JET PROPULSION LABORATORY
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Interim Design and Development Report

POWER CONDITIONING AND CONTROL SYSTEM

Prepared for:

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1. INTRODUCTION

This interim technical report contains detailed information on a power conditioning and control system for a cesium bombardment ion engine, and a thorough technical discussion describing the development to date of specific circuitry for: (1) the power supplies that provide the arc, vaporizer, magnet, valve and reservoir heater, power; (2) the positive and negative high voltage supplies that furnish power to the ion beam and accelerator electrode; (3) the programmer that provides the sequencing required for operation of the engine; (4) the instrumentation and signal conditioning; and (5) the packaging techniques for laboratory test.

The technical objectives, as outlined in the Statement of Work, are as follows:

The contractor shall:

A. Provide for circuit design, component selection, component and circuitry technique evaluation, procurement of selected components, mechanical design and thermal analysis of a lightweight power conditioning and control system, which shall be capable of integration with a cesium bombardment ion engine developed under Lewis Research Center, Contract No. NAS3-5250, with Electro-Optical Systems, Inc. In the performance of this effort, the contractor shall accomplish the following:

1) Design a lightweight power conditioning and control system to operate a cesium bombardment thrustor utilizing a solar panel power supply or a simulated solar panel power supply as prime power source.

2) Perform thermal analysis and mechanical design associated with the packaging of the electronic system design, required by Paragraph A, so that it will operate at steady state rated level in a vacuum.

3) Perform part selection, utilizing flight type components wherever possible and evaluate circuitry techniques and components in the laboratory, as required.

These technical objectives provide for the design of a lightweight and reliable power conditioning and control system which will be operated in conjunction with a cesium bombardment engine in a lifetime test of at least 500 hours duration.

Additionally, this design effort is a part of a joing effort by NASA, JPL and the Air Force to establish the feasibility of a solar electric propulsion spacecraft. As a result of this relationship, the efforts on this program have, as an additional design constraint, the requirement to be compatible with the overall spacecraft design effort which is being carried on in parallel at EOS. The technical discussion which follows points out areas in which compromises have been required in order to best meet both the test constraints and the proper representation of a spacecraft electric propulsion system.

2. TECHNICAL DISCUSSION

2.1 Design Considerations

2.1.1 Engine Interface

The bombardment engine, developed under Lewis Research Center, Contract NAS3-5250, was designed to operate steady state at approximately two kilowatts of input power with a specific impulse of 6940 seconds. The nominal operating voltages, currents and input power for the engine functions were tabulated in the initial design and development report, dated 31 March 1965. Consideration was also given to an additional candidate power level of 750 watts. Present discussion, in relation to the cesium bombardment engine and the operating point for tests, indicate that a one kilowatt engine input power level with a specific impulse between 4000 and 5000 seconds would be the most desirable. Negotiations with the Lewis Research Center, in regard to the testing of the engine developed under Contract NAS3-5250, have resulted in a follow-on contract, No. NAS3-7112, which specifies an operating specific impulse of 5000 seconds at 1000 watts of engine input power. The engine input requirements associated with this operating point form the basis for the electrical requirements of the power conditioning system and are not incompatible with spacecraft design study assumptions.

2.1.2 Electrical Requirements

2.1.2.1 Engine

The engine parameters associated with a specified specific impulse of 5000 seconds, engine input power of 1000 watts, and an assumed mass efficiency of 91 percent are as follows:

FUNCTION	SYMBOL	ENGINE INPUT
Accelerating Potential	V+	2.0 KV D.C.
Accel. Electrode Pot.	V-	0.6 KV D.C.
V- Supply Current	I-	0.0048 Amps.
V+ Supply Current	I+	0.408 Amps.
Beam Current	I_b	0.403 Amps.
Arc Voltage	V_a	6.9 VDC
Arc Current	I_a	20 Amps.
Beam Power	P_b	0.806 KW
Drain Power	P_d	0.013 KW
Magnet Power	P_m	0.010 KW
Arc Power	P_a	0.138 KW
Vaporizer Power	P_v	0.010 KW
Neutralizer Power	P_n	0.021 KW
Total Power	P_t	0.998 KW
Specific Impulse	I_{sp}	4990 Seconds
Mass Efficiency	N_m	91%
Pwr. to Thrust Ratio	Kw/lb	142.2 watts/mlb
Thrust	F	6.8 mlbs.

2.1.2.2 Power Conditioning System

The power conditioning electrical requirements are a function of the engine input requirements and the engine operating characteristics. The parameters tabulated represent those required from the power conditioning system at the nominal operating point:

SUPPLY FUNCTION	VOLTS E_o	AMPS I_o	PERCENT REGULATION	PERCENT RIPPLE
+ High Voltage	2000	.408	None	± 5
- High Voltage	-600	.0048	None	± 5
Arc Supply	6.9	20	25A Current Limit	± 1 Volt
Engine Vaporizer	3.94	2.54	Beam I Reg.	A.C.
Engine Cathode	8.0	10	None	A.C.
Magnet Supply	5.0	2.0	I ± 1.0	± 1.0
Reservoir Heater	20	8	None	A.C.
Valve Heater	6	9	None	A.C.
Neut. Cathode	2.8	2.8	None	A.C.
Neut. Vaporizer	3.5	3.5	Beam E Reg.	A.C.
Neut. Bias	17	0.5	± 10	± 5.0

A more detailed discussion of each individual supply is included in later sections.

2.1.3 Prime Power Source

The prime power in this application is to be obtained from a solar panel array or an equivalent simulator. The output characteristics of a solar panel are such that to obtain maximum utilization of power output, the load must be properly matched to the solar panel. Otherwise, instabilities and loss of regulation can result. The requirement to match the load to the prime power source output characteristics places design constraints upon load. Consideration has been given this problem and the resulting power conditioning and control system design embodies a combination of input characteristics so that instabilities and loss of regulation will not occur, and inefficiencies will be minimized.

2.1.4 Mechanical Requirements

Included in the mechanical design effort is the packaging of the electronic system into a configuration so that the system will operate steady state in a vacuum environment. The configuration to be used for test purposes will provide a cooling which is closely analogous to space conditions. Primary cooling will be accomplished by radiation. A more detailed description of the packaging approaches is included in later sections.

2.1.5 Environmental Requirements

The environmental requirements for this power conditioning and control system have been selected so as to be consistent with typical launch and space conditions in the laboratory. Past experience with systems operating in a vacuum environment has established additional guidelines for environmental requirements. The following environmental parameters have been adopted as the environmental design requirements for this system:

Atmospheric Pressure - $< 10^{-4}$ Torr.

Temperature Range - 50°C to $+125^{\circ}$C.

Vibration -

Random Gaussian - 15 to 200 CPS .03g @ 2 CPS constant sweep for 5 minutes.

Sinusoidal -

5 - 20 CPS	¼" double amplitude
20 - 400 CPS	10 g
400 - 1500 CPS	15 g
1500 - 3000 CPS	15 g

along three mutually perpendicular axis constant sweep. From 5 to 3000 CPS in not less than 45 minutes.

Acceleration - 15 g static acceleration in both directions along each of two major thrust axis. Mandatory duration 10 minutes each direction.

Shock - 30 g ½ sine wave 6 millisecs. Three times in each of three mutually perpendicular axis.

The actual system to be built for testing will not be a flight system and, therefore, not packaged for flight vibration, acceleration and shock conditions. However, all components were selected with consideration given to all environmental parameters.

2.1.6 Parts - Selection and Evaluation

Components required for the system have been selected from the JPL Specification ZPP-2061-PPL-F, High Reliability and Preferred Parts lists. In instances of circuit design which required components not listed on the above references or for which alternates were not available or acceptable, it was necessary to select components from other sources. Such selections were made with conservative engineering judgement as to maximum reliability, safety factors, multiplicity of sources, and general flight quality and suitability of each component.

At such time as actual hardware is developed and fabricated, each component in this class will be reexamined for possible replacement by new additions to the reference documents, or will be evaluated and tested for continued applicability in the equipment.

2.2 Electrical Design

2.2.1 Design Philosophy

In the design of power conditioning and control systems for electric propulsion engines, there are a number of basic approaches which can be taken and any of these approaches can be implemented with a variety of active circuit elements. However, the choice of the basic approach and the active devices to be used is dependent upon efficiencies, power to weight ratios and reliabilities that can be obtained. These parameters are used as the basis for selection of components and circuits. The major design tasks associated with such power conditioning and control systems are: power conversion, regulation and control, and programming. The following paragraphs list the relative merits of several methods of accomplishing each of the indicated functions along with the method selected for this program.

2.2.1.1 Power Conversion

Three Phase Master Conversion.

The DC output from the power source is converted into three phase square wave power, which is applied to an AC bus. The separate engine function power supplies derive power from this AC bus. With a three phase system, transformers can be used with high mass utilization efficiency. Output control can be accomplished in the transformer primaries by use of either silicon-controlled rectifiers (SCR's) or saturable reactors. Rectification frequencies are high, allowing the use of smaller filtering components. However, the circuitry required to convert prime DC to regulated three phase AC becomes quite complex, and the number of components involved is appreciable, adding to system weight and compromising system reliability. For these reasons, this method is not considered usable for this application.

Single Phase Master Conversion. All of the power required by the system is converted from DC to AC by a single converter and distributed throughout the system in this AC form. In general, primary static conversion will result in high efficiencies, since more efficient use of transformers, transistors and other circuit components is realized when the design is optimized for conversion only and does not perform regulation and control. However, the system imposes restrictions for regulation and control of multiple output voltages which reduce the advantage in efficiency gained by optimizing conversion. In addition, a system designed to deliver a kilowatt of power would require switching transistors capable of 18 amperes collector current in the steady state condition. Allowing for a 50 percent safety factor, this would require at least 50 ampere transistor capability. The availability of such devices is extremely limited in the silicon types and the switching speed capability is also limited. For these reasons, this method was not considered usable for this application.

Local Conversion. The primary DC power is converted, as required, for operation of the specific engine functions at the point of usage. This technique allows the separation of regulation and control in addition to including them in the DC to AC conversion step. This design approach is extremely useful where the system requires separate regulation and control of multiple output voltages. However, since in the design considered here only a few of the engine functions require regulated outputs. Therefore, a system utilizing only local conversion was not considered an optimum choice.

Hybrid Conversion. A compromise between single phase master conversion and local conversion would, in this case, appear most desirable. Using this approach, the merits of both systems can be applied, where needed, to obtain the greatest efficiencies and reliabilities. This technique utilizes converters to furnish square wave AC power for several functions and also utilizes local conversion. This approach is to be used in this application.

Conversion Frequency. The switching frequency at which the converters operate is in part dependent upon the capability of switching devices for the "worst case" power demand conditions. Consideration of the switching characteristics of available devices, along with transformer losses and weight as a function of frequency, is required to select an optimum frequency of operation for converters. This subject is discussed in detail in later sections.

2.2.1.2 Regulation and Control

The methods of regulation and control employed can best be discussed at the point of usage. Since a hybrid system of conversion is being used in this application, several types of regulation and control are used. These methods were selected in each instance because design evaluation indicated they are the best for the function to be performed.

2.2.1.3 Programming

Electric propulsion engines require that application of power to the engine functions occur in proper sequence. Different flight operational requirements will determine the method used to obtain the programmed commands. A flight requiring a few cycles of long term ion engine operation will utilize signals from the vehicle master programmer/sequencer. The local programmer simulating the spacecraft sequencer has been included in this system for endurance testing in the laboratory. Table I summarizes the engine operating requirements for each phase sequence.

TABLE I

BOMBARDMENT ENGINE OPERATING REQUIREMENTS												
ENGINE FUNCTIONS	PHASE I			PHASE II			PHASE III			PHASE IV		
	Engine Heatup			Engine Heatup			Engine Cleanup			Engine On		
	30 Minutes			30 Minutes			5 Minutes			Operate		
	E Volts	I Amps	P Watts	E Volts	I Amps	P Watts	E Volts	I Amps	P Watts	E Volts	I Amps	P Watts
+ High Voltage	x	x	x	x	x	x	x	x	x	2000	.408	816
- High Voltage	x	x	x	x	x	x	x	x	x	600	.0048	3
Arc Power	10	x	x	7.5	18	135	7.5	18	135	6.9	20	138
Magnet Power	5.0	2.0	10.0	5.0	2.0	10.0	5.0	2.0	10.0	5.0	2.0	10.0
Engine Vaporizer	x	x	x	5	5	25	x	x	x	3.16	3.16	10.0
Engine Cathode	8	10	80	x	x	x	x	x	x	x	x	x
Neutralizer Vaporizer	x	x	x	x	x	x	3.16	3.16	10.0	3.16	3.16	10.0
Neutralizer Cathode	2.24	2.24	5.0	2.24	2.24	5.0	2.24	2.24	5.0	2.24	2.24	5.0
Neutralizer Bias	x	x	x	x	x	x	12	0.5	6.0	8	0.1	.8
Reservoir	20	8	160	20	8	160	x	x	x	x	x	x
Valve	6	9	54	x	x	x	x	x	x	x	x	x
Drain Power	x	x	x	x	x	x	x	x	x	x	x	10.0
Total Power	x	x	309	x	x	335	x	x	166	x	x	993

2.2.2 Low Voltage Inverter System

2.2.2.1 General Discussion

The primary design objectives of the low voltage inverter system are minimum weight, maximum efficiency and high reliability. A main parameter affecting both weight and efficiency is operating frequency. A survey of available transistors was conducted using an IBM 1620 digital computer to determine the transistor losses as a function of frequency (see Fig. 1). In addition, transformer performance characteristics have been evaluated over the frequency range of from 1 KC to 100 KC.

An optimum frequency of operation can be determined from an analysis of overall converter losses with consideration of component weight and size. As the operating frequency is increased, size and weight of the power transformer will decrease, but transistor and X-F losses will increase, resulting in a trade of size and weight for efficiency.

In the transformer design, a choice can be made in selection between nickel-iron grain oriented alloys and ferrite materials. The nickel-iron-alloys, such as Deltamax and Orthonol, offer an advantage in weight since they can be operated at higher flux densities than ferrites. In larger sizes of tape wound cores, the thinner gauges of tape are not available for operation at higher frequencies. This precludes the use of the tape wound cores at frequencies about 10 KC. A preliminary investigation was made on transformer core materials for high

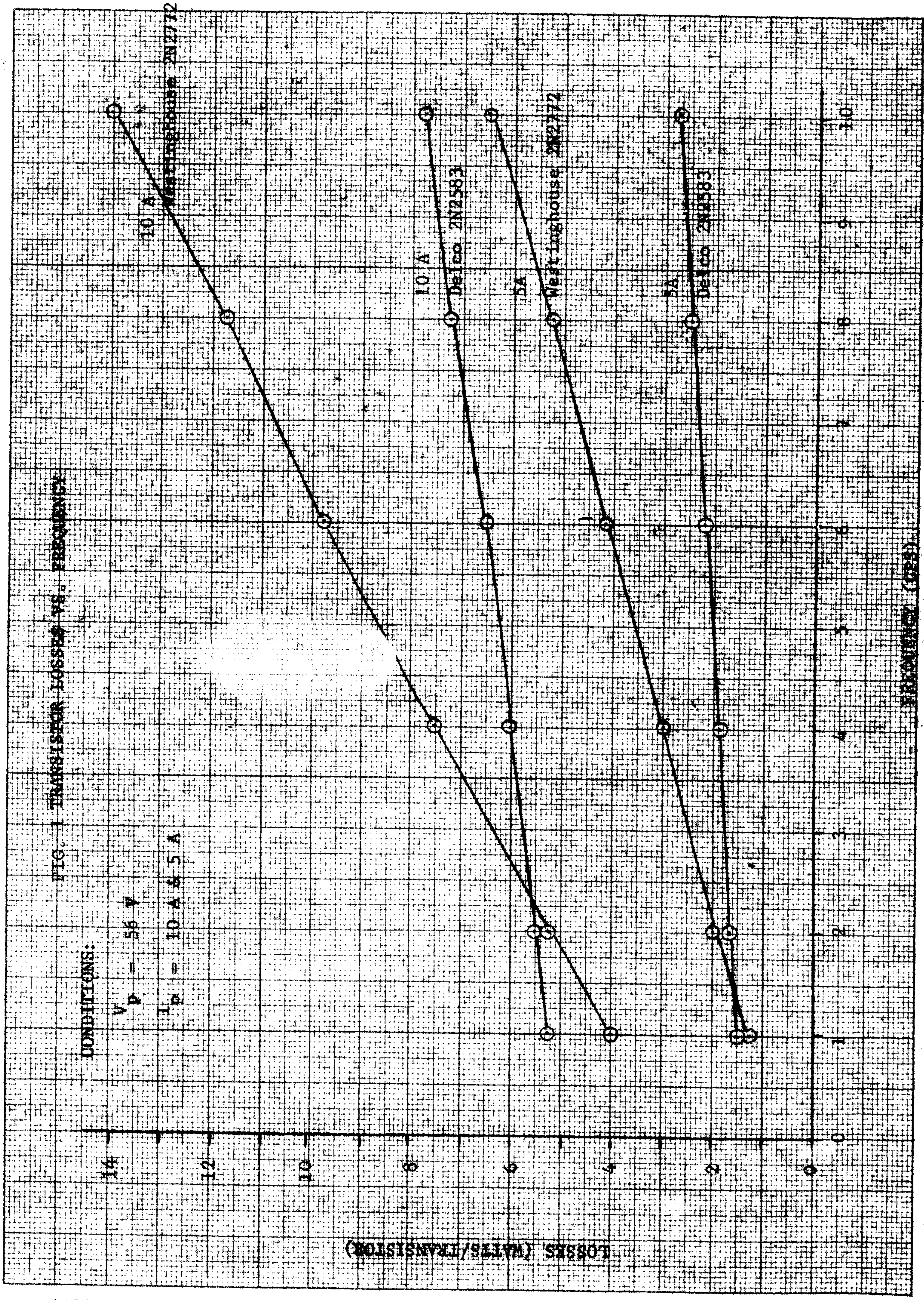

FIG. 1 TRANSISTOR LOSSES VS. FREQUENCY
CONDITIONS:
V_p = 56 V
I_p = 10 A & 5 A
10 A Westinghouse 2N2772
10 A Delco 2N2583
5A Westinghouse 2N2772
5A Delco 2N2583
LOSSES (WATTS/TRANSISTOR)
FREQUENCY
0
2
4
6
8
10
12
14
0
1
2
3
4
5
6
7
8
9
10

frequency operation. Plots of core weight vs. frequency and core loss vs. frequency were made. These are shown in Figs. 2 and 3 respectively. The core materials considered are:

1. 50 percent nickel alloy
2. 80 percent nickel alloy
3. Square loop ferrite
4. High flux density ferrite.

Examination of Fig. 2 shows that the 50 percent nickel alloy material has the lowest core weight for a given frequency. Core weights for the other three materials are very much the same at any given frequency.

Ferrites have the definite advantage of low core losses when operated at high frequency. Fig. 3 shows the core losses of various core materials as a function of frequency. The ferrites show core losses that run 20 percent of the best available nickel-iron alloy. This low core loss for ferrite material would indicate that ferrites would be the core material of choice in a high frequency DC-DC converter.

Based on the data presented in Figs. 2 and 3, a selection of the operating frequency of the converter should be made as high as possible consistent with the minimum allowable losses. Besides the transformer core and copper losses, there are the switching and saturable losses of the power transistors.

A selection of other frequencies can also be made at the expense of efficiency or weight. If the frequency is increased, the transformer losses will increase as shown in Fig. 3. If the frequency is lowered, the weight of the transformer will increase as shown in Fig. 2 .

