

# A Voltage Gain-Controlled Modified CFOA And Its Application in Electronically Tunable Four-Mode All-Pass Filter Design

Norbert Herencsar, Jaroslav Koton, Abhirup Lahiri, Bilgin Metin, and Kamil Vrba

**Abstract**—This paper presents a new active building block (ABB) called voltage gain-controlled modified current feedback amplifier (VGC-MCFOA) based on bipolar junction transistor technology. The versatility of the new ABB is demonstrated in new first-order all-pass filter structure design employing single VGC-MCFOA, single grounded capacitor, and three resistors. Introduced circuit provides all four possible transfer functions at the same configuration, namely current-mode, transimpedance-mode, transadmittance-mode, and voltage-mode. The pole frequency of the circuit can be easily tuned by means of DC bias currents. The theoretical results are verified by SPICE simulations based on bipolar transistor arrays AT&T ALA400-CBIC-R process parameters.

**Keywords**—Voltage gain-controlled modified CFOA, MCFOA, electronically tunable filter, four-mode circuit, all-pass filter.

## I. INTRODUCTION

After the second-generation current conveyor (CCII) was introduced by Sedra and Smith in 1970 [1], it became the most versatile active building block (ABB) used for analog signal processing and the is basic ABB of many other active elements such as the composite current conveyor [2] done by an interconnection of two CCII, which was recently introduced as modified current feedback operational amplifier (MCFOA) [3]–[7] or the conventional CFOA [8] (CCII followed by unity gain voltage buffer - UGVB). It should be noted that, the MCFOA is different from the conventional CFOA defined in [8], since the W terminal current of the MCFOA is copied to the Y terminal in the opposite direction. However, it is well known that the Y-terminal current of the conventional CFOA is equal to zero. Short list of additional CCII-based ABBs is the following: the second-generation current-controlled conveyor (CCCII) [9], where the intrinsic resistance of X-terminal can be tuned, the differential difference CC (DDCC) [10] and its more versatile derivative the so-called universal current conveyor (UCC) [11]–[14], the dual-X CCII (DXCCII) [15], which is an interconnection of CCII and inverting CCII

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in which the Y-terminal is joined, the current differencing buffered amplifier (CDBA) [16] employing current differencing unit (CDU) based on two CCII and UGVB, or the universal voltage conveyor (UVC) [17] based on two CCII and differential UGVB.

Recently the further research has been focused on CCII-based ABBs employing operational transconductance amplifier (OTA) [18] at their output stage. Probably the most known active element from this group is the current differencing transconductance amplifier (CDTA) [19], but other versatile elements such as the current-conveyor transconductance amplifier (CCTA) [20], where CCII is followed by an OTA, the differential-input buffered and transconductance amplifier (DBTA) [21] in which an interconnection of two CCII are followed by UGVB and OTA, the current follower transconductance amplifier (CFTA) [22], which employs a CCII with grounded Y-terminal and an OTA, the current backward transconductance amplifier (CBTA) [23], which is a specific interconnection of CCII and OTA, or the  $z$ -copy current-controlled current inverting transconductance amplifier (ZC-CCCITA) [24] have also received considerable attention.

For easy tunability of the circuit parameters using electrical signals by either voltage and/or current and to increase the universality of the conventional CCII, the electronically-tunable CCII (ECCII) [25], [26], programmable current amplifier (PCA) [27], K-gain CCII [28], the voltage and current gain CCII (VCG-CCII) [29], and the variable gain current conveyor (VGCCII) [30] were introduced.

In this paper we present a novel ABB called voltage gain-controlled modified current-feedback operational amplifier (VGC-MCFOA). The VGC-MCFOA joins the voltage gain control feature of the VCG-CCII in the conventional MCFOA [3]–[7]. To demonstrate the usefulness of the VGC-MCFOA, a new first-order all-pass filter (AFP) structure is proposed, which operates in current-mode (CM), transimpedance-mode (TIM), transadmittance-mode (TAM), and voltage-mode (VM), respectively. To prove the theoretical analysis, SPICE simulations based on bipolar transistor arrays AT&T ALA400-CBIC-R process parameters are given.

