

Realization of Novel Cascadable Current-Mode All-pass Sections

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Abstract: This paper introduces four new resistorless circuits of first-order current-mode all-pass filter (CMAPF) based on dual-X current conveyor transconductance amplifier (DXCCTA). All the four circuits use a single DXCCTA and a capacitor for their realization. The main features of the proposed CMAPFs are: use of minimum active and passive components, resistorless realization, electronically adjustable pole frequency, easily cascadable, good sensitivity performance with respect to active and passive elements, low total harmonic distortion of output current (0.74%) and good operating frequency range (39.2 MHz). The non-ideal analysis of the proposed circuits has also been explored. Moreover, two applications of the proposed first-order CMAPF in terms of second order CMAPF and current-mode quadrature oscillator are also presented. HSPICE simulations have been carried out with 0.18 μm CMOS process parameters to validate the proposed circuits.

Keywords: All-pass Filter, Quadrature Oscillator, Electronic Tuning, Cascadable, DXCCTA.

1 Introduction

THE wide range of applications of all-pass filters (APFs) have made them important cell for communication systems. Their main applications include phase shifters while maintaining amplitude to be constant, generation of quadrature signals or multiphase signals [1] and design of high-Q band-pass circuits. Both the voltage-mode as well as current-mode first-order APFs based on numerous high performance active elements are presented in the literature [2-23]. However, current-mode circuits have gained significant attention because they have increased bandwidth, simple structures of circuits, widespread dynamic range and low power consumption [24]. Therefore, designers have shown interest to develop the first order current-mode all-pass filter (CMAPFs) while focusing on reducing the number of active and passive elements, resistor less

realization, ease of cascability, electronic tunability of pole frequency etc.

The circuits of APFs presented in [2,4-9,11-16,18,19] have the advantage of cascability and do not have any component matching restrictions. Additionally the circuits presented in [2-5,7,9,12,14,15,17,18,20,21,23] are based on single active element and circuits presented in [2,3,9-16,19] have the property of electronic tuning. However, these APFs suffer from one or more of the following weaknesses; no resistorless realization [4-7,9,17,18,20-23], no electronic tuning [4-8,17-18,20-23], use of more than one passive components [4-7,9,17,18,20-23] and lower pole frequency [2-23].

In this paper four novel circuits of CMAPF that comprise single dual-X current conveyor transconductance amplifier (DXCCTA) [25] and a capacitor are proposed. Thus, no external resistor is required for their realization. The input impedance of the proposed CMAPFs is low and output impedance is high. Therefore, the proposed circuits are easily cascadable. Moreover, the pole frequency is electronically tunable. The sensitivity performance of the proposed circuits is also found good. Furthermore, the presented CMAPFs exhibit low total harmonic distortion (THD) for the output current. A comparison of the presented CMAPFs with few of the previously

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reported APFs is shown in Table. 1. Two applications in terms of second order CMAPF and current-mode quadrature oscillator (CM-QO) are also given. HSPICE simulation results using 0.18 μm are depicted to examine the performance of the presented circuits.

2 DXCCTA Basics

DXCCTA [25] combines the feature of both DXCCII [26] and OTA. The symbol and CMOS implementation of DXCCTA are shown in Fig. 1(a) and (b), respectively. In CMOS implementation of DXCCTA, the MOS transistors, M_1 - M_{20} form the input stage, DXCCII and MOS transistors, M_{21} - M_{32} form the output stage, OTA. The current mirror formed by M_{14} and M_{15} forces the drain currents of M_1 and M_2 to be equal. Thus, voltage of $X+$ terminal becomes equal to voltage

of Y terminal. The transistors M_3 - M_5 , M_{16} - M_{18} form the inverting voltage follower which causes the voltage at terminal, Y to be followed in inverting manner by voltage at terminal, $X-$. Moreover, the gate voltages of M_{12} and M_{13} are equal which causes the current at terminal, $X+$ to be followed by current at terminal, $Z+$. Similarly the current of terminal, $X-$ is transferred to the terminal, $Z-$. In OTA stage, the MOS transistors, M_{21} - M_{22} , M_{25} - M_{26} form a single input balanced output differential amplifier. The current mirrors formed by MOS transistors M_{24} - M_{25} , M_{26} - M_{27} and M_{31} - M_{32} cause the output of differential amplifier to appear at terminal, $O+$. The MOS transistors, M_{23} , M_{28} - M_{30} form an inverter thus, current at $O-$ terminal flows in opposite direction of current at $O-$ terminal. The following hybrid matrix gives the port relationships of DXCCTA

Table 1 Comparison of the proposed CMAPF with few of the already existing APFs.

