in Figure 29. The schedule spans 46 months, from start of the program until all equipment is shipped to LLL for installation. The first 24 months are spent developing the system and the last 22 months producing the equipment for MFTF-B. Important assumptions had to be made in developing this schedule. These assumptions are considered to contain elements of risk and thus the schedule could be subject to considerable variation.

The schedule in Figure 28 assumes that the feasibility of the gyrotron tube at a power level of 250 kW will have been proven before the program starts. Provided this is so, then it should be possible to further develop the tube into a manufacturable product. The schedule calls for two tubes (possibly subspecification) to be made available at months 18 and 20 for checking out the ECRH system equipment. This time period is comparable to that required to develop high power traveling wave tubes for a radar system. When operation of the gyrotrons with the transmitter test set is evaluated and the set is operational, it can be made available for the design verification of the high-power microwave components in an integrated system test.

The rough order of magnitude cost estimates which were developed are shown in Table 6. These cost estimates are purely engineering estimates based on past experience. Therefore, these costs are subject to wide variation when a carefully prepared cost estimate is developed. Effects of inflation are not included.

Table 5
WORK BREAKDOWN STRUCTURE

1. Program Management
 - Program Manager
 - Contract Administration
 - Procurement
 - Financial Analysis
 - Schedules and Plans
2. Project Engineering and Equipment Integration
 - Design Guidelines/Subsystem Specification
 - Standards
 - Safety
 - Test Requirements
 - Test Plan
 - Quality Assurance
 - Environmental Impact
3. Development
 - 3.1 Gyrotron
 - 3.2 Transmitter Test Set
 - 3.3 Waveguide Components
4. Gyrotron Transmitter Equipment
 - 4.1 Gyrotron Cabinet
 - (1) Gyrotron
 - (4) Magnet P.S. #1 (Main)
 - (2) Magnet P.S. #2 (Gun)
 - (1) Magnet P.S. #3 (Collector)
 - Mod Anode Regulator and Control
 - Cooling Assembly
 - Cabinet Shell and Accessory
 - 4.2 Switch Tube Cab
 - Switch Tube and Assembly
 - Driver and P.S.
 - Pre-driver and P.S.
 - Cooling
 - Isolation Transformer
 - Cabinet Assembly
 - 4.3 Engineering and Drafting
5. Waveguide Equipment
 - (2) Window Arc Detector
 - (2) Diagnostic Coupler
 - (3) 90 Degree Bends
 - Undesired Mode Absorber
 - Mode Converter
 - Vacuum Window
 - N₂ Pressurizer Post
 - 2.5" Dia. Waveguide
 - Load Switch
 - Calorimeter Load
6. Documentation and Data
7. Installation and Checkout
8. Field Test

Figure 27. Program Plan MFTF-B ECRH Equipment

ECRH DEVELOPMENT AND PRODUCTION SCHEDULE

Figure 28. ECRH Development and Production Schedule

Table 6
MFTF-B ECRH
Cost Estimate

Program Management	0.3M
Project Engineering and Equipment Integration	0.5M
Development	
Gyrotron	Not Costed
Gyrotron Transmitter Test Set	1.5M
Waveguide Components	0.6M
Gyrotron Transmitter Equipment	
Gyrotron Cabinet* 8 x \$200K	1.6M
Switch Tube Cabinet 2 x \$253K	.5M
Engineering and Drafting	1.2M
Waveguide Equipment	1.3M
Documentation and Data	Not Costed
Installation and Checkout	Not Costed
Site Testing	Not Costed

Also not Costed: Modification to H.V.D.C. P.S.

Monitor and Control
Modular Switch and Crowbar
Site Cooling System

*Includes gyrotron and magnets at \$100K.