DESIGN AND OPERATION OF A 45 μs REPETITIVELY PULSED 12 MW ELECTRON BEAM FOR A CO₂ LASER

Randy D. Curry

A Thesis Submitted for the Degree of PhD at the University of St Andrews

1992

Full metadata for this item is available in St Andrews Research Repository at:
http://research-repository.st-andrews.ac.uk/

Please use this identifier to cite or link to this item:
http://hdl.handle.net/10023/10972

This item is protected by original copyright
DESIGN AND OPERATION OF A
45 µs REPETITIVELY PULSED
12 MW ELECTRON BEAM FOR A
CO₂ LASER

A thesis presented by
R. D. Curry, M.S.E.E.
to the
University of St. Andrews
in application for the degree of
Doctor of Philosophy
January 1992
DECLARATION

I hereby certify that this thesis has been composed by me, and is a record of my original work. The material presented in this thesis has not previously been presented for a higher degree.

Randy D. Curry
CERTIFICATION

I certify that Randy D. Curry has spent nine terms of research work under my supervision, that he has fulfilled the conditions of Ordinance No. 16 (St. Andrews University), and that he is qualified to submit the following thesis in application for the degree of Doctor of Philosophy.

Dr. A. Maitland
Department of Physical Sciences
University of St. Andrews
ACKNOWLEDGEMENT

It is with great appreciation that I acknowledge the encouragement and advice given by Dr. A. Maitland in preparation of this thesis. I also gratefully acknowledge the help and advice of Professor H. Menown, who has over the years contributed to my knowledge of thyatrons in high energy applications.

The success of a project of the size described in this thesis owes much to many dedicated engineers and scientists in addition to the author. Some of these significant contributors are listed in the two papers of Appendix A and I wish to thank them for the crucial help they generously gave me from time-to-time. I should also like to thank Steve Payton for his input.

I would also particularly like to express my appreciation for the help and advice given by Dr. L. Schlitt, Mr. L. Demeter and Mr. W. Weseloh during the research program described. The help and advice of my wife Noreen Curry is also sincerely appreciated.
AUTHOR’S CAREER

Randy D. Curry was born in Lubbock, Texas in 1960. After attending local public schools, he completed his secondary education in 1978. Upon graduation, he attended Texas Tech University and completed a Bachelor’s of Science Degree in 1982. Following completion of his undergraduate degree, a higher degree was undertaken. In 1984 a Master’s of Science Degree was completed at Texas Tech University, with an emphasis in pulse power engineering.

Mr. Curry began his research career as an undergraduate research assistant. While working on a Bachelor of Science Degree, he supported numerous research programs in the areas of plasma physics, CO₂ laser heating of plasmas, mass spectrometry of spark-gap by-products and tokamak research. Upon entering the Master’s program at Texas Tech University, Mr. Curry’s research career continued in the area of triggered surface discharge switch physics.

Since completing his Master’s of Science degree, Mr. Curry has worked for three commercial companies. While working in the private sector, Mr. Curry has completed research programs in the area of high voltage opening switch research (9 MV), high power modulator development for both, excimer laser and electron beam accelerators, and laser-triggered picosecond risetime solid-state switches. As both author and co-author, Mr. Curry has completed over 20 publications during his research career.
ABSTRACT

The parameter space over which electron beams operate for pumping CO₂ lasers has been significantly extended. A repetitive pulsed power/electron beam system for the generation of 45 µs, 220 keV, electron beams for CO₂ lasers has been designed and implemented which can operate over a 6:1 impedance range. The system consists of a 150 kV - 250 kV modulator and a large area, (2500 cm²) 5-20 mA/cm² hot-cathode electron-beam gun. The system is designed and fabricated to be used in conjunction with an electron beam sustained laser and is capable of operating continuously at up to 10 pps.

The following thesis examines, in detail, the experimental and theoretical investigation of the modulator/electron beam subsystems which comprise the electron beam system. Parameters of the CO₂ laser which determine the electron beam parameter space, and thus the modulator characteristics, are reviewed. Both cold-cathode and hot-cathode pulsed current parameters are reviewed, and the cathode characteristics upon which the hot-cathode was selected are discussed. The 132 thoriated tungsten filament, grid-controlled, hot-cathode gun used to generate the required 5-20 mA/cm², 250 cm length source of electron beam with a spatial flatness of ± 10% is shown.

Based on the hot-cathode electron beam gun design, the theoretical and practical design characteristics of the hollow-cathode, thyratron switched, 1800 J, 10 pps, 250 kV modulator are shown. Design equations which allow
selection of the modulator operation range, and selection of the modulator matching resistors used to match the fixed impedance modulator to the 6:1 electron beam impedance range with a pulse droop of less than 0.7% and a pulse flatness of ± 10% are derived. Moreover, a detailed discussion of the circuit simulations used to both tune and optimize the modulator’s impedance tapered, transformer-coupled Type E PFN to the four electron beam operating ranges are shown. Practical layout and design of the modulator is discussed along with the analysis of the modulator’s inductor design and the grounding and shielding principles employed in the system. Component spacings required to insure reliable modulator operation at 250 kV are also considered.

The design and implementation of the 45 µs, SCR switched, transformer-coupled PFN based grid-pulser are discussed. The grid pulser characteristics including the 105-650 V, 28-166 A, pulse required to control the 5-20 mA/cm² hot-cathode current density are examined. Theoretical design equations are derived to allow, the Miller capacitance, and thus the time varying impedance of the grid-cathode region to be predicted. Circuit simulations showing the grid-pulser circuit interaction with the constant perveance, Miller capacitance dominated, grid-cathode load are discussed. Final layout of the SCR-switched, grid-pulser including design of the spiral inductors used for the PFN is reported. Specific layout details are shown which allow the grid pulser to operate reliably while floating at the 220 kV electron
beam acceleration pulse.

A discussion of the subsystem experimental integration phase conducted at Pulse Sciences, Inc. is also presented in the following thesis. Test results of the modulator and grid-pulser into their respective electron beam loads are compared with the circuit simulations. Relevant vacuum flashover physics issues found during the experimental integration of the modulator, grid-pulser, and electron-beam gun are considered. Specific discussions of the vacuum gap and vacuum insulator physics which effect the operation of the large area electron beam gun are reported. Included in this discussion is the experimental cold-cathode conditioning procedure used to condition the electron-beam gun to 250 kV. In-situ, hot-cathode filament carburization and activation procedures developed during the integration phase of the program are also presented, with a short discussion of the background mass spectra measured during the experiments. Finally, the experimental characterization of the spatial uniformity of the electron beam is reported along with a short summary of the investigation results.
**TABLE OF CONTENTS**

Chapter

1 Introduction 1
   1.1 Motivation for High Reliability 1
   System Design 5
   1.2 Introduction to CO\textsubscript{2} Laser Discharges 5
   1.3 Overview of Thesis 7
   References for Chapter 1 10

2 Design Considerations for the Generation of E-Beams in the 1-100 Microsecond Ranges 12
   2.1 Introduction and Requirements 12
   2.1.1 Cold and Thermionic Cathode Process 15
   2.1.2 Thoriated Tungsten Cathodes 22
   2.1.3 Activation of Thoriated Tungsten Cathode Filaments 23
   2.2 Overview of the Pulse Power and Hot Cathode Electron Beam System 24
   2.3 Modulator Current and Voltage Requirements 38
   References for Chapter 2 44

3 Circuit Modelling and Design of the 250 kV Modulator 46
   3.1 Introduction to Pulse Forming Networks for Generation of Electron Beams 46
   3.2 Impedance Matching of Pulse Forming Networks into a Constant Current Load 53
3.3 Circuit Design of the 250 kV, 45 μs Modulator for the Constant Current, Grid-Controlled Cathode Load 62

3.3.1 Circuit Analysis and Simulation of a 52 μs, 5.5 Ω Tapered Impedance Type E PFN 72

3.4 Layout and Design of the 250 kV, 52 μs Modulator 89

References for Chapter 3 100

4 Circuit Modelling and Design of the Grid-Pulser 102

4.1 Grid-Pulser Design Considerations 102

4.2 Analytic Model Derivation of the Grid-to-Cathode Load 105

4.3 Simulation of the Grid-Pulser into the Grid-Cathode load 113

4.4 Grid-Pulser PFN Layout and Testing 126

References for Chapter 4 138

5 Integration and Experimental Evaluation of the Modulator/Electron Beam System 139

5.1 Experimental Cathode Conditioning Procedure 139

5.2 Modulator Characterization into a Resistive Load 140

5.2.1 Phase I: Experimental Modulator Check-Out 140

ix
LIST OF FIGURES

2.1 Photograph and layout of grid controlled electron beam gun with control-extraction grid.

2.2 Complete cathode assembly to which the high voltage 150 kV - 250 kV acceleration pulse is applied.

2.3 Layout of high voltage filament transformer bushing and cathode in relation to its vacuum chamber and Hibachi-electron beam window.

2.4 Simplified system layout showing the inter-relationship of the modulator grid pulser and electron beam gun.

2.5 Photograph of high voltage modulator mounted above its oil tank.

2.6 Block diagram of modulator and support components.

2.7 Schematic of pulse power system (modulator) including resonantly charged power supply, pulse forming network, step-up transformer and high voltage coaxial cables linking the modulator to the electron beam gun.
2.8 Energy deposition distribution versus the penetration depth of electrons for various electron energies.

2.9 Transmission and absorption curves for 150 keV - 200 keV electrons.

3.1 Equivalent forms for five-section Guillemin voltage-fed network.

3.2 Six section Type E PFN showing the mutual inductance (M) between inductors.

3.3 Approximate models of PFN used to select initial modulator operating ranges.

3.4 Fixed impedance, 52 μs, 6 Ω PFN shown along with the transformer model and stray capacitance, used to model the modulator circuit.

3.5 Tapered impedance, 52 μs, 5.5 Ω PFN model derived from iterative circuit simulation.

3.6 Normalized load voltage of fixed impedance PFN and tapered impedance PFN response into a constant current load.
3.7 Impedance tapered load voltage across the constant current load over the four operating ranges.

3.8 Simplified modulator schematic showing grid-pulser and matching resistors.

3.9 Simulated secondary modulator voltage \( V_{\text{AK}} \) into a pulsed constant current load.

3.10 Simulated modulator current into matching resistors and pulsed constant current load.

3.11 Photograph of modulator showing high voltage step-up transformer, PFN, and matching resistor.

4.1 Thevenin equivalent model of the electron beam system including the modulator and grid pulser.

4.2 Block diagram of the grid pulser and the constant perveance load into which it operates.

4.3 Block diagram of the grid pulser including the charging system, control system, and 15 section PFN.
4.4 Initial fixed impedance PFN used for the computer calculations. All values are on the secondary of the 1:2 step-up transformer.

4.5 Final tuned grid pulser PFN values from the circuit simulations. All values are on the output of the 1:2 step-up transformer.

4.6 Simulated grid-pulser PFN waveform across the constant perveance, grid-cathode load.

4.7 Grid pulser and constant perveance grid cathode load.

4.8 Photograph of grid-pulser including a side-view of the PFN.

4.9 Photograph of initial grid-pulser PFN being discharged into a 4 Ω load.

4.10 Computer simulation of grid pulser PFN response into 4 Ω load (solid line) and the actual tuned response of the PFN into the 4 Ω load (dashed line).

4.11 Photograph of final tuned grid-pulser.
5.1 Measured response of modulator into the matching resistive load; 10 μs/div, 40 kV/div.

5.2 Comparative response of modulator output into a resistive load and the circuit simulation of the modulator response into a resistive load.

5.3 The measured grid-pulser response into the space-charge limited load represented by the grid-cathode load. No acceleration voltage or electron beam is present.

5.4 Normalized modulator voltage waveforms. Solid trace is the modulator pulse into a resistive load. Dashed trace is the modulator pulse into a constant current load.

5.5 Faraday cup diagnostic setup.

5.6 Spatial uniformity of cathode including pulse-to-pulse power supply non-reproducibility, 200 keV, 18 mA/cm² current density.

B.1 Listing of the basic program used to verify the modulator operating range.

C.1 Listing of Modulator circuit simulation code.

D.1 Listing of the grid-pulser circuit simulation code.
CHAPTER 1
INTRODUCTION

1.1 Motivation for High Reliability System Design

New requirements have significantly influenced the pulse power field over the past 10 years, both in the United States and abroad. Industrial as well as laboratory pulse power systems developed for megawatt to terawatt power levels have traditionally been based upon Marx bank technology [1]. Pulsed laboratory research projects over 200 kV have relied upon Marx bank charged pulse forming lines or Marx run-down circuits [2]. Variations of the Marx circuit have been configured to generate microsecond or multi-microsecond megavolt pulses [3]. Marx bank technology, however, does not lend itself to repetitive requirements. Although megavolt Marxes have been designed and demonstrated, gas flow requirements, as well as the large number of switches, are quite complex [4]. In addition, use of charging resistors in the Marx circuit at rep-rates above about 10 Hz present problems which further complicate the design and reliability of the system. Lifetimes of $10^4-10^6$ shots for the spark gaps do not meet present commercial or scientific requirements.

Recent advances in closing switches, such as thyristors and silicon controlled rectifiers (SCR) have significantly improved the reliability of repetitive
switches to meet emerging pulse power requirements. High voltage transformer coupled modulators based on these new switches are replacing older single-shot Marx systems. Hollow anode thyatrons used for pulsed lasers can extend the lifetime of modulators designed to deliver acceleration voltages to high voltage vacuum diode loads [5]. Unusual open circuit and short circuit fault modes require that the closing switch and modulator components conduct large reverse and forward currents frequently, while operating reliably. New pulse transformer technology, when coupled with this emerging switch technology, results in high reliability repetitive pulse generators capable of megavolt, megawatt operation at kiloamperes under 10-100 pps rep-rate operation [6].

Electron beam pumped CO₂ lasers are one example of systems which require repetitive modulator pulse power systems which have long lifetimes. Electron beam pumped CO₂ laser systems require both a modulator-electron beam system and a high-current - low-voltage modulator to "sustain" the gas discharge. A modulator-electron beam system is used to preionize the laser gas. A high energy 150 keV - 200 keV, 5-20 mA/cm² electron beam is injected into the laser gas volume. The injected electron beam stabilizes the discharge, changing the E/N ratio of the CO₂ laser gas. (A more detailed discussion follows in Section 1.2.) A high-current, low-energy modulator (20-40 kV) is used to "pump" the CO₂ laser discharge imparting energy to the laser gas.
The following thesis examines the theoretical and experimental design of a repetitive modulator/electron beam system built to pump a CO$_2$ laser [Table 1.1]. The 150 keV - 200 keV electron beam system we describe significantly extends the parameter range of electron beams used to pump CO$_2$ lasers. The modulator-electron beam system, as outlined in Table 1.1, allows both the electron beam energy and the current density to be varied independently [7]. The electron beam current density can be varied from 5-20 mA/cm$^2$ and the electron beam energy varied from 150-200 keV. This is unique in that it allows the electron beam energy to be tuned to different CO$_2$ laser gas mixtures being investigated. Concurrently, as shown in Table 1.1, the modulator-electron beam system was designed to operate from a single-shot mode up to 10 pps, also allowing gas heating effects to be explored over the 250 cm cathode length, while maintaining a ± 10% electron beam spatial profile. Moreover, the modulator-electron beam system was designed to investigate the unexplored regime of long (45 μs) laser pulse-widths, with a ± 10% variation in electron beam energy during the pulse [7].

The subsystems which comprise the modulator-electron beam system were theoretically and experimentally investigated during the course of the program. The major subsystems, namely the modulator system used for acceleration of the electron beam, the grid-pulser used to control the electron-beam current and pulsewidth, the hot-cathode electron beam source, and the final vacuum
### Table 1.1
**Summary of Electron Beam and Modulator Parameters**

#### 1.1A Electron Beam System Specifications

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cathode Area</td>
<td>2500 cm²</td>
</tr>
<tr>
<td>E-Beam Length</td>
<td>250 cm</td>
</tr>
<tr>
<td>E-Beam Width</td>
<td>10 cm</td>
</tr>
<tr>
<td>Post Foil Current Density</td>
<td>5-20 mA/cm²</td>
</tr>
<tr>
<td>Electron Beam Uniformity</td>
<td>± 10%</td>
</tr>
<tr>
<td>Post Foil Average Electron Energy</td>
<td>150-200 keV</td>
</tr>
<tr>
<td>Beam Pulsewidth (± 10%)</td>
<td>45 µs</td>
</tr>
<tr>
<td>Maximum E-Beam Rise and Fall-Time</td>
<td>4.5 µs</td>
</tr>
<tr>
<td>Pulse Repetition Rate</td>
<td>10⁸ pps continuous</td>
</tr>
<tr>
<td>System Lifetime</td>
<td>10 shots</td>
</tr>
</tbody>
</table>

#### 1.1B Modulator System Specifications

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Load Voltage (± 10%)</td>
<td>150 kV - 220 kV (250 kV peak)</td>
</tr>
<tr>
<td>Modulator Current</td>
<td>20-160 A</td>
</tr>
<tr>
<td>Peak Power</td>
<td>37.5 MW</td>
</tr>
<tr>
<td>Energy Stored</td>
<td>1800 J</td>
</tr>
<tr>
<td>Pulse-width</td>
<td>45 µs</td>
</tr>
<tr>
<td>Electron Beam Load Impedance</td>
<td>1.7 - 11 kΩ</td>
</tr>
<tr>
<td>Thyatron Voltage</td>
<td>35 kV</td>
</tr>
<tr>
<td>Thyatron Peak Current</td>
<td>3000 A</td>
</tr>
<tr>
<td>Thyatron di/dt</td>
<td>4.3 x 10⁸ A/sec</td>
</tr>
<tr>
<td>Repetition Rate</td>
<td>single shot to 10 pps</td>
</tr>
<tr>
<td>Lifetime</td>
<td>1 x 10⁶ pulses</td>
</tr>
</tbody>
</table>
insulator physics-engineering parameters found during the integration phase of the program are reported in the following thesis. Additionally, in-situ activation and carburization procedures developed are reported along with the spatial characterization measurements of the electron beam.

1.2 Introduction to CO₂ Laser Discharges

The modulator-electron beam system described in this thesis was specifically tailored to meet the requirements of an electron beam sustained atmospheric pressure CO₂ laser. Specific laser gas calculations are not within the purview of this thesis due to the proprietary nature of the laser design and the laser gas mixture. However, as an introduction to high power, atmospheric pressure, electron beam sustained, CO₂ lasers, the discharge parameters governing the laser design are discussed.

Lasers of the CO₂ type fall into two categories. The first category is the self-sustained CO₂ laser whereby the gas discharge is maintained by the electric field across the electrodes [8]. These lasers have a discharge E/N ratio of roughly 4-6 x 10⁻¹⁶ V·cm⁻² [9]. Based on Lowke’s efficiency calculations this limits the total power efficiency of the discharge to 40% based on a CO₂-N₂-He mixture of 1:7:30 [10]. As demonstrated by Verdeyen, when the electrical efficiency and quantum efficiency is included in the calculation a "wall plug" efficiency of less than 22% is possible [9] (however an
efficiency of 32.2% is possible as will be discussed below). Moreover, self-sustained discharges of long pulse-widths tend to constrict, thus forming arcs when operated at high powers and excessively high current densities.

To increase the overall system efficiency, electron-beam sustained CO₂ lasers have been developed. An electron beam with an energy of 100-200 keV is used to ionize the laser gas. A separate discharge is used to "pump" the laser gas [8]. The discharge is non self-sustaining without the electron beam. With the electron beam as an ionization source the laser gas can be utilized at E/N ratios on the order of 1-2 x 10⁻¹⁶ V-cm⁻². This results in an increase in the overall efficiency of the laser systems. As the E/N ratio is reduced from 4-6 x 10⁻¹⁶ V-cm⁻² to 1-2 x 10⁻¹⁶ V-cm⁻² the total power coupled into the gas increases from 40% to 85% [10]. For an electron-beam sustained CO₂ laser discharge Verdeyen has shown that an overall "wall plug" system efficiency of 32.2% is possible [9]. This is a rather important result if 100-200 kW of average power is being coupled into the laser discharge.

At atmospheric pressure in an electron beam sustained CO₂ laser, approximately 0.2 J/cm³ specific energy is imparted to the laser gas [11]. Typical electron densities in these discharges are on the order of 10¹¹-10¹³ electrons per cm³ [12]. At these electron densities the recombination rate for electrons is ~10⁻⁷ cm³/sec [12]. Thus the rate of ionization required to
balance losses is $10^{15}$-$10^{19}$ ionizations/cm$^3$/sec. From the ionization rate, the ionization power of the electron beam can be computed. Daugherty has shown that the external ionization source, or in this case the electron beam density, is one to four orders below that of the primary laser discharge [12].

Although the external electron beam source is one to four orders of magnitude below that of the primary laser discharge, to first order, the electron beam uniformity determines the spatial uniformity of the laser discharge. The spatial distribution, however, is also a function of the external electric field, the primary electron dose distribution, and the gas characteristics. Smith has used Monte Carlo calculations to calculate the electric field profile resulting from the injection of an electron beam into a gas discharge volume [13, 14]. The calculations are complicated and the electric field calculated depends upon the gas mixture employed.

An electron beam source which could accommodate a wide variety of different discharge conditions was built and tested. As shown in Table 1.1 the modulator-electron beam system was capable of providing 150 keV-220 keV electrons over current densities of 5-20 mA/cm$^2$. This parameter range is applicable to the $1-2 \times 10^{-16}$ V-cm$^2$ conditions prevalent in electron beam sustained lasers [10].

1.3 Overview of Thesis

To meet the 5-20 mA/cm$^2$ current density required for
the 2500 cm² electron beam area, both cold cathode and thermionic cathode technologies were reviewed. Chapter 2 discusses the relevant parameters effecting cold cathode and thermionic cathode design in these large area, low current density cathodes. Cold cathode space-charge limited emission is discussed along with design criteria for thermionic cathodes. Based on the design trade-off for the cathode, an overview of the modulator-electron beam system is presented. The modulator conceptual design along with the conceptual design of the 45 µs thermionic cathode is shown. Finally, the electron beam characteristics required for ionization of the laser gas are discussed. The modulator voltage characteristics as a function of these electron beam parameters are derived.

Chapter 3 presents a detailed design study of the thyatron-switched modulator used to accelerate the electron beam to 150 keV-220 keV energies. The representative types of PFN’s available to the modulator engineer are presented along with their applicability to transformer coupled modulators. Based on the final selection of the modulator geometry, a circuit model for the modulator-electron beam circuit is derived. Circuit simulations used to optimize the modulator characteristics into a constant current (thermionic) load are shown. The modulator design and layout derived from the PFN trade-off study and the circuit simulations is reviewed.

In order to control the electron beam turn-on, turn-off, and flatness characteristics, a grid controlled
thermionic cathode was selected. Chapter 4 describes the constant perveance load into which the grid pulser must operate. A model for the time-varying Miller capacitance is derived to simplify the circuit simulations. A circuit model for the grid pulser and grid-cathode requirement is presented. As will be noted in Section 4.4, the purpose of this model is to guide the practical design features of the grid-pulser design. Relevant circuit calculations used to optimize the grid-pulser interaction are described, along with the final grid pulser layout. Initial tests of the grid pulser into a resistive load are compared with the theoretical circuit simulations and used to confirm the accuracy of the circuit model.

Test results of the integrated modulator-electron beam system are described in Chapter 5. Test results of the modulator characteristics into a resistive load and the electron beam load are compared with the model simulations. Final integration tests on the modulator-electron beam system are described in detail. Cold cathode conditioning procedures developed during the "check-out" of the system are described. Relevant vacuum breakdown physics for vacuum insulators and large area vacuum gaps are shown, and compared with experimental results found during the check-out of the electron-beam gun. Finally, electron beam uniformity measurements using Faraday cup diagnostics to verify the required $\pm 10\%$ electron beam uniformity are presented.
REFERENCES

CHAPTER 1


CHAPTER 2

DESIGN CONSIDERATIONS FOR THE GENERATION OF E-BEAMS IN THE 1-100 MICROSECOND RANGES

2.1 Introduction and Requirements

Traditionally, modulators operate at voltages within the range from several kilovolts up to the megavolt range. In the 20-100 kV range, they are usually switched either by a high current gas spark gap or a thyratron. New developments in ceramic thyratron technology have dramatically affected the selection of modulator switches. Until recently, high average current thyratrons which could handle large voltage reversals were not commercially available. At low-to-modest currents standard thyratrons were used with protection circuits designed to limit reverse currents and voltages under mismatched load conditions [1]. Reverse current protection often consists of snubber diodes or reverse thyratrons connected across the main forward conducting thyratron. This protection circuitry is primarily designed to prevent reverse current damage to the forward conducting thyratron. Ultimately, the cost of the modulator was increased significantly. Thus, spark gaps were usually more attractive in modulators designed for high reversal regimes or at high peak currents (~ 10 kA) and repetition rates of ≤ 100 pps.

Excimer lasers as well as electron beam accelerators operate under mismatched load conditions. Excimer laser
drivers are typically designed to accommodate a collapsing load impedance, whereby high reverse currents and voltages are generated as arcs develop in the laser gas. Similarly, both cold cathode and hot cathode electron beam diodes used in accelerators display many of the characteristics of excimer lasers. During the electron beam pulse, plasma generated on the cathode may result in closure of the diode, thus shorting out the anode-cathode gap. The vacuum diode region may also short if the diode arcs, or a vacuum bushing flashes. Grid controlled electron beam guns (thermionic or cold cathode) further complicate the modulator’s load characteristics. Because gridded electron beam guns require a positive grid pulse to turn on the beam, open circuit fault modes can result if the grid pulser fails to fire. Open circuit conditions can also occur as the diode emitting material ages. If thoriated tungsten filaments are used, as in the the case of many hot cathodes, the thorium can be effectively depleted from the surface by contaminants or ion bombardment. As a result, electron emission may drop rapidly. Likewise, open circuit fault modes can be present when cold cathodes are employed. Often, felt is used for the emitter. As the felt ages, emission may degrade and thus change the impedance of the diode, or even result in an open circuit load condition. In both cases, conventional snubber diode technology cannot tolerate the high di/dt, dv/dt, reverse currents and voltages at even modest repates. New thyratrons, developed for use in excimer laser modulators, avoid many of the limitations of traditional
thyratrons. These new, hollow anode, ceramic thyratrons are able to handle 50-60% current reversal without a decrease in lifetime, thus eliminating the requirement for reverse protection devices in many cases. This advantage is quite important for modulators intended for applications other than those of laser drivers.

The factors which directly influence the design of a given modulator are the type of cathode (cold or thermionic), the output voltage and current, and the system lifetime reliability specification. Marx charged Blumlein or coaxial high voltage pulse power systems have traditionally been designed to deliver only about $10^4$ shots or so, whereas, commercial and scientific applications now require lifetimes of $10^6 - 10^9$ shots without failure. A simple calculation emphasizes the need for new design approaches to meet the requirements of new applications of pulse power technology. If one of the systems designed to deliver $10^4$ shots is operated at 10 pps or 36,000 shots per hour, the system will require maintenance three times an hour, a rather impractical situation. Developmental laboratory experiments were easily accommodated by such Marx bank charged coaxial systems designed over the past 20 years. Unfortunately these systems are not adequate for present system lifetime requirements.

To achieve pulsed power systems which operate at high repetition rates with long lifetime and at high energies, new design approaches and philosophies need to be developed. The following pages review some of the design
philosophies now available to the pulse power engineer and present new technologies which have been developed specifically for multi-kilojoule modulators. One such system is then described in the following thesis. As part of a CO$_2$ laser development program a 45 µs, 150 keV-220 keV electron beam pulse was required to sustain the laser discharge. To facilitate this requirement an integrated pulse power modulator and electron beam system was developed. The load characteristics, which influenced the modulator design, as well as the selection of the major modulator components are reviewed. An integrated pulse power design based upon a system trade-off study is presented. Characteristics of both cold cathodes and hot cathodes which strongly influenced the final cathode design are reviewed. Behavior of the cathode material under the required 45 µs pulse-width is described.

2.1.1 Cold and Thermionic Cathode Processes

Cold cathode diodes are perhaps the most difficult to characterize for long pulses. The diode current density of an ideal vacuum diode is described by the Child-Langmuir equation,

\[ J = 2.33 \times 10^{-6} \frac{V_{AK}^{3/2}}{d^2} \text{A/cm}^2 \]  

(2.1)

where $V_{AK}$ (volts) is the accelerating voltage applied to the anode-cathode gap, $d$ (cm) the anode-cathode
separation, and $J$ the current density in A/cm$^2$. Tripoli [2] empirically derived a modified version of the Child-Langmuir equation for cold-cathode diodes under microsecond pulse conditions. He noted that the anode-cathode gap was essentially time dependent due to plasma which originates in the gap and closes it at velocity $u$(cm/µs). The modified Child-Langmuir equation which results is

$$J = 2.33 \times 10^{-6} \frac{V_{AK}^{3/2}}{(d-ut)^2} \text{ A/cm}^2.$$ (2.2)

The equation is relevant for voltages below 250 kV. Above 250-500 kV, relativistic correction factors are required to account for particle interaction.

Relativistic electron beam geometries used for flash x-ray radiography, gamma radiation simulation, and reflex diode experiments utilize cold cathodes [3]. The diode's cathode may be a velvet material or stainless steel alloy. In any event, the pulses of megavolt simulation machines are several hundreds of nanoseconds in length. The plasma generated at the cathode surface does not have time to short the anode-cathode gap during a 100 ns pulse. For longer pulses, the reported 1-2 cm/µs velocity of the cathode plasma limits cold cathode operating regimes to pulse-widths of about 2 µs [4].

Cold cathode field emission of electrons was theoretically calculated by Fowler and Nordheim [5]. Electrons are trapped by the potential barrier existing in metallic conductors. Electron emission from the surface
of the metal occurs when the electrons are subjected to
electrostatic fields of the order of $10^6$-$10^7$ V/cm, or are
thermally excited by heating the cathode to high
temperatures. A tungsten cathode at about $2500^\circ$ K emits
$0.5$-$1.0$ A/cm$^2$. Cold cathodes utilize electrostatic field
emission. Fowler and Nordheim showed that the current
density $J$ is strongly dependent on the surface electric
field $E_s$, the work function of the metal $\varphi_w$, the band gap
energy $\varphi_B$ and the Fermi energy $\varphi_m$ as shown in Eq. 2.3
below [6],

$$J = CE_s^2 e^{-D/E_s}$$  \hspace{1cm} (2.3)

where we have,

$$C = 6.2 \times 10^{-6}/\varphi_B \quad (\varphi_m/\varphi_w)^{1/2}$$

and

$$D = 6.8 \times 10^9 \quad \varphi_w^{3/2}$$

The Fowler-Nordheim equation has been verified experi-
mentally and has been shown to describe accurately the
emission current up to about $10^6$-$10^8$ A/cm$^2$.

Above $10^8$ A/cm$^2$, the emitting surface site vaporizes
(explodes) at surface fields corresponding to $10^7$-$10^8$
V/cm. The emission current described by the Fowler-
Nordheim equation departs significantly from experimental
data above the explosive emission regime. A time-
dependent emission from the cathode has been observed
which is independent of the applied electric field and the
anode-cathode separation. Mesyats [7,8,9] has postulated
that micro-enhancements at the surface of the cold cathode may be responsible for both the observed plasma ejected from cold cathodes, and departure of the theoretical emission current from experiments [10]. Macroscopically, the surface of a metallic or graphite emitter may appear smooth. Microscopically, large numbers of surface enhancements are present which result in many field emission sites at the surface of a cathode under vacuum conditions. Although theoretical calculations assume that, macroscopically, the entire surface of a cathode emits current uniformly, in reality, the tips of "whiskers" may emit current densities well above the calculated macroscopic current density. Rapid joule heating results in evaporation of material from the tip of the emitter. Subsequent ionization of the evaporated metallic material results in an expanding plasma. Observed 1-2 cm/μs expansion velocities explain the time dependent behavior of long microsecond pulsed diodes [11]. The resulting impedance collapse of the cold cathode diode affects the design of the high voltage modulator used to generate the accelerating voltage. At any discrete point in the pulse, the space-charge limited current of the anode cathode gap is described by Equation 2.2. At a particular instant in time, the diode can be modelled as a constant current source. Over the length of the pulse, the current may increase by a factor of 2-3 times over the initial value due to the changing anode-cathode separation. As the plasma flare expands toward the anode, the effective separation d is significantly reduced,
resulting in a decreasing diode impedance. Eventually, the diode is shorted by the plasma closure thus causing voltage and current reversal across the pulse power modulator components.

