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Analysis of Switching Regulator Power Supply

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ANALYSIS OF SWITCHING REGULATOR POWER SUPPLY

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CHAPTER 1

INTRODUCTION TO THE POWER SUPPLY

Every electrically driven load requires a stable power source, whether the load is a sensitive dc motor, a scanning electron microscope, X-ray equipment, a computer, or measuring instrumentation. Oddly enough, power supplies are usually the last item considered when designing or purchasing equipment, despite the fact that it is impossible to operate any electronic load without one.

In addition, it is impossible to get optimum performance from a load coupled to a mismatched power supply. This kind of relationship breeds some undesirable phenomena-overheating, arcing, spurious noise generation, annoying overload shutdowns, even blowing up of load or power supply. Worse, inadequate stability in an incorrectly specified power supply can completely mask the specified accuracy or sensitivity of the driven load.

The diversity of power supplies available makes it almost possible to have a specific unit on hand (or at nearly off-the-shelf readiness) for any given load and space requirement. Therefore, the end-user is in the unique position of having as many choices as time permits him to consider. And if the kind of power supply needed cannot be selected from stocked items, a custom power supply designer will build exactly what is required. Obviously, the plenitude of possibilities means the user must decide which power supply most closely meets his needs, and for a decision some foreknowledge is necessary.

The regulator is the stabilizing heart of the power supply - the electronics that maintains the output voltage within its specified limits, no matter what happens on the line or at the load. Regulator types used in low-voltage dc power sources are linear (series pass), switching, and ferroresonant.

As the least expensive of the three major power regulator types, the ferroresonant regulator has few parts (which means high reliability), good input/output isolation (4 to 5kV) and built-in overload and short-circuit protection. On the negative side, the ferroresonant regulator has 1.) bulky, heavy power transformer, limiting its minimum size; 2.) poor output voltage regulation with line frequency change, 3.) poor load regulation with varying load current requirements, 4.) and a large leakage flux field, which can create magnetic field pickup problems. Ferroresonant regulators are best applied where load current demand remains relatively constant.

Medium-priced with wide application, the conventional linear, series-pass regulator has many positive attributes. It provides excellent (0.1 per cent) regulation, produces low ripple and output noise, and shows high reliability. Its main disadvantage is its low efficiency, because it is inherently a heat-dissipative system. The series regulator usually contains paired collector-coupled Darlington transistors, which supply the load current. Semiconductor technology has brought series regulators and over-voltage protectors to minimum size as monolithic components such as TO3 cans, but these must be mounted on large heat sinks, because heat is the killer of semiconductors; this limits the ultimate output wattage of the devices. A 50-60 Hz power transformer, the input to the series, limits the minimum size of this type of power source. Series regulators are most competitive and efficient below the 100 Watt output level at high voltage with low current, but switching regulators will be increasingly utilized for low-voltage, high-current applications, due to their superior operating efficiency.

Switching power technology developed as a solution to problems arising from a need to shrink size and weight of equipment, the special power needs of miniaturized "smart" instrumentation, and the growing demand for increased efficiency and energy conservation. The switching regulator (switcher) can operate at 80 per cent efficiency, compared with about 30 per cent for conventional linear regulators. The switched regulator delivers more input power to the load and dissipates less heat; thus, switchers can tolerate greater component packing density and still produce up to four times the wattage in the same space as conventional linear types. However, switching regulators cost more; the technology is still young, and there are some inherent problems. Being more complex, the switching regulator is generally the least reliable of the regulators. Because the switching regulator utilizes a square wave generator/chopper comprising transistors operated in the switching mode (usually running at 20 to 30 kHz), unavoidable high-frequency noise is by-product. This should be considered when driving any microvolt-sensitive load, such as communications equipment. The most common method of achieving voltage regulation with the switcher is by varying the on-off time ratio of the switching transistors. The output of the transistors is fed to a small high-frequency ferrite transformer, rectifier, choke and filter (there is no heavy line-side transformer in a switcher). The switcher provides good input/output isolation and fairly good regulation.

The switching power supply's increased efficiency, reduced size, and lower weight are strong inducements that are persuading customers to pay up to 30 per cent more for a switcher at that level power. Semiconductor economics, in the form of lower-price power transistors and control circuits, is shrinking the price, which is being pressed by spiraling materials costs (copper and steel), and assembly costs. The crossover point for switching power supplies is nearing the 100 Watt level, indicating that it is now at about 200 Watts, down from 300 Watts few years ago.

Switching regulators have become increasingly popular in new equipment designs, not only in aerospace and defence applications, but in computers, industrial process control systems, instrumentation, and communications. Several studies indicate that half of all ac/dc power supplies will be switchers by 1980.

CHAPTER 2

ANALYSIS OF INDUCTIVE SWITCHING CIRCUITS

There are a number of useful circuit configurations which utilize inductive switching. Depending upon load requirements, we can choose the proper configuration. In this chapter, the single ended switching regulator, which is most efficient and also requires least components, will be discussed.

The basic configuration of the switching regulator is shown in Fig. 2.1. The control circuit causes the switcher to switch ON and Off, with pulse-width modulation of the resulting square wave. Two models of the operation are possible: 1.) OFF time remains constant while the ON time is changed to correspond to the varying load power, thus causing the frequency to vary. 2.) The frequency is fixed while the proportion of ON time is varied.

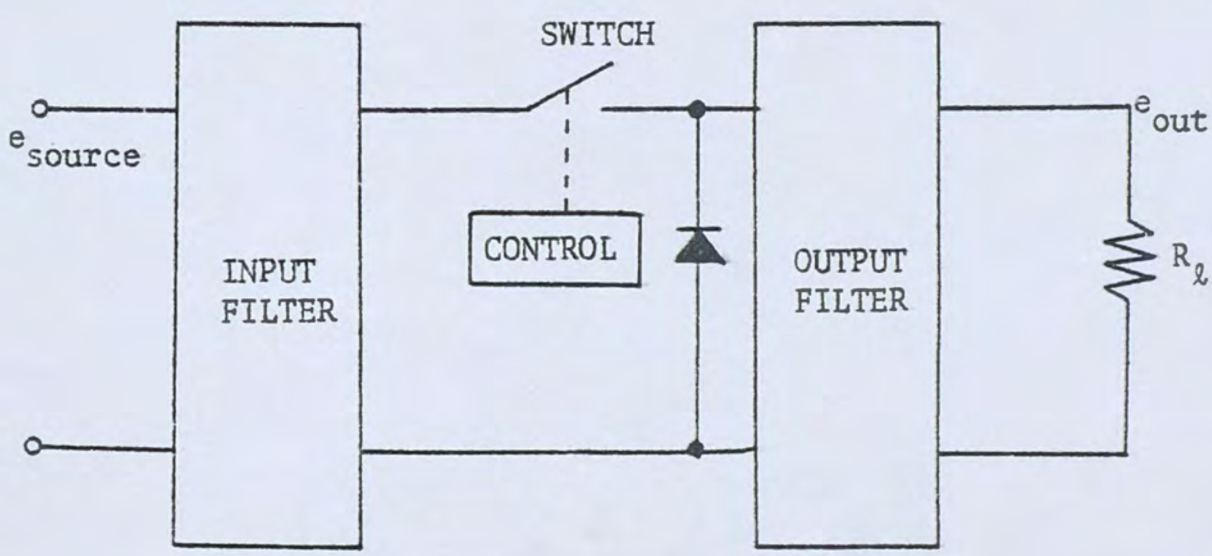


Fig. 2.1 Basic Switching Regulator

The pulse-width modulated wave is then filtered through a low-pass filter, and the resulting direct current is controlled in amplitude. Ideally the efficiency of such a regulator is 100%. In actual practice, the efficiency of the regulator is limited by losses in the switch and inductor, but these losses are in principle considerably less than those found in most other regulating approaches.

The input filter in the switching regulator design serves three purposes: to smooth output spikes and high frequency transients with large peak values and small volt-second integrals; to eliminate input ripple having frequency components at or near the modulating frequency of the switcher (which would produce low frequency components by heterodyning); and to attenuate AC components produced by transistor switching. The switcher chops the DC output of the input filter in such a way as to deliver constant volt-second energy pulses to the intergrater.

2.1 OUTPUT FILTER

There is a wide variety of filters available with which to improve the switcher output waveforms. The objective of the filter at the output of a switcher is the reduction or attenuation of harmonics appearing at the load. The general approach is to provide a shunt path for the harmonic currents with a series impedance across which the harmonic voltages appear. This general arrangement is shown in Fig. 2.2.

The attenuation of any given harmonic depends on the ratio of the parallel load-shunt impedance to the total impedance at that frequency, that is

$$\frac{E_\ell}{E_i} = \frac{Z_o}{Z_1 + Z_o} \tag{2.1}$$

$$\text{where } Z_o = \frac{Z_2 Z_\ell}{Z_2 + Z_\ell}$$

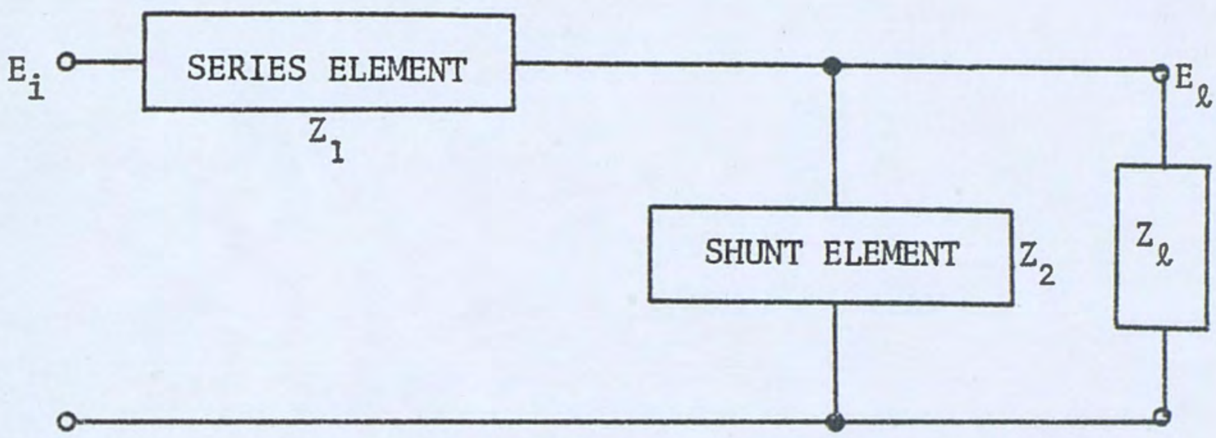


Fig. 2.2 Basic Configuration of Output Filter

The simplest form of an efficient filter is the single section L-C filter shown in Fig. 2.3. This filter consists of series inductance and a shunt capacitance.

