LINEAR TECHNOLOGY

The LT1206: A 60MHz, 250mA Current-Feedback Amplifier

by William Jett

Introduction

The LT1206 current feedback amplifier is Linear Technology's high-speed solution for driving low-impedance loads. The part combines 60MHz bandwidth with a guaranteed 250mA output current, operation with ±5V to ±15V supplies, and optional compensation for capacitive loads, making it well suited for driving multiple cables and other difficult loads. A shutdown feature drops the supply current to less than 100µA when the part is turned off. Thermal shutdown and short-circuit protection are included in the device. The LT1206 is available in a variety of packages ranging from the surface-mount SO8 to the 7-pin TO220 power package.

Capacitive Loads

Driving capacitive loads presents two challenges to the IC designer: maintaining the stability of the amplifier loop and supplying enough current to slew the capacitor. The effect of a load capacitance can be thought of in terms of the stability of the output stage and that of the entire loop. A capacitive load can cause the output stage of an amplifier to peak or even to oscillate. The effect of this peaking on the overall response depends on the feedback components. Sometimes additional networks are used to reduce the peaking; a resistor added in series with the load capacitance will isolate the output stage, or a series RC snubber network added in parallel with the load will swamp out the load capacitance (Figure 1). The first approach causes an increase in output impedance and reduction of swing. In the second approach, power is wasted in the snubber network, reducing the current available to slew the load capacitance. The LT1206 uses an optional internal network connected to an additional pin to compensate for capacitive loads. Adding a 0.01µF bypass cap between the COMP pin and the OUTPUT pin connects the
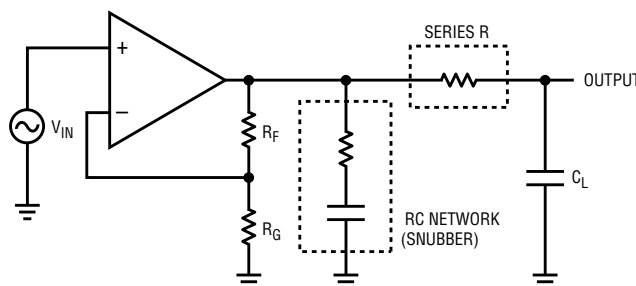
Figure 1. Conventional approaches to driving capacitive lines

Service, Customer Satisfaction, and All That

by Richard Markell

I first came to Linear Technology five years ago. In those days we had two applications engineers (not including Jim Williams, who played guru to all) and about 300 people. Since my first days at Linear Technology, the emphasis has been on quality. We don't publish circuits unless they have been breadboarded and extensively tested. The customer is our primary concern. Phone calls must be returned as soon as possible. Circuit breadboards are sometimes sent back and forth across the country several times before the customer is satisfied that the circuit works correctly.

The word got out. Customers, competitors, and even the stock market discovered LTC. Customers discovered that we mean what we say about customer support and quality; that is, we don't take it for granted.

Now we have several more applications engineers as well as many more FAEs. They all work for you, the customer. If you don't believe that, give it a try.

This issue features a bevy of new products from LTC. LTC's first power-factor corrector (PFC) circuit is high-lighted and, to make life easier for "ye olde editor," Carl Nelson answers the age-old question, "What is Power Factor, Anyway?" The LT1248 power-factor corrector can provide power factors of greater than 99.5% over a 10:1 load range—quite an achievement!

Next are two more switching regulators, which warp through the 90% efficiency barrier. The LTC1147 is an 8-pin, current-mode buck DC/DC converter with output current to 1 amp and efficiencies greater than 90% over two decades of output current. The LTC1174 is an easy-to-use step-down converter, which requires only a few external components to construct a complete high-efficiency converter. The LTC1174 requires only 130μA quiescent current with no load and a tiny 1μA in shutdown. An internal current limit is pin selectable to 340mA or 600mA. Both of the new switchers make use of Burst Mode™ operation to maximize efficiency at low current levels.

The high-speed area continues to blossom with the introduction of the LT1206, a 60MHz current-feedback amplifier with a guaranteed 250mA of output current. The LT1206 is optimized to drive multiple video cables. It features excellent differential gain and phase performance specifications. Additionally, the LT1206 draws less than 100μA when turned off.

This issue features a bumper crop of Design Ideas. A high-current synchronous switcher circuit converts 5 volts to 3.3 volts at 10 amps and another switcher converts 12–36 volts to 5 volts at 5 amps output current. The LT1228 makes a reappearance in this issue in an article on how to optimize a video gain control. The growing world of RF and wireless communications is represented with a circuit to generate protected gate bias for GaAs power amplifiers for RF transmissions. Noise generation and noise generators are here with short sections on noise generation and noise for communications channel testing (i.e., filters). Finally, there is a circuit that provides a constant 5 volt output from 3.5 to 40 volt inputs without the use of a transformer. The circuit hacking continues.

FAE Cameo: Hasti Foroutan

LTC now has nineteen Field Application Engineers (FAEs) spread throughout the world to assist customers in the design and selection of circuits available from LTC. All of our FAEs are available by phone and, in certain situations, in person, to help you design your circuitry. This space will profile one FAE per issue.

Hasti Foroutan works out of our Southwest Area office. She covers part of southern California, including Orange and San Diego Counties and parts of Los Angeles and Riverside Counties. She has BSEE and MBA degrees from the University of California at Irvine. Hasti's expertise is in the areas of switching power supply and interface design. She also has extensive experience in servo-control systems.

Hasti once designed the entire power management circuit of a palm-top computer on a flight from San Jose to Orange County. The whole circuit worked the first time it was powered up on the customer's printed circuit board. This design was powered from four AA cells and consisted of a dual output (5 volts and 3.3 volts) switching supply for the logic, a constant-current battery charger, a –12V supply for V_{PP} flash-memory programming, –24 volts for LCD contrast, plus low-battery detect and microprocessor watchdog and reset functions.

Hasti is our first female FAE and, as might be expected, a number of eyebrows are ever so slightly raised when she first visits customers. Hasti views this as a challenge to be overcome and feels that once she gains the customer's confidence, any preconceptions disappear.

Hasti enjoys traveling and is, in fact, a world traveler. She has been to many countries in Europe, North Africa, and the Middle East, and has visited exotic cities in Greece and Turkey. At home in sunny southern California Hasti enjoys tennis and horseback riding, but she keeps all of us at the factory honest by attending finance and business seminars in her spare time. Hasti can be reached through LTC's Southwest Regional Sales Office as listed on the back of this magazine. ⌐⌐

The LT1248 Power-Factor Corrector

by Carl Nelson

What is Power Factor, Anyway?

Power factor, as it relates to AC-mains usage, is basically the ratio of RMS load power to the product of line voltage and RMS current. The two are not the same because the line current or its harmonics may be phase shifted with respected to line voltage.

Figure 1 shows typical waveforms for an electronic load (a TV, stereo, computer, or the like) that full-wave rectifies the line voltage to feed a large filter capacitor. Line current flows only at the peak of the line voltage, generating large harmonic content and making the RMS current much higher than would be predicted by load power. After power-factor correction, the current waveform mimics that seen with a resistive load.

Clever engineers will note that a switching power supply driving a constant load actually looks like a negative resistance to the mains. As the mains voltage decreases, current increases to maintain constant load power. Power factor, however, is concerned with what happens in one cycle of the line. The fact that RMS currents increase with decreasing line voltage when averaged over many cycles is a separate issue. It is interesting to note, however, that this may become a thorn in the side of power companies as electronic loads become a higher percentage of total load. The use of rolling brownouts to reduce peak load will actually result in higher mains current.

The LT1248 was designed to handle power-factor correction for off-line switching supplies by generating a 380–400V DC output using a simple boost topology. This DC supply is then used as a raw input to a conventional switching supply, which isolates and steps down the voltage to provide the final load voltages. The chip can directly drive an external MOSFET at power levels up to 300W with AC input voltages varying over the full international range of 90 to 270VAC. The boost topology was chosen because it is very efficient and uses a minimum of parts, and because the inductor acts as a natural filter to isolate switching ripple from the AC line. Nevertheless, the LT1248 is versatile enough to also implement inverting and flyback topologies with very little change in the control circuitry.

One way to implement power-factor correction is to use constant switch on-time and an inductance low enough to ensure discontinuous-mode operation under all load and input-voltage conditions. If switch on-time does not vary during one cycle of the line, the average current drawn from the line will vary directly with voltage and will, therefore, look resistive. This approach has the disadvantage of very high peak currents, which lower efficiency, require a larger switch, and create more EMI. A second approach forces the inductor to operate just at the border of discontinuous mode and uses a multiplier to make peak current track the line-voltage waveform. This method reduces peak currents but has the additional disadvantage of variable frequency.

The LT1248 uses a fixed-frequency approach that measures actual line current and forces switch duty cycle to adjust itself dynamically to make line current track line voltage. This technique gives high power factor with low harmonic distortion and works independently of continuous- or discontinuous-mode operation, so that peak currents are kept low and wide variations in load can be easily accommodated. A multiplier inserted between the error amplifier and the PWM logic has line voltage as one input, forcing current to track voltage.

The circuit in Figure 2 shows the LT1248 generating a power-factor corrected 382VDC bus suitable as an input for a standard, off-line switching regulator. The line filter replaces the filter normally used in switching regulators; it can be simpler and lower in cost than the normal filter because of the filtering action of the boost inductor. A standard bridge provides full-wave rectified power to the boost input. Note that no capacitor is used at the output of the bridge. A capacitor here would create harmonic line currents. A boost architecture is created by inductor L1, diode D1, and switch Q1. The basic DC equations for a boost circuit are:

$$V_{OUT} = \frac{V_{IN}}{1-DC} \qquad I_{IN} = \frac{(V_{OUT})(I_{OUT})}{V_{IN}}$$
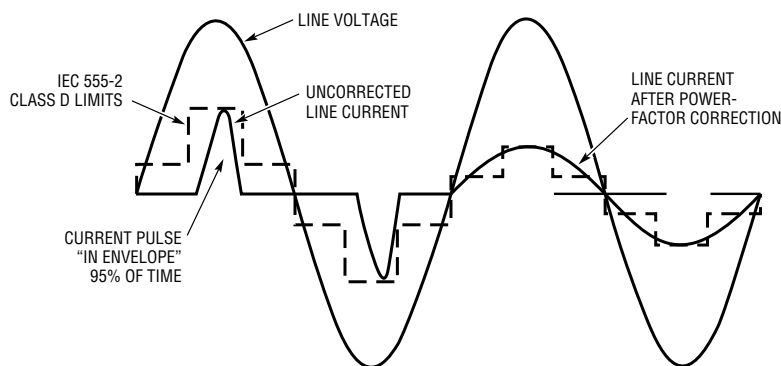$$(DC = \text{duty cycle})$$

Figure 1. Typical waveforms for an electronic load

If the loop functioned as a normal boost converter and held the output voltage constant with a constant load, input current would be highest at low input voltage and lowest at high input voltage. This is exactly what we don't want—the voltage and current should track. Remember, however, that power factor is concerned with what happens in one cycle of the line, i.e., whether current tracks voltage over a 16ms time frame. The fact that average current (over many cycles) does not track voltage is irrelevant. This special boost converter cannot violate the basic equations shown on page three for averaged conditions, but it can force current to track voltage during one line cycle. It does this by making the response time of the error amplifier very slow. Over many line cycles, the error amplifier can correct the switch drive to keep output voltage constant, but within one

cycle the multiplier forces line current to follow line voltage.

One consequence of power-factor correction is that the current delivered to the output capacitor is equivalent to full-wave rectified 60Hz AC, and therefore the capacitor has a healthy amount of 60Hz ripple on it. This is the direct result of power-factor correction and is independent of the topology used to achieve correction. From a practical standpoint, this is not usually a problem, but it does have one important implication: It is theoretically impossible to combine the power-factor function with the switching-regulator function to achieve a power-factor corrected output that does not contain large amounts of 60Hz ripple.

Loop Description

The basic purpose of the loop is to sense line current and force it to track

line voltage, while maintaining a constant DC output voltage. First I will describe the circuit that forces line current to follow line voltage. A current proportional to line voltage is generated with R5. This current is lightly filtered with C4 to remove any switching ripple and then fed into a multiplier. The output of the multiplier (I_M) is the product of this current and a second current proportional to error-amplifier output squared. (The squaring of error amplifier output does not affect power-factor correction per se; it is done for loop-stability considerations—more on this topic later.) Over one line cycle, the error amplifier output is assumed to be steady, so I_M is a full-wave rectified 60Hz sine wave with an amplitude dependent on error-amplifier output. This rectified sine wave current is fed through R2 to develop a voltage that becomes the reference to which line
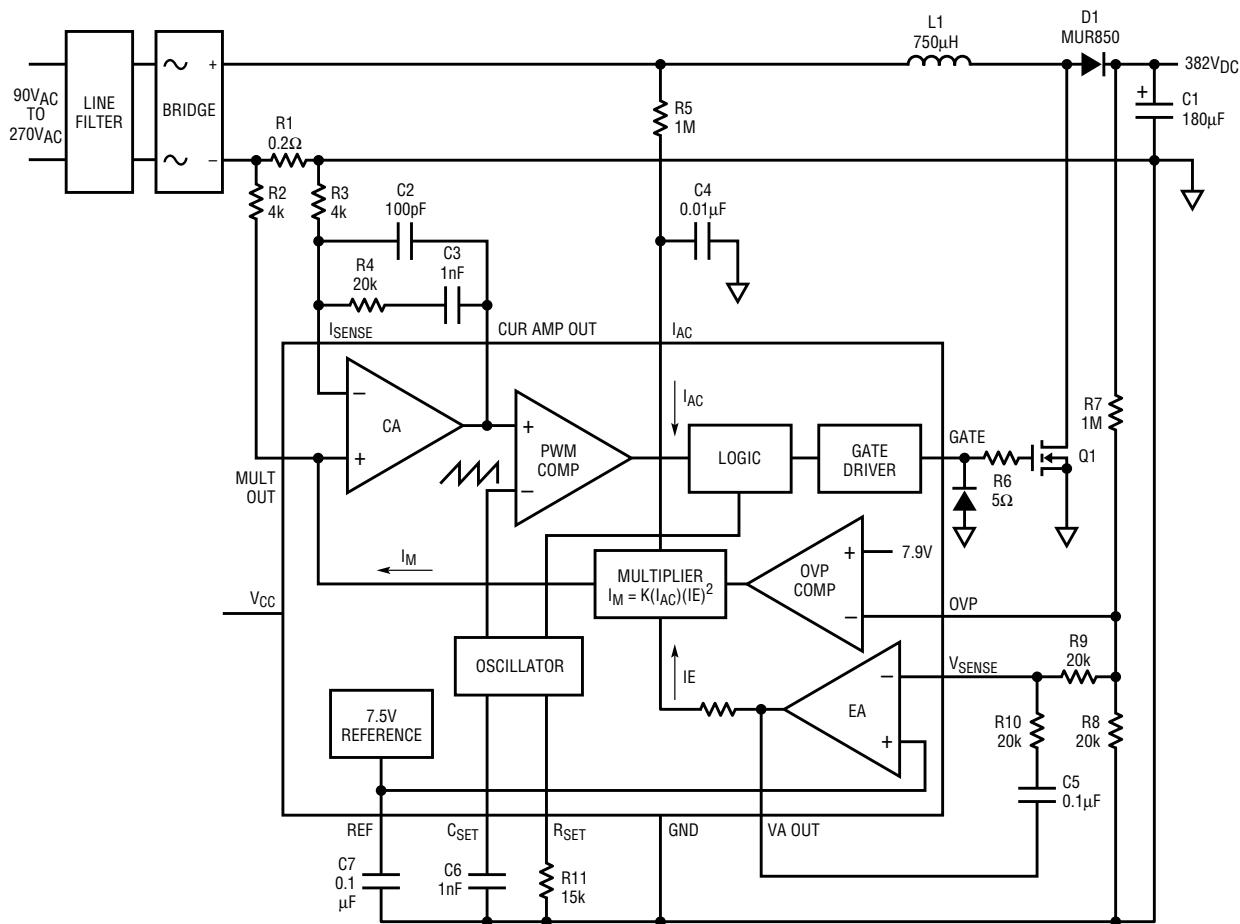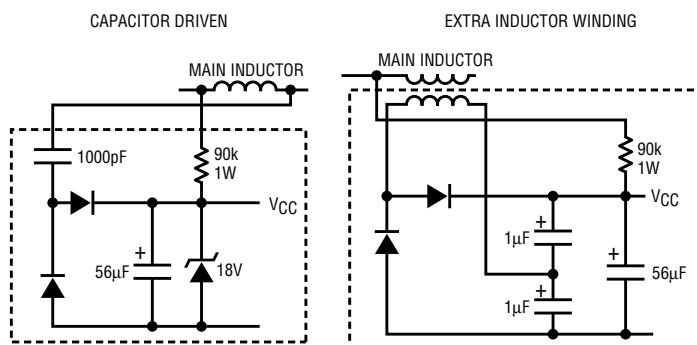

Figure 2. 25W–300W power-factor corrected supply schematic

CAPACITOR DRIVEN   EXTRA INDUCTOR WINDING

Figure 3. Generating the LT1248's supply voltage

current is compared. Notice that the current amplifier CA has its inputs connected to this reference voltage and to a second resistor (R3), which is connected to the opposite side of R1. If CA operates as a true op amp, it must somehow force the voltage across R1 to be equal to the reference voltage across R2. There is no voltage drop across R3 other than amplifier bias current. C2, C3, and R4 are included to filter switching ripple from CA's output.

