design ideas

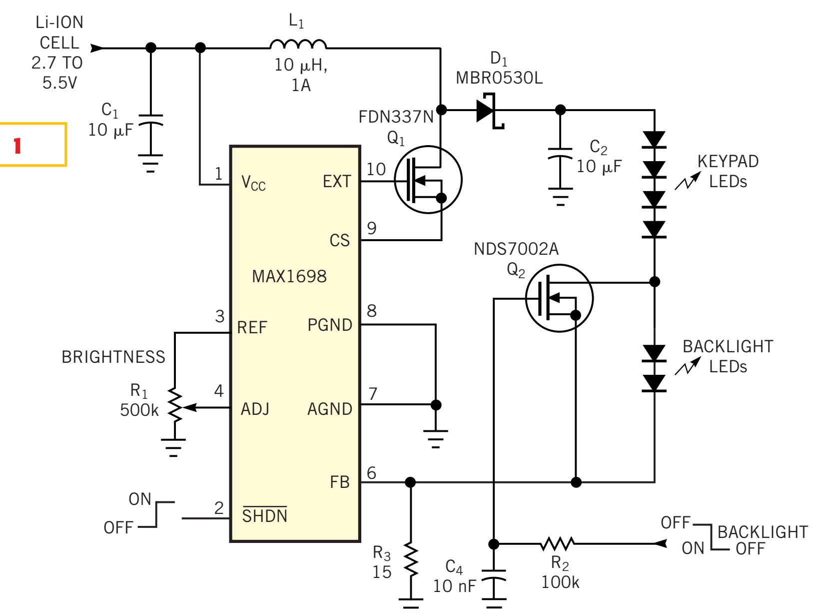
Edited by Bill Travis and Anne Watson Swager

Single FET controls LED array

Len Sherman, Maxim Integrated Products, Sunnyvale, CA

White-LED backlights are gaining acceptance because they offer higher reliability and simpler drive circuitry than backlights based on CCFL (cold-cathode-fluorescent-lamp) and EL (electroluminescent) technology. As a result, white-LED backlights are increasingly common in PDAs (personal digital assistants), cell phones, digital cameras, and other portable devices. A design in which the display requires backlighting for extended periods needs an efficient circuit that drives the LEDs with a controlled current and eliminates the wasted power associated with current-limiting resistors. **Figure 1** shows a switch-mode boost design that regulates current instead of voltage. Because all the LEDs are connected in series, they all receive the same current without the need for ballasting resistors. Identical currents help achieve uniform intensity. And, because the output current is low (20 mA in this case), the output-filter capacitance, C_2, can be smaller than for a load consisting of parallel-connected LEDs.

Figure 1



When this circuit turns off the backlight LEDs, the keypad LEDs remain on with no change in intensity.

The circuit's 90% conversion efficiency offers a distinct power-saving advantage over resistor-limited and linearly regulated designs. It might appear that a series-LED connection is unsuitable for applications in which some (but not all) LEDs must be off. A cell phone, for example, sometimes needs that capability for occasions when the display is off but the keypad remains lit. Or, a PDA might need to play a sound file while maintaining illumination in the buttons but not the display. In the circuit of **Figure 1**, switching off individual LEDs or groups of LEDs is not a problem, even with series drive. Applying a logic-high level to the gate of a simple MOSFET switch, Q_2, turns off a subset of LEDs by shunting their current. The remaining LEDs (for the keypad, for example) remain on, and their intensity remains constant because IC_1 regulates their current, by sensing the voltage across R_2 (300 mV at full brightness). When the circuit turns the LEDs on and off, the R_2-C_4 network at the gate of Q_1 slows the load changes sufficiently to prevent transients in the LED drive current. Other features include adjustable intensity via the ADJ pin and full shutdown via the $\overline{\text{SHDN}}$ pin.

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Circuit protects battery from overdischarge

Martin Wuzik, Implex AG Hearing Technology, Ismaning, Germany

All electronic circuits powered by a battery discharge the battery. In some applications, it is undesirable to overdischarge the battery, because it could irreversibly reduce the battery's capacity and the number of discharge/charge cycles. The circuit in **Figure 1** protects a single NiMH (nickel-metal-hydride) cell by disconnecting the load from the battery. **Figure 2** shows the output voltage, V_{SYS}, versus the input voltage, V_{BAT}. For this NiMH battery, the switching points are 1.1 and 1.3V. If the battery discharges and V_{BAT} drops below 1.1V, Q_1 switches off ,and the node Main Circuit disconnects from the battery. In that case, the battery's only load is the pair of voltage detectors IC_1 and IC_2 from Ricoh (www.ricoh-usa.com). The load current of one detector is typically 800 nA, so the battery drain is 1.6 μA. The user must now charge the battery. Once the battery charges and the voltage reaches 1.3V, the load reconnects to the battery and remains connected as long as V_{BAT} stays above 1.1V.

