BOOST REGULATOR DESIGN & DEVELOPMENT

TECHNICAL REPORT NO. TR-1001

FINAL REPORT

Published February, 1969

JPL Contract No. 952293

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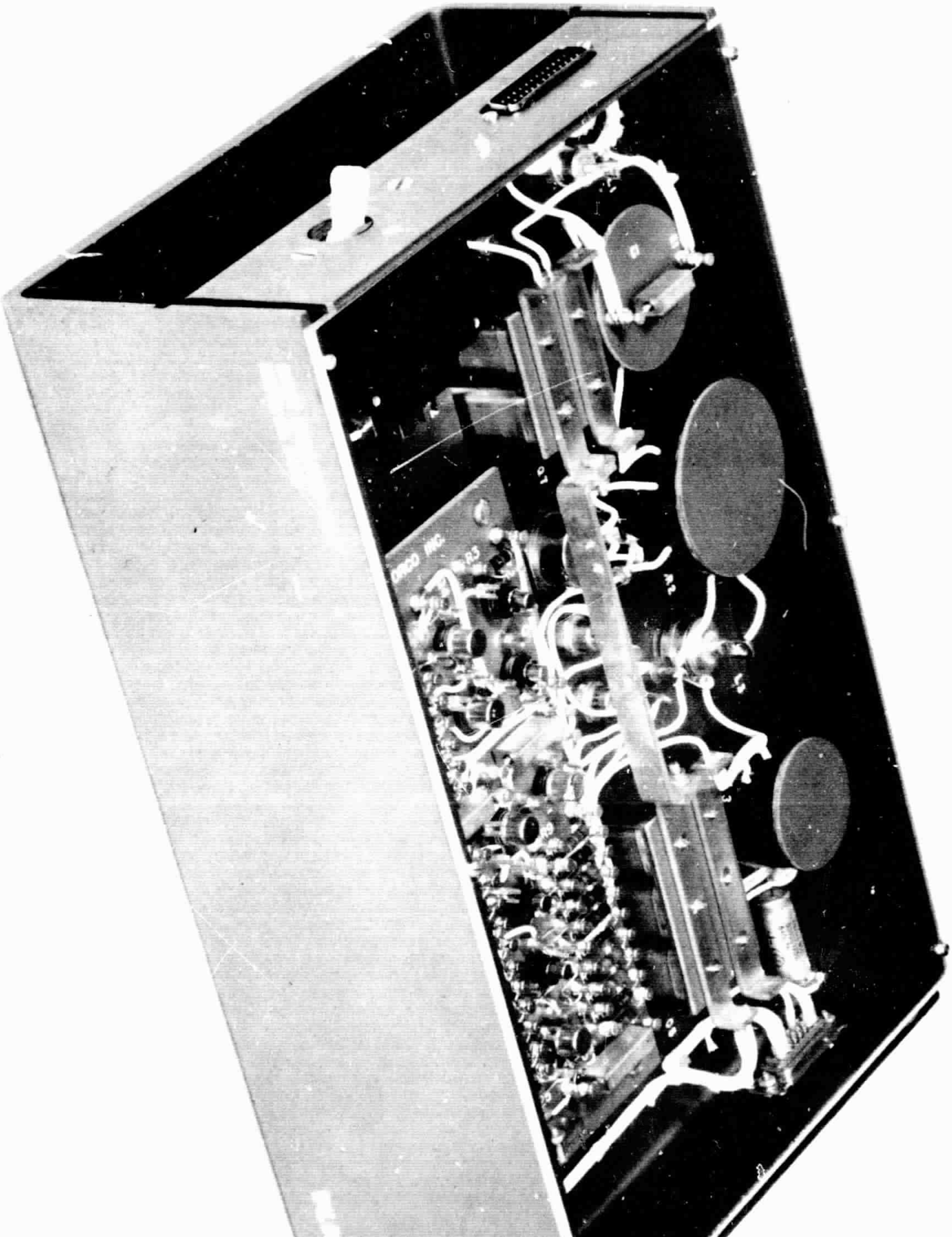

STATEMENT

OF

TECHNICAL CONTENT

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I

ABSTRACT

This report covers the theoretical and empirical criteria used in the design of a Pulse and Frequency Modulated Boost Regulator (the so-called Improved Boost Regulator or IBR).

All of the operating modes are discussed, and criteria are derived to enable the design of such regulators, given the source and load requirements. Specifically, criteria are developed for sizing the output and input capacitors, the charging inductance, and the starting and drive circuit components. Also included are specific design examples and a comparison of predicted to measured performance.

II

INTRODUCTION

A. BACKGROUND

The particular Boost Regulator (BR) design studied under this contract, was conceived as operational within the general constraints of a "Mariner-type" power subsystem. The constraints of a Mariner system are:

(1) A wide input voltage range, from 25VDC to 50VDC.

(2) High efficiency over a wide load range, e.g. low standby losses.

(3) Non-Synchronous operation.

(4) Low parts-count, less than the present (Mariner '69) design which uses 80 parts.

(5) Low weight.

An improved version of a Boost Regulator (IBR) design was achieved within the general constraints cited above. The basic IBR design was developed for the Capsule System Advanced Development (CSAD) program, by Wilorco, under sub-contract to the Jet Propulsion Laboratory (1). The design studied under this contract (2) is similar to the CSAD design in all major respects, however some refinements have been added.

During this effort particular emphasis was placed on achieving a design capable of operation to no-load at high efficiency.

B. PURPOSE

The purpose of this sub-contract was:

(1) To verify the CSAD design technique at several higher power levels.

(2) To codify and systematize as much of the design technique as possible.

(3) To design and construct three breadboard models of the IBR at the 100 watt, 200 watt, and 400 watt load power level.

(4) To compare the results of tests on the models with the results predicted by the design equations.

Every effort has been made to maintain a consistent set of units within the body of this report. The following units are used exclusively:

(a) Inductance, microhenries

(b) Capacitance, microfarads

(c) Resistance, ohms

(d) Time, microseconds

(e) Voltage, volts or millivolts

(f) Current, amperes or milliamperes

Although no sterilization or high-impact requirements were imposed on this program, we have been aware of the possibility that an IBR may be utilized in a planetary-landing mission. Thus all of the functional component parts used in this design are amenable to severe temperature and mechanical shock environments.

C. REFERENCES

(1) Capsule System Advanced Development (CSAD) Sub-Contract #952003

(2) Improved Boost Regulator Development (IBR) Contract #952293

(3) NASA Technical Report CR-898

D. ACKNOWLEDGMENT

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III

TECHNICAL DISCUSSION

A. THEORY OF OPERATION

1. General

The circuit described herein is a free-running, blocking oscillator, whose power output capability is controlled by controlling the amount of input energy absorbed each half-cycle. We will first describe the operation of the open-loop version of the oscillator, and then discuss the closed-loop regulated version.

For positive starting, a separate starting circuit must be incorporated, and to conserve power, our starting circuit turns itself OFF after the output reaches its nominal level of operation. The starting circuit, although interesting, will not be treated as a part of the basic oscillator circuit, but will be discussed in detail further in this section.

Figure 1 shows the schematic diagram of the open-loop oscillator at its simplest.

With Q1 turned ON, current starts to flow in N1 in accordance with the relation:

$$I = \int i \, dt = \frac{1}{L} \int_0^T e \, dt$$

Performing the indicated integration and substituting $E_s = e$, and assuming that the transistor's saturation voltage is zero, we obtain:

$$I = \frac{E_s}{L} T$$

Thus the current in transistor Q1 increases linearly with time.

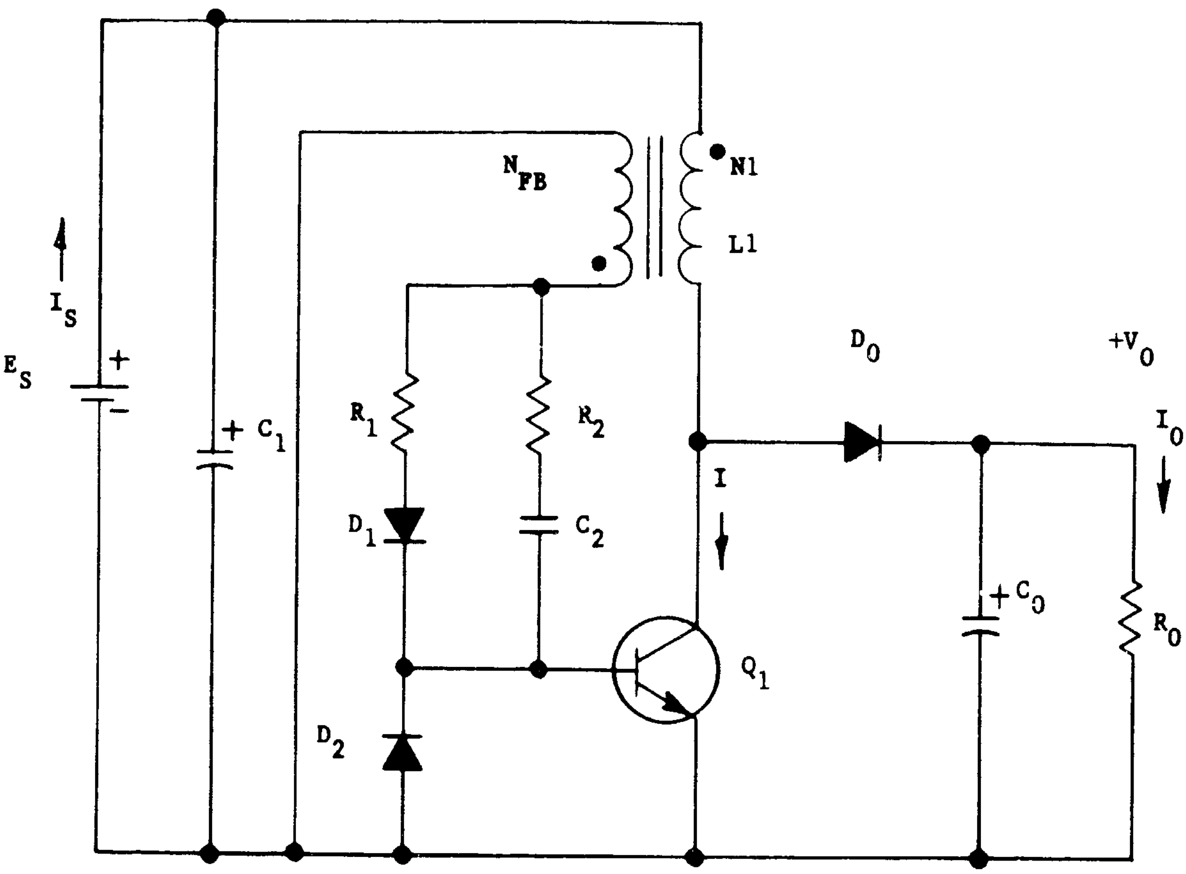


FIGURE 1. BASIC BOOST-OSCILLATOR.

As current (I) flows in the primary N1, a voltage is generated across the secondary, or feedback, winding NFB. This feedback voltage continues the action performed by the starting cirucit and maintains Q1 saturated. At some pre-determined peak-current level ($I = I_p$) the circuit switches Q1 OFF and CR3 starts to conduct. At this instant, all of the energy previously stored in the primary inductance (LI) is now started down its discharge path into the output capacitor, where it is stored and delivered into the load, where it is dissipated. The charge-discharge cycle is then repeated by virtue of the base-circuit elements. The action of the base-circuit elements is described later. During the inductor's (L1) charging period, energy stored in Co is provided to the load. The inductor current wave-form is shown in Figure 2.

To see just how all of the required actions take place in the real world, we refer to Figure 3. Figure 3 shows the circuit elements required to perform all of the operating cycle, except for the starting circuitry:

With the power-switching transistor (Q1) initially turned ON, a feedback voltage appears across the feedback winding NFB with the polarity shown in Figure 3a. This feedback voltage drives an instantaneous current through the base-drive capacitor (C2) and also drives a continuous current through the control circuit to maintain Q1 in an ON condition. With Q1 ON, the current through N1 continues to increase in a linear fashion. This interval is defined as the "charge period"; when energy is stored in L1.

$t_1 - t_0$ = Starting circuit delay, present on 1st cycle only.

$t_2 - t_1$ = Charge Time, Tchg.

$t_3 - t_2$ = Discharge Time, Tdis.

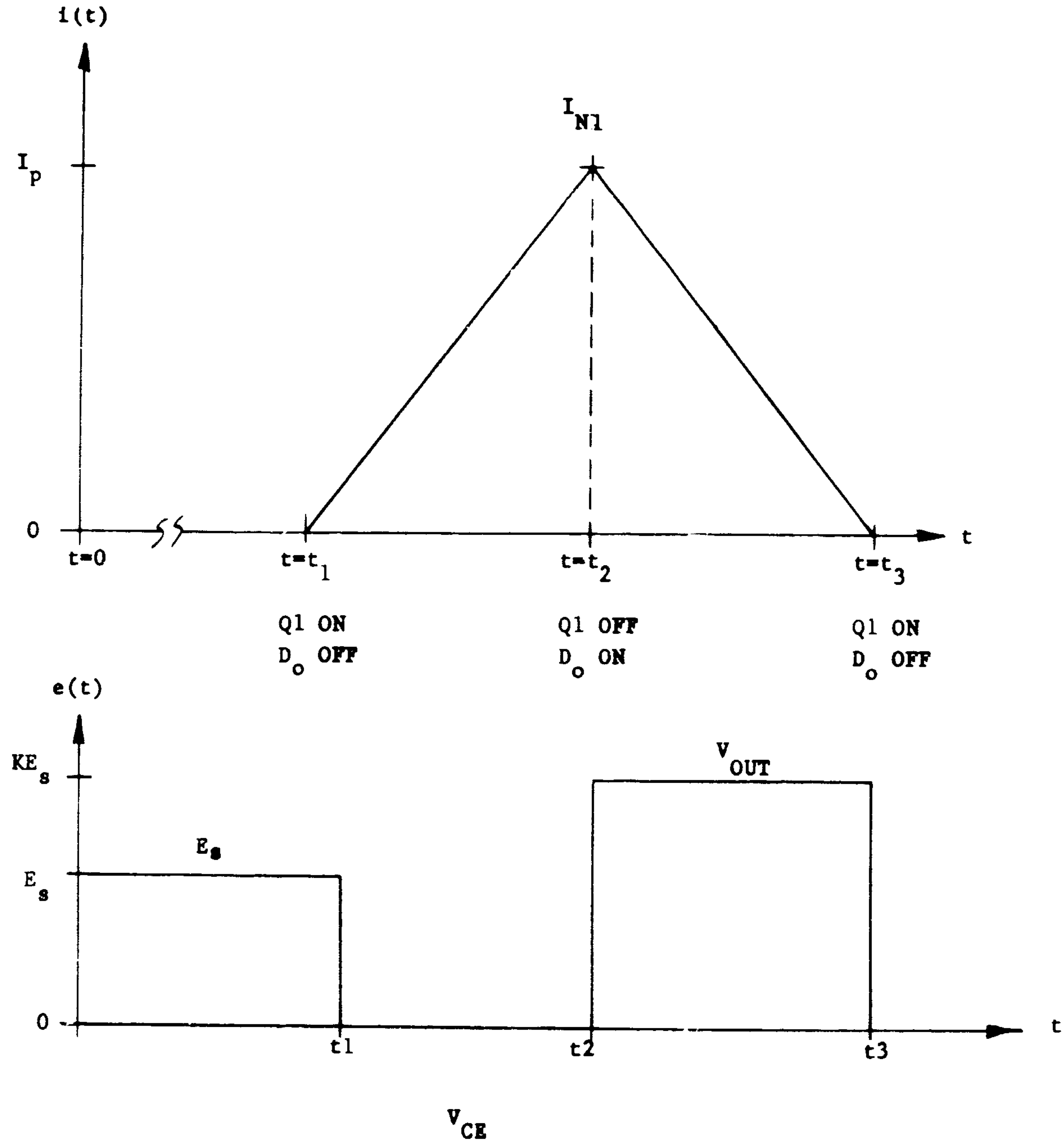


FIGURE 2. PRINCIPAL WAVEFORMS AT TURN-ON

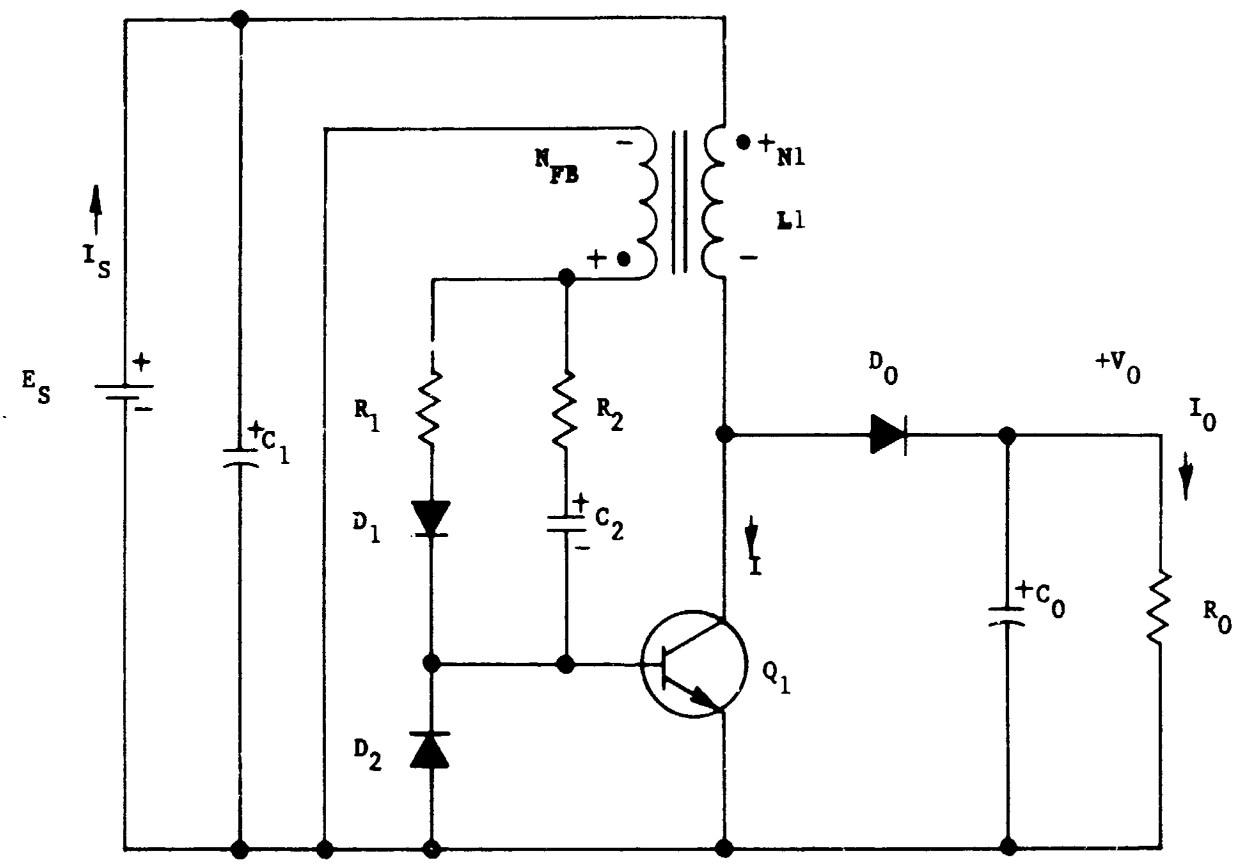


FIGURE 3a. POLARITIES DURING CHARGE INTERVAL.

Eventually, as L1 charges, the current through QICE becomes sufficiently high so that the base current (limited by the control circuit resistance, R1) is no longer able to maintain Q1 saturated.

Thus Q1 starts to turn OFF and thereby starts to sustain voltage across itself. This reduces the voltage available to N1, and thereby reduces the feedback voltage. As the feedback voltage is reduced the base-drive current is reduced, and Q1 is forced further into its OFF state. This sequence is regenerative and Q1 is completely turned OFF in very rapid fashion.

With Q1 entirely OFF, no current flows through it; however L1 was charged to a peak-current value, Ip. This peak-current must flow somewhere, and fortunately a path is available through diode Do, capacitor Co, and load Ro. However, the sense of the voltage across N1 has now reversed, because L1 has become a source of energy for Co and Ro. Thus the current through N1 during this discharge period is decreasing, and therefore the feedback voltage across NFB has a reversed polarity as shown in Figure 3b.

As shown in Figure 3b, the feedback voltage is now series-additive with the voltage previously stored on capacitor C2. Thus a current is driven in the reverse direction through diode D2 and base-region capacitance CB, limited only by the base resistor R2. This current performs its main function by discharging C2 and CB, and then charging C2 to a polarity as shown in Figure 3c, in expectation of the next charging cycle.

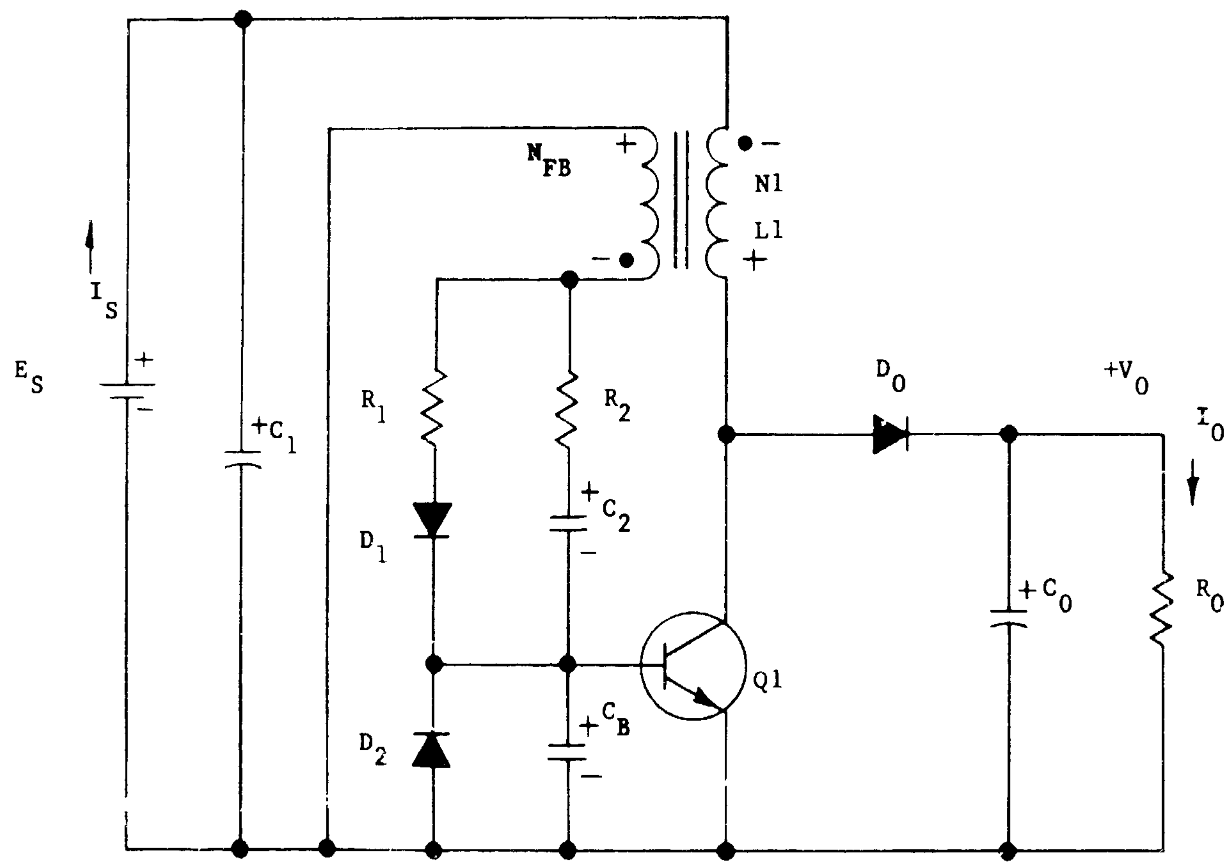


FIGURE 3b. POLARITIES DURING TRANSISTOR Q1 TURN-OFF TIME, SHOWING THE BASE-JUNCTION CAPACITANCE (C_B).

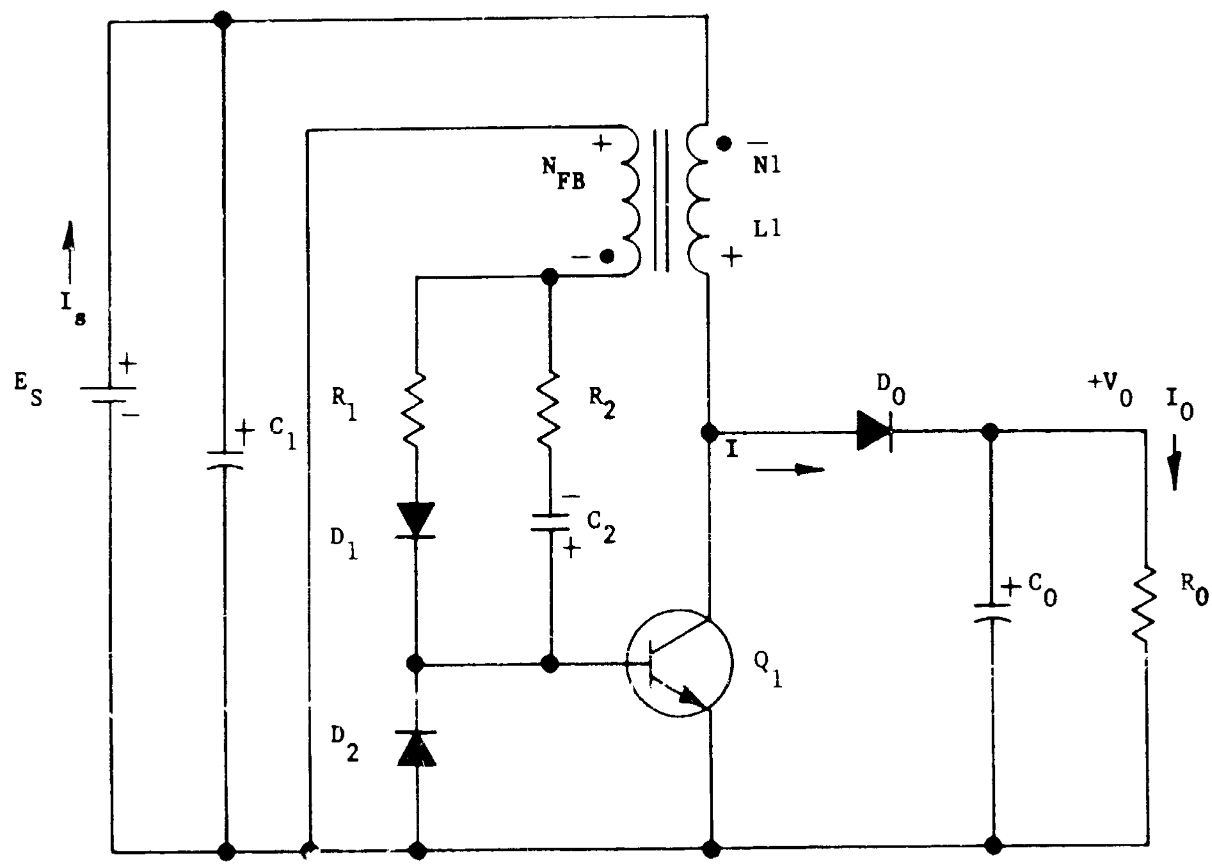


FIGURE 3C. POLARITIES DURING THE DISCHARGE INTERVAL.

When the primary inductance L1 discharges into the output circuit, the current through N1 eventually reaches zero. Thus the feedback voltage goes to zero and capacitor C2 can discharge through QICE to turn Q1 ON and to start the next charge period.

As noted, the charge period and discharge period are independent and this fact leads to the method of derivation of the required circuit components and operating parameters. (See Appendixes for details of all derivations.)

To regulate the output properly implies that the average (DC) voltage on capacitor Co is constant and that the ripple (PPAC) voltage on Co is relatively small.

DC regulation is achieved by sensing the output voltage, comparing this value to a reference voltage, amplifying the difference (error) voltage, and thereby controlling the amount of DC current delivered to the base of the power-switching transistor Q1 from feedback winding NFB (see Figure4). By reducing the base-drive current to Q1 (by increasing R1), we immediately limit the peak-current, Ip, during the charge half-cycle, decreasing the energy stored and subsequently decreasing the energy deliverable to the load.

The same basic equations may be written to describe both Oscillator and Regulator operation:

(a) For the charge interval, Q1 conducting:

$$E_s = L_1 \frac{di}{dt} \qquad \text{-------------------------} (1)$$

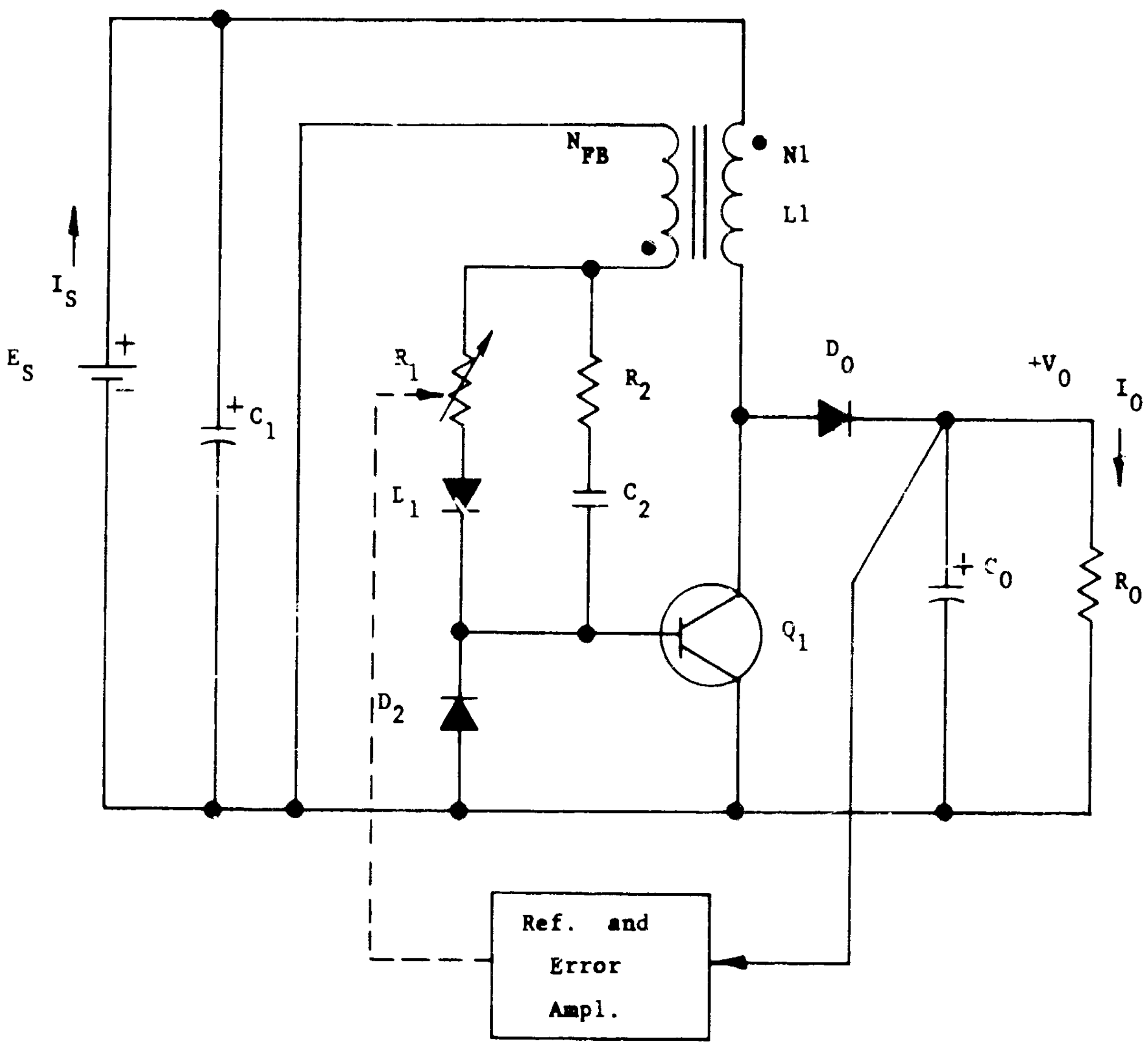


FIGURE 4. BASIC BOOST-REGULATOR.

(b) For the discharge interval, Q1 non-conducting:

$$V_o - E_s = L_1 \frac{di}{dt} \qquad \text{---------------------- (2)}$$

Substituting the definitions given above into equations (1) and (2) we obtain general expressions for the charge and discharge times:

$$T_{chg} = \frac{L_1 I_P}{E_s} \qquad \text{---------------------- (3)}$$

$$T_{dis} = \frac{L_1 I_P}{V_o - E_s} \qquad \text{---------------------- (4)}$$

Adding equations (3) and (4) yields an expression for the total period:

$$T = L_1 I_P \left[\frac{1}{E_s} + \frac{1}{V_o - E_s} \right] \qquad \text{------------- (5)}$$

Equation (5) is the basic relation governing the operation of the Boost Oscillator or Boost Regulator. From equation (5) and other considerations, we will derive design criteria for both open-loop (Oscillator) and closed-loop (Regulator) operation.

By substituting the geometrically-derived expression for the peak current:

$$I_p = 2 I_o \frac{T}{T_{dis}}$$

or $$I_p = 2I_o(1-\delta)^{-1} \qquad \text{--------------- (6)}$$

where $\delta = \dfrac{T_{chg}}{T}$ = duty-cycle,

into equation (5) we obtain;

$$T = 2L_1 I_o (1-\delta)^{-1}\left[\frac{1}{E_s} + \frac{1}{V_o - E_s}\right] \qquad \text{-------- (7)}$$

Now substituting the definition of load current;

$$I_o = \frac{V_o}{R_o} \qquad \text{---------------------------- (8)}$$

into equation (7) yields;

$$T = \frac{2L_1}{R_o}(1-\delta)^{-1}\left[\frac{1}{E_s} + \frac{1}{V_o - E_s}\right] \qquad \text{-------- (9)}$$

Finally by normalizing the input voltage with respect to the output voltage, and solving for the inductance, we obtain;

$$L_1 = \frac{R_o T}{2}(1-\delta)u(1-u) \qquad \text{-------- (10)}$$

where $u = \dfrac{E_s}{V_o}$, the normalized input voltage.

In the open-loop case, the ratio "u" is constant in any given design, but is dependent on duty-cycle in the following simple fashion; $u + \delta = 1$ ---------------------- (11)

In the closed-loop case, the ratio "u" varies over the range;

$$\frac{E_{s1}}{V_o} \leq u \leq \frac{E_{s2}}{V_o} ,$$

and the duty-cycle as well as the period vary with input voltage.

2. Oscillator Operation

Because the normalized input voltage is constant in this mode of operation, we rewrite equation (10) in terms of duty-cycle only and thereby delete the normalized input voltage from the expression for inductance. This yields:

$$L_1 = \frac{R_o T}{2}\delta(1-\delta) \qquad \text{-------------- (12)}$$

Thus we see that once the load (Ro), operating frequency (periodT), and duty-cycle (δ) are selected, the charge inductance is uniquely determined by equation (12). Conversely, once the charge inductance is known, the period and duty-cycle are seen to be independent of the input voltage. However in any particular design, the duty-cycle will be controlled by the feedback circuit parameters, and as the duty-cycle is varied, both the duty-cycle (and thereby the output voltage) and the period will vary.

Once L1 is calculated, equation (12) may be used to find the period at any duty-cycle. The minimum period for the simple oscillator is obtained when $\delta = 0.333$.

The output voltage is related to the duty-cycle as given below:

$$V_o = E_s(1-\delta)^{-1} \qquad \text{---------------- (13)}$$

For convenience we also define the "boost ratio" as:

$$\frac{V_o}{E_s} = (1-\delta)^{-1} = u^{-1} \qquad \text{---------------- (14)}$$

It remains then to relate the duty-cycle to the base-drive parameters to complete the steady-state description of the oscillator.

By definition;

$$I_P = \beta I_B \qquad \text{-------------------------- (15)}$$

where β is the current-gain of the switching transistor, and I_B is the base-current.

In terms of the feedback (base) circuit parameters, we obtain;

$$I_P = \beta \frac{V_{FB}}{R_1} = \beta \frac{N_{FB}}{N_1} \frac{E_S}{R_1} \qquad \text{------------- (16)}$$

Substituting the definition of Ip, equation (6), into equation (16) we obtain;

$$\frac{2V_o}{R_o}(1-\delta)^{-1} = \beta \frac{N_{FB}}{N_1} \frac{E_S}{R_1} \qquad \text{------------- (17)}$$

Rewriting equation (17) and substituting equation (14) yields;

$$(1-\delta)^2 \beta \frac{N_{FB}}{N_1} \frac{R_o}{R_1} = 2 \qquad \text{------------- (18)}$$

Solving equation (18) for the duty-cycle, gives;

$$\delta = 1 - \left(\frac{2N_1 R_1}{\beta N_{FB} R_o}\right)^{1/2} \qquad \text{------------- (19)}$$

From the definition of duty-cycle, we have the following restriction on its range;

$$0 \leq \delta \leq 1 \qquad \text{------------------ (20)}$$

This inequality implies the upper bound;

$$\frac{2N_1 R_1}{\beta N_{FB} R_o} \leq 1 \qquad \text{------------------ (21)}$$

and the fact that all of the parameters (β , N2, N1, Ro, and R1) are positive numbers implies the lower bound;

$$0 \leq \frac{2N_1 R_1}{\beta N_{FB} R_o} \qquad \text{------------------ (22)}$$

Taking the upper limit yields;

$$R_1 \leq \beta \frac{R_o}{2} \frac{N_{FB}}{N_1} \qquad \text{------------------ (23)}$$

The relationship between parameters expressed by inequality (23) above gives the allowable range of the theoretical base-drive resistance for proper operation.

It is also convenient to calculate the "boost-ratio" for the oscillator in terms of the circuit parameters;

$$\text{Boost-Ratio} = \frac{V_o}{E_S} = \left(\frac{\beta R_o N_{FB}}{2R_1 N_1}\right)^{1/2} \qquad \text{------- (24)}$$

3. Regulator Operation

The same expression, equation (10), may be used to calculate the required value of charge inductance (L1) for the Regulator-mode of operation, as was used to calculate L1 in the Oscillator-mode of operation. It is also desireable to calculate the operating frequency at other input voltages. This is accomplished by simply rewriting equation (10) in terms of the normalized input voltage and solving for the period;

$$T = \frac{2L_1}{R_o}\mu^{-2}(1-\mu)^{-1} \qquad \text{--------------- (25)}$$

The minimum period Tx occurs at a value of $\mu = \mu_x = 0.626$.

It is preferable to maintain a duty-cycle no greater than 50% to optimize the transfer of energy from input to output, and to minimize filter requirements.

To accomplish this it is necessary, where $V_o > 2E_s$, to provide an overwind on the charging inductor.

Figure 5 shows the circuit configuration with an overwind. To calculate the value of charge inductance required by this configuration necessitates some recalculation. The detail calculation is given in appendix A, and the result is given below;

$$L_1 = \frac{R_o T}{2}\,\frac{\mu^2\delta^2}{1-\mu} \qquad \text{-------------------- (26)}$$

Once the charge inductance is calculated from equation (26) at the selected values of load, frequency, and duty-cycle at minimum input voltage, we can calculate the overwind, frequency and duty-cycle at any value of input voltage.

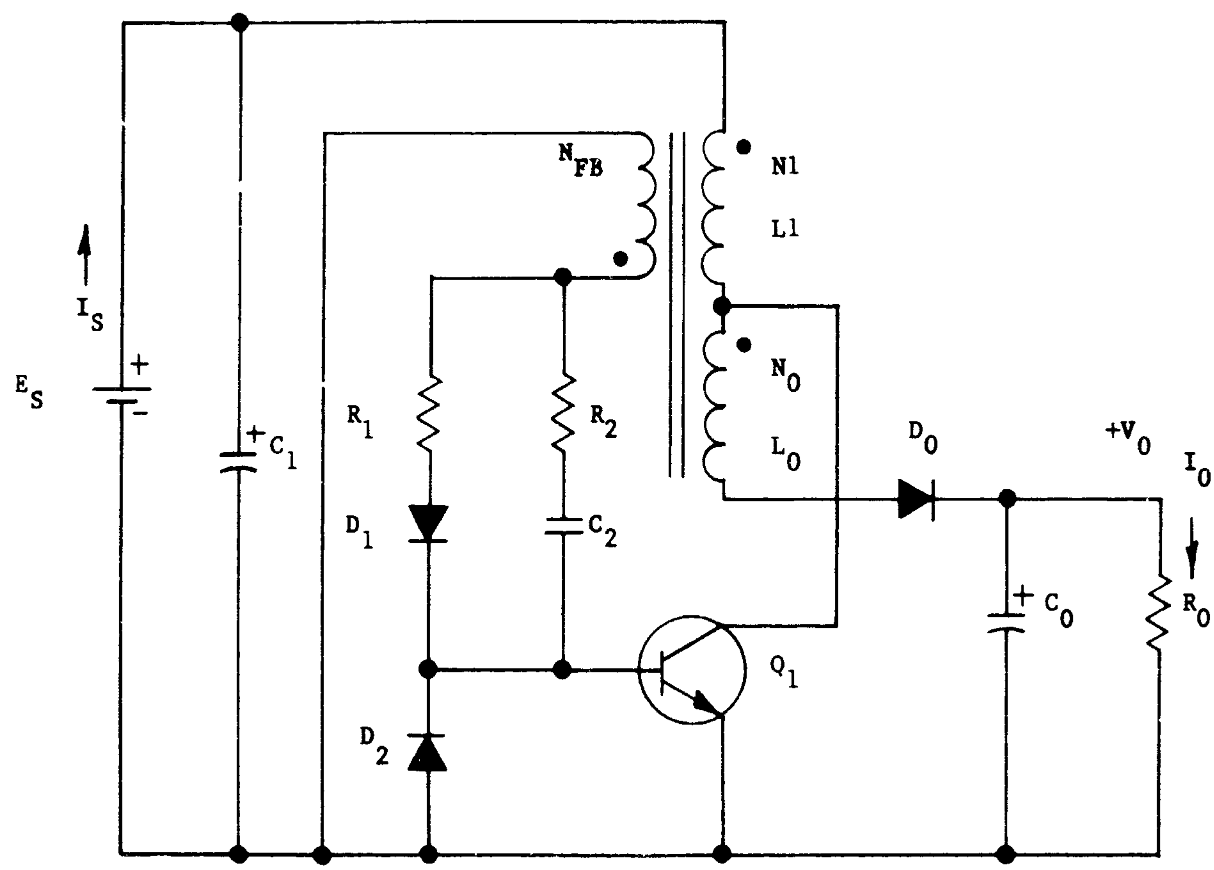


FIGURE 5. BOOST-OSCILLATOR WITH AN OVERWIND.

All of the formulas required to perform the calculations are listed below:

$$L_1 = \left(\frac{T_1 R_{o,min}}{2}\right)\left(\frac{\mu_1^2 \delta_1^2}{1-\mu_1}\right) \qquad (27)$$

$$n = \left(\frac{1}{\mu_1} - 1\right)\left(\frac{1}{\delta_1} - 1\right) \qquad (28)$$

$$T = \left(\frac{2L_1}{R_o}\right)\left(\frac{1-\mu}{\mu^2}\right)\left(1 + \frac{\mu n}{1-\mu}\right)^2 \qquad (29)$$

$$\delta = \left(1 + \frac{\mu n}{1-\mu}\right)^{-1} \qquad (30)$$

$$R_o \equiv V_o / I_o \qquad (31)$$

The minimum period (maximum oscillator frequency) is calculated in Appendix B, and exists at a value of $\mu = \mu_x$ and;

$$\mu_x = \frac{(8n+1)^{1/2} - 3}{2(n-1)} \qquad (32)$$

As is evident from expression (29), the period is inversely proportional to the load resistance. As the load current decreases, the operating frequency rises, and one could expect an infinitely-high frequency at zero load current. Luckily, the switching losses rise rather dramatically as the load current reduces below about 10%, and these internal losses prevent any runaway condition from damaging the switching transistor.

