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## LT1251/LT1256 Video Fader and DC Gain-Controlled Amplifier

by William H. Gross

### Introduction

Variable-gain amplifiers are used extensively to provide effects and level adjustments in video systems. The simplest and most common effect is the fade to black. This is created by gradually reducing the signal gain to zero. Another, more specialized kind of video mixing is the dissolve, where one signal is reduced while another is increased. The result is the familiar fading of one scene into another. The dissolve is accomplished by a circuit block called the fader. The fader circuit has two signal inputs, a control input, and a single output. Of course, if the second input signal is zero, the scene fades to black.

The ideal fader would have well-defined gain versus control voltage, sufficient bandwidth for video, low distortion, and a DC output level that is independent of the control signal. The most difficult design issue in making a fast, variable-gain amplifier is obtaining good gain accuracy; the gain must be linear with respect to the control signal and it must not change with temperature.

The LT1251/LT1256 is a two-input, one-output current-feedback amplifier with a linear control that sets the amount each input contributes to the output. The gain-control circuitry is trimmed for absolute accuracy at wafer sort and is completely temperature compensated.

### The Video Fader

A block diagram of the basic fader is shown in Figure 1. The control signal varies the position of the potentiometer such that at one extreme the output contains only IN1 and at the other extreme it contains only IN2. The control is linear; i.e., for the control signal at 50%, the output is the sum of one half of IN1 and one half of IN2. If both inputs are the same, the output is independent of the control signal. The design of the controlled potentiometer is the most challenging aspect of fader design.

For zero or 100% control input, it is important that the video fader completely attenuates the appropriate input signal. Most of the time the fader is operating with only one input signal at the output. In this condition, the off signal must not bleed

*continued on page 27*

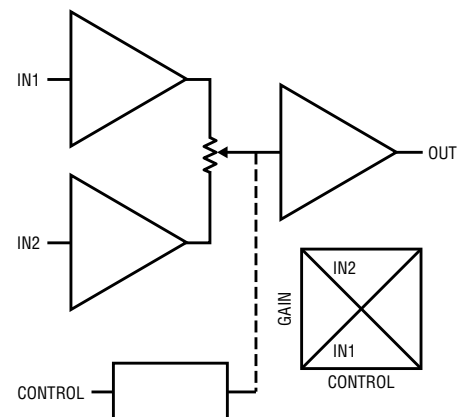


Figure 1. Basic fader circuit

# Morphing the Factory Applications Staff

by Rich Markell

Much has been written of late about the popular visual effect called "morphing." An example is when Odo on Star Trek exquisitely turns himself into a small mouse and scampers around the space station unnoticed, then returns as Odo, with the evidence he needs to convict Quork of running a gambling operation.

Linear Technology is not (yet) in the business of creating special effects for Hollywood, but as time goes by and more engineers are added to the factory applications staff, the personality mix becomes more and more diverse. Suppose the personalities of the factory Application Engineers were all metamorphosed into one. How would this event be perceived on the phone or in person? What would it be like?

You'd probably step out of your office to meet a nice man who'd have pens, calculators, and barometric sensors in one pocket and, perhaps, gardening tools in the other. He'd probably be wearing a multicolored propeller beanie, but the propeller would be partially broken off and have CK722's or nuvistors attached to it. The vision fits and is only partly a joke. When you finally sit down to

discuss circuits with our morphed "all-around engineer," you'd find someone with a broad knowledge of LTC's product line. He could help you design the world's smallest, flattest, most efficient switcher or a "Better-than-Bessel" filter. He could evaluate your HDSL application and suggest an A/D to fit, or tell you which video mux to use to switch between HDTV sumo wrestling and luge. Finally, at the end of the day, perhaps he'd morph back into a mouse and crawl into Williams' 547 for the night.

Our lead article in this issue features a great new video-product family, the LT1251/LT1256. These products incorporate a 30MHz video fader and a DC gain-controlled amplifier into a single IC. The parts have greater than 80dB signal-to-noise ratios with good differential phase and gain response. These devices are perfect for new multimedia computer boards, video products for both the professional and the consumer, and a variety of other circuits. We also introduce a new micropower DC/DC converter, the LT1302. Designed for battery-powered applications, no other part can deliver so much power from an SO-8 package. The LT1302

operates from two, three, or four cells and can provide up to 600 milliamps at 5 volts or up to 1 amp from a 3.3 volt supply.

Also featured in this issue is the LTC1152. Designed in Singapore, the LTC1152 is a zero-drift, rail-to-rail input, rail-to-rail output swing operational amplifier. The part operates from 2.7V to 14V of total supply voltage. The amplifier has a 1MHz gain-bandwidth product and can plug directly into any 8-pin op amp socket. The LTC1066-1, introduced in these pages, is the first monolithic filter that combines RC active techniques with switched-capacitor technology. The LTC1066-1 is an eighth-order, elliptic lowpass filter with only 1.5 millivolts (max.) DC offset and 14 bits of DC gain linearity.

ADCs are again prominent in this issue, as we highlight the LTC1285 and LTC1288. These two converters are small, SO-8 12-bit converters with serial interfaces for digitizing sensors or pen screen inputs, and for use in cellular phones.

We conclude with a palette of Design Ideas in this issue as well as a variety of New Device Cameos.

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## **FAE Cameo: Jon Dutra**

LTC now has twenty-two Field Application Engineers (FAEs) worldwide to assist our customers however possible. Jon Dutra is one of two FAEs in our Silicon Valley Northwest Sales Office. He now covers Northern Nevada, Idaho, and San Francisco Bay Area customers with company names beginning in the letters A-K.

Early in his career, Jon designed thermocouple measurement systems, data-acquisition systems, switching power supplies, and video-speed analog systems. He has been an active

ham radio operator since age 14 and is a licensed but inactive pilot.

In over five years with LTC he has had many interesting experiences. "I love working with customers, helping them solve their engineering problems in elegant, cost effective ways," he says.

Currently, Jon is spending about 50% of his time with regulator issues; interface products, filters, op amps, references, and A/Ds consume the remainder. When Jon is not in the field, he can often be found in the lab, building and testing some new cir-

cuit, sometimes just for fun, but usually for a specific customer. One such breadboard turned into a \$1.7 million order and helped him earn the new award of LTC FAE of the Year.

Jon and his wife Barbara have been married for eight years. They have two children, Thomas, 6, and Brittany, 5, who keep their parents very busy. He enjoys tennis, gardening, building "things," and playing with his children. Jon can be reached through the LTC Northwest Sales office listed on the back of this magazine.

**LT**

# The LTC1152 Rail-to-Rail Operational Amplifier

by Dave Dwelley

## Introduction

Over the past few years, the term “rail-to-rail” has become a common phrase in op-amp advertising. Generally, this implies an output stage that can swing to within millivolts of either power supply. Many CMOS and BiCMOS parts make this claim; there are even a few all-bipolar designs that can come within a saturation voltage of the rails (approximately 0.7 volts), close enough to rail-to-rail for most designers. Some of these parts include inputs whose common-mode ranges include the negative power supply rail; this generally earns the op amp the additional tag line “single supply.” All this is well and good, but it obscures an important point: nearly all the parts boasting “rail-to-rail” performance don’t include input common mode to the positive rail.

## Rail-to-Rail Input CMR

Common-mode range (CMR) is one of those specs buried deep in the data sheet that few people look at. Perhaps, but rail-to-rail input common-mode range is the distinction that separates true rail-to-rail op amps from plain vanilla op amps with fancy output stages. Take one of the more common op-amp circuits, the unity-gain follower (Figure 1). Most designers are familiar with this circuit for a couple of reasons; there are no pesky resistor-ratio formulas to figure the gain and you get predictable performance as long as you remember to pick a unity-gain-stable op amp. There is, however, a hidden

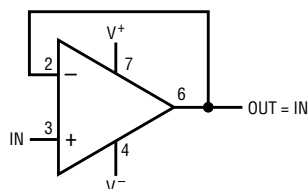


Figure 1. Unity-gain follower

trap with this circuit. Operating from a single 5V supply, most “single supply, rail-to-rail” op amps will follow the input all the way to ground. However, as we approach the positive supply, the output stops swinging! We’re not even close to the rail yet!

What happens in this situation is that the input exceeds the op amp’s CMR long before the output gets to the rail. When op amps ran from  $\pm 15V$  supplies, input common-mode range generally ran out a couple of volts away from either rail, but no one seemed to mind too much; giving away 3 or 4 volts out of 30V of possible input swing didn’t seem like too much of a compromise. That same op amp running from a single 5V supply now has only 1V of CMR remaining—suddenly it’s very significant. Newer parts that can common mode to V– get the bottom end of the range back, but most run out of CMR around 1.5V below V+. That’s still a large portion of the total input range; such a part running from a single 2.7V supply ( $3V \pm 10\%$ ) gives away more than half of its possible input range. To further aggravate the situation, many op amps get weird when you exceed their common-mode ranges; the front-end devices can turn completely off, causing second-order effects that can do strange things to the output stage

and cause problems with the feedback loop. To make a true rail-to-rail follower, you need a rail-to-rail input CMR, rail-to-rail output-swing op amp that works over a wide range of supply voltages—like the LTC1152.

## The Secret

The LTC1152 is a CMOS, zero-drift, rail-to-rail input CMR, rail-to-rail output-swing operational amplifier that will work from 2.7V to 14V total supply voltage. It achieves rail-to-rail input CMR by using a self-contained charge pump to generate an internal voltage regulated to about 2V higher than V+. This allows it to use a conventional PMOS front-end structure running from this internal supply (Figure 2). Input signals at V+ are still well below this internal supply, allowing the front end to amplify them without level shifting or extra front end devices. This same front end inherently includes V– in its common-mode range, allowing full rail-to-rail CMR from a traditional front-end structure. CMR typically extends about 0.3V beyond either rail before leakages start to form across the parasitic clamp diodes and affect the input impedance.

The charge pump and all of its support circuitry, including the capacitors, are included on the die; no

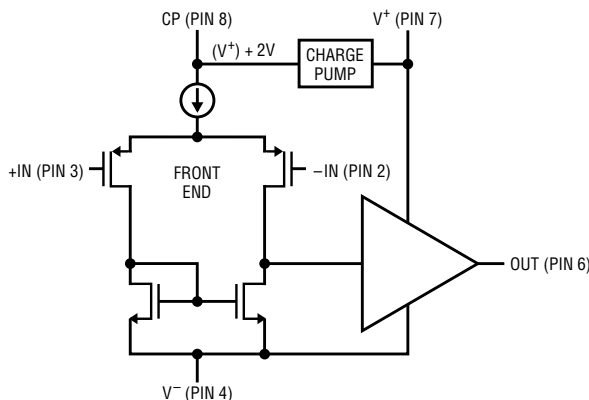


Figure 2. LTC1152 functional diagram

external components are required. Additionally, the charge pump typically runs at 4.7MHz, well above the 1MHz gain bandwidth of the LTC1152; this ensures that very little charge-pump feedthrough actually reaches the output pin. Sensitive applications can further reduce feedthrough by connecting an external bypass capacitor between the charge pump output at pin 8 and V+ at pin 7; a 0.1 $\mu$ F cap will pretty much wipe out any remaining charge-pump noise at the output. The self-contained charge pump also allows the LTC1152 to conform to the industry-standard op-amp pinout. It can plug into any standard 8-pin op-amp socket, either standard DIP or SO8, provided any trim circuitry on pins 1, 5, and 8 is removed. The charge pump is necessarily quite small to fit inside the package; as a result, it's not a good idea to try to drive any external loads (other than bypass capacitors) from pin 8. By the same token, don't connect the bypass capacitor to ground or V-; the more volts the charge pump has to put into the capacitor, the longer it will take for the LTC1152 to start up.

**The Rest of the Secret**

In addition to the charge-pump front end, the LTC1152 has a zero-drift architecture, adapted from LTC's family of zero-drift op amps. Like the other family members, the LTC1152 constantly corrects its own offset and drift errors for optimum DC performance. The net result is an offset voltage spec of 10 $\mu$ V maximum, 100nV/ $^{\circ}$ C maximum drift over temperature, minimum PSRR of 105dB over temperature, and minimum 115dB CMRR over temperature, all over the entire rail-to-rail input range. Additionally, the LTC1152 maintains this performance over a single-supply voltage range of 2.7V to 14V, or a dual-supply range of  $\pm$ 1.35V to  $\pm$ 7V, allowing it to run from most standard digital supplies as well as split analog supplies. In other words, a follower built with an LTC1152CS8 will have an output voltage guaranteed to be

within 25 $\mu$ V of the input, in Singapore or in Lillehamer (over temperature), all errors accounted for, with almost any power supply that's handy, until the output runs into either power-supply rail. Try that with any other op amp!

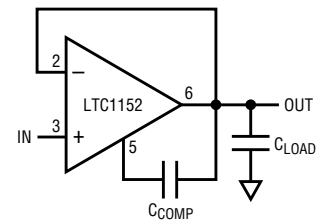
As with all zero-drift op amps, the LTC1152 pays for its exceptional DC performance by exhibiting aliasing behavior at its internal clocking frequency. In the case of the LTC1152, that clocking frequency is about 2.3kHz (actually, it's exactly the charge pump frequency divided by 2048). AC input signals near this frequency will generate aliasing products, with their magnitude dependent on the closed-loop gain of the circuit configuration. As a rule of thumb, the aliasing products will be about (80dB - the closed-loop gain) below the input signal. In the case of a unity-gain follower with the closed-loop gain of 0dB, the sum and difference frequencies will be attenuated by about 80dB—pretty far down. For higher gain configurations, they may be more prominent. Because the zeroing clock is divided down from the charge pump oscillator, there are no interference products or "beat frequencies" between the charge pump and the auto-zero circuit. Many applications, especially those with relatively low gain, can use the LTC1152 as a wideband amplifier all the way up to its 1MHz gain-bandwidth product, without ever noticing that it is an autozeroed amplifier.

**Rail-to-Rail Output**

The LTC1152's input stage wouldn't be much use without an equally rail-to-rail output stage. The output stage is powered off the hard power supply rails, not the internal charge pump; although this prevents it from swinging outside the rails (an LT1026 charge-pump chip can be used if you need to do that), it allows the output to provide much more current than could be supplied with an all-internal charge pump. The LTC1152 output will swing right up to either rail when unloaded; the

open-loop output impedance is about 190 $\Omega$ , limiting the output swing with load to that of the resistor divider formed between this 190 $\Omega$  and the load impedance. Very high value load resistors allow the output to swing closer than a millivolt to either rail; a 1k load will swing to about +4.2V with a 5V supply. The output current is limited to about  $\pm$ 20mA under short-circuit conditions.

Single-supply applications with the load referenced to ground will swing all the way to ground due to the pulldown effect of the load; the only error is the half-wave rectified input noise amplified by the closed-loop gain of the circuit. This can be as low as 1 $\mu$ V for unity gain circuits—that's pretty close to ground. Similarly, circuits with the output loaded to V+ will swing all the way to V+.



**Figure 3. Externally compensating the LTC1152**

The output stage is optimized to drive capacitive loads up to about 10,000pF. Larger capacitive loads can be driven by externally compensating the LTC1152 (Figure 3). Connecting a 1000pF capacitor from pin 5 (COMP) to pin 6 (OUT) ensures unity-gain stability with loads up to 1 $\mu$ F; 0.1 $\mu$ F between COMP and OUT allows the LTC1152 to drive as much capacitance as you can put on it. The trade-off is speed; large output compensation capacitors work by moving the compensation pole lower in frequency, directly affecting the gain-bandwidth product. With a 1000pF compensation capacitor, the LTC1152 has a gain-bandwidth product of about 20kHz.

## Shutdown Mode

The LTC1152 includes a shutdown feature that disables the part, puts the output into a high-impedance state, and drops the supply current from 2.2mA to about 1 $\mu$ A. All this is accomplished by pulling pin 1 low, either with CMOS logic running from the same supplies as the LTC1152 or with an open collector/open drain device. Additionally, the shutdown pin thresholds are designed so that, when the LTC1152 is run from dual supplies, CMOS logic running from the same positive supply and ground can interface directly to the shutdown pin. Pin 1 includes an internal pullup to ensure that the part stays active if the pin is left floating. This pullup current increases when the part is active to ensure that capacitive feedthrough from fast-moving signals at pin 2 (the amplifier's negative input) does not inadvertently couple to the shutdown circuit and shut the LTC1152 down.

## Applications

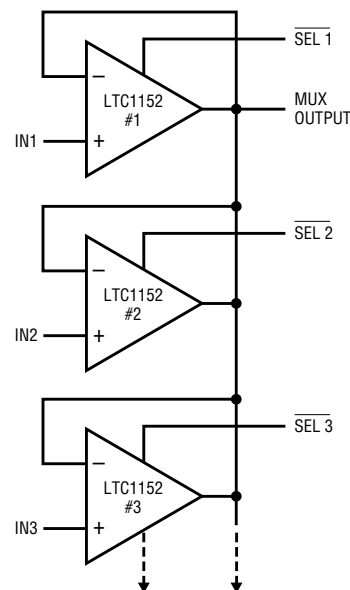
The combination of features provided by the LTC1152 make it well suited for instrumentation applications that require high DC precision and maximum dynamic range. With a single 5V supply, the LTC1152 can process signals from its 2 $\mu$ V<sub>P-P</sub> noise floor all the way to 5V<sub>P-P</sub>; almost 128dB, or better than 21 bits at DC. It is also well suited for use as a precision unity-gain buffer for transducers operated from single supplies; the pA-level bias currents and tiny offset voltage will preserve signals from even very high impedance devices without degradation, and the exceptional CMRR over the entire input range allow it to pick out small signals buried under large common-mode interference. The capacitive load capabilities of the output stage allow it to drive long runs of cable, making it a good buffer amplifier for remotely located sensors. The wide input common-mode range allows the LTC1152 to be used to sense current in either power supply rail, while the excellent DC precision allows extremely small

value sense resistors to be used, minimizing the effect on the rest of the circuit.

The high-impedance output state in shutdown mode can be used to multiplex several signals into one by tying the outputs of multiple LTC1152s together (Figure 4). The active channel is selected by enabling the corresponding LTC1152 with its SD pin and disabling the others. The low output impedance of the active channel prevents the feedback networks of the disabled channels from causing errors in the output. When all channels are disabled, the mux output becomes high-impedance if the individual channels are configured as followers; any other amplifier configurations will load the output with the feedback resistor network. Enabling more than one channel at once will cause unusual behavior as two active channels fight with each other; the LTC1152's current-limited output will protect the amplifiers from damage, but the output may not be what you expected.

## Layout Precautions

As with all zero-drift amplifiers, circuits using the LTC1152 must be laid out with some care, or external parasitics will cause DC errors much greater than those caused by the LTC1152 itself. Thermocouples created where the solder meets the copper PC board traces can create temperature-drift errors as much as ten times greater than the 10nV/°C due to the LTC1152. Leakages from leftover solder flux or cheap PC board material can swamp the input bias current due to the chip and cause bizarre low-level behavior at high impedance inputs (such as the positive input of a unity gain follower). This topic is covered in some detail in the LTC1152 data sheet; some of the things that can cause errors are surprising. Read the discussion to get that last unaccounted microvolt out of your system. Although the LTC1152 will work fine in sloppy layouts, it won't give all the DC precision it is capable of. In particular, don't expect to see microvolt/



**Figure 4. Multiplexing four LTC1152s using the shutdown pin**

picoamp performance if you build your test circuit on a plug-in proto-board. If you're a regular reader of this magazine, you've already read about how using a ground plane will improve your circuit's performance and spice up your life in other ways, so I won't go into detail. Just use one.

