MAXIM Engineering Journal

News Briefs

■ **MAXIM REPORTS 38TH CONSECUTIVE QUARTER OF INCREASED REVENUES AND 36TH CONSECUTIVE QUARTER OF INCREASED EARNINGS**

Maxim Integrated Products, Inc., reported record net revenues of $66,628,000 for the third quarter ended March 31, 1995, compared to $40,572,000 for the same period a year ago. This represents a 64.2% gain in net revenues from the same quarter a year ago and the 38th consecutive quarter of increased revenues. Net income of $10,124,000 (or $0.30 per share) for the quarter marked the 36th consecutive increasingly profitable quarter for Maxim and compared to net income of $6,318,000 (or $0.20 per share) for the same quarter in fiscal 1994.

Operating income for the quarter was 22.3% of net revenues, again one of the industry's highest. Gross margin increased to $38,977,000 (58.5% of net revenues) from $23,687,000 (58.4% of net revenues) for the same quarter in fiscal 1994.

In the three months ended March 31, 1995, the Company recorded a $10.9 million charge to operating income relating to the Company's program announced in Q2 to modernize its equipment and manufacturing facilities. The charge relates to a cumulative adjustment for depreciation, as a result of changing estimates of useful lives associated with equipment that management estimates will be replaced or substantially upgraded over the next 3 years. Approximately $6.9 million and $4.0 million of the charge was allocated to cost of goods sold and research and development expense, respectively, based on the nature of the equipment affected.

Jack Gifford, Chairman, President, and CEO, commented: "This was an exceptionally strong quarter for Maxim. Bookings across all geographic regions and across all product lines remained at record levels. The number of individual orders received during this quarter increased approximately 100% over the same quarter last year. This is a strong indication that our growth is coming from new customers and new applications for our products. We generated over $8 million of cash after purchasing $11.9 million of capital equipment and purchasing $3.2 million of Maxim common stock."

Gifford continued: "Our conversion from 4" to 6" wafers at our acquired Beaverton, Oregon, facility is on track. We anticipate that over 50 percent of our Q4 and Q1 wafer requirements will be satisfied from this facility."

Maxim has been selected by the New York consulting firm Stern Stewart as one of "America's Best Wealth Creators." The selection is based on "market value added" (MVA), described by *Fortune* magazine as "the best way yet for evaluating how well a company performs its basic mission of creating wealth for shareholders." MVA is the difference between the total market value of the company's stock and its invested capital. The difference, when positive, constitutes the increase in value of the capital provided by lenders and shareholders.

Maxim has once again been selected by *Business Week* as one of "America's 1000 Most Valuable Companies," ranked by market value. According to *Business Week*, "investors today are clearly favoring only those companies likely to continue growing even if the overall economy slows. That's the case for a blessed handful of companies, for which 1994 wasn't a topsy-turvy year at all. Most were in the technology sector, where demand for semiconductors, computers, and related equipment shows no signs of letting up."

Versatile waveform generator operates from 0.1Hz to 20MHz

As a single-chip source of high-frequency waveforms, the MAX038 can serve both as the core of a moderate-performance waveform generator, and as a card-level waveform source for electronic systems in which the cost of a dedicated waveform generator may be prohibitive.

The MAX038 is a precision, high-frequency function generator that produces accurate sine, square, triangle, sawtooth, and pulse waveforms with a minimum of external components. The internal 2.5V reference (plus an external capacitor and potentiometer) lets you vary the signal frequency from 0.1Hz to 20MHz. An applied ±2.3V control signal varies the duty cycle between 10% and 90%, enabling the generation of sawtooth waveforms and pulse-width modulation.

A second frequency-control input—used primarily as a VCO input in phase-locked-loop applications—provides ±70% of fine control. This capability also enables the generation of frequency sweeps and frequency modulation. The frequency and duty-cycle controls have minimal interaction with each other.

All output amplitudes are $2V_{p-p}$, symmetrical about ground. The low-impedance output terminal delivers as much as ±20mA, and a two-bit code applied to the TTL-compatible A0 and A1 inputs selects the sine, square, or triangle output waveform:

A0	A1	Waveform
X	1	Sine wave
0	0	Square wave
1	0	Triangle wave

(X = Don't care)

To synchronize MAX038 operation with other devices in the system, the internal oscillator produces a TTL-compatible SYNC output, whose duty cycle remains constant at 50% regardless of the duty cycle set for the output waveform. The MAX038's internal phase detector enables such synchronization as well. It also enables the demodulation of frequency-modulated signals.

Details of operation

By alternately charging and discharging an external capacitor, the MAX038's relaxation oscillator produces simultaneous square and triangle waves. An internal sine-shaping circuit converts the triangle to a low-distortion, constant-amplitude sinewave. The sine, square, and triangle waves are applied to an internal multiplexer that lets you select the output waveform according to the state of address lines A0 and A1. The output amplitude remains constant at ±1V regardless of wave shape or frequency (**Figure 1**).

See **Figure 2** for the MAX038's block diagram and operating circuit. Powered from ±5V, the device consumes 400mW and has a nominal output frequency set by the oscillator capacitor C_F. Coarse deviations from that frequency are made by varying the IIN current from 2µA to 750µA, a range of 375:1 (**Figure 3**). As shown, the IIN current can be derived using the on-board 2.5V reference and an external fixed or variable resistor.

To adjust the frequency digitally, connect a voltage-output DAC to IIN via a series resistor (**Figure 4**). The converter output ranges from 0V at zero to 2.5(255/256)V at full-scale. Current injected by the converter into IIN, therefore, ranges from 0µA to 748µA. The 2.5V reference and 1.2MΩ resistor inject a constant 2µA, so (by superposition) the net current into IIN ranges from 2µA (at a code of 0000 0000) to 750µA (at 1111 1111). The quad-DAC IC operates from 5V or ±5V. As described below, it can also provide digital control of FADJ and DADJ.

For fine adjustments (±70%), apply a control voltage in the range ±2.3V to the frequency adjust (FADJ) terminal (**Figure 5**). Both FADJ and IIN have wide bandwidths that allow the output frequency to be modulated at a maximum rate of about 2MHz (**Figure 6**). As the more linear input, IIN is preferred for open-loop frequency control. As the voltage input, FADJ is better suited for use in a phase-locked loop. For digital control of FADJ, configure a DAC and external op amp (as in Figure 4) to produce an output ranging from -2.3V (0000 0000) to 2.3V (1111 1111).

Duty cycle (the percentage of time that the output is positive) can be adjusted in the range 10% to 90% by applying a ±2.3V control signal to the duty-cycle-adjust terminal DADJ (**Figure 7a**). This signal changes the ratio of charge current to discharge current for the C_F capacitor while maintaining a nearly constant output frequency (**Figure 7b**).
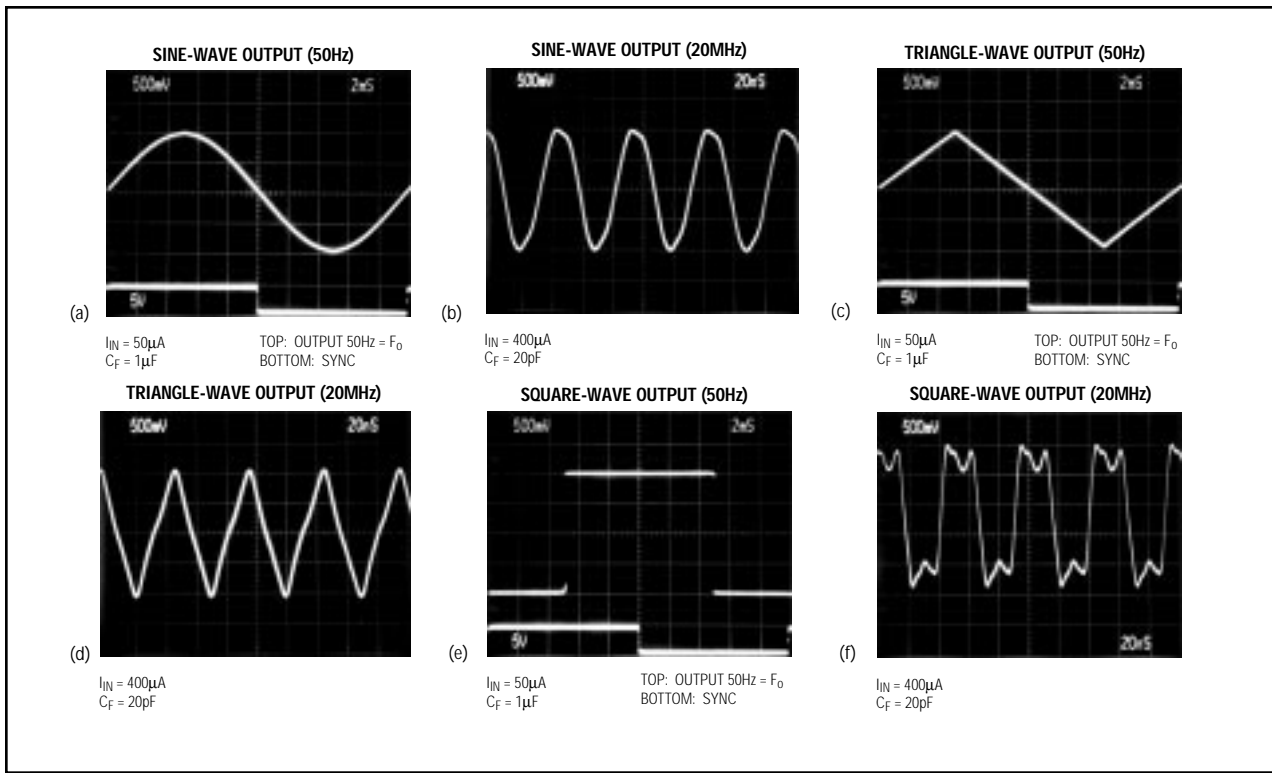
SINE-WAVE OUTPUT (50Hz)

(a) $I_{IN} = 50\mu A$
$C_F = 1\mu F$
TOP: OUTPUT 50Hz = F_o
BOTTOM: SYNC

SINE-WAVE OUTPUT (20MHz)

(b) $I_{IN} = 400\mu A$
$C_F = 20pF$

TRIANGLE-WAVE OUTPUT (50Hz)

(c) $I_{IN} = 50\mu A$
$C_F = 1\mu F$
TOP: OUTPUT 50Hz = F_o
BOTTOM: SYNC

TRIANGLE-WAVE OUTPUT (20MHz)

(d) $I_{IN} = 400\mu A$
$C_F = 20pF$

SQUARE-WAVE OUTPUT (50Hz)

(e) $I_{IN} = 50\mu A$
$C_F = 1\mu F$
TOP: OUTPUT 50Hz = F_o
BOTTOM: SYNC

SQUARE-WAVE OUTPUT (20MHz)

(f) $I_{IN} = 400\mu A$
$C_F = 20pF$

Figure 1. Among waveforms available at the MAX038 output are the sine (a, b), triangle (c, d), and square (e, f).