Graphic comparisons of the results attained empirically to those calculated from equations (29), (30), and (32) are shown in Figures 6 and 7, respectively. As is evident, correlation is excellent and the design equations are deemed useful and accurate.

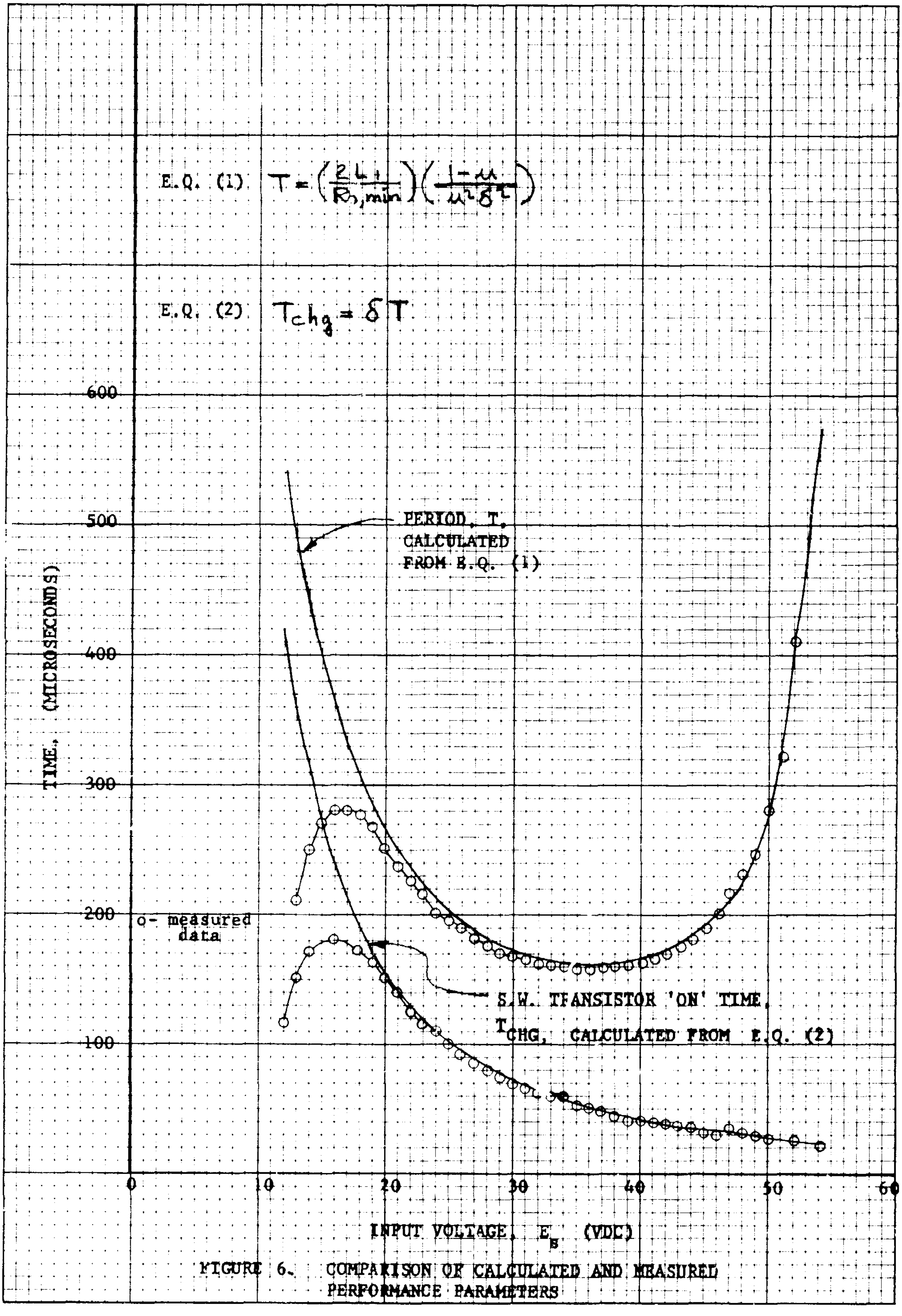


FIGURE 6. COMPARISON OF CALCULATED AND MEASURED PERFORMANCE PARAMETERS

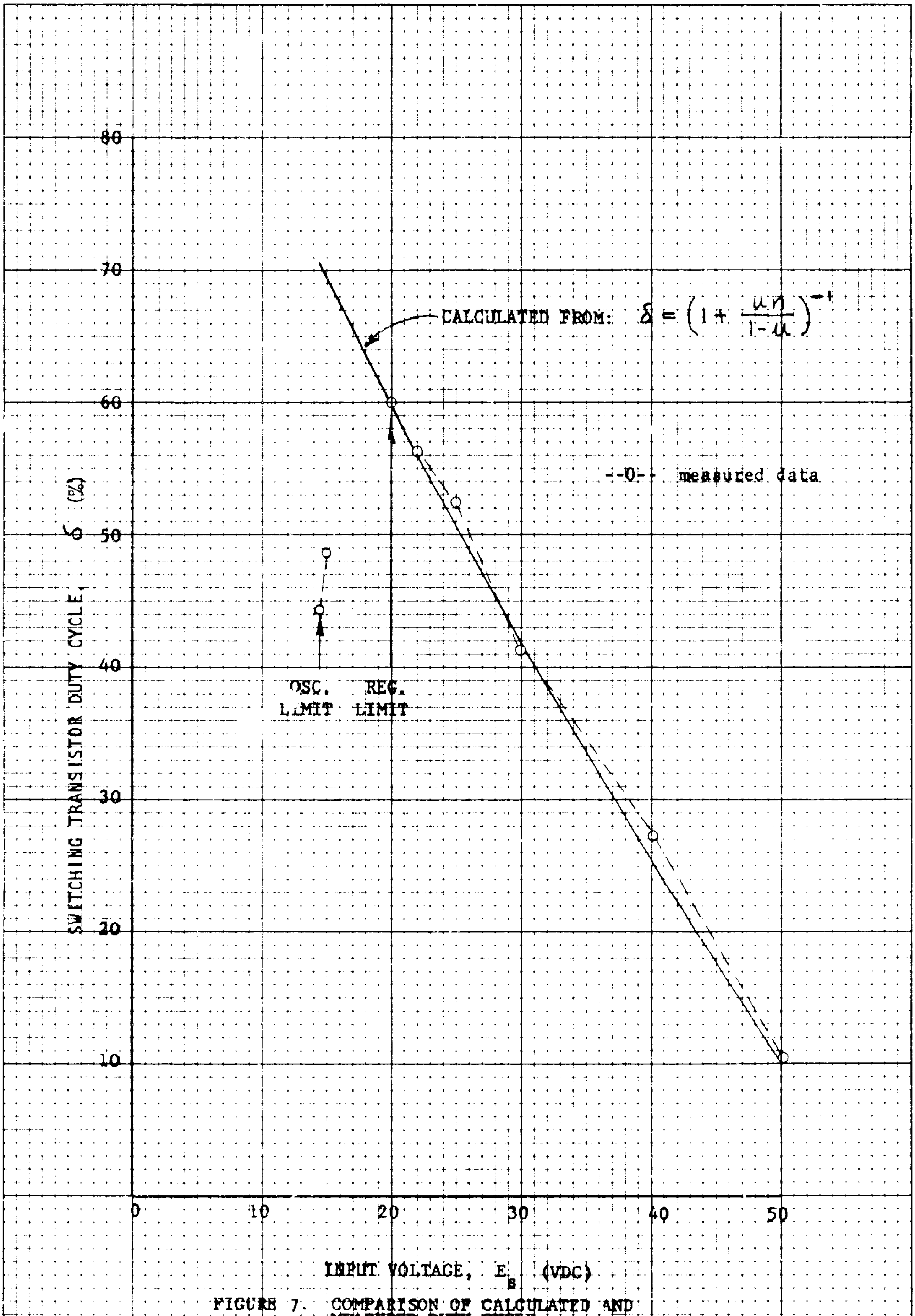


FIGURE 7. COMPARISON OF CALCULATED AND MEASURED DUTY-CYCLE.

4. <u>Efficiency Consideration</u>

To take into account all of the individual loss terms in the circuit is a long and tedious process. But, if we lump all of the losses into two components, we can effectively account for the distributed losses and still write our descriptive equations as simply as before.

Let us assume that the transistor switch (Q1) and the output rectifier (Do) both have a constant drop of one volt regardless of current level and temperature. Then equations (1) and (2) may be rewritten to include the effect of these voltage drops as;

$$E_s - 1 = L_1 \frac{di}{dt} \quad \text{------------------ (33)}$$

$$V_o + 1 - E_s = (L_1 + L_o)\frac{di}{dt} \quad \text{------------------ (34)}$$

From equations (33) and (34) we may derive an expression for the period, once again;

$$T = \frac{2nL_1 I_o}{1-\delta}\left[\frac{1}{E_s-1} + \frac{n}{V_o-(E_s-1)}\right] \quad \text{---- (35)}$$

The period as given by equation (35) is of the same form as that given by equation (5). Thus all previously derived expressions are valid, provided only that we replace every "Es" term by "Es-1". Or, we may simply redefine the normalized input voltage as:

$$\mu \equiv \frac{E_s - 1}{V_o} \quad \text{-------------------------- (36)}$$

In effect, the definition (36) implies losses of about 8% at an input of 25VDC and losses of about 4% at an input of 50VDC. As will be seen below, these numbers present fair approximations to their measured counterparts.

A comparison of the actual, measured efficiency to an assumed, constant efficiency is shown in Figure 8. The figure shows the input current as a function of the input voltage under constant rated load conditions for the 100W IBR. We believe that the differences between the measured and calculated values are minor, and that a constant efficiency is a fair approximation.

However by defining the normalized input voltage as in expression (36) above, a much closer approxination is obtained.

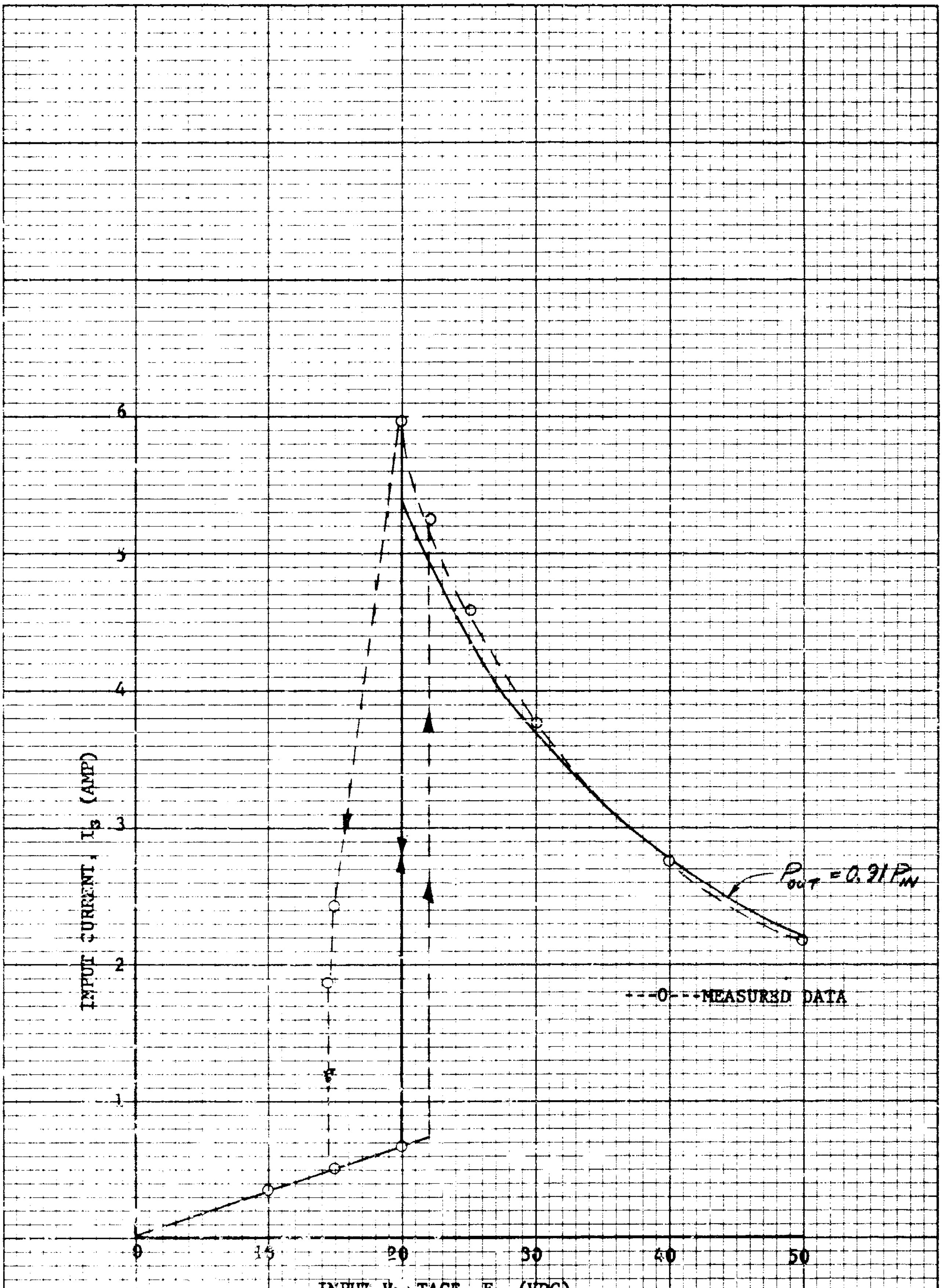


FIGURE [illegible] COMPARISON OF MEASURED TO CALCULATED INPUT VOLT-AMPERE CHARACTERISTIC FOR THE 100W IBR.

5. Starting Considerations

Initial starting is accomplished by means of a simple unijunction (UJT) transistor relaxation oscillator. The "starting circuit delay" shown in Figure 2 is equal to the period of the UJT oscillator. This period is not critical, and may be set as desired in the "millisecond" or "second" time ranges. Once the circuit has been turned ON for the first time (i.e. Q1 saturated), and the proper voltage is present at the output terminals, the differential error-amplifier produces a signal to inhibit further oscillation of the UJT.

To minimize standby losses, the "inhibit" in our design actually opens the DC power line to the oscillator. If the output voltage drops below its regulated value, the UJT oscillator automatically resumes operation.

For the circuit to start, transistor Q1 must be held ON for a sufficiently long time so that the voltage on the charge inductor's feedback winding has the time to build to the full feedback voltage. This means that capacitor C2 must be large enough to continue to conduct for the full build-up time.

Resistor R2 should be sized to limit the base-drive current to the peak-current value divided by the linear current-gain of Q1.

To calculate the voltage build-up time-constant, we first calculate the series limiting resistor R2:

$$R_2 = \frac{V_{FB,min} - V_{BE,sat}}{(I_{P1}/\beta)} \qquad \text{-------------- (37)}$$

The effective inductance of the feedback winding is:

$$L_{FB} = L_1 \left(\frac{N_{FB}}{N_1} \right)^2 \qquad \text{--------------------- (38)}$$

Then the feedback voltage build-up time constant is LFB/R2; the RC time-constant in the base circuit should be much larger that this;

$$R_2 C_2 \gg \frac{L_{FB}}{R_2} \approx \frac{4 L_{FB}}{R_2}$$

$$\text{and} \quad C_2 \geqq \frac{4 L_1}{R_2^2} \left(\frac{N_{FB}}{N_1} \right)^2 \qquad \text{------------- (39)}$$

6. <u>Charging Inductor Design</u>

In sizing the charge inductor, we must know the following:

(a) Inductance value,

(b) DC current level,

(c) RMS current level, and

(d) AC voltage and frequency.

The inductance value is calculable from the equations previously given. The DC current level is determined by the load, input voltage, and efficiency. The RMS currents are determined by virtue of the DC currents and the circuit wave-forms. These currents determine the inductor's wire size. The AC voltage is calculable from the input voltage swing, but is generally not a determining factor in sizing the charge inductor.

All of the currents in the primary and overwind sections of the charge inductor are tabulated below:

	INDUCTOR WINDING	CURRENT IN TERMS OF I_o		
		PEAK	AVERAGE	RMS
	PRIMARY, N1 (Ip1)	$\frac{2n}{1-\delta}$	$1+\frac{n\delta}{1-\delta}$	$\frac{1.152}{(1-\delta)^{1/2}}\left(1+\frac{n^2\delta}{1-\delta}\right)^{1/2}$
	OVERWIND N_o (Ip2)	$\frac{2}{1-\delta}$	1	$\frac{1.152}{(1-\delta)^{1/2}}$
50% DUTY-CYCLE	PRIMARY	$4n$	$1+n$	$1.63\,(1+n^2)^{1/2}$
	OVERWIND	4	1	1.63
NO OVERWIND	PRIMARY N1	4	2	2.31

TABLE 1.

7. Semiconductor Discussion

The switching transistor must be specified to have current-gain at collector currents as high as the value of:

$$I_{P1} = \frac{2n}{1-\delta} I_o \qquad \text{---------------} \quad (40)$$

Although the switching transistor operates at the average current level as far as power dissipation is concerned, the gain at Ipl is important in terms of the base-drive power requirements.

Thus it is preferable to use two, or more, parallel transistors to obtain high gain, than to stretch the capabilities of a single power transistor by operating with low $\beta\ (\beta < 20)$. Some additional power is lost in base-current splitting resistors, but this is more than made up by the decrease in saturated drop and increase in β of each transistor in the parallel configuration.

The problem of paralleling transistors does not arise in the circuits considered herein until the output power level exceeds about 400 watts. At this output power level, the peak transistor current (Ipl) exceeds thirty-two amperes, and one rapidly approaches that rarified atmosphere of power transistor state-of-the-art.

Fortunately the wire size for the primary of the charge inductor becomes proportionately large. Fortunately; because by utilizing parallel strands of wire for the primary winding we arrange for automatic sharing of collector current in the paralleled power transistors.

Furthermore as the state-of-the-art is approached in fast, high current power transistors, so also is the state-of-the-art approached in fast recovery, high current power rectifiers.

Again, because of the winding requirements, the output rectifiers may be paralleled and automatically achieve current sharing. Figure 9 illustrates this connection principle.

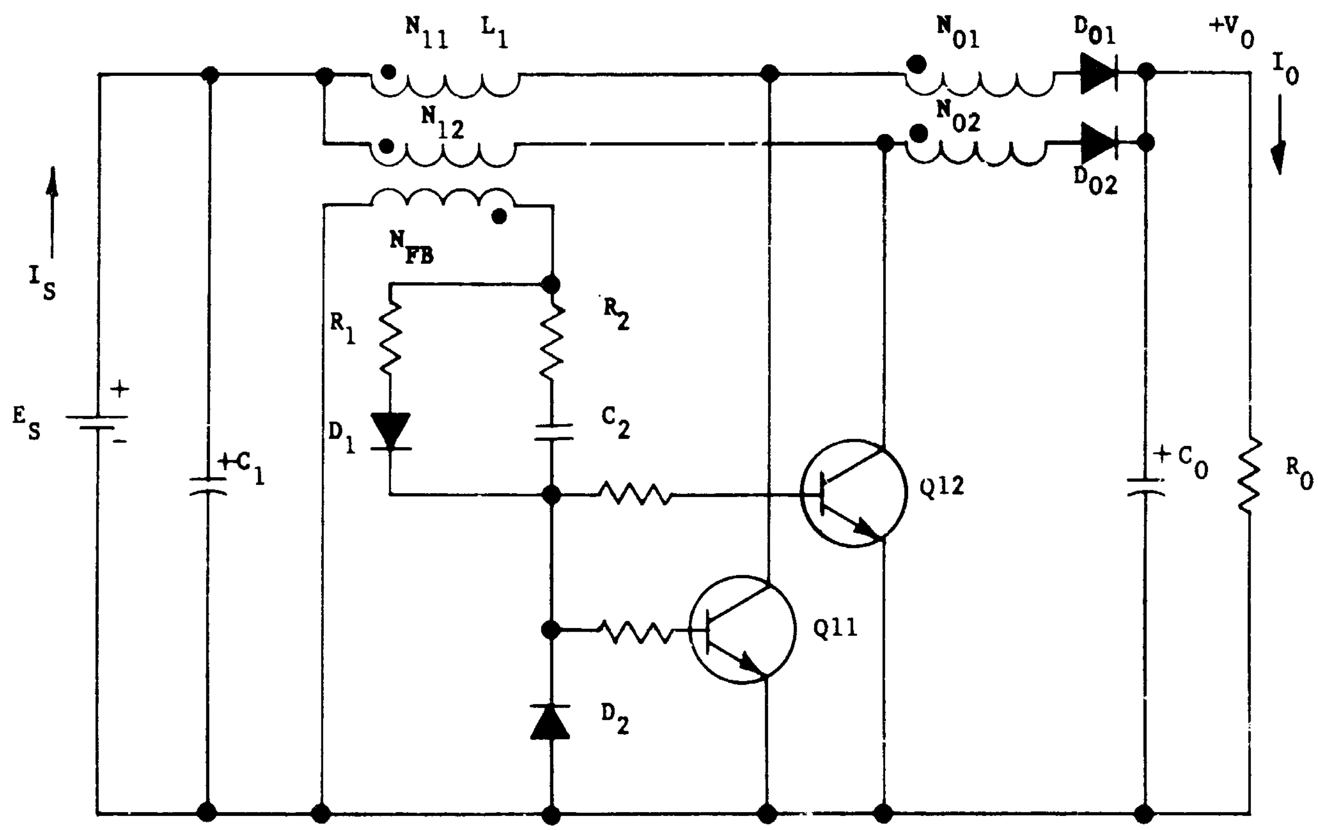


FIGURE 9. BOOST-OSCILLATOR WITH MILTIPLE POWER SECTION.

B. DESIGN CONSIDERATIONS

1. General

In the design effort reported herein, the following functional requirements were specified:

Input Voltage Range:	25 to 50	volts DC
Efficiency, Minimum:	90	%
Regulation:	± 0.5	%
Output Voltage:	56	volts DC
Output Ripple:	50	MV RMS

In specifying all of the above, nothing is implied regarding operating frequency, synchronization or circuit protection. Although the Boost Regulators to be designed are intended to provide the DC power for the 2.4KHZ inverters aboard a spacecraft, the BR's have sufficiently low ripple output so that operating frequency is not a sub-system design criterion. Synchronization is important only when it is necessary to phase specific noise bursts to minimize total DC bus noise content. Because the intended spacecraft and the BR design itself have sufficient decoupling, this also is unnecessary.

Circuit protection for the spacecraft's power subsystem is on the component (sub-assembly) level, and redundancy exists in terms of a "Main" BR and a "Standby" BR.

Thus the choice of operating frequency was made on the basis of convenience. A frequency of 5KHZ was selected as the operating frequency (T1) at the minimum input voltage (25VDC) and maximum load (100w, 200w, and 400w). At this frequency the switching losses at full load should be minimal and the weight of inductors and capacitors should also be minimal. This conclusion is based on a study by TRW Systems (Reference 3). This study actually shows an optimum (minimum-weight to maximum-efficiency ratio) at a frequency nearer to 6KHZ, but for the sake of convenience 5KHZ was finally settled-on as the design-point frequency (period T1).

2. Charge Inductor

All three charge inductances were calculated using equation (27). Certain of the parameters in equation (27) are independent of the load level. These parameters are tabulated in Table II, and are used in all of our subsequent design calculations.

The charge inductance, average, RMS, and peak currents, core numbers, turns, and wire sizes, for each inductor were calculated, using the equations developed in section A, and are summarized in Table III.

Results of the design calculations were verified in detail tests of the 100 watt BR breadboard. The period (T), the duty-cycle (δ), and the charge time (Tchg) were all calculated over the range of input voltage (25VDC to 50VDC). These parameters are plotted as the solid lines in Figure 6 and Figure 7. Actual data as measured on the 100 watt breadboard is plotted as circles on the same two figures. Correlation is adequate from 20 volts DC input to 52 volts DC input.

No degradation of performance occurred over the 20V to 52V input range except for a slight decrease of efficiency at 20 volts input.

All three breadboards exhibited the same general performance characteristics; although detailed data, other than verification of contract requriements, was recorded for the 100 watt IBR breadboard only.

Parameter	Symbol	Value	Units
Minimum Normalized Input Voltage	u_1	0.428	—
Overwind Ratio	n	1.335	—
Maximum Duty-Cycle	δ_1	0.50	—
Period at u_1	T_1	200	microseconds
Minimum Period	Tx	156	microseconds
Value of u at Tx	u_x	0.626	—
Value of δ at Tx	δ_x	0.309	—
Maximum Normalized Input Voltage	u_2	0.875	—
Minimum Duty-Cycle	δ_2	0.097	—
Period at u_2	T_2	278	microseconds

TABLE II.

Parameter	Symbol	Value for Breadboard: 100W	200W	400W	Units
Charge Inductance:	L1	250	125	63	microhenry
Overwind Current:	I1	1.78	3.57	7.14	amp avg
	Io, RMS	3.07	6.14	12.3	amp rms
	Ip2	7.14	14.3	28.6	amp peak
Primary Current:	Is	4.17	8.34	16.7	amp avg
	Is, RMS	6.56	13.2	26.4	amp rms
	Ip1	9.54	19.1	38.2	amp peak
Core part no:	---	55083-M4	55090-M4	(2)55716-M4	---
Primary Turns:	N1	55	38	21	---
Wire Size:	---	(2)#17	(4)#17	(8)#17	AWG No.
Overwind Turns:	No	17	12	7	---
Wire Size:	---	(2)#17	(4)#17	(8)#17	AWG No.
Feedback Turns:	NFB	11	7	5	---
Wire Size:	---	#22	#20	(2)#17	AWG No.

TABLE III.

3. Output Capacitor

During the charge period, output diode Do is back-biased and all of the load current is supplied by the output capacitor, Co. If Co is sufficiently large, the load current will be essentially constant. Then the consequent drop in output voltage is due to the charge lost from the capacitor. The output capacitor's voltage drop during the inductor's charge period is:

$$v_o^- = \Delta V_o = \frac{I_o \Delta T}{C_o} = \frac{I_o T_{chg}}{C_o} \text{ -------------- } (41)$$

where v_o^- is the negative-going portion of the output ripple waveform.

During the discharge period, the inductor's transient current flows through capacitor Co. At the start of this interval, the instantaneous output voltage rises suddenly because of the step-current into Co. This corresponding voltage-step is caused by the effective series resistance (ESR) of the capacitor's structure. As the current into Co starts flow with constant-slope (Ip2 / Tdis), the capacitor voltage continues to rise as a function of its integrated current. We can sum these two effects and write an expression for the positive-going peak of the output ripple waveform:

$$v_o^+ = (I_{P2})(ESR) + \frac{I_{P2}}{C_o T_{dis}} \int_0^{t_x} (T_{dis} - t)\,dt \text{ --------------- } (42)$$

where: $0 \leq t_x \leq T_{dis}$

Equation (41) may be used to calculate the instantaneous positive-going peak output voltage (v_o^+) at any point during the inductor's discharge interval. We can also calculate the maximum value of v_o^+ by utilizing the measured value of ESR and by integrating to tx = Tdis. Performing the indicated integration we obtain:

$$v_o^+ = 0.16 + 0.073 = 0.233 \text{ VPK}$$

Because v_o^+ is greater than v_o^-, the peak-to-peak ripple voltage on Co is identically v_o^+. Table IV below summarizes the calculated and measured values of these ripple parameters for the 100W IBR:

	Calculated	Measured	UNITS
v_o^-	0.135	0.15	VPK
v_o^+	0.233	0.24	VPK
Ripple	0.233	0.24	VPP

TABLE IV

Photographs of the ripple waveforms are given in Appendix F.

4. Semiconductors

Several different types of semiconductors, both transistor and rectifier devices, were tested to determine their applicability in the IBR circuit configuration.

Of the transistors tested only the following were acceptable in terms of saturated drop, switching speed, and current gain:

SOLITRON 2N3599
RCA 2N3265
WESTINGHOUSE 1776-0860

By using any of these transistors it was possible to achieve the (minimum) efficiency specification limit of 90%.

To gain confidence in these devices, and to permit competition between manufacturers (to reduce price or to increase quality) it was decided that these devices would each be used in one of the IBR breadboards. Because the 2N3599 and the 2N3265 were available from the beginning of the program, they were used for the 100 watt and 200 watt units respectively. It should be noted that to achieve at least 90% efficiency in the 200W unit, two parallel 2N3599's or 2N3265's were required.

The 1776-0860 came to our attention, and became available, during the 400 watt IBR design phase, thus it was selected for this IBR.

In attempting to find the best rectifier available for use in the IBR's, we evaluated normal fast-recovery rectifiers, transistor base-collector diodes, and ion-implantation diodes. Of all the devices checked, the Westinghouse IN3911 was best. This was decided initially on the basis of DC test data comparing DC forward voltage drop for the various devices (see Figure 10).

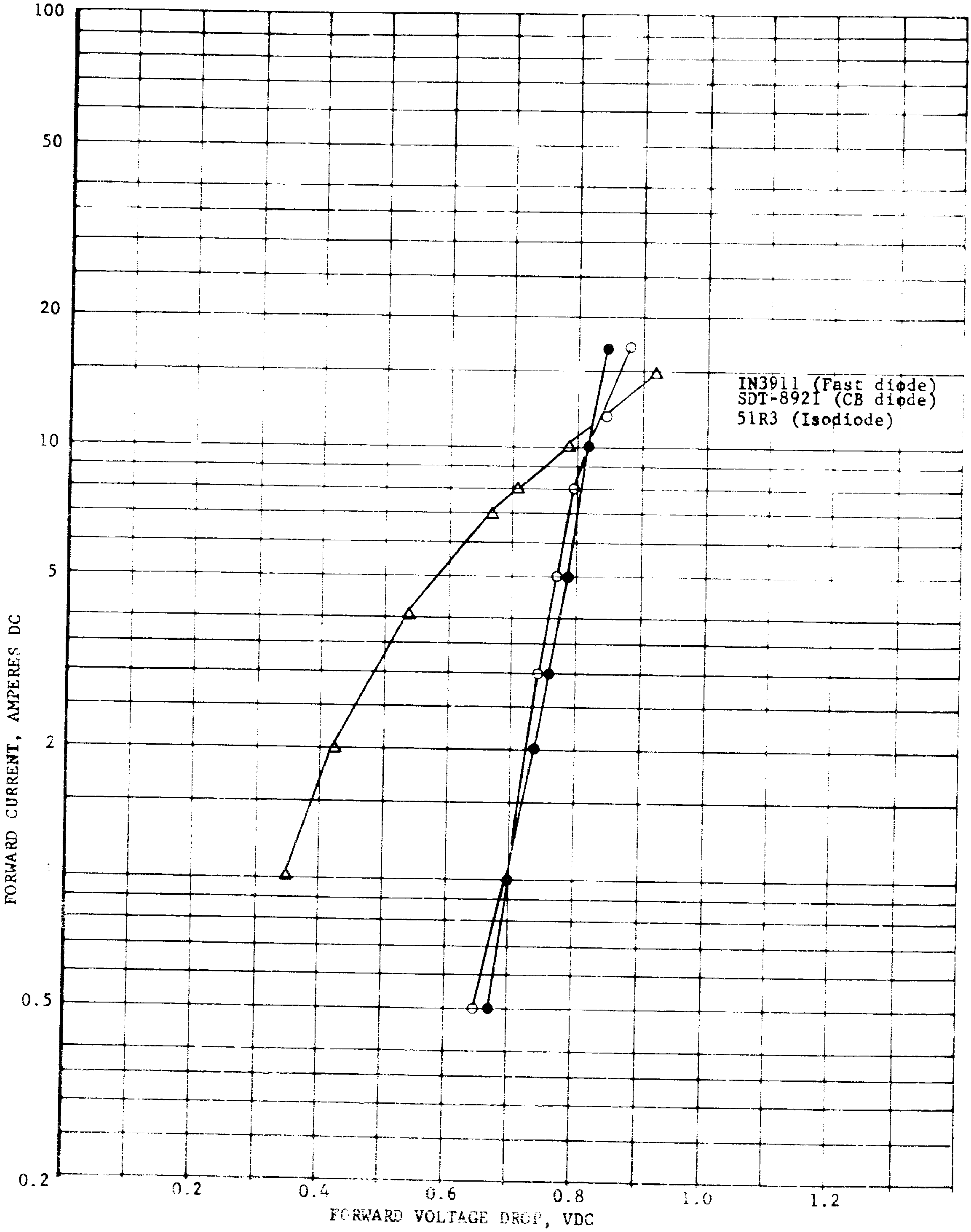


FIGURE 10. COMPARISON OF FORWARD DROP CHARACTERISTICS

C. TEST RESULTS

All three IBR's were designed and constructed in similar fashion. Figures 11, 12, and 13 show the layout of the three breadboards.

Drawings number 5507, 5535, and 5536 are the schematic diagrams of the 100W, 200W, and 400W IBR's respectively. For convenience, each parts list is placed just behind its respective schematic diagram and layout photograph.

Each IBR breadboard was subjected to both steady-state and dynamic tests, as well as to an investigation of all pertinent waveforms.

Efficiency curves are given in Figures 14, 15, and 16. Typical steady-state regulation and stability test results are given in Figure 17 for the 400W IBR. Total regulation in each case as well as ripple and efficiency were well within the specification requirements.

Results of output impedance tests are given in Figures 18, 19, and 20.

Results of the transient input and transient load tests were recorded photographically and these photos are given in appendix . In general load transients from 10% to 50% (of rated load) or from 50% to 100% (of rated load) had peak excursions of less than 1.0 volt (less than 2% of the nominal output) and in all cases returned to within $\pm$280 mv ($\pm\frac{1}{2}$% of nominal output) in less than 5.0 milliseconds.

Response to input line transients (from 25v to 38v and from 35v to 50v was uniformly excellent. All output voltage excursions were well within the steady-state regulation band ($\pm$280 mv).

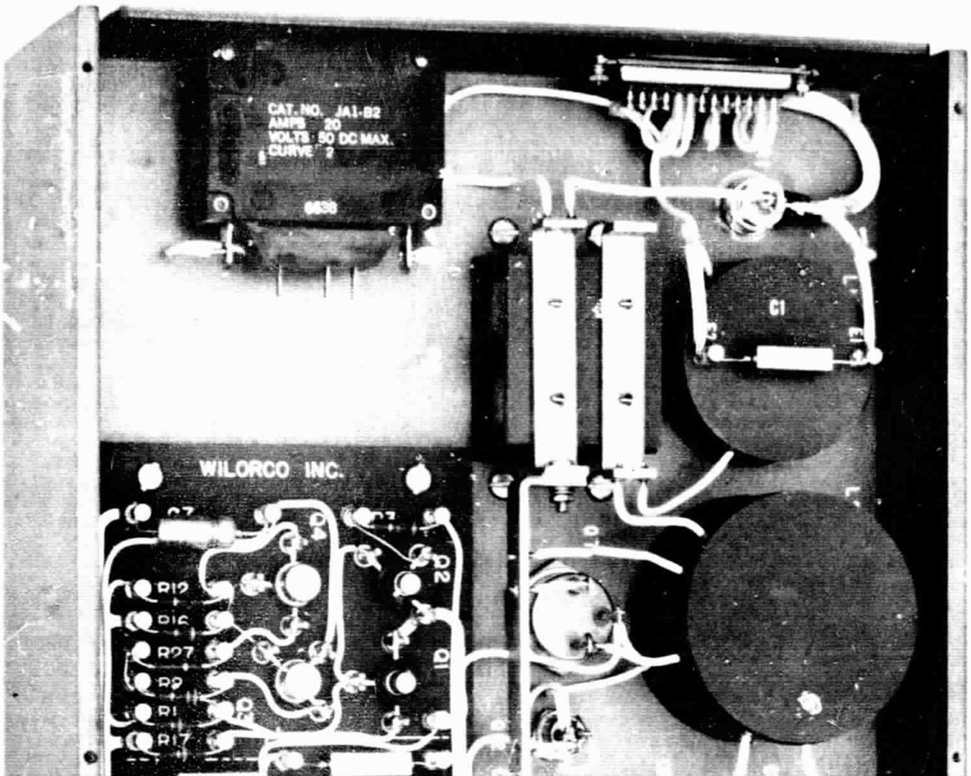
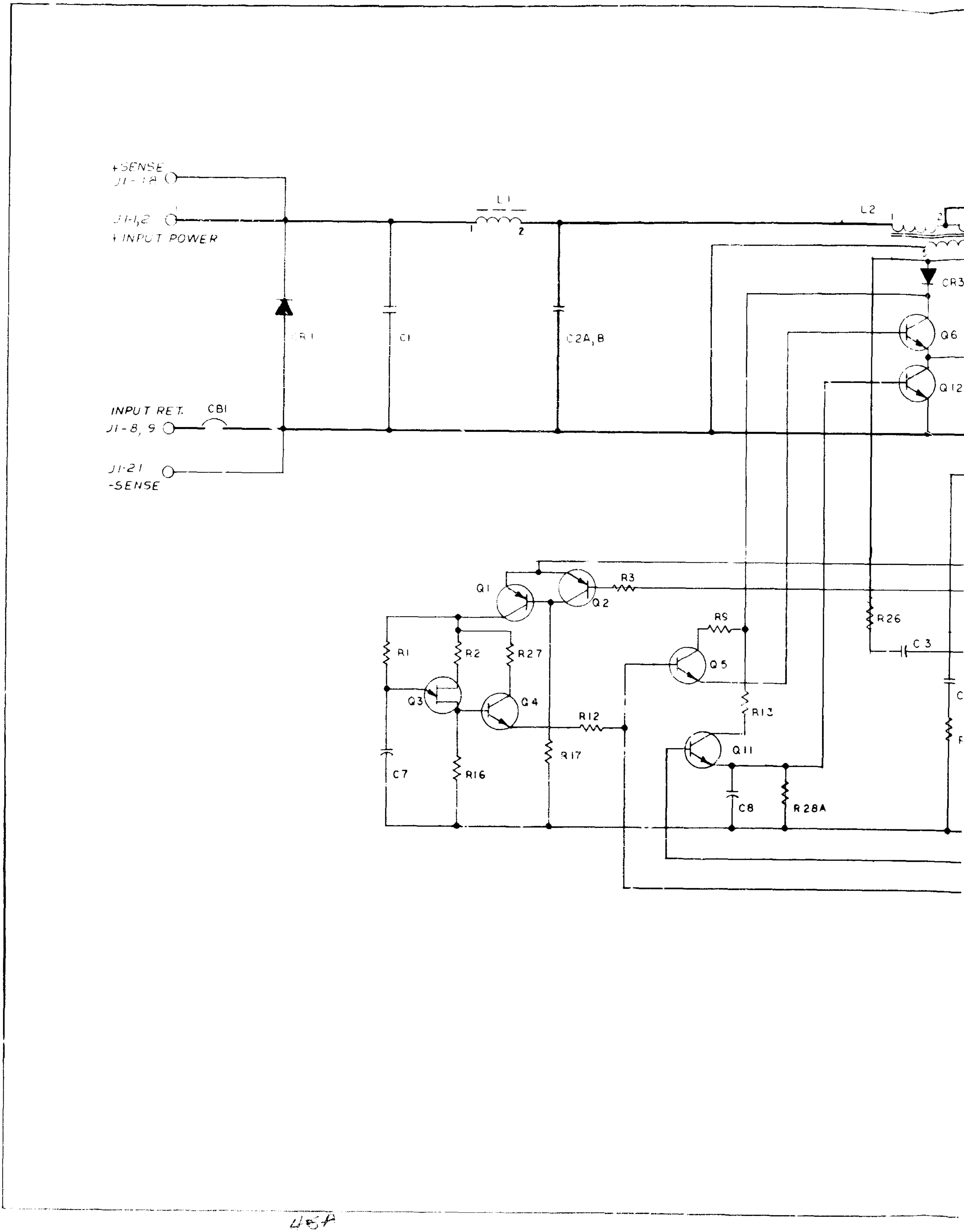
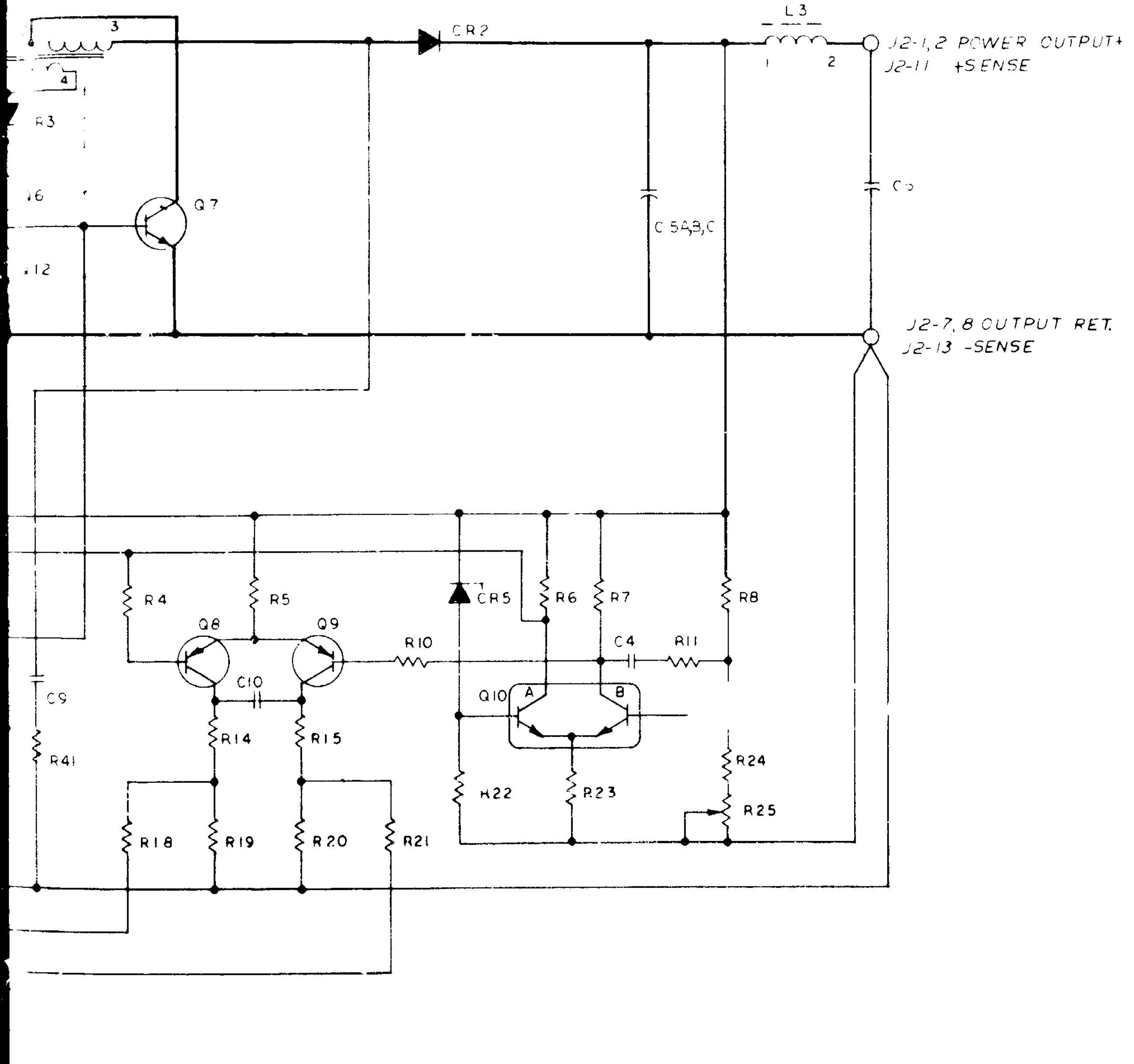


FIGURE 11. LAYOUT OF THE 100W. IBR BREADBOARD


+SENSE
J1-18
J1-1,2
+INPUT POWER
INPUT RET.
J1-8, 9
CB1
J1-21
-SENSE
CR1
C1
L1
C2A, B
L2
CR3
Q6
Q12
Q1
Q2
R3
R26
C3
R5
Q5
R1
R2
R27
Q3
Q4
R12
R13
Q11
R17
C7
R16
C8
R28A

REV.	DESCRIPTION	DATE	APPROV.



