CMOS Current-controlled Current Differencing Transconductance Amplifier And Applications To Analog Signal Processing

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Introduction

- 5-terminals active element, namely current differencing transconductance amplifier (CDTA) seems to be a versatile component in the realization of a class of analog signal-processing circuits, especially analog frequency filters.
- The purpose of our reference paper is to design and synthesize a modified-version CDTA, which is newly named current controlled current differencing transconductance amplifier (CCCDTA) and using a CMOS technology.

Concept of CCCDTA

- CCCDTA properties are similar to the conventional CDTA, except that input voltages of CCCDTA are not zero and the CCCDTA has finite input resistances *R*p and *R*n at the p and n input terminals, respectively.

$$\begin{bmatrix} V_p \\ V_n \\ I_z \\ I_x \end{bmatrix} = \begin{bmatrix} R_p & 0 & 0 & 0 \\ 0 & R_n & 0 & 0 \\ 1 & -1 & 0 & 0 \\ 0 & 0 & 0 & \pm g_m \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ V_x \\ V_z \end{bmatrix}.$$

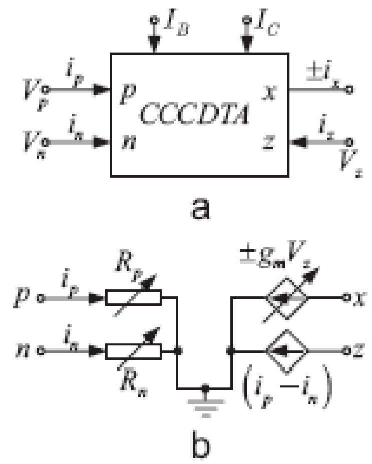

Fig. 1. CCCDTA: (a) symbol and (b) equivalent circuit.

Finite Input Resistance of CDC

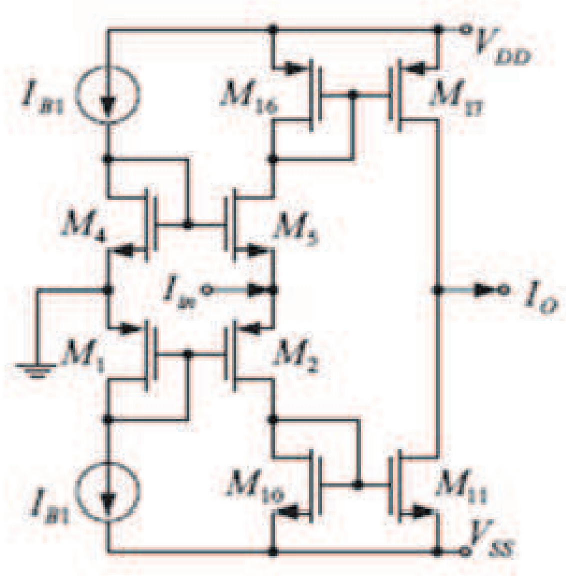

Fig. 2. Unity gain current amplifier with finite input resistance.

$$R_{in} = \frac{1}{g_{m2} + g_{m5}} - \frac{I_{B1}}{I_{in}(g_{m2} + g_{m5})}\left(\frac{g_{m2}}{g_{m1}} - \frac{g_{m5}}{g_{m4}}\right)$$

$$g_{m1} = g_{m4} \text{ and } g_{m2} = g_{m5} = g_m$$

$$R_{in} = \frac{1}{2g_m} = \frac{1}{\sqrt{8\beta_n I_{B1}}}.$$

$$I_O = -\alpha I_{in} + \varepsilon,$$

$$\alpha = \frac{g_{m5}g_{m17}}{g_{m16}(g_{m2} + g_{m5})} + \frac{g_{m2}g_{m11}}{g_{m10}(g_{m2} + g_{m5})}$$

$$\varepsilon = \left\{ \begin{array}{l} \frac{I_{B1}g_{m5}g_{m17}}{g_{m4}g_{m16}} - \frac{I_{B1}g_{m2}g_{m11}}{g_{m1}g_{m10}} \\ + \frac{I_{B1}}{(g_{m2} + g_{m5})}\left(\frac{g_{m2}}{g_{m1}} - \frac{g_{m5}}{g_{m4}}\right) \\ \times \left(\frac{g_{m5}g_{m17}}{g_{m16}} + \frac{g_{m2}g_{m11}}{g_{m10}}\right) \end{array} \right\}$$