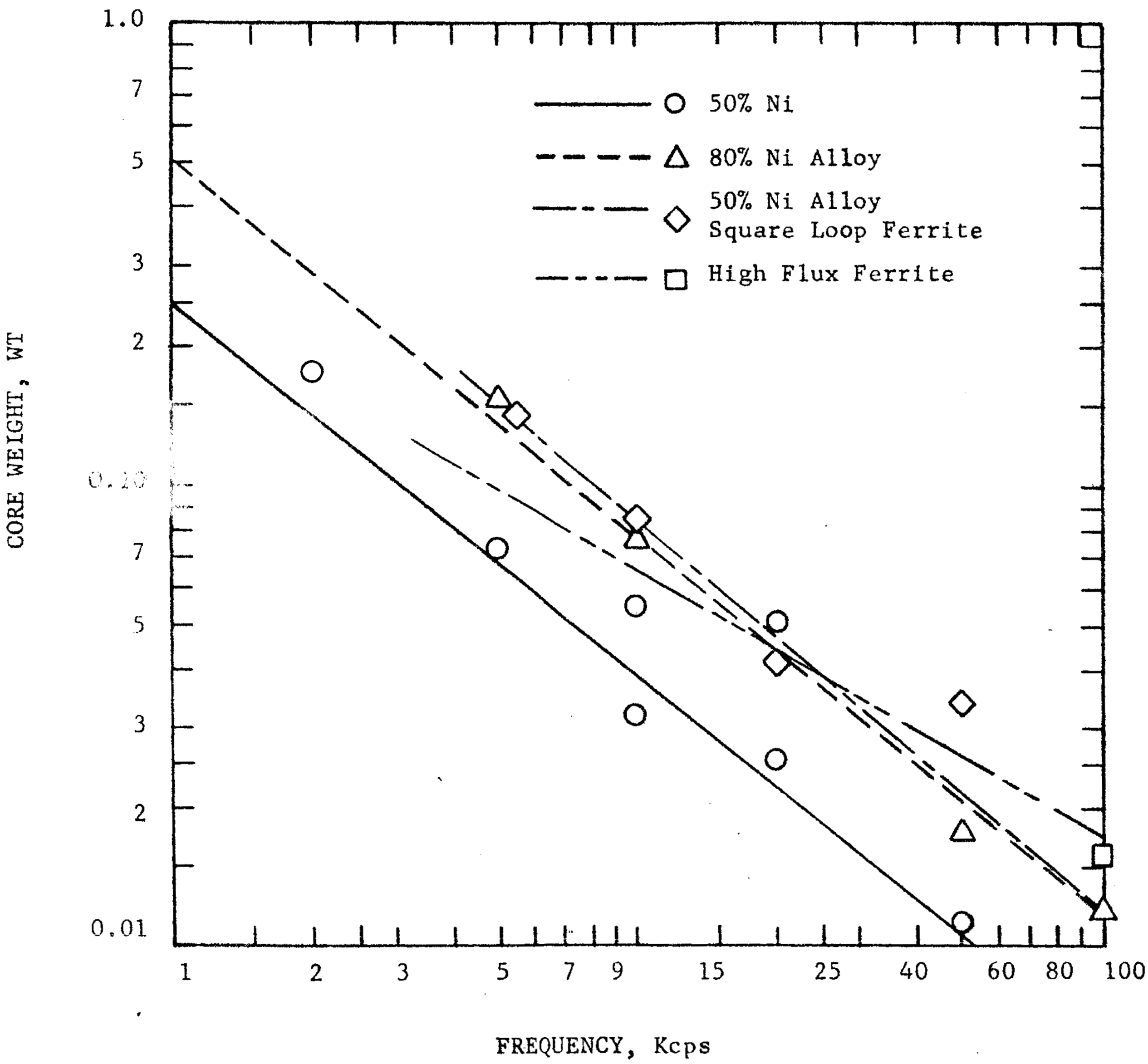

FIG. 2 CORE WEIGHTS VS FREQUENCY FOR GENERAL CORE MATERIALS

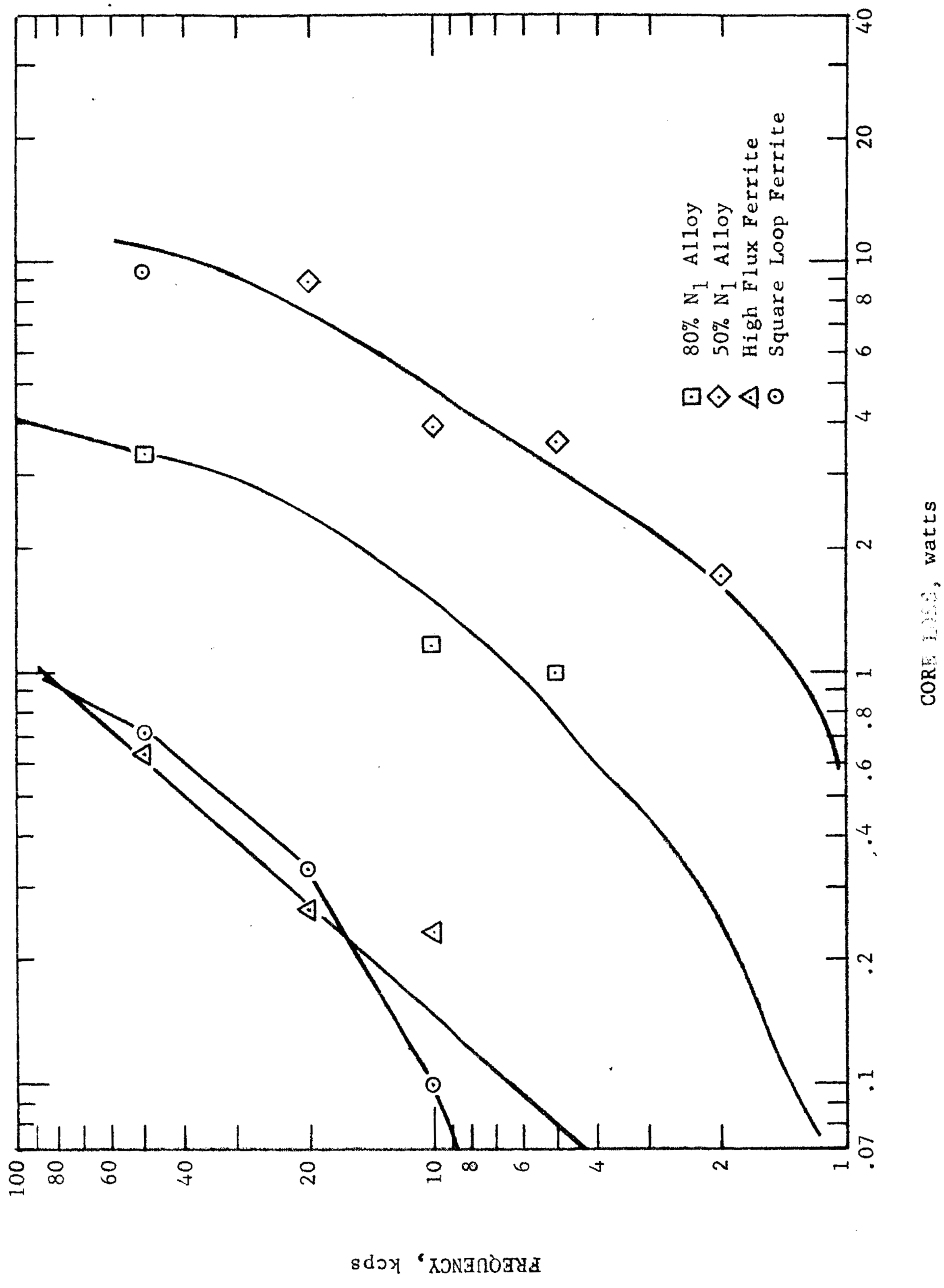

FIG. 3 CORE LOSS VS FREQUENCY FOR VARIOUS CORE MATERIALS

Since the maximum recommended operating frequency for tape wound cores is 10 KC, this frequency was selected as the operating frequency. Use of a ferrite core, for the same core weight, would mean a frequency of operation of 25 KC to 40 KC, Fig. 2. At present, there is no transistors available which could be used efficiently in the L.V. inverter system. The transistors which are available are either too slow with the required BV_{ceo}, or if the switching speed is adequate, BV_{ceo} is low.

2.2.2.2 Circuit Evaluation

The low voltage inverter system is divided into three sections. One section is used to power the reservoir and valve heaters, one to supply regulated direct current for the regulation and control circuits, and the last section supplies regulated AC power for the remaining low voltage engine functions.

Dividing the L.V. inverter system into three sections permits higher overall efficiency since each section of the L.V. inverter system can be optimized to its load. The division of output power in this manner requires each switching amplifier to switch lower collector currents. This allows the use of faster transistor switches, thus lowering the switching losses.

The use of silicon controlled rectifiers was considered in the early stages, but were ruled out because of their excessive switching and saturation losses at frequencies above 5 KC. SCR's were also considered unreliable because of the difficulty in commutating an SCR inverter in the presence of inductive loads.

One method of obtaining high efficiency power regulation from a DC source is the utilization of pulse width modulation. The mechanism selected for pulse width modulation is the use of a quasi-square wave. This quasi-square wave can be generated in two ways. The first way is to drive a magnetic amplifier with a square wave and by varying the current in the control winding, vary the time that the mag-amp is on. This generates the typical quasi-square wave. The primary advantage in this scheme is basically its simplicity. No amplifiers, F/F's, delay multi's, nand gates, etc., are required. The regulation loop is simplified and reliability of the circuit can be high due to its simplicity.

There are, however, some disadvantages to the use of mag-amps in this application. If, for example, the cores of the mag-amp are not matched and/or do not track each other with changes in temperature, the quasi-square wave output will not be symetrical about the zero axis. The expected mismatch of the cores in the mag-amps can be on the order of two to five percent nominal. Examination of the quasi-square wave shows that the dissymmetry in the quasi-square wave is directly proportional to the mismatch of the mag-amp cores. This causes a DC bias in the transformer driving waveform. This DC bias is equal approximately to the degree of mismatch in the cores. The DC power represented by the DC bias cannot be transformed to perform useful work and must, therefore, be dissipated as heat. This power loss will subtract from efficiency of the L.V. inverter system lowering efficiency about two to five percent. Also, the weight of the L.V. inverter system will increase because the heat generated will have to be dissipated in a heat sink. This additional heat load will increase the weight of the heat sinks required by the L.V. inverter system. There would also be a 10 percent increase in transformer weight.

Further investigation into the dynamic switching characteristics of the mag-amp would be necessary to determine whether the switching rise and fall time would be sufficiently fast enough to permit efficient switching. Also, the problem of ringing in the mag-amp and its associated circuits would have to be investigated.

The second method for generating a quasi-square wave is the addition of two square waves, one of which is delayed in time reference to the other. This scheme has a number of advantages. Its primary advantage is excellent output waveform symmetry. This occurs because the square waves are derived from F/F's. The two square waves are first generated in pulse form at twice the desired operating frequency. The series of pulses, which are to become the delayed square wave, are delayed at this time. Then, the two series of pulses, one direct and one delayed, each drive a F/F. The use of a F/F insures that the square waves will actually have an exact 50 percent duty cycle and, therefore, when they are added together, a symmetrical quasi-square wave is generated.

The disadvantage of this scheme is its high parts count. This would tend to indicate that the reliability of this circuit would be lower. The use of high reliability parts and "worst case" circuit design will insure a reliability adequate for the planned mission.

Also, substantial size reduction of the unit could be made using microcircuits. At present, the use of microcircuits is not anticipated. There is not enough data available on their performance in a high noise environment and further investigation would be required before consideration of their use in the system.

A unique feature of the low voltage inverter is that the power output transformer has multiple secondaries; this eliminates the requirement for separate transformers, reducing weight and increasing overall efficiency.

2.2.2.3 Circuit Description

The low voltage inverter system is graphically illustrated in the functional block diagram, Fig. 4, and electrically outlined by the schematic, Fig. 5.

The low voltage inverter is a high power, high efficiency device whose output circuit consists of two switching power amplifiers operated in parallel. Each power amplifier is driven by a square wave, one of which is delayed in time with respect to the other. The 20 KC clock generates the drive for the inverter. The drive is divided by two and drives one of the two power amplifiers. The delayed drive is also driven by the 20 KC clock. The delayed drive receives an error signal from the voltage reference and control amplifier. The error voltage changes the time delay between the two square waves such that the RMS output voltage of the inverter remains constant under conditions of varying input voltages. The output voltage of the inverter is sampled at the output transformer and drives the voltage reference and control amplifier.

The actual generation of the quasi-square wave output can be accomplished in two ways: (1) Add the two square waves in the output of the switching amplifiers, and (2) Add them together in a driver stage and use the resulting quasi-square wave to drive the switching amplifier.

Driving the switching amplifier with a quasi-square wave is difficult. The problem is that when both switching amplifiers are turned off during the dwell time of the quasi-square wave, the reactive energy stored in the reactive load will tend to flow back into the switching amplifier and thus present high voltages across the now turned off switches. Insertion of a reversed or free-wheeling diode between the collectors of the switching amplifier transistors and their supply voltage will prevent the generation of these high voltages. This solution

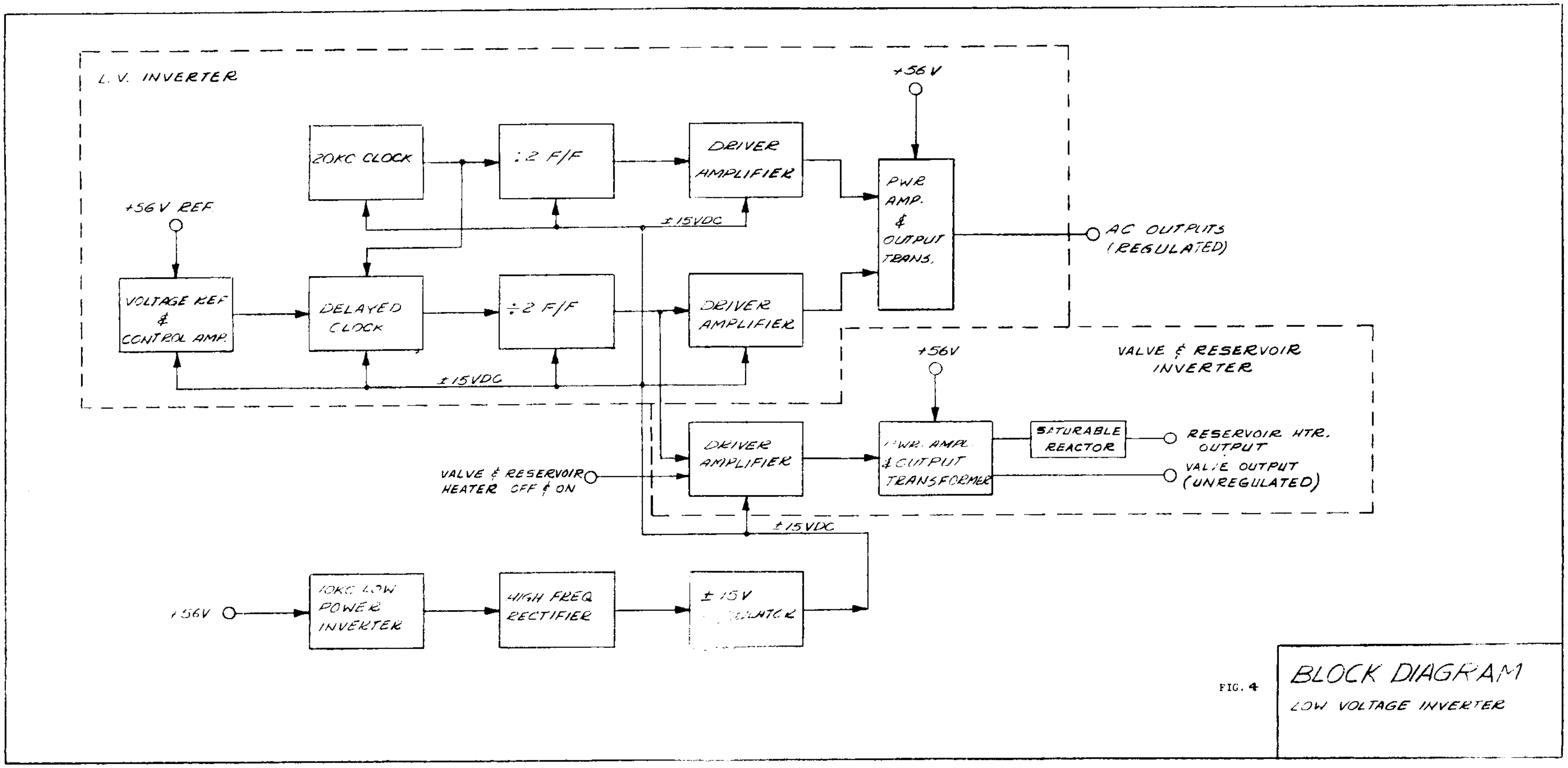

FIG. 4 BLOCK DIAGRAM OF LOW VOLTAGE INVERTER SYSTEM

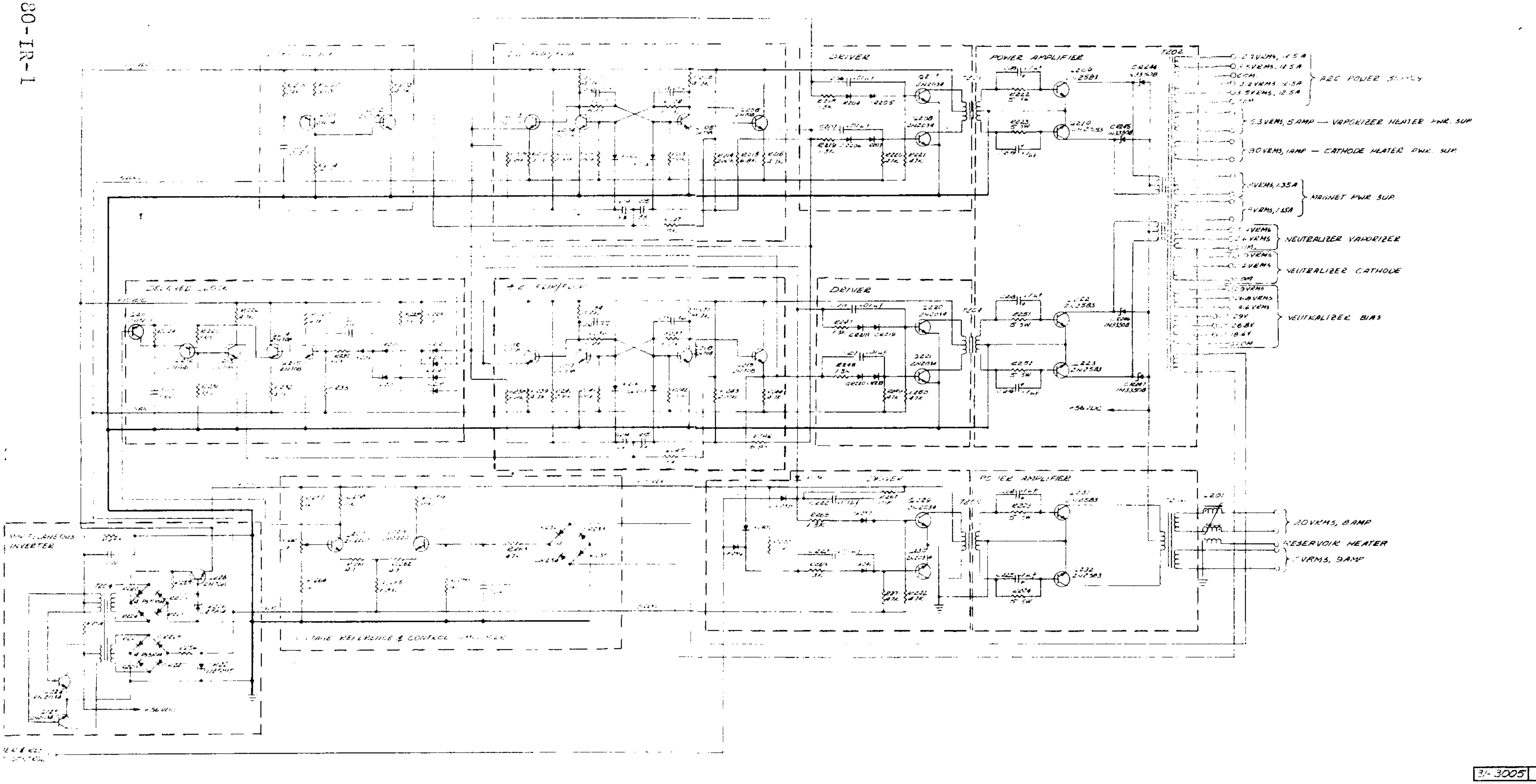

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FIG.   SCHEMATIC LOW VOLTAGE INVERTER SYSTEM

9 LAST TRANSISTOR NO. Q232
8 LAST DIODE NO. CR247
7 LAST CAPACITOR NO. C225
6 LAST RESISTOR NO. R274
5 ALL SEMICONDUCTOR DIODES MC 1231
4 ALL CAPACITOR VALUES ARE IN PICOFARADS
3 ALL RESISTOR VALUES ARE IN OHM ± 1/2 WATT
2 SYSTEM SCHEM DWG 31-3010
1 REF PARTS LIST 31-3005
NOTE: UNLESS OTHERWISE SPECIFIED

ELECTRO-OPTICAL SYSTEMS, INC.
SCHEMATIC DIAGRAM
LOW VOLTAGE INVERTER
UNCL
E 31-3005
12705

lowers the efficiency of the output stage because the free-wheeling diodes, during their conduction period, will dissipate appreciable power.

The power dissipated in the free-wheeling diode is equal to:

$$P_{rev} = I_{rev} E_{diode}$$

P_{rev} = power dissipated in the free-wheeling diode.

I_{rev} = reverse current generated by the inductive load.

E_{diode} = Forward voltage drop of a silicon diode $\approx$ 1 volt.

Assuming the reverse current at "worst case" to be equal to the load current which is 5 amps:

$$P_{rev} = (5A)(1V) = 5 \text{ watts.}$$

In the dual switching amplifier, the individual collector currents are reduced by one-half the power dissipated in the free-wheeling diodes during dwell time which is:

$$P_{rev} = (2.5A)(1V) = 2.5 \text{ watts each amplifier}$$

or 5 watts total.