## II. CIRCUIT DESCRIPTION

The voltage gain-controlled modified current feedback operational amplifier (VGC-MCFOA) is a five-terminal ABB and its circuit symbol is shown in Fig. 1(a). Compared to the conventional MCFOA presented in [3]–[7], its voltage transfer

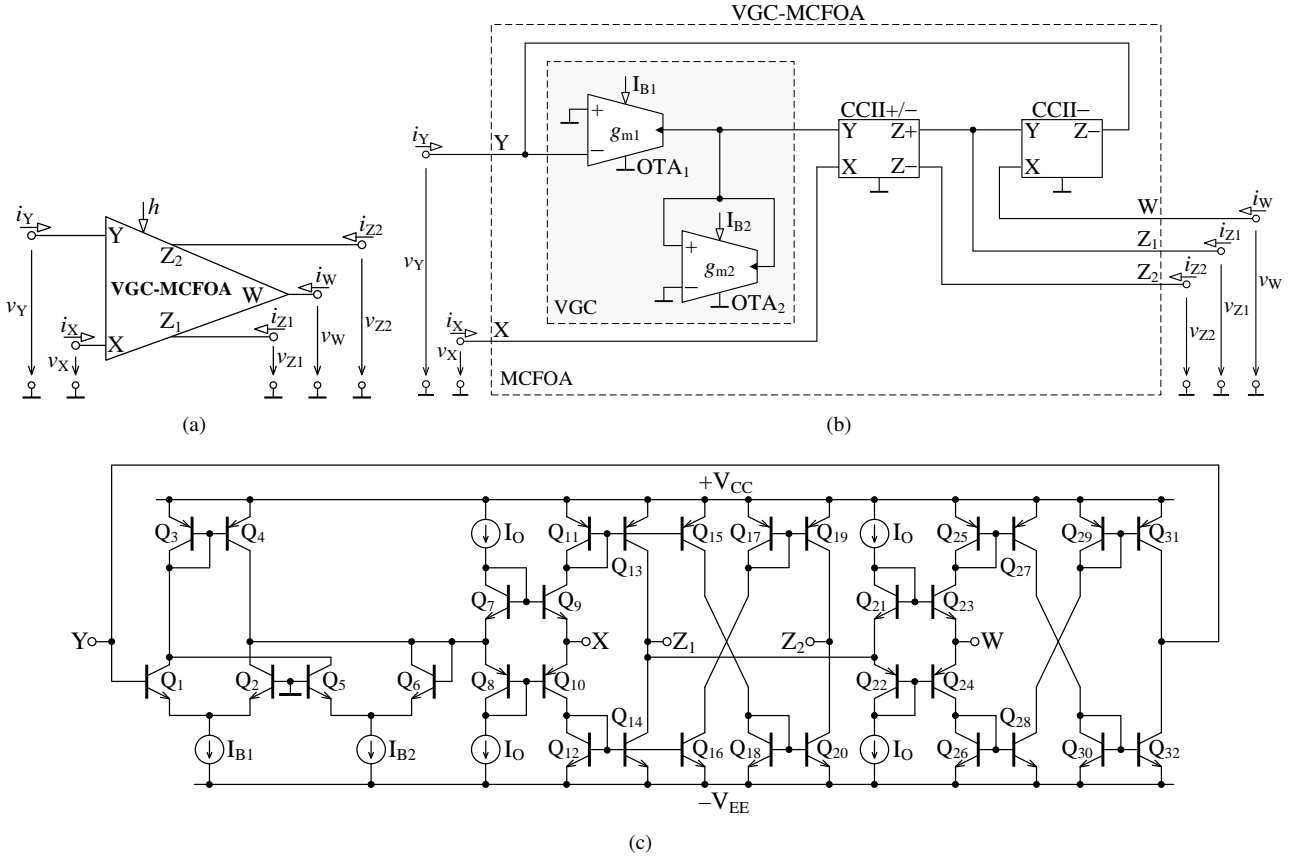


Fig. 1. Bipolar implementation of VGC-MCFOA

from the Y to X terminal can be easily electronically tuned by means of the voltage gain  $h$ . Hence, the relations between the individual terminals of the VGC-MCFOA can be described by the following hybrid matrix:

$$\begin{bmatrix} i_Y \\ v_X \\ i_{Z1} \\ i_{Z2} \\ v_W \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & -\alpha_1 \\ h\beta_1 & 0 & 0 & 0 & 0 \\ 0 & \alpha_2 & 0 & 0 & 0 \\ 0 & -\alpha_3 & 0 & 0 & 0 \\ 0 & 0 & \beta_2 & 0 & 0 \end{bmatrix} \begin{bmatrix} v_Y \\ i_X \\ v_{Z1} \\ v_{Z2} \\ i_W \end{bmatrix}. \quad (1)$$

The frequency-dependent non-ideal current gains  $\alpha_j$  for  $j = \{1, 2, 3\}$  and voltage gains  $\beta_k$  for  $k = \{1, 2\}$  are ideally equal to unity. Using a single-pole model [4], they can be defined as:

$$\alpha_j(s) = \frac{\alpha_{oj}}{1 + s\tau_{\alpha_j}}, \quad (2)$$

$$\beta_k(s) = \frac{\beta_{ok}}{1 + s\tau_{\beta_k}}, \quad (3)$$

where  $\alpha_{oj}$  and  $\beta_{ok}$  are DC current and voltage gains of the element, respectively. The bandwidths  $1/\tau_{\alpha_j}$  and  $1/\tau_{\beta_k}$  on the order of a few gigad/s in current technologies are ideally equal to infinity. At low and medium frequencies i.e.,  $f \ll (1/(2\pi)) \times \min\{1/\tau_{\alpha_j}, 1/\tau_{\beta_k}\}$ , Eqs. (2) and (3) turn to:

$$\alpha_j(s) \cong \alpha_{oj} = 1 - \varepsilon_{\alpha_{ij}}, \quad (4)$$

$$\beta_k(s) \cong \beta_{ok} = 1 - \varepsilon_{\beta_{vk}}, \quad (5)$$

where  $\varepsilon_{\alpha_{ij}}$  and  $\varepsilon_{\beta_{vk}}$  are the current and voltage tracking errors, whereas  $|\varepsilon_{\alpha_{ij}}| \ll 1$  and  $|\varepsilon_{\beta_{vk}}| \ll 1$ , respectively.

The basic idea for implementation of the proposed VGC-MCFOA is shown in Fig. 1(b), where the  $OTA_1$  and  $OTA_2$  are used to control the voltage gain  $h$  and two  $CCII+/-$  represent the conventional MCFOA. Subsequently, the bipolar implementation of the VGC-MCFOA is shown in Fig. 1(c). The voltage gain control stage is formed by two simple differential pair amplifiers (transistors  $Q_1-Q_6$ ) and transistors  $Q_7-Q_{32}$  form the two  $CCII+/-$  based MCFOA, respectively. Here it is worth mentioning that the voltage gain control of VGC-CCII [29] was implemented using the same technique. For the implementation in Fig. 1(b) the voltage gain  $h$  can be expressed as:

$$h = \frac{g_{m_{1,2}}}{g_{m_{5,6}}}, \quad (6)$$

where  $g_{m_{1,2}} = \frac{I_{B1}}{2V_T}$  and  $g_{m_{5,6}} = \frac{I_{B2}}{2V_T}$ . Here, the  $V_T$  is the thermal voltage (approximately 26 mV at 27°C) and the  $I_{B1}$  and  $I_{B2}$  are control currents adjusting the transconductances  $g_{m_{1,2}}$  and  $g_{m_{5,6}}$ , respectively. Therefore, the voltage gain  $h$  in (6) can be given as:

$$h = \frac{I_{B1}}{I_{B2}}. \quad (7)$$

From (7) it is obvious that the proposed VGC-MCFOA can be easily adjusted electronically by either  $I_{B1}$  and/or  $I_{B2}$  currents.