## Conclusions

The LTC1152 comes about as close as possible to ideal op amp performance at DC and low frequencies: lots of output current, virtually no input current, low power-supply current (especially when shut down), and no DC error over the entire supply range, input and output. Additionally, it's simple to use; it uses the standard op-amp pinout in both DIP and SO8, and the extra functional pins will mind their own business if you chose to ignore them. If you use the extra features, the LTC1152 can do things most op amps can only dream of: sense signals over the entire power supply range, drive huge capacitive loads, shut down to virtually zero supply current while putting the output in a high-impedance state, and create that most elusive of circuits, the true rail-to-rail follower.



# LT1302 Micropower DC/DC Converter Delivers Unprecedented Power Levels in an Eight-Lead SOIC Package

by Steve Pietkiewicz

## Introduction

The LT1302 micropower, DC-to-DC converter IC achieves new levels of performance. No other IC converter on the market today is capable of efficiently delivering so much power in such a small package. The internal, low-loss NPN switch can handle current in excess of 2A with a drop of just 300mV; the device can deliver 5V at up to 600mA from a 2V input or up to 1A from a 3.3V supply. Designed for battery-powered applications, the LT1302 can operate from a two-,

three-, or four-cell input and can deliver far more output power than other micropower DC/DC converter ICs. Finally, designers of battery-operated systems can get the output power they've been looking for without resorting to complex, space-consuming, and expensive discrete solutions.

The LT1302 maintains high efficiency over a wide range of load current, thanks to automatic Burst Mode™ operation. Quiescent current

is only 200µA and the shutdown pin can be activated to further reduce supply current to just 15µA. Inductor and capacitor size are kept small because the operating frequency is in the 200kHz–400kHz range, allowing all components to be surface-mounted. Designers of PDAs, digital cellular phones, portable transmitting devices, or other systems needing high efficiency over a broad output power range will appreciate the performance of the LT1302.

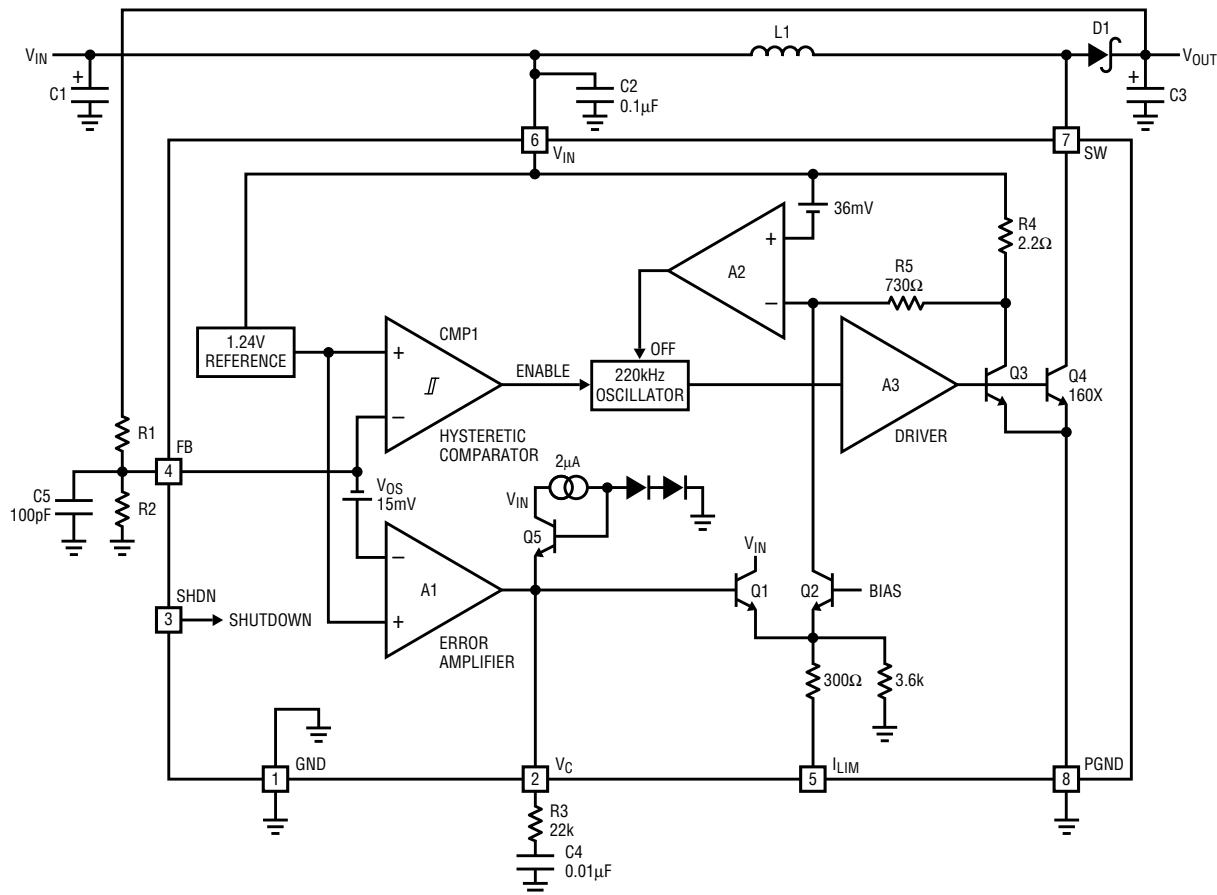


Figure 1. Block Diagram: LT1302

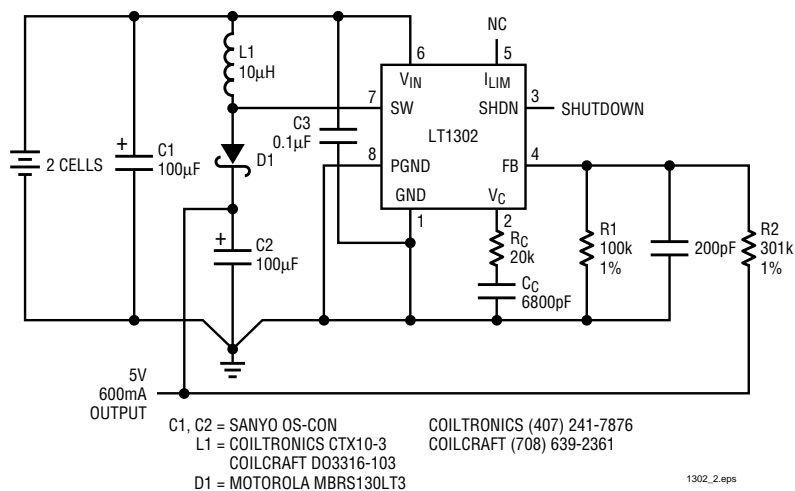


Figure 2. Two or three cell to 5V converter delivers 600mA, 1A from 3.3V supply

### Operation

The LT1302's operation can best be understood by examining the block diagram in Figure 1. The LT1302 operates in one of two modes, depending on load. With light loads, comparator CMP1 controls the output; with heavy loads, control is passed to error amplifier A1. Burst Mode™ operation consists of monitoring the FB pin voltage with hysteretic comparator CMP1. When the FB voltage, related to the output voltage by external attenuator R1 and R2, falls below the 1.24V reference voltage, the oscillator is enabled. Switch Q4 alternately turns on, causing current buildup in inductor L1, then turns off, allowing the built-up current to flow into output capacitor C3 via D1. As the output voltage increases, so does the FB voltage; when it exceeds the reference plus CMP1's hysteresis (about 5mV) CMP1 turns the oscillator off. In this mode, peak switch current is limited to approximately 800mA by A2, Q2, and Q3. Q2's current, set at 34µA, flows through R5, causing A2's negative input to be 25mV lower than VIN. This node must fall more than 36mV below VIN for A2 to trip and turn off the oscillator. The remaining 11mV is generated by Q3's current flowing through R4. Emitter-area scaling sets Q3's collector current to 0.625% of switch Q4's current. When Q4's current is 800mA, Q3's current is 5mA, creating an 11mV drop across R4

which, added to R5's 25mV drop, is enough to trip A2.

When the output load is increased to the point where the 800mA peak current cannot support the output voltage, CMP1 stays on and the peak switch current is regulated by the voltage on the VC pin (A1's output). VC drives the base of Q1. As the VC voltage rises, Q2 conducts less current, resulting in less drop across R5. Q4's peak current must then increase in order for A2 to trip. This current-mode control results in good stability and immunity to input voltage variations. Because this is a linear, closed-loop system, frequency compensation is required. A series RC from VC to ground provides the necessary pole-zero combination.

### Applications

#### Two or Three Cell to 5V Converter

Figure 2 shows a two or three cell to 5V DC/DC converter that can deliver up to 600mA from a two-cell input (2V minimum), or up to 900mA from a three-cell input (2.7V minimum). R1 and R2 set the output voltage at 5V. The 200pF capacitor from FB to ground aids stability; without it the FB pin can act as an antenna and pick up dV/dt from the switch node, causing some instability in switch current levels at heavy loads. L1's inductance value is not critical; a minimum of 10µH is suggested in

two-cell applications (although this guideline is ignored in the two-cell-to-12V circuit shown later). Lower values typically have less DC resistance and can handle higher current. Transient response is better with low inductance, but more output current can be had with higher values. Peak current in Burst Mode™ operation increases as inductance decreases, due to the finite response time of the current sensing comparator in the LT1302. The Coilcraft DO3316 series inductors have been found to be excellent in terms of performance, size, and cost, but their open construction results in some magnetic flux spray; try Coiltronics' OCTAPAC series if EMI is a problem. Transient response with a load step of 25mA to 525mA is detailed in Figure 3. There is no overshoot upon load removal because switching stops entirely when output voltage rises above the comparator threshold. Undershoot at load step is less than 5%. The circuit's efficiency at various input voltages is shown in Figure 4.

Although efficiency graphs present useful information, a more "real-world" measure of converter performance comes from battery-lifetime

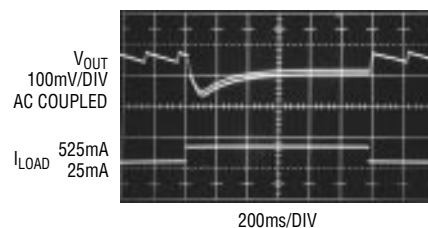


Figure 3. Transient response of DC/DC converter with 2.5V input. Load step is 25mA to 525mA.

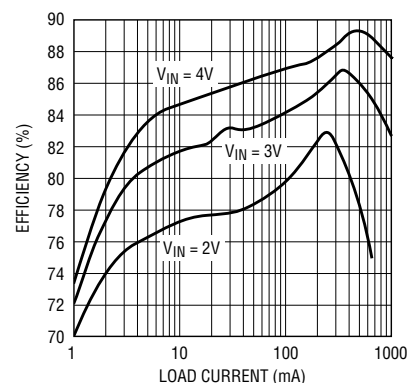
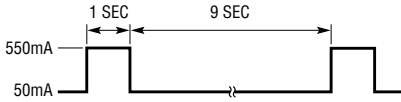


Figure 4. Efficiency of Figure 2's circuit.

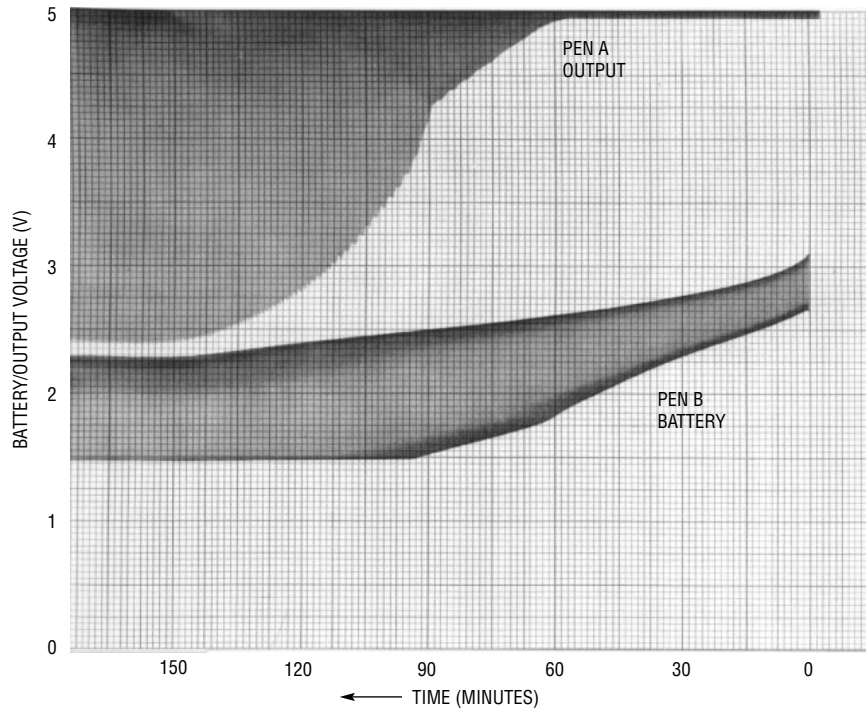


**Figure 5. Load profile for battery-life curves in Figures 6 and 7.**

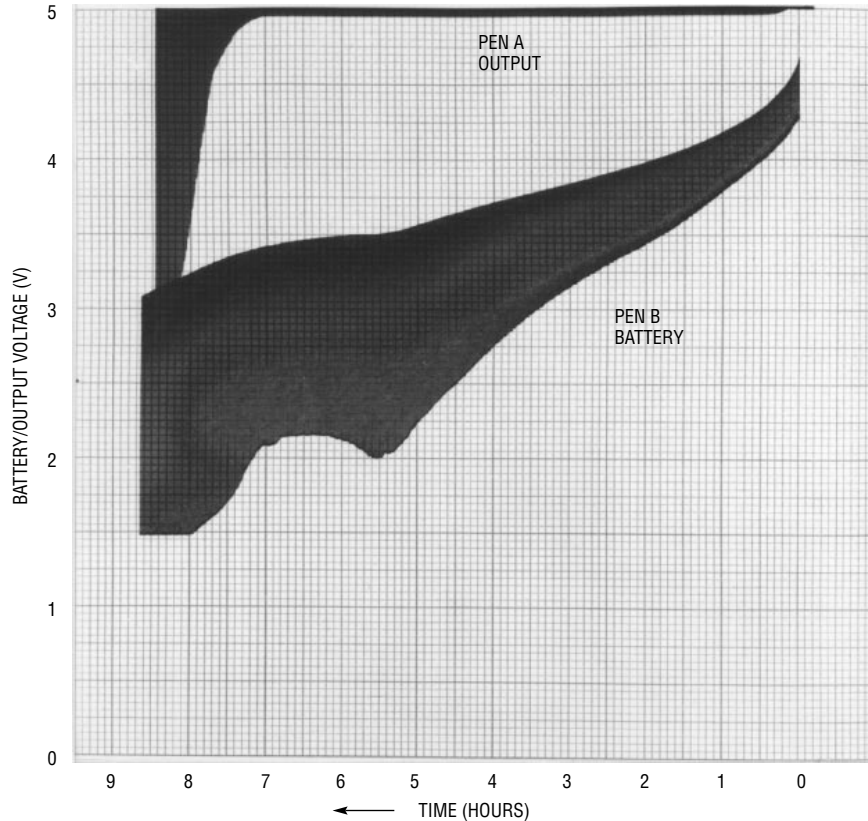
chart recordings. Many systems require high power for a short time, for example, to spin up a hard disk or transmit a packet of data. Figures 6, 7, and 8 present battery-life data with a load profile of 50mA for 9 seconds and 550mA for 1 second, as detailed in Figure 5. At the chart speeds used, individual ten-second events are not discernable and the battery voltage appears as a very thick line. Figure 6 shows operating life using a two-cell alkaline (Eveready E91) battery. Battery voltage (pen B) drops 400mV as the output load changes from 50mA to 550mA. Battery impedance (330mΩ when fresh) can be derived from this data. After 63 minutes, the battery voltage drops substantially below 2V when the output load is 550mA, causing the output voltage (pen A) to drop. The output returns to 5V when the load drops to 50mA. The LT1302's undervoltage lockout prevents the battery voltage from falling below 1.5V until the battery is completely discharged (not shown on the chart).

A three-cell alkaline battery has a significantly longer life, as shown in Figure 7. Note that the time scale here is one hour per inch. Usable life is about 7.3 hours, a sevenfold improvement over the two-cell battery. Again, battery impedance causes the battery voltage (pen B) to drop as the load changes from 50mA to 550mA. The increasing change between the loaded and unloaded battery voltage over time is due to both increased current demand on the battery as its voltage decreases and increasing battery impedance as it is discharged.

Replacing the two-cell alkaline with a two-cell NiCad (AA Gates Millennium) battery results in a surprise, shown in Figure 8. Although these AA NiCad cells have one-fourth the



**Figure 6. Two-cell alkaline battery to 5V converter with load profile of Figure 5 gives 63 minutes operating life. Battery life decreases when 550mA load is applied; impedance is 330 milliohms when fresh. Output voltage drops at 550mA load after 63 minutes, but converter can still deliver 50mA.**



**Figure 7. Three-cell alkaline battery to 5V converter with pulsed load has 7.3 hours operating life.**



energy of AA alkaline cells, operating life is 2.8 times greater with the 50mA/550mA load profile. Dramatically lower battery impedance (80mΩ for the NiCad, versus 330mΩ for the alkaline) is the cause. Battery voltage (pen B) drops just 100mV as the output load changes from 50mA to 550mA, compared to 400mV for alkalines. Additionally, impedance stays relatively constant over the life of the battery. This comparison clearly illustrates the limitations of alkaline cells in high-power applications.

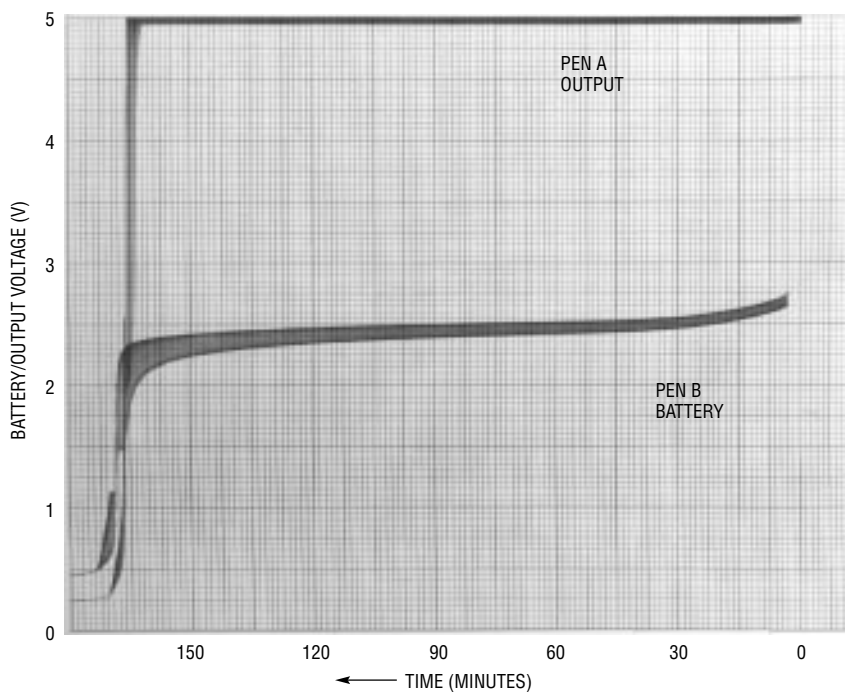
### Two-Cell-to-12V Converter

Portable systems with PCMCIA interfaces often require 12V at currents of up to 120mA. Figure 9's circuit can generate 12V at over 120mA from a two-cell battery. Operating the converter in continuous mode requires a higher duty cycle than the LT1302 provides, so a very low inductance (3.3μH) must be used in order to provide enough output current in discontinuous mode. Efficiency for this circuit is in the 70–80% range, as Figure 10's graph shows. Battery life at this power level would be short with a continuous load, but the most common application for this voltage/current level, flash memory programming, has a rather low duty factor. Maximum output current versus input voltage is shown in Figure 11. To operate this circuit from a three-cell battery, change L1's value to 6.8μH. This will result in lower peak currents, improving efficiency substantially.