If the free-wheeling diodes do not conduct, and the inductive reverse current is actually conducted by the saturated transistors, the power dissipated in the switching amplifier is then:

$$P_{amp} = R_{sat} I_{rev}^2$$

$$P_{amp} = (50M\Omega)(2.5A)^2 = .313 \text{ watts}$$

P_{amp} = power dissipated in the switching amplifier during dwell time.

R_{sat} = saturation resistance of transistor.

Since there is more power dissipated in the free-wheeling diode than in the transistors, the overall efficiency can be increased by using the transistors of the two switching amplifiers to carry the reverse current. Let the transformer add the square waves from the switching amplifiers in its secondary, as shown in Fig. 6. In. Fig. 6, time 1 Q_1 and Q_3 are on and power is delivered to the reactive load L_1 and R_1. Collector and load currents flow as shown. During dwell time, Fig. 6, time 2 Q_1 and Q_4 are on and no power is delivered to the load. If the load was resistive, no current would flow in the transistor collectors. In this case, the load is inductive and a reactive current I_r will flow in the opposite direction to I_1 at time 1. This current flow, when reflected to the primaries, tends to make current flow in the collector circuits of the transistors as shown by I'_{r1} and I'_{r2}. In the case of Q_1, the current flow I'_{r1} is opposite to the normal current flow of the transistor. If Q_1 has a reverse β of 0 (worst case), Q_1 cannot conduct any current. In the case of Q_4, the reverse current I_{r2} is in the direction of normal current flow and Q_4 will conduct. In fact, Q_4 can and does conduct all of the current I_r returned by the inductive load. In time 3, Fig. 6, Q_2 and Q_4 are on and power is delivered to the load, and in time 4, Fig. 6, Q_3 is the transistor which conducts all of the current returned by the load.

The design of the switching amplifiers accommodates this extra collector current by increasing the drive required by the amplifier transistors. The extra drive required, about one watt, is much less than the power dissipated by the free-wheeling diodes, which is about five watts.

Addition of the two square waves in the switching amplifier outputs will have a higher overall efficiency. The output switching amplifiers will always have one transistor turned on. The reverse current, which will flow during dwell time, will therefore be forced to flow in one of the two turned on output transistors. Thus, this inductive current can perform useful work without dissipating power in a free-wheeling diode.

20 KC Clock. The 20 KC clock consists of a UJT, 2N491, relaxation oscillator and a pulse amplifier. The period of the relaxation oscillator is determined by the time constant of R201 and C201. The time constant is given by the equation:

$$R \approx RC \ln \left[\frac{1}{1-\eta}\right]$$

Where R and C are the UJT emitter time constant, and η is the intrinsic standoff ratio, R202 is selected to provide first order temperature compensation and is determined from the equation:

$$R = \frac{0.7R_{bb}}{\eta V}$$

Where R_{bb} is the interbase resistance, and V is the interbase voltage, Q202 is a pulse amplifier whose output is a 15V pulse and is used as a trigger for the $\div 2$ flip-flops.

The $\div 2$ flip-flop is a standard RST flip-flop whose outputs are buffered by emitter followers. The output of the flip-flop is not adequate to drive the driver, and the emitter follower provides the necessary power gain. The set and reset inputs of both flip-flops are connected together and are driven by Q and $\overline{Q}$ outputs of the undelayed flip-flop.

Driver. The driver block is driven by the buffered Q and $\overline{Q}$ outputs of the $\div 2$ flip-flop. The use of the Q and $\overline{Q}$ outputs provides a push-pull signal. The outputs of the amplifier transistors, Q207, Q208, Q220 and Q221, are matched to the low impedance input of the power amplifier by T200 and T201. The base resistors limit the base drive to the driver transistors and the base capacitors are used to neutralize the store base charge of the driver transistors. The base diodes perform three functions: (1) They offset Veb of the driving emitter follower, (2) Provide some noise immunity, and (3) act as a disconnect diode when the input signal goes to ground allowing the emitter base junction to be reverse biased from a negative voltage source.

Power Amplifier. The power amplifier is a push-pull switching amplifier. There are two identical power amplifiers in the low voltage inverter. The addition of their outputs to form the quasi-square wave output occurs in the output transformer secondary. The base resistors are used to control the base drive and the base capacitors are used to improve the fall time of the transistors. The output power amplifier obtains its collector supply voltage directly from the primary power source. The transistors selected were 2N2583. These transistors have one of the highest BV_{ceo}'s of any transistors available.

The output transistors may see a maximum voltage of 2 x 84 or 168 volts, the 84 volts being the maximum solar panel output voltage. Allowing a voltage derating of 50 percent, the minimum BV_{ceo} of the transistor is 336 volts. The 2N2583 is rated at a BV_{ceo} of 500 volts and a $VCE_{(sus)}$ of 325 V minimum. This insures that the transistors will not experience voltage breakdown in the inverter, even when the inverter is operating into highly reactive loads.

Voltage Reference and Control Amplifier. The voltage reference and control amplifier is used to generate an error voltage when the RMS output voltage changes from nominal. The output voltage is sampled by a separate winding on the output transformer. This AC voltage is full wave rectified and the resultant signal integrated. This voltage is compared to a zener reference CR 231 by a differential amplifier Q227 and Q228. The output voltage of the differential amplifier drives the delayed clock.

Stability of the regulation feedback loop is accomplished by using the RC integrater as the dominate time constant of the feedback loop. The RC time constant is selected such that the gain of the loop is below unity before the phase shift reaches 180°. No calculations are presented here since it is intended that Nyquist diagrams or Bode plots will be prepared from data taken from an operating breadboard. Since the loop contains digital elements, some of which have highly non-linear transfer characteristics, it is felt that any mathematical discussion at this time would have to be confirmed empirically.

<u>Delayed Clock</u>. The delayed clock consists of a synchronizing gate, a UTJ delay, current control and a pulse amplifier. The synchronizing gate is two AND gates and an OR gate conntected to an inverter which is buffered by an emitter follower. The emitter follower, Q214, is required to supply the necessary charge current to C210 for minimum time delay, when Q211 is saturated. The inputs of the synchronous gates are connected to the Q and $\overline{Q}$ outputs of ÷/2/flip-flop No's. 1 and 2. Such that:

$$\overline{C} = Q_1 Q_2 + \overline{Q} \, \overline{Q}_2$$

When C goes true, the time delay starts. The output of the synchronizing gate then supplies +15 V to the UJT delay circuit. At the end of the timing cycle, the UJT fires and C then goes false. The maximum delay is set by R224 and the minimum delay is set by R224/R_{sat} Q211 + Z_o Q214. The delay is variable between these two limits. If at the initial turn-on of the low voltage inverter the two ÷ flip-flops are not in synchronization, the set and reset inputs of ÷/2 flip-flop No. 2 will not allow the ÷ flip flop No. 2 to change state when the UJT delay circuit fires thus forcing the two flip-flops into synchronization.

<u>Miscellaneous Converter</u>. The low power inverter is of the self-saturating type. The inverter derives

its power from the primary power source. Its output is rectified and supplies, after regulation, the ±15 volts required by the low voltage inverter. A simple emitter follower regulator referenced from a Zener diode is used for the +15 volts. The -15 volt output is simply Zener regulated. This ±15 volts is made available to the other power conditioning system.

2.2.2.4 Circuit Parameters

Input Voltage:

Nominal	56 Volts
Minimum	50 Volts
Max. Operational	70 Volts
Absolute Maximum	84 Volts

Output Voltages:

All outputs regulated ±2%.

1. 9.0 V RMS tapped at 5.1 V RMS at 12.5 Amps., 2 outputs each.
2. 5.3 V RMS at 5 Amps.
3. 8.0 V RMS at 10 Amps.
4. 2 each - 8 V RMS at 1.35 Amps.
5. 2 each - 3.9 V RMS tapped at 2.6 V RMS at 3.5 Amps.
6. 29 V RMS tapped at 26.8 V RMS and 18.6 V RMS at 0.5 Amps. with center taps for all individual outputs.

Losses:

All losses based on input power of 220 watts maximum.

Transformer (Output)	2.85%
Output Transistors	1.82%
Output Transistor Base Loss	.22%
±15 V Supply Including Inverter Loss.	4.33%
TOTAL	8.82%

Efficiency:

$100 - 8.82 = 91.18\%$

Weight:

Output Transformer	16.0 oz.
Power Transistors (4)	4.0 oz.
19 each TO-18 Transistors	0.4 oz.
Driver Transformer	6.0 oz.
Printed Circuit (est.)	4.0 oz.
Low Power Inverter	8.6 oz.
	39.0 oz.

2.2.3 Reservoir and Valve Heater Power Supply

2.2.3.1 General

The reservoir and valve heater power supply is a low output voltage medium power driven DC-AC inverter. This inverter is used to supply unregulated AC to the cesium reservoir and valve heater during engine heatup.

2.2.3.2. Circuit Description

Reference Fig. 4 block diagram. This inverter is basically a part of the low voltage inverter. It is considered separately since it functions only during engine heatup. The inverter consists of a switching power amplifier and driver for the power amplifier. The input square wave is taken from the ÷/2 flip-flop output and amplified by the driver. The driver is a push-pull amplifier transformer coupled to the power amplifier. The output transformer has two secondaries which supply the heater power to the cesium reservoir and valve. The output of the inverter which powers the cesium reservoir can be turned off and on by a pair of AND gates whose inputs are the drive signal and the programmer command. The ±15 volts required by the driver are supplied by the low power inverter. The power amplifier runs off the primary power source.

2.2.3.3 Circuit Parameters

Input Voltage:

Nominal	56 Volts
Minimum	50 Volts
Maximum Operating	70 Volts
Absolute Maximum	84 Volts

Output Voltages:

Reservoir	20 V RMS at 8 Amps controlled by a saturable reactor.
Valve	6 V RMS at 9 Amps.

Losses:

All losses based on input power of 220 watts.

Output Transformer		2.85%
Output Transistors		.71%
Output Transistor Base		.11%
Driver Stage		1.25%
	TOTAL	4.92%

Efficiency:

95.08%

Weight:

Output Transformer	16.0 oz.		
Power Transistors	2.0 oz.		
Driver Transformer	3.0 oz.		
TOTAL WEIGHT	21.0 oz.	=	1.31 lbs.

The total weight does not include structure or heat sinks.

2.2.4 Magnet Supply

2.2.4.1 General Discussion

The magnet supply is a direct current supply with an output capability of 5.5 VDC and 2.7 amperes maximum for a total DC power output of approximately 15 watts. ±1% current regulation is required. Four approaches for obtaining the desired output have been considered. They are:

A. Series transistor regulated supply.

B. Self-saturating mag-amp regulated supply.

C. Silicon controlled rectifier regulated supply.

D. Saturable reactor regulated supply.

2.2.4.2 Circuit Selection and Evaluation

A - Consider the schematic of a transistor regulator, Fig. 7. Unregulated DC is applied to the regulator from a transformer rectifier. The DC is applied to a transistor series regulator Q_1. Output current is sampled by the current transformer CT1. The output of the CT1 is rectified and a voltage is obtained which is directly proportional to the output current. This voltage is compared to a stable reference CR1 by means of a differential amplifier. The resulting difference voltage is an error signal which will vary directly with the output current. Use of this error signal to control the base drive of the series regulating transistor Q1 provide constant current regulation.

B - Fig. 8 is a schematic diagram of a self-saturating mag-amp regulator in a full wave center tap configuration. AC from the low voltage inverter output transformer secondary is applied to the regulator and the input to the rectifiers CR1 and CR2 is controlled by the mag-amp M1. Output current is sensed by the current transformer CT1. The secondary output of CT1 is rectified referenced to a Zener diode

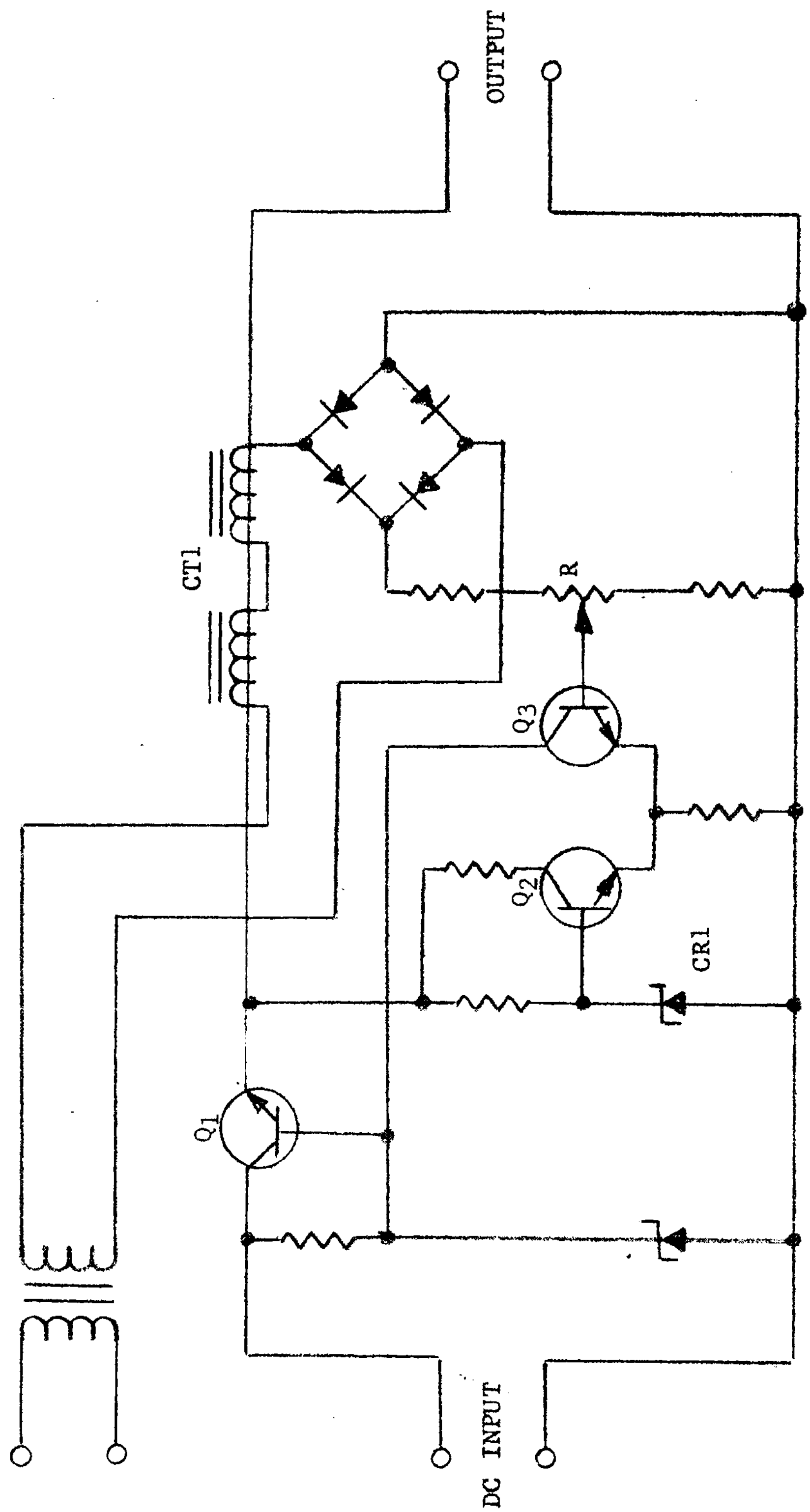

FIG. 7 SCHEMATIC TRANSISTOR REGULATOR

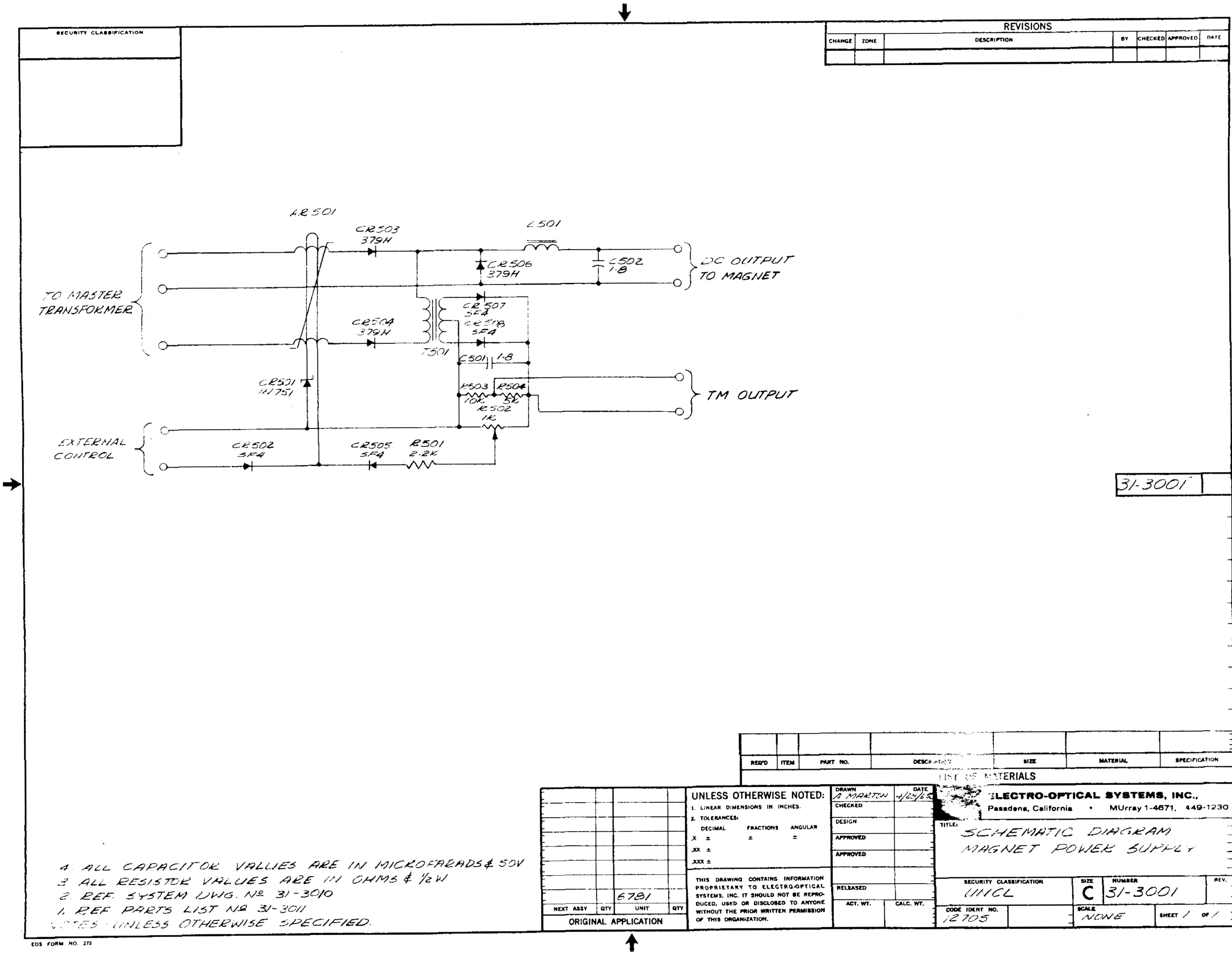

FIG. 8 SCHEMATIC MAGNET SUPPLY

CR8 and applied to the control winding of the mag-amp. Since the voltage produced by CT1 is directly proportional to the load current, its application to the control winding provides current regulated output. External control of the supply can be accomplished through a diode OR gate to the control winding of the mag-amp.

C. - The silicon controlled rectifier regulator is illustrated by the schematic Fig. 9. This design approach is quite similar to the self-saturating mag-amp. The silicon controlled rectifiers replace the mag-amps and a small mag-amp is used to develop synchronized square wave drive signals for the SCR gates. A current transformer is used to sense the output load current and develop an AC voltage proportional to the load current. This signal, when applied to the gate drive mag-amp control winding,controls the firing angle of the SCR's and thus controls the output current providing current regulation.

D - Fig. 10 is a schematic representation of the saturable reactor design approach. The saturable reactor is located in the output transformer primary. A current transformer senses the output current and provides a DC output voltage which is inversely proportional to the output current. This voltage, when applied to the control winding of the saturable reactor, will control the output current providing the required current regulation.

Each of the preceding design approaches have advantages and disadvantages. The relative merits of these approaches can best be seen by direct comparison; therefore, Table II has been prepared which compares these design approaches. Refering to the table, it can be seen that the transistor regulator approach, while simpler, has more losses; the saturable reactor regulator approach is heavier and would require an extra transformer; the SCR design approach is the

	#1 TRANSISTOR REGULATOR	#2 SELF SATURATING MAG. AMP.	#3 SCR REGULATED	#4 SATURABLE REACTOR
RELIABILITY	This type of regulation is less reliable than the other designs because of transient susceptibility and all components are at the +H.V. potential.	Good.	Fair.	Good.
WEIGHT	Differential in weight between these designs vary slightly at this power level - 15 to 25 watts.			Heaviest design, although only slightly.
EFFICIENCY	Transistors have excessive dissipation. Efficiency is quite low because of the high transistor losses.	At a power level of 15 to 25 watts, the differential in efficiency between these designs is negligible.		
ADVANTAGES	Simple, lightweight, fast response.	High efficiency, isolation from +H.V. easy to obtain, not susceptible to transient damage, high reliability.	High efficiency, isolation from +H.V. easy to obtain, lightweight.	High reliability, less susceptible to transient damage, control circuits not at high voltage.
DISADVANTAGES	Difficult to isolate control circuitry from +H.V., high power losses, highly susceptible to transients.	Slightly heavier than transistor or SCR designs. More filtering required for these designs.	Susceptible to transients.	Highest weight, lower efficiency than SCR or SSMA designs, higher control required, slowest response.

