

Analyzer tests reverse-recovery behavior of diodes

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Testing the reverse-recovery behavior of diodes normally requires complex testing gear. You must be able to establish the forward-conduction conditions, the blocking state, and the transition between the two. You also need a means of extracting the characteristics from the resulting waveform. In short, a specialist should handle this complex job; it is not something you routinely control in the field.

This fact explains why engineers generally prefer to rely on published data.

Checking the reverse-recovery time yourself could be advantageous, however, if testing were simple and straightforward. Such a setup would enable you to compare devices from different manufacturers under identical conditions and test devices having no such specification, such as substrate diodes of driver ICs, zener diodes, and stan-

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ard rectifiers. (Because of the number of combinations of the test parameters, a direct comparison of the data is rare-

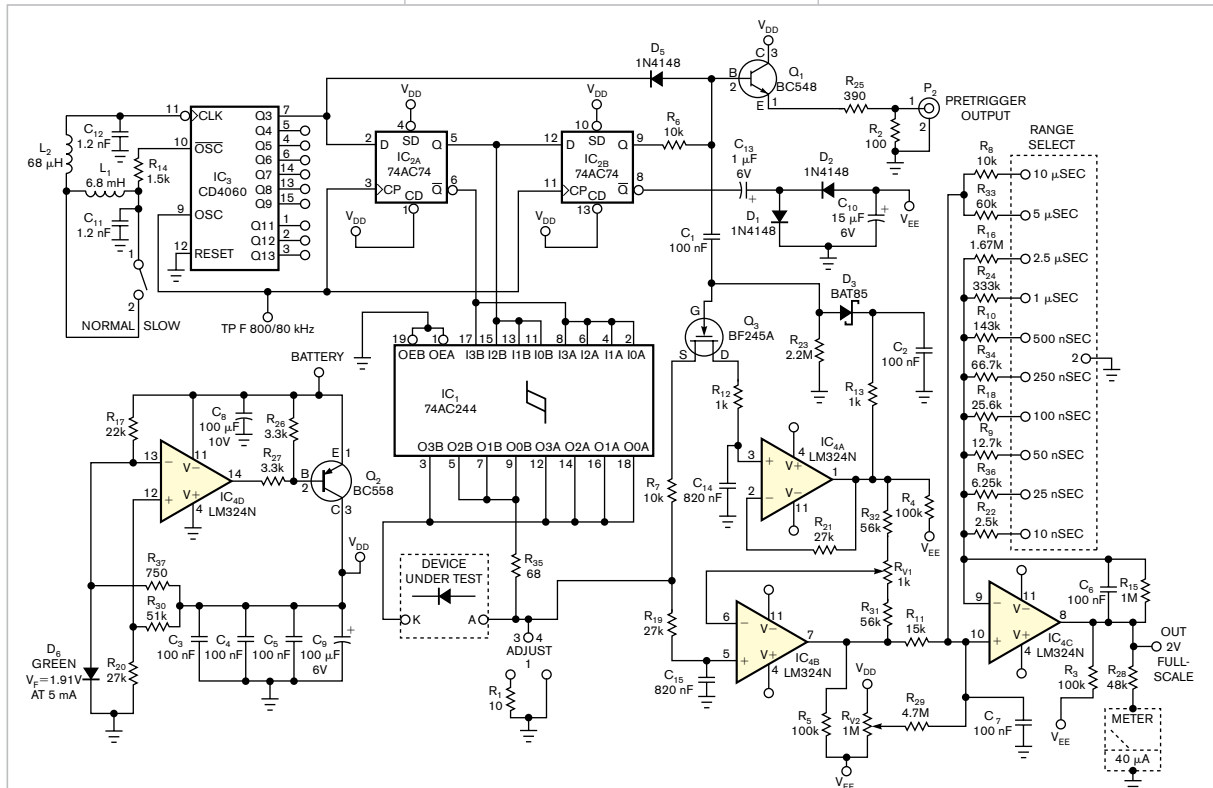


Figure 1 This diode-recovery test setup allows you to compare devices from different manufacturers under identical conditions.

ly possible.) Note that shorter reverse-recovery time is not necessarily better. Slow diodes can be useful, too. They can generate small dead times, improve the efficiency of converters, and provide other benefits (**Reference 1**).

