NASA CR-54216

An Experimental Model of a 2KW, 2500 Volt Power Converter for Ion Thrustors Using Gate Controlled Switches in Two Phase-Shifted Parallel Inverters

by

F. A. Elder
L. E. Staley

prepared for

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

Contract NAS 3-5917

WESTINGHOUSE ELECTRIC CORPORATION
AEROSPACE ELECTRICAL DIVISION
LIMA, OHIO

Copies of this report can be obtained from:

National Aeronautics and Space Administration
Office of Scientific and Technical Information
Washington, D.C., 20546
Attention: AFSS-A

NOTICE

SUMMARY REPORT

An Experimental Model of a 2KW, 2500 Volt Power Converter for Ion Thrustors Using Gate Controlled Switches in Two Phase-Shifted Parallel Inverters

by

F. A. Elder

L. E. Staley

prepared for

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

May 17, 1965

Contract NAS3-5917

Technical Management
NASA Lewis Research Center
Cleveland, Ohio
Spacecraft Technology Division
Bernard L. Sater

WESTINGHOUSE ELECTRIC CORPORATION
AEROSPACE ELECTRICAL DIVISION
Box 989,
Lima, Ohio 45801

(UNCLASSIFIED)

May 17, 1965

An Experimental Model of a 2KW, 2500 Volt Power Converter
for Ion Thrustors Using Gate Controlled Switches
in Two Phase-Shifted Parallel Inverters

Prepared by:

F. A. Elder

L. E. Staley

Approved by:

R. P. Putkovich

H. B. James

(UNCLASSIFIED)

PREFACE

The technical work described in this report was performed at the Westinghouse Research Laboratories by the Electrical Systems and Power Conditioning Group. The authors wish to acknowledge the technical assistance offered by Messrs. W. H. Beck, J. C. Engel and Dr. P. F. Pittman and the technical guidance of Mr. R. P. Putkovich. The support and cooperation of Messrs. J. Motto, R. Prunty and L. Rice of the Westinghouse Semoconductor Division was particularly helpful during the investigation of GCS operating characteristics. The assistance and guidance provided by Mr. G. W. Ernsberger of the Westinghouse Aerospace Electrical Division and Mr. B. L. Sater of the NASA Lewis Research Center was especially appreciated.

An Experimental Model of a 2KW, 2500 Volt Power Converter for Ion Thrustors Using Gate Controlled Switches in Two Phase-Shifted Parallel Inverters

by

F. A. Elder

L. E. Staley

ABSTRACT

28950

The development of an ion thrustor power supply utilizing gate controlled switches (GCS) is described. A GCS inverter is used to convert 160 V.D.C. to 2500 V.D.C. Output voltage regulation is accomplished by varying the phase angle between two inverter stages. Overall efficiency at the rated output of 2 KW is 82.6%. Control circuitsare incorporated in the system to prevent failure due to overloads and to provide automatic recycling. The total weight of the system components is 44.5 pounds. Recommendations for achieving a system weight of 24 pounds and an efficiency approaching 90% are made. The results of this program show that the GCS is a useful switching device for power conversion applications.

author

TABLE OF CONTENTS

TABLE OF CONTENTS

TABLE OF CONTENTS

SECTION PAGE

LIST OF FIGURES

LIST OF FIGURES
(Continued)

LIST OF TABLES

An Experimental Model of a 2KW, 2500 Volt Power Converter for Ion Thrustors Using Gate Controlled Switches in Two Phase-Shifted Parallel Inverters

by

F. A. Elder

L. E. Staley

Westinghouse Electric Corporation

SUMMARY

This document is the Summary Report of the work performed during Phase I of Contract NAS 3-5917. The purpose of this work was to evaluate the merits of the new Gate Controlled Switch (GCS) in a low weight high efficiency power converter for Ion Thrustors. Phase I included the design, fabrication, and extensive tests of two models of the converter. The Breadboard Model proved feasibility of the proposed concepts, and the Experimental Model incorporated the final design into a suitable package for delivery to the customer. The criteria for the final design included high efficiency and low weight compatible with the overall system performance. Extensive tests were performed on the Experimental Model to determine its compliance with the work statement requirements on safety factors and reliability, both at full rated load and during transient short circuit loads.

In the Experimental Model a full load efficiency of 82.6% was obtained with a total circuit component weight of 44.5 pounds. The effiency is lower and the weight is somewhat higher than expected; however further development could yield an efficiency of about 90% and a total weight of 24 pounds, using the present work statement. Further reductions in weight to approximate 12 pounds with a slight decrease in efficiency could be achieved by eliminating the input filter, the overcurrent time delay, and increasing the frequency.

The Experimental Model proved its reliability by successfully surviving the application of approximately 2000 short circuit applications, and over 200,000 "Blink-Off" "Soft-On" cycles during complete testing. Component temperatures during any testing did not exceed safe limits nor did any degradation of the system occur.

I. INTRODUCTION

A 2KW, 2500 volt power converter employing gate controlled switches (GCS) in two parallel inverters was developed to fulfill the objectives of the contract work statement. These objectives were to employ the use of new technology, where practical, to achieve minimum system weight, high efficiency, and high reliability by applying appropriate safety factors. The purpose of this work was to advance the technology of power conditioning for Ion Thrustors. This includes providing a range of adjustments of operating characteristics to achieve optimum matching of the power converter to the Ion Thrustor. The Ion Thruster is characterized by frequent and severe electrical breakdowns throughout, which must be controlled and extinguished. The power converter: provides current regulation into the electrical breakdown; provides "Blink-Off" for an adjustable length of time to allow the breakdown to extinguish if the breakdown persists; and provides automatic restart in a "Soft-On" manner to reestablish operation without causing further electrical breakdown. Although operation with an Ion Thrustor was not part of this contract, the ability of the power converter to provide the above functions was demonstrated with a load simulator having electrical breakdown devices.

II. PRELIMINARY CONVERTER DESIGN CONSIDERATIONS

The preliminary converter design considerations were the choice of the power switching device; the choice of input voltage; the choice of the operating frequency; and the choice of the basic circuits. All decisions were based on two prime objectives. First, the overall objective was to accomplish the conversion of low voltage D.C. power to regulated high voltage D.C. power with a system having low weight and high efficiency. The other objective was to provide high reliability by applying adequate derating factors. For the experimental model a minimum safety factor of two for current and voltage and a minimum safety factor of four in allowable power dissipation for semiconductors was to be a design objective. With these two objectives in mind the four areas of choice were resolved as follows:

A. CHOICE OF GCS

It was decided in responding to NASA RFP-STD-213 that a GCS inverter system was capable of fulfilling the specified performance requirements. This decision was based on the generally superior characteristics of the GCS, when compared to other power switching devices. The comparison in Table II-1 shows the fast switching times of the GCS, which could be used to achieve good efficiency at high frequency and thereby yield a small system weight. The comparison also shows that because the voltage rating of the GCS is relatively high, a safety factor of 2 to 1 could be obtained without using devices connected in series. The high forward voltage drop of the GCS was known to be a disadvantage, but this could be compensated for by using the GCS at higher voltages and lower currents for a given power output. Table II-1 compares the numerical values of GCS characteristics in other power switching devices, and is part of the information on which the decision to use GCS's is based.

Table II-2 lists the parameters of Table II-1 in terms of advantages and disadvantages of the GCS and other power switching devices when

TABLE II-1 CHARACTERISTICS OF FOUR POWER SWITCHING DEVICES

	Germanium Power Transistors	Silicon Power Transistors	Silicon Controlled Rectifiers	Gate Controlled Switches
Forward Voltage Drop in Volts	0.1 to 0.2	0.5 to 1	1 to 1.5	3 to 4
Forward Blocking Voltage in Volts	50	120	1600	400
Reverse Blocking Voltage in Volts	5 to 10	10 to 20	1600	400
Switching Speeds in Microseconds	1 to 2	1 to 2	0.5 to 1	0.1 to 0.5
Rated Junction Temperature in Degrees Centigrade	90	150 to 200	125	125
Leakage Currents in Milliamperes Except *in Microamperes	80 to 100	5 to 15	*0.5 to 2	0.5 to 2
Degree of Control	Full Control with the Base Lead	Full Control with the Base Lead	Turn-on Control with the Gate Lead	Full Control with the Gate

TABLE II-2 ADVANTAGES AND DISADVANTAGES OF FOUR POWER SWITCHING DEVICES FOR ION THRUSTOR POWER CONVERTER APPLICATION

	Advantages	Disadvantages
1. Germanium Transistors	a. Low Forward Voltage Drop	a. Low Forward Blocking Voltage b. Low Rated Junction Temperature c. High Leakage Currents d. Requires Continuous Drive through the Base Lead to Maintain the On State
2. Silicon Transistors	a. Low Forward Voltage Drop b. Relatively High Forward Blocking Voltage c. High Rated Junction Temperature d. Good Switching Speeds	a. Moderate Leakage Current b. Required Continuous Drive throgh the Base Lead to Maintain the On State
3. Silicon Controlled Rectifier	a. Nominal Forward Voltage Drop b. Pulse Turn-On with Gate Lead c. Good Switching Speeds	a. Forced or Natural Commutation Required for Turn-Off
4. Gate Controlled Switch	a. High Forward Blocking Voltage b. Pulse Turn-On and Pulse Turn-Off with Gate Lead c. High Rated Junction Temperature c. Fast Switching Speeds	a. High Forward Voltage Drop when Applied to Low Voltage Circuits

these devices are applied to the proposed power supply. The final decision to use the GCS was based on the information contained in both tables.

It was expected that an experimental 25 ampere GCS would be available for this program; however, this device did not become available by the time the contract was awarded. It was then decided that the Westinghouse 242ZP, 10 amp rms, 700 volt, GCS should be used for the inverter system. A second alternative was the Motorola MGCS 821-6 or MGCS 925-6 devices, which are both 5 amp rms, 400 volt GCS's. The disadvantages in using these devices are apparent since they can handle only 500 watts of power compared to 1750 watts for the 242ZP device.

B. CHOICE OF INPUT VOLTAGE

After the GCS was decided upon, the input voltage was selected which would be compatible with the GCS. Because of the 2 Kw power requirement, the input current had to be balanced with the choice of input voltage, and both quantities had to have a 2 to 1 safety factor applied to them when compared with the current and voltage ratings of the GCS's. The proposed concept of two square-wave inverters, phase shifted with respect to each other and series addition of their output transformer secondary voltages, requires that each inverter handle 1 Kw of power. (See Figure II-1.) Taking into account the safety factor, the GCS is allowed 5 amps rms of current on a 100% duty cycle or 7 amps rms for a 50% duty cycle. Each half of an inverter carries the load current on alternate half-cycles so the D.C. input current to an inverter could be 7 amps. The voltage into each inverter would then be 1000/7 = 143 volts. To allow for the 3 to 4 volts GCS forward drop and other circuit losses an input voltage of 150 volts was selected. The two proposed bridge inverters could then use 10 amp rms, 300 volt, GCS's carrying 7 amps and blocking 150 volts on alternate half-cycles. The highest voltage input compatible

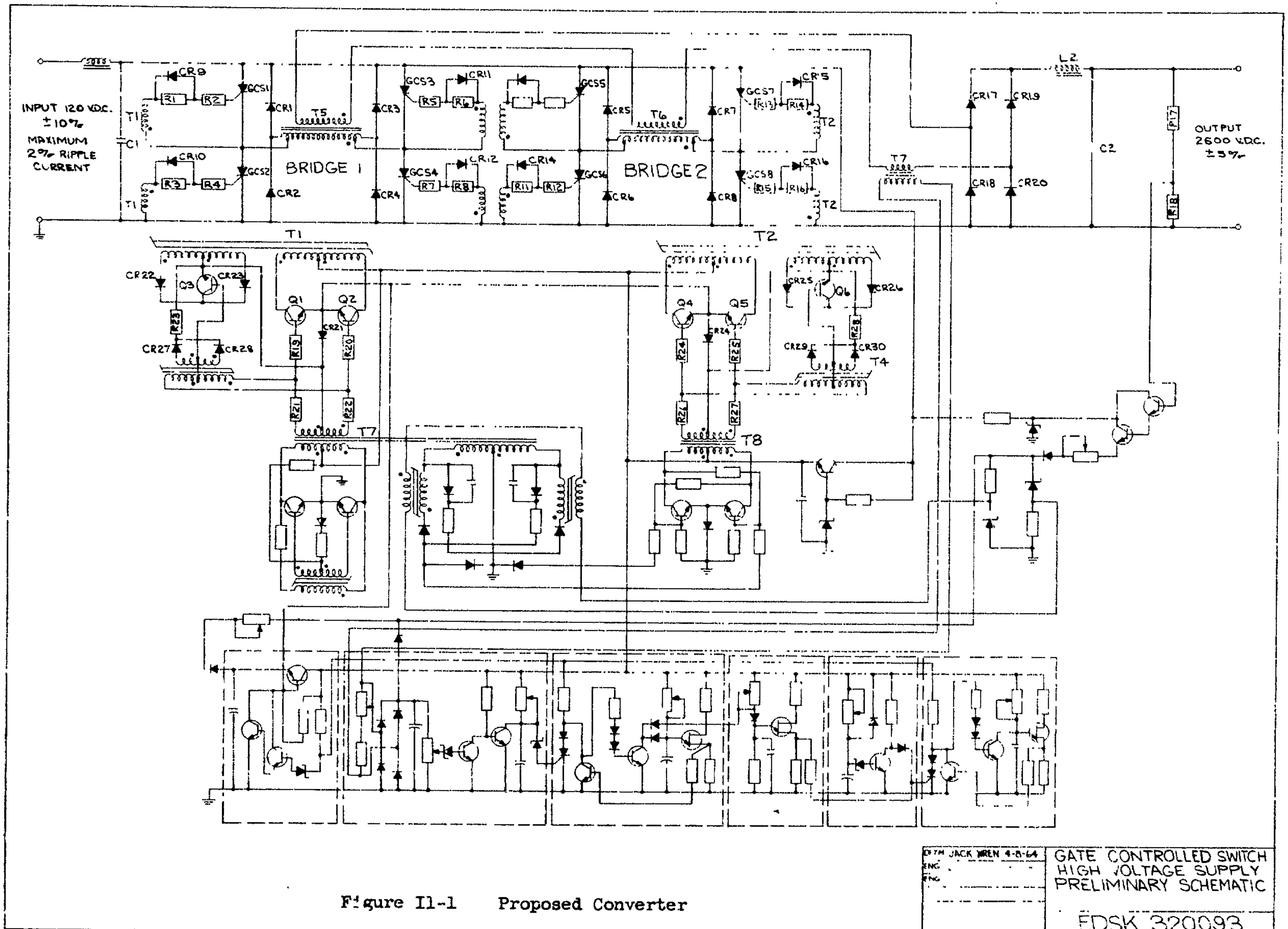


Figure II-1 Proposed Converter

with the bridge circuit and available GCS is 350 volts for a 700 volt device. The inverter current for this case need only be 3 amps to achieve 1 Kw per inverter of output power. Assuming 4 volts forward voltage drop in each GCS, the power lost due to GCS conduction dissipation in two bridges at 150 volts is 112 watts and in two bridges at 300 volts is 56 watts. This shows that for the bridge inverter the input voltage should be high and the current low to minimize losses.

The final choice of input voltage was determined after consideration of the basic inverter circuits and their relative advantages. The input voltage was finally chosen to be approximately 150 volts and the reasons for this are explained in a later section.

C. CHOICE OF FREQUENCY

The optimum operating frequency of the inverter depends on the switching times of the GCS. The switching times for the GCS had not been completely determined; therefore, it was not possible to make a meaningful frequency calculation early in the project. The switching times listed in Table II-1 were those observed in other types of circuits. Switching times of the GCS are very much dependent on a magnitude of current and voltage, and especially on the circuit impedance (i.e.,amount of circuit inductance). Final frequency selection would have to be done with actual circuit hardware and the measurement of actual GCS switching times in the circuit. Choosing of the operating frequency was done tentatively on the basis of past experience with semiconductors and similar circuits. An operating frequency of 1000 cps was chosen to take advantage of the magnetic component weight reduction due to high frequency and to minimizing the loss in efficiency due to the expected switching losses of the GCS.

The early circuit tests indicated that the GCS switching times varied from about 0.1 microsecond to several microseconds. Because of the variance and because of other difficulties encountered in applying the GCS, the calculation of optimum operating frequency was delayed. It was felt that more information about the GCS and its operation in the circuit was needed before accurate frequency calculations could be made. The experimental work to gather operating data on the GCS was conducted in an inverter system operating at 1000 cps.

D. CHOICE OF BASIC CIRCUITS

1. Inverter Circuit

The proposed circuit of two square-wave bridge inverters, phase-shifted with respect to one another for voltage regulation, was further analyzed in terms of the GCS actually available for use. This disclosed that two bridge inverters employing 8 GCS's would have higher losses at a given input voltage, would be more complex due to the greater number of devices and would require more complicated gate circuits. (The actual importance of the GCS gate circuits and their complication became apparent after some experience with them had been acquired.) An alternate to the bridge inverter is the center-tapped inverter which employs only two GCS's per inverter. The basic concept of two inverters with phase-shift is not changed by the center-tapped inverter; the output transformer secondary voltages are still square-waves which are phase-shifted for voltage regulation. In the center-tapped inverter only 4 GCS's are needed for the two inverters to provide 2 Kw of output; thereby reducing the complexity of the system. A center-tapped inverter handling 1 Kw and using 10 amp rms, 700 volt, GCS's would operate with a 150 volt input at 7 amps. The safety factor on voltage would be slightly greater than 2 to 1 since the maximum blocking voltage applied to the GCS would be 2 (150) or 300 volts. The current safety factor would also be 2 to 1 because the GCS conducts 7 amps for only one-half cycle, which is equivalent to 5 amps rms. Thus the maximum capabilities of the GCS (700 volts and 10 amps rms) are fully utilized and the 2 to 1 safety factor is provided.

The total losses for two center-tapped inverters employing 4 GCS's would be 56 watts compared to 112 watts for two bridge inverters (at 150 volts input and 4 volts forward drop per GCS). The savings of 56 watts amounts to 2.8% of the rated output power and is a step towards achieving maximum efficiency. Further savings are accomplished by the reduction of 8 separate gate circuits for 8 GCS's in two bridge converters, to 2 separate gate circuits for 4 GCS's in two center-tapped inverters. In the bridge circuit, separate and isolated drive circuits are required for the GCS. Because of the difference between turn-on and turn-off times of the GCS, the turn-on drive signal must be delayed in the bridge circuit to prevent a momentary short circuit during the switching period. In the two center-tapped inverters the common cathode connection of two GCS's permits simplification of the gate circuits. It is no longer necessary to delay the firing of the GCS to account for differences between turn-on and turn-off times. The transformer primary winding in series with the GCS offers sufficient inductance to prevent build up of current during the switching period overlap. Therefore, the drive circuits of the center-tapped inverter are fewer and simpler than those of the bridge inverter.

There is also a savings in the power requirements from the low voltage control; first because only 4 GCS's are driven, and second because the number of gate circuits are fewer and more simplified making them more efficient. The estimated savings in power is 44 watts or 2.2% of the rated output power.

A further advantage of 4 GCS's in two center-tapped inverters and their attendant simpler drive circuits, is that the system becomes more reliable. The fewer parts establish a longer probable system life.

The overall advantages of two center-tapped inverters over two bridge inverters in this application are: the fewer number of GCS's required; the lower circuit losses at a given input voltage, fewer, simpler, and more efficient gate circuits; and increased system reliability. For these reasons the use of the center-tapped inverter was employed in this application. Evaluation of the inverter system was based on application of the presently available Westinghouse 242Z GCS. Any improvements, changes, or use of different devices that become available may offer sufficient reason for reconsidering the bridge circuit or any other means of converting low voltage D.C. to regulated high voltage D.C. power.

2. Chopper

The proposed method of obtaining low voltage D.C. for the control circuits (See Figure II-1.) appeared obviously inadequate when the available GCS's required raising the input voltage to 150 volts D.C. The inadequacy was further compounded by the larger than expected power consumption of the control circuits. To obtain approximately 50 watts of control power at 20 volts would require 325 watts to be dissipated in the class A type transistor voltage regulator originally proposed. A chopper was selected as a more efficient means of obtaining the low voltage D.C. The chopper circuit selected is that shown in Figures IV-1 and App. B-1. This circuit produces a well-regulated, low-ripple output voltage of a value determined by the voltage of Zener diode CR15.

3. Master Oscillator

The original form of the master oscillator in the proposed circuit uses a saturating core to determine the frequency. This concept was retained in the final system. Other choices of basic circuits and the particular need for many forms of isolated power to supply them, prompted an increase in the power handling capability of the master oscillator. Since most of the circuits required isolation from each other and from the input terminals, it was decided to put all control power through the master oscillator and use isolated secondary windings to supply control circuits. The chopper then would not have to step down as far in voltage and the master oscillator could work at higher input voltages.

4. Schmitt Trigger

The main reason for choosing the Schmitt trigger for the phase shifting circuit was its speed of response. The original work statement called for regulation of the output current into a short circuit to a value which would not cause damage to any circuit components. In order to maintain safe limits on component currents it would be necessary to phase-shift from full in-phase to full out-of-phase as rapidly as possible. The Schmitt trigger was considered to offer an effective means of providing rapid phase-shift control.

5. Voltage Regulator and Overcurrent Regulator

Circuits for accomplishing regulation were to be investigated during the development program. The discussion of the development of these circuits points out some of the methods tried.

6. GCS Drive Circuit

The proposed GCS drive circuits were modified after the particular requirements and characteristics of the 242Z GCS were better understood. Reference is made to Section III of this report, where all GCS gate circuits examined during the development of this system are described.

III. CIRCUIT DEVELOPMENT

After the initial choice of basic circuits, the development of these circuits began. Work was directed towards perfecting the basic circuits or finding specific faults and substituting other circuits. The following discussions point out the highlights of the development of specific circuits.

A. INVERTER DEVELOPMENT

The major reason for investigating the use of GCS's in an inverter is that this new device offers a combination of the desirable features of both transistors and silicon controlled rectifiers. The GCS can be switched on and off and does not require a continuous drive current to maintain conduction. Unlike an SCR, the GCS does not require forced commutation, thus permitting elimination of commutating capacitors and inductors from the inverter circuit. During the early development for this project very successful operation of the GCS was being achieved. However, as more circuitry was added to the breadboard and as more severe operating conditions were tried, GCS failures began occurring. A thorough check of the power dissipation in the device and its operating currents and voltages as compared to its ratings showed reasonable margins. Further investigation disclosed a severe unbalance of inverter transformer magnetizing current in each half-cycle of operation. This unbalance was due to a D.C. component of voltage applied to the transformer which was causing it to operate near saturation. As the D.C. input voltage was increased, the transformer magnetizing current increased and finally caused failure of the GCS. The possible causes of the D.C. component of voltage were; unequal duration of half-cycles, unequal forward voltage drop of the GCS's, or unequal switching times of the GCS's. Similar problems had been recognized in previous transistor and SCR inverter circuits. The solutions used in earlier inverters were then tried in the GCS inverter

First a larger air gap was placed in the inverter transformer. This rounds the saturation knee of the hysteresis loop and increases the magnetizing current. Exhaustive inverter tests showed that the air gap had to be increased to the point where the magnetizing current plus full load current exceeded the rating of the GCS. The increased magnetizing current was also severely increasing the size of the output transformer because of the larger wire required. Different magnetic materials were also considered; especially those with a rounded magnetization curve. However, these materials would probably have failed to solve the problem for the same reason the increased air gap failed to work.

Another method of forcing a balance of the inverter half-cycle currents was to connect a low value resistor in series with the D.C. input to the inverter. By introducing a greater impedance in series with the transformer, the unbalance in current would be reduced and saturation would be avoided. This resistor helps to balance the volt-seconds applied to the core each half-cycle by producing a larger voltage drop when the current is increased during a half-cycle. Experimental tests of this method proved that it worked, but the value of resistance and the resulting power loss prohibited use of this method.

Consideration was given to the possibility of sensing half-cycle currents and controlling GCS half-cycle conduction times to maintain balance. This method was not tried because of the expected complexity of the necessary control circuits.

A solution was finally found and it is shown in the inverter circuit of Figure III-1. The added components are C1, L1, R1, R2, and CR3. The balancing of inverter currents by the addition of C1 and L1 is as follows: Assume that current I1 is greater than current I2 at the end of their respective half-cycles. When GCS1 is being turned off, the larger I1 flows through C1 and reverses the voltage of C1 in a relatively short time. This is explained by the following equation:

$$e = \frac{1}{C} \int i \, dt$$

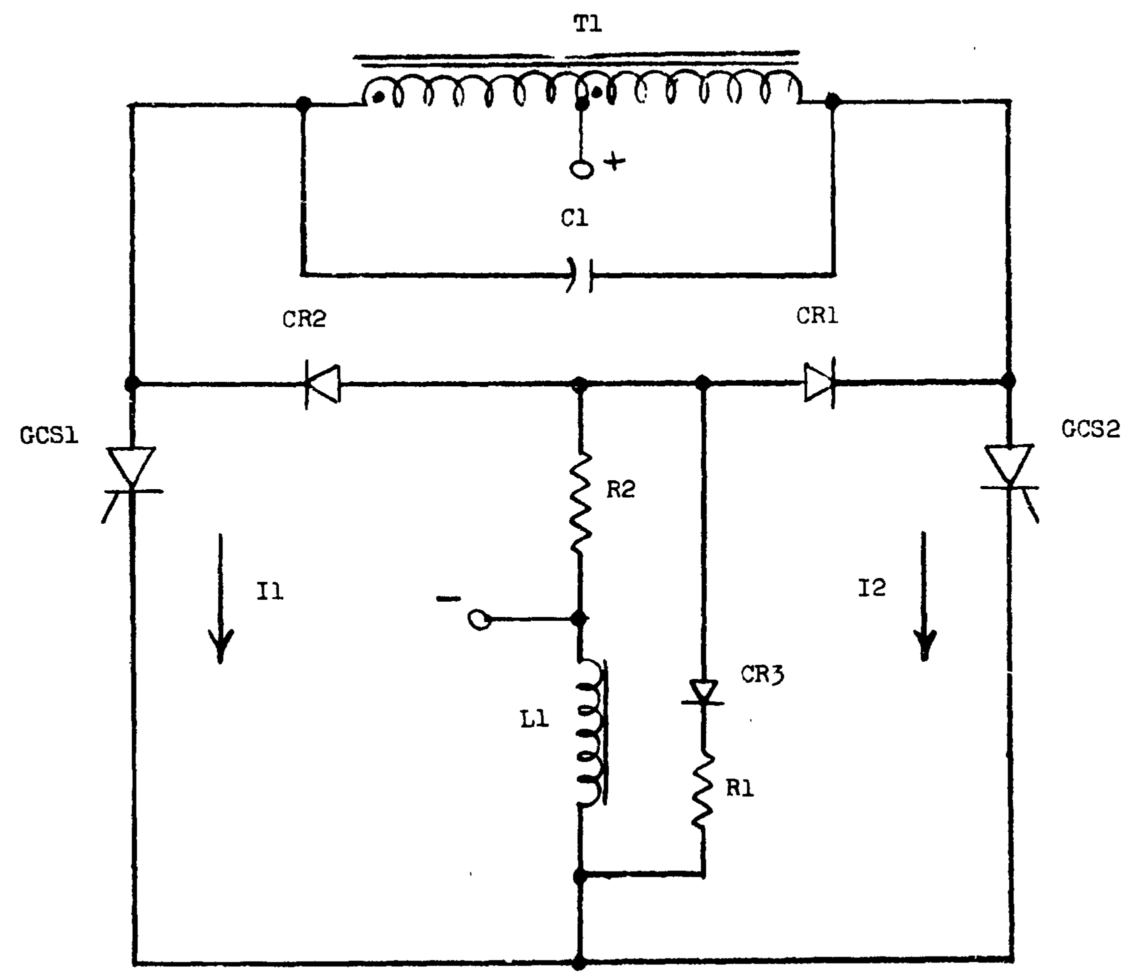


Figure III-1. Inverter Circuit Component Additions

If e and C are fixed, then the larger the value of i the smaller dt becomes. The voltage across the transformer T1 reaches its peak magnitude sooner and maintains it longer in this half-cycle because the large I1 reversed the voltage of C1 relatively fast. The volt-seconds of the present polarity applied to the transformer will therefore be larger and tend to increase the current I2. If, however, I2 remains small when it is time to turn-off CGS2, then the small current will take longer to reverse the charge on C1 (see equation above). It will take longer to reach the peak voltage of the opposite polarity in this new half-cycle. Therefore, at the end of this new half-cycle, the transformer will have a smaller magnetizing current. The current I1 is then smaller than the cycle before. The effect of C1 tends to equalize the two currents I1 and I2 by changing the half-cycle volt-seconds applied to the transformer T1. The inductor L1 is provided to present an impedance between the inverter and the source while switching is taking place, so that C1 can successfully perform its function. In addition, the impedance added by L1 also aids in balancing the half-cycle currents. The original experimental value of capacitance chosen was 0.5 microfarads and the original line choke was 5.0 millihenries. After further investigation the capacitance was reduced 0.05 microfarads. Because this capacitor was an important factor in determining transient conditions during GCS switching, considerable attention was given to the selection of an optimum value. Because the line inductor was less critical and time for further investigation was limited, equal effort was not devoted to optimizing the inductor. Further development may permit this inductor to be significantly reduced in size.

