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Generalized mixed mode universal filter realization using current controlled conveyors

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Abstract

Most of the work done on current controlled conveyor (CCCII) based continuous time filters has been in either current mode or voltage mode domain. Architectures that can work in mixed mode may be useful from IC realization viewpoint and application adaptability. This paper presents two generalized mixed mode configurations that use two DO-CCCII/MO-CCCII, two capacitors and two resistors. The circuits are capable of realizing all generic filter functions: low pass, band pass, high pass, notch and all pass. PSPICE simulation results are presented.
Keywords: Filter, Current conveyor, mixed mode

1 Introduction

In recent past there has been a great emphasis on designing active filters using current controlled conveyors (CCCIIs) [1]. Universal filters are useful class of filter since they permit realization of different filter functions with the same topology depending on the port/ports used. It provides a versatile, simple and cost effective solution to the integrated circuit manufacturers. Already a number of universal filter structures, [1-14] and references cited therein, based on current conveyor have been reported. However, most of the structures can be classified either as voltage mode or current mode [1-8] and a little work has been done in the domain of mixed mode universal/multifunction filters [9-14].

Generalized mixed mode filter structures i.e. a single structure which can realize voltage mode (i.e. both the input and output as voltage), current mode (i.e. both the input and output as current), trans-admittance mode (i.e. input as voltage and output as current) and trans-impedance mode (i.e. input as current and output as voltage) universal/multifunction filter are limited [10-14]. Comparison of the previous works [9-14] are made in Table 1. The proposed structure is obtained from the structure of refs. [7, 8]. The refs. [7, 8] are analysed only for CM mode responses, whereas the proposed circuit is obtained for a generalized mixed mode operation by inserting voltage signals at suitable terminals and a little alteration of circuit. Applications where power consumption and adaptation in integrated circuit environment are important, the number of active and passive components employed is of prime concern.

In this paper two new generalized mixed mode universal filters are presented that employ only two dual output current controlled conveyors(DO-CCCIIs), one resistor and two capacitors and another one uses one DO-CCCII, one multi-output CCCII (MO-CCCII), two capacitors and resistors each. The proposed structures realize all the standard functions of generalized mixed mode universal filter i.e. LP, BP, HP, notch (NF) and AP in all the four mixed modes (CM, VM, Trans-impedance mode, Trans-admittance mode). Most of the responses are available in both non-inverted (N.I) and inverted (I) form. This feature is useful in some applications such as the design of inverting/non-inverting BPF in hearing aid for selective amplification or attenuation of audio signal [15]. The filter, under all operations, exhibits low active and passive sensitivities. The workability of the proposed structure has been confirmed by PSPICE simulations.

Table-1

Related Work	No. Of Components Used			Functions Perform				No. of Input Current used	Orthogonal tunability of ω_0, Q_0 and ω_0/Q_0
	Current Conveyor/ CFOA	R	C	Current Mode	Transimpedance Mode	Transadmittance Mode	Voltage Mode		
Ref. 9	4 CCIIs	4	2	------	LP, BP, HP	------	-----	1	Possible
Ref. 10	4 CCCIIs	-	2	LP,BP, HP	LP, BP, HP	LP, BP, HP, NF	LP,BP, HP,NF	2	Possible
Ref. 11	7 CCIIs	8	2	LP,BP, HP,NF APF	LP ,BP, HP, NF, APF	LP ,BP, HP, NF, APF	LP,BP, HP,NF APF	3	Possible
Ref. 12	4 CFOAs	9	2	--do—	--do--	---do---	--do--	1	Possible
Ref. 13	3 CCIIs	3	4	--do—	--do--	---do---	--do--	3	Possible
Ref. 14	6 CCIIs	3	2	LP,BP, HP	LP, BP, HP	BP, HP	LP,BP, HP	1	Possible
Present work	2 CCCIIs	2	2	LP,BP, HP,NF APF	LP , BP, HP, NF, APF	LP, BP, HP, NF, APF	LP,BP, HP,NF APF	4*	Possible

* The number of input currents can be reduced to 3 by connecting a switch in series with resistor R_{out} between I_{out2} and ground in Fig. 1 [as used in Ref.13]

2 Circuit description

The proposed network of Fig.1 is based on employing DO-CCCIIs. The DO-CCCIIs have high impedance y terminal i.e. $i_y = 0$. Port relationship using standard notations can be represented as

$$v_{X_i} = v_{Y_i} + i_{X_i}|R_{Xi}(I_{0i})|, \quad i_{Z_i+} = i_{X_i}, \text{ and } i_{Z_i-} = -i_{X_i} \tag{1}$$

where $R_{Xi} = \mathrm{V}_T/\ 2\mathrm{I}_{0i}$, V_T is the thermal voltage, I_{0i} is bias current of CCCII and $i = 1, 2$.

