Flyback converters: solid-state solution to low-cost switching power supplies

For the user, close attention to design detail will yield a high-performance system with low production costs

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☐ Long a fixture in high-power applications like mainframes and their peripherals, the solid-state switching power supply is only now making its mark in low-power applications where the low-cost linear supply predominates. Momentum for this thrust comes from simplified switcher designs that lower the cost of converting an ac input to multiple dc outputs.

Typically, most switching converters have been custom designs, tailored to solve a particular problem or set of problems. When equipment makers ran into size, weight, or cooling restrictions as they designed supplies into their gear, they ordered the custom switchers.

With the new standard converters, these manufacturers are finding they can profit from such advantages as improved portability and the elimination of cooling fans, without sacrificing economy. Thus switchers are beginning to show up in volume in such low-power applications as microcomputer systems, home-computer video terminals, and small commercial computer systems.

Minimizing cost

The new switchers may be the product of a common design philosophy, but their different makers do not agree on how to minimize costs. Nor do they agree on what cost-performance tradeoffs to make in most cases. The original-equipment manufacturer, faced with this controversy, can resolve it for his particular applications — but only if he is familiar with some of the basic design considerations in this field.

What he will find for his consideration is such designs as the Boschert OL25 series. These general-purpose, low-power switching supplies are designed to compete directly against linear supplies by reconciling the lowest possible cost with acceptable performance in most cases.

These switchers (Fig. 1) are an adaptation of the flyback, or ringing choke converter, generally used in television receivers' horizontal circuits, xenon flash units, and other very low-cost applications. Other techniques are used by other switcher manufacturers, but the flyback approach minimizes the costs of both magnetic and semiconductor components. The new Boschert line attacks the problem of high regulation costs by novel power-control methods.

The initial choice

In theory, all switchers are alike; in practice, their designers can choose among a number of basic approaches (Fig. 2). Usually they select the one that provides the required performance with no need for added components or a major engineering effort. Often, a better decision is to select a lower-cost and -performance basic circuit and to design in the higher performance with an engineering effort that also seeks to minimize production costs.

Since the switcher essentially is an electronic system rather than a traditional power supply, it should be evaluated as a system, rather than as the sum of its

1. The big punch. The OL25 switcher measures 2.5 by 4 by 6 inches and weighs 12 ounces; yet it delivers a hefty 25 watts of switching power. An adaptation of the flyback converter, the design minimizes both magnetic and semiconductor costs.

components. Coming into play are such considerations as consolidation of functions, regulation and ripple filtering, the number of power stages, and energy storage.

Most applications require multiple power outputs. Major savings may be achieved combining a function necessary for each output stage in a shared circuit. For example, the filtering required after the power switch might be combined into a filter circuit at the outputs.

The OEM designer may find that he must allow for

tradeoffs in regulation and ripple to take advantage of the switcher's benefits without cost penalties. For example, the designer of a digital system wants a precise + 5-volt output to power most of the system.

Almost all switchers directly sense the + 5-v output and regulate it to ± 1% in order to keep it well within the typical digital component tolerance of ± 5%. However, auxiliary outputs usually are semi-regulated: that is, they are dependent upon the regulation provided by the control circuitry for the ± 5-v output.

The rationale for semi-regulation is that acceptable performance for low- and medium-power uses can be achieved on the auxiliary outputs without added power-handling stages. One circuit then can regulate all input voltage effects for all outputs, and acceptable load regulation can be met by minimized output impedances.

A fact of life

Ripple in switching power supplies tends to raise designers' emotions, for it is inescapable. The fast switching characteristics of these units means the phenomenon is inherent, so it is pointless to strive for the lowest possible ripple at any cost.

However, every converter output has a filter, which generally can be designed to reduce ripple to an acceptable level at a nominal cost. Ripple voltage levels within 2% of the dc output voltage usually are low enough for the ripple to become buried in the normal system noise.

