How to Design High Order Filters with Stopband Notches Using the LTC1562 Operational Filter (Part 2)

by Nello Sevastopoulos

This is the second in a series of articles describing applications of the LTC1562 connected as a lowpass, highpass or bandpass filter with added stopband notches to increase selectivity. Part 1 (Linear Technology VIII:2, May 1998, pp. 28–31) described one method of coupling the four Operational Filter™ building blocks of the LTC1562 to design an 8th order lowpass filter with two stopband notches. Part 2 expands the technique of Part 1 to design an 8th order bandpass filter with two stopband notches.

Throughout this series of articles, notches will be generated by first summing the input signal with a 180 degree out-of-phase signal appearing at the output(s) of the LTC1562 Operational Filter and second, by adjusting the summation gains to yield a zero sum.

Part 1 showed one proprietary method of creating notches in the stopband of a lowpass filter. The essence of this method is briefly revisited in Figure 1, where two of four Operational Filter sections are coupled to form a 4th order lowpass filter with one stopband notch. The notch is obtained by summing the input signal, \( V_{IN} \), with the output, \( V_{1A} \), into the inverting node of the next section of the IC. The two signals, \( V_{IN} \) and \( V_{1A} \), will tend to cancel each other at a frequency where they are 180 degrees out of phase. The cancellation will be complete if the amplitudes of \( V_{IN} \) and \( V_{1A} \) yield equal (and opposite) currents at the summing junction of the op amp of Figure 1, that is if:

\[
R_{IN2} = R_{FF2} \cdot \left( \frac{R_{Q1}}{R_{IN1}} \right) \quad (1)
\]

In Figure 1, the lead capacitor \( C_{IN1} \) raises the frequency where a 180 degree phase shift occurs above the center frequency of the 2nd order section \( f_0 \). The resulting notch frequency is then higher than the cutoff frequency of the 4th order filter.

Figure 1 can be easily modified to make the frequency of the notch lower than the center frequency of the 2nd order section from which it is derived. This is useful in bandpass filters where an unwanted frequency lower than the center frequency of the filter must be rejected. This is shown in Figure 2, where the input signal is summed with output \( V_{2A} \) instead of output \( V_{1A} \). The frequency of the resulting notch is:

\[
f_{N2} = f_{01} \cdot \sqrt{1 - \frac{R_{1}}{R_{Q1}} \cdot \frac{C_{1}}{C_{IN1}} \cdot \frac{R_{21}}{R_{IN1}}} \quad (2)
\]

\( R_1 = 10k; \ C = 159.15pF \)

and the gain conditions dictating Equation 1 now translate to:

\[
R_{IN2} = R_{FF2} \cdot \left( \frac{R_{Q1}}{R_{1}} \cdot \frac{C_{IN1}}{C} \right) \quad (3)
\]

The circuit of Figure 2 can be used to build a 4th order bandpass filter with one notch below its center frequency. Such a filter can simultaneously detect a tone and reject an unwanted frequency located in the vicinity of the passband.
DESIGN IDEAS

The notch techniques of Figures 1 and 2 will be referred as “feedforward.” This is necessary to separate these techniques from others to be shown later, in Part 3 of this series of articles.

The feedforward notch technique of Figure 2 can be advantageously combined with Figure 1 to realize sharp bandpass filters with two stopband notches: one notch below and one above the center frequency. Filters of this type can be very selective, although they are quite cumbersome to design. A step-by-step design procedure is illustrated below.

A Practical Example

An 8th order 100kHz bandpass filter is realized, through FilterCAD™ for Windows® (available at no charge from Linear Technology—see the “Design Tools” page in this issue), by cascading four 2nd order sections of equal Q. The –3dB band-edges are arithmetically symmetric with respect to the filter’s 100kHz center frequency and signals below 80kHz and above 125kHz are attenuated by 60dB or more. Figure 3 shows the theoretical amplitude response and Table 1 shows the desired filter parameters, namely, the center frequencies, Qs and notch frequencies. The filter of Figure 3/Table 1 can be realized by decomposing the 8th order realization into two independent 4th order filter sections and then cascading these two 4th order sections, which is an easier task than designing an 8th order elliptic bandpass filter all at once. FilterCAD, in custom mode, should be used to perform this operation. Figure 4 and Table 2 show the filter decomposition and the cascading sequence; note the left and right notches. Figure 5 uses the LTC1562 Operational Filter to realize the filter of Figure 3 as decomposed in Figure 4. The design is split into two 4th order sections. The algorithm to calculate the external passive components is outlined below.

