Electronically Tunable Resistorless Mixed Mode Biquad Filters

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Abstract. *This paper presents a new realization of electronically tunable mixed mode (including transadmittance- and voltage-modes) biquad filter with single input, three outputs or three inputs, single output using voltage differencing transconductance amplifier (VDTA), a recently introduced active element. It can simultaneously realize standard filtering signals: low-pass, band-pass and high-pass or by selecting input terminals, it can realize all five different filtering signals: low-pass, band-pass, high-pass, band-stop and all-pass. The proposed filter circuit offers the following attractive feature: no requirement of inverting type input signal which is require no addition circuit, critical component matching conditions are not required in the design, the circuit parameters ω_0 and Q can be set orthogonally or independently through adjusting the bias currents of the VDTAs, the proposed circuit employs two active and minimum numbers of passive components. Furthermore, this filter was investigated from the point of view of limited frequency range, stability conditions, effects of parasitic elements and effects of non-ideal and sensitivity. Thus, taking these effects and conditions into consideration, working conditions and boundaries of this filter are determined. We also performed Monte Carlo, THD and noise analyses. Simulation results are given to confirm theoretical analyses.*

Keywords

Mixed mode filter, electronically tunable filter, biquad filter.

1. Introduction

Current-mode universal active components have two distinct advantages: they provide wide bandwidths and high slew rate. On the other hand, many of today's analog signal processing applications require voltage-mode operation. Universal or multifunction filters exhibit a useful class of filters since they permit realization of different filter functions with the same topology depending on the port used. It provides a versatile, simple and cost effective solution to the integrated circuit manufacturer.

Several transadmittance-mode (TAM) single input, three outputs (SITO) and three inputs, single output (TISO) and voltage-mode (VM) TISO biquad filters based on these newly introduced active elements have been presented in literature [1-10]. Transadmittance-mode SITO filters reported in literature [1-3] employ three active blocks that is CCII, FTFN and FDCCII, two capacitors and using three resistors in [1], [2], three NMOS transistors in [3]. Furthermore, in [1], [2] ω_0 and Q cannot be controlled electronically, also ω_0 and Q cannot be adjusted from each other independently. In ref [4] a transadmittance- and voltage-mode TISO filter has been presented which contains four OTAs and two capacitors. Furthermore, this filter needs a minus input voltage signal to realize AP filter and requires matching condition between transconductance values, also ω_0 and Q cannot be adjusted independently from each other. Shah, Quadri and Iqbal [5] proposed a transadmittance TISO filter using two CDTAs, two capacitors and two resistors. In addition to, this filter circuit has to match between resistors and transconductance value for realizing band-stop (BS) and all-pass (AP) filter. The work by Abuelma'atti [6] reported mixed-mode TISO filter circuit employing four DO-CCCIIs and two capacitors. However this filter circuit cannot realize AP filter. Kumngern, Knobnob and Dejhan [7] proposed a voltage mode TISO biquad filter using six OTAs and two capacitors. Moreover, ω_0 and Q cannot be adjusted from each other independently. The configuration [8] implements voltage mode TISO all second-order functions and employs two DDCCs, two capacitors and two resistors. However, this proposed filter needs a minus input voltage signal to realize AP filter, also ω_0 and Q which cannot be controlled electronically cannot be adjusted independently from each other. Kacar, Yesil and Noori [9] proposed two voltage mode TISO biquad filters. The first proposed filter employs two VDBAs and two capacitors. However, this proposed filter needs a minus input voltage signal to realize AP filter, also ω_0 and Q cannot be adjusted from each other independently. The second proposed filter includes two VDBAs, two capacitors and one resistor. Shah and Malik [10] also proposed a voltage mode TISO biquad filter using FTFN and CTA, two capacitors and resistor. Furthermore, this filter requires matching condition to realize AP filter, also ω_0 cannot be controlled electronically.

A summary of the performance parameters of some recently reported filters in [1-10] and the proposed filter are given in Tab. 1. A careful inspection of Tab. 1 reveals that the proposed filter possess superior features according to recently reported filters in [1-10].

Mode of filters	Ref.	No.of Active Elements	No.of Passive Elements	Matching conditions	Independent tunable Q	Electronically Tunable ω_0	Additional comparison
SITO TRC	[1]in 2001	3 CCII+	3R/2C	No	No	No	-
	[2] in 2004	3 PFTFN	3R/2C	No	No	No	-
	[3]in	3 FDCCII	2C	No	No	Yes	-
	The proposed filter	**2 VDTA**	**2C**	**No**	**Yes**	**Yes**	-
TISO TRC	[4] in 2010	5 OTA	2C	No	No	Yes	-
	[5] in 2007	2 CDTA	2R/2C	Yes	No	Yes	-
	[6] in 2003	4 CCCII±	2C	No	Yes	Yes	This filter cannot realize APF
	The proposed filter	**2 VDTA**	**2C**	**No**	**Yes**	**Yes**	**-**
TISO VM	[7] in 2010	6 OTA	2C	No	No	Yes	-
	[8] in 2004	2 DDCC	2R/2C	No	No	No	The filter needs of the inverting input voltage for APF
	[9]a in 2012	2 VDBA	2C	No	No	Yes	The filter needs of the inverting input voltage for APF
	[9]b in 2012	2VDBA	1R/ 2C	No	Yes	Yes	-
	[10] in 2005	FTFN, CFA	3R/2C	Yes	Yes	No	-
	The proposed filter	**2 VDTA**	**2C**	**No**	**Yes**	**Yes**	**-**

Tab. 1. Comparison of proposed filters with those of previous circuits.

New designs are provided for researchers with these new proposed active elements. In [11] present active elements are reviewed and several new active elements are introduced. One of these newly introduced active elements is voltage differencing transconductance amplifier (VDTA). Yesil, Kacar and Kuntman [12] presented a CMOS implementation and application of voltage mode filter. Previously CMOS implementation of the proposed structure is employed with basic floating current sources (FCS) and its supply voltage is ±0.9 V. Besides, several types of circuits which are voltage mode single input, five outputs filter, current mode SITO filter, current mode sinusoidal oscillator and grounded-floating inductance simulator, are presented [13-16].

This work presents a new realization of electronically tunable mixed mode (including transadmittance- and voltage-modes) biquad filters with single input, three outputs or three inputs, single output using voltage differencing transconductance amplifier (VDTA). This proposed circuit provides both SITO transadmittance-mode and TISO transadmittance- and voltage-mode filter. Furthermore, the topology proposed can generate the standard filter functions (low-pass (LP), band-pass (BP) and high-pass (HP)), all filter functions namely LP, BP, HP, BS and AP without use of the inverting input terminals, respectively. The features of the proposed circuit are that the proposed mixed mode biquad filter employs two VDTAs and two capacitors; no requirement with the component choice conditions to realize specific filtering functions; natural frequency can be adjusted electronically by biasing current and quality factor can be adjusted as independent frequency; it has very low active and passive sensitivities. Furthermore, for this filter, effects of limited frequency, stability conditions and effects of parasitic elements and effects of non-ideal and sensitivity were investigated. Thus, taking these effects and conditions into consideration, working conditions and boundaries of this filter are determined.

