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Current-mode Multi-input Single-output Filter based on CCCDTAs

***Abstract**. This paper presents three current-mode multi-input single-output filters based on the current controlled current differencing transconductance amplifier (CCCDTA). They can be set to serve as standard filter functions. The pole frequency and quality factor parameters are such that filters can be electronically controlled and independently tuned via input bias current of the CCCDTAs. All output filters have high-impedances in current-mode configuration. Their performances were simulated by using PSPICE which resulted in a correlation with the theoretical prediction.*

Streszczenie.** W artykule zaprezentowano trzy filtry bazujące na current controlled current differencing transconductance amplifier CCCDTA. Filtry mogą być strojone za pośrednictwem wejścuowego prądu bias układu CCCDTA. **Wielowejściowy filtr bazujący na układzie CCCDTA

Keywords: Current-mode, multi-input single-output, CCCDTA, grounded capacitor.
Słowa kluczowe: filtr, układ CCCDTA,

Introduction

Analog filters are useful in analog signal processing. For example, they are used at the front-end of ADCs or smoothing on the output of DACs in audio systems, Tx/Rx in radio systems, and instrument and measurement systems [1-3]. Also, current-mode circuits have apparently been more popular in the synthesis and development of filters than voltage-mode circuits, since they have the advantages of simple circuitry, greater linearity, wider dynamic ranges, a high slew rate, high speed, etc. [4-10] as well as current-mode circuits which consume low voltage and low current thus yielding low power consumption.

Recently, the main research topic for analog filters has required five response standard filters containing a highpass filter (HPF), a bandpass filter (BPF), a lowpass filter (LPF), a bandreject filter (BRF) or a notch filter and an allpass filter (APF) in the same configuration or circuit topology. Particularly, the multi-input single-output (MISO) filters have been of interest in reports in the literature [11-27,29-30]. The various realizations of MISO filters based-on different high performance active building such as current differencing transcoductance amplifiers (CDTA), current controlled current differencing transcoductance amplifiers (CCCDTAs), current-controlled current conveyors II (CCCII), current-controlled current conveyor transconductance amplifiers (CCCCTA), voltage differencing transconductance amplifiers (VDTA), differential voltage current conveyor transconductance amplifiers (DVCCTA) etc. As mentioned above [11-23,24-27], the advantages and disadvantages are as follows. The circuits in [11-14,16-17] exhibit the advantage of high output impedance which can drive loads directly, but external resistors have to be used. Although the external resistors can be replaced by digital potentiometers, the circuits become more complicated. Also, the proposed filter circuits in [11,13-14,16,21] cannot be tuned to the pole frequency and quality factor electronic ally that are improper for use in [2,17-19,26] can be electronically tuned by adjusting the transconductance gain (g_m) in which the pole frequency or quality factor were adjusted but they suffered from the use of external resistors which cause difficulties in circuit integration [28]. Furthermore, the filters in [17] used a single active element which is smart and complex, but they require additional current gains or two outputs (2 g_m) for the circuits to become complicated. The filter circuits in [11,13-16,20-25,27] can be electronically adjusted to the pole frequency and quality factor since they use grounded capacitors which are suitable for IC implementation. However, they suffer from the use of excessive active elements.

The purpose of this paper is to introduce current-mode MISO filters made of two CCCDTAs. Their advantages are shown as follows:

1. The proposed circuits can be operated as standard functions of filters including HPF, LPF, BPF, BRF and APF with the selection via input signals.
2. The pole frequency and quality factor of the proposed filters can be electronically tuned through the input bias current fed to the CCCDTAs.
3. The quality factor can be independently tuned without effect on the pole frequency.
4. The output ports of the proposed filter have high-impedances, which allow easy drive load or cascade in current-mode configuration.
5. The proposed circuits used only two grounded capacitors, which are suitable for integrated circuit implementation.

