Application Report

SBOA121–November 2009

Making the Most of a Low-Power, High-Speed Operational Amplifier

Xavier Ramus — *High-Speed Products*

ABSTRACT

High-speed, high-performance operational amplifiers tend to be associated with high power dissipation. This application note compares the relative performance of several low-power, high-speed operational amplifiers and describes trade-offs to balance performance with low quiescent power dissipation.

Contents

List of Figures

1 Introduction

Achieving undeniably portable instrumentation requires the use of a low power dissipation operational amplifier. It is extremely difficult to balance various low-power application circuits while maintaining high system performance. This difficulty is partly the result of the ac performance degradation, such as slew rate and bandwidth, as the amplifier quiescent current is reduced. In this application report, we first start with bandwidth performance comparison, then move to harmonic distortion, and see what actual losses we must address with low quiescent current devices. This document concludes with the analysis of several application circuits as well as reviewing the advantages of each amplifier topology for each application.

2 Bandwidth Performance Comparison

A useful practice for voltage-feedback amplifiers (VFAs) is to do a side-by-side comparison of the gain bandwidth product (GBP) and the quiescent current for a given op amp.

Note that for different op amp architectures such as the current-feedback amplifier (CFA), this approach does not work because a CFA does not have a GBP. Instead, the bandwidth of the highest gain shown in the product data sheet is used to mirror the GBP definition of the VFA.

The method used to derive the GBP from the data sheet is provided below for a high-speed VFA. It is extremely rare to have the unity-gain bandwidth be equal to the gain bandwidth product. This rarity is as much the result of the practice of not having overcompensated amplifiers as it is the effects of package parasitics that extend the unity-gain bandwidth. For high-speed amplifiers, you must measure the GBP at gain of 40dB or greater in the open-loop gain graph. The example performance graph shown in Figure 1 is taken from the OPA890 data sheet.

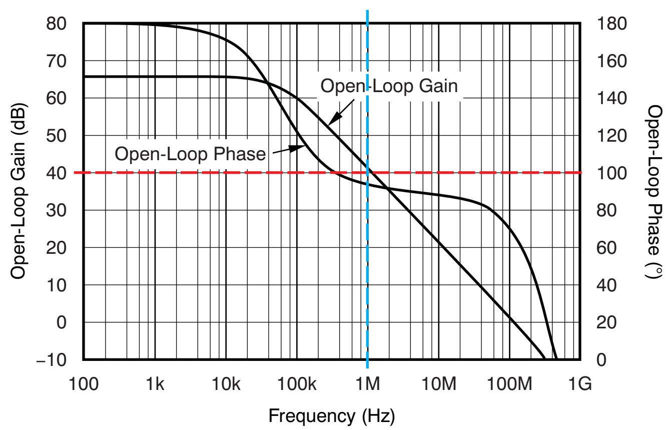

Figure 1. OPA890 Open-Loop Gain and Phase Performance

In Figure 1, at the intersection of the red line and the blue line, you measure the GBP; in the case of the OPA890, this measurement results in a bandwidth of 1.3MHz for a 100V/V gain (40dB), or 130MHz GBP. This number is consistent with the number reported in the ±5V electrical specifications table of the device data sheet.

Table 1 summarizes the bandwidth and quiescent current information for various low-power, high-speed VFAs and CFAs.

Table 1. VFA and CFA Comparison

VFA Device	Gain Bandwidth Product (MHz)	Quiescent Current (mA)	
OPA890	130	1.1	
OPA2889	75	0.46	
THS4281	38	0.8	
CFA Device	**Bandwidth (MHz)**	**Gain[1] (V/V)**	**Quiescent Current (mA)**
OPA684	71	100	1.7
OPA683	35	100	0.94

(1) Highest gain recommended for this device.

By comparing the OPA890 with the OPA683, we can see the clear advantage of using a current-feedback amplifier for high-gain applications. If we tried to use the OPA890 in a 100V/V application, we would achieve a 1.3MHz, –3dB bandwidth. This performance pales in comparison with the 35MHz bandwidth of the OPA683, or for slightly more quiescent power dissipation, the 71MHz bandwidth of the OPA684.

