

# A NEW APPROACH TO SWITCHING REGULATORS

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This article describes a 24-Volt, 3-Ampere switching mode supply. It operates at 20 kHz from a 120 Vac line with an overall efficiency of 70%. New techniques are used to shape the load line. The control circuit uses a quad comparator and an opto coupler and features short circuit protection.



**MOTOROLA Semiconductor Products Inc.**

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## INTRODUCTION

The function of a switching regulator is to convert an unregulated dc input to a regulated dc output. The method used to accomplish this function differs dramatically from the conventional series pass regulators. In a switching regulator, the power transistor is used in a switching mode rather than a linear mode. Efficiencies are usually 70% or better which is about double that of the series pass regulator. High-frequency switching regulators offer considerable weight and size reductions and better efficiency at high power over conventional 60 Hz transformer-coupled, series regulated power supplies.

However, a debate continues on the value of switching type supplies as contrasted with conventional series pass regulators. One argument, frequently advanced when discussing switching regulators, is that the electro-magnetic interference (EMI) problems overcome most of the gains in efficiency. Eliminating high-frequency transients, not present in the linear case, adds a new dimension to the design of switching regulators. Controlling this undesired output actually requires very little bulk weight. Basically the necessary techniques come from good RF practice. Layout and lead length play significant roles when designing effective switching circuits.

EMI can be reduced to acceptable levels giving compact, fast and efficient control of power. Switching regulators have found wide use in aerospace and portable applications where power is expensive. Low loss ferrite cores for transformers and chokes, the use of high permeability magnetic alloys for shielding, and the wide range of miniature semiconductor and IC devices for the switching and regulation circuitry have contributed to the success of switching type supplies. The block diagram of a switching regulator is shown in Figure 1. This circuit regulates by switching the

series transistor to either the ON or OFF condition, the duty cycle determining the average dc output. Duty cycle is adjusted in accordance with a feedback proportional to the difference between the dc output and a reference voltage.

Switching is usually a constant frequency just above the audible range (20 kHz typical), although some varieties show variable frequency with changing line and load. Higher frequencies are generally less efficient since transistor switching losses and ferrite core losses increase. Low frequency unibase transistors ( $f_T$ -200 kHz) are sufficient for series pass regulators; switching regulators must use epibase and triple diffused devices ( $f_T$ -4 MHz) to operate efficiently. For the ultimate in efficiency, even higher frequency annular and double diffused devices can be used, ( $f_T > 30$  MHz). Darlington transistors are also used in these switching mode supplies; they can provide an overall gain in efficiency through lower base drive requirements even though their saturation voltages are higher than single chip devices.

A fast recovery rectifier or Schottky barrier diode is used as the free wheeling clamp diode to keep the switching transistor load line within SOA limits and to increase efficiency at these frequencies. Other products used in these supplies are shown in Table 1 below.

**TABLE 1**  
Semiconductors Used in Switching Regulators

POWER DEVICES	CONTROL IC's	
	DIGITAL	LINEAR
Power Transistors	Gates Flip-Flops	Op Amps Comparators
Rectifiers	Monostable Multivibrator	Timers Regulators

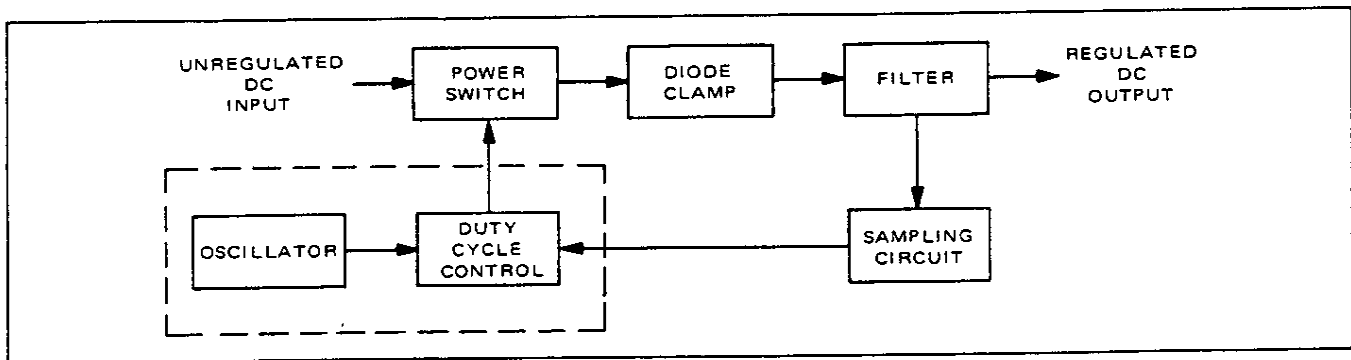


FIGURE 1 - SWITCHING REGULATOR

Circuit diagrams external to Motorola products are included as a means of illustrating typical semiconductor applications; consequently, complete information sufficient for construction purposes is not necessarily given. The information in this Application Note has been carefully checked and is believed to be entirely reliable. However, no responsibility is assumed for inaccuracies. Furthermore, such information does not convey to the purchaser of the semiconductor devices described any license under the patent rights of Motorola Inc. or others.

## TYPICAL CIRCUITS

Various circuit configurations have been used in switching regulator designs. All have the following common elements:

1. Switching transistor
2. Clamp diode
3. LC filter
4. Logic or control block.

Representative circuits are shown in Figure 2. It should be noted that none of these circuits offers isolation between the line and load. (However, the one transistor design highlights simplicity and economy.) It is usually desirable to have at least one line in common with the input and output to reduce ground loops. The one line approach also determines whether the output voltage will be considered positive or negative. However, by definition this is academic since most circuits will operate from either supply since the input and output grounds are usually isolated. This is suitable for the most general use of switching regulators which power industrial controls and other logic circuits. In circuits 2(a) and 2(b) the logic

operates from the load voltage. These circuits are not self-starting and provisions must be made to operate from the line during starting and short circuits. In circuits (c) and (d) the logic operates continuously from the line and is isolated from the load. The sense and feedback elements must be electrically isolated for this reason. An opto coupler or an equivalent device is ideal for this purpose.

Circuits (b) and (d) are generally used in line operated supplies because economical high voltage NPN transistors are available whereas PNP types are not. Of these two, circuit (d) is the most popular because the logic is tied directly to the series switch, and switching can be much more efficient.

Driver transformers are also used in many designs to interface between the logic and switching transistor. In such a case, there are no circuit constraints on the transistor type, and it may be an NPN or PNP device.

## THEORY OF OPERATION

The high efficiency of switching regulators is a result of operating the series transistor in a switching mode. When the transistor is switched ON, full input voltage is applied to the LC filter; when it is OFF, the input voltage is zero. With the transistor turned on and off for equal amounts of time (50% duty cycle), the dc load voltage will be half the input voltage. The output voltage ( $V_O$ ) will always equal the input voltage ( $V_{in}$ ) times the duty cycle ( $D$ ) as follows:

$$V_O = D V_{in}$$

Varying the duty cycle will therefore compensate for changes in the input voltage; this technique is used to obtain a regulated output voltage.