### III. PROPOSED ALL-PASS FILTER

#### A. Ideal Case Study

The proposed four-mode APF is shown in Fig. 2. Considering the ideal VGC-MCFOA (i.e.  $\alpha_j$  and  $\beta_k$  are unity), based on the input selected two following cases can be considered:

*Case I:* If  $I_{in1} = I_{in2} = I_{in}$ ,  $V_{in} = 0$  (grounded), and assuming  $R_2 = R_3 = R$ , then we can obtain the following transfer functions (TFs):

$$T_{CM}(s) = \frac{I_{out}}{I_{in}} = \frac{sCR_1 - h}{sCR_1 + h} = \frac{I_{B2}sCR_1 - I_{B1}}{I_{B2}sCR_1 + I_{B1}}, \quad (8)$$

$$T_{TIM}(s) = \frac{V_{out}}{I_{in}} = R \cdot \frac{sCR_1 - h}{sCR_1 + h} = R \cdot \frac{I_{B2}sCR_1 - I_{B1}}{I_{B2}sCR_1 + I_{B1}}. \quad (9)$$

*Case II:* If the input of the APF is  $V_{in}$ ,  $I_{in1} = I_{in2} = 0$ , and assuming  $R_1 = R_2 = R$ , then for the circuit the following TFs can be obtained:

$$T_{TAM}(s) = \frac{I_{out}}{V_{in}} = -\frac{1}{R} \cdot \frac{sCR_3 - h}{sCR_3 + h} = -\frac{1}{R} \cdot \frac{I_{B2}sCR_3 - I_{B1}}{I_{B2}sCR_3 + I_{B1}}, \quad (10)$$

$$T_{VM}(s) = \frac{V_{out}}{V_{in}} = -\frac{sCR_3 - h}{sCR_3 + h} = -\frac{I_{B2}sCR_3 - I_{B1}}{I_{B2}sCR_3 + I_{B1}}. \quad (11)$$

Thus, from Eqs. (8)–(11) it is seen that by suitable selection of input and output all four possible modes, i.e. current-, transimpedance-, transadmittance-, and voltage-mode first-order APF can be realized with the same circuit topology.

The phase responses of TFs in (8) and (9) are calculated as follows:

$$\begin{aligned} \varphi_{CM}(\omega) = \varphi_{TIM}(\omega) &= 180^\circ - 2\arctg\left(\frac{1}{h} \cdot \omega CR_1\right) = \\ &= 180^\circ - 2\arctg\left(\frac{I_{B2}}{I_{B1}} \cdot \omega CR_1\right), \end{aligned} \quad (12)$$

and phase responses of TFs in (10) and (11) are given as:

$$\begin{aligned} \varphi_{TAM}(\omega) = \varphi_{VM}(\omega) &= -2\arctg\left(\frac{1}{h} \cdot \omega CR_3\right) = \\ &= -2\arctg\left(\frac{I_{B2}}{I_{B1}} \cdot \omega CR_3\right). \end{aligned} \quad (13)$$

Hence, the phases of TFs in (8) and (9) alter from  $180^\circ$  to  $0^\circ$  while according to (10) and (11) the phase shift change between  $0^\circ$  to  $-180^\circ$ , respectively.