IC_1 is a voltage detector with a 1.3V setpoint and a push-pull output. IC_2 has a 1.1V setpoint. An important difference between the two detectors is that IC_2 has an open-drain output. If the battery voltage drops but remains within the 1.1 to 1.3V range, IC_1's output is low, and Q_2 switches off. Q_3 switches on because IC_2's output is still in the high-impedance state. If V_{BAT} drops below 1.1V, IC_2's output switches low, Q_3 turns off, and, as a result, Q_1 also switches off. As soon as V_{BAT} drops below 1.1V, the load disconnects from the battery. The load reconnects to the battery only when the battery charges to a voltage higher than 1.3V. At voltages of 1.1 to 1.3V, IC_2 cannot switch on Q_3 because the IC's output is an open-drain type and V_{SYS} is low. IC_1's output must assume a high state to switch on Q_2 and to finally switch on Q_1 on. The transistors are low-threshold MOSFETs from Supertex (www.supertex.com). The circuit uses no trimming resistors. You can select IC_1 and IC_2 off the shelf with 100-mV steps and 2% switching-point accuracy. You can adapt the circuit for the higher voltages of Li-ion batteries by selecting the voltage detectors.

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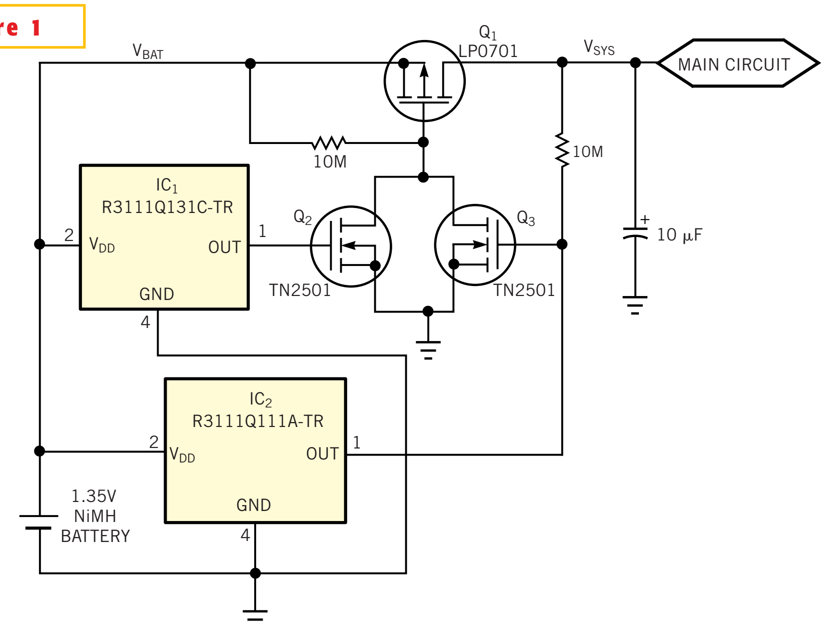
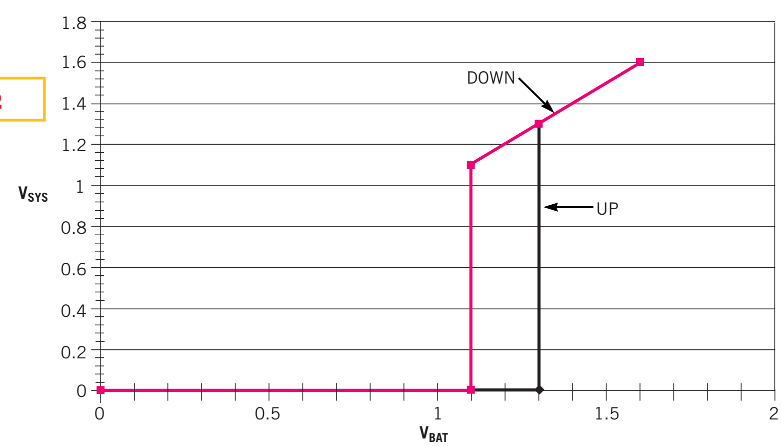
Figure 1



A simple circuit prevents excessive discharge of NiMH cells.