Repetitive operation of the switching transistors at a fixed duty cycle produces the steady state waveforms shown in Figure 3. With the switch closed, inductor current ( $I_L$ ) flows from the input voltage ( $V_{in}$ ) to the load. The difference between the input and output voltage ( $V_{in}-V_O$ ) is applied across the inductor. This causes  $I_L$  to increase during this time.

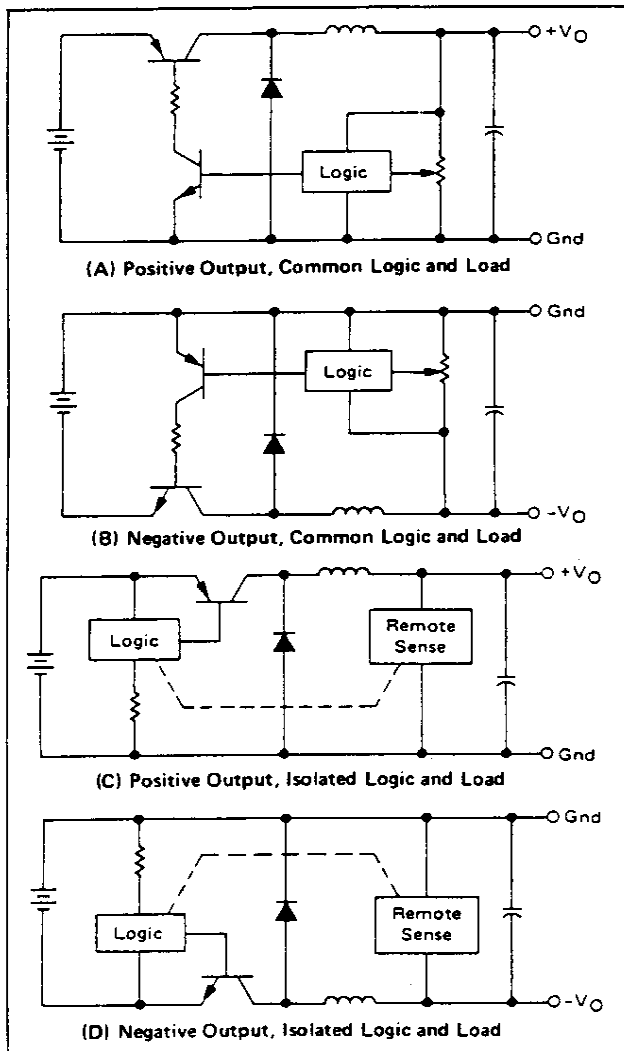


FIGURE 2 - SWITCHING REGULATOR CIRCUIT VARIATIONS

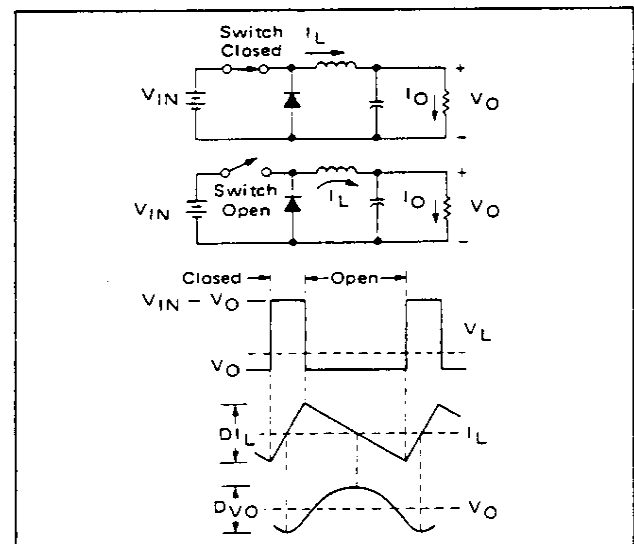


FIGURE 3 - THEORETICAL WAVEFORMS

With the switch open, stored energy in the inductor forces  $I_L$  to continue to flow to the load and return through the free wheeling diode. The inductor voltage is now reversed and is approximately equal to  $V_O$ . During this time,  $I_L$  decreases.

The average current through the inductor equals the load current. Since the capacitor keeps  $V_O$  constant, the load current ( $I_O$ ) will also be constant. When  $I_L$  increases above  $I_O$ , the capacitor will charge and when  $I_L$  drops below  $I_O$ , the capacitor will discharge. These waveform inflection points are indicated in Figure 3. The end results of steady state operation are as follows:

1. The average inductor voltage will be zero but a wide variation from  $(V_{in}-V_O)$  to  $V_O$  will be experienced.
2. The dc current flowing through the inductor will equal the load current. A small amount of sawtooth ripple will also be present.
3. The dc voltage on the capacitor is equal to the load voltage. A small amount of ripple (quasi-sine wave) will also be present here.

Transient operation must consider changes in  $V_{in}$  and  $I_O$ . Input voltage changes are automatically compensated for by appropriate duty cycle variations in a closed loop system. Input regulation and ripple rejection are dependent on loop gain but are generally adequate as indicated later.

Changes in  $I_O$  are more difficult to compensate for and load transient response is generally poor. Changes in  $I_O$  are compensated for with temporary duty cycle changes. For example, a change in load from half to full will result in the following:

1. Duty cycle increases to its maximum (the transistor may just stay ON).
2. The inductor current takes many cycles to increase to its new dc level.
3. Duty cycle returns to its original value.

## DESIGN

The circuit in this article was designed to supply a regulated 24 Vdc output from a 120 Vac line. Typical load variations are from 1.5 to 3 Amperes and the line from 100 to 140 Vac. In addition to providing good regulation, the circuit is efficient and has short circuit protection. The design of this regulator utilized the functional blocks shown in Figure 4. Since it is line operated, an input rectifier and filter are required to convert the incoming ac to dc. The power switch uses a

high voltage transistor operating at a switching frequency of 20 kHz to control the pulse width of the voltage applied to the LC output filter.

The control portion of this design uses an oscillator, a "one shot" comparator, and a feedback block which also includes an opto coupler. The oscillator generates the fixed frequency of operation. The resulting output forms a clock pulse for the comparator. The second input to the comparator is derived from the feedback amplifier. This amplifier senses the output voltage and controls the duty cycle of the "one shot" comparator. The output of the comparator is then fed directly to the power switch to complete the loop. Additional details regarding the design and operation of these blocks are provided in the following sections.

## INPUT RECTIFIER AND FILTER

The input rectifier and filter consists of a bridge rectifier and filter capacitor. To improve the efficiency series limiting resistance is not specifically used; starting current surges are higher because of this and a 12 Ampere rectifier is therefore required even though the average output current is less than one Ampere. A 150  $\mu$ F capacitor was used to limit the input ripple to under 20 V peak-to-peak with a full load.