Consequently, the zero ( $\omega_z$ ) and pole ( $\omega_p$ ) frequencies of all four TFs can be found as:

$$\omega_{(CM,TIM)z} = \omega_{(CM,TIM)p} = h \cdot \frac{1}{CR_1} = \frac{I_{B1}}{I_{B2}} \cdot \frac{1}{CR_1}, \quad (14)$$

$$\omega_{(TAM,VM)z} = \omega_{(TAM,VM)p} = h \cdot \frac{1}{CR_3} = \frac{I_{B1}}{I_{B2}} \cdot \frac{1}{CR_3}. \quad (15)$$

From Eqs. (14) and (15) it is clearly seen that the pole/zero frequency values can be easily tuned by means of the bias currents  $I_{B1}$  and/or  $I_{B2}$ .

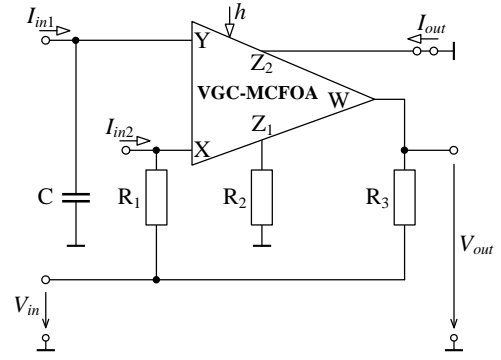


Fig. 2. Proposed electronically tunable all-pass filter using VGC-MCFOA

#### B. Non-Ideal Analysis

Taking into account non-idealities of the VGC-MCFOA, TFs (8) and (9) in *Case I* of the filter convert to:

$$\begin{aligned} T_{CM}(s) &= \frac{I_{out}}{I_{in}} = \alpha_3 R_3 \cdot \frac{sCR_1 - \beta_1 h}{sCR_1 R_3 + \alpha_1 \alpha_2 \beta_1 \beta_2 h R_2} = \\ &= \alpha_3 R_3 \cdot \frac{I_{B2}sCR_1 - I_{B1}\beta_1}{I_{B2}sCR_1 R_3 + I_{B1}\alpha_1 \alpha_2 \beta_1 \beta_2 R_2}, \end{aligned} \quad (16)$$

$$\begin{aligned} T_{TIM}(s) &= \frac{V_{out}}{I_{in}} = \frac{\alpha_2 \beta_2 R_2 R_3 \cdot (sCR_1 - h\beta_1)}{sCR_1 R_3 + \alpha_1 \alpha_2 \beta_1 \beta_2 h R_2} = \\ &= \frac{\alpha_2 \beta_2 R_2 R_3 \cdot (I_{B2}sCR_1 - I_{B1}\beta_1)}{I_{B2}sCR_1 R_3 + I_{B1}\alpha_1 \alpha_2 \beta_1 \beta_2 R_2}, \end{aligned} \quad (17)$$

and non-ideal phase responses from TFs (16) and (17) can be expressed as:

$$\begin{aligned} \varphi_{CM}(\omega) = \varphi_{TIM}(\omega) &= 180^\circ - \arctg\left(\frac{I_{B2}}{I_{B1}} \cdot \frac{\omega CR_1}{\beta_1}\right) - \\ &- \arctg\left(\frac{I_{B2}}{I_{B1}} \cdot \frac{\omega CR_1 R_3}{\alpha_1 \alpha_2 \beta_1 \beta_2 R_2}\right). \end{aligned} \quad (18)$$

The zero and pole frequencies in Eq. (14) change to:

$$\omega_{(CM,TIM)z} = \frac{I_{B1}}{I_{B2}} \cdot \frac{\beta_1}{CR_1}, \quad \omega_{(CM,TIM)p} = \frac{I_{B1}}{I_{B2}} \cdot \frac{\alpha_1 \alpha_2 \beta_1 \beta_2 R_2}{CR_1 R_3}. \quad (19)$$