Ref. No.	ABB used	ABB count	Passive components count	Resistorless realization	Cascadable	Component matching constraints	Electronic tunability	CMOS technology (μm)	Operational frequency (MHz)
2	DV-DXCCII	1	1	Yes	Yes	No	Yes	0.18	27
3	EX-CCCII	1	1	Yes	No	No	Yes	0.25	1.9
4	MO-DXCCII	1	2	No	Yes	No	No	0.35	1.57
5	DX-MOCCII	1	2	No	Yes	No	No	0.35	6.36
6	DO-CCII	2	2	No	Yes	No	No	0.5	30.3
7	DXCCII	1	2	No	Yes	No	No	0.35	1.51
8	DO-CCII	2	1	Yes	Yes	No	No	0.13	6.37
9	CBTA	1	2	No	Yes	No	Yes	0.18	1
10	CCCII	2	1	Yes	No	No	Yes	0.25	2
11	CDTA	2	1	Yes	Yes	No	Yes	NA	0.318
12	CDTA	1	1	Yes	Yes	No	Yes	0.35	1.35
13	CCCCTA	2	1	Yes	Yes	No	Yes	NA	0.2
14	CCTA	1	1	Yes	Yes	No	Yes	NA	0.284
15	CFCTA	1	1	Yes	Yes	No	Yes	0.25	2
16	ACA & CF	3	1	Yes	Yes	No	Yes	0.18	1
17	MO-DXCCII	1	3	No	Yes	Yes	No	0.18	21.3
18	DXCCII	1	2	No	Yes	No	No	0.5	25
19	DDCC	2	1	Yes	Yes	No	Yes	0.5	1.17
20	CCII	1	3	No	No	Yes	No	NA	0.001
21	CCIII*	1	4	No	No	Yes	No	NA	0.001
22	DVCC*	2	3	No	No	Yes	No	0.5	1.5
23	CCII*	1	3	No	No	Yes	No	NA	110
P	DXCCTA	1	1	Yes	Yes	No	Yes	0.18	39.2

Abbreviations:- DV-DXCCII: Differential voltage dual-X second generation current conveyor, EX-CCCII: Extra-X current controlled conveyor, MO-DXCCII: Multi-output dual-X second generation current conveyor, DX-MOCCII: dual-X second generation multioutput current conveyor, DO-CCII: Dual output second generation current conveyor, DXCCII: Dual-X second generation current conveyor, CBTA: Current backward transconductance amplifier, CCCII: current controlled current conveyor, CDTA: Current differencing transconductance amplifier, CCCCTA: Current controlled current conveyor transconductance amplifier, CCTA: current conveyor transconductance amplifier, CFCTA: Current follower cascaded transconductance, DDCC: Differential difference current conveyor, DVCC: Differential voltage current conveyor, ACA & CF: Adjustable current amplifier & current follower, DXCCTA: Dual-X current conveyor transconductance amplifier, ABB: Active building block, NA: Not applicable, *: 1st structure, P: Proposed.

$$\begin{bmatrix} I_Y \\ V_{X+} \\ V_{X-} \\ I_{Z+} \\ I_{Z-} \\ I_{O+} \\ I_{O-} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ -1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & g_m \\ 0 & 0 & 0 & -g_m \end{bmatrix} \begin{bmatrix} V_Y \\ I_{X+} \\ I_{X-} \\ V_{Z-} \end{bmatrix} \quad (1)$$

The transconductance, g_m in Eq. (1) depends on the bias current I_B and it is expressed as $g_m = \sqrt{k_n I_B}$ where $k_n = \mu_n C_{OX} (W/L)$ is the physical parameter of MOS transistors. It is well known that physical parameter, k_n is temperature dependent [27] thus, the transconductance, g_m is also temperature dependent and its value decreases with increase in temperature.