## OUTPUT FILTER

The output filter is also quite simple as shown in Figure 5. Calculations of the values for L and C are more involved but quite conventional. The appropriate design information is also shown in Figure 5. To design the filter

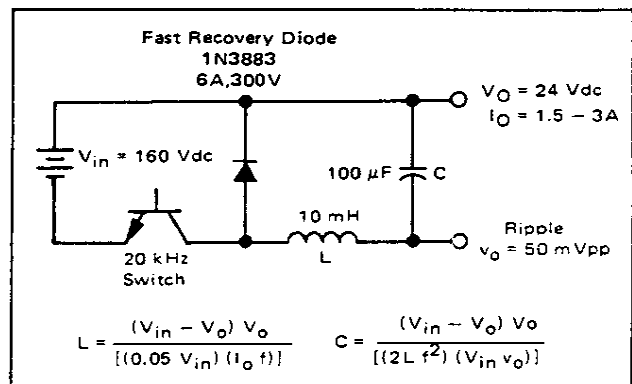


FIGURE 5 – OUTPUT FILTER AND DESIGN SPECIFICATIONS

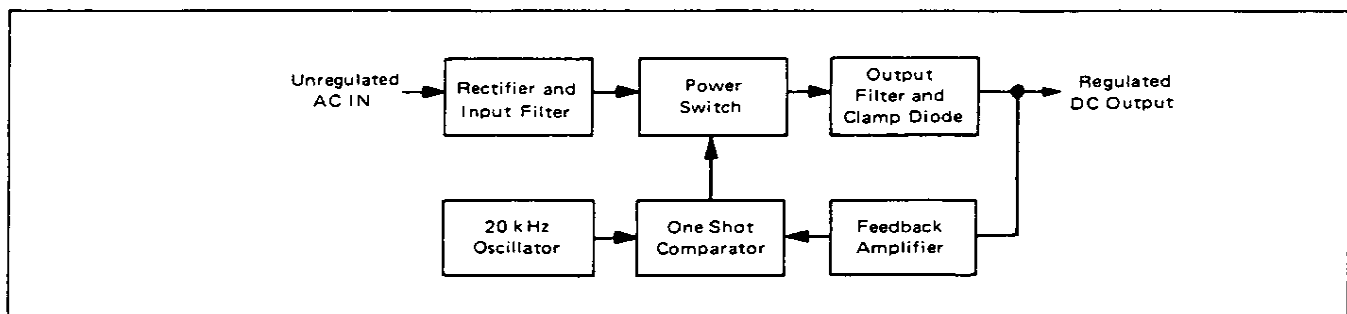


FIGURE 4 – SWITCHING REGULATOR BLOCK DIAGRAM

approximate values for the following must be known.

1. Input voltage,  $V_{in}$
2. Output voltage,  $V_o$
3. Frequency,  $f$
4. Output current,  $I_o$  (and range)
5. Output ripple  $v_o$  (peak-to-peak)

A general "rule of thumb" is to restrict current variations in the choke ( $L$ ) to 10% of the load current. Variations are restricted to 5% in this case since a regulation range of load current of at least a 2 to 1 ratio is required. Smaller values of inductance could be used to improve transient response as long as the choke does not run dry (allowing choke current to drop to zero during each alternation). A value for  $L$  can be found using:<sup>(1)</sup>

$$L = \frac{(V_{in} - V_o) V_o}{0.05 I_o V_{in} f} \text{ (for } \Delta I_L = 5\% I_o \text{)}$$

$$L = \frac{(160 - 24) 24}{0.05 (3) 160 (20 \text{ k})} = 6.8 \text{ mH.}$$

The nearest standard choke available in this design case was a 10 mH air core choke. A standard Triad C 58U or an equivalent custom designed ferrite core may be used if a small air gap is provided in the latter case to prevent saturation at currents up to 5 Amperes dc; this is the short circuit limit.

To keep  $v_o$  below 50 mV, a value for  $C$  can be found using:<sup>(1)</sup>

$$C = \frac{(V_{in} - V_o) V_o}{(2L) f^2 V_{in} (v_o)}$$

$$C = \frac{(160 - 24) 24}{2 (10 \text{ mH}) (20 \text{ kHz})^2 160 (50 \text{ mV})} = 51 \mu\text{F.}$$

The dissipation factor of most electrolytic capacitors is too high for this application; because of this, two 50  $\mu\text{F}$  capacitors were used to obtain the desired performance. A more expensive stacked foil electrolytic or solid tantalum capacitor may also be used. High frequency bypass capacitors and an additional high frequency LC filter are used in similar designs to prevent switching spikes from reaching the load. This approach was not required in this design because of load line shaping which will be described later. Operation at 20 kHz requires a fast recovery diode. A 6 Ampere, 300 Volt, 1N3883 was chosen to handle up to 5 Ampere free-wheeling currents and peak line voltage.

## POWER SWITCH

The power switch includes a push-pull driver which provides the interface between the integrated circuit drive signal and the actual power switch, a high voltage NPN power transistor. Power to this stage and the logic is provided by a 12 Volt step-down transformer. The IC logic is in the position indicated earlier in the discussion of Figure 2(d). The main advantage of this layout is that the logic ground is common to the emitter of the power switch. Because of this, conventional speed-up capacitors can be used to couple the drive signal to the base terminal to improve switching speeds. This is especially important in this particular design. Because of the large step-down in voltage and the high switching frequency, ON times will be very short. The minimum switching time is found using the peak  $V_{in}$  of 200 V (high line) instead of 160 V.

The earlier discussion of switching regulator theory indicated the duty cycle would be approximately:

$$D = V_o/V_{in} = 24/200 = 12\%.$$

At 20 kHz, the ON-time will be 12% of 50 microseconds or approximately 6  $\mu\text{sec}$ . With simple resistive termination of the base emitter junction, storage time alone can be as much as 10  $\mu\text{sec}$ . For this reason, a special drive circuit and a high frequency 250 volt, 8 Ampere, 2N6306 power transistor were required for this application. This device has good saturation voltages at 3 Amperes with excellent switching speeds. The drive circuit shown in Figure 6 does the job of switching the 2N6306 at 3 Amperes and 20 kHz from the 120 Vac line. To do this, two unique design innovations were used.

1. An artificial negative bias supply was created from the single positive supply available to improve fall time.
2. Current limiting was added to the base current to limit overdrive and reduce storage time.