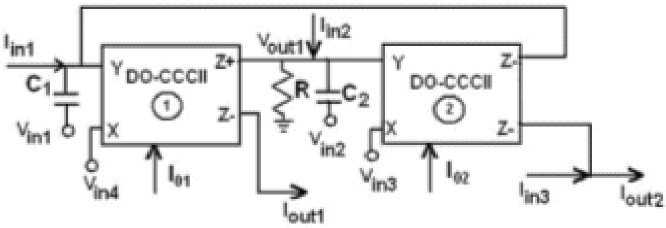

Figure 1: Proposed mixed mode circuit using DO-CCCIIs.

Routine analysis yields the transfer functions as

$$I_{out1} = \frac{N_{v1}(s) + N_{i1}(s)}{D(s)}, \ I_{out2} = \frac{N_{v2}(s) + N_{i2}(s)}{D(s)}, \ V_{out1} = \frac{N_{v3}(s) + N_{i3}(s)}{D(s)} \tag{2}$$

where

$$\begin{aligned}
&N_{i1}(s) = I_{in2} - (G + sC_2)R_{x2}I_{in1}, \\
&N_{v1}(s) = sC_2V_{in2} - (G + sC_2)V_{in3} - (V_{in1} - V_{in4})(G + sC_2)R_{x2}sC_1 \\
\\
&N_{i2}(s) = I_{in3}D(s) - sC_1R_{x1}I_{in2} - I_{in1}, \\
&N_{v2}(s) = -(V_{in1} - V_{in4})sC_1 + sC_1GR_{x1}V_{in3} + s^2C_1C_2R_{x1}(V_{in3} - V_{in2}) \\
\\
&N_{i3}(s) = sC_1R_{x1}R_{x2}I_{in2} + I_{in1}R_{x2}, \\
&N_{v3}(s) = V_{in3} + (V_{in1} - V_{in4})R_{x2}sC_1 + s^2C_1C_2R_{x1}R_{x2}V_{in2} \\
\\
&D(s) = R_{x1}R_{x2}C_1C_2s^2 + sGR_{x1}R_{x2}C_1 + 1
\end{aligned} \tag{3}$$

The functionality of the circuit of Fig.1 can be enhanced by replacing second DO-CCCII by a MO-CCCII (Fig.2). The analysis of circuit of Fig. 2 gives an additional voltage term as follows.

$$V_{out2} = \frac{N_{v4}(s) + N_{i4}(s)}{D(s)} \tag{4}$$

where

$$\begin{aligned} &N_{i4}(s) = R_{out}(I_{in4}D(s) - sC_1R_{x1}I_{in2} - I_{in1}) \\ &N_{v4}(s) = R_{out}(-(V_{in1} - V_{in4})sC_1 + sC_1GR_{x1}V_{in3} + s^2C_1C_2R_{x1}(V_{in3} - V_{in2})) \end{aligned} \tag{5}$$

From above equations one can see that specializations in the numerator result in filter functions as presented in Table 2 and Table 3 below for circuit of Fig. 2.

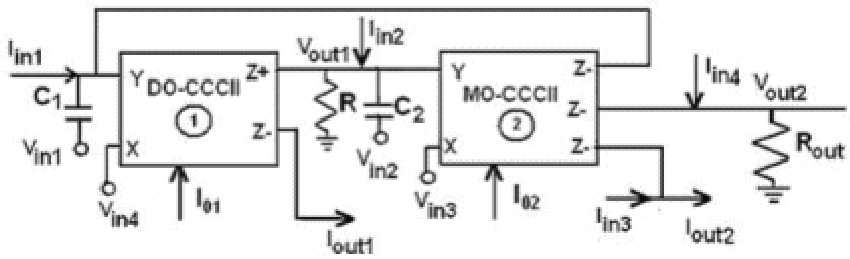

Figure 2: Proposed generalized mixed mode universal filter using DO-CCCII and MO-CCCII.

Case I. With V_{in1}= V_{in2}= V_{in3}= V_{in4}=0, we obtain current mode (CM) response and trans-impedance mode responses under the conditions as shown in Table 2.

Case II. With I_{in1}= I_{in2}= I_{in3}= I_{in4}=0, we obtain voltage mode (VM)and trans-admittance mode responses under the conditions as shown in Table 3.

We see in Table 2 and Table 3 that all the universal filter responses such as LP, BP, HP, Notch(NF) and all pass(AP) are obtained for all the four modes (VM, CM, Trans-impedance Mode, Trans-admittance Mode) of generalized mixed mode operations. However, in the case of trans-admittance mode, the notch responses are obtained as low-pass notch and high-pass notch and all pass response is obtained as low-pass-all-pass response. Tables also reveal that most of the outputs are available both in non-inverting (N.I) and inverting (I) modes which may be useful in some applications [15].

