The Designer's Guide for Switching Power Supply Circuits and Components by R J Haver

Introduction

The Switching Power Supply continues to increase in popularity and is one of the fastest growing markets in the world of power conversion. Its performance and size advantages meet the needs of today's modern and compact electronic equipments and the increasing variety of components directed at these applications makes new designs even more practical.

This guide is intended to provide the designer with an overview of the more popular inverter circuits, their basic theory of operation, and some of the subtle characteristics involved in selecting a circuit and the appropriate components. Also included are valuable design tips on both the major passive and active components needed for a successful design. Finally, a complete set of selector guides to Motorola's Switchmode components is provided which gives a detailed listing of the industry's most comprehensive line of semiconductor products for switching power supplies.

Comparison with Linear Regulators

The primary advantages of a switching power supply are efficiency, size, and weight. It is also a more complex design, cannot meet some of the performance capabilities of linear supplies and generates a considerable amount of electrical noise. Switchers are being accepted in the industry, particularly where size and efficiency are of prime importance, because its performance is still adequate for most applications and is often cost competitive in the 50 W power level and above. Because the switcher's passive components such as transformers and filters are smaller, they are almost always lower in cost than the high power (100 W) linear regulators. However, active component count is high (70 to 140 devices) and remains high regardless of the output power rating. This makes it less cost effective at the lower power levels. Switchers have been significantly cost reduced in recent years because designers have been able to simplify the control circuits with new, cost effective integrated circuits and have found even lower cost alternatives in the passive component area.

A performance comparison chart of switching versus linear supplies is shown in Figure 1. Switcher efficiencies run from 70 to 80% but occasionally fall to (60–65%) when linear post regulators are used for the auxillary outputs. Some linear power supplies on the other hand, are operated with up to 50% efficiency but these are areas where line variations or hold-up time problems are minimal. Most linears operate with typical efficiencies of only 30%. The overall size reduction of a 20 kHz switcher is about 4:1 and newer designs in the 100 to 200 kHz region end up at about 8:1 (versus a linear). Other characteristics such as static regulation specs are comparable, while ripple and load transient response are usually worse. Output noise specs can be somewhat misleading. Very often a 500 mV switching spike at the output may be attenuated considerably at the load itself due to the series inductance of the connecting cables and the additional filter capacitors found in many logic circuits. In the future, the noise generated at higher switching frequencies (100–500 kHz) will probably be easier to filter and the transient response will be faster. Hold-up time is the switchers inherent ability to regulate over wide variations in input voltage. It is easier to store energy in high voltage capacitors (200–400 V) than in the lower voltage (20–50 V) filter capacitors common to linear's power supplies. This is due to the fact that the physical size of a capacitor is dependent on its CV product while energy storage is proportional to CV^2.

Popular Inverter Configurations

A switching power supply is a relatively complex circuit as is shown by the four basic building blocks of Figure 2. It is apparent here that the heart of the supply is really the high frequency inverter. It is here that the work of chopping the rectified line at a high frequency (20 kHz) is done. It is here also that the line voltage is transformed down to the correct output level for use by logic or other electronic circuits. The remaining blocks support this basic function. The 60 Hz input line is rectified and filtered by one block and after the inverter steps this voltage down, the output is again rectified and filtered by another. The task of regulating the output voltage is left to the control circuit which closes the loop from the output to the inverter. Most control circuits generate a fixed frequency internally and utilize pulse width modulation techniques to implement the desired regulation. Basi-

FIGURE 1 — 20 kHz Switcher versus Linear Performance

Parameter	Switcher	Linear
Efficiency	75%	30%
Size	2.0 W/IN^3	0.5 W/IN^3
Weight	40 W/lb.	10 W/lb.
Cost 200–500 W*	$1.00/W	$1.25/W
Cost 50–150 W*	$1.50/W	$1.50/W
Line and Load Regulation	0.1%	0.1%
Output Ripple V_{P-P}	50 mV	5.0 mV
Noise V_{P-P}	50–200 mV	—
Transient Response	1 ms	20 µs
Hold-Up Time	20–30 ms	1–2 ms

*Based on 1980 Cost Figures

cally, the on-time of the square wave drive to the inverter is controlled by the output voltage. As load is removed or input voltage increases, the slight rise in output voltage will signal the control circuit to deliver shorter pulses to the inverter and conversely as the load is increased or input voltage decreases, wider pulses will be fed to the inverter.

The inverter configurations used in today's switchers actually evolved from the buck and boost circuits shown in Figures 3A and 3B. In each case the regulating means and loop analysis will remain the same but a transformer is added in order to provide electrical isolation between the line and load. The forward converter family which includes the push-pull and half bridge circuits evolved

FIGURE 2 — Functional Block Diagram — Switching Power Supply

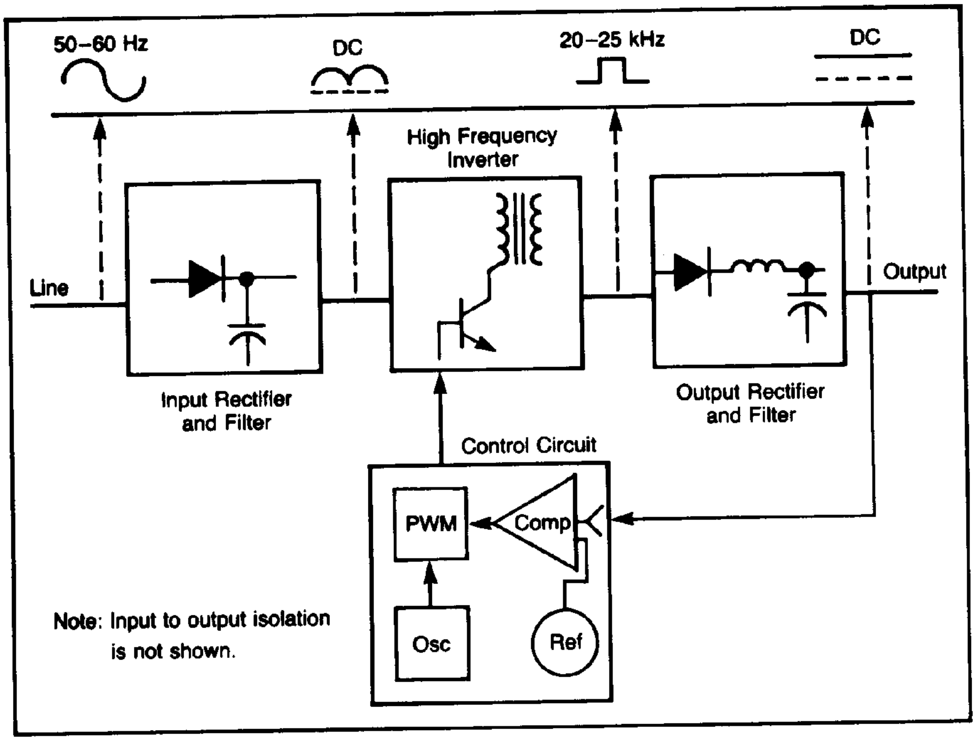
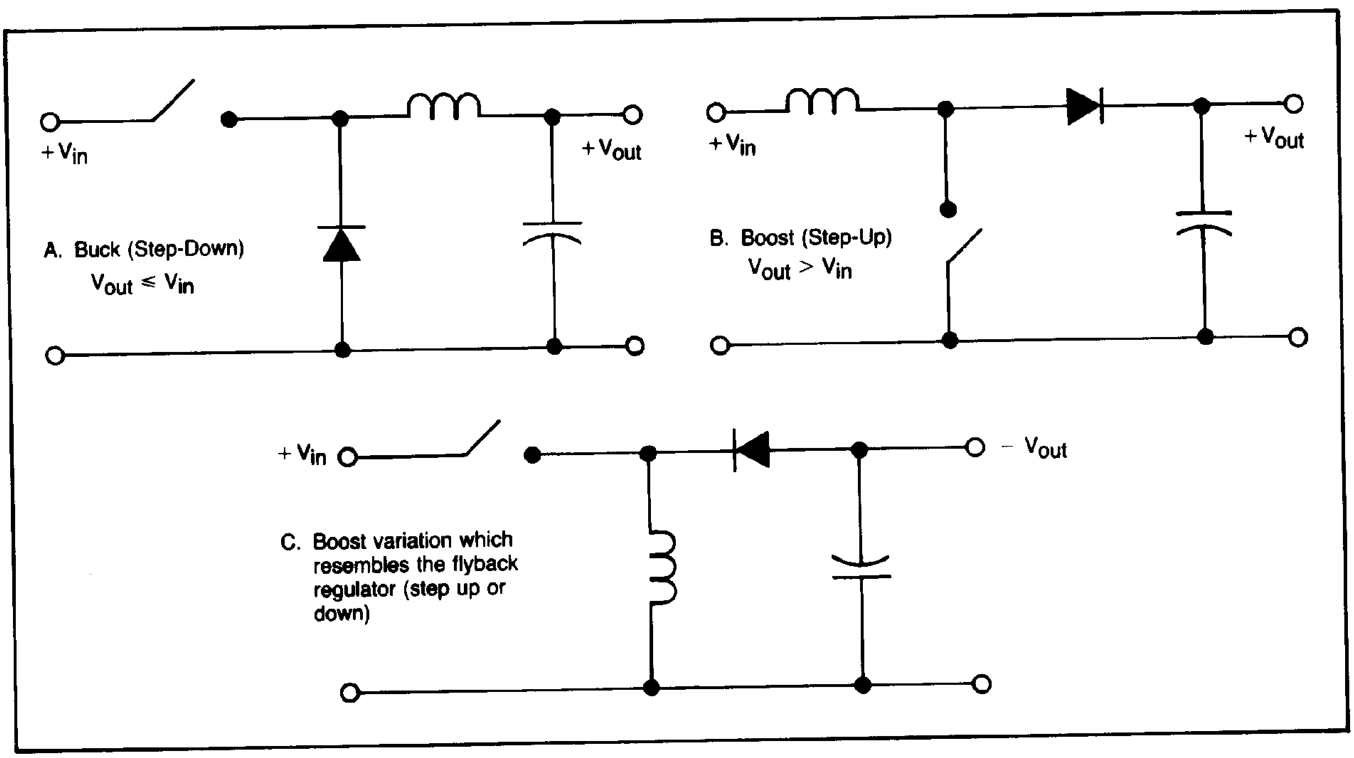


FIGURE 3 — Nonisolated DC-DC Converters



from the buck regulator (Figure 3A). And the newest switcher, the flyback converter, actually evolved from the boost regulator. The buck circuit interrupts the line and provides a variable pulse width square wave to a simple averaging LC filter. In this case, the first order approximation of the output voltage is $V_{out} = V_{in} \times$ duty cycle and regulation is accomplished by simply varying the duty cycle. This is satisfactory for most analysis work and only the transformer turns ration will have to be adjusted slightly to compensate for IR drops, diode drops, and transistor saturation voltages.

Operation of the boost circuit is more subtle in that it first stores energy in a choke and then additively delivers this energy with the input line to the load. However, the flyback regulators which evolved from this configuration delivers only the energy stored in the choke to the load. This method of operation is actually based on the boost variation model shown in Figure 3C. Here, when the switch is opened, only the stored inductive energy is delivered to the load. The true boost circuit can also regulate by stepping up (or boosting) the input voltage whereas the variation or flyback regulator can step the input voltage up or down. Analysis of the boost regulator begins by dealing with the choke as an energy storage element which delivers a fixed amount of power to the load:

$$P_o = 1/2\,L\,I^2 f_o$$

where I = the peak choke current

f_o = the operating frequency

and L = the inductance

Because it delivers a fixed amount of power to the load regardless of load impedance (except for short circuits), the boost regulator is the designer's first choice in photo-flash and capacitive-discharge (CD) automotive ignition circuits to recharge the capacitive load. It also makes a good battery charger. For an electronic circuit load, however, the load resistance must be known in order to determine the output voltage:

$$V_o = \sqrt{P_o R_L} = I\sqrt{\frac{L f_o R_L}{2}}$$

where R_L = The load resistance

In this case, the choke current is proportional to the on-time or duty cycle of the switch and regulation for fixed loads simply involves varying the duty cycle as before. However, the output also depends on the load which was not the case with buck regulators and results in a variation of loop gain with load.

