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A new filter configuration using current feedback op-amp

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Abstract

A new general configuration which employs the current feedback op-amp and realizes second-order lowpass, bandpass and highpass responses is given. The proposed filters have the attractive advantage that the effects of the stray capacitance *Cz* and *Rz* can be absorbed within the filter components in all of the three cases. The proposed lowpass filter has the additional advantage of having the two capacitors grounded and the effect of *Rx* can also be absorbed within one of the filter resistors. PSpice simulation results that demonstrate the effects of a nonideal current feedback op-amp are included. Experimental results are also given. © 1998 Published by Elsevier Science Ltd. All rights reserved.

1. Introduction

The application of the current feedback op-amp (CFOA) [1–4] in the realization of second-order highpass, bandpass and lowpass filters has recently been reported in the literature [5,6]. The lowpass filter given in [5] requires an additional buffer stage and it employs two floating capacitors. The lowpass filter given in [6] employs one grounded *C* and one floating *C* and does not need a buffer. The bandpass filter given in [6], however, realizes only real axis pole, that is $Q < 0.5$, which can be achieved by any passive RC circuit. A lowpass filter based on the Sallen–Key circuit and realized using a single CFOA has been reported very recently [7]; one of the two capacitors, however, is floating. The applications of the CFOA in realizing filters based on the simulation of a grounded inductor [8] as well as floating inductor circuits [9,10] have also been considered in the literature. Several multiple CFOA filter circuits that employ grounded capacitors have been reported in [7,11] and are suitable for high-Q applications. In some applications it may be desirable to have an economical single CFOA low-Q or medium-Q filters [12], especially in cases which require maximally flat magnitude lowpass or highpass responses, or wide-bandwidth bandpass filters. The low-Q filters are defined as those with $Q \leq 2$, and the medium-Q filters have $Q \leq 20$ [12].

In this paper a new single CFOA configuration which realizes second-order bandpass, lowpass and highpass transfer functions is introduced. The proposed bandpass filter employs the minimum number of passive elements, namely two resistors and two capacitors; a third resistor or capacitor is needed to provide independent control on the center frequency gain of the filter. The lowpass filter has the attractive advantage of having both capacitors grounded. The proposed configuration is suitable for low-Q and medium-Q applications. The basic advantage of the proposed structure is that the effects of *Cz* and *Rz* are absorbed by the filter components, besides which the effect of *Rx* is also absorbed by one of the resistors in the lowpass filter as well as in one of the two proposed bandpass filters.

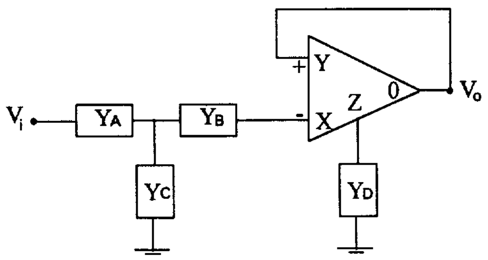

Fig. 1. The proposed general configuration.

2. The proposed configuration

Fig. 1 represents the proposed general configuration which employs the CFOA as the active building block. The X terminal, which is also defined as the inverting input terminal, is characterized by a very low input impedance. The Y terminal, which is also defined as the non-inverting input, has a very high input impedance. The two outputs Z and O exhibit a very high and a very low output impedance, respectively.

The voltage transfer function of this circuit is given by:

$$T(s) = \frac{-Y_A Y_B}{(Y_A + Y_B + Y_C)Y_D - (Y_A + Y_C)Y_B} \quad (1)$$

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Table 1
The ω_0 and the Q passive sensitivities of the bandpass filter I and the lowpass filter

Filter	S	x	C_1	C_2	R_1	R_2	R_3
	—	$S^{\omega_0}_x$	$-\frac{1}{2}$	$-\frac{1}{2}$	$-\frac{1}{2}$	$-\frac{1}{2}$	—
Bandpass Fig. 2(a) with $R_3=\infty$	Equal C	S^Q_x	$Q^2-\frac{1}{2}$	$\frac{1}{2}-Q^2$	$-\frac{1}{2}$	$\frac{1}{2}$	—
	Equal R	S^Q_x	$\frac{1}{2}$	$-\frac{1}{2}$	$\frac{1}{2}-Q^2$	$Q^2-\frac{1}{2}$	—
Lowpass Fig. 3(a)	Equal R	$S^{\omega_0}_x$	$-\frac{1}{2}$	$-\frac{1}{2}$	0	-1	0
		S^Q_x	$\frac{1}{2}$	$-\frac{1}{2}$	$\frac{1}{2}$	$2Q^2-1$	$\frac{1}{2}-2Q^2$

This configuration is capable of realizing bandpass, lowpass and highpass voltage responses as explained next.

3. The bandpass filter I

Taking $Y_A=1/R_1$, $Y_B=sC_1$, $Y_C=0$, $Y_D=sC_2+(1/R_2)$, the transfer function becomes:

$$T(s)=\frac{-sC_1R_2}{s^2C_1C_2R_1R_2+s(C_1R_1+C_2R_2-C_1R_2)+1} \tag{2}$$

which realizes a bandpass response with ω_0 and Q given by:

$$\omega_0=\frac{1}{\sqrt{C_1C_2R_1R_2}},\quad Q=\frac{\sqrt{C_1C_2R_1R_2}}{C_1R_1+C_2R_2-C_1R_2} \tag{3}$$

To realize a specified ω_0 and Q the following two designs are considered:

Equal C design: taking $C_1=C_2=C$, the design equations for R_1 and R_2 are obtained as:

$$R_1=\frac{1}{\omega_0QC}\ \text{and}\ R_2=\frac{Q}{\omega_0C} \tag{4}$$

Equal R design: taking $R_1=R_2=R$, the design equations for C_1 and C_2 are given by:

$$C_1=\frac{Q}{\omega_0R}\ \text{and}\ C_2=\frac{1}{\omega_0QR} \tag{5}$$

In both cases the magnitude of the gain at ω_0 is given by:

$$T_0\equiv|T(j\omega_0)=Q^2 \tag{6}$$

From eq. (3) the ω_0 and the Q passive sensitivities can be obtained and are summarized in Table 1 for the equal C as well as for the equal R design.

