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MAXIM REPORTS RECORD REVENUES AND EARNINGS FOR THE FIRST QUARTER OF FISCAL 2001

Maxim Integrated Products, Inc., (MXIM) reported record net revenues of $285.1 million for its fiscal first quarter ending September 23, 2000, a 58.3% increase over the $180.0 million reported for the same quarter a year ago. Net income increased to a record $93.3 million in the first quarter, compared to $58.4 million last year, a 59.8% increase. Diluted earnings per share were $0.29 for the first quarter, a 52.6% increase over the $0.19 reported for the same period a year ago.

During the quarter, cash and short-term investments increased by $71.8 million after paying $95.5 million for 1,445,000 shares of the Company's common stock, $7.1 million for properties and facilities, and $30.3 million for wafer fab and test equipment. Accounts receivable decreased by $2.7 million to $144.5 million, while inventories increased $5.4 million to $64.0 million during the quarter.

Gross margin for the first quarter increased to 70.5%, compared to 70.1% in the fourth quarter of last fiscal year. Research and development expense was $46.7 million or 16.4% of net revenues in the first quarter, compared to $45.3 million or 17.6% of net revenues in the fourth quarter. During the quarter, the Company recorded a writedown, primarily of equipment, of $14.9 million to cost of goods sold and $3.5 million to research and development expense. Additionally, the Company recorded a $3.0 million charge to selling, general and administrative expense related to technology licensing matters.

The Company recorded an expense in the first quarter of approximately $3.7 million in Medicare taxes on employees' realized stock option gains, compared to the fourth quarter expense of $2.4 million. Prior to the third quarter of fiscal 2000, these tax payments were recorded within stockholders' equity as an offset against the proceeds received from the exercise of stock options.

First quarter bookings were $339 million. This was above the Company's estimate of customer consumption for the first quarter of the current fiscal year. Turns orders received in the quarter were $98 million, compared to $108 million received in the prior quarter (turns orders are customer orders that are for delivery within the same quarter and may result in revenue within the same quarter if the Company has available inventory that matches those orders).

First quarter ending backlog shippable within the next 12 months was approximately $443 million, including approximately $353 million requested for shipment in the second quarter of fiscal 2001. The Company's fourth quarter ending backlog shippable within the next 12 months was approximately $420 million, including approximately $314 million that was requested for shipment in the first quarter. All of these backlog numbers have been adjusted to be net of cancellations and estimated future U.S. distribution ship and debit pricing adjustments.

Jack Gifford, Chairman, President, and Chief Executive Officer, commented on the results: "During the first quarter, we reached record revenues and earnings levels. Our balance sheet remains strong, with cash and short-term investments increasing over $70 million during the quarter. Annualized return on average stockholders' equity of 32.3% is one of the highest in the industry. New product introductions for the first quarter of our product introduction year were above our plan."

Mr. Gifford continued: "First quarter bookings moderated to a level closer to but significantly above estimated end market consumption for our products. On a related subject, I am often asked about 'where the industry is in the cycle.' My answer is that companies like Maxim focus relatively little on short-term purchasing patterns affected by inventory accumulation or shortage (which can often contribute to the cause of the cycles in question), but much more on the correctness of our product market targeting and the long-term success of our products in these chosen growth product markets. These are the key factors that have determined our growth rates for the last 15 years and that will, I believe, determine them in the future. Although bookings may fluctuate in the short term due to ordering patterns, we believe that customer acceptance of Maxim products is strong, and our expectations for fiscal 2001 and our future growth remain unchanged."

Proper layout and component selection control EMI

Most portable devices include a regulator or other form of power supply, and the lower supply voltages associated with smaller lithography ICs have mandated these power circuits in many nonportable devices as well. Although not fully understood by many designers, the trade-offs among different types of regulators and power supplies can have a major effect on battery life, compliance with EMI/EMC regulations, and even the basic operation of a product under design. The following overview covers the mechanisms and physical principles governing the generation and propagation of electrical noise in power supplies.

Voltage regulators

The most common power converter is the voltage regulator. It accepts a voltage that varies over a given range and generates an output voltage that does not vary. Regulators comprise two main categories: switching types and all others, mainly the linear and shunt types. Unlike switching regulators, linear and shunt types are limited because their output voltage must remain less than the input voltage. Also, the efficiency of most switching regulators is better than that of an equivalent linear or shunt regulator. Yet, the low noise and simplicity of linear/shunt types makes an attractive alternative to switching regulators.

The simplest type of voltage regulator is a shunt regulator, which adjusts current through a resistor to drop the input voltage to a regulated output level. Zener diodes function similarly, but power dissipation in a zener is high, and its load regulation (change in output voltage with change in load current) is poor. Some shunt regulators let you set the regulation voltage with a voltage divider, but those types usually appear as building blocks in more complex regulators or power supplies. In general, shunt regulators are appropriate for low-power systems in which the variation of load current is small. This narrow application range can be expanded, however, by adding an active pass element (usually a bipolar transistor) that transforms the shunt into a linear regulator.

Linear voltage regulators

Linear voltage regulators use an active pass element (bipolar or MOSFET) to drop the input voltage down to the regulated output voltage. Among these devices, the low-dropout (LDO) types have become popular in the last decade. Dropout refers to the minimum difference (between input and output voltage) that sustains regulation. Dropout voltages as high as 1V have been called LDO, but 100mV to 300mV is more typical.

Because a linear regulator's input current is approximately equal to its output current, its efficiency (output power divided by input power) is a function of the output/input voltage ratio. Thus, dropout is important because lower dropout means higher efficiency. If input voltage is much higher than output voltage, however, or if it varies widely, then the maximum efficiency is difficult to achieve. Another function of LDO regulators (to be discussed) is to serve as a barrier to the noise generated by a switching regulator. In that role, the LDO regulator's low-dropout characteristic improves the circuit's overall efficiency.

Switching regulators

If the performance of a linear or shunt regulator is not adequate for the application, then the designer must turn to a switching regulator. However, along with improved performance come the drawbacks of larger size and cost, greater sensitivity to (and generation of) electrical noise, and a general increase in complexity.

Noise generated by a switching regulator or power supply can emerge through conduction or radiation. Conducted emissions can take the form of voltage or current, and each of these can be further categorized as common-mode or differential-mode conduction. To complicate matters, the finite impedance of connecting wires enables voltage conduction to cause current conduction and vice versa, and differential-mode conduction to cause common-mode conduction and vice versa.

In general, you can optimize a circuit to reduce one or more of these emissions. Conducted emission usually poses a greater problem for fixed systems than for portable systems. Because the portable device operates from batteries, its load and source have no external connections for conducting emissions.

To understand the source of noise in a switching regulator, you must first understand its operation. Descriptions of the many types of switching regulators are beyond the scope of this article. But, generally, a switching regulator converts the source voltage/current to load voltage/current by employing active elements (transistors and diodes) to shuttle current through storage elements (inductors and capacitors). To illustrate, the MAX1653 DC-DC converter controller forms a typical synchronous-rectified step-down converter (**Figure 1**).

During normal operation, the circuit conducts current from input to output when the high-side switch (N1) is on, and continues conducting through the inductor when N1 is off and the synchronous rectifier (N2) is on. First-order approximations of the current and voltage waveforms (**Figure 2**) make a flawed assumption that all the components are ideal.

Because N1 is on only part time, the input source and input capacitor (C_{IN}) see discontinuous currents. C_{IN} supplies the excess current (I_L - I_{INPUT}) while N1 is on, and it stores charge from the input current while N1 is off. If C_{IN} were infinite, with zero equivalent series resistance (ESR) and zero equivalent series inductance (ESL), the voltage across it would remain constant during these partial charge and discharge cycles. Actual voltage fluctuates over each cycle, of course. The current pulses divide between C_{IN} and the input source, based on the relative conductances at or above the converter's switching frequency.

One way to eliminate these conducted emissions is the brute force approach: connect low-impedance bypass capacitors at the input. A more subtle approach, however, can save cost and board area: add impedance between the source and the converter, making sure the necessary DC current can pass. The best impedance is an inductor, but ensure that the converter's input impedance remains low up to its loop crossover frequency. (The loop crossover for most DC-DC switching converters is between 10kHz and 100kHz.) Otherwise, input-voltage fluctuations can destabilize the output voltage.

Current ripple on the output capacitor (C_{OUT}) is much less than on C_{IN}. Its amplitude is lower, and (unlike the input capacitor) its current is continuous and therefore has less harmonic content. Normally, each turn of the coil is covered with wire insulation, forming a small capacitor between each pair of turns. Adding these parasitic capacitors in series forms a small equivalent capacitor in parallel with the inductor, which provides a path for conduction of current impulses to C_{OUT} and the load. Thus, the discontinuous edges of the voltage waveform at the switching node (LX) conduct high-frequency current to C_{OUT} and the load, usually resulting in spikes on the output voltage, with energy in the 20MHz to 50MHz range.

Often the load for this type of converter is some form of microelectronics susceptible to conducted noise, and fortunately, the converter's conducted noise is easier to control at the output than at the input. As for the input, output-conducted noise can be controlled by very-low-impedance bypassing or secondary filtering. Be cautious of secondary (post) filtering, however. Output voltage is a regulated variable in the control loop, so an output filter adds delay or phase (or both) to the loop gain,

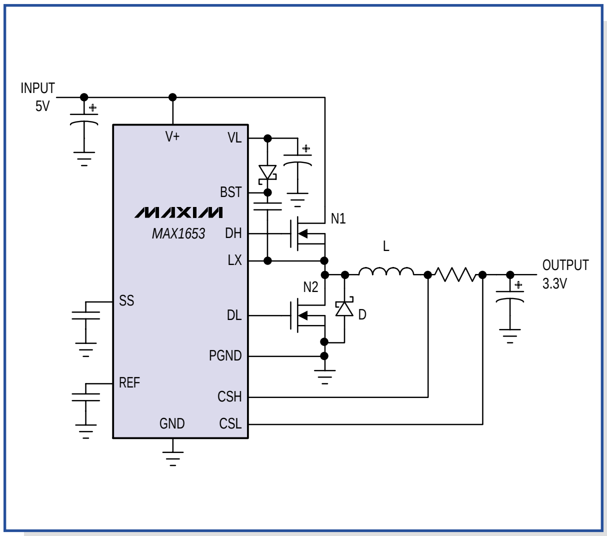

Figure 1. This illustrative step-down switching regulator features an external switching transistor (N1) and synchronous rectifier (N2).

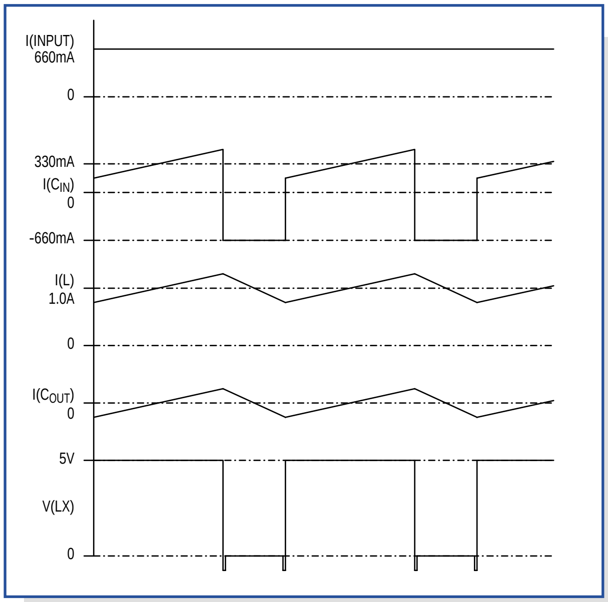

Figure 2. These waveforms from the Figure 1 circuit are based on an assumption of ideal components.

possibly destabilizing the circuit. If a high-Q LC post filter is placed after the feedback point, the inductor's resistance will degrade the load regulation, and transient load currents may cause ringing.

Other topologies

Other switching-converter topologies have problems similar to those of the step-down converter. A step-up converter, for example (**Figure 3**), has the basic structure of a step-down converter but with inputs and outputs swapped. Thus, problems at the input of a step-down converter apply to the output of a step-up converter as well, and vice versa.