Turn-off of the power switch is accomplished by forcing the IC to a logic low. This turns on the 2N6034 and creates a path for reverse base current. The 10  $\mu\text{F}$  capacitor which is used as a reverse bias supply then removes stored base charge from the 2N6306 and forces it to turn-off. Reverse base current flows for about 2  $\mu\text{sec}$ . The diodes are added to prevent the capacitor from discharging into the 10 ohm resistor once the base is

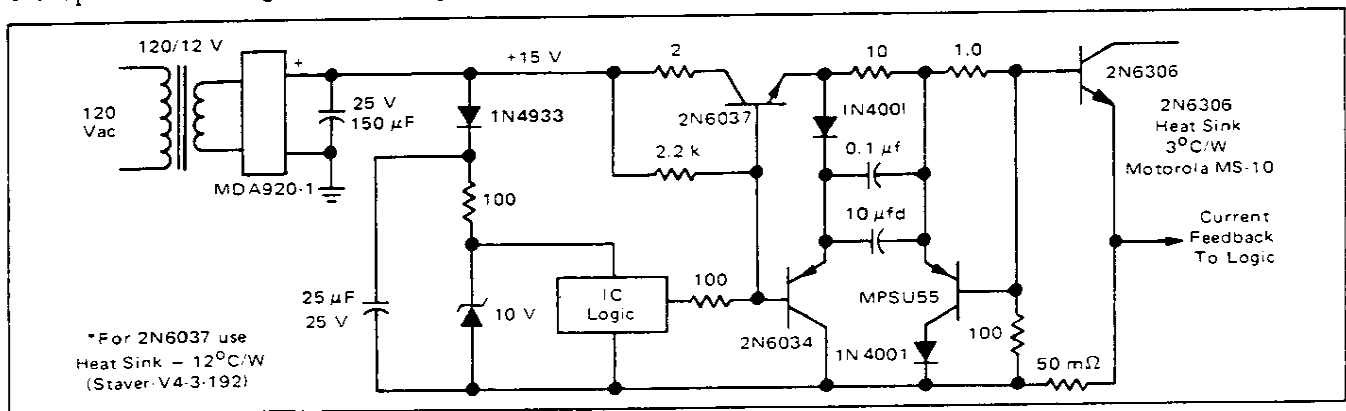
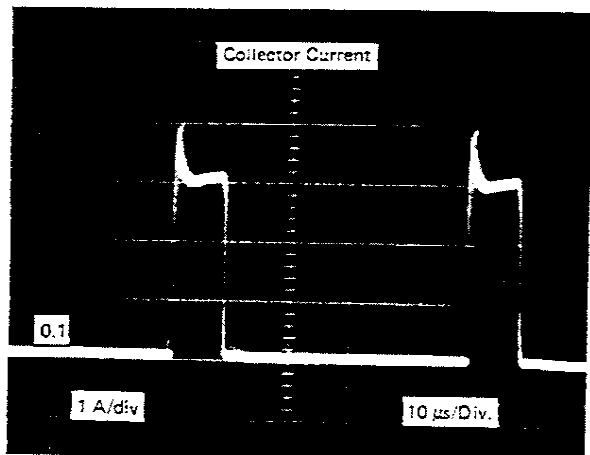


FIGURE 6 - POWER SWITCH AND LOW VOLTAGE SUPPLY

cleared. This minimizes the amount of charge that must be replaced during each alternation.

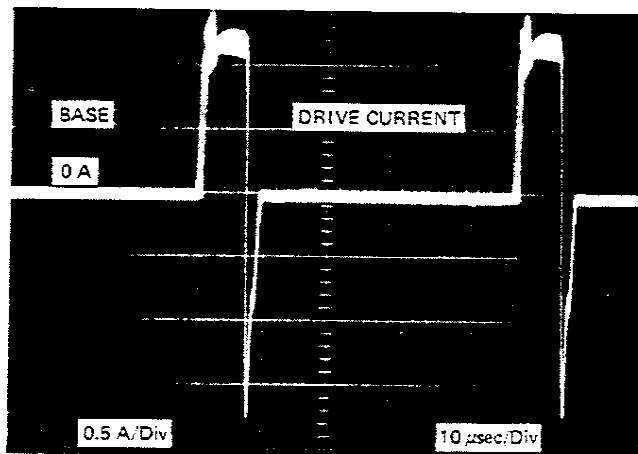
When the logic goes high, the 2N6037 turns on and forward base current flows through the 10 ohm resistor turning on the 2N6306. The 2N6034 and 2N6037 are both plastic Darlington power transistors. However, the 2N6037 should be used with an appropriate heat sink since it supplies both forward base drive for the device and recharge current for the 10  $\mu$ F capacitor. With this

high gain Darlington, recharge takes approximately 2  $\mu$ sec. Most of the recharge current is bypassed around the 2N6306 base by the MPSU55; this device limits forward base drive to about 1 Ampere. This particular drive level was chosen to keep storage time low and at the same time ensures that the 2N6306 will remain saturated even under short circuit conditions. The actual operation waveforms of the power switch and drive circuit are shown in Figures 7 and 8, respectively.

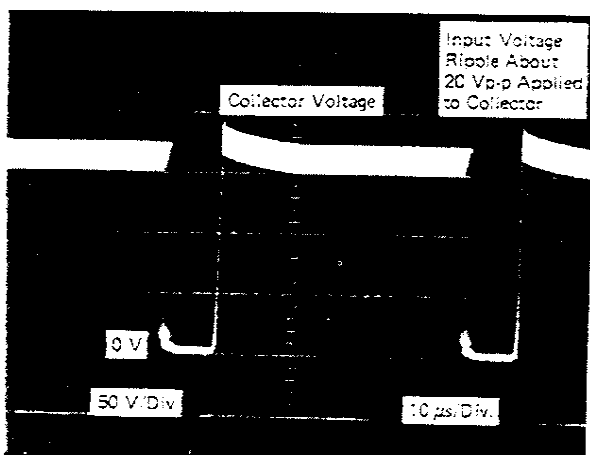


Collector Output Current

Both Measured  
With 120 Vrms Input  
and Output  
24 V 3 A



Measured  
With 120 Vrms Input  
and Output  
24 V 3 A



Collector Voltage

FIGURE 7 – POWER SWITCH WAVEFORMS

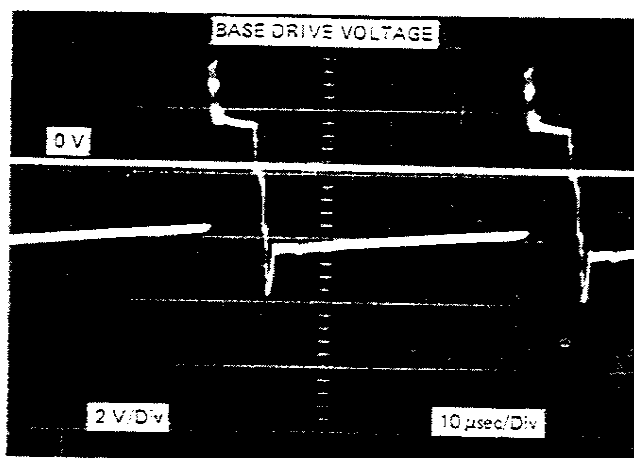


FIGURE 8 – DRIVE CIRCUIT WAVEFORMS

## FEEDBACK AMPLIFIER

Because the load ground and logic ground are isolated, it is necessary to use a 4N28 opto coupler for the feedback amplifier. The schematic is shown in Figure 9. The 1N5290 is a 0.5 mA current limiting diode and causes the voltage drop across the 39 k sense resistor to remain fixed at about 20 V. This unique design causes changes in load voltage to appear directly at the base of the 2N5088 and results in sensing without attenuation. Load voltage changes produce a linear change in LED current and a proportional voltage change at the sense terminal of the logic circuitry.

