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# Reduction of Operational Amplifiers Finite Gain Effects in Switched-Capacitor Low-Pass Notch Biquads

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**Abstract:** An combined approach for reducing the errors in the notch frequency $f_z$, in the pole frequency $f_p$, in the quality factor $Q_p$ and in the amplitude $H_p$ at the pole frequency of switched-capacitor low-pass notch biquads is presented. At first, the conventional integrators in the biquads are replaced with gain- and offset-compensated integrators. Subsequently, the errors $\Delta f_z/f_z$, $\Delta f_p/f_p$, $\Delta Q_p/Q_p$ and $\Delta H_p/H_p$ are minimized by modifying the values of the integrating capacitances and of the appropriately chosen zero-forming and pole-forming capacitances. The effectiveness of this approach is demonstrated by designing two low-pass notch biquad topologies which realize the same transfer function.

**Keywords:** filters, gain- and offset- compensation, operational amplifiers, switched-capacitor integrators

## 1 Introduction

One of the important nonideal properties which influence the performance of the switched-capacitor (SC) circuits is the finite dc gain $A$ of the operational amplifiers (op amps). In filters, the finite gain causes errors in both the amplitude and the phase responses [1, 2]. From the point of view of simplifying amplifier design and improving high-frequency capability several gain- and offset- compensated (GOC) SC building blocks (integrators, gain stages, sample-and–hold circuits) have been reported in the literature. The phase error $\theta(\omega)$ of the GOC integrators is proportional to $1/A^2$ (in a conventional integrator this is a simple inverse dependence $1/A$). In most of the GOC integrators proposed the reduction in phase error was obtained at the expense of increased gain error m$(\omega)$.

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The SC low-pass notch (LPN) biquads are commonly realized using a feedback loop containing one inverting and one noninverting conventional integrators.

The gain errors of the integrators affect the notch frequency $f_z$ and the pole frequency $f_p$ of the biquads, while the phase errors affect the zero $Q$-factor $Q_z$, the pole $Q$-factor $Q_p$, the attenuation $H_z$ at the notch frequency and the magnitude $H_p$ of the biquad transfer function at the pole frequency.

A combined approach for minimization the effects of op amps finite gain in switched-capacitor bandpass biquads has been presented in [3].

In this paper the approach proposed in [3] is adapted for reducing the errors in the notch frequency $f_z$, in the pole frequency $f_p$ , in the quality factor $Q_p$ and in the amplitude $H_p$ at the pole frequency of SC LPN biquads with low but precisely known and stable op amps dc gain [4]. The effectiveness of this approach is verified for two LPN biquad topologies which realize the same transfer function.

## 2 Proposed design approach

The $z$-domain transfer function of the LPN biquads has the general form

$$H(z) = \frac{N(z)}{D(z)} = k\frac{1 + a_1 z^{-1} + a_2 z^{-2}}{1 + b_1 z^{-1} + b_2 z^{-2}}, \tag{1}$$

which for any pairs of complex conjugate zeros and poles can be rewritten as

$$H(z) = k\frac{1 - 2\, R_z\, \cos\theta_z\, z^{-1} + R_z^2 z^{-2}}{1 - 2R_P \cos\theta_P z^{-1} + R_P^2 z^{-2}}. \tag{2}$$

Here, $R_z$ is the radius and $\theta_z$ is the angle to the zero; $R_p$ and $Q_p$ correspond to the pole radius and angle, respectively. From (2) the following relationships for the notch frequency $f_z$ , the zero $Q$-factor $Q_z$ , the pole frequency $f_p$ and the pole $Q$-factor $Q_p$ can be derived:

$$f_z = \frac{f_s}{2\pi}\sqrt{\theta_z^2 + (\ln R_z)^2}, \quad Q_z = -\frac{\pi f_z/f_s}{\ln R_z} \tag{3}$$

and

$$f_p = \frac{f_s}{2\pi}\sqrt{\theta_p^2 + (\ln R_p)^2}, \quad Q_p = -\frac{\pi f_p/f_s}{\ln R_p}, \tag{4}$$

where $f_s$ is the sampling frequency.