3 Proposed First-order Current-mode All-pass Filters

The newly proposed first-order current-mode all-pass filters, CMAPF-I, CMAPF-II, CMAPF-III and CMAPF-IV are shown in Fig. 2(a), (b), (c) and (d), respectively. The CMAPF-I and CMAPF-III employ one DXCCTA and one capacitor only whereas, CMAPF-II and CMAPF-IV employ one multiple output DXCCTA along with a single capacitor. It is to be mentioned that in case of CMAPF-III and CMAPF-IV the voltage of Z+ terminal (V_{Z+}) is transferred in terms of current to the O+ and O- terminals instead of voltage of Z- terminal (V_{Z-}). Therefore, a little change in CMOS implementation (terminal Z+ is connected to the gate of M_{21} instead of terminal Z-) is needed in case of

CMAPF-III and CMAPF-IV. Using the port relationships of DXCCTA given in Eq. (1), all the proposed CMAPFs are analyzed and transfer function obtained after analysis is given as follows:

$$\frac{I_{out}}{I_{in}} = \frac{s - \frac{g_m}{C}}{s + \frac{g_m}{C}} \quad (2)$$

The pole frequency, ω_0 and frequency dependent phase angle, $\varphi(\omega)$ are given as

$$\omega_0 = \frac{g_m}{C} = \frac{\sqrt{k_n I_B}}{C} \quad (3)$$

$$\varphi(\omega) = \pi - 2 \tan^{-1} \left(\frac{\omega C}{g_m} \right) \quad (4)$$

It is to be noted from Eq. (3) that electronic adjustment of ω_0 is possible by means of bias current I_B . Furthermore, the input impedance of all CMAPFs is low and output impedance of all CMAPFs is high. Thus all the proposed CMAPFs are easily cascadable.

3.1 Non-Ideal Analysis

The following equation gives the port relationships for the non-ideal model of DXCCTA

$$\begin{aligned} V_{X+} &= \beta_1 V_Y, & V_{X-} &= -\beta_2 V_Y, \\ I_{Z+} &= \alpha_1 I_{X+}, & I_{Z-} &= \alpha_2 I_{X-}, \\ I_{O+} &= \gamma_1 g_m V_{Z-}, & I_{O-} &= -\gamma_2 g_m V_{Z-} \end{aligned} \quad (5)$$

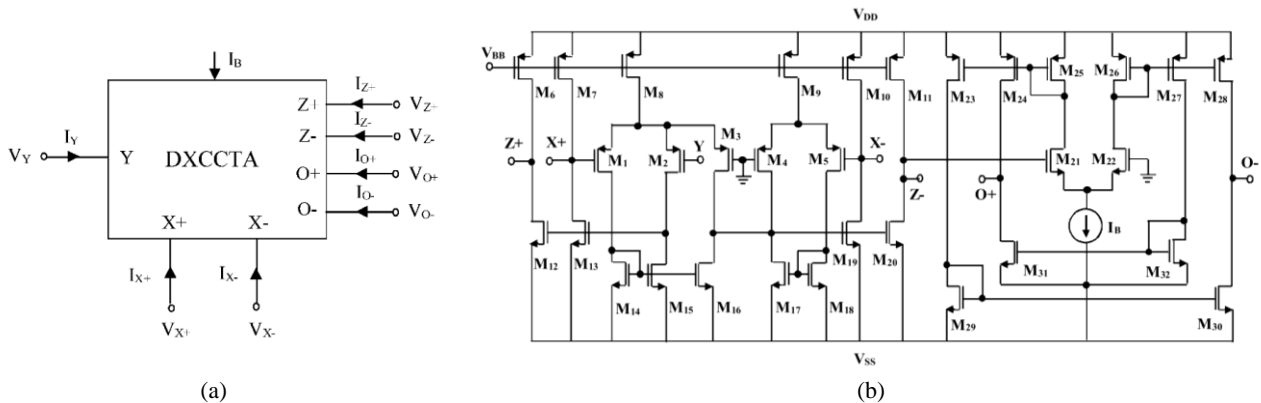


Fig. 1 a) Symbol of DXCCTA, b) CMOS implementation of DXCCTA [25].

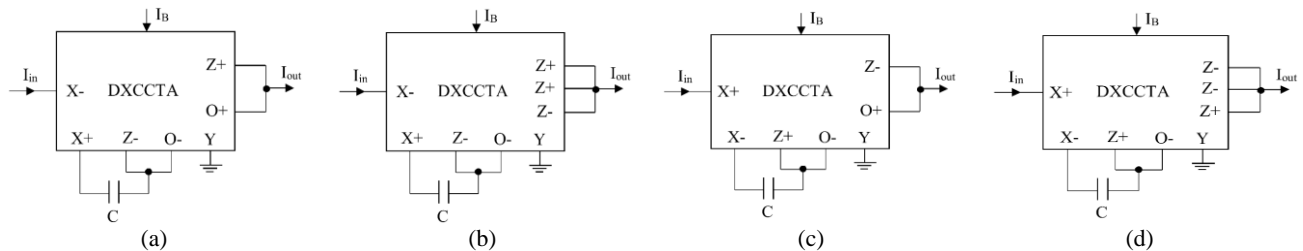


Fig. 2 Proposed first-order current-mode all-pass filter circuits: a) CMAPF-I, b) CMAPF-II, c) CMAPF-III and d) CMAPF-IV.