Step-down converters are limited because their output voltage must be less than the input voltage. Similarly, a step-up converter's output voltage must be greater than its input voltage. This requirement is problematic when the output voltage falls within the input voltage range. Flyback converter topology solves this problem (**Figure 4**).

Because currents at the input and output are both discontinuous, making conducted emissions more difficult to control, flyback converter noise is generally worse than that of a step-up or step-down type. Another problem with this converter is that current in each transformer winding is discontinuous. These discontinuities act with the transformer's leakage inductance to produce high-frequency spikes, which can conduct to other circuits. Physical separation of the primary and secondary windings causes this leakage inductance. Thus, the leakage inductance is caused by magnetic fields in the air (because fields in the core couple both the primary and secondary windings). Spikes due to the leakage inductance therefore cause magnetic-field radiation.

Single-ended primary inductance converter (SEPIC) topology also solves the problem of overlapping input and output voltages. Similar to a flyback circuit, the SEPIC converter connects a capacitor between the transformer primary and secondary windings (**Figure 5**). This capacitor, which provides a path for current in the primary and secondary windings during the time that flyback currents are off, improves the flyback circuit by making the primary and secondary currents continuous. On the other hand, adding input or output capacitance to a flyback circuit can often improve its emissions sufficiently to make that topology just as acceptable. However, if conducted and radiated noise are expected to be a problem, the SEPIC circuit may be preferable to the flyback.

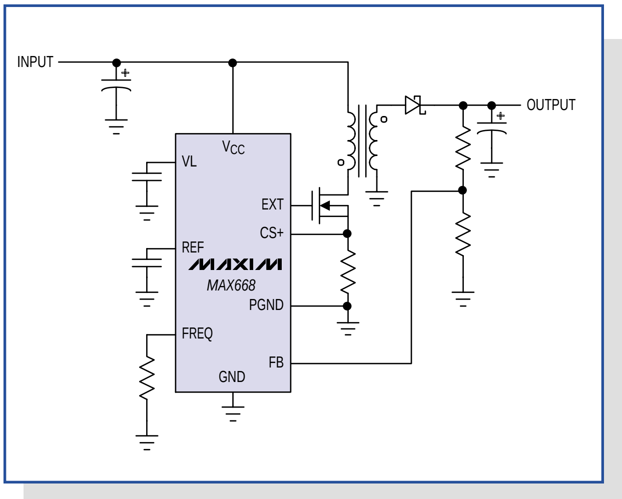

Figure 4. A flyback regulator maintains regulation for inputs that range above and below the output voltage.

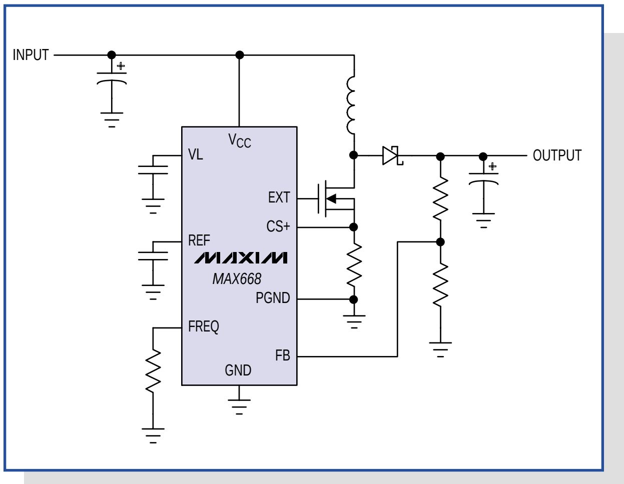

Figure 3. This step-up switching regulator lacks synchronous rectification, but is otherwise similar to the step-down type, with inputs and outputs swapped.

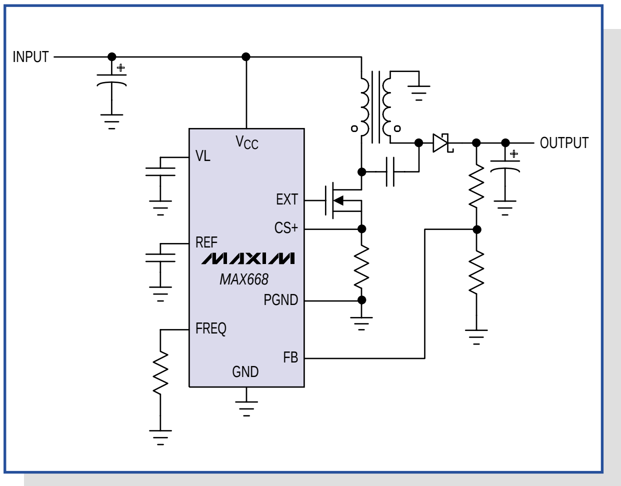

Figure 5. Otherwise similar to a flyback regulator, the SEPIC has continuous primary and secondary currents that generate less noise.

Linear post regulation

For some applications in which output noise must be minimized, the efficiency deficit of using a linear regulator is not acceptable. A switching regulator followed by a linear post regulator may be suitable in these cases. The post regulator attenuates high-frequency noise generated by the switching regulator, resulting in noise performance approaching that of a lone linear regulator. Because most voltage conversion occurs in the switching regulator, however, the efficiency penalty is much smaller than for a lone linear regulator.

This scheme can also replace flyback and SEPIC converters in applications for which the input and output voltages overlap. The step-up converter operates when the input is less than the output, and the linear regulator operates when the input is greater than the output. A step-up converter and LDO linear regulator can be combined in a single IC (**Figure 6**). This device also includes a track mode in which the step-up converter output voltage is always 300mV above the LDO output voltage. As a result, the LDO regulator maintains sufficient power-supply rejection ratio (PSRR) and headroom (input minus output voltage) to attenuate noise from the step-up converter under all conditions.

Common-mode noise

By definition, common-mode conduction is in phase on both connections of the input or output. Typically, it poses a problem only for fixed systems that have a path to earth ground. In a typical off-line power supply with common-mode filters (**Figure 7**), the main source of common-mode noise is the MOSFET. The MOSFET is usually a major power-dissipating element in the circuit and requires a heatsink.

For a TO-220 device, the heatsink tab would connect to the MOSFET drain, and in most cases the heatsink would conduct current to earth ground. Because the MOSFET is both insulated and electrically isolated from the heatsink, it has some capacitance to earth ground. As it switches on and off, the rapidly changing drain voltage drives current through the parasitic capacitance (C_{P1}) to earth ground. Because the AC line has low impedance to earth ground, these common-mode currents flow from the AC input to earth ground. The transformer also conducts high-frequency current through the parasitic capacitance (C_{P2A}, C_{P2B}) between its isolated primary and secondary windings. Thus, noise can be conducted to the output as well as the input.

In Figure 7, the common-mode conducted noise is attenuated by the common-mode lowpass filters between the noise source (the power supply) and the input or output. Common-mode chokes (CML1, CML2) are generally wound on a single core with the polarity shown. Load current and the line current driving the power supply are both differential-mode currents (i.e., current flowing in

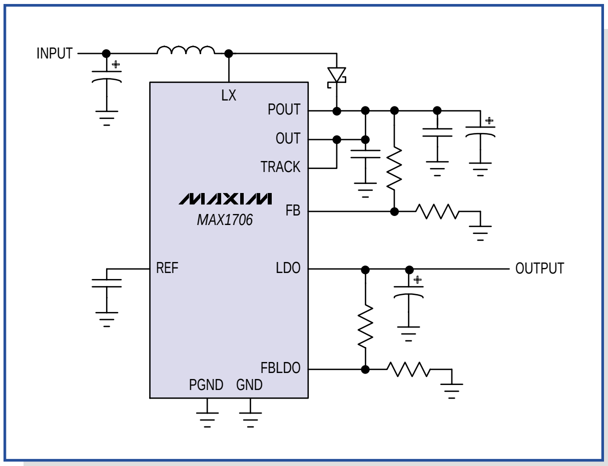

Figure 6. As a third option for maintaining regulation when the input range overlaps the output voltage, this IC combines a switching regulator (for step up) and a linear regulator (for step down).

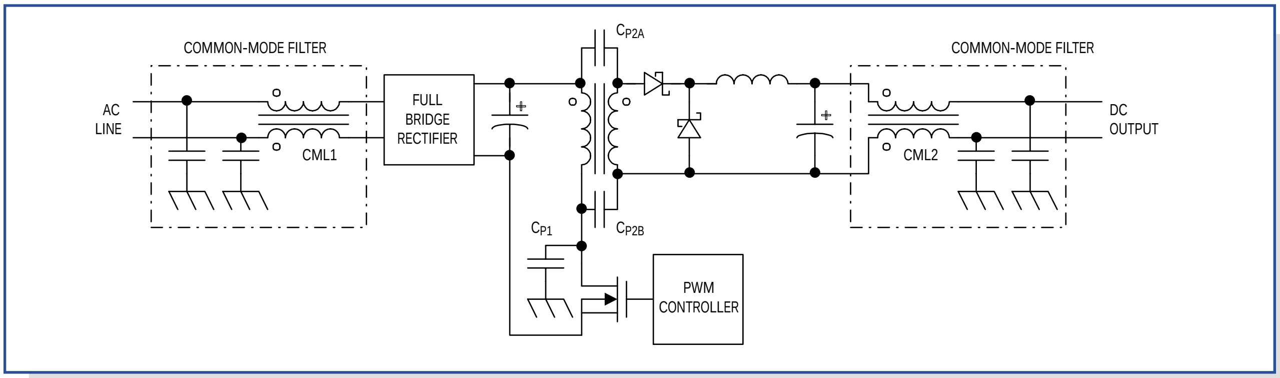

Figure 7. Common-mode filters in this typical off-line power supply reduce noise common to both sides of the input and output.

one line flows out the other). By winding the common-mode chokes on a single core, the fields due to differential-mode currents cancel, allowing use of a smaller core because very little energy is stored in it.

Many of the common-mode chokes designed for off-line power supplies are wound with physical separation between the windings. This construction adds differential-mode inductance, which also helps to reduce the conducted differential-mode noise. Because the core links both windings, fields due to differential-mode current and differential-mode inductance are in the air rather than the core, which may produce radiated emissions.

Common-mode noise generated in the power supply's load may be conducted through the power supply to the AC line through parasitic capacitance (C_{P2A}, C_{P2B}) in the transformer. A Faraday shield in the transformer (a ground plane between primary and secondary) can reduce this noise (**Figure 8**). The shield forms capacitors from the primary and secondary to ground, and these capacitors shunt common-mode currents to ground rather than allowing them to pass through the transformer.

Just as conducted emissions can be in the form of voltage or current, radiated emissions can be in the form of electric or magnetic fields. Because fields exist in space rather than in conductors, however, there is no distinction between differential and common-mode fields. An electric field exists in the space between two potentials, and a magnetic field exists around a current traveling through space. Both fields can exist in a circuit because capacitors store energy in electric fields and inductors/transformers store/couple energy in magnetic fields.

Electric fields

Because an electric field exists between two surfaces or volumes with different potentials, it is relatively easy to contain the electric-field noise generated within a device by surrounding the device with a ground shield. Such shielding is common practice in the construction of CRTs, oscilloscopes, switching power supplies, and other devices with fluctuating high voltages. Another common practice is the use of ground planes on circuit boards. Electric fields are proportional to the potential difference between surfaces and inversely proportional to the distance between them. They exist, for instance, between a source and any nearby ground plane. Multilayer circuit boards, therefore, let you shield circuitry or traces by placing a ground plane between them and any large potential.

However, be cautious of capacitive loading on high-voltage lines when using ground planes. Capacitors store energy in electric fields, so placing the ground plane near a conductor forms a capacitor between the conductor and ground. A large dV/dt signal on the conductor can cause large conducted currents to ground, thereby degrading the conducted emissions while controlling the radiated emissions.

If electric-field emissions are present, the most likely culprit is the highest potential in the system. In power supplies and switching regulators, beware of the switching transistors and rectifiers because they normally have high potentials and may also have large surface areas due to heatsinking. Surface-mount devices may have this problem too, because they often require lots of circuit-board copper for heatsinking. In that case, also beware of capacitance between any large-area heatsink plane and the ground plane or a power-supply plane.