How does the current amplifier force line current? Like any standard PWM loop, comparator C1 compares a triangle-wave signal from the oscillator to the CA output to generate a square wave whose duty cycle is proportional to CA output. This signal flows through some logic to drive a power MOSFET. Line current will be a function of switch duty cycle, so the loop is closed and CA can force the voltage across R1 to follow the voltage across R2. This action must have a bandwidth many times greater than 60Hz to avoid distortion, but much less than the typical 100kHz switching frequency to reject switching ripple.

The second part of the loop requirement is to maintain a constant DC output voltage. Voltage is sensed with the R7/R8 divider and compared to a 7.5 volt reference by the error amplifier. Local AC feedback around the error amplifier, consisting of R10 and C5, gives the amplifier very slow response, but over several cycles of the line it can adjust the magnitude of the line current reference I_M to whatever value is needed to maintain the output voltage at

382VDC. The fact that it takes several cycles of the line to correct for any load or line variations is important and necessary. If the amplifier could respond significantly in one line cycle, it would create distortion in the I_M reference current with resulting line-current distortion.

The output of the multiplier is made proportional to the square of the error amplifier output to minimize variations in voltage control-loop gain. The worst case for loop stability tends to be at high line voltage with light loads. The error amplifier output is low in this condition and the squaring term produces lower loop gain with improved response.

Chip Start-Up and Supply Voltage

During normal operation, power for the LT1248 can be drawn from an auxiliary winding on the power inductor or transformer. Start-up, however, is a problem, because the auxiliary power is not available until switching starts. The solution is to make the IC have a low supply current until its input voltage reaches some threshold—say, 16V. Then the chip turns on to normal operation and draws normal operating current. A large supply bypass capacitor holds the supply voltage up long enough to allow power from the auxiliary winding to develop. The trickle current needed to charge the capacitor is taken from a large resistor connected to the rectified mains. This is often referred to as "burp" starting.

In older designs, burp starting required several milliamps of trickle

current. This was annoying because the resistor had to be about 40kΩ to ensure start-up at low line voltage, but at high line voltage, the resistor had about 250V across it and dissipated almost 2W. The LT1248 has a start-up current of approximately 250µA, so much lower trickle currents can be used. Supply efficiency is improved and heat buildup is significantly reduced.

Auxiliary windings on boost-converter inductors do not generate constant output voltages like those on flyback transformers. The voltage across the inductor is equal to the input voltage in one polarity and is equal to output voltage minus input voltage in the opposite polarity. Both of these signals are unregulated and vary widely. The solution is to peak detect both waveforms and sum them. The result is a voltage proportional to the regulated output voltage. Figure 3 shows how this is done. The 90k resistor supplies trickle start-up current.

Also shown in Figure 3 is a new technique for IC supply voltage that eliminates the need for an extra inductor winding. It uses capacitor charge transfer to generate a constant current source, which feeds a zener diode. Current to the zener is equal to $(V_{OUT} - V_Z)(C)(f)$, where V_Z is the zener voltage and f is the switching frequency. For $V_{OUT} = 382V$, $V_Z = 18V$, $C = 1000pF$, and $f = 100kHz$, zener current will be 36mA. This is enough to operate the IC, including the extra current drawn by gate-charge drive to the FET.

Protection Features

At Linear Technology, we pride ourselves on making "bustproof" ICs, so we have included several features on the LT1248 to handle potentially destructive conditions, regardless of their origins. There is a peak-current-limiting comparator that overrides the current regulating loop if, for some reason, the loop attempts to increase switch currents to destructive levels. This is a backup to the main current limiting function

LTC1174: A High-Efficiency Buck Converter

by San-Hwa Chee
and Randy Flatness

Introduction

Not long ago, linear regulators were the choice for low-power applications and applications where heat dissipation was not a problem. Recently, with the trend towards portability, users started to take a hard look at switching regulators for these applications. The main advantage of switching regulators over linear regulators is higher efficiency, which translates to longer battery life.

The LTC1174 is a 8-pin SOIC, "user-friendly" step-down converter. (A DIP package is also available.) Only four external components are needed to construct a complete, high-efficiency converter. With no load, it requires only 130µA of quiescent current; this decreases to a mere 1µA upon shutdown. The LTC1174 is protected against output shorts by an internal current limit, which is pin selectable to either 340mA or 600mA. This current limit also sets the inductor's peak current. This allows the user to optimize the converter's efficiency depending upon the output current requirement.

In dropout conditions, the internal 0.9Ω (at a supply voltage of 9V) power P-channel MOSFET switch is turned on continuously (DC), thereby maximizing the life of the battery source.

(Who says a switcher has to switch?) In addition to the features already mentioned, the LTC1174 boasts a low battery detector. Moreover, the LTC1174 comes in three versions: the LTC1174-5 (5V output), the LTC1174-3.3 (3.3V output), and the LTC1174 (adjustable). All versions function down to an input voltage of 4V and work up to an absolute maximum of 13.5V.

Efficiency

Figure 1 shows a practical LTC1174-5 circuit with a minimum of components. Efficiency curves for this circuit at two different input voltages are shown in Figure 2. Note that the efficiency is 94% at a supply voltage of 6V and load current of 175mA. This makes the LTC1174 attractive to all power sensitive applications and shows clearly why switching regulators are gaining dominance over linear regulators in battery-powered devices.

If higher output currents are desired, pin 7 (I_{PGM}) can be connected to V_{IN}. Under this condition, the maximum load current is increased to 450mA. The resulting circuit and efficiency curves are shown in Figures 3 and 4, respectively.
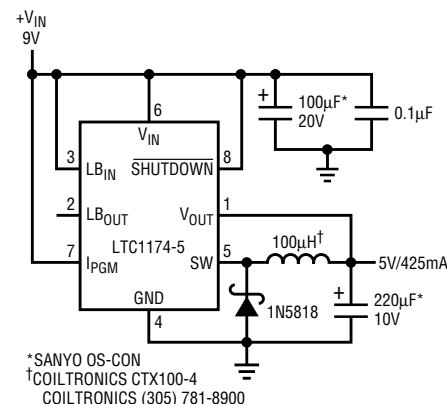
Figure 3. Typical application for higher output currents

What is Burst Mode™ Operation?

To maximize efficiency, the LTC1174 was designed to go into sleep mode once the output voltage reaches regulation. When this occurs, a majority of the internal circuitry is turned off, reducing the quiescent current from 0.45mA to 130mA. Under these conditions, all the load current is supplied by the output capacitor. When the output voltage drops by an amount equivalent to the hysteresis of the comparator, the LTC1174 wakes up and starts switching again until the output reaches regulation. The process then
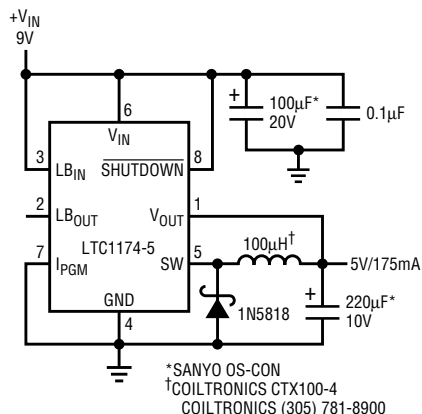
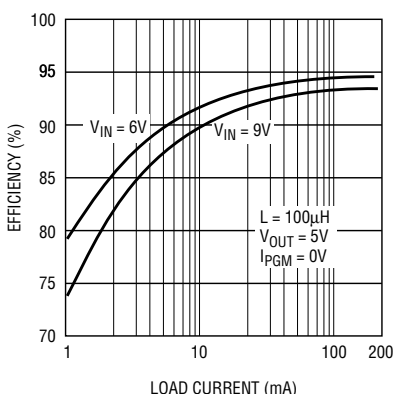Figure 1. Typical application for low output currents

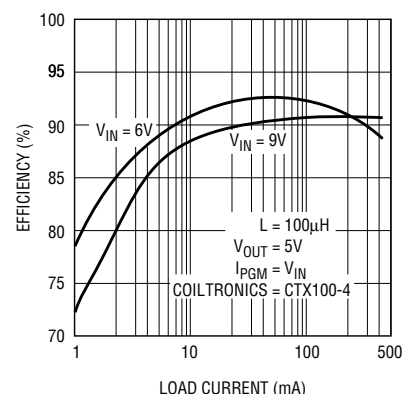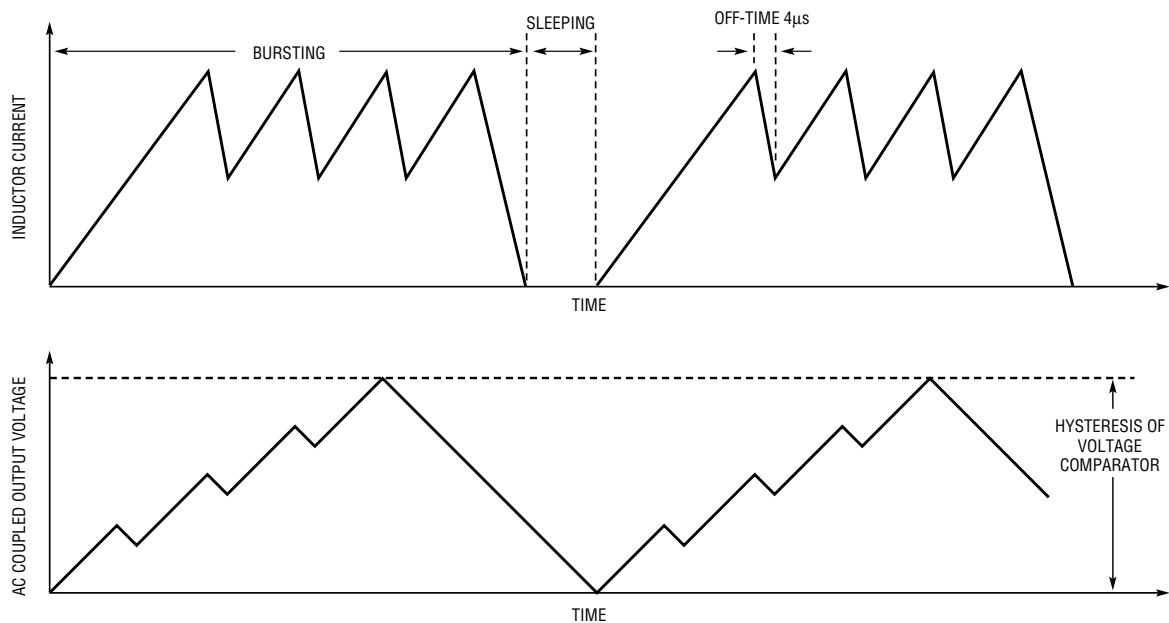Figure 2. Efficiency vs. load current

Figure 4. Efficiency vs. load current

Burst Mode™ is a trademark of Linear Technology Corporation

Figure 5. LTC1174 inductor current and output voltage at medium load

repeats. The inductor's current and output voltage waveforms (Figure 5) reveal how the LTC1174 utilizes Burst Mode™ operation.

As the load current increases, the time spent in sleep decreases. Eventually the LTC1174 goes into continuous-mode operation, never going to sleep. Any increase in load current past this point will cause the output to drop out of regulation.

Constant Off-Time Architecture

Another technical point to note is that the LTC1174 uses a constant off-time architecture. This off-time is set to 4ms when the output is in regulation; otherwise it is inversely proportional to the output voltage. In the extreme condition, the P-channel is turned on at a 100% duty cycle to make the LTC1174 a low-dropout regulator. The advantage of this scheme is that the inductor's ripple current is predictable and well controlled, making the selection of inductor much easier. The inductor's peak-to-peak ripple current is inversely proportional to the inductance. With a 50mH inductor and a 5V output voltage, the ripple current is 440mA. If a lower ripple current is desired, a larger inductor can be used. The inductance value is governed by the current requirement of the load.

The minimum value of L can be determined by the following:

$$L \ (\text{Henrys}) = \frac{4 \times 10^{-6}(V_{OUT} + V_D)}{2(I_{PEAK} - I_{OUT})}$$
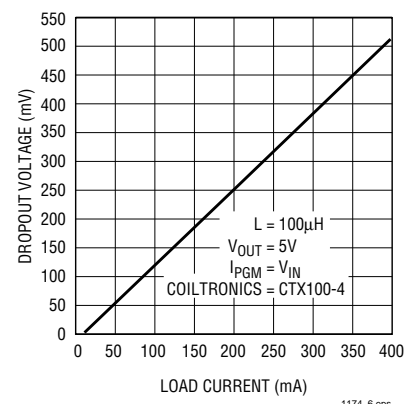
where I_{PEAK} = 340mA or 600mA, depending on the condition of I_{PGM} and V_D = diode forward drop. As the inductance is increased from this minimum value, the ripple current decreases. This also increases the maximum load current, since the inductor's peak current is fixed.

Another consequence of constant off-time architecture is that the switching frequency is related to the input voltage. For an input voltage range of 6V to 12V, with an output voltage of 5V, the operating frequency varies from about 42kHz to 146kHz. For example, with an input voltage of 9V, the switching frequency is approximately 110kHz. At this high operating frequency, a small inductor value can be used.
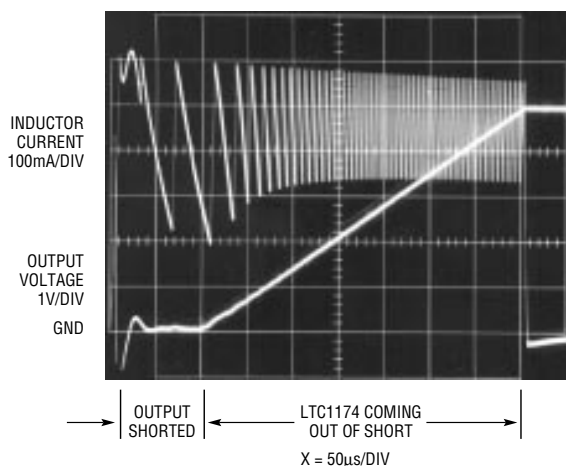
100% Duty Cycle in Dropout Conditions

When the input voltage decreases, the switching frequency decreases. With the off-time constant, the on-time is increased to maintain the same peak-to-peak ripple current in the inductor. Ultimately, a steady state
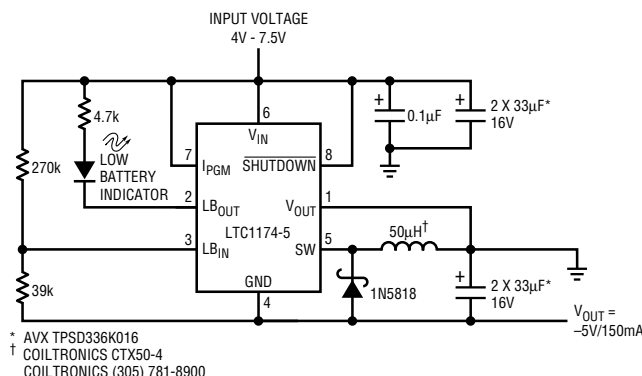
condition will be reached where Kirchoff's Voltage Law determines the dropout voltage. When this happens, the P-channel power MOSFET is turned on at DC (100% duty cycle). The dropout voltage is then governed by the load current multiplied by the total DC resistance of the MOSFET, the inductor, and the internal 0.1Ω current-sense resistance. Figure 6 shows the dropout voltage as a function of output current. Note that for a load current of 400mA, the dropout voltage is only 0.51V. This is comparable to what linear regulators can offer. Unlike linear regulators, where ground current varies with load current, the LTC1174's ground current is a constant 450µA.

