Inside SWITCHING POWER SUPPLIES

HARRY L. TRIETLEY

Learn the basics of switching regulators-the heart of switching power supplies.

THE ADVANTAGES OF REGULATED switching power supplies are too great to ignore. These versatile supplies are well known for their high efficiency, cool operation, small size, and the ability to work with a wider range of input voltages than their linear counterparts. Once limited to high-power or high-efficiency applications, they are now finding their way into low-power, low-cost consumer goods.

Because the control elements used in switching regulators are always either fully on or fully off, they have low power consumption and require little or no heat sinking. Small-size high-frequency transformers can be used, and, since regulation efficiency is not too affected by the input-to-output voltage differential, it's possible to handle two-to-one input variations, such as 115/230 volt operation.

Switching regulators do, however, have disadvantages. A primary drawback is their complexity, and therefore circuit cost. They also exhibit failure modes not seen in simple linear regulators, and can radiate substantial electromagnetic interference (EMI) if not properly designed. Fortunately, a number of IC's have been developed that not only include most of the complex circuitry, but also overcome common failure modes, which well look at later. Now we'll concentrate on basics.

Switching regulator **basics**

Let's begin by reviewing a linear (non-switching) regulator as shown in Fig. 1. Op-Amp IC1 compares feedback voltage V_{FB} to reference voltage V_{REF}. If V_{FB} is too high, Q1's base voltage decreases or, if it's too low, the base voltage increases, until V_{FB} equals V_{REF}. At equilibrium, Q1's emitter-collector voltage drop equals Vu,,,, $- V_{REG}$. The transistor power dissipation, W, equals $(V_{UNREG} - V_{REG}) \times I$.

A well-designed linear regulator can provide excellent regulation and transient response, low noise and ripple, and complete freedom from EMI. It does, however, waste power in the regulating transistor, especially at high load currents. Regulating widely-varying inputs is a problem because power dissipation increases as Vu,,,, goes up.

Now let's look at how pulse-width modulation (PWM) controls voltage. As shown in Fig. 2, Q1 is alternately turned on and off by the PWM control circuitry The output is R-C filtered to obtain a DC average. If Q1 is always off, the output voltage will be zero: if it is always on, the output will equal the input. The output voltage will be proportional to the duty cycle, which is the ratio of the "on" time to the total period.

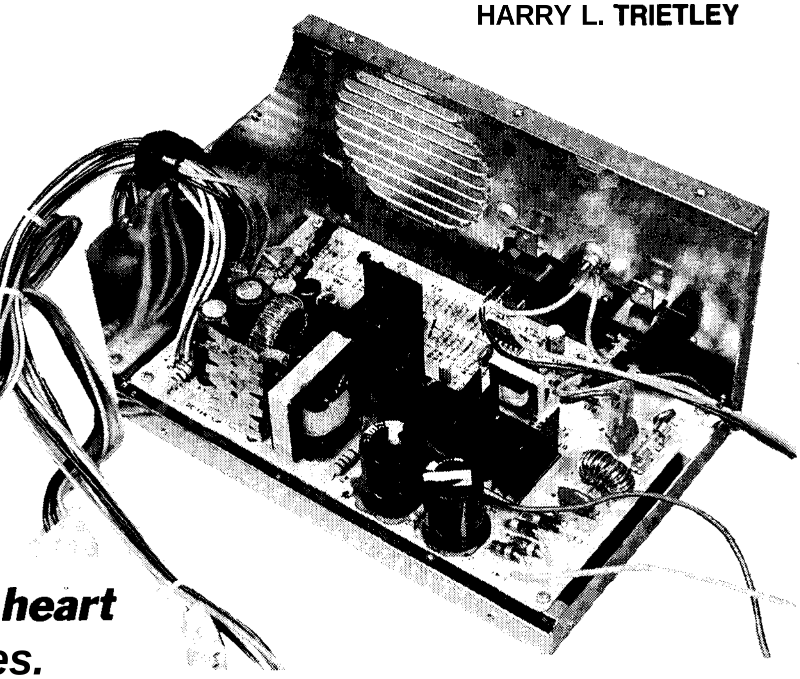
$$V_{OUT} = V_{IN} \times (T_{ON}/T_{TOTAL})$$
$$= V_{IN} \text{ x Duty Cycle}$$

If Q1 were ideal (no voltage drop in the on state) it would dissipate no power. Actual voltage drop varies depending on the transistor and the current level, but is usually less than or equal to one volt. (Power FET's respond better in high-current applications.) Power dissipation still occurs in filter resistor R1, reducing overall circuit efficiency

Practical circuits

To reduce resistive power losses, switching regulators use L-C, rather than R-C filters, as shown in Fig. 3. When Q1 is on, V_{UNREG} is applied to inductor L1 and D1 is reverse-biased. The inductive current supplies the load and also charges output capacitor C2.

When Q1 turns off, the inductive current continues, flowing through D1. The diode conducts until the inductor current reduces to zero, or until Q1 is again turned on, whichever occurs



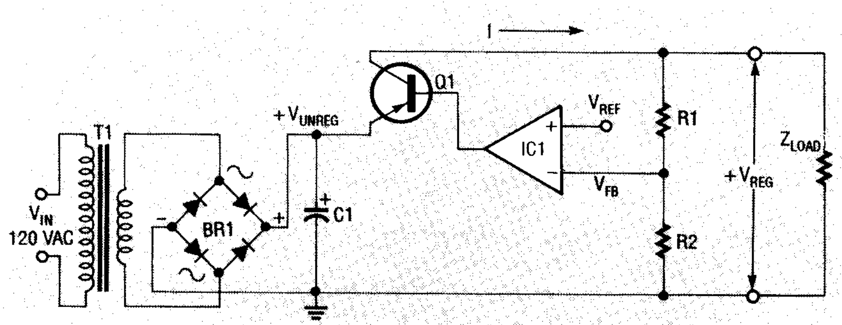


FIG. 1—A LINEAR REGULATOR CONTROLS conduction through a regulating transistor to maintain the required output.

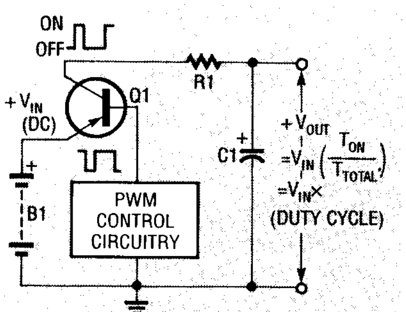


FIG. 2—PULSE WIDTH MODULATION (PWM) produces an output proportional to the duty cycle.

first. Inductor L1 smooths the on-off current from Q1, while C2 further evens out the load voltage. Complex switching-regulator control circuitry varies the duty cycle to keep the feedback voltage equal to the reference voltage. A circuit of that type is commonly called a "buck" converter because it bucks, or reduces, the input voltage.

The output current is greater than the input current because the inductive current continues while Q1 is turned off. For an ideal circuit, the regulator would be 100% efficient, meaning ($V_{REG} \times I_{OUT}$) equals ($V_{UNREG} \times I_{IN}$. In reality, however, circuit efficiency is typically about 80%.

Figure 4 shows a flyback, or "boost" converter. It operates much like a TV flyback, but with feedback added to control the output voltage. When Q1 is on, an inductive current builds up in L1. When Q1 turns off, the inductive current flows through diode D1, into C2 and the load. Since L1 is supplying current, its output side is positive and its voltage adds to, or "boosts" V_{UNREG}. The output voltage is, therefore, greater than the input voltage.

The longer Q1 is on, the higher the inductive current becomes. If Q1's off time is not long enough to reduce the current to zero, it will build even higher during the next on cycle, raising the output even higher. Conversely, if the on time becomes very short, the circuit will act as if the input is connected directly to the output, and the output will equal the input (no boost). Mathematically, ignoring losses,

$$V_{REG} = V_{UNREG}/(1 - \text{Duty Cycle}).$$

As with the buck converter, feedback is used to control the duty cycle for the desired output. One thing to keep in mind when using a boost converter is that its input current can be much higher than its output current.

The control circuitry

Let's get inside the control IC a little. To keep things simple, we'll leave out protective and problem-correcting circuitry and focus on the basic pulse-width control.

Figure 5 shows a typical circuit consisting of a clock (R-C oscillator), comparator, reference, error amplifier, flip-flop, output gate, and switching transistor. The error amplifier's output is proportional to the difference between the reference and feedback voltages. IC's vary greatly in detail, but generally use the principles discussed below.