Table 2.

Function Type	*Case I*			
	CM mode		Trans-impedance mode	
	input	output	input	output
Low pass	$I_{in1}=0, I_{in2}=I_{in}$	I_{out1}(N.I.)	$I_{in2}=0, I_{in1}=I_{in}$	V_{out1}(N.I.)
	$I_{in3}=I_{in2}=0, I_{in1}=I_{in}$	I_{out2}(I.)	$I_{in4}=I_{in2}=0, I_{in1}=I_{in}$	V_{out2} (I.)
Band pass	$I_{in3}=I_{in1}=0\ \ I_{in2}=I_{in}$	I_{out2} (I.)	$I_{in1}=0, I_{in2}=I_{in}$	V_{out1}(N.I.)
			$I_{in4}=I_{in1}=0,\ I_{in2}=I_{in}$	V_{out2}(I.)
High pass	$I_{in3}=I_{in2}=I_{in1}=I_{in}, R_{x2}=R$	I_{out2}(N.I.)	$I_{in4}=I_{in2}=I_{in1}=I_{in}\ \ R_{x2}=R$	V_{out2}(N.I.)
Notch	$I_{in1}=0, I_{in3}=I_{in2}=I_{in}\ R_{x2}=R$	I_{out2}(N.I.)	$I_{in1}=0,\ I_{in4}=I_{in2}=I_{in}\ R_{x2}=R$	V_{out2}(N.I.)
All pass	$I_{in1}=0, I_{in3}=I_{in2}=I_{in}\ 2R_{x2}=R$	I_{out2} (N.I.)	$I_{in1}=0, I_{in3}=I_{in2}=I_{in}\ 2R_{x2}=R$	I_{out2} (N.I.)

(Note: N.I= Non-inverting, I= Inverting)

Table 3.

Function Type	*Case II*			
	VM mode		Trans-admittance mode	
	input	output	input	output
Low pass	$V_{in4}=V_{in2}=V_{in1}=0$, $V_{in3}=V_{in}$,	V_{out1}(N.I.)	$V_{in4}=V_{in1}=0$, $V_{in3}=V_{in2}=V_{in}$	I_{out1}(I.)
Band pass	$V_{in4}=V_{in3}=V_{in2}=0$, $V_{in1}=V_{in}$	V_{out1}(N.I.)	$V_{in4}=V_{in3}=V_{in1}=0, V_{in2}=V_{in}$	I_{out1}(N.I.)
	$V_{in3}=V_{in2}=V_{in1}=0$, $V_{in4}=V_{in}$	V_{out1}(I.)	$V_{in4}=V_{in3}=V_{in2}=0, V_{in1}=V_{in}$	I_{out2}(I.)
	$V_{in3}=V_{in2}=V_{in1}=0$, $V_{in4}=V_{in}$	V_{out2}(N.I.)	$V_{in3}=V_{in2}=V_{in1}=0, V_{in4}=V_{in}$	I_{out2}(N. I.)
	$V_{in4}=V_{in3}=V_{in2}=0$, $V_{in1}=V_{in}$	V_{out2}(I.)	$V_{in4}=V_{in1}=0, V_{in3}=V_{in2}=V_{in}$	I_{out2}(N. I.)
High pass	$V_{in4}=V_{in3}=V_{in1}=0, V_{in2}=V_{in}$	V_{out1}(N.I.)	$V_{in4}=V_{in3}=V_{in1}=0, V_{in2}=V_{in}$	I_{out2}(I.)
	$V_{in4}=V_{in3}=V_{in2}=0$, $V_{in1}=V_{in}$	V_{out2}(I.)		
Notch	$V_{in4}=V_{in1}=0, V_{in3}=V_{in2}=V_{in}$	V_{out1}(N.I.)	$V_{in4}=0, V_{in3}=V_{in2}=V_{in1}=V_{in}$ $R_{x1}=2R$, $R_{x2}=R$	$I_{out1}+I_{out2}$(I.)
All pass	$V_{in1}=0, V_{in2}=V_{in3}=V_{in4}=V_{in}$,Rx1=R	V_{out1}(N.I.)	$V_{in4}=0, V_{in3}=V_{in2}=V_{in1}=V_{in}$ $R_{x1}=3R$, $R_{x2}=R$	$I_{out1}+I_{out2}$(I.)