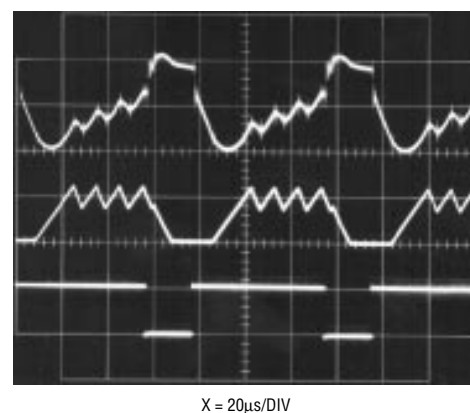Figure 6. Dropout voltage vs. output current

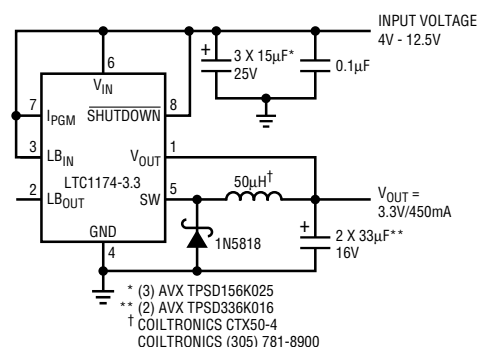Figure 7. Short-circuit and start-up response of the LTC1174

Figure 8. Load transient response

Figure 9. Positive to –5V converter with low-battery detection

Figure 10. 5V to 3.3V output application

Good Start-Up and Transient Behavior

The LTC1174 exhibits excellent start-up behavior when it is initially powered-up or recovering from a short-circuit. This is achieved by making the off-time inversely proportional to the output voltage when the output is still in the process of reaching its regulated value. When the output is shorted to ground, the off-time is extended long enough to prevent any possibility of a build up in the inductor's current. When the short is removed, the output capacitor begins to charge and the off-time gradually decreases. Note the absence of overshoot when the LTC1174 comes out of a short-circuit, as shown in Figure 7. When the output reaches its regulated value, the off-time is fixed at 4μs. The initial power-up waveform is similar to Figure 7.

In addition, the LTC1174 has excellent load-transient response. When the load current drops suddenly, the feedback loop responds quickly by turning off the internal P-channel switch. Sudden increases in output current will be met initially by the output capacitor, causing the output voltage to drop slightly. As mentioned above, tight control of inductor's current means that output-voltage overshoot is virtually eliminated (see Figure 8).
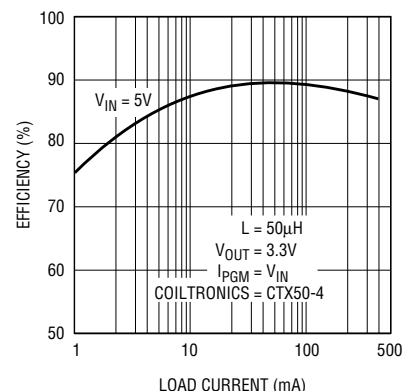
Typical Applications

Positive-to-Negative Converter

The LTC1174 can easily be set up for a negative output voltage. The LTC1174-5 is ideal for –5V outputs, as this configuration requires the fewest components. Figure 9 shows the schematic for this application with low-battery detection capability. The LED will turn on at input voltages below 4.9V. The efficiency of this circuit is 81% at an input voltage of 5V and output current of 150mA.

A 5V to 3.3V Converter

The LTC1174-3.3 is ideal for applications that require 3.3V at less than 450mA. A minimum board area, surface mount 3.3V regulator is shown in Figure 10. Figure 11 shows that this circuit can achieve efficiency greater than 85% for load currents between 5mA and 450mA. ⎇⎇

Figure 11. Efficiency vs. load current

The New SO8 LTC1147 Switching-Regulator Controller Offers High Efficiency in a Small Footprint

by Randy Flatness

Introduction

The LTC1147 switching-regulator controller is the latest addition to Linear Technology's high efficiency, step-down DC/DC converter family. This 8-pin controller uses the same current-mode architecture and Burst Mode™ operation as the LTC1148/LTC1149 but without the synchronous switch. Ideal for applications requiring up to 1 amp, the LTC1147 shows 90% efficiencies over two decades of output current.

The LTC1147, like the other members of the LTC1148/LTC1149 family, automatically changes from current-mode operation at high output currents to Burst Mode™ operation at low output currents. The wide operating range is illustrated by the typical efficiency curve of Figure 1. All members of the LTC1147/LTC1148/LTC1149 family are capable of operation at 100% duty cycle, providing very low dropout operation, and all have built-in current limiting. Load and line regulation are excellent for a wide variety of conditions, including the transition from Burst Mode™ operation to continuous mode operation.
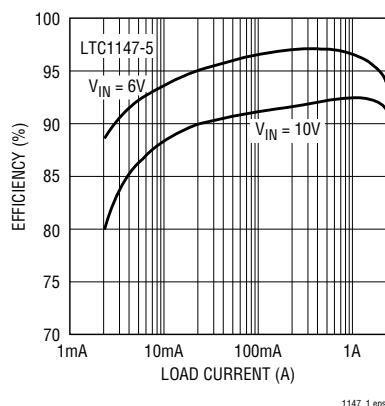

Figure 2. This LTC1147 5V to 3.3V converter achieves 92% efficiency at 300mA load current

Figure 1. Greater than 90% efficiency is obtained for load currents of 20mA to 2A (V_{IN} = 10V)

High Efficiency in a Small Area

The 8-pin SOIC package and a few external components make high efficiency DC-to-DC conversion feasible in the extremely small board space available in today's portable electronics. (A DIP package is also available.) An ideal application for the LTC1147 is dropping 5V to 3.3V locally on a PC board. If a linear regulator with a 1 amp output current is used, the power dissipation will exceed 1.7 Watts. This is unacceptable when there is no way to remove the heat from an enclosed space. The LTC1147 5V to 3.3V converter shown in Figure 2 has 85% efficiency at 1A output, with efficiencies greater than 90% for load currents up to 500mA. Using the LTC1147 reduces the power dissipation to less than 500mW. The efficiency plotted as a function of output current is shown in Figure 3.

Giving Up the Synchronous Switch

The decision whether to use a non-synchronous LTC1147 design or a fully synchronous LTC1148 design requires a careful analysis of where losses occur. The LTC1147 switching-regulator controller uses the same
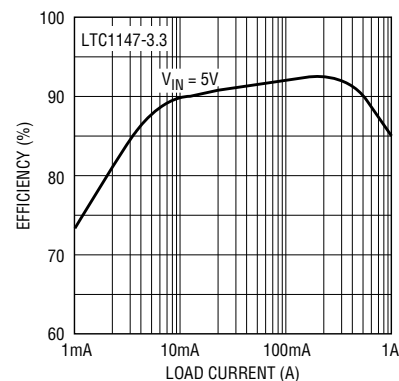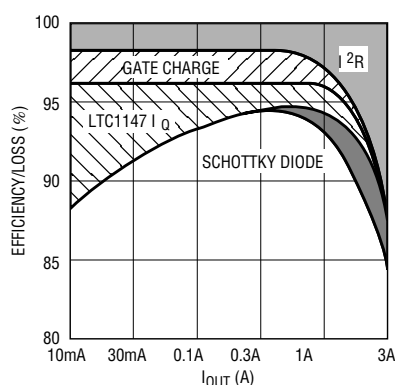

Figure 3. The LTC1147 5V to 3.3V converter provides better than 90% efficiency from 20mA to 500mA of output current

Figure 4. Low current efficiency is enhanced by Burst Mode™ operation. Schottky diode loss dominates at high output currents

loss-reducing techniques as the other members of the LTC1148/LTC1149 family. The non-synchronous design saves the N-channel MOSFET gate-drive current at the expense of increased loss due to the Schottky diode.

Figure 4 shows how the losses in a typical LTC1147 application are apportioned. The gate-charge loss (P-channel MOSFET) is responsible for the majority of the efficiency lost in the mid-current region. If Burst Mode™ operation was not employed, the gate-charge loss alone would cause the efficiency to drop to unacceptable levels at low output currents. With Burst Mode™ operation, the DC supply current represents the only loss component that increases almost linearly as output current is reduced. As expected, the I^2R loss and Schottky-diode loss dominate at high load currents.

In addition to board space, output current and input voltage are the two primary variables to consider when deciding whether to use the LTC1147. At low input-to-output voltage ratios, the top P-channel switch is on most of the time, leaving the Schottky diode conducting only a small percentage of the total period. Hence, the power lost in the Schottky diode is small at low output currents. This is the ideal application for the LTC1147. As the output current increases, the diode loss increases. At high input-to-output voltage ratios,

the Schottky diode conducts most of the time. In this situation, any loss in the diode will have a more significant effect on efficiency and an LTC1148 might therefore be chosen.

Figure 5 compares the efficiencies of LTC1147-5 and LTC1148-5 circuits with the same inductor, timing capacitor, and P-channel MOSFET. At low input voltages and 1A output current, the efficiency of the LTC1147 differs from that of the 1148 by less than two percent. At lower output currents and high input voltages, the LTC1147's efficiency can actually exceed that of the LTC1148.

Ideal for Low Dropout Applications

Because the LTC1147 is so well suited for low input-to-output voltage ratio applications, it is an ideal choice for low dropout designs. All members of the LTC1148/LTC1149 family (including the LTC1147) have outstandingly low dropout performance. As the input voltage on the LTC1147 drops, the feedback loop extends the on-time for the P-channel switch (off-time is constant) thereby keeping the inductor ripple current constant. Eventually the on-time extends so far that the P-channel
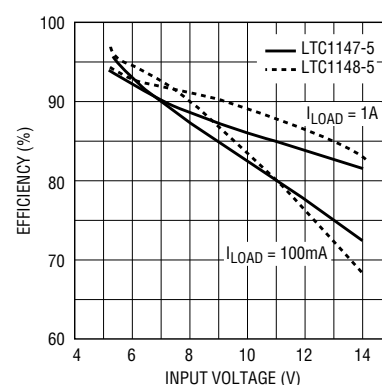

Figure 5. At high input voltages combined with low output currents, the efficiency of the LTC1147 exceeds that of the LTC1148

MOSFET is on at DC or at a 100% duty cycle.

With the switch turned on at a 100% duty cycle, the dropout is limited by the load current multiplied by the sum of the resistances of the MOSFET, the current shunt, and the inductor. For example, the low dropout 5V regulator shown in Figure 6 has a total resistance of less than 200mΩ. This gives it a dropout voltage of 200mV at 1A output current. At input voltages below dropout the output voltage follows the input. This is the circuit whose efficiency is plotted in Figure 1.
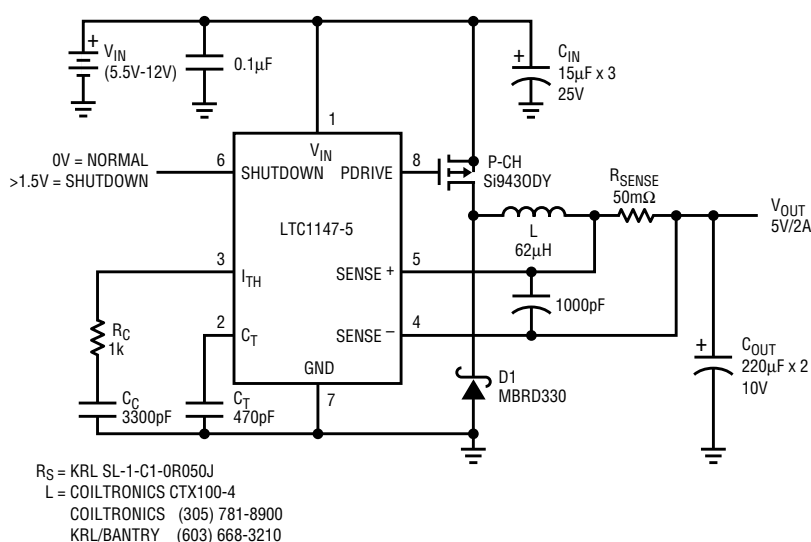
Figure 6. The LTC1147 architecture provides inherent low dropout operation. This LTC1147-5 circuit supports a 1A load with the input voltage only 200mV above the output.
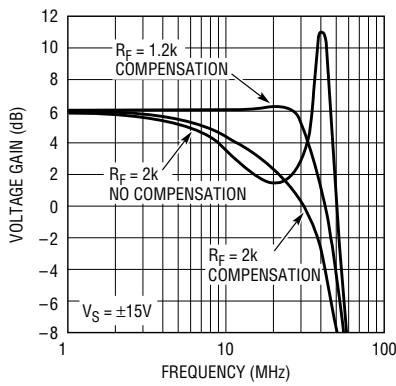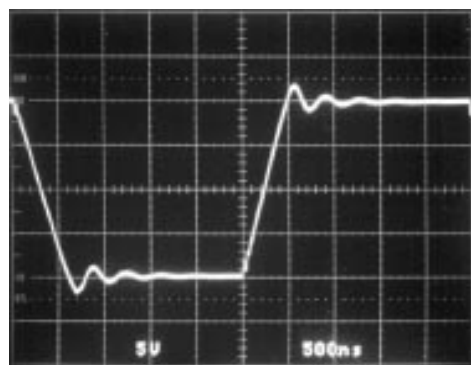
Figure 2. Frequency response, A_V = +2, C_L = 200pF

Figure 4. LT1206 distribution amplifier

LT1206, continued from page 1

network. This network has the effect of smoothing the output peaking so that, with the correct feedback resistor, the overall response is flat. Figure 2 shows the effect of the optional compensation network on frequency response with C_L = 200pF. Using a 1.2kΩ feedback resistor and the compensation, the overall response is flat within 0.35dB to 30MHz.
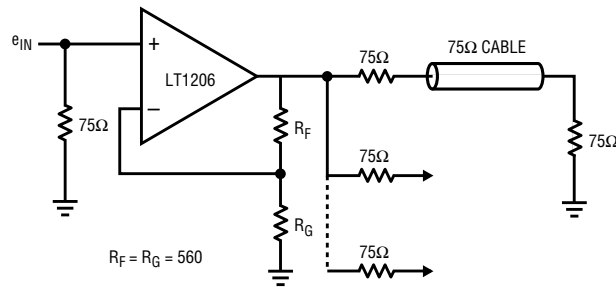
Although the optional compensation works well with capacitive loads, it simply reduces the bandwidth when it is connected with resistive loads. For instance, with a 30Ω load, the bandwidth drops from 55MHz to 35MHz when the compensation is connected. Hence, the compensation was made optional.

Although the small signal response with a capacitive load depends on the feedback components and the optional compensation network, the large signal response is limited by the maximum output current. In the fastest configuration, the LT1206 is capable of a slew rate of 1V/ns. The current required to slew a capacitor at this rate is 1mA per picofarad of capacitance, so a 10,000pF cap would require 10A. The large signal behavior with C_L = 10,000pF is shown in Figure 3. The slew rate is about 40V/μs, determined by the current limit of 400mA.