At the beginning of each clock cycle a pulse resets the flip-flop, turning Q1 on, and the oscillator begins along the positive slope of the ramp. When the ramp exceeds the error signal, the comparator's output goes high, setting the flip-flop and turning Q1 off. The higher the error signal, the longer Q1 remains on.

The output voltage dividers in Figs. 3 and 4 are designed so that the feedback voltage, V_{FB}, equals the reference voltage, V_{REF}, when the output reaches the desired level. If the output voltage goes too high, the error voltage decreases, reducing the duty cy-

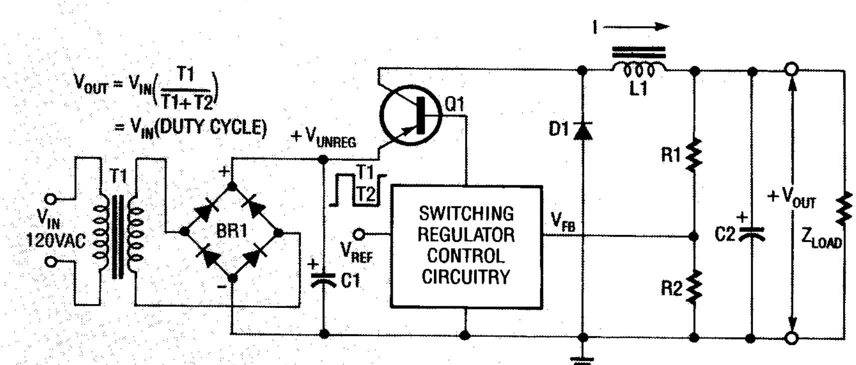


FIG. 3—MOST SWITCHING REGULATORS use L-C filtering to eliminate resistive power losses.

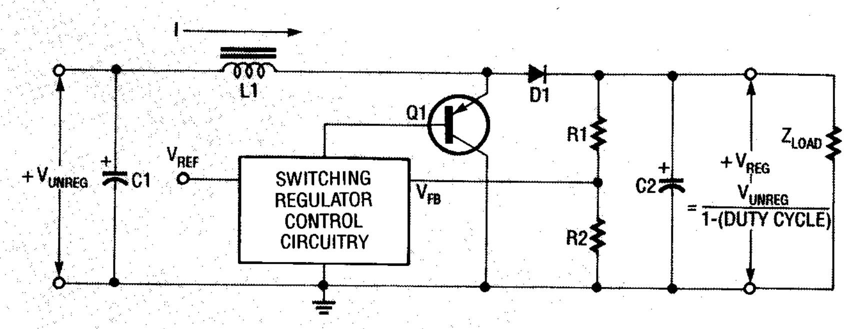


FIG. 4—A FLYBACK, OR "BOOST," converter produces an output voltage higher than its input.

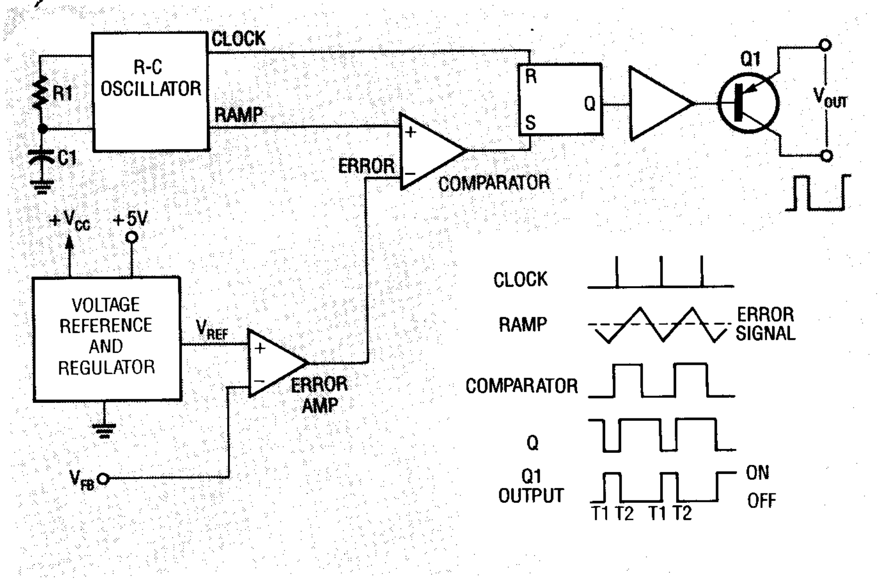


FIG. 5—THIS CIRCUIT SHOWS the heart of a pulse-width modulator, without its protective and problem-correcting circuitry.

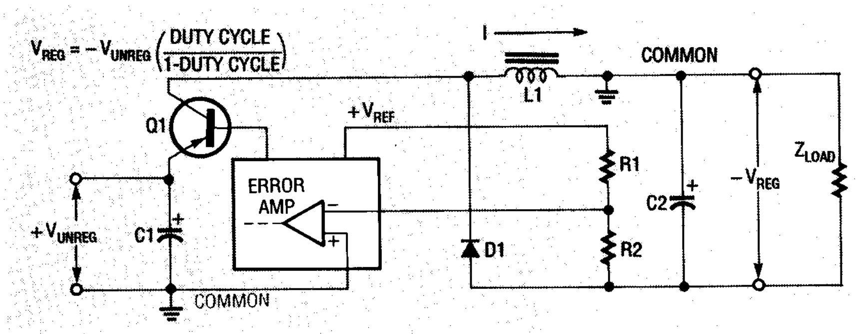


FIG. 6—A "BUCK-BOOST" CONVERTER produces a negative output from a positive input.

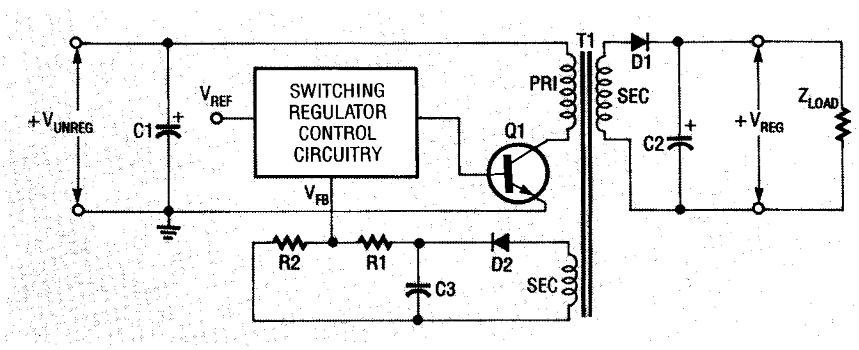


FIG. 7—IN A TRANSFORMER-COUPLED flyback converter a separate winding on the transformer allows output regulation while still maintaining input/output isolation.

cle and, therefore, the output. On the other hand, if the output voltage drops, the error voltage and duty cycle increases until the output returns to its design value. Now let's examine some more circuits and explore the differences between them.

A voltage inverter

Figure 6 shows a "buck-boost" converter which produces a negative output from a positive input. That circuit is similar to the circuit shown in Fig. 3. In Fig. 6, however, the right side of the inductor is connected to common and the feedback network is different. The name "buck-boost" comes from the fact that the output can be lower or higher than the input.

When Q1 turns on, the input voltage is applied to inductor L1, causing an increasing current. Unlike the "buck" regulator, that current does not flow through the load while Q1 is on. When Q1 turns off, the inductive current continues, flowing through D1 and charging output capacitor C2 with a negative voltage. The current continues until it reduces to zero, or until Q1 again turns on, whichever comes first. The longer the duty cycle, the higher the inductive current and, therefore, the higher the output voltage. The regulated output, V_{REG}, ignoring power losses, is

$-V_{UNREG} \times [(\text{Duty Cycle})/1-(\text{Duty Cycle})]$.

Transformer coupling

Up to this point, the circuits we've seen don't provide input-to-output isolation. They also suffer a second, less obvious shortcoming—an unbalanced current in the inductor, which produces a DC flux in the core, leading to saturation at lower power levels. Unbalanced operation of the power inductor requires larger, gapped cores to support the necessary magnetic fields. On the other hand, those circuits are simpler and use fewer components than the ones we're about to examine.

Primary-to-secondary isolation can be accomplished by using a transformer as shown in Fig. 3. High-power line-frequency transformers, though, are bulky and expensive. Since the pulse-width-modulation circuitry runs at high frequencies, it's often more efficient to transformer-couple and rectify its output pulses.