This Design Idea presents a tester that, using only a handful of inexpensive, standard components, allows you to check reverse-recovery time. The test conditions are fixed for simplicity, to normalize the tests and to provide a common standard for comparison purposes. These conditions are compatible with 99% of the devices susceptible to test. The tester's forward current is just low enough to be safe with small switching diodes but high enough to overcome the capacitive effects in larger devices.

A diode-resistor AND gate lies at the heart of the circuit; the gate's diode is the DUT (device under test, **Figure 1**). IC₁ buffers flip-flop IC_{2A}, which derives the antiphase square waves that drive this gate. R₃₅ sets the DUT's forward current to approximately 75 mA. With an ideal diode, the gate's output would always stay low, because one of the inputs is always low. But a real diode remains conductive after the transition, generating a positive pulse across R₃₅. Instead of using the brute-force approach of directly measuring this pulse width, the circuit uses a subtler scheme. The R₁₉/C₁₅ network averages the pulse and amplifies and displays the resulting voltage. Because the measurement frequency is fixed at 50 kHz, a correct scaling factor is all that is necessary.

A real diode also has a forward voltage, which you would average with the result. Q₃ takes care of this problem by sampling this forward voltage through IC_{4A} and subtracting it from the output voltage through R₃₂. Vary-

ing the gain of amplifier IC_{4C} sets the various ranges. In this case, the ranges are in a 1, 2.5, 5 sequence, which suit the salvaged galvanometer this circuit uses as an indicating device. You could easily create other ranges by adapting the values of R₈ through R₂₂. The big advantage of this measuring method is that it handles only dc or low-frequency signals, requiring no fast comparators or samplers, yet it can resolve a few hundreds of picoseconds.

The built-in oscillator of IC₃ generates the clock. The clock frequency is 800 kHz and divides down to produce the 50-kHz reference at Q₃. An optional slow mode is available for those needing to test devices slower than 5 μ sec. The insertion of coil L₁ decreases the clock frequency to 80 kHz and enables you to measure reverse-recovery time as fast as 50 μ sec. IC₂ generates the test waveforms and shifts the 50-kHz signal at the clock rate. The leading and trailing states then exit through the D₃/R₆ AND gate to produce a sampling pulse that centers on the conduction period. Because the sampling occurs far from any transitions, it need not be particularly fast or accurate. C₁ transfers the sampling pulse and provides a convenient pretrigger signal, which Q₁ buffers. This option enables a comfortable observation of the waveform when you connect an oscilloscope to the anode of the DUT.

The unused output, Pin 8 of IC_{2B}, feeds a negative-voltage generator, serving as a bias source for the outputs of IC₄ to let them reach a true zero. The measurement circuits receive power from a 9V battery by a supply encompassing IC_{4D}. An LED serves as a reference to the 5.5V and provides some temperature compensation because the reverse-recovery time de-

pends highly on ambient temperature.

You can make some adjustments to the circuit. For example, with no diode inserted, you can short the Adjust testpoint 1 to 4. In the 10- or 25-nsec range, R_{V2}, which is 0 nsec in the range, to get a midscale reading. Move the short to Adjust testpoint 3, R_{V1}, and R_{V1}, thereby providing V_F cancellation, to read the same value. Repeat the procedure until the reading is independent of the position of the short. The adjustment interacts with the zero due to the offset of the amplifiers.

Now, you have eliminated the effect of V_F. You can adjust the 0 nsec by shorting Adjust testpoints 1 and 4 and adjusting R_{V2} to read zero on the 10-nsec range. This adjustment yields 0 nsec with a typical offset of 1 to 2 nsec in the positive direction. Residual skew in the timing and charge-injection effects cause this offset. Normally, this offset should not be a problem, because it is small, stable, and constant. If you need an absolute accuracy down to the picosecond, you have to test a known, ultrafast diode, such as an FD700 or a BAY82, and adjust the 0 nsec to read the actual value. If you lack access to such a diode, you can always arbitrarily shift the value by 1.5 nsec. This adjustment is normally sufficient to reach a ± 500 -psec accuracy. Schottky diodes are unsuitable. Despite their low recovery time, they generate a nonzero reading because of their relatively high capacitance and non-negligible leakage currents. Low-capacitance, mixer-type diodes are too fragile for this tester. **EDN**