5456		-4	-3	-2	-1	ITEM	PART NO.	DESCRIPTION	MANUFACTURER
UNIT NO.	NEXT ASSY							LIST OF MATERIALS	

UNLESS OTHERWISE SPECIFIED DIMENSIONS ARE IN INCHES TOLERANCES ON
LINEAR TOL: FRACT. ± ; .XX ± ; .XXX ±
ANGULAR TOL ±
SURFACE ROUGHNESS
REMOVE ALL BURRS & BREAK ALL SHARP CORNERS

MATERIAL
MATERIAL SPEC
FINISH
FINISH SPEC

DRAFTSMAN: [signature] DATE: 28 Jan 68
CHECKED
DESIGN APP
CLIENT APP
COLONEL

WILORCO INCORPORATED
LONG BEACH, CALIFORNIA

TITLE: SCHEMATIC DIAGRAM
100W BOOST REGULATOR

SCALE
SHEET OF
SIZE: D
DWG NO.: 5507
REV

COMPONENT PARTS LIST

JET PROPULSION LABORATORY
CALIFORNIA INSTITUTE OF TECHNOLOGY

PL. 5456 REV.

SHEET 1 OF 4

FOR 100W BOOST REGULATOR

Colonel 9-3-68

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
Q1,2,8,9	TRANSISTOR					2N3963 (FAIRCHILD)
Q3	TRANSISTOR					2N490B (G.E.)
Q4,11	TRANSISTOR					2N2102 (RCA)
Q6,12	TRANSISTOR					SDT7B01 (SOLITRON)
Q7	TRANSISTOR					2N3599 (SOLITRON)
Q10	DUAL TRANSISTOR					2N2060 (FAIRCHILD)
Q5	TRANSISTOR					SDT7403 (SOLITRON)
CR1	RECTIFIER					MR1122 (MOTOROLA)
CR2	RECTIFIER					1N3911 (WESTINGHOUSE)
CR3	RECTIFIER					HF5B (HUGHES)
CR5	ZENER TC					1N4573A (MOTOROLA)
C1,8	CAPACITOR	.15 MFD	100V			DG1-154E (ELECTRON PRODUCTS)
C2,A B	CAPACITOR	440 MFD	100V			69F1164 (G.E.)
C3	CAPACITOR	.068 MFD	100V			DG1-683E (ELECTRON PRODUCTS)
C4	CAPACITOR	.39 MFL	100V			DG1-394E (ELECTRON PRODUCTS)
C5,A C	CAPACITOR	440 MFD	100V			69F1164 (G.E.)
C6	CAPACITOR	110 MFD	100V			69F475 (G.E.)
C7	CAPACITOR	9 MFD	125V			137D905C2125FO (SPRAGUE)
C9	CAPACITOR	.01 MFD	200V			96P1039252 (SPRAGUE)

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COMPONENT PARTS LIST

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CALIFORNIA INSTITUTE OF TECHNOLOGY

PL. 3456 ____ REV.

SHEET 2 OF 4

FOR 100W BOOST REGULATOR

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
C10	CAPACITOR	.022 MFD	100V			96P2239152
R1,17	RESISTOR	100K	½W		5%	RC20GF104J
R2	RESISTOR	4.7K	½W		5%	RC20GF472J
R3	RESISTOR	82K	½W		5%	RC20GF823J
R4	RESISTOR	1K	½W		5%	RC20GF102J
R5	RESISTOR	150 ohm	½W		5%	RC20GF201J
R6,7	RESISTOR	1.5K	½W		5%	RC20GF152J
R8	RESISTOR	6.34K	1/8W		1%	RN60C1341
R9,12	RESISTOR	100 ohm	½W		5%	RC20GF101J
R10,18,21	RESISTOR	1K ohm	½W		5%	RC20GF102J
R11	RESISTOR	100 ohm	½W		5%	RC20GF101J
R14,15	RESISTOR	9.1K	½W		5%	RC20GF682J
R16	RESISTOR	270 ohm	½W		5%	RC20GF271J
R19	RESISTOR	390 ohm	½W		5%	RC20GF391J
R20	RESISTOR	750 ohm	½W		5%	RC20GF751J
R22	RESISTOR	49.9K	1/8W		1%	RN60C4992
R23	RESISTOR	24K	½W		5%	RC20GF243J
R24	RESISTOR	47.5K	1/8W		1%	RN60C4752
R25	POTENTIOMETER	5K				3250L-501-502 (BOURNS)

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COMPONENT PARTS LIST

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CALIFORNIA INSTITUTE OF TECHNOLOGY

PL. 5456 ___ REV.
SHEET 3 OF 4

FOR 100W BOOST REGULATOR

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
R26	RESISTOR	15 ohm	½W		5%	RC20GF150J
R27	RESISTOR	1.0F	½W		5%	RC20GF103J
R13	RESISTOR	10 ohm	½W		5%	RC20GF100J
R28A	RESISTOR	220 ohm	½W		5%	RC20GF221J
CB1	CIRCUIT BREAKER		5A			JAI-B2 (HEIN.)
L1	CHOKE ASSEMBLY					5469 (VILORCO)
L2	CHOKE ASSEMBLY					5468 (VILORCO)
L3	CHOKE ASSEMBLY					5470 (VILORCO)
	TERM. BD					5476 (VILORCO)
	TERM. BD					5484 (VILORCO)
	INSULATOR					5473-1 (VILORCO)
	INSULATOR					5473-3 (VILORCO)
	BRKT-CAPACITOR					5471 (VILORCO)
	BRKT-CAPACITOR					5477 (VILORCO)
	BUS BAR					5474-2 (VILORCO)
	BUS BAR					5474-1 (VILORCO)
	BUS BAR					5478 (VILORCO)
	BUS BAR					5479 (VILORCO)
	BUS BAR					5475 (VILORCO)
	CHASSIS					5482 (VILORCO)

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CALIFORNIA INSTITUTE OF TECHNOLOGY

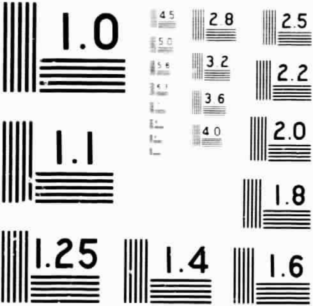
PL. 5456 REV.

SHEET 4 OF 4

FOR 100W BOOST REGUALTOR

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
	COVER-CHOKE					5485-1 (WILORCO)
	COVER-CHOKE					5485-2 (WILORCO)
	SUB CHASSIS					5481 (WILORCO)
	COVER					5483 (WILORCO)
R41	RESISTOR	22 ohm	½W		5%	RC20GF220J

JPL 0204 JULY 61 A



MICROCOPY RESOLUTION TEST CHART

NATIONAL BUREAU OF STANDARDS - 1963

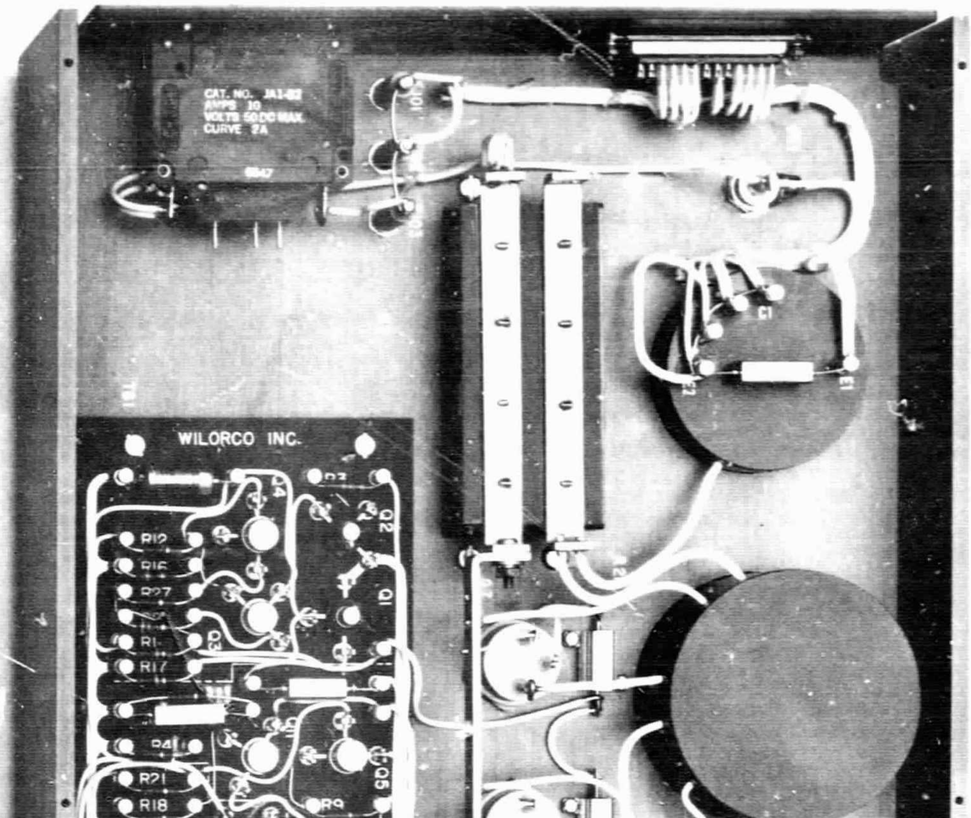
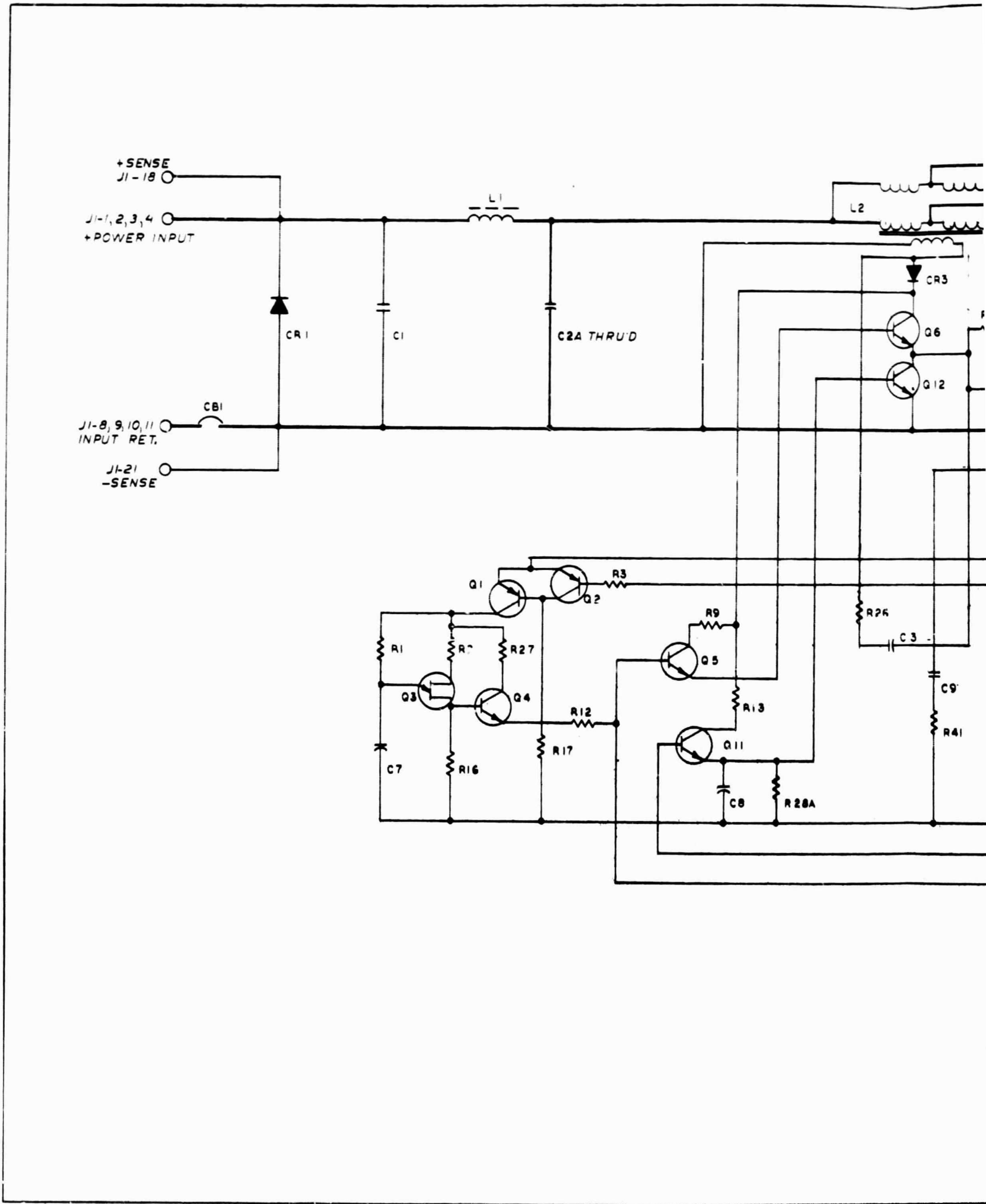
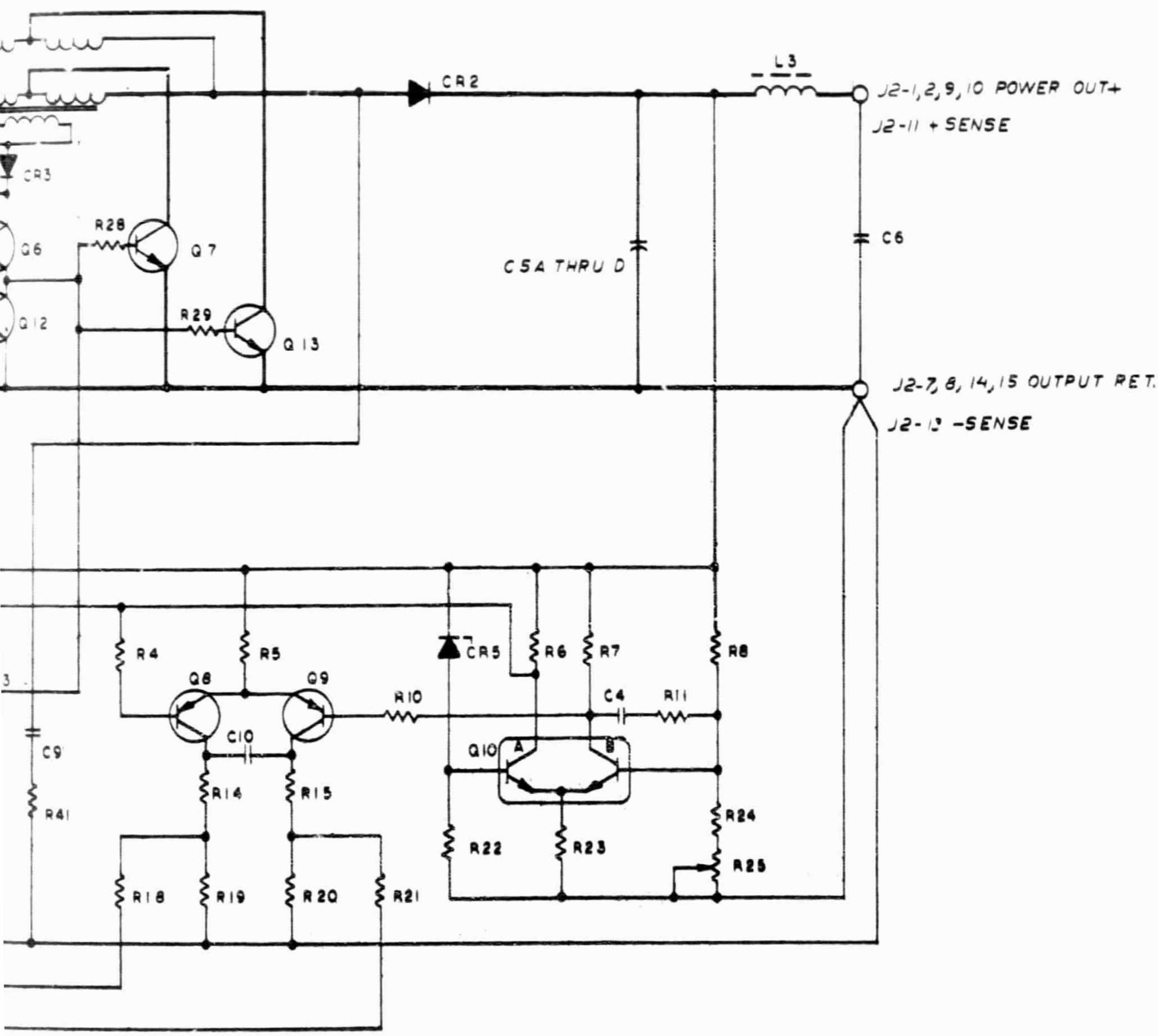


FIGURE 12. LAYOUT OF THE 200W. IBR BREADBOARD



+SENSE
J1-18
J1-1,2,3,4
+POWER INPUT
L1
L2
CR3
CR1
C1
C2A THRU D
Q6
Q12
CB1
J1-8,9,10,11
INPUT RET.
J1-21
-SENSE
Q1
Q2
R3
R9
R26
C3
R1
R27
Q5
Q3
Q4
R12
R13
C9
R41
R17
Q11
C7
R16
C8
R28A



5457

LIST OF MATERIALS

REMOVE ALL BURRS & BREAK ALL SHARP CORNERS

COLOWEL

WILORCO INCORPORATED
LONG BEACH, CALIFORNIA

SCHEMATIC DIAGRAM
200W BOOST REGULATOR

SCALE
SHEET OF

D 5535

COMPONENT PARTS LIST

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CALIFORNIA INSTITUTE OF TECHNOLOGY

PL. 5457 REV.
SHEET 1 OF 4

FOR 200W BOOST REGULATOR

Colonel 9-3-68

REF. DESIG.	COMPONENT DESCRIPTION NAME	VALUE	RATING	MATERIAL	±%	MFG. OR VENDER'S NAME AND NO.
Q1,2,8,9	TRANSISTOR					2N3963 (FAIRCHILD)
Q3	TRANSISTOR					2N490B (G.E.)
Q4,11	TRANSISTOR					2N2102 (RCA)
Q6,12	TRANSISTOR					SDT7B01 (SOLITRON)
Q10	DUAL TRANSISTOR					2N2060 (FAIRCHILD)
Q5	TRANSISTOR					SDT7403 (SOLITRON)
CR1	RECTIFIER					MR1122 (MOTOROLA)
CR2	RECTIFIER					1N3911 (WESTINGHOUSE)
CR3	RECTIFIER					HF5B (HUGHES)
CR5	ZENER TC					1N4573A (MOTOROLA)
C1,8	CAPACITOR	.15 MFD	100V			DG1-154E (ELECTRON PRODUCTS)
C2,A D	CAPACITOR	440 MFD	100V			69F1164 (G.E.)
C3	CAPACITOR	.068 MFD	100V			DG1-683E (ELECTRON PRODUCTS)
C4	CAPACITOR	.39 MFD	100V			DG1-394E (ELECTRON PRODUCTS)
C5,A D	CAPACITOR	440 MFD	100V			69F1164 (G.E.)
C7	CAPACITOR	9 MFD	125V			137D905C2125F0 (SPRAGUE)
C9	CAPACITOR	.01 MFD	200V			96P1039252 (SPRAGUE)
C10	CAPACITOR	.022 MFD	100V			96P2239152
R1,17	RESISTOR	100K	½W		5%	RC20GF104J

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PL. 5457 REV.
SHEET 2 OF 4

FOR 200W BOOST REGULATOR

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
R2	RESISTOR	4.7K	½W		5%	RC20GF472J
R3	RESISTOR	82K	½W		5%	RC20GF823J
R4	RESISTOR	1K	½W		5%	RC20GF102J
R5	RESISTOR	150 ohm	½W		5%	RC20GF201J
R6,7	RESISTOR	1.5K	½W		5%	RC20GF152J
R8	RESISTOR	6.34K	1/8W		1%	RN60C6341
R9,12	RESISTOR	100 ohm	½W		5%	RC20GF101J
R10,18,21	RESISTOR	1K ohm	½W		5%	RC20GF102J
R11	RESISTOR	100 ohm	½W		5%	RC20GF101J
R14,15	RESISTOR	6.8K	½W		5%	RC20GF682J
R16	RESISTOR	270 ohm	½W		5%	RC20GF271J
R19	RESISTOR	390 ohm	½W		5%	RC20GF391J
R20	RESISTOR	750 ohm	½W		5%	RC20GF751J
R22	RESISTOR	49.9K	1/8W		1%	RN60C4753
R23	RESISTOR	24K	½W		5%	RC20GF243J
R24	RESISTOR	47.5K	1/8W		1%	RN60C4752
R25	RESISTOR	5K				3250L-501-502 (BOURNS)
R27	RESISTOR	1.0K	½W		5%	RC20GF102J
R13	RESISTOR	10 ohm	½W		5%	RC20GF100J
R28A	RESISTOR	220 ohm	½W		5%	RC20GF221J

COMPONENT PARTS LIST

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PL. 5457 ____ REV. [ ]
SHEET 3 OF 4

FOR 200W BOOST REGULATOR

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
CB1	CIRCUIT BREAKER		10A			JA1-B2 (HEIN.)
Q7,13	TRANSISTOR					2N3265 (RCA)
C6	CAPACITOR	440 MFD	100V			69F1164 (G.E.)
C4	CAPACITOR	.39 MFD	100V			DG1-394E(ELECTRON PRODUCTS)
L1	CHOKE ASSEMBLY					5494 (WILORCO)
L2	CHOKE ASSEMBLY					5495 (WILORCO)
L3	CHOKE ASSEMBLY					5496 (WILORCO)
	TERM. BD					5500 (WILORCO)
	TERM. BD					5484 (WILORCO)
	INSULATOR					5473-2 (WILORCO)
	INSULATOR					5473-4 (WILORCO)
	BRKT-CAPACITOR					5472 (WILORCO)
	BRKT-CAPACITOR					5505 (WILORCO)
	BUS BAR					5498-1 (WILORCO)
	BUS BAR					5498-2 (WILORCO)
	BUS BAR					5499 (WILORCO)
	BUS BAR					5497 (WILORCO)
	BUS BAR					5475 (WILORCO)
	CHASSIS					5503 (WILORCO)
	COVER-CHOKE					5483-3 (WILORCO)

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COMPONENT PARTS LIST

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PL. 545 ____ REV. [ ]
SHEET 4 OF 4

FOR 200W BOOST REGULATOR

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
	COVER-CHOKE					5485-4 (WILORCO)
	SUB CHASSIS					5502 (WILORCO)
	COVER					5504 (WILORCO)
R26	RESISTOR	15 ohm	½W		5%	RC20GF150J
R28,29	RESISTOR	1 ohm	5W		3%	RH5 (DALE)
R41	RESISTOR	22 ohm	½W		5%	RC20GF220J

JPL 0204 JULY 61 A

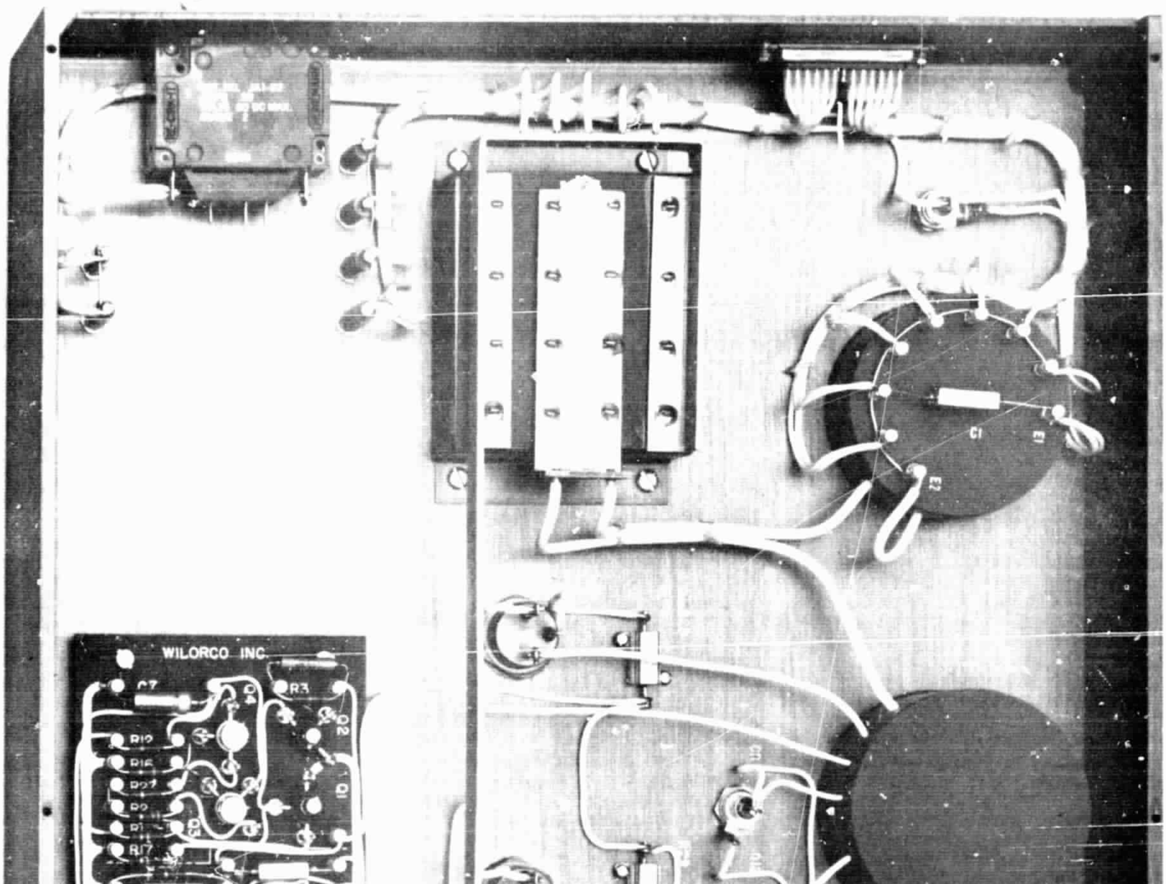
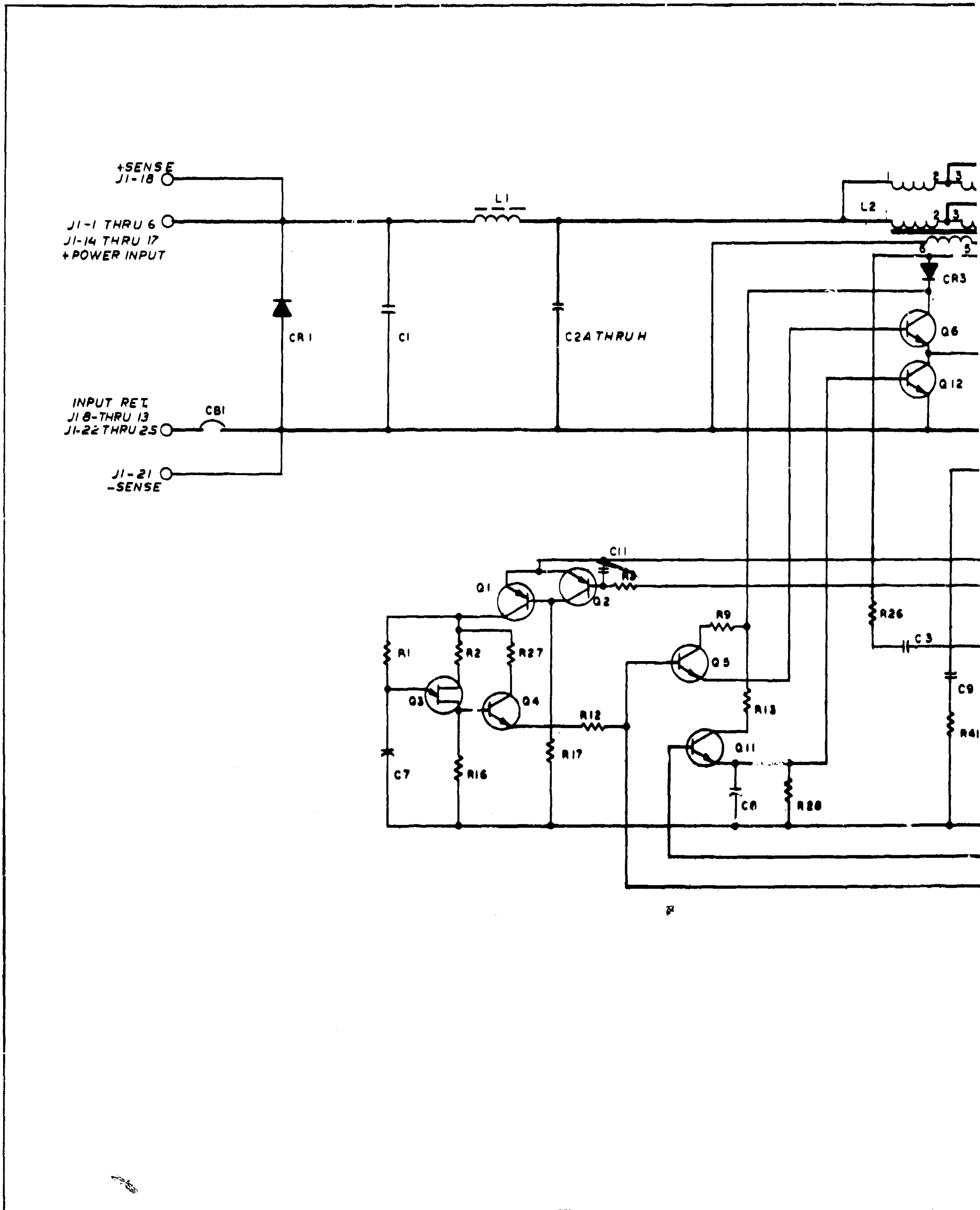
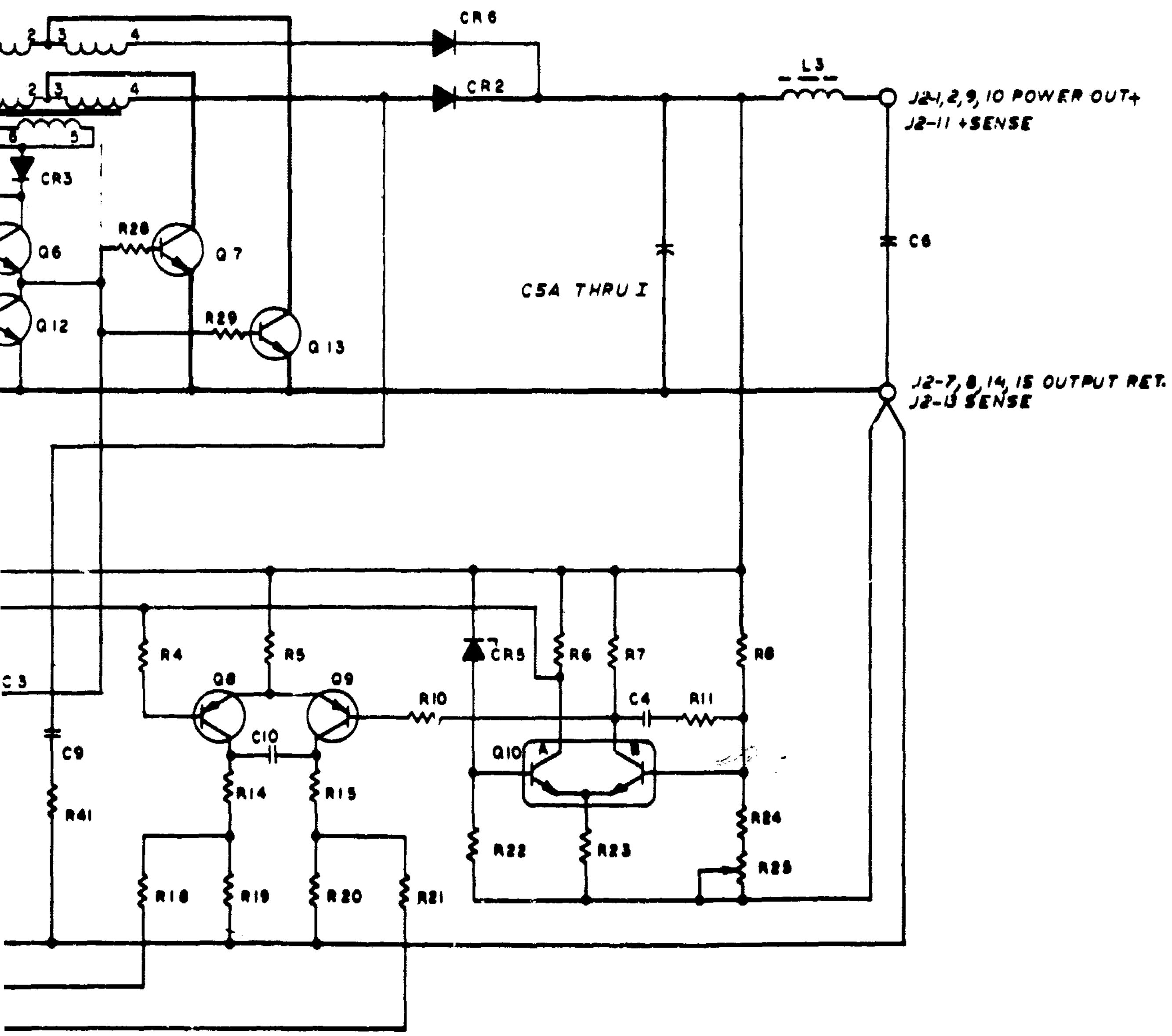


FIGURE 13. LAYOUT OF THE 400W. IBR BREADBOARD


+SENSE
J1-18
J1-1 THRU 6
J1-14 THRU 17
+POWER INPUT
INPUT RET.
J1 8-THRU 13
J1-22 THRU 25
J1-21
-SENSE
CB1
CR1
C1
L1
C2A THRU H
L2
CR3
Q6
Q12
C11
R3
Q1
Q2
R9
R26
C3
Q5
R1
R2
R27
Q3
Q4
R12
R13
C9
R41
Q11
R17
C7
R16
C8
R28

52A



WILORCO INCORPORATED
LONG BEACH, CALIFORNIA

SCHEMATIC DIAGRAM
400W BOOST REGULATOR

D 5536

COMPONENT PARTS LIST

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SHEET 1 OF 4

FOR 400W BOOST REGULATOR

COLONEL

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
Q1,2,8,9	TRANSISTOR					2N3963 (FAIRCHILD)
Q3	TRANSISTOR					2N490B (G.E.)
Q4,11	TRANSISTOR					2N2102 (RCA)
Q6,12	TRANSISTOR					SDT7B01 (SOLITRON)
Q10	DUAL TRANSISTOR					2N2060 (FAIRCHILD)
Q5	TRANSISTOR					SDT 7403 (SOLITRON)
CR1	RECTIFIER					MR1122 (MOTOROLA)
CR2	RECTIFIER					1N3911 (WESTINGHOUSE)
CR3	RECTIFIER					HF5B (HUGHES)
CR5	ZENER TC					1N4573A(MOTOROLA)
CR6	RECTIFIER					1N3911 (WESTINGHOUSE)
C1,8	CAPACITOR	.15 MFD	100V			DG1-154E (ELECTRON PRODUCTS)
C2,A H	CAPACITOR	440 MFD	100V			69F1164 (G. E.)
C3	CAPACITOR	.068 MFD	100V			DG1-683E (ELECTRON PRODUCTS)
C5,A I	CAPACITOR	440 MFD	100V			69F1164(G.E.)
C7	CAPACITOR	9 MFD	125V			137D905C2125F0 (SPRAGUE)
C9	CAPACITOR	.01 MFD	200V			96P1039252 (SPRAGUE)
C10	CAPACITOR	.022 MFD	100V			96P2239152
C11	CAPACITOR	.01 MFD	200V			162103
R1,17	RESISTOR	100K	½W		5%	RC20GF104J

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COMPONENT PARTS LIST

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SHEET 2 OF 4

FOR 400W BOOST REGULATOR

COLONEL

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
R2	RESISTOR	4.7K	½W		5%	RC20GF472J
R3	RESISTOR	82K	½W		5%	RC20GF823J
R4	RESISTOR	1K	½W		5%	RC20GF102J
R5	RESISTOR	150 ohm	½W		5%	RC20GF201J
R6,7	RESISTOR	1.5K	½W		5%	RC20GF152J
R8	RESISTOR	6.34K	1/8W		1%	RN60C6341
R9	RESISTOR	100 ohm	½W		5%	RC20GF101J
R10,18,21	RESISTOR	1K ohm	½W		5%	RC20GF102J
R11	RESISTOR	100 ohm	½W		5%	RC20GF101J
R14,15	RESISTOR	9.1K				RC20GF682J
R16	RESISTOR	270 ohm	½W		5%	RC20GF271J
R19	RESISTOR	390 ohm	½W		5%	RC20GF391J
R20	RESISTOR	750 ohm	½W		5%	RC20GF751J
R22	RESISTOR	49.9K	1/8W		1%	RN60C4753
R23	RESISTOR	24K	½W		5%	RC20GF243J
R24	RESISTOR	47.5K	1/8W		1%	RN60C4753
R25	POTENTIOMETER	5K				3250L-501-502 (BOURNS)
R26	RESISTOR	15 ohm	½W		5%	RC20GF150J
R27	RESISTOR	1.0K	½W		5%	RC20GF103J
R13, R12	RESISTOR	10 ohm	½W		5%	RC20GF10RJ

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SHEET 3 OF 4

FOR 400W BOOST REGULATOR

COLONEL

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
R28A	RESISTOR	100 ohm	½W		5%	RC20GF100J
R28	RESISTOR	.1 ohm				RH5 (DALE)
R29	RESISTOR	.1 ohm				RH5 (DALE)
CB1	CIRCUIT BREAKER		20A			JA1-B2 (HEIN.)
Q7,13	TRANSISTOR					1776-0360 (WESTINGHOUSE)
C6	CAPACITOR	440 MFD	100V			69F1164 (G.E.)
C4	CAPACITOR	.39 MFD	100V			DG1-394E (ELECTRON PRODUCTS)
R41	RESISTOR	15 ohm	½W		5%	RC20GF150J

COMPONENT PARTS LIST

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SHEET 4 OF 4

FOR 400W BOOST REGULATOR

COLONEL

REF. DESIG.	COMPONENT DESCRIPTION					MFG. OR VENDER'S NAME AND NO.
	NAME	VALUE	RATING	MATERIAL	±%	
	BRACKET-CAPACITOR					5541 (WILORCO)
	BRACKET-CAPACITOR					5544 (WILORCO)
	BUS BAR					5538 (WILORCO)
	BUS BAR					5539 (WILORCO)
	BUS BAR					5537 (WILORCO)
	BUS BAR					5540 (WILORCO)
	CHASSIS					5548 (WILORCO)
	COVER					5549 (WILORCO)
	SUB CHASSIS					5547 (WILORCO)
	INSULATOR					5542-1 (WILORCO)
	INSULATOR					5542-2 (WILORCO)
	TERMINAL BOARD					5484 (WILORCO)
	TERMINAL BOARD					5546 (WILORCO)
	COVER-CHOKE					5545-1 (WILORCO)
	COVER-CHOKE					5545-2 (WILORCO)
	CHOKE ASSEMBLY					5551 (WILORCO)
	CHOKE ASSEMBLY					5552 (WILORCO)
	CHOKE ASSEMBLY					5553 (WILORCO)

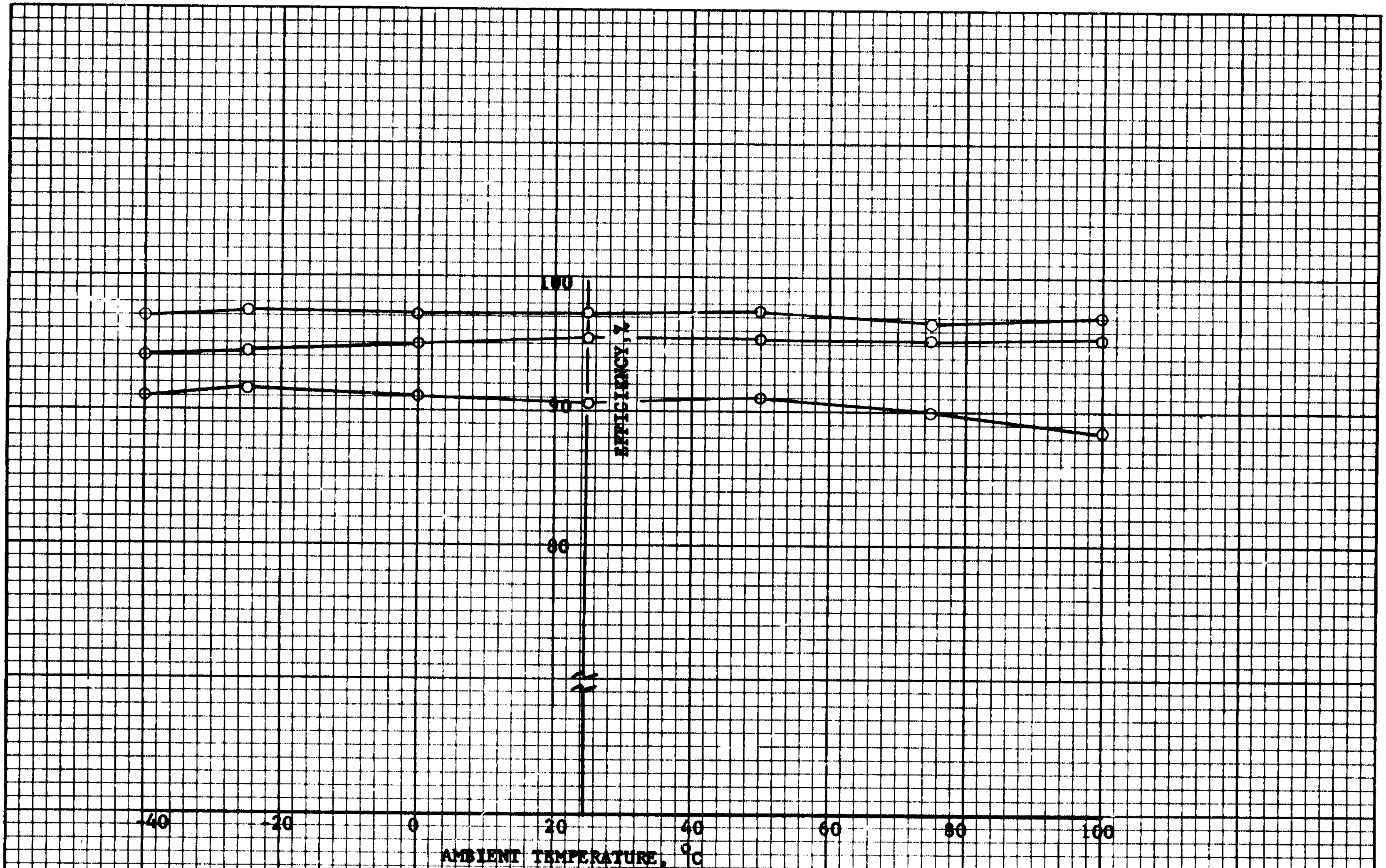


FIGURE 14. OPERATING EFFICIENCY, 100 WATT IBR.