The 0.5 mA diode is considered to be a zero temperature coefficient unit. Also, the output level of the opto coupler of 1 mA is chosen to minimize its temperature dependence. A photo Darlington coupler was tried for increased gain; however, it introduced low frequency oscillations because of its slow speed and thus, could not be used.

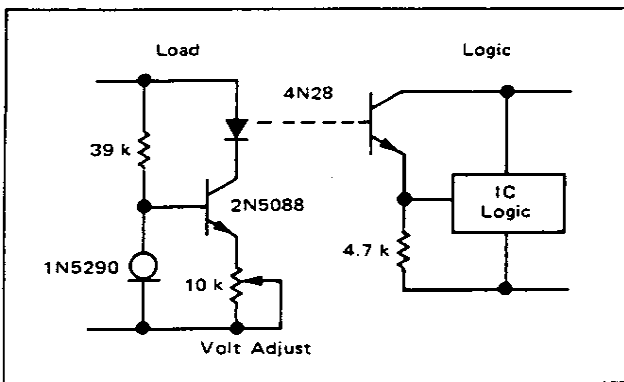


FIGURE 9 – FEEDBACK AMPLIFIER WITH OPTO COUPLER

## OSCILLATOR, COMPARATOR, AND CURRENT LIMIT CIRCUITS

In addition to the feedback amplifier, the control circuit for this switching regulator also requires an oscillator, comparator, and current limiter as part of its control circuitry. These last three functions are all made possible by the use

of a single IC chip, the MC3302 quad comparator.

The schematic diagram of the oscillator, comparator, and current limit circuits is shown in Figure 10. Comparator 1 is used as a 20 kHz oscillator. It supplies a sawtooth output which operates between the voltage limits defined by the 100 k positive feedback resistor and the logic supply voltage. The second comparator takes this output and compares it to the feedback signal to produce a variable duty cycle output pulse for the power switch. The timing diagram in Figure 11 illustrates how this happens. In normal operation, the feedback signal is a constant dc voltage which is between the limits of the oscillator sawtooth. When the sawtooth exceeds the feedback threshold, comparator 2 switches to a high output level. The comparator is reset when the sawtooth drops back below the feedback signal. In actual practice, the comparator output pulse is delayed by 2  $\mu$ sec because of internal propagation time. This phase shift has no effect upon regulation. Variations in the output and feedback signals will still produce compensated changes in the power switch pulse width.

The remaining two comparators in Figure 10 are used to initiate current limiting action. Comparator 3 senses the over current and triggers comparator 4 which is used as a "one shot" multivibrator. The zero input voltage threshold is derived from the reference voltage of 0.24 V from comparator 3. Power dissipation in the 50 m $\Omega$  sense resistor is therefore kept at a minimum. With this combination of reference voltage and current resistor,

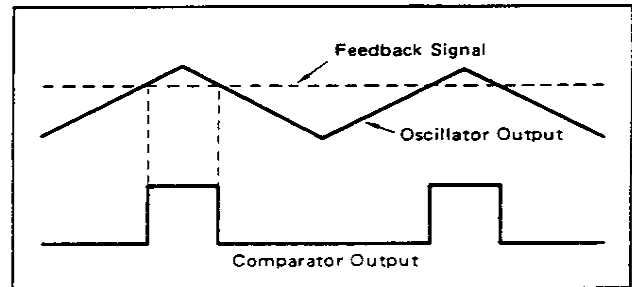


FIGURE 11 – TIMING DIAGRAM

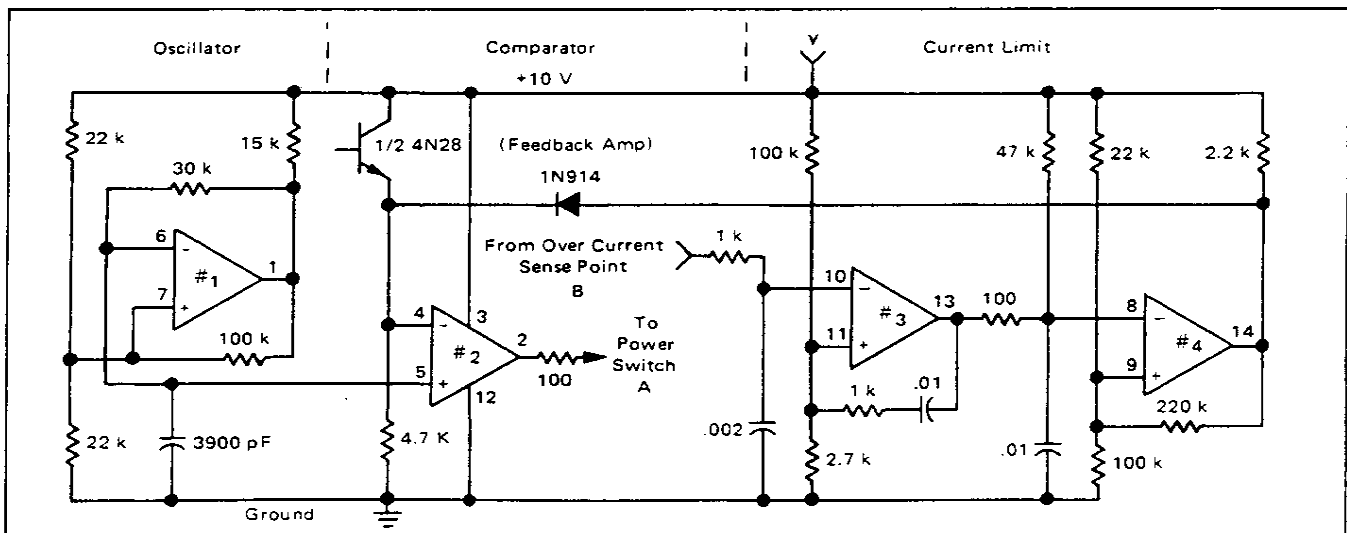


FIGURE 10 – CONTROL CIRCUIT WITH MC3302 QUAD COMPARATOR

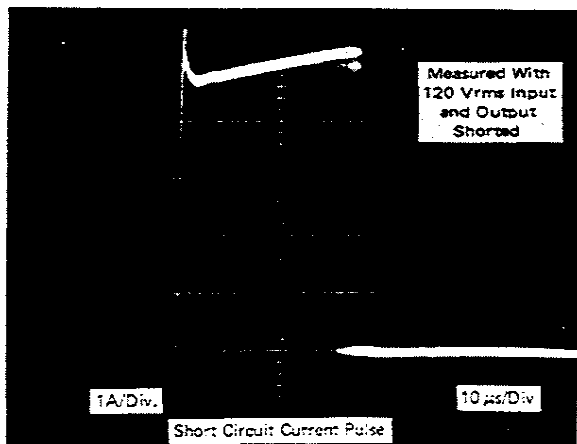
current limiting occurs at approximately 4 Amperes. When this level is reached (due to overload or short circuit), the "one shot" capacitor is discharged and a millisecond time interval is initiated. Positive ac feedback is used to ensure that this capacitor does completely reset. An RC filter is also provided to prevent inductive transients from accidentally shutting the regulator down.