In most high-power applications, semi-regulation on auxiliary outputs does not suffice. Therefore, the basic converter's cost should be minimized, and the savings should be used for such performance-improving functions as a linear post-regulator and a filter on each auxiliary output. The quasi-square-wave converter of Fig. 2a is the choice here, because it requires only one power conversion, minimizing semiconductor costs.
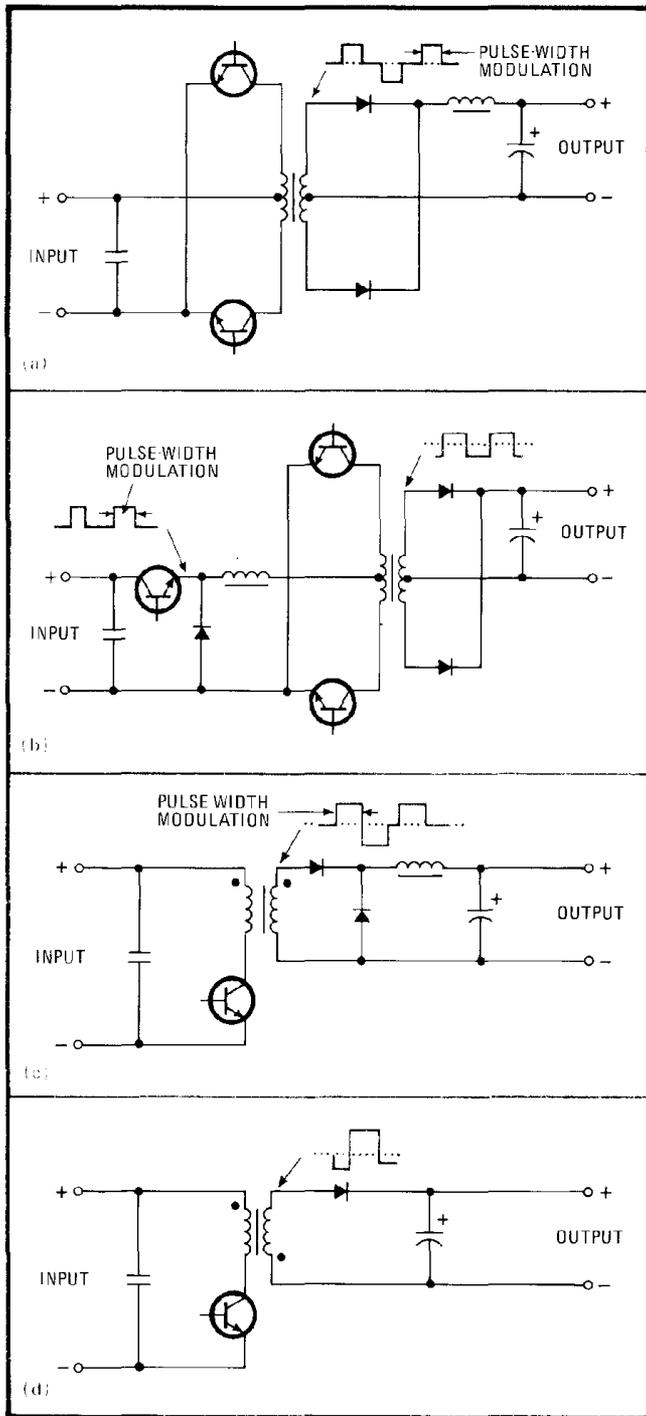
For medium-power applications, the two-stage switcher (Fig. 2b) is the best choice, because its improved power conditioning makes semi-regulation practical. Post-regulators are eliminated and filtering requirements are minimized, but at the expense of additional semiconductor devices in the basic unit.