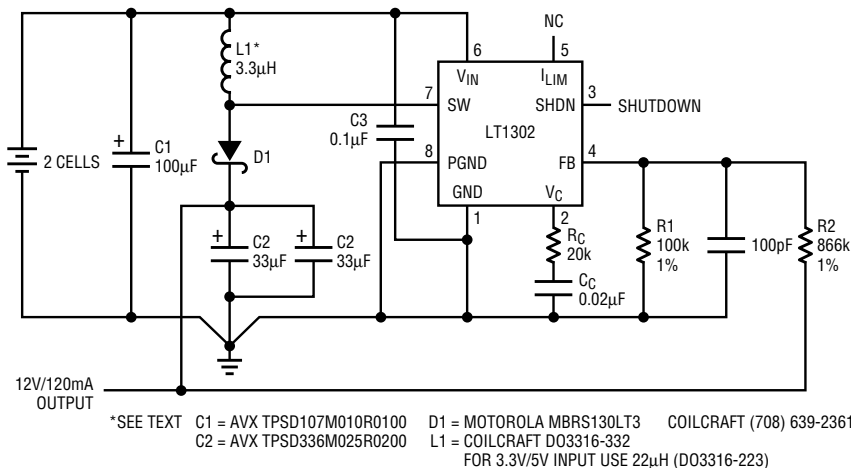
By changing L1's value to 22μH, the circuit will operate from a 3.3V or 5V supply. Up to 350mA can be generated from 3.3V; 600mA can be delivered from a 5V input. Efficiency, pictured in Figure 12, exceeds 80% over much of the load range, and peaks at 89% with a 5V input.

### Single-Cell to 5V/150mA Converter

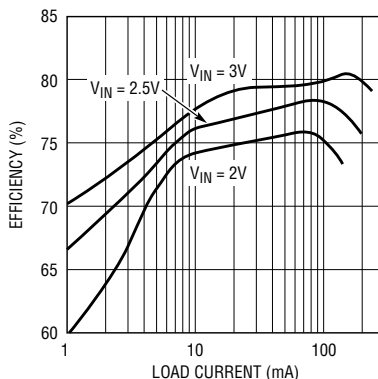
Stand-alone, single-cell converters can typically provide no more than 40mA–50mA at 5V from a single cell.



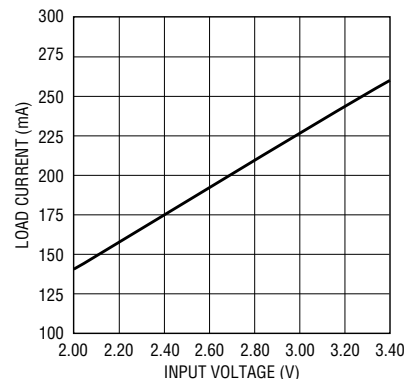
**Figure 8.** Two-cell NiCad battery to 5V converter shows dramatically lower ESR of NiCads compared to alkalines. Battery impedance is 80 milliohms. Although the 600 milliampere hour NiCad has one fourth the energy of 2.4 Amp/hour alkalines, with 50mA/550mA loads NiCads outlast alkalines by a factor of 2.8. Low cell impedance is maintained until the battery is completely discharged.



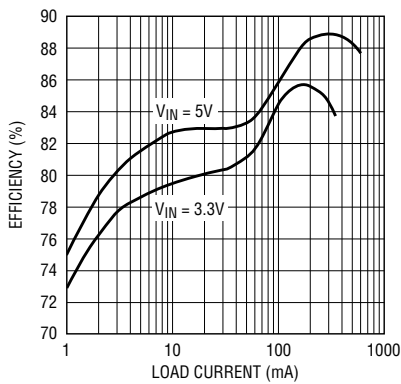
**Figure 9.** Two-cell to 12V DC/DC converter delivers 120mA. Changing L1's value allows operation from 3.3V/5V supply.



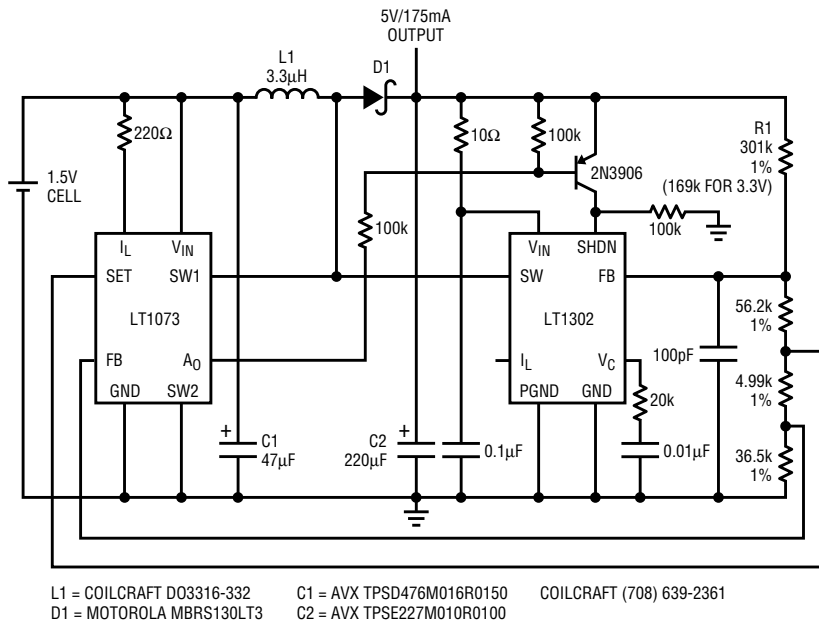
**Figure 10.** Two-cell to 12V converter efficiency



**Figure 11.** Maximum load current of two-cell to 12V converter versus input



**Figure 12. 3.3V/5V to 12V converter efficiency**



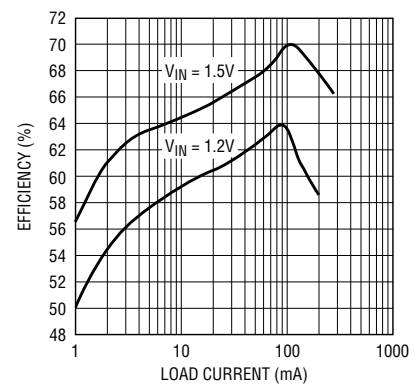
**Figure 13. Single-cell to 5V converter delivers 150mA. Changing R1 to 169K provides 3.3V at 250mA.**

When more power is required, the LT1302 can be used in conjunction with a single-cell device.<sup>1</sup> Figure 13's circuit operates from a single cell and delivers 5V at 150mA output. Although the LT1302 requires a minimum  $V_{IN}$  of 2V, single-cell operation can be achieved by powering the LT1302 from the 5V output. At start-up,  $V_{OUT}$  is equal to the cell voltage minus a diode drop. The LT1073 initially puts the LT1302 in its shutdown state. The LT1073 switches L1, causing L1's current to alternately buildup and dump into C2. When  $V_{OUT}$  reaches approximately 2V, the LT1073's SET pin goes above 212mV, causing AO to go low. This pulls the LT1302's SHDN pin low, enabling it. The output, now bootstrapped by the much higher power LT1302, quickly reaches 2.4V. When the LT1073's FB pin reaches 212mV, its switching action stops. The brief period when the LT1073 and LT1302 are switching simultaneously has no detrimental effect. When the output reaches 5V, the LT1073 has ceased switching. Circuit efficiency is in the 60–70% range, as shown in Figure 14.

### Three-Cell to 3.3V/12V Buck-Boost Converter

Obtaining 3.3V from three cells is not a straightforward task; a fresh battery measures over 4.5V and a fully depleted one 2.7V. Since battery voltage can be both above and below the output, common step-up (boost) or step-down (buck) converters are inadequate. Figure 15's circuit provides an efficient solution to the problem using just one magnetic component, and also provides an auxiliary 12V output. When the LT1302's switch is on, its SW pin goes low, causing current buildup in T1D and T1E (windings are paralleled to achieve lower DC resistance). D1's anode goes to  $-V_{IN}$  because of the phasing of T1C/T1A relative to T1D/T1E. C1 is charged to  $V_{IN}$ . When the switch opens, SW flies high to a voltage of  $V_{IN} + V_{OUT} + V_{DIODE}$ . Energy is transferred to the output by magnetic coupling from T1D/T1E to T1C/T1A, and by current flowing through C1. During this flyback phase, T1A/T1C has 3.3V plus a diode drop across the windings. T1B, which has a 3:1 turns ratio, has approximately

10V–11V impressed upon it. T1B "stands" on the 3.3V output, resulting in about 13–14V at the input of the LT1121 linear regulator, which then precisely regulates the 12V output. Since this output is not directly regulated by the LT1302, it cannot be loaded without having at least a small load on the directly regulated 3.3V output. The LT1121 can be turned off by pulling its SHDN pin low, isolating the load from the output. Figure 16 shows the circuit's efficiency for various input voltages.



**Figure 14. Single-cell to 5V converter efficiency**

**Construction Hints**

The high-speed, high-current switching associated with the LT1302 mandates careful attention to layout. Follow the suggested component placement in Figure 17 for proper operation. High-current functions are separated by the package from sensitive control functions. Feedback resistors R1 and R2 should be close to the feedback pin (pin 4). Noise can easily be coupled into this pin if care is not taken. If the LT1302 is operated

off a three cell or higher input, R3 (2Ω) in series with V<sub>IN</sub> is recommended. This isolates the device from noise spikes on the input voltage. Do not put in R3 if the device must operate from a 2V input, as input current will cause the LT1302's input voltage to go below 2V. The 0.1μF ceramic bypass capacitor C3 (use X7R, not Z5U) should be mounted as close as possible to the package. Grounding should be segregated as illustrated.

C3's ground trace should not carry switch current. Run a separate ground trace up under the package as shown. The battery and load return should go to the power side of the ground copper. Adherence to these rules will result in working converters with optimum performance. ⚡

<sup>1</sup> Williams, Jim. "200ma Output, 1.5 to 5V Converter." *Linear Technology III:1* (February, 1993) p. 17

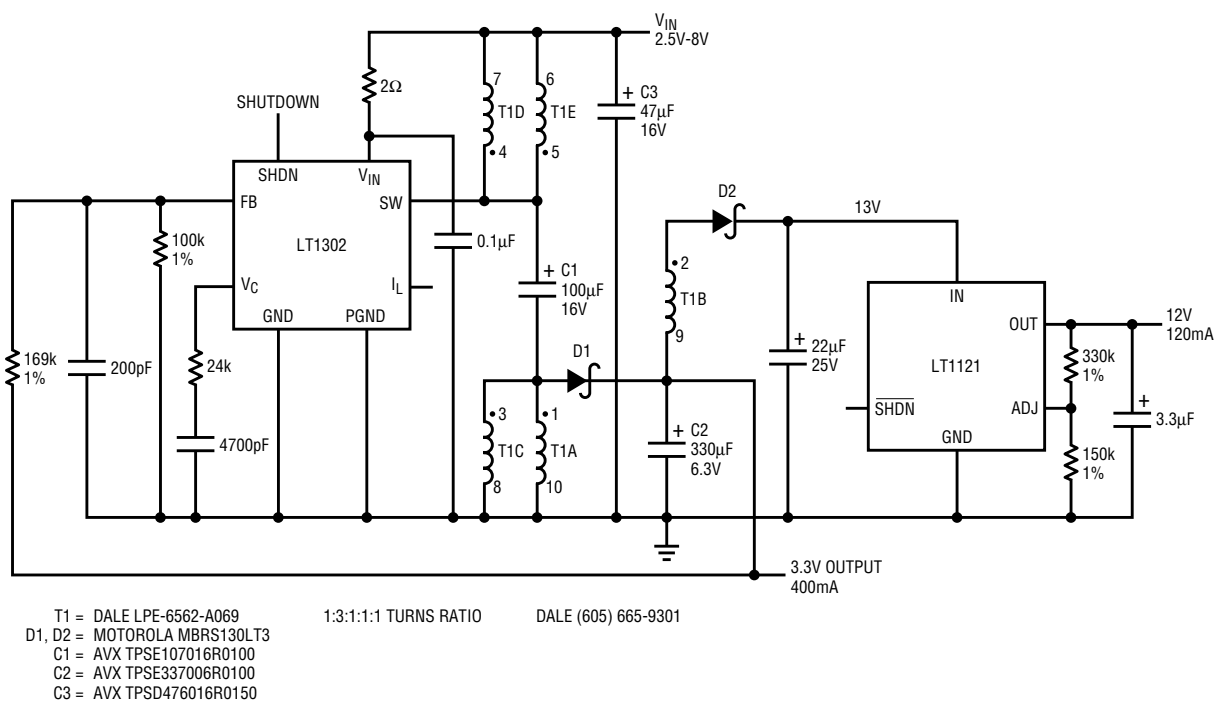


Figure 15. Three-cell to 3.3V Buck-Boost converter with auxiliary 12V regulated output

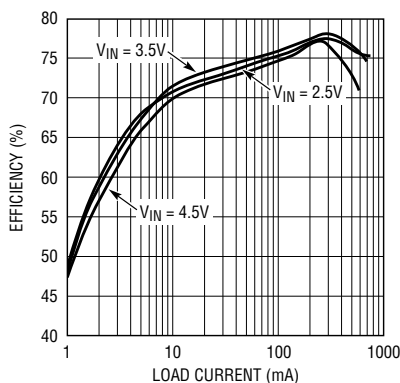


Figure 16. 3.3V buck-boost converter efficiency

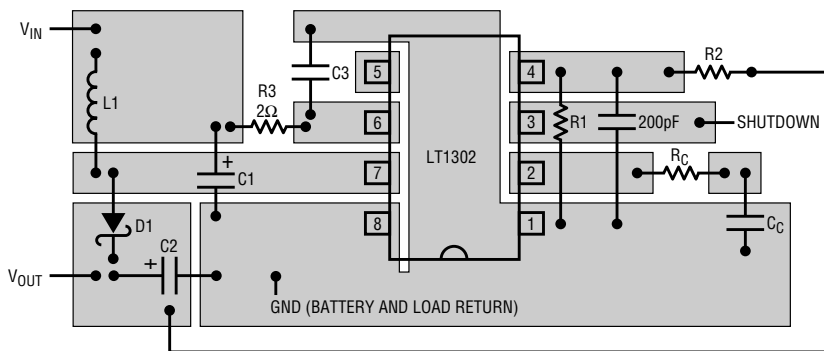


Figure 17. Suggested component placement for LT1302

# The LTC1066-1: Fourteen Bit DC Accurate Elliptic Lowpass Filter

by Nello Sevastopoulos

## Introduction

The LTC1066-1 is the first monolithic filter that combines RC active techniques with switched-capacitor technology to achieve outstanding DC and AC performance. The LTC1066-1 is an eighth-order filter with clock-tunable cutoff frequency to 100kHz, 1.5mV maximum DC offset over the commercial temperature range, and 14 bits of DC gain linearity.

## The Technology: How Is It Done?

Figure 1 shows the block diagram of the LTC1066-1 (a patent is pending on the device). A high-speed, precision op amp at the device's input, A1, performs the following DC tasks: A1 stores the offset voltage of the eighth-order switched capacitor filter, the offset voltage of the output unity gain buffer, A2, and the DC voltage drop across the feedback resistor,  $R_F$ , across its feedback capacitor,  $C_F$ . The DC output offset of the overall filter is then equal to the DC offset of the input precision op amp, minus the voltage drop across  $R_F$ . The small voltage drop across  $R_F$  is caused by the op amp's input bias current.

A1 also performs the following AC tasks. For frequencies above the cutoff frequency ( $1/2\pi R_F C_F$  of the external RC), the input op amp serves two purposes. First, it buffers the input signal,  $V_{IN}$ , and, second, it isolates or disconnects the IC's output,  $V_{OUT}$ , from the input terminal of the internal switched-capacitor network as the input frequency is increased. With increasing frequency, the AC gain of the DC correcting loop tends towards zero and the frequency response characteristics of the LTC1066-1 are dictated only by the transfer function of the internal switched-capacitor network. The transition between DC and AC is very critical and extreme care is taken to make it as transparent to the user as possible.

It is important to appreciate the stringent requirements on the input op amp: A1 must not only have excellent DC characteristics but must also be able to handle high-frequency common-mode signals without introducing distortion.

The output high-frequency op amp, A2, buffers the switched-capacitor network and maintains low output impedance over a wide range of frequencies. The low output impedance

preserves the stop-band characteristics of the internal switched-capacitor network. The DC offset of the output op amp is corrected by the loop, as discussed above. Op amp A2 should have at least the same AC performance as the input op amp.

## LTC1066-1 DC Performance: Let the Precision Op Amps Show How Good They Really Are

The LTC1066-1 features both DC accuracy and low output  $V_{OS}$ . Filter users often confuse DC accuracy with low DC offset. DC accuracy preserves the DC information of the input signal. DC accuracy does not imply low DC offset. If the filter output offset does not change with the DC value of the input voltage, the filter is DC accurate.

Large output DC offsets limit the filter's dynamic range. This is especially true with low power-supply voltages. Adjusting the output offset over a wide temperature range can be cumbersome and costly. The DC performance of the LTC1066-1 is primarily dictated by the DC characteristics of the input precision op

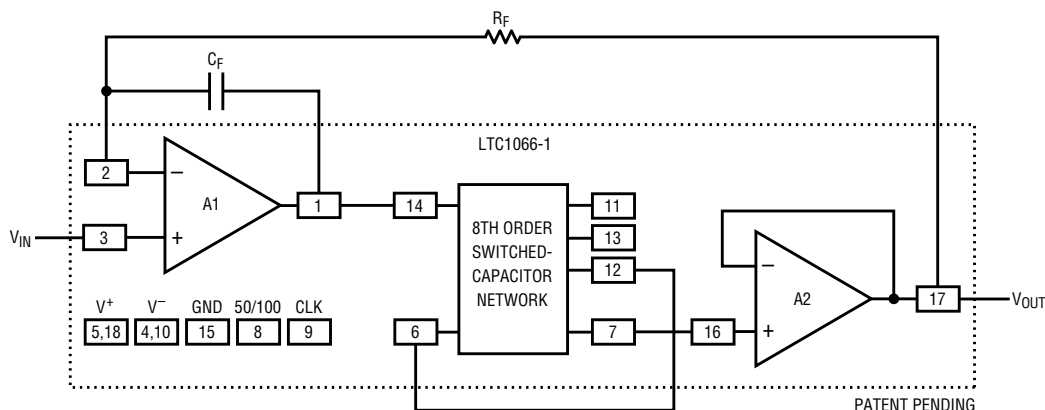
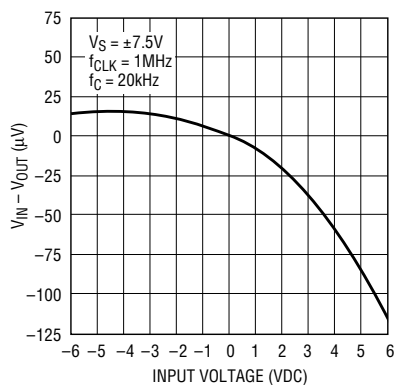


Figure 1. LTC1066 functional diagram



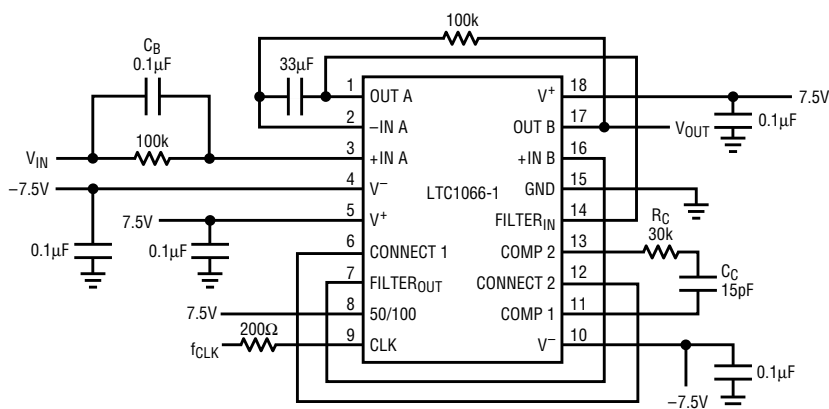
**Figure 2. DC gain linearity of LTC1066-1 is better than 14 bits**

amp. DC input voltages in the vicinity of half the total power supply of the filter are processed with exactly 0dB (1V/V) of gain. Using a  $\pm 7.5V$  supply, a  $\pm 5VDC$  input signal can be processed with better than 14-bit DC gain linearity. Figure 2 displays the  $(V_{IN(DC)} - V_{OUT(DC)})$  error for an input range of +6 to -6 volts. The LTC1066-1 cutoff frequency was set at 20kHz.