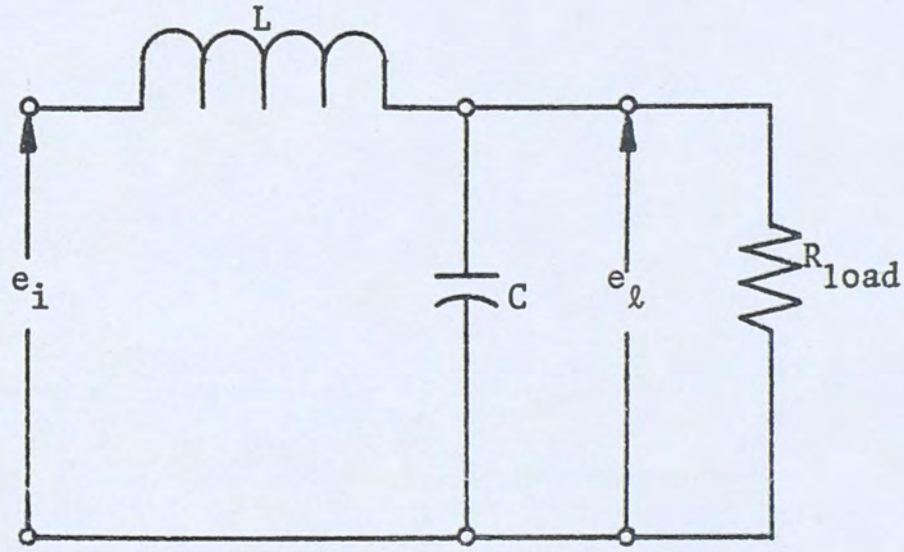


Fig. 2.3 Single L-C Section Filter

The transfer function for the general form of the single section L-C filter will be developed. Assume the load is purely resistive and L and C are loss-less.

$$\frac{E_\ell}{E_i}(s) = \frac{\dfrac{\frac{1}{sC} R_\ell}{R_\ell + \frac{1}{sC}}}{sL + \dfrac{\frac{1}{sC} R_\ell}{R + \frac{1}{sC}}} = \frac{1}{s^2CL + sL/R_\ell + 1} \qquad (2.2)$$

or

$$\frac{E_\ell}{E_i}(j\omega) = \frac{1}{-\omega^2LC + j\omega L/R_\ell + 1} \qquad (2.3)$$

Let's consider the standard transfer function of the second order system

$$\frac{E_\ell}{E_i} = \frac{1}{s^2/\omega_n^2 + 2\delta s/\omega_n + 1} \qquad (2.4)$$

where ω_n: resonant frequency

δ: damping ratio

The general form of the frequency response

$$\frac{E_\ell}{E_i}(j\omega) = \frac{1}{-\omega^2/\omega_n^2 + j2\delta\omega/\omega_n + 1} \qquad (2.5)$$

Comparing this equation to (2.3) we see that

$$\omega_n^2 = \frac{1}{LC} \qquad (2.6)$$

$$\delta = \frac{1}{2R}\,\frac{L}{C} \qquad (2.7)$$

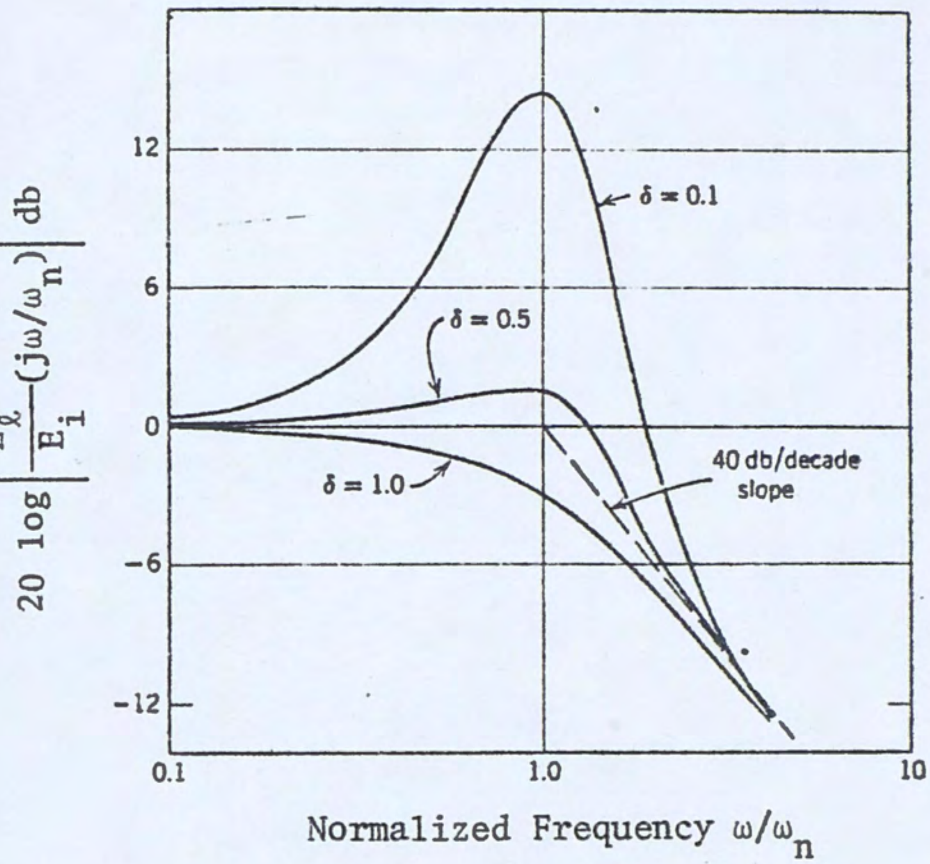


Fig. 2.4 Gain of Filter vs. Frequency

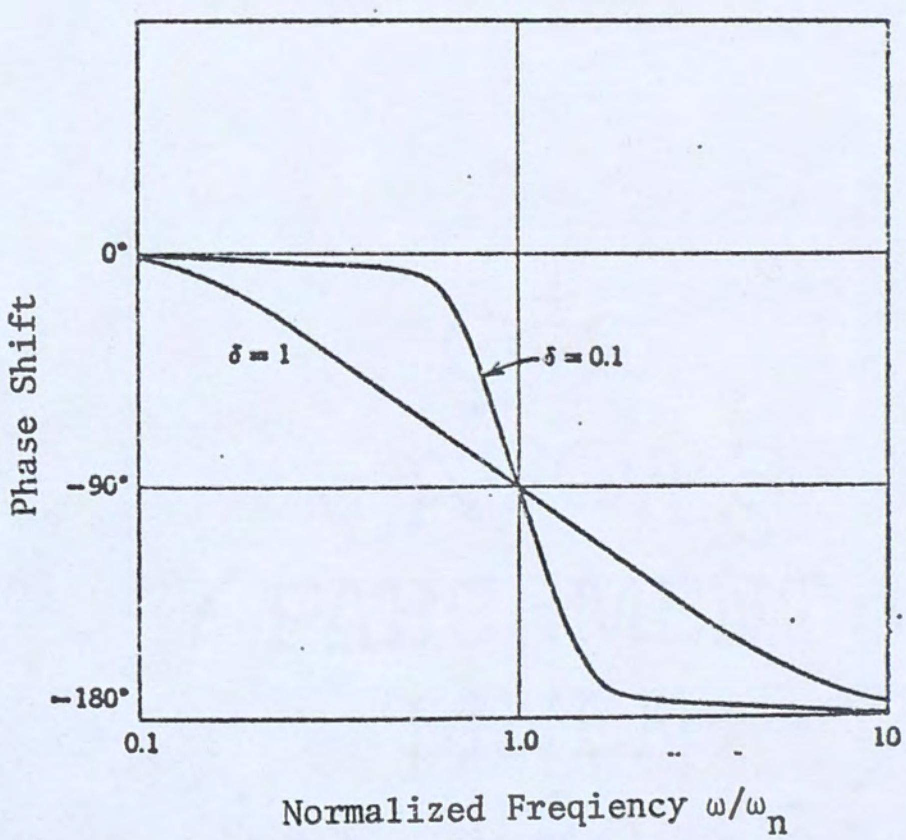


Fig. 2.5 Phase Shift of Filter vs. Frequency

The log of the magnitude of E_ℓ/E_i from Equation 2.5, plotted against ω/ω_n, the non-dimensional frequency term, on a log scale, is the frequency response curve. The frequency response with various values of damping ratio, δ, is shown in Fig. 2.4. Note that at low values of frequency, $\omega/\omega_n << 1$, the magnitude of the transfer function, is unity $\{\log E_\ell/E_i(j\omega/\omega_n) = \log 1 = 0\}$. The rate of increase in attenuation with increasing frequency, above $\omega/\omega_n = 1$, approaches 40 db per decade.

The phase shift through the filter as a function of frequency is shown in Fig. 2.5 in the same non-dimensional form as used in Fig. 2.4. There are two curves shown for different values of δ. The phase shift of the single section L-C filter is zero at very low frequencies, and approaches 180° at high frequencies.

In general, when an L-C filter sufficiently attenuates the lowest harmonic present in the output, the higher harmonics are automatically adequately reduced to a permissible level. This results because a practical filter is usually designed so that its resonant frequency ω_n is below the lowest harmonic to be attenuated. Thus, the harmonics present occur in the frequency range where the filter attenuation curve has a 40 db/decade slope as shown in Fig. 2.4 (normalized frequency > 2). The form of the response curve indicates that the attenuation at a given frequency is determined by this frequency's relation to the resonant frequency of filter.

Where higher values of harmonic attenuation are required, cascaded L-C filters become attractive. This general trend continuous with an increasing number of stages of filters. In most practical cases, however, the need of harmonic attenuation is satisfied with not more than a two-stage filter.

Before entering into the detailed analysis of the two-stage filter, it is interesting to consider some factors which may influence the decision of whether to use a cascaded filter. First, servo-system stability considerations are important. When the filter involved is included in a closed-loop voltage regulating system, the presence of two L-C networks can cause a great deal of difficulty in stabilizing. Another consideration is the over-all size and construction cost of the filter. Although a cascaded filter may appear theoretically desirable at one value of attenuation to minimize the size and construction, the cost of the extra components may elevate the actual value of attenuation at which cascaded filters become advantageous. The transfer function for the two stage filter, shown in Fig. 2.6, is

$$\frac{E_\ell}{E_i}(s) = \frac{1}{(s^2C_1L_1 + 1)\ (s^2L_2C_2 + \dfrac{sL_2}{R_\ell} + 1) + \dfrac{sL_1}{R_\ell} + s^2L_1C_2} \tag{2.8}$$

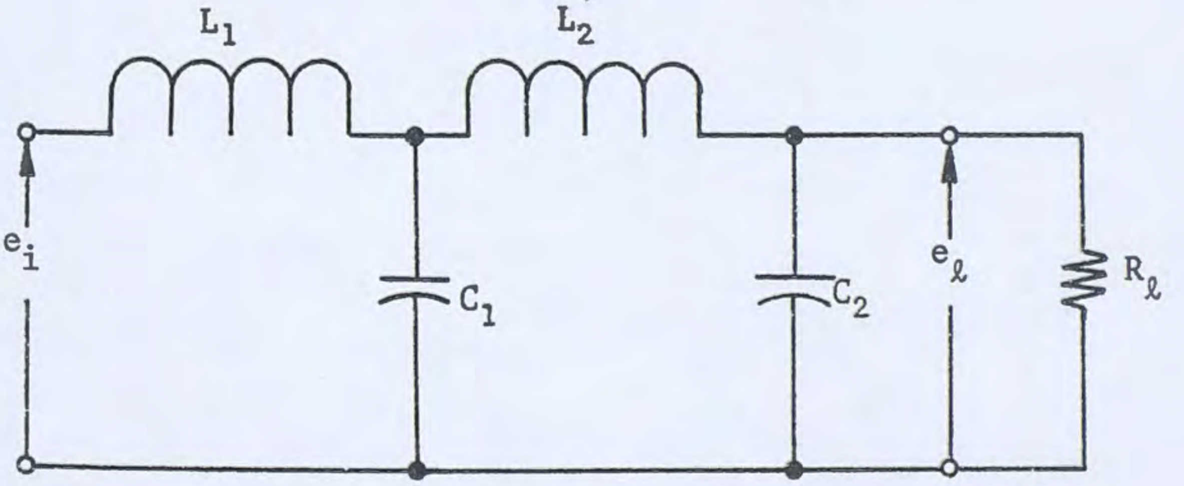


Fig. 2.6 Cascaded L-C Filter

To determine the attenuation at which the two-stage and single-stage filters are equivalent, assume that the cascaded filters, shown in Fig. 2.6, have equal values of inductances and capacitances, each equal to half the values in the single section components.

