IC timers control dc-dc converters

by P. R. K. Chetty
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An integrated-circuit timer such as the MC1455 can be used as the control element in a simple dc-to-dc converter regulator. Shown below are a current step-up converter regulator and a polarity-reversing voltage step-up converter regulator. Both are regulated to within 0.5% for load currents of 300 milliamperes, and have a ripple of less than 5 mA.

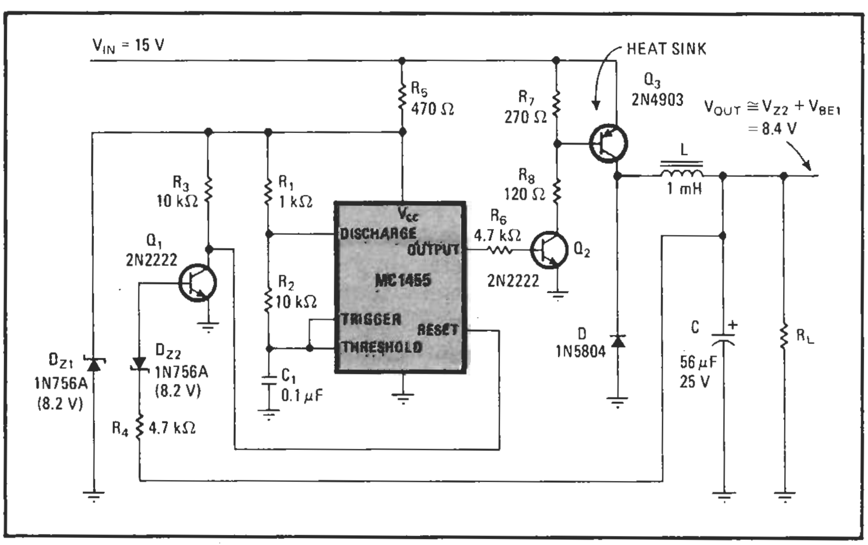
In these circuits the MC1455 operates as an astable multivibrator, turning the pass transistor on and off to keep the output-filter capacitor charged to the desired output voltage. Overvoltage is prevented by a feedback arrangement that turns off the multivibrator when the capacitor voltage reaches a predetermined level.

The astable-mode connection of the timer causes the voltage across capacitor C_1 to oscillate between $V_{CC}/3$ and $2\ V_{CC}/3$ at a frequency of approximately $1.44/(R_1+R_2)C_1$—about 1.3 kilohertz. The maximum operating voltage of the timer is 16 volts, but here its V_{CC} is clamped at 8.2 V by zener diode D_{Z1}. The input voltage therefore can have any value within the ratings of the pass transistor and the filter capacitor.

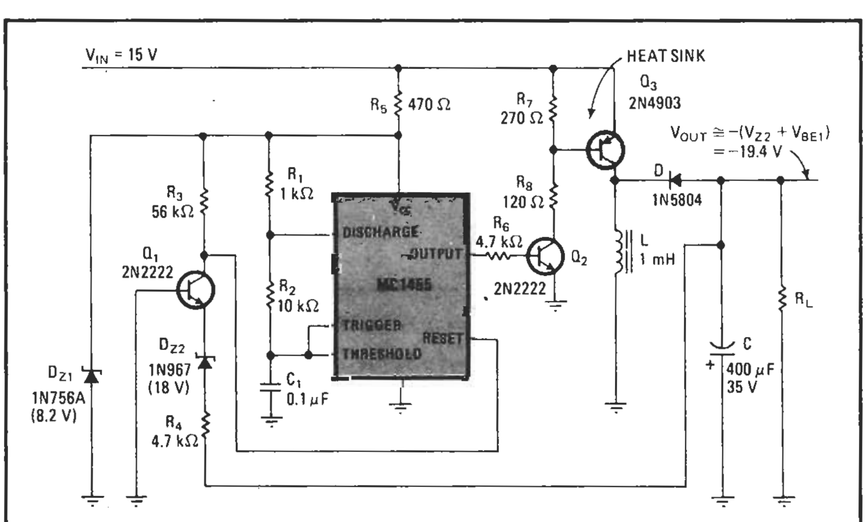
Figure 1 shows the current step-up converter regulator. When the output of the control timer is high, transistor Q_2 is turned on and therefore pass transistor Q_3 is turned on. Collector current from Q_3 flows through inductor L into the load and the filter capacitor. When the output of the timer goes low, the transistors turn off. Diode D commutates the current flow flowing through the inductor when Q_3 switches off. If there were no feedback circuit, the output voltage would depend upon the input voltage and the duty cycle.

The feedback circuit consists of R_4, zener diode D_{Z2}, transistor Q_1, and R_3, Whenever the output voltage exceeds $(V_{Z2}+V_{BE1})$, Q_1 turns on and drives the reset terminal of the 1455 low. The transistors Q_2 and Q_3 therefore stay off, allowing the output voltage to decrease. Thus the output voltage V_{out} is maintained approximately equal to $(V_{Z1}+V_{BE1})$.

