AN-752
Application Note

AN 80-WATT SWITCHING REGULATOR FOR CATV AND INDUSTRIAL APPLICATIONS

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This application note describes a 24-Volt, 3-Ampere switching, regulated power supply that operates above 18 kHz from a 40-to 60-Volt, 60-Hz square wave source (CATV power line from a ferroresonant transformer) or a dc standby source with input output isolation. The control circuit consists of a dual operational amplifier and a linear integrated circuit timer which are used to vary the on time of a new high-speed power transistor. The circuit provides good efficiency, good regulation, low output ripple and incorporates input and output over voltage shutdown protection.

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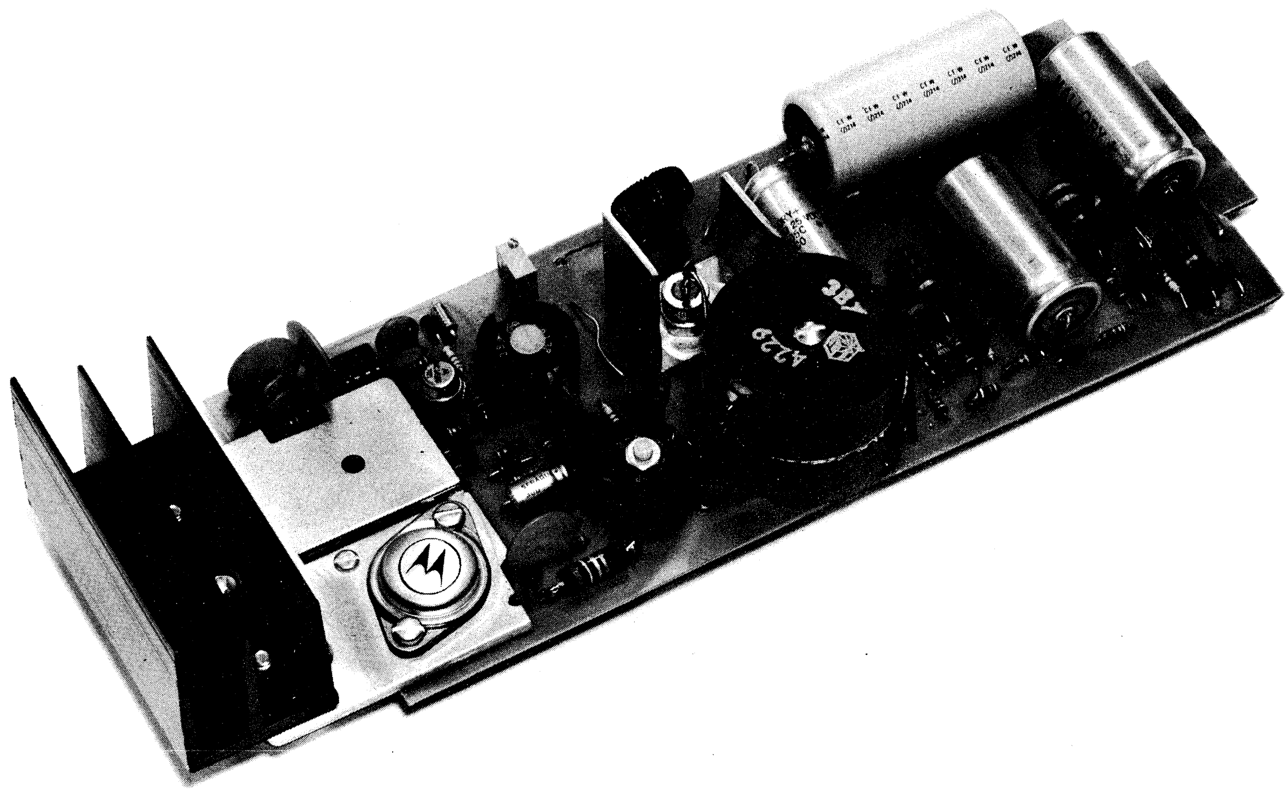

INTRODUCTION

Switching regulated power supplies are growing in popularity in a wide variety of applications. The efficient operation and small size of a switching supply are very attractive features in most applications. This is especially true in cable TV (CATV) systems where small size, weight and efficiency are prime considerations. Although the regulator discussed in this application note was designed to meet specific CATV requirements, the basic approach, including line operation, can be applied to many other applications in the industrial marketplace. A brief discussion of line operation will be included at the end of this application note.

A block diagram of a typical CATV trunk station is shown in Figure 1. The input consists of the RF information signal and a 60-Hz square wave from a ferroresonant transformer. The 60-Hz square wave is used to power the RF circuitry via the regulated dc supply contained within the trunk station. The RF processing block can contain a forward amplifier, a reverse amplifier, a bridger amplifier and AGC circuitry representing a load current (I_L) of as much as 3 Amperes.

In many existing systems the output level of the ferroresonant transformer is 30 Volts and the dc regulator is required to work over a 20-to 30-Volt range depending on the physical distance between the transformer and the trunk station. In newer systems, this output is 60 Volts with the regulator required to work over 40-to 60-Volt range. Ideally, it would be desirable to have the dc power supply work over a 20-to 60-Volt range so that existing 30 Volt systems could be easily retrofitted with any new regulator designs and fewer power supplies would be required. However, some compromise in performance and cost would be required to obtain this wide range of operation. These compromises will be discussed briefly in the design section; in the meantime, the discussion will be limited to a specific design developed to meet the specifications shown in Table I.

TABLE I – CATV Trunk Station Supply Design Specifications

Input Voltage – 40 V to 60 V Squarewave @ 60 Hz
120 Hz Input Ripple – 5 V Peak-to-Peak
Input Transient – 120 V for 200 ms
Output Current – 0.3 A to 3 A
Output Voltage – 20 to 27 Vdc (Adjustable W/1% Regulation)
Output 20 kHz Ripple – < 14 mV Peak-to-Peak
Output 120 Hz Ripple – < 14 mV Peak-to-Peak
Input-Output Isolation

Figure 2-A shows a series switching regulator approach that is very popular in many applications and is used in some existing CATV systems. There are two major disadvantages with this approach. The basic circuit does not provide input-output isolation and the dc input voltage must be greater than the ouptut voltage. To alleviate these problems, a fairly expensive 60 Hz transformer is required on the input as shown. The "ringing" choke converter approach shown in Figure 2-B eliminates the 60 Hz transformer, provides isolation, can be designed to operate over a wide range of input voltages and can operate from a dc stand-by source. Because of these inherent advantages, this approach was used to develop a switching regulated power supply for CATV trunk staions.

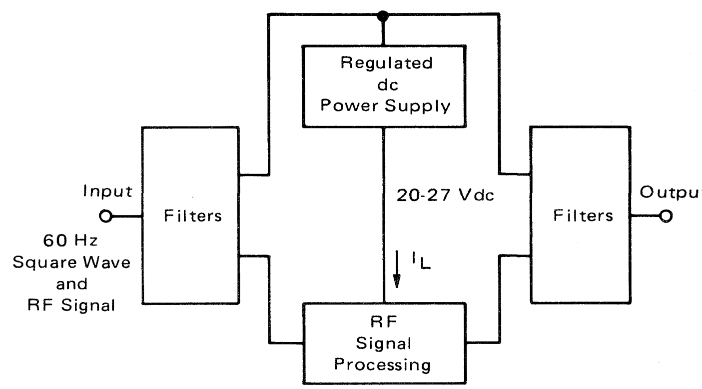

FIGURE 1 – CATV Trunk Station Block Diagram

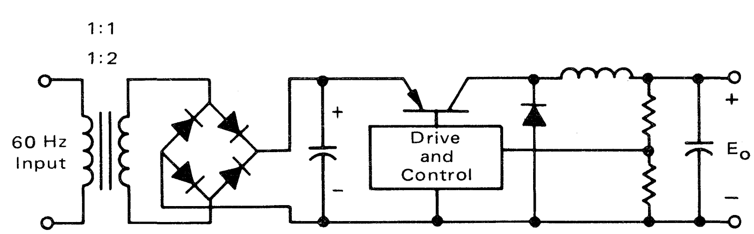

FIGURE 2-A – Series Switching Regulator

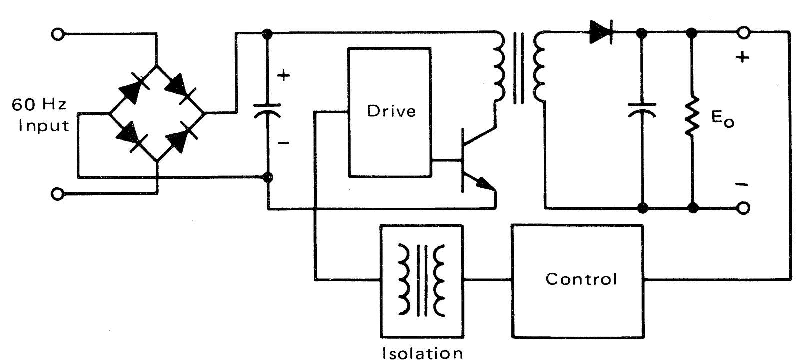

FIGURE 2-B – "Ringing" Choke Switching Regulator

FIGURE 2 – Switching Regulators

THEORY OF OPERATION AND CIRCUIT DESCRIPTION

Power Conversion and Regulation

A functional schematic of the circuit which uses a dual operational-amplifier (MC1458) and an integrated circuit timer (MC1455) to control a new high-voltage, high-speed power transistor (2N6546) is shown in Figure 3. The concept, as in all switching regulators, is to operate the power transistor between saturation and the off state at a high frequency (preferably above the audible range) and vary either the on time or off time, or both, to maintain a constant output voltage. For this particular design, the off time is fixed and the on time is varied.

At start up, the base drive supplied through R1, saturates the 2N6546 and the full input voltage is applied across the primary winding (W1) of the power transformer. With this voltage constant, the current ramps-up linearly until the 2N6546 is switched off. While the power transistor is on, the current transformer represented by W4 provides adequate base drive to keep the 2N6546 in saturation ($I_C/I_B = 10$). At the same time, the secondary (W2) is phased so that diode CR4 is reversed biased and no current flows in the secondary. This causes all of the energy absorbed during the on time of the power transistor to be stored in the primary magnetic field.

When the power transistor is switched off, the transformer polarities reverse, diode CR4 is forward biased, and the energy stored in the primary is transferred to the secondary. The output capacitor, C_O, charges up to the required output voltage and must be large enough to supply worst case load current during the time that diode CR4 is in the blocking state (on time of the power transistor).

