A new filter topology for analog high-pass filters

By Mark Fortunato

Analog Field Applications Manager

The analog circuit designer today has liter-
ally dozens of circuit topologies available to
implement filters, from the venerable
Sallen-Key (SK) filter, in use for well over
fifty years,[1] to more esoteric and hard-to-
pronounce filters such as the Mikhael-
Bhattacharyya (MB) filter or the Padukone-
Mulawka-Ghausi (PMG) filter.[2] Each of
these filters has advantages and disadvan-
tages relative to its cousins. Nearly all of the
filter topologies used today were developed
in the 1950s, '60s, and '70s.[2–6]. Can we
come up with a filter topology that has an advantage over
all the many topologies that have been in use for decades?
For some specific needs, the new topology presented here
has some unique advantages.

Almost all the common high-pass filters (HPFs) tend to
have one thing in common—capacitors in series with the
forward signal path. For most applications, having capaci-
tors in the signal path is not a problem. However, there
are applications where such capacitors can be problematic.
For example, in broadband low-noise circuits such as many
audio circuits, there is a need to keep resistance values,
and thus noise, low. These applications also often call for
high-pass functions that roll off at low frequencies, below
10 Hz in some audio applications. These cases can thus
call for very large capacitor values. Large-value capacitors
tend to be very expensive or have voltage coefficients and
other non-idealities that can ruin the fidelity of the signals
being passed through them.

Another limitation of the HPFs commonly used is that
the entire filter circuit is implemented separately and

Figure 1. Gain block with high-order LPF
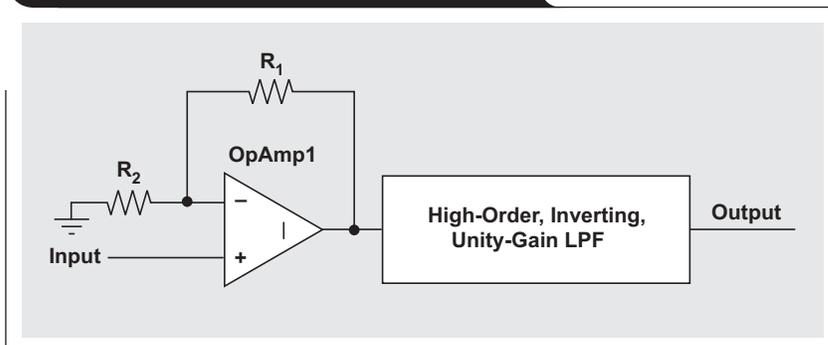
placed either before or after another circuit functional
block. Sometimes a filter in front and one after a block is
needed. The technique we will discuss here allows an
engineer to design a circuit without consideration of the
high-pass function required. After the circuit is designed,
another circuit can be "wrapped around" the original one
that will cause the overall circuit to have a high-pass func-
tion without affecting the operation of the original circuit
at frequencies above the high-pass rolloff.

Adding a high-pass function to block DC offsets

Figure 1 is a simplified schematic of a circuit with a gain
block driving a high-order low-pass filter (LPF) in a signal-
processing application. In this example there is an offset in
the input signal and an offset caused by the filter, both of
which must be removed. Typically a designer would place
a capacitor in series with the input and the output as
shown in Figure 2. For many applications this approach is
just fine; but for some applications, this simple AC-coupling
scheme can cause problems. Besides the reasons already