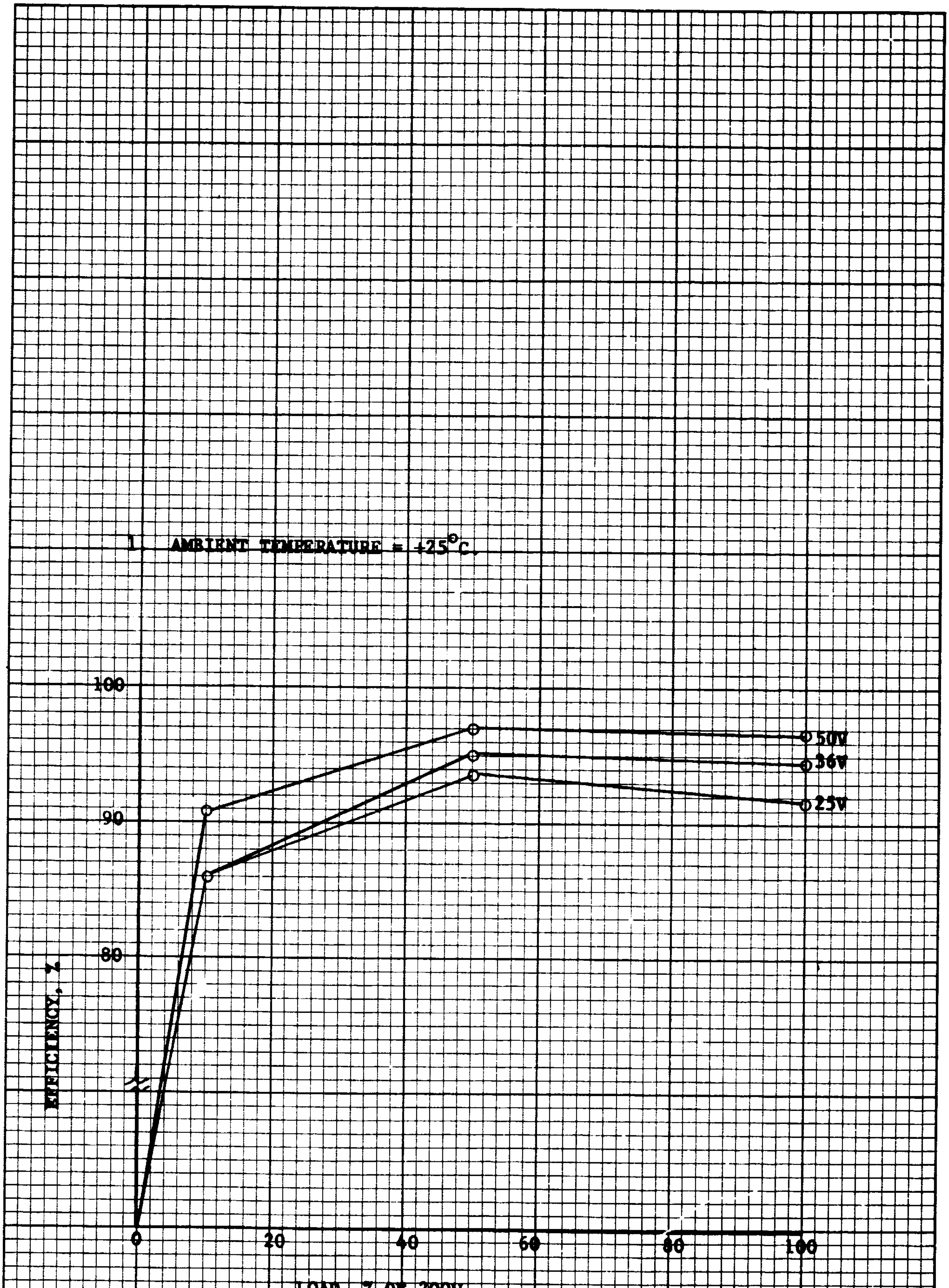


FIGURE 15. EFFICIENCY AS A FUNCTION OF LOAD FOR THE 200W IBR.

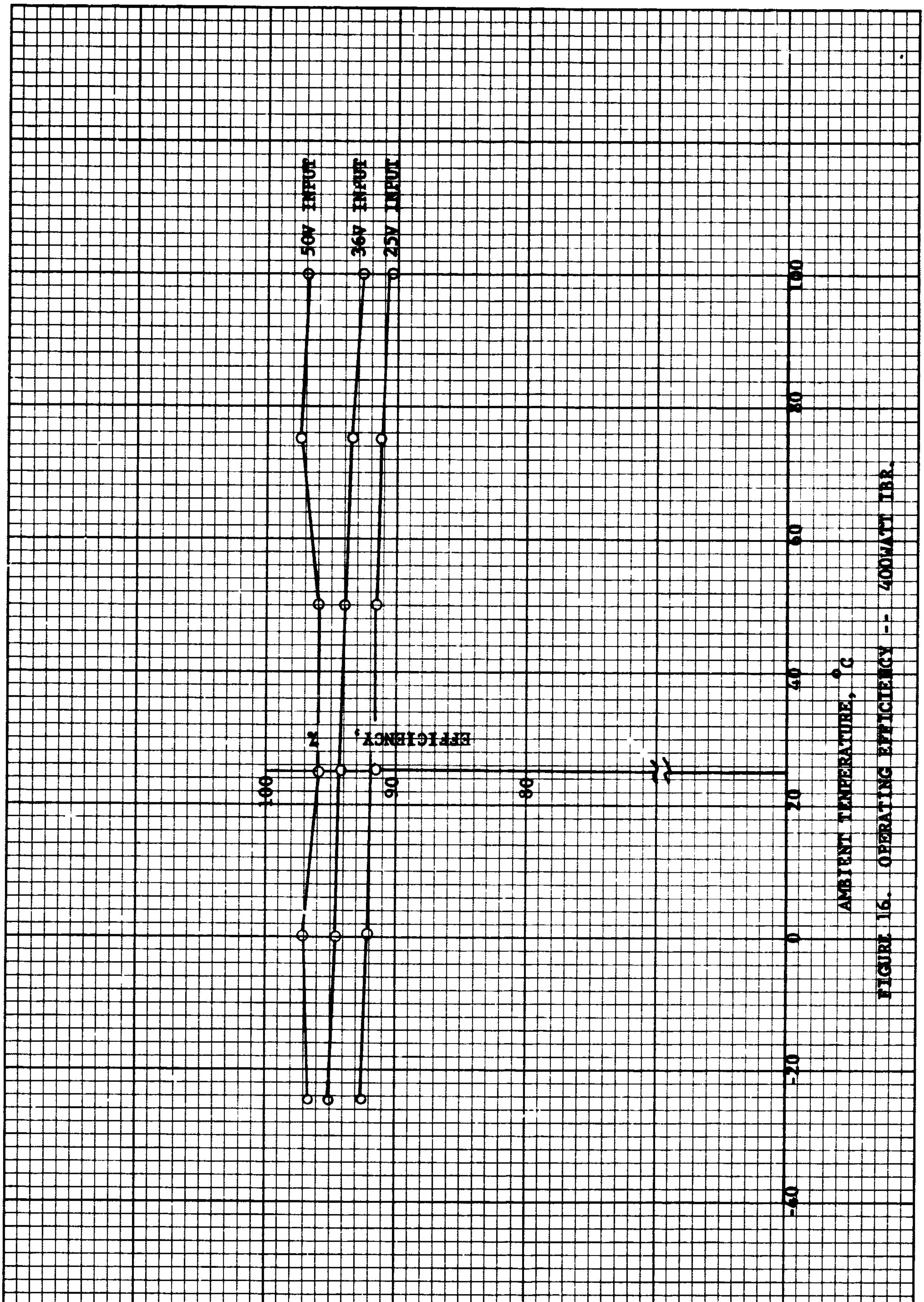


FIGURE 16. OPERATING EFFICIENCY -- 400WATT IBR.

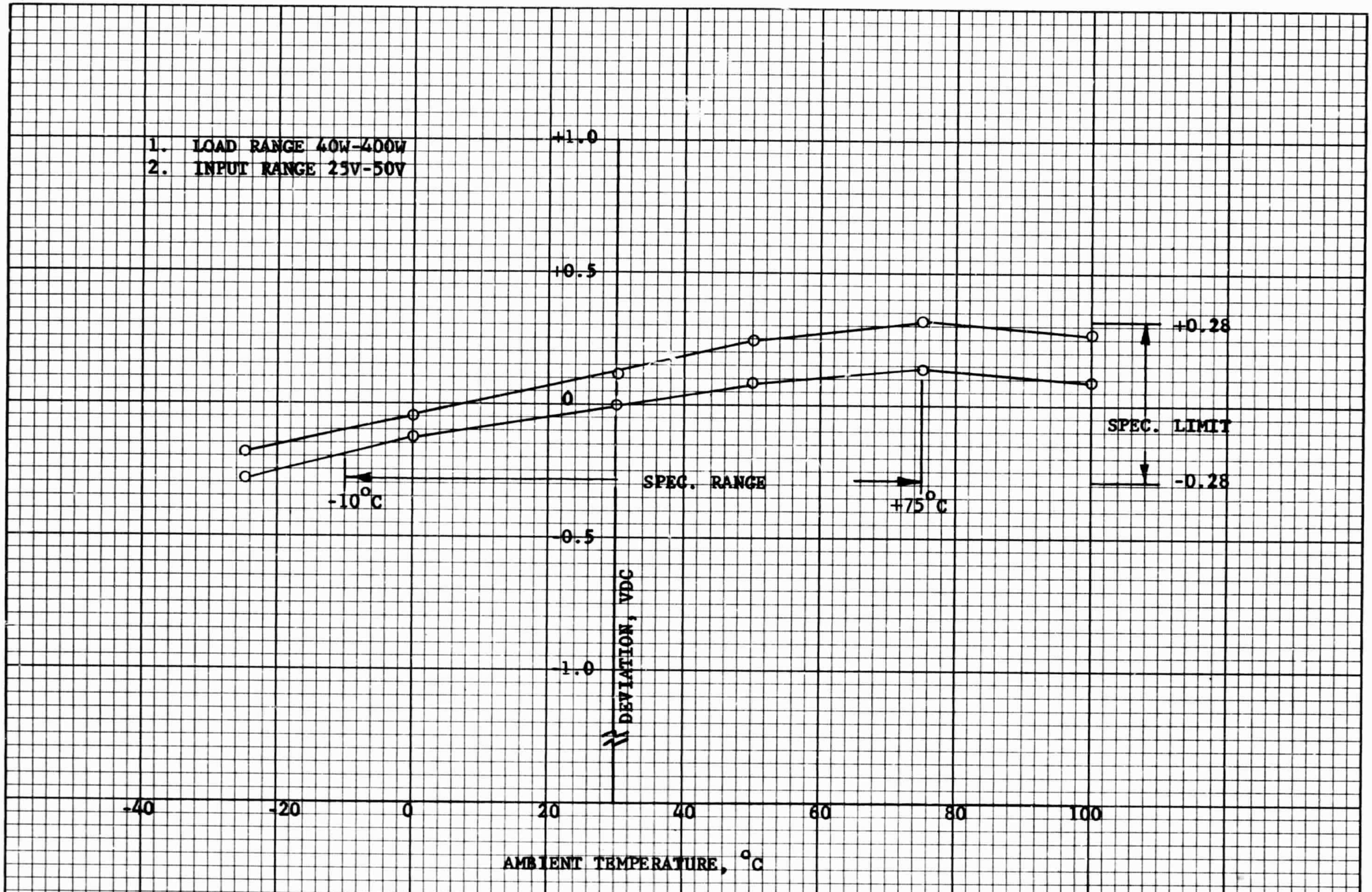


FIGURE 17. REGULATION BAND FOR THE 400 WATT IBR.

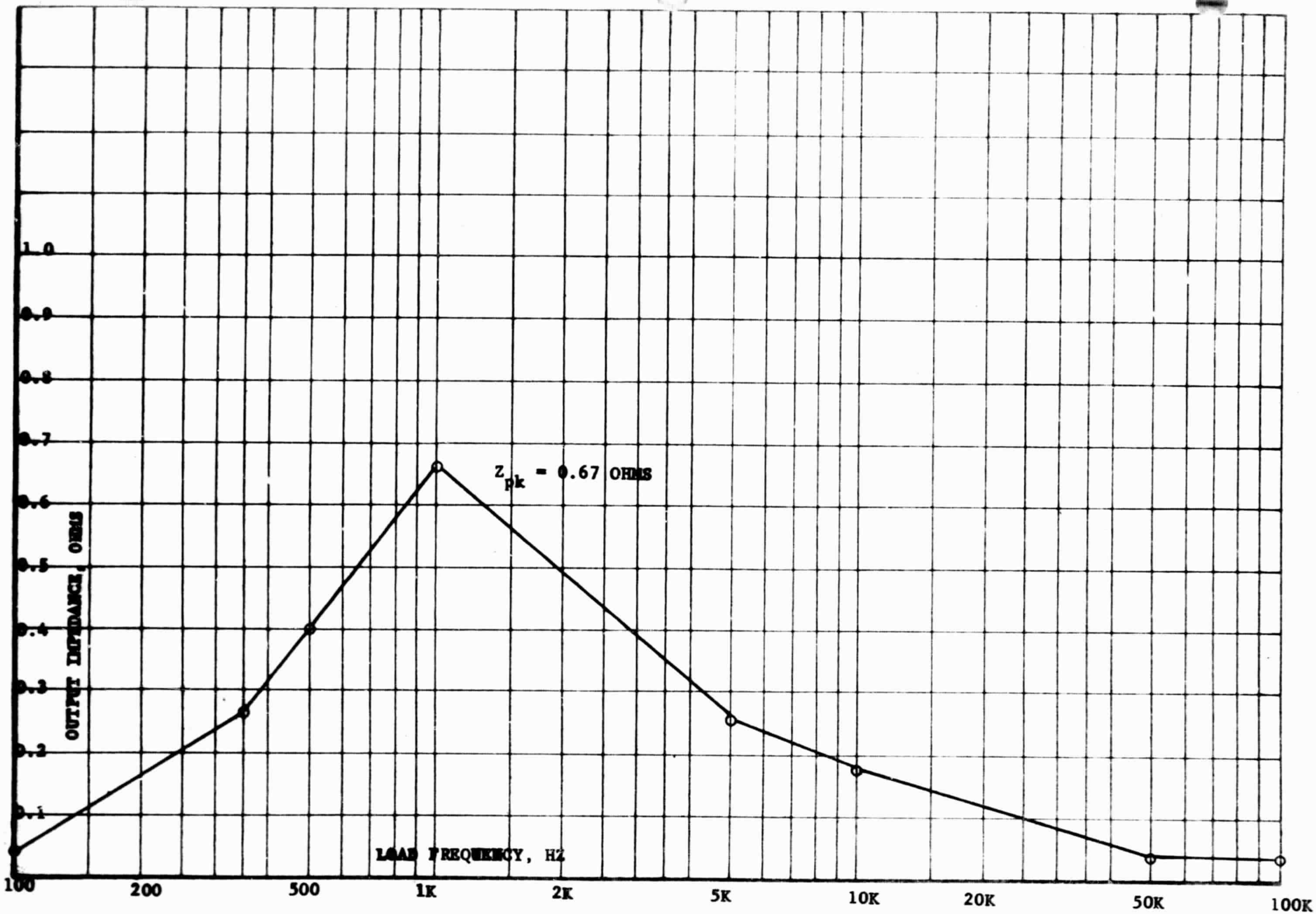


FIGURE 18. OUTPUT IMPEDANCE, 100 WATT IBR.

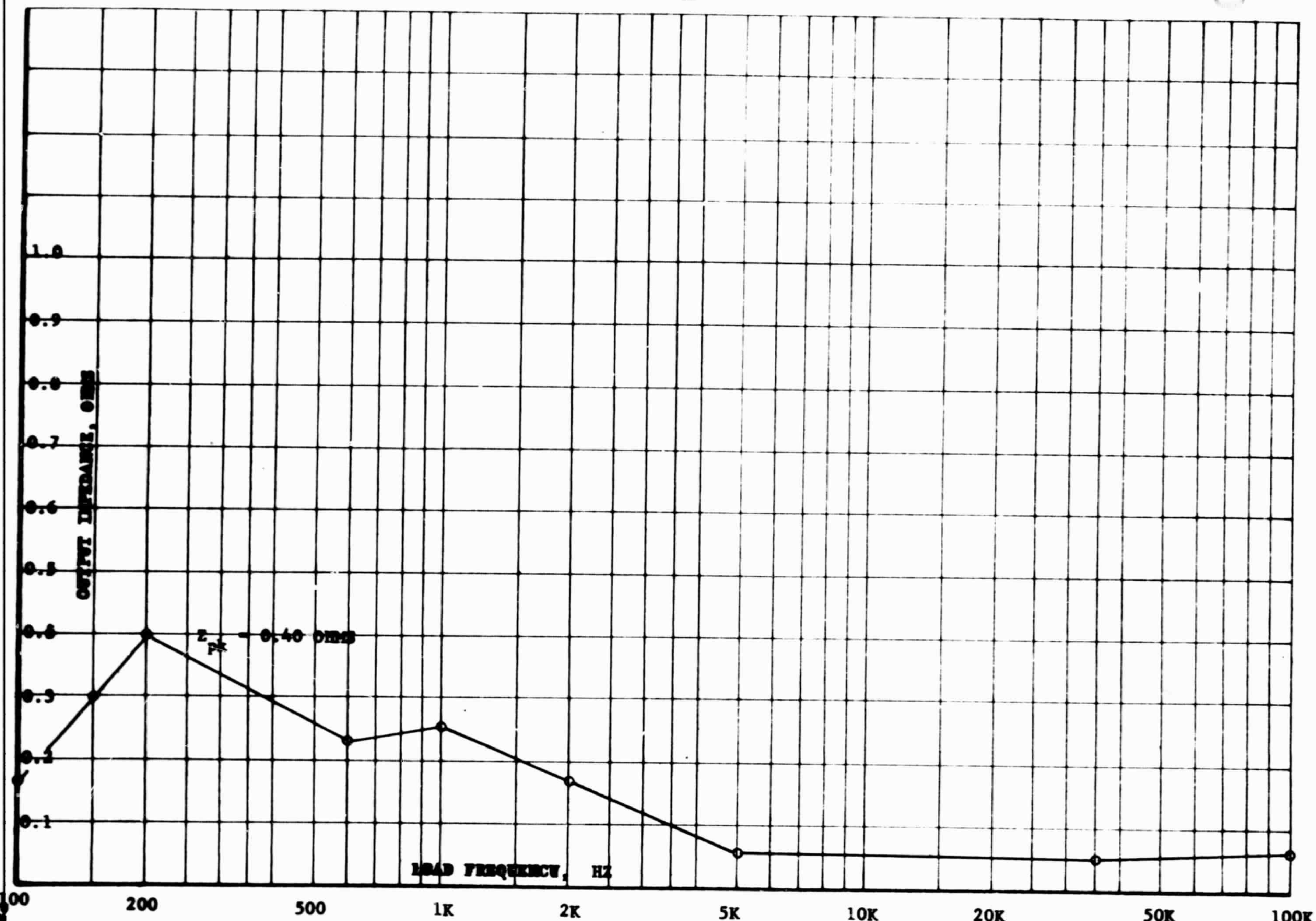


FIGURE 19 -- OUTPUT IMPEDANCE, 200WATT IBR.

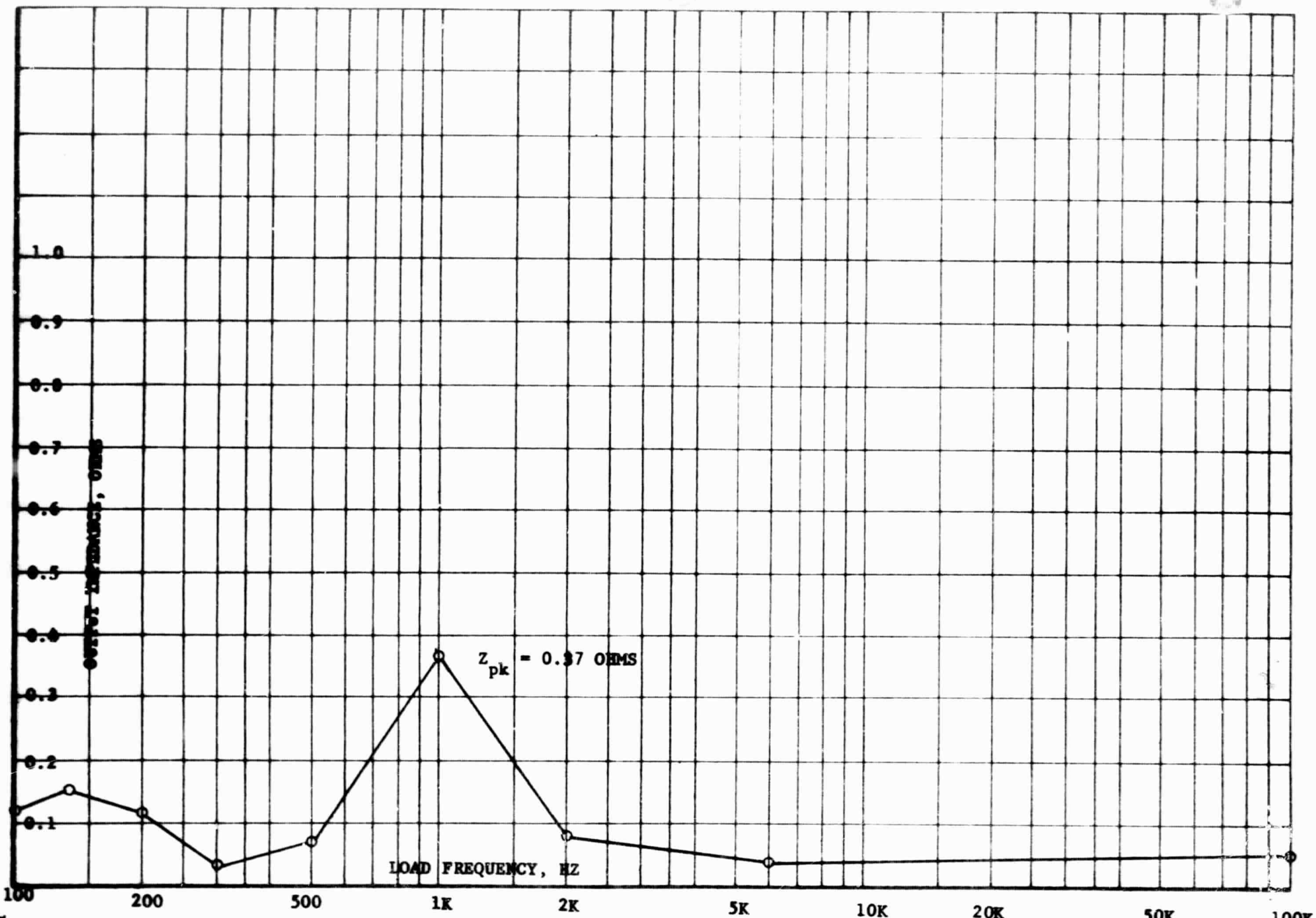


FIGURE 20. OUTPUT IMPEDANCE, 400 WATT IBR.

Response waveforms were also recorded for step-input transients. These photographs illustrate how the IBR turns ON; the input voltage appears at the output terminals quite rapidly, then we see the starting-circuit (UJT) delay, and then the actual IBR oscillator's turn ON transient.

D. LIST OF SYMBOLS

C	Capacitance, microfarads
C1	Input capacitance
C2	Feedback capacitance
Co	Output capacitance
D1	Feedback diode
D2	Steering or Clamping diode
Do	Output diode
E	Potential source, volts
Es	Power source voltage
e	Instantaneous voltage
i	Instantaneous current
I	Average current, amperes
I1	Average current during charge interval
I2	Average current during discharge interval
Ip1	Peak current during charge interval
Ip2	Peak current during discharge interval
Io	Average output current into load
L	Inductance, microhenries
L1	Charge inductance
L2	Discharge inductance
Lo	Overwind inductance
n	Overwind ratio
N	Number of turns of wire
N1	Number of turns on charge winding

N2	Number of turns on overwind
NFB	Number of turns on feedback winding
P	Power, watts
Po	Output power
Ps	Source power
Q1	Switching transistor
R	Resistance, ohms
R1	Feedback resistance
R2	Current limiting resistance
Ro	Load resistance
t	An instant of time
t1	Start of charge interval
t2	End of charge interval and beginning of discharge interval.
t3	End of discharge interval
to	Time of application of input (source) voltage
T	The period of oscillation
Tchg	The charge time interval
Tdis	The discharge time interval
T1	The period of oscillation at the lowest input voltage
T2	The period of oscillation at the highest input voltage
Tx	The minimum period of oscillation
u	The normalized input voltage
u1	Lowest value of u
u2	Highest value of u

ux	Value of u at Tx
v	Instantaneous voltage drop
V	Average voltage, volts
Vo	Average output voltage
VFB	Average feedback voltage during Tchg
β	Transistor current-gain
η	Efficiency
δ	Duty-Cycle
δ_1	Maximum duty-cycle (at u1)
δ_2	Minimum duty-cycle (at u2)
δ_x	Duty-cycle at Tx
Es1	Lowest power source voltage
Es2	Highest power source voltage

IV

RECOMMENDATIONS

A. PACKAGING

Because of their extremely low parts weight, we believe that the IBR's may be packaged to yield specific power numbers greater than 30 watts per pound. Table V below summarizes the estimated physical characteristics achievable.

A program to package one or more of the IBR's in a flight configuration seems to be worthwhile.

B. INTEGRATION

It is also of some interest to determine if a configuration in which the IBR drives an Inverter can be regulated from the Inverter's output terminals. This question has to do with the stability of the resultant closed-loop. If this configuration proves feasible it will provide the means to maintain the 50VAC, 2.4KHZ, Inverter output to much tighter regulation limits than are presently achievable.

Two kinds of protective schemes are required by the Power Subsystem:

1. Protection of the raw (solar array and battery) bus from short-circuits.
2. Protection of the inverter output from a short-circuit of one particular load.

Protection of the raw DC bus may be accomplished by insertion of a transistor switch between the raw bus and the IBR input. This switch would remain saturated for all operating currents, but would enter a current-limiting mode in the event of an overload or short-circuit at its output terminals. We envision a current limiting scheme in which a DC current transformer is biased to a preset maximum value. This saturable transistor, in conjunction with others, can also be used to accomplish those switching functions presently necessitating the Kinetics switch. This latter fact may be particularly useful in missions having extremely long life, such as the "Grand Tour".

Magnetic protection of individual spacecraft loads is also practical at the 2.4KHZ operating frequency of the inverter. A reactor in series with a given load is biased to a preset maximum (gate) current level; as the load tries to draw more current, the reactor starts to absorb (input) voltage thus creating a current-limited input. The obvious advantage of this scheme is that reset is automatic.

It seems to us mandatory that these areas be investigated in a thorough manner. One result of the investigation may be the increase of the spacecraft's power subsystem frequency from 2.4KHZ to 4.8KHZ.

If this frequency increase is adopted, a second result may be the replacement of the inverter's switching relays by all solid-state switching. This is because it becomes practical to use more magnetic components at the higher frequencies.

C. THEORY

A direct analog of the IBR is a "bucking" regulator configuration. It is possible to build a Buck Regulator using the same feedback.

mechanism used on the IBR.

A program to investigate the efficiency, and to write the operating equations for the Buck Regulator is a natural outgrowth of the IBR investigation. Buck Regulators may be particularly useful where solar-cell panels reach very low (steady-state) temperatures and where maximum power is required at the low temperature (i.e. at high bus voltages).

A program should also be instituted to investigate the buck-boost configuration. This is the configuration in which the output is taken from an isolated secondary winding and thus allows the output voltage to be, in effect, unrelated to the input voltage range.

D. PROGRAMMING

It is also possible to write a computer program which will size all major circuit components as a function of output power and operating frequency. This will require that the computer has available a "library" of data on capacitors, cores, wire, and power semiconductors. Such a library will be useful in itself as an invaluable aid to the power design engineer.

	100W	200W	400W	UNITS
COMPONENT WEIGHT	2.3	4.1	8.0	lb
CASE WEIGHT	0.9	1.7	2.1	lb
UNIT WEIGHT	3.2	5.8	10.1	lb
UNIT VOLUME	80	130	220	cu. in.
UNIT DENSITY	0.040	0.045	0.046	lb/cu. in.
DISS DENSITY	0.14	0.17	0.20	w/sq. in.
SPECIFIC POWER	31	35	40	w/lb.
DISS/SURFACE AREA	0.10	0.14	0.22	w/sq. in.
FREE-AIR TEMP RISE	18	23	32	^oC

TABLE V.

APPENDIX A

Derivation of Operating Equations

During the charge interval the voltage on the inductor is essentially equal to the input voltage, therefore;

$$E_S = L_1 \frac{\Delta I}{\Delta T} = L_1 \frac{I_{P1}}{T_{chg}} \qquad \text{-----------------(1)}$$

During the discharge interval the voltage on the inductor (including the overwind) is essentially equal to the output voltage less the input voltage, therefore;

$$V_0 - E_S = L_2 \frac{\Delta I}{\Delta T} = L_2 \frac{I_{P2}}{T_{dis}} \qquad \text{-----------------(2)}$$

Solving equations (1) and (2) for the charge and discharge times, gives;

$$T_{chg} = L_1 I_{P1} / E_S \qquad \text{-----------------(3)}$$

and, $$T_{dis} = L_2 I_{P2} / (V_0 - E_S) \qquad \text{-----------------(4)}$$

Adding equations (3) and (4), we obtain the total period;

$$T = \frac{L_1 I_{P1}}{E_S} + \frac{L_2 I_{P2}}{V_0 - E_S} \qquad \text{-----------------(5)}$$

Substituting $I_{P1} = n I_{P2}$ and $L_2 = n^2 L_1$ into equation (5) and collecting terms, yields;

$$T = n L_1 I_{P2} \left[\frac{1}{E_s} + \frac{n}{V_0 - E_s} \right] \text{----------------(6)}$$

Substituting $I_{P2} = 2I_0 / (1-\delta)$ (derived in Appendix D) into equation (6) yields;

$$T = \frac{2 n L_1 I_0}{1-\delta} \left[\frac{1}{E_s} + \frac{n}{V_0 - E_s} \right] \text{----------------(7)}$$

Substituting $I_0 = V_0 / R_0$ into equation (7), and rearranging;

$$T = \frac{2 n L_1}{R_0 (1-\delta)} \left[\frac{V_0}{E_s} + \frac{n V_0}{V_0 - E_s} \right] \text{----------------(8)}$$

By defining a "normalized" input voltage as; $\mu = E_s / V_0$ ----------------(9)

and substituting the definition (9) into equation (8) gives;

$$T = \frac{2 n L_1}{R_0 (1-\delta)} \left[\frac{1}{\mu} + \frac{n}{1-\mu} \right] \text{----------------(10)}$$

Substituting the expression for the overwind ratio (derived in Appendix C) into equation (10), results in the following simple form of the period;

$$T = \left(\frac{2 L_1}{R_0} \right) \left(\frac{1-\mu}{\mu^2 \delta^2} \right) \text{----------------(11)}$$

To obtain any stationary values of T requires that T be stated in terms of the independent variable "u" and the constant parameters. Thus we rewrite equation (11) by substituting equation (C2);

$$T = \left(\frac{2L_1}{R_0}\right)\left(\frac{1-u}{u^2}\right)\left(1 + \frac{un}{1-u}\right)^2 \qquad \text{----------------(12)}$$

APPENDIX B

Calculation of Operating Frequency Range

Once having selected the operating frequency under a given set of conditions, the frequency under any other set of conditions is calculable from equation (A12), in terms of the period;

$$T = \left(\frac{2L_1}{R_o}\right)\frac{(1-u+un)^2}{u^2(1-u)} \qquad \text{-----------------} (1)$$

$$= \frac{2 \times 288 \times 10^{-6}}{31.4} F_1(u)$$

$$T = 18.3\, F_1(u) \quad \text{microseconds} \qquad \text{---------------} (2)$$

$$\text{where } F_1(u) \equiv \frac{(1-u+un)^2}{u^2(1-u)} \qquad \text{---------------------} (3)$$

The expression F1 (u) can be minimized with respect to the normalized input voltage. Differentiating F1 (u) with respect to u, gives;

$$\frac{dF_1(u)}{du} = 2(n-1)(1-u+un)(u^2-u^3)^{-1} - (1-u+un)(2u-3u^2)(u^2-u^3)^{-2} \qquad \text{------------} (4)$$

Letting the derivative in equation (4) vanish and collecting terms yields;

$$u_x^2 + \frac{3}{n-1}u_x - \frac{2}{n-1} = 0 \qquad \text{--------------------} (5)$$

Solving equation (5) for ux yields;

$$u_x = \frac{(8n+1)^{1/2} - 3}{2(n-1)} \qquad \text{--------------------} (6)$$

Substituting the value of overwind ratio (n=1.335) calculated in Appendix C into equation (6) yields;

$$u_x = 0.626\,.$$

Thus the maximum operating frequency occurs at an input voltage, Es,x;

$$E_{s,x} = 1 + u_x V_0 \qquad \text{-------------------- (7)}$$

$$= 1 + 0.626 \times 56$$

$$E_{s,x} = 36.2 \text{ VOLTS.}$$

It should be noted that the operating frequency is directly proportional to the load resistance. Thus as one approaches the no-load condition the operating frequency becomes very high and is dependent only on the internal fixed losses and on component response limitations.

APPENDIX C

Calculation of Duty-Cycle and Overwind Ratio

The duty-cycle, δ , may be calculated directly from its definition;

$$\delta \equiv T_{chg}/T$$
$$= 1 - T_{dis}/T$$

Substituting from equations (A4) and (A5);

$$\delta = 1 - \frac{\left(\frac{L_2 I_{P2}}{V_o - E_s}\right)}{\left(\frac{L_1 I_{P1}}{E_s}\right) + \left(\frac{L_2 I_{P2}}{V_o - E_s}\right)} \quad \text{---------------(1)}$$

Simplifying equation (1) results in the following expression for the duty cycle;

$$\delta = \left(1 + \frac{un}{1-u}\right)^{-1} \quad \text{-----------------(2)}$$

Equation (2) may be rewritten to obtain the overwind ratio required by the circuit;

$$n = \left(\frac{1}{\delta} - 1\right)\left(\frac{1}{u} - 1\right) \quad \text{-----------------(3)}$$

For our particular case the overwind ratio is calculated at a minimum input voltage (determined by external requirements) and a maximum duty-cycle (selected by the designer);

$$n = \left(\frac{1}{0.5} - 1\right)\left(\frac{1}{0.428} - 1\right)$$

$$n = 1.335$$

We may now use the formula for duty-cycle, equation (2), to calculate the duty-cycle at any input voltage.

For example, we may calculate the duty-cycle at the maximum frequency of operation (Tx);

$$\delta_x = \left(1 + \frac{0.626 \times 1.335}{1 - 0.626}\right)^{-1}$$

$$\delta_x = 30.9\% .$$

And at the maximum input voltage (U_2);

$$\delta_2 = \left(1 + \frac{0.875 \times 1.335}{1 - 0.875}\right)^{-1}$$

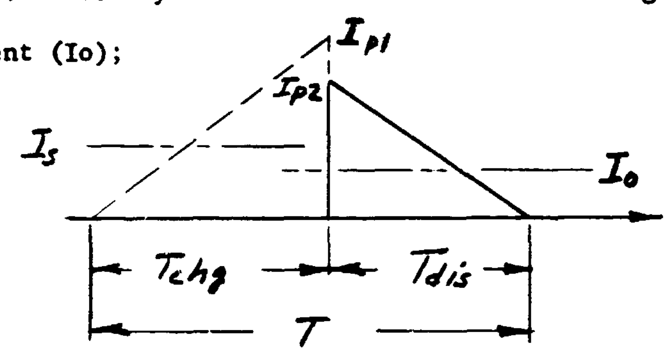
$$\delta_2 = 9.7\% .$$

APPENDIX D

Calculation of Currents

1. **Peak Current During Discharge --**

The peak current during the inductor's discharge interval (Ip2) is calculated directly from its relation to the average output, or load, current (Io);



$$I_o T = \frac{1}{2} I_{p2} T_{dis} \qquad \text{------------------ (1)}$$

$$I_{p2} = \frac{2 I_o T}{T_{dis}} = \frac{2 I_o}{1-\delta} \qquad \text{------------------ (2)}$$

2. **Peak Current During Charge--**

$$I_{p1} = n I_{p2} = \frac{2 n I_o}{1-\delta} \qquad \text{------------------ (3)}$$

3. **Average Input (Source) Current --**

$$I_s T = \frac{1}{2} I_{p1} T_{chg} + I_o T \qquad \text{------------------ (4)}$$

$$I_s = n I_o \frac{\delta}{1-\delta} + I_o \qquad \text{------------------ (5)}$$

$$I_s = I_o \left(1 + \frac{n\delta}{1-\delta}\right) = \frac{I_o}{\mu} \qquad \text{------------------ (6)}$$

4. RMS Currents in Windings --

For any sawtooth waveforms, the RMS value may be expressed as follows;

$$I_{RMS} = I_{PK}\left(\frac{T_{ON}}{3T}\right)^{1/2} \quad \text{--------------- (7)}$$

Applying this relation to our charging inductor's waveforms, yields;

From equations (2) and (7);

$$I_{0,RMS} = \frac{2I_0}{1-\delta}\left(\frac{T-T_{chg}}{T}\right)^{1/2} \quad \text{--------------- (8)}$$

$$= \frac{2I_0}{1-\delta}\left(\frac{1-\delta}{3}\right)^{1/2} \quad \text{--------------- (9)}$$

$$= 2I_0\left[3(1-\delta)\right]^{-1/2} \quad \text{---------------(10)}$$

From equations (2), (3), and (7);

$$I^2_{S,RMS} = I^2_{1,RMS} + I^2_{0,RMS} \quad \text{---------------(11)}$$

$$= \left(\frac{2nI_0}{1-\delta}\right)^2\frac{\delta}{3} + \frac{(2I_0)^2}{3(1-\delta)} \quad \text{--------------- (12)}$$

$$= \frac{4I_0^2}{3(1-\delta)}\left(1+\frac{n^2\delta}{1-\delta}\right) \quad \text{---------------(13)}$$

$$= \frac{2I_0}{\sqrt{3(1-\delta)}}\left(1+\frac{n^2\delta}{1-\delta}\right)^{1/2} \quad \text{--------------- (14)}$$

5. **Relationship of Peak Currents --**

The energy stored in inductor L_1 is essentially equal to the energy released through inductor L_2. Thus we have;

$$\frac{1}{2} L_1 I_{p1}^2 = \frac{1}{2} L_2 I_{p2}^2$$

But $L_2 = n^2 L_1$;

$$L_1 I_{p1}^2 = n^2 L_1 I_{p2}^2$$

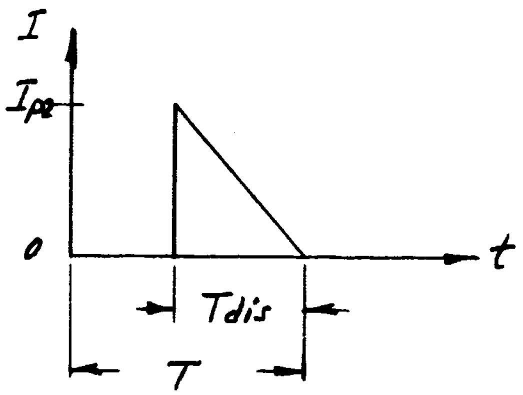
and $I_{p1} = n I_{p2}$ ------------------(15)

APPENDIX E

Boost-Regulator Inductor Design

This analysis was performed by D. Hopper of JPL. It is cur opinion that it is a useful adjunct to the design engineer who is less interested in canonical forms and more interested in answers.

From analysis, we have the waveform below. Also the decay can be assumed to be linear:



$T = \frac{1}{f}$ (period)

$$I_{avg} = I_o = \frac{1}{2} I_{p2} \frac{T_{dis}}{T} \qquad \text{-----------------(1)}$$

$$I_{p2} = 2 I_o \frac{T}{T_{dis}} \qquad \text{-----------------(2)}$$

Because $\sum v = 0$

$$E_S + V_{L2} - V_D - V_o = 0 \qquad \text{-----------------(3)}$$

Where E_S = input voltage,
V_{L2} = Inductor voltage,
V_D = diode forward drop,
and V_o = output voltage.

