Application Report

SNOA374B–October 1996–Revised April 2013

OA-29 Low-Sensitivity, Highpass Filter Design with Parasitic Compensation

ABSTRACT

This application report covers the design of a Sallen-Key highpass biquad. This design gives low component and op amp sensitivities. It also shows how to compensate for the op amp's bandwidth (pre-distortion) and parasitic capacitances. A design example illustrates this method. These biquads are also called KRC or VCVS [voltage-controlled, voltage-source].

Contents

List of Figures

All trademarks are the property of their respective owners.

Copyright © 1996–2013, Texas Instruments Incorporated

1 Introduction

Changes in component values over process, environment and time affect the performance of a filter. To achieve a greater production yield, the filter needs to be insensitive to these changes. This application report presents a design algorithm that results in low sensitivity to component variation. See reference [6] for information on evaluating the sensitivity performance of your filter.

To achieve the best production yield, the nominal filter design must also compensate for component and board parasitics. The components are pre-distorted (reference [5]) to compensate for the op amp bandwidth. This application report expands the pre-distortion method in reference [5] to include compensation for parasitic capacitances. This method is valid for either voltage-feedback or current-feedback op amps.

2 Parasitic Compensation

To pre-distort your filter components and compensate for parasitic capacitances:

1. Use the method in reference [5] to include the op amp's effect on the filter response. The result is a transfer function of the same order whose coefficients include the op amp group delay (τ_{oa}) evaluated at the passband edge frequency (f_c).
2. For all parasitic capacitances in parallel with capacitors:
   - Add the capacitors together
   - Simplify the resulting coefficients
   - Use the sum of time constants form for the coefficients when possible
3. For all parasitic capacitances in parallel with resistors:
   - Replace the resistor R_x in the filter transfer function with the parallel equivalent of R_x and C_p.

     $$\frac{R_x \leftarrow R_x}{\left(1+R_xC_ps\right)}, s = j\omega$$

   - Alter this impedance to a convenient form and simplify:
   - *Do not create new terms* (a coefficient times a new power of s) in the transfer function *after simplifying*
   - The most useful approximations are:

     $$\frac{R_x}{\left(1+R_xC_ps\right)} \approx R_x\left(1-R_xC_ps\right)$$

     $$\approx R_xe^{-R_xC_ps}$$

     These approximations are valid when:

     $$\frac{\omega << 1}{\left(R_xC_p\right)}$$

   - Convert (1 + R_xC_ps) to the exponential form (a pure time delay) when it multiplies, or divides, the entire transfer function
   - Do not change the gain at $\omega \approx \omega_p$ in allpass sections
   - When simplifying, discard any terms that are products of the error terms ($k\tau_{oa}$ and R_xC_p); they are negligible
   - Use the sum of time constants form for the coefficients when possible

   Use an op amp with adequate bandwidth (f_{3dB}) and slew rate (SR):

$$f_{3dB} \geq 10f_H \tag{1}$$

$$SR > 5f_HV_{peak} \tag{2}$$

where f_H is the highest frequency in the passband of the filter, and V_{peak} is the largest peak voltage. This increases the accuracy of the pre-distortion algorithm. It also reduces the filter's sensitivity to op amp performance changes over temperature and process. Make sure the op amp is stable at a gain of A_V = K.

Copyright © 1996–2013, Texas Instruments Incorporated

3 KRC Highpass Biquad Design

The biquad shown in Figure 1 is a Sallen-Key highpass biquad. V_{in}needs to be a voltage source with low output impedance.