Cold cathode materials based on Mesyats' postulate have been developed which consist of many enhanced emission sites. It is well known that black felt, a synthetic based material, has a surface consisting of many carbon loaded fibers. In an electric field the tips of the fibers provide the field enhancement necessary to cause field emission. Experiments conducted with 1 µs pulse lengths at 150 kV, have shown that the lifetime of the felt emitter depends on the average current density and the number of emission sites available [12]. Lifetimes of $10^7-10^8$ shots have been demonstrated at average current densities of 1 A/cm$^2$.

When the $10^8$ cathode/system lifetime is coupled with the 45 µs pulse-width required for the electron beam CO$_2$ laser, cold cathode materials do not meet the system requirements. The explosive-emission characteristics of the cold cathode materials limit the pulse-width to several microseconds, and require cathode current densities below approximately 1 A/cm$^2$ if the lifetime of $10^8$ is to be achieved. However, if substantially shorter lifetimes are permitted, velvet cathodes can operate at $10^3-10^4$ A/cm$^2$.

In contrast, for a diode with a thermionic cathode, impedance characteristics are invariant during the applied pulse if the cathode is space-charge limited or grid
controlled. Based on theoretical calculations, Dushman [13] derived an equation to describe the available emission current density as a function of the surface work function ($\varphi_w$) and the surface temperature ($T$). The Dushman equation is

$$J_s = A T^2 e^{-\varphi_w / kT}, \ \text{A/cm}^2,$$

(2.4)

where we have

$$A = \frac{2\pi q m_e k^2}{\hbar^2}, \ \text{A/cm}^2 \cdot \text{deg}$$

The Dushman or Richardson-Dushman equation has been experimentally verified for metals including tungsten and thoriated tungsten.

Dushman showed that the emission current density of a thermionic cathode is exponentially proportional to the work function, $\varphi_w$, of the cathode. The prevalent 4-5 eV work function of metals requires excessive operating temperatures to realize 1 A/cm$^2$ current densities. Because of its high work function (4-5 eV), tungsten requires temperatures of about 2900 K to achieve modest (0.5 A - 1 A/cm$^2$) current densities [14]. Such high temperatures necessitate complex cathode support structures designed to insulate and survive at the elevated temperatures. Various surface coatings, along with selectively introduced impurities, have been developed which lower the work function of metals and thereby reduce the cathode operating temperatures to the
range $1000^\circ$ K-2000$^\circ$ K. Among these cathodes are BaO and SrO coated filaments and thoriated tungsten filaments.

To achieve cathode operating temperatures in the range $1000^\circ$ - $1500^\circ$ K, the work function $\varphi_w$ of the emitter must be reduced to 1-2 eV. Two well known coatings, BaO and SrO, lower the work function of the surface so that the operating temperatures are in the range $1000^\circ-1500^\circ$ K. Emission current densities at these reduced temperatures are typically $1-10$ A/cm$^2$ [15]. Because the lower cathode temperatures result in lower thermal cycling stresses in the surrounding support structure as well as significantly less heater power, oxide coatings have been widely employed in vacuum tube and gas tube environments [16]. One such application includes thyratrons [17]. The use of the oxide coated cathodes in thyratrons decreases the heater power as well as the thermal loading of the gas. Moreover, the ceramic or glass housing temperature is lowered, thus reducing the thermal stresses on the metal-to-ceramic seals.

Inherently, difficulty in achieving uniform emission from coatings may occur if the coating has not been uniformly applied. Additionally, during the lifetime of a cathode, flaking of a coating may occur if the coating is not uniform. Dispenser cathodes, however, avoid this problem. Dispenser cathodes consist of a metallic matrix into which a low work function material has been diffused. As the cathode material "ages", the low work function material diffuses to the surface of the cathode. One such dispenser cathode material uses BaO impregnated into a
tungsten matrix [18]. The impregnation of a metallic matrix avoids several problems such as flaking associated with coated cathodes but, like coated cathodes, dispenser cathodes are subject to poisoning by trace contaminants. Thus, use of oxide coated as well as dispenser cathodes is primarily limited to small area cathodes of about a hundred square centimeters, under vacuum conditions which are closely controlled.

Since both oxide coated and dispenser cathodes are easily poisoned by trace contaminants, their operation can be quite expensive when incorporated into large area electron guns. Trace background contaminants of water vapor can cause rapid reduction of emission current and permanent "poisoning" of the oxide coating. Clean room assembly and special vacuum cleaning techniques are required to assure long lifetime operation of both dispenser and oxide based cathodes.

2.1.2 Thoriated Tungsten Cathodes

Thoriated tungsten cathodes offer an intermediate alternative to both oxide and pure metal cathodes. Thoriated tungsten (Th-W) alloys have proved to be the most versatile and durable under adverse operating conditions. Electron guns with areas over 5000 cm² have been developed for electron beam pumping of laser gases. Large area electron guns (~ 2000-3000 cm²) have been demonstrated which use 10-20 mil thoriated tungsten filaments [19]. When operated at 1800°-2000° K, reported
lifetimes of thoriated tungsten filaments approach 2000 hours even after repeated cycling to atmosphere. In addition, residual gas analysis has shown that trace amounts of CO, CO$_2$, and H$_2$O do not significantly degrade the emission characteristics of the filaments.

2.1.3 Activation of Thoriated Tungsten Cathode Filaments

Early investigators found that the addition of 1-2% ThO$_2$ to tungsten significantly lowered the work function to 2.63 eV, after activation of the filament [20]. In contrast to cold cathodes, thermionic cathodes require an activation procedure to drive the emitter material to the surface of the cathode. Thorium and tungsten have work functions of 3.47 eV and 4.53 eV, respectively. To activate Th-W filaments, the filaments are heated to approximately 2500° K for one minute. During the activation procedure, the ThO$_2$ is dissociated to give Th and O$_2$. Residual gas analysis during the activation procedure confirms the evolution of O$_2$ during the process. After activation, the filament temperature is rapidly reduced to about 2100° K. The filaments are then held at 2100° K for 20-30 minutes to allow the thorium to diffuse to the surface of the tungsten filaments. A Th-W monatomic film is formed on the surface thus decreasing the work function to 2.63 eV and increasing the emission characteristics of the thermionic cathode at lower operating temperatures. The original activation process utilizes only a small fraction of the ThO$_2$. Because only
a thin monatomic film is required to lower the work function of the surface, the activation process uses only several parts per thousand of the ThO₂ contained in the tungsten matrix. Thus, thoriated tungsten filaments poisoned by ion sputtering or contamination are renewable by subsequent reactivation.

If long-lifetime (~2000 hours) of the Th-W filaments is required, additional surface treatment is desirable. Prior to activation, the surface of the filament requires carburization [21]. Experiments backed by detailed microscopic examination of the Th-W surface have demonstrated that a tungsten carbide film is necessary to prevent deactivation by background contamination. Carburization of the Th-W filament is done by introducing a hydrocarbon gas such as methane, acetylene or propane while heating the filaments. At the elevated filament temperature the hydrocarbon reacts at the surface of the filament and forms a thin layer of tungsten carbide.

2.2 Overview of the Pulse Power and Hot Cathode Electron Beam System

An electron beam system was investigated to evaluate the applicability of multi-microsecond pulsed cathodes. Following the cathode study, a grid controlled, thermionic electron beam gun, with a pulsed acceleration voltage was selected to meet the 45 μs pulse-width required to ionize and sustain a CO₂ laser discharge (Table 1.1). Specific goals of the study included the development of a compact
grid pulser for control of the electron beam and a high reliability 150 kV-250 kV pulser capable of a $10^8$ shot lifetime as shown in Table 1.1. The electron beam system consisted of five distinct sections: (1) thermionic grid controlled electron beam gun, (2) 250 kV thyratron switched modulator for acceleration of the electron beam, (3) compact grid pulser for control of the electron beam, (4) 1 kHz heater supply and circuitry to provide power to the cathode filaments, and (5) an interactive control system for control and shutdown of the system. Physics/engineering parameters and trade-offs which affect the operation, design, and lifetime of the high voltage system are shown.

To obtain a uniform electron beam along the 250 cm cathode length, discrete thoriated tungsten filaments, control grids, and extraction grids were used (Figure 2.1). The thoriated tungsten filaments operate at a temperature of 1800-1990$^\circ$K. The 132 filaments require 6000-8000 W of heater power. The electron emission from each 15 mil thoriated tungsten filament is controlled by applying a positive pulse to the hemicylindrical extraction grid which encloses each filament, Figure 2.1. A second planar grid, at the same potential as the extraction grid, is positioned to be uniformly illuminated by the electrons emerging from the extraction grid. The uniform beam of electrons which emerge from the field-free control-grid/extraction-grid region is then accelerated by a high voltage pulse which is applied between the cathode and anode. The accelerated electron beam traverses the
Figure 2.1 Photograph and layout of grid controlled electron beam gun with the control-extraction grid.
anode-cathode gap, and penetrates the geometrically transparent Hibachi-foil structure separating the vacuum chamber from the laser cavity. The electron beam is then used to preionize the CO$_2$ laser cavity. The many discrete filament arrays which comprise the overall cathode assembly are enclosed in a polished, smooth, cowling (Figure 2.2). The electro-polished stainless steel cowling minimizes surface enhancements which could lead to uncontrolled cold cathode emission at the applied 150 kV-250 kV acceleration potentials.

The high voltage cathode is supported by two high vacuum, ceramic bushings, Figure 2.3. The accelerating potential is supplied through one bushing along with heater power and a grid pulse. The bushing and cathode are cooled during operation by oil flowing along their outside surfaces. The respective high voltage pulse, negative grid bias, grid pulse, and filament power are routed through three parallel cables, which originate at the modulator.

The pulse power system consists of a 150 kV-250 kV modulator, a grid pulser floating at the accelerating potential, Figure 2.4, and a 8 kW, 1 kHz filament circuit to deliver the filament power at the cathode potential. With the exception of the commercial high voltage power supplies and a remotely located control system, each pulse power system is housed in an oil-filled tank located 6 m from the electron beam gun (Figure 2.5). Three 6 m, 300 kV, solid dielectric coaxial cables feed their respective signals to the gun through an oil-filled housing located
Figure 2.2 Complete cathode assembly to which the high voltage 150 kV - 250 kV acceleration pulse is applied.

Figure 2.3 Layout of high voltage filament transformer bushing and cathode in relation to its vacuum chamber and Hibachi-electron beam window.
Figure 2.4 Simplified system layout showing the interrelationship of the modulator grid pulser and electron beam gun.
Figure 2.5  Photograph of high voltage modulator mounted above its oil tank.
on the back of the cathode vacuum chamber.

The high voltage modulator consists of a charging system which resonantly charges a lumped-element pulse forming network (PFN) to approximately 30-32 kV (Figure 2.6). The modulator output pulse is initiated by triggering the 35 kV hollow anode, ceramic thyratron. The thyratron discharges the PFN through the primary of a 1:19, 250 kV, iron-core pulse transformer to produce the required cathode voltage (Fig. 2.7).

Selection of the 1:19 transformer ratio was based upon the peak operating voltage of the thyratron, the required 220-230 keV electron beam energy, and the PFN-load impedance mismatches present. The thyratron operating voltage of 30-32 kV was selected on the basis of its 35 kV rating. Stray capacitance across the secondary of the transformer was also included in these calculations. From these factors, a 1:19 transformer ratio was calculated. It should also be mentioned that a fixed impedance PFN could not meet the E-beam acceleration voltage droop or ripple requirements when the stray capacitance and the transformer droop were considered together. An impedance tapered Type E PFN was designed to compensate for the circuit droop characteristics.

The impedance tapered PFN is impedance tailored along its length to compensate for the stray capacitance across the secondary of the transformer core. The details of this calculation and design are given in Section 3.2 (Impedance Matching of Pulse Forming Networks into a Constant Current Load). Independent electron beam energy
Figure 2.6 Block diagram of modulator and support components.
Figure 2.7  Schematic of pulse power system (modulator) including resonantly charged power supply, pulse forming network, step-up transformer and high voltage coaxial cables linking the modulator to the electron beam gun.
and current were required by the laser manufacturer in order to investigate the laser gas parameter range. Different laser gas mixtures require different electron dose profiles, and thus different electron beam energies. Concurrently, independent adjustment of the electron beam current is required to match the gas preionization rate to the discharge characteristics. Thus, the modulator must operate satisfactorily into a load which may vary over a wide range. To accommodate these independent voltage and current requirements over the wide parameter range of output currents, output load matching resistors were included as part of the modulator.

Details of the modulator design arising from these requirements are given in Section 3.3 (Circuit Design of a 250 kV, 45 μs Modulator for the Constant Current Grid-Controlled Cathode Load). Initial calculations showed a 6:1 operating impedance range, for 50 A to 125 A currents, at 150 kV-250 kV voltages. The 6:1 impedance range was well outside the normal 2:1 impedance variation imposed by reverse voltage ratings of the components. Also since the modulator has an output impedance comparable to that of the constant current beam, the output voltage of the modulator is dependent on the beam current and the matching resistor selected. This effect is discussed in Section 3.3 and 3.3.1. Thus, the PFN charge voltage and resistor selection interact to determine the amplitude of the output voltage pulse. To determine accurately the modulator voltage and current versus the applied grid voltage, a circuit model of the high voltage pulser and
electron beam load was constructed. The circuit model developed allows the anode-cathode current to be predicted for a given grid voltage and allows the response of the anode-cathode pulse to be ascertained when varying the relative timing of the modulator output and the grid pulse with respect to the turn-on of the electron beam. Section 3.3.1 (Circuit Analysis of a 52 μs, 5.5 Ω, Tapered Impedance Type E PFN), discusses details of the model derived and the synergistic circuit effects which interact. The computer model constructed also allows an accurate determination of the energy (reverse voltage), reflected from impedance mismatches. Based upon the model results, protection circuits were designed to limit reversal on components during normal operation and under fault mode conditions. In order to prevent over-voltage of the high voltage bushings, and the electron gun cathode-anode gap in the event that the modulator was operated outside of the normal setting, a circuit model of the system was needed. A transmission line model was devised to allow selection of the output resistors before operation of the system. Both the transmission line code and the circuit model based on SPICE, a sophisticated circuit analysis code, are discussed in Chapter 3, under Section 3.3 and 3.3.1.

A compact grid pulser (800 V, 45 μsec, 10 pps) was developed to provide the extraction/control grid pulse for the electron gun. In order to reference the grid pulse to the cathode potential, the grid pulser is floated at the cathode acceleration voltage (Fig. 2.4). Because the
grid pulser is at the output voltage of the modulator, it is oil-immersed and located in the modulator tank (Fig. 2.5).

The grid pulser utilized a PFN switched by a high power solid state silicon controlled rectifier. The PFN is discharged into a 2:1 step-up transformer to provide the 400 V-800 V pulse for turn-on of the electron beam. The risetime and pulse shape of the grid pulse were tailored so that the interactions of the modulator pulse and turn-on characteristic of the electron beam did not degrade the applied anode-cathode voltage pulse. To understand this grid pulser/electron beam interaction, a model was developed to determine the time-varying impedance of the grid during turn-on of the beam. The model describing the behavior of the grid impedance is presented along with the main circuit modeling techniques in Section 4.2 and 4.3. As part of the grid pulser design, an SCR crow-bar switch was included to control the length of the grid pulse. Layout and implementation/testing of the grid pulser is described in Section 4.4. The ability to control the length of the current pulse is important because the electron beam interacts significantly with the foil below energies of 125 keV; the electron beam must be shut off before the applied cathode potential falls below 125 kV. To reduce the interaction of the electron beam with the foil, the beam is turned off by crow-barring the grid pulse. This effectively shuts off the electron beam. Moreover, the grid of the cathode is biased negatively so that the displacement current
which flows in the grid-to-cathode circuit is negligible during the risetime of the anode-cathode pulse. If displacement currents were to cause the grid to be positively biased, the low energy electrons emitted during the risetime of the anode-cathode pulse would be absorbed by the foil window. Eventually the foil would overheat under repetitive operation, and finally develop pinholes or rupture.

A filament circuit was adopted which can deliver a maximum of 8 kW to the cathode’s multiple filament arrays and the grid pulser, both of which are at cathode potential. The voltage across each filament must be limited in order to prevent local filament-to-grid potentials, which may be on the order of the grid-to-filament control pulse voltage. Thus, filament power must be delivered at high current (500 A) and low voltage (15 V), which is not compatible with available high voltage coaxial cables. Because the copper conductor of each high voltage coaxial cable is limited in size, excessive losses would result if the filament power was delivered to the filaments at low voltage. The filament circuit chosen allows the power to be transported to the electron beam guns at high voltage (400-480 V) and low current (15 A). The high voltage is then reduced to values consistent with the cathode requirements. In order to reduce the size and weight of the filament step-down transformer, which must be mechanically and electrically isolated from ground, a 1 kHz power system is used rather than the normal 50-60 Hz line frequency. A circuit was developed which allows the
filament power to be transformed to cathode potential by means of the bifilar secondary winding on the high voltage modulator pulse transformer. Circuits to isolate the 1 kHz power supply from transients were also devised during the study.

2.3 Modulator Current and Voltage Requirements

The modulator's output voltage and current are determined by the transparency of the Hibachi-foil structure and the width of the laser chamber. The electron beam which emerges from the field-free control/extraction grid region is accelerated by the anode-cathode voltage. To penetrate the foil window separating the cathode vacuum chamber and laser cavity, electron energies above about 125 keV are required. Furthermore, the post-foil electrons are also required to fully ionize the laser gas. Typical CO₂ laser chambers may range from 2 cm in width to well over 20 cm. Thus, the required penetration depth or required range of the electrons may be over 30 cm (equivalent range in air) if the foil window thickness is included.

The penetration depth of electrons versus energy are shown in Fig. 2.8, along with their respective tabulated ranges for various foil window thicknesses in Table 2.1. Figure 2.8 illustrates the energy deposition versus penetration depth for electrons of various energies [22]. The extrapolated range ($R_e$) of electrons is given in mg/cm², which is the ratio of the density of the medium to
Figure 2.8 Energy deposition distribution versus the penetration depth of electrons for various electron energies [Ref. 22].
Table 2.1
Tabulated areal thickness, electron ranges and penetration depth for Al and Ti foil windows

<table>
<thead>
<tr>
<th>Foil Thickness</th>
<th>Areal Thickness</th>
<th>Electron Equivalent Range</th>
<th>Electron Energy</th>
<th>Electron Penetration depth (cm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ti foil (4.54 g/cm$^3$)</td>
<td>17.7 micron (0.7 mil)</td>
<td>8 mg/cm$^2$</td>
<td>6.2 cm</td>
<td>150 keV</td>
</tr>
<tr>
<td></td>
<td>25 micron (1 mil)</td>
<td>11.5 mg/cm$^2$</td>
<td>8.9 cm</td>
<td>200 keV</td>
</tr>
<tr>
<td>Al foil (2.7 g/cm$^3$)</td>
<td>17.7 micron (0.7 mil)</td>
<td>4.7 mg/cm$^2$</td>
<td>3.64 cm</td>
<td>150 keV</td>
</tr>
<tr>
<td></td>
<td>25 micron (1 mil)</td>
<td>6.7 mg/cm$^2$</td>
<td>5.19 cm</td>
<td>200 keV</td>
</tr>
</tbody>
</table>

A. Tabulated areal thickness for 17.7 and 25 micron Al and Ti foils with their equivalent range compared with air.

<table>
<thead>
<tr>
<th>Foil Thickness</th>
<th>Electron Energy</th>
<th>Electron Penetration depth (cm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ti</td>
<td>150 keV</td>
<td>6.6 $\times 10^{-3}$</td>
</tr>
<tr>
<td>Ti</td>
<td>200 keV</td>
<td>1.1 $\times 10^{-2}$</td>
</tr>
<tr>
<td>Al</td>
<td>150 keV</td>
<td>1.1 $\times 10^{-2}$</td>
</tr>
<tr>
<td>Al</td>
<td>200 keV</td>
<td>1.85 x 10</td>
</tr>
<tr>
<td>Air</td>
<td>150 keV</td>
<td>23.25</td>
</tr>
<tr>
<td>Air</td>
<td>200 keV</td>
<td>38.7</td>
</tr>
</tbody>
</table>

B. Calculated Al and Ti foil thickness required to stop 150 keV and 200 keV electrons. Penetration depth of electrons in air has been included as a reference.
the electron penetration distance (Fig. 2.8). That is, the density of a substance or gas, for example a foil window, multiplied by the thickness of the medium gives the equivalent mass thickness in mg/cm$^2$. The curves shown are for no specific medium and must be modified by taking into account the material the electrons are traversing. Even a thin foil window represents a sizable penetration depth to an electron having an energy of 150 keV. As an example a 17.8 micron (0.7 mil) titanium foil has mass thickness of 8 mg/cm$^2$ (density 4.54 g/cm$^3$) which is equivalent to 6.26 cm of air with a mass thickness of 1.29 mg/cm$^3$. Thus, a minimum acceleration voltage of 125-150 kV is required to both penetrate the foil window and ionize a laser cavity of modest transverse dimension. The range for an electron of 150 keV is typically 20 cm in an electron-beam sustained CO$_2$ laser gas-mixture at STP.

The penetration or range calculations do not predict the amount of current which is transmitted through the foil for a given energy. However, if the electron beam is monoenergetic, the fraction of electrons transmitted through the foil versus foil thickness can be calculated. Because the incident energy of the electron beam along with the density of the medium determine the number of reflected or backscattered electrons, Monte Carlo computer codes can be used to calculate the number of electrons reflected (transmitted) for a given set of boundary conditions. For titanium foils, Fig. 2.9 shows the transmitted and reflected fractions of electrons versus foil thickness at 150 keV and 200 keV. For a 17.8
Figure 2.9 Transmission and absorption curves for 150 keV - 200 keV electrons.
micron (0.7 mil) thickness of titanium foil the transmission coefficient is 70% and 88% for respective 150 keV and 200 keV electrons. Unfortunately, the electron beam is not monoenergetic. Due to the stray capacitance across the output of the modulator and cathode, the acceleration voltage pulse is not flat. Furthermore, as the cathode voltage and current are changed along with the output matching resistor, ripple will appear on the pulse. As a baseline design, a ± 10% modulator voltage flatness over the 6:1 load impedance range at 150 kV-230 kV was chosen.

Similarly, post-foil currents of 12.5 A-50 A were required to ionize the laser gas. For 150 keV electron beam energies, a 17.8 micron titanium foil intercepts 30% of the beam. In addition, if the electron beam trajectory is not perpendicular to the Hibachi, some of the prefoil cathode current will be intercepted by the long side-walls of the Hibachi. Moreover, the geometric transparency or amount of the open area, versus total Hibachi area must be included in the calculations of current. Thus, the actual number of electrons cannot be fully estimated. Instead, the "average" voltage during the pulse is used to estimate the electron current transmitted through the foil. As a baseline design, a total transparency of 0.4 for the Hibachi and foil was calculated. This sets the prefoil cathode current range to 30-125 A for the required 5-20 mA/cm² electron-beam current density in the laser-gas.
REFERENCES
CHAPTER 2


CHAPTER 3
CIRCUIT MODELLING AND DESIGN
OF THE 250 KV MODULATOR

3.1 Introduction to Pulse Forming Networks
For the Generation of Electron-Beams

To accelerate the electron beam to the 150 keV - 220 keV energies required to sustain the CO₂ laser discharge (Table 1.1), a modulator based on a thyratron switched pulse forming network (lumped element transmission line) was selected.

The discrete or lumped approximation of transmission lines are known as Guillemin or pulse forming networks (PFN)[1]. Six fundamental voltage-fed networks were derived by Guillemin to approximate the rectangular pulse produced by a charged transmission line when it is discharged into a fixed load impedance (Fig. 3.1). Similar networks had been previously derived from Rayleigh’s principle which argued that physical realization of distributed parameters can be approximated by ordinary differential equations [1]. Unfortunately, the networks derived from Rayleigh’s principle inherently suffer from overshoot and oscillation. The overshoot and oscillation are the result of the finite number of sections which are used to approximate the rectangular pulse produced by a distributed transmission line when it is discharged into a fixed impedance load. Guillemin however, found that the overshoot or inherent discontinuity in both the pulse shape and mathematical
Figure 3.1 Equivalent forms for five-section Guillemin voltage-fed network. Multiply the values of the inductances by $Z_{NT}$ and the values of the capacitances by $\tau/Z_N$. The inductances are in henries and the capacitances in farads if pulse duration is expressed in seconds and network impedance $Z_N$ in ohms [1].
approximation could be eliminated if pulse waveforms with finite risetime were used to derive the lumped parameters for the pulse forming network. He then derived the corresponding circuit values for parabolic and trapezoidal (risetime) waveforms. Guillemin also found that the number of sections in the PFN varies with the required risetime, pulse flatness (overshoot), and fall-time. The five networks shown in Fig. 3.1 were derived and approximate a trapezoidal waveform with a risetime which is 8% of the pulse-width. Inherently the risetime, pulse flatness, and fall-time determine the number of sections needed in the pulse-forming-network.

The six fundamental pulse forming networks and their applicability to electron beam generators are reviewed below. The most commonly employed pulse forming networks are the Type A, Type C, and a subset of the Type D network known as a Type E PFN (Fig. 3.2). The Type A topology differs significantly from the topology of the more common Type C and Type E networks. The Type A PFN relies upon resonant filter sections to produce a rectangular pulse shape. A storage capacitor is charged and then discharged into the uncharged resonant tank circuits which comprise the Type A network. The pulse width is determined by the total series circuit inductance as shown by the five section, Type A PFN of Fig. 3.2. Subsequently the Type A has been used for electron beam drivers operating at energies above 100 keV [2]. A common Marx-bank generator or high voltage capacitor can be discharged into the uncharged filter or
Figure 3.2 Six section Type E PFN showing the mutual inductance (M) between inductors
resonant tank sections to produce a trapezoidal pulse shape. The Marx-bank series inductance can also be tailored to give the pulse width desired. The Type A has some disadvantages which should be mentioned. Because the charged primary capacitance is discharged into the uncharged filter sections, resonant circuit considerations require the filter section capacitors be rated at twice the original charge voltage. Moreover, the inductances in the filter section are usually low in value and difficult to tune to the desired inductance value derived by circuit simulation calculations.

The Type C is the easiest pulse forming network to design. Given a desired pulse waveform, a Fourier decomposition of the waveform is used to derive each individual branch network. The discharge current through each branch network is given by

$$i(t)_V = V_{ch} \sqrt{C_L/L_V} \sin \left( t/\sqrt{L_V C_V} \right),$$

(3.1)

where $V = 1, 3, 5, \ldots$ The branch number $V$ corresponds to the odd series of the Fourier series for a trapezoidal or parabolic waveform [1]. Each branch capacitance is determined from the individual sequential coefficient in the Fourier series. Because each individual branch network is unique for the Type C PFN, the branch capacitances are different. The first branch of the network stores the majority of the energy in the network. As such, the capacitor is large and may be physically unrealizable as a single unit. The cost of fabricating
the network is also significantly higher than for a Type E, for many different values of capacitors are required. The individual branch inductances may also be large and have large stray capacitances which complicate the design of the PFN. Thus for commercial applications, the Type C network has fundamental problems.

The Type D Guillemin network is the most interesting of the network topologies (Fig. 3.1)[1]. The equal capacitance values derived for the PFN network represent a significant advantage over the Type C PFN. The Type D network, however, has the inherent disadvantage of requiring negative inductance values in each discharge branch of the network. The theoretical derivation of the Type D network requires that a physically unrealizable negative inductance be placed in series with each section capacitance. A physically realizable Type E network was derived to replace the Type D network [1]. The Type E PFN, as shown in Fig. 3.2 consists of equal capacitances like the Type D; however, the negative inductances are not required. Rather, the mutual inductances associated with a continuously wound or coupled inductance approximate the negative inductance normally associated with a Type D network. The Type E network, when designed for a fixed impedance load, is usually designed to have equal inductance between adjoining capacitors. Both the output and opposing end branch inductances, however, are made 20-30% higher than the middle section inductances [1]. This results in an important advantage if transformer coupling of the PFN is required, as in an
electron beam system. If the leakage inductance of the transformer is included as part of the total inductance of the PFN output section, the pulse risetime and the pulse shape can be tuned easily by varying the inductance of each PFN section. However, in many designs the mutual inductance cannot be physically realized or fully utilized if a transformer is used in the modulator. In addition, the conventional Type E PFN has a constant impedance along its length, and does not lend itself to time varying loads, charging large stray capacitance or pulse transformer droop if a rectangular pulse is desired.

To accommodate time varying impedance loads "hybrid" Type E PFN's have been developed. The first such hybrid PFN was developed to drive constant pervenance loads such as klystrons [3]. Neal has documented the design of one such 10-section PFN used to drive klystron loads at the SLAC accelerator facility [3]. Constant capacitance was used in each PFN section. The impedance of the PFN was varied using adjustable inductors mounted orthogonally to avoid mutual coupling. This allowed the PFN to be easily tuned into the transformer/constant perveance electron beam load presented by the klystron [3].

Curry adopted a similar approach to the design of hybrid Type E PFN's transformer coupled into collapsing electron beam loads [4]. The hybrid type PFN was folded into a transmission line geometry to lower the inductive feed to the PFN sections [4]. In addition, the inductance of each section was weakly and mutually
coupled to compensate for the capacitor series inductance [4]. The output PFN section inductance was mounted orthogonally to prevent mutual coupling. This allowed the transformer leakage inductance to be used as part of the last PFN section inductance [4].

To meet the electron beam modulator design requirements into the constant current load as outlined in Table 1.1., a hybrid Type E PFN was utilized similar to that reported for a cold cathode load [4]. The hybrid Type E PFN selected has the section inductances weakly coupled mutually; however, the last PFN inductor is mounted orthogonal to prevent mutual coupling [5]. The detailed calculation process utilized for the PFN/modulator design is presented in the following sections. It should be mentioned that an alternate approach to Type E PFN design has been reported by Ranon but not experimentally verified [6].

3.2 Impedance Matching of Pulse Forming Networks into a Constant Current Load

Although pulse forming networks and fixed impedance transmission lines are designed to discharge into a matched load impedance, the load can be time varying as in the case of the constant-current, thermionic cathode load. As the anode-cathode voltage varies, the thermionic load maintains a constant-current. Thus the impedance of the anode-cathode load varies in time. Moreover, the pulse-forming-network may be required to
operate into wide ranges of currents and voltages as shown in Table 1.1. The PFN-based modulator, however, is limited to an impedance range of about 2:1 by commercially available high energy capacitors. Although a 30% voltage reversal (2:1 impedance range) is tolerable, most present-day high energy capacitor manufacturers prefer less voltage reversal. For example, 20% if $10^8$ shot lifetimes are to be achieved. This, of course, necessitates either engineering a scheme to change the impedance of the pulse-forming-network to match the load impedance range, or the inclusion of parallel resistors to match the load impedance to the PFN impedance. The simplest method of matching the PFN to the wide load ranges is to use parallel resistors. As mentioned, the PFN impedance could be changed to match various load ranges. Unfortunately, many bulky mechanical switches would be required to switch the PFN capacitors or the inductors.