Principle and Operation

The research of current controlled current differencing transconductance amplifiers or CCCDTAs was conducted by [31] in 2006 which was developed from the CDTA [32]. It has the finite input resistance: R_p and R_n at the input port p and n, respectively. The intrinsic resistance is equal and can be controlled by the bias current I_{B1}. The symbols and equivalent circuits of CCCDTAs are illustrated in Fig.1 a) and b), respectively. The characteristics of the ideal CCCDTA are represented as shown in the following hybrid matrix:

$$\begin{bmatrix} V_p \\ V_n \\ I_z \\ I_x \end{bmatrix} = \begin{bmatrix} R_p & 0 & 0 & 0 \\ 0 & R_n & 0 & 0 \\ 1 & -1 & 0 & 0 \\ 0 & 0 & 0 & \pm g_m \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ V_x \\ V_z \end{bmatrix} \tag{1}$$

For the CCCDTA implemented by a BJT technology, the intrinsic resistances (R_p and R_n) and the transconductance gain g_m can be shown by the following equation:

$$R_p = R_n = \frac{V_T}{2I_{B1}}, \quad \text{and} \tag{2}$$

$$g_m = \frac{I_{B2}}{2V_T}, \tag{3}$$

where: V_T - thermal voltage, I_{B1} and I_{B2} - DC bias current.

The V_T can be approximately measured 26 mV at room temperature. I_{B1} and I_{B2} are the DC bias currents used to control the intrinsic resistances and transconductance gains of CCCDTAs, respectively.

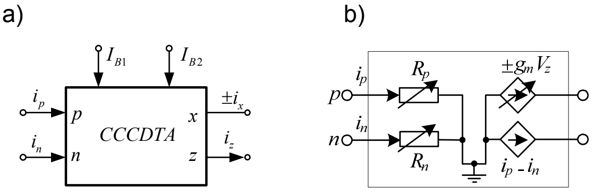

Fig.1. CCCDTA a) Symbol , b) Equivalent circuit.

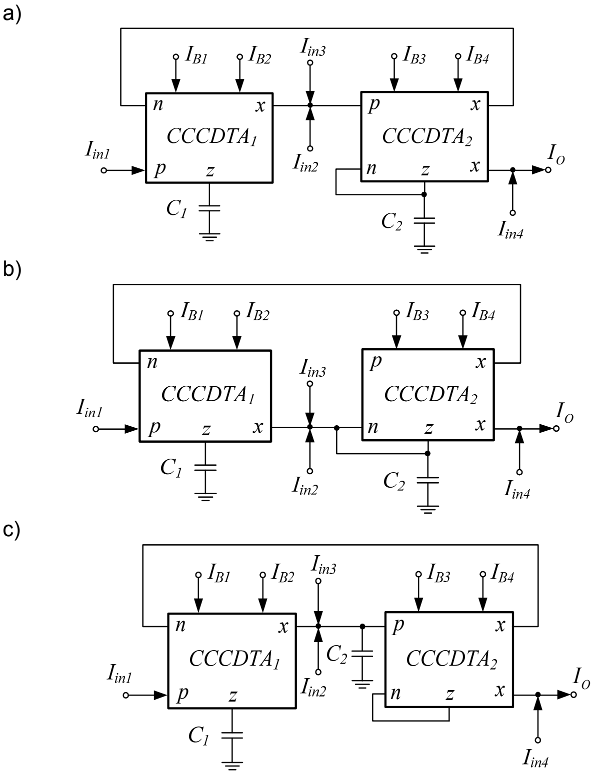

Fig.2. Proposed current-mode MISO filters.

Proposed current-mode MISO filters

The proposed current-mode MISO filters are illustrated in Figs. 2 a), b) and c) and implemented by using CCCDTAs. They provide two CCCDTAs and two grounded capacitors. In addition, the grounded capacitors are suitable for IC implementation since they compensate for the parasitic capacitances at the nodes/terminals of the circuits [4-7]. However, some input-impedances of the proposed filters are not low. They are required the high output-impedance ofcurrent source. The circuits in Figs. 2 a) and b) can be analysed by using equation (1), the current gain transfer function of the proposed filter is demonstrated as shown by the following equation:

$$I_O = \frac{\left\{ \begin{aligned} & \frac{-g_{m1}g_{m2}}{C_1C_2} I_{in1} + s\frac{g_{m2}}{C_2}\left(I_{in2}+I_{in3}\right) \\ & + \left(s^2 + s\frac{2}{C_2R_{n2}} + \frac{g_{m1}g_{m2}}{C_1C_2}\right) I_{in4} \end{aligned} \right\}}{s^2 + s\frac{2}{C_2R_{n2}} + \frac{g_{m1}g_{m2}}{C_1C_2}} . \tag{4}$$