TABLE II - COMPARISON OF DESIGN APPROACHES FOR MAGNET SUPPLY.

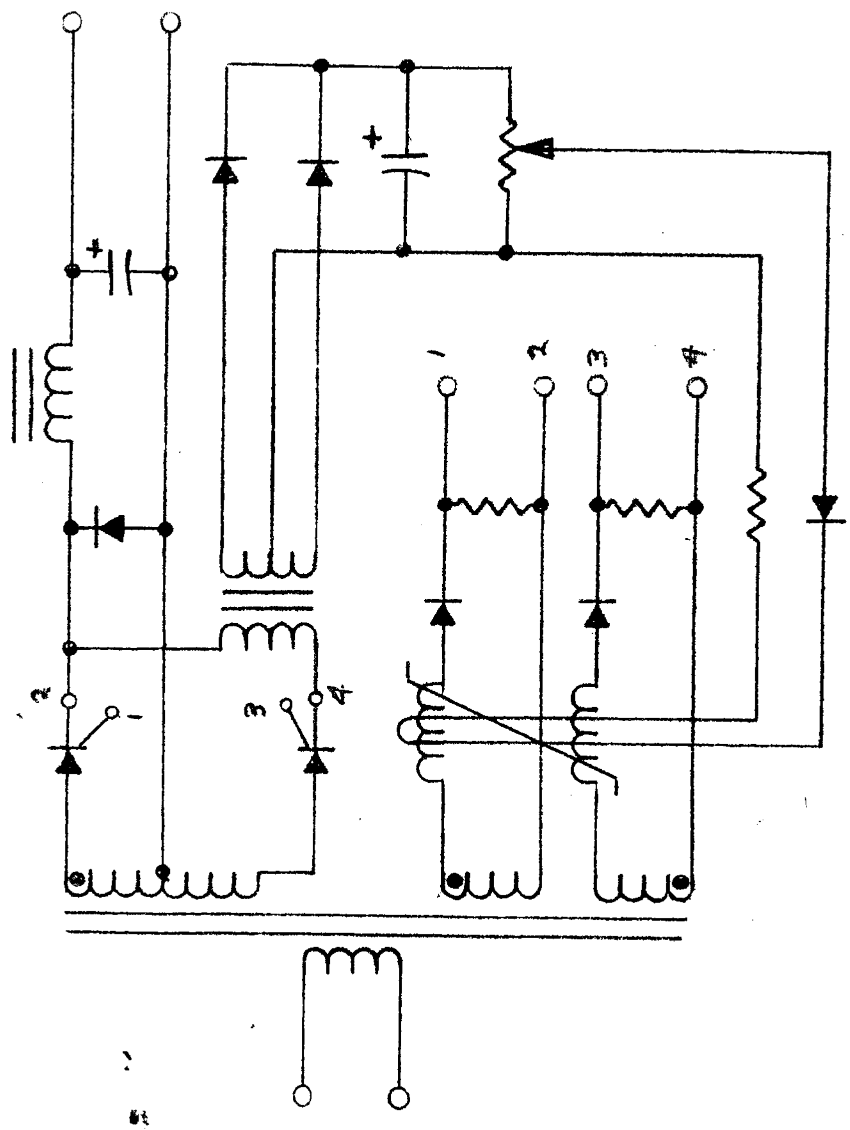

FIG. 9 SCHEMATIC SILICON CONTROLLED RECTIFIER REGULAR

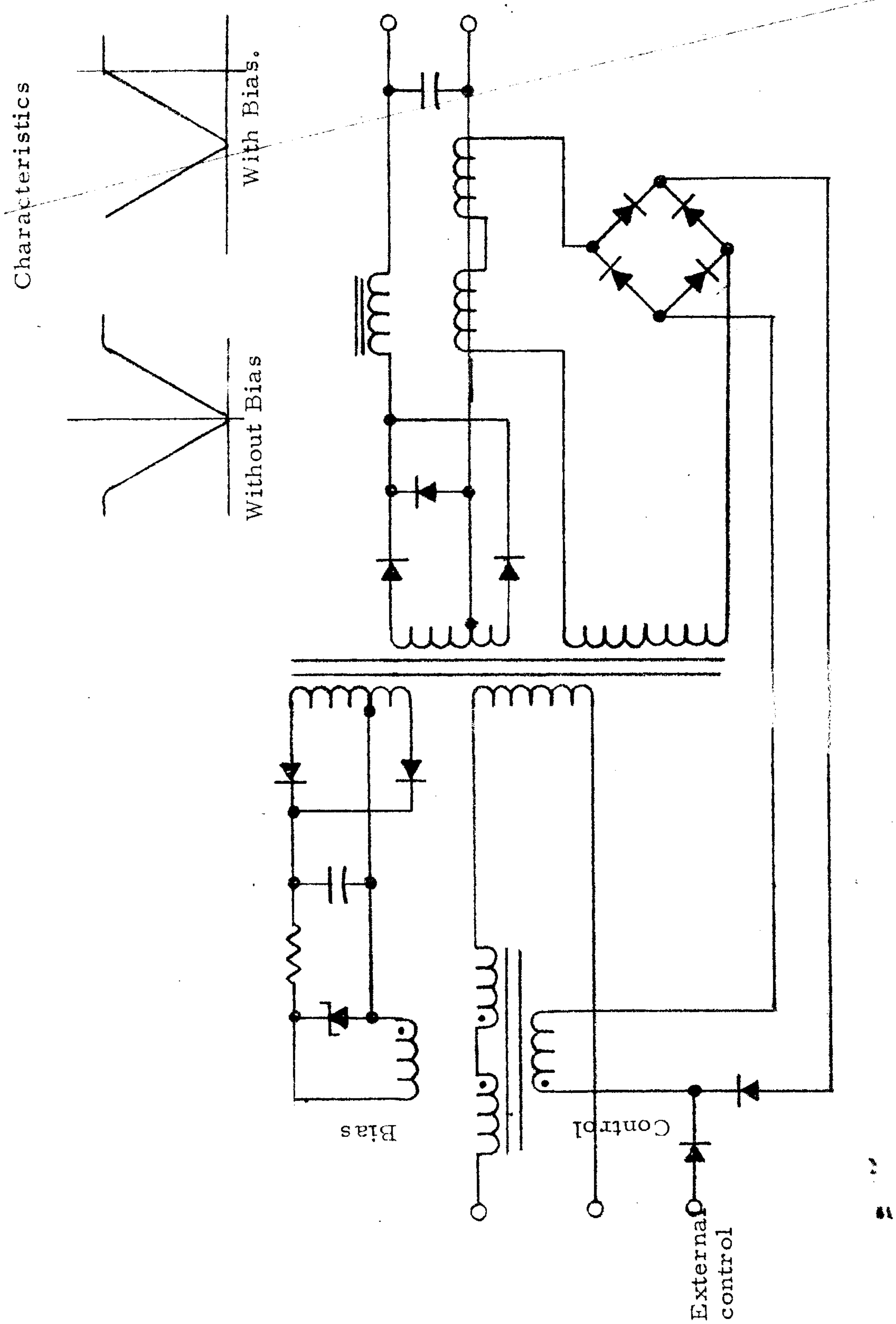

FIG. 10 SCHEMATIC SATURABLE REACTOR REGULATOR

lightest of the design approaches, however, requirements for transient protection and increase of switching losses at higher frequencies precludes its use. The best overall design approach is the self-saturating mag-amp, and this has been designed for the magnet supply application.

2.2.4.3 Circuit Parameters

Output:

5.5 VDC   1% ripple

2.7 amps. max.

14.85 watts max.

Regulation:

Current regulated $\pm 1\%$.

Losses:

Reactors	0.3 V x 1.35 A x 2 = 0.810 watts
Diodes	0.7 V x 1.35 x 2 = 1.9 watts
Switching losses	.1 x 0.95 x 2 = 0.19 watts
Filtering losses	0.1 V x 2.7 = 0.27 watts
Control losses	.022 x 15/8 = 0.41 watts
Total losses	3.58 watts

Conversion Efficiency:

$= (14.85/14.85 + 3.58) \times 100$

$= 14.85/18.44 = 80.6\%$

Rect. Diodes

Rect. Diodes	3.6 grams	3 required	10.8 grams
Sig. Diodes	2 grams	6 required	12.0 grams
Zener Diodes	2 grams	1 required	2.0 grams
Mag-amps.	10 grams	2 required	20.0 grams
Current Transformers	30 grams	1 required	30.0 grams
Capacitors	30 grams	2 required	60.0 grams

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Total Weight 134.8 grams
= 0.297 lbs.

2.2.5 Arc/Cathode Heater Power Supply

2.2.5.1 General Discussion

The gas discharge or electron bombardment ion engine produces ions in the plasma region of a gas collision type ion source. The gas collision ion source is a low pressure arc where ionization is due, almost entirely, to collisions between atoms and electrons with a kinetic energy greater than the ionization potential level. The arc supply required for this application has an output capability of 25 amperes maximum. Current limiting is employed to limit the current to 20 amperes when the arc resistance is between 0.3 and 0.4 ohm. A cathode heater power supply is also provided with an output capability of 8 volts AC at 10 amperes. Since the ions are obtained by the collision of electrons with atoms, control of the ion output can control the cathode heater temperature. Under normal operating conditions, there is sufficient thermal coupling from other engine heat sources to maintain the cathode temperature at, or slightly above the desired temperature. During this period, power is not required and must be controlled. Such control is effected by sensing the arc current and using the resultant output to turn the cathode heater power supply on when arc current is below 10 amperes and off when the arc current is above 15 amperes.

Several approaches were studied for the supply. The simplest possible approach, that of placing a high reactance in the AC portion of the circuitry sufficient to provide 5 percent regulation, disregarding master transformer considerations, could be built with an efficiency of 82 percent.

However, 3 KVA of reactive power would have to be produced. Therefore, this approach was not considered further.

A conventional self-saturating mag-amp design promises approximately 83 percent efficiency and a separate design for this system was prepared.

Regardless of the system used, if conventional power rectifiers are used, efficiency is limited because of the losses in the rectifiers. A design to partially circumvent the rectifiers losses is a synchronous switching rectifier. This system, as shown by schematic and the following paper design, has an efficiency of 88 percent using "worst case" parameters. The system produces a higher efficiency because the transistor losses are less than a power rectifier.

Several improvements on the circuit are planned based on breadboard work. The first is optimizing the drive circuitry for the transistors used. It is believed efficiency can be increased above predicted efficiency by this approach.

It is evident that paralleling switching transistors will increase efficiency. In fact, a work-out reveals efficiency to 92 percent are possible using three transistors in parallel. The drawback is that perfectly matched transistors are not available and therefore would not share the loads equally. The outputs of each supply would be connected in parallel. This system would be somewhat heavier and perhaps less reliable. The latter being a debatable statement, but certainly would require more components and circuit connections.

2.2.5.2 Design Considerations

The input is a quasi-square wave

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formed by adding two square waves; one delayed with respect to the other, depending on the input voltage level. At nominal input, the square wave has an "off" time of 26 percent each half cycle, or 54° "off" and 126°on. In a quasi-square wave;

$$E\ RMS = E_{peak}\ ft$$

$$E_{ave} = E_p\ ft$$

$$E\ RMS = E_{ave}\ \frac{\sqrt{ft}}{ft}$$

where f = frequency

t = time "on"

ft = "on" time % (½~)

At nominal then, $E\ RMS = E_{ave}\ \frac{\sqrt{.76}}{.76} = 1.14$

If additionally, the mag-amp also must reduce the "on" time by 25 percent more to exercise control from 6 - 8 volts, the "on" time at nominal will be approximately 50 percent, and E RMS = 1.4 E_{ave}. This number is used in Arc DC and mag-amp design. The factor 1.29 (60%) was used for the current transformer because it has adjustable output.

2.2.5.3 Parts Selection and Evaluation

Transistor Considerations. The only transistor presently available with desirable characteristics for this application is the Honeywell MHT 8300 series. The manufacturer will supply a modified 8303 with a $V_{ce(sat)}$ of .5 VDC max. at 20 amp Ic and 2A Ib. This silicon NPN's fast switching time and other ratings make it the choice for the arc design.

The bulk of the losses occur in the conduction state, which is the reason for the effort to obtain low $V_{ce(sat)}$. These losses are reduced considerably if a parallel technique is used.

The reverse current losses are very low and are not considered in the total loss column.

Diodes selected have typical reverse recovery time of 70 nanoseconds, their losses are insignificant at the low level of operation. They are derated at least 3:1 in the arc supply design.

Switching Loss:

$$P_c = \frac{1}{3} \frac{Ts}{T} I_{cs}^2 \text{ RL max.}$$

$$T_s = \frac{Ton + Toff}{2}$$

$$P_c = \frac{1}{3} \left(\frac{1.7}{100} \times 20^2 \times .4\right) = .905$$

Saturation Loss:

$$P_s = \frac{1}{2} I_{ce} Vce(sat)$$

$$= \frac{20 \times .5}{2} = 5.0$$

Base Drive:

$$P_b = \frac{1}{2} V_b I_b$$

$$= \frac{2 \times 2}{2} = \frac{2.0}{}$$ loss/transistor = .905 + 5 + 2 = 7.905 watts

Total loss =7.905 x 2 = 15.8 watts

<u>Filter Inductor</u>. Because of the very low load impedance, a choke offers the most as a choice between a choke and a capacitor filter. The ripple permissible is one volt RMS. At low line voltage and high arc impendance, the rectifier output would not require any filter to meet the ripple spec. However, at high line and low arc impedance, there is need

for a filter.

Conditions that produce $\frac{E\ RMS}{E_{ave}}$ = 1.4 are assumed for inductor design.

Ep-p across L = 11.2; I = 20 A

$$E = -L_c \frac{di}{dt}$$

$$L_c = \frac{25 \times 5.6 \times 10^{-6}}{20} = 7\ \mu h$$

To provide a measure of short circuit protection and adequate filtering, a much larger choke will be designed with 50μh as a goal.

2.2.5.4 Circuit Description

The design combines a pair of transistors Q301 and Q302 (see schematic Fig. 11) with a small self-saturating half-wave magnetic amplifier driving the base-emitter junction. An in-phase voltage is taken from the master transformer and applied to the base to drive the transistor to the saturated "on" state in synchronism with the source voltage, producing a DC output via full wave rectification. The system produces a higher efficiency because the transistor losses are less than a power rectifier and the base drive mag-amp losses are less than the losses of mag-amps located in a high current portion of the circuit.

The mag-amps have two control windings, plus a bias winding. The bias winding is not operated from a stable source as might be expected, but its control is very limited and intended to merely remove the effect of the reverse diode current. One control winding is provided for an overload signal to reduce the arc power in the event there is a high voltage

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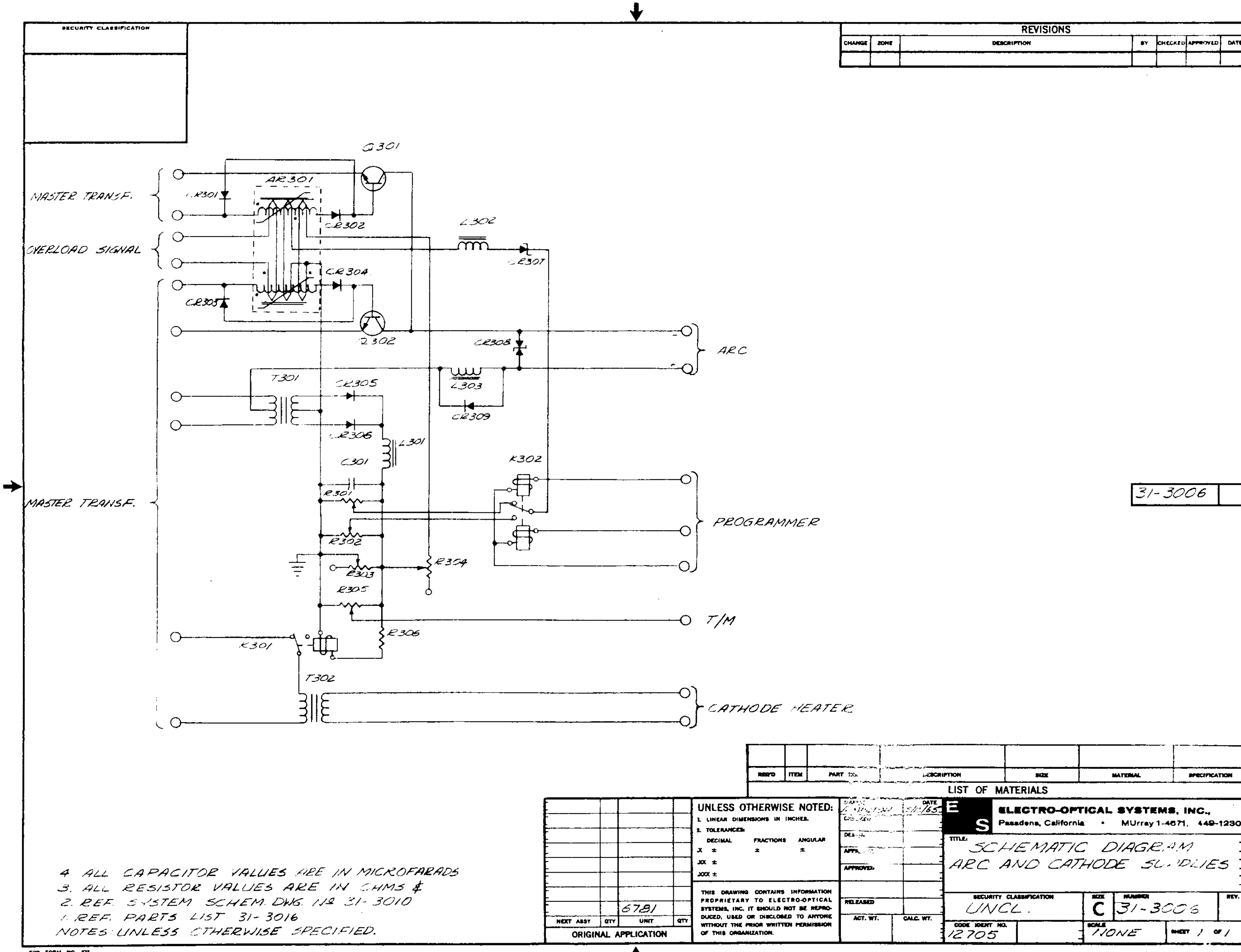


FIG. 11 SCHEMATIC ARC/CATHODE HEATER SUPPLY

arc-over. The second control winding operates from a current sensing transformer and exercises control after a predetermined current has been exceeded, limiting the output current.

The mag-amp has two cores, the gate winding of each being oriented so no fundamental voltages are induced in the control windings. The control windings are common to both cores. Diodes CR302 and CR304 pass the proper polarity drive current for the transistors, CR301 and CR303 provide a reverse turn-off drive for the transistors to minimize reverse leakage currents and switching times.

T301 is the current sensing transformer, the system includes a full wave rectifier-filter, and is heavily damped with resistance to obtain linear operation, but the total power consumed is only slightly over one watt.

K301 is energized by the current sensing supply whenever the arc current is greater than 15 amperes. The relay is de-energized whenever the arc current drops below 10 amps.

The arc current is filtered by a single choke, L303. Diode CR309 is used to dissipate any inductive voltage generated, should the arc be extinguished suddenly. CR308 is used to clamp the entire circuit at a relatively low voltage if large transients are generated externally by high voltage arc-overs.

K302 permits the arc current to be regulated at a second level during engine warmup; it is controlled by the programmer.

T302 is the source of cathode heater current. It is used to permit switching at a low current level and also to permit insulating the relay from the high voltage.

The mag-amp control winding, current sensing circuitry and all other switching is accomplished at essentially ground potential.

Circuit Efficiency:

Losses -

Transistors	15.8	watts
Mag-Amp	.12	watts
Mag-Amp Diodes	1.2	watts
Current Sensing &	1.1	watts
DriverFilter	.6	watts
Total System Losses	18.82	watts

Nominal power is 140 watts.

$$\text{Eff.} = \frac{140}{140 + 18.82} \times 100 = 88\%$$

Above does not include cathode heater supply as it is not used during run phase of engine.

2.2.5.5 Circuit Parameters

Total weight of electronic components is 1.23 lbs.

Circuit Weight (does not include hardware or structure):

Transistors	50.2 gm.	
Diodes	38.5 gm.	
Relays	42.3 gm.	
AR301(M.A.)	15.3 gm.	
T301	10.5 gm.	
T302		.29 lb.
L303		.58 lb.
Resistors	0.2 gm.	
Cap	7.6 gm.	
Totals	164.6 gm.	0.87 lb.