The performance of the circuit in Fig. 1 is as follows:



1. Converted and regulated. Dc-to-dc converter includes IC timer for regulation. The MC1455, connected as free-running multivibrator, switches Q_3 on and off. If output gets too high, feedback circuit drives timer reset low to hold switch off. Regulation is less than 0.5% at 300 mA, and ripple is less than 5 mA. Output voltage is lower than input voltage, so current can be stepped up.



2. Polarity reversed. Positions of inductor, commutating diode, and feedback elements are changed here for negative output voltage. This circuit arrangement can step up magnitude of either voltage or current; components chosen here provide voltage step-up. Regulation is same as before.

STOP pulse. Meanwhile the cycle counter passes to states 2 through 10.

The next clock puts the cycle counter into state 11, but the gate detects this and clears the BUSY flip-flop. This in turn raises the READY line, resets the cycle counter, and puts the shift register back into the LOAD mode. Thus, the transition from state 10 to the READY mode proceeds asynchronously within a few nanoseconds. During this transition the shift-register output remains high because a logic 1 is loaded from the V_{CC} line.

Transmission at 10 characters per second results if a new character is provided within one clock period (9.09 ms) of this READY indication. Even if a new character is received immediately, however, the output will remain at 1 and transmission will not begin until the next clock. This insures a minimum stop pulse duration of two clock periods. If no character is received, the converter will wait in the READY mode indefinitely.

The following modifications adapt the circuit to the Baudot code. Delete the left-hand 74165, and connect the SI and A inputs of the right-hand 74165 to V_{CC}. Then replace the 7410 gate with a 7404 inverter driven off the 7493's D output (the A output now connects only to B_{in}; B and C outputs are left with no connection). □

Switching converter raises linear regulator's efficiency

by Sadeddin Kulturel
Istanbul, Turkey

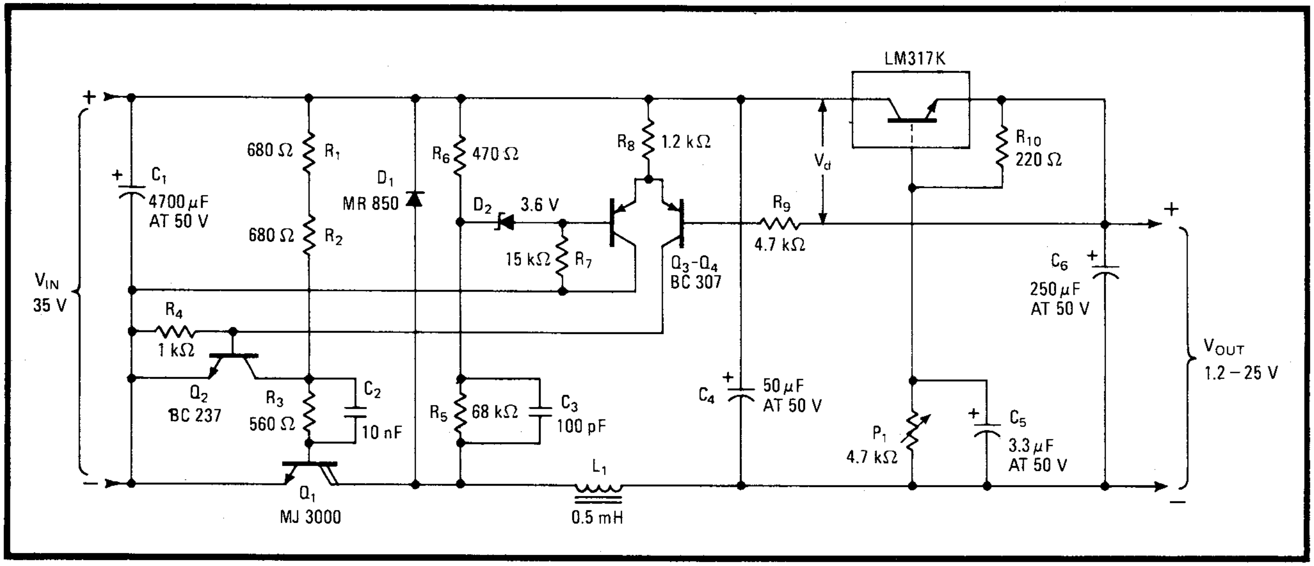
The low ripple and fast recovery of a series-pass voltage regulator can be attained at the high efficiency of a switching regulator if both are combined. In this circuit, the performance is achieved by using the switching circuit as a preregulator for the linear element.

As shown in the illustration of the general circuit, which is designed to transform the 35-volt raw input into a well-regulated output, heat dissipation across the LM317K series element can be reduced if it is made to handle a switched, rather than a continuous, input. Here, the switching regulator is formed by transistors Q_1–Q_4, D_1, and L_1. During power up, Q_1, driven through R_1–R_3, is brought into saturation. Q_2 remains off and Q_3 is turned on.

Switching occurs when V_d equals 3.6 volts, which is D_2's zener voltage. Q_4 then turns on, as does Q_2, and Q_3 is turned off.