Equation (3) yields:

$$V_{L2} = V_o + V_D - E_s \qquad \text{-----------------(4)}$$

From $V_L = L\dfrac{di}{dt}$;

$$L_2 = V_{L2}\frac{dt}{di} = V_{L2}\frac{\Delta t}{\Delta i}$$

$$= (V_o + V_D - E_s)\left[\frac{T_{dis}}{2I_o\left(\frac{T}{T_{dis}}\right)}\right]$$

$$L_2 = (V_o + V_D - E_s)\frac{T_{dis}^2}{2I_o T} \qquad \text{-----------------(5)}$$

When transistor Q1 is ON, $\sum v = 0$;

$$E_s - V_{L1} - V_{sat} = 0 \qquad \text{-------- --------(6)}$$

$$V_{L1} = E_s - V_{sat} \qquad \text{-----------------(7)}$$

From Transformer Equations,

$$V = N\frac{d\phi}{dt} \times 10^{-8} \qquad \text{-----------------(8)}$$

Also $\dfrac{d\phi}{dt} = K\dfrac{di}{dt}$ where (K= a constant)

Because $\frac{di}{dt}$ is assumed linear, $\frac{d\phi}{dt}$ can be assumed to be linear.

$$d\phi_{dis} = \left(\frac{V_{L2}}{N_2}\right) dt \qquad \text{-----------------(9)}$$

$$\Delta\phi_{dis} = \left(\frac{V_{L2}}{N_2}\right) T_{dis} \qquad \text{-----------------(10)}$$

$$d\phi_{chg} = \left(\frac{V_{L1}}{N_1}\right) dt \qquad \text{-----------------(11)}$$

$$\Delta\phi_{chg} = \left(\frac{V_{L1}}{N_1}\right) T_{chg} \qquad \text{-----------------(12)}$$

Because

$$\Delta\phi_{dis} = \Delta\phi_{chg} \qquad \text{-----------------(13)}$$

$$\left(\frac{V_{L2}}{N_2}\right) T_{dis} = \left(\frac{V_{L1}}{N_1}\right) T_{chg} \qquad \text{-----------------(14)}$$

or

$$N_1 = N_2 \left(\frac{V_{L1}}{V_{L2}}\right)\left(\frac{T_{chg}}{T_{dis}}\right) \qquad \text{-----------------(15)}$$

and

$$N_1 = N_2 \left(\frac{E_S - V_{sat}}{V_0 + V_D - E_S}\right)\left(\frac{T_{chg}}{T_{dis}}\right) \qquad \text{-----------------(16)}$$

From the relation that the charge energy is equal to the discharge energy;

$$\frac{1}{2} L_1 I_{p1}^2 = \frac{1}{2} L_2 I_{p2}^2 \qquad \text{-----------------(17)}$$

which yields;

$$\frac{I_{p1}}{I_{p2}} = \frac{N_2}{N_1} = n \qquad \text{-----------------(18)}$$

and

$$I_{p1} = 2 n I_0 \frac{T}{T_{dis}} \qquad \text{-----------------(19)}$$

The equations basic to design are:

$$L_2 = \left(V_o + V_D - E_s\right)\left(\frac{T_{dis}^2}{2 I_o T}\right) \qquad \text{-----------------(20)}$$

$$N_1 = N_2\left(\frac{E_s - V_{sat}}{V_o + V_D - E_s}\right)\left(\frac{T_{chg}}{T_{dis}}\right) \qquad \text{-----------------(21)}$$

$$I_{p1} = \left(\frac{N_2}{N_1}\right)\left(2 I_o\right)\left(\frac{T}{T_{dis}}\right) \qquad \text{-----------------(22)}$$

APPENDIX F

Photographic Data for the 100W IBR

FIGURE 1. Input Current, 25V

2. Input Current, 38V

3. Input Current, 50V

4. Input Capacitor Voltage, 25V

5. Input Capacitor Voltage, 38V

6. Input Capacitor Voltage, 50V

7. Primary Current in Charging Choke, 25V

8. Primary Current in Charging Choke, 38V

9. Primary Current in Charging Choke, 50V

10. Overwind Current in Charging Choke, 25V

11. Overwind Current in Charging Choke, 38V

12. Overwind Current in Charging Choke, 50V

13. Emitter Current in Switching Transistor (Q1), 25V

14. Emitter Current in Switching Transistor (Q1), 38V

15. Emitter Current in Switching Transistor (Q1), 50V

16. Collector Emitter Voltage (Q1), 25V

17. Collector Emitter Voltage (Q1), 38V

18. Collector Emitter Voltage (Q1), 50V

19. Feedback Winding Voltage, 25V

20. Feedback Winding Voltage, 38V

21. Feedback Winding Voltage, 50V

22. Feedback Current in Charging Choke, 25V

23. Feedback Current in Charging Choke, 38V

24. Feedback Current in Charging Choke, 50V

25. Output Capacitor Voltage, 25V

26. Output Capacitor Voltage, 38V

27. Output Capacitor Voltage, 50V

28. Input Current Transient

29. Load Transient Response, 10%-50%, 25V

30. Load Transient Response, 50%-100%, 25V

31. Load Transient Response, 10-50, 38V

32. Load Transient Response, 50-100, 38V

33. Load Transient Response, 10-50, 50V

34. Load Transient Response, 50-100, 50V

35. Load Transient Response, 10-50, 25V (200W)

36. Load Transient Response, 50-100, 25V (200W)

37. Load Transient Response, 10-50, 38V (200W)

38. Load Transient Response, 50-100, 38V (200W)

39. Load Transient Response, 10-50, 50V (200W)

40. Load Transient Response, 50-100, 50V (200W)

41. Line Transient Response, 0V-50V

42. Line Transient Response, 0V-25V

43. Line Transient Response, 0V-25V/50V (200W)

44. Line Transient Response, 25V-38V No Load

45. Line Transient Response, 35V-50V No Load

46. Line Transient Response, 38V-25V Full Load

47. Line Transient Response, 25V-50V Full Load

48. Line Transient Response, 25V-38V Full Load

49. Line Transient Response, 50V-25V Full Load

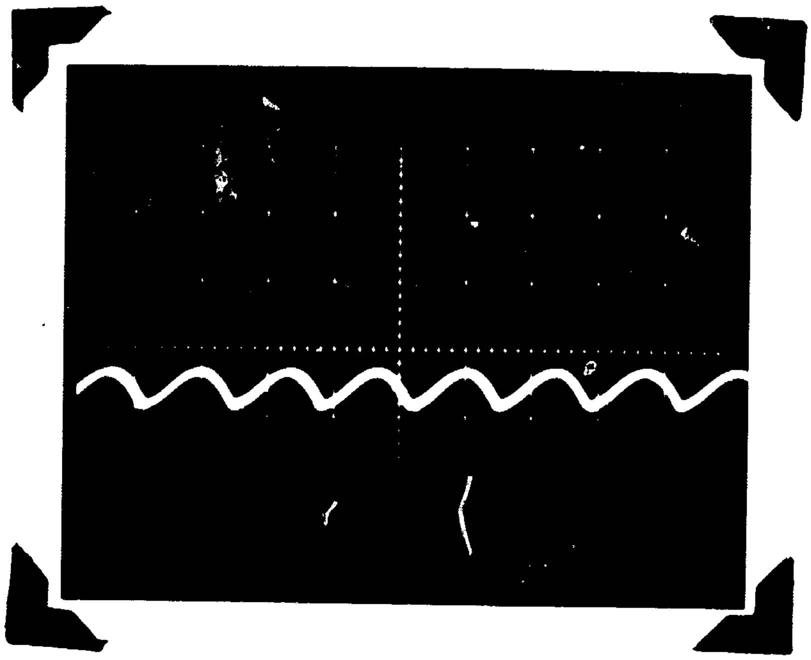
50. Line Transient Response, 38V-25V No Load

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORMS

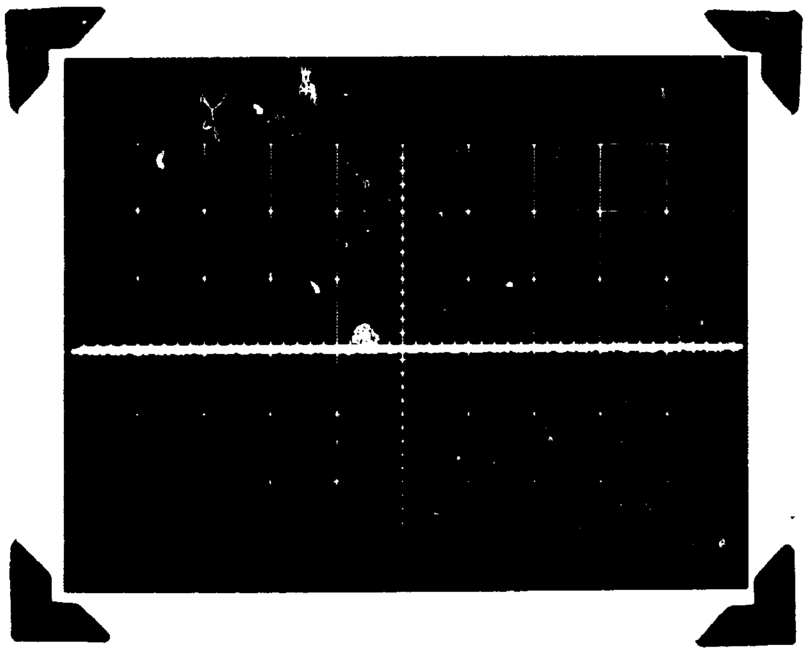
L_1 INPUT CURRENT @50 V

200 ma/cm



200 us/cm

20 ma/cm



100 us/cm

10% LOAD

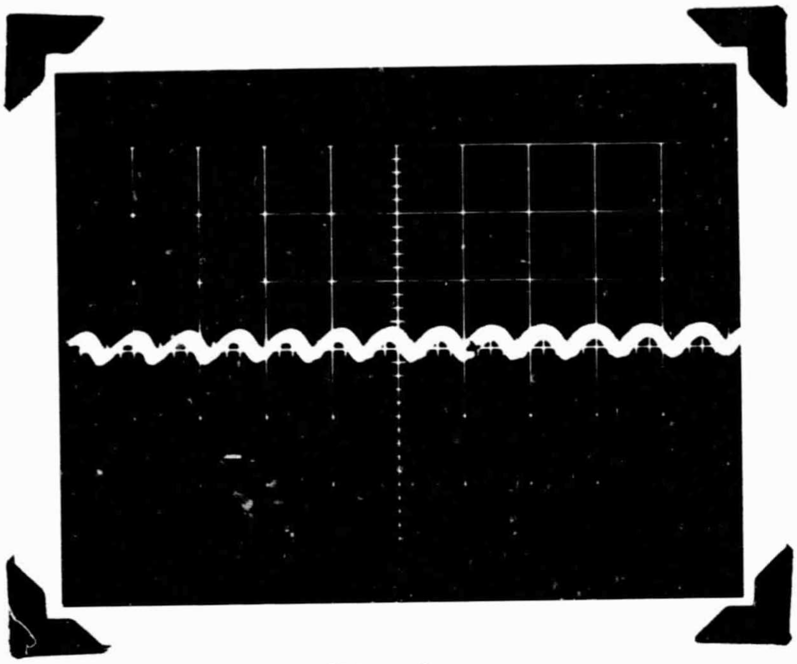
FIGURE 1

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORMS

L_1 INPUT CURRENT @38 V

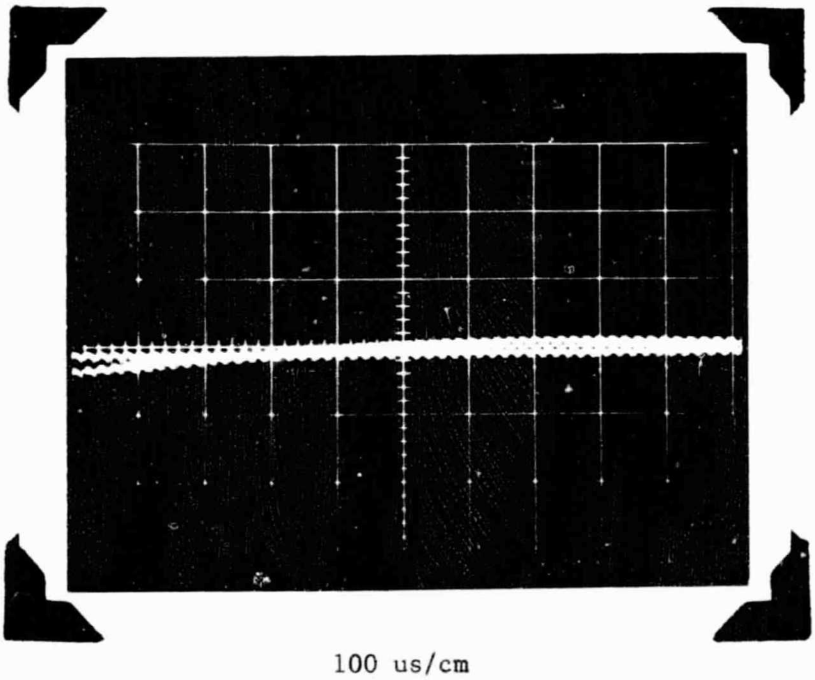
200 ma/cm



200 us/cm

100% LOAD

20 ma/cm



100 us/cm

10% LOAD

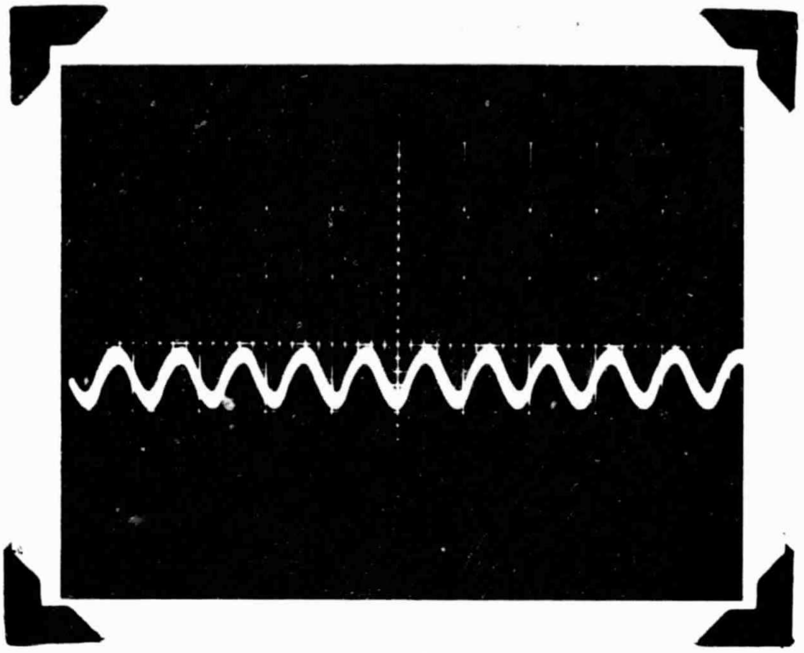
FIGURE 2

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORMS

L_1 INPUT CURRENT @25 V

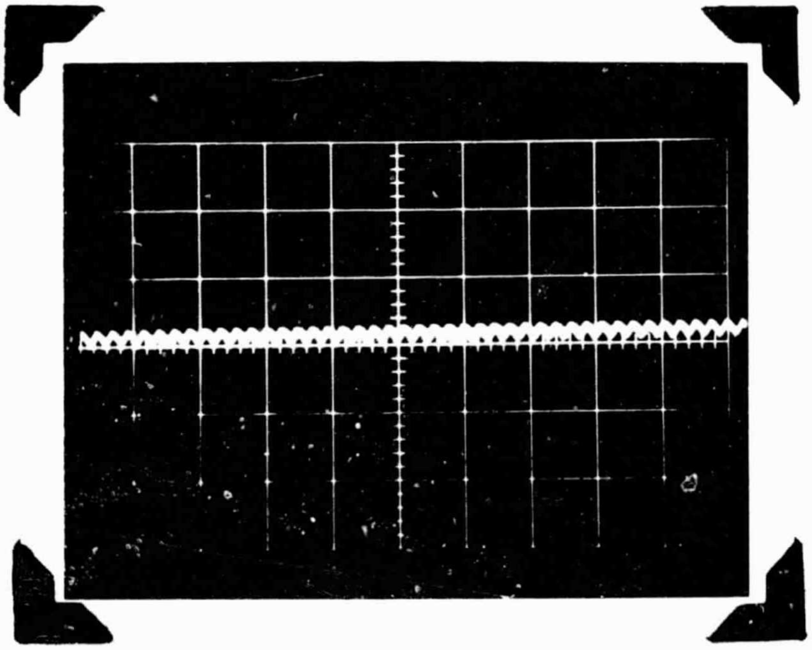
200 ma/cm



200 us/cm

100% LOAD

20 ma/cm



100 us/cm

10% LOAD

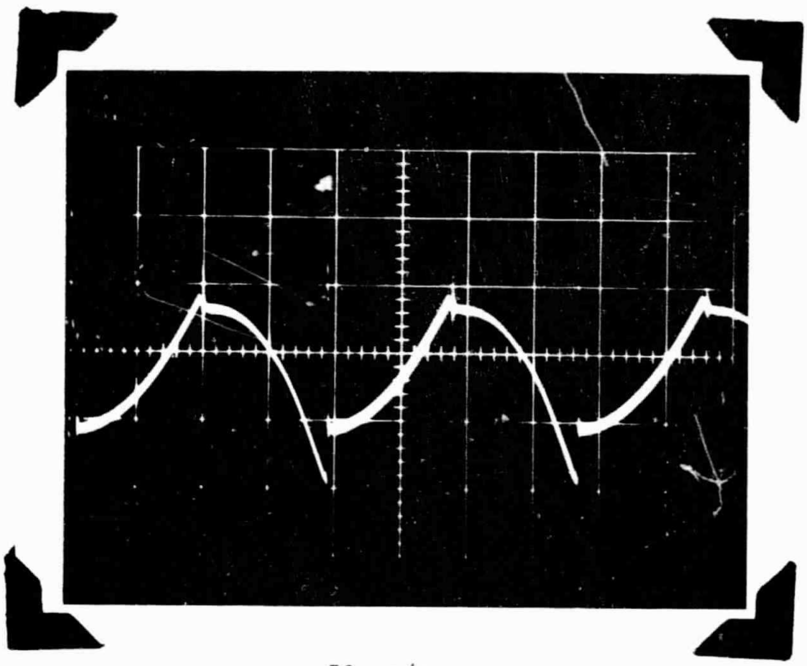
FIGURE 3

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CAPACITOR VOLTAGE @ 25V

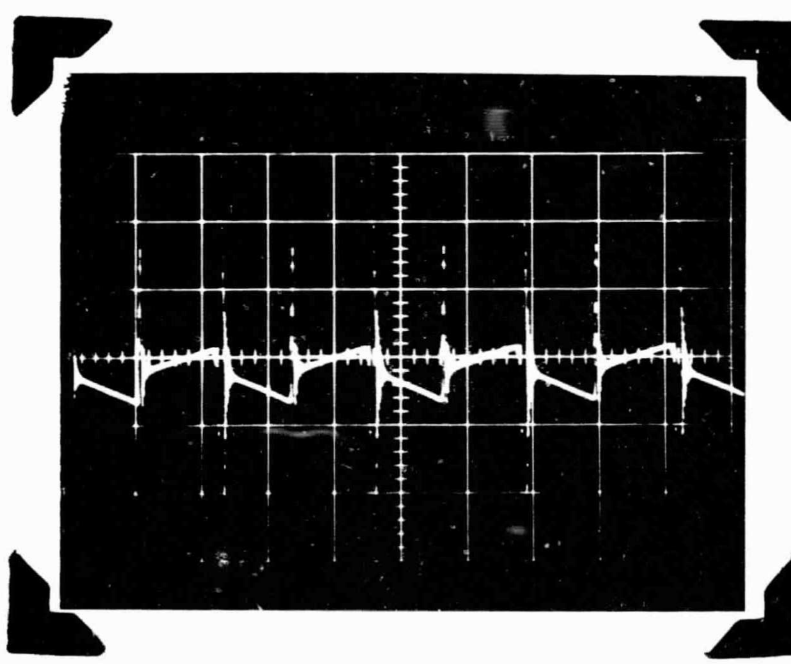
.2V/cm



50 us/cm

100% LOAD

.1V/cm



10 us/cm

10% LOAD

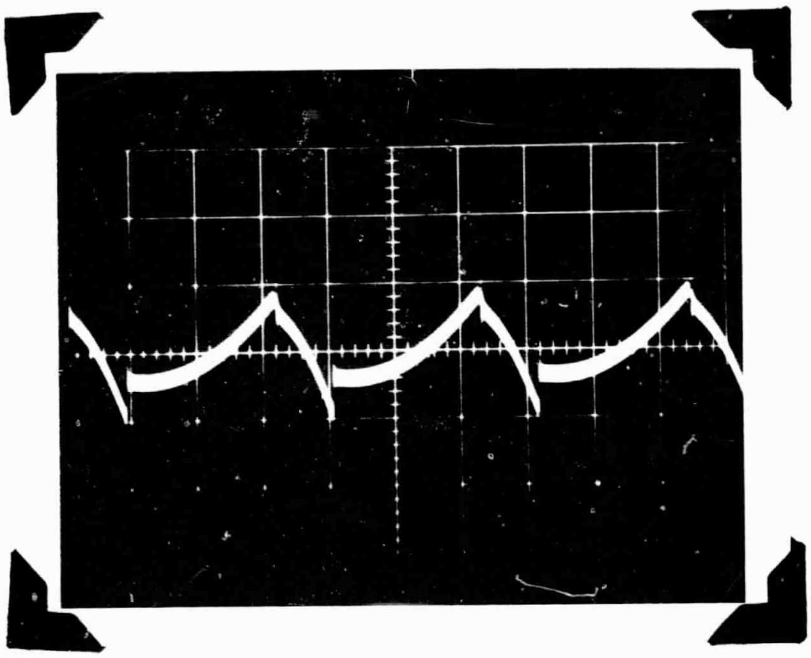
FIGURE 4

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CAPACITOR VOLTAGE @ 38V

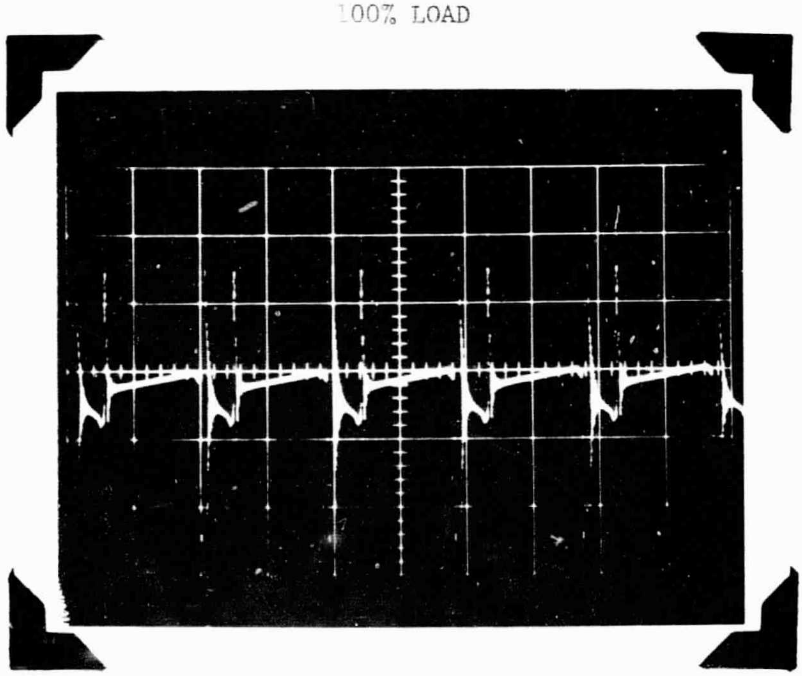
.2V/cm



50 us/cm

100% LOAD

.1V/cm



10 us/cm

10% LOAD

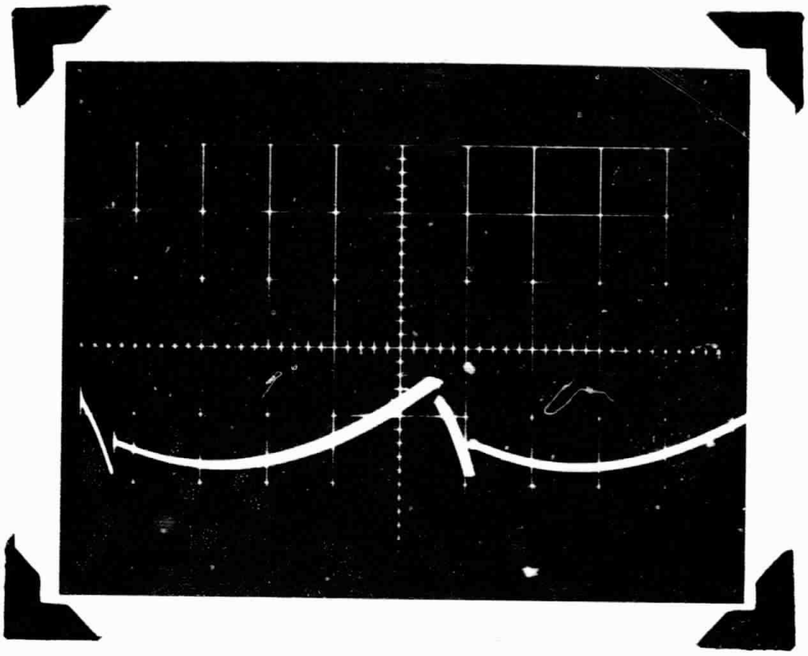
FIGURE 5

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CAPACITOR VOLTAGE @ 50V

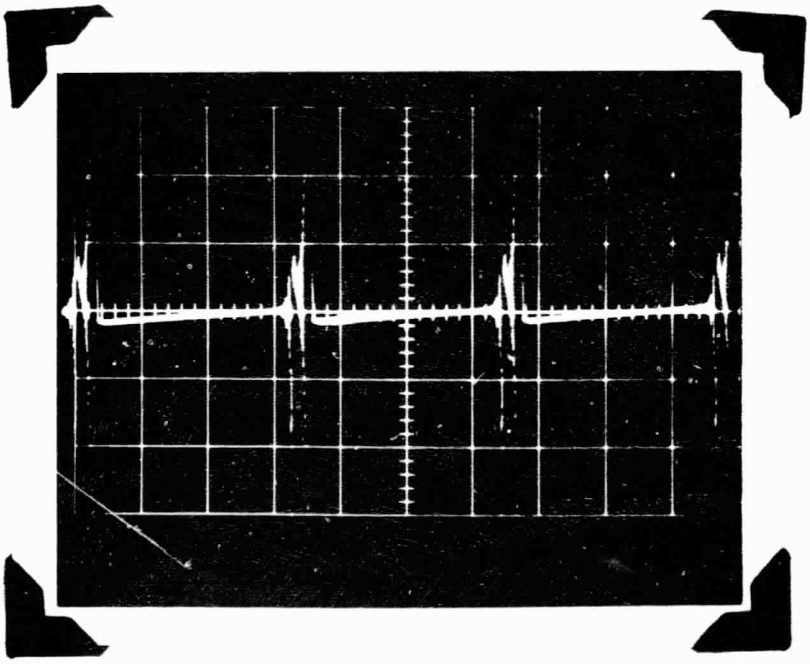
.2V/cm



50 us/cm

100% LOAD

.1V/cm



10 us/cm

10% LOAD

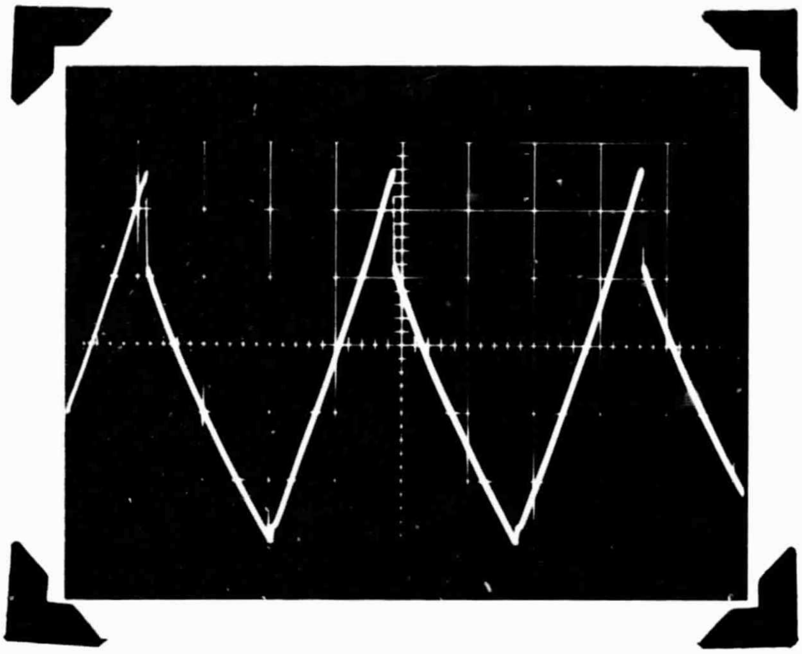
FIGURE 6

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CURRENT CHARGING CHOKE @ 25V

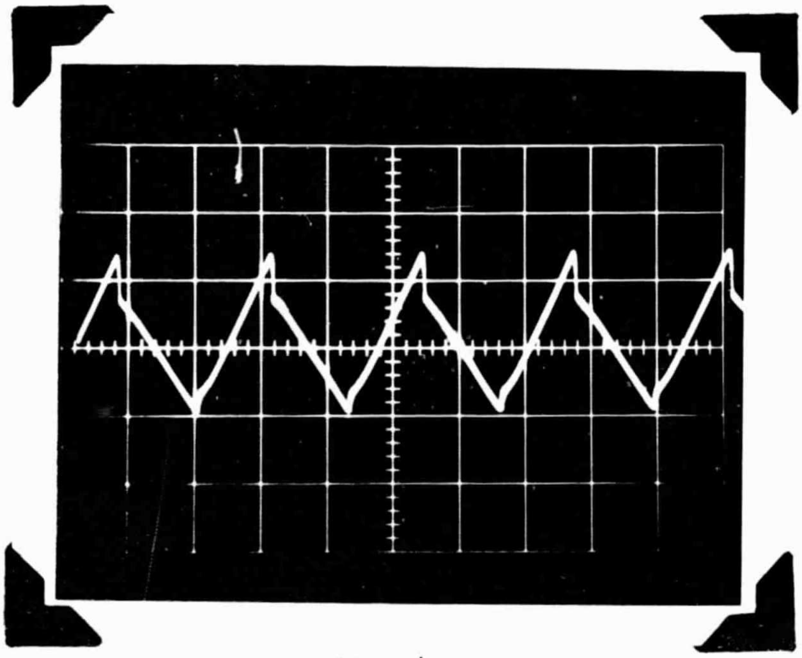
2.0 A/cm



50 us/cm

100% LOAD

.5 A/cm



10 us/cm

10% LOAD

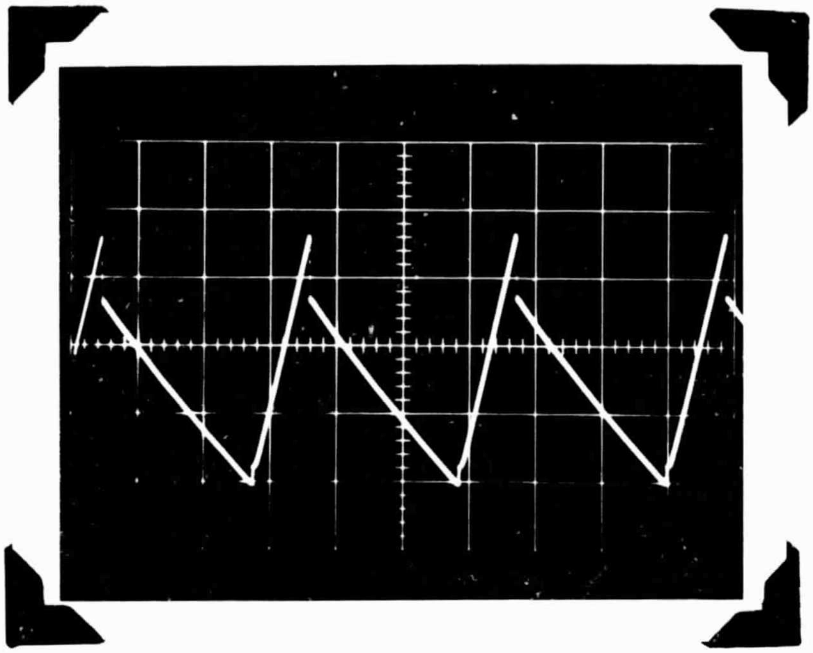
FIGURE 7

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CURRENT CHARGING CHOKE @ 38V

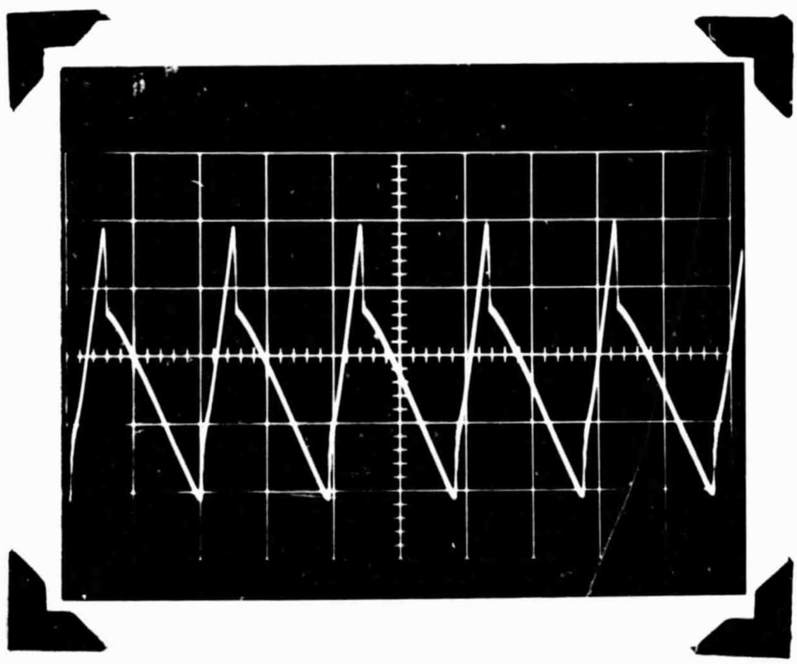
2.0 A/cm



50 us/cm

100% LOAD

.2A/cm



10 us/cm

10% LOAD

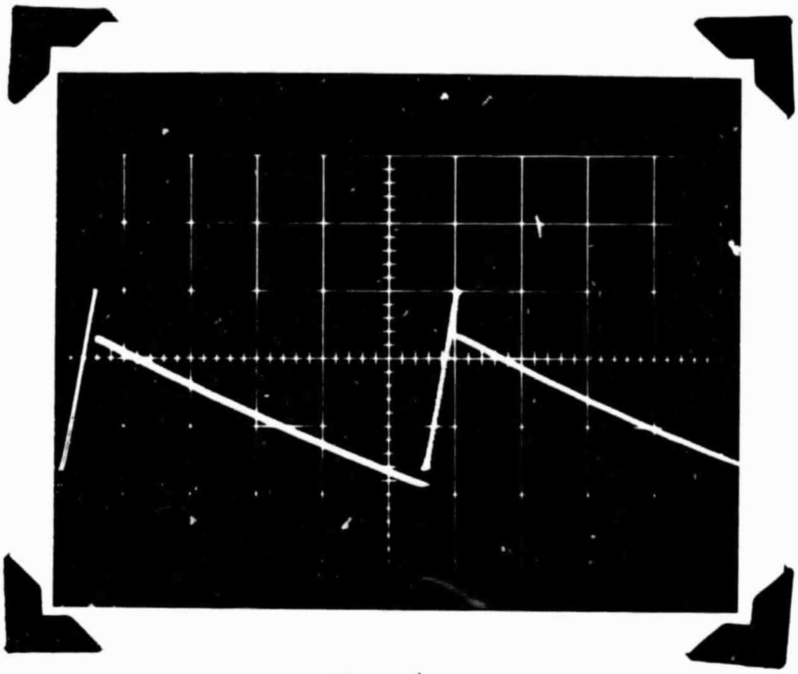
FIGURE 8

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CURRENT CHARGING CHOKE @ 50V

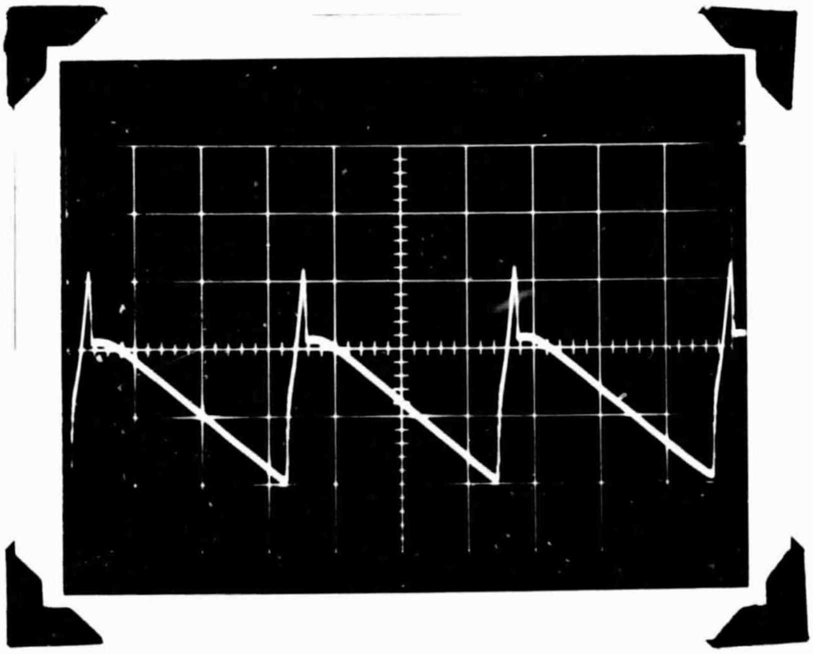
2.0 Amp/cm



50 us/cm

100% LOAD

.2 A/cm



10 us/cm

10% LOAD

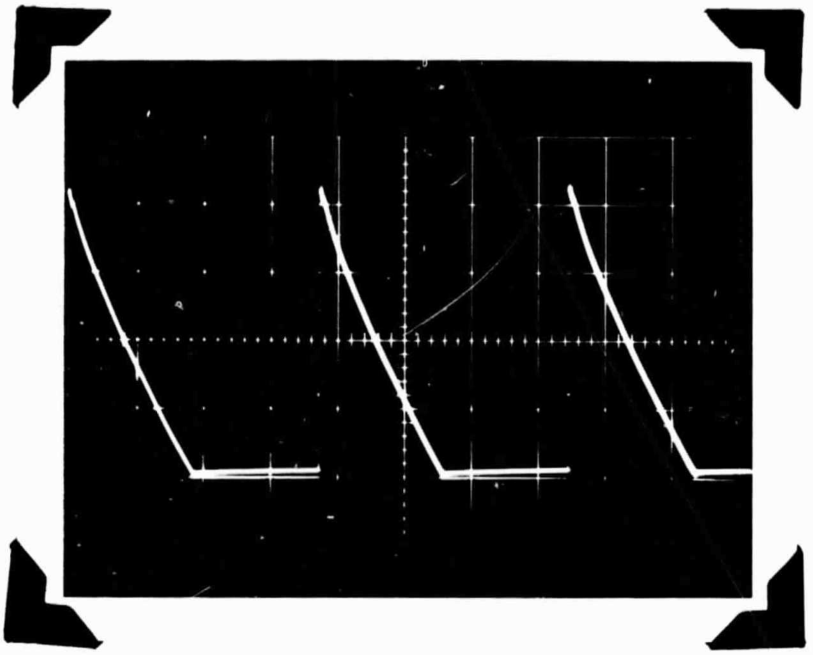
FIGURE 9

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

OUTPUT CURRENT CHARGING CHOKE @ 25V

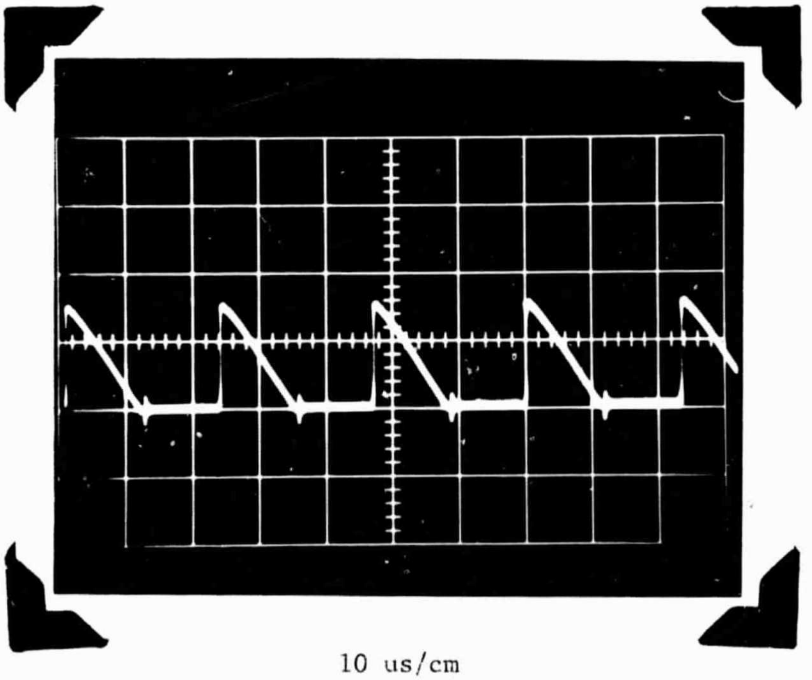
2.0 Amp/cm



50 us/cm

100% LOAD

.5 Amps/cm



10 us/cm

10% LOAD

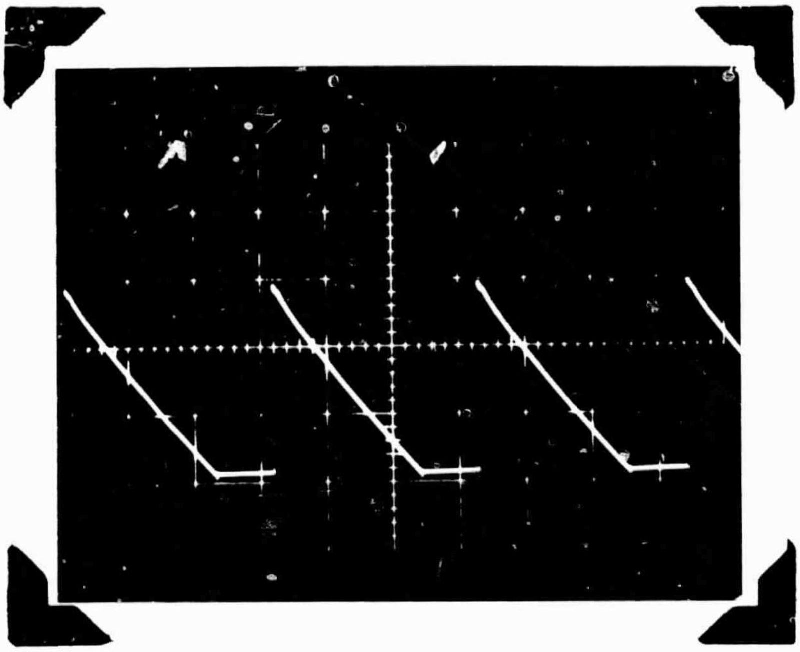
FIGURE 10

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

OUTPUT CURRENT CHARGING CHOKE @ 38V

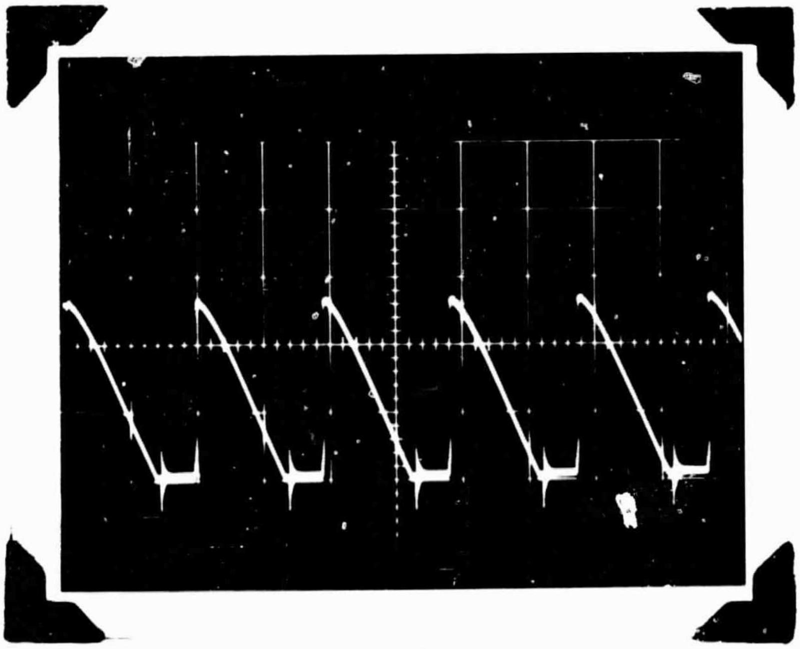
2.0 Amp/cm



50 us/cm

100% LOAD

.2 Amp/cm



10 us/cm

10% LOAD

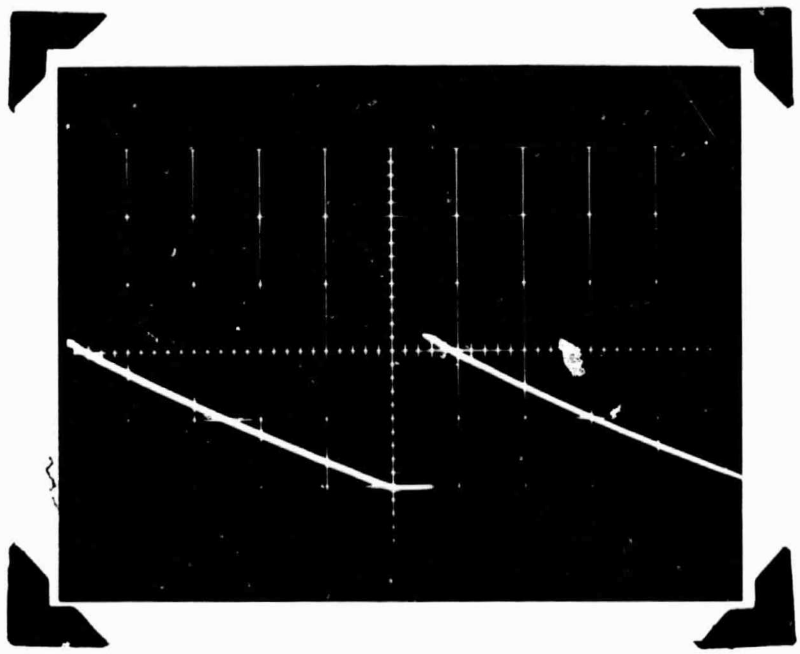
FIGURE 11

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

OUTPUT CURRENT CHARGING CHOKE @ 50V

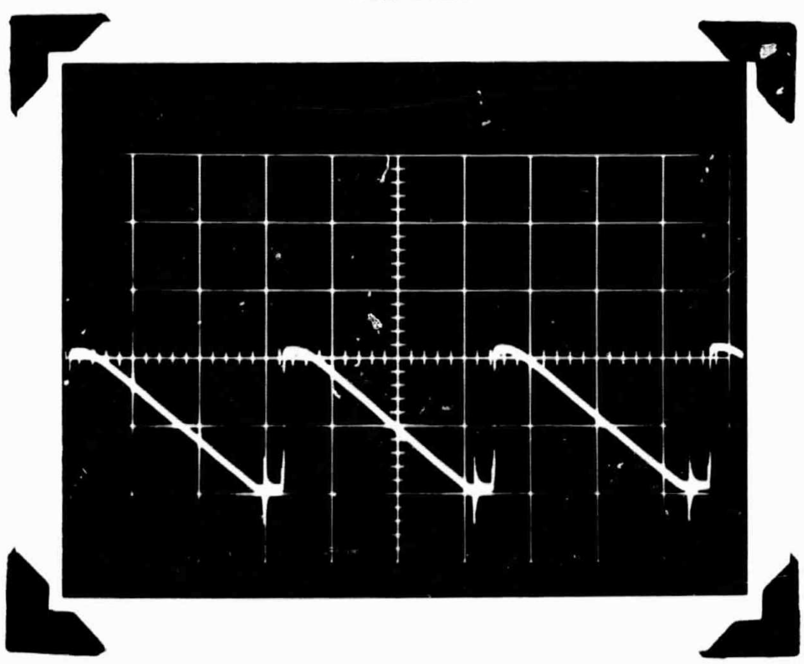
2 Amp/cm



50 us/cm

100% LOAD

.2 Amp/cm



10 us/cm

10% LOAD

FIGURE 12

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

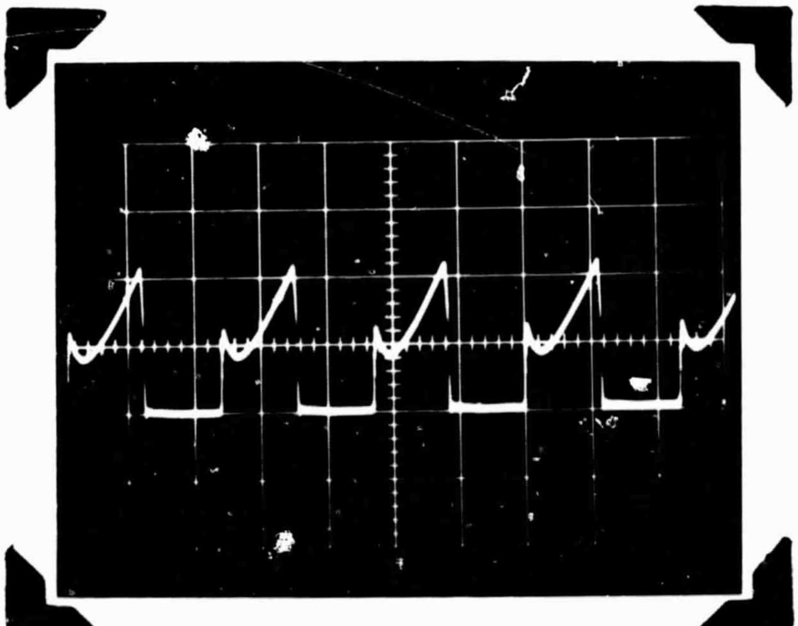
EMITTER CURRENT SWITCHING TRANSISTOR @ 25V