Use of the capacitor and line choke required the addition of other components to the inverter. The line choke stores energy and if the current is interrupted it generates large voltage spikes. Resistor R1 and diode CR3 were added to the circuit (see Figure III-1) to provide a low impedance path for line choke current to flow during the GCS switching period when both GCS's may be off. This prevents excessive voltage spikes across the GCS's when they are turning off. The charging of the capacitor through the line choke causes a build up of current which is undesirable. A method of preventing this build up is to dissipate the stored energy in resistor. This is accomplished by resistor R2 in Figure III-1. The dissipation of this resistor in this application is relatively high because when the two inverters are phase-shifted there is an interval when current is not supplied to the load. During this interval the current is maintained through the line choke and flows in the alternate path provided by R2.

In the first breadboard circuits the rate-of-rise of voltage (dV/dt) across the GCS was controlled by an RC circuit. This RC network also increased the turn-off capability of GCS circuit by momentarily shunting the current from the GCS as turn-off was occurring. The Westinghouse Semiconductor Division recommends that the diode-resistor-capacitor shunt path allow approximately 5 microseconds for the full blocking voltage of the device to be reached. Therefore C3 (See Figure III-2) must pass current for 5 microseconds and the voltage of C3 must not exceed 350 volts at the end of that interval. With a load current of 8 amps, then:

$$e = \frac{1}{C}\int i \, dt$$

and

$$C3 = 0.114 \text{ microfarads}$$

Allowing a safety factor of 2, the current would be 16 amps and C3 becomes approximately 0.25 microfarads. The voltage rating of this capacitor should be at least 600 WVDC. The resistor R6 limits the discharge current of C3 through GCS1 when GCS1 turns on. For a charging voltage of 350 volts on C3, the power rating of R6 should be

$$P = \frac{1}{2} C V^2 \times f$$

or

$$P \approx 15 \text{ watts}$$

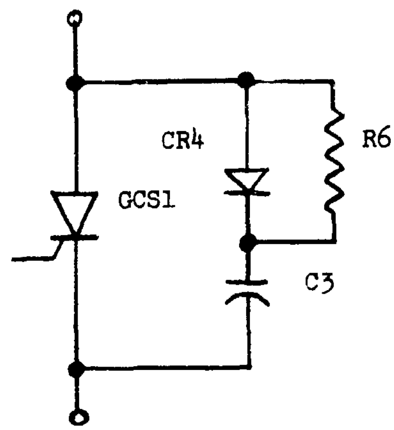


Figure III-2. GCS dV/dt Circuit

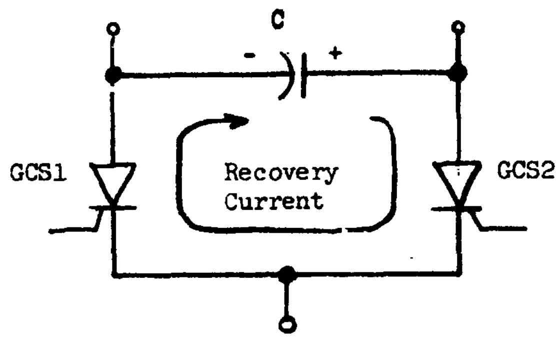


Figure III-3. Recovery Current

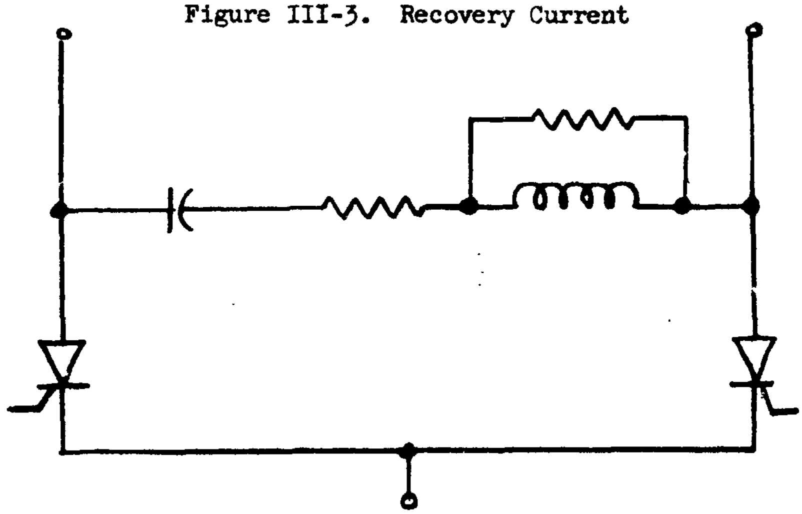


Figure III-4. Inverter dI/dt Circuit

The diode CR4 must handle peak repetitive currents of 8 amps for 5 to 10 microseconds once every cycle of the inverter frequency. It must have a blocking voltage of about 700 volts to allow a 2 to 1 safety factor. Since standard power diodes normally have long recovery times, the use of fast recovery diode is recommended to provide more reliable service. A 6-amp, 600-volt, 200 nanosecond maximum recovery time, Type 478M diode was originally used.

Later tests with this RC circuit showed that it introduced a serious problem in inverter operation. The capacitor C3 does not always charge to the full voltage as GCS1 turns off; then when GCS2 turns on, transformer action forces a rapid increase of the voltage on C3 which causes a large current spike to flow through GCS2. Because of the resonant charging effect produced by the leakage inductance of the transformer, the voltage on C3 builds up excessive voltage across GCS1. These circuits were removed and capacitor C1 in Figure III-1 was used to provide a controlled dV/dt for the GCS. The elimination of these circuits removed 60 watts of power dissipation from the two inverter circuits.

Capacitor C1, however produced a large reverse recovery current in the GCS. Figure III-3 illustrates that when GCS1 turns-off and GCS2 turns-on, the voltage on C1 is applied across GCS1. This causes a spike of current to flow through GCS2 to recover GCS1 to the blocking state. The first attempted solution was the addition of small saturating inductors in the anode of each GCS. These did not perform adequately and so were replaced with an inductor in series with the capacitor. The circuit used is that shown in Figure III-4. The inductor limits the rate of current build up (dI/dt) and two resistors prevent ringing to excessive voltages.

The installation of the final dI/dt circuits prompted the re-addition of the dV/dt circuits. The dV/dt protection afforded by the capacitor only, from anode to anode of the GCS's was no longer available. The component values of the new dV/dt circuits were experimentally deterimined to provide adequate protection for worst case operation. The final components are smaller in value and size than previously used and their losses are lower.

As pointed out in the above discussion, a number of problems were encountered during the development of the GCS inverter. The major problems were unbalanced transformer magnetizing currents and large dV/dt and dI/dt conditions. Solutions were found after careful investigation of operating conditions in the breadboard. The resulting breadboard circuit was capable of providing 1 Kw per inverter at a 1000 cps operating frequency.

B. CHOPPER DEVELOPMENT

The initial choice of the chopper circuitry proved entirely practical for this application. The only work required in this area was the final design of circuit components to cover the range of input voltage and to design the output filter components. The addition of one capacitor C40 to improve the response and the regulation of the chopper is shown in Figure App. B-1.

C. MASTER OSCILLATOR DEVELOPMENT

Development work on the master oscillator, after the initial choice of the basic circuit, is described in the experimental model electrical design under master oscillator improvements.

D. SCHMITT TRIGGER DEVELOPMENT

Various circuits were considered for providing the phase-shifted drive signals to the inverters. A Schmitt trigger circuit was selected because it could provide very rapid phase control. Fast response was considered necessary to maintain reliable inverter operation during the short circuit condition. If the output were shorted the inductor and capacitor of the high voltage filter would feed the load for a few milliseconds but to limit the inverter current it would be necessary to shift the inverters 180° out of phase within a one or two half-cycles. The Schmitt triggers operated this fast, but it was difficult to maintain stability. A longer time constant in the sensing network slowed this response to 3 to 5 half-cycles and prevented instability. The output inductor would necessarily have to limit the build up of current to the load so the inductor was increased to its present size.

Consideration was given to have a single Schmitt trigger running at double the oscillator frequency and driving a binary count-down circuit to feed the phase-shifted pair of inverters. However, this was discarded as difficulties were foreseen when the system attempted to phase-shift rapidly.

E. OVERCURRENT CIRCUITRY DEVELOPMENT

The first proposed current sensing circuit used two ring cores fed with a square wave and having their output windings connected in opposition. One core had a square loop characteristic and had a winding which was fed with the output current. Under normal conditions both cores were unsaturated so the net result was practically zero output. In the event the output increases to 1 ampere (25% overload) the square loop core saturates and a D.C. output is obtained. This signal trips an 'AND' gate which interrupts the Schmitt trigger operation and drives inverter 2 out of phase with inverter 1. Removing the output short circuit restores operation of the Schmitt triggers and the 'AND' gates are released. When it was decided to have the po[illegible] supply work without a proportional output current control, this system was discarded.

Other methods were tried for overcurrent sensing and one which proved to be extremely successful was to use a pair of transistors in push-pull with their bases driven with a square wave. The D.C. voltage applied to the collectors is developed across a resistor placed in series with the output of the high voltage supply. As the output current increased a large output was produced. This output was rectified and fed back to the Schmitt triggers to provide proportional phase control. When the series input transistors were added to the inverter circuit, it was no longer desirable or necessary to sense and control the current with this type of circuit.

F. VOLTAGE REGULATOR DEVELOPMENT

Voltage regulation of the inverters is accomplished by phasing back one pair of inverters with respect to the others. The output voltage needed to be sensed proportionally and a D.C. signal fed back to the phase shifters. Several circuits were tried before arriving at the present system. One circuit which was tried was to sense the high voltage A.C. being fed into the bridge rectifier. This was not highly successful because the peaks of A.C. voltage always remained the same and spikes of voltage at narrow conduction angles produced errors. The present system consists of a bleeder network across the high voltage D.C. output, feeding a Shockley diode pulse transformer network and a reference Zener diode, which produced successful results.

G. GCS DRIVE CIRCUIT DEVELOPMENT

The progressive steps in the development of the GCS gate circuitry is shown by Figures III-5 a, b, c; III-5 d,e,f; and III-6. These figures display the various forms of GCS gate circuits considered. For simplicity, only the disadvantages of each circuit are given. In Figure III-5a an early drive circuit is shown in which delay of the turn-on gate current was provided by a saturating core. A problem with this method was that magnetizing current flowing into the gate would turn the GCS on. The circuit of Figure III-5b was not satisfactory because the high reverse voltage drop of the GCS gate caused the Shockley diode to fire on the turn-off pulse, thus shunting the current away from the GCS gate. The circuit of Figure III-5c solved the problem of the above circuit, but the turn-on waveform is slow in rising due to the inductance of the transformer winding. Figure III-5d shows one proposed drive circuit that was considered but dismissed because of its complexity. Likewise, were the circuits of Figure III-5 e and f dismissed.

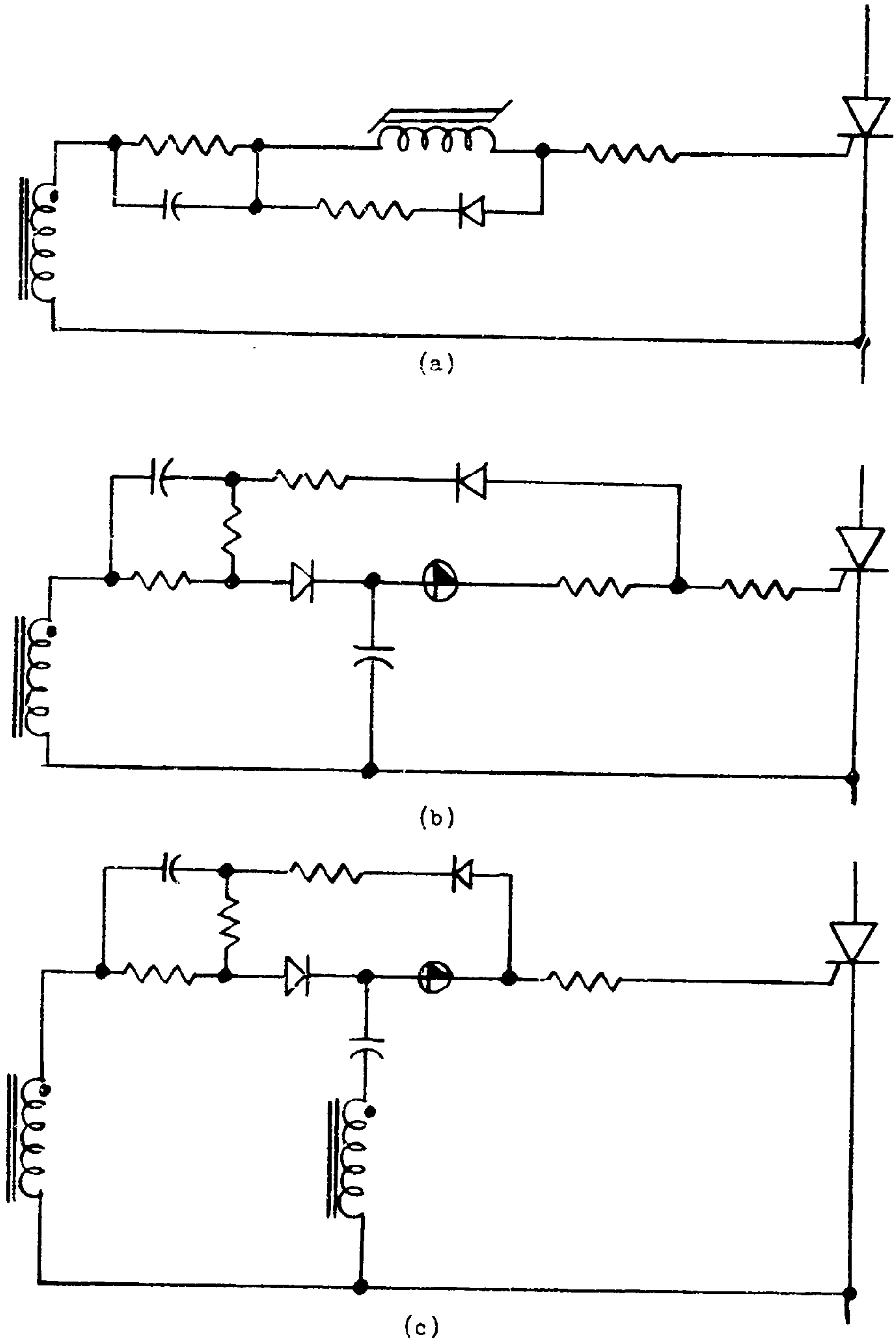
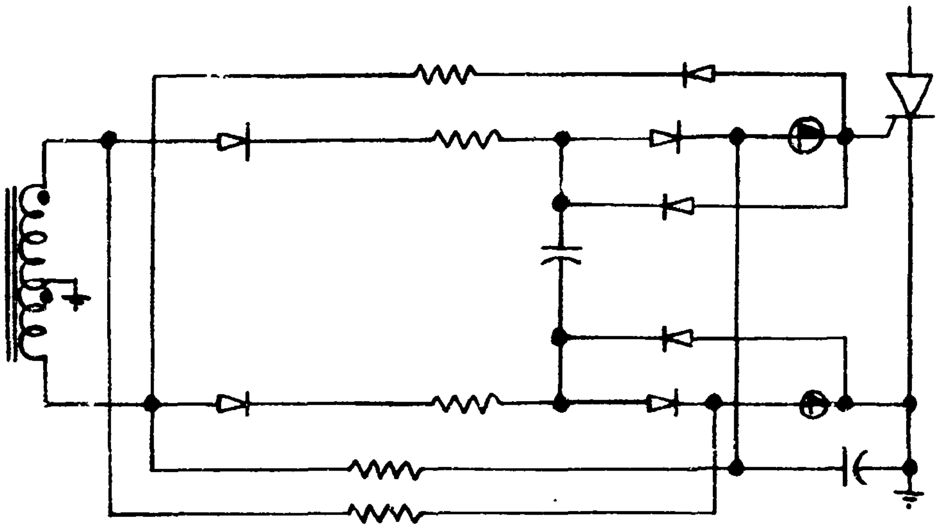
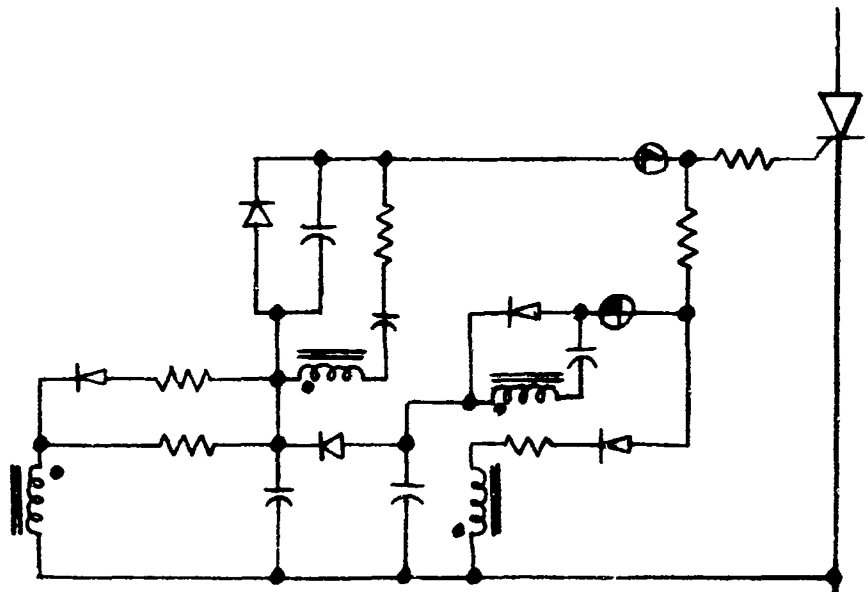


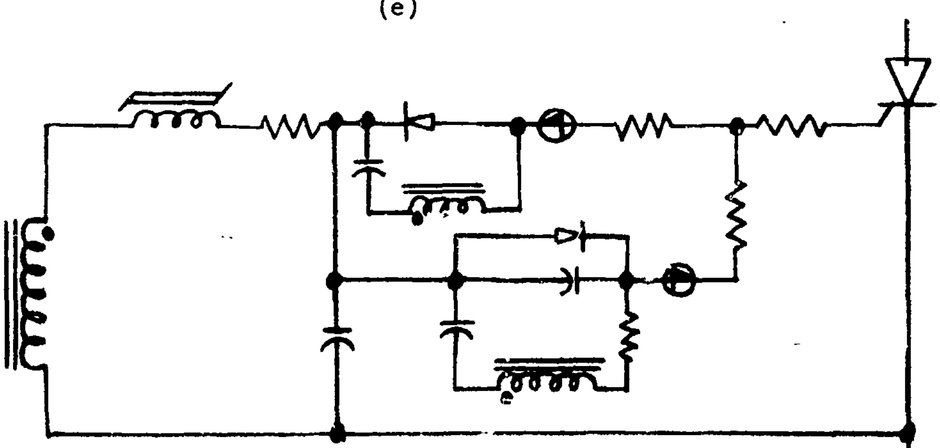
Figure III-5. Experimental GCS Drive Circuits



(d)



(e)



(f)

Figure III-5. Experimental GCS Drive Circuit (Cont'd)

The final drive circuit chosen is shown in Figure III-6. Once the need for delay of one GCS turning on after the turn-off of the other GCS was removed, a common gate pulse could be used. The pulse is produced by breakover of the Shockley diode in the full wave diode bridge. The two capacitors provide a low impedance so that the pulse magnitude can be high. The shunt capacitor by-passes the inductance of the transformer secondary and the series capacitor forms the pulse. The center-tapped secondary winding and the resistor-diode networks provide bias to the GCS; current bias to keep it on during its conducting half-cycle and voltage bias to keep it off during its blocking half-cycle.

H. "SOFT-ON" CIRCUIT DEVELOPMENT

The "Soft-On" circuit acquired the form shown in Figure IV-1 as soon as the necessary input criteria to the Schmitt trigger were established. The breadboard model "Soft-On" circuit was not capable of being reset in the minimum "Blink-Off" time period of 2 milliseconds. In order to remedy this situation the "Soft-On" circuit was changed to that shown in Figure App. B-1 This change provided the required rapid reset feature, and was the only major development performed on this circuit.

I. "BLINK-OFF" CIRCUIT DEVELOPMENT

The original method of initiating "Blink-Off" was by the timing out of the overcurrent time delay circuit. After a predetermined time of operating in a current limiting mode into an overload condition the circuit would "Blink-Off". The addition of the series power transistors Q26 and Q29 eliminated the overcurrent regulator circuit and provided a new form of current regulation. The method of sensing overcurrent was then changed to that shown in Figure App. B-1. When the input current to an inverter reaches a certain point, the series power transistor pulls out of saturation and limits the maximum input current to that value. The voltage appearing on the collector of the series transistor is then fed into an RC integrating circuit. When

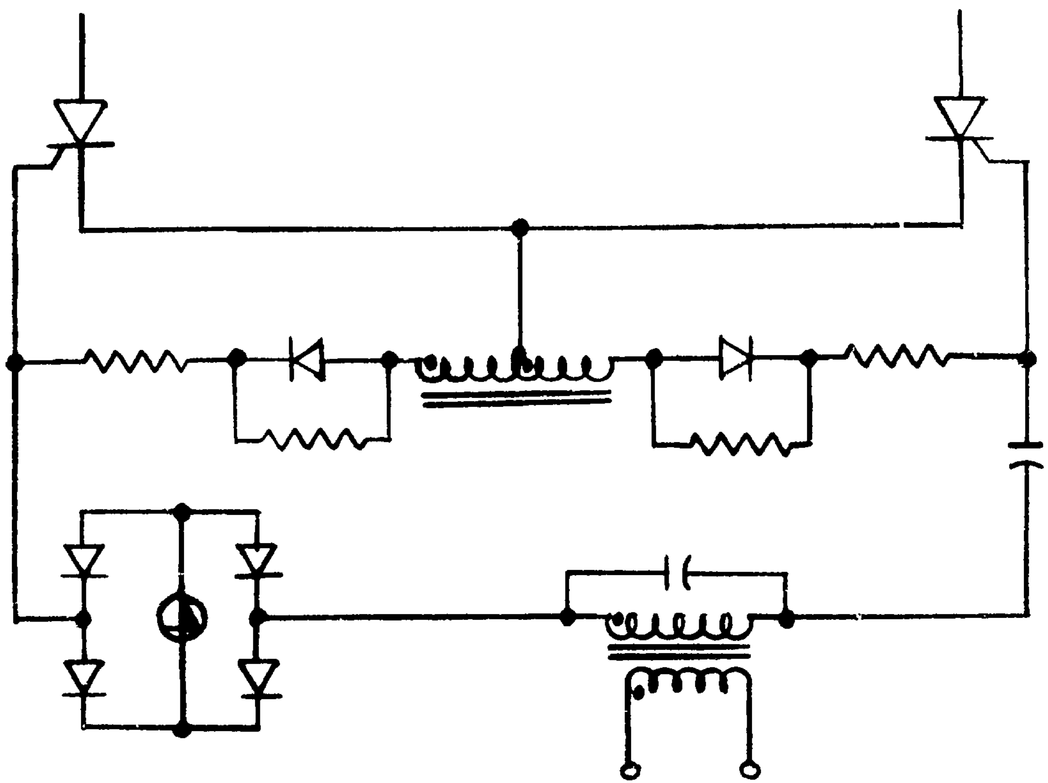


Figure III-6. Final GCS Drive Circuit

the voltage on the capacitor reaches the voltage of the four layer breakover device, a pulse is formed which is delivered to SCR Q37 to initiate "Blink-Off". The input circuit to the gate of Q37 is the only major developmental change; the rest of the "Blink-Off" circuit which provides the timing function has not been greatly changed.

IV. BREADBOARD MODEL EVALUATION

The results of the Breadboard Model testing are segregated into five categories: performance of control and protection functions, static load tests, component electrical stress tests, gate controlled switch failures, and conclusions and recommendations.

A. PERFORMANCE OF CONTROL AND PROTECTION FUNCTIONS.

1. Start-up Time Delay

With the on-off switch (SW1) (see Figure IV-1) in the on position, after power is applied to the input terminals of the unit, there is 200 milliseconds of delay before the system comes "soft-on" and starts supplying output voltage. The major portion of this delay is in the chopper circuit, where 160 milliseconds are required to charge the filter capacitor. Further delay is in the Shockley diode and SCR start circuit on the input to the master oscillator. Added delay in the master oscillator results due to charging of the rectifier supply filter capacitors. Thus the master oscillator is started and stabilized, all d-c supplies are on, control circuits are energized and prepared, and timing circuits are ready to initiate proper sequential operation of the system. The start-up time delay only operates when input power is first applied or whenever the input power is removed and then reapplied. In the experimental model the start-up time delay remains at about 200 milliseconds because of the addition of CR 102 and Q27 (See Figure App. B-1).

2. "Soft-On" Time Delay

The "soft-on" time delay, determined in part by resistors R75, R76 and capacitors C45, C46, is adjustable between 2.0 milliseconds and 2.0 seconds in accordance with the work statement requirement. The "soft-on" circuit controls the rate at which the two power inverters are brought in phase. The two inverters are in phase within 10 degrees at the end of the timing interval chosen. Measurements of "soft-on" time constant range in the experimental model were determined by the length of time the output took to reach 2500 volts.