where, β_1, β_2 and α_1, α_2 are the voltage transfer gains (Y to $X+$, and Y to $X-$) and current transfer gains ($X+$ to $Z+$, and $X-$ to $Z-$), respectively. The γ_1 and γ_2 are the transconductance inaccuracies from $Z-$ to $O+$ terminal and from $Z-$ to $O-$ terminal, respectively. It is to be further mentioned that in case of CMAPF-III and CMAPF-IV, the expressions of I_{O+} and I_{O-} are given as follows:

$$I_{O+} = \gamma_1 g_m V_{Z+} \quad \text{and} \quad I_{O-} = -\gamma_2 g_m V_{Z+} \quad (6)$$

Considering the non-ideal port relationships given in Eq. (5) and (6), the all four CMAPFs are reanalyzed and provide the following transfer functions

$$\text{CMAPF-I: } \frac{I_{out}}{I_{in}} = \alpha_1 \alpha_2 \left[\frac{s - \frac{\gamma_1 g_m}{\alpha_1 C}}{s + \frac{\gamma_2 g_m}{C}} \right] \quad (7)$$

$$\text{CMAPF-II: } \frac{I_{out}}{I_{in}} = \alpha_2 (2\alpha_1 - 1) \left[\frac{s - \frac{\gamma_2 g_m}{(2\alpha_1 - 1)C}}{s + \frac{\gamma_2 g_m}{C}} \right] \quad (8)$$

$$\text{CMAPF-III: } \frac{I_{out}}{I_{in}} = \alpha_1 \alpha_2 \left[\frac{s - \frac{\gamma_1 g_m}{\alpha_2 C}}{s + \frac{\gamma_2 g_m}{C}} \right] \quad (9)$$

$$\text{CMAPF-IV: } \frac{I_{out}}{I_{in}} = \alpha_1 (2\alpha_2 - 1) \left[\frac{s - \frac{\gamma_2 g_m}{(2\alpha_2 - 1)C}}{s + \frac{\gamma_2 g_m}{C}} \right] \quad (10)$$

Equations (7) to (10) provide the following pole frequency

$$\omega_0 = \frac{\gamma_2 g_m}{C} \quad (11)$$

The sensitivities of ω_0 with respect to active components and capacitor are given as

$$S_{\gamma_2}^{f_0} = S_{g_m}^{f_0} = 1, \quad S_C^{f_0} = -1 \quad \text{and} \quad S_{\beta_1, \alpha_1, \alpha_2, \gamma_1}^{f_0} = 0 \quad (12)$$

Equation (12) reveals that none of the sensitivity is in excess of unity in magnitude. Therefore, the proposed CMAPFs enjoy good sensitivity performance.

4 Simulation Results

HSPICE simulation results using 0.18 μm CMOS process parameters have been carried out to validate the proposed circuits of CMAPF. The W/L ratios of MOS transistors are given in Table 2. The minimum length of transistors is selected (just double of the technology used) keeping in view the practical aspects of

integration. The supply voltages of ± 1.25 V are used and biasing voltage used is $V_{BB} = 0.42$ V. The proposed filter is designed for 39.78 MHz frequency. The bias current is fixed to $I_B = 26$ μA for which value of g_m is 0.5 mS. The value of capacitor used is $C = 2$ pF. Fig. 3 shows the simulated phase response along with the magnitude response. A 90° phase shift occurs at pole frequency of $f_0 = 39.2$ MHz (1.47% error) which is very close to the theoretically designed pole frequency. The transient responses of input and output at pole frequency are shown in Fig. 4(a) and their simulated frequency spectrums are depicted in Fig. 4(b). The Lissajous pattern shown in Fig. 4(c) confirms the quadrature relationship between input and output. The output waveform at pole frequency shows a total harmonic distortion (THD) of 0.74%. Additionally, the variation of THD (%) with the amplitude of input current is plotted in Fig. 5. It is to be noted from the Fig. 5 that for an amplitude range up to 50 μA , the THD is not higher than 2%.