If $g_{m1} = g_{m4}$, $g_{m2} = g_{m5}$, $g_{m10} = g_{m11}$ and $g_{m16} = g_{m17}$, then $I_O = -I_{in}$

Finite Input Resistance of CDC

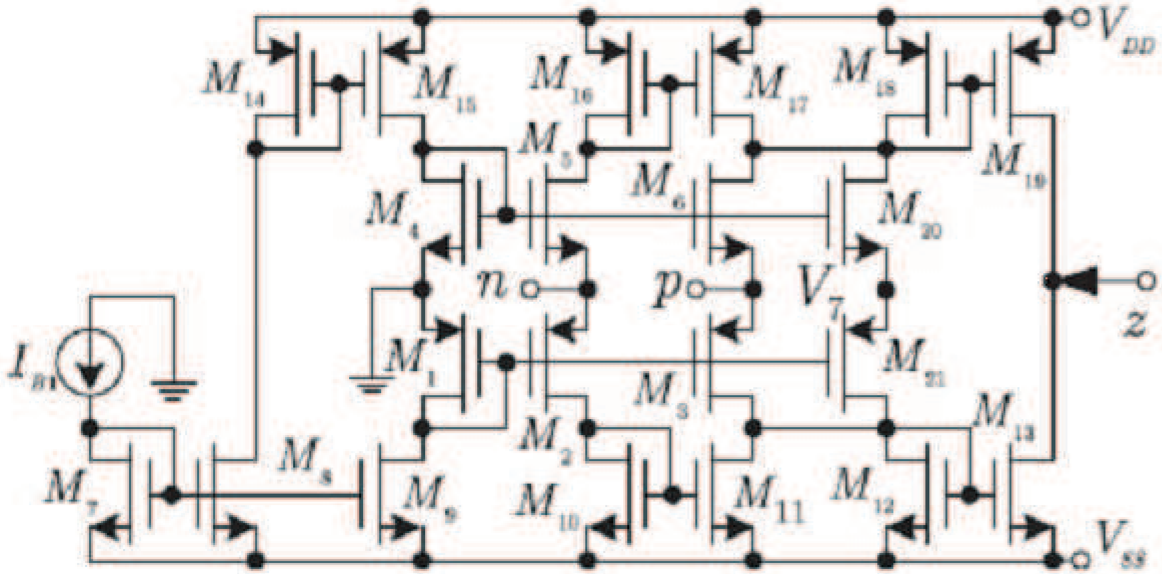

Fig. 3. A current differencing circuit which has finite input resistances.

Finite Input Resistance of CDC

- The current differencing circuit which has finite input resistances is shown in Fig. 3. The circuit implementation consists of mixed translinear loops (M1–M6). The mixed loops are DC biased by I_{B1} using current mirrors (M7–M10 and M14–M16). The output z-terminal that generates the current difference of p and n terminals is realized using transistors (M11–M13 and M17–M21)

$$I_{\mathrm{z}} = \alpha_{\mathrm{p}} I_{\mathrm{p}} - \alpha_{\mathrm{n}} I_{\mathrm{n}} + \varepsilon. \quad \Rightarrow \quad I_{\mathrm{z}} = I_{\mathrm{p}} - I_{\mathrm{n}}$$

$$\alpha_{\mathrm{p}} = \frac{g_{\mathrm{m6}} g_{\mathrm{m19}}}{g_{\mathrm{m18}}(g_{\mathrm{m6}} + g_{\mathrm{m3}})} + \frac{g_{\mathrm{m13}} g_{\mathrm{m3}}}{g_{\mathrm{m12}}(g_{\mathrm{m6}} + g_{\mathrm{m3}})},$$