L_1 = inductance of single stage filter

C_1 = capacitance of single stage filter

$L_{11} = L_{22}$ = inductance of two-stage filter

$C_{11} = C_{22}$ = capacitance of two-stage filter

$$L_{11} = L_{22} = L_1/2 \tag{2.9}$$

$$C_{11} = C_{22} = C_1/2 \tag{2.10}$$

The transfer function of a single L-C filter in the unloaded case is

$$\frac{E_\ell}{E_i}(s) = \frac{1}{s^2L_1C_1 + 1} \tag{2.11}$$

The transfer function of the cascaded L-C filter in the unloaded case, from (2.8), (2.9), and (2.10) is

$$\frac{E_\ell}{E_i}(s) = \frac{1}{\left(\frac{s^2C_1L_1}{4}\right)^2 + 3\left(\frac{s^2L_1C_1}{4}\right) + 1} \tag{2.12}$$

Expressing the transfer function (2.11) and (2.12) in terms of $j\omega$, and making use of relationship between L_1, C_1 and resonant frequency $\omega_n\left(\omega_n^2 = \frac{1}{L_1C_1}\right)$, we get

single: $$\frac{E_\ell}{E_i}(j\omega) = \frac{1}{-\frac{\omega^2}{\omega_n^2} + 1} \tag{2.13}$$

cascaded:
$$\frac{E_\ell}{E_i}(j\omega) = \frac{1}{\frac{\omega^4}{16\omega_n^4} - \frac{3\omega^2}{4\omega_n^2} + 1} \tag{2.14}$$

To determine the value of attenuation at which the two filters are equivalent, the magnitude of the two transfer functions are equated.

$$\left| -\frac{\omega^2}{\omega_n^2} + 1 \right| = \left| \frac{\omega^4}{16\omega_n^2} - \frac{3}{4}\frac{\omega^2}{\omega_n^2} + 1 \right| \tag{2.15}$$

When ω is greater then ω_n, the left hand term must be written $\omega^2/\omega_o^2 - 1$ so that both side of the equation will produce positive results

$$\frac{\omega^2}{\omega_n^2} - 1 = \frac{\omega^4}{16\omega_n^2} - \frac{3}{4}\frac{\omega^2}{\omega_n^2} + 1$$

$$\frac{\omega^4}{16\omega_n^4} - \frac{7}{4}\frac{\omega^2}{\omega_n^2} + 2 = 0 \tag{2.16}$$

Solving for the ratio ω/ω_n, we get

$$(\omega/\omega_n)^2 = 26.8 \text{ or } 1.2 \tag{2.17}$$

This indicates that the two filters are equivalent at two points. Actually, the lower number arises because of equivalence in the resonant peaking region. Substituing $(\omega/\omega_n)^2 = 26.8$ into (2.13) or (2.14), we have

$$\frac{E_\ell}{E_i}(j\omega) = \frac{1}{-25.8}$$

$$\frac{E_\ell}{E_i}(j\omega) = \frac{1}{25.8\underline{/180^\circ}} \tag{2.18}$$

Equation 2.18 indicates that in filter applications where attenuations of up to approximately 26 : 1 are required, the single-stage filter is most desirable. In cases where attenuations greater than 26 : 1 are required, the cascaded filter should be considered. It is important to note that these conclusions resulted for this specific case where loading on the output of the filters was neglected and where the cascaded filter sections were assumed identical to each other. In addition, it should be reemphasized that many practical considerations will affect the decision of whether single section or cascaded filters should be used for particular applications.

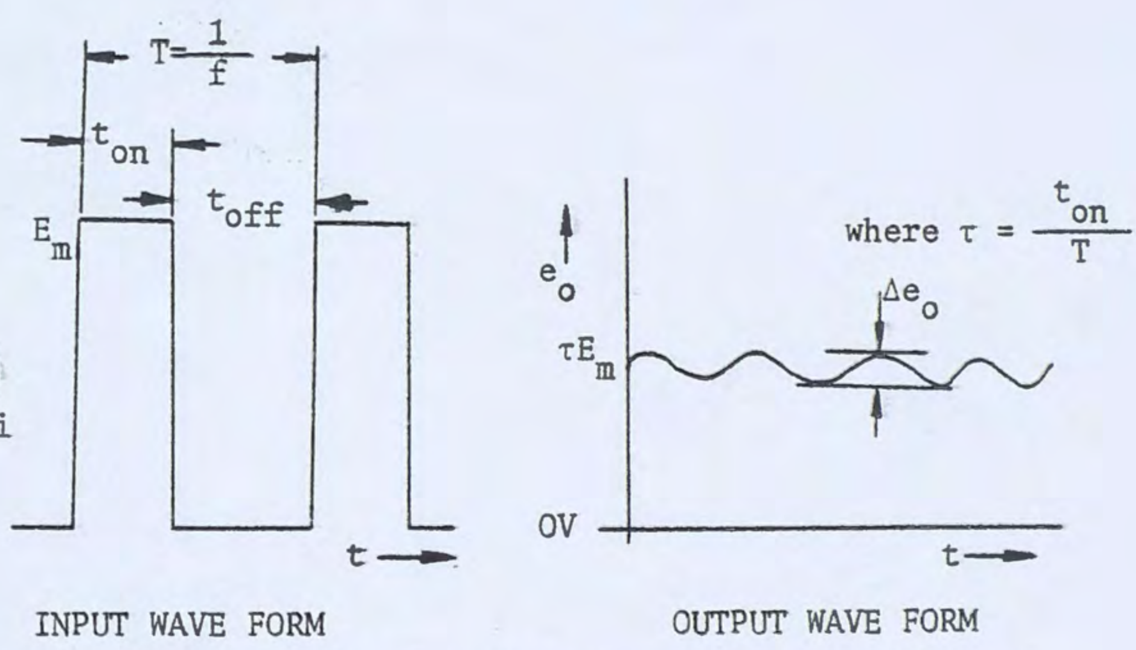


Fig. 2.7 Input and Output Wave Forms of Output Filter

2.2 OUTPUT RIPPLE AND STEP RESPONSE

A squarewave voltage is produced by switching action. The dc voltage is switched alternately from ON to OFF, producing a square wave voltage output to the output filter. The input and output wave forms are shown in Fig. 2.7.

The input square wave form can be written as an odd periodic function, with the form

$$e_i(t) = \frac{a_o}{2} + \sum_{n=1}^{\infty} b_n \sin n\omega t \qquad (n = 1,2,3,\ldots) \tag{2.19}$$

$$\text{where } b_n = \frac{\omega}{\pi} \int_{-\pi}^{\pi} e(t) \sin n\omega t \, dt \quad (n = 1,2,3,\ldots) \tag{2.20}$$

We apply the square wave with three harmonic components to the single section L-C filter. And, now we write the output voltage $e_o(t)$ in the form

$$e_o(t) = E_o^{(0)} + E_o^{(1)} \sin \omega t + E_o^{(2)} \sin 2\omega t + E_o^{(3)} \sin 3\omega t \tag{2.21}$$

Using Equation 2.3, we evaluate the transfer function at the first, second, and the third harmonics of the series, obtaining

$$N(j\omega) = \frac{1}{1 - \omega^2 LC + j\omega L/R} \tag{2.22}$$

$$N(2j\omega) = \frac{1}{1 - 4\omega^2 LC + j2\omega L/R} \tag{2.23}$$

$$N(3j\omega) = \frac{1}{1 - 9\omega^2 LC + j3\omega L/R} \tag{2.24}$$

And we can get

$$E_o^{(0)} = \tau E_m$$

$$E_o^{(1)} = N(j\omega) \left(\frac{b_1}{\pi} \angle 180^\circ\right) \tag{2.25}$$

$$E_o^{(2)} = N(j2\omega) \left(\frac{b_2}{\pi} \angle 180^\circ\right) \tag{2.26}$$

$$E_o^{(3)} = N(j3\omega) \left(\frac{b_3}{\pi} \angle 180^\circ\right) \tag{2.27}$$

Finally we get the output,

$$e_o(t) = \tau E_m$$

$$+ \tau\frac{E_m}{\pi} \quad c_1 \frac{1}{\sqrt{(1 - \omega^2LC)^2 + (\omega L/R)^2}} \cos\,(\omega t + \angle A_1)$$

$$+ c_2 \frac{1}{\sqrt{(1 - 4\omega^2LC)^2 + (2\omega L/R)^2}} \cos\,(2\omega t + \angle A_2)$$

$$+ c_3 \frac{1}{\sqrt{(1 - 9\omega^2LC)^2 + (3\omega L/R)^2}} \cos\,(3\omega t + \angle A_3) \quad (2.28)$$

Hence, the dc component ripple of the first harmonic is

$$\frac{(\Delta e_o)\ \text{1st harmonic}}{e_{dc}} = \frac{c_1/\pi}{\sqrt{(1 - \omega^2LC)^2 + (\omega L/R)^2}} \quad (2.29)$$

For typical values of ω, L, C and R, second and third harmonic terms are so much smaller as to be negligible. Considering Equation 2.29, we see the ripple is

$$\frac{\Delta e_o}{e_{out}} \simeq \frac{1}{\omega^2LC} \quad (2.30)$$

Thus we see ripple can be reduced by high frequency operation or large product of LC.

Let us apply the unit step function to the second order system, which has the standard form as expressed in Equation 2.4.

The observed response to the unit step is

$$e_o(t) = 1 - \frac{1}{\sqrt{1 - \delta^2}}\, e^{-\delta\omega_n t} \sin\,(\sqrt{1 - \delta^2}\,\omega_n t + \Phi) \quad (2.31)$$

$$\text{where } \Phi = \tan^{-1}\left(\frac{\sqrt{1 - \delta^2}}{\delta}\right)$$

Fig. 2.8 shows a plot of $e_o(t)$ as a function of normalized time $\omega_n(t)$ for various of the damping ratio, δ. Considering Equation 2.7. which is $\delta = \frac{1}{2R}\frac{1}{C}$, and Fig. 2.8, we see that no-load operation or low L/C ratio are undesirable, causing large (sometimes intolerable) overshoot in input current and output voltage, especially at startup (when the circuit is first energized).

