AN ORDINARY OP AMP CAN PROVIDE WIDE VOLTAGE SWINGS, BUT YOU MUST CONSIDER DESIGN ISSUES AND TRADE-OFFS.

Bootstrapping your op amp yields wide voltage swings

Getting 100V p-p from a monolithic op amp is just one example of what you can achieve by bootstrapping power supplies. “Bootstrapping” in this context is simply a method of controlling a device’s supply voltages based on its output.

In the circuit of Figure 1, the system supply voltages, \( V_{CC} \) and \( V_{EE} \), are fixed, but the device supply voltages, \( V_{CO} \) and \( V_{EO} \), change dynamically as a function of \( V_{OUT} \). The op amp can then cover peak-to-peak voltage swings far greater than the total voltage you apply across its supply rails.

The maximum voltage that you can apply across a monolithic op amp’s supply rails, which the manufacturer’s IC process determines, is generally around 30 to 40V. Figure 2 illustrates some results in which the voltage difference, \( V_{CO} - V_{EO} \), remains constant at approximately 30, and the absolute voltages, \( V_{CO} \) and \( V_{EO} \), swing more than 70V to follow \( V_{OUT} \). Two emitter followers and two resistor pairs generate \( V_{CO} \) and \( V_{EO} \) (Figure 1). (The two diodes shown are added merely to improve output voltage swing as described in the following circuit analysis.)

WHY BOOTSTRAP?

Op amps offer a simple and effective alternative to discrete-transistor designs and have proved their usefulness in a range of applications. However, some applications require output-voltage swings greater than those that a standard monolithic op amp can generate.

The most direct approach to achieving these wide voltage swings is to design the amplifier using discrete transistors. This approach allows you the flexibility to customize the amplifier for the application. You can also easily achieve high output power with this method. However, discrete-transistor designs require more of a designer’s time and effort than other approaches and require more parts, complicating manufacturing. It is also difficult to achieve precision in these designs because of device matching and temperature gradients.

High-voltage op-amp modules provide an alternative that considerably eases a designer’s task. These devices are just as easy to use as monolithic op amps but are generally in the form of hybrid modules, thereby allowing high-voltage (and often high-power) operation. One strong advantage of these modules over discrete designs is that they have factory-specified performance, relieving the designer’s task in characterizing performance. The most significant disadvantage of these hybrid modules is their cost. Also, far more monolithic op amps than hybrid op amps are available. A hybrid often cannot meet the performance demands of a design. In this case, bootstrapping techniques open the list of available devices to many hundreds.

Bootstrapping designs require more effort but are significantly lower cost than high-voltage op-amp modules. An all-discrete design might offer you still lower cost, but the additional design and characterization effort you must make often offsets the lower cost.

A typical bootstrapping circuit uses fixed system-supply voltages, \( V_{CC} \) and \( V_{EE} \), but the device-supply voltages, \( V_{CO} \) and \( V_{EO} \), change as a function of the output voltage.
this cost reduction. A variety of monolithic op amps is available, and each features fully factory-specified performance that you can apply even when a bootstrapping network surrounds the op amp. Extending the voltage range of standard op amps by bootstrapping offers you flexibility and maintains a “canned” set of performance parameters.

For any high-voltage-amplifier design, you should consider all three techniques. This article offers a detailed look at bootstrapping, the least documented method of the three (Table 1).

HOW DOES BOOTSTRAPPING WORK?

Ignoring the diode drops and $V_{BE}$ drops for a moment, you can express $V_{CO}$ and $V_{EO}$ in Figure 1 as:

\[ V_{CO} = \frac{V_{CC}R_1 + V_{OUT}R_2}{R_1 + R_2} \]  
\[ V_{EO} = \frac{V_{EE}R_3 + V_{OUT}R_4}{R_3 + R_4} \]

Next, add the effects of transistor $V_{BE}$, but omit the optional diodes from the circuit, and you get a more realistic representation of the device supply voltages:

\[ V_{CO} = \frac{V_{CC}R_1 + V_{OUT}R_2}{R_1 + R_2} - 0.6 \]  
\[ V_{EO} = \frac{V_{EE}R_3 + V_{OUT}R_4}{R_3 + R_4} + 0.6 \]

In this case, you solve for the maximum output voltage that you can achieve with an ideal rail-to-rail op amp as:

\[ V_{OUT} = V_{CC} - 0.6 \]

Thus, the peak output voltage increases by $0.6 \times (R/R_{V})V$.