The circuit can be modified by adding one more resistor R_3 as shown in Fig. 2(a) in order to realize a specified gain at ω_0 equal to T_0 smaller than Q^2. In this case the transfer

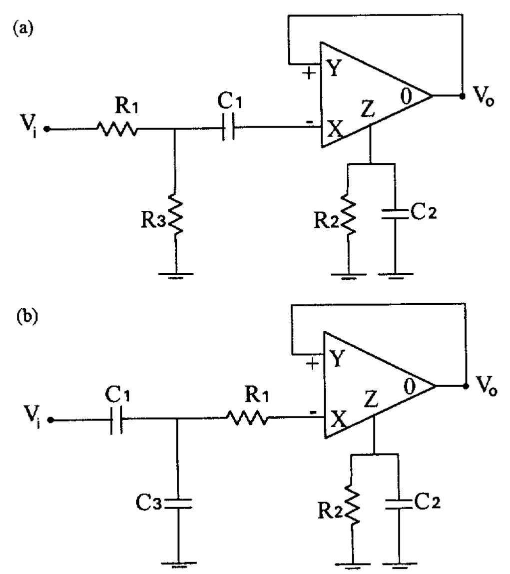

Fig. 2. (a) The bandpass filter I. (b) The bandpass filter II.

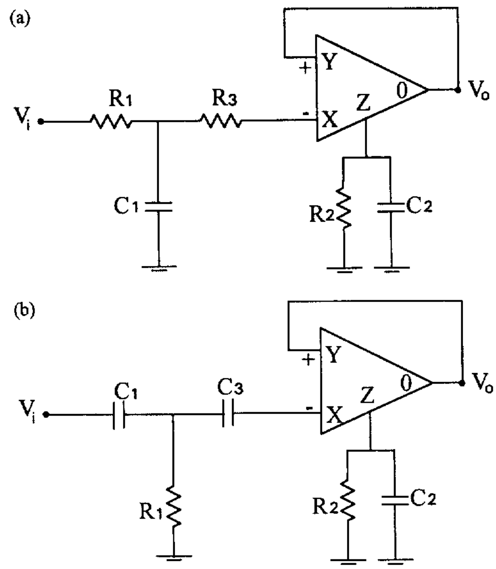

Fig. 3. (a) The lowpass filter. (b) The highpass filter.

function is given by:

$$T(s)=\frac{-sC_1R_2}{s^2C_1C_2R_1R_2+s\left(C_1R_1+(C_1R_1+(C_2R_2-C_1R_2)\left(1+\frac{R_1}{R_3}\right)\right)+\left(1+\frac{R_1}{R_3}\right)} \tag{7}$$

Taking $C_1 = C_2 = C$ the design equations are given by:

$$R_1=\frac{Q}{\omega_0CT_0},\ R_2=\frac{Q}{\omega_0C},\ R_3=\frac{Q}{\omega_0C}\left(\frac{1}{Q^2-T_0}\right) \tag{8}$$

3.1. The bandpass filter II

Using the RC:CR transformation, a second bandpass filter circuit is obtained and shown in Fig. 2(b). The transfer function in this case is given by

$$T(s)=\frac{-sC_1R_2}{s^2C_2(C_1+C_3)R_1R_2+s(C_1R_1+C_3R_1+C_2R_2-C_1R_2-C_3R_2)+1} \tag{9}$$

One of the recommended designs in this case is to take

$$C_1+C_3=C_2 \tag{10}$$

In this case the ω_0, Q and T_0 of the filter are given by:

$$\omega_0=\frac{1}{C_2\sqrt{R_1R_2}},\ Q=\sqrt{\frac{R_2}{R_1}}\text{ and }T_0=\frac{C_1R_2}{C_2R_1} \tag{11}$$

The design equations for specified ω_0, Q and T_0 are given by:

$$R_1=\frac{1}{\omega_0QC_2}\text{ and }R_2=\frac{Q}{\omega_0C_2} \tag{12a}$$

$$C_1=\frac{T_0C_2}{Q^2}\text{ and }C_3=C_2-C_1 \tag{12b}$$

where C_2 is chosen much larger than Cz in order to eliminate its effect on the filter performance. It is also possible to compensate the effect of Cz, by taking the design value of C_2 equal to its theoretical value minus Cz. The improvements that can be achieved in the magnitude and phase characteristics will be demonstrated in Section 4.

An alternative recommended design is the equal R design; in this case the design equations for specified ω_0, Q and T_0 are given by:

$$R_1=R_2=R \tag{13}$$

$$C_1=\frac{T_0}{\omega_0QR},\ C_2=\frac{1}{\omega_0QR}\text{ and }C_3=C_2\left[Q^2-T_0\right]$$

where R_1 is chosen much larger than Rx in order to eliminate its effect on the filter performance. It is also possible to compensate the effect of Rx by taking the design value of R_1 equal to its theoretical value plus Rx.

3.2. The lowpass filter

Taking $Y_A = 1/R_1, Y_B = 1/R_3, Y_C = sC_1$ and $Y_D = sC_2 + (1/R_2)$, the grounded capacitor lowpass filter shown in Fig. 3(a) is obtained, which has the following transfer function:

$$T(s)=\frac{-\frac{1}{C_1C_2R_1R_3}}{s^2+s\left(\frac{1}{C_1R_1}+\frac{1}{C_2R_2}+\frac{1}{C_1R_3}-\frac{1}{C_2R_3}\right)+\frac{1}{C_1C_2R_1R_2}\left(1+\frac{R_1}{R_3}-\frac{R_2}{R_3}\right)} \tag{14}$$

For a specified ω_0 and Q, the design equations are given by:

Equal R design: taking

$$R_1=R_2=R_3=R$$

$$C_1=\frac{2Q}{\omega_0R}\text{ and }C_2=\frac{1}{2Q\omega_0R} \tag{15}$$

In this case dc gain $= -1$.