The output regulation is accomplished by sensing both the output and input voltages and varying the on time of the power transistor to supply more or less energy depending on the output and input conditions. The output voltage is sensed via a resistor divider network and applied to the non-inverting input of A1 (½ MC1458). This feedback voltage ($K1\ V_O$) is compared to the voltage reference (V_{ref}), the difference is amplified by A1 and A2 (½ MC1458) and the result appears at the output of A2 as a positive dc level (Vdc).

The positive portion of the signal from W3 is proportional to V_I ($K2V_I$) and is integrated by A2 producing a negative ramp at the output of A2. The slope of this ramp is proportional to the input voltage and the starting point (Vdc) is proportional to the input and output voltage. As the slope and the starting point vary, the time required to reach the threshold voltage of the MC1455 varies, thereby, varying the on time of the 2N6546.

When the output voltage of A2 goes below the threshold voltage of the MC1455, its output goes high and turns on a control transistor which pulls the base of the 2N6546 to ground and turns it off. A pulse transformer is used between the control circuitry and the power transistor to maintain input-output isolation. The off time is fixed by the MC1455 and is set to allow complete transfer of energy under worst case conditions (heavy load and low input voltage). When the MC1455 times out, the 2N6546 is allowed to turn on again and the cycle repeats.

A complete schematic of the circuit is shown in Figure 4. Note that the control circuitry (A1, A2, and A3) is powered from the output circuit. This eliminates the need for a separate regulated supply but requires some additional start up circuitry as the control circuitry does not operate until the output voltage is established.

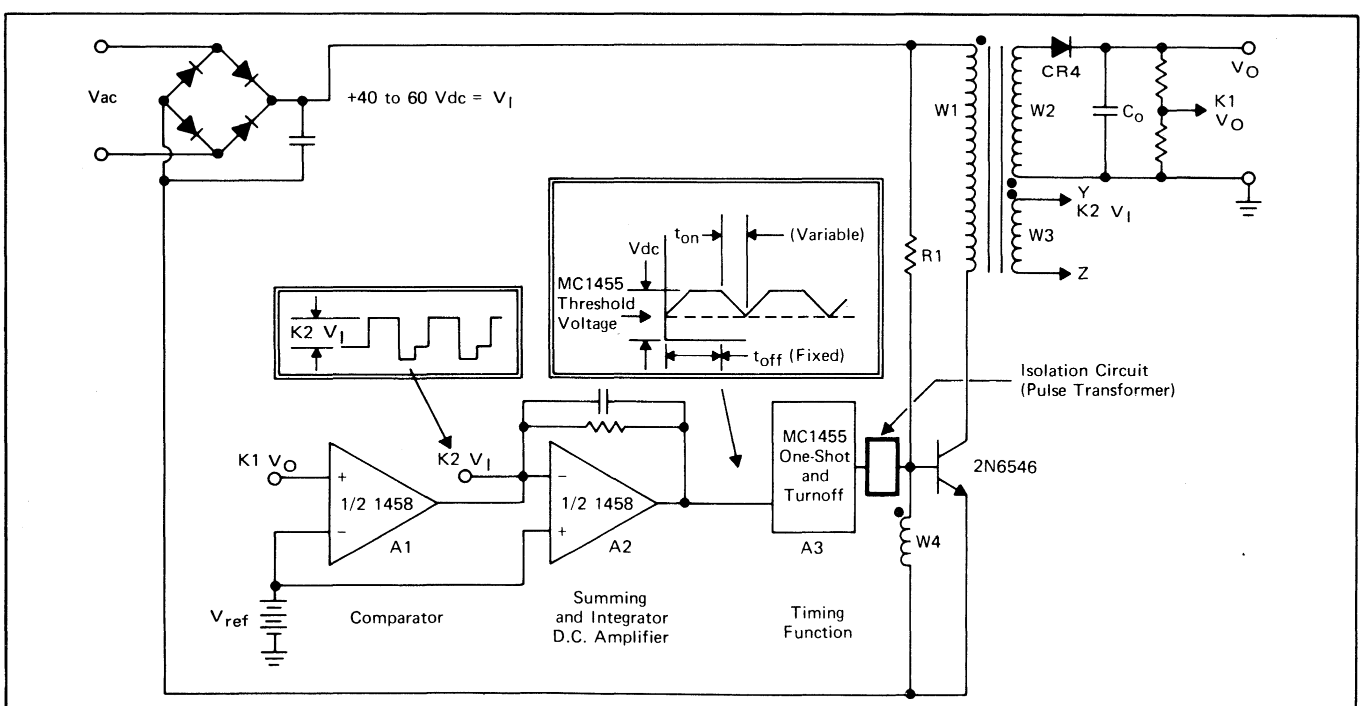

FIGURE 3 — Switching Regulator Block Diagram

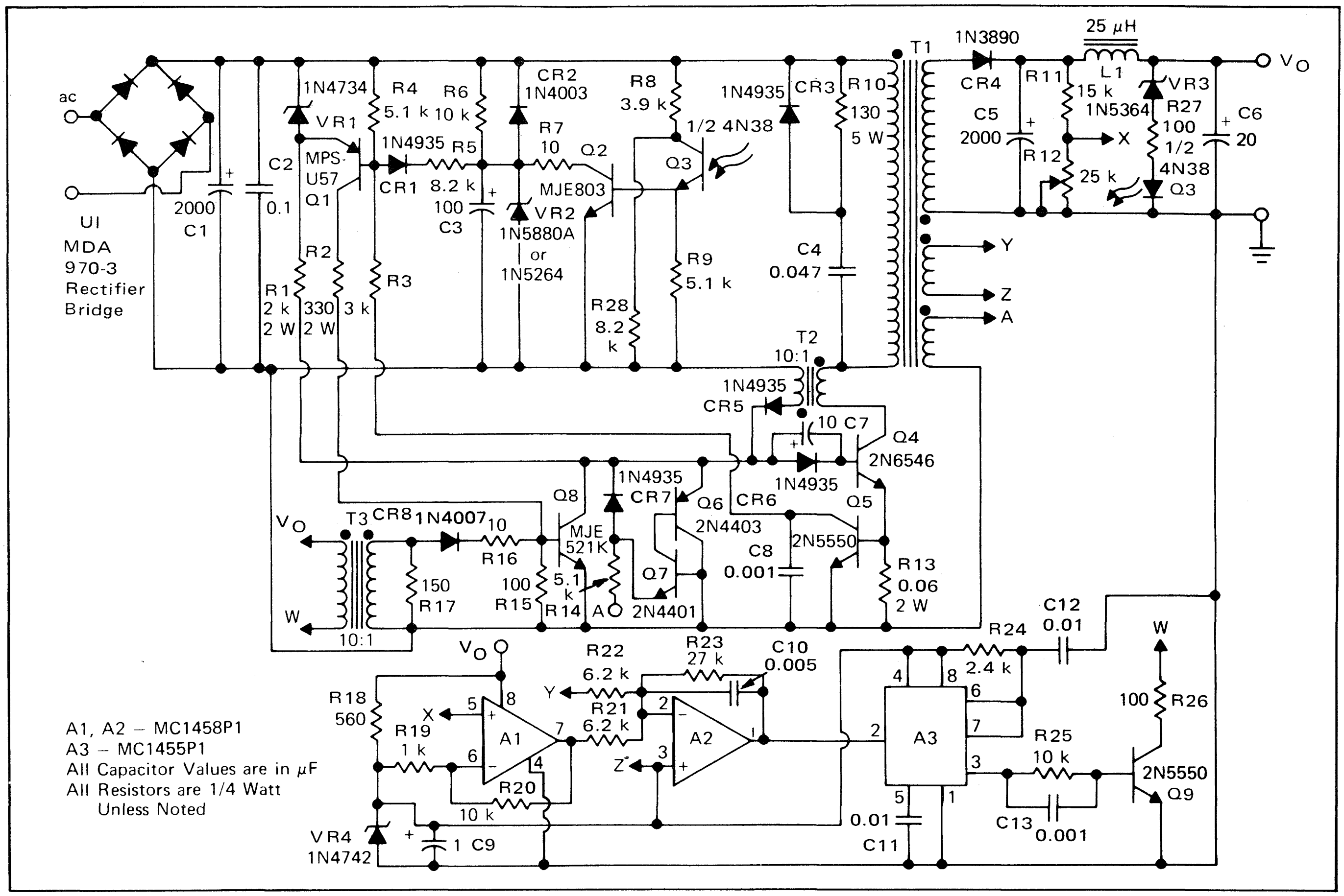

FIGURE 4 — 80-Watt CATV Switching Supply

Start Up Circuit

Figure 4-A is a partial schematic showing the start-up circuitry. Initially the power transistor is allowed to run at a higher current than required under full load conditions to insure start-up with heavy loads. The current in Q4 increases linearly, as in normal operation, until sufficient voltage is developed across R13 to turn on Q5. Q5 provides base drive for Q1 which drives Q8 into saturation and shunts base drive away from Q4. This action, as in normal operation, causes an abrupt change in Q4 collector current, a reversal of transformer polarities and transfer of energy from the primary of T1 to the secondary. Since the current through Q4 and R13 decreases rapidly, some sort of hysteresis is required to prevent Q4 from rapidly turning back on and oscillating about some high dc current level. Auxiliary winding A, Q6, and Q7 provide this hysteresis. During the on time, winding A is positive and provides a small amount of base current to Q4 through R14 and CR7. When the transformer polarities reverse, winding A goes negative turning on Q6 and Q7 which holds the base of Q4 at ground until the secondary has completely discharged. When the secondary current decays to a fairly low level, Q4 is allowed to turn on again and the cycle is repeated until the control circuit takes over. If the output is shorted, this start-up circuit operates at a low duty cycle and protects Q4. However, short circuit currents will appear in the output circuit and could result in component damage if the short circuit condition is prolonged. Continuous short circuit operation could also overstress the start-up circuitry, which was designed for intermittent operation only.

Input Over Voltage Shut Down Circuit

This portion of the supply is designed to shut down if a transient appears on the input line. If the supply was allowed to operate under this condition, the power transistor would have to withstand a voltage level equal to the transient voltage, plus the kick back voltage that appears during turn off. If the circuit is shut down, the components need only block the peak transient voltage that occurs.