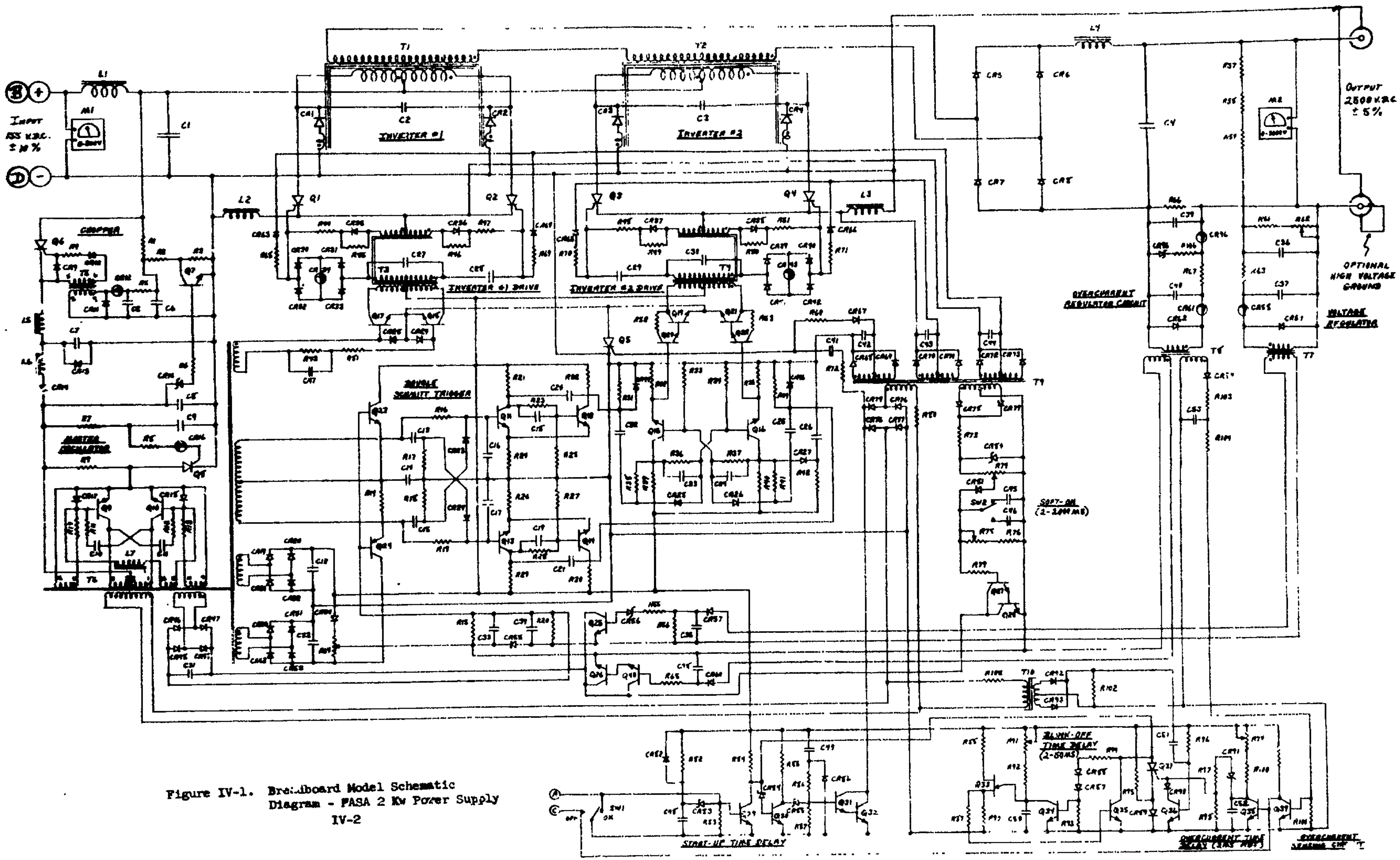


Figure IV-1. Breadboard Model Schematic Diagram - FASA 2 Kw Power Supply

3. Overcurrent Time Delay

When the output exceeds the trip point of the overcurrent regulator circuit, the Shockley diode relaxation oscillator begins supplying a signal; part of which is conducted to the overcurrent sensing circuit. This signal initiates the overcurrent time delay. After this delay it is possible to "blink-off" the system. The actual "blink-off" is inhibited by a synchronizing circuit. The synchronizing circuit allows the main power inverter's switching transients to be over before "blink-off" occurs. Thus the synchronizing circuit imposes discrete steps of about 0.6 milliseconds in the overcurrent time delay. The overcurrent time delay is adjustable from 2 to 50 milliseconds in accordance with the work statement requirement. Later this circuitry was eliminated because of requested changes in the work statement requirements.

4. "Blink-Off" Time Delay

The "blink-off" time delay is adjustable from 1 to 50 milliseconds. This is the time between "blink-off" of the system, due to the overcurrent time delay circuit, and the start of "soft-on". An important point to note here is that the output current may or may not fall to zero during the off time of the system (i.e., 2 to 50 milliseconds). In the experimental model "blink-off" was extended to a range of 25 to 500 milliseconds.

5. On-Off Switch

The on-off switch and control terminals A and C are provided for external control of the system. With the switch in the on position, or control point A shorted to the minus input terminal, the operation of the system is as described in Section IV.A.1. When the switch is moved to the off position or when control point C is shorted to the minus input terminal, the operation is as described below. With SW1 in the off position, the base of Q38 is shorted to ground simulating an overcurrent in the output to the overcurrent time delay circuit. The operation then proceeds as described in Section IV.A.3 until the system is automatically shut down by "blink-off". The controls then lock up the operation to the off mode. Moving the switch back to the on position brings the control immediately to the "soft-on" condition.

B. STATIC LOAD TESTS.

1. Input Current Ripple

The input ripple from the source was measured using a T and M Research Products current viewing resistor (model number SBNC-2-01). The current ripple wave form is shown in Figure IV-2. The allowable ripple is 2% rms and 5% peak. For a nominal input current of 14 amps at full load, the rms ripple allowable is 0.28 amps and peak is 0.70 amps. The waveform of Figure IV-2 shows a peak-to-peak ripple of 0.25 amps. Thus the system meets the requirement of the work statement. The input filter may be reduced to conserve weight and still meet the 2% rms and 5% peak input ripple current requirement.

2. The Output Voltage Ripple

The output voltage ripple was observed under full-load and no-load. The input voltage was maintained at 155 volts D.C. Figure IV-3 (a) and (b) show the full-load voltage ripple and no-load voltage ripple, respectively. The full-load output voltage ripple was measured using a thermocouple voltmeter connected in series with a high voltage capacitor. The high voltage capacitor was approximately ten times the capacitance of the output filter capacitor. The measured ripple was 4.5 volts rms, or 0.18%. The 5% maximum ripple voltage in the work statement was thus satisfied.

3. Efficiency

The efficiency of the breadboard model was measured and is graphically shown in Figure IV-4. The full-load efficiency for 155 volts D.C. input and rated output is 90%.

4. Regulation

The ability of the system to regulate the output voltage to 2500 volts D.C. $\pm$ 5% from 30% load to 100% load and to regulate to 2500 volts D.C. $\pm$ 10% at no-load in accordance with the work statement was measured.

Amplitude Scale: 0.1 amps/cm.

Time Scale: 0.5 milliseconds/cm.

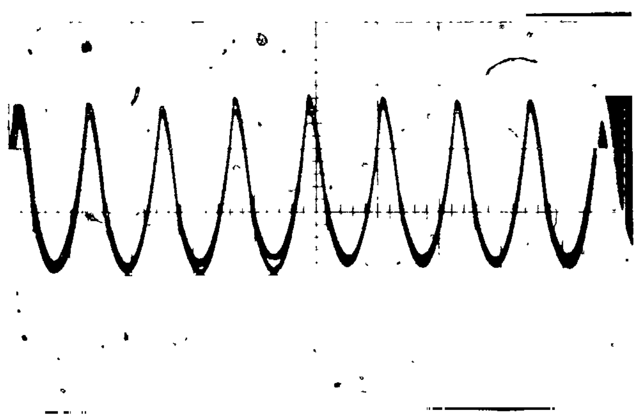
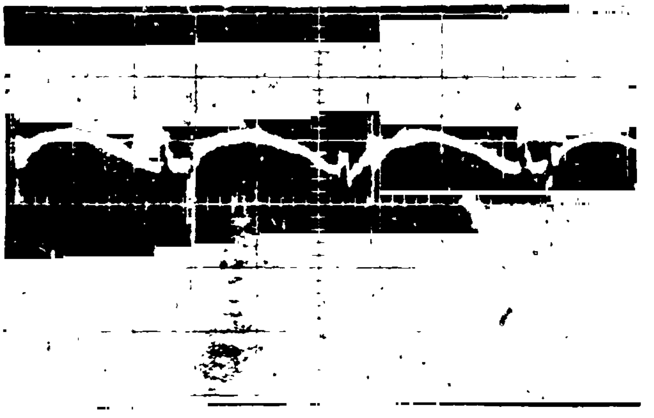


Figure IV-2. Input Direct Current Ripple at Full Load

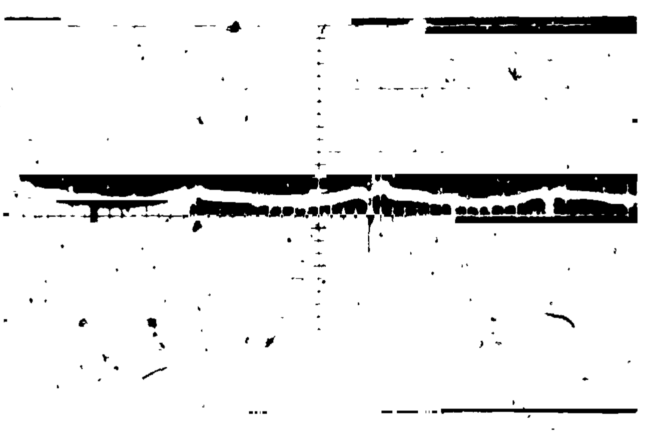
Amplitude Scale: 20 volts/cm.

Time Scale: 0.2 milliseconds/cm.

For Both Pictures



(a) Full Load



(b) No Load

Fi. . IV-3. Full Load and No Load Output Voltage Ripple

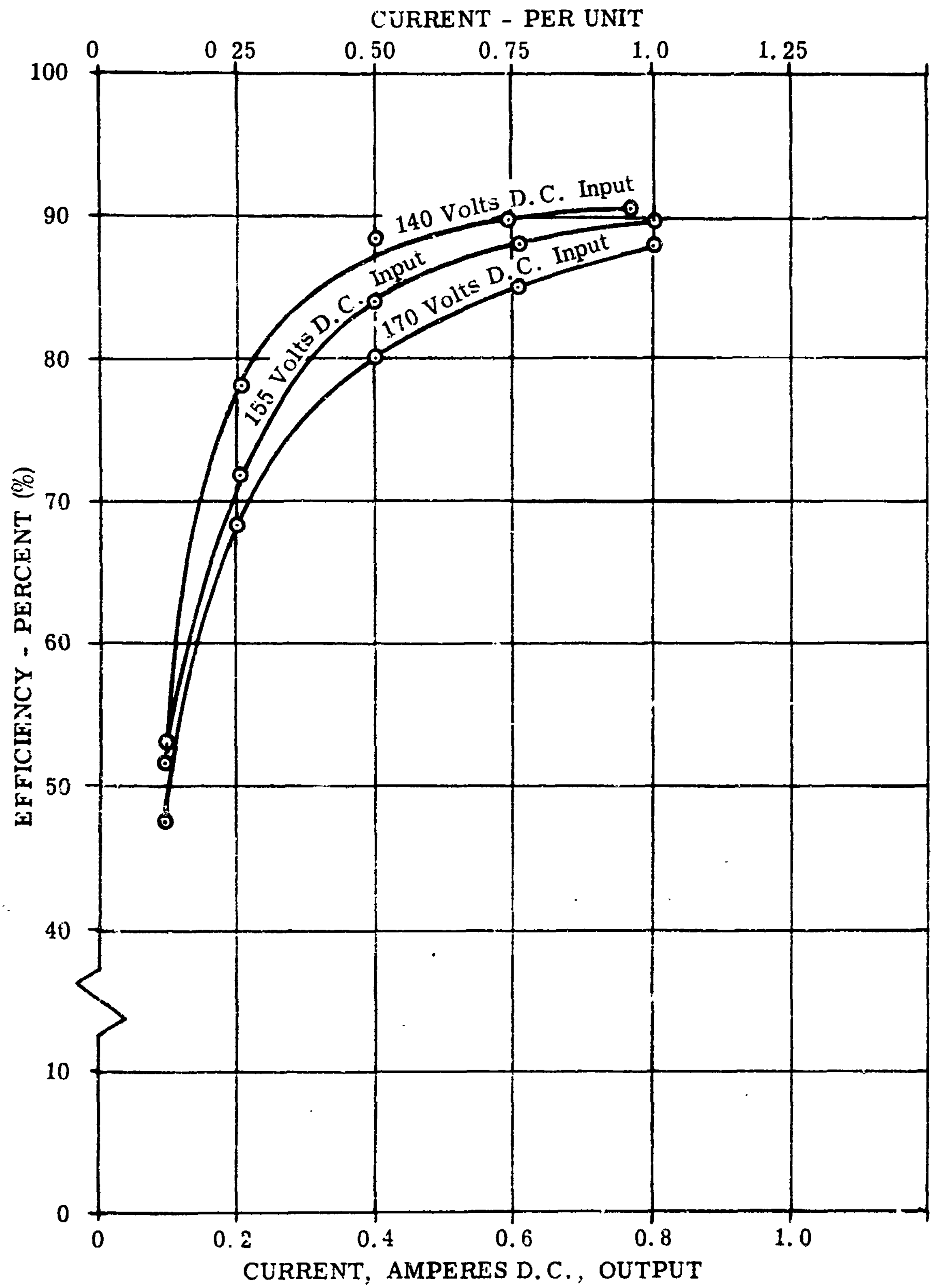


Figure IV-4
Efficiency Versus Load for 2 KW, 2500 Volt D.C.
Breadboard Model Power Converter

The results are shown in Figure IV-5 for three conditions of input voltage; 140 volts D.C., 115 volts D.C. and 170 volts D.C. The results show that the system meets the regulation requirements of the work statement.

5. Safety Factor Applied to Inverter GCS's

The work statement goal of a 2 to 1 safety factor on current and voltage for semiconductors was checked in the breadboard model for the critical component, the GCS. The test results showed that with the present rating GCS devices a safety factor of about 1.75 to 1 could be achieved. The voltage and current for one GCS in each of the two power inverters was observed on an oscilloscope and photographed. The oscillograms are shown in Figures IV-6 (a) and (b), and IV-7 (a) and (b). The peak voltage across a GCS is 400 volts and the rms current through a GCS is 5.35 amps (see Figure IV-8). The voltage safety factor was 1.75 to 1. The rms current safety factor was 1.87 to 1. The resulting D.C. voltage across the high voltage rectifier bridge is shown in Figure IV-9. To achieve a higher safety factor with available GCS's would require using a greater number of devices. It is felt that the added complexity associated with additional devices is not warranted in the present development program, particularly since higher power devices which would provide greater safety factors are under development and will be available in the future.

C. COMPONENT ELECTRICAL STRESS TESTS.

The component electrical stress tests on the breadboard model were limited by the inability of the GCS's to take large overload currents during short circuit transients. Components have since been added to reduce the surge current through the GCS. One change involved making the output choke L4 larger in inductance. Obviously, this increased system weight. A second change, considered during breadboard tests was the addition of a transistor in series with the input to the inverters. This transistor would be fully on or saturated during normal operation, but would regulate the current by pulling out of saturation during overloads. Later in the breadboard develop-

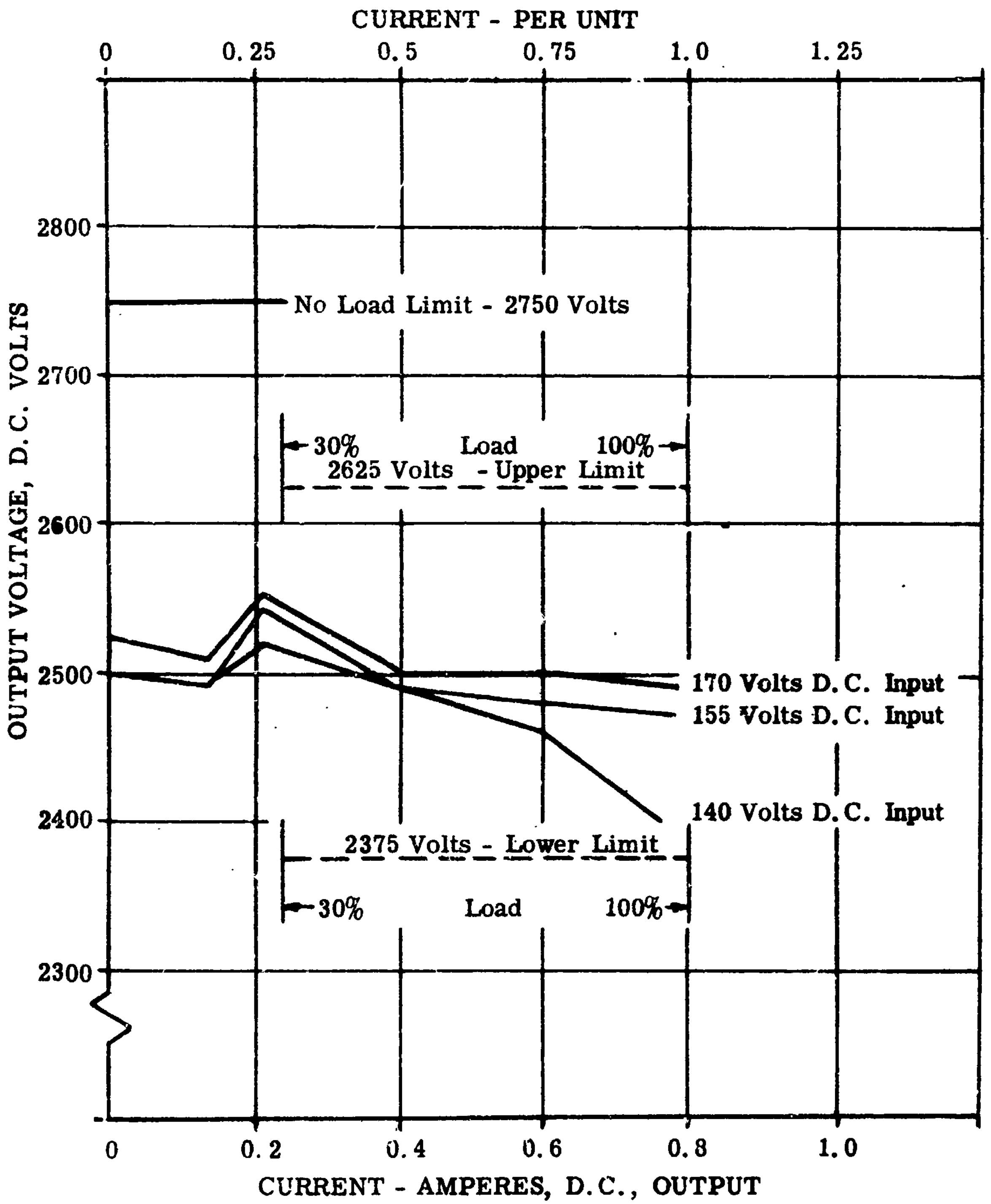
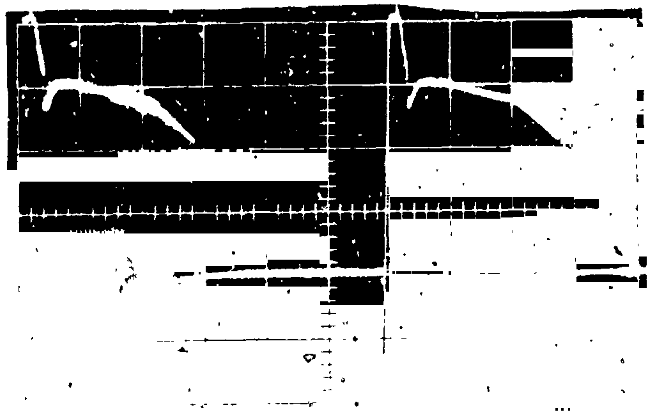


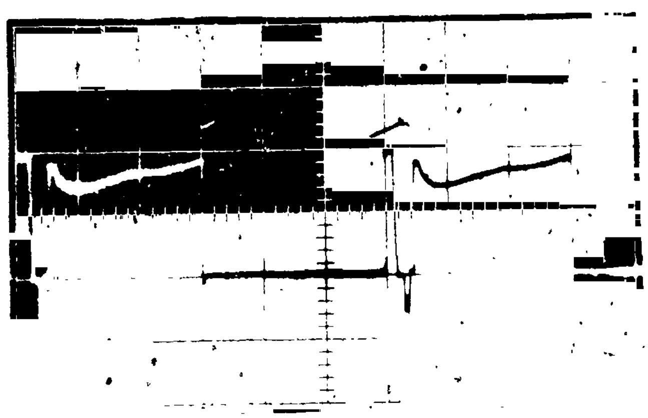
Figure IV-5.

Voltage Regulation Curves for 2 KW, 2500 Volt D.C.
Breadboard Model Power Converter



(a) GCS Voltage V_F

Amplitude Scale: 100 volts/cm.

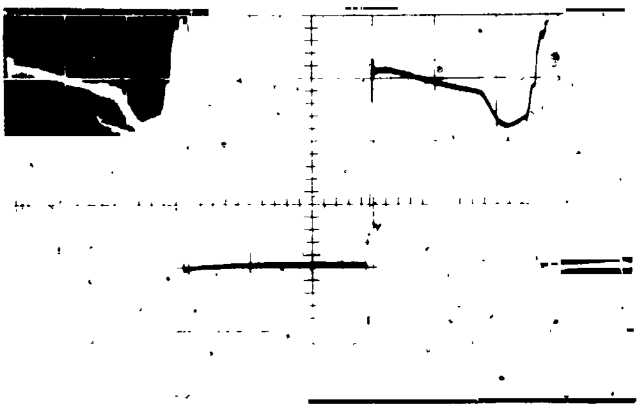


(b) GCS Current I_F

Amplitude Scale: 5.0 amps/cm.

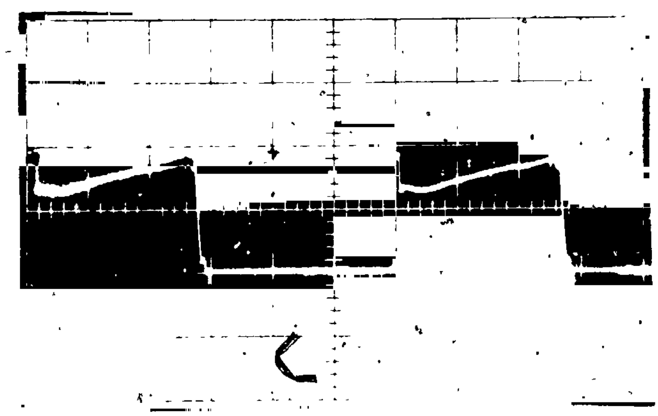
Both Pictures at 0.2 milliseconds/cm.

Figure IV-6. GSC Voltage and Current in Inverter No. 1



(a) GCS Voltage V_F

Amplitude Scale: 100 volts/cm.



(b) GCS Current I_F

Amplitude Scale: 5.0 amps/cm.

Both Pictures at 0.2 milliseconds/cm.

Figure IV-7. GCS Voltage and Current in Inverter No. 2

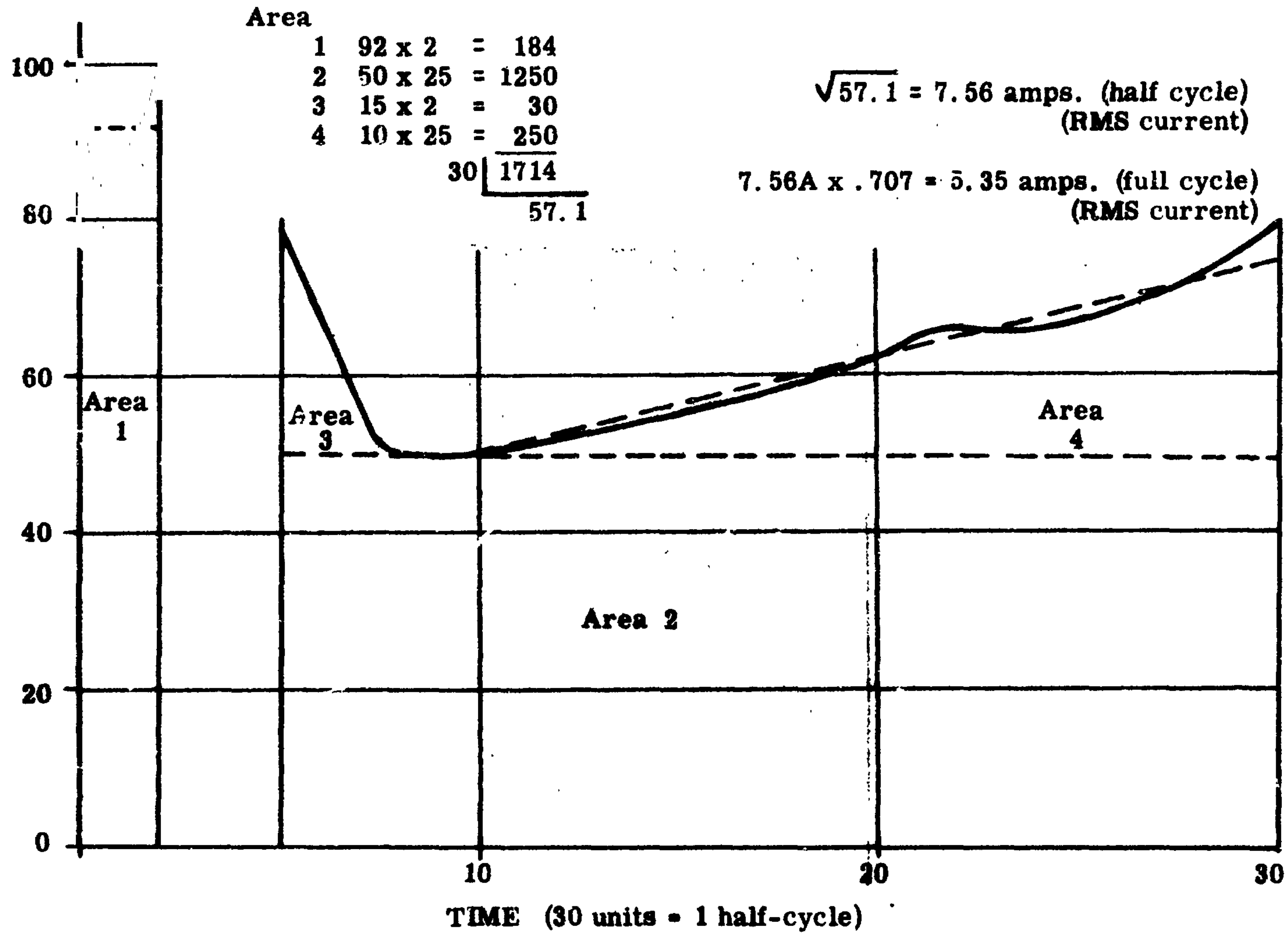


Figure IV-8

Determination of RMS Current of Figure IV-6(b)

Amplitude Scale: 1000 volts/cm.

Time Scale: 0.2 milliseconds/cm.