For both regulators, transient response or responses to step changes in load are very difficult to analyze. They lead to what is termed a "load dump" problem. This requires that energy already stored in the choke or filter be provided with a place to go when load is abruptly removed. Practical solutions to this problem include limiting the minimum load and using the right amount of filter capacitance to give the regulator time to respond to this change.

Flyback and Forward Converters

To take advantage of the regulating techniques just discussed and also provide isolation, a total of seven popular configurations have evolved and are illustrated in Figures 4 and 7. Each circuit has a practical power range or capability associated with it as follows:

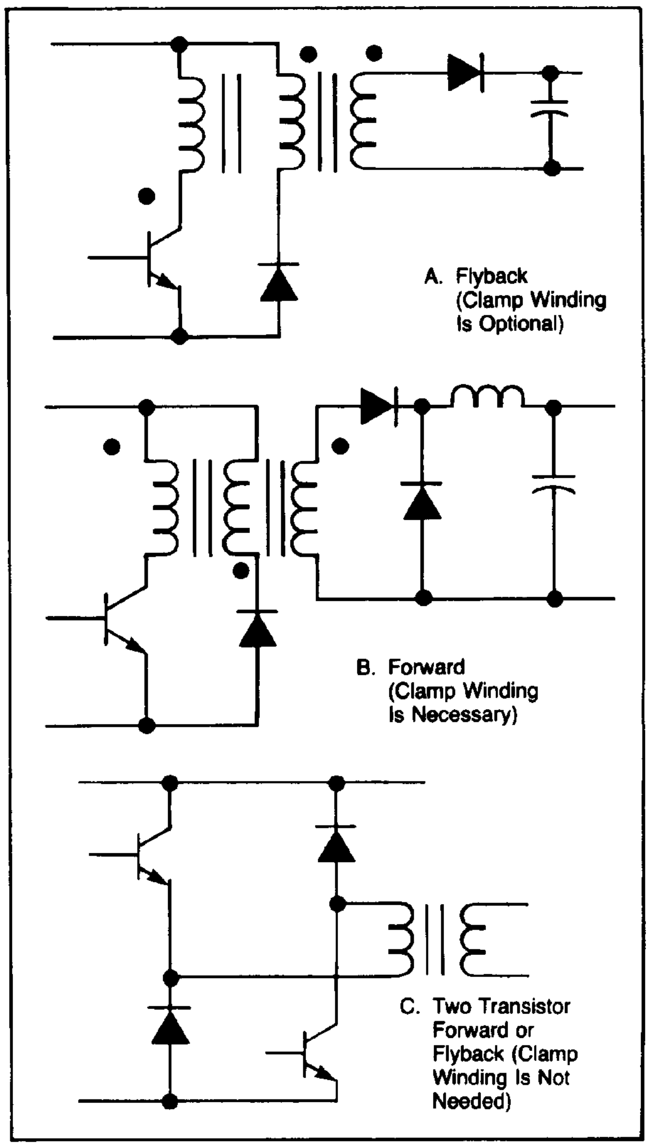
Circuit	Power Range	Motorola Reference
Flyback	50 to 100 watts	EB-87
Forward	100 to 200 watts	
Push-Pull	200 to 500 watts	EB-88, AN-737A
Half Bridge	200 to 500 watts	EB's 86 & 100, AN-767
Full Bridge	500 to 2000 watts	EB-85

First to be discussed will be the low power (20–200 W) converters which are dominated by the single transistor circuits shown in Figure 4. All of these circuits operate the magnetic element in the unipolar rather than bipolar mode. This means that transformer size is sacrificed for circuit simplicity.

Flyback

The flyback (alternately known as the "ringing choke") regulator stores energy in the primary winding and dumps it into the secondary windings (see Figure 4A). A clamp winding is usually present to allow energy stored

FIGURE 4 — Low Power Popular (20–200 W) Converter Configurations



in the leakage reactance to return safely to the line instead of avalanching the switching transistor. The operating model for this circuit is the boost circuit variation discussed earlier. The flyback is the lowest cost regulator (except at high power levels) because output filter chokes are not required since the output capacitors feed from an energy source rather than a voltage source. It does have higher output ripple than the forward converters because of this. However, it is an excellent choice when multiple output voltages are required and does tend to provide better cross regulation than the other types. In other words changing the load on one winding will have little effect on the output voltage of the others.

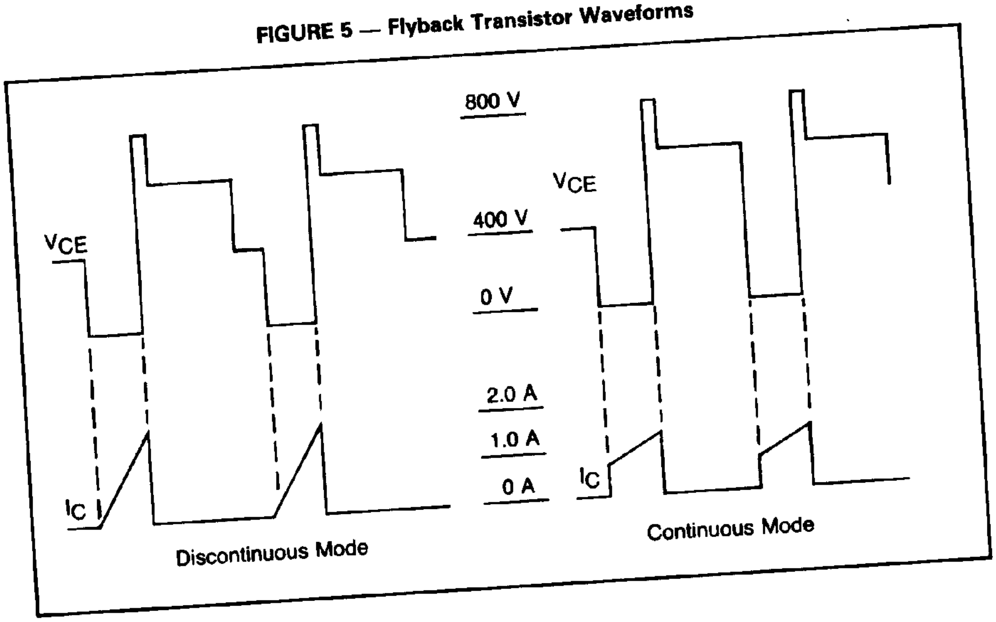
A 120/220 Vac flyback design requires transistors that block twice the peak line plus transients or about 1.0 kV. Presently variations of the 1200 to 1500 V horizontal deflection transistors are used here. These bipolar devices are relatively slow ($t_f = 1.0\ \mu s$) and tend to limit efficient operating frequencies to 20 to 30 kHz. The availability of 1000 V TMOS FETs will permit operation at much higher frequencies. Faster 1.0 kV bipolar transistors are also planned in the future and will provide another design alternative. The two transistor variations of this circuit (Figure 4C) eliminates the clamp winding and adds a transistor and diode to effectively clamp peak transistor voltages to the line. With this circuit a designer can safely use the faster 400 V to 500 V bipolar or FET Switchmode transistors and push operating frequencies considerably higher. There is a cost penalty here over the single transistor circuit due to the extra transistor, diodes and base drive circuitry.

operating in the continuous mode. This is generally an advantage for the transistor in that it needs to switch only half as much peak current in order to deliver the same power to the load. In many instances, the same transformer may be used with only the gap reduced to provide more inductance. Sometimes the core size will need to be increased to support the higher LI product (2 to 4 times) now required because the inductance must increase by almost 10 times to effectively reduce the peak current by two. In dealing with the continuous mode, it should also be noted that the transistor must now turn on from 500 to 600 V rather than 400 V level because there no longer is any dead time to allow the flyback voltage to settle back down in the input voltage level. Generally it is advisable to have V_{CEO} (SUS) ratings comparable to the turn-on requirements.

The flyback converter stands out from the others in its need for a low inductance, high current primary. Conventional E and pot core ferrites are difficult to work with because their permeability is too high even with relatively large gaps (50 to 100 milli-inches). The industry needs something better (like powered iron) that will provide permeabilities of 60 to 120 instead of 2000 to 3000 for this application.

The single transistor forward converter is shown in Figure 4B. Although it initially appears very similar to the flyback, it is not. The operating model for this circuit is actually the buck regulator discussed earlier. Instead of storing energy in the transformer and then delivering it to the load, this circuit uses the transformer in the active or forward mode and delivers power to the load while

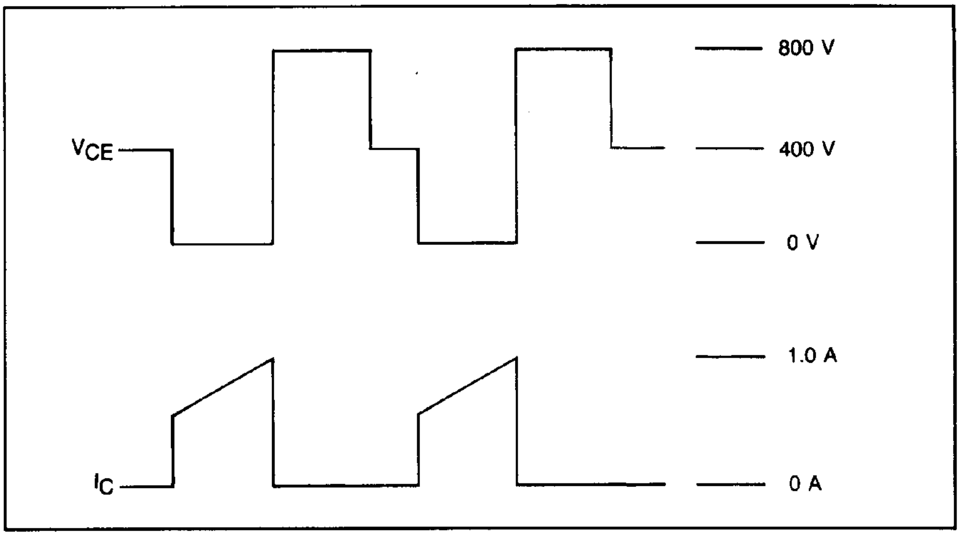
FIGURE 5 — Flyback Transistor Waveforms



A subtle variation in the method of operation can be applied to either of these circuits. The difference is referred to as operation in the discontinuous or continuous mode and the waveform diagrams are shown in Figure 5. The analysis given in the earlier section on boost regulators dealt strictly with the discontinuous mode where all the energy is dumped from the choke before the transistor turns on again. If the transistor is turned on while energy is still being dumped into the load, the circuit is

the transistor is on. The additional output rectifier is used as a freewheeling diode from the LC filter and the third winding is actually a reset winding. It generally has the same turns as the primary, (is usually bifilar wound) and does clamp the reset voltage to twice the line. However its main function is to return energy stored in the magnetizing inductance to the line and thereby reset the core after each cycle of operation. Because it takes the same time to set and reset the core, the duty cycle of this circu