A typical condition for CATV systems has been specified in Table I as 120 Volts for 200 ms. If much higher transients are expected, some other means of protection such as a zener clamp across the dc input, would have to be used in addition to or instead of the shut down circuit. The partial schematic of Figure 4-B shows the portion of the circuit used for over voltage shut down. The value of zener VR2 is chosen to clamp the base of Q1 at a voltage slightly higher than the normal operating voltage. When the input voltage exceeds this level plus the zener drop of VR1, Q1 begins to conduct and turns on Q8 shunting the base drive away from Q4. Q4 will be held off as long as the input voltage is high and will automatically restart when the over voltage condition disappears. Resistor R2 and transistor Q1 were designed for low duty cycle (< 1%) operation and would have to be increased in power handling capability if prolonged operation is required.

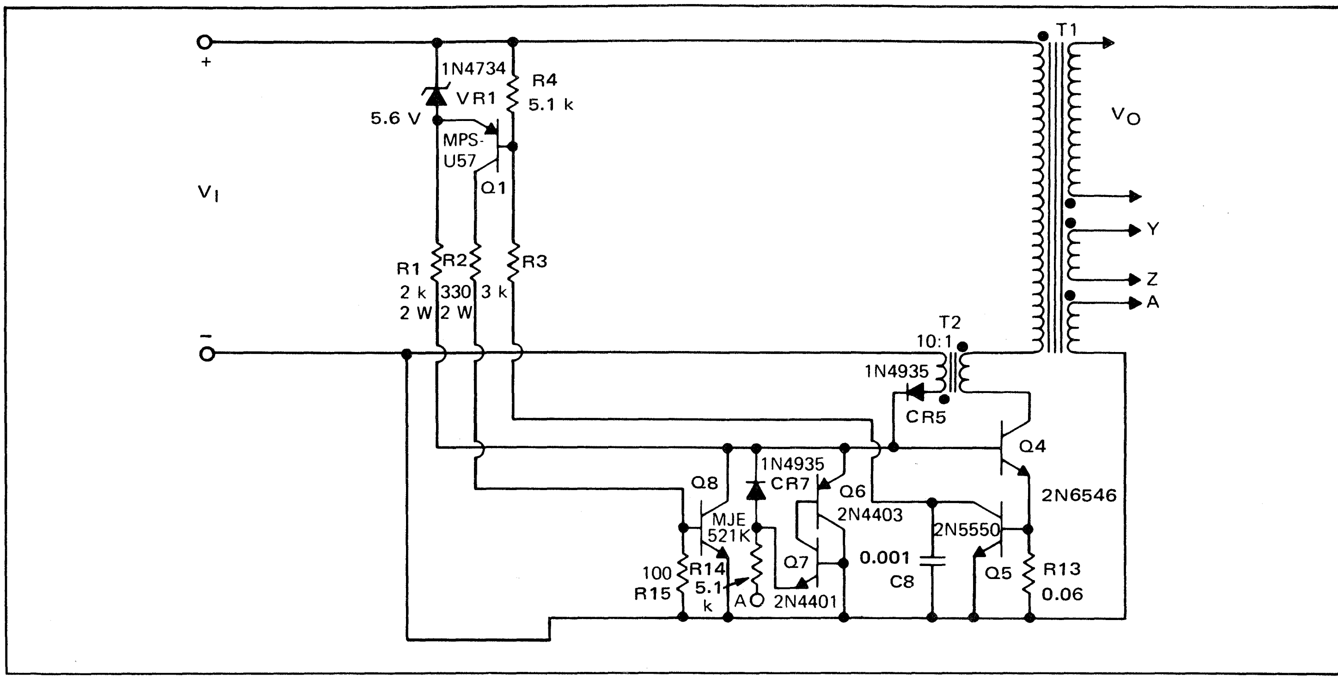

FIGURE 4-A – Start-Up Circuit

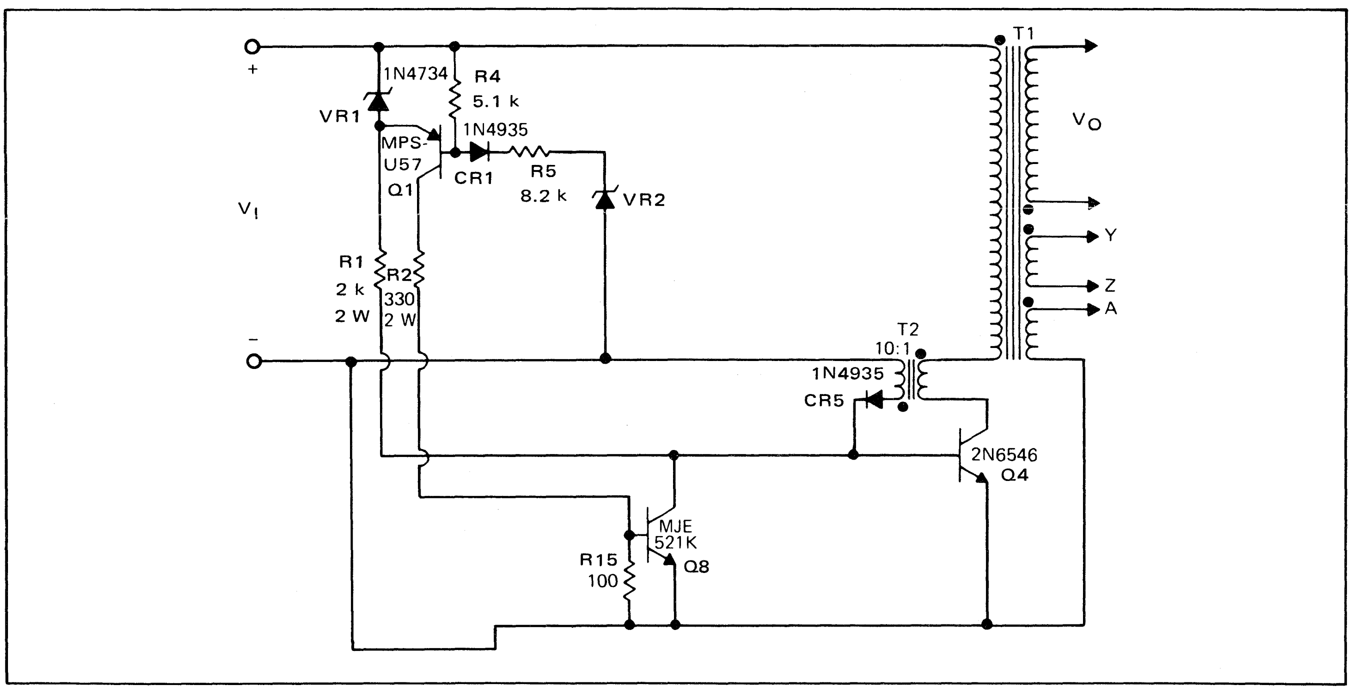

FIGURE 4-B – Input Over Voltage Shut Down Circuit

Output Over Voltage Shut Down Circuit

Under no load conditions the output voltage will rise rapidly at start-up. This situation can cause some problems since the control circuit is designed to work within a very narrow range centered around the output voltage level. The control circuit is saturated in either the high or low state when the output voltage is outside of this range and Q4 operates under start up conditions. If the output voltage passes through this range before the control circuit can react, it would continue to increase until some component breaks down. The circuit components shown in the partial schematic of Figure 4-C are used to guard against this and other situations that could result in loss of control. The circuit is designed to shut down the supply when an over voltage condition exists at the output and automatically restart when this condition disappears. When the output voltage exceeds the zener voltage of VR3, the optoelectronic coupler, Q3 (used for input-output isolation), turns on, driving a darlington transistor (Q2) into saturation. This provides base drive for Q1 and discharges C3 through Q2. Q1 turns on, drives Q8 into saturation which holds Q4 off and prevents the supply from operating.

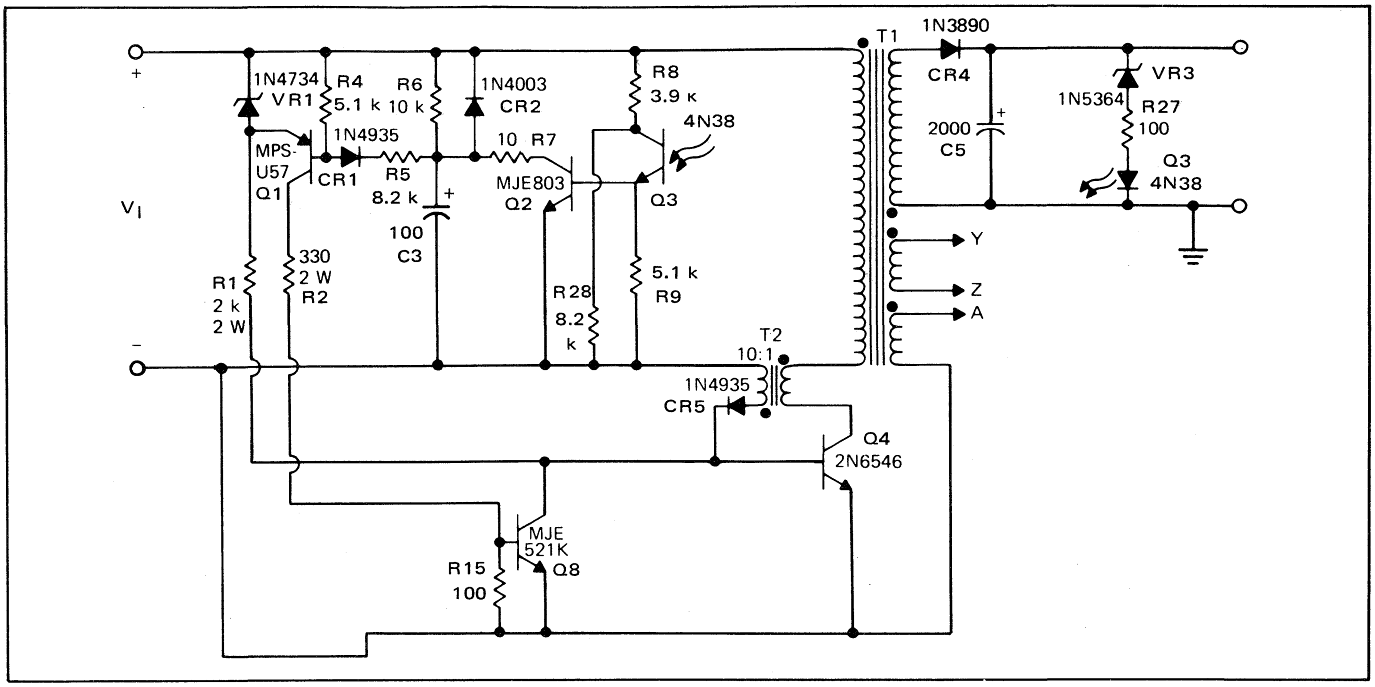

FIGURE 4-C — Output Over Voltage Shut Down Circuit

When the output voltage drops below the zener voltage of VR3, Q3 and Q2 turn off, but C3 has to recharge to a level close to the input voltage before Q1 and Q8 turn off and allow Q4 to return to normal operation. The delay time represented by the recharging of C3 is needed to insure that the output voltage has decayed to a value below the desired output level. This will allow the output voltage to pass through the control range again when the circuit is restarted.