Furthermore, the phase responses of output current for different values of bias current ($I_B = 20, 25,$ and 30 μA) are shown in Fig. 6. The pole frequency varying against the bias current I_B is plotted in Fig. 7. The bias current is varied from 10 μA to 40 μA with 5 μA step size and a frequency range of 14 MHz to 56.1 MHz is achieved for this range of bias current. Next, the performance of the proposed circuit is tested for temperature variation. The input and output waveforms at different temperatures (25°C, 50°C, 75°C and 100°C) are shown in Fig. 8. The plot of variations of pole frequency and THD against the temperature is also shown in Fig. 9. As discussed in Section 2, the transconductance, g_m decreases with increase in temperature therefore, the pole frequency also decreases with increase in temperature since pole frequency is directly proportional to g_m as to be seen from Eq. (3). Fig. 9 shows that pole frequency decreases from 39.2 MHz to 33 MHz as temperature is increased from 25°C to 100°C. Moreover, THD of the output current varies from 0.74% to 0.92% when temperature is varied from 25°C to 100°C. It is to be observed that THD is less than 1% for a wide range of temperature variation.

Table 2 W/L ratios of MOS transistors.

MOS Transistors	W(μm)/L(μm)
M ₁ , M ₂	0.72/0.36
M ₃ , M ₄ , M ₅ , M ₂₄ , M ₂₅	1.44/0.36
M ₁₄ , M ₁₅	1.34/0.36
M ₁₆ , M ₁₇ , M ₁₈	2.4/0.36
M ₆ , M ₇ , M ₈ , M ₉ , M ₁₀ , M ₁₁ , M ₁₂ , M ₁₃ , M ₁₉ , M ₂₀	4.8/0.36
M ₂₁ , M ₂₂	3.6/0.36
M ₂₃ , M ₂₆ , M ₂₉ , M ₃₀	2.88/0.36
M ₂₇ , M ₂₈ , M ₃₁ , M ₃₂	0.36/0.36

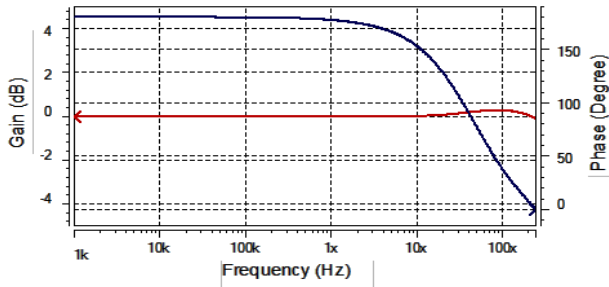


Fig. 3 Phase response and magnitude response.

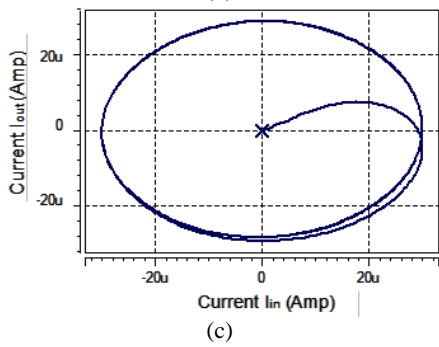
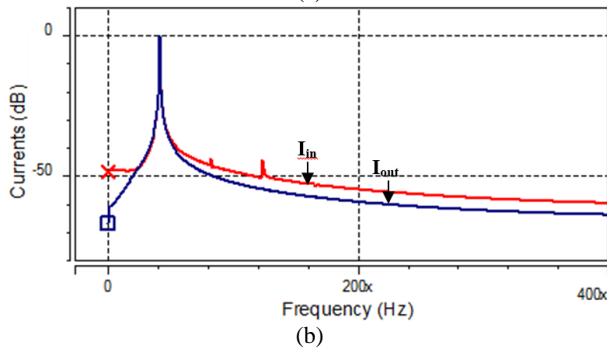
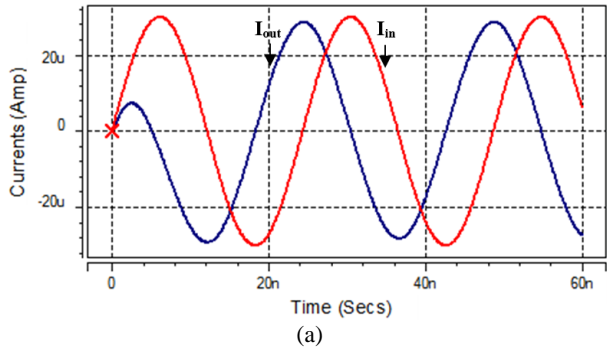


Fig. 4 a) Transient responses of input and output at pole frequency, b) Frequency spectrums and c) Lissajous pattern at pole frequency.