To satisfy the voltage and current requirements of the electron beam source, the modulator must operate into a 6:1 impedance range. To accommodate this requirement high voltage parallel resistors were selected to match the modulator impedance to the constant-current load impedance. Selection of the matching resistors, however, is not straightforward. Both the load-range and value of the matching resistor interact and, ultimately, influence the dynamic impedance range over which the PFN is matched to the load. To simplify the selection of both the operating range and the value of the parallel resistors
required, a set of equations is required. Given the average impedance of the PFN, \( R_{PFN} \), the average load resistance, \( R_{load} \), and the allowable reflection coefficient, \( \Gamma_l \), the matching resistor \( R_{par} \) is given by:

\[
R_{par} = \frac{R_{PFN} R_{load} + \Gamma_l R_{PFN} R_{load}}{\Gamma_l (R_{load} + R_{PFN}) - R_{load} + R_{PFN}} . \tag{3.2}
\]

Equation 3.2 is derived using a simplified model of the modulator. As a first approximation the PFN is modeled as a perfect transmission line, Fig. 3.3. An average impedance is assumed for the load resistance, \( R_{load} \), and the parallel matching resistor, \( R_{par} \). Because Eq. 3.2 was derived to select the parallel matching resistor for this impedance range, the time variation of the load during the applied voltage pulse has been disregarded. As a verification of the predicted operating ranges and matching resistor combinations, a circuit simulation code must be employed. In reality, the stray capacitance shunting an electron beam load may change the characteristic impedance of the load during the initial risetime of the pulse. As a first approximation, these time varying load characteristics have been ignored. Rather, an average value for the load resistance during the applied pulse voltage has been assumed.

To further simplify the derivation, the transmission line model is reduced to a voltage source with a characteristic impedance of the PFN, \( R_{PFN} \) (Fig. 3.3b). Combining the equations for both the reflection
Perfect T-Line (Characteristic Impedance $Z_0$) (approximation of PFN)

A. Simple approximation of a pulse forming network.

$$R_{load_p} = \frac{R_{load} R_{par}}{R_{load} + R_{par}} \quad \Gamma_1 = \frac{R_{load_p} - R_{PFN}}{R_{load_p} + R_{PFN}}$$

B. Model used for initial computer modeling.

Figure 3.3 Approximate models of PFN used to select initial modulator operation ranges.
coefficient and the parallel load-matching resistor combination gives Eq. 3.2. Equation 3.2 is required to select the parallel matching resistance having been given the maximum allowable reflection coefficient (voltage reversal), the average load resistance and the average PFN impedance. It should be noted that Eq. 3.2 simplifies to Eq. 3.3 for $\Gamma_1 = 0$,

$$R_{\text{par}} = \frac{R_{\text{PFN}} \cdot R_{\text{load}}}{R_{\text{load}} - R_{\text{PFN}}}.$$  \hspace{1cm} (3.3)

With some algebraic manipulation, Eq. 3.3 reduces to the equation for the parallel combination of $R_{\text{load}}$ and $R_{\text{par}}$.

Given $R_{\text{par}}$, one additional equation is required to predict the load resistance range. That is, given a parallel matching resistor, the modulator's load range can be calculated. Equation 3.4 below was independently derived to calculate the load resistance range based on a fixed PFN impedance, $R_{\text{PFN}}$, the selected parallel resistance $R_{\text{par}}$, and the maximum or desired reflection coefficient $\Gamma_1$.

$$R_{\text{load}} = \frac{R_{\text{PFN}} \cdot R_{\text{par}} + \Gamma_1 \cdot R_{\text{PFN}} \cdot R_{\text{par}}}{R_{\text{par}} - \Gamma_1 \cdot R_{\text{par}} - \Gamma_1 \cdot R_{\text{PFN}} - R_{\text{PFN}}}.$$  \hspace{1cm} (3.4)

Note that Eq. 3.4 calculates the allowable load impedance for a given matching resistor $R_{\text{par}}$. Thus, the operating range for the modulator can be calculated for a given voltage reversal, $\Gamma_1$.  
To illustrate the advantages in using Eqs. 3.2 - 3.4 for designing a modulator, a review of the design procedure is necessary. Three conditions should be calculated and satisfied before the design procedure is initiated. First the impedance of the PFN must be known or calculated from the system requirements. Additionally, the allowable reflection coefficient or voltage reversal must also be derived from the capacitor and thyatron specifications. A reflection coefficient of -20% is not unusual and is consistent with reverse voltage ratings of capacitors. Consequently, the minimum and maximum required load impedance \( R_{\text{load(min)}} \) and \( R_{\text{load(max)}} \) must also be known a priori. The second condition is that a negative voltage reflection rather than a positive reflection should be chosen. The negative voltage reflection is desirable for two reasons. The negative reflection can be damped out with snubber diodes if necessary. Moreover, a positive reflection results in a "staircase" type waveform across the load, which effectively increases the deionization time of the thyatron by a large factor. A reverse voltage (or reflection coefficient) thus allows the system to recover more quickly and so reduce the probability of thyatron "hang-up". The third condition is perhaps the most difficult to calculate. The required efficiency at the peak power transfer of the modulator must be set. Because, the efficiency of the system is integral to the selection of the parallel resistance \( R_{\text{par}} \) at the highest peak modulator current and voltage output,
the dynamic impedance range must be calculated from Eqs. 3.2 - 3.4. The parallel combination, \( R_{\text{loadp}} \), of the load impedance, \( R_{\text{load}} \), and the parallel matching resistor, \( R_{\text{par}} \), is defined as

\[
R_{\text{loadp}} = \frac{R_{\text{load}} \ast R_{\text{par}}}{R_{\text{load}} + R_{\text{par}}}.
\]  

(3.5)

Similarly, the reflection coefficient is defined by

\[
\Gamma_1 = \frac{R_{\text{loadp}} - R_{\text{PFN}}}{R_{\text{loadp}} + R_{\text{PFN}}}. 
\]  

(3.6)

Thus, as \( R_{\text{par}} \) is increased or decreased with respect to the load impedance \( R_{\text{load}} \), the efficiency of the system varies. If an efficiency of 80-90% is required, this sets the minimum matching resistance, \( R_{\text{par}} \), at four to five times the minimum load resistance, \( R_{\text{load(min)}} \).

As a simple design example, assume that an 80% efficiency is required at the peak modulator voltage and current to minimize the power dissipated in the parallel matching resistance. Thus, \( R_{\text{par}} \) is about four times the minimum load resistance \( R_{\text{load(min)}} \). Because there is a one-to-one correspondence between the load impedance \( R_{\text{par}} \) and \( \Gamma_1 \), the maximum allowable load impedance for a given \( \Gamma_1 \) in this range can be calculated using Eq. 3.4. The maximum load impedance follows from the assumption that the load is always slightly undermatched with respect to the PFN impedance. Thus, for \( \Gamma_1 \approx 0 \), Eq. 3.4
sets the maximum allowable load resistance $R_{load(max)}$ for a given range. Table 3.0 summarizes the relative values and ranges of $R_{par}$, $R_{load}$, $R_{loadp}$ and $\Gamma_1$.

Table 3.0

$R_{load(min)} \leq R_{load} \leq R_{load(max)}$

$R_{loadp(min)} \leq R_{loadp} \leq R_{loadp(max)}$

$-20\% < \Gamma_1 < 0$

$(R_{par(min)} = 4 * R_{load(min)})$

To accommodate the required 6:1 impedance range, three to four parallel matching resistor values are required, along with three to four modulator operating ranges. The initial calculated impedance range $R_{load(min)} < R_{load} < R_{load(max)}$ may cover a small dynamic range only. For example we may have $R_{load(max)}/R_{load(min)} \approx 1.4:1$. Subsequently, additional calculations may be required. Once an initial $R_{par}$ and impedance range is calculated, additional matching resistors and allowable values for $R_{load}$ can be computed using the aforementioned equations. It should also be noted that the initial $R_{load(max)}$ becomes the minimum load resistance $R_{load(min2)}$ for the next range. Thus, using Eq. 3.2 and $\Gamma_1 = -20\%$, a second matching resistor, $R_{par2}$ can be defined. Similarly, $R_{load(max2)}$ can be set from Eq. 3.4 and $\Gamma_1 \approx 0$. An iterative analysis is required if a low $\Gamma_1$ is desired over the entire impedance range. Since there is no unique solution to the
equations, the designer is free to make various "trade-offs" to arrive at a sound design.

To facilitate the iterative calculations a transmission line program was written utilizing Eqs. 3.2 to 3.6. The computer program calculates $R_{\text{LOAD}}'$, $\Gamma_1'$, efficiency $\eta_{\text{EFF}}'$, current $I_{\text{PAR}}'$ through $R_{\text{PAR}}'$, current $I_{\text{LOAD}}$ through the load, and the voltage across the load, given the peak pulse voltage, the minimum and maximum load resistances, and the PFN impedance $R_{\text{PFN}}'$ and $R_{\text{PAR}}'$. The computer program listed in Appendix B, is written in IBM Basic. Because the attainment of a "good value" of efficiency in a continuous repetitive system is important in reducing component heating, a theoretical "best case" efficiency calculation was also included in the program. For the program, "efficiency" was defined as the ratio of power delivered to the load to the total power delivered by the modulator to the constant current load. Equation 3.7 defines the efficiency as

$$\eta_{\text{EFF}} = \frac{I_{\text{LOAD}}^2 R_{\text{LOAD}}}{I_{\text{LOAD}}^2 R_{\text{LOAD}} + I_{\text{PAR}}^2 R_{\text{PAR}}}$$  \hspace{1cm} (3.7)

for the simplified model shown in Fig. 3.3.
3.3 Circuit Design of the 250 kV, 45 μs Modulator for the Constant Current, Grid-Controlled Cathode Load

Both independent control of the electron beam energy (acceleration voltage) and electron beam current density were required for a CO₂ laser. To meet the 187 kV to 250 kV (150 keV - 220 keV, ± 10%), 45 μs pulse requirements imposed by the electron beam specifications, a line type modulator was designed using Eqs. 3.2-3.4. Design goals included operation of the modulator over a 20 A - 160 A current range into a 1.7 kΩ to 11 kΩ electron beam load. The specifications of a 45 μs pulse-width along with the 6:1 impedance range and lifetime of 10⁸ shots strongly influenced the modulator topology.

To preionize the CO₂ laser discharge as well as sustain the discharge, a 150 keV - 220 keV, 12.5 A - 50 A a post-foil electron beam current was required (Table 1.1). However, because the electron beam is vacuum immersed, some loss of electron energy and current is common. As the beam traverses the foil-Hibachi structure which links the vacuum chamber to the laser chamber, a considerable fraction of the electron beam may be intercepted. To compensate for the electron losses and allow operation into the parallel matching resistors, the modulator was designed to supply 20 A - 160 A. An overview of the electron losses is shown in Table 3.1.
Table 3.1

<table>
<thead>
<tr>
<th>Description</th>
<th>Calculated Transparency</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. Foil Electron Loss at Low Energies (30%)</td>
<td>70%</td>
</tr>
<tr>
<td>2. Geometric Hibachi Transparency (25%)</td>
<td>75%</td>
</tr>
<tr>
<td>3. Electron Loss from Divergence and Interception from Hibachi Ribs (10%)</td>
<td>90%</td>
</tr>
<tr>
<td><strong>Total Electron Transparency</strong></td>
<td>~ 47%</td>
</tr>
</tbody>
</table>

Calculation of the actual time-varying impedance into which the modulator operates is difficult. The impedance history is complicated by the non-linear electron energy loss through the foil and the ripple appearing on the anode cathode acceleration pulse. Without a detailed circuit simulation the ripple on the anode-cathode modulator pulse is difficult to deduce. However, some rather simple approximations do allow an initial estimate of the average impedance range of the modulator to be calculated. At low electron energies (150 keV), a 20-30% loss of electron energy is common when the beam passes through a 0.7 mil (17.78 microns) titanium foil. Thus, the modulator is required to operate from 187 kV - 250 kV when the ± 10% ripple specification is included. Note that the average modulator voltage that is required varies from about 150 kV to 220 kV with peak voltages required to be in the range from 187 kV to 250 kV. When the voltage range is considered in conjunction with the required electron beam current, the effective cathode impedance range is 1.7 kΩ to 11 kΩ. The modulator load impedance is somewhat less
than the 1.7 kΩ value.

The PFN impedance was based upon the modulator thyratron specifications. Electron beam guns present a harsh load to a conventional thyratron. Even under normal operating conditions an electron gun will short out occasionally. The subsequent high reverse voltage and current can damage a conventional thyratron. Moreover, when the modulator operates into a mismatched load (for which it is designed) current and voltage reversals of 20% or so are common. Commercial, hollow anode thyratrons which can handle such reversals under continuous operation are available [7]. These hollow anode thyratrons can conduct large reverse currents, even up to 80% of the forward current. During forward conduction, plasma is stored in anode slots. Upon reversal, the anode plasma provides a source of electrons. Extensive lifetime testing of both 35 kV and 50 kV hollow anode thyratrons had demonstrated lifetimes of $10^8$ pulses at 1 μs pulse widths [8]. However, no performance or lifetime data has been published for multi-microsecond, continuous operation of hollow anode thyratrons, much less for 45 μs pulse widths. A 35 kV hollow anode, CX1574C was chosen as the primary switch in the modulator. To prevent thyratron prefires the PFN charge voltage was limited to 32 kV. Scaling of the thyratron’s 1 μs, 10-12 kA current rating to 45-50 μs pulse widths from empirical curves set the peak thyratron discharge current at 3 kA - 4 kA [9].
On the basis of the thyratron ratings, the modulator was configured around a 5.5 Ω, 32 kV, PFN and was impedance matched to the load with a 1:19 iron-core transformer (Fig. 2.7). Utilizing Eqs. 3.2-3.4, an iterative analysis was used to determine both the impedance range and the required matching resistances for the modulator. At the highest modulator power (or current level), an 80% (or better) efficiency was required. As aforementioned, to achieve the desired 80% modulator efficiency, the first matching resistor $R_1$ is chosen to be about four times the minimum load (electron beam) resistance of 1700 Ω or about 7000 Ω. Once the matching resistor $R_{\text{par}} \approx 7000$ Ω was calculated, Eq. 3.4 was used to predict the maximum operating impedance of the range (in this case, 2600 Ω, Fig. 2.7). This design is based on two assumptions. The first assumption is that the characteristic load impedance driven by the modulator is less than the modulator output impedance. In this case, the output or secondary impedance of the modulator is $5.5 \Omega \times (19)^2$ (or 1985 Ω). This impedance is satisfactory at the low end of the operating range for $R_{\text{loadp}}$ is 1367.8 Ω when the 7000 Ω is considered in combination with the 1700 Ω electron-beam load. The highest impedance of the operating range is calculated using Eq. 3.4. Secondly, it is assumed that at the maximum impedance in an operating range the reflection coefficient is zero, or slightly negative. From Eq. 3.4 the maximum impedance of the operating range is 2600 Ω. On the basis of a preliminary calculation, the modulator
operating impedance ranges were calculated with Eqs. 3.2 to 3.4 and are summarized in Table 3.2, along with the required matching resistors.

Table 3.2

<table>
<thead>
<tr>
<th>Electron Beam Load Impedance Range (Ω)</th>
<th>Electron Beam Dynamic Impedance Range</th>
<th>( R_{\text{par}}(Ω) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>-20% &lt; ( R_1 ) &lt; 0%</td>
<td></td>
<td>( R_{\text{par}} )</td>
</tr>
<tr>
<td>1700 &lt; ( R_L ) &lt; 2600</td>
<td>1.4:1</td>
<td>7071 (D) (( R_1 ))</td>
</tr>
<tr>
<td>2600 &lt; ( R_L ) &lt; 3760</td>
<td>1.4:1</td>
<td>3800 (C) (( R_1, R_2 ))</td>
</tr>
<tr>
<td>3760 &lt; ( R_L ) &lt; 7229</td>
<td>1.9:1</td>
<td>2560 (B) (( R_1, R_3 ))</td>
</tr>
<tr>
<td>7229 &lt; ( R_L ) &lt; 11,000</td>
<td>1.9:1</td>
<td>1951 (A) (( R_1, R_2, R_3 ))</td>
</tr>
</tbody>
</table>

Matching Resistors

\[ R_1 = 7071 \, \Omega \quad R_2 = 8214 \, \Omega \quad R_3 = 4012 \, \Omega \]

To confirm the preliminary impedance calculations, the basic transmission line program (Appendix B), written from Eqs. 3.2 - 3.6 was utilized. Although Eqs. 3.2 - 3.6 allow the impedance ranges to be calculated easily, calculation of the peak voltage and current into the electron-beam load is much more difficult. Because the cathode current is controlled only by the grid pulse, a slight change in anode-cathode voltage \( V_{AK} \) results in a dramatic change in the total load impedance \( R_{\text{loadp}} \). Rather than account for the constant current characteristics of the electron-beam load in the transmission line program, a slightly different philosophy was adopted. The computer program was written to allow the minimum and maximum electron-beam load impedances (\( R_{\text{load}} \)) to be input along with the PFN impedance, \( R_{\text{PFN}} \), and \( R_{\text{par}} \). The theoretical peak voltage
is then specified (e.g., 250 kV) for a matched load condition. Given the impedance parameters and the peak voltage, the program then tabulates the voltage across the load, the current through the load, the parallel matching resistance ($R_{loadp}$ combination), the reflection coefficient ($\gamma$, $\Gamma_1$) and the modulator efficiency for a given range. In confirmation of the preliminary calculations, the four operating ranges were calculated using the program. Tables 3.3 - 3.6 summarize the results of the calculations. Note that the calculated ranges satisfy the design criteria. That is, the load is always slightly undermatched and that the reflection coefficient is kept below -18%. Moreover, at peak power, the efficiency of the modulator is about 80%.

After final calculation of the modulator operation ranges, a detailed circuit design is required to tailor the PFN characteristics to the electron-beam load. Because a change in load impedance affects the flatness, width, risetime, and fall-time of the pulse, a circuit analysis code is utilized to tune the inductance and capacitance of each section of the PFN. Moreover, the risetime of the cathode current and phasing of the cathode current (grid-pulse), control the overshoot present on the anode-cathode voltage pulse. Details of the circuit simulation and the analysis of circuit simulation results are given in the following sections. In essence, we use the simulation results to assist in tuning the modulator to the desired load.
Table 3.3
Calculated electron beam system load impedance, range one.

\[ V_{\text{PULSE}} = 250000 \]
\[ R_{\text{LOAD MIN}} = 1700 \]
\[ R_{\text{LOAD MAX}} = 2600 \]

\[ R_{\text{LOAD}} \quad R_{\text{LOADP}} \quad \text{GAMMA} \]
\[ \begin{array}{ccc}
1700 & 1367.816 & -0.1840793 \\
1790 & 1425.484 & -0.1640578 \\
1880 & 1481.982 & -0.1450882 \\
1970 & 1537.347 & -0.1270895 \\
2060 & 1591.612 & -0.1099892 \\
2150 & 1644.809 & -9.372154E-02 \\
2240 & 1696.972 & -7.822723E-02 \\
2330 & 1748.124 & -6.345241E-02 \\
2420 & 1798.302 & -4.934804E-02 \\
2510 & 1847.529 & -3.586955E-02 \\
2600 & 1895.833 & -2.297615E-02 \\
\end{array} \]

\[ V_{\text{LOAD}} \quad I_{\text{LOAD}} \quad I_{\text{PAR}} \quad \text{EFF} \]
\[ \begin{array}{cccc}
203980.2 & 119.9883 & 29.14003 & .8045977 \\
208985.5 & 116.7517 & 29.85508 & .7963596 \\
213728 & 113.6851 & 30.53257 & .7882883 \\
218227.6 & 110.7755 & 31.17538 & .7803791 \\
222502.7 & 108.011 & 31.7861 & .7726269 \\
226569.6 & 105.3812 & 32.36709 & .7650273 \\
230443.2 & 102.8764 & 32.92046 & .7575758 \\
234136.9 & 100.4879 & 33.44813 & .750268 \\
237663 & 98.20785 & 33.95186 & .7430998 \\
241032.6 & 96.02893 & 34.43323 & .7360674 \\
244256 & 93.94459 & 34.89371 & .7291666 \\
\end{array} \]

\[ R_{\text{PFN}} = 1985 \]
Parallel Resistance, \( R_{\text{PAR}} = 7000 \)
No. of Points = 11
Table 3.4
Calculated electron beam system load impedance, range two.

<table>
<thead>
<tr>
<th>( R_{\text{LOAD}} )</th>
<th>( R_{\text{LOADP}} )</th>
<th>( \text{GAMMA} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>2600</td>
<td>1543.75</td>
<td>-0.1250443</td>
</tr>
<tr>
<td>2716</td>
<td>1583.917</td>
<td>-0.1123824</td>
</tr>
<tr>
<td>2832</td>
<td>1622.678</td>
<td>-0.1004308</td>
</tr>
<tr>
<td>2948</td>
<td>1660.107</td>
<td>-8.913136E-02</td>
</tr>
<tr>
<td>3064</td>
<td>1696.27</td>
<td>-7.843206E-02</td>
</tr>
<tr>
<td>3180</td>
<td>1731.232</td>
<td>-6.828636E-02</td>
</tr>
<tr>
<td>3296</td>
<td>1765.051</td>
<td>-5.865233E-02</td>
</tr>
<tr>
<td>3412</td>
<td>1797.782</td>
<td>-4.949229E-02</td>
</tr>
<tr>
<td>3528</td>
<td>1829.476</td>
<td>-4.077206E-02</td>
</tr>
<tr>
<td>3644</td>
<td>1860.183</td>
<td>-3.246069E-02</td>
</tr>
<tr>
<td>3760</td>
<td>1889.947</td>
<td>-2.453014E-02</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>( V_{\text{LOAD}} )</th>
<th>( I_{\text{LOAD}} )</th>
<th>( I_{\text{PAR}} )</th>
<th>( \text{EFF} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>218738.9</td>
<td>84.13036</td>
<td>57.56288</td>
<td>0.59375</td>
</tr>
<tr>
<td>221904.4</td>
<td>81.70265</td>
<td>58.3959</td>
<td>0.5831799</td>
</tr>
<tr>
<td>224892.3</td>
<td>79.41112</td>
<td>59.18218</td>
<td>0.5729795</td>
</tr>
<tr>
<td>227717.2</td>
<td>77.24463</td>
<td>59.92557</td>
<td>0.5631298</td>
</tr>
<tr>
<td>230392</td>
<td>75.19321</td>
<td>60.62947</td>
<td>0.5536131</td>
</tr>
<tr>
<td>232928.4</td>
<td>73.24793</td>
<td>61.29695</td>
<td>0.5444126</td>
</tr>
<tr>
<td>235336.9</td>
<td>71.40076</td>
<td>61.93077</td>
<td>0.535513</td>
</tr>
<tr>
<td>237626.9</td>
<td>69.64447</td>
<td>62.5334</td>
<td>0.5268996</td>
</tr>
<tr>
<td>239807</td>
<td>67.97251</td>
<td>63.1071</td>
<td>0.518559</td>
</tr>
<tr>
<td>241884.8</td>
<td>66.37893</td>
<td>63.6539</td>
<td>0.5104782</td>
</tr>
<tr>
<td>243867.5</td>
<td>64.85837</td>
<td>64.17565</td>
<td>0.5026456</td>
</tr>
</tbody>
</table>

\( V_{\text{PULSE}} = 250000 \) \( R_{\text{PFN}} = 1985 \)
\( R_{\text{LOAD \ MIN}} = 2600 \) \( \text{Parallel Resistance, } R_{\text{PAR}} = 3800 \)
\( R_{\text{LOAD \ MAX}} = 3760 \) \( \text{No. of Points} = 11 \)
Table 3.5
Calculated electron beam system load impedance, range three.

\[ V_{\text{PULSE}} = 250000 \quad R_{\text{PFN}} = 1985 \]

\[ R_{\text{LOAD MIN}} = 3760 \quad \text{Parallel Resistance, } R_{\text{PAR}} = 2560 \]

\[ R_{\text{LOAD MAX}} = 7229 \quad \text{No. of Points} = 11 \]

<table>
<thead>
<tr>
<th>R_{\text{LOAD}}</th>
<th>R_{\text{LOADP}}</th>
<th>GAMMA</th>
</tr>
</thead>
<tbody>
<tr>
<td>3760</td>
<td>1523.038</td>
<td>-0.1316867</td>
</tr>
<tr>
<td>4106.9</td>
<td>1576.994</td>
<td>-0.1145442</td>
</tr>
<tr>
<td>4453.8</td>
<td>1625.614</td>
<td>-9.953612E-02</td>
</tr>
<tr>
<td>4800.7</td>
<td>1669.65</td>
<td>-8.628739E-02</td>
</tr>
<tr>
<td>5147.6</td>
<td>1709.722</td>
<td>-7.450566E-02</td>
</tr>
<tr>
<td>5494.5</td>
<td>1746.343</td>
<td>-6.396008E-02</td>
</tr>
<tr>
<td>5841.4</td>
<td>1779.94</td>
<td>-5.446583E-02</td>
</tr>
<tr>
<td>6188.3</td>
<td>1810.872</td>
<td>-4.587311E-02</td>
</tr>
<tr>
<td>6535.199</td>
<td>1839.444</td>
<td>-3.805936E-02</td>
</tr>
<tr>
<td>6882.099</td>
<td>1865.917</td>
<td>-3.092329E-02</td>
</tr>
<tr>
<td>7228.999</td>
<td>1890.514</td>
<td>-2.438031E-02</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>V_{\text{LOAD}}</th>
<th>I_{\text{LOAD}}</th>
<th>I_{\text{PAR}}</th>
<th>EFF</th>
</tr>
</thead>
<tbody>
<tr>
<td>217078.3</td>
<td>57.7336</td>
<td>84.79622</td>
<td>.4050633</td>
</tr>
<tr>
<td>221364</td>
<td>53.9005</td>
<td>86.4703</td>
<td>.3839866</td>
</tr>
<tr>
<td>225116</td>
<td>50.5447</td>
<td>87.93593</td>
<td>.3649947</td>
</tr>
<tr>
<td>228428.2</td>
<td>47.58227</td>
<td>89.22975</td>
<td>.3477931</td>
</tr>
<tr>
<td>231373.6</td>
<td>44.94786</td>
<td>90.3803</td>
<td>.3321397</td>
</tr>
<tr>
<td>234010</td>
<td>42.58987</td>
<td>91.41015</td>
<td>.3178348</td>
</tr>
<tr>
<td>236383.6</td>
<td>40.46694</td>
<td>92.33732</td>
<td>.3047111</td>
</tr>
<tr>
<td>238531.7</td>
<td>38.5456</td>
<td>93.17645</td>
<td>.2926283</td>
</tr>
<tr>
<td>240485.2</td>
<td>36.79844</td>
<td>93.93951</td>
<td>.2814672</td>
</tr>
<tr>
<td>242269.2</td>
<td>35.2028</td>
<td>94.6364</td>
<td>.2711261</td>
</tr>
<tr>
<td>243904.9</td>
<td>33.73979</td>
<td>95.27536</td>
<td>.261518</td>
</tr>
</tbody>
</table>
Table 3.6
Calculated electron beam system load impedance, range four.

\[ V_{\text{PULSE}} = 250000 \quad R_{\text{PFN}} = 1985 \]
\[ R_{\text{LOAD MIN}} = 7229 \quad \text{Parallel Resistance, } R_{\text{PAR}} = 1951 \]
\[ R_{\text{LOAD MAX}} = 11000 \quad \text{No. of Points} = 11 \]

<table>
<thead>
<tr>
<th>( R_{\text{LOAD}} )</th>
<th>( R_{\text{LOADP}} )</th>
<th>( \Gamma )</th>
</tr>
</thead>
<tbody>
<tr>
<td>7229</td>
<td>1536.359</td>
<td>- .1274056</td>
</tr>
<tr>
<td>7606.1</td>
<td>1552.72</td>
<td>- .1221917</td>
</tr>
<tr>
<td>7983.2</td>
<td>1567.839</td>
<td>- .1174163</td>
</tr>
<tr>
<td>8360.3</td>
<td>1581.852</td>
<td>- .1130265</td>
</tr>
<tr>
<td>8737.399</td>
<td>1594.876</td>
<td>- .1089771</td>
</tr>
<tr>
<td>9114.499</td>
<td>1607.012</td>
<td>- .1052302</td>
</tr>
<tr>
<td>9491.599</td>
<td>1618.348</td>
<td>- .1017531</td>
</tr>
<tr>
<td>9868.698</td>
<td>1628.961</td>
<td>-9.851761E-02</td>
</tr>
<tr>
<td>10245.8</td>
<td>1638.918</td>
<td>-9.549941E-02</td>
</tr>
<tr>
<td>10622.9</td>
<td>1648.278</td>
<td>-9.267736E-02</td>
</tr>
<tr>
<td>11000</td>
<td>1657.092</td>
<td>-9.003286E-02</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>( V_{\text{LOAD}} )</th>
<th>( I_{\text{LOAD}} )</th>
<th>( I_{\text{PAR}} )</th>
<th>( \text{EFF} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>218148.6</td>
<td>30.17688</td>
<td>111.8138</td>
<td>.2125272</td>
</tr>
<tr>
<td>219452.1</td>
<td>28.85212</td>
<td>112.4818</td>
<td>.2041414</td>
</tr>
<tr>
<td>220645.9</td>
<td>27.63878</td>
<td>113.0938</td>
<td>.1963923</td>
</tr>
<tr>
<td>221743.4</td>
<td>26.52338</td>
<td>113.7563</td>
<td>.1892099</td>
</tr>
<tr>
<td>222755.7</td>
<td>25.49451</td>
<td>114.1752</td>
<td>.1825343</td>
</tr>
<tr>
<td>223692.5</td>
<td>24.54248</td>
<td>114.6553</td>
<td>.1763138</td>
</tr>
<tr>
<td>2245651.7</td>
<td>23.659</td>
<td>115.1008</td>
<td>.1705032</td>
</tr>
<tr>
<td>225370.6</td>
<td>22.83691</td>
<td>115.5154</td>
<td>.1650634</td>
</tr>
<tr>
<td>226125.2</td>
<td>22.07004</td>
<td>115.9022</td>
<td>.15996</td>
</tr>
<tr>
<td>226830.7</td>
<td>21.35299</td>
<td>116.2638</td>
<td>.1551627</td>
</tr>
<tr>
<td>227491.8</td>
<td>20.68108</td>
<td>116.6027</td>
<td>.1506448</td>
</tr>
</tbody>
</table>
3.3.1 Circuit Analysis and Simulation of a 52 μs, 5.5 Ω Tapered Impedance Type E PFN

The baseline design for the pulse forming network utilized a 52 μs, 5.5 Ω, six section Type E PFN. The 52 μs pulse-width was determined from an equivalent square wave argument. Because a PFN is an imperfect approximation of a distributed transmission line, the PFN is inferior with respect to both risetime and fall-time.

The documented risetime of a five-section PFN is about 8% into a resistive load [1]. The modulator PFN however utilized a six-section line working into a constant current/resistive load. Large stray capacitance across the electron-beam load was also present due to the 2500 cm² cathode area (Table 1.1). As a "first-cut" analysis it was assumed that the risetime and fall-time would be about 7 μs, or about 6.5% of the flat part of the pulse. An equivalent square-wave argument was then used to determine the pulse-width used to calculate the PFN capacitance. Donaldsen has shown an equivalent square-wave based upon the required pulse risetime and fall-time can be used to calculate the required PFN capacitance [10]. That is, for the initial calculations, we assume that the rise- and fall-time is 7 μs. If it is assumed that the rise- and fall-times are approximately linear at the FWHM points, a 52 μs equivalent square-wave pulse can be used to calculate the modulator PFN capacitance. The characteristic impedance of the PFN can be calculated in a similar manner. It was earlier noted
that a PFN having an average impedance of 5.5 Ω could be switched by available commercial thyratrons. If the average impedance of the PFN is 5.5 Ω, then the output PFN section may have a slightly higher output impedance, e.g., 6.5 Ω, and the opposing PFN section, a lower impedance, e.g., 4.5 Ω. A 6 Ω PFN impedance affords an initial intermediate starting point for the circuit simulations.