$$\alpha_{\mathrm{n}} = \frac{g_{\mathrm{m5}} g_{\mathrm{m19}} g_{\mathrm{m17}}}{g_{\mathrm{m16}} g_{\mathrm{m18}}(g_{\mathrm{m2}} + g_{\mathrm{m5}})} + \frac{g_{\mathrm{m2}} g_{\mathrm{m11}} g_{\mathrm{m13}}}{g_{\mathrm{m10}} g_{\mathrm{m12}}(g_{\mathrm{m2}} + g_{\mathrm{m5}})}$$

Transconductance Amplifier

- Transistors M22 and M23 function as a differential amplifier to convert an input voltage to an output current. M24 and M25 work as a simple current mirror when I_{B2} is an input bias current. When V_{in} is applied, this makes I_{D22} and I_{D23} flowing in M22 and M23, respectively.

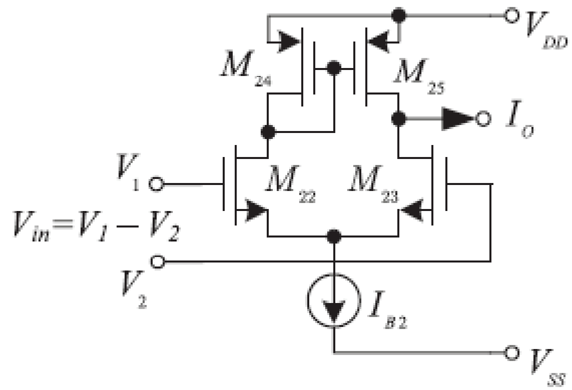

Fig. 4. The simple transconductance amplifier.

Transconductance Amplifier

$$I_O = \beta_1 g_{m22} V_1 - \beta_2 g_{m23} V_2 + \varepsilon,$$

$$\beta_1 = \frac{g_{m25}}{g_{m24}} + \frac{(g_{m23} g_{m24} - g_{m22} g_{m25})}{g_{m24}(g_{m22} + g_{m23})},$$

$$\beta_2 = 1 - \frac{(g_{m23} g_{m24} - g_{m22} g_{m25})}{g_{m24}(g_{m22} + g_{m23})}$$

If $g_{m22} = g_{m23} = g_m$ and $g_{m24} = g_{m25}$

$$I_O = g_m (V_1 - V_2)$$

- If V_2 node is grounded and V_1 is connected to the z terminal. Thus, the output current is