As Q_2 turns on, Q_1 switches off, and because of the positive voltage spike created by L_1, load current is momentarily forced through D_1 as V_d decreases. When V_d reaches the lower hysteresis threshold of Q_3 as established by R_5 and R_6, Q_2 and Q_4 turn off, and Q_1 turns on, completing the switching cycle. With the supply's negative path restored, V_d rises until it reaches V_2, and the process is repeated.

The linear regulator can be of any type, including a three-terminal, nonadjustable device. Note that a switching current regulator can be formed if the regulator is replaced by a resistor. In that case, the switching current will be $I_S = V_Z/R$. □



Mixed mode. Switched and linear regulators are combined to form a unit that has the advantages of both—low ripple, fast response, and high efficiency. Here a switched circuit serves as a preregulator for the linear series-pass element, the LM317K.

Reverse magnetic spike drives switching transistor efficiently

by John Klimek
Pretoria, South Africa

Driving switching transistors efficiently is a problem faced by power-supply designers, and it will take some time for V-MOS transistors to replace bipolar ones in high-current power-supply applications. However, starting conduction with a reverse magnetic spike can both cut design costs and improve switching efficiency.

The conventional method of using a simple resistor-limited drive generates high power losses, and a Darlington transistor connection is lossy due to high $V_{ce(sat)}$ and increased switching times. In addition, a current transformer drive is efficient but lacks the ability to start and stop the current cycle in a power transistor.

The current transformer, T_1, is the heart of this circuit. Transistor Q_1 conducts when V_{in} is low and supplies about 5 milliamperes into the 100-turn transformer winding. When V_{in} becomes high, Q_1 turns off and the stored magnetic energy develops a reverse spike simultaneously on all three transformer windings.

The spike through the 10-turn winding starts to conduct Q_2, which is now in the current-transformer self-driving mode. When V_{in} goes low again, Q_1 starts to conduct and along with D_2 shunts the voltage across the 100-turn winding, thereby terminating the driving current of Q_2. This shunting is followed by the termination of the main current. Once the main current is stopped, the magnetizing current is supplied to the 100-turn winding and the entire cycle repeats.

The high turns ratio of 100:1 enhances switching efficiency because only 150 mA of current is shunted by Q_1 to stop 15 amperes through Q_2. In addition to efficiency, the circuit provides excellent insulation between the switching transistor and driving circuitry. □

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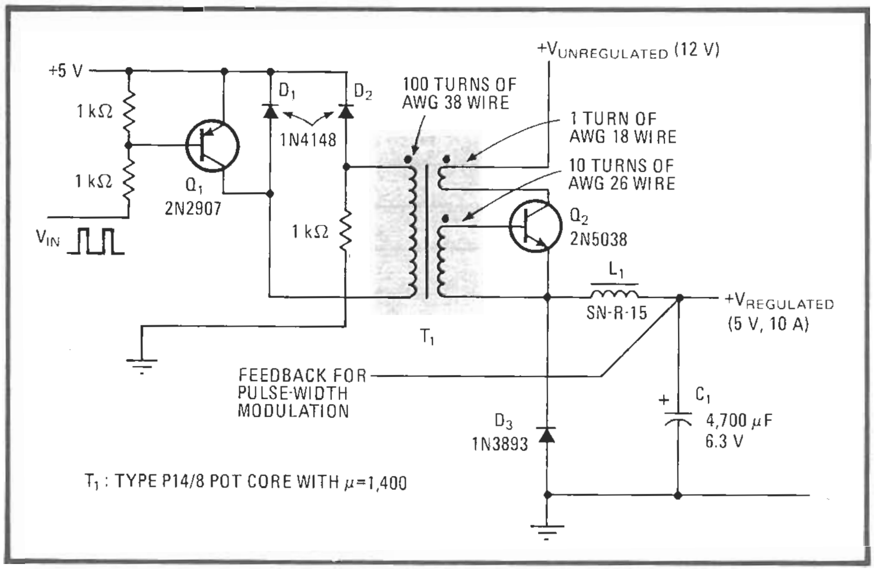

Drive. The current transformer T_1 reduces ohmic and switching losses, thereby driving the transistor efficiently. In addition to an efficient high-current drive, it also provides good insulation between the switching transistor and driving circuitry.

EDITED BY BRAD THOMPSON
AND FRAN GRANVILLE

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READERS SOLVE DESIGN PROBLEMS

Extend low-output-voltage switching regulator's input range

Hua (Walker) Bai, Linear Technology, Milpitas, CA

Internal operating voltages in electronic devices continue to decrease, but input-source voltages don't change. As the difference between input and output voltages increases, so does the improvement in efficiency that a switching regulator offers. Unfortunately, as a switched-mode step-down converter's output voltage decreases, the decrease imposes limitations on the circuit's input-voltage range. This Design Idea shows how to extend a low-output-voltage step-down converter's input-voltage range.

