

MAXIM Engineering Journal

Volume Twenty

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News Briefs

■ MAXIM REPORTS RECORD REVENUES AND EARNINGS FOR THE QUARTER AND THE FISCAL YEAR

Maxim Integrated Products, Inc., reported record net revenues of \$76.0 million for the fourth quarter ended June 30, 1995, compared to \$44.1 for the same period a year ago. This represents a 72.3% gain in net revenues from the same quarter a year ago. Net income was \$11.5 million (or \$0.34 per share) for the quarter, compared to net income of \$6.8 million (or \$0.21 per share) for the same quarter in fiscal 1994.

Maxim ended fiscal 1995 with net revenues of \$250.8 million, a 63.0% increase over fiscal 1994 net revenues of \$153.9 million. Operating income for fiscal 1995 was \$57.2 million, a 60.7% increase over the \$35.6 million reported in 1994. Income per share for fiscal 1995 was \$1.17, compared to \$0.76 in fiscal 1994, an increase of 53.9%. Stockholders' equity grew to \$178.7 million at June 30, 1995 from \$130.2 million at June 30, 1994, an increase of 37.3%. Total assets increased to over \$256 million.

Cost of sales in the fourth quarter results included approximately \$3.3 million of costs related to the completion of the Company's transition to 6" wafers. Also, selling, general and administrative expenses increased significantly over the prior quarter as a result of the Company's international expansion and certain one-time costs associated with technology licensing matters.

During the quarter, Maxim's cash and short-term investments increased by \$14.0 million or \$0.41 per share. For the year, the Company increased cash by \$43.9 million or \$1.32 per share, while investing over \$39.4 million in capital equipment and repurchasing approximately \$11.9 million of its common stock.

Inventories were at a record low level of 55 days, representing a substantial improvement over the Q494 level of 91 days. Accounts receivable levels were at 33 days and continue to remain among the lowest in the industry.

Worldwide demand for the Company's products is also at record levels across all geographic regions and all sales channels. Acceptance of new products in fiscal 1995 continued to accelerate, with customers designing in Maxim's new products at a much faster rate than previously experienced. Maxim has doubled its customer base over the past year and increased the direct OEM portion of its business to over 40%. The demand for the Company's products in the third and fourth quarters far exceeded the Company's record 68% growth in revenues reported in the second half of fiscal 1995.

Jack Gifford, Chairman, President and CEO commented, "1995 was a tremendous success for Maxim, its shareholders and employees. We successfully integrated the operations acquired from Tektronix in May 1994, we increased our manufacturing capacity by four-fold, and we set new records for both time-to-market and acceptance rates of our new products by new and existing customers."

Gifford continued, "I am especially pleased that the acquisition of the Tektronix fabrication facility uniquely positions Maxim to take advantage of the worldwide shortage in precision analog products to grow our market share while reinforcing our commitment to our customers. Although we are confident of our long-term growth rates, notwithstanding the bulge in revenues during the next three quarters, we will continue to control spending consistent with our sustainable business levels to ensure that we can continue to deliver consistent revenue and earnings performance in 1997 and beyond."

During the quarter, Maxim was named Dataquest's "European Niche Vendor of 1995." This title reflects Maxim's high level of quality and customer support in the highly competitive European market.

New current-sense amplifiers aid measurement and control

The conventional current-measurement circuit—a discrete difference amplifier fed by the signal voltage across a current-sense resistor—is giving way to smaller, less expensive integrated circuits. The monolithic MAX471, for example, integrates the amplifier with a 35mΩ sense resistor and is capable of sensing continuous currents as high as 3A.

A similar device, the MAX472, allows measurement of virtually any current by connecting current-sense and gain-setting resistors external to the package.

Both amplifiers sense the *high-side* load current and produce a current-source output that may be referenced to ground. That capability lets you generate a ground-referenced output voltage, proportional to load current, by connecting a single resistor between the output and ground. The MAX471 has a fixed internal gain that delivers 500μA per ampere of sensed current, and the MAX472 lets you choose the current-sense and gain-setting resistors for an output range of 0mA to 1.5mA. The MAX472 offers more flexibility, but requires a bit more design work.

High-side sense resistors (between power source and load) are preferable to a low-side resistor between the load and power-source return. Low-side sense resistors produce voltage drops that can disrupt the system's ground connection and complicate the charging and power-management circuitry.

Internal workings

The MAX471 and MAX472 contain two cross-connected difference amplifiers (Figure 1), each driving an npn transistor. Current through the sense resistor develops a voltage that turns one amplifier off and the other on. The active amplifier's npn transistor then produces a current that flows from the more positive side of the sense resistor, through the gain-setting resistor (RG1 or RG2), and through the transistor to the OUT terminal.

An opposite-polarity current through R_{SENSE} turns this diff-amp off and the other on, producing an output

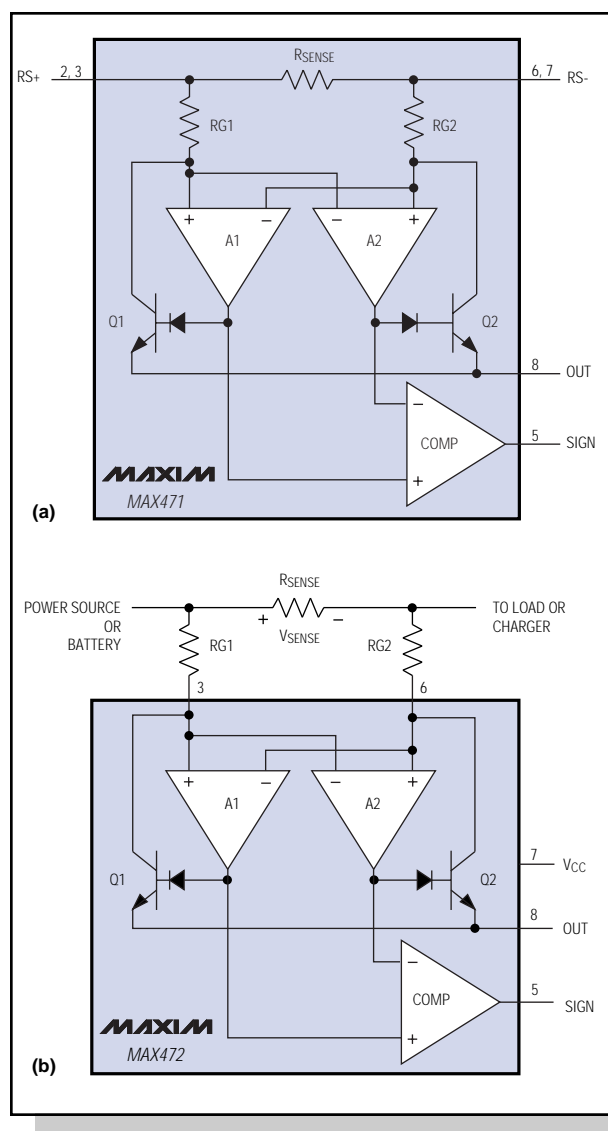


Figure 1. Internal current-sense and gain-setting resistors enable the MAX471 (a) to measure currents up to 3A. For higher current levels, add external sense and gain resistors to the otherwise similar MAX472 (b).

current that always flows in the same direction. Direction of the sensed current is indicated by the SIGN output. This open-drain output simplifies the interface to logic circuitry operating at supply voltages within the IC's 3V to 36V supply range.

Gain-setting resistors RG1 and RG2 (internal to the MAX471 and external to the MAX472) determine the ratio between the sensed current and the OUT current. These resistors should be matched to assure similar accuracy in both directions. Assuming RG1 = RG2 = RG, you can calculate the desired ratio simply by dividing the gain-resistor value into the sense-resistor value:

$$I_{OUT} / I_{SENSE} = R_{SENSE} / R_G$$

The MAX471 ratio is internally set at 0.5mA/A, which produces an output of 1.5mA when the sensed current is 3A. The MAX472 lets you select sense and gain resistors to measure virtually any level of current, provided the maximum at OUT does not exceed 1.5mA.

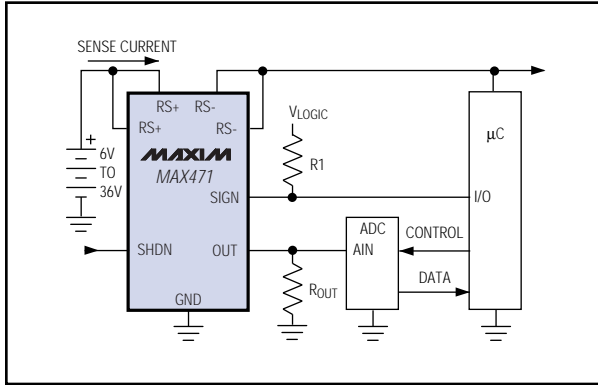


Figure 2. An A/D converter enables the µC to read the MAX471's current-source output.

The role of ADCs in current measurement

The ICs described above were designed primarily for measuring current flow from batteries. If the flow is fairly steady, you can measure the corresponding output voltage (developed across R_{OUT} in **Figure 2**) with an A/D converter. Each digital sample then represents a "snapshot" of the current flowing through R_{SENSE} . To detect the direction of current flow, you can either sample the SIGN output or connect SIGN to an I/O line on the controller.

If the measured current is not steady, you can integrate it by replacing R_{OUT} with a capacitor. Knowing that $I_{AVERAGE} = C\Delta V/\Delta t$, you can allow the capacitor to charge to a certain voltage, take a measurement, divide by the time between measurements, and then discharge the capacitor (with a shunt switch) and start over. Current flow during the capacitor-discharge interval (dead time) is lost to the measurement, but that effect is minimal if the discharge time is a small fraction of charge time.

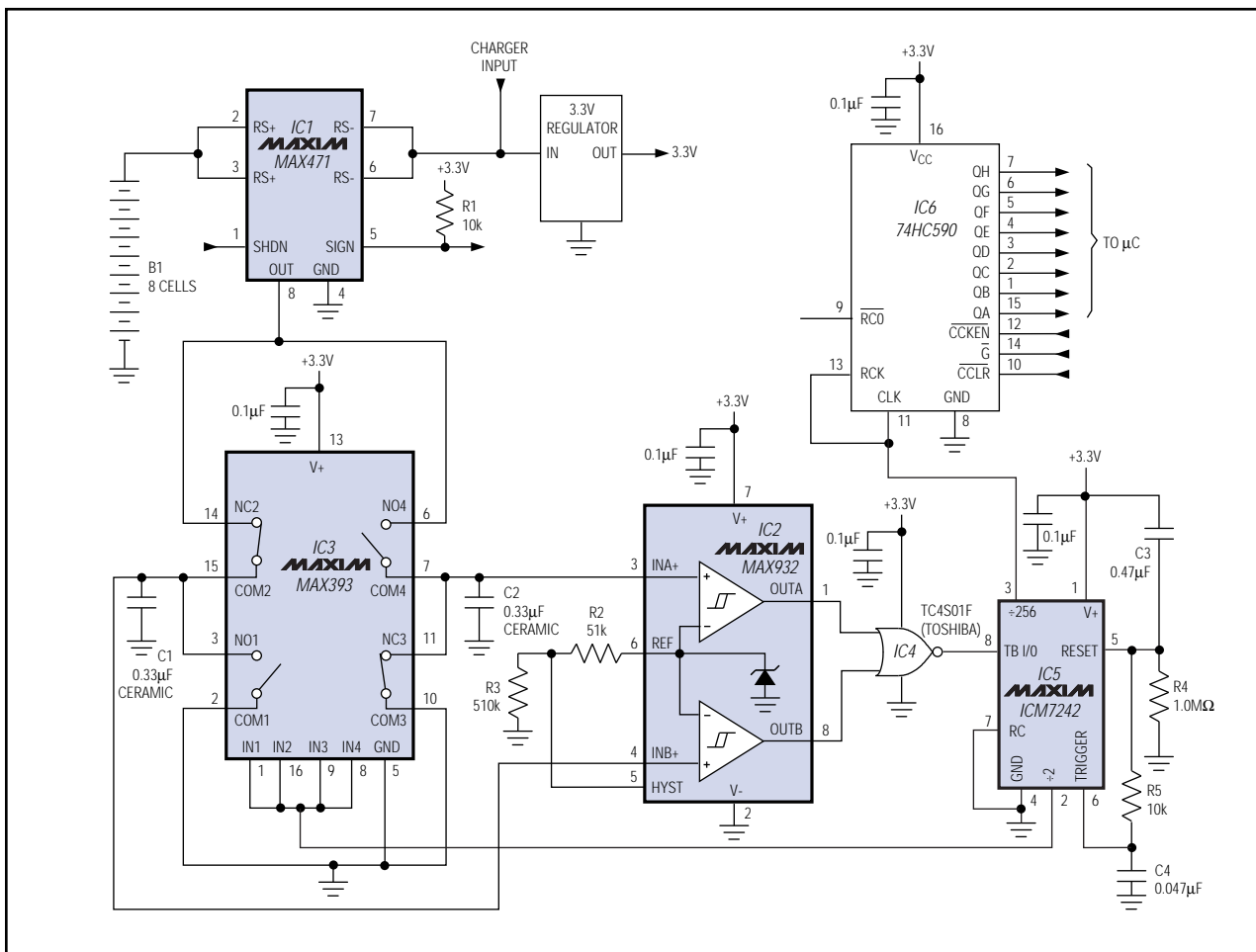


Figure 3. By switching between two integrating capacitors (C1 and C2) this battery fuel-gauge circuit monitors a continuous battery discharge with minimal error.