For small ratios $f_z/f_s$, $f_p/f_s$ and high $Q$-factors the frequencies $f_z$, $f_p$ and the $Q$-factors $Q_z$, $Q_p$ are approximately given by

$$f_z \approx \frac{f_s}{2\pi}\sqrt{1 + a_1 + a_2}, \quad Q_z \approx \frac{\sqrt{1 + a_1 + a_2}}{1 - a_2} \tag{5}$$

and

$$f_p \approx \frac{f_s}{2\pi}\sqrt{1+b_1+b_2}, \quad Q_p \approx \frac{\sqrt{1+b_1+b_2}}{1-b_2}. \tag{6}$$

For standard design of LPN biquads the op amps gain values are assumed to be infinite. Then, the coefficients in (1) are functions of the capacitances only and $Q_z \to \infty$ $(a_2 = 1)$. In this case, from (5), (6) and (1) the logarithmic sensitivities of the $f_z$, $f_p$, $Q_p$ and of the amplitude $H_p$ at the pole frequency to the capacitances can be obtained.

The proposed combined approach for minimization of the errors $\Delta f_z/f_z$, $\Delta f_p/f_p$, $\Delta Q_p/Q_p$, $\Delta H_p/H_p$ and for enhancement of $Q_z$ consists in the following consecutive steps, in which the calculations are made for the nominal value $A_o$ of the op amps dc gain $A$:

**Step 1**. First, to reduce the effect of op amp imperfections ( dc gain $A$ and offset voltage $V_{OS}$) the conventional integrators in the LPN biquad considered are replaced with Nagaraj-86 [5] and Ki-89 [6] GOC SC integrators. These simple bi-phase integrators form an excellent GOC integrator-pair without using extra clock phases or holding circuits to satisfy the sampling conditions. The reduced phase errors of the GOC integrators provide a reduction in the errors $\Delta Q_p/Q_p$, $\Delta H_p/H_p$ and enhance the quality factor $Q_z$.

**Step 2**. The gain error $m(\omega)$ of the integrators is equivalent to an element value variation $\Delta C_i$ of the integrating capacitance$C_i$ . If the finite dc gain $A_o$ is known, the value of $C_i$ can be replaced with $C_i' = C_i(1+m)$ in the two integrators of the GOC biquad, thereby essentially reducing the gain errors $m(\omega)$ [7]. This prewarping technique automatically provides a reduction in the notch frequency error $\Delta f_z/f_z$ and in the pole frequency error $\Delta f_p/f_p$ of the biquad considered.

**Step 3**. For some LPN biquad configurations the compensation in Step 2 is not sufficiently effective. In this case the errors $(\Delta f_z/f_z)_2$ and $(\Delta f_p/f_p)_2$, obtained in the previous step, can be further reduced by modifying one zero-forming capacitance $C_z$ and one pole-forming capacitance $C_{p1}$. The new capacitance values $C_z'$ and $C_{p1}'$ are the solutions of the equations

$$S_{C_Z}^{f_z}\left(\frac{C_z'}{C_z}-1\right) = -\left(\frac{\Delta f_z}{f_z}\right)_2 \tag{7}$$

and

$$S_{C_{p_1}}^{f_p}\left(\frac{C_{p1}'}{C_{p_1}}-1\right) = -\left(\frac{\Delta f_p}{f_p}\right)_2. \tag{8}$$

The sensitivities $S_{C_Z}^{f_z}$ and $S_{C_{p_1}}^{f_p}$ are calculated using (5) and (6) for the initial values of the capacitances and infinite op amp dc gain.

**Step 4**. The errors $(\Delta Q_p/Q_p)_2$ and $(\Delta H_p/H_p)_2$ obtained in Step 2, or the errors $(\Delta Q_p/Q_p)_3$ and $(\Delta H_p/H_p)_3$ obtained in Step 3, can be further minimized by modifying one another pole-forming capacitance $C_{p2}$. This capacitance is chosen such that the following relations hold:

$$S_{C_{p_2}}^{Q_p} \approx S_{C_{p_2}}^{H_p}, \quad \left|S_{C_{p_2}}^{f_p}\right| \ll 1, \quad \left|S_{C_{p_2}}^{f_z}\right| \ll 1. \tag{9}$$

The new capacitance value $C'_{p2}$ is calculated from the equations

$$S_{C_{p_2}}^{Q_p}\left(\frac{C'_{p_2}}{C_{p_2}}-1\right) = -\left(\frac{\Delta Q_p}{Q_p}\right)_r \tag{10}$$

and

$$S_{C_{p_2}}^{H_p}\left(\frac{C'_{p_2}}{C_{p_2}}-1\right) = -\left(\frac{\Delta H_p}{H_p}\right)_r. \tag{11}$$

The terms $(\Delta Q_p/Q_p)_r$ and $(\Delta H_p/H_p)_r$ on the right hand sides are the errors $(\Delta Q_p/Q_p)_2$ and $(\Delta H_p/H_p)_2$ obtained in Step 2, or the errors $(\Delta Q_p/Q_p)_3$ and $(\Delta H_p/H_p)_3$ obtained in Step 3.

