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LOW INPUT VOLTAGE CONVERSION-REGULATION FROM UNCONVENTIONAL PRIMARY AND SECONDARY SOURCES

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JANUARY 1968

GSFC

GODDARD SPACE FLIGHT CENTER
GREENBELT, MARYLAND

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ABSTRACT

The primary source power for some proposed deep space probe missions will be provided by radioisotopic fueled thermoelectric generators (RTG's). However RTG's are power-limited and storage is needed to supply peak power requirements. The use of a single electrochemical cell battery, when combined with voltage stepup DC to DC conversion can have advantages as a long lifetime storage system. This paper presents the principal results obtained from study and circuit development in the area of low input voltage conversion-regulation (LIVCR) from unconventional primary (RTG's) and secondary (battery) sources. An analysis was made of the source, converter, and load interrelationships for power-limited primary sources and low impedance secondary sources. A new concept and implementation of a circuit technique to improve low input voltage inverter crossover reliability, efficiency, and source ripple, by optimizing inverter full-cycle base driving conditions was developed. A configurational current feedback driven inverter approach which protects against output transformer saturation destructive current surges was developed which has important use for low impedance source LIVCR applications. To confirm the analysis, applications demonstrating the design approaches developed are presented in the paper along with waveform and numerical data. It is shown that the combining of sources each with radically different output characteristics can be functionally-integrated into a power system using the DC to DC converter-regulators specially developed, which provide conditions operationally suitable to each source subsystem while enhancing the reliability and efficiency of the overall power supply system.

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CONTENTS

Page

I. INTRODUCTION 1

II. LOAD DEMAND, PRIMARY SOURCE, AND ENERGY STORAGE CONSIDERATIONS 1

A. Primary Sources and No Energy Storage Requirement 1

B. Electrochemical Storage 3

1. Conventional Multiple Cell Battery 3
2. Single Cell Battery 3

III. SOURCE-INVERTER/CONVERTER-LOAD INTERRELATION . . . 3

A. High-Impedance/Power-Limited Sources 3

B. Low Impedance Sources. 4

IV. PUSH-PULL OSCILLATOR INVERTER, OVERLAP AND SWITCHING, CROSSOVER CONSIDERATIONS 6

A. Voltage Feedback Driven Inverters. 6

B. Current Feedback Driven Oscillator Inverters 7

C. Current Feedback Driven Oscillator Inverters with Source Voltage Controlled Switching Rate 8

D. Inverter Crossover, Source Ripple, and Filtering 8

V. A NEW BASE CURRENT CONTROL AND SHAPING CIRCUIT TECHNIQUE TO IMPROVE INVERTER/CONVERTER CROSSOVER PERFORMANCE AND REDUCE SOURCE RIPPLE 10

A. Concept, Implementation, and Performance. 10

B. Effect of Load on Inverter Period 15

C. An Alternate Oscillator Inverter Configuration with Current Feedback Drive and Source Voltage Controlled Switching Rate 19

D. Voltage Feedback Driven Inverter Application and Limitation 20

CONTENTS (Continued)

		Page
E.	The Inductor	20
F.	Characteristics and Advantages	20
VI.	PROTECTION AGAINST DESTRUCTIVE POWER TRANSFORMER SATURATION CURRENT SURGES IN CURRENT FEEDBACK INVERTERS/CONVERTERS	23
A.	High-Impedance/Power-Limited Source and Primary Winding Current Feedback Driven Inverters	23
B.	Low Impedance Source and Primary Winding Current Feedback Inverters	25
C.	Low Impedance Source and Secondary Winding Current Feedback Driven Inverters	25
D.	Power Limited Source and Secondary Winding Current Feedback Driven Inverters	28
VII.	VERIFICATION OF ANALYSIS	28
A.	Improved Crossover Technique Waveforms	30
B.	Protection Against Excessive Current Surges in Event of Output Transformer Saturation	34
	1. Half Cycle of Transformer Saturation	34
	2. Portion of Half Cycle of Saturation	36
VIII.	AN APPLICATION OF LOW INPUT VOLTAGE CONVERSION-REGULATION (LIVCR) FROM AN UNCONVENTIONAL SECONDARY ELECTROCHEMICAL SOURCE	37
IX.	RTG PRIMARY SOURCE COMBINED WITH A SECONDARY BATTERY SOURCE	40
X.	CONCLUSIONS	46
XI.	REFERENCES	47

ILLUSTRATIONS

Figure		Page
1	Typical Radioisotopic Fueled Thermoelectric Generator (RTG) Source Voltage/Current (v/i) Characteristics with Loading . .	2
2	Transformer (T_1) Primary Winding Current Feedback Driven Inverter Configuration	5
3	Idealized Low Impedance Source Voltage/Current (v/i) Characteristics with Loading	6
4	Typical Voltage Feedback Driven Push-Pull Oscillator Inverter Crossover	7
5	Typical Current Feedback Driven Push-Pull Oscillator Inverter Crossover	8
6	Typical Source (v/i) Ripple	9
7a	Improved Crossover Circuit	11
7b	Improved Crossover Circuit Idealized Waveforms	12
8	Effect of Load on Inverter Period	16
9a	Method of Varying Effective Inductance of L with Load	17
9b	Conducting Half-Cycle Base Currents Versus Inverter Period (T)	17
10	Improved Crossover Technique in an Alternate Push-Pull Oscillator Inverter Configuration with Current Feedback Drive and Source Voltage Controlled Switching Rate	19
11	Improved Crossover Technique in a Voltage Feedback Driven Push-Pull Oscillator Inverter with Source Voltage Controlled Switching Rate	21
12	Improved Crossover Technique Idealized Base Drive Waveforms	22

ILLUSTRATIONS (Continued)

Figure		Page
13	Power Supply Using Current Feedback Driven Converter-Regulator (Optionally with Source Voltage Controlled Switching Rate) for Use with Power-Limited Sources	24
14	Power Supply Using a Current Feedback Driven Converter (Optionally with Source Voltage Controlled Switching Rate) for Use with Low Impedance Sources	27
15a	Inductance L and Inverter Frequency Versus Inverter Input Current	29
15b	Peak i_L Current Versus Inverter Input Current	30
16	Improved Crossover Technique Waveforms	31
17	Output Transformer Saturation, Waveforms	35
18	GSFC In-house Developed Low Input Voltage Converter-Regulator, and Prototype Silver Zinc ($A_g Z_n$) Battery (Yardney HR-150(S)-1)	38
19	Preassembled GSFC In-house Developed Low Input Voltage Converter-Regulator	39
20	Automatic Battery Charge/Discharge Cycler, Laboratory Lifetest Arrangement.	41
21a	Single-Cell Silver-Zinc Battery, 24 Hour Charge-Discharge Regime	42
21b	Single-Cell Silver-Zinc Battery, 12 Hour Charge-Discharge Regime	43
22	Typical Power Demand Profile	44
23	RTG (Primary Source) and Battery (Secondary Source) Power Supply System, Block Diagram	45

LOW INPUT VOLTAGE CONVERSION-REGULATION FROM UNCONVENTIONAL PRIMARY AND SECONDARY SOURCES

I. INTRODUCTION

With the increasing availability of new and improved primary and secondary power sources, it is necessary to consider the interrelated source, inversion/conversion, regulation, load and energy storage requirements aspects which will enable development of long lifetime power supplies for space applications using such sources. Future mission load profiles will require power supplies which can supply average and peak output power demands with high overall efficiency and reliability while also imposing the constraint of minimum size and weight. These aspects have been considered and in-house R & D effort conducted to develop conversion, regulation, and related power supply circuitry which functionally-integrate the overall power system requirements effectively, efficiently and reliably, while operating from high and/or low impedance type of newly developed unconventional sources. The principal results obtained from this work is presented in this report.

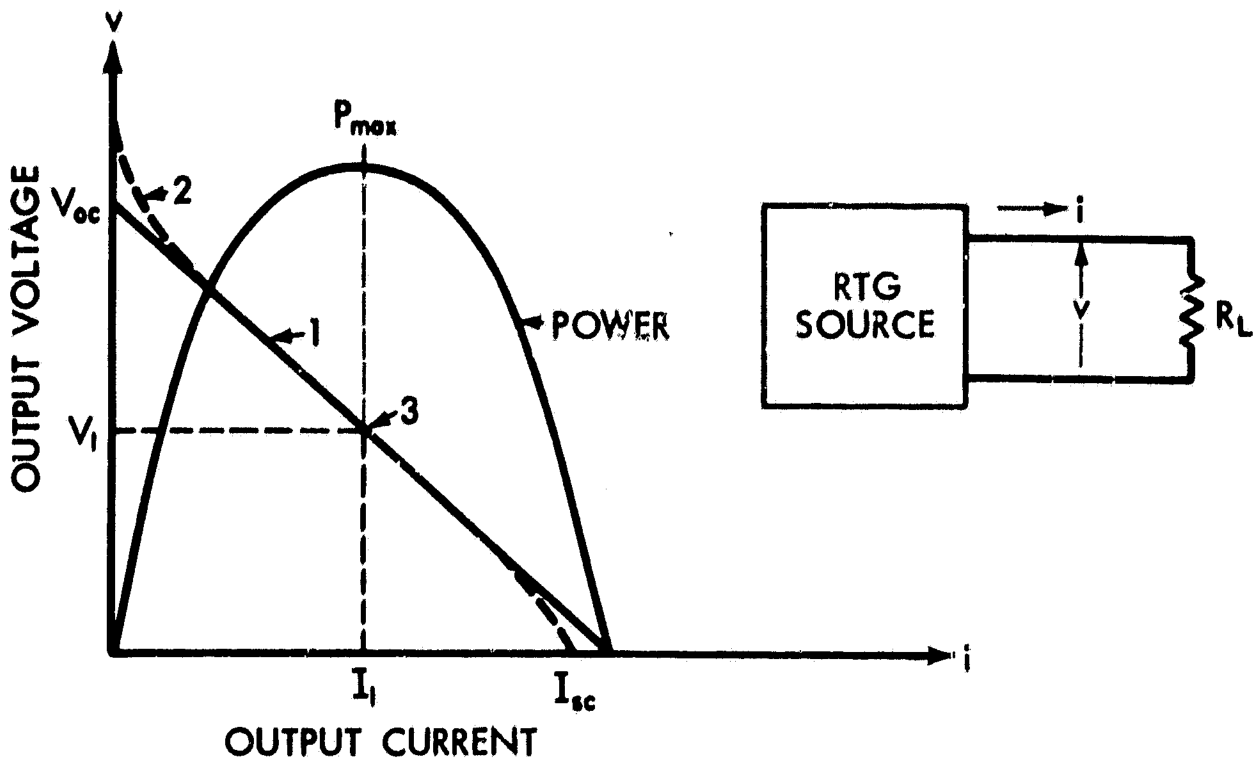
II. LOAD DEMAND, PRIMARY SOURCE, AND ENERGY STORAGE CONSIDERATIONS

A. Primary Sources and No Energy Storage Requirement

For the power supply, where no energy storage is used, the primary source or sources, must be sized large enough to handle the peak output power demands. Primary sources can be employed singly, in multiples, or in combination with other types of primary sources to satisfy overall power requirements. For example, a radioisotopic fueled thermoelectric generator source (RTG) may be combined with a solar-photovoltaic source providing a hybrid-primary power source system (reference 1).

Typically illustrated in Figure 1 are the loading characteristics of a radioisotopic fueled thermoelectric generator (RTG) which is probably the most highly developed unconventional primary source being considered for future deep space probe mission use. For the case of conditioned higher output voltages obtained from the inherently low voltage primary sources, and where no requirement for energy storage is imposed, the achievable power outputs to the load are determined principally by:

1. The source maximum output power capability



1. Assumed Linear Resistance (R_G) Simulation
2. Typical RTG v/i Characteristic
3. V_1/I_1, Matched-Load, Maximum Power Output Condition ($R_L = R_G$)

Figure 1. Typical Radioisotopic Fueled Thermoelectric Generator (RTG) Source Voltage/Current (v/i) Characteristics with Loading

2. Source to equivalent load matching

3. Source efficiency and conversion-regulation losses

4. Source decay

Necessarily in this case the source output capability must be sized large enough to handle the peak power demands, even if these peak power demands are occasional and having high power demand excursions. If the average power output requirements are much smaller than the occasional peak power requirements, the source size becomes excessively large, and inordinately expensive (the large source size can be wasteful of scarce isotope fueling). Besides being uneconomical the requirement for large sources may present a greater thermal and radioactivity problem to the spacecraft circuits and mission experiments. Also large radioisotopic sources present greater abort or reentry radiological hazards.

It has been postulated and analytically demonstrated (reference 2) that by using low output voltage primary source operation, and DC to DC converter voltage step-up, the overall primary source reliability can be improved.

B. Electrochemical Storage

1. Conventional Multiple Cell Battery

General practice has been to use the current limiting voltage regulator which follows the voltage step-up DC to DC converter as a battery charging source of a multiple cell battery storage system. By this conventionally employed method, peak power demands in excess of the prime energy source output power capability would then be supplied from the stored energy of the high voltage secondary energy source battery system. The energy storage requirement on the high voltage side of the converter-regulator utilizes a large number of series-connected electrochemical cells. For precise control of such a battery system a complex battery charge/discharge sensing and control implementation is required. The cycle and lifetime reliability of a battery system is reduced as the number of series-connected cells is increased. For multicell battery applications redundancy or remote activation provisions also become exceedingly complex. Increasing the operating voltage reduces power conditioning constraints but imposes severe problems of reliability in the battery storage system. The reliability needed for long battery system lifetime and cycle life probably cannot be achieved with large series-connected battery strings, and their attendant complex sensing and control circuitry (reference 3).

2. Single Cell Battery

By reducing the complexities of the charge/discharge control circuitry and the redundant battery activation and switching circuitry through the use of a minimal number of series connected cells, the storage system reliability and therefore the overall power system reliability can be improved. Preferably a single cell approach enables precise battery charge/discharge monitoring and control. The increased size and weight of conversion due to low voltage operation may be properly justified considering the reduction of size, weight and complexity of equivalent higher voltage storage system requirements.

III. SOURCE-INVERTER/CONVERTER-LOAD INTERRELATION

A. High-Impedance/Power-Limited Sources

Figure 1 typically illustrated the decreasing output voltage with increasing output current which results in the power-limited characteristic. The thermoelectric generator and thermionic diodes can be placed in this source category.

The loading effects on source reliability have already been discussed in some detail for the case of radioisotopic fueled thermoelectric generator sources in reference 1.

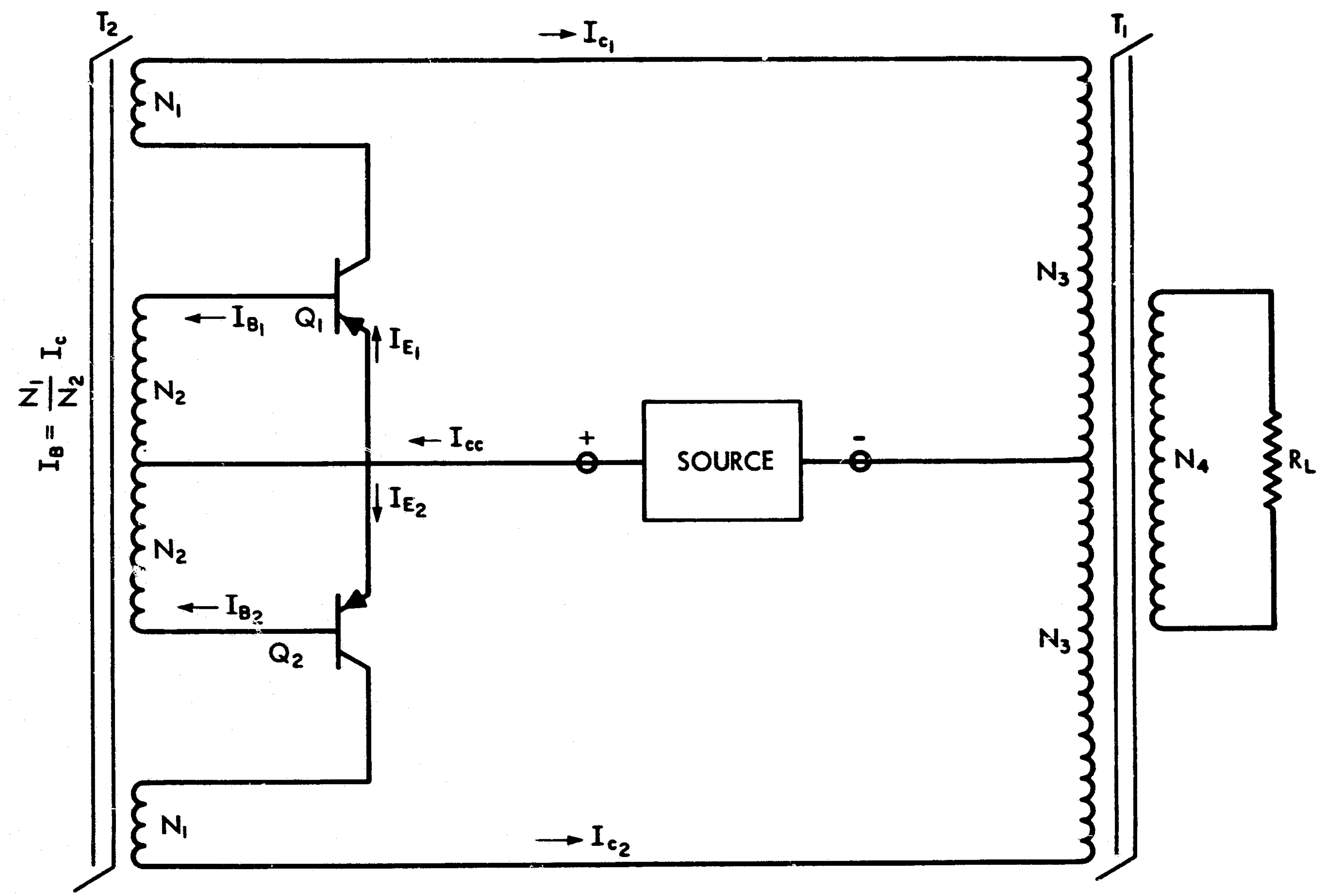
It will be appreciated from Figure 1 that the source characteristic with loading inherently limits the maximum source/inverter current. The voltage step-up inverter used can be designed using power transformer (T_1) primary winding current feedback as typically illustrated by Figure 2. For the high-impedance/power-limited source the primary winding feedback configuration is highly efficient and has enabled a reliable, efficient, functionally-integrated power system using an RTG primary source which is highly resistant to damage from abnormal loading (reference 1).

B. Low Impedance Sources

Certain secondary energy sources, for example, high energy content silver-zinc batteries have essentially unlimited power capability. That is, the battery can supply much higher power levels than that demanded in the normally loaded condition. The idealized output voltage/current characteristic of such a low impedance source is typically illustrated by Figure 3. The essentially constant output voltage for most of its output characteristic gives the steadily rising power output. The current increases with loading, until in the limit, destructive current levels are reached, resulting in catastrophic failure.

When such a source energizes a voltage step-up inverter it is absolutely necessary that no inherent facet of inverter loading operation present a short circuiting or excessive load to the low impedance battery source. Even transient or momentary source/converter/loading shorting effects can allow destructive or component lifetime reducing spike current flows to be reached.