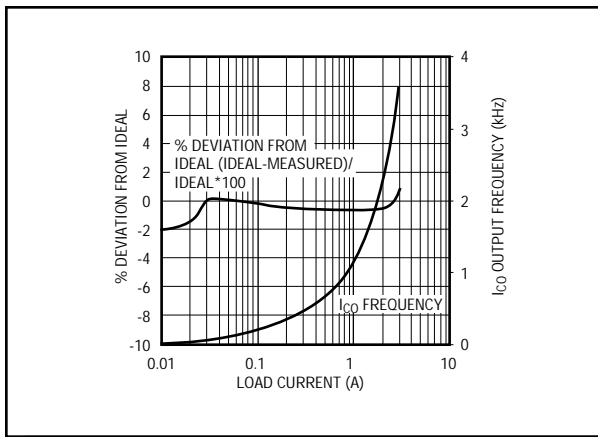


Figure 4. For low values of load current, the measurement error in Figure 3 deviates because of offset error in the current-sense amplifier (IC1).

A circuit with two charging capacitors (**Figure 3**) lets you integrate the current without creating a dead interval. The relationship $\Delta Q = C\Delta V$ lets you calculate the total charge removed from the battery or added to it. For a given comparator threshold voltage (1.182V in this case), you must balance the effects of capacitance needed (to store charge between sampling intervals), capacitor leakage, and the full-scale output current from IC3.

The Figure 3 circuit converts current to frequency and counts the resulting pulses. As an example, a full-scale current of 3A from the battery causes IC2's current-source output (pin 8) to source 1.5mA. Analog switches in IC3 steer this 1.5mA to one of two 0.33 μ F ceramic capacitors while discharging the other to ground. When the charging capacitor reaches 1.182V (the threshold voltage of dual comparator IC2) the corresponding comparator output goes high.

These low-to-high comparator transitions are summed by the NOR gate (IC4) and fed to IC5, whose divide-by-two output drives the analog switches to simultaneously discharge one capacitor and divert charging current to the other. IC5 serves as a prescaler, dividing the NOR gate output by 256 and feeding it to an 8-bit parallel counter (IC6). By periodically reading this counter's three-state outputs, the μ P can track net cumulative charge taken from the battery:

$$\Delta Q_C = C\Delta V,$$

where ΔQ_C is the maximum charge accumulated by either capacitor C, and ΔV is the corresponding capacitor voltage (i.e., 1.182V, which triggers the capacitor's discharge). Then, ΔQ_C is proportional to charge taken from the battery (ΔQ_B) by the same factor

that scales IC1's current-source output (0.5mA/A):

$$\Delta Q_C = (0.5\text{mA/A}) \times \Delta Q_B.$$

Rearranging terms,

$$\Delta Q_B = 2000\Delta Q_C = 2000C\Delta V.$$

Each negative-going transition at the NOR-gate output represents Q_B of charge from the battery, and 256 of these transitions produce one clock cycle into IC6, representing:

$$\begin{aligned} 256\Delta Q_B &= 256(2000C\Delta V) \\ &= 256(2000)1\mu\text{F}(1.182\text{V}) \\ &= 0.605 \text{ coulombs.} \end{aligned}$$

Thus, to assure a continuous monitoring of current from the battery, the system must read the three-state outputs of IC6 at least once during each of its 256-count cycles. These cycles vary according to the rate of battery discharge, but the maximum read interval is determined by the maximum anticipated discharge rate (3A in this case). The consequent 1.5mA output from IC3 charges each capacitor to 1.182V in $CV/i = (0.33\mu\text{F})(1.182\text{V})/1.5\text{mA} = 260\mu\text{s}$. This interval is multiplied by 256 in the IC5 prescaler and again by 256 in the IC6 counter: $260\mu\text{s}(256)(256) = 17$ seconds maximum. You can extend this maximum interval with a larger value for the two capacitors.

Measurement linearity remains relatively constant for load currents between 100mA and 3A. Below 100mA, measurement errors increase due to the effect of offset current in the current-sense amplifier (**Figure 4**).

Current control

Though designed primarily for current measurements, the MAX471/MAX472 amplifiers also lend themselves to current-control applications. These include current sources, overcurrent-protection circuits, and battery chargers. The circuit of **Figure 5**, for example, is a current-source battery charger capable of sourcing 2.5A with efficiency better than 90%. Again, the MAX471 senses current on the battery's high side, allowing use of a common ground as in automotive applications.

IC1 is a current-mode buck-regulator controller whose drive outputs (DH and DL) control two external n-channel MOSFETs. Their on-resistances are lower than those of equivalent p-channel MOSFETs, so the n-channel devices dissipate less power for given amounts of channel current. Positive gate drive for the high-side MOSFET (Q1) is generated by a charge pump in IC1. Buck regulators limit their maximum output to V_{IN} , so removing the battery while this charger is operating will not cause a dangerous rise in V_{OUT} .

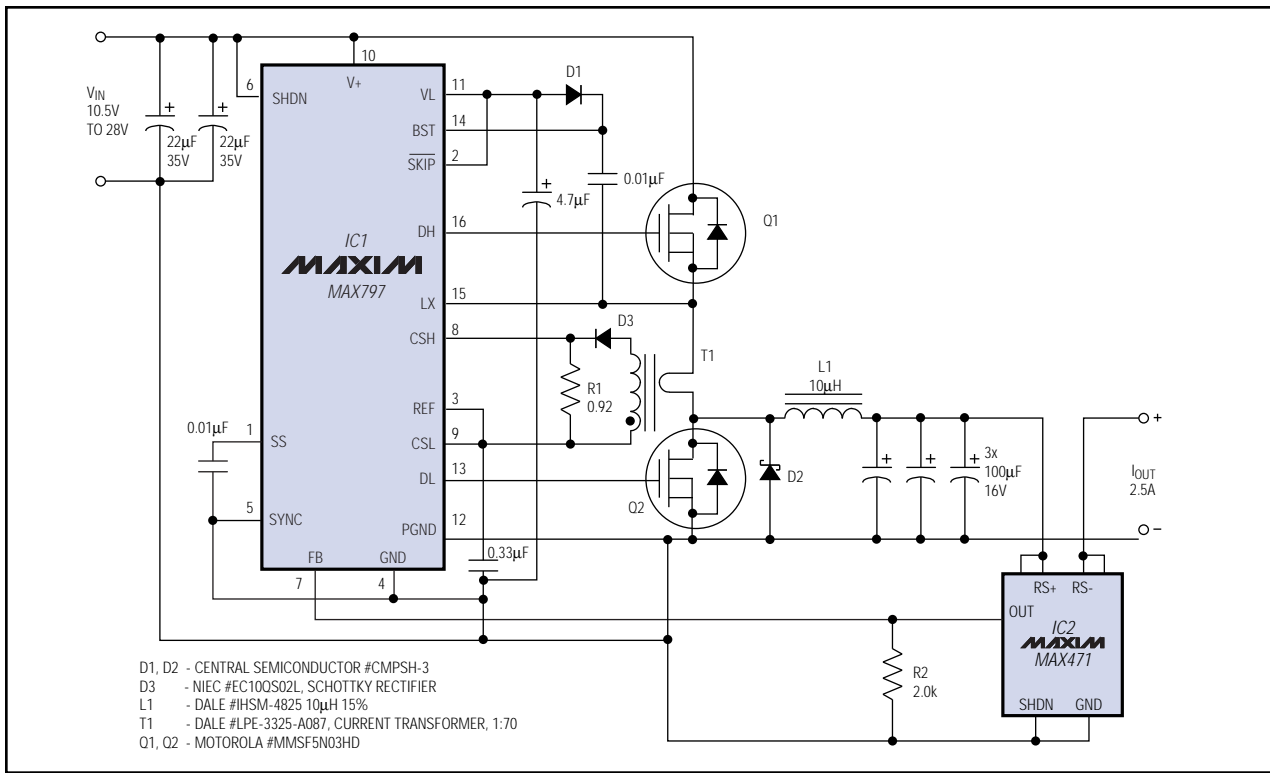


Figure 5. A high-side current-sense amplifier (IC2) enables this 2.5A, 90%-efficient, current-source battery charger to have a common ground with the battery.

The current-sense amplifier (IC2) senses I_{OUT} via an internal sense resistor and produces a smaller but proportional output current. The external resistor R2 then produces the feedback voltage required by IC1. For digital control of I_{OUT} , you can switch to other output resistors using an analog-switch array, or one or more small FETs such as the 2N7002 (not shown). On-resistance errors contributed by these FETs are not significant because IC2 produces a low output current.

Current through Q1 flows through the primary of the current transformer (T1), whose secondary directs a reduced current through the sense resistor R1. The result (compared with a conventional sense resistor in series with Q1) is lower power dissipation and better efficiency. This circuit operates with efficiency as high as 96% (Figure 6). Efficiency decreases with output voltage, because at lower voltages the fixed power level required to drive the FETs and IC represents a larger percentage of the total.

Step-up battery charger

Battery charging in many portable applications is controlled by a μP or μC (Figure 7). The processor issues CHARGE $\overline{ON/OFF}$ and $\overline{FAST/TRICKLE CHARGE}$

commands, and IC2 monitors the charging current. IC1 is a step-up switching regulator that boosts the applied 5V to a level necessary for supplying the combined charging current and system-load current. The 5V supply must include short-circuit protection for this application.

IC2's output current (pin 8) is proportional to the current through sense resistor R9, reduced by a factor of 10^{-4} (which equals the value of R9 divided by the value common to R7 and R10). Q3 and Q4 are on during a fast

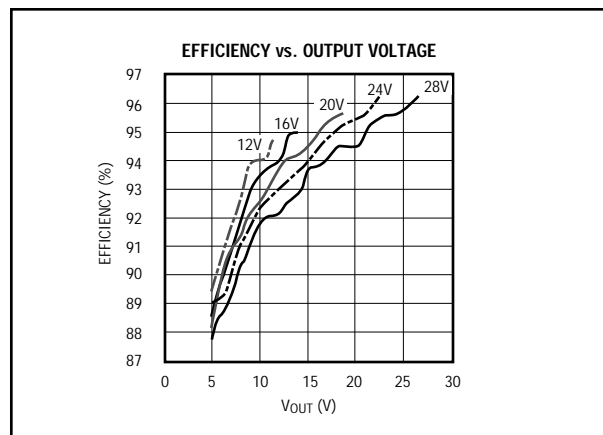


Figure 6. A fixed level of power dissipation in the battery charger of Figure 4 causes efficiency to decline with output voltage.