The LTC1066-1 output DC offset,  $V_{OS(OUT)}$ , is measured with the input grounded and with dual, symmetrical power supplies.  $V_{OS(OUT)}$  is typically 100 microvolts and is optimized for the combination  $R_F = 20k\Omega$ ,  $C_F = 0.1\mu F$  (see block diagram). The  $V_{OS}$  temperature drift is  $7\mu V/^\circ C$ .

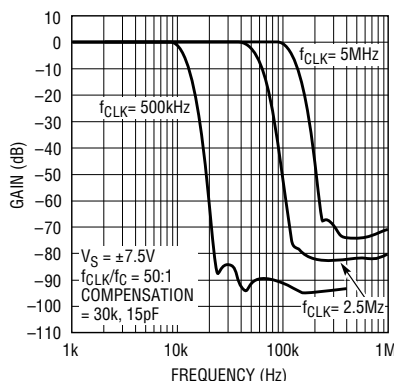
$V_{OS}$  can be calculated from the formula:

$$V_{OS(OUT)} = V_{OS(OA1)} - R_F I_B$$



MAXIMUM OUTPUT VOLTAGE OFFSET = 4mV, DC LINEARITY =  $\pm 0.0063\%$ ,  $T_A = 25^\circ C$ .  
 THE PIN 6 TO 12 CONNECTION SHOULD BE UNDER THE IC AND SHIELDED BY AN ANALOG SYSTEM GROUND PLANE.  
 RC COMPENSATION BETWEEN PINS 11 AND 13 REQUIRED ONLY FOR  $f_{CUTOFF} > 50kHz$ .  
 THE 33µF CAPACITOR IS A NONPOLARIZED, ALUMINUM ELECTROLYTIC,  $\pm 20\%$ , 16V (NICHICON UUPIC 330MCRIGS OR NIC NACEN 33M16V 6.3 x 5.5 OR EQUIVALENT).

**Figure 4. DC-accurate, 10Hz to 100kHz, eighth-order elliptic lowpass filter,  $f_{CLK}/f_C = 50:1$**



**Figure 3. LTC1066-1 amplitude versus frequency**

Where  $I_B$  is the input bias current of A1 (approximately 60nA).

The filter DC output offset voltage is, for all practical purposes, independent of the clock frequency.

### LTC1066-1 AC Performance

#### Clock Tunability

An external clock tunes the cutoff frequency of the internal switched-capacitor network. The device has been optimized for a clock-to-cutoff-frequency ratio of 50:1. The internal double sampling greatly reduces the risk of aliasing.

The maximum obtainable cutoff frequency,  $f_{CUTOFF(MAX)}$ , depends on power supply, clock duty cycle, and temperature;  $f_{CUTOFF(MAX)}$  does not depend on the value of the external

resistor/capacitor combination  $R_F C_F$ . For the commercial temperature range (0 to  $+70^\circ C$ ) and  $\pm 7.5V$  supplies, the maximum obtainable cutoff frequency is 100kHz (see Figure 3). The  $R_C C_C$  compensation, as shown in Figure 4, is needed only for cutoff frequencies above 60kHz. The data detailed in Figure 3 reveals the important fact that for a cutoff frequency of 100kHz, the stopband attenuation still remains greater than 70dB for input frequencies up to 1MHz.

The minimum obtainable cutoff frequency depends on the  $R_F C_F$  time constant of the servo loop. For a given  $R_F C_F$  time constant, the minimum obtainable cutoff frequency of the LTC1066-1 is:

$$f_{CUTOFF(MIN)} = 250 / (2\pi R_F C_F)$$

$$f_{CUTOFF(MIN)} = 100kHz$$

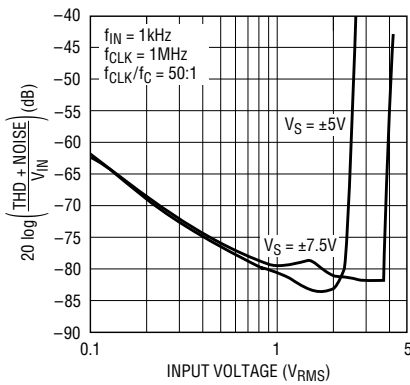
For instance, if  $R_F = 20k\Omega$ ,  $C_F = 1\mu F$ ,  $f_{CUTOFF(MIN)} = 2kHz$ , and  $f_{CLOCK(MIN)} = 100kHz$ .

Under these conditions, a clock frequency below 100kHz will "warp" the passband gain by more than 0.1dB. Please see the LTC1066-1 data sheet for more details.

Figure 4 shows an application allowing clock tunability from 10Hz to 100kHz. The  $R_C C_C$  frequency compensating components maintain a flat passband for cutoff frequencies between 50kHz and 100kHz. The input resistor reduces the output DC offset caused by the op amp bias current through the 100k $\Omega$  feedback resistor. The measured DC offset and the gain nonlinearity are 4mV and  $\pm 0.0063\%$  (84dB), respectively. The 0.1 $\mu F$  bypass capacitor,  $C_B$ , helps keep the total harmonic distortion of the filter from being degraded by the 100k $\Omega$  input resistor. The frequency compensation components (30k $\Omega$ , 15pF) maintain a flat passband for cutoff frequencies all the way up to 100kHz.

#### Dynamic Range

The LTC1066-1 wideband noise is 100 microvolts RMS. Figure 5 shows the noise plus distortion versus RMS input voltage at 1kHz. With a  $\pm 5V$



**Figure 5. LTC1066-1 dynamic range**

supply, the filter can swing  $\pm 2.5V$  (5V full scale) with better than 0.01% distortion plus noise. The maximum signal-to-noise ratio, in excess of 90dB, is achieved with  $\pm 7.5V$  supplies. Unlike previous monolithic filters, the data shown in Figure 5 is taken without using any input or output op amp buffers. The output buffer of the LTC1066-1 can drive a 200 $\Omega$  load without dynamic-range degradation.

**Aliasing and Anti-Aliasing**

All sampled-data systems will alias if their input signals exceed half the sampling rate, but aliasing for high-order, band-limited, switched-capacitor filters need not be a serious problem. The LTC1066-1, when operating with a 50:1 clock-to-cutoff-frequency ratio, will have significant aliasing only for input signals centered around twice the clock frequency and its even multiples. Figure 6 shows the input frequencies that will generate aliasing at the filter output. For instance, if the filter is tuned to a 50kHz cutoff frequency using a 2.5MHz clock, significant aliasing will occur only for input frequencies of 5MHz  $\pm$  50kHz. The filter user should be aware of the spectrum at the input to the filter. Next, an assessment should be made as to whether a simple, continuous-time anti-aliasing

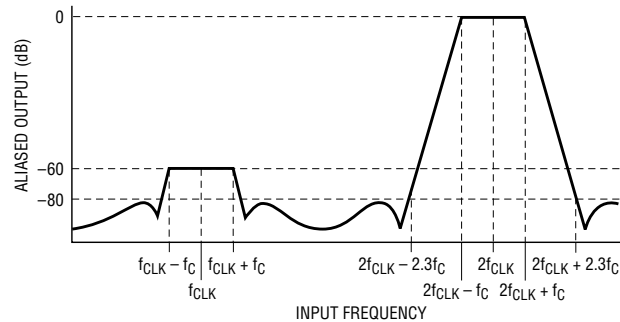
filter in front of the LTC1066-1 is required. The anti-aliasing filter should do precisely what it is meant to do, that is, provide band-limiting. The anti-aliasing filter should not degrade the DC or AC performance of the LTC1066-1.

For fixed-cutoff-frequency filter applications, the anti-aliasing function is quite trivial. Figure 7 shows the precision input op amp used to perform both the DC-accurate function of the LTC1066-1 and as a second-order, Butterworth anti-aliasing lowpass filter. The cutoff frequency of the RC anti-aliasing filter is set three times higher than the cutoff frequency of the LTC1066-1. For the example

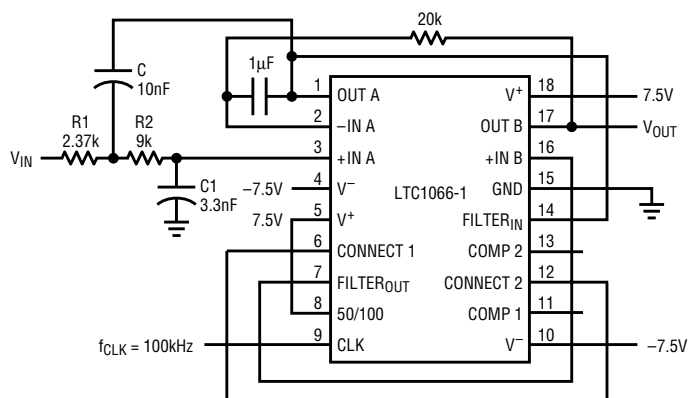
shown in Figure 7, the input anti-aliasing filter provides a 62dB attenuation at twice the clock frequency of the switched-capacitor filter.

**Conclusions**

The LTC1066-1 crams filter performance usually found in multiple-package, RC active designs into a single 18-pin SOIC. The filter is an eighth-order elliptic low pass filter with  $f_{CUTOFF}$  useful to 100kHz. It boasts true 14-bit gain linearity along with DC accuracy. The LTC1066-1 will replace larger, more expensive, and less accurate solutions in instrumentation, data acquisition, and other types of circuitry. **LT**



**Figure 6. Aliasing versus frequency  $f_{CLK}/f_C = 50:1$  (pin 8 to V+) Clock is a 50% duty-cycle square wave.**



**Figure 7. Adding a 2-pole Butterworth input anti-aliasing filter. Set  $C1 = 0.33C$ ,  $R2 = 3.8 \times R1$ ;  $f-3dB$  (input anti-aliasing) =  $0.8993/(2\pi R1C)$**

# Micropower, 12-Bit, SO-8 ADCs Now Available for Three-Volt Systems

by William C. Rempfer and Marco Pan

## Portable/Battery Systems: a Hot Item

One of the fastest growing high-tech markets today is that for portable and battery-operated systems. Manufacturers are scrambling to introduce new products into this growing area. Cellular phones, portable computers, personal digital assistants (PDAs), and portable industrial equipment are expected to be among the hottest new products. Even the financial community is aware of this trend, looking to invest in companies that make products for these markets. The companies that can capitalize on this demand can reap huge benefits.

Component suppliers are also working very hard to develop products that provide value to the system manufacturers. "Providing value" simply means helping system designers and their companies succeed against their competition and succeed in providing the best solutions to their customers. What these system designers need most from components is small size, low power, low cost, and 3V operation.

## A/D Converters in Portable Systems

Portable systems use A/D converters. Many use ADCs directly, as when digitizing the pen-screen input in a pen-based computer or digitizing the sensor output in a portable gas meter or detector. Others use ADCs more

indirectly, for example to monitor the system's health by monitoring voltages and temperatures inside an instrument. Regardless of the use of the ADC, system designers have had a real struggle getting small ADCs at low enough power levels and at low cost. To get it all to work on 3V is a further difficulty.

LinearTechnology recently brought some relief by introducing the world's first 12-bit, micropower ADCs in SO-8 packages: the LTC1286 and LTC1298. These two converters provide the micropower, small-size, low-cost conversion eagerly awaited by designers of 5V systems. But there is also a growing need for 3V ADCs. Now LTC is meeting that need by releasing a similar pair of ADCs designed for 3V systems: the LTC1285 and LTC1288. This article will discuss some of the system designer's challenges and how these converters meet them.

## Meeting the ADC Needs of Portables

### Small Size: SO-8

Until recently, it has been impossible to get *any* 12-bit ADC in an SO-8 sized package. Designers had to suffer with converters in the larger sized packages such as 16- and 20-pin SOICs and 8-pin DIPs. The 8-pin de-

vices (such as the LTC1292/LTC1297) were the most popular because they were the smallest available at the time. Their serial I/O made them more efficient than parallel devices. However, they were still much larger than the desired SO-8 and were not surface mountable.

The first SO-8, micropower, 12-bit ADCs on the market were the LTC1286/LTC1298, introduced by LTC for 5V systems. The two new products, the LTC1285 and LTC1288, bring the same features to 3V systems, again for the first time. As Figure 1 shows, the LTC1285 is a single-channel, 12-bit device. The LTC1288 has 2 channels. These devices are already seeing great success in portable applications because of their size. The total board space of the 12-bit ADC (with a single 0.1 $\mu$ F bypass chip capacitor) is just 0.08 square inches. The height is also very small, at 1.7mm.

## Micropower Operation: 160 $\mu$ A and Auto-Shutdown

Micropower, 12-bit A/D conversion is another area that has not been addressed until recently. In the past, designers had to cycle the ADCs on and off to try to achieve micropower performance. Figure 2 shows an example of this method. A higher power ADC is switched on and off to reduce power drain during inactive periods. This approach has three drawbacks. First, it requires switching hardware and a switching signal. Second, while the ADC power supply is switched off, any digital or analog inputs to the ADC need to be disabled or switched off to prevent forward biasing the substrate diodes of the ADC (the ADC

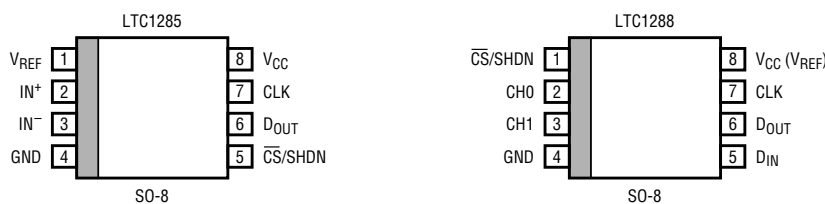
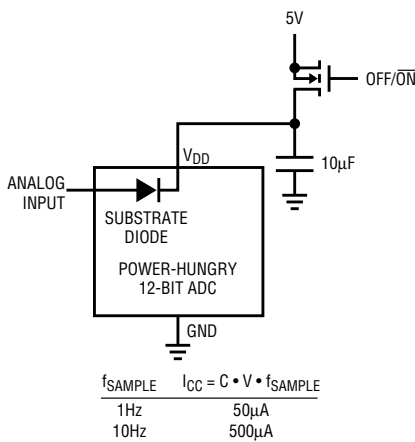


Figure 1. These micropower, 3V, 12-bit ADCs come in SO-8 packages and offer one or two input channels



**Figure 2. High-side switching a power-hungry ADC takes hardware and wastes power. Repeatedly switching the required bypass capacitor consumes 500 microamps even when taking readings at only 10Hz.**

will load the signals tied to its inputs and this can cause latchup problems or excessive current drain). Third, most power-hungry ADCs require a large bypass capacitor to be connected at the supply pin. The charging and discharging of this capacitor in each power-down cycle draws a very large current, ruining the dissipation of the circuit even at very low sample rates. For example, a 10µF bypass capacitor switched at a 10Hz rate consumes 500µA.

The first solution to the problem was provided with auto-shutdown on low-to-medium power converters (e.g., the LTC1297). Here, the ADC is tied directly to the supply and shuts itself off whenever it is not converting. The switching hardware is eliminated, the need to disable input signals is removed, and the current drain in the bypass capacitor is eliminated because the cap is never switched. The power drain is simply the "ON" current of the converter multiplied by the duty cycle (the percentage of time it is on). The auto-shutdown is invisible to the user, which is nice.

In many new systems, operating power becomes more important because the ADC is in use much of the time. The LTC1286/LTC1298 were the first devices on the market to offer the ideal combination: micropower

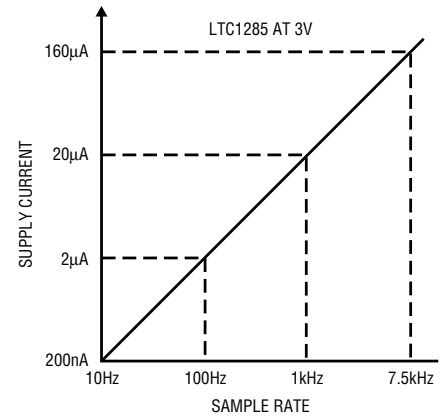
operating currents and auto-shutdown. The new LTC1285/LTC1288 bring this same performance to 3V. Drawing only 160µA at full speed, they provide the lowest power alternative available. Figure 3 shows the current drain from a 3V supply versus sample rate. The current drops linearly with sample rate from 160µA at 7.5kHz to 2µA at 100Hz. At 10Hz it draws only 200nA, compared to the 500µA of the switched ADC example of Figure 1.

### 3V Operation Guaranteed

Before 3V ADCs became available, designers had to have a 5V supply on the board. There was a strong incentive to eliminate the 5V supply entirely. Now that is possible. LTC has had a family of 3V ADC products for some time (as shown in Table 1), but this is the first time anyone has offered the combination of 12-bits, micro-power, SO-8 and 3V all in one device. The LTC1285/1288 are designed, specified, and tested for 3V.

### Low Cost by Design

Cost is a big concern for designers of portable and battery-operated systems. These systems are typically high-volume, low-cost products, so the components they use must be suitably priced. The LTC1285/LTC1288 are designed to meet this need. They use an architecture that guarantees excellent differential linearity, low drift, and no missing codes, without expensive calibration, trimming, or processing. They are priced



**Figure 3. The auto shutdown feature of the LTC1285/1288 makes the ADC's supply current drop as its sample rate is reduced. The 160 microamp drain at 7.5kHz drops to 2 microamps at 100Hz.**

to make them attractive to high volume users.

### Conclusion

Designers have had trouble finding 12-bit ADCs which meet any one of the four criteria we've talked about: small size, micropower operation, 3V operation, and low cost. But to find an ADC which meets all four constraints at once has been impossible until now. There are huge opportunities for system designers in the portable and battery operated marketplace, if they can meet the challenges required of them. The LTC1285 and LTC1288 can help any designer facing this challenging situation succeed. **LT**

**Table 1. The 3V ADC family now includes the LTC1285/LTC1288, which combine 12-bits, micropower operation, 3V, and SO-8 in one device.**

3V Device	Bits	Speed	Channels	Package	Supply Current	Auto Shutdown?
LTC1096/8	8	20KSPS	1/2	SO-8	100µA	Yes
LTC1196	8	383KSPS	1	SO-8	2mA	No
LTC1198	8	287KSPS	2	SO-8	2mA	Yes
LTC1283	10	15KSPS	8	SO-20	150µA	No
LTC1287	12	30KSPS	1	8-DIP	1mA	No
LTC1289	12	25KSPS	8	SO-20	1mA	Software
LTC1282	12	140KSPS	1	SO-24	4mA	No
<b>LTC1285/8</b>	<b>12</b>	<b>7.5KSPS</b>	<b>1/2</b>	<b>SO-8</b>	<b>120µA</b>	<b>Yes</b>



# A Low-Power, Low-Voltage CCFL Power Supply

by Steve Pietkiewicz

Most recently published CCFL driver circuits require an input supply of 7V to 20V and are optimized for bulb currents of 5mA or more. This precludes lower power operation from two- or three-cell batteries often used in PDAs, palmtop computers, and the like. A CCFL power supply that operates from 2 to 6V is shown in Figure 1. This circuit can drive a small (75mm) CCFL over a 100µA to 2mA range.