The following two cases are considered:

a) The capacitance $C'_{p2}$ is the solution of equation (10), i.e.,

$$C'_{p2} = C_{p2}\left[1-\frac{(\Delta Q_p/Q_p)_r}{S_{C_{p_2}}^{Q_p}}\right]. \tag{12}$$

b) The capacitance $C'_{p2}$ is the average of the solutions of (10) and (11), i.e.,

$$C'_{p2} = C_{p2}\left\{1-0.5\left[\frac{(\Delta Q_p/Q_p)_r}{S_{C_{p_2}}^{Q_p}}+\frac{(\Delta H_p/H_p)_r}{S_{C_{p_2}}^{H_p}}\right]\right\}. \tag{13}$$

## 3 Application of the proposed approach

The approach proposed is illustrated by means of two LPN biquad topologies. For comparative purposes the two biquads are designed to fulfil the same specifications:

$$f_p \cong 1\,\text{kHz}, \quad f_z \cong 2.5\,\text{kHz}, \quad Q_p \cong 30, \quad H_p \cong 28\,\text{dB}, \quad f_s \cong 12.5\,\text{kHz}.$$

Fig.1 shows the circuit schema of the Martin and Sedra's type-I LPN biquad with conventional integrator [8]. The relative capacitance values are: $C_1$=2.027, $CF_1$=1, $CB_1$=1, $CB_2$=1, $C_2$=59.207, $CF_2$=29.945 and $CF_3$=10.687.

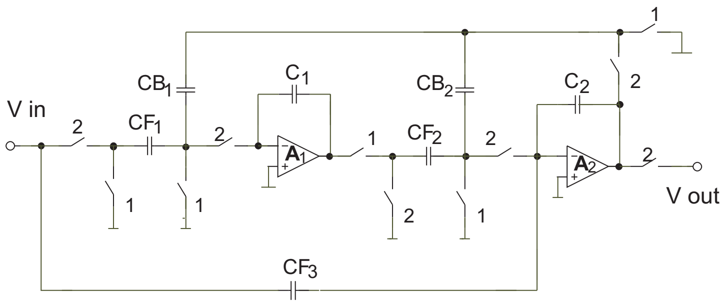


Fig. 1. Martin and Sedra's type-I LPN biquad with conventional integrators.

The ideal $z$-domain transfer function is

$$H_{id}^{22}(z) = -\frac{CF_3}{C_2}\frac{z^{-2} - \left[2 - \frac{CF_1 CF_2}{C_1\, CF_3}\right] z^{-1} + 1}{z^{-2} - \left[2 + \frac{CB_2}{C_2} - \frac{CB_1 CF_2}{C_1 C_2}\right] z^{-1} + \frac{CB_2}{C_2} + 1}. \tag{14}$$

The frequencies $f_z$ and $f_p$, and the quality factor $Q_p$ are approximately given by

$$f_z \approx \frac{f_s}{2\pi}\sqrt{\frac{CF_1 CF_2}{C_1 CF_3}}, \qquad f_p \approx \frac{f_s}{2\pi}\sqrt{\frac{CB_1 CF_2}{C_1(C_2+CB_2)}}$$
$$Q_p \approx \sqrt{\frac{CB_1 CF_2 (C_2 + CB_2)}{C_1 CB_2^2}}. \tag{15}$$

It was found that for nominal op amps gains $A_{0_1} = A_{0_2} = 100$ the deviations of $f_z$, $f_p$, $Q_p$ and $H_p$ from the ideal case are:

$$(\Delta f_z/f_z)_c = -0.852\%, \quad (\Delta f_p/f_p)_c = -1.427\%,$$
$$(\Delta Q_p/Q_p)_c = -47.33\%, \quad (\Delta H_p/H_p)_c = -47.25\%.$$

The zero $Q$-factor and the attenuation at $f_z$ are $Q_{zc} = 128.18$ and $H_{zc} = -56.78$dB.

According to the approach proposed the first integratior in the conventional biquad (Fig.1) is replaced with the Nagaraj-86 integrator and the second integrator-with the Ki-89 integrator. The resulting filter is shown in Fig.2, where $C_{h1} = 1$ and $C_{h2} = C_2$.