$$I_O = g_m V_z$$

where

$$g_m = \sqrt{\beta_n I_{B2}}.$$

The Proposed CCCDTA

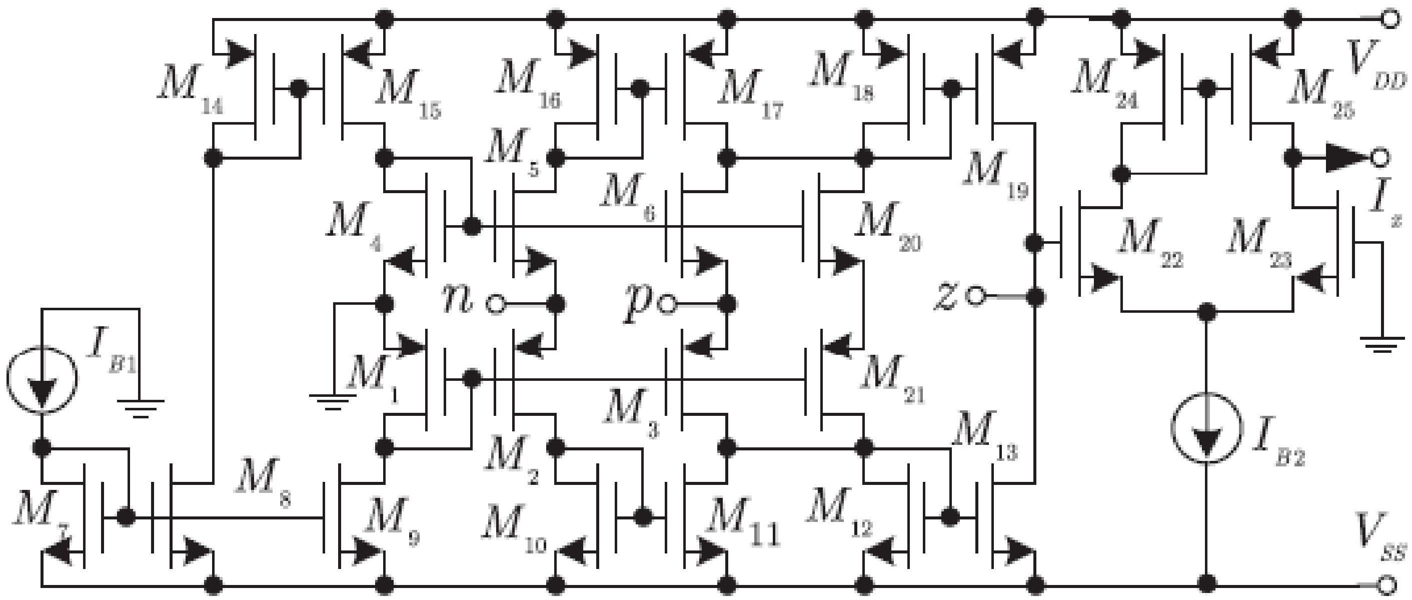

Fig. 5. Proposed current-controlled current differencing transconductance amplifier.

Simulation Results for CCCDTA

Table 1. Dimensions of the MOS transistors

CMOS transistors	$W(\mu m)/L(\mu m)$
M1–M3, M21	4/0.5
M4–M6, M20	2/0.5
M7–M13	5/0.5
M14–M19	15/0.5
M22–M23	30/1.5
M22–M23	15/1.5

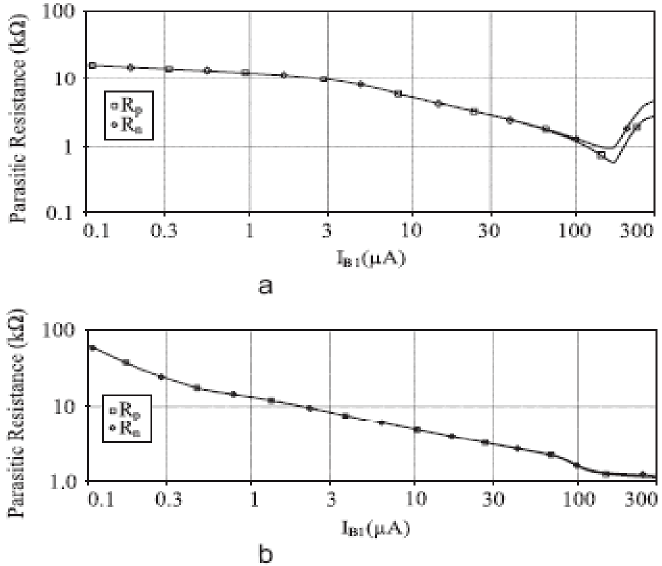

Fig. 7. Parasitic resistances at input terminals relative to I_{B1}.