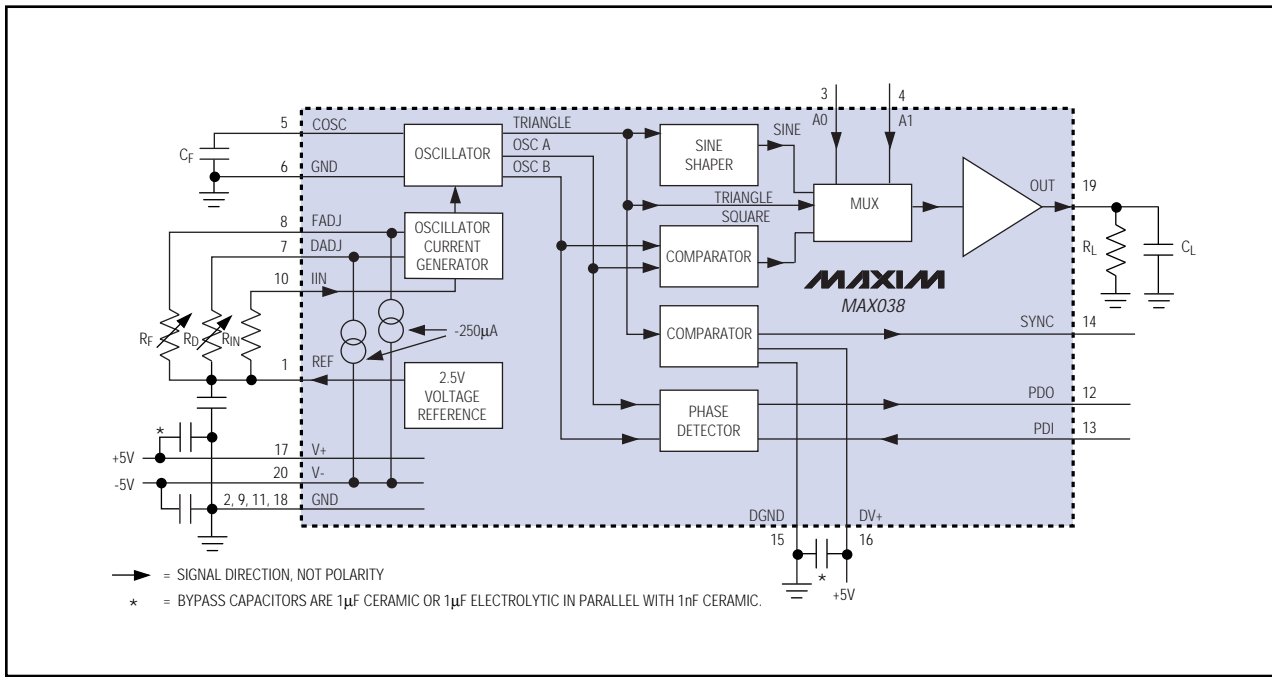*Figure 2. This figure combines the block diagram and basic operating circuit for the MAX038 0.1Hz-to-20MHz waveform generator.*
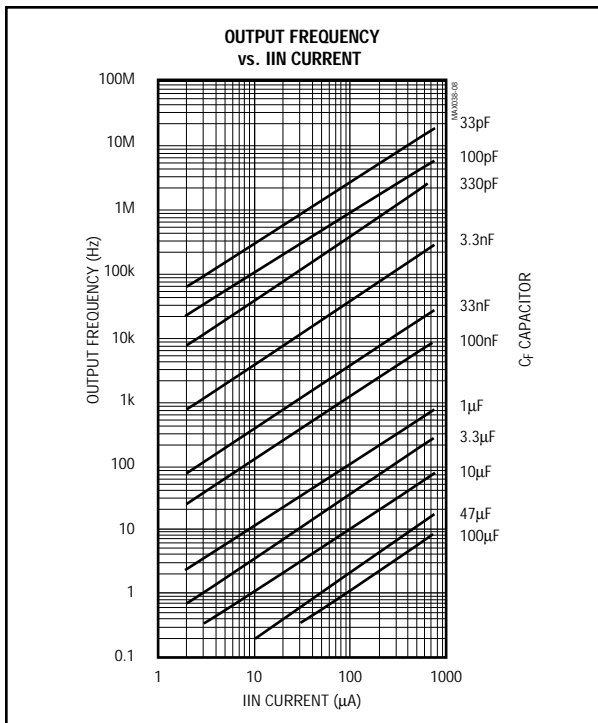
OUTPUT FREQUENCY
vs. IIN CURRENT

Figure 3. The output frequency in Figure 2 varies with current into IIN and the value of C_F.

The DADJ input also lets you minimize distortion in the output sine wave. Minimum distortion occurs at a duty cycle of exactly 50%, but the typical duty cycle (with $V_{DADJ} = 0V$) is 50% ±2%. By applying a small control voltage (typically less than ±100mV) to DADJ, therefore, you can set the exact 50% symmetry that minimizes distortion (see insert, **Figure 8**).

The source driving DADJ must supply a constant 250µA (see Figure 2). The temperature coefficient of this internal current sink is unimportant for op amps and other low-impedance sources, but is significant when using a variable resistor as shown. Thus, variable resistors suit manual operation only, in which the operator can correct errors through readjustment. Like FADJ, DADJ has a 2MHz bandwidth and ±2.3V range. It can be digitally controlled with an identical circuit (Figure 4). **Figure 9** shows the duty-cycle modulation that results when a triangle wave is applied to DADJ.
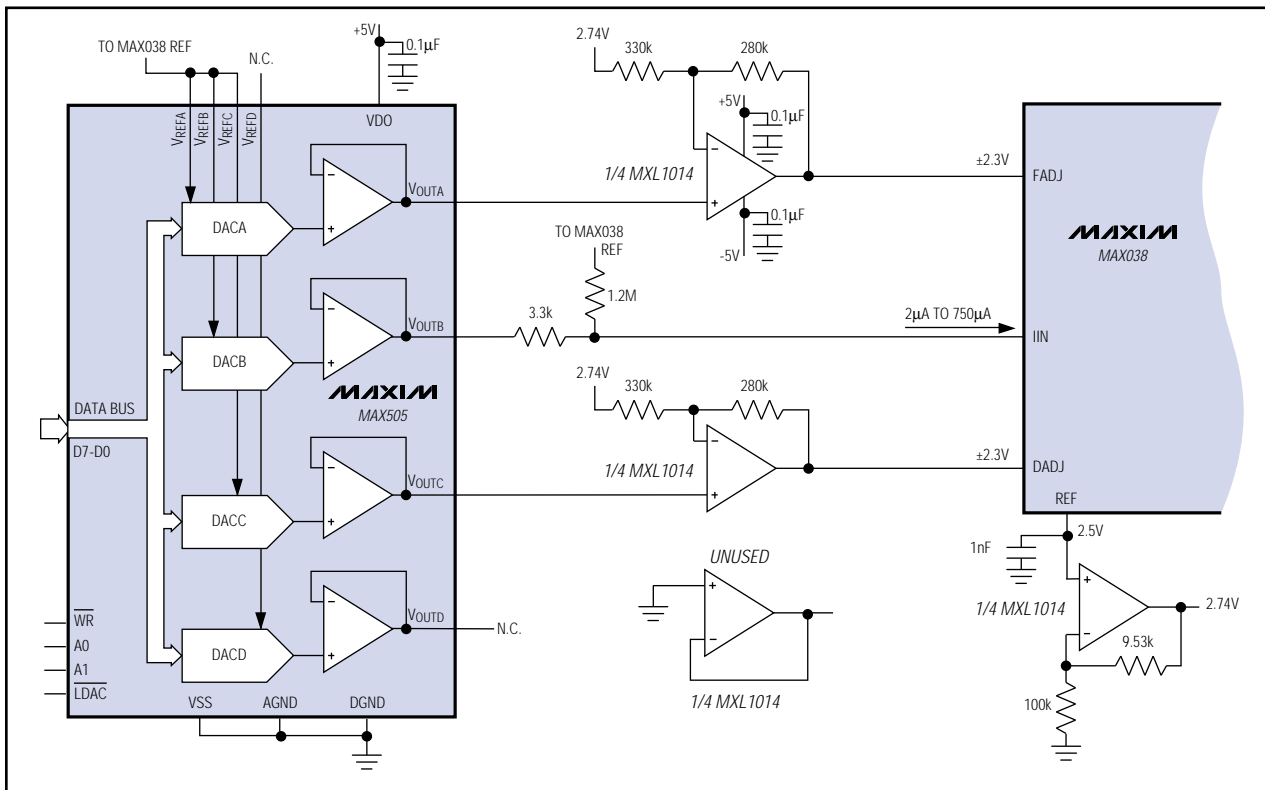

Figure 4. Three 8-bit D/A converters easily provide digital control of the two frequency-adjust inputs (IIN and FADJ) and the duty-cycle-adjust input (DADJ).
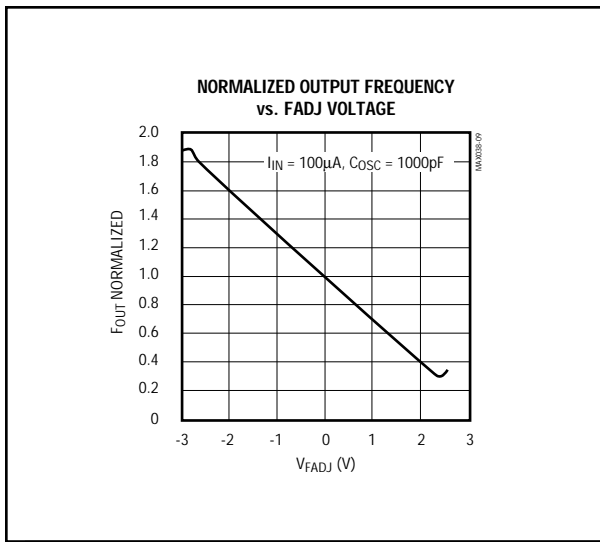
NORMALIZED OUTPUT FREQUENCY vs. FADJ VOLTAGE

$I_{IN} = 100\mu A$, $C_{OSC} = 1000pF$

Figure 5. For fine control of the output frequency, apply a control voltage in the range ± 2.3V to FADJ.