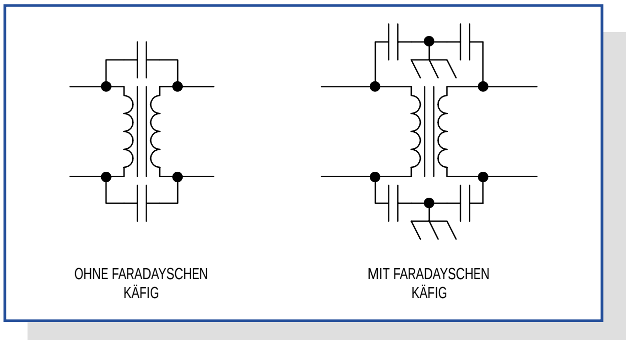

Figure 8. A Faraday shield between primary and secondary blocks common-mode noise that would otherwise pass through the transformer's parasitic interwinding capacitance.

Magnetic fields

Electric fields are relatively easy to contain, but magnetic fields are a different proposition. Enclosing a circuit in high-μ material can provide an effective shield, but that approach is difficult and costly. Usually, the best way to control magnetic-field emissions is to minimize them at the source. In general, this requires that you choose inductors and transformers designed to minimize radiated magnetic fields. Equally important, the circuit board layout and interconnect wiring should be configured to minimize the size of current loops, especially in high-current paths. Not only do high-current loops radiate magnetic fields, they also increase the inductance of conductors, which can cause voltage spikes on lines that carry high-frequency current.

Inductors

Circuit designers inexperienced in transformer or inductor design are likely to choose off-the-shelf transformers and inductors. Still, a bit of knowledge about magnetics can enable a designer to choose the optimal components for an application.

The key to reducing inductor emissions is in the use of high-μ material for keeping the field in the core and out of the surrounding space. Magnetic fields have a proportionally higher density in higher-μ material. This is much like parallel conductances: a 1S conductance (i.e., a 1Ω resistor) in parallel with a 1mS conductor (1kΩ resistor) has 1000 times the current of the 1mS conductor. A magnetic field density divides in a ratio of 1000:1 between a 1000μ, 1in^2 core and a 1μ, 1in^2 core. High-μ materials cannot store a lot of energy, so for compact inductors you must employ a high-μ core with an air gap.

To understand why, see **Figure 9**. The B-field (X-axis) is proportional to V × t/N, where N is the number of turns. The H-field (Y-axis) is proportional to N × i. Thus, the slope of the curve (proportional to μ) is also proportional to the inductance (L = V/[di/dt]). Adding a gap to this ferrite (or any other high-μ core) reduces the slope, thereby lowering the effective μ and consequently the inductance. Inductance decreases by the change in slope, maximum current increases by the change in slope, and the saturation B-field remains the same. Therefore, the maximum energy (1/2 LI^2) stored in the inductor increases. This increase can also be illustrated by applying a voltage to the inductor and noting the amount of time to reach Bsat. Energy stored in the core is the integral of (V × i)dt. Because the current associated with a gapped core is higher for the same voltage and time, the corresponding level of stored energy is higher.

Gapping the core, however, increases magnetic-field radiation in the space around the inductor. Bobbin cores, for example, whose large air gap makes them notorious generators of magnetic-field radiation, are generally avoided in some noise-sensitive applications for that reason. The bobbin core—a bobbin-shaped piece of ferrite—is one of the simplest and cheapest types of gapped ferrite core. Wire is wound around the center post to make an inductor. Costs are low because the wire can be wound directly around the core with no extra work other than terminating the wire. In some cases, the wires are terminated on a metalized area at the bottom of the core, allowing the inductor to be surface mounted. In other surface-mounted components, the inductor is mounted on a ceramic or plastic header to which the wires are terminated.

Some manufacturers put ferrite shields around the bobbin core to help reduce field emissions. This measure helps, but it also reduces the gap and therefore reduces the energy that can be stored in the core. Because the ferrite itself can store very little energy, a small gap is often retained between shield and core, which allows some unwanted radiation of magnetic fields in this type of inductor. Depending on the level of acceptable emissions, however, the bobbin core may be a good compromise between cost and EMI.

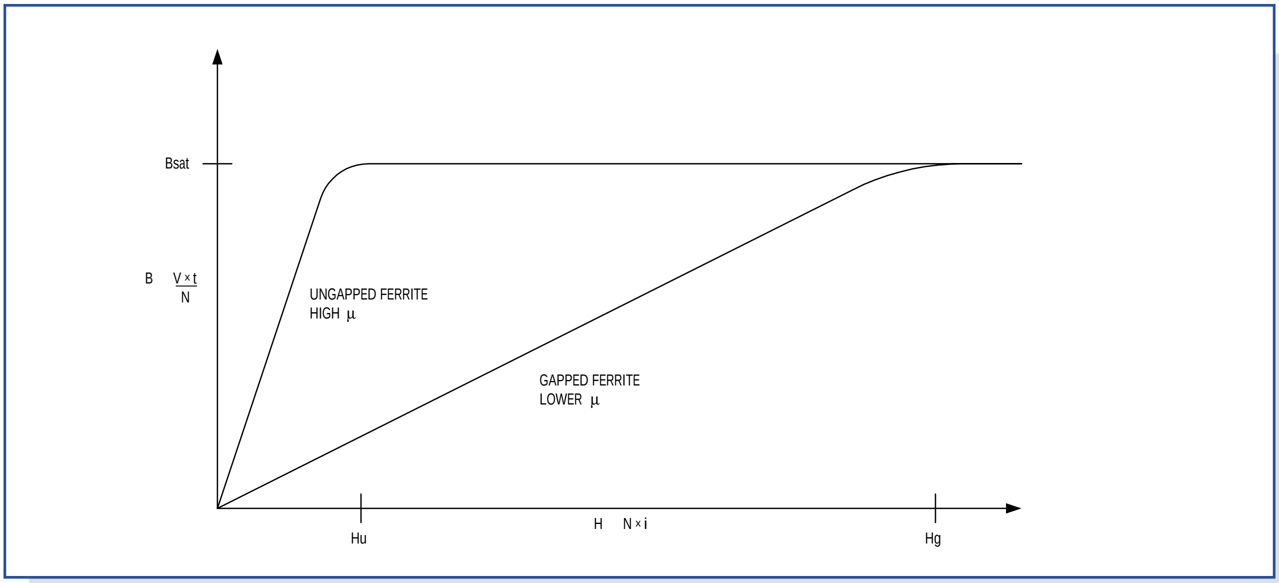

Figure 9. Gapping a ferrite core forces magnetic flux out of the core and allows the inductor or transformer to store energy in a field surrounding the device.

Various other core shapes can be gapped (or not) according to requirements of the application. For example, pot cores, E-I cores, and E-E cores all have center legs or posts that can be gapped (**Figure 10**). Gapping the center of the core, which is completely surrounded by the coil, helps reduce the emissions radiated from the air gap. These inductors are usually more expensive because the coil must be wound separately from the core, and the core part is assembled around the coil. For easy design and assembly, cores can be purchased with a pregapped center leg.

Perhaps the best core for reducing radiated emissions is the distributed-gap toroid. This core is made by pressing a mixture of filler and high-µ metal powder into the doughnut shape of a toroid. The grains of metal powder, separated by nonmagnetic filler, have small air gaps between them that create an overall “air gap” evenly distributed throughout the core. The coil is wound through the center and around the outside of this core, making the field travel in a circle along the middle of the coil. As long as the coil is wound around the whole circumference of the toroid, it shields the outside by completely surrounding the magnetic field.

The loss in a typical distributed-gap toroidal core is sometimes higher than for gapped ferrites because metal grains in the toroid are susceptible to eddy currents that generate heat and reduce power-supply efficiency. Toroids are also expensive to wind because wire must be fed through the center of the core. Machines can do this, but they are slower and more expensive than traditional coil-winding machines.

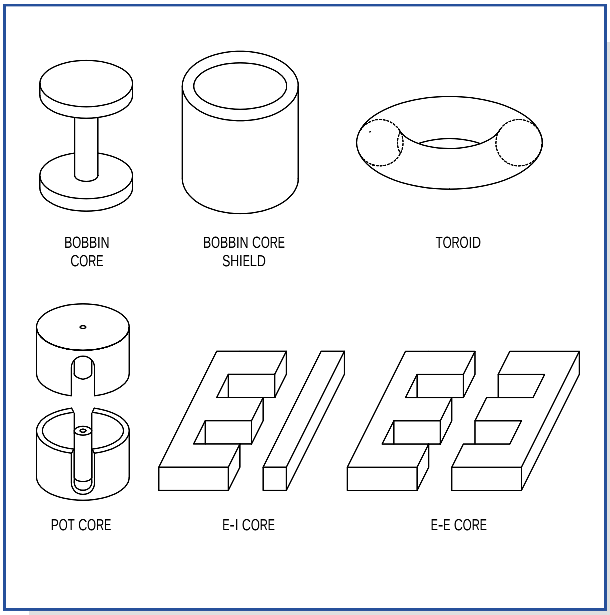

Figure 10. Different core geometries offer trade-offs among energy storage, field emission, and ease of assembly. All can be gapped.

Some ferrite toroidal cores have a discrete air gap. The resulting magnetic-field emissions are higher than those of distributed-gap cores, but typical gapped toroids have lower losses because they contain the field better than other discretely gapped ferrite cores. The coil reduces emission by shielding the gap, and the toroidal shape helps to keep the field inside the core.

Transformers

Transformers have many limitations in common with inductors because they are wound on the same cores. Still, some issues are unique to transformers. The performance of actual transformers can approach that of an ideal transformer—coupling voltage from primary to secondary with a ratio of voltages proportional to the ratio of turns in each winding.

An equivalent transformer circuit (**Figure 11**) models the interwinding capacitance as C_{WA} and C_{WB}. The problem posed by these parameters is mainly that of common-mode emissions in isolated power supplies. Winding capacitances C_P and C_S are small and usually negligible at the operating frequencies of switching power supplies and regulators. Magnetizing inductance L_M is important because too much magnetizing current can cause the transformer to saturate. As for inductors, saturation increases the magnetic-field emission from transformers. Saturation also causes higher core loss, higher temperature (with the possibility of thermal runaway), and a degradation of coupling between the windings.

Leakage inductance is caused by a magnetic field that links one winding but not the other. Although some

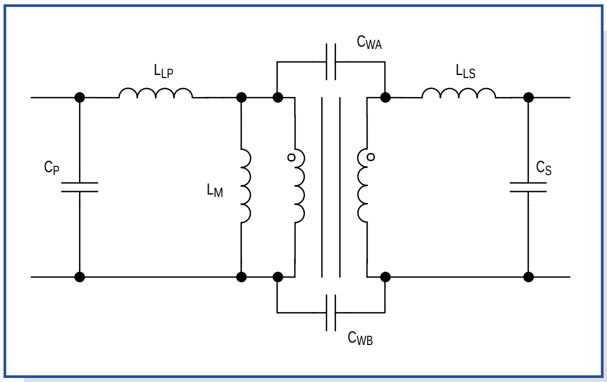

Figure 11. Parasitic elements in the equivalent model for a transformer modify its ideal behavior.

coupled inductors and transformers (like the common-mode choke discussed earlier) are designed for a high level of this parameter, leakage inductances L_{LP} and L_{LS} are the most problematic parasitic elements in a switching power supply. Magnetic flux that links two windings couples those windings together. All transformer windings are around the core, so any leakage inductance is outside the core, in the air, where it can cause magnetic field emissions.

Another problem with leakage inductance is the large voltage generated when the current changes quickly, as it does in the transformer of most switching power supplies. Such voltage can overstress the switching transistor or rectifier. Dissipative snubbers (usually a series resistor and capacitor) are often used to control this voltage by dissipating the energy of the voltage spike. On the other hand, some switching devices are designed to withstand repetitive avalanche breakdowns and can dissipate the energy without external snubbers.