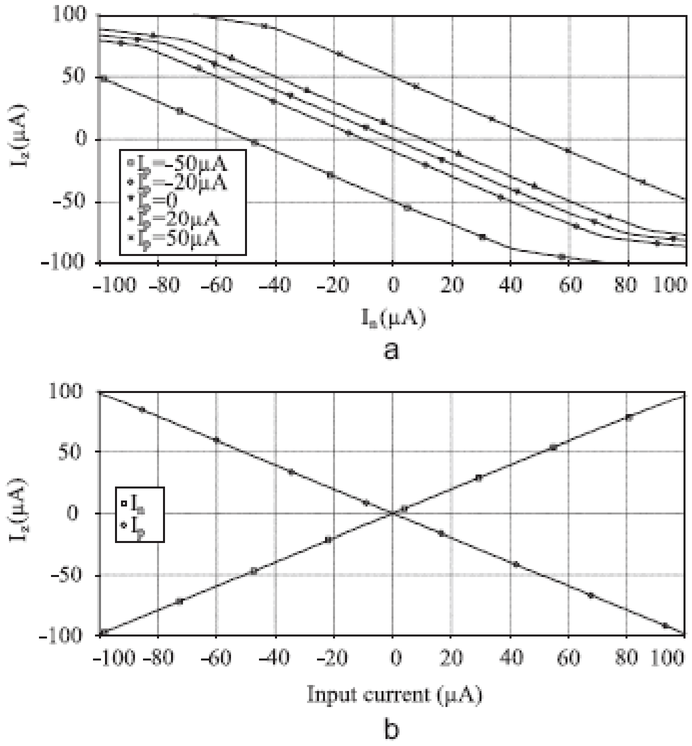

Fig. 6. DC transfer characteristics of the CCCDTA.

Simulation Results for CCCDTA

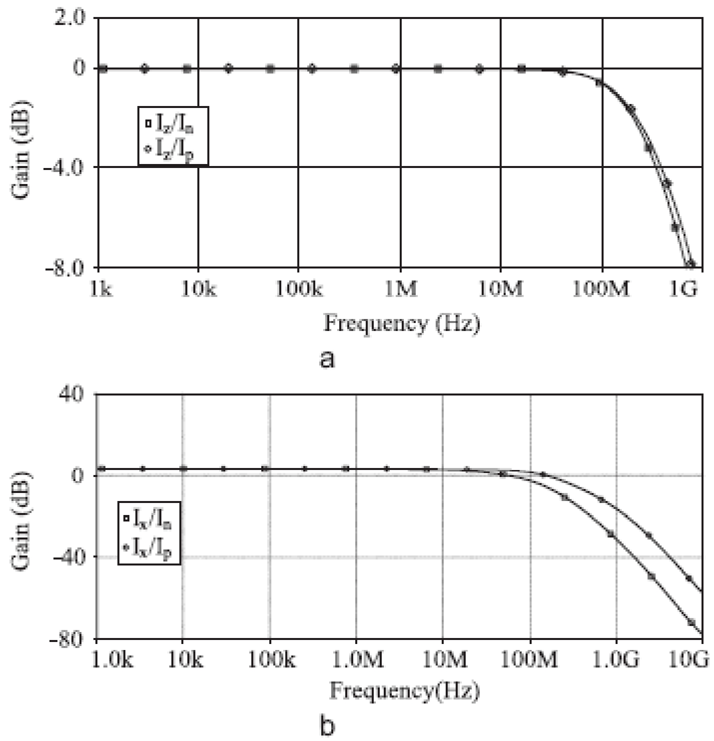

Fig. 8. Frequency responses at output terminals: (a) z terminal and (b) x terminal.

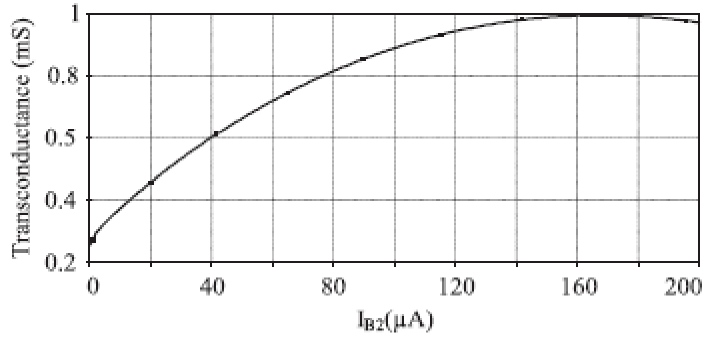

Fig. 9. Transconductance value relative to I_{B2}.