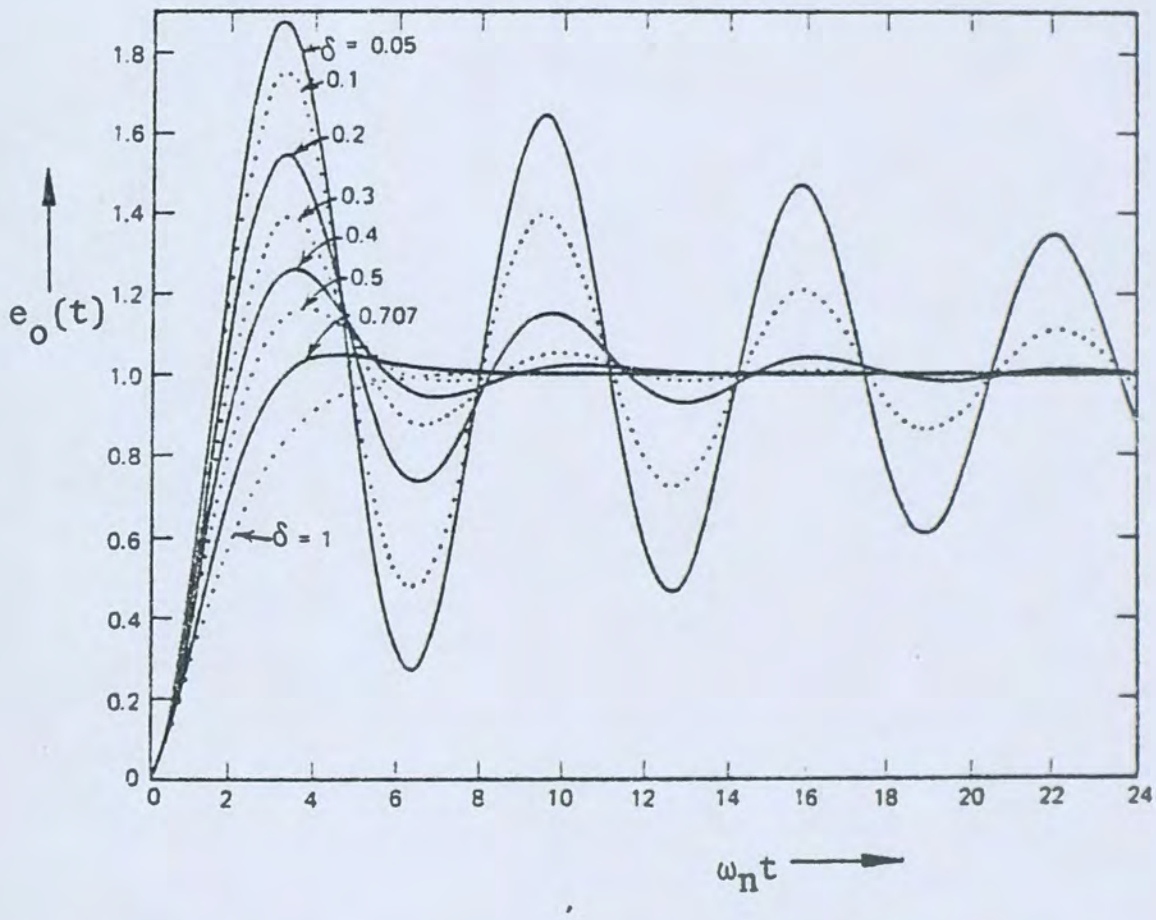


Fig. 2.8 Step Responses of The Second Order System

It is relatively simple to determine the required L-C product to attenuate a harmonic a given amount. However, the values of L and C are still undetermined. The required trade-off, a study of the interrelation between the values of L and C their affect on the switcher, will be disscussed in Chapter 3. Although the results may not provide a means of selecting the optimum values of L and C for a specific situation, they will indicate the trends present so that intelligent choice may be made in any particular application.

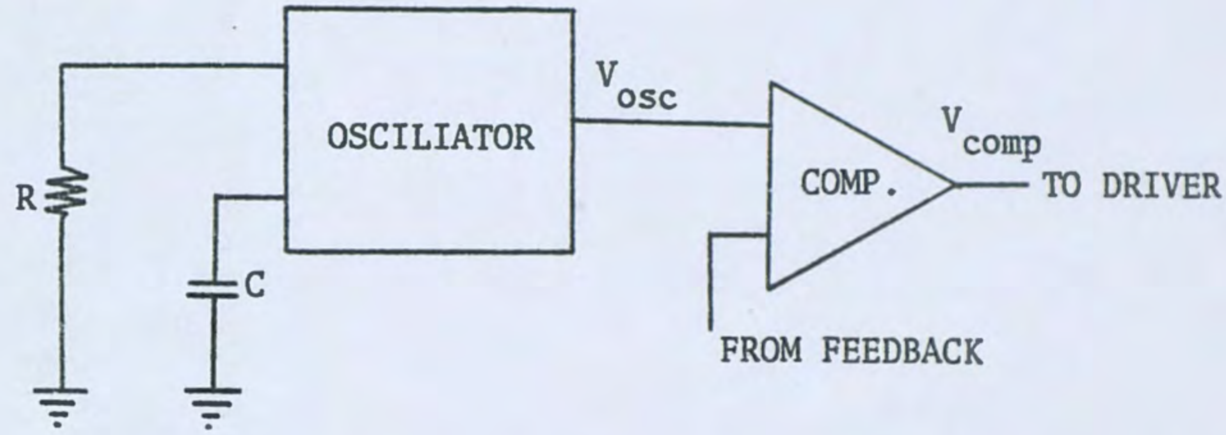
2.3 CONTROL CIRCUIT

Various methods exist for the control of voltage amplitude by means of time modulation of the switched waveform. Two basically different types of modulation may be used; both are known as pulse-width modulation, although they are in no way similar. For the purposes of this discussion, the two types will be referred to as "pulse-duration" modulation and "pulse-frequency" modulation.

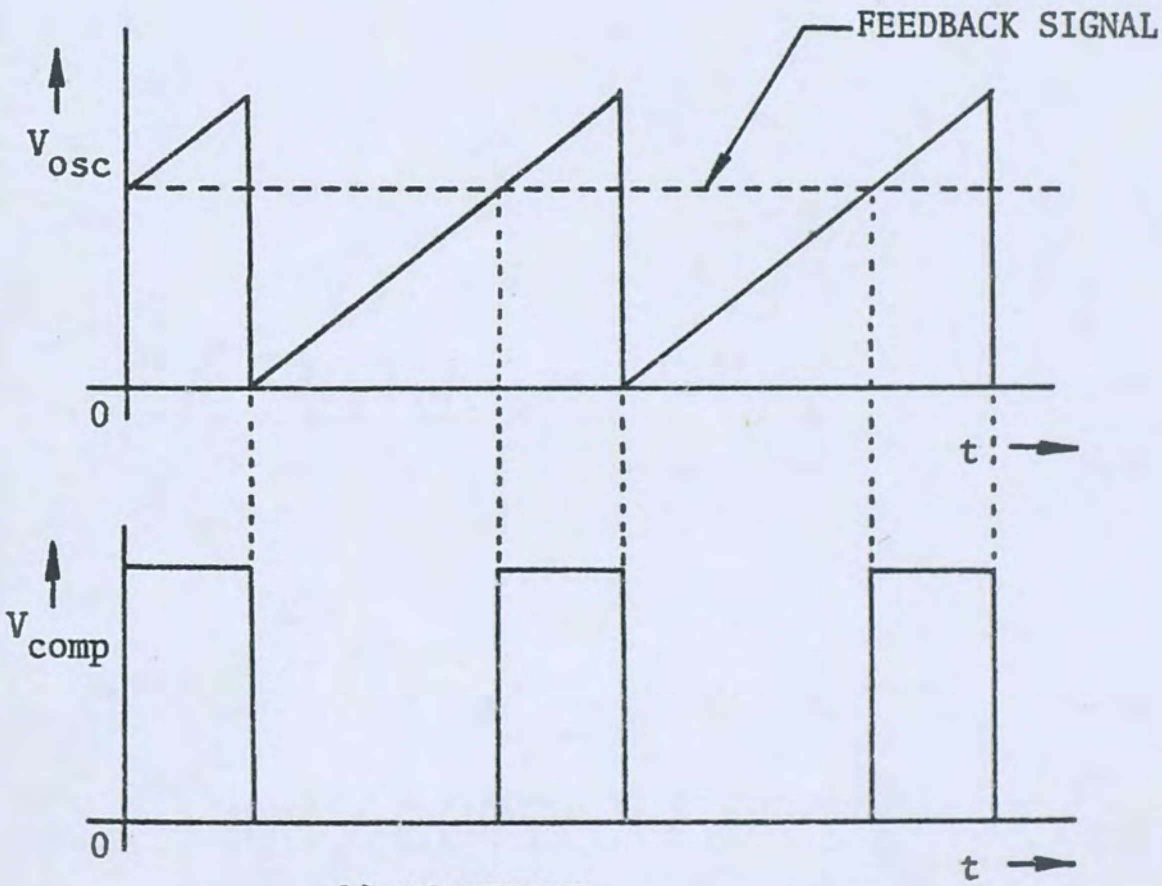
When the modulation technique is adopted for dc/dc conversion, its operation is not disturbed by dynamic variations of the input voltage. Large input voltage transients cause only small variations in the output voltages, regardless of the frequency content of the transients.

One possible method for the basic pulse width modulation is shown in Fig. 2.9. Basically it consists of oscillator and comparator. The oscillator in the circuit in Fig. 2.9a uses an external resistor, R, to establish a constant charging current into an external capacitor, C. This constant current charging gives a linear ramp voltage. Then the comparator takes this output and compare it to the feedback signal to produce a variable duty cycle output pulse for the power switch.

The timing diagram in Fig. 2.9b 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, the comparator switches to a high output level. The comparator is reset when the sawtooth drops back below the feedback signal. The output of comparator is a square wave as shown.



a) CIRCUIT DIAGRAM



b) WAVEFORMS

Fig. 2.9 Basic Pulsewidth Modulator

CHAPTER 3

DESIGN CONSIDERATIONS

This chapter will outline the procedures used in the design of the switching regulator. It is in this area that power supply design engineers tend to disagree. The rules for selection of parameter values are various depending on applications. In military applications, the designer usually find himself in situations where he is forced to design for minimum size and weight, maximum possible efficiency, extremes of operating temperature, minimum interference or line noise, or combinations of these. Each of these restrictions can dictate ground rules for component selection, since the optimizing of one condition may call for trade-offs in the others. This chapter is a summary of various design procedures, and evaluations made of some of these trade-offs.

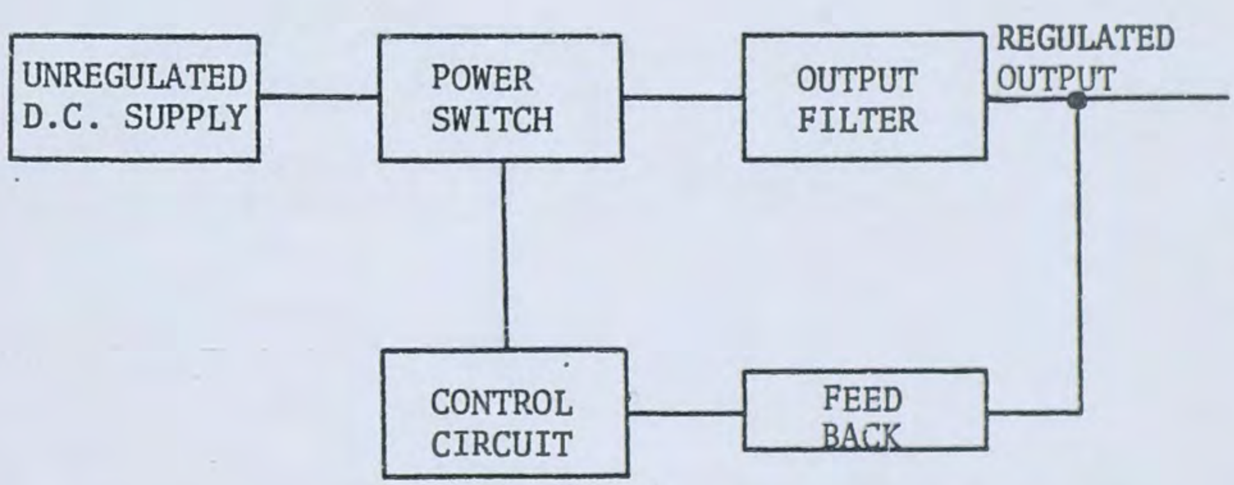


Fig. 3.1 Switching Regulator Block Diagram

This is usually chosen to be somewhere between 5 kHz and 100 kHz, depending upon size and weight requirements and availability of transistors and diodes with high switching speeds for the power required. As the operating frequency is increased, semiconductors must be capable of faster switching, since the losses during switching become a greater portion of regulator losses. The cost of the proper semiconductors also increases with operating frequency, and efficiency usually decreases. These disadvantages are offset, however, by the reduction in size of the choke and capacitor needed. In general, operating frequencies are chosen between 10 kHz and 40 kHz, since for average applications these offer the best compromise with higher current. Normally the frequency be greater than 20 kHz to eliminate noise at audible pitch.