Figure 2



The load disconnects from the battery when the voltage drops below 1.1V and reconnects when the battery charges above 1.3V.

design ideas

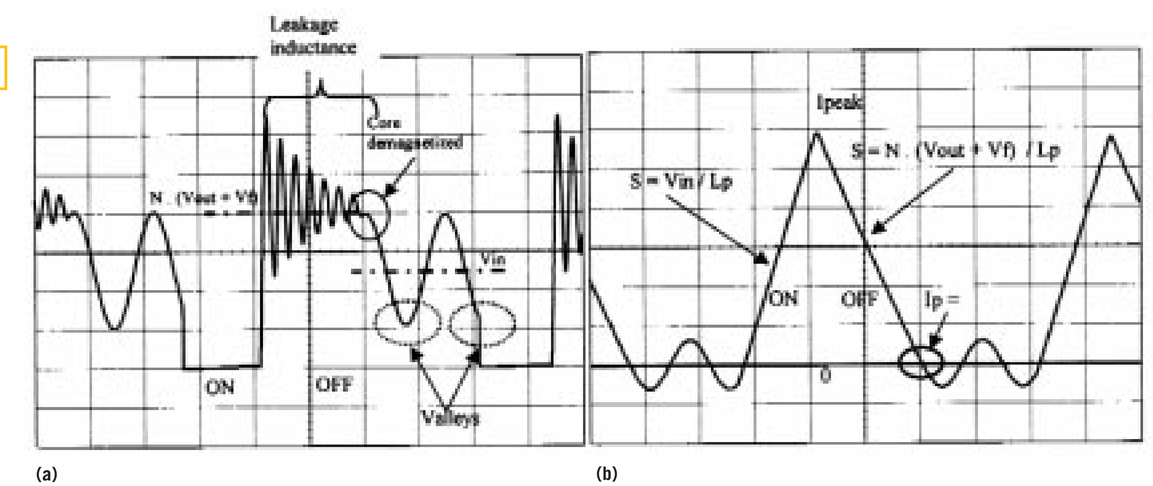
Two diodes change demagnetization-signal polarity

Christophe Basso, On Semiconductor, Toulouse, Cedex, France

Power-supply designers usually like flyback converters to operate in DCM (discontinuous-conduction mode) rather than in CCM (continuous-conduction mode). In DCM, the flyback converter is a first-order system at low frequencies, which eases the feedback-loop compensation. You can use a low-cost secondary rectifier, thanks to soft blocking conditions. In DCM, I_P goes to zero, and the diode stops conducting, whereas the power-switch turn-on event in CCM forces the diode to brutally stop conducting. Also in DCM, valley switching ensures minimum switching losses that C_{OSS} and all the parasitic capacitances bring.

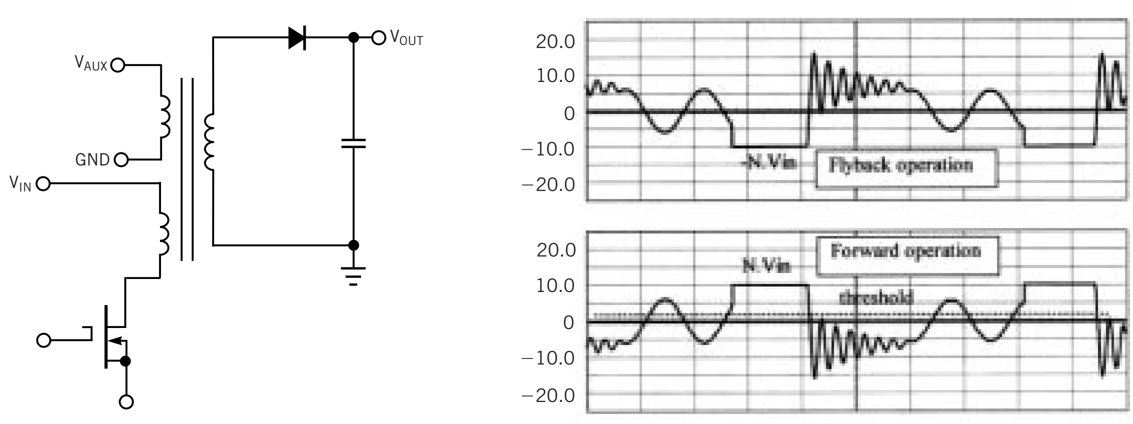
In valley switching, or QR (quasiresonant) operation, the curve of the drain-source voltage, V_{DS}, of a typical flyback converter, shows that when the power switch closes, you observe a low level due to the $R_{DS}(ON)\times I_P$ product (**Figure 1a**). At the switch opening, V_{DS} rises quickly and starts to ring at a high frequency because of the leakage-inductance presence. During this time, the primary current transfers to the secondary, and a reflected level of $N\times(V_{OUT}+V_F)$ appears on the MOSFET drain, where N is the secondary-to-primary turns ratio, V_{OUT} is the output voltage, and V_F is the secondary-diode forward drop. As soon as the primary current has fallen to zero in DCM operation, the transformer core is fully demagnetized (**Figure 1b**). The drain