A switching-mode step-down regulator, such as Linear Technology's (www.linear.com) LT1936 (IC_1), includes an internal high-side NPN power transistor between its input, V_{IN}, and switched-output (SW) pin. For highest efficiency, the high-side NPN transistor requires a base voltage that's higher than the input voltage. The circuit of **Figure 1** works best for output voltages greater than 3V. A charge pump comprising diode D_2 and capacitor C_5 maintains the voltage at the Boost pin 3V above V_{IN}. When IC_1's internal power transistor switches off, the voltage at SW goes to ground through D_1. Boost capacitor C_5 charges to 3V supplied from V_{OUT} through D_2. When the power transistor turns on, the voltage at SW jumps to V_{IN}, and the voltage at the Boost pin jumps to V_{IN}+3V, which provides sufficient head room to drive the power transistor into saturation for greatest efficiency.

However, output voltages below 2.8V no longer provide sufficient drive volt-

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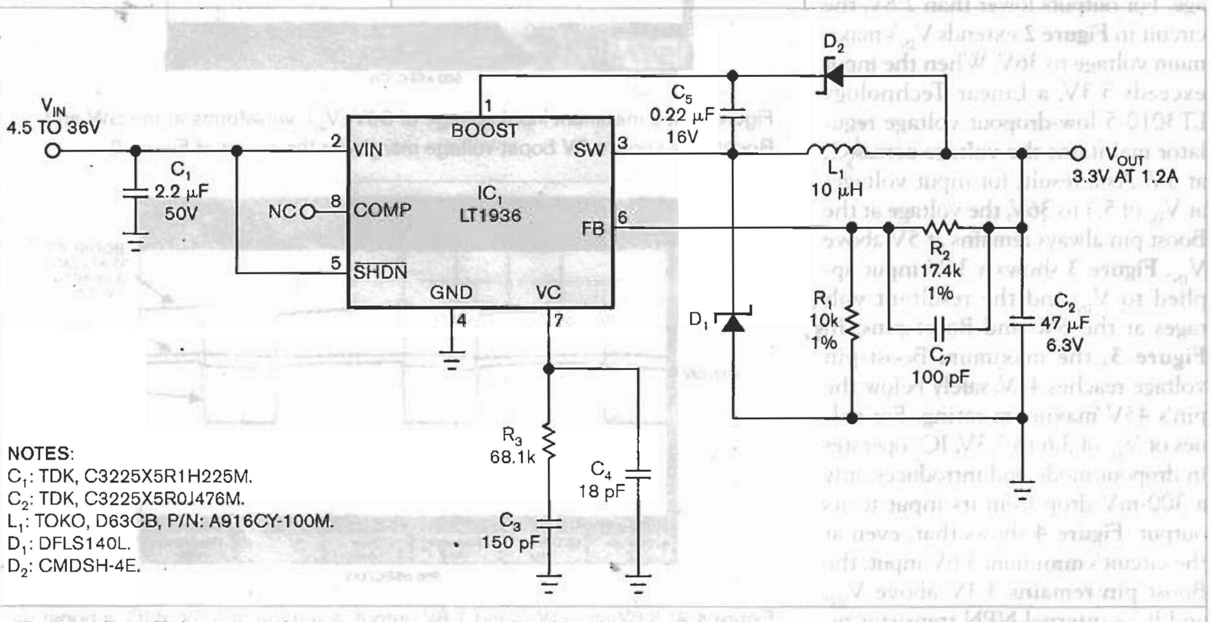

Figure 1 For efficient operation at output voltages of 3.3V or higher, a charge pump comprising D_2 and C_5 provides a voltage boost that provides sufficient drive for IC_1's internal switching transistor.

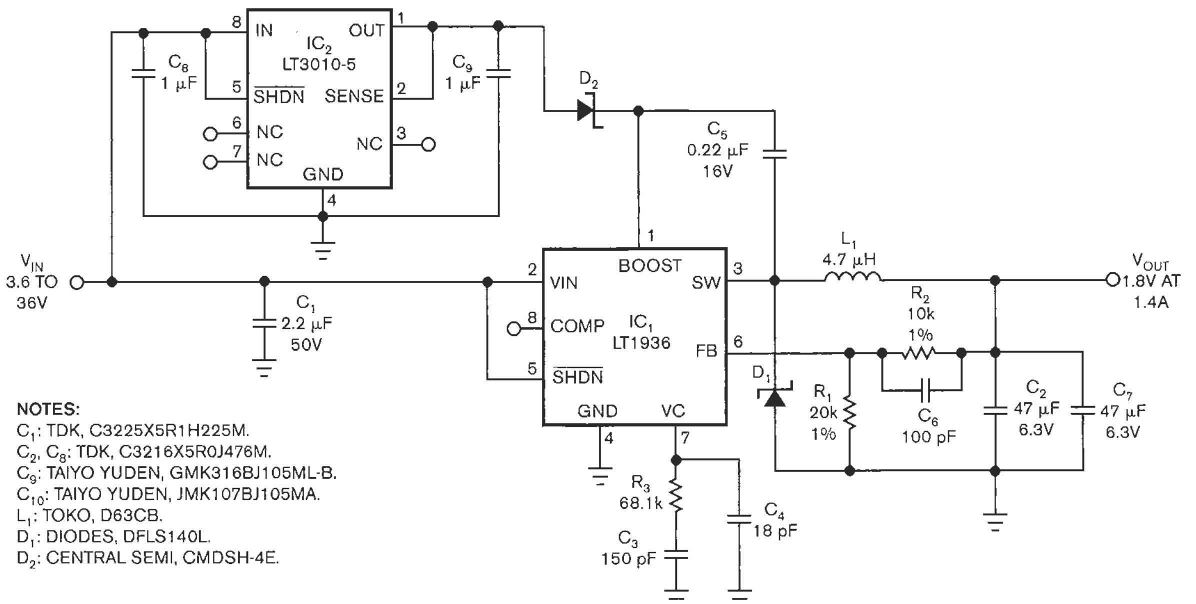

Figure 2 At outputs as low as 1.8V, efficient operation at low input voltages benefits from an added low-dropout regulator, IC_2, for the Boost pin, which extends the circuit's input-voltage range.