DESIGN CONSIDERATIONS AND EQUATIONS

Input Rectifier and Filter

This portion of the circuit consists of a bridge rectifier assembly and a filter capacitor. A 12 Ampere bridge was used, even though the average current drawn by the supply is less than 3 Amperes, to improve efficiency under start up and heavy load conditions. A 2000 μF capacitor was used to keep the input ripple below 5 V peak-to-peak.

Power Transformer (T1)

The power transformer design is a critical part of this power supply, and requires special consideration. The design begins by assuming that the primary winding is a linear power inductor, and calculating the inductance and operating current level required to meet the system specifications.

System Specifications

P_{out} = 80 Watts (27 V @ 3 A)
V_I = 40 to 60 Volts

Assumed Specifications
f_{min} = 18 kHz, t = 55 μs
η = Efficiency = 80%

The required input power based on the assumed efficiency is:

$$P_{in} = P_{out} \times \frac{1}{\eta} = 80/0.8 = 100 \text{ Watts}$$

and the maximum energy (W) per cycle is

$$W = P_{in} \times t = 100 \text{ Watts} \times 55\ \mu\text{s/cycle} = 5500\ \mu\text{J/cycle}$$

This is the energy that must be stored in the primary during the on time of the power transistor. Knowing this energy, the primary inductance can be calculated as follows. The energy stored in an inductor is

$$W = \tfrac{1}{2} L I^2_{pk} \qquad (1)$$

and the voltage-current relationship is

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

When V is constant, the peak current reached during a specific time interval $\triangle$t, assuming zero initial current is

$$I_{pk} = \frac{V\triangle t}{L} \qquad (2)$$

Substituting equation (2) into equation (1) and solving for L

$$W = \tfrac{1}{2} L \left(\frac{V\triangle t}{L}\right)^2$$

and

$$L = \frac{(V\triangle t)^2}{2W} \qquad (3)$$

The minimum frequency occurs at maximum load and low input voltage. If we limit the duty cycle at the low frequency to 50 or 60%, Δt can be chosen to be 30 μs. Substituting the known quantities into equation (3) we get:

$$L = \frac{(40 \times 30 \times 10^{-6})^2}{2\,(5500 \times 10^{-6})} = 130\ \mu H$$

With this inductance and the fact that the maximum peak current occurs at maximum duty cycle, equation (2) can be used to calculate the current required.

$$I_{pk}\,(max) = \frac{V \Delta t}{L} = \frac{40\,(30 \times 10^{-6})}{130 \times 10^{-6}} = 9.2\ \text{Amp}$$

This is the peak operating current at maximum load conditions. However, during start up, I_{PK} will reach 10 Amperes which is set by R13 shown in Figure 4-A. At this point the material, shape and size of the transformer core must be chosen. In this design example a ferrite pot core was used because it was readily available and would give good performance at reasonable cost and size. However, system requirements may dictate a different choice and the designer should carefully investigate available materials and cores before making a final decision. Ferroxcube's 3B7 material was chosen for this design example and the following approach was used to determine the minimum core size.

First determine the required $A_C A_{CB}$ product from

$$*A_C A_{CB} \geqslant \frac{1.3\ Po}{f\ B_{max}} \times 10^6\ (cm^4) \qquad (4)$$

where A_C = Available Winding Area of Core (cm^2)
A_{CB} = Effective Area of Core Bobbin (cm^2)
P_{out} = Output Power (Watts)
f = Frequency (Hz)
B_{max} = Saturation Flux Density (Guass)

For 3B7 material B_{max} = 3800 Gauss, therefore:

$$A_C A_{CB} \geqslant \frac{1.3\ (80)}{(18 \times 10^3)\ (3.8 \times 10^3)} \times 10^6 = 1.52\ cm^4$$

Two available cores that met this requirement were Ferroxcube Part Nos. 3622P-L00-3B7 and 4229P-L00-3B7. The specifications for these parts are shown in Table II.

From the $A_C A_{CB}$ product, the 3622 core could be used, however, the 4229 core was chosen to allow some flexibility in the design. Having decided on a core size and material, the minimum number of primary turns can now be determined from

$$Np = \frac{Lp\ I\,pk}{A_e\ B_{max}} \times 10^8 \qquad (5)$$

where: Np = Primary Turns
Lp = Primary Inductance (Henries)
I pk = Peak Primary Current (Amperes)
A_e = Effective Area of Core (cm^2)
B_{max} = Saturation Flux Density (Gauss)

Since the saturation flux density of ferrite materials decreases with increasing temperature, the value of B_{max} should be adjusted in equation (5) to provide for high temperature operation. Based on the published high temperature data for Ferroxcube 3B7 material, B_{max} was set at 2000 Gauss instead of 3800 Gauss (B_{max} @ 25°C).

Calculating N_p from equation (5)

$$Np = \frac{(130\ \mu H)\ (10\ A)}{(2.66\ cm^2)\ (2000\ \text{Gauss})} \times 10^8 = 24\ \text{Turns}$$

The required air gap can be determined using the following relationships for the effective flux path.

$$l_e = \frac{0.4\ \pi Np I\,pk\ \mu_{av}}{B_{max}} \qquad (6)$$

and

$$l_e = l_m + \mu_{av}\ l_g \qquad (7)$$

where

l_e = Effective Flux Path (cm)
Np = Primary Turns
μ_{av} = Average Permeability
I pk = Peak Primary Current (Amperes)
B_{max} = Saturation Flux Density (Gauss)
l_m = Length of Magnetic Path (cm)
l_g = Length of Air Gap (cm)

The average permeability for 3B7 material, as shown in Table II is 1900, and calculating l_e from equation (6)

$$l_e = \frac{(0.4)\ (3.14)\ (24)\ (10)\ (1900)}{2000} = 286\ cm$$

TABLE II

Core Material	Core Number	A_e (CM^2)	A_C (CM^2)	A_{CB}	lm (cm)	B_{max} Gauss	μAV	$A_C A_{CB}$ (cm^4)
	3622	2.02	1.08	0.116 in^2 0.748 cm^2	5.78	3800	1900	1.51
3B7	4229	2.66	1.94	0.217 in^2 1.40 cm^2	6.81	3800	1900	3.72
	2213	0.635	0.420	0.046 in^2 0.297 cm^2	3.12	3800	1900	0.189

*The equations used in this approach are from Reference 7.

Substituting this value for l_e and the value for l_m from Table II into Equation (7) we get

$$286 \text{ cm} = 6.81 \text{ cm} + (1900)\, l_g$$

$$l_g = \frac{2.86 - 6.81}{1900} = 0.147 \text{ cm} = 0.058 \text{ in}$$

Now the minimum primary to secondary winding ratio can be determined by considering the maximum allowable discharge time in the secondary circuit. Using the previous assumption of a 30 μs on time at 18 kHz, the secondary circuit must discharge within 25 μs, and the turns ratio can be found from the following relationships

$$V_S = \frac{L_S dI_S}{dt_{off}} \approx \frac{L_S \triangle I_S}{\triangle t_{off}} \qquad (8)$$

$$L_S = L_P/(N_P/N_S)^2 \qquad (9)$$

$$I_S = I_P(N_P/N_S) \qquad (10)$$

Substituting Equations (9) and (10) into (8) and solving for (N_P/N_S) min

$$V_S \approx \frac{[L_P/(N_P/N_S)^2]\ [\triangle I_P(N_P/N_S)]}{\triangle t_{off}}$$

$$V_S \approx \frac{(L_P)(\triangle I_P)}{(N_P/N_S)(\triangle t_{off})}$$

$$\frac{N_P}{N_S} \quad \frac{(L_P)(\triangle I_P)}{(V_S)(\triangle t_{off})}$$

$$\left(\frac{N_P}{N_S}\right)_{min} \approx \frac{(130 \times 10^{-6} h)(9.2\ A)}{(27\ V)(25 \times 10^{-6} s)} = 1.77$$

Solving for the secondary turns

$$N_S = N_P/1.77$$
$$= 24/1.77 = 13.55 \text{ (used 13 turns)}$$

The number of turns for the control windings (Y-Z and A) shown in Figure 4 were chosen to give reasonable drive signals. Winding A performs a polarity sensing function and can be 1:1 with a series resistor to limit the current. Winding Y-Z drives the integrator and was set at a 6:1 ratio to give a 10-Volt signal at the maximum input voltage. Once the number of turns for each winding is known, the wire size can be chosen and the fit can be checked. The wire sizes for the primary and secondary windings were chosen so that their ampere per square inch ratings were less than 1500 A/square inch. Almost any convenient wire size down to 40 gauge can be used for the control windings as the currents in these windings are very low. Number 16 gauge H. F. wire was used for the primary and secondary and number 26 gauge for the control windings.

Using the turns per square inch factor for each wire size, we can determine the fit as follows.

$$A_{CB} \geqslant \frac{N_P}{T16} + \frac{N_S}{T16} + \frac{N_{Y-Z}}{T26} + \frac{N_A}{T26}$$

where

N_P = Primary Turns
N_S = Secondary Turns
N_{Y-Z} = Turns On the Y-Z Winding
N_A = Turns On the A Winding
T16 = Turns Per Square Inch for #16 Gauge
T26 = Turns Per Square Inch for #26 Gauge

$$A_{CB} \geqslant \frac{24}{327} + \frac{13}{327} + \frac{4}{2932} + \frac{24}{2932} = 0.122 \text{ sq. in.}$$

From Table II, A_{CB} for the 4229 core is 0.217 square inch, therefore, this design should fit on the 4229 core. The complete specifications for T1 are shown in Table III.