It should also be noted that the source may be a combination of several primary or secondary sources, combined in such a manner that a low impedance source configured with other higher impedance sources, can enable a resultant low impedance source being presented to the inverter/converter terminals. When such a low impedance essentially unlimited power source energizes the voltage step-up inverter, protection against excess current surges is necessary against possible abnormal conditions of:



T_2 - SATURATES, IS FREQUENCY DETERMINING ELEMENT

T_1 - NON-SATURATING OPERATION

Figure 2. Transformer (T_1) Primary Winding Current Feedback Driven Inverter Configuration

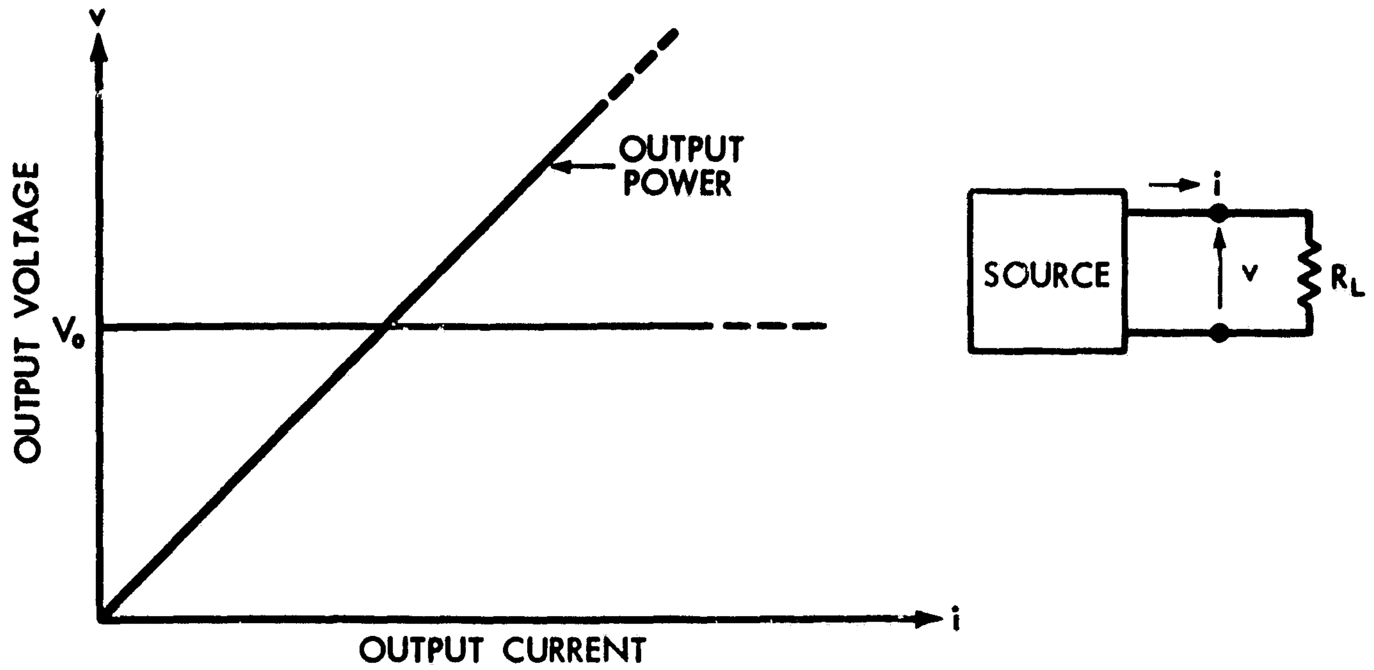


Figure 3. Idealized Low Impedance Source Voltage/Current (v/i) Characteristics with Loading

1. Transistor storage time created overlap

2. Output transformer core magnetic saturation

3. Load short circuit.

Circuit approaches have been developed in-house at GSFC which can efficiently, and with simple reliable electronic implementation inhibit excess abnormal source/inverter/converter/regulator, or load current surges. These designs which enable the use of efficient load related current feedback inverter circuitry are presented in Sections V and VI of this report.

IV. PUSH-PULL OSCILLATOR INVERTER, OVERLAP AND SWITCHING, CROSSOVER CONSIDERATIONS

Low-voltage high-current transistor junctions are large in area, which results in their having long rise, fall, and storage times. These transistor characteristics result in transistor switching and circuit overlap losses, which reduce reliability as well as efficiency.

A. Voltage Feedback Driven Inverters

Figure 4 typically illustrates the Voltage Feedback Push-Pull Oscillator Inverter Crossover condition. Assuming a constant load, if there were no

Figure 4. Typical Voltage Feedback Driven Push-Pull Oscillator Inverter Crossover

transistor storage time created overlap or other circuit switching problems, the inverter common leg or source current would be the continuous I_{cc} value. However when the off-transistor is driven on, because of storage time of the off-going transistor, a transitory condition occurs during crossover where both collector currents are flowing simultaneously. These collector currents buck in the power transformer primary and a low impedance loading condition exists for the duration of the off-going transistor storage time interval. The resultant current surge spiking is typically illustrated in Figure 4. For this time interval, current flow from the source is limited only by the collector-to-emitter impedances and the impedance of the power source. The large current spike, multiplied by the full voltage of the power source is power lost. Efficiency is reduced, and excessive instantaneous transistor dissipation can compromise transistor reliability, particularly in the case of a very low-impedance source.

In addition to transistor storage time created overlap, transistor rise and fall times and other circuit component switching characteristics influence the transient switching interval. The current/voltage spiking condition existent determines the amount of filtering required before and after the conversion process.

B. Current Feedback Driven Oscillator Inverters

Figure 5 typically illustrates the current feedback driven push-pull oscillator inverter crossover conditions. Although transistor storage time exists

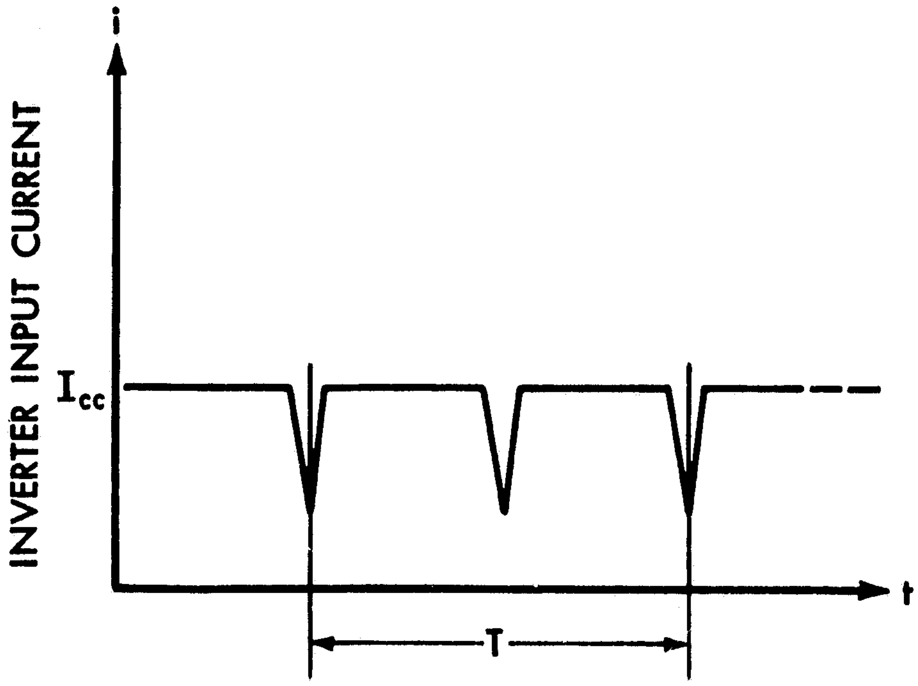


Figure 5. Typical Current Feedback Driven Push-Pull Oscillator Inverter Crossover

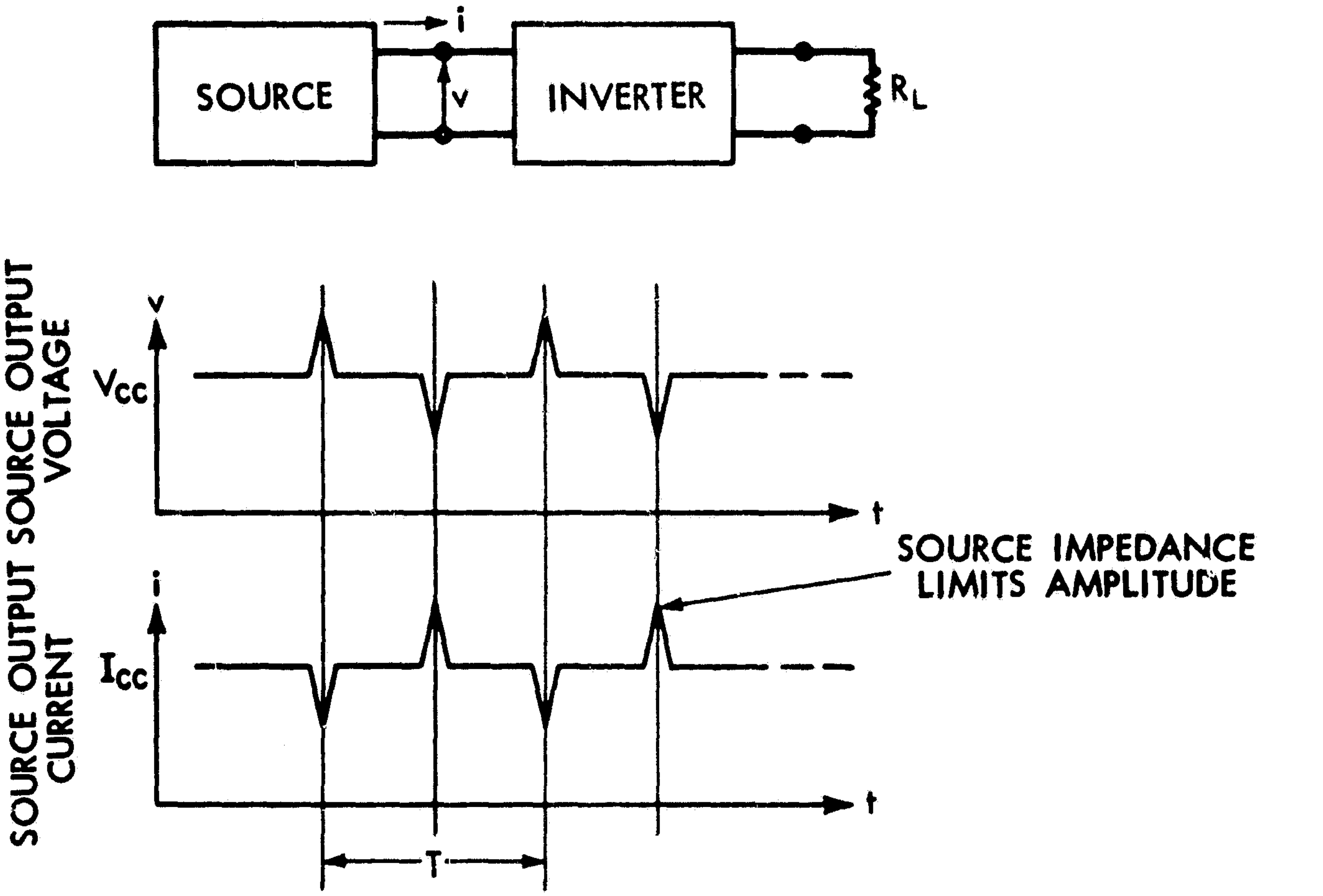
it is much less than for the voltage feedback drive case as overdriving is reduced because of the load related base drive relationship. It is important also to note that inverter switching is not initiated until whatever storage time exists is essentially over. The transient reduction of source current as illustrated is due to rise and fall time of transistors and other circuit components. Filtering before and after the conversion stage is needed to reduce the voltage/current ripple produced by this crossover result. Particularly in the case of a high impedance source, transistor reliability is compromised by voltage stresses from the resulting source voltage spiking.

C. Current Feedback Driven Oscillator Inverters with Source Voltage Controlled Switching Rate

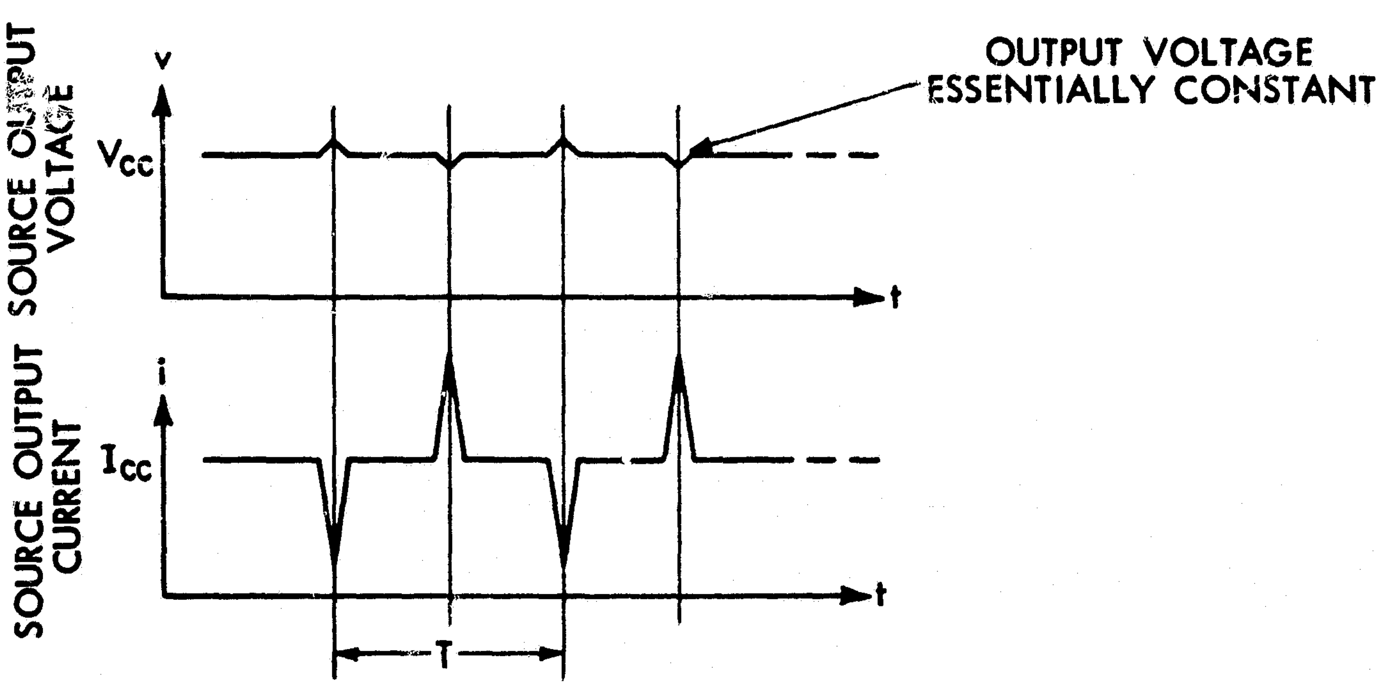
The crossover conditions for these inverter configurations are a combination of the individual characteristics as described previously in Sections A and B. The exact condition resulting is a combined function of: the positive current feedback; the negative voltage feedback; circuit component switching characteristics; and input/output filtering.

D. Inverter Crossover, Source Ripple, and Filtering

Figure 6 typically demonstrates a range of source v/i ripple brought about by the crossover characteristics of inversion and as influenced by the source impedance. In the case of a low impedance source, or where a combination of



HIGH IMPEDANCE SOURCE



LOW IMPEDANCE SOURCE

Figure 6. Typical Source (v/i) Ripple

primary and secondary sources give an effective low impedance source, the overlap high current spiking resulting from transistor storage time can be a transistor and a source reliability and an efficiency reducing condition. In the case of a high impedance source, the crossover created source v/i ripple is also a transistor and source, efficiency and reliability reducing factor. Any required use of additional filtering may be impractical from a size and weight viewpoint, as well as from a reliability viewpoint, considering the additional components needed and their associated voltage/power stresses.

V. A NEW BASE CURRENT CONTROL AND SHAPING CIRCUIT TECHNIQUE TO IMPROVE INVERTER/CONVERTER CROSSOVER PERFORMANCE AND REDUCE SOURCE RIPPLE

The crossover performance of push pull oscillator inverters/converters using inherently slow switching as well as long storage time (low saturation resistance germanium units) transistors can be improved by using circuit approaches which optimize overall inverter/converter crossover performance.

A previously developed circuit technique (reference 4) utilized a base-to-base connected inductance to reduce the storage time interval. This was only partially effective, a serious problem being, that the conducting transistor drive was reduced prematurely causing a nonsaturated, high loss condition prior to completion of inverter switchover. Additionally the method could not deal with the more serious storage time problems of inverter transistors where the inverter is subject to variable loading or is operated from a variable input voltage source.

A new and basic concept which effectively deals with the total cycle base drive condition has been developed at GSFC. The method is applicable in output transformer primary and secondary winding current feedback driven push-pull oscillator inverters with source voltage controlled switching rate. The method is also applicable with voltage feedback driven oscillator inverters with source voltage controlled switching rate. Low input voltage inverters/converters using this circuit approach have proven effective, efficient, and reliable in both low and high impedance source applications.

