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THE PERKIN-ELMER CORPORATION
AEROSPACE DIVISION
2855 Metropolitan Place, Pomona, California 91767

FINAL REPORT
ION PUMP POWER SUPPLY
VOLUME 3 OF 6
FOR
COMBINED STUDY PROGRAM

By Michael F. Hagen

March 1971

Perkin-Elmer SPO 30006
NASA Contract Number NAS1-9469

Prepared for
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
LANGLEY RESEARCH CENTER
Langley Station
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ABSTRACT

A program was conducted in which an instrument system concept was studied to optimize the application of a mass spectrometer as a sensor for monitoring the primary atmospheric constituents, as well as atmospheric contaminants, on board a manned spacecraft. The program was divided into six individual studies representing the primary system parts complementing the spectrometer: A Carbon Monoxide Accumulator Cell (Volume 1), an Ion Pump (Volume 2), an Ion Pump Power Supply (Volume 3), an Inlet Leak (Volume 4), an Ion Source (Volume 5), and an Undersea Atmospheric Analyzer (Volume 6). The principle goal of the combined study program was the achievement of an instrument concept of minimum power, weight and size without compromising the minimum detection limits of the instrument.

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SUMMARY

The high voltage Ion Pump Power Supply for the Combined Studies Program was designed to meet the contract requirements of NASA Langley, Contract NAS1-9469, to operate a four liter per second ion pump efficiently and reliably. Specifically, the ion pump power supply has demonstrated compliance to the specification requirements for a minimum power (under 30 watts) breadboard power supply capable of reliably starting the ion pump at pressures of 1×10^{-3} torr. Breadboard testing has shown that the maximum measured input power to the Ion Pump Power Supply is approximately twenty watts, which met and exceeded the design goal of thirty watts maximum. Additional testing with a test load simulating a four liter per second ion pump at a pressure of 1×10^{-3} torr demonstrated the ability of the power supply to provide sufficient output power capability to reliably start the pump. An added feature of the power supply is its ability to withstand momentary short circuits on its output without damage.

The high voltage power supply described in this report has therefore satisfactorily met the NASA Langley Research Center requirements under Contract NAS1-9469.

INTRODUCTION

A combined study program was initiated under Contract NAS1-9469 with NASA Langley Research Center to develop an inlet leak and ion pump assemblies for a Magnetic Mass Spectrometer System. This system was to be used, ultimately, in monitoring and controlling a multi-gas atmosphere in a spacecraft cabin.

Ion pumps are the most practical method of producing and maintaining high vacuums for mass spectrometers intended for space flight use. This design report therefore, describes the development and testing of a high voltage power supply necessary to operate the four liter per second ion pump used in the Magnetic Mass Spectrometer System for NASA Langley Research Center, Contract NAS1-9469,

Since the end use of a mass spectrometer system is for spaceflight use, an efficient yet practical high voltage power supply was required. The contract design goal was twofold: First, to maintain the total input power to the Ion Pump Power Supply at less than thirty watts; and second, to be capable of reliably starting the ion pump at pressures as high as 1×10^{-3} torr. These design goals were to be verified by manufacturing and testing a breadboard Ion Pump Power Supply.

DESIGN DESCRIPTION

The Ion Pump Power Supply for the Combined Studies Program under NASA Langley Contract NAS1-9469 has been designed to operate a four liter per second ion pump efficiently and reliably. In general, the most difficult task of designing an Ion Pump Power Supply is that of providing the large current operating range needed at various pump pressures. To ensure that the pump does not dissipate excessive power, voltage regulation is required as a function of pump pressure. As an example, a typical change of pump pressure would be from 1×10^{-3} torr (starting pressure) to 1×10^{-6} torr (running pressure). At a constant voltage, this differential in pump pressure represents approximately the same change in pump current, or a factor of 1000 (1000 times). When pump pressure is low (representing running pressure) keeping the voltage high is advantageous, and when pump pressure is high (representing starting pressure) keeping the voltage low is advantageous. Pump performance is thereby optimized. Therefore, some method of voltage control is obviously necessary. Additionally, the voltage-current relationship is nonresistive; and therefore a possible current increase of ten could be required with a voltage increase of only three. This effect is due to momentary high pressure (outgassing) in the pump, which causes a nonlinear change of the voltage level as a function of pump current.

Several possible design approaches were examined to resolve these problems; however, none singularly provided the required performance. A decision was made based on the importance of the tradeoffs, and a power supply was developed to meet the majority of the system performance goals. Three of the design approaches investigated are summarized in the following paragraphs. The dual mode supply, selected as the best compromise to meet the performance goals, will then be described in detail.

Series Impedance Approach

The simplest method to control the output voltage is a supply with a series impedance. This can be achieved using resistance or current limiting circuitry depending on the output characteristics desired. Although this method has worked very well in other applications, the requirements for the combined studies program specify a starting pressure of 1×10^{-3} torr and an operating pressure of 1×10^{-6}. In order to provide 850 volts for starting at this high pressure and five kilovolts for normal operation, excess input power would be required and a highly inefficient system would result.

Duty Cycle Control Approach

Another approach is that of utilizing a series switching regulator to duty-cycle control the input voltage with feedback from the ion pump current. With this method, practically any required output characteristic curve can be implemented. The major difficulty, however, is the design of an efficient converter and rectifier circuit to accommodate both heavy currents at low voltages and light currents at high voltages. If a standard full-wave bridge rectifier circuit is utilized, a very high step-up ratio (20 V to 5 kV) would be required in the power transformer.

Two particular problems are encountered in this type of high voltage DC/DC converter transformer. These are: First, excessive interwinding capacitance; and second, high-voltage insulation. The interwinding capacitance causes heavy switching losses in the converter transistors. The need for high voltage insulation requires larger physical size, resulting in increased core and dielectric losses.

Switching Preregulator Approach

If a single high voltage multiplier were used with a switching preregulator, other factors appear. Voltage multipliers perform well under light loading, but heavier loading requires voltage multipliers with proportionately more capacitance. This capacitance reflects to the primary of the converter as the square of the turns ratio, resulting in heavy switching losses similar to those expected from excessive interwinding capacitance in the Duty Cycle Control Approach.

Dual Mode Supply Approach

The Ion Pump Power Supply design for the Combined Studies Program has taken a dual mode optimum approach. Two supplies have been designed; one for high-voltage/low-current and one for low-voltage/high-current. This approach provides good efficiency upon starting and running. After an initial high-presuure start at a low-voltage (850 V), the pressure drops because of pumping. When the pump pressure reaches the maximum expected ion pump operating pressure, the supply switches to the high voltage mode (5 kV). This insures maximum pumping speed during operation and optimizes the pump efficiency. Switchover hysteresis as shown in Figure 1, is designed into the logic circuitry to allow for greater current on the 850 V supply output before switching to the high voltage supply. Additionally, a time delay function is incorporated to prevent the supply from switching to the low-voltage mode due to current transients larger than the steady-state switchover current.

Although the dual mode ion pump power supply is more complex than a single supply, the tradeoff between high efficiency at heavy and light loading versus reliability, cost and size is justifiable. Also, a savings of as much as forty watts will result over a single mode supply utilizing any one of the previously described design approaches.

CONTRACT REQUIREMENTS

The contractual requirements for the Combined Studies Ion Pump Power Supply specify a minimum power (under 30 W) power supply capable of starting the ion pump at a pressure of 1 x 10^{-3} torr. The deliverable unit shall consist of a breadboard assembly of the Ion Pump Power Supply. Documentation shall be provided to completely specify the assembled equipment.

DUAL MODE POWER SUPPLY SPECIFICATIONS

Input Voltage: 20 Vdc $\pm$1%

Input Current: 0.88 A measured under short circuit conditions.

Output Voltage:

a. Low Voltage Mode: 850 V Nominal
750 V Min at 12 mA

b. High Voltage Mode: 5000 V Nominal
3900 V Min at 1.8 mA
From Worst Case Analysis (WCA)

Output Ripple Voltage:

a. Low Voltage Mode: 6 V p-p

b. High Voltage Mode: 190 V p-p

Dual Mode Switchover Points:

a. Nominal Lower Trip Point: 0.1 mA $\pm$5% after set

b. Nominal Upper Trip Point: 1 mA $\pm$5% after set

SYSTEM DESCRIPTION - DUAL-MODE POWER SUPPLY

The System Block Diagram for the Ion Pump Power Supply is shown in Figure 2. The input power is filtered for electromagnetic interference (EMI) considerations. The small drive oscillator produces base drive for the power converters and supplies isolated dc voltages for operation of the current sensor and base drive switchover circuits. The input and output grounds are isolated by operating the current sensor on the secondary ground and switching base drive with a relay. The current sensor relay switches base drive between the low voltage (850 V) and high voltage (5 kV) converters, depending on ion pump current. Figure 1 shows the output voltage versus current. The difference in the switchover point, in switching from high to low and low to high voltage, is termed "hysteresis".

The hysteresis is provided to allow for the extra current requirement of the Ion Pump Power Supply when switched from the low to high voltage mode. The direction of the arrows in Figure 1 indicates that the response of the supply is due to a change in load when in either the high or low voltage mode. When in the low voltage mode, as in starting the ion pump, the current must decrease to one-tenth of a milliampere before switchover to the high voltage mode. Once in the high voltage mode, the supply will provide up to one milliampere before switching back to the low voltage mode. The switchover time is approximately one second; a time delay is introduced to prevent transients from switching modes unnecessarily. For the one second before switching, the high voltage supply is current limited to approximately two milliamperes. The low voltage supply is current limited to provide twelve milliamperes.

CIRCUIT DESCRIPTION - DUAL MODE POWER SUPPLY

The Ion Pump Power Supply consists of the following circuits; the drive oscillator, the current sensor and the power converters.

Drive Oscillator

The drive oscillator (Figure 3) supplies base drive for the DC/DC converters as well as dc voltages for the current sense circuitry. Using the small free-running oscillator (10 kHz) to drive the power converter transformers allows the converters to be operated in their linear region. This reduces current spiking and switching losses in the converter chopper transistors. Power line interference (EMI) and transformer core loss are both reduced using this configuration. The drive oscillator transformer is designed on a low-loss (square permeability 80) core. The chopper transistors (Q1 and Q2) are high frequency types, which in conjunction with the transformer (T1) operate on only 100 milliwatts of power.

When power is first applied, current flows through R3 and into the center tap of the feedback winding of T1. Q1 or Q2 will turn on, depending on which has the most gain. Due to the polarity of the windings on T1, the opposite transistor is turned off. After a period of one-half the oscillator's operating period the transformer core saturates, the magnetic flux collapses and all voltages fall toward zero. The ON transistor is turned off and the OFF transistor, no longer being reversed biased, becomes susceptible to turn on. The voltage overshoot caused by core saturation is enough to turn on the previously OFF transistor. The opposite transistor becomes reverse biased and remains off until core saturation, thus completing one cycle of operation.

Frequency of oscillation is mainly dependent on core flux capability, input voltage and primary turns by the following expression; frequency is affected only slightly by loading converter transistors and core squareness ratio:

$$f = \frac{E \times 10^8}{\phi_T N_P}$$

where, E = Applied rms input voltage

ϕ_T = Total flux capability

N_P = Number of turns on driven winding

Resistors R1 and R2 are chosen to provide sufficient base drive to saturate Q1 and Q2 under full load. Diodes CR1 and CR2 alternately provide a path to ground for base drive. R3 is a starting resistor, chosen to provide enough current to ensure oscillation. On the secondary of T1 are two windings, one supplies base drive for the main power converters and the other winding is rectified and filtered to provide plus or minus seven and one-half volts for the current sensor. This winding is insulated for high voltage, since it is common to the secondary high voltage ground. This maintains primary-to-secondary ground isolation.

Ion Pump Current Sensor

The ion pump current is sensed to provide the mode switchover command to the power converters. (See Figure 4.) The voltage-current relationship for the sensor has been explained in the system description and was shown in Figure 2.

Ion pump current is sensed in the ground side of the high voltage supply by R35. Since R35 is a one kilohm resistor, a one volt per milliampere sense voltage to load current relationship is established. The sense amplifier as a National LM301A operated on plus or minus seven and one-half volt supplies. Capacitors C26 and C27 bypass transients around R35. The diodes (CR31 to CR33) protect the input of the LM301A from overvoltage caused by high load currents. Zener diode CR34 provides supply line transient protection for AR1. Resistors R26 and R29 set the reference voltage on Pin 2 of AR1 to either plus one-tenth or minus one-tenth volt, depending on the state of the output transistor Q4. The sense amplifier changes to the opposite state when the load current is 100 microamperes or one milliampere, corresponding to the plus one-tenth or minus one-tenth volt reference on the inverting input of AR1. R30 maintains the source resistance for the inverting input relatively low. Resistor R31 sets the source impedance for the noninverting input approximately equal to that of the inverting input. This is done in order to minimize the effects of drift in amplifier input bias currents. From the manufacturer's data on the LM301A, capacitor C4 rolls off the frequency response of AR1 to an open zero gain point of about 300 Hertz.

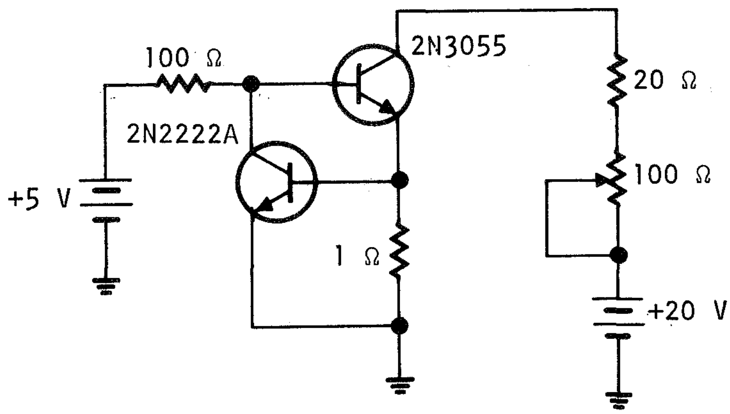
Capacitors C15 and C16, along with R28 provide positive feedback around the LM301A to enable the circuit to latch more positively, thus stabilizing the system during switching. Capacitor C2 and resistor R11 form a unidirectional time delay. When a heavy current transient occurs and the five kilovolt converter is operating, it is desirable to have the high voltage supply remain energized for about one second before switching to the 850 volt supply. This eliminates unnecessary switching back and forth on transient currents. The high voltage converter will current limit at two milliamperes during this period before switchover, so that no component overstressing will be encountered.

When the output of AR1 changes from +V saturation to -V saturation, as when a heavy load occurs, the capacitor C13 changes from +V saturation to $- V_{BE}$ Q6 (ON) $+V_{CR17}$(ON). The RC product of R23 and C13 thus determines the time delay for switchover from the high to low voltage mode. When switching the other direction, the time delay is not so prevalent. This is because the voltage on C13 only has to change about six-tenths of a volt to turn Q6 from on to off. Transistors Q3, Q4 and Q6 provide the small base current needed by Q6, R23 can be made large to reduce the value of C13 needed for the one second time delay. Transistor Q5 switches the positive reference voltage from CR16 into the summing mode of AR1. Capacitor C12 bypasses the drive to Q5, eliminating relay chatter during switching. Diodes CR16 and CR18 are used strictly as low current reference voltages. They eliminate the need for regulated voltage for the high current relay K1. R6 keeps K1 (12 V relay) from drawing excessive current when operated between plus and minus seven and one-half volts. Diodes CR7 and CR8 along with C6, act as a transient suppressor across K1. CR9 and CR10 provide protection from false turn on of Q3 by the V_{CE}(sat) of Q4. CR11 allows the collector of Q4 to rise when Q3 is on. R7 and R11 provide a path for collector-to-base leakage current, and are chosen sufficiently low to prevent forward biasing of their respective transistor.

High Voltage and Low Voltage Converters

The high voltage converter (see Figure 5) delivers five kilovolts at currents up to two milliamperes. The dc input voltage (20V) is converted to an 850 volt (rms) squarewave by the chopper transistors (Q13 and Q14) and the high voltage transformer (T3). The transformer output is multiplied to five kilovolts dc by the X6 rectifier circuit. The output current is limited by the one ohm resistors (R19, R20) and the transistors (Q11, Q12) connected in parallel with the bases of the chopper transistors. As chopper collector current increases, the voltage drop across the one ohm resistors turn on the current limiting transistors. Base drive for the choppers is diverted to ground and Q13 and Q14 are biased into the active region. This produces the sharp current limiting point, only a slight slope in the voltage vs current graph can be seen. This small change in voltage (< 5%) as current increases is due to the one ohm sense resistors. The capacitors across the sense resistors improve the squareness of the collector waveforms on the chopper transistors.