Simultaneously, the proposed MISO filter in Fig. 2 c) shows that the current gain can exhibit transfer function as follows:

$$I_O = \frac{\left\{ \begin{aligned} & \frac{-g_{m1}g_{m2}}{2C_1C_2} I_{in1} + s\frac{g_{m2}}{2C_2}\left(I_{in2}+I_{in3}\right) \\ & + \left(s^2 + s\frac{1}{C_2R_{p2}} + \frac{g_{m1}g_{m2}}{2C_1C_2}\right) I_{in4} \end{aligned} \right\}}{s^2 + s\frac{1}{C_2R_{p2}} + \frac{g_{m1}g_{m2}}{2C_1C_2}} . \tag{5}$$

According, to Eqs. (4) and (5), the current output will be obtained five standard second-order filters, inverting LPF non-inverting HPF, BPF, BRF and APF functions by selection/connection with the input signals. Also, the input can be selecting/connecting to obtain output filter response from all MISO filters are shown in Table 1.

Table 1. Input selections to obtain output filter response from all proposed circuits.

Filter Response	Input Selection			
I_O	I_{in1}	I_{in2}	I_{in3}	I_{in4}
LPF	1	0	0	0
HPF	1	1	0	1
BPF	0	1	0	0
BRF	0	1	0	1
APF	0	1	1	1

With regard to Table 1, the proposed filters have some drawbacks when they are operated in HPF, BPF and APF. They are required to give simultaneious and equal input signals. In this case, the input signals can be fed simultaneously/equally by using the current mirrors circuit.

The pole frequency (ω_p) and quality factor (Q_p) of Figs. 2. a) and b), are shown in Eqs. (6) and (7), respectively.

$$\omega_p = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} , \quad \text{and} \tag{6}$$

$$Q_p = \frac{R_{n2}}{2}\sqrt{\frac{C_2 g_{m1}g_{m2}}{C_1}} . \tag{7}$$

In Fig. 2 c) the ω_p and Q_p can be analyzed as

$$\omega_p = \sqrt{\frac{g_{m1}g_{m2}}{2C_1C_2}} \quad \text{and} \tag{8}$$

$$Q_p = R_{p2}\sqrt{\frac{C_2 g_{m1}g_{m2}}{2C_1}} . \tag{9}$$

Substituting the parasitic resistance $R_{p2}=R_{n2}=V_T/\ 2I_{B3}$, transconductance gain $g_{m1}=I_{B2}/2V_T$ and $g_{m2}=I_{B4}/2V_T$ into Eqs. (6) - (9), and the ω_p and Q_p can be expressed as

$$\omega_p = \frac{1}{2V_T}\sqrt{\frac{I_{B2}I_{B4}}{C_1C_2}} , \quad \text{and} \tag{10}$$

$$Q_p = \frac{1}{8I_{B3}}\sqrt{\frac{C_2 I_{B2}I_{B4}}{C_1}} , \tag{11}$$

Similarly, the ω_p and Q_p in Eq. (8) – (9) can be shown as

$$\omega_p = \frac{1}{2V_T}\sqrt{\frac{I_{B2}I_{B4}}{2C_1C_2}} , \quad \text{and} \tag{12}$$

(13) $$Q_p = \frac{1}{4I_{B3}}\sqrt{\frac{C_2 I_{B2} I_{B4}}{2C_1}}.$$

It is clearly seen that the three current-mode MISO filters have the same ability andthe ω_p and the Q_p can be electronically tuned by adjusting the DC bias current of the CCCDTAs. Moreover, the Q_p can be independently tuned with DC bias current I_{B3} without any effect on the ω_p. Also, all high impedances of current output only use grounded capacitors.

Sensitivities of proposed filters.

