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THESIS

SWITCHING MODE POWER SUPPLIES,

BY

David W. Beard

Sept 80

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REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1. REPORT NUMBER	2. GOVT ACCESSION NO. AD-A093 296	3. RECIPIENT'S CATALOG NUMBER
4. TITLE (and Subtitle) Switching Mode Power Supplies		5. TYPE OF REPORT & PERIOD COVERED Master's Thesis; Sept., 1980
		6. PERFORMING ORG. REPORT NUMBER
7. AUTHOR(s) David W. Beard		8. CONTRACT OR GRANT NUMBER(s)
9. PERFORMING ORGANIZATION NAME AND ADDRESS Naval Postgraduate School Monterey, California 93940		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS
11. CONTROLLING OFFICE NAME AND ADDRESS Naval Postgraduate School Monterey, California 93940		12. REPORT DATE Sept., 1980
		13. NUMBER OF PAGES 94
14. MONITORING AGENCY NAME & ADDRESS(if different from Controlling Office) Naval Postgraduate School Monterey, California 93940		15. SECURITY CLASS. (of this report) Unclassified
		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited.

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

18. SUPPLEMENTARY NOTES

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

Power Supply, Voltage Regulator

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

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Switching Mode Power Supplies

by

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Submitted in partial fulfillment of
the requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

NAVAL POSTGRADUATE SCHOOL
September 1980

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ABSTRACT

The subject of switching mode power supplies was examined. A comparison between linear regulators and switching mode power supplies was made to show the options available for the various types of convertors. Two switching mode power supplies were constructed and tested. The operating efficiency of both systems was found to be more than eighty percent over the specified input voltage and load current conditions. The switching mode power supply circuits required additional components in the design due to the complex pulse width regulation, pulse amplification, and self start circuit required for the system operation. The switching mode power supply is an inherent source of radio frequency interference, thus the complexity of the problem was examined with several methods presented that minimized the RFI of the system.

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I. INTRODUCTION

Since the invention of the electron tube there has been a need to supply a source of power to operate electronic circuits. As the complexity of electronic circuits increased, the need for a better regulated, higher power output supply was evident. It was in this era that vaccum tube regulators were developed. Thyratrons handled the job of AC line preregulation with triodes and pentodes used as series regulators. These power supplies, with a large transformer and a filament power supply, were bulky and inefficient.

Semiconductor components in the 1960's generated requirements for better regulation in the equipment in which they were used, and also provided the means of accomplishing that regulation at low cost. These semiconductor supplies used precision zener diodes and power transistors to obtain highly reliable power supplies that represented a quantum leap over electron tube technology. In the late 1960's, equipment continued to shrink in size and integrated circuit voltage regulators were developed that presented a smaller package size and incorporated protection circuits on the chip. These linear regulator systems were very compact, easy to design and implement. For high power systems a great deal of consideration must be given to heat dissipation, either by large heat sinks or by cooling fans.

Switching mode power supplies were an outgrowth of DC to DC electro-mechanical systems such as the vibrating reed switch. These switching mode power supplies utilizing integrated circuit oscillators and control

circuitry were developed in the middle 1970's. This technology produced power supplies that were much more efficient than ny other type of regulated power supply.

The first section of this thesis deals with the types of linear power supplies; the use of low power on card regulators, and some practical power supply circuits used in modern equipment, with efficiency calculations.

The succeeding sections present information on the use of switching mode power supplies; the theory of operation; the types of supplies used; several designs of practical circuits; and the advantages and disadvantages of the switching mode power supply.

II. LINEAR POWER SUPPLIES

The simplest form of a modern power supply consists of a transformer, rectifier, and a filter section. This unregulated power supply suffers from three major limitations. Any change in the input line voltage results in a similar change of the output voltage. This variation is defined as line regulation or as static regulation. The second type of variation is due to the change of the load resistance or load current. This variation is defined as load regulation or dynamic regulation. The third variation is a result of a change in the operating temperature. The ouput voltage drifts as the components within the power supply warm up. This variation is defined as thermal regulation. The drift in each component is measured as the temperature coefficient in units of parts per million per °C. This factor is usually larger in semiconductor devices than passive components.

For many applications any large change in supply voltage is unacceptable for system operation. Regulated power supplies were developed to maximize the line regulation, load regulation and thermal regulation. That is to minimize the changes in the output voltage due to these variables.

A. ZENER REGULATORS

The most basic regulator circuit developed was the Zener diode circuit. The diode operates in the avalanche or breakdown region thus the device acts as an element whose resistance is inversely proportional to the current flowing through the device. The Zener diode is shunting the load, thus a portion of the current flows to the load resistance.

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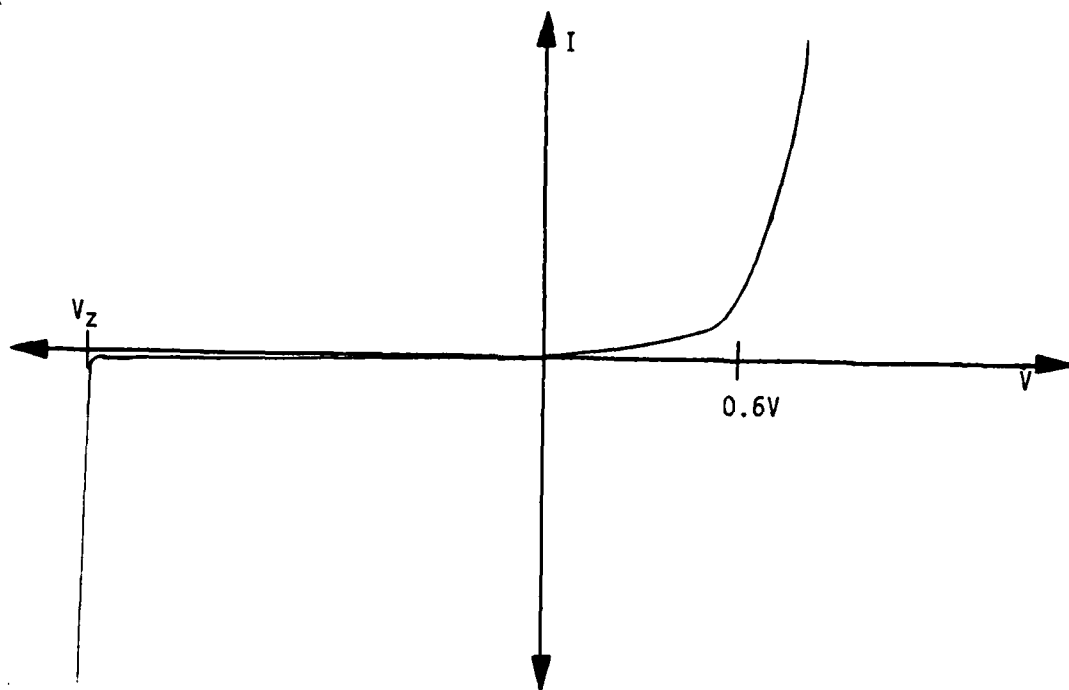
Figure 1. Zener Diode I-V Characteristics

The I-V characteristics of a typical Zener diode is shown in Fig. 1. The values of the zener voltage can be from 1.8 volts to several hundred volts with a power dissipation of 250 mW up to 50 W or more.
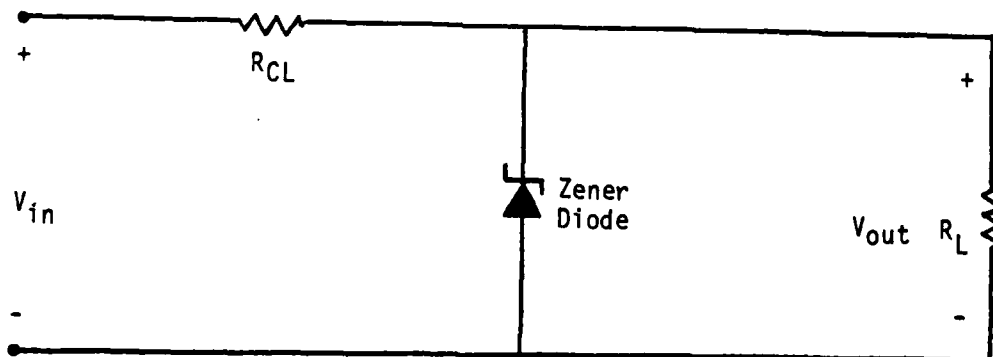
Figure 2. Zener Voltage Regulator

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To design a zener regulator as in Fig. 2, the characteristics of the load resistance must be known. The maximum and minimum load currents must be known.

Max load conditions:

$$I_{L\ max} \simeq I_{in} \qquad (1)$$

$$I_z \simeq 0 \qquad (2)$$

Minimum load conditions:

$$I_{in} = I_z + I_{L\ min} \qquad (3)$$

The current limiting resistor, R_{CL} must be selected to limit the current in the zener diode. This resistor must provide adequate current to the load for minimum input voltages.

$$V_{in\ min} = V_z + I_{in}R_{CL} \qquad (4)$$

The power rating of zener diode is determined by finding the maximum current through the device, during maximum input voltage conditions and minimum load conditions. [1]

$$I_{z\ max} = \frac{V_{in\ max}}{R_{CL}} - I_{L\ min} \qquad (5)$$

$$P_z = I_{z\ max} R_z \qquad (6)$$

Thus for a load that requires 5.1 V at 200 ma, switching from no load to full load, and a supply with a 9 VDC ± 2 V, the component values are then:

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$$I_{L\ max} = 200\ mA \qquad\qquad (7)$$

$$I_{L\ min} = 0 \qquad\qquad (8)$$

$$V_{in\ max} = 11\ V \qquad\qquad (9)$$

$$V_{in\ min} = 7\ V \qquad\qquad (10)$$

$$V_z = 5.1\ V \qquad\qquad (11)$$

$$R_{CL} = \frac{V_{in\ min} - V_z}{I_{in}} \qquad\qquad (12)$$

$$R_{CL} = \frac{((9-2) - 5.1)\ V}{0.2\ A} \qquad\qquad (13)$$

$$R_{CL} = 9.5\ \Omega$$

$$I_{z\ max} = \frac{(9+2)\ V}{9.5\ \Omega} - 0 \qquad\qquad (14)$$

$$I_{z\ max} = 1.16\ A$$

$$P_z = (5.1\ V)\ (1.16\ A) \qquad\qquad (15)$$

$$P_z = 5.9\ W$$

$$P_R = \frac{((9+2) - 5.1)^2\ V^2}{9.5\ \Omega} \qquad\qquad (16)$$

$$P_R = 3.7\ W$$

Thus for a 200 mA maximum load, a 10 W zener diode is selected. A 5 W current limit resistor was required. The efficiency for the regulator under maximum load conditions is given by:

$$\eta = \frac{P_L}{V_{in} \cdot I_{in}} \qquad\qquad (17)$$

Where

$$I_{in} = \frac{V_{in} - V_z}{R_{CL}} \qquad\qquad (18)$$

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$$\eta = \frac{1.02 \text{ W}}{9 \text{ V} \cdot 0.41 \text{ A}} \qquad\qquad (19)$$

$$\eta = \frac{1.02 \text{ W}}{3.69 \text{ W}} = 27.6\%$$

The dynamic impedance of the Motorola 1N3996R, 5.1 V, 10 W zener

diode is 0.7 ohms. Then a zener with a current change of 410 mA exhi-

bits a voltage change of $(0.41 \text{ A})(0.7 \mu) = 287$ mV. This load regulation

is not adequate for many circuits. For worst case conditions (i.e. *maximum*

load with minimum input voltage compared to minimum load with maximum in-

put voltage) results in an output voltage difference of 810 mV.

For a system whose raw DC voltage varies to a greater extent, the

system performance is dependent on the worst case input voltage.

Table I. Operating Conditions and Component Values for
Typical Zener Regulators.

Input Voltage	9 V ± 2 V	9 V ± 3 VDC
Minimum Input Voltage	7 V	6 V
R_{CL}	9.5 Ω	4.5 Ω
$I_{z\ max}$	1.16 A	2.67 A
P_z	5.9 W	13.6 W
η efficiency (max load)	27.6%	13%
Output Regulation	±405 mV	±933 mV
Output Voltage	5.1 V ± 8%	5.1 V ± 18%

Thus the maximum system performance, and minimum system cost was

achieved *for equipment whose input voltage variations are smaller* as

shown in Table I.

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For most applications, the zener regulator is unacceptable due to the inefficiency of the device, especially with less than the maximum load. In low power applications where a stable reference voltage is required, the zener is an attractive solution.

B. TRANSISTOR SHUNT REGULATORS

The transistor shunt regulator (Fig. 3) operates in much the same manner as the zener regulator. The transistor acts as the shunt element; the base current is supplied by a zener diode. The output voltage of this regulator is the base to emitter junction voltage (0.6 V for silicon transistors) plus the zener voltage. The transistor is used due to the larger current values and power dissipation of this device.
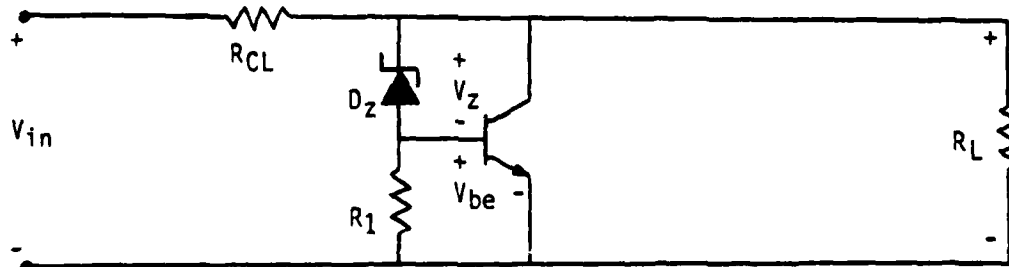
Figure 3. Transistor Shunt Regulator

The transistor shunt regulator, like the zener regulator, draws a total current from the supply that is proportional to the unregulated input voltage. The regulating element divides the current between itself and the load. For most systems the efficiency of this type of regulator is unacceptable, however for specialized systems this system may prove to be the most cost effective solution.

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C. SERIES PASS REGULATORS

The series pass regulator was developed to provide better regulation than the other linear regulators in use. The regulator is improved by the high gain negative feedback amplifier. Most linear voltage regulator systems are of this type. Until the development of high-current, low forward-drop transistors, usable as high-frequency single-pole switches, it was the main and practically the only voltage-regulating technique up to power levels of 1,000 W. [2]

The series pass regulator is usually comprised of a stable reference source, a differential amplifier for gain in the feedback loop, and a series pass element, usually a power transistor. (Fig. 4)
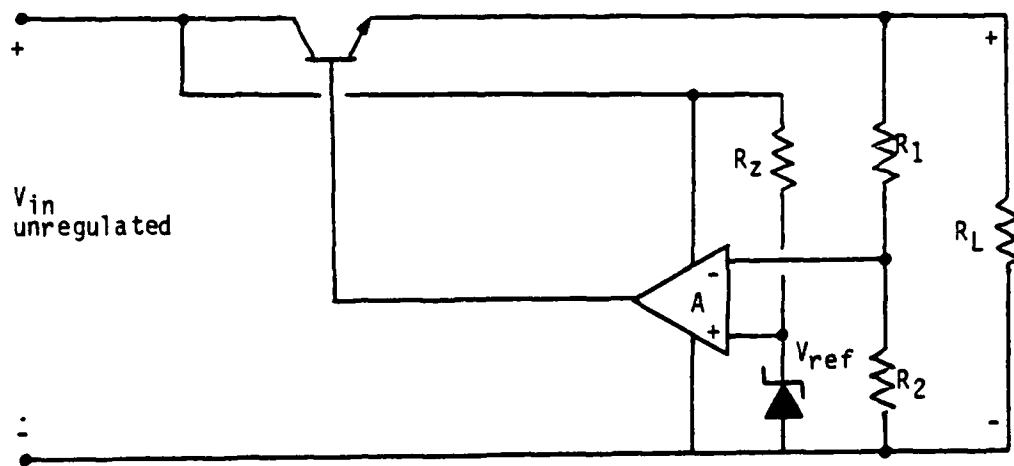
Figure 4. Series Pass Regulator With Feedback

The output is kept at a nearly constant voltage by using the series-pass element as a variable resistor. As the input voltage varies, the effective resistance changes to minimize any voltage change at the load. The transistor base current is supplied by an amplifier. A voltage divider supplies a fraction of the output voltage to the inverted input

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of the differential amplifier. The noninverted input is connected to a stable reference voltage. The differential amplifier produces a voltage that is proportional to the difference between the reference supply and the output sample. This amplified difference must be fed to a current gain stage and a DC level shifter to supply the base current to the series pass element. This negative feedback loop drives the error voltage to nearly zero, thus forcing the sampled output voltage to be essentially equal to the reference voltage. The feedback gain allows a very accurate output voltage regulation. [3]

The series pass element must be selected to handle the maximum current of the load. The input current is equal to the current required by the load, neglecting a small bias current required for the feedback loop and pass element. Thus the series pass regulator is more efficient than the shunt regulator under conditions of less than maximum load.