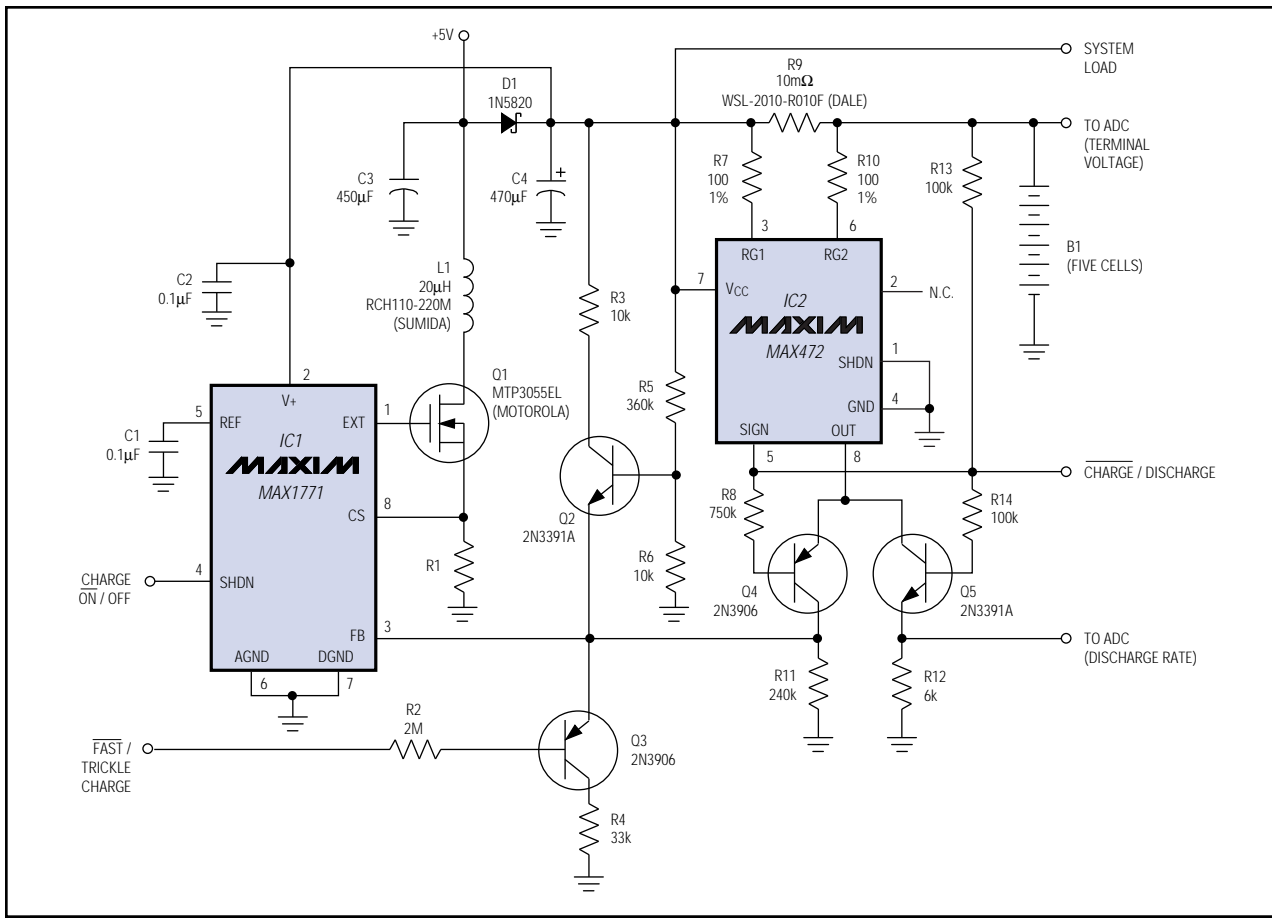


Figure 7. Under control of a microprocessor, this step-up battery charger delivers battery current and load current at the same time.

charge, so the output current flows through the parallel combination of R11 and (approximately) R4. The resulting feedback voltage to the boost converter (pin 3) maintains the R9 fast-charge current at 500mA. This feedback also enables the regulator to supply as much as 500mA of load current in addition to the charging current. Q2 limits the battery voltage to 10V (2V per cell).

An external processor and multi-channel A/D converter monitor the battery's terminal voltage during a fast charge. When the A/D senses a change of slope in this voltage, the processor terminates charging by asserting a high on $\overline{\text{FAST/TRICKLE CHARGE}}$. Q3 turns off, causing a rise in the feedback (to IC1) that lowers the charging current to the trickle-charge rate of approximately 60mA.

If the boost converter shuts down, or if the sum of load current and charging current exceeds the boost converter's output capability, the charging current reverses as current flows out of the battery. IC2 indicates this reversal via its open-collector SIGN output, pulled high by R13, which turns off Q4 and turns on Q5.

Current through R12 then produces a voltage proportional to the battery's discharge current: 5A through R9 produces a full-scale response of 3V across R12.

By integrating this voltage over time (sampling at fixed intervals and multiplying each sample by the associated time interval), the system can monitor energy removed from the battery. Based on this measurement and the terminal-voltage measurement, the processor can then re-initiate a fast charge (by asserting $\overline{\text{FAST/TRICKLE CHARGE}}$ low) before the battery reaches its end of life.

Switched, digitally controlled current source

The variable current source of **Figure 8** generates 0A to 5A outputs with a compliance range of 4V to 30V. It has two advantages over conventional current sources: the 12-bit D/A converter (IC2) makes it digitally programmable, and its switch-mode step-down regulator (IC1) is more efficient than the alternative linear pass transistor. Applications include battery charging and dc motor control.

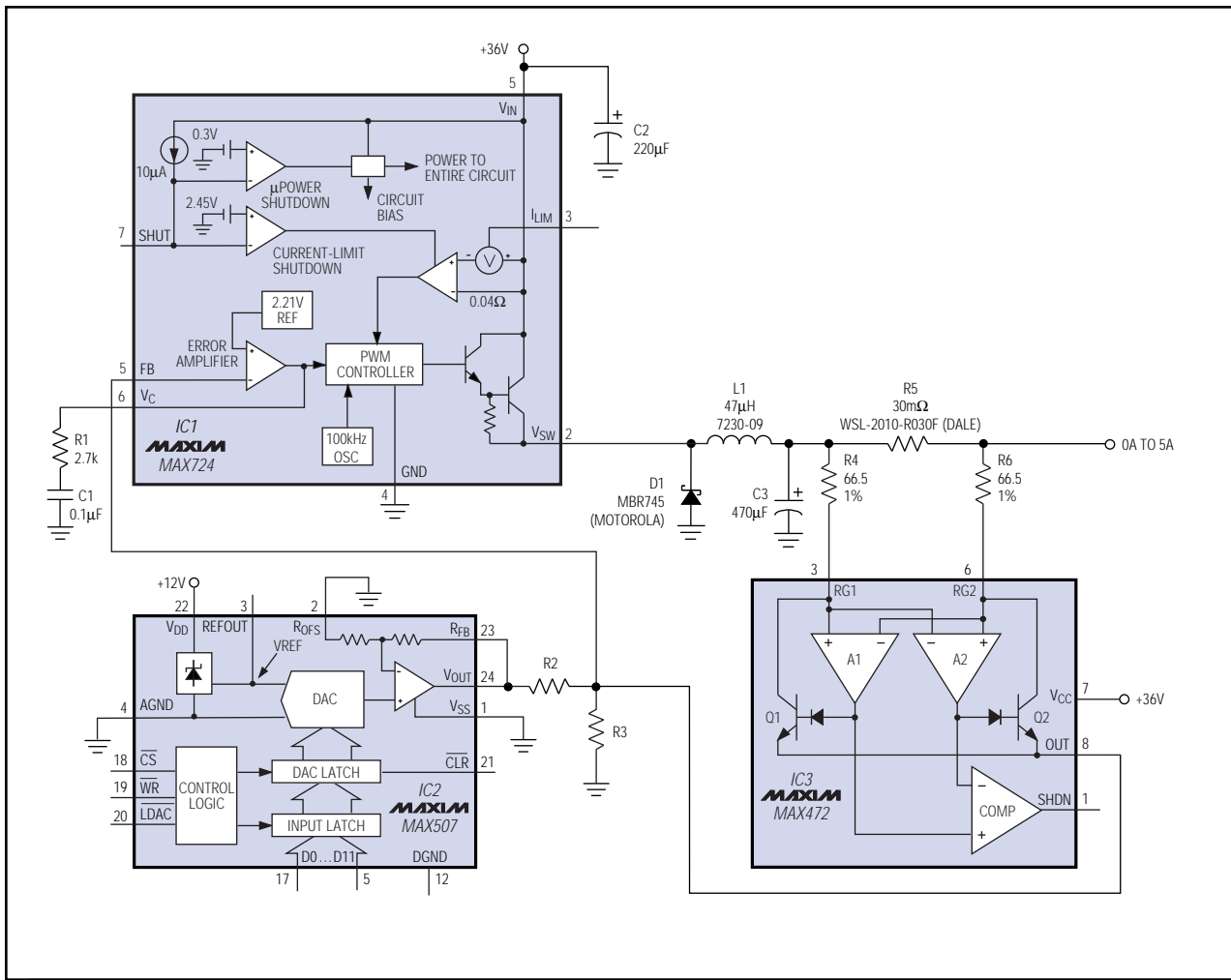


Figure 8. This 0A to 5A variable current source features an efficient, switch-mode step-down regulator (IC1). A 12-bit D/A converter (IC2) makes the source digitally programmable.

IC3 (the current-sense amplifier) senses output current as a voltage drop across R5 and produces a proportional signal current at pin 8. Thus, the regulator's feedback voltage (at pin 1 of IC1) is set by the D/A converter and modified by IC3's current feedback, which flows across the parallel combination of R2 and R3. This current feedback opposes any change in load current due to a change in load resistance.

The D/A converter generates 0V to 10V, producing a source current that varies inversely with code: FFF_{HEX} (10V from the DAC) produces 0mA, and 000_{HEX} (0V from the DAC) produces 5A. For a given programmed level, the actual output varies somewhat with load resistance and the corresponding compliance voltage. When tested at 1.5A, for instance, the output of the circuit deviates about 15mA (i.e., 1% of 1.5A) for compliance voltages between 10V and 20V (Figure 9).

You can easily reconfigure the circuit for other ranges of output current (I_{SOURCE}) by resizing R2 and R3:

$$I_{SOURCE} = \frac{2217 [(V_{FB} (R2 + R3) - R3V_{DAC})]}{R2R3},$$

where $V_{FB} = 2.21V$ and V_{DAC} ranges from 0V to 10V. Values for R2 and R3 are defined by the desired range for I_{SOURCE} : $V_{DAC} = 0V$ for the high value of I_{SOURCE} , and $V_{DAC} = 10V$ for the desired low value of I_{SOURCE} . Substituting these two sets of values in the equation yields two equations, to be solved simultaneously for the values of R2 and R3.

Variable linear current source

By converting current to a feedback voltage you can transform a low-dropout linear voltage regulator into a current regulator (Figure 10). The control input

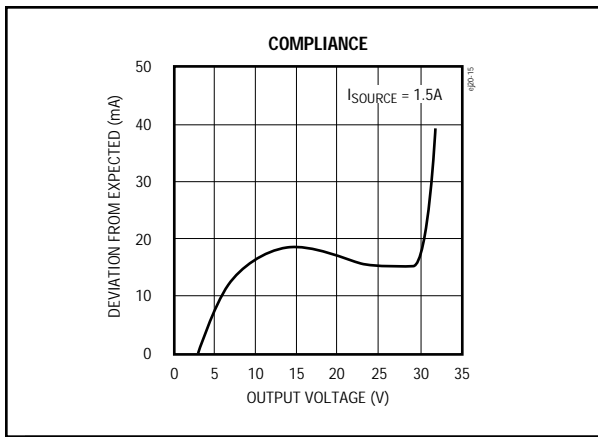


Figure 9. The Figure 8 current source varies only 2mA or so over the compliance range 10V to 20V.