The circuit uses an LT1301 micro-power DC-to-DC converter IC in conjunction with a current-driven, Royer-class converter comprising T1, Q1, and Q2. When power is applied along with intensity-adjust voltage  $V_A$ , the LT1301's  $I_{LIM}$  pin is driven slightly positive, causing maximum switching current to flow through the IC's internal switch pin (SW). L1 conducts current, which flows from T1's center tap, through the transistors, into L1. L1's current is deposited in switched fashion to ground by the regulator's action.

The Royer converter oscillates at a frequency set primarily by T1's characteristics (including its load) and the 0.068µF capacitor. L1 sets the magnitude of the Q1-Q2 tail current, and hence, T1's drive level. The 1N5817 diode maintains L1's current flow when the LT1301's switch is off. The

0.068µF capacitor combines with L1's characteristics to produce sine-wave voltage drive at the Q1 and Q2 collectors. T1 furnishes voltage step-up and about 1400V<sub>P-P</sub> appears at its secondary. Alternating current flows through the 22pF capacitor into the lamp. On positive half-cycles the lamp's current is steered to ground via D1. On negative half-cycles the lamp's current flows through Q3's collector and is filtered by C1. The LT1301's  $I_{LIM}$  pin acts as a zero summing point, with about 25µA bias current flowing out of the pin into C1. The LT1301 regulates L1's current to maintain equality of Q3's average collector current, representing one-half the lamp current, and R1's current, represented by  $V_A/R1$ . When  $V_A$  is set to zero, the  $I_{LIM}$  pin's bias current forces about 100µA bulb current.

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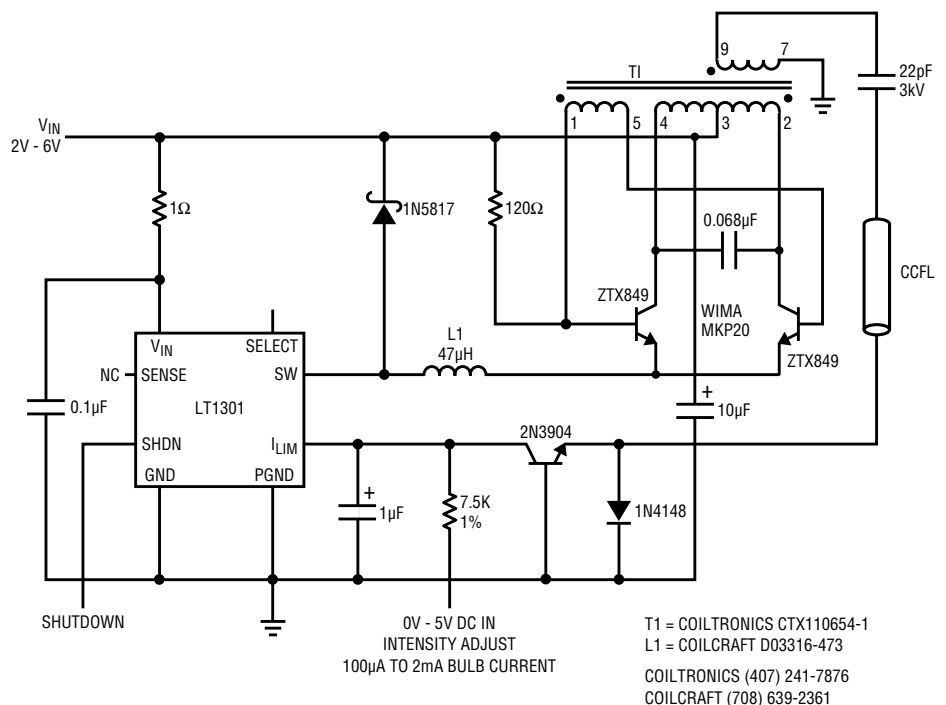


Figure 1. CCFL power supply

# LTC1262 Generates 12V for Programming Flash Memories without Inductors


by Anthony Ng and Robert Reay

Flash memories require a +5V  $V_{CC}$  supply and an additional +12V supply for write or erase cycles. The +12V supply can be a system supply, or be generated from the +5V supply using a DC-DC converter circuit. Single-supply flash memories (i.e., those with the +12V converter built-in) are available, but these memories have lower capacities and slower write/erase performance, and are therefore not as popular as memories without a built-in 12V supply. Flash memories require that the +12V supply be regulated to within 5% and not exceed the permitted maximum voltage (14V for Intel ETOX™ memories). The LTC1262 offers a simple and cost effective +12V programming supply to meet these requirements.

Figure 1 shows a typical application circuit. The only external components needed are four surface-mount capacitors. The LTC1262 uses a triple charge-pump technique to convert +5V to +12V. It operates from 4.75V to 5.5V and delivers 30mA while regulating the 12V output to within 5%. The TTL-compatible SHDN

pin can be driven directly by a microprocessor. When the SHDN pin is taken high (or floated), the LTC1262 enters shutdown mode. In this state, the supply current of the LTC1262 is reduced to 0.5 microamps typical and the +12V output drops to  $V_{CC}$ . When SHDN is taken low, LTC1262 leaves shutdown mode and the output rises to 12V without any potentially harmful overshoot (see Figure 2).

The LTC1262 is available in both 8-pin DIP and narrow SO packages. With small surface-mount capacitors, the complete +12V supply takes up very little space on a printed circuit board. In power sensitive applications, such as PCMCIA flash cards for portable PCs, the LTC1262

shutdown current is low enough to preclude the need for external switching devices when the system is inactive. 

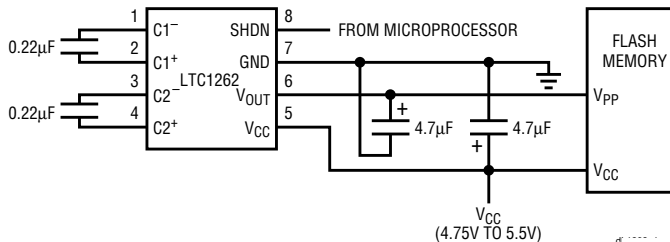


Figure 1. Typical LTC1262 application circuit

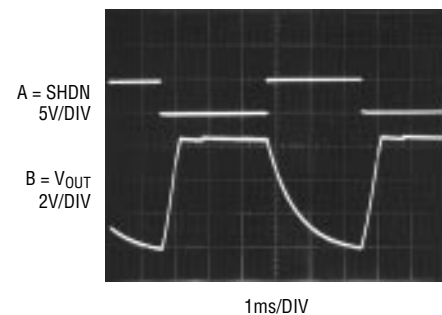


Figure 2. LTC1262 taken in and out of shutdown

# Active-Negation Bus Terminators

by Dale Eagar

High-speed data buses require transmission-line techniques, including termination, to preserve signal integrity. Lost data on a bus can be attributed to reflections of the signals from the discontinuities of the bus. The solution to this problem is proper termination of the bus.

Early designs of bus terminators were passive (see Figure 1). Passive termination works great, but wastes

lots of precious power, especially when the bus is not being used.

The ideal solution is a voltage source capable of both sourcing and sinking current. Such a voltage source, with termination resistors, is shown in Figure 2. This is called active negation. Active negation uses minimal quiescent current, essentially providing only the power needed to properly terminate the bus.

## Active-Negation Bus Terminator Using Linear Voltage Regulation

The active-negation circuit shown in Figure 3 provides the power to the output at an efficiency of about 50%; the rest of the power is dissipated in either Q1 or U1, depending on the polarity of the output current.

The circuit will source or sink current. Current is sourced from the

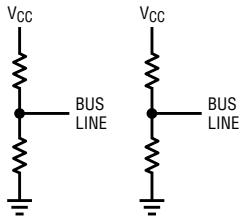


Figure 1. Passive-termination technique

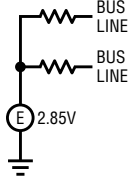


Figure 2. Active-negation termination technique

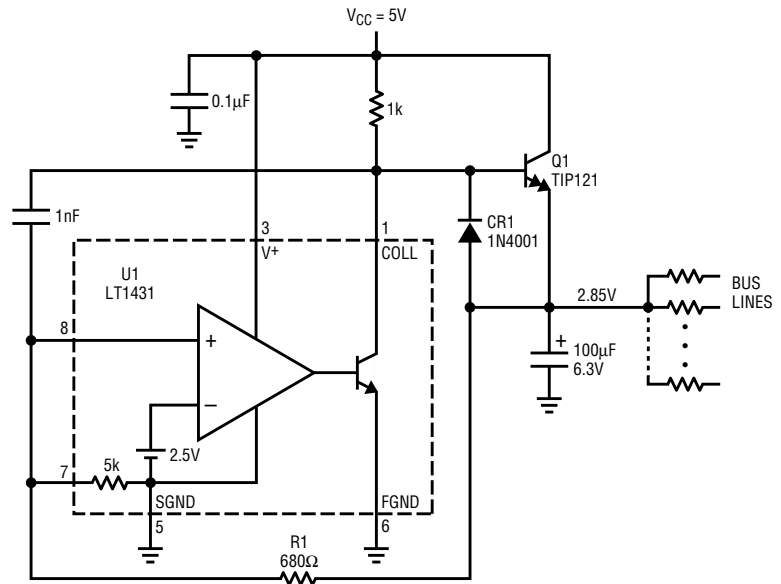


Figure 3. Linear active-negation voltage source

+5V supply through Q1, an NPN Darlington, to the output. The sink current flows through CR1 into the collector (pin 1) of the LT1431, and to ground. The LT1431 regulates a scaled version of the output voltage against the internal 2.5V bandgap reference, driving the base of Q1 or drawing current through CR1 to regulate the output voltage. R1 and the internal 5kΩ resistor of the LT1431 scale the output voltage.

### Switching-Power-Supply, Active-Negation Network

The switching, active-negation terminator shown in Figure 4 is a synchronous switcher. This solution further reduces dissipation and therefore achieves higher efficiency. This type of switcher can both source and sink current.

The switching power supply operates as follows. The 74AC04 hex inverters (U1 and U2) form a 1MHz, variable-duty-factor oscillator. The duty factor is controlled by the output of the regulator, U3, and is maintained at the ratio of  $2.85V/V_{IN}$ .  $V_{IN}$  is the +5V supply, which powers U1, U2, and U3. The output voltage is the average voltage of the square wave ( $V_{IN} \times$  duty factor) from the outputs of

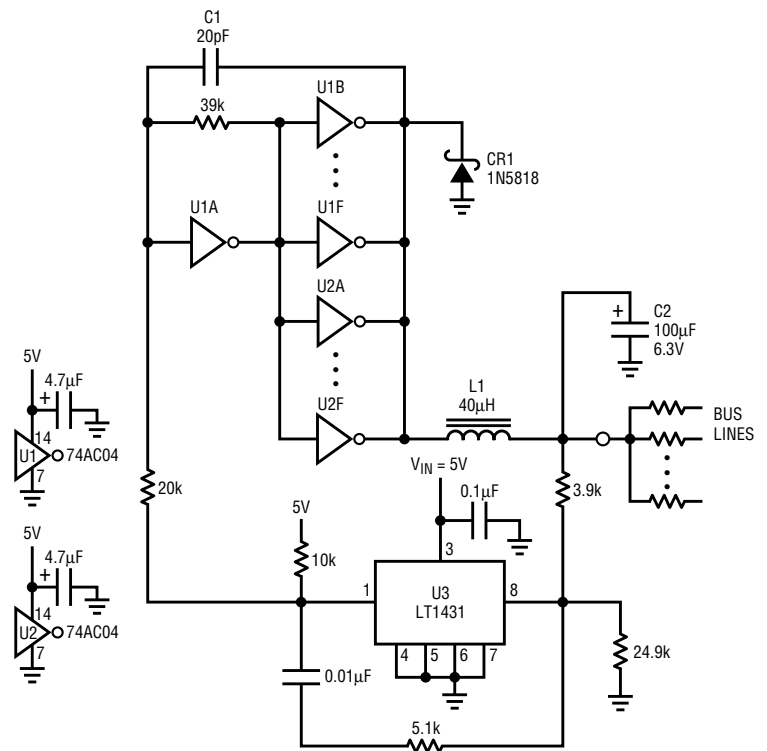


Figure 4. Switching active-negation terminator

U1B–U1F and U2A–U2F. L1 and C2 filter the AC component of the 0-to-5V signal, yielding a DC output voltage of 2.85V.

CR1 is added to prevent latchup of U1 and U2 during adverse conditions.

A logic gate could easily be added to the oscillator to add a disable function to this terminator, further lowering the quiescent power when termination is not needed.

# Extending Op Amp Supplies to Get More Output Voltage

by Dale Eagar

We often hear of applications that require high-output-voltage, low-output-impedance amplifiers. Here is a topology that allows you to extend an op amp's output voltage swing while still maintaining its short circuit protection. The trick is to suspend the op amp between two MOSFET source followers so that the supply voltages track the op amp's output voltage (see Figure 1). The circuit shown in Figure 1 will perform very nicely with any run-of-the-mill ideal op amp. The problem is in the lead times of ideal op amps—they just keep getting pushed out to later dates.

Non-Ideal op amps have realistic lead times and can be made to work in the extended-supply mode. They have bandwidth limitations in both CMRR and PSRR. The circuit shown in Figure 2 implements the extended supply as shown in Figure 1 and has several additional components: C1 is added to decouple the supply, improving high frequency PSRR; R3 and R5 decouple the gates of Q1 and Q2 from AC ground, preventing Q1 and Q2 from running off together to redirect local air traffic; R1, R2, and C4 form a snubber to de-Q the two pole system formed by the Miller capacitance of Q1 and Q2 and the high frequency CMRR of IC1; additionally, R4, R6, C2, C3, Z1, and Z2 form the two 15V voltage sources (E1 and E2 in Figure 1); CR1 and CR2 are protec-

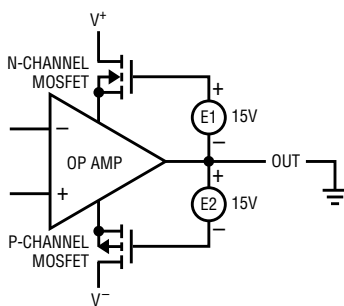


Figure 1. Block diagram of suspended-supply op amp

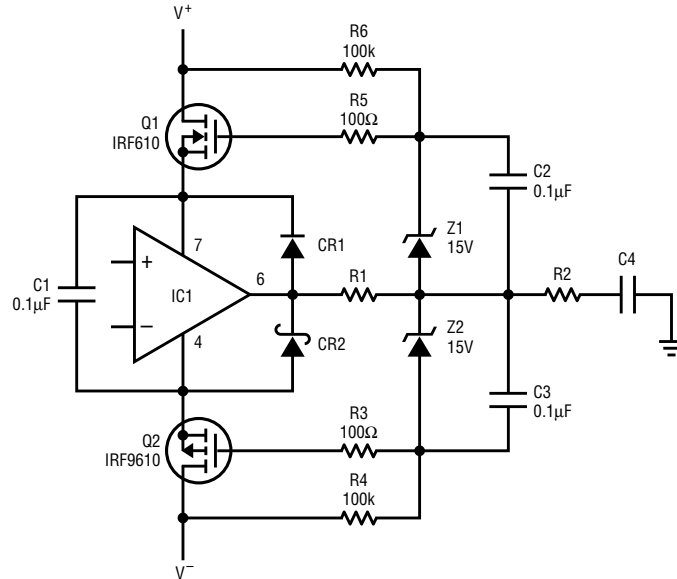


Figure 2. Detailed high-voltage op amp

tion diodes that allow the output to be instantaneously shorted to ground when the output is at any output voltage.

The values of R1, R2, and C4 vary with the MOSFETs' Miller capacitance and with the high frequency CMRR of the op amp used. They are selected to minimize the overshoot in the step response of the amplifier.

## High-Voltage, High-Frequency Amplifier

Using the LT1227 current-feedback amplifier (CFA) in the extended-supply mode as shown in Figure 2, it is relatively easy to get a 1MHz power bandwidth at 100V<sub>P-P</sub> (see Figure 3 for component values). This circuit has short-circuit protection and is stable into all capacitive loads.

## If One Is Good, Are Two Better?

Dual and quad op amps can also be configured with extended supplies, although the design gets just a wee bit tricky. When extending supplies of multiple stages and/or complete

circuits, some design rules need to change. Op amp circuits generally require a ground against which to reference all signals. The problem encountered when using extended-supply mode is that "ground" is swinging through the common-mode range of the op amp and beyond. This raises the following question: "If I cannot reference the signals to ground, to what can I reference them?" The answer? "Use the output as the signal reference." This works for all stages except the last stage, where

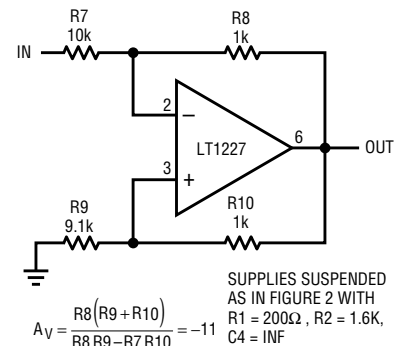


Figure 3. High-speed suspended op amp

SUPPLIES SUSPENDED AS IN FIGURE 2 WITH  
 $R1 = 200\Omega$ ,  $R2 = 1.6K$ ,  
 $C4 = \text{INF}$

$$A_v = \frac{R8(R9 + R10)}{R8R9 - R7R10} = -11$$

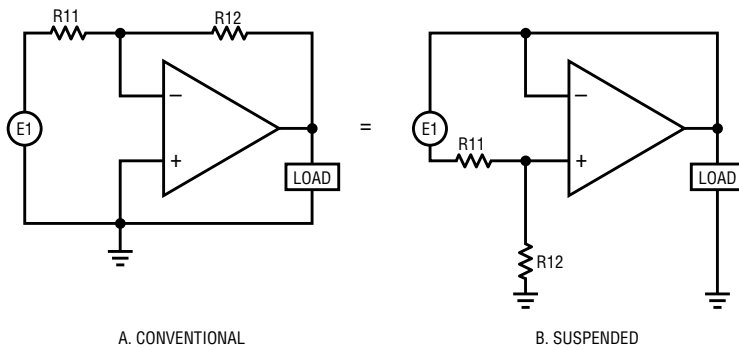


Figure 4. Inverting amplifiers (A. conventional/B. suspended)

comes in. After a few phone calls to make sure that the transformer maker shipped the right one, the engineer (face covered with egg) asks if anyone needs a rather large paperweight. The engineer (still wiping egg from his face) then decides to use switching power supply technology to solve this “simple” problem.

Here is a simple ring-tone generator that can be turned on and off with a logic signal. It has a fully isolated output, is short-circuit protected, and can be powered by any input voltage from 3V to 24V.