The performance parameters of the GOC biquad for $A_{0_1} = A_{0_2} = 100$ are

$$\begin{aligned}
&(\Delta f_z/f_z)_1 = -1.4033\%, \quad (\Delta f_p/f_p)_1 = -2.0818\%,\\
&(\Delta Q_p/Q_p)_1 = -1.4044\%, \quad (\Delta H_p/H_p)_1 = -1.1374\%\\
&Q_{z1} = 13000, \quad H_{z1} = -96.91\,\text{dB}, \quad H(2500\,\text{Hz}) = -45.87\,\text{dB}.
\end{aligned} \tag{16}$$

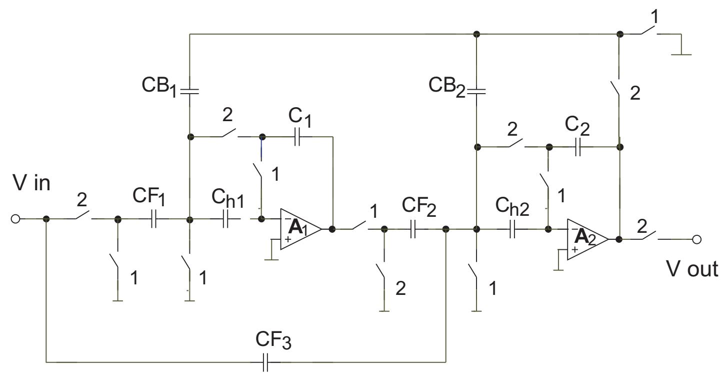


Fig. 2. Martin and Sedra's type-I LPN biquad with GOC integrators.

First, for reducing the errors $(\Delta f_z/f_z)_1$ and $(\Delta f_p/f_p)_1$ of the GOC biquad, the integrating capacitances $C_1$ and $C_2$ are modified according to the expressions [9]

$$C_1' = \left(C_1 - \frac{CF_1 + CB_1 + C_{h_1}}{A_{0_1}}\right)\left(1 + \frac{1}{A_{0_1}}\right)^{-1} \tag{17}$$

and

$$C_2' = \left(C_2 - \frac{CB_2 + CF_2 + CF_3}{A_{0_2}}\right)\left(1 + \frac{1}{A_{0_2}}\right)^{-1}. \tag{18}$$

For $A_{0_1} = A_{0_2} = 100$ one obtains $C_1' = 1.977228$ and $C_2' = 58.208594$. The corresponding performance parameters of the biquad are

$$\begin{aligned} &(\Delta f_z/f_z)_2 = -3.25.10^{-5}\%, \quad (\Delta f_p/f_p)_2 = -2.075.10^{-3}\%, \\ &(\Delta Q_p/Q_p)_2 = -0.98282\%, \quad (\Delta H_p/H_p)_2 = -0.98410\%, \\ &Q_{z_2} = 12867, \quad H_{z_2} = -96.68\,\text{dB}, \quad H(2500\,\text{Hz}) = -84.35\,\text{dB}. \end{aligned} \tag{19}$$

From (15) and (14) one finds

$$S^{Q_P}_{CB_2} = -\frac{2C_2 + CB_2}{2(C_2 + CB_2)} = -0.99170, \quad S^{f_P}_{CB_2} = -\frac{CB_2}{2(C_2 + CB_2)} = -8.305.10^{-3},$$

$$S^{H_P}_{CB_2} = -1.00162.$$

Nearly the same errors $(\Delta Q_p/Q_p)_2$ and $(\Delta H_p/H_p)_2$, on one hand and nearly the same sensitivities $S^{Q_P}_{CB_2}$ and $S^{H_P}_{CB_2}$ on the other hand suggest that the errors $(\Delta Q_p/Q_p)_2$ and $(\Delta H_p/H_p)_2$ can be further reduced by modifying the capacitance $CB_2$. The choice of this capacitance is based also on the low sensitivity $S^{f_P}_{CB_2}$.