Resistors R17 and R18 were later added in series with the bases of the current limit transistors to limit base drive. This was necessary because of the peak charging current required by the X6 voltage multiplier. As a result, the current limiting is not so pronounced as with the test setup used to establish the current limit test circuit below (see Figure 6). The resistor "softens" the knee of the turn on of the limiting transistors. Diodes CR14 and CR15 were also added to suppress negative transients caused by switching the transformer T3.



Current Limit Test Circuit (See Figure 6)

The low voltage converter consists of the same primary circuit and transformer as that of the high voltage converter. The secondary voltage is bridge rectified to provide 850 volts dc at a minimum of 12 milliamperes. The outputs of the two converters are paralleled together with isolation diodes in series with the 850 volt supply. When operating on the five kilovolt mode, these diodes isolate the 850 volt and five kilovolt circuitry. The negative outputs of both supplies are paralleled and connected to secondary ground through R32 and the current sense resistor.

Resistors R4 and R5 current limit the base drive to the converter transistors. Since, under maximum load (short circuit conditions), both the high and low voltage converters draw the same power, the base drive requirements are identical. Thus, the same base resistors may be used for both converters. Relay K1 (driven by the current sense circuit) is utilized to switch base drive to either DC/DC converter.

The high voltage transformer (used in both converters) is designed for low loss and high reliability. The core is a Magnetics Incorporated Square Permeability 80 type. The high voltage secondary winding is progressively bank wound to reduce winding to winding voltage. Before the primary is wound, the secondary is completely vacuum encapsulated and baked out. This ensures the proper insulation from primary to secondary and also keeps the primary from physically disturbing the secondary. The completed transformer is finally vacuum-dip finished to seal out moisture.

Worst Case Analysis

The schematics of Figures 3, 4, and 5 have been used in performing worst case analysis on each of the three circuits. The following worst case parameters have been used:

a.	Input voltage	20 V ±1%
b.	1% Resistors*	±2%
c.	5% Carbon Resistor*	±22.5%
d.	Ambient Temperature	25 - 40°C

*Derating applied to basic tolerance to account for worst case end of life parameter variation.

<u>Worst Case Analysis: Drive Oscillator (See Figure 3)</u>.- The Drive Oscillator must be checked for operation under maximum load and minimum input voltage and over the operating temperature range. Maximum and minimum drive voltages to the converter and sensor circuit must be established.

The worst case input voltage for the Ion Pump Power Supply must first be determined.

$$(1) \quad \underline{V_{CC}} = \left[V_{CC(NOM)}\right] \underline{[\text{Tol on 20 V}]} - \overline{I}_{20\ V}\, \overline{R}_{INDUCTOR}$$

$$(2) \quad \overline{V_{CC}} = \left[V_{CC(NOM)}\right] \overline{[\text{Tol on 20 V}]}$$

where: $V_{CC(NOM)} = 20\ V \pm 1\%$

$$\left.\begin{array}{ll} (1) & \underline{V_{CC}} = 19.6 \\ (2) & \overline{V_{CC}} = 20.2 \end{array}\right\} \begin{array}{l} \text{Max and} \\ \text{Min } V_{CC} \end{array}$$

$\overline{R}_{INDUCTOR} = 0.25$ ohm

$\overline{I}_{20\ V} = 0.8$ A

Next the maximum and minimum transformer winding voltages will be determined.

<u>Winding #1:</u> Drive Oscillator base drive windings 4-5-6

$$(3)\quad \overline{V_{W1}} = \left\{\left[\overline{V_{CC}} - \underline{V_{CE(SAT)}}\,(Q1,Q2) - \underline{I_C}\,(Q1,Q2)\,\underline{R_{W(PRI)}}\right] \times \overline{\text{Ratio}}\,(PRI:W1)\right\} - (\underline{I_{W1}}\,\underline{R_{W1}})$$

$$(4)\quad \underline{V_{W1}} = \left\{\left[\underline{V_{CC}} - \overline{V_{CE(SAT)}}\,(Q1,Q2) - \overline{I_C}\,(Q1,Q2)\,\overline{R_{W(PRI)}}\right] \times \underline{\text{Ratio}}\,(PR1:W1)\right\} - (\overline{I_{W1}}\,\overline{R_{W1}})$$

where:

Ratio (PRI: W1)	= 8:1 $\pm$3%
$\overline{V_{CE(SAT)}}_{(Q1,Q2)}$	= 0.3 V
$\underline{V_{CE(SAT)}}_{(Q1,Q2)}$	= 0 V
$\overline{V_{CC}}$	= 20.2 V
$\underline{V_{CC}}$	= 19.6 V
$\overline{R_{WI}}$	= 3 ohms
$\underline{R_{WI}}$	= 2 ohms
$\overline{R_{W(PRI)}}$	= 6 ohms
$\underline{R_{W(PRI)}}$	= 4 ohms
$\overline{I_{WI}}$	= 5 mA
$\underline{I_{WI}}$	= 2 mA

$\overline{I_{C(Q1,Q2)}}$ = 65 mA

$\underline{I_{C(Q1,Q2)}}$ = 0

(3) $\overline{V_{WI}}$ = 2.60 V

(4) $\underline{V_{WI}}$ = 2.27 V

Max and min voltages for drive oscillator base drive

<u>Winding #2</u>: Power Converter Base Drive Windings 7-8-9

$$(5)\quad \overline{V_{W2}} = \left\{\left[\overline{V_{CC}} - \underline{V_{CE(SAT)}}\ (Q1,Q2) - \underline{I_{C(Q1,Q2)}}\ \underline{R_{W(PRI)}}\right] \times \overline{\text{Ratio}}\ (PR1{:}W2)\right\} - \left(\underline{I_{W2}}\ \underline{R_{W2}}\right)$$

$$(6)\quad \underline{V_{W2}} = \left\{\left[\underline{V_{CC}} - \overline{V_{CE(SAT)}}\ (Q1,Q2) - \overline{I_{C(Q1,Q2)}}\ \overline{R_{W(PRI)}}\right] \times \underline{\text{Ratio}}\ (PR1{:}W2)\right\} - \left(\overline{I_{W2}}\ \overline{R_{W2}}\right)$$

where: Ratio (PRI: W2) = 5:1 $\pm$3%

$\underline{R_{W2}}$ = 4 ohms

$\overline{R_{W2}}$ = 5 ohms

$\overline{I_{W2}}$ = 50 mA

$\underline{I_{W2}}$ = 25 mA

(5) $\overline{V_{W2}}$ = 4.06 V

(6) $\underline{V_{W2}}$ = 3.42 V

Max and min voltages for converter base drive

Winding #3: ±7.5 Vdc supply windings 10-11-12

$$(7)\quad \overline{V_{W3}} = \left\{ \left[\overline{V_{CC}} - \underline{V_{CE(SAT)}}\,(Q1,Q2) - \underline{I_{C(Q1,Q2)}}\;\underline{R_{W(PRI)}} \right] \times \overline{\text{Ratio}}\,(PRI:W3) \right\} - (\underline{I_{W3}}\;\underline{R_{W3}})$$

$$(8)\quad \underline{V_{W3}} = \left\{ \left[\underline{V_{CC}} - \overline{V_{CE(SAT)}}\,(Q1,Q2) - \overline{I_{C(Q1,Q2)}}\;\overline{R_{W(PRI)}} \right] \times \underline{\text{Ratio}}\,(PRI:W3) \right\} - (\overline{I_{W3}}\;\overline{R_{W3}})$$

where: Ratio (PRI: W3) = 2.35 ±3%

$\underline{R_{W3}}$ = 1.5 ohms

$\overline{R_{W3}}$ = 2.5 ohms

$\underline{I_{W3}}$ = 30 mA

$\overline{I_{W3}}$ = 60 mA

(7) $\overline{V_{W3}}$ = 8.82 V

(8) $\underline{V_{W3}}$ = 7.71

Max and min ac voltages for ±7.5 Vdc supplies

From the above ac voltages the maximum and minimum ±7.5 Vdc outputs will be calculated.

$$(9)\quad \overline{V\ \pm 7.5\ V} = \overline{V_{W3}} - \underline{V\ \text{diode}}$$

$$(10)\quad \underline{V\ \pm 7.5\ V} = \underline{V_{W3}} - \overline{V\ \text{diode}}$$

where: $\overline{V_{W3}}$ = 8.82 V

$\underline{V_{W3}}$ = 7.71 V

$\underline{V\ diode}$ = 0.5 V

$\overline{V\ diode}$ = 1.0 V

(9) $\overline{V\ \pm 7.5\ V}$ = 8.32 V

(10) $\underline{V\ \pm 7.5\ V}$ = 6.71 V

Max and min ±7.5 Vdc supply voltages

Beta requirements for all transistors must be checked. A power summary must first be developed to determine maximum chopper collector current.

Power Summary for Drive Oscillator

Load	Maximum Voltage (V)	Maximum Current (mA)	Maximum Power (mW)
Base Drive for Drive Oscillator	2 (2.60)	9.3	48
Power Converter Base Drive	4.06	46.3	188
±7.5 Vdc Supplies	8.82 V	60	1058
Core Loss	20 V	10	200
TOTAL LOAD POWER			1.94 W

The total load power for the drive oscillator is shown in the Power Summary.

The minimum necessary Beta for the chopper transistors (Q1 and Q2) must first be determined.

$$(11)\quad \underline{B_{Q1}} = \frac{\overline{I_{CQ1}}}{\underline{I_{BQ1}}}$$

$$(12)\quad \overline{I_{CQ1}} = \frac{\overline{P_{DRIVE\ OSC}}}{\underline{V_{CC}}}$$

$$(13)\quad \underline{I_{BQ1}} = \frac{2\,\underline{V_{W1}} - \overline{V_{FD1}} - \overline{V_{BE(Q2)}}}{2\,\overline{R1}}$$

$$(14)\quad \underline{B_{Q1}} = \frac{\overline{P_{DRIVE\ OSC}}}{\underline{V_{CC}}\left[\dfrac{2\,\overline{V_{W1}} - \underline{V_{FD1}} - \underline{V_{BE(Q2)}}}{2\,\overline{R1}}\right]}$$

where: $\underline{B_{Q1}} = 50$

$\overline{P_{DRIVE\ OSC}} = 1.94$ W

$\underline{V_{CC}} = 19.6$ V

$\underline{V_{FD1}} = 1$ V

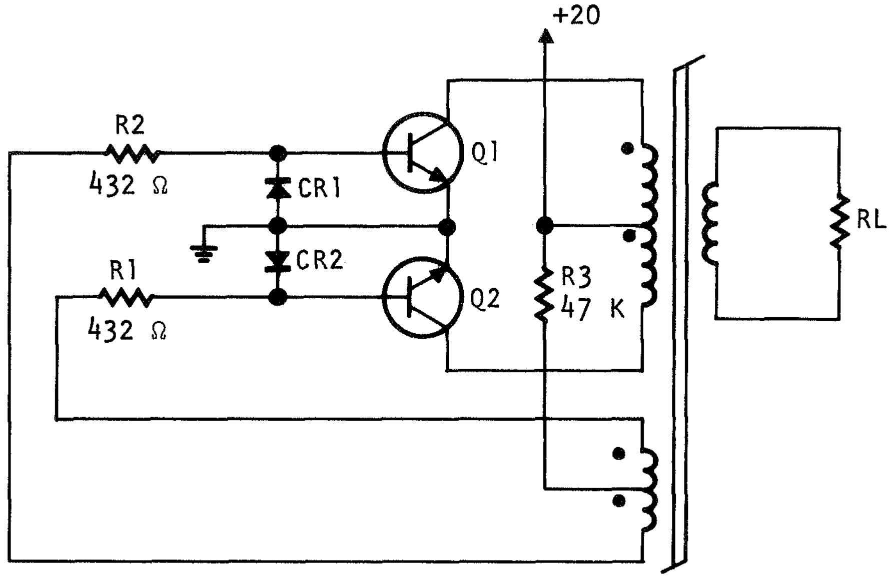
$\underline{V_{BE(Q2)}} = 1$ V

$\overline{R1} = 440$ ohms

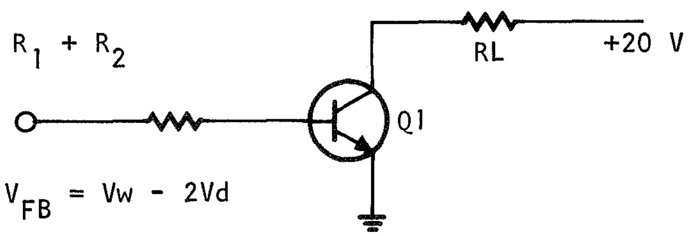
therefore: $\underline{B_{Q1}} = 26.4$

The minimum available Beta from the PG1074 is 50; the maximum needed is 26.4: therefore, there is sufficient gain.

The drive oscillator must be checked to ensure starting under worst case conditions. The minimum worst case open loop gain must be shown to at least 1.0.



Drive Oscillator Circuit



Drive Oscillator Equivalent Circuit

To find the gain of the equivalent circuit:

(15) $\Delta V_{OUT} = \Delta I_C \, R_L$

$\Delta I_C = \Delta I_B \, B$

$\Delta V_{OUT} = \Delta I_B \, B \, R_L$

Sine $\Delta I_B = \dfrac{\Delta V_{IN}}{R_{IN}}$

(17) $\dfrac{\Delta V_{OUT}}{\Delta V_{IN}} = \dfrac{B \, R_L}{R_{IN}}$

(18) $R_{IN} = (R1 + R2) + R_{B'} + (B + 1) \, R_E$

(19) $Gain_{(VOLTAGE)} = \dfrac{B \, R_L}{(R1 + R2 + R_{B'} + (B + 1) \, R_E}$

and $R_E = \dfrac{26}{I_E \text{ (mA)}}$ ohms

$R_{B'} = 100$ ohms or less for $I_E > 1$ mA

$R_L = 198$ ohms (from power summary drive oscillator)

therefore:

(20) $Gain_{(VOLTAGE)} = \dfrac{B \, R_L}{(R1 + R2) + R_{B'} + \dfrac{(B + 1) \, 26}{I_E}}$

From (3)

(21) $\underline{Gain_{(VOLTAGE)}} = \dfrac{\underline{B} \, \underline{R_L}}{(\overline{R}1 + \overline{R}2) + \overline{R}_{B'} + \dfrac{(\underline{B} + 1) \, 26 \, K^*}{\underline{I_E}}}$

where: K^* modifies the constant 26 to allow for worst case temperature.

Since $R_E \cong \frac{26}{I_E} = \frac{KT}{q\ I_E}$ at 25°C

(22) K = Boltmann's Constant

T = Temperature °K

q = Electronic charge

at 40°C

(23) $$R_E \cong \frac{26}{I_E}\ \frac{313}{298} = \frac{27.3}{I_E} \text{ ohms}$$

Since minimum gain appears at minimum emitter current, this occurs during startup when the starting resistor (R3) supplies the only base current.

$\therefore$ During starting

(24) $$\underline{I_E} = \underline{B}\ \underline{I_B}$$

where: $\underline{B} = 40$

(25) $$\underline{I_B} = \frac{V_{\underline{CC}} - \overline{V}_{\underline{BE}(Q1,Q2)}}{\overline{R_B} + \frac{\left(\overline{R1,2} + \overline{R_{TRANS\ W1}}\right)}{2}}\ \frac{1}{2}$$

The multiplying factor of one-half assumes the worst case condition of current splitting equally between Q1 and Q2.