The sensitivities of the active and passive elements of MISO filters of the proposed circuits in Figs. 2 a), b) and c) are low as shown in Eqs. (14) and (15)

(14) $$S_{V_T}^{\omega_p} = -1,\ S_{I_{B2},I_{B4}}^{\omega_p} = \frac{1}{2},\ S_{C_1,C_2}^{\omega_p} = -\frac{1}{2},$$

(15) $$S_{I_{B3}}^{Q_p} = -1,\ S_{I_{B2},I_{B4},C_2}^{Q_p} = \frac{1}{2},\ S_{C_1}^{Q_p} = -\frac{1}{2}.$$

It is noticeable that the relative sensitivities of the proposed MISO filters are equal or less than unity in magnitude.

Non-ideal effects

The non-ideal effects of CCCDTAs are decreasing performance of proposed filters. These can be detailed as follow:

a) The voltage and current tracking errors.

The voltage and current tracking errors modify the characteristics of CCCDTAs as

(16) $$\begin{bmatrix} V_p \\ V_n \\ I_z \\ I_x \end{bmatrix} = \begin{bmatrix} R_p & 0 & 0 & 0 \\ 0 & R_n & 0 & 0 \\ \alpha_p & -\alpha_n & 0 & 0 \\ 0 & 0 & 0 & \pm\beta g_m \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ V_x \\ V_z \end{bmatrix}.$$

The tracking error of the current transfer gains from p, n terminals to z terminal are α_p and α_n, respectively. β is the tracking error of the transconductance gains. The parameters of the proposed filters shown in Figs.2. a) and b) have been modified to

(17) $$\omega_p = \frac{1}{2V_T}\sqrt{\frac{\beta_1\beta_2 I_{B2} I_{B4}\alpha_{n1}}{C_1 C_2}},$$

(18) $$Q_p = \frac{1}{(1+\alpha_{n2})4I_{B3}}\sqrt{\frac{\beta_1\beta_2 I_{B2} I_{B4} C_2 \alpha_{n1}}{C_1}}$$

Then, the parameters of proposed filter in Fig. 2 c) become

(19) $$\omega_p = \frac{1}{2V_T}\sqrt{\frac{\beta_1\beta_2 I_{B2} I_{B4}\alpha_{n1}}{2C_1 C_2}},$$ and

(20) $$Q_p = \frac{1}{4I_{B3}}\sqrt{\frac{\beta_1\beta_2 I_{B2} I_{B4} C_2 \alpha_{n1}}{2C_1}}.$$

b) The parasitic elements

These topics explain the effect of the parasitic resistances and parasitic capacitances at high impedance ports which are z port and x port of CCCDTAs. These parasitic capacitances have downgraded the performance of the proposed filters. The ω_p and Q_p of the proposed filters in Figs.2.a) and b) can be obtained, respectively, as

(21) $$\omega_p = \sqrt{\frac{g_{m1}}{C_1' C_2'}\left\{\frac{G_{z1}}{g_{m1}}\left(G_{z2} + 2G_{n2}\right) + g_{m2} G_{x1}' G_{x2}'\right\}},$$

(22) $$Q_p = \frac{1}{2G_{n2}} \frac{\sqrt{\frac{C_2' g_{m1}}{C_1'}\left\{\frac{G_{z1}}{g_{m1}}\left(G_{z2} + 2G_{n2}\right) + g_{m2} G_{x1}' G_{x2}'\right\}}}{\left(\frac{G_{z2} R_{n2}}{2} + \frac{C_2' G_{z1}}{2C_1' G_{n2}} + 1\right)}.$$

Where

$$C_1' = C_1 + C_{z1}, C_2' = C_2 + C_{z2}, G_{n1} = \frac{1}{R_{n1}}, G_{p2} = \frac{1}{R_{p2}},$$

$$G_{n2} = \frac{1}{R_{n2}}, G_{x1} = \frac{1}{R_{x1}}, G_{x2} = \frac{1}{R_{x2}}, G_{z1} = \frac{1}{R_{z1}},$$

$$G_{z2} = \frac{1}{R_{z2}},$$

$$G_{x1}' = \frac{G_{p2}}{G_{x1} + sC_{x1} + G_{p2}},\ G_{x2}' = \frac{G_{n1}}{G_{x2} + sC_{x2} + G_{n1}}.$$