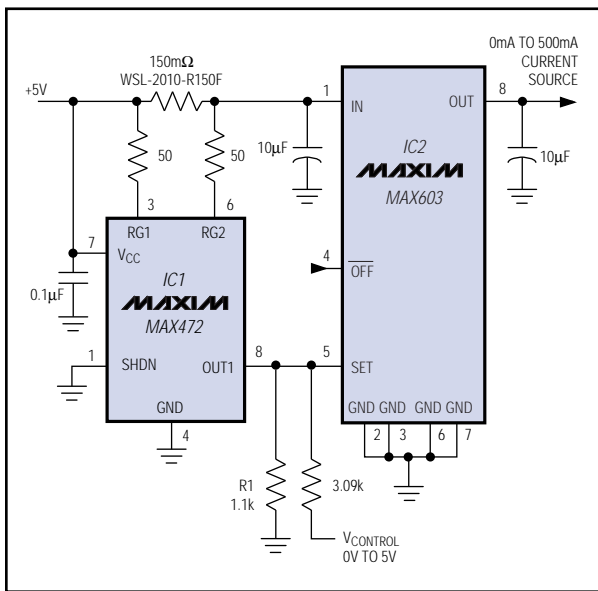


Figure 10. This current source (similar to that of Figure 8) varies linearly from 0mA to 500mA in response to the applied control voltage.

$V_{CONTROL}$ determines the output current: applying 5V sets this current to zero, and applying 0V sets it to 250mA. Intermediate voltages from a D/A converter (or the buffered output of a potentiometer) let you control the output current digitally or manually.

This circuit has a compliance range of 0V to 4.7V when powered from 5V, but it also operates from supply voltages as high as 11V, if you take care not to exceed the maximum power-dissipation rating for the regulator package. This rating is 1.8W at room temperature. Ignoring a tiny operating current, the chip dissipation equals the voltage drop from pin 4 to pin 6 times the current through the chip via these pins.

Under the worst-case conditions of 250mA output current, output shorted to ground, and a dissipation limit of 1.5W (the 1.8W package rating less a safety factor), the circuit's input voltage can be as high as 6V. ($6V \times 250mA = 1.5W$.) Excessive power dissipation causes the thermal-protection circuitry to turn on and off, producing a pulsed output current as the internal temperature oscillates about the thermal trip point.

Polarity-reversal/ forward-current protection

The Figure 11 circuit protects a battery-operated system in two ways: Q1 prevents damage due to the flow of reverse current that otherwise occurs when the battery is installed backwards, and Q3 prevents the excessive flow that otherwise occurs with a sudden load increase or short circuit.

A properly installed battery fully enhances Q1 by pulling its gate more than 5V below the source. If the battery is installed backwards, Q1 is off because the gate is positive with respect to the source. Regardless of battery polarity, the body diodes of Q1 and Q3 are oriented to assure that no current can flow when either device is off. Both FETs have low on-resistance.

Current-sense amplifier IC2 senses the load current flowing between its RS+ and RS- terminals and develops a proportional voltage across R8. During normal operation, the comparator outputs are high and Q3 remains on.

When the load current exceeds a limit set by R8 (i.e., $I_{LIMIT} = 2000V_{TH}/R8$, where 2000 is the sense amplifier's gain and V_{TH} is the comparators' input threshold ($1.182V \pm 2\%$), the B comparator output goes low, turns off Q4, turns off Q3, and disconnects the battery from its load. At the same time, Q6 provides positive feedback by pulling the comparator input up to the collapsing supply rail, latching Q3 off as the supply voltage drops.

A short-circuited output turns off IC2 by removing the voltage at pins 6 and 7 (3V is the minimum for proper operation). IC1's B comparator loses control because the R8 voltage goes to zero, but comparator A is able to shut off Q3 by turning off Q5. Q2 speeds the Q3 turn-off time to about 10μs, and when Q3 is off, the circuit draws about 2μA. (To restore power, you press S1.) During normal operation the battery current varies with its terminal voltage: 200μA at 5V, 230μA at 6V, 300μA at 8V, and 310μA at 10V.

(Circle 1)

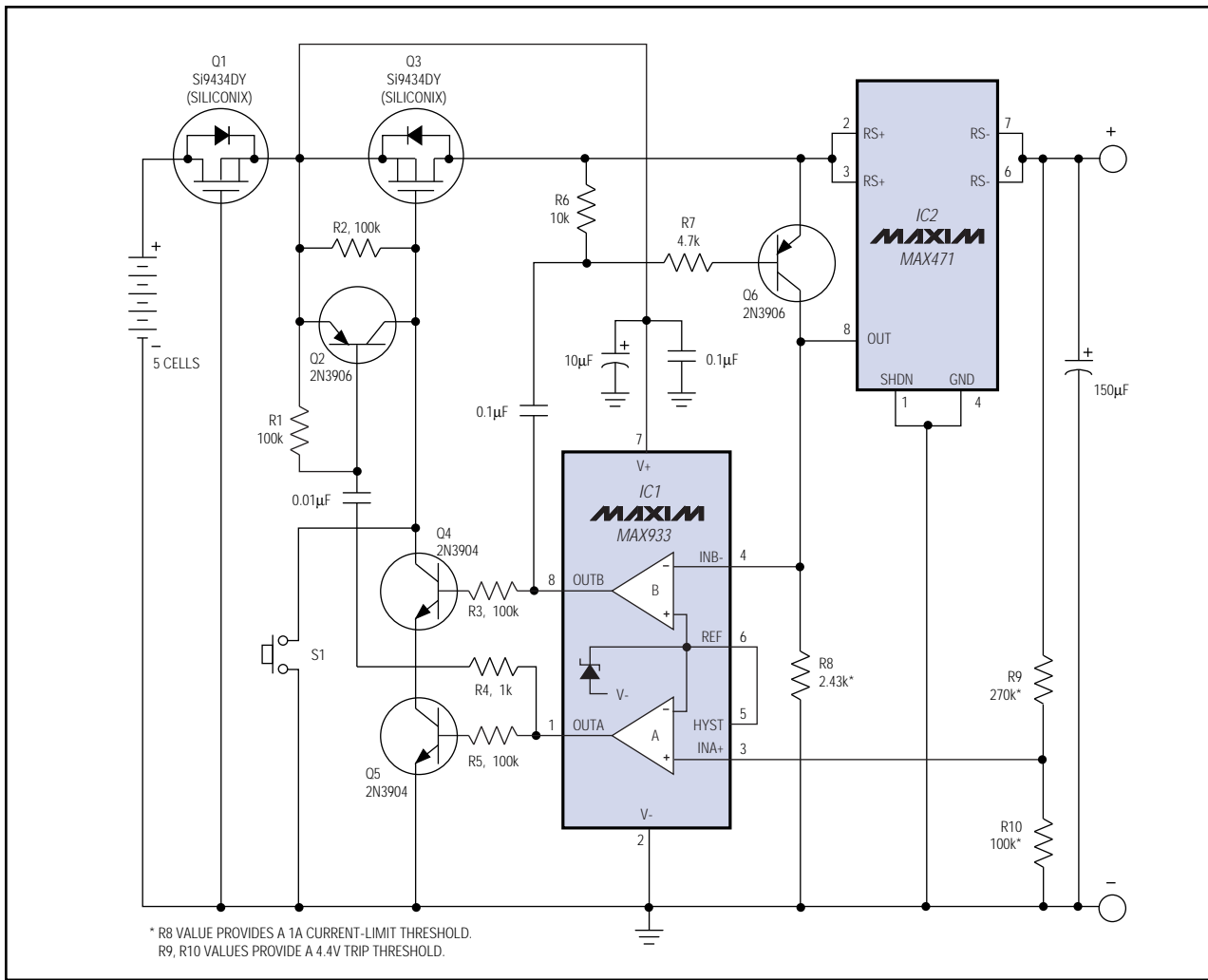


Figure 11. This load-protection circuit prevents current flow in response to excessive loads, output short circuits, and wrong-polarity connections to the battery.

DESIGN SHOWCASE

Three-input supply powers 3.3V portables

The single-ended primary-inductance converter (SEPIC) of **Figure 1** accepts input voltages ranging from 3V to over 6V, and produces a regulated 3.3V, 200mA output. It accepts an input voltage from three sources: a 5V dc wall cube, a 3-cell AA battery, and a lithium backup battery.

Unlike conventional boost regulators, whose battery current continues to flow during shutdown (unless you add a cutoff switch), this circuit's output turns fully "off" in response to a shutdown command. And, unlike flyback-transformer regulators and combination step-up/linear regulators, the Figure 1 circuit requires no transformer. Coils L1 and L2 should be the same type and have the same value, but coupling between them is not required. They can be wound on the same core for convenience, but the circuit works equally well if they are completely separate.

Capacitor C3 couples energy to the output and requires low ESR to handle the high ripple currents. Conversion efficiency with a low-ESR Sanyo OSCON capacitor is 85%, which is 3% higher than that

available with less-expensive 1 μ F ceramic capacitors. Tantalum capacitors are not recommended because high ripple currents cause them to self-heat.

During normal operation, the ac adapter's 5V output powers the circuit and turns off Q1. Disconnecting the adapter removes 5V, turns on Q1, and allows the three AA cells to provide power. If the 3.3V output drops below 3.0V, a low-battery comparator in IC1 alerts the system by driving LBO low. And for backup, a diode-OR connection allows an optional lithium battery (coin cell B2) to provide load current at the 3.3V output.

As an added twist, diode D2 provides a supply voltage for IC1 (pin 8) by capturing the switching pulses at LX (pin 7). This voltage (approximately the sum of V_{IN} and V_{OUT}) improves start-up capability under full load, and improves low- V_{IN} efficiency by boosting gate drive to the internal switching MOSFET. Maximum V_{IN} is limited to about 12V, which easily accommodates the 3-cell battery.

A related idea appeared in the 4/13/95 issue of EDN.

(Circle 2)

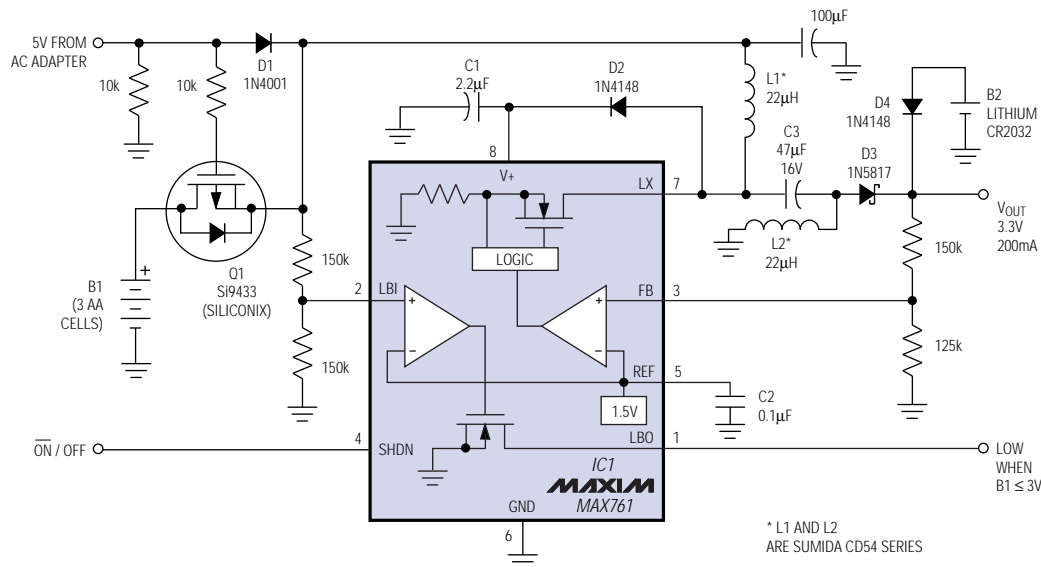


Figure 1. When the wall cube is disconnected, this SEPIC regulator draws power from a 3-cell battery. A lithium battery (optional) backs up the 3.3V output.

DESIGN SHOWCASE

Latching regulator prevents deep discharge of battery

A rechargeable battery's load should be removed at the point of complete discharge, to avoid a further (deep) discharge that can shorten its life or destroy it altogether. Because a battery's terminal voltage recovers when its load is removed, you can't simply disconnect the load when the terminal voltage dips below the established threshold and then re-connect it when the voltage returns above that threshold. Such action may produce chatter in the disconnect switch.