FREQUENCY MODULATION USING IIN

(a)

TOP: OUTPUT
BOTTOM: I_{IN}

FREQUENCY MODULATION USING FADJ

(b)

0.5V
0
-0.5V

TOP: OUTPUT
BOTTOM: FADJ
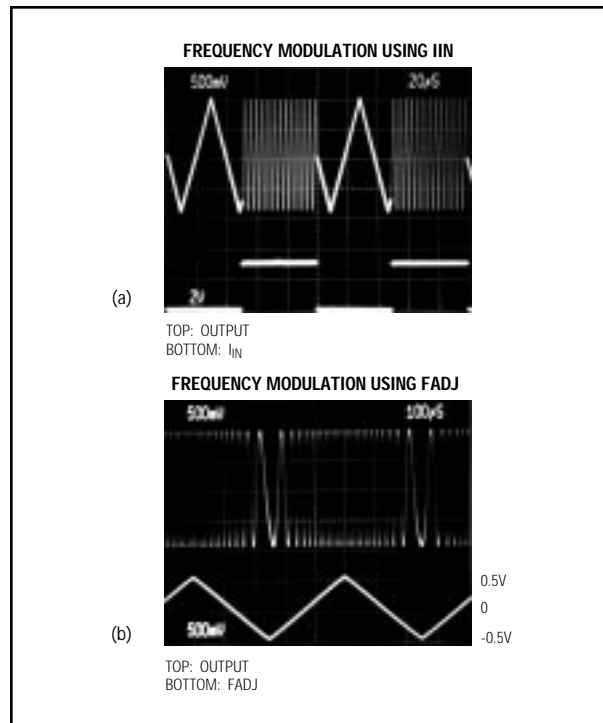
Figure 6. The MAX038 inputs IIN and FADJ allow gross (a) and fine (b) adjustments of output frequency.

DUTY CYCLE vs. DADJ VOLTAGE

$I_{IN} = 200\mu A$

(a)

NORMALIZED OUTPUT FREQUENCY vs. DADJ VOLTAGE

$I_{IN} = 10\mu A$
$I_{IN} = 25\mu A$
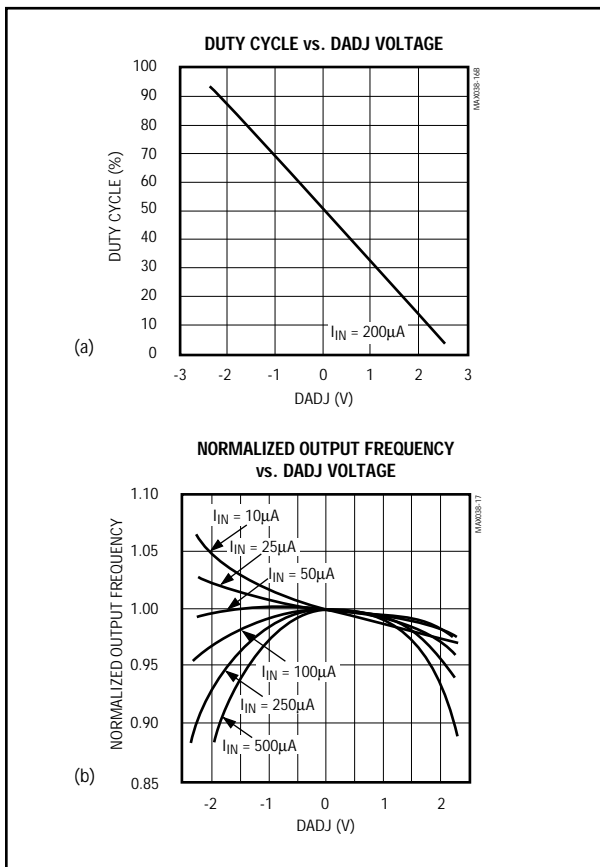$I_{IN} = 50\mu A$
$I_{IN} = 100\mu A$
$I_{IN} = 250\mu A$
$I_{IN} = 500\mu A$

(b)

Figure 7. An independent DADJ control voltage in the range ± 2.3V adjusts the duty cycle (a) with little effect on frequency (b).

$$F_0 = \frac{2 \times 2.5V}{R_{IN} \times C_F}$$

PRECISION DUTY-CYCLE ADJUSTMENT CIRCUIT
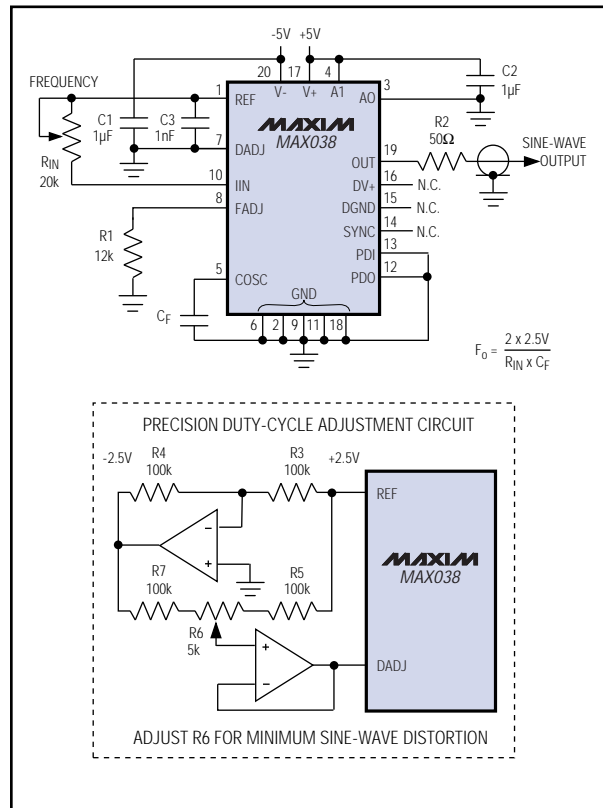
ADJUST R6 FOR MINIMUM SINE-WAVE DISTORTION

Figure 8. This circuit lets you adjust for the minimum sine-wave distortion available at a 50% duty cycle.

Phase-locked operation

The MAX038's internal phase detector is intended primarily for use in phase-locked-loop (PLL) configurations. In **Figure 10a**, for example, the phase detector in IC2 enables that device to synchronize its operation with that of IC1. You connect the applied reference signal to IC2's TTL/CMOS-compatible phase-detector input (PDI) and connect the phase-detector output (PDO) to the input (FADJ) of the internal voltage-controlled oscillator. PDO is the output of an exclusive-OR gate—a mixer—which produces rectangular current waveforms at frequencies equal to the sum and difference of the PDI frequency and the MAX038 output frequency. These waveforms are integrated by C_{PD} to form a triangle-wave voltage output at PDO (**Figure 10b**). The 10Ω/100pF pair at PDI limits that pin's rate of rise to 10ns.
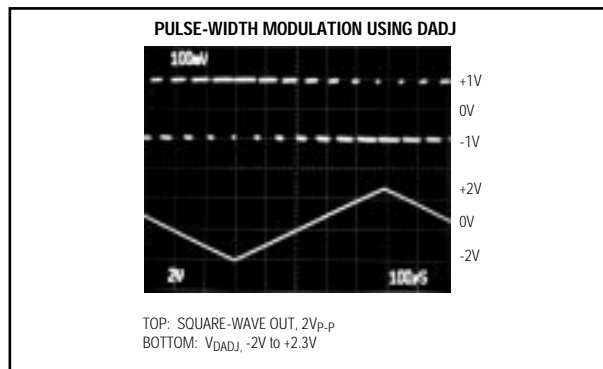

PULSE-WIDTH MODULATION USING DADJ

TOP: SQUARE-WAVE OUT, 2V$_{P-P}$
BOTTOM: V$_{DADJ}$, -2V to +2.3V

Figure 9. A triangle wave applied to DADJ provides linear variations in the duty cycle.

The PDO current-pulse levels are 0µA and 500µA, with a duty cycle that approaches 50% when PDI and the output are in phase quadrature (90° out of phase). Otherwise, the duty cycle approaches 100% when the phase difference approaches 180°, and 0% when the phase difference approaches 0°. R_{PD}, C_{PD}, and R_Z comprise a filter that determines the PLL frequency response.

At the SYNC output is a square wave of fixed 50% duty cycle, whose rising edge coincides with the rising edge of an output sine or triangle wave as it passes through zero volts. If the output is a square wave, SYNC's rising edge occurs at the mid-point of the positive portion, causing SYNC to lead the output by 90°.

SYNC lets you slave one MAX038 to another by providing a TTL-compatible square wave at the phase-detector input (PDI), as required by the slaved device (IC2). On the other hand, SYNC isn't available if a MAX038 is synchronized with sine or triangle waves from other sources. For those cases, the PDI input must be driven by a comparator (as shown in **Figure 11**) to square up the signal and provide the appropriate level shifting.