From Eq. (19), the active and passive sensitivities of zero and pole frequencies are given as:

$$S_{I_{B1}, \beta_1}^{\omega_{(CM,TIM)z}} = -S_{I_{B2}, C, R_1}^{\omega_{(CM,TIM)z}} = 1, \quad S_{\alpha_1, \alpha_2, \alpha_3, \beta_2, R_2, R_3}^{\omega_{(CM,TIM)z}} = 0, \quad (20)$$

$$S_{I_{B1}, \alpha_1, \alpha_2, \beta_1, \beta_2, R_2}^{\omega_{(CM,TIM)p}} = -S_{I_{B2}, C, R_1, R_3}^{\omega_{(CM,TIM)p}} = 1, \quad S_{\alpha_3}^{\omega_{(CM,TIM)p}} = 0, \quad (21)$$

and it is evident that the sensitivities of active parameters and passive components for  $\omega_{(CM,TIM)z}$  and  $\omega_{(CM,TIM)p}$  are at maximum unity in relative amplitude. The same study can also be done for the *Case II* with similar results.

### IV. SIMULATION RESULTS

First, the proposed VGC-MCFOA in Fig. 1(c) has been further investigated in SPICE software. In the design the transistor model parameters NR100N (NPN) and PR100N (PNP) of bipolar arrays ALA400-CBIC-R from AT&T were used

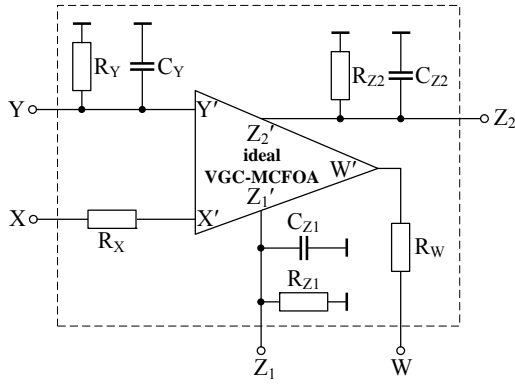


Fig. 3. Model of the VGC-MCFOA including parasitic elements

TABLE I  
PARAMETERS OF VGC-MCFOA SHOWN IN FIG. 1(C) (NOTE: # at  $h = 1$ )

Parameters	Values
$i_Y/i_W$ gain ( $\alpha_1$ )	1.001
$v_X/v_Y$ gain ( $\beta_1$ )#	0.995
$i_{Z1}/i_X$ gain ( $\alpha_2$ )	0.982
$i_{Z2}/i_X$ gain ( $\alpha_3$ )	1.001
$v_W/v_{Z1}$ gain ( $\beta_2$ )	0.999
$i_Y/i_W$ $f_{-3dB}$	49.16 MHz
$v_X/v_Y$ $f_{-3dB}$ #	108.15 MHz
$i_{Z1}/i_X$ $f_{-3dB}$	89.22 MHz
$i_{Z2}/i_X$ $f_{-3dB}$	67.34 MHz
$v_W/v_{Z1}$ $f_{-3dB}$	879.47 MHz
$R_Y$	55.35 k $\Omega$
$C_Y$	1.919 pF
$R_X$	42.01 $\Omega$
$R_W$	36.96 $\Omega$
$R_{Z1}$	94.31 k $\Omega$
$C_{Z1}$	2.047 pF
$R_{Z2}$	97.61 k $\Omega$
$C_{Z2}$	1.272 pF

[31]. The DC supply voltages are  $+V_{CC} = -V_{EE} = 2.5$  V. Bias current  $I_O = 400 \mu\text{A}$  has been chosen and  $I_{B1}$ ,  $I_{B2}$  were set to  $101 \mu\text{A}$  and  $100 \mu\text{A}$ , respectively, to obtain voltage gain  $h = 1$  precisely. The maximum values of terminal voltages and terminal currents without producing significant distortion were determined to be  $\pm 106.7$  mV and  $\pm 16.23$  mA, respectively. Evaluated DC current and voltage gains,  $f_{-3dB}$  frequencies of transfers, and values of the X and W terminal parasitic resistances (in series) and Z and Y terminal parasitic resistances and capacitances (in parallel) shown in Fig. 3 are given in Table I. The total power dissipation of the proposed VGC-MCFOA was found to be 23.1 mW.