The individual section inductance and capacitance of the Type E PFN can be derived from the equations for a distributed transmission line. That is, given the pulse width and impedance of the network, the total capacitance and inductance comprising the PFN can be calculated. For a transmission line, the pulse-width is given by Eq. 3.8,

\[
\tau = 2 \sqrt{L_T/C_T},
\]

(3.8)

where \( L_T \) is the total PFN inductance and \( C_T \) the total PFN capacitance. The factor of 2 arises from the manner in which a transmission line discharges into a fixed impedance load. If a charged transmission line is discharged into a matched impedance, the pulse voltage across the load is 50% of the charge voltage, but the pulse-width is twice the one-way transit time of the transmission line. The corresponding average impedance of the PFN is given by Eq. 3.9,

\[
R_{PFN} = \sqrt{L_T/C_T}.
\]

(3.9)
Combining Eqs. 3.8 and 3.9, we get

\[ C_T = \tau/(2 \times R_{PFN}), \]  

\[ L_T = R_{PFN} \times \tau/2, \]  

where \( R_{PFN} \) is the average impedance of the PFN and \( \tau \), the required pulse-width \([1]\). For a six-section, Type E PFN, the individual section capacitance is simply \( C_N = C_T/6 \).

As a starting point for the circuit simulations, a 6 Ω, 52 μs fixed impedance PFN was chosen. As shown in Fig. 3.4, the respective inductance and capacitance of each section are approximately equal. The first and last section inductances are adjusted to be about 20% higher than the middle section inductances \([1]\). From Eqs. 3.10 and 3.11 the total calculated inductance is 156 μH and the total capacitance is 4.44 μF.

Roark has shown that it is possible to derive the differential equations for the PFN and to model it into a nonlinear load using Fourier design techniques \([11]\). However, for our purpose a much simpler analysis was chosen. The initial 6 Ω, 52 μs PFN circuit along with constant current/resistive load, the associated stray circuit capacitance and the switch inductance was "coded" into a commercial circuit analysis code (SPICE). The PFN was then tuned on the basis of the SPICE-calculated voltage and current response of the line when working into the electron-beam load (constant-current load).
Figure 3.4 Fixed impedance, 52 μs, 6 Ω PFN shown along with the transformer model and stray capacitance, used to model the modulator circuit.
The modulator/electron beam system specifications complicate the circuit simulation. As part of the system specification, the modulator is 6 m removed from the cathode. High voltage coaxial cables connect the modulator to the electron beam gun. The coaxial cables which link the modulator to the cathode introduce a large amount of stray capacitance across the output of the modulator. Moreover, the cathode is grid-controlled. The cathode is turned on only above an anode-cathode voltage of 135 kV to prevent deposition of electrons into the laser's foil window. Thus the pulsed current source represents a time varying load which is partially capacitive, resistive, and a constant-current (load) during the anode-cathode accelerator pulse. During the risetime of the anode-cathode acceleration voltage the impedance of the load is determined only by a combination of the parallel matching resistor selected and the stray capacitance across the secondary of the transformer (Fig. 3.4). Until the electron beam has been "turned on" by the grid pulse, the modulator current is determined only by $R_{\text{par}}$ and the secondary stray capacitance (cable capacitance). Once the electron beam current, or in this case, the constant-current load, is "turned on" the modulator's load then changes to what is, predominantly, a resistive/constant-current load. The charged cable capacitance and the other stray capacitance across the modulator's output do however interact with the modulator circuit and to a lesser extent determine the anode-cathode voltage pulse flatness once the electron beam has
been "turned on". To model these interactions accurately, the circuit model must include the stray capacitance, the parallel matching resistor and the constant-current load. The model derived for the modulator/electron beam system, as shown in Fig 3.4, includes the stray load capacitance, the pulsed current source, and the parallel matching resistor to simulate the physical load present in the modulator/electron beam system.

Intuitively the coaxial cable linking the electron beam gun to the modulator complicates the modelling of the PFN response to the constant current load. Because the cable impedance is low compared to the driving impedance of the modulator, the cable, or in this case the three 67 Ω parallel cables, can be modeled as a lumped capacitance and inductance. Of course, this approximation is valid only when the transit time of the cables is much less than the risetime of the electron beam. For the modulator of Fig. (2.7) the three 67 Ω cables have a combined impedance of 22.3 Ω and a one-way transit time of 30 ns. When the cable impedance is compared with the 1985 Ω driving impedance of the modulator, the cables appear as a capacitance. The high cable capacitance dramatically alters the risetime of the cathode-anode acceleration voltage and must be used in the model. The 60 ns two-way transit time of the coaxial cables is substantially less than the theoretical 2 μs risetime of the electron beam current. Thus, the simple relation \( C = \frac{\tau}{Z_{\text{cable}}} \) gives the
equivalent cable capacitance of 1.34 nF. It was found that the cable capacitance is the dominant secondary capacitance across the transformer. Additional, stray secondary capacitance is introduced by the large 2500 cm\(^2\) cathode structure, and the grid-pulser case-to-ground. The cathode stray capacitance to ground is approximately 320 pF. The grid-pulser case-to-ground contributes an additional 150 pF. Similarly, the large secondary transformer winding has an effective capacitance of 200 pF. Rather than include all the discrete lumped capacitance values as part of the circuit model, the values can be summed. The 320 pF cathode capacitance, when combined with the cable capacitance, gives a total secondary stray capacitance of 1.66 nF. The stray secondary transformer capacitance and the grid-pulser capacitance have an equivalent 350 pF. To further simplify the model, the stray capacitance across the secondary of the modulator, the parallel matching resistor and the constant current load are referred to the primary of the model. When the 1.66 nF and 350 pF capacitance are referred to the primary of the modulator, the respective values are 0.6 \(\mu\)F and 0.126 \(\mu\)F as shown in Fig. 3.4. The combined 0.726 \(\mu\)F uncharged stray capacitance is equivalent to one PFN section capacitance, and must be charged during the risetime of the anode-cathode pulse. This, of course, complicates the tuning of the PFN to the desired \(\pm 10\%\) pulse flatness specifications and increases the amount of overshoot present on the leading edge of the applied cathode
Along with the lumped cable capacitance, the cable inductance was also calculated. Using the relation \( L = \tau Z_{\text{cable}} \), the equivalent cable inductance is 559 nH. When referred to the primary of the transformer, the equivalent cable inductance is 1.55 nH. Compared to the transformer 11 \( \mu \text{H} \) leakage inductance and the 26 \( \mu \text{H} \) PFN output inductance, the series cable inductance can be neglected.

The effect of the non-ideal pulse transformer through which the PFN is coupled to the load must also be included in the circuit model. An ideal pulse transformer would have an infinite risetime, no droop and no secondary capacitance. Although the transformer secondary capacitance has already been included in the model, the leakage inductance and self-inductance of the transformer must also be included in the model. These inductances account for the finite risetime and the droop of the pulse introduced by a physically realizable transformer [12]. For an ideal transformer the coupling coefficient \( K \) is 1; in reality \( K \) is somewhat less than unity. Since \( K \) is almost always less than unity for a physical transformer, not all of the magnetic flux generated by the primary is coupled to the secondary [1]. The associated leakage inductance introduced by this imperfect coupling increases the risetime of the transformer [12]. Several models of transformers have been developed which include the effect of this leakage inductance [13]. The model used for the circuit
simulation symmetrically distributes the leakage inductance about the transformer's self-inductance $L_p$ as shown in Fig. (3.4). For the 1:19, 250 kV, 52 μs pulse transformer selected, the primary leakage and self-inductances are 11 μH and 18.6 mH, respectively. Using an average load impedance and the average impedance of the PFN the droop is 0.7% and does not significantly effect the pulse-width or flatness [13].

SPICE, a commercial circuit simulation, was used to model the response of the circuit of Fig. 3.4 to the pulsed current load presented by the electron beam. During the simulations it was found that the untapered PFN of Fig. 3.4 was unable to meet the 45 μs pulse-width required by the CO$_2$ laser. As part of the iterative design process, the impedance of the PFN was tapered, as shown in Fig. 3.5. Tapering of the PFN impedance was performed by increasing the inductance at the front-end of the PFN, and lowering the inductance at the back of the PFN. Because each section of the PFN has a characteristic impedance given by the section inductance and capacitance, $Z_{section} = \sqrt{L_{section}/C_{section}}$, the impedance of the PFN is decreased along its length. The voltage across the load increases with time in response to tapering the PFN impedance. Figure 3.6 shows the normalized voltage waveforms of both the untapered and tapered PFN response to the pulsed current source load. The dashed line represents the response of the untapered PFN (Fig. 3.4) to a constant current source load (Fig. 3.6). During the initial risetime of the pulse, the stray
Figure 3.5 Tapered impedance, 52 µs, 5.5 Ω PPN model derived from iterative circuit simulations.
Figure 3.6  Normalized load voltage of fixed impedance PFN and tapered impedance PFN response into a constant-current load. The dashed line is the fixed impedance PFN voltage into the constant-current load. The solid line is the tapered impedance PFN voltage into a constant-current load.
secondary capacitance is charged. As the electron beam turns on, the stray secondary capacitance begins to discharge. The voltage pulse, or in this case, the anode-cathode voltage begins to droop, as seen in Fig. 3.6. This voltage droop can be reduced by decreasing the PFN impedance along its length, which effectively charges the stray capacitance by increasing the pulse voltage slope. Analysis of the voltage waveform from the untapered PFN circuit model also indicated that the pulse width corresponding to a ± 10% flatness specification was only 40 μs in width. The 40 μs pulse-width was 12.5% short of the desired 45 μs pulse-width.

As mentioned, an iterative tuning procedure was adopted during the course of the design process. Because of the risetime or turn-on time of the current source (electron beam load), phasing of the current source with respect to the anode-cathode voltage is important. As aforementioned, the load impedance seen by the modulator is complicated and varies in time. The load is not only capacitive and resistive, but also contains a constant-current load. To model the turn-on of the electron beam load, a constant current source with a 2 μs risetime (0-100%), 45 μs flat-top, and 4.5 μs fall-time was used. The constant current source was delayed 3.5 μs into the anode-cathode pulse (load voltage). Physically, this represented the electron beam turning on above a nominal 135 keV energy level. This prevents the low energy electrons from being deposited into the foil and subsequent overheating of the foil.
The final version of the tapered PFN is shown in Fig. 3.5. After several "tuning" iterations, during which small changes were made to the PFN section inductors, the simulated response of the modulator/PFN met the required ± 10% voltage flatness specification, as well as the required pulse-width of 45 µs (90-90). The normalized voltage pulse (Fig. 3.6) produced by the tapered PFN is contrasted with the voltage waveform of the untapered PFN of Fig. 3.4. The impedance tapering increases the usable portion of the waveform by 5 µs. Moreover, the rising slope of the pulse increases the average voltage or in the case of an electron beam, the average electron energy seen by the laser gas. The iterative tuning process allowed the voltage waveform to be tuned to ± 8.5%, 1.5% better than the required specification.

To understand the subtle differences produced by varying the impedance of the load slightly, the normalized waveforms produced at the maximum and minimum impedances (Table 3.2) of the four ranges were overlaid (Fig. 3.7). Three characteristic differences are apparent after review of the eight overlaid voltage waveforms. The phase of the maximum and minimum inflections or oscillations shift with respect to the initial peak of the waveform. This effect is accentuated on the third oscillation at the tail end of the pulse. As the load impedance changes, the rate at which the stray secondary capacitance discharges into the electron beam load also changes. Inherently, this also effects the fall-time of the pulse and the reversal across the
Figure 3.7 Tapered impedance PFN (normalized) load voltage across the constant-current load over the four operating ranges (Table 3.2), showing the subtle waveform differences between operating ranges.
anode-cathode gap. The reverse voltage was also found to be higher than originally calculated. The 30% voltage reversal as noted from the circuit simulations was 10% higher than the anticipated theoretical value. To decrease these high voltage reversals across the anode-cathode gap, a 375 kV - 600 Ω snubber diode-resistor pair was later added to the modulator design (Fig. 3.8).

The circuits of Figs. 3.4 and 3.5 reference all component values to the primary side of the transformer. This allowed the circuit simulation to be simplified and allowed the interaction of the PFN and uncharged stray capacitance to be better understood while tuning the circuit. A full listing of one of the circuit simulation codes is provided in Appendix C, with a detailed explanation of the code listing. After final tuning of the PFN, the primary voltage was scaled to represent the modulator secondary voltage (1:19) and plotted on an expanded scale. The secondary or electron beam voltage along with a reference current beam is shown in Fig. 3.9. As aforementioned, the grid-controlled electron beam is not turned on until after the anode-cathode voltage reaches a minimum value of -135 kV. Similarly, the grid pulse, or in the case of the simulation, the current source, is turned off prior to the anode-cathode voltage falling below -135 kV or, in this case, -205 kV. Figure 3.9 shows the simulated average -234 kV voltage into a parallel matching resistor $R_{\text{par}} = 4542$ and a constant current (84.7 A) load. The total load impedance is 2737 Ω. Correspondingly, the current supplied by the modulator
Figure 3.8 Simplified modulator schematic showing grid-pulser and matching resistors.
Figure 3.9 Simulated secondary modulator voltage ($V_{AK}$) into a pulsed constant-current load. The turn-on time of the pulse constant-current load is shown for reference only (dashed line).
is shown in Fig. 3.10. The simulated current pulse shown in Fig. 3.10 is the modulator current supplied to the electron beam (84.7 A), the current through the matching resistor (51.5 A), and the displacement current required to charge the stray capacitance across the secondary of the transformer. The initial, slow-rising, low current section of the waveform corresponds to the displacement current required to charge the secondary stray capacitance. The fast 2 μs linear pulse current risetime observed in the circuit simulation coincides with the modeled turn-on time of the beam current. Similarly, the linear 4.5 μs fall-time is a consequence of the turn-off time or fall-time of the beam model's current. Uniquely, the total current waveform supplied by the modulator is not linear with voltage. That is, an increase in voltage does not correspond to linear increase in modulator current, as would be observed into a strictly resistive load. Thus the modulator's circuit interaction with its load is complex. A small change in the PFN impedance can result in large observed changes in the modulator's output waveform, as seen from the circuit simulations.

3.4 Layout and Design of the 250 kV, 52 μs Modulator

The size and component layout of the modulator is driven by four components; the 250 kV, 1:19 step-up transformer, the 52 μs 1800 J, six-section, 5.5 Ω PFN, the charging inductor, and the grid pulser. The most massive component found in the modulator is the iron-
Figure 3.10 Simulated modulator current into matching resistors and pulsed constant-current load.
core, 250 kV pulse transformer. The 66.0 cm x 66.0 cm x 81.2 cm transformer size is a result of the volt-second product and the insulation spacing required by the 250 kV secondary of the transformer. To accommodate the 52 μs, 250 kV pulse, a volt-second product of 14 is required to prevent saturation of the core. The transformer pulse specifications include the calculated 8.25 μs (10-90) risetime, the 45 μs flat-top and the 14 μs (90-10) fall-time of the anode-cathode voltage. In order to reduce the amount of silicon iron-core required by the transformer, the full delta-B swing of the core is utilized. A 50 A DC reset current opposite in direction to the PFN current is used to reset the B-H loop of the core. The reset current is provided by a 12 V, 50 A power supply isolated from the 28-30 kV, 52 μs PFN pulse by a 100 mH isolation inductor. The 100 mH isolation inductor along with a few filter capacitors prevent transients from coupling into the power supply and allow the 12 V, 50 A DC power supply to be connected to the primary of the 1:19 step-up transformer. The modulator layout, along with the high voltage step-up transformer, is shown in Fig. 3.11.

Final aspects of the design of the 250 kV pulse transformer were settled by size and cost considerations. Although to first order the core volume is determined by the energy switched per pulse, core heating or core loss influence the transformer design [1]. It is well known that both ferromagnetic core loss and hysteresis loss depend upon laminate eddy currents [14]. The eddy
Figure 3.11 Photograph of modulator showing high voltage step-up transformer, PFN, and matching resistor.
current loss is strongly dependent on the thickness of the core laminate as shown by Charap and Winter [15, 16]. As the laminate thickness decreases the core loss decreases [15, 16]. However, laminate thickness directly determines the core cost and thus the final transformer cost [14]. Lord has reported multi-megawatt, multi-microsecond design and operation of both 1-mil to 4-mil laminate silicon iron core transformers [17]. It has also been found that substantial cost increases are incurred when the laminate thickness is decreased [14]. Based on this analysis a 4-mil laminate thickness was selected for the transformer. The estimated core loss for the pulse transformer was estimated to be less than 2%, or 360 watts when transferring 1800 J/pulse at 10 pps [18].

A photograph of the transformer and other circuit components is shown in Fig. 3.11. To increase the volt-second product of the transformer and reduce the overall size of the transformer the two secondaries are connected in series. The long pulse-width (52 μs) and the high (11 μH) leakage inductance of the transformer allow the two secondaries to be connected in series. One secondary winding is designed to operate up to about 150 kV. The second secondary winding is then connected to the first for operation up to 250 kV.

The secondary of the transformer is wound in the bifilar arrangement. The bifilar secondary winding allows 480 V, 1 kHz power to be routed to the electron gun filaments (Fig. 3.8). The secondary coil was wound
with 16 gauge wire. Because the filament power is delivered at 480 V and low current, heating of the secondary winding is negligible. To insure reliable operation of the transformer, oil insulation is required. The oil immersion provides both efficient heat sinking and high voltage insulation of the iron core, and primary-secondary windings.

The breakdown field for oil is both time dependent and polarity sensitive [19]. Liquids have been found to have a $t^{-1/3}$ breakdown dependence in the microsecond to nanosecond range [19]. The breakdown dependence of oil is polarity sensitive if the circuit drive impedance is above 10 $\Omega$ [19]. For an applied negative pulse the breakdown strength of transformer oil is about 1.7 times that of a positive pulse [19]. The negative electric field breakdown for transformer oil is calculated to be $-158 \text{ kV/cm}$ for a 100 cm$^2$ area if a 50 $\mu$s pulse is applied. To prevent flashover or breakdown between the primary and secondary, the transformer’s peak electric fields are designed for operation at about $-100 \text{ kV/cm}$ (well below the calculated breakdown field). Moreover, both secondaries of the 50 $\mu$s, 250 kV transformer are graded with respect to the electric field distribution. The inductance of the transformer winding grades the electric field, from turn-to-turn. Corona rings integral to the design of the transformer grade the electric field at the ends of the secondary winding, and prevent the primary-to-secondary electric field from exceeding $-100 \text{ kV/cm}$. 
Generous spacing between the 50 μs step-up transformer, and the surrounding low voltage components was included to insure breakdown-free operation of the modulator. Included in this design strategy were all the components or assemblies floating up to 250 kV. Connections to the secondary of the transformer and the grid pulser were also included in this list. A generous 15 cm spacing was selected between these high voltage components and their surrounding low voltage components as the design criterion. Although some geometries allow the peak electric field to be calculated, the many components in a modulator of this type make the calculation difficult. Sharp points associated with bolted connections can increase the field enhancement by a factor of five or more. In addition, it should be noted that the modulator was specifically designed to be raised and lowered into an oil tank (Fig. 2.5). During the lifting and lowering process, the oil can become aerated. Thus, after reviewing the breakdown formulae for oil, a conservative spacing of 15 cm was allowed around all high voltage components. In addition, connections to all the high voltage components were designed with electric field reduction in mind. Copper tubing, 5 cm in diameter, along with specially machined end-pieces allowed easy connection to the last turn of the pulse transformer. The large diameter copper tubing was also used to connect the high voltage matching resistors, the high voltage snubber diodes and the grid pulser. The large diameter tubing limited the resultant
electric fields to below -125 kV/cm.

The six 0.74 μF capacitors in the PFN were securely mounted using welded steel rails. Each 0.74 μF capacitor was conservatively rated at 35 kV and 3000 A peak discharge currents. Under continuous 10 pps operation and 20% voltage reversal, the manufacturer's estimated lifetime of the PFN capacitors is $10^9$ shots. To reduce the effect of stray inductance upon the modulators, output waveform and series inductance (ESL) of each capacitor was limited to 350 nH. The ESL included the inductance of the two capacitor feed-through connections and the internal inductance of the capacitor pads. In order to limit the power dissipated in the capacitors during each pulse, the internal series resistance of each capacitor was specified to be 20 mΩ or less.

To minimize ohmic losses from the PFN inductors, copper tubing 0.635 cm in diameter was used to fabricate the inductors. Inductors, $L_1 - L_5$, of the pulse forming network were fabricated on a continuous 16.5 cm diameter lucite form. The number of turns required for each inductor was calculated from Nagoaka’s formulae as given by

$$L = \frac{\pi \mu r^2 N^2 K}{b}$$

where $b$ is the length of the inductor, $r$ the average radius, $\mu$ the permeability of free space, and $K$ a correction factor for magnetic fringing effects. An
iterative calculation was used to arrive at both the diameter of the inductor form used and the required number of turns per inductor. The length of each inductor was calculated to be 30.4 cm with a radius of 8.25 cm. The calculated number of turns per inductor is summarized in Table 3.7. The theoretical value listed in Table 3.7 is the required PFN inductance from the circuit simulations. The calculated value is from Nagoaka's formulae with $K \approx 0.8$ in the range of interest [20]. The actual value shown in Table 3.7 is the measured inductance after tuning of the inductor. Using a Z-meter which digitally displays inductance, $L_1 - L_6$ was tuned by varying the pitch of the windings, the spacing, and the length. Insulated spacer blocks were then used to securely hold the inductor turns in place.

Table 3.7

<table>
<thead>
<tr>
<th>Inductor</th>
<th>Theoretical Value</th>
<th>Calculated Value</th>
<th>Actual Value</th>
<th>No. of Turns</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1$</td>
<td>16.6 $\mu$H</td>
<td>17.8 $\mu$H</td>
<td>18.6 $\mu$H</td>
<td>16</td>
</tr>
<tr>
<td>$L_2$</td>
<td>23.9 $\mu$H</td>
<td>22.6 $\mu$H</td>
<td>22.6 $\mu$H</td>
<td>18</td>
</tr>
<tr>
<td>$L_3$</td>
<td>27.6 $\mu$H</td>
<td>27.92 $\mu$H</td>
<td>26.3 $\mu$H</td>
<td>20</td>
</tr>
<tr>
<td>$L_4$</td>
<td>28.0 $\mu$H</td>
<td>27.92 $\mu$H</td>
<td>27.8 $\mu$H</td>
<td>20</td>
</tr>
<tr>
<td>$L_5$</td>
<td>29.1 $\mu$H</td>
<td>30.78 $\mu$H</td>
<td>30.9 $\mu$H</td>
<td>21</td>
</tr>
<tr>
<td>$L_6$</td>
<td>25.0 $\mu$H</td>
<td>25.19 $\mu$H</td>
<td>23.9 $\mu$H</td>
<td>19</td>
</tr>
</tbody>
</table>

With the exception of $L_1$ and $L_6$, the measured value of inductance is within 6% of the calculated value. The output inductor $L_6$ was conservatively made less than the
required theoretical value. At the time of fabrication, a review of the lead lengths required to connect $L_6$ to $C_6$ and $L_6$ to the pulse transformer, showed that about 1 $\mu$H of stray inductance was present. The inductance of $L_6$ was correspondingly reduced by about 1 $\mu$H. The increased inductance of $L_1$ was a result of adding extra lead length to connect to $C_1$. A value of 18.6 $\mu$H was measured for $L_1$. The calculated value from Nagoaka's formula was 17.8 $\mu$H or 4.5% low from the measured value.

The pulse forming network is resonantly charged from two 14.5 $\mu$F filter capacitors $C_{CH}$ (Fig. 3.8). To limit the complexity of the modulator, a passive charging scheme was selected. A 1 H, 20 kV, iron-core charging inductor along with a series 50 kV, 6 A average charging diode, resonantly charges the PFN in 6.6 ms. The charging inductor limits the peak current to 3 A, (18 kV charge voltage) and allows a commercial diode to operate reliably at 10 pps. In addition to limiting the peak charge current, the 1 H charging choke allows the thyatron to recover in the inter-pulse period.

During layout of the modulator, a single-point grounding philosophy similar to that described by Morrison [21] was adopted to eliminate ground loops. With the exception of the 250 kV pulse transformer secondary winding ground, and the 375 kV snubber diode stack, all ground tie points terminate on the modulator ceiling. A 30 cm wide ground plane does however link the secondary winding of the transformer to the single-point, ceiling ground plane.
The modulator’s thyatron is oil-immersed and mounted from the ceiling ground plane. To prevent overheating of the thyatron, forced oil cooling was provided. A pump located external to the modulator tank circulates oil drawn from the modulator tank through a set of nozzles located on the walls of the modulator. The oil is first cooled using a water/oil heat exchanger. The cooled oil is then directed at both the thyatron anode and cathode. This prevents overheating of the thyatron under the long pulse conditions present.

The modulator was built as a self-contained assembly. The welded, steel-frame assembly allowed the modulator to be built on the laboratory floor and then lowered on to supporting hydraulic lifts (Fig. 2.5). When flush with the oil tank, the modulator tank measured 219 cm (H) x 231.64 cm (L) x 231.64 cm (W). When filled with oil the modulator weighs 12,727 Kg and contains about 7,570 liters of oil.
REFERENCES
CHAPTER 3


18. Private Communication, Magne Stangenes, Stangenes Industries.


4.1 Grid Pulser Design Considerations

The design of the grid pulser is inherently more sophisticated than that of the modulator. The grid pulser specifications are imposed by the required ± 10% electron beam flatness specification (Table 1.1). The grid pulser design is further complicated by the space-charge-limited current load into which it must operate. Because the grid-cathode electron beam (current) load is space-charge-limited, the grid-cathode load is governed by the Child-Langmuir equation [1]. The grid impedance is therefore proportional to $V^{-1/2}$, thus the grid-pulser is required to operate into a constant perveance load. Two important implications arise from this fact. To meet the 5-20 mA/cm$^2$ post foil current requirements the grid pulser must operate over a wide voltage and current range. Consequently, the varying impedance complicates the pulse risetime, pulsewidth, and flatness of the grid-cathode control pulse.

Concurrent to the voltage dependent impedance presented by the grid-cathode load, a voltage dependent capacitance is present in gridded electron beam guns. This voltage dependent capacitance is classically known as the Miller capacitance [2]. If an equivalent circuit model of the triode is considered, the input capacitance across the grid-cathode region is described
where \( C_{gk} \) is the grid-to-cathode capacitance, \( C_{gA} \) the grid-to-anode capacitance, and \( A_0 \) the gain of the triode circuit. The grid-cathode input capacitance, \( C_i \), can thus exceed the original \( C_{gk} \) if the electron gun has a high current or voltage gain. If a modest grid-to-anode capacitance of even 90 pF for the 2500 cm\(^2\) cathode is considered in conjunction with a voltage gain of 500 - 600, the grid-cathode capacitance \( C_{gk} \sim 1.7 \) nF can be exceeded by a factor of thirty. To be consistent with this conclusion, any model derived for the grid pulser circuit simulations must also include the Miller effect [2]. Unfortunately, existing circuit simulation programs cannot be configured easily to model this capacitance in terms of the voltage gain \( A_0 \). As part of the investigation, an alternative model which describes the capacitance in terms of the modulator impedance, the grid voltage, and the anode current was derived. The model, described in section 4.1 allows the input grid capacitance to be simulated on existing circuit simulation programs. Both the modeling techniques and the capacitance model for the grid-cathode load are discussed. Relevant physical parameters which describe the interaction of the grid pulser with the 2500 cm\(^2\) cathode are also described in the following sections.
The voltage and current parameter space over which the grid pulser is required to operate is difficult to characterize theoretically. Derivation of the grid voltage and current is not straightforward. If the 2500 cm² electron beam gun geometry is considered, the filament surface area which emits current can not be calculated easily (Fig. 2.1A). In classic triode tubes, the extraction grid is normal to the entire surface of the cathode filament. Thus, the entire filament surface contributes to the extracted electron current, and the Child-Langmuir equation can be used to calculate the required grid control voltage. Unfortunately, the concentric hemi-cylindrical extraction grid-filament geometry is not easily analyzed using closed form solutions. Although a rough approximation of the grid voltage and current requirements can be ascertained from geometrical arguments, a computer code is necessary. In order to calculate the space-charge-limit and the electron trajectories, an electron particle trajectory code known as EGUN was utilized [3]. Using the results of the particle code, the geometrical transparency (0.6-0.7) of both the control and extraction grids, and the 0.75 cm spacing between the filament and extraction grid, the grid voltage and current were derived [3]. The corresponding grid pulser current and voltage range is 28 A - 166 A and 185 V - 650 V, respectively. The corresponding grid-cathode impedance over which the grid pulser must operate is 4 Ω to 6.6 Ω. Concurrent with these
impedance requirements, the grid pulse voltage must have a $\pm 6\%$ voltage flatness, a 2 $\mu$s risetime, 4 $\mu$s fall-time, and a nominal 45 $\mu$s pulse-width.

4.2 Analytic Model Derivation of the Grid-to-Cathode Load

Modeling of circuit or pulse-forming-network response to resistive or linear voltage loads is well documented. Available circuit simulation codes include network elements which allow simulation of these cases [4, 5, 6]. However, when the physical load departs dramatically from a resistive load (for example a plasma load or a space-charge-limited electron beam load), alternate mathematical models are required which can be implemented with available circuit simulation codes. During the investigation of the design of the grid pulser for the 2500 cm$^2$ cathode, it was found that available circuit simulation codes could not easily model the grid-to-cathode, space-charge-limited electron beam load, or the inherent Miller capacitance arising from the triode geometry. Because the grid-cathode current is proportional to $V_{gk}^{3/2}$, the impedance history of the grid-to-cathode pulse is non-linear. Moreover, the grid-cathode load includes a time varying capacitance which greatly accentuates the non-linear load characteristics over which the grid pulser operates. As the grid-cathode voltage changes slightly so does the Miller capacitance. This results
in a slight displacement current in addition to the space-charge-limited current. In order to develop a circuit model which takes into account the physical interaction between the grid pulser, the modulator and its space-charge-limited load, a Thevenin equivalent model of the system is derived. The model includes the modulator, the grid pulser, and the finite fixed capacitance between the grid and anode as shown in Fig. 4.1. The model derived from this Thevenin equivalent circuit allows the voltage dependent Miller capacitance to be modeled as a function of the circuit parameters, independent of the closed circuit gain of the system.

Two assumptions were made prior to the derivation using the Miller capacitance model. Physically the grid possesses a finite geometrical transparency. The transparency results in the grid intercepting a small portion of the cathode-to-anode current. However, the intercepted anode current represents a small portion of the current flow between the grid and cathode, and can be disregarded in order to simplify the capacitance model. The second assumption is dictated by the physics of electron flow. In order to extract electrons from the cathode, the grid must be positive with respect to the cathode potential. Thus, the model must take into account the current supplied by the grid pulser in order to extract current from the cathode.

Based on the above assumptions, the equations for the capacitance model can be developed from Fig. 4.1. Because the model assumes that the grid remains
Figure 4.1 Thevenin equivalent model of the electron beam system including the modulator and grid pulser.
positive and stable during anode voltage fluctuations, the grid current \( I_g \) must cancel the displacement current through \( C_{ga} \) or \( I_g = -I_c \) [7]. By definition, the displacement current through \( C_{ga} \) can also be written as

\[
I_c = C (\dot{V}_p - \dot{V}_g), \tag{4.2}
\]

where \( \dot{V}_p \) is the time derivative of the anode voltage and \( \dot{V}_g \) is the derivative with respect to time of the grid voltage. Correspondingly, the current \( I_m \) supplied by the modulator is the result of the anode-cathode current \( I_p \) and the displacement current \( I_c \) through the grid-anode capacitance, \( C_{ga} \). Thus, the modulator current can be written as

\[
I_m = I_p + I_c. \tag{4.3}
\]

Similarly, a representative expression for the anode current can be defined as

\[
I_p = \alpha V_g^{3/2} \tag{4.4}
\]

where \( \alpha \) is the perveance of the grid-anode space [1]. It should be noted that \( \alpha \) is derived from the geometry of the electron beam gun, and to first order, is independent of current and voltage variation. For the electron beam cathode of area 2500 cm\(^2\), we have \( \alpha \approx 0.01 \) for the grid-to-cathode region.
One additional equation is required before a grid-cathode circuit model can be derived. To deduce the interaction of the modulator and grid pulser, the anode voltage in terms of the modulator voltage, current, and impedance is required. Utilizing the Kirchhoff voltage equations and the Thevenin model shown in Fig. 4.1, the anode voltage can be written as

\[ V_P = V_m - I_m Z, \]  

(4.5)

where the anode voltage is a function of the modulator output impedance \( Z \), the modulator current \( I_m \) and the modulator voltage \( V_m \). Differentiation of Eq. 4.5 with respect to time gives

\[ \dot{V}_P = \dot{V}_m - \dot{I}_m Z, \]  

(4.6)

where \( \dot{V}_P \) and \( \dot{V}_m \) are the derivatives with respect to time of the anode and modulator voltage, respectively, and \( \dot{I}_m \) is the time derivative of the modulator current. Equation 4.6 can be simplified to give

\[ \dot{V}_P = -\dot{I}_m Z. \]  

(4.7)

During normal operation of a grid-controlled electron beam gun, the grid pulse is applied near the "top" of the pulse at \( \dot{V}_m \approx 0 \). Differentiation of Eq. 4.3, and substitution for the derivative of the modulator current yields
\[ \dot{V}_p = -Z (\dot{I}_p + \dot{I}_c). \] (4.8)

Substitution of Eq. 4.8 into Eq. 4.2 allows the displacement current \( I_c \) to be expressed in terms of the time-varying anode current, \( \dot{I}_p \), and \( \dot{I}_c \), the time-varying current through \( C_{ga} \), and the time-varying grid voltage \( \dot{V}_g \). Thus the displacement current \( I_c \) can be written as

\[ I_c = -C_{ga} (Z \dot{I}_p + Z \dot{I}_c + \dot{V}_g). \] (4.9)

Equation 4.9 defines the displacement current through \( C_{ga} \) as a function of the time-varying anode current, the time-varying displacement current through \( C_{ga} \) and the time-varying grid voltage \( \dot{V}_g \) [7]. However, further manipulation of Eq. 4.9 is required before the Miller capacitance model can be derived in a form which can be "coded" into a circuit simulation.