Figure 1



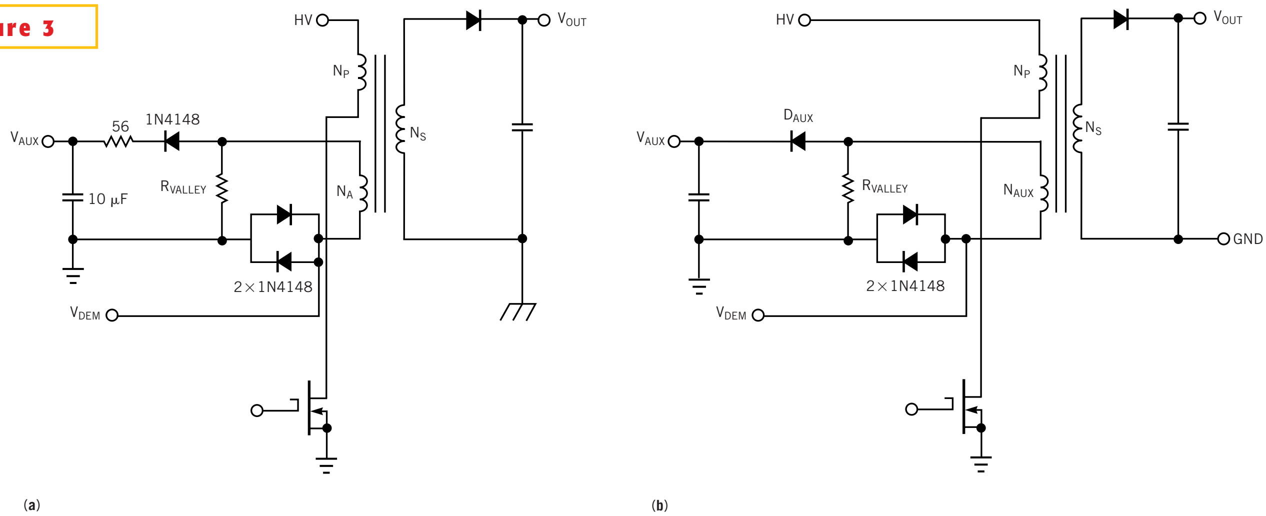
A typical drain-source waveform of a flyback converter shows high-frequency ringing (a). In DCM operation, the primary current ramps up and down to zero (b).

Figure 2



An auxiliary winding (a) lets you observe the flux image in the transformer's core for both flyback and forward operation (b).

Figure 3



A simple component arrangement allows forward-mode detection with a flyback-like PWM controller (a) or flyback-mode detection with a forward-like controller (b).

branch starts to ring but at a lower frequency than in **Figure 1a** because the primary inductance, L_P, is now involved.

This natural oscillation exhibits the following frequency value, where C_{LUMP} represents all of the circuit's parasitic capacitances, such as C_{OSS} and the stray capacitance from the transformer.

$$F_{RING} = \frac{1}{2\pi\sqrt{L_P \bullet C_{LUMP}}}.$$

As with any sinusoidal signal, there are peaks and valleys. When you restart the switch in the valley, all the parasitic capacitance values are at their lowest possible levels. Also, the capacitive losses, which are equal to $1/2 \times C_{LUMP} \times V_{DS}^2 \times F_{SW}$, are small because the MOSFET is no longer the seat of turn-on losses, which removes the usual turn-on parasitics. That is the secret of QR operation.