### Ring-Tone Generator

### How It Works

Suspended along with the dual op amp in Figure 5 are two voltage references and an oscillator. Keep in mind when referring to Figure 5 that the node labeled “A” is the output; this is the reference common for the references, the oscillator, and the first lowpass filter (U1a). The two references VR1 and VR2 produce  $\pm 2.5V$ . The oscillator U2, running on the  $\pm 2.5V$  references, produces a 20Hz square wave rail-to-rail. U1a is a second-order, Sallen-and-Key lowpass filter that knocks off the sharp edges, presenting the somewhat smoothed signal at point “B.”

Next comes the tricky stuff. U1b is a second-order, multiple-feedback (MFB) lowpass filter/amplifier that

using the output as the reference would simply discard the signal. In the last stage, ground is effectively the output and the feedback resistor is R12. This is shown in Figures 4a and 4b. Figure 4a shows a conventional inverting amplifier where the input and output signals are referred to ground. Figure 4b shows the equivalent circuit implemented in the extended-supply mode.

Here are two rules for design in the extended-supply mode, which will be demonstrated in the next application:

Rule 1: When designing multiple stages in the extended-supply mode, reference the signals of all stages except the last to the output of the last stage.

Rule 2: Invert the signal using the circuit in Figure 4b at the last stage.

Ring-tone generators are sine-wave-output, high-voltage inverters for the specific purpose of ringing telephone bells. In decades past, the phone company generated their ring tones with motor generator sets with the capacity to ring numerous phones simultaneously. Often, ring tones are 20Hz at 90V with less than 10mA per bell output current capability. Since the power supplied is low, one would think that the task is minimal. This is not always so. “It’s simple—no problem,” is often heard in response to queries about ring-tone generators. “Just hook a couple of logic-level FETs to two spare output bits of the microprocessor and hook their drains to the primary side of a transformer, with the center tap hooked to 5V, or 12V, or whatever.” At this point everyone is happy until the transformer

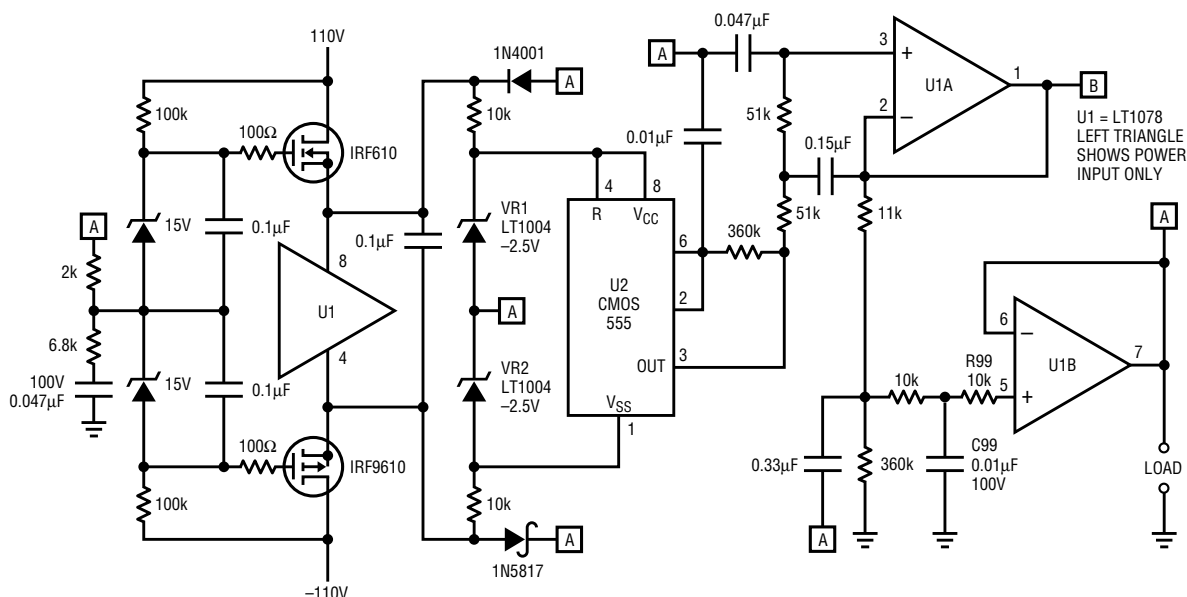


Figure 5. Ring-tone generator: oscillator, filter, and driver.

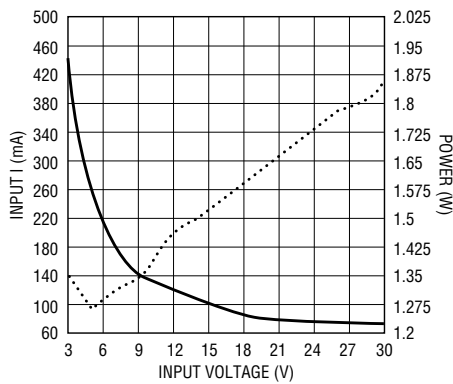


Figure 6. Input-current and input power versus input voltage while ringing one bell, for circuit shown in Figure 5.

performs four functions: first, it subtracts the voltage at point “A” (its own output voltage) from the voltage at point “B” (the incoming signal), forming a difference that is the signal; second, it filters the difference signal with a two-pole lowpass filter, smoothing out the last wrinkles in the signal; third, it amplifies the filtered difference signal with a gain of 34; and fourth, it references the amplified signal to ground, forming the output.

Note that R99 shown in Figure 5 is there to protect the input of U1b in the event that the output is shorted when the output voltage is very high. This measure is necessary because the bottom end of C99 is connected to ground, and C99 could have up to 100V across it. When the output is shorted to ground from a high voltage, R99 limits the current into the input of U1b to an acceptable level.

This circuit, when coupled with the switching power supply shown in Figure 7, implements a fully isolated sine-wave ring-tone generator.

The input current and power versus input voltage for the combination ring-tone generator (Figures 5 and 7) are shown in Figure 6. The output waveform (loaded with one bell) is shown in Figure 8, and the harmonic distortion is shown in Figure 9.

Although somewhat tricky at first, extended-supply mode is a valuable tool to get out of many tight places. There is also a great deal of satisfaction to be gathered when making it work, for those of you who love a technical challenge.

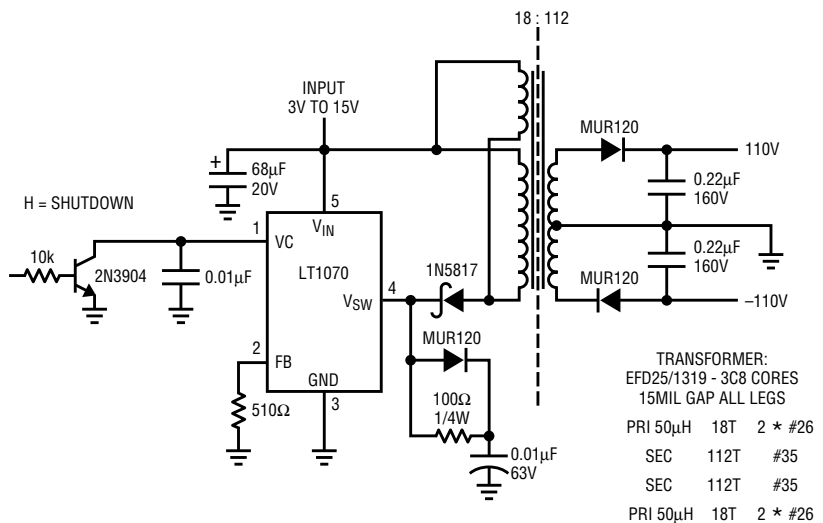


Figure 7. High-voltage power supply for ring-tone generator

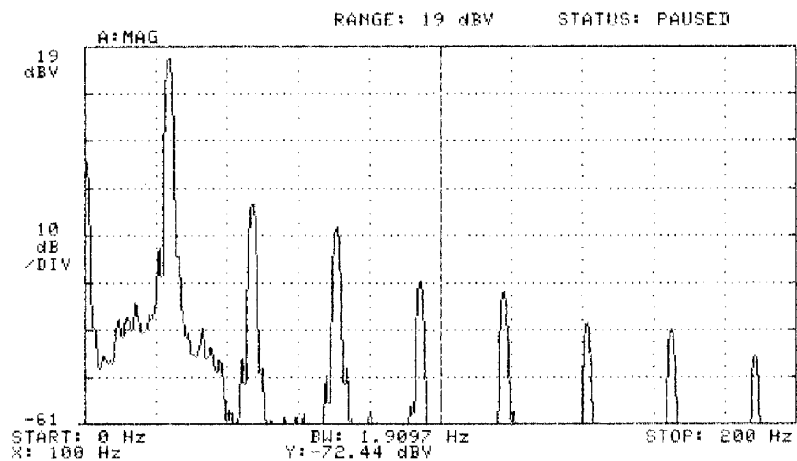


Figure 8. Ring-tone generator frequency-spectrum plot

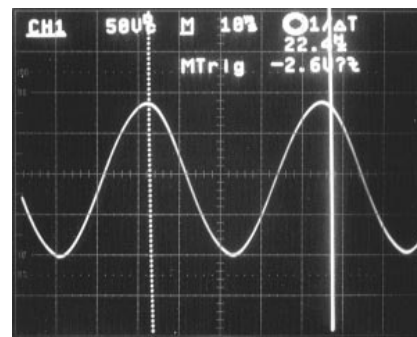


Figure 9. Sine-wave output from ring-tone generator

# Simple PCMCIA VPP Socket Switching for Line-Operated Systems

by Doug La Porte

PCMCIA card sockets are not only for portable systems. Many line-operated systems are using PC cards for easier system software upgrades or feature additions. For line-operated systems requiring a PCMCIA card socket, there is often a higher voltage supply (>13V) that can be converted to the required VPP voltages with a linear regulator and some additional logic. Switching supply configurations could be used, but often EMI considerations and circuit simplicities outweigh the need for better efficiency. Figure 1 shows a circuit that uses a commonly available 24V supply and the LT1121 adjustable regulator.

## The Circuit

The circuit uses a simple linear voltage regulator to achieve four VPP voltage states: 0V, 5V, 12V, and a high-impedance state. The LT1121 has the ability to supply the necessary currents at the VPP pin, and has current- and thermal-limiting features. The limiting protection can be very important, because the designer can never be sure of the condition of the PC card being plugged into the socket. To achieve the high-imped-

ance state, the LT1121's shutdown feature is used. In this mode, the circuit consumes only 16 $\mu$ A. The 0V state is attained by switching Q1 while the LT1121 is shutdown. The 5V and 12V states are achieved by grounding R3 and R4, respectively. The capacitor C2 slows the rise and fall times of the output voltage to minimize overshoot and assures compliance with flash memory requirements.

## The Logic

The logic shown in Figure 1 interfaces directly to PCMCIA logic controllers. There are many ways to implement the required logic. The method shown can be implemented using some spare logic from a PLD, gate array, or ASIC. The pull-down resistors R5 and R6 force the output to the high-impedance state if the 5V logic supply fails.

The three-state output HC126 gates must be CMOS. The three-state condition allows the unselected resistor to float, and a CMOS logic low level is required to pull the selected resistor to ground. These two gates should be located close to the resistors and the LT1121. If spare logic

cannot be close to the LT1121, the EN0 and EN1 signals can be buffered remotely to drive two logic level MOSFET's close to the LT1121, as shown in Figure 2.

## Thermal Considerations

Connecting the LT1121 directly to the 24V supply is not recommended. The maximum voltage rating of the part is not exceeded in this configuration, but a minimum of 12V will be dropped across the regulator. At a current draw of 120mA the part will have to dissipate 1.44W of power. The "A" suffix SO-8 part specified here has a junction-to-ambient thermal resistance of 60°C/W (that of the "non-A" suffix SO-8 part is 120°C/W) with a maximum operating junction temperature of 125°C. This would put the maximum operating ambient temperature at 38°C. This is usually not acceptable. Two 4.3V, 1W zener diodes lower the voltage drop across the LT1121A. In this configuration the input voltage to the LT1121 will be at about 15V to 16V, depending on the current requirement, greatly decreasing the power dissipation of the part.  $\blacktriangleleft$

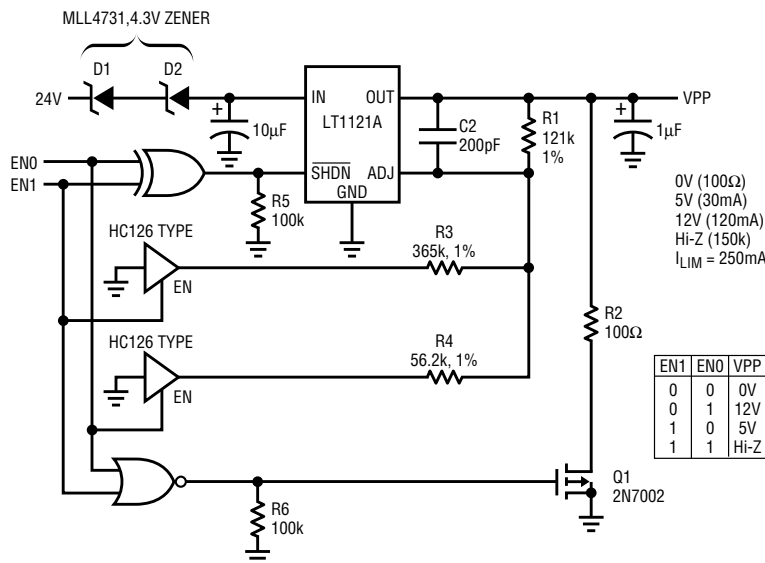


Figure 1. 24V-input PCMCIA socket VPP switch/regulator

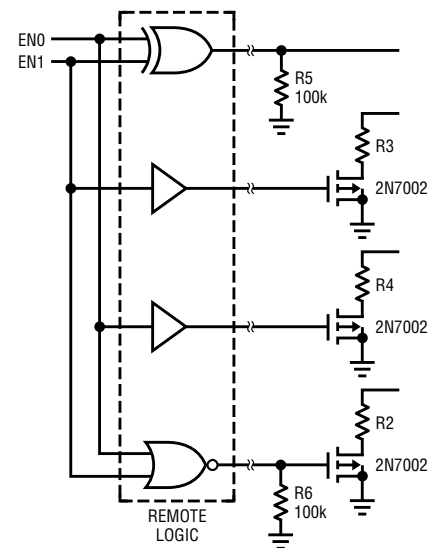
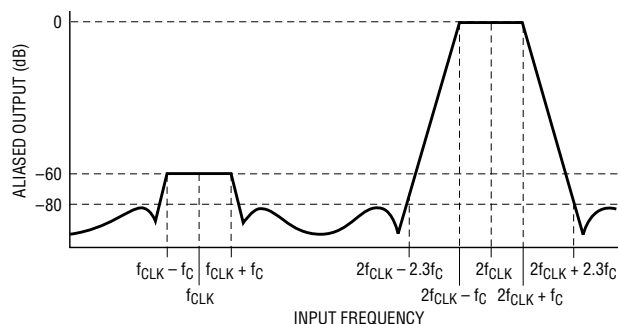


Figure 2. Using additional MOSFETs for remote logic interface

# DC-Accurate, Clock-Tunable, Lowpass Filter with Input Anti-Aliasing Filter

by Philip Karantzalis



**Figure 1. Aliasing versus frequency  $f_{CLK}/f_C = 50:1$  (pin 8 to V+); Clock is a 50% duty-cycle square wave.**

In a sampled-data system the sampling theorem says that if an input signal has any frequency components greater than one half the sampling frequency, aliasing errors will appear at the output. In practice, aliasing is not always a serious problem. High-order switched-capacitor lowpass filters are band-limited and significant aliasing occurs only for input signals centered around the clock frequency and its multiples.

Figure 1 shows the LTC1066-1 aliasing response when operated with a clock-to- $f_C$  ratio of 50:1. With a 50:1 ratio, the LTC1066-1 samples its input twice during one clock period and the effective sampling frequency is twice the clock frequency. Figure 1 shows that the maximum aliased output is generated for inputs in the range of  $2 \times f_{CLK} \pm f_C$ . ( $f_C$  is the cutoff frequency of the LTC1066-1.) For instance, if the LTC1066-1 is programmed to produce a cutoff frequency of 20kHz with a 1MHz clock, maximum aliasing will occur only for input signals in the narrow range of 2MHz  $\pm$ 20kHz and its multiples.

The simplest anti-aliasing filter is a passive, first-order lowpass RC filter. The -3dB frequency of the RC filter should be chosen so that the passband of the RC filter does not influence the passband of the

flat passband of the LTC1066-1 at 10kHz (the passband attenuation of the 50kHz RC filter for frequencies less than 10kHz is less than 0.2dB). If

LTC1066-1. For the situation when the LTC1066-1 clock frequency is 500kHz, an RC filter with the -3dB frequency set at 50kHz attenuates by 26dB any possible aliasing inputs in the range 1MHz  $\pm$ 10kHz. The passband shape of the 50kHz RC filter does not degrade the

the LTC1066-1 is clock tuned to a cutoff frequency of 5kHz (with a clock frequency of 250kHz), the 50kHz RC filter will provide 20dB attenuation for aliasing inputs in the range of 500kHz  $\pm$ 5kHz. Therefore, a first-order lowpass RC filter will attenuate all aliasing signals to the LTC1066-1 by a minimum of 20dB for a clock-tunable range of one octave.

For added anti-aliasing bandwidth, a first-order, lowpass RC filter can be tuned by the clock signal of LTC1066-1 to follow the cutoff frequency of the higher-order filter. The circuit is shown in Figure 2. The circuit operation is as follows. The six comparators

**Table 1. Component calculations for the circuit in Figure 3.**

- Definitions:
1. The cutoff frequency of the LTC1066-1 is abbreviated as  $f_C$ .
  2.  $f_{C\ LOW}$  is the lowest cutoff frequency of interest.
  3. A range of five octaves is from  $f_{C\ LOW}$  to  $32 \times f_{C\ LOW}$ .

Component Calculations:

$$\frac{1}{2\pi R_F C_F} = \frac{f_{C\ LOW}}{250} \quad R_{IN} = R_F \text{ (if } R_F \text{ can be chosen as 20k)}$$

$R_{IN}$  and  $C_{IN}$  are not needed)

$$C_1 = \frac{1}{f_{C\ LOW}} \mu F \quad (f_{C\ LOW} \text{ in Hz); } R_1 = 1k$$

$$C_2 = C_1 \pm 5\%; C_3 = 2 \times C_1 \pm 5\%; C_4 = 4 \times C_1 \pm 5\%; C_5 = 8 \times C_1 \pm 5\%$$

$$C_P = 50pF; R_P = \frac{10^5}{50 f_{C\ LOW}} k\Omega$$

$$C_A = 0.047\mu F; R_A = \frac{5 \times 10^5}{50 f_{C\ LOW}} k\Omega$$

Example:

for a five-octave range from 1kHz to 32kHz—  
 $f_{C\ LOW} = 1kHz$

Let  $C_F = 1\mu F \pm 20\%$ , then  $R_F = 40.2k\Omega \pm 1\%$ .  $R_{IN} = R_F = 40.2k\Omega \pm 1\%$ ,  $C_{IN} = 0.1\mu F$

$C_1 = 0.001\mu F \pm 5\%$ ,  $C_2 = 0.001\mu F \pm 5\%$ ,  $C_3 = 0.0022\mu F \pm 5\%$

$C_4 = 0.0039\mu F \pm 5\%$ ,  $C_5 = 0.0082\mu F \pm 5\%$

$C_P = 50pF$ ,  $R_P = 2k\Omega \pm 5\%$ ,  $C_A = 0.047\mu F$ ,  $R_A = 10k\Omega \pm 1\%$

$C_P = 50pF$ ,  $R_P = 2k\Omega$ ,  $C_A = 0.047\mu F$ ,  $R_A = 10k\Omega$



inside the LTC1045 detect the clock frequency. The clock signal of the LTC1066-1 is converted to a pulse output whose duty cycle changes with clock frequency. The average voltage of the pulse signal is delivered to a four-window comparator whose outputs drive the four analog switches of

the LTC202. When the LTC1066-1 clock frequency increases or decreases by more than one octave ( $2x$  or  $x/2$ ), a capacitor is switched in or out of the first order lowpass filter formed by resistor R1 ( $1k\Omega$ ) and capacitor C1. The  $-3dB$  frequency of the lowpass RC filter is therefore doubled or halved if the cutoff frequency of the LTC1066-1 is doubled or halved. Resistor R1 and capacitors C1 through C5 allow the lowpass RC filter to be tuned over a range of five octaves, providing at least 20dB attenuation

to any LTC1066-1 input signals in the range  $2 \times f_{CLK} \pm f_C$  (the RC filter also attenuates all aliasing signals near any multiples of the clock frequency).