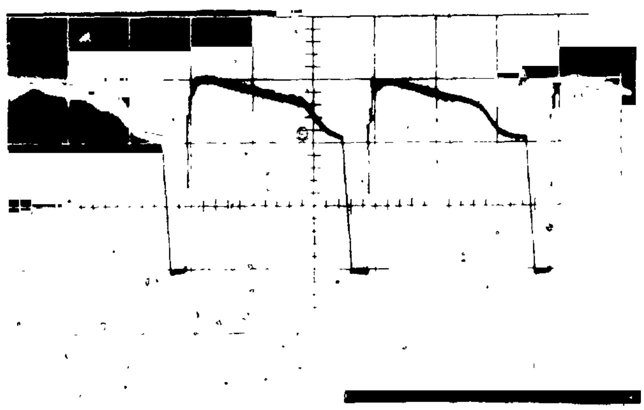


Figure IV-9. High Voltage Rectifier Output Voltage ment a transistor was added in series with each inverter to permit operation of the system during transient short circuit conditions. In the first breadboard the system was able to handle a suddenly applied load of 500 ohms, which is equal to about one-sixth the rated load resistance of 3125 ohms. Under this condition the system was able to phase back and regulate the output D.C. current to 0.85 amps or about 5% overload. It was also possible to short the output terminals with the system de-energized and then slowly raise the input voltage from zero to nominal input voltage. The converter, however, would not withstand the sudden application of a short circuit.

D. GATE CONTROLLED SWITCH FAILURES

During the course of breadboard development several GCS failures occured. Some failures were due to external causes such as accidental grounding of circuits through oscilloscopes or shorted connections, while others resulted from a malfunction of the inverter circuit. Of the total 15 failures, five were accidentally caused and the remaining ten were caused by circuit malfunction. When the inverter GCS failed to turn-off at the end of its half-cycle and the opposite GCS turned on, the current through the devices would rise to a large value and would fail the device. Likewise, if the output transformer saturated, a large current would flow through the GCS.

Fast-acting current limiting fuses were used to protect the device; however, it was found that these fuses would not clear rapidly enough to prevent device failure. To provide better protection for the devices, a solid state circuit breaker, which could interrupt within a few microseconds, was added to the breadboard system. After the addition of a static circuit breaker, no further device failures were encountered.

The majority of the circuit malfunctions which caused GCS failures were the result of output transformer saturation. When slightly different volt-seconds were applied to output transformer primary winding during successive half-cycles, the core would eventually become saturated and require more excitation current, than the GCS could turn-off. The currents through each GCS were balanced by the addition of C2, C3, L2 and L3. (See Figure IV-1). Assume that the current I1 through Q1 exceeds the current I2 through Q2 at the end of their half-cycles. When Q1 is being turned off, the larger current I1 flows through C2 and reverses its charge in a relatively short time.

$$e = \frac{1}{C} \int i \, dt$$

If e and C are fixed, then the larger the value of i the smaller dt becomes. The voltage across the transformer T1 reaches peak magnitude sooner and maintains it longer in this half-cycle because the large I1 reversed the charge on C2 relatively fast. The volt-seconds of the present polarity applied to the transformer will therefore be larger and tend to increase the current I2. If, however, I2 remains small when it is time to commutate Q2, then the small current will take longer to reach peak voltage of the opposite polarity in the new half-cycle. Therefore, at the end of this new half-cycle the transformer will have a smaller magnetizing current. The current I1 is then smaller than the cycle before. The total effect of C2 is to equalize the two currents I1 and I2 by changing the half-cycle volt-seconds applied to the transformer T1. The inductor L2 is provided to present an impedance between the inverter and the source while switching is taking place, so that C2 can successfully perform its function.

E. CONCLUSIONS AND RECOMMENDATIONS

The breadboard circuitry development indicated that the performance of the GCS power conditioning system could approach the goals specified in the work statement. Use of the GCS in the type of inverter system originally selected and proposed for the application presented rather difficult problems. The major problems with the GCS was that unless circuit operation was restricted to those values of current and voltage which the device can reliably turn off, failure of the circuit would occur. To maintain operation within these restrictions, additional circuitry was required to control the transient conditions which are beyond the GCS turn-off capability. During the breadboard development solutions to many of the problems associated with the application of the relatively new gate controlled switches were found. Further effort after the breadboard evaluation was required to achieve satisfactory inverter operation with suddenly applied output short circuits.

V. GATE CONTROLLED SWITCH EVALUATION

The Gate Controlled Switch was considered a "critical component" during the program and therefore warranted further detailed examination: The following are the results of that examination.

A. GATE CURRENT TO FIRE

The gate current to fire was measured for eleven GCS's and the results are shown in Table V-1. The method of measurement was to apply a long time duration low amplitude positive gate pulse. The pulse magnitude was then slowly increased until the GCS would continually fire. This current was noted and is the gate current to fire. The pulse time duration was then shortened until the device began to fail to fire and then increased just enough to ensure continued firing. This is reported as the pulse width to fire. The conditions for this test are 25°C, a 200 volt input and a 20 ohm load.

The actual gate current pulses used in the ion engine power supply are typically those shown in Figure V-1 (a) and (b). The comparison of these actual waveforms with the results of Table V-1 indicates that the positive gate current pulses supplied to the GCS by the system drive are more than adequate to fire any GCS of the type used.

B. GATE CURRENT TO TURN-OFF

The same test set-up used above was used to test the eleven GCS's for the magnitude and duration of the negative gate current pulse required to turn the device off. The negative gate current pulse amplitude was set at 4 amps and slowly reduced until the GCS failed to turn-off. The magnitude was then increased enough to ensure turn-off and the duration shortened to the point where the GCS still managed to turn-off. These two quantities are listed as the gate current to turn-off and the pulse width to turn-off, respectively in Table V-1. The conditions of the test are 25°C, 200 volt input and a 20 ohm load.

Table V-1. Test Results of Eleven GCS's

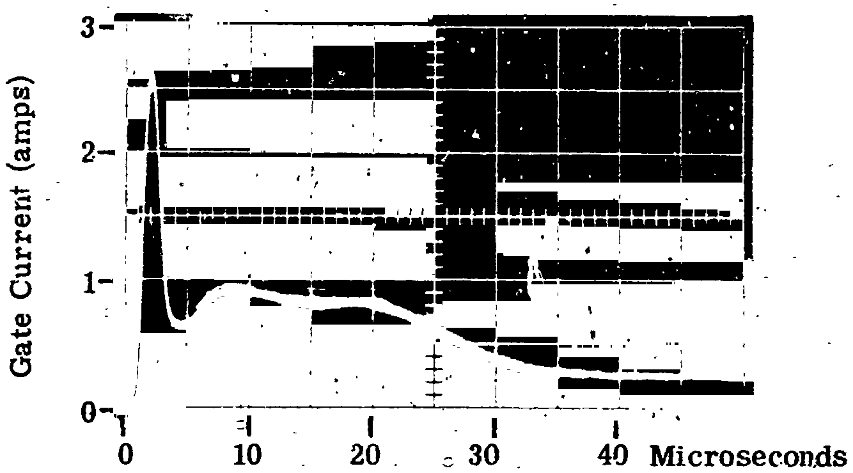
GCS No.	V_{fB} (volts)	V_{rB} (volts)	t_{on} (μsec)	t_{off} (μsec)	I_{hold} (amps)	V_f at 5 amps (volts)	V_f at 10 amps (volts)	$I_{g\ to\ f}$ (amps)	$\Delta t_{to\ f}$ (μsec)	GCS Δt_{on} (μsec)	$I_{gT\ off}$ (amps)	$\Delta t_{T\ off}$ (μsec)	GCS Δt_{off} (μsec)
1	800	500	2	4	0.36	2.2	2.7	0.1	6	8	2	20	8
2	700	450	2	12	0.36	2.4	2.9	0.06	10	12	0.3	14	13
3	650	850	2.5	-*	0.32	3.0	3.9	0.1	19	20	0.36	14	14
4	800	870	3	1.5	0.45	3.1	3.6	0.12	15	15	0.42	27	27
5	750	750	3	5	0.38	2.3	2.8	0.1	15	20	0.22	26	20
6	750	850	5	4	1.1	3.6	4.1	0.25	13	14	0.36	28	24
7	750	900	3	-*	0.2	3.0	3.5	0.1	9	10	2.4	13	13
8	750	900	3	4	0.45	2.4	2.9	0.12	15	15	0.4	25	25
9	600	300	3	-*	0.3	2.8	3.6	0.2	5	9	0.2	30	30
10	720	720	2	0.5	0.3	2.6	3.0	0.08	8	8	0.5	18	17
11	610	900	5	2	0.5	3.7	4.7	0.2	15	15	0.4	25	25

*Unit needs external shunt capacitance to help turn it off.

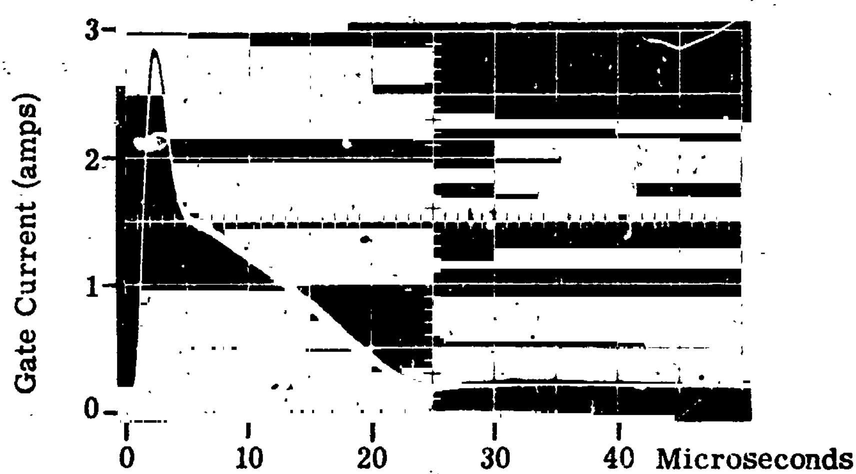
Legend

V_{fB} - forward voltage at 1 ma of forward leakage current.
V_{rB} - reverse voltage at 1 ma of reverse leakage current.
t_{on} - turn on time with 0.5 amp drive 200 volt 10 amp load.
t_{off} - turn off time with 4.0 amp drive 200 volt 10 amp load.
I_{hold} - minimum anode current device can sustain with no gate drive.
V_f - forward voltage drop of the device while conducting.
$I_{g\ to\ f}$ - minimum gate current pulse amplitude to ensure turn on.
$\Delta t_{to\ f}$ - minimum gate current pulse time duration at $I_{g\ to\ f}$ to ensure turn on.
GCS Δt_{on} - turn on time of GCS at $I_{g\ to\ f}$ and $\Delta t_{to\ f}$.
$I_{gT\ off}$ - minimum gate current pulse amplitude to ensure turn off.
$t_{T\ off}$ - minimum gate current pulse time duration at $I_{gT\ off}$ to ensure turn off.
GCS Δt_{off} - turn off time of GCS at $I_{gT\ off}$ and $\Delta t_{T\ off}$.
$I_{g\ to\ f}$, $\Delta t_{to\ f}$, GCS Δt_{on}, $I_{gT\ off}$, $\Delta t_{T\ off}$, and GCS Δt_{off} taken for 200 volts 10 amps.

As measured with a model No. SBNC-2-01 T and M Research Products Current Viewing Resistor



(a)



(b)

Figure V-1. Typical Turn On Gate Current Pulse

The inverter drive circuits produce negative gate current pulses to turn off as shown in Figure V-2 (a) and (b). The recommended pulse waveform and a magnitude is shown in the Westinghouse Semiconductor Division's specification sheet for the 242 ZP gate current requirements for turn off of 8 amps. (See Figure V-9). The waveforms of Figure V-2 compared to this waveform show more than adequate turn-off ability at 3 amps.

C. FORWARD VOLTAGE DROP

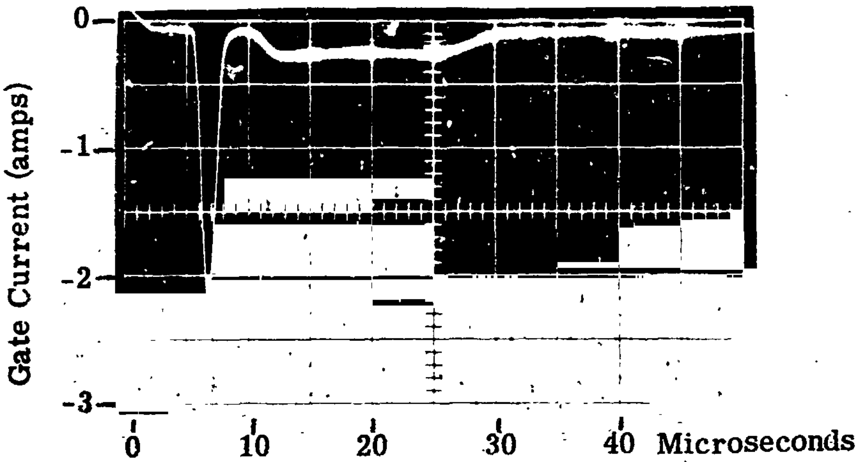
The GCS forward voltage drop versus the forward current was displayed on a scope and photographed. The picture shown in Figure V-3 if for device No. 3 of Table V-1 and it shows the general shape of this V_f versus I_f plot for all devices. Forward voltage drop was measured for all eleven GCS's at current levels of 5 and 10 amps. The forward voltage drop at 5 amps can be compared to the 3.0 volt figure given in the GCS specification sheet. The forward voltage drop at 10 amps can be used in GCS power loss calculations for actual system operation. Measurement is done with the circuit shown in Figure V-4. The measurement at 5 amps show that some devices exceed the specified value of 3.0 volts. These measurements were performed at 25°C.

D. SWITCHING TIMES

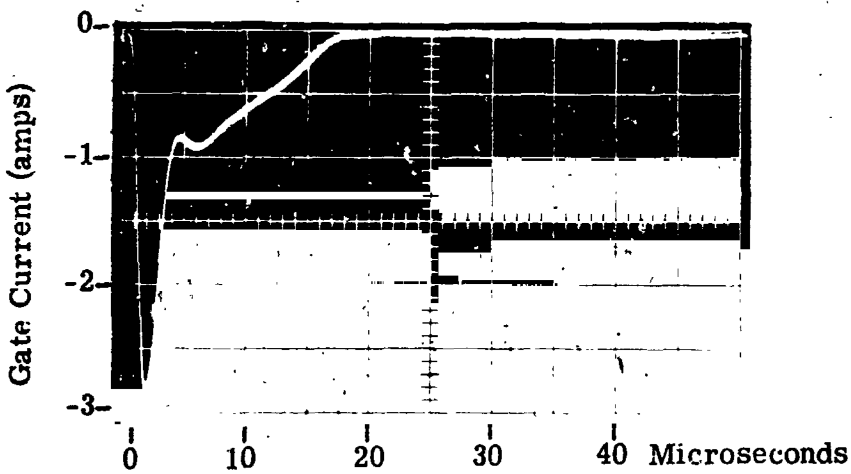
The switching times of the eleven GCS's were measured and the results are reported in Table V-1. One of the GCS's (GCS No. 4) was further checked by taking pictures of its turn-on and turn-off switching time in both the test circuit and the inverter.

In the test circuit the GCS turns on and turns off 10 amps into a 20 ohm load. The drive conditions are 0.5 amps drive for turn-on and 4.0 amps for turn-off. In the inverter circuit, the current that the device must turn-on and turn-off is shown in Figure V-5.

As measured with a model No. SBNC-2-01 T and M Research Products Current Viewing Resistor



(a)



(b)

Figure V-2. Typical Turn Off Gate Current Pulses

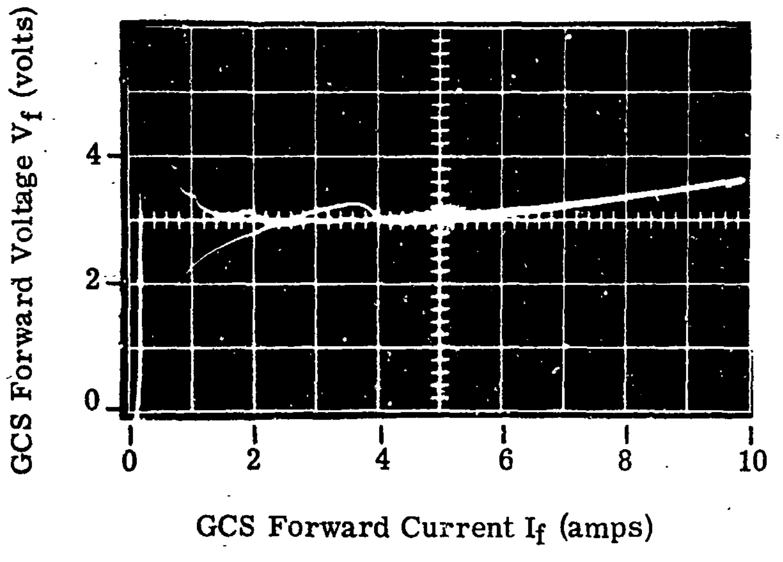


Figure V-3. Typical GCS V_f Versus I_f, Inverter GCS No. 4

(As measured in circuit of Figure V-4)

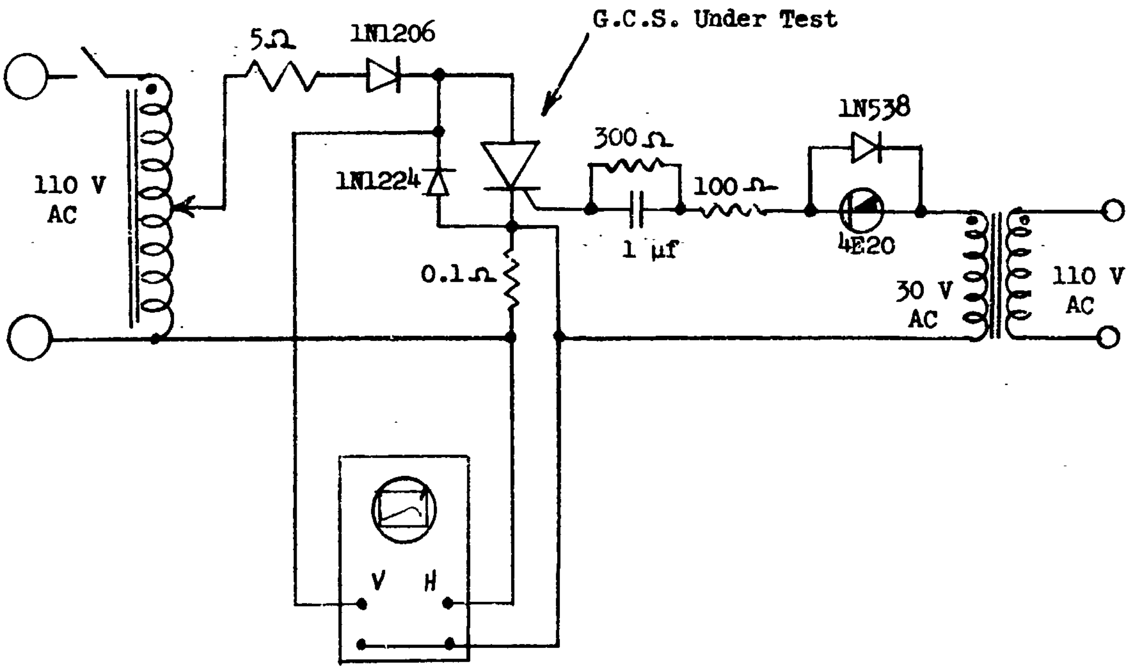


Figure V-4. Forward Voltage Versus Forward Current Test Circuit

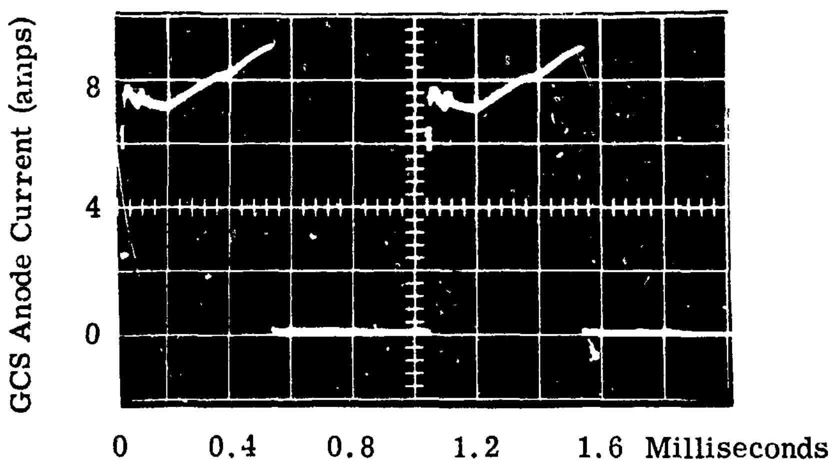


Figure V-5. GCS No. 4 Inverter Anode Current at Full Load

(As measured with a model No. SBNC-2-01 T and M Research Products Current Viewing Resistor)

Figure V-6 (a) shows the turn-on switching time of GCS No. 4 in the test circuit and Figure V-6 (b) shows the turn-off switching time. Figure V-7 shows switching in the inverter circuit under full load. The switching time in the inverter circuit can be seen to be faster on turn-on and comparable at turn-off to those measured in the test circuit.

E. POWER DISSIPATION

The power dissipation, in the four GCS's used in the inverter at full load, was measured by measuring each GCS stud temperature and comparing it to a known stud temperature versus power dissipation plot. A typical plot for GCS No. 4 is shown in Figure V-8. The full load power dissipation for the four inverter GCS's are 11, 12, 20, and 16 watts.

F. SPECIAL TESTS

1. Additional Static and Dynamic GCS Operating Characteristics

Further evaluation of the Westinghouse type 242 gate controlled switch was conducted by our Device Evaluation and Application Group. This evaluation is included to provide supplementary information on the characteristics of the GCS.

The following static and dynamic characteristics of four different GCS's were measured:

a. Forward and Reverse Blocking Voltages (Table V-2)

The blocking voltages are considered to be those voltages at which the device's leakage current reaches 1 ma. Blocking voltages were measured at two temperatures (25 and 100°C).

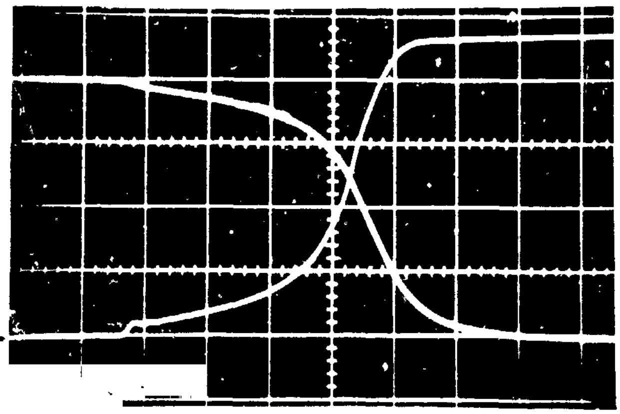
b. Forward Voltage Drop (Table V-2)

The forward anode to cathode voltage drops of the devices were measured with a 10 amp current at 25°C.

Time Scale: 1.0 microseconds/division

Amplitude Scale: Voltage at 50 volts/cm
Current at 2 amps/cm

V

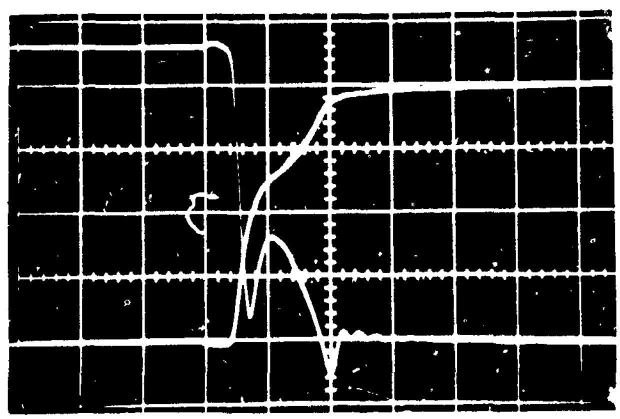


OV & OA

I

(a)

Turn On



V

OV & OA

I

(b)

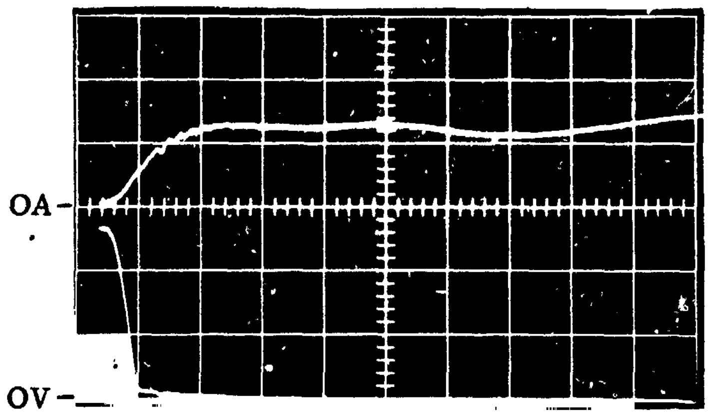
Turn Off

Figure V-6. T.O.T. Tester Switching Times of GCS No. 4

Time Scale: 1.0 microseconds/division

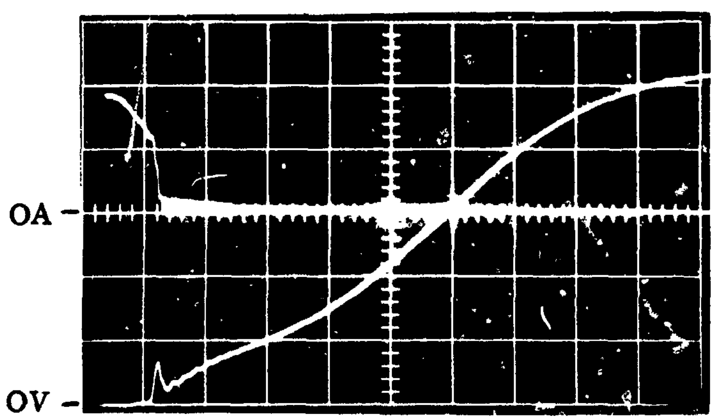
Amplitude Scale: Voltage 100 volts/division
Current 5 amps/division

Current as measured with a model No. SBNC-2-01 T and M Research Products Current Viewing Resistor



(a)

Turn On



(b)

Turn Off

Figure V-7. Inverter Switching Times of GCS No. 4

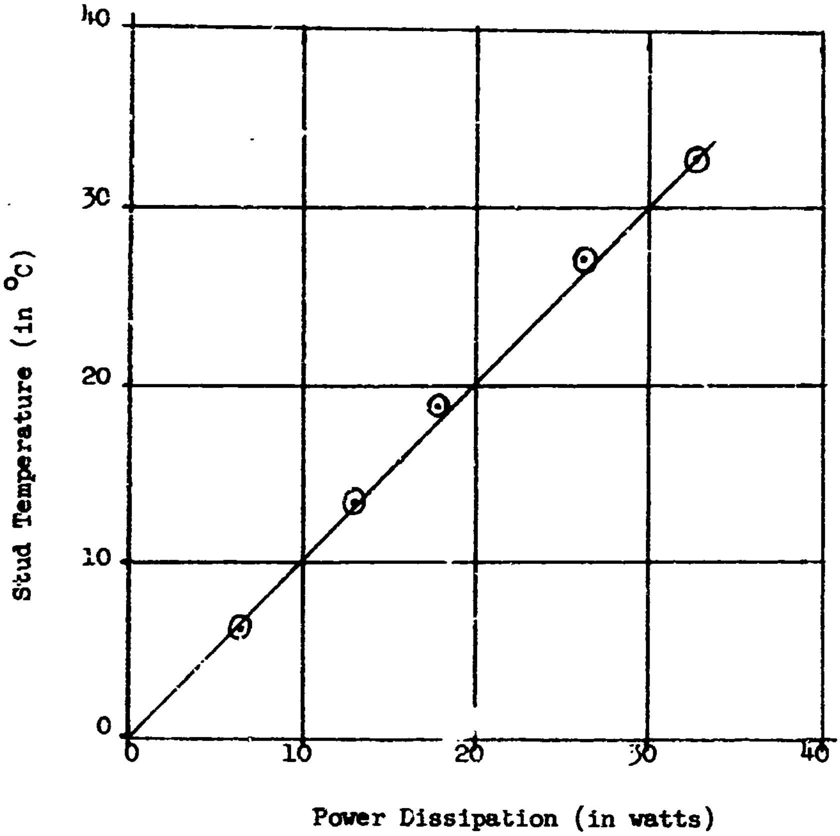


Figure V-8. Stud Temperature Versus Power Dissipation for GCS No. 4

Table V-2. GCS Static Characteristics

GCS Unit No.	Forward Blocking Voltage (volts)		Reverse Blocking Voltage (volts)		Forward Voltage Drop at 10 Amp (volts)	Holding Current (amp)	Thermal Impedance °C/W
	25°C	100°C	25°C	100°C	25°C	25°C	
7	900	>1000	900	>1000	2.7	0.24	0.7
21	160	150	400	320	2.8	0.16	0.5
97	250	380	300	450	4.0	0.76	0.8
98	360	350	300	350	3.0	0.18	0.9

Table V-3. GCS Dynamic Characteristics

GCS Unit No.	Turn On Time (μsec) 100 volts, 10 amp Load	Turn Off Time (μsec) 100 volts, 10 amp Load
	25°C	100°C
7	3.0	5.5
21	2.0	6.5
97	2.5	1.0
98	1.9	5.8

c. Holding Current (Table V-2)

The holding current is defined to be the minimum value of anode current which is required to keep the device in an on state following the removal of the turn-on pulse. As with SCR's, if the anode current is reduced below this value, the internal regeneration ceases and the device will turn off. This was measured at a 25°C temperature.

d. Thermal Impedance (Table V-2)

The thermal impedance from junction to case was measured. This is determined dynamically by measuring the junction and case temperatures as well as the power dissipation of the device while it is carrying a large average forward current.

e. Turn On Time (Table V-3)

This was measured with a 100 volt, 10 amp anode load and a 1/2 amp gate turn on pulse. The measurements were made for a 25°C temperature.

f. Turn Off Time (Table V-3)

This was measured with a 100 volt, 10 amp anode load and a 4 amp gate drive. The measurements were made for a 100°C temperature. It was found that if the turn-off pulse were reduced to 2 amp the turn-off time increased about 50 percent.