Figure 2. Adding simple AC-coupling stages
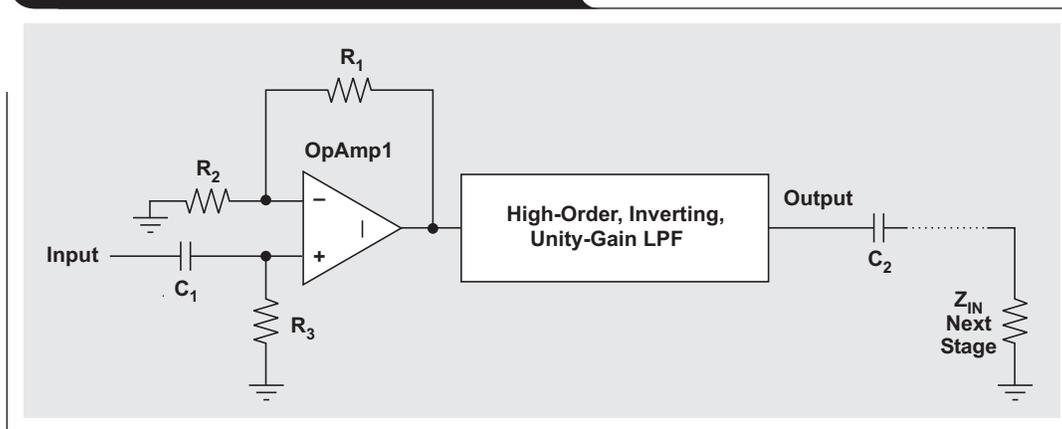
discussed, the characteristics of these high-pass stages may adversely affect the signal-processing function that is the primary purpose of the overall circuit. Each of these AC-coupling stages creates a single real (i.e., simple) pole at the frequency determined by the applicable RC time constant. Especially as the number of AC-coupling stages in the signal chain increases, the composite high-pass function is unlikely to be the optimal one for the full circuit and system. More commonly, a filter with complex pole pairs is necessary to optimize the HPF response.

Using servo feedback rather than blocking capacitors

In our example, let's assume that the input signal presents the largest of the offsets and that the filter's output offset, though small, is objectionable for the latter stages. Let's further assume that the desirable HPF function is that of a single pole. If we eliminate the input AC-coupling filter, the output filter will certainly remove all DC offsets for the subsequent stages; but then that input offset will cause the signal applied to and processed by the filter to be significantly shifted "off center," which can cause significant distortion.

An old technique referred to as "servo feedback" is often used in cases like this. This technique provides AC coupling, removing all offsets at the output of the circuit without putting any circuitry in series with the amplifier/filter chain of Figure 1. Servo feedback is fully covered in Reference 7.

The feedback path added in Figure 3 is an inverting integrator. The integrator output is fed to the inverting terminal of the input amplifier so that the overall loop has negative feedback. Assuming that the rolloff of the LPF is at least a decade above the desired high-pass rolloff, we can treat the LPF as a flat gain block for the purposes of calculating the high-pass response. A simple analysis shows that we have added a high-pass function with the transfer function

$$\frac{V_{OUT}}{V_{IN}} = -G_{LPF}\frac{R_1 + R_2}{R_2}\frac{S\dfrac{R_2R_3C_1}{R_1G_{LPF}}}{1 + S\dfrac{R_2R_3C_1}{R_1G_{LPF}}}, \qquad (1)$$

where G_{LPF} is the absolute value of the LPF gain. G_{LPF} is a high-pass function with a 3-dB (pole) frequency of

Figure 3. Single-pole HPF using servo feedback
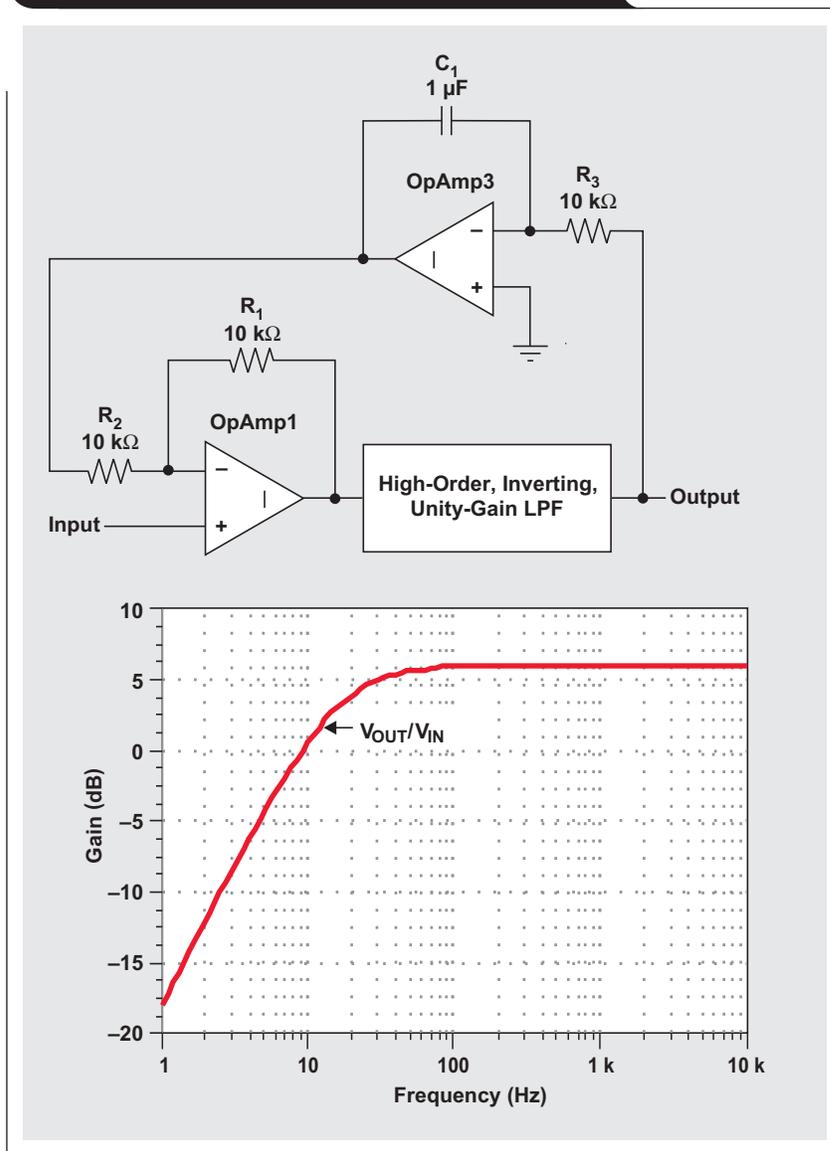
Since the LPF in this circuit has a gain of –1, we get the frequency response shown in Figure 3 with a pole frequency of 15.92 Hz.