The ω_0 and the Q passive sensitivities are given in Table 1, from which it is seen that a lowpass Butterworth response will have Q sensitivities to any of the passive circuit components equal to ± 0.5.

3.3. The highpass filter

Fig. 3(b) represents the highpass filter which can be easily obtained from the lowpass filter by applying the RC:CR transformation.

For the case of interest, namely the equal C design, the transfer function is given by:

$$T(s)=\frac{-s^2C^2R_1R_2}{s^2C^2R_1R_2+2sCR_1+1} \tag{16}$$

It is seen that the magnitude of the high frequency gain is equal to one.

For a specified ω_0 and Q the design equations are given by:

$$C_1=C_2=C_3=C \tag{17a}$$

$$R_1=\frac{1}{2\omega_0QC}\text{ and }R_2=\frac{2Q}{\omega_0C} \tag{17b}$$

Expressions for the ω_0 and Q passive sensitivities similar to those given in Table 1 can be obtained easily, and are not included here to limit the length of the paper.

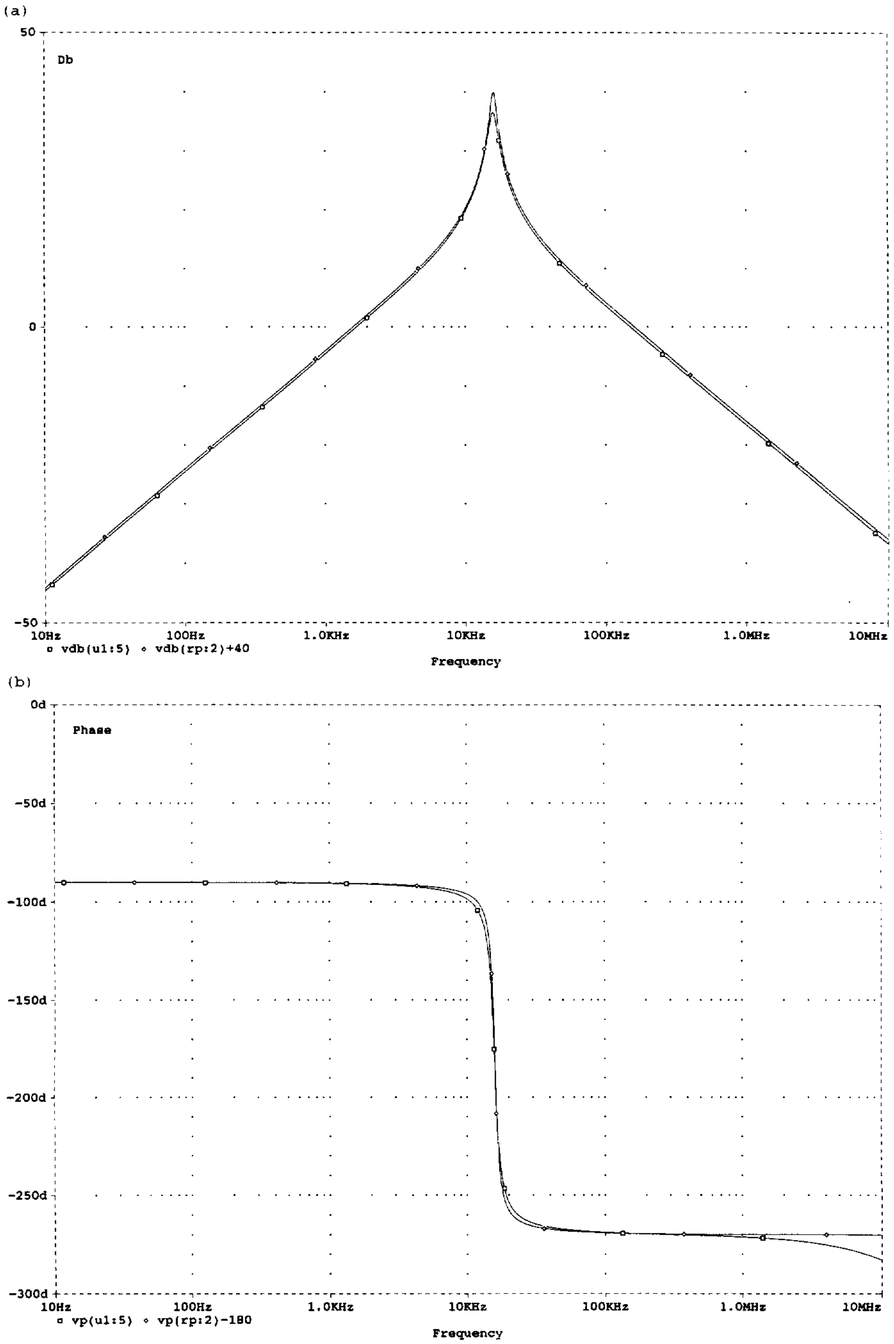

Fig. 4. (a) The magnitude characteristics of the bandpass filter I, with $R_3 = \infty$. (b) The phase characteristics of the bandpass filter I, with $R_3 = \infty$.

4. Effects of non-ideal CFOA

The proposed filter configuration has the attractive advantage that the effects of Cz and Rz can always be absorbed in the filter components C_2 and R_2 in all types of filters considered here. The lowpass filter as well as the bandpass filter II have the additional advantage that Rx can also be absorbed in the resistor R_3 in the lowpass filter or in R_1 in the case of the bandpass filter II. Of course the bandpass filter I with $R_3 = \infty$ has the same advantage that the effect of Rx can be absorbed in R_1.