Buffers and Cable Drivers

The combination of a 60MHz bandwidth, 250mA output-current capability, and low output impedance make the LT1206 ideal for driving multiple video cables. One concern when driving multiple transmission lines is the effect of an unterminated (open) line on the other outputs. Since the unterminated line creates a reflected wave that is incident on the output of the driver, a non-zero amplifier output impedance will result in crosstalk to the other lines. Figure 4 shows the LT1206 connected as a distribution amplifier. Each line is separately terminated to minimize the effect of reflections. For systems using composite video, the differential gain and phase performance are also important and have been considered in the internal design of the device. The differential phase and differential gain performance versus supply is shown in Figures 5 and 6 for 1, 3, 5, and 10 cables. Figure 7 shows the output impedance versus frequency. Note that at 5MHz the output impedance is only 0.6Ω.

Although the wide bandwidth and high output drive capabilities of the LT1206 make it a natural for video circuits, these characteristics are also useful for audio applications. Figure 8 shows the LT1206 combined with the LT1115 low-noise amplifier to form a very low-noise, low-distortion audio buffer with a gain of 10. With a 32Ω load and a 5Vrms output level (780mW), the THD+noise for the circuit is 0.0009% at 1kHz, rising to 0.004% at 20kHz. The frequency response is flat to 0.1dB from DC to 600kHz, with a –3dB bandwidth of 4MHz. The circuit is stable with capacitive loads of 250pF or less.

A_V = +2, R_F = R_G = 3k, V_S = ±15V, C_L = 0.01μF

Figure 3. LT1206 large signal response
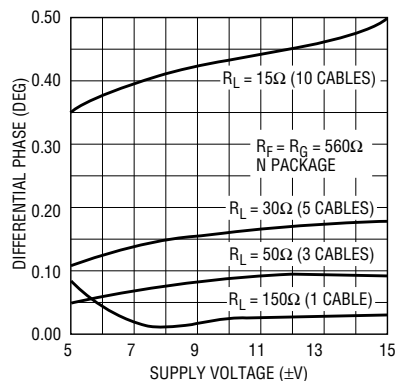
Figure 5. Differential phase vs. supply voltage
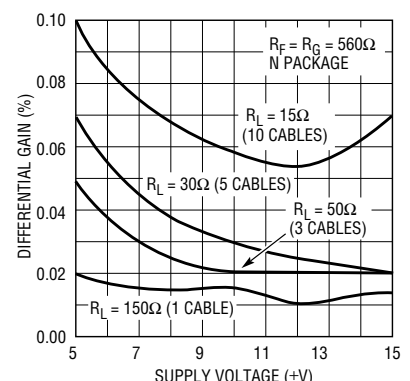
Figure 6. Differential gain vs. supply voltage
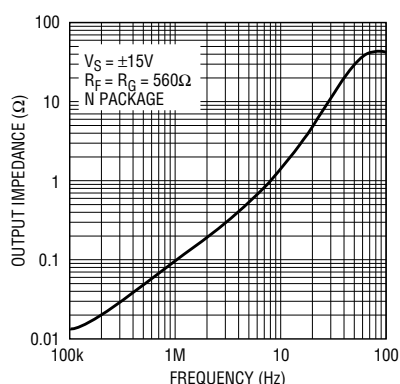
Figure 7. Output impedance vs. frequency

Circuit Description

The LT1206 uses a current-feedback topology and LTC's complementary bipolar process to obtain excellent speed and power gain with simple circuitry. Since both the NPNs and PNPs are fast devices, both are used in the signal path. Figure 9 shows a simplified schematic. The input stage consists of devices Q1–Q4. Devices Q5–Q8 form very fast current mirrors. The output stage consists of complementary emitter followers Q9 and Q12, followed by complementary Darlington followers Q10–Q11 and Q13–Q14. Diodes D1 and D2 act as level shifts, providing the proper AB bias levels for the output devices Q11 and Q14. The optional compensation network for capacitive loads consists of C_C and R_C. When the COMP pin is left open, devices Q15 and Q16 act as a bootstrap, preventing the compensation network from loading the output of the current mirrors. When used, a bypass capacitor (0.01µF) AC shorts the COMP pin to the OUTPUT pin. A DC short cannot be used because of the large currents that would flow in the COMP pin during an output short circuit.

The shutdown pin provides a dual function; first, it can be used to turn off the biasing for the amplifier, reducing the quiescent current to less than 100µA, and second, it can be used to control the quiescent current in normal operation. All of the biasing for the LT1206 is derived from the collector current of Q18, which results in the supply current being proportional to the shutdown pin current. With the shutdown pin grounded, Q17 limits Q18 current to about 500µA, resulting in a supply current of approximately 20mA. Supply current can be reduced by putting a resistor from the shutdown pin to ground. The voltage across the resistor is then V_S–3Vbe. The supply current is approximately 40 times the current that flows in the shutdown pin. For example, to set the supply current to 10mA on ±5V supplies, use a 12kΩ resistor.
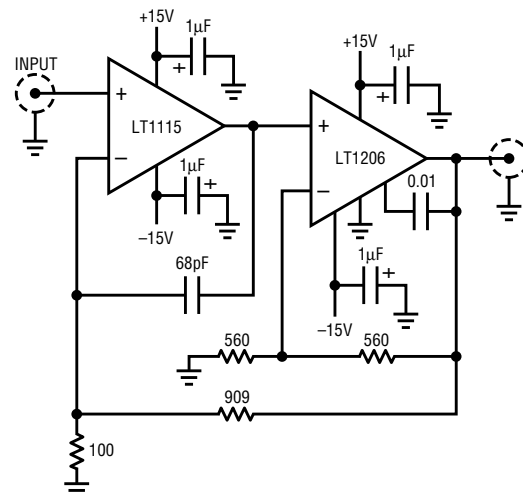
Figure 8. Low noise × 10 buffered line driver

Performance

Table 1 summarizes the major performance specifications of the LT1206 on ±5V and ±15V supplies.

Conclusions

The LT1206 combines high output current with wide bandwidth to form an effective solution for driving low-impedance loads. Stability problems with capacitive loading have been solved by a novel compensation scheme. Package options range from the surface mount SO8 to the 7-pin TO220. *LƮ*
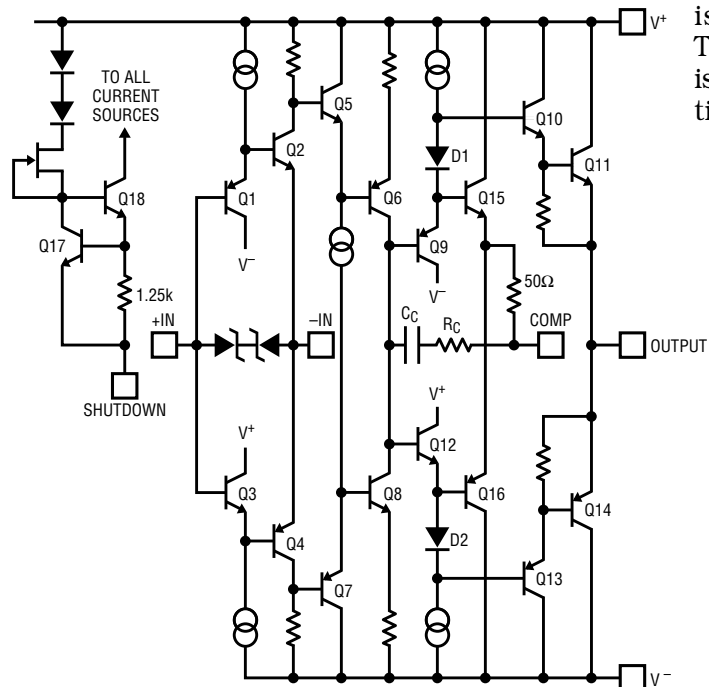
Figure 9. Simplified schematic

Table 1. LT1206 Performance

Parameter	Conditions	$V_S = \pm 5V$	$V_S = \pm 15V$
Bandwidth	$A_V = +2$, $R_L = 50$	45MHz	60MHz
Bandwidth	$A_V = +2$, $R_L = 30$	40MHz	55MHz
Bandwidth	$A_V = +2$, $R_L = 10$	30MHz	38MHz
Slew rate	$A_V = -1$, $R_L = 100$	500V/µs	1000V/µs
Slew rate	$A_V = -1$, $R_L = 50$	450V/µs	700V/µs
Slew rate	$A_V = +2$, $R_L = 50$	300V/µs	600V/µs
Minimum output current		250mA	250mA
Maximum input offset voltage		10mV	10mV
Maximum inverting input current		60µA	60µA
Nominal supply current		18mA	20mA

provided by a current clamp on the I_M output.

To guard against output over voltage conditions, an over-voltage-comparator monitors DC output voltage and immediately stops switching action if the output voltage rises more than 10% above its intended value. Over-voltage is a real possibility with power-factor designs because of the slow error amplifier response time. Remember that power delivered to the output increases as the square of input voltage for frequencies above the bandpass of the control loop (20Hz). A sudden load shedding coincident with increased line voltage can create a situation where the error amplifier cannot reduce current quickly enough to prevent overshoot at the output. The over-voltage comparator provides an extra level of protection against such unforeseen conditions.

An internal gate clamp limits gate drive if the supply voltage climbs above 16V, ensuring reliable MOSFET operation during extended periods of high chip supply voltage.

Additional Features

The enable/sync pin does double duty as an on/off input and as a synchronizing port. The threshold for shutdown is 2.5V, and that for sync is 5.5V, so the two functions can be made non-interactive. The LT1248 is sometimes synchronized to the main switcher or to a system clock to prevent beat frequencies or to eliminate specific time-slot noise spikes.

The SS pin acts as clamp on the reference input to the error amplifier during startup. By connecting a capacitor to SS, the output can be programmed to ramp up in a controlled fashion.

Performance

Figure 4 shows the performance of the LT1248 both in terms of power factor and harmonic distortion. Power factor remains above 99.5% over a ten-to-one load-current range. This allows the supply to be used in "green" applications, where it will be used in both full-power and low-power modes. Harmonic distortion also remains low over a wide load range, even when measured past the 20th harmonic. Agency guidelines may actually be more concerned with harmonic content than power factor, so a well-designed power-factor corrector must perform well in both areas.

The LT1248 fits in 16-pin, narrowbody, dual-in-line and surface mount packages. It uses a high-frequency IC process to achieve fast switching with only 8mA supply current. Switching frequency can extend up to 300kHz. A flexible topology, low external parts count, and the ability to handle both low- and high-power applications make this new IC attractive for a wide variety of applications. An 8-pin version with frequency set internally at 100kHz (LT1249) will also be available. *LT*
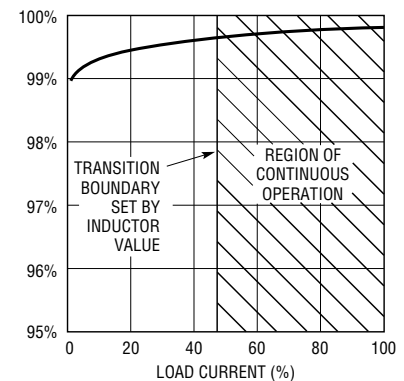

Figure 4. Power factor vs. load current

High Efficiency Step-Down Regulator Family

Table 1 shows the members of the LTC1147/LTC1148/LTC1149 family and several of their applications. The LTC1147 is available in both fixed 3.3V and fixed 5V versions and is available in both 8-pin DIP and 8-pin SOIC surface mount packages. For other output voltages, the LTC1148 adjustable is suggested. (If the synchronous switch is not needed in a low current application, the N-channel MOSFET can be eliminated and a Schottky catch diode can be used.)

Conclusion

The LTC1147 adds even more versatility to Linear Technology's family of high efficiency step-down regulator controllers. Optimized for low current applications, the LTC1147 saves board space and cost over synchronous designs with only a small reduction in efficiency. The high performance of this controller is ideal for extending the battery life of the newest portable electronics. *LT*

Table 1. LTC1147/LTC1148/LTC1149 family applications

	LTC1147-3.3	LTC1147-5	LTC1148-3.3	LTC1148-5	LTC1148	LTC1149-3.3	LTC1149-5	LTC1149
Continuous input voltage ≤ 48V						✔	✔	✔
Continuous input voltage ≤ 13.5V	✔	✔	✔	✔	✔			
Low dropout 5V		✔		✔			✔	
Adjustable/ multiple output					✔			
5V to 3.3V	✔			✔				
Minimum board area	✔	✔						

Ultra-Low-Power CMOS RS232 Transceiver Achieves 10kV ESD Protection, Eliminates Latch-Up

by Ricky Chow
and
Robert Reay

The LTC1337 is an ultra-low-power RS232 transceiver with supply current up to 20 times lower than other CMOS devices. Using proven technology borrowed from LTC's low-power RS485 products, the LTC1337 achieves ultra-low power consumption without sacrificing ESD protection and latch-up immunity.

The single 5 volt, three-driver/five-receiver RS232 transceiver draws only 300µA of supply current in the no-load condition and 1µA in the shutdown mode. The charge pump requires only four 0.1µF capacitors and can supply up to 12mA of extra current to power external circuitry. The transceiver can operate up to 120k baud with a 1000pF cap in parallel with a 3kΩ load, and both driver outputs and receiver inputs can be forced to ±25V and withstand multiple 10kV ESD strikes.

Latch-Up Eliminated

The main sources of latch-up in CMOS RS232 transceivers have been eliminated by incorporating circuit techniques borrowed from the LTC485 family. Latch-up can occur in conventional CMOS output stages when the output of the driver is forced beyond the supply rails. If the output of the traditional CMOS output stage of Figure 1 is driven above V+ or
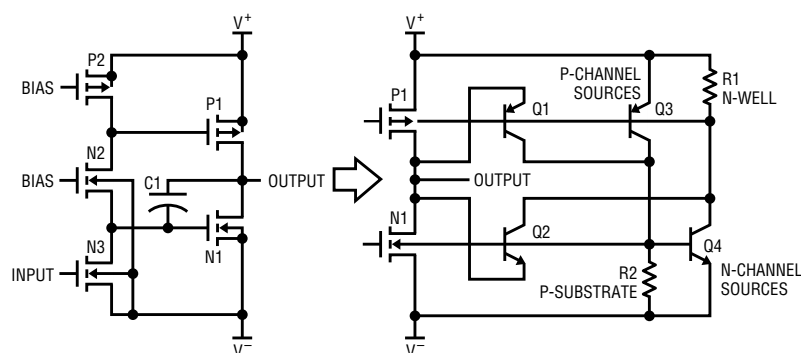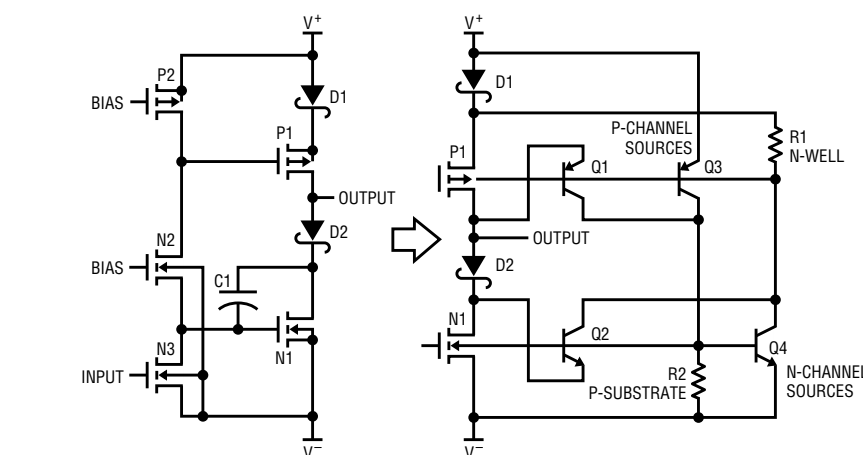

Figure 2. Simplified LTC1337 driver output stage and parasitic devices

below V–, the parasitic P+/N-well/P-substrate PNP or the N+/P-substrate/N-well NPN (Q1, or Q2, respectively) will turn on.