FIGURE 6 — Forward Converter Transistor Waveforms

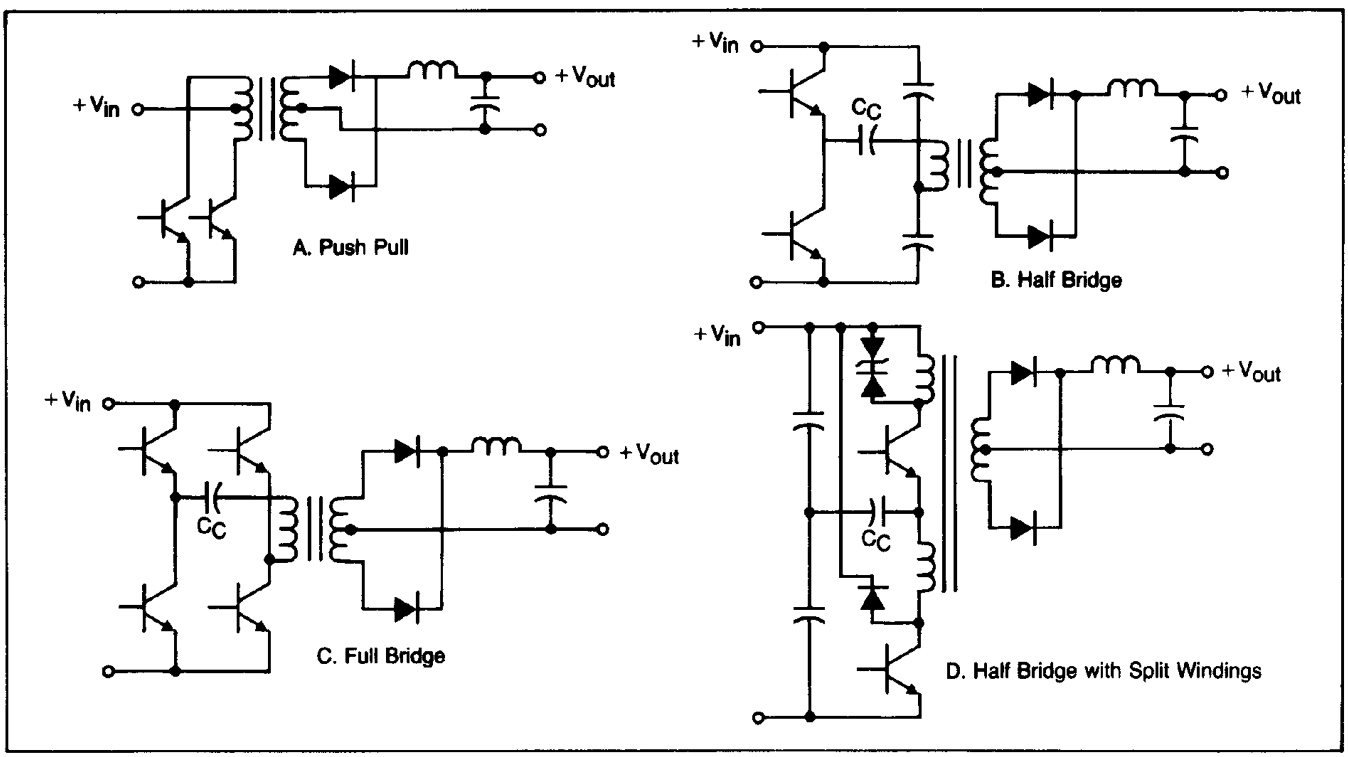


cannot exceed 50%. This also is a very popular low power converter and like the flyback is practically immune from transformer saturation problems. Transistor waveforms shown in Figure 6 illustrate that the voltage requirements are identical to the flyback. For the single transistor versions, 400 V turn-on and 1.0 kV blocking devices like the 1200 to 1500 V transistors are required. The two transistor circuit variations shown in Figure 4C again adds a cost penalty but allows a designer to use the faster 400 to 500 V devices. With this circuit, operation in the discontinuous mode refers to the time when the load is reduced to a point where the filter choke runs "dry". This means that choke current starts at and returns to zero during each cycle of operation. Most designers prefer to avoid this type of mode because of higher ripple and noise even though there are no adverse effects on the components themselves. Standard ferrite cores work fine here and in the high power converters as well. In these applications, no gap is used as the high permeability (3000) results in the desirable effect of very low magnetizing current levels.

Push-Pull and Bridge Converters

The high power circuits shown in Figure 7 all operate the magnetic element in the bipolar or push-pull mode and require 2 to 4 inverter transistors. Because the trans-

FIGURE 7 — High Power Popular Converter Configurations (100 W–1.0 kW)



formers operate in this mode they tend to be almost half the size of the equivalent single transistor converters and thereby provide a cost advantage over their counterparts at power levels of 100 to 1.0 kW.

Push-Pull

The push-pull converter shown in Figure 7A is one of the oldest converter circuits around. It's early use was in low voltage inverters such as the 12 Vdc to 120 Vdc power source for recreational vehicles and in dc to dc converters. Because these converters are free running rather than driven and operate from low voltages, transformer saturation problems are minimal. In the high voltage off-line switchers, saturation problems are common and difficult to solve. The transistors are also subjected to twice the peak line voltage which requires the use of relatively slow 1.0 kV transistors. Both of these draw backs have tended to discourage designers of off-line switchers from using this configuration.

Half and Full Bridge

The most popular high power converter is the half bridge (Figure 7B). It has two clear advantages over the push-pull and became the favorite rather quickly. First, the transistors never see more than the peak line voltage and the standard 400 V fast Switchmode transistors that are readily available may be used. And second, and probably even more important, transformer saturation problems are easily minimized by use of a small coupling capacitor (about 2 – 5 μF) as shown. Because the primary winding is driven in both directions, a full wave output filter, rather than half, is now used and the core is actually utilized more effectively. Another more subtle advantage of this circuit is that the input filter capacitors are placed in series across the rectified 220 V line which allows them to be used as the voltage doubler elements on a 120 V line. This still allows the inverter transformer to operate from a nominal 320 V bus when the circuit is connected to either 120 V or 220 V. Finally, this topology allows diode clamps across each transistor to contain destructive switching transients. The designer's dream, of course, is for fast transistors that can handle a clamped inductive load line at rated current. And a few (like the Switchmode III series from Motorola) are beginning to appear on the market. However, the older designs in this area still end up using snubbers to protect the transistor which sacrifices both cost and efficiency.

The effective current limit of today's low cost TO-3 discrete transistors (250 mil die) is somewhere in the 10 to 20 A area. Once this limit is reached, the designer generally changes to the full bridge configurations shown in Figure 7C. Because full line rather than half is applied to the primary winding, the power out can be almost double that of the half bridge with the same switching transistors. Power Darlington transistors are a logical choice to higher power control with current, voltage and speed capabilities allowing very cost and performance effect designs. Another variation of the half bridge is the split winding circuit shown in Figure 7D. A diode clamp can protect the lower transistor but a snubber or zener clamp must still be used to protect the top transistor from switching transients. Because both emitters are at an ac ground point, expensive drive transformers can now be replaced by lower cost capacitively coupled drive circuits.

Component Design Tips

Transformers

With respect to transformer design, many of today's designers would say don't try it. They'd advise using a consultant or winding house to perform this task and with good reason. It takes quite a bit of time to develop a feel for this craft and be able to use both experience and intuition to find solutions to second and third order problems. Because of these subtle problems, most designers find that after the first paper design is done, as many as four or five lab iterations may be necessary before the transformer meets the design goals. However, there is a considerable design challenge in this area and a great deal of satisfaction can be obtained by mastering it.

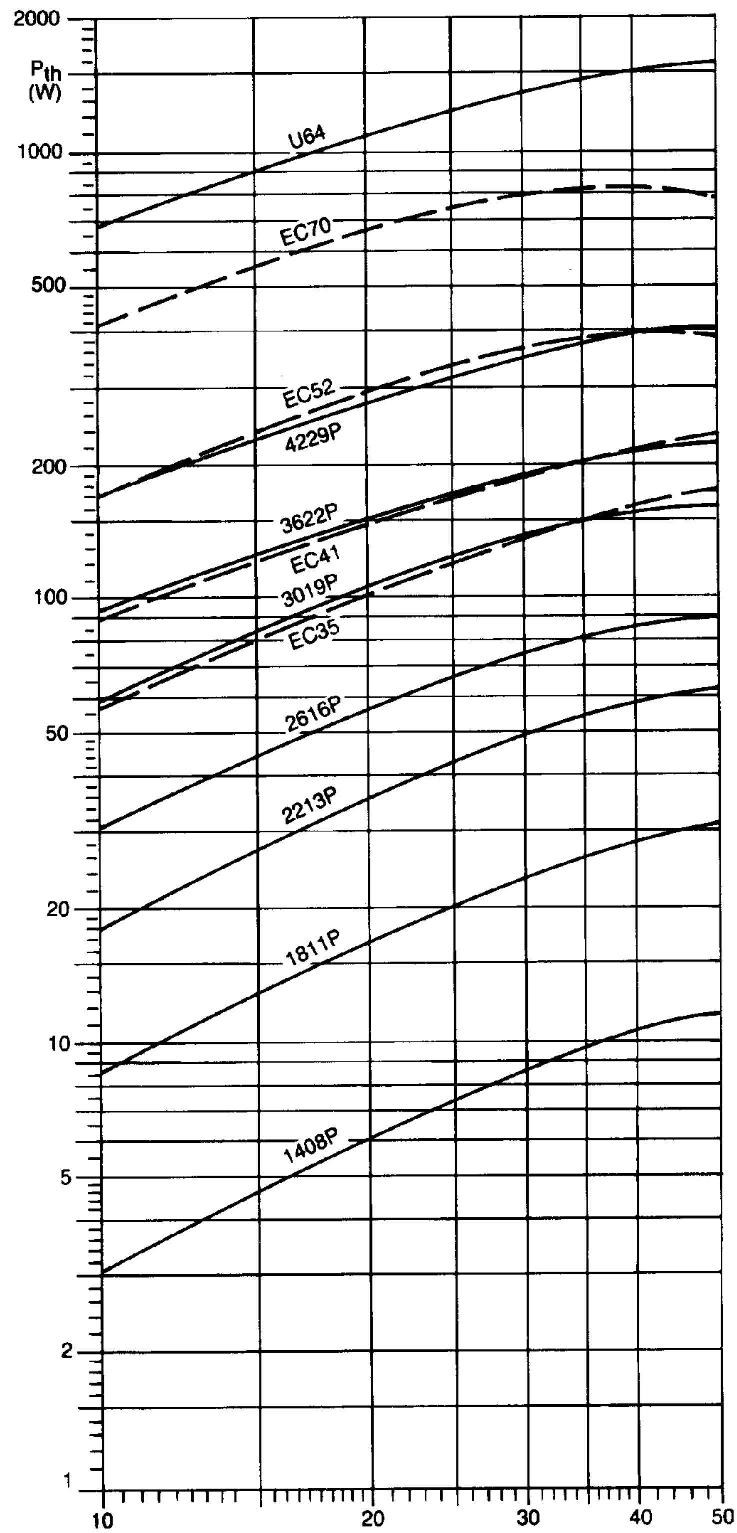
This component design, as does all others, begins by requesting all available literature from the appropriate manufacturers and then following this up with phone calls when specific questions arise. A partial list of companies is shown in Figure 8. Designs below 50 W generally use pot cores but for 50 W and above E cores are preferred. E cores expose the windings to air so that heat is not trapped inside and make it easier to bring out connections for several windings. Remember that flyback designs require lower permeability cores than the others. The classic approach is to consult manufacturers charts like the one shown in Figure 9 and then to pick a core with the required power handling ability. Both E and E-C (E cores with a round center leg) are popular now and they are available from several manufacturers. EC cores offer a performance advantage (better coupling) but standard E cores do cost less and are also used in these applications. Another approach that seems to work equally as well is to do a paper design of the estimated windings and turns required. Size the wire for 500 circular mils (CM) per amp and then find a core that has the required window area for this design. Now, before the windings are put on, it is a good idea to modify the turns so that they fit on one layer on an integral number of layers on that bobbin. This involves checking the turns per inch of the wire against the bobbin length. The primary generally goes on first and then the secondaries. If the primary hangs over an extra half layer, try reducing