The new capacitance value $CB_2'$ is given by the expression

$$CB_2' = CB_2 \left[ 1 - \frac{(\Delta Q_p/Q_p)_2}{S_{CB_2}^{Q_P}} \right]. \tag{20}$$

One obtains $CB_2' = 0.9900895$. The corresponding performance parameters of the biquad are

$$(\Delta f_z/f_z)_3 = -3.171.10^{-5}\%, \quad (\Delta f_p/f_p)_3 = 6.355.10^{-3}\%,$$
$$(\Delta Q_p/Q_p)_3 = -5.928.10^{-3}\%, \quad (\Delta H_p/H_p)_3 = 6.873.10^{-7}\%,$$
$$Q_{z_3} = 12867, \quad H_{z_3} = -96.68\,\text{dB}, \quad H(2500\,\text{Hz}) = -84.35\,\text{dB}.$$

The capacitance $CB_2'$ can be made equal to the unit capacitance. Then, the new values of the capacitances $C_2, CF_2$ and $CF_3$ are

$$C_2 = 58.791247, \quad CF_2 = 30.244742 \quad \text{and} \quad CF_3 = 10.793974.$$

By rounding-off the values of the capacitances to the third digit after the decimal point we finally obtain

$$C_1 = 1.977, \quad CF_1 = 1, \quad CB_1 = 1, \quad C_{h_1} = 1, \quad C_2 = 58.791,$$
$$CB_2 = 1, \quad CF_2 = 30.245, \quad CF_3 = 10.794, \quad C_{h_2} = 58.791.$$

Table 1 summarizes the performance parameters of the GOC Martin and Sedra's type-I LPN biquad with rounded-off capacitances and gain variation $A_{0_1} = A_{0_2} = 100 \pm 8$.

Table 1. Performance parameters of the GOC Matrin and Sedra's type-I LPN biquad with rounded-off capacitances

| A | 92 | 100 | 108 |
|---|---|---|---|
| $(\Delta f_z/f_z)$ [%] | $-0.1164$ | $6.884.10^{-3}$ | 0.1122 |
| $(\Delta f_p/f_p)$ [%] | $-0.1699$ | 0.0128 | 0.1690 |
| $(\Delta Q_p/Q_p)$ [%] | $-0.2156$ | $1.727.10^{-4}$ | 0.1734 |
| $(\Delta H_p/H_p)$ [%] | $-0.1860$ | $6.113.10^{-3}$ | 0.1596 |
| $Q_z$ | 10906 | 12865 | 14997 |
| $H(f_z)$ [dB] | $-95.26$ | $-96.68$ | $-97.99$ |
| $H$(2.5kHz) [dB] | $-68.42$ | $-81.29$ | $-66.31$ |

The GOC-version of the conventional Huang and Sansen's LPN biquad (Fig.5 from [10]) is shown in Fig.3. The relative capacitance values are: $C_1 = 2.02$, $C_{1S} = 27.79$, $CF_1 = 1$, $CB_1 = 1$, $CB_2 = 1$, $C_2 = 5.633$, $CF_2 = 2.79$, $CF_3 = 1$, $C_{h_1} = 1$ and $C_{h_2} = C_2$.

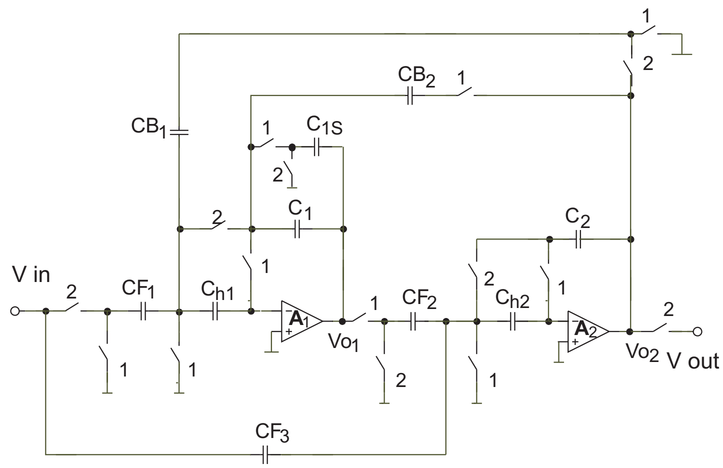


Fig. 3. Huang and Sansen's LPN biquad with GOC integrators.