It may also be noted that all the responses as jotted down in case-I and case-II will also be obtained for circuit of Fig.1 except V_{out2}. All the filters are characterized by

$$\omega_0 = (\frac{1}{R_{x1}R_{x2}C_1C_2})^{1/2}, \quad \frac{\omega_0}{Q_0} = \frac{1}{RC_2} \text{ and } Q_0 = R(\frac{C_2}{R_{x1}R_{x2}C_1})^{1/2} \tag{6}$$

We notice that Q_0 can be independently controlled keeping ω_0 unchanged through the grounded resistor R for LP and BP functions in case-I and through R and then R_{x1} and R_{x2} for HP, Notch and AP functions. Similarly, in case-II, the LP, BP and HP functions through R and for Notch and AP functions through R and then R_{x1} and R_{x2}. It may be noted that the resistances R_{x1} and R_{x2} can easily be adjusted to the required values by externally controlling the bias currents (I_{01} and I_{02}) of DO-CCCIIs/MO-CCCII. The parameter ω_0 can also be adjusted electronically by controlling bias currents of DO-CCCIIs/MO-CCCII without disturbing ω_0/Q_0.

3 Effect of current controlled conveyor non-idealities

The frequency performance of the filter circuit may deviate from the ideal one due to nonidealities. The nonidealities effects may be categorized in two groups. The first comes from frequency dependence of internal current and voltage transfers of DO-CCCII/MO-CCCII. Due to the non-ideality, the port relations of Eq. (1) may be expressed as

$$v_{X_i} = v_{Y_i}\beta_i(s) + i_{X_i}|R_{Xi}(I_{0i})|, \quad i_{Z_{i+}} = i_X\alpha_{i+}(s) \text{ and } i_{Z_{i-}} = -i_X\alpha_{i-}(s) \tag{7}$$

where $i = 1$ and 2, β_i(s) and [α_{i+}(s) , α_{i-}(s)] denote voltage and current transfer functions respectively, which can be modeled as

$$\beta_i(s) = \frac{\beta_i}{s+\omega_{\beta_i}}, \quad \alpha_{i+}(s) = \frac{\alpha_{i+}}{s+\omega_{\alpha_{i+}}} \text{ and } \alpha_{i-}(s) = \frac{\alpha_{i-}}{s+\omega_{\alpha_{i-}}}.$$

The terms β_i, α_{i+}, and α_{i-} represent low frequency values of β_i(s), α_{i+}(s), and α_{i-}(s) respectively and $\omega_{\beta_i}$and $\omega_{\alpha_i}$are respectively pole frequencies of voltage and current transfer functions. The transfer functions represented in Eqs. (2) and (4) for various current and voltage modify as:

$$\begin{array}{ll} I_{out1}|_n = \frac{N_{vn1}(s)+N_{in1}(s)}{D_n(s)}, & I_{out2}|_n = \frac{N_{vn2}(s)+N_{in2}(s)}{D_n(s)}, \\ V_{out1}|_n = \frac{N_{vn3}(s)+N_{in3}(s)}{D_n(s)}, & V_{out2}|_n = \frac{N_{vn4}(s)+N_{in4}(s)}{D_n(s)} \end{array} \tag{8}$$

where

$$N_{in1}(s) = \frac{\alpha_{1-}\alpha_{2-}\beta_1\beta_2 I_{in2}}{(s+\omega_{\alpha_{1-}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})(s+\omega_{\beta_2})} - \frac{\alpha_{1-}\beta_1(G_2+sC_2)R_{x2}I_{in1}}{(s+\omega_{\alpha_{1-}})(s+\omega_{\beta_1})}$$

$$N_{vn1}(s) = \frac{sC_2\alpha_{1-}\alpha_{2-}\beta_1\beta_2 V_{in2}}{(s+\omega_{\alpha_{1-}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})(s+\omega_{\beta_2})} - \frac{\alpha_{1-}\alpha_{2-}\beta_1(G_2+sC_2)V_{in3}}{(s+\omega_{\alpha_{1-}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})}$$

$$-\frac{\alpha_{1-}(G_2+sC_2)R_{x2}}{(s+\omega_{\alpha_{1-}})}\left(\frac{\beta_1 sC_1V_{in1}}{(s+\omega_{\beta_1})} - sC_1V_{in4}\right)$$

$$N_{in2}(s) = I_{in3}D_n(s) - \frac{\alpha_{2-}\beta_2 sC_1R_{x1}I_{in2}}{(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_2})} - \frac{\alpha_{1+}\alpha_{2-}\beta_1\beta_2 I_{in1}}{(s+\omega_{\alpha_{1+}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})(s+\omega_{\beta_2})}$$

$$N_{vn2}(s) = -\frac{\alpha_{1+}\alpha_{2-}\beta_2}{(s+\omega_{\alpha_{1+}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_2})}\left(\frac{\beta_1 sC_1V_{in1}}{(s+\omega_{\beta_1})} - sC_1V_{in4}\right)$$