The collector current of Q1 or Q2 will then turn on the SCR structure formed by R1, R2, Q3, and Q4. The SCR structure forms a short-circuit path between V+ and V–, and remains on even though the output is brought back within the supply rails (i.e., Q1 or Q2 turns off). This will cause the charge pump to collapse and excessive current to flow. The latch-up state can only be stopped by turning off the power.

By incorporating Schottky diodes into the output stage of the LTC1337

driver (simplified in Figure 2), the triggering source of the latch-up is eliminated. When the output is forced above V+ or below V–, Schottky diode D1 or D2 will reverse bias and prevent current flow. By preventing Q1 or Q2 from turning on, the trigger source of latch-up is eliminated.

ESD Protection

ESD protection is provided by a specialized ESD protection circuit connected to the receiver inputs and driver outputs (Figure 3). The protection cell is designed to turn on and absorb the ESD pulse energy, but to remain off during normal operation. Figure 4 shows the I-V curve of the ESD protection cell.

When the voltage reaches about 23V with respect to ground, the ESD cell starts to turn on; when it reaches about 28V the cell snaps back and conducts a large amount of current while limiting the voltage to 5V. During an ESD pulse, the ESD cell turns on and clamps the pin at 5V until all of the energy has been dissipated, then returns to its off state. With the voltage clamped at 5V, the power dissipation is kept to a minimum and

Figure 1. Conventional CMOS output stage and parasitic devices

Switching Regulator Provides Constant 5V Output from 3.5 – 40V Input, without a Transformer

by Brian Huffman

A common switching-regulator requirement is to produce a constant output voltage from an input voltage that varies above or below the output voltage. This is particularly important for extending battery life in battery-powered applications. Figure 1 shows how an LT1171 switching regulator IC, two inductors, and a "flying" capacitor can generate a constant output voltage that is independent of input voltage variations. This is accomplished without the use of a transformer. Inductors are preferred over transformers because they are readily available and more economical.

The circuit in Figure 1 uses the LT1171 to control the output voltage. A fully self-contained switching regulator IC, the LT1171 contains a power switch as well as the control circuitry (pulse-width modulator, oscillator, reference voltage, error amplifier, and protection circuitry). The power switch is an NPN transistor in a common-emitter configuration; consequently, when the switch turns on, the LT1171's V_{SW} pin is connected to ground. This power switch can handle peak switch currents of up to 2.5A.

Figure 2 shows the operating waveforms for the circuit. In this architecture, the capacitor C2 serves as the single energy transfer device between the input voltage and output voltage of the circuit. While the LT1171 power switch is off, diode D1 is forward biased, providing a path for the currents from inductors L1 and L2. Trace A shows inductor L1's current waveform and trace B is L2's current waveform. Observe that the inductor current waveforms occur on top of a DC level.

The waveforms are virtually identical because the inductors have identical inductance values and the same voltages are applied across them. The current flowing through inductor L1 is not only delivered to the load, but is also used to charge C2. C2 is charged to a potential equal to the input voltage.

When the LT1171 power switch turns on, the V_{SW} pin is pulled to ground and the input voltage is applied across the inductor L1. At the same time, capacitor C2 is connected across inductor L2. Current flows from the input-voltage source through inductor L1 and into the LT1171. Trace C shows the voltage at the V_{SW} pin, and Trace D is the current flowing through the power switch. The catch diode (D1) is reversed biased, and capacitor C2's current also flows through the switch, through ground, and into inductor L2. During this
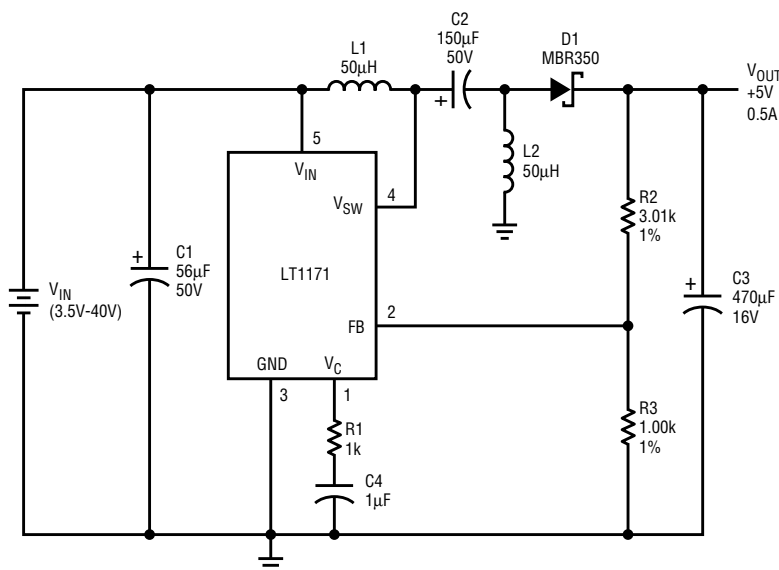
EQ. 1: $V_{OUT} = 1.25V (1 + R2/R3)$

C1 = NICHICON (AL) UPL1H560MEH, ESR = 0.250Ω, I_{RMS} = 360mA
C2 = NICHICON (AL) UPL1H151MPH, ESR = 0.100Ω, I_{RMS} = 820mA
C3 = NICHICON (AL) UPL1C471MPH, ESR = 0.090Ω, I_{RMS} = 770mA
L1, L2 = COILTRONICS CTX50-4, DCR = 0.090Ω,
    COILTRONICS (305) 781-8900

Figure 1. LT1171 provides constant +5V output from 3.5V to 40V input. No transformer is required

interval, C2 transfers its stored energy into inductor L2. After the switch turns off, the cycle is repeated.

Another advantage of this circuit is that it draws its input current in a triangular waveshape (see Trace A in Figure 2). The current waveshape of the input capacitor is identical to the current waveshape of inductor L1, except that the capacitor's current has no DC component. This type of ripple injects only a modest amount of noise into the input lines because the ripple does not contain any sharp edges.

Figure 3 shows the efficiency of this circuit for a 0.5A load and maximum output current for various input voltages. The two main loss elements

are the output diode (D1) and the LT1171 power switch. A Schottky diode is chosen for its low forward voltage drop; it introduces a 10% loss, which is relatively constant with input-voltage variations. At low input voltages the efficiency drops because the LT1171 power switch's saturation voltage becomes a higher percentage of the available input supply.

This circuit can deliver an output current of 0.5A at a 3.5V input voltage. This rises to 1A as input voltage is increased. Above 20V, higher output currents can be achieved by increasing the values of inductors L1 and L2. Larger inductances store more energy, providing additional current to the load. If 0.5A

of output current is insufficient, use a higher current part, such as the LT1170.

The output voltage is controlled by the LT1171 internal error amplifier. This error amplifier compares a fraction of the output voltage, via the R1–R2 divider network shown in Figure 1, with an internal 1.25V reference voltage, and varies the duty cycle until the two values are equal. (The duty cycle is determined by multiplying the switch-on time by the switching frequency.) The RC network (R1 and C4 in Figure 1) connected to the V_C pin provides sufficient compensation to stabilize this control loop. Equation 1 (on the schematic) can be used to determine the output voltage.

LT
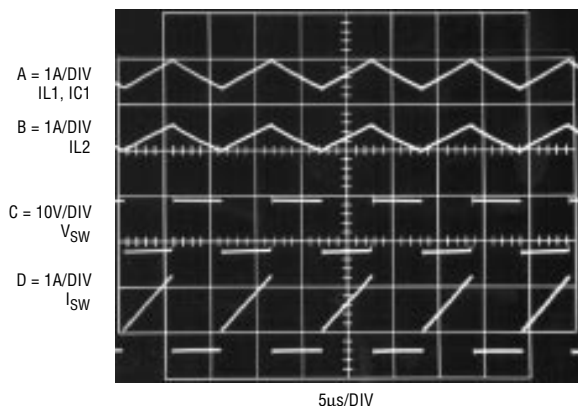
Figure 2. LT1171 switching waveforms
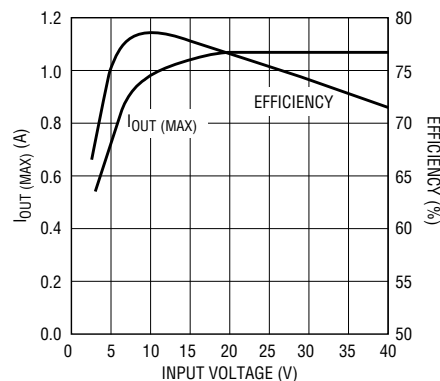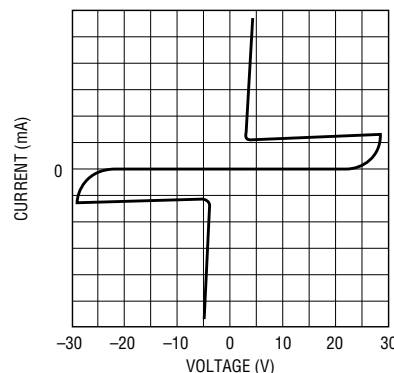
Figure 3. Efficiency and load characteristics for various input voltages
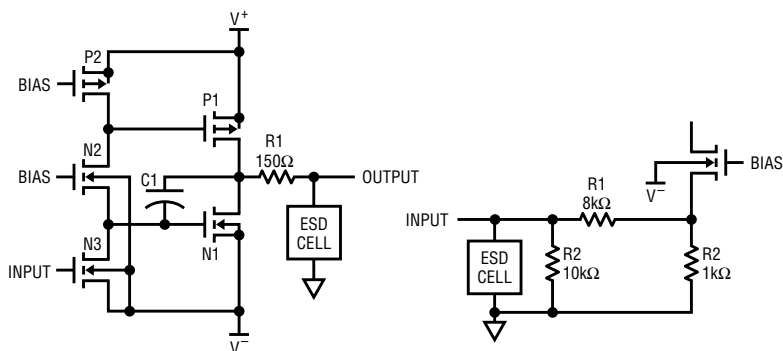
Figure 3. LTC1337 ESD protection

Figure 4. ESD protection cell IV curve

the CMOS devices are kept below the oxide-rupture voltage. To provide further protection, the receiver input goes through a 9-to-1 resistive divider before going to any CMOS devices. The ESD cell provides protection against over 10kV of human-body-model ESD strikes. *LT*

Optimizing A Video-Gain-Control Stage Using the LT1228

by Frank Cox

Video automatic-gain-control (AGC) systems require a voltage- or current-controlled gain element. The performance of this gain-control element is often a limiting factor in the overall performance of the AGC loop. The gain element is subject to several, often conflicting restraints. This is especially true of AGC for composite color video systems, such as NTSC, which have exacting phase- and gain-distortion requirements. To preserve the best possible signal-to-noise ratio (S/N),[1] it is desirable for the input signal level to be as large as practical. Obviously, the larger the input signal, the less the S/N will be degraded by the noise contribution of the gain-control stage. On the other hand, the gain-control element is subject to dynamic-range constraints, and exceeding these will result in rising levels of distortion.

Linear Technology makes a high-speed transconductance (g_m) amplifier, the LT1228, which can be used as a quality, inexpensive gain-control element in color video and some lower-frequency RF applications. Extracting the optimum performance from video AGC systems takes careful attention to circuit details.

As an example of this optimization, consider the typical gain-control circuit using the LT1228 shown in Figure 1. The input is NTSC composite video, which can cover a 10dB range,

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Extracting the optimum performance from video AGC systems takes careful attention to circuit details

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from 0.56 volt to 1.8 volt. The out-put is to be 1 volt peak-to-peak into 75 ohms. Amplitudes were measured from peak negative chroma to peak positive chroma on an NTSC modulated ramp test signal (see sidebar).

Notice that the signal is attenuated 20:1 by the 75 ohm attenuator at the input of the LT1228, so the voltage on the input (pin 3) ranges from 0.028 to 0.090 volt. This is done

to limit distortion in the transconductance stage. The gain of this circuit is controlled by the current into the I_{SET} terminal, pin 5 of the IC. In a closed-loop AGC system the loop-control circuitry generates this current by comparing the output of a detector[2] to a reference voltage, integrating the difference and then converting to a suitable current. The measured performance for this circuit is presented in Table 1.

All video measurements were taken with a Tektronix 1780R video-measurement set, using test signals generated by a Tektronix TSG 120. The standard criteria for characterizing NTSC video color distortion are the differential gain and the differential phase. For a brief explanation of these tests see the sidebar "Differential Gain and Phase." For this design exercise the distortion limits were set at a somewhat arbitrary 3% for differential gain and 3° for differential phase. Depending on conditions, this should be barely visible on a video monitor.

Figures 2 and 3 plot the measured differential gain and phase, respectively, against the input signal level
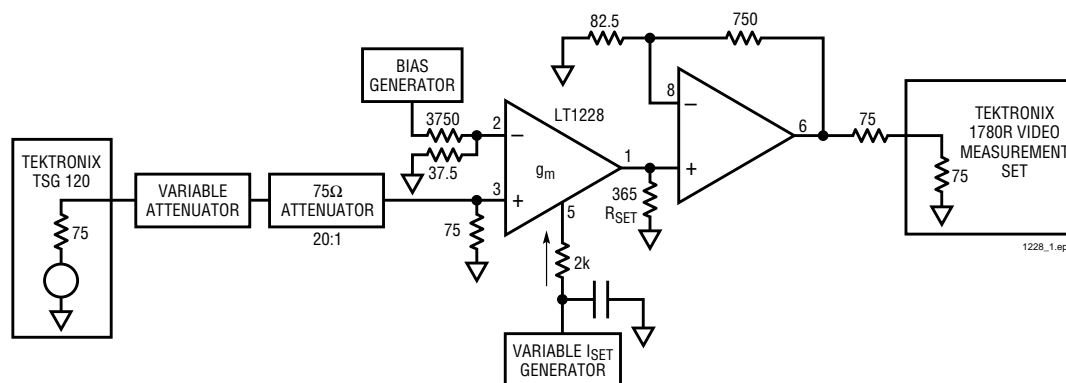

Figure 1. Schematic diagram

(the curves labeled "A" show the uncorrected data from Table 1). The plots show that increasing the input signal level beyond 0.06 volt results in a rapid increase in the gain distortion, but comparatively little change in the phase distortion. Further attenuating the input signal (and consequently increasing the set current) would improve the differential gain performance but degrade the S/N. What this circuit needs is a good tweak!

Optimizing for Differential Gain

Referring to the small signal transconductance versus DC input voltage graph (Figure 4), observe that the transconductance of the amplifier is linear over a region centered around zero volts.[3] The 25°C g_m curve starts to become quite nonlinear above 0.050 volt. This explains why the differential gain (see Figure 2, curve A) degrades so quickly with signals above

this level. Most RF signals do not have DC bias levels, but the composite video signal is mostly unipolar.

Video is usually clamped at some DC level to allow easy processing of sync information. The sync tip, the chroma reference burst, and some chroma-signal information swing negative, but 80% of the signal that carries the critical color information (chroma) swings positive. Efficient use of the dynamic range of the LT1228 requires that the input signal have little or no offset. Offsetting the video signal so that the critical part of the chroma waveform is centered in the linear region of the transconductance amplifier allows a larger signal to be input before the onset of severe distortion. A simple way to do this is to bias the unused input (in this circuit the inverting input, pin 2) with a DC level.

In a video system it might be convenient to clamp the sync tip at a more negative voltage than usual.

Clamping the signal prior to the gain-control stage is good practice because a stable DC reference level must be maintained.

The optimum value of the bias level on pin 2 used for this evaluation was determined experimentally to be about 0.03 volt. The distortion tests were repeated with this bias voltage added. The results are reported in Table 2 and Figures 2 and 3 (curves B). The improvement to the differential phase is inconclusive, but the improvement in the differential gain is substantial.

References

1. Signal to noise ratio, S/N = 20 x log(RMS signal/ RMS noise).

2. One way to do this is to sample the colorburst amplitude (the nominal peak-to-peak amplitude of the colorburst for NTSC is 40% of the peak luminance) with a sample-and-hold and peak detector.