2. Proposed Circuit

The circuit symbol of the proposed active element, VDTA, is shown in Fig. 1, where *V*P and *V*N are input terminals and Z_P, Z_N, X+ and X- are output terminals. All terminals exhibit high impedance values. The terminals relations of a VDTA as shown in Fig. 1 can be characterized by

$$\begin{bmatrix} I_{Z_P} \\ I_{Z_N} \\ I_{X+} \\ I_{X-} \end{bmatrix} = \begin{bmatrix} \beta_{ZP}g_{mF} & -\beta_{ZP}g_{mF} & 0 \\ -\beta_{ZN}g_{mF} & \beta_{ZN}g_{mF} & 0 \\ 0 & 0 & \beta_{X+}g_{mS} \\ 0 & 0 & -\beta_{X-}g_{mS} \end{bmatrix} \begin{bmatrix} V_{VP} \\ V_{VN} \\ V_{ZP} \end{bmatrix}. \quad (1)$$

where ideally $\beta_{ZP} = \beta_{ZN} = \beta_{X+} = \beta_{X-} = 1$ which represents the tracking error for the first and second stages of VDTA. g_{mF} and g_{mS} are the first and second transconductance gains of the VDTA, respectively.

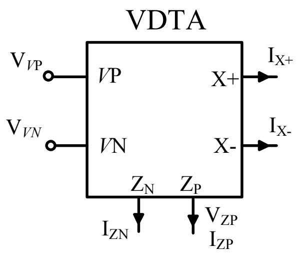

Fig. 1. The circuit symbol of the VDTA.

According to input terminals, an output current at Z_P and Z_N terminals are generated. The intermediate voltage of

Z_P terminal is converted to output currents. Input and output transconductance parameters of VDTA element in the circuit are determined by the transconductance of output transistors. It can be approximated as

$$g_{mF} \cong \frac{g_1 g_2}{g_1 + g_2} + \frac{g_3 g_4}{g_3 + g_4} \;, \tag{2a}$$

$$g_{mS} \cong \frac{g_5 g_6}{g_5 + g_6} + \frac{g_7 g_8}{g_7 + g_7} \tag{2b}$$

where g_i is the transconductance value of the i^{th} transistor defined by

$$g_i = \sqrt{I_{Bi} . \mu_i . C_{OX} \left[\frac{W}{L} \right]_i} \;,$$

μ_i is (i=n, p) the mobility of the carrier for NMOS (n) and PMOS (p) transistors, C_{OX} is the gate-oxide capacitance per unit area, W is the effective channel width, L is the effective channel length and I_{B1} and I_{B2} are the bias currents of the first and second stages of VDTA.

g_{mFJ} and g_{mSJ} are belonging to parameters of the first and second transconductance stages of the J^{th} VDTA. Between transconductance gains (Z_P/Z_N, X+/X-) of the VDTA tracking errors occur due to the fact that upper and lower bias current mirrors cannot be matched and depend on values of output resistance. The first stage and the second stage can simply be implemented by floating current sources (FCS) [17]. Output resistance of the traditional FCS stage is not high enough for some applications. To increase the output impedance improved FCS stage [18] can be used but while the output resistance value is improved with this structure, output voltage swing drops up to V_{DSsat}.

In this paper, basic FCS stages will be used for CMOS realization of VDTA and are shown in Fig. 2 [12]. Differently from CMOS realization of VDTA in previous study [12], both only Z_N terminal has been indicated in Fig. 2 and ideal current sources are replaced with swing cascode current mirror. According to basic current mirrors, the advantages of wide swing cascode current sources are superior to high accuracy and high output resistance. Also, minimum output voltage of wide swing cascode current mirror is $2V_{DSsat}$ [19].

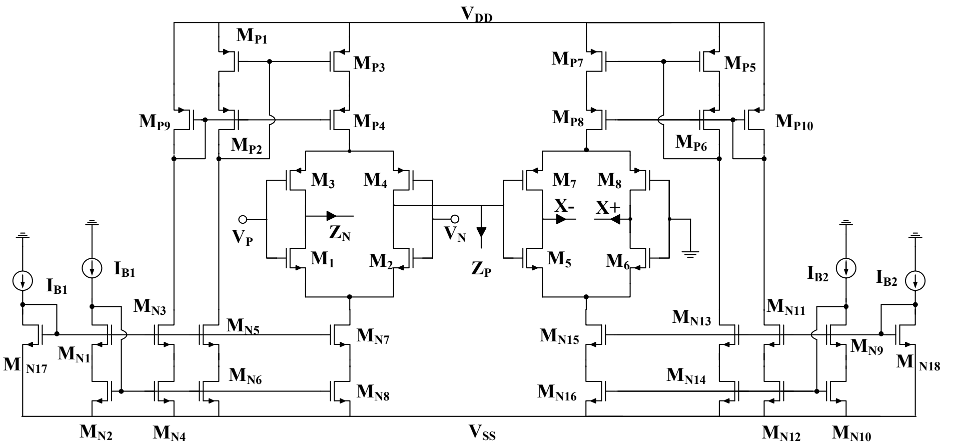

Fig. 2. The CMOS implementation of VDTA.

An application of the VDTA is shown in Fig. 3 where both single input, multi outputs and three inputs, single output filter topologies are provided. Besides, its application obtains two different modes that are voltage- and transadmittance-modes.

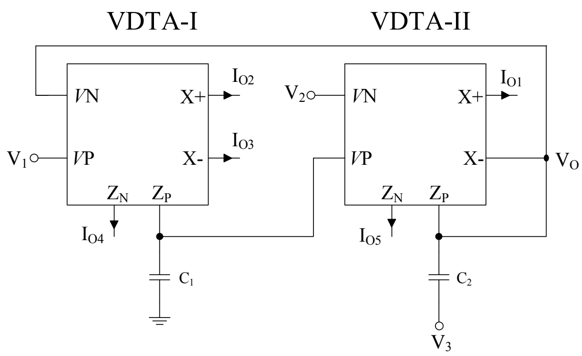

Fig. 3. Application of the proposed VDTA.

The proposed circuit analyses yield for SITO filter topology of transadmittance-mode transfer functions shown in (3) to (6). If $V_2 = V_{IN}$ and $V_1 = V_3 = 0$, then

Non-inverting LP: $$\frac{I_{O2}}{V_{IN}} = \frac{g_{mF1} g_{mS1} g_{mF2} / C_1 C_2}{s^2 + s g_{mS2}/C_2 + g_{mF1} g_{mF2}/C_1 C_2} \,. \tag{3}$$

Inverting LP: $$\frac{I_{O3}}{V_{IN}} = -\frac{g_{mF1} g_{mS1} g_{mF2} / C_1 C_2}{s^2 + s g_{mS2}/C_2 + g_{mF1} g_{mF2}/C_1 C_2} \,. \tag{4}$$

Inverting BP: $$\frac{I_{O4}}{V_{IN}} = -\frac{s g_{mF1} g_{mF2} / C_2}{s^2 + s g_{mS2}/C_2 + g_{mF1} g_{mF2}/C_1 C_2} \,. \tag{5}$$

Non-inverting HP: $$\frac{I_{O5} + I_{O1}}{V_{IN}} = \frac{s^2 g_{mF1}}{s^2 + s g_{mS2}/C_2 + g_{mF1} g_{mF2}/C_1 C_2} \,. \tag{6}$$

After above filter outputs are obtained simultaneously, BS and AP can be obtained with combining of definite currents, that is, $I_{O5}+I_{O1}+I_{O2}$ for output of BS filter, $I_{O5}+I_{O1}+I_{O2}+I_{O4}$ for output of AP filter.