FIGURE 8 — Partial List of Core (C) and Transformer (T) Manufacturers

Company	Location	Code
Ferroxcube Inc.	Saugerties, N.Y.	C
Indiana General	Keasby, N.J.	C
Stackpole	St. Marys, PA.	C
TDK	El Segundo, CA.	C
Pulse Engineering	San Diego, CA.	T
Coilcraft	Cary, IL.	T

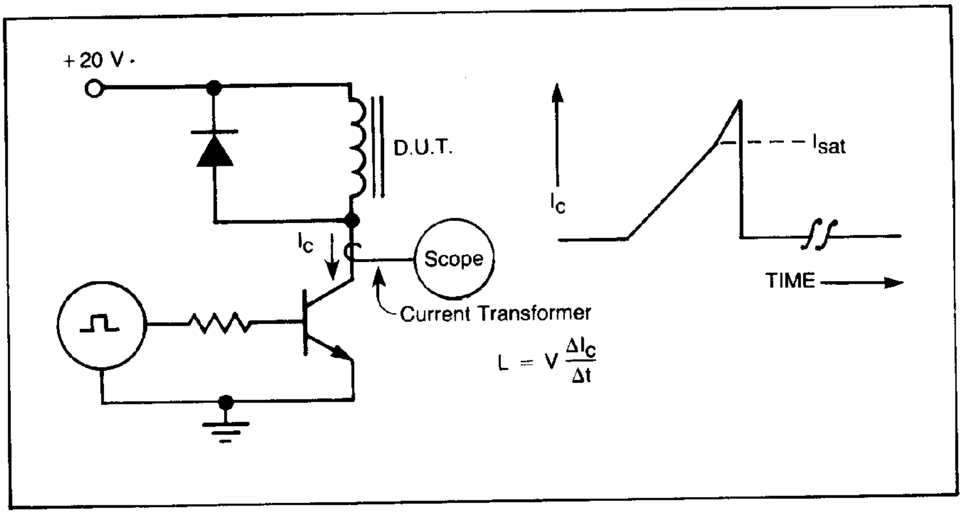
the turns or the wire size. Conversely, if the secondary does not take up a full layer, try bifilar winding (parallel) using wire half the size originally chosen; i.e., 3 wire sizes smaller like 23 vs. 20. This technique ultimately results in the use of foil for the higher current (20 A) low voltage windings. Most windings can be separated with 3 mil mylar (usually yellow) tape but for good isolation, cloth is recommended between primary and secondary.

FIGURE 9 — Core Selection for Bridge Configurations Compliments of Ferroxcube



Note: Power handling decreases by a factor of 2 in forward and 4 in flyback configurations.

FIGURE 10 — Simple Coil Tester



Finally, once a mechanical fit has been obtained, it is time for the circuit tests. The voltage rating is strictly a mechanical problem and is one of the reasons why U.L. normally does not allow high voltage bifilar windings. The inductance and saturating current level of the primary are inherent to the design, and should be checked in the circuit or other suitable test fixture. Such a fixture is shown in Figure 10 where the transistor and diode are sized to handle the anticipated currents. The pulse generator is run at a low enough duty cycle to allow the core to reset. Pulse width is increased until the start of saturation is observed (I_{SAT}). Inductance is found using:

$$L = E/(di/dt)$$

In forward converters, the transformer generally has no gap in order to minimize the magnetizing current (I_M). For these applications the core should be chosen large enough so that the resulting LI product insures that I_M at operating voltages is less than I_{SAT}. For flyback designs, a gap is necessary and the test circuit is useful again to evaluate the effect of the gap. The gap will normally be quite large

$$L_g >> L_m/u$$

where:
- L_g = gap length
- L_m = magnetic path length
- u = permeability

Under this stipulation, the gap directly controls the LI parameters and doubling it will decrease L by two and increase I_{SAT} by two. Again, the anticipated switching currents must be less than I_{SAT} when the core is gapped for the correct inductance.

Transformer tests in the actual supply are usually done with a high voltage dc power supply on the primary and with a pulse generator or other manual control for the pulse width (such as using the control IC in the open loop configuration).

Here the designer must recheck three areas:

1. No evidence of core saturation
2. Correct amount of secondary voltage
3. Minimum core or winding heat rise

If problems are detected in any of these areas, the ultimate fix may be to redesign using the next larger core size. However, if problems are minimal, or none exist, it is possible to stay with the same core or even consider using the next smaller size.

Transistors

The initial selection of a transistor for a switcher is basically a problem of finding the one with voltage and current capabilities that are compatible with the application. For the final choice performance and cost tradeoffs among devices from the same or several manufacturers have to be weighed. Before these devices can be put in the circuit, both protective and drive circuits will have to be designed.

Motorola's first line of devices for switchers were trademarked "Switchmode" transistors and introduced in the early 70's with data sheets that provided all the information that a designer would need including reverse bias safe operating area (RBSOA) and performance at elevated temperature (100°C). The first series was the

FIGURE 11 — Motorola High Voltage Switching Transistor Technologies

Family	Typical Device	Typical Fall Time	Approximate Switching Frequency
SWITCHMODE I	2N6545 MJE13005 MJE12007	200–500 ns	20K
SWITCHMODE II	MJ13081	100 ns	100K
SWITCHMODE III	MJ16010	50 ns	200K
TMOS	MTP5N40	20 ns	500K

2N6542 through 2N6547, TO-3 and was followed by the MJE13002 through MJE13009 series in a plastic TO-220 package. Finally, high voltage (1.0 kV) requirements were met by the metal MJ8500 thru MJ8505 series and the plastic MJE8500 series. And just recently, Motorola introduced the two new families of "Switchmode" transistors shown in Figure 11. The Switchmode II series is an advanced version of Switchmode I that features faster switching. Switchmode III is the state of today's bipolar art with both exceptional speed and RBSOA. Here, device cost is somewhat higher, but system costs may be lowered because of reduced snubber requirements and higher operating frequencies. A similar argument applies to Motorola TMOS Power FETs. These devices make it possible to switch efficiently at higher frequencies (200 to 500 kHz) but the main selling point is that they are easier to drive. This latter point is the one most often made to show that systems savings are again quite possible even though the initial device cost is higher.

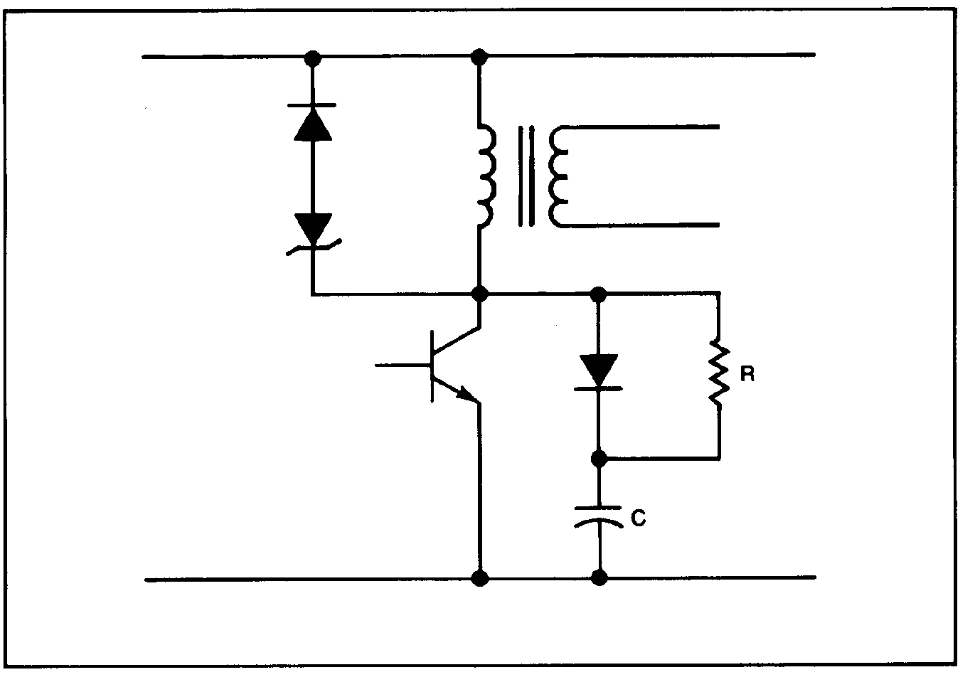
Figure 12 is a review of the transistor voltage requirements for the various off-line converter circuits. As illustrated, the most stringent requirement for single transistor circuits (flyback and forward) is the blocking or V_{CEV} rating. Bridge circuits, on the other hand, turn-on and off from the dc bus and their most critical voltage is the turn-on or V_{CEO} (SUS) rating.

Most Switchmode transistor load lines are inductive during turn-on and turn-off. Turn-on is generally inductive because the short circuit created by output rectifier reverse recovery times is isolated by leakage inductance in the transformer. This inductance effectively snubs most turn-on load lines so that the rectifier recovery (or short circuit) current and the input voltage are not applied simultaneously to the transistor. Sometimes primary interwinding capacitance presents a small current spike but usually turn-on transients are not a problem. Turn-off transients due to this same leakage inductance, however, are almost always a problem. In bridge circuits, clamp diodes can be used to limit these voltage spikes. If the resulting inductive load line exceeds the transistors reverse bias switching capability (RBSOA) then an RC network may also be added across the primary to absorb some of this transient energy. The time constant of this network should equal the anticipated switching time of the transistor (100 ns to 1.0 μs). Resistance values of 100 to 1000 ohms in this RC network are generally appropriate. Trial and error will indicate how low the resistor has to be to provide the correct amount of snubbing. For single transistor converters, the snubber shown in Figure 13 is generally used. Here slightly different criterion are

FIGURE 12 — Power Transistor Voltage Chart

Line Voltage	Circuit			
	Flyback, Forward or Push-Pull		Half or Full Bridge	
	V_{CEV}	$V_{CEO(sus)}$	$V_{CEO(sus)}$	V_{CEV}
220	850	400	400	400
120	450	200	200	200

FIGURE 13 — Zener Clamp and Snubber for Single Transistor Converters



used to define the R and C values:

$$C = \frac{I\, t_f}{V}$$

where I = The peak switching current
t_f = The transistor fall time
V = The peak switching voltage (Approximately twice the dc bus)

also $R = t_{on}/C$ (it is not necessary to completely discharge this capacitor in order to obtain the desired effects of this circuit)

where t_{on} = The minimum on-time or pulse width

and
$$P_R = \frac{CV^2 f}{2}$$

where P_R = The power rating of the resistor
and f = The operating frequency

Most of today's transistors that are used in 20 kHz converters switch slow enough so that most of the energy stored in the leakage inductance is dissipated by the snubber or transistor causing little voltage overshoot. Higher speed converters and transistors present a slightly different problem. In these newer designs snubber elements are smaller and voltage spikes from energy left in the leakage inductance may be a more critical problem depending on how good the coupling is between the primary and clamp windings.