In a symmetrical system (in which ground is equidistant between $V_{CC}$ and $V_{EE}$), let $R_3=R_1$ and $R_4=R_2$. Making this substitution in equations 6 and 7, you can see that the difference between $V_{CO}$ and $V_{EO}$ is constant if you assume that $V_{CC}$ and $V_{EE}$ are constant.

\[ V_{CO} - V_{EO} = \frac{R_1}{R_1 + R_2} (V_{CC} - V_{EE} - 1.2) \]

So, for the example in Figure 2, where $V_{CC}=60V$, $V_{EE}=-60V$, $R_1=10$ kΩ, and $R_2=28$ kΩ, the voltage across the op amp remains constant at about 31V throughout the 100V p-p swing of the output.

As with all op-amp applications, you must ensure that the voltage at the noninverting input always remains within the device’s common-mode input range. Whereas this task is trivial in standard op-amp circuits with fixed power supplies, it requires more insight for bootstrapping configurations, in which the op-amp supply rails change with the output. Even as $V_{CO}$ and $V_{EO}$ change, $V_{IN}$ must always remain between them (Figure 2). You must guarantee this situation by design, or a latched condition might occur. To ensure that your design meets input common-mode range under all conditions, you must address dc conditions, transient conditions, phase reversal, and power-up conditions.

DC CONDITIONS

When considering dc gain, remember that the feedback network of a bootstrapped op-amp circuit works in the same way as that of any other op-amp gain stage. The gain of the Figure 1 circuit is simply $A_v = \frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_1}{R_2}$. In configurations in which $V_{CC} - V_{EE}$ is less than twice $V_{CO} - V_{EO}$, you can run the circuit at any gain, including inverting gains. But for wider system supply rails and to achieve wider output swings, you must use a noninverting configuration and carefully select gain. If you set gain too high, you will exceed the op amp’s input common-mode range, which will likely result in latch-up of the bootstrap network.

This simulation shows the circuit of Figure 1, providing a 100V p-p sine-wave using an AD820 op amp.

\[ V_{EO} = (V_{EE} + 0.6)R_3 + V_{OUT}R_4 \]

\[ R_3 + R_4 \]

<table>
<thead>
<tr>
<th>TABLE 1—TECHNIQUES FOR ACHIEVING HIGH-VOLTAGE AMPLIFICATION</th>
</tr>
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<tbody>
<tr>
<td><strong>High-voltage op-amp module</strong></td>
</tr>
<tr>
<td><strong>Bootstrapping with monolithic op amp</strong></td>
</tr>
<tr>
<td><strong>Discrete-transistor monolithic op amp</strong></td>
</tr>
</tbody>
</table>
in Figure 2 would cause $V_{CO}$ to exceed $V_{CM}$ at its peak and $V_{CO}$ to exceed $V_{CM}$ on the negative side. This situation clearly violates the op amp's input common-mode range in that both power supplies are farther from ground than its input. Luckily, you can easily avoid this condition. With a low enough gain, the output stage saturates before the input stage, and the power-supply rails stop increasing before they exceed the input (Figure 3).

Assuming that you have a symmetrical system with positive gain (in which $R_c$ is “grounded” halfway between $V_{CC}$ and $V_{EE}$, the following two equations are sufficient to ensure that you avoid the above condition:

\[ A_V \leq \frac{V_{CC} - 0.6}{V_{CC} - 0.6 - (V_{CO} - V_{EO}) + V_{HRH}} \]  
\[ A_V \leq \frac{V_{EE} + 0.6}{V_{EE} + 0.6 + (V_{CO} - V_{EO}) - V_{HRH}} \]

$V_{HRH}$ is the op amp’s input head-room voltage—the difference between its positive power supply and its resulting maximum common-mode input voltage—on the high side, and $V_{HRH}$ is the input head-room voltage on the low side.