Table 2. Conclusions of CCCDTA parameters

Parameters	Values
Power consumption	1.48 mW
−3 dB Bandwidth	282 MHz (I_z/I_n), 311 MHz (I_z/I_p), 50MHz (I_z/I_n), 142 MHz (I_x/I_p)
Input current linear range	−100 μA to 100 μA
R_n and R_p ranges	821 Ω–25.1 kΩ
Input bias current range for controlling R_n and R_p	10nA to 180 μA
Transconductance	0.25–1 mS
Switching time delay	2 ns
Input bias range for controlling transconductance amplifier	1 nA to 100 μA
R_z	1.03 MΩ
R_x	999.01 kΩ

Simulation Results for CCCDTA-comments

- The PMOS and NMOS transistors employed in the proposed circuit were simulated by using the parameters of a 0.35 m TSMC CMOS technology with ±1.5V supply voltages.
- It is seen that the adjustable I_{B1} current for controllable parasitic resistances is about 10 nA -180 μA range and the resistances due to positive and negative input voltages are slightly different. The bias current more than 180 μA affects MOS transistors operation to saturate and then the CCCDTA cannot be controlled by the input parasitic resistances.
- CCCDTA is linear in $-100\ \mu A < I_n\ (I_p) < 100\ \mu A$ range.
- The obtained maximum transconductance is about 1 mS.

Application Examples – Current mode biquad filter

- Current mode biquad filter employs only one active element and 2 grounded capacitors, which is easy to fabricate.

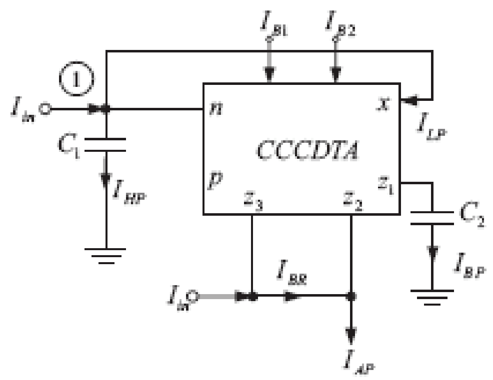

Fig. 10. Universal biquad filter based on the CCCDTA.

$$\frac{I_{HP}}{I_{in}} = \frac{s^2}{s^2 + s\frac{1}{C_1 R_n} + \frac{g_m}{C_1 C_2 R_n}},$$

$$\frac{I_{LP}}{I_{in}} = \frac{g_m / C_1 C_2 R_n}{s^2 + s\frac{1}{C_1 R_n} + \frac{g_m}{C_1 C_2 R_n}}$$

and

$$\frac{I_{BP}}{I_{in}} = \frac{s / C_1 R_n}{s^2 + s\frac{1}{C_1 R_n} + \frac{g_m}{C_1 C_2 R_n}}.$$

Application Examples – Current mode biquad filter

- The pole frequency can be adjusted by I_{B1} and I_{B2} without affecting the quality factor by keeping the ratio I_{B1} and I_{B2} to be constant.

$$\frac{I_{BS}}{I_{in}} = \frac{s^2 + g_m/C_1C_2R_n}{s^2 + s\frac{1}{C_1R_n} + \frac{g_m}{C_1C_2R_n}}$$

$$\frac{I_{AP}}{I_{in}} = \frac{s^2 - s/C_1R_n + g_m/C_1C_2R_n}{s^2 + s\frac{1}{C_1R_n} + \frac{g_m}{C_1C_2R_n}}.$$

$$\omega_0 = \left(\frac{\beta_n\sqrt{8I_{B1}I_{B2}}}{C_1C_2}\right)^{1/2}, \quad Q_0 = \left(\frac{C_1\sqrt{I_{B2}}}{C_2\sqrt{8I_{B1}}}\right)^{1/2}.$$