age to fully saturate IC_1's switching transistor, and the circuit's efficiency suffers due to increased voltage drop across the transistor. In this situation, connecting D_2's anode to V_{IN} instead of V_{OUT} doubles the Boost pin's voltage to twice the value of V_{IN} but limits V_{IN} to 20V to avoid exceeding the Boost pin's allowable maximum voltage. For outputs lower than 2.8V, the circuit in **Figure 2** extends V_{IN}'s maximum voltage to 36V. When the input exceeds 5.3V, a Linear Technology LT3010-5 low-dropout voltage regulator maintains the voltage across C_9 at 5V. As a result, for input voltages at V_{IN} of 5.3 to 36V, the voltage at the Boost pin always remains at 5V above V_{IN}. **Figure 3** shows a 36V input applied to V_{IN} and the resultant voltages at the SW and Boost pins. In **Figure 3**, the maximum Boost-pin voltage reaches 41V, safely below the pin's 43V maximum rating. For values of V_{IN} of 3.6 to 5.3V, IC_2 operates in dropout mode and introduces only a 300-mV drop from its input to its output. **Figure 4** shows that, even at the circuit's minimum 3.6V input, the Boost pin remains 3.3V above V_{IN}, and IC_1's internal NPN transistor receives sufficient drive voltage for saturated operation. EDN

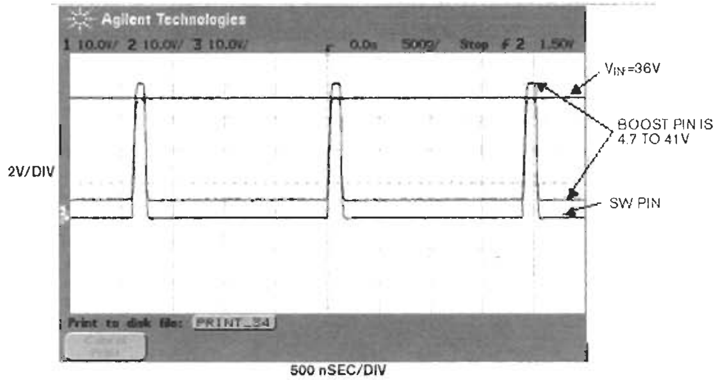

Figure 3 At a maximum input voltage of 36V (V_{IN}), waveforms at the SW and Boost pins show a 5V boost-voltage margin for the circuit of Figure 2.

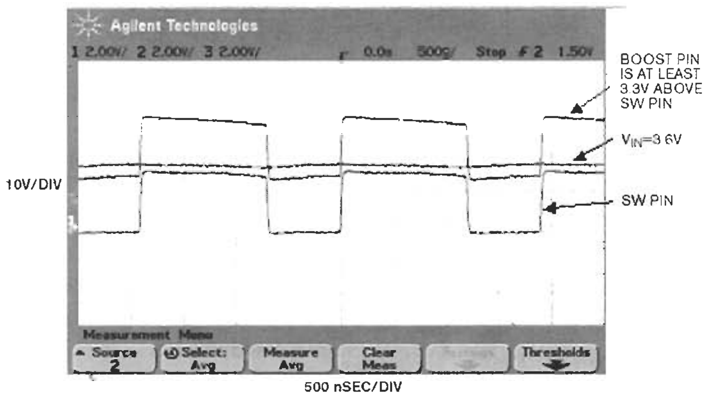

Figure 4 At 3.6V input (V_{IN}) and 1.8V output, a voltage of 3.3V at IC_1's boost pin ensures that IC_1's internal switch still operates in saturated mode.

Switching regulator reduces motor brake's power consumption

Alain Minoz, Elekta Instrument AB, Stockholm, Sweden

For safety reasons, a motor that drives a safety-critical electromechanical assembly often includes an electromagnetic brake on its drive shaft. The brake typically comprises a solenoid coil that actuates a mechanical clutch, and, when you power it, the brake allows the drive shaft to rotate. Although simple and robust, the brake requires a lot of energy to release the clutch and then much less energy to remain actuated.

Measurements show that a brake rated for 24V dc requires a minimum of 18V to release and as little as 8V holding voltage. Substituting those numbers into the equation $P_{COIL}=V^2/R_{COIL}$ shows that, while actuated, the brake consumes less than a quarter of the power required for its initial release. Conversion of excess release power into heat normally doesn't pose problems. However, a precision positioner that uses a brake mounted on a long drive screw can suffer from unacceptable errors if the heat expands the drive screw and alters the assembly's position.