Zener Diodes

If necessary, protection from voltage spikes may be obtained by adding a zener and rectifier across the primary as shown in Figure 13. Here Motorola's 1 W and 5 W zener lines with ratings up to 200 V can provide the clamping or spike limiting function. If the zener must handle most of the power, its size can be estimated using:

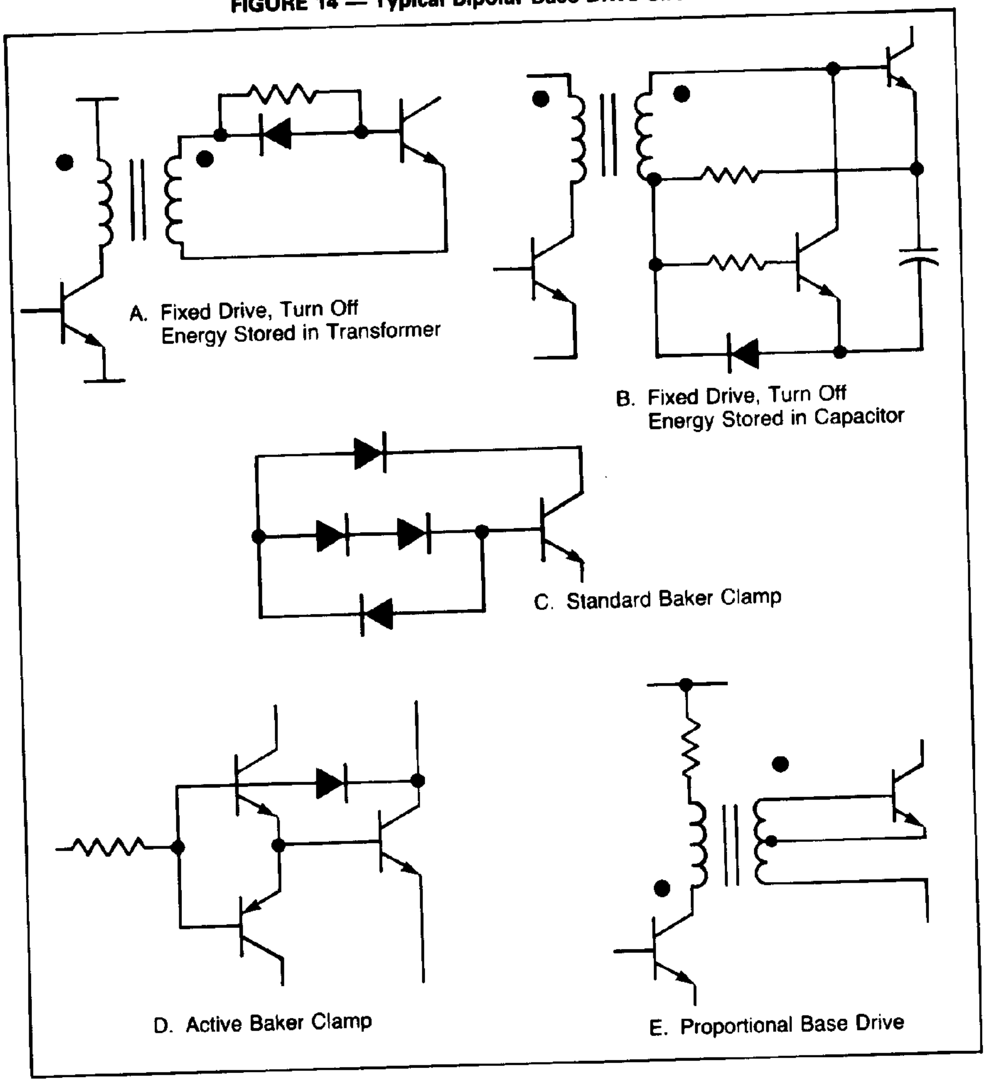
$$P_Z = \frac{L_L I^2 f}{2}$$

where P_Z = The zener power rating
and L_L = The leakage inductance (measured with the clamp winding or secondary shorted)
I = Peak collector current
f = Operating frequency

Mosorb Transient Suppressors

Distinction is sometimes made between devices trademarked Mosorb (by Motorola Inc.), and standard zener/avalanche diodes used for reference, low-level regulation and low-level protection purposes. It must be emphasized that Mosorb devices are, in fact, zener diodes. The

FIGURE 14 — Typical Bipolar Base Drive Circuits



basic semiconductor technology and processing are identical. The primary difference is in the applications for which they are designed. Mosorb devices are intended specifically for transient protection purposes and are designed, therefore, with a large effective junction area that provides high pulse power capability while minimizing the total silicon use. Thus, Mosorb pulse power ratings begin at 600 watts — well in excess of low power conventional zener diodes which in many cases do not even include pulse power ratings among their specifications.

MOVs, like Mosorbs, do have the pulse power capabilities for transient suppression. They are metal oxide varistors (not semiconductors) that exhibit bidirectional avalanche characteristics, similar to those of back-to-back connected zeners. The main attributes of such devices are low manufacturing cost, the ability to absorb high energy surges (up to 600 joules) and symmetrical bidirectional "breakdown" characteristics. Major disadvantages are: high clamping factor, an internal wear-out mechanism and an absence of low-end voltage capability. These limitations restrict the use of MOVs primarily to the protection of insensitive electronic components against high energy transients in applications above 20 volts, whereas, Mosorbs are best suited for precise protection of sensitive equipment even in the low voltage range — the same range covered by conventional zener diodes.

Drive Considerations

There are probably as many base drive circuits for bipolars as there are designers. Ideally, the transistor would like just enough forward drive (current) to stay in or near saturation and reverse drive that varies with the amount of stored base charge such as a low impedance reverse voltage. Many of today's common drive circuits are shown in Figure 14. The fixed drive circuits of 14A and 14B tend to emphasize economy, while the Baker clamp and proportional drive circuits of 14C, 14D and 14E emphasize performance over cost.

FET drive circuits are another alternative. The standard that has evolved at this time is shown in Figure 15. This transformer coupled circuit will produce forward and reverse voltages applied to the FET gate which vary with the duty cycle as shown. For this example, a V_{GS} rating of 20 V would be adequate for the worst case condition of high logic supply (12 V) and minimum duty cycle. And yet, minimum gate drive levels of 10 V are still available with duty cycles up to 50%. If wide variations in duty cycle are anticipated, it might be wise to consider using a semi-regulated logic supply for these situations. Finally, one point that is not obvious when looking at the circuit is that FETs can be directly coupled to many ICs with only 100 mA of sink and source capability and still switch efficiently at 20 kHz. However, to achieve switching efficiently at higher frequencies, several amps of drive may be required on a pulsed basis in order to quickly charge and discharge the gate capacitances. A simple example will serve to illustrate this point and also show that the Miller effect, produced by C_{DG}, is the predominant speed limitation when switching high voltages (see Figure 15B). A FET responds instantaneously to changes in gate voltage and will begin to conduct when the threshold is reached (V_{GS} = 2 to 3 V) and be fully on with V_{GS} = 7 to 8 V. Gate waveforms will show a step at a point just above the threshold voltage which varies in duration depending on the amount of drive current available which determines both the rise and fall times for the drain current. To estimate drive current requirements, two simple calculations with gate capacitances can be made:

1. $I_M = C_{DG}dv/dt$ and

and 2. $I_G = C_{GS}dv/dt$

FIGURE 15A — Typical Transformer Coupled FET Drive

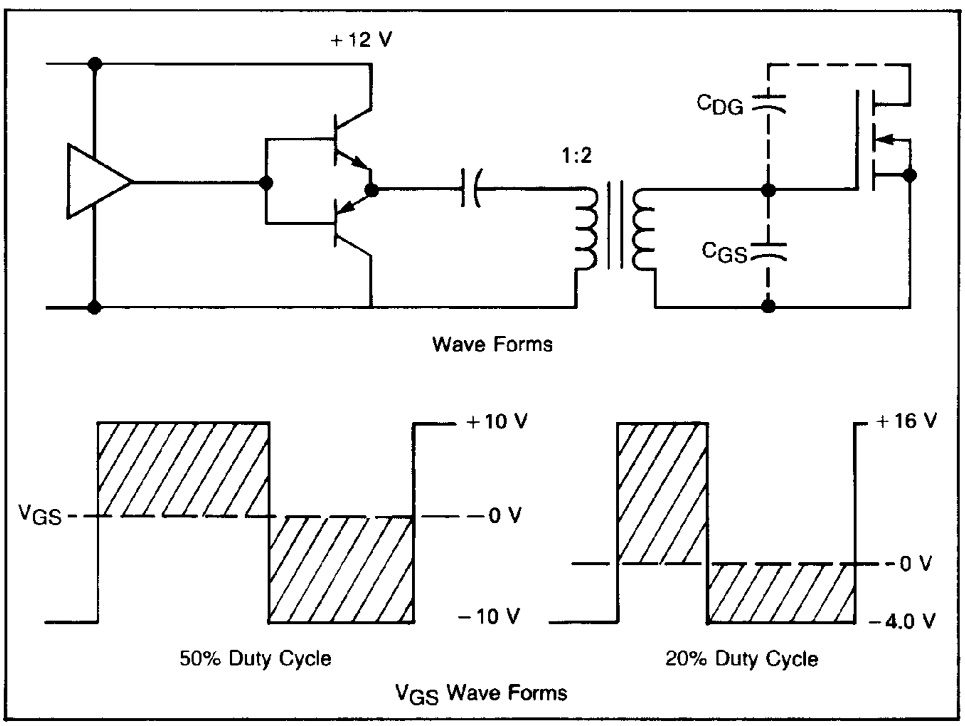
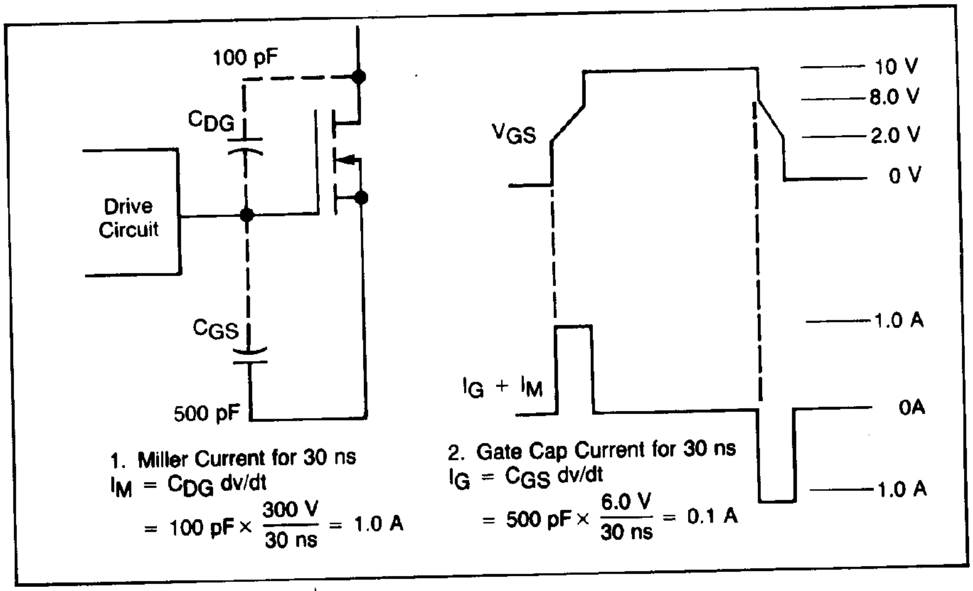


FIGURE 15B — FET Drive Current Requirements



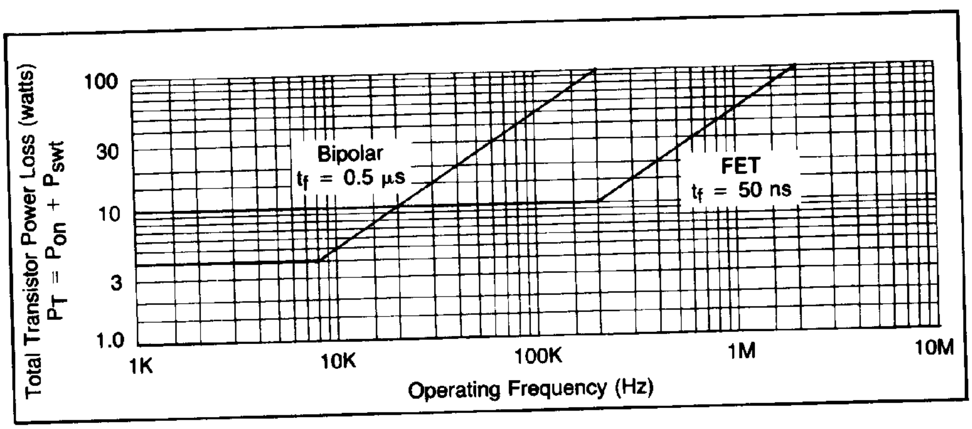
I_M is the current required by the Miller effect to charge the drain-to-gate capacitance at the rate it is desired to move the drain voltage (and current). And I_G is usually the lesser amount of current required to charge the gate-to-source capacitance through the linear region (2 to 8 V). As an example, if 30 ns switching times are desired at 300 V where C_{DG} = 100 pF and C_{GS} = 500 pF, then

$$I_M = 100\ pF \times 300\ V/30\ ns = 1.0\ A \text{ and}$$
$$I_G = 500\ pF \times 6\ V/30\ ns = 0.1\ A$$

This example shows the direct proportion of drive current capability to speed and also illustrates that for most devices, C_{DG} will have the greatest effect on switching speed and that C_{GS} is important only in estimating turn-on and turn-off delays.

are analyzed using Figure 16. Here, typical power losses for 5 A switching transistors versus frequency are shown. The FET (and bipolar) losses were calculated at 100°C rather than 25°C because on resistance and switching times are highest here and 100°C is typical of many applications. These curves are asymptotes of the actual device performance, but are useful in establishing the "break point" of various devices, which is the point where saturation and switching losses are equal. Since this is as low as 10K for some bipolars, it is possible that a FET even with high on-voltages can be competitive efficiency-wise at 200 kHz. The faster Switchmode II and III bipolar products would fall somewhere between the curves shown and therefore, be more competitive with FETs at the higher operating frequencies.