REFERENCE

1 Vlemincq, Louis, "Slow diodes or handy timing devices?" *EDN*, Sept 16, 2004, pg 83, www.edn.com/article/CA450598.

High-power LED drivers require no external switches

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As the latest generation of new LEDs achieves higher levels of power and efficiency, use of these de-

vices extends to new areas, such as flashlights and vehicular applications. High-power LEDs are finding use even

in ambient lighting, long the sole province of incandescent bulbs and fluorescent tubes. A current source is the best way to power LEDs. Because most energy sources, including batteries, generators, and industrial mains, look more like voltage sources than current sources, LEDs require that you insert some

electronic circuitry between them and the source of power. This circuitry can be as simple as a series resistor, but a better choice, considering energy efficiency and other factors, is a high-efficiency, voltage-fed current source. For LEDs with currents greater than 0.35A, an inductive switching regulator is usually the best choice.

This Design Idea presents a series of circuits based on single-power-IC switching regulators, with efficiency and miniaturization as the main objectives. The circuits' designers approach these objectives by minimizing the use of large components, such as external power transistors, switches, high-value capacitors, and current-sense resistors, and by maintaining regular operation by delivering constant, high-intensity light over as extended a range as possible.

The circuits in **figures 1** through **3** are suitable for applications in which the power source comprises three or four alkaline, NiMH (nickel-metal-hydride), or NiCd (nickel-cadmium) cells. Those in **figures 4** and **5** are for vehicular applications in which the nominal line voltage for the power-distribution system is 12, 24, or 42V. The circuits of **figures 4** and **5** are also useful in industrial systems that include a 24V distribution line for control and emergency subsystems and in telecom applications for which the system power is distributed as a -48V line.

The designers of these circuits based them on the same concept: a fully integrated, single-die-IC switching regulator and a micropower operational amplifier. The op amp drives the 1.25V feedback terminal on the IC. Although that node targets the topology of a standard voltage regulator, the op amp matches it to the much smaller current-sense voltage and the slightly different topology of a current regulator. None of the circuits requires the use of external power switches. The design eliminates the use of the large-valued filter capacitors you usually find in a switching regulator, because there is no need to smooth out high-frequency ripple in the LED current. Common to all circuits is the option of adding a dimming capability by introducing adjustable bias at an op-amp input through

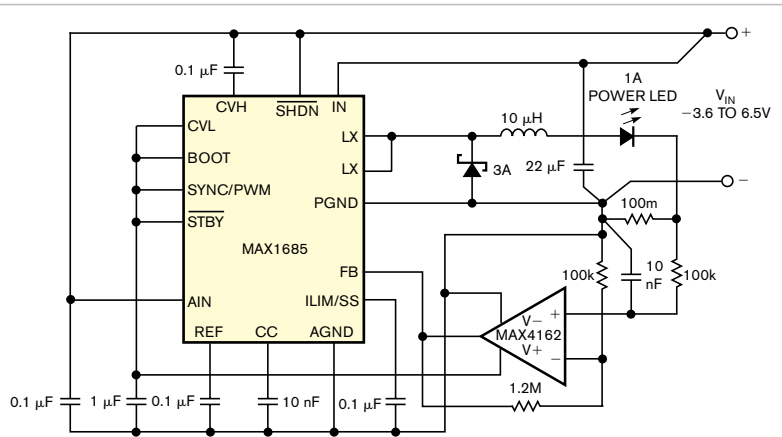


Figure 1 This miniature, 1A, high-power LED driver operates on 3.6 to 6.5V.

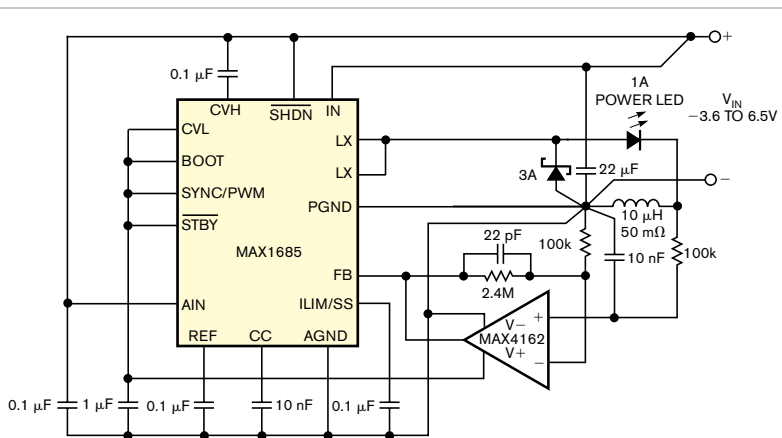


Figure 2 Similar to the circuit in Figure 1, this miniature, 1A, LED driver operates on 3.6 to 6.5V but requires no current-sense resistor.