The internal phase detector can also demodulate frequency-modulated signals (Figure 11). In this circuit, the output of one MAX038 is being frequency modulated with a 10kHz sine wave. The ±34mV signal at FADJ of IC1 causes a ±1% variation in the output frequency (±100kHz), and the rate of variation is 10kHz. The comparator assures a proper square wave for IC3's PDI input as mentioned above. The two MAX038s are set to the same center frequency.
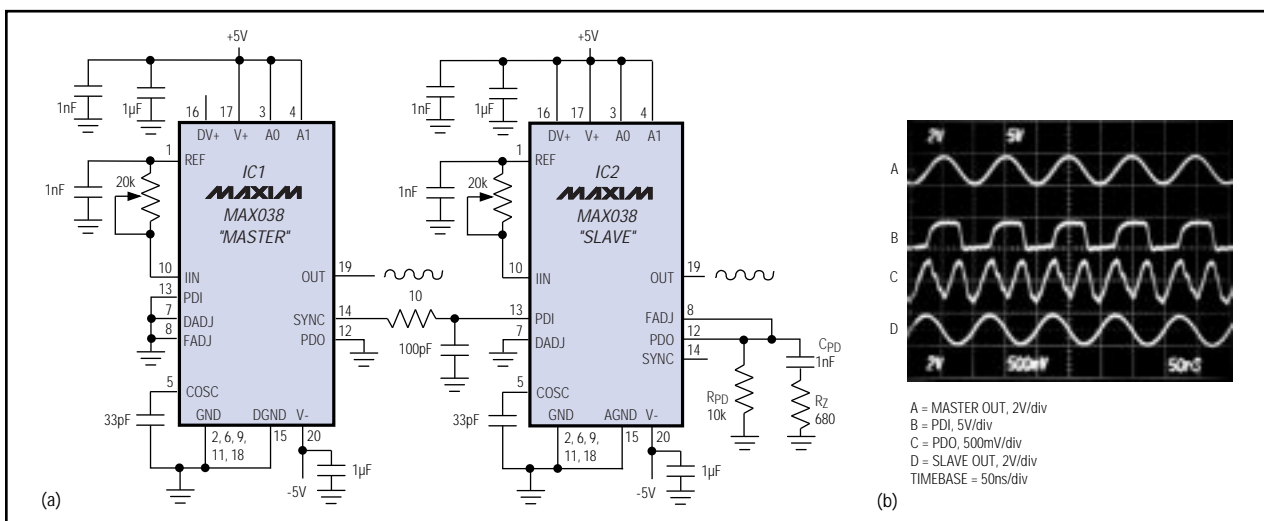

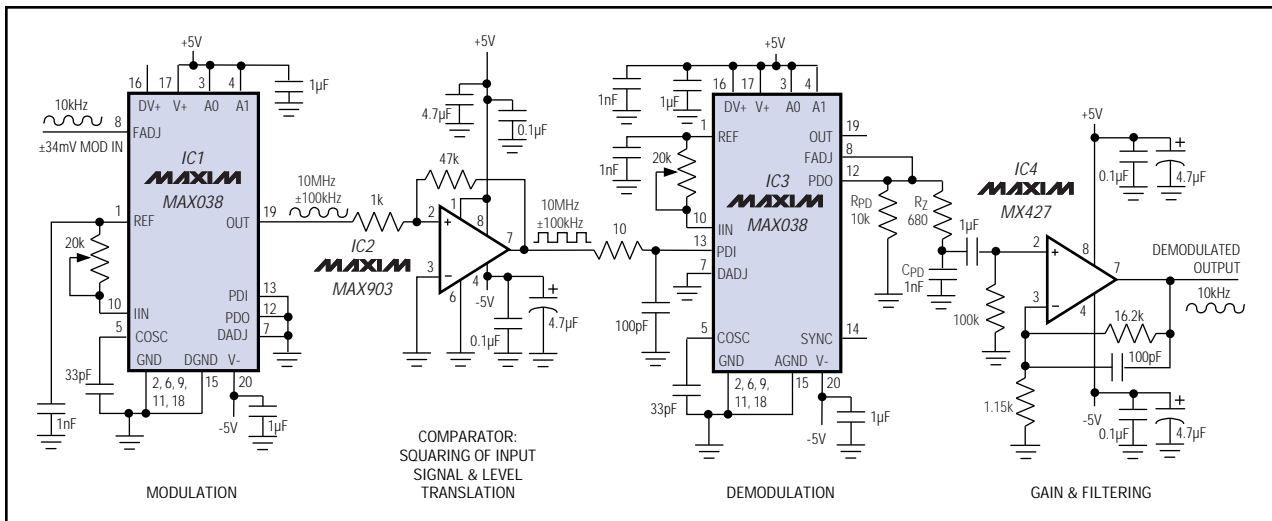*Figure 10. The SYNC output lets you synchronize one MAX038 to another.*

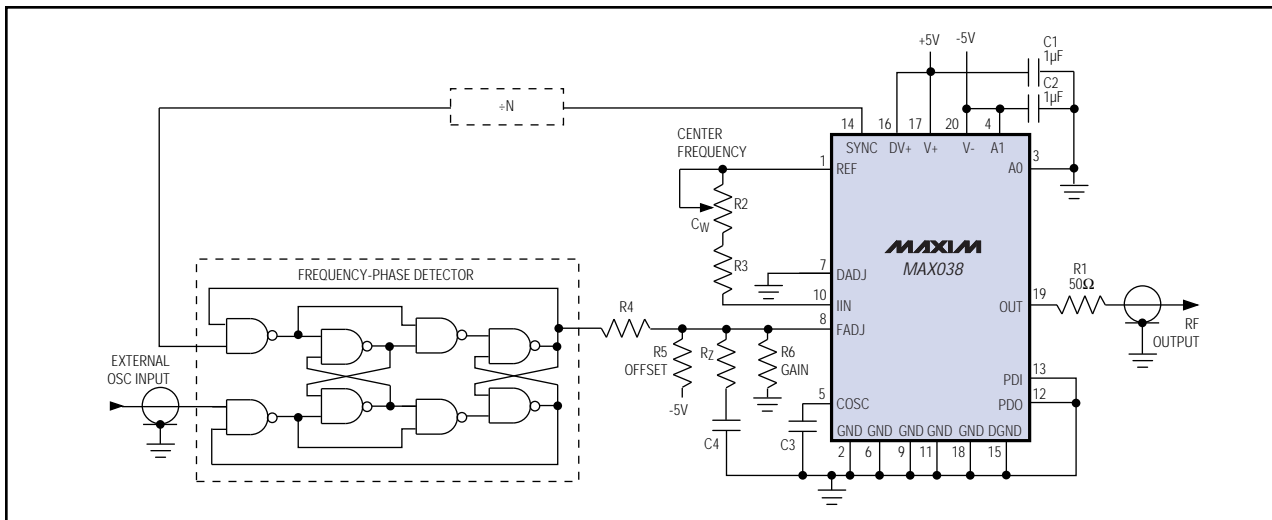*Figure 11. This circuit modulates a 10MHz carrier with 10kHz, and then recovers the modulating signal.*

Figure 12. This discrete-gate frequency-phase detector allows use of an external ÷N circuit, and assures that the PLL will lock only to the fundamental (not a harmonic) of the applied frequency.

The frequencies at IC3's phase-detector output are the sum and difference of the frequencies at PDI and OUT. Thus, with appropriate cutoff frequency and gain, the lowpass filter (IC4) passes only the original 10kHz signal to the demodulated output. The pole for this filter is set by the 16.2kΩ and 100pF components. As in Figure 10a, the frequency response for IC3's PLL is set by R_{PD}, C_{PD}, and R_Z (see **Sidebar**).

When the loop is in lock, PDI is in approximate phase quadrature with the output signal. Also when in lock, the duty cycle at PDO is 50%, and PDO's average output current is 250µA. The current sink at FADJ demands a constant 250µA, so PDO outputs above and below that

level develop a bipolar error voltage across R_{PD} that drives the FADJ voltage input. Note—the MAX038's internal phase detector is a phase-only detector, producing a PLL whose frequency-capture range is limited by the bandwidth of its loop filter. For wider-range applications, consider an external phase-frequency detector.

To gain the advantages of a wider capture range and an optional ÷N circuit (which allows the PLL to lock onto arbitrary multiples of the applied frequency), you can introduce an external frequency-phase detector such as the 74HC4046 or the discrete-gate version shown in **Figure 12**. Unlike phase detectors that may lock to harmonics of

the applied signal, the frequency-phase detector locks only to the fundamental. In the absence of an applied frequency, its output assumes a positive dc voltage (logic "1") that drives the RF output to the lower end of its range as determined by resistors R4–R6. These resistors also determine the frequency range over which the PLL can achieve lock. Again, R4–R6, C4, and R_Z determine the PLL's dynamic performance.

Frequency synthesizer

The MAX038 and four other ICs can form a crystal-controlled, digitally programmed frequency synthesizer that produces accurate sine, square, or triangle waves in 1kHz increments over the range 8kHz to 16.383MHz (**Figure 13**). Each of the 14 manual switches (when open) makes the listed contribution to output frequency: opening only S0, S1, and S8, for example, produces an output of 259kHz.

The switches generate a 14-bit digital word that is applied in parallel to the D/A converter (IC2) and a ÷N circuit in IC1. IC1 also includes a crystal-controlled oscillator and high-speed phase detector, which form a phase-locked loop with the voltage-controlled oscillator in IC5.

The DAC and dual op amp (IC4) produce a 2µA-to-750µA current that forces a coarse setting of the IC5 output frequency—sufficient to bring it within capture range of the PLL. This loop, in which the phase detector in IC1 compares IC5's SYNC output with the crystal-oscillator frequency divided by N, produces differential-phase information at PDV and PDR. IC3 then filters and converts this information to a ±2.5V single-ended signal, which, when summed with an offset and applied to FADJ, forces the SIGNAL OUTPUT frequency to the exact value set by the switches.

Applying coarse frequency control with the DAC and the IIN terminal of IC5 (pin 10) gives the fine-control input (FADJ) a reasonably fast response to switch changes. The 50MHz, 50Ω lowpass output filter passes 16MHz sine, square, and triangle waves with reasonable fidelity, while blocking high-frequency noise generated by the ÷N circuit.

The MAX038 is priced at $10.37 (100 pcs, FOB USA).

(Circle 1)

Phase-locked-loop analysis

The following discussion borrows heavily from Chapter 2 of "Phaselock Techniques," 2nd Edition, 1979, by Floyd M. Gardner.

Phase-locked loops such as those operating in Figures 10a and 11 can be modeled by a phase detector, loop filter, and voltage-controlled oscillator (VCO) as shown in **Figure A**. The input-signal phase is $\phi_i(t)$ and the VCO-output phase is $\phi_o(t)$. Assuming that the loop is locked, the phase detector is linear, and the phase detector's output current (i_d) is proportional to the phase difference between its inputs, then:

(1) $i_d = K_d(\phi_i - \phi_o),$

where K_d, called the phase-detector gain factor, is measured in units of amperes per radian. For the MAX038, $K_d = 3.18*10^{-4}$ A/rad.