FIGURE 16 — Typical Switching Losses at 5 A and T_J = 100°C



Aside from its unique drive requirements, a FET is very similar to a bipolar transistor. Today's 400 V FETs complete with bipolar transistors in many switching applications. They are faster and easier to drive, but do cost more and have higher saturation, or more accurately, "on" voltages. The performance or efficiency tradeoffs

Rectifiers

Once components for the inverter section of a switcher have been chosen, it is time to determine how to get power into and out of this section. This is where the all important rectifier comes into play. The input rectifier is generally a bridge that operates off the ac line and into a capacitive filter. For the output section, most designers use Schottkys for efficient rectification of the low voltage, 5 V output windings and for the higher voltage, 12 to 15 V outputs, the more economical fast recovery or ultrafast diodes are used.

4. If I_S <Ip, consider either increasing the limiting resistor (R_S) or utilizing a larger diode.

In the output section where high frequency rectifiers are needed, there are several types available to the designer. In addition to the Schottky (SBR) and fast recovery (FR), there is also an ultra fast recovery (UFR) which fills the gap between the 50 V Schottky and the 600 V fast recovery lines. Comparative performance for devices with similar current ratings is shown in Figure 18. The obvious point here is that lower forward voltage improves efficiency and lower recovery times reduces turn losses in the switching transistors, but the tradeoff is higher cost. As stated earlier, Schottkys are generally used for 5.0 V outputs and fast recovery devices for 12 V outputs and greater. The ultra fast is competing primarily with the Schottky in those applications where cost is more important than efficiency. Of these devices, only the Schottky may need special handling. Ten years ago Schottkys were very fragile and could fail short from either excessive dv/dt (1.0 to 5.0 volts per nanosecond) or reverse avalanche. Present day devices, however, have something similar to Motorola's "guard ring" and internal zener which minimizes these earlier problems and reduces the need for RC snubbers and other external protective networks.

FIGURE 17 — Choosing Input Rectifiers

	SBR	UFR	FR
V_F	0.5–0.6	0.9–1.0	1.2–1.4
t_{rr}	10 ns	25 ns	150 ns
t_{rr} FORM	"SOFT"	"ABRUPT"	"EITHER"
V_R	30–50 V	50–150 V	50–600 V

NOTES: 1. Low V_F improves efficiency
2. Low t_{rr} reduces transistor switching losses
3. Soft (verses abrupt) recovery reduces noise

For the process of choosing an input rectifier, it is useful to visualize the circuit shown in Figure 18. To reduce cost, most earlier approaches of using choke input filters, soft start relays (Triacs), or SCRs to bypass a large limiting resistor have been abandoned in favor of using small limiting resistors or thermistors and a large bridge. The bridge must be able to withstand the surge currents that exist from repetitive starts at peak line. The procedure for finding the right component and checking its fit is as follows:

1. Choose a rectifier with 2 to 5 times the average I_O required
2. Estimate the peak surge current (I_p) and time (t) using:

$$I_p = \frac{1.4\,V_{in}}{R_S} \qquad t = R_S C$$

Where V_{in} is the RMS input voltage
R_S is the total limiting resistance, and
C is the filter capacitor size

3. Compare this current pulse to the sub cycle surge current rating (I_S) of the diode itself. If the curve of I_S versus time is not given on the data sheet, the approximate value for I_S at a particular pulse width (t) may be calculated knowing:
   - I_{FSM} – the single cycle (8.3 ms) surge current rating and using.
   - $I^2\sqrt{t} = K$ which applies when the diode temperature rise is controlled by its thermal response as well as power (i.e., $T = K'P\sqrt{t}$ for $t < 8$ ms. This gives:

$$I_S{}^2\sqrt{t} = I^2{}_{FSM}\sqrt{8.3\text{ ms}} \text{ or}$$

$$I_S = I_{FSM}\left(\frac{8.3\text{ ms}}{t}\right)^{1/4}, \text{ t is in milliseconds}$$

Power Triacs and Inrush Current Limiting

Many high current PWM switching supplies operate directly off the ac line. They have very large capacitive input filters with high inrush surge currents. The line circuit breaker and the rectifier bridge must be protected during turn-on.

Surge current limiting can be accomplished by adding R_S and an SCR "short" after charging C1 as shown in Figure 19, or by phase controlling the line voltage with a Triac.

FIGURE 18 — Output Rectifier Type Comparisons

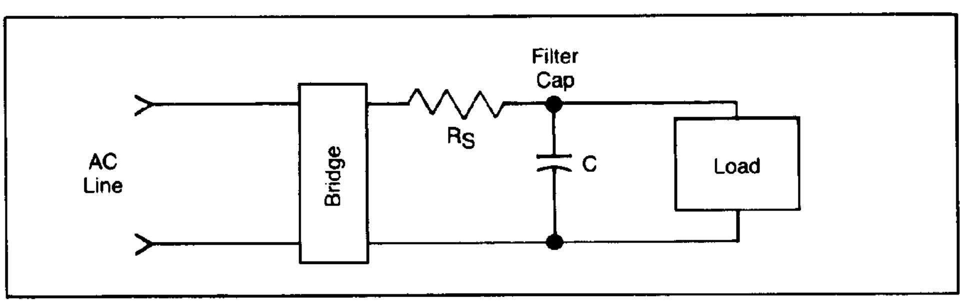
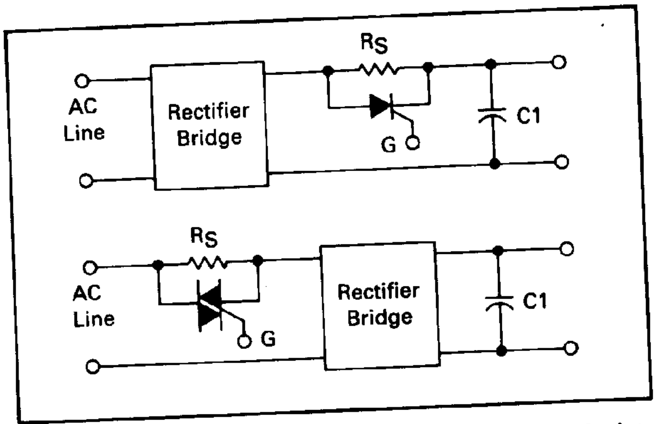


FIGURE 19 — Surge Current Limiting For A Switching Power Supply



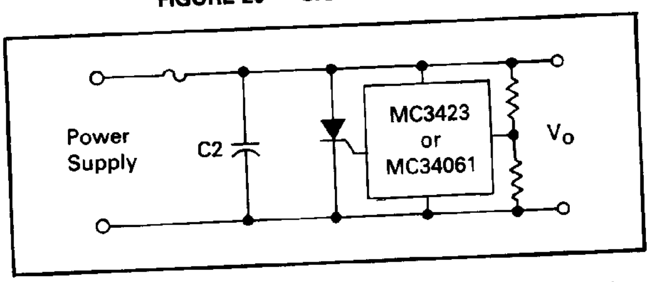
For further information, see EB-78 and MC3420 data sheet.

Power SCRs for Crowbar Applications

Linear and switching power supplies can be protected from overvoltage with a crowbar circuit. For linear supplies, the pass transistor can fail shorted, allowing high line transformer voltage to the load. For switching power supplies, a loose or disconnected remote sense lead can allow high voltage to the load.

The crowbar circuit, shown below ignores noise spikes but will fire the SCR when a valid overvoltage condition is detected. The SCR will discharge C2 and either blow the fuse or cause the power supply to shut down.

FIGURE 20 — Crowbar Circuit



For further information, see AN-568 and MC3423 data sheet.

Capacitors and Filters

In today's 20 kHz switchers, aluminum electrolytics still predominate. The good news is that most have been characterized, improved, and cost reduced for this application. The input filter requires a voltage rating that depends on the peak line voltage; i.e., 400 to 450 V for a 220 V switcher. If voltage is increased beyond this point, the capacitor will begin to act like a zener and be thermally destroyed from high leakage currents if the rating is exceeded for enough time. In doubler circuits, voltage sharing of the two capacitors in series can be a problem. Here extra voltage capability may be needed to make up for the imbalances caused by different values of capacitance and leakage current. A bleeder resistor is normally used here not only for safety but to mask the differences in leakage current. The RMS current rating is also an important consideration for input capacitors and is an example of improvements offered by today's manufacturers. Earlier "lytics" usually lacked this rating and often overheated. Large capacitors that were not needed for performance were used just to reduce this heating. However, today's devices like the swedged variety from Mepco-Electra offer lower thermal resistance, improved connection to the foil and good RMS ratings. A partial list of manufacturers that supply both high voltage input and the lower voltage output capacitors for switchers is shown in Figure 21. Most of the companies offer not only the standard 85°C components, but devices with up to 125°C ratings which are required because of the high ambient temperatures (55 to 85°C) that many switchers have to operate in, many times without the benefit of fans.

For output capacitors the buzz word is low ESR (equivalent series resistance). It turns out that for most capacitors even in the so-called "low ESR" series, the output ripple depends more on this resistance than on the capacitor value itself. Although typical and maximum ESR ratings are now available on most capacitor designed for switchers, the lead inductance generally is not specified except for the ultra-high frequency four terminal capacitors from some vendors. This parameter is responsible for the relatively high switching spikes that appear at the output. However, at this point in time, most designers find it less costly and more effective to add a high frequency noise filter rather than use a relatively expensive capacitor with low equivalent series inductance (ESL).

These LC noise or spike filters are made using small powdered iron toriods (1/2 to 1″ OD) with distributed windings to minimize interwinding capacitance. And the output is bypassed using a small 0.1 μF ceramic or a 10 to 50 μF tantalum or both. Larger powered iron toriods are often used in the main LC output filter although the higher permeability ferrite C and E cores with relatively large gaps can also be used. Calculations for the size of this component should take into account the minimum load so that the choke will not run "dry" as stated earlier.