Simulated voltage gain  $h$  responses between Y and X terminals is demonstrated in Fig. 4. In case of (A), the external bias current  $I_{B1}$  has been varied in large interval from  $10 \mu\text{A}$  to  $1$  mA (equal to gain  $h = 0.1$  to  $10$ ) at constant  $I_{B2} = 100 \mu\text{A}$ . From Fig. 4 it can be clearly seen that due to the above mentioned non-idealities of the VGC-MCFOA, the obtained voltage gain is in reduced range  $0.101 \div 8.71$ . Hence, to overcome the the large variation of control current  $I_{B1}$  and simultaneously obtain the same gain range i.e.  $h = 0.1$  to  $10$ , the control current  $I_{B1}$  has been varied in reduced interval

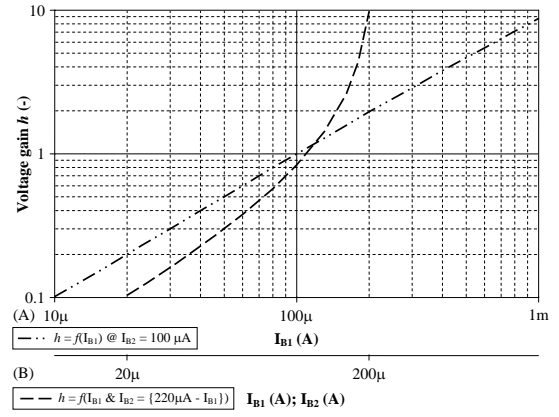


Fig. 4. Voltage gain  $h$  responses vs. applied bias currents

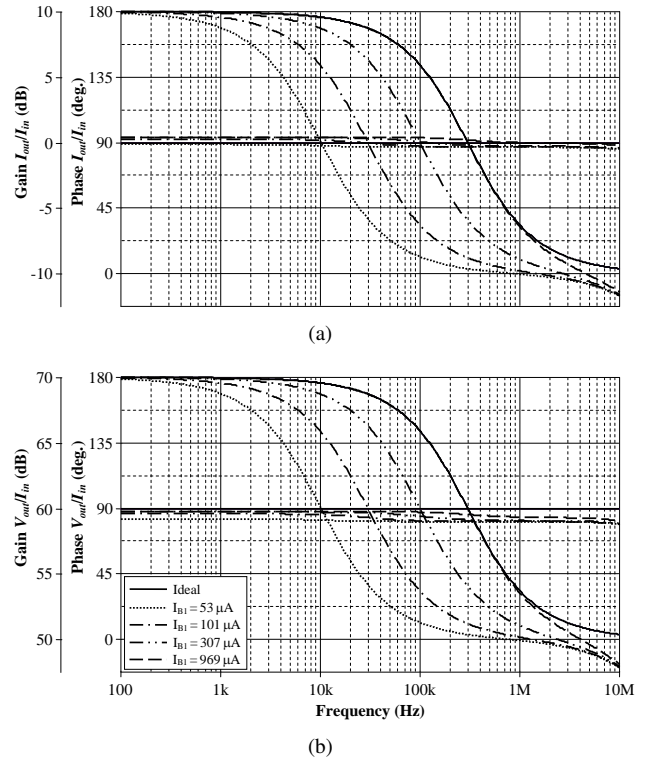


Fig. 5. Electronical tunability of the pole frequency by bias current  $I_{B1}$  based on case (A) at constant  $I_{B2} = 100 \mu\text{A}$ : (a) current-mode, (b) transimpedance-mode first-order APF responses

from  $20 \mu\text{A}$  to  $200 \mu\text{A}$  together with  $I_{B2}$  according to Eq.  $\{220 \mu\text{A} - I_{B1}\}$ . In this case (B), the obtained voltage gain is in range  $0.102 \div 9.869$ , which is much closer to the theoretical one.