Figure 7 shows a simple transformer-coupled flyback converter. The control circuit is the same as in Fig. 4, but the inductor has been replaced with a flyback transformer. The on-time primary current builds up flux, which collapses when Q1 turns off. The collapsing field induces voltages in both secondaries, one of which produces the output while the other provides an isolated feedback voltage. Although this is a simple circuit, it still produces a net DC current in the transformer. The output voltage, V_{REG}, can be expressed as

$V_{UNREG} \times N \times [(\text{Duty cycle})/1-(\text{Duty Cycle})]$

where N equals the transformer turns ratio.

A circuit known as a "forward

converter" (Fig. 8) is better suited for high-power supplies. When Q1 turns on, the unregulated input is applied to the first winding and D1 is reversed-biased. The primary current begins to rise and a voltage is induced in the output winding. Output current flows through D2 and L1.

When Q1 turns off, the collapsing field induces reverse-polarity voltages in all three windings. Since Q1 is off and D3 is reverse-biased, their windings carry no current. Current flows through the middle winding, known as a "reset" winding, and D2 becomes forward-biased. During that time the inductive current in L1 flows through D3.

As long as D2 conducts, the reset winding is connected to the' input voltage. That condition continues until the current reduces to zero. There are two advantages of that circuit: the average primary current is zero, and the winding voltages are well-defined during the off portion of the cycle. A smaller core can be used, and high flyback voltages are not a problem. To maintain zero average current, the on time must never be longer than the off time, so the duty cycle is limited to 50%. The output voltage, V_{REG}, is

$$V_{UNREG} \times N \times \text{Duty Cycle.}$$

The output and input grounds are tied together in Fig. 8 for proper feedback voltage. To provide input-to-output isolation it is also necessary to isolate the feedback. We will discuss ways to do that in a future issue.

Finally, the push-pull circuit shown in Fig. 9 is similar to a DC-to-DC inverter, but with pulse-width modulation added. That circuit provides the best efficiency in high-power converters.

The primary winding's center tap is connected to V_{UNREG}. Transistors Q1 and Q2 are under the control of the switching regulator circuit. They are alternately pulsed on, connecting first one end of the primary and then the other to common. Raising the duty cycle increases the average applied voltage, and therefore the output voltage. Each transistor's duty cycle is limited to 50%. (we must not have both turned on at once). but since there are two, the overall duty cycle can approach 100%. Again, isolated

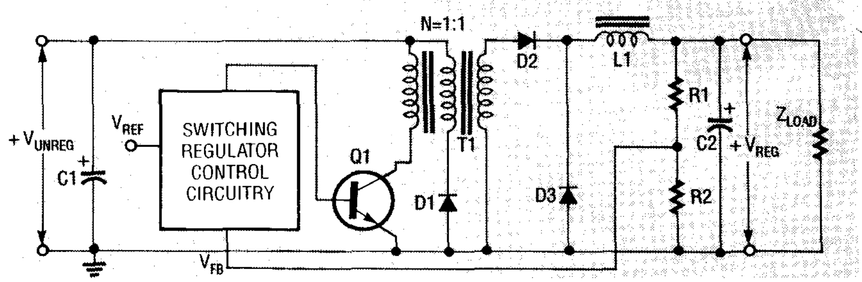


FIG. 8—IN A FORWARD CONVERTER direct-coupled feedback provides optimum regulation, but no input-output isolation.

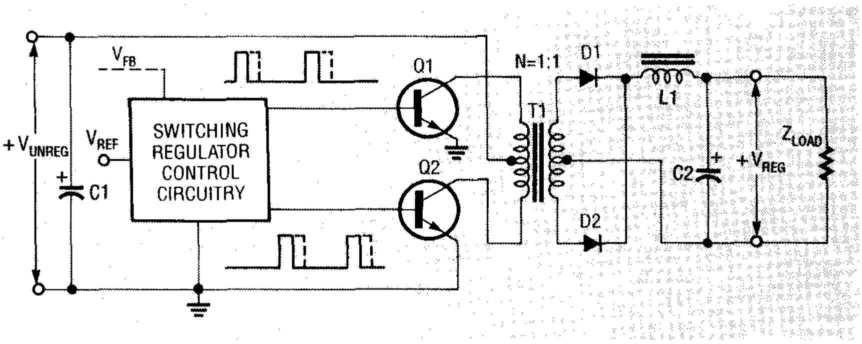


FIG. 4-A PUSH-PULL CONVERTER, similar to a DC-DC inverter but with pulse-width modulation, provides best efficiency for high-power supplies.

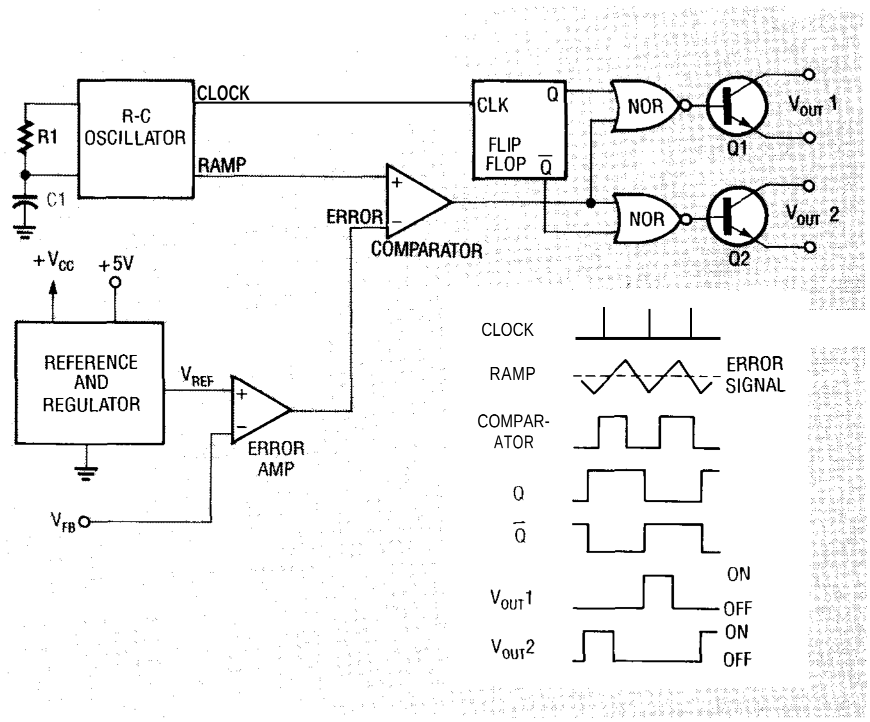


FIG. IO-ADDING A STEERING FLIP-FLOP and a pair of NOR gates produces a pulse-width modulator with push-pull output.

feedback is needed if input-to-output isolation is required. The output voltage, V_{REG}, is the same as the forward converter

$$V_{UNREG} \times N \times \text{Duty Cycle.}$$

Controlling the two transistors requires a change in the control circuitry, so let's examine the IC again. Figure 10 is similar to Fig. 5, with output-steering circuitry added. The clock pulses toggle the steering flip-flop. At the start of each cycle, when the comparator's output is low, the NOR gate whose Q input is low will turn on. The other remains off until the start of the next cycle toggles the flip-flop. Figure 10

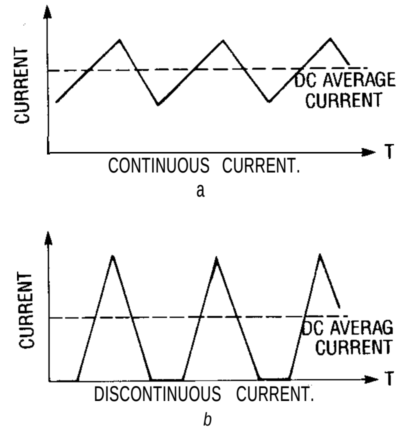


FIG. II-CONTINUOUS *(a)* and discontinuous (b) inductor current.

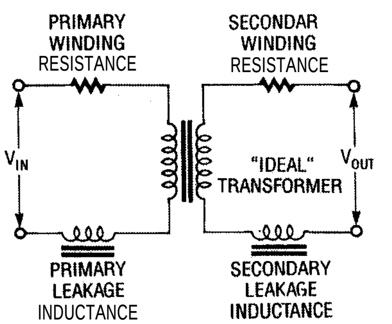


FIG. 12-A TRANSFORMER MODEL showing winding resistances and leakage inductances. Stray capacitances and core losses are not included.

shows the timing waveforms.

An IC of that type is very versatile, and can be used in all the circuits we have examined. Single-output control is implemented by simply paralleling Q1 and Q2. For forward converters, the 50% duty cycle limitation is easily provided by using only Q1 as the drive.

Which one should I use?