A. Concept, Implementation, and Performance

The new technique involves using an inductor (L) in series with the normally used frequency determining saturable reactor (SR) in the negative voltage feedback, frequency determining, loop. This is illustrated in Figure 7. Including the proper inductance in the frequency determining loop network enables

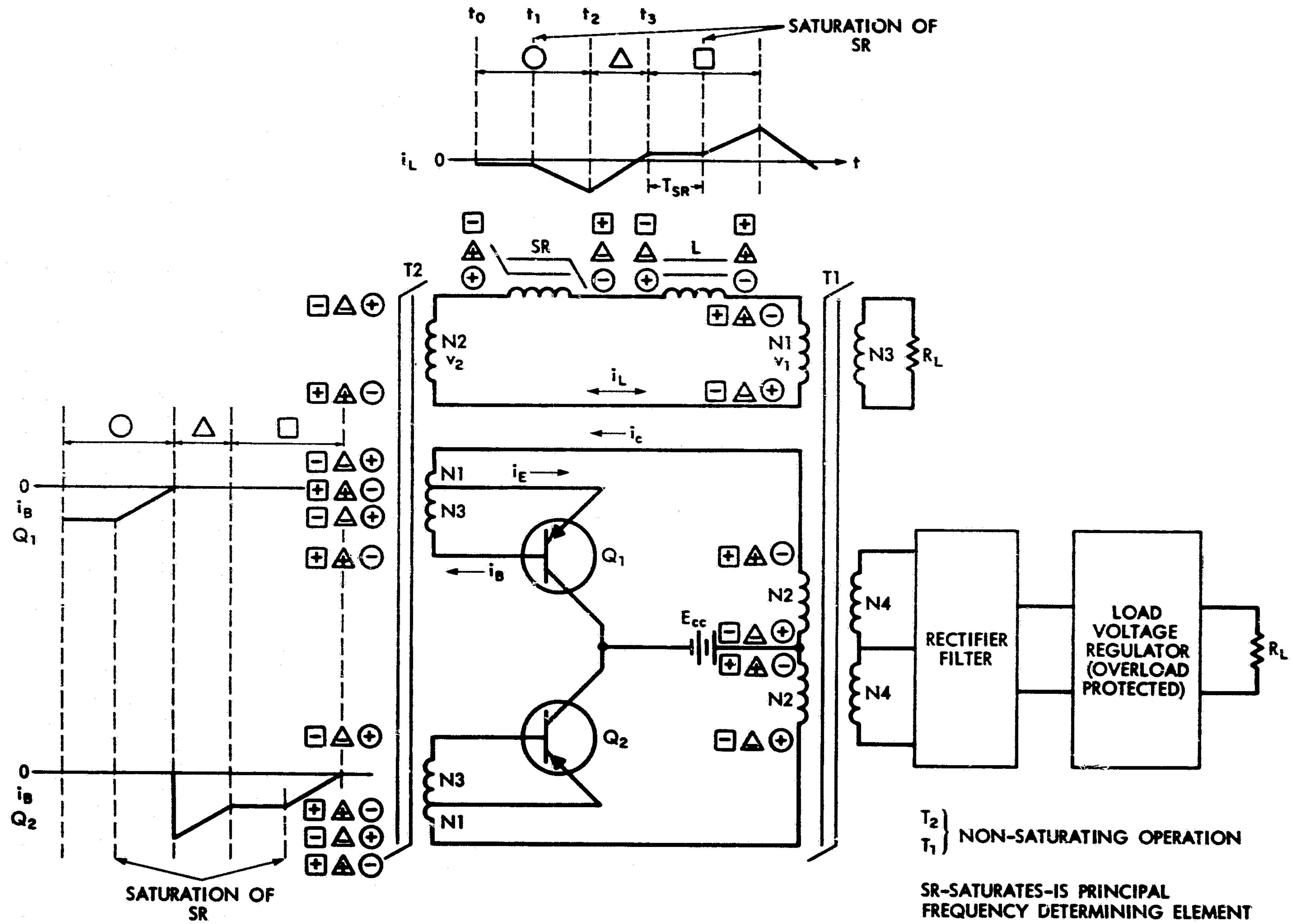


Figure 7a. Improved Crossover Circuit

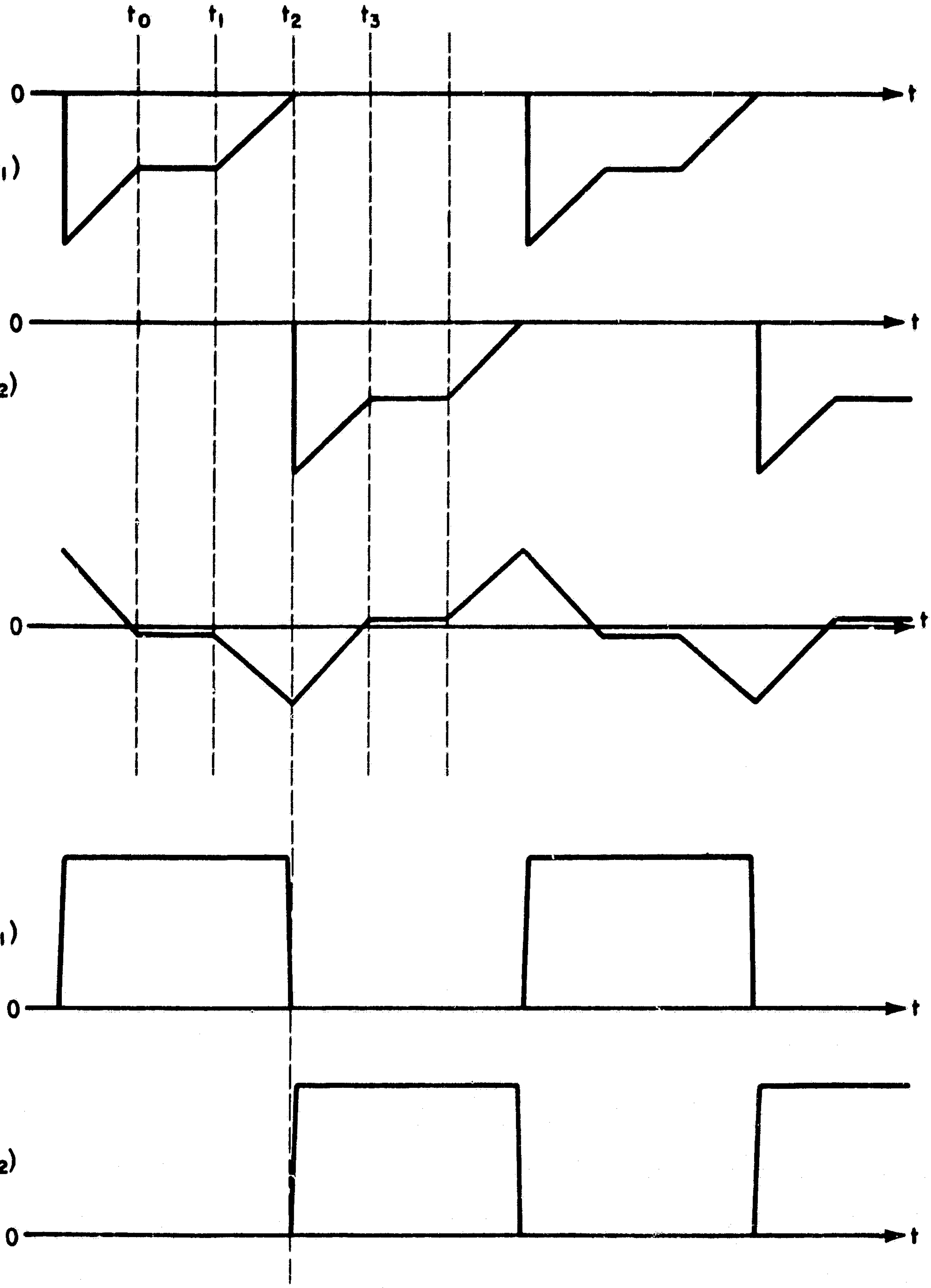


Figure 7b. Improved Crossover Circuit Idealized Waveforms

an electronically complex but smoothly controlled and very efficient switching action of the inverter that results in improved crossover. This switching action will be described for the several discrete intervals comprising one-half cycle of inverter operation.

$t_0 t_1$ Interval

In the circuit of Figure 7a the initial assumptions are made that the SR is not saturated and Q1 is the conducting transistor. v_1 is source-voltage-related negative voltage feedback from the power output transformer T_1, and v_2 is voltage induced in the N_2 winding of the current transformer T_2. v_1 is made much larger than v_2 to make the frequency of operation essentially proportional to source (E_{cc}) voltage. For the initial condition that the saturable reactor (SR) is not saturated, the circuit of Figure 7a operates as a normal current feedback driven, push-pull, transformer coupled output, power oscillator, DC to square wave AC inverter or DC to DC converter. During the interval that the SR is not saturated, the frequency determining network loop impedance is high, and the positive feedback base current (i_B) is related to the collector current (i_C) by

$$i_B = i_C \frac{N_1}{N_3} \qquad \text{(eq 1)}$$

$t_1 t_2$ Interval

Referring to Figures 7a, and 7b, v_1 and v_2 are in phase and act to saturate the SR at t_1. After saturation, the SR impedance is negligibly small. During the interval t_1 to t_2 the i_L current in the frequency determining loop rises and the i_L value at t_2 is given by

$$i_L(t_2) = \frac{v_1 + v_2}{L}(t_2 - t_1) + i_L(t_1) \qquad \text{(eq 2)}$$

where t_1 = time of SR saturation
t_2 = inverter switchover
$i_L(t_1)$ = SR magnetizing current (negligibly small).
v_1 and v_2 are assumed constant in the interval t_1 to t_2. The i_L current is illustrated in Figures 7a and 7b.

During the interval t_1 to t_2, the N_2 and N_3 windings of T_2 share i_C current according to the relationship

$$i_B = i_C \frac{N_1}{N_3} - i_L \frac{N_2}{N_3} \qquad \text{(eq 3)}$$

where $i_L \frac{N_2}{N_3}$ is the Q_1 base current reduction effected during the interval t_1 to t_2. Since i_C is fixed by the load, any increase of i_L results in reduced i_B.

Inverter Switchover (t_2)

The rising i_L current, as i_B approaches zero, in accordance with eq 3, is limited to

$$i_L(\max) = i_C \frac{N_1}{N_2} \qquad \text{(eq 4)}$$

where i_C is fixed by the load. When i_L approaches this transformer ratio reflected $i_B = 0$ value, i_L cannot increase further, and voltage reversal of the inductance occurs, which also reverses the polarities of the base driving transformer initiating switchover. Switchover thus occurs when the stored base charges effect is very minimal.

$t_2 t_3$ Interval

The current rise in L during the interval t_1 to t_2 results in energy storage in its magnetic field. One part of this stored energy is from the N_2 winding of T_2. The remaining part of the stored energy is from the N_1 winding on T_1. It is the energy withdrawal from the source of energy $N_1 T_2$ which reduces the available energy in the base circuit of Q_1. The stored energy in L is recovered during the interval t_2 to t_3. One part of this recovered energy ($p_2 dt$ of v_2) is delivered to the switching-on transistor Q_2 providing base overdrive during the interval t_2 to t_3. The remaining part of the recovered stored energy ($p_1 dt$ of v_1) is delivered to the N_1 winding of the output transformer T_1 and supplies the load.

At time t_2, Q_2 switch-on has been initiated. During the interval t_2 to t_3, normal current feedback drive is supplied to the base of Q_2 as a result of Q_2 collector current flow in the N_1 winding of T_2. Also, during the interval t_2 to t_3 in addition to the normal i_B drive (Equation 1), an overdrive is supplied by the recovered stored energy in L delivered to the N_2 winding of T_2. This additional drive is $i_L \frac{N_2}{N_3}$. This overdrive condition brings Q_2 quickly into full conduction. The total base drive on Q_2 during the interval t_2 to t_3 is

$$i_B = i_C \frac{N_1}{N_3} + i_L \frac{N_2}{N_3} \qquad \text{(eq 5)}$$

Figures 7a and 7b illustrate the Q_1 and Q_2 base drive conditions.

At time t_3 the SR is reset (becomes a high impedance), and an identical half cycle of operation ensues with Q_2 as the conducting transistor.

B. Effect of Load on Inverter Period

As shown by Figures 8a and 8b, the circuit can be implemented to perform in two separate modes of operation, these are:

1. A constant period is obtained when L is reduced appropriately as a function of load.

2. A variable period is obtained when the load varies and L is held constant.

Implementation 1 above is desirable in most cases to enable minimization of power output transformer weight and size, and to avoid power output transformer core saturation which could occur with lowered frequency of operation.

A method of varying L with load is illustrated by Figure 9a. In Figure 9a the I_{cc} current which is a function of load varies the permeability of the core material of inductance L, thereby modifying its effective inductance value with load.

Two half cycles of base current conduction compromising inverter period T are typically illustrated in Figure 9b where

$$T = 2T_{SR} + 4T_L \qquad \text{(eq 6)}$$

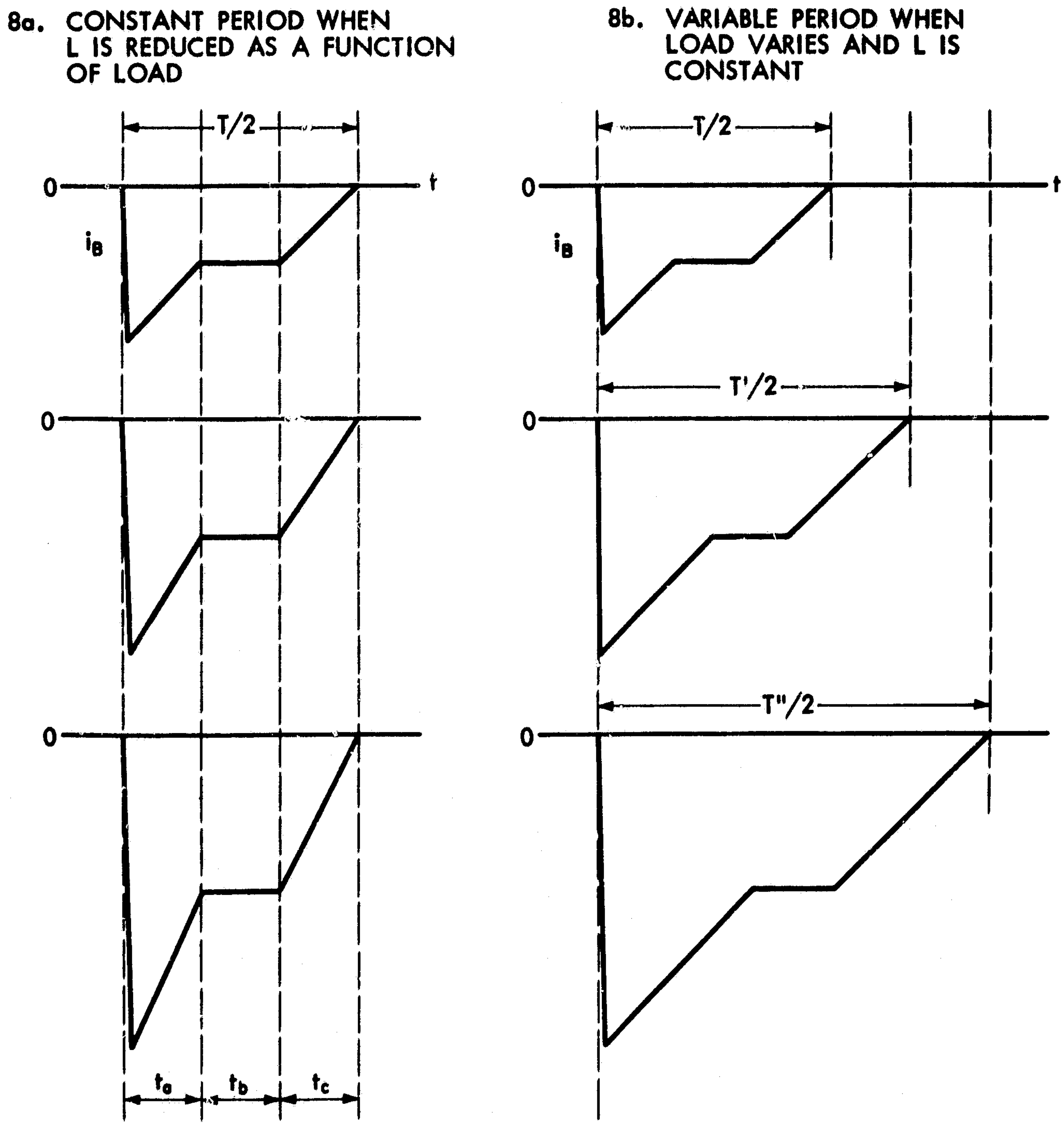
T_{SR} is independent of loading. Thus if T is to be made constant with load, T_L must be a constant with load.

$$i_L(t_2) = \frac{v_1 + v_2}{L}(t_2 - t_1) + i_L(t_1) \qquad \text{(eq 2)}$$

where
$i_L(t_1)$ = negligibly small SR magnetizing current
$i_L(t_2) = i_L(\max)$ at switchover
v_2 is a function of load, but is also negligibly small
$(t_2 - t_1) = T_L$