The leakage inductance of a transformer can be determined by shorting the secondary and measuring the inductance at the primary. This measurement includes any secondary leakage inductance coupled through the transformer, but in most cases, such leakage must be accounted for anyway because it adds to the primary voltage spike. The corresponding spike energy is calculated as $E = 1/2\ LI^2$, so power lost to the leakage inductance is the energy of each spike multiplied by the switching frequency: $P = 1/2\ LI^2 f$.

Transformer requirements depend on the power-supply topology. Topologies that directly couple energy across the transformer—such as half-bridge, full-bridge, push-pull, or forward converters—require a very high magnetizing inductance to prevent saturation. The transformer primary and secondary simultaneously conduct current in these circuits, directly coupling the energy through the transformer. Because little energy is stored in the core, the transformer can be smaller. These transformers are typically wound on an ungapped core of ferrite or other high-µ material.

Other power-supply topologies require that the transformer core store energy. The transformer in a flyback circuit stores energy through the primary in the first half of the switching cycle. In the second half of the cycle, energy is retrieved and fed to the output through the secondary. As is true for inductors, an ungapped high-µ core is not suitable for storing energy in a transformer. Instead, the core must be discretely gapped or have a distributed gap. The resulting component will be larger than one with an equivalent ungapped core, but it eliminates the need for an extra inductor, which can save cost and space.

Layout

Component selection is very important in controlling EMI, but the circuit-board layout and interconnects are equally important. Especially for the high-density, multilayer circuit boards often used in switching power supplies, layout and component placement are critical to the circuit's proper operation and interaction. The power switching can cause large dV/dt and di/dt signals in the circuit board traces, which cause compatibility problems by coupling to other traces. Compatibility problems and expensive circuit-board revisions can be avoided, however, by taking extra care in the layout of critical paths.

A distinction can be made between radiated and conducted emissions in a system, but the distinction blurs when talking about interference in a circuit board and wiring. Adjacent traces that couple electric fields also conduct currents through parasitic capacitance. Likewise, traces that are coupled by magnetic fields act somewhat like transformers. These interactions can be described in terms of lumped components or through field theory. Which approach to take depends on which method more accurately describes the interaction.

Crosstalk

Two or more conductors in close proximity are capacitively coupled, so large voltage changes on one will couple currents to the other. If the conductor's impedance is low, the coupled currents generate only small voltages. Capacitance is inversely proportional to the distance between the conductors and proportional to the area of the conductors, so the conducted noise can be minimized by keeping the area of the adjacent conductors small, and their separation large.

Another method for reducing the coupling between conductors is to add a ground plane or shield. A ground trace (or in some cases a power-supply bus or other low-impedance node) between conductors can prevent their interaction by capacitively coupling them to ground instead of to each other. But exercise caution. Traces carrying fast dV/dt changes, positioned close to a plane with high-impedance interconnect to ground, can couple these changes to the ground plane. In turn, the ground plane can couple the signals to sensitive lines, thereby exacerbating rather than helping the noise problem. If the ground plane doesn't carry large currents, it may be tempting to connect it to ground through a small wire. The high inductance of a small wire, however, can cause the ground plane to look like a high impedance to fast-changing voltages.

You must ensure that a ground plane does not inject noise into sensitive parts of a circuit. Input and output bypass capacitors, for example, often pass current through a ground plane, and the high-frequency current components can affect sensitive circuitry. To prevent this problem, circuit boards often include separate planes for power and signal grounds. Connected at a single point, these planes minimize the noise injected into signal ground by potentials generated across the power ground plane. This practice is similar to that of a star ground in which all components connect to ground at a single point (all traces leave that point in a “star” pattern). The star ground has the same effect as separate power- and signal-ground planes, but it isn’t practical for large, complex circuit boards that include many grounded components.

If a node is known to be sensitive to injected noise, then traces and wires connected to that node should be routed away from nodes with high-voltage changes. If that isn’t possible, add a good ground or a shield. Good capacitive bypassing of the node can also decrease its susceptibility to crosstalk. Normally, a small capacitor connected between the node and ground or between the node and a power-supply bus forms a suitable bypass.

When choosing the bypass capacitor, make sure it has a low impedance over the range of frequencies that are potentially problematic. ESR and ESL may cause the impedance to be higher than expected at high frequencies, so the low ESR and ESL of ceramic capacitors is attractive for bypass applications. The ceramic dielectric has a large effect on performance as well. Higher capacitance dielectrics (such as Y5V) may allow large changes in capacitance over voltage and temperature. At maximum-rated voltage, capacitors made with these ceramics may exhibit as little as 15% of their unbiased capacitance. A smaller capacitance value with a better dielectric, yielding crosstalk attenuation that is less dependent on bias and temperature, will in many cases provide better and more consistent bypassing.

The placement of bypass capacitors is also critical. To attenuate high-frequency noise, you must route the signals in question through the bypass capacitor. In **Figure 12a**, the length of trace in series with the capacitor adds to its ESR and ESL, increasing the impedance at high frequencies and reducing the capacitor’s effectiveness as a high-frequency bypass. A better layout (**Figure 12b**) routes the trace through the capacitor, so the traces’ stray ESR and ESL aid the bypass capacitor’s filter action rather than degrading it.

Some nodes should not be bypassed because doing so changes their frequency characteristics. An example is the feedback resistor-divider. In most switching power supplies, a resistive feedback divider drops the output voltage down to a level acceptable to the error amplifier. A large bypass capacitor added to this feedback node forms a pole with the resistance of that node. Because the divider is part of the control loop, this pole becomes part of the loop characteristic. If the pole frequency is less than one decade above the crossover frequency, its phase or gain effects can adversely affect the loop stability.

Inductance

Some currents in a switching power supply switch on and off quickly. Stray inductance in those current paths may induce large noise voltages that couple into sensitive circuitry and stress the components. Lines carrying DC currents seldom cause problems because DC does not cause voltage spikes or couple AC to other traces. For instance, a line in series with an inductor is not a problem because the stray inductance is much smaller than the inductor value. The large series inductance also prevents discontinuities in the current.

If a circuit produces discontinuous currents, try to prevent the current from traveling in large loops. Large loops of current produce larger values of inductance, thereby increasing any consequent magnetic field radiation. This caveat applies to component placements

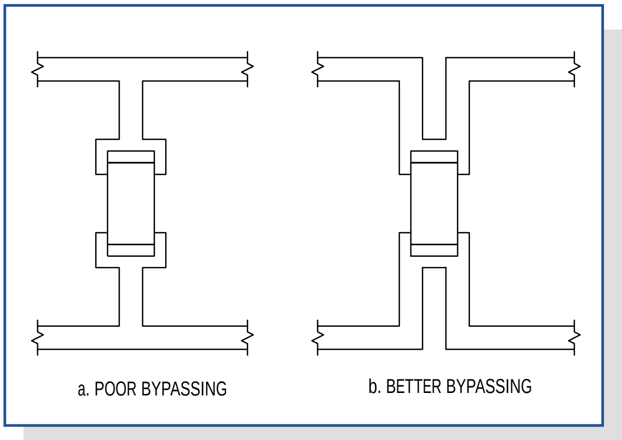

Figure 12. Poor bypass connections (a) add trace inductance and resistance to the capacitor. In the better connection (b), trace parasitics add to the capacitor’s filter action.

as well because current usually switches between active devices such as transistors and diodes.

Consider the step-down converter in Figure 1. When the high-side MOSFET switch (N1) is on, current travels through the input, N1, the inductor, and the load. After N1 turns off, the diode (D) conducts current until the synchronous rectifier (N2) turns on. Current then flows through N2 until it turns off, then, again, the diode carries current until the cycle restarts. Note that currents through the inductor and output capacitor are continuous and therefore should not be major contributors to noise. If N1, N2, and D are placed some distance from each other, the surrounding fields must shift quickly in response to the rapid current changes within them. Because the voltage generated is proportional to the change in magnetic field with time ($d\Psi/dt$), these rapid field fluctuations can generate large voltage spikes.

Note that the input source and output load carry high-frequency currents. These currents should pass through the input and output bypass capacitors; otherwise, they are conducted through the input or output lines, or both (see the *Common-mode noise* section). The impedance of input and output bypass capacitors is important. They should be large enough to keep the impedance low at input and output, but larger capacitors (tantalum or electrolytic types, for instance) have higher ESR and ESL than smaller ceramic types. So, you must ensure that the capacitors' impedance is sufficiently low at the frequencies of concern.

One alternative is to parallel a ceramic capacitor with an electrolytic or tantalum capacitor because the ceramic has lower impedance at high frequencies. In most cases, however, that arrangement is no better than multiple electrolytic or tantalum capacitors in parallel to reduce ESR and ESL, or multiple ceramic capacitors in parallel to increase the total capacitance.

Measuring INL and DNL for high-speed ADCs

Manufacturers have recently introduced high-performance analog-to-digital converters (ADCs) that feature outstanding static and dynamic performance. You might ask, "How do they measure this performance, and what equipment is used?" The following discussion lends insight on techniques for testing two of the accuracy parameters important for ADCs: integral nonlinearity (INL) and differential nonlinearity (DNL).

INL and DNL, although not the most important electrical characteristics that specify the high-performance data converters used in communications and fast data-acquisition applications, gain significance in the higher-resolution imaging applications. Unless you regularly work with ADCs, you can easily forget the exact definitions and importance of these parameters. The following section serves as a brief reminder.

INL and DNL definitions

DNL error is defined as the difference between an actual step width and the ideal value of one LSB (**Figure 1a**). For an ideal ADC, in which the differential nonlinearity coincides with DNL = 0LSB, each analog step equals 1LSB (1LSB = $V_{FSR}/2^N$, where V_{FSR} is the full-scale range, and N is the resolution of the ADC), and the transition values are spaced exactly 1LSB apart. A DNL error specification of ≤1LSB guarantees a monotonic transfer function with no missing codes. An ADC's monotonicity is guaranteed when its digital output increases (or remains constant) with an increasing input signal, thereby avoiding sign changes in the slope of the transfer curve. DNL is specified after the static gain error has been removed. Its definition is:

$$DNL = \left| [(V_{D+1} - V_D) / V_{LSB\text{-}IDEAL} - 1] \right|$$

where $0 < D < 2^N - 2$.

V_D is the physical value corresponding to the digital output code D, N is the ADC resolution, and $V_{LSB\text{-}IDEAL}$ is the ideal spacing for two adjacent digital codes. By adding noise and spurious components beyond the effects of quantization, higher DNL values usually limit the ADC's dynamic performance in terms of signal-to-noise ratio (SNR) and spurious-free dynamic range (SFDR).

INL error is described as the deviation, in LSB or percent of full-scale range (FSR), of an actual transfer function from a straight line. The INL-error magnitude then depends directly on the position chosen for this straight line. Two definitions are commonly used: "best-straight-line INL" and "endpoint INL" (**Figure 1b**):

- Best-straight-line INL provides information about offset (intercept) and gain (slope) error, plus the position of the transfer function. It determines, in the form of a straight line, the closest approximation to the ADC's actual transfer function. Although the line's exact position is not clearly defined, this approach yields the best repeatability, and it serves as a true representation of linearity.
- Endpoint INL passes the straight line through endpoints of the converter's transfer function, thereby defining a precise position for the line. Thus, the straight line for an N-bit ADC is defined by its zero- (all zeros) and full-scale (all ones) outputs.

The best-straight-line approach is generally preferred because it produces better results. INL is measured after both static offset and gain errors have been nullified and can be described as follows:

$$INL = \left| [(V_D - V_{ZERO}) / V_{LSB\text{-}IDEAL}] - D \right|$$

where $0 < D < 2^N - 1$.

V_D is the analog value represented by the digital output code D, N is the ADC's resolution, V_{ZERO} is the minimum analog input corresponding to an all-zero output code, and $V_{LSB\text{-}IDEAL}$ is the ideal spacing for two adjacent output codes.

Generic setup for testing static INL and DNL

INL and DNL can be measured with either a quasi-DC voltage ramp or a low-frequency sine wave as the input. A simple DC (ramp) test may incorporate a logic analyzer, a high-accuracy DAC (optional), a high-precision DC source for sweeping the input range of the device under test (DUT), and a control interface to a nearby PC or X-Y plotter.