One method of solving the problem involves actuating the brake by applying 24V dc for a brief interval and then reducing the holding voltage to 12V. Under these conditions, the brake dissipates only a quarter of the initial power and thus operates at a reasonable temperature. **Figure 1** shows the influence of actuation voltage on the brake's temperature. As expected, lowering the voltage after actuation drastically lowers the brake's temperature and therefore its effects on the positioning screw.

Figure 2 shows one obvious voltage-reduction approach, which uses relays and a power resistor to halve the voltage applied to the brake. Setting the current-limiting resistance, R_{POWER}, equal to the brake's solenoid resistance, R_{BRAKE}, reveals a few problems. First, the power resistor must dissipate as much power as the brake solenoid's coil. Second, the relays and power resistor occupy considerable space on a pc board. Third, proportioning the values of the R_1C_1 delay circuit's components to achieve a few seconds' delay can prove difficult.

Figure 3 shows another approach, which uses the actuator coil's inductance and replaces relays with an IC. The voltage you apply to the brake need not be continuous, and applying a PWM (pulse-width-modulated) voltage works as well as applying a dc holding voltage because the coil's inductance integrates the current pulses.

A switched-mode voltage regulator can provide an inexpensive and effective PWM-drive voltage. For example, National Semiconductor's (www.national.com) LM2575 adjustable regulator, IC_1, operates over a 7 to 40V range and includes an on/off-control input and a high-impedance feedback input, but any other switching-regulator IC with these two characteristics would also serve. Resistors R_1 and R_2 determine the holding voltage (**Figure 4**). Capacitor C_3 filters the PWM signal to a dc voltage at the feedback input and also maintains the feedback input for a few seconds during start-up at ground, forcing the regulator to deliver the full input voltage to actuate the brake. Diode D_1 quickly discharges the capacitor when the regulator switches off, diode D_2 clamps the switch-off transient voltage that the brake's actuating coil produces, and diode D_3 protects IC_1 against reverse voltage. Photocoupler IC_2 isolates the brake controller from the control circuit.

During start-up, the duration of the regulator's 24V actuation-pulse output fluctuates from 1 to 4 seconds (**Figure 4**). Fortunately, the variation has no impact on the circuit's function but could

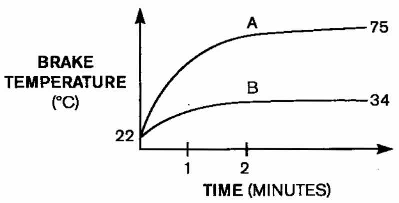

Figure 1 Under continuous operation with 24V applied, the brake's temperature stabilizes at 75°C, or 53°C above ambient temperature (Curve A). Applying a 24V actuation pulse for a few seconds and then applying a 12V holding voltage stabilizes the brake at 34°C, or only 12°C above ambient temperature (Curve B).

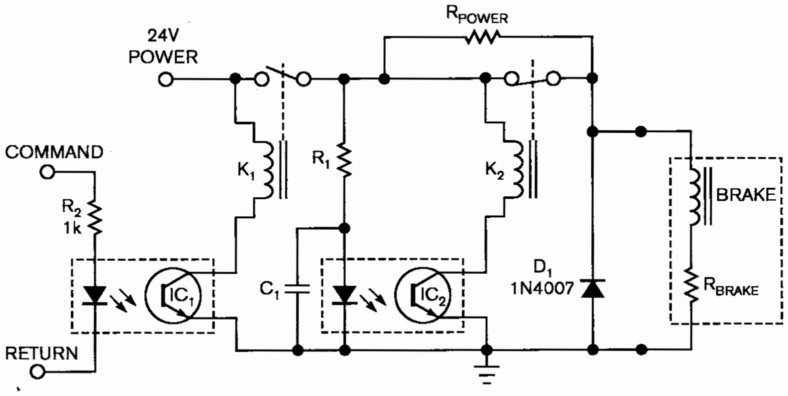

Figure 2 Actuating the brake release trips relay K_1 and applies 24V to the brake. An RC network delays K_2's actuation. When normally closed relay K_2 opens, resistor R_{POWER} reduces the voltage applied to the brake to the holding level.