We have examined six circuits-three without transformers (buck, boost, and buck-boost) and three with (flyback, forward, and push-pull). Let's take some time now to compare the advantages and drawbacks of those techniques.

Transformer-coupled circuits are more flexible in stepping voltages up and down, and can provide input-to-output isolation. Negative outputs require only reversal of the rectifier diodes, and multiple secondaries can be used to provide multiple output voltages. The main drawback of transformer-coupled circuits is the cos. and the size of the transformer itself.

The choice among transformerless circuits is often simple. Use the buck circuit (Fig. 3) for voltage stepdown, where the output is lower than the input; the boost circuit (Fig. 4) for step-up; or the buck-boost circuit (Fig. 6) for polarity inversion. All three use the same number of components and have similar control requirements. One performance difference is worth noting: the buck converter tends to have lower output ripple because the inductor aids in filtering the output current.

When designing those circuits you must take into account the peak voltages and currents in the transistors and diodes to ensure that those components operate within their specified ratings. The buck converter operates with lower peak currents than the others, due to the filtering action of the inductor. Peak currents in the transistor and diode equal the output currents, while the peak voltages equal the input voltages.

In a boost converter, peak transistor and diode currents, I,,,, equal

$$I_{OUT} \times (V_{OUT}/V_{IN}).$$

The peak voltage kquals the output voltage.

In a buck-boost supply, the peak current, I,,,, equals

$$I_{OUT}/1\text{-Duty Cycle.}$$

The peak voltage equals the sum of the input and output voltages.

One drawback of the boost circuit should be mentioned. Because the input is directly connected to the output through the inductor and diode, it is not possible to use short-circuit limiting in the circuit.

The flyback converter (Fig. 7) retains the advantages (cost and simplicity) and drawbacks (hi&h peak currents, high ripple, and DC coil current) of a transformer-coupled circuit. It's the best choice when a simple, low-cost circuit is needed to regulate up to tens of watts. Peak switch current, I,,,, of a flyback converter is

$$I_{OUT} \times (N \times V_{IN} + V_{OUT})/V_{IN}$$

Forward and push-pull converters (Figs 8. and 9) are best for regulating higher power, whether isolation is needed or not. Both require extra windings, inductors and circuitry, but both provide the transformer with a balanced current. Also, both produce lower output-ripple current than the flyback. As a result, smaller transformers and filter components may be used. Input peak and output ripple currents are higher in the forward converter, because its duty cycle is limited to under 50%. Both are well-suited for use at tens to hundreds of watts, but for highest power (especially above 1000 watts) a push-pull converter should be chosen.

Discontinuous operation

For most efficient operation in any of the circuits we've discussed, the inductor current should flow continuously: otherwise, ripple currents will increase and regulation may suffer. That effect is most apparent in transformerless circuits. Those circuits depend on energy stored during the on cycle being transferred to the output when the transistor turns off. If the inductance is too low, all of its stored energy will be transferred to the output before the transistor turns back on.

Continuous operation results when the peak-to-peak ripple current in the inductor is less than twice the inductor's load. or DC average current: in other words, when the inductance is large enough that the negative excursion of its ripple never reaches zero. Figure 11-a shows continuous operation. while Fig. 1 l-b shows discontinuous operation. For example, in the buck converter of Fig. 3. continuous operation means that inductor current is always flowing into the load. Maintaining continuous operation in a switching regulator is usually a simple matter of choosing a large enough inductor.

Discontinuous operation normally occurs at low output loads. when the DC current is so low that the negative excursion cannot be kept above zero. Fortunately. discontinuous operation is not disastrous. only annoying, if it only happens under abnormally light loading. A decrease in regulation and increase in ripple are the usual result.

On the other hand, if the problem occurs under heavy loads due to poor design (improper inductor selection) the result may be core saturation, excessive current spikes and destruction of components such as the switching transistor.

Inductors and transformers

Let's finish our discussion by looking at inductors and transformers. The design of switching-regulator magnetics is a complex subject which we cannot cover completely in this article. We will, however, briefly discuss some of the more important concepts such as physical size, construction, ratings, and leakage inductance.

Our first consideration is size. The inductance of a choke or transformer must be large enough to keep ripple current within acceptable bounds and to maintain continuous operation. The core must not saturate at its highest current. Some of the design tradeoffs include size, power, filtering and transient response. Larger inductances and cores provide highest power and lowest ripple, but with slow recovery from transients.

Cores should be a ferrite material or powdered iron-laminations are not suitable for high-frequency operation. Toroidal cores minimize EM1 because they tend to be self-shielding. Air gaps usually are needed to prevent saturation with unbalanced DC currents. The gap reduces the core's permeability, requiring larger structures to achieve the required inductance. When buying an inductor or transformer make sure it is rated for the frequencies and DC currents you will be applying to it. The affect of saturation could be the destruction of switching transistors, control IC's or other components in the circuit.

An approximate inductance value can be calculated from basic inductor theory. Inductor current increases linearly with time when a DC voltage is applied

$$\Delta I = E \times T/L$$

where AI is the change in current in amps, E is the applied voltage in volts, T is time in seconds, and L is inductance in henrys.

If your circuit operates at a frequency in hertz equal to 1/T, the maximum voltage across the inductor is E and you want to design for a peak-to-peak ripple current of AI, the inductor value can be found by

$$L = E/2(\Delta I)f$$

It's best to start with a little extra inductance, then optimize it experimentally Inductor values in the medium to high microhenries are common.

Switching regulators operate at high frequencies and fast risetimes, and switching transients can produce peak voltages higher than the values given earlier. Transformers with switched primary currents are the main source of that problem.

A major source of primary-side spikes is leakage inductance. Figure 12 shows a transformer model including winding resistances and leakage inductances. (Winding capacitances are not shown.) In an ideal transformer there would be perfect magnetic coupling between the primary and secondary. A voltage spike could not appear across the primary unless a proportional spike was seen on the secondary If there was a load across the secondary, especially when a capacitor is used, spikes would not occur.

In reality, a small portion of the flux produced by the primary is not coupled to the secondary. Electrically, that means that a small part of the primary's inductance is not coupled to the secondary, and vice-versa. Transformer leakage inductance is represented in Fig. 12. Switched primary currents produce spikes in the leakage inductance.

Leakage inductance can be minimized, but not completely eliminated, by proper transformer design. The best approach is a bifilar winding, where the primary and secondary are wound together, their wires intermixed in the same coil. That may not be possible in transformers requiring high primary-to-secondary breakdown voltages. It's sometimes necessary to add Zener diodes and/or small capacitors, across the primary to protect the switching transistors and diodes.

In the second and final part of this article, we'll look at some more protective and safeguard circuitry provided in switching regulator IC's. We'll also examine some IC families with which you should be familiar. **R-E**

Inside SWITCHING POWER SUPPLIES

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We'll take an in-depth look at two switching regulator IC families, with some applications, and guide you through basic troubleshooting techniques.

IN OUR LAST EDITION, WE EXAMINED the basics of switching regulator power supplies. Now we'll dive into some real-world applications. Well examine the 3524/5/7 and 3842 IC families in detail, summarize others, and show some typical circuits. In the process we'll study how to select components for those circuits and learn more about how switching regulator IC's are protected against such problems as startup current surges, undervoltage, and overload. We'll finish up with some basic troubleshooting hints.

Let's first start off with an explanation of the standard nomenclature used in naming the IC's we will discuss in this article. The first digit "1" indicates full military temperature range of -55 to 150°C "2" indicates an industrial temperature range of -25 to 85°C, and "3" is a commercial temperature range of 0 to 70°C. Hereafter, we will refer only to the commercial version IC's, with prefix "3." Keep in mind that all those IC's are also available in military and industrial versions. A suffix of "A," "B," or "C" indicates an enhanced version of the IC, which we will discuss in more detail later in this article.

Manufacturers may use many different prefixes, some of which include

SG-Signetics, SGS-Thomson, Motorola, and Linear Technology.
CS—Cherry Semiconductor.
XR-Exar.
CA-GE-RCA.
IC—IPS.
LTSG-Linear Technology.
LM-National Semiconductor.
UC-Unitrode, Motorola, Linear Technology, and Signetics.
UD-SGS-Thomson.
IP—IPS.
LAS-Lambda.