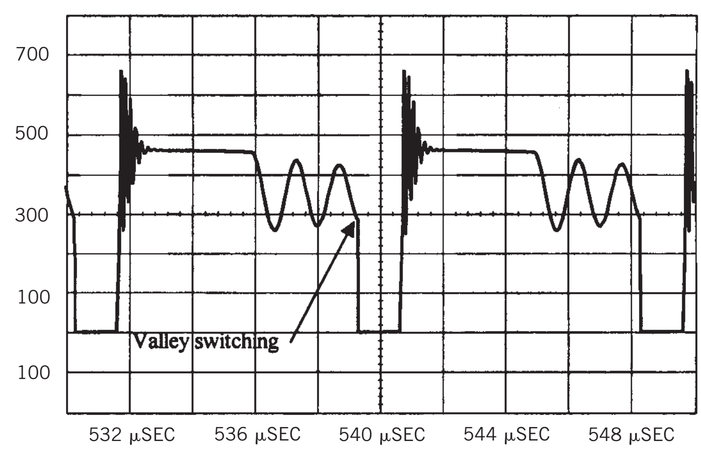
You can easily observe the core flux through an auxiliary winding (**Figure 2a**). Thanks to the coupling between the windings, the auxiliary section delivers a voltage image of the core's flux through the following formula:

$$V_{AUX} = N \bullet \frac{d\phi}{dt}.$$

Now, you can wire the winding either in flyback operation, as the power winding, or in forward operation. The observed signals look the same but have different polarity (**Figure 2b**). Note that both signals center about ground. The problem lies in the fact that most PWM controllers accept only the flyback polarity. Typical examples include the MC-33364 and MC44608 (www.onsemi.com). In battery-charger applications, you usually wire the auxiliary winding—the one that self-supplies the controller and gives the demagnetization signal—in forward mode. The reason is simple: When the battery you charge is close to 0V, the auxiliary windings are also nearly 0V because both windings are coupled in flyback mode. By operating in forward mode, whatever happens on the secondary side is invisible, and the voltage is always there to supply the controller. However, the demagnetization signal now has the wrong polarity, and the controller doesn't restart at the core's reset event.

Figure 3a shows a way around this problem. You still wire the winding for forward operation, but you add two extra diodes in series with the winding. At

Figure 4



When you properly adjust the time constant using R_{VALLEY}, the switch restarts in the middle of the valley.

the switch closing, you apply $N \times V_{HV}$, where N is the ratio between the auxiliary winding, N_A, and the primary winding, N_P. You clamp V_{DEM} to −0.6V, and the current circulates through R_{VALLEY}. At the switch opening, the voltage reverses and becomes positive but clamped to 0.6V on V_{DEM}. When this level collapses, the PWM controller reactivates the power switch.

You can implement this same type of circuit for PWM controllers that need a forward demagnetization signal but for which you would like to operate the auxiliary winding in flyback mode (**Figure 3b**). The problem and the cure are similar.

When you properly select R_{VALLEY}, this resistance naturally combines with sense-pin internal capacitance to add switch delay right in the middle of the wave (**Figure 4**).

Some controllers exhibit different demagnetization threshold levels. The MC33364 starts at around 1V, and the MC44608 toggles at 65 mV. Because of the diodes, you clamp V_{DEM} between ±600 mV, which could not trigger the MC33364. A small offset from the internal reference to the demagnetization pin brought by a 150-kΩ resistor and a typical R_{VALLEY} of 10 kΩ have provided good circuit operation.

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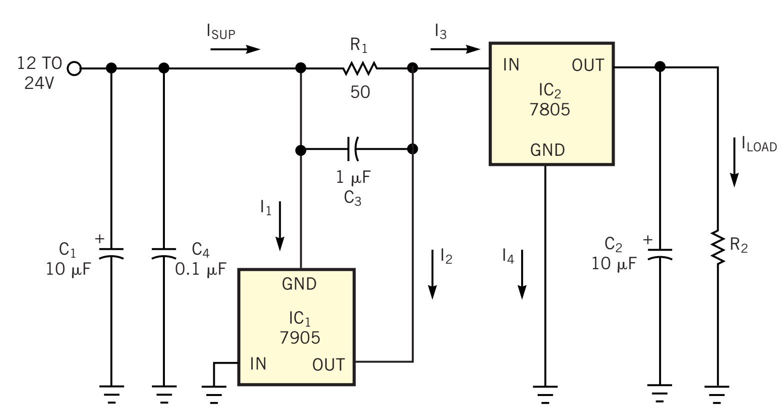
Simple scheme keeps current drain constant