TISO filter topology of the proposed circuit analyses yield transadmittance-mode transfer function shown in (7) and voltage-mode transfer function shown in (8).

$$I_{O1} = \frac{V_3 s^2 g_{mS2} - V_2 s g_{mF2} g_{mS2}/C_2 + V_1 g_{mF1} g_{mF2} g_{mS2}/C_1 C_2}{s^2 + s g_{mS2}/C_2 + g_{mF1} g_{mF2}/C_1 C_2}, \quad (7)$$

$$V_O = \frac{V_3 s^2 - V_2 s g_{mF2}/C_2 + V_1 g_{mF1} g_{mF2}/C_1 C_2}{s^2 + s g_{mS2}/C_2 + g_{mF1} g_{mF2}/C_1 C_2}. \quad (8)$$

Depending on the voltage status of V_1, V_2 and V_3 in the numerator of (7) and (8) one of the following five filter functions is realized:

(i) LP: $V_1 = V_{IN}$, $V_2 = V_3 = 0$,

(ii) BP: $V_2 = V_{IN}$, $V_1 = V_3 = 0$,

(iii) HP: $V_3 = V_{IN}$, $V_1 = V_2 = 0$,

(iv) BS: $V_1 = V_3 = V_{IN}$, $V_2 = 0$,

(v) AP: $V_1 = V_2 = V_3 = V_{IN}$.

The natural frequency (ω_0) and quality factor (Q) can be given as follows;

$$\omega_0 = \sqrt{\frac{g_{mF1} g_{mF2}}{C_1 C_2}}, \quad (9)$$

$$Q = \frac{1}{g_{mS2}} \sqrt{\frac{C_2 g_{mF1} g_{mF2}}{C_1}}, \quad (10)$$

It is obviously found that, from (9) and (10) the quality factor can be adjusted by value of g_{mS2} without affecting the natural frequency. In addition, from (9) it should be remarked that the natural frequency can be electronically adjusted by bias currents (I_B).

3. Different Analyses of the Proposed Filter

3.1 Effects of the Limited Frequency and Stability Analyses

The limited bandwidth of transconductance gains of the VDTA may affect the stability and working conditions of the proposed circuit. Thus, we apply Routh-Hurwitz stability criteria to examine effect of frequency dependent transconductance gains. We assume that transconductance gains have only single corner frequencies that are denoted as ω_{pi} (i = F1,F2,S1,S2). It is well known that transconductance values of the VDTA are a frequency dependent parameter and can be expressed with one pole model as shown below [20], [21]

$$g_{mi}(s) = \frac{g_{m0i}}{1+\frac{s}{\omega_{Pi}}} = g_{m0i}\frac{\omega_{Pi}}{\omega_{Pi}+s} \cong g_{m0i}(1 - s\tau_i), \quad \omega\tau << 1 \quad (11)$$

where g_{m0ki} is the zero frequency transconductance gain, ω_{pki} is the parasitic pole frequency and $\tau_{ki} = 1/\omega_{pki}$ is parasitic delay. k is belonging to corresponding transconductance stage names of i^{th} VDTA which is F and S.

The proposed filter is shown in Fig. 3. Routine analyses of the proposed filter have given transfer functions in (3)-(6). Using (11) the reanalysis of SITO filter topology of transadmittance-mode transfer functions yields,

$$\frac{I_{O2}}{V_{IN}} = \frac{g_{m0F1} g_{m0F2} g_{m0S1}(1 - s\tau_{F1})(1 - s\tau_{F2})(1 - s\tau_{S1})/C_1 C_2}{\Delta s}, \quad (12)$$

$$\frac{I_{O3}}{V_{IN}} = -\frac{g_{m0F1} g_{m0F2} g_{m0S1}(1 - s\tau_{F1})(1 - s\tau_{F2})(1 - s\tau_{S1})/C_1 C_2}{\Delta s}, \quad (13)$$

$$\frac{I_{O4}}{V_{IN}} = -\frac{s g_{m0F1} g_{m0F2}(1 - s\tau_{F1})(1 - s\tau_{F2})/C_2}{\Delta s}, \quad (14)$$

$$\frac{I_{O5} + I_{O1}}{V_{IN}} = \frac{s^2 g_{m0F2}(1 - s\tau_{F2})}{\Delta s}. \quad (15)$$

And again, using (11) the reanalysis of TISO filter topology of transadmittance- and voltage-mode transfer functions yields (16)-(17),

$$I_{O1} = \frac{V_3 s^2 g_{m0S2}(1 - s\tau_{S2})}{\Delta s} - \frac{V_2 s g_{m0F2} g_{m0S2}(1 - s\tau_{F2})(1 - s\tau_{S2})/C_2}{\Delta s} + \frac{V_1 g_{m0F1} g_{m0F2} g_{m0S2}(1 - s\tau_{F1})(1 - s\tau_{F2})(1 - s\tau_{S2})/C_1 C_2}{\Delta s}, \quad (16)$$

$$V_O = \frac{V_3 s^2}{\Delta s} + \frac{V_2 s g_{m0F2}(1 - s\tau_{F2})/C_2}{\Delta s} + \frac{V_1 g_{m0F1} g_{m0F2}(1 - s\tau_{F1})(1 - s\tau_{F2})/C_1 C_2}{\Delta s} \quad (17)$$

where Δs is the characteristic equation and is given by

$$\Delta s = s^2\left(1 - \frac{g_{m0S2}\tau_{S2}}{C_2} + \frac{g_{m0F1} g_{m0F2}\tau_{F1}\tau_{F2}}{C_1 C_2}\right) + s\left(\frac{g_{m0S2}}{C_2} - \frac{g_{m0F1} g_{m0F2}}{C_1 C_2}(\tau_{F1} + \tau_{F2})\right) + \frac{g_{m0F1} g_{m0F2}}{C_1 C_2}. \quad (18)$$

In this case, the modified natural frequency (ω_0') and quality factor (Q') can be given as follows

$$\omega_0' = \sqrt{\frac{\frac{g_{m0F1} g_{m0F2}}{C_1 C_2}}{1 + \frac{g_{m0F1} g_{m0F2}\tau_{F1}\tau_{F2}}{C_1 C_2} - \frac{g_{m0S2}\tau_{S2}}{C_2}}}, \quad (19)$$

$$Q' = \frac{\sqrt{\frac{g_{m0F1} g_{m0F2}}{C_1 C_2} \Big/ \left(1 + \frac{g_{m0F1} g_{m0F2}\tau_{F1}\tau_{F2}}{C_1 C_2} - \frac{g_{m0S2}\tau_{S2}}{C_2}\right)}}{\frac{g_{m0S2}}{C_2} - \frac{g_{m0F1} g_{m0F2}}{C_1 C_2}(\tau_{F1} + \tau_{F2})}. \quad (20)$$

From (18), it can be seen that due to the limited bandwidth of transconductance gains of VDTA, stability problem of the proposed filter is found. In order to operate the circuit as a stable filter the following conditions should be satisfied,

$$\frac{g_{\mathrm{m0S2}}\tau_{\mathrm{S2}}}{C_2} << 1 + \frac{g_{\mathrm{m0F1}}g_{\mathrm{m0F2}}\tau_{\mathrm{F1}}\tau_{\mathrm{F2}}}{C_1 C_2}\,, \tag{21}$$

$$\frac{g_{\mathrm{m0F1}}g_{\mathrm{m0F2}}\left(\tau_{\mathrm{F1}}+\tau_{\mathrm{F2}}\right)}{C_1} << g_{\mathrm{m0S2}}\,. \tag{22}$$

The modified natural frequency and quality factor is a little affected when in (21)-(22) conditions are satisfied. Therefore, the effect of the parasitic delay is negligible.