Equation 4.9 can be manipulated using the definition of the anode current or \( I_p = aV_g^{3/2} \) (Eq. 4.4). Differentiation of Eq. 4.4 with some simple algebra gives

\[ \dot{I}_p = \left( \frac{3}{2} \frac{I_p}{V_g} \right) \dot{V}_g \] (4.10)

Substitution of \( I_g = -I_c \) and Eq. 4.10 into Eq. 4.9 gives
Inspection of Eq. 4.11 allows the Miller effect to be written using the Thevenin circuit parameters. The first term of Eq. 4.11 describes the voltage-dependent grid current as a function of the grid voltage, anode current, and modulator impedance. By use of similarity arguments, the first term can be shown in the form

\[ I_g = C_{\text{eff}} \dot{V}_g \]  

(4.12)

where \( C_{\text{eff}} \) is the voltage-dependent Miller capacitance. Thus the Miller capacitance from Eq. 4.11 can be written as

\[ C_{\text{eff}} = C_{\text{ga}} \frac{3 Z I_p}{2 V_g} + 1 \]  

(4.13)

where \( Z \) (ohms), \( I_p \) (amps), and \( V_g \) (voltage) are as previously defined. The second term of Eq. 4.11 describes the resistive-capacitive decay of the grid-to-anode capacitance [7]. Because the second term is small compared to the first term, it can be neglected.

Equation 4.13 describes the Miller effect or Miller capacitance in terms of the circuit parameters. However, before Eq. 4.13 can be implemented it must be rewritten. At the initialization of a circuit program or \( t = 0^+ \), Eq. 4.13 has a mathematical singularity when \( V_g = 0 \). However, this singularity is removed by
combining Eq. 4.4 and Eq. 4.13 to give
\[ C_{\text{eff}} = C_{\text{ga}} \left[ \frac{3Z \alpha v_g^{1/2}}{2} + 1 \right] \] (4.14)

Although Eq. 4.14 cannot be directly entered into SPICE, a simple polynomial expansion of Eq. 4.15 allows the equation to be incorporated into the circuit simulation code.

Concurrent with Eq. 4.14, the space-charge characteristics of the grid-cathode region must also be incorporated into the circuit simulation model. The model developed to evaluate the Miller capacitance neglects the current flow in the grid-cathode region. To first order, the impedance of the grid-cathode region is

\[ Z_{\text{gk}} = \frac{V_{\text{gk}}}{I_{\text{gk}}} \] (4.15)

Substitution of the Child-Langmuir equation
\[ I_{\text{gk}} = a_{\text{gk}} v_g^{3/2} \] (4.16)
into Eq. 4.15 gives

\[ Z_{\text{gk}} = \frac{1}{a_{\text{gk}} V_{\text{gk}}} v_g^{1/2} \] (4.17)

where \( Z_{\text{gk}} \) is the voltage-dependent impedance of the grid-cathode region, and \( a_{\text{gk}} \) the perveance of the grid-cathode region. The perveance \( a_{\text{gk}} \) is approximately 0.08-0.1 for the grid-cathode region. This includes
the transparency of the grids.

Utilizing Eqs. 4.14 and 4.17, the model shown in Fig. 4.2 was constructed to allow the grid pulser response to the grid-cathode load characteristics to be determined.

Section 4.3 presents the circuit simulation results of the grid-pulser response to the grid-cathode model. As part of the investigation, the results of Section 4.3 are compared with the grid-pulser operation into the grid-cathode load, and an analysis of the results is presented.

4.3 Simulation of the Grid Pulser into the Grid-Cathode Load

Integral to the performance of an electron beam system, the grid pulser parameters determine the risetime of the electron beam, the peak current, the flatness of the current pulse, the pulse-width, and finally the fall-time of the beam. As part of the investigation, a grid pulser having a 2 μs risetime, a nominal 45 μs pulse-width, and a 2-4 μs fall-time was required. Additionally, a grid pulse with a ± 6% flatness specification and a controllable pulse-width were required to meet the constant current load parameters imposed by the modulator circuit simulations and the electron beam parameters (Table 1.1). A block diagram of the grid pulser is shown in Fig. 4.3. The grid pulser is based on a fifteen-section pulse-
Figure 4.2 Block diagram of the grid pulser and the constant perveance load into which it operates.
Figure 4.3 Block diagram of the grid pulser including the charging system, control system, and 15-section PFN.
forming-network of fixed impedance which is transformer-coupled to the grid-cathode load. The pulse-forming-network is switched by a high voltage, high current SCR. Moreover, the pulse-width is controlled by a crowbar SCR. The crowbar SCR allows the PFN to operate into the $4.4 \, \Omega - 6 \, \Omega$ impedance presented by the grid-cathode region under mismatched load conditions. Residual voltage left on the PFN by the mismatched impedance is effectively discharged, thus allowing the forward conducting SCR to recover during the interpulse period.

Design and analysis of the grid pulser topology was based on a trade-off study. The study reviewed the size constraints of the grid pulser (35 cm x 45 cm x 30 cm), as well as the life specification requirement of $10^8$ pulses. Based upon these considerations a pulse-forming-network impedance matched to the grid-cathode load by a pulse transformer was selected. Because of severe size limitations, the design of the grid pulser utilized an SCR as the switch. From a lifetime standpoint, a thyatron switch was quite attractive, however the ancillary support systems were too bulky. After review of available SCR's, the voltage of the PFN was set at 800 V. Moreover, the peak current rating of commercially available, fast risetime SCR's was limited to 400 A. Within these constraints, a $1.4 \, \Omega$ pulse-forming-network impedance matched by the 1:2 step-up transformer to the grid-cathode load was selected. In addition to matching the impedance of the pulse-
forming-network to the space-charge-limited load, the transformer reduces the transients coupled into the grid pulser during fault-mode conditions. Inherent to electron beam guns are fast transients generated by arcing in the electron beam gun. The resultant transient is coupled into the grid pulser by the coaxial cables linking the modulator and grid pulse to the electron gun. Transformer coupling of the pulse forming network and the SCR switch substantially reduce the transients coupled into the grid pulser thus increasing the life of the solid state switches in this application. Additionally, a negative bias is required to prevent premature turn-on of the electron beam during the risetime of the modulator. Transformer coupling allows the bias voltage to be capacitively coupled through the transformer secondary to the grid.

Analysis and design of the pulseforming network is complicated by the non-linear voltage dependent load characteristics and the associated Miller capacitance. Both the risetime and the pulse flatness interact with the load characteristics. To first order, the risetime of the applied grid pulse is a function only of the risetime of the network section nearest the load. Consistent with the 2 μs risetime requirements, as well as the nominal 45 μs "flat-top" specification, a 49.5 μs (equivalent square-wave), 1.4 Ω, fifteen section pulse forming network was chosen as the initial design. Utilizing Eqs. 3.11 and 3.13, a first approximation of the capacitance and inductance for the
pulse-forming-network was calculated; the total capacitance and inductance was 17.7 μF and 34.6 μH, respectively.

Because of the non-linear load characteristics, a detailed circuit simulation of the pulse-forming-network response into the grid-cathode load is required. As a first approximation, the pulse forming network is constructed as a fixed impedance network. That is, the total capacitance and inductance of the network is divided equally among the sections (Fig. 4.4). To reduce the complexity of the computer circuit calculations, the pulse forming elements are referenced to the secondary of the 1:2 step-up transformer. The capacitance is, of course, scaled down by a factor of four and the inductance of each section scaled up by a factor of four. The model of the grid-cathode load which was derived in Section 4.2 is used to simulate the Miller capacitance and the voltage dependent grid impedance. As shown in Fig. 4.4, the circuit simulation consists of many elements and is quite complex, interactively.

As part of the investigation, an iterative approach was utilized similar to the one adopted for tuning the modulator PFN to its constant current load. Iterative calculations over all four electron beam operating ranges were required to tune the PFN. The resultant pulse forming network is shown in Fig. 4.5, along with the load model used in the calculation. For clarity, the pulse forming network capacitance and
Figure 4.4 Initial fixed impedance PFN used for the computer calculations. All values are on the secondary of the 1:2 step-up transformer.

\[ C_1 = 0.299 \, \mu F \]
\[ C_2 - C_{15} = 0.299 \, \mu F \]

\[ L_1 = 9.39 \, \mu H \]
\[ L_2 = 9.39 \, \mu H \]
\[ L_3 = 9.39 \, \mu H \]
\[ L_4 = 9.39 \, \mu H \]
\[ L_5 = 9.39 \, \mu H \]
\[ L_6 = 9.39 \, \mu H \]
\[ L_7 = 9.39 \, \mu H \]
\[ L_8 = 9.39 \, \mu H \]
\[ L_9 = 9.39 \, \mu H \]
\[ L_{10} = 9.39 \, \mu H \]
\[ L_{11} = 9.39 \, \mu H \]
\[ L_{12} = 9.39 \, \mu H \]
\[ L_{13} = 9.39 \, \mu H \]
\[ L_{14} = 9.39 \, \mu H \]
\[ L_{15} = 6.4 \, \mu H \]

(does not include 3.4 \, \mu H cable inductance)
Figure 4.5 Final tuned grid pulser PFN values from the circuit simulations. All values are on the output of the 1:2 step-up transformer.
inductance are shown on the secondary of the pulse transformer. The primary PFN values can, of course, be found by scaling the capacitance and inductance values in Fig. 4.5 by a factor of four. Figure 4.6 shows the final grid-cathode waveform (circuit code is listed in Appendix D). The final simulated waveform was flat to ±6% with a pulse-width of 45.5 μs. The waveform overshoot and front oscillation is a result of the non-linear load characteristics. The oscillation amplitude, however, changes over the operating range.

Because the Miller capacitance is a function of both the modulator output impedance and the grid-cathode voltage (Eq. 4.14), the voltage dependent capacitance varies over the electron beam operating range, thus changing the risetime and overshoot on the grid-cathode pulse. As an example of the load capacitance variation, the calculated $C_{eff}$ for the highest current range is

$$C_{eff} = [26 \times V_{gk}^{1/2} + 1] \times 90 \text{ pF}$$

or 59.7 nF. Comparatively, the calculated $C_{eff}$ for the lowest current range is

$$C_{eff} = [17 \times V_{gk}^{1/2} + 1] \times 90 \text{ pF}$$

In both cases the Miller capacitance is more than ten-fold higher than the 1.7 nF combined cable and electron gun capacitance. When the Miller capacitance is
Figure 4.6 Simulated grid-pulser PFN waveform across the constant preveance, grid-cathode load.
included with the 1.7 nF cable capacitance, the grid-pulser must "drive" 61.4 nF. Moreover, comparison of $C_{eff}$ for the two operating ranges shows that the grid-cathode load capacitance varies by a factor of roughly 2.5. Inherently, this changes the PFN risetime and overshoot over the entire electron gun operating range. Concurrently, the amplitude of the overshoot varies, and oscillation on the front of the grid-cathode waveform is dependent upon the electron beam operating range.

Intuitively, the complex interaction of the pulse-forming-network and the grid-cathode load can be understood by examining the physics of the circuit. During the risetime of the grid pulse, the pulse-forming-network interacts with the Miller capacitance. Review of the equivalent circuit of Fig. 4.7 shows that the first PFN section capacitance is much larger than the equivalent grid capacitance. When the grid-pulser PFN is triggered, the voltage across the Miller capacitance "rings" up as a result of the C-L-C circuit formed. The impedance of the grid-cathode region does however tend to damp out the oscillations. As the grid-cathode voltage increases across the grid, the voltage-dependent impedance of the region decreases, and thus helps to damp out circuit oscillations. As a consequence, the Miller capacitance charge is drained-off as the grid-cathode load decreases in impedance. However, because of the initial charge stored in the Miller capacitance, some oscillation on the front-end
A. Output equivalent circuit of the grid pulser, and the grid-cathode load.

B. Equivalent circuit showing the dependence of risetime upon the PFN section closest to the grid. All values are on the secondary of the grid pulse.

Figure 4.7 Grid-pulser and constant perveance grid cathode load.
of the pulse persists until the Miller capacitance is fully discharged. A close examination of circuit operating parameters shows that, indeed, this is the case. As an example, we consider the regime of the highest electron beam currents. The grid pulser operates at a voltage of 650 V. The subsequent Miller capacitance is 59.7 nF from Eq. 4.18. When the Miller capacitance is considered in conjunction with the first section inductance, the impedance of the circuit is 12.67 Ω from the \(\sqrt{L/C}\) value. However, from Eq. 4.17, the grid impedance is 3.92, or a factor of 3.2 lower than the circuit impedance. Thus, the grid-cathode impedance is low enough to damp out oscillation during the applied grid-cathode pulse. However, during the initial risetime of the grid-cathode pulse, the impedance of the region is quite high. Thus, the pulse tends to overshoot slightly, while the grid-cathode impedance decreases.

A similar circuit condition exists throughout the operating range of the grid-pulser. As an example, consider the 5 mA/cm\(^2\) or lowest range of the electron beam. In this range the grid pulser operates at about 250 V. The consequent Miller capacitance is 24.2 nF. When the impedance of the circuit is calculated using the equivalent output PFN inductance of 9.8 μH, the initial circuit impedance is approximately 20.1 Ω. The calculated parallel impedance of the grid-cathode region is 6.3 Ω, from Eq. 4.17. Uniquely, the damping ratio stays constant. This results as the Miller
capacitance decreases at the lower operating range, thus decreasing the oscillation seen on the simulated grid-cathode voltage waveform.

4.4 Grid Pulser PFN Layout and Testing

The small physical size of the grid pulser necessitates careful layout and design of the grid pulser PFN. As a consequence of the low (1.4 Ω) PFN impedance, a folded PFN design was selected to minimize the stray inductance inherent in the PFN layout. Moreover, a folded PFN design, that is a PFN which is physically folded in geometry, was required to accommodate the small physical size of the compact grid pulser design. Consistent with these design requirements, a PFN design was investigated based upon spiral inductors. Classically, PFN design and layout has relied upon solenoidal construction of the PFN inductors as described in Chapter 3.0. However, a preliminary design calculation using Nagoaka’s formulae showed that a physical length of 30 cm was required to accommodate one of the PFN section inductances required by the PFN [8]. This length, as well as the large number of inductors needed, presupposed the use of solenoidal inductor design. As a result of the spiral inductor design and the folded PFN design which makes calculation of the mutual inductances difficult, experimental investigation of the final PFN design was emphasized.
The folded PFN is based on a "split" design. Seven sections of the PFN are accessible from the bottom of the grid pulser and the remaining eight sections accessible from the top. Figures 4.8A and 4.8B are photographs of the grid-pulser and PFN. An "open" construction of the PFN allows easy removal for tuning of the inductor and capacitor values. Although not apparent in Fig. 4.8, the PFN is built on a common plastic frame. Common tie rails are insulated from ground and easily removed with the entire PFN assembly.

High density reconstituted mica capacitors rated at 1200 V were used for the PFN. Each capacitor was measured with a Z-meter. Maximum variation in capacitance was found to be 5%. The capacitance per section was 1.24 \( \mu \)F (cf. the calculated value of 1.18 \( \mu \)F, Section 4.3, Fig. 4.5), and the capacitance of the last section was 1.7 \( \mu \)F (cf. 1.6 \( \mu \)F, Fig. 4.5), measured and calculated values differ by only 6%.

A spiral inductor geometry was adopted for the PFN inductor. To optimize the inductor design, Eq. 4.20 was used:

\[
L = 40 N^2 r^2 K \text{ (nH),} \quad (4.20)
\]

where we have

\[
k = \frac{r}{r + \Delta r(2/3) + w}
\]
A. Side view of Grid-Pulser

B. Bottom view of the grid-pulser showing view of the removable PFN.

Figure 4.8 Photograph of grid-pulser including a side-view of the PFN.
and $N$ is the number of turns, $r$ the average radius of the inductor, $\Delta r$ the build-height, and $w$ the width of the inductor [9]. The simulation, calculated and final inductance values are shown in Table 4.1. The inductance values shown in Table 4.1 assume a width of 1.27 cm. The build-height assumes that the inductors are constructed with copper ribbon of $1.27 \times 10^{-2}$ cm thickness, insulated on both sides with kapton tape of thickness $8.89 \times 10^{-3}$ cm. To trim the inductance the form radius was shimmed.

During the inductor construction, it was found that the theoretical inductance varied by 8% from the desired value (Table 4.1). Trimming of each individual inductance was difficult and prone to error. The individual inductance values were measured with a Z-meter having a resolution limited to $\pm 200$ nH. Inherently, the representative inductances were 8% below the original calculated value. Although the spiral inductor offers a compact design, accurate trimming of the inductance to the desired value is quite difficult. However, as a starting point, the grid-pulser PFN was constructed from the values shown in Table 4.1.

To tune the PFN, a 4 $\Omega$, non-inductive load was placed across the output of the impedance matching transformer. Although the charge voltage on the PFN was varied throughout the tests, a minimum of 300 volts was used to check pulse fidelity through the iron-core pulse transformer. The pulse shape was also checked at
Table 4.1
PFN inductance values including simulation, calculated and measured values. All values refer to the primary of the grid-pulser.

<table>
<thead>
<tr>
<th>Requirement</th>
<th>Simulation Value</th>
<th>Calculated Value</th>
<th>Measured Value*</th>
</tr>
</thead>
<tbody>
<tr>
<td>L₁ = 1.96 μH</td>
<td>L₁ = 1.86 μH</td>
<td>L₁ = 1.8 μH</td>
<td></td>
</tr>
<tr>
<td>L₂ = 2.05 μH</td>
<td>L₂ = 2.05 μH</td>
<td>L₂ = 2.1 μH</td>
<td></td>
</tr>
<tr>
<td>L₃ = 2.14 μH</td>
<td>L₃ = 2.21 μH</td>
<td>L₃ = 2.2 μH</td>
<td></td>
</tr>
<tr>
<td>L₄ = 2.25 μH</td>
<td>L₄ = 2.21 μH</td>
<td>L₄ = 2.3 μH</td>
<td></td>
</tr>
<tr>
<td>L₅ = 2.32 μH</td>
<td>L₅ = 2.21 μH</td>
<td>L₅ = 2.3 μH</td>
<td></td>
</tr>
<tr>
<td>L₆ = 2.45 μH</td>
<td>L₆ = 2.35 μH</td>
<td>L₆ = 2.5 μH</td>
<td></td>
</tr>
<tr>
<td>L₇ = 2.45 μH</td>
<td>L₇ = 2.8 μH</td>
<td>L₇ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₈ = 2.45 μH</td>
<td>L₈ = 2.8 μH</td>
<td>L₈ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₉ = 2.45 μH</td>
<td>L₉ = 2.8 μH</td>
<td>L₉ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₁₀ = 2.45 μH</td>
<td>L₁₀ = 2.8 μH</td>
<td>L₁₀ = 2.5 μH</td>
<td></td>
</tr>
<tr>
<td>L₁₁ = 2.45 μH</td>
<td>L₁₁ = 2.8 μH</td>
<td>L₁₁ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₁₂ = 2.45 μH</td>
<td>L₁₂ = 2.8 μH</td>
<td>L₁₂ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₁₃ = 2.45 μH</td>
<td>L₁₃ = 2.8 μH</td>
<td>L₁₃ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₁₄ = 2.45 μH</td>
<td>L₁₄ = 2.8 μH</td>
<td>L₁₄ = 2.6 μH</td>
<td></td>
</tr>
<tr>
<td>L₁₅ = 2.45 μH</td>
<td>L₁₅ = 2.8 μH</td>
<td>L₁₅ = 2.6 μH</td>
<td></td>
</tr>
</tbody>
</table>

*Includes leakage of transformer and correction for cable inductance.
both the minimum and maximum pulse voltages to ascertain the transformer pulse fidelity.

The initial test waveform into the resistive 4 Ω load is shown in Fig. 4.9. Two relevant characteristics are obvious after review of Fig. 4.9. The most obvious is the impedance variation along the length of the PFN. As the impedance of each section varies, the voltage across the fixed resistance load changes. Figure 4.9 shows this effect clearly. After the pulse risetime the voltage slowly increases. The test voltage waveform levels out and then drops abruptly at approximately 21 μs. Although the waveform voltage contribution by each individual section can be identified by the ripple on the waveform, the quantitative variation of each section impedance is difficult to calculate analytically. As the PFN discharges into the load, internal voltage reflections occur along the length of the PFN. As a result, forward and backward travelling waves are present in the PFN. These voltage reflections are a result of the abrupt impedance changes, section to section.

The second most prominent characteristic of the load voltage waveform is the voltage "dip" (Fig. 4.9). Because of the folded PFN design, an abrupt impedance mismatch occurs between Section 7 and Section 8 of the PFN. Measurements done during the tuning process have shown that this impedance mismatch results from the inductive connection between Section 7 and Section 8. Also in pulse-forming-networks of low impedance,
Figure 4.9 Photograph of initial grid-pulser PFN being discharged into a 4 Ω load; 10 μs/div, 50 V/div.
relatively small inductances introduced by the long lengths of lead significantly influence the pulse shape. Long lead lengths can easily introduce stray inductances of the same order as the PFN inductance itself, and can thus significantly alter the output waveform.

In order to simplify the PFN tuning procedure, the PFN response into a resistive load was simulated using SPICE. The voltage-dependent load of Figure 4.5 was replaced with a 4 Ω resistor.

The resultant waveform is shown in Figure 4.10. Because of the impedance tapering required to tune the PFN to the desired ± 6% flatness specification into the constant perveance grid load, the pulse has a distinctive slope when working into a resistive load. That is, the top of the pulse has a gradual negative slope as a result of its impedance tapering (long pulses are notorious for this, anyway).

To match the actual PFN characteristics to those of the simulation (Fig. 4.10), an experimental approach was chosen. During the investigation it was found that small adjustment of the individual section capacitance or inductance would change the PFN waveform. Tuning was also complicated by the internal voltage reflections present in the PFN. Trimming of the section capacitances and inductance would change the waveform characteristics slightly. The folded PFN design also contributed to this problem. As a result of the abrupt connection between Sections 7 and 8,
Figure 4.10 Computer simulation of grid-pulser PFN response into 4 Ω load (solid line) and the actual tuned response of the PFN into the 4 Ω load (dashed line).
standing waves were present along the length of the PFN. The reflected voltage waves add in the two discrete halves of the PFN, as demonstrated by the two peaks in Fig. 4.9. Of course, the consequential internal reflection make systematic identification of individual section impedance changes impossible. To circumvent this problem, an active tuning procedure was adopted. A 0.1 μF capacitor was mounted on an insulated handle. While operating at 10 pps, the capacitor was then inserted across the PFN section under test. During testing the waveform was displayed on an oscilloscope and the PFN waveform monitored. The section which required trimming could be identified easily by a small change in the ripple or oscillation present on the top of the waveform. Once identified, the section impedance was trimmed either inductively or capacitively.

The PFN was systematically tuned from front to back. Once the "problem" section was identified, the inductance or capacitance could easily be trimmed with small hollow ferrite slugs or discrete 0.1 μF capacitors. During the investigation it was found that an iterative tuning procedure was necessary, due to the interaction between sections. If the final PFN waveform (dashed line of Fig. 4.10) is compared with the circuit simulation (solid line), the two waveforms agree nicely. Both 10-90% risetimes are 2 μS. However, as shown, the experimental waveform displays much less overshoot than the simulation. If the actual
waveform of Fig. 4.11 is reviewed, we see that this is the case. It was found that the transformer coupling reduced the high frequency overshoot and oscillation. At high frequencies, the iron-core of the coupling transformer becomes lossy, thus damping out unwanted overshoot [10]. Because the circuit simulation did not include a detailed model for the transformer, the experimental waveform was found to be better than the circuit simulations indicated. However, as noted in Section 1.3, the prime purpose of the simulation studies was to guide the practical design features so that the performance specifications could be attained or bettered. The excellent performance achieved clearly demonstrated the success of this strategy.
A. Final grid-pulser response into a 4 Ω load
50 kV/div; 10 μs/div.

B. Expanded view of grid-pulser waveform into
a 4 Ω load; 5 μs/div, 50 V/div.

Figure 4.11 Photograph of final measured grid-pulser waveform.
REFERENCES
CHAPTER 4


9. J. Fockler and Ian Smith, private communications.

5.1 Experimental Approach Background

An experimental approach similar to that adopted for the grid pulser test was selected for the modulator-electron beam system investigation. The modulator response was first characterized into a resistive load using the matching resistors integral to the modulator (Fig. 3.8). Utilizing the experimental voltage and current measurements, the waveforms were compared with the modulator circuit simulations. The results of the investigation and the analysis are presented in Section 5.2.

Full voltage and current tests of the integrated modulator-electron beam system were used to experimentally verify the system models. Upon integration of the electron beam system, voltage, current, and Faraday scanning cup diagnostics were employed to "check-out" the system. Concurrently, the investigation included diagnosis of the foil window response to the high energy pulse loading present during continuous operation at 10 pps. The response of both aluminum and titanium foils are reviewed. Both the operational characteristics and the system analysis results are presented in Sections 5.3 and 5.4.
5.2 Modulator Characterization into a Resistive Load

The verification tests to "check-out" the modulator prior to integration with the electron beam system were divided into two phases. The first phase was tailored to "tuning" up the modulator on a single shot basis to prevent damage to major modulator components. Attention was focused on the currents and voltages delivered by the pulse-forming-network while working into a resistive load.

The second phase of the modulator test program was designed to investigate the modulator performance at full average power. Subsequently, the modulator's reliability at full voltage and current was determined. The performance of the major modulator components under full peak power was also investigated. Both the thyratron operation and the high voltage resistor operation were reviewed. Predominant failure mechanisms studied included voltage flashover of the high voltage resistors, oil insulation integrity of the high voltage step-up transformer, and thyratron performance under continuous operation.

5.2.1 Phase 1: Experimental Modulator Check-Out

As part of the Phase 1 experimental tests, the high voltage matching resistors were calibrated with the 250 kV voltage probe integral to the modulator and a commercially available current probe, calibrated to ± 2%. Voltage and current measurements of the modulator output into the
matching resistors were verified under single-shot conditions in the 40-100 kV range. During calibration of the resistor stacks the modulator’s voltage was restricted to this operating range to prevent component damage from excessive reversal.

Analysis of the matching resistor voltage and current calibration data showed a strong variation of resistance with voltage. This voltage coefficient of resistance was not surprising for it is well known that carbon-ceramic resistors have a strong voltage coefficient when the electric field exceeds several kV/cm. The voltage coefficient was found to be linear in the 40-100 kV range. On the basis of the experimental data, an additional number of resistors were added to the matching resistor stacks.

The modulator voltage was increased and the resistor response in the 175 kV - 200 kV voltage range checked. When the average electric fields across the resistor stacks were in the range 3.6 - 4.1 kV/cm, the resistance was found to be unstable in concurrent tests. This was a rather unexpected result. Test data from the manufacturer showed a 7.9 kV/cm to 10 kV/cm breakdown field in oil. An examination of the resistor stack’s surface did not reveal tracking, even though high energy modulator discharges had occurred.

After examination of the matching resistor geometry, the unstable resistance was attributed to an unforeseen geometrical problem. The resistor stacks were mounted vertically to insure a good current contact. Compressible, bronze wire screen was used between the resistors. The
vertical geometry of the resistor stacks, along with the propensity of the screen to trap air, was found to be the problem. Upon raising and lowering the modulator into the oil bath, air was trapped between the resistors. Once lowered into oil, the air escaped, clinging to the surface of the resistors. The volume of air trapped between the resistors is easily calculated. If the screen thickness of 0.076 cm is considered in conjunction with the 2.54 cm inside diameter and 11.1 cm outside diameter of the screen, the volume of trapped air is 1.74 cm$^3$, when a geometrical screen transparency of 25% is used for the calculation. Due to the complex geometry it is difficult to accurately calculate the actual diameter of the escaping air bubbles. If an average diameter of 0.1 cm is assumed for the escaping gas bubbles then approximately 3,323 bubbles will escape from one of the resistor screen interfaces before the interface is fully outgassed. The above is, of course, a worst-case calculation.

Breakdown as a result of gaseous inclusions in an insulating fluid has been reported previously [1,2]. Both Alston and Kao have developed theoretical formulae for the observed breakdown fields in liquid dielectrics. It is interesting to note that the inception of the breakdown process is attributed to the electric field in the gaseous inclusion or gas bubble exceeding the breakdown field of the gas. Alston derived the following formula for the electric field $E_b$
as a function of the liquid's dielectric constant $\varepsilon_1$ and the applied electric field $E_o$. Assuming that the average field $E_o$ across the resistors is 4.1 kV/cm, and that the dielectric constant of the transformer oil is 2.5, Eq. 5.1 gives 5.2 kV/cm for the electric field within the gas bubble. It is rather interesting to note that the observed instability of the resistance cannot be solely attributed to breakdown of one gas bubble. The calculated field of 5.12 kV/cm is well below the 25-30 kV/cm breakdown field of air [3], but as shown below, a field enhancement factor of about 30 can be expected.

Two alternate explanations have been offered for the unstable matching resistor characteristics found at 175 kV - 200 kV. The first is based on the dielectric mismatch which occurs at the surface of the resistors if air is present in oil. The dielectric constant of air is one, that of oil is about 2.5, and the dielectric constant of the carbon-ceramic resistor matrix is in the range of 4-8. Thus, a small air bubble clinging to the resistor surface can create a very localized field enhancement. If a bubble of diameter 0.1 cm is assumed, the field enhancement is on the order of 30. The ensuing streamer formation in oil results in flashover of the carbon-ceramic-oil interface. Thus, one or two resistors in a resistor stack could "flashover" without causing a breakdown of the entire stack.
The second explanation is based on the calculated volume of trapped air. If the air slowly escaped from the sixty interfaces found in the matching resistor stacks, consecutively generating a stream of bubbles along any one resistor interface, the stream of bubbles could be aligned by electrostatic forces and so cause a breakdown channel to develop. Again, it is likely that flashover could occur erratically at one or two resistors.

Once identified, the resistor flashover problem was easily eliminated. Each resistor in the resistor stack included a 2.54 cm hole. Prior to lowering the modulator into oil, a hand-pump was used to force oil down the center of the individual matching resistor stacks. The bottom resistor was plugged, forcing a high pressure oil stream into the space between resistors. Further tests at full 200 kV - 250 kV operating voltage showed stable resistance characteristics. The parallel combination of all three resistor stacks were found to be 1700 Ω, or 15% lower than the calculated value of 1951 Ω at 220 kV. The voltage coefficient of the resistors was found to be 15-20% at an average electric field of 4.1 kV/cm.