From equation 2,

$$T_L = \frac{L i_L(\max)}{v_1} \qquad \text{(eq 7)}$$



DURING t_a $\quad i_b = i_c \dfrac{N_1}{N_3} + i_L \dfrac{N_2}{N_3}$

DURING t_b $\quad i_B = i_c \dfrac{N_1}{N_3}$

DURING t_c $\quad i_B = i_c \dfrac{N_1}{N_3} - i_L \dfrac{N_2}{N_3}$

i_b VARIES AS i_c WHICH IS A FUNCTION OF TOTAL LOADING

Figure 8. Effect of Load on Inverter Period

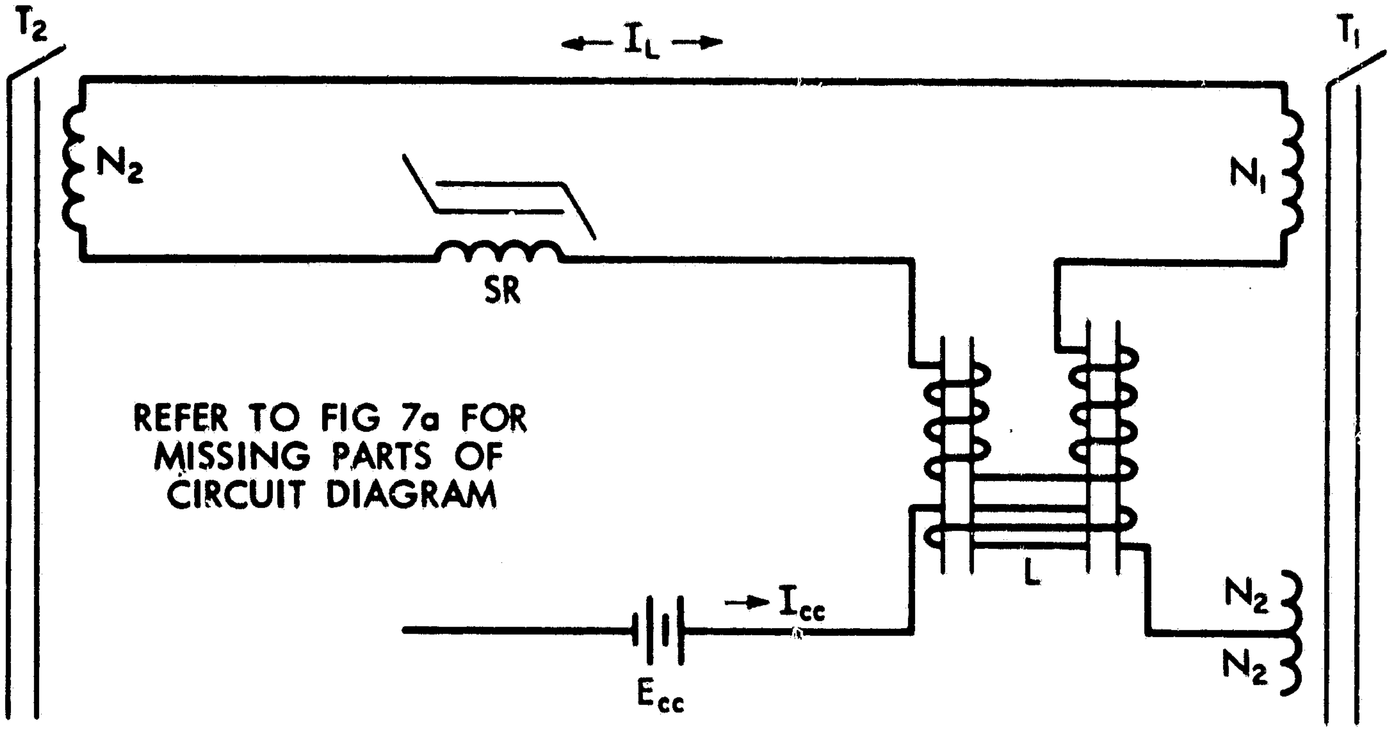


Figure 9a. Method of Varying Effective Inductance of L with Load

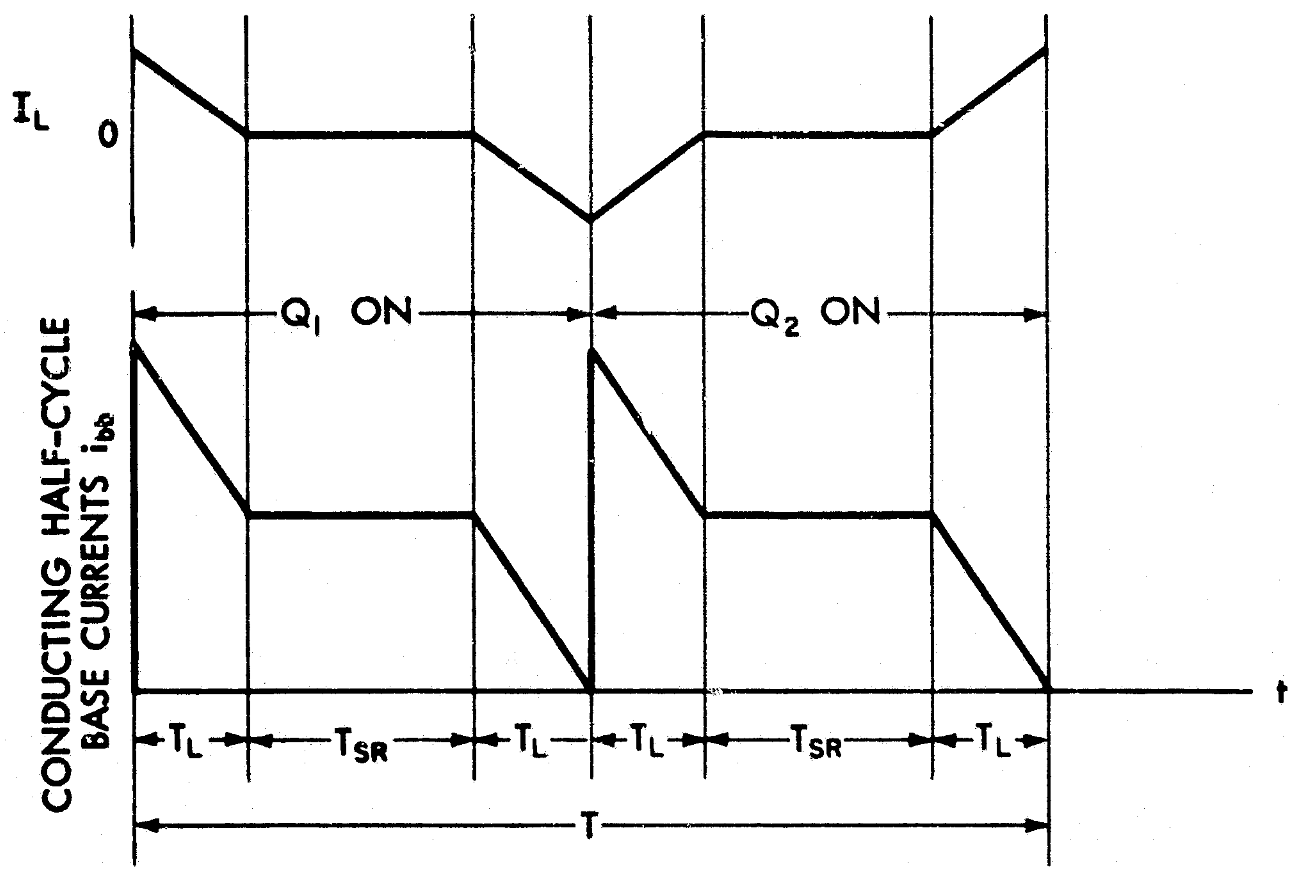


Figure 9b. Conducting Half-Cycle Base Currents Versus Inverter Period (T)

From equations 4 and 7, equating i_L(max) values, obtains

$$T_L = \left(\frac{1}{v_1}\right)\left(\frac{N_1}{N_2}\right)\left(Li_c\right) \qquad \text{(eq 8)}$$

Since v_1 is assumed constant in the interval t_2-t_1, a necessary condition for T_L to be constant is therefore

$$Li_c = Li_{cc} = K \qquad \text{(eq 9)}$$

The optimum T_L is selected at the worst crossover condition which is the maximum load condition. The $Li_{cc} = K$ relationship has been physically closely realized (Figure 15a) from i_{cc} (max) to i_{cc} (max)/4 values. Equation 9 indicates that approaching $i_{cc} = 0$, the $Li_{cc} = K$ relationship is not physically realizable. Below $\frac{i_{cc}(\text{max})}{4}$, crossover ripple represents a negligible condition.

The value of L should be selected or controlled so that the rate of i_B decrease results in optimum inverter switchover. It is necessary to achieve minimum stored base charges at switchover for minimum storage time (thus minimum overlap). While one selected value of L can be made to work reasonably well over most expected load range conditions, it is, however, better to control L as a function of load. This procedure has the advantage that the frequency can be maintained essentially proportional to input voltage, while the frequency variation with load is minimized.

The approximate value of L required is obtained by solving equation 2 for L. By assuming that all N_1T_2 (Figure 7) energy is being delivered to N_2T_2 at time t_2, then

$$i_L(t_2) = i_c \frac{N_1}{N_2} \qquad \text{(eq 4)}$$

Also, assuming that the SR magnetizing current is negligibly small, then

$i_L(t_1) = 0$, and from Equation (2)

$$L = \frac{(v_2 + v_1)(t_2 - t_1)N_2}{i_c N_1} \qquad \text{(eq 10)}$$

where $(t_2 - t_1)$ approximates the maximum storage time (at i_c max) of the transistors Q_1 or Q_2. The optimum values of L required can be readily determined experimentally by adjustment of the value of L to effect minimum input ripple conditions (current ripple and voltage ripple are functions of source impedance). The optimum L values will also result in optimization of overall efficiency.

C. An Alternate Oscillator Inverter Configuration with Current Feedback Drive and Source Voltage Controlled Switching Rate

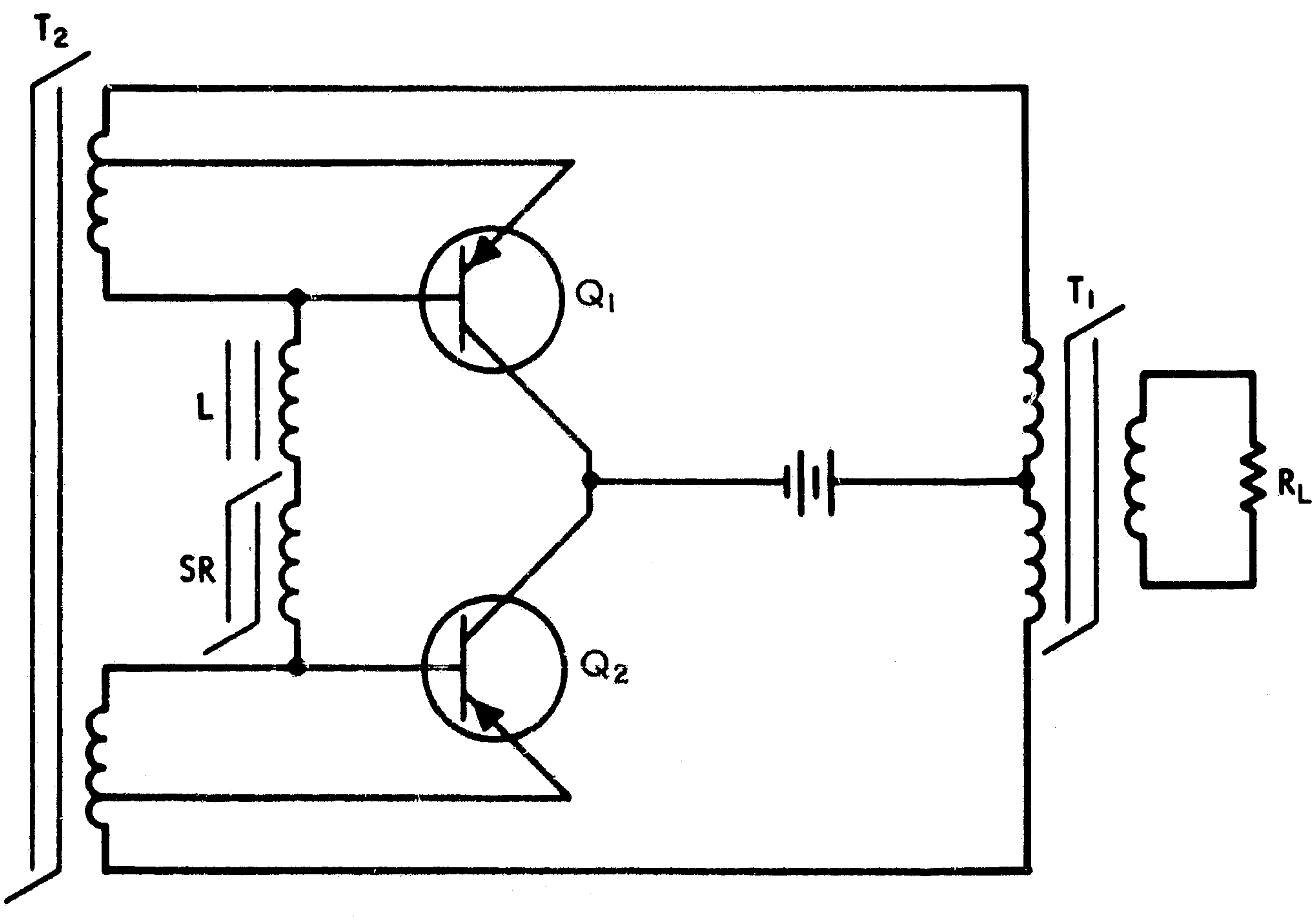


Figure 10. Improved Crossover Technique in an Alternate Push-Pull Oscillator Inverter Configuration with Current Feedback Drive and Source Voltage Controlled Switching Rate.

The principle of operation of this inverter circuit is essentially the same as the previously explained inverter circuits of the current feedback driven type. The possible improvement in circuit rise time characteristics by the elimination of some transformer bandwidth limitation makes this configuration approach useful.

D. Voltage Feedback Driven Inverter Application and Limitation

The new crossover technique can also be used to advantage in purely voltage feedback driven, push-pull power oscillator inverter circuits. Figure 11 shows an example of its use in a positive voltage feedback base drive application. Referring to Figure 11, the voltage v_{FB} provides positive feedback voltage which maintains either Q_1 or Q_2 in conduction. The operation of the frequency determining loop and base current shaping action is similar to the cases presented for the current feedback drive applications. However, the absolute value of i_B in this case is not a function of i_C (therefore load), as in the current feedback base drive cases presented previously.

E. The Inductor

The operation and explanation of the new concept and implementation has been with the use of a physical, lumped parameter inductance in series with the main frequency determining element SR. Other physical arrangements are possible by using different transformer core materials, core slitting, residual or introduced distributed inductance.

F. Characteristics and Advantages

Referring to Figure 12 it is observed that the new circuit approach for improved Inverter Crossover:

1. Separates and controls base current reduction (off-going transistor) and base current overdrive (on-coming transistor) to two distinct and separate time intervals. The initiation of switchover occurs only after stored base charges are minimized and conditions for switching optimized.

2. The base current of the switching-off transistor 1 is reduced in a controlled manner, and as a function of load (note A) prior to switchover initiation, to a minimal value at the instant of switchover (note B). There is a minimal storage time existent at switchover thus reducing the storage time created overlap problem.

3. The overdrive occurs after switchover (note C); this results in a rapid rise time of the switching-on transistor thus minimizing the crossover problem. The overdrive current of the switching-on transistor 2 (note C) is a function of load current (twice the normal load-related drive).

4. The base voltages contain no serious peak amplitudes (note D & E). This eliminates transistor reliability problems due to $V_{EB_{REV.}}$ (Max.) limitations.

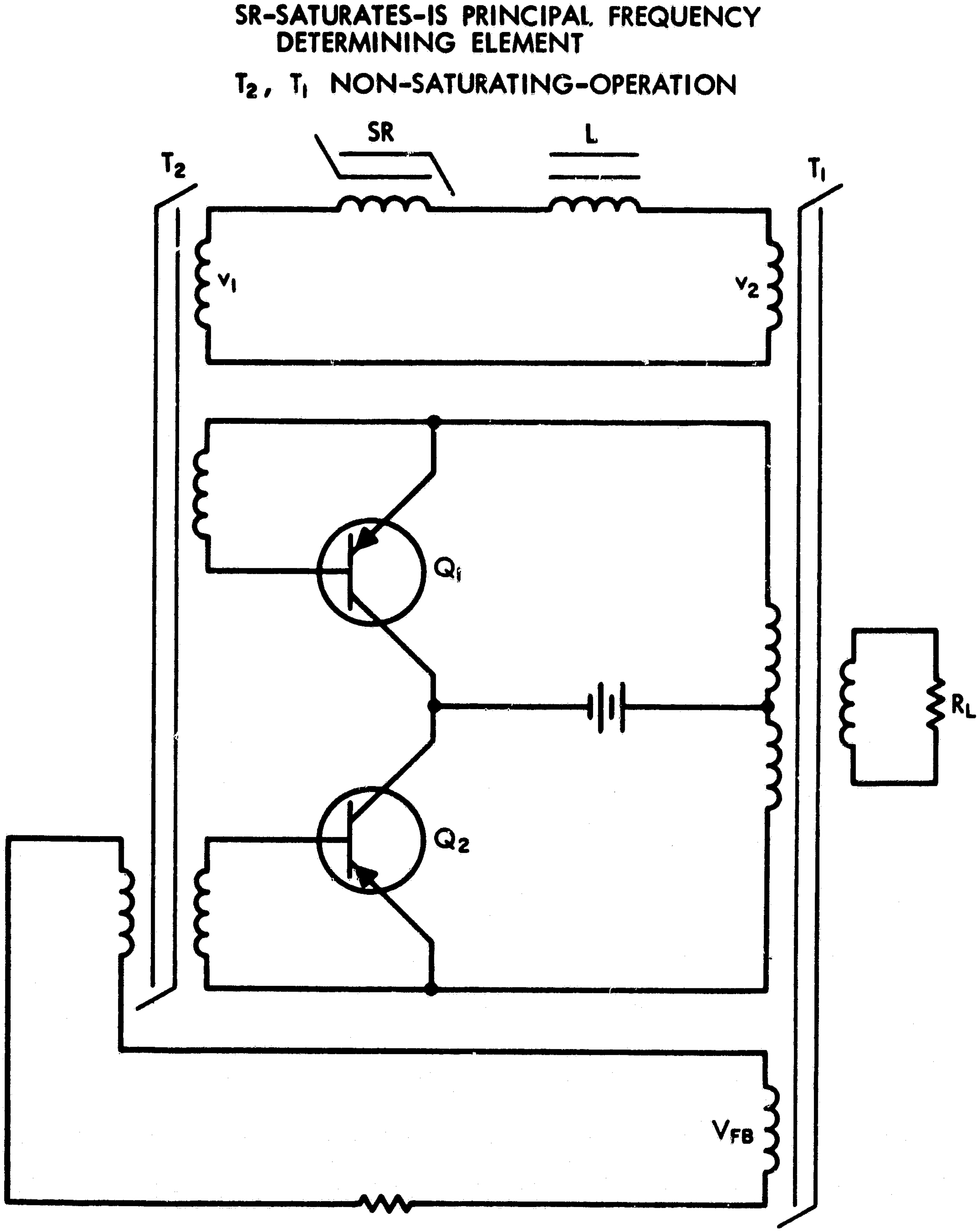


Figure 11. Improved Crossover Technique in a Voltage Feedback Driven Push-Pull Oscillator Inverter with Source Voltage Controlled Switching Rate

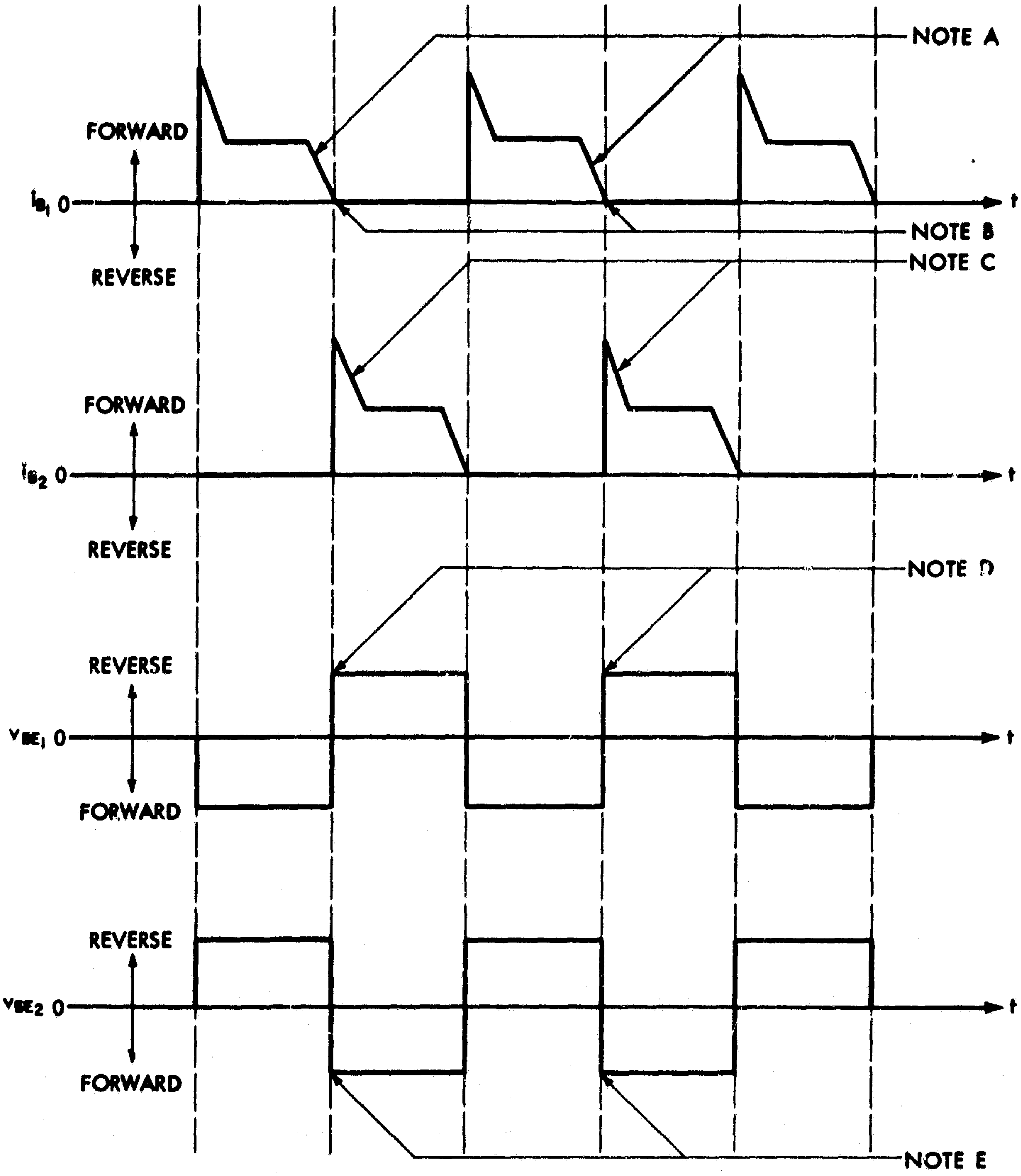


Figure 12. Improved Crossover Technique Idealized Base Drive Waveforms

5. The technique is nondissipative except for minimally small wire and core losses, essentially all stored energy being retrieved, therefore it is very efficient.