The voltage of a discharged cell returns almost to the level of a fully charged cell, so hysteresis can't necessarily compensate for the recovery effect either. What's needed is a circuit that disconnects the load from the battery and keeps them separate until an external signal (such as that from a battery charger or pushbutton switch) indicates that the battery has been recharged or replaced.

Such a circuit can enlist the low-battery comparator in a low-dropout linear regulator (Figure 1). In this circuit, the low-battery comparator and error amplifier share the internal reference and the external resistor divider. With the resistor values shown, the

low-battery output (LBO) goes low and disconnects both the battery and load when the output falls eight percent below its nominal value. The battery and load then remain disconnected until commanded otherwise by S1.

Two factors enable the latching action in this circuit: the low-battery comparator remains active during shutdown (most regulators deactivate this comparator during shutdown), and the circuit monitors the regulated output voltage instead of the battery voltage (regulator voltage can't recover until the regulator is turned back on).

The circuit also provides a $\overline{\text{POWER FAIL}}$ signal (LBO, pin 1) that goes low 50ms before the output is turned off (Figure 2). This signal can provide a controlling microprocessor time to perform house-keeping and shutdown functions. When LBO goes low, C1 discharges through R3 until the $\overline{\text{STBY}}$ input reaches its threshold (1.15V). The IC then enters its standby mode and disconnects the battery. IC1 is a linear regulator capable of sourcing 150mA with a 350mV dropout voltage. It has a 10 μ A standby current and accepts input voltages to 11.5V.

A related idea appeared in the 3/16/95 issue of EDN.

(Circle 3)

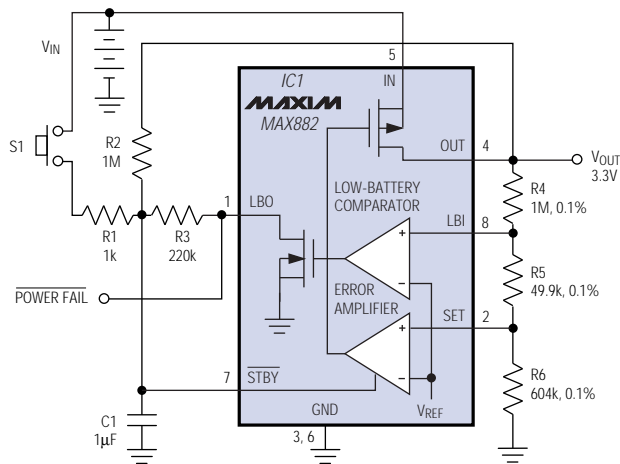


Figure 1. To protect the battery, this circuit disconnects the load before the battery enters deep discharge. To reconnect, you must press S1.

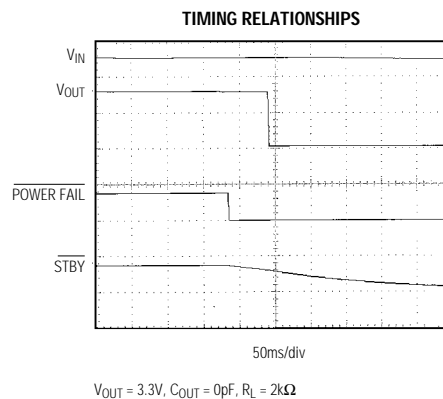


Figure 2. These waveforms illustrate timing relationships in the circuit of Figure 1.

DESIGN SHOWCASE

Notch filter is dc-accurate

Most active filters exhibit noise, distortion, gain error, and dc offset, but a filter topology that separates the dc and ac paths can eliminate the last two of these unwanted behaviors completely (**Figure 1a**).

The dc path has no op amps, and therefore no dc offset. Nor does it have a dc gain error, other than -6dB of attenuation caused by the R_{IN}/R_{LOAD} divider. (This attenuation is absent for applications that omit the R2 termination.) The ac path consists of a capacitor (C1) and a synthetic inductor composed of two wideband transconductance amplifiers (WTAs) and their associated components. The result is an active circuit that emulates the passive filter of **Figure 1b**.

Simulating the inductance avoids the use of an actual inductor, which can (among other problems) act as a transmitting and receiving antenna for EMI. The equivalent inductance L_{EQ} is $C/[(gm1)(gm2)]$, where $gm1$ and $gm2$ are transconductances produced by IC1 and IC2. The inductance value can be large if $(gm1)(gm2) \ll 1$, but one end of the network must always connect to ground. Each gm is set by an

external resistor (R3 or R5) according to the relationship $gm = 8/R$.

For optimum noise performance, the gm values should allow a full range of output swing for each WTA. You should start with equal gm values, and simulate the filter in Spice using “g” elements for the amplifiers. Observe the peak voltage amplitude at each WTA output while sweeping the frequency at least one decade above and below the filter’s corner frequency (3.2kHz in this case).

The peak value across the inductor (at pin 13 of IC2) is demanded by the filter and cannot be changed, so you adjust the peak value at IC1’s pin 13 to match. Let K equal the ratio of these peak values ($VO1_{pk}/VO2_{pk}$). Gain is proportional to transconductance, so divide $gm1$ by K and multiply $gm2$ by K. Then, rerun Spice with the new gm values to verify that the peaks are equal and the filter shape has not changed.

The filter—consisting of the source/load connection shunted by the series network of C1 and the

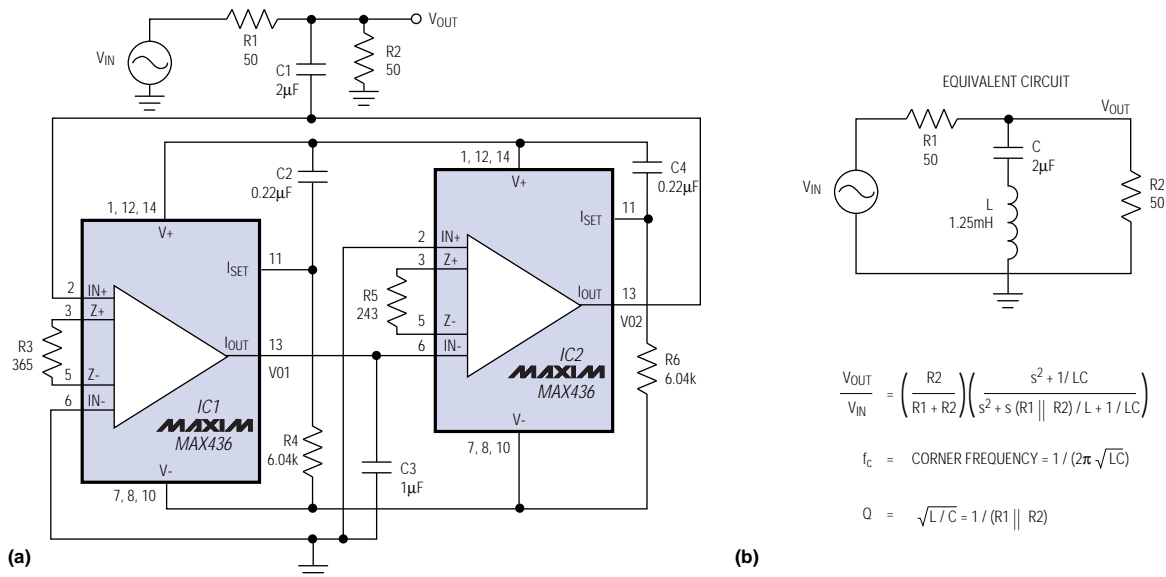


Figure 1. The circuit below C1 constitutes a synthetic inductor that is part of a dc-accurate notch filter (a). The equivalent passive filter is shown in b.

$$\frac{V_{OUT}}{V_{IN}} = \left(\frac{R2}{R1 + R2} \right) \left(\frac{s^2 + 1/LC}{s^2 + s(R1 \parallel R2)/L + 1/LC} \right)$$

$$f_c = \text{CORNER FREQUENCY} = 1 / (2\pi \sqrt{LC})$$

$$Q = \sqrt{L/C} = 1 / (R1 \parallel R2)$$

synthetic inductor—was tested on a network analyzer that includes the 50Ω R_{IN} and R_{LOAD} resistors. At the 3.2kHz corner frequency, it produces a 2nd-order notch response whose rejection (ideally infinite) is about 40dB (**Figure 2**).

High-frequency error is dominated by parasitic capacitance between the output of the synthetic inductor and ground. Though small, this error increases as the parasitic reactance approaches the parallel combination of the source and load resistances. To minimize error in the frequency response, you should keep these resistances small with respect to the WTA's $3k\Omega$ output impedance.

A related idea appeared in the 3/2/95 issue of EDN.

(Circle 4)

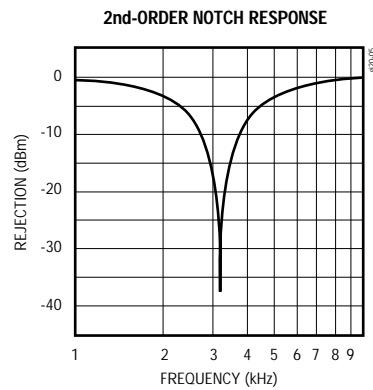


Figure 2. This 3.217kHz, 2nd-order notch response is produced by the Figure 1a circuit.

DESIGN SHOWCASE

Linear regulator converts 3.3V to 2.9V

Linear regulators (vs. switching types) are often the best choice for generating supply voltages below 3.3V. At low output voltage and moderate load currents, the linear type is less expensive and requires less space, yet maintains reasonable efficiency. The circuits of **Figure 1**, for example, have efficiencies higher than 87%.

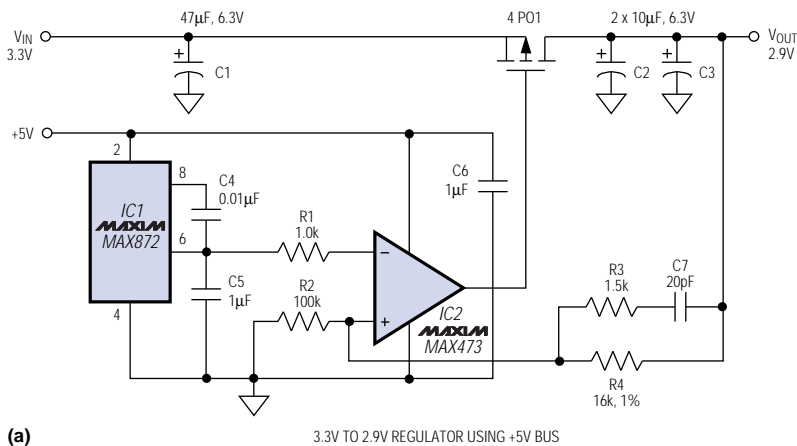
Dropout voltage for the available “low-dropout” linear regulators may not be low enough for the load current desired. The Figure 1 circuits, therefore, employ low-threshold p-channel MOSFETs that

drop only 300mV at 2A. If available, a 5V bus should power the op amp and the 2.5V reference (**Figure 1a**). The higher rail voltage improves transient response by enabling the op amp to drive the MOSFET faster. (Power-management circuitry in the newer processors can change load current within tens of nanoseconds.)

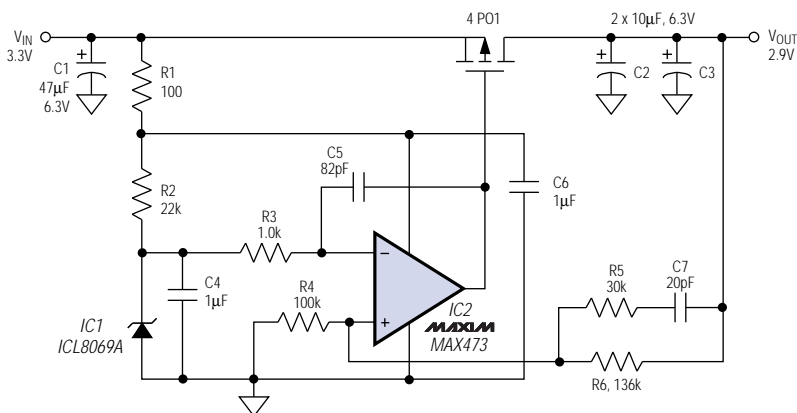
The 10MHz op amp allows common-mode inputs from ground to within 1.9V of the positive rail, so 3.3V operation allows insufficient room for a 2.5V reference. With a 1.2V reference (**Figure 1b**) the 3.3V circuit provides $\pm 3\%$ initial output accuracy and $\pm 5\%$ regulation (for the fixture tested) over the allowed extremes for line, load, temperature, and supply-voltage transients. The Figure 1a circuit measured $\pm 4\%$ for the extremes and $\pm 2\%$ for initial accuracy. Both circuit outputs change less than 1mV for a 5% V_{IN} change, and only 3mV for a 0A to 2A load change (see **Figure 2** for the effect of a 1A to 2A load change).