Meanwhile, proposed filter in Fig. 2 c) can be reanalyzed with the ω_p and Q_p as

(23) $$\omega_p = \sqrt{\frac{g_{m1} g_{m2}}{2C_1' C_2}\left(\frac{2G_{z1} G_{p2}}{g_{m1} g_{m2}} + \frac{G_{x1}' G_{x2}'}{\left(sC_{z2} + \frac{G_{z2}}{G_{n2}} + 1\right)}\right)},$$

(24) $$Q_p = \frac{1}{G_{p2}\left(\frac{C_1' + C_2 G_{z1}}{G_{p2}}\right)} \sqrt{\frac{g_{m1} g_{m2} C_1' C_2}{2}\left(\frac{2G_{z1} G_{p2}}{g_{m1} g_{m2}} + \frac{G_{x1}' G_{x2}'}{\left(sC_{z2} + \frac{G_{z2}}{G_{n2}} + 1\right)}\right)}$$

where

$$C_1' = C_1 + C_{z1}, G_{n1} = \frac{1}{R_{n1}}, G_{p2} = \frac{1}{R_{p2}}, G_{n2} = \frac{1}{R_{n2}},$$

$$G_{x1} = \frac{1}{R_{x1}}, G_{x2} = \frac{1}{R_{x2}},\ G_{z1} = \frac{1}{R_{z1}}, G_{z2} = \frac{1}{R_{z2}},$$

$$G_{x1}' = \frac{G_{p2}}{G_{x1} + sC_{x1} + G_{p2}},\ G_{x2}' = \frac{G_{n1}}{G_{x2} + sC_{x2} + G_{n1}}.$$

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3. Simulation Results

To confirm the theoretical analysis of the proposed current-mode MISO filters, the proposed current-mode MISO filter circuit as shown in Fig. 2.a) was chosen as a simulation example. Being simulated through the PSPICE program, the PNP and NPN transistors employed the proposed circuit by using the parameters of the PR200N and NR200N bipolar transistors of ALA400 transistor array from AT&T. The internal construction of the CCCDTA in Fig.3 is used for simulations. The CCCDTAs for the proposed filter were biased with I_{B2}=I_{B4}= 200 μA, I_{B3}= 25 μA and ± 2.5 V. The external grounded capacitors are set as

$C_1 = C_2 = 1\mathrm{nF}$. The first result in Fig.4 depicts the dynamic range of the designed filter when it operates in LPF. The current input I_{in1} was varied from -60 µA to 60 µA.

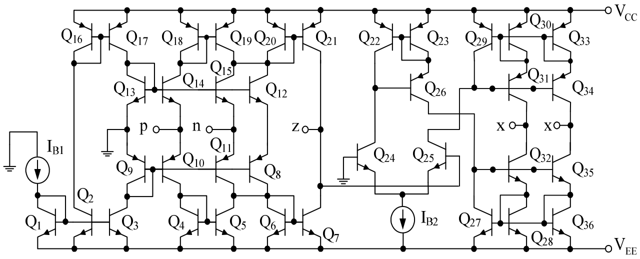


Fig. 3 The internal construction of CCCDTA

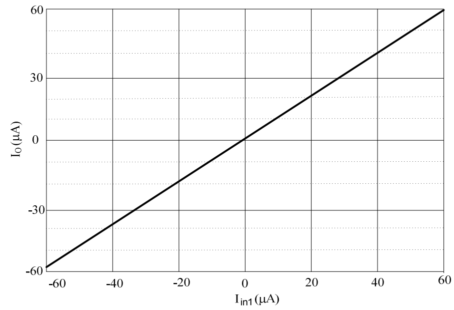
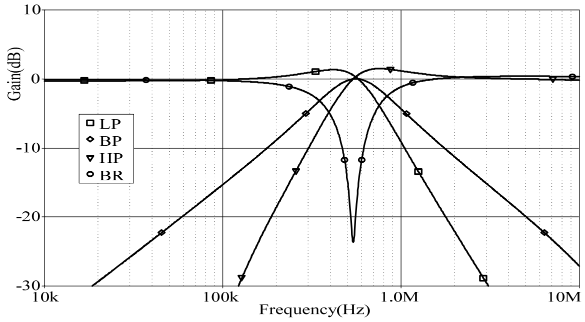


Fig.4. The dynamic range of the design filter.