6. Treats the transistor as a current controlling device, which it is.

7. Can be used with either power output transformer primary or secondary winding current feedback inverters or in voltage feedback driven inverters since the switchover is controlled by the frequency determining loop functioning interval only.

8. Enables total circuit switching time performance to approach closely the intrinsic transistor component rise and fall time.

9. Permits minimizing of overlap and switching inverter crossover ripple over a wide range of load demand and input voltage.

10. Provides proper base drive current shaping for optimum inverter performance over the total cycle.

11. The new crossover technique is useful in the case of both high and low impedance sources. The base current control and shaping as used with push-pull oscillator inverters effectively improves total inverter crossover performance, and eliminates most of the transient voltage and current spiking conditions usually present at the source output. Output ripple conditions at the load are also reduced without the need of using additional filtering. Thus minimal input/output ripple can be obtained without incurring a size and weight penalty, and with improved reliability.

VI. PROTECTION AGAINST DESTRUCTIVE POWER TRANSFORMER SATURATION CURRENT SURGES IN CURRENT FEEDBACK INVERTERS/CONVERTERS

A. High-Impedance/Power-Limited Source and Primary Winding Current Feedback Driven Inverters

Figure 13 is a power supply using a load related current feedback inverter, with source voltage related negative voltage feedback controlling the switching rate. Included in the frequency determining loop is the inductor L which improves the total crossover performance with regard to transistor storage time created overlap. Primary transformer winding (T_1) current feedback is used to keep the inverter transistors well into saturation limiting collector to emitter dissipation (P_{CE}). In the presence of abnormal saturation of the power

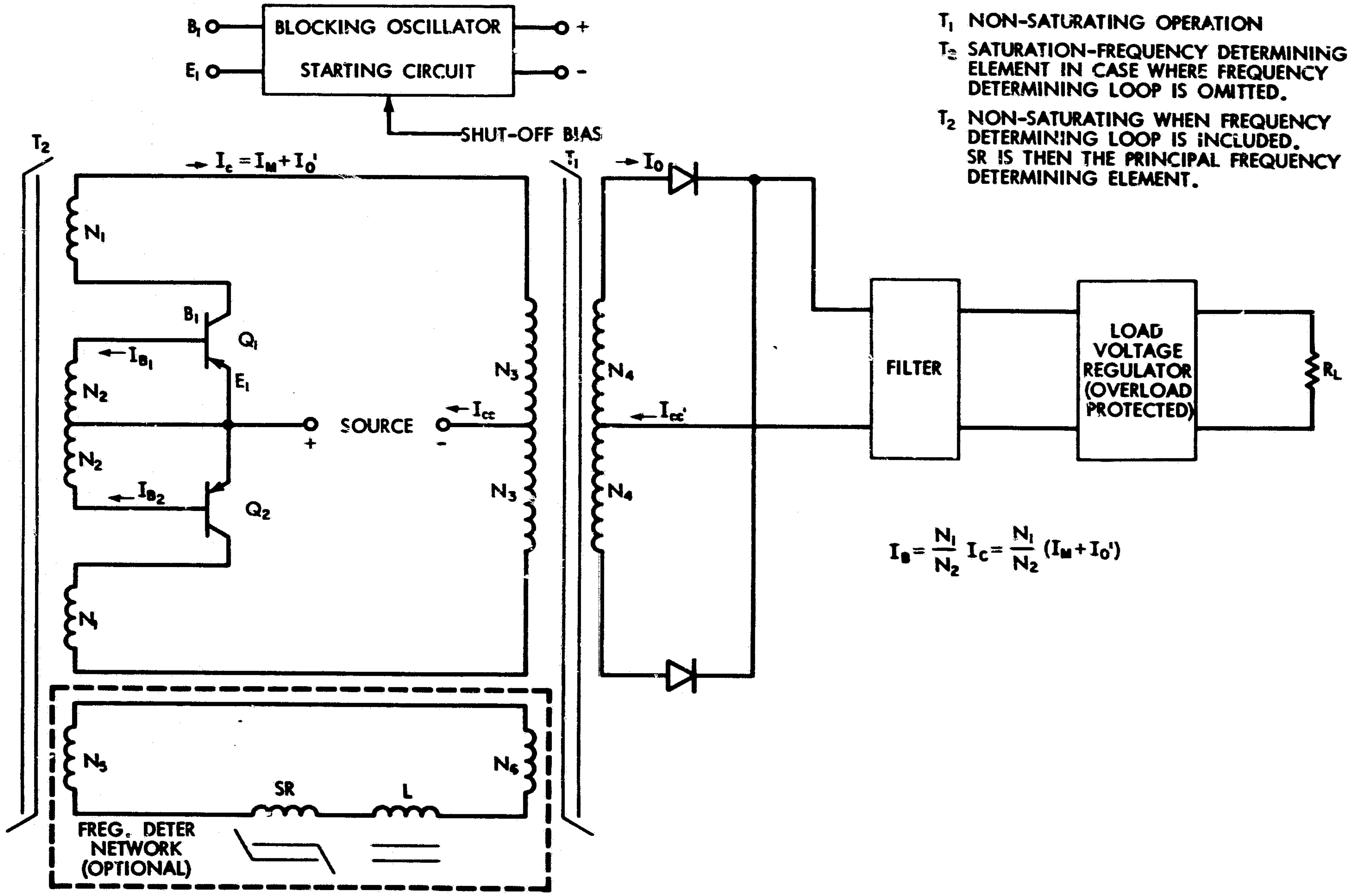


Figure 13. Power Supply Using Current Feedback Driven Converter-Regulator (Optionally with Source Voltage Controlled Switching Rate) for Use with Power-Limited Sources

transformer (T_1) the maximum current amplitude is limited to the source maximum current capability, which is approximately twice the normal load current for matched-load maximum output power conditions. All inverter/converter components can be sized to meet this maximum source current capability. Such a system can be made very reliable. A detailed description of this type power system is included in reference 1 therefore it will not be further discussed in this report. However it should be noted that the frequency determining network consisting of $N_6(T_1)$, L, SR, and $N_5(T_2)$ can be omitted if a straight current feedback drive operation is more applicable. For such a case T_2 is designed to saturate and be the inverter frequency determining element. Since V_{EB} does not change much with loading, the straight current feedback drive inverter frequency rises only slightly with increased loading.

B. Low Impedance Source and Primary Winding Current Feedback Inverters

In cases of transient, or steady state, short or excess loading conditions, with a low impedance essentially unlimited power source, the source output current is not limited. This leads possibly to catastrophic destruction of the source or inverter, if the power output transformer primary winding current feedback drive inverter circuit (typically illustrated in Figures 2 and 13) is used. The problem here is that in the presence of abnormal main power transformer core magnetic saturation, the transformer magnetizing current requirement is a part of the load-related feedback current. The current feedback maintains the transistor collector to emitter impedance low despite the increasing magnetizing current requirements due to saturation. The core saturation effectively short circuits the transformer windings. The only source current limiting is by the collector-emitter impedance, which however, is maintained low by the current feedback. Thus when the power transformer primary winding current feedback driven inverter is used, the destructive current level problem is additionally compounded in the presence of main power transformer core magnetic saturation. To prevent this catastrophic failure possibility, it is necessary to unload the source from the possible short circuiting effects of output transformer core saturation.

C. Low Impedance Source and Secondary Winding Current Feedback Driven Inverters

Pearlman (reference 5) originated the secondary winding current feedback inverter. By using secondary winding current feedback, an inverter configurational approach has been successfully developed to reduce catastrophic failure possibility of excess current flow in the case of abnormal main power transformer magnetic core saturation. This application will be described for

(1) inverter applications using a load-related current feedback driven inverter, as well as for, (2) the load-related current feedback driven inverter with source voltage related frequency of operation.

Figure 14 illustrates a power system in which a load related current feedback driven inverter is used. In case 1 (frequency determining loop omitted), T_2 is a saturable base transformer whose design, and the V_{BE} voltage of the conducting transistor determine the frequency of operation. The use of a secondary transformer winding current feedback configuration separates the core magnetization current requirements from the load positive current feedback line. In normal operation base current drive is a function of load current where

$$I_B = I_o \frac{N_1}{N_2} \qquad \text{(eq 11)}$$

The collector current $I_c = I_m + I_o'$
where I_o' = primary side reflected load current. In the presence of main power transformer saturation, I_M increases, but feedback current I_0 diminishes. Thus load feedback base current drive of the conducting transistor is reduced, effectively decoupling the short circuit loading caused by T_1 core saturation, and preventing destructive current flow through the source, transformer, and collector to emitter loop. By this method load-related current feedback drive use has been enabled while inserting a self-protecting mechanism from excess magnetization current in the event of the main power transformer magnetic core saturation.

The secondary winding current feedback inverter approach also protects against destructive source/inverter surge currents due to transistor storage time created overlap, since drive is removed if both transistors should conduct simultaneously. Some tolerable source current ripple is experienced in both the cases of saturation and transistor overlap occurrence.

The Figure 14 inverter schematic can be modified by addition of the frequency determining loop as shown by the dotted enclosure. This addition enables a power system using a load-related positive current feedback inverter with source-related negative voltage feedback controlled inverter switching (case 2). That is, the inverter frequency is now a direct function of source voltage being determined mainly by the voltage time integral of the saturable reactor SR. It is important to note that this inverter design now is protected against T_1 core saturation effects, as well as against transistor storage time created inverter overlap, and has good crossover performance enabling a minimum of source v/i ripple. However, as also illustrated in Figure 14, for complete system protection, the use of an overload protected load voltage regulator on the output side

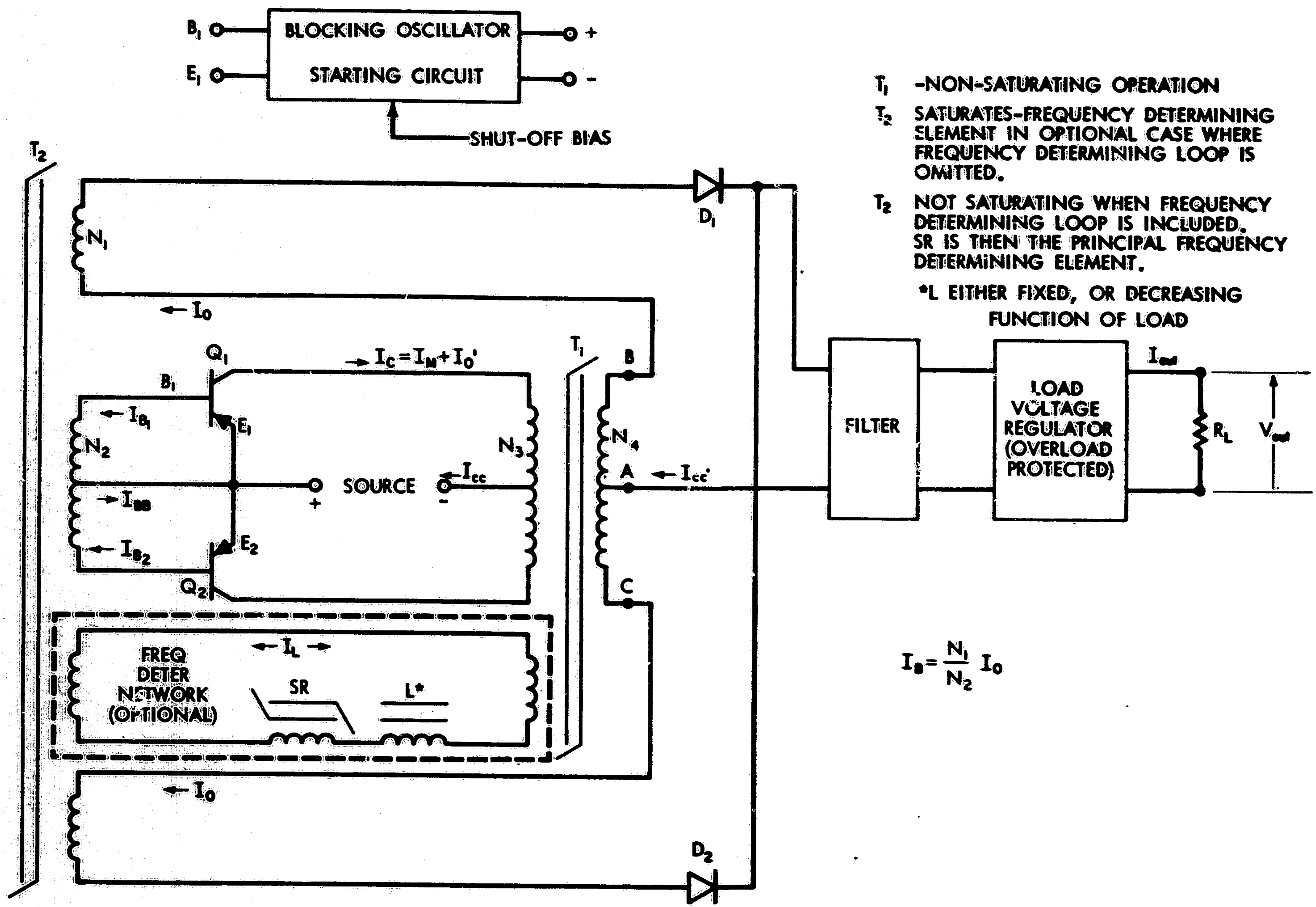


Figure 14. Power Supply Using a Current Feedback Driven Converter (Optionally with Source Voltage Controlled Switching Rate) for Use with Low Impedance Sources

of this power system is mandatory to protect both source and converter against output load short circuit conditions which otherwise would also allow destructive current levels in a current-feedback-driven-converter low-impedance source application.

D. Power Limited Source and Secondary Winding Current Feedback Driven Inverters

In the case of a power-limited source a properly designed secondary winding current feedback driven inverter/converter can also be safely operated into short circuit output load conditions without the need of additional overload protecting circuitry. Thus the secondary winding current feedback driven inverter/converter is more universally applicable with regard to source impedance considerations.

VII. VERIFICATION OF ANALYSIS

The waveforms, graphs and other information presented in this section relate to the implementation wherein inductance L is changed, as previously shown in Figure 9a, maintaining the inverter frequency essentially constant with varying load. The LIVCR partially schematic and partially block diagram is shown in Figure 14.

A graph of the measured inductance L versus input current of the breadboarded LIVCR unit is shown in Figure 15a. As also shown in Figure 15a, relatively constant frequency has been effected over the major portion of the loading range with the inductance L values generated. An $LI = K$ curves (eq 9) as shown in Figure 15a is needed to better the frequency stability with load. The $K = 8400 \times 10^{-6}$ value is based upon selecting the optimum $L = 210 \times 10^{-6}$ Henries with 40 amperes flowing in the two turn bias winding. From a reliability and practicality viewpoint all that is required in this case is to assure that the frequency is not lowered with load to the point of allowing output transformer saturation to occur.

While the inverter period has been made relatively constant with loading, it is important to note also in Figure 15a that the inverter frequency is a direct function of input voltage as desired to prevent output transformer saturation with increased source voltage.

By using an additional bias winding on L, independent and precise vernier frequency control of the inverter is possible with the improved crossover circuit technique. This feature may be advantageous in other applications.

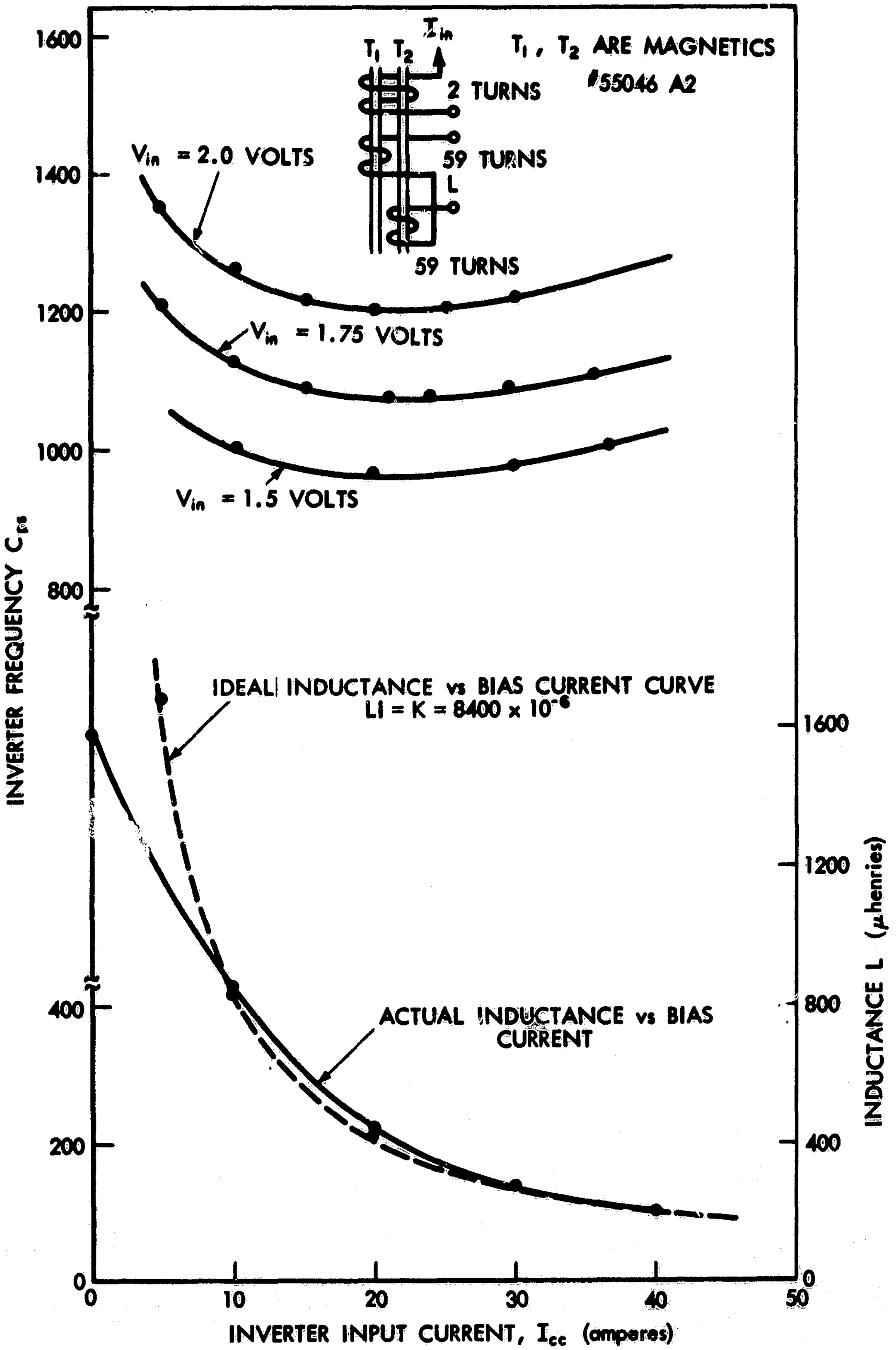


Figure 15a. Inductance L and Inverter Frequency Versus Inverter Input Current

The measured i_L(Max) current is plotted in Figure 15b as a function of inverter input current. i_L(Max) which closely approximates di_L is a direct function of the load related i_{cc} current. The resulting base drive reduction of the off-going transistor, as well as the base overdrive of the on-coming transistor, due to i_L current flow, are also directly load-related functions.