BASE DRIVE TRANSFORMER (T2)

The base drive transformer was designed as a low-level pulse transformer. The winding ratio of this transformer should be chosen so that the power transistor operates at a forced gain low enough to maintain a reasonable saturation voltage at peak collector currents. In this design a 1:10 ratio was used which will provide 1 Ampere of base drive at 10 Amperes operating current (starting conditions). Figure 5 shows the base drive circuit and the voltage and current waveforms for transformer T2. As shown in waveform B, the secondary voltage during the on time of the power transistor is equal to two diode drops, a V_{BE} drop and the drop across R13. These voltages increase with increasing base current; worst case conditions are shown in Figure 5. An important consideration in the design of this transformer is to keep the percentage droop at the operating frequency as low as possible. This will insure that the voltage and current pulses are transformed with minimum loss. The percentage droop of a pulse transformer is defined by the following relationship:

$${}^*P_D = \frac{100\, R\, t_p}{L_m} \qquad (11)$$

where

P_D = Percentage Drop (Percent)
R = Effective Primary Circuit Resistance (Ohms)
t_p = Width of the Pulse Being Reproduced (Seconds)
L_m = Magnetizing Inductance of the Transformer (Henries)

The effective primary resistance (R) in equation (11) consist of several components as shown in the transformer low frequency equivalent circuits of Figures 5-A and 5-B.

*The equations used in this approach are from Reference 7.

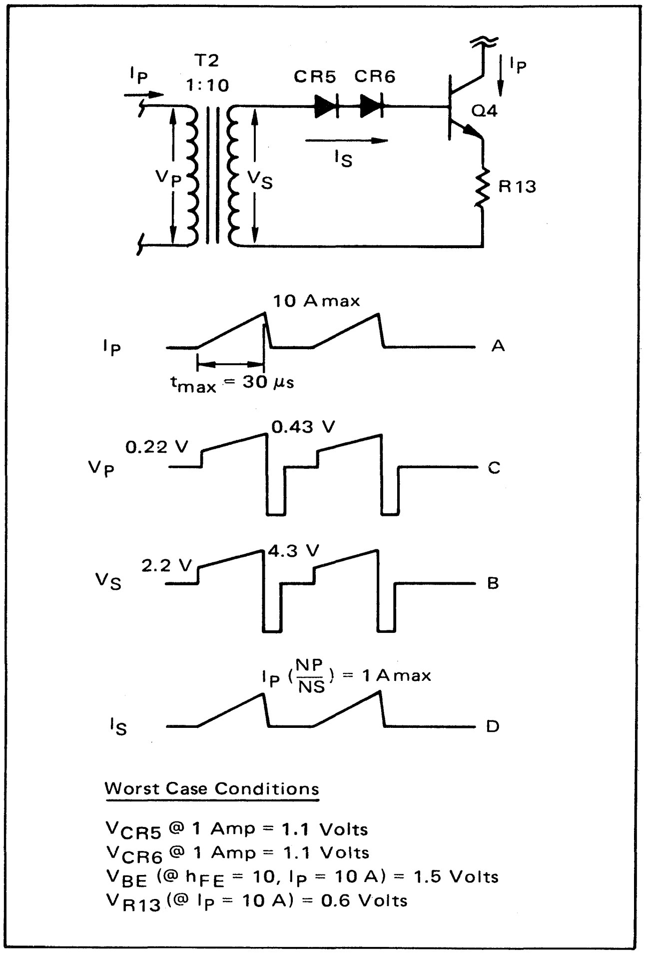

FIGURE 5 — Base Drive Transformer (T2)

From Figure 5-B, R is equal to $(R_g + R_p)$ in parallel with the parallel combination of R_C and $(n^2R_S + n^2R_L)$. However, a simplified approximation for R can be obtained if the following assumptions are made.

a) $R_S \ll R_L$
b) R_C is large (meaning core loss is low)
c) $R_g \gg R_P$
d) $R_g \gg n^2R_L$

In this application these are reasonable approximations and

$$R = n^2R_L$$

From Figure 5

$$R_L = \frac{V_S}{I_S} = \frac{4.3\text{ V}}{1\text{ A}} = 4.3\ \Omega$$

$$n = \frac{1}{10}$$

Therefore:

$$R = \left(\frac{1}{10}\right)^2 (4.3) = 0.043\ \Omega$$

With this value of R and assuming a droop of 1%, the minimum value of L_m at the maximum pulse width can be determined from Equation (11).

$$L_m = \frac{100\ R\ t_p}{P_D} = \frac{100\ (0.043)\ (30 \times 10^{-6})}{1} = 129\ \mu\text{H}$$

Now the core can be sized using the same approach that was used in the design of T1.

$$A_CA_{CB} \geqslant \frac{1.3\ P_{out}}{f\ B_{max}} \times 10^6\ (\text{cm}^4) \qquad (4)$$

From Figure 5 P_{out} = (4.3 V) (1 A) = 4.3 Watts

Therefore:

$$A_CA_{CB} \geqslant \frac{1.3\ (4.3) \times 10^6}{(18 \times 10^3)(2 \times 10^3)} = 0.155\ \text{cm}^4$$

The smallest available pot core meeting this requirement was Ferroxcube Part No. 2213P-L00-3B7. The specifications for this part are also shown in Table II. The number of primary turns can now be determined from

$$L_m = 0.4\ \pi N_P^2 \mu_{av} \left[\frac{A_e}{l_m}\right] \times 10^{-8}$$

solving for N_P

$$N_P = \sqrt{\frac{L_m}{0.4\ \pi\ \mu_{av} \left[\frac{A_e}{l_m}\right] \times 10^{-8}}}$$

$$N_P = \sqrt{\frac{129 \times 10^{-6}}{(0.4)\ (3.14)\ (1.9 \times 10^3) \left[\frac{0.635}{3.12}\right] \times 10^{-8}}}$$

$$= \sqrt{26.5} = 5.1\ \text{Turns, Used 5 Turns}$$

and

$$N_S = \frac{N_P}{n} = 5/(1/10) = 50\ \text{Turns}$$

Using the same current per square inch value as before, #16 gauge wire was chosen for the primary and #24 gauge for the secondary. Checking the fit,

$$A_{CB} \geqslant \frac{N_P}{T16} + \frac{N_S}{T24}$$

$$A_{CB} \geqslant \frac{5}{327} + \frac{50}{1893} = 0.0416\ \text{in sq.}$$

The A_{CB} for the 2213 core is 0.046 in. sq. from Table II. The complete specifications for T2 are shown in Table III.

TABLE III – Transformer Specifications

T1:	Primary	– 24 Turns of #16 AWG
	Secondary	– 13 Turns of #16 AWG
	Y-Z Winding	– 4 Turns of #24 AWG
	A Winding	– 24 Turns of #24 AWG
	Pot Core	– Ferroxcube #4229P-L00-3B7
	Bobbin	– Ferroxcube #4229FID
	Air Gap	– 0.058 in.
T2:	Primary	– 5 Turns of #16 AWG
	Secondary	– 50 Turns of #24 AWG
	Pot Core	– #2213P-L00-3B7 Ferroxcube
	Bobbin	– #2213FID Ferroxcube
	Air Gap	– 0
T3:	Primary	– 330 Turns of #32 AWG
	Secondary	– 33 Turns of #26 AWG
	Pot Core	– #2213P-L00-3B7 Ferroxcube
	Bobbin	– #2213FID
	Air Gap	– 0

CONTROL CIRCUIT PULSE TRANSFORMER (T3)

The procedure used to design this transformer is exactly the same as the procedure used to design the base drive transformer (T2). Figure 6 shows the specific circuit components used to determine the output requirement for T3. During the initial turn-off of Q4, Q8 is required to sink I_{B2}, plus I_{B1} from the base drive transformer. I_{B1} will be present until the collector current begins to decrease (depending on the storage time of Q4).

Therefore, initially

$$I_{CQ8max} = I_{B1} + I_{B2}$$

$$= \frac{I_{CQ4}}{10} + \frac{I_{CQ4}}{10} \quad \text{(assuming } I_{B1} = I_{B2} = I_C/10\text{)}$$