Layout is critical if the circuits are to accommodate transient rise times faster than 100ns. The regulator should be close to the load, and V_{OUT} should be sensed at the load. The op amp and reference should have a single-point ground to prevent ground bounce and noise from upsetting the feedback loop.

Because the initial edge of any transient will be absorbed by the 10 μ F output capacitors, these components (such as the Sanyo OS-



(a) 3.3V TO 2.9V REGULATOR USING +5V BUS



(b) STAND-ALONE 3.3V TO 2.9V REGULATOR

Figure 1. These linear regulators generate 2A at 2.9V, either with the help of a 5V bus (a) or from 3.3V alone (b).

CON types shown) must have very low ESR. To minimize the effect of lead inductance in surface-mount capacitors, you can reduce the lead lengths virtually to zero by routing output current right through the capacitors' metallized contacts.

A related idea appeared in the 3/16/95 issue of EDN.

(Circle 5)

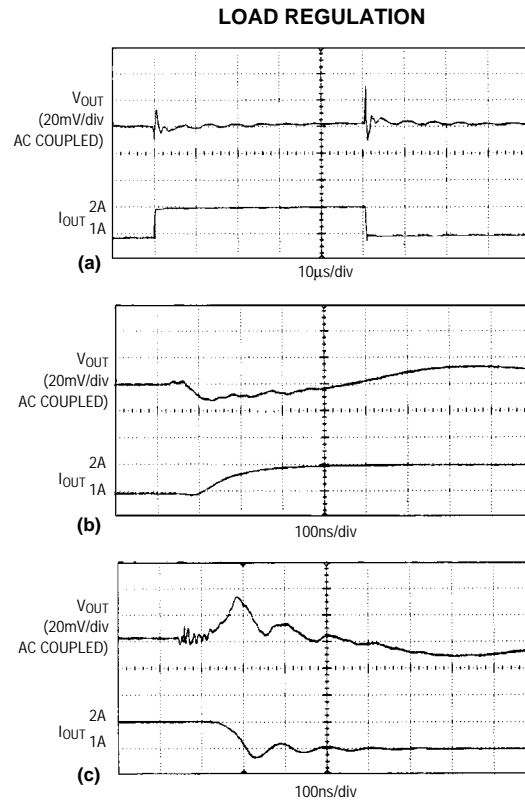


Figure 2. A step change in load current (a) for the Figure 1 circuits is stretched from 10µs to 100ns per division for the 1A to 2A case (b) and the 2A to 1A case (c). Vertical resolution is 20mV/div for the output voltage (top traces) and 1A/div for the load current (bottom traces).

DESIGN SHOWCASE

Data multiplexer adds cursor to LED-display driver

A data multiplexer (IC2) enables the LED-display driver in **Figure 1** to highlight any selected digit by intensifying its brightness. This cursor function enables use of the display for data entry as well as readout.

The operator of an environmental chamber, for instance, can set a desired temperature by entering that value via the display. Four “buttons” are required, but are not shown: left/right for selecting the digit to be intensified, and up/down for changing the digit’s value (the buttons can be read by a microcontroller). Driving $\overline{\text{CURSOR}}$ high then removes the cursor effect (by disabling IC2), and allows the display to continue monitoring temperature as it moves toward the new set point.

IC1 controls as many as eight 7-segment digits (eight segments including the decimal point) by scanning them sequentially and producing a value for each according to data stored in the chip via its serial interface. Each segment-driver output is a current source that delivers approximately 100 times the current entering I_{SET} (pin 18). Thus, you can change the brightness of a given digit by altering the I_{SET} current as that digit is scanned. (IC1 also provides a 16-level digital brightness control via 4-bit pulse-width modulation of the segment currents.)

The corresponding segments in each digit are bused together externally. The cathodes of all LED segments in a digit are bused together internally, allowing the digit to be turned on by drawing current from its common node with a logic-zero signal. Unselected digit-driver lines remain high. The digit-scan rate is approximately 1300/second.

When you apply a 3-bit digit-select code to IC2 while driving $\overline{\text{CURSOR}}$ low, the multiplexer connects the corresponding digit signal to terminal Y and its complement to terminal W (pin 6). Thus, selecting a particular digit for cursor intensification drives W high during that digit scan, placing R1 and R3 in parallel and driving more current into I_{SET} . (When W is low, R3 robs current from I_{SET} .) If $\overline{\text{CURSOR}}$ remains high, the digits exhibit uniform maximum brightness because W remains high for all of them.

Each of IC1’s eight digit-driver outputs can sink LED currents as high as 320mA, but these outputs remain logic-compatible with the digital inputs of IC2. Even at 320mA, the digit-driver output voltages remain below the multiplexer’s guaranteed low-level input voltage (V_{IL}).

A related idea appeared in the 3/30/95 issue of EDN.

(Circle 6)

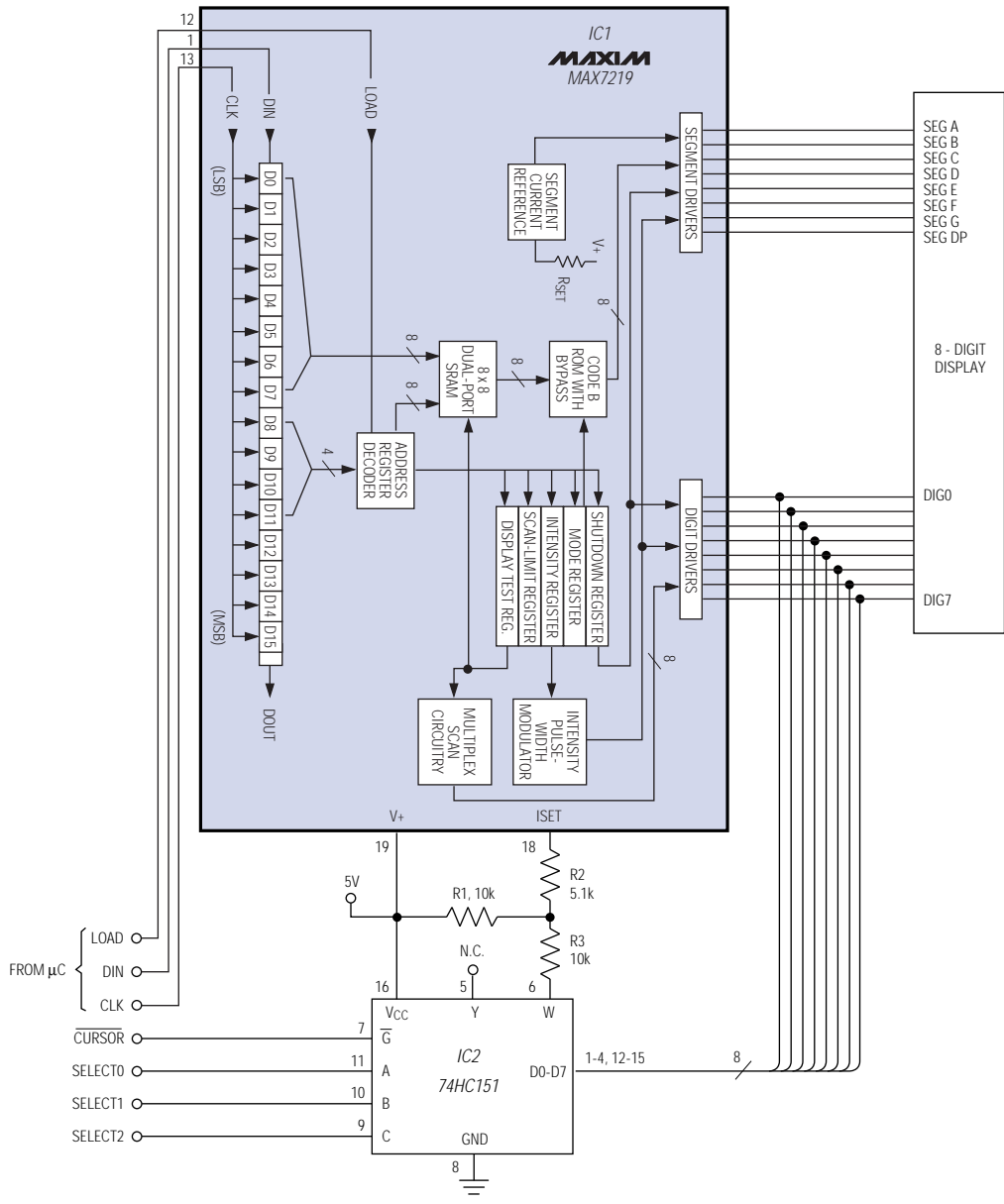


Figure 1. The digital multiplexer in this 8-digit display (IC2) provides a cursor function that intensifies the selected digit.

NEW PRODUCTS

Two-wire, octal DAC saves μ P I/O lines

The MAX521 is the first available octal DAC with a 2-wire serial interface. Containing eight 8-bit voltage-output D/A converters, the MAX521 minimizes I/O lines to the μ P and simplifies signal routing by allowing multiple DACs and other devices (including 2-wire EEPROMs) to share the same 2-wire bus. The bus is compatible with conventional microprocessors and the I²C™ serial-data standard.

Operating from a single 5V supply, the MAX521 is ideal for digital adjustments of offset and gain. Its eight voltage outputs

swing rail to rail, and its five independent reference inputs (with ranges that include the positive rail) allow the outputs to assume different full-scale levels. The device has a 10 μ A shutdown mode and a power-on reset that sets all outputs to zero. Double buffering allows all addressed DACs to be updated simultaneously, and the serial interface with internal software protocol allows data rates to 400kbps.

Other 8-bit, 2-wire D/A converter ICs, with performance similar to that of the MAX521, are the following: the dual-DAC MAX519 (two reference inputs), the dual-DAC MAX518 (8-pin SO package), and the single-DAC MAX517. The quad-DAC MAX520 (ultra-low 100 μ A maximum

supply current) will be available after October 1995.

The MAX521 comes in 20-pin DIPs and in space-saving 24-pin SO and SSOP packages (the SSOP is the smallest package available for an octal DAC). The MAX520 and MAX519 come in 16-pin DIP and SO packages, and the MAX518 and MAX517 come in 8-pin DIP and SO packages. All are available in versions tested for the commercial (0°C to +70°C) and extended-industrial (-40°C to +85°C) temperature ranges (contact the factory for MIL grades). Starting prices (1000 up, FOB USA) range from \$2.10 for the MAX517 to \$4.95 for the MAX521.

I²C is a trademark of Philips Corp. (Circle 7)

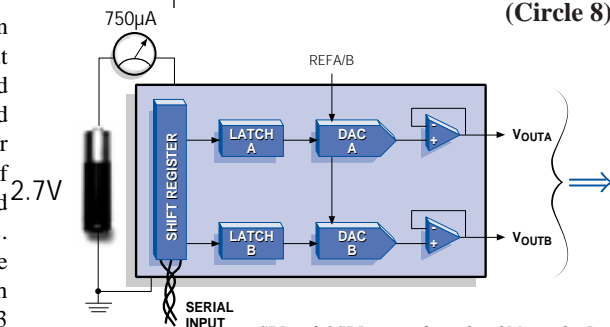
Serial-input, dual 8-bit DAC comes in SO-8 package

The MAX522 is a dual 8-bit, low-power D/A converter whose ultra-low supply current (1mA operating; 1 μ A shutdown) and space-saving 8-pin SOIC and DIP packages are ideal for portable and battery-powered applications. Operating from a single supply voltage in the range 2.7V to 5.5V, it has buffered voltage outputs and a 5MHz, 3-wire serial interface compatible with the SPI™, QSPI™, and Microwire™ interface standards.