A. Improved Crossover Technique Waveforms

The LIVCR waveforms are shown for the fully-loaded case in Figures 16a through 16j, and waveforms for the case of half-loading are shown in Figure 16a[1] through 16j[1].

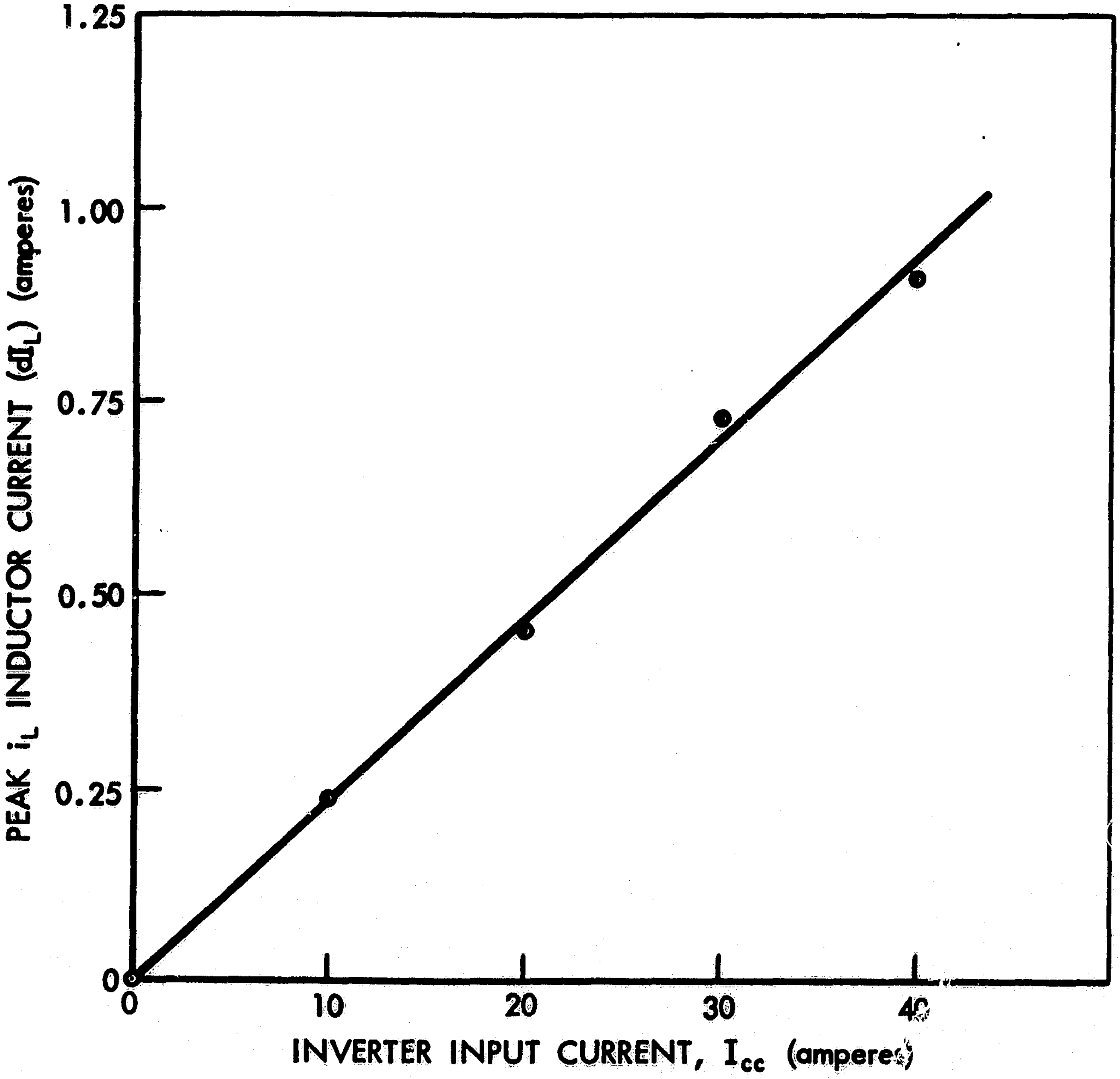
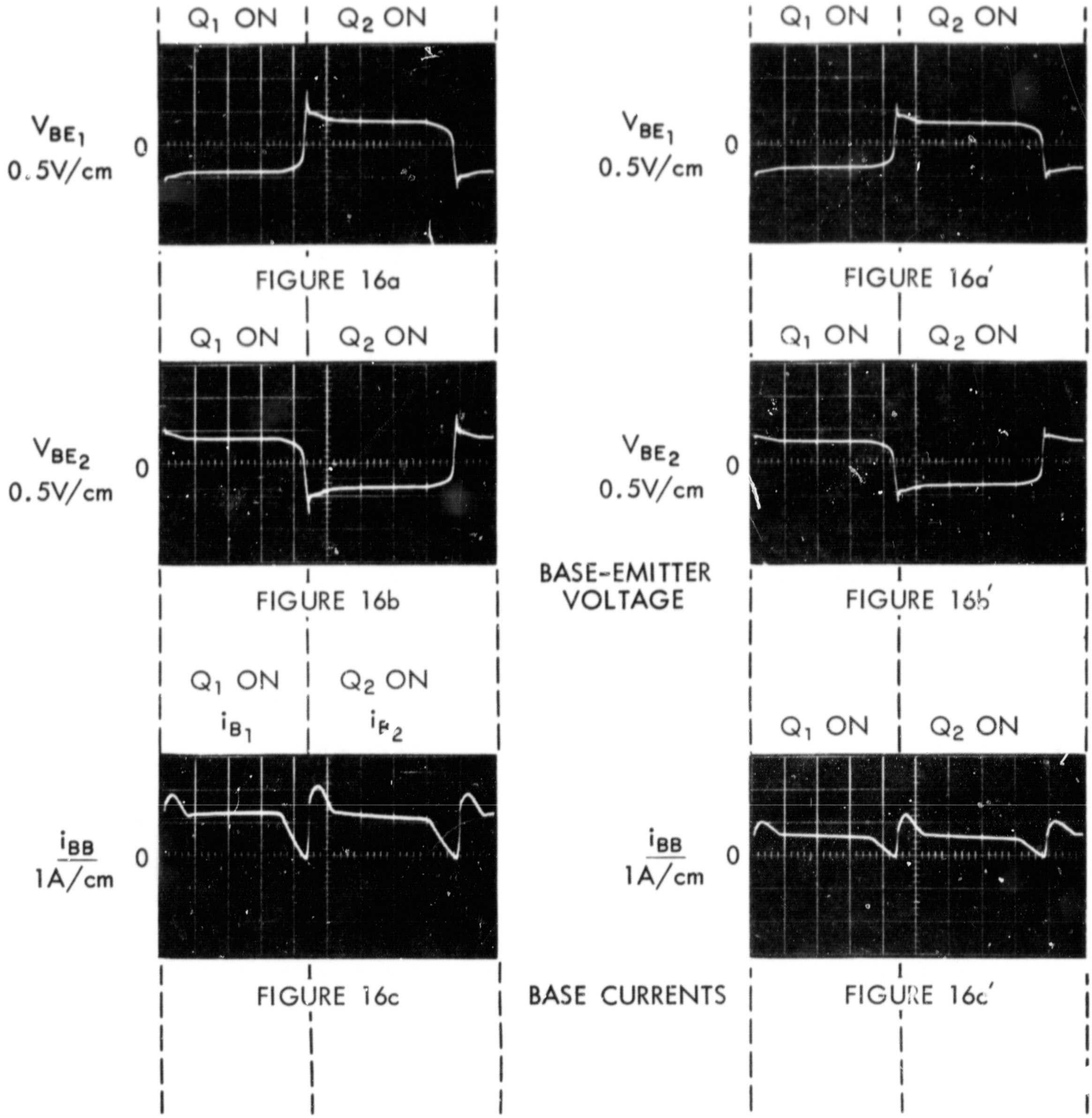


Figure 15b. Peak i_L Current Versus Inverter Input Current



$R_L = 15\Omega$	$R_L = 30\Omega$
$V_{in} = 1.5\,V$	$V_{in} = 1.5\,V$
$I_{in} = 40.2\,A$	$I_{in} = 19.8\,A$
$V_o = 26.2\,V$	$V_o = 26.2\,V$
$I_o = 1.76\,A$	$I_o = 0.88\,A$
$E_{ff} = 76.5\%$	$E_{ff} = 77.6\%$
125μSEC/cm SWEEP	125μSEC/cm SWEEP

Figure 16. Improved Crossover Technique Waveforms

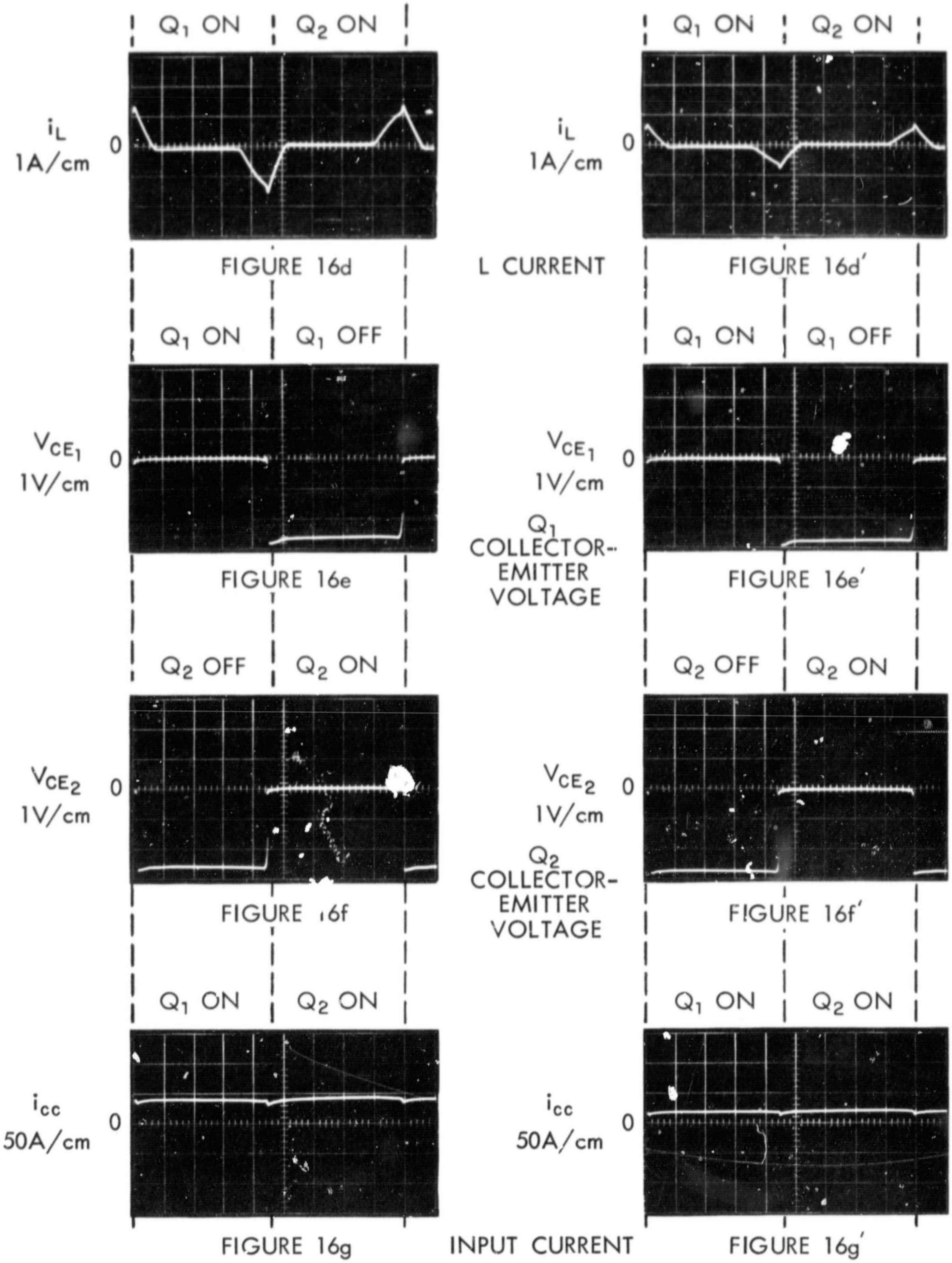


Figure 16. Improved Crossover Technique Waveforms (continued)

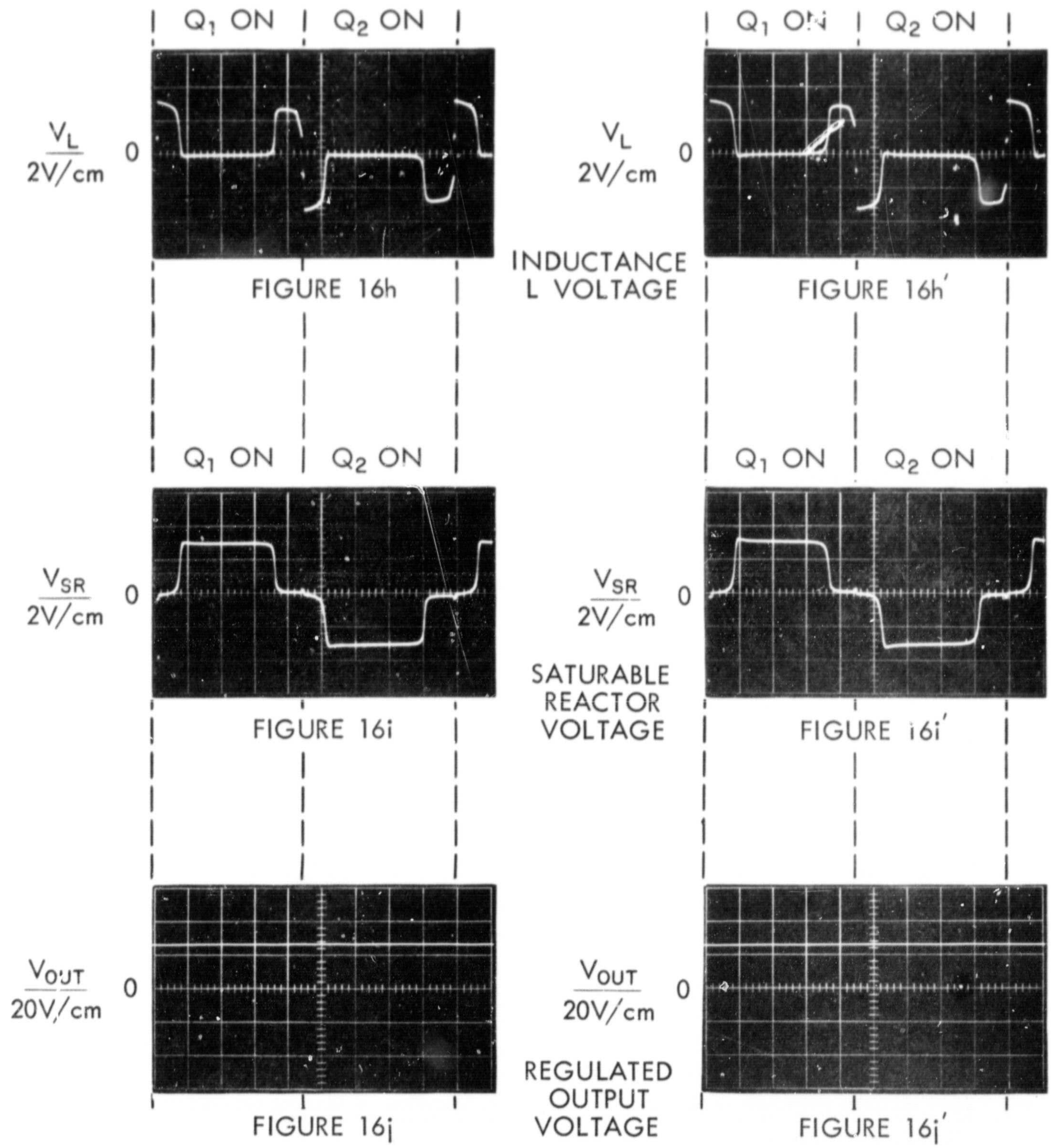


Figure 16. Improved Crossover Technique Waveforms (continued)

Direct comparison of the load-related full-cycle base drive condition is illustrated by Figures 16a, b, c, and 16a^1, b^1, c^1.

The load-related i_L current which produces the base drive reduction or overdrive is illustrated by Figures 16d and 16d^1. Figures 16d and 16d^1 also show that the frequency contribution of i_L (same T_L) has been made essentially constant with load through controlling L by means of the load-related input current.

Figures 16e, f and 16e^1, f^1 shows the fast rise and fall times of the collector voltage. The good source current continuity shown by Figures 16g and 16g^1 evidence the improved crossover technique effectiveness in enabling fast rise and fall times of Q_1 and Q_2 collector current, and also most importantly showing that storage time created overlap has been eliminated. Very low saturation resistance Solitron 2204 germanium transistors were used in the low input voltage DC to square wave AC inverter section of the LIVCR. Two paralleled transistors were used on each side of the inverter for Q_1 and Q_2 respectively (see Figure 19). When subject previously to standard square wave (forward and reverse) base voltage drive, these transistors showed considerable rise, storage, and fall times (reference 6). However when used in the inverter incorporating the new crossover technique full-cycle base drive control, very fast total circuit switching times are realized.

The low current ripple at crossover is obtained without the need of any filtering of the input current. Figures 16i and 16i^1 illustrate that the frequency contribution of the saturable reactor SR (V_{SR} constant with load) is solely a function of input voltage as desired.

The photographic waveforms closely correlate the analysis presented in Section V. The waveforms also show that the optimum full-cycle-drive improves overall efficiency as well as reliability by reducing the component voltage/power stresses over the full cycle. Most important however is the effective reduction of voltage/power stress peaks during the crossover transitory interval.

B. Protection Against Excessive Current Surges in Event of Output Transformer Saturation

1. Half Cycle of Transformer Saturation

Figure 17a is the inverter collector to collector voltage waveform illustrating a case of transformer saturation for a half cycle. The condition was induced by severely unbalancing the base to emitter resistances, causing the output transformer to saturate frequently on one side of its B-H magnetic characteristic.

$V_c - V_c$ 0
1V/cm

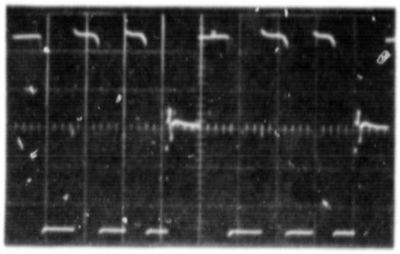


FIGURE 17a

i_b 0
1A/cm

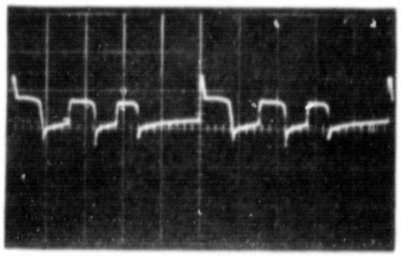


FIGURE 17b

i_{cc} 0
50A/cm

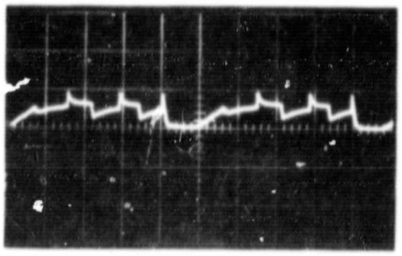


FIGURE 17c

i_{cc}' 0
1A/cm

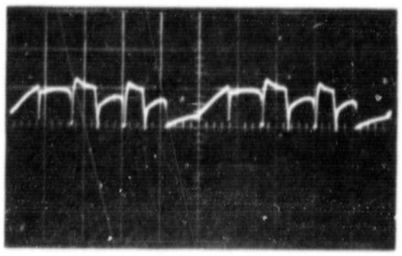


FIGURE 17d

HALF CYCLE
SATURATION CONDITION

$E_{in} = 1.5$V

$I_{in} = 22.1$ A (Average)

$E_o = 26.146$ V Regulated

$I_o = 0.875$ A

$E_{ff} = 69\%$

FREQUENCY = 998 cps.