In order to obtain a practical realization that closely approximates the theoretical one, the Q of each 2nd order section will be lowered by 15%. (Please consult the LTC1562 final data sheet.)

In order to follow the long and tedious algorithm below, consider the intuitive outline: We need to calculate the following set of passive components for the first 4th order section: $R_{IN1}$, $C_{IN1}$, $R_{21}$, $R_{Q1}$, and $R_{IN2}$, $R_{FF2}$, $R_{22}$ and $R_{Q2}$. $R_{21}$, $R_{Q1}$, $R_{22}$ and $R_{Q2}$ are easily calculated via the expression for the center frequency, $f_{Oi}$, and $Q_i$ for the 2nd order section “i.” The expression for the notch, equation (2), involves the product of $R_{IN1} \cdot C_{IN1}$, so neither component can be calculated separately. Instead, $R_{IN1}$ is calculated by considering the maximum gain (which occurs around the center frequency $f_{Q1}$) at either node V1A or V2A. This controls premature internal clipping. Once $R_{IN1}$ is set, $C_{IN1}$ is easily calculated via equation (2) for the lower band notch. Similarly, equation (3) defines the ratio of $R_{IN2}$ to $R_{FF2}$, so neither of these components can be calculated independently of the other. $R_{FF2}$ is calculated by considering the gain factor (“GAIN”) of the 4th order filter section at the V1B output (Figure 1/Table 2). Once $R_{FF2}$ is set, $R_{IN2}$ is calculated via equation (3).

Table 1. Parameters of the four sections of an 8th order, 100kHz bandpass filter

<table>
<thead>
<tr>
<th>$f_0$</th>
<th>Q</th>
<th>$f_N$</th>
<th>Q$_N$</th>
<th>Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>99.9687e3</td>
<td>10.0000</td>
<td>129.2814e3</td>
<td>77.3023e3</td>
<td>BP</td>
</tr>
<tr>
<td>96.9964e3</td>
<td>10.0000</td>
<td>129.2814e3</td>
<td>77.3023e3</td>
<td>LPN</td>
</tr>
<tr>
<td>103.0322e3</td>
<td>10.0000</td>
<td>77.3023e3</td>
<td>100.0000e3</td>
<td>HPN</td>
</tr>
</tbody>
</table>

Table 2. Filter decomposition and cascading sequence

| $f_{Qi}$ = 96.9964k | Q$_i$ = 10 |
| $f_{Q2}$ = 103.0322k | Q$_2$ = 10 |
| $f_{Q3}$ = 100k | Q$_3$ = 10 |
| $f_{Q4}$ = 129.2814k | Q$_4$ = 10 |
| $H(s) = GAIN \cdot N(s)/D(s)$ |
| $GAIN = 0.2823$ |
| $N(s) = A1s^2 + 9072 \cdot 10^9$ |
| $A1 = 62.8122 \cdot 10^3$ |
| $H(s) = GAIN \cdot N(s)/D(s)$ |
| $GAIN = 0.1788$ |
| $N(s) = A1s^2 + 659 \cdot 83 \cdot 10^9$ |
| $A1 = 62.8319 \cdot 10^3$ |
The same design method is later repeated to derive the passive components for the second 4th order section:

1. Calculate the passive components of the of the first 4th order section

\( f_{O1} = 96.9964 \text{kHz}, \ Q = 8.5, \ f_{O2} = 99.9687 \text{kHz}, \ Q = 8.5, \ f_{n2} = 77.3 \text{kHz} \)

1. Calculate the center frequency-setting resistor, \( R_{21} \):

For details, please refer to the LTC1562 data sheet.

\[ R_{21} = \left( \frac{100 \text{kHz}}{f_{O1}} \right)^2 \cdot 10k = 10.629k \]

(choose the closest 1% value, \( R_{21} = 10.7k \ (1\%) \))

2. Calculate the Q-setting resistor, \( R_{Q1} \):

For details, please refer to the LTC1562 data sheet.

\[ R_{Q1} = Q1 \sqrt{R_{21} \cdot 10k} = 87.925k \]