The loop filter determines dynamic performance for the loop. It transforms the phase-detector error current (i_d) into an error voltage (v_c) that determines the VCO's output frequency. The VCO's deviation from center frequency (f_o) is:

(2) $\Delta f = \dfrac{K_o v_c}{2\pi},$

where K_o is the VCO gain factor, and has units of rad/sec-V. For the MAX038, $K_o = 0.2915\omega_o$, where $\omega_o = 2\pi f_o$. Since frequency is the derivative of phase, the VCO's operation may be described as:

(3) $\dfrac{d\phi_o}{dt} = K_o v_c.$

Taking the Laplace transform of equation 3:

(4) $L\left[\dfrac{d\phi_o(t)}{dt}\right] = s\phi_o(s) = K_o V_c(s).$

Therefore:

(5) $\phi_o(s) = \dfrac{K_o V_c(s)}{s}.$

Thus, the VCO output phase is linearly related to the integral of the control voltage. Taking the Laplace transform of equation 1:
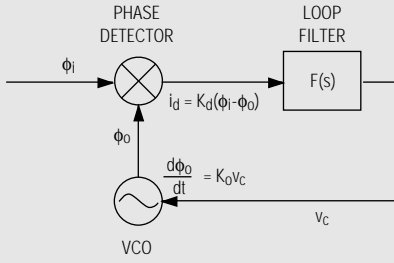
(6) $I_d(s) = K_d[\phi_i(s) - \phi_o(s)].$

Figure A. Phase-lock loop, basic block diagram.

Figure B. Loop filter.

And from Figure A:

(7) $V_c(s) = F(s)I_d(s),$

where F(s) is the transfer function for the loop filter (**Figure B**). Combining the equations 5, 6, and 7 gives the loop equations 8, 9, and 10:

(8) $H(s) = \dfrac{\phi_o(s)}{\phi_i(s)} = \dfrac{K_o K_d F(s)}{s + K_o K_d F(s)},$

(9) $\dfrac{\phi_i(s) - \phi_o(s)}{\phi_i(s)} = \dfrac{\phi_e(s)}{\phi_i(s)} = \dfrac{s}{s + K_o K_d F(s)} = 1 - H(s),$

(10) $V_c(s) = \dfrac{s K_d F(s)\phi_i(s)}{s + K_o K_d F(s)} = \dfrac{s\phi_i(s)}{K_o} H(s),$

where H(s) is the closed-loop transfer function and $\phi_e(s)$ is the phase error. The transfer function for the loop filter (Figure B) is F(s):

$$F(s) = \frac{V_{OUT}(s)}{I_{IN}(s)} = \frac{R_{PD}(s C_{PD} R_Z + 1)}{s C(R_{PD} + R_Z) + 1}$$

(11)

$$= \frac{R_{PD}(s\tau_2 + 1)}{s\tau_1 + 1},$$

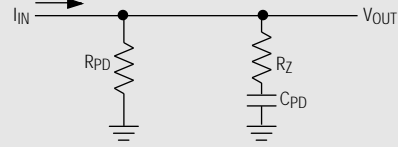where $\tau_1 = (R_{PD} + R_Z)C_{PD}$, and $\tau2 = R_Z C_{PD}$.

Substituting the F(s) expression into equation 8 gives:

(12) $H(s) = \dfrac{\dfrac{K_o K_d R_{PD}(s\tau_2 + 1)}{\tau_1}}{s^2 + \dfrac{s(1 + K_o K_d R_{PD}\tau_2)}{\tau_1} + \dfrac{K_o K_d R_{PD}}{\tau_1}},$

which can be written as:

(13) $H(s) = \dfrac{s\left(\zeta\omega_n - \dfrac{\omega_n^2}{K_o K_d R_{PD}}\right) + \omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2},$

where ω_n is the loop's natural frequency and ζ is its damping factor:

$$\omega_n = \left(\frac{K_o K_d R_{PD}}{\tau_1}\right)^{\frac{1}{2}},$$

$$\zeta = \frac{1}{2}\left(\frac{K_o K_d R_{PD}}{\tau_1}\right)^{\frac{1}{2}}\left(\tau_2 + \frac{1}{K_o K_d R_{PD}}\right).$$

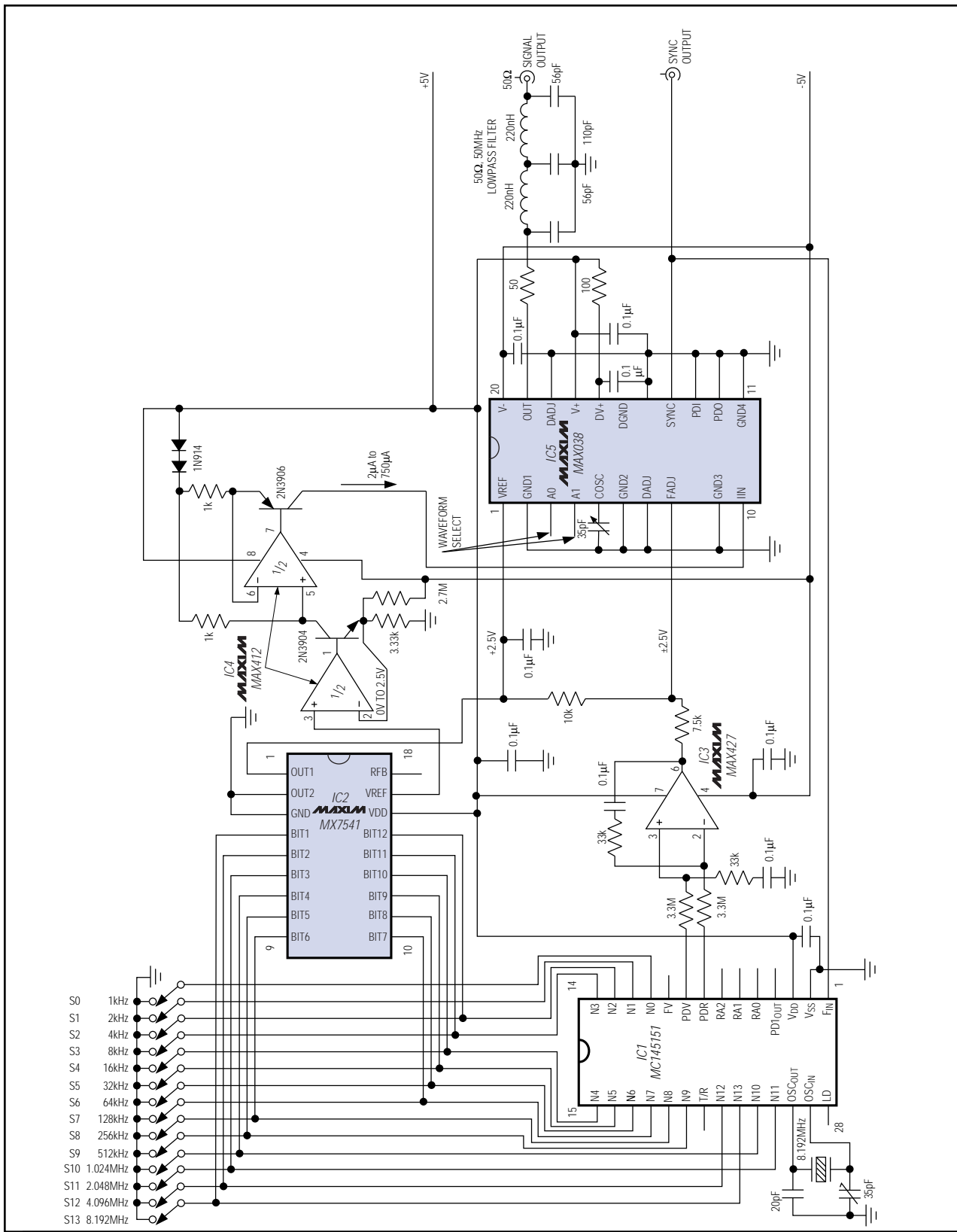As in conventional 2nd-order systems, ζ should be greater than 0.707.

Figure 13. This manually programmed digital frequency synthesizer can step from 8kHz to 16.383MHz in 1kHz increments.

Derive 5V from four AA cells

Four AA cells in series—a common power source for portable instruments—produces a battery voltage that declines from about 6V when fresh to about 4V when discharged. For 5V regulators, this above-and-below variation complicates the circuit design.

Flyback-transformer circuits can convert 6V-to-4V inputs to a regulated 5V, as can a step-up (boost) converter followed by a linear regulator. (For $\pm$5V requirements, you can choose either circuit plus a charge pump.) If, however, the instrument is fully portable and the battery voltage can float, a less complicated inverter circuit can easily generate the 5V or $\pm$5V rails. Moreover, the inverter's single switching frequency simplifies filtering and precludes the generation of beat frequencies.

The inverter circuit substitutes a transformer with two matched windings for the usual inductor (**Figure 1a**). When IC1's internal switch turns off, the circuit impresses V_{OUT} plus a diode drop across each winding. With a proper choice of reference connection as shown, the second (right-hand) winding can generate an additional supply voltage (-5V in this case).

V_{OUT} (pin 8) is the feedback connection. For stability, the regulated output (5V in this case) should have the heavier load. It usually does, because the negative rail in most systems is only a bias supply. But, if your system demands more load current from the -5V output, you should reconnect the second winding to produce the 5V output, as shown in **Figure 1b**.

The transformer should have side-by-side bifilar windings for best coupling, but an off-the-shelf (non-bifilar) transformer such as the Coiltronics CTX20-4 gives acceptable performance (**Table 1**). The V-value (nominally -5V) depends on the load currents and the transformer turns ratio (which may deviate from 1:1). Loads of 5mA to 50mA @ V- and 50mA @ 5V, for example, cause a V- change of less than 300mV—less than that expected from a charge pump. When unloaded, V- increases due to the rectification of ringing that occurs when D2 turns on.
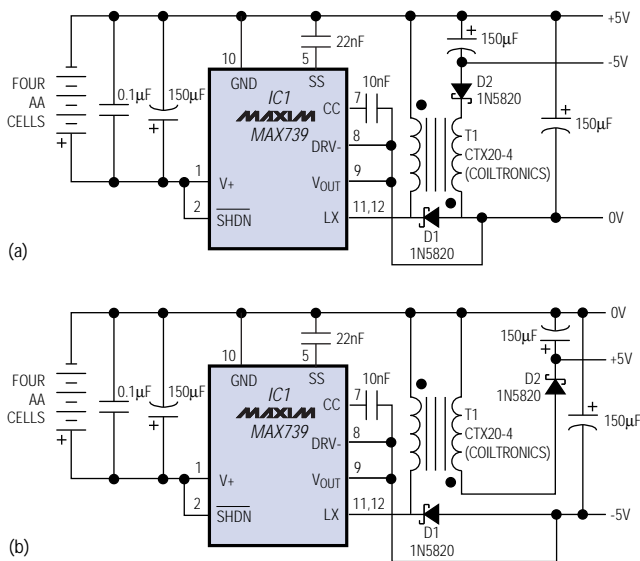
(Circle 2)

Figure 1. A 1:1 transformer enables this inverting switching regulator to generate -5V (a). Different connections for the right-hand coil are recommended if -5V supplies the heavier load current (b).