The transfer function is:

$$\frac{V_o}{V_{in}} \approx \frac{H_\infty \left(\frac{1}{\omega_p^2}\right) s^2}{1 + \left(\frac{1}{(\omega_p Q_p)}\right) s + \left(\frac{1}{(\omega_p^2)}\right) s^2}$$

$$K = 1 + \frac{R_f}{R_g}$$

$$H_\infty = K$$

$$\frac{1}{(\omega_p Q_p)} = R_5 C_1 + R_5 C_3 - R_4 C_3 (K - 1)$$

$$\frac{1}{\omega_p^2} = R_4 R_5 C_1 C_3$$

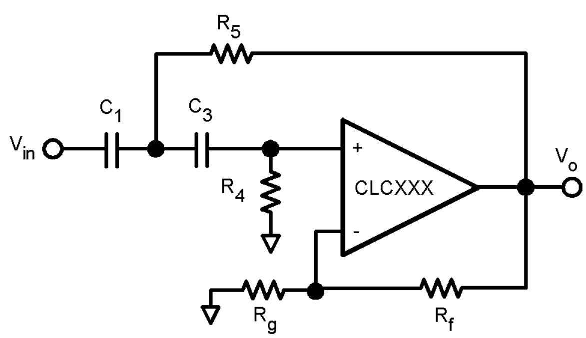

Figure 1. Highpass Biquad

To achieve low sensitivities, use this design algorithm:

1. Partition the gain for good Q_p sensitivity and dynamic range performance:
   - Use a low noise amplifier before this biquad if you need a large gain
   - Select K with this empirical formula:

     $$K = \begin{cases} 1, & 0.1 \le Q_p \le 1.1 \\ \frac{2.2\,Q_p - 0.9}{Q_p + 0.2}, & 1.1 < Q_p < 5 \end{cases}$$

     These values also reduce the op amp bandwidth's impact on the filter response. This biquad's sensitivities are too high when $Q_p \ge 5$
2. Select an op amp with adequate bandwidth (f_{3dB}) and slew rate (SR):
   $f_{3dB} \ge f_H f_{3dB} \ge 10 f_c SR > 5 f_H V_{peak}$
   where f_H is the highest signal frequency, f_c is the corner frequency of the filter, and V_{peak} is the largest peak voltage. Make sure the op amp is stable at a gain of A_v = K.
3. For current-feedback op amps, use the recommended value of R_f for a gain of A_V = K. For voltage-feedback op amps, select R_f for noise and distortion performance. Then set R_g for the correct gain:

   $$R_g = \frac{R_f}{(K - 1)}$$

Copyright © 1996–2013, Texas Instruments Incorporated

4. Initialize the resistance level $\left(R = \sqrt{R_4 R_5}\right)$.Increasing R will:
   - Increase the output noise
   - Reduce the distortion
   - Improve the isolation between the op amp outputs and C_1 and C_3
   - Make the parasitic capacitances a larger fraction of C_1 and C_3

5. Initialize the capacitance level $S_{\alpha_i}^{H_\infty}$, and the component ratios

$$\left(c^2 = \frac{C_3}{C_1} \text{ and } r^3 = \frac{R_5}{R_4}\right):$$

$$c = \frac{1}{\left(\omega_p R\right)}$$

$$c^2 = 0.10$$

$$r^2 = \max\left\{0.10, \left(\frac{1+\sqrt{1+4Q_p^2}\left(1+c^2\right)(K-1)}{2\cdot Q_p\cdot\left(1+c^2\right)/c}\right)\right.$$

6. Recalculate C^2 and initialize the capacitors:

$$c^2 = \left(\frac{2\cdot r\cdot Q_p}{1+\sqrt{1+4Q_p^2}\left(K-1-r^2\right)}\right)^2$$

$$C_1 = \frac{C}{c}$$

$$C_3 = cC$$

7. Set C_1 and C_3 to the nearest standard values.
8. Recalculate C, C^2, R and r^2:

$$C = \sqrt{C_1 C_3}$$

$$c^2 = \frac{C_3}{C_1}$$

$$R = \frac{1}{\left(\omega_p C\right)}$$

$$r^2 = \left(\frac{1+\sqrt{1+4Q_p^2\left(1+c^2\right)(K-1)}}{2\cdot Q_p\cdot\left(1+c^2\right)/c}\right)^2$$

9. Calculate the resistors:

$$R_4 = \frac{R}{r}$$

$$R_5 = rR$$

The component sensitivity formulas are in the table below. The sensitivities formulas are in the table below. The sensitivities to α_i = K are a measure of this biquad's sensitivity to the op amp group delay (reference [5]). To evaluate this biquad is sensitivity performance, use the method in reference [6].