3.2 INPUT SOURCE AND INPUT FILTER

The source of power is specified and would not be one of the design choices. When output power is 500 Watts or less, the single phase 120 Volt line is generally used. For larger supplies, three-phase power from a 208 V. or 220 V. source is used because energy storage in filter capacitors is more economical at these higher voltages. Tolerances on line voltage are typically ±10%. Operation beyond these points can be prevented by voltage crowbars and shutdown circuits.

The input rectifier and filter consists of a bridge rectifier and filter capacitor. Unlike designs using the series regulator, larger input variation is permissible since the switcher regulates by switching the transistor to either the ON or OFF condition.

Since many switching supplies are operated directly off the rectified high Vac line with capacitive input filters, some means of preventing rectifier failure due to inrush surge currents is usually necessary. As shown in Fig. 3.2, a series resistor, Rs, is used to provide inrush current limiting. After the filter capacitor is charged, the SCR conducts (shorts Rs out of the circuit), eliminating its otherwise large power dissipation.

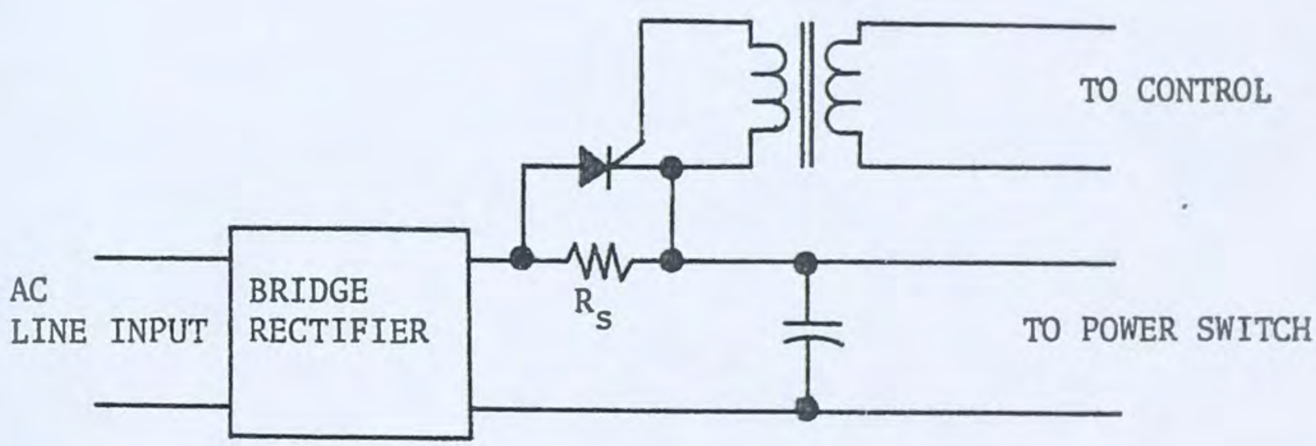


Fig. 3.2 Input Section

3.3 SWITCHER

The performance capabilities of the power supply are determined by the switcher, because they are the parts least able to withstand overloads, such as those caused by load faults or misuses. Because of their efficiency and reliability, transistor switches are ideally suited to the manipulation and control of large amounts of power.

The primary considerations when selecting a power transistor for switch-mode applications are voltage and current ratings, switching speed, and energy handling capability. In this section, these specifications will be discussed.

Switching time is a factor in determination of the maximum power that the transistor is capable of dissipating. The switching times, t_r(rise time) and t_f(fall time), are of prime consideration in selection of a transistor to be used as the switch. The frequency cut-off characteristics of the transistor must be high compared to the actual switching frequency. If the transistor cannot switch rapidly between the states of saturation and cut off, excessive junction heating will result. Therefore the frequency cut off characteristics of the transistor should be from five to ten times the frequency of oscillation used in the power supply.

Collector breakdown voltage should be higher than the supply voltages. When the transistor turns off, the inductive load maintains current through the transistor or load while the collector to emitter voltage V_{ce} rises to the value of the input voltage V_{in}. This condition alone results in a power pulse which is increased by the additional pulse created because the diode does not conduct immediately when the voltage across it reverses.

The collector to emitter saturation voltage is also important because it determines part of the power loss in the circuit, namely the dissipation of transistor during the ON period. The leakage current is important because the transistor essentially conducts this amount of current during the OFF period and thus increases total dissipation.

Generally to maximize the efficiency of a power converter under load, the transistor should switch the maximum voltage possible. Because of junction heating there is a maximum collector current which can be switched, and this is independent of the supply voltage. Therefore with a given collector current the power output will increase directly with increased supply voltage.

3.4 OUTPUT FILTER

A fundamental part of the switching regulator is the output filter. The switching regulator filter can take on various forms depending upon the load requirements. However, if a wide range of voltage and current is required, and LC filter is used in combination with a commutating diode.

Having determined the operating frequency, we must proceed to find specific values for L and C. As we have seen in Equation 2.30, this LC product can be achieved with any L/C ratio. There are, however several practical economic and performance considerations that apply to L and C values. It is favorable to push in the direction of small L and large C for the following reasons:

1. An inductor will have considerable weight and volume compared to a capacitor with equal energy storage capacity. Small L and large C, within the limit defined below, will usually result in the lowest cost, weight, and size.
2. The rate of increase of current, di/dt, is determined by L and the voltage across it (e_{in} - e_{out}), as follows:

$$di/dt = \frac{e_{in} - e_{out}}{L}$$

Hence small L and large C results in better transient behavior with step changes in load current. As mentioned in Chapter2, one major objection to a low L/C ratio is that it causes large (sometimes intolerable) overshoot in input current and output voltage on starting. The soft start feature of practical circuits shown in Chapter 4, effectively controls the startup transient, thereby protecting all components and minimizing voltage overshoot. With the soft start circuit, this problem is eliminated and no longer pertains to the selection of L and C values.

To determine the inductance of the choke, the inductance value used is a compromise between the need for a high value to limit peak currents and thus permit good transistor utilization, and the need for a low value to permit fast response to sudden current demands. Refering to Fig. 3.3 and its associated equations, the peak to peak ripple current through the inductor, Δi, is inversely proportional to the inductance. As L is made smaller, Δi determines how small L is permitted to be, as follows:

Before design work can begin, a value must be chosen for i_{max}. This is the level of current in the choke when the switching transistor turns off. The current build up in the choke during the transistor ON time is equal to, Δi, $2 \times (i_{max} - i_{load})$, since $\Delta i/2 = i_{max} - i_{load} = i_{load} - i_{min}$. Trade-offs encountered here are the need of a large amount of inductance if the current rise is to be held to a

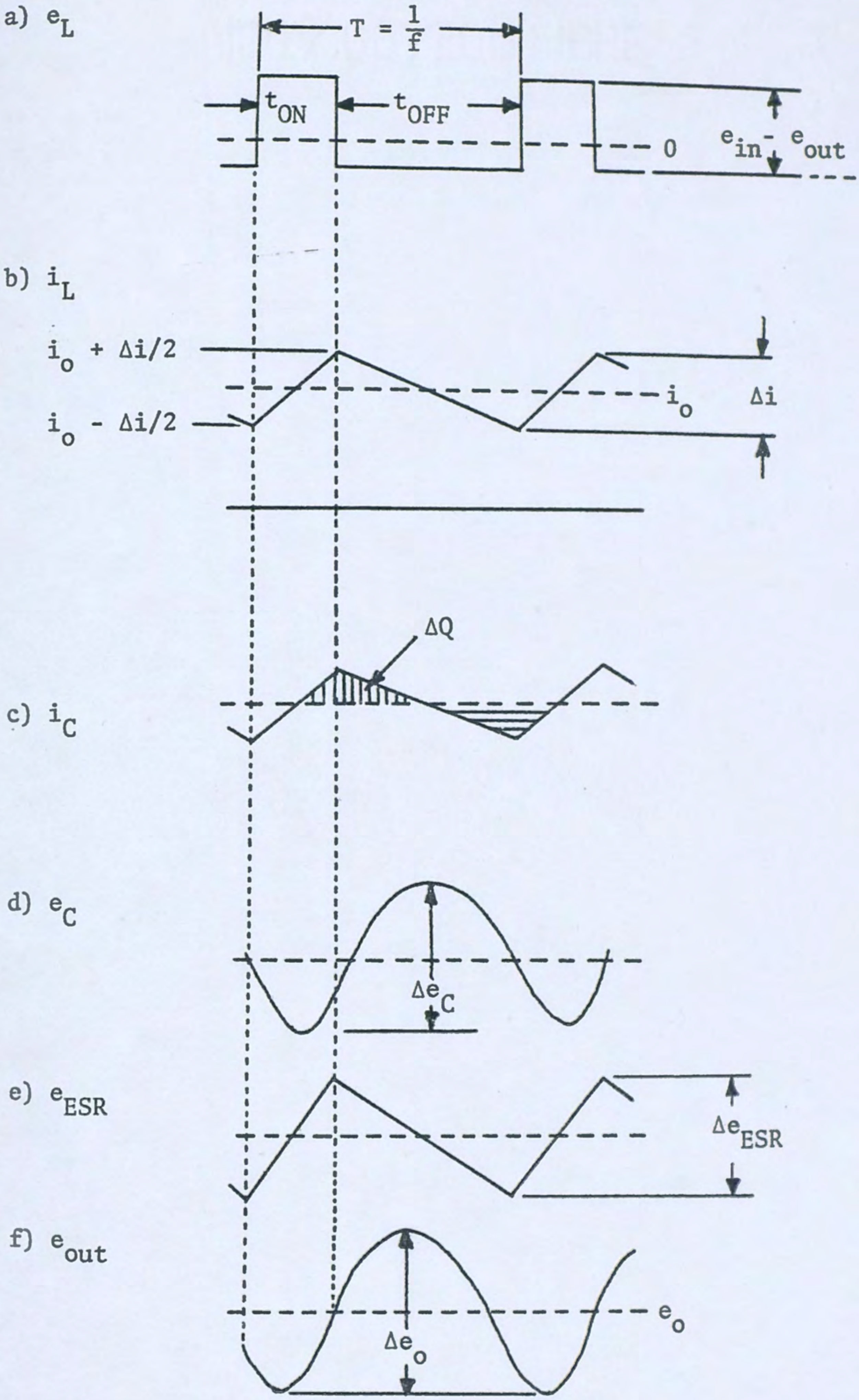


Fig. 3.3 Switching Regulator Wave Forms

minimum. If a small inductor is required, i_{max} can be permitted to rise to as much as twice i_{load}, but the transistor and the diode must now switch that increased current. If $\Delta i/2$ is permitted to become larger than i_{load}, the minimum inductor current becomes a negative value. This is impossible, since neither the switching transistor nor commutating diode will conduct. Therefore the switching regulator goes into a discontinuous mode of operation which is perfectly safe, but the regulation with output current changes becomes poor. For a smoothly operating system, i_{max} is normally chosen to be 10% to 30% higher than i_{load}. The worst case consideration to insure that discontinuous operation does not occur is to make $\Delta i/2$ equal to the minimum load output current.