2.Amps/cm

50 us/cm

100% LOAD

.5 Amp/cm



10 us/cm

10% LOAD

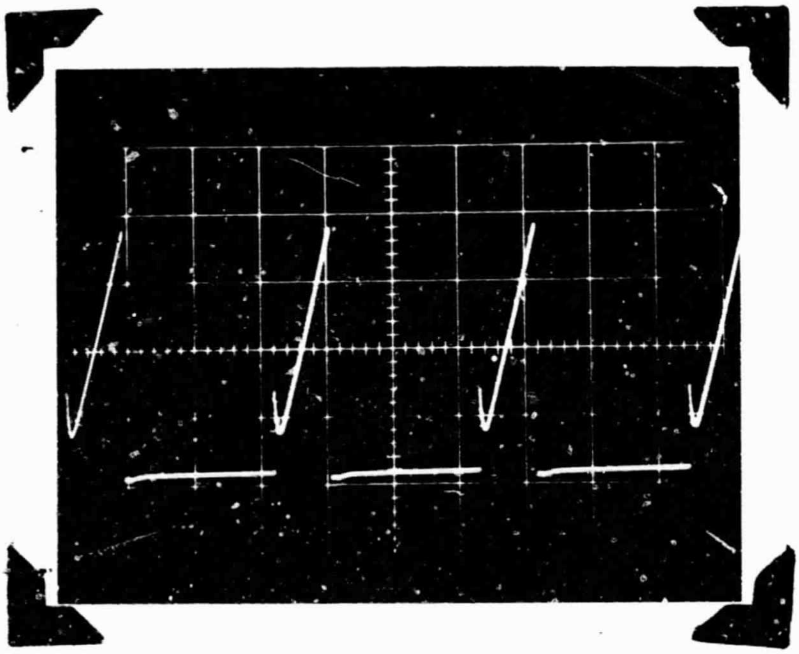
FIGURE 13

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

EMITTER CURRENT SWITCHING TRANSISTOR @ 38V

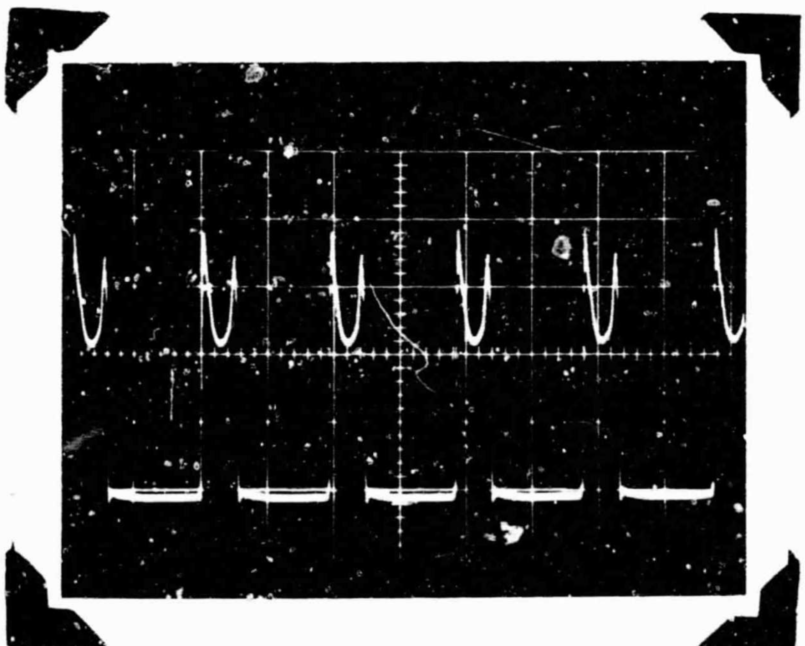
2.Amp/cm



50 us/cm

100% LOAD

.2 Amp/cm



10 us/cm

10% LOAD

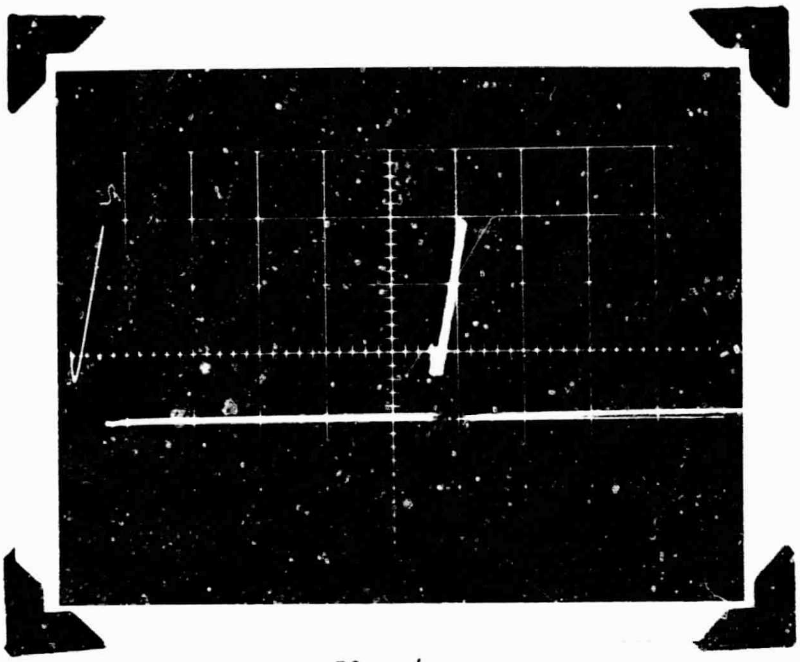
FIGURE 14

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

EMITTER CURRENT SWITCHING TRANSISTOR @ 50V

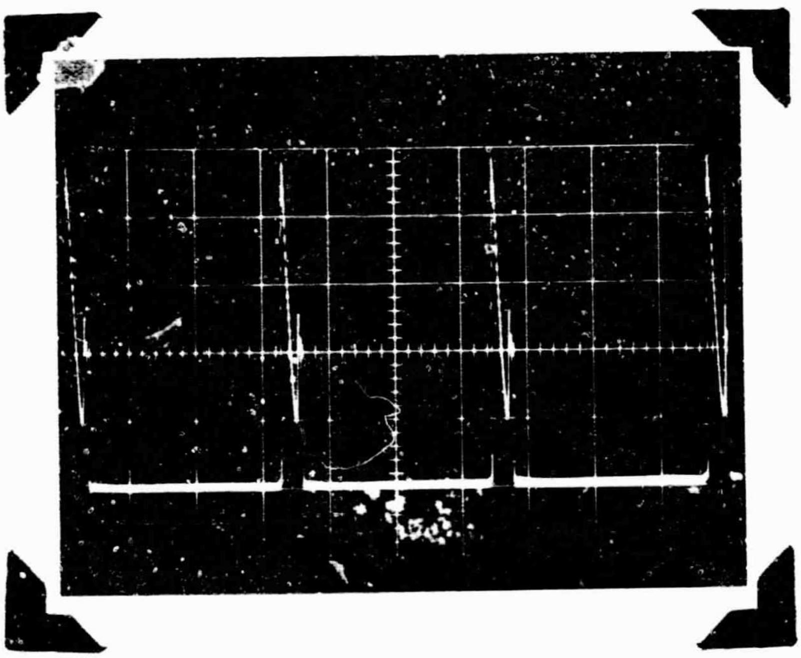
2.Amps/cm



50 us/cm

100% LOAD

.2AMPS/cm



10 us/cm

10% LOAD

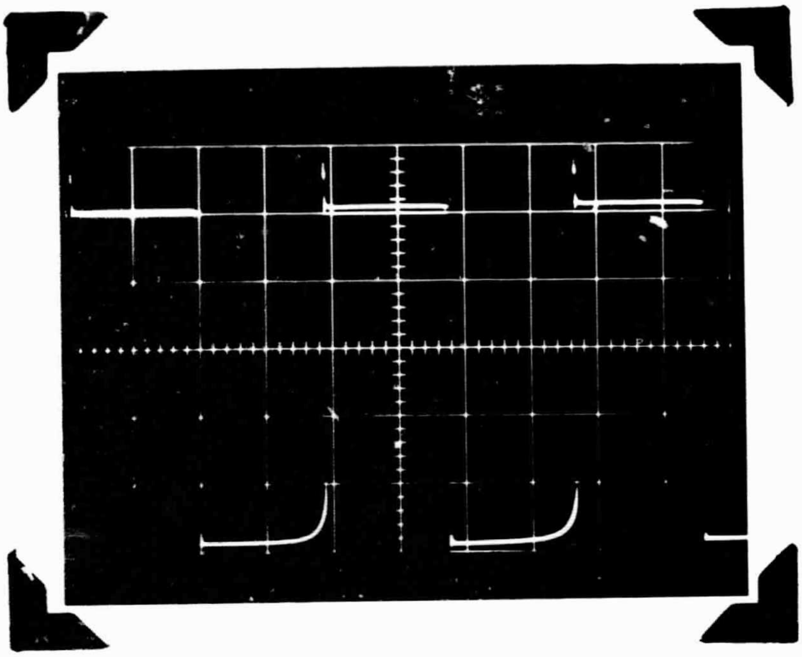
FIGURE 15

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

V.C.E. SWITCHING TRANSISTOR @ 25V

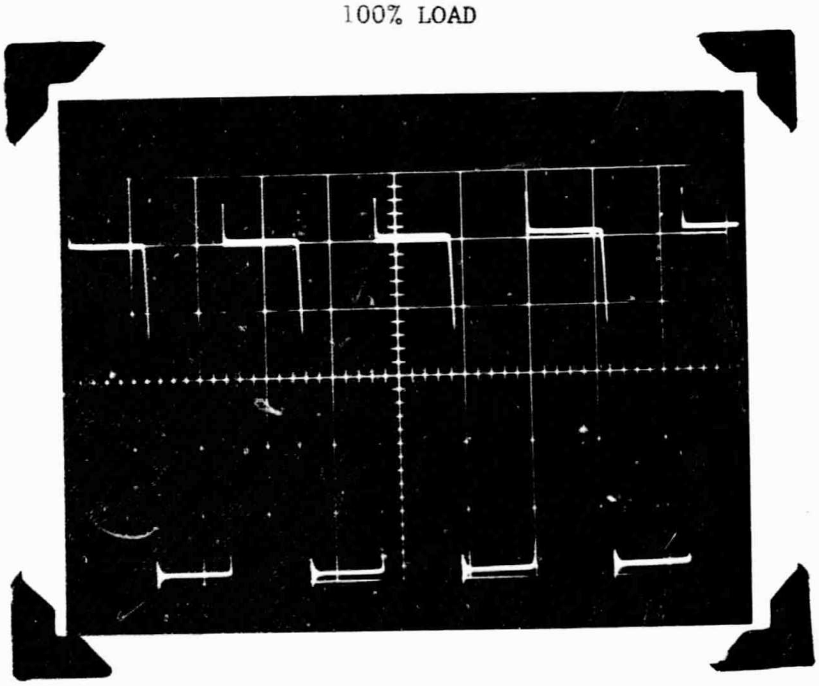
10V/cm



50 us/cm

100% LOAD

10V/cm



10 us/cm

10% LOAD

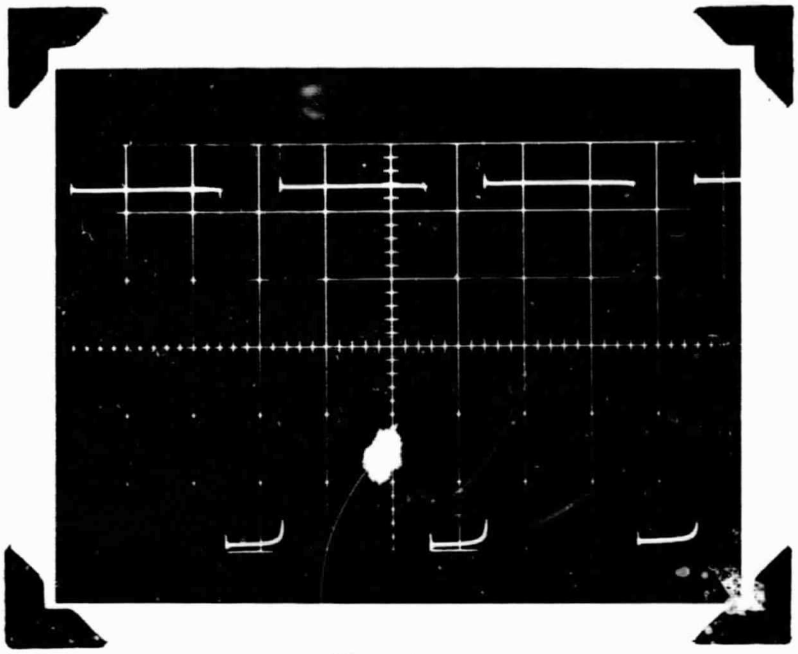
FIGURE 16

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

V.C.E. SWITCHING TRANSISTOR @ 38V

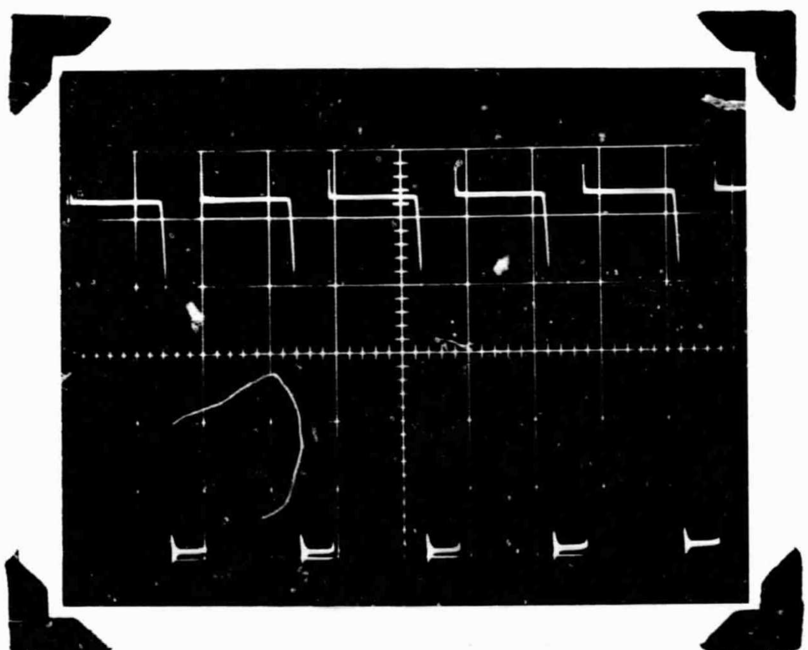
10V/cm



50 us/cm

100% LOAD

10V/cm



10 us/cm

10% LOAD

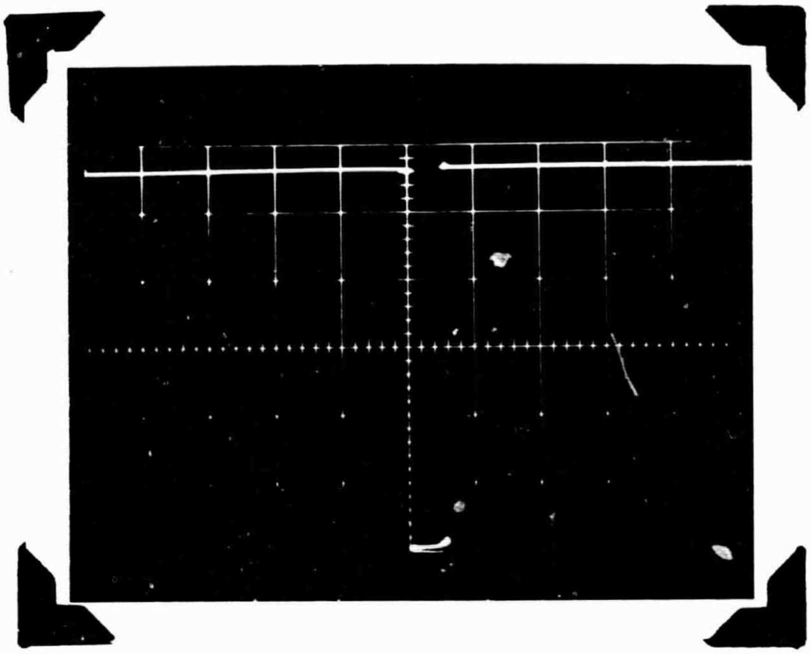
FIGURE 17

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

V.C.E. SWITCHING TRANSISTOR @ 50V

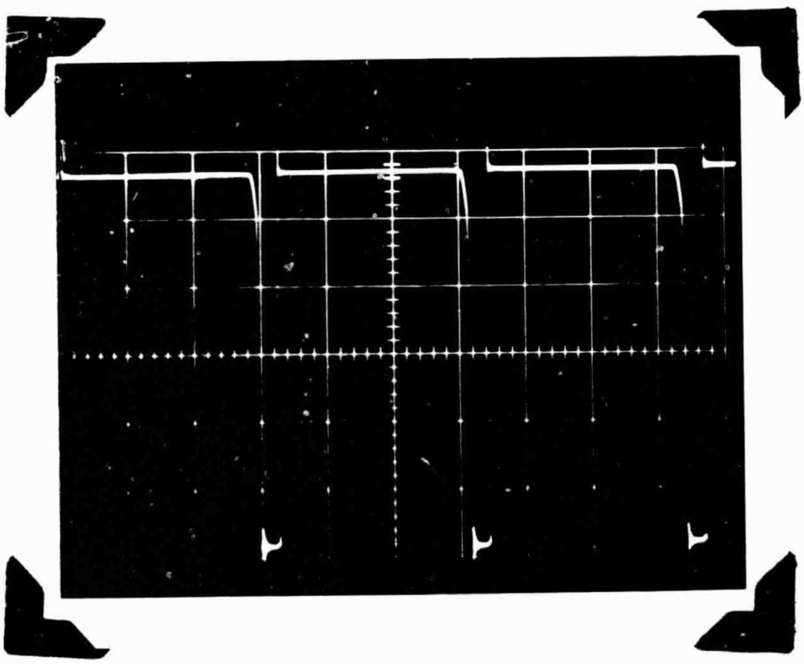
10V/cm



50 us/cm

100% LOAD

10V/cm



10 us/cm

10% LOAD

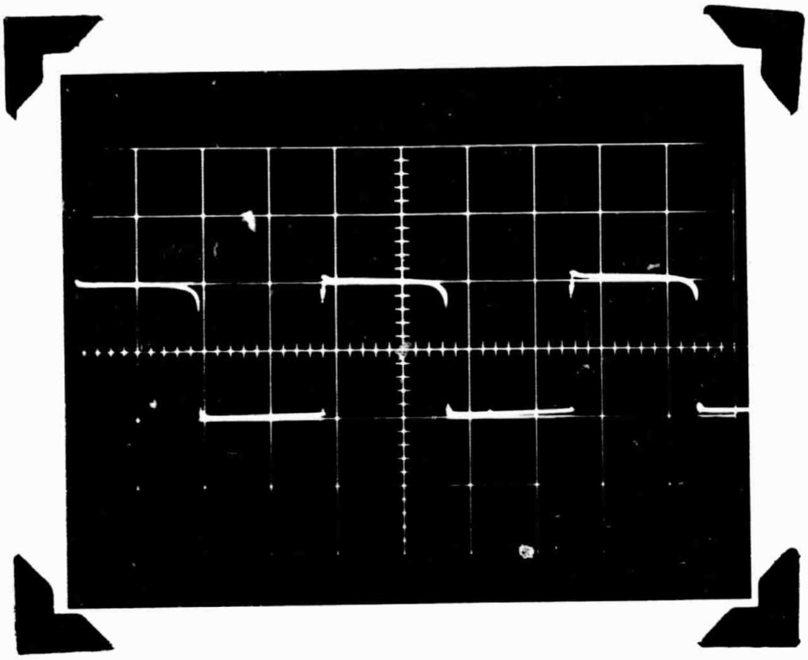
FIGURE 18

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

FEEDBACK WINDING VOLTAGE @ 25V

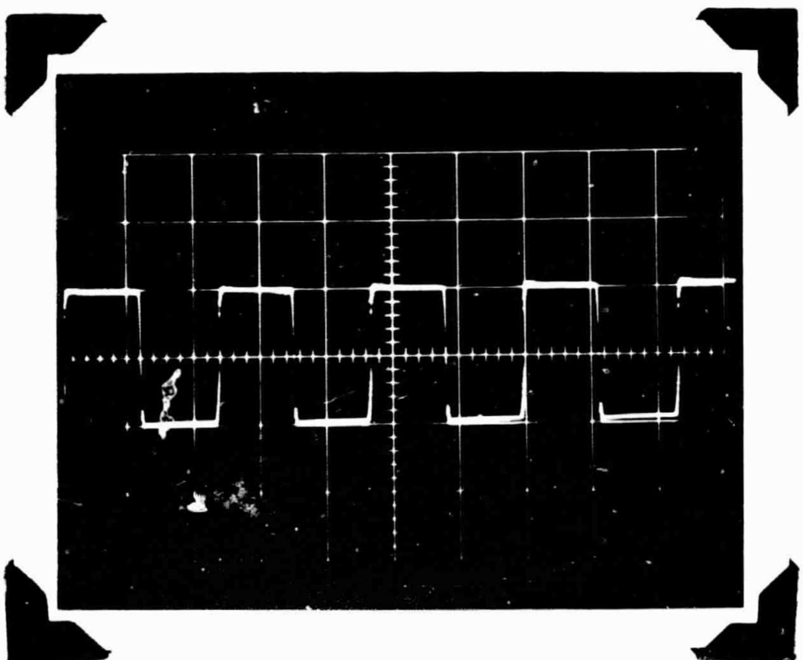
5V/cm



50 us/cm

100% LOAD

5V/cm



10 us/cm

10% LCAD

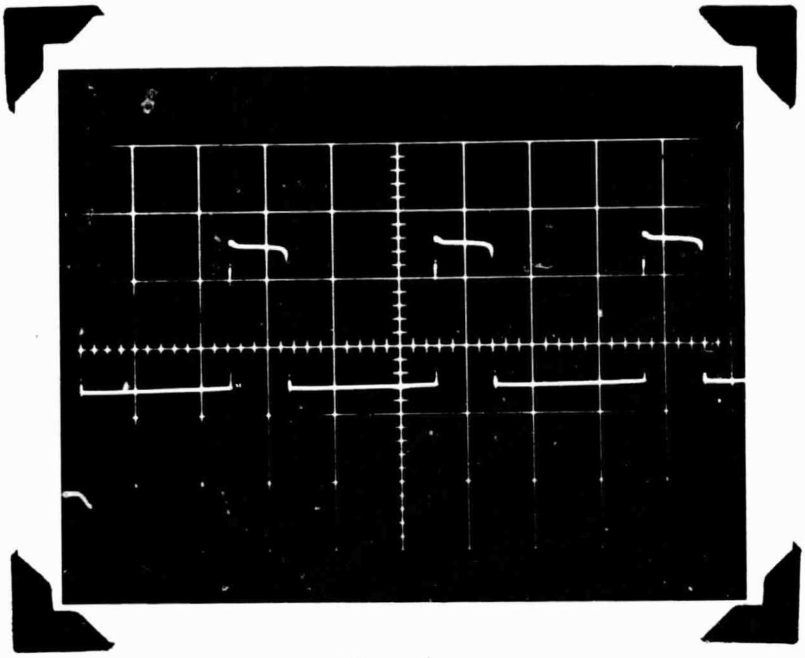
FIGURE 19

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

FEEDBACK WINDING VOLTAGE @ 38V

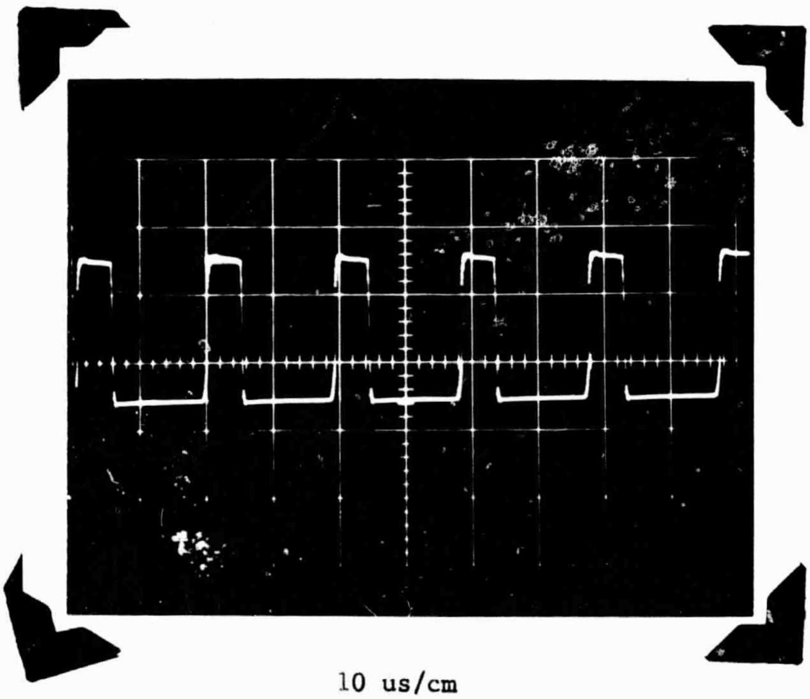
5V/cm



50 us/cm

100% LOAD

5V/cm



10 us/cm

10% LOAD

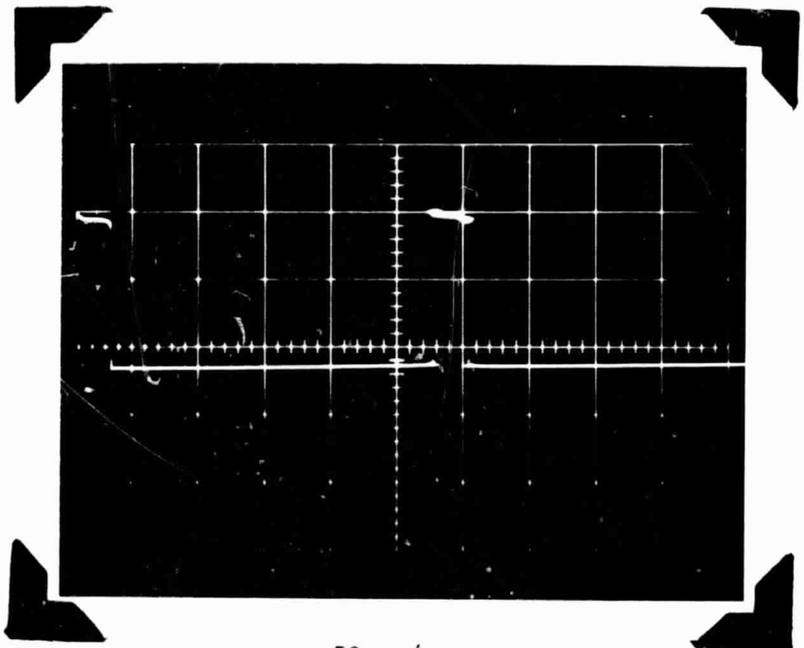
FIGURE 20

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

FEEDBACK WINDING VOLTAGE @ 50V

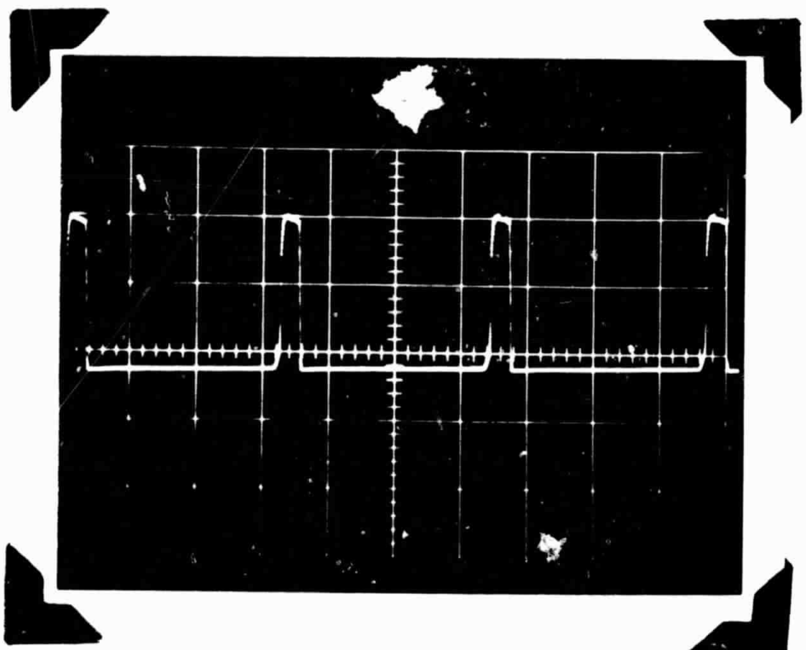
5V/cm



50 us/cm

100% LOAD

5V/cm



10 us/cm

10% LOAD

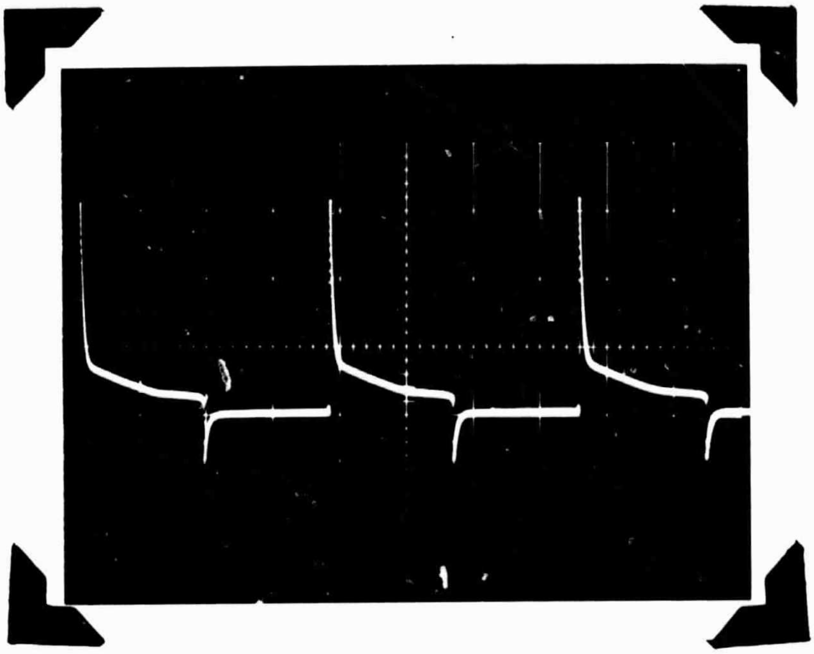
FIGURE 21

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

FEEDBACK CURRENT CHARGING CHOKE @ 25V

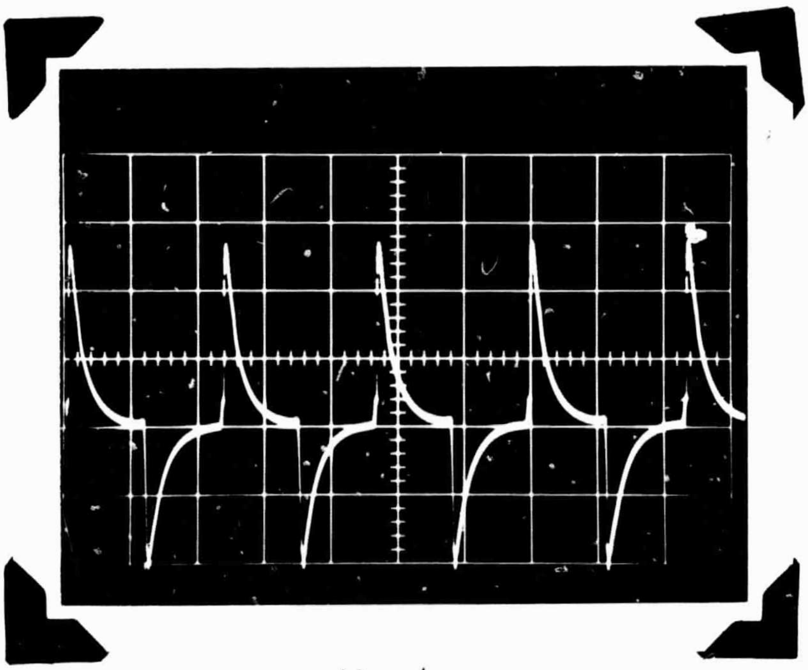
.5 Amps/cm



50 us/cm

100% LOAD

.2 Amps/cm



10 us/cm

10% LOAD

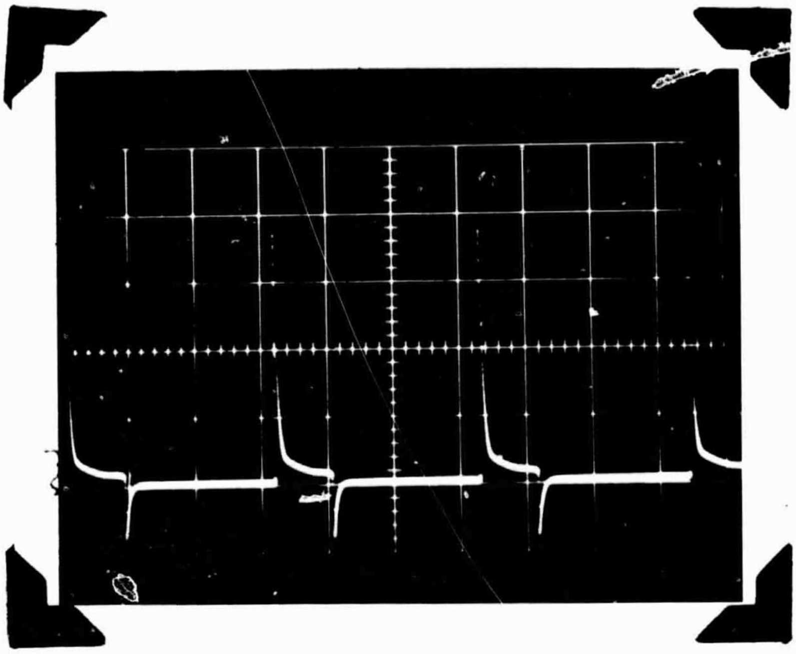
FIGURE 22

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

FEEDBACK CURRENT CHARGING CHOKE @ 38V

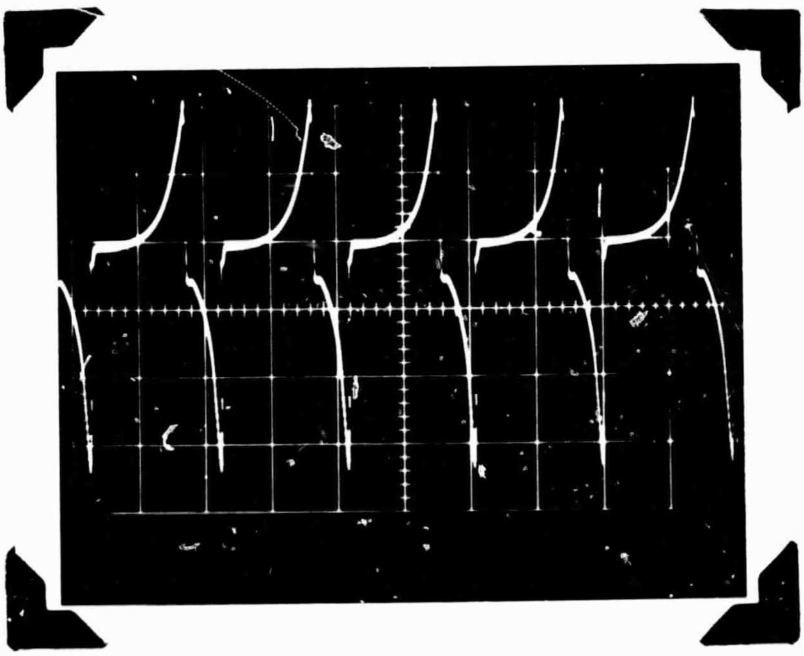
.5 Amps/cm



50 us/cm

100% LOAD

.2 Amps/cm



10 us/cm

10% LOAD

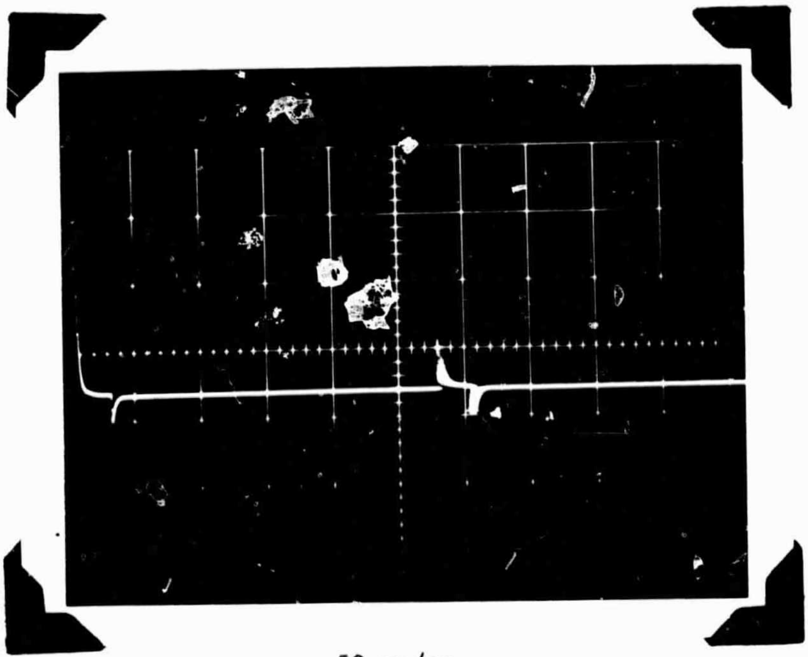
FIGURE 23

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORMS

FEEDBACK CURRENT CHARGING CHOKE @ 50V

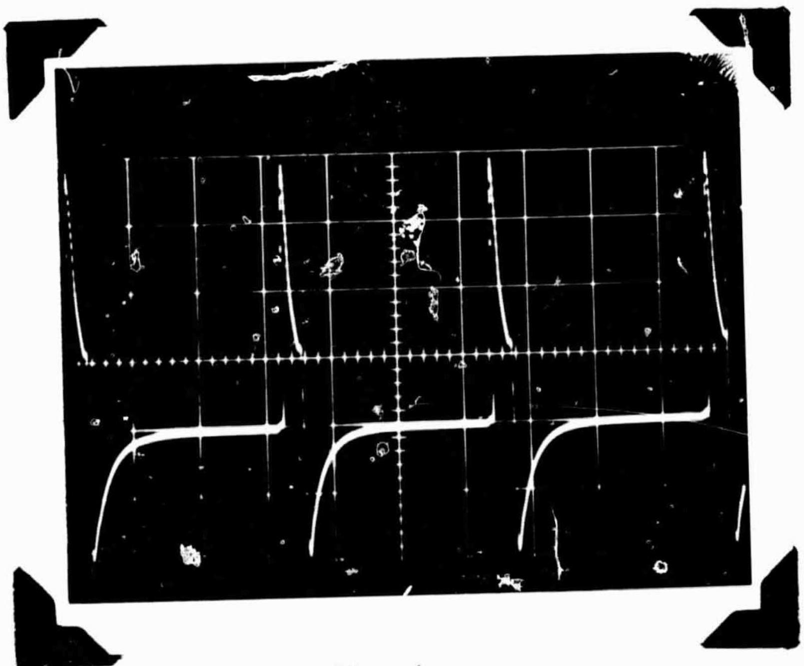
.5 Amps/cm



50 us/cm

100% LOAD

.2 Amps/cm



10 us/cm

10% LOAD

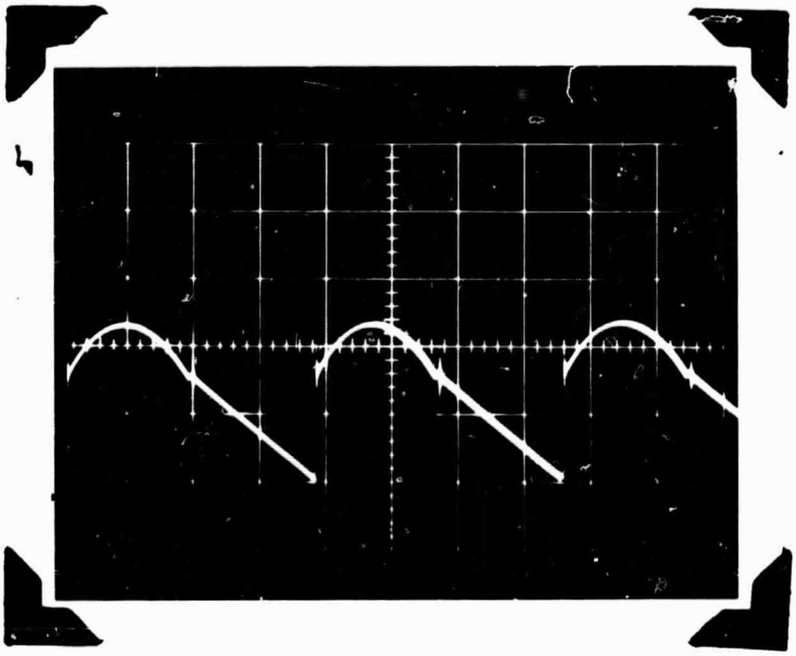
FIGURE 24

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

CHARGING CAPACITOR VOLTAGE @ 25V

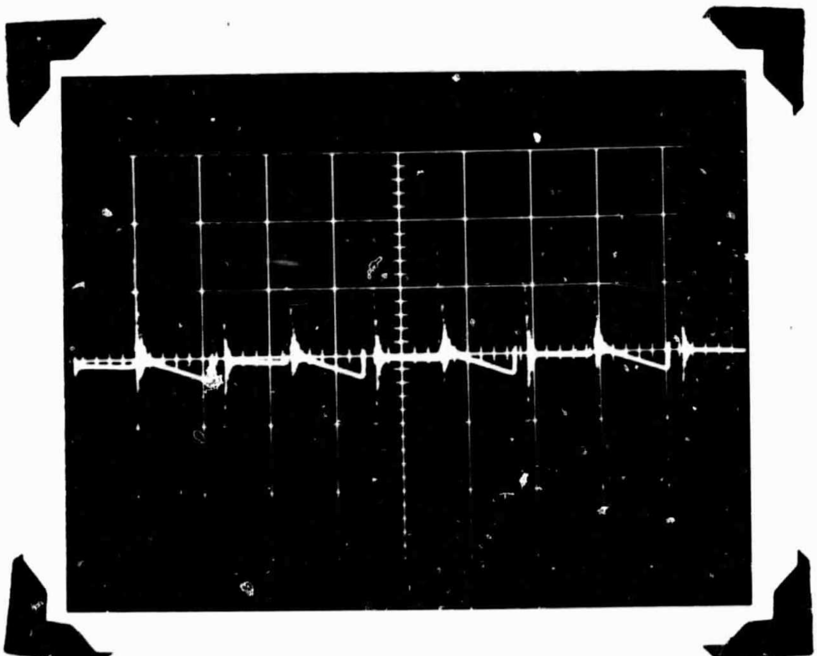
.1V/cm



50 us/cm

100% LOAD

.05V/cm



10 us/cm

10% LOAD

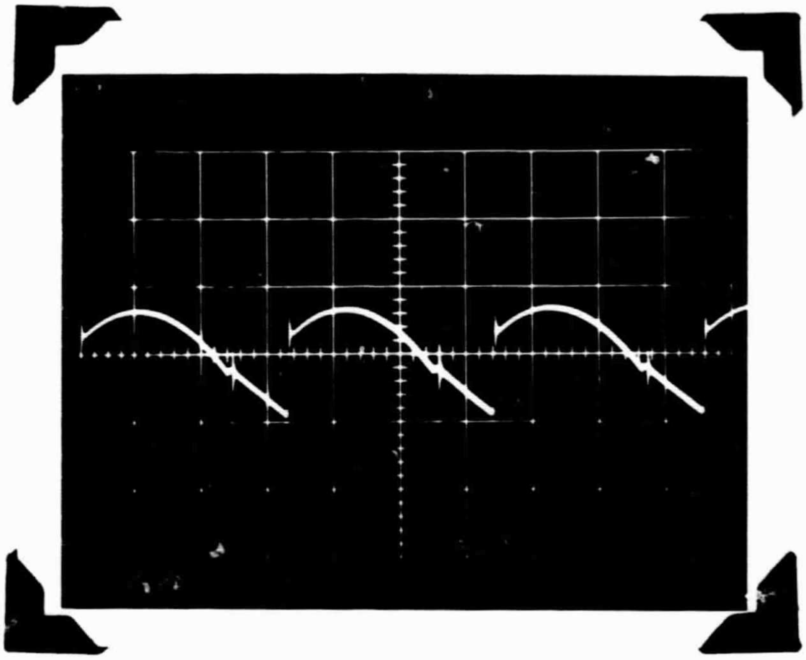
FIGURE 25

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

CHARGING CAPACITOR VOLTAGE @ 38V

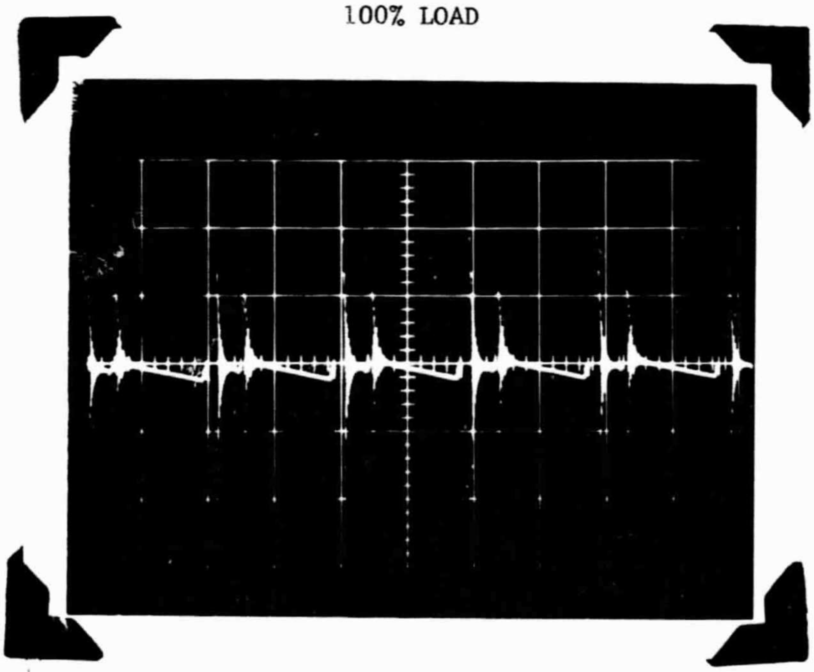
.1V/cm



50 us/cm

100% LOAD

.05V/cm



10 us/cm

10% LOAD

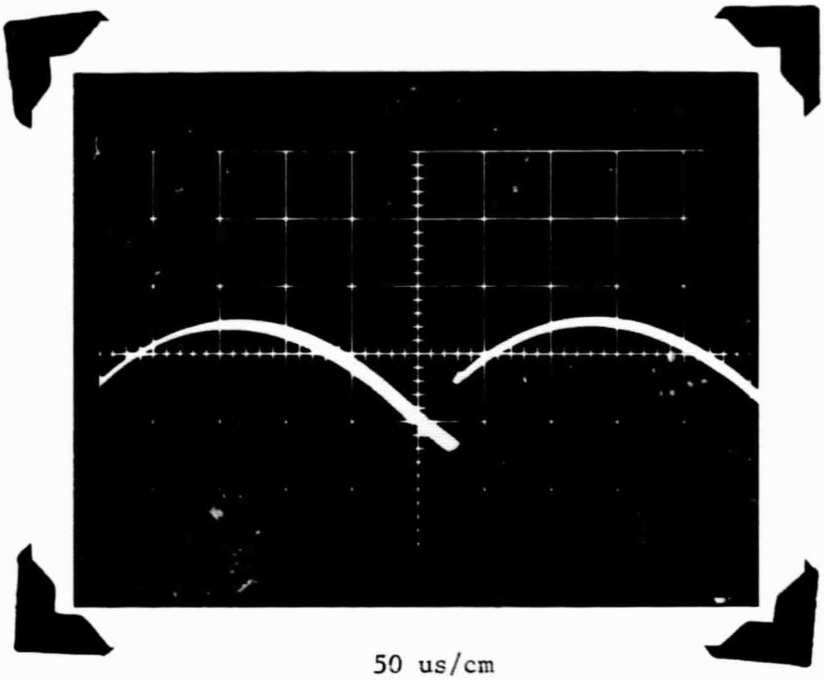
FIGURE 26

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

CHARGING CAPACITOR VOLTAGE @ 50V

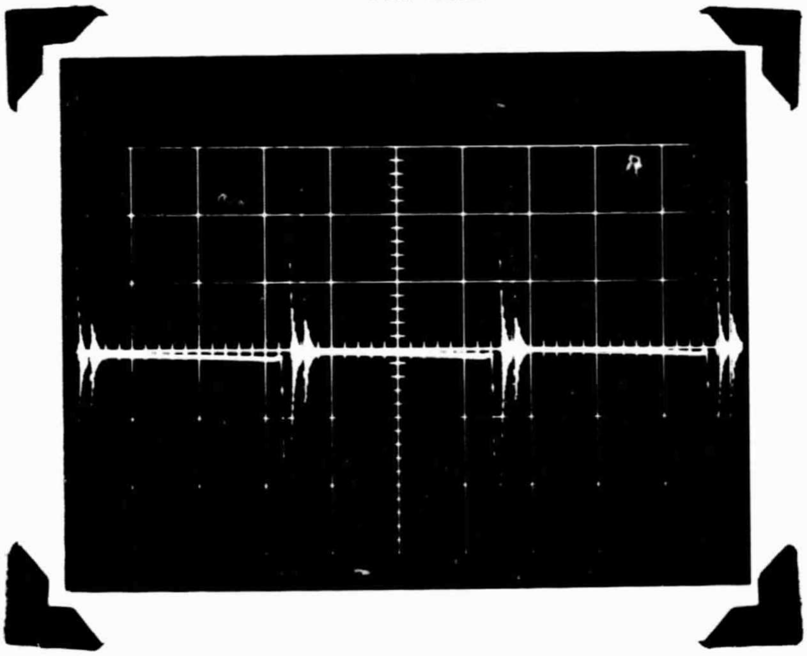
.1V/cm



50 us/cm

100% LOAD

.05V/cm



10 us/cm

10% LOAD

FIGURE 27

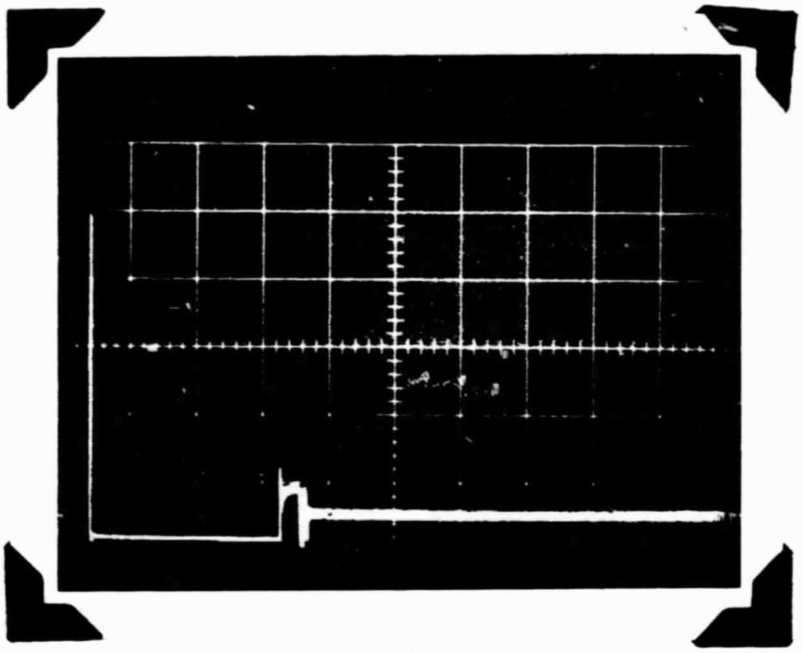
IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