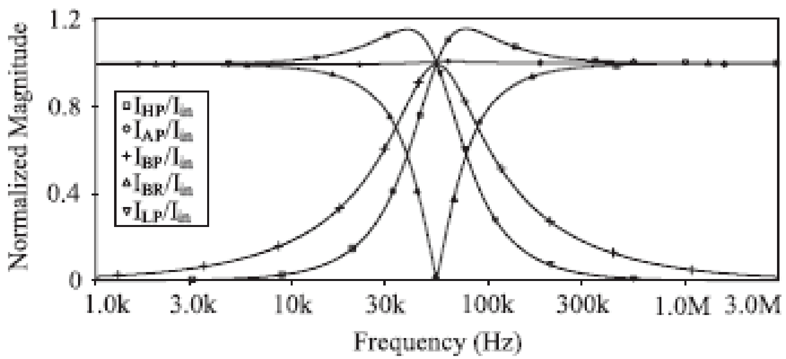

Fig. 11 Simulated results of circuit

Application Examples – Current mode biquad filter

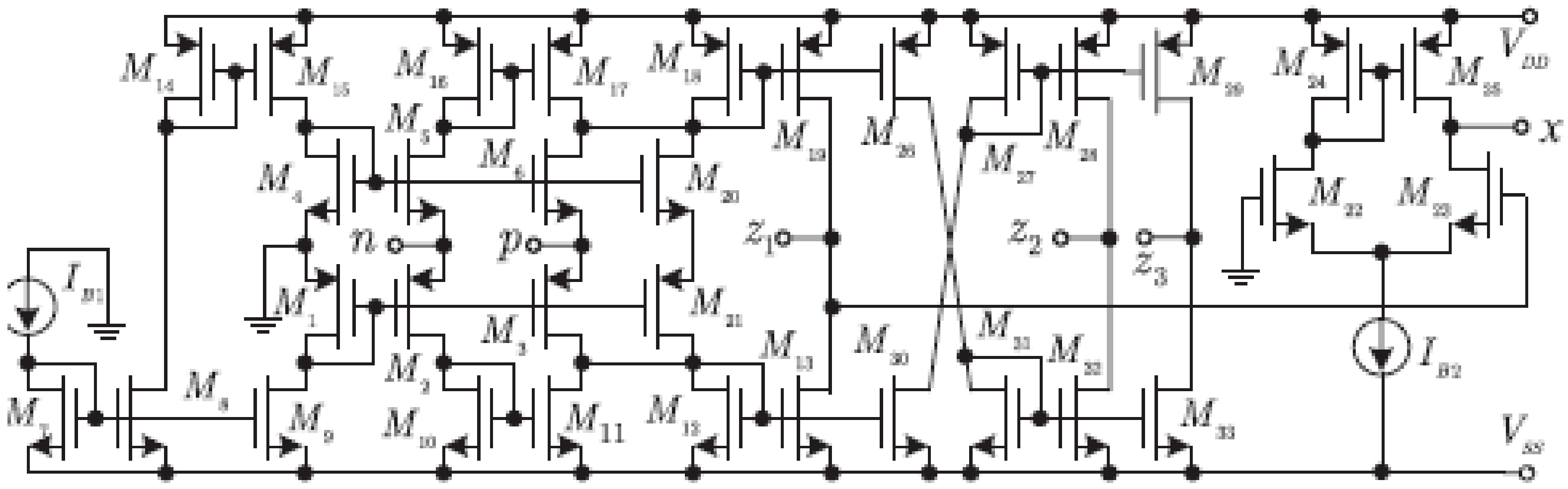

Fig. 12. Internal construction of MO-CCCDTA.