You can achieve greater gains than those that the above equations allow by cascading multiple stages. Alternatively, you can configure one stage to operate at higher gains using a later-described method.

**TRANSIENT CONDITIONS**

Once you select gain to keep $V_{CM}$ within the op amp’s common-mode input range under dc conditions, you must consider transient signals. The op amp’s output has a finite slew rate, and its supplies are a function of its output. Thus, a step function at the op amp’s input can easily exceed the amp’s supply range. You should not directly apply a square wave to the op amp because it would exceed the device’s supplies when the op amp was just beginning to slew. To avoid the latched condition that this situation might cause, place a slew limit on the signal feeding the amplifier to limit transients to less than or equal to the op amp’s slew rate (Figure 4). To guarantee adequate limitation with a simple RC filter, choose the following RC time constant:

\[ R_{IN}C_{IN} \geq \frac{V_{STEP}}{SR} \]

If you properly select gain, the amplifier’s output will saturate before its input common-mode range is violated.

**POWER-UP CONDITIONS**

Because bootstrapped amplifiers are sensitive to latch-up, you must pay additional attention to power-supply sequencing of these circuits. For instance, if the positive rail comes up a few milliseconds before the negative rail, it can send the device supply voltages, $V_{CO}$ and $V_{EO}$, toward the positive rail while the input remains at ground, thereby violating the op amp’s input common-mode range. The best way to avoid the latch-up condition that this situation can cause is to keep the input at ground potential and simultaneously bring the power supplies up (Figure 5).

**EXPANDING POSSIBILITIES**

The common theme of the above points of concern is the op amp’s input common-mode range. With proper attention to this detail, you can create bootstrapping circuits with wide-ranging configurations that go far beyond these simple examples.

Consider, for instance, a design with high gain and wide output swing. If you need greater gain from a single stage than the gain you can achieve with the circuit in Figure 1, then you may find the circuit in Figure 6 useful. In this configuration,
The gain from $V_{CM}$ to $V_{OUT}$ is the largest possible gain you can achieve based on equations 14 and 15, which are identical to equations 10 and 11:

$$A_{OUT/CM} = \frac{V_{OUT}}{V_{CM}} = 1 + \frac{R_F}{R_G}, \quad (13)$$

$$A_{OUT/CM} \leq \frac{V_{CC} - 0.6}{V_{CC} - 0.6 - (V_{CO} - V_{EO}) + V_{THRL}}, \quad (14)$$

and

$$A_{OUT/CM} \leq \frac{V_{EE} + 0.6}{V_{EE} + 0.6 + (V_{CO} - V_{EO}) - V_{HRL}}. \quad (15)$$

If a negative value appears on the right side of these inequalities, then you can operate the circuit in Figure 1 at any gain, and you need not add $R_B$ to the circuit. Otherwise, set $A_{OUT/CM}$ to the highest gain that the above inequalities allow. You can then increase the overall gain of the stage from $V_{IN}$ to $V_{OUT}$ to virtually any gain by adding the resistor, $R_B$. The expression for this overall gain is:

$$A_{OUT/IN} = \frac{V_{OUT}}{V_{IN}} = \frac{R_G R_B + R_F R_B}{R_G R_B - R_F R_B} = 1 + \frac{R_E}{R_G} \frac{1}{1 - \frac{R_E}{R_B}}. \quad (16)$$

But for you to solve for $R_F$ and $R_B$, it is easier to express the relation in terms of the two gains:

$$R_F = \frac{A_{OUT/IN}}{A_{OUT/CM}} \cdot \frac{1}{1 - \frac{A_{OUT/CM}}{A_{OUT/IN}}}. \quad (17)$$

The condition for equations 16 and 17 is that $A_{OUT/IN}$ must be greater than $A_{OUT/CM}$, and thereby that $R_F/R_B$ must be greater than $R_E/R_G$. If you do not meet this condition, the gain equation will “blow up,” indicating the circuit’s instability.