(choose the closest 1% value, \( R_{Q1} = 86.6k \ (1\%) \))

3. Calculate the input resistor \( R_{IN1} \) from the following expression(s):

a. if \( f_{O1} \leq 100\text{kHz} \) (for LTC1562)

\[ R_{IN1} = Q1 \cdot R_{21} \cdot \sqrt{\frac{1}{Q1^2} \left( 1 - \frac{f_{n2}^2}{f_{O1}^2} \right)} \]

\[ R_{IN1} = 95.56k \]

Although not applicable for this example, thoroughness dictates mentioning the case below:

b. if \( f_{O1} \geq 100\text{kHz} \) (for LTC1562)

\[ R_{IN1} = \frac{R_{Q1} \cdot R_{21}}{\left( 1 + \frac{f_{n2}^2}{f_{O1}^2} \right)} \]

\[ R_{IN1} = 110k \]

Make sure, in either case 3a or 3b, that \( R_{IN1} \) is greater than \( R_{21} \), that is, the DC gain at pin 3 in Figure 5 is less than unity; if not set \( R_{IN1} = R_{21} \) and proceed to step 4a.

The expression for \( R_{IN1} \) sets the gain at \( f_{O1} \) equal to unity at the node of maximum swing (V1A or V2A). Note that, for high Qs, the gain at \( f_{O1} \) is the maximum gain. If you know the spectrum of the signals that will be applied to the filter input, and if internal gains higher than unity will be allowed, the value of \( R_{IN1} \) can be reduced to improve the input signal-to-noise ratio.

4a. Use the value of \( R_{IN1} \), calculated above, and calculate the value for the input capacitor \( C_{IN1} \) from the notch equation (2).

\[ C_{IN1} = \frac{1}{\left( 1 - \frac{f_{n1}^2}{f_{O1}^2} \right)} \cdot \frac{R_{IN1}}{R_{Q1}} \cdot C \]

(\( f_{n1} < f_{O1} \); \( C = 159.15\text{pF} \))

\[ C_{IN1} = 5.639\text{pF} \]

Use the commercially available NPO type 0402 surface mount capacitor with the value nearest the ideal value of \( C_{IN1} \) calculated above. For instance, for \( C_{IN1} \), choose an off-the-shelf 5.6pF capacitor.

4b. Recalculate the value of \( R_{IN1} \) after \( C_{IN1} \) is chosen.

\[ R_{IN1} = \left( \frac{C_{IN1}(\text{ideal})}{C_{IN1}(\text{NPO,0402})} \right) \cdot C_{IN1}(\text{ideal}) \]

\[ R_{IN1} = 95.3k \ (1\%) \]

5. Calculate the frequency- and Q-setting resistors \( R_{22}, R_{Q2} \), as done in steps 1 and 2, above. Choose the closest 1% standard resistor values.

\( R_{22} = 10k \ (1\%); \ R_{Q1} = 84.5k(1\%) \)

6. Calculate the feedforward resistor, \( R_{FF2} \):

\[ \frac{1}{R_{FF2} \cdot C} = \frac{\text{Gain} \cdot A1}{C} \]

The values for parameter \( \text{Gain} \cdot A1 \) are provided by FilterCAD; they relate to the coefficients of the numerator of the transfer function \( V_{1B}/V_{IN} \) in Figure 1; a passband AC gain of unity is assumed (see Table 2). Please note that, for a lowpass case, as in Part 1 of this article series, the value of \( \text{Gain} \cdot A1 \) is the DC gain of the filter and its value can be easily set without software assistance.

Equating the numerator of the filter transfer function with the values provided by FilterCAD:

\[ \frac{V_{1B}}{V_{IN}} = \frac{s(s^2 + \omega_n^2)}{(R_{FF2} \cdot C) \cdot D(s)} = \frac{\text{Gain} \cdot A1(s^2 + A2)}{D(s)} \]

\[ \text{Gain} = 0.2823 \]

\[ A1 = 62.8122 \cdot 10^3 \]

\[ A2 = (2\pi f_o)^2 = 235.9 \cdot 10^9 \]

\[ R_{FF2} = 1/(\text{Gain} \cdot A1 \cdot C) = 354.35k; \]

\[ C = 159.15\text{pF} \]

\[ R_{FF2} = 357k(1\%) \]
II. Calculate the passive components of the second 4th order section

1. \[ f_{o3} = 100kHz, \quad Q3 = 8.5, \quad f_{o4} = 103.0322kHz, \quad Q4 = 8.5, \quad f_{n4} = 129.2814kHz \]

Except for the bandpass gain calculations, the algorithm will be the same as the lowpass design of Part 1 of this article.