If the setup includes a high-accuracy DAC (much higher than that of the DUT), the logic analyzer can monitor offset and gain errors by processing the ADC's output data directly. The precision signal source creates test voltages for the DUT by sweeping slowly through the input range of the ADC from zero-scale to full-scale. Once reconstructed by the DAC, each test voltage at the ADC input is subtracted from its corresponding DC level at the DAC output, producing a small voltage

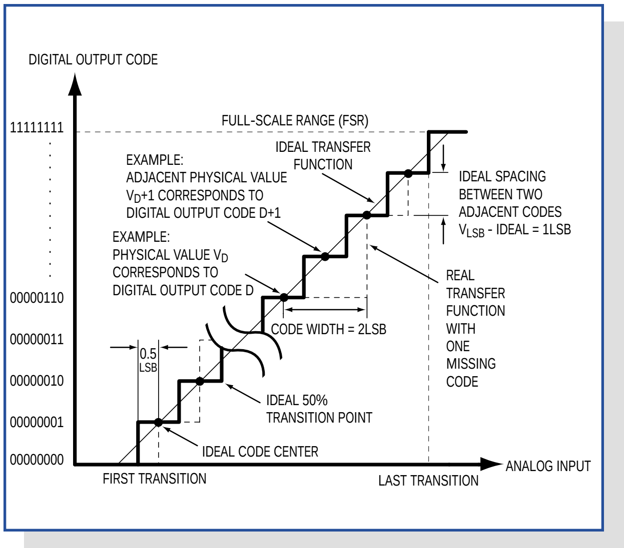

Figure 1a. To guarantee no missing codes and a monotonic transfer function, an ADC's DNL must be ≤ 1LSB.

The right transition

A transition voltage is defined as the input voltage that has equal probabilities of generating either of the two adjacent codes. The nominal analog value—corresponding to the digital output code generated by an analog input in the range between a pair of adjacent transitions—is defined as the midpoint (50% point) of this range. If the limits of the transition interval are known, this 50% point is easily calculated. The transition point can be determined at test by measuring the transition interval limits and then dividing the interval by the number of times each of the adjacent codes appears within it.

difference (V_{DIFF}) that can be displayed with an X-Y plotter and linked to the INL and DNL errors. A change in quantization level indicates differential nonlinearity, and a deviation of V_{DIFF} from zero indicates the presence of integral nonlinearity.

Integrating analog servo loop

Another way to determine static linearity parameters for an ADC, similar to the preceding but more sophisticated, is an integrating analog servo loop. This method is usually reserved for test setups that focus on high-precision measurements rather than speed.

A typical analog servo loop (**Figure 2**) consists of an integrator and two current sources connected to the ADC input. One source forces a current into the integrator, and the other serves as a current sink. A digital magnitude comparator connected to the ADC output controls both

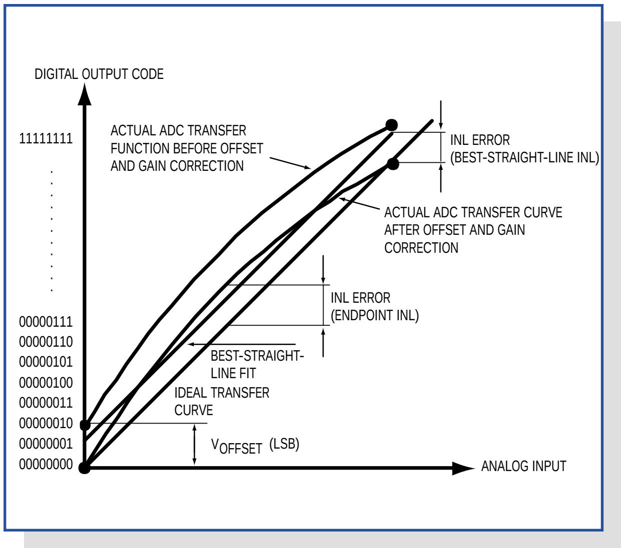

Figure 1b. Best-straight-line and endpoint fit are two possible ways to define the linearity characteristic of an ADC.

Transfer function

The transfer function for an ideal ADC is a staircase in which each tread represents a particular digital output code and each riser represents a transition between adjacent codes. The input voltages corresponding to these transitions must be located to specify many of an ADC's performance parameters. This chore can be complicated, especially for the noisy transitions found in high-speed converters, and particularly for digital codes that are near the final result and changing slowly.

Transitions are not sharply defined as shown in Figure 1b, but are more realistically presented as a probability function. As the slowly increasing input voltage passes through a transition, the ADC converts more and more often to the next adjacent code. By definition, the transition corresponds to that input voltage for which the ADC converts with equal probability to each of the flanking codes.

current sources. The other input of the magnitude comparator is controlled by a PC, which sweeps it through the 2^N - 1 test codes for an N-bit converter.

If the polarity of feedback around the loop is correct, the magnitude comparator causes the current sources to "servo" the analog input around a given code transition. Ideally, this action produces a small triangle wave at the analog inputs. The magnitude comparator controls both rate and direction for these ramps. The integrator's ramp rate must be fast when approaching a transition, yet sufficiently slow to minimize peak excursions of the superimposed triangular wave when measuring with a precision digital voltmeter (DVM).

For INL/DNL tests on the MAX108, the servo-loop board connects to the evaluation board through two headers (**Figure 3**). One header establishes a connection between the MAX108's primary (or auxiliary) output port and the magnitude comparator's latchable input port (P). The second header ensures a connection between the servo loop (the magnitude comparator's Q port) and a computer-generated digital reference code.

The fully decoded decision resulting from this comparison is available at the comparator output P > QOUT and is then passed on to the integrator configurations. Each comparator result controls the logic input of the switch independently and generates voltage ramps as required to drive succeeding integrator circuits for both DUT inputs. This approach has advantages, but it also has several drawbacks:

- The triangular ramp should have low dV/dt to minimize noise. This condition generates repeatable numbers but results in long integration times for the precision meter.
- Positive- and negative-ramp rates must be matched to arrive at the 50% point, and the low-level triangular waves must be averaged to achieve the desired DC level.
- Integrator designs usually require careful selection of the charge capacitors. To minimize potential errors due to the capacitors' "memory effect," integrator capacitors with low dielectric absorption are recommended.
- Accuracy is proportional to the integration period and inversely proportional to the settling time.

A DVM connected to the analog integrated servo loop measures the INL/DNL error vs. digital output code

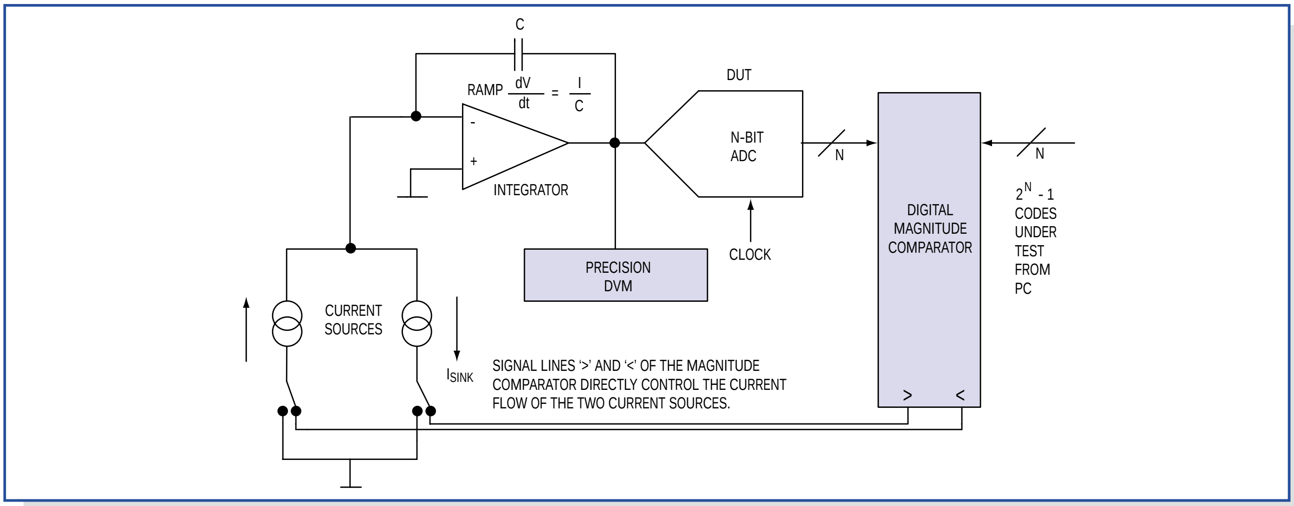

Figure 2. This circuit configuration is an analog integrating servo loop.

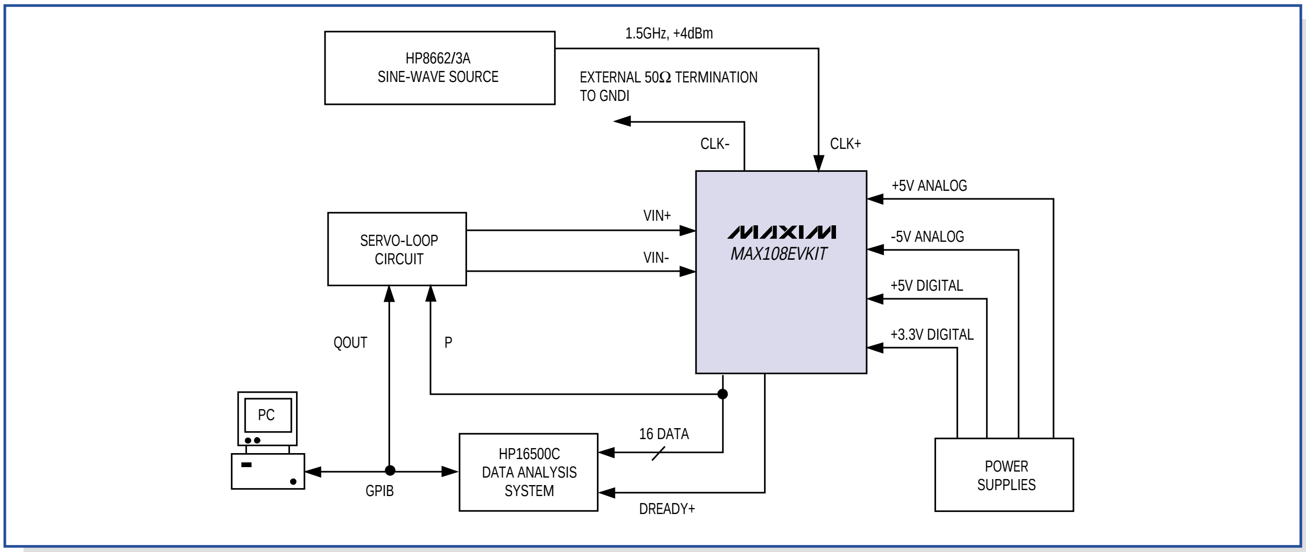

Figure 3. With the aid of the MAX108EVKIT and an analog integrating servo loop, this test setup determines the MAX108's INL and DNL characteristics.

(**Figures 4a, 4b**). Note that a parabolic or bow shape in the plot of INL vs. digital output code indicates the predominance of even-order harmonics, and an "S" shape indicates the predominance of odd-order harmonics.

To eliminate negative effects in the previous approach, you can replace the servo loop's integrator section with an L-bit successive-approximation register (SAR) that captures the DUT's digital output codes, an L-bit DAC, and a simple averaging circuit. Combined with the magnitude comparator, this circuit forms a SAR-type converter configuration (**Figure 5**) in which the magnitude comparator programs the DAC, reads its outputs, and performs a successive approximation. Meanwhile, the DAC presents a high-resolution DC level to the input of the N-bit ADC under test. In this case, a 16-bit DAC was chosen to trim the ADC to 1/8LSB accuracy and obtain the best possible transfer curve.

The advantage of an averaging circuit is apparent when noise causes the magnitude comparator to toggle and become unstable, as it does on approaching its final result. Two divide-by counters are included in the averaging circuit. The "reference" counter has a period of 2M clock cycles, where M is a programmable integer governing the period (and hence the test time). A "data"counter, which increments only when the magnitude comparator output is high, has a period equal to one half of the first 2M - 1 cycles.