Table 1. V_{OUT} (V+ and V-) vs. V_{IN} and R_{LOAD} for Figure 1b

INPUT VOLTAGE (V)	INPUT CURRENT (mA)	V+ LOAD (Ω)	V+ (V)	V- LOAD (Ω)	V- (V)	EFFICIENCY (%)
6	1.68	NONE	5.07	NONE	6.55	—
6	62	100	5.08	NONE	10	69.37
6	68	100	5.08	1000	5.68	71.16
6	124	100	5.08	100	5.41	74.02
5	1.8	NONE	5.08	NONE	6.48	—
5	74.5	100	5.08	NONE	10	69.28
5	82.5	100	5.08	1000	5.69	70.41
5	151	100	5.08	100	5.42	73.09
4	1.8	NONE	5.1	NONE	7	—
4	95	100	5.1	NONE	10	68.45
4	105	100	5.1	1000	5.71	69.69
4	196	100	5.1	100	5.3	68.31

Note: 1000Ω load gives 5mA at output.
100Ω load gives 50mA at output.

DESIGN SHOWCASE

One-cell battery drives high-intensity LEDs

The forward voltage for high-intensity LEDs (1.5V to 2.5V) is too large for operation with one-cell batteries. The **Figure 1** circuit overcomes this limitation with a boost-regulator technique—it drives controlled current pulses through the LED regardless of the LED's forward voltage (**Figure 2**), and operates on input voltages from 6.2V to below 1V. The circuit is useful for bicycle lights, beacons, alarms, flashlights, and low-power indicators.

IC1 is normally part of a regulated boost converter, but in this case it simply transfers energy without regulating the output. Omission of the usual rectifier and output filter capacitor makes a compact circuit, as does the high switching frequency (about 175kHz), which enables the use of a tiny surface-mount inductor measuring only 3.2mm by 4.6mm by 2.6mm high.

Programming resistor R1 sets the LED intensity by setting a peak current for the inductor and LED—an approach that avoids the power dissipation and intensity variation (vs. battery voltage) associated with a series current-limiting resistor. A 10kΩ value for R1 sets the approximate peak at 75mA, and the average LED current at about 26mA.

A shutdown command turns off the OUT terminal completely, even if cell voltage exceeds the LED's forward voltage, by turning off the diode internal to IC1. (During shutdown, most step-up converters exhibit a troublesome dc path from the battery through the coil and diode to the load.) This circuit draws about 8µA during shutdown and about 60mA during normal operation. It operates for 35 hours continuously on one AA (or R4 size) alkaline cell.
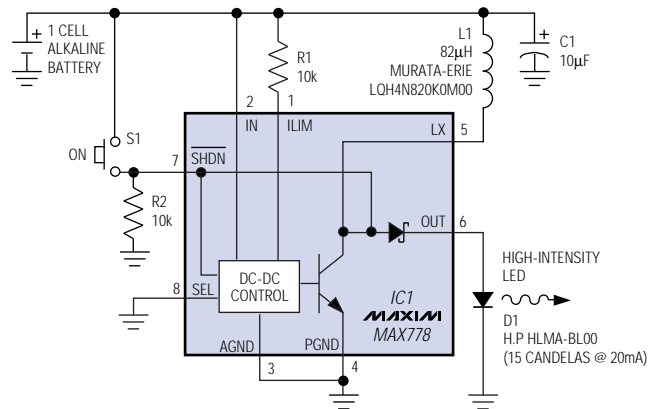
(Circle 3)

Figure 1. Unlike most dc-dc boost converters, this circuit drives high-intensity LEDs from a single cell without rectifiers or filter capacitors.
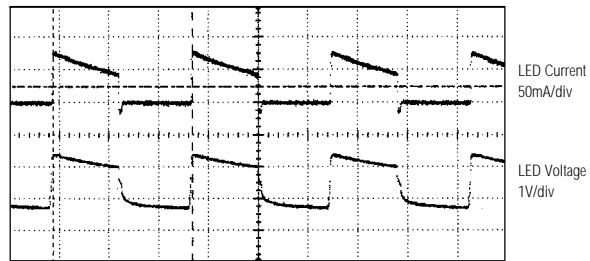
Figure 2. These waveforms for LED voltage and current differ from those of a typical dc-dc converter, but their high frequency assures an absence of visible flicker.

Analog switches operate with 3V or 5V supplies

By adding a single component to a 3V-only or 5V-only board, you can operate conventional CMOS analog switches with performance approaching that specified with ±15V supplies. This means fast switching, low on-resistance, CMOS/TTL compatibility, low power consumption, and a signal range ($\pm V_{CC}$) that exceeds the input supply range (V_{CC} to ground).

Simply add a charge-pump voltage converter (IC1), which produces $\pm 2V_{CC}$ outputs from a V_{CC} input. These unregulated voltages ensure reliable switch operation for V_{CC} levels as low as 3V. Logic thresholds for the switch remain unaffected.

A V_{CC} of 3V (for instance) produces ±6V rails for the switch (IC2), resulting in on-resistance <30Ω, switching times <200ns, leakage <0.1nA, and I_{CC} current <0.5mA. Raising V_{CC} to 5V produces ±10V rails, resulting in on-resistance <20Ω, switching times <150ns, leakage <0.4nA, and I_{CC} current <1.3mA.

IC1 can easily power additional switches and/or low-power op amps, but more than a few milliamps of load current degrades performance by lowering the unregulated supply rails.
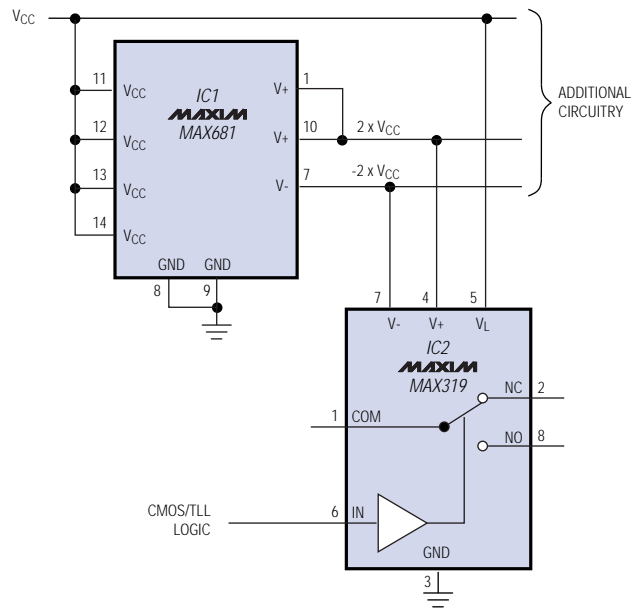

Figure 1. The charge pump (IC1) provides a local bipolar power supply for the CMOS analog switch (IC2).

(Circle 4)

Quick-charge IC routes step-up switcher output to battery

Simplicity and high efficiency favor the step-down, switch-mode battery charger over other configurations, but sometimes the available supply voltage is too low for the battery's terminal voltage. Examples include charging a backup battery from the supply it backs up, and using an automobile battery to charge a large number of cells in series.

Such applications can be handled by a step-up switching regulator in conjunction with a fast-charge controller (**Figure 1**). This circuit can charge five or more NiCd or NiMH cells from a 5V input. Because cell voltages vary greatly during the charge cycle, you should resist the temptation to simply boost V_{IN} above the highest terminal voltage for the battery being charged (and then dropping the boosted

voltage back to battery level with a linear regulator). That approach is inefficient and generates heat.

This circuit also supplies load current while charging the battery. Because load current bypasses the battery and the R2 sense resistor, it has little effect on the charge current. (As an option to improve efficiency when the charger is not in use, you can greatly reduce the voltage drop across R2 by adding a low-$r_{DS(ON)}$ MOSFET in parallel.)

The IC1 connection shown, which holds its feedback input (pin 3) below the internal set-point, causes the step-up converter to act as a current source producing its maximum output current. Then, replacing the external current-sense resistor with a power MOSFET (Q2) allows IC2's DRV output to set this output current via the Q3-Q4 current mirror, which controls Q2's gate voltage and on-resistance.

To configure the overall circuit for a different range of input voltage, adjust R1 (if necessary) so the current into pin 15 of IC2 remains between 5mA and 20mA. This condition ensures that the shunt regulator in IC2 maintains the internal V+ at approximately 5V. If V_{IN} exceeds 16V, you should remove the direct connection to IC1 and power that chip with V+ from IC2 instead. Output power should not exceed 20W for the Q1 device shown, and R2 adjusts the charging current:

$$I_{CHARGE} = 0.25 / R2.$$

To charge batteries with other than six cells in series, re-program IC2 according to instructions in the data sheet. The zener voltage of D1, which prevents overvoltage on capacitor C1 if the battery is removed during charging, should approximate (in volts) twice the numerical cell count. C1 helps to smooth the otherwise chopped output current from IC1; take care not to exceed its ripple-current rating (the ripple current is approximately $I_{CHARGE} + I_{LOAD}$).
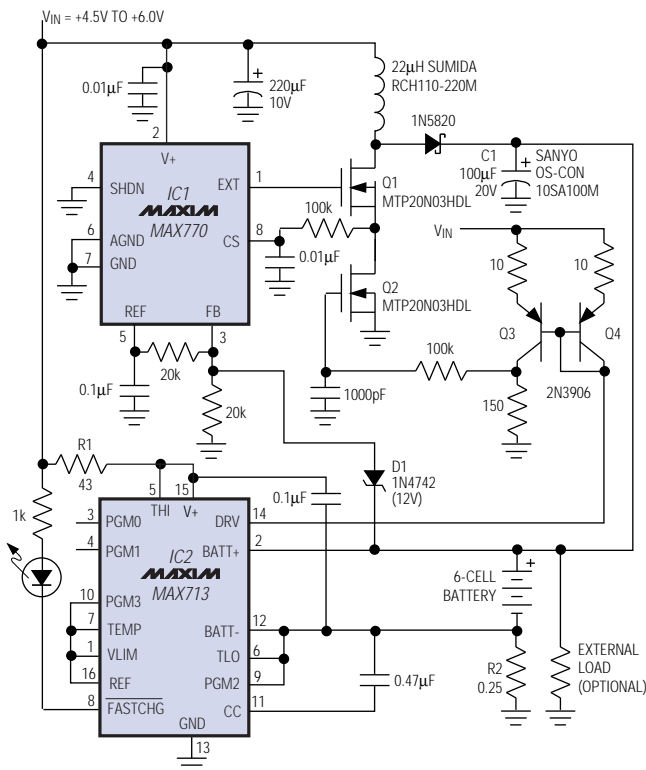

Figure 1. A step-up switching regulator (IC1) and fast-charge controller (IC2) charge multi-cell batteries with speed and efficiency.