.87 + 164.6 x .0022 = 1.23 lbs.

2.2.6 Vaporizer Heater Power Supply

2.2.6.1 General Discussion

The vaporizer heater supply is an AC supply with a maximum output of five volts at five amperes RMS. The design approaches applied to the magnet power supply also apply here since both supplies are relatively close in power level. All of the arguments in favor of the magnet supply design approach also apply for the vaporizer heater power supply. The output is not regulated but is controlled externally by either the arc current or the beam current, but not by both simultaneously.

2.2.6.2 Circuit Description

Figure 12 is the schematic of this design. The vaporizer input voltage is stepped to the desired level by a transformer (T_1). This voltage is then applied to the gate windings of mag-amp M_1. The control windings of the mag-amp are driven by the arc current or the beam current signals which determine the AC output of the supply. Output current is sensed by a current sensing network (CT1 and associated components) where a DC voltage is developed and applied to the control windings of the mag-amp. CR8 and CR9 form an OR gate. When the load current is above a set value, the voltage developed at CR9 exceeds that of CR8. This makes the cathode of CR8 more positive than its anode and the supply is in a current regulation mode. The supply will be protected against overloads and short circuits in this manner. The setting of R2 determines the maximum current that can be drawn from the vaporizer supply. Another control (No. 2) is obtainable through a separate bias winding. R4 will limit the current through this winding.

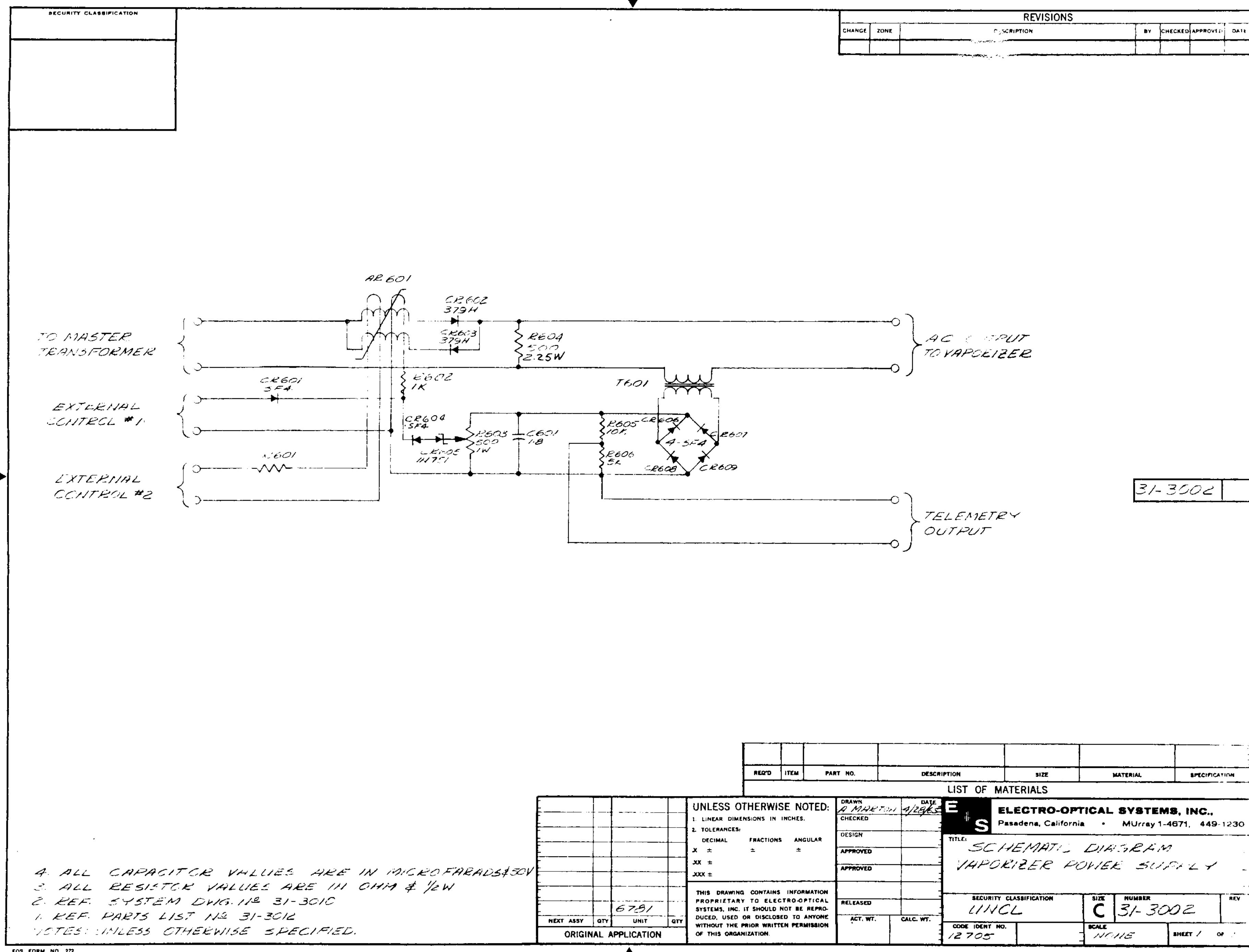


FIG. 12 SCHEMATIC VAPORIZER HEATER SUPPLY

2.2.6.3 Circuit Parameters:

Power Losses:

Reactors 0.3V x 2.5 x 2	= 1.5 watts
Rectifier Diodes 0.7V x 2.5 x 2	= 3.5 watts
Switching losses 0.1 x 3.5	= 0.35 watts
Bleeder Resistor	= 0.5
Control Losses 0.04 x 10	= 0.4
Total Losses	= 6.25 watts

Efficiency

input = 25 + 6.25 = 31.25 watts

output = 25

efficiency = 25/31.25 x 100 = 80.5%

Weight

Reactors	40 gms.	2 req.	80 g gms.
Rect. Diodes	3.6 gms.	2 req.	7.2 gms.
Signal Diodes	2 gms.	7 req	14.0 gms.
Resistors	5 gms.	3 req.	15.0 gms.
Current Transformer	40 gms.	1 req.	40.0 gms.
Capacitors	30 gms.	1 req.	30.0 gms.

Total Weight 186.2 gms. =

0.410 lbs.

2.2.7 Programmer

2.2.7.1 General Discussion

Proper operation of the EOS electron bombardment cesium engine requires that power be applied in a particular sequence. The cesium fuel must be preheated to liquid form, electron emitting cathodes must be preheated, electrodes must be heated to optimum temperatures, and high voltage must be applied at the proper time.

The following program is required to operate the EOS electron bombardment engine:

Time (Minutes)	Command	Function
T + 0 –	Start Engine Command	Initiates Preheat I Phase
	1. Misc. Inverter ON	1. Starts low voltage inverter, activates cathode heater, magnet power, arc power and neutralizer cathode heater.
	2. Reservoir - Valve Inverter ON	2. Applies power to Reservoir and Valve Heaters.
T + 30	Timer Command	Start Preheat II Phase
	1. Turn on Engine Vaporizer Heater with feedback control from arc current.	1. Arc current is started with feedback control to warm engine chamber and electrodes.
	2. Turn off **Valve Heater.**	2. No longer needed, since valve **is** opened by this time.
T + 60	Timer Command	Clean Engine & Start Neutralizer System
	1. Turn off Vaporizer Heater.	1. Stop cesium flow to engine chamber.

	2. Turn on Neutralizer Vaporizer Heater and regulate to simulated probe voltage.	2. Start Neutralizer system.
T + 65	<u>Timer Command</u>	<u>Turn on Engine Beam</u>
	1. Turn on High Voltage Supplies.	1. Start Beam.
	2. Turn on Engine Vaporizer and switch to beam current feedback.	2. Beam regulation.
	3. Switch Neutralizer regulation to beam potential control.	3. Neutral beam current regulation.
T + ____	<u>System Command</u>	<u>Shut Down Engine</u>
	1. Turn off Misc. Inverter.	1. Shuts down Low Voltage Inverter.
	2. Turn off High Voltage Inverter.	2. High Voltage off.

If the engine is to be used for a single cycle, long term operation as a possible power source for long distance space flights, this engine turn-on cycle could be derived as part of the master vehicle program in the form of a subroutine, thus eliminating the need for a system programmer. Since flight mode has not been determined, a flight programmer has been designed for this power conditioning system, which will accept an engine start/stop command and function accordingly.

2.2.7.2 Programmer Design Considerations

The programmer in this system is used only in support of the laboratory test. An electro-mechanical timer was selected as being the simplist and most reliable way of performing this function.

Mechanical Timers. A mechanical timer has been selected as the timing device to be used in the programmer for this system, because of its proven reliability. These devices have been used in previous EOS ion engine power conditioning systems very successfully and have operated both in the lab and in space without failure. They have been designed to meet space environments and have successfully operated in a nuclear radiation environment. Most important, they are unaffected by high energy transients caused by high voltage arcing at the engine.

These mechanical devices are motor driven type timers using cam operated switches to produce the desired pulse duration, switching sequence and cycle time. It consists essentially of a D.C. motor, adjustable cams affixed to a cam shaft, miniature micro-switches and a gear train connecting the motor to the cam shaft. An RF filter is used to minimize noise generated by the D.C. motor. The total device is enclosed in a hermetically sealed case and weight is ten ounces.

The cycle time is determined by the motor used and the gear train. One switch in the timer is used to control the motor and indicate the end of a cycle. The cams can be adjusted to energize the command switches at any time during the timing cycle. Figure 13 is a schematic diagram of the mechanical timer to be used.

2.2.7.3 Programmer Functional Description

The programmer has been designed to turn on/off the electron bombardment cesium engine upon receiving on/off commands in the form of positive 28 VDC pulses from the spacecraft command system or from the ground test equipment. Upon receiving an "engine on" command, relay K101 will direct the system primary D.C. power to the mechanical timer motor to start the engine turn-on sequence. Since

56

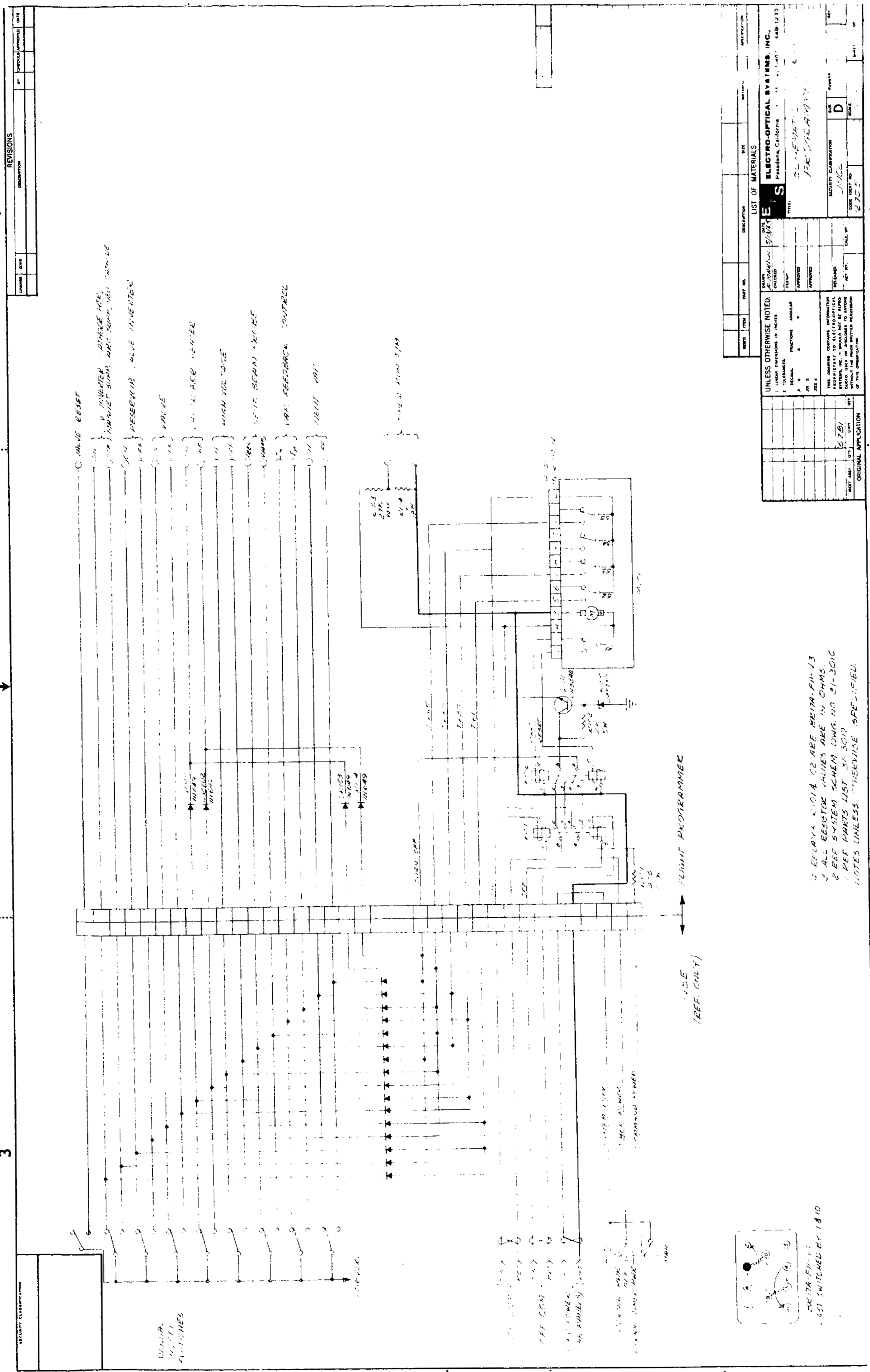


FIG. 13 SCHEMATIC PROGRAMMER

primary power in this case is derived from solar panels, the voltage is expected to vary from 50.4 VDC to 84 VDC. To obtain some degree of accuracy and repeatability, the timer motor voltage is regulated by a simple emitter follower regulator consisting of Q101, CR105, and R102.

As the timer cams are advanced by action of timer motor, each timer switch is activated to provide the system commands required. Each command is routed to the appropriate power supplies in the system by way of the programmer interface connector. Blocking diodes are placed in series with the command lines to allow manual control of the system during ground tests without interaction. Since these diodes are not necessary during flight, a flight connector will be used to provide connections between the timer and the system directly.

As the timer completes the "turn-on" cycle, the internal motor switch, S1, is activated. S1, in turn, energizes relay K102, which turns off motor power. The programmer is now in a static state, drawing no power from the prime power source while the engine is in full operation.

A "system off" command is given to shut the engine down. This command energized relay K101, which applies a turn-off command to the system relays by way of relay K102. This action also energizes the timer motor again. The timer will now reset itself back to the zero position where it will again energize K102 and thus turn off its own operating power. The programmer is again inactive and is ready to accept the next "system on" command for the next engine turn-on sequence.

Diodes, CR101 through CR104, are used to control the vaporizer heater power supply at the required times and prevent command interaction to the other system functions.

Resistor, R101, is a series dropping resistor to limit power dissipation in the command relays throughout the system. Command duration is minimized to three minutes. This minimum pulse period is limited by timer cam configuration and reliable switch activation parameters.

Resistors R103 and R104 provide a telemetry signal to indicate that voltage is applied to the timer motor and is, therefore, operating.

2.2.7.4 Design Description

Mechanical Timer. The mechanical timer design aspects are important, since it is the heart of the programmer. The timer is designed to the following specifications:

A. Motor: Direct current permanent magnet field -
(Start) 28 Volts, 100 milliamps maximum
(Running) 45 milliamps nominal

B. Switches: Precision snap action type SPDT -
Contact Rating: 2.5 amps inductive, 5 amps resistive at 28 VDC

C. Cycle Time: 75 minutes $\pm$ 10% at 25°C
75 minutes $\pm$ 15% - 55°C to +125°C

D. Switch Settings: $\pm$ 2% of specified time in relation to total cycle time.

E. Life and cycles to be specified according to system operating requirements.

F. Environmental specifications according to overall system requirements, stated elsewhere in this report.

G. Entire unit to be hermetically sealed and purged with dry nitrogen.

A prototype unit has been fabricated and functionally checked successfully.

The Voltage Regulator. An emitter follower voltage regulator is used to regulate the solar panels power to the timer motor. This is necessary to maintain timer accuracy, since command times are in direct relationship to the timer motor speed. The emitter follower design was selected to provide sufficient regulation and still minimize power dissipation and component sizes and weight. Regulation of 4% can be expected over the environmental range. This is adequate since the engine programming accuracy is not stringent.

All components selected in the design allow a 100% safety factor in the worst case condition.

Resistor R101 is in series with the system relay command power controlled by the timer. Its value is selected to limit power dissipation in the relay coils. This was necessary because of the wide input voltage range of 50 - 84 VDC and the design limitations of the system relays.

The system relays have 3 K dual coils which are rated for 50 VDC operation. Maximum recommended power dissipation of these relays at $125^{\circ}C$ is 1.25 watts. The value of R101 will vary if the number of relays to be commanded by one timer switch exceeds the maximum of 5 presently used.

2.2.8 Neutralizer System

2.2.8.1 General

The neutralizer consists of a cathode, a cesium vaporizer or feed system and an engine beam potential probe. A.C. power is required for the cathode and vaporizer heaters while D.C. bias power is required for the beam potential probe.

Internal feedback regulation is not required in any of these power supplies, as the vaporizer is controlled by the beam or probe potential in a system feedback loop. Regardless of heater power and bias voltage variations, the control of the vaporizer heater by the probe (or beam) potential regulates the system for maximum neutralization, since the beam (or probe) potential is inversely proportional to the degree of neutralization. The feedback is such to increase the vaporizer power, and hence electrons available for neutralization of the beam with increasing beam potential.

2.2.8.2 Circuit Evaluation

It was found that efficiency is higher utilizing the low voltage inverter transformer secondary windings, since the iron losses (and weight) of the separate transformer technique are eliminated. The copper losses in this case may be higher or lower, depending upon the relative number of turns required to provide the output voltage. Any increase in copper losses, is, however, offset by the difference in iron losses as they are the greater portion of the power lost. The iron losses of the low voltage inverter transformer are constant and independent of load or output voltage.

With this reasoning, it was decided to derive the required power supply voltages directly from secondary windings on the low voltage inverter transformer.

Neutralizer Cathode. The requirements of the neutralizer cathode heater are relatively simple: an unregulated source of A.C. voltage with turn-on and turn-off capabilities and instrumentation for remote measuring of the output voltage and current. This fixed the configuration of this power supply to that of a SPDT latching relay to connect and disconnect the cathode heater from the master transformer secondary winding.

A measure of the rms level of the output voltage (quasi square wave) is obtained by rectifying a sample of the output voltage and then integrating this with an RC filter.

A measure of the output current is obtained by means of a current transformer connected in series with line. The transformer steps up the small voltage drop appearing across its primary circuit. The primary and secondary voltages are proportional to the line current. The secondary voltage is rectified and filtered to obtain a D.C. voltage proportional to the rms value of the load current.

Continuous adjustment of the output voltage is not necessary. Once the optimum voltage for a particular engine is determined, it can be set permanently. This will be accomplished by multiple taps on the master inverter transformer secondary.

Beam Probe Bias Supply. The bias supply consists of a simple full wave rectifier with turn-on and turn-off capabilities.

A full wave rectifier with center tapped transformer secondary was selected over the simpler full wave bridge rectifier as the diode losses, which were the greatest portion of the circuit losses, were double those of the two diode circuit. The center tap secondary winding, though requiring twice the number of turns as the full wave bridge circuit, will have the same copper losses.

A choke input LC filter was selected for its load regulating characteristic and because the choke will also prevent high initial surge currents that occur with a capacitor input filter.

Instrumentation for measuring the output voltage consists of a simple resistive voltage divider. Instrumentation for measuring the D.C. output current, however, presents a much more complex problem. A low value of resistance could be inserted in series with the output to measure the current, but this is impractical and wasteful of power, since a full scale output of 5 volts is required. A low resistance shunt could also be used to measure the current and the resulting voltage drop across it amplified by a solid state D.C. amplifier.

A stable solid state D.C. amplifier, however, is complex and would have questionable reliability. Instead, the magnetic transductor for measuring the load current was selected. These devices are inherently stable and highly reliable and can be designed to have low power loss.

Vaporizer Heater. The requirements of the vaporizer heater supply are to have turn-on and turn-off capabilities and to supply continuously controllable power to the vaporizer heater. This continuous control must be accomplished by application of a positive D.C. voltage for use in a feedback control loop. In addition, instrumentation must be provided for remote measurement of the output voltage and current (rms values).

Turn-on and turn-off capabilities are provided by a double pole, double throw latching relay, which opens the control and bias windings of a saturable reactor, which is used to control the output voltage. This technique was used rather than switching the line directly so that a lower rating and smaller relay might be used.

Silicon controlled rectifiers were considered as the active element to control the output power, but their switching losses are excessive at the operating frequency (10 Kc). In addition, problems are also encountered in properly driving two SCR's to obtain full wave switching, since a balanced A. - C. line was not used.

The use of transistors as series regulating or switching elements was also considered, but also found to have excessive power losses.