For reliable regulation, the system must be designed with a minimum voltage drop across the series element that is in the linear region of operation. For typical silicon power transistors, a minimum of 2.0 V across the series element is usually adequate. That is, the minimum input voltage must be greater than the output voltage plus the 2 volt minimum collector to emitter voltage shown in Fig. 5.
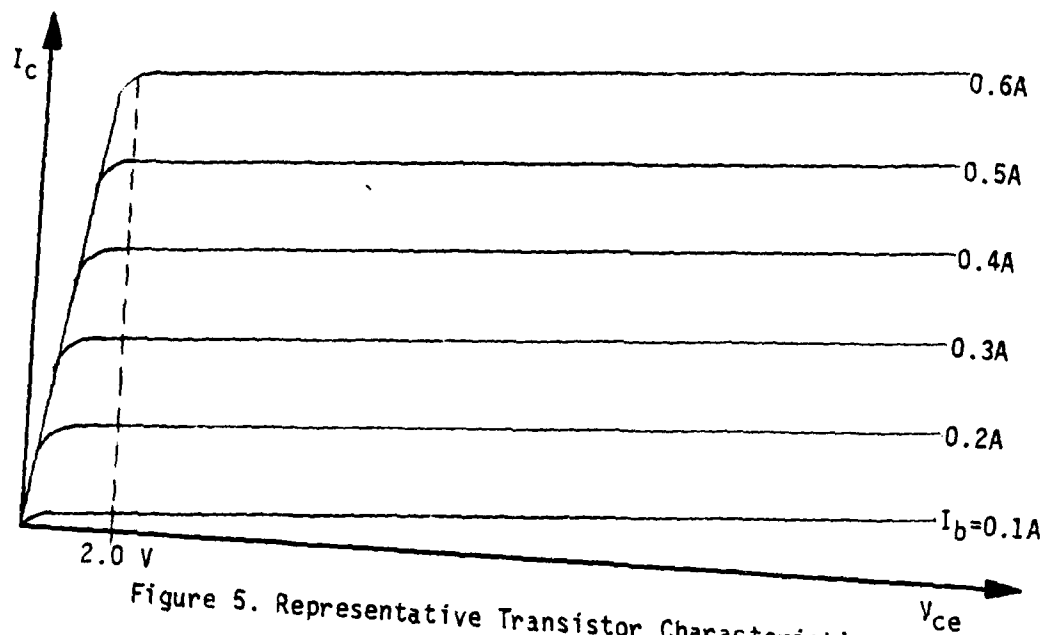
Figure 5. Representative Transistor Characteristics

The series pass regulator efficiency is dependent on the input voltage. The maximum efficiency occurs at the minimum input voltage, conversely the minimum efficiency occurs at the maximum input voltage. For an input voltage of V_{in} with a tolerance of T percent:

$$V_{in} (1 - 0.01T) \geq V_{out} + 2 \qquad (20)$$

Minimum Efficiency:

$$n = \frac{V_{out}}{V_{in}} = \frac{-V_{out}}{\left(\frac{1 + 0.01T}{1 - 0.01T} \right)(V_{out} + 2)} \qquad (21)$$

$$\frac{V_{out}}{V_{in\ max}} = \left(\frac{1 - 0.01T}{1 + 0.01T} \right) \left(\frac{V_{out}}{V_{out} + 2} \right) \qquad (22)$$

Maximum Efficiency:

$$n = \frac{V_{out}}{V_{in\ min}} = \frac{V_{out}}{V_{out} + 2} \qquad (23)$$

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The minimum input voltage is selected such that $V_{in\ min} = V_{out} + 2$. Thus the efficiency is maximum for systems whose input voltage has a small tolerance and whose output voltage is large.

For a system that requires 5.0 V at 200 mA with a raw DC input of 9 VDC ± 2 V, the minimum efficiency is found by solving for Eq. (22). A graph of the average efficiency versus output voltage for various input voltage tolerances is shown in Fig. 6.

$$\eta_{avg} = \frac{V_{out}}{V_{in}} = \frac{5}{9} \text{ or } 55.6\% \qquad (24)$$

Figure 6. Efficiency vs. Output Voltage
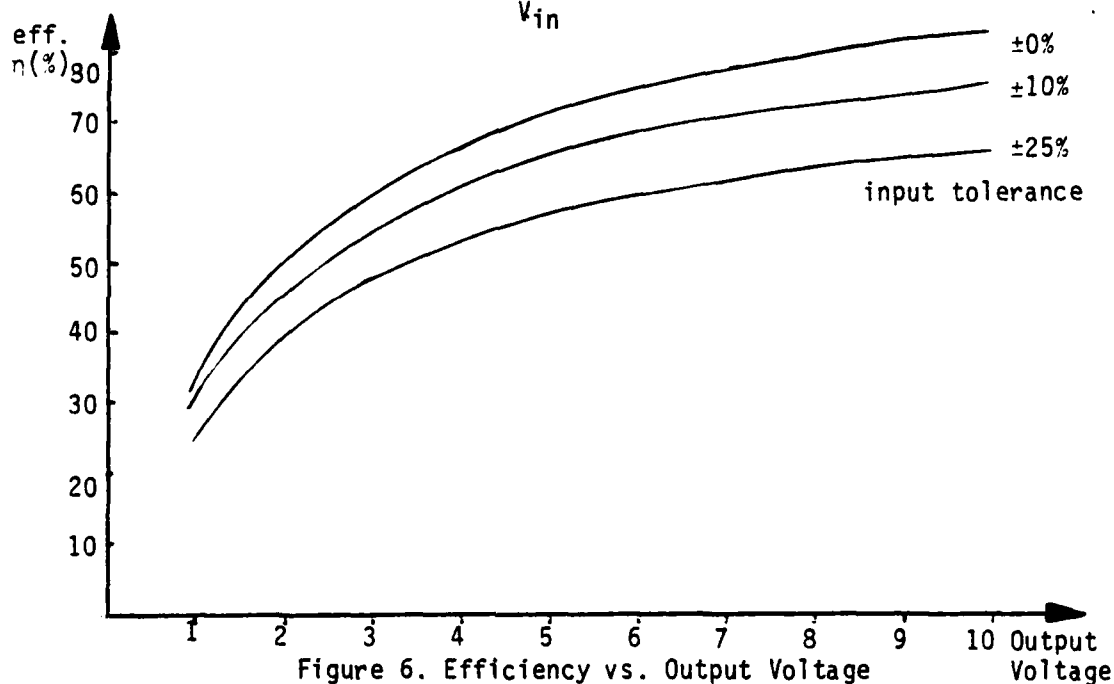
The average efficiency is based on the average input voltage selected such that the input voltage at the minimum is two volts greater that the output voltage.

The output voltage variation of the series pass regulator is due to the variation of the input voltage, changes in the load current,

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and thermal drift. The regulation can be calculated once the device parameters are known. The regulation was calculated by Eqs. (25) through (37) with reference to Fig. 7.
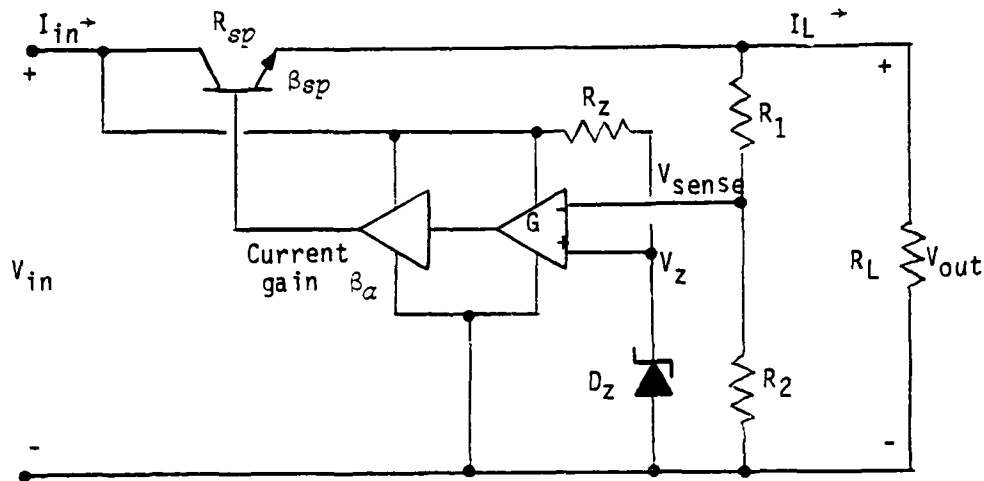
Figure 7. Series Pass Regulator Used for Equations

The parameters to be used are:

R_{sp} = Dynamic Impedance, slope $\Delta V_C/\Delta I_C$ of the V_C-I_C curve at the operating point of the series pass transistor.

β_α = Current gain of the driver circuit.

g_m = Transconductance of the differential amplifier.

β_{sp} = Current gain of the series pass transistor.

A = Open-loop voltage gain of the feedback loop.

V_z = Reference Voltage.

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$$V_{out} = \frac{V_{in} R_L}{R_L + R_{sp}} - A\left(\frac{V_{out} R_2}{R_1 + R_2} - V_z\right) \tag{25}$$

$$V_{out} = V_{in}\left(\frac{R_L}{R_L + R_{sp}}\right) + V_z A \tag{26}$$

For

$$R_L < R_{sp}$$

and

$$\left(\frac{R_2}{R_1 + R_2}\right) A \gg 1 \tag{27}$$

Then

$$V_{out} = \frac{V_{in} R_L (R_1 + R_2)}{R_{sp} A R_2} + V_z \frac{(R_1 + R_2)}{R_2} \tag{28}$$

$$V_{out} \simeq \frac{V_z (R_1 + R_2)}{R_2} \tag{29}$$

Line regulation is given by:

$$\frac{\Delta V_{out}}{\Delta V_{in}} = \frac{R_L}{R_{sp}} \cdot \frac{(R_1 + R_2)}{A R_2} \tag{30}$$

where

$$A = R_L g_m \beta_a \beta_{sp} \tag{31}$$

Load regulation is given by:

$$\Delta I_{out} = \Delta V_{sense} g_m \beta_a \beta_{sp} \tag{32}$$

where

$$V_{sense} = \frac{R_2}{R_1 + R_2} V_{out} \tag{33}$$

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and

$$\Delta V_{sense} = \left(\frac{R_2}{R_1 + R_2} \right) \Delta V_{out} \qquad (34)$$

since

$$\Delta I_{out} = \left(\frac{R_2}{R_1 + R_2} \right) \Delta V_{out} \; g_m \; \beta_a \; \beta_{sp} \qquad (35)$$

rearranging gives

$$\Delta V_{out} = \frac{R_1 + R_2}{R_2} \cdot \frac{\Delta I_{out}}{g_m \; \beta_a \; \beta_{sp}} \qquad (36)$$

substituting Eqs. (32) and (34) yields

$$\Delta V_{out} = \left(\frac{R_1 + R_2}{R_2} \right) \frac{R_L}{A} \; \Delta I_{out} \qquad (37)$$

The values of R_1 and R_2 are selected to supply an adequate current to the differential amplifier input. The ratio of R_1 to R_2 is determined by knowing the output voltage and selecting a suitable reference voltage source. Zero temperature coefficients for zener diodes can be obtained in the voltage ranges from approximately 4.9 V to 5.8 V. Thus to minimize the thermal drift of the power supply, a zener diode may be selected and reverse biased at the required current to achieve zero thermal drift due to the zener reference. This method of selecting the zener diode is only appropriate for power supplies whose output voltage is greater than this zero coefficient voltage range. $(V_z \leq V_{out})$ [4]

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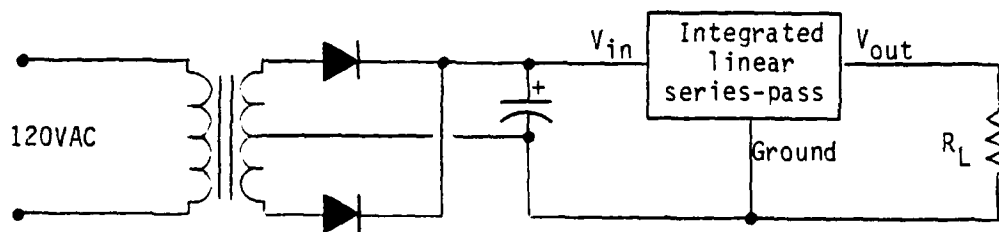
Figure 8. Power Transformer and Regulator System

For a power supply operated from the AC line, a transformer, rectifier and filter are used to supply an unregulated DC voltage to the regulator circuit. (Fig. 8) A transformer is selected with an adequate volt·ampere rating. The turns ratio is determined such that for low line conditions, the output voltage is at least 2 V greater than the required output voltage. The minimum input voltage occurs at the lowest line input, at the bottom of the ripple waveform. The minimum output voltage of the capacitive filter is due to the line variation, capacitor value and one diode drop for fullwave center tapped transformer (two diode drops for a full wave bridge). The voltage drop for typical rectifiers is approximately 1 V.

To find the proper capacitor value we use the equation:

$$\frac{I_{Load}}{C} = \frac{dV}{dt} \tag{38}$$

For a full wave rectifier the filter capacitor must supply the current to the load for one half the 60 H_z input period minus the conduction angle of the rectifiers as shown in Fig. 9. For low ripple values, Δt approaches 8.33 msec.

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$$C = \frac{I_{Load}}{\Delta V_{out}} \quad \Delta t$$

$$C = \frac{I_{Load}}{\Delta V_{out}} \quad (8.33 \times 10^{-3}) \text{ Farads}$$

or

$$C = \frac{I_{Load}}{\Delta V_{out}} \quad (8.33 \times 10^{3}) \text{ } \mu F \qquad (39)$$
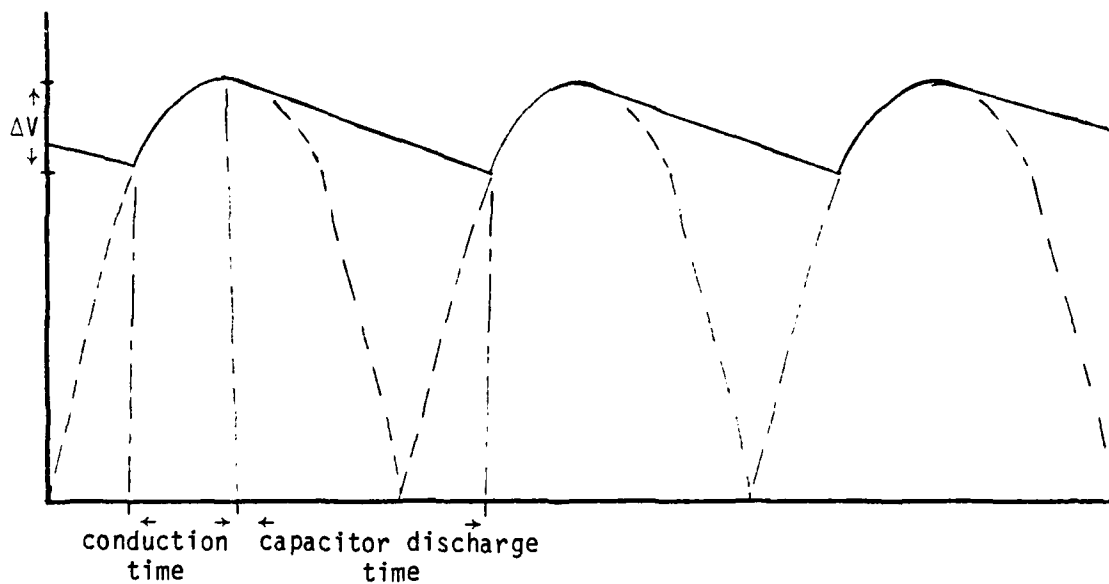
Figure 9. Output of Fullwave Rectifier And Capacitive Filter

The overall power supply efficiency must reflect the losses due to the input transformer, the losses in the rectifiers, and the losses in the regulator. The efficiency in the 60 Hz transformer, for ratings of 1 VA to 100 VA are typically 95% to 99%. The losses in a fullwave bridge rectifier was found to be near the assumption of 2 V. Thus the power loss in watts is equal to twice the current. The efficiency of the regulator has been found to be dependent on the input voltage.

For a 5 V, 2 A power supply with a 10% ripple on the capacitive filter output, operating from the AC line that varies from 90 V to 130 V, the overall system efficiency was found to be 39.5% at 115 V line voltage. Results are tabulated in Table II.