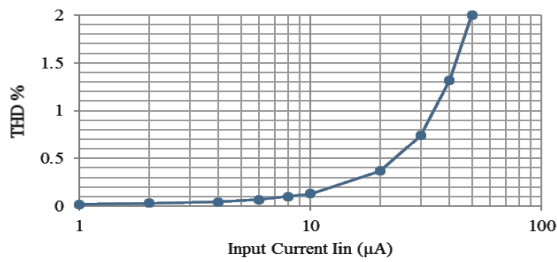


Fig. 5 THD variation against input current I_{in} .

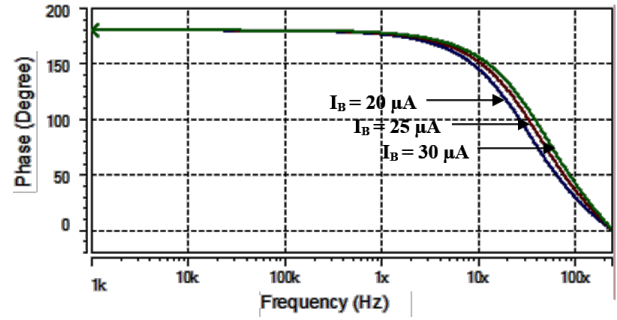


Fig. 6 Phase responses at different bias currents.

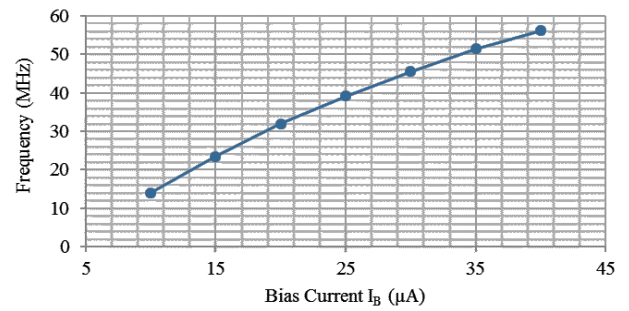


Fig. 7 Pole frequency variation against bias current, I_B .

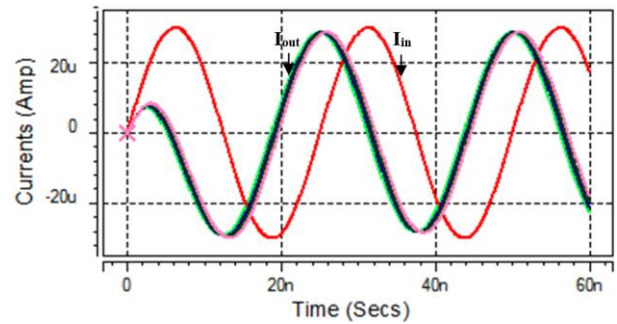


Fig. 8 Input and output waveforms at different temperatures (25°C, 50°C, 75°C and 100°C).

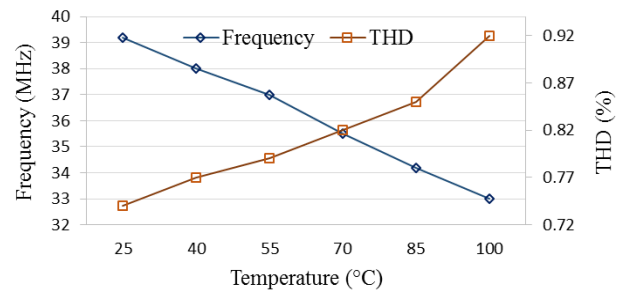


Fig. 9 Pole frequency and THD variations against temperature.