The use of a simple saturable reactor to control the vaporizer heater voltage was selected on the basis of low power losses and high reliability. This technique has several advantages:

the control circuits are electrically isolated from the output circuit; multiple control windings may be used to obtain many desirable characteristics; and saturable reactors are extremely rugged.

In addition to turn-on and turn-off capabilities, means are provided to switch the vaporizer control from a fixed simulated beam potential to the actual beam potential. This capability is provided by a DPDT latching relay, which switches the saturable reactor control winding between these two points on command from the system programmer.

The instrumentation used is identical to that used on the cathode heater supply. To obtain a measure of the A.C. output voltage a sample of it is rectified and filtered, providing a D.C. output voltage proportional to the output rms voltage. A quasi current transformer in the return side of the A.C. line samples the output current. The small voltage drop across the transformer primary is stepped up by the transformer producing an A.C. voltage of several volts. This voltage is rectified and filtered to give a D.C. voltage proportional to the output current.

2.8.3 Circuit Description

A functional block diagram of the neutralizer is shown in Fig. 14. Here it can be seen that the cathode heater is independent of the other supplies.

The probe bias supply is a common full wave rectifier and filter. The probe is fed through a series resistance so that the output or probe voltage is lower than the power supply output. The true beam voltage is fed back to control the vaporizer output power during normal operation. During the warmup mode the power supply output is used to simulate the probe potential. Selection of these modes is accomplished by means of a DPDT latching relay.

The vaporizer heater consists of a saturable reactor which controls the output by means of the beam probe potential. These items comprise a feedback control loop which operates to maximize neutralization.

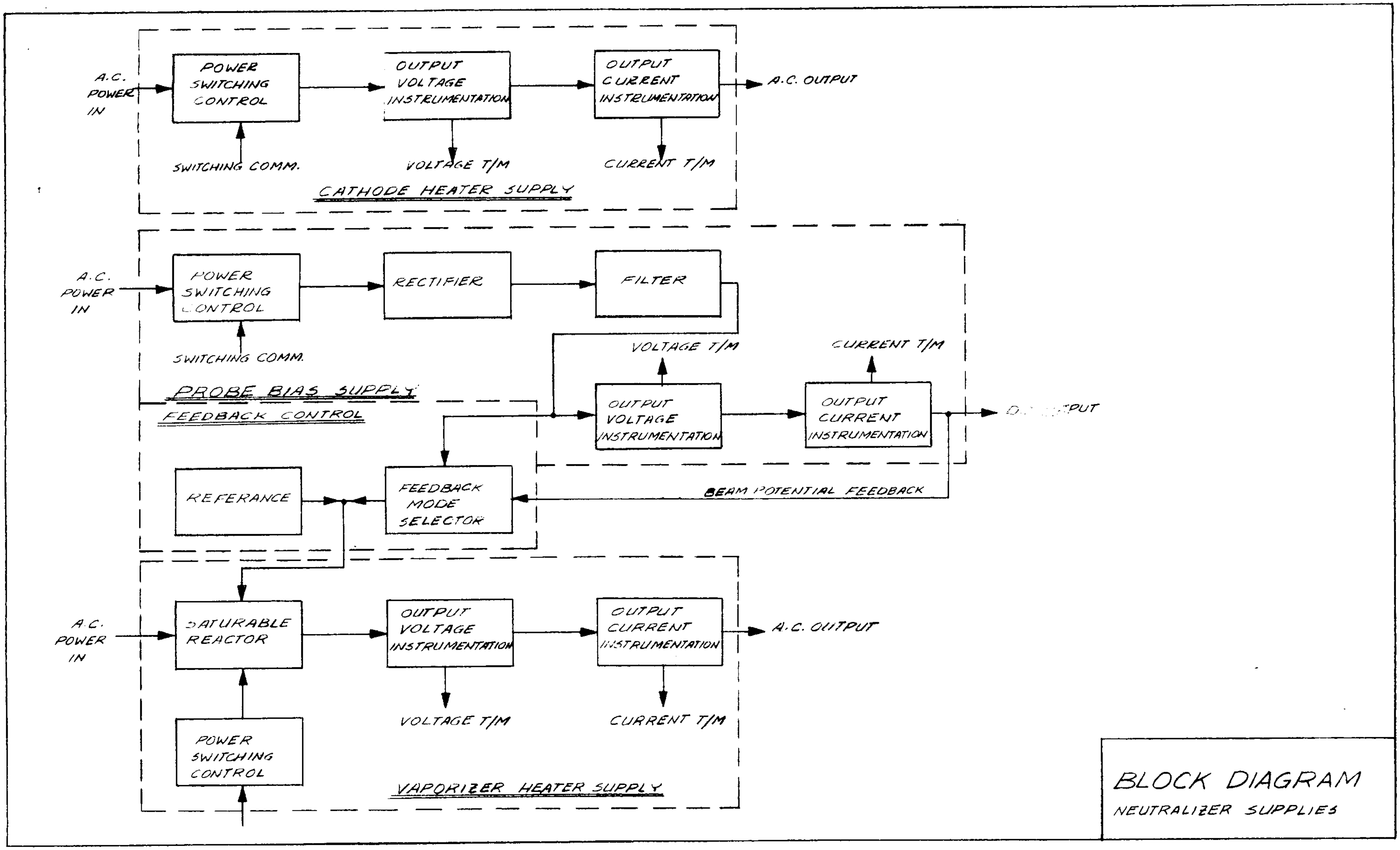


FIG. 14 FUNCTIONAL BLOCK DIAGRAM NEUTRALIZER

In addition, each supply has instrumentation available for making remote measurements of all the output voltages and currents.

The detailed operation of the neutralizer is easily explained with reference to the neutralizer schematic in Fig. 15.

The cathode heater receives A.C. power from the low voltage inverter at a frequency of 10 kc. The voltage is adjustable to two levels at 2.6 and 3.9 volts by changing taps on the output of the inverter transformer. A more flexible form of adjustment is not necessary as the adjustment is permanent once the engine is set up.

Proportional D.C. analog voltages give a measure of the output voltage and current. CR404 rectifies the output voltage, which is a quasi square wave. R410 and C402 comprise an integrating filter, which averages the rectified output from CR404. Referring to Fig. 16, ideally the rms value of a quasi square wave is:

while the average value of the half rectified wave is:

$$E_{ave} = \frac{E}{2}\left(1 - \frac{2t_D}{T}\right)$$

The fact that the average value is not related to the rms value in a linear manner is immaterial, since calibration curves of the average voltage, as a function of the rms value of the output voltage, will be available. The voltage drop across CR404 (and CR406 in current circuit) will also contribute to the nonlinear relationship. Resistors R409 and R419 provide discharge paths for the filter capacitors in the event the circuit is operating and the instrumentation circuits are not in use.

Transformer T401, in series with the load

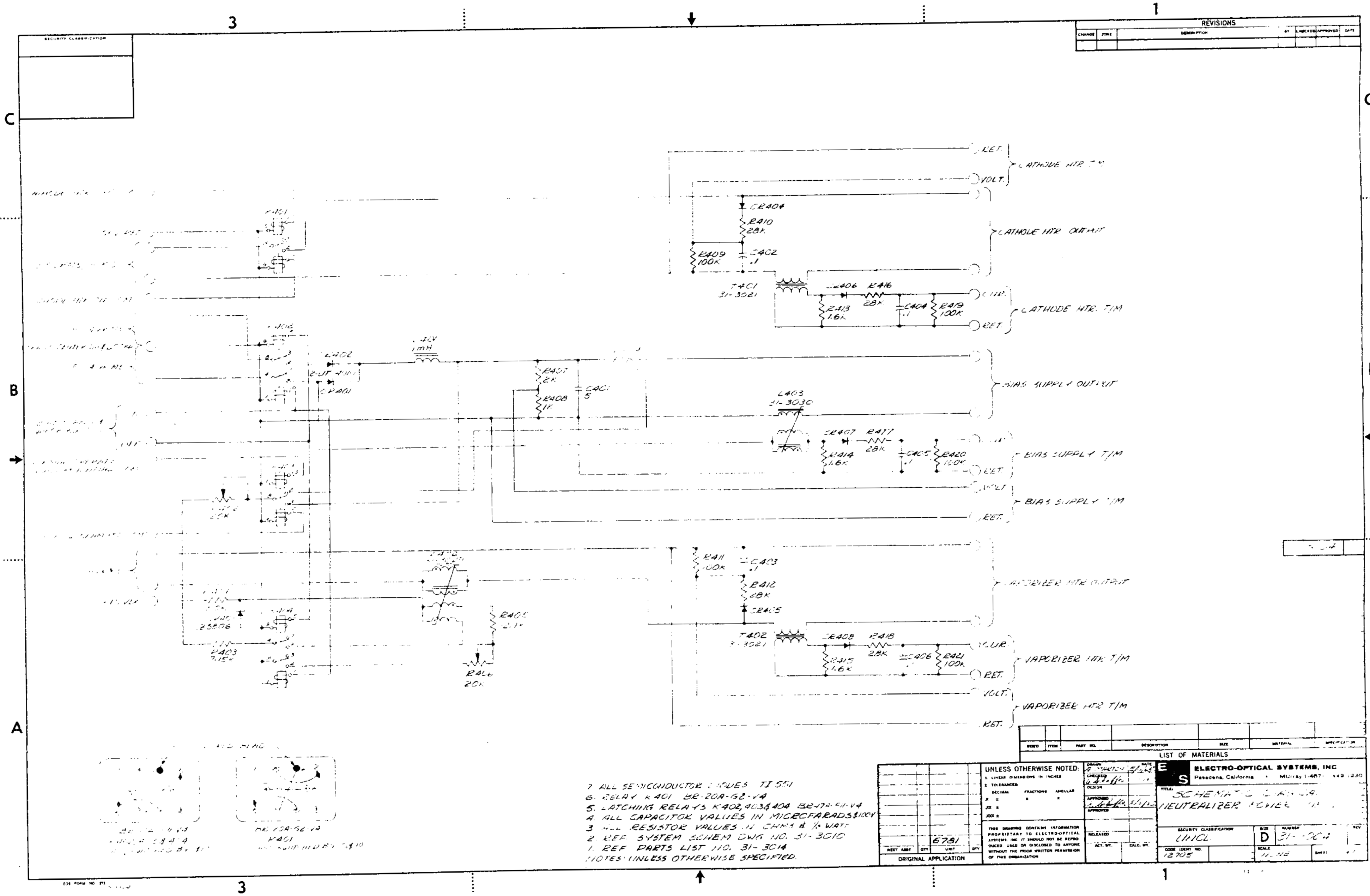


FIG. 15 SCHEMATIC NEUTRALIZER

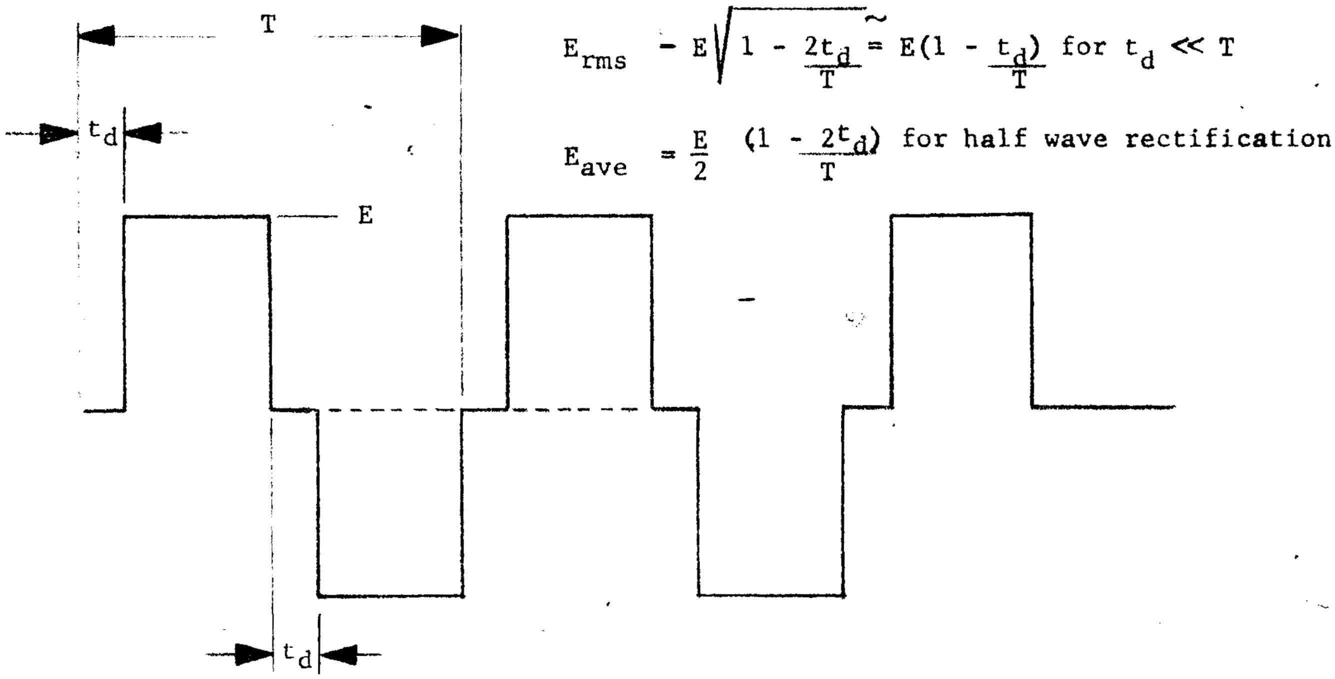


FIGURE 16 QUASI-SQUARE WAVE AVAILABLE FROM LOW VOLTAGE INVERTER.

impedance, provides a means of measuring the output current of the cathode heater. T401 is a current transformer that reflects a very low impedance into its primary circuit and, thus, has little effect upon the output circuit current. The output current develops a small voltage drop across the transformer primary, which is stepped up to several volts on the secondary side. The voltage developed across R413, which proportional to the primary current, is rectified by CR406. The resulting rectified voltage is filtered and aceraged by R416 and C404. This D.C. voltage is a measure of the rms value of the output current.

Bias Supply. A.C. power for the bias supply is provided by a center tapped winding on the low voltage inverter transformer. Three taps provide 9, 13, 14 volts rms. Relay K402 switches power to the circuit applying power to rectifying diodes CR401 and CR402. L401 and C401 form a choke input filter, which will reduce the output ripple voltage to one percent or less. Resistors R407 and R408 form a D.C. voltage divider for remote measurement of the power supply output voltage. R404 is a series dropping resistor used to feed the bias voltage to the beam probe, yet allow the probé potential to drop to lower values.

Saturable reactor L403 controls an alternating current through R414. The driving voltage is derived from one side of the center tapped driving transformer. The voltage drop across R414 is rectified by CR407 and integrated and filtered by R417 and C405. Bias supply current passing through the control winding of L403 controls the current through R414, and hence the D.C. voltage across C405. This voltage is a measure of the bias supply current.

The beam probe voltage and the bias supply voltage are each fed to the vaporizer heater circuit through the contacts of relay K403 which selects the feedback mode.

Vaporizer Heater. A.C. power is supplied the vaporizer heater from a multiple tapped secondary winding on the low voltage inverter transformer. Voltages available are 2.6 and 3.9 volts rms.

Relay K404, which is controlled by the system programmer, switches power to the control and bias windings of L402. When the control and bias windings are opened, the cores of L402 are unable to saturate, holding the device in an off condition. In addition, L402 provides continuous control of the vaporizer output power.

The control winding of L402 receives a driving current from the beam probe potential or from the bias supply output through relay K403. Reference diode CR403 acts as a voltage reference for the feedback signal. Potentiometer R402 changes the gain and bias point of L402. In order that the gain may be adjusted without disturbing the bias point, a bias winding of opposite polarity is provided. It is controlled by potentiometer K406. A stable bias voltage for the bias winding is provided by reference diode CR403.

Instrumentation for remote measurement of output voltage is provided by CR405, R412, C403, while T402, R415, CR408, R418, and C406 provide means of measuring the output current. These circuits are identical to those used in the cathode heater.

2.2.8.4 Circuit Parameters

The important circuit parameters are listed in the following chart. All quantities shown were calculated for maximum power output, but bo not include relay coil losses.

Circuit	Input Volts Rms	Input Amps Rms	Output Volts	Output Amps	Component Weight, Oz.	Losses, Watts	Efficiency %
Cathode Heater	2.6 & 3.9	2.6 & 3.9	2.6 & 3.9 Vrms	2.6 & 3.9 Vrms	1.84	.081	99%
Bias Supply	9 to to 14	.14 to .42	9 to 14 VDC	0 to .5 A D.C.	1.37	.65 *	90%
Vaporizer Heater	2.6 & 3.9	0 to 3.9	0 to 3.9	0 to 3.9	1.29	.47 to .61	96% to 97%

* Losses due to R404 not included as it is considered as part of load impedance.

2.2.9 High Voltage Power Supply

2.2.9.1 General Discussion

The high voltage power supply consists of a positive high voltage section capable of 2000 volt D.C. output at 408 mA nominal and a negative high voltage section capable of -600 volt output at 20 mA maximum. Neither of these outputs need to be regulated.

The design of the high voltage supply was approached from the standpoint of providing the necessary high voltages by the most efficient and reliable method that would yield minimum weight and package size. Efficiency considerations dictated the use of fast switching transistors with low saturation characteristics. Efficiency was also a major concern in the design of the necessary transformers. Core materials with low loss characteristics were selected and design considerations minimized the copper losses. Reliability considerations have resulted in utilizing semiconductors at one-half their rated voltage and current ratings. Passive components have been used at one-half their power and voltage ratings. Consideration was also given to the frequency of operation. The prime consideration was to minimize package weight, size and power losses. The selection of operating-frequency was based on transistor losses and transformer losses and weight.

A specific feature of the power supply is that it utilizes one converter and one output transformer with two secondary windings to provide the two high voltage outputs. In the past, separate converter circuits were used to produce the two high voltages. In so doing, two separate modules were required which increased system weight and reduced overall efficiency. In the approach to be used, both output voltages are provided from one converter circuit, thereby maximizing efficiency and minimizing size and weight.

2.2.9.2 Design Considerations

The input and output requirements for the converter are as follows:

A. Input:

Voltage: 56 VDC Nominal

84 VDC Maximum

B. Output:

1) Voltages: +2.00 KVDC @ 408 ma.

-600 VDC @ 20 ma.

2) Total Power: 816 Watts

3) Efficiency: 90%

4) Size: Minimum

5) Weight: Minimum

C. Other Requirements:

1) Provide overload protection for both voltages.

2) Provide both output voltages from one converter (one transformer).

3) Provide telemetry outputs.

2.2.9.3 Parts Selection and Evaluation

The criteria used in the selection of component parts was reliability. The final unit can only be as reliable as its component parts.

All semiconductors, except high voltage rectifiers, will be operated at no greater than one-half their voltage and no greater than one-half their current ratings for their operating temperature. Semiconductors will be operated at estimated function temperature of at least 25°C below their rated temperature.

High voltage rectifiers will be rated at peak inverse voltages of at least two and one-half times their use rating and for current at least two and one-half times their use rating. In the event of overloads, the rectifiers will not be stressed above one-half their peak voltage and current ratings for the overload conditions.

Other components will be operated at no greater than one-half their voltage or current rating.

Wherever possible, component parts that are listed on the JPL approved parts list have been used.

Because of the nature of the design, specific requirements have precluded the use of JPL approved parts in some cases. The output switching transistors, for example, have not been approved by JPL. The Delco 2N2583 transistors were absolutely necessary to complete the design of the power supply.

2.2.9.4 Circuit Selection and Evaluation

The philosophy employed in considering design approaches for the high voltage supply stipulated that the basic circuit be reliable, highly efficient, have minimum size and weight. Since the object of the high voltage supply is to convert a low D.C. voltage to a high D.C. voltage, the simplest design would be as shown in the block diagram of Fig. 17. The diagram shows the low D.C. voltage coupled to a switching system. The switch shown serves to chop the input D.C. power to a square wave-alternating waveform. The square wave voltage is then stepped up to a high voltage by means of a transformer. On the secondary of the transformer, the alternating voltage is rectified and then filtered to provide the required high D.C. voltage. This method of providing a high D.C. voltage can be made very reliable, highly efficient and can result in a small, lightweight package if the proper components are used; in particular, highly efficient switches, transformers and rectifiers are required for the system.

Recognizing that a D.C. to D.C. converter system was required, several design approaches were considered. Because of the power level (800 watts), only driven converters were given consideration.

The input conditions for the D.C. to D.C. converter are as follows:

E_{in} = 56 VDC nominal and 84 VDC max.

I_{in} = 15 amps max.