In this section the effect of Rx on the bandpass filter I with a finite R_3 is considered. Taking the effect of Rx into consideration and for equal C design, the transfer function of the circuit of Fig. 2(a) becomes:

$$T(s) = \frac{-sCR_2}{s^2C^2R_1R_2\left(1+\frac{Rx}{R_{13}}\right) + sCR_1\left(1+\frac{Rx}{R_{13}}\right) + \frac{R_1}{R_{13}}} \tag{18a}$$

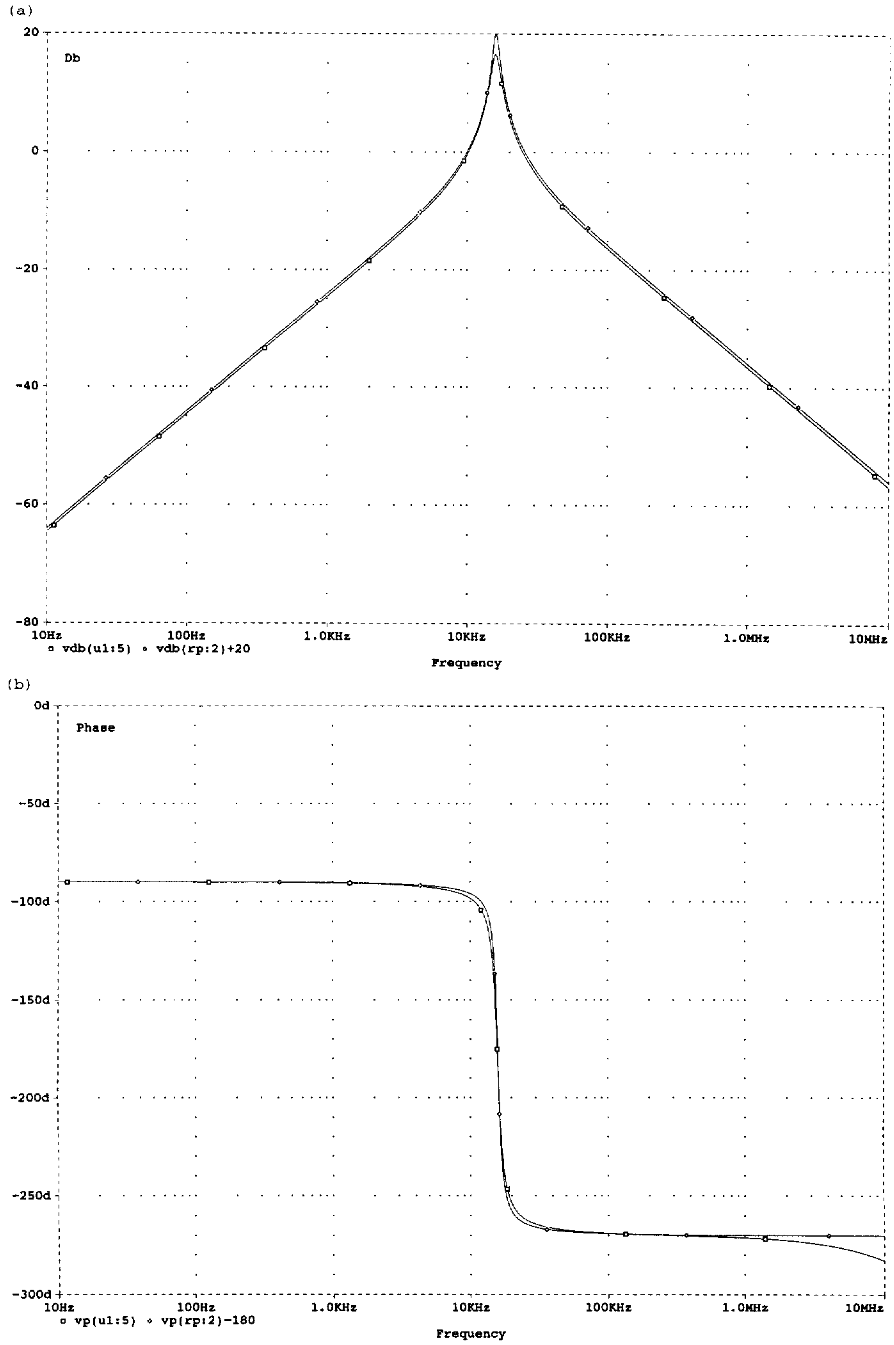

Fig. 5. (a) The magnitude characteristics of the bandpass filter I, with $T_0 = 10$. (b) The phase characteristics of the bandpass filter I, with $T_0 = 10$.

where

$$R_{13} = \frac{R_1 R_3}{R_1 + R_3} \tag{18b}$$

Thus, the actual values of ω_0, Q and T_0 are given by:

$$\omega_{0a} = \frac{1}{C\sqrt{R_2(R_{13}+Rx)}}, \quad Q_a = \sqrt{\frac{R_2}{R_{13}+Rz}} \text{ and } T_{0a} \tag{19}$$

$$= \frac{R_2 R_{13}}{R_1(R_{13}+Rx)}$$

It is seen that the effect of Rx is to reduce ω_0, Q and T_0. In order to minimize the effect of Rx on the filter characteristics, it is necessary to take $R_{13} >> Rx$. It is also possible to compensate completely the effects of Rx on ω_0 and Q, by taking the design value of R_{13} equal to its theoretical value minus Rx. The center frequency gain, however, will be slightly reduced from its ideal value.