INPUT CURRENT TRANSIENT

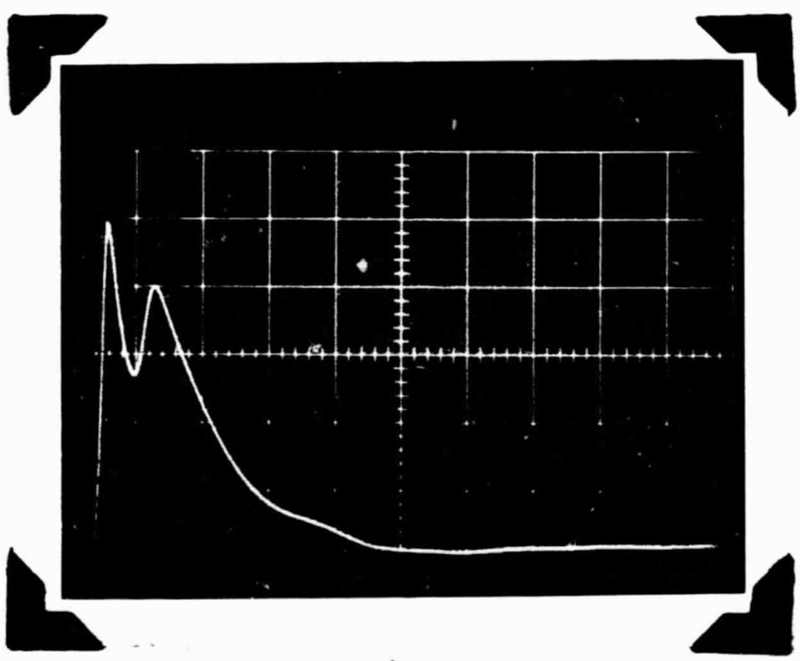
MEASURED ACROSS .1 Ω

1 V/cm



100 ms/cm

1 V/cm



1 ms/cm

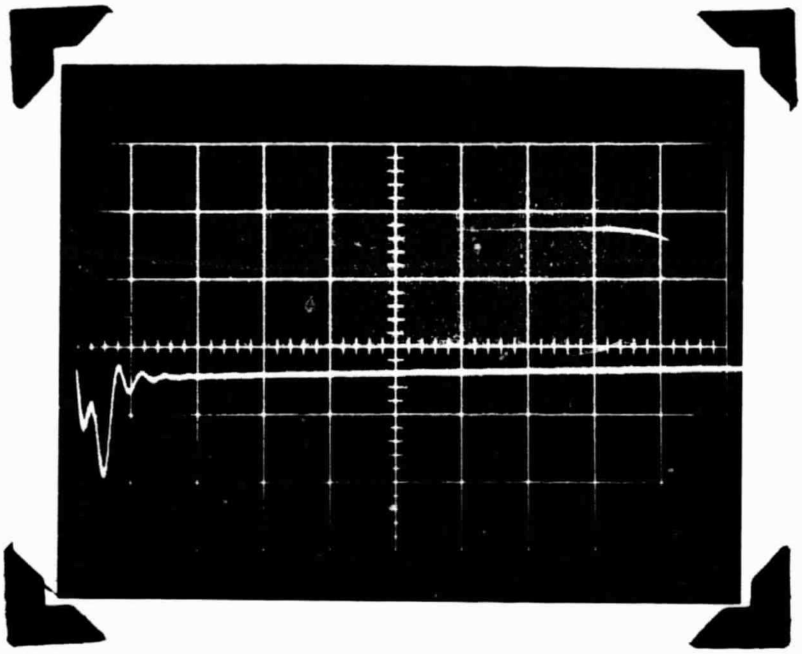
SAME AS ABOVE BUT SWEEP
SPEED INCREASED

FIGURE 28

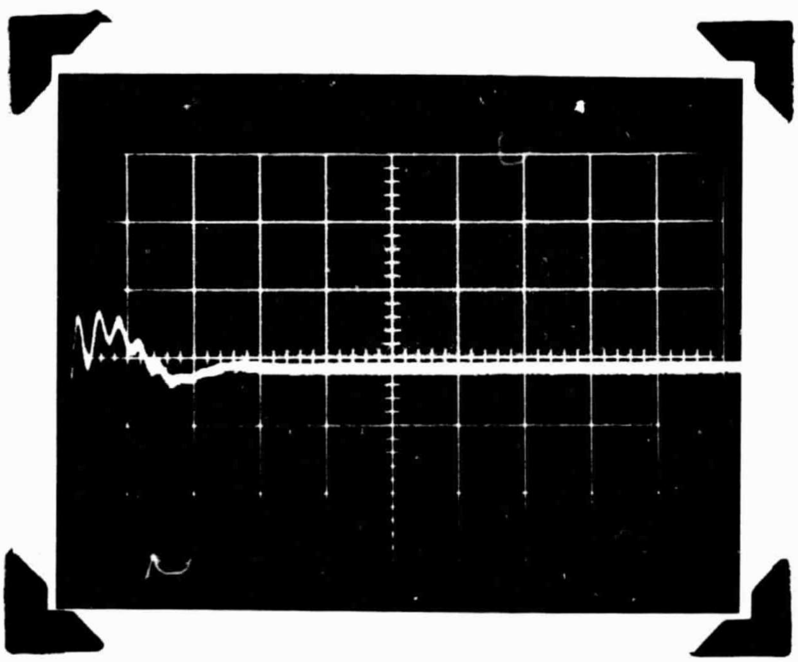
BOOST REGULATOR 100W

TRANSIENT RESPONSE

25 VOLTS INPUT VERTICAL 0.5V/cm HORIZONTAL 5.0ms/cm



10% to 50%



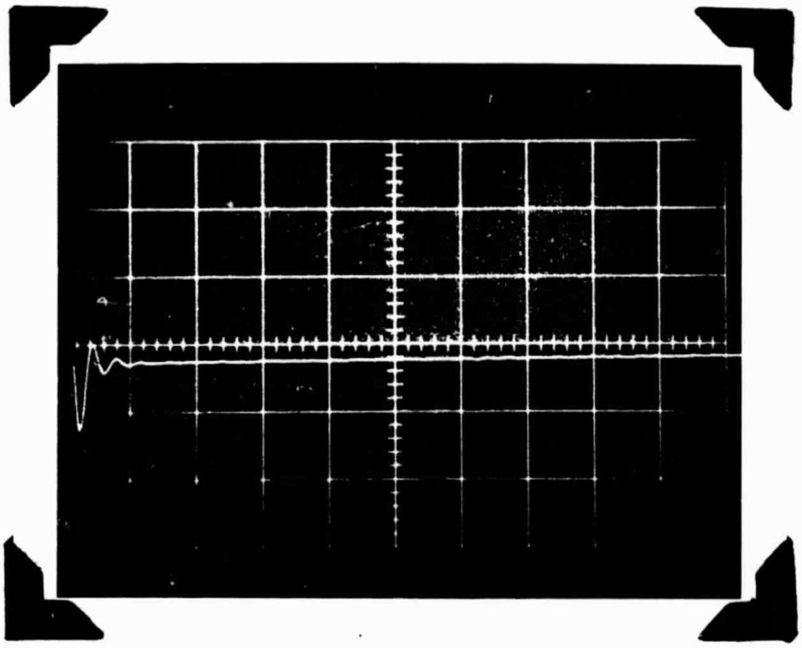
50% to 10%

FIGURE 29

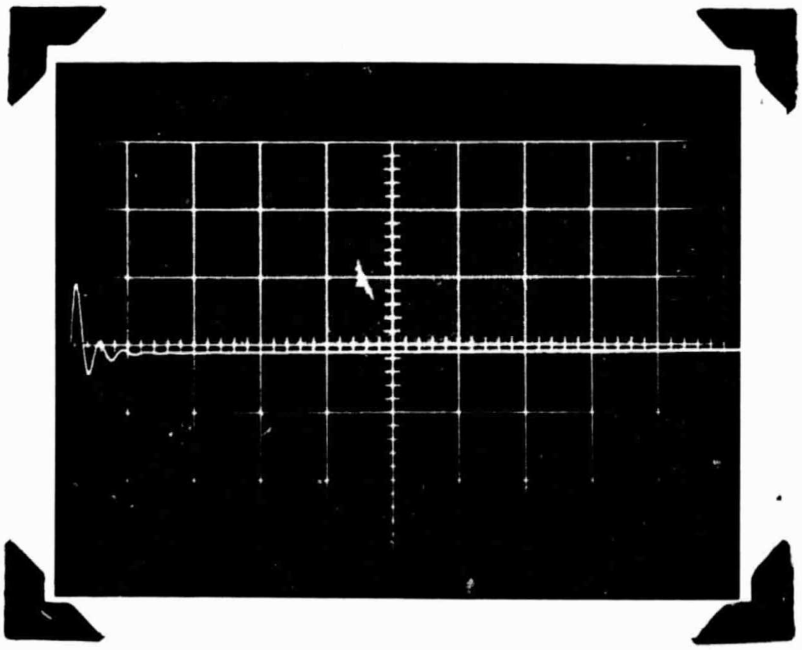
BOOST REGULATOR 100W

TRANSIENT RESPONSE

25 VOLTS INPUT VERTICAL 0.5V/cm HORIZONTAL 5.0ms/cm



50% to 100%



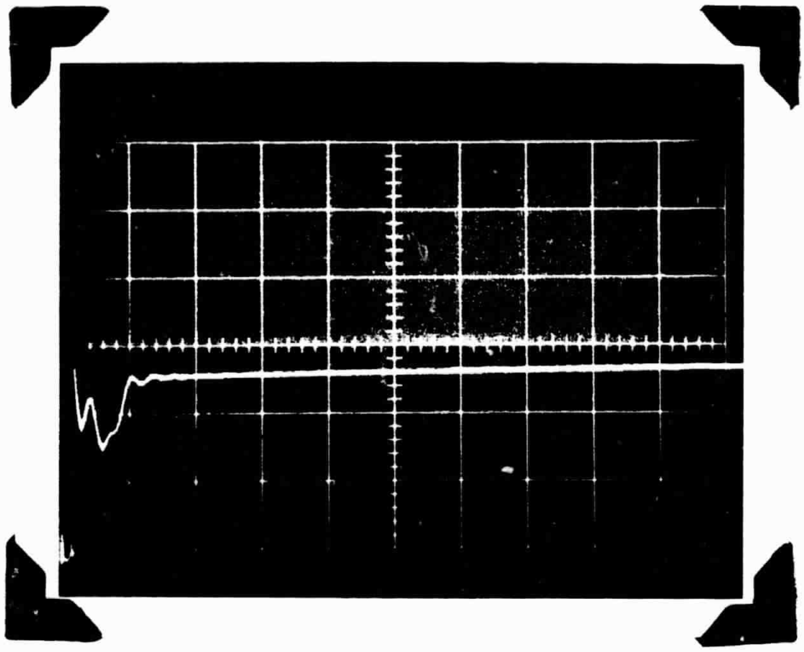
100% to 50%

FIGURE 30

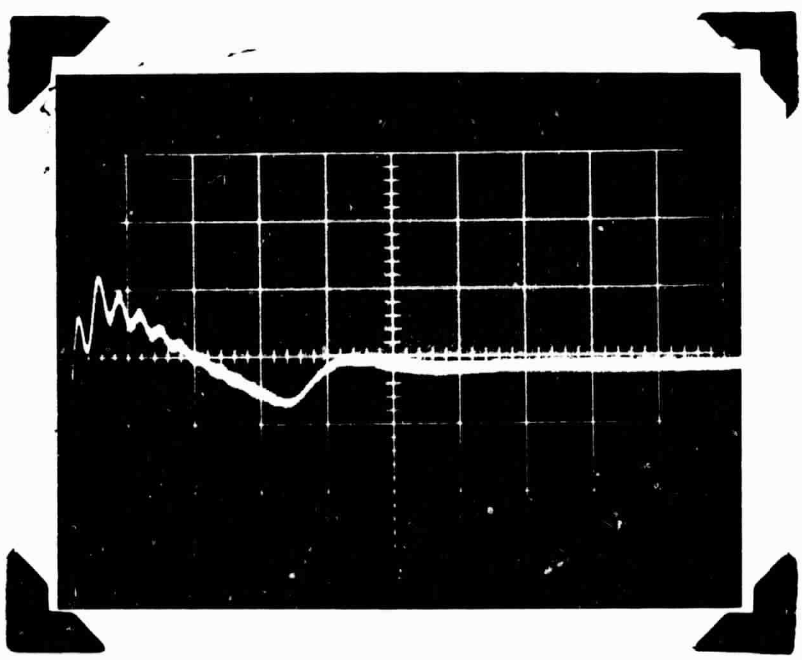
BOOST REGULATOR 100W

TRANSIENT RESPONSE

38 VOLTS INPUT VERTICAL 0.5V/cm HORIXONTAL 5.0ms/cm



10% to 50%



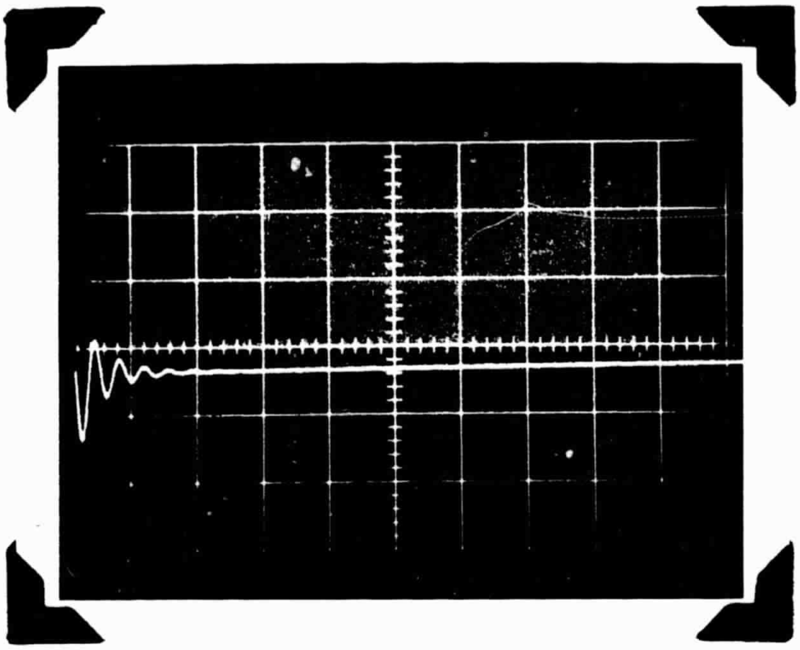
50% to 10%

FIGURE 31

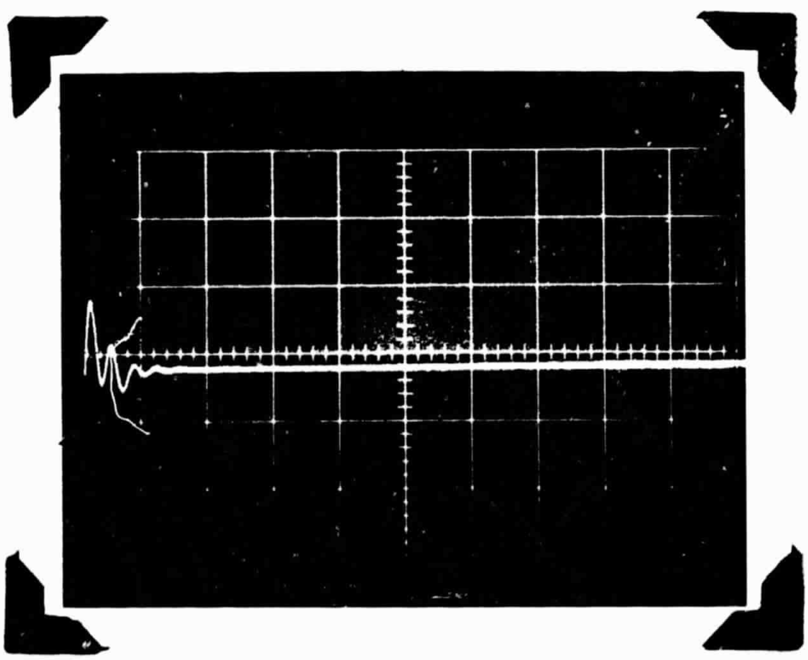
BOOST REGULATOR 100W

TRANSIENT RESPONSE

38 VOLTS INPUT VERTICAL 0.5V/cm HORIZONTAL 5.0ms/cm



50% to 100%



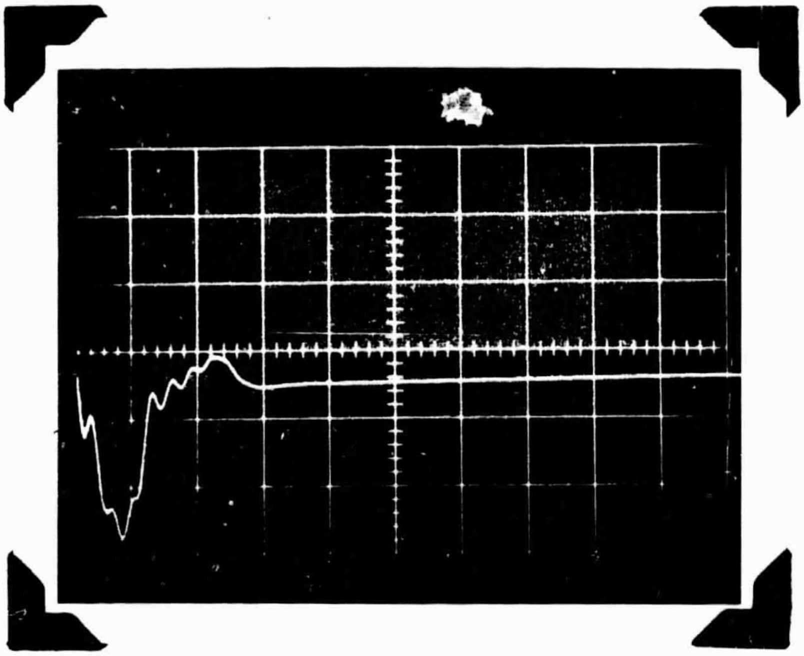
100% to 50%

FIGURE 32

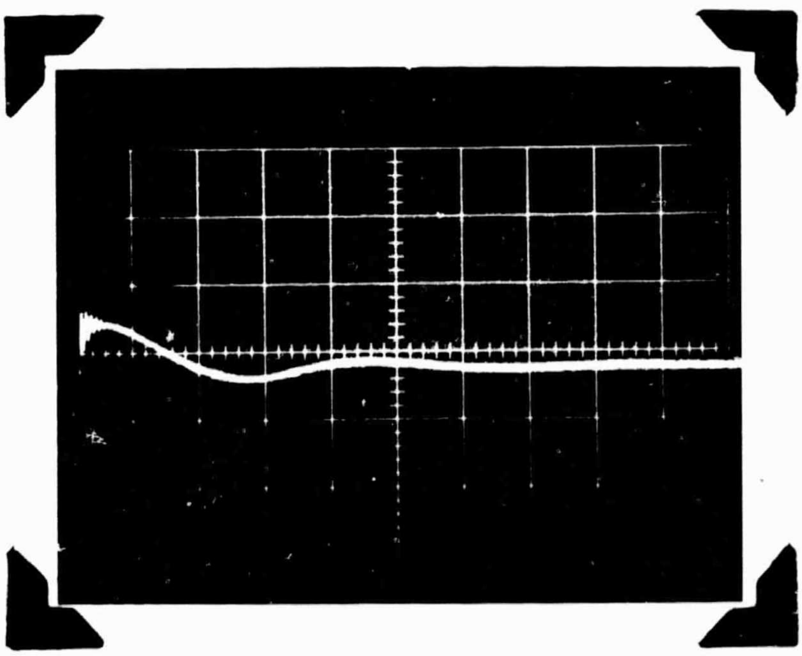
BOOST REGULATOR 100W

TRANSIENT RESPONSE

50 VOLTS INPUT VERTICAL 0.5V/cm HORIZONTAL 5.0ms/cm



10% to 50%



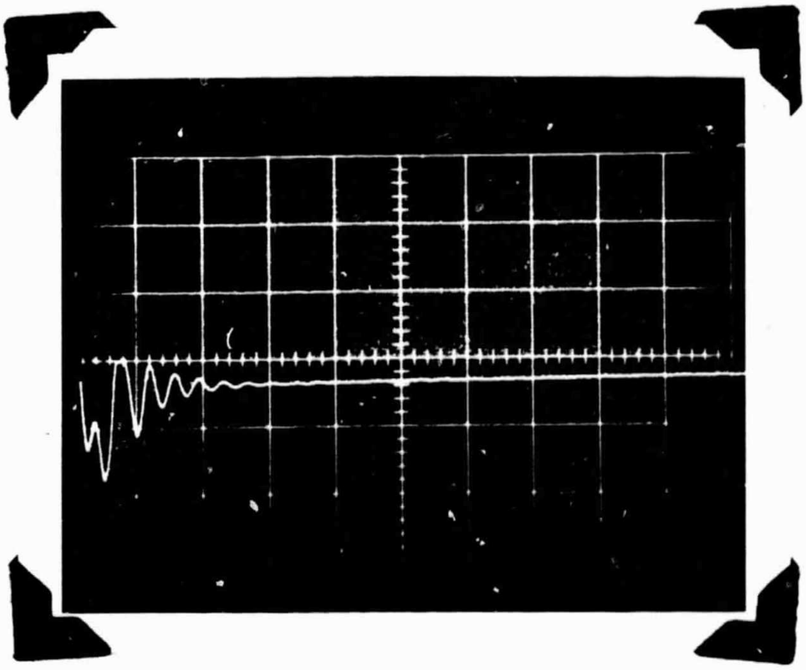
50% to 10%

FIGURE 33

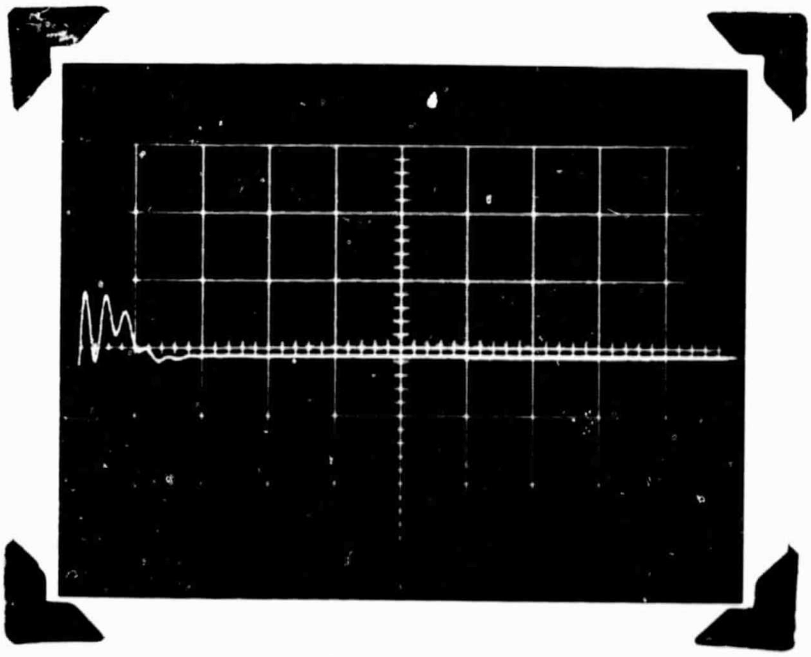
BOOST REGULATOR 100W

TRANSIENT RESPONSE

50 VOLTS INPUT VERTICAL 0.5V/cm HORIZONTAL 5.0ms/cm



50% to 100%



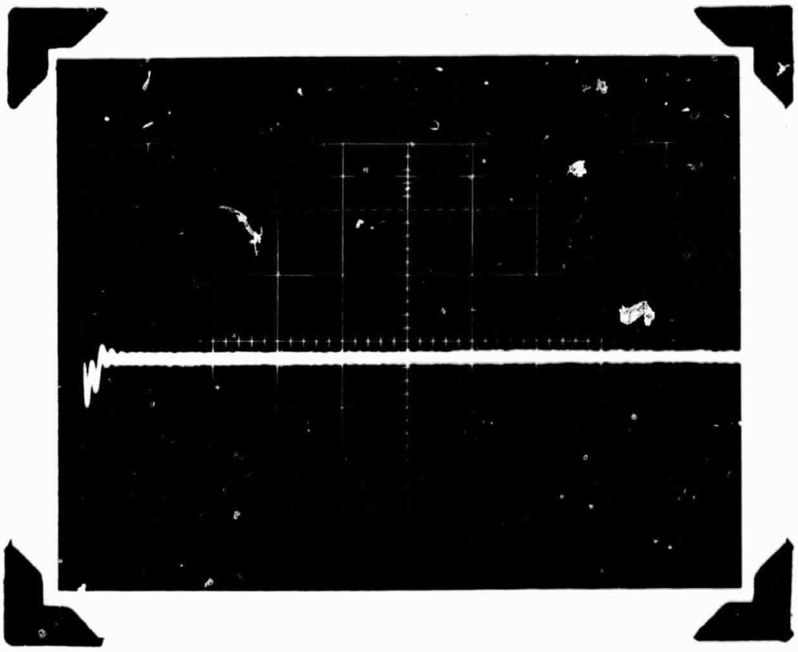
100% to 50%

FIGURE 34

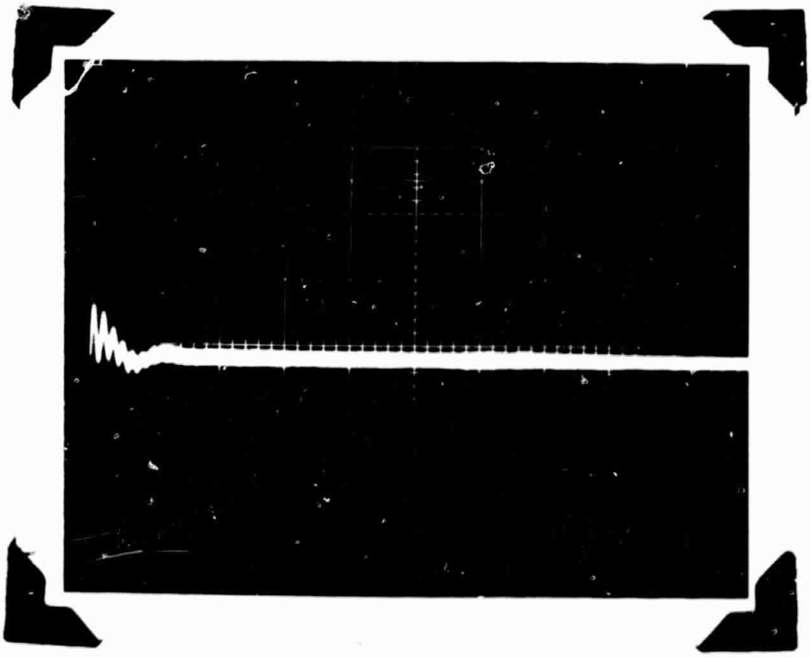
BOOST REGULATOR 200W

TRANSIENT RESPONSE

25 VOLTS INPUT VERTICAL 1V/cm HORIZONTAL 10ms/cm



10% to 50%



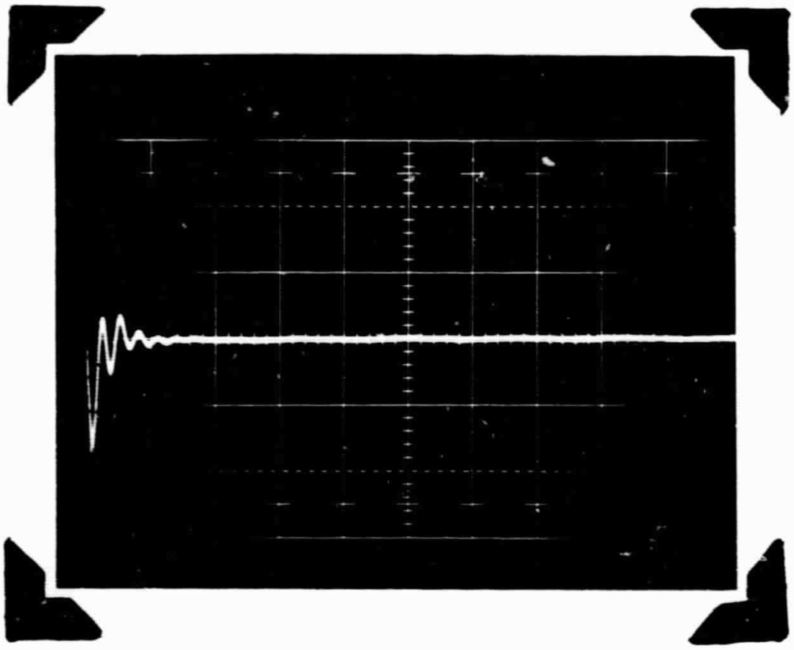
50% to 10%

FIGURE 35

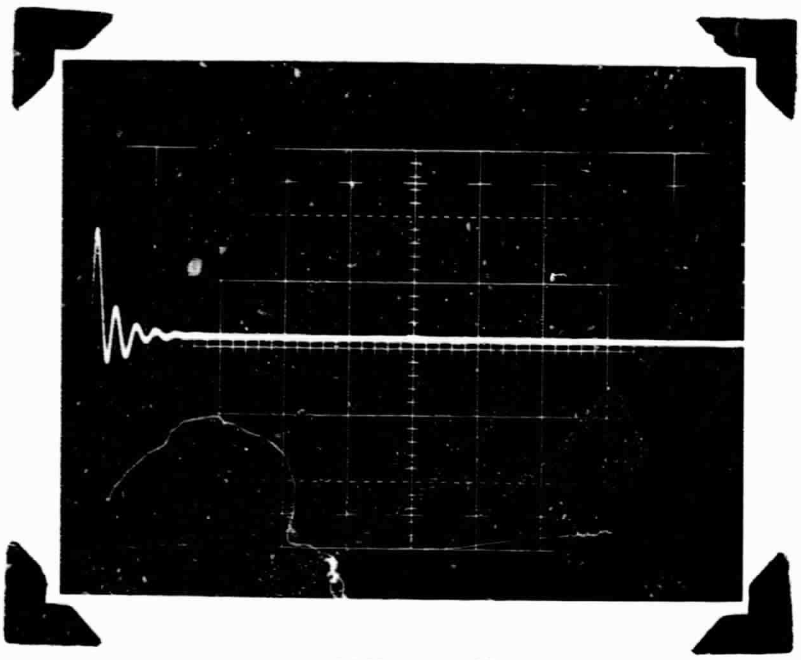
BOOST REGULATOR 200W

TRANSIENT RESPONSE

25 VOLTS INPUT VERTICAL .5V/cm HORIZONTAL 5 ms/cm



50% to 100%



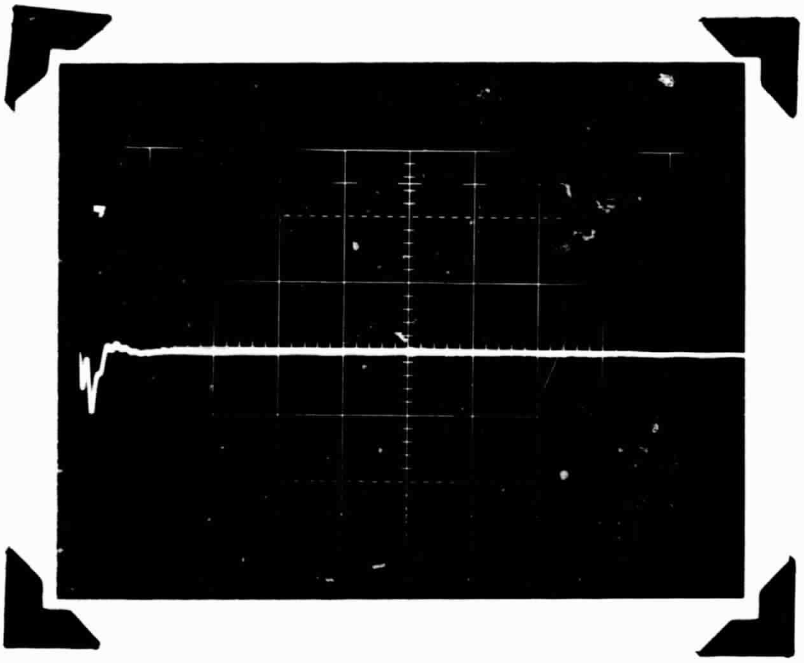
100% to 50%

FIGURE 36

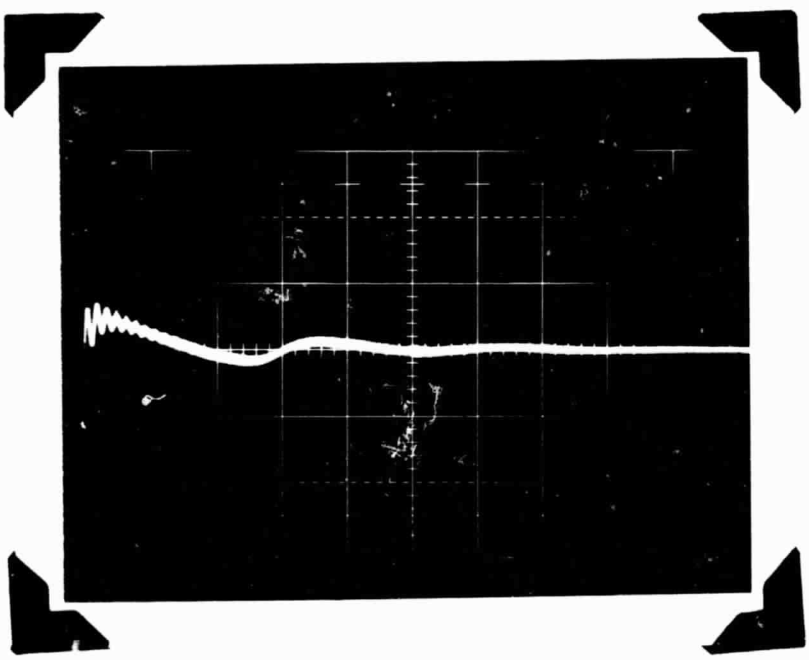
BOOST REGULATOR 200W

TRANSIENT RESPONSE

36 VOLTS INPUT VERTICAL 1V/cm HORIZONTAL 10 ms/cm



10% to 50%



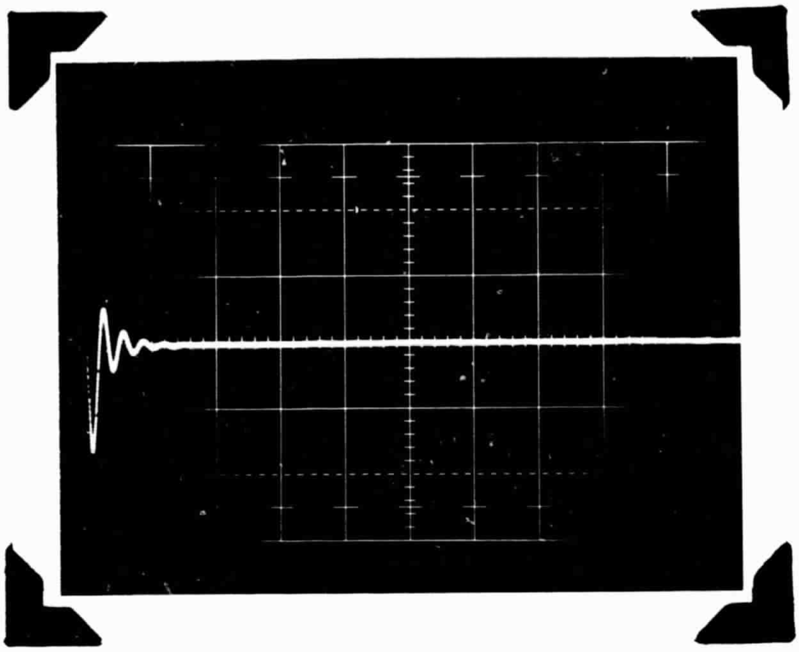
50% to 10%

FIGURE 37

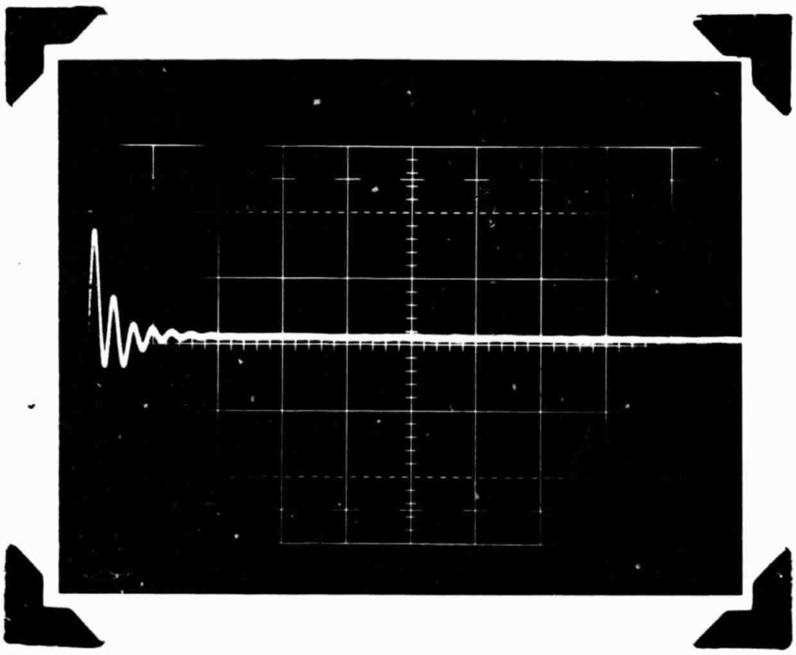
BOOST REGULATOR 200W

TRANSIENT RESPONSE

36 VOLTS INPUT VERTICAL .5V/cm HORIZONTAL 5 ms/cm



50% to 100%



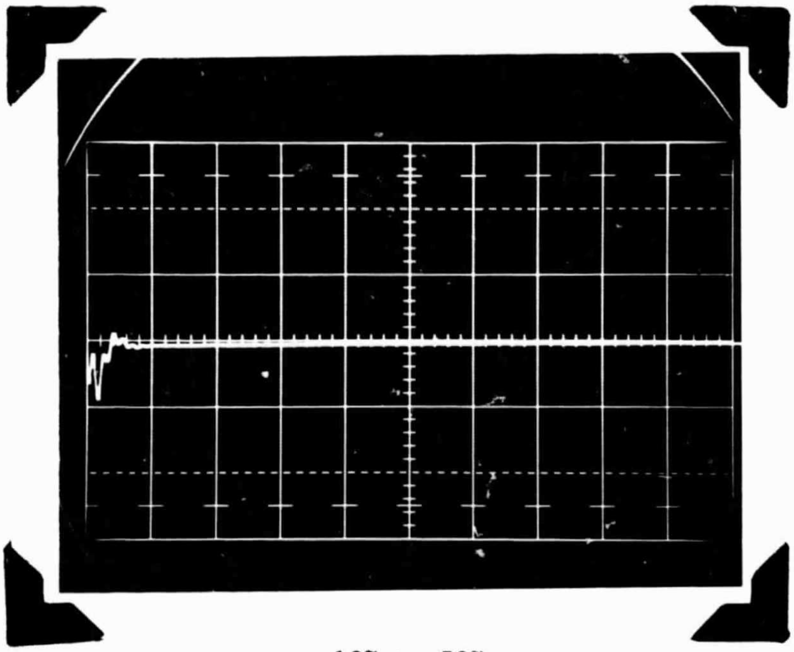
100% to 50%

FIGURE 38

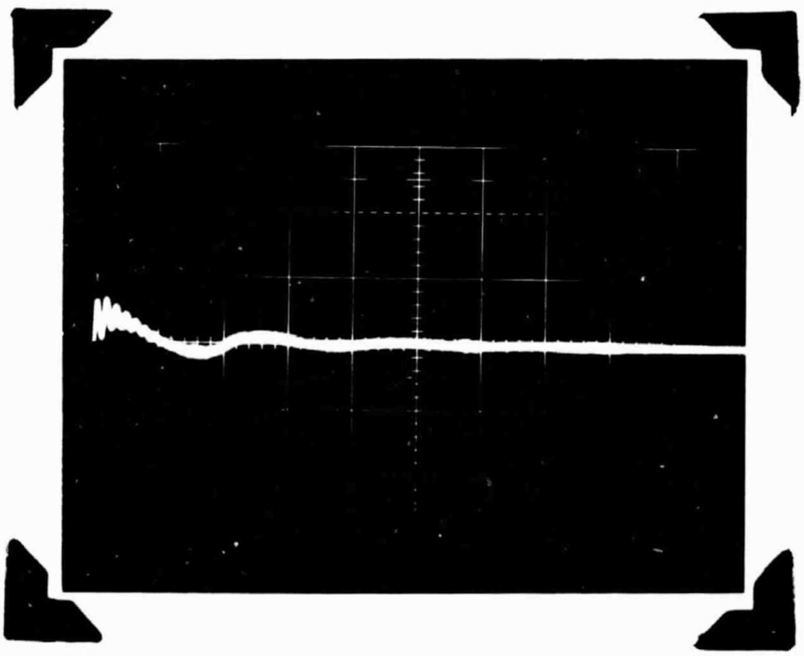
BOOST REGULATOR 200W

TRANSIENT RESPONSE

50 VOLTS INPUT VERTICAL 1V/cm HORIZONTAL 10 ms/cm