Note that all of these devices are capable of operating on a +3V supply; at this low supply voltage, though, only the THS4281 has the full output voltage swing because it is a rail-to-rail input/output (RRIO) op amp.

The application described in Figure 2 takes advantage of the low-power CFA OPA684 for a video summing amplifier. In this configuration, the OPA684 provides greater than 120MHz bandwidth. If a VFA (such as the OPA890) were used instead, the maximum bandwidth would be limited to 15.5MHz as a result of the gain bandwidth product dependency for the VFA architecture. To achieve the same bandwidth as the OPA684, a 1GHz GBP VFA would be required. Although such devices are readily available, the quiescent current would have to be increased tenfold.

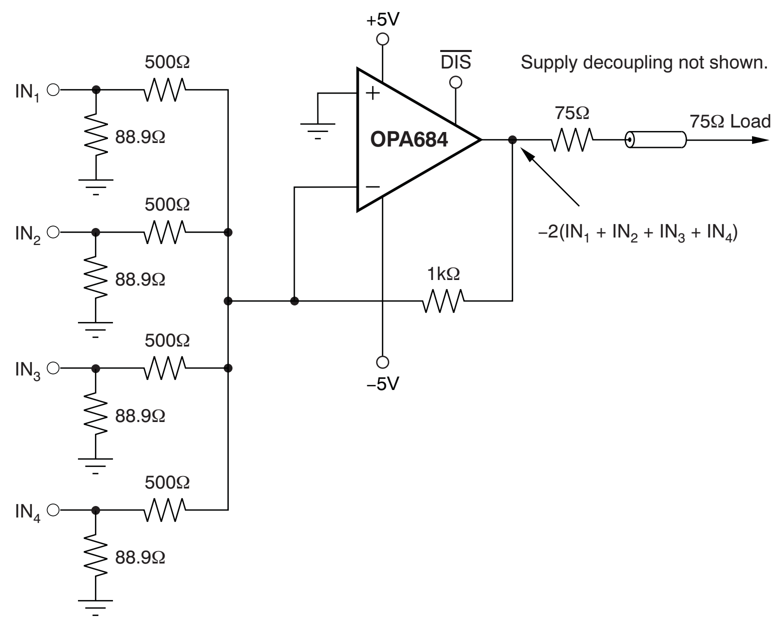

Figure 2. Video Summing Amplifier Circuit

Figure 2 shows a typical inverting summing application where four sources are summed through 500Ω gain resistors while also including an 88.9Ω terminating impedance, to present a 75Ω input impedance to each source. The gain for each channel is –2V/V to the output pin and –1V/V to the matched load. For the

OPA684, the extremely low inverting input impedance ensures non-interactive summing for all of the channels. The amplifier bandwidth is largely independent of variations in the gain setting elements, and instead depends primarily on the feedback resistor value. This type of circuit may be used to sum numerous signals together or, where the earlier stages can be disabled, to allow multiple channels to be brought together with only the active channel passing on to the output.

3 Harmonic Distortion Comparison

Harmonic distortion for both VFAs and CFAs depends on the loop gain. For a VFA, the loop gain is the difference between the noise gain of the closed-loop amplifier and the open-loop gain; for a CFA, it is the difference between the compensation element and the open-loop transimpedance gain. Figure 3 shows the loop gain performance for a VFA, while Figure 4 illustrates the loop gain performance for a CFA.

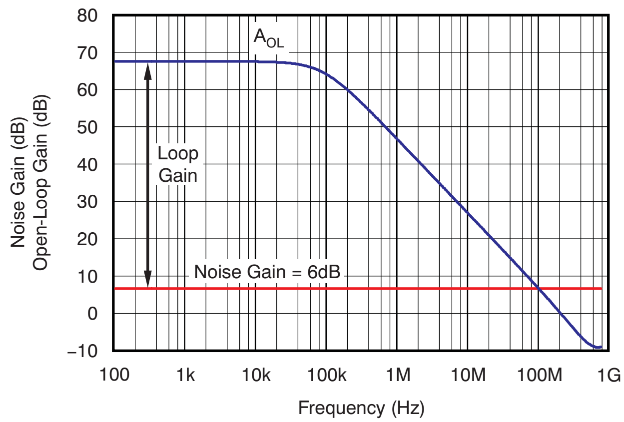

Figure 3. VFA Loop Gain (Typical Performance)