The current buildup in the choke is calculated as a function of time and the voltage across the choke. The voltage is simply the difference between the input and output.

The classic formula is

$$E = L\frac{di}{dt} \tag{3.1}$$

Rearranging, we get

$$L = \frac{dt}{di} E \tag{3.2}$$

E is now the voltage across the choke, di is the current change, and dt is the on time, which is the duty cycle times the period.

$$E = e_{in} - e_{out}$$

$$di = 2 \times (i_{max} - i_{load}) = \Delta i$$

$$dt = \frac{e_{out}}{e_{in}} \frac{1}{f}$$

Substituting, we get

$$L = \frac{(e_{in} - e_{out})\, e_{out}}{2(i_{max} - i_{load}) f\, e_{in}} = \frac{e_{out}\, t_{off}}{\Delta i} \tag{3.3}$$

The calculated inductance must appear in the physical choke with DC current flowing in the winding. The design process consists of selecting the proper core and determining wire size, number of turns, copper losses, and temperature rise.

Ferrite and Mo-Permalloy powder are excellent core materials for the switching regulator inductor. Since the rms AC current through the inductor is small compared to the DC current, AC losses in the winding and core losses will be negligible compared with DC winding losses.

Selection of the proper core to meet specification is a process concerned with two factors: (1) The core must provide the desired inductance without saturating magnetically at the maximum peak overload current. In this respect each core has a specific (Li^2) saturation energy storage capability. (2) The core must have a window area for the winding which admits the number of turns necessary to obtain the required inductance with a wire size which will result in acceptable DC losses in the winding at the full load output current.

The following equations provide the basis for this design approach. Equation 3.4 defines the value of inductance in terms of basis core parameters and the total number of turns wound on the core:

$$L = N^2 A_1\ 10^{-6}\ \text{mh} \tag{3.4}$$

$$\text{where } A_1 = 0.4\pi\mu \frac{A_e}{\ell_e} \times 10 \text{ mh for 1000 turns} \tag{3.5}$$

μ: effective permeability

ℓ_e: effective magnetic path

A_e: effective magnetic cross section area.

Any specific core has a maximum ampere-turn capability limited by magnetic saturation of the core material.

$$(Li^2)_{sat} = \frac{(B_{sat})^2 \; A_e^2 \; 10^{-4}}{A_1} \tag{3.6}$$

The core selected for an application must have an $(Li^2)_{sat}$ value greater than Li^2_{max} to insure the core will not saturate under maximum peak overload conditions.

The final step is to determine the requirements for the capacitor C and ESR(equivalent series resistance) values which will result in the desired output ripple voltage, Δe_o.

With L, Equation 3.3, and Equation 2.29 we can calculate minimum required C value. To simplify, we use Equation 2.30 initially.

$$C = \frac{e_{out}}{\Delta e_o \omega^2 L} \tag{3.7}$$

Total output ripple voltage, Δe_o, is the sum of the waveforms is Fig. 3.3d and 3.3e. When designing a switching regulator to operate at frequencies greater than 20 kHz in order to achieve small size and low cost in the L and C filter elements, the ESR of the capacitor usually dominates completely. Even when high quality capacitors (low ESR) are employed, it is usually necessary to use a larger capacitance value than would otherwise be required in order to realize the ESR needed to achieve the ripple objective of the design.

From Fig. 3.3e:

$$ESR_{max} = \frac{\Delta e_{ESR}}{\Delta i}$$

$$= \frac{\Delta e_{o\ max}}{\Delta i} \tag{3.8}$$

With high frequency operation, capacitor ESR usually dominates, forcing the use of C value much greater than C_{min} (Equation 3.7) in order not to exceed ESR_{max}.

CHAPTER 4

DESIGN EXAMPLE

In designing a switching regulator power supply, the following parameters will normally be predefined. Specific values shown for each parameter will be used as the basis for a design example.

e_{in}: 208 Vac ±15% 60 Hz single phase input voltage

e_o : 125 Vdc output voltage

Δe_o: 1% peak to peak line/load regulation

$i_{o\ max}$: 15A output current, full load

$i_{o\ min}$: 4A output current, minimum load

The complete switching regulator circuit is shown in Fig. 4.1. The first step in the design is to decide on the operating frequency of switching regulator. Operation above 20 kHz is desirable to eliminate the possibility of audio noise. In this design 25 kHz constant frequency is used. The constant frequency is most used because of the more predictable performance of magnetic components.

4.1 LINE RECTIFIER AND FILTER

This portion of the circuit consists of a bridge rectifier assembly and a filter capacitor. A 35 Ampere bridge is used, even though the average current drawn by the supply is less than 15 Amperes, to improve efficiency under start up and heavy load conditions. Because of the fast response of the switching regulator, input line variations, including line ripple, are filtered by pulse-width modulation. It is, however, still necessary to keep the input voltage level within the voltage regulation range of the control circuit, and this requires the use of a filter capacitor. In actual operation, 176 Volt lines (worst case) will have about 30 Volts peak to peak ripple at full load using filter capacitor of 1900 μF.

Now we have the lowest and the highest input voltage levels to the switcher, which are 200 Volt dc and 335 Volt dc. With these two conditions, we find maximum and minimum duty cycle of the switcher.

To get maximum duty cycle:

$$(e_{input\ min} - e_{loss}) \times t_{on\ max} = T \times e_{out}$$

where $e_{input\ min} = 200$ Volt

$e_{loss} \simeq 3$ Volt

$T = 40\ \mu s$

$$t_{on\ max} = 25.4\ \mu s \tag{4.1}$$

To get minimum duty cycle:

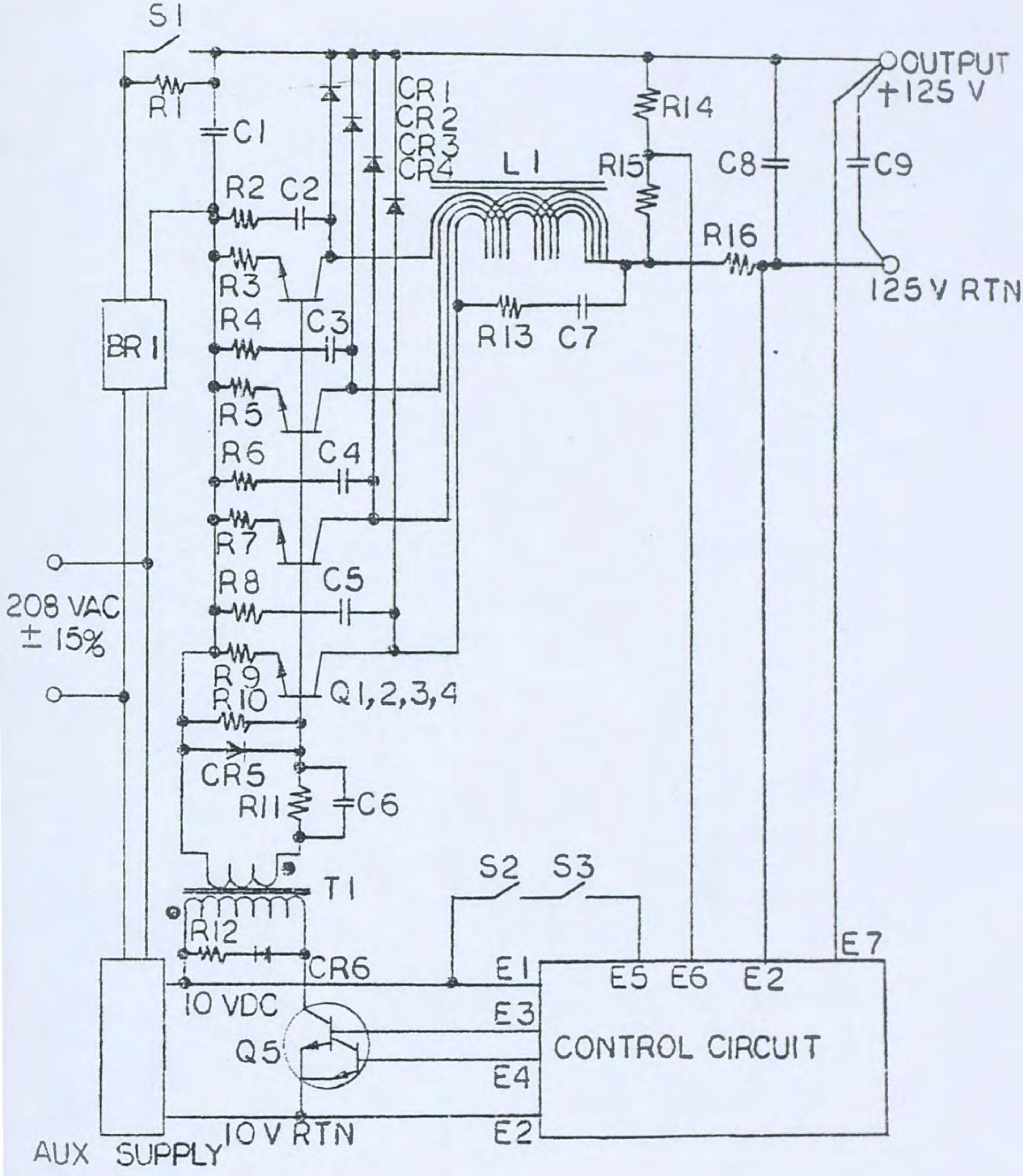
$$(e_{input\ max} - e_{loss}) \times t_{on\ min} = T \times e_{out}$$

where $e_{input\ max} = 335$ Volt

$e_{loss} \simeq 3$ Volt

$T = 40\ \mu s$

$$t_{on\ min} = 15.1\ \mu s \tag{4.2}$$



PART LIST: See Fig. 4.2

Fig. 4.1 Switching Regulator Power Supply

CAPACITORS

C1	1900μF, 450V
C2,3,4,5	0.01μF, 1kV
C6	5μF, 25V
C7	0.001μF, 1kV
C8	2900μF, 200V
C9	1μF, 200V

RESISTORS

R1	75, 25W
R2,4,6,8,13	150, 5W, ±1%
R3,5,7,9	0.2, 5W, ±1%
R10, 12	10, 5W
R11	0.5, 10W, ±1%
R14	100k, ½W, ±1%
R15	442, ½W, ±1%
R16	0.05, 25W, ±1%

DIODES

CR1,2,3,4	MR1376
CR5,6	IN5415
BR1	MDA3508

TRANSISTORS

Q1,2,3,4	2N6583
Q5	2N6353

INDUCTOR, L1

4 core 55110 A2, Magnetics Molypermalloy Powder Core
55 turns
18 AWG 4 wire

TRANSFORMER, T1

3 core 55310 A2, Magnetics Molypermalloy Powder Core
Primary: 60 turns, 23 AWG wire
Secondary: 20 turns, 18 AWG wire

Fig. 4.2 Electrical Part List

4.2 OUTPUT FILTER

The output filter is the heart of the switching regulator. To get good regulation and high performance, we have to choose the proper inductor and capacitor. The power inductor design is a critical part of this regulator, and requires special consideration.

From Equation 3.3

$$L = \frac{(e_{in} - e_{out})\, e_{out}}{2(i_{max} - i_{load}) f\, e_{in}}$$

$$= \frac{e_{out}\, t_{off}}{\Delta i} \tag{4.3}$$

In our example, $i_{o\ min} = 4A$, $i_{max} = 15A$. Caculating $\Delta i = 0.5 \times i_{o\ max}$, $\Delta i = 7.5A$, now that $t_{off\ max} = 24.9\ \mu s$ and $\Delta i = 7.5A$ have been determined, L_{need} is

$$L_{need} = 415\ \mu H \tag{4.4}$$

This 415 μH inductance with DC bias current 22.5A is needed. Using 4 of 55110 A2 magnetics molypermalloy powder core, this minimum required inductance can be achieved with 55 turns.