3.2 Effects of Parasitic Elements Analyses

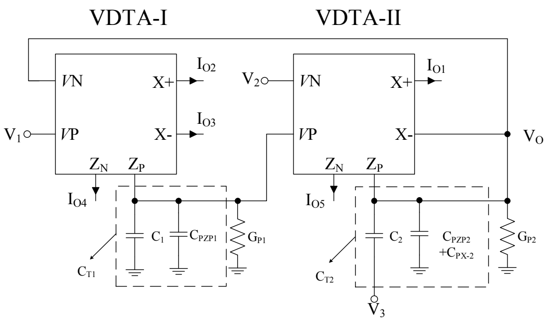

Fig. 4. Parasitic impedances of VDTA affecting filter of Fig. 3.

In this section, the network model given in Fig. 4 is adopted for studying parasitic properties. For example, the proposed filter is taken into consideration with parasitic elements. Effects on quality factor and natural frequency of the proposed circuit are investigated. According to the model given in Fig. 4, SITO filter topology of transadmittance-mode transfer functions can be recalculated,

$$\frac{I_{O2}}{V_{IN}} = \frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}g_{\mathrm{mS1}}/C_{\mathrm{T1}}C_{\mathrm{T2}}}{\Delta sp}\,, \tag{23}$$

$$\frac{I_{O3}}{V_{IN}} = -\frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}g_{\mathrm{mS1}}/C_{\mathrm{T1}}C_{\mathrm{T2}}}{\Delta sp}\,, \tag{24}$$

$$\frac{I_{O4}}{V_{IN}} = -\frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}\left(sC_{\mathrm{T1}}+G_{\mathrm{P1}}\right)/C_{\mathrm{T1}}C_{\mathrm{T2}}}{\Delta sp}\,, \tag{25}$$

$$\frac{I_{O5}+I_{O1}}{V_{IN}} = \frac{s^2 g_{\mathrm{mF2}} + s g_{\mathrm{mF2}}\left(G_{\mathrm{P1}}/C_{\mathrm{T1}}+G_{\mathrm{P2}}/C_{\mathrm{T2}}\right) + g_{\mathrm{mF2}}G_{\mathrm{P1}}G_{\mathrm{P2}}/C_{\mathrm{T1}}C_{\mathrm{T2}}}{\Delta sp}\,. \tag{26}$$

As analog procedures are applied for TISO filter topology of transadmittance- and voltage-mode transfer functions

$$I_{O1} = \frac{\mathrm{V}_3 g_{\mathrm{mS2}}\left(s^2 C_2/C_{\mathrm{T2}} + sC_2 G_{\mathrm{P1}}/C_{\mathrm{T1}}C_{\mathrm{T2}}\right)}{\Delta sp} - \frac{\mathrm{V}_2 g_{\mathrm{mS2}}g_{\mathrm{mF2}}\left(s/C_{\mathrm{T2}} + G_{\mathrm{P1}}/C_{\mathrm{T1}}C_{\mathrm{T2}}\right)}{\Delta sp} + \frac{\mathrm{V}_1 g_{\mathrm{mS2}}\, g_{\mathrm{mF1}}g_{\mathrm{mF2}}/C_{\mathrm{T1}}C_{\mathrm{T2}}}{\Delta sp}\,, \tag{27}$$

$$V_O = \frac{\mathrm{V}_3\left(s^2 C_2/C_{\mathrm{T2}} + sC_2 G_{\mathrm{P1}}/C_{\mathrm{T1}}C_{\mathrm{T2}}\right)}{\Delta sp} - \frac{\mathrm{V}_2 g_{\mathrm{mF2}}\left(s/C_{\mathrm{T2}} + G_{\mathrm{P1}}/C_{\mathrm{T1}}C_{\mathrm{T2}}\right)}{\Delta sp} + \frac{\mathrm{V}_1\, g_{\mathrm{mF1}}g_{\mathrm{mF2}}/C_{\mathrm{T1}}C_{\mathrm{T2}}}{\Delta sp} \tag{28}$$

where the denominator (Δsp) can be expressed as

$$\Delta sp = s^2 + s\left(\frac{g_{\mathrm{mS2}}}{C_{\mathrm{T2}}} + \frac{G_{\mathrm{P2}}}{C_{\mathrm{T2}}} + \frac{G_{\mathrm{P1}}}{C_{\mathrm{T1}}}\right) + \frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}}{C_{\mathrm{T1}}C_{\mathrm{T2}}} + \frac{G_{\mathrm{P1}}\left(g_{\mathrm{mS2}}+G_{\mathrm{P2}}\right)}{C_{\mathrm{T1}}C_{\mathrm{T2}}} \tag{29}$$

and total capacitances and total conductances are denoted analogously as $C_{T1} = C_1 + C_{PZP1}$, $C_{T2} = C_2 + C_{PZP2} + C_{PX-2}$, $G_{P1} = G_{PZP1}$ and $G_{P2} = G_{PZP2} + G_{PX-2}$. C_{PZPi}, $C_{PX\text{-}2i}$ and G_{Pi} are parasitic capacitances and conductance of i^{th} VDTA elements, respectively. From (29), the modified natural frequency ($\omega_0^{"}$) and quality factor ($Q^{"}$) can be calculated as follows,

$$\omega_0^{"} = \sqrt{\frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}}{C_{\mathrm{T1}}C_{\mathrm{T2}}} + \frac{G_{\mathrm{P1}}\left(g_{\mathrm{mS2}}+G_{\mathrm{P2}}\right)}{C_{\mathrm{T1}}C_{\mathrm{T2}}}}\,, \tag{30}$$

$$Q^{"} = \frac{\sqrt{\dfrac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}}{C_{\mathrm{T1}}C_{\mathrm{T2}}} + \dfrac{G_{\mathrm{P1}}\left(g_{\mathrm{mS2}}+G_{\mathrm{P2}}\right)}{C_{\mathrm{T1}}C_{\mathrm{T2}}}}}{\dfrac{g_{\mathrm{mS2}}}{C_{\mathrm{T2}}} + \dfrac{G_{\mathrm{P2}}}{C_{\mathrm{T2}}} + \dfrac{G_{\mathrm{P1}}}{C_{\mathrm{T1}}}}\,. \tag{31}$$