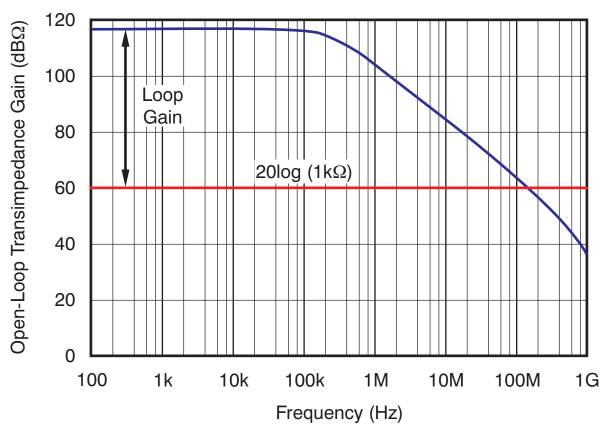

Figure 4. CFA Loop Gain (Typical Performance)

By comparing Figure 3 and Figure 4, we can see that a CFA has less loop gain than a VFA for low-gain operation if comparing dB to dBΩ directly. This characteristic indicates that a CFA may maintain better distortion at higher gains than does a VFA if only the gain resistor is changed. Conversely, at low gains, the VFA achieves much better distortion than the CFA if the gain is adjusted using only the gain resistor. Note that for both architectures, loop gain varies with frequency. The CFA maintains a higher loop gain with higher frequency than does a VFA.

To the first order, the loop-gain of a CFA depends on the feedback resistor. To the second order, it depends on the noise gain and the inverting input resistance, as can be seen in the following equation.

$$R_{COMP} = R_F + r_I \times NG$$

With:

- R_{COMP}: Total compensation of a CFA
- R_F: Feedback resistor
- r_I: Inverting input resistance
- NG: Noise Gain

As the NG increases, R_F can be reduced to optimize the bandwidth.

As an example, the relationship between distortion and the amplifier architecture for low gain performance is shown in the harmonic distortion graphs for the OPA2889 and OPA683. These plots, tested under the same conditions to 1MHz, are shown in Figure 5 and Figure 6, respectively.

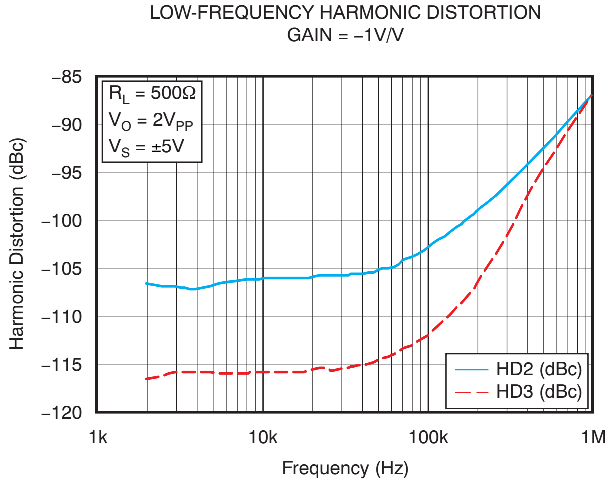

Figure 5. OPA2889 Low-Frequency Harmonic Distortion (Gain of −1V/V)