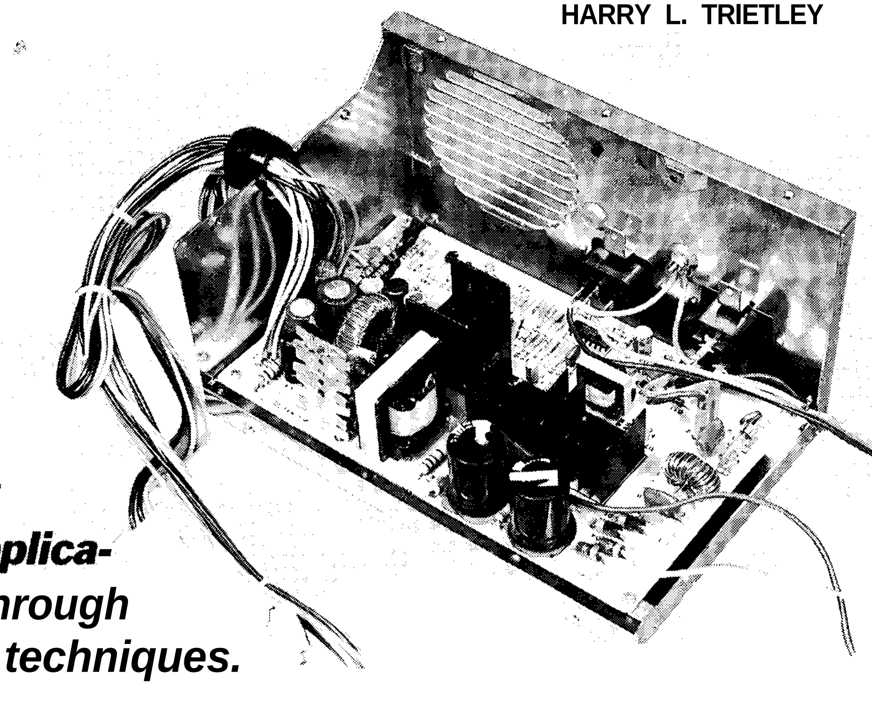
The SG3524/5/6/7 IC

Figure 1 shows the internal circuit of the switching regulator IC 563524. In that circuit, the oscillator produces both ramp and pulse outputs. Ignoring the current limit (CL) and shutdown circuits for the moment, the comparator's output goes high when the ramp exceeds the output of the error amplifier. The NOR gates then go low, turning the output transistors off.

Each NOR gate can be high only when its three inputs are low. The oscillator output toggles the flip-flop, enabling one gate, and then the other to respond to the comparator. That action gates one transistor on at a time, providing push-pull operation. The selected transistor turns on at the start of each cycle, and turns off as soon as the ramp exceeds the error signal. At the end of each cycle, the oscillator pulse momentarily forces both gates low, protecting against the possibility of both transistors being on at the same time.

The current-limit amplifier protects against current overloads. Its output is an open-collector type-open-circuit when high, pull-down to ground when low. The current-limit amplifier and the shutdown transistor can be used to force the comparator output high, shutting down both transistors.

Figure 2 shows the SG3524 in a simple DC-DC converter. The oscillator frequency of about 60 kHz is set by R5 and C2. (The flip-flop divides the push-pull output frequency to 30 kHz.) The current-limit amplifier goes low when its input exceeds 0.2 volts, limiting R11's current to 2 amps in case of overload or transformer



saturation. Transistors Q1 and Q2 are used for switching transformer current. (The on-chip transistors are rated at only 100 mA.) Supply pulses produced by the circuit are filtered by C4.

The output of the error amplifier is proportional to the difference between the reference input (pin 2) and the feedback (pin 1). If the output increases, the error voltage drops. The ramp then reaches the error voltage more quickly and the transistors turn off sooner, until the output is reduced back to 5 volts. Since the feedback voltage and ground are directly connected, input-to-output isolation is not provided.

Resistors R6 and R7 limit the current through the internal drive transistors, which are used to switch Q1 and Q2. Frequency compensation for closed-loop stability is provided by R10 and C3. Transistors Q1 and Q2 should be high-speed switching power transistors rated at least 5 amps and 60 volts. Shottky or fast-recovery diodes should be

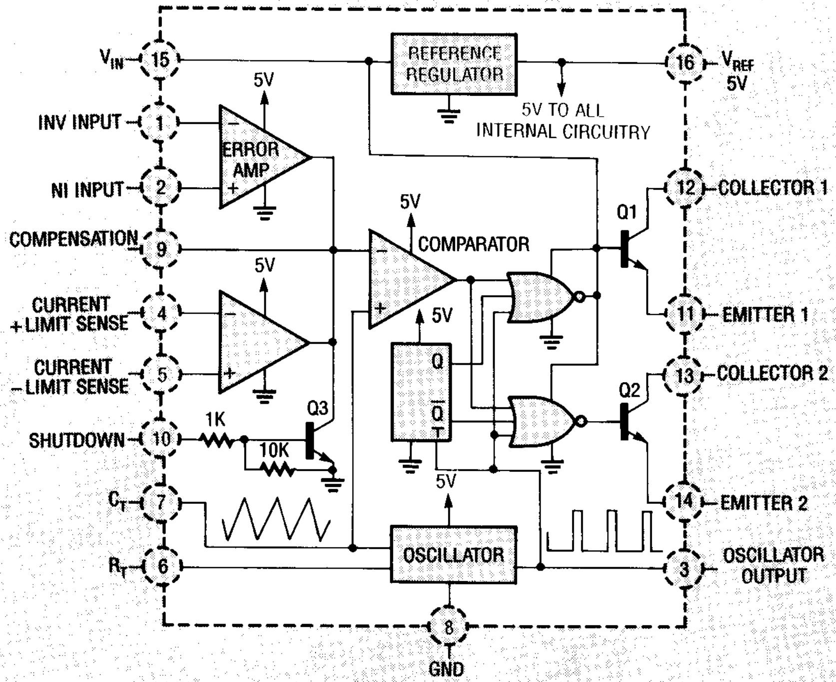


FIG. 1—INTERNAL BLOCK DIAGRAM of the SG3524 switching regulator IC.

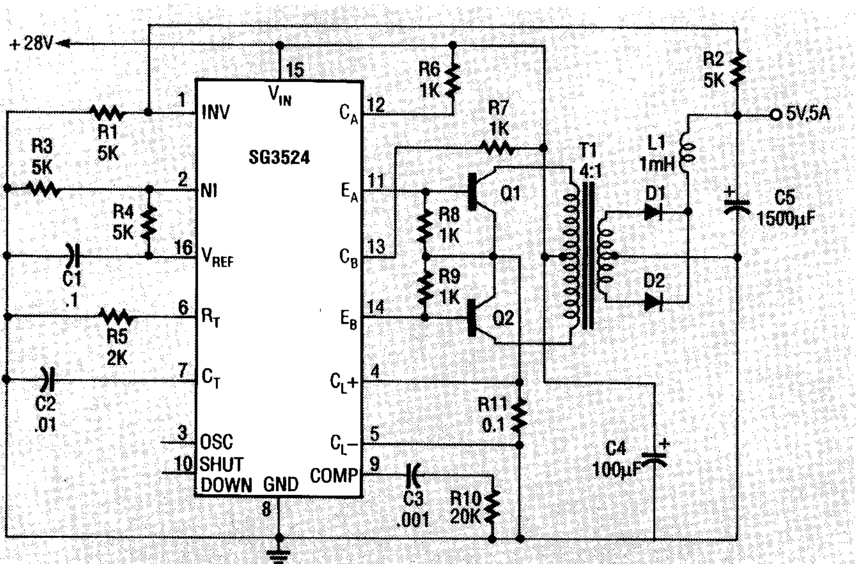


FIG. 2—THE SG3524, TWO SWITCHING transistors, a transformer and a few components create a complete 5-volt push-pull switching regulator.

used for D1 and D2. Because the output is balanced, the transformer core does not need to be gapped, a small ferrite core will do.

At high frequencies, the equivalent series resistance (ESR) of filter capacitor C5 is higher than its capacitive impedance. Low series-resistance electrolytics should be used, preferably capacitors designed specifically for switching supplies.

The enhanced SG3524A

Figure 3 shows the enhanced version SG3524A, which is pin-compatible and interchangeable with the non-A version. The enhanced version adds an undervoltage lockout circuit which disables the regulator until its input rises above 8 volts. That holds current drain to standby levels during turn-on, guarding against problems during startup, surges, and brownouts. A pulse-width modulator latch is also added, which eliminates multiple pulsing in noisy environments. Set by the comparator and reset by the clock pulse, it can switch only once per comparison cycle.

Further protection is provided by thermal protection circuitry (not shown). Performance specifications also are improved-the 5-volt reference is trimmed more closely (± 1%) and the error amplifiers output can swing up to the 5-volt rail.