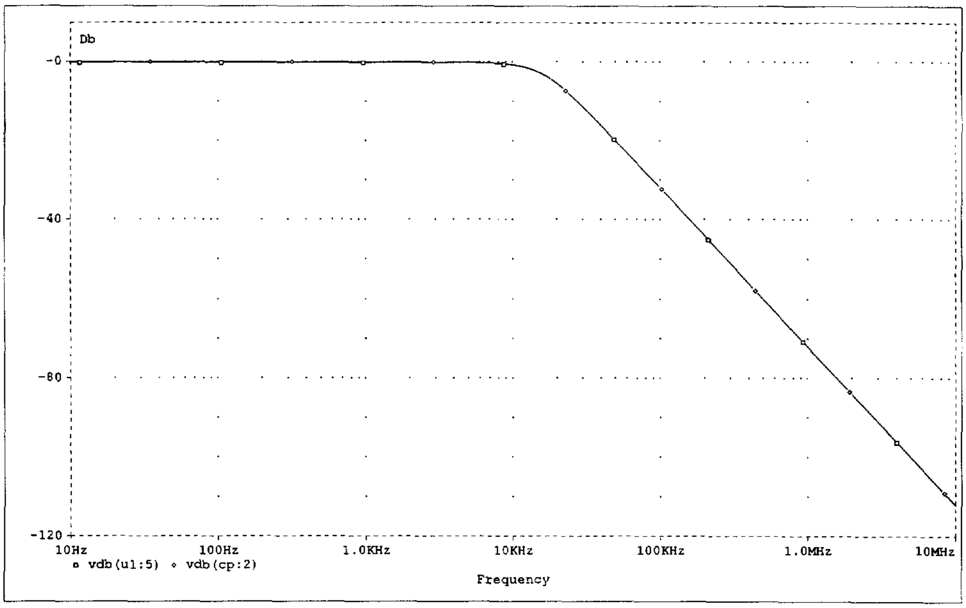

Fig. 6. The magnitude characteristics of the Butterworth lowpass filter.

Similar expressions can be obtained for the effect of Rx on the highpass filter characteristics.

5. Simulation and experimental results

PSpice simulations for the circuits of Figs. 2 and 3 have been carried out using the AD844A from Analog Devices biased with ± 9 V.

Fig. 4 represents the magnitude and phase responses for the circuit of Fig. 2(a) designed to realize $\omega_0 = 100$ krad/sec and $Q = 10$ and using the equal C design by taking $C_1 = C_2 = 1$ nF, $R_1 = 1$ kΩ and $R_2 = 100$ kΩ. The simulated responses are compared with the ideal responses obtained from a passive RLC filter designed for the same ω_0 and Q (by adding 40 dB to the magnitude of the gain and subtracting 180° from the phase, so that comparison can be meaningful). It is seen that a slight difference exists between the two responses owing to the non-ideal CFOA.

Fig. 5 represents the magnitude and phase responses of the bandpass filter of Fig. 2(a) designed for $\omega_0 = 100$ krad/sec, $Q = 10$ and $T_0 = 10$. The simulations are based on taking $C_1 = C_2 = 1$ nF, $R_1 = 10$ kΩ, $R_3 = 1.111$ kΩ and $R_2 = 100$ kΩ.

Fig. 6 represents the simulated magnitude response of the lowpass filter of Fig. 3(a) designed for $\omega_0 = 100$ krad/sec and $Q = 0.707$ (Butterworth response), by taking $R_1 = R_2 = R_3 = 10$ kΩ, $C_1 = 1.414$ nF and $C_2 = 0.707$ nF. It is seen that the magnitude response coincides exactly with the ideal lowpass response.

Fig. 7(a) represents the simulated magnitude response of the highpass filter of Fig. 3(b) designed for $\omega_0 = 100$ krad/sec, $Q = 0.707$, by taking $C_1 = C_2 = C_3 = 1$ nF, $R_1 = 7.07$ kΩ and $R_2 = 14.14$ kΩ. It is seen that the response coincides with the ideal response up to 1 MHz and starts to deviate from the ideal response due to the Rx of the CFOA. The high frequency magnitude response can be slightly improved by proper magnitude scaling, as shown in Fig. 7(b), which is based on taking $C_1 = C_2 = C_3 = 0.1$ nF, $R_1 = 70.7$ kΩ and $R_2 = 141.4$ kΩ.

In order to demonstrate the improvements that can be achieved in the filter characteristics when the effects of the dominant stray component are compensated, the following PSpice simulations are included.

Fig. 8 represents the magnitude and phase characteristics of the bandpass filter II of Fig. 2(b) designed to realize $\omega_0 = 100$ krad/sec, $Q = 10$ and $T_0 = 10$ by taking $C_1 = 0.1$ nF, $C_3 = 0.9$ nF, $R_1 = 1$kΩ and $R_2 = 100$ kΩ. Taking the effect of $Cz = 5.5$ pF into consideration, the design value of C_2 is taken as 994.5 pF. It is seen that the simulated responses are very close to the ideal responses.

Fig. 9 represents the magnitude and phase characteristics of the bandpass filter I of Fig. 2(a) with $R_3 = \infty$, designed to realize $\omega_0 = 100$ krad/sec, $Q = 20$, by taking $C_1 = 1$ nF, $C_2 = 995$ pF, $R_1 = 500$ Ω and $R_2 = 200$ kΩ. It is seen that the simulated responses are very close to the ideal responses.