The data counter executes a count only when the magnitude counter's output is high. Together, the reference and data counters average the number of highs and lows, store the result in a flip-flop, and pass it on to the SAR register. This procedure is repeated 16 times (in this case) to generate the complete digital output code word. Like the previous method, this one has advantages and disadvantages:

- The test setup's input voltage is defined digitally, allowing easy modification of the number of samples over which the result is to be averaged.
- The SAR approach provides a DC level rather than a ramp at the DUT's analog input.
- As a disadvantage, the DAC in the feedback loop sets a finite limit on the input voltage resolution.

Dynamic testing of INL and DNL

To assess an ADC's dynamic nonlinearity, you can apply a full-scale sinusoidal input and measure the converter's SNR over its entire full-power input bandwidth. The theoretical SNR for an ideal N-bit converter (subject only to quantization noise, with no distortion) is:

$$SNR_{dB} = N \times 6.02 + 1.76$$

Embedded in this figure of merit are the effects of glitches, integral nonlinearity, and sampling-time uncertainty. You can obtain additional linearity information by performing the SNR measurement at a constant frequency and as a function of the signal amplitude. Sweeping the entire amplitude range, for example, from zero to full scale and vice versa, produces large deviations from the source signal as source amplitude approaches the converter's full-scale limit. To determine the cause of

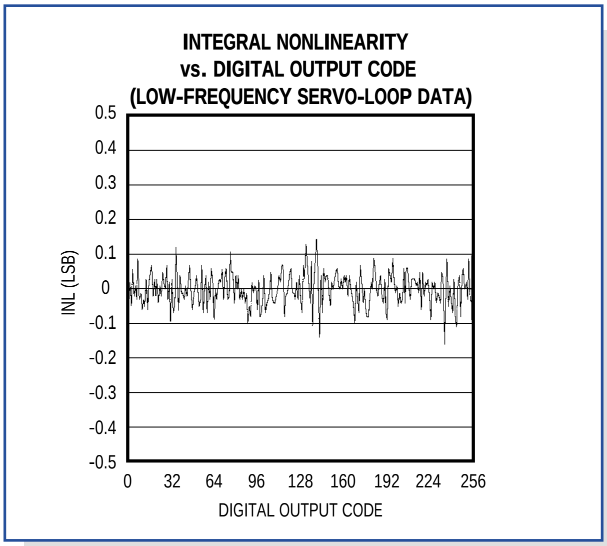

Figure 4a. This plot shows typical INL for the MAX108 ADC, captured with the analog integrating servo loop.

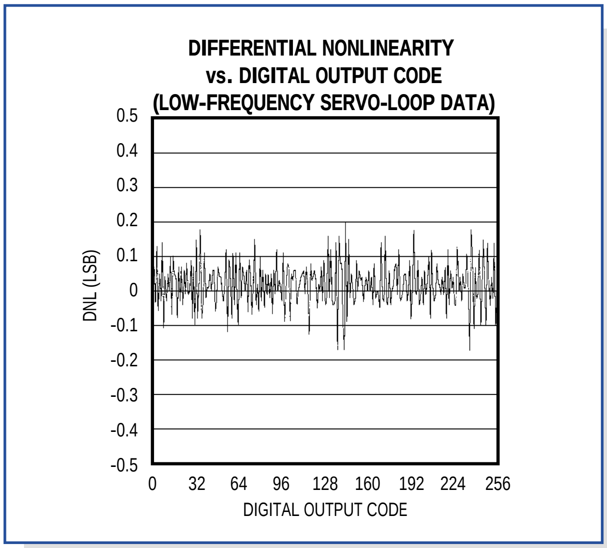

Figure 4b. This plot shows typical DNL for the MAX108, captured with the analog integrating servo loop.

these deviations—while ruling out the effects of distortion and clock instability—use a spectrum analyzer to analyze the quantization error as a function of frequency.

Countless other approaches are available for testing the static and dynamic INL and DNL of both high- and low-speed data converters. The intent herein is to give interested readers insight on the generation of powerful typical operating characteristics (TOCs), using tools and techniques that are simple, but smart and precise.

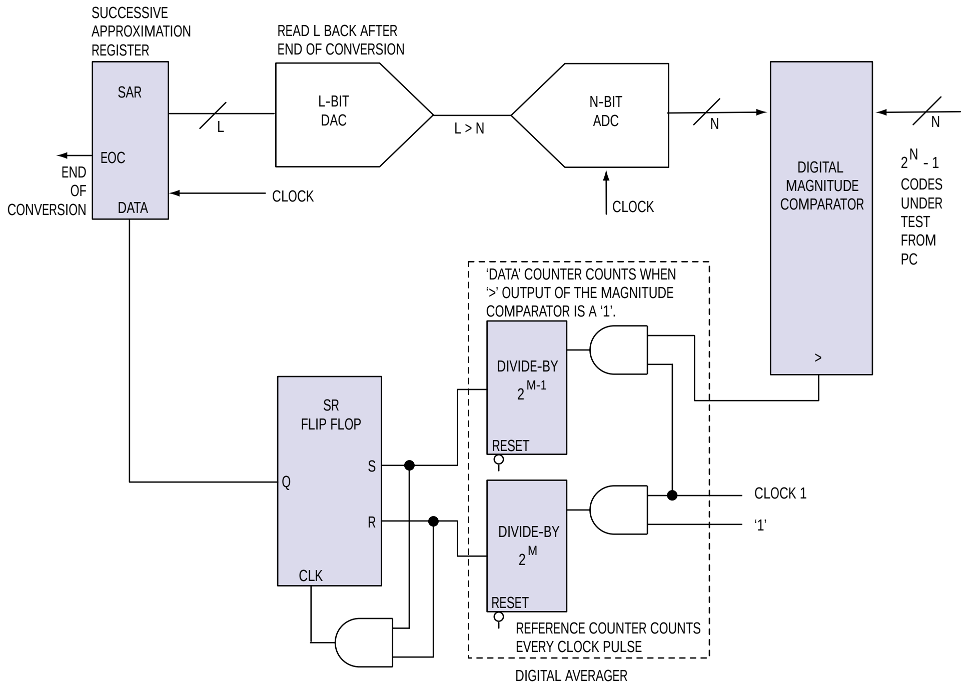

Figure 5. Successive approximation and a DAC configuration replace the integrator section of the analog servo loop.

SAR converter

A SAR converter works like the old-fashioned chemist's balance. On one side is the unknown input sample, and on the other side is the first weight generated by the SAR/DAC configuration (the most significant bit, which equals half of the full-scale output). If the unknown weight is larger than 1/2FSR, this first weight remains on the balance and is augmented by 1/4FSR. If the unknown is smaller, the weight is removed and replaced by a weight of 1/4FSR.

The SAR converter then determines the desired output code by repeating this procedure N times, progressing from MSB to LSB. N is the resolution of the DAC in the SAR configuration, and each weight represents one binary bit.

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NEW PRODUCTS

10-bit, 80Msps, low-power ADC delivers 73dBc SFDR at Nyquist

A new family of low-power (3.0V) 10-bit ADCs is capable of digitizing wide-bandwidth inputs to 80Msps (MAX1448), 60Msps (MAX1446), and 40Msps (MAX1444). Unlike comparable ADCs, the MAX1448 achieves a full 58.5dB SNR, low -69dBc THD, and 73dBc SFDR at the Nyquist input frequency (40MHz), while consuming only 120mW (typ).

Operating from a 2.7V to 3.6V analog supply, these converters target high-speed communication, imaging, and instrumentation applications that require wide bandwidth, good linearity, and excellent dynamic performance at low power. To provide a range of interface options, the separate digital-supply terminal allows output operation from 1.7V to 3.6V. Three-state outputs present data in the straight offset-binary format.

The MAX1448's innovative, differential pipelined architecture allows a 2Vp-p input voltage and supports both differential and single-ended input configurations. Its fully differential input T/H amplifier minimizes the need for external components while providing a full-power bandwidth >400MHz, and its low input capacitance (5pF) enables excellent dynamic performance. The MAX1448 also includes a precision 2.048V bandgap reference that can be overdriven with an external reference voltage—a flexible arrangement that automatically ensures the correct DC-bias levels for AC-coupled applications.

Pin-compatible, lower-speed versions with the same excellent specifications are also available: the 60Msps, 90mW MAX1446, and the 40Msps, 57mW MAX1444. A shutdown mode lowers supply current to 5µA during idle periods. The MAX1448 comes in a space-saving 32-pin TQFP package specified for the extended temperature range (-40°C to +85°C). Prices start at $8.35 (1000–up, FOB USA). An evaluation kit including the MAX1448 is available for $119.50.

3.0V to 5.5V, 2.5Gbps transimpedance amp with +0dB overload

The MAX3864, a 3V to 5.5V transimpedance amplifier with +0dBm overload, is ideal for short-to-long reach SDH/SONET applications that require -24dBm to +0dBm optical dynamic range. The MAX3864's 30mil x 50mil die size and 110mW power dissipation make it ideal for applications in a TO-56 header. With a 2.5Ω transimpedance gain over 1.9GHz of bandwidth, it makes an excellent choice for 2.488Gbps SDH/SONET extended-temperature applications.

The MAX3864 is available as dice and in an 8-pin SO package specified for the extended temperature range. Prices start at $7.95 for the MAX3864E/D and $8.95 for the MAX3864ESA (1000–up, FOB USA). An evaluation kit is available to minimize design time.

Low-cost, low-power DAC families in SOT23

The MAX5360–MAX5365 and MAX5380–MAX5385 families are ultra-low-power, 6- and 8-bit voltage-output DACs. Available in miniature SOT23 packages, they provide a simple 2-wire (MAX5360/61/62 and MAX5380/81/82) (or 3-wire, MAX5363/64/65 and MAX5383/84/85) serial interface that handles clock rates up to 400kHz (or 10MHz). They operate from a single +2.7V to +5.5V supply, offer Rail-to-Rail® precision internal amplifiers, and draw ultra-low supply currents (130µA typ). A 1µA shutdown mode further reduces power consumption. Small size and low power suit the devices for use in portable and battery-powered equipment. A fast-mode, I²C-compatible serial interface (MAX5360/61/62 and MAX5380/81/82) reduces interconnect complexity when communicating with multiple devices.

Internal to each chip is a current-steering DAC, a class AB buffer amplifier capable of driving resistive and capacitive loads, an internal bandgap reference (MAX5360/61/63/64 and MAX5380/81/83/84 only), and a proprietary deglitch circuit that eliminates power-up glitches at the DAC output. An internal power-on reset circuit ensures that outputs power up to 0V and remain there until a valid write takes place. The MAX5362/63 and MAX5382/85 reference is derived from the power-supply inputs, allowing the devices to provide the widest dynamic output range (0V to V_{SUPPLY}). The MAX5360/61/63/64 and MAX5380/81/83/84 provide supply-independent output levels and high PSRR specifications without an external reference. Maximum linearity is specified at ±1LSB for all devices.

Prices (1000–up, FOB USA) start at $0.65 for the MAX5360–MAX5365 and $0.95 for the MAX5380–MAX5385.

Rail-to-Rail is a registered trademark of Nippon Motorola, Ltd.

I²C is a trademark of Philips Corp.

Low-RON analog switches optimized for ±5V applications

The MAX4675–MAX4679 single-pole/single-throw (SPST) CMOS analog switches are designed for precision signal switching and routing in systems with bipolar ±5V supplies. Applications include xDSL modems, ATE, and data-acquisition systems. The single-switch MAX4675/MAX4676 have 4Ω (max) R_{ON}, and the quad-switch MAX4677/MAX4678/MAX4679 have 2Ω (max) R_{ON}. All operate from dual ±2.7V to ±5.5V supplies or from a single +2.7V to +5.5V supply, making them suitable for use in portable equipment.

The MAX4675/MAX4676 are available in a 6-pin SOT23 package specified for the extended temperature range. Prices start at $0.65 (2500–up, FOB USA). The MAX4677/MAX4678/MAX4679 come in a 16-pin TSSOP package, also specified for the extended range. Prices start at $2.15 (1000–up, FOB USA).