2. Turning Off 50 Amps to 50 Volts

The last test performed was to measure the maximum current which can be turned off by the device. For this test a power supply capable of providing 100 volts, 100 amp pulses having a duration of 250 microseconds and a repetition rate of 60 cps was used.

All four units tested were able to turn off 50 amperes being supplied to a 1 ohm load. Two units tested at higher currents failed when turning off 70 amperes to a 1 ohm load. These tests indicate that the devices are capable of turning off currents considerably higher than the forward current rating of the devices.

3. Turning Off 15 Amps to 300 Volts

As was reported in the Second Monthly Progress Report, a test was performed to prove the ability of the GCS to turn off more than rated current of 8 amperes peak in the inverter circuit. The results of this test on four units showed that the GCS could turn off 15 amps to 300 volts. The conclusion was that a nearly 2 to 1 safety factor was built into the inverters for this GCS characteristic.

G. POSSIBLE GCS FAILURE MODES

The conditions which may induce failure of the Type 242 device have not yet been well defined. While it is known that exceeding the published steady state ratings of the device can lead to failure, the effects of specific voltage and current conditions in causing degradation or total failure of the device have not been determined. In particular, the effect of high gate turn-off currents, the effect of reverse anode voltage with forward or reverse gate current, and the limitations on dv/dt and di/dt are not well defined. In the breadboard, varying effects were observed, depending on the GCS being used, and an attempt was made to obtain further data on the characteristics of the devices to determine if these effects would explain the failures.

1. Excessive Negative Gate Current

The device anode current in the inverter displayed the expected half-cycle current waveshape. However, 30 to 40 microseconds after turn-off it again conducted forward anode current. This current lasted about 20 microseconds at 6 to 7 amps and then disappeared for the rest of that half-cycle. The high turn-off current, the presence of reverse anode voltage, and high dv/dt were suspected to be causes of this condition.

This device was removed from the inverter and checked in a separate test circuit to determine if the effect could be reproduced. The device displayed the following characteristics: as the turn-off gate current was increased, the anode current changed from the normal and expected turn-off condition to one similar to that observed in the inverter. This test indicated that increasing the gate current to obtain improved turn-off performance may produce unfavorable effects during the turn-off transients.

2. Negative Anode Voltage and Negative Anode Current

Effects of forward and reverse gate current with reverse anode voltage were observed and it was found that the reverse leakage current of the device increased significantly when gate current of either polarity was flowing. This increased anode current may flow only in localized regions of the device junctions and as a result cause excessive heating which leads to degradation of characteristics or failure. The increased leakage caused by forward gate current is expected and occurs not only in the GCS but also in SCR's. The higher leakage produced by reverse gate current was not expected and did not occur with all devices which were checked.

Since the unexpected effects described above were observed in only a few devices, they can probably be eliminated by improved control of the fabrication process and by appropriate acceptance tests. When this device is in volume production, it is very likely that the quality and characteristics of the device will be more consistent.

3. DV/DT and DI/DT

Limits for dv/dt and di/dt have not been specified for the type 242 device; however, limits of 200 volts/microseconds and 15 amps/microseconds have been suggested by the Westinghouse Semiconductor Division. No failures have been directly traced to excessive dv/dt and di/dt, but under certain circuit conditions, high values of these transients have been observed. The large value of dv/dt and di/dt (1000 volts/microsecond and 40 amps/microsecond) observed were a direct consequence of the fast switching speed of these devices.

To prevent further failure of devices due to the above possible causes, modifications were made in the inverter circuit to eliminate reverse anode voltage, and to limit dv/dt and di/dt. The major change was a reduction in the capacitance connected acorss the transformer primary to reduce and practically eliminate the reverse anode voltage condition. High dv/dt caused by rapid switching of the GCS was reduced by connecting a capacitive circuit in parallel with the device to control the rate of application of forward voltage. High di/dt has been reduced by appropriate additions of inductance to the circuit. More selective screening tests are also being employed to detect devices displaying unusual characteristics. The above changes appear to have improved the reliability of the circuit and to have reduced device failures.

H. CONCLUSIONS AND RECOMMENDATIONS

The results of the Gate Controlled Switch Evaluation led to the following conclusions:

1. The current pulses supplied by the ion engine power supply to the GCS's for turn-on and turn-off are more than adequate as compared to the GCS specifications.

2. The forward voltage drop of the GCS's are generally below the specified value but a few exceed that value. Higher forward drop does not harm circuit operation but does increase circuit losses, and puts a lower limit on maximum system temperature.

3. The switching times (on and off) of the GCS are very rapid as operated in the ion engine power supply. This is due to the good gate drive current pulses provided.

4. The power dissipation within the GCS's when operated at rated full load are not excessive. The heat sink size must be selected to ensure adequate safety factor on the GCS junction temperature.

5. The special tests of this report indicate that the present GCS is capable of turning off much more than the specified maximum eight amperes. One test showed the GCS could turn off 50 amps and this would more than satisfy a 2 to 1 safety factor on this device specification for normal GCS operation in the ion engine power supply.

6. GCS failure modes investigated in this report show that the GCS can display unusual characteristics which could lead to destructive failure of the GCS in this application. It is believed that good acceptance tests and tighter quality control could eliminate the GCS's that displayed the failure modes described in this report.

WESTINGHOUSE ELECTRIC CORPORATION
Youngwood, Pennsylvania

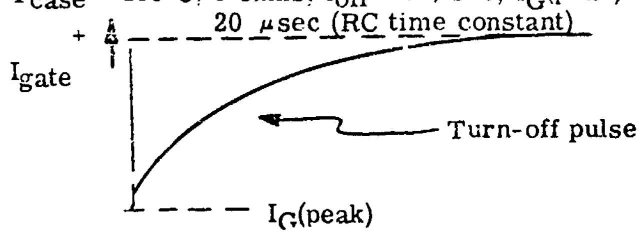
ADVANCED ENGINEERING DATA

MAXIMUM RATINGS & CHARACTERISTICS FOR THE TYPE 242ZB-P GATE CONTROLLED RECTIFIER

0.25" Stud Package (9/16" Hex)

	Symbol	Value
Forward RMS current, amperes	I_F	10
Fwd. Blocking Voltage, Volts, V_{FB}	B D F H K M P	100 200 300 400 500 600 700
Reverse Blocking Voltage, Volts, V_{FB}		100 200 300 400 500 600 700
Peak anode Gate Turn-Off Current, Amperes	I_{ATO}(max)	8.0

T_{case} = 100°C; 5 ohms; t_{off} = 5 μsec; I_G(peak) = 1.0 amps



	Symbol	Value
Peak 1/2 cycle, Δ current amps	I_{FM}(surge)	20
Forward blocking and reverse leakage current at T_J = 125°C, mAdc	I_{FB}, I_{RB}	10
Forward voltage drop at I_F = 5.0Adc at T_J = 25°C, Vdc	V_F	3.0
Thermal resistance, junction to case, °C/W	θ_{JC}	1.4
Operating temperature, °C	T_J	-65 to +125
Storage temperature, °C	T_{stg}	-65 to +150
Turn-off energy, watt-sec	E_{STO}	1×10^{-3}
Peak forward gate power watts	P_{GM}	10
Average reverse gate power, watts	P_{GR}(Av)	5.0
Peak forward gate current, amperes	i_{GF}	4.0
Peak reverse gate current, amperes	i_{GR}	4.0
Average reverse gate current, ma	I_{GR}	400 ma
Gate current to trigger at T_J = 25°C, ma	I_{GT}	200 ma
Gate voltage to trigger at T_J = 25°C, volts	V_{GT}	2
Stud Torque, Dry, In-Lbs.	----	30

Δ at 60 cycles per second

Figure V-9. Published GCS Characteristics

VI. EXPERIMENTAL MODEL EVALUATION

A. EXPERIMENTAL MODEL DESIGN DETAIL

1. Mechanical Construction

The overall mechanical construction criteria include isolation of the high voltage components for personal safety, accessible circuit components for measurement, and isolation of power circuit components form control circuit components for reasons of heat and electromagnetic radiation. A picture of the experimental model is shown in Figure VI-1. The high voltage cage is behind the front panel. On the right and left sides are mounted the boards which contain the control circuitry. Behind the high voltage cage are the series transistors and heat sinks. The larger components such as inverter output transformers and input filter capacitors are also between the high voltage cage and the rear wall. The gate control switches are at the rear on separate insulated heat sinks which form the rear wall of the cabinet.

2. Electrical Design Modifications

a. Series Transistor Current Regulation

The breadboard model tests showed the inability of the original circuit to withstand the application of a short circuit to its output. Considerable time was spent solving this problem, both prior to and following the breadboard evaluation tests. At first effort was mainly directed towards altering the control circuits to achieve the desired operation during the short circuit. Modification and improvement of control circuits did not provide a solution and the search for a solution was shifted to simpler, but less efficient, power circuit alterations.

The solution finally accepted was the use of a power transistor in series with the D C. input to each inverter. The transistors are operated in a current regulation mode and thus, during the application of a short circuit to the inverter, GCS currents do

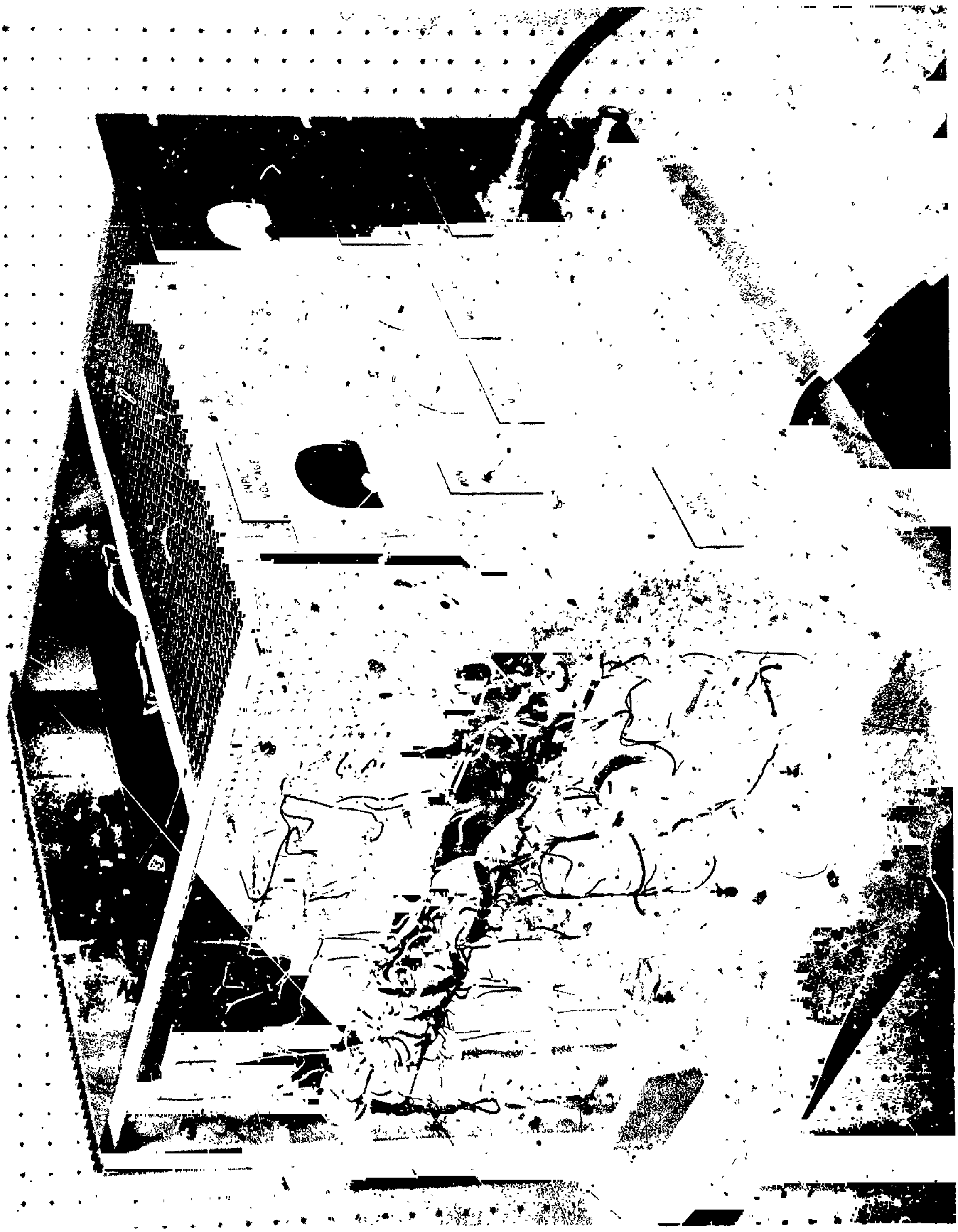


Figure VI-1. Experimental Model Picture

not exceed reasonable levels. Each inverter has its own series transistor for two reasons: 1) the power transistor available can only handle the current and voltage requirements of one inverter with adequate safety margin, and 2) the regulation of the total input D.C. current would allow excessive current in a single GCS if an unbalance should occur.

The results of tests after the circuit additions showed that the transistor current regulation was not compatible with the overcurrent regulator circuit operation. A request was then made for permission to eliminate current regulation into a short circuit and to allow the unit to "blink-off" immediately under overcurrent conditions.

b. "Blink-Off" Modification

One advantage of the series transistor current regulator is that it provides a simple method of accomplishing "Blink-Off". It is much easier to turn-off two transistors than to furnish two pairs of GCS's with their turn-off drive requirements at a relatively high repetition rate. Also, when the transistors go into the current regulation mode upon overload, they provide a convenient signal indicating the overload condition. In the new circuit the voltage across the series transistors is sensed, and when it exceeds a predetermined level for a certain length of time, "Blink-Off" is initiated.

c. Start-Up Time Delay Improvement

The Start-Up Time Delay circuitry of the breadboard model was complicated and as tests revealed, was not necessary; so it was removed. Two other components, Q27 and CR102, were added to prevent the two series transistors (Q26 and Q29) from turning on during the initial application of input D.C. power. After start-up, these components only function when input D.C. power is removed and reapplied.

d. "Blink-Off" and "Soft-On" Circuit Changes

A request was made for a change in the work statement of the contract to lengthen the minimum "Soft-On" time constant from 2 milliseconds to 50 milliseconds and the minimum "Blink-Off" time delay from 2 milliseconds to 25 milliseconds. The reasons for these requests were: 1) to allow adequate safety margin on the power dissipated in Q26 and Q29, 2) to prevent system instability which might occur when the time constants were adjusted to values near the operating frequency of the system.

e. Master Oscillator Improvement

An important improvement was made in the master oscillator circuitry to reduce the number of components and improve the operation. It was found that the full voltage starting feature was not needed for successful start of the low voltage control circuits. The components were removed and a simple, master oscillator start circuit was added. The removal of the previous master oscillator starting circuit and its replacement with the simpler circuit resulted in a savings of weight, and increased efficiency and reliability.

f. Inverter Drive Interaction Problem

Because of the nature of the inverter GCS gate drive requirements there appeared an interaction problem between the two drive circuits. This problem developed as the secondary circuit components of transformers T3 and T4 were altered to produce higher peak current pulses to the GCS gates. These pulses are supplied by each primary circuit and in turn come from the common D C. supply of the primary circuits. The problem resulted in the inability to phase shift the drive signals to inverter No. 2. The large current pulses drawn by the drive of inverter No. 1 from the common D.C. Supply caused the driver of inverter No. 2 to switch at those times, independent of No. 2's own drive signal from the Schmitt trigger and flip-flop. The signals from the flip-flop were increased in an attempt to overcome the interaction, but this did not solve the problem. The final solution was the addition of two RC filter networks. The two RC filters isolated the drive circuits from each other on high speed transients, thus preventing the interaction problem.

g. Improved Drive to Inverter No. 2

Later efforts to improve the magnitude and wave shape of the GCS gate current pulses in inverter No. 2 resulted in a modification of the flip-flop output circuitry. As a result of these modifications, the magnitude and wave shape of the GCS gate current pulses can be adjusted to the desired form.

h. Improved GCS Gate Current Pulse Wave Shaping

Final achievement of the required GCS gate current pulse magnitude and wave shape required the addition of a resistor in each inverter drive circuit. Each resistor limits the magnitude of current that can flow in the transformer secondary and damps out ringing of the transformer.

i. GCS dI/dt Circuit Additions

Detailed observations of the inverter operation revealed the need for limiting the rate of rise of current (dI/dt) in the GCS's. Consultation with Westinghouse Semiconductor Division personnel disclosed the practical limit of 15 amps/μsec. rating of the GCS. Component additions were made to the circuit to achieve the necessary dI/dt limiting.

j. Reinstalling GCS dV/dt Circuits

The same detailed observations of the inverter operation mentioned above and the modifications made there necessitated reinstalling the dV/dt circuits used very early in this project. There is one important difference however; the capacitance value and therefore the diode and resistor are very much smaller. This reduction is possible as a result of turning the blocking GCS on at the same time the conducting GCS is being turned off. The reduced component sizes and reduced power losses are a great savings over the original design. Reinstalling the GCS dV/dt circuits protects the GCS's by holding dV/dt below the value given in the Critical Components Test Report (200 volts/µsec.).

k. Inverter Line Choke Energy Dissipation

The method of removing "trapped-energy" in the inverter line chokes L2 and L3 is by dissipation. This method is an alternate to the method used in the breadboard where a tap on the output transformer was used to reclaim the energy. This method is less efficient but is more reliable than the transformer tap. The inverter line choke energy dissipation is a simple reliable method of eliminating trapped energy.

l. Output Short Circuit Surge Current Limiting

The work statement of this contract did not specify any output short circuit surge current limits. Since the overcurrent regulator sensing resistor was no longer used, there were no internal components to provide surge current limiting. To provide a limit, a resistor was inserted in series with high voltage filter capacitor to limit surge current into a short circuit.

B. STATIC LOAD TEST DATA

Experimental model tests were performed in the laboratory with the set up as shown in Figure VI-2. This shows the experimental model in the lower center of the picture. At the upper right is the three-phase, 230-volt, 60 cps variable autotransfromer used to supply adjustable input voltage. Slightly to the left is the three-phase, six-diode rectifier bridge used to obtain D.C. from the 60 cps source. Below the rectifier is the capacitor of a single L-section LC filter (the inductor is not shown) which was used to smooth the D.C. so that very

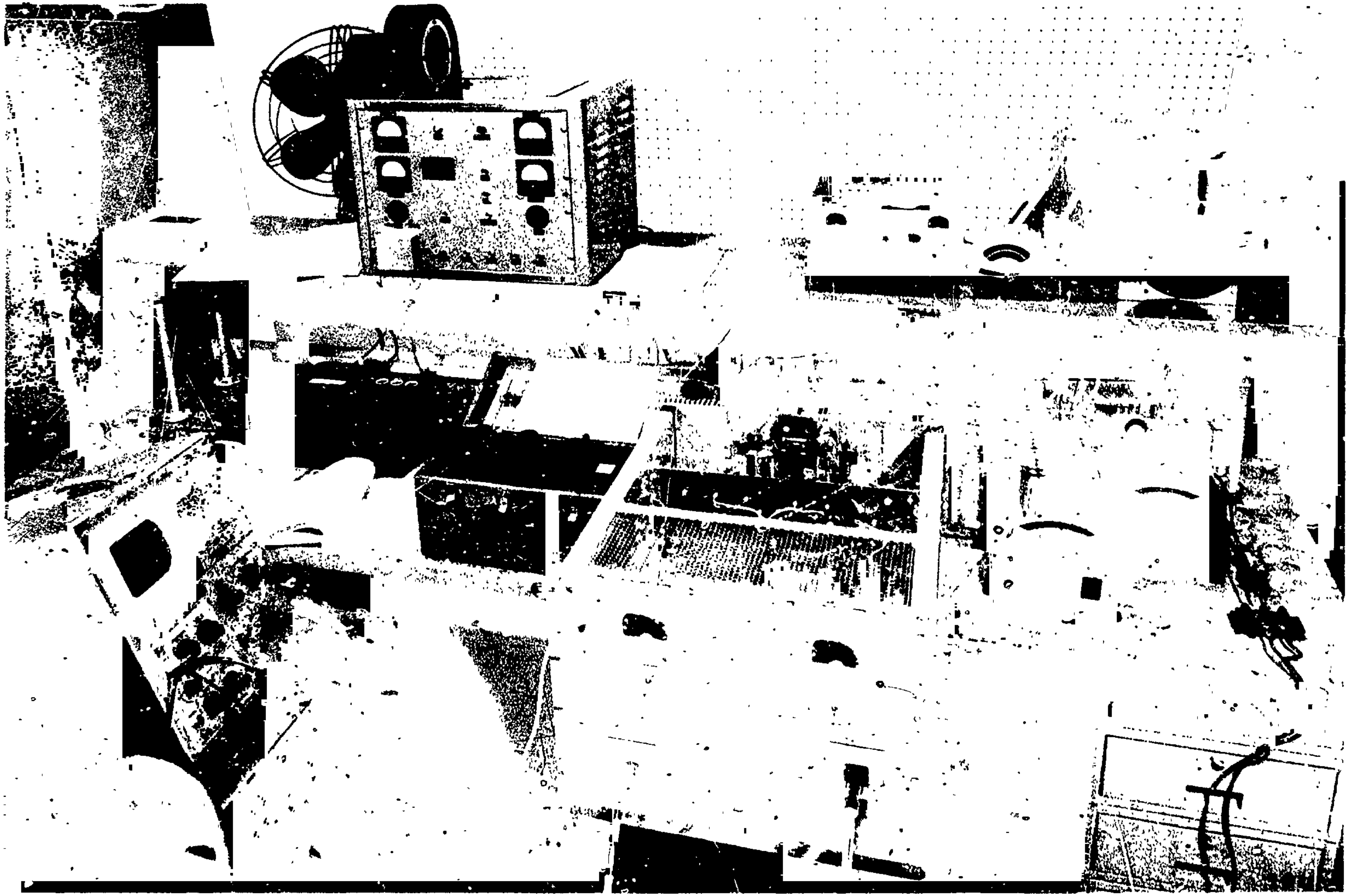
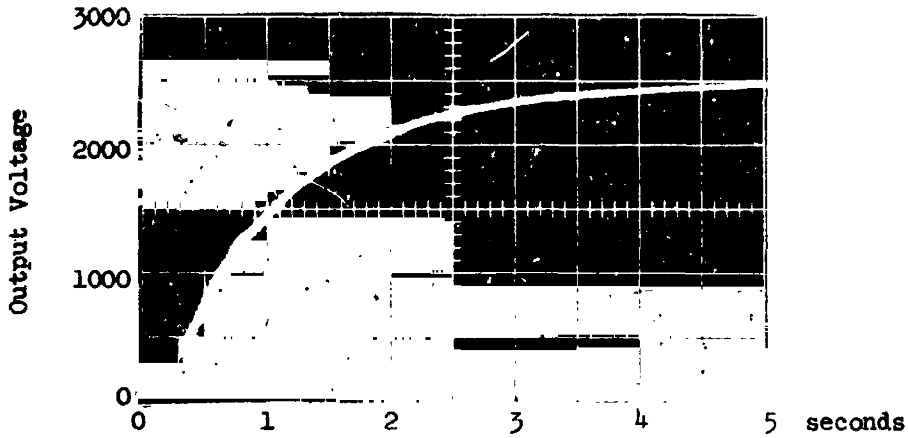


Figure VI-2. Laboratory Setup for Static Load and Electrical Stress Tests little ripple would appear at the input to the experimental model. Thus, the input current ripple measurements and circuit operation is not affected by the 60 cps power. To the left of the rectifier is the load simulator. The load simulator was assembled as part of Contract NAS 3-5918 and consists of a vacuum triode (RCA Type 7C24/5762) for variable resistive loads and a hydrogen thyratron (Amperex Type 6268) and a vacuum relay (Eimac Type VS-2) for application of short circuit transients. The following is a description of test data taken under static loading of the experimental model.