Fig. 10 represents the magnitude and phase characteristics of the bandpass filter II of Fig. 2(b) designed to realize $\omega_0 = 100$ krad/sec, $Q = 20$ and $T_0 = 100$. Using the equal R design, and taking $C_1 = 50$ nF, $C_2 = 0.5$ nF, $C_3 = 150$ nF,

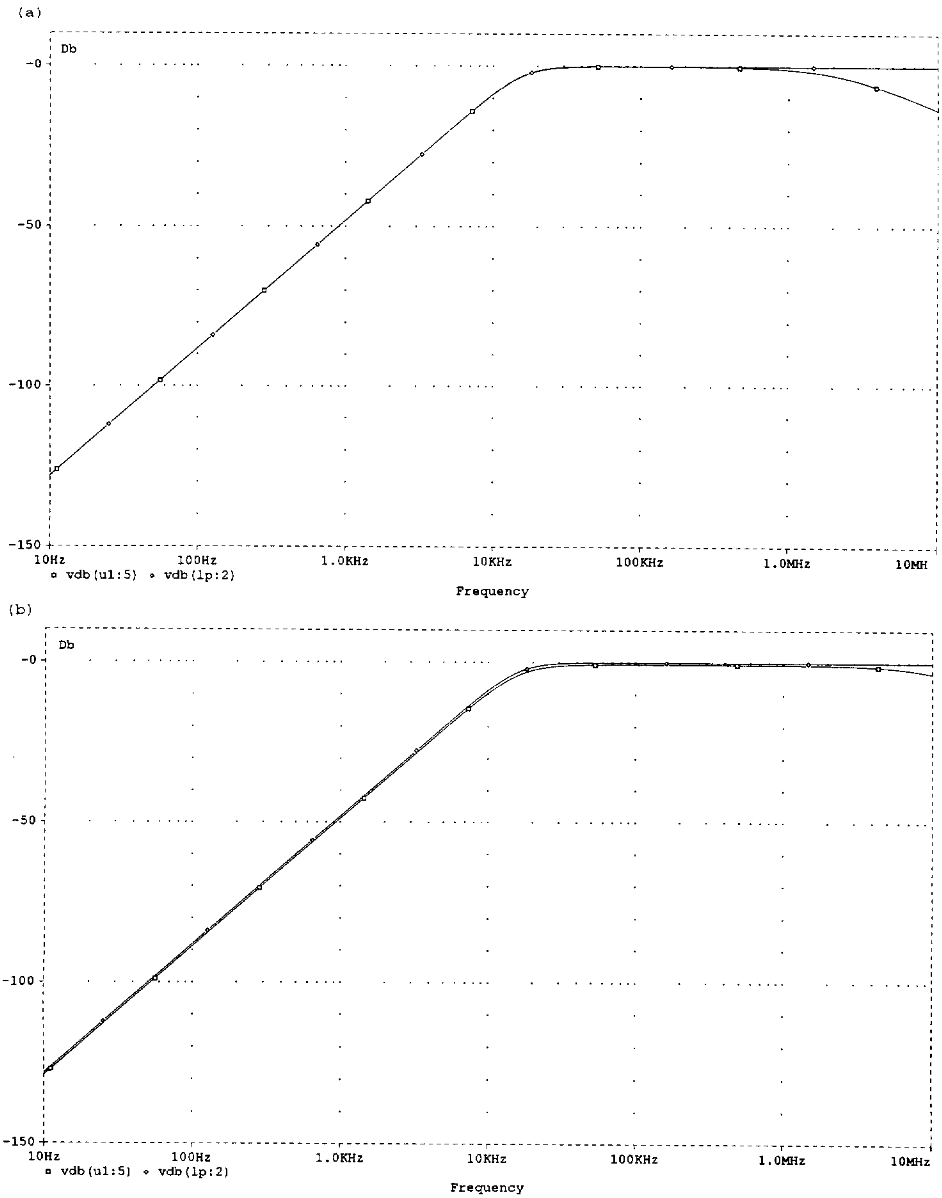

Fig. 7. The magnitude characteristics of the Butterworth highpass filter.

$R_1 = 1\text{k}\Omega$ and $R_2 = 1\ \text{k}\Omega$. For the equal R design, the dominant parasitic element in this case is Rx. Taking the effect of $Rx = 65\ \Omega$ into consideration, the design value of R_1 is taken as 935 Ω. From the simulations it is seen that excellent filter characteristics are obtained.

Several experimental measurements have been taken in order to verify the performance of the filters. The bandpass filter of Fig. 2(a) was realized with $R_1 = 100\ \text{k}\Omega$, $R_2 = 500\ \text{k}\Omega$, $R_3 = 25\ \text{k}\Omega$ and $C_1 = C_2 = 1$ nF. The center frequency gain was measured and found to be equal to 5.5 approximately, and the measured center frequency = 1.7 kHz.

The lowpass filter of Fig. 3(a) was realized with $C_1 = 2$ nF, $C_2 = 1$ nF and $R_1 = R_2 = R_3 = 10\ \text{k}\Omega$. The dc gain was measured and found to be -1 and the measured $f_{3\text{dB}} = 13.1$ kHz, which is slightly higher than its theoretical value.

It is seen that the experimental results confirm the results of the theoretical analysis, except for a slight error due to the tolerances in the discrete resistors and capacitors used in the laboratory.