$$f = \frac{\dfrac{R_1}{R_2}G_{LPF}}{2\pi R_3C_1}. \qquad (2)$$

This servo technique solves the problem of putting capacitors in series with the signal path, eliminates the need for multiple high-pass stages, and allows a designer to add a high-pass function to a gain/low-pass block without modifying the block itself. However, this technique is capable of implementing only simple poles and thus does

Figure 4. New second-order high-pass topology

(a) Second-order servo feedback

(b) Same circuit redrawn to be in-line
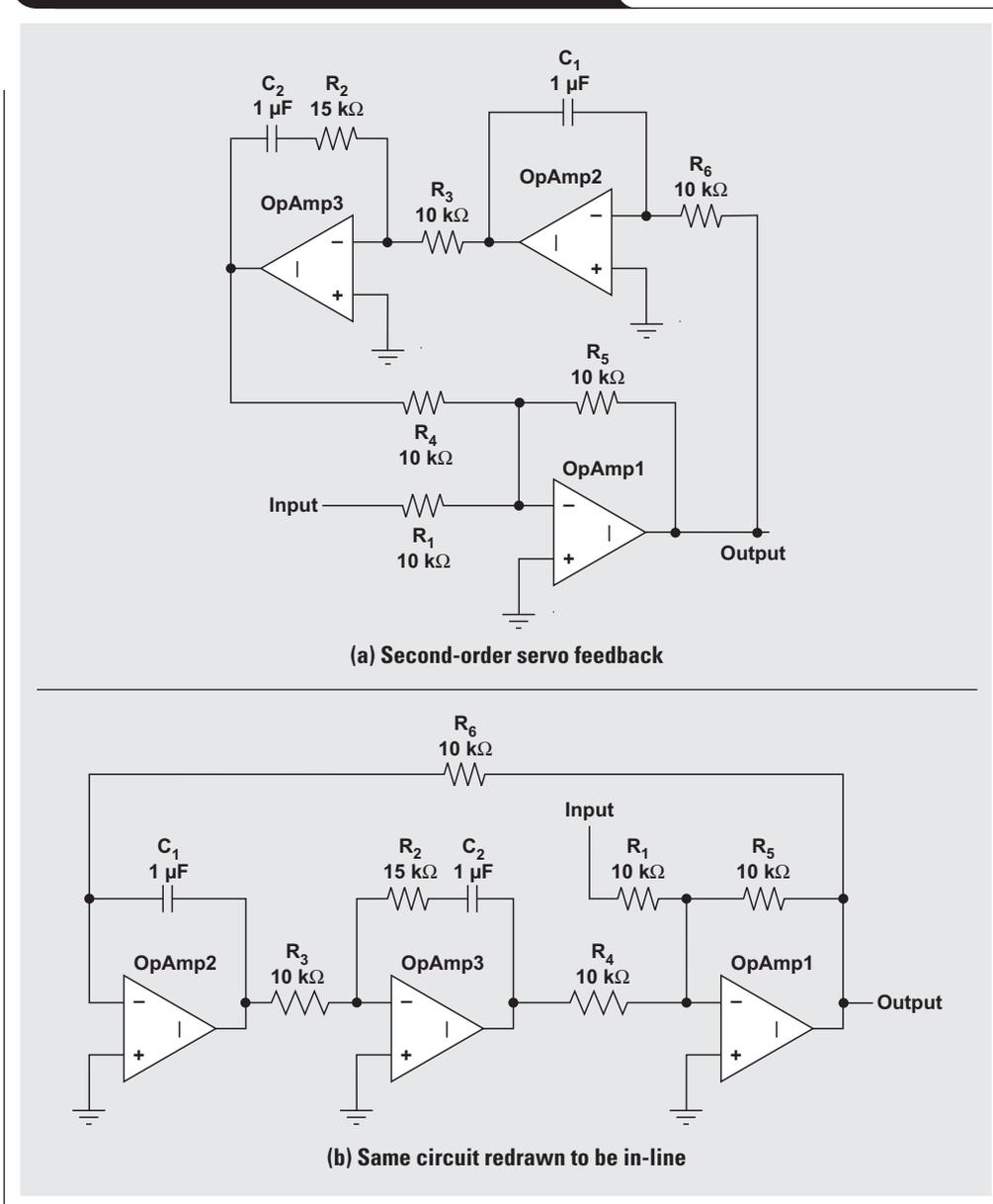
not allow us to create complex pole pairs. Therefore we need a similar technique that will provide a complex second-order function.