From (30) and (31) it can be seen that effective values of conductances and capacitances are increased by parasitic impedances as a result $\omega_0^{"}$ and $Q^{"}$ decrease according to ω_0 and Q at (9)-(10). The parasitic effects on the natural frequency and quality factor can be avoided by choosing

$$C_1 >> C_{PZP1}\,, \tag{32}$$

$$C_2 >> C_{PZP2} + C_{PX-2}\,. \tag{33}$$

3.3 Effects of Non-ideal and Sensitivity Analyses

Taking the tracking errors of the VDTA into account in (1), the proposed circuit analyses can be recalculated for SITO filter topology of transadmittance-mode transfer functions and TISO filter topology of transadmittance- and voltage mode transfer functions shown in (34) to (39), respectively,

$$\frac{I_{O2}}{V_{IN}} = \frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}g_{\mathrm{mS1}}\beta_{\mathrm{ZP1}}\beta_{\mathrm{ZP2}}\beta_{\mathrm{P1}}/C_1C_2}{\Delta sn}\,, \tag{34}$$

$$\frac{I_{O3}}{V_{IN}} = -\frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}g_{\mathrm{mS1}}\beta_{\mathrm{ZP1}}\beta_{\mathrm{ZP2}}\beta_{\mathrm{N1}}/C_1C_2}{\Delta sn}\,, \tag{35}$$

$$\frac{I_{O4}}{V_{IN}} = -\frac{s g_{\mathrm{mF1}}g_{\mathrm{mF2}}\beta_{\mathrm{ZN1}}\beta_{\mathrm{ZP2}}/C_2}{\Delta sn}\,, \tag{36}$$

$$\frac{I_{O5}+I_{O1}}{V_{IN}} = \frac{s^2 g_{\mathrm{mF2}}\beta_{\mathrm{ZN2}} + s g_{\mathrm{mF2}}g_{\mathrm{mS2}}\left(\beta_{\mathrm{ZN2}}\beta_{\mathrm{N2}} - \beta_{\mathrm{ZP2}}\beta_{\mathrm{P2}}\right)/C_2}{\Delta sn}\,, \tag{37}$$

$$I_{O1} = g_{\mathrm{mS2}}\frac{\mathrm{V}_3 s^2\beta_{\mathrm{P2}} - \mathrm{V}_2\, s g_{\mathrm{mF2}}\beta_{\mathrm{ZP2}}\beta_{\mathrm{P2}}/C_2 + \mathrm{V}_1\, g_{\mathrm{mF1}}g_{\mathrm{mF2}}\beta_{\mathrm{ZP1}}\beta_{\mathrm{ZP2}}\beta_{\mathrm{P2}}/C_1C_2}{\Delta sn}\,, \tag{38}$$

$$V_O = \frac{\mathrm{V}_3 s^2 - \mathrm{V}_2\, s g_{\mathrm{mF2}}\beta_{\mathrm{ZP2}}/C_2 + \mathrm{V}_1\, g_{\mathrm{mF1}}g_{\mathrm{mF2}}\beta_{\mathrm{ZP1}}\beta_{\mathrm{ZP2}}/C_1C_2}{\Delta sn} \tag{39}$$

where Δsn is the characteristic equation and is given by

$$\Delta sn = s^2 + \frac{s g_{\mathrm{mS2}}\beta_{\mathrm{N2}}}{C_2} + \frac{g_{\mathrm{mF1}}g_{\mathrm{mF2}}\beta_{\mathrm{ZP1}}\beta_{\mathrm{ZP2}}}{C_1C_2}\,. \tag{40}$$

From (37), it is seen that transfer function is affected by non-ideal parameters. To decrease this influence, tranconductance gains of between Z_P and Z_N terminals should be necessarily equal, similarly between X+ and X- terminals, too. This equality is obtained by current mirror that has high output resistance. According to the denominator of the non-ideal model, the modified natural frequency and quality factor are obtained by

$$\omega_0''' = \sqrt{\frac{g_{mF1} g_{mF2} \beta_{ZP1} \beta_{ZP2}}{C_1 C_2}}, \tag{41}$$

$$Q''' = \frac{1}{g_{mS2}\beta_{N2}} \sqrt{\frac{g_{mF1} g_{mF2} C_2 \beta_{ZP1} \beta_{ZP2}}{C_1}}. \tag{42}$$

Evaluation of (41)-(42) shows that the transconductance inaccuracy factor β slightly affects the value of the modified natural frequency and quality factor. By using (41)-(42), the passive and active sensitivities of ω_0''' and Q''' can be derived in (43)-(44)

$$\begin{aligned} &S^{\omega_0}_{g_{mF1}} = S^{\omega_0}_{g_{mF2}} = -S^{\omega_0}_{C_1} = -S^{\omega_0}_{C_2} = S^{\omega_0}_{\beta_{ZP1}} = S^{\omega_0}_{\beta_{ZP2}} = \frac{1}{2} \\ &S^{\omega_0}_{g_{mS1}, g_{mS2}, \beta_{P1}, \beta_{P2}, \beta_{N1}, \beta_{N2}, \beta_{ZN1}, \beta_{ZN2}} = 0 \end{aligned} \tag{43}$$

$$\begin{aligned} &S^{Q}_{g_{mF1}} = S^{Q}_{g_{mF2}} = S^{Q}_{C_2} = -S^{Q}_{C_1} = S^{Q}_{\beta_{ZP1}} = S^{Q}_{\beta_{ZP2}} = \frac{1}{2} \\ &S^{Q}_{g_{mS2}} = S^{Q}_{\beta_{N2}} = -1, \; S^{Q}_{g_{mS1}, \beta_{ZN1}, \beta_{ZN2}, \beta_{P1}, \beta_{P2}, \beta_{N1}} = 0 \end{aligned} \tag{44}$$

It can be seen that regarding the sensitivities of the natural frequency and quality factor all of the passive and active variables are very low and less than or equal to unity in magnitude.

4. Simulation Results

We perform simulations by using SPICE program with TSMC CMOS 0.18 µm process [22]. The circuit in Fig. 3 is simulated by using the CMOS VDTA in Fig. 2 which is the similar as in [12]. Only Z_N terminal has been indicated in Fig. 2 and ideal current sources are replaced wide swing cascode current mirror. The dimensions of the MOS transistor for the VDTA implementation are given in Tab. 2. Supply voltages and bias currents are taken as $V_{DD} = -V_{SS} = 1.5$ V and $I_{B1,2} = 40$ µA. Simulation results show that this choice yields the transconductance values of VDTA as $g_{mF} \approx g_{mS} \approx 150$ µA/V.

Transistors	W(µm)	L(µm)
M_1, M_2, M_5, M_6	0.72	0.36
M_3, M_4, M_7, M_8	3.6	0.36
M_{N1}-M_{N16}	5.04	0.9
M_{N17}, M_{N18}	1.26	0.9
M_{P1}-M_{P8}	25.2	0.9
M_{P9}, M_{P10}	6.3	0.9

Tab. 2. Transistors aspect ratio of VDTA.

The proposed SITO transadmittance-mode, TISO transadmittance- and voltage-modes biquad filters in Fig. 3 was designed for 1 MHz center frequency and quality factor of $Q = 1$ by choosing $C_1 = C_2 = 23.5$ pF and taken $I_{B1,2} = 40$ µA. Fig. 5 presents the simulated responses of SITO transadmittance-mode biquad filters.