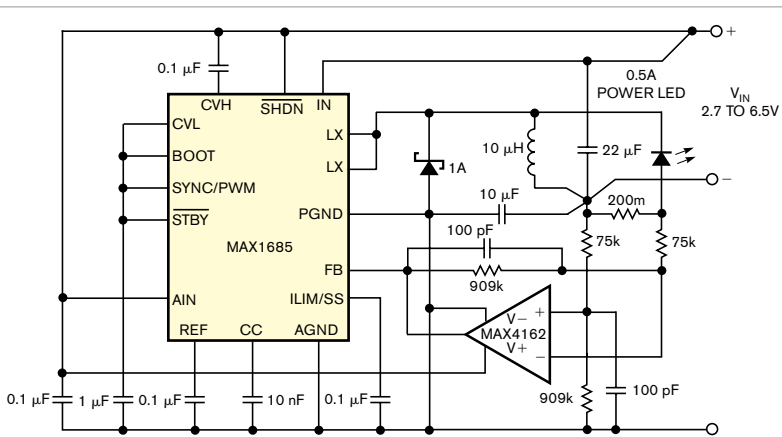


Figure 3 Another miniature, high-power LED driver delivers 0.5A and operates on 2.7 to 6.5V.

a resistor and a potentiometer powered from the internal regulator—the VD or CVL terminal, depending on the IC.

A high-frequency switching regulator powers the basic regulator circuit for LEDs (Figure 1). It operates with input voltages of 3.6 to 6.5V, drives a single LED with currents as high as 1A, and uses a current-sense resistor to control the current-regulation loop. The circuit of Figure 2 is similar, but, in place of a current-sense resistor, it employs the parasitic resistance of the inductor as a current-sensing element. Like the circuit in Figure 1, it operates with 3.6 to 6.5V inputs and drives one LED with currents as high as 1A.

For the single-LED circuit of Figure 3, the starting voltage of the MAX1685 defines the input range, which goes as low as 2.7V. Its maximum current capability is 0.5A versus 1A for the circuits in figures 1 and 2. The upper operating limit remains 6.5V. Once this circuit is operating, it maintains power to the LED even for input voltages as low as 1.7V. Applications for the circuits of figures 1, 2, and 3 include headlights, flashlights, and any other portable lights powered by three or four alkaline primary cells, three or four NiMH/NiCd secondary cells, or a single lithium secondary cell.

The circuits of figures 4 and 5 operate over 8 to 50V. Assuming a 12V system in which all the components are properly specified, these circuits can survive load dumps, thanks to the 76V absolute maximum rating for the IC's input-power terminal, V_{IN} . The maximum available current is 1A, and the circuits can drive as many as three LEDs in series, provided that you increase the lower limit of the operating range to 11.5V. These two circuits are

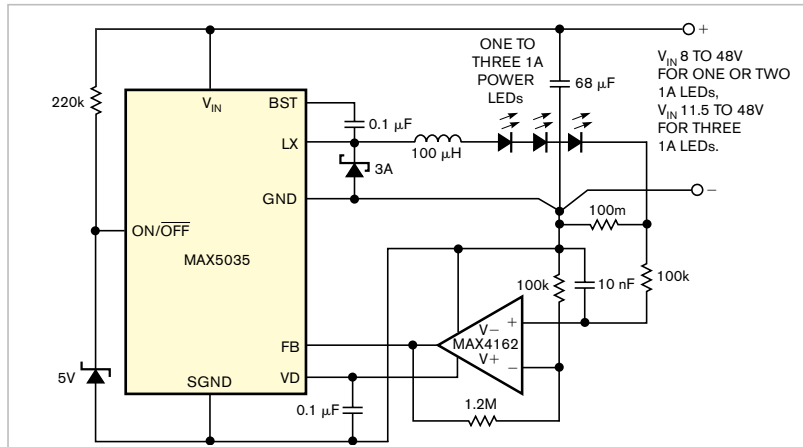


Figure 4 This miniature, 1A, LED-driver circuit operates on 8 to 50V and drives as many as three LEDs in series.