Maximum source and sink currents are 5mA for the "DAC A" buffer and 0.5mA

for the "DAC B" buffer. Each output swings within 0.5V of ground and V_{CC}. To help reduce the shutdown power consumption, the DAC's reference input is internally disconnected from the V_{REF} pin during shutdown.

The serial input feeds an internal shift register that holds eight data bits and four control bits (load and shutdown commands for each DAC). A rising edge of \overline{CS} then loads the 8-bit word into either or both DACs. (The MAX522's software format is compatible with the MAX512 and MAX513 triple 8-bit DACs.)



(Circle 8)

The MAX522 comes in 8-pin plastic DIP and SO packages, in versions tested for the commercial (0°C to +70°C) and extended-industrial (-40°C to +85°C) temperature ranges. Prices start at \$2.25 (1000 up, FOB USA).

8-channel muxes withstand overvoltage to 40V beyond supply rails

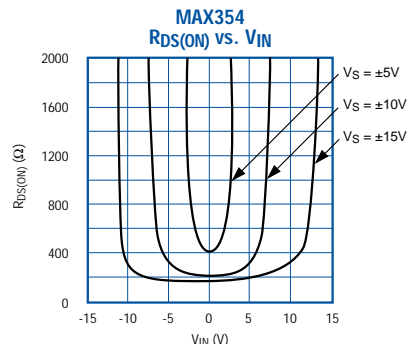
The MAX354 fault-protected analog multiplexer has eight single-ended channels, and the MAX355 has four differential channels. For both, the eight internal switches each consist of an n-channel, p-channel, and n-channel MOSFET in series—a structure that opens automatically in response to fault voltages. The result is overvoltage protection that operates to 40V beyond either supply rail during power-up, power-down, and fault conditions. It also protects sensitive external circuitry by clamping the output voltage near the nominal supply rail.

On-resistances are 350 Ω maximum, with symmetry that enables demultiplexing as well as multiplexing applications. Input leakage currents are less than 0.5nA at +25°C and less than 5nA at +85°C. MAX354/MAX355 multiplexers operate either from a single supply of 4.5V to 36V or dual supplies of \pm 4.5V to \pm 18V. Power consumption is less than 1.5mW. All digital inputs have 0.8V and 2.4V thresholds that ensure TTL/CMOS compatibility without pull-up resistors. Break-before-make switching action is guaranteed.

MAX354/MAX355 multiplexers are improved pin-compatible replacements for MAX358/MAX359, DG508A/DG509A, and DG458/DG459 devices; they are also pin-compatible with ADG508F/ADG509F multiplexers. The MAX354 and MAX355

come in 16-pin DIP and wide-SO packages, in versions tested for the commercial (0°C to +70°C), extended-industrial (-40°C to +85°C), and military (-55°C to +125°C) temperature ranges. Prices start at \$2.45 (1000 up, FOB USA).

(Circle 9)



NEW PRODUCTS

Quad, SPDT analog switch has low on-resistance

The MAX394 is a new quad SPDT analog switch fabricated with Maxim's low-voltage silicon-gate process. It operates from a single supply in the range 2.7V to 15V, or a bipolar supply in the range $\pm 2.7V$ to $\pm 8V$. The four independent switches have low on-resistances (35Ω maximum, 17Ω typical), guaranteed to match within 2Ω . Each channel is guaranteed flat within 4Ω over the analog signal range.

MAX394 switches turn on in less than 130ns and turn off in less than 75ns, with typical break-before-make intervals of 10ns. Design improvements guarantee extremely low charge injection (10pC), low power consumption ($10\mu W$), and the ability to withstand electrostatic discharge (ESD) in excess of 2000V. Low quiescent current ($>1\mu A$ with the inputs high or low) makes the MAX394 ideal for portable applications.

Logic inputs are TTL and CMOS compatible, and guaranteed to switch in the range 0.8V to 2.4V for supply voltages of 8V and below. For supply voltages above 8V, the switching range is 0.8V to 4V. Logic inputs and the switched analog signals can range anywhere between the supply rails without damage.

The MAX394 is pin compatible with the MAX333. It comes in 20-pin DIP, SSOP, and wide-SO packages, in versions tested for the commercial ($0^\circ C$ to $+70^\circ C$), extended-industrial ($-40^\circ C$ to $+85^\circ C$), and military ($-55^\circ C$ to $+125^\circ C$) temperature ranges. Prices start at \$3.19 (1000 up, FOB USA).

(Circle 10)

8-channel muxes have latchable address inputs

The MAX382 analog multiplexer has eight single-ended channels, and the MAX384 has four differential channels. Both are CMOS devices with latchable, TTL/CMOS-compatible channel-address inputs. They operate from a single supply of 2.7V to 16.5V or dual supplies of $\pm 3V$ to $\pm 8V$. Power consumption is less than $10\mu W$.

MAX382/MAX384 multiplexers guarantee low charge injection (10pC maximum) and less than 2.5nA of off leakage at $+85^\circ C$. On-resistances are

matched to within 4Ω maximum (within a device) and guaranteed low (100Ω maximum). Further, both devices are guaranteed (per MIL-STD-883, Method 3015.7) to withstand electrostatic discharge (ESD) in excess of 2000V. Pinouts are compatible with the industry-standard DG428/DG429 and DG528/DG529 multiplexers.

MAX382/MAX384 multiplexers come in 18-pin DIP and wide-SO packages, in versions tested for the commercial ($0^\circ C$ to $+70^\circ C$), extended-industrial ($-40^\circ C$ to $+85^\circ C$), and military ($-55^\circ C$ to $+125^\circ C$) temperature ranges. Prices start at \$3.00 (1000 up, FOB USA).

(Circle 11)

NiCd/NiMH fast-charger is the smallest available

The MAX2003A[†] is the most compact and the most complete stand-alone fast-charger available for NiCd (nickel cadmium) and NiMH (nickel-metal hydride) rechargeable batteries. It enables a single circuit to safely fast-charge both battery types. Available in 16-pin narrow SO packages for compact designs, the MAX2003A can be configured either as an efficient switch-mode current regulator or as a cost-effective gating controller for an external current source.

Improvements over the standard MAX2003 include a pulsed trickle charge (for longer battery life) and a temperature-slope-sensing interval that varies with charge rate (to improve the termination accuracy). The MAX2003A employs one

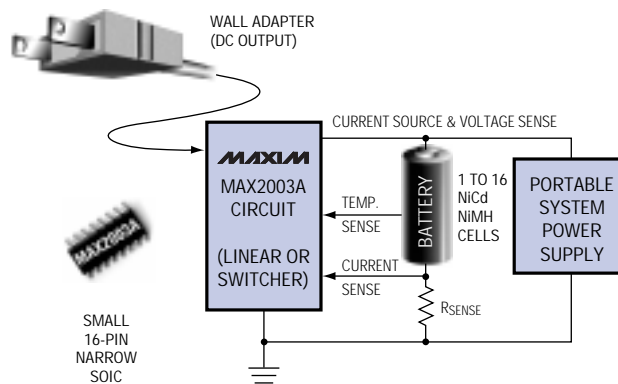
of five methods to terminate fast charges: temperature slope, negative delta voltage, maximum temperature, maximum time, and maximum voltage. As a safety feature, it inhibits a fast charge until the battery voltage and battery temperature are within acceptable limits.

To condition the battery and achieve more accurate capacity measurements, the MAX2003A provides a switch-activated discharge-before-charge option. Other options include top-off charging and direct drivers for LED status lights.

The MAX2003A comes in 16-pin narrow-SO, DIP, and wide-SO packages, tested for the commercial ($0^\circ C$ to $+70^\circ C$) temperature range. Prices start at \$3.80 (1000 up, FOB USA). An evaluation kit (MAX2003AEVKIT-SO) is available to reduce design time.

[†] Future Product, contact factory for availability.

(Circle 12)

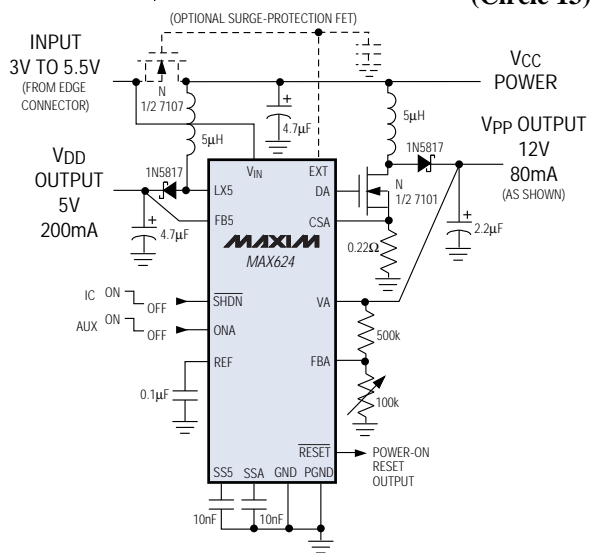


NEW PRODUCTS

Dual-output, 1MHz boost converter fits thin PCMCIA cards

The MAX624 is a miniaturized power supply for memory cards and "solid-state-disk" drives. It accepts inputs from 3V to 5.5V and produces two regulated outputs: the main output of 5V \pm 4% @ 200mA, and an auxiliary output that is adjustable \pm 2% from 5V to 30V. The high internal switching frequency (1MHz) allows use of tiny external components.

Small filter capacitors and an internal soft-start function reduce current surges at start-up, making the chip ideal for low-profile, hot-insertion PCMCIA cards. Power consumption is minimal, thanks to on-board power management and high-efficiency conversion (85% for the main 5V output). For additional power savings, you can shut down the entire chip (lowering the quiescent supply current to 40 μ A), or just the adjustable regulator (leaving the 5V output active).



(Circle 13)

If the input voltage drops below 2.8V, the chip's RESET output signals the system's controller by going low and remaining low for 4ms after V_{IN} returns above 2.8V. The MAX624 smoothly ramps the supply voltage applied to the card by controlling an optional, external, high-side n-channel MOSFET, and thereby prevents the inrush currents that would otherwise occur during hot insertions.

The MAX624 comes in a 16-pin narrow-SO package, tested for the industrial temperature range (-25°C to +85°C). Prices start at \$5.25 (1000 up, FOB USA).

Smallest GaAsFET bias generators also have lowest noise (1mVp-p)

The MAX840/MAX843/MAX844 charge pumps convert positive input voltages to low-noise negative outputs for biasing the GaAsFET power amplifier in a cellular telephone's RF transmitter. They offer the smallest size, lowest noise, and lowest operating voltage available for these applications.

Packaged in tiny 8-pin SOICs, the devices save additional space with a low external-component count. Requiring only three 0.22 μ F capacitors and one 4.7 μ F capacitor, they yield complete circuits that occupy less than 0.1in². The \pm 5% regulated output has a typical ripple voltage of only 1mVp-p, with switching frequency guaranteed in the range 80kHz to 120kHz.

The wide input range for these dc-dc converters (2.5V to 10V) allows them to operate from diverse power sources that include one or two lithium-ion cells, a battery of three to five NiCd or NiMH cells, and a plug-in adapter or charger.

The MAX840 preset output is -2V, adjustable from -0.5V to -9.4V using two external resistors. MAX843 and MAX844 outputs are adjusted over this range with an externally applied positive control voltage. Guaranteed output currents are 3mA. The MAX844 has a second negative output (unregulated) that remains active during shutdown and is suitable for use as a low-power bias supply for LCDs. MAX840 and MAX843 devices shut down completely and draw less than 1 μ A.

The MAX840/MAX843/MAX844 come in 8-pin SO packages, in versions tested for the industrial (-25°C to +85°C) and extended-industrial (-40°C to +85°C) temperature ranges. Prices start at \$1.75 (1000 up, FOB USA). To speed your designs, Maxim offers a fully assembled evaluation kit (MAX840EVKIT-SO).