While the "one shot" device is recharging, the output of comparator 4 remains high to simulate excessive output voltage at this feedback sense point (comparator 2); drive pulses are completely inhibited during this down time. Comparator 4 is reset and drive pulses continue when the "one shot" device reaches an 8 volt threshold. Positive feedback to this reference prevents circuit oscillations as the threshold is approached. If the short is removed, the switching regulator will automatically reset into a full load. If the short remains, operation (with 30  $\mu$ s 4 Ampere pulse) continues at an audible 1 kHz rate set by the "one

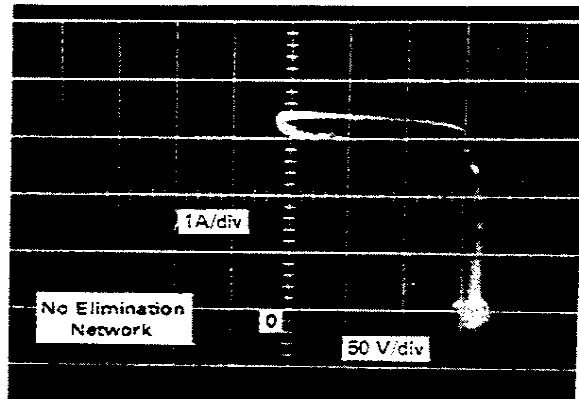
shot" multivibrator. The short circuit collector currents are shown in Figure 12.

### LOAD LINE SHAPING

Load line shaping can be used to improve reliability and reduce EMI.<sup>(1)(2)</sup> Shaping is basically done by using reactive elements to absorb what would normally be switching losses. Because resistors are used to dissipate the power stored in these reactive elements, there is no appreciable improvement in overall circuit efficiency. However, because safe operating area (SOA) stresses on the switching transistor are reduced, cooler and more reliable operation results. Illustrations of the load line before and after are shown in Figure 13 and 14. A secondary but very desirable effect can also be noticed at the output. Switching spikes, due to the abrupt recovery of the fast recovery diode and fast switching speed of the power transistors, are reduced from a 0.6 Volt to a 0.1 Volt peak. Because



Short Circuit Current Measured at the Collector with Current Probe



Load Line

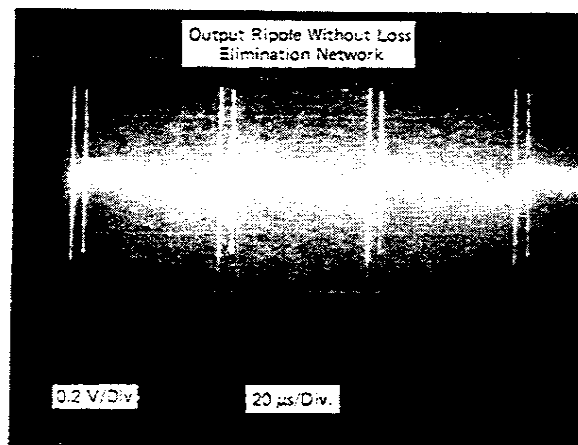
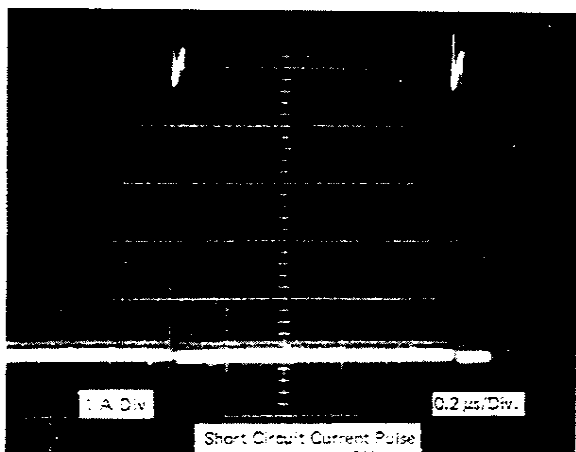


FIGURE 12 – SHORT CIRCUIT CURRENTS

FIGURE 13 – BEFORE LOAD LINE SHAPING



these spikes are reduced to acceptable levels, additional high frequency output filters are not required. Calculations of the power losses reveal the following:

TABLE 2 – Switching Power Losses (Watts)

Power	Transistor Only	Transistor with Snubber		Total
		Transistor	Snubber	
turn on	2.4	0.8	1.8	2.6
turn off	3.0	1.0	1.9	2.9
Total	5.4	1.8	3.7	5.4

The calculations show the dramatic improvement. The remaining power loss of approximately 0.4 watt, due to base and collector saturation voltages of 1.0 and 0.5 Volt respectively, is unaffected by these changes.

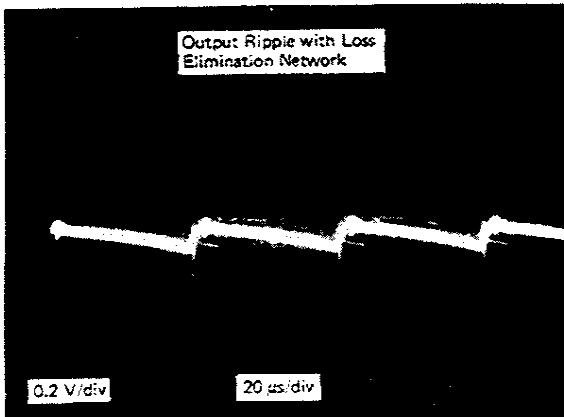
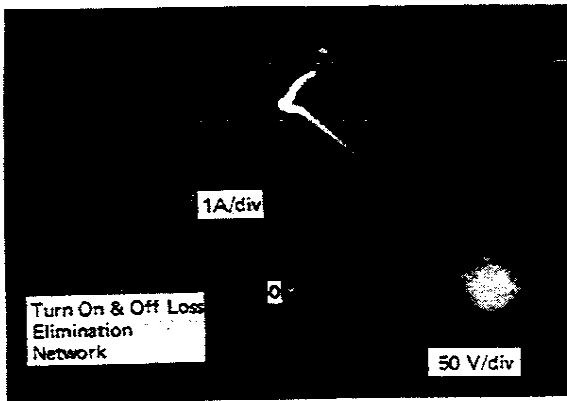


FIGURE 14 – AFTER LOAD LINE SHAPING

The actual loss elimination networks used in this design are shown in Figure 15 along with the design equations. During turn-on the inductor ( $L_x$ ) must block the input voltage ( $V_{in}$ ) during the transistor turn-on time ( $t_r$ ). Diode current ( $I_o$ ) which has been flowing in the free-wheeling diode and inductor prior to this time, will decrease to zero. Measured current rise and fall times on the transistor are 0.3 and 0.4  $\mu\text{sec}$  respectively. Therefore:

$$L_x = \frac{V_{in} t_r}{I_o} = \frac{160 (0.3 \mu\text{sec})}{3} = 16 \mu\text{H}.$$

The value used was 20  $\mu\text{H}$ . The inductor should not saturate at rated load to be effective and current must build back up to  $I_o$  during the free-wheeling time of 40  $\mu\text{sec}$ . The resulting inductance/resistance ( $L/R$ ) time constant of 4  $\mu\text{sec}$  ensures this condition. Therefore:

$$R_L = L_x / \gamma = 20 / 4 = 5 \Omega \text{ and}$$

$$P_R = \frac{1}{2} L I^2 f = \frac{1}{2} \times 20 \mu \times (3)^2 \times 20 \text{ k} = 1.8 \text{ W}$$

Using a 4.7 ohm resistor was practical since the collector voltage overshoot was limited to approximately 15 V.