5 Applications of the Proposed CMAPFs

5.1 Second Order CMAPF

To verify the property of cascability of the proposed first order CMAPF-I, a second order CMAPF is realized by cascading the two first order CMAPF-I as shown in Fig. 10. The circuit employs two DXCCTA and two capacitors. The transfer function obtained from the analysis of circuit is given as

$$\frac{I_{out}}{I_{in}} = \left(\frac{sC_1 - g_{m1}}{sC_1 + g_{m1}} \right) \times \left(\frac{sC_2 - g_{m2}}{sC_2 + g_{m2}} \right) \quad (13)$$

The pole frequency and phase angle obtained from Eq. (13) are given as

$$\omega_0 = \left(\frac{g_{m1} \times g_{m2}}{C_1 \times C_2} \right)^{\frac{1}{2}} \quad (14)$$

$$\varphi(\omega) = -2 \tan^{-1} \left(\frac{\omega C_1}{g_{m1}} \right) - 2 \tan^{-1} \left(\frac{\omega C_2}{g_{m2}} \right) \quad (15)$$

It is clear from Eq. (15) that a second order all-pass filter provides a phase shift of 0° to -360° when frequency is varied from 0 to ∞.

The circuit of second order CMAPF is designed for a pole frequency of 39.7 MHz. The bias currents used are $I_{B1} = I_{B2} = 26 \mu A$ for which g_m is 0.5 mS. Two capacitors used are $C_1 = C_2 = 2 \text{ pF}$. Fig. 11 shows the simulated magnitude response and phase response. It is to be noted that -180° phase shift arises at pole frequency of 39.2 MHz (1.47% error) which is almost similar to theoretically designed pole frequency. Additionally the simulated transient responses of input and output at pole frequency and their frequency spectrums are shown in Fig. 12(a) and (b), respectively. It is to be clearly seen from the Fig. 12(a) that input and output have a phase difference of 180° at the pole frequency.

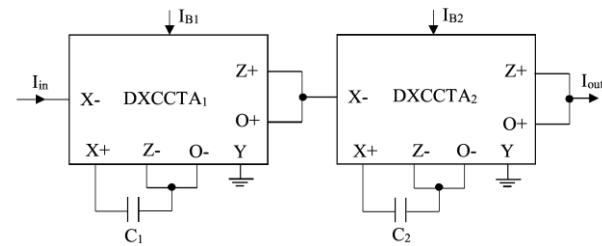


Fig. 10 Circuit of the proposed second order CMAPF.

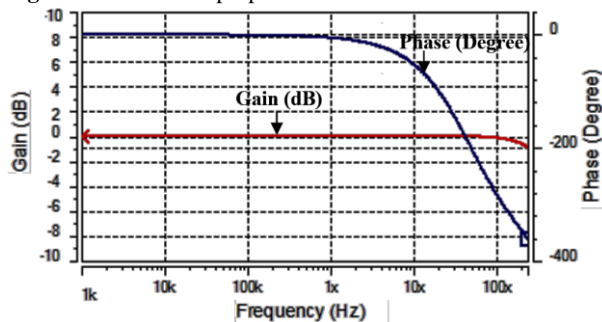


Fig. 11 Simulated phase and magnitude responses.

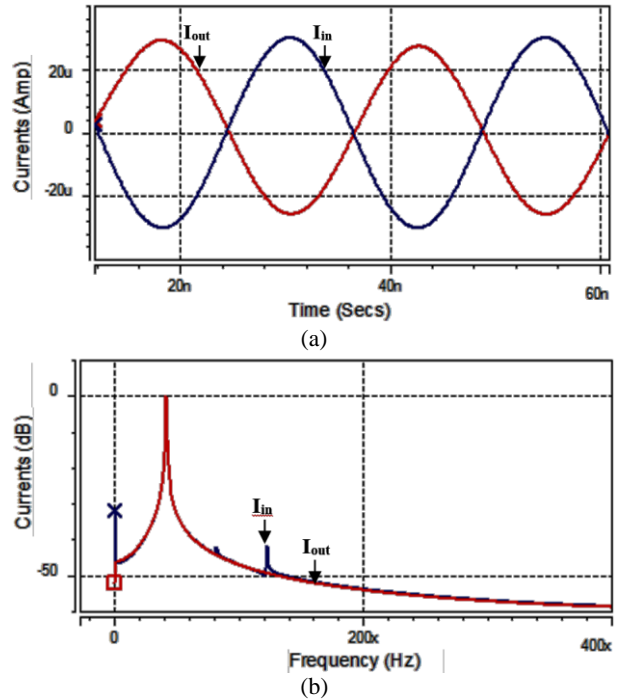


Fig. 12 a) Transient responses of input and output and b) Frequency spectrums.