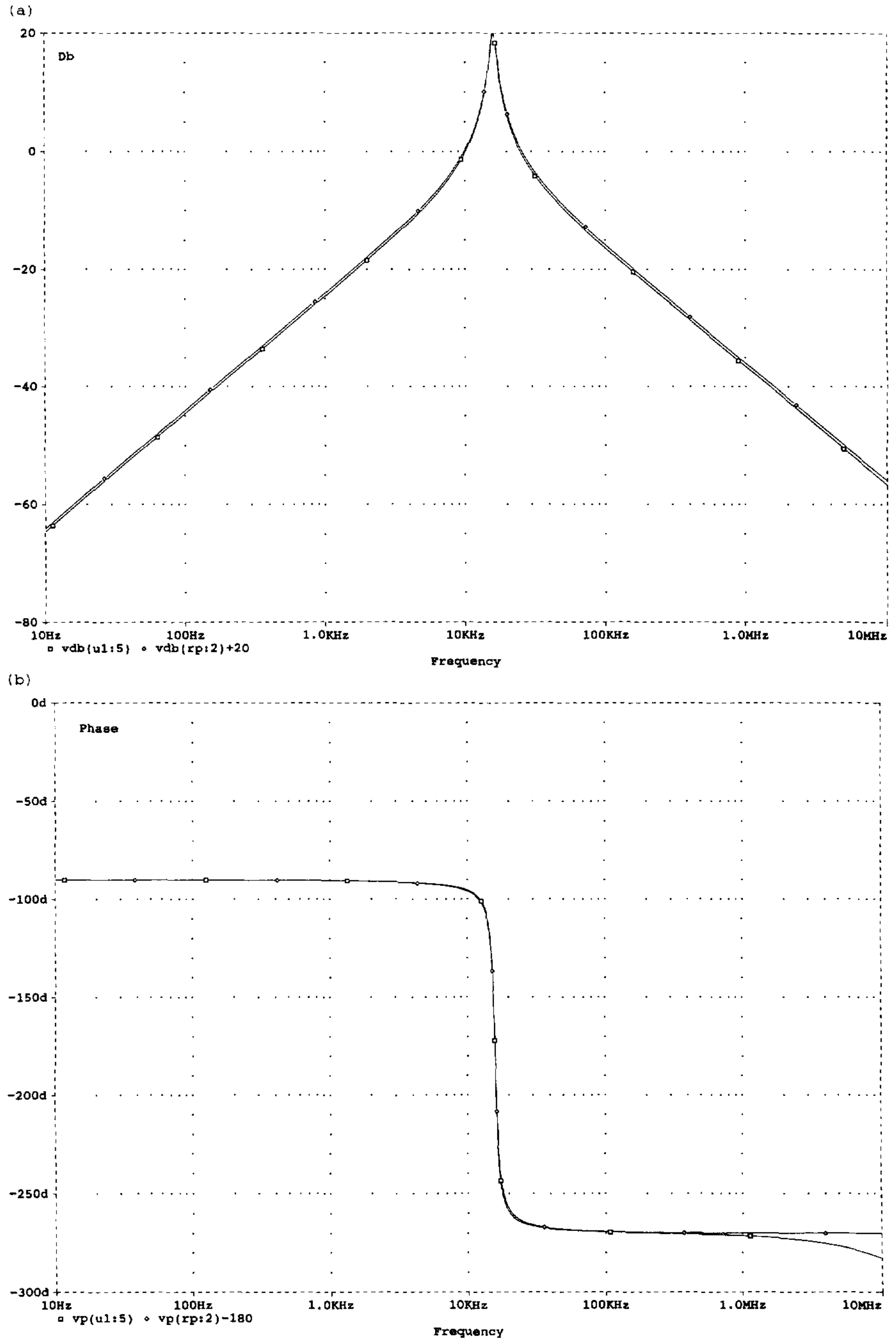

Fig. 8. (a) The magnitude characteristics of the compensated bandpass filter II, with $Q = 10$ and $T_0 = 10$. (b) The phase characteristics of the compensated bandpass filter II, with $Q = 10$ and $T_0 = 10$.

6. Conclusions

A new generalized configuration using a single CFOA has been introduced. The proposed configuration is capable of realizing lowpass, bandpass and highpass responses of low-Q and medium-Q values [12]. The proposed lowpass filter has the attractive advantages of using grounded capacitors and the circuit parameters can absorb the effects of Rx, Cz and Rz completely. The simulations and the experimental results indicate almost ideal characteristics.