$$+\frac{\alpha_{2-}(G_2+sC_2)sC_1R_{x1}V_{in3}}{(s+\omega_{\alpha_{2}-})} - \frac{\alpha_{2-}\beta_2}{(s+\omega_{\alpha_{2}-})(s+\omega_{\beta_2})}s^2C_1C_2R_{x1}V_{in2}$$

$$N_{in3}(s) = sC_1R_{x1}R_{x2}I_{in2} + \frac{\alpha_{1+}\beta_1R_{x2}I_{in1}}{(s+\omega_{\alpha_{1+}})(s+\omega_{\beta_1})}$$

$$N_{vn3}(s) = \frac{\alpha_{1+}\alpha_{2-}\beta_1V_{in3}}{(s+\omega_{\alpha_{1+}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})} + \frac{\alpha_{1+}\beta_1 sC_1R_{x2}V_{in1}}{(s+\omega_{\alpha_{1+}})(s+\omega_{\beta_1})}$$

$$-\frac{\alpha_{1+}sC_1R_{x2}V_{in4}}{(s+\omega_{\alpha_{1+}})} + s^2C_1C_2R_{x1}R_{x2}V_{in2}$$

$$N_{in4}(s) = R_{out}(I_{in3}D_n(s) - \frac{\alpha_{2-}\beta_2 sC_1R_{x1}I_{in2}}{(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_2})}$$
$$-\frac{\alpha_{1+}\alpha_{2-}\beta_1\beta_2 I_{in1}}{(s+\omega_{\alpha_{1+}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})(s+\omega_{\beta_2})})$$

$$N_{vn4}(s) = -R_{out}(\frac{\alpha_{1+}\alpha_{2-}\beta_2}{(s+\omega_{\alpha_{1+}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_2})}\left(\frac{\beta_1 sC_1V_{in1}}{(s+\omega_{\beta_1})} - sC_1V_{in4}\right)$$

$$+\frac{\alpha_{2-}(G_2+sC_2)sC_1R_{x1}V_{in3}}{(s+\omega_{\alpha_{2}-})} - \frac{\alpha_{2-}\beta_2}{(s+\omega_{\alpha_{2}-})(s+\omega_{\beta_2})}s^2C_1C_2R_{x1}V_{in2})$$

$$D_n(s) = sC_1(G_2+sC_2)R_{x1}R_{x2} + \frac{\alpha_{1+}\alpha_{2-}\beta_1\beta_2}{(s+\omega_{\alpha_{1+}})(s+\omega_{\alpha_{2-}})(s+\omega_{\beta_1})(s+\omega_{\beta_2})} \tag{9}$$

The above equations clearly indicate that the pole frequencies of voltage and current transfer functions of DO-CCCIIs/MO-CCCII affect the overall filter response. The effect can however be ignored if the operating frequencies are chosen sufficiently smaller than voltage and current transfer pole frequencies of DO-CCCII/MO-CCCII.

The second group of nonidealities comes from parasitics of DO-CCCII/MO-

CCCII comprising of resistances and capacitances connected in parallel at terminals y and z (i.e. R_y, C_y, R_z, C_z) and inductance L_x in series to R_x at terminal x [17].The effects of these parasitics on filter response depend strongly on circuit topology. In the presence of these parasitics the circuit given in Fig. 2 modifies to Fig. 3 where

$$\begin{aligned} C_{1p} &= C_{y1}//C_{z2}, \qquad C_{2p} = C_{z1}//C_{y2}, \\ G_{1p} &= 1/\left(R_{y1}//R_{z2}\right), \text{ and } G_{2p} = 1/\left(R_{y2}//R_{z1}\right). \end{aligned}$$

The inductance is ignored in Fig. 3 as it affects the frequency response only at very high frequency.

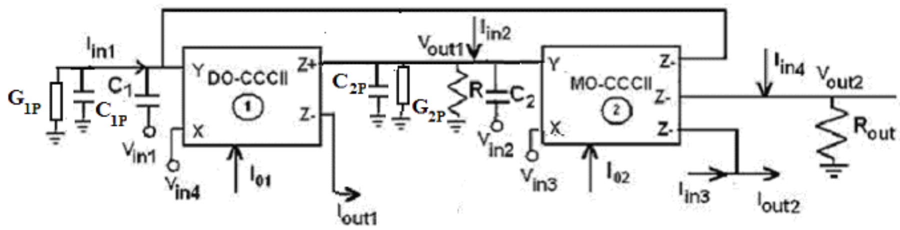

Figure 3: Proposed filter structure including parasitics.