(Circle 15)

One-cell battery at 0.9V powers smallest high-power boost converter

The MAX866 and MAX867 dc-dc boost converters are the smallest high-power, high-efficiency devices available for generating regulated outputs from single-cell voltage inputs. The MAX866 lets you pin-select an output of 3.3V or 5V, and the MAX867 lets you set an output between 2.7V and 6V using two external resistors. Both devices are guaranteed to start at 0.9V and deliver up to 9mA with an input of 1.2V. Once started, they guarantee operation down to 0.5V. Typical efficiencies are 80%.

MAX866 and MAX867 converters come in tiny, 8-pin μ MAX packages only 1.11mm high and 0.023in² in area. Switching frequencies as high as 250kHz enable use of small external components, resulting in an overall circuit area of just 0.2in². Because high-frequency switching also lowers the peak current and output-voltage ripple, these converters are excellent for use in pagers, remote controls, cameras, and other low-voltage miniature equipment. Each IC features a 1 μ A shutdown mode and a built-in low-battery detector.

Both converters are tested for the extended-industrial temperature range (-40°C to +85°C). Prices start at \$1.76 (1000 up, FOB USA). A fully assembled evaluation kit (MAX866EVKIT-MM) is available to speed your designs.

(Circle 14)

NEW PRODUCTS

Four-pin reset ICs include manual reset

The MAX811 (active-low reset) and MAX812[†] (active-high reset) are the smallest available μ P-reset ICs with manual-reset inputs. Requiring no external components, they come in SOT-143 packages that occupy only 98x120 mils of board area.

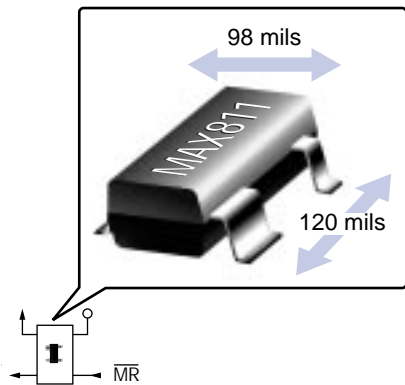
Each asserts a reset to the μ P whenever V_{CC} falls below a preset threshold, maintaining the reset for at least 140ms after V_{CC} rises back above the threshold. Short transients on V_{CC} are ignored. (The MAX811's active-low reset is guaranteed to be in the correct state for V_{CC} down to 1V.) Intended for 3V, 3.3V, and 5V systems, the MAX811 and MAX812 provide excellent reliability and low cost by eliminating the need for external components and adjustments.

MAX811/MAX812 devices come in five versions, each distinguished by a different trip threshold and suffix letter. Choosing the appropriate threshold lets you monitor (for instance) a 3V \pm 5% system containing 3V \pm 10% ICs, or a 5V \pm 5% system containing 5V \pm 10% ICs. Low supply currents (3 μ A at 3V or 3.3V, and 8 μ A at 5V) make the MAX811 and MAX812 devices ideal for portable and battery-powered applications.

The MAX811 and MAX812 come in SOT-143 packages, specified for the commercial (0°C to +70°C) temperature range. Prices start at \$0.99 (1000 up, FOB USA).

[†] Future Product, contact factory for availability.

(Circle 16)



Most-accurate μ P supervisors have \pm 1% thresholds

MAX814/MAX815/MAX816 supervisory circuits improve the reliability of microprocessor (μ P) systems because their high-accuracy (\pm 1%) reset thresholds eliminate the need for manual trimming. The ICs assert active-low $\overline{\text{RESET}}$ outputs (which remain low for 200ms after V_{CC} returns to normal) during power-up, power-down, and brownout conditions. They also include a manual-reset input ($\overline{\text{MR}}$), and an independent threshold detector that can monitor another power-supply voltage, generate a power-fail warning, or detect a low battery voltage. All devices are available in 8-pin DIP and SO packages.

The internal reset comparators ignore fast transients on V_{CC} , but guarantee correct outputs for V_{CC} levels down to 1V. Low supply currents (75 μ A maximum) make the MAX814/MAX815/MAX816 ideal for portable and battery-powered systems.

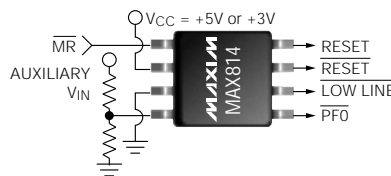
MAX814 and MAX815 supervisors offer one of four fixed reset thresholds, each with a tolerance of \pm 1%: 4.85V to 4.75V, 4.75V to 4.65V, 4.60V to 4.50V,

and 3.06V to 3.00V. The MAX816 threshold is set with an external resistor divider. MAX814 and MAX816 supervisors provide active-high $\overline{\text{RESET}}$ s in addition to the $\overline{\text{RESET}}$ output. The MAX815 includes a watchdog circuit that monitors software execution, and issues a warning when 1.56 seconds elapse without digital transitions on its input line.

The MAX814 has a 2-stage power-fail warning in which a separate low-line comparator compares V_{CC} to a threshold 60mV above the reset threshold. Thus, a declining supply voltage triggers an early warning that allows the system to prepare itself for the complete loss of V_{CC} .

The MAX814/MAX815/MAX816 come in 8-pin DIP and SO packages, in versions tested for the commercial (0°C to +70°C) and extended-industrial (-40°C to +85°C) temperature ranges. Prices start at \$4.39 (1000 up, FOB USA).

(Circle 17)



Spread-spectrum transmitter suits 900MHz ISM band

The MAX2402 transmitter integrates a double-balanced mixer, variable-gain stage, and 20dBm power amplifier on a single bipolar die. Operating from 800MHz to 1000MHz, it suits many applications in portable and stationary wireless design—particularly spread-spectrum systems in the 902MHz to 928MHz ISM band that employ direct-sequence or frequency-hopping techniques.

The LO (local oscillator) inputs, internally self-biased and capacitively coupled, lead to an amplifier that accepts differential or single-ended signals in the range -10dBm to 6dBm. The modulation input (MOD) has a 200MHz bandwidth and accepts input signals at any level within the supply range (4.75V \leq V_{CC} \leq 5.5V). This input buffer is linear from (approximately) 1.5V to 3.5V.

MOD and LO signals are applied to a double-balanced Gilbert-cell mixer optimized for carrier-frequency rejection. Next is a variable-gain stage that provides more than 35dB of adjustment, and a class AB power amplifier that provides more than 20dBm of output power (more than 100mW into 50 Ω). The power amplifier has a bias adjustment that allows the user to make trade-offs between efficiency and harmonic distortion. A power-down function lowers the supply current below 1 μ A in less than 10 μ s.

The MAX2402 comes in a 20-pin SSOP package screened for the extended-industrial temperature range (-40°C to +85°C). Prices start at \$3.78 (1000 up, FOB USA). Maxim also offers an evaluation kit (MAX2402EVKIT-SO) that saves time and design effort.

(Circle 18)

NEW PRODUCTS

70MHz quadrature modulator/demodulator operates from 3V

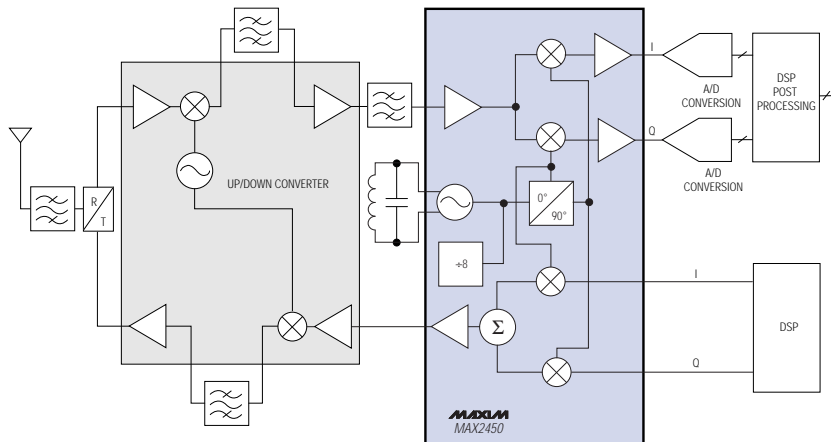
The monolithic MAX2450[†] includes a quadrature modulator and demodulator with supporting oscillator and divide-by-8 prescaler. It operates from a single 3V supply and draws only 8mA. Applications include 900MHz digital cordless phones, GSM and North American cellular radios, wireless LANs, VHF transceivers, and digital satellite communications.

The modulator accepts 10MHz differential I and Q baseband signals (with amplitudes to 1.2Vp-p) and produces differential IF outputs as high as 70MHz. The demodulator accepts single-ended IF

inputs and produces the differential I and Q outputs. This demodulator circuitry provides 49dB of voltage-conversion gain, automatic cancellation of baseband offset, and a thin-film 400Ω input impedance for matching to an external IF filter.

To minimize spurious feedback, the MAX2450's internal oscillator is set at twice the IF (typically 140MHz) via external tuning components. The oscillator and associated phase shifters produce differential quadrature signals exhibiting amplitudes within ±0.3dB and a phase imbalance no greater than ±3°. A command to the CMOS-compatible power-down control lowers the supply current to 20μA.

The MAX2451[†] includes a demodulator, oscillator, and prescaler. The



Clock-recovery/data-retiming IC includes phase/frequency detector

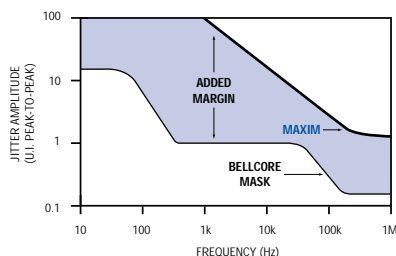
The MAX3270 provides clock recovery and data retiming for 155bps and 622bps applications in SDH/SONET (Synchronous Digital Hierarchy/Synchronous Optical Network) and ATM (Asynchronous Transfer Mode) systems. It meets Bellcore and CCITT specifications for jitter tolerance, which ensures error-free data recovery.

Recovered clock and data signals are phase-aligned by a fully integrated phase-locked loop (PLL). The PLL has a phase/frequency detector that eliminates the need for an external reference clock. The MAX3270 includes an output-

frequency monitor for detection of PLL acquisition or loss of input data. Differential-ECL inputs and outputs make the device less susceptible to noise in a high-frequency environment.

The MAX3270 is available in a 44-pin MQFP package, screened for the extended-industrial (-40°C to +85°C) temperature range. Prices start at \$39.50 (1000 up, FOB USA).

(Circle 20)



MAX2452[†] includes a modulator, oscillator, and prescaler.

MAX2451 and MAX2452 devices come in 16-pin narrow-SO packages; the MAX2450 comes in a 20-pin SO package. All are screened for the commercial (0°C to +70°C) temperature range.

[†] Future Product, contact factory for availability.

(Circle 19)

High-gain RF amplifier delivers 100mW in 900MHz band

The MAX2430[†] RF power amplifier operates on 3V to 5V and delivers more than 100mW (20dBm) in the 800MHz to 1000MHz band. This capability enables the low-cost device to serve as an output amplifier for cordless telephones and other ISM applications, and as a pre-driver for higher-power applications such as cellular telephones.

Designed for Class AB operation, the MAX2430 consists of a large power transistor driven by a capacitively coupled gain stage and a driver stage. The input impedance is matched to 50Ω (VSWR ≤1.5) and the overall power gain is guaranteed >30dB. To save power during the "idle slots" in time-division multiple-access (TDMA) transmissions, a TTL/CMOS-compatible command (EN low) can throttle the MAX2430 supply current to below 10μA in only 1μs.

Another input pin (V_B) has various uses. The simplest is to connect an external capacitor from V_B to ground, which programs the RF output's ramp rate during turn-on and turn-off. For other applications, a diode to ground maintains the RF input match during turn-off while reducing the quiescent current from 50mA to 25mA. More generally, driving V_B with a control voltage of 0V to 2.2V adjusts the bias and power gain in the output stage. V_B also enables more complex control arrangements.

The MAX2430 comes in a 16-pin narrow-SO package, screened for the industrial (0°C to +85°C) temperature range.

[†] Future Product, contact factory for availability.

(Circle 21)