(Circle 5)

Low-battery monitor delays system shutdown

The circuit of **Figure 1** gives an early warning of declining battery voltage. Then, to allow a controlling processor time for emergency housekeeping chores such as the storage of register data, the circuit delays system shutdown by a specified time interval (rather than waiting for battery voltage to decline further, to a specified lower level). Circuit components are chosen for low quiescent current, which protects discharged cells by minimizing the battery drain during shutdown: IC1 draws 1μA, IC2 draws 3μA, and R1/R2 draw 3μA, for a total shutdown current of about 7μA.

Also vital to the application is the tight tolerance (±1%) on IC2's comparator threshold, which allows precise monitoring of the NiCd battery's flat discharge characteristic. Positioning the low-battery warning right at the knee of this discharge curve enables a maximum extension of battery life.

IC1 is a low-dropout linear regulator that supplies 250mA of output current and drops only 350mV at 200mA. IC2 is a combination dual comparator and ±1%-accurate voltage reference. When V_{BATT} falls below the threshold set by R1 and R2, OUTB (pin 8 of IC2) goes high. This high level serves as a low-battery warning while charging C1 through R3. When the INA voltage at pin 3 reaches the internal reference level (1.182V ±1%), OUTA (pin 1) issues the shutdown command to IC1.

As an example, set the voltage threshold to 0.9V per cell for a 6-cell stack (5.4V). Then, 5.4V[R2 / (R1 + R2)] = 1.182V. Let R1 = 1MΩ; R2 then equals 280kΩ. Use 287kΩ. You can add ±25mV of hysteresis to this threshold by setting R4 = 49.9kΩ and R5 = 2.4MΩ, as explained in the MAX923 data sheet.

Assume 1MΩ for R3, then calculate C3 using the following equation:

$$V_{TH} = V_{OUTB}(1 - e^{-t/\tau})$$

where V_{TH} is the threshold voltage, V_{OUTB} is the output of the internal comparator (assume 4.9V), and τ = R3C1. Solving this equation for a one-second delay (t = 1) yields τ = 3.6sec. Therefore, C1 = 3.6μF.

As an alternative, you can choose a standard value for C1 such as 3.9μF, which also yields a delay time of about one second. A good low-leakage capacitor for this application is the surface-mount Novacap (p/n 1825Z395K250 for 3.9μF). Note that C1 must be fully discharged for the circuit to provide the full delay. C1 becomes charged while the system is in shutdown, and then requires about 6 seconds to discharge completely.
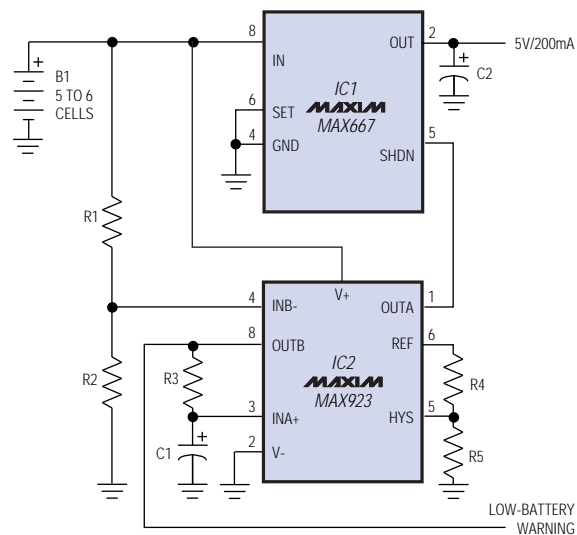
(Circle 6)

Figure 1. *This 12V regulator issues a warning when the battery voltage is low, and shuts itself down approximately one second later. The shutdown current is about 7μA.*

NEW PRODUCTS

Fast 16-bit ADC has 10μA shutdown

The MAX195—a serial-output, sampling A/D converter (ADC)—requires less operating power than any other 16-bit, 16-pin ADC. It requires only 80mW maximum, even at the maximum throughput of 85ksps. At lower sample rates it can reduce power dissipation even further by entering a 10μA shutdown mode between samples.

The MAX195 comes in 16-pin SOICs and plastic DIPs, which occupy considerably less board space than the 28-pin SOICs and 44-pin PLCCs usually associated with high-speed 16-bit ADCs. Thus, the MAX195 is ideal for portable instrumentation, medical equipment, and other applications for which battery life and small size are critical.

Internal self-calibration circuitry achieves 16-bit linearity and monotonicity by cancelling the internal offset voltage and by fine-adjusting the capacitor values on an internal, binary-weighted capacitive D/A converter. Calibration occurs during power-up and whenever commanded via the RESET input.

The MAX195 operates on ±5V, with separate analog and digital supply terminals that minimize the effects of digital noise. The input circuit is pin-selectable for unipolar (0V to $+V_{REF}$) or bipolar (-V_{REF} to $+V_{REF}$) inputs. Separate terminals for the main and serial clocks allow two modes of operation: the offset-binary, serial-data bits can be transmitted as generated, or stored and shifted out (on the DOUT terminal, at rates to 5MHz) during the next signal acquisition.

The MAX195 is available 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 $24.00 (1000 up, FOB USA). **(Circle 7)**

85ksps with 5V or ±5V Input Ranges
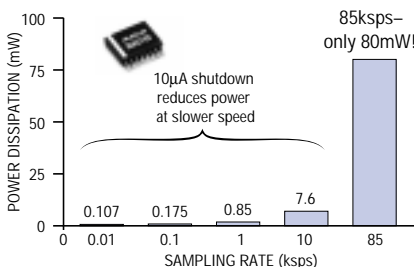
Multi-range, overvoltage-tolerant 12-bit DAS has 100ksps throughput

The MAX197/MAX199 12-bit data-acquisition systems (DAS) operate on single 5V supplies, yet the MAX197 accepts analog inputs that extend below ground and above the supply rail. This capability enables it to interface with sensors powered by ±12V, ±15V, and 4mA-to-20mA current loops. Analog ranges for the eight input channels can be programmed independently and "on the fly" to accommodate differences in signal level from channel to channel. MAX197 ranges are ±5V, ±10V, 0V to 5V, and 0V to 10V; MAX199 ranges are ±V_{REF}/2, ±V_{REF}, 0V to V_{REF}, and 0V to V_{REF}/2. Switching from the ±10V range to the 0V-to-5V range, for example, gives an effective dynamic resolution of 14 bits.

Each device includes an 8-channel multiplexer, 5MHz track/hold, 12-bit A/D converter, and 4.096V reference. Each is "overvoltage tolerant," meaning the selected channel is not perturbed by inputs to ±16.5V on the idle channels, nor does overvoltage affect supply current in the power-down modes. Typical operating power is 30mW.

Operation is software-programmable via an 8-bit, three-state, bidirectional data port and standard microprocessor (μP) interface. An 8-bit control word selects the input channel (one of eight), input range (two unipolar and two bipolar), power-down mode (one of two), internal or external clock, and whether the sample acquisitions are triggered internally or externally.

Both ICs include reference-buffer amplifiers that allow ±1.5% adjustments. Both produce a standard interrupt signal that alerts the μP when a conversion ends and data is available. Data-access and bus-release timing is compatible with most popular μPs, and all logic inputs and outputs are TTL/CMOS compatible.

The MAX197 and MAX199* come in 28-pin DIP, SO, and SSOP 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. **(Circle 8)**


* Future Product
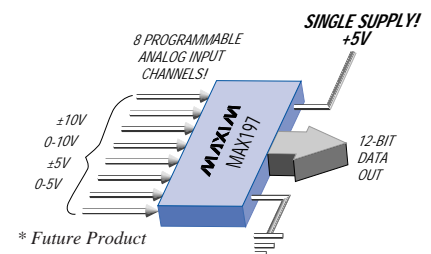
Analog switches suit battery-operated systems

The MAX320–MAX325 dual SPST (single-pole/single-throw) analog switches are designed for either unipolar or bipolar power supplies. The MAX320/MAX321/MAX322 operate from ±3V to ±8V, and the MAX323/MAX324/MAX325 operate from 2.7V to 16V. All have low leakage (100pA), fast switching (150ns t_{ON}, 100ns maximum t_{OFF}), low charge injection (5pC), and ESD protection to 2000V per MIL-STD-883, Method 3015.7.

The MAX320 has two normally open (NO) switches and the MAX321 has two normally closed (NC) switches. The MAX322 includes one NO and one NC switch. Powered from ±5V supplies, they offer 1.25mW power consumption, 35Ω maximum on-resistance, 2Ω (max) matching between channels, and 4Ω (max) variation over the allowed signal range within a channel.

MAX323 switches are NO, MAX324 switches are NC, and the MAX325 has one NO and one NC switch. Powered from a 5V supply, they have extremely low power consumption (5μW), 60Ω maximum on-resistance, 2Ω (max) matching between channels, and 6Ω (max) variation within a channel.

MAX320–MAX325 switches come in 8-pin DIP, SO, and μMAX 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 $0.99 (1000 up, FOB USA). **(Circle 9)**
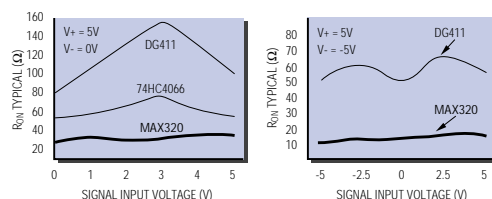
NEW PRODUCTS

Fast-charge controller charges NiCd and NiMH batteries

The MAX2003 applies discharge conditioning and fast charging to nickel-cadmium (NiCd) and nickel-metal-hydride (NiMH) batteries. When configured as a switch-mode current regulator, its efficient energy transfer minimizes power dissipation. When configured as a gating controller for an external current source, it minimizes the component count.