As with the first example, this circuit requires slew-rate limiting for transient signals. If the slew rate of the incoming signal exceeds $1 R_F/V_{THRL}$ times the slew rate of the op amp, then you should add $C_{IN}$ to form a slew-limiting RC time constant,

$$f_p = \frac{1}{2 \pi C_{IN} R_B - \frac{R_F}{2 \pi C_{IN} R_B}}. \quad (20)$$

To limit noise bandwidth, you should place $f_p$ as low as possible without affecting desired signals.

**OFFSETS, NOISE, AND NONIDEAL BEHAVIOR**

Because of the two feedback networks, an error analysis of the circuit in Figure 6 is somewhat more complex than for a basic op-amp gain stage. Because the mechanics of the error analysis are beyond the scope of this article, the following **equations** omit derivations. For consistency, all errors are referred to the output. To obtain input-referred errors, you simply divide by the gain of the stage. Noise gain is also the gain that amplifies the op amp’s input offset voltage to the output. At low frequencies, including dc, the noise gain of the circuit in Figure 6 is:

$$A_N = \frac{V_{OUT(NOISE)}}{V_{NOISE}} = \frac{(R_F + R_G) (R_B + R_B)}{R_C R_B - R_F R_B}, \quad (21)$$

where $V_{NOISE}$ can be either the op amp’s input offset voltage (for dc analysis) or the op amp’s input-voltage noise. For wideband voltage-noise analysis, the following pole and zero further define this transfer function from $V_{NOISE}$ to $V_{OUT}$:

$$f_p = \frac{1}{2 \pi C_{IN} R_B - \frac{R_F}{2 \pi C_{IN} R_B}}, \quad (22)$$

$$f_z = \frac{1}{2 \pi C_{IN} R_B} + \frac{1}{2 \pi C_{IN} R_B}. \quad (23)$$

To determine output error due to op-amp offset voltage, you simply multiply the op amp’s $V_{OS}$ by the circuit’s dc noise gain, $A_N$ (Equation 21). You should use the same approach with the op amp’s low-frequency (1/f) noise to refer it to the output. Solving for wideband rms output noise is more complex, but you can simplify the task if $f_p$ and $f_z$ are far enough...
apart to let you assume a single-pole noise roll-off. In this case, simply multiply the op amp's input voltage-noise density by \( A_n \sqrt{f_p} \) to obtain the resulting output-referred rms noise voltage (Reference 1).

The effects of the op amp's input bias currents and current noise are similar to those of offset voltage and voltage noise in that you translate them into output-referred voltage errors. One difference between these effects is that, in the case of current errors, both the inverting and noninverting inputs induce separate errors in the output. You can sum—as a "root sum of squares" for noise—all output-referred errors to obtain a total output error.

For dc and low frequencies, you translate the input current errors to output referred voltage errors using the following equations:

\[
V_{\text{OUT}(\text{NOISE+})} = \frac{R_{\text{G}}R_{\text{B}}(R_{\text{B}} + R_{\text{G}})}{R_{\text{G}}R_{\text{B}} - R_{\text{IN}}R_{\text{IN}}} I_{\text{NOISE+}}; \quad (24)
\]

\[
V_{\text{OUT}(\text{NOISE-})} = \frac{R_{\text{F}}R_{\text{G}}(R_{\text{B}} + R_{\text{G}})}{R_{\text{G}}R_{\text{B}} - R_{\text{F}}R_{\text{IN}}} I_{\text{NOISE-}}; \quad (25)
\]

where \( I_{\text{NOISE+}} \) is the input current noise or input bias current at the noninverting input, \( I_{\text{NOISE-}} \) is the same for the inverting input, and \( V_{\text{OUT}(\text{NOISE+})} \) and \( V_{\text{OUT}(\text{NOISE-})} \) are the output-referred errors that result from each.

Again, for wideband-noise analysis, you must consider the effects of \( C_{\text{IN}} \). For both the noninverting and inverting current-noise transfer functions, a pole appears at:

\[
f_p = \frac{R_{\text{G}}R_{\text{B}} - R_{\text{IN}}R_{\text{F}}}{2\pi C_{\text{IN}}R_{\text{IN}}R_{\text{G}}R_{\text{B}}}; \quad (26)
\]

which applies to equations 24 and 25. For the inverting current noise only, a zero appears at:

\[
f_z = \frac{R_{\text{B}} + R_{\text{IN}}}{2\pi C_{\text{IN}}R_{\text{IN}}}; \quad (27)
\]

which applies only to Equation 25.