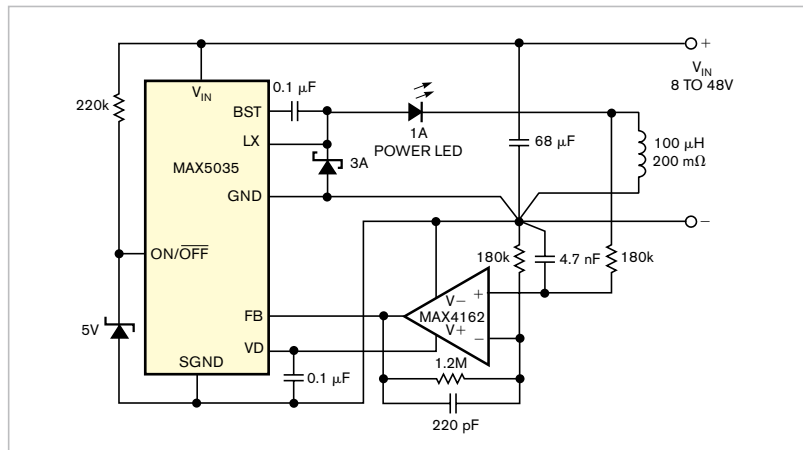


Figure 5 Otherwise similar to the circuit in Figure 4, this circuit requires no sense resistor.

similar, except for the use of the inductor resistance as a current sensor in Figure 5. The disadvantage of using the inductor resistance in this way is the resulting dependence of output current on temperature, due to the large temperature coefficient of copper re-

sistivity. The inductor winding is made of copper, and its dc resistance has a first-order temperature coefficient of 3.9 parts/1000°C. As a result, the regulated current decreases about 4% for each 10°C increase in operating temperature. EDN

Perform PSRR testing with analyzers having no dc-bias ports

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An amplifier's PSRR (power-supply-rejection ratio) is among the most commonly characterized pa-

rameters when analyzing the performance of an op amp. Examples of some noise sources on an amplifier's power-

supply pins include parasitic supply-line traces, their interaction with currents that the amplifier draws, and the noise that switching circuits sharing the same supply create. Both sources produce voltage-amplitude variations reproduced as noise signals at the amplifier's input pins.