During turn-off, the capacitor ( $C_x$ ) must supply load current ( $I_o$ ) while the transistor is turning OFF ( $t_f$ ). The capacitor will have charged to the input voltage and will now discharge to zero. Therefore:

$$C_x = \frac{I_o t_f}{V_{in}} = \frac{3 (0.4 \mu\text{sec})}{160} = 7500 \text{ pF}.$$

Because of parasitic circuit capacitance, only 4700 pF is required. The capacitor must recharge during the on-time of the transistor to 10  $\mu\text{sec}$ . The RC time constant of 1.0  $\mu\text{sec}$  is chosen to ensure this condition. Therefore:

$$R_c = \tau / C_x = 1 \mu\text{sec} / 4700 \text{ pF} = 210 \Omega.$$

and

$$P_R = \frac{1}{2} C V^2 f = \frac{1}{2} \times 7500 \text{ pF} \times (160)^2 \times 20 \text{ k} = 1.9 \text{ W}$$

Using a 150 ohm resistor is practical because the collector current overshoot is limited to approximately 1 Ampere. Thus, parasitic capacitance has no effect on this recharge time. In actual practice, the values of  $L_x$  and  $C_x$  may be empirically adjusted to obtain the desired load line effect.

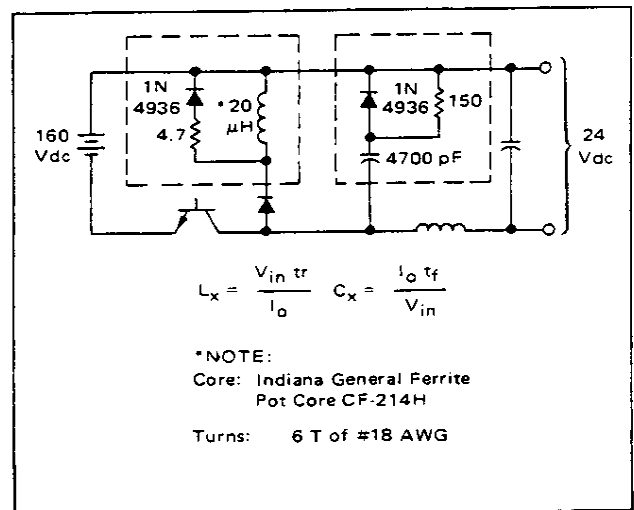


FIGURE 15 – LOSS ELIMINATION NETWORKS

## CONCLUSION

This article has discussed the circuitry used to implement a line operated switching regulator as shown in Figure 16. The semiconductor components highlighted in this design include:

1. A 2N6306 power transistor. This device switches 3 Amperes from 200 Volt and is fast enough to handle the 20 kHz square waves generated by the design.
2. The 2N6034 and 2N6037 complementary power Darlington. These devices switch up to 3 Amperes and provide a single stage interface between the IC logic and the 2N6306.
3. An 4N28 opto coupler. This device provides an electrically isolated feedback signal from the load to the IC logic. It is fast enough to operate in a 20 kHz switching frequency loop.
4. The MC3302P quad comparator. The comparator itself provides all the logic functions necessary for voltage regulation and current limiting in a single package.

Several unique design innovations have been presented. The most significant design highlight is reactive load line shaping. By using techniques from a Bell Laboratories report,<sup>(1)(2)</sup> switching power losses are reduced by a factor of 3. The same reactive components are also responsible for a 6 to 1 reduction in output noise.

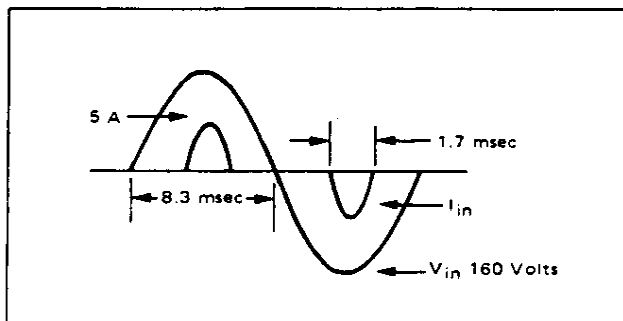
The ripple content of the output voltage (24 Volts dc) is shown in Figure 17. The 60 Hertz ripple frequency has been reduced significantly by a factor of 100 to yield a 200 mV peak-to-peak maximum ripple component. The ripple factor is approximately 10 mV per volt. The 20 kHz output ripple is only 50 mV peak-to-peak; this is less than 0.2% of the output voltage.

The output voltage variations due to line and load changes are as follows:

TABLE 3 — Line versus Load Variation Comparisons

Regulation	Conditions	$\Delta V_o$	Percent Regulation
Load	$I_o = 1.5$ to 3 Ampere $V_{in} = 120$ Volts ac	0.2 V	0.8%
Line	$V_{in} = 100$ to 140 Volts ac $I_o = 3$ Amperes	0.7 V	3%

The efficiency is measured at rated load (24 Volts, 3 Amperes) with 120 Volt input. The oscilloscope display of input voltage and current is shown below:



The input voltage varied from 150 to 170 Volts during the power pulse but is considered constant at 160 Volts. The current pulse is essentially a sine wave. This allows the use of a peak-to-peak comparison on an average ratio of  $2/\pi$  in calculating the total average current and input power as shown:

$$P_{in} = E_p \frac{2}{\pi} I_p \frac{\text{Pulse Width}}{\text{Period}}$$

$$= (160 \text{ V}) \frac{2}{\pi} (5 \text{ A}) \frac{1.7 \text{ ms}}{8.3 \text{ ms}} = 103 \text{ watts.}$$

The overall efficiency can then be found using:

$$\text{Efficiency (\%)} = \frac{P_o}{P_{in}} 100$$

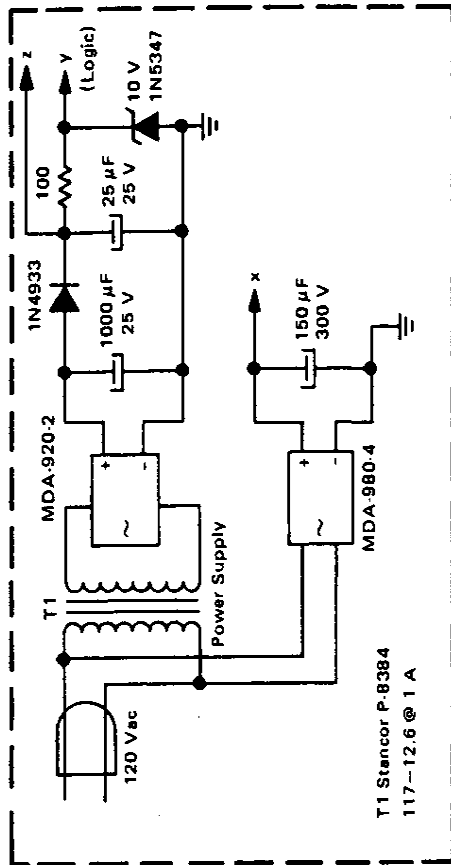
$$= \frac{(24 \text{ V})(3 \text{ A})}{103} 100 = 70\%$$

As was indicated earlier, this is about twice that of a series-pass regulator.