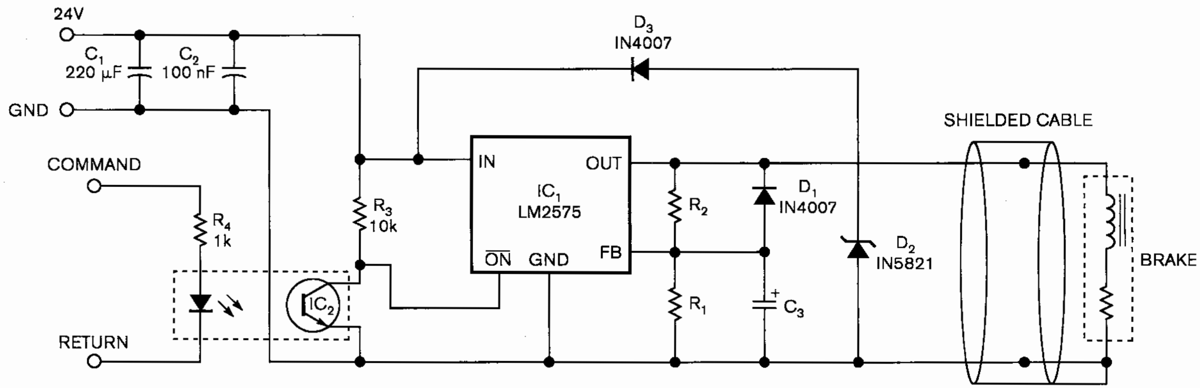

Figure 3 Applying the command input switches on PWM regulator IC_1, and capacitor C_1 holds IC_1's feedback input low, applying a maximum output voltage of 24V to the brake until C_1 fully charges. As the feedback voltage slowly rises to 1.23V, the regulator's output voltage decreases to approximately 12V, the brake's nominal holding voltage.

present a problem if another application requires a precisely timed actuation pulse. After start-up, the regulator delivers a 12V holding voltage, reducing the power demand to one-quarter of the start-up value. As a bonus, the circuit uses inexpensive components, occupies only a few square centimeters of pc-board area, and eliminates the need for two electromechanical relays. Wiring for the PWM-drive voltage can radiate electrical noise unless the circuit is adjacent to the brake. For remote installation, use a shielded twisted-pair cable to minimize noise radiation.

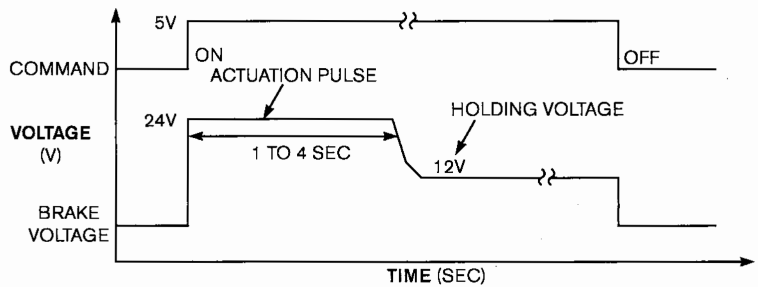

Figure 4 After the actuation pulse applies full voltage to the brake, the regulator's output gradually decreases to the nominal holding voltage.

Controller+MOSFET Equals Smaller ORing Solutions

By David Morrison, Editor in Chief

The use of Schottky diodes to connect redundant power supplies in parallel for load sharing — a technique known as ORing — offers a simple means of protecting the system against a power-supply failure. However, because Schottky diodes introduce voltage drops and power losses, they are replaced in some applications with power MOSFETs, which are controlled in such a manner to act like ideal diodes. Because the MOSFETs selected have low $R_{DS_{ON}}$, their conduction losses are a fraction of a Schottky's losses, and their voltage drop is typically negligible. Consequently, heatsinking requirements are reduced with a MOSFET.

Naturally, the tradeoff in using a MOSFET for ORing power supplies is added complexity because an additional controller is needed. However, with that added complexity, designers not only get reduced power losses, voltage drops and easier thermal management, they also have the benefit of additional forms of circuit protection that are easily built into the MOSFET controllers, also known as ideal diode controllers or ORing controllers. Two recent product announcements illustrate how ORing controllers are evolving to shrink the controller and MOSFET footprint, while also providing greater functionality.

Linear Technology's LTC4358 is a single high-voltage ideal diode controller with a co-packaged 5-A (20-mΩ) n-channel MOSFET. The LTC4358 regulates the forward voltage drop across the internal MOSFET to ensure smooth switchover from one path to another without oscillation. If a power source fails or is shorted, the LTC4358 ensures a fast 500-ns turn-off to minimize reverse current transients. The device is offered in 4-mm × 3-mm 14-pin DFNs and 16-lead TSSOPs.

Designed to operate with a 9-V to 26.5-V supply, the LTC4358 shrinks pc-board footprint versus two-chip controller+MOSFET designs. Alison Steer, product marketing manager for mixed-signal products at Linear Technology, describes the space savings versus an existing solution.

"Our single ideal diode controller, LTC4357, with an external SO-8 FET takes at least 0.125 in^2," Steer explains. "The LTC4358 TSSOP package consumes about one-third that space, and the DFN, half that again. This is only for the components — extra area is needed for heat dissipation. Choosing an external FET with similar $R_{DS_{ON}}$ as the LTC4358 results in the same thermal area requirement. For 5 A, a single-sided board required 2 in. × 2 in. of area."

Pricing for the LTC4358 begins at $2.10 each in 1000-piece quantities.

Picor, a subsidiary of Vicor, has introduced the Cool-ORing family of active-ORing controller ICs and full-function active-ORing solutions. Offered in various IC package styles, the controller chips are designed to drive industry-standard n-channel MOSFETs. Meanwhile, the modules co-package the company's controllers with optimized MOSFETs in 5-mm × 7-mm land-grid arrays (LGAs).

