Single-CDTA (Current Differencing Transconductance Amplifier) Current-Mode Biquad Revisited

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Abstract: This paper presents a current-mode (CM) second-order filter, employing one so-called "Z Copy – Current Differencing Transconductance Amplifier" (ZC-CDTA), two grounded capacitors, and one virtually grounded resistor. The filter provides two current outputs of low-pass (LP) and band-pass (BP) types for driving independent loads, and one high-pass (HP) current output, flowing through one of the working capacitors. Other implementations of various types of transfer functions are proposed by means of commonly used methods of input/output current summing. Their critical analysis is accomplished from the point of view of tracking errors and dynamic range optimization. Results of the PSpice simulation, which utilizes the CDTA model on the level of CMOS transistor structure, are enclosed.

Keywords: - CDTA, PSpice, Current-mode filters.

1 Introduction

A number of applications of the current differencing transconductance amplifier (CDTA) [1] can be found in the literature. Its input-to-output behavior predetermines this circuit component particularly for current-mode signal processing. The difference of input currents of the low-impedance *p* and *n* terminals flows out of the *z* terminal as current I_z. The voltage between the *z* terminal and the ground is transformed via a multiple-output operational transconductance amplifier (OTA) into a set of currents I_x. Fig. 1 shows the CDTA+- version [1] with a pair of bi-directional currents $+ I_x$ and $– I_x$.

The application potential of the CDTA can be increased by extending the circuit with an auxiliary *zc* (Z Copy) terminal, which provides a copy of current I_z (see Fig. 1). Such a CDTA is called ZC-CDTA (Z Copy - CDTA) [2].

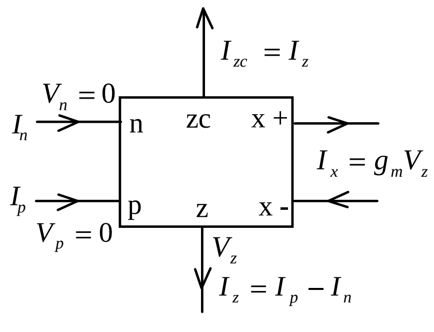

Fig. 1: Active CDTA element containing the "zc" terminal, which provides a copy of the current I_z.

A number of papers deal with the design of 2nd order CDTA-based filters. Most of them contain five [3], four [4], [5], three [6], [7], or two [1], [8-11] CDTAs. A biquad, employing a single CDTA, has been published in [12]. It provides the BP and HP outputs or LP and BP outputs after the RC:CR transformation. However, an analysis shows that the filter consists of a passive 1st order current divider and an active 1st order CDTA-based filter. That is why the transfer functions contain only two real poles, and the quality factor cannot exceed the value 0.5. Another drawback of the filter providing the HP and BP outputs consists in utilizing the floating capacitor. Together with the parasitic resistance of the p-terminal of CDTA, it forms an RC cell which degrades the filter behavior in the high-frequency region.

Single-CDTA biquads employing two grounded capacitors and one resistor, virtually grounded through the low impedance CDTA terminal, have been proposed in [1]. However, they can only serve as low-pass and band-pass filters, and the output currents flow via the working resistances. In [13], the circuit from [1] is complemented with one resistor, namely R_3, and such a structure is described as a universal current-mode biquad, providing four basic transfer functions (low-pass, band-pass, high-pass, and band-reject) simultaneously. However, this circuit labors with the following drawbacks: 1) Only the output current of the low-pass section flows to an independent load. All the remaining output currents flow through the working impedances and thus they cannot be directly utilized without negative influence on the filter behavior. 2) The additional resistor R_3 does not represent any increase of the filter versatility and thus it can be removed from the filter structure without decreasing its functionality.