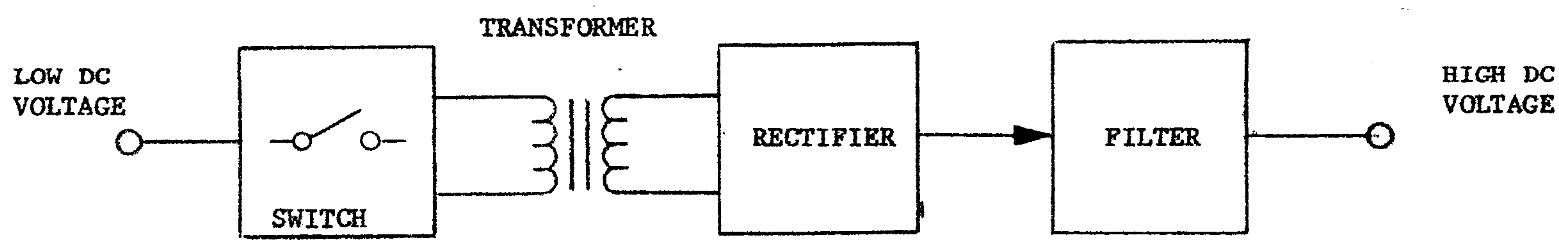


FIGURE 17. BLOCK DIAGRAM OF A BASIC DC-TO-DC CONVERTER SYSTEM

The push-pull converter was first considered. A block diagram of the circuit is shown in Fig. 18. This circuit consists of an oscillator circuit which serves as the time base for the system. The oscillator output then feeds two out of phase signals to the drive circuits shown. The driver circuits then power the output power amplifier stages. It is in the power amplifier stages that the input power is switched. The voltage step-up transformation is accomplished by the transformer. The rectifier circuit shown serves to rectify the output of the transformer and the necessary filtering of ripple is accomplished by the output filter.

The push-pull approach has several advantages:

A. It is an uncomplicated means of converting D.C. voltages.

B. It is a highly efficient method when the proper transistors and transformers are employed.

C. It is a highly reliable conversion method when high reliability transistors are used within their rated specifications.

D. It is a proven conversion method. Many well designed systems have been used successfully in space applications.

One disadvantage of the push-pull approach is that the transistors must be able to withstand twice the supply voltage. For this application, transistors must have collector to emitter breakdown voltage greater than (2 x 84) 168 volts. Allowing for 50% of their rated voltage, the transistor required must have a break-down of 336 volts. Most power transistors on the market today have collector to emitter breakdown voltages considerably less than 336 volts.

The bridge converter approach was also considered for this application. A block diagram of the circuit is shown in Fig. 19. The oscillator serves as the converter time base. The out-of-phase signals from the oscillator serve as input signals to the driver circuits. The drivers, in turn, provide input power to the power amplifiers. The power amplifiers are connected in a bridge configuration. The output transformer steps up the voltage from the power amplifiers to the

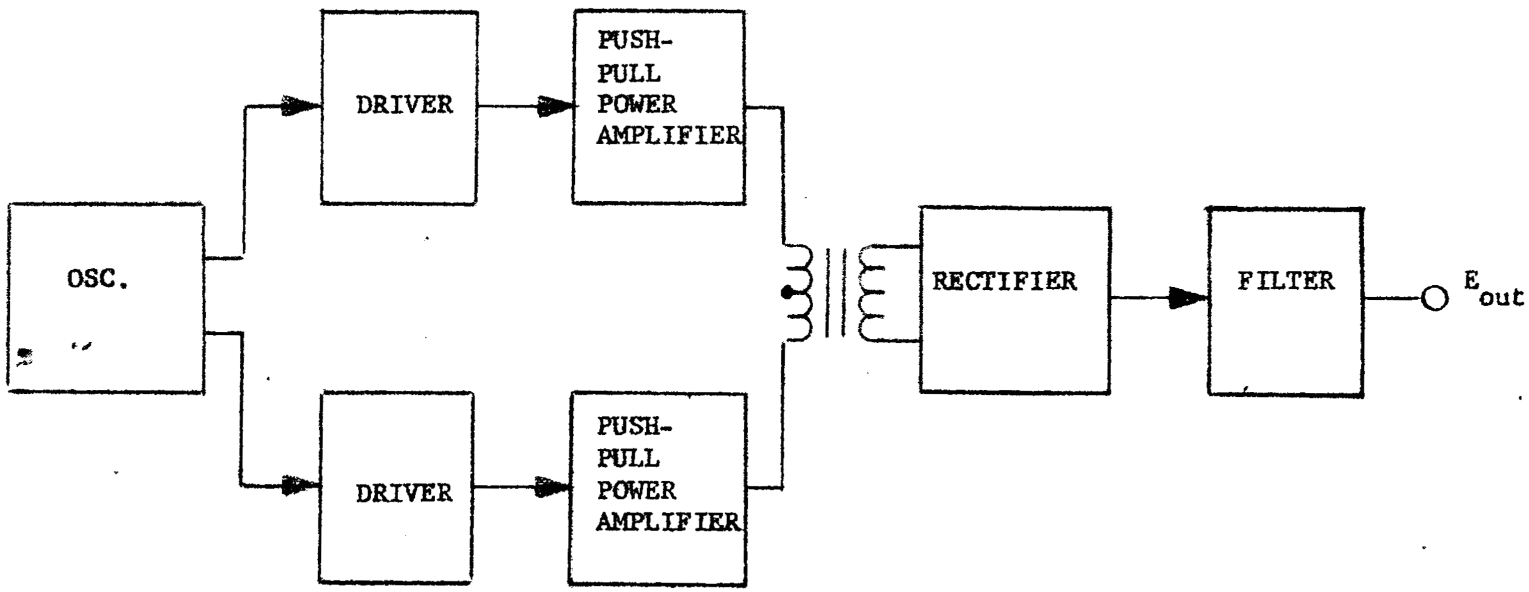


FIGURE 18. BLOCK DIAGRAM OF A PUSH-PULL DC-DC CONVERTER

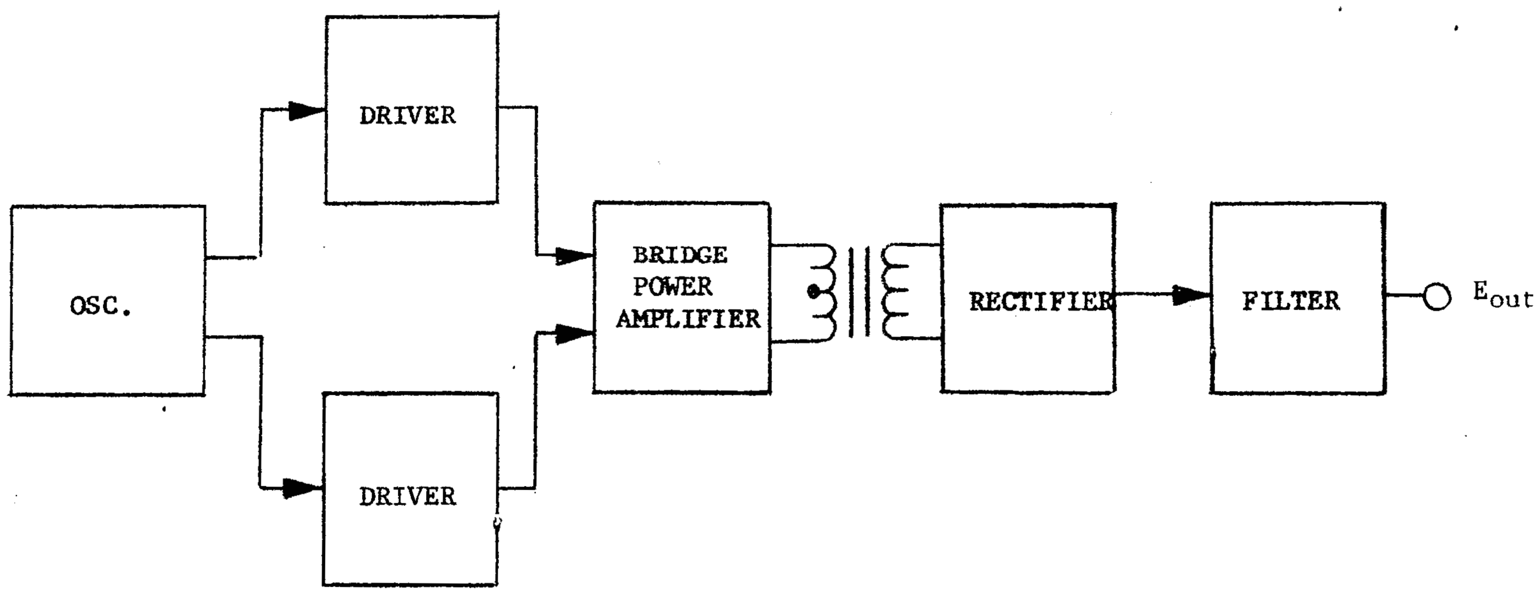


FIGURE 19. BLOCK DIAGRAM OF A BRIDGE DC-DC CONVERTER

rectifier circuit where the signal is rectified. Output ripple is reduced by means of the filter.

The bridge converter approach has one distinct advantage: the main advantage of this circuit is that each transistor in the circuit is subjected to only the supply voltage during cutoff. Thus, the bridge circuit will operate safely with twice the number of transistors than a push-pull circuit does for the same power and voltage ratings.

The disadvantages of the bridge circuit are:

A. It requires twice the number of transistors; therefore, the transistor losses are greater.

B. The overall efficiency is less than that of a push-pull circuit.

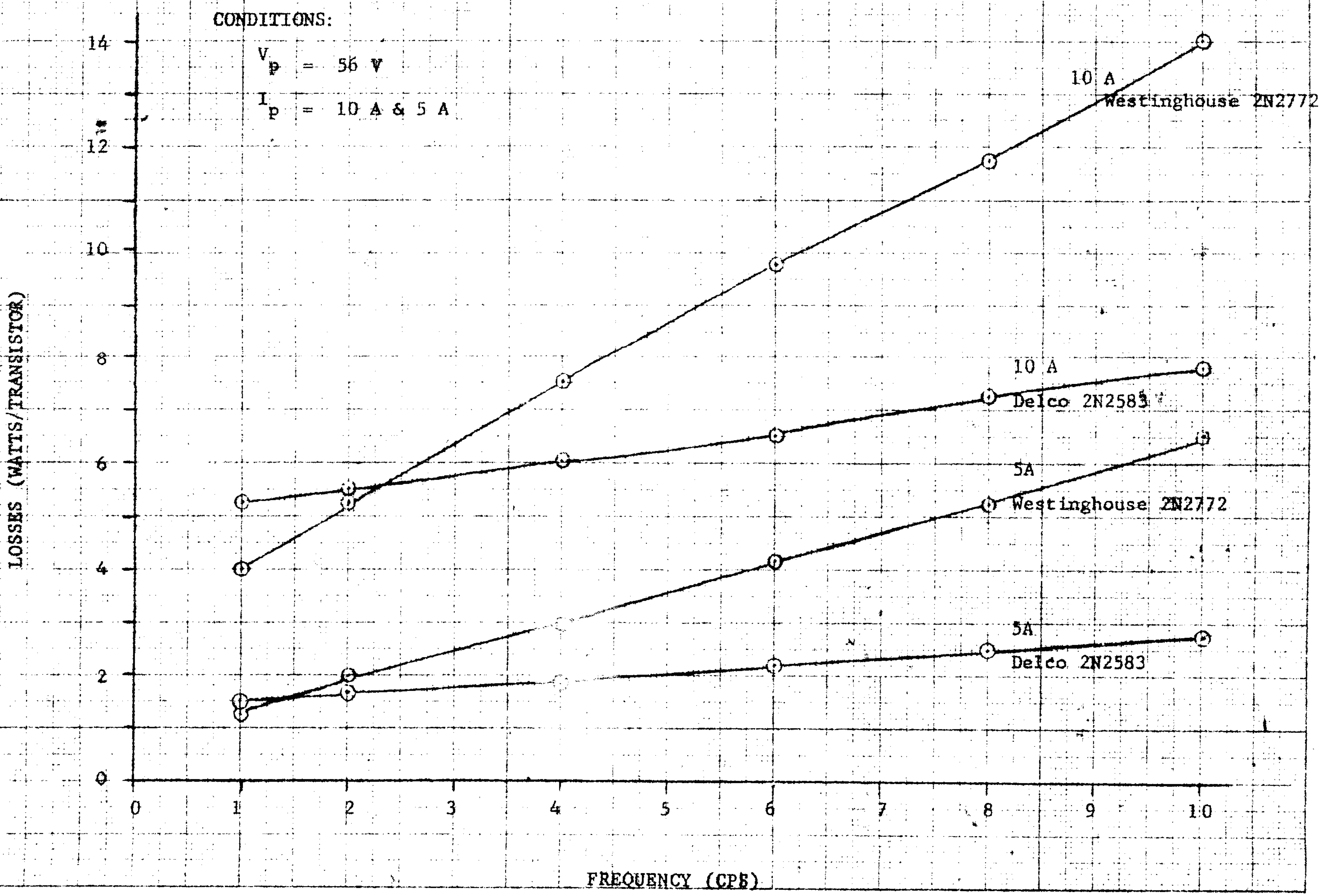
C. The circuit would be bulkier and heavier.

Before selecting the design approach, consideration was given to the power transistors. The required transistors had to switch 15 amperes of current. Utilizing the 50% safety factor, the current rating of the transistor would be 30 amperes. In considering the voltage requirements, the push-pull inverter required a 336 volt transistor, whereas the bridge circuit requirement was only 168 volts., but the transistor losses are doubled.

To provide the margin of safety required for voltage and current, two transistors were considered. One was the Westinghouse 2N2772 and the other was the Delco 2N2583.

A comparison of losses was made as a function of frequency for current levels of 10 and 5 amps. The results are shown in Fig. 20. It can be seen that the Delco transistor is the more efficient unit at both current levels. An added advantage is the fact that the Delco unit can be obtained with breakdown voltages of 400 and 500 volts; making it possible to employ the transistors in a push-pull circuit. On the other hand, the Westinghouse transistors are rated at 250 volts maximum, requiring them to be used in the less efficient bridge circuit. The only advantage of the Westinghouse transistor in this application was the 30 ampere current rating, which would not have

FIGURE 20. TRANSISTOR LOSSES VS. FREQUENCY



required paralleling of transistors. The Delco transistor, on the other hand, as rated for 10 amperes maximum so that three transistors would be paralleled for each switch of a push-pull circuit.

As a result of the comparison made, the Delco 2N2583 transistor was selected for the power transistors of the high voltage power supply. Paralleled units will be necessary, but they provide a considerable margin of safety with the high voltage breakdown rating. The prime reason for their selection, however, was because of their high efficiency as power switches.

Having made the selection of power switching transistors, the next task was that of selecting the most reliable and efficient method of converting the low D.C. voltage to the necessary high voltages. Since the push-pull method is the more efficient approach, it was selected as the best design approach. The power transistors selected are also compatible with the design approach chosen. The combination of power transistors and design approach selected will result in a highly efficient and reliable power supply.

2.2.9.5 Circuit Description

Figure 21. shows a block diagram of the high voltage power supply. The diagram shows all of the individual stages. A discussion of the various stages is contained in the following paragraphs. Reference to the schematic diagram, Fig. 22., will also aid in understanding the circuit. The 10 kc clock shown in the block diagram is used as the system time base. The circuit employed is an astable multivibrator. The output from this circuit is a 10 kc square wave.

The output from the multivibrator circuit is used to drive a binary circuit where the input signal is divided by two. The circuit, composed of Q701 and Q702, utilizes positive feedback in such a manner that the two transistors tend toward opposite states, one off and one on. Base triggering is used for the circuit.

The 5 kc output waveforms from the binary collector circuits are coupled to two emitter follower circuits composed

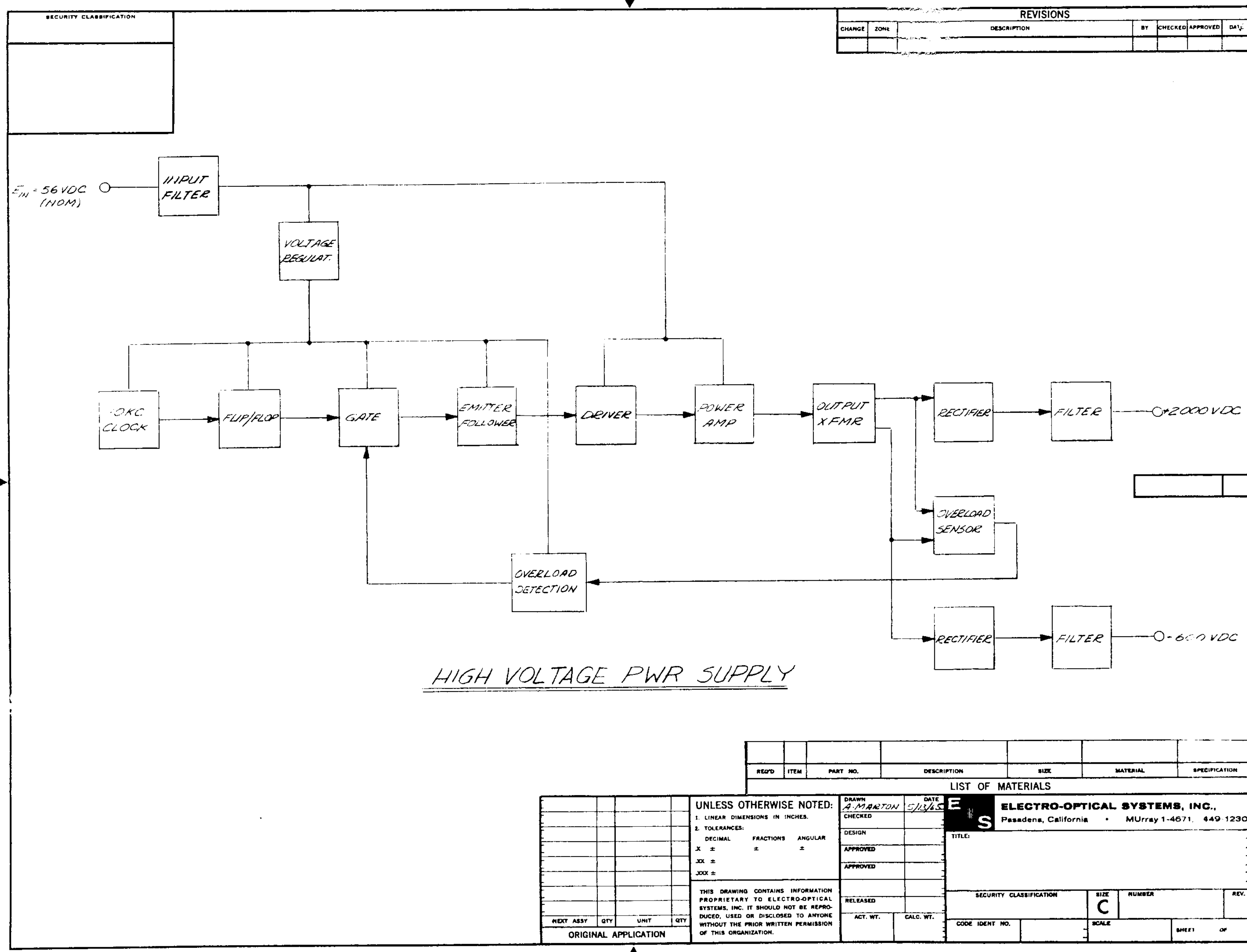


FIG. 21 BLOCK DIAGRAM HIGH VOLTAGE SUPPLY

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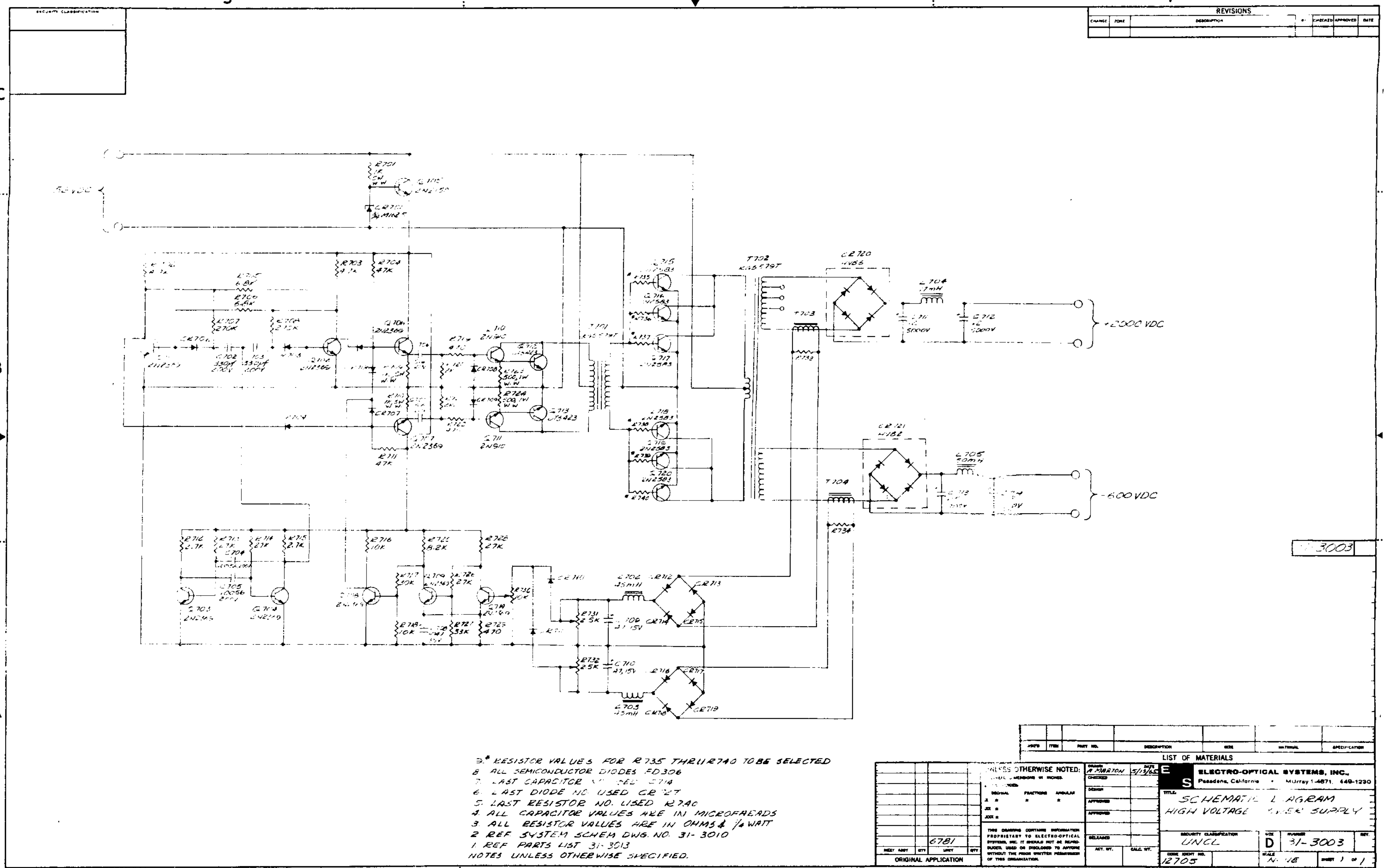

FIG. 22 SCHEMATIC HIGH VOLTAGE SUPPLY

of Q708 and Q707. These circuits act as impedance matching circuits. They present a high impedance to collectors of the binary circuit and have a low output impedance at their emitter circuits.