Considering the parasitics outlined above the expressions in Eqs. 3 and 5 modify to

$$\begin{aligned} N_{in1}(s) &= I_{in2} - (G_{2eq} + sC_{2eq})R_{x2}I_{in1} \\ N_{vn1}(s) &= sC_2V_{in2} - (G_{2eq} + sC_{2eq})V_{in3} \\ &\quad -(sC_1V_{in1} - (G_{1p} + sC_{1eq})V_{in4})(G_{2eq} + sC_{2eq})R_{x2} \end{aligned}$$

$$\begin{aligned} N_{in2}(s) &= I_{in3}D_n(s) - (G_{1p} + sC_{1eq})R_{x1}I_{in2} - I_{in1} \\ N_{vn2}(s) &= -(sC_1V_{in1} - (G_{1p} + sC_{1eq})V_{in4}) \\ &\quad +(G_{2eq} + sC_{2eq})(G_{1p} + sC_{1eq})R_{x1}V_{in3} - sC_2(G_{1p} + sC_{1eq})R_{x1}V_{in2} \end{aligned}$$

$$\begin{aligned} N_{in3}(s) &= (G_{1p} + sC_{1eq})R_{x1}R_{x2}I_{in2} + R_{x2}I_{in1} \\ N_{vn3}(s) &= V_{in3} + sC_1R_{x2}V_{in1} \\ &\quad -(G_{1p} + sC_{1eq})R_{x2}V_{in4} + (G_{1p} + sC_{1eq})sC_2R_{x1}R_{x2}V_{in2} \end{aligned}$$

$$
\begin{aligned}
&N_{in4}(s) = R_{out}(I_{in3}D_n(s) - (G_{1p} + sC_{1eq})R_{x1}I_{in2} - I_{in1}),\\
&N_{vn4}(s) = R_{out}(-(sC_1V_{in1} - (G_{1p} + sC_{1eq})V_{in4})\\
&\qquad +(G_{2eq} + sC_{2eq})(G_{1p} + sC_{1eq})R_{x1}V_{in3} - sC_2(G_{1p} + sC_{1eq})R_{x1}V_{in2})\\
&D_n(s) = (G_{1p} + sC_{1eq})(G_{2eq} + sC_{2eq})R_{x1}R_{x2} + 1,\\
&G_{2eq} = G + G_{2p}, \text{where } G = 1/R\\
&C_{1eq} = C_{1p}//C_1, \quad C_{2eq} = C_{2p}//C_2, \text{ when } V_{in1} = V_{in2} = 0
\end{aligned}
\tag{10}
$$

A careful investigation of the above equations and Fig. 3 reveal that in the proposed topology parasitic conductance G_{2p} is appearing in parallel to G, hence can be compensated by pre-distorting value of external passive component R. Similarly the effects of parasitic capacitance C_{1P} and C_{2P} can be compensated by pre-distorting C_1 and C_2 when $V_{in1}= V_{in2}= 0$. To further investigate the effect of parasitics and to obtain corresponding approximate design criterion, we have taken a case of low pass current mode (CM) filter. The current mode low pass response may be obtained as

$$
I_{out1}(s)|_n = \frac{I_{in2}}{(G_{1p} + sC_{1eq})(G_{2eq} + sC_{2eq})R_{x1}R_{x2} + 1} \tag{11}
$$

Here,

$$
\frac{\omega_0}{Q_0}|_n = \frac{G_{1p}}{C_{1eq}} + \frac{G_{2p}}{C_{2eq}} + \frac{G}{C_{2eq}}. \tag{12}
$$

Let us consider $C_1 = C_2 = C \;>> $ parasitic capacitances. Hence we can write $C_{1eq} \cong C_{2eq} \quad \cong C_1 = C_2 = C$.

So, the ideal Band Width can be written as

$$
\frac{\omega_0}{Q_0} = \frac{G}{C} \cong \frac{G}{C_{2eq}} \tag{13}
$$

We find from Eqs.(12) and (13) that $\frac{\omega_0}{Q_0}|_n$ will be approximately equal to ideal case if we choose

$$
\frac{\omega_0}{Q_0} \cong \frac{G}{C_{2eq}} \;>>\; \frac{G_{1p}}{C_{1eq}} + \frac{G_{2p}}{C_{2eq}}
$$

or,

$$
C >> [(G_{1p} + G_{2p})\frac{Q_0}{\omega_0} = C_{D1}(say)] \tag{14}
$$

Similarly

$$\omega_0^2|_n = \frac{1}{R_{x1}R_{x2}C_{1eq}C_{2eq}} + \frac{G_{1p}(G_{2p}+G)}{C_{1eq}C_{2eq}}$$

will be approximately equal to the ideal case if we choose

$$\omega_0^2 \cong \frac{1}{R_{x1}R_{x2}C_{1eq}C_{2eq}} >> \frac{G_{1p}(G_{2p}+G)}{C_{1eq}C_{2eq}},$$

or,

$$\omega_0^2 >> (\frac{G_{1p}G_{2p}}{C_{1eq}C_{2eq}} + \frac{G_{1p}\omega_0}{C_{1eq}Q_0})$$