$\underline{V_{CC}} = 17.6$ V, $\overline{V_{BE(Q1,Q2)}} = 1$ V

R3 = 57.5 K $R_{TRANS\ W1} = 3$ ohms

$\overline{R1},2 =$ 432 ohms

$\underline{B}_{(Q1,Q2)} = 40$

$$\therefore \left.\begin{array}{l} \underline{I_B} \cong 0.32 \text{ mA} \\ \underline{I_E} = 9.6 \text{ mA} \end{array}\right\} \text{Min base and emitter current for Q1, Q2}$$

Solving for the gain of the equation 21

$$\text{Gain (V) Circuit} = 0.3 \left.\right\} \begin{array}{l}\text{Min gain (V)}\\ \text{of Q1, Q2}\end{array}$$

This is only the gain of the transistor circuit, the transformer winding ratio and voltage drops of the diodes and transistors must be accounted for.

$$(26) \quad \underline{\text{Gain (V) Total Open Loop}} = \left[\text{Gain (V) Circuit}\right] \times$$

$$\left[\underline{\text{Gain (V) Transformer}}\right] \times \left[\overline{\text{Atten}} \text{ (V) due to } V_{BE}, V_D\right]$$

where:

$$\underline{\text{Gain (V) Transformer}} = 0.242 \text{ (From timing converter worst case analysis)}$$

$$(27) \quad \underline{\text{Atten (V) due to } V_{BE} \text{ and diode drop} = \underline{V_W} - 2V_D}$$

$$\text{and } \underline{V_W} = 4.75 \text{ V}$$

$$\overline{V_D} = 1.0 \text{ V}$$

$$\underline{\text{Gain (V) Total open loop}} = 1.3 \left.\right\} \begin{array}{l}\text{Min worst case open}\\ \text{loop gain during starting.}\end{array}$$

The preceding analysis has shown that, worst case, there is sufficient loop gain for the timing converter to start reliability.

<u>Worst Case Analysis: Ion Pump Current Sensor</u>.- The worst case trip points for the current sensor must be determined. The circuit diagram is shown in Figure 4 and an equivalent circuit to the input circuitry of the current sensor is shown in Figure 7.

Voltage, V1, is generated by the ion pump current flowing through resistor R35. V4 and V5 are the voltage sources that determine the circuit trip points. Voltage source V5 sets V3 nominally to minus one volt. V4 is switched in to the node at V3 when the minus one volt trip point is desired. The variable resistors R26 and R29 allow selection of current trip points and provide a means to compensate for initial tolerances on components and supply voltages.

The maximum and minimum current trip points (settable by R26 and R29) are to be determined. The two discrete levels of V1 are defined below:

$V1_{(ON)}$ = V1 when relay K1 is energized

$V1_{(ON)}$ nominal = -1.0 V

$V1_{(OFF)}$ = V1 when relay K1 is not energized

$V1_{(OFF)}$ nominal = -0.1 V

Solving for $V1_{(ON)}$ maximum and minimum: (S1 open in model)

(28) (1) $$\left|\overline{V1_{(ON)}}\right| = \left|\overline{V3_{(ON)}}\right| + \overline{V_{OS(A1)}} + \overline{I_{OS(A1)}}\;\overline{R_{EQ(ON)}} + \overline{I_{BIAS(A1)}}\left|\left(\overline{R_{EQ(ON)}} - R31\right)\right|$$

(29) (2) $$\left|\underline{V1_{(ON)}}\right| = \left|\underline{V3_{(ON)}}\right| - \overline{V_{OS(A1)}} - \overline{I_{OS(A1)}}\;\overline{R_{EQ(ON)}} - \overline{I_{BIAS(A1)}}\left|\left(\overline{R_{EQ} - R21}\right)\right|$$

where: $\underline{R_{EQ(ON)}} = (\underline{R25} + \underline{R29})\,||\,\underline{R30}$

$\left|\overline{V3_{(ON)}}\right|$ and $\left|\underline{V3_{(ON)}}\right|$ must be determined with S1 open:

(30) (3) $$\left|\overline{V3_{(ON)}}\right| = \overline{V5}\,\frac{\overline{R30}}{\underline{R29} + \underline{R25} + \overline{R30}}$$

(31) (4) $$\left|\underline{V3_{(ON)}}\right| = \underline{V5}\,\frac{\underline{R30}}{\overline{R29} + \overline{R25} + \underline{R30}}$$

where: $\overline{V5}$ = 5.355 V

$\underline{V5}$ = 4.845 V

$\overline{R30} = 10.2\ K \quad \overline{R29} = 55\ K \quad \overline{R25} = 13.25\ K$

$\underline{R30} = 9.8\ K \quad \underline{R29} = 0 \quad \underline{R25} = 12.75\ K$

$$\left.\begin{array}{l} |\overline{V3_{(ON)}}| = -2.38\ V \\ |\underline{V3_{(ON)}}| = -0.61\ V \end{array}\right\} \text{Max and min Value } V3_{(ON)}$$

Solving equations (28) (29)

where: $\overline{V}_{OS(A1)} = 10\ mV \quad \overline{R31} = 6.79\ k$

$\underline{I}_{OS(A1)} = 70\ nA \quad \underline{R31} = 6.50\ k$

$\underline{R_{EQ(ON)}} = 5.55\ k$

$\overline{I}_{BIAS(A1)} = 300\ nA \quad R_{EQ(ON)} = 8.86\ k$

$$\left.\begin{array}{ll} (28) & |\overline{V1_{(ON)}}| = 2.41\ V \\ (29) & |\underline{V1_{(ON)}}| = 0.59\ V \end{array}\right\} \text{Max and min trip voltages with relay energized}$$

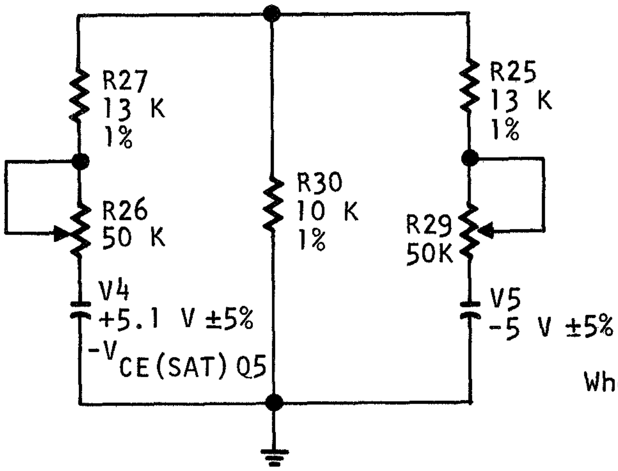
Solving for $V1_{(OFF)}$ maximum and minimum: (S1 closed in model)

$$(32) \quad |\overline{V1_{(OFF)}}| = |\overline{V3_{(OFF)}}| + \overline{V}_{OS(A1)} + \overline{I}_{OS(A1)}\ \overline{R}_{EQ(OFF)} + \overline{I}_{BIAS(A1)}\ |\overline{R}_{EQ(OFF)} - \underline{R31}|$$

$$(33) \quad |\overline{V1_{(OFF)}}| = |\underline{V3_{(OFF)}}| - \overline{V}_{OS(A1)} - \overline{I}_{OS(AL)}\ \overline{R}_{EQ(OFF)} - I_{BIAS(A1)}\ \left|\left(R_{EQ(OFF)} - \underline{R31}\right)\right|$$

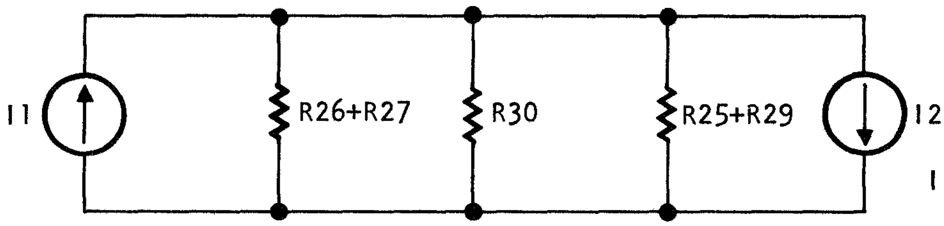
$\overline{V3_{(OFF)}}$ and $\underline{V3}$ must be determined with S1 closed: The equivalent circuit is shown below.

Equivalent Circuit for Solution of $V3_{(OFF)}$



Where: $V_{CE(SAT)}$ Q5 = 0.25

$V_{CE(SAT)}$ Q5 = 0

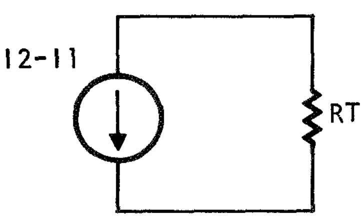


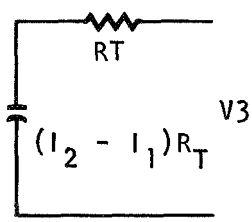
$$I_1 = \frac{V4}{R26+R27}$$

$$I_2 = \frac{V5}{R25+R29}$$

which simplifies to: (R2 + R27) || R30 || (R25 + R29)

(R26 + R27)||R30||(R25 + R29)





$$(34)\quad \left|\overline{V3_{(OFF)}}\right| = \left[\frac{\overline{V5}}{\underline{R25} + \underline{R29}} - \frac{\underline{V4}}{\overline{R26} + \overline{R27}}\right] \left[(\overline{R26} + \overline{R27}) \parallel \overline{R30} \parallel (\underline{R25} + \underline{R29})\right]$$

$$(35)\quad \left|\underline{V3_{(OFF)}}\right| = \left[\frac{\underline{V5}}{\overline{R25} + \overline{R29}} - \frac{\overline{V4}}{\underline{R26} + \underline{R27}}\right] \left[(\underline{R26} + \underline{R27}) \parallel \underline{R30} \parallel (\overline{R25} + \overline{R29})\right]$$

where:

$\overline{V4}$	= 5.355 V	$\overline{R27}$	= 13.25 k
$\underline{V4}$	= 4.595 V	$\underline{R27}$	= 12.75 k
		$\overline{R26}$	= 55 k
		$\underline{R26}$	= 0
$\overline{V5}$	= 5.355 V	$\overline{R25}$	= 13.25 k
$\underline{V5}$	= 4.845 V	$\underline{R25}$	= 12.75 k
		$\overline{R29}$	= 55 k
		$\underline{R29}$	= 0
		$\overline{R30}$	= 10.2 k
		$\underline{R30}$	= 9.8 k

$$\left.\begin{array}{ll} (34) & \left|\overline{V3_{(OFF)}}\right| = 1.85\ \text{V} \\ (35) & \left|\underline{V3_{(OFF)}}\right| = 1.80\ \text{V} \end{array}\right\} \text{Max and min values } \left|V3_{(OFF)}\right|$$

NOTE: $\underline{V3_{(OFF)}}$ = Neg

The maximum and minimum (settable) values of $V1_{(OFF)}$ can be determined.

Although $\underline{V3_{(OFF)}}$ solved for an Equation (8) in a positive value, the minimum usable trip level is zero volts. The contribution of offset and bias current is approximately twelve millivolts in the analysis for $V3_{(ON)}$. It will be dropped from the solutions of $V3_{(OFF)}$, since it is negligible.

$$\left.\begin{array}{lll}(32) & \overline{V1_{(OFF)}} = 1.85\ \text{V} \\ (33) & \underline{V1_{(OFF)}} = 0\end{array}\right\}\ \text{Max and min trip voltage with relay OFF.}$$

Considering the tolerance on the one kilohm current sense resistor, the maximum and minimum values of the two current trip levels are as follows:

$$\overline{I_{TRIP\ (RELAY\ ON)}} = 2.46\ \text{mA}$$

$$\underline{I_{TRIP\ (RELAY\ ON)}} = 0.58\ \text{mA}$$

$$\overline{I_{TRIP\ (RELAY\ OFF)}} = 1.89\ \text{mA}$$

$$\underline{I_{TRIP\ (RELAY\ OFF)}} = 0\ \text{mA}$$

The above values indicate the initial setability of the current trip points.

After initial set, the drift due to temperature is another consideration to be made. The drift of the initial set point results mainly from changes in the plus or minus five volt reference voltages.

The temperature coefficient of the zener diode (1N5523B) is not specified in the manufacturer's data. The basic tolerance, however, is plus or minus five percent and a maximum change of plus or minus three percent for temperature and bias current variations have been assumed as worst case. Other items responsible for drift are the current limit adjustment potentiometers. They have an accuracy of plus or minus two percent after set from zero to fifty-five degrees centigrade. The overall worst case drift in current trip point is, therefore, plus or minus five percent.

The accuracy of the sensor could be greatly increased by using select-at-test resistors instead of potentiometers. Also, more accurate temperature compensated zener diodes could be utilized to further reduce drift error.

To complete the current sensor worst case analysis, the transistors are checked for sufficient base drive under worst case conditions. Q1 is first checked for sufficient base drive.

$$(36)\quad (1)\quad \underline{B_{Q3}}\ \text{(Required)} = \frac{\overline{I_{CQ3}}}{I_{BQ3}}$$

$$(37)\quad (2)\quad \overline{I_{CQ3}} = \left[\frac{\overline{V(+7.5)} - \overline{V(-7.5)}}{\underline{R_{RELAY}} + \underline{R6}}\right]$$

$$(38)\quad (3)\quad \underline{I_{BQ3}} = \left(\frac{\left[\overline{V(+7.5)} - \overline{V(-7.5)}\right]\left(\overline{VD9} + \overline{VD11}\right) - \overline{V_{BE(SAT)Q3}}}{\overline{R8}}\right) - \left(\frac{\overline{V_{BE(SAT)Q3}}}{\underline{R7}}\right)$$

where:

$\underline{R_{RELAY}}$	=	189 ohms	$\overline{V(+7.5)}$	=	8.32 V
$\underline{R6}$	=	98 ohms	$\overline{V(-7.5)}$	=	8.32 V
$\overline{R8}$	=	6.25 k	$\overline{VD9}$	=	1.0 V
$\underline{R7}$	=	36.4 k	$\overline{VD11}$	=	1.0 V
			$\overline{V_{BE(SAT)Q3}}$	=	1.0 V

(38) $\underline{I_{BQ3}}$ = 2.15 mA

(37) $\overline{I_{CQ3}}$ = 58 mA

(36) $\underline{B_{Q3}}$ = 27

Since Q3 has a minimum available Beta of 80, there is sufficient base drive.

Q2 is checked for sufficient base drive:

$$(39)\quad \underline{B_{Q2}}\ \text{(REQUIRED)} = \frac{\overline{I_{CQ2}}}{\underline{I_{BQ2}}}$$

$$(40)\quad \overline{I_{CQ2}} = \frac{\overline{V(+7.5)} - \overline{V(-7.5)} - \underline{VD11} - \underline{V_{CE(SAT)Q4}}}{\underline{R8}}$$

$$(41)\quad \underline{I_{BQ2}} = \frac{V(-7.5) - \overline{V_{BE(SAT)Q4}} - \overline{V_{CE(SAT)Q6}}}{\overline{R12}} - \frac{\overline{V_{BE(SAT)Q4}}}{\underline{R11}}$$

where:

$$\underline{R8} = 3.95\ \text{k}$$

$$\overline{R12} = 27\ \text{k}$$

$$\underline{R11} = 36.4\ \text{k}$$

$$\overline{V_{BE(SAT)Q4}} = 1\ \text{V}$$

$$\overline{V_{CE(SAT)Q6}} = 0.25\ \text{V}$$

$$\underline{V_{CE(SAT)Q4}} = 0$$

$$(41)\quad \underline{I_{BQ4}} = 0.26\ \text{mA}$$

$$(40)\quad \underline{I_{CQ4}} = 4.1\ \text{mA}$$

$$(39)\quad \underline{I_{Q4}}\ \text{(REQUIRED)} = 15.8$$

Since Q4 has a minimum available Beta of 80, there is sufficient base drive.

Q6 is checked for sufficient base drive:

$$(42)\quad \underline{B_{Q6}}\ \text{(REQUIRED)} = \frac{\overline{I_{CQ6}}}{\underline{I_{BQ6}}}$$

(43) $$\overline{I_{CQ6}} = \frac{\overline{V(-7.5)} - \underline{V_{BE(SAT)Q4}} - \underline{V_{CE(SAT)Q6}}}{\underline{R12}} - \frac{\underline{V_{BE(SAT)Q4}}}{\overline{R11}}$$

(44) $$I_{BQ6} = \frac{\underline{V_{O(A1)}} - \overline{V_{BE(SAT)Q6}}}{\overline{R23}} - \overline{V_{D17}} - \frac{\overline{V_{BE(SAT)(Q6)}}}{\underline{R22}}$$

where:

$\overline{V_{BE(SAT)Q6}}$	=	1V	$\underline{R12}$	=	17 k
$\underline{V_{CE(SAT)Q6}}$	=	0	$\overline{R11}$	=	57.7 k
$\underline{V_{O(A1)}}$	=	5 V	$\overline{R22}$	=	80.2 k
$\overline{V(-7.5\ V)}$	=	-8.32 V	$\overline{V_{D17}}$	=	1 V
			$\underline{R22}$	=	77.5 k

(44) $\underline{I_{BQ6}}$ = 0.025 mA

(43) $\overline{I_{CQ6}}$ = 0.43 mA

(42) $\underline{B_{Q6}}$ = 17.2

Since Q6 has a minimum available Beta of 60, there is sufficient base drive.