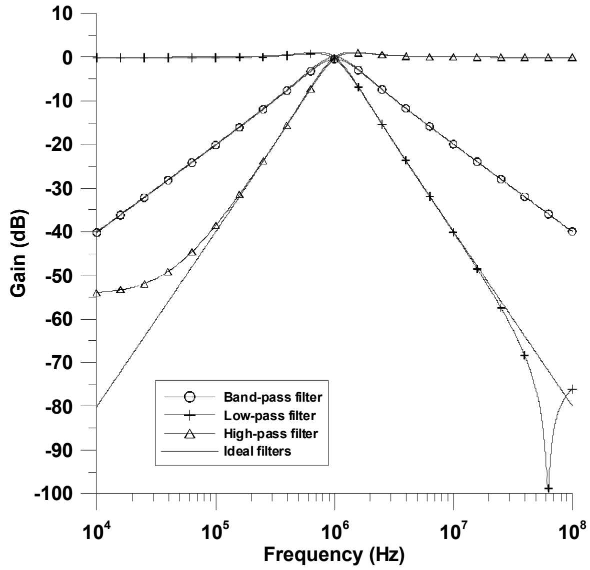

Fig. 5. Gain-frequency responses of normalized SITO transadmittance-mode LP, BP and HP filters.

Frequency responses of TISO transadmittance- and voltage-mode low-pass, band-pass, high-pass and band-stop filters are shown in Fig. 6 and Fig. 8, respectively. Fig. 7 and Fig. 9 represent the simulated all-pass phase–frequency and amplitude–frequency responses of the proposed TISO transadmittance- and voltage-mode filters, respectively.

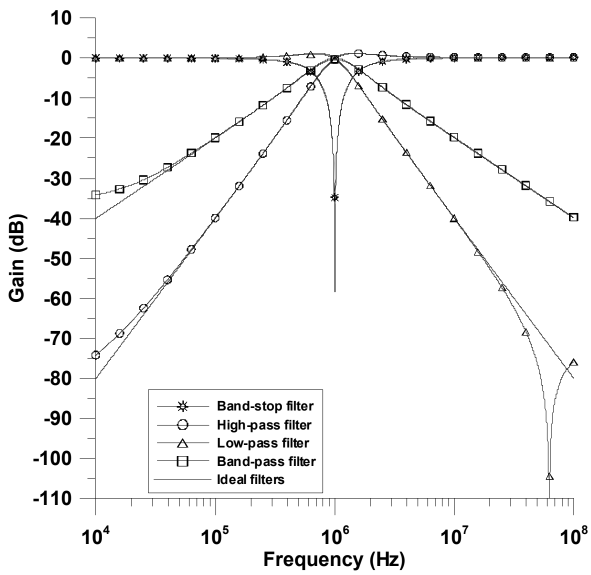

Fig. 6. Gain-frequency responses of normalized TISO transadmittance-mode LP, BP, HP and BS filters.

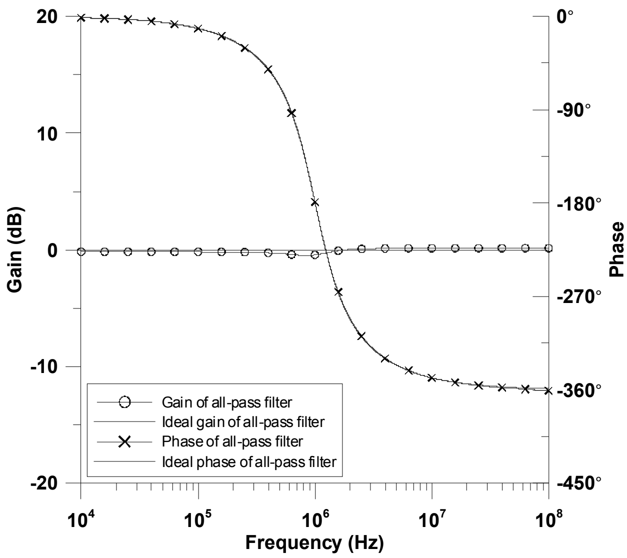

Fig. 7. Gain-frequency responses of normalized TISO transadmittance-mode AP filter.

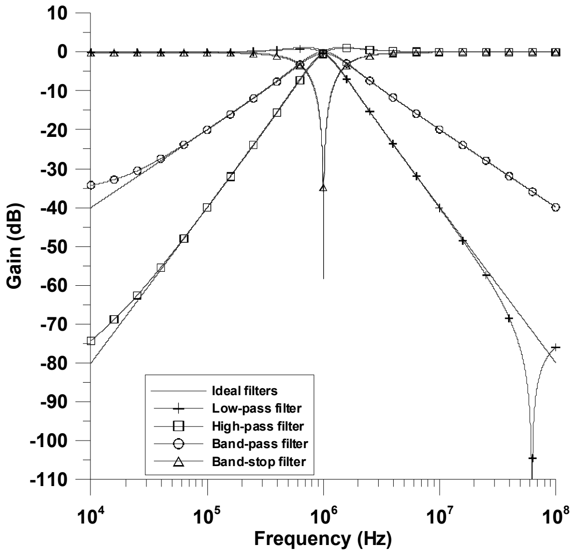

Fig. 8. Gain-frequency responses of TISO voltage-mode LP, BP, HP and BS filters.

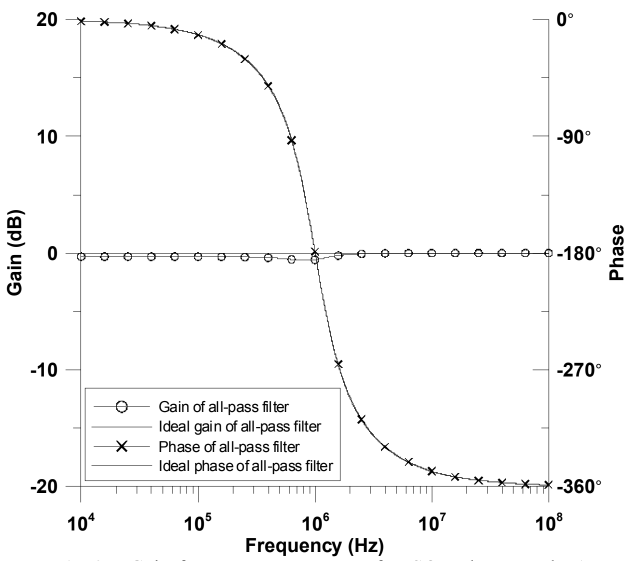

Fig. 9. Gain-frequency responses of TISO voltage-mode AP filter.

The electronic tunability of frequency response is provided by changing all the biased currents from 10 μA to 80 μA which is shown in Fig. 10. The electronic tunability of quality factor is provided by assigning $I_{B11} = I_{B21} = I_{B12} = 40$ μA and varying the biased current I_{B22} from 5 μA to 60 μA which is shown in Fig. 11 (I_{B1i} and I_{B2i} are belonging to bias current of ith VDTA element).

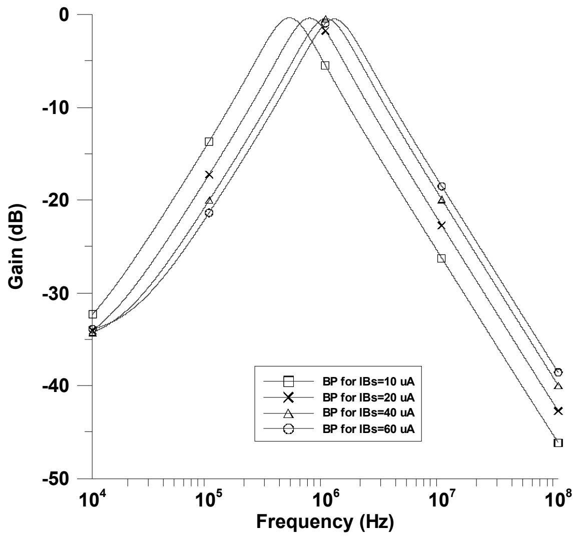

Fig. 10. Electronic tunability of frequency responses of TISO voltage mode BP filter.