Let's look at one more member of this family, and an application. Figure 4 shows the workings of the SG3525A/7A. The 3525A and 3527A differ only in their output logic; the 3525A is low when off, while the 3527A is high when off. (The pinouts of the 3525A/7A do not match those of the 3524 IC series.)

Operation is similar to the 3524, but with added features. The oscillator has a sync input, making it easy to lock the frequencies of several supplies, eliminating problems with beat frequencies in multiple-supply boards or systems. The shutdown circuit (also included in the 3524A) and soft-start feature simplify the design of protective circuitry, as will be seen in the next application. The totem pole (push-pull) outputs, rated at maximum 500 mA, provide fast,

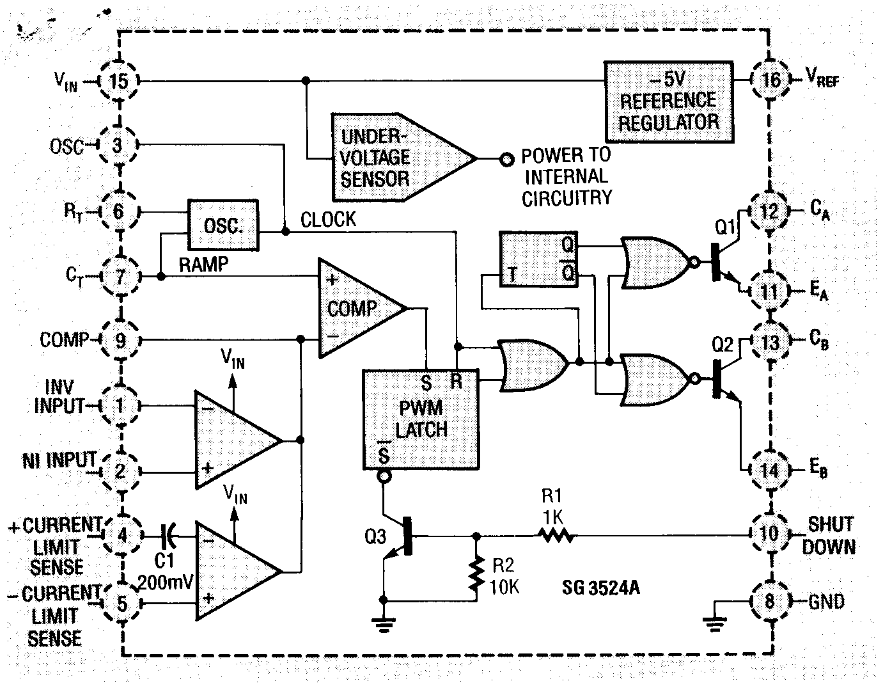


FIG. 3—THE SG3524A IMPROVES the basic device by adding undervoltage lockout, a pulse-width modulator (PWM) latch, thermal protection, and improved reference accuracy.

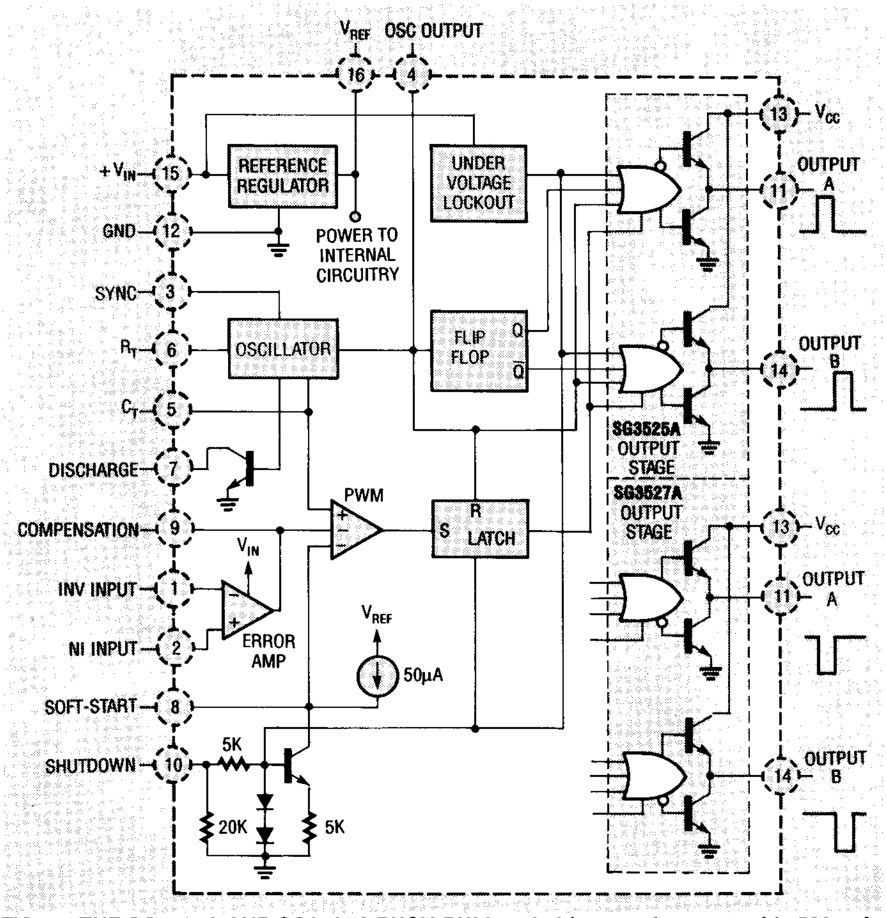


FIG. 4—THE SG3525A AND SG3527A PUSH-PULL switching regulators provide 500-mA, totem-pole outputs and oscillator synchronization.

solid switching for high and low transitions. The 3524's separate current-limit amplifier has been omitted.

Figure 5 shows a 15-watt DC-DC converter. The 200-kHz frequency (100 kHz final output) is set by R2–C2. The internal discharge transistor (pin 7) allows control of the discharge time at the end of each ramp. That provides an ensured stoptime between output pulses so that, even with switching delays, both transistors cannot be on at once. A 47-nanosecond time constant is provided by R16-C2.

The 5-volt reference (pin 16) is connected to the noninverting input (pin 2) by current limiting resistor R3, while C9 provides high-frequency bypassing. Negative feedback voltage is divided by Rl-R4, dividing the 6-volt output down to 5 volts. The basic theory of operation is similar to that of Fig. 2; the ramp is compared to the error signal to control the on-off switching of the outputs. On each cycle, the internal flip-flop selects either output A or B. The selected output is switched high at the start of each ramp and reset to low by the latch when the ramp voltage exceeds the error amp's output. As in Fig. 2, the direct feedback connection means there is no input to output isolation.

Compensation for closed-loop stability is provided by R6, R7 and C4. Switching spike currents are limited by R10, R11, and R12 in the output stages. Components C5 and R17 act as a "snubber," limiting switching transients from the primary.

When input power is first applied, Q1 will be off and soft-start capacitor C3 will be discharged. As C3 is charged from the internal 50-μA current source, its voltage will rise, gradually increasing the voltage to the pulse-width modulator (PWM) control's bottom input. That gradually increases the amount of time per cycle that the output is turned on, providing a "soft" rise of the output voltage, which allows the filter capacitors to charge slowly, reducing startup current surges.

If R9's current exceeds about 3 amps (0.7-volt drop), Q1 will turn on, energizing the shutdown circuit which pulls pin 8 low and discharges C3. As the current drops below 3 amps, Q1 turns off, C7 discharges, the shutdown input goes low and the soft-start capacitor provides a "soft" recovery for the power supply.