Table II. Losses in a Linear Regulator System, Experimental Results

Power Input to System	25.3 W
Transformer Power Loss	0.5 W
Rectifier Power Loss	4.1 W
Regulator Power Loss	10.7 W
Power Delivered to Load	10.0 W

Most modern equipment utilize integrated circuit voltage regulators. These voltage regulators can be the fixed output voltage type or externally adjustable voltage regulators. These regulators can be used for currents of up to three amperes. The devices are packaged in a variety of cases inlcuding the TO-3 package, TO-220 plastic package, TO-39 metal can, TO-5, TO-99, 8 pin DIP and a flat package. These units can be used alone or with an external power transistor for increased current capabilities. These devices often include protection

circuits such as short-circuit protection, current foldback or current limit, thermal protection, and overvoltage protection. [5]

Integrated circuit voltage regulators allow large systems to be powered by a distributed power supply. A large transformer and filter section serves the entire system. This raw DC or coarsely regulated DC is fed to the various subsections. Each subsection will have an integrated voltage regulator on the printed circuit board. Any voltage variations or low frequency noise on the power bus is removed by the final on-card regulator.

III. BASIC SWITCHING MODE POWER SUPPLIES

A. PRINCIPLES OF OPERATION

The switching mode power supply (Fig. 10) is basically a chopper cir-
cuit, a high frequency transformer or inductor, rectifier and an output
filter network.

Modern switching mode circuits evolved from primitive DC to DC con-
vertors. These circuits consisted of a mechanical switch that vibrated
at audio frequencies. This switch chopped the input current, thus acting
as a DC to square wave convertor. This square wave output was connected
to a transformer. The secondary of the transformer was then rectified
and filtered. These systems were often used as a DC to DC step-up con-
vertor in battery powered systems such as automobile radios built in the
1940's and 50's.

Integrated circuit technology has provided an integrated circuit os-
cillator and switching transistors as a replacement for the vibrating
reed switch. The switching frequency is most often between 20 KHz and
50 KHz. Ferrite materials are used for the transformer cores with fast
recovery Schottky rectifiers used on the secondary of the transformer.
Switching mode power supplies can be used to operate from the AC line.
These off-line systems employ a rectifier and filter to convert the 60 Hz
line frequency to DC. The DC to DC step-down convertor is used to supply
the required output voltage level.

Switchers are smaller and lighter in weight than comparable linear
supplies. The reduced size is due to the use of transformers and

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inductors at 20 KH$_z$ rather than the 60 Hz line frequency. The second savings in size is due to smaller heat sinks and the elimination of fans (or use of smaller fans) required to cool the system. This reduction is due to the increased efficiency of the switcher. Typical efficiencies are on the order of eighty percent.
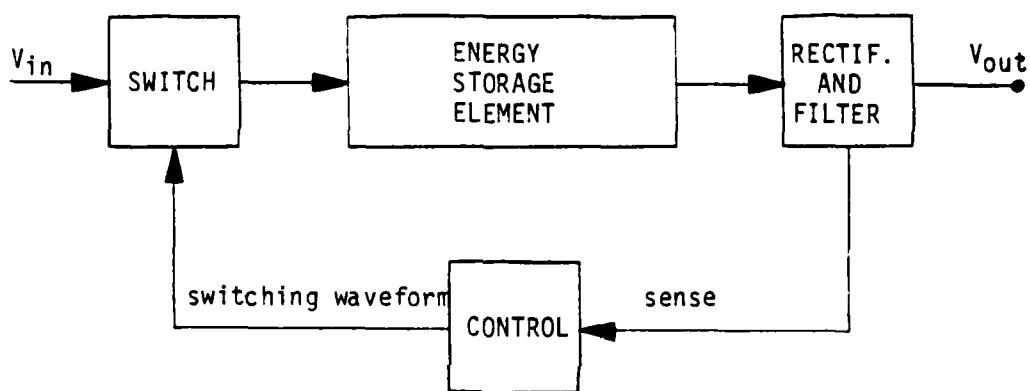
Figure 10. Basic Switching Mode Power Supply

B. ENERGY STORAGE IN INDUCTORS

One of the most important components in the switching mode power supply is the transformer or inductor. For switchers operating in the 20 KHz to 50 KHz range, the power transformer design and core selection are critical for high efficiency and low RFI operation. The transformer operates as an energy storage device. The input voltage is applied across the primary winding of the transformer when the switching transistor saturates. This voltage causes a current to start flowing through the primary winding such that:

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$$e = L \frac{di}{dt} \tag{40}$$

where   e = instantaneous voltage
                L = Inductance in Henries

This current builds linearly during the period that the transistor is saturated. This changing current produces a proportional changing flux:

$$e = -N \frac{\partial \phi}{\partial t} (10^{-8}) \tag{41}$$

where   N = turns in the winding
                ϕ = flux in maxwells

The changing flux induces a voltage across the windings of the secondary. The use of ferrite cores increases the coupling of the windings. The core losses are due to hysteresis resulting from the reversal of magnetic domains in the core and eddy current losses due to current flowing in the core induced by the changing flux. [6]

The magnetic flux in any cross section of the magnetic field is found from the expression:

$$\phi = \int B \cos\alpha \, dA \quad \text{in webers} \tag{42}$$

where α = angle between the direction of flux
                density and the normal at each point
                over the surface.

In high permeability cores, most of the magnetic flux is confined to the core. The flux is distributed nearly uniformly through the cross-sectional area of the core. The flux density is calculated by:

$$B = \frac{\phi}{A} \tag{43}$$

where   B = flux in density in gauss
          A = the cross section area cm^2
          ϕ = flux in webers   [7]

The flux density is limited by the core material.

$$B = \mu H$$

$$B = \mu \frac{4\pi \, Ni}{\ell} \tag{44}$$

where μ = Permeability of the core
          H = Magnetizing force in oersteds
          i = Instataneous current in amperes
          ℓ = Length of magnetic circuit in cm

The selection of the core material was based on the minimum core losses, heat dissipation, core volume, and cost. The core material must have a high permeability for operation at 20 KHz, maintaining a small package size.

One of the materials used in switcher inductive cores is a metal tape wound core. These cores consist of a metal allow tape (iron and nickel or silicon alloys) wound in an annular segment around a toroidal bobbin. The tape thickness used is ½ or 1 mil. These cores have a higher maximum usable flux density than ceramic ferrite cores. The most commonly used core is a ferrite core. The ferrite cores are ceramic ferromagnetic mixtures of iron oxide and zinc, nickel or manganese oxides. The powdered oxides are shaped and fired in a kiln. The ferrite cores exhibit lower losses and less cost than metal tape cores with the disadvantage of a slightly larger size.

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Ferrite cores come in a variety of shapes. The most common shapes shown in Fig. 11 have cross sections such as U, E, cup or torroids.

| U section | E section | Cup Core | Torroid Core |
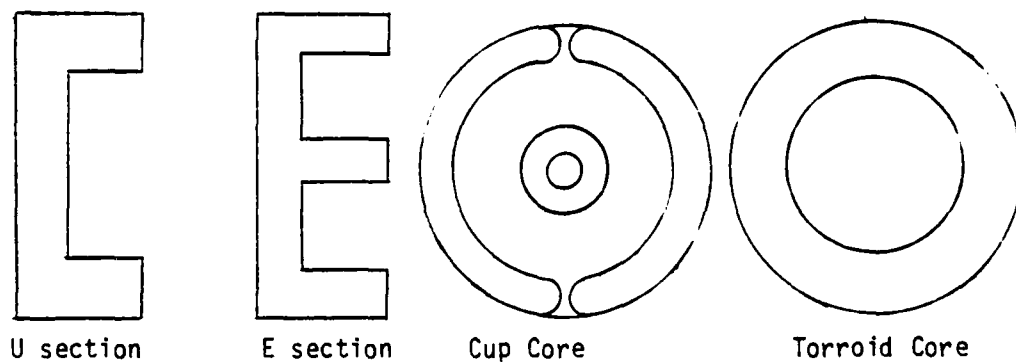
Figure 11. Core Geometry

The U, E, and cup cores can be stacked to achieve high power transformers. The E and cup cores come with bobbins making winding easier. In this work the core selection was based on core types available and the ease of winding. Torroid cores were selected due to ready availability and necessity of hand winding the limited number of devices constructed.

Complete design calculations of the inductive elements of a switching mode power supply are contained in Section VIII. F.

IV. SWITCHING MODE POWER SUPPLY TYPES

A. FORWARD CONVERTORS

The forward convertor (Fig. 12) switches current through and inductor
to the load. The energy storage inductor is in series with the load.
The supply current is switched through the inductor to the load. The
magnitude of the supply current, hence the input power, is determined by
the inductor value and the duration of the switching pulse. The current
flowing during this pulse charges the filter capacitor, and power is
delivered to the load.
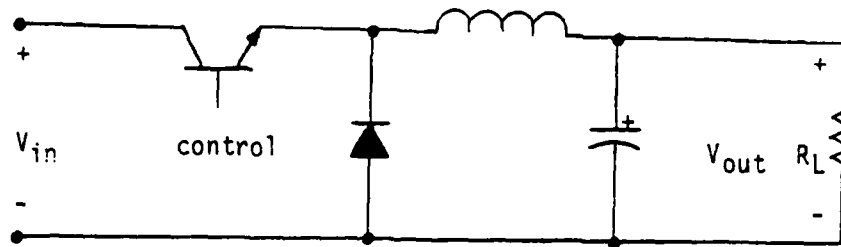
Figure 12. Basic Forward Convertor

To provide line isolation, a transformer must be used between the
switching transistor and the load. (Fig. 13)
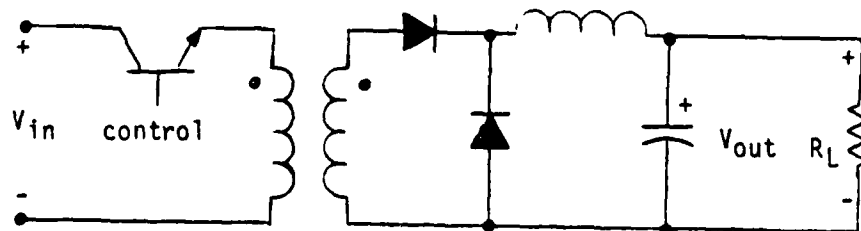
Figure 13. Forward Convertor With Line Isolation

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The major advantage of the forward convertor is that output ripple
is minimized due to the inductor being in series with the load. The
major disadvantages are the larger inductor value needed for a given
power output, and the added complexity over a flyback convertor when
isolation is required. [8]

B. FLYBACK CONVERTOR

The flyback convertor (Fig. 14) was so named because the energy
dumped into the inductive element is discharged to the load during
the flyback period or rest period of switching transistor.
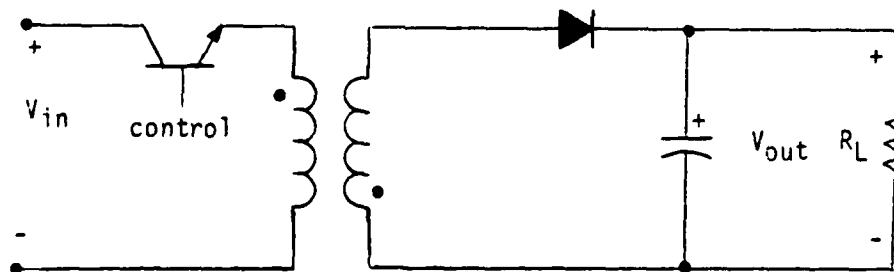
Figure 14. Flyback Convertor

Step-up or step-down convertors can easily be designed using fly-
back convertors by selection of the turns ratio of the transformer.
Multiple outputs require the addition of a secondary winding, rectifier,
and filter capacitor for each output. The flyback convertor provides
line isolation allowing voltage inversion by connection the ground
lead to either output terminal. [9]

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The flyback convertor suffers from output ripple that requires a large capacitor to provide adequate filtering. The higher ripple voltage is due to the smaller inductance of the secondary side of the transformer.

C. PUSH-PULL CONVERTOR

The third variation of switchers is the push-pull convertor. The circuit (Fig. 15) consists of a center-tapped power transformer. Each side of primary winding is switched into the circuit on alternate half cycles.

Push-pull convertors exhibit lower average currents in each side of primary winding than the flyback convertors. The peak current values are approximately equal however. The average currents in each of the switching transistors and rectifiers are half that in comparable convertors, thus cooling is easier.
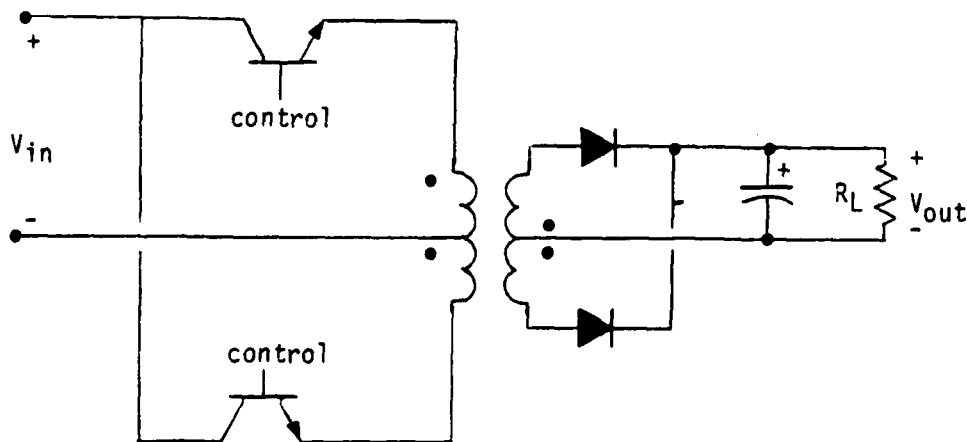
Figure 15. Push-Pull Convertor

The drive circuit of the push-pull convertor must be designed
such that both transistors will never overlap in the conduction cycle.
Any overlap due to finite rise and fall times of the switching transis-
tor will cause a current surge through the transistors, destroying one
or both devices.

One method developed to prevent on-state overlap is to introduce
a known dead time. This dead time must be a little longer than the
worst case transistor storage time. Most of control range was re-
quired for the regulation of the output, leaving less than ten percent
of the period for the dead time. This factor limits the operating
frequency of the push-pull circuit. [10]

A dynamic delay-sensing circuit has been developed by Ferranti
Electronics Limited. This circuit consisted of two fast-acting inhibits
connected to a pulse-steering gate that was connected to the base
drive circuit. The cross-coupled circuit was connected to a sense
circuit that detected when the transistors were at cut off. The oppo-
site transistor base current would not be applied until that condition
was met. This protective circuit compensated for transistor storage
delays and allowed safe operation at higher frequencies due to the
feedback control. [11]

Push-pull convertor circuits used are more complex than other
convertor designs. The added complexity is due to twice as many turns
required for the transformer, two switching transistors, the drive and
protection circuit each requires, and two rectifiers for each output
voltage required. Due to the added complexity and cost disadvantage,

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plus the lower average currents in the primary side of each switch, push-pull convertors are used only for higher power applications.

The output ripple is significantly reduced due to two switching pulses per period. Thus a higher frequency component of ripple and a smaller capacitor value is required in the output filter.

V. VOLTAGE REGULATION IN SWITCHING MODE POWER SUPPLIES

There are two basic methods of voltage regulation, with several
variations within the general types. The regulator must control the
amount of current dumped into the inductive elements within the convertor.

A. ON/OFF REGULATION

The simplest regulation technique operates by gating a fixed pulse
width whenever the output voltage drops. The period of oscillation is
dependent on the load current. The minimum period is fixed by an external
capacitor. This minimum period, or maximum duty cycle, limits the maximum
load current.

The Fairchild µA78S40 Universal Switching Regulator Subsystem consists
of an oscillator and a comparator, connected to an AND gate that is connected
to an RS flip-flop. The noninverting comparator input is connected to the
on-chip reference source. The inverting input is connected to a resistive
voltage divider that senses the output voltage. When the output voltage
drops, the comparator output goes high, gating the oscillator square wave
to the flip-flop thus providing a base current and switching current through
the inductive element. The oscillator is blocked when the output voltage
increases. [12]

The on/off regulator operation requires a large filter section to
attenuate the ripple voltage. The ripple increases when the load is re-
duced to less than maximum because the system is operating at a reduced
frequency.

B. PULSE WIDTH MODULATION REGULATION

Pulse width modulation regulators operate by increasing the pulse width when the output voltage drops. The Signetics NE5560 switched mode power supply control circuit, SG3524 regulating pulse width modulator and several other devices utilize a pulse width regulation technique.