$$I_{CQ8max} = \frac{9.2\ A}{10} + \frac{9.2\ A}{10} = 1.84\ \text{Amp}$$

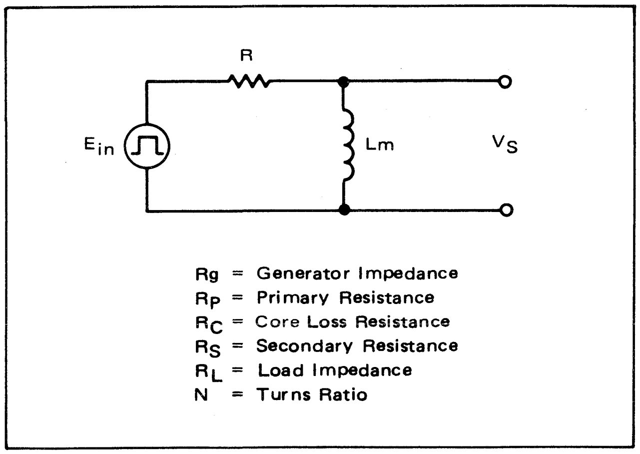

FIGURE 5-A

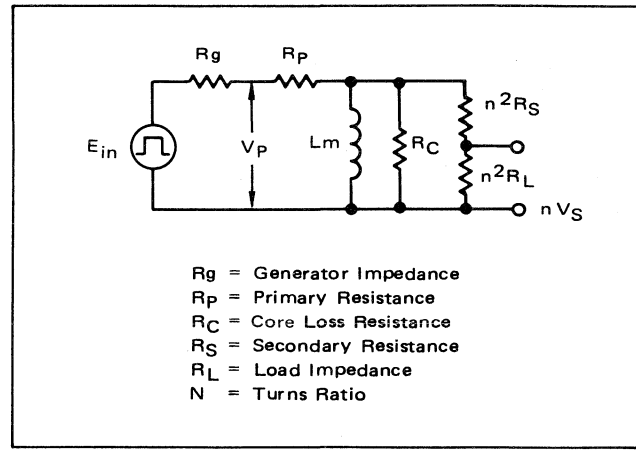

FIGURE 5-B – Low Frequency Equivalent Circuits

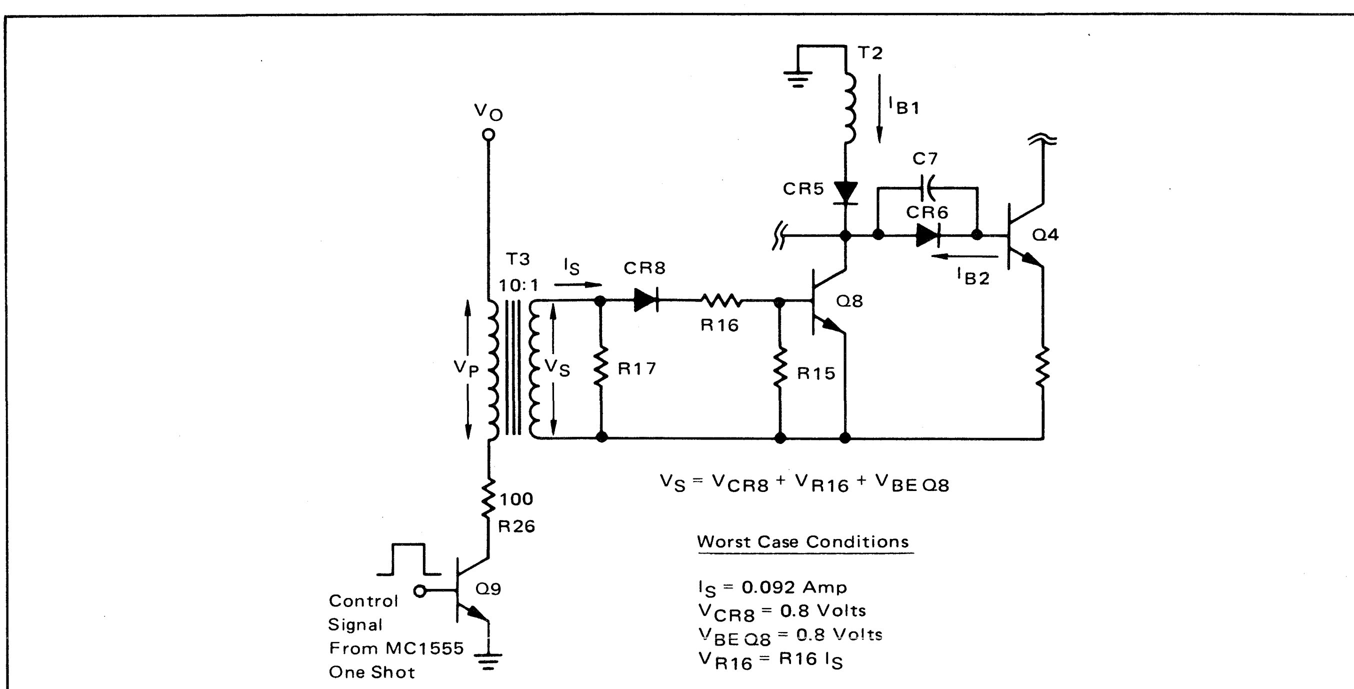

FIGURE 6 – Control Transformer Circuit (T3)

If Q8 is operated at a forced gain of 20 under worst case conditions, then

$$I_S = \frac{I_{CQ8max}}{20} = \frac{1.84 \text{ Amp}}{20} = 0.092 \text{ Amp}$$

The winding ratio of T3 can be chosen so that the control circuitry works at a fairly low current level and adequate voltage is applied to the base circuit of Q8. CR8 is used to protect the base-emitter junction of Q8 from reverse voltages induced during turn-off and R16 is used to limit the base drive during the on time. Using a standard low-speed diode such as the 1N4003 for CR8 provides some reverse bias for Q8 during turn-off and improves its switching time. The reverse bias appears because of the reverse current of the diode during reverse recovery. In this example the winding ratio was set at 10:1 and R26 was added to limit the current in Q9 during possible fault conditions. Solving for the primary voltage (V_P)

$$V_P = V_O - V_{R26} - V_{CE(sat)\,Q9}$$
$$= V_O - (I_S/n)(100) - V_{CE(sat)Q9}$$

With the output voltage of the supply adjusted to its low value (V_O = 20 Volts) and assuming a saturation voltage of 0.3 Volts for Q9 gives

$$V_{P(min)} = 20 - (0.092/10)(100) - 0.3 = 19 \text{ Volts}$$

and

$$V_{S(min)} = V_P/N = 19/10 = 1.9 \text{ Volts}$$

knowing the minimum V_S, R16 can be set to provide the required base drive. From Figure 6

$$V_S = V_{CR8} + I_S R16 + V_{BE\,Q8}$$

$$R16 = \frac{V_S - V_{CR8} = V_{BE\,Q8}}{I_S} = \frac{1.9 - 0.8 - 0.8}{0.092}$$

$$= 3.26\ \Omega \text{ (used } 3\ \Omega\text{)}$$

At higher supply output voltages the base drive and operating current of the control transistor (Q9) will increase but will still be within the ratings of the devices.

The reflected primary impedance can be calculated, based on the secondary requirements as follows:

$$R = \frac{V_S}{I_S}(n)^2 = \frac{1.9}{0.092}(10)^2 = 2065\ \Omega$$

substituting this value into Equation (11), allowing a 10% droop and solving for L_m

$$P_D = \frac{100\ Rt_p}{L_m} \qquad (11)$$

or

$$L_m = \frac{100(2065\ \Omega)(25 \times 10^{-6}\ \mu s)}{10} = 516 \text{ mH}$$

Calculating the turns and wire size using the same procedure that was used for T2 resulted in the transformer specifications shown in Table III for T3. Sizing the core based on output power requriements indicated that a smaller core could be used; however, in this example the 2213 core was used for convenience.

POWER TRANSISTOR (Q4)

Most of the important power transistor requirements can be established based on the design of the power transformer (T1) and the base drive transformer (T2). Q4 must operate at approximately 9 Amperes with a forced gain of 10 and reasonably low saturation voltage.

The required blocking voltage for Q4 can be determined from the following relationship.

$$BV_{CEX} > V_I(max) + n_{T1}\ V_O(max) \qquad (12)$$

where:

BV_{CEX} = The Collector-to-Emitter Breakdown Voltage With Specified Base Circuit Conditions
V_I (max) = 80 Volts Before Shutdown Occurs
n_{T1} = T1 Winding Ratio $\simeq$ 2
V_O (max) = 27 Volts

Therefore

$$BV_{CEX} > 80 \text{ V} + (2)(27)$$
$$BV_{CEX} > 134 \text{ Volts}$$

This voltage will appear across Q4 after it is completely off. However, during the actual turn-off, Q4 will be required to handle maximum current at some higher voltage with the base-emitter junction reversed biased. The voltage reached during the turn off transition will depend on the snubber network used across the primary of T1 (which will be discussed in some detail later). To properly design the snubber network, the device capability during reversed bias turn off must be established. The $E_{S/b}$ rating (secondary breakdown energy in the reversed bias mode) shown on many power transistor data sheets guarantees the device capability with a specific load inductance. However, data has shown that the $E_{S/b}$ rating decreases significantly with increasing inductance. Also, in most applications the voltage is limited and does not reach the avalanche voltage of the device, so that an $E_{S/b}$ rating does not necessarily apply. On the data sheet for the 2N6546, used in this application, a new specification has been added which is shown in Figure 7. The specification states that the device can switch rated current at a specific clamped voltage (i.e. if the collector voltage is clamped at or below the specified value, the transistor can switch rated current regardless of inductive load). With this information the circuit designer can optimize the clamping network and still be assured that the device is being used within its safe operating area.

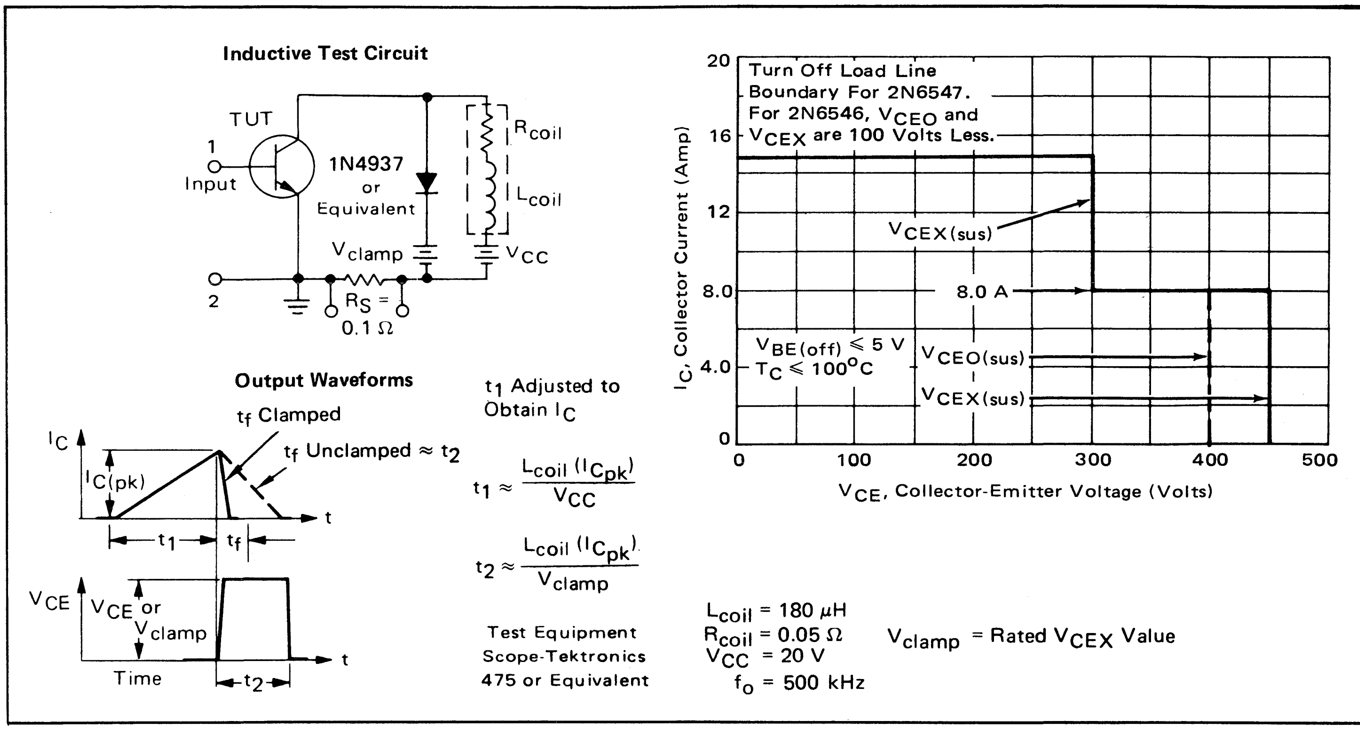

FIGURE 7 — Switching Safe Operating Area — V_{CE} (Clamped)

The switching time of Q4 (especially the fall time) is important and should be as low as possible to reduce power losses during turn-off. CR6 and C7 shown in Figure 4 were added to provide some reverse bias during turn-off and improve the switching of Q4.