5.2 Current-Mode Quadrature Oscillator

A common technique to realize an oscillator circuit is to use an APF together with an integrator. The proposed current-mode quadrature oscillator (CM-QO) is also realized using CMAPF-I and an integrator formed with the help of DXCCTA. Fig. 13 shows the circuit of CM-QO. The characteristic equation obtained from the analysis of the circuit can be found as

$$s^2 + s \frac{(g_m RC_2 - C_1)}{RC_1 C_2} + \frac{g_m}{RC_1 C_2} = 0 \quad (16)$$

The oscillation frequency (f_0) and condition of oscillation (CO) from Eq. (16) are found as

$$f_0 = \frac{1}{2\pi} \sqrt{\frac{g_m}{RC_1 C_2}} \quad (17)$$

$$\text{CO: } g_m RC_2 \leq C_1 \quad (18)$$

For equal value of capacitors the CO in Eq. (18) becomes as

$$\text{CO: } g_m R \leq 1 \quad (19)$$

The quadrature current outputs I_1 and I_2 are related as

$$I_1 = j \omega RC_2 I_2 \quad (20)$$

Equation (20) confirms the quadrature relationship between the two current outputs.

The proposed CM-QO is designed for an oscillation

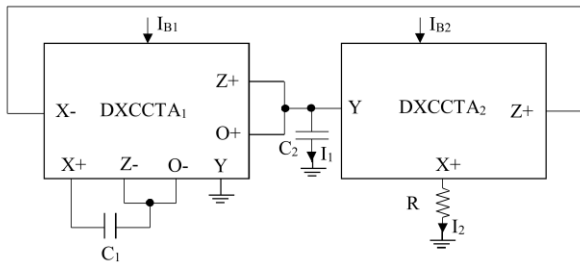


Fig. 13 Circuit of the proposed CM-QO.

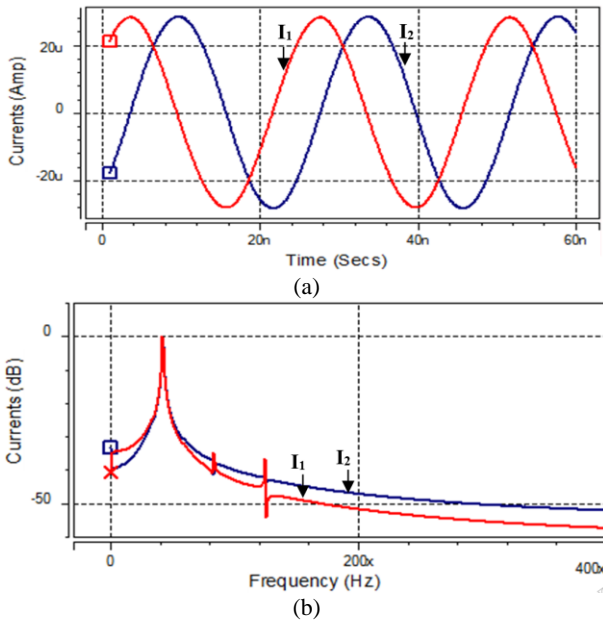


Fig. 14 a) Transient responses of quadrature current outputs and b) Frequency spectrums.

frequency of 42 MHz with bias current $I_{B1} = I_{B2} = 26 \mu\text{A}$. The passive components used are $C_1 = C_2 = 2 \text{ pF}$ and $R = 1.8 \text{ K}\Omega$. Fig. 14(a) and (b) show the transient responses of quadrature current outputs and their frequency spectrums, respectively. The simulated oscillation frequency (41.5 MHz) is found very close to the designed oscillation frequency.

6 Conclusion

Four resistorless circuits of first order CMAPF based on single DXCCTA and one capacitor have been introduced in this paper. The proposed circuits use minimum number of active and passive components and can be cascade easily to realize the higher order of all-pass filter. Moreover, the pole frequency is adjustable through bias current, I_B . The circuits have the good sensitivity performance and they also exhibit low THD up to a reasonable range of input current. Furthermore, the applications of the proposed first order CMAPF in terms of second order CMAPF and CM-QO are presented. All the proposed circuits have good operational frequency. HSPICE simulation results have been depicted to validate the theoretical analyses.

Furthermore, the proposed CMAPF circuits are expected to find applications in single side band suppressed carrier modulation (SSB-SC) circuits and in phase equalizers. The proposed CM-QO can also be expected useful in many applications such as single side band generation, quadrature mixers, selective voltmeters etc. in the areas of communication and measurement systems.

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