A new circuit topology implements a complex pole pair

Figure 4 shows just such a circuit, drawn two different ways. Figure 4a shows a gain block with the frequency-dependent part of the circuit wrapped around it like the first-order servo filter discussed earlier. Figure 4a is very similar to the previous schematics except that there are two integrators in the feedback path, one of which has an

added resistor included to set the Q of the second-order function. Figure 4b shows the identical circuit just shifted around to be in-line. Anyone familiar with three-op-amp biquad circuits such as the Kerwin-Huelsman-Newcomb (KHN) and Tow-Thomas (TT) filters will see a distinct similarity. In fact, this topology is the same as the TT filter except that, rather than having a resistor in parallel with C_1, it has R_2 in series with C_2.

The end result of this subtle change from the TT filter is that, whereas the TT filter implements an LPF and a band-pass filter (BPF) but no HPF, this circuit can implement an HPF and a BPF but no LPF. In our new circuit the HPF

output is the node labeled "Output" in both Figures 4 and 5. Figure 5 identifies the BPF output as "BPF_{OUT}."

The transfer function, pole frequency, and Q for our new HPF are respectively given by

$$\frac{V_{OUT}}{V_{IN}} = -\frac{R_5}{R_1} \frac{S^2 C_1 C_2 R_3 R_6 \frac{R_4}{R_5}}{S^2 C_1 C_2 R_3 R_6 \frac{R_4}{R_5} + SC_2 R_2 + 1}, \quad (3)$$

$$f_0 = \frac{1}{2\pi \sqrt{C_1 C_2 R_3 R_6 \frac{R_4}{R_5}}}, \quad (4)$$

$$\text{and } Q = \frac{1}{R_2} \sqrt{\frac{R_4}{R_5} R_3 R_6 \frac{C_1}{C_2}}. \quad (5)$$

Using sensitivity analysis as covered in References 2, 5, 6, and 8, we find that, as with the KHN and TT filters, all sensitivities of f_0 or Q to the passive components are 1 or lower. It is quite difficult to get lower sensitivities.