NEW PRODUCTS

Clickless/popless stereo headphone drivers achieve >90dB PSRR at 22kHz

The MAX4298 speaker/headphone driver amplifier is designed for use in harsh environments where board space is at a premium and the digital supply is noisy. Patented design techniques achieve an ultra-high power-supply rejection ratio (PSRR) across the audio signal band while delivering high, rail-to-rail output current. The device drives the highly capacitive loads encountered when driving long cables terminated by remote loads, such as notebook headphones. The MAX4298 is fully compliant with PC '99 standards.

The MAX4298 achieves >90dB PSRR at 22kHz, and drives a 1.5V_{RMS} signal into a 10kΩ load with 0.0008% THD+N and 1.2V_{RMS} into 32Ω headphones with only 0.02% distortion. The MAX4298 operates from a single +4.5V to +5.5V supply. Available in space-saving 10-pin µMAX and 14-pin SO packages, the MAX4298 features the best available clickless/popless power-up, power-down, mute, and unmute capability. Prices start at $0.85 (1000–up, FOB USA).

180MHz, dual-supply op amps are 16-bit accurate

The single MAX4430/MAX4431 and dual MAX4432/MAX4433 op amps feature wide bandwidth, 16-bit 37ns settling time, and low-noise/low-distortion operation. The MAX4430/MAX4432 are compensated for unity-gain stability and have a small-signal -3dB bandwidth of 180MHz. The MAX4431/MAX4433 are compensated for closed-loop gains of +2 or greater and have a small-signal -3dB bandwidth of 215MHz.

The MAX4430–MAX4433 op amps require only 11mA of supply current per amplifier while achieving 125dB open-loop gain. Voltage-noise density is a low 2.8nV/$\sqrt{Hz}$ and provides SFDR of 100dB (4V_{P-P}) at 1MHz. These characteristics make the op amps ideal for driving modern, high-speed 14- and 16-bit ADCs.

The MAX4430–MAX4433 deliver output currents to 60mA with wide voltage swings capable of driving an ADC with 4V or more of input dynamic range. Their voltage-feedback architecture serves many applications that otherwise require current-feedback amplifiers. The MAX4430/MAX4431 come in a space-saving 5-pin SOT23 package, and the MAX4432/MAX4433 come in an 8-pin µMAX package. Prices start at $2.25 (1000–up, FOB USA).

8:1 and dual 4:1 analog muxes have 3.5Ω R_{ON}

The MAX4638/MAX4639 CMOS ICs are suitable for use as 8:1/dual 4:1 analog multiplexers or demultiplexers. Each device operates from a single 1.8V to 5V supply or from dual ±2.5V supplies. When operating from a single 5V supply, the guaranteed R_{ON} is 3.5Ω. Other specifications are -75dB off-isolation, -85dB crosstalk (output to each off channel), 18ns t_{ON}, 7ns t_{OFF}, and a 0.25nA (max) guaranteed leakage current at +25°C.

All channels guarantee break-before-make switching, and the 1.8V to 5.5V operating range makes the devices ideal for use in battery-powered, portable instruments. Each handles rail-to-rail analog signals and provides bidirectional operation. All control inputs are TTL/CMOS compatible. Channel selection is encoded with the standard BCD format, and an enable input allows cascading of multiple devices.

The MAX4638/MAX4639 come in 16-pin SO and TSSOP packages. Prices start at $1.31 (1000–up, FOB USA).

Tiny ICs monitor and control fan speed

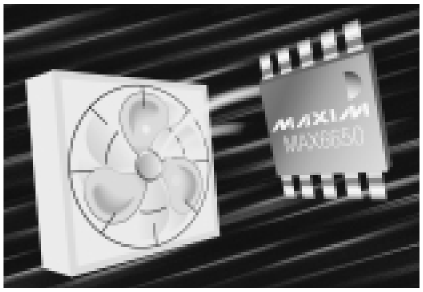
The MAX6650/MAX6651 fan-controller ICs offer a simple way to regulate and monitor the speed of a 5V/12V brushless DC fan with built-in tachometer. Communicating through an I²C-compatible/SMBus™ interface, the MAX6550/MAX6551 automatically force the fan's tachometer frequency (proportional to fan speed) to match a value selected and programmed by the system. The chip controls fan speed using low-noise, linear regulation of voltage across the fan with an external MOSFET or bipolar transistor.

The MAX6650 regulates the speed of a single fan by monitoring the fan's tachometer output. The MAX6651 also regulates the speed of a single fan, but it contains additional tachometer inputs that allow it to monitor up to four fans, controlling them as a single unit when the fans are operated in parallel.

Both devices contain general-purpose input/output (GPIO) pins that serve as digital inputs, digital outputs, and hardware interfaces. Capable of sinking 10mA, these open-drain I/O terminals can be used to drive an LED. In case of software failure, one of the GPIO pins can be configured to turn the fan on fully. Besides controlling fan speed, the MAX6550/MAX6551 devices monitor fan behavior and generate an alert at a GPIO pin when they detect a fault condition.

Packaged in a space-saving 10-pin µMAX, the MAX6650 is the smallest integrated fan controller available today. The MAX6651 comes in a small, 16-pin QSOP package. Prices start at $2.03 for the MAX6650 and $2.25 for the MAX6651 (1000–up, FOB USA).

SMBus is a trademark of Intel Corp.



NEW PRODUCTS

Dual, universal, switched-capacitor filters save space and power

The MAX7490/MAX7491 include two second-order switched-capacitor building blocks in a 16-pin QSOP. Both filter blocks are capable of generating all second-order functions: bandpass, lowpass, highpass, notch (band-reject), complex zeros, and allpass functions. Three of these functions are available simultaneously. Excellent accuracy and stability enable the MAX7490/MAX7491 to eliminate the complex and costly tuning normally required in high-order filter production.

Operating from a single supply of +5V (MAX7490) or +2.7V (MAX7491) and drawing only 3.5mA, these filters deliver rail-to-rail performance (within 0.2V of the supply rails). Quiescent currents drop to 0.2µA in shutdown. Center frequencies are clock-tunable from 1Hz to 40kHz, with a clock-to-corner ratio of 100:1. Two clocking options are available: self-clocking through the use of an external capacitor, or external clocking for tighter control of the center frequency. The MAX7490/MAX7491 double-sampling architecture places the sampling-to-corner frequency ratio at 200:1.

The MAX7490/MAX7491 are available in a 16-pin QSOP package specified for the commercial (0°C to +70°C) and extended (-40°C to +85°C) temperature ranges. Prices start at $2.15 (1000–up, FOB USA).

SOT23 LCD bias supplies deliver up to 28V

The MAX1605/MAX1606 step-up DC-DC converters employ an internal 0.5A switch to deliver up to 30V for LCD bias applications in cell phones, personal digital assistants (PDAs), palmtop computers, and other handheld portable equipment.

The MAX1605 comes in an ultra-small 6-pin SOT23 package. The MAX1606 comes in a small 8-pin µMAX package and includes an extra internal switch that completely disconnects output from input during shutdown.

Both devices let you reduce the output ripple and component size by limiting the peak inductor current to 500mA, 250mA, or 125mA (user set). Switching frequencies to 500kHz allow the use of tiny surface-mount components. Both devices have low 18µA quiescent supply current. They operate from a +2.4V to +5.5V V_{CC} supply, but the battery voltage can range from 0.8V to V_{OUT}.

The MAX1605/MAX1606 are specified for the extended temperature range (-40°C to +85°C). Prices start at $1.55 (MAX1605) and $1.75 (MAX1606) (1000–up, FOB USA). Preassembled evaluation kits with recommended external components are available to help reduce design time.

High-voltage 36V DC-DC step-down controllers in µMAX

The MAX1744/MAX1745 step-down DC-DC controllers operate with input voltages as high as 36V. They step down to a pin-selectable 3.3V or 5V output (MAX1744) or an adjustable output from 1.25V to 18V (MAX1745). These devices employ an external P-channel MOSFET to allow an output-power capability exceeding 50W. Packaged in a small, 10-pin µMAX, they set new space-saving standards for high-voltage, high-power, DC-DC step-down conversion.

The MAX1744/MAX1745 have a proprietary, current-limited control scheme that achieves over 90% efficiency over a wide range of output current. Their high switching frequencies (up to 300kHz) allow very small external components. Quiescent current is a low 90µA and drops to only 4µA in shutdown. A 100% maximum duty cycle allows the lowest possible dropout voltage.

The MAX1744/MAX1745 are specified for the extended temperature range (-40°C to +85°C). Prices start at $2.75 (1000–up, FOB USA). A preassembled evaluation kit with recommended external components is available to help reduce design time.

1MHz, low-voltage step-down regulators include synchronous rectification

The MAX1742/MAX1842/MAX1843 PWM step-down DC-DC converters provide high-efficiency, low-voltage outputs for notebook and subnotebook computers. The converters feature internal synchronous rectification for high efficiency and reduced component count. The internal 90mΩ PMOS switch and 70mΩ NMOS synchronous-rectifier switch easily deliver continuous load currents up to 1A. Maximum current thresholds are 1.3A for the MAX1742, 3.1A for the MAX1842, and 2.7A for the MAX1843. All devices achieve efficiencies as high as 95%. All provide preset output voltages of +2.5V, +1.8V, or +1.5V, or adjustable outputs from +1.1V to V_{IN}.

The current-mode, constant off-time, PWM control scheme includes an idle mode that maintains high efficiency during light-load operation. The scheme allows switching frequencies up to 1MHz, which allows the user to optimize trade-offs between efficiency, output switching noise, component size, and cost. Also included is an adjustable soft-start to limit surge currents during startup, a 100% duty-cycle mode for low-dropout operation, and a low-power shutdown mode that disconnects the input from the output and reduces supply current below 1µA.

The MAX1742/MAX1842 come in a 16-pin QSOP package specified for the extended temperature range (-40°C to +85°C). Prices start at $3.71 (1000–up, FOB USA). The MAX1843 comes in a 28-pin QFN package. Prices start at $3.89 (1000–up, FOB USA).

NEW PRODUCTS

Li+ battery chargers with internal power switch deliver 1.5A

The MAX1757/MAX1758 are stand-alone battery chargers for one to four Li+ cells. They regulate battery voltage with accuracy better than 0.8% and conversion efficiency up to 90%. A complete internal state machine safely controls the charging sequence. These step-down, switch-mode DC-DC converters use an internal, high-side, N-channel FET power switch to provide accurate charging currents up to 1.5A and to sustain high efficiency over a wide input voltage range. The MAX1757 input voltage limit allows operation up to 14V (up to 3 cells), and the MAX1758 input limit allows operation to 28V (up to 4 cells). Both devices feature a 98% duty cycle to allow use of lower voltage input adapters.

These devices use two loops to regulate the voltage set point and charging current, making automatic transitions between current regulation (during fast-charge) and voltage regulation (near full-charge). To service the system load during charging, an additional control loop monitors total current drawn from the input source. By lowering the charging current to prevent overloading the input supply when the system load increases, this loop allows use of a lower-cost wall adapter.

A built-in safety timer automatically terminates charging once the adjustable time limit has been reached. Battery temperature is continuously monitored by an external thermistor to prevent charging if the battery is too hot or too cold. The MAX1757/MAX1758 safely precharge near-dead cells if the battery voltage is below 2.5V per cell (battery voltage must rise above this threshold before fast-charge can begin). Open-drain outputs drive external LEDs to indicate fast-charge, full-charge, and fault conditions.

The MAX1757/MAX1758 come in a space-saving 28-pin SSOP package specified for the extended temperature range (-40°C to +85°C). Prices (1000–up, FOB USA) start at $3.25 (MAX1757) and $3.52 (MAX1758). Preassembled evaluation kits with recommended external components are available to help reduce design time.

Input-current-limited battery charger is 0.7% accurate

The MAX1772 is a highly integrated, multichemistry battery charger that simplifies the construction of accurate and efficient chargers. User-programmed or hardwired, it employs analog inputs to control the charging current and voltage. Its buck topology with synchronous rectification achieves high efficiency.