The results shown in Fig. 5 a) are a simulation of the frequency response (LP, BP, HP and BR) of the proposed filter that gives an input connection as in Table 1. In addition, the simulations of gain and phase responses of the AP filter are displayed in Fig.5 b). Their ω_p is 559.75 kHz. The electronic tuning of *the* ω_p of BP response can be confirmed in Fig. 6, while being kept at a ratio of $I_{B2}=I_{B4}$ and $I_{B2}=8I_{B3}$. To be more practical, the bias currents I_{B2} and I_{B4} can be set equally by using a current mirror circuit. Also, the ratio of bias currents I_{B2} and I_{B3} can be set up with the current-mode analog multiplier/divider circuit which is presented in [33]. It can be seen that the ω_p of BP response varies as 284.44 kHz, 559.75 kHz and 1.08 MHz, respectively, without the influence of Q_p. In this case, the bias currents I_{B2} and I_{B4} are varied as $I_{B2} = I_{B4} =$ 100 µA, 200 µA and 400 µA , respectively. The electronic tuning of Q_p is demonstrated in Fig. 7 with a variable bias current I_{B3}. The Q_p varies as 2, 1 and 0.5 while the bias current I_{B3} varies as 12.5 µA*, 25* µA and 50 µA, respectively. Also Q_p can be freely tuned with non-interactive of the ω_p. The time-domain performance of the BP response is shown in Fig. 8, which is a sinusoidal input current signal of 559.75 kHz with peak-to-peak values of 40 µA. The total harmonic distortion (THD) of the BP response on the various sinusoidal input current signals at a peak-to-peak amplitude of 559.75 kHz are depicted in Fig.9. It is found that the THD value is minimized by 0.25% at a 40 µA peak-to-peak amplitude.

For the implementation of the proposed MISO, the tolerances of passive capacitors caused the deviation of the ω_p. The Monte Carlo analysis can be used to determine the deviations from these tolerances. Fig. 10. shows the results of a Monte Carlo analysis of BP response with 10% tolerances to all the capacitors by using the Gaussian probability distribution and 100 trials. The histograms of the possible spread of ω_p are exhibited in Fig.11. It can be seen that the mean and median of pole frequencies are 574.057 kHz and 574.750 kHz, respectively. The standard deviation is 39.458 kHz. The minimum and maximum of ω_p are 473.346kHz and 657.635kHz, respectively. However, the these deviations can be compensated for or minimized by properly tuning the bias current of the CCCDTAs.

a)



b)

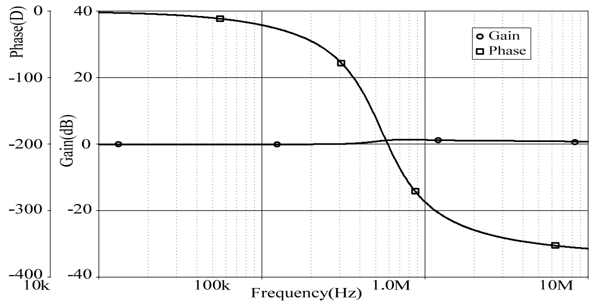


Fig.5. The simulated frequency response of the current-mode MISO filter a) BP, LP, HP and BR ,b) AP