The voltage and current waveforms for Q4 are shown in Figure 8. The load line is shown in Figure 9.

LOAD LINE SHAPING NETWORK

Load line shaping is used to reduce transistor dissipation during switching and to protect the device from overstress conditions. The load line shaping network is shown in Figure 10. A conservative approach in designing this network is to choose C4 to limit the rate of rise of voltage across Q4 so that the collector voltage reaches the rated clamp voltage at a time greater than the fall time of the power transistor. This will insure that the power transistor is operating within its safe operating area (SOA).

Therefore:

$$C4 = \frac{I_{pk}\ t_V}{V_{Clamped}} \qquad (13)$$

where:

I_{pk} = Peak Current In the Primary (Amperes)

t_V = Time to Reach Clamp Voltage (should be greater than the inductive fall time of Q4) (Seconds)

$V_{Clamped}$ = Rated Clamp Voltage (Volts)

To calculate C4min use $t_V = t_{f\,max} = 1\ \mu s$

$$C4\,min = \frac{(10)\,(1.0 \times 10^{-6})}{350} = 0.029\ \mu F \text{ (used } 0.047\ \mu F)$$

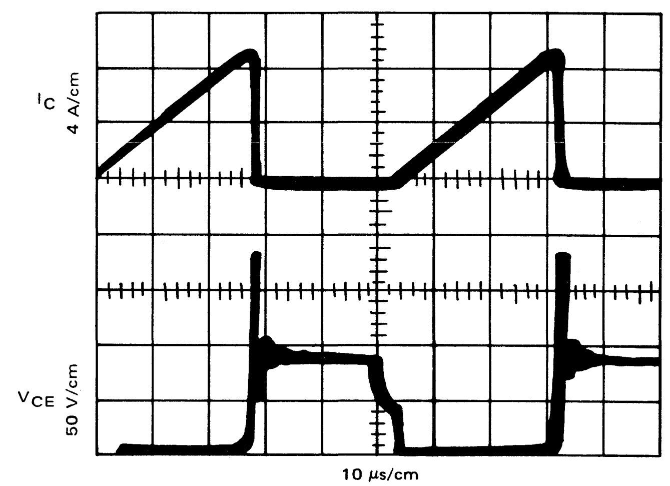

FIGURE 8 — Voltage and Current Waveform for Q4

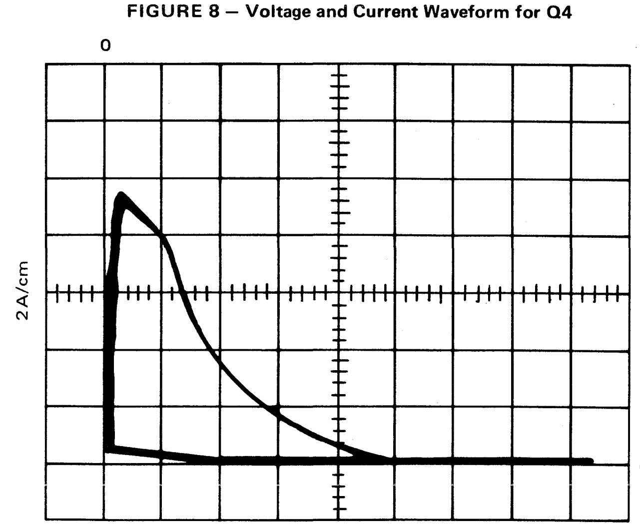

FIGURE 9 — Q4 Load Line

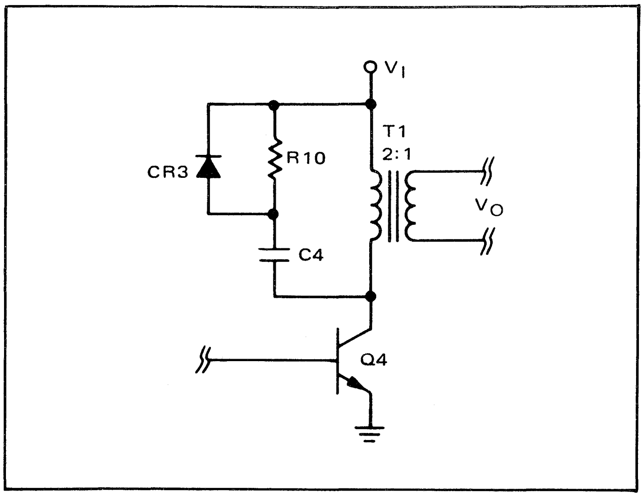

FIGURE 10 — Load Line Shaping Network

R10 should be selected so that C4 charges close to the input voltage during the on time, but limits the additional current supplied to Q4 at turn on. The minimum on time which occurs at minimum load, is approximately 6 μs.

Therefore:

$$R10C4 = 6 \times 10^{-6}\ s$$
$$R10 = 6 \times 10^{-6}/(0.047 \times 10^{-6}) = 128\ \Omega \text{(used } 130\ \Omega)$$

The major portion of the power dissipation in R10 occurs during the charging cycle and can be estimated as follows:

$$P_{D_{avg}} = \frac{V_I{}^2\ K\tau}{R10\ T} \tag{14}$$

where

- K = A Constant Used to Transfer the Exponential Power Pulse in R10 Into An Equivalent Rectangular Pulse and Is $\cong 0.5$
- τ = RC Time Constant (Seconds)
- T = Period at Maximum Frequency (Seconds)

Substituting and Solving for $P_{D_{avg}}$ gives

$$P_{D_{avg}} = \frac{(60)^2(0.5)(6 \times 10^{-6})}{(130)(31 \times 10^{-6})} = 2.7 \text{ Watts}$$

CONTROL CIRCUIT

The most important design considerations for this portion of the circuit (see Figure 11) are the dc gain from the supply output to the output of the integrator (A2), the integrator time constant and the time constant of the MC1455 (A3). The zener reference diode (VR4) is not critical in this design because the output is adjustable. However, if R11 and R12 are fixed, then the combined tolerances of R11, R12 and VR4 will have to be less than the specified output voltage variation.

The integrator time constant (R22C10 from Figure 11) can be set to equal the maximum on time of the power transistor (30 μs at low input and heavy loads). If a convenient value is chosen for C10 then R22 can be determined. In this design, C10 was chosen to be 0.005 μF then

$$R22 = \frac{t_{on}}{C10} = \frac{30 \times 10^{-6}\ \mu s}{0.005 \times 10^{-6}} = 6\ k\Omega$$

At low input, the Y-Z voltage to the integrator A2 during the on time of Q4 is

$$+V_{Y\text{-}Z} = \left(\frac{N_{Y\text{-}Z}}{N_P}\right)(V_{I\ low}) = (1/6)(40\ V) = 6.67 \text{ Volts}$$

and the output of A2 will have a slope of $-\frac{+V_{Y\text{-}Z}}{R22C10}$

$$= -\frac{6.67}{30 \times 10^{-6}}$$

At low input, maximum load conditions, the output of the integrator will change by 6.67 Volts in 30 μs. As the threshold voltage of the MC1455 is 1/3 V_{CC} or 4 Volts, the maximum integrator voltage must be 4 + 6.67 = 10.67 Volts at the beginning of the on time (Vdc shown in Figure 11).

At the other extreme of high input, the Y-Z voltage during the on time of Q4 is

$$+V_{Y\text{-}Z} = (1/6)\ (60\ V) = 10 \text{ Volts.}$$

The integrator output will therefore have a slope of $-\frac{10}{30 \times 10^{-6}}$. The on time of Q4 under the conditions of high input, minimum load, can be calculated from the relationship of energy stored during on time to energy needed during the complete cycle.

If we assume an efficiency of 50% at minimum load of 8 Watts, input power will be 16 Watts. Required energy per cycle is

$$W = P_{in}\ (t_{on} + t_{off}) = 16\ (t_{on} + 25 \times 10^{-6}) \text{ Joules.}$$

Energy stored during Q4 on time is

$$W = \frac{(V_I\ t_{on})^2}{2L} = \frac{(60\ t_{on})^2}{2 \times 130 \times 10^{-6}}$$

As these two energys must be the same, we can equate and solve for t_{on}.

$$\frac{(60\ t_{on})^2}{2 \times 130 \times 10^{-6}} = 16\ (t_{on} + 25 \times 10^{-6})$$

$$3600\ t_{on}^2 = (4160 \times 10^{-6})\ (t_{on} + 25 \times 10^{-6})$$

$$3.6 \times 10^3\ t_{on}^2 - 4.16 \times 10^{-3}\ t_{on} - 1.04 \times 10^{-7} = 0$$