Q5 is checked for sufficient base drive:

(45) $$B_{Q5\ (REQUIRED)} = \frac{\overline{I_{CQ5}}}{\underline{I_{BQ5}}}$$

(46) $$\overline{I_{CQ5}} = \frac{\overline{V(+5\ V)} - \underline{V_{CE(SAT)Q5}} - \underline{V_{TRIP}}}{\underline{R27}}$$

$$(47) \qquad \underline{I_{BQ5}} = \frac{\underline{V(+5\ V)} - \overline{V_{BE(SAT)Q5}} - \overline{V_{CE(SAT)Q4}}}{\overline{R21}}$$

where:

$\overline{V(+5\ V)} = 5.35\ V \qquad \overline{R21} = 184\ k$

$\underline{V_{TRIP}} = 0 \qquad \underline{R27} = 12.7\ k$

$\underline{V_{CE(SAT)Q5}} = 0$

$\overline{V_{BE(SAT)Q5}} = 1$

$\overline{V_{CE(SAT)Q4}} = 0.25\ V$

(47) $\underline{I_{BQ5}} = 0.222\ mA$

$\overline{I_{CQ5}} = 0.42\ mA$

(45) $\underline{B_{Q5}} = 19.1$

Since Q5 has a minimum available Beta of 60, there is sufficient base drive.

The relay K1 is checked for worst case maximum and minimum voltage:

$$(48) \qquad \overline{V_{RELAY}} = \left[\overline{V(+7.5\ V)} - \overline{V(-7.5\ V)}\right]\left(\frac{\overline{R_{RELAY}}}{\underline{R6} + \overline{R_{RELAY}}}\right)$$

$$(49) \qquad \underline{V_{RELAY}} = \left[\underline{V(+7.5\ V)} - \underline{V(-7.5\ V)}\right]\left(\frac{\underline{R_{RELAY}}}{\overline{R6} + \underline{R_{RELAY}}}\right)$$

where:

$\overline{V(+7.5)} = 8.32\ V; \quad \overline{V(-7.5)} = -8.32\ V$

$\underline{V(+7.5)} = +6.71; \quad \underline{V(-7.5)} = -6.71\ V$

$\overline{R_{RELAY}} = 244$ ohms; $\overline{R6} = 102$ ohms

$\underline{R_{RELAY}} = 189$ ohms; $\underline{R6} = 98$ ohms

(49) $\underline{V_{RELAY}} = 8.72\ V$

(48) $\overline{V_{RELAY}} = 11.9\ V$

Manufacturer's data guarantees relay pull in at six and eight-tenths volts at twenty-five degrees centigrade and nine and three-tenths volts at 125 degrees centigrade. This would indicate approximately seven and two-tenths volts at the maximum temperature of forty degrees centigrade.

Thus, there is sufficient relay voltage worst case.

Worst Case Analysis: Power Converters (See Figure 5).- The 850 volt and five kilovolt supplies shall be checked worst case for minimum voltage and current output before current limiting occurs. From data on the diode 2N4401; five-tenths of a volt, $V_{BE(ON)}$, will produce less than one-tenth milliampere collector current. Since the chopper base current is about thirty milliamperes, the one-tenth milliampere drawn will not cause noticeable limiting. The following worst case analysis assumes a minimum of five-tenths of an ampere collector current with no effects of current limiting.

850 volt Converter:

(50) $$\underline{V_{850\ V}} = \left[\underline{V_{PRI(T2)}}\right]\left[\underline{ratio}(PRI:\ 850\ V)T2\right]\left[\underline{Winding\ Tol}\right] - \overline{2\ V_{DIODE}} - \overline{I_{850} R}_{(SEC)T2}$$

(51) $$\underline{V_{PRI(T2)}} = \underline{V_{CC}} - \overline{V_{CE(SAT)Q9,Q10}} - \overline{V_{BE(ON)Q7,Q8}} - I_c\ \overline{R_{PRI(T2)}}$$

where:	$\overline{I_{850\ V}}$	$\cong$	12 mA
	$\underline{V_{CC}}$	=	19.6 V
	$\overline{R_{INDUCTOR}}$	=	1/4 ohm
	$\overline{V_{DIODE(H.V.)}}$	=	10 V at 100 mA
	$\overline{V}_{BE(ON)Q7,Q8}$	=	0.5 V
	$\underline{\text{Ratio}}_{(PRI:\ 850\ V)T2}$	=	43.4: 1
	$\overline{V}_{CE(SAT)Q9,Q10}$	=	0.5 V
	$\underline{\text{winding tolerance}}$	=	0.98
	$\overline{R_{(PRI)T1}}$	=	0.2 ohms
	$\overline{R_{(SEC)T1}}$	=	175 ohms
	$\overline{I_{20\ V}}$	=	0.7 A
	I_C	=	0.5 A(Defined)
(50)	V_{850}	=	763 V Min output voltage

To solve for minimum output current under these conditions, solve for the power transferred to the secondary of the transformer.

(52) $$\underline{P_{SEC}} = \underline{V_{PRI(T1)}} \times I_c \times \text{Efficiency}$$

where: From equation (51)

$\underline{V_{PRI(T1)}}$ = 18.47 V

$I_C \quad = \quad 0.5 \text{ A}$

$\text{Efficiency T1} \quad = \quad 0.95$

(52) $\therefore \quad \underline{P_{SEC}} \quad = \quad 8.77 \text{ W}$

(53a) $$\underline{I_{OUTPUT}} = \frac{\underline{P_{SEC}}}{\underline{V_{SEC}}}$$

where: $\underline{V_{SEC\ 850}} \quad = \quad 785 \text{ V}$

(53b) $\underline{I_{OUTPUT}} \quad = \quad 11.2 \text{ mA}$

From this analysis it has been shown that the worst case output voltage and the current at that voltage is 763 volts dc at eleven and two-tenths milliamperes.

Next, the chopper transistors must be checked for sufficient Beta. The maximum Beta needed is determined by dividing the worst case maximum collector current by the worst case minimum base current.

(54) $$\underline{B} = \frac{\overline{I_{C(Q9,Q10)}}}{\underline{I_{B(Q9,Q10)}}}$$

For $\overline{I_C}$, five-tenths of an ampere is used, because it was used in the previous worst case analysis for minimum output levels.

(55) $$\underline{I_{B(Q9,Q10)}} = \frac{\underline{V_{BASE}}_{\text{ DRIVE WIND}} - \overline{V_{BE(Q9,Q10)}} - \overline{R15}\ \overline{I_C}}{\overline{R4}}$$

where: $\overline{I_{C(Q9,Q10)}} \quad = \quad 0.5 \text{ A}$

$\underline{V_{BASE}}_{\text{ DRIVE WINDING}}$ (from worst case analyzer - Drive Osc) $\quad = \quad 3.42 \text{ V}$

$\overline{V_{BE(Q9,Q10)}} \quad = \quad 1 \text{ V}$

$\overline{R15}$ = 1.02 ohms

$\overline{R4}$ = 80.5 ohms

(55) $\underline{I_{B(Q9,Q10)}}$ = 23.8 mA

(54) $\underline{B}_{(Q9,Q10)}$ = 21

The minimum Beta for the chopper transistors is 40; the Beta required by the circuit is 21. Therefore, there is sufficient gain.

Five Kilovolt Converter.- Since the high voltage converter is the same as the low voltage, except for the X6 multiplier, the analysis is almost the same. The minimum transformer output voltage determined in the previous worst case analysis can be multiplied by six to yield a minimum value for the high voltage. However, there are some additional losses in the X6 multiplier due to transients. Since these losses are dependent on circuit capacitance, waveform squareness and their factors, they are impossible to predict. While a detailed Fourier analysis might be performed, the circuit parameters would still be questionable. Therefore, this analysis is considered only as a rough calculation. The actual worst case output voltage should be determined by subtracting the difference between nominal (5 kV) and calculated worst case voltage (4.6 kV at 1.8 mA) from the actual test data taken.

(56) $$\underline{V_{5\ kV}} = 6X\ \underline{V_{SEC\ 850}} - \overline{V_{DIODE}} - \overline{I_{850\ V}}\ \overline{R_{(SEC)T3}}$$

(57) $$\underline{I_{5\ kV}} = \frac{\underline{P_{SEC}}}{6X\ \underline{V_{SEC\ 850\ V}}}$$

where:

$\underline{V_{SEC}}$ = 785 V

$\overline{V_{DIODE}}$ = 10 V

$\overline{I_{850}}$ = 12 mA

$\overline{R_{(SEC)T3}}$ = 175 ohms

$\underline{P_{SEC}}$ = 8.87 W

(56) $\underline{V_{5\ kV}}$ = 4625 V

(57) $\underline{I_{5\ kV}}$ = 1.88 mA

$\left.\right\}$ Min output voltage and current at that voltage

The minimum calculated output of the high voltage supply is four and six-tenths kilovolts; the minimum current at this point is one and eight-tenths milliamperes. Since this is 400 volts below nominal; the actual worst case output can be figured from Figure 2 as 3900 volts.

The Beta requirement for the high voltage converter is identical to the requirement for the low voltage.

The remaining item to be worst case analyzed in the power converter is the power dissipation of the chopper transistors. From data on the 2N4401, typically a $V_{BE(ON)}$ of seven-tenths of a volt will cause thirty milliamperes collector current. This would cause severe current limiting of the chopper transistor since its total base drive current is about thirty milliamperes. At a $V_{BE(ON)}$ of eight-tenths of a volt, the 2N4401 would typically draw 500 milliamperes of collector current. The circuit current limit point would definitely be before this point, so worst case V_{BE} is chosen at eight-tenths of a volt. At eight-tenths volt, $V_{BE(ON)}$, the chopper transistor is drawing eight-tenths of a milliampere. Assuming full voltage across the transistor (which would be worst case), fifteen and six-tenths of a watt are dissipated in the chopper transistor. Considering the one-half duty cycle, each transistor worst case power dissipation is seven and eight-tenths watts. The Thermalloy 6166B heat sink used on each chopper transistor would rise to 126 degrees centigrade if the supply was short circuited (worst case condition). Since the heat sink is mounted on the power supply chassis, the temperature rise will not actually reach 126 degrees centigrade. The chopper transistor is rated for twenty watts dissipation at 100 degrees centigrade case temperature. While the case temperature may be somewhat over 100 degrees centigrade, the maximum power is seven and eight-tenths watts. The transistors are therefore well within their ratings at short circuit conditions.

CONCLUSION

Ion Pump Power Supply Test Results

The results of the Ion Pump Power Supply tests are summarized in Figures 8 through 11. Figures 8 and 9 show the output voltage vs current relationships. The graphs in Figures 10 and 11 show the input current and output ripple voltage vs the output load current. These graphs are included to demonstrate the major supply characteristics with respect to the operating conditions.

The mode switchover points have been designed and adjusted to one milliampere and one-tenth of a milliampere as required; however, for experimentation purposes, the switchover point can be varied by adjustment of R26 and R29. Adjustment is most readily accomplished by using a DVM and monitoring the voltage drop across the current sense resistor, R35.

With the ion pump supply operating in the high voltage mode (low output current) adjust R29 for one volt. Next, load the supply until it switches to the low voltage mode. Then adjust R26 to one-tenth of a volt. The one volt and one-tenth of a volt readings correspond to one milliampere and one-tenth of a milliampere, but can be varied at any operating point consistent with Figures 8 and 9. Since the current sense resistor is a one kilohm, a one milliampere per volt relationship exists.

During the ion pump supply tests, the output was short circuited several times. The supply first current limited on the high voltage mode and the switching (with a delay of about one second) down to the low voltage mode and again current limited. When the short was removed, the supply returned to normal operation. While this short circuit protection has demonstrated its effectiveness, it was designed for intermittent operation only. Therefore, the user is cautioned not to purposely short or overload the supply.

The Combined Studies Ion Pump Power Supply has demonstrated compliance to the specification requirements for a minimum power (under 30 watts) breadboard power supply capable of starting the ion pump at a pressure of 1×10^{-3} torr.

The deliverable unit has been tested and the test results documented in Figures 8 through 11 of this report. Additionally, the circuit schematic (Figure 12) of Appendix A and photographs (Figure 13) of Appendix B completely document the breadboard deliverable unit.

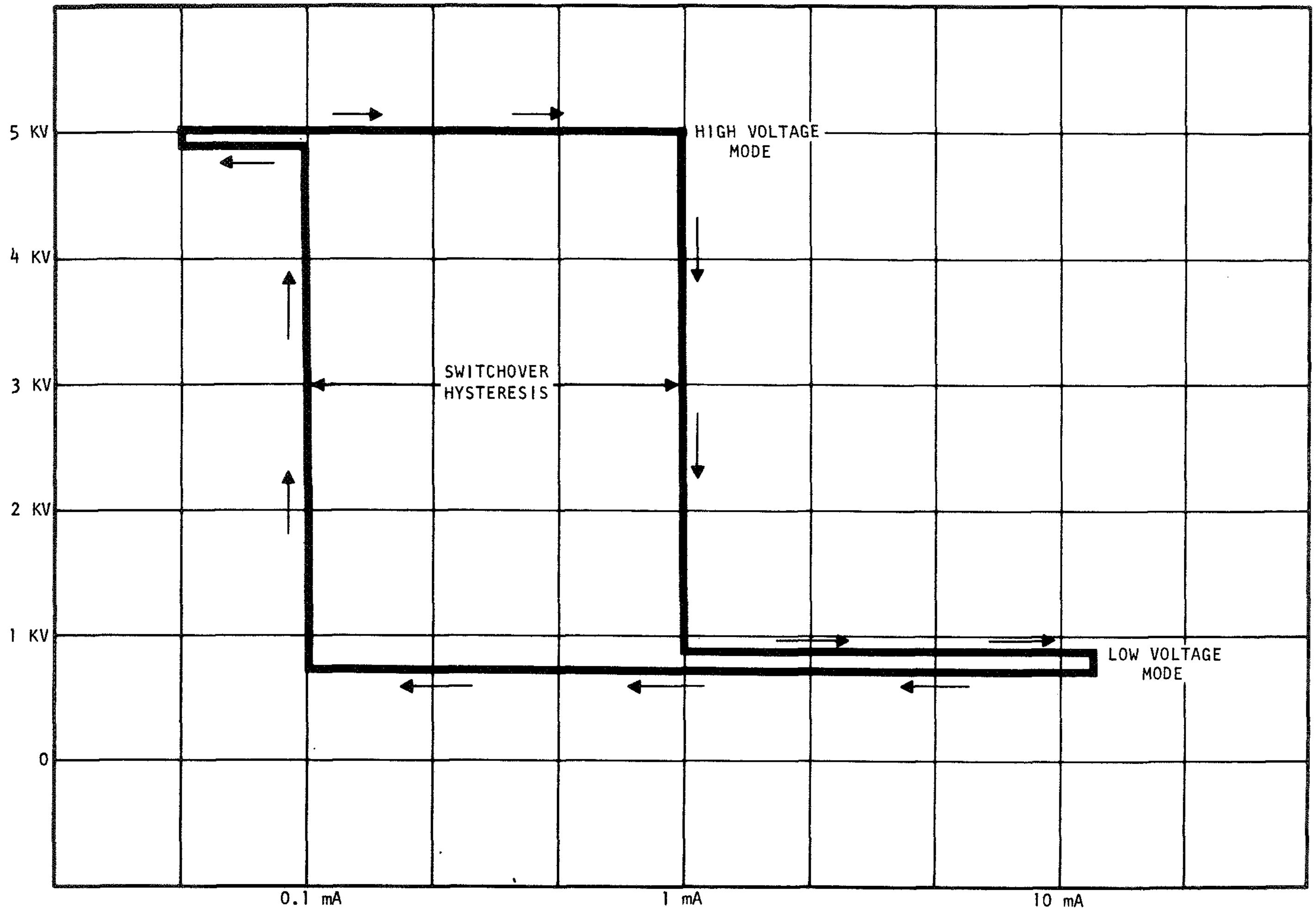


FIGURE 1.- Dual Mode Power Supply Output Characteristics - Voltage versus Current