The NE5560 circuit (Fig 16) consists of a sawtooth oscillator, pulse width modulator, RS latch, and an internal reference source. The output of the sawtooth generator is internally connected to the noninverting input of the pulse width modulator. The feedback sense voltage and reference voltage are connected to an operational amplifier. When the output voltage of the convertor drops below the desired level, the output of the error amplifier is inversely proportional to the output voltage. This error amplifier is internally connected to an inverting input of the pulse width modulator comparator. When the error voltage is maximum, the input to the modulator is minimum thus the sawtooth waveform is greater in magnitude than the error level for a longer period of time, producing a rectangular waveform whose duty cycle is proportional to the error voltage. When the error voltage decreases, the duty cycle decreases as shown in Fig. 17.
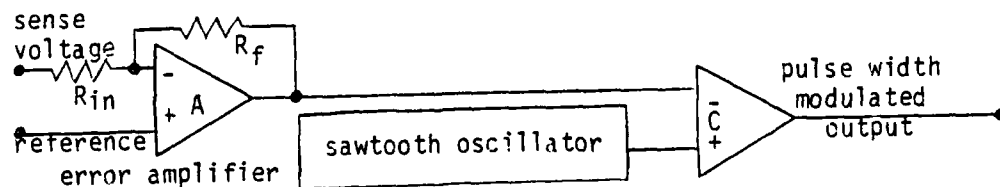
Figure 16. Basic Pulse Width Modulation Circuit Block Diagram
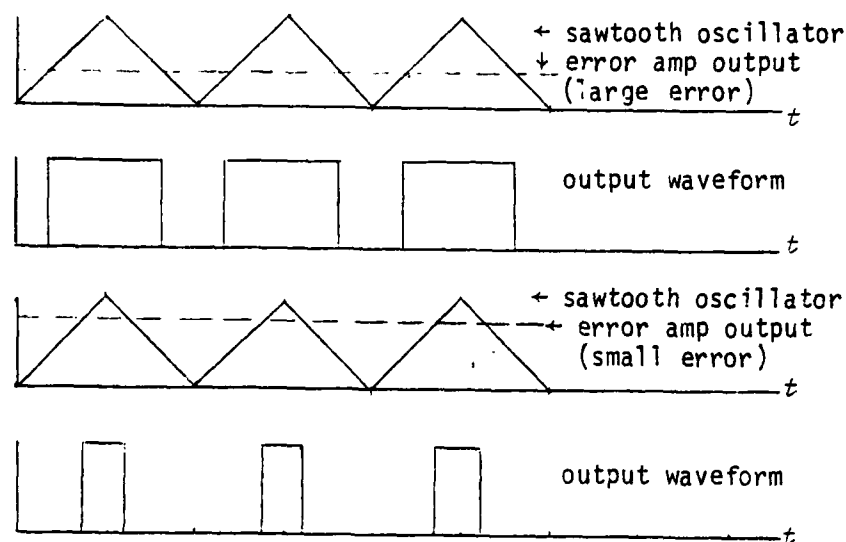
Figure 17. Pulse Width Modulation Waveforms

The NE5560 circuit incorporates additional control and protection circuits that limit the maximum duty cycle. The maximum duty cycle is selected by the designer and implemented with two external resistors. 13

The pulse width modulation technique is used in the majority of regulator control circuits. It produces an output that is at a constant fundamental frequency, thus ripple filtering is easier to accomplish than the on/off regulator. The disadvantage of the pulse width modulation circuits is the additional complexity. Fortunately the complexity is contained within the integrated circuit. The design steps are similar for each type of regulation system.

C. PSUEDO REGULATION FOR MULTIPLE OUTPUT SUPPLIES

There are two methods of regulation developed for multiple output power supplies. One method is to operate as two or more independent

convertors, each with its own control circuits. Each output is fully regulated and fault protected. This system achieves regulation specifications comparable to that of single output systems. Ripple is reduced by using pulse synchronization of all oscillators. All of the control circuits produce a pulse at the same time, the duration of each pulse being dependent on the load for that output.

The disadvantage of this regulation technique is an increase in the parts count, by a factor of two for a convertor with two output voltages.

A simpler form of multiple output regulation is a technique called pseudo regulation. In this system one control circuit is used with a feedback loop from one of the output voltages. For a second output voltage an additional secondary winding is added to the power transformer with the number of turns dependent on the desired output voltage. A rectifier and filter capacitor is used for each output as shown in Fig. 18.
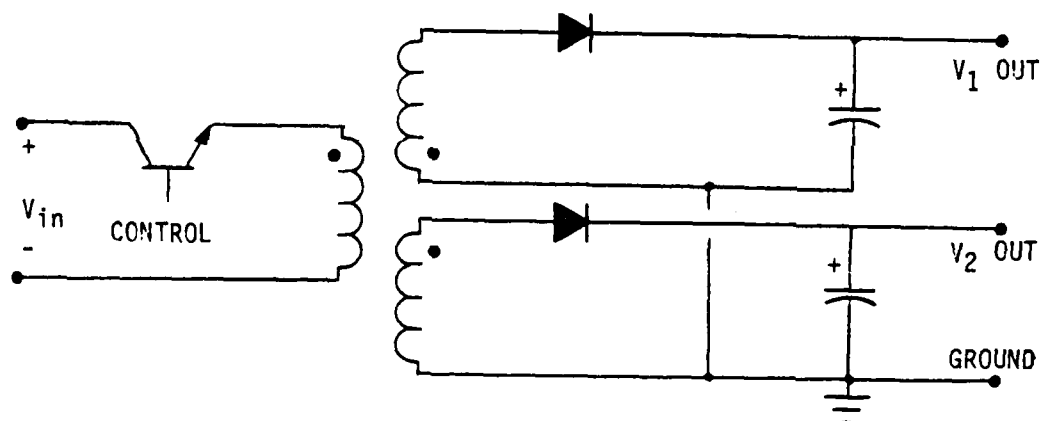
Figure 18. Multiple Output Regulator

The major advantage in pseudo regulation is the ease of adding additional outputs by adding only two components and an additional secondary winding. The cost of multiple output supplies are minimized. [14]

The limiting factor in this regulation technique is the poor load regulation. If the load currents of each output change by the same percentage, the regulation is acceptable, however when only one load current is changed the regulation degrades causing a change in voltage of up to 30%. The line regulation of the technique equals that of a single output supply. In systems where the regulation on the addition ouputs is not critical or where the load changes by small amounts or at the same time, this technique may be acceptable.

In power supply systems that have a low power secondary output, a linear series pass voltage regulator as in Fig. 19 may be used to improve regulaiton with an addition of one component at the expense of a decrease in efficiency.
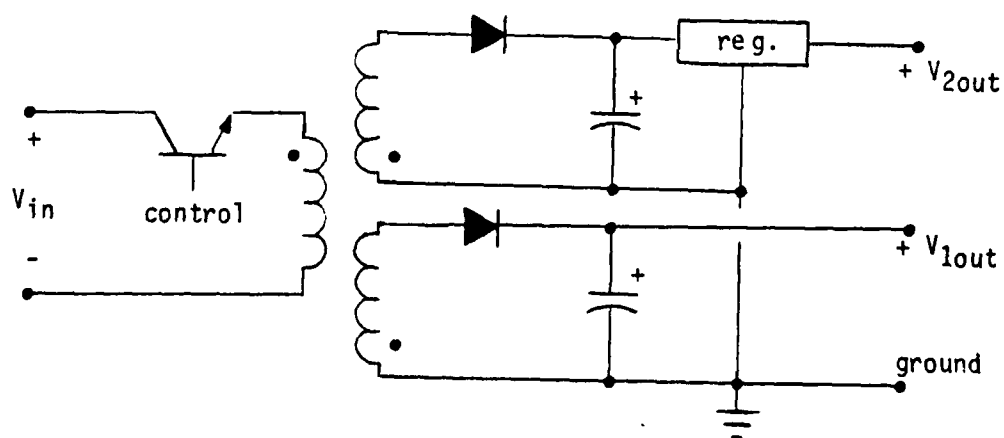
Figure 19. Multiple Output with Linear Regulator

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VI. BASIC SWITCHER DESIGNS

Switching mode power supplies can be configured to obtain a voltage that is greater than, less than, or of opposite polarity than the input voltage. These output levels can be obtained by using a flyback convertor with a transformer, setting the output voltage by the turns ratio and making the ground connection to the output to achieve a voltage inversion or noninversion as in Fig. 20.

(a)

(b)
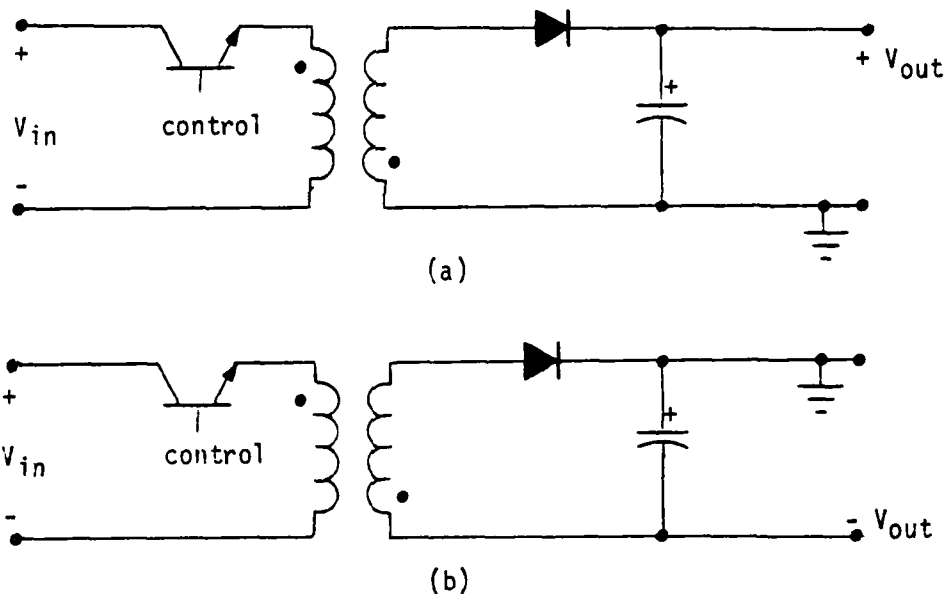
Figure 20. Flyback Outputs (a)Noninverting, (b)Inverting

For forward convertors, the output level depends on the placement of the inductor and free wheeling diode.

A. STEP-DOWN CONVERTORS

The step-down convertor (Fig. 21) is the most common circuit used. The convertor provides a low voltage output to the load.

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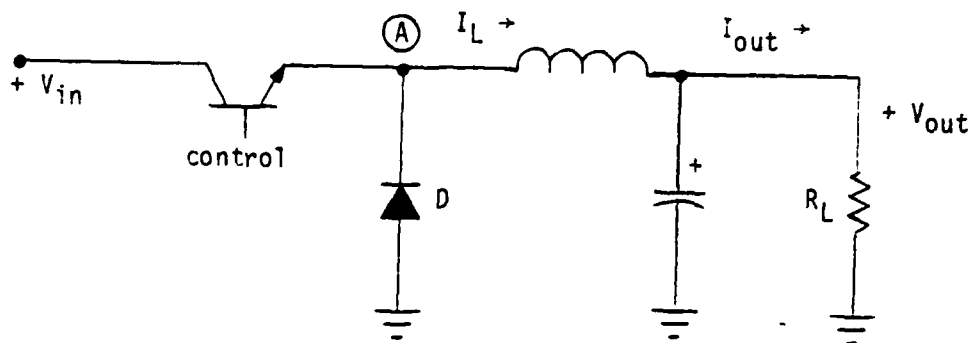
Figure 21. Step-Down Convertor

When the transistor switches on, the voltage of point A rises to
the supply voltage minus the transistor saturation voltage. The
supply current flows into the inductor, charging the capacitor and
supplying a current I_{out} to the load. The current through the inductor
increases linearly during the on period. The voltage drop across
the inductor is equal to the voltage at point A, less the output vol-
tage. When the transistor opens, the current through the transistor
stops. The current through the inductor continues to flow, linearly
decreasing, pulling point A below ground by one diode drop, with the
diode conduction. The duty cycle of the circuit is adjusted via feed-
back, to provide the required output voltage. The average current
through the inductor is equal to the load current (Fig. 22a). The
supply current is a linearly increasing function during the time the
transistor is on and zero when the transistor is off (Fig. 22b). The
average current from the supply is dependent on the duty cycle. The

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inductor used must be sufficiently large to provide a load current
during the off time. A larger value inductor reduces the peak currents
at the end of the one time.

The freewheeling diode conducts during the off time. The diode
limits the potential across the inductor, providing a current path
for the inductor. The capacitor filters the output voltage reducing
the ripple component.
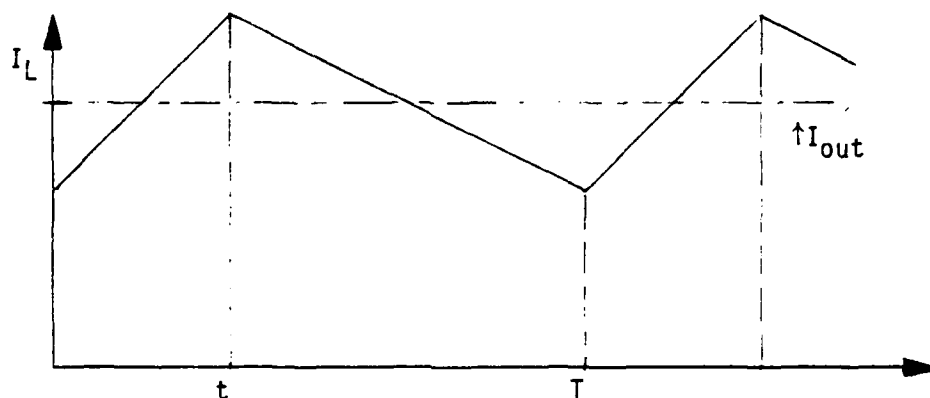
(a)
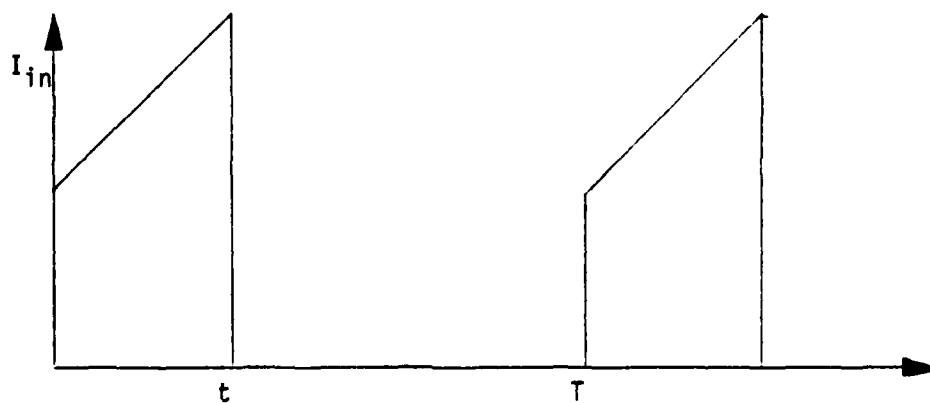
(b)

Figure 22. Current Waveforms of the Step-down Convertor
        (a) Inductor Current
        (b) Current through the Switching Transistor

B. STEP-UP CONVERTOR

The step-up convertor (Fig. 23) operates with the inductor in series with the supply. The transistor is connected from the inductor to ground.
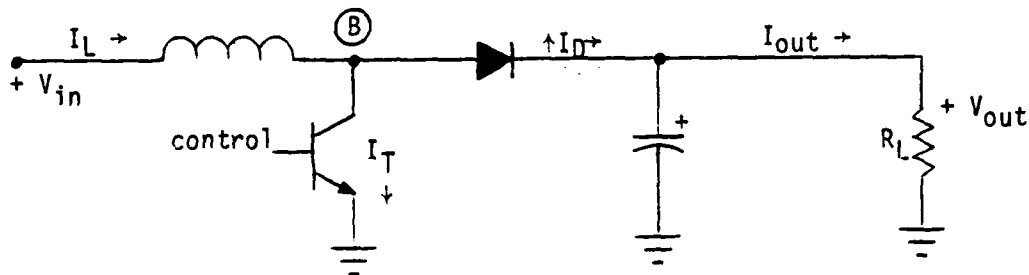
Figure 23. Step-up Convertor

The transistor is turned on, the voltage at point B goes to $V_{ce\ sat}$ above ground. The supply current begins flowing, increasing linearly (Fig. 27). When the transistor turns off, the voltage at point B rises, forward biasing the diode charging the capacitor and furnishing current to the load. The current in the inductor continues to flow, linearly decreasing. Current is drawn from the supply during the entire period. The voltage of point B rises to one diode drop above the output voltage. The transistor must turn on again prior to the current in the inductor decreasing to zero. The inductor value must be selected to supply current during the off-period.
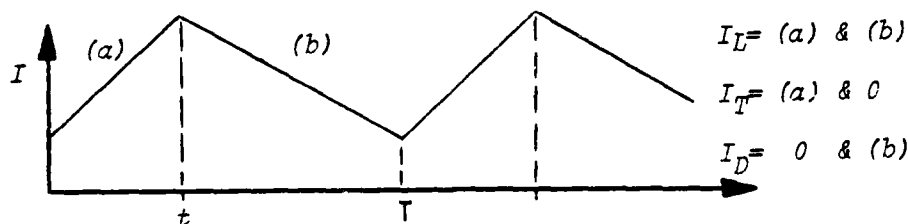
Figure 24. Current Waveforms of the Step-Up Convertor

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These high efficiency step-up convertors are used in many battery powered systems and in systems that require a higher voltage than the system DC bus. These convertors are very compact and simple to utilize.