3. Notice also that the linear region expands with higher temperature. Heating the chip has been suggested.

Table 1. Measured performance data (uncorrected)				
Input (volts)	I_{SET}(ma)	Differential Gain	Differential Phase	S/N
0.03	1.93	0.5%	2.7°	55dB
0.06	0.90	1.2%	1.2°	56dB
0.09	0.584	10.8%	3.0°	57dB

Table 2. Measured performance data (corrected)					
Input (volts)	Bias Voltage	I_{SET}(ma)	Differential Gain	Differential Phase	S/N
0.03	0.03	1.935	0.9%	1.45°	55dB
0.06	0.03	0.889	1.0%	2.25°	56dB
0.09	0.03	0.584	1.4%	2.85°	57dB
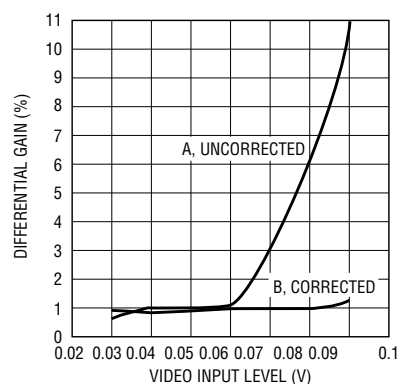
Figure 2. Differential gain vs. input level
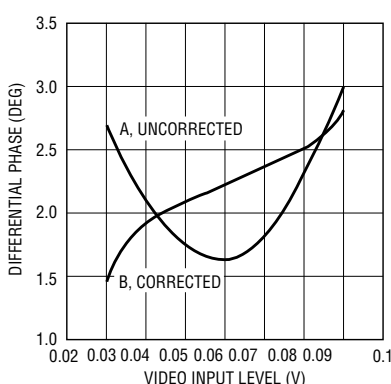
Figure 3. Differential phase vs. input level
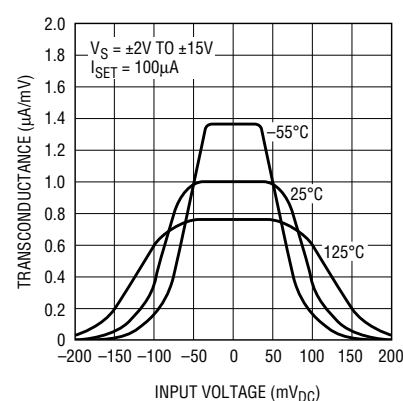
Figure 4. Small-signal transconductance vs. DC input voltage

Differential Gain and Phase

Differential gain and phase are sensitive indications of chroma-signal distortion. The NTSC system encodes color information on a separate subcarrier at 3.579545MHz. The color subcarrier is directly summed to the black and white video signal. (The black and white information is a voltage proportional to image intensity and is called luminance or luma.) Each line of video has a burst of 9 to 11 cycles of the subcarrier (so timed that it is not visible) that is used as a phase reference for demodulation of the color information of that line. The color signal is relatively immune to distortions, except for those that cause a phase shift or an amplitude error to the subcarrier during the period of the video line.

Differential gain is a measure of the gain error of a linear amplifier at the frequency of the color subcarrier. This distortion is measured with a test signal called a modulated ramp (shown in Figure 5). The modulated ramp consists of the color subcarrier frequency superimposed on a linear ramp (or sometimes on a stair step). The ramp has the duration of the active portion of a horizontal line of video. The amplitude of the ramp varies from zero to the maximum level of the luminance, which, in this case, is 0.714 volt. The gain error corresponds to compression or expansion by the amplifier (sometimes called "incremental gain") and is expressed as a percentage of the full amplitude range. An appreciable amount of differential gain will cause the luminance to modulate the chroma, causing visual chroma distortion. The effect of differential gain errors is to change the saturation of the color being displayed. Saturation is the relative degree of dilution of a pure color with white. A 100% saturated color has 0% white, a 75% saturated color has 25% white, and so on. Pure red is 100% saturated, whereas pink is red with some percentage of white and is therefore less than 100% saturated.

Differential phase is a measure of the phase shift in a linear amplifier at the color subcarrier frequency when the modulated ramp signal is used as an input.
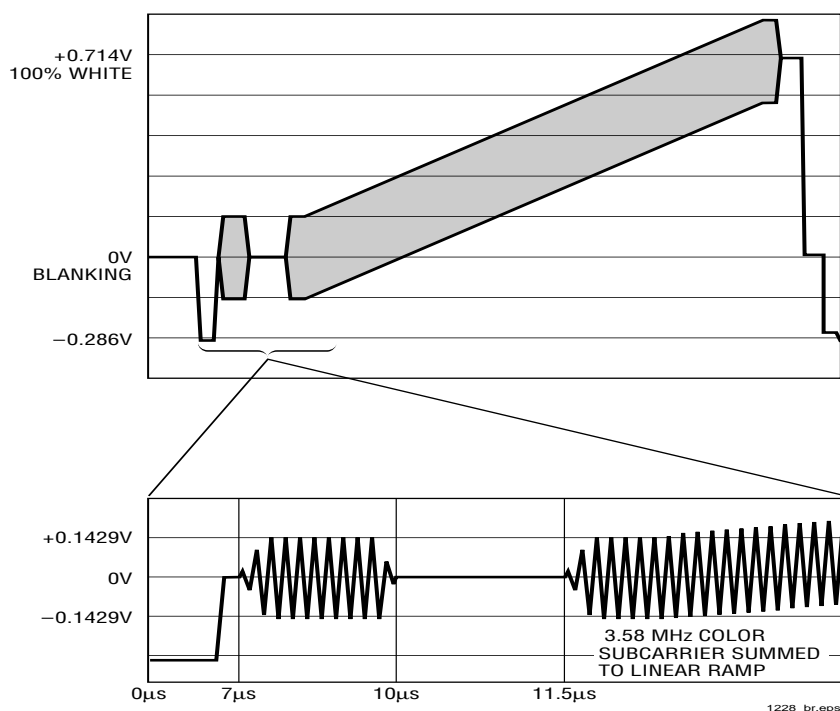
The phase shift is measured relative to the colorburst on the test waveform and is expressed in degrees. The visual effect of the distortion is a change in hue. Hue is the that quality of perception which differentiates the frequency of the color, red from green, yellow-green from yellow, and so forth.

Three degrees of differential phase is about the lower limit that can unambiguously be detected by observers. This level of differential phase is just detectable on a video monitor as a shift in hue, mostly in the yellow-green region. Saturation errors are somewhat harder to see at these levels of distortion—3% of differential gain is very difficult to detect on a monitor. The test is performed by switching between a reference signal, SMPTE (Society of Motion Picture and Television Engineers) 75% color bars, and a distorted version of the same signal, with matched signal levels. An observer is then asked to note any difference.

In professional video systems (studios, for instance) cascades of processing and gain blocks can reach hundreds of units. In order to maintain a quality video signal, the distortion contribution of each processing block must be a small fraction of the total allowed distortion budget[4] (the errors are cumulative). For this reason, high-quality video amplifiers will have distortion specifications as low as a few thousandths of a degree for differential phase and a few thousandths of a percent for differential gain. *LT*

Figure 5. NTSC test signal

References

4. From the preceding discussion, the limits on visibility are about 3° differential phase, 3% differential gain. Please note that these are not hard and fast limits. Tests of perception can be very subjective.

Noise Generators for Multiple Uses

A Broadband Random Noise Generator

by Jim Williams

Filter, audio, and RF-communications testing often require a random noise source. Figure 1's circuit provides an RMS-amplitude regulated noise source with selectable bandwidth. RMS output is 300 millivolts, with a 1kHz to 5MHz bandwidth, selectable in decade ranges.

Noise source D1 is AC coupled to A2, which provides a broadband gain of 100. A2's output feeds a gain-control stage via a simple, selectable lowpass filter. The filter's output is applied to A3, an LT1228 operational transconductance amplifier. A1's output feeds LT1228 A4, a current-feedback amplifier. A4's output, which is also the circuit's output, is sampled by the A5-based gain-control configuration. This closes a gain control loop to A3. A3's I_{SET} current controls gain, allowing overall output level control.

Figure 2 plots noise at a 1MHz bandpass, whereas Figure 3 shows RMS noise versus frequency in the same bandpass. Figure 4 plots similar information at full bandwidth (5MHz). RMS output is essentially flat to 1.5MHz, with about ±2dB control to 5MHz before sagging badly.

A Diode Noise Generator for "Eye Diagram" Testing

by Richard Markell

The circuit that Jim Williams describes evolved from my desire to build a circuit for testing communications channels by means of "eye diagrams." (See *Linear Technology*, Volume I, Number 2 for a short explanation of the eye diagram.) I wanted to replace my pseudo-random code generator circuit, which used a PROM, with a more "analog" design—one that more people could build without specialized components. What evolved was a noise source sampled by a very fast comparator (see Figure 5). The comparator outputs a random pattern of 1's and 0's.

The noise diode (an NC201) is filtered and amplified by the LT1190 high-speed operational amplifier (U1). The output feeds the LT1116 (U2), a 12ns, single-supply, ground-sensing comparator. The 2kΩ pot at the inverting input of the LT1116 sets the threshold to the comparator so that a quasi-equal number of 1's and 0's are output. U3 latches the output from U2 so that the output from the comparator remains latched throughout one clock period. The two-level output is taken from U3's Q0 output.

The additional circuitry shown in the schematic diagram allows the circuit to output four-level data for PAM (pulse amplitude modulation) testing. The random data from the two-level output is input to a shift register, which is reset on every fourth clock pulse. The output from the shift register is weighted by the three 5kΩ resistors and summed into the LT1220 operational amplifier from which the output is taken. The filter network between the 74HC74 output and the 74HC4094 strobe input is necessary to ensure that the output data is correct.
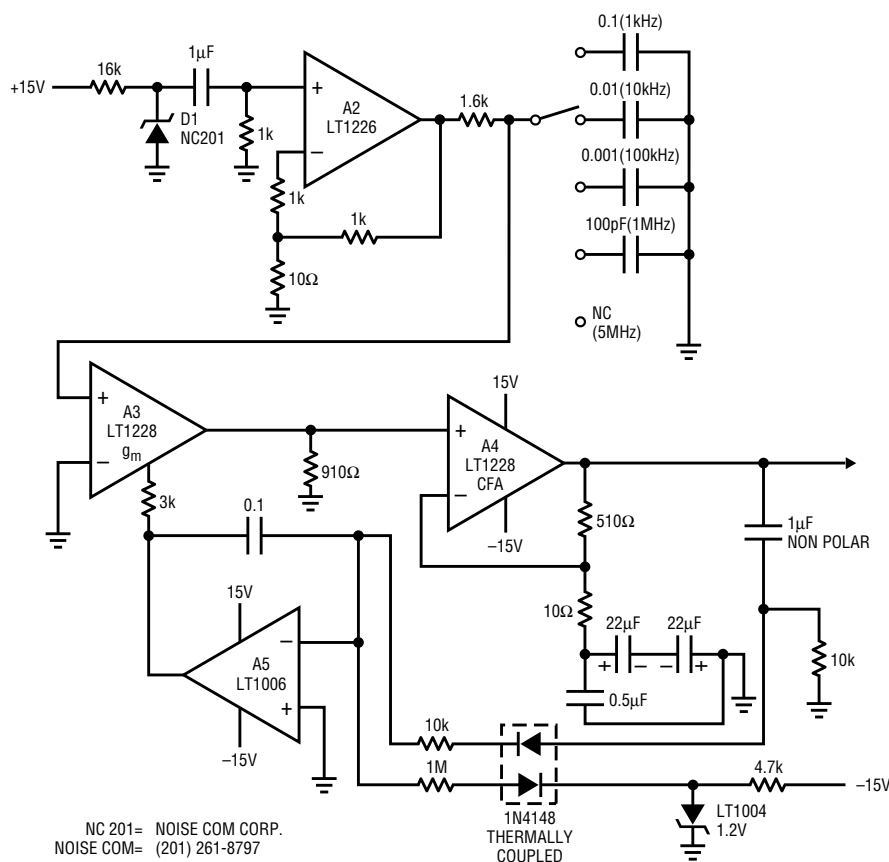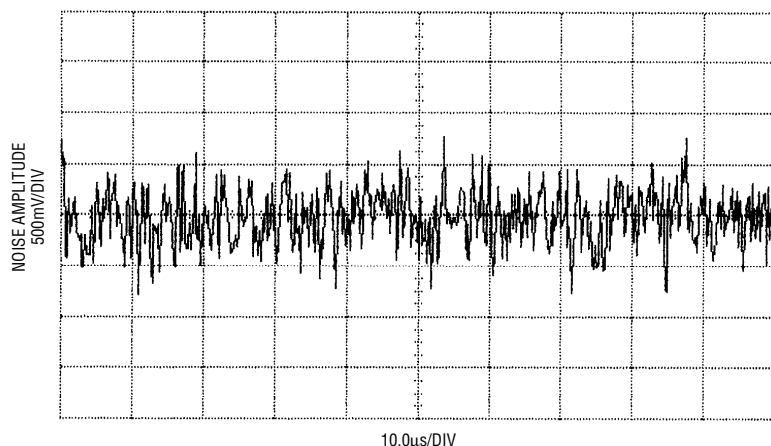

Figure 1. Broadband random noise generator schematic
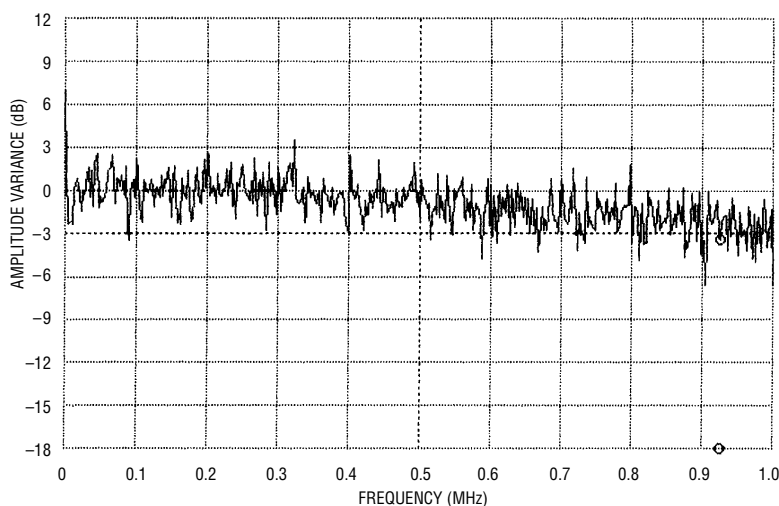
Figure 2. Noise amplitude at 1MHz bandpass

Figure 3. RMS noise vs. frequency at 1MHz bandpass
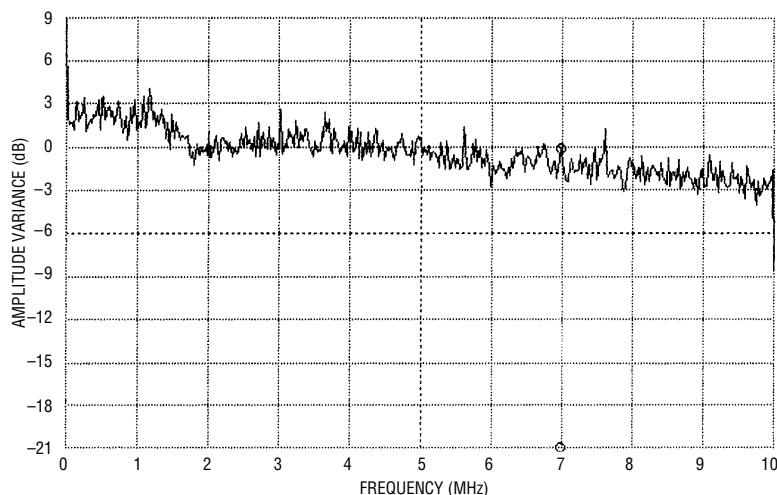

Figure 4. RMS noise vs. frequency at 5MHZ bandpass

Symmetrical White Gaussian Noise

by Bent Hessen-Schmidt,
NOISE COM, INC.

White noise provides instantaneous coverage of all frequencies within a band of interest with a very flat output spectrum. This makes it useful both as a broadband stimulus and as a power-level reference.