This particular line operated switching regulator has many advantages over a conventional series-pass design; since the efficiency is high, less heat sinking is required. Because the switching regulator operates directly from the line, the need for a 60 Hertz transformer is eliminated. The size and weight reductions are therefore quite significant and are becoming increasingly important in many applications. The main drawback of most uncompensated switching regulators is noise. This drawback can be eliminated by using load line shaping. Load line shaping not only reduces the high frequency electrical noise associated with switching mode supplies but also improves the reliability of the power stage. An opto coupler is used in the 20 kHz feedback loop to maintain a high degree of line and load isolation. A single package quad comparator is used to reduce component quantity for both pulse width control and short circuit protection.

This article has demonstrated a significant method of designing a regulated switching supply with 70% overall efficiency. The approach describes and illustrates how to decrease component counts and avoid using costly, cumbersome and inefficient transformer-operated industrial power supplies, while providing complete short circuit protection.

The author wishes to acknowledge Paul Fletcher for his assistance in preparing this note.



Overall Eff. = 70%    Input    Output  
 100 V-140 V    24 Vdc @ 3 A  
 Mounted On  
 • ① Motorola MS-10 (3°C/W)  
 Heat Sink  
 • ② Staver V4-3-192 (12°C/W)  
 Heat Sink  
 • ③ Custom Design Aircore  
 10mH @ 5 A or TRIAD C-58u

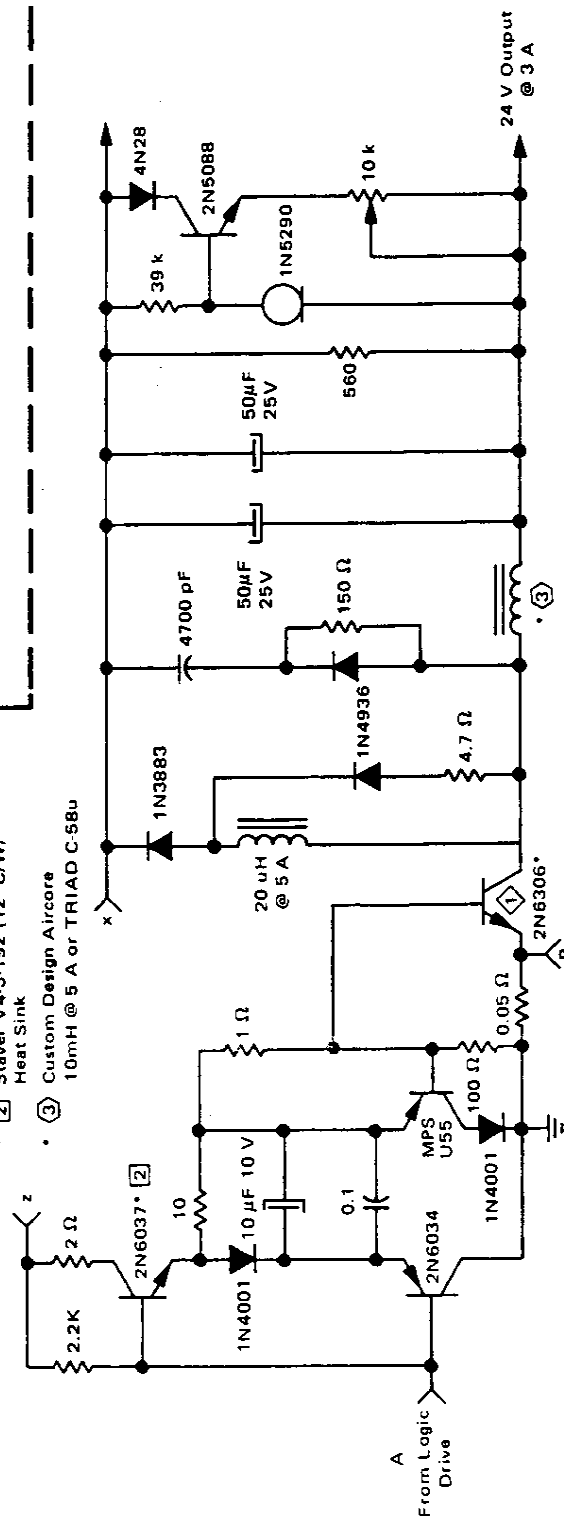
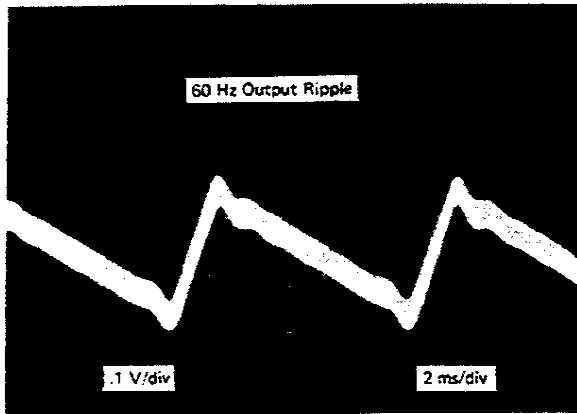


FIGURE 16 — Complete Regulator Schematic (for logic see Figure 10)



NOTE:  
Both Measured  
With 120 Vrms Input  
and Output Ripple with Attendant  
20 kHz ripple 24 V at 3 A

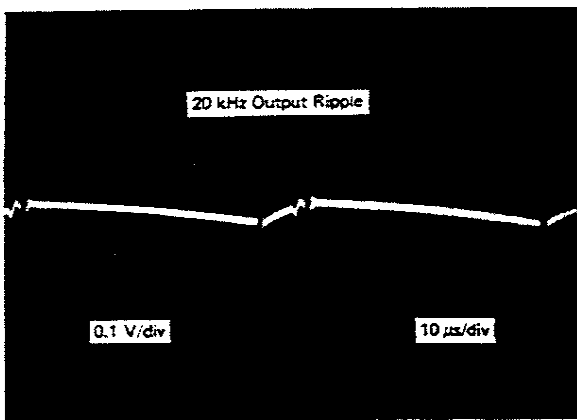



FIGURE 17 – OUTPUT RIPPLE WAVEFORMS

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