C. INVERTING CONVERTOR

The function of the inverting convertor (Fig. 25) design is to supply a voltage that is of the opposite polarity of the source. The magnitude of the output voltage can be greater than or less than the input voltage.
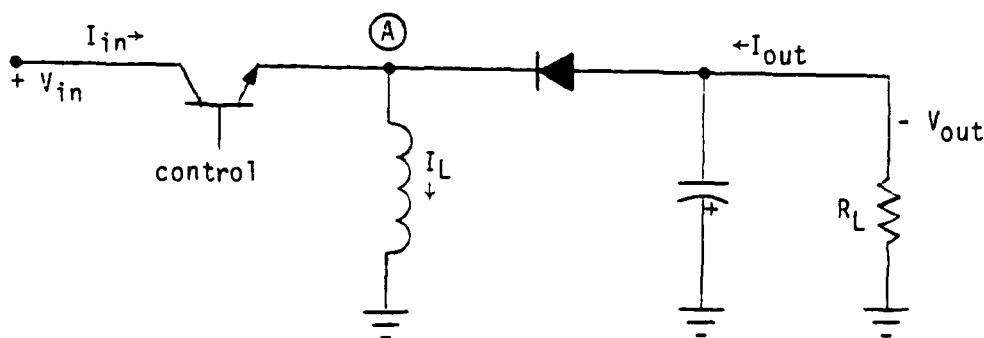
Figure 25. Inverting Convertor

The convertor operates by causing a current to begin flowing in the inductor, increasing linearly when the transistor switch closes. The voltage of point A is $V_{ce\ sat}$ less than the supply voltage. When the transistor switch opens, the current in the inductor continues to flow, decreasing linearly. The voltage of point A drops to one diode drop below the output voltage, (point A being more negative then the output). The diode is forward biased, with a current flowing through it and the inductor. The output capacitor filters the output ripple. The average inductor current is equal to the sum of the load current and the supply current.

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The inverting convertor system is a practical subsystem in complex equipment that requires a negative voltage for system operation. The compact size allows the invertor to be used as an on card convertor.[12,15]

VII. DESIGN OF A MULTIPLE OUTPUT ON/OFF REGULATOR

A. DESIGN SPECIFICATIONS

The object was to replace a linear voltage regulator used in an instrument comprised of digital and analog circuits. The existing linear regulator suffered from excessive heat buildup, thus requiring an exten- heat sink and fan for cooling. The heat buildup of the test set degraded the system performance in prolonged use. The choices in the design were to increase the package size to provide adequate cooling and thermal isolation, or to design a high efficiency switching power supply that would fit into the restricted volume within the package. The high efficiency switcher would operate cooler thus alleviating the thermal problems.

Table III. Power Supply Specifications

Outputs:

+5 VDC	2.0 A	10 W
+12 VDC	0.5 A	6 W
-12 VDC	0.5 A	6 W

Total Output Power 22 W

AC input: 115 VAC or 230 VAC, 50 to 60 Hz
Line Variation ± 10%

Output Ripple: ± 50 mV$_{p-p}$

Regulation: ± 5% for the ±12 V outputs
± 1% for the ± 5 V output

The existing linear system (Fig. 26) used a 50-60 Hz step down trans- former with two primary windings and three secondary windings.
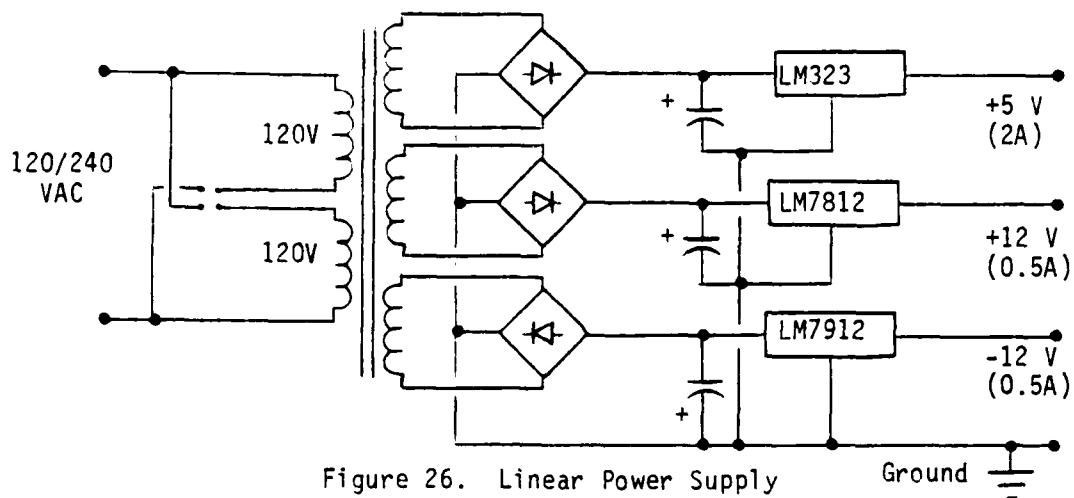
Figure 26.   Linear Power Supply

To simplify design and minimize cost it was decided to utilize a
step-down transformer, replacing the linear series pass regulators
with step-down switching regulators. The system required load regulation
on each output, thus three regulators would be required (Fig. 27).

The input step-down transformer selected was a dual primary, dual
secondary unit. A 25 VA transformer was used, a smaller transformer
than the linear supply required. The Fairchild μA78S40 Universal
Switching Regulator Subsystem integrated circuit was selected because
it incorporated a switching transistor capable of 1.5 A and an internal
diode rated at 1.0 A thus minimizing external components in the low
current supplies. The transformer was selected to have an output volt-
age of 16 Vrms. The secondary voltage selected must be high enough
to be greater than the output voltage by approximately two volts,
during low line voltage conditions. The high voltage condition was
limited by the maximum IC supply voltage of 40 V.

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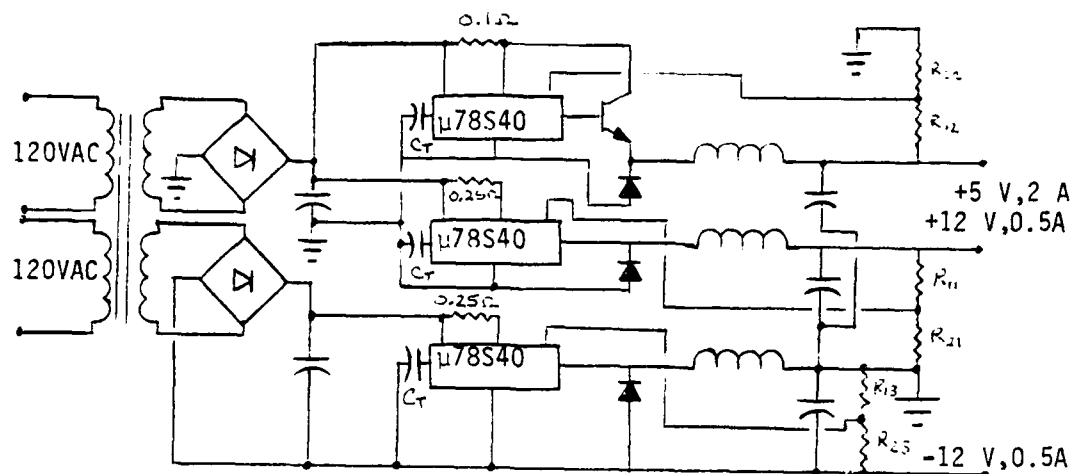
Figure 27. 22 Watt, +5 and ±12 VDC Switching Mode Power Supply

Two secondary output were required for the power supply. The inversion circuit was not used because it would have required an external transistor and diode. Both secondaries of the transformer were used, each with a fullwave bridge and filter capacitor. The second winding was connected to the step-down convertor to provide -12 V. The most positive output of this step-down convertor was connected to the output ground; the most negative output provided -12 VDC. This was accomplished due to the isolation between the secondary windings.

B. INTEGRATED CIRCUIT OPERATION

The +5 V, 2A convertor required the use of an external transistor to switch the higher peak currents. One version of this supply was built and tested using three μA78S40 circuits operating in parallel. Figure 28 illustrates parallel operation of two circuits. The oscillator required a timing capacitor that was three times the nominal value.

The auxiliary operational amplifier was used as a comparator, connected
to each of the main comparators, causing each circuit to regulate and
and generate switching pulses at the same time. On a cost basis this
configuration was rejected in favor of one regulator circuit with a
transistor. If the integrated circuit cost decreases for the µA78S40
due to large production runs, using multiple regulators in parallel may
become cost effective.
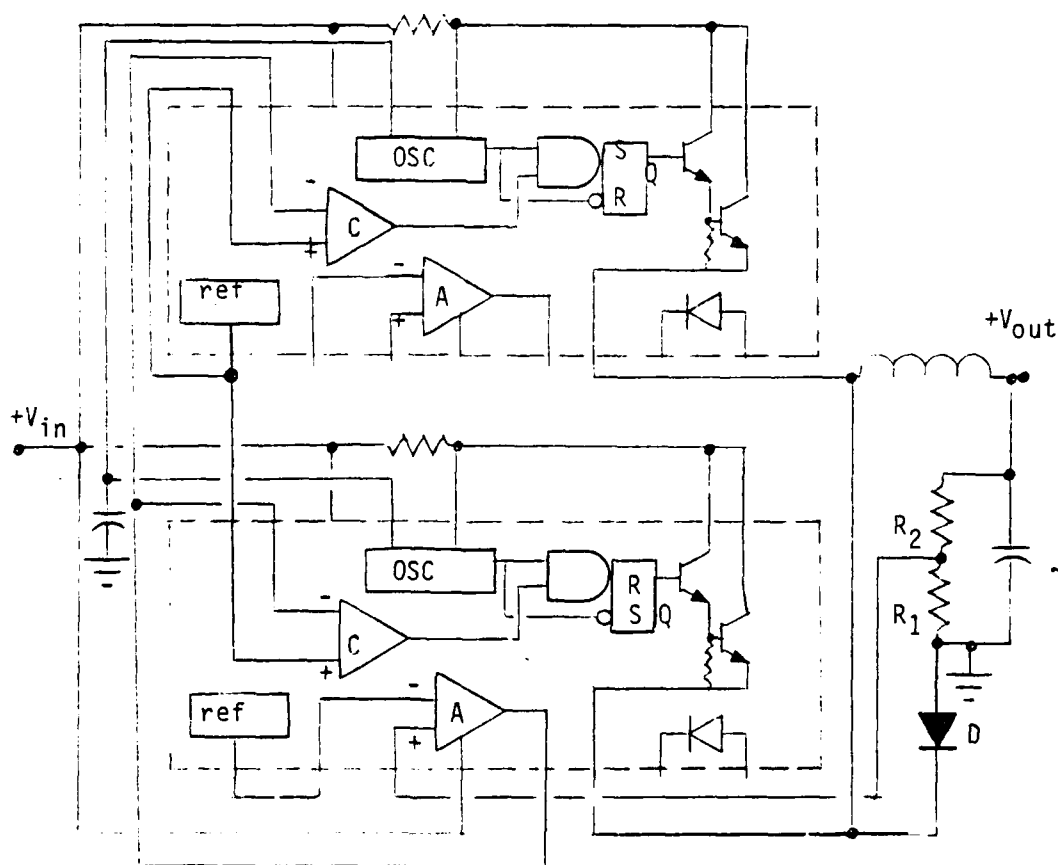
Figure 28. Parallel Operation of the µA78S40

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The ±12 V circuits were constructed with the on-chip diode connected. The diode was rated at 1.0 ampere, however above a current output of 350 milliamperes, the integrated circuit heated excessively. The heating caused a loss of accurate regulation. This thermal drift was corrected by connecting an external freewheeling diode. This relieved the thermal load from the integrated circuit. The typical voltage drop across the diode was specified at 1.25 to 1.5 volts. With an average current of 350 milliamperes, this resulted in a power dissipation of 437.5 to 525 milliwatts for the test conditions. The output transistor saturation voltage of 1.1 to 1.3 volts was also a prime source of heat.

The diode used was a fast recovery Schottky diode, Varo type VSK 530; rated at 5 amperes continuous with a peak inverse voltage of 30 V. The low voltage supply consisted of seven external components, for each section. The only external semiconductors used were the fast recovery diodes and one transistor. Two resistors were used in the feedback loop, connected as a voltage divider. The sense voltage is connected to pin 10, (the inverting input of the comparator). A timing capacitor was selected for a nominal frequency of 20 KH_z under full load (0.02 µF). The output filter capacitors (electrolytic 1000 µF, 15 VDC for the ±12 V outputs; 2200 µF, 6.3 VDC for the 5 V output) were selected to keep the output ripple within specifications. The final external component used was the short circuit sense resistor. This resistor sets the limit of the peak current. The integrated circuit measures the voltage drop acorss this resistor. The circuit current limits when the voltage across this resistor reaches 250 to

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350 millivolts, limiting the peak current through the circuit. A
resistor value of 0.25Ω was selected to limit the peak current to
approximately 1.0 ampere on the ±12 V supply.

C. INDUCTOR SELECTION AND DESIGN

For the step-down convertor, the value of the inductor must be
selected to be at least equal to:

$$L \geq \left(\frac{V_{out} + V_d}{I_{peak}} \right) t_{off} \qquad (45)$$

where:

V_d = voltage drop across the diode

I_{peak} = peak current $\simeq$ twice output current

t_{off} = off time, set by timing capacitor

and

$$t_{off} \ (\mu sec) = \frac{C_T \ (\mu F)}{45 \times 10^{-5}} \qquad [16] \qquad (46)$$

For the design specifications of the power supply:

$$t_{off} = 44 \ \mu sec \qquad (47)$$

$$L = \frac{(12 + 1.0) \ V}{1.0 \ A} \ (44 \ \mu sec) \qquad (48)$$

thus the inductor must be at least

$$L = 572 \ \mu H \qquad (49)$$

An inductor value of 1 mH was selected to reduce the peak currents
and to allow for any tolerances in the timing capacitor, diode
drop, or inductor. The inductor must handle an average current equal
to the output current. This limits the wire size of the inductor used.

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A safe value of 500 circular mils per ampere will insure a reliable inductor with minimal heat rise. For a 500 milliampere output, this required at least # 26 AWG. A toriod inductor of 1 mH was easily constructed. The device has a diameter of 1.20 inches and a height of 0.375 inches, the size being reasonable for this circuit.

D. ALTERNATE CIRCUIT IMPLEMENTATION

The power supply operation was satisfactory, meeting the design specifications. The only problem that existed was a single source supplier and long delivery times for the Fairchild µA78S40 integrated circuit. There existed no pin for pin replacement for this device. Therefore an alternate power supply circuit was designed around an available regulator chip. The 1524/2524/3524 family of Regulating Pulse Width Modulators, manufactured by Signetics, Silicon General and National Semiconductor with a similar device MC3420 by Motorola, was selected. The Motorola MC3420 was not pin for pin compatible however the circuit operation and general specification were similar. The Signetics SG3524 was used in the circuit.