1. \[ R_{23} = (100kHz/f_{o3})^2 \cdot 10k = 10k \text{ (1\%)} \]
2. \[ R_{93} = Q3 \sqrt{R_{23} \cdot 10k} = 85k, \quad R_{93} = 84.5k \text{ (1\%)} \]
3. Calculate the input resistor \( R_{IN3} \) from the following expression(s):

3a. If \( f_{o3} \leq 100kHz \) (for LTC1562)

\[
R_{IN3} = Q3 \cdot R_{23} \cdot \sqrt{\frac{1}{\omega_{N3}^2} + \left(1 - \frac{f_{o3}^2}{f_{n4}^2}\right)^2 \cdot Q3^2} \quad (8)
\]

\[
R_{IN3} = 302.41k
\]

3b. If \( f_{o3} \geq 100kHz \) (for LTC1562)

\[
R_{IN3} = R_{93} \cdot \sqrt{\frac{1}{\omega_{N3}^2} + \left(1 - \frac{f_{o3}^2}{f_{n4}^2}\right)^2 \cdot Q3^2} \quad (9)
\]

For \( f_{o3} = 100kHz \), as in the example above, either expression can be used. Note that the expression for \( R_{IN3} \) in 3b, above, is the same as expression for \( R_{IN1} \) shown in Part 1 of this article.

4a. Use the theoretical value for \( R_{IN3} \), calculated above, and calculate the value of the input capacitor \( C_{IN3} \) from the notch equation (2) of part 1 of this article; for convenience this is repeated below:

\[
Q_{IN3} = C \cdot \frac{R_{Q3}}{R_{IN3}} \cdot \left(1 - \frac{f_{o3}^2}{f_{n4}^2}\right) \quad (10)
\]

\[
C_{IN3} = 17.86\mu F
\]

Use a commercially available NPO-type 0402 surface mount capacitor with the value nearest the ideal value of \( C_{IN3} \) calculated above. For instance, \( C_{IN3} = 18\mu F \).

4b. Recalculate the value for \( R_{IN3} \) calculated in step 3a after \( C_{IN3} \) is chosen.

\[
R_{IN3} = 294k \text{ (1\%)}
\]

5. Calculate the frequency- and Q-setting resistors, R24 and RQ4, as done in steps 1 and 2, above. Choose the nearest 1% standard value.

\[
R_{24} = 9.42k; \quad R_{24} = 9.53k \text{ (1\%)}
\]

6. Calculate the feedforward resistor, RFF4. First equate the numerator of the 4th order filter transfer function with the values provided by FilterCAD (see Table 2):

\[
\frac{V_{OUT}}{V_{1B}} = \frac{s + \omega_{Q3}^2}{\omega_{Q3}^2 \cdot s^2 + \omega_{N4}^2} \quad (11)
\]

\[
\text{GAIN} \cdot A1s \cdot (s^2 + \omega_{N4}^2) \quad \text{D(s)}
\]

\[
\text{THEN } R_{FF4} = \frac{1}{\text{GAIN} \cdot A1} \cdot \frac{1}{\omega_{Q3}^2 \cdot \omega_{N4}^2} \quad \text{D(s)}
\]

\[
\text{GAIN} = 0.1788
\]

\[
A1 = 62.8319 \cdot 10^3
\]

\[
R_{FF4} = 334.64k, \quad \text{choose } R_{FF4} = 332k \text{ (1\%)}
\]

7. Solve for \( R_{IN4} \) by using equation (1) of Part 1 of this article, which dictates the gain condition for the occurrence of a notch. For convenience, this gain condition is repeated below.

\[
R_{IN4} = R_{FF4} \cdot \frac{R_{Q3}}{R_{ING}} \quad (12)
\]

\[
R_{IN4} = 95.422k; \quad R_{IN4} = 95.3k \text{(1\%)}
\]

Experimental Results

Figure 6 shows the measured amplitude response of the filter of Figure 5. The values of the passive component are as calculated above and as shown in Figure 5. The measured amplitude response closely approximates the ideal response as synthesized by FilterCAD. The peak frequency with standard 1% resistor values and 5% capacitor values is 100.65kHz (0.65% off). The higher frequency notch, although it shows a respectable depth of 70dB, is not as well defined as the notch below the filter's center frequency, yet the –65dB bandwidth is as predicted by FilterCAD. The 10dB lack of the upper band notch depth is due to the finite speed of the internal op amps; they cause the practical 180 degree phase shift frequency and the gain at V1A’s output to depart slightly from the theoretical calculations.