Peter Güttler, APS Software Engineering GmbH, Cologne, Germany

It is sometimes advantageous to keep the overall current consumption of an electronic device constant. A large, seven-segment display, for example, draws nearly zero current when no segment is on to hundreds of milliamps when fully lit. This heavily varying current can cause EMI problems when a device receives its power through long cables from a remote power supply. The low-parts-count circuit in **Figure 1** keeps current consumption constant. IC_2 is an ordinary three-terminal regulator that supplies 5V to the load, R_2. IC_2 draws a total current $I_3 = I_{LOAD} + I_4$. (I_4 is approximately 8 mA, the quiescent current of IC_2). The negative three-terminal voltage regulator, IC_1, maintains 5V across R_1. The current through R_1 is $I_2 + I_3$. So, $I_2 = 5V/R_1 - I_3$, and total supply current $I_{SUP} = I_1 + 5V/R_1$. I_1 is approximately 2 mA, the quiescent current of IC_1. If the load draws more current, IC_1 reduces I_2 and vice versa.

This regulation works well as long as I_3 is smaller than $5V/R_1$. If the load draws more current, IC_1 stops regulating and the voltage drop across R_1 rises above 5V. This example sets R_1 at 50Ω, setting the supply current, I_{SUP}, to approximately 102 mA. C_1 and C_4 are input-filter capacitors, C_2 improves ripple rejection, and C_3 provides stability. Note that R_1 dissipates $(5V)^2/R_1$ and must have an adequate power rating. IC_1 and IC_2 may require heat-sinking. The minimum supply voltage for this circuit is 12V. (The minimum input voltage for IC_2 = 7V + IC_1's reference voltage.) If your application cannot tolerate the 5V drop across R_1, try using an LM337 with a 1.25V reference voltage for IC_1.

Figure 1



This circuit maintains a constant supply current of approximately 102 mA.

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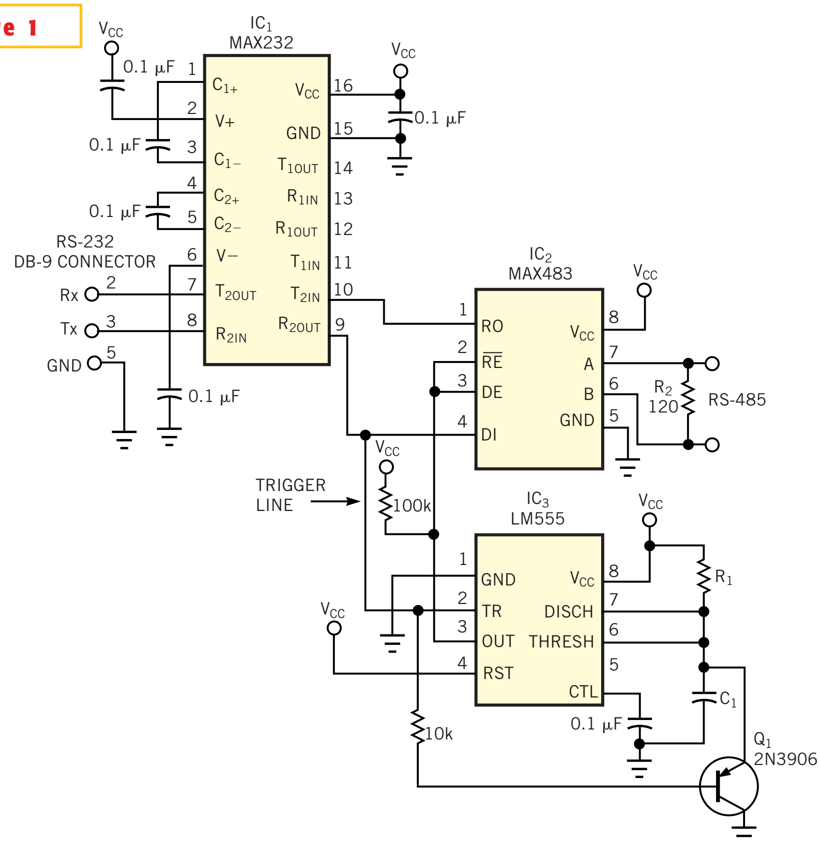
RS-232/485 converter has automatic flow control