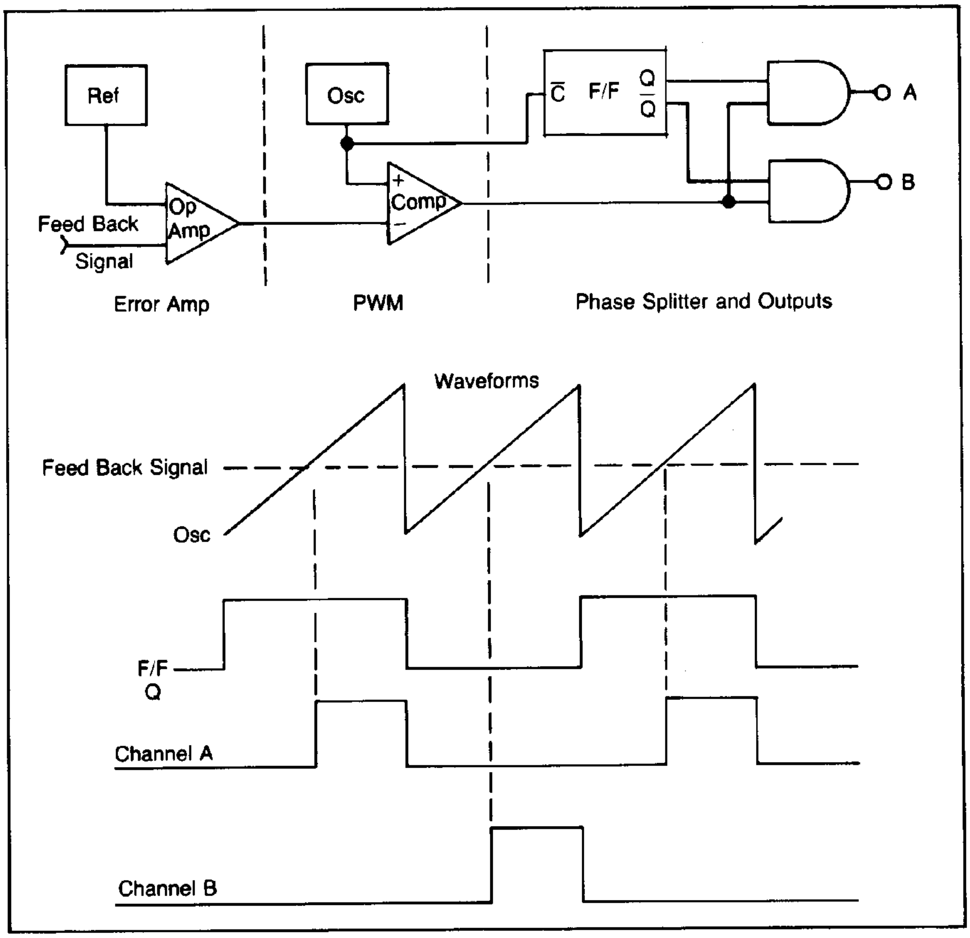
FIGURE 21 — Partial List of Capacitor Companies

Company (U.S.)	Location
Sprague	North Adams, MA
MEPCO/Electra	Columbia, SC
Cornell-Dublier	Sanford, NC
Sangamo	Pickens, SC
Mallory	Indianapolis, IN

Control Circuits

Ten years ago, discrete control circuits were in use and only bits and pieces of ICs could be found. Since that time, various semiconductor companies recognized the designers' needs for a dedicated control IC and now a variety of these circuits are on the market and widely used. They may provide the designer with a cost incentive over the discrete or a simpler control circuit or both. Internally, most of these resemble the functional configuration shown in Figure 22. The basic regulating function is performed in the pulse width modulator (PWM) section. Here, the dc feedback signal is compared to a fixed frequency sawtooth (or triangular) waveform. The result

FIGURE 22 — Basic SM Control IC



is a variable duty cycle pulse train which, with suitable buffer or interface circuits can be used to drive the power switching transistor. Some ICs provide only a single output while others provide the phase splitter shown to alternately pulse two output channels. In this latter case provisions are usually made either internally or by wire "or" ing the outputs to convert the dual to a single output channel. Additionally most ICs provide the error amplifier section shown as a means to process, compare and amplify the feedback signal.

Features required by a control IC vary to some extent because of the particular needs of a designer and on the circuit configuration chosen. However, most of today's current generation ICs have evolved with the capabilities or features listed in Figure 23. It is primarily the cost differences in these parts that determines whether all or only part of these features will be incorporated. Most of these are evident to the designer who has already started comparing data sheets, except perhaps for the hiccup and feed forward features. The hiccup terminology is

FIGURE 23 — Desirable Features of Switchmode Control ICs

- PROGRAMMABLE (TO 500 kHz) FIXED FREQUENCY OSCILLATOR
- LINEAR PWM SECTION WITH DUTY CYCLE FROM 0 TO 100 %
- ON BOARD ERROR AMPLIFIERS
- ON BOARD REFERENCE REGULATOR
- ADJUSTABLE DEAD TIME
- UNDERVOLTAGE (LOW V_{CC}) INHIBIT
- GOOD OUTPUT DRIVE (100 TO 200 mA)
- OPTION OF SINGLE OR DUAL CHANNEL OUTPUT
- UN-COMMITTED OUTPUT COLLECTOR AND EMITTER OR TOTEM POLE DRIVE CONFIGURATION
- SOFT START
- CURRENT LIMITING WITH "HICCUP MODE" AS BACKUP
- SYNC CAPABILITY

used to indicate a current limited mode of operation like a dead short that would produce excessive switching losses if the inverter were to continue to free run. The hiccup function interrupts the output pulses for a set amount of time and then allows a restart (usually with an audible amount of noise). Such a switcher will hiccup indefinitely until the short is removed. Feed forward is a feature that was added to improve the ripple rejection or line regulation or both. These ICs feed a sample of the input voltage directly to the PWM section. In this way, the switcher can anticipate the need to vary pulse width in response to a change in the input and not have to wait for the normal delays before its effect would be seen in the output. A partial list of control (and supervisory) ICs available to today's designer is given in Figure 24. Because low cost and second sources are important, parts like the TL494 (available from TI and Motorola) and Silicon General's 1524 (available from a variety of sources) have already captured a large share of this market. There is also a more limited number of applications that will require the high performance capabilities and will pay the cost premiums for parts like the SG1526, NE5560 and the ZN1066.

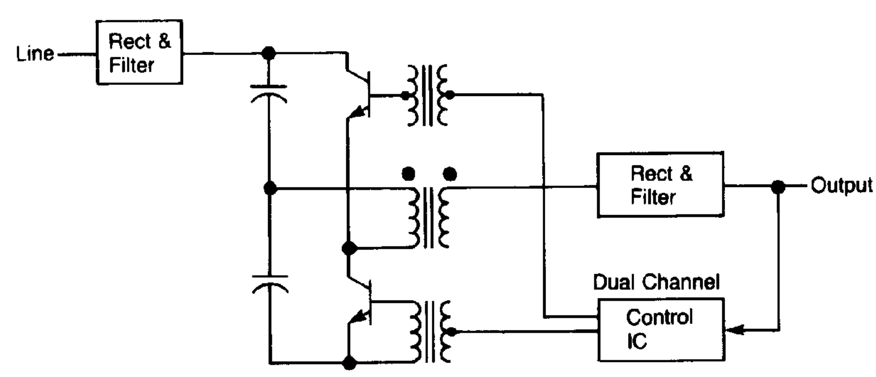
Today there is a need for a simple, low cost, single channel control IC for low power (20 to 200 W) applications like Motorola's MC34060. This component would be used to run the low power flyback type configurations and probably would be part of a three chip rather than single chip system. The differences in these two approaches are illustrated in Figure 25.

FIGURE 24 — Partial Listing of SM ICs

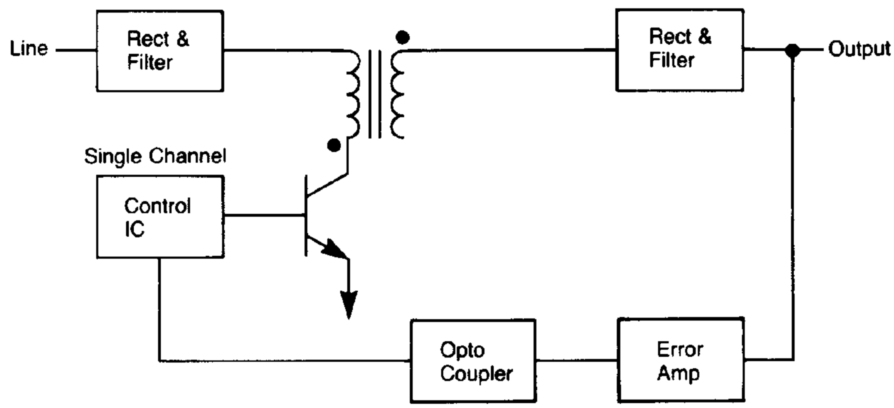
Single Channel Control ICs		Dual Channel Control ICs	Protection/Interface/Supervisory Circuits	
TL497	TEA1001	TL494*	MC3423*	OVP
NEC1042	SL442	SG1526*	TL431*	REF, OP AMP
uA78s40*	TDA240	SG1525/1527*	MC3424*	OUVP
SE5560	MC34060*	MC3420*	MC3425*	OUVP
SE5561	TDA4600*	SG1524	MMH0026*	DRIVER
		ZN1066		

***NOTE:** Either available from or soon to be introduced by Motorola

FIGURE 25 — Control Circuit Topologies



A. Single Chip System — Drive Transformer Isolation



B. Three Chip System — Opto Coupler Isolation

When it is necessary to drive two or more power transistors, drive transformers are a practical interface element and are driven by the conventional dual channel IC just discussed (Figure 25A). In the case of a single transistor converter, however, it is usually more cost effective to directly drive the transistor from the IC (Figure 25B). In this situation, an optocoupler is commonly used to couple the feedback signal from the output back to this control IC. And the error amplifier in this case is nothing more than an op amp and reference.

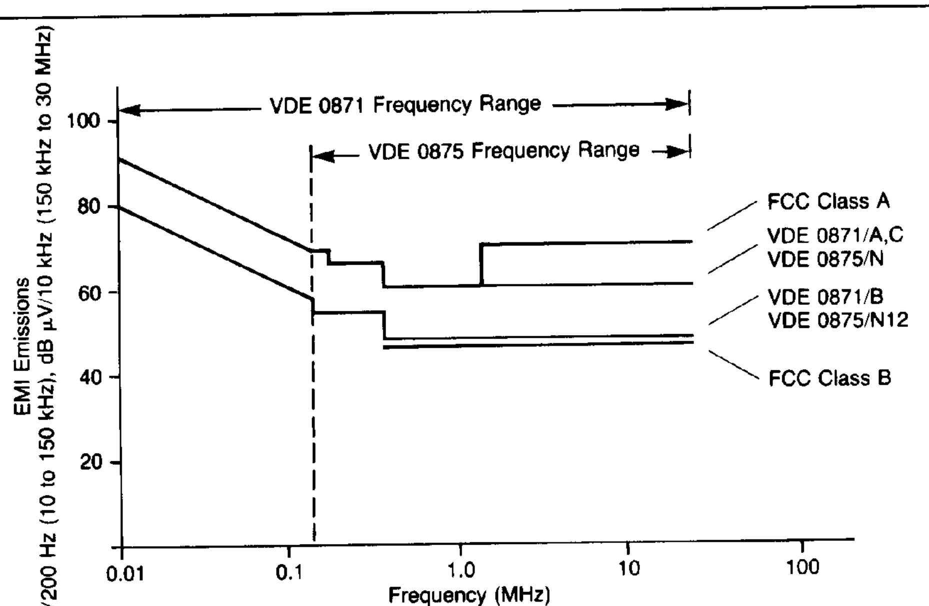
The Future

The future offers a lot of growth potential for switchers in general and low power switchers (50 – 200 watts) in particular. The latter are responding to the growth in microprocessor based equipment as well as computer peripherals. Today's configurations have already been challenged by the sine wave inverter which reduces noise and improves transistor reliability but does effect a cost penalty. Also, a trend to higher switching frequencies to reduce size and cost even further has begun. The latest bipolar can operate efficiently up to 100 kHz and the FET seems destined to own the 200 to 500 kHz range. These newer switchers have not yet realized a significant cost savings primarily because of deficiencies in the passive component area.