Of course, this filter could be used as a separate filter block like any other HPF topology. In the first example given earlier, we could add a three-op-amp circuit like this to the front of the gain block/filter section and another following the filter. In this case, our new HPF topology would have no real advantage over the KHN filter, and the only

Figure 5. Transfer functions for the new filter
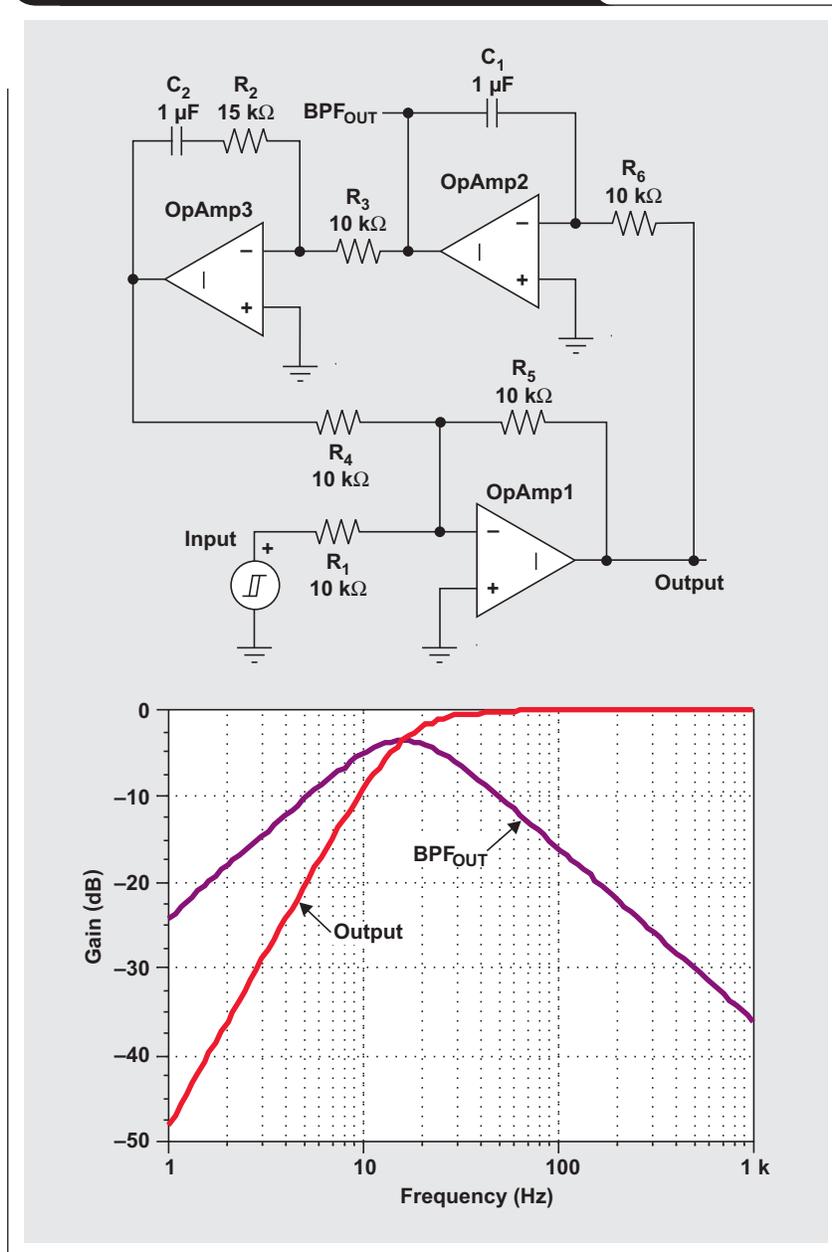
advantage it would have over any other common topology is that it can implement an HPF with no capacitors in the forward signal path.

The unique feature that this filter provides is that it is easily applied around a gain block to add a high-pass function without putting any additional circuitry in the forward signal path. Later we will see that there is a variation of this circuit that works with an inverting amplifier while feeding back to the non-inverting terminal, and that there are other variations that work with non-inverting gain blocks. All of these variations feed back to only one input.

Applying this technique to a non-inverting amplifier

If the gain block to which we want to add a high-pass function is non-inverting, we can use the variation of the topology shown in Figure 6.

The transfer function for this circuit, Equation 6, is identical to that of the circuit in Figure 5, except that the gain term is $(R_4 + R_5)/R_4$ rather than $-R_5/R_1$:

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_4 + R_5}{R_4} \frac{S^2 C_1 C_2 R_3 R_6 \dfrac{R_4}{R_5}}{S^2 C_1 C_2 R_3 R_6 \dfrac{R_4}{R_5} + S C_2 R_2 + 1} \quad \textbf{(6)}$$

Since the rest of Equation 6 is the same as Equation 3, the pole frequency and Q for this HPF variation are also given by Equations 4 and 5.