The SG3524 circuit operated by using a fixed running frequency, regulating by varying the pulse width. The circuit (Fig. 29) consisted of dual output transistor with the option of a push-pull convertor. This option was not exercised in the interest of circuit simplicity. The output transistors were rated at a maximum current of 100 mA each. Thus external switching transistors were required for each of the circuits.[17]
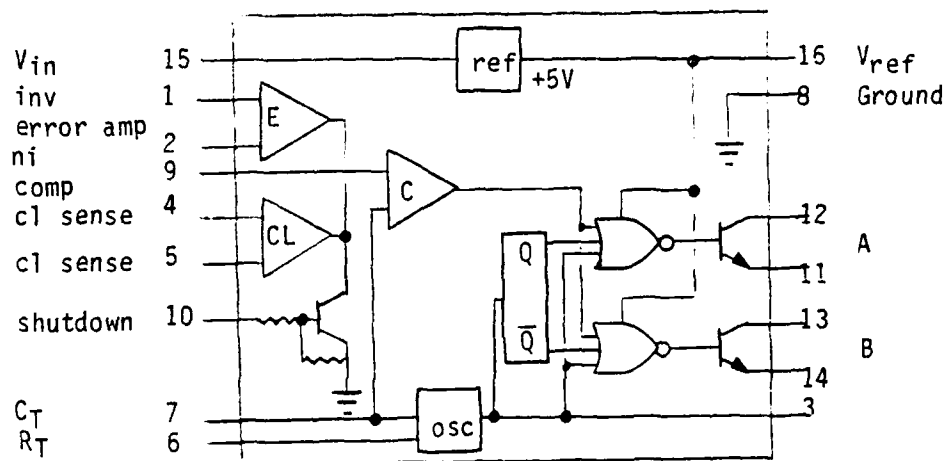
Figure 29. SG3524 Convertor Block Diagram

The circuit required an R-C phase compensation network from pin 9 to ground (0.1 µF capacitor and a 11 KΩ resistor in series) for loop stability.
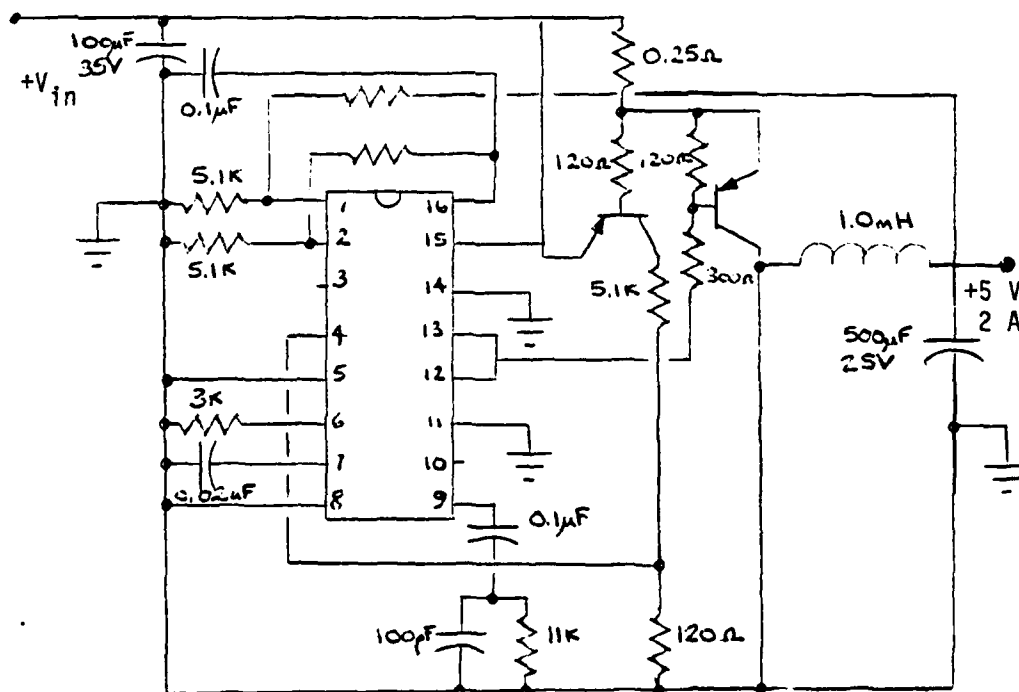
Figure 30. SG3524 Step-Down Convertor

The circuit (Fig. 30) met the design specification. The output ripple
was less than that of the μA78S40 circuit. The advantage of the SG3524
circuit was the set operating frequency. Under conditions of no load
or partial load, the frequency did not change. This condition made it
easier to filter the output ripple and contain the RFI. This circuit is
promising for audio circuits due to the operating frequency always being
above the audio frange, minimizing interference.

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VIII. DESIGN OF A 35 WATT, 5 VOLT DC TO DC CONVERTOR

A. DESIGN SPECIFICATIONS

The design requirements were to design a DC to DC convertor with the specifications shown in Table IV.

Table IV. 35 Watt DC/DC Convertor Design Specifications.

Input:	48 VDC, positive ground operating range 42 V to 56 V maximum input current 2 ampere
Output:	5.2 VDC negative ground Variable from 4.8 to 5.2 V (preset) 7 Amps full load regulation ±1% (52 millivolts) no load to full load ripple ±1% (52 millivolts)
Physical Dimensions:	
	$4\frac{1}{2}$" W x $8\frac{1}{2}$" L x 1" H, with fingers for a 24/48 card connector
Temperature Range:	-10°C to +50°C with no change in performance
Protection Circuits:	
	Overcurrent (short-circuit) protection. Overvoltage protection.

This circuit presented several requirements that complicated the design. The invertered polarity input, high input voltage, relatively high current output, and physical dimensions presented the most difficult criteria.

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B. DESIGN IMPLEMENTATION

The circuit was designed as a pulse width modulated flyback circuit with the Signetics NE5560 Switched-Mode Power Supply Control Circuit used as the control element. The NE5560 (Fig. 31) incorporated several functions that gave it a great deal of flexibility. The circuit included a soft-start feature that causes the output voltage to increase slowly when first turned on, eliminating voltage overshoot and limiting the surge current. The circuit also inludes a current limit sense that protects the system in event of an overload or short circuit condition. An over-voltage circuit protects the system, shutting down the circuit when the output voltage rises above a preset limit. A remote on/off feature allows a TTL control signal to power up or turn off the circuit. This would allow flexibility of designing CPU power supplies that could con-serve power by turning on the power to an auxillary device only when the section was required. [18,19]
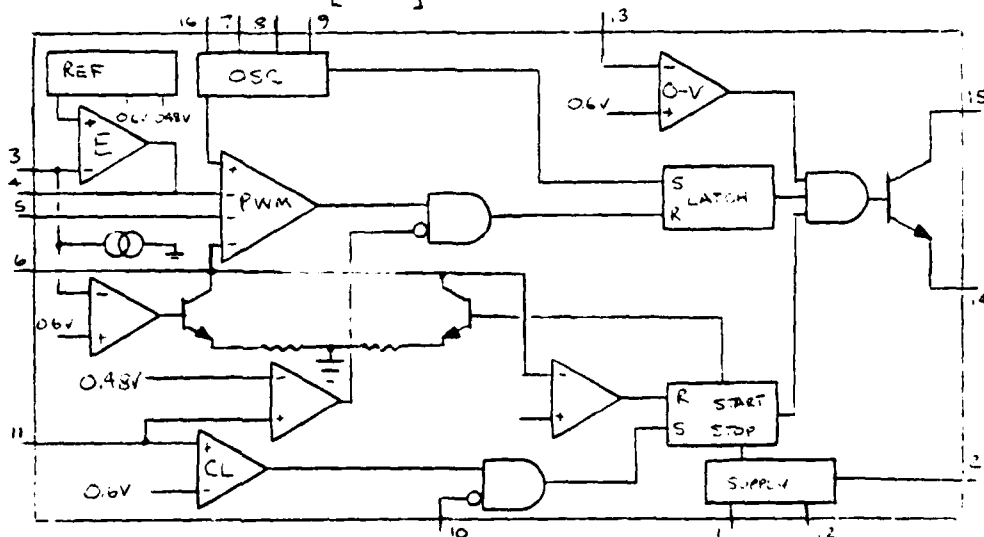
Figure 31. Signetics NE5560 Block Diagram

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The integrated circuit maximum input supply voltage is 18 V. This limit required the use of an efficient method of supplying a voltage for the integrated circuit and driver circuit.

The flyback convertor configuration was selected due to the smaller inductor value required and the smaller surge current in the input side of the system during turn on.
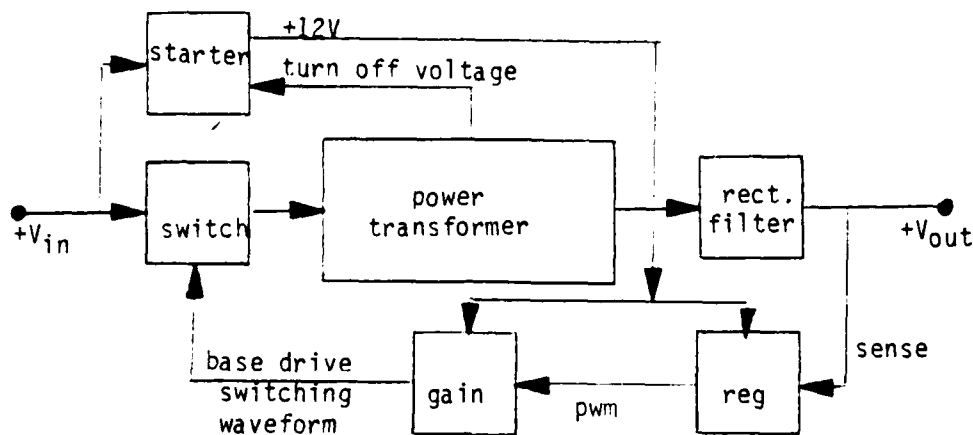
Figure 32. 35 W DC to DC Convertor, Block Diagram

C. STARTER CIRCUIT OPERATION

Figure 32 shows a block diagram of the convertor. To minimize regulation drift, circuit complexity, and to maximize loop stability, the regulator circuit was placed on the output side of the circuit. A simple resistive voltage divider was all that was required in the feedback loop. A capacitor was used for phase compensation. The chip and driver voltage were provided by an additional winding on the power transformer, once the system was operating. The problem was to provide a voltage to start the circuit. This voltage had to be inverted in polarity from the -48 V input (positive ground). The 48 V potential was too great

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for most integrated circutis. To solve this problem a starter circuit was built that consisted of a 555 astable multivibrator and a high frequency transformer. The 555 was designed to operate at a forty percent duty cycle, connected to a transformer with a DC blocking capacitor. The 555 chopper operated at 20 KHz. The 200 milliampere output of the 555 was supplied from the-48 V line with a series dropping resistor and a transistor switch. When the system was first turned on, the 555 chopper produced a +11.5 volt output that powered the NE5560 and driver stage. This circuit drove the transistor switching circuit. When the main circuit came on-line, the 5 volt output was available. When the circuit was operating, a capacitor was slowly charged with a long RC time constant. The charging began when the magnetic flux had built up in the power transformer. The capacitor continued charging until a zener diode turned off the transistor switch connected to the 555. The 555 then ceased to operate, the system continued to operate because of the +12 V winding on the power transformer that provided the voltage for the regulator/driver.

The starter circuit was inefficient, wasting power in dropping resistor (35 volts at 200 milliamperes) however the starter operated for only three to fifteen seconds each time the device was turned on.
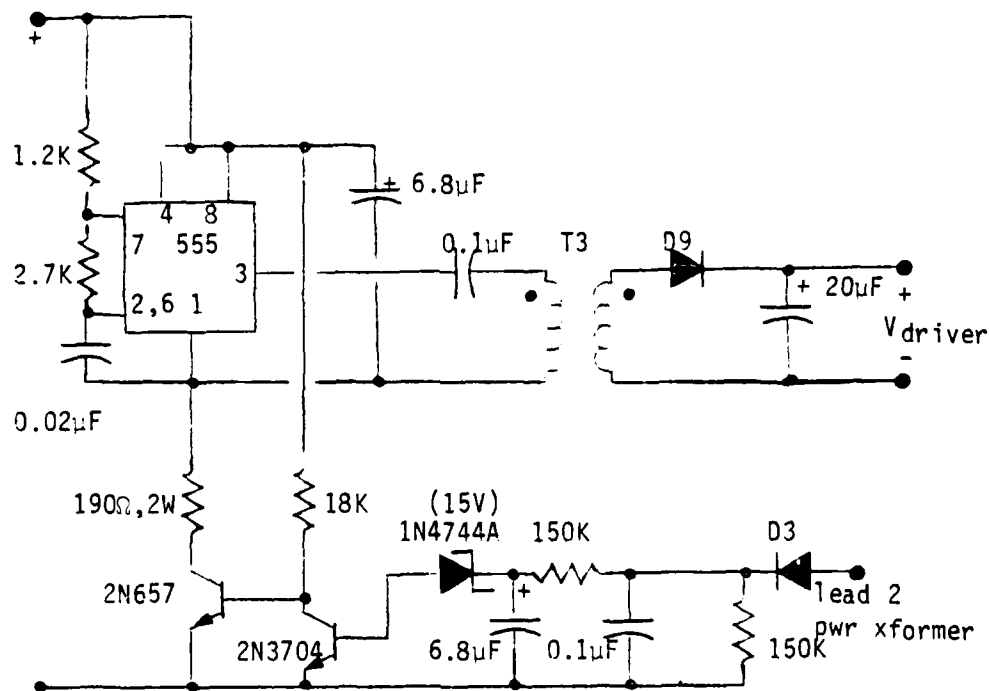
Figure 33. Starter Circuit

As shown in Fig. 33, the zener diode 1N4744A (15 V) provided a snap-off for the starter, letting the capacitor charge to 15 V before the diode avalanched. This snap action switched the 555 off by saturating the 2N3704 transistor, turning off the 2N657 transistor. This quick switching kept the transistors out of the active region. If the transistors were to remain in the active region, the circuit would have gone into an un-stable mode, as follows: the starter would have run (producing +12 V); the regulator chip would have begun running and the output would have started increasing. The power transistor switching would cause the tran-sistor in the starter circuit to become active, decreasing the 555 current and decreasing the +12 V regulator supply. The oscillation would continue

at a rate of approximately 1 cycle per second with the output voltage of 0.4 to 1.2V. The oscillation would cause the transistor switch 2N657 (TO-5) to heat up, destroying the transistor. This unstable mode was dependent on the load resistance and input voltage. The instability was most severe in the worst case conditions, low voltage (-42 V) input and maximum load (7 ampere) current. The addition of the 1N4744A zener diode prevented the start-up oscillation under those worst case conditions.

D. POWER TRANSISTOR SELECTION

The most critical component of the switching mode power supply was the switching transistor. The higher the input voltage, the more critical this component became. The transistor had to have an adequate V_{ceo} rating to avoid breakdown during the turn off of the inductor current. At a 48 V input, the maximum collector to emitter voltage measured was 150 volts.

The current rating of the transistor had to be greater than the peak current of the inductor. The peak current was approximately twice the average current or approximately 4 amperes. The rise and fall times of the transistor are important for efficient operation while the thermal resistance of the device is important in high power output. The thermal resistance was not a factor in the selection of the power transistor for this application.

The transistor selected was the General Electric NPN Power Switching transistor D64VE3. This transistor provided up to a two percent increase in efficiency over other TO-3 switching transistors.

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The base drive of the switching transistor was provided by a driver transistor and a pulse transformer. To insure transistor saturation, a base current of 400 milliamps was selected. [20]

E. TRANSISTOR PROTECTION CIRCUIT

In Fig. 34, the current built linearly when the transistor switched current through an inductive load. When the transistor switch was opened, the current in the inductor continued to flow, decreasing linearly. This caused the voltage at the emitter to drop.
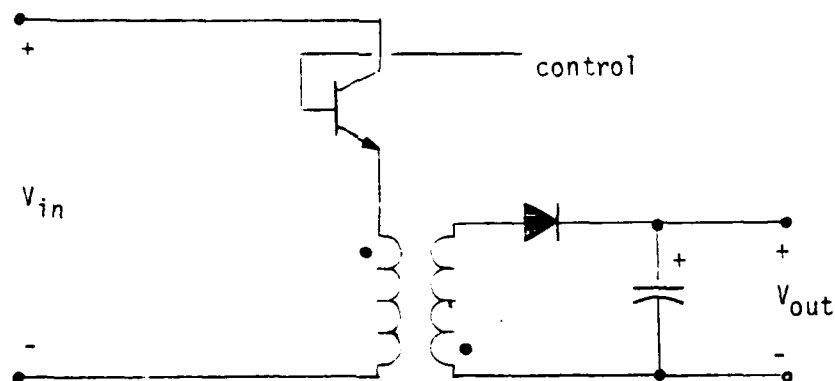
Figure 34. Transistor Circuit, No Protection Against Breakdown
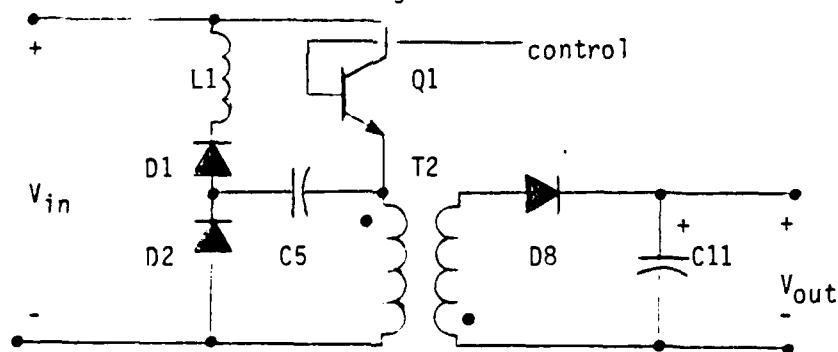
Figure 35. Transistor Protection Circuit

The transistor would have broken down without some provision to limit the voltage drop. The solution was to dump charge into the emitter by discharging a capacitor. In Fig. 35, C_5 limited the potential across the transformer. When the transistor was switched on, the capacitor was discharged. When the transistor opened, the potential of the emitter dropped, C_5 was charged through D_2. When the transistor saturated, C_5 discharged through D_1 and L_1.
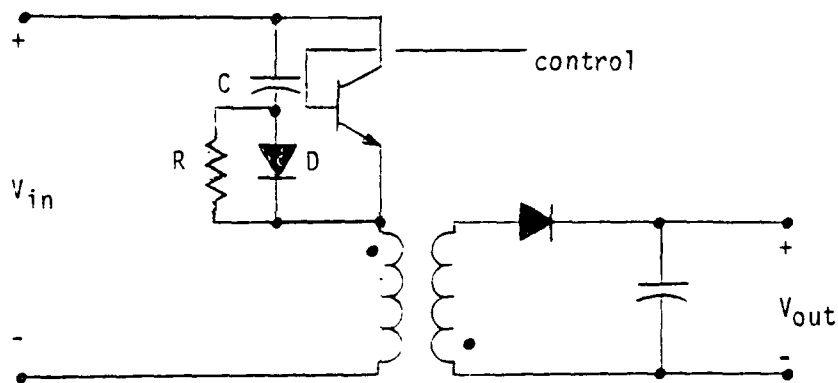
Figure 36. Alternative Transistor Protection Circuit

The inductor stored energy, dumping current back into the supply. The protective circuit in Fig. 36 operated in the same basic manner. The snubber resistor dissipated the power during the on/off transient. This element resulted in an average power loss of 1 W, decreasing the overall efficiency of the system. The inductive protective circuit transfers and stores energy, with very little power dissipation. Diode D_1 is a general purpose silicon diode.