Copyright © 1996–2013, Texas Instruments Incorporated

α_i	$S^{H_\infty}_{\alpha_i}$	$S^{\omega_p}_{\alpha_i}$	$S^{Q_p}_{\alpha_1}$
C_1	0	$-\frac{1}{2}$	$-\left(Q_p \cdot \frac{r}{c} - \frac{1}{2}\right)$
C_3	0	$-\frac{1}{2}$	$\left(Q_p \cdot \frac{r}{c} - \frac{1}{2}\right)$
R_4	0	$-\frac{1}{2}$	$\left((K-1) \cdot Q_p \cdot \frac{c}{r} + \frac{1}{2}\right)$
R_5	0	$-\frac{1}{2}$	$-\left((K-1) \cdot Q_p \cdot \frac{c}{r} + \frac{1}{2}\right)$
R_f	$\frac{K-1}{K}$	0	$\left((K-1) \cdot Q_p \cdot \frac{c}{r}\right)$
R_f	$-\frac{K-1}{K}$	0	$-\left((K-1) \cdot Q_p \cdot \frac{c}{r}\right)$
K	1	0	$\left(K \cdot Q_p \cdot \frac{c}{r}\right)$

4 KRC Highpass Biquad Parasitic Compensation

To pre-distort this biquad, and compensate for the [parasitic] non-inverting input capacitance of the op amp (C_{ni}), do the following (see Appendix A for the derivation of the formulas):

1. Start the iterations by ignoring the parasitics:$\tau^2 = 0\tau^2_4 = 0$
2. Estimate the pre-distorted values of ω_p and Q_p ($\omega_{p(pd)}$ and $Q_{p(pd)}$ that will compensate for τ_{oa} and C_{ni}:

$$\frac{\omega_{p(pd)} = \omega_{p(nom)}}{\sqrt{1-\tau_4^2\omega_{p(nom)}^2}}$$

$$\frac{Q_{p(pd)} = Q_{p(nom)}}{\left(\frac{\omega_{p(pd)}}{\omega_{p(nom)}} - Q_{p(nom)}\tau_2\omega_{p(pd)}\right)}$$

Where $\omega_{p(nom)}$ and $Q_{p(nom)}$ are the nominal values of ω_p and Q_p

3. Recalculate the resistors and capacitors using $\omega_{p(pd)}$ and $Q_{p(pd)}$:

$$\frac{1}{\omega_{p(pd)}^2} = R_4R_5C_1C_3$$

$$\frac{1}{\left(\omega_{p(pd)}Q_{p(pd)}\right)} = R_5C_1 + R_5C_3 - R_4C_3(K-1)$$

Section 5 accomplishes this by recalculating R and r^2, then R_4 and R_5:

$$R = \frac{1}{\left(\omega_{p(pd)}C\right)}$$

$$r^2 = \left(\frac{1+\sqrt{1+4Q_{p(pd)}^2\left(1+c^2\right)(K-1)}}{2 \cdot Q_{p(pd)} \cdot \left(1+c^2\right)/c}\right)^2$$

$$R_4 = \frac{R}{r}$$

$$R_5 = rR$$

Copyright © 1996–2013, Texas Instruments Incorporated

4. Calculate the resulting parasitic correction factors:$\tau^2 = R_4C_{ni}\tau^2{}_4 = K\tau_{oa}R_4C_3 + R_4R_5(C_1 + C_3)C_{ni}$
5. Calculate the resulting filter response parameters ω_p and Q_p

$$\frac{\omega_p = \omega_{p(pd)}}{\sqrt{1+\tau_4^2\omega_{p(pd)}^2}}$$

$$Q_p = \frac{Q_{p(pd)}}{\left(\frac{\omega_p}{\omega_{p(pd)}} + Q_{p(pd)}\tau_2\omega_p\right)}$$

6. Repeat steps 2-5 until:

$\omega_p = \omega_{p(nom)}$

$Q_p = Q_{p(nom)}$

7. Estimate the high frequency gain:

$$\frac{H_\infty \approx K}{\left(1+\tau_4^2\omega_{p(pd)}^2\right)}$$

If this reduces the gain too much, then repartition the gain.