To reduce the AC adapter's cost and to avoid overloading it when the load and battery charger are supplied simultaneously, the MAX1772 lets you program the maximum current drawn from the AC adapter. MAX1772 outputs can monitor the battery-charging current, AC-adapter current, and whether the adapter is present. The MAX1772 easily provides 4A to charge two to four Li+ cells in series. When charging, it makes an automatic transition from regulating current to regulating voltage.

The MAX1772 comes in a space-saving 28-pin QSOP package specified for the extended temperature range (-40°C to +85°C). Prices start at $5.04 (1000–up, FOB USA).

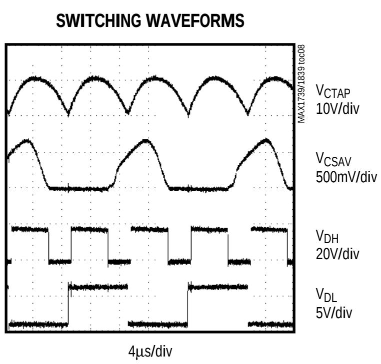
CCFL backlight controllers have wide brightness range

The MAX1739/MAX1839 controllers are optimized for driving cold-cathode fluorescent lamps (CCFLs) using the industry-proven Royer oscillator inverter architecture. The Royer architecture maximizes CCFL life by providing near-sinusoidal drive waveforms over the entire input range. In the MAX1739/MAX1839, this architecture is optimized to achieve high efficiency, maximize the dimming range, and work over a wide input voltage range (4.6V to 28V).

These devices monitor and limit the transformer center-tap voltage to ensure minimal voltage stress on the transformer, thereby increasing its operating life and easing its design requirements. The controllers also provide protection against lamp outs, buck shorts, and other fault conditions.

The controllers achieve a 50:1 dimming range by adjusting the lamp current and simultaneously "chopping" the CCFL on and off, using a digitally adjusted pulse-width modulated (DPWM) method. CCFL brightness can be controlled either by an applied analog voltage or through an SMBus-compatible 2-wire interface (MAX1739 only). The MAX1739/MAX1839 drive an external high-side N-channel power MOSFET and two low-side N-channel power MOSFETs, all synchronized to the Royer oscillator. An internal 5.3V linear regulator powers the MOSFET drivers and most of the internal circuitry.

The MAX1739/MAX1839 are available in a space-saving 20-pin QSOP package specified for the extended temperature range (-40°C to +85°C). Prices start at $3.85 (1000–up, FOB USA).



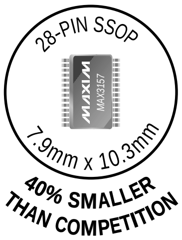
NEW PRODUCTS

Isolated RS-422/RS-485 transceiver in a surface-mount package

The MAX3157 RS-422/RS-485 transceiver is ideal for noisy industrial applications and large local area networks. Its ±50V isolation protects the device against ground differentials that can interrupt data transmission or destroy traditional RS-485 devices. The MAX3157 is a low-cost, surface-mount alternative to optoisolated transceivers. To ensure data integrity without using external components, it also features true fail-safe receiver inputs that guarantee a logic-high output level when the inputs are open or shorted or the bus is idle.

The MAX3157 operates from a single +5.0V supply and features a 25µA shutdown mode. Control pins for receiver phase and transmitter phase allow users to compensate for improper wiring by reversing the output-signal polarity. The device offers pin-selectable half- or full-duplex operation to provide maximum compatibility with existing RS-422/RS-485 networks. The MAX3157 features slew-rate limiting to reduce EMI and reflections and is guaranteed to transfer data up to 250kbps.

The MAX3157 comes in space-saving 28-pin SSOP and DIP packages specified for the commercial and extended temperature ranges. Prices start at $3.50 (1000–up, FOB USA).



3V, 1µA multiprotocol interfaces

The MAX3160/MAX3161/MAX3162 are the world's first 3V, 1µA multiprotocol ICs to combine RS-232 and RS-485/RS-422 compliance in a single chip. Shutdown mode lowers supply current to 1µA while the receivers remain active. Designed for networking, point of sales, and industrial equipment, these ICs expand end-product capability by allowing systems to communicate using the RS-232 or RS-485/RS-422 standard. The MAX3160 is optimized for multiprotocol operation on a single interface bus. The MAX3161 is a better choice for applications with a single UART and separate interface bus. The MAX3162 is ideal for applications such as protocol translators that require both RS-232 and RS-485/RS-422 protocols.

RS-232 or RS-485/RS-422 operation is pin programmable on the MAX3160/MAX3161. The MAX3162's dedicated RS-232 and RS-485/RS-422 I/O pins allow data transmission using both protocols simultaneously. All devices feature a low-dropout output stage that enables operation from a single +3V to +5.5V supply. Supply current drops to 1µA in shutdown mode while the RS-232 receiver inputs remain active. Other MAX3160 family features include pin-programmable slew-rate limiting to reduce EMI, pin-programmable half- or full-duplex operation (MAX3160/MAX3161), pin-programmable FAST operation allowing RS-232-compliant data rates up to 1Mbps or RS-485/RS-422 data rates up to 10Mbps, and true fail-safe receivers that ensure a logic high output when the inputs are shorted or open.

All devices are available in space-saving SSOP packages specified for the commercial and extended temperature ranges. Prices start at $5.50 (1000–up, FOB USA).

LVDS in SOT23 achieves >250ps output-pulse skew

The MAX9110–MAX9113 low-voltage differential signaling (LVDS) line drivers and receivers are designed for high-speed applications requiring minimum power dissipation, board space, and radiated noise. As Maxim's first general-purpose LVDS products, they offer the smallest size and the lowest pulse skew (250ps) available in such devices. Their exceptional performance maximizes the imaging resolution and speed of printers and copiers, including those that employ dithering and grey-scale techniques to enhance image quality. Low-distortion switching enables data rates exceeding 500Mbps, making these products ideal for signal and clock distribution in networking and telecom applications.

The MAX9110/MAX9112 are single and dual transmitters, and the MAX9111/MAX9113 are single and dual receivers. All devices operate from a single +3.3V supply, conform to the EIA/TIA-644 LVDS standards, and feature the low-power and low-EMI benefits of LVDS (31mW power consumption for the dual-channel MAX9112 and 350mV differential output-voltage swings for all four devices).

These devices form a pin-compatible plug-in upgrade to the National Semiconductor DS90LV017/018/027/028. All are available in a standard 8-pin SO package as well as a tiny 8-pin SOT23 package (less than 43% of the footprint of an SO) for applications requiring minimum space and PCB cost. Prices (1000–up, FOB USA) start at $0.81 for the MAX9110/MAX9111 and $1.78 for the MAX9112/MAX9113.

SOT23, low-power µP supervisors with battery backup

The MAX6361–MAX6364 supervisory circuits reduce the complexity and number of components required for power-supply monitoring and battery control functions in µP systems. These circuits improve system reliability and accuracy compared to separate-IC or discrete-component alternatives. MAX6361–MAX6364 functions include µP reset, backup battery switchover, and power failure warning.

These supervisors operate from supply voltages as low as +1.2V. The factory-preset, reset-threshold voltages range from 2.32V to 4.63V. The devices provide a manual reset input (MAX6361), watchdog timer input (MAX6362), battery-on output (MAX6363), and an auxiliary adjustable reset input (MAX6364). In addition, each part type is offered in three reset-output versions: an active-low push-pull reset, an active-low open-drain reset, and an active-high open-drain reset.

The MAX6361–MAX6364 are available in a small, 6-pin SOT23 package. Prices start at $1.50 (2500–up, FOB USA).

NEW PRODUCTS

2.8V single-supply, cellular-band linear power amplifier

The MAX2251 low-voltage linear power amplifier (PA) is designed for TDMA/AMPS dual-mode phone applications operating with a 2.8V to 4.5V supply voltage. It comes in an ultra-compact (2.06mm x 2.06mm) chip-scale package (CSP) and delivers over +30dBm of linear power in TDMA operation (41% typical efficiency). An on-chip shutdown capability lowers the operating current to 1µA (typ), thereby eliminating the need for an external supply switch.

The MAX2251 includes a power detector. It needs no external reference voltage or bias circuitry and requires only a few external matching components. The use of external bias resistors eliminates the waste of “safety-margin” current and ensures the highest possible efficiency at all power levels by allowing current throttleback at the lower output-power levels. Gain variation for ambient temperatures from -40°C to +85°C is only ±0.9dB. Prices start at $1.95 (1000–up, FOB USA).

Low-noise SiGe amplifier in UCSP package has variable IIP3

The MAX2374 is a silicon-germanium (SiGe), switchable-gain, variable-linearity, low-noise amplifier (LNA) designed for cellular-band, code-division multiple-access (CDMA) applications. It is also suitable for high-dynamic-range, low-noise applications such as TDMA and PDC. It provides a high intermodulation intercept point (IIP3) that can be adjusted for specific requirements through an external resistor.

The MAX2374’s high-gain mode optimizes system sensitivity, and its low-gain mode optimizes system linearity. To achieve small size, high gain, and low noise (1.5dB noise figure), the LNA is packaged in a tiny ultra-chip-scale package (UCSP) with six solder bumps. The LNA operates from a +2.7V to +5.5V single supply and draws just 8.5mA while achieving a +6.2dBm input IIP3. Shutdown mode lowers the supply current to <1µA. Prices start at $ 0.89 (1000–up, FOB USA).

Complete dual-band quadrature transmitters

The MAX2366 dual-band, triple-mode transmitter for cellular phones represents the most integrated and architecturally advanced device available for this application. It takes a differential I/Q baseband input and upconverts to IF through a quadrature modulator and IF variable-gain amplifier (VGA). The signal is then routed to an external bandpass filter and upconverted to RF through an SSB mixer and RF VGA. The signal is further amplified with an on-board PA driver.

The basic functional blocks of this IC include dual IF synthesizers, dual RF synthesizers, a local oscillator (LO) buffer, and a 3-wire programmable bus. The MAX2367 supports single-band, single-mode (PCS) operation, and the MAX2368 supports single-band cellular dual-mode operation. The supply-voltage range is 2.7V to 5.5V.

Two IF voltage-controlled oscillators (VCOs), two IF ports, two RF LO input ports, and three PA-driver output ports allow the use of a single receive IF frequency and split-band PCS filters for optimum out-of-band noise performance. Select an operational mode by loading data on the SPI™/QSPI™/ MICROWIRE™-compatible 3-wire serial bus, and the PA drivers let you eliminate up to three RF SAW filters. Charge-pump current, sideband rejection, IF/RF gain balancing, standby, and shutdown are also controlled with the serial interface.

The MAX2366/MAX2367/MAX2368 are available in a 48-pin QFN exposed-pad package specified for the extended temperature range (-40°C to +85°C). Prices start at $5.85 (1000–up, FOB USA).

SPI/QSPI are trademarks of Motorola, Inc.

MICROWIRE is a trademark of National Semiconductor Corp.

Industry-first, 3.5GHz SiGe LNA provides high linearity and application flexibility

The MAX2645 SiGe LNA is the industry’s first low-noise amplifier optimized for 3.4GHz to 3.8GHz wireless local loop, wireless broadband access, and digital microwave radio applications. Manufactured with Maxim’s advanced high-frequency SiGe process, the MAX2645 features low noise figure, high gain, adjustable third-order intercept point (IP3), and a logic-controlled gain step function. These capabilities make it ideal for use as a 1st- and 2nd-stage receiver LNA, a PA predriver in the transmitter, or an LO buffer amp.

Typical MAX2645 performance includes 14.4dB gain, 2.3dB noise figure, and +4dBm input IP3, all with just 9.2mA of supply current. The gain step feature reduces the LNA gain by 24dB while increasing the input IP3 to +13dBm, thereby improving the receiver front-end performance under high input signal levels. This mode lowers supply current to only 3mA. The IP3 is also adjustable through an off-chip bias resistor. When used as a PA predriver, for example, the amplifier input IP3 can be adjusted to +11.5dBm (+26dBm output IP3) for 13mA.

The supply-voltage range is +3V to +5.5V, and a logic-controlled shutdown reduces supply current to 1µA. The MAX2645 is offered in a miniature 10-pin µMAX package with exposed paddle. Prices start at $1.25 (1000–up, FOB USA). A fully assembled evaluation kit is available to help reduce design time.