The circuit in Figure 2 can be used for any clock-tunable, five-octave range for cutoff frequencies from 10Hz to 80kHz (with  $\pm 5$  volt supplies for LTC1066-1) or for cutoff frequency as high as 100kHz (with  $\pm 8$  volt supplies for the LTC1066-1). For cutoff frequencies greater than 50kHz, a 15pF capacitor in series with a 30k $\Omega$  resistor should be connected between pins 11 and 13 of the LTC1066-1 to minimize passband gain peaking. Table 1 provides a design guide for choosing the component values of  $R_A$ ,  $R_P$ ,  $R_F$ ,  $R_{IN}$ ,  $C_F$ , C1 through C5,  $C_P$ , and  $C_A$ .

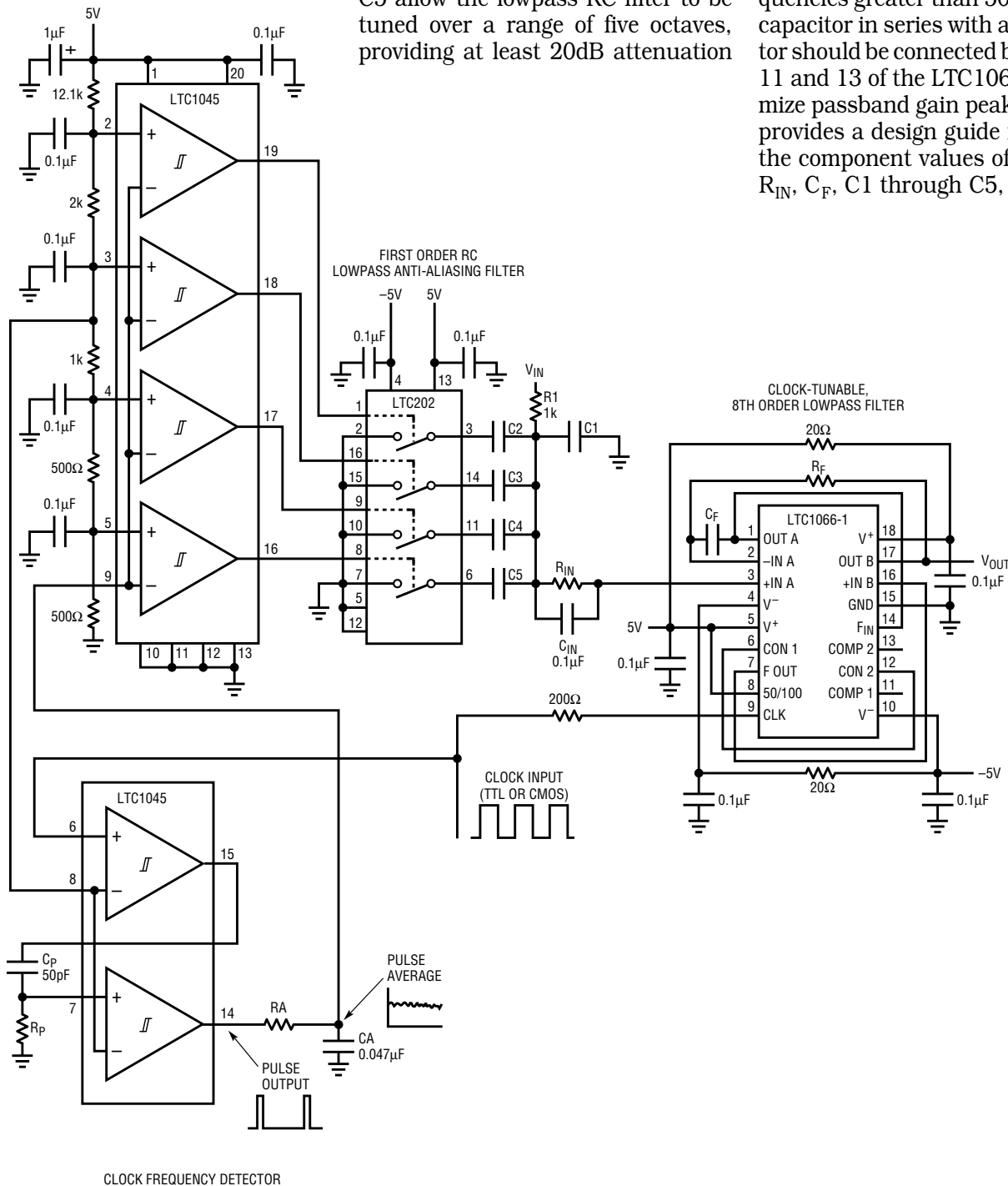


Figure 2. DC-accurate, clock-tunable lowpass filter with input anti-aliasing.

# Regulated Charge-Pump Power Supply

by Tommy Wu

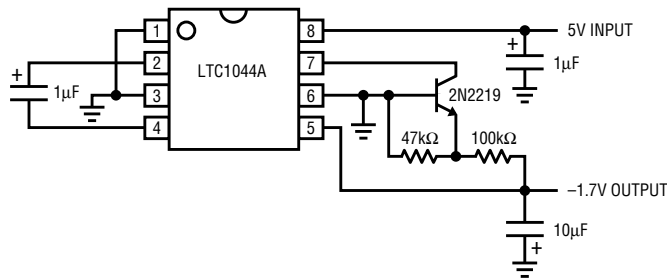


Figure 1. Regulated charge pump

The circuit shown in Figure 1 uses an LTC1044A charge pump inverter to convert a +5 volt input to a -1.7V potential as required for a certain LCD panel. Output regulation is provided by a novel feedback scheme, which uses components Q1, R1, and R2. Without feedback, the charge pump would simply develop approximately -5V at its output. With feedback applied,  $V_{OUT}$  charges in the negative direction until the emitter of Q1 is biased by the divider comprising R1 and R2. Current flowing in the collector tends to slow the LTC1044A's internal oscillator, reducing the available output current. The output is thereby maintained at a constant voltage.

In this application less than 5mA output current is required. As a result, charge-pump capacitor C1 is reduced to 1 microfarad from the usual 10 microfarads. Curves of output voltage with and without feedback are shown in Figure 2. The equivalent

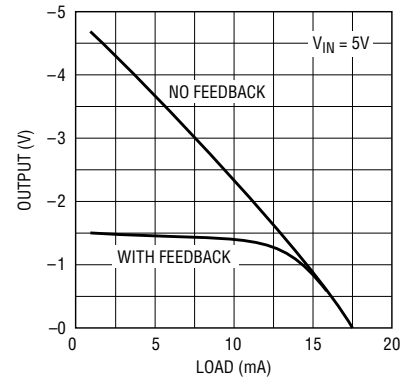


Figure 2. Effect of feedback on output voltage

output impedance of the charge pump is reduced from approximately 100 ohms to 5 ohms.

A variety of output voltages within the limits of the curve in Figure 2 can be set by simply adjusting the  $V_{BE}$  multiplier action of Q1, R1, and R2. Tighter regulation or a higher tolerance could be obtained by adding a reference or additional gain, at the expense of increased complexity and cost.

# Low-Noise Wireless-Communications Power Supply

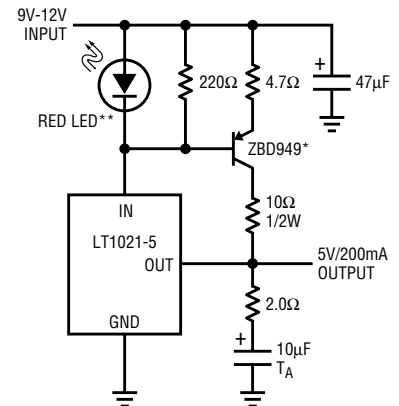
by Mitchell Lee and Kevin Vasconcelos

Shown in Figure 1 is a 5V power supply we designed for a synthesizer oscillator. The original design used a three-terminal regulator, but the regulator's voltage noise contributed to excessive phase noise in the oscillator, leading us to this solution. Of prime importance is noise over the 10Hz-to-10kHz band. Careful measurements show a 40dB improvement over standard three-terminal regulators.

The regulator is built around a 5V, buried-zener reference. It is the buried zener's inherently low noise that makes the finished supply so quiet. Measured over a 10Hz-to-10kHz band, the 5V output contains just  $7\mu V_{RMS}$  noise at full load. The 10Hz-

10kHz noise can be further reduced to  $2.5\mu V_{RMS}$  by adding a  $100\mu H$ ,  $1000\mu F$  output filter. The noise characteristics of the reference are tested and guaranteed to a maximum of  $11\mu V$  over the band of interest.

An external boost transistor, the ZBD949, provides gain to meet a 200mA output current requirement. Current limiting is achieved by ballasting the pass transistor and clamping base drive. Although our application only requires 200mA, it is possible to extend the output current to at least 1A by selecting an appropriate ballast resistor and addressing attendant thermal considerations in the pass transistor.



\* ZETEX INC (516) 864-7630  
\*\* GLOWS IN CURRENT LIMIT. DO NOT OMIT.

Figure 1. Ultra low noise 5V, 200mA Supply Output Noise is 7 microvolts RMS over a 10Hz to 10kHz bandwidth. Reference noise is guaranteed less than 11 microvolts RMS. Standard, three-terminal regulators have 100 times the noise and no guarantees.

LT1251/LT1256, continued from page 1

through and cause ghosts. The most common way to ensure that only the selected input is on is to overdrive the control input below zero and above 100%. Unfortunately, this means that the gain is not a simple function of the control signal. The LT1251 eliminates this problem with special circuitry that ensures that, for a control signal of 2% or 98%, only one input is on and the other is completely off. The LT1256 does not have this special circuit and is linear from zero to 100%.

The LT1251/LT1256 has flat response (0.1dB) from DC to 5MHz and low differential gain and phase (0.1% and 0.1°) for composite video applications. The signal-to-noise ratio (5MHz bandwidth) is 80dB referenced to 1V. The -3dB bandwidth is greater than 30MHz, ideal for computer RGB applications, and the absolute gain accuracy is better than 3%. The LT1251/LT1256 operate on a single +5V supply as well as on dual supplies from ±5V to ±15V. The output is able to drive all resistive loads, including a doubly terminated cable.

The LT1251/LT1256 can accommodate a variety of control signals and levels. Zero to one or zero to ten volts are common, as are bipolar signals around zero. Some systems use current inputs or voltages connected to resistors feeding the summing node of an op amp. Often, in variable-gain amplifier applications, several control inputs are summed together. In

order to make the LT1251/LT1256 compatible with as many systems as possible, the input and output of an internal op amp are available to the user.

**The LT1251/LT1256 has flat response from DC to 5MHz and low differential gain and phase for composite video applications...**

**The -3dB bandwidth is greater than 30MHz, ideal for computer RGB applications**

The closed-loop feedback topology of the LT1251/LT1256 minimizes control feedthrough and DC shift at the output. The typical feedthrough is only 2.5mV peak-to-peak and the DC shift between inputs is guaranteed to be less than 4.5mV.

**Circuit Description**

Figure 2 shows the block diagram of the LT1251/LT1256 signal path. The potentiometer of Figure 1 has been replaced with the mathematical equivalent blocks: K, 1 - K, and summation. K is a constant determined by the control circuit and can be any value between 0 and 1. To make the LT1251/LT1256 more versatile, the

maximum gain for each input is set by external resistors. The input stages use current feedback, and the external resistors set both the gain and bandwidth.

For ideal op amp parameters and  $R_{F1} = R_{F2}$ , the equation for the gain of Figure 2 is:

$$V_0 = K \times IN1 \times A_{V1} + (1 - K) \times IN2 \times A_{V2}$$

where  $A_{V1} = 1 + R_{F1}/R_{G1}$  and  $A_{V2} = 1 + R_{F2}/R_{G2}$

This shows that as K goes from 0 to 1, the output fades linearly from input 2 times its gain, to input 1 times its gain. The gains are set by the external resistors in the same way that the gain of an op amp is set. For the inverting case,  $A_{V1}$  and  $A_{V2}$  are  $-R_F/R_G$ , again just like a standard op amp.

The complete gain equation with all the gory details is derived in the LT1251/LT1256 data sheet.

Figure 3 shows the control-circuit block diagram. The LT1251/LT1256 consists of two identical voltage-to-current converters (V-to-I); each V-to-I contains an op amp, an NPN transistor and a resistor. The converter on the right generates a **full scale** current,  $I_{FS}$ ; the one on the left generates a **control** current,  $I_C$ . The ratio  $I_C/I_{FS}$  is K; K ranges from a minimum of zero (when  $I_C$  is zero) to a maximum of one (when  $I_C$  is equal to or greater than  $I_{FS}$ ). The parameter K determines the gain from each signal input to the output.

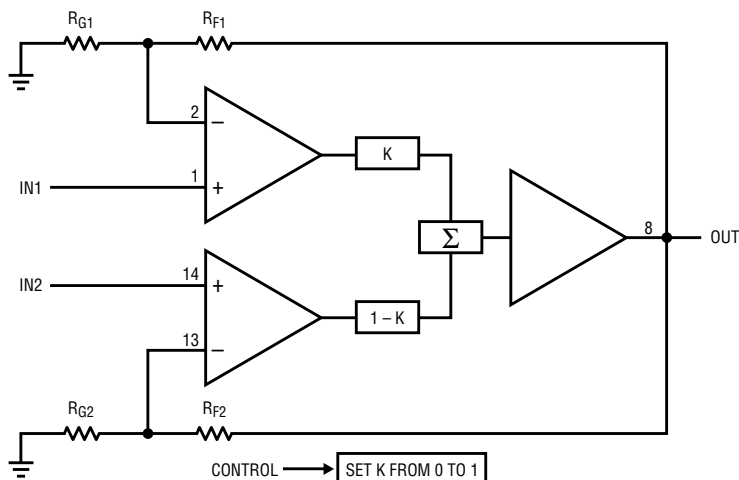


Figure 2. LT1251/1256 signal-path block diagram

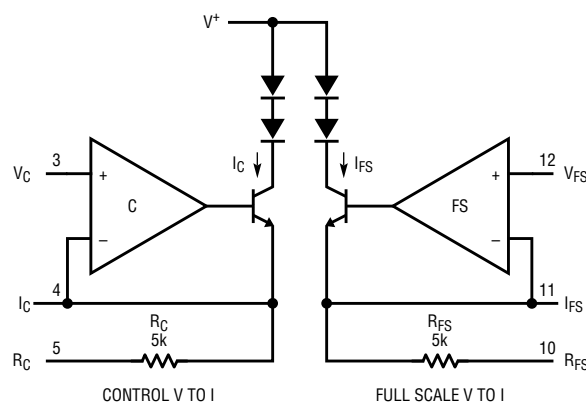


Figure 3. LT1251/1256 Control circuit block diagram

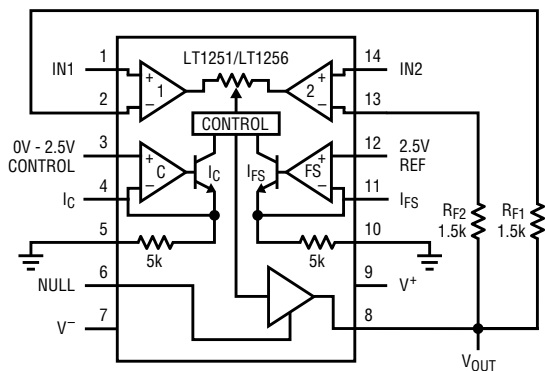


Figure 4. Two-input video fader

The op amp in each V-to-I converter drives the transistor until the voltage at the inverting input is the same as the voltage at the noninverting input. If the open end of the resistor (pin 5 or 10) is grounded, the voltage across the resistor is the same as the voltage at the noninverting input. The emitter current is therefore equal to the input voltage ( $V_C$ ) divided by the resistor value ( $R_C$ ). The collector current is essentially the same as the emitter current; it is the ratio of the two collector currents that sets the gain.

The LT1251/LT1256 is tested with pins 5 and 10 grounded and a full-scale voltage of 2.5V applied to  $V_{FS}$  (pin 12). This condition sets  $I_{FS}$  at approximately 500 $\mu$ A; the control voltage ( $V_C$ ) is applied to pin 3. When the control voltage is negative or zero,  $I_C$  is zero and  $K$  is zero. When  $V_C$  is 2.5V or greater,  $I_C$  equals  $I_{FS}$  and  $K$  is one. The gain of channel one ranges from 0% to 100% as  $V_C$  increases from zero to 2.5V. The gain of channel two moves the opposite way, from

100% down to 0%. The worst-case error in  $K$  (the gain) is  $\pm 3\%$ , including initial accuracy and temperature effects. By using a 2.5V full-scale voltage and the internal resistors, no additional errors need be accounted for.

In the LT1256,  $K$  changes linearly with  $I_C$ . To ensure that  $K$  is zero,  $V_C$  must be  $-15$ mV or lower to overcome the worst-case control-op-amp offset. Similarly, to ensure that  $K$  is 100%,  $V_C$  must be 3% larger than  $V_{FS}$ , based on the guaranteed gain accuracy.

To eliminate the overdrive requirement, the LT1251 has internal circuitry that senses when the control current is at 5% and sets  $K$  to 0%; similarly at 95% it sets  $K$  to 100%. The LT1251 guarantees that 2% (50mV) input ensures  $K = 0$  and 98% (2.45V) results in  $K = 100\%$ .

The operating currents of the LT1251/LT1256 are derived from  $I_{FS}$  and therefore the quiescent current is a function of  $V_{FS}$  and  $R_{FS}$ . An approximate formula for the supply current is:

$$I_S = 1\text{mA} + 24 \times I_{FS} + V_S/20\text{k}$$

where  $V_S$  is the total supply voltage between pin 9 and 7. Using the internal resistors (5k) with  $V_{FS}$  equal to 2.5V results in  $I_{FS}$  equal to 500 $\mu$ A and a supply current of 14.5mA on  $\pm 15$ V supplies. The supply current can be reduced by reducing  $I_{FS}$ , but the slew rate and bandwidth will also be reduced. There is no reason to use a larger value of  $I_{FS}$ .

Performance

Figure 4 shows the LT1251/LT1256 configured as a fader with unity gain. A full-scale voltage of 2.5V is applied to  $V_{FS}$  (pin 12) and the control input drives pin 3. Figure 5 shows the typical linearity of LT1256 gain; the worst-case error, including temperature effects, is  $\pm 3\%$ . Figure 6 shows the frequency response for both the ON and OFF channels. At 5MHz the OFF channel is down 60dB to prevent ghosts from haunting us.