In this paper, a current-mode filter, starting from the topologies in [1] and [13], and consisting of a single ZC-CDTA, two grounded capacitors and one resistor, which is virtually grounded through the low-impedance CDTA input, is described. No Q-limitation from [12] exists for this filter. The biquad provides simultaneously a pair of LP and BP current outputs into independent loads and an HP current output to the working capacitor. Potential HP and BR (band-reject) outputs to independent loads are also discussed from the points of view of the precision of implemented transfer function [14] and the upper bound of the filter dynamic range.

2 Circuit description

The proposed single-ZC-CDTA biquad is shown in Fig. 2. Considering the ideal model of the CDTA, the output currents I_{o1}, I_{o2} and I_c of the filter are as follows:

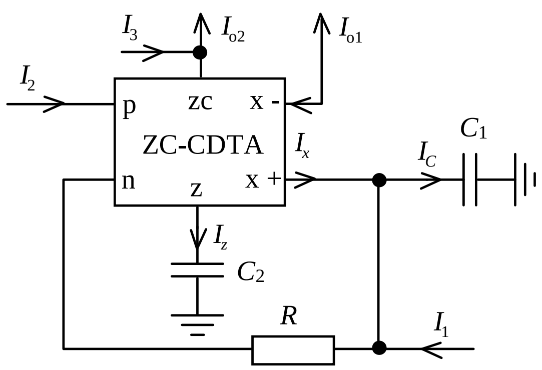

Fig. 2: The CM biquad.

$$I_{o1} = \frac{b_0 I_1 - (b_0 + as) I_2}{D(s)},$$
$$I_{o2} = \frac{-b_1 s I_1 + (b_1 s + s^2) I_2}{D(s)} + I_3, \qquad (1\text{ a, b, c})$$
$$I_c = \frac{s^2 I_1 + as I_2}{D(s)},$$

where

$$D(s) = s^2 + b_1 s + b_0, \qquad (2)$$

$$b_1 = \frac{1}{RC_1},\ b_0 = \frac{g_m}{RC_1 C_2},\ a = \frac{g_m}{C_2} \qquad (3)$$

Natural frequency ω_0, quality factor Q, and bandwidth $B = \omega_0/Q$ are given by

$$\omega_o = \sqrt{\frac{g_m}{RC_1 C_2}},\ Q = \sqrt{g_m R \frac{C_1}{C_2}},\ B = \frac{1}{RC_1}. \qquad (4\text{ a, b, c})$$

The sensitivities of ω_o and Q to variations in g_m, R, C_1, and C_2 are $\pm 1/2$, the sensitivities of B to variations in R and C_1 are -1. As obvious from Eqs. (1), the circuit in Fig. 2 can provide several transfer functions, depending on the type of input current excitation I_{in}:

a) $I_{in} = I_1$, $I_2 = I_3 = 0$ (5)

$$I_{o1} = I_{LP} = \frac{b_0}{D(s)} I_{in},\ I_{o2} = I_{BP} = \frac{-b_1 s}{D(s)} I_{in},$$
$$I_c = I_{HP} = \frac{s^2}{D(s)} I_{in}. \qquad (6\text{ a, b, c})$$

Alternatively,

$$I_{HP} = I_{in} + I_{o2} - I_{o1} = I_{in} + I_z + I_x. \qquad (7)$$

b) $I_{in} = I_1 = I_2$, $I_3 = 0$ (8)

$$I_{o1} = I_{BP} = -\frac{as}{D(s)} I_{in},\ I_{o2} = I_{HP} = \frac{s^2}{D(s)} I_{in},$$
$$I_c = I_{BP+HP} = \frac{as + s^2}{D(s)} I_{in} \qquad (9\text{ a, b, c})$$

c) $I_{in} = I_1 = I_3$, $I_2 = 0$ (10)

$$I_{o1} = I_{LP} = \frac{b_0}{D(s)} I_{in},\ I_{o2} = I_{BR} = \frac{s^2 + b_0}{D(s)} I_{in},$$
$$I_c = I_{HP} = \frac{s^2}{D(s)} I_{in} \qquad (11\text{ a, b, c})$$