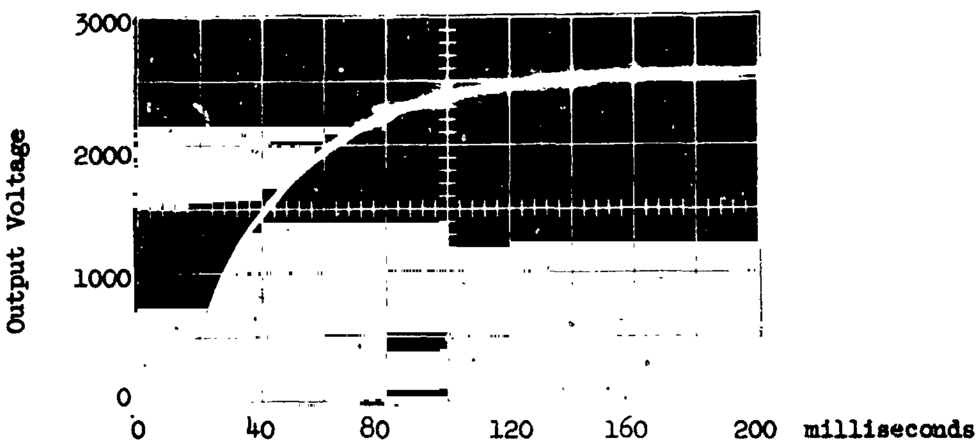
1. Soft-On Time Constant

Measurements of the limits of adjustment of the "Soft-On" time constant were made on the experimental model in the following manner. It was first set up and operated with full rated load at rated input voltage, and then shut down by the switch on the unit. Next, the "Soft-On" controls were set to the limit desired for the test. With a high voltage probe and oscilloscope the output voltages were measured. The oscilloscope was then set for internal D.C. level triggering of a single sweep and the sweep time adjusted for the "Soft-On" time constant under consideration. Single sweep oscillograms were then taken. The results of the "Soft-On" measurements are shown in Figures VI-3 and VI-4. The limits of the "Soft-On" time constant are 24 milliseconds minimum and 5000 milliseconds maximum. The "Soft-On" time constant is defined as the length of time it takes the output voltage to reach 2500 volts starting from zero volts.

Two things affect the minimum "Soft-On" time constant. Output filter components L4, R108, and C4 establish a certain minimum time constant of delay even to a step function of output voltage from the inverters. Furthermore, there is a slight time constant built into the application of drive current to the 2N1814 power transistors Q26 and Q29. This time constant was provided to prevent any starting malfunction of the inverters due to the relatively instantaneous switching of these transistors from the blocking state to the conducting state, and to prevent destructive voltage transients on the transistors themselves caused b instantaneously stopping the current in an

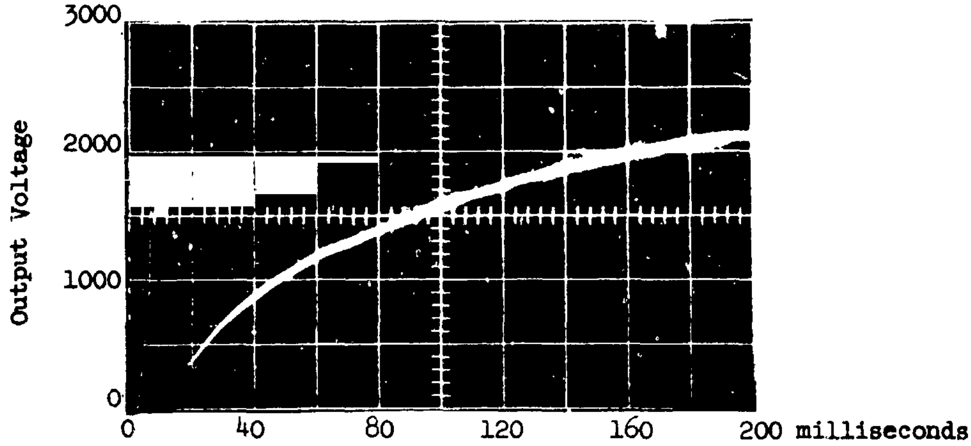


a. Output Voltage Rise (R75 = 100 K Ohms)

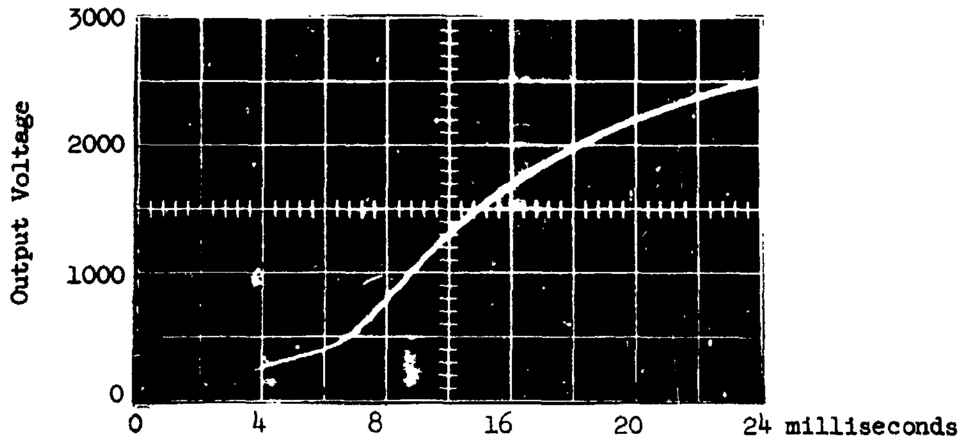


b. Output Voltage Rise (R75 = 0 Ohms)

Figure VI-3. Soft-On Time Constant Range with 54 Microfarad Capacitor C46 + C45
(Full Load on Experimental Model)



a. Output Voltage Rise (R75 = 100 K Ohms)



b. Output Voltage Rise (R75 = 0 Ohms)

Figure VI-4. Soft-On Time Constant Range
with 4 Microfarad Capacitor C45
(Full Load on Experimental Model)

inductive circuit. This was a precautionary measure and adequate test data has not been obtained to prove its worth. In summary, the minimum "Soft-On" time constant is affected by the output filter component values, which are fixed by other criteria, and the time constant built into the drive circuits of Q26 and Q29.

2. Output Voltage Regulation and Efficiency

For the minimum, the rated, and the maximum input voltage, the load was varied from no load to full load in appropriate increments to obtain voltage regulation and efficiency curves. The results of the efficiency data, Figure VI-5, shows a maximum full load efficiency of 84.5% (corrected for meter calibration error) at minimum input voltage. This value is lower than the value obtained with the breadboard, mainly because of the addition of the current limiting transistors Q26 and Q29, and negative feedback resistors R42 and R68. These components combined account for about 100 watts of losses or 4% reduction in efficiency.

The regulation curves for the various input voltages are shown in Figure VI-6. With minimum input voltage and full load (0.8 amps) the output voltage is 2490 volts. This value is above the minimum specified by the work statement. The regulation is within the specified limits for both load and input voltage variations with a no load upper limit of 2640 volts.

3. Input Current Ripple

The measurement of the input current ripple was carried out at full load (0.8 amps) and rated input voltage. The precision non-inductive current measuring resistor (made by T and M Research Products) was inserted in series with the input D.C. source to the experimental model. The resulting photographs appear as Figure VI-7. The peak-to-peak value of ripple current is 0.16 amps or 0.08 amps peak. This is well within the work statement limit of 5% peak or 0.77 amps peak. The rms value of input ripple current is only about 0.04 amps which

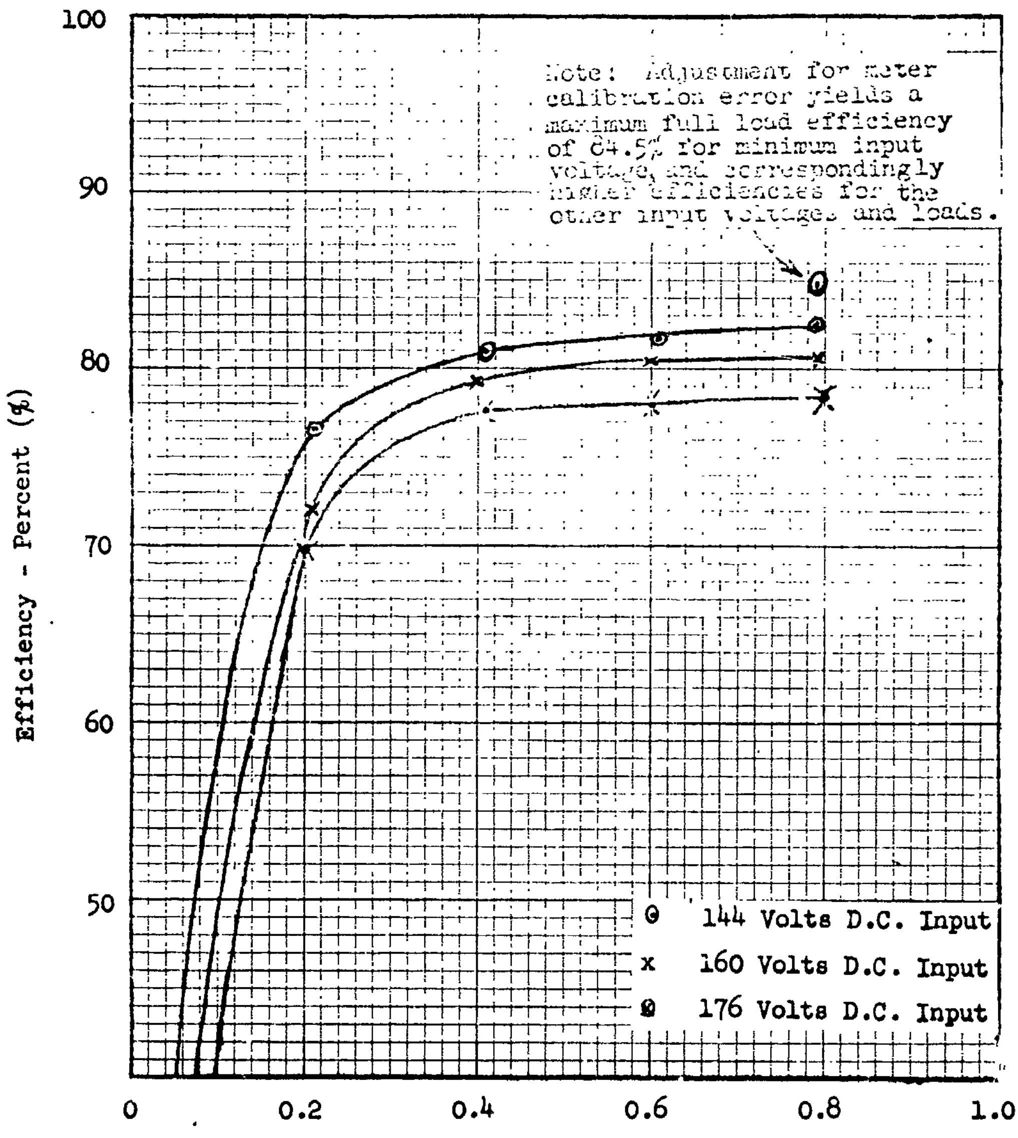


Figure VI-5 Experimental Model Efficiency Plotted as a Function of Load

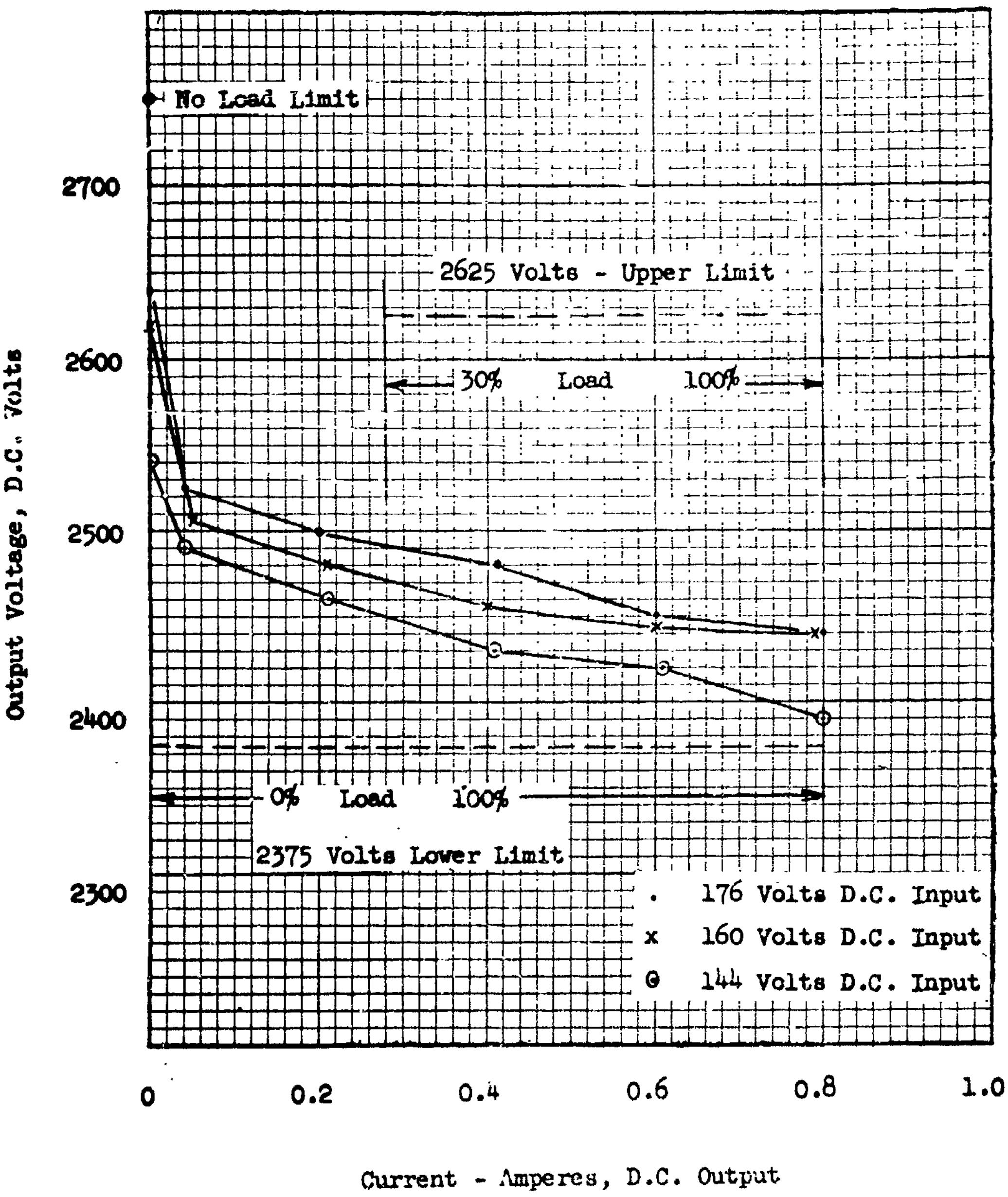
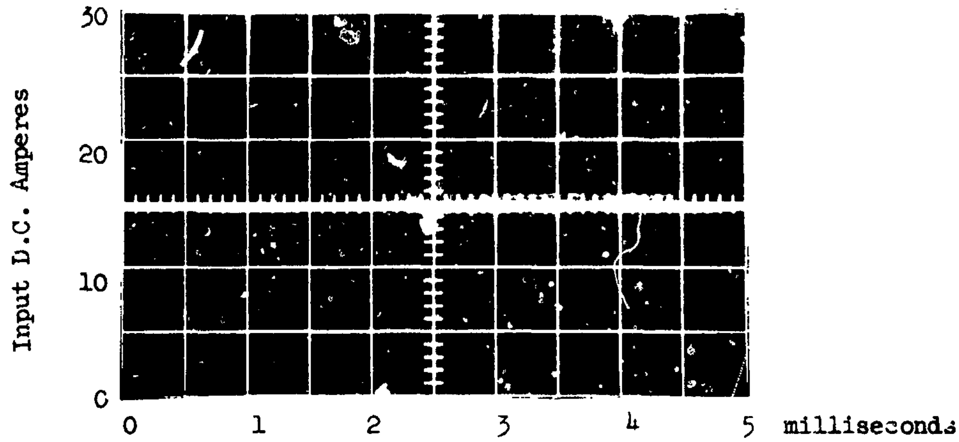
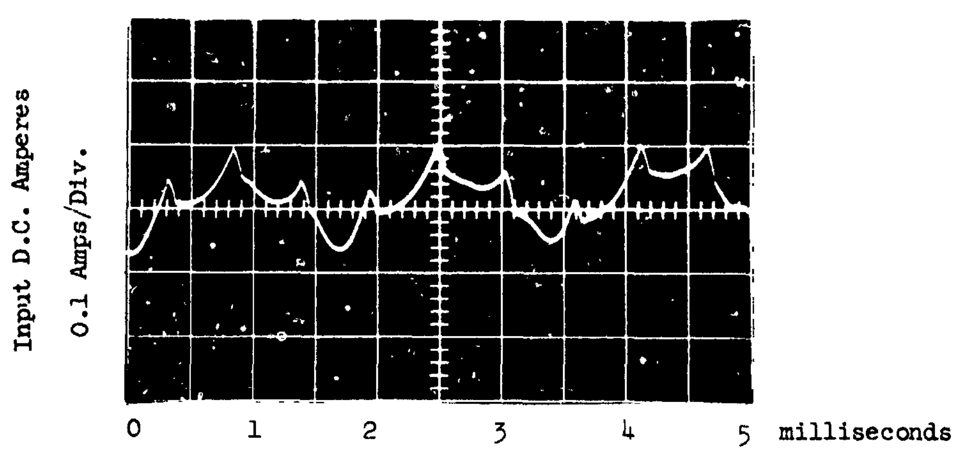


Figure VI-6 Experimental Model Output Voltage Plotted as a Function of Load



a. Full Load Input Current Ripple



b. Full Load Input Current Ripple

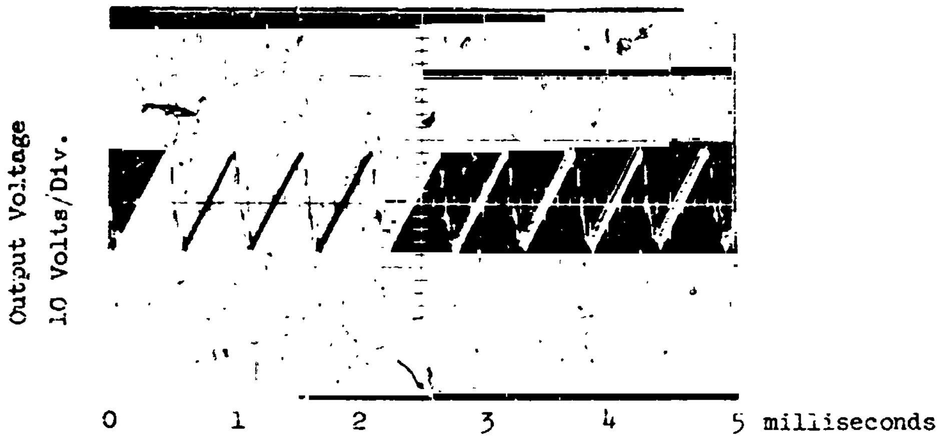
Figure VI-7. Input Current Ripple with Full Load on the Experimental Model is also within the specified limit of 2% rms or 0.31 amps. The very low value of input ripple current is the result of using commercial grade aluminum electrolytic capacitors having a two-to-one safety factor on the ripple current through each capacitor. This resulted in more capacitance than is necessary to make an adequate filter. Commercial grade capacitors were used because of their considerably lower cost and the desire to conserve funds for exploring the more important technical problems.

4. Output Voltage Ripple

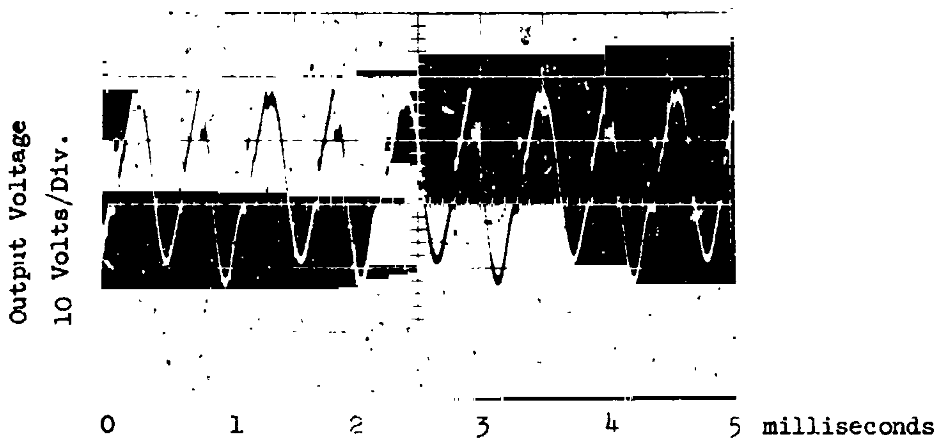
Output voltage ripple was measured at rated input voltage, for two load conditions. The results of these tests are displayed as Figure VI-8. The method was an oscilloscope connected across the terminals of a VTVM which was connected to read output ripple voltage. The D.C. component of output voltage was blocked by capacitors in series with the VTVM. The VTVM readings at no load and full load show a maximum rms ripple voltage of 10 volts at no load. This ripple amounts to 0.4% of the nominal output D.C.voltage and is very much below the 5% or 125 volts rms ripple allowed in the work statement. The ripple is very low because the output filter had been designed primarily to limit the rate of current build up in a short circuit load. These large values of filter components are no longer needed and could be considerably reduced to achieve a size and weight reduction.

C. COMPONENTS ELECTRICAL STRESS TEST DATA

The data and tables presented in this section show the safety factors applied to the electrical stresses of the components. The stresses of only the major semiconductor components are given. Other components stresses such as ressistors, capacitors, and semiconductors in circuits not affected by the power load transients have been omitted. The stress levels of these components have been designed to be less than one-half of their rating.



a. Full Load Output Voltage



b. No Load Output Voltage

Figure VI-8. Output Voltage Ripple with Full Load and No Load on Experimental Model

1. Component Safety Factors for 2 Kw Rated Load

For these measurements the experimental model was set up for full load (0.8 amps) operation at rated input voltage. The stresses of one component of each set of similarly stressed components were then measured. All current measurements were made with the T and M Research Products current viewing resistor. Voltages were measured with a standard oscilloscope probe, except for the high voltage rectifier where a high-voltage oscilloscope probe was used. The waveforms were displayed on a Tektronix Type 533 oscilloscope and the Type 53/54D preamp was used for all single trace voltage and current waveforms. Table VI-1 is a summary of the components examined for stress; the electrical ratings of the components to which the 2 to 1 safety factor was applied; and the measured maximum stress of that component.

The "critical components" of this system show a minimum safety factor of 1.67 to 1 for this electrical stress at rated load in the experimental model. The low safety factor of 1.25 to 1 for the peak gate power of GCS Q3 has been discussed with our Semiconductor Division personnel and it has been determined that the GCS gate power can be considerably higher than the published value without having any destructive effects. The high voltage rectifier is considered adequate for this application with a minimum safety factor of 2.5. The current regulating transistors (2N1814's) have a minimum safety factor of 1.6, but they perform their function with little trouble as evidenced by the endurance tests.

2. Component Safety Factors with Hydrogen Thyratron and Vacuum Relay Transient Breakdown Devices

The experimental model was operated in a similar manner and with the same equipment as that described above. The oscilloscope traces were observed during the interval of application of the hydrogen thyratron short circuit and again for the vacuum relay short circuit. The oscilloscope was set for a single sweep, triggered just prior to the closing of the short, so that an expanded view of the few cycles of inverter operation during the short might be obtained.

VI-1. COMPONENT SAFETY FACTORS WITH RATED LOAD

Component	Rating	Electrical Stress	Safety Factor
GCS Q1 (WX242ZP)	I_F Max-10 A	6 A RMS	1.67
	V_{FB} Max=700 V	400 V Pk	1.75
	T_J Max=125°C	75°C	50°C
	I_{GF} Pk=4 A	2 A Pk	2.0
	I_{GR} Pk=4 A	1.25 Pk	3.2
	P_{GM} = 10 W	6.0 W	1.67
	P_{GR} Av=5 W	.6 W	8.3
	I_{GR} Av=0.4 A	.03 A	13.3
GCS Q3 (WX242ZP)	I_F Max = 10 A	6 A RMS	1.67
	V_{FB} Max= 700	390 Pk	1.8
	T_J Max=125°C	77°C	48°C
	I_{GF} Pk=4 A	2 A Pk	2
	I_{GR} Pk=4 A	1.1 Pk	3.6
	P_{GM}= 10 W	8.0 W	1.25
	P_{GR} Av=5 W	.5 W	10
	I_{GR} Av=0.4 A	.03 A	13.3
Rectifier CR8 (FSPF80W)	PIV 8000 V	2800 V	2.9
	I_{AV} 2.0 A	0.8 A	2.5
Diode CR1 (478M)	PIV 600 V	410 V	1.5
	I_{AV} 6.0	0.35 A	17
Diode CR3	PIV 600 V	360 V	1.7
	I_{AV} 6.0 A	0.3 A	20
Transistor Q26	V_{CE} Max 300 V	20 V on 190 V off	1.6
	I_C Max=30 A	8 A	3.8
	P_C Watts 250 W	25 W	10
Transistor Q29 (2N1814)	V_{CE} Max 300 V	26 V on 190 V off	1.6
	I_{CE} Max=30 A	8 A	3.8
	P_C Watts 250 W	22 W	11.4

TABLE VI-1. COMPONENT SAFETY FACTORS WITH RATED LOAD
(Continued)

Component	Rating	Electrical Stress	Safety Factor
Transistor Q8 (2N1015F)	V_{CE} Max 300 V	190 V off	1.6
	I_C Max 7.5 A	0.5 A	15
Transistor Q30 (2N1015F)	V_{CE} Max 300 V	190 V off	1.6
	I_C Max 7.5 A	0.5 A	15
Diode CP67 (478M)	PIV 600 V	140 V	4.3
	I_{AV} 6.0	0.2 A	30
Diode CR68 (478M)	PIV 600 V	210 V	2.9
	I_{AV} 6.0	1.0 A	6.0
SCR Q6 (2N1777)	V_{FB} Max 400	160 V	2.5
	I_F AV 4.7	.3 A	16
Transistor Q9 (2N1016E)	V_{CE} 250 V	100 V	2.5
	I_C 7.5 A	.5	15
Transistor Q18 (2N1016B)	V_{CE} 100 V	40 V	2.5
	I_C 7.5 A	.25 A	30
Transistor Q32 (2N1016A)	V_{CE} 60 V	23 V	2.6
	I_C 7.5 A	1.0 A	7.5
Transistor Q23 (2N3054)	V_{CE} 90 V	29 V	3.1
	I_C 2 A	.05 A	40
Transistor Q24 (2N1039)	V_{CE} 40 V	34 V	1.3
	I_C 1	.02 A	50

The results results are summarized in Table VI-2. Data for only components directly affected by the short circuit are given. Worst case for all was used for the stress levels induced by the hydrogen thyratron or vacuum relay. However, the differences between the two types of short circuit are very slight.

The safety factor on GCS forward rms current had dropped to 1.5 from 1.67 at rated load. However, under the conditions of the short circuit some allowance should be made for the higher surge current rating of the GCS when determining this safety factor. The other component safety factors remain at permissible levels for these tests.

The components used in the experimental model were readily available manufacturer's catalog parts. It is quite possible that the manufacturer could supply higher voltage components by special request. This would improve the safety factors of the components showing a voltage stress in excess of the objective 2 to 1 derating.