5 Design Example

The circuit shown in Figure 2 is a 3rd-order Butterworth highpass filter. Section A is a buffered single pole section, and Section B is a highpass biquad. Use a voltage source with low output impedance, such as the CLC111 buffer, for V_{in}:

The nominal filter specifications are:

f_c = 50MHz—(passband edge frequency)

f_s = 10MHz—(stopband edge frequency)

f_H = 200MHz—(highest signal frequency)

A_p = 3.0dB—(maximum passband ripple)

A_s = 40dB—(minimum stopband attenuation)

H_∞ = 0dB—(passband voltage gain)

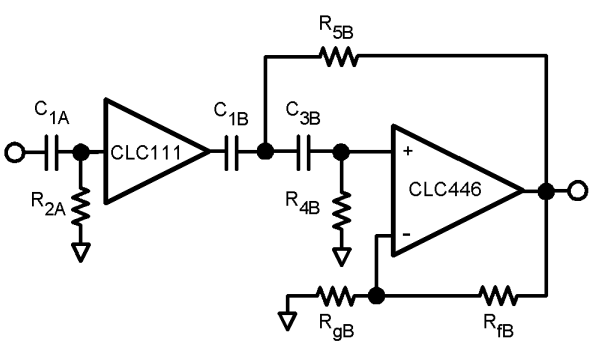

Figure 2. Highpass Filter

Copyright © 1996–2013, Texas Instruments Incorporated

The 3rd-order Butterworth filter [1-4] meets our specifications. The pole frequencies and quality factors are:

Section	A	B
$\omega_p/2\pi$ [MHz]	50.00	50.00
Q_p [ ]	–	1.000

Overall Design

1. Restrict the resistor and capacitor ratios to:
2. $0.1 \le c^2, r^2 \le 10$
3. Use 1% resistors (chip metal film, 1206 SMD)
4. Use 5% capacitors (ceramic chip, 1206 SMD)
5. Use standard resistor and capacitor values

Section A Design and Pre-distortion:

1. Use the CLC111. This is a close-loop buffer.
   - f_{3dB} = 800MHz > f_H = 200MHz
   - f_{3dB}= 800MHz > $10f_c$= 500MHz
   - SR = 3500V/µs, while a 200MHz, $2V_{pp}$ sinusoid requires more than 100V/µs
   - $\tau_{oa} \approx$ 0.28ns at 10MHz
   - $C_{ni(111)}$ = 1.3pF (input capacitance)
2. Select R_{2A} for noise, distortion and to properly isolate the CLC111's output and C_{1A}. The pre-distorted value of R_{2A}, that also compensates for $C_{ni(111)}$, is [5]:

$$R_{2A} = \frac{\left(\dfrac{1}{\omega_p - \tau_{oa}}\right)}{\left(C_{1A} + C_{ni(111)}\right)}$$

The results are in the table below:

- The Initial Value column shows ideal values that ignore any parasitic effect
- The Adjusted Value column shows the component values that compensate for $C_{ni(111)}$ and CLC111 is group delay (τ_{oa})
- The Standard Value column shows the nearest standard 1% resistors and 5% capacitors

Component	Initial	Value Adjusted	Standard
C_{1A}	30pF	30pF	30pF
R_{2A}	106Ω	92.8Ω	93.1Ω
$C_{ni(111)}$	–	1.3pF	1.3pF

Copyright © 1996–2013, Texas Instruments Incorporated

Section B Design:

1. Since Q_p = 1.000, set K_B to 1.00
2. Use the CLC446. This is a current-feedback op amp
   - f_{3dB} = 400MHz > f_H = 200MHz
   - f_{3dB} < $10f_c$ = 500MHz; the design will be sensitive to the op amp group delay
   - SR = 2000V/µs > 1000V/µs (see Item #1 in "Section A Design")
   - τ_{oa} ≈ 0.56ns at 10MHz
   - $C_{ni(446)}$ = 1.0pF (input capacitance)
3. Use the CLC446's recommended R_f at A_v = 1.0:
   - R_{fB} = 453Ω
   - Then leave R_{gB} open so that K_B = 1.00
4. Initialize the resistor level:
   - R ≈ 100Ω
5. Initialize the capacitor level, and the component ratios:

$$C \approx \frac{1}{2\pi(50.00\text{MHz})\cdot(100\Omega)} = 31.83\text{pF}$$

$$c^2 \approx 0.1000$$

$$r^2 \approx \max\{0.10, 0.0826\} = 0.1000$$

6. Recalculate C^2 and initialize the capacitors: C^2 ≈ 0.127 C_{1B} ≈ 89.3pF C_{3B} ≈ 11.3pF
7. Set the capacitors to the nearest standard values: C_{1B} ≈ 91pF C_{3B} ≈ 11pF
8. Recalculate the capacitor level and ratio, and the resistor level and ratio:
9. Calculate the resistors: R_{4B} = 324Ω R_{3B} = 31.2Ω
10. The sensitivities for this design are:

$$C = \sqrt{(91\text{pF})\cdot(11\text{pF})} = 31.64\text{pF}$$

$$c^2 = \frac{(11\text{pF})}{(91\text{pF})} = 0.1209$$

$$R = \frac{1}{2\pi(50.00\text{MHz})\cdot(31.64\text{pF})}$$

$$= 100.6\Omega$$

$$r^2 = 0.1056$$

αi	$S_{\alpha_i}^{H_\infty}$	$S_{\alpha_i}^{\omega_p}$	$S_{\alpha_i}^{Q_p}$
C_{1B}	0.00	−0.50	−0.39
C_{3B}	0.00	−0.50	0.39
R_{4B}	0.00	−0.50	0.50
R_{5B}	0.00	−0.50	−0.50
R_{fB}	0.00	0.00	0.00
R_{gB}	0.00	0.00	0.00
K	1.00	0.00	1.12

Copyright © 1996–2013, Texas Instruments Incorporated

Section B Pre-distortion:

1. The design gives these values:
   - $\omega_{p(nom)} = 2\pi(50.00\text{MHz})$
   - $Q_{p(nom)} = 1.000$
   - $K_B = 1.00$
   - $C_{1B} = 91\text{pF}$
   - $C_{3B} = 11\text{pF}$
2. Iteration 1 shows the initial design results. Iterations 2-4 pre-distort R_{4B} and R_{5B} to compensate for the CLC446's group delay, and for $C_{ni(446)}$:

Iteration #		1	2	3	4
$\frac{\omega_{p(pd)}}{2\pi}$	[MHz]	50.00	59.73	56.81	57.54
$Q_{p(pd)}$	[ ]	1.000	0.9320	0.9561	0.9505
R	[Ω]	10.6	84.22	88.54	87.42
r^2	[Ω]	0.0962	0.1108	0.1053	0.1065
R_{4B}	[Ω]	324.3	253.0	272.9	267.9
R_{5B}	[Ω]	31.21	28.03	28.73	28.53
τ_2	[ns]	0.324	0.253	0.273	0.268
τ_4	[ns]	1.741	1.511	1.575	1.559
$\frac{\omega_p}{2\pi}$	[MHz]	43.87	51.96	49.52	50.13
Q_p	[ ]	1.034	0.984	1.003	0.999

   - The midband gain estimate is:

$$H_\infty \approx 0.770[\text{V/V}].\ \text{Iteration } 1 \approx 0.759\ [\text{V/V}].\ \text{Iteration } 4 \quad (3)$$

   - The simulations gave a lower value for H_∞. Increasing K could help overcome this loss, but would also increase the sensitivities.
3. The resulting components are:

Component	Initial	Value Adjusted	Standard
C_{1B}	91pF	91pF	91pF
C_{3B}	11pF	11pF	11pF
$C_{ni(446)}$	–	1.0pF	1.0pF
R_{4B}	324Ω	268Ω	267Ω
R_{5B}	31.2Ω	28.5Ω	28.7Ω
R_{fB}	453Ω	453Ω	453Ω
R_{gB}	∞	∞	∞

Figure 3 and Figure 4 show simulated gains. The curve numbers are:

1. Ideal (Initial Design Values, $\tau_{oa} = 0$, $C_{ni} = 0$)
2. Without pre-distortion (Initial Design Values, $\tau_{oa} \neq 0$, $C_{ni} = 0$)
3. With pre-distortion (Pre-distorted Values, $\tau_{oa} \neq 0$, $C_{ni} = 0$)

Copyright © 1996–2013, Texas Instruments Incorporated

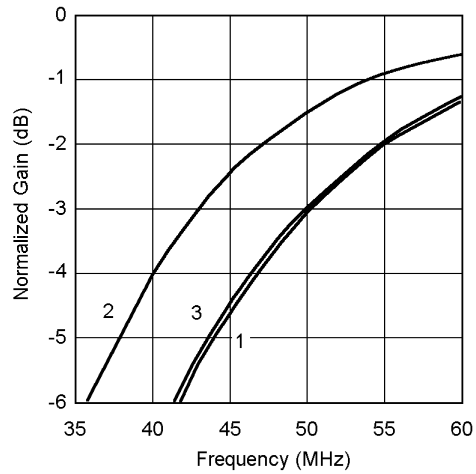

Figure 3. Simulated Filter Magnitude Response

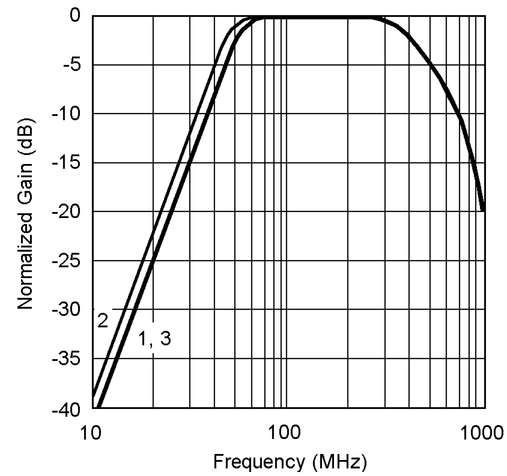

Figure 4. Simulated Filter Magnitude Response

6 SPICE Models

SPICE Models are available for most of Comlinear's amplifiers. These models support nominal DC, AC, AC noise and transient simulations at room temperature.

We recommend simulating with Comlinear's SPICE model to:

- Predict the op amp's influence on filter response
- Support quicker design cycles

Include board and component parasitic models to obtain a more accurate prediction of the filter's response.

To verify your simulations, we recommend bread-boarding your circuit.

7 Summary

This application report contains an easy to use design algorithm for a low sensitivities Sallen-Key highpass biquad. Designing for low ω_p and Q_p sensitivities gives:

- Reduced filter variation over process, temperature and time
- High manufacturing yield
- Lower component cost

A low sensitivity design is not enough to produce high manufacturing yields. This Application Note shows how to compensate for the op amp bandwidth, and for the [parasitic] input capacitance of the op amp. This method also applies to any other component or board parasitics. The components must also have low enough tolerance and temperature coefficients.

Copyright © 1996–2013, Texas Instruments Incorporated

Appendix A Derivation of Pre-distortion and Parasitic Capacitance Compensation Formulas

To pre-distort this filter, and compensate for the [parasitic] input capacitance of the op amp C_{ni}):