However, the costs of medium-power semiconductors are reasonable, and the two-stage design shown trades off the cost of added devices against savings down the line. Also, it needs only one primary inductor, whereas the quasi-square-wave unit needs one for each output.
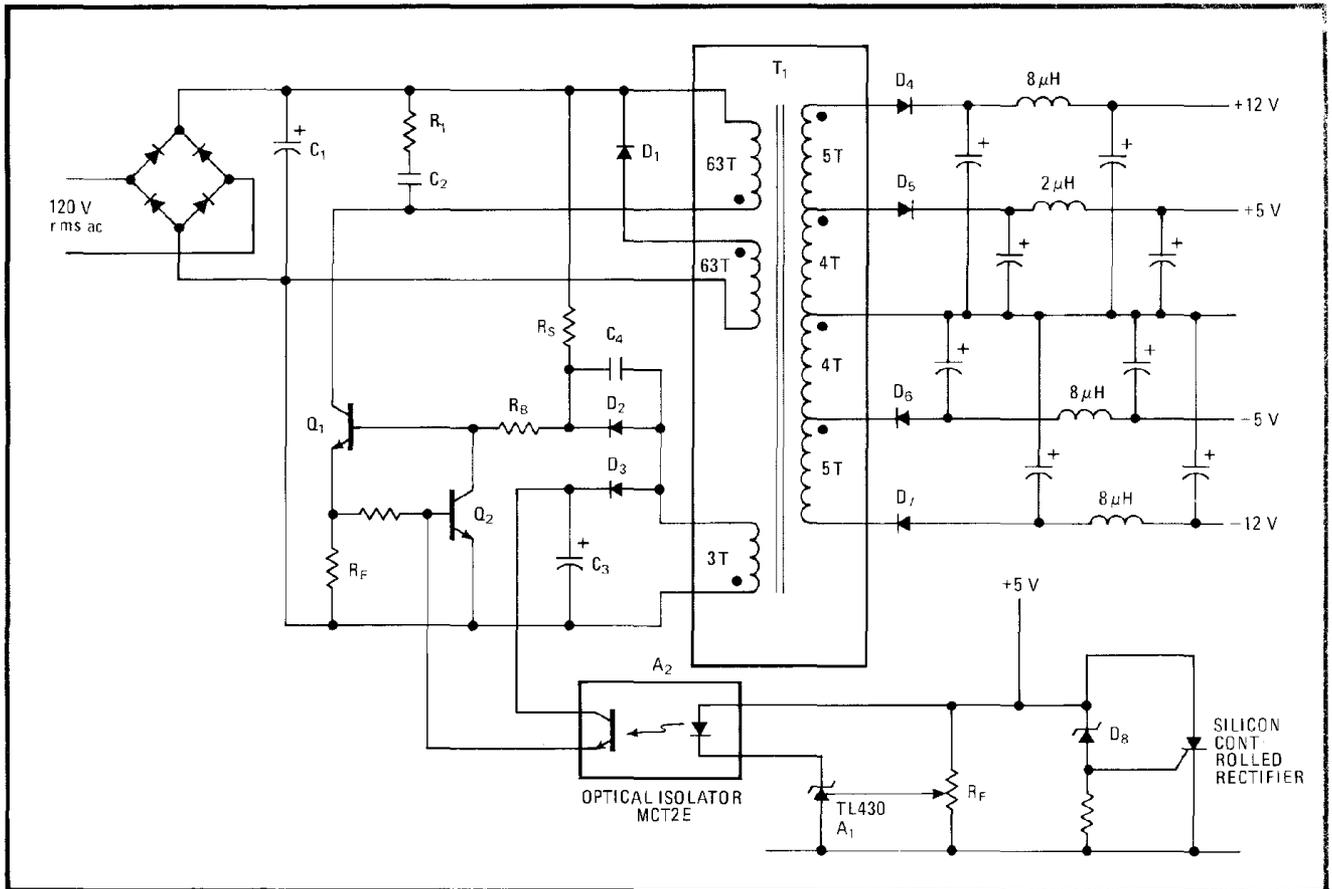
Where flyback fits in

Forward and flyback converters (Figs. 2c and d, respectively) could be used for medium-power applications, but they are too expensive at present component costs. A forward converter needs a large inductor in each output for energy storage. A flyback converter does not need such inductors, but its utilization of power semiconductor devices is less efficient, requiring the use of relatively expensive components.

For low-power applications, efficient utilization of components is less urgent, because the low-power semiconductor devices used are relatively inexpensive. Moreover, the OEM designer can choose a switcher from a number of converters with a single power switch and

2. Switching converters. Low cost makes the quasi-square-wave converter (a) attractive. For medium power, the two-stage switchers (b) excel. Forward and flyback converters (c and d) require clever design to offset their higher-priced parts.

3. The OL25 power supply. A novel control circuit to adjust current flow during operation is governed by a feedback from the +5-V output through A₁ and A₂ to transistors Q₁ and Q₂. The ±5-V outputs are regulated; the ±12-V are semi-regulated.

one output change.

Forward converters have been preferred over flyback switchers because of their relatively higher performance. However, a flyback-based power supply can offer equivalent performance to a forward unit if its designer plows back part of the savings in magnetic and semiconductor costs into design improvements. The savings in output inductors more than offsets the slight increase in transformer complexity needed to combine the isolation and storage functions.

Low-cost energy

Yet this more complex double-duty transformer contributes heavily to the flyback converter's major advantage: lower-cost energy storage. Often overlooked as so fundamental a fact of switcher design, energy storage represents a large chunk of total system costs, as well as being a key to the unit's efficiency.