John Howard, Kw Aware, Ventura, CA

RS-485 COMMUNICATIONS can provide longer range and better noise immunity than RS-232, as well as multidrop capability. Because it does not have separate transmit and receive lines, RS-485 requires flow control. RS-232/485 converters often use one of the RS-232 handshaking lines to control direction, but several communications-software packages do not support flow control. The circuit in **Figure 1** is an RS-232/485 converter that uses the transmitted signal itself to control the flow. The circuit uses MAX232 and MAX483 interface circuits, IC_1 and IC_2 from Maxim Integrated Products (www.maxim-ic.com) to convert between the ICs' respective signal levels and logic levels. Because both ICs invert the signal, the circuit preserves the original sense of the signal. The MAX483 is normally in the receiving mode. When transmission begins, the signal triggers IC_3, the LM555 timer, which in turn toggles IC_2's DE and RE lines, putting the chip into the transmitting mode.

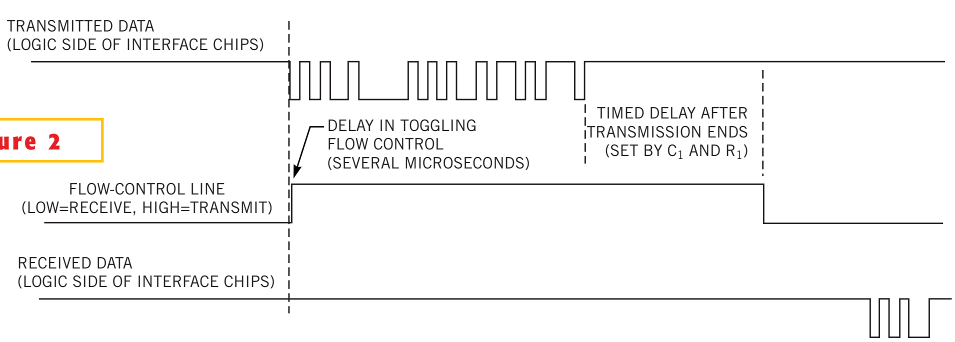
Q_1, the 2N3906, fully discharges C_1 each time the trigger line goes low, restarting the timing cycle. The values of R_1 and C_1 determine how long IC_3 maintains the transmitting mode after transmission ends. This interval should be long enough such that the converter doesn't switch directions while sending characters containing long sequences of zeros. On the other hand, it shouldn't be so long that the converter misses received characters. The interval T in seconds is $T=1/R_1C_1$, where R_1 is in ohms, and C_1 is in farads. The flow control responds within a few microseconds after transmission commences, so the converter does not miss any bits at low and medium data rates. The application for this circuit operates at 14,400 bps. **Figure 2** shows the timing of the serial and flow-control lines. The entire circuit can fit into a DB-25 (or even a DB-9) back shell.

Figure 1



Automatic flow control makes RS-232/485 conversion easy.

Figure 2



R_1 and C_1 determine how long the transmitting mode lasts.

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Circuit provides accurate RTD measurements

Tito Smailagich, ENIC, Belgrade, Yugoslavia

THE CIRCUIT IN **Figure 1** is an efficient measuring circuit for PT100 RTD elements. IC_1 provides an accurate 2.5V output and, together with P_1 and R_1, also provides a stable 1-mA current to the RTD element. The output of IC_2 is −0.1V. P_2 provides a zero adjustment (at 0°C) for IC_3, an amplifier with a gain of 25. P_3 provides a gain adjustment. If, for example, you replace the RTD element by a fixed resistance of 124.78Ω (the RTD's resistance in **Table 1** at 64°C), you would trim P_3 to obtain 0.64V output. Tolerance values for Class A and B elements are ±0.35 and ±0.8°C, respectively. You can use the standard values from **Table 1** or, if you need more accuracy, you can calibrate the PT100 element in a controlled-temperature environment.

