Photomultipliers, avalanche photodiodes, ultrasonic transducers, condenser microphones, radiation detectors, and similar devices require high-voltage, low-current biasing. Additionally, the high voltage must be free of noise. A common requirement is less than 1 mV of noise, and these devices sometimes require noise to be less than a few hundred microvolts. Switching-regulator configurations cannot normally achieve this performance level without employing special techniques. One aid to achieving low noise is the fact that load currents rarely exceed 5 mA. This freedom permits the use of output-filtering methods that are otherwise impractical.

This article describes a variety of circuits featuring outputs of 200 to 1000V with less than 100 µV of output noise in a 100-MHz bandwidth. Special techniques, most notably power stages that minimize high-frequency harmonic content, enable this performance. Although sophisticated, all these examples use standard, commercially available magnetic components. This provision should help you quickly arrive at a manufacturable design.

Before proceeding any further, understand that you should use caution in the construction, testing, and use of the circuits this article describes. High-voltage, lethal potentials are present in these circuits. Use extreme caution in working with and making connections to these circuits. Again, these circuits contain dangerous, high-voltage potentials.

**Resonant Royer-based converters**

The resonant Royer topology suits low-noise operation due to its sinusoidal power delivery ([Reference 1](#) and [Reference 2](#)). The resonant Royer is attractive because transformers for LCD-backlight service are readily available. These transformers are available from multiple sources, well-proven, and competitively priced. *Figure 1*’s resonant Royer topology achieves 100-µV-p-p noise at 250V output by minimizing high-frequency harmonics in the power-drive stage. The self-oscillating resonant Royer circuitry comprises Q₂, Q₃, C₁, T₁, and L₁. Current flow through L₁ causes the T₁, Q₂, Q₃, and C₁ circuitry to oscillate in resonant fashion, supplying sine-wave drive to T₁’s primary with resultant sinelike high voltage appearing across the secondary.

T₁’s rectified and filtered output feeds back to amplifier-reference A₁, which biases the Q₁ current sink, completing a control loop around the Royer converter. L₄ ensures that Q₁ maintains constant current at high frequency. Milliampere-level output current allows the presence of a 10-kΩ resistor in the output filter. This resistor greatly aids filter performance with minimal power loss. The low current requirements permit certain freedoms in the output filter and feedback network (see sidebar “Feedback considerations in high-voltage dc/dc converters”). The RC path to A₁’s negative input combines with the 0.1-µF capacitor to compensate A₁’s loop. D₅ and D₆, low-leakage clamps, protect A₁ during start-up and transient events. Although *Figure 2*’s collector waveforms are distorted, no high-frequency content is present.
The circuit’s low harmonic content combines with the RC-output filter to produce a transcendentally clean output. Output noise (Figure 3) is just discernible in the monitoring instrumentation’s 100-µV noise floor Reference 3).

Figure 4’s variant of Figure 1 maintains 100-µV output noise and extends the input-supply range to 32V. Q₁ may require heat-sinking at high input-supply voltage. Converter and loop operation remains the same as in Figure 1, although Figure 4 re-establishes compensation components to accommodate the LT1431 control element.

The previous resonant Royer examples use linear control of converter current to furnish harmonic-free drive. The trade-off is decreased efficiency, particularly as input voltage scales. You can improve efficiency by employing switched-mode current drive to the Royer converter. Unfortunately, such switched drive usually introduces noise. However, you can counter this undesirable consequence.

Figure 5 replaces the linearly operated current sink with a switching regulator. The Royer converter and its loop are the same as in Figure 4; Figure 6’s transistor-collector waveshape (Trace A) is similar to that of the other circuits. The high-speed, switched-mode current-sink drive (Trace B) efficiently feeds L₁. This switched operation improves efficiency but degrades output noise. Figure 7 shows switching-regulator harmonic clearly responsible for 3-mV-p-p output noise—about 30 times greater than that of the linearly operated circuits.

Careful examination of Figure 7 reveals almost no Royer-based residue. Switching-regulator artifacts dominate the noise. Eliminating this switching-regulator-originated noise and maintaining efficiency requires special circuitry, but this circuitry is readily available (Figure 8). The resonant Royer converter and its loop are reminiscent of the circuits in the preceding figures. The fundamental difference is the LT1534 switching regulator that uses controlled transition times to retard high-frequency harmonic and maintain efficiency. This approach blends switching and linear-current-sink benefits (Reference 3). Rᵥ and Rᵢ set the voltage and current-transition rate, respectively, which represents a compromise between efficiency and noise reduction.

Figure 9’s Royer collector waveshape (Trace A) is nearly identical to the one that Figure 5’s circuit produces. Trace B, depicting LT1534-controlled transition times, markedly departs from its Figure 5 counterpart. These controlled transition times dramatically reduce output noise (Figure 10) to 150 µV p-p—a 20-fold improvement over Figure 7’s LTC3401-based results.