As with voltage-noise analysis, you can simplify the transfer function by assuming a single-pole roll-off. In this case, to determine the rms output noise resulting from input current noise, you can simply multiply the values obtained from equations 24 and 25 for wideband noise density by \( A_n \sqrt{f_p} \) where \( f_p \) comes from Equation 26.

You have considered offset voltage and two bias currents as sources of dc error and voltage noise and two current noises as sources of noise, and you have referred each source of error to the output. Now, you must sum these errors. For dc errors, this summing takes the form of a simple sum of the magnitude of each error term. But the noise errors add instead as a root sum of squares; that is:

\[
V_{\text{TOTAL}} = \sqrt{V_1^2 + V_2^2 + V_3^2}. \quad (28)
\]

In most cases, either the op amp's voltage noise or one of its current noises dominates the total output noise. The smaller output-referred noise terms generally make a negligible contribution to total noise. However, you should also consider the Johnson noise of the signal path resistors, \( R_{\text{G}}, R_{\text{IN}}, \) and \( R_{\text{B}} \) (Reference 1).
ence 1). As with the other noise sources, you should refer these noises to the output and add them as a root sum of squares with the other output noise values.

LOW DEVICE-SUPPLY VOLTAGE

It may be intuitively clear that it is best to set the op amp’s device supply voltage, \( V_{CC} - V_{EE} \), near its maximum specified operating voltage when you are bootstrapping for wide signal swings. But to explicitly show how a device’s supply voltage affects performance, consider the following configurations. Both have ±60V system power supplies, and both require a gain of 10. However, the device supply is 30V in one case and 10V in the other (see Tables 2 and 3, respectively, at www.ednmag.com). To design these two circuits, you select \( R_1 \) and \( R_2 \) using Equation 9 to achieve the desired device supply voltage.

You then choose \( R_1 \) and \( R_2 \) to achieve a gain as high as possible from \( V_{CM} \) to \( V_{OUT} \) and to meet the conditions of Equations 14 and 15. For this example, assume op-amp input head room of 1V from each rail in both cases. Equation 17 then gives us the ratio of \( R_{IN} / R_1 \) that will result in a gain of 10 from \( V_{IN} \) to \( V_{OUT} \) and the two circuits are complete. To determine the noise gain of each, use the component values in Equation 21. The results of this exercise show that, by reducing by one-third the device’s supply voltage, you almost quadruple the output error.

OP-AMP SELECTION

Bootstrapping is a way to use just about any monolithic op amp to output wide signal swings. Even so, you should choose an op amp that can operate with a fairly high supply voltage to begin with; the above comparison shows the advantage of this approach. Hundreds of commonly available op amps operate from 30 or 40V power supplies, so avoid choosing one with a maximum \( V_{EE} \) of only 5 or 10V. Beyond that, individual system requirements determine what precision, speed, and other parameters the op amp should offer. Some important parameters to consider include the following:

Output current drive: At high-voltage swings, even a 2-kΩ load can pull significant current.

Input bias current: Large-value resistors in feedback paths make it useful to have a FET input stage.

Slew rate: Slew-rate limiting can distort large-magnitude ac signals.

Table 4 at www.ednmag.com shows a few op amps from Analog Devices suitable for bootstrapping, including both FET and bipolar input stage devices. This list is not comprehensive. Each bootstrapping application has a unique set of requirements for op-amp parameters, so if what you require doesn’t appear on this list, look through various vendors’ op-amp offerings. Regardless of your application, the choice of op amp requires knowledge of design requirements. Use the equations to answer the following questions:

What slew rate does your design need? (See Equations 19 and 20.)

How will offset voltage and bias current affect output error? (See Equations 21, 24, and 25.)

Will a rail-to-rail input stage or wider device supply (maximum \( V_{SD} \)) significantly improve performance? (See Equations 9, 14, 15, 17, 21, 24, and 25.)

Also keep in mind that a FET input stage can allow you to use larger value resistors in the feedback network with minimal impact on total output error.