Alternatively,

$$I_{HP} = I_{o2} - I_{o1} \qquad (12)$$
$$I_{BP} = I_{in} - I_{o2} \qquad (13)$$

d) $I_{in} = I_1/2 = I_3$, $I_2 = 0$ (14)

$$I_{o1} = I_{LP} = \frac{2b_0}{D(s)} I_{in}, \; I_{o2} = I_{AP} = \frac{s^2 - b_1 s + b_0}{D(s)} I_{in},$$
$$I_c = I_{HP} = \frac{2s^2}{D(s)} I_{in} \tag{15 a, b, c}$$

Note that the p terminal of the CDTA is not utilized in cases a), c), and d). In case b), the excitation current I_2, flowing into the p terminal, can be substituted by an opposite current flowing out of the n terminal. Then the p terminal can be omitted. This fact can be used to simplify the input circuitry of the CDTA for this concrete application.

Case a) represents a single-input multiple-output filter, in which the low-pass and band-pass outputs are provided simultaneously to independent loads. The high-pass output current flows through capacitor C_1 and thus it cannot be practically utilized. Alternatively, a copy of this current can be obtained by combining other currents according to (7).

The cases b, c), and d) represent double-input multiple-output filters with concrete matching conditions between the input currents.

The above combinations of input excitations (8), (10), (14) and the summing of output currents (7), (12), (13) in order to obtain several types of transfer functions are frequently used in the literature [8], [15], [16]. However, the corresponding error analysis of the consequences of matching errors has not been published. Also, the problems of dynamic range optimization are often disregarded. That is why the following sections describe an error analysis from this point of view.

3 Matching error analysis

Eqs. (7) and (5) show that HP output current can be obtained by linear combination of the input current and currents $I_x = -I_{o1}$ and $I_z = I_{o2}$. Circuit implementation consists in adding/subtracting copies of these currents. These copies are normally produced with certain matching errors. As a first approximation, one can regard these errors as frequency independent and being in accordance with the following equations:

$$I'_{HP} = \beta_{in} I_{in} + \beta_z I_{o2} - \beta_x I_{o1}, \tag{16}$$

where

$$\beta_{in} = 1 - \varepsilon_{in}, \; \beta_z = 1 - \varepsilon_z, \; \beta_x = 1 - \varepsilon_x, \tag{17}$$

and ε_{in}, ε_z, and ε_x are tracking errors, which are zero in the ideal case.

Substitution of (1) and (5) into (16) and a short arrangement yield

$$I'_{HP} = \beta_{in} \frac{s^2 + b'_1 s + b'_0}{D(s)}, \tag{18}$$

where

$$b'_1 = b_1 (1 - \frac{\beta_z}{\beta_{in}}), \; b'_0 = b_0 (1 - \frac{\beta_x}{\beta_{in}}). \tag{19 a, b}$$

It is obvious from (18) that tracking errors cause an additional transfer zeros in the HP transfer function. These zeros can be of a different character, depending on the mutual relationship between the tracking errors. In addition, the β_{in} parameter affects the filter gain.

In the first instance, consider the case of $\beta_x < \beta_{in}$ and $\beta_z < \beta_{in}$. The natural frequency ω_z and the quality factor Q_z of the parasitic zero points are as follows:

$$\omega_z = \omega_0 \sqrt{1 - \frac{\beta_x}{\beta_{in}}}, \; Q_z = Q_0 \frac{\sqrt{1 - \frac{\beta_x}{\beta_{in}}}}{1 - \frac{\beta_z}{\beta_{in}}}. \tag{20 a, b}$$

If $\beta_z < \beta_{in}$ (or $\beta_z > \beta_{in}$), the real part of the zero points is negative (or positive).