Characterizing PSRR over frequency

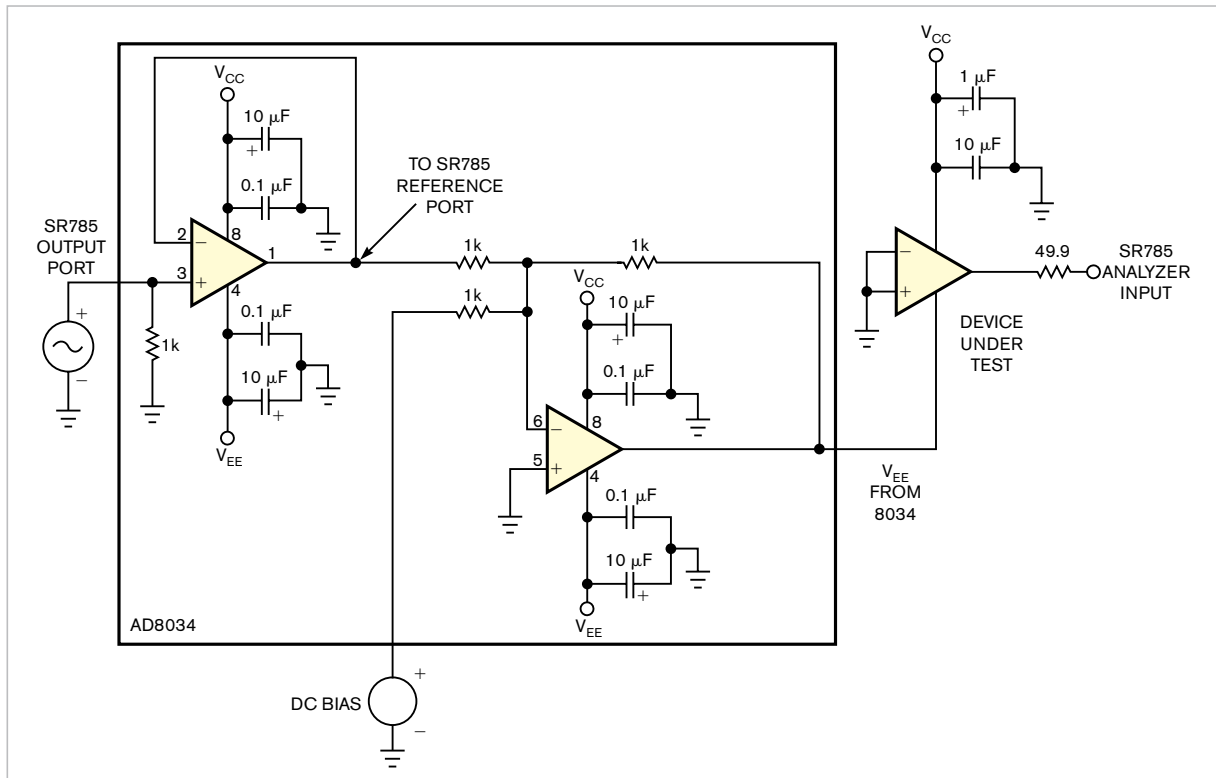


Figure 1 The AD8034 sums the analyzer's reference output and a dc bias, supplying the negative-supply voltage for the device under test to perform negative-PSRR analysis.

commonly involves the use of analyzers equipped with a dc-bias port, such as Agilent's (www.agilent.com) 8753. To measure negative PSRR, for example, the amplifier's $-V_s$ pin comes through Port 1, with the negative dc voltage through the bias port, of the 8753 with a superimposed sinusoid. To complete the measurement, you measure the amplifier's output on Port 2. Unfortunately, the 8753 doesn't measure frequencies below 30 kHz because of the limitations of the analyzer's internal bias, T. Additionally, most PSRR-versus-frequency plots begin at frequencies far below 30 kHz.

An alternative technique would involve the use of an analyzer that has no dc-bias port but that can characterize frequency response as low as 10 or even 1 Hz. One such analyzer is the Stanford Research Systems (www.thinksrs.com) SR785, which can make measurements better than -120 dB. One way of approaching this problem is to connect the output port of the SR785 to a buffer/inverting-summer circuit

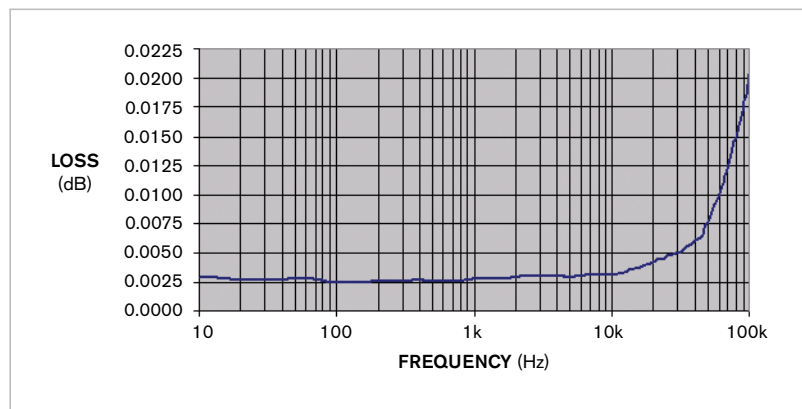


Figure 2 You can neglect any loss the AD8034 incurs. The response of the AD8034 buffer/inverting summer from 10 Hz to 10 kHz is approximately flat.

constructed with an Analog Devices (www.analog.com) AD8034.

Figure 1 illustrates a negative-PSRR test-circuit configuration. Pin 3 connects to the SR785 source-output port. Pin 1, which is V_{OUT} of the buffer amplifier, connects to the reference port of the SR785. Here, the first amp isolates the output port of the SR785 from

the dc bias and provides the sinusoidal output. The second amplifier within the AD8034 sums the dc bias and sinusoid, which it uses to feed the DUT's (device under test's) negative-supply pin. The **figure** omits all bypass capacitors at the DUT's negative-supply pin. A 1-kΩ resistor from Pin 3 to ground prevents the noninverting input from

floating. The positive terminal of the external dc-power supply feeds Pin 6 through a 1-kΩ resistor. Connecting the DUT's output to Channel 2A of the SR785 completes the test-circuit configuration.