CONSIDER A COMPOSITE APPROACH

When looking at the op amps in Table 4, you might notice that the devices with the greatest output current drive, \( I_{OUT} \), are not those with the best dc precision, \( V_{CM} \) and \( I_v \). This situation is largely true for all op amps. However, an alternative exists for precision applications needing more than a few tens of milliamperes of output current drive. By connecting two op amps in a composite configuration (Figure 8), you can use the slew rate and output current drive capabilities of one device and the dc precision of the other. Because the input amp globally closes the feedback loop, the imprecise output amp can provide the necessary muscle without adding error to the system.

The feedback network, consisting of \( R_p, R_n, R_{IN}, R_{NB}, C_{NB} \), and the bootstrap network, \( R_1 \) through \( R_4 \) plus two diodes in Figure 8, are the same as those in Figure 6. Further, the equations for calculating component values and error terms apply equally to both figures. The only component value in Figure 8 that the Figure 6 analysis omits is the new feedback resistor, \( R_{FB} \). Anyone familiar with current feedback op amps will recognize the importance of this resistor. If you use an op amp other than the AD811 you have to choose the value of \( R_4 \) based on the op amp’s data sheet, in which a table of recommended values for different gains generally appears. Values larger than those recommended reduce the op amp’s bandwidth and slew rate. Values
smaller than those recommended degrade stability, possibly resulting in oscillations. If you use a voltage-feedback op amp as the output device, the value of $R_F$ can be as low as 0Ω or a shorted connection.

Table 5 (at www.ednmag.com) shows a few op amps that fit well into Figure 8’s circuit. An AD825 or OP97 performs well as the input device, and an AD811 or AD815 current-feedback device offers advantages in the output stage. Current-feedback op amps are generally poor choices for bootstrapping because of their sensitivity to feedback-network impedance. However, they make excellent output amps in this composite circuit because they connect in a simple unity-gain mode, and the input amp handles the high-impedance global-feedback paths.

The output current drive, $I_{OUT}$, of the input amp makes no difference because the output op amp drives the load (Table 5). Also, the input offset voltage, $V_{OS}$, of the output amp causes no output error because the input amp closes the global-feedback loop. By carefully selecting two op amps for a composite configuration, you can achieve performance that is unavailable from any single device.

TIE UP LOOSE ENDS

You must now consider the transistors that form the bootstrapping network itself. You have as much flexibility in selecting these transistors as you have in selecting the op amps. The primary concerns are:

- Breakdown voltages, $V_{CB}$ and $V_{CE}$ (an important parameter for obvious reasons);
- $h_{FE}$: Higher beta allows larger value resistors for $R_1$ through $R_4$.
- Power dissipation: In most cases, the transistors must dissipate more power than the op amps.

Tests for this article use ZTX653 and ZTX753 from Zetex (www.zetex.com). These transistors have $h_{FE}$ greater than 100 and breakdown voltages greater than 100V. Their power dissipation is not of great concern for test purposes, because the test circuits were not driving significant loads. In place of the diodes in series with $R_1$ and $R_3$, the tests simply used these same transistors in a “diode-connected” configuration.

You can use MOSFETs instead of bipolar transistors. The primary disadvantage of MOSFETs in these circuits is headroom. Therefore, if your design doesn’t require that its output voltages approach the system supply rails, then MOSFETs might be a good choice. When you use MOSFETs, you should, if possible, replace the diodes that are in series with $R_1$ and $R_3$ with a few series diodes to approximate the MOSFET’s $V_{GS}$ voltage. Alternatively, you can diode-connect a second pair of MOSFETs in place of the diodes to achieve the same basic function.

As with most analog circuits, lack of adequate power-supply decoupling can appreciably degrade performance, especially when transient signals need accurate amplification. Unlike typical op-amp circuits, however, a bootstrapped circuit has local power supplies, $V_{CO}$ and $V_{EE}$, that move dynamically with signal voltage. Therefore, you should not decouple these nodes directly to ground. The best place for local bypass capacitors, in this case, is between the bases of the two bootstrap-network transistors (Figures 1, 6, and 8). In addition to this local decoupling of the device supplies, $V_{CO}$ and $V_{EE}$, you should decouple the system supplies, $V_{CC}$ and $V_{EE}$, directly to ground.