Independently of the above conditions, if the following condition is true

$$\left(1 - \frac{\beta_z}{\beta_{in}}\right)^2 > 4 \frac{Q^2}{\omega_0} \left(1 - \frac{\beta_x}{\beta_{in}}\right), \tag{21}$$

then the parasitic zero points are complex, otherwise they are real.

The character of transfer zero, which depends on the random relationships between the values β_{in}, β_x, and β_z, determines the course of frequency response in the stop-band and transition regions: a monotonic gain increase with increased frequency for real zeros, and a transmission notch for complex zeros. In any case, the low-frequency gain K(0) and high-frequency gain K(∞) are as follows:

$$K(0) = \beta_{in} - \beta_x, \; K(\infty) = \beta_{in}, \tag{22}$$

and their ratio

$$\frac{K(0)}{K(\infty)} = 1 - \frac{\beta_x}{\beta_{in}}. \tag{23}$$

For the current CMOS technologies, the tracking errors are usually not less than several per cent of the nominal values. For ε_{max} = MAX{ε_{in}, ε_x, ε_z}, the worst-case analysis of (23) leads to the result

$$\left|\frac{K(0)}{K(\infty)}\right|_{\min} = \left|1 - \frac{1+\varepsilon_{\max}}{1-\varepsilon_{\max}}\right| = \frac{2\varepsilon_{\max}}{1-\varepsilon_{\max}}. \tag{24}$$

For ε_{max} = 1%, this relative parasitic low-frequency gain is approximately 0.0202, i.e. -33.9dB. This undesirable value is decreased to 0.001, i.e. -60dB for ε_{max} = 0.1%. However, such a small tracking error is unrealistic in CMOS technology.

Fig. 3 shows the results of computer simulation of the HP responses of the filter from Fig. 2 when the HP output is constituted via Eq. (7). The circuit parameters are designed as follows:

$$C_1 = C_2 = 20\text{pF},\ R = 563\ \text{Ohms},\ g_m = 888\mu\text{s}. \tag{25}$$

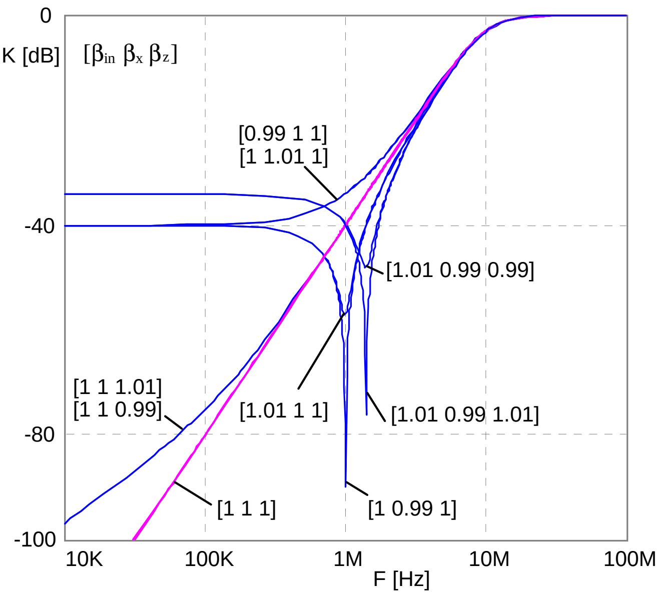

Fig. 3: The HP responses affected by different tracking errors. PSpice analysis with ideal model of the CDTA.