Maintaining gain-bandwidth product with an inverting amplifier

In Figure 5 we demonstrated this filter technique for an inverting amplifier. Note that both the input signal and the feedback signal are applied to the inverting terminal via R_1 and R_4, respectively. While the addition of R_4 to add the HPF feedback has no effect on the nominal forward gain for the gain block, it does increase the division ratio of the feedback path of the gain block from

$$\frac{R_1}{R_1 + R_5} \quad \text{to} \quad \frac{R_1 \| R_4}{R_1 \| R_4 + R_5}, \quad \textbf{(7)}$$

where $R_1 \| R_4$ represents that R_1 is parallel to R_4. This change in local feedback around op amp 1 has the effect of decreasing the effective gain-bandwidth product (GBWP) of the op amp by the same amount. In our case, $R_1 = R_4$, which means the GBWP will be decreased by 33%.

For an inverting amplifier, rather than feeding back to the inverting terminal as in Figure 5, we can feed back to the non-inverting terminal, thus avoiding decreasing the effective GBWP of the op amp. Figure 7 shows this variation.

Notice that R_2 is missing in this configuration. Recall that R_2 was needed to add a zero to the feedback path, which allowed the Q to be set to a reasonable value; without R2, the Q of the circuit would always be very high. In this circuit we get a zero by reconfiguring the op amp 3 stage from inverting to non-inverting. This reconfiguration was

Figure 6. Applying the technique to a non-inverting amplifier
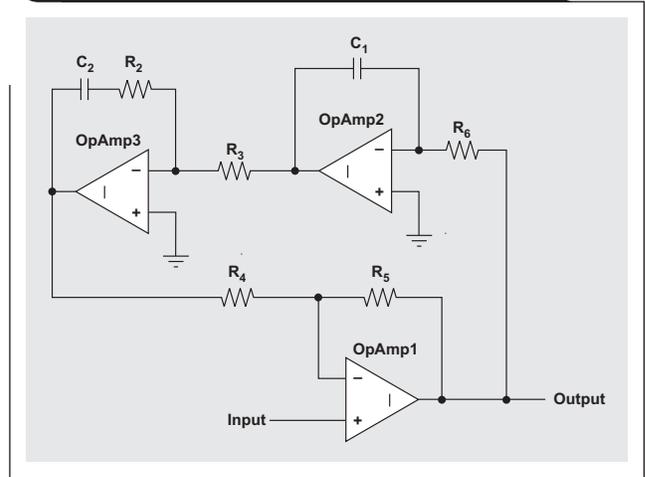
Figure 7. Inverting amplifier with feedback to non-inverting terminal
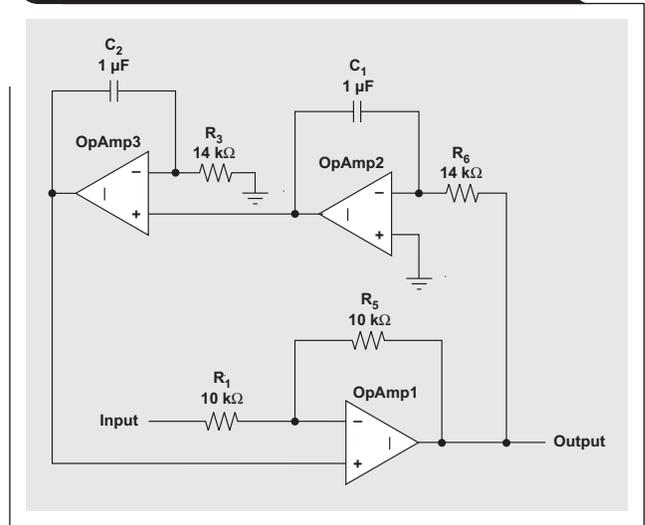
necessary to keep the feedback negative. The resulting transfer function, along with the pole frequency and Q, are respectively given by

$$\frac{V_{OUT}}{V_{IN}} = -\frac{R_5}{R_1} \frac{S^2 C_1 C_2 R_3 R_6 \dfrac{R_1}{R_5 + R_1}}{S^2 C_1 C_2 R_3 R_6 \dfrac{R_1}{R_5 + R_1} + S C_2 R_3 + 1}, \quad \textbf{(8)}$$

$$f_0 = \frac{1}{2\pi \sqrt{C_1 C_2 R_3 R_6 \dfrac{R_1}{R_5 + R_1}}}, \quad \textbf{(9)}$$

$$\text{and } Q = \sqrt{\frac{R_1}{R_1 + R_5} \frac{R_3}{R_6} \frac{C_1}{C_2}}. \quad \textbf{(10)}$$