Symmetrical white Gaussian noise is naturally generated in resistors. The noise in resistors is due to vibrations of the conducting electrons and holes, as described by Johnson and Nyquist.[1,2] The distribution of the noise voltage is symmetrically Gaussian, and the average noise voltage is:

$$\overline{V}_n = 2\sqrt{kT\int R(f)\,p(f)\,df} \qquad (1)$$

where:

   $k =$ 1.38E–23 J/K (Boltzmann's constant)
   $T =$ temperature of the resistor in Kelvin
   $f =$ frequency in Hz
   $h =$ 6.62E–34 Js (Planck's constant)
   $R(f) =$ resistance in ohms as a function of frequency

$$p(f) = \frac{hf}{kT[\exp(hf/kT) - 1]} \qquad (2)$$

$p(f)$ is close to unity for frequencies below 40GHz when T is equal to 290°K. The resistance is often assumed to be independent of frequency, and $\int df$ is equal to the noise bandwidth (B). The available noise power is obtained when the load is a conjugate match to the resistor, and it is:

$$N = \frac{\overline{V}_n^{\,2}}{4R} = kTB \qquad (3)$$

where the "4" results from the fact that only half of the noise voltage and hence only 1/4 of the noise power is delivered to a matched load.

Equation 3 shows that the available noise power is proportional to the temperature of the resistor; thus it is often called thermal noise power. Equation 3 also shows that white noise power is proportional to the bandwidth.

An important source of symmetrical white Gaussian noise is the noise diode. A good noise diode generates a high level of symmetrical white Gaussian noise. The level is often specified in terms of excess noise ratio (ENR).

$$\text{ENR (in dB)} = 10\text{Log}\frac{(T_e - 290)}{290} \quad (4)$$

T_e is the physical temperature that a load (with the same impedance as the noise diode) must be at to generate the same amount of noise.

The ENR expresses how many times the effective noise power delivered to a non-emitting,

non-reflecting load exceeds the noise power available from a load held at the reference temperature of 290°K (16.8°C or 62.3°F).

The importance of a high ENR becomes obvious when the noise is amplified, because the noise contributions of the amplifier may be disregarded when the ENR is 17dB larger than the noise figure of the amplifier (the difference in total noise power is then less than 0.1dB). The ENR can easily be converted to noise spectral density in dBm/Hz or $\mu V/\sqrt{Hz}$ by use of the white noise conversion formulas in Table 1.

Table 1. Useful white noise conversion

dBm $= \text{dBm/Hz} + 10\log(BW)$
dBm $= 20\log(\overline{V}n) - 10\log(R) + 30dB$
dBm $= 20\log(\overline{V}n) + 13dB$ for R = 50ohms
dBm/Hz $= 20\log(\mu Vn\sqrt{Hz}) - 10\log(R) - 90dB$
dBm/Hz $= -174\text{dBm/Hz} + \text{ENR}$ for ENR > 17dB

When amplifying noise it is important to remember that the noise voltage has a Gaussian distribution. The peak voltages of noise are therefore much larger than the average or RMS voltage. The ratio of peak voltage to RMS voltage is called crest factor, and a good crest factor for Gaussian noise is between 5:1 and 10:1 (14 to 20dB). An amplifier's 1dB gain-compression point should therefore be typically 20dB larger than the desired average noise-output power to avoid clipping of the noise.

For more information about noise diodes, please contact NOISE COM, INC. at (201) 261-8797. *LT*

References

1. Johnson, J.B. "Thermal Agitation of Electricity in Conductors," *Physical Review*, July 1928, pp. 97-109.
2. Nyquist, H. "Thermal Agitation of Electric Charge in Conductors," *Physical Review*, July 1928, pp. 110–113.
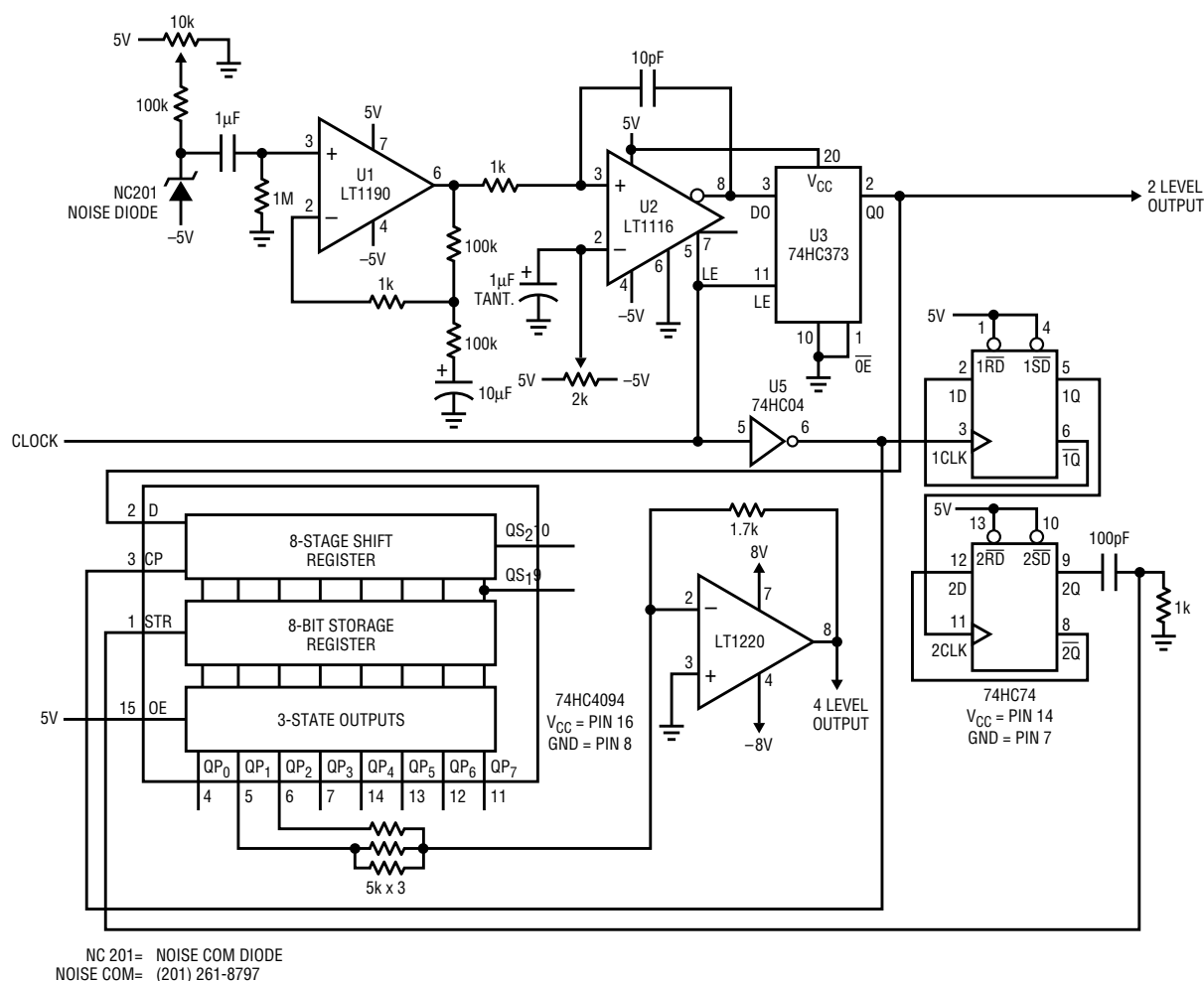
Figure 5. Pseudo-random code generator schematic diagram

High-Current Synchronous Switcher Converts 5V to 3.3V at 90% Efficiency

by Milt Wilcox

The next generation of microprocessors used in desktop computers and workstations will consume prodigious amounts of current at 3.3V. Often a high-current, offline 5V supply is already available. The problem then becomes how to generate 3.3V from the available supply without the need for costly and space-consuming heat sinks. For example, dropping 5V to 3.3V in a linear regulator would consume a minimum of 17W.

The circuit in Figure 1 efficiently converts 5V to 3.3V at output currents up to 15A and features soft-start and short-circuit protection. The efficiency at 10A output is 91% (see Figure 2), meaning that only 7.25A is drawn from the 5V input. Because the 3.3W-conversion power loss is fairly evenly divided among R_S, L1, and the four MOSFETs, heat sinking is not required for operation at 10A. At 15A output, total losses are still only 7W, requiring minimal heat sinking.

At start-up, the circuit in Figure 1 uses the charge pump built into the LT1158 synchronous, N-channel MOSFET driver to pump the gates of the top MOSFET switch above V_{IN}. Start-up current is controlled by the same LT1158 protection loop that provides short-circuit protection. The DC level of the CMOS 555 triangle-wave oscillator relative to the LT1158 input threshold sets the PWM duty cycle during normal operation. The LT1431 contains the reference and error amplifier that controls the DC level of the triangle wave via the CMOS 555 supply pin.

When the output is shorted, the LT1158 current-sense comparator overrides the LT1431 output to create a current-mode protection loop. Current limit is approximately $100mV/R_S$, or approximately 20A for the Figure 1 circuit. During a short, the duty cycle drops to a very low
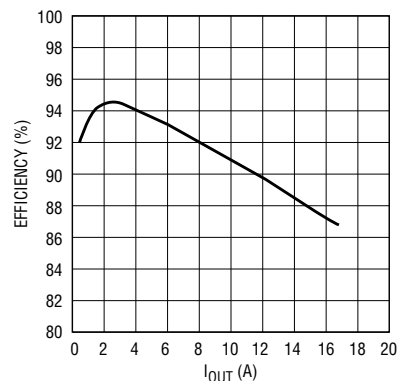

Figure 2. Efficiency for 5V to 3.3V high-current synchronous switcher

value, increasing the dissipation in each bottom MOSFET to approximately 3W. Therefore, 20°C/W heat sinking should be provided for the bottom MOSFETs if continuous short-circuited operation is required.
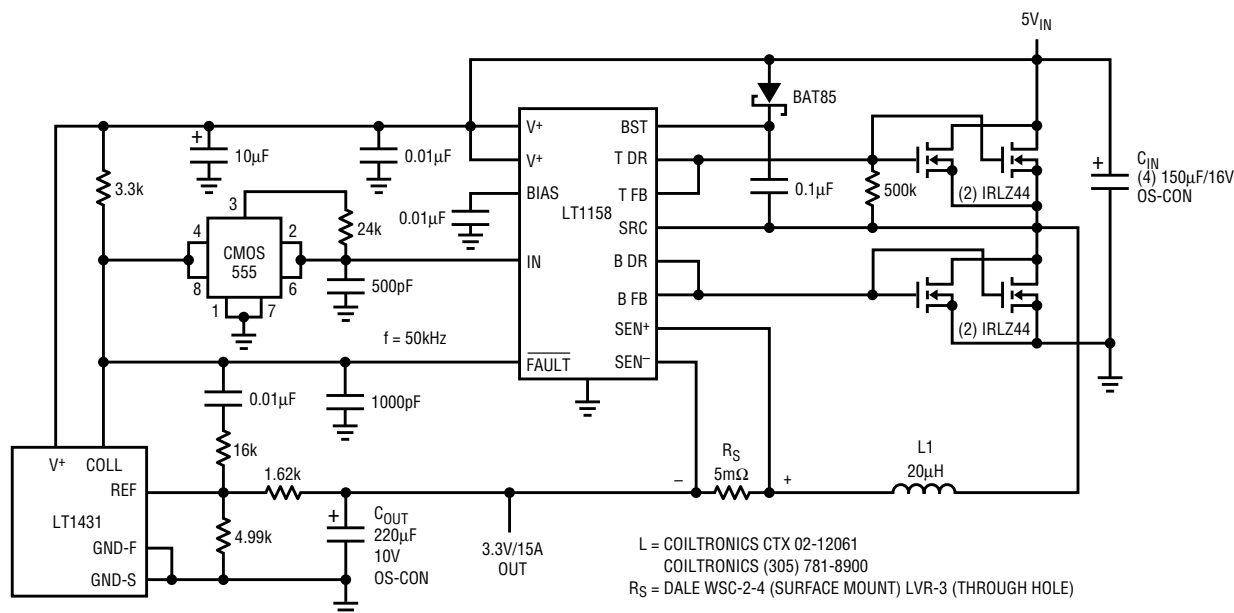
LT

Figure 1. 5V to 3.3V/15A high-efficiency switching regulator

Protected Bias for GaAs Power Amplifiers

by Mitchell Lee

Portable communications devices such as cellular telephones and answer-back pagers rely on small, GaAsFET-based 0.1 to 1.0W RF amplifiers as the transmitter output stage. The main power device requires a negative gate-bias supply, which is not readily available in a battery-operated product. The circuit shown in Figure 1 not only develops a regulated negative gate bias, it also switches the positive supply, protects against the loss of gate bias, limits power dissipation in the amplifier under high standing-wave ratio (SWR) conditions, and protects against amplifier failures that might otherwise short-circuit the battery pack.

Negative bias is supplied by an LTC1044 charge-pump inverter, and the amplifier's positive supply is switched by an LTC1153 electronic circuit breaker. An open-collector switch can be used to turn the LTC1044 inverter off by grounding the OSC pin (7). When off, the LTC1044 draws only 2μA.

The negative output from the LTC1044 is sensed by a 2.5V reference diode (IC2) and Q2. With no negative bias available, Q2 is off and Q3 turns on, pulling the LTC1153's control input low. This shuts off the GaAs amplifier. Total standby power, including the LTC1044, is approximately 25μA.
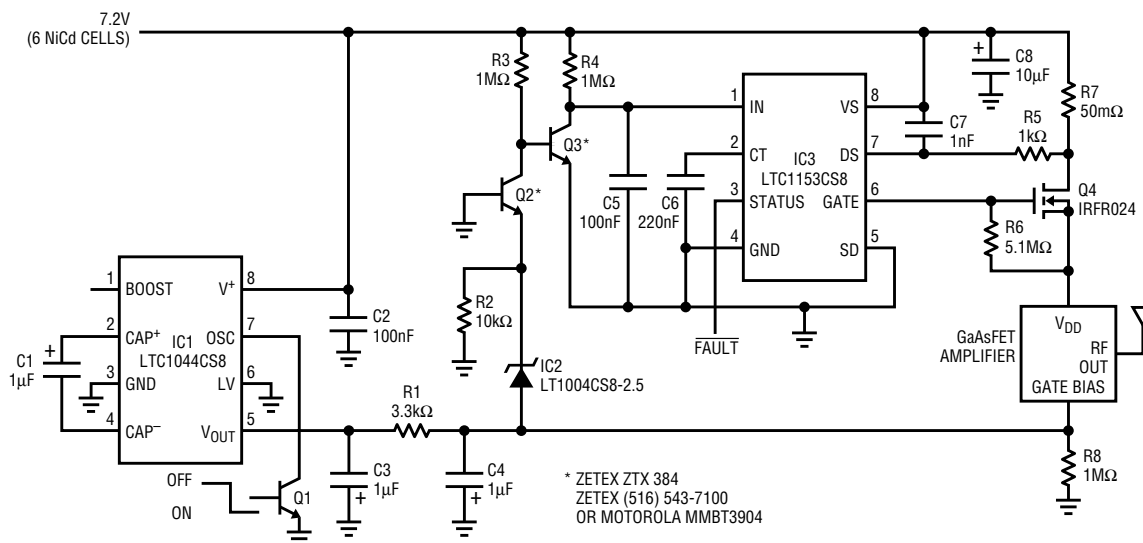
If the LTC1044's OSC pin (7) is released, a negative output, nearly equal in magnitude to the battery input voltage, appears at V_{OUT} (pin 5). The negative bias is regulated by R1, IC2, and Q2's base-emitter junction. Q2 saturates, shutting Q3 off and thereby turning the LTC1153 on.

The LTC1153 charges the N-channel MOSFET (Q4) gate to 10V above the battery potential, switching Q4 fully on. Power is thus applied to the GaAs amplifier.

The nominal negative bias is –3.2V, comfortably assuring the –2.5V minimum specified for the amplifier. Total quiescent current, exclusive of the GaAs amplifier drain supply, is approximately 1.5mA in the "on" state.