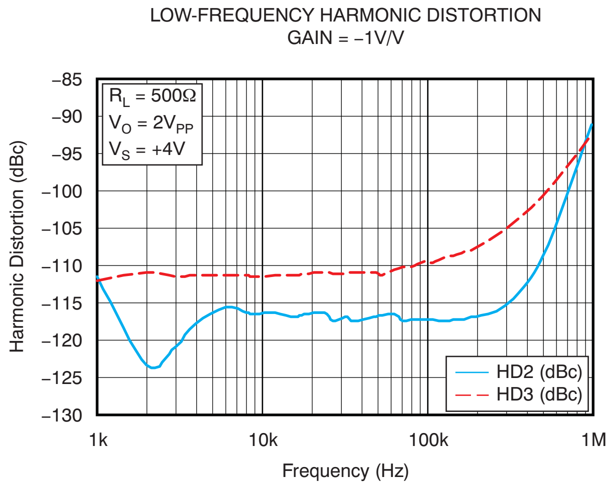

Figure 6. OPA683 Low-Frequency Harmonic Distortion (Gain of −1V/V)

Note that as the open-loop gain for a VFA and the open-loop transimpedance gain for a CFA decrease, the harmonic distortion deteriorates.

The open-loop transimpedance gain of the OPA683 is shown in Figure 7. Here, the roll-off frequency of the open-loop transimpedance gain matches the increase in distortion in Figure 6. The same relation between roll-off of the open-loop gain and corner frequency at which the distortion increases can be observed for the OPA2889, as Figure 8 illustrates.

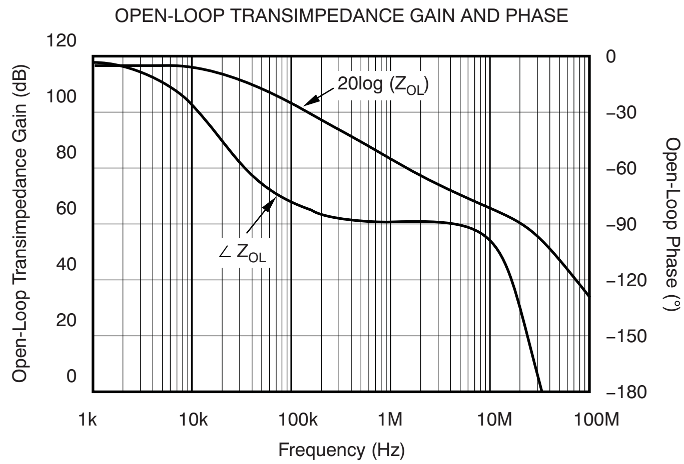

Figure 7. OPA683 Open-Loop Transimpedance Gain Performance

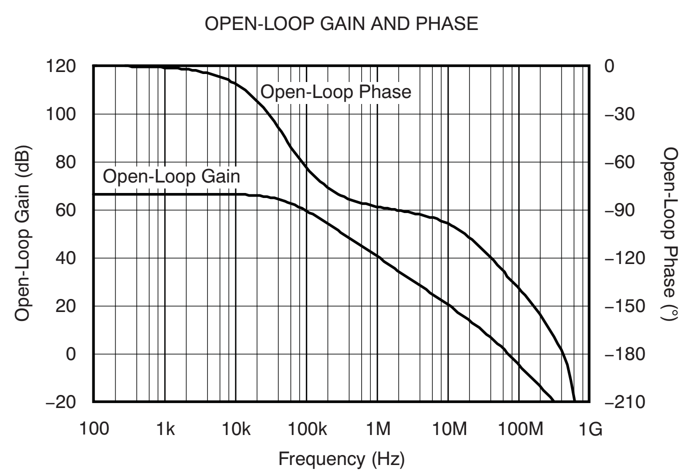

Figure 8. OPA2889 Open-Loop Gain Performance

Keeping in mind that CFAs do have an advantage for high-gain circuits, how does the OPA683 compare to a high-bandwidth, decompensated VFA? Figure 9 and Figure 10, respectively, compare the OPA847 to the OPA683.