Reference


Author biographies

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**TABLE 2 - KEY DESIGN PARAMETERS**

FOR A 30V DEVICE SUPPLY CONFIGURATION

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<td>$R_2 = R_4$</td>
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**TABLE 3 - KEY DESIGN PARAMETERS**

FOR A 10V DEVICE SUPPLY CONFIGURATION

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<th>parameter</th>
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**TABLE 4 - SOME BOOTSTRAP FRIENDLY OP AMPS AND THEIR KEY PARAMETERS**

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<th>$V_{HRL}$</th>
<th>$V_{OHRH}$</th>
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<th>$I_{OUT}$</th>
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<td>0.01V</td>
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<td>15mA</td>
<td>1.9MHz</td>
<td>3V/µs</td>
<td>36V</td>
</tr>
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<td>10pA</td>
<td>1.5V</td>
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<td>1.6V</td>
<td>1.6V</td>
<td>26mA</td>
<td>46MHz</td>
<td>140V/µs</td>
<td>36V</td>
</tr>
<tr>
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<td>1mV</td>
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<td>2V</td>
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<td>2.4V</td>
<td>50mA</td>
<td>24MHz</td>
<td>250V/µs</td>
<td>36V</td>
</tr>
<tr>
<td>AD845</td>
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<td>700µV</td>
<td>750pA</td>
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<td>2V</td>
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<td>2.5V</td>
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<td>16MHz</td>
<td>100V/µs</td>
<td>36V</td>
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<td>25V/µs</td>
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<td>50V/µs</td>
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<td>50mA</td>
<td>50MHz</td>
<td>350V/µs</td>
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</tr>
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<td>AD841*</td>
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<td>3V</td>
<td>5V</td>
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<td>2V</td>
<td>2V</td>
<td>25mA</td>
<td>600KHz</td>
<td>0.3V/µs</td>
<td>44V</td>
</tr>
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<td>2nA</td>
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<td>0.09V</td>
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<td>15V/µs</td>
<td>36V</td>
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<td>2.2V</td>
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<td>2.8V/µs</td>
<td>44V</td>
</tr>
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<td>Bip</td>
<td>50µV</td>
<td>1.2nA</td>
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<td>30µV</td>
<td>30pA</td>
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<td>1V</td>
<td>20mA</td>
<td>900KHz</td>
<td>0.2V/µs</td>
<td>40V</td>
</tr>
</tbody>
</table>

* Device is immune to potential phase reversal.

**TABLE 5 - SOME COMPOSITE-FRIENDLY OP AMPS AND THEIR KEY PARAMETERS**

<table>
<thead>
<tr>
<th>part #</th>
<th>type</th>
<th>$V_{OS}$</th>
<th>$I_B$</th>
<th>$V_{HRH}$</th>
<th>$V_{HRL}$</th>
<th>$V_{OHRH}$</th>
<th>$V_{OHRL}$</th>
<th>$I_{OUT}$</th>
<th>GBP</th>
<th>SR</th>
<th>maxV $S$</th>
</tr>
</thead>
<tbody>
<tr>
<td>AD825*</td>
<td>FET</td>
<td>1mV</td>
<td>10pA</td>
<td>1.5V</td>
<td>1.5V</td>
<td>1.6V</td>
<td>1.6V</td>
<td>26mA</td>
<td>46MHz</td>
<td>140V/µs</td>
<td>36V</td>
</tr>
<tr>
<td>OP97*</td>
<td>Bip</td>
<td>30µV</td>
<td>30pA</td>
<td>1V</td>
<td>1V</td>
<td>1V</td>
<td>1V</td>
<td>20mA</td>
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<td>0.2V/µs</td>
<td>40V</td>
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<td>3V</td>
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<td>140MHz</td>
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<tr>
<td>AD815*</td>
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<td>1V</td>
<td>1V</td>
<td>750mA</td>
<td>120MHz</td>
<td>900V/µs</td>
<td>36V</td>
</tr>
</tbody>
</table>

* Device is immune to potential phase reversal.