Using the bipolar implementation of the VGC-MCFOA the proposed APF from Fig. 2 has also been simulated in the SPICE software. The ideal and simulated gain and phase responses and electronical tunability of both current- and transimpedance-mode transfers based on case (A) discussed above i.e. by the bias current  $I_{B1}$  at constant  $I_{B2} = 100 \mu\text{A}$ , are demonstrated in Fig. 5. In the simulations the passive element values were selected as  $C = 5$  nF,  $R_1 = R_2 = R_3 = 1$  k $\Omega$  and the voltage gain  $h$  has been varied as  $h = \{0.53; 1;$

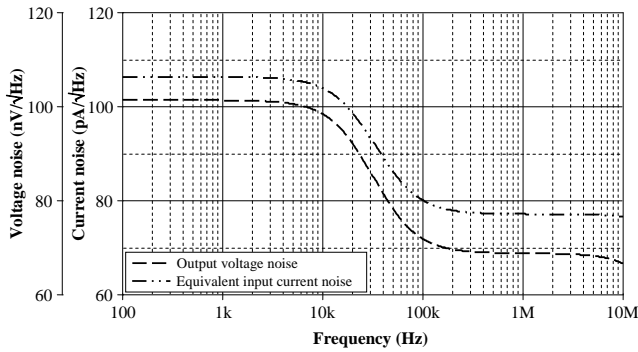


Fig. 6. Output and equivalent input noise variations vs. frequency at  $h = 1$  ( $f_0 = 29.9$  kHz) for the transimpedance-mode first-order APF response

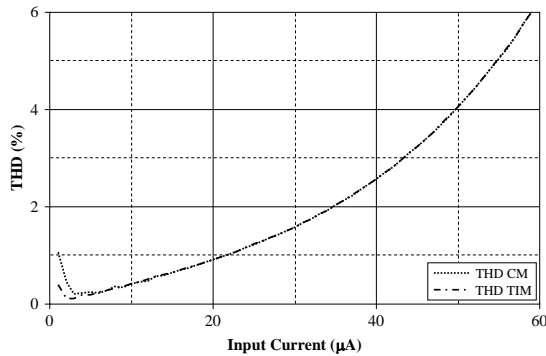


Fig. 7. THD variation of the proposed APF for both current- and transimpedance-mode responses against applied input current at  $f_0 = 29.9$  kHz

2.95; 8.46} to set the pole frequency of the proposed circuit as  $f_0 \approx \{10.2; 29.9; 100; 300\}$  kHz. Using the INOISE and ONOISE statements, the input and output noise behavior with respect to frequency has also been simulated, as it is shown in Fig. 6. The output noise and equivalent input noise at pole frequency ( $f_0 \approx 29.9$  kHz) were found as  $86.13 \text{ nV}/\sqrt{\text{Hz}}$  and  $93.38 \text{ pA}/\sqrt{\text{Hz}}$ , respectively. The THD variation for both responses with respect to amplitude of the applied sinusoidal input current at pole frequency of 29.9 kHz (filter parameter:  $I_{B1} = 101 \mu\text{A}$ ) is shown in Fig. 7. An input with the amplitude of  $30 \mu\text{A}$  yields for both responses THD values of 1.59%. From the simulations it is evident that the gain and phase characteristics of the filter are in good agreement with theory and the deviations are caused by the non-idealities of the active element used.

## V. CONCLUSION

In this paper, a novel ABB called voltage gain-controlled MCFOA, which joins the voltage gain control feature of the voltage and current gain CCII in the conventional MCFOA. The usefulness of the tunable feature in the introduced VGC-MCFOA is demonstrated in four-mode first-order all-pass filter design. Since the capacitor in the circuit is grounded, the proposed filter is attractive for integration. The pole frequency can successfully be tuned in wide frequency range by means of external bias currents. The SPICE simulations confirm the theoretical assumptions.

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