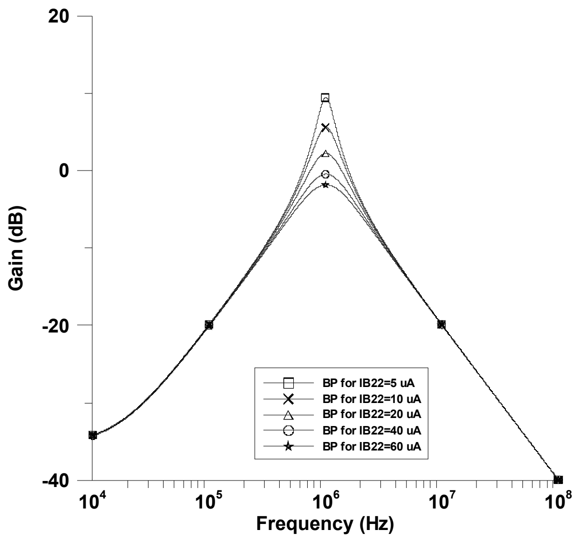

Fig. 11. Electronic tunability of quality factor of TISO voltage mode BP filter.

The behavior of internal structure of VDTA in Fig. 2 with respect to width (W) and length (L) of transistors M_1-M_8 mismatching and process parameters V_{T0} and t_{OX} have been evaluated through statistical analysis results by utilizing the well-known Monte Carlo analysis. Monte Carlo simulation is performed five hundred times for the TISO voltage-mode BP filter in Fig. 3. The Monte Carlo analyses with 5% Gauss deviation of t_{OX} and relevant transistors' width and length and 10% Gauss deviation of V_{T0} are achieved as in Fig. 12 where center frequency and quality factor variations of the TISO voltage-mode BP filter in Fig. 3 are given. According to Monte Carlo simulations, the standard deviations of the center frequency and quality factor are only 0.66% and 0.88%, respectively. It is shown that the proposed filter has low sensitivity to the mismatch between input transistors at FCS stage.

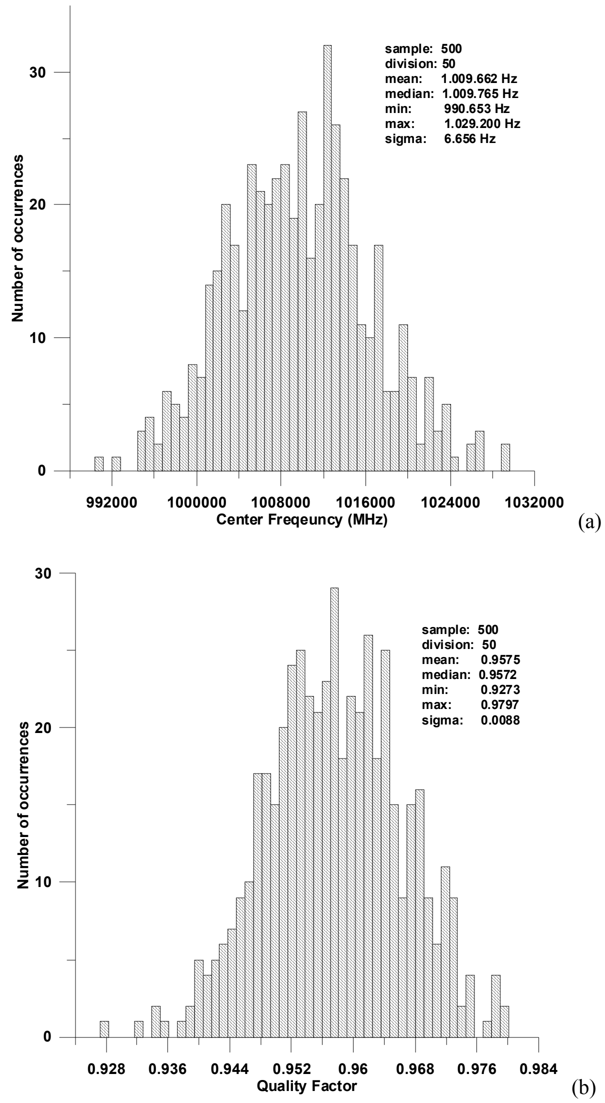

Fig. 12. The histograms of (a) center frequency and (b) quality factor distribution of the TISO voltage-mode BP filter (500 times Monte Carlo Simulation).

The THD result for the voltage-mode BP filter for Fig. 3 is given in Fig. 13 which clearly shows that for an input signal lower than peak-to-peak 430 mV, the THD remains in acceptable limits thus confirming the practical utility of the proposed circuit i. e. 3%. As can be seen, there is very good agreement between theory and simulations.

Output noise behaviors for SITO and TISO transconductance- and TISO voltage- mode BP filters of the proposed filter with respect to frequency have also been simulated, as it is shown in Fig. 14. For these filters, the equivalent output noises at operating frequency $f_0 = 1$ MHz are found as 4.97 pA√Hz, 5.4 pA√Hz and 28.4 nV√Hz, respectively.

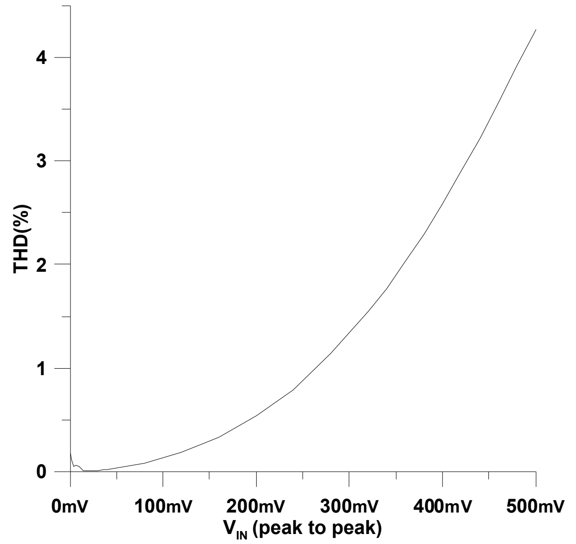

Fig. 13. Total harmonic distortion (THD) of the voltage-mode BP filter for an input signal at 1 MHz.