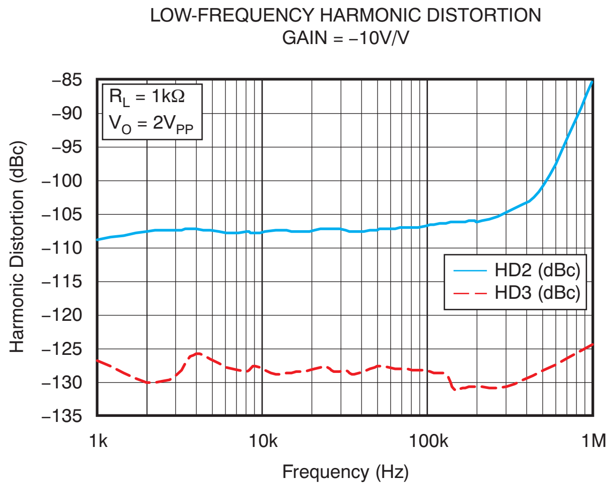

Figure 9. OPA847 Low-Frequency Harmonic Distortion (Gain of –10V/V)

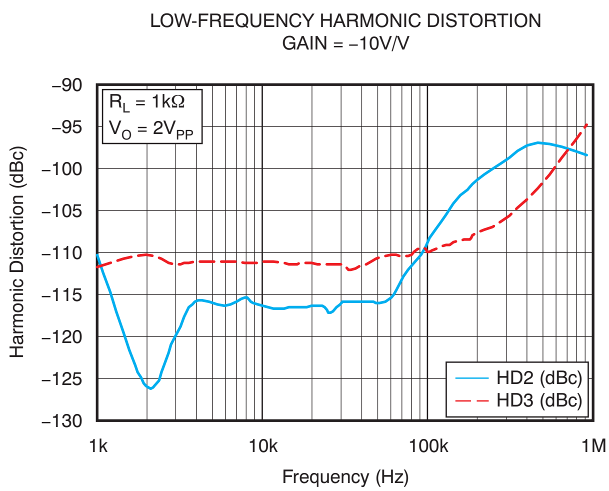

Figure 10. OPA683 Low-Frequency Harmonic Distortion (Gain of –10V/V)

Recall that at a gain of 100V/V, the OPA683 continues to have a bandwidth of 35MHz. The OPA847, with a GBP of 3900MHz, would achieve 39MHz for the same gain. Distortion, as shown in Figure 9, is very close at 100kHz to that of the OPA683; but the OPA847 requires 18.1mA to achieve this level of performance, whereas the OPA683 requires only 0.94mA.

Why then would we use the OPA847?

This amplifier, with its much higher quiescent current, also offers many specifications that the OPA683 cannot achieve—such as very low noise and (for a high-speed amplifier) relatively good dc precision. On the other hand, the OPA683 has more drive capability.

All in all, the final application dictates the amplifier requirements; but low-power devices should not necessarily be excluded in favor of higher dissipation amplifiers when they have adequate performance for the end application.

4 Low-Power Filtering

There are several common filtering architectures, among which two active-filter approaches are prominent: the multiple-feedback filter and the Sallen-Key filter.

4.1 MFB Filters

The *multiple-feedback filter* (MFB) is also called an infinite gain filter because the operational amplifier operates as an integrator. A typical fully-differential MFB filter is shown in Figure 11 with a 2MHz Butterworth filter frequency response shown in Figure 12.