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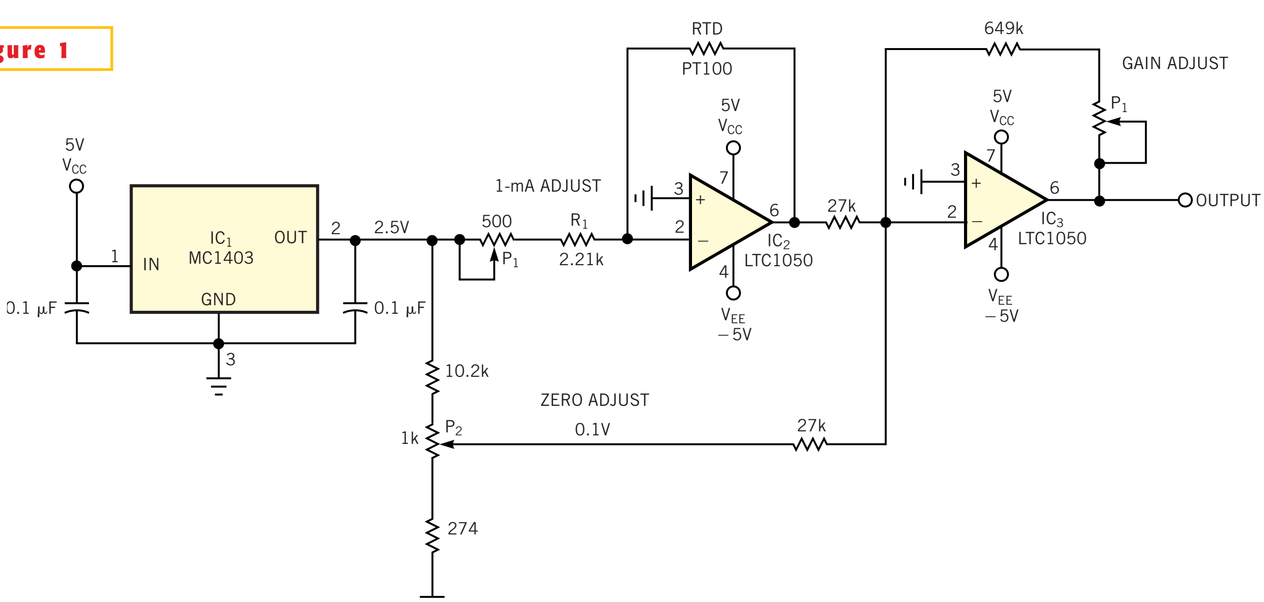
TABLE 1—RESISTANCE VERSUS TEMPERATURE FOR PT100 RTD ELEMENT

°C	Ω	°C	Ω	°C	Ω	°C	Ω	°C	Ω
−20	92.16	6	102.34	32	112.45	58	122.47	84	132.42
−19	92.55	7	102.73	33	112.83	59	122.86	85	132.8
−18	92.95	8	103.12	34	113.22	60	123.24	86	133.18
−17	93.34	9	103.51	35	113.61	61	123.63	87	133.57
−16	93.73	10	103.9	36	114	62	124.01	88	133.95
−15	94.12	11	104.29	37	114.38	63	124.39	89	134.33
−14	94.52	12	104.68	38	114.77	64	124.78	90	134.71
−13	94.91	13	105.07	39	115.15	65	125.16	91	135.09
−12	95.3	14	105.46	40	115.54	66	125.54	92	135.47
−11	95.69	15	105.85	41	115.93	67	125.93	93	135.85
−10	96.09	16	106.24	42	116.31	68	126.31	94	136.23
−9	96.48	17	106.63	43	116.7	69	126.69	95	136.61
−8	96.87	18	107.02	44	117.08	70	127.08	96	136.99
−7	97.26	19	107.4	45	117.47	71	127.46	97	137.37
−6	97.65	20	107.79	46	117.86	72	127.84	98	137.75
−5	98.04	21	108.18	47	118.24	73	128.22	99	138.13
−4	98.44	22	108.57	48	118.63	74	128.61	100	138.51
−3	98.83	23	108.96	49	119.01	75	128.99	101	138.88
−2	99.22	24	109.35	50	119.4	76	129.37	102	139.26
−1	99.61	25	109.73	51	119.78	77	129.75	103	139.64
0	100	26	110.12	52	120.17	78	130.13	104	140.02
1	100.39	27	110.51	53	120.55	79	130.52	105	140.4
2	100.78	28	110.9	54	120.94	80	130.9	106	140.78
3	101.17	29	111.29	55	121.32	81	131.28	107	141.16
4	101.56	30	111.67	56	121.71	82	131.66	108	141.54
5	101.95	31	112.06	57	122.09	83	132.04	109	141.91
								110	142.29

Figure 1



This circuit provides accurate temperature measurements using a PT100 RTD element.