The full-function modules (PI2121, PI2122, PI2123 and PI2125) provide a compact alternative to separate controller+MOSFET combinations as well as competing ORing modules. According to Picor, the 35-mm^2 footprint of the LGA is up to 70% smaller than existing solutions. Moreover, these devices are said to require less thermal derating (no derating up to 60°C ambient) than some of the existing modules.

The small size of the new modules is partly attributed to their use of proprietary MOSFET die developed for Picor. For example, the PI2121, an 8-V, 24-A device suitable for ≤ 5-V bus applications, contains a MOSFET with an $R_{DS_{ON}}$ of 1.5 mΩ typ. To achieve that value of on-resistance in a discrete MOSFET, it may be necessary to select a FET that is larger than Picor's controller+MOSFET module, according to Carl Smith, director of strategic marketing and business development at Picor.

The two other modules include the PI2123, a 15-V, 15-A solution suitable for ≤ 9.6-V bus applications, and the PI2125, a 30-V, 12-A solution suitable for ≤ 12-V bus applications. Those two devices feature MOSFETs with on-resistances of 3 mΩ and 5.5 mΩ, respectively.

The controller ICs (PI2001, PI2002 and PI2003) and the modules target a range of bus voltages, provide scalability for different current levels and feature fast dynamic response to input power-source failures. Reverse-current turn-off delay time for the PI2121, PI2123 and PI2125 modules and for the PI2001 controller is 160 ns typical. Programmable undervoltage and overvoltage detection, overtemperature detection and active low-fault flag output are common functions to all seven devices. But then several other features such as load disconnect and performance levels are specific to certain family members.

The full-function modules start at $1.98 each in 10,000-piece quantities. The controllers are available in 3-mm × 3-mm 10-lead TDFNs and 8-lead SOICs with prices starting at $0.76 each in 10,000-piece quantities. **PETech**

Power-Supply Failures Are Mostly Preventable

By Tom Skopal, Vice President, Sales & Marketing,
Acopian Power Supplies, Easton, Pa.

Many and probably most power-supply failures are easily preventable. They are most frequently the result of overstressing the supply with heat (either ambient or self-generated), transients or overloading. If you're a power-supply designer, many of these causes may be obvious to you. But don't assume they are obvious to your customers.

Or, if you're specifying power supplies for use in your systems, be aware of the not-so-obvious pitfalls such as misleading reliability specs or low-cost power supplies with specifications that may be too good to be true. By following the guidelines listed here, most power-supply failures can be prevented.

Be sure that air can circulate freely around and through the supply (if necessary, use a fan), and don't block the vented surfaces of enclosed supplies. If the input power is likely to contain large transients — for example, from motors and high currents being switched — use transient suppressors on the troublemakers when possible and put an input filter on the supply.

Continuously drawing more current than that for which the supply is rated will likely result in an early failure. This is particularly true when the environment is hot and the manufacturer's derated specification for that temperature is not heeded.

As with most other aspects of life, you get what you pay for. ... Relatively low-priced power supplies usually have components stressed to much higher levels than higher-priced supplies.

Frequently, supposed "failures" are actually misapplication problems. Be certain the power supply can handle not only the steady-state operating current, but also the maximum surge current the load may draw. Most power supplies incorporate electronic current limiting on their outputs and may simply "lock up" when the load tries to draw a substantial surge. This is especially true with high capacitance loads and nonlinear loads such as incandescent lights.

Another simple consideration is your input source. Are you certain the actual ac input voltage is always within the specified range? Keep in mind that a supply with a 105-Vac to 125-Vac input range most likely won't work properly (or at all) in Japan, where the nominal is 100 Vac.

As with most other aspects of life, you get what you pay for. You won't get the highest-reliability supply for the lowest price. Relatively low-priced power supplies usually have components stressed to much higher levels than higher-priced supplies, so they have less operating margin and are likely to fail more quickly.

Heatsinking also may be skimpy, resulting in semiconductors running hot. Note that many low-priced units have a warranty period of less than a year, which says something about their manufacturers' confidence in their product.

And be realistic when it comes to your expectations of reliability. Mean-time-between-failure (MTBF) calculations are often based solely on parts counts, without taking into account the effects of temperature and other stresses. So don't place a great deal of confidence in those ratings.

For example, how can the MTBF of a power supply be 500,000 hours when the electrolytic capacitors in it typically degrade in less than 10 years, which is only 88,000 hours? If you must have a power-supply function for many years without risk of failure, consider redundancy. The outputs of two power supplies, each capable of individually supporting the load, can be connected in parallel (through diodes, to prevent interaction) so that if one drops out, the other will continue to power the load without interruption while the failed supply is being repaired or replaced.

Even when the individual power supplies each have a modest MTBF rating, redundancy can raise the effective MTBF of the power system to astronomical levels. Of course it's important to continuously monitor the outputs of both, so that a failure doesn't go undetected. **PETech**

Tom Skopal *has been at Acopian Technical Company for almost 40 years and has written numerous articles for electronics publications. He received a BSE degree from Drexel University and has done extensive graduate work at both Drexel University and Seton Hall. He was the recipient of an Idea of the Year award from* Electronic Design *magazine.*