Power transformer T1 is wound on an EE25 ferrite core (0.25-inch center leg). It feeds a

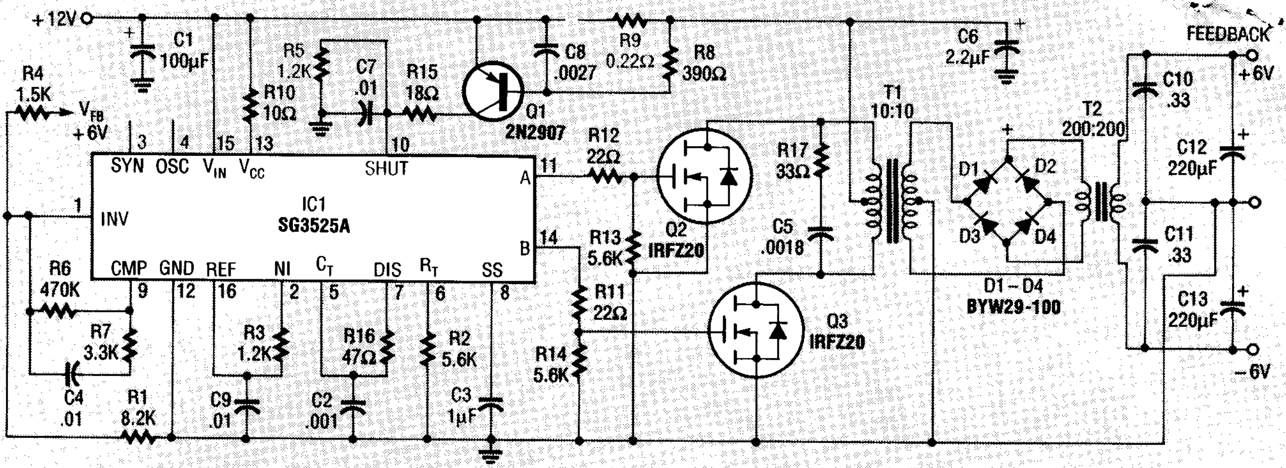


FIG. 5—A PUSH-PULL SWITCHING REGULATOR produces plus and minus 6 volts at 15 watts.

conventional full-wave bridge, providing + and – outputs. Coupled inductor T2, consisting of two coils wound on a cylindrical ferrite core, and the output capacitors filter the output to 50 millivolts peak-to-peak. Transistors Q2 and Q3 are 50-volt, 5-amp, N-channel power MOSFET's. Fast-recovery diodes must be used in the rectifier due to the high frequency: Dl-D4 are 100-volt, 8-amp diodes with 35-nanosecond recovery.

Current-mode regulators

We now turn to a different class of switching regulators-current mode. Although the basic operating theory remains the same (pulse-width modulation), current-mode switching regulators differ in that the internal ramp is eliminated. In its place, the ramp-like increase in the transformer's inductive current is used for control.

Figure 6 shows the basics of a current-mode comparator. The pulse from an R-C clock sets the flip-flop, producing a high output. FET Q1 turns on and transformer current begins to flow. As the inductive current ramps upward, the feedback from current-sensing resistor R2 increases. Eventually the feedback voltage equals the error amplifiers output, at which point the comparator resets the flip-flop, Q1 then turns off until the next clock pulse.

As with previous regulators, the feedback voltage, V_{FB}, represents the filtered output. If the feedback becomes lower or high-

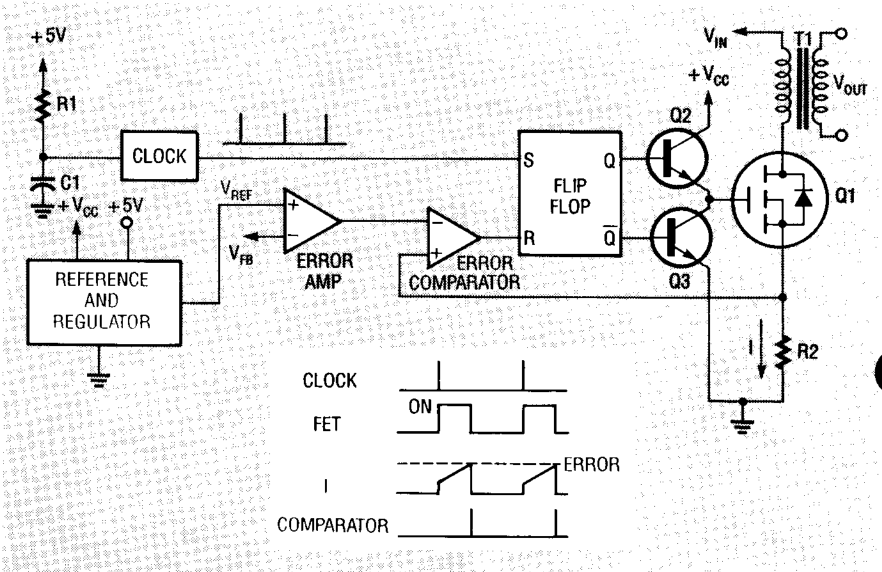


FIG.6—A CURRENT-MODE comparator uses the current feedback signal as the ramp, providing pulse-by-pulse current limiting.

er than the reference voltage, the error signal will increase or decrease accordingly, increasing or decreasing the on time until the proper voltage is restored.

Current-mode regulation offers two major advantages; pulse-by-pulse current limiting, and feedforward line regulation. Notice that the circuit in Fig. 6 contains no current-sensing comparator. Instead, each current pulse ends as soon as it exceeds the level set by the error amplifier. No matter what the cause of overload, whether transformer saturation, an output short, or input overvoltage, the circuit will limit current instantly Pulse-by-pulse limiting also eliminates the need for a separate soft-start circuit.

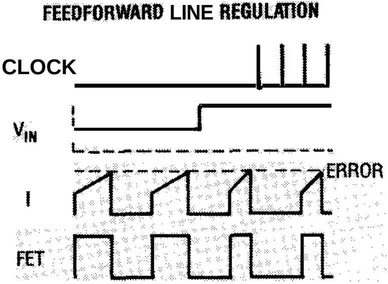


FIG. 7—FEEDFORWARD compensation of input variations is achieved when the ramp rate of the transformer's primary current increases as the input voltage increases.

Feedforward line regulation is illustrated by the waveforms shown in Fig. 7. With a fixed load, the input voltage suddenly increases. On the very next pulse, the inductive current, I, ramps more quickly due to the increased transformer voltage. Since the feedback and the error signal have not changed, the limit is reached more quickly and the pulse width becomes shorter. Changes in line voltage are, therefore, compensated before they have a chance to affect the output.

UC3842/3/4/5

Figure 8 shows the block diagram of current-mode PWM controller IC UC3842. Compared with the circuit in Fig. 6, the UC3842 adds an undervoltage lockout and an output NOR gate. The undervoltage lockout, with hysteresis, disables the output pulses until V_{CC} rises above 16 volts. Once started, it will not drop out unless V_{CC} goes below 10 volts, a feature which prevents constant toggling between "operate" and "lockout." When disabled, the output (pin 6) goes to a high-impedance state. A "bleeder" resistor should be connected from pin 6 to ground to prevent leakage current from turning the switching FET on.

The output NOR gate implements lockout, but also serves another protective function. When the oscillator pulse is high, the NOR output will be low, the OR output high, and pin 6 low. The output cannot go high until the clock goes low. The clock is set up so that timing capacitor Cl charges through R1, and discharges through the constant current sink. By choosing a larger capacitor and smaller resistor, the charging time (clock low) can be decreased and the discharge time (clock high) increased. That allows you to establish the maximum on time, or duty cycle, which is especially important in circuits where duty cycles higher than 50% can lead to transformer core saturation.

The D2–D4-R1-R2 network between the error amplifier and the current-sensing comparator reduces the error signal so that excessive power is not lost in the current-sensing resistor. The one-volt Zener diode clamps the error signal so the maximum turn-off level will never exceed one volt.

UC3843 is similar to the 3842 but has a lower lockout voltage.

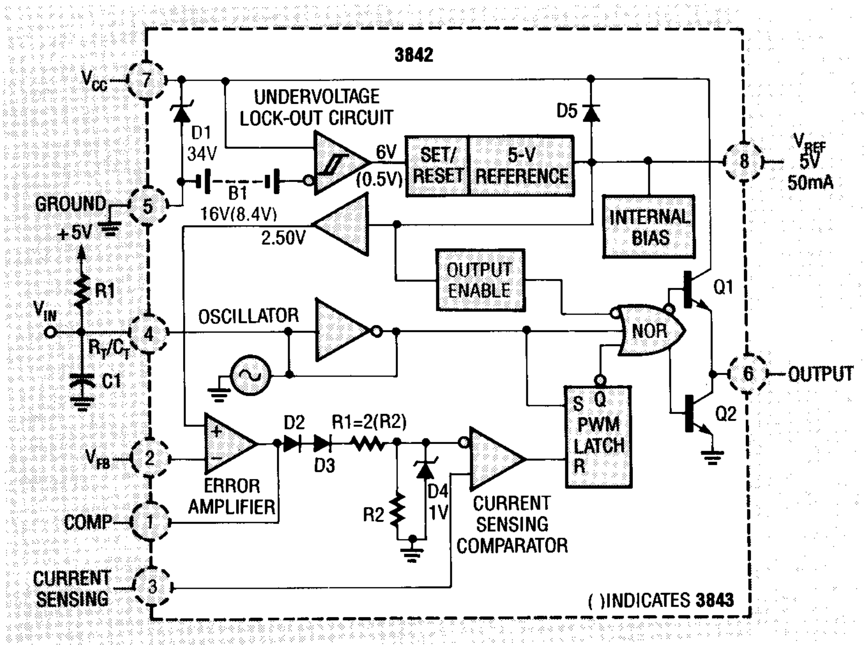


FIG. 8—INTERNAL BLOCK diagram of the UC3842 current-mode switching regulator IC. The UC3843 IC is similar but operates with a lower undervoltage lockout.