The first term in R. H. S. can be neglected, so

$$C >> [\frac{G_{1p}}{\omega_0 Q_0} = C_{D2}(say)] \tag{15}$$

Thus by choosing

$$C = C_1 = C_2 >> \max(C_{D1}, C_{D2}) \tag{16}$$

the effect of parasitic impedance can be practically eliminated and thus filter may approach towards ideal response. However, the maximum frequency of operation will be limited by poles of current (f_α) and voltage (f_β) transfers which are simulated to be 61.2MHz and 215.6MHz respectively for the proposed circuit.

It can also be easily evaluated to show that the active and passive sensitivities of pole ω_0 and pole Q_0 are within unity in magnitude. Thus the proposed structures can be classified as insensitive.

4 Simulation results

To validate the theoretical predictions, the proposed circuit of Fig. 2 is simulated with PSPICE using translinear current conveyor [1] and typical parameters of bipolar transistors PR100N (PNP) and NR100N (NPN) [16] with supply voltages of ±2.5 volts. It is found that to practically eliminate the effect of parasitics of CCCII the value of $C = C_1 = C_2$ has to satisfy equation 16. The parasitic resistances, capacitances and inductance and values of f_α and f_β are simulated to be R_y = 91 KΩ, R_z = 384 KΩ, C_y = 5.28 pF, C_z = 2.28 pF, L_x= 0.2 μH, 61.2MHz and 215.6MHz respectively for CCCIIs bias current = 100 μA. The corresponding values of C_{D1} (Eq.14) and C_{D2} (Eq.15) are calculated to be 21.4 pF and 10.71 pF respectively

Table 4.

Designed(ideal) value of f_0	$I_{01} = I_{02} = I_{03}$	C	Calculated $Max(C_{D1},C_{D2})$	$\frac{C}{Max(C_{D1},C_{D2})}$	f_0(simulated)
1.27MHz	100µA	1nF	21.40 pF	46.70	1.26MHz
1.27MHz	90µA	0.9nF	20.33 pF	44.26	1.25MHz
1.27MHz	80µA	0.8nF	19.24 pF	41.58	1.24MHz
1.27MHz	70µA	0.7nF	18.19 pF	38.48	1.095MHz
1.27MHz	60µA	0.6nF	16.90 pF	35.50	0.929MHz
1.27MHz	50µA	0.5nF	15.64 pF	31.96	0.724MHz
1.27MHz	40µA	0.4nF	14.20 pF	28.16	0.529MHz
1.27MHz	30µA	0.3nF	12.52 pF	23.96	0.378MHz
1.27MHz	20µA	0.2nF	9.30 pF	21.50	0.242MHz
1.27MHz	10µA	0.1nF	7.00 pF	14.28	0.119MHz
1.27MHz	5µA	0.05nF	4.50 pF	11.11	0.0669 MHz
1.27MHz	1µA	0.01nF	1.20 pF	08.33	0.01448MHz

for bias current = 100μA. In order to show the effects of parasitics on f_0, the simulation of low pass filter was carried out for $f_0 = 1.27$ MHz at $Q_0 = 1$. Table 4 shows the various values of capacitance $C(= C_1 = C_2)$ and corresponding ideal and simulated values of f_0. It is evident form Table 4 and Fig. 4 that when values of C are low, the simulated value of f_0 deviate strongly from ideal value and when C is about 45 times of $\max(C_{D1}, C_{D2})$, the simulated f_0 is very closed to ideal value. Hence to get an output almost free from parasitics effect, the value of $C = C_1 = C_2$ should be about 45 times of max (C_{D1}, C_{D2}) or more.

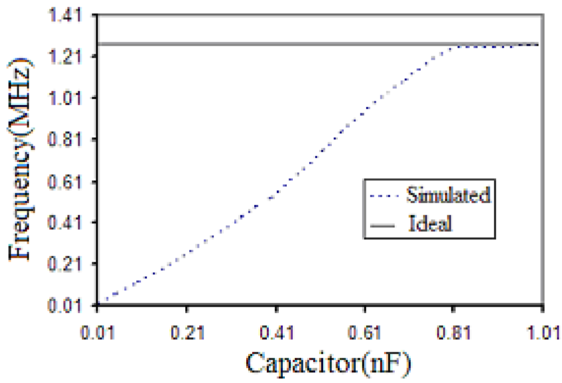

Figure 4: Dependence of frequency on capacitor (C) values.