F. POWER TRANSFORMER DESIGN

The power transformer must be chosen to have a large enough inductance to store the required energy during the on period, and be able to supply the required load current without the windings overheating causing excessive losses.

The inductance of the transformer primary winding was calculated from the switching current waveform shown in Fig. 37.

Figure 37. Current in the Transformer Primary

$$v = L \frac{di}{dt} \tag{50}$$

$$\Delta I_L = \frac{V_{in}}{L} \pm 1 \tag{51}$$

where:

V_{in} = source voltage

L_u = inductance in H

t = on time

T = period

I_p = peak current in inductor

ΔI_L = current increase during switching

I_1 = initial current

I_{avg} = average current from source

δ = duty cycle t/T

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$$I_1 t + \frac{\Delta I_L t}{2} = I_{avg} T \qquad (52)$$

Substituting Eq. (51) into (52)

$$I_1 t + \frac{V_{in}}{2L_0} t^2 = I_{avg} T \qquad (53)$$

$$I_1 + \frac{V_{in}}{2 L_0} t = I_{avg} \frac{T}{t} \qquad (54)$$

$$I_p = \Delta I_L + I_1 \qquad (55)$$

Rearranging Eq. (55)

$$I_1 = I_p - \Delta I_L \qquad (56)$$

Substituting Eq. (56) into Eq. (54) and simplifing

$$I_p = I_{avg} \frac{T}{t} + \frac{V_{in}}{2L_0} \qquad (57)$$

For $I_1 = 0$, then

$$\frac{V_{in}}{2L_0} t = I_{avg} \frac{T}{t} \qquad (58)$$

Solving for L_0,

$$L_0 = \frac{V_{in} t^2}{2I_{avg}T} \qquad (59)$$

Then

$$I_p = I_{avg} \frac{T}{t} + 2I_{avg}\frac{T}{t} \qquad (60)$$

$$I_p = 3 I_{avg}\frac{T}{t} \qquad (61)$$

For the supply:

$$V_{in} = 48 \text{ V} \qquad (62)$$

Input power is

$$P_{in} = \frac{1}{\eta} P_{out} \qquad (63)$$

$$P_{in} = \frac{1}{0.80}(35 \text{ W})$$

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Then

$$P_{in} = 43.75 \text{ W} \tag{64}$$

$$I_{avg} = 0.91 \text{ A} \tag{65}$$

$$T = \frac{1}{20 \text{ KHz}} = 50 \text{ μsec} \tag{66}$$

$$t = 0.3 \text{ T} \tag{67}$$

$$t = 15 \text{ μsec (nominal operating conditions)} \tag{68}$$

Solving for Eq. (59)

$$L_0 = \frac{(48 \text{ V}) (15 \text{ μsec})^2}{2(0.91 \text{ A}) (50 \text{ μsec})} \tag{69}$$

$$L_0 = 119 \text{ μH}$$

To minimize the peak current, select a transformer with an inductance greater than L_0.

For $L = 2L_0$, $L = 0.24$ mH

Solving for I_p, from Eq. (57)

$$I_p = 2I_{avg} \frac{T}{t} \tag{70}$$

$$I_p = 6.07 \text{ A} \tag{71}$$

For $L = 3L_0$, $L = 0.36$ mH

$$I_p = I_{avg} \frac{T}{t}(1 + \frac{2}{3}) \tag{72}$$

$$I_p = 5.06 \text{ A} \tag{73}$$

For larger inductor values, ΔI_L decreases as I_1 increases as shown in Fig. 38.
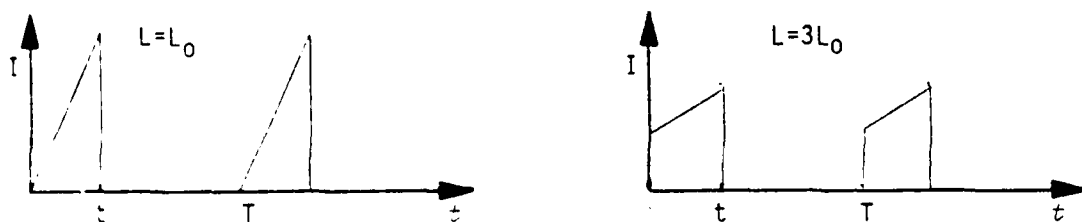
Figure 38. Peak Transformer Currents

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For calculating the turns ratio, use Eq. (74)

$$L \frac{\Delta I_L}{(T-t)} = n(V_{out} + V_d) \quad\quad (74)$$

where

V_{out} = Output Voltage

V_d = drop across diode

n = turns ratio

Substituting from Eqs. (51) and (74)

$$L \frac{V_{in}}{L(T-t)} t = n (V_{out} + V_d) \quad\quad (75)$$

$$\frac{V_{in}}{(\frac{T}{t} - 1)} = n (V_{out} + V_d) \quad\quad (76)$$

Solving for n

$$n = \frac{V_{in}}{(V_{out} + V_d)(\frac{T}{t} - 1)} \quad\quad (77)$$

For the supply:

$$n = \frac{48V}{(5+1)V \left(\frac{50\mu s}{15\mu s} - 1 \right)} = 3.43$$

The pulse width will vary to regulate for changes in the line voltage and load currents. The pulse width was selected at 15 µsec for efficient operation, a trade off between losses in the diode and in the transistor. The maximum pulse width must be set by an external resistor to the NE5560 such that it will provide operation during low line conditions.

Solve for Eq. (76)

$$t = \frac{T}{1 + \frac{V_{in}}{n(V_{out} + V_d)}} \quad\quad (78)$$

For low line voltage conditions (V_{in} = 42 V)

$$t = \frac{50 \text{ } \mu sec}{1 + \frac{42}{3.43(5+1)}} \tag{79}$$

$$t = 16.5 \text{ } \mu sec \tag{80}$$

$$\delta = 33\% \tag{81}$$

To allow for component tolerances the maximum permissible duty cycle of the regulator was limited to thirty five percent.

The power transformer selected was wound on a ferrite toroid core. The transformer was selected with an inductance of 0.4 mH. The turns ratio selected was 3.0. The wire size was selected to limit the current density to 500 circular mils per ampere. This required an area of 2530 circular mils for a peak current of 5.06 A. This required an area of 2530 circular mils for a peak current of 5.06 A. This required the use of 16 AWG (2580 circular mils). The peak current in the secondary winding was calculated to be 11.8 A. The required area was 5900 circular mils. This required the use of 12 AWG, or multiple strands of a smaller gauge wire. To ease the winding, bifilar conductors of 16 AWG were chosen for the secondary.

For ferrite cores, the magnetic flux must be limited to approximately 2000 Gauss. A core is then selected that will have the required inductance of 0.4 mH and the physical size of the transformer compatible with the application. The core selected was Ferrocube part number 55121-A2-GR (u=60) with 45 turns on the primary, 15 turns on the secondary (bifilar) Thirty additional turns of 22 AWG were added for the integrated circuit and driver voltage supply.

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Several different power transformers were built and tested with different turns ratios and inductance values. The efficiency varied by one percent or less for each transformer.

G. NE5560 SWITCHED MODE POWER SUPPLY CONTROL CIRCUIT

The integrated circuit used for the power supply was the Signetics NE5560. This device required a supply voltage of 10 to 18 V. The circuit generated a pulse whose duration was proportional to an error voltage. The error voltage was generated by a feedback sense voltage and a reference voltage. The error signal was connected to an internal high gain amplifier and was then fed to the pulse width modulator section. The gain of the amplifier was set at 200 with an external feedback resistor. The gain of the amplifier sets the degree of regulation. For a closed loop gain of 200 and a duty cycle of thirty percent, the regulation was 0.3 percent. A higher gain reduced the loop stability of the system. This high gain created a small linear range that the circuit operated. When the output dropped enough to drive the error voltage out of the linear range, the pulse width went to the maximum. The maximum duty cycle was set by an external resistor to be approximately thirty five percent. When the output voltage rose above the desired value, the error voltage decreased the pulse width to a minimum of approximately 0.5 usec or one percent. The circuit regulated on a pulse to pulse basis. The fast acting feedback loop controlled the pulse width continually within the 50 usec period.

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The output of the regulator chip was connected to the base of silicon NPN transistor 2N657. This transistor provided a current gain. The NE5560 output stage was rated at 40 mA. The output stage consisted of an NPN transistor with the collector and emitter available on pins 15 and 14 respectively. This pin out allowed the flexibility of connecting the output transistor to a separate supply voltage if required (+5 V for a TTL driver for example)

H. DRIVER STAGE AND PULSE TRANSFORMER

As shown in Fig. (39a) the output stage of the integrated circuit was connected to the +12 volt output through a collector resistor of 2.2 KΩ. The emitter was connected to a 1.3 KΩ resistor to ground. The base of the driver transistor was connected to the emitter (pin 14). This provided a base drive of 5.5 mA. The driver transistor was connected to a pulse transformer. When the driver transistor saturates, a current is switched through the pulse transformer, building linearly. The inductance of the pulse transformer was calculated to be 0.9 mH. The secondary of the pulse transformer was connected to the base and emitter leads of the power switching transistor.

The pulse transformer served three purposes. The first was to provide isolation from the line to load. The regulator must control the base of the switching transistor, on the line side. The pulse transformer provided this isolation.

The second function the pulse transformer provided was to produce a current gain to insure adequate base drive. The pulse transformer

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was a step-down transformer, converting a twelve volt switching waveform to a high current one volt pulse. The turns ratio required to keep the power transistor saturated was 11.43 (160 turns on the primary, 14 turns on the secondary). The peak current in the primary winding was calculated to be 173 mA. This resulted in a secondary current peak value of 1.98 A as shown in Fig. (39 b). The transformer was connected in phase. A current through the secondary is produced at the same time the primary current flows. The secondary current linearly increases thus more base current is provided at the end of the switching pulse, when the additional current is needed most.

The final function of the pulse transformer is to improve the switching waveform. At the end of the switching pulse, the induced voltage in the secondary goes negative. Charge in the base emitter junction is quickly removed, thus the switching transient for the power transistor is improved.

Some switching power supplies operate with the control element on the primary side of the supply. An optoisolator consists of an LED and a phototransistor. Typical devices provide 500 volts of isolation or more packaged in a TO-5 can or 8 pin DIP. An optoisolator will not provide current gain or improve the switching transient by quickly removing the base current. Optoisolators may drift with temperature or over the long term operation.

For this application, the pulse transformer was the simplest method to provide base drive. A protective circuit was added to the primary to avoid transistor breakdown.

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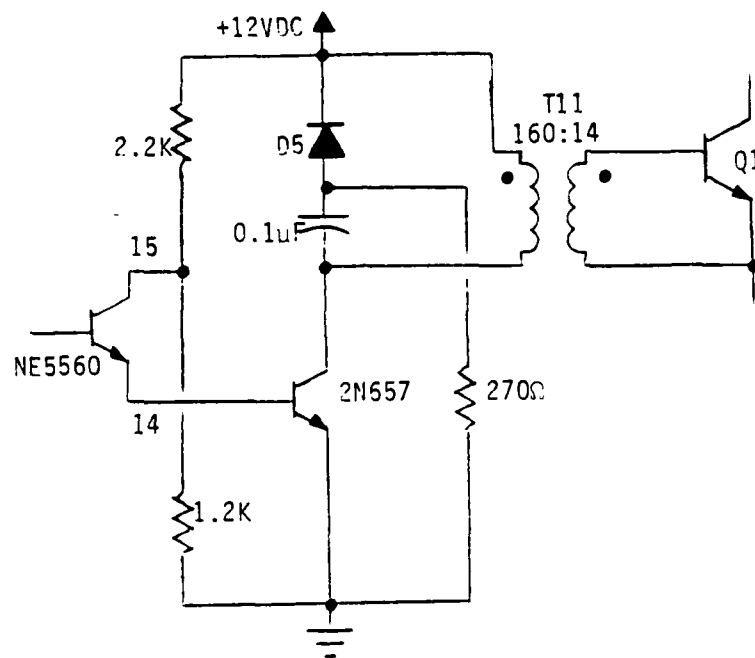
(a)

$$e = L \frac{di}{dt}$$

$$\Delta I = I_p = \frac{(12 - V_{ce\ sat})\ V\ (15\ \mu sec)}{0.9\ mH}$$

$$I_p = 173\ mA$$

$$I_{sec\ peak} = 1.98\ A$$
(b)

Figure 39 (a) Base Drive Circuit
              (b) Base Drive Current Calculations

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I. FEEDBACK STABILITY

For accurate regulation and stability a phase shift network is re-
quired on the feedback loop. This phase compensation is accomplished
with a capacitor in parallel with the feedback resistor for the error
amplifier (pin 3 and 4). For precise regulation, a large DC gain is
required. A gain of 200 was selected to keep the output within the
limited regulation specifications. This gain caused system oscillation
due to output ripple being amplified and injected into the pulse width
modulator. The insertion of the capacitor created a low pass filter
behavior. The capacitor selected was 0.47 μF electrolytic. This has
a high frequency roll off at 0.7 Hz. This resulted in a voltage gain
of 0.003 at the 20 KHz ripple frequency (-50.5 dB).

The capacitor value selected was a compromise between excellent loop
stability and rapid response or corrections in the pulse width for a
change in output voltage. The soft start feature kept the circuit from
voltage overshoot during turn on, which relieved part of the requirement
for a rapid response.

The nature of the load and changes in the load current must be known
when designing the power supply feedback loop.

J. SYSTEM EFFICIENCY

The completed power supply met the design specification. Appendix
A shows the detailed circuit. The overall system efficiency was greater
that required.

Table V. Operating Conditions and Efficiency

V_{in}	I_{in}	V_{out}	I_{out}	P_{in}	P_{out}	$\eta(\%)$
42	0.185	5.19	1.2	7.78	6.23	80.1
42	0.516	5.19	3.5	21.7	18.2	83.8
42	1.075	5.19	7.0	45.25	36.3	80.5
48	0.162	5.19	1.2	7.77	6.23	80.2
48	0.459	5.19	3.5	22.0	18.2	82.6
48	0.941	5.19	7.0	45.2	36.3	80.4
56	0.139	5.19	1.2	7.79	6.23	80.0
56	0.399	5.20	3.5	22.3	18.2	81.5
56	0.809	5.20	7.0	45.3	36.4	80.4

Minimum input voltage*

17.3	0.461	5.18	1.2	7.97	6.22	78.0
29.2	0.780	5.18	3.5	22.8	18.1	79.6
33.2	1.363	5.17	7.0	45.3	36.2	80.0

*Minimum operating voltage for system to remain in regulation. The system failed to self start for input voltages less than 39 volts, however once the system was operating, the regulator continued to function as the input voltage was decreased to the value shown.

The majority of the system loss was due to the voltage drop of the rectifier. The Schottky rectifier had a voltage drop of 0.7 to 1.0 volts, dependent on the current. This resulted in a power dissipation of seven watts under full load conditions. The rectifier was located on an aluminum heat sink that was separate from the rest of the system. The power transistor is a source of power loss due to a saturation volt-

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age of 0.7 volts during the on period and a leakage current of 0.1 mA. The remainder of the system losses were due to the integrated circuit, the driver stage, the power transformer and the filter capacitor. The filter capacitor loss resulted from the equivent series resistance and the ripple voltage.

Table VI.   Convertor Power Budgets

Power Input	45.2 W
Switching Transistor Loss	2.1 W
Rectifier Loss	4.9 W
Filter Capacitor Loss	0.3 W
Other Losses	1.6 W
Power Delivered to Load	36.3 W

IX. RADIO FREQUENCY INTERFERENCE SUPPRESSION

The most serious problem with switching mode power supplies is the radio frequency interference they generate. The fast rise times in switching high currents creates high frequency noise through the HF band. The RFI is produced by direct radiation and by the conduction of interfering currents through the input and output terminals. The suppression of direct radiation can be accomplished by shielding, however the suppression of conduction RFI requires more extensive design considerations throughout the design stages of the circuit.

A. MINIMIZING RFI BY CIRCUIT LAYOUT

Radio Frequency Interference can be produced by any stray capacitances or ground loops within the circuit. This includes inter-wiring capacitance, capacitance from the wiring to ground, and interwinding capacitance in all power transformers and pulse transformers.

1. Wiring Layout

Good wiring practice should be exercised as the printed circuit board is designed. The high current paths should be kept short and wide, minimizing the lead inductance and conductor resistance. When locating high currents switching devices on the printed circuit board, the forward and return traces should be kept close together to cancel magnetic fields. The feedback loop should be isolated from the switching waveforms.