The next step is to determine the requirements for the capacitor. From Equation 3.7, we can find minimum required value of output capactor.

$$C_{min} = 386\ \mu F \tag{4.5}$$

The most defficult component selection problem for high frequency switching regulator applications is to find and specify an output capacitor with suitable low ESR. Most tantalum and aluminum electrolytic capacitor types do not have ESR specifications (probably because ESR is not very good). Anyhow to get a low value of ESR, we choose a 2900 μF capactior.

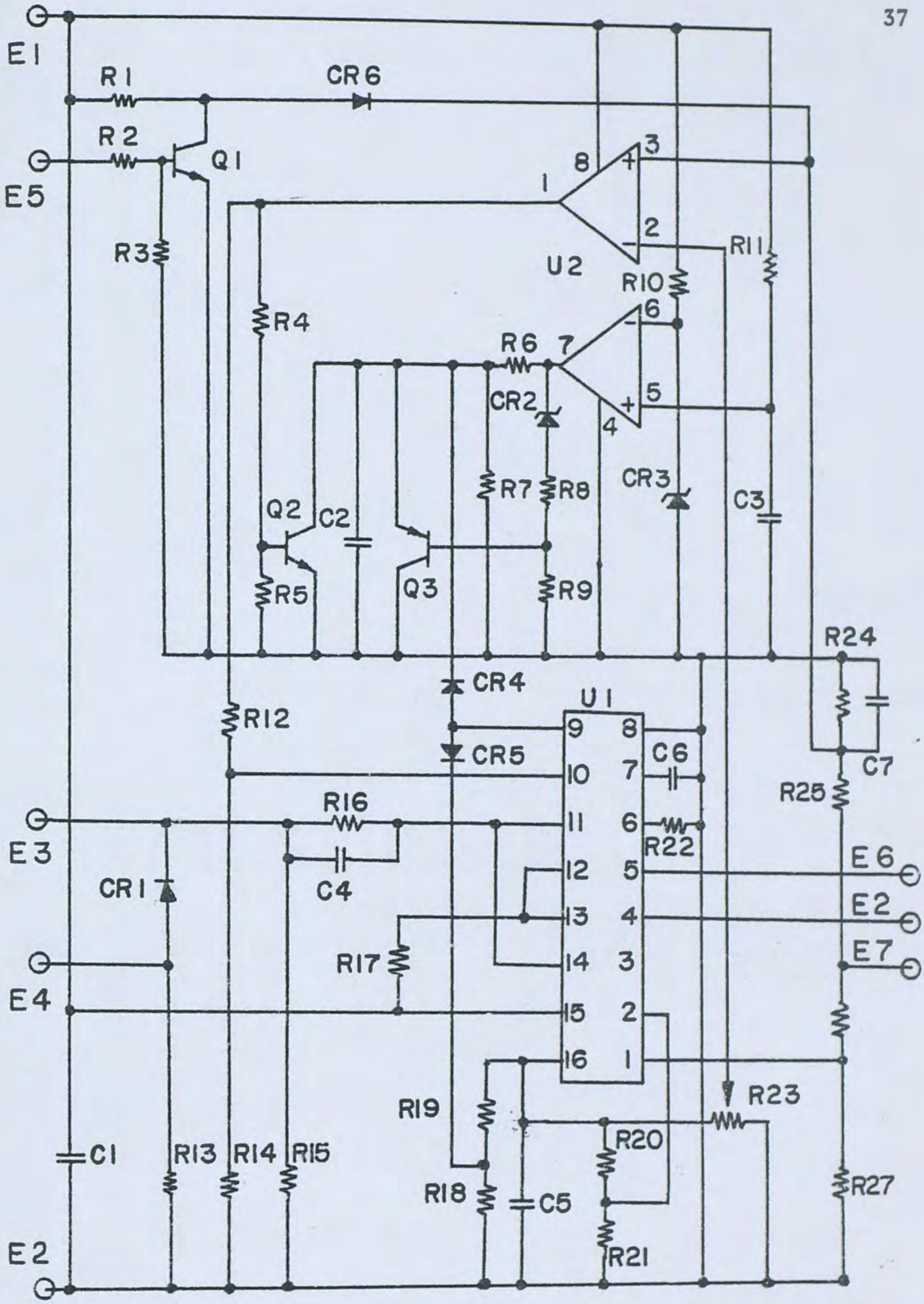
4.3 POWER SWITCHING COMPONENTS

Fast switching transistors and diodes are required to maintain good efficiency in high frequency switching regulators. Voltage ratings of the power switching transistor and catch diode must be greater than the maximum input voltage, 350 Vdc, including any transient voltages that may appear at the input of the switching regulator. Transistor 2N6583 has the following characteristics:

V_{cBo}	$500V_{dc}$	Minimum
$V_{ceo(sus)}$	$400V_{dc}$	Minimum (I_c = 200mA)
$V_{ce(sat)}$	$1.5V_{dc}$	Maximum (I_c = 7A, I_B = 1.4A)
t_r	150_{nsec}	Maximum (V_{ce} = 300V, I_c = 7A, I_{B1} = 1A, I_{B2} = 1.4A)
t_s	$2.0_{\mu sec}$	Maximum (V_{ce} = 300V, I_c = 7A, I_{B1} = 1A, I_{B2} = 1.4A)
t_f	150_{nsec}	Maximum (V_{ce} = 300V, I_c = 7A, I_{B1} = 1A, I_{B2} = 1.4A)
T_j	200^oC	Maximum

Also we pick up the MR1376 fast recovery rectifier.

Our application requires 22.5 ampere switching current. There is no such high speed transistor. We must parallel the switching regulator output stage with 4 transistors to achieve higher output current with high speed. Operating the switching regulator output stage parallel requires some thought, particulary the sharing of current between the transistors. The current sharing is provided by equal resistors R3, R4, R5, and R6 when the bases of the switchers are connected together as shown in Fig. 4.1



PART LIST: See Fig. 4.4

Fig 4.3 Control Circuit

CAPACITORS

C1,4,5	0.33μF, 25V
C2	500μF, 10V
C3	33μF, 10V
C6	0.02μF, 15V
C7	0.1μF, 15V

DIODES

CR1,4,5,6	1N4148
CR2	1N751A
CR3	1N823A

I.C.

U1	SG3524
U2	LM158

RESISTORS

R1, 2,3,12,14	10k, ¼W
R4,5,6,7,8,9,10	4.7k, ¼W
R11	100k, ¼W
R13,15	1k, ¼W
R16,17	470, ¼W
R18,22	3.3k, ¼W
R19,24	2.2k, ¼W
R20,21	4.87k, ¼W, 1%
R23,27	5k
R25,26	100k, ½W, 1%

TRANSISTOR

Q1,2	2N2222
Q3	2N2907

SG3524 BLOCK DIAGRAM

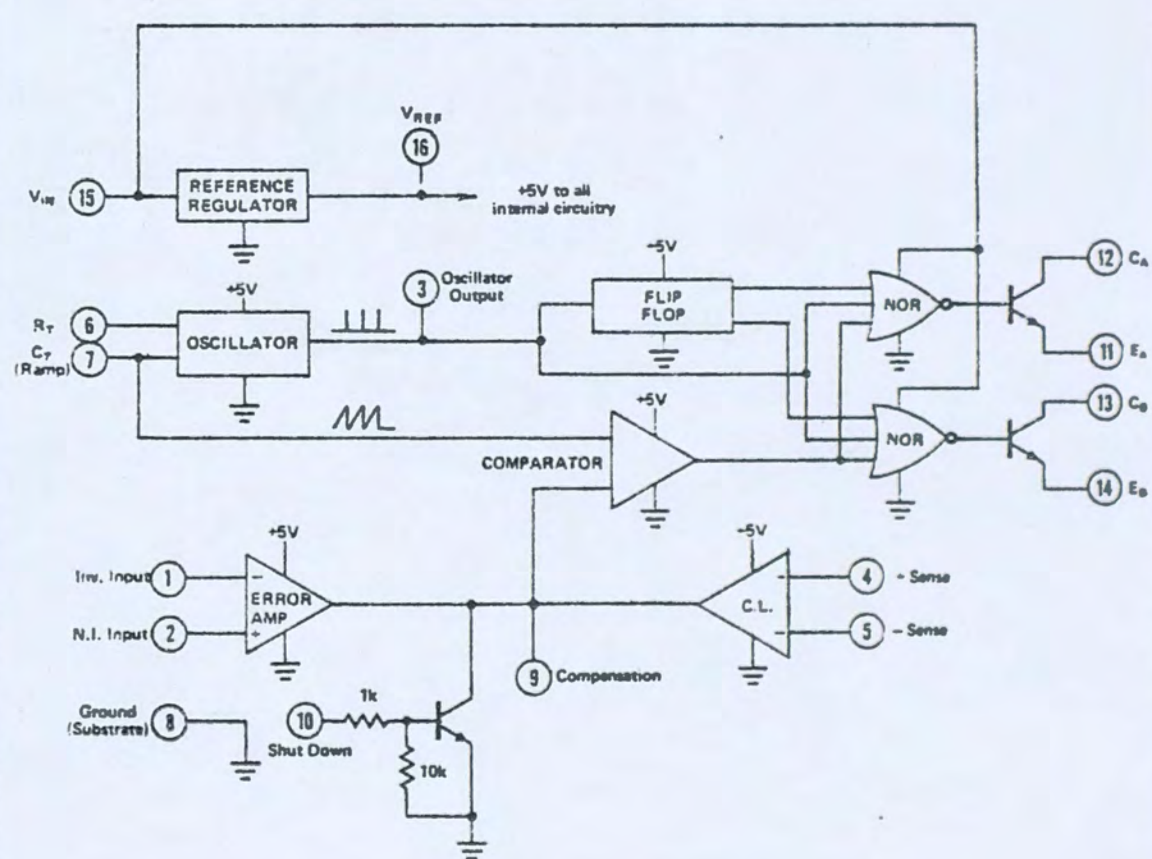


Fig. 4.4 Electrical Part List

4.4 CONTROL CIRCUIT

The control circuit is shown in Fig. 4.3. The control portion of this design uses a monolithic integrated circuit which contains voltage reference, error amplifier, comparator, oscillator, and current limiting and shut down circuitry.

The oscillator generates the fixed frequency of operation. The oscillation period is approximately t = R22C6 where t is in micro-seconds when R22 is in Ohms and C6 is in micro-Farads. The resulting output forms a clock pulse for the comparator. The second input to the comparator is derived from the error amplifier. This error amplifier senses the output voltage and controls the duty cycle of the one shot comparator. And the resulting pulse width controlled output is then fed to the base drive transformer in pulse width controlled switching mode at 25 kHz. The output of the drive transformer is then fed to the main switch to complete the loop.

A key factor to successful operation of high voltage switching lies in the drive waveforms supplied to the power switches. Drive signals must have fast rise and fall times.