The growth pattern predicted at this time can possibly be impacted by noise problems. Originally governed only by MIL specs and the VDE in Europe, now (effective October 1981) the FCC has released a set of specifications that apply to electronic systems which often include switchers (see FCC Class A in Figure 26). It seems probable, however, that system engineers or power supply designers will be able to add the necessary line filters and EMI shields without evoking a significant cost penalty which would slow the growth of switchers.

The most optimistic note concerning switchers is in the component area. Switching power supply components have actually evolved from components used in similar applications. And it is very likely that newer and more mature products specifically for switchers will continue to appear over the next several years. The ultimate effect of this evolution will be to further simplify, cost reduce and increase the reliability of these designs.

FIGURE 26 — Noise Limits



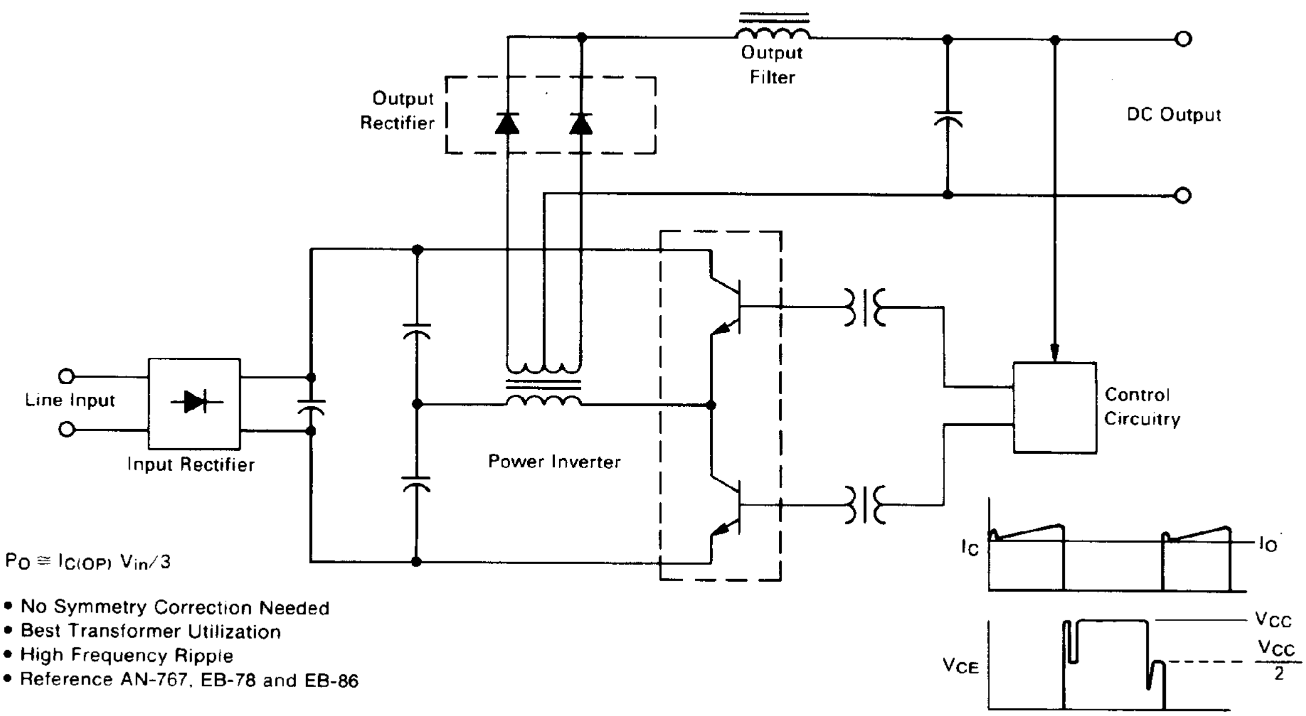
Notes: 1. FCC Class A covers commercial, Class B covers residential.
2. Also SCE VDE 0871/0875 for noise and VDE 0730 or UL478 for safety.

Basic Switching Power Supply Configurations

Minimum device voltage rating recommended for these two circuits:
See Tables 1, 2 and 3, for recommended devices.

V_{in} V_{ac}	V_{DSS} or $V_{CEO(sus)}$ V_{dc}
120	200
220	400
380	600

FIGURE 1 — Basic Half-Bridge Configuration



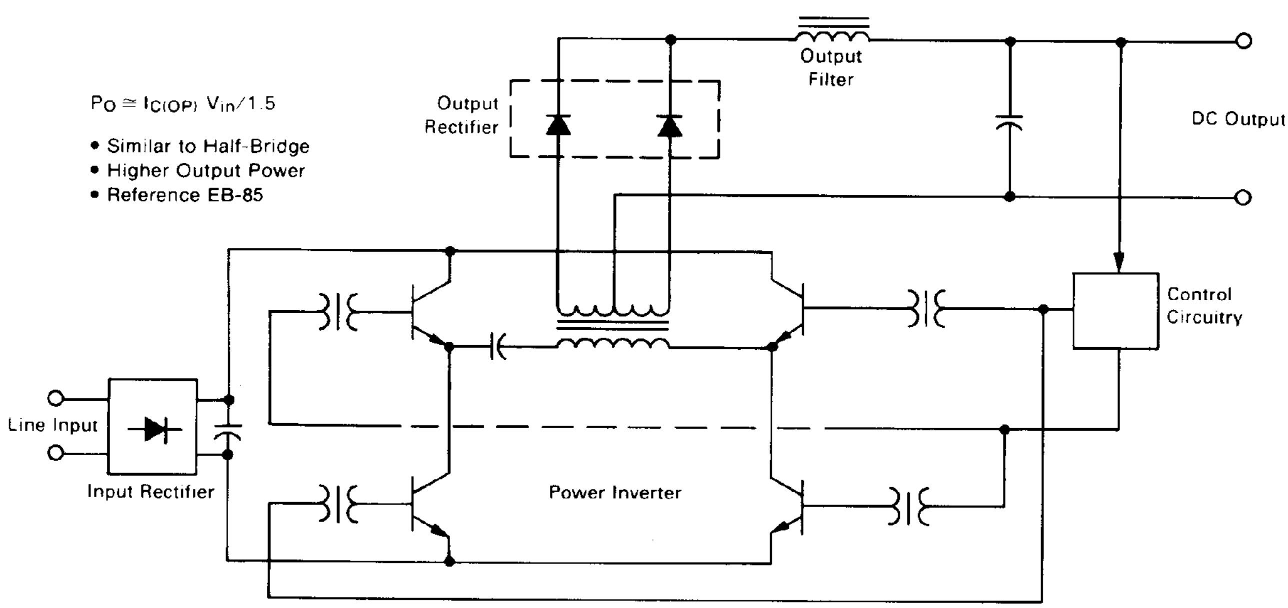
$P_O \cong I_{C(OP)} V_{in}/3$

- No Symmetry Correction Needed
- Best Transformer Utilization
- High Frequency Ripple
- Reference AN-767, EB-78 and EB-86

FIGURE 2 — Basic Full-Bridge Configuration

$P_O \cong I_{C(OP)} V_{in}/1.5$

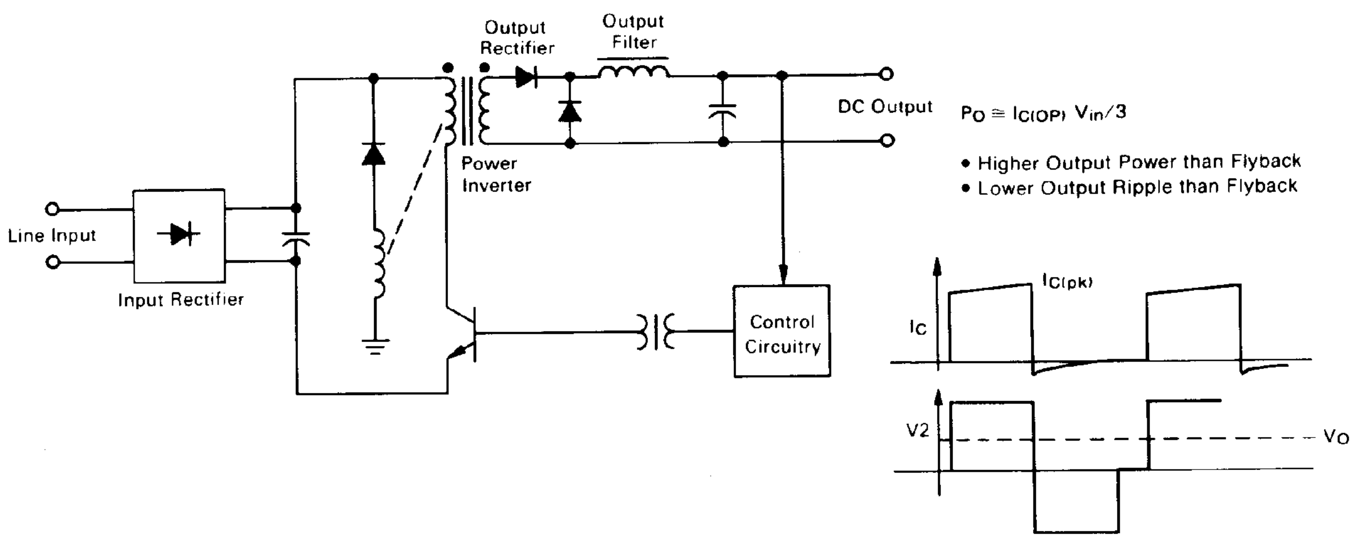
- Similar to Half-Bridge
- Higher Output Power
- Reference EB-85



Minimum recommended device voltage rating for these three circuits:
See Tables 4 and 5, for recommended devices.

V_{in} V_{ac}	V_{DSS} or V_{CEV} V_{dc}
120	450
220	850

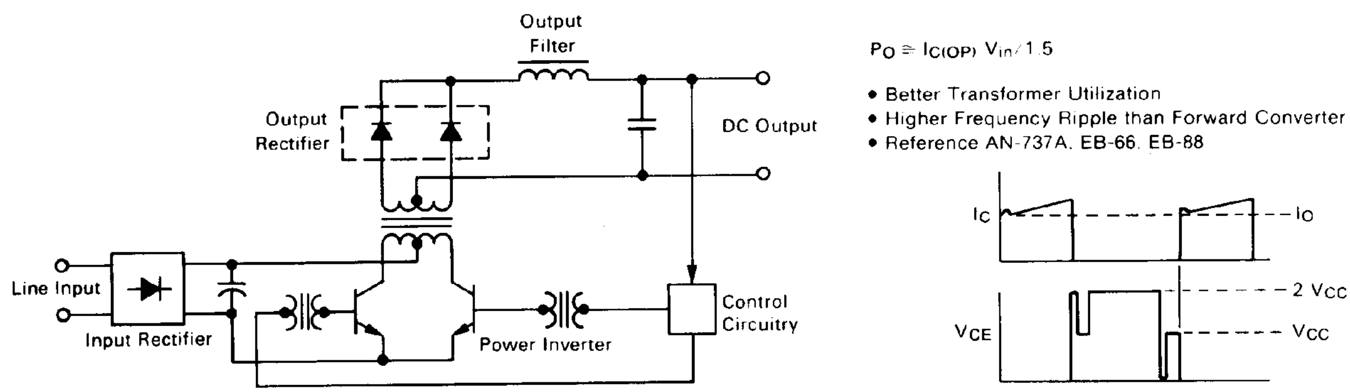
FIGURE 3 — Basic Forward Converter



$P_O \cong I_{C(OP)}\ V_{in}/3$

- Higher Output Power than Flyback
- Lower Output Ripple than Flyback

FIGURE 4 — Basic Push-Pull Configuration



$P_O \cong I_{C(OP)}\ V_{in}/1.5$

- Better Transformer Utilization
- Higher Frequency Ripple than Forward Converter
- Reference AN-737A, EB-66, EB-88

FIGURE 5 — Basic Flyback Configuration

$P_O \cong I_{C(OP)}\ V_{in}/6$

- No Filter Inductance
- Low Power Multiple Output Systems
- Reference EB-87