With this configuration there is no component that can independently set the Q. Also, since R_1 and R_5 set the gain of the forward amplifier, we cannot use them to set the pole frequency or the Q. Instead, C_1, C_2, R_3, and R_6 have to be manipulated to set both the pole frequency and the Q. Fortunately, f_0 is a function of the product of these four component values, while Q is a function of the ratio of the resistors and the ratio of the capacitors. These mathematical relationships make it fairly easy to choose the right component values. We can simply set $R_3 = R_6$ and $C_1 = C_2$, then choose their values to set f_0. Then the desired Q can be set by altering the ratio of the resistors and/or the capacitors while keeping their products constant.

Implementing higher-order HPF functions

Since the new second-order filter topology and the older first-order servo technique can both be used to wrap a high-pass function around a gain block and/or an LPF function, we can use any number of these in a signal chain to create a composite HPF function of any practical order we want.

Figure 8 shows how we have created a third-order filter by combining the first-order servo feedback HPF function from Figure 5 with the second-order circuit from Figure 7.

While many DC-blocking applications can be readily handled with the insertion of capacitors in the signal path, and many others can be satisfied with older circuit topolo-

Figure 8. Combining two sections for a third-order HPF
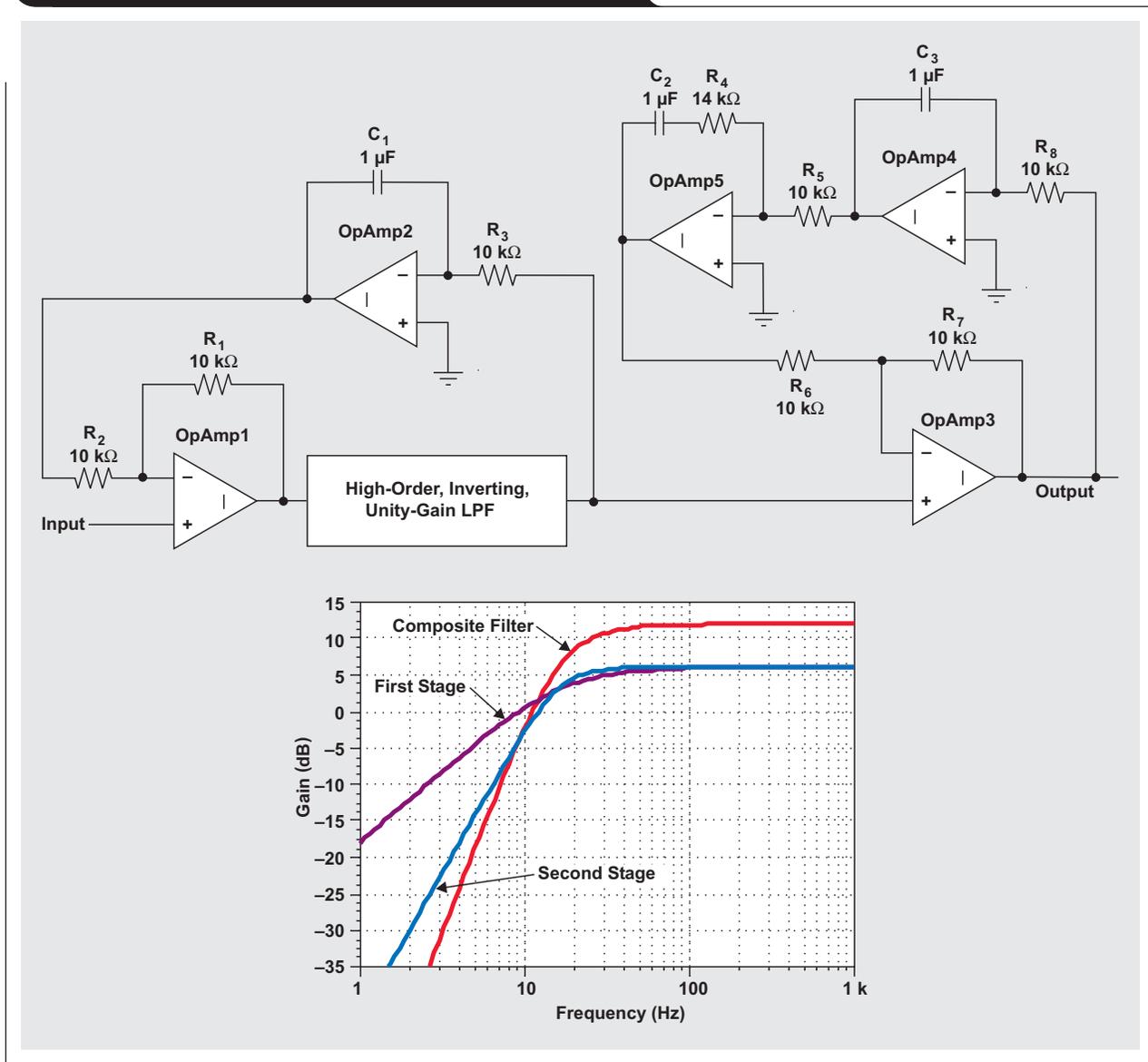
gies, the first-order servo feedback HPF and the new second-order HPF we have described can provide a great advantage in some applications due to their unique features. These features include the ease of adding to gain/LPF blocks without adding circuitry in the signal path or modifying the gain/LPF blocks, and the ability to implement HPF functions without capacitors in the signal path. The combination of the two topologies provides the ability to implement HPFs of higher orders while maintaining these advantages.