The performances parameters of the GOC biquad for $A_{0_1} = A_{0_2} = 100$ are

$$\begin{aligned} &(\Delta f_z/f_z)_1 = -1.6892\%, \quad (\Delta f_p/f_p)_1 = -2.3416\%, \\ &(\Delta Q_p/Q_p)_1 = -0.2035\%, \quad (\Delta H_p/H_p)_1 = 0.058\% \\ &Q_{z_1} = 6622, \quad H_{z_1} = -91.04\,\mathrm{dB}, \quad H(2500\,\mathrm{Hz}) = -44.02\,\mathrm{dB}. \end{aligned} \tag{21}$$

For the modified values of the integrating capacitances $C_1' = 1.970297$ and $C_2' = 5.539703$ the performance parameters of the biquad are

$$\begin{aligned} &(\Delta f_z/f_z)_2 = -0.28253\%, \quad (\Delta f_p/f_p)_2 = -0.25599\%, \\ &(\Delta Q_p/Q_p)_2 = 0.07455\%, \quad (\Delta H_p/H_p)_2 = 0.0660\%, \\ &Q_{z_2} = 6625, \quad H_{z2} = -90.91\,\mathrm{dB}, \quad H(2500\,\mathrm{Hz}) = -58.52\,\mathrm{dB}. \end{aligned} \tag{22}$$

The errors $(\Delta f_z/f_z)_2$ and $(\Delta f_p/f_p)_2$ can be further reduced by modifying the zero-forming capacitance $CF_1$ and the pole-forming capacitance $CB_1$. The new capacitance values $CF_1'$ and $CB_1'$ are the solutions of the equations

$$S_{CF_1}^{f_z}\left(\frac{CF_1'}{CF_1} - 1\right) = -\left(\frac{\Delta f_z}{f_z}\right)_2, \quad S_{CB_1}^{f_p}\left(\frac{CB_1'}{CB_1} - 1\right) = -\left(\frac{\Delta f_p}{f_p}\right)_2. \tag{23}$$

One obtains $CF_1' = 1.0056506$ and $CB_1' = 1.0051199$, for which the performance parameters of the biquad are

$$\begin{aligned} &(\Delta f_z/f_z)_3 = 3.961.10^{-2}\%, \quad (\Delta f_p/)_3 = 1.792.10^{-3}\,\%, \\ &(\Delta Q_p/Q_p)_3 = 0.3300\%, \quad (\Delta H_p/H_p)_3 = 0.3972\%, \\ &Q_{z_3} = 6629, \quad H_{z_3} = -90.92\,\mathrm{dB}, \quad H(2500\,\mathrm{Hz}) = -88.19\,\mathrm{dB}. \end{aligned} \tag{24}$$

The errors $(\Delta Q_p/Q_p)_3$ and $(\Delta H_p/H_p)_3$ can be reduced by modifying the capacitance $CB_2$. The new capacitance value $CB_2'$ is the average of the capacitance $(CB_2')_{Q_P}$ and $(CB_2')_{H_P}$ calculated from the equations

$$S_{CB_2}^{Q_P}\left(\frac{(CB_2')_{Q_P}-CB_2}{CB_2}\right)=-\left(\frac{\Delta Q_p}{Q_p}\right)_3 \tag{25}$$

and

$$S_{CB_2}^{H_P}\left(\frac{(CB_2')_{H_P}-CB_2}{CB_2}\right)=-\left(\frac{\Delta H_p}{H_p}\right)_3. \tag{26}$$

This results in $CB_2'=1.0036366$.

Table 2 summarizes the performance parameters of the GOC Huang and Sansen's LPN biquad with rounded-off capacitances and gain variation $A_{0_1}=A_{0_2}=100\pm 8$.

Table 2. Performance parameters of the GOC Huang and Sansen's LPN biquad with rounded-off capacitances

| A | 92 | 100 | 108 |
|---|---|---|---|
| $(\Delta f_z/f_z)$ [%] | $-0.0800$ | 0.0682 | 0.1949 |
| $(\Delta f_p/f_p)$ [%] | $-0.2032$ | $2.133.10^{-3}$ | 0.1777 |
| $(\Delta Q_p/Q_p)$ [%] | $-0.22606$ | $-0.062$ | 0.0603 |
| $(\Delta H_p/H_p)$ [%] | $-0.0791$ | 0.0622 | 0.1653 |
| $Q_z$ | 5619 | 6631 | 7728 |
| $H(f_z)$ [dB] | $-89.50$ | $-90.93$ | $-92.25$ |
| $H(2.5\text{kHz})$ [dB] | $-67.44$ | $-77.25$ | $-64.24$ |

## 4 Conclusion

A combined approach for reducing the effects of op amps finite gain in switched-capacitor low-pass notch biquads has been presented. The effectiveness of the approach proposed has been demonstrated by two low-pass notch biquad topologies which realize the same transfer function. The filters with modified capacitances have approximately an order smaller of magnitude relative errors

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