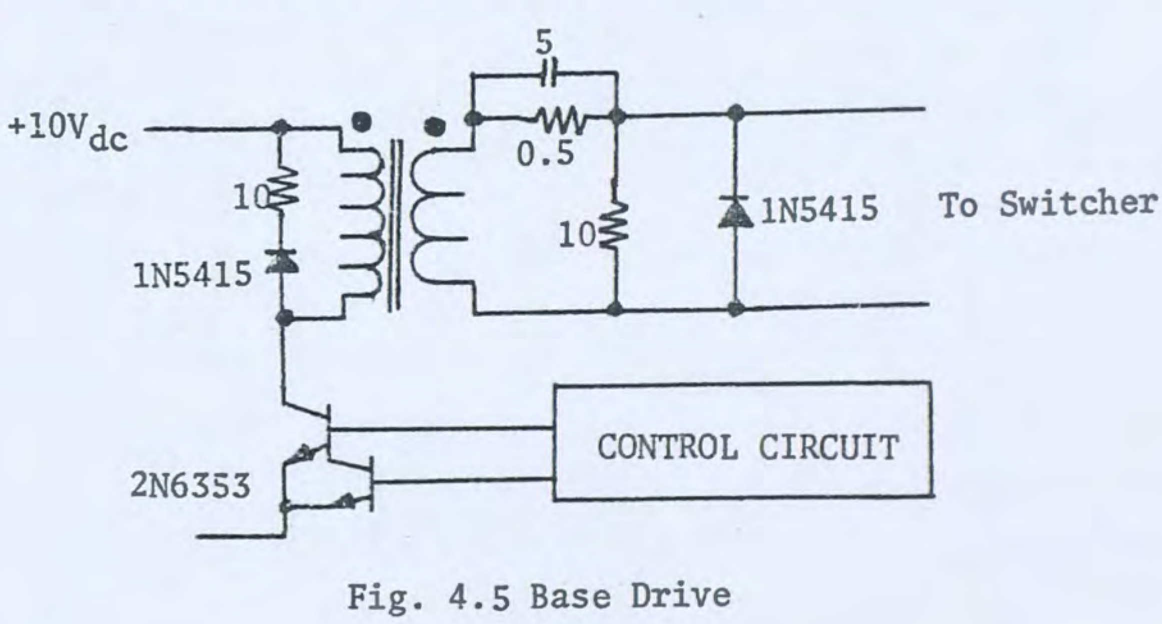


Fig. 4.5 Base Drive

The base drive transformer, shown in Fig. 4.5, 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 resonable saturation voltage at peak collector currents. In this design a 1 : 3 ratio was used which will provide 2.8 ampere of base drive at peak switcher current.

Fig. 4.6 shows the relationship of the voltages and current in the base and collector of a main power switcher. It is common practice to use a forced gain ratio(β) which is a function of the collector current and saturation characteristics of the switching transistor. The magnitude of the reverse drive, however, is determined by the turn-off characteristics and the energy capability of the device.

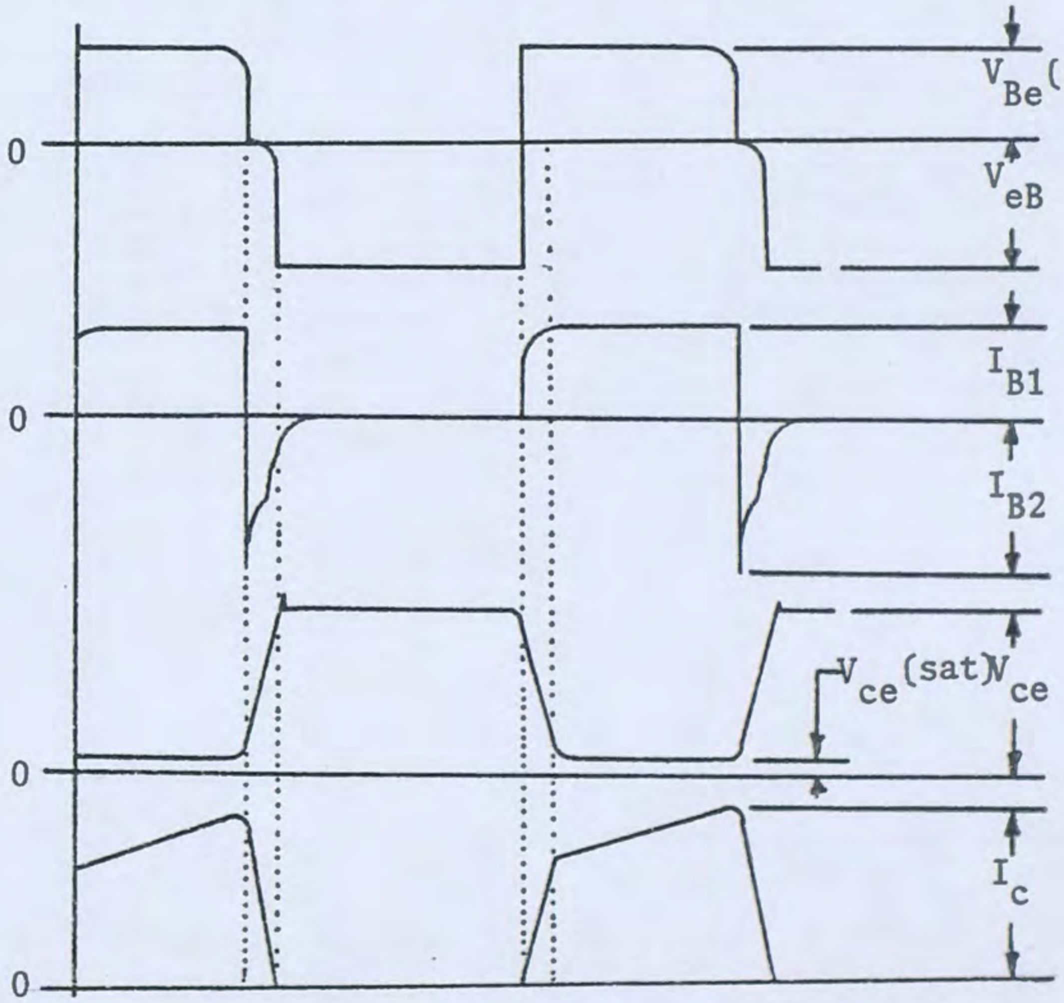


Fig. 4.6 Switching Transistor Wave Form

Thus, as I_{B2} is increased, storage and fall times are decreased, but V_{eB} is increased. It follows, then, that for practical purposes, the reverse drive circuit must provide high I_{B2} at low V_{eB} and there for must exhibit low source impedance. The circuit shown in Fig. 4.3 fulfills the requirements for low source impedance for I_{B2}.

4.5 PERFORMANCE FEATURES

Soft Start

As mentioned in Chapter 2, we may see intolerable overshoot in input current and output voltage when the circuit is first energized. Further, for reliable operation of any switching regulator, it is essential to stay within the safety ratings of the power switches. Under normal steady-state conditions, this is easy to achieve, but most failures occur during startup or shut down.

In the control circuit, as long as U1 pin 9 is 0 Volt, there is no output. As capacitor C2 charges through $R6 \| R7$, with R11C3 delay time, the output pulse slowly increases from zero to the point where the feedback loop takes control. By R18 and R19, maximum pulse width on-time is fixed.

Fault Protection

Whenever the output is too high, comparator U2 pin 1 is high. This high output of comparator U2 pin 1 shutdowns pulse width modulator output and turns on transistor Q2. Thus capacitor C2 is fully discharged. And then, the soft start circuitry provides automatic reset with soft startup. If the output goes too high again it will shut down and reset by itself. The output waveform is shown in Fig. 4.7.

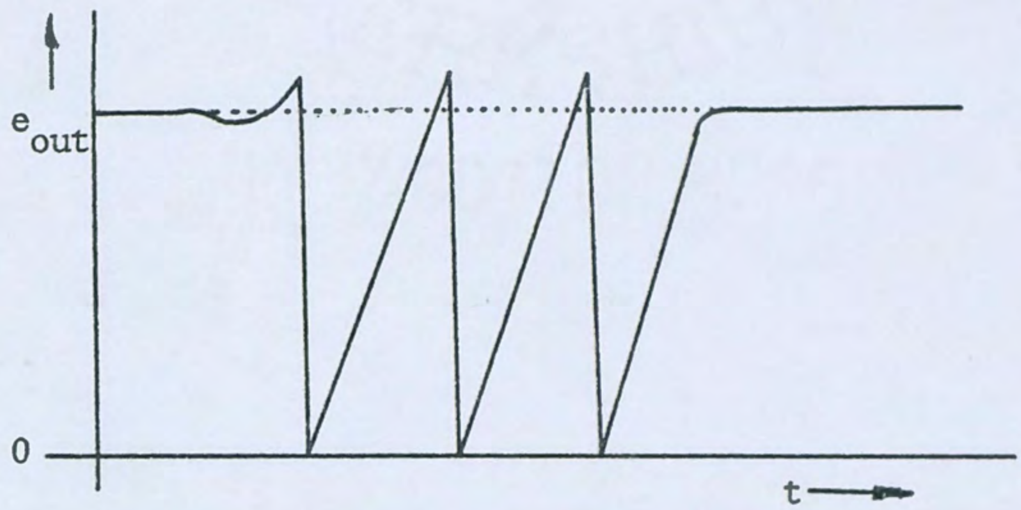


Fig. 4.7 False State Output Wave Form

Foldback current limiting is provided. Maximimum output current is determined by

$$i_{max} = \frac{1}{R16} V_{TH} + \frac{e_o R15}{R15 + R14} \qquad (4.6)$$

where $V_{TH} = 200$ mV

$i_{max} = 15A$

Finally, an interlock circuit is provided. If the regulator is not connected to the load, the regulator will not be able to turn on. This circuitry can be used for many purposes.

Efficiency

The efficiency of a switching regulator depends upon the factors given in the following equation

$$\text{Efficiency} = \frac{P_{out}}{P_{out} + P_{loss}} 100\% \qquad (4.7)$$

where

$P_{out} = 1875$ Watt

$P_{loss} = P_{input\ filter} + P_{switcher} + P_{diode} + P_{output\ filter} + P_{aux}$

$P_{input\ filter} = 50$ Watt for low line input

$P_{switcher} = 50$ Watt for transistor, snubber network and current sharing resistor

$P_{diode} = 20$ Watt

$P_{output\ filter} = 40$ Watt for full load

$P_{aux} = 40$ Watt for aux supply

Efficiency $\simeq 90\%$ (4.8)

CHAPTER 5

RESULTS AND RECOMMENDATIONS FOR FURTHER STUDY

The purpose of this investigation was to analyze the switching power supply. As a result the main part of this investigation was restricted to the analysis of the single ended switching regulator.

Several other type switching mode circuits are available to the power supply designer. The most popular circuits in use today are the single ended switching regulator, the flyback converter, and the push-pull converter.

As we have seen, switching power supply circuits not only increase efficiency but they also reduce power supply size and weight substantially. Compared to conventional series pass regulator supplies, volume and weight reductions of 10 : 1 and efficiencies of 90% are not uncommon.

However, switching power supplies have several disadvantages. The most significant disadvantage of the switching technique is the noise and RFI produced. Input filters and the layout of components can reduce this to acceptable levels.

Noise radiation from an enclosed power supply is rarely a problem. Noise difficulties arise from signals conducted through the AC power lines or radiated by those lines. Shielding the power line or bypassing the conducted noise with small ceramic capacitors at the source of the AC power will minimize this interference.

Another disadvantage is the slower response of the switching supply to changing load demands. For a step change load current from 50% of rating to full load, the 20 kHz switching supply has about a millisecond recovery time. This compares 20 to 50 μsec response times for conventional series-pass types. Total excursion time is just as important as the time it takes to recover, and when powering fast digital logic circuits even 20 to 50 μsec can be a lifetime.

Because of its greater complexity, there are a number of problems which need to be understood better in order that the switching type power supply may find more wide-spread application. Because switching supplies have much more sophisticated circuitry, there is more chance for misapplications on the part of the user. This should decrease as the user becomes more knowledgeable about switching supplies.

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