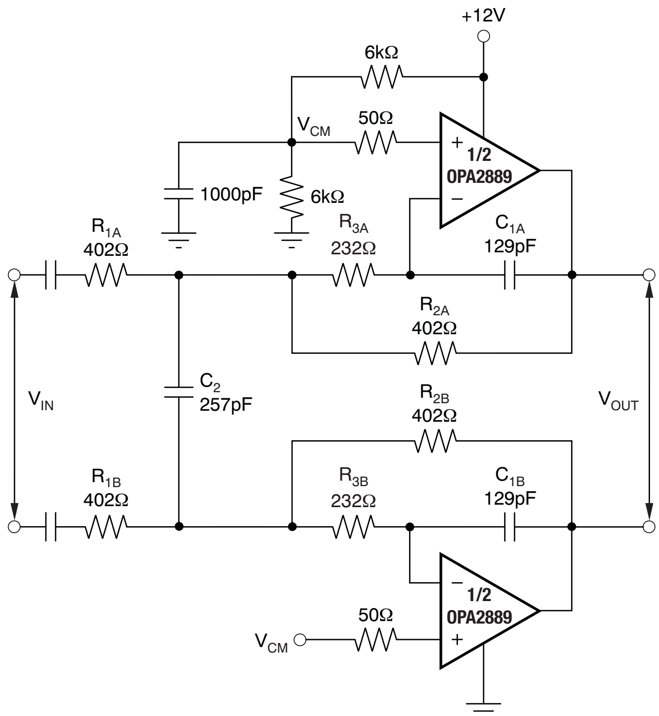

Figure 11. MFB Filter Topology Using the OPA2889

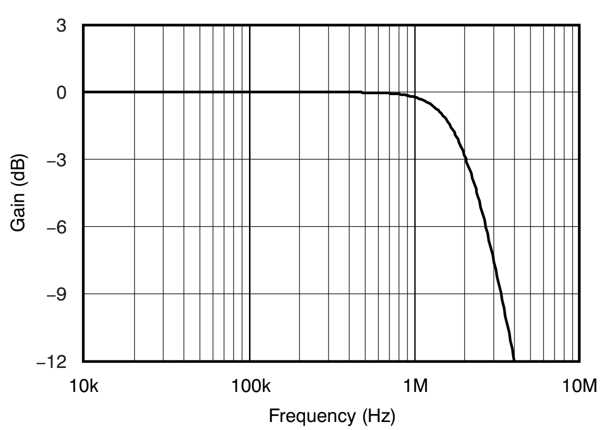

Figure 12. MFB Filter Frequency Response

In the circuit shown in Figure 11, the noninverting inputs are used to set the common-mode voltage (V_{CM}) at the output of the differential circuit. In this case, it is generated by two 6kΩ resistors that set the reference at the mid-supply, and bypassed by a 1nF capacitor. This capacitor eliminates the high-frequency noise contribution of the 6kΩ resistors. A 50Ω series resistance on the noninverting input helps isolate any LC parasitic elements and avoids potential oscillations. The common-mode gain for each of the amplifier is 1V/V. Consequently, a unity-gain stable amplifier is required. A non-unity-gain stable amplifier may be used in this application to provide better distortion, if necessary, at the expense of additional components in order to provide the adequate noise gain shaping to the common-mode gain of the amplifier.

Note that because of the capacitance across the feedback path, a VFA is required for this filter. CFA use is not recommended with this filter architecture. The filter design here is a Butterworth filter and has been implemented using the following component ratios:

- $f_o = \dfrac{1}{2\pi RC}$
- $R_1 = R_2 = 0.65 \bullet R$
- $R_3 = 0.375 \bullet R$
- $C_1 = C$
- $C_2 = 2 \bullet C$

One advantage of this type of filter is the independent setting of characteristic pulsation and the quality coefficient ω_o and Q. This topology is normally used in filters that have high Qs and require a high gain.

4.2 Sallen-Key Filters

A Sallen-Key topology is normally used to set the gain independently of the filter. In a unity-gain configuration, this filter is usually applied in filters with high-gain accuracy and low Q. A typical Sallen-Key filter circuit, as Figure 13 illustrates, shows a single-supply, low-pass filter with a 4V/V gain. The dc noise gain is set to 1V/V because of the series capacitor in the gain. With this dc isolation for the gain, the bias can be easily set by a resistor divider at the filter input. This technique is accomplished here with two 12.5kΩ resistors. The filter resistors and capacitors have been adjusted to provide a Butterworth (Q = 0.707) response with a $\omega_0 = 2\pi \bullet 5\text{MHz}$. This approach gives a flat passband response with a –3dB cutoff at 5MHz.