$R_L = 30\,\Omega$

$V_c - V_c$ 0
1V/cm

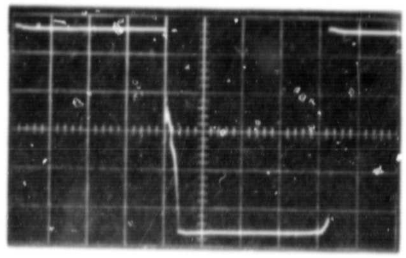


FIGURE 17e

i_b 0
1A/cm

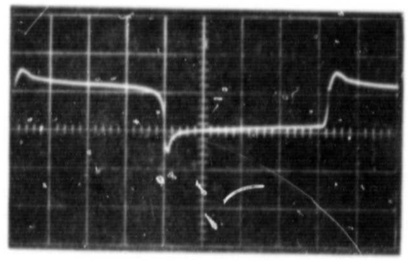


FIGURE 17f

i_{cc} 0
50A/cm

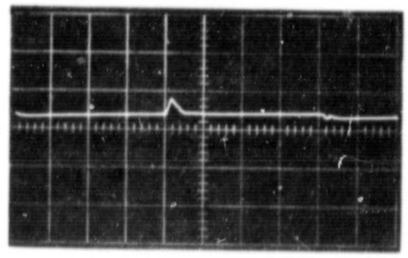


FIGURE 17g

i_{cc}' 0
1A/cm

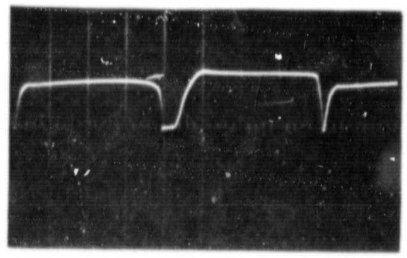


FIGURE 17h

END OF HALF CYCLE
SATURATION CONDITION

$E_{in} = 1.5$V

$I_{in} = 21.6$ A (Average)

$E_o = 26.21$ V Regulated

$I_o = 0.875$ A

$E_{ff} = 71\%$

FREQUENCY = 1168 cps.

$R_L = 30\,\Omega$

Figure 17. Output Transformer Saturation, Waveforms

Figure 17b shows that the base drive of the normally-on transistor has been completely removed for the half cycle of transformer saturation.

Figure 17c confirms that when the base drive was removed the input current (i_{cc}) went to zero for the interval of transformer saturation. The uneven collector currents result from the purposely introduced unbalanced base to emitter resistance.

Figure 17d is the i_{cc}' current in the secondary winding feeding the filter and load voltage regulator. Despite the severe LIVC output transformer saturations occurring, the regulated output load voltage remains normal for the saturation conditions illustrated, and reliable continued LIVCR performance results. The average input current level rises somewhat because of reduced efficiency caused by the abnormal saturation.

2. Portion of Half Cycle of Saturation

Figure 17e illustrates a less severe case of transformer saturation occurring near the end of a half cycle of operation. Such a case is typical when minor unbalance conditions exist.

Figure 17f shows that the base drive of the inverter on-transistor is terminated immediately upon occurrence of core saturation.

Figure 17g shows that only a tolerable transitory increase of input current resulted with the occurrence of saturation.

Figure 17h is the secondary current feeding the filter and load voltage regulator. The regulated output voltage remains normal despite the LIVC purposely created transformer saturations conditions.

The inverter configurational protection technique which uses output transformer secondary winding current feedback enabled effective decoupling of the source from the short circuit loading occasioned by output transformer saturation for all cases tested. Nondestructive continued operation of the LIVC resulted despite the severe saturation creating unbalance conditions purposely introduced. Normal continued regulated output of the load voltage regulator was also typical of all cases tested. The waveform data and other experimental results obtained corroborate the output transformer secondary winding current feedback protection technique analysis presented in Section VI.

The use of main power transformer secondary winding current feedback has enabled the development of a LIVCR using a load-related current feedback driven inverter without danger of regenerative buildup of source/inverter current to destructive levels in case of abnormal transformer core magnetic saturation. It is significant that the method automatically provides this protection without im· ›sing any compromising penalties, principally: without need of using lossy, bulky, expensively manufactured magnetic cores or core materials. The method is reliable, time independent, foolproof, and has minimal additional circuitry requirements. As shown in Figure 14 the method is also compatible for use with the improved inverter crossover technique and when so combined it enables improved overall system reliability and efficiency in both low and high impedance source applications.

VIII. AN APPLICATION OF LOW INPUT VOLTAGE CONVERSION-REGULATION (LIVCR) FROM AN UNCONVENTIONAL SECONDARY ELECTROCHEMICAL SOURCE

The improved crossover technique presented in Section V, and the automatic protection in the event of output transformer saturation technique presented in Section VI, have been incorporated into a LIVCR model designed to operate from a rechargeable sealed silver-zinc single cell battery storage system. A partially block and partially schematic diagram of the LIVCR was given in Figure 14. The GSFC in-house developed and built LIVCR unit and a prototype battery (Yardney A_gZ_n HR-150(s)-1) being used in this application are shown in Figure 18.

The internal construction of the LIVCR is shown in Figure 19.

The LIVCR consists of a low input voltage inverter section which efficiently inverts and voltage transforms the low battery voltage to a higher square wave AC voltage. The square wave AC voltage is then rectified and filtered completing the low battery voltage step-up conversion to the higher usable DC output voltage values. In the low input voltage converter (LIVC) section, the lowest usable battery voltage under discharge conditions is stepped-up to the minimal input voltage required for proper operation of the follow-on load voltage regulator.

The overload protected, high efficiency, follow-on load voltage regulator used is based largely upon a Honeywell Inc. design (reference 7). The load voltage regulator section maintains the output load voltage at 28 volts ±2% for LIVC stepped up battery voltages in the range of 1.4 to 2.0 volts.

A constant value of L was used in this LIVC design. This results in inverter frequency variation with loading (as shown in Figure 8b), but since this particular application presents a nominally constant power load to the inverter the simpler constant L implementation is used.

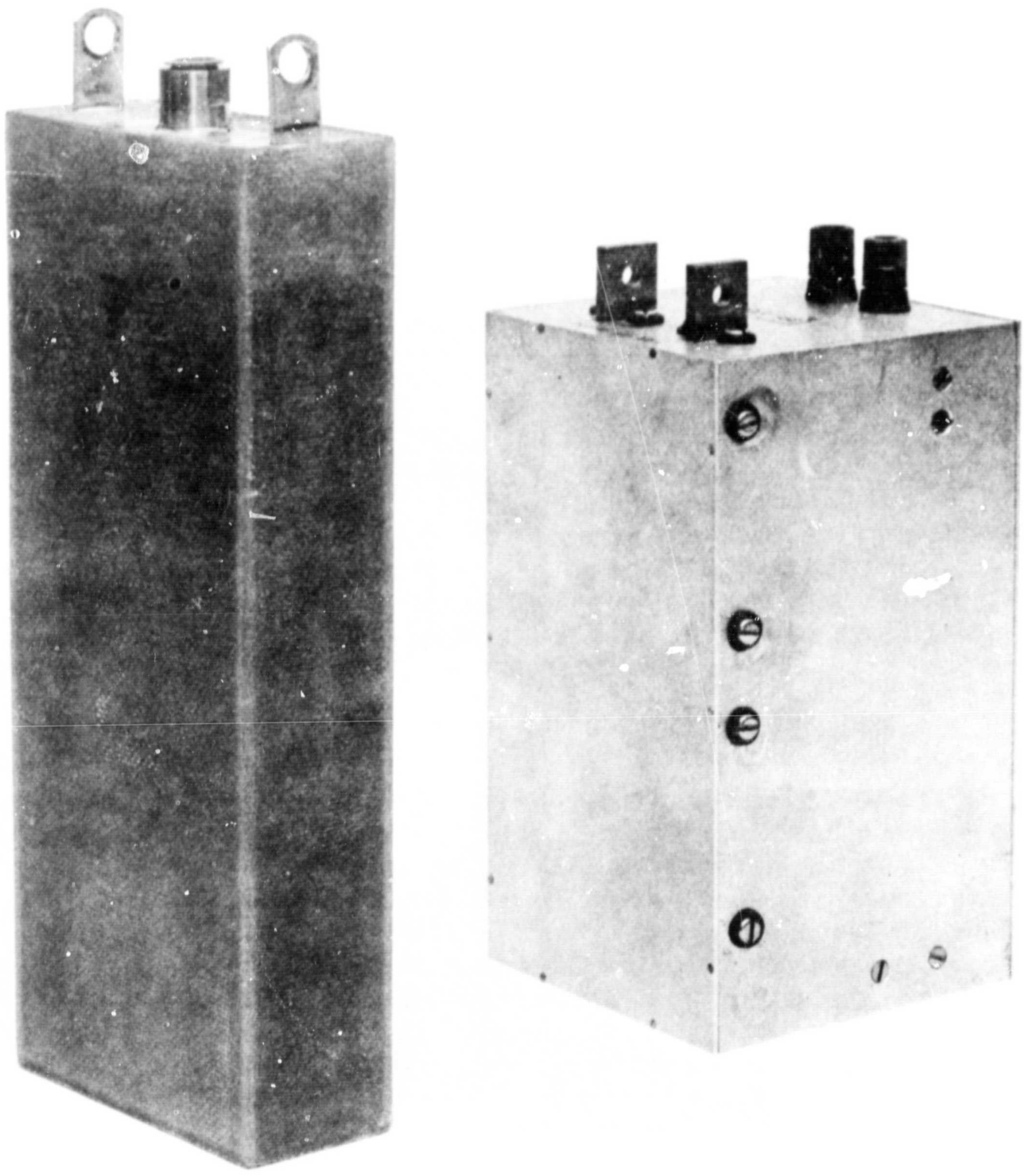


Figure 18. GSFC In-House Developed Low Input Voltage Converter-Regulator, and Prototype Silver Zinc (A_gZ_n) Battery (Yardney HR-150 (S)-1)

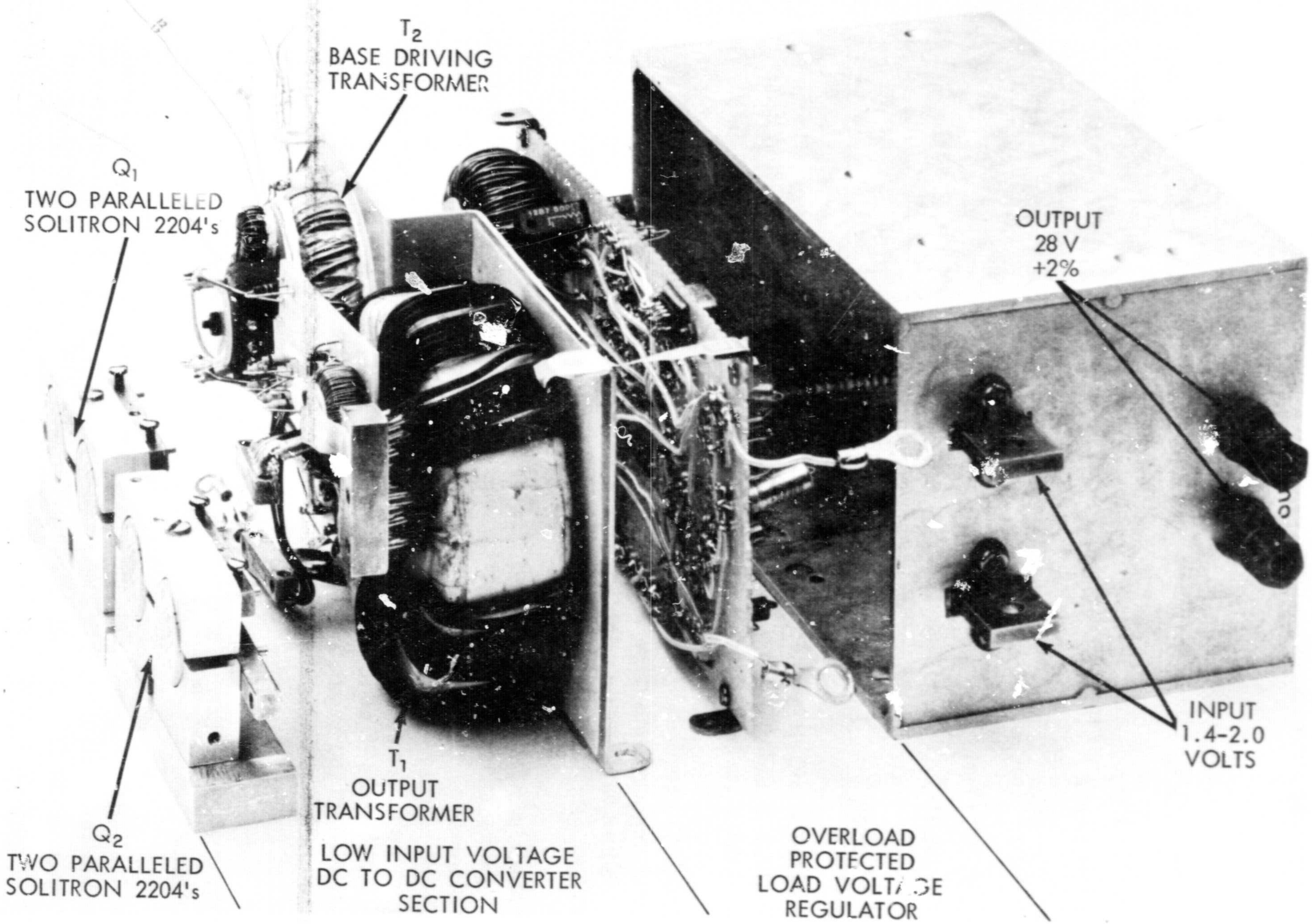


Figure 19. Preassembled GSFC In-House Developed Low Input Voltage Converter-Regulator

Some typical performance data obtained is given in Table 1.

Table 1

Input Voltage (V_{in})	1.45 Volts	1.75 Volts	1.98 Volts
Input Current (I_{in})	41.3 Amperes	35.9 Amperes	33.2 Amperes
Power Input (P_{in})	60 Watts	62.8 Watts	65.7 Watts
Output Current (I_{out})	1.65 Amperes	1.68 Amperes	1.69 Amperes
Output Voltage (V_{out})	27.72 Volts	28.15 Volts	28.43 Volts
Output Power (P_{out})	45.7 Watts	47.3 Watts	48.1 Watts
Overall Efficiency	76.1%	75.3%	73.3%

The LIVCR efficiency has been optimized at the approximately 1.45 V plateau voltage of the prototype silver-zinc battery used since operation is at this voltage for a large part of the battery discharge.

Figure 20 shows the battery, and the LIVCR, in an automatic cycling laboratory test arrangement designed to test the lifetime and cycle life of the battery, as well as to determine the lifetime and cyclic performance characteristics of the LIVCR.

Typical battery data (reference 8) obtained for a 50 percent depth of discharge, 24 hour charge-discharge regime is shown in Figure 21a. A 25 percent depth of discharge, 12 hour charge-discharge regime is shown in Figure 21b. As shown the battery is charged using a method in which constant current is supplied to the battery which then tapers off sharply as the battery voltage limit of 1.99 ±0.010 volts is reached. The battery discharges into the LIVCR which is fixed-loaded on the regulated output voltage side, thus presenting a constant power demand condition to the battery.

IX. RTG PRIMARY SOURCE COMBINED WITH A SECONDARY BATTERY SOURCE

Typical mission power demand profiles containing both average and peak power demands as illustrated in Figure 22 are postulated to demonstrate the system's use of the LIVCR approaches previously developed in this report.

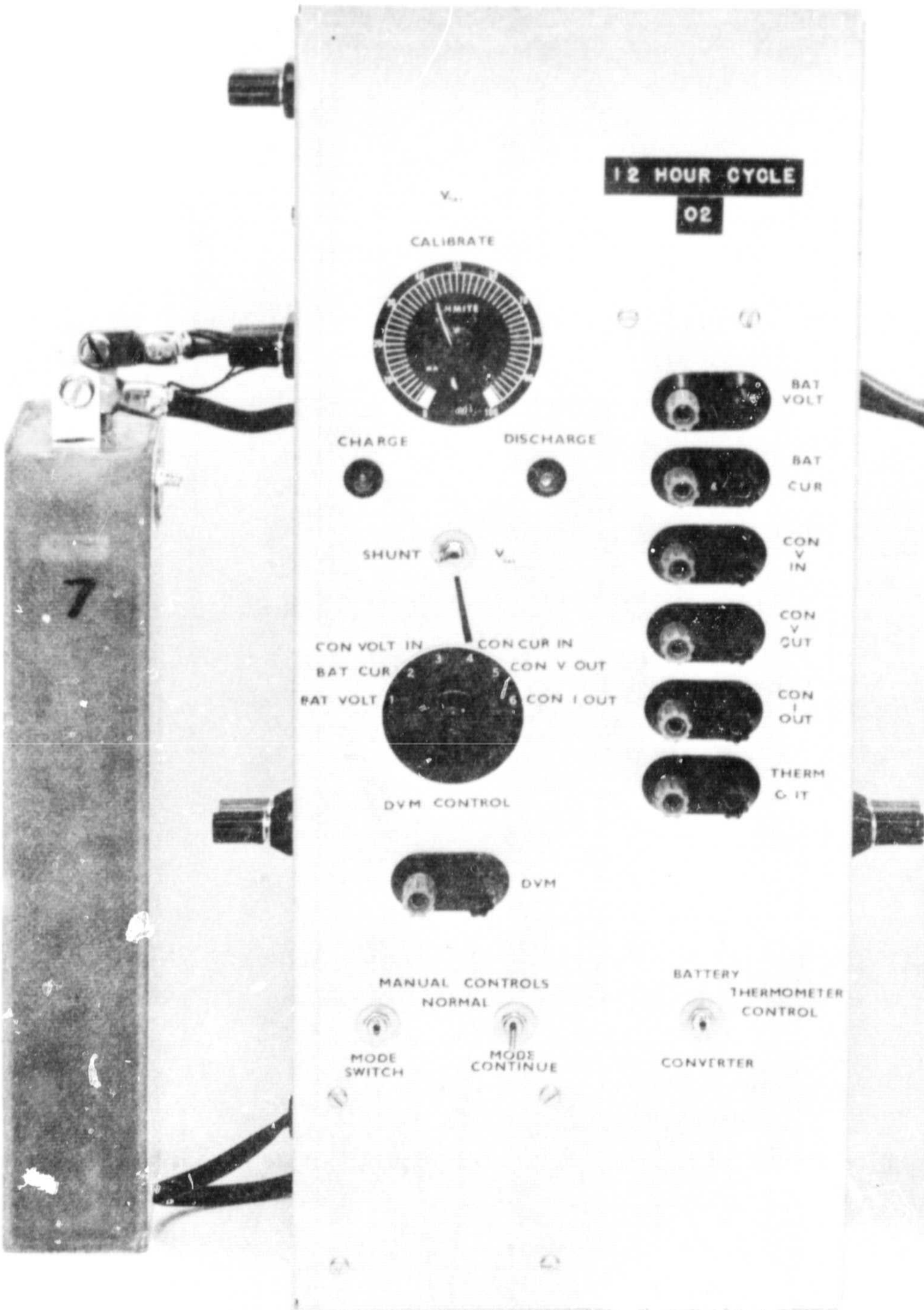


Figure 20. Automatic Battery Charge/Discharge Cycler, Laboratory Lifetest Arrangement