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2. RFI produced by Heat Sink Current

The capacitance between a TO-3 transistor and its heat sink when a mica insulating washer is used is approximately 100pF. This capacitance produced RFI that was conducted through the heat sink which was usually connected to the ground lead.

There are several solutions to this problem that were developed. One solution was to enclose the heat sink within a shield, connecting the shield to ground. The heat sink was not connected to ground.

Another solution was to connect the heat sink to the emitter lead. The effect was to connect a collector to emitter capacitor. The currents within the heat sink were minimized thus RFI was reduced.

A third alternative was to construct a screen between the transistor and the heat sink. This screen is connected to the supply and insulated from the transistor and heat sink.

The final alternative was to place the transistor within the circuit such that its collector was common to the input line common. The heat sink was grounded and the transistor was connected to the heat sink. This method requires the use of an NPN transistor for positive ground input systems and a PNP transistor for negative ground input systems. This eliminates the collector to ground capacitance thus minimizing switching switching currents within the heat sink.

B. RFI PRODUCED BY THE POWER TRANSFORMER

1. Interwinding Capacitance as a Source of RFI

Another path for unwanted RFI currents is the interwinding capacitance in the power transformer. A transformer wound with primary and

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secondary conductors alternating in layers to increase magnetic coupling, also increases the interwinding capacitance. The value of the capacitance is usually on the order of 100 pF or less. It extremely difficult to isolate these currents.

One method that has been used is the use of a thin copper screen or shield that was placed between the primary and secondary windings. The screen was then connected to the supply line, returning any currents generated. On higher power tranformers two screens may be used. One screen was connected to the supply line and the other screen was connected to the output line. In layered transformers multiple screens would be required.

2. RFI Produced by the Transformer Core

The ferrite transformer core may be a source of RFI currents if the capacitance between the core and primary is substantial. The RFI current is generated in the core and flows to the output via the hold down clamp to ground. This can be minimized by insulating the core from the ground and tying it to the supply line.

C. RFI GENERATED BY THE OUTPUT RECTIFIER

The use of fast recovery output rectifiers generates a large RFI current that flows to the output when the rectifier snaps off.

The typical fall time of the reverse current at snap off is 10 ns. This high frequency current surge can result in ringing, depending on

transformer leakage inductnace, wiring inductance, and stray capacitance
This snap off ringing is most pronounced at 50 MHz to 100 MHz. This
ringing is minimized by using soft recovery diodes or by connecting a
small capacitor in series with a resistor across the diode. A value of
0.01 µF with a 10Ω resistor was used in parallel with a VARO VSK 1530
Schottky rectifier for the 5 volt, 7 ampere output with good results. [21,22]

D. RFI CONDUCTED TO THE LOAD AND LINE

The most severe RFI problems are the RFI currents that are on the
line or on the output bus. The RFI on the output bus will disrupt the
operation of the system when it is used with audio or radio equipment
that operate in those frequency ranges.

The most difficult RFI to trace is when the RFI currents are gener-
ated on the line side. The currents can back up the input line and
interfere with any equipment that operates from the ac line. When com-
plex systems are in use, it is often difficult to trace the problem to
the offending unit. Typical safe RFI values are shown in Table VII as
measured by the circuit shown in Fig. 40

Table VII. RFI Line Isolation Requirements [23]
RFI voltages cannot exceed the following ranges (referenced to 1 V)

+84dB	150 KHz to 200 KHz
+80dB	200 KHz to 500 KHz
+70dB	500 KHz to 30 MHz
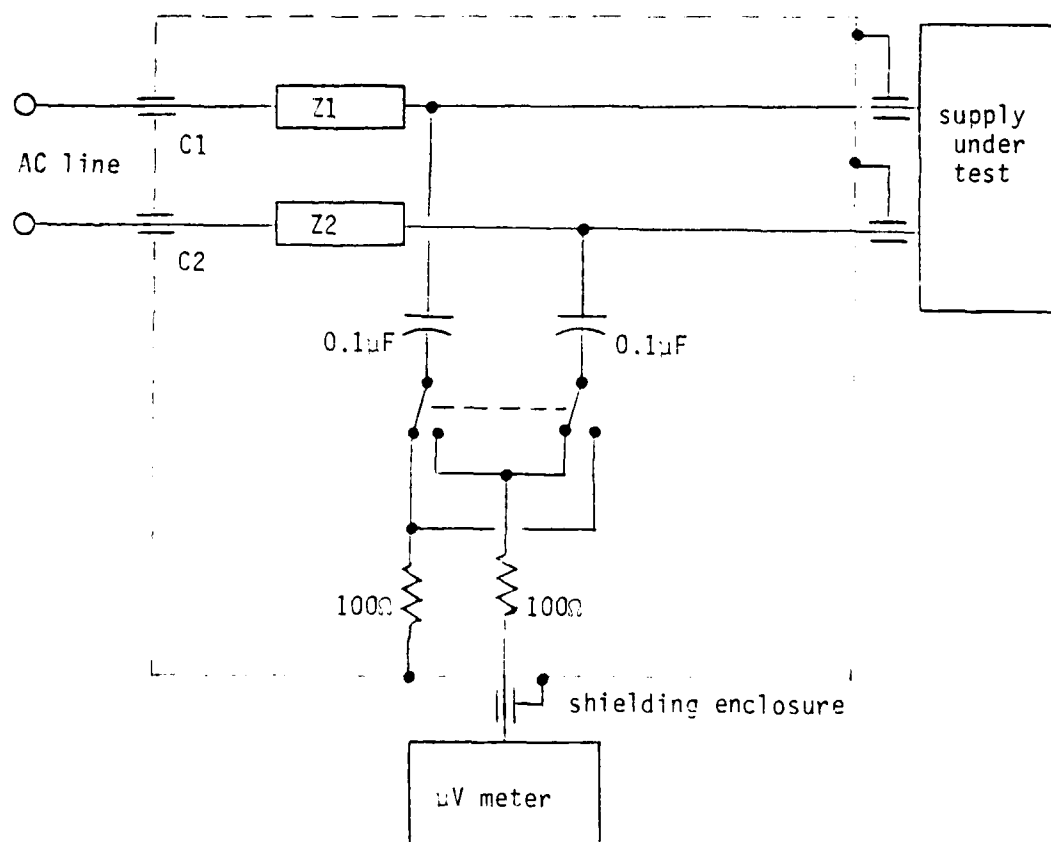
C1, C2 - Coaxial lead-through capacitors = 0.5µF

Z1, Z2 - 500µH

Figure 40. Line Isolating Network for Measuring RFI

E. PULSE SYNCHRONIZATION

The simplest method of eliminating RFI problems when the supply is connected to systems that use a CRT or a master clock operating in the

15-30 KHz range is to synchronize the regulator oscillator with the horizontal oscillator of the CRT or the master clock. This permits the power supply oscillator to saturate the power transistor during the period when it will not interfere with the operation of the system for which it provides power.

For the pulse synchronization technique to operate, a control signal must be available from the CRT or system clock and that signal must be connected to the regulator chip. There are several regulator circuits that have a synchronization or modulation input available. Not all regulators have this provision, hence an interface circuit must be designed if one of these regulator circuits are chosen.

The pulse synchronization technique does not limit the RFI currents that are injected into the ac line. An isolation filter must be provided when there is a possibility that the supply may cause interference in other systems that share the ac line.

X. CONCLUSION

It has been shown that switching mode power supplies are efficient DC to DC convertors. Convertors used either a conventional step-down transformer and a rectifier/filter section connected to a switching regulator, or a high voltage rectifier with the regulator operating from the line voltage.

The switching mode power supply is a flexible subsystem that can provide an inverted polarity output voltage. The switcher is a very compact device that minimizes the heat problems that plague linear supplies.

Two practical switching supplies were constructed and tested. The supplies were able to meet the design specifications on all counts. The supplies were different from each other and from other switching supplies presently manufactured. The two supplies were different in several aspects, showing the methods of switched mode power supply operation and regulation techniques.

The complexity of the switching mode supply is greater than linear regulators, thus the cost for switchers is usually greater and the failure rate would be expectedly higher. The cooler operating temperature may eliminate the need for a fan and extensive heat sink, making the cost reasonable for some applications. The cooler operation also improves the operating life of most components, thus the failure rate may be closer to that of a linear system then what was first expected. The

operating life of the supplies was not tested, thus no data on failure rates were generated.

Table VIII. give the disadvantages and advantages of switching mode supplies versus linear supplies.

The RFI question should be addressed early in the design stage. For specialized systems, sensitive equipment or low signal level amplifiers/ receivers at antenna sites, the presence of any measurable level of RFI may preclude the use of a switching mode power supply anywhere near the site. For most other requirements the use of a switcher can be considered.

The largest advantage of the switcher is its high efficiency. The high efficiency allows its use in battery power devices or in other systems where a limited or expensive source of power is available.

Table VIII. Linear and Switching Mode Advantages and Disadvantages

LINEAR SUPPLY	SWITCHING MODE SUPPLY
Larger, due to heatsink and 60 Hz transformer	Small size
Heavy	Light weight
Low parts count	Complex, more components
Inefficient $\eta<50\%$	High efficiency $\eta\approx80\%$
Low ripple (60 Hz)	Higher ripple ($\approx$ 20 KHz)
Fast transient response	Slow transient response
Low Cost	Moderate to high cost (Decreasing)
Low system noise	RFI
Limited on chip protective circuits	On chip protective circuits
Simple design and implementation	More difficult to design and construct
May require cooling fan	Usually does not require cooling fan
Switches required for power up/down	TTL programable power up/down
Voltage overshoot	Soft start
Single output	Multiple outputs
Difficult to incorporate in UPS systems without many additional circuits	Often incorporated in UPS systems. (Uninterruptible Power Supplies)

APPENDIX A 5 V, 35 W DC to DC CONVERTOR

Parts List

Capacitors

C1, C4 - 6.8 μF, 50 V, electrolytic
C2 - 0.02 μF, 100 V, ceramic disc
C3, C5, C6, C7, C19 - 0.1 μF, 100 V, ceramic disc
C8 - 0.47 μF, 50 V, electrolytic
C9, C12 - 20 μF, 50 V, electrolytic
C10 - 0.0033 μF, 50 V, ceramic disc
C11 - 5700 μF, 6.3 V, electrolytic, low ESR type
     Mallory VPR572U6R3J2L or equiv.
C13 - 0.002 μF, 100 V, ceramic disc
C14 - C17 - 8 μF, 150 V, electrolytic
C18 - 20 μF, 15 V, electrolytic, optional filter cap for -5 V output

Diodes

D1, D3, D5, D6, D9 - General Purpose Silicon Diodes, 100mA, 100 PIV
D2 - Switching Diode, 3 A, 200 PIV, RG3D or equiv.
D4- Zener Diode, 14 V, 100 mW, 1N4744A
D7 - Silicon Diode, 100mA, 25 PIV
D8 - Schottky Fast Recovery Rectifier, 1N5827 or Varo VSK1530
D10 - Zener Diode, 5.6 V, 100mW, 1N4734A
D11 - Silicon Rectifier, 1A, 50 PIV, optional for -5 V output

Resistors (all carbon composition 10% unless otherwise noted)

R1, R11, R13, R14, - 1, 2K, ¼W
R2- 2.7 K, ¼W
R3-190Ω, 2W
R4, R9 - 18 K, ¼W
R5 - 270Ω, ¼W
R6 - 2.2 K, ¼W
R7 - 20 K, ¼W
R8 - 12 K, ¼W
R10 - 470 K, ¼W
R12 - 500Ω, 1 turn trimpot, Bourns 3386-T-1-501 or equiv.
R15 - 3.3 K, ¼W
R16 - Dropping Resistor for -5 V output
R17, R18 - 150 K, ¼W

Transistors and Integrated Circuits

Q1 - Switching Transistor, Silicon NPN, TO-3, GE type D64VE3

Q2- Q3 - Silicon NPN, TO-5, (Vce at least 60 V), 2N657
IC1 - Timer, National LM555CN, 8 pin DIP
IC2 - Switched-Mode Power Supply Control, Signetics NE5560N, 16 pin DIP
Q4 - Silicon NPN, 2N3704

Inductors and Transformers

L1 - 20 mH, Core 55120-A2-GR (μ=60), 530 turns of # 35 AWG

T1 - Pulse Transformer, Base Drive Circuit: Core 55120-A2-GR, 0.9 mH
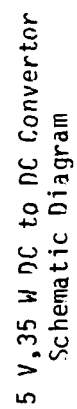   Primary   (1-2): 160 turns   #26 AWG
                      Secondary   (3-4):   14 turns   #24 AWG

T2 - Power Transformer, 0.4 mH
                      Primary   (1-2):   45 turns   #16 AWG
                      Secondary   (3-4):   15 turns   #16 AWG (bifilar wound)
                      Secondary   (5-6):   30 turns   #22 AWG

T3 - Starter Transformer, Core 55120-A2-GR, 0.53 mH
                      Primary   (1-2): 123 turns   #28 AWG
                      Secondary   (3-4): 123 turns   #28 AWG

5 V, 35 W DC to DC Convertor
Schematic Diagram

90
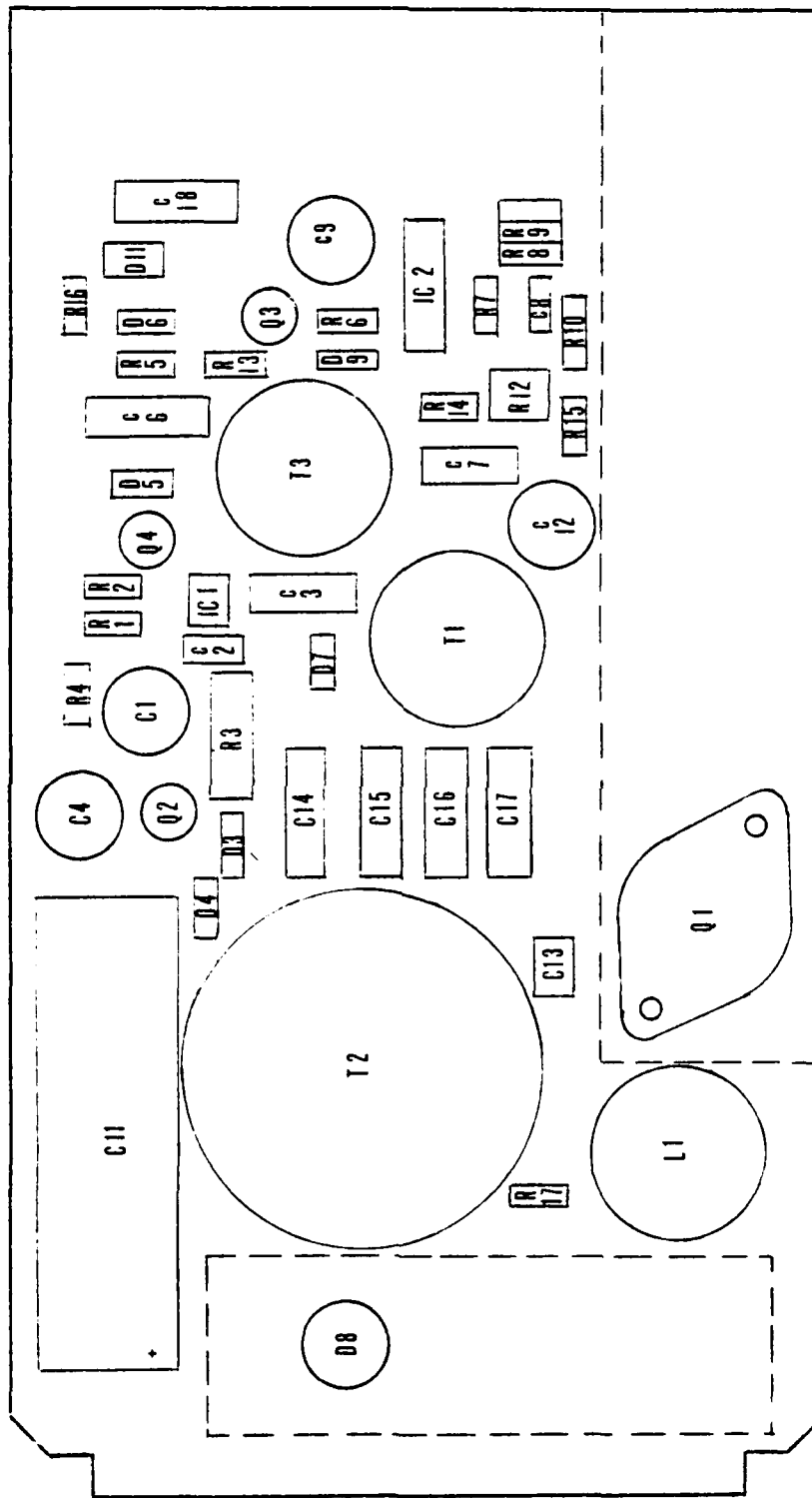
5 V, 35 W DC to DC Convertor

Component Layout

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