The control path has a 10MHz bandwidth. Figure 7 shows the

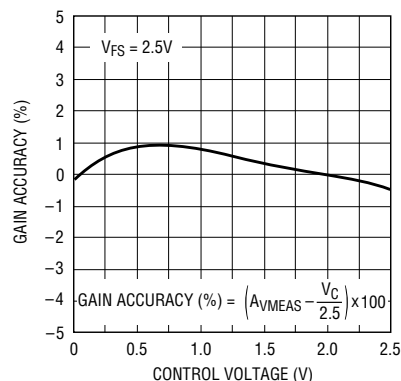


Figure 5. LT1256 Gain accuracy versus control voltage

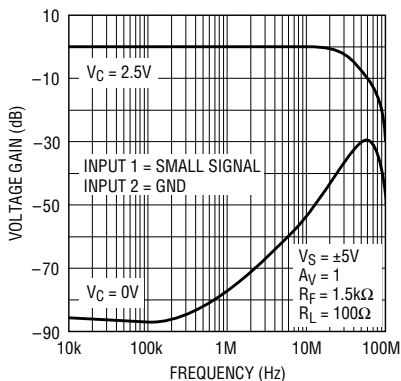


Figure 6. LT1251/1256 Gain versus frequency

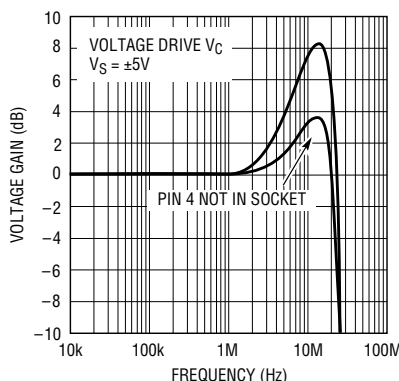


Figure 7. LT1251/1256 Control-path bandwidth

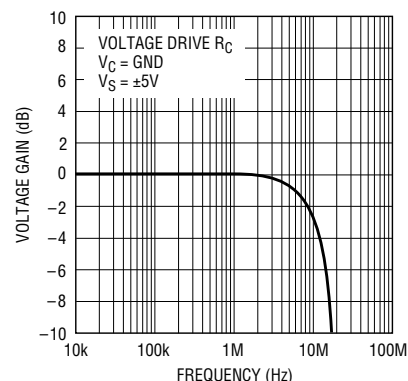


Figure 8. LT1251/1256 Control-path bandwidth

response while driving pin 3; the peaking is caused by the stray capacitance from pin 4 to ground. This capacitance is in parallel with the internal 5k resistor between pins 4 and 5. Grounding pin 3 and driving pin 5 eliminates the effect of the stray capacitance to ground, since pin 4 becomes a virtual ground, as when keying on a color or luminance level. Figure 8 shows the true response of the control path. The control path is fast enough for quick switching between signals, as when keying on a color or luminance level. The control path introduces only a small (50mV), short (50ns) glitch when switched quickly.

A summary of the LT1251/LT1256 performance operating on  $\pm 5V$  supplies in the configuration shown in Figure 4 is given in Table 1.

### Applications

Grounding IN2 of the LT1256 in Figure 4 results in a two-quadrant multiplier. Figure 9 shows the two-quadrant multiplier being used as an AM modulator. The output will deliver +10dBm into 50 ohms. The LT1077 op amp senses the LT1256 output DC and drives the null pin, eliminating any DC at the output. The null-pin voltage is nominally 100mV above the negative supply and therefore the op amp output must be able to swing within a few millivolts of the negative supply. Without the LT1077, the worst-case DC output voltage is 50mV.

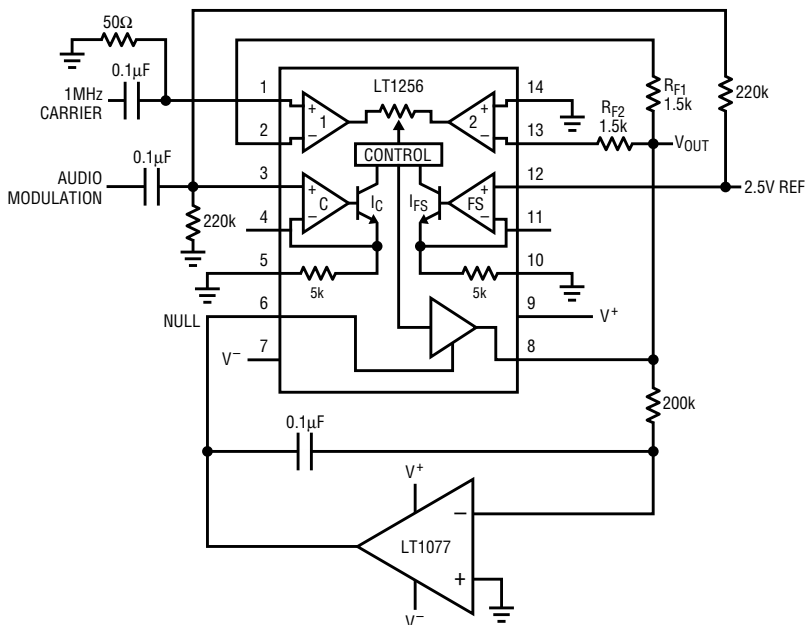


Figure 9. AM modulator with DC-output nulling circuit

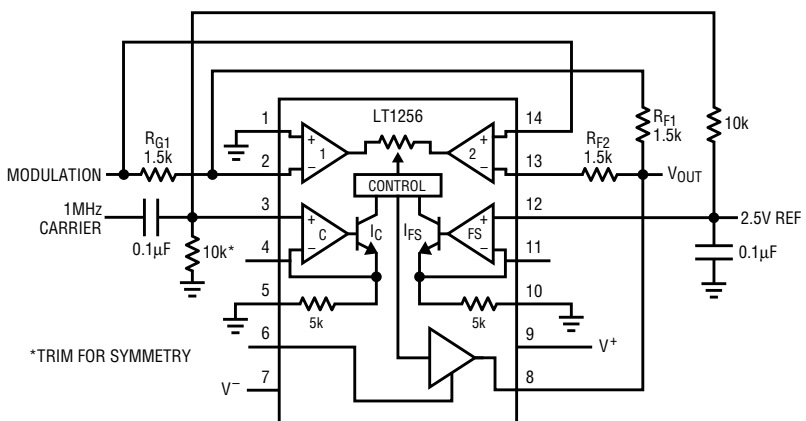


Figure 10. Four-quadrant multiplier used as a double-sideband, suppressed-carrier modulator

Table 1. LT1251/LT1256 performance summary

Slew Rate (@ $\pm 2V$ , $R_L = 150\Omega$ )	300V/ $\mu$ s
Full-Power Bandwidth ( $1V_{RMS}$ )	30MHz
Small-Signal Bandwidth	30MHz
Differential Gain (NTSC, $R_L = 150\Omega$ )	0.1%
Differential Phase (NTSC, $R_L = 150\Omega$ )	0.1DEG
Total Harmonic Distortion (1kHz, $K = 1$ )	0.001%
(1kHz, $K = 0.5$ )	0.01%
(1kHz, $K = 0.1$ )	0.4%
Rise Time, Fall Time	11 ns
Overshoot	3%
Propagation Delay	10 ns
Settling Time (0.1%, $V_O = 2V$ )	65 ns
Quiescent Supply Current	13.5mA

By operating one input stage in an inverting configuration and the other in a noninverting configuration and driving both inputs, the LT1256 becomes a four-quadrant multiplier. Figure 10 shows the four-quadrant multiplier being used as a double-sideband, suppressed-carrier modulator. The LT1077 DC-output-nulling circuit could be added if necessary.

The LT1251/LT1256 can be used to implement numerous other functions, including voltage-controlled filters, phase shifters and oscillators. Squaring and limiting circuits can be designed by feeding the output or input into the control pins. Gamma correction and other compression circuits are created in a similar manner. The applications are limited only by the designer's imagination.  $\blacktriangleleft$

# New Device Cameos

## **LTC1347 Ultra-Low-Power 5V-Powered RS232 Transceiver with Five Receivers Active in SHUTDOWN Mode**

The LTC1347 is a new three-driver, five-receiver RS232 interface transceiver with an integral charge-pump power generator for single 5V supply operation. The circuit has exceptionally low power consumption: 300 $\mu$ A in normal operation, and only 80 $\mu$ A in SHUTDOWN mode. All five receivers remain active during SHUTDOWN to allow monitoring of all data lines for incoming data. The LTC1347 is pin compatible with the LT1137A, and operates with only four 0.1 $\mu$ F charge-pump capacitors.

The LTC1347 meets or exceeds all EIA/TIA-232 specifications. Operation at data rates up to 120k baud is guaranteed. Slew rate with a 3k $\Omega$  2500pF load is a minimum of 3V/ $\mu$ s. When powered down or in SHUTDOWN, the driver outputs remain high impedance for line voltages to  $\pm$ 25V.

Like all Linear Technology RS232 transceivers, the LTC1347 is protected against  $\pm$ 10kV ESD strikes to the RS232 inputs and outputs. This integrated ESD resistance saves the expense and space of external protection devices.

The circuit is available in 28-lead DIP, SOIC, and SSOP packages.

## **The LTC1348—True RS232 from 3.3V**

The LTC1348 is a three-driver, five-receiver RS232 transceiver designed to work from a single 3.3V supply. It uses a voltage-tripling charge pump to generate true RS232 output swings from a supply as low as 3.0V, while drawing only 500 $\mu$ A of quiescent current—the lowest in the industry. It also features low-current shutdown and receiver keep-alive modes for additional power savings. The charge pump requires only five space-saving 0.1 $\mu$ F capacitors to operate. The

transceiver will operate at speeds up to 50kbaud with the maximum 2500pF, 3k $\Omega$  load, or as fast as 120kbaud with a 1000pF, 3k $\Omega$  load. It will withstand repeated ESD strikes of up to  $\pm$ 10kV at the driver outputs and receiver inputs without damage.

The transceiver operates in one of four modes: normal, receiver-alive, receiver-disable, and shutdown. In normal or receiver-disable modes, supply current is only 500 $\mu$ A with all RS-232 outputs unloaded. In receiver-alive mode, all five receivers are kept alive and the supply current is reduced to 12 $\mu$ A. Shutdown drops supply current to less than 1 $\mu$ A. All RS-232 outputs assume high impedance states in shutdown or receiver-alive modes, or when the power is off. The receiver outputs assume high impedance states in receiver-disable or shutdown modes. The LTC1348 is available in 28-pin DIP and SSOP packages.

## **LTC1481/LTC1483—Lowest Power RS485 Transceivers Yet**

The LTC1481 and LTC1483 are the latest members of Linear Technology's growing family of CMOS RS485 interface devices. Both the LTC1481 and LTC1483 are pin compatible with the industry standard 75176 pinout. They achieve significant power savings by cutting the quiescent current when the receiver is active and by providing a shutdown mode that reduces the current consumption to below 1 $\mu$ A. The LTC1481 and LTC1483 have a maximum quiescent current of 120 $\mu$ A in receiver-active mode, more than four times lower than any other RS485 transceiver. The low power consumption of the LTC1481 and LTC1483 makes them the lowest supply current RS485 transceivers available today. Similar to the other members of Linear's 485 family, the LTC1481

and LTC1483 use a unique fabrication process and design that includes Schottky diodes in series with the MOS output transistors, allowing the outputs to maintain high impedance when forced up to  $\pm$ 7V beyond the supply rails or when the power is off.

The LTC1481 features half-duplex operation at up to 5M baud, with receiver input propagation delay of less than 200ns. The LTC1483 features the same receiver speed as the LTC1481, but its driver slew rate is deliberately slowed down to reduce EMI levels in the transmitted signal. Both the LTC1481 and LTC1483 will enter shutdown mode if the driver and receiver are disabled at the same time. The LTC1481 and LTC1483 are offered in 8-pin DIP and SOIC packages, in both commercial and industrial temperature grades.

## **Introducing the LT1169 Dual Op Amp—High Input Impedance with Low Voltage and Current Noise.**

The LT1169 is a low-noise, precision, dual-JFET operational amplifier. The low voltage noise, (6nV/ $\sqrt{\text{Hz}}$ ), is better than that of most bipolar op amps and the low bias current, (2pA), is better than that of most JFET op amps. Unlike most monolithic JFET op amps, the input bias current is essentially independent of common-mode voltage ( $\Delta I_B = 2\text{pA}$  over the input common-mode range of -10 volts to 13 volts). The device has a very high input resistance ( $10^{13}\Omega$ ) and the input capacitance is less than 2pF, assuring high gain linearity when buffering AC signals from high impedance transducers. The combination of low voltage and current noise makes the LT1169 the first choice for amplifying low-level signals from very high impedance transducers.

The LT1169 is unconditionally stable for gains of one or more, even

with 1,000pF capacitive loads. The low  $I_B$  of JFET op amps inherently results in low current noise at the expense of other parameters, but this is not the case with the LT1169. Each individual amplifier is 100% tested for voltage noise, slew rate (4.1V/microsecond), and gain bandwidth product (4.8MHz). Other key features are 0.65mV  $V_{OS}$  and a voltage gain of three million.

A full set of matching specifications is provided for precision instrumentation amplifier front ends. Specifications at  $\pm 5$  volt supply operation will also be provided.

### **LTC1159 High Efficiency Synchronous Regulator Controller**

The LTC1159 is the newest member of Linear Technology's family of above 90%-efficiency, stepdown DC-to-DC converters. The LTC1159 features an extremely wide, 4V-to-40V input operating-voltage range and reduced supply currents. The quiescent current while regulating the output is 250 $\mu$ A, and current in shutdown drops to only 20 $\mu$ A. The combination of low supply currents and high-input voltage capability is ideal for battery-powered applications that require high-voltage AC wall adapters.

A unique EXTVCC pin on the LTC1159 allows the MOSFET drivers and control circuitry to be powered from an external source, such as the output of the regulator itself. Deriving control and driver power from  $V_{OUT}$  improves efficiency at high input voltages because the resulting current drawn from  $V_{IN}$  is scaled by the duty cycle of the regulator. During start-up or short-circuit conditions, operating power is supplied by an internal 4.5V low-dropout linear regulator. This regulator automatically turns off when the EXTVCC pin is pulled above 4.5V.

The LTC1159, like other members of the LTC1148 family, automatically switches to Burst Mode™ operation at low output currents to maintain greater than 90% efficiency over two

### **LTC in the News...**

#### **Another Record Quarter for Linear Technology**

Thanks again to your outstanding support, Linear Technology posted record sales and earnings for the third quarter of fiscal 1994, which ended April 3, 1994.

Robert H. Swanson, Jr., LTC President and CEO, announced that sales increased 33% and profits increased 59% over the third quarter of fiscal 1993. Sales were a record \$51,667,000 and net income was \$15,217,000, or 40 cents per share, compared with sales of \$38,806,000 and net income of \$9,571,000, or 26 cents per share, for the third quarter of fiscal 1993.

According to Bob Swanson, "We have once again reported record financial results. The general business climate is good. Geographically our sales were particularly strong overseas, fueled largely by offshore manufacturing for U.S. companies. Our business continues to be broadly based across end markets, with some additional strength this quarter in a variety of solutions for desktop and notebook computers. Our operating income exceeded 40% of sales for the first time in our history and we generated an additional \$8.8 million in cash and short-term investments."

In its "1994 CEO of the Year" competition, *Financial World* magazine presented LTC President & CEO Bob Swanson a bronze award in the Electronics and Semiconductor category, ranking him with the CEO of Intel.

*UPSIDE* magazine included Linear Technology at 76th on its list of the best 200 technology companies in the U.S.

The *Los Angeles Times* will include Linear Technology in its 1994 "Top 100 Companies in California." The *San Francisco Chronicle* included LTC in its "Top 100 in the Bay Area." The *Chronicle* ranked LTC fourth in return on sales, 18th in return on equity, 36th in market value and 40th in growth among the biggest companies in the Bay Area. Linear Technology also figured prominently in the *San Jose Mercury* list of the "Silicon Valley 150," ranking third in return on sales, 19th in return on equity, 19th in market capitalization, 24th in profit and 73rd in sales.

*The Wall Street Transcript*, a magazine published for stock market broker/analysts, published the results of a poll in which more than 40 analysts selected LTC President Bob Swanson for a bronze medal among all CEOs in the semiconductor industry. **LT**

decades of load current range. The LTC1159 also features a constant off-time architecture and complementary power MOSFETs drivers. This combination results in a switching regulator with a dropout voltage lower than that of most linear low-dropout regulators.

To illustrate the flexibility of the LTC1159, the data sheet shows applications for high-efficiency 2.5V/5A, 3.3V/2.5A, 5V/10A, and 12V/5A regulators. It also includes a high-efficiency, dual-output application for obtaining up to 17W combined output power from 3.3V and 5V outputs.

The LTC1159 is available in fixed 5V, fixed 3.3V, and adjustable versions. Package options include 16-lead DIP, 16-lead narrow SOIC, and 20-lead SSOP. **LT**

**For further information on the above or any of the other devices mentioned in this issue of *Linear Technology*, use the reader service card or call the LTC literature service number: 1-800-4-LINEAR. Ask for the pertinent data sheets and application notes.**

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Information furnished by Technology Corporation is believed to be accurate and reliable. However, Linear Technology makes no representation that the circuits described herein will not infringe on existing patent rights.

## DESIGN TOOLS

### Applications on Disk

#### NOISE DISK

This IBM-PC (or compatible) program allows the user to calculate circuit noise using LTC op amps, determine the best LTC op amp for a low noise application, display the noise data for LTC op amps, calculate resistor noise, and calculate noise using specs for any op amp. Available at no charge.

#### SPICE MACROMODEL DISK

This IBM-PC (or compatible) high density diskette contains the library of LTC op amp SPICE macromodels. The models can be used with any version of SPICE for general analog circuit simulations. The diskette also contains working circuit examples using the models, and a demonstration copy of PSPICE™ by MicroSim. Available at no charge.

## Technical Books

**1990 Linear Databook** — This 1440 page collection of data sheets covers op amps, voltage regulators, references, comparators, filters, PWMs, data conversion and interface products (bipolar and CMOS), in both commercial and military grades. The catalog features well over 300 devices. \$10.00

**1992 Linear Databook Supplement** — This 1248 page supplement to the *1990 Linear Databook* is a collection of all products introduced since then. The catalog contains full data sheets for over 140 devices. The *1992 Linear Databook Supplement* is a companion to the *1990 Linear Databook*, which should not be discarded. \$10.00

**Linear Applications Handbook** — 928 pages full of application ideas covered in depth by 40 Application Notes and 33 Design Notes. This catalog covers a broad range of "real world" linear circuitry. In addition to detailed, systems-oriented circuits, this handbook contains broad tutorial content together with liberal use of schematics and scope photography. A special feature in this edition includes a 22 page section on SPICE macromodels. \$20.00

**1993 Linear Applications Handbook Volume II** — Continues the stream of "real world" linear circuitry initiated by the *1990 Handbook*. Similar in scope to the 1990 edition, the new book covers Application Notes 41 through 54 and Design Notes 33 through 69. Additionally, references and articles from non-LTC publications that we have found useful are also included. \$20.00

**Interface Product Handbook** — This 200 page handbook features LTC's complete line of line driver and receiver products for RS232, RS485, RS423, RS422 and AppleTalk® applications. Linear's particular expertise in this area involves low power consumption, high numbers of drivers and receivers in one package, 10kV ESD protection of RS232 devices and surface mount packages. Available at no charge.

**Monolithic Filter Handbook** — This 234 page book comes with a disk which runs on PCs. Together, the book and disk assist in the selection, design and implementation of the right switched capacitor filter circuit. The disk contains standard filter responses as well as a custom mode. The handbook contains over 20 data sheets, Design Notes and Application Notes. \$40.00

**SwitcherCAD Handbook** — This 144 page manual, including disk, guides the user through SwitcherCAD—a powerful PC software tool which aids in the design and optimization of switching regulators. The program can cut days off the design cycle by selecting topologies, calculating operating points and specifying component values and manufacturer's part numbers. \$20.00

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