Figure 5(a) shows the simulated and theoretical results for low pass, band pass and high pass responses for a pole frequency of f_0= 1.27 MHz, quality factor of Q_0= 1 and the component values of C = C_2= C_3= 1nF and $I_{01} = I_{02} = 100\mu$A. Tunability of Q_0 with ω_0 unchanged for band pass filter is shown in Fig. 5(b). This is designed at f_0= 1.27 MHz with C_1= C_2= 1nF, R_{x1}= 0.125K (i.e. $I_{01} = 100\mu$A), R_{x2}= 0.125K (i.e. I_{02}= 100μA) and R = 0.125K (Q_0=1), 0.625K (Q_0=5), 1.25K (Q_0=10). The simulation and theoretical results agree quite well.

The proposed circuit is also tested for dynamic range i.e. up to the level of input signal for which the output is within the permitted distortion level. The response is obtained for band pass current mode filter by applying a sinusoidal current input at f_0= 1.27 MHz with $C_1 = C_2 =$ 1nF, R= 125Ω and $I_{01} = I_{02} = 100\mu$A. The result in Fig. 6 shows that for large range of input signal level, output distortion is within acceptable limit of the order of THD = 5%. It shows that the proposed circuit is also useful for even large signal.

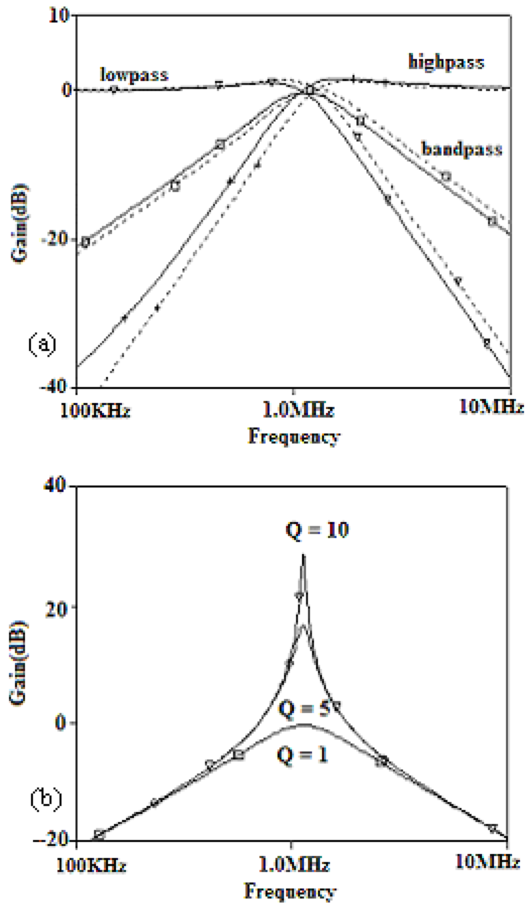

Figure 5: (a) Response of current mode (current-input current-output) filter of Fig. 2 at I_{out2}. Solid line - simulated, dashed line - theoretical. (b) Orthogonal tunability of Q_0 and ω_0.

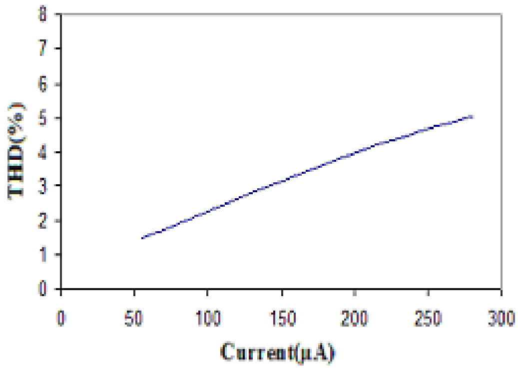

Figure 6: Dependence of output total harmonic distortion (%THD) on input signal amplitude.

5 Conclusion

In this paper, two new generalized mixed mode universal filter architectures are proposed that can be used as VM (voltage-input voltage-output) or CM (current-input current-output) or Trans-impedance mode(current-input voltage-output) or Trans-admittance mode (voltage-input current-output). The circuits use either two DO-CCCIIs, two capacitors and one resistor or one DO-CCCII, one MO-CCCII, two capacitors and resistors each. However the proposed circuits, like any other multi-input filter [10,11,13], require additional active elements to feed same current at multiple input nodes. They have the following attractive features:

1. They can realize all the standard functions of universal filter i.e. LP, BP, HP, Notch and AP.
2. Both the ω_0 and Q_0 and ω_0 and ω_0/Q_0 are orthogonally tunable.
3. Active and passive sensitivities of ω_0, Q_0 and ω_0/Q_0 are low and within unity in magnitude.
4. Most of the responses are available in both non-inverted and inverted form. This feature is useful in some applications [15].

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