Solving this equation results in $t_{on} \simeq 6\ \mu s$

In 6 μs, with high input voltage, the integrator output will change by

$$-\frac{10}{30 \times 10^{-6}} \times 6 \times 10^{-6} = -2 \text{ Volts.}$$

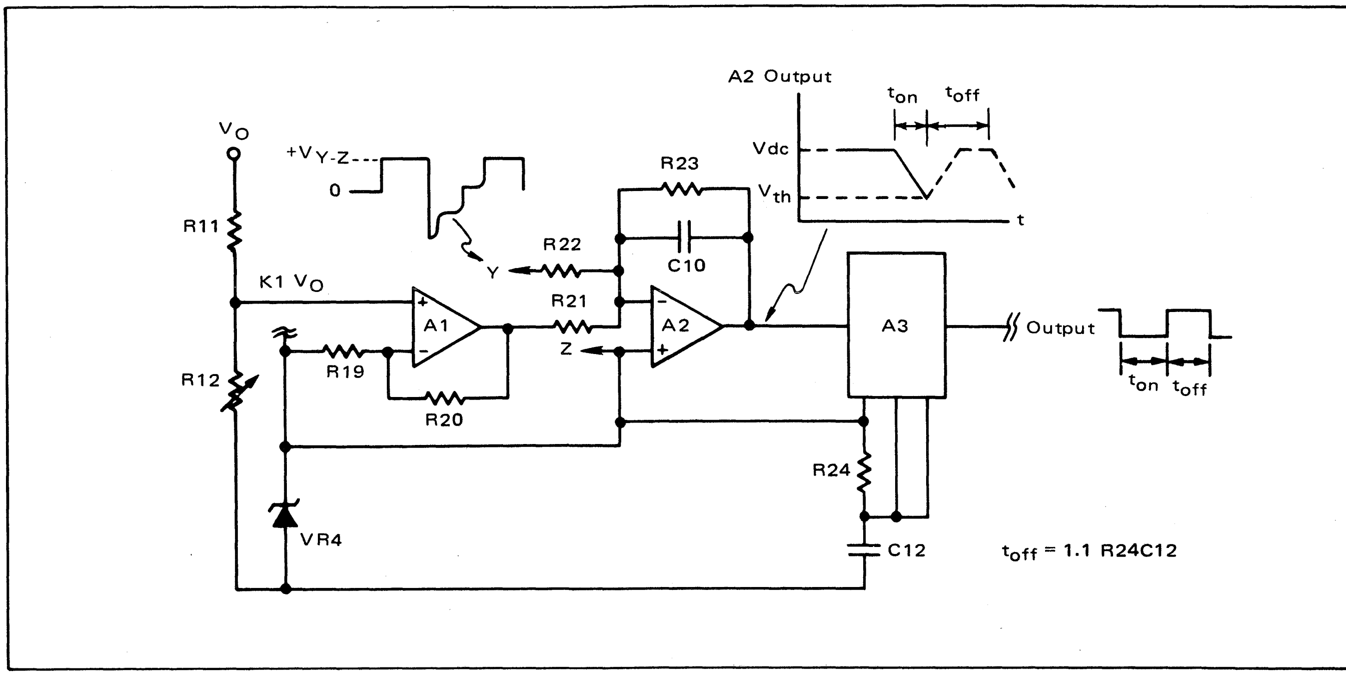

FIGURE 11 — Control Circuit

The maximum integrator voltage under high input, minimum load conditions is therefore 4 + 2 = 6 Volts.

The dc level of the integrator output voltage must change by 10.67 -6 = 4.67 Volts when going from worst case maximum load to worst case minimum load. The dc gain of the control circuit determines how much of a change in V_O will be required to cause 4.67 Volts change at the integrator output. If we choose 1% regulation over worst case line and load, then KV_O, where $K = \frac{R12}{R11 + R12}$, cannot change more than 1%, or 0.12 Volts. Minimum control circuit gain may then be calculated as

$$0.12\,A = 4.67$$
$$A = 38.9.$$

DC gain of the control circuit shown in Figure 4 is 11 x 4.35 = 47.9 which gives regulation of

$$\frac{4.67}{12 \times 47.9} = 0.008 = 0.8\%$$

for worst case line and load changes.

The time constant for the MC1455 is set to equal the off-time which was chosen to be 25 μs. From the MC1455 data sheet and Figure 4

$$1.1\ R24C12 = 25\ \mu s$$

A convenient value can be chosen for one of the components, which fixes the second one. For the circuit of Figure 4, R24 = 2.4 kΩ, C12 = 0.01 μF.

OUTPUT FILTER

The output filter section shown in Figure 12 is required to maintain the high frequency ripple at or below 14 mV peak-to-peak. To reach this level with a single capacitor filter (C5) at heavy loads would require approximately 7500 μF. In this example we chose to reduce the size of C5 and add an LC section (L1 and C6) to achieve the design specification. If 50 mV is allowed across C5 then

$$C5 = \frac{I_{O\,max}\ t_{on(max)}}{V_{Ripple}}$$

$$= \frac{(3A)\ (30 \times 10^{-6})}{50 \times 10^{-3}} = 1800\ \mu F \text{ (used } 2000\ \mu F)$$

The L1C6 section has to attenuate the 50 mV signal to 14 mV or less. C6 should be chosen so that its impedance (X_C) at the operating frequency is much less than the minimum load impedance (R_L). This will reduce the effect of the load on the attenuation process since the impedance of C6 (X_C) in parallel with the load impedance (R_L) will be approximately equal to the impedance of C6. Setting $X_{C(max)} \leqslant 10\%\ R_{L(min)}$ will give reasonable results. For this example

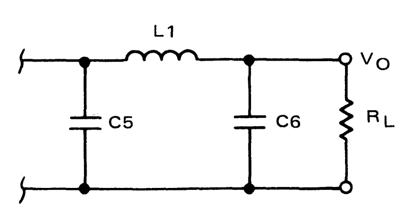

FIGURE 12 — Output Filter Section

$$R_{L(min)} = \frac{V_{O\,min}}{I_{O\,max}} = \frac{20}{3} = 6.67\ \Omega$$

and

$$X_{C(max)} \leqslant (0.1)(6.67) = 0.667\ \Omega$$

solving for C6 from

$$X_{C\,(max)} = \frac{1}{2\,\pi\, f_{min}\, C6\, min}$$

gives

$$C6\ min = \frac{1}{2\,\pi f_{min}\, X_{C6\,max}} = \frac{1}{2(3.14)(18\times10^3)(0.667)}$$

$$= 13.3\ \mu F \text{ (used } 20\ \mu F)$$

L1 can be determined from the following relationship for the required filter attenuation.

$$14\ mV \geqslant \Delta V_O \simeq \frac{X_{C6}}{X_{L1} + X_{C6}} \Delta V_I$$

solving for X_{L1}

$$X_{L1} + X_{C6} = \frac{X_{C6} \Delta V_I}{\Delta V_O}$$

$$X_{L1} = X_{C6}\left(\frac{\Delta V_I}{\Delta V_O} - 1\right) = 0.667\left(\frac{50}{10} - 1\right) = 2.7\ \Omega,$$

assuming $\Delta V_O = 10$ mV instead of 14 mV

and from

$$X_{L1} = 2\pi f L1$$

$$L1 = \frac{X_{L1}}{2\pi f} = \frac{2.7}{2(3.14)(18\times10^3)} = 24\ \mu H \text{ (used } 25\ \mu H)$$

The core for this inductor can be designed using the same procedure used for the power transformer. Since $I_{max} = I_{dc} + I_{ac} \cong 3$ A, #16 gauge wire was chosen. Then:

$$^*A_C A_{CB} = \frac{A_X\, L\, I_{max}}{0.8\, B_{max}} \times 10^8\ (cm^4)$$

where A_X = cross sectional area of the wire in cm^2

*The equations used in this approach are from Reference 7.

$$A_C A_{CB} = \frac{(0.013\ cm^2)(25\times10^{-6}h)(3\ A)}{(0.8)(2000\ g)} \times 10^8\ cm^4$$

$$= 0.06\ cm^4$$

From Table II, core #2213-L00-3B7 should work. The number of turns can be calculated from

$$N = \frac{L I_{max}}{A_e B_{max}} \times 10^8 = \frac{(25\times10^{-6})(3\ A)}{(0.635)(2000)} \times 10^8$$

$$= 5.9 \text{ (used 6 turns)}$$

checking the fit:

$$A_{CB} \geqslant \frac{N}{T16} = \frac{6}{327} = 0.018 \text{ in. sq.}$$

Since this A_{CB} for the 2213 core is 0.046 in. sq., this winding should fit.

Checking the magnetic path length required,

$$l_e = \frac{(0.4)\pi N I_{max}\, \mu_{AV}}{B_{max}} = \frac{(0.4)(3.14)(6)(3)(1900)}{2000}$$

$$= 21.5\ cm$$

$$l_e = l_m + \mu l_g$$

$$21.5\ cm = 4.5\ cm + (1900)\ l_g$$

$$17 = 1900\ l_g$$

$$l_g = 0.0089\ cm = 0.002\ ins$$

PERFORMANCE AND CONCLUSIONS

The overall performance of the circuit is summarized in Table IV

TABLE IV –

Output Current – 0.3 A to 3 A
Output Voltage – 20 to 27 Vdc (Adjustable)
20 kHz Ripple – < 15 mV Peak-to-Peak
120 Hz Ripple Rejection – > 60 dB (<14 mV Peak-to-Peak)
Load Regulation – < 1%
Line Regulation – < 0.5%
Efficiency – > 75%

The excellent 120 Hz ripple rejection can be attributed to the control technique used, which senses both the output and input voltage. The circuit performed well to an ambient of 70°C. This could be improved with a change in core material and/or shape, additional heat sinking and possibly a selection of critical devices.

As was mentioned in the introduction, the circuit can be designed to operate over a wider range of input voltages with certain trade offs. Three possible approaches to accomplish this and their effects on the operation are listed below.

1. USING A SIMILAR TRANSFORMER DESIGN; at lower input voltage, the minimum operating frequency would be lower and the maximum operating current would be higher. This would require a redesign of the drive circuit, a higher current power transistor, an increase in the off time to insure that secondary completely discharges, some modification to the transformer to prevent saturation and larger filter elements.
2. T1 COULD BE REDESIGNED TO OPERATE AT THE SAME CURRENT LEVEL, BUT AT A MUCH LOWER FREQUENCY WITH LOWER INPUT VOLTAGES. This would require either an increase in off-time or a change in winding ratio.
3. T1 COULD BE REDESIGNED TO OPERATE AT THE SAME CURRENT LEVEL AND FREQUENCY WITH LOWER INPUT VOLTAGES. This would require a reduction in off time, a higher turns ratio, higher voltage capability for the power transistor, higher peak currents in the secondary, larger output filter elements and higher frequency operation at high input voltages.

The second approach appears to offer the least amount of compromise. The major disadvantage is the lower frequency of operation.

In other applications such as line operation and lower output power systems the basic circuit could be scaled up or down depending on the requirements. Line operation may be possible with the same power device (2N6546) used in this application as it has high voltage blocking capability. The maximum operating current in a line operated system would be much lower for equivalent output power requirements because of the high input voltage. This would also reduce the drive requirements for the power transistor. On the other hand, higher output power could be achieved with a line operated version if the operating current is held constant (200 to 300 Watts may be achievable). The voltage capability of the input bridge and portions of the control circuitry would have to be increased for line operation. Various resistors would have to be increased to limit currents and power dissipation.

In summary, the basic approach is very flexible and can be tailored to specific system requirements by following the design approach shown in this application note and adjusting component values to meet the design specifications.

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