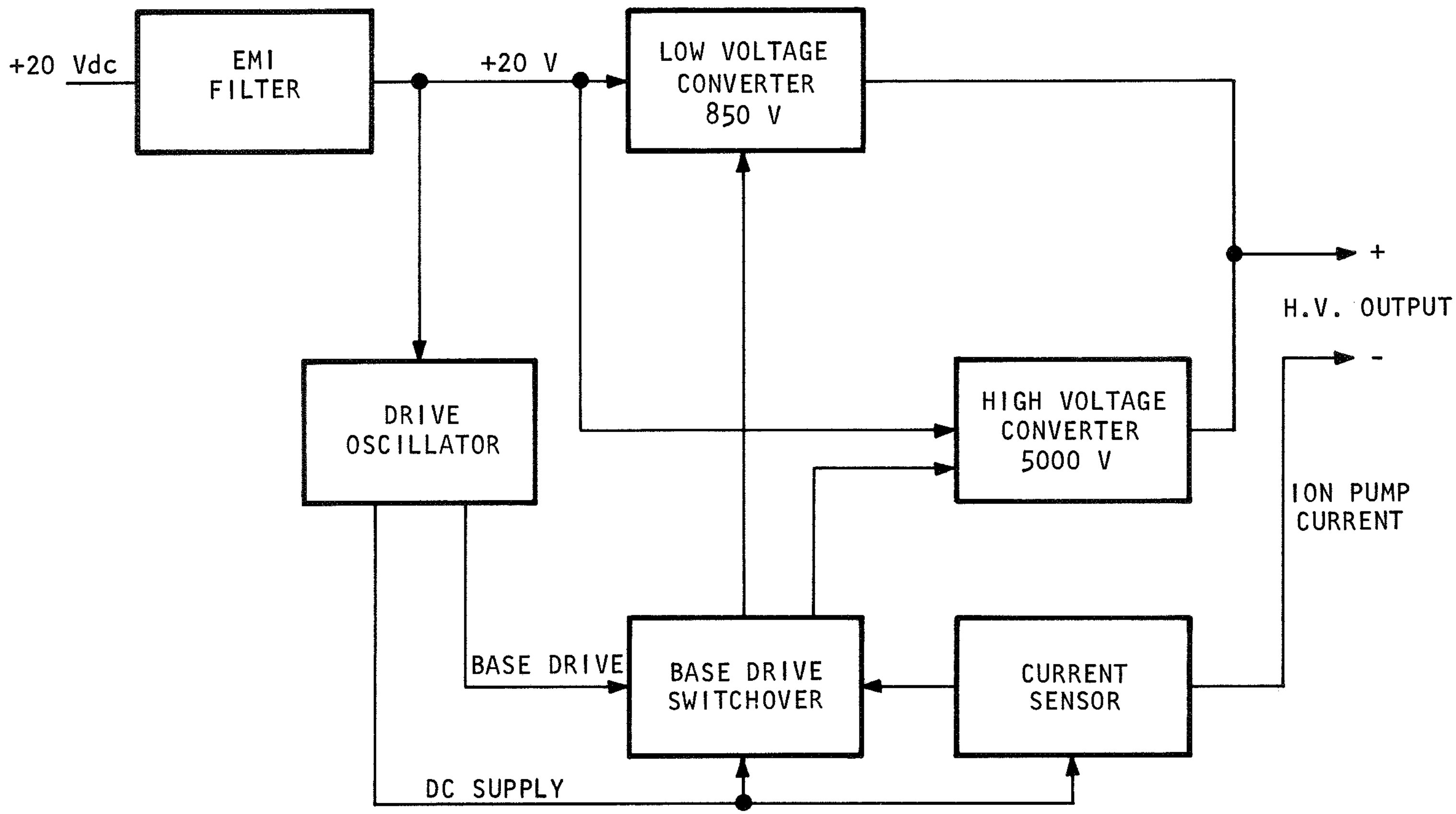
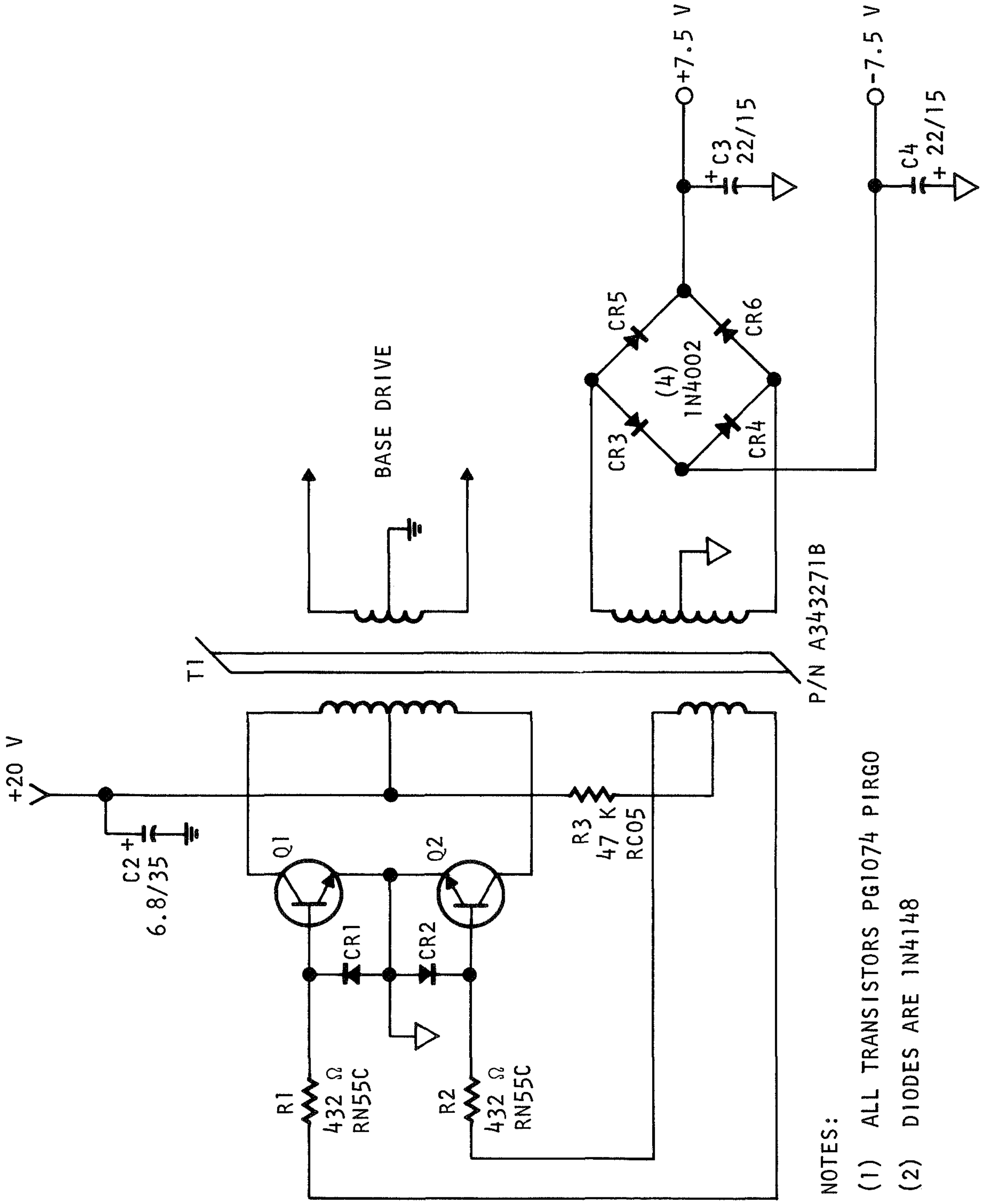


FIGURE 2.- System Block Diagram of Dual Mode Ion Pump Power Supply

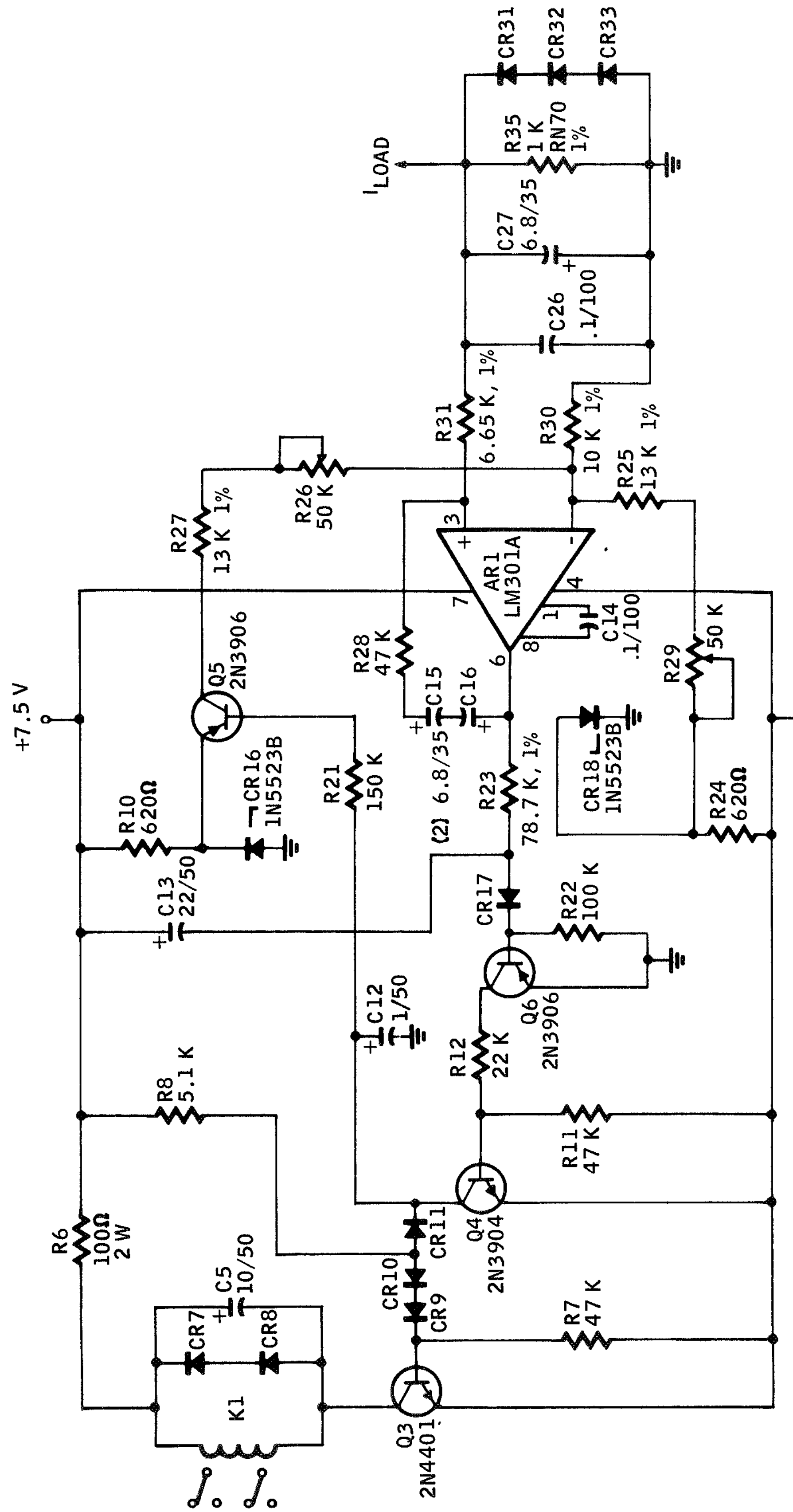


NOTES:

(1) ALL TRANSISTORS PG1074 PIRGO

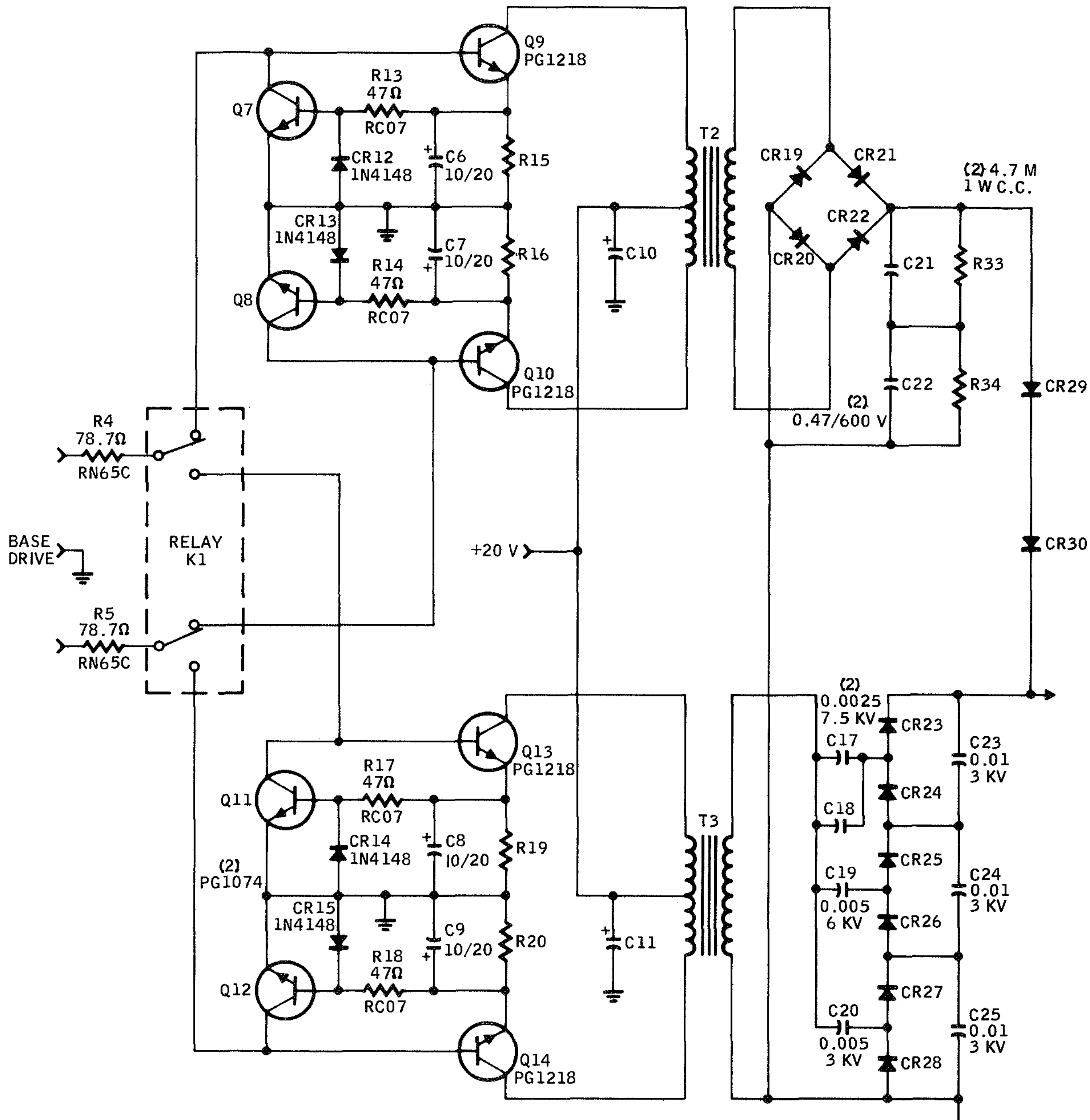
(2) DIODES ARE 1N4148

FIGURE 3.- Ion Pump Power Supply Drive Oscillator



NOTES, UNLESS SPECIFIED:
1. ALL RESISTORS RC05
2. ALL 1% RESISTORS RN55C

FIGURE 4.- Ion Pump Power Supply Current Sensor



NOTES, UNLESS SPECIFIED:
1. ALL DIODES SCE40
2. ALL TRANSISTORS 2N4401
3. ALL CAPACITORS 6.8-35 V
4. ALL RESISTORS 1Ω 1W IN.W.
5. TRANSFORMERS P/N 343270