D. ENDURANCE TEST DATA

The fulfillment of the work statement requirement of 10 hours of operation into a dummy load was accomplished by two endurance tests of the experimental model. Continuous full-load was applied to the experimental model for 7 hours and component temperature rises recorded. Cycled full-load operation and vacuum relay short circuit operation was performed for an additional 7 hours and again component temperatures were recorded. The combination of the continuous full-load and cycled short circuit load for a total of 14 hours was considered adequate to comply with the 10 hour requirement.

The maximum component temperature recorded was 121°C for the output filter inductor L4. This is a relatively low temperature compared to the permissible operating temperature for this component. The possible weight reductions in Table VI-3 include this consideration in determining the values stated. Semiconductor components maintained an adequate margin of safety, with the hottest of these components being the two power transistors (2N1814's) at 61°C and 63°C. The GCS's ran cool with a maximum recorded stud temperature of 52°C for Q3.

TABLE VI-2. COMPONENT SAFETY FACTORS WITH HYDROGEN THYRATRON AND VACUUM RELAY TRANSIENT BREAKDOWN DEVICES

Component	Rating	Electrical Stress	Safety Factor
GCS Q1 (WX242ZP)	V_{FB} Max 700 V	450 V	1.6
	I_F Max 10 A	6.5 A	1.5
GCS Q3 (WX242ZP)	V_{FB} 700 V	430 V	1.6
	I_F Max 10 A	6.5 A	1.5
Diode CR8 (FSPF80W)	PIV 8000 V	3000 V	2.7
	I_{AV} 2.0 A	.95 A	2.1
Diode CR1 (478M)	PIV 600 V	440 V	1.4
	I_A 6 A	2.0 A	3.0
Diode CR3 (478M)	PIV 600 V	400 V	1.5
	I_A 6 A	3.0 A	2.0
Transistor Q26 (2N1814)	V_{CE} Max 300 V	190 V	1.6
	I_C Max 30 A	8.5 A	3.5
Transistor Q8 (2N1015F)	V_{CE} Max 300 V	190 V	1.6
	I_C Max 7.5 A	0.5 A	15
Transistor Q29 (2N1814)	V_{CE} Max 300 V	190 V	1.6
	I_C Max 30 A	9 A	3.3
Transistor Q30 (2N1015F)	V_{CE} Max 300 V	190 V	1.6
	I_C Max 7.5 A	0.4 A	1.9
Diode CR67 (478M)	PIV 600 V	140 V	4.3
	I_{AV} 6.0	1.2 A	5.0
Diode CR68 (478M)	PIV 600 V	250 V	2.4
	I_{AV} 6.0	1.4 A	4.3

Although the transient loading during the cycled tests is severe, the resulting low duty cycle causes fewer average watts of dissipation in each device. The "Blink-Off" and "Soft-On" controls were adjusted for approximately 10 "Soft-On's" per second resulting in about 137,000 complete "Blink-Off" and "Soft-On" cycles during this test. There were also 310 short circuit transients in this 7 hours of testing. The actual total number of short circuit transients during the life of the experimental model is about 2000, including all tests performed on the inverter. The "Blink-Off" and "Soft-On" cycles exceed 200,000 for all inverter tests. There is no apparent degradation of components or circuit operation from any of these tests. The only noticeable change in operation during the full load endurance test was an increase in the noise level of the magnetic devices. This can be remedied in future power inverters by using epoxy cement to firmly bond the C-cores.

E. WEIGHT AND EFFICIENCY

A tabulation of the weight of all major components or groups of similar components in the experimental model is given in Table VI-3. The total component weight of the experimental model is 44.48 lbs.; this includes an estimated weight of 1.0 lb. for the small miscellaneous components not specifically listed in the table. Because of other problems, sufficient time was not available to optimize these components for the final circuit configuration (especially L4 which remains at the design used when the output current was regulated by the overcurrent regulator control circuit). The last column of the table points out estimated weight reductions in circuit components that might be achieved considering the new method of current regulating; the low temperature rise of components during the endurance tests; and other circuit changes pointed out in the conclusions and recommendations.

The distribution of total power loss of 420.2 watts among the various components is given in Table VI-4. At full load and rated input voltage the efficiency is 82.6% (corrected for meter errors).

Table VI-3. COMPONENT WEIGHT ANALYSIS

Line	Item	Component Description	Qty.	Weight/Item (lbs.)	Total Weight of Component Type (lbs.)	% of all Component Weight	Weight Subtotals	Percentage Subtotals	Possible Weight Reduction (lbs.)
1	L1	Input Choke	1	0.81	0.81	1.82	0.81	1.82	
2	L2,L3	Line Choke	2	1.75	3.5	7.87	4.31	9.69	3.0
3	L4	Output Choke	1	16.19	16.19	36.35	20.5	46.04	10.0
4	L5	Chopper Choke	1	0.5	0.5	1.12	21.0	47.16	
5	L6	Chopper Choke	1	1.25	1.25	2.81	22.25	49.97	0.25
6	L7	Saturating Choke	1	0.25	0.25	0.56	22.5	50.53	0.1
7	L8,L9	Inverter Choke	2	0.1	0.2	0.45	22.7	50.98	
8	T1,T2	Output Transformer	2	4.25	8.5	19.1	31.2	70.08	1.0
9	T3,T4	Driver Transformer	2	0.69	1.38	3.10	32.58	73.18	0.5
10	T5,T7,T8	Pulse Transformer	3	0.12	0.36	0.81	32.94	73.99	
11	T6	Oscillator Transformer	1	1.25	1.25	2.81	34.19	76.80	0.5
12	T9	Soft-On Transformer	1	0.2	0.2	0.45	34.39	77.25	0.1
13	T10	Drive Transformer	1	0.2	0.2	0.45	34.59	77.70	0.15
14	C1	Input Capacitor	2	1.39	2.78	6.25	37.37	83.95	2.0
15	C2,C3	Inverter Capacitor	2	0.125	0.25	0.56	37.62	84.51	
16	C4	Output Capacitor	1	0.63	0.63	1.41	38.25	85.92	
17	C8	Chopper Capacitor	1	0.88	0.88	1.98	39.13	87.90	0.65
18	C34	Voltage Reg. Capacitor	1	0.2	0.2	0.45	39.33	88.35	0.15
19	C38,C39	Driver Capacitor	2	0.87	1.74	3.91	41.07	92.26	1.25
20	CR5,6,7,8	High Voltage Rectifier	1	0.2	0.2	0.45	41.27	92.71	
21	Q26,Q29	Transistors 2N1814	2	0.5	1.0	2.25	42.27	94.96	
22	Q8,9,10 17,18,19,21 30,32	Transistors 2N1016 (Series)	9	0.065	0.585	1.32	42.855	96.28	0.35
23		All 25 watt Resistors	19	0.033	0.627	1.41	43.482	97.69	0.41
24		All Other Components		1.0	1.0	2.25	44.482	99.94	20.41

TABLE VI-4. COMPONENT LOSS ANALYSIS

Line	Item	Component or Circuit Description	Qty.	Loss/Item or Circuit (watts)	Total Loss of Components or Circuits (watts)	% of all Component Losses
1	CR36,97,98,99 Q1,2,3,4	G.C.S. + Diodes	4	17.0	68.0	16.2
2	L4	Output Choke	1	60.0	60.0	14.3
3	T1,T2	Output Transformer	2	21.0	42.0	10.0
4	R57,58,59, 61,62	Voltage Feed Back Resistance Bleed.	1	13.0	13.0	3.1
5	CR5,6,7,8 R108	High Voltage Rectifier High Voltage Capacitor Resistor	1	13.0 2.0	13.0 2.0	3.1 0.5
6	M2	Meter Resistor	1	2.0	2.0	0.5
7	R8, R9, R60,65, R11,12, 66,67	Energy Absorbing Resistors	8	9.2	73.6	17.5
8	L1	Input Choke	1	5.0	5.0	1.2
9	L5	Chopper Inductor	1	0.5	0.5	0.1
10	L6	Chopper Inductor	1	7.5	7.5	1.7
11	T6	Oscillator Transformer	1	12.0	12.0	2.9
12	All Others	Control Circuits and Miscellaneous other Components		20.0	20.0	4.8
13	Q8,26,29,30	Series Transistors	2 Pairs	34.0	68.0	16.2
14	R42, R68	Current Sensing Resistors	2 Pairs	16.8	33.6	8.0
					420.2	100.0

Since the GCS's for the experimental model are mounted on heatsinks different from those of the breadboared, a typical temperature rise versus power dissipation was measured again. The results of this measurement is shown in Figure VI-9. From this data it was determined that the GCS loss is about 17 watts per device. The other component losses were measured or calculated as required. The series transistors (Q26 and Q29) and their resistors (R42 and R68) account for a total of 101.6 watts or 24.2% of all circuit losses.

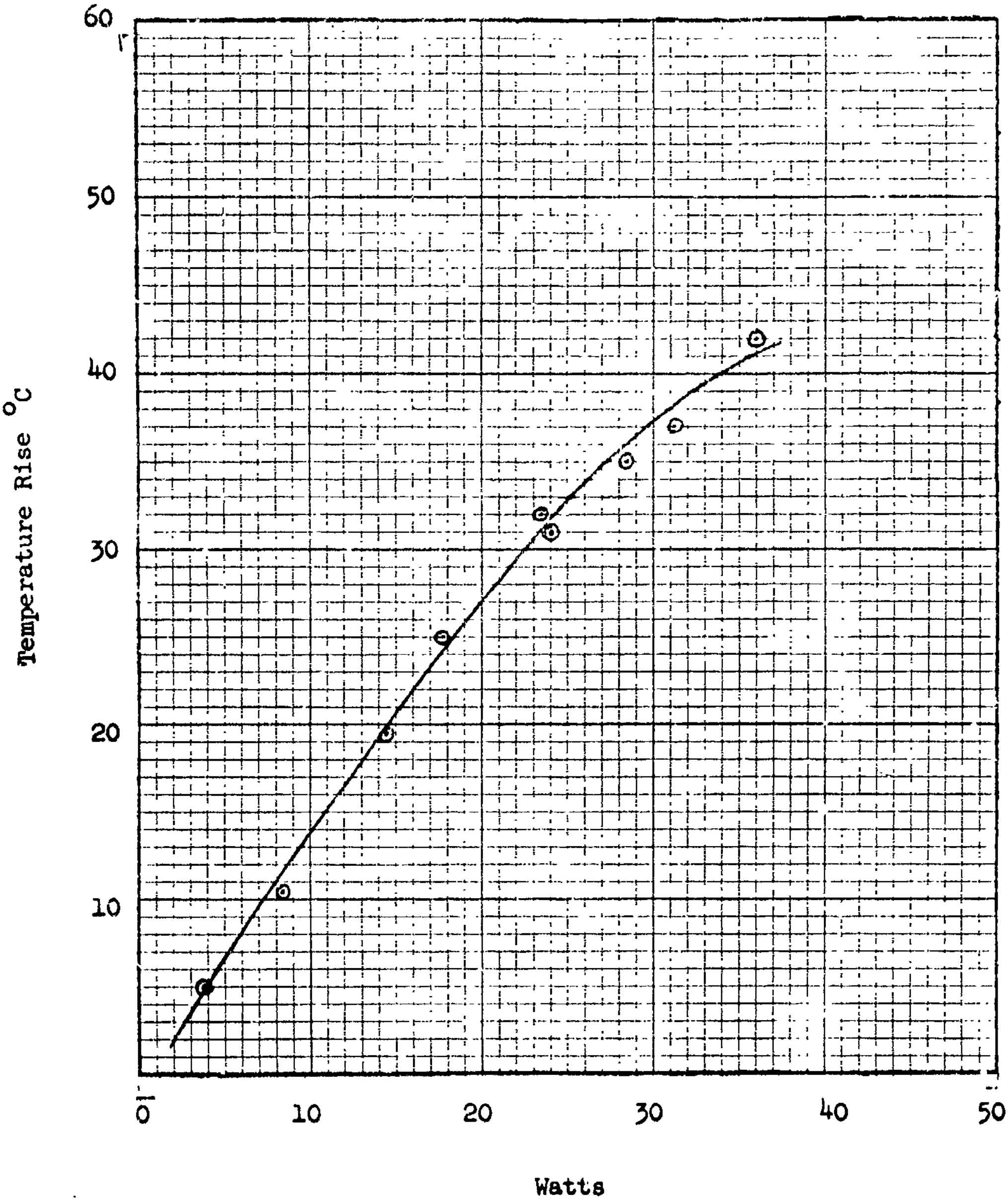


Figure VI-9. Temperature Rise vs. Power Dissipated for A Typical GCS on Its Heatsink

VII. CONCLUSIONS AND RECOMMENDATIONS

Final test and evaluation of the ion engine thrustor power supply employing gate controlled switches shows that the originally specified performance requirements can be very closely matched. The techniques of using two inverters operated with a variable phase shift has proven to be an effective means of converting power from low-voltage D.C. to high voltage, regulated D.C.

For the experimental model a safety factor of 2 to 1 was a design goal. With currently available GCS's a minimum safety factor of about 1.5 on current and 1.6 on voltage was obtained. The power transistors have a safety factor of about 1.5. Since these components were obtained in the highest ratings available any improvement in safety factor would require use of parallel or series devices or changes in the input voltage and the number of inverter stages used.

The ability of the inverter to operate continuously at full load and through a large number of short circuit "Blink-Off" cycles proves the reliability of components and circuit operation. There were no component failures or malfunctic ... the inverter during the experimental model tests.

The experimental model does not fulfill original expectations with regard to weight and efficiency. The weight (44 pounds) may be reduced to about 24 pounds and the efficiency of 82.6% may be raised as high as 90%. By eliminating the input filter and the overcurrent time delay and increasing the frequency it is estimated that the component weight could be further reduced to about 12 pounds with the efficiency still near 90%.

During the program a number of problems were encountered in utilizing the gate controlled switch. These problems arose because this was a new device being applied in a new application and because the device characteristics were subject to change as improved device fabrication techniques were being employed. To cope with these problems various circuits were added to the system during development and a very conservative design approach was taken. Due to lack of time it was not possible to evaluate the need for these added circuits or to redesign components which were excessively large. Consequently the experimental model does not represent the ultimate capabilities of a GCS inverter system for this application.

In the present system the following potential improvements exist:

a. Elimination of transistors Q26 and Q29; this would remove approximately 20% of the losses.
b. Reduction of output filter inductor to value needed to meet only the ripple requirement (this assumes that no overcurrent time delay is required); this would reduce the weight by about 10 pounds.
c. Reduce or possibly eliminate the inverter input inductors.
d. Simplification and reduction of complexity of low power control circuits.

To fully exploit the capabilities of the GCS, further circuitry development should be conducted. The high speed switching capabilities of the devices as observed during this program indicate that new circuits which use the devices at high frequency should be investigated. Circuits employing inductive discharge, for example, offer a simple and effective means of converting low voltage D.C. to high voltage D.C. This technique and other new techniques can offer smaller and more efficient conversion systems for space application.

APPENDIX A. DESCRIPTION OF EXPERIMENTAL MODEL OPERATION

The 2 kw high voltage power supply utilizes two GCS inverters to provide a regulated output voltage of 2500 volts D.C. ± 5 percent when connected to a supply voltage of 160 volts D.C. ± 10 percent. (See Figure App. B-1). The input voltage is converted by one pair of GCS's Q1 and Q2 to approximately 1400 volts square wave output at 1000 cycles per second. The second pair of GCS's Q3 and Q4 also provide a 1400 volt square wave output; however, they are driven slightly out of phase with the first pair. The secondaries of the output transformers T1 and T2 are connected in series to produce a 2500 volt square wave A.C. This voltage is then rectified by a high voltage bridge rectifier (CR5, 6, 7, and 8) and filtered by an LC filter consisting of L4 and C4 to produce the desired D.C. output voltage. Voltage regulation is achieved by phasing back GCS's Q3 and Q4.

1. Low Level Control Voltage

A low level D.C. voltage for control and protection functions is obtained by using an SCR Q6 in a D.C. chopper circuit. The drive for the SCR is produced by a Shockley diode CR12 discharging a capacitor C5 through a transformer T5; diode CR11 dissipates the reverse energy and allows the circuit to recover. The secondary of T5 feeds the gate of SCR Q6 which turns on at a rate dependent on the Shockley diode circuit oscillation frequency. The SCR is turned off again by the voltage reversal across C7 due to the oscillation of C7 and L5. Smoothing of the chopped pulses is accomplished by L6 and C8. A voltage reference is provided by Zener diode CR15, and an error signal is fed through R6 into the base of transistor Q7, which shunts R3. The amplified error signal varies the voltage at the junction of R1 and R2 to control the oscillation rate of the Shockley diode circuit. The output D.C. voltage is thereby regulated to about 50 volts by Zener diode CR15.

Upon application of voltage to the input terminals, the free-running Shockley diode oscillator provides 50 volts D.C. to the master oscillator, which oscillates at approximately 1000 cycles per second. The frequency is controlled by a saturable reactor L7 which reverses the drive to the transistor bases when it saturates. Taps are provided on L7 to change the frequency. The oscillator consists in part of transistors Q9 and Q10, and transformer T6. Transformer T6 provides various output voltages for drive and control purposes. SCR Q27 allows all the D.C. voltages to rise to full value and establish the correct drive to the inverters before turning on the series transistors Q26 and Q29. Shockley diode CR102 turns on SCR Q27 when the positive D.C. supply reaches a sufficient level to operate the circuits.

2. Main Power stages and Output Circuit

Power transformers T1 and T2, gate controlled switches Q1, Q2, Q3 and Q4 are the main power stage components. L4 and C4 serve as output filter. In the power stages Q1 and Q2 are driven on and off by transformer T3 and silicon transistors Q17 and Q18. Drive is provided for these transistors directly from the master oscillator. The other pair of GCS's Q3 and Q4 are driven by transformer T4 and silicon transistors Q19 and Q21. Drive is provided for these transistors from a phase shifting circuit consisting of a double Schmitt trigger and a flip-flop circuit. When both inverters are operating into a normal full load (i.e. 800 mA), the bridge consisting of Q3 and Q4 is being switched on and off almost in phase with bridge Q1 and Q2 so the outputs of transformers T1 and T2 are added thus giving an output of 2500 volts. Output voltage regulation is achieved by varying the phase shift between the two inverters. This is discussed in the section on Voltage Regulation.

The secondaries of the power transformers are fed into a full wave rectifier bridge CR5-8. The rectified output is filtered by the filter consisting of L4 and C4. Resistor R108 is in series with C4 to limit the discharge rate of C4 when a short circuit occurs at the output terminals.

To help commutation, capacitors C2 and C3 have been added across the primary of each transformer; however, to prevent the current from rising at an excessive rate through the GCS's, inductors L8 and L9 have been added. Also resistors R8, R9, R11 and R12 are added to damp the oscillations which tend to occur. A network across each GCS, as shown by CR4, R100 and CR96, reduces the rate of voltage rise across the GCS and aids the turn-off. When Q1 has been conducting and is being turned off so that it starts to block voltage, C41 starts to charge through diode CR96, and the anode voltage slowly rises to the blocking voltage. The current is thus diverted from the GCS to the capacitor during the switching transient.

Inductors L2 and L3 were used to provide a limiting impedance between the source and inverters to prevent excessive current in the GCS's during switching. Their use required the addition of components to dissipate their stored energy. Stored energy is dissipated in R60, R65, R66, and R67. Diodes CR67, CR68, and CR1 through CR4 provide the path for discharging the energy of the inductors. After the series transistors Q26 and Q29 were added, the need for the inductors and the added components became questionable. Series transistors Q26 and Q29 protect the power supply by series regulation of inverter D.C. current. They are normally in a saturated condition by application of a positive bias on their bases. The absolute maximum potentials of the bases of Q8 and Q30 above ground are held fixed by CR72-CR71 and CR86-CR85, respectively. When excessive inverter current increases the voltage drop across R42 or R68, the emitter voltage of Q26 or Q29 approaches the limit imposed on the base potential and thereby removes drive from these transistors, pulling them out of saturation and regulating the inverter current.

3. Phase Control Network

The magnitude of the output voltage is determined by the phase angle between the two inverters. For maximum output GCS's Q3 and Q4 are running practically in phase with Q1 and Q2. For zero output GCS's Q3 and Q4 are running 180° of phase with GCS's Q1 and Q2 so the outputs of the transformers are in opposition. The phase difference between inverters is controlled by the D.C. voltage applied to the input of the double Schmitt trigger circuit. A square wave output is fed from the master oscillator via a winding on the oscillator transformer to an integrating network consisting of R16 and C16; or its counterpart R19 and C17, which operates on the negative half-cycle. Capacitor C16 charges through R16 to form the ramp voltage which is fed to the base of transistor Q11. When this voltage exceeds by 0.5 volt, the voltage developed across R24, due to current flowing through Q12, Q11 starts to conduct. Voltage applied to the base of Q12 is reduced by the current flowing through R21 and Q11, which reduces the voltage dropped across R23. Q11 therefore rapidly turns on and Q12 rapidly turns off. As long as the voltage applied to the base of Q11 remains above this triggering voltage, Q11 stays on and Q12 remains off. The voltage from the oscillator must reverse to allow this potential to fall and reset the trigger. When the voltage reverses, C17 charges through R19 and actuates the other Schmitt trigger consisting of Q13 and Q14.

The D.C. voltage fed into the center tap of the transformer T6 winding modified the rate at which C16 or C17 charges. Therefore, the trigger can be either delayed or advanced depending upon whether the input D.C. voltage is positive or negative. A positive voltage is normally fed into this point which keeps the Schmitt triggers almost at the point of triggering. A small voltage from the positive going square wave is all that is necessary to activate the trigger. The Schmitt triggers therefore are running practically in phase with the master oscillator. This control voltage is supplied at the emitter of the transistors Q23 and Q24. A negative voltage being fed into the bases of Q23 and Q24 turns off Q23 and turns on Q24 to make the voltage at the center tap of the transformer more negative. This delays the rise of the voltage ramp being fed to the Schmitt triggers and, therefore, phase shifts the square wave. This is the normal action when an over-voltage signal is being detected. Negative going signals are fed from the voltage regulator control or from the "Soft-On" control. Signals from either of these sources are negative and increase the phase difference. The ramp then has to start at this large negative voltage and build up to the triggering voltage of the Schmitt trigger. The ramp starts steeply but as capacitor C16 charges, the slope decreases and the ramp voltage may fail to reach the triggering level. To prevent this, a spike was added at the extreme end of the ramp to insure that the trigger did operate and produce a 180° phase shift. The spike is formed by taking a signal from the opposite side of the drive winding and differentiating it with C15 and R18, or C13 and R17. This feature insures that both Schmitt triggers operate correctly and that they cannot be driven past the 180° phase shift position.

4. Phase Control Flip-Flop Circuit

When Q11 or Q13 are driven on, as previously described, a negative going signal is fed through capacitor C20 or C21 to the flip-flop circuit comprised of transistors Q15 and Q16. The positive going signal which is evident when Q11 or Q13 are off is suppressed by diode CR44 and CR45. Assume Q16 in the flip-flop is conducting. A negative going signal fed to its base through C22 would turn it off and as Q15's base is fed from the collector of Q16, it would turn on. A negative going signal being fed through C25 reverses the above. The flip-flop is therefore being driven on and off in accordance with signals from the Schmitt triggers. Note that only the leading edge of the original oscillator waveform is being used for control. The phase shifting network is therefore independent of any noise or spurious signals as the leading edges of the square waves are the only signal which will activate the network.

5. Isolation and Driver Circuit (Phase Shifted Pair)

Transistors Q36 and Q38 are driven by the voltage developed across R32 and R35. Four diodes CR90-94 and two resistors R52 and 53 provide isolation and prevent any feedback from the following power stages from causing malfunction of the flip-flop circuit. Q36 and Q38 drive a pair of power transistors Q19 and Q21 through an impedance matching transformer T10.

6. Inverter Drive Circuits

Inverter 1 is driven by a pair of power transistors Q17 and Q18 through transformer T3. Drive for these transistors is supplied by a winding on the oscillator transformer. Transformer T3 has two secondary windings; one winding supplies a full half cycle positive gate bias current of about 400 milliamperes to the gates of the GCS's through the networks consisting of R44, CR35, R45, CR36, R46 and R47. The other winding supplies a sharp spike current of about 2 or 3 amps at the beginning of each half-cycle to both GCS gates to ensure fast turn-on and turn-off of the GCS's. This current is supplied through C28, R38, C27, a bridge rectifier consisting of 4 diodes CR30-33; and a Shockley diode CR34. Voltage across the bridge builds up when the transformer square wave of voltage switches until the breakover voltage of the Shockley diode CR34 is reached; then C27 and C28 rapidly discharge into the gates of Q1 and Q2, turning one on and the other off. Similar action takes place in inverter 2 with T4 supplying the drive through identical circuits.

7. Voltage Regulator

To obtain a feedback signal for voltage regulation the output voltage is measured across resistors R57, R58, R59, R61, R62. A portion of this voltage is tapped off and fed to a Shockley diode relaxation oscillator consisting of R63, C37 and CR58. Capacitor C37 charges until the break-down voltage of the Shockley diode is reached and then it discharges through the primary winding of pulse transformer T7. Diode CR59 dissipates the inductive energy from the transformer. The pulse repetition rate of the Shockley diode oscillator is dependent upon the magnitude of the voltage and the RC time constant. Potentiometer R62 sets the operating voltage of the oscillator and thus sets the point at which output voltage regulation begins. The output of transformer T7 is rectified by CR57 and smoothed by C35, and then fed to the base of transistor Q25 through a resistor R55 and Zener diode CR56. As the output voltage rises the repetition rate of the Shockley relaxation oscillator increases causing the voltage applied to the base of Q25 to increase. Current from Q25 charges capacitor C34 and causes a larger voltage drop across R15, which drives the bases of Q23 and Q24 more negative. The phase angle is thus increased and consequently reduces the output voltage. C35 and C34 provide appropriate filtering to prevent oscillation. Resistor R64 is set at full CCW which allows the control transistor Q23 and Q24 to operate the Schmitt triggers almost in phase with the input square wave.

8. "Blink-Off" and "Soft-On" Control

When any condition exists which would cause excessive current to flow in either inverter the circuit will automatically "blink-off." The "blink-off" action is initiated by the voltage drop appearing across transistors Q26 and Q29. If the voltage from collector to emitter of either transistor exceeds the breakover voltage of Shockley diode CR89, a pulse is fed through transformer T8 to trigger the "blink-off" time delay circuit, to turn-off transistors Q26 and Q29, and to prepare the "soft-on" circuit for the next cycle. (See Figure App. A-1). Transistors Q26 and Q29 are turned off by SCR Q37 firing and turning off Q4 which then turns on transistor Q32. When Q32 is on, the drive voltage to transistors Q26 and Q29 is reversed because of the path created through CR87, Q32, R86 and R87, and the emitter-base junctions of Q26 and Q29. Also when SCR Q37 is on Q34 turns off and C50 charges at a rate determined by R91, R92, and R109. When the firing point of unijunction transistor Q33 is reached, it fires and turns on Q35, allowing SCR Q37 to recover to its blocking state. This action ends the "blink-off" period.