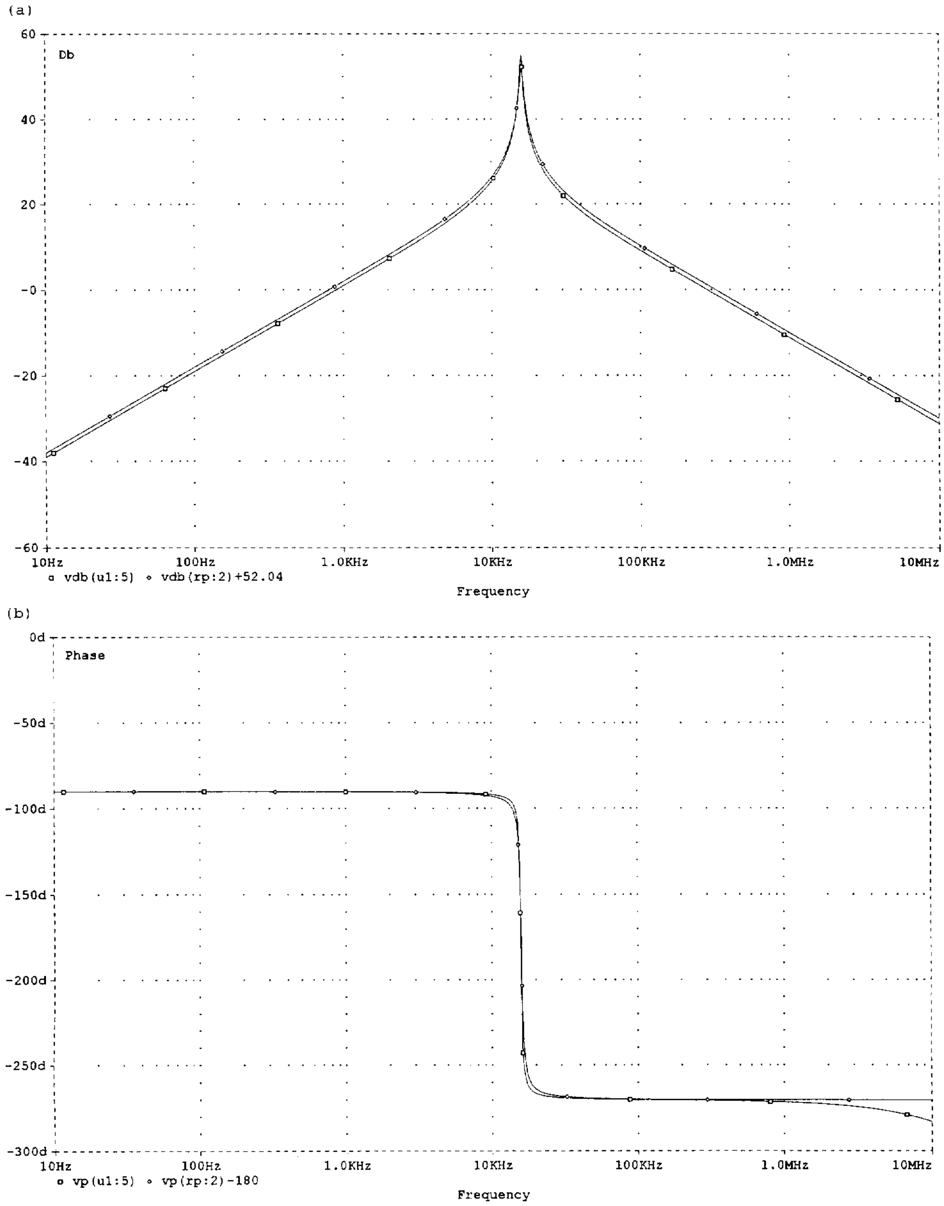

Fig. 9. (a) The magnitude characteristics of the compensated bandpass filter I, with $R_3 = \infty$ and $Q = 20$. (b) The phase characteristics of the compensated bandpass filter I, with $R_3 = \infty$ and $Q = 20$.

Two bandpass filter circuits that are related by the *RC*:*CR* transformation are also given. Compensation can be achieved in the equal *C* design case by modification of the value of C_2 to absorb the effect of *Cz*. On the other hand, in the equal *R* design case, R_1 in the bandpass filter II (or R_{13} in bandpass filter I) should be modified to compensate the effect of *Rx*. Simulation results that demonstrate the filter characteristics are included. A highpass filter is also included. PSpice simulations and experimental results are given.

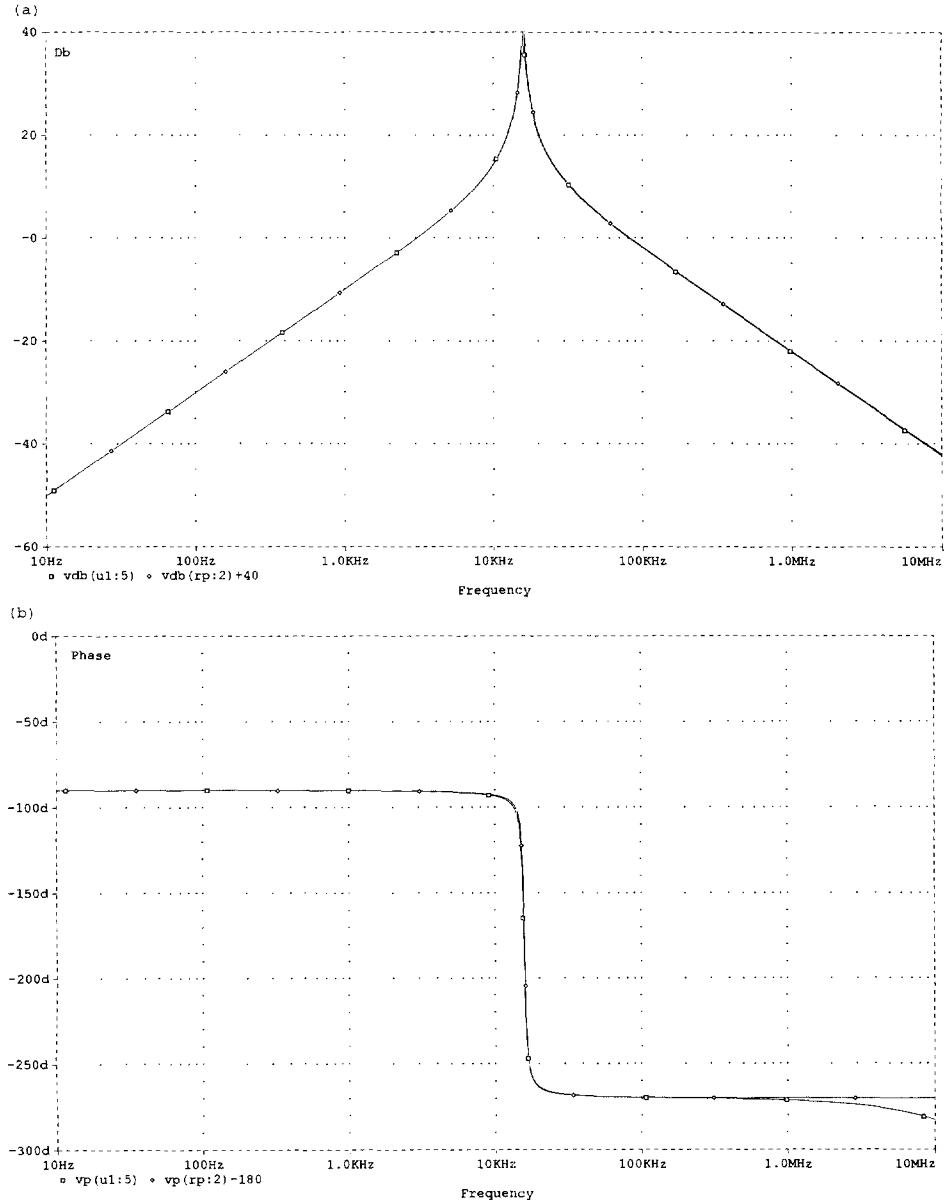

Fig. 10. (a) The magnitude characteristics of the compensated bandpass filter II, with $Q = 20$ and $T_0 = 100$. (b) The phase characteristics of the compensated bandpass filter II, with $Q = 20$ and $T_0 = 100$.

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