1. Use the method in reference [5] to include the op amp's effect on the filter response. The result is:

$$\frac{V_o}{V_{in}} \approx \frac{H_\infty\left(\frac{1}{\omega_p^2}\right)s^2}{1+\left(\frac{1}{(\omega_p Q_p)}\right)s+\left(\frac{1}{\omega_p^2}\right)s^2} \cdot e^{-\tau_{oa}s}$$

   where the op amp group delay (τ_{oa}) is evaluated at the passband edge frequency (f_c), and:

$$\frac{1}{(\omega_p Q_p)} = R_5C_1 + R_5C_3 - R_4C_3(K-1)$$

$$\frac{1}{\omega_p^2} = R_4R_5C_1C_3 + K\tau_{oa}R_4C_3$$

$$K = \frac{1+R_f}{R_g}$$

$$H_\infty = K$$

2. Since C_{ni} is in parallel with R_4, replace R_4 with the parallel equivalent of R_4 and C_{ni}:

$$\frac{R_4 \leftarrow R_4}{(1+R_4C_{ni}s)}$$

$$\frac{V_o}{V_{in}} \approx \frac{H_\infty\left(\frac{R_4C_3(R_5C_1+K\tau_{oa})}{1+R_4C_{ni}s}\right)s^2 \cdot e^{-\tau_{oa}s}}{\left(1+\left(\frac{R_4C_3(1-K)}{1+R_4C_{ni}s}+R_5(C_1+C_3)\right)s + \left(\frac{R_4C_3(R_5C_1+Kt_{oa})}{1+R_4C_{ni}s}\right)s^2\right)}$$

Copyright © 1996–2013, Texas Instruments Incorporated

3. After simplifying, we obtain:

$$\frac{V_o}{V_{in}} \approx \frac{H_\infty\left(\frac{1}{\omega_p^2}\right)s^2}{1+\left(\frac{1}{\left(\omega_p Q_p\right)}\right)s+\left(\frac{1}{\omega_p^2}\right)s^2} \cdot e^{-\tau_{oa}s}$$

where :

$$\frac{1}{\left(\omega_p Q_p\right)} = t_1 + t_2$$

$$\tau_1 = R_5C_1 + R_5C_3 - R_4C_3\,(K-1)$$

$$\tau_2 = R_4C_{ni}$$

$$\frac{1}{\omega_p^2} = \tau_3^2 + \tau_4^2$$

$$\tau_3^2 = R_4R_5C_1C_3$$

$$\tau_4^2 = K\tau_{oa}R_4C_3 + R_4R_5\left(C_1 + C_3\right)C_{ni}$$

$$K = \frac{1+R_f}{R_g}$$

$$H_\infty = \frac{K \cdot \left(\tau_3^2\right)}{\left(\tau_3^2 + \tau_4^2\right)}$$

Copyright © 1996–2013, Texas Instruments Incorporated

Appendix B Bibliography

1. R. Schaumann, M Ghausi and K. Laker, *Design of Analog Filters: Passive, Active RD, and Switched Capacitor.* New Jersey: Prentice Hall, 1990.
2. A. Zverev, *Handbook of FILTER SYNTHESIS.* John Wiley & Sons, 1967.
3. A. Williams and F. Taylor, *Electronic Filter Design Handbook*. McGraw Hill, 1995.
4. S. Natarajan, *Theory and Design of Linear Active Networks*. Macmillan, 1987.
5. *OA-21 Component Pre-Distortion for Sallen Key Filters Application Report* (SNOA369)
6. *OA-27 Low-Sensitivity, Lowpass Filter Design Application Report* (SNOA372)
7. *OA-28 Low-Sensitivity, Bandpass Filter Design with Tuning Method Application Report* (SNOA373)

Note: The circuits included in this application note have been tested with Texas Instruments parts that may have been obsoleted and/or replaced with newer products. Please refer to the CLC to LMH conversion table to find the appropriate replacement part for the obsolete device.

Copyright © 1996–2013, Texas Instruments Incorporated

IMPORTANT NOTICE

Texas Instruments Incorporated and its subsidiaries (TI) reserve the right to make corrections, enhancements, improvements and other changes to its semiconductor products and services per JESD46, latest issue, and to discontinue any product or service per JESD48, latest issue. Buyers should obtain the latest relevant information before placing orders and should verify that such information is current and complete. All semiconductor products (also referred to herein as "components") are sold subject to TI's terms and conditions of sale supplied at the time of order acknowledgment.