To complete the Phase 1 single-shot modulator tests, an analysis of the modulator PFN response into a resistive load was verified using circuit simulations. The measured response of the modulator into the 1700 Ω matching resistor load is shown in Fig. 5.1. The measured risetime of the 204 kV pulse was 5.74 μs and the measured 90-90 pulse-width was 46 μs. The required pulse-width was 45 μs. The similarity in waveshape of the normalized circuit
Figure 5.1 Measured response of modulator into the matching resistive load; 10 μs/div, 40 kV/div.
simulation, Fig. 5.2, shows a close correspondence between the simulated and measured waveforms. However, we note that a significant difference does occur late in the pulse, particularly at the falling edge.

The experimental data indicates that the modulator voltage actually increases during the pulse. This is contrary to the circuit simulation results which indicate a pulse droop rather than an increase in the voltage. The increasing voltage characteristic is easily explained by one physical factor which was not included in the circuit simulation. In the physical layout of the PFN, mutual inductance is present. The mutual inductance coupling the adjacent PFN section inductors reduce the equivalent series inductance of each section capacitor. It has been shown by Guillemin that the mutual inductance between coils is equivalent to a negative inductance modeled in series with the equivalent series inductance of each PFN section capacitor [4]. Thus, less voltage is seen to be dropped across the stray section inductances (so increasing the voltage across the output of the PFN) when the mutual inductance is considered.

From the Phase 1 experimental program, it was determined that the modulator voltage waveform did indeed meet the pulse-width and pulse flatness specification. The increasing voltage waveform characteristics are an advantage if a constant current load is to be driven. Thus, no tuning of the modulator PFN was required.
Figure 5.2 Comparative response of the modulator output into a resistive load and the circuit simulation of the modulator response into a resistive load (normalized voltage). The circuit simulation is shown by the solid line. The measured modulator response into a resistive load is shown by the dashed curve.
5.2.2 Phase II: Full Power Modulator Tests and Thyatron Tests

One apparent trend did become obvious during continuous modulator operation. The thyatron performance was dependent on the triggering scheme, the heater and reservoir settings, and the anode and cathode cooling. The unusually long pulse (52 μs) along with the 30 kV, 3 kA circuit parameters distinctly influenced the thyatron behavior. A conventional single grid triggering scheme was tried, but the thyatron performance was erratic. A dual trigger scheme was established to stabilize the thyatron discharge. A high current 300 V, 5 A, 1 μs pulse was used to preionize grid #1 of the CX1574C hollow anode thyatron. Following the low voltage, high current preionization pulse, a 1 kV, 1 μs, 20 A pulse is applied to grid #2. The low voltage grid #1 pulse provides "seed" electrons to stabilize the thyatron discharge. The delay between the preionization pulse and the main 1 kV trigger pulse was determined to be critical for reliable triggering of the thyatron. A 1.5 - 1.8 μs delay of the grid #2 pulse with respect to the preionization was found to both reduce the thyatron jitter and eliminate all prefires. Exact delay times varied with the three thyatrons tested, and required fine tuning to the individual thyatron. The heater and reservoir voltage settings were also found to be critical. During continuous 10 pps operation, the heater and reservoir voltage settings were optimized to 6.1 V and 4.0 V, respectively. At higher heater and reservoir settings,
prefires and high jitter operation was noted. The long 52 μs discharge was found to result in anode and cathode heating. Directed oil flow across both the anode and cathode was found to stabilize the thyratron operation. The unstable thyratron performance is attributed to the gas dynamics of a thyratron. Under high-repetition rates or high energy discharge conditions it has been shown that pressure gradients result from anode heating [5]. The reported gas gradients destabilize the thyratron discharge. Additionally, it is well known that continuous operation of a thyratron results in elevated heating at the cathode [6]. Forced cooling of the cathode removes this additional heat component, and stabilizes the thyratron performance.

5.3 Electron Beam System Investigation

Following the modulator "check-out" a four-part program was established to integrate the electron beam system. The four-part program included the following: (1) the integration of the modulator and grid pulser with the electron-beam gun, (2) Lifetime testing of the vacuum/foil interface of the laser operating at full power. (3) Verification of the electron beam system over the required 5-20 mA/cm² (12.5-50 A into the gas) current range to establish electron beam adjustability, and (4) Characterization of the electron beam spatial uniformity. The following discussion highlights the key aspects of each test phase of the program showing the inherent physics
issues relevant to the successful demonstration of an electron beam system.

Prior to application of high voltage, characterization of the grid pulser into the space-charge limited grid-cathode load was required. Although the grid pulser response into a resistive load verified the circuit simulation model, a space-charge limited load is non-linear with voltage. Inherently the PFN response can change substantially into a non-linear load. Utilizing the fiber optically coupled voltage and current monitors integral to the grid pulser, operation of the grid pulser into the electron beam gun load was verified. As shown in Fig. 5.3, the grid pulser waveform has minimal overshoot even at the grid pulser. Concurrently, it was found that the grid waveform was flat to within 4.8% (if the overshoot is not considered), well within the required 6% grid pulse specification. Inherently, as shown in Fig. 5.3, some overshoot is present on the waveform. Experimentally, it was verified that the overshoot at the grid-cathode load is reduced substantially. The cable inductance linking the grid-pulser to the electron beam gun and the Miller capacitance of the grid-cathode region reduce the overshoot by filtering the high frequency components. Based on the initial grid-cathode waveforms, no tuning of the grid pulser was required.

As part of the investigation, the modulator response into the electron beam load was required. Initial single shot, hot cathode electron beam tests indicated both voltage-dependent and time-dependent vacuum breakdown.
Figure 5.3 The measured grid-pulser response into the space-charge limited load represented by the grid-cathode load. No acceleration voltage or electron beam is present.

Top Trace: Voltage across the grid-cathode load 360 V/div, 10 µs/div.

Bottom Trace: Current into the grid-cathode load 40 A/div, 10 µs/div.
Vacuum breakdown was evidenced by rapid collapse of the cathode-anode acceleration voltage and increasing cathode current. The rapid collapse of the cathode-anode voltage was similar to that found in vacuum gap breakdown experiments [7]. However, exact breakdown mechanisms were difficult to identify. Both cathode-anode plasma closure and/or voltage flashover of the ceramic high voltage bushing could result in collapse of the cathode-anode voltage. Rather than damage the cathode by multiple high energy breakdowns, a cold cathode field emission conditioning procedure was adopted, for both the cathode and high voltage bushing condition process.

5.3.1 Cold Cathode Conditioning Procedure

To avoid breakdown at low voltage and further deconditioning of the cathode-anode gap, a low-energy cold cathode conditioning process was developed on the basis of the initial hot cathode-modulator tests and the observed conditioning processes reported for small vacuum gap research. That is, the initial 100 kV modulator-hot cathode tests indicated breakdown at cathode-anode fields of 13.3 kV/cm and high voltage insulator breakdown at 1.97 kV/cm. Even if field enhancements in the system are taken into account, the electrical field at which breakdown occurs is a factor of -10-100 below that reported previously [8,9]. Concurrently it appeared that the 1800 J PFN pulse deconditioned the vacuum system further, degrading the reliability of the electron beam
system by increasing the number of breakdowns. The deconditioning effect under high energy is a well known effect and has been previously reported by R. Hackman [10]. Based on these very preliminary results, a conditioning process was developed which limited the breakdown energy to about 10 J. The process which is discussed in the following section was developed on the basis of cold cathode emission experiments. To prevent damage to the thoriated tungsten filaments, the cathode and high voltage bushing were conditioned at ambient temperature. Thus, gas poisoning of the hot thoriated tungsten filaments was avoided.

Investigation of both small area vacuum gaps and large area gaps, as reported by Kuenning and Hackam, have shown that the optimum energy required to condition vacuum electrodes is about 1-10 J [10, 11]. Several mechanisms which are believed to be responsible include "cratering" and whisker growth which may lead to high field enhancements, reducing the voltage hold-off of a vacuum gap [12]. Thus, a low energy conditioning procedure, such as the use of a current-limited DC power supply, appeared to be more appropriate than using the modulator to condition the cathode.

On the basis of this observation, a DC conditioning procedure independent of the high energy modulator pulse was chosen. Initial calculations showed that the cathode-anode capacitance was approximately 320 pF. At 100 kV and 250 kV the integral cathode capacitance stores 1.6 J and 10 J, respectively. Based on these findings, a circuit was
chosen to charge the cathode-anode capacitance using a 1 mA, 250 kV, current limited power supply. The conditioning circuit did however require a current limiting resistor. The 250 kV DC voltage programmable power supply is connected to the high voltage oil bushing on the electron beam gun by a 300 kV coaxial cable. The 67 Ω cable is 10 m long and has an equivalent capacitance of 736 pF. The energy stored in the cable at 100 kV and 250 kV is 3.68 J and 23 J. To prevent the cable energy from discharging into the cathode-anode region during the conditioning process a current limiting resistor is required. Experimental investigations have shown that a 1-10 MΩ resistor is sufficient to isolate both the power supply and power supply cable from the discharge [10]. A 10 MΩ isolation resistor was selected on the basis of this data. However, the 10 MΩ resistor does not limit the current associated with charging the cathode-anode capacitance. The power supply output impedance is programmable using an internal current program mode and voltage mode. To control the charge time of the cathode-anode capacitance the current output of the power supply is programmable up to 1 mA.

To setup the system for the cold cathode conditioning of the 250 cm cathode, as well as the high voltage ceramic bushing, the cables linking the modulator to the electron beam system were removed. The DC power supply cable and limiting resistor were installed in place of the modulator cables.
As part of the conditioning procedure, the power supply voltage was raised to \(-100\) kV over a 10 minute interval. At \(-100\) kV, an of emission current of \(0.5\) mA was noted. During the conditioning process, the background pressure of the vacuum chamber rose from \(5 \times 10^{-7}\) to \(3 \times 10^{-6}\) torr as monitored by a cold-cathode ion gauge. Generation of soft x-rays and occasional voltage breakdown in the vacuum chamber was also observed. Over a four hour period the conditioning voltage was increased to \(-150\) kV. At \(-150\) kV, a \(1\) mA emission current was measured. At \(-150\) kV, it was also noted that \(1\) mA of emission current resulted in frequent vacuum breakdown occurrences. That is, voltage breakdown was noted every 10-15 seconds. It was found that reduction of the power supply current to \(0.6\) mA reduced the frequency of breakdown to once every 1-10 minutes. At a power supply current of \(0.6\) mA the power supply voltage was ramped up to \(-250\) kV over an 8 hour period.

Initial conditioning of a new cathode required a 12-16 hour period. After initial conditioning both the cathode and high voltage bushing could be exposed to atmosphere for a day or two without total loss of conditioning. Subsequent conditioning after exposure to atmosphere required only 1-2 hours of full 250 kV DC hold-off. A "reworked" cathode which had been removed from the vacuum chamber would condition after 7-8 hours. It has been hypothesized that the 7-8 hours conditioning procedure required for a reworked cathode resulted from minute scratches and/or the presence of dust particles due to handling of the cathode.
For a fully DC conditioned system it was observed that only 0.2-0.3 mA of cold cathode emission current was noted at 250 kV, for the ~ 2500 cm$^2$ cathode area.

5.3.2 Experimental Evaluation of the Modulator Electron Beam System

The system tests were resumed following the cold-cathode conditioning procedure. Initial pulse tests at -150 kV demonstrated both single shot and full 10 pps repetitive rate reliability. Based on these data points, single-shot full voltage tests were tried. At -220 kV no voltage breakdown of the electron-beam system was noted. Full power 10 pps tests at -220 kV and 20 mA/cm$^2$ (110 A) emission current levels demonstrated the required performance of both the modulator and the electron beam system (Table 1.1). The cathode voltage and current are shown in Fig. 5.3. The constant current load characteristics are illustrated by the lower trace of Fig. 5.3. Although the peak modulator voltage varies from 198 kV to 224 kV, the grid controlled 110 A cathode current remains constant. Thus, the modulator's load impedance varies during the pulse.

To demonstrate the load characteristics imposed by the constant current load, the normalized modulator response into a resistive load and into the constant current load are shown in Fig. 5.4. The constant current load tends to accentuate the pulse characteristics. Both voltage ripple and overshoot are more prominent in the constant current
Figure 5.4 Normalized modulator voltage waveforms. Solid trace is the modulator pulse into a resistive load. Dashed trace is the modulator pulse into a constant-current load.
load case. One additional characteristic is noted on the constant current load pulse. During the fall-time of the voltage pulse, the electron beam current is turned off by the grid-cathode pulse to prevent damage to the foil interface. This prevents the low energy electrons in the "tail" of the pulse from being deposited into the foil. When the cathode current is turned off, the load seen by the modulator is transformed from a predominantly constant current load to a resistive load. Once the cathode current is turned off, the modulator dissipates the remaining pulse energy into the modulator matching resistor. The transition from the constant current load to the resistive load tends to decrease the amount of droop on the pulse. This effect also increases the pulse-width slightly.

During the checkout phase of the electron beam system, the cathode and the high voltage vacuum insulator, voltage reliability was investigated. An interesting trend manifested itself while the system was operated at -200 kV. After DC high voltage conditioning of the cathode and vacuum bushing to -250 kV, the cathode and bushing would only stay conditioned for two weeks if no voltage was applied to the system. Further experiments revealed that the DC conditioning of the vacuum system degraded if DC voltage was not reapplied for one or two days after the main conditioning procedure. However, intermittent, pulsed cathode operation of the system required no prior or additional conditioning up to two weeks between system tests. During the tests a vacuum integrity of better than $5 \times 10^{-6}$ Torr was maintained using a cryogenic vacuum pump.
As part of the experimental program, the high voltage reliability of the pulsed cathode, and high voltage bushing was also investigated. Three one-hour full power tests at 224 kV, 10 pps rep-rate and 20 mA/cm² (110 A) cathode current demonstrated high voltage reliability. During three one-hour tests only four or five vacuum breakdowns were noted. For the system to operate reliably without vacuum breakdowns required a vacuum pressure of better than 5 x 10⁻⁶ Torr. During several one hour tests of the system, pressure rises from 5 x 10⁻⁷ to 5 x 10⁻⁶ torr were observed. Above 5 x 10⁻⁶ Torr frequent vacuum breakdown discharges occurred under repetitive conditions.

5.3.3 Applicability of the Cold Cathode Tests to the Hot Cathode Results

Analysis of the cold cathode conditioning procedure and its applicability to the pulsed, hot cathode system tests is complicated by a large number of parameters. For the cold cathode conditioning procedure, field emission and subsequent vacuum arcing has been hypothesized to be the predominant conditioning mechanism. However, the presence of the two high voltage ceramic (alumina) support bushings also play a role in the conditioning process. That is, charge must be applied to the insulator surface before the theoretical voltage hold-off can be achieved [14].

The experimental parameter space is further complicated by the geometry of the electron gun and the Hibachi foil support structure. The front of the electron
gun is comprised of an eighty percent transparent mesh which makes up the control grid. The mesh forms an equipotential surface although the surface electric fields are increased due to the field enhancement arising locally at the wires of the mesh. The ribbed Hibachi structure (anode) also increases the field at the anode surface. These field enhancements cause field emission, especially during the conditioning process.

It should be mentioned that during the conditioning process, surface contamination such as adsorbed gases may be removed. As a precaution against layers of gross contamination prior to installation, the electron gun, vacuum insulators, vacuum chamber, and Hibachi structure were thoroughly degreased and cleaned using vacuum compatible solvents. However, this cleaning process does not prevent gas monolayers (or multilayers) from forming during installation at atmospheric pressure. These layers, which form as a result of being exposed to gas, may also be desorbed during the conditioning process. The desorbed layers can result in a vacuum pressure rise. Conversely, the pressure rise may result in plasma and metal vapor being present during the cold cathode conditioning process. Also, as vacuum arcs occur, metal is sputtered from both the anode and cathode (electron gun).

The hot cathode test parameter space is even more complicated than that of the cold cathode tests. The presence of heated thoriated tungsten filaments implies elevated temperature operation of the cathode structure, the ceramic bushings, the anode or Hibachi, the vacuum
chamber, O’ring surfaces and the foil interface. Thus, these surfaces may desorb gases for several weeks before coming to equilibrium [15]. The presence of an electron beam in the cathode-anode gap further complicates the comparison between the hot and cold cathode tests. An electron beam can easily ionize gases on surfaces [15]. In the event that gas layers are desorbed, a local Paschen minimum may occur near the surface, leading to a local breakdown. Also, whiskers or enhanced microprotrusions can rapidly be heated to vapour temperatures, evolving both gas and metal vapours.

During hot cathode operation, low energy 200 - 500 eV electrons are extracted from the thoriated tungsten surface by the extraction grid. These low energy electrons are focused in the extraction control grid region. After leaving this region, they are accelerated in the control grid-anode (Hibachi) space to 150 - 250 keV. For the most part, the grids are quite transparent. However, 30-40 percent of the electrons emitted by the filaments are intercepted by the grid structure. The intercepted electrons can cause desorption and ionization of gas layers if they are on the grid surfaces. Thus the geometry and physical operation of the hot cathode and cold cathode tests make quantitative analysis of the two experiments quite difficult. However, the contribution of key physics parameters can be estimated. Common parameters as well as parameters unique to each experiment are discussed in the following paragraphs.
As aforementioned, field emission leading to collapse of the anode-cathode voltage was observed prior to cold cathode conditioning. It is also interesting to note that anode-cathode voltage collapse occurred at a pulsed voltage of -150 kV. The voltage breakdown across the vacuum bushing and/or the cathode-anode gap may have resulted from field emission. This hypothesis is supported by the high voltage DC conditioning tests. Upon application of -90 kV to the cathode-grid structure, about 1 mA of current emission was measured.

Electron emission at these low voltages has been previously noted by other investigators. Vacuum electron emission at 37 kV/cm has been documented for various metals with thin oxide layers [16]. However, this electron emission process can be discounted with the following considerations of electron generation rate. The peak calculated cathode field at 90 kV is 48 kV/cm. This includes a field enhancement factor of four based upon published field enhancement corrections for a mesh above a planar surface [17]. At 37 kV/cm, the reported electron generation rate is on the order of \(10^3 \text{ to } 10^6\) electrons per second [16]. Experimentally an electron generation rate of \(6.25 \times 10^{15}\) electrons per second was observed during the cold cathode conditioning tests. This is many orders of magnitude above that reported for thin film electron emission, even though the surface fields are comparable. Thus, it is unlikely that electron emission based solely upon the presence of oxide films is responsible for the emission current.
Two explanations are offered based on vacuum field emission experiments reported in the literature. Although the macroscopic field enhancement of four was computed from the geometry, the microscopic surface enhancements can easily be an order of magnitude above this number. Investigations have shown that, prior to conditioning, many surfaces have clumps or microprotrusions present on the metallic surface [18]. These clumps or microprotrusions can give rise to field enhancement factors on the order of 10 to 1000 times that over macroscopic field enhancements [19]. The electron beam gun however did not lend itself to the generation of data which could be used to determine the field enhancements present on the cathode surface. As previously mentioned, the anode-cathode electric field was within a factor of 7 of the high voltage bushing field. Thus, the field emission current measured included both leakage current across the insulator surface as well as the electron emission.

Based on the microscopic field enhancements reported in the literature, it has been hypothesized that clumps or microprotrusions were partially responsible for initiation or emission of electrons at the cathode surface. If even a modest field enhancement of 7.5 is considered for the protrusions, subsequent surface electric fields of 360 kV/cm can result at applied voltages of -90 kV, or so. This could, of course, easily explain the high electron emission currents of 1 mA for it is well known that $10^6$ A/cm$^2$ - $10^7$ A/cm$^2$ field emission current densities can result from surface fields above 250 kV/cm.
Once emission begins, the microprotrusions or clumps begin to heat. At this point three things may occur. The microprotrusion may heat to a temperature above which the metal may have a substantial vapour pressure. The field enhancement at the tip of the microprotrusion is subsequently reduced as metal is melted away [21]. Once the many sharp projections present are removed, field emission stops and the vacuum gap is conditioned [21]. The conditioning process at the surface is somewhat more complicated than this would imply. Two additional factors may result. If the plasma or metal vapour evolved closes the anode-cathode gap, a vacuum arc may occur. The energy of the vacuum arc must, however, be limited to less than 10 J for conditioning to occur. If the arc energy is less than 10 J, the field enhancement may be removed, and that particular emission site conditioned [19]. Above about 50 J, surface damage may be enough to effectively decondition the surface. It is likely that vacuum arcs of small energy were one of the conditioning processes present in the cold cathode experiments. As the voltage was raised, subsequent vacuum gas pressure rise was noted along with the arcing.

There is, however, an additional phenomenon which may be observed if microprotrusions, "whiskers" or clumps are present when an electric field is applied. Cranberg reported the clump theory in 1952 [22]. If the clumps or surface imperfections are loosely held to the surface and become charged, an electric field may accelerate the charged clumps toward the anode or cathode. Or, as the
microprotrusions heat from the deposited energy they may pinch off to form a clump and be accelerated toward the anode where it may arrive with sufficient energy to vaporize the clump. In any event, vacuum arcing may develop in the evaporating clump. This effect is only valid if the anode field is less enhanced than that at the cathode surface. Maitland further expanded upon these results, showing that the observed breakdown mechanism is rather complicated, and may be a result of other factors [23]. The clump theory however may not account for the effects seen in large vacuum gaps at the initiation of the discharge. The clump initiation theory has not been confirmed for the breakdown mechanism found in the cold cathode or hot cathode experiments.

Although it is believed that surface imperfections are primarily responsible for the field emission seen during conditioning, one alternate explanation has arisen based upon research reported by Prichard [24]. That is, on the basis of Prichard’s research, an ion exchange multiplication mechanism could also account for the voltage breakdown observed in vacuum gaps with electrode separations of 5 cm or more. In 1972, Prichard investigated the voltage breakdown behavior of 1-15 cm vacuum electrode gaps at voltages up to 400 kV. When DC voltage was applied using a Pelletron generator, the voltage breakdown of vacuum gaps varied linearly with electrode separations of 1-5 cm. At electrode spacings of 5-15 cm a constant breakdown or conditioned voltage of 230 kV was observed. Moreover, Prichard also found that emission current was
present when voltages as low as 90 kV were applied to unconditioned electrodes, even though the electrode spacing was on the order of 5-15 cm [24]. A similar behavior was observed during the cold-cathode conditioning process.

Research by Prichard also confirmed a pressure-voltage breakdown dependence similar to that reported by other investigators at small gap spacings. Admission of gas with a subsequent rise in background gas pressure increased the voltage hold-off of both the large and small cathode-anode gaps [24]. However, in contrast to these findings, Prichard observed that the 5-15 cm gaps did not follow the Cranberg model. Based upon this observation, the initiation of the discharges observed (for the 5-15 cm spacings) during the experiment cannot be attributed to clumps.

To investigate the observed breakdown process, Prichard injected ions into the gap using sodium chloride films applied to the electrodes. Mass spectra and Faraday cups were used to confirm the presence of ions. The discharge and predischarge currents were closely correlated with the presence of negative ions emitted from the cathode surface rather than electrons. To confirm this, Prichard also injected electrons into the large area gap. Upon injection of electrons, the gap voltage and the frequency of breakdowns remained constant. This indicated that large area gap discharges may arise not as a function of electron generation, but as a result of ion initiated current.

Upon review of the cold cathode test results described in this thesis, similarities are noted between Prichard’s experiments and the conditioning experiment. Initial
emission current was measured at 90 kV at a anode-cathode gap of 7 cm when the hot-cathode was conditioned. It is interesting to note that the emission current at these relatively voltages and large vacuum gap spacings was also found by Pritchard [24]. During the cold-cathode conditioning procedure, the conditioned voltage attained slowly increased to 245-250 kV. Although the DC power supply would supply 260 kV, the large 7.5 cm gap would not condition above 250 kV. Moreover, frequent vacuum discharges were noted during the cold-cathode conditioning process. When the power supply current, and thus the cathode conditioning current, was limited to 0.4-0.5 mA, the vacuum discharge frequency decreased by a factor of 4 to 5, or to about one discharge every 2-4 minutes. Thus the frequency of the vacuum breakdown process of large area, large vacuum gaps spacings may be a function of the ion multiplication process as found by Pritchard [24]. By reducing the power supply current, the ion multiplication (or ionization) process may be decreased, thus decreasing the frequency of vacuum breakdown. A similar process was postulated by Pritchard based upon his Pelletron experiments [24]. When the Pelletron charging current was decreased, the frequency of discharges decreased although the voltage across the vacuum gap remained constant at 230 kV (5-15 cm spacings). The voltage breakdown frequency dependence on conditioning current in Prichard’s experiments was found to be a result of an ion exchange multiplication process [24]. Negative ions are emitted from the cathode during conditioning. The negative ions
bombard the anode which thus yields secondary ions which are then accelerated back onto the cathode to create further anode directed ions. This ion multiplication process is primarily dependent on the anode-cathode current and, to second order, the cathode voltage. If either the cathode-current or cathode-voltage is decreased, ion multiplication drops and the generation of secondary ions falls below unity.

The vacuum breakdown dependence upon current and voltage suggests that a particle exchange process may lead to a vacuum breakdown. The positive-ion hypothesis has however been ruled out, due to the frequency dependence of vacuum breakdown upon current [16]. The positive-ion, negative-ion process however is similar to the observed vacuum breakdown dependence upon current seen during the cathode conditioning experiments. The positive-ion, negative-ion hypothesis suggests the vacuum breakdown occurs when the following inequality is satisfied

\[(AB + GH) \geq 1\]  \hspace{1cm} (5.2)

where A is the average number of positive ions produced by one electron, B the average number of secondary electrons produced by one of these positive ions, G the average number of negative ions produced by one positive ion, and H the average number of positive ions produced by one negative ion [16]. The G-H products have been measured at 250 kV and have been found to be less than unity or 0.51, 0.25, and 0.25 for copper, aluminum and steel electrodes.
respectively [16]. Thus, the positive-ion, negative-ion process depends quite strongly upon the vacuum gap voltage and current prior to breakdown. If either positive or negative ion production falls as a result of a current or voltage decrease, the probability of vacuum breakdown decreases. Thus, it could be hypothesized that the dependence of the vacuum breakdown frequency upon current was a result of a particle exchange process.

The similarity between Prichard’s results and the cold-cathode conditioning tests are remarkable. The frequency of the vacuum breakdowns in both experiments were dependent on the conditioning current. Moreover, emission current at 90 kV was noted in both experiments even though the electrode spacings were quite large (5-15 cm).

It is, however, unlikely that the observed vacuum breakdowns were a result of electron emission alone. The mean free path of electrons in vacuum is hundreds of centimeters even at modest energies of 100 eV [25]. At ~ 100 keV the mean-free path of electrons is thousands of centimeters. Thus it is likely that an ion exchange multiplication process did indeed contribute to the voltage breakdowns observed during the anode-cathode conditioning process. Further research is required to substantiate or confirm the presence of this phenomenon in the cold-cathode conditioning process.
5.3.4 Vacuum Insulator Flashover Conditioning Phenomenon

Other vacuum discharge processes may also have influenced the conditioning procedure. The most likely is the conditioning of the two ceramic bushings, which support the cathode. As aforementioned, the average calculated electric field across the ceramic bushings is within a factor of seven of the average cathode-anode electric field. Thus, it is quite difficult to "unfold" the leakage current across the insulator and the cold cathode field emission current. However, many of the vacuum conduction processes observed for vacuum gaps are similar to those of vacuum insulator processes. For completeness, a review of vacuum breakdown phenomena across insulator surfaces and their relevance to the conditioning process is discussed.

Prior to conditioning of the vacuum system, single shot, hot cathode tests confirmed the presence of cathode-anode voltage collapse. The fall-time of the voltage collapse was not measured, however. The response time (risetime) of the voltage monitor was 200-300 ns. Moreover, the voltage monitor was at a distance of 6 m from the vacuum system. Although the voltage data indicated a fall-time somewhat less than 200 ns, the formation time of the arc or breakdown could not be determined accurately. It should be noted that the collapse of the cathode-anode voltage was quite fast. Thus, insulator flashover cannot be ruled out as the possible initiator of a complete anode-cathode breakdown process.
In 1960, Ian Smith, while at Aldermaston (AWRE) investigated the formation time of discharges across vacuum insulators [26]. Smith found that the breakdown process was quite rapid. In fact, even across 50 cm long insulators, a discharge could form in 20 ns. Based on Smith’s work, Watson proposed an electron avalanche mechanism which explains the rapid growth of the discharge across vacuum insulators [27]. Watson surmised that thermionic electrons emitted off of the cathode strike the insulator surface. Secondary emission greater than unity leads to a net positive charge on the insulator surface. This positive charge further enhances the electric field near the cathode. If the surface field and cathode field are above the threshold for avalanche, electron avalanche conditions ensue which lead to voltage breakdown across the insulator surface. Watson also postulated that the dielectric constant as well as the shape of the insulator with respect to the applied electric field influence the breakdown process. This was supported by the result obtained by Smith while at Aldermaston. Experiments by Smith showed a voltage breakdown curve dependent on the angle of the insulator with respect to the electric field. It should also be mentioned that both Smith and Watson found a strong conditioning effect for nanosecond pulse-widths. Initial test samples would "flashover" well below their theoretical limits, but would subsequently condition to a set breakdown voltage.

Milton substantiated the conditioning effect reported by Smith and Watson for nanosecond pulse-widths [28]. It
is, however, interesting to note that Milton was unable to find a conditioning effect for microsecond pulse-widths. Milton did find a DC conditioning effect for insulators as previously noted by Gleichauf [29].

Gleichauf observed a DC conditioning phenomenon similar to that observed during the DC conditioning of the cathode assembly and the high voltage bushing. Gleichauf found that a series of breakdowns would gradually increase the voltage hold-off of the insulator. Removal of the conditioning voltage led to a decrease in the ability of the insulator to hold voltage [29]. Subsequent application of voltage resulted in fast reconditioning of the insulator. Prior history of the insulator also greatly effected the time which was required to recondition the insulator. That is, the time between reconditioning determined the initial breakdown voltage and the time required to condition the insulator back to full hold-off voltage. The same loss of conditioning was also found if the insulator was brought up to atmospheric pressure, irrespective of the time for which it was exposed to atmosphere [29].

A similar effect was found during the cold cathode DC conditioning process. Removal of the DC voltage was found to "decondition" the insulator. If the DC conditioning voltage was re-applied to the electron beam system, breakdown at a substantially lower level than that previously observed was found, although the system would condition much more quickly. It is also interesting to note that DC conditioning was also lost when the electron
beam system was exposed to ambient atmospheric pressure. This is, of course, similar to the results reported by Gleichauf [29].

The electron beam experimental results described herein were also similar to those reported by Smith. The ceramic (Alumina) bushings, 50.8 cm in length and 25.4 cm in diameter, were of the same length as those tested by Smith [26]. The pulsed breakdown conditions occurred during the risetime of the applied 50 μsec pulse. As previously mentioned, the voltage diagnostics did not have the response-time to resolve the formation time of the breakdown process. However, the flashover process did appear to be quite fast, and in review of Smith's experiments, surface flashover in vacuum can occur in 20-30 ns even across 50 cm long insulators.

It should also be mentioned that the rise in background pressure during the cold cathode conditioning process may favor ceramic bushings conditioning as one of the predominant mechanisms leading to high voltage vacuum flashover both prior to conditioning and during conditioning. Electron avalanches over insulator surfaces have been shown to desorb gas both prior to surface flashover and as a result of surface breakdown [30]. R.A. Anderson and Brainard confirmed this experimentally by measuring the mass spectra of the background gas and the time of flight of ions generated at the insulator surface [30]. Pillai and Hackman derived a model based on Watson's results and the experimental evidence presented by Anderson and Brainard [31, 32, 33]. Pillai and Hackman postulated
that as electrons bombard the insulator surface, molecules on the surface (CO, H₂, N₂, H₂O, etc.) are copiously desorbed. As the molecules are liberated from the surface, the probability of development of charge avalanches is further increased.