FIGURE 5.- Dual Mode Ion Pump Power Supply Power Converters

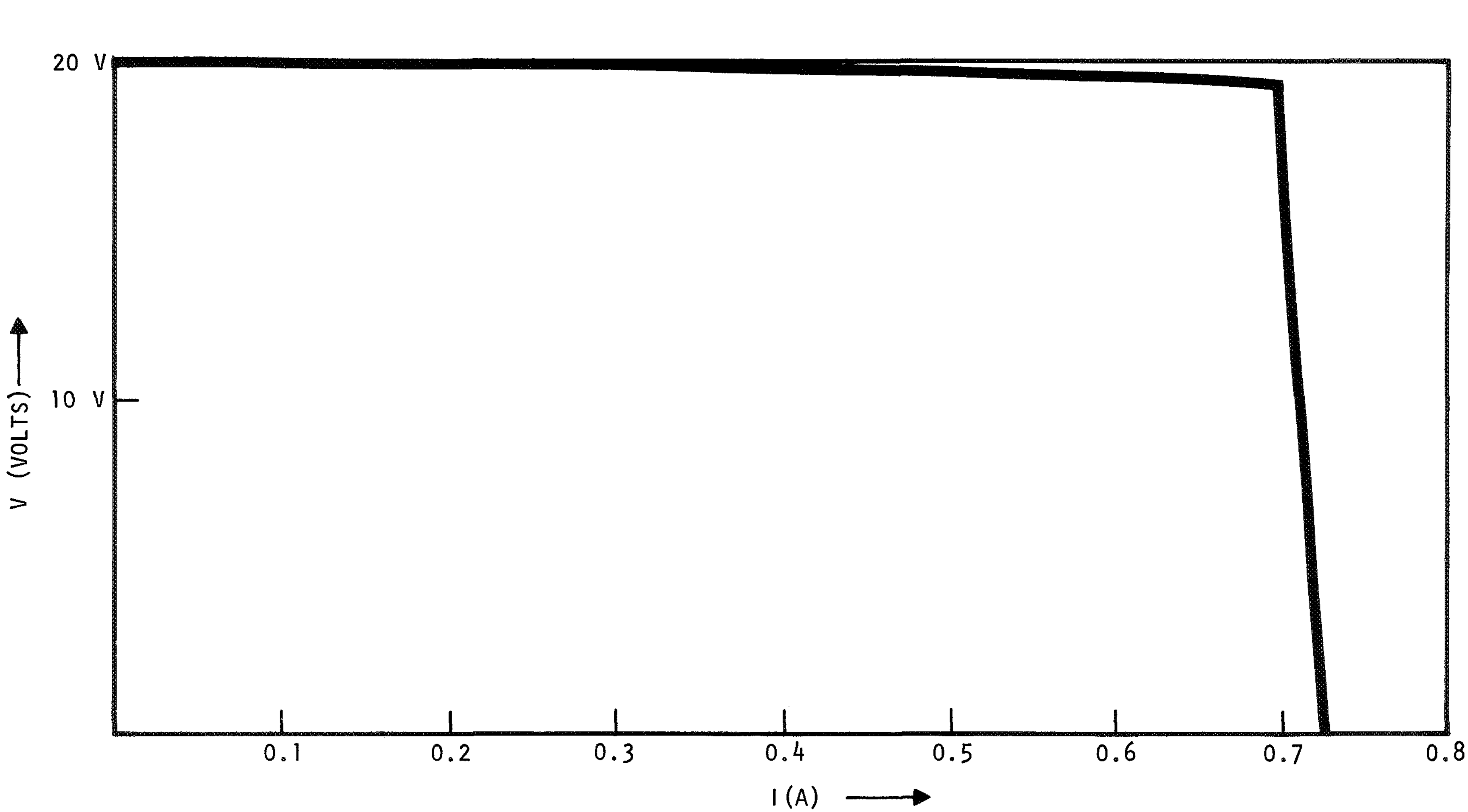


FIGURE 6.- Graph - Current Limit Circuit Test Results for Dual Mode DC/DC Converter Input Voltage vs Current

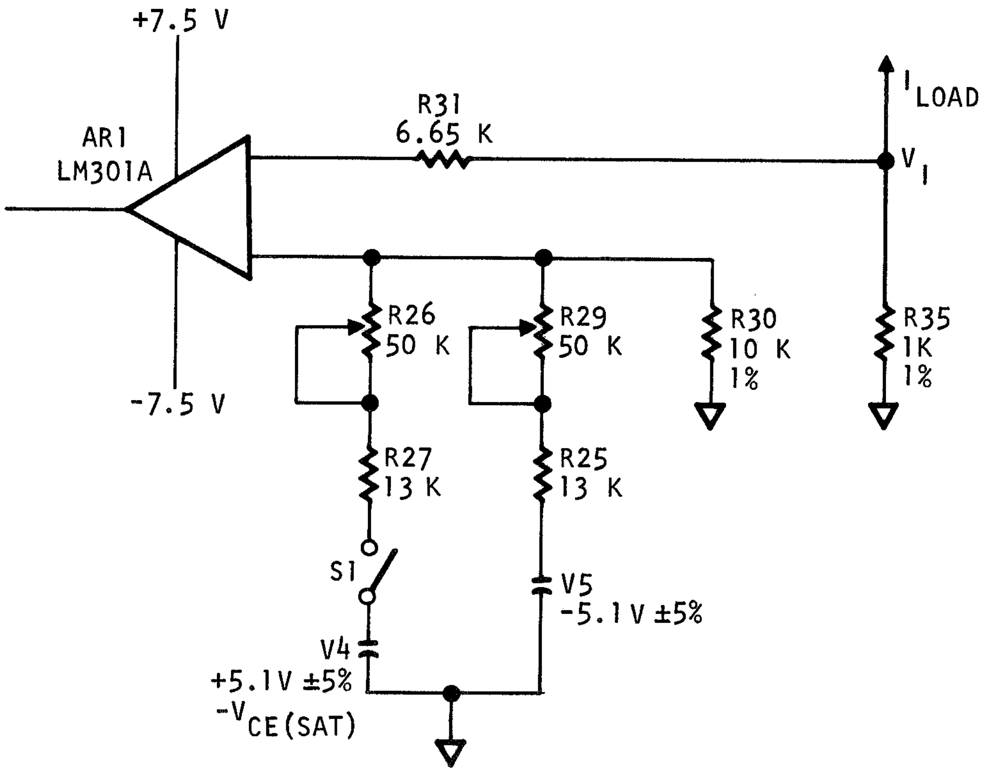


FIGURE 7.- Current Sensor - Equivalent Circuit Input Circuitry

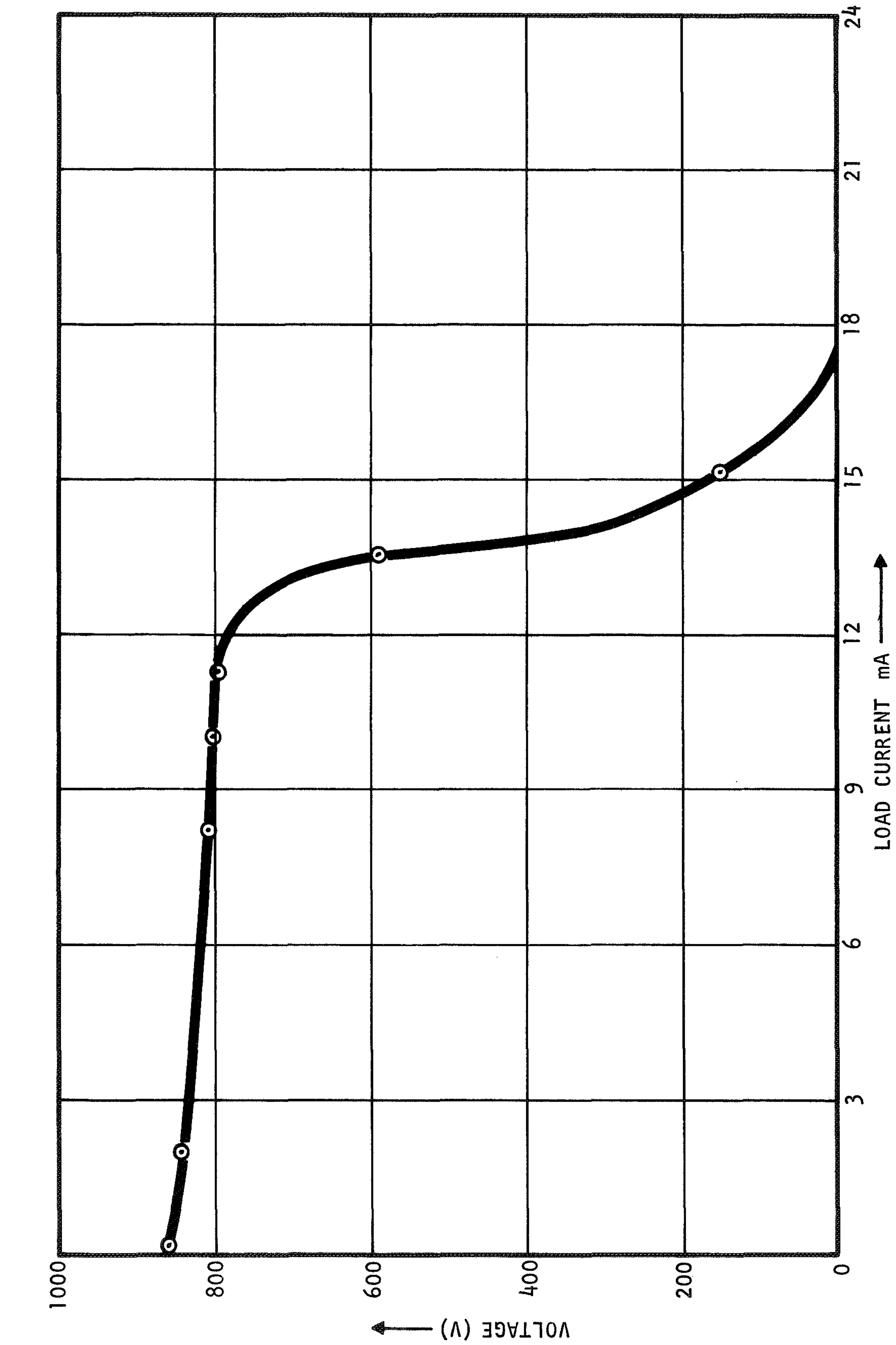


FIGURE 8.- Graph - 850 V Supply-Voltage vs Current

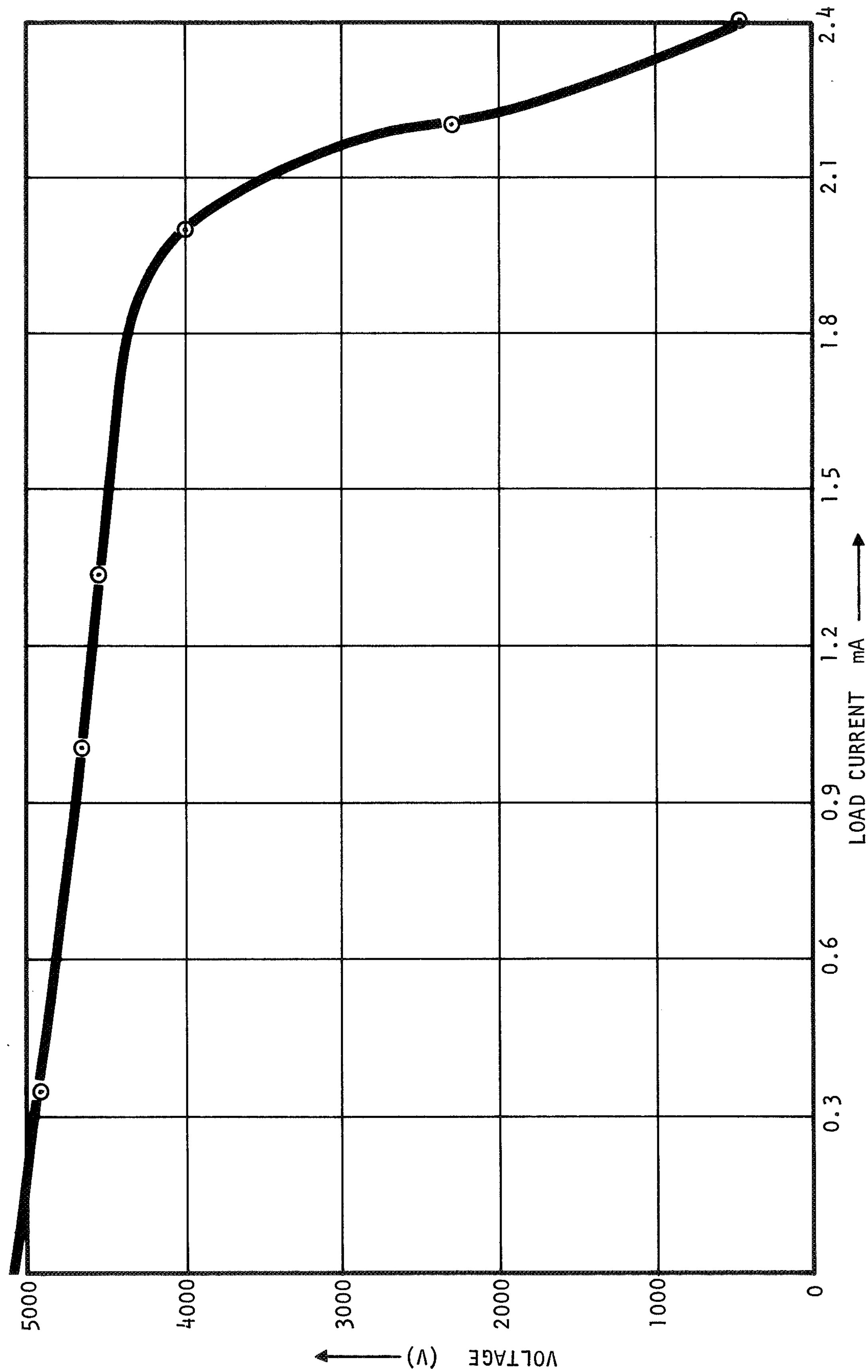


FIGURE 9.- Graph - 5000 V Supply-Voltage vs Current

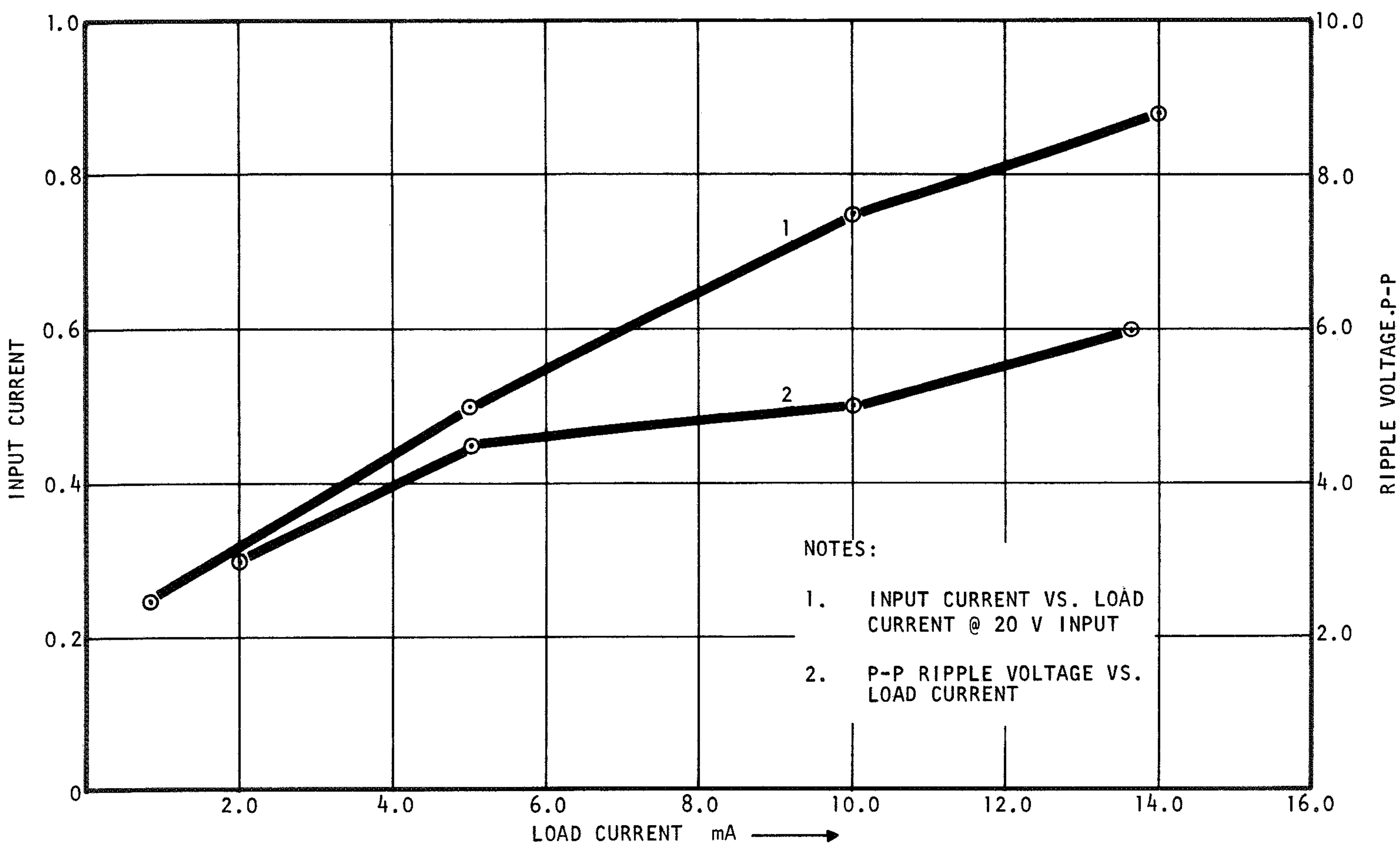


FIGURE 10.- Graph - 5000 V Supply (See Notes)