The corresponding biquad parameters are f_0 = 10MHz, Q = 0.707. High-pass frequency responses are analyzed for various combinations of the values β_{in}, β_x, and β_z. The combination [1 1 1] corresponds to an ideal response without tracking errors, whereas the combination [1.01 0.99 1.01] reflects the worst case (24) for ε_{max} = 1%. Note that rather small modifications are caused only by keeping precise tracking of input and x-terminal copies and considering only the tracking error of z-copy (the cases [1 1 1.01] and [1 1 0.99]). As follows from Eqs. (18) and (19), this case represents a modification of the numerator of the transfer function from the ideal form "s^2" to "$s(s+b_1')$", where $b_1' = \pm 0.01 b_1$. In other words, the original double zero point s = 0 is split to a single zero s = 0 and one real zero which absolute value is one percent of the coefficient b_1. The impact on the frequency response, shown in Fig. 3, is the same for positive and negative tracking error ε_z. Figure 3 generally demonstrates the fact that the method of creating the HP output as a combination of current signals according to (7) is worthless except the cases of extremely small tracking errors.

A similar analysis can be performed for cases (8), (10) and (14), in which the filter responses are sensitive to fulfilling the matching conditions between multiple current inputs, with similar negative influences on the frequency responses. That is why the only filter sections which do not suffer from matching problems are LP and BP sections (6 a, b) with a single current input.

4 Dynamic range optimization

Linear operation of the CDTA is limited by three factors: 1) The range of voltage V_z which provides the linear regime of the internal OTA, 2) The maximum currents I_p, I_n, I_z for linear operation of the input current differencing unit, 3) The maximum current I_x, which can be estimated as a product of V_{zmax} and transconductance g_m.

The optimization of the upper bound of the dynamic range can be symbolically described as follows:

$$\{MAX(I_p, I_n, I_z) | f \in <0, \infty)\} \le I_{CDU\max},$$

and

$$\{MAX(V_z) | f \in <0, \infty)\} \le V_{z\max}, \tag{26 a, b, c}$$

or

$$\{MAX(I_x) | f \in <0, \infty)\} \le I_{x\max} = g_m V_{z\max},$$

where I_{CDUmax}, I_{xmax} and V_{zmax} are determined by the nonlinear characteristics of the CDTA.

In addition, we should also take care of the maximum value of other nodal voltages of the circuit, concretely the voltage across the capacitance C_1.

The single-input version (5) is a typical case of biquad with unity-gains of LP section at low frequencies, HP section at high frequencies, and BP section at natural frequency ω_0. It is well-known that such parameters are accompanied by gains of LP and HP sections being Q at the natural frequency.

A simple analysis of the filter with the input defined by Eq. (5) leads to the conclusion that the transfer function from the input current to the voltage across C_1 has a band-pass character with the maximum gain at the natural frequency. The value of the maximum transimpedance is R. The transimpedance between the input current and the voltage across C_2 is of low-pass character with the low-frequency value of $1/g_m$, which is increased Q times at the natural frequency. That is why the product g_mR should be designed also from the point of view of the upper bound of dynamic range.

A similar analysis of the remaining variants of filter excitations b), c), and d) in Section 1 leads to the conclusion that variant b) in particular should not be used due to inconvenient signal dynamics. For example, the band-pass section with the I_x output has a gain of Q^2 at the natural frequency whereas the high-pass gain at the same frequency is typically Q.

5 Results of computer simulation

The performance of the proposed filter was verified using the simulation in PSpice. According to the analysis from Sections 3 and 4, the single-input excitation has been chosen, which does not suffer from the matching and dynamic range problems. The CDTA model from [17] was used, employing the n-well CMOS process TSMC 0.35µm. In [17], the CDTA was designed for a bandwidth of about 1GHz. The transconductance was set to 888µS via a bias current of 40µA. The values of passive components were used as shown in Eq. (25).

The results of AC analysis are given in Fig. 4. Both the natural frequency and the quality factor are in accordance with the proposed values.

6 Conclusion

A 2nd - order current-mode filter employing a single CDTA is described in the paper as a revisited version of circuits described in [1], [12], and [13]. The remaining filter components are two grounded capacitors and one virtually-grounded resistor. This topology provides, without the necessity of modifying the input current gate or the RC:CR transformation, a pair of high-impedance current outputs of the BP and LP types. The output current I_{HP} flows through the working capacitance.