It is rather intriguing to note that a rapid rise in the vacuum background pressure was found during the DC high voltage conditioning process. If, indeed, the model presented by Pillai and Hackman is valid, desorbed gases or molecules released by electron-ion bombardment may account partially for the chamber pressure rise.

Pulsed hot cathode tests done prior to conditioning showed voltage breakdown characteristics similar to those measured during the DC conditioning process. Thus, it is likely that both the vacuum gap processes and the ceramic insulator characteristics discussed were responsible for the low voltage breakdown behavior during the applied 50 μsec pulse. Pillai and Hackman have shown that DC and pulsed voltage breakdown are the same for insulators which are straight [33]. Thus, the mechanism for 50 μsec pulse-widths may be the same as those for DC conditions.

One other factor should be mentioned. During full temperature operation of the hot cathode, a pressure of 3 x 10⁻⁶ Torr was noted prior to application of high voltage. The background gas was analyzed using a residual gas analyzer (quadrupole mass spectrometer, DYCOR Industries). Residual gas analysis showed four prominent mass spectra. Carbon Monoxide (CO) accounted for 31.2% of the background gas, with water vapor (H₂O), Carbon (C) and
carbon dioxide comprising 12.5%, 5.6%, and 3.12%, of the measured mass spectra, respectively. No hydrocarbons or silicon components were observed; this absence is probably due to the use of a cryogenic pump. The remaining 48% of the spectra consisted of contaminants too small to resolve with the residual gas analyzer.

The mass spectra measured are believed to result from gas desorption for it is known that insulator surfaces and vacuum surfaces desorb CO₂, CO, H₂O, and N₂ [34]. Thoriated tungsten filaments are also known to desorb CO when heated [35].

During pulsed operation, electrons are extracted from the filament surface by the extraction grid and control grid of the electron beam gun. The 100-500 eV electrons then enter the control grid anode-cathode region and are accelerated to 150-220 keV. To determine if the low energy electrons could possibly ionize the residual gas present in the system, the mean free path of 100 eV electrons was calculated. The collisional cross-section of CO is \(10^{-16}\) cm² [36]. At a pressure of \(3 \times 10^{-6}\) Torr and an average gas temperature of 500°K the mean free path of the electrons was calculated to be on the order of \(10^4-10^5\) cm. Thus, it is unlikely that vacuum breakdown as a result of the background gas was responsible for the pulsed low voltage flashover observed.

Consistent with this result it is likely that both insulator flashover and/or vacuum gap flashover was the primary mechanism responsible for the pulsed low voltage breakdown originally observed.
5.4 Verification of Electron Beam Uniformity and Suitability of Foil Windows for Extraction of the Electron Beam

To verify the current density of the electron beam into a gas, filament emission current and electron beam uniformity were measured with a scanning Faraday cup in air. Concurrently the response and lifetime of both $2.54 \times 10^{-3}$ cm aluminum and $1.78 \times 10^{-3}$ cm titanium foils were tested. Lifetime of the foils were deduced by monitoring the foils for pin-hole leaks. The over-pressure stress arising from the electron beam energy deposition into a gas and the chemical reaction of the gas with the foil were found to have an effect on foil performance. The results of the tests are reported below and their effect on the final system parameters is discussed.

5.4.1 Test and Evaluation of Aluminum and Titanium Windows for Extraction of the Electron Beam

Two different foil interfaces were tested during the experiments. Both aluminum and titanium foil were the primary foils of choice. From a system standpoint, aluminum foil was the most desirable. Aluminum foil has a much lower areal density than titanium, thus less electron beam energy is deposited into the foil. The electron beam energy (voltage) can thus be lowered to reduce the stress on the modulator and electron beam components. The lower areal density of the aluminum foil also implies less
scattering of the electron beam as the beam traverses the foil. Less scattering results in a more uniform electron beam and less divergence through the foil. From a practical standpoint, aluminum has several distinct disadvantages. The first disadvantage is the chemical reactivity of aluminum. Aluminum is known to react chemically with oxygen and nitrides even at low temperatures, whereas titanium reacts only at temperatures over 315°C. Titanium is fairly inert in oxygen at the 93°C operating temperature of the foil-Hibachi-gas interface.

The second disadvantage of aluminum arises from the material strength. Titanium can operate at stress levels of 80,000 psi. Aluminum foil has a stress-strain level two to three times lower than titanium and is thus less resistant to pin-hole formation.

To resolve the question of which foil material to use for the final CO₂ laser tests, both aluminum and titanium foils were tested at 20 mA/cm into air at STP. Foil lifetime and performance was based on the time required for pin-holes to develop. A procedure based on the effusion rate of gases was developed to detect pin-holes using the mass spectrometer (residual gas analyzer). Gases of low molecular weight effuse more rapidly than do those of high molecular weight [37]. Thus, by determining the lowest weight background gas which results from atmospheric leaks, a qualitative estimate of the time required for pin-holes to develop could be surmised. At atmospheric pressure, the lowest molecular weight gas is helium. Before the start of each foil test the background helium level was measured as
a calibration level. The background helium level was found to be in the $10^{-9}$ to $10^{-8}$ Torr range.

Aluminum foil was the first material tested. At a 1-2 pps repetition rate and 20 mA/cm$^2$ beam current, the aluminum foil quickly degraded over 2-3 hours of operation. A factor of 2-3 rise in helium background gas was found after 2-3 hours. During operation into air a white discoloration was visually observed on the foil surface. Further tests of the aluminum foil at full 10 pps repetitive rate quickly degraded the vacuum chamber pressure to the point where tests had to be discontinued. The chamber pressure rose to $-8 \times 10^{-6}$ Torr at which time the tests were terminated. The rather short operating lifetime of aluminum foil (5-hours) was found to be impractical from the system level. In fact, the relatively fast chamber pressure rise was found to degrade the filament emission level probably by a process in which the background oxygen strips the thorium surface from the filaments. Prior to continuation of the foil tests, a carburization procedure was developed to reactivate the filaments (Appendix E).

Following reactivation of the filaments, a titanium foil was mounted on the Hibachi structure. The vacuum helium partial pressure was measured as a reference. Initial 2-3 hour operation of the foil at 1-2 pps and 20 mA/cm$^2$ current density did not show a change in the helium partial pressure. Thus no pin-holes were detected in the initial foil tests. Three, one-hour tests at 10 pps and 20 mA/cm$^2$ were then done to verify the foil lifetime.
No degradation during the accumulated 180,000 shot tests was noted. Based on these results, titanium foil was selected for the system integration phase.

5.4.2 Electron Beam Uniformity Measurements

In order to measure both the filament emission current and the electron beam uniformity along the Hibachi length a scanning Faraday cup was installed 5 cm from the foil. The Faraday cup consisted of either 1 cm$^2$ or 10 cm$^2$ charge collector coupled back to the electron beam chamber ground. The current through the change collector was monitored with a Pearson commerical current probe. The charge collector was motorized to allow uniformity measurements while operating the electron beam system (Fig. 5.5). This allows the Faraday cup to move at a velocity corresponding to the Hibachi-rib/filament-filament distance between pulses. The threaded shaft upon which the Faraday cup moved was coupled to a potentiometer. A power supply coupled to the potentiometer allowed the X trace on an oscilloscope to be synchronized to the position of the Faraday cup (Fig. 5.5). It should also be mentioned that a variable speed motor was used to drive the Faraday cup.

Before characterizing the electron beam uniformity, stationary Faraday cup measurements were used to verify the filament emission current. For a measured beam current density in air of 20 mA/cm$^2$ the current density at each filament was 2.55 A/cm$^2$. The corresponding grid voltage was 557 V. Indendent pyrometric measurements during the
Figure 5.5 Faraday Cup Diagnostic Setup.
experiments showed a filament temperature of $1900^\circ K$. Both the $2.55 \text{ A/cm}^2$ emission current and filament temperature were found to be within the established parameter space for $3.81 \times 10^{-2} \text{ cm (15 mil)}$ thoriated tungsten filaments, for lifetimes of several thousand hours [38].

Utilizing the stationary cup measurements the adjustability of the post foil electron beam current density was also confirmed. Electron beam current density was confirmed for beam energies of 150 to 200 keV in air. The measured electron beam density was found to meet or exceed all desired electron beam specifications (Table 1.1) of $5 \text{ mA/cm}^2 - 20 \text{ mA/cm}^2 (12.5 - 50 \text{ A})$.

To conclude qualification of the electron beam system, scanning Faraday cup measurements were done to verify electron beam uniformity along the length of the 250 cm cathode. Both $1 \text{ mA/cm}^2$ and $10 \text{ mA/cm}^2$ charge collectors were used to verify uniformity over both the 10 cm wide electron beam aperture and the centre of the electron beam. A $\pm 10\%$ variation of electron beam current density along the cathode length was measured. Both the $1 \text{ mA/cm}^2$ and $10 \text{ mA/cm}^2$ measurements are shown in Fig. 5.6.

Upon review of Fig. 5.6 a fine repeatable structure in the electron beam current density is noted. This repeatable structure has been attributed to thermal expansion of the cathode. Thermal expansion of the cathode results in a small misalignment of the extraction grids with respect to the Hibachi inter-rib spacing. The electron beam is partially intercepted by the deep inter-rib geometry, reducing the geometric transparency of the
Figure 5.6 Spatial uniformity of cathode including pulse-to-pulse power supply non-reproducibility, 200 keV, 18 mA/cm² current density.
Hibachi-foil interface. The ± 10% electron beam uniformity however, met the desired system specifications.

5.5 Summary

Based upon detailed design calculations, an electron beam system which expands the parameter space for electron beam pumped CO$_2$ lasers has been built and tested. Two such systems have presently been installed in CO$_2$ lasers and are operational. To date, the systems have accumulated well over $5 \times 10^6$ shots. Moreover, the system has met or exceeded the design specifications. At present, both electron beam guns have been operated well outside their intended parameter range both above $20 \text{ mA/cm}^2$ and below $5 \text{ mA/cm}^2$ current densities. Both electron beam systems have been adapted to these requirements. Moreover, over 1500 hours has been demonstrated on a set of filaments before replacement. Similarly, the $1.788 \times 10^{-3} \text{ cm}$ titanium foil interface has proven to be extremely trouble free. Concurrently the pulse power modulators have required no maintenance, thus proving the suitability of hollow-anode thyatrons for use in electron beam systems, even with pulse durations of 52 µsec.
REFERENCES
CHAPTER 5


6. Private Communication; Tom Churchill, Spectra Technology, Seattle, WA.


APPENDIX A

REVIEW PAPERS

The following papers review the 12 MW E-beam system. The first paper reviews the essential aspects of the design, construction and performance of the electron beam system in its final form as incorporated into the CO$_2$ laser. The second paper gives relevant details which established the basic technology from which the final system was developed.
A REPETITIVE PULSED POWER SYSTEM FOR THE GENERATION OF 45 µS, 200 keV ELECTRON BEAMS
FOR CO₂ LASERS


Pulse Sciences, Incorporated
600 McCormick Street
San Leandro, California 94577

ABSTRACT

A two module electron beam source operating over a wide range of output parameters has been designed and fabricated to be used in conjunction with a pair of electron beam sustained CO₂ lasers. Each module comprised a grid-controlled thermionic electron beam gun incorporating a compact grid pulser for control of the electron beam, a 250 kV thyratron switched modulator for acceleration of the electron beam, 1 kHz filament heater, and a complex computerized control system. Table 1 summarizes the overall system specifications.

Table 1. Summary of electron beam requirements.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cathode Area</td>
<td>2500 cm²</td>
</tr>
<tr>
<td>Electron Beam Length</td>
<td>250 cm</td>
</tr>
<tr>
<td>Electron Beam Width</td>
<td>10 cm</td>
</tr>
<tr>
<td>Post Foil-Current Density</td>
<td>5-20 mA/cm²</td>
</tr>
<tr>
<td>Current Density Uniformity</td>
<td>± 10%</td>
</tr>
<tr>
<td>Post Foil Electron Average</td>
<td>150-200 keV</td>
</tr>
<tr>
<td>Energy</td>
<td></td>
</tr>
<tr>
<td>Beam Pulselength (±10%)</td>
<td>45 µm</td>
</tr>
<tr>
<td>Maximum E-beam Rise and Fall Time</td>
<td>4.5 µm</td>
</tr>
<tr>
<td>Pulse Repetitive Rate</td>
<td>10 pps</td>
</tr>
<tr>
<td>System Lifet ime</td>
<td>10⁶ shots</td>
</tr>
<tr>
<td>Distance</td>
<td>6 meters</td>
</tr>
</tbody>
</table>

The high voltage cathode assembly employs 132 thoriated tungsten filaments distributed over the area of the 250 cm × 10 cm output window. The cathode assembly including the control grids is supported by two high voltage ceramic busbars in a stainless steel vacuum chamber. For acceleration of the electron beam, a pulsed 150-200 kV voltage is applied by a thyratron switched modulator, between the cathode assembly and anode window. The cathode current is controlled by a 200-800 V grid pulser. The grid pulser which is fiber-optically programmed is floated at the cathode potential and is located in the modulator tank.

The 250 kV modulator provides 45 µs, 150-250 kV acceleration pulses at 10 pps. Electron beam load impedances can be accommodated over a 4:1 range. A 1100 J, 35 kV pulsed network (PFN) is coupled to the electron beam gun by a 1:19 step-up transformer. Design and modulating parameters are discussed which assure a 10⁶ shot lifetime.

*This work was sponsored under Contract No. F19628-81-C-0011 by Spectra Technology, Inc.
**Present address: Power Spectra, Inc., Fremont, CA.
**Present address: Leland Schlitt Consulting Services, Livermore, CA.

The grid pulser employs a SCR switched, 15 section FPM. Design and modelling of the grid to cathode load, as well as the transient response of the grid pulser with the electron beam is presented. Computer models are used which allow accurate modelling of the V-T grid to cathode load impedance.

INTRODUCTION

The low current density, 45 µm pulselength and the requirement for independent control of electron beam energy and current, preclude the use of cold cathode electron sources. A grid controlled thermionic gun employing an array of directly heated filaments with individual extraction grids was selected. In the patented triode configuration, (1) precise control of the electron flow is achieved by decoupling the functions of current extraction and electron acceleration in the anode to cathode gap (A-K). To obtain uniform electron extraction from the 132, 15 mil diameter thoriated tungsten filaments, each filament is combined with a concentric extraction grid. As shown in Figure 1, this hemicylindrical grid extracts a space charge limited flow from the thoriated tungsten filament and injects the electrons into a field-free region established between the extraction grid and a common planar control grid. The control grid is positioned so that the extracted electrons uniformly illuminate its surface. The resultant spatially uniform electron beam is accelerated in the A-K gap. The high 30 kV/cm acceleration fields, are separated from the low extraction fields by the intergrid field-free region. While the total energy of the electrons depends on the combined effect of the extraction and acceleration fields, the current density is only a function of the applied grid pulse voltage.

![Dual grid controlled electron beam gun configuration](image)

Figure 1. Dual grid controlled electron beam gun configuration.

The 132 filament/grid arrays are assembled on a baseplate as shown in Figure 2. Self-tensioning springs incorporated into each filament-grid array prevent droop of the fila-
ments as they expand at the 1900-2000 K temperature of the emitter. Although the filament to extraction grid electron flow is space charge limited, the A-K gap requires operation well below space charge limits. Nominaly the control grid of the hot cathode assembly is located 7.5 cm from the anode. At 150 kV operating voltage, the space charge limit is 2.41 A/cm², two orders of magnitude greater than the 20 mA/cm² beam current density. Consequently, space charge effects in the anode-grid region are negligible. Because the cathode assembly is pulsed to 150-250 kV with respect to the vacuum chamber, field enhancements can lead to cold cathode emission, and must be avoided by eliminating sharp emitting surfaces. This is accomplished by enclosing the 135 discrete filament arrays in an electro-polished stainless steel cowl (Figure 3). The cowl also assures uniform acceleration fields between the cathode assembly and anode.

Figure 2. Photograph of filament extraction grid arrays partially assembled onto the cathode reflector.

Figure 3. Cross-section of cathode assembly and electro-polished enclosure.

Fifteen-all-diameter, 2% thoriated tungsten filaments were selected to obtain the required 2000 hour lifetime. Thoriated tungsten alloys have demonstrated durability under adverse operating conditions, and are relatively inexpensive. Large area electron guns (2000-5000 cm²) have used thoriated tungsten filaments with demonstrated lifetimes of over 2000 hours (2) even after repeated cycling to atmosphere. In addition, trace amounts of CO, CO₂, and H₂O vapor do not noticeably degrade the emission characteristics of the filaments.

The high voltage cathode is supported by two high vacuum, ceramic bushings (Figure 4). During operation, a vacuum pressure of 1-3 x 10⁻⁸ torr is maintained to prevent high voltage breakdown and degradation of filament emission. The accelerating potential is supplied through one bushing along with filament power and the grid pulse. The high voltage acceleration pulse, grid pulse, and filament power are separately routed through three parallel 6 meter coaxial cables which originate at the modulator. Since the voltage drop along each filament must be small compared to the grid-to-filament control pulse voltage, filament power must be delivered at high current (500 A), and low voltage (15 V), which is not compatible with the use of commercially available high voltage coaxial cable. The filament power is transported by the modulator electron beam gun cables at high voltage (400-480 V), and low current (15 A), and stepped down to 15 V at the electron beam gun. In order to reduce the size and weight of the filament step-down transformer, which must be mechanically and electrically isolated from ground, a 1 kHz power system is used rather than the normal 50-60 Hz line frequency. As illustrated in Figure 4, the transformer is located in the high voltage bushing assembly. The grid pulser is also powered from this system.

**MODULATOR SYSTEM DESIGN**

The modulator which provides the cathode acceleration pulse is based on a thyatron switched, Type E pulseforming network (Figure 5) (2). Table 2 outlines the respective specifications of the modulator as well as the thyatron voltage and current.

The pulseforming network is resonantly charged and then discharged through a step-up pulse transformer. Fault mode conditions including a shorted load present minimal problems to the hollow anode thyatron, (ESV 1574C), primary switch. Protection from open circuit fault modes is provided by a metal oxide variator stack across the output of the step-up transformer. This type of modulator has a fixed impedance. Since the modulator is required to operate into a wide range of output loads, matching resistors can be paralleled with the electron beam load. The matching resistors reduce the overall efficiency of
the system at low beam currents, however they provide a means of matching the PFN to the electron beam load.

**Figure 5.** Schematic of modulator layout and filament circuits.

**Table 2.** Summary of modulator and thyatron voltage and current specifications derived from the electron beam gun requirements.

| Load Voltage: | 150-200 kV |
| Modulator Current: | 20-150 A |
| Peak Power: | 37.5 kW |
| Energy Stored: | 1800 J |
| Electron Beam Load: | 1.7-11 kN |
| Impedance: | 11 kV |
| Thyatron Voltage: | 35 kV |
| Thyatron Peak Current: | 3000 A |
| Thyatron di/dt: | 4.3 × 10^5 A/sec |
| Rep-Rate: | Single shot to 100pps |
| Lifet ime: | 1 × 10^6 pulses |

During normal operation a 2:1 impedance mismatch, or 10% voltage reversal, is quite tolerable to a line-type modulator. The baseline design assumed a 15% voltage reversal, allowing for the conservative philosophy required to accommodate the 10^6 shot lifetime. A simplified model for the modulator was analyzed to select the required matching resistors and the subsequent current and voltage operating ranges.

The effective electron beam impedance requires some explanation. During the applied high voltage pulse the current emitted from a grid-controlled cathode is essentially constant. The emission current from the cathode is not the current which reaches the laser gas volume. Approximately 60% of the current is intercepted by the ribs and the anode foil. At low electron energies, 150 keV, approximately 20-10% of the electron energy is lost as the beam goes through the anode foil. The load voltage was therefore assumed to vary from 187-250 kV. This accounts for the energy lost in the anode foil and a 15% load reflection coefficient. The values shown in Table 2 include the intercepted beam current, and the current through the matching resistors. The equivalent electron beam load impedance varies from 350 to 11 kN, with a 6:1 variation in impedance range.

Table 3 shows the calculated values. The calculated operating ranges are based on two assumptions. The first assumption is that a negative reflection is preferable to a positive reflection. This implies that the system be matched at the low current and of each operating range (highest impedance). The second assumption is that a 80% efficiency for the modulator is desired at the highest operating current level. This sets R_par at about a 4:1 ratio with respect to the lowest load resistance.

**Table 3.** Operational and calculated matching resistance and dynamic impedance range of modulator.

<table>
<thead>
<tr>
<th>Electron Beam Load Impedance Range (Ω)</th>
<th>Electron Beam Dynamic Impedance Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>-106 &lt; R &lt; 0</td>
<td>R_par</td>
</tr>
<tr>
<td>1900 &lt; R &lt; 1600</td>
<td>1701 (0) (A)</td>
</tr>
<tr>
<td>1300 &lt; R &lt; 1200</td>
<td>1265 (0) (A)</td>
</tr>
<tr>
<td>750 &lt; R &lt; 720</td>
<td>645 (0) (A)</td>
</tr>
<tr>
<td>375 &lt; R &lt; 290</td>
<td>305 (0) (A)</td>
</tr>
<tr>
<td>Maximum Resistor</td>
<td>R_1 = 1251</td>
</tr>
<tr>
<td>R_2 = 8215</td>
<td></td>
</tr>
</tbody>
</table>

A detailed circuit model was evaluated which included the full PFN transfer characteristics and the inherent stray capacitance and inductance found in a physical layout. To accommodate the four operating ranges shown in Table 3, a 32 μs, 5.5-ohm Type E PFN was designed. The six-section PFN is impedance-tapered to compensate for the transformer pulse drop and to provide a 45 μs (90-90%) pulsewidth with a ± 8.5% flatness specification. A simplified circuit model is shown in Figure 6. The pulseforming network was designed to operate at a 32 kV peak charge voltage in conjunction with the available 35 kV hollow anode thyatron. A 1:19, 250 kV iron core transformer matches the low voltage PFN to the electron beam load. All parameters are referenced to the primary (Figure 6).

**Figure 6.** Simplified circuit model used to simulate the tapered pulseforming network response to a constant current load.

The electron beam load impedance history was simulated with a pulsed current source using SPICE, a circuit simulation code. Because the grid controlled cathode current is a function of the grid voltage and not the applied cathode voltage, the load at the modulator appears as a constant current source. A fixed resistive load is not a good approximation of a space-charge-limited vacuum diode or a grid controlled cathode. Rather a pulsed current source is used to simulate the beam load. The current source rise time and fall time were chosen to be 2 μs and 4.5 μs. An interactive circuit simulation and tuning procedure allowed the phasing of the current pulse and modulator pulse to be adjusted for a pulse flatness of ± 0.5%.

Figure 7 shows the normalized A-K, or load voltage waveforms of both a 6-ohm constant impedance PFN and a 5.5-ohm average tapered PFN (Figure 6) response to a constant current load. The dashed line represents the constant impedance PFN response to the electron beam load, or in the case of the simula-
tion, a pulsed current source. The response of the fixed impedancePFN to the pulsed current load is not unexpected because the coaxial cables linking the modulator to the electron beam source represent an uncharged stray secondary capacitance on the order of one PFN section capacitance (Figure 6). As the electron beam turns on, the stray secondary capacitance begins to discharge and the A-K voltage droops. Although the modulator is still supplying charge to the stray capacitance a constant impedancePFN cannot supply enough charge to compensate for the voltage droop. The voltage droop across the load can however be compensated by tapering the impedance of the PFN along its length. If the leakage inductance of the step-up transformer is considered as an integral part of the PFN output inductance, the tapered impedance of the PFN varies from 5.9-ohms down to 6.7-ohms. The tapered impedance allows the PFN to compensate for the load voltage droop as shown in Figure 7. In addition, tapering of the PFN impedance during the tuning process demonstrated the required 45 pulsewidth. Although the step-up transformer did not predict the subtle differences shown in the voltage pulse from varying the load impedance.

GRID PULSER SYSTEM

The design of the grid pulser is inherently more complicated than that of the modulator. To meet the required acceleration voltage flatness, a grid pulser with a 47 μs pulsewidth and a 2 μs risetime was required. A 1-ohm, 15 section, Type E PFN was chosen. The 15 section PFN was selected to meet the ± 6% grid voltage flatness requirement imposed by the circuit simulations and electron beam energy specifications. In addition, the PFN was designed to have a 2 μs risetime and 4 μs fall-time corresponding to the electron beam or pulsed current source used in the modulator circuit simulations. The modeling of the grid pulser PFN response is not straightforward. Because the grid-cathode current is space-charge-limited, Miller capacitance significantly affects the grid pulser risetime [6]. Miller capacitance, or the voltage dependent capacitance across the grid-cathode region does not lend itself well to transient modeling, on available circuit codes such as SPICE.

A rather simple model of the modulator and grid circuit interaction does allow the voltage dependent capacitance of the system to be modeled with SPICE. A simplified equivalent circuit of the modulator and grid circuit was constructed as shown in Figure 9. The model assumes two conditions. The first condition is that the grid does not intercept current flowing between the cathode and anode. In reality, this is not the case for the grids have a geometrical transparency and do intercept a small fraction of this current. How-
The voltage dependent grid capacitance and the voltage dependent grid impedance was calculated using a curve fitting program independent of SPICE. The polynomials describing the voltage dependent components were then used in SPICE to model the response of the PFN to the grid circuit (Figure 10).

Prior to integration with the electron beam system, the grid pulser was tested using a resistive load. To tune the grid pulser PFN to the desired flatness and pulse shape, additional circuit simulations were required. The voltage dependent load was replaced with a 4-ohm resistive load. The waveforms generated from the circuit simulations were then used to tune the PFN rise time and pulse flatness. The inductance of the spiral PFN inductors were tuned by inserting small, hollow, ferrite cores into the center of the inductors. Additional tuning was done by placing small discrete capacitors across the fixed PFN capacitors. As shown in Figure 11, the PFN rise time (dashed line) corresponds well to the circuit simulation rise time (solid line). The PFN waveforms however does not exhibit the oscillations shown by the circuit simulation. Because the step-up transformer acts as a low-pass filter, the oscillations predicted by the circuit simulation are reduced substantially. The waveform dV/dt of the PFN however matches the circuit simulation closely (Figure 11).

The grid pulser is located in the oil filled modulator tank and fires at cathode potential. The grid pulser includes the 1-ohm PFN which is impedance-matched by a 1:2 step-up transformer to the 4-ohm grid cathode impedance. A series regulated power supply which is programmable by fiber-optic link is used to charge the PFN (Figure 12). The 1-ohm PFN is discharged by a 800 V SCR through the output step-up transformer. To insure that the PFN SCR turns off, a second crowbar SCR is triggered to crowbar the PFN pulse. The crowbar circuit also allows the grid pulser width to be controlled.

![Figure 10. Block diagram of SPICE model used to simulate the grid pulser PFN response to the voltage dependent grid capacitance.](image-url)

![Figure 11. Simulated response of grid pulser PFN versus measured voltage response into a four-ohm load.](image-url)
inductor, the grid modulator is driven by four components: the tank, ensuring 7.2 ft. high x 7.6 ft. long x 7.6 ft. wide. When filled with oil, the modulator weighs about 14 tons and contains about 2000 gallons of oil. Oil cooling for the thyratron, and matching resistors is assured by directed oil nozzles and an integral oil cooling system. The modulator is linked to the electron beam gun system by three coaxial cables, each rated at 300 kV peak voltage.

Figure 13. Photograph of modulator layout.

Before tuning the modulator PPM, the modulator was integrated with the electron beam gun. Initial electron beam cathode tests indicated plateaus closure of the cathode to anode. The 1800 J modulator PPM made pulse conditioning of the cathode impossible. Rather than damage the cathode for multiple high energy breakdowns, a dc high voltage conditioning procedure was adopted. A 1 mA, 250 kV current limited power supply was used to condition the cathode. Initial conditioning of the cathode and high voltage bushings required 12-15 hours. Subsequent conditioning of the cathode after opening the chamber to atmospheric pressure required 1-2 hours.

Checkout of the system over the full operating range verified the accuracy of the modeling. Independent adjustment of the current and electron energy (cathode voltage) was achieved. Reliability of the electron beam system was also tested at full power. The modulator when operated continuously at 10pps, 210 kV, and full 160 A output current, performed to specifications. In addition, no tuning of the modulator PPM or grid modulator was required. The cathode voltage and current are shown in Figure 14. Although the circuit modeling indicated a trend in the operating range, tuning adjustment of the cathode grid pulse was required to tune the overshoot present on the cathode voltage pulse. The circuit calculations had predicted a slightly faster dv/dt on the rise of the cathode pulse. The slower rise time, due to stray PPM inductance and higher cathode capacitance, resulted in a longer required delay time between the applied cathode voltage and the grid pulse. A delay time of 5 µs was required versus the model's 3.5 µs prediction. In addition, a higher displacement current due to the under-estimation of the cathode to anode capacitance was noted. The initial slow current rise time seen on the current waveform of Figure 14 is a result of displacement current. Even though a negative bias was applied to the control and extraction grid, adjustment of the bias had no effect on the displacement current noted on the electron beam current waveforms.

Figure 14. Top trace: cathode acceleration voltage, 1 A/div showing the modulator response to the constant current load. Bottom trace: electron beam current, 1 A/div at 15 A/cm² average current density.

An electron beam current density of 5-20 mA/cm² was verified by operation into air. A dc current generated by a motor-driven runner traversed the length of the hibachi during repeat operation of the system. A ±10% variation of the beam current, regardless of beam length was demonstrated thus meeting full system specifications as outlined in Table 1. As shown in Figure 15, the electron beam current density is ±10% when the modulator pulse-to-pulse voltage variation of ± 2% is included. The fine repeatable structure along the length of the structure is due to thermal expansion of the cathode. Thermal expansion of the cathode results in a small misalignment of the extraction grids with respect to the hibachi support ribs. The resulting electron beam is partially intercepted by the deep inter-rib geometry, reducing the geometric transparency of the hibachi. However, priorization of the laser gas is unaffected and the electron beam system met full system specifications.

Figure 15. Spatial uniformity of cathode including pulse-to-pulse power supply/nongood reproducibility, 200 keV, 20 mA/cm² peak current.

Two such electron beam systems have been integrated with the customer's CO₂ laser.
A compact 250 kV modulator has been designed by Pulse Sciences, Inc. (PSI) which can easily operate under these stringent conditions. The modulator was designed as a general purpose driver for a high impedance thermionic diode. The modulator was delivered as a turn-key system complete with integral oil pumping and storage capability. A partial block diagram (Figure 1) illustrates some of the key features of the system. An automatic control system capable of monitoring all essential conditions of the ancillary support equipment and fault modes was designed into the system. Although the control system is discussed in a following section, a brief summary is required before proceeding. Because the modulator was designed as a stand-alone test facility for diode development, complete automatic control was required. The control system was designed to monitor and control charging voltage, detect short circuit and open circuit conditions, and provide shutdown in case of reoccurring fault modes. In addition, the control system was designed to monitor the status of the ancillary equipment (auxiliary supply, PPM charge voltage, etc.) and provide shutdown to avoid damage to the key components of the system including the thyatron, the step-up transformer, the main power supply and the high voltage bus. To illustrate the complexity and the key design issues, the following sections provide a detailed discussion of the modulator and controls.

Figure 1. Block diagram of the modulator and support systems.

MODULATOR DESCRIPTION

The modulator was designed to provide 50-250 kV, 1 ms pulses to a 400-ohm load. The design utilized a 35 kV, 15 kA hollow anode thermion (ENV 154T) to discharge a 1.8-ohm tapered waveform network through an iron core pulse transformer (15:1 step-up) (Figure 2). The pulses forming network was resonantly charged from a 5 kF capacitor. DC charged to approximately 17 kV. To provide for single-pulse operation, a trickle charge supply, adjustable from 0-35 kV was also provided.