Energy is stored when the power-switching transistor turns on and is delivered to the load when the transistor turns off. Switch-control regulation thus replaces the highly inefficient energy-dissipation techniques used to regulate linear power supplies.

However, most switchers must store their energy in expensive magnetic components. By storing energy in the isolation transformer, the flyback converter eliminates this component expense. Moreover, it needs the fewest components for each basic power-handling function. In fact, one component suffices for each function, because
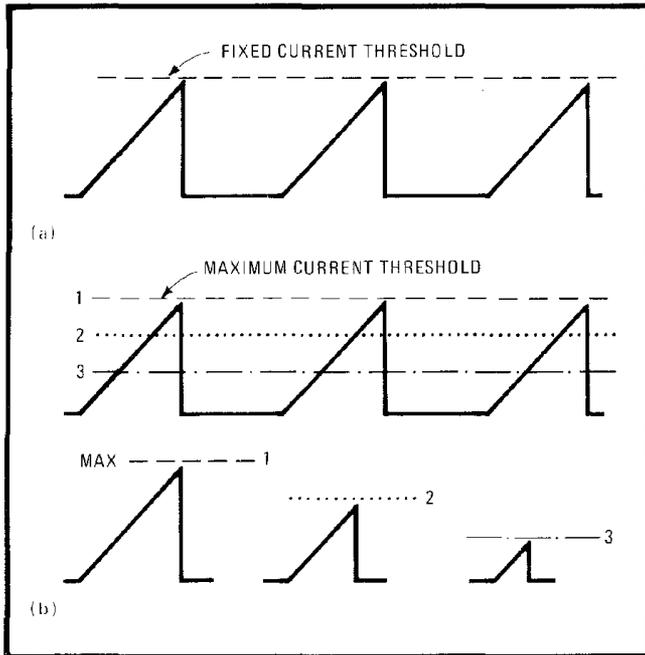
the flyback converter is a single-ended circuit.

As the Boschert OL25 series shows, the result can be a versatile design (Fig. 3) that provides 25 watts maximum in four continuous outputs over worst-case line-power conditions. The basic model has ±5-v regulated outputs and ±12-v semi-regulated outputs. A modular design approach allows outputs to be tailored for optional voltages throughout a ±40-v range.

The main challenge in flyback design is to minimize the cost of the control circuitry. In general, controlling the output power is achieved by a blocking oscillator power stage that changes state whenever total positive feedback gain from the switching section through the transformer exceeds unity. The blocking oscillator then usually delivers a fixed amount of power.

Controlling the power

The OL25 operates as a blocking oscillator under the following control law: output is linearly proportional to current flowing in the primary circuit at the time transistor Q₁ in Fig. 3 turns off. The simplifying assumption is made that output power is independent of input line voltage and operating frequency, so far as the control loop is concerned. Then the control-loop model for the power stage is simply a current source driving the output capacitances.

To adjust power flow during operation, the OL25 implements a novel control circuit. The new technique provides excellent regulation resulting in significant

4. Modulated threshold. The basic oscillator's primary current is limited by a fixed threshold (a), but the OL25's circuit (b) implements a modulated threshold technique that results in a variable duty cycle; hence, the power outputs are constantly regulated.

5. Flux modes. The OL25 converter operates in the discontinuous flux mode because of better cost tradeoffs, despite the greater average power delivery in the continuous mode operation. Also, the continuous mode operation requires complex control circuitry.

savings in the components required for control circuitry. Moreover, the circuitry occupies little more than 10% of the total board area.

In a basic blocking oscillator (Fig. 4a), primary current rises linearly until a preset current threshold is reached, then drops to zero until the end of the cycle. In the OL25 (Fig. 4b), the current threshold is modulated, resulting in a variable duty cycle and power adjustment. The control circuit is designed to home in on that duty cycle, ensuring proper regulation. The circuit programs the current level at which the power-stage feedback exhibits a gain greater than unity. Thus, switching proceeds at the proper duty cycle.
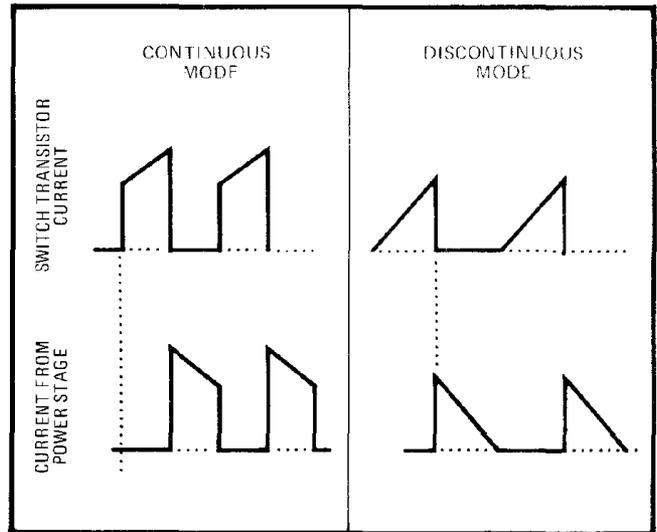
Greater stability