10% to 50%



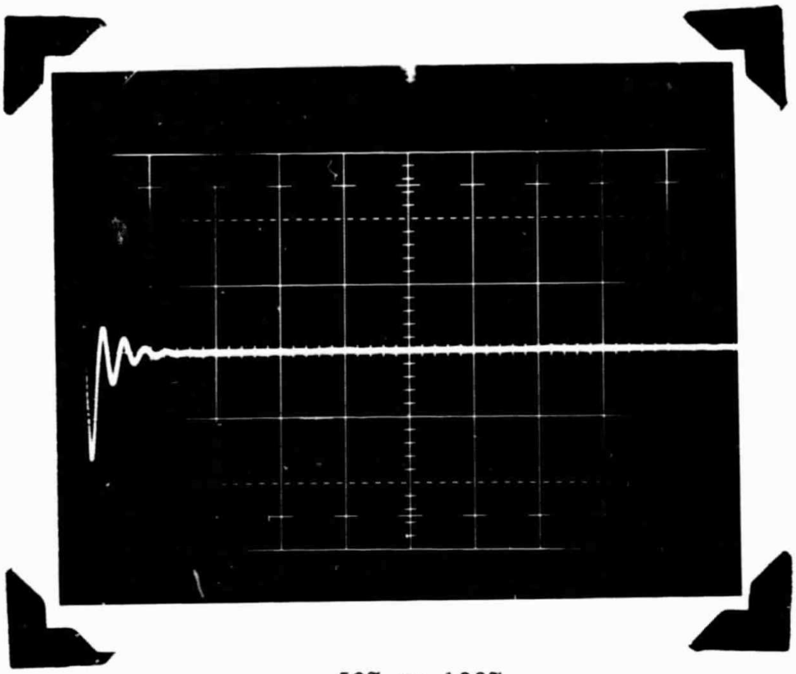
50% to 10%

FIGURE 39

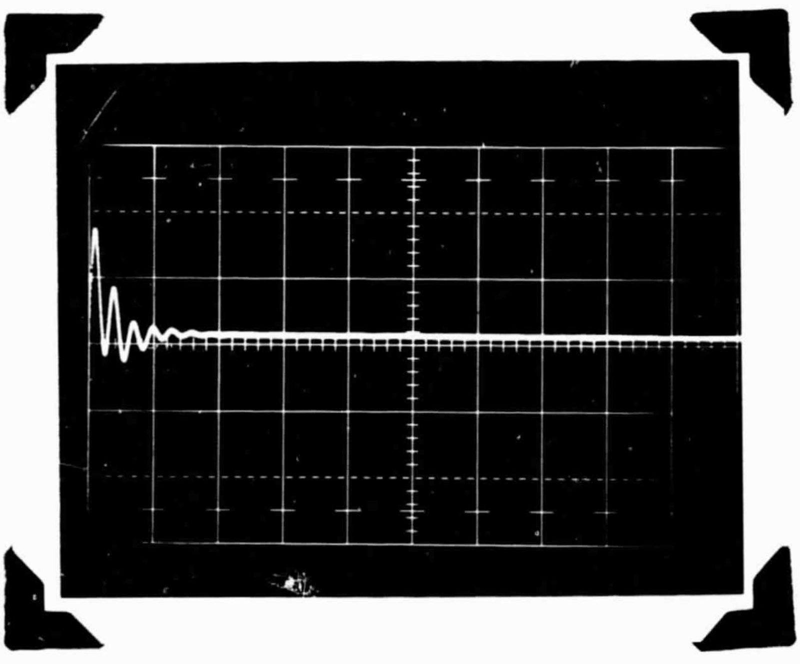
BOOST REGULATOR 200W

TRANSIENT RESPONSE

50 VOLTS INPUT VERTICAL .5V/cm HORIZONTAL 5ms/cm



50% to 100%



100% to 50%

FIGURE 40

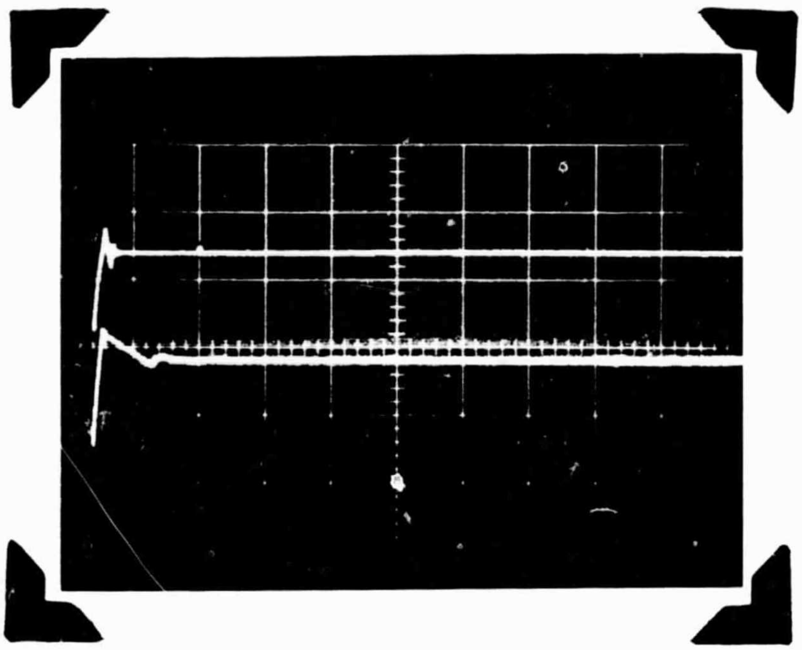
IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

VERTICAL 20V/cm HORIZONTAL 100 ms/cm

OUTPUT WAVEFORM



TOP 0-50V F/L

BOTTOM 0-50V 10% LOAD

FIGURE 41

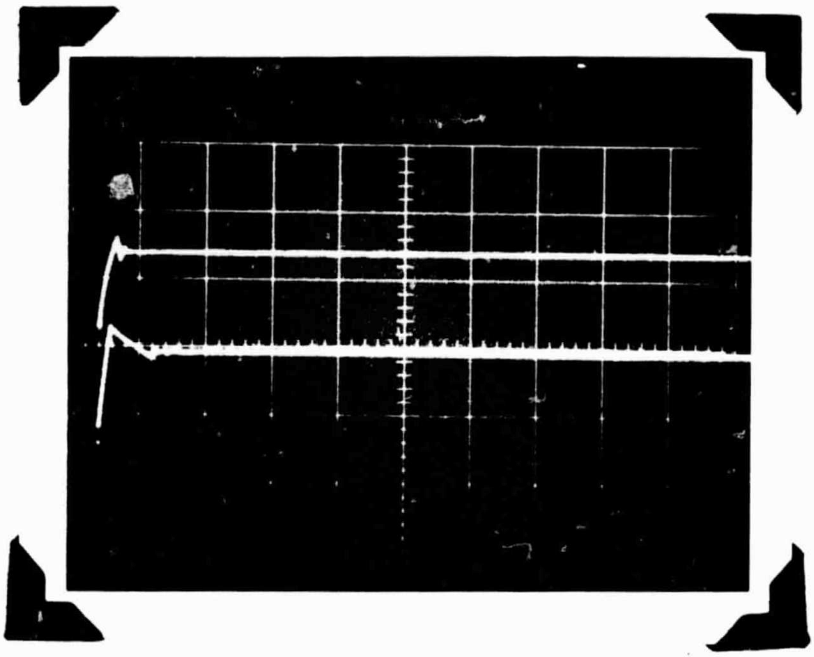
IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

VERTICAL 20V/cm HORIZONTAL 100 ms/cm

OUTPUT WAVEFORM



TOP 0-25V F/L

BOTTOM 0-25V 10% LOAD

FIGURE 42

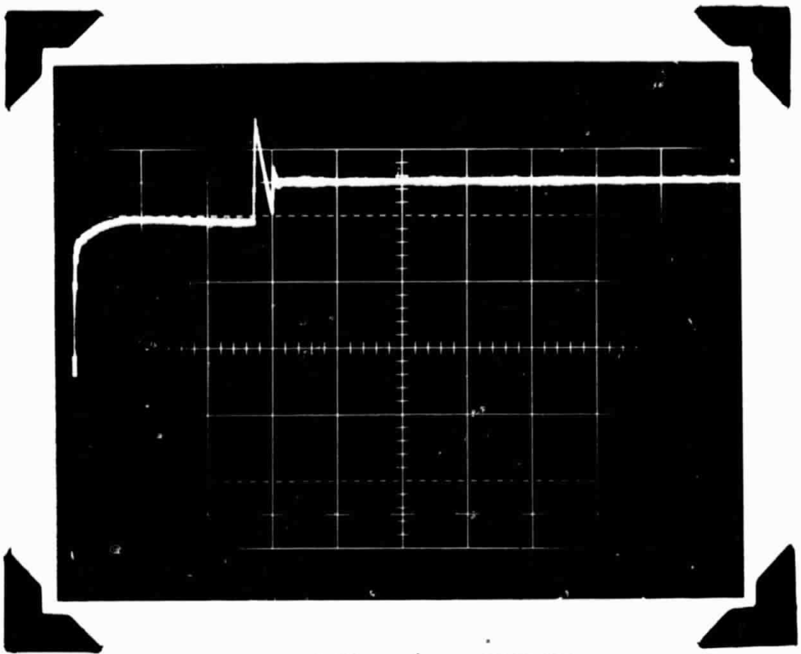
IMPROVED BOOST REGULATOR 200W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

FULL LINE STEP CHANGE 3 ms

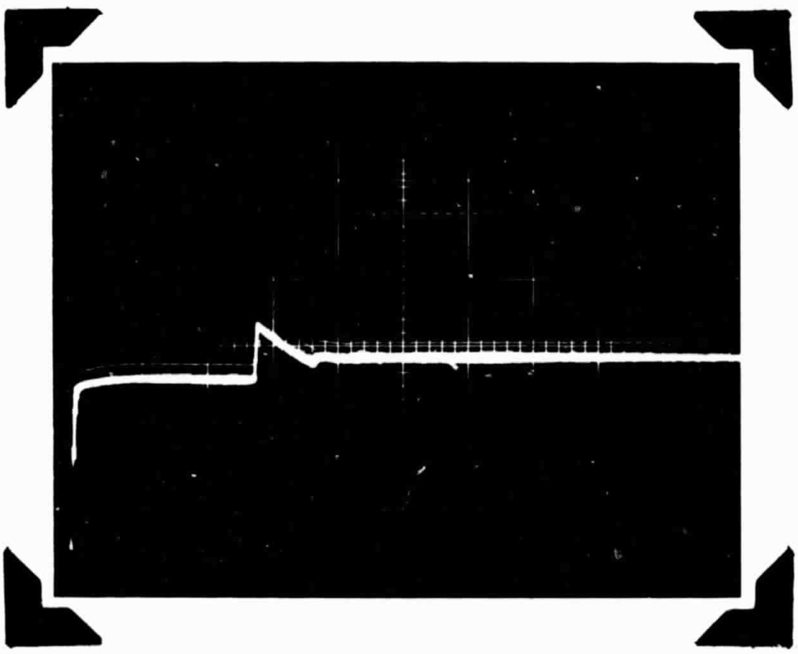
10 V/cm



100 ms/cm OUTPUT

50% LOAD

20 V/cm



100 ms/cm OUTPUT

10% LOAD

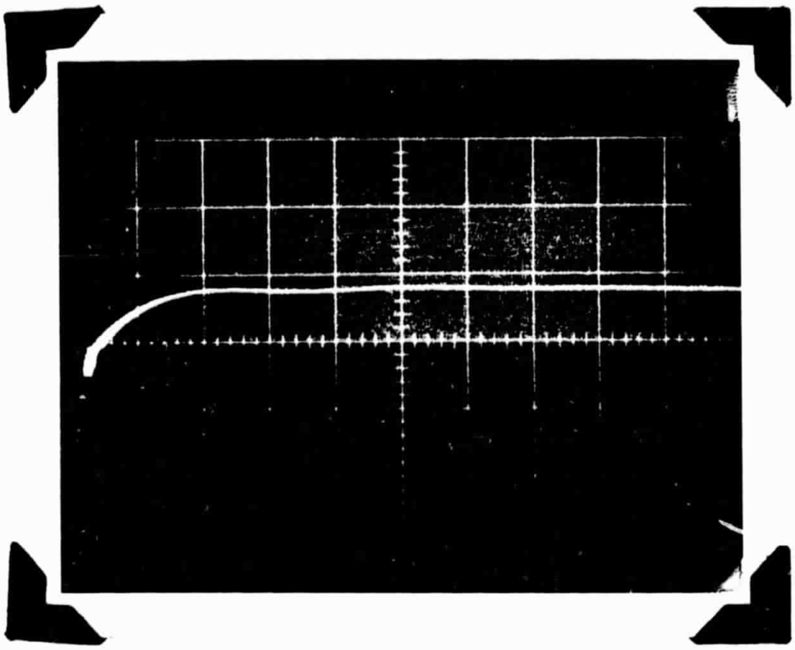
FIGURE 43

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

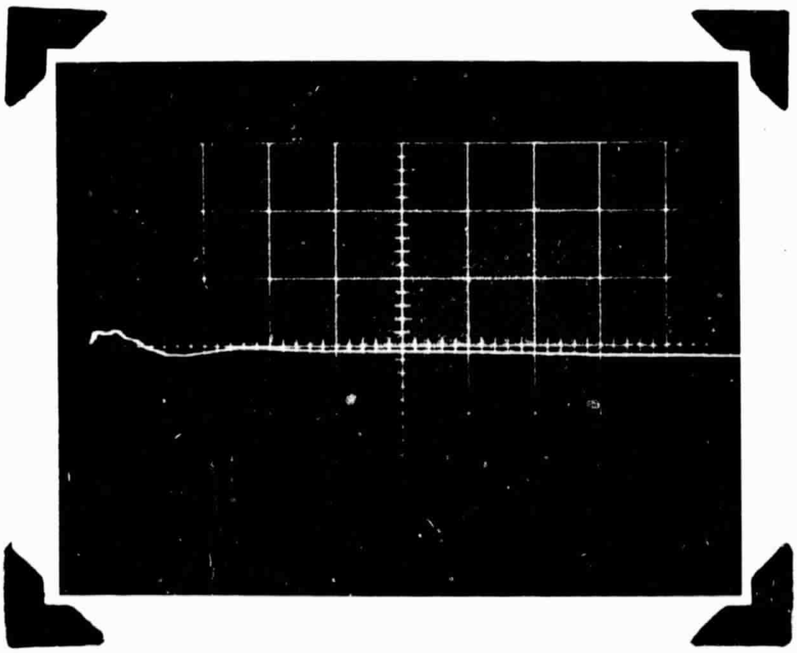
10 V/cm



20 ms/cm INPUT

25V to 38V

.1 V/cm



5 ms/cm OUTPUT N/L

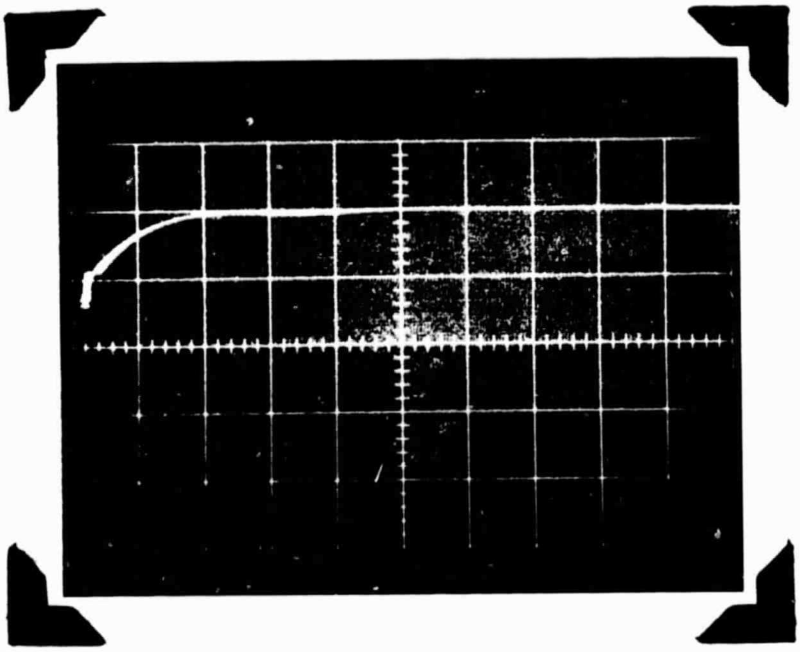
FIGURE 44

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

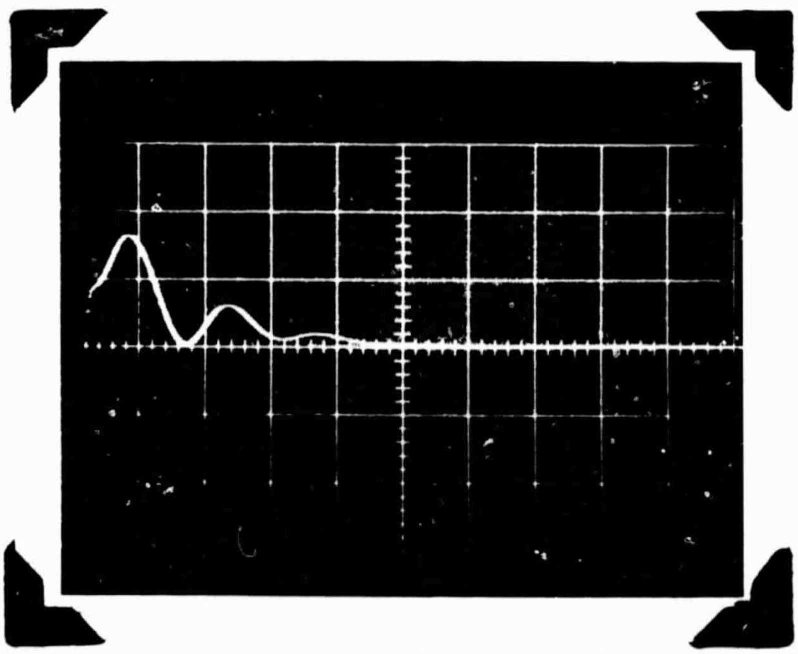
10 V/cm



20 ms/cm

35.4V to 50V INPUT

.1V/cm



50 ms/cm OUTPUT N/L

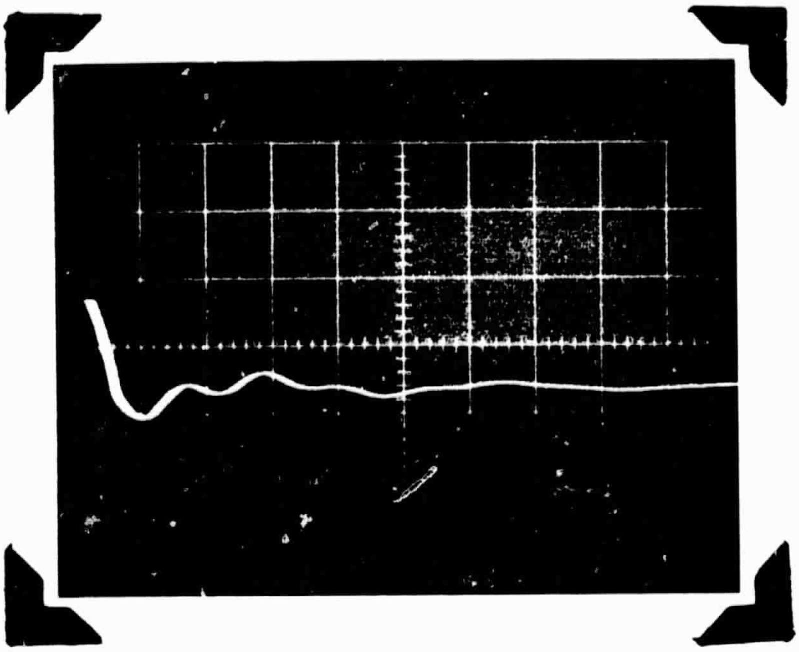
FIGURE 45

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

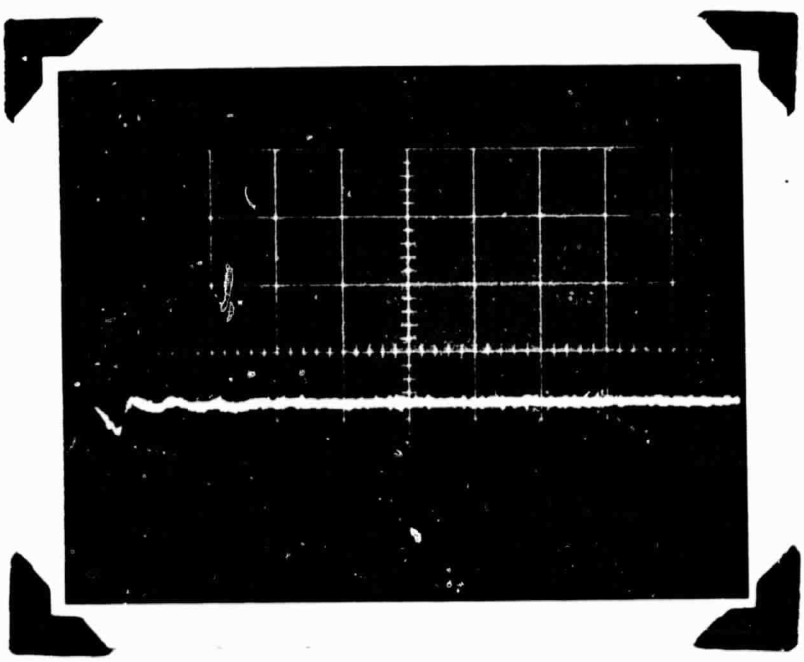
10 V/cm



50 ms/cm

38 to 25 V INPUT

.05 V/cm



100 ms/cm OUTPUT F/L

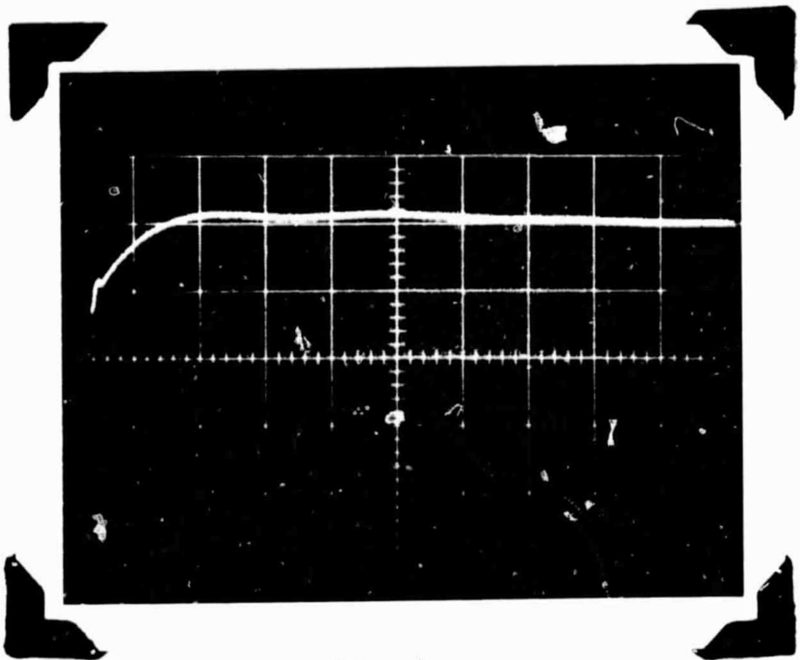
FIGURE 46

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

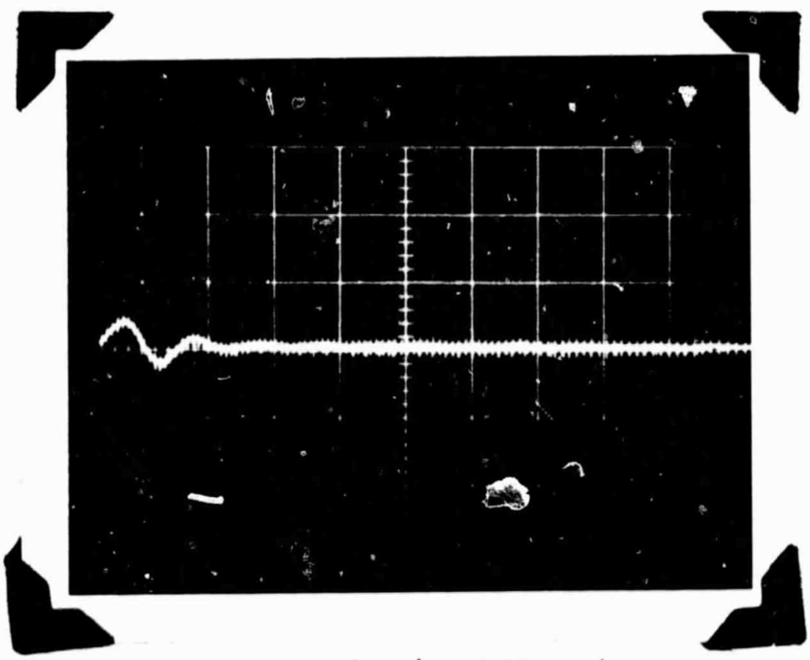
20 V/cm



20 ms/cm

35.4V to 50V INPUT

.1V/cm



2 ms/cm OUTPUT F/L

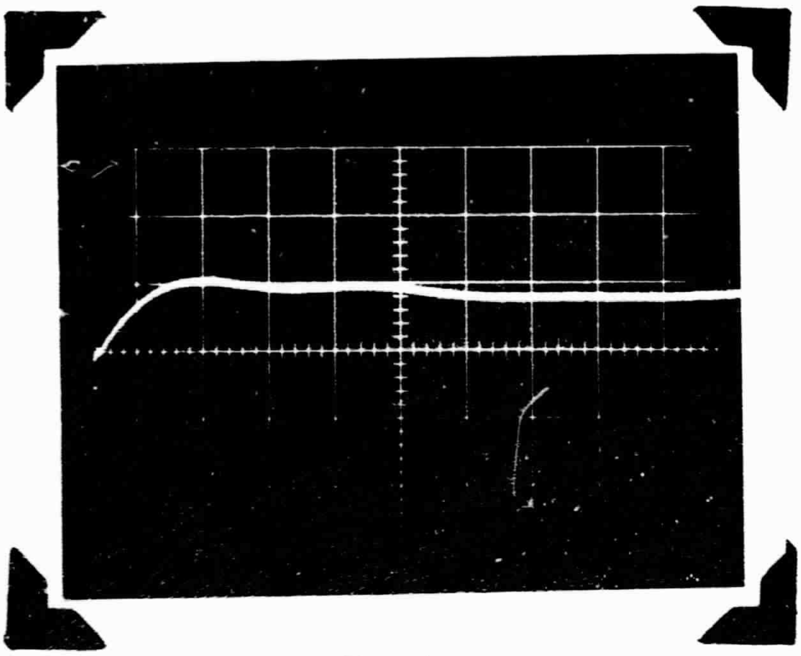
FIGURE 47

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

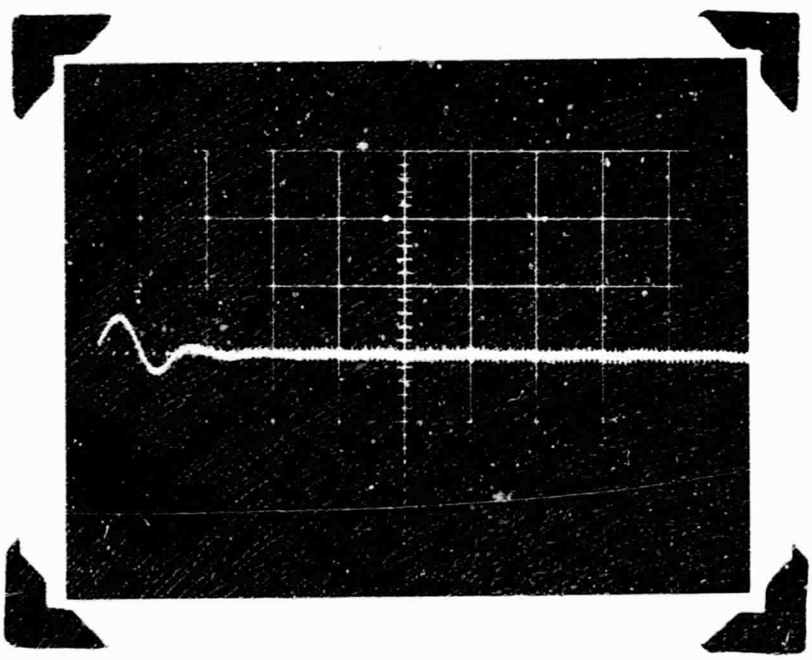
10 V/cm



20 ms/cm

25 to 38V INPUT

.1 V/cm



2 ms/cm OUTPUT F/L

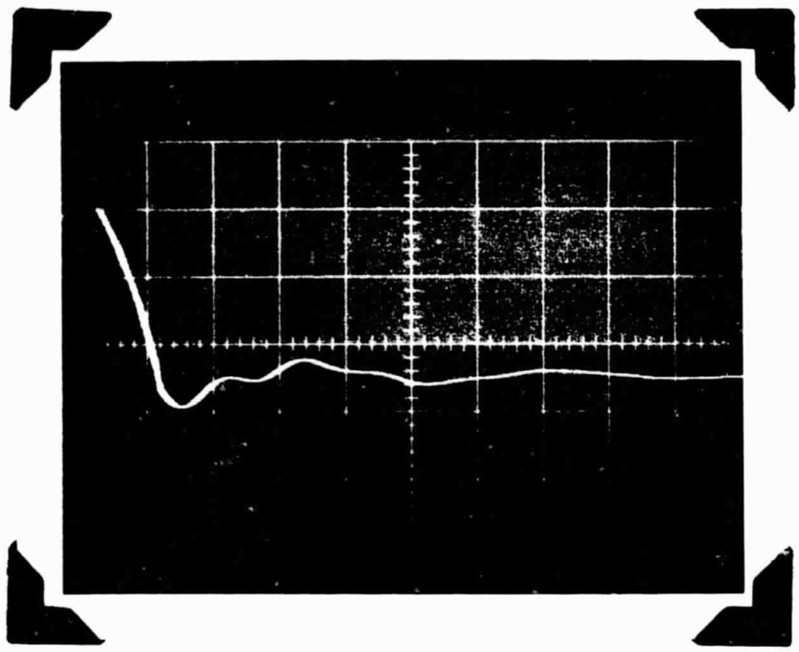
FIGURE 48

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

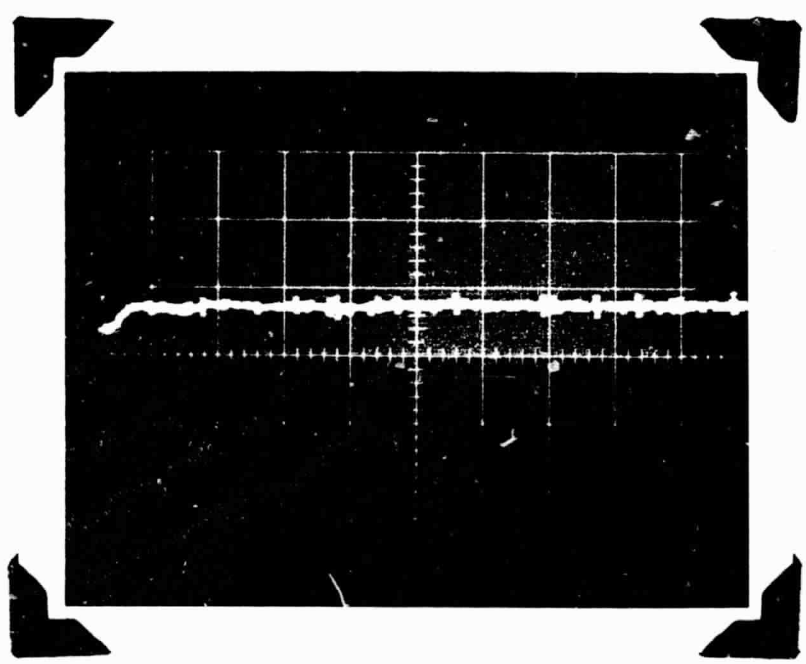
10 V/cm



50 ms/cm INPUT

50V to 25V

.05 V/cm



50 ms/cm OUTPUT F/L

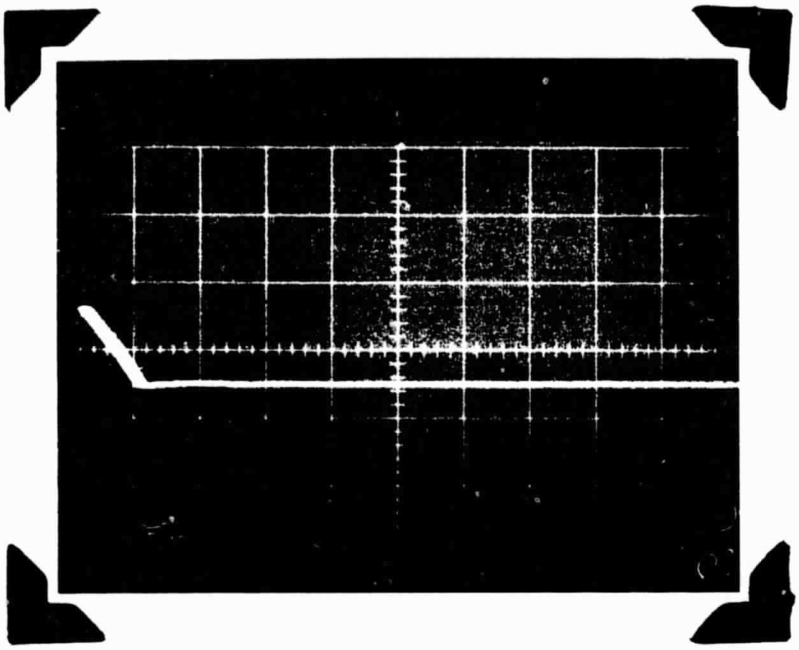
FIGURE 49

IMPROVED BOOST REGULATOR 100W

TYPICAL WAVEFORM

RESPONSE TO LINE TRANSIENT

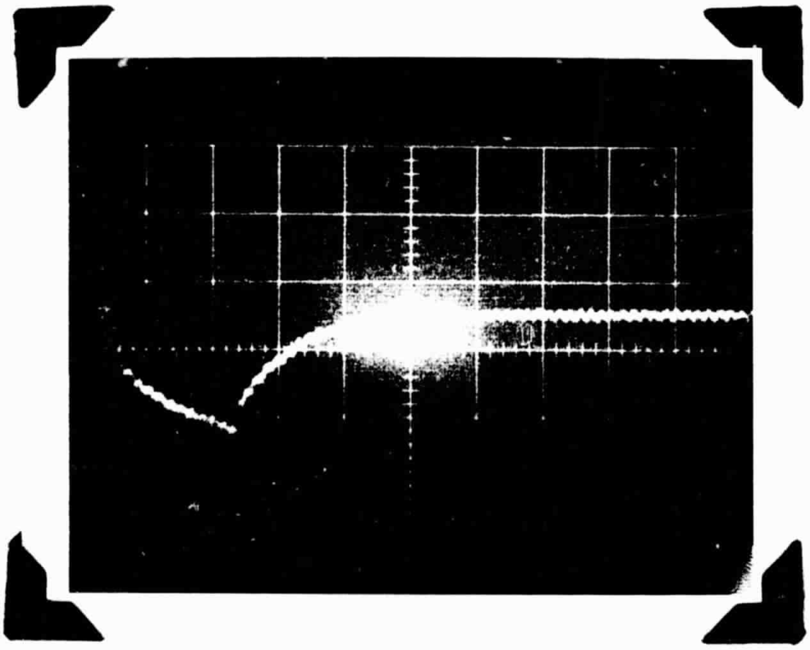
10 V/cm



200 ms/cm INPUT

38V to 25V

5 mv/cm



100 ms/cm OUTPUT N/L

FIGURE 50