Returning to the point when Q32 was conducting it should be noted that Q32 also allows the oscillator voltage to be applied to transformer T9. (See Figure App. B-1). The output of T9 charges capacitors C45 or C46 to establish the voltages required for the "soft-on" period. As soon as voltage appears across C45 or C46, transistor Q28 begins to conduct. This causes the control transistors Q23 and Q24 to change the D.C. level of Schmitt trigger and thereby increase the phase angle between inverters. The phase angle is held at 180° until the end of the "blink-off" period. When Q32 turns off transformer T9 is no longer energized and the voltage of C45 or C46 begins to decay through R75 and R76. As this voltage decreases transistor Q28 begins to turn-off thus changing the control signal which sets the phase angle. The setting of R75 and selection of either C45 or C46 determines the rate at which the phase angle changes and thereby sets the rate at which the system output voltage increases. This completes the "soft-on" cycle. If an overcurrent condition exists during or immediately after the "soft-on" period the unit will again "blink-off" and the cycle will be repeated.

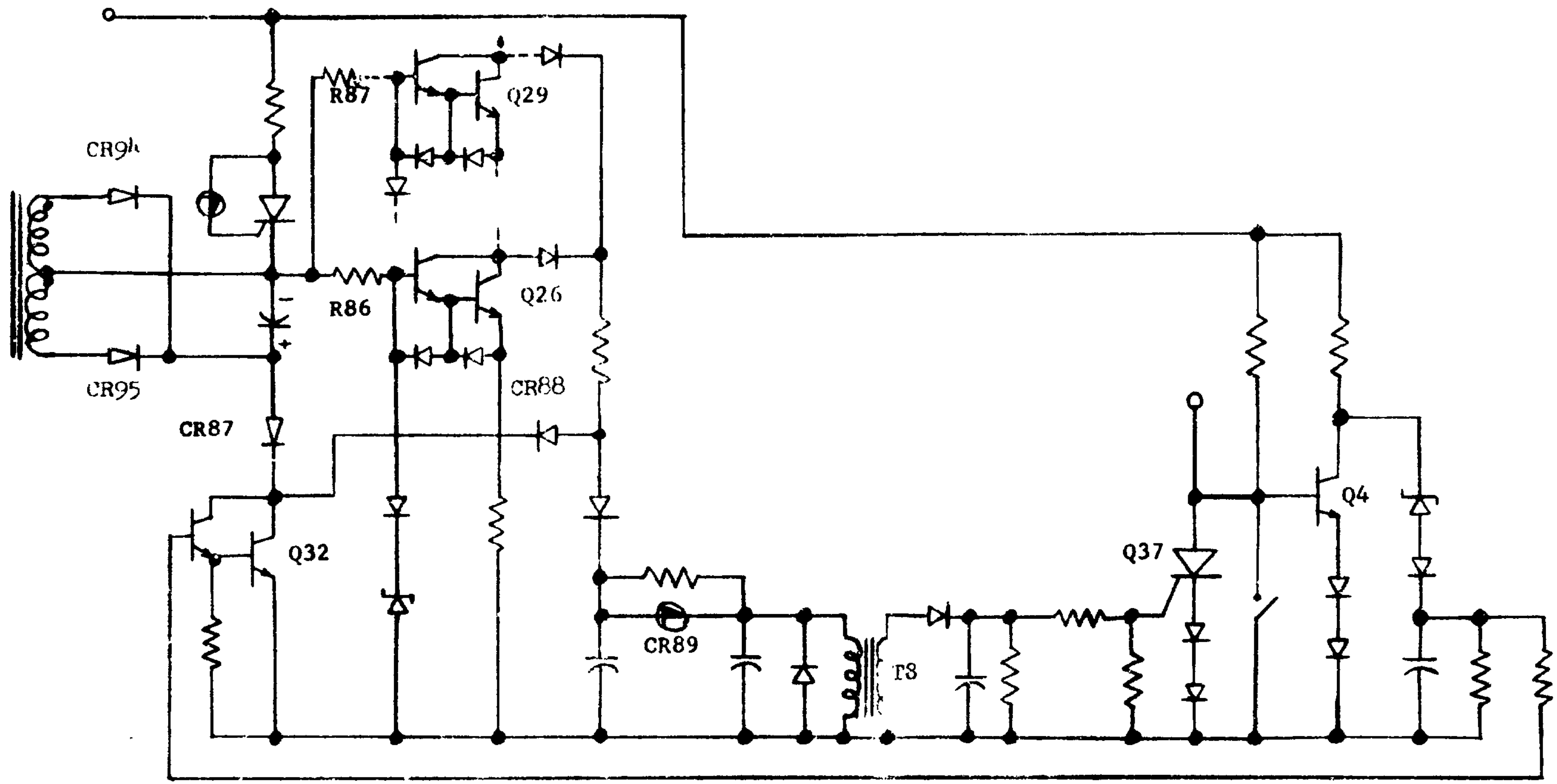


Figure App. A-1 Circuit for "Blink-Off"

APPENDIX B

EXPERIMENTAL MODEL SCHEMATIC AND PARTS LIST

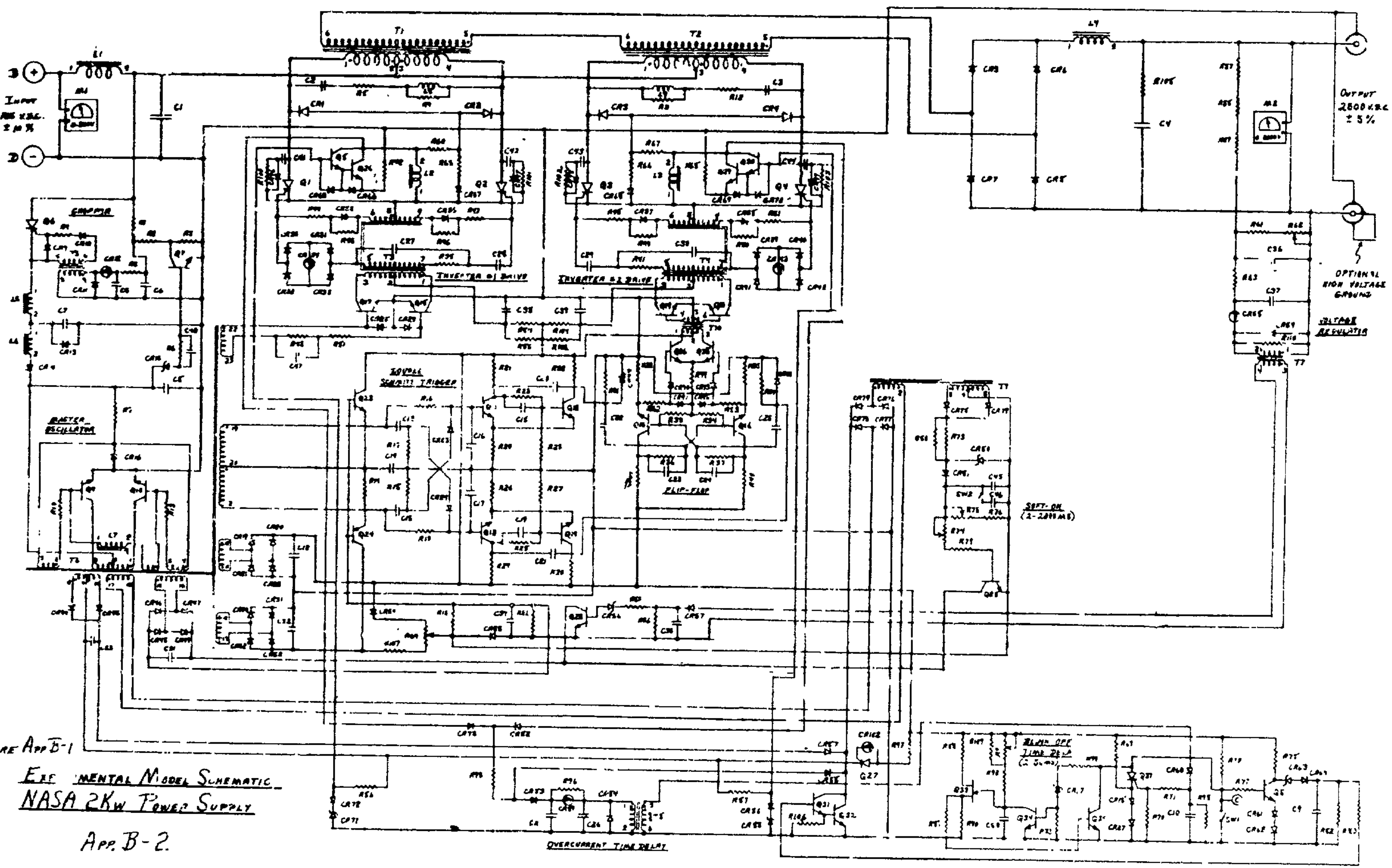


FIGURE APP B-1

EXPERIMENTAL MODEL SCHEMATIC

NASA 2KW POWER SUPPLY

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY

Quantity	Component Number	Capacitance (μfd)	Volts (w.v.d.c.)	Material	Manufacturer
2	C1	1000	350	Alum. Elec.	Sangamo
1	C2	0.05	1000	Paper	Aerovox
1	C3	0.05	1000	Paper	Aerovox
1	C4	0.25	5000	Paper	Dearborn
1	C5	0.15	200	Paper	Sprague
1	C6	50	50	Alum. Elec.	Cornell-Dubilier
2	C7	1.0 (2 reqd.)	600	Paper	Electron Products
1	C8	1500	100	Alum. Elec.	Sangamo
1	C9	4	150	Alum. Elec.	Cornell-Dubilier
1	C10	0.1	200	Paper	Aerovox
1	C11	0.25	200	Paper	Aerovox
1	C12	50	50	Alum. Elec.	Cornell-Dubilier
1	C13	0.1	200	Paper	Aerovox
1	C14	2.0	100	Paper	Electron Products
1	C15	0.1	200	Paper	Aerovox
2	C16	1.0	200	Paper	Electron Products
2	C17	1.0	200	Paper	Electron Products
1	C18	0.25	200	Paper	Aerovox
1	C19	0.25	200	Paper	Aerovox
1	C20	0.005	600	Paper	Aerovox
1	C21	0.005	600	Paper	Aerovox
1	C22	0.001	600	Paper	Aerovox
1	C23	0.05	200	Paper	Aerovox
1	C24	0.05	200	Paper	Aerovox

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Capacitance (µfd)	Volts (w.v.d.c.)	Material	Manufacturer
1	C25	0.001	600	Paper	Aerovox
1	C26	0.01	200	Paper	Aerovox
1	C27	0.47	200	Paper	West.-Cap.
1	C28	0.47	200	Paper	West.-Cap.
1	C29	0.47	200	Paper	West.-Cap.
1	C30	0.47	200	Pape	West.-Cap.
1	C31	10	50	Alum. Elec.	Cornell-Dubilier
1	C32	50	50	Alum. Elec.	Cornell-Dubilier
1	C33	100	25	Alum. Elec.	Cornell-Dubilier
1	C34	1500	35	Elec.	Sangamo
1	C35	50	50	Alum. Elec.	Cornell-Dubilier
1	C36	2.0	200	Paper	Aerovox
1	C37	0.1	200	Paper	Aerovox
1	C38	8000	50	Elec.	Sangamo
1	C39	8000	50	Elec.	Sangamo
1	C40	2	100	Paper	Electron Prod.
1	C41	0.05	600	Paper	Aerovox
1	C42	0.05	600	Paper	Aerovox
1	C43	0.05	600	Paper	Aerovox
1	C44	0.05	600	Paper	Aerovox
1	C45	4.0	150	Paper	Cornell-Dubilier
1	C46	50	50	Alum. Elec.	Cornell-Dubilier
2	C47	2.0	100	Paper	Electron
	C4	Not Used			
	C49	Not Used			
1	C50	4	150	Alum. Elec.	Cornell-Dubilier

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY

(Continued)

Quantity	Component Number	Resistance	Watts	Material	Manufacturer
1	R1	10 K	2 W	carbon composition	
1	R2	1.5 K	1/2 W	carbon composition	
1	R3	15 K	1 W	carbon composition	
1	R4	100	1 W	carbon composition	
1	R5	9.1 K	1/2 W	carbon composition	
1	R6	2 K	1/2 W	carbon composition	
1	R7	5.1 K	1 W	carbon composition	
1	R8	10	20 W	wire wound	Ohmite
1	R9	100	5 W	wire wound	Daleohm
1	R10	20	2 W	carbon composition	
1	R11	100	5 W	wire wound	Daleohm
1	R12	10	20 W	wire wound	Ohmite
1	R13	20	2 W	carbon composition	
1	R14	51	1 W	carbon composition	
1	R15	1 K	1 W	carbon composition	
1	R16	500	10 W	wire wound	Daleohm
1	R17	1 K	1/2 W	carbon composition	
1	R18	1 K	1/2 W	carbon composition	
1	R19	500	10 W	wire wound	Daleohm
1	R20	10 K	2 W	carbon composition	
1	R21	820	1 W	carbon composition	
1	R22	2 K	1 W	carbon composition	
1	R23	15 K	1/2 W	carbon composition	
1	R24	50	1 W	wire wound	
1	R25	3 K	1/2 W	carbon composition	
1	R26	50	1 W	wire wound	Daleohm
1	R27	3 K	1/2 W	carbon composition	
1	R28	15 K	1/2 W	carbon composition	
1	R29	820	1 W	carbon composition	
1	R30	2 K	1 W	carbon composition	

PARTS LIST FOR 2 KW [illegible] VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Resistance	Watts	Material	Manufacturer
1	R31	10 K	1/2 W	carbon composition	
1	R32	240	1 W	carbon composition	
1	R33	10 K	1/2 W	carbon composition	
1	R34	10 K	1/2 W	carbon composition	
1	R35	240	1 W	carbon composition	
1	R36	10 K	1/2 W	carbon composition	
1	R37	10 K	1/2 W	carbon composition	
1	R38	10	10 W	wire wound	Daleohm
1	R39	820	1 W	carbon composition	
1	R40	820	1 W	carbon composition	
1	R41	10	10 W	wire wound	Daleohm
2	R42	.1	25 W	wire wound	Daleohm (series)
1	R43	100	1 W	carbon composition	
1	R44	20	1 W	carbon composition	
1	R45	1 K	1/2 W	carbon composition	
1	R46	1 K	1/2 W	carbon composition	
1	R47	20	1 W	carbon composition	
1	R48	20	1 W	carbon composition	
1	R49	1 K	1/2 W	carbon composition	
1	R50	1 K	1/2 W	carbon composition	
1	R51	20	1 W	carbon composition	
1	R52	240	1 W	carbon composition	
1	R53	240	1 W	carbon composition	
1	R54	10 K	1/2 W	carbon composition	
1	R55	1 K	1 W	carbon composition	
1	R56	22 K	1 W	carbon composition	
1	R57	160 K	10 W	wire wound	Daleohm
1	R58	160 K	10 W	wire wound	Daleohm
1	R59	160 K	10 W	wire wound	Daleohm
4	R60	5	25 W	wire wound	Daleohm (series-parallel)

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Resistance	Watts	Material	Manufacturer
1	R61	4.3 K	1/2 W	carbon composition	
1	R62	10 K	2 W	carbon composition potentiometer	
1	R63	10 K	1/2 W	carbon composition	
1	R64	1.5 K	2 W	carbon composition potentiometer	
2	R65	5	25 W	wire wound	Daleohm (parallel)
2	R66	5	25 W	wire wound	Daleohm (parallel)
4	R67	5	25 W	wire wound	Daleohm (series parallel)
2	R68	.1	25 W	wire wound	Daleohm (series)
1	R69	2 K	1/2 W	carbon composition	
1	R70	10 K	1/2 W		
1	R71	1.3 K	1/2 W	carbon composition	
1	R72	5.1 K	1/2 W	carbon composition	
1	R73	33	2 W	carbon composition	
1	R74	100 K	2 W	carbon composition potentiometer	
1	R75	0.1 meg	2 W	carbon composition potentiometer	
1	R76	1 K	1/2 W	carbon composition	
1	R77	10 K	1/2 W	carbon composition	
1	R78	510	2 W	carbon composition	
1	R79	1.5 K	1/2 W	carbon composition	
1	R80	33	2 W	carbon composition	
2	R81	10	2 W	carbon composition	(parallel)
1	R82	1 K	1 W	carbon composition	
1	R83	3 K	1/2 W	carbon composition	
1	R84	10	2 W	carbon composition	
1	R85	10	2 W	carbon composition	
1	R86	51	2 W	carbon composition	
1	R87	51	2 W	carbon composition	

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Resistance	Watts	Material	Manufacturer
1	R88	470	1/2 W	carbon composition	
1	R89	270	1/2 W	carbon composition	
1	R90	47	1/2 W	carbon composition	
1	R91	25 K	2 W potentiometer	carbon composition	
1	R92	3.0 K	1/2 W	carbon composition	
1	R93	51 K	1/2 W	carbon composition	
1	R94	10 K	1/2 W	carbon composition	
1	R95	10 K	10 W	wire wound	Daleohn
1	R96	100 K	1/2 W	carbon composition	
2	R97	20	25 W	wire wound	Daleohm (series)
1	R98	1 K	1/2 W	carbon composition	
2	R99	10	1 W	carbon composition	(parallel)
1	R100	1 K	10 W	wire wound	Daleohm
1	R101	1 K	10 W	wire wound	Daleohm
1	R102	1 K	10 W	wire wound	Daleohm
1	R103	1 K	10 W	wire wound	Daleohm
1	R104	10	2 W	carbon composition	
1	R105	10	2 W	carbon composition	
1	R106	100	1/2 W	carbon composition	
1	R107	750	1 W	carbon composition	
1	R108	1 K	25 W	wire wound	Daleohm
1	R109	62 K	1/2 W	carbon composition	
1	R110	1 K	1/2 W	carbon composition	

PARTS LIST FOR 2 KW 2[illegible]00 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Diode No.	Volts	Amps		Manufacturer
1	CR1	478M	600 V	6 A		Westinghouse
1	CR2	478M	600 V	6 A		Westinghouse
1	CR3	478M	600 V	6 A		Westinghouse
1	CR4	478M	600 V	6 A		Westinghouse
1	CR5	FSPF80W	8000 V	1.6 A		Solitron
1	CR6	FSPF80W	8000 V	1.6 A		Solitron
1	CR7	FSPF80W	8000 V	1.6 A		Solitron
1	CR8	FSPF80W	8000 V	1.6 A		Solitron
1	CR9	1N645	225 V	400 MA		Clevite
1	CR10	1N645	225 V	400 MA		Clevite
1	CR11	1N645	225 V	400 MA		Clevite
1	CR12	4E20-28	20 V	Shockley		Clevite
1	CR13	1N3671A (404 S)	800 V	12 A		Westinghouse
1	CR14	1N1204A	400 V	12 A		Westinghouse
1	CR15	1N978B	51 V	400 MW	Zener	Motorola
1	CR16	1N645	225 V	400 MA		Clevite
1	CR17	1N747	3.6 V	400 MW	Zener	Motorola
1	CR18	1N645	225 V	400 MA		Clevite
1	CR19	1N1124A	200 V	3 A		I.R.
1	CR20	1N1124A	200 V	3 A		I.R.
1	CR21	1N1124A	200 V	3 A		I.R.
1	CR22	1N1124A	200 V	3 A		I.R.
1	CR23	1N645	225 V	400 MA		Clevite
1	CR24	1N645	225 V	400 MA		Clevite
	CR25	Not Used				
	CR26	Not Used				
1	CR27	1N645	225 V	400 MA		Clevite
1	CR28	1N645	225 V	400 MA		Clevite
1	CR29	1N645	225 V	400 MA		Clevite
1	CR30	1N645	225 V	400 MA		Clevite

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Diode No.	Volts	Amps	Manufacturer
1	CR31	1N645	225 V	400 MA	Clevite
1	CR32	1N645	225 V	400 MA	Clevite
1	CR33	1N645	225 V	400 MA	Clevite
1	CR34	4E20-28	20 V	Shockley	Clevite
1	CR35	1N645	225 V	400 MA	Clevite
1	CR36	1N645	225 V	400 MA	Clevite
1	CR37	1N645	225 V	400 MA	Clevite
1	CR38	1N645	225 V	400 MA	Clevite
1	CR39	1N645	225 V	400 MA	Clevite
1	CR40	1N645	225 V	400 MA	Clevite
1	CR41	1N645	225 V	400 MA	Clevite
1	CR42	1N645	225 V	400 MA	Clevite
1	CR43	4E20-28	20 V	Shockley	Clevite
1	CR44	1N645	225 V	400 MA	Clevite
1	CR45	1N645	225 V	400 MA	Clevite
1	CR46	1N645	225 V	400 MA	Clevite
1	CR47	1N645	225 V	400 MA	Clevite
1	CR48	1N645	225 V	400 MA	Clevite
1	CR49	1N645	225 V	400 MA	Clevite
1	CR50	1N645	225 V	400 MA	Clevite
1	CR51	1N645	225 V	400 MA	Clevite
1	CR52	1N645	225 V	400 MA	Clevite
1	CR53	1N645	225 V	400 MA	Clevite
1	CR54	1N645	225 V	400 MA	Clevite
1	CR55	1N645	225 V	400 MA	Clevite
1	CR56	1N3016	6.8 V	1 W Zener	Motorola
1	CR57	1N645	225 V	400 MA	Clevite
1	CR58	4E20-28	20 V	Shockley	Clevite
1	CR59	1N645	225 V	400 MA	Clevite
1	CR60	1N645	225 V	400 MA	Clevite

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Diode No.	Volts	Amps		Manufacturer
1	CR61	1N645	225 V	400 MA		Clevite
1	CR62	1N645	225 V	400 MA		Clevite
1	CR63	1N747	3.6 V	400 MW	Zener	Motorola
1	CR64	1N645	225 V	400 MA		Clevite
1	CR65	1N645	225 V	400 MA		Clevite
1	CR66	1N645	225 V	400 MA		Clevite
1	CR67	478 M	600 V	6 A		Westinghouse
1	CR68	478 M	600 V	6 A		Westinghouse
1	CR69	1N645	225 V	400 MA		Clevite
1	CR70	1N645	225 V	400 MA		Clevite
1	CR71	1N3993A	3.9 V	10 W	Zener	Motorola
1	CR72	1N645	225 V	400 MA		Clevite
1	CR73	1N649	600 V	400 MA		G.E.
1	CR74	1N1220	200 V	750 MA		Westinghouse
1	CR75	1N1220	200 V	750 MA		Westinghouse
1	CR76	1N1220	200 V	750 MA		Westinghouse
1	CR77	1N1220	200 V	750 MA		Westinghouse
1	CR78	1N2611	200 V	750 MA		Motorola
1	CR79	1N2611	200 V	750 MA		Motorola
1	CR80	1N2813B	15 V	50 W	Zener	Motorola
1	CR81	1N645	225 V	400 MA		Clevite
1	CR82	1N649	600 V	400 MA		G.E.
1	CR83	1N645	225 V	400 MA		Clevite
1	CR84	1N645	225 V	400 MA		Clevite
1	CR85	1N3993A	3.9 V	10 W		Motorola
1	CR86	1N645	225 V	400 MA		Clevite
1	CR87	1N1124A	200 V	3 A		I.R.
1	CR88	1N645	225 V	400 MA		Clevite
1	CR89	4E20-28	20 V	Shockley		Clevite
1	CR90	1N645	225 V	400 MA		Clevite

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Diode No.	Volts	Amps	Manufacturer
1	CR91	1N645	225 V	400 MA	Clevite
1	CR92	1N645	225 V	400 MA	Clevite
1	CR93	1N645	225 V	400 MA	Clevite
1	CR94	1N1218	100 V	1.6 A	Westinghouse
1	CR95	1N1218	100 V	1.6 A	Westinghouse
1	CR96	1N1443	1000 V	1.6 A	Westinghouse
1	CR97	1N1443	1000 V	1.6 A	Westinghouse
1	CR98	1N1443	1000 V	1.6 A	Westinghouse
1	CR99	1N1443	1000 V	1.6 A	Westinghouse
	CR100	Not Used			
	CR101	Not Used			
1	CR102	4E20-28	20 V	Shockley	Clevite

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Transistor	Volts	Amps		Manufacturer
1	Q1	242 ZP	700 V	10 A	GCS	Westinghouse
1	Q2	242 ZP	700 V	10 A	GCS	Westinghouse
1	Q3	242 ZP	700 V	10 A	GCS	Westinghouse
1	Q4	242 ZP	700 V	10 A	GCS	Westinghouse
1	Q5	2N2102	80 V	1 A		R.C.A.
1	Q6	2N1777	400 V	4.7 A	SCR	G.E.
1	Q7	2N2201	100 V_{CE}	500 MA		G.E.
1	Q8	2N1015F	300 V	7.5 A		Westinghouse
1	Q9	2N1016E	250 V	7.5 A		Westinghouse
1	Q10	2N1016E	250 V	7.5 A.		Westinghouse
1	Q11	2N2270	60 V	1 A		R.C.A.
1	Q12	2N2270	60 V	1 A		R.C.A.
1	Q13	2N2270	60 V	1 A		R.C.A.
1	R14	2N2270	60 V	1 A		R.C.A.
1	Q15	2N2270	60 V	1 A		R.C.A.
1	Q16	2N2270	60 V	1 A		R.C.A.
1	Q17	2N1016B	100 V	7.5 A		Westinghouse
1	Q18	2N1016B	100 V	7.5 A		Westinghouse
1	Q19	2N1016B	100 V	7.5 A		Westinghouse
	Q20	Not Used				
1	Q21	2N1016B	100 V	7.5 A		Westinghouse
	Q22	Not Used				
1	Q23	2N3054	60 V	4 A		R.C.A.
1	Q24	2N1039	60 V	.5 A		Texas Instruments
1	Q25	2N2102	80 V	1 A		R.C.A.
1	Q26	2N1814	300 V	30 A		Westinghouse
1	Q27	2N1771	60 V	4.7 A		Westinghouse
1	Q28	2N2102	80 V	1 A		R.C.A.
1	Q29	2N1814	300 V	30 A		Westinghouse
1	Q30	2N1015F	300 V	7.5 A		Westinghouse

PARTS LIST FOR 2 KW 2500 VOLT DC POWER SUPPLY
(Continued)

Quantity	Component Number	Transistor	Volts	Amps	Manufacturer
1	Q31	2N2102	80 V	1 A	R. C. A.
1	Q32	2N1016A	60 V	7.5 A	Westinghouse
1	Q33	2N491 B	Unijunction		G. E.
1	Q34	2N2102	80 V	1 A	R. C. A.
1	Q35	2N2102	80 V	1 A	R. C. A.
1	Q36	2N3054	60 V	4 A	R. C. A.
1	Q37	2N2324A	100 V	1 A	G. E.
1	Q38	2N3054	60 V	4 A	R. C. A.

Miscellaneous

1	SW1	SPST Switch
1	SW2	SPST Switch
1	M1	Voltmeter 0 - 150 V.D.C. Triplett 327T
1	M2	Voltmeter 0 - 3000 V.D.C. Triplett 327T
1		Input Receptacle Amphenol MS/AN 3102A-20-24P
2		Output Receptacle Amphenol UG-560/U #82-805
1	L1	Input Filter Inductor
2	L2 and L3	Inverter Line Inductors
1	L4	Output Filter Inductor
1	L5	Chopper Ringing Inductor
1	L6	Chopper Smoothing Inductor
1	L7	Saturating Core Inductor
2	L8 and L9	Inverter dI/dt Inductors
2	T1 and T2	Output Transformers
2	T3 and T4	Driver Transformers
1	T5	Pulse Transformers
1	T6	Oscillator Transformer
1	T7	Pulse Transformer
1	T8	Pulse Transformer
1	T9	Soft-On Transformer
1	T10	Impedance Matching Transformer