In addition to reducing cost, the technique enhances stability. It results in a 90° phase lag of output currents at high frequencies, instead of the 180° shift characteristic of direct duty-cycle control.

This primary current control simplifies feedback compensator design. Switching frequency for this design is around 20 kilohertz, and operating frequency is inversely proportional to the output power.

The blocking oscillator turns on when the energy stored in the magnetic field of transformer T_1 in Fig. 3 is approximately zero. It turns off at an energy level determined by the base drive voltage, the emitter resistor R_E (the current-sense resistor), and the primary inductor. Stored energy is released to the outputs via rectifiers D_4 through D_7. When the energy in T_1 has been drained to approximately zero, the switching cycle repeats.

Transistor Q_1 provides the desired adjustment of power flow. Q_1 and Q_2 are part of a current-limiting circuit that varies the duty cycle. The adjustment is governed by feedback from the ±5-v output through

amplifiers A_1 and A_2, so that a rise in voltage above +5 v produces a compensating reduction in output power and voltage.

Q_1 saturates and the transformer-inductor primary current starts its linear climb. The current increases until the rising voltage across R_E reduces the base current enough for Q_1 to operate in the linear class A mode. Q_1's constant-collector-current characteristics cause current limiting, so the rate of change of current in T_1 decreases; the voltages across the primary and base windings decrease; and Q_1 is driven off.

Now the energy stored in T_1's magnetic field must escape. The voltage on the windings reverses in polarity and increases in magnitude until a decay current path is found. The output rectifiers D_4 through D_7 conduct before input rectifier D_1 conducts. D_1 clamps the leakage inductance spike on the primary switch to the input source voltage.

Q_1 remains off until all energy is drained from T_1 and the output rectifier currents go back to zero. T_1 then rings back with the primary inductance and C_2's capacitance until Q_1 is again biased class A and turns on to repeat the cycle. Output power is a linear function of the current flowing in Q_1 as it comes out of saturation.

Lowering control costs

A major savings was achieved in the feedback control circuit by using +5-v–compatible integrated circuits. One is an adjustable zener diode, the TL 430, a 3-pin package used as an error amplifier. The conventional control circuit is a 16-pin integrated circuit. The other is MCT2E, an optical isolator that does double duty as an optoisolator and as part of the control circuit.

The TL 430 (A_1 in Fig. 3) meets the requirements for feedback control in switching regulators with on-chip functions like the high-gain operational amplifier and a voltage reference. In this application, the +5-v output provides both power and feedback signal. The TL 430's 2.75-v internal reference voltage is compared with the

output feedback signal at the wiper of potentiometer R_F.

The optoisolator, A_2, provides the necessary ac line isolation. The alternative would be placing a small transformer in the control loop. In the main power path, T_1 provides the required isolation.

A_1, A_2, and the current-limiting circuit control the amount of power flowing to the secondary. A_1 operates as an inverting transconductance amplifier. When the +5-v output voltage rises, A_1 drives the light-emitting diode in A_2 harder, thereby controlling the current through the latter's output transistor.

The current from A_2 prebiases the base of Q_2, reducing the current required to turn that transistor on. When Q_2 is biased on by a rise in voltage across resistor R_E, Q_1 turns off. The result is output regulation through adjustment of peak current and thereby output power.

The silicon controlled rectifier to the right of A_1 in Fig. 3 is part of a temperature-stabilized crowbar circuit that provides over-voltage protection. The circuit acts as a short when the 5.1-v zener diode, D_8, is overcome and the voltage at the SCR gate exceeds 0.8 V. The SCR selected assures the high rate of current change needed to discharge the output capacitances.