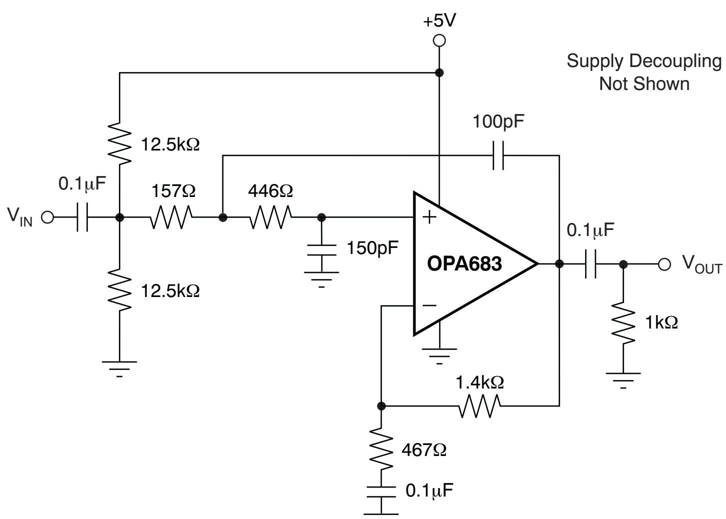

Figure 13. Single-Supply Sallen-Key Filter Implementation

The OPA683 provides an exceptionally capable gain block for implementing Sallen-Key-type filters. Typically, the bandwidth interaction with gain settings for low-power amplifiers constrain these filters to using unity-gain amplifiers. Because the OPA683 holds very high bandwidth to high gains, however, applications that provide signal gain as well as the desired filter shape are easily implemented.

Figure 14 shows an example of a 5MHz, second-order, low-pass filter where the amplifier provides a voltage gain of 4V/V. This single-supply implementation (applicable to single +12V operation as well) consumes only 5.1mW quiescent power.

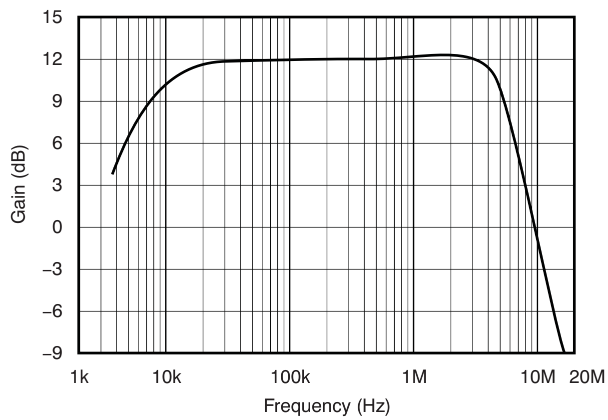

Figure 14. Butterworth Low-Pass Filter Frequency Response

5 Multiplying DAC Transimpedance Amplifier

Multiplying digital-to-analog converters (DACs), such as the DAC7822, can make good use of the low-power, high slew rate VFA. A typical circuit, shown in Figure 15, shows the OPA890 used as the output transimpedance driver. Note that a CFA can be used in transimpedance applications; however, because the feedback element is also the compensation element, a CFA would generally provide very little flexibility in a design where bandwidth is important. Additionally, in a VFA transimpedance application, a feedback capacitance controls the frequency response peaking and achieves unconditional stability. Such a technique is not recommended for CFAs because it may lead to instability under the best case conditions.

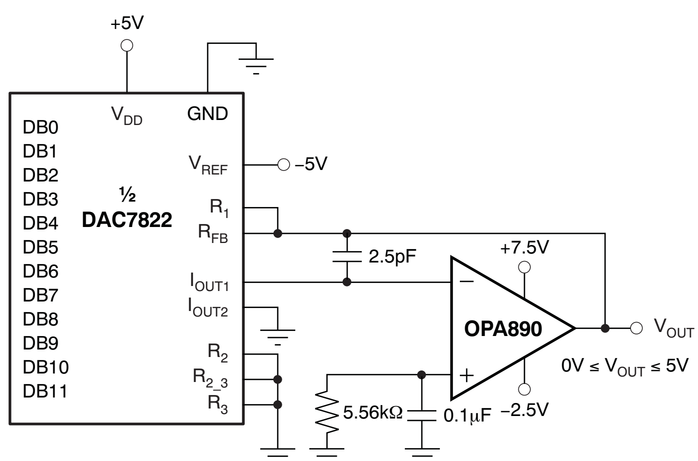

Figure 15. Multiplying DAC Transimpedance Driver

In the schematic illustrated in Figure 15, the transimpedance gain is set by a resistor internal to the DAC7822. The 2.5pF capacitance was selected to achieve a flat frequency response. On the inverting node, a 5.56kΩ resistor is used to minimize dc offset error. A 0.1μF capacitor was placed in parallel to this resistance to minimize the noise contribution for frequency above 300Hz.