Building the buffer/inverting summer with an AD8034 dual amplifier is a good choice because it has a supply range of 5 to 24V; a signal-frequency response well beyond 1 MHz; and a large capacitive-load-drive capability, allowing you to neglect the capacitance of test cables. Further, the AD8034 can deliver as much as 40 mA of load current.

To instill confidence that this buffer/inverting summer configuration works, **Figure 2** proves that you can neglect any loss that the AD8034 incurs. The **figure** demonstrates that the response of the AD8034 buffer/inverting summer of 10 Hz to 10 kHz is approximately flat with a loss of only 0.0025 dB, and the loss from 10 to 100 kHz is approximately 0.024 dB. **Figure**

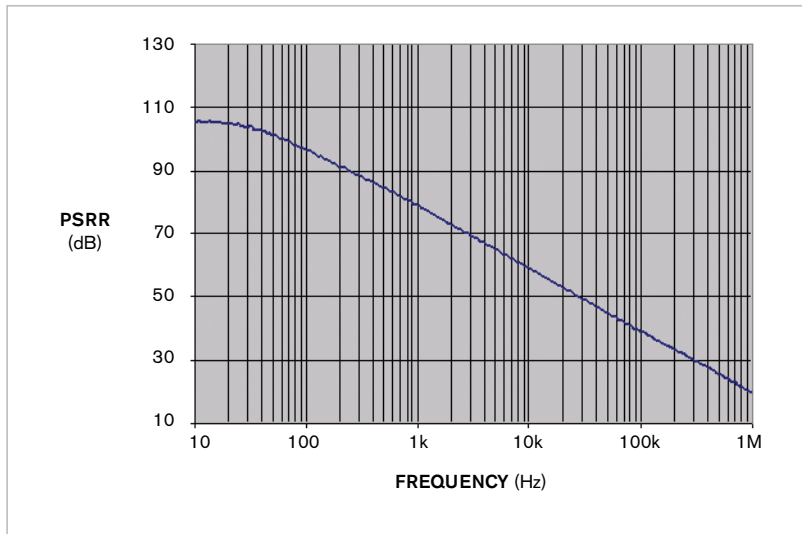


Figure 3 The negative-PSRR test results show responses beyond 100 kHz.

3 shows negative-PSRR test results. The Hewlett-Packard (www.hp.com) HP8753 provides the PSRR-versus-frequency responses beyond 100 kHz. You can measure positive PSRR (**fig-**

ures 4 and 5) by connecting Pin 3 to the SR785's output port. Pin 1, V_{OUT} of the buffer amplifier, connects to the reference port of the SR785. Here, you use the first amp to isolate the output