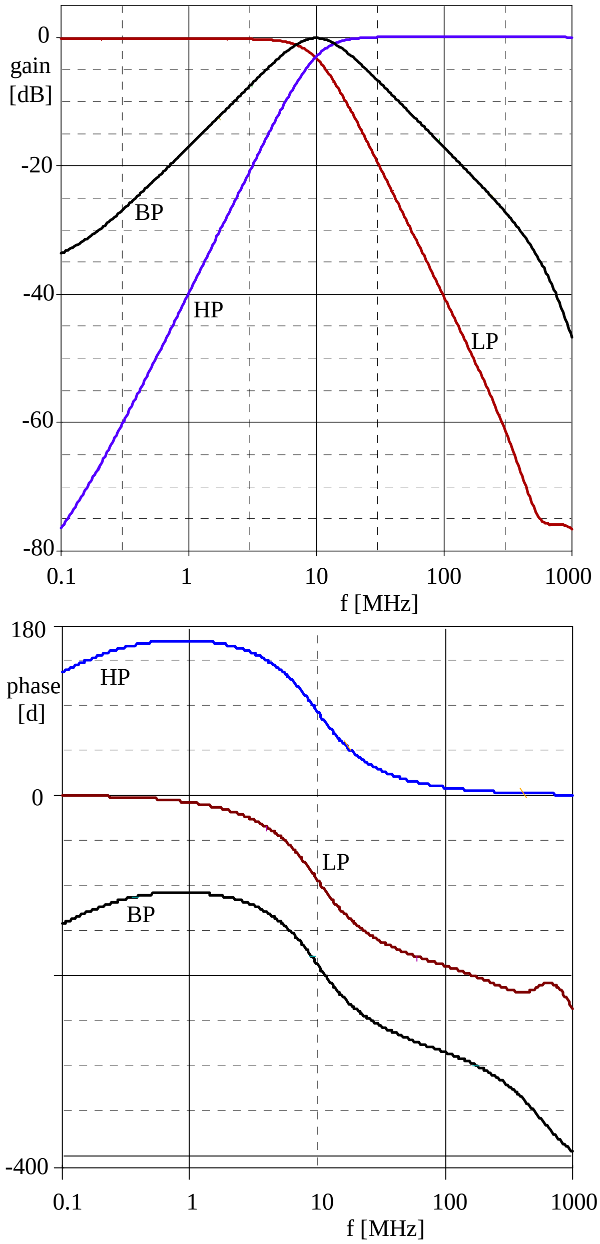

Fig. 4: Results of PSpice simulation of proposed circuit.

Where appropriate, the HP output can be implemented, together with other types of filter response, as an output for driving an independent load, using one of the known circuit techniques which are based on the combination of multiple-current inputs and summing the output currents. In the paper, these techniques are critically analyzed from the point of view of tracking errors and filter dynamic range. As a conclusion, the techniques, utilizing the copies of currents, are not recommended owing to the high sensitivities of resulting transfer functions to the tracking errors of these copies. In addition, the single-CDTA biquads do not provide a sufficient number of degrees of freedom for their proper optimization. The only variant has been selected which does not suffer simultaneously from tracking errors and dynamic range problems. Its PSpice simulation on the level of transistor CMOS structure of the CDTA confirms the theoretical conclusions. The sensitivities of the filter are low.

As results from Eqs. (4), the natural frequency ω_0 can be tuned independently of B by changing the transconductance g_m. However, the single-CDTA biquad version does not provide more freedom in electronic tuning, e.g. the orthogonal adjustment of ω_0 and Q.

After avoiding the above commented sensitive techniques of multiple current inputs and summing the output currents, this filter does not provide the high-pass output to an independent load. Such output can be added via a special current sensing technique [18]. However, this solution represents an auxiliary circuitry. From this point of view, two- or three-CDTA biquads could be more advantageous realizations of applications which require high-pass filter sections.

Acknowledgments

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