References

For more information related to this article, you can download an Acrobat Reader file at www-s.ti.com/sc/techlit/ *litnumber* and replace *"litnumber"* with the **TI Lit. #** for the materials listed below.

Document Title	TI Lit. #
1. R.P. Sallen and E.L. Key, "A practical method of designing active filters," *IRE Trans. on Circuit Theory*, Vol. CT-2, pp. 74–85, Mar. 1955.	—
2. L.P. Huelsman and P.E. Allen, *Introduction to the Theory and Design of Active Filters* (New York: McGraw-Hill, 1980).	—
3. Rolf Schaumann, Michael A. Soderstrand, and Kenneth R. Laker, Eds., *Modern Active Filter Design* (New York, IEEE Press, 1981).	—
4. Sanjit K. Mitra, Ed., *Active Inductorless Filters* (NewYork, IEEE Press, 1971).	—

Document Title	TI Lit. #
5. Aram Budak, *Passive and Active Network Analysis and Synthesis* (Boston: Houghton Mifflin Co., 1974).	—
6. M.S. Ghausi and K.R. Laker, *Modern Filter Design: Active RC and Switched Capacitor* (Englewood Cliffs, N.J.: Prentice-Hall, 1981).	—
7. R. Mark Stitt, "AC Coupling Instrumentation and Difference Amplifiers," Application Report sboa003	
8. Mark Fortunato, "Circuit Sensitivity: With Emphasis on Analog Filters," Texas Instruments Developer Conf. 2007 . . . . .sprp524	
9. J. Tow, "A step-by-step active-filter design," *IEEE Spectrum*, Vol. 6, pp. 64–68, Dec. 1969.	—
10. L.C. Thomas, "The biquad: Part I – Some practical design considerations," *IEEE Trans. on Circuit Theory*, Vol. CT-18, pp. 350–357, May 1971.	—

Related Web site

amplifier.ti.com

Products	
Amplifiers	**amplifier.ti.com**
Data Converters	**dataconverter.ti.com**
DSP	**dsp.ti.com**
Interface	**interface.ti.com**
Logic	**logic.ti.com**
Power Management	**power.ti.com**
Microcontrollers	**microcontroller.ti.com**

Applications	
Audio	**www.ti.com/audio**
Automotive	**www.ti.com/automotive**
Broadband	**www.ti.com/broadband**
Digital control	**www.ti.com/digitalcontrol**
Military	**www.ti.com/military**
Optical Networking	**www.ti.com/opticalnetwork**
Security	**www.ti.com/security**
Telephony	**www.ti.com/telephony**
Video & Imaging	**www.ti.com/video**
Wireless	**www.ti.com/wireless**

SLYT299