The main power supply was sized to provide for inter pulse recharge of the previously mentioned 5 kF capacitor. The 30 kW power rating of the supply

CH1262-4/86/0000-0031501.00 © 1986 IEEE
The thyatron (LED 1574C) was selected for peak voltage hold-off, peak current and the ability to conduct high reverse currents. The thyatron was rated at peak currents of 15 kA and 7.5 kA reverse currents. Lifetime tests published on 11 kA operating conditions, reported minimum lifetime in excess of $1 \times 10^{10}$ shots at 11 kA and $10^6$ A/ sec. In addition, the thyatron had also been tested at 500 pps. Design goals of $1 \times 10^6$ shots with 10 kA peak currents and an anticipated $\Delta t/\Delta t = 5 \times 10^4$ A/sec at 100 pps were well within the thyatron specification test parameters. During short circuit fault modes, large reverse currents (10-12 kA) could possibly occur in the primary discharge loop. Although this was slightly higher than specification, tradeoff studies done, showed that the thyatron would conduct these reverse currents. The thyatron in the short circuit condition was also required to withstand several reversals until the voltage was damped out by the smaller diode circuit placed across the PFN (Figure 2).

A folded stripline PFN was constructed to provide a low impedance, low inductance PFN required to drive the 400-ohm diode load through a step-up transformer. The pulseforming network was a hybrid of the type E PFN and the Rayleigh PFNs. The pulseforming network shown in Figure 3 illustrates this. The required flatness of 2% for a 1 ms pulse was not easily achieved, along with the desired residence of 200 nsec. The inductance of the selected capacitors (ESL) and the output capacitance of the step-up transformer interacted and further complicated the design tradeoffs. Originally, 7.5 nF reconstituted mica capacitors were selected for the PFN capacitors. The equivalent series inductance of the capacitors was specified to be less than 60 nH. Unfortunately, in-house tests revealed approximately ESL's of 290 nH per capacitor. In low impedance PFNs, high ESL can dramatically affect the output pulse, resulting in early drop off and large variations in the pulse. Original circuit simulations had taken into account the ESL drop of the transformer and the large secondary stray capacitance (150-200 pF), and had shown that the high ESL would shorten the pulse. Early tests with the modulator revealed that the output pulse-width was indeed shorter by 80 ns, 920 ns in duration. To compensate for the high series inductance (ESL) (Figure 3), an eighth section was added. Although the inductor of the extra section (Section 1) is not mutually coupled to the other inductors of the PFN, the extra section did extend the pulsewidth to the required 1 ms pulsewidth and the required flatness. Typically when an extra section is added to a PFN, droop in the pulse is normal. However, the stripline design of the PFN reduced the stray inductance in the network to a minimum (Figure 4) allowing the extra section to compensate for tail end droop. If the construction of the PFN is reviewed in Figure 4, the advantages of utilizing stripline construction is easily understood. The stripline produces a low inductance connection between capacitors and reduces the stray inductance normally present in the intersection connections.

![Figure 2. Schematic of 250 kV modulator, showing the resonant charge system, PFN and some of the protection circuitry designed into the modulator.](image)

- $C_{1-2} = 53 \text{ nF} \pm 5\%$
- $L_1 = 25 \text{ nH}$
- $L_2 = 137 \text{ nH}$
- $L_3 = 151 \text{ nH}$
- $L_4 = 167 \text{ nH}$
- $L_5 = 167 \text{ nH}$
- $L_6 = 167 \text{ nH}$
- $L_7 = 167 \text{ nH}$
- $L_8 = 280 \text{ nH}$
- $E_{SL1} = 40 \text{ nH}$

$N_{1-4} = N_{10} = N_{15} = N_{20} = 10 \text{ nH}$

![Figure 3. Modified type-E PFN showing the approximate values of the section inductances.](image)

![Figure 4. An outline drawing of the PFN showing the low inductance stripline design of the PFN.](image)
Each section of the network consisted of a module housing seven of the 7.5 kF capacitors. Three mounting holes were provided in the modules. Coils constructed from 1/4" copper tubing were used as the inducitors, and mounted from these holes. Early inductance calculations using Henagan's inductance formulas and Terman's formulas showed that the inductances were not easily adjusted by varying the length of the coils [7,4]. A fairly unique solution was devised. Because the PPN was required to be tunable to compensate for load changes (different diode impedances), tunability was an important consideration. In addition to the low inductance provided by a strip-line design, the low inductance mounting of the capacitors close to the section inducers allowed an additional inducer to tune the first few sections of the PPN. The third inducer introduced into the network was mounted between the two coils which met the primary section inductance. To help in the review of the PPN outline the currents are shown in Figure 4.

The output of the PPN was stepped up by a 15:1 step-up transformer. In addition the iron core transformer provided matching between the 1.8 kV PPN and the 400 kV load. The transformer (manufactured by Stanganelli Industries) was designed to provide a 1.5, 250 kV pulse with less than 2.5% drop. To accommodate these specifications, the primary leakage inductance was approximately 100 mH. With the calculated output stray capacitance of 160 pF, the estimated risetime was 100 ns. The full 15 swing of the core was utilized by biasing the primary with a reset current. A 55 A current from a 5 V power supply resets the transformer before each pulse. Protection circuitry provided isolation between the high voltage on the primary and the low voltage side of the supply. In addition, a quadrilateral secondary winding was provided to supply power to a 2 kW heater winding biased at +2 kV with respect to the 250 kV output pulse.

Fault mode protection circuits designed into the modulator provided protection to the thyatron, the pulse transformer and the high voltage banking. No clamping circuits were used across the thyatron. However, a similar network was placed across the thyatron in the event of a fault mode (Figure 2). High voltage diodes along with series limiting resistors were placed across the PPN to prevent reversals from occurring on the PPN capacitors. Tests showed that the high voltage diodes were capable of conducting 10 kA currents and the 5 x 10^12 A/sec dI/dt's present. A metal oxide variactor stack consisting of 150 GE 1000 LA 40 kV's was used to provide a surge load to the output of the transformer at 300 kV. No loading of the output pulse was noted at lower voltages, by the stack. Although the stack under normal operating conditions does not load the pulse, the MOD stack has a finite lifetime under normal operating conditions. The calculated lifetime of the stack was approximately 1 x 10^12 shots. For long lifetime modulators this could conceivably be a problem. A design goal for the modulator was 1 x 10^15 shots without maintenance. The estimated lifetime of the modulator is 1 x 10^12 - 1 x 10^15 shots under repeated burst mode operation (excluding the MOD stack).

The circuit of Figure 3 was designed into four stages all modular in design. The modulator was housed in an oil tank mounted atop an oil reservoir (Figure 5). Associated controls were designed into individual mounted control cabinets. The power supply was provided in a beet sealed oil tank totally self-contained. The low inductance mounting and the layout of the modulator's internal components are shown in Figure 5. The oil tank, oil reservoir, rack mounted controls and the modulator frame are shown in the photograph in Figure 6.

Figure 5. A cross section drawing of the modulator showing the low inductance connections.

Figure 6. Photograph of the modulator and associated support hardware. Showing the modulator (top) supported above the oil tank and reservoir. The control system is shown to the right.

**COMMAND, CONTROL AND TRIGGERING SYSTEM**

The control and trigger system are housed primarily in a 19" rack. The system is divided into the system power control chassis, the 30 kW, 20 kV supply control chassis, the 30 kW, 40 kV supply chassis, the thyatron master/reservoir chassis, the system interconnect/junction chassis and the master control chassis (Figure 7). Additional elements are located in NFI boxes located on the modulator tank (Figure 5). The system incorporated proven design techniques for noise suppression and immunity in pulsed power environments. The system was designed to allow integration into a larger system with a safety/interlock system.
Figure 7. Block diagram showing the interconnections of the control system.

The control system performs the following major functions:
- Monitor and display of all safety interlocks.
- Monitor and display of all support and subsystem interlocks.
- Control and remote programming of all high voltage supplies.
- Control and monitoring of all crowbars and clamps.
- Generation and sequencing of all low-level trigger signals.
- Control of all system parameters such as rep-rate, number of pulses, charge voltage, etc.
- Display of system status.
- Real-time display of measured PFP voltage.
- Fault detection and auto-ABORT control functions.
- Orderly shutdown of all systems upon auto-ABORT command.

Figure 8 shows the main control panel and the functions monitored by the controller.

Conditions of all interlocks, fault detection circuitry, and subsystems are displayed real-time using tri-color LED's. Green denotes that the interlock is connected. Red denotes that the interlock is not satisfied or broken, and yellow denotes that the interlock has been bypassed.

Selection of charge voltage, number of pulses and rep-rate is accomplished by the use of thumbwheel switches. Voltage and pulse resolution is 0.1 kV, 1 pulse, and 0.1 Hz respectively. Both internal and external triggering is available but rep-rate limited to 110 Hz maximum. A low jitter pulse arriving approximately 500 ns before system output is provided for timing. The front panel also displays the quantity of actual shots fired during a burst mode or single shot operation.

Three system modes are available: single-shot, burst, and slow-rap. The burst mode allows up to 99 pulses at up to 107 Hz using an internally generated or external pulse train. The slow-rap mode allows continuous operation at 1.2 Hz or less.

Real-time scaled digital display of measured PFP voltage is accomplished using a fast gated analog to digital converter/display unit triggered from a fire delay signal.

The system design approach incorporates features that limit single shot fault modes to energy/voltage levels within component specifications. Therefore, fault mode protection philosophy is based on prevention of subsequent pulses occurring were a repeat of the particular fault mode is likely. All fault modes detected are identified via front panel indicator displays. Fault levels are auto-scaled with requested PFP voltage, thus no external programming is required when operating at different voltages.

Auto-ABORT detection circuits for output high, (load open), output low (load shorted), PFP high, PFP low (thyatron latchup), varistor stack overcurrent and power supply faults are provided which are fast enough to prevent a subsequent shot from being fired at the maximum rep-rate of 100 pps. The front panel indicators identify which circuit caused the auto-ABORT. Automatic control of all crowbars, power supplies and trigger amplifier are provided.

The high voltage power supply fault detect circuit detects overcharging of the PFP network. Other power faults (such as overcurrent) will break the power supply interlock, and result in an orderly abort of the system.

Pulse forming network and thyatron faults are detected using a fast gated comparator circuit in conjunction with a high impedance PFP voltage monitor. By detection of an abnormal PFP voltage level, shot to shot faults such as thyatron latchup and are detected. This fault results in an immediate abort sequence, allowing quick shutdown of the modulator.

The output of the modulator is shunted with a varistor stack (Figure 3). This circuit element prevents tube flashover in the event of an open circuit load. A current shunt on the varistor stack and fast comparator circuit are used to detect this fault.

The modulator is provided with a fast resistive monitor. The output of this monitor is fed into a fast gated pulse integral detection circuit (not shown in Figure 7). Detection of an abnormally high integrated voltage signal indicates an open load.
fault. Detection of a low pulse integrated voltage signal indicates tube flashover or shorted fault conditions.

The gated thyatron trigger generator is triggered by a low level (±25 V) pulse from the master control unit. Because of the high rate of current rise in the PFN discharge circuit, a presynchronization pulse scheme was utilized to enhance thyatron performance. This pulse and the delayed trigger pulse are generated by solid-state circuitry incorporating over-repet and noise immunity elements. All grid bias voltages and required delays are generated internally. The thyatron reservoir was set to 6.65 V and the reservoir voltage set to 3.85 V. Although the specified levels given by EVV are 6.6 V and 6.5 V for the thyatron under normal operating conditions, the thyatron at full charge voltage latch-up was used when the factory settings were used. With the heater and reservoir settings used, the thyatron recovered well within the recharge period.

**System Tests**

Testing was performed into a resistive load. The thermionic diode indicated in Figure 2 was replaced by a 400-ohm fixed load. These step-amp resistors 1° in diameter, 127° long, oil integrator) were mounted on the output of the transformer. In addition the vacuum bushing was left connected to evaluate the impact of the total stray capacitance on the output of the transformer. Along with the matched load tests, short circuit and open circuit tests were performed. During these tests, fault mode interrupts in the main control system were bypassed, to allow rep-rate fault mode testing. In the latter an in-situ voltage monitor was used along with a Pearson Probe (110 A) to evaluate the systems voltage and current performance.

Figure 9 shows the superposition of 23 output voltage pulses taken at 1 pps rep-rate. Specifically, these were shots 15, 12, 22, 5, 2 in a 3600 shot sequence. The PFN charge voltage was 25 kV which produces a 200 kV output pulse into a matched load. Pulses-to-pulses voltage variation is influenced by the trickle charge supply which has sufficient time to provide some regulation at rep-rates of 1 pps or less.

![Figure 9. Superposition of 23 output voltage pulses taken at 1 pps. PFN charge voltage = 25 kV. Scales are 46 kV/div and 200 ns/div.](image)

The voltage and current are shown in Figure 10, into a 400-ohm load. The photograph is an overlay of 50 shots during a burst mode test. The peak current was approximately 400 A peak. In addition, the peak output voltage was approximately 280 V, 13% during the 1 us pulse (PFN charge voltage 11 kV, 2 kA). Note the slight difference in the current and voltage waveshapes. Because the output of the circuit was capacitance-buffered rather than inductance limited, some slow oscillation or droop was seen on the voltage waveform. The current was flat to better than ±5%. Notice also that some amplitude variation is present on the pulses. This was a result of the power supply voltage variation and not a result of the discharge circuit.

![Figure 10. Voltage and current measurements through a 400-ohm load. The top trace is the current through the load, 200 ns/div; 200 A/div. The bottom trace is the voltage across the load, 200 ns/div, 42.5 kV/div, under 100 pps 50-pulse burst mode operation.](image)

The modulator was also tested under open circuit and short circuit fault modes. The resistive load wasshorted out, and the modulator was fired several times into the short (Figure 11). The peak current was 1300 A versus the calculated current of 1275 A. The resulting thyatron current was 19.5 kA. No degradation in the modulators' performance was noted, although approximately 100 shorted shots were fired. The open circuit tests tested the validity of using high voltage MOV stacks as protection devices during open circuit shots. The secondary was open circuited by disconnecting the load. Figure 12 shows an initial transient of 460 kV present for approximately 150 ns before the MOV stack clamps the voltage to 300 kV (PFN charge voltage 35 kV). No degradation in the performance of the modulator was noted during these tests, indicating the feasibility of using passive protection schemes. In addition, because the high voltage transient only lasted for several hundred ns, no vacuum insulator breakdown was noted during the tests.

![Figure 11. Short circuit current on the secondary of the transformer, 200 ns/div, 400 A/div (33 kV PFN charge voltage).](image)
Figure 12. Voltage across the secondary of the transformer during an open circuit shot. During the tests the Vpp = 35 kV. The initial transient is a result of the HV turn-on time, 2 us/div. 115 kV/div.

SUMMARY

In summary, the modulator and control system met all of the design specifications. In addition, the minimum output voltage of 250 kV was exceeded by 15 kV (Vpp = 33 kV) and could approach 280 kV (Vpp = 35 kV). The risetime of the output voltage pulse was 220 ns (10-90) compared to the calculated 200 ns risetime. This was a result of stray capacitance and a tradeoff in the tunability of the PPM. Moreover, the folded strip-line design of the PPM provided tunability output of the modulator desirable to a general purpose test stand. The control system performed as designed, allowing for complete automatic control and fault mode protection.

ACKNOWLEDGMENTS

Phil Chapmey along with Jim Fockler and Reins Lackner provided many valuable suggestions which were gratefully appreciated. In addition, Reins Lackner worked long hours in support of the program and should be commended. Pulse Sciences would also like to acknowledge the help and advice provided by Bu Memon and the staff at IEV.

REFERENCES


APPENDIX B

BASIC VERIFICATION PROGRAM
APPENDIX B

The computer program used to verify Eq. 3.4 and the modulator design analysis is listed in Fig. B.1. All input values are on the secondary of the modulator. The theoretical best case or matched impedance voltage, $V_{\text{pulse}}$ is input in volts. The calculated maximum and minimum electron beam impedance $R_{\text{LOAD \ min}}$ and $R_{\text{LOAD \ max}}$ are entered in ohms. ThePFN impedance is referred to the secondary of the modulator. The number of points which are to be calculated between $R_{\text{LOAD \ min}}$ and $R_{\text{LOAD \ max}}$, is also entered upon running the program.

The "BASIC" computer program then tabulates the following:

RLOAD: The load impedance at each calculation point.

RLOADP: The parallel impedance combination of the electron beam load (RLOAD) and the parallel matching resistor.

GAMMA: The reflection coefficient $\Gamma_1$ at the load.

VLOAD: The load voltage based on the input pulse voltage $V_{\text{pulse}}$, and GAMMA the tabulated reflection coefficient at the load.
ILOAD: The current into the "load" or the electron beam current.

IPAR: The current into the parallel resistor.

EFF: The calculated efficiency of the modulator as given by Eq. 3.7.

The program is written in BASIC. It has been run under both IBM BASIC-A and Microsoft QBASIC (DOS V5.0).
Listing of the basic program used to verify the modulator operating ranges.

```
10 'THIS PROGRAM CALCULATES THE VOLTAGE ON
20 'LOAD USING REFLECT. COEFF'S AT EACH IMPEDANCE
30'
35 DEFINT X,N
40 DEFSNG E,G,I,R,V
50 DIM RLOAD!(20)
60 DIM RLOADP!(20)
70 DIM GAMMA!(20)
80 DIM VLOAD!(20)
90 DIM ILOAD!(20)
95 DIM IPAR!(20)
100 DIM EFF!(20)
105 DIM RP1!(20)
106 DIM RP!(20)
110 CLS
120 LINE INPUT "THE NAME OF THE FILE";FILENAME$
130 INPUT "WHAT IS THE PULSE VOLTAGE";VPULSE!
140 INPUT "WHAT IS THE MINIMUM LOAD RESISTANCE";RLOADMIN!
150 INPUT "WHAT IS THE MAXIMUM LOAD RESISTANCE";RLOADMAX!
155 INPUT "WHAT IS THE PFN'S RESISTANCE";RPFN!
156 INPUT "WHAT IS THE PARALLEL RESISTANCE";RPAR!
160 INPUT "WHAT IS THE NUMBER OF POINTS";N
167 LPRINT "..:,... ".
185 N=N+1
190 FOR X=2 TO N
195 S=X-1
200 RLOAD!(X)=RLOAD!(S)+DEL!
210 NEXT
220 FOR X=1 TO N
230 RLOADP!(X)=(RPAR! * RLOAD!(X))/(RPAR! + RLOAD!(X))
240 GAMMA!(X)=(RLOADP!(X)-RPFN!)/(RLOADP!(X)+RPFN!)
250 VLOAD!(X)=(1-ABS(GAMMA!(X)))*VPULSE!
260 ILOAD!(X)=VLOAD!(X)/RLOAD!(X)
270 IPAR!(X)=VLOAD!(X)/RPAR!
280 RP1!(X)=(ILOAD!(X)^2)*RLOAD!(X)
290 RP!(X)=((IPAR!(X)^2)*RPAR!)+((ILOAD!(X)^2)*RLOAD!(X))
300 EFF!(X)=RP1!(X)/RP!(X)
310 NEXT
330 LPRINT ,FILENAME$
340'
350'
360'
370'
380 LPRINT ",VPULSE =";VPULSE!
390 LPRINT ",RLOAD MIN=";RLOADMIN!
400 LPRINT ",RLOAD MAX=";RLOADMAX!
410 LPRINT ",RPFN=";RPFN!
420 LPRINT ",PARALLEL RESISTANCE,RPAR=";RPAR!
430 LPRINT ",NO. OF POINTS=";N
440 LPRINT "RLOAD" TAB(25) "RLOADP" TAB(45) "GAMMA"
```
450 FOR X = 1 TO N
460 LPRINT RLOAD!(X) TAB(25) RLOADP!(X) TAB(45) GAMMA!(X)
470 NEXT
480 ,
490 ,
500 ,
510 ,
515 LPRINT "VLOAD" TAB(25) "ILOAD" TAB(45) "IPAR" TAB(64), "EFF"
520 FOR X = 1 TO N
530 LPRINT VLOAD!(X) TAB(25) ILOAD!(X) TAB(45) IPAR!(X)
       TAB(64) EFF!(X)
540 NEXT
560 END
APPENDIX C

MODULATOR CIRCUIT SIMULATION MODEL
APPENDIX C

The circuit model program used to model the modulator's response into a pulsed current source is listed in Fig. C.1. All circuit elements are on the primary of the modulator. The capacitances C1-C6 and L1-L5 model the pulse forming network. The LPL PFN inductor includes the stray circuit inductance and the output PFN stage inductance.

To simulate the thyatron switch closure, a switch model has been included. The charged PFN is switched into the constant current load by GS. The voltage controlled current source is controlled by the voltage source VG5 into the resistive load RGS.

The output pulse transformer of the modulator is modeled by LT1, LT2, LP, and CST. The leakage inductance LT1 and LT2 are symmetrically distributed about the self inductance LP of the transformer.

The stray capacitance across the output of the pulse transformer is modeled by CST. Additional stray circuit capacitance from cables and the large cathode area are represented by CSL.

The pulsed current source into which the modulator must operate is given by ILO. The pulsed current source is "timed" to the risetime of the modulator's output pulse. A piecewise (PWL) linear table was used to specify ILO's pulsed current, timing and peak current value. The peak ILO value of 1.61 has been normalized to the circuit model parameters.
The parallel matching present across the electron beam current ILO is modeled by RPL.

The remaining simulation options control the plot and print functions. When this model is run under Intusoft's SPICE (PC version) simulation code, the load voltage and current are plotted in graphical form.*

*Intusoft, P.O. Box 6607, San Pedro, CA 90734-6607.
Figure C.1

Listing of modulator circuit simulation code.

MODEL OF MODULATOR INTO A CONSTANT CURRENT LOAD
D1 1 2 DMOD
R1 1 0 5.5
* MODEL OF PFN
C1 2 0 .734UF IC=28
L1 2 4 16.6UH
C2 4 0 .734UF IC=28
L2 4 6 23.9UH
C3 6 7 0 .734UF IC=28
L3 6 8 27.5UH
C4 8 10 0 .734UF IC=28
L4 8 12 28UH
C5 10 12 0 .734UF IC=28
L5 10 12 0 29.1UH
C6 12 14 0 .734UF IC=28
LPL 12 14 25UH
* END PFN MODEL
RBS 14 16 1E06
GS 14 16 POLY(2) 14 16 120 0 0 0 0 .0 1.0 IC= 1E-03,1000
VGS 120 0 PWL(0.0 0.5NS 0.1NS 10NS 1000)
RGS 120 0 10
* TRANSFORMER MODEL
LT1 16 18 5.5UH
LT2 18 20 5.5UH
LP 18 0 18.7MH
* END TRANSFORMER MODEL
CST 22 0 126.5NF
VC2 22 24 0.0
VC1 22 24 0.0
CSL 22 0 0.65UH
* MODEL OF CONSTANT CURRENT LOAD
ILO 24 0 PWL(0.0 0.0 3.5US 0.0 5.5US 1.61 53.5US 1.61 58US .1)
* END MODEL OF CONSTANT CURRENT LOAD
RPL 24 0 12.6
.MODEL DMOD D
.TRAN .5US 100US UIC
.PLOT TRAN V(24)
.PLOT TRAN V(2)
.PLOT TRAN I(VC1)
.END
APPENDIX D

GRID-PULSER CIRCUIT SIMULATION MODEL
APPENDIX D

The SPICE circuit model used to simulate the grid pulser's response into the space-charge limited grid-cathode load is shown in Fig. D.1. All circuit elements are on the secondary of the grid-pulser.

Capacitances C1-C15 model the PFN capacitances. The ESR of each capacitor was also included in the simulation. The capacitors ESR is modeled with RC1-RC15. The PFN inductors are represented by LC1-LC15. Inductor LC15 includes the transformer leakage inductance and the coaxial cable inductance between the grid-pulser and the cathode.

The Child-Langmuir load which is given by Eq. 4.17 is modeled by GRL, a voltage controlled current source. The parameters shown for GRL were determined by curve fitting Eq. 4.17 to a polynomial in the voltage range of interest, or 0-550 V in this case.

The Miller capacitance of the grid-to-cathode region is modeled by a voltage dependent capacitance, CL2. Equations 4.18 to 4.19 were "curve-fitted" to polynomials in the region from 0-550 V. The polynomial coefficients were then used to specify the voltage-dependent Miller capacitance CL2.

The parallel resistor integral to the grid-pulser was included in the model. The 21 Ω resistor is represented by RP.
The remaining simulation options control the plot and print functions. When this model is run under Intusoft's PC-SPICE version, the load voltage and current are plotted in graphical form.*

*Intusoft, P.O. Box 6607, San Pedro, CA 90734.
Figure D.1
Listing of the grid-pulser circuit simulation code.

MODEL OF GRID PULSER
*PFN MODEL
C1 2 0 .386UF IC=380
RGB3 2 0 100000
RC1 2 4 .0372
LC1 4 6 7.86UH
C2 6 0 .293UF IC=380
RC2 6 8 .0372
LC2 8 10 8.21UH
C3 10 0 .293UF IC=380
RC3 10 12 .0372
LC3 12 14 8.57UH
C4 14 0 .293UF IC=380
RC4 14 16 .0372
LC4 16 18 9.0UH
C5 18 0 .293UF IC=380
RC5 18 20 .0372
LC5 20 22 9.3UH
C6 22 0 .293UF IC=380
RC6 22 24 .037
LC6 24 26 9.8UH
C7 26 0 .293UF IC=380
RC7 26 28 .037
LC7 28 30 9.8UH
C8 30 0 .293UF IC=380
RC8 30 32 .037
LC8 32 34 9.8UH
C9 34 0 .293UF IC=380
RC9 34 38 .037
LC9 38 40 9.8UH
C10 40 0 .293UF IC=380
RC10 40 42 .037
LC10 42 44 9.8UH
C11 44 0 .293UF IC=380
RC11 44 46 .037
LC11 46 48 9.8UH
C12 48 0 .293UF IC=380
RC12 48 50 .037
LC12 50 52 9.8UH
C13 52 0 .293UF IC=380
RC13 52 54 .037
LC13 54 56 9.8UH
C14 56 0 .293UF IC=380
RC14 56 58 .037
LC14 58 60 9.8UH
C15 60 0 .293UF IC=380
RC15 60 62 .037
LC15 62 64 9.8UH
* END PFN MODEL
* MODEL OF CHILD-LANGMUIR LOAD
GRL 64 66 64 66 1E-04 9.103E-02 4.917E-04 -2.194658E-07
IC=.0001,200
VC2 66 0 0.0
* END MODEL OF CHILD LANGMUIR LOAD
* MODEL OF MILLER CAPACITANCE
CL2 64 0 POLY 2.302E-09 2.11E-10 -9.623E-13 1.899E-15
   IC=.0001,200
* END MODEL OF MILLER CAPACITANCE
RP 64 0 21
.TRAN .5US 56US UIC
.PLOT TRAN V(64)
.PLOT TRAN I(VC2)
.END
APPENDIX E

FILAMENT CARBURIZATION AND ACTIVATION PROCEDURES
APPENDIX E

INTRODUCTION

Both carburization and activation procedures were developed during the experimental phase of the program. During the course of the investigation a set of activated thoriated tungsten filaments were stripped of their emission surface due to pin-hole leaks which developed in a foil window under test.

Attempts to reactivate the filaments failed. Based on similar results by Jenkins [1], it is believed that oxygen contaminants "poisoned" the low work function thorium surface. In order to reactivate the 132, 3.81 x 10^{-2} cm (15 mil) diameter cathode filaments, the surface must be recarburized. This process deposits a tungsten carbide (W\textsubscript{2}C) surface onto the filament prior to precipitation of thorium out of the matrix and onto the surface [2]. The W\textsubscript{2}C surface slows the evaporation rate of the thorium as it deposits out onto the filament surface. This allows a monolayer of thorium to build up on the surface of the filament [1, 2].

The foil window was replaced prior to carburization and reactivation was done with a water cooled dummy anode. Quartz, optically clear, windows mounted in the dummy anode, allowed the filament surface temperature to be measured using pyrometry. Based on these pyrometric temperature curves, the filament surface temperature as a function of the filament’s current and voltage was established. The filament temperature as a function of voltage and current are shown in Table E.1. For these
tests the 1 kHz heater supply was replaced with a DC, 0-40 V, 700 A welding supply.

CARBURIZATION PROCEDURE

The carburization procedure developed was based on the work of Jenkins and Schneider [1, 2]. The carburization procedure as follows, utilizes propane as the carburization gas. Both "hardware store" grade propane and laboratory grade propane have been tried with reasonable filament lifetimes of 1000-2000 hours. Surface area of each filament is approximately 1.1 cm$^2$.

The 2500 cm$^2$ cathode is slowly pre-heated over a 10-15 minute interval to 2300 K (2027°C). This corresponds to a filament current of 480 A. This initial pre-heating phase prevents mechanical warping of the cathode due to thermal stress. The cathode is left at 2300 K for 10 minutes to allow the mechanical components to reach thermal equilibrium. The power supply voltage (or the filament current) is then reduced to zero in about 20 seconds. This prevents rapid cooling of the cathode. Vacuum during this phase is maintained at $1-2 \times 10^{-6}$ Torr. The vacuum pump is then isolated from the vacuum system by closing a gate valve. Propane is then injected into the cathodes vacuum chamber bringing the cathode background pressure to 2 Torr.

A mechanical roughing pump is then started: The background propane pressure is then reduced to between 1000-500 millitorr. At this pressure the temperature of the cathode is brought back up to 2300 K (2027°C). This
allows the carburization of the filament surfaces to proceed. The cathode is maintained at a temperature of 2300 K for 3 minutes, 30 seconds. The roughing pump has pumped the background pressure down to between 50-100 millitorr in this time period. At this time the gate valve to the cryogenic vacuum pump is opened and the chamber pressure is reduced to 1-2 x 10^-6.

The cathode voltage and current is slowly reduced to zero over 20 minutes, and the cathode is allowed to cool unless activation of the filaments is desired.

ACTIVATION PROCEDURE

The activation procedure is also based on the work of Jenkins and Schneider [1, 2]. Following carburization, the 2% thoriated tungsten filaments are activated, to reduce the work function of the filament to 2.63 eV [1].

The cathode temperature is slowly raised from room temperature to 2073 K (1800°C, 410 A, 132 filaments) over a 10 minute interval to prevent mechanical "warping". The cathode is left at this temperature for 10-20 minutes to allow components to come to thermal equilibrium. The vacuum pressure is maintained at or below 5 x 10^-6 Torr during the activation process. This prevents inadvertent poisoning of the filament surfaces.

To "flash" the filaments, the filaments temperature is raised to 2500°-2600° K (2227-2327°C, 600-660A), in about 20 seconds. The cathode is maintained at this temperature for 60 seconds. This step reduces the ThO₂ in the filament to Th and O₂.
After 60 seconds at 2500°-2600° K, the filament temperature is reduced to 2100° K (1827°C, 450 A). The filaments are held at this temperature for 20-30 minutes to precipitate the thorium (TH) out of the tungsten matrix. This allows a thin monolayer of thorium atoms to deposit out onto the filament surface.

Following this activation procedure the cathode temperature is reduced to 1800°-1900° K for cathode operation.

REFERENCES


## TABLE E.1

Filament Temperature vs. Welder (Power Supply) Voltage and Current

<table>
<thead>
<tr>
<th>Filament Temperature</th>
<th>I\text{welder}</th>
<th>V\text{welder}</th>
</tr>
</thead>
<tbody>
<tr>
<td>1250°C</td>
<td>250 A</td>
<td>6.5 V</td>
</tr>
<tr>
<td>1330°C</td>
<td>250 A</td>
<td>8 V</td>
</tr>
<tr>
<td>1400°C</td>
<td>300 A</td>
<td>11 V</td>
</tr>
<tr>
<td>1480°C</td>
<td>330 A</td>
<td>14 V</td>
</tr>
<tr>
<td>1590°C</td>
<td>350 A</td>
<td>15.5 V</td>
</tr>
<tr>
<td>1625°C</td>
<td>380 A</td>
<td>17 V</td>
</tr>
<tr>
<td>1800°C</td>
<td>410 A</td>
<td>21 V</td>
</tr>
<tr>
<td>1840°C</td>
<td>450 A</td>
<td>23 V</td>
</tr>
<tr>
<td>2250°C</td>
<td>570-600 A</td>
<td>30 V</td>
</tr>
<tr>
<td>2300°C</td>
<td>~660 A</td>
<td>33 V</td>
</tr>
</tbody>
</table>