Note that in order to achieve maximum amplitude, the supply voltages were set at +7.5V and –2.5V for the amplifier, allowing the output voltage to swing between 0V and 5V. Figure 16 shows the frequency response of this circuit.

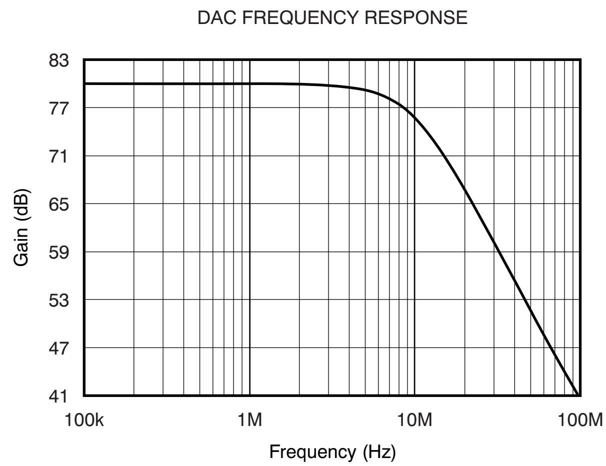

Figure 16. Multiplying DAC Frequency Response Using the OPA890 as Output Stage

Now look more closely at the interaction of the OPA890 with the DAC7822. We can see that the output impedance out of the DAC7822 is not high impedance, but varies between 4.5kΩ and 22.1kΩ (excluding code 000h) for a 10kΩ nominal V_{REF} input resistance. The I_{OUT} resistance changes are directly related to the code change. This relatively low impedance has multiple effects when an operational amplifier is used.

Some of the effects that apply to all amplifier technologies are:

- the noise gain of the amplifier changes for each code;
- the output offset voltage of the amplifier changes for each code because of the input offset voltage;
- the input offset current cannot be cancelled. The effects of the input bias current can be reduced, but not eliminated, by selecting a CMOS amplifier instead of a bipolar amplifier, thus affecting the total output offset voltage of the amplifier with each code.
- The noninverting pin of the amplifier must be tied to ground and cannot be used to create a dc offset to center the output voltage to any desired value.

The following analysis excludes the input offset current. The total output offset voltage variations, as a result of the code changing internally to the DAC, can be expressed as:

$$\Delta V_{OSO} = +\Delta NG \bullet \{ [(R_F \parallel R_{OUT1}) - R_S] + V_{OS}\}$$

where:

$$4.5k\Omega \le R_{OUT1} \le 22.1k\Omega$$

$$R_F = 10k\Omega$$

Using this value, the variation of the parallel combination of R_F and R_{OUT1} can be constrained to:

$$4.19k\Omega \le (R_F \parallel R_{OUT1}) \le 6.88k\Omega$$

In order to optimize the bias current cancellation, R_S is selected to be the average of those limiting numbers, or:

$$R_S = \frac{(6.88k\Omega + 4.19k\Omega)}{2} = 5.56k\Omega$$

Looking at the variation for each code, the total error (when including all codes) is ~3.9 LSB for the OPA890.

Notice that most of the error occurs at the first few codes (0, 1, 2). Excluding these codes from the analysis then yields the result shown in Table 2.

Table 2. DC Accuracy vs Code

Codes	Total Error Because of V_{OS} and I_B
All codes	3.9 LSB
Excluding code 0	2.5 LSB
Excluding code 0 and 1	2 LSB
Excluding code 0, 1 and 2	1.83 LSB

Note that 1LSB = 1.221mV in the circuit shown in Figure 13.

Eliminating a few codes increases the resolution of this high-speed, low-power, multiplying DAC transimpedance amplifier solution.

6 Conclusion

Relatively high bandwidth and high slew rate devices are now available with low power dissipation. This report has discussed the some of the trade-offs between various architectures as well as offered a better understanding of the specification, along with suggestions and solutions for low-power applications.