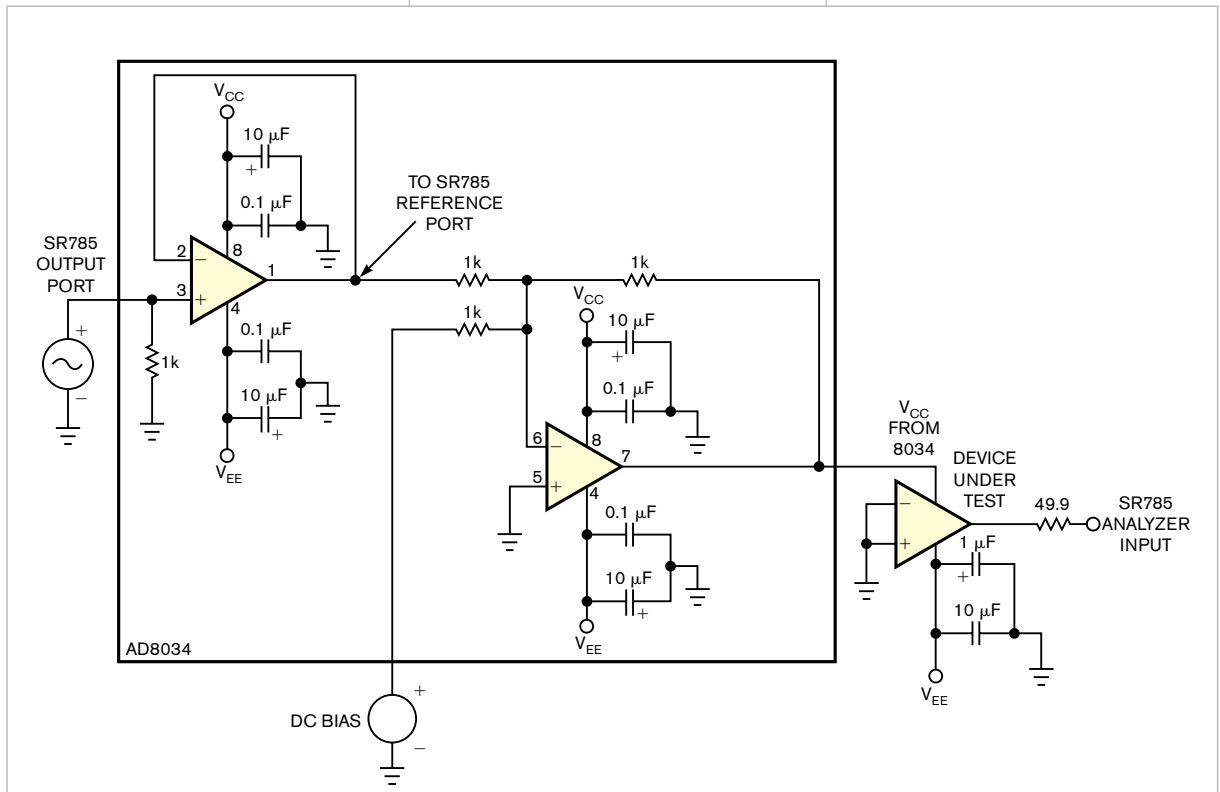


Figure 4 This test setup measures the positive PSRR. Note that bypass capacitors are absent from the device under test's positive-supply input.

port of the SR785 from the dc bias and provide the sinusoidal output. The second amplifier within the AD8034 sums the dc bias and sinusoid, which you use to feed the DUT's positive-supply pin. You must remove all bypass capacitors at the DUT's positive-supply pin. A 1-k Ω resistor from Pin 3 to ground prevents the noninverting input from floating. Feed the negative terminal of the external dc-power supply to Pin 6 through a 1-k Ω resistor. Connecting the DUT's output to Channel 2A of the SR785 completes the test-circuit configuration.

For the AD8034, assume that the DUT has a maximum supply voltage of $\pm 15\text{V}$, that you need to test negative PSRR, and that the DUT supplies $\pm 10\text{V}$. If you want to accommodate the maximum SR785 output of 5V peak, the first amplifier of the AD8034 needs enough head room to avoid clipping the $\pm 5\text{V}$ signal from the output port of the SR785. In this case, a supply setting for the AD8034 of 6 and -16V is sufficient to prevent any problems.

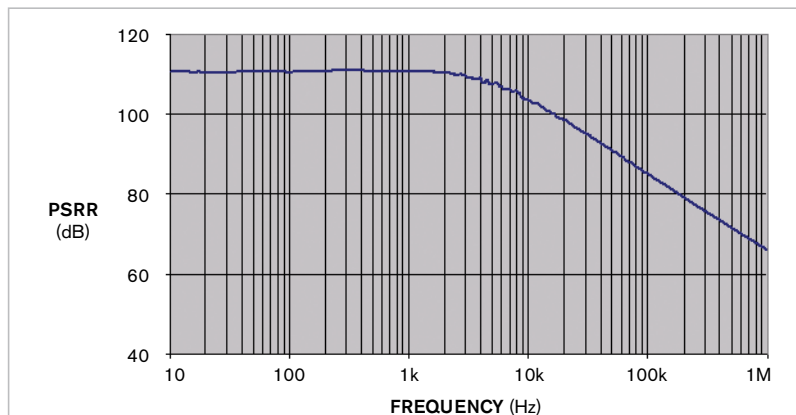


Figure 5 The positive-PSRR test results show responses beyond 100 dB.

This amount provides enough head room to accommodate the first amp of the AD8034, which handles a $\pm 5\text{V}$ signal centered at ground. The -16V accommodates the dc bias of -10V and the $\pm 5\text{V}$ signal centered at -10V at the output of the second amplifier of the AD8034. Positive PSRR is similar: Just set the AD8034 supplies to 16 and -6V for this example.

You might consider using separate

power supplies for the DUT and the AD8034 to simplify matters. However, you can use the same dc-power supply for the DUT to provide the dc-bias voltage at Pin 6 of the AD8034 buffer/inverting summer. Choose the output voltage of the SR785 or whichever analyzer you use so that the DUT operates within its linear region of operation. You can apply this technique to other applications. [EDN](#)