Figure 11 is essentially identical to Figure 8, except that it produces a -1000V output. A, provides low impedance, inverting feedback to the LT1534. Figure 12a’s output noise measures less than 1 mV. As before, resonant Royer ripple dominates the noise; no high-frequency content is detectable. It is worth noting that this noise figure proportionally improves with increased filter-capacitor values. For example, Figure 12b indicates only 100-µV noise with 10-times-higher filter-capacitor values, although the capacitors are physically large. The original values represent a reasonable compromise between noise performance and physical size.

**Push-pull converters**

Controlled transition techniques are also directly applicable to push-pull architectures. Figure 13 uses a controlled transition push-pull regulator in a simple loop to control a 300V output converter. Symmetrical-transformer drive and controlled switching-edge times promote low output noise. The D₁ through D₄-connected damper further minimizes residual aberrations. In this case, the output filter uses inductors, although you could employ appropriate resistor values. Figure 14 displays smooth transitions at the transformer secondary outputs. (Trace A is T₁ Pin 4, and Trace B is T₁ Pin
7.) The absence of high-frequency harmonic results in extremely low noise. Figure 15’s fundamental-related output residue approaches the 100-µV measurement noise floor in a 100-MHz bandpass. This performance is spectacularly low noise in any dc/dc converter, and certainly in one providing high voltage. Here, at 300V output, noise represents less than 1 part in 3 million.

Figure 16 is similar, except that output range varies from 0 to 300V. An LT3439, which contains no control elements, replaces the LT1533. It simply drives the transformer with 50%-duty-cycle, controlled switching transitions. A1, Q1, and Q2 enforce feedback control by driving current into T1’s primary center tap. A1 compares a resistively derived portion of the output with a user-supplied control voltage. These values produce a 0 to 300V output in response to a 0 to 1V control voltage. An RC network from Q2’s collector to A1’s positive input compensates the loop. Collector waveforms and output-noise signature are nearly identical to those in Figure 13. Output noise is 100 µV p-p over the entire 0 to 300V output range.

**Flyback converters**

You don’t usually associate flyback converters, with their abrupt, poorly controlled energy delivery, with low-noise output. However, careful magnetic selection and layout can provide surprisingly good performance, particularly at low output current. Figure 17’s design provides 200V from a 5V input (Reference 4 and Reference 5). The scheme is a basic inductor-flyback-boost regulator with some important deviations. Q1, a high-voltage device, resides between the LT1172 switching regulator and the inductor. This approach permits the regulator to control Q1’s high-voltage switching without undergoing high-voltage stress. Q1, operating as a cascode with the LT1172’s internal switch, withstands L1’s high-voltage flyback events (Reference 6 through Reference 10).

Diodes associated with Q1’s source-terminal clamp, L1, originated spikes arriving through Q1’s junction capacitance. The high voltage is rectified and filtered, forming the circuits’ output. The ferrite bead and 100 and 300Ω resistors aid filter efficiency (Reference 11 and Reference 12). Feedback to the regulator stabilizes the loop and the Vc-pin network provides frequency compensation. A 100-kΩ path from L1 bootstraps Q1’s gate drive to about 10V, ensuring saturation. The output-connected diode provides short-circuit protection by shutting down the LT1172 if the output is accidentally grounded.

Figure 18’s traces A and C are LT1172 switch current and voltage, respectively. Q1’s drain is Trace B. Current-ramp termination results in a high-voltage flyback event at Q1’s drain. A safely attenuated version of the flyback appears at the LT1172 switch. The sinusoidal signature, due to inductor ring-off between conduction cycles, is harmless. Figure 19, output noise, comprises low-frequency ripple and wideband, flyback-related spikes measuring 1 mV p-p in a 100-MHz bandpass.

In a transformer-coupled flyback circuit, the transformer secondary provides voltage step-up referred to the flyback-driven primary (Figure 20). The 4.22-MΩ resistor supplies feedback to the regulator, closing a control loop. A 10-kΩ, 0.68-µF filter network attenuates high-frequency harmonic with minimal voltage drop. Figure 21 clearly shows flyback-related transients in the output noise, although they are within 300 µV p-p.

The circuit in Figure 22 employs the LT3468 photoflash-capacitor charger as a general-purpose, high-voltage dc/dc converter. Normally, the LT3468 regulates its output at 300V by sensing T1’s flyback-pulse characteristic. This circuit allows the LT3468 to regulate at lower voltages by truncating its charge cycle before the output reaches 300V. A1 compares a divided-down portion of the output with the program input voltage. When the output-derived potential at A1’s negative input exceeds the program voltage at A1’s positive input, A1’s output goes low, shutting down the LT3468.
The feedback capacitor provides ac hysteresis, sharpening A₁’s output to prevent chattering at the trip point. The LT3468 remains shut down until the output voltage drops low enough to trip A₁’s output high, turning it back on. In this way, A₁’s duty cycle modulates the LT3468, causing the output voltage to stabilize at a point that the program input determines.