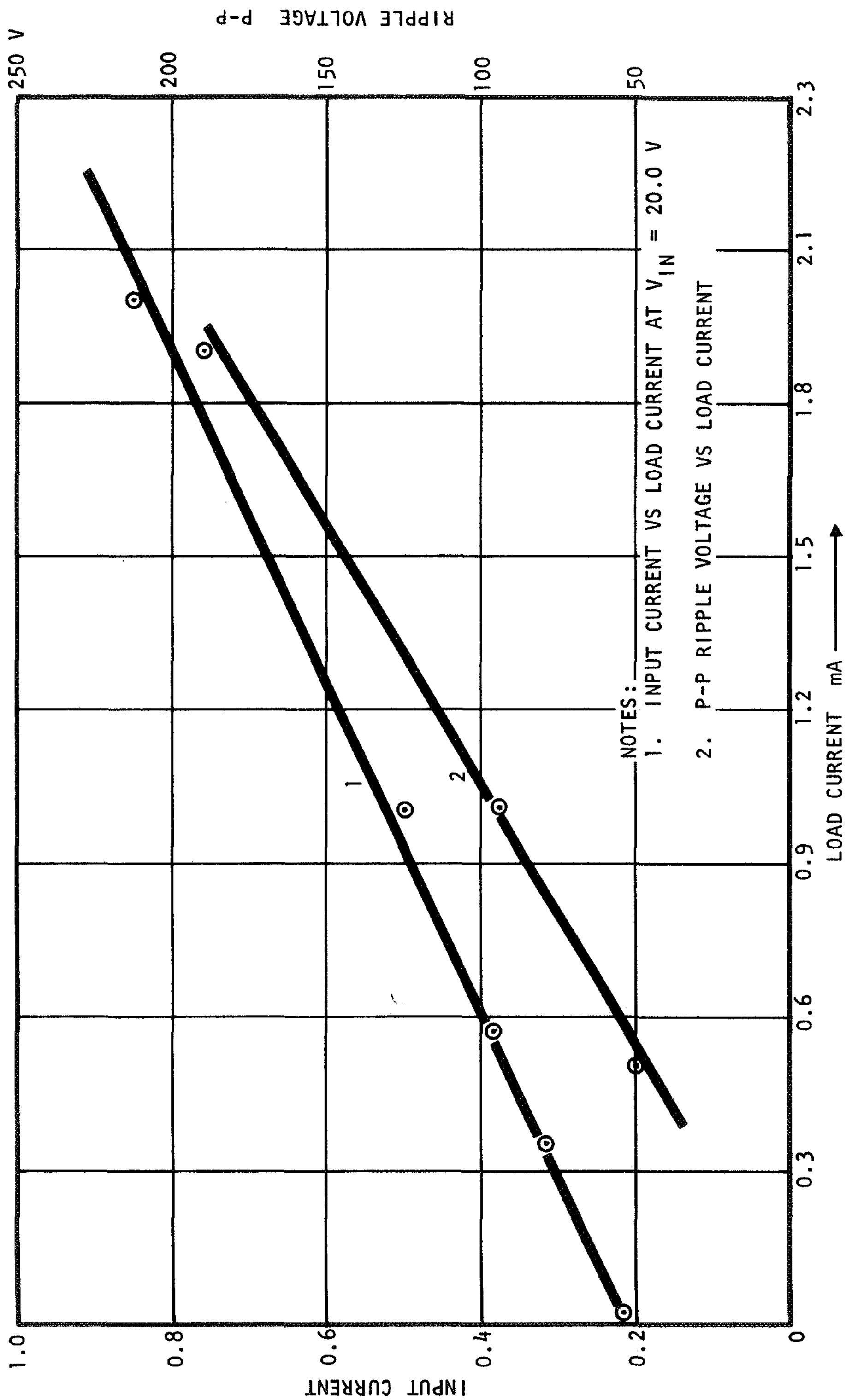


FIGURE 11.- Graph - 500 V Supply (See Notes)

APPENDIX A

ION PUMP POWER SUPPLY SCHEMATIC DIAGRAM

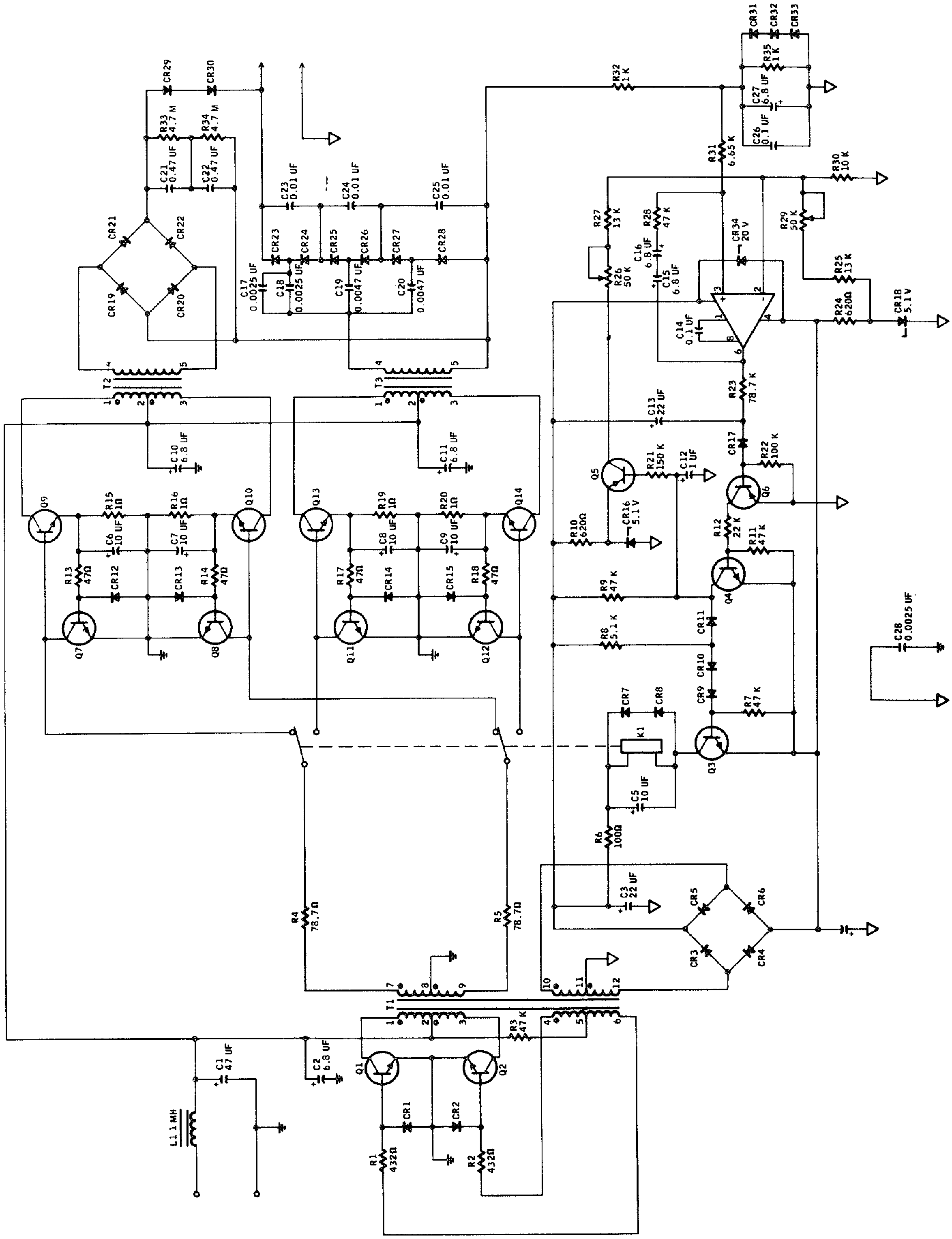


FIGURE 12.- Schematic, Ion Pump Power Supply

APPENDIX B

ION PUMP POWER SUPPLY ASSEMBLY LAYOUT

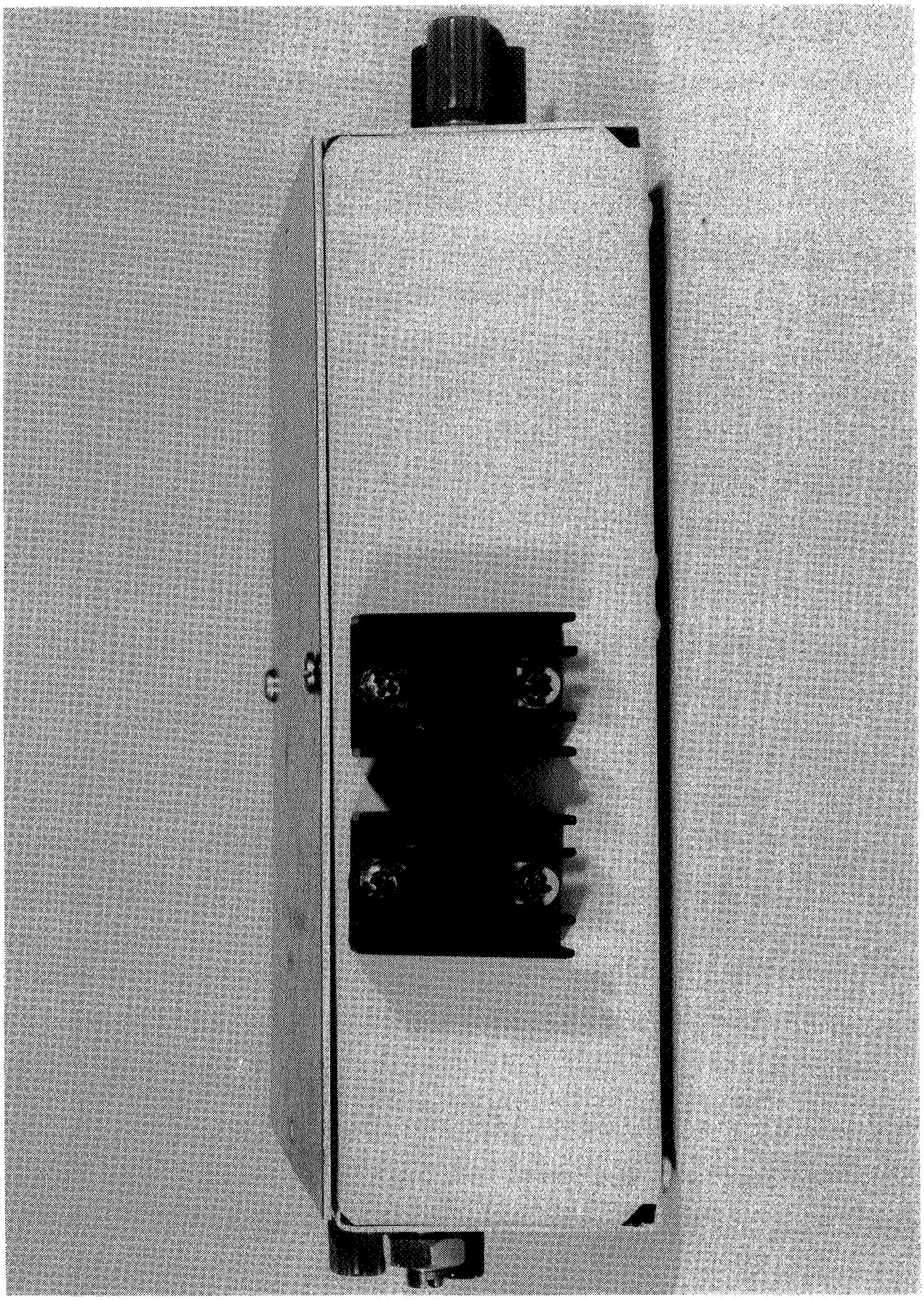


FIGURE 13.- Ion Pump Power Supply (Sheet 1 of 6)

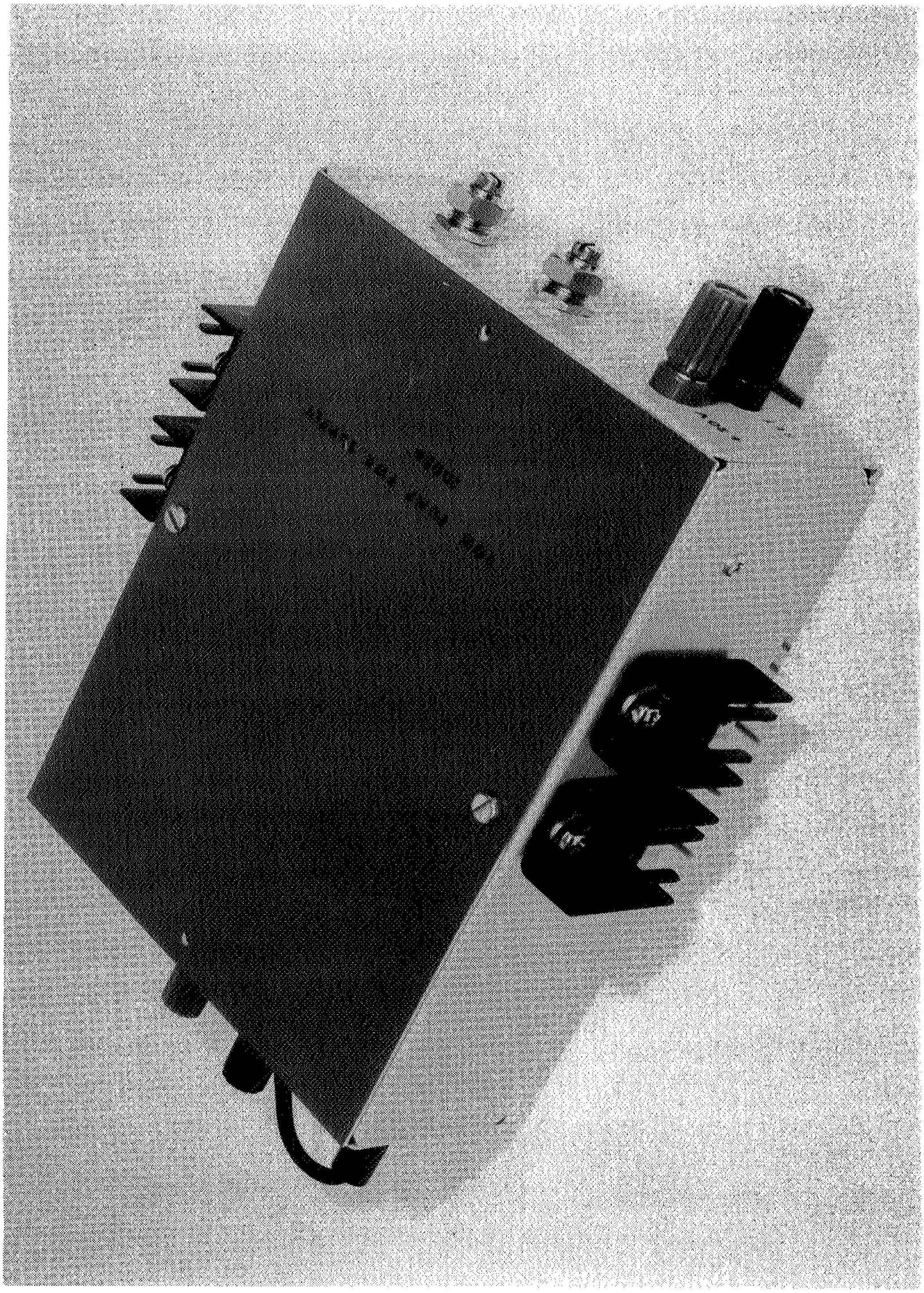


FIGURE 13.- Ion Pump Power Supply (Sheet 2 of 6)