TI warrants performance of its components to the specifications applicable at the time of sale, in accordance with the warranty in TI's terms and conditions of sale of semiconductor products. Testing and other quality control techniques are used to the extent TI deems necessary to support this warranty. Except where mandated by applicable law, testing of all parameters of each component is not necessarily performed.

TI assumes no liability for applications assistance or the design of Buyers' products. Buyers are responsible for their products and applications using TI components. To minimize the risks associated with Buyers' products and applications, Buyers should provide adequate design and operating safeguards.

TI does not warrant or represent that any license, either express or implied, is granted under any patent right, copyright, mask work right, or other intellectual property right relating to any combination, machine, or process in which TI components or services are used. Information published by TI regarding third-party products or services does not constitute a license to use such products or services or a warranty or endorsement thereof. Use of such information may require a license from a third party under the patents or other intellectual property of the third party, or a license from TI under the patents or other intellectual property of TI.

Reproduction of significant portions of TI information in TI data books or data sheets is permissible only if reproduction is without alteration and is accompanied by all associated warranties, conditions, limitations, and notices. TI is not responsible or liable for such altered documentation. Information of third parties may be subject to additional restrictions.

Resale of TI components or services with statements different from or beyond the parameters stated by TI for that component or service voids all express and any implied warranties for the associated TI component or service and is an unfair and deceptive business practice. TI is not responsible or liable for any such statements.

Buyer acknowledges and agrees that it is solely responsible for compliance with all legal, regulatory and safety-related requirements concerning its products, and any use of TI components in its applications, notwithstanding any applications-related information or support that may be provided by TI. Buyer represents and agrees that it has all the necessary expertise to create and implement safeguards which anticipate dangerous consequences of failures, monitor failures and their consequences, lessen the likelihood of failures that might cause harm and take appropriate remedial actions. Buyer will fully indemnify TI and its representatives against any damages arising out of the use of any TI components in safety-critical applications.

In some cases, TI components may be promoted specifically to facilitate safety-related applications. With such components, TI's goal is to help enable customers to design and create their own end-product solutions that meet applicable functional safety standards and requirements. Nonetheless, such components are subject to these terms.

No TI components are authorized for use in FDA Class III (or similar life-critical medical equipment) unless authorized officers of the parties have executed a special agreement specifically governing such use.

Only those TI components which TI has specifically designated as military grade or "enhanced plastic" are designed and intended for use in military/aerospace applications or environments. Buyer acknowledges and agrees that any military or aerospace use of TI components which have ***not*** been so designated is solely at the Buyer's risk, and that Buyer is solely responsible for compliance with all legal and regulatory requirements in connection with such use.

TI has specifically designated certain components as meeting ISO/TS16949 requirements, mainly for automotive use. In any case of use of non-designated products, TI will not be responsible for any failure to meet ISO/TS16949.

Products		**Applications**	
Audio	www.ti.com/audio	Automotive and Transportation	www.ti.com/automotive
Amplifiers	amplifier.ti.com	Communications and Telecom	www.ti.com/communications
Data Converters	dataconverter.ti.com	Computers and Peripherals	www.ti.com/computers
DLP® Products	www.dlp.com	Consumer Electronics	www.ti.com/consumer-apps
DSP	dsp.ti.com	Energy and Lighting	www.ti.com/energy
Clocks and Timers	www.ti.com/clocks	Industrial	www.ti.com/industrial
Interface	interface.ti.com	Medical	www.ti.com/medical
Logic	logic.ti.com	Security	www.ti.com/security
Power Mgmt	power.ti.com	Space, Avionics and Defense	www.ti.com/space-avionics-defense
Microcontrollers	microcontroller.ti.com	Video and Imaging	www.ti.com/video
RFID	www.ti-rfid.com		
OMAP Applications Processors	www.ti.com/omap	**TI E2E Community**	e2e.ti.com
Wireless Connectivity	www.ti.com/wirelessconnectivity		

Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265

Copyright © 2013, Texas Instruments Incorporated