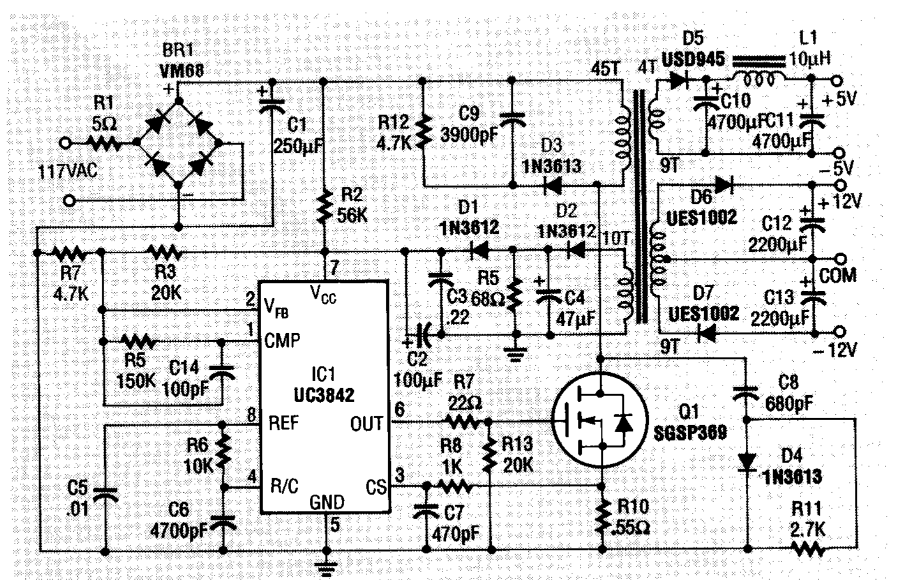


FIG. 9—THIS OFF-LINE CURRENT-MODE regulator produces isolated +5 volts and ±12 volts from 117-volt line power.

Intended for use at lower voltages, it operates at 8.4 volts, and drops out at 7.9 volts. UC3844 and UC3845 (not shown) have one added feature; a flip-flop which disables the output on alternate clock cycles. That guarantees the duty cycle will always be less than 50% for circuits where that is critical.

An off-line flyback converter

Figure 9 shows an SGS-Thomson UC3842 IC in an "off-line" flyback regulator. The circuit provides + 5 volts at 4 amps and ± 12 volts at 300 mA, and can deliver 27 watts.

The term "off-line" means that the regulator is on the primary side of the transformer and operates directly "off the line." The primary advantage of such a circuit is that large amounts of power can be coupled through a small, high-frequency transformer. Line operation requires high-voltage transistors and diodes, and prevents direct coupling between the output and the feedback circuit.

The line voltage is rectified and filtered by BR1 and Cl. Initial startup current to the IC is provided by R1. The UC3842's under-voltage lockout circuitry prevents startup until the voltage on C2 reaches 16 volts. The 50-kHz operating frequency is set by R6-C6, with a maximum duty cycle of about 95%. The internal 5-volt supply is filtered by C5 to eliminate switching spikes. Current-mode feedback is provided by R10, while C14 and R5 are used for frequency compensation.

Once the circuit has started, voltage feedback comes from the 10-turn control winding. The voltage at pin 2 is compared to the internal 2.5-volt reference. The voltage difference increases or decreases the duty cycle until the voltage at pin 7 equals 13.1 volts. Allowing for diode voltage drops, that corresponds to a peak voltage of about 14.6 volts on the control winding. The control-to-secondary turns ratios are chosen to produce 5- and 12-volt DC outputs. Notice that control is from the control winding's voltage, the outputs are only indirectly regulated. Power losses due to currents in the windings, diodes and inductor will affect the outputs. Five-volt regulation is 10% accurate, while the ±12-volt regulator has 5% accuracy

Transistor Q1 is a 500-volt, 5-amp power MOSFET The diodes are fast-recovery diodes. A "snubber" network is formed by D3-C9-R12 to hold turn-off spikes below Q1's breakdown voltage. Snubber D4-C8-R11 slows the turn-off rise time until Q1's current has had a chance to decay,

Transformer design is important: the air gap must be large enough to prevent core saturation but small enough to maintain the required inductance. (Note that an air gap is not needed in balanced push-pull circuits.) In the Fig. 9 circuit, an EC35 ferrite core is used (3/8-inch dia. center leg, Ferroxcube EC35-3C8) with a 0.5 mm gap in the center leg.

The primary winding consists of 45 turns of 26 AWG wire. The 12-volt windings are each 9 turns of 30 AWG wire, wound together (bifilar). The 5-volt secondary is only 4 turns, but instead of using a heavier gauge wire, four bifilar, 4-turn windings of 26 AWG wire are used, with their ends connected in parallel. The control (feedback) winding consists of two bifilar, parallel 10-turn 30 AWG windings. Now let's take a look at how an optoisolator can be used in a switching regulator.

Optocoupled feedback

Optocouplers provide a convenient way of coupling isolated feedback. Figure 10 shows a circuit in which the 5-volt secondary of a switching regulator is controlled. If the output goes above 5 volts, the inverting input decreases below 2.5 volts and the optocoupler's LED current decreases. That decreases the couplers output transistor current, increasing V_{FB} until the isolated output returns to 5 volts.

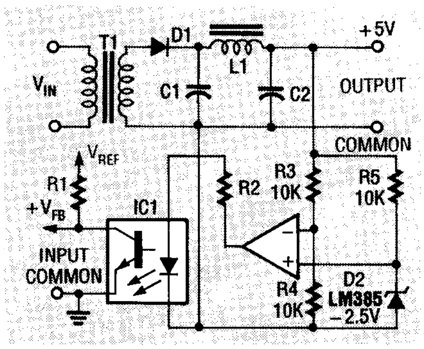


FIG. 10—OPTOCOUPLER FEEDBACK allows precise control of an isolated output.

A wide selection of IC's

Once a new IC technology is established, the offerings multiply as designs advance and the market expands. Switching regulators are no exception. Voltage mode, current mode, single-ended and push-pull IC's cover a wide variety of power levels and user-specific applications.

Table 1 summarizes some of the many IC families available. Most of the devices shown can be multiple-sourced. The part number prefixes vary from manufacturer to manufacturer, and many offer additional, proprietary devices.

It's not possible to fully describe all devices in an abbreviated table, but the listing should help direct you to data sheets for IC's to meet your needs. The 8-pin devices tend to be simpler to apply, while the 16-pin and larger IC's generally offer more complicated protective and "housekeeping" features.

The 3524/5/7 and 3842-7 families have been fully covered in this article. The 4191-3 family, with its low operating voltage and 200-µA current drain, is ideal for battery and micropower applications. Companion micropower device 4391 provides regulated negative outputs from positive supplies. LT1070 is the only IC in the listing housed in a power IC package.

Troubleshooting hints

When troubleshooting switching regulators, always begin with the obvious. Check for input power and output shorts, broken wires, defective connectors, solder bridges, defective solder joints, bad copper traces, scorched components, and so on. It's surprising how often a good visual inspection can uncover a problem.

Make sure you have a data sheet, pinouts of the control IC, and a circuit schematic, preferably with voltages and waveforms. There is such a wide variety of IC's and operating modes that it's difficult to troubleshoot on an intuitive basis. Figure 11 shows a "generic" block diagram, which may help you to think through the circuit function-by-function.

When breadboarding temporary components, remember that switching regulators produce fast, high-current pulses. Conductor size and lead dress are important. The input filter capacitor should be close to the IC, not a foot away. If the main source of power is at a distance, add a several hundred microfarad input bypass capacitor next to the IC.

Even though you may understand the operation of switching regulators, troubleshooting them can be difficult. The IC and its circuitry perform many functions, and the failure of one can cause improper operation of the rest. For example, failure of the feedback circuit may lead to overvoltage, overcurrent, and shutdown by one of the protective features. Is the circuit dead, unstable or out of regulation? That alone may often narrow the search to one particular part of the circuit.

The following hints may help you pinpoint the problem to a specific area of the circuit. After the visual inspection, check the output for shorts or overloads and check the input source, rectifier, filter, and transformer.