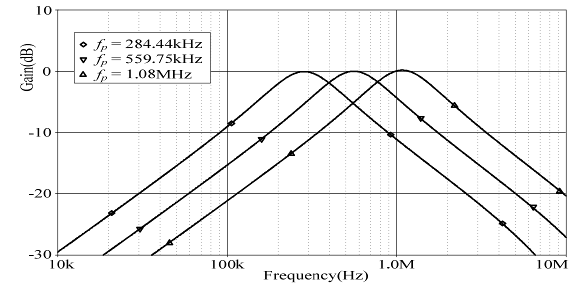


Fig.6. The simulated BP response with different ω_p

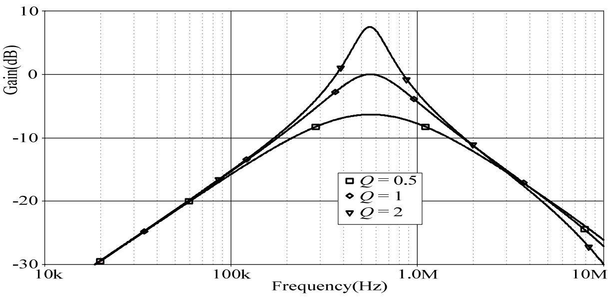


Fig.7. The simulated BP response with different *Q*

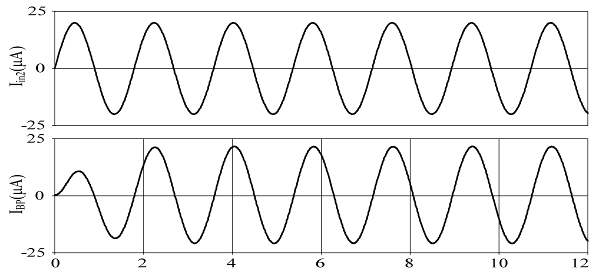


Fig.8. The simulated BP response in time-domain

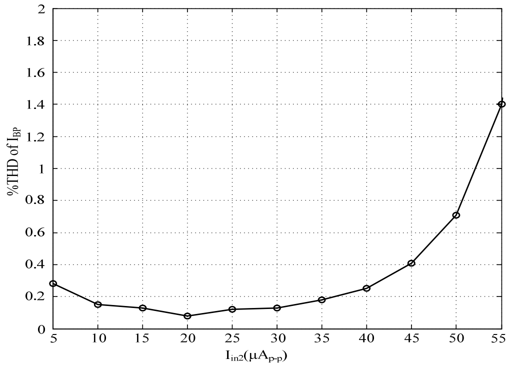


Fig.9. %THD versus peak-to-peak amplitude of the applied sinusoidal signal at 559.75kHz

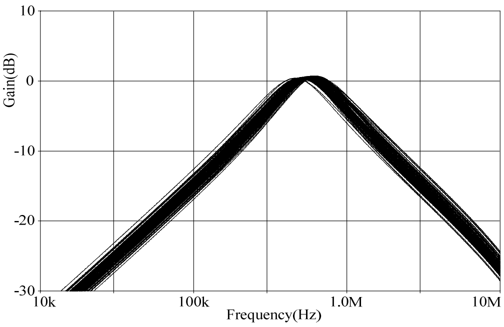


Fig.10. The Gaussian probability variation of the BP response with a Monte Carlo analysis

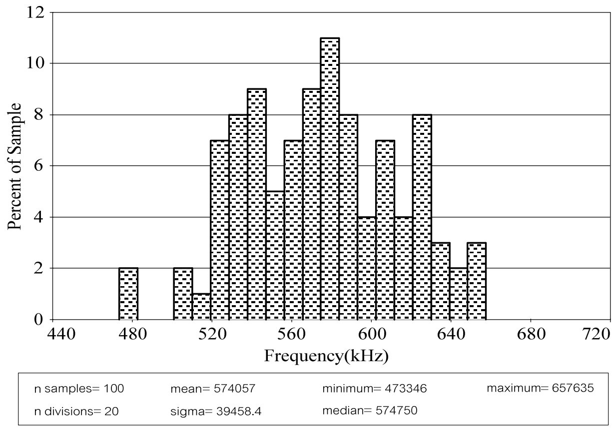


Fig.11. The histograms of the possible spread of ω_p

4. Conclusions

The three circuits of current-mode MISO filters using CCCDTAs were presented; they consist of two CCCDTAs and two grounded capacitors. The advantage of the proposed circuits is that their pole frequency and quality factor is electronically adjustable by input bias currents of the CCCDTAs. The output impedances of a filter is high which is easy to cascade or drive the circuit loads. The proposed filters are easier for IC implementation since the grounded capacitors can be scaled down by the area of IC and compensated by the parasitic capacitance at the node/ports of the circuits. The PSPICE simulation is used to verify its function by comparing it with a theoretical analysis . The simulation results demonstrate function and futures very well.

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