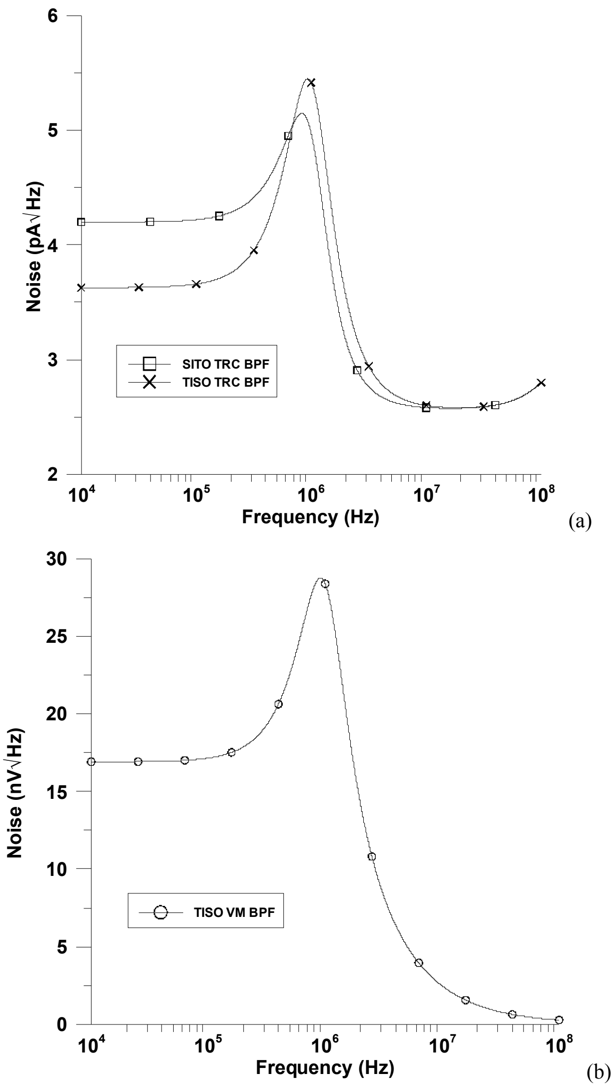

Fig. 14. Output referred (a) noise responses of SITO and TISO transconductance-mode BP filters, (b) noise response of TISO voltage-mode BP filter.

5. Conclusion

In this paper, a new mixed mode biquad filter configuration employing two VDTAs and two capacitors has been presented. The filter topology can generate the standard filter functions and also, by selecting input terminal, the proposed filter can generate all five different voltage- and transadmittance-mode filters. It possesses the following properties: (i) ability of realizing the LP, BP, HP, BS and AP filter responses without any component matching condition, (ii) very low active and passive sensitivities, (iii) no requirement of inverting type input signal which requires no addition circuit, (iv) the circuit parameters ω_0 and Q can be set orthogonally or independently through adjusting the bias currents of the VDTAs. Moreover, for this filter effects of limited frequency, stability conditions and effects of parasitic elements and effects of non-ideal and sensitivity were investigated. Thus, taking these effects and conditions into consideration, working conditions and boundaries of this filter are determined. Taking into account deviations of process parameters and dimensions of MOS transistors, performances of the proposed filter also have been investigated by the Monte Carlo analysis. Simlation results are given to demonstrate the effectiveness of the proposed filter.

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References

[1] TOKER, A., CICEKOGLU, O., OZCAN S., KUNTMAN, H. High-output-impedance transadmittance type continuous-time multifunction filter with minimum active elements. *Int. J. Electronics,* 2001, vol. 88, no. 10, p. 1085-1091.

[2] SHAH, N. A., IQBAL, S. Z., PARVEEN, B., SITO high output impedance transadmittance filter using FTFNs. *Analog Integrated Circuits and Signal Processing,* 2004, vol. 40, p. 87–89.

[3] EL-ADAWY, A., SOLIMAN, A. M., ELWAN, H. O. A novel fully differential current conveyor and applications for analog VLSI. *IEEE Transactions on Circuits and Systems–II: Analog and Signal Digital Processing,* 2000, vol. 47, no. 4, p. 306-313.

[4] LEE, C. N. Multiple-mode OTA-C universal biquad filters. *Circuits Syst Signal Process*, 2010, vol. 29, p. 263–274.

[5] SHAH, N. A., QUADRI, M., IQBAL, S. Z. CDTA based universal transadmittance filter. *Analog Integr Circ Sig Process*, 2007, vol. 52, p. 65–69.

[6] ABUELMA'ATTI, M. T. A novel mixed-mode current-controlled current-conveyor-based filter. *Active and Passive Elec. Comp.,* 2003, vol. 26, p. 185–191.

[7] KUMNGERN, M., KNOBNOB, B., DEJHAN, K. Electronically tunable high-input impedance voltage-mode universal biquadratic filter based on simple CMOS OTAs. *Int. J. Electron. Commun. (AEU)*, 2010, vol. 64, p. 934–939.

[8] TEMIZYUREK, C., MYDERRIZI, I. A novel three-input one-output voltage-mode universal filter using differential difference current conveyor (DDCC). In *Proceedings of the 12th IEEE Mediterranean Electrotechnical Conference (MELECON 2004).* 2004, vol. 1, p. 103–106.

[9] KAÇAR, F., YEŞIL, A., NOORI, A. New CMOS realization of voltage differencing buffered amplifier and its biquad filter applications. *Radioengineering,* 2012, vol. 21, no. 1, p. 333-339.

[10] SHAH, N. A., MALIK, M. A. Voltage/current-mode universal filter using FTFN and CFA. *Analog Integrated Circuits and Signal Processing,* 2005, vol. 45, p. 197–203.

[11] BIOLEK, D., SENANI, R., BIOLKOVÁ, V., KOLKA, Z. Active elements for analog signal processing: Classification, review, and new proposals. *Radioengineering,* 2008, vol. 17, no. 4, p. 15-32.

[12] YEŞIL, A., KAÇAR, F., KUNTMAN, H. New simple CMOS realization of voltage differencing transconductance amplifier and its RF filter application. *Radioengineering,* 2011, vol. 20, no. 3, p. 632-637.

[13] SATANSUP, J., PUKKALANUN, T., TANGSRIRAT, W. Electronically tunable single-input five-output voltage-mode universal filter using VDTAs and grounded passive elements. *Circuits Syst Signal Process*, 2012, DOI 10.1007/s00034-012-9492-0.

[14] SATANSUP, J., TANGSRIRAT, W. Single VDTA-based current-mode electronically tunable multifunction filter. In *4th International Science, Social Science, Engineering and Energy Conference.* Petchburi (Thailand), December 2012.

[15] PRASAD, D., BHASKAR, D. R. Electronically controllable explicit current output sinusoidal oscillator employing single VDTA. *International Scholarly Research Network ISRN Electronics*, 2012, Article ID 382560, 5 pages, doi:10.5402/2012/382560

[16] PRASAD, D., BHASKAR, D. R. Grounded and floating inductance simulation circuits using VDTAs. *Circuits and Systems,* 2012, vol. 3, p. 342-347.

[17] ARBEL, F., GOLDMINZ, L. Output stage for current-mode feedback amplifiers, theory and applications. *Analog Integrated Circuits and Signal Processing,* 1992, vol. 2, no. 3, p. 243–255.

[18] ALTUN, M., KUNTMAN, H. Design of a fully differential current mode operational amplifier with improved input-output impedances and its filter applications. *Int. Journal of Electronics and Communications (AEU),* 2008, vol. 62, no. 3, p. 239 to 244.

[19] ALLEN, P. E., HOLBERG, D. R. *CMOS Analog Circuit Design*. New York: Oxford, 2002.

[20] PETERSON, K. D., NEDUNGADI, A., GEIGER, R. L. Amplifier design consideration for high frequency monolithic filters. In *Proceedings of European Conference on Circuit Theory and Design*. 1987, p. 321-326.

[21] PROMMEE, P., PATTANATADAPONG, T. Realization of tunable pole-Q current-mode OTA-C universal filter. *Circuits, Systems, and Signal Processing,* 2010, vol. 29, no. 5, p. 913–924.

[22] MINAEI, S., YUCE, E. Novel voltage-mode all-pass filter based on using DVCCs. *Circuits, Systems and Signal Processing*, 2010, vol. 29, no. 3, p. 391-402.

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