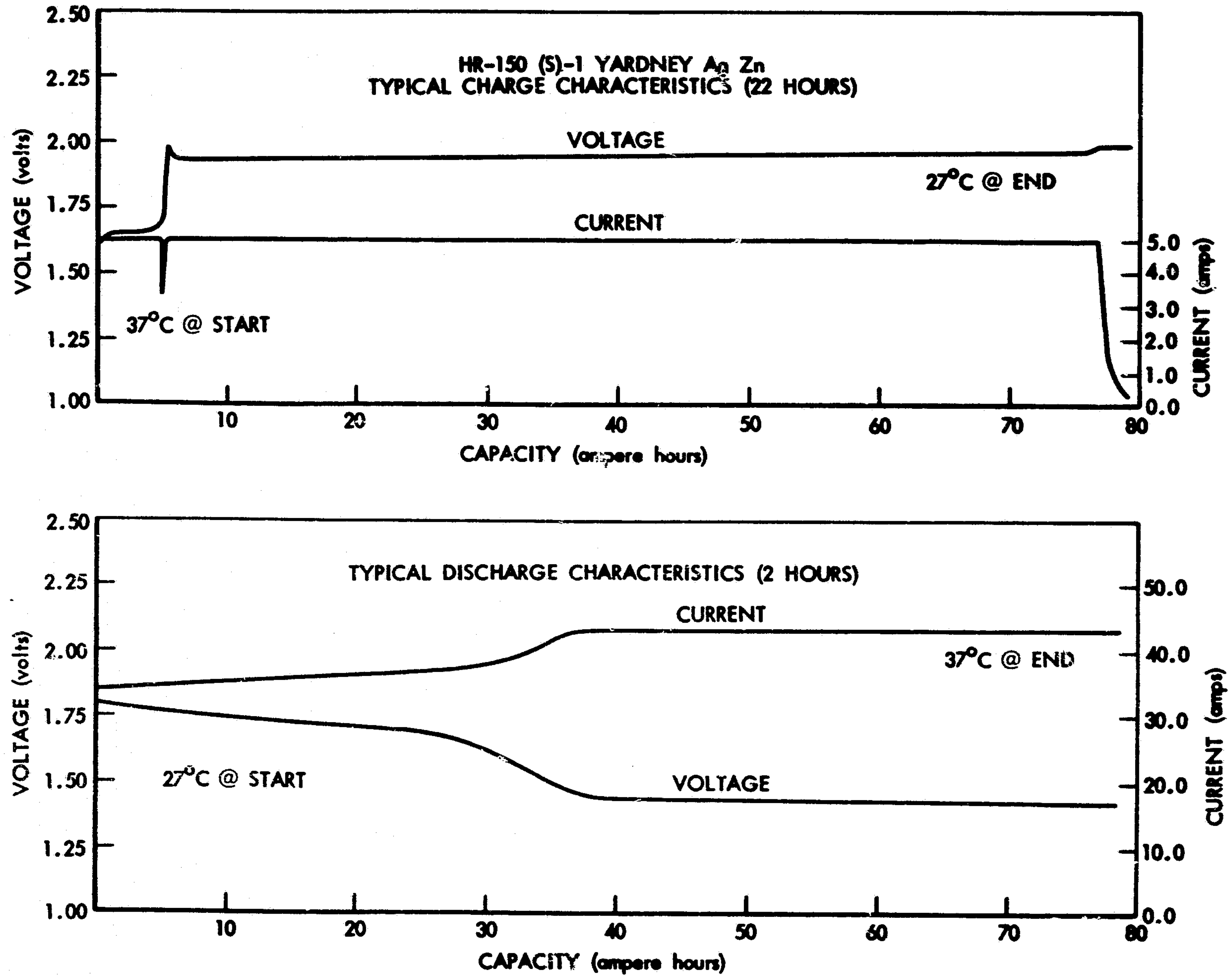


Figure 21a. Single-Cell Silver-Zinc Battery, 24 Hour Charge-Discharge Regime

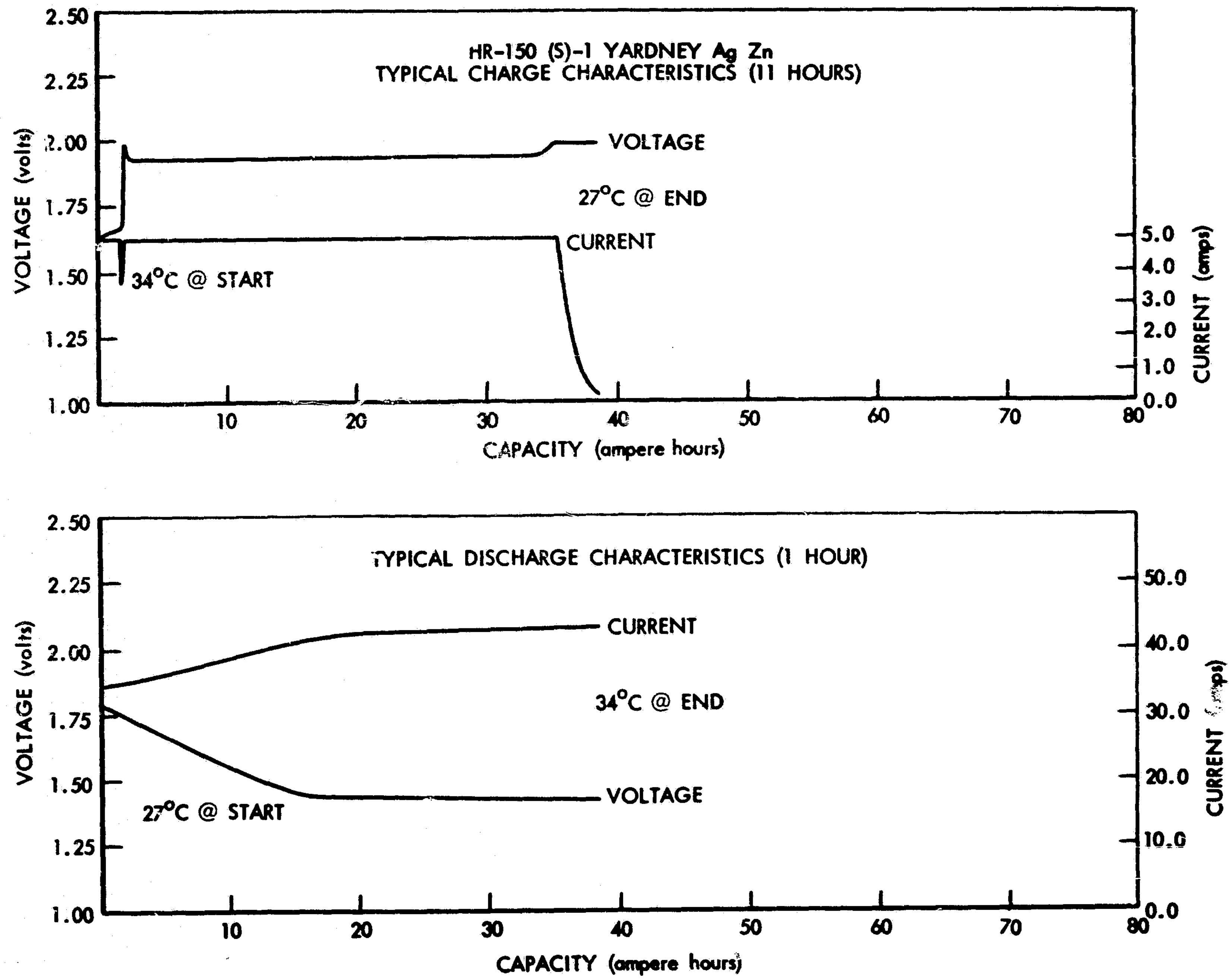


Figure 21b. Single-Cell Silver-Zinc Battery, 12 Hour Charge-Discharge Regime

The power supply configuration block diagram of Figure 23 is appropriate to satisfy the assumed power demand profiles and will be considered for use in this example. In this one of many possible systems capable of supplying the average and peak power requirements, it is assumed that when the output load power demand exceeds the primary source (RTG) limited output power capability the secondary (battery) source will supply the difference between the output load demand and the primary source output power capability. Also when the output load demand is less than the primary source capability, the primary source alone will supply the output load demand. The primary source will additionally supply power for charging the battery storage system.

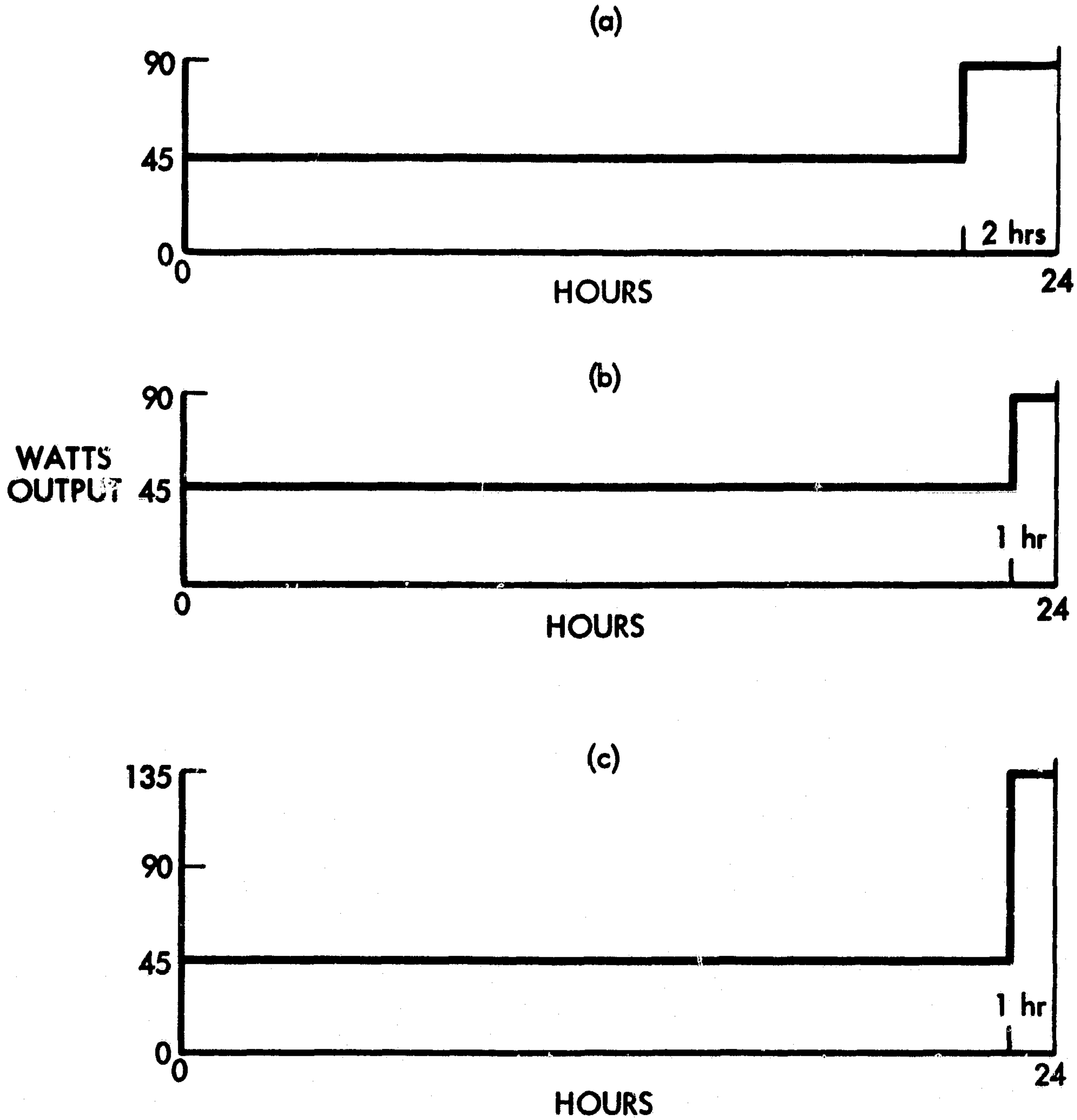


Figure 22. Typical Power Demand Profile

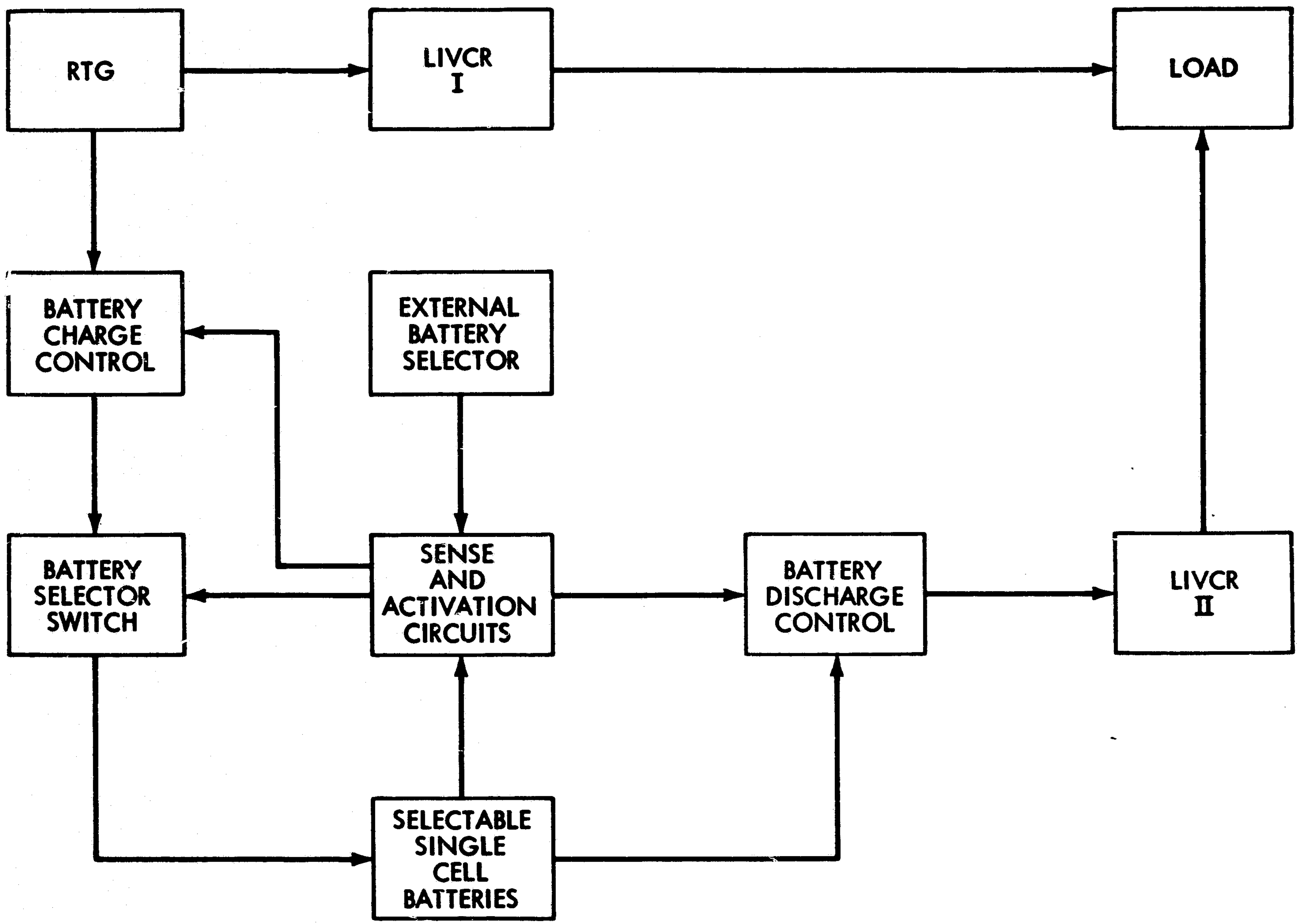


Figure 23. RTG (Primary Source) and Battery (Secondary Source) Power Supply System, Block Diagram

In the system shown in Figure 23 the power path through LIVCR I provides the average output power demand requirements. The LIVCR I block operating from the power-limited RTG source can be reliably and efficiently implemented using the output transformer primary winding current feedback type inverter which is briefly discussed in Section III A. of this report and which is more fully described in reference 1. Also highly advantageous in the inverter electronic implementation is the use of the improved crossover technique described in Section V. The power path through the LIVCR II block operating from the low impedance battery source supplies the output power demands in excess of the RTG limited output power capability. The use of output power transformer secondary winding current feedback and the improved crossover technique in the inverter section as described in Sections VI and V respectively, are highly advantageous in the electronic implementation of this block. From a systems viewpoint, the inherent LIVCR I output voltage fall-off when the RTG peak output power capability is exceeded (reference 1), enables LIVCR II to automatically and efficiently supply the load power demands in excess of the RTG peak output power capability. The LIVCR II regulated output voltage is set slightly lower than the LIVCR I regulated output voltage at RTG peak power output and minimum battery discharge with maximum RTG utilization is achieved.

For deep space probe applications battery activation may be delayed until greater distances require higher transmitter power levels and the storage system lifetime can be extended when necessary by switching out the expended battery and by switching in remotely activated redundant batteries (reference 8). The single cell storage system makes this practical as well as reliable because the complexity of cell switching, charge/discharge control, and battery condition sensing circuitry is much reduced when compared with multiple battery storage systems.

X. CONCLUSIONS

1. The use of redundant, remotely activated, single cell batteries, combined with low input DC voltage conversion-regulation to higher DC voltage directly usable at the load, can improve the reliability of storage systems needed for long lifetime, high peak power requirement, future mission applications.

2. The load-related optimum full-cycle base current control of the improved inverter crossover technique, is effective in essentially eliminating the deleterious efficiency/reliability effects of transistor storage time created overlap, by reducing peak voltage/power stresses during the inverter crossover interval. Improvement in the efficiency/

reliability of the voltage stepup inverters/converters as well as of the primary and secondary power sources themselves has been achieved.

3. The inverter configurational protection technique which uses output transformer secondary winding current feedback enables effective decoupling of the source from the short circuit loading occasioned by output transformer saturation. Nondestructive continued operation of the LIVCR is possible despite severe saturation-creating unbalance conditions. Normal continued regulated output of the load voltage regulator resulted in all cases of transformer saturation tested.

4. The particularized basic LIVCR design approaches developed to operate from either high or low impedance sources have successfully enabled sources with different output characteristics to be used singly or to be functionally combined in a manner favorable to each source subsystem as well as the overall power supply system. The specialized designs developed have resulted in improvement in overall efficiency, performance, and reliability.

XI. REFERENCES

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