For the sake of perfection, the notch depth can be easily restored by tweaking the value of \( R_{93} \); the new \( R_{93} \) will be 75k. This is shown with dashed lines in Figure 6. This, however, lowers the passband gain by the ratio of the new to the old \( R_{93} \) value, that is, by about –1.0dB (you cannot fool mother nature). Depending on the application, the 10dB of additional notch depth for 1.5dB of passband gain loss may be a reasonable trade. The passband gain can also be corrected by lowering the values of either pair, \( (R_{FF2}, R_{IN2}) \) or \( (R_{FF4}, R_{IN4}) \), by the same amount (1.5dB). In Figure 6, the gain was restored to 0dB by changing the values of \( R_{IN2}, R_{FF2} \) to 93.1k and 300lk respectively.

The total integrated noise was an impressively low 69\mu VRMS, allowing a signal-to-noise ratio well in excess of 80dB. The input signal-to-noise ratio can be further increased if the pass-
band gain can be higher than 0dB or if internal nodes are allowed to have gains higher than 0dB. Please contact the LTC Filter Design and Applications Group for further details. The low noise behavior of the filter makes it useful in applications where the input signal has a wide voltage range. This is true provided the filter magnitude response does not change with varying input signal levels, that is, the filter gain is linear. The gain linearity measured at the 100kHz theoretical center frequency of the filter is shown in Figure 7. The gain is perfectly linear for input amplitudes up to 1.25V_RMS (3.5V_p-p) so an 84dB dynamic range can be claimed. The input signal, however, can reach amplitudes up to 3V_RMS (8.4V_p-p, 92dB SNR) with some reduction in gain linearity.

References
4. LTC1562 Final Data Sheet.

Acknowledgments
Philip Karantzalis and Nello Sevastopoulos of LTC’s Monolithic Filter Design and Applications Group contributed to the application examples.

SW, V_BAT and GND in Figure 2 will help in spreading the heat and will reduce the power dissipation in conductors and MOSFETs.

Other Applications
The LT1505 can also be used in other system topologies, such as the telecom application shown in Figure 5. The circuit in Figure 5 uses the battery to supply peak power demands. By doing so, the required peak power from the wall adapter can be much lower than the peak power required by the load. The wall adapter has to supply the average power only.

Conclusion
The LT1505 is a complete, single-chip battery charger solution for today’s demanding charging requirements in high performance laptop applications. The device requires a small number of external components and provides all necessary functions for battery charging and power management. High efficiency and small size allow for easy integration with the laptop circuits. Also, by adding a simple external circuit, charging can be easily controlled by the host computer, allowing for more sophisticated charging schemes.

Step-Down Conversion, continued from page 30
lower cost LTC1430A replacing the LTC1649. The LTC1430A does not include the 3.3V to 5V charge pump and requires a 5V supply to drive the external MOSFET gates. The current drawn from the 5V supply depends on the gate charge of the external MOSFETs but is typically below 50mA, regardless of the load current on the 2.5V output. The drains of the Q1/Q2 pair draw the main load current from the 3.3V supply. The remaining circuit works in the same manner as in Figure 1. Efficiency and performance are virtually the same as the LTC1649 solution, but parts count and system cost are lower.

In a 3.3V to 2.5V application, the steady-state, no-load duty cycle is 76%. If the input supply drops to 3.135V (3.3V – 5%), the duty cycle requirement rises to 80% at no load, and even higher under heavy or transient load conditions. Both the LTC1649 and the LTC1430A guarantee a maximum duty cycle of greater than 90% to provide acceptable load regulation and transient response. The standard LTC1430 (not the LTC1430A) can max out as low as 83%—not high enough for 3.3V to 2.5V circuits. Applications with larger step-down ratios, such as 3.3V to 2.0V, can use the circuit in Figure 3 successfully with a standard LTC1430.