*Figure 23*’s 250V-dc output (Trace B) decays down about 2V until A₁ (Trace A) goes high, enabling the LT3468 and restoring the loop. This simple circuit works well, regulating over a programmable 0 to 300V range, although its inherent hysteretic operation mandates the unacceptable 2V output-ripple noted. The loop-repetition rate varies with the input voltage, output setpoint, and load, but the ripple is always present.

The circuit in *Figure 24* greatly reduces ripple amplitude, although complexity increases. The circuit’s postregulator reduces the output ripple and noise of *Figure 22*’s circuit to only 2 mV. A₁ and the LT3468 are identical to *Figure 23*’s circuit, except for the 15V zener diode in series with the 10-MΩ/100-kΩ feedback divider. This component causes C₁’s voltage, and hence Q₁’s collector, to regulate 15V above the V_PROGRAM input-dictated point. The V_PROGRAM input also routes to the A₂-Q₂-Q₁ linear postregulator. A₂’s 10-MΩ/100-kΩ feedback divider has no zener diode, so the postregulator follows the V_PROGRAM input with no offset. This arrangement forces 15V across Q₁ at all output voltages. This figure is high enough to eliminate undesirable ripple and noise from the output and keep Q₁’s dissipation low.

Q₃ and Q₄ form a current limit, protecting Q₁ from overload. Excessive current through the 50Ω shunt turns on Q₃. Q₃ drives Q₄, shutting down the LT3468. Simultaneously, a portion of Q₁’s collector current turns on Q₄ hard, shutting off Q₁. This loop dominates the normal regulation feedback, protecting the circuit until you remove the overload.

*Figure 25* shows just how effective the postregulator is. When A₁ (Trace A) goes high, Q₁’s collector (Trace B) ramps up in response. Note the LT3468’s switching artifacts on the ramp’s upward slope. When the A₁-LT3468 loop is satisfied, A₁ goes low and Q₁’s collector ramps down. The output postregulator (Trace C), however, rejects the ripple, showing only 2 mV of noise. The slight blurring of the trace derives from A₁-LT3468 loop jitter.

**Circuit characteristics**

*Table 1* summarizes and notes the salient characteristics of the circuits in this article. This *table* is only a generalized guideline and not an indicator of capabilities or limits. Too many variables and exceptions exist to accommodate the categorical statement the *table* implies. The interdependence of circuit parameters makes summarizing or rating various approaches a hazardous exercise. There is simply no intellectually responsible way to streamline the selection and design process if you want optimum results. A meaningful choice must be the outcome of laboratory-based experimentation. Too many interdependent variables and surprises exist for a systematic, theoretically based selection. Tables such as this one seek authority through glib simplification, and simplification is disaster’s deputy. Nonetheless, *Table 1*, in all its glory, lists input-supply range, output voltage, and current, along with comments for each circuit.

**References**

2. Bright, Pittman, and Royer, “Transistors as On-Off Switches in Saturable Core Circuits,”
Feedback considerations in high-voltage dc/dc converters

A high-voltage dc/dc-converter-feedback network is a study in compromise. The appropriate choice of such a network depends on the application. Considerations include desired output impedance, loop stability, transient response, and high-voltage-induced overstress protection. Figure A through Figure E list typical options. The circuit in Figure A requires no special commentary; the one in Figure B adds an ac-lead network for improved dynamics. Diode clamps protect the feedback node from the capacitor's differentiated response. The circuit in Figure C, a low-ripple, two-section filter, slows transient response, but a lead network provides stability. Resistor R, outside the loop, sets dc output impedance. The circuit in Figure D encloses R within the dc loop, lowering output resistance but delaying loop transmission. A feedback capacitor supplies corrective leading response. The circuit in Figure E moves the feedback capacitor to the filter input, further extending Figure D circuit's leading response. The circuit in Figure F replaces filter resistor R with an inductor, lowering output resistance but introducing parasitic shunt capacitance, which combines with capacitor-loss terms to degrade filtering. The inductor also approximates a transformer secondary, vulnerable to stray flux pickup with resulting increased output noise (Reference A).

A common concern in any high-voltage-feedback network is reliability. You must carefully choose
your components, and you must use and strictly adhere to conservative voltage ratings. Although you can easily ascertain component ratings, more subtle effects, such as ill-suited board material and board-wash contaminants, can be reliability hazards. Long-term electromigration effects can have undesirable results. You should consider every potential unintended conductive path as an error source and plan your layout accordingly. You must anticipate operating-temperature, altitude, humidity, and condensation effects. In extreme cases, it may be necessary to rout the board under components operating at high voltage. Similarly, it is common practice to use several units in series to minimize voltage across the output-connected feedback resistor. Contemporary packaging requirements emphasize a tightly packed layout that may conflict with high-voltage standoff requirements. You must carefully review this trade-off; otherwise, reliability will suffer. Do not underestimate the potentially deleterious—or even disastrous—effects of environmental factors, layout, and component choice over time. You must think clearly to avoid unpleasant surprises.

Reference