Short circuits or over-current conditions in the GaAs amplifier can damage the circuit board, the batteries, or both. The LTC1153 senses the amplifier's supply current and turns Q4 off if it is over 2A. After a timeout period set by C6 (200ms) the LTC1153 tries again, turning Q4 on. If the amplifier's supply current is still too high, the LTC1153 trips off again. This cycle continues until the fault condition is cleared. Under fault conditions the LTC1153's STATUS pin (3) is low. As soon as the fault is cleared, the LTC1153 resets and normal operation is restored. *LT*

Figure 1. Schematic diagram

High-Current, Synchronous, Step-Down Switching Regulator

by Brian Huffman

The new LTC1149 can drive an external, N-channel MOSFET to achieve high-efficiency, high-current power conversion. The standard P-channel approach is preferred at output currents of under 2A; however, P-channel MOSFETs become a dominant loss element at higher output currents, limiting overall circuit efficiency. Consequently, N-channel MOSFETs are better suited for use in high-current applications, since they have a substantially lower on-resistance than comparably priced P-channels. The best P-channels have an on-resistance of 60mΩ, whereas N-channels with on-resistances less than 25mΩ are readily available. The circuit shown in Figure 1 uses the low-loss characteristics of N-channel MOSFETs, providing efficiency in excess of 90% at an output current of 5A.

The circuit's operation is as follows: the LTC1149 provides a PDRIVE output (Pin 4) that swings between ground and 10V, turning Q3 on and off. While Q3 is on, the N-channel MOSFET (Q4) is off because its gate is pulled low by Q3, through D2. During this interval, the NGATE output (pin 13) turns the synchronous switch (Q5) on, creating a low-resistance path for the inductor current.

Q4 turns on when its gate is driven above the input voltage. This is accomplished by bootstrapping capacitor C2 off the drain of Q4. The LTC1149 V_{CC} output (Pin 3) supplies a regulated 10V output that is used to charge C2 through D1 while Q4 is off. With Q4 off, C2 charges to 5V during the first cycle in Burst Mode™ operation and to 10V thereafter.

When Q3 turns off, the N-channel MOSFET is turned on by the SCR-connected NPN-PNP network (Q1 and Q2). Resistor R2 supplies Q2 with enough base drive to trigger the SCR. Q2 then forces Q1 to turn on,
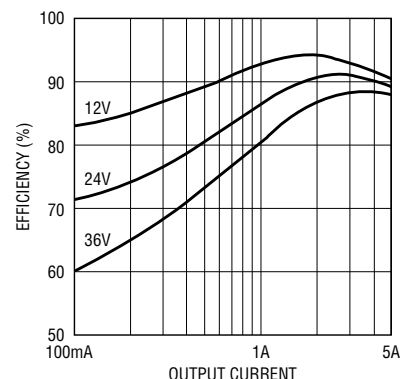
Figure 2. LTC1149-5 (12V–36V to 5V/5A) high-current buck

supplying more base drive to Q2. This regenerative process continues until both transistors are fully saturated. During this period, the source of Q4 is pulled to the input voltage. While Q4 is on, its gate-source voltage is approximately 10V, fully enhancing the N-channel MOSFET.

Efficiency performance for this circuit is quite impressive. Figure 2 shows that for a 12V input the efficiency never drops below 90% over the 0.6A to 5A range. At higher input voltages efficiency is reduced due to transition losses in the power MOSFETs. For low output currents, efficiency rolls off because of quiescent current losses. *LT*

C3 (TA) LOW ESR
C_IN NICHICON (AL) UPL1J102MRH, ESR = 0.027Ω, I_RMS = 2.370A
C_OUT SANYO (OS-CON) 10SA220M, ESR = 0.035Ω, I_RMS = 2.360A
Q1 PNP, BVCEO = 30V
Q2 NPN, BVCEO = 40V
Q3 SILICONIX NMOS, BVDSS = 60V, RDS_ON = 5Ω

Q4, Q5 NMOS, BVDSS = 60V, RDS_ON = 0.05Ω
D1, D2 SILICON, VBR = 75V
D3 MOTOROLA SCHOTTKY, VBR = 60V
R_SENSE = KRL NP-2A-C1-0R020J, P_D = 3W
L1 = COILTRONICS CTX50-5-52, DCR = 0.21Ω, IRON POWDER CORE
ALL OTHER CAPACITORS ARE CERAMIC

Figure 1. LTC1149-5 (12V–36V to 5V/5A) using N-channel MOSFETs

New Device Cameos

LT1180A, LT1181A, and LT1130A Family RS232 Transceivers

Three new RS232 interface transceivers provide enhanced performance and fault tolerance compared to existing devices. The LT1180A and LT1181A are tw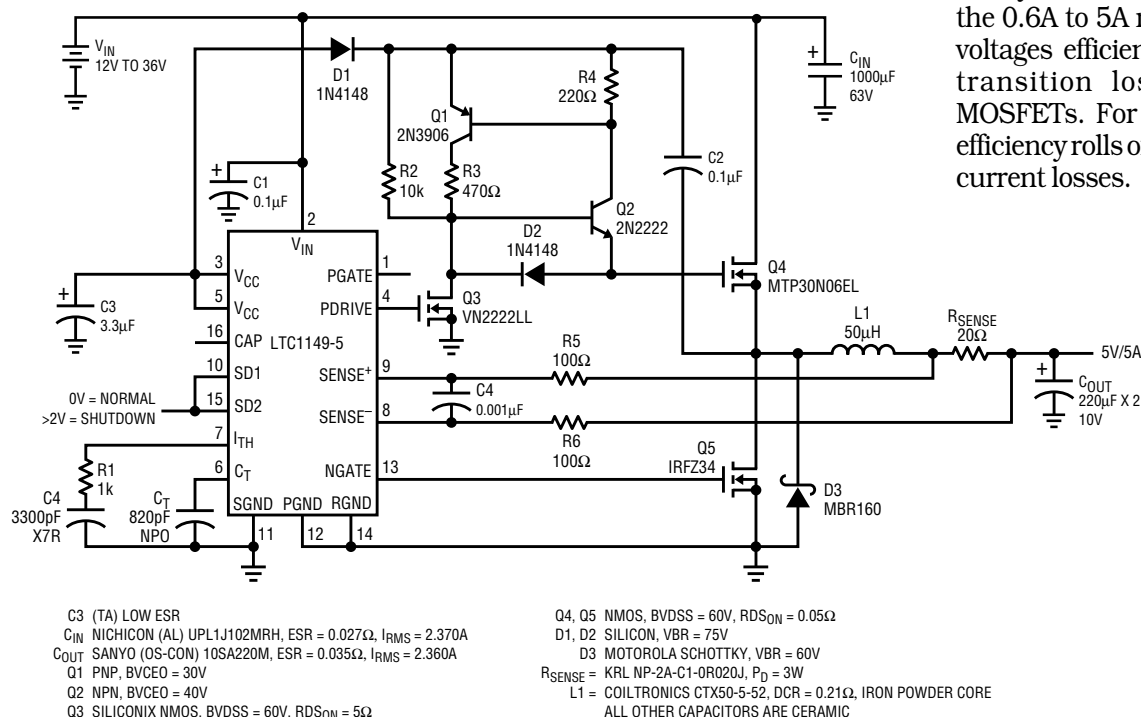o-driver/two-receiver transceivers, pin compatible with the existing LT1080/LT1081 and LT1180/LT1181. The LT1130A family consists of eleven devices that are pin compatible with the LT1130 family. Members of the LT1130A family provide up to five drivers and five receivers with a charge pump in a single package.

All of the new circuits feature ±10kV ESD protection on the RS232 line pins and operate to 120k baud while driving up to 2500pF loads. The on-chip charge-pump power generators are capable of using low-cost 0.1μF capacitors to generate RS232 levels from standard, 5V power supplies. Power consumption is reduced by as much as 43% from the earlier devices, while SHUTDOWN and DRIVER DISABLE operating modes allow further power savings in some systems.

All of these circuits are available in Plastic DIP or SOIC packages. The LT1137A and LT1138A are also available in 28-lead SSOP packages.

The LTC1163 Triple 1.8V to 6V MOSFET Driver

The LTC1163 triple 1.8V to 6V gate driver makes it possible to switch either supply- or ground-referenced loads through low $R_{DS(ON)}$ N-channel switches from as little as 1.8V (two discharged cells). The LTC1163 contains three on-chip charge pumps so that less expensive, lower $R_{DS(ON)}$ N-channel MOSFETs can be used to replace high-side P-channel switches, which generally perform poorly below 5V and cannot operate down to 1.8V.

The three charge pumps require no external components and have been designed to be very efficient. The standby current with the three inputs switched off is typically 0.01 microamps. The quiescent current rises to 95 microamps per channel with the input turned on and the charge pump producing 11V from a 3.3V supply.

Micropower operation, coupled with a power supply range of 1.8V to 6V, makes the LTC1163 ideal for 2- to 4-cell battery-powered applications. The LTC1163 is also well suited for 3.3V and 5V nominal supply applications. The LTC1163 is available in both 8-pin DIP and 8-pin SOIC packaging. The pin-out has been optimized as a "flow-through" configuration, with the three inputs and ground pin on one side and the three outputs and supply pin on the other.

LT1204 Four-Input Video Multiplexer with Current-Feedback Amplifier

The LT1204 is a four-input video multiplexer designed to drive 150Ω cables and to expand easily into larger routing systems. Wide bandwidth, high slew rate, and low differential gain and phase make the LT1204 ideal for broadcast-quality routing. Gain flatness is 0.1dB at 40MHz for A_V = 2. Channel separation and disable isolation are 90dB at 10MHz. The channel-to-channel output switching transient is only 40mV, with a 50ns duration, making the transition imperceptible on high-quality monitors.

A unique feature of the LT1204 is its ability to expand into larger routing matrices. This is accomplished by bootstrapping the feedback resistors in the disable condition, raising the true output impedance of the circuit. This feature negates the effects of poor cable terminations in large systems.

A large input range of ±6V makes the LT1204 ideal for general-purpose analog signal selection and multiplexing. The new multiplexer operates on ±5V to ±15V supplies and a shutdown feature reduces the supply current to just 1mA. The part is available in a 16-lead plastic DIP (N) package or a 16-lead wide SO.

LT1381 5V-Powered RS232 Transceiver in a 16-Lead, Narrow-Body SO Package

The LT1381 is a new two-driver/ two-receiver RS232 interface transceiver with an integral charge-pump power generator. The circuit is available in 16-lead narrow SOIC packages, reducing board space by 27% over existing devices. The circuit is also available in a 16-lead plastic DIP.

The circuit contains an on-chip charge-pump generator, which uses small 0.1μF ceramic chip capacitors to supply RS232 output drive levels while operating from a single 5V power supply. Power consumption is a low 13mA max. RS232 I/O pins are protected from ESD transients in excess of ±10kV, eliminating the need for expensive TranZorbs. Driver outputs are capable of driving 2500pF loads to data rates in excess of 120k baud. Driver outputs are high impedance when powered down, and do not load the RS232 line as do CMOS devices.

LT1332 Wide-Supply-Range, Low Power RS232 Transceiver

The LT1332 is an extremely low power RS232 transceiver intended for systems that operate at low voltage yet require true RS232 output levels. The LT1332 is designed to be powered from an external switching regulator, which may be used elsewhere for power conditioning. In a typical application, the LT1332 shares

the regulator's positive output, while charge is capacitively pumped from the regulator's switch pin to the negative supply. Schottky rectifiers built into the LT1332 simplify the charge-pump design.

The LT1332/switcher combination generates fully compliant RS232 signal levels from inputs as low as 2V (e.g., two dead AA batteries) or as high as 6V (e.g., a poorly regulated 5V supply). The LT1109A can deliver greater than 100mA of output current, making it an excellent choice for powering FLASH memory, while the surplus power available to the LT1332 is sufficient to drive long cables, heavy loads, or mice. When operated unloaded, the LT1332 draws 1mA of supply current. While shut down, the LT1332 typically draws 40µA of supply current which keeps one receiver active for detecting start-up signals.

If external RS232 supplies are available (6.6V < V+ < 13.2V; –13.2V < V– < –6.6V), the LT1332 can be used as a stand-alone unit. When used in this way, the LT1332's low supply current makes it an attractive alternative to the LT1141.

The LT1332 is a complete RS232 serial port with three drivers and five receivers, arranged in a flow-through architecture similar to that of the popular LT1137. Advanced driver output stages operate up to 120k baud while driving heavy capacitive loads. New ESD structures on the driver outputs and receiver inputs make the LT1332 resilient to multiple ±10kV strikes, eliminating costly transient suppressors.

The LT1332 requires two external charge pump caps for the negative supply, but only small decoupling caps are needed on the positive supply inputs. The low parts count helps reduce board space. The LT1332 is available in SO, SSOP, and DIP packages.

LT1259 Dual/LT1260 Triple Current-Feedback Amplifiers

The LT1259 contains two independent 100MHz current-feedback amplifiers, each with a shutdown pin. These amplifiers are designed for excellent linearity while driving cables and other low impedance loads. The LT1260 is a triple version, especially suited to RGB video applications. These amplifiers operate on all supplies from single 5V to ±15V and draw only 5mA per amplifier when active.

When shut down, the LT1259/LT1260 amplifiers draw zero supply current and their outputs become high impedances. The amplifiers' turn-on times are only 70ns and their turn-off times are 140ns. In portable equipment and systems using several amplifiers, the shutdown feature reduces system power without the problems normally associated with power supply switching. The shutdown feature also allows the outputs of several amplifiers to be wired together to make a video mux amp.

Dual and triple amplifiers significantly reduce costs compared with singles; board space and the number of insertions are reduced and fewer supply-bypass capacitors are required. The wide bandwidth and high slew rate of these amplifiers make driving RGB signals easy. Only two LT1260s are required to make a complete, two input RGB MUX and cable driver.

The LT1259 is available in 14-pin DIPs and narrow S14 surface mount packages. The LT1260 is available in 16-pin DIPs and narrow S16 surface mount packages. *LT*

LTC in the News . . .
Third Quarter Earnings

Thanks to the continued patronage of our customers and acceptance of our products and application solutions in the marketplace, Linear Technology Corporation recorded its 30th consecutive quarter of increased sales and income levels. Linear Technology Corporation net sales for the third quarter of 1993 reached a record $38,806,000, a 27% increase over the third quarter of 1992. Third quarter net income also reached a record $9,571,000 or $0.26 per share, an increase of 46% over last year (adjusted to reflect the two-for-one stock split distributed on November 24, 1992).

Rankings & Ratings

Magazines and newspapers annually rank leading companies on the bases of sales growth, profitability, return on equity and other criteria. Linear Technology has traditionally done very well in all of the ratings for which the company qualified. This year is no exception. Three of the year's recent evaluations are summarized below.

In *Business Week*'s "Top 1000: America's Most Valuable Companies," based on its market value, Linear ranked 812th in the U.S. compared to 853rd last year, when LTC first appeared on this list. This ranking placed Linear Technology Corporation best in its class and 10th among all semiconductor manufacturers.

Financial World magazine ranked Linear Technology Corporation 1st among all semiconductor companies and 22nd among its "Top 200 Growth Companies in the U.S." ranking for 1993.

Upside magazine is a national magazine published in San Francisco that focuses on high technology companies. In this year's "Upside 100" list, Linear Technology placed 5th in market value growth among all semiconductor companies and 31st within the 800 technology companies in the survey.

Industry statistical analysts at **In-Stat** in Scottsdale, Arizona, named Linear Technology "the best financially managed manufacturer in the semiconductor industry" in its annual Kachina Awards.

Bob Swanson, President of Linear Technology, shares the honor of *Financial World* "CEO of the Year" with the CEO of Loral and Andy Grove of Intel.

LT

For further information on the above or any other devices mentioned in this issue of Linear Technology, use the reader service card or call the LTC literature-service number: (800) 637-5545. Ask for the pertinent data sheets and application notes.

DESIGN TOOLS
Applications on Disk

NOISE DISK
This IBM-PC (or compatible) progam 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 232 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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