Stored algorithms enable the MAX2003 to serve as a stand-alone charger. These algorithms provide five independent methods for terminating a fast charge:

temperature slope, negative-delta voltage, maximum temperature, maximum time, and maximum voltage. To ensure safety, the chip inhibits initiation of a fast charge until the battery's voltage and temperature are within acceptable limits. When configured for the appropriate method of charge termination, a single MAX2003 circuit can fast-charge both NiCd and NiMH batteries.

The MAX2003 also provides a switch-activated discharge-before-charge option that preconditions the battery. Other features include an optional "top-off" charge after the

fast charge, and LED-driver outputs that indicate charge status and acceptable-temperature status.

As a direct replacement for the bq2003, the MAX2003 is available in 16-pin DIP and wide-SO packages, tested for the commercial (0°C to +70°C) temperature range. It also comes in a space-saving narrow-SO package. An evaluation kit (MAX2003EVKIT-SO) saves time and design effort. Prices for the MAX2003 start at $3.25 (1000 up, FOB USA).

(Circle 10)

Smallest-available isolated power driver fits Type I PCMCIA cards

- *450kHz switching frequency permits 1.5mm transformer height*

The MAX845 isolated power driver comes in an ultra-small μMAX package and drives miniature transformers with profiles as low as 1.5mm. Output filtering (when necessary) is simplified by alternate switching of the internal power MOSFETs, whose combined load current—interrupted only by short (100ns) break-before-make intervals—is nearly constant. The result is lower values of input-current ripple and output-voltage ripple.

The MAX845 can drive small transformers because its switching frequency (450kHz minimum) is high. Because the frequency varies linearly with V_{CC}, the chip

applies to the transformer an ET product that is constant at 5V-μs. (ET is the time a given voltage can be applied before the transformer saturates.) By maintaining a tightly controlled ET product, the MAX845 delivers maximum power to the primary of a minimum-size transformer, without risk of saturation.

The transformer's driven primary has a center tap connected to a dc supply voltage of 3.3V or 5V, and its secondary can be wound to produce positive or negative isolated outputs capable of delivering 750mW. A logic-controlled shutdown mode lowers the 1mA quiescent current to only 0.4μA.

A fully assembled evaluation kit (MAX845EVKIT-MM), which includes a sample of the recommended transformer, is available as an aid in design and evaluation. The MAX845 comes in 8-pin μMAX and SOIC packages specified for the extended-industrial (-40°C to +85°C) temperature range. Prices start at $1.63 (1000 up, FOB USA).

(Circle 11)

MAX845: Smallest Isolated DC-DC Converter
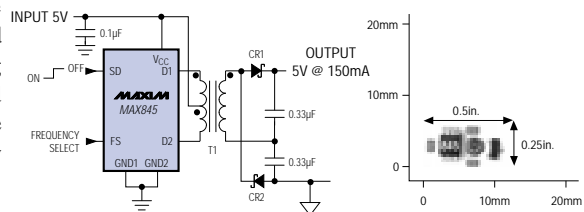
200mA, 1.5W linear regulators are world's smallest

Low dropout voltage and ultra-low supply current enables the MAX882/MAX883/MAX884 linear regulators to extend battery life. They deliver up to 200mA (even for junction temperatures to +125°C), and come in 8-pin SOIC packages that dissipate as much as 1.5W. Supply currents are 15μA maximum over temperature, regardless of output current.

Supply currents remain low because the internal MOSFET requires no base current, unlike a conventional regulator's pnp pass transistor, which saturates and draws excessive base current as V_{IN} approaches the level of V_{OUT}. MOSFETs also provide low dropout voltage—the lowest input-output differential for which regulation can be sustained. MAX883 on-resistance at 5V

is 1.1Ω, so for 200mA I_{OUT} the typical dropout voltage is only 220mV. At 3.3V and 200mA the typical dropout is 320mV.

The MAX882 offers a 7μA standby mode that disables V_{OUT} but keeps alive the biasing circuitry and the low-battery comparator. MAX883 and MAX884 regulators offer a complete shutdown that turns off all circuitry, reducing their supply currents to less than 1μA. All three devices include a low-battery detector, short-circuit and reverse-current protection, and thermal shutdown.

MAX882 and MAX884 output voltages are preset at 3.3V; the MAX883 is preset at 5V. All three employ Maxim's

Dual-Mode™ operation, which allows the use of external resistors to set (in addition to the preset levels) any V_{OUT} level between 1.25V and 11V. (As higher-I_{OUT} alternatives, MAX603/MAX604 linear regulators can deliver 500mA.)

The MAX882/MAX883/MAX884 are available in 8-pin DIPs, as well as a special high-power, 8-pin SO package that can dissipate 1.5W at +70°C (vs. 0.47W for a conventional package). The devices come in versions tested for the commercial (0°C to +70°C) and extended-industrial (-40°C to +85°C) temperature ranges. Prices start at $1.45 (1000 up, FOB USA).

(Circle 12)

The MAX882 Family: Lowest Supply Current at All Loads
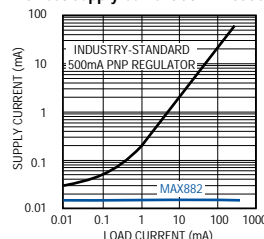
NEW PRODUCTS

20µA RS-485 transceivers (as many as 256) share one 2-wire bus

The full-duplex MAX1482 and half-duplex MAX1483 are low-power transceivers suitable for both RS-485 and RS-422 communications. Each includes one receiver and one driver, whose output is slew-rate limited to minimize EMI and the line reflections caused by improperly terminated cables. Data rates are guaranteed to 250kbps.

The transceivers operate on 5V and draw 20µA supply currents. Shutdown lowers the supply current to 0.1µA. The receiver input impedances guarantee 96kΩ minimum, which is 1/8 of a unit load as defined by the RS-485 standard. RS-485 allows up to 32 unit-load transceivers (each with 12kΩ input impedance) on a single cable; therefore the cable can support as many as 256 MAX1482 or MAX1483 transceivers.

Short-circuit protection for the driver outputs includes current limiting and thermal-shutdown circuitry that places the outputs in a high-impedance state. The receivers guarantee logic-high outputs when their input circuits are open.

The MAX1482 comes in 14-pin DIP or SO packages, and the MAX1483 comes in 8-pin DIP, SO, or µMAX (half the size of an 8-pin SO) packages. Both transceivers are available in versions tested for the commercial (0°C to +70°C) and extended-industrial (-40°C to +85°C) temperature ranges. Prices start at $1.25 (1000 up, FOB USA).

(Circle 13)

Reduce RS-485 Supply Current to 20µA
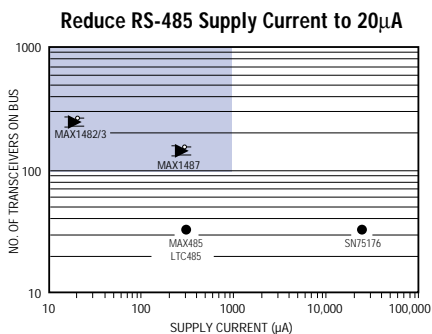
## 2.5Msps RS-485 transceivers (as many as 128) share one 2-wire bus

The half-duplex MAX1487 is a low-power, 2.5Mbps serial-data transceiver for RS-485 and RS-422 systems. It operates on 5V and incudes one driver and one receiver. The receiver's guaranteed 48kΩ input impedance ($1/4$-unit load as defined by the RS-485 standard) allows one pair of wires to carry as many as 128 transceivers—four times the number allowed with a standard RS-485 transceiver of 12kΩ unit load. The driver's output slew rate enables data transmissions to 2.5Mbps.

When either unloaded or fully loaded but with drivers disabled, the MAX1487 draws just 230µA of quiescent current. Its three-state driver output is short-circuit protected and includes thermal-shutdown circuitry to prevent excessive power dissipation. As a fail-safe feature, the receiver output guarantees a logic high when its input circuit is open.

The MAX1487 is screened for the commercial (0°C to +70°C) and extended-industrial (-40°C to +85°C) temperature ranges. It comes in an 8-pin DIP or SO package, or in a µMAX package, which occupies only half the space of an 8-pin SO. Prices start at $1.25 (1000 up, FOB USA).

(Circle 14)
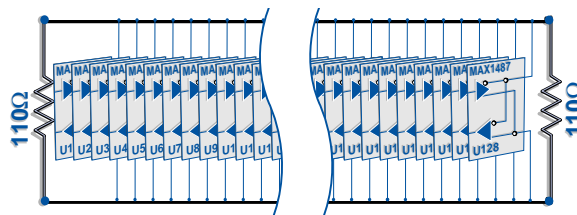
5V, fiber-optic transceiver chipset handles data rates to 1Gbps

The MAX3260/MAX3261/MAX3262 devices comprise a complete chipset for fiber-optic transceivers, for use in 1063Mbps Fibre Channel and 622Mbps Sonet/SDH (synchronous optical network/synchronous digital hierarchy) applications. The three chips (preamplifier, postamplifier, and laser driver) operate on a single 5V supply.

The MAX3260 is a high-speed transimpedance amplifier whose extended dynamic range is useful in optical receiver systems with as much as 25dB of input signal range. It consumes only 115mW when the output is ac coupled. With a 50Ω output termination, the consumption is 300mW.

The MAX3261 is a 1.2Gbps laser-diode driver whose complementary enable inputs provide an interface with open-fiber-control (OFC) architecture—a feature not found in other 1.2Gbps laser-diode drivers. For transmitters that include a monitor photodiode, the chip's automatic power control (APC) circuit maintains constant laser power despite changes in laser efficiency. Other MAX3261 features include a TTL-compatible indicator for laser failure, a programmable slow-start circuit that prevents laser damage, and a temperature-compensated reference voltage that provides simple programming of the laser current.

The MAX3262 limiting amplifier's high gain and wide bandwidth suit the device for use as a post amplifier in fiber-optic receivers with data rates to 1Gbps. Gain is adjustable from 33dB to 48dB. At 48dB, the MAX3262 can amplify signals as small as 6mVp-p, and its outputs are PECL-compatible (i.e., Pseudo-ECL, which operates between 0V and 5V) Complementary loss-of-signal (LOS) outputs interface with the OFC circuitry and provide chatter-free detection of signal loss. LOS hysteresis for any programmed level (nominally 2.5dB) preserves a balance between noise immunity and dynamic range. The chip's offset-correction function reduces pulse-width distortion.

MAX3260, MAX3261, and MAX3262 devices are available in die form for the commercial (0°C to +70°C) temperature range. Prices for the MAX3260 start at $9.25 (100 up, FOB USA).

(Circle 15)