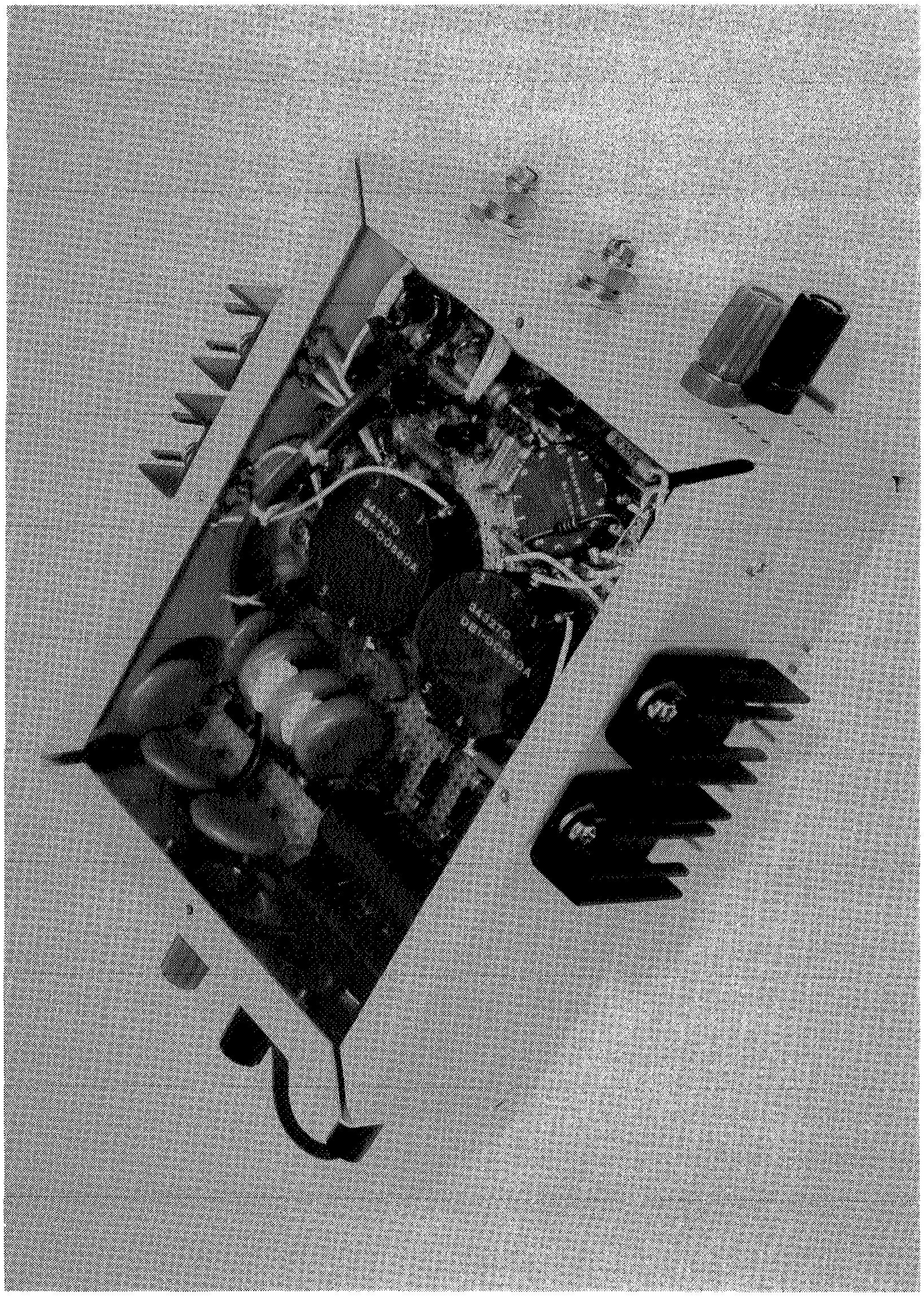


FIGURE 13.- Ion Pump Power Supply (Sheet 3 of 6)

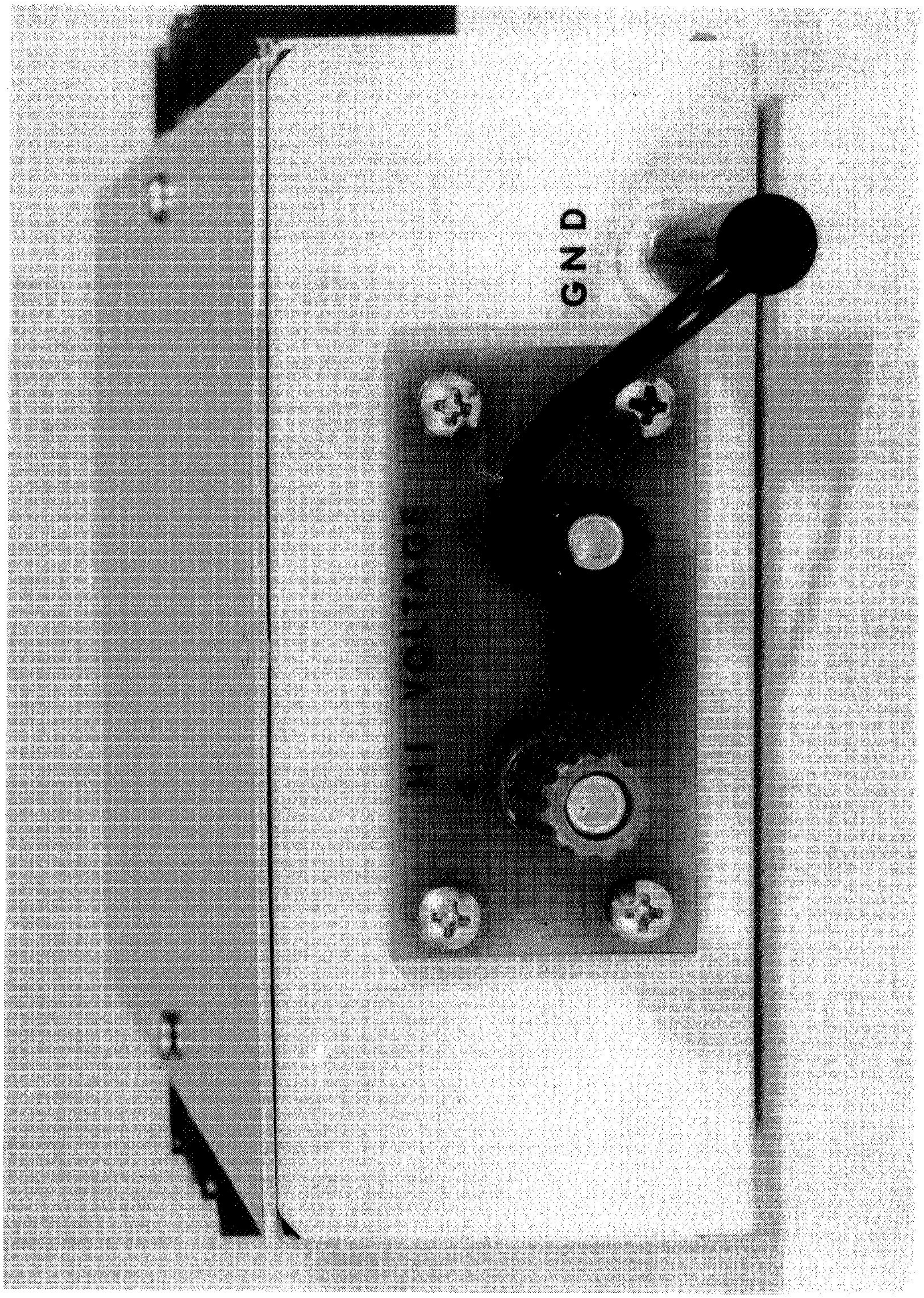


FIGURE 13.- Ion Pump Power Supply (Sheet 4 of 6)

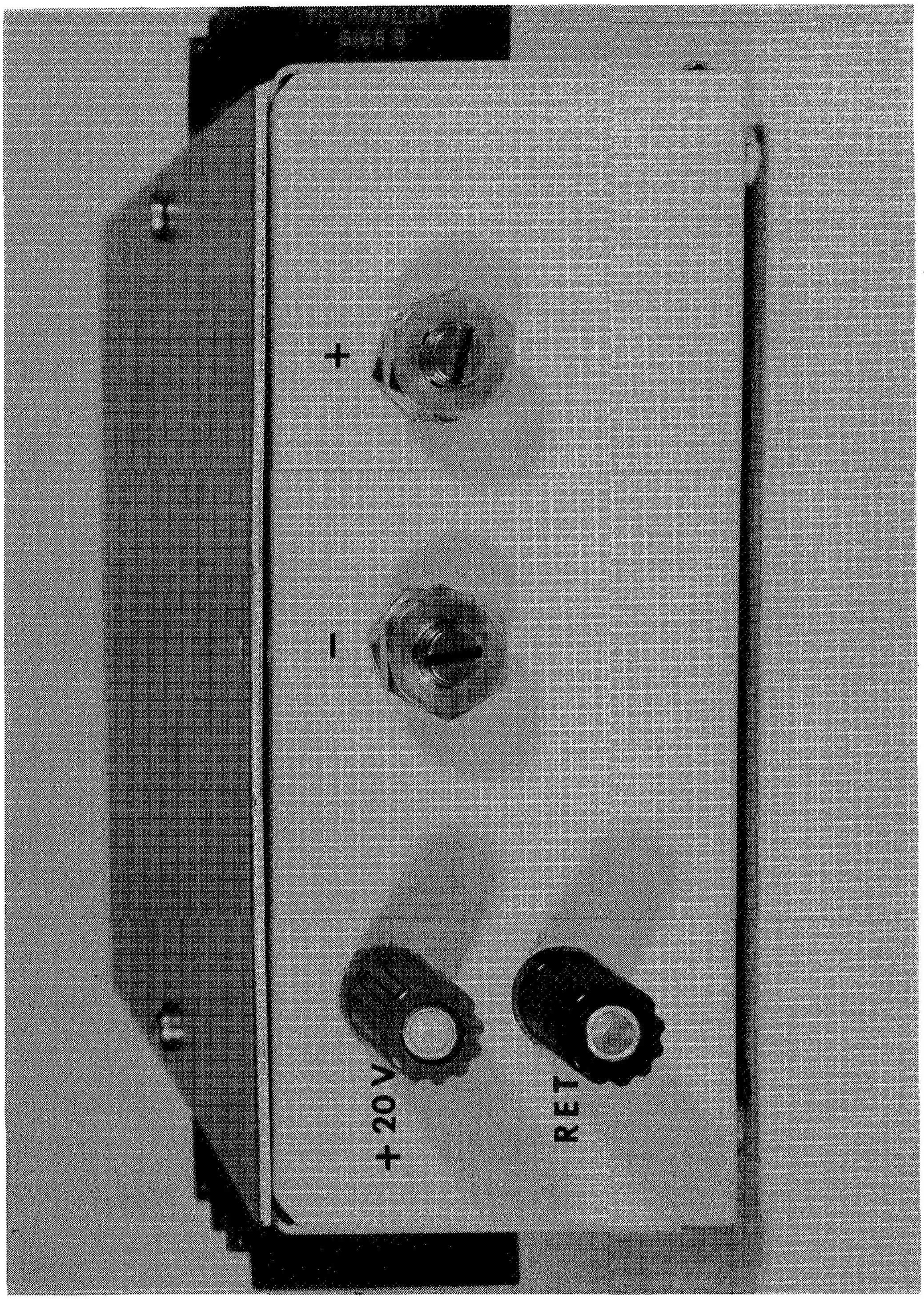


FIGURE 13.- Ion Pump Power Supply (Sheet 5 of 6)

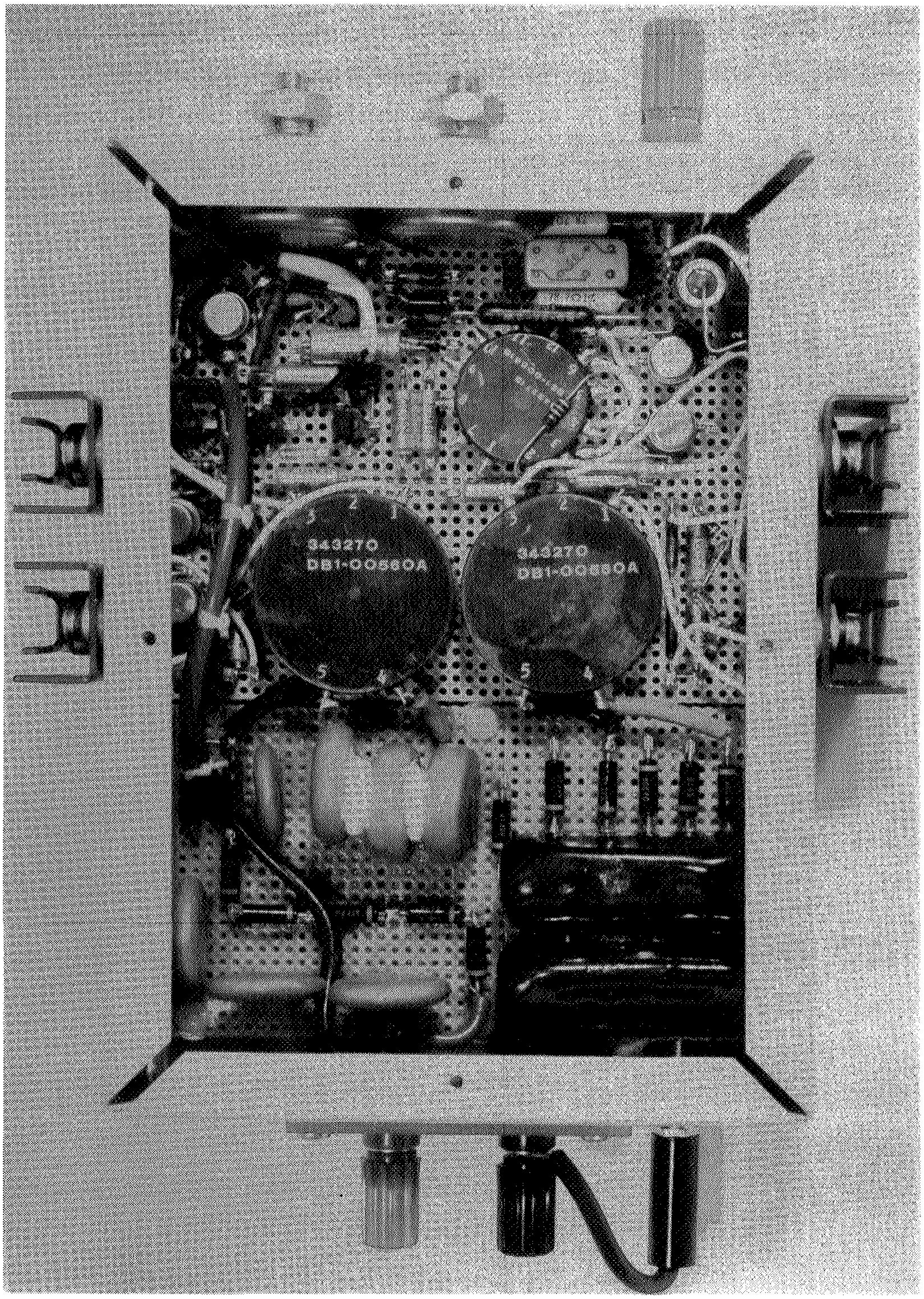


FIGURE 13.- Ion Pump Power Supply (Sheet 6 of 6)