Accounting for filters

Each output filter in Fig. 3 is shown as a π-section filter with an inductor. This is not a general requirement, but it may be needed in some applications to minimize ripple on one or more outputs. Hence, the output stages are designed to accept a small air-core inductor and the additional capacitor required by a π-section filter.

The input filter is standardized. It has sufficient capacity to maintain maximum power output for 16 milliseconds after a line-power interruption.
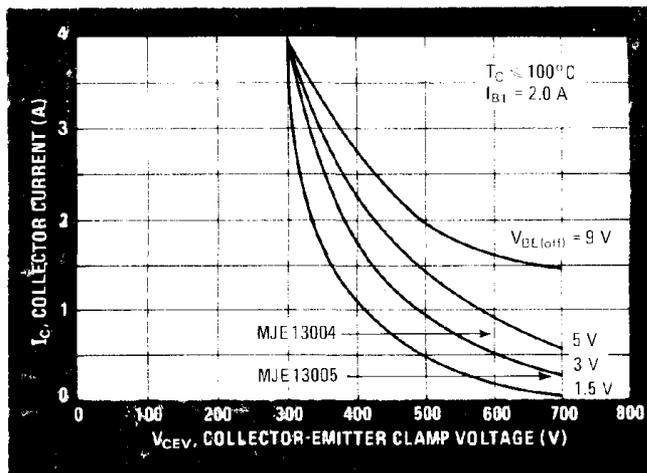
For energy storage, the flyback converter's transformer must be designed in as a multi-winding inductor rather than as a transformer. All power-inductor parameters become critical, and the winding geometry must be planned carefully to minimize leakage inductance.

In designing the transformer-inductor, some general criteria must be met. The magnetic circuit path is usually gapped in the one place where energy is stored. For the most part, the path operates over a flux density range from zero to maximum, typically 3,200 gauss for power ferrite at 100°C.

Core size determines the number of turns and the turns ratio can be derived from a steady-state design equation. For the Boschert OL25, an Electrical Plastics M1187-2 core is used. The turns for the primary and secondary coils are shown in Fig. 3.

The OL25 converter operates in a discontinuous flux mode. This provides a better cost tradeoff then the continuous mode since it reduces the volume of the magnetics components, and control-circuitry costs.

In the continuous mode, the ac flux is small compared to the dc component. Hence the current flowing in the switch transistor at turn-off is much the same as at turn-on, and the average power delivered to the load is greater than for discontinuous operation for a given peak transistor current (Fig. 5). However, the continuous mode requires relatively complex oscillator and control circuitry, shooting costs up.

6. Safe operation. Q_1 was chosen from Motorola's MJE 13004 family of power transistors because it exhibited a reverse bias that allows a proper margin of safety in the converter design. The 75-W npn transistors are designed for high-speed switching applications.

Selecting the power switching transistor (Q_1 in Fig. 3) is a matter of determining the required duty cycle. The limiting factor is the safe operating area of Q_1's reverse bias. Consider the steady-state design equation:

$$\frac{\text{volts} \times \text{seconds}}{\text{turns}}(\text{on}) = \frac{\text{volts} \times \text{seconds}}{\text{turns}}(\text{off})$$

for the on and off states of the ac-power-line input side of the converter. If just the primary winding of the transformer-inductor is considered, the number of primary turns drops out, and the voltage stress on Q_1 can be analyzed.

Using two equations

The voltage stress equals the source voltage plus the primary flyback voltage. If on time equals off time (a 50% duty cycle), flyback voltage equals source voltage. Thus, Q_1 must hold off twice the source voltage plus any spike voltage due to leakage inductance.

On the other hand, keeping the duty cycle as high as practical reduces the amount of current the transistor must pass. The switch current's dc component is:

$$I_{\text{Switch}} = \frac{\text{power output}}{V_{\text{in}} \times \eta \times \text{duty cycle}}$$

where η is efficiency. These two equations give voltage-current tradeoffs for various power levels and transistor specifications.

Since reliable operation is the most important consideration, peak voltages must be kept within the limits of presently available low-cost transistors. Compensating for the increases in input voltage by reducing the duty cycle accomplishes this.

One OL25 version for 110-v application operates at a 40%-50% duty cycle, and another version for 220-v application operates at a 20%-25% duty cycle. For both, the transistors are selected from the Motorola MJE13004 family of 75-w npn silicon power transistors. Designed for high-speed switching in inductive circuit applications, the MJE13004 has the reverse-bias safe operating area shown in Fig. 6. □