The out-of-phase output signals from the emitter follower circuits are then used as input signals to the push-pull driver circuit, composed of Q710, Q711, Q712 and Q713. Emitter follower circuits are used to power the driver output transistors, Q712 and Q713. The transistors used in the driver circuit are used as saturated switches. The square wave output from the driver circuit is then coupled to the base circuit of the push-pull power amplifier consisting of Q715, Q716, Q717, Q718, Q719 and Q720. The transformer secondary circuit delivers the required voltage and current to saturate the output power transistors.

The power amplifier stage uses a trifilar wound transformer (T702). The transformer has three parallel primary windings. The windings are commonly terminated at the collectors of three parallel power transistors. In effect, three transistors are paralleled for each side of the push-pull circuit. Each transistor will ultimately carry one-third of the total collector current when properly balanced.

There are certain advantages to using a trifilar transformer. The main advantage of the trifilar scheme is that it permits using smaller wire for each winding. This insures tighter coupling to the core, which results in a smaller transformer. To optimize the output circuit power, transistors will be matched as closely as possible for parameters such as turn-on time (t_{on}), turn-off time (t_{off}), collector to emitter saturation voltage (V_{ce}sat), base-emitter saturation voltage (V_{be}sat), and current gain (h_{fe}). Matching of these parameters will insure a close approximation to a single switch by three parallel transistors. In addition, balancing resistors will be used in the base circuits of the output transistors to compensate for any variations in transistor gain characteristics. This will ensure current division through the power transistors.

The output transformer (T702) has two secondary windings. The high voltage winding delivers 2000 volts and the other winding delivers 600 volts. The design of the transformer will insure that no breakdown will occur between either secondary windings or secondary to primary, or secondary to core. A 10,000 volt insulation is provided.

The first secondary winding is rectified by a full wave bridge consisting of diodes CR720, CR721, CR722 and CR723. The ripple of the output voltage is filtered by the pi-type network consisting of C711, C712 and L704. The inductor used in the output filter serves another purpose in the circuit. In case of a short on the output of the power supply, it would serve as a current limiter in the circuit. The fault current will not rise instantaneously, but as determined by the time constant of the inductor and the output resistor, this time constant would then give the overload protection circuit time to sense the overload and turn off the supply,

The second secondary winding delivers 600 volts. This voltage is also rectified by a full wave bridge circuit, consisting of CR724, CR725, CR726 and CR727. The output ripple is minimized by the output filter consisting of capacitors C713, C714 and inductor L705. As in the case of the positive high voltage output, the inductor used in the filter network also serves to limit the fault currents in case of a heavy overload on the supply. The load current will not rise instantaneously, as determined by the inductor L705 and the output resistance on the output.

Overload currents from both output windings are sensed by means of current transformers T703 and T704. Current transformer T703 senses the current through the 2000 volt winding and T704 senses the current through the 600 volt winding. The outputs from the current transformers are rectified by full wave rectifier bridges composed of CR712, CR713, CR714 and CR715 for the 2000 volt output and CR716, CR717, CR718 and CR719 for the 600 volt output. Averaging filters composed of C709 and L702 and C710 and L703 are used to filter the output from the bridge rectifiers. Potentiometers R731 and R732 serve to adjust for any percentage of overload current required. The voltage at the potentiometers will be a linear function of the currents through the secondaries of the output transformer.

The outputs from both overload sensing circuits are then gated by an OR circuit. If either the 2000 volt circuit of the 600 volt circuit are overloaded, an output will appear across resistor R730. This will enable turning off the power supply when either circuit is overloaded.

The output from the overload sensing circuits is then used to fire the Schmitt trigger circuit composed of Q709 and Q714. Adjustment of the level of input voltage to the trigger circuit is provided by potentiometer R730. The output from the Schmitt trigger is put through saturated switch, Q708. When the switch is open, the collector voltage back biases two diodes of an AND circuit. In this condition, the signal from the binary circuit is gated through the AND circuit to the rest of the system to provide the proper D.C. voltages at the output. When the Schmitt trigger circuit fires, the switch is saturated or closed so that both diodes in the AND circuits that were previously back biased become forward biased and subsequently the output from the binary circuit is not gated or allowed to pass to the rest of the system. As a result, the output voltages from the power supply are reduced to zero. Under normal operating conditions, the gating circuit permits a driving signal to be removed instantaneously from the output transistors. This then prevents causing damage to the power switches of the power supply.

Power to the low level circuitry is provided by a one transistor series regulator circuit, composed of Q705, CR701 and R701. This circuit provides a regulated 16 VDC.

Telemetry outputs from the power supply will also be provided. High voltage current information will be obtained from the output of the filter network, C709 and L702. Low voltage current information will be obtained from the output of the filter network, composed of C710 and L703. A zero to 5 VDC signal will be provided from both circuits.

The output voltages from the power supply will be monitored by means of the step-down transformers shown across the secondary windings of the output transformers. These transformers will be provided with the proper insulation to prevent voltage breakdown between primary to secondary and primary to core. The secondary circuits of the transformers will be rectified and filtered. The monitoring circuits will furnish 0 to 5 VDC outputs to correspond to the 0 to 2000 VDC and 0 to 600 VDC high voltage outputs, respectively.

2.2.9.6 Circuit Parameters

Compoent Power Losses:	Transistors	16.0 Watts
	Transformers	31.0
	Diodes	10.5
	Filter	2.0
	Driver	1.0
	Ckt. Losses	4.0
		64.5 Watts

Component Weight:	Transistors	280 gms.
	Capacitors	264 gms.
	Current Transformers	20 gms.
	Inductors	50 gms.
	Bridge	150 gms.
	Output Transformers	876 gms.
	Low Power Circuits	10 gms.
		1650 gms. = 58 oz. = 3.63 lbs.

Expected Circuit Efficiency:

819.0 Output

64.5 Losses

883.5 Watts Input

$= \frac{819.0}{883.5}$ 92.7%

2.3 Packaging

The electronics system is to be operated in the form of a breadboard in a vacuum tank. Two things are required of the mechanical design: adequate cooling and ease of testing and servicing. To accomplish this, the package configuration is as shown in Fig. 23. It consists of an open fram holding a stack of drawers and shelves. The shelves are fixed and are cooled by a flow of liquid nitrogen through them. The surfaces of the shelves are coated with an acrylic enamel to improve their radiation absorptivity in the infrared region. The drawers consist of single thin plates mounted with quickly detachable fasteners to conventional slide mechanisms. Connectors are used on each drawer. The primary interconnecting cable harness is mounted to frame and the service loops are kept in order by cable retractors opposite each drawer. This permits a compact package; yet each drawer, which contains one or more space complete circuit modules, may be pulled out and serviced without disoonnecting it electrically. Packging density is such that the system will function in air without the aid of LN_2 cooling.

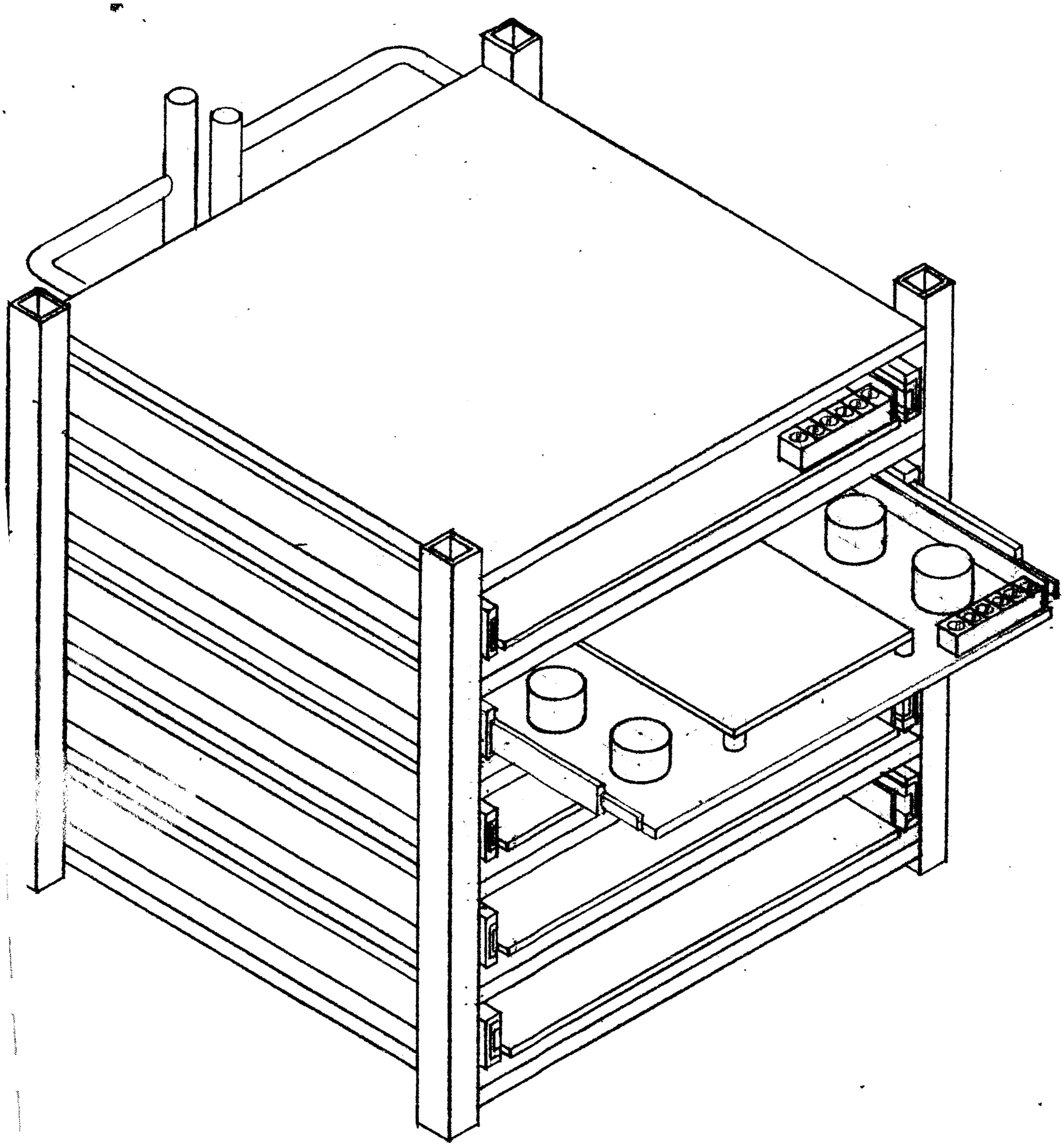

FIG. 23 PACKAGE CONFIGURATION

3. SUMMARY

3.1 System Block Diagram

The overall system is graphically illustrated by the functional block diagram, Fig. 24. Included on this drawing are the commands required, and the time at which the switching sequence takes place for the laboratory endurance test configuration.

3.2 System Parameters

3.2.1 Test System

The following chart illustrates weights, losses, input, output and efficiencies of the individual units which comprise the power conditioning and control system for the laboratory endurance test configuration.

Unit	Weight	Input	Losses	Output	Efficiency
Low Voltage Inverter	2.44	201	17.8	183.24*	91.2
Reservoir Valve Heater	1.31	225*	11.9*	214	95.1
Magnet Supply	0.297	12.4	2.4	10	80.6
Arc/Cathode Htr. Supply	1.23	157.0	19.0	138	88
Vaporizer Supply	.41	12.4	2.4	10	80.3
Neutralizer Supply	.28	1.44	.1	1.34	94
High Voltage Supply	3.63	883.5	64.5	819	92.7
TOTALS:	9.6	1084.5	106.2	978.34	90.2

Specific Weight = 9.6/1084.5 = 8.85 lbs/kw

The totals shown represent the power required by the engine during normal operation. The individual power figures marked with an asterisk (*) are not included in the totals, because they do not contribute to engine output during normal operation.

3.2.2 Flight System

The flight configuration of the power conditioning and

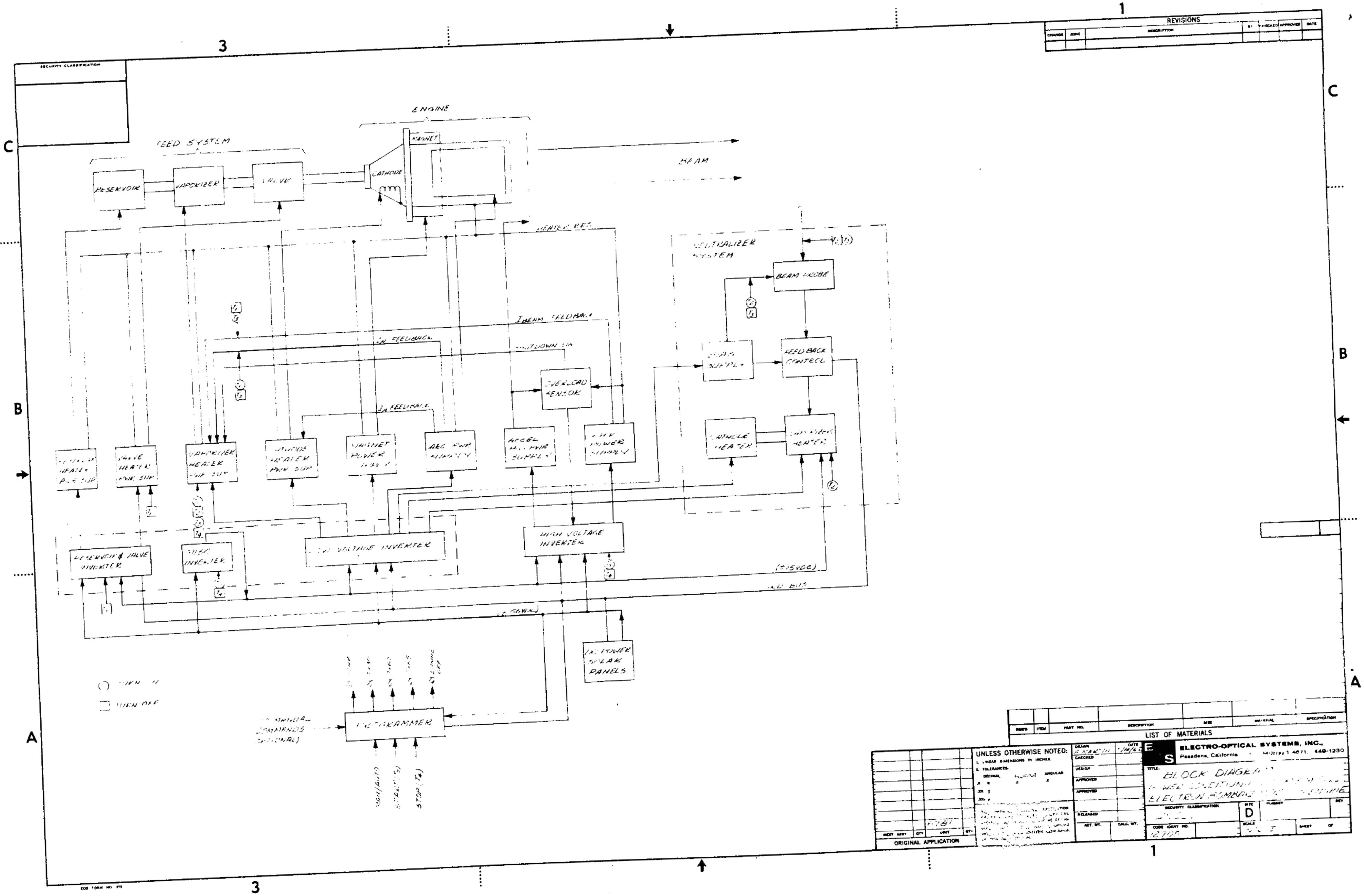

FIG. 2 SYSTEM BLOCK DIAGRAM

control system differs from the above described laboratory test configuration in the following ways:

A. A neutralizer for each individual engine is not required for a multiple engine concept. Sufficient neutralization can be obtained from one neutralizer for several engines. In this analysis each neutralizer is conservatively assigned to three engines. This eliminates 0.2 lbs. from the test system weight and a loss of 1.0 watts.

B. A reservoir heater is required for a test system, to insure a liquid state of the cesium prior to engine turn on. In the flight system this will be achieved in other manners, since the reservoir is surrounded by the electronics, only slight compartment heating is required to maintain the cesium at liquid temperature. Therefore, a reservoir heater is not required. The valve heater output is then added to the low voltage inverter transformer. This results in a weight reduction of 1.1 lbs.

C. In the test system, telemetry outputs are furnished for all engine functions. The majority of these outputs are required for analysis at engine performance and for diagnostic purposes in the event of a malfunction. For a flight system only pertinent information, such as beam current, plus high voltage, minus high voltage and arc current are required. The elimination of all telemetry functions, except those required for the flight system, will result in a weight reduction of 0.25 pounds minimum.

D. Because of the output characteristics of the solar panel, consideration must be given to the inclusion of an energy storage system to provide the transient energy required for the switching inverters. A filter containing inductance and capacitance in an L section filter is one method of providing this energy. The component weight of a filter section, capable of storing sufficient energy is a maximum 1.8 lbs. This weight can be modified by dividing the loads into three engine groups and separating the load demand of each engine system in that group by 120^{o} increments over the entire switching cycle. Then, the actual transient demand at any one time is one-third the total transient power required by the propulsion system. A minimum reduction of 0.8 lbs. in filter weight can be realized. The final filter weight for each power conditioning and control system is 1.0 lb.

E. Printed circuit boards, weighing 0.25 lbs., were included as part of the low voltage inverter. This must be considered part of packaging and not as electronic component weight.

F. A further design iteration of the arc cathode heater power supply has resulted in a weight reduction of 0.15 lbs. in magnetic elements.

G. Further design iterations of all magnetic components with examination of derating factors employed, core areas, copper areas, losses and reliability will result in additional weight advantages. A minimum of 0.5 lbs. reduction can reasonably be expected.

The following chart illustrates the weight, input power, losses, output power, efficiencies and weight advantage of each individual power supply which comprise the flight power conditioning and control system:

Unit	Weight	Input	Losses	Output	Effic.	Advantage
Low Voltage Inverter	2.19	201	17.8	183.24	91.2	.25
Reservoir Valve Htr.	0.21	56.8*	2.8*	54*	95.1	1.10
Magnet Supply	0.30	12.4	2.4	10.0	80.6	---
Arc/Cathode Supply	1.08	157.4	19.4	138.0	87.6	0.15
Vaporizer Supply	.41	12.4	2.4	10	80.3	---
Neutralizer	.08	.44	.03	.41	94.0	0.20

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Unit	Weight	Input	Losses	Output	Effic.	Advantage
High Voltage Supply	4.63	885.5	66.5	819.0	92.5	-1.00
TOTALS:	8.90	1085.5	107.6	977.4	90	0.70
Telemetry Weight	- .25					
Magnet Redesign	- .50					
TOTAL WEIGHT:	8.15					

Specific Weight: 8.15/1085.5 = 7.52 lbs./kw

The preceeding statistics relating to weight refer to electronic components only. Assuming a packaging factor of 50%, the overall specific weight for the power conditioning and control system for a flight system is 15 lbs./kw at an efficiency of 90%.