Application Examples – Current mode biquad filter

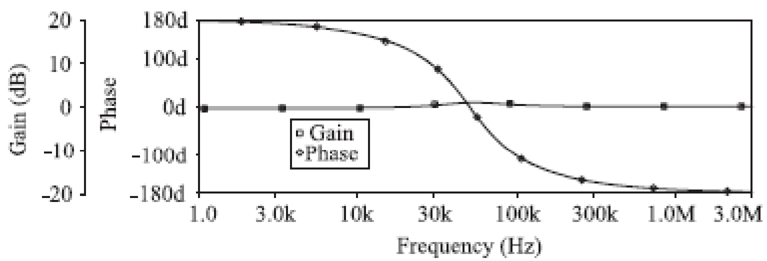

Fig. 13. Gain and phase responses of all-pass function.

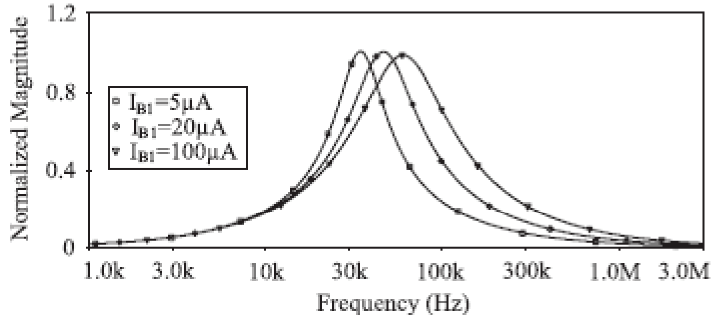

Fig. 14. Band-pass responses with varying I_{B1} of circuit in Fig. 13.

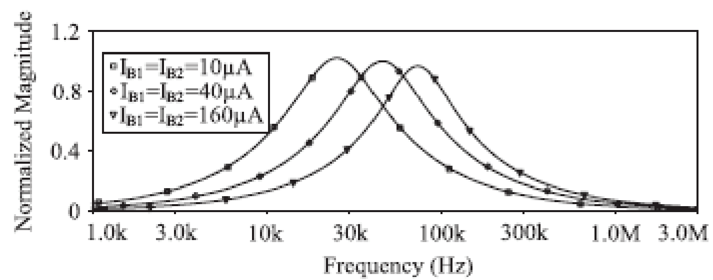

Fig. 15. Band-pass responses for different values of I_{B1} and I_{B2}.

Application Examples – Floating Inductance Simulator

- The inductance value L_{eq} can be electronically adjusted by either I_{B1} or I_{B2}.

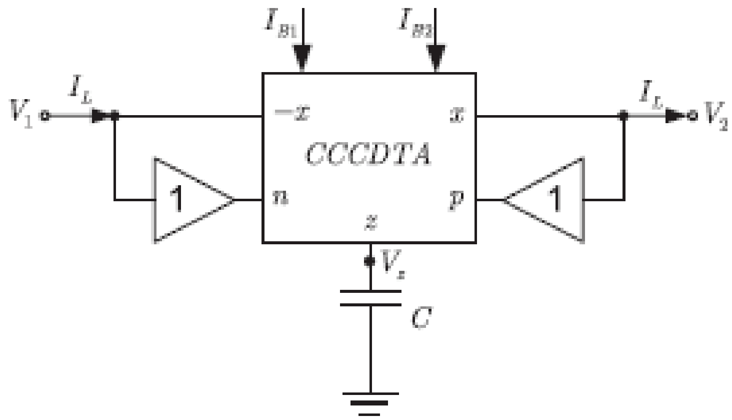

Fig. 16. Floating inductance simulator using CCCDTA.

$$I_z = \frac{V_1 - V_2}{R},$$

$$R_n = R_p = R.$$

$$V_z = \frac{I_z}{sC} = \frac{V_1 - V_2}{sCR}.$$

$$I_x = I_L = \frac{g_m(V_1 - V_2)}{sCR}.$$

$$Z_{in} = \frac{V_1 - V_2}{I_L} = \frac{sCR}{g_m}.$$

$$L_{eq} = \frac{CR}{g_m} = \frac{1}{\beta_n\sqrt{8 I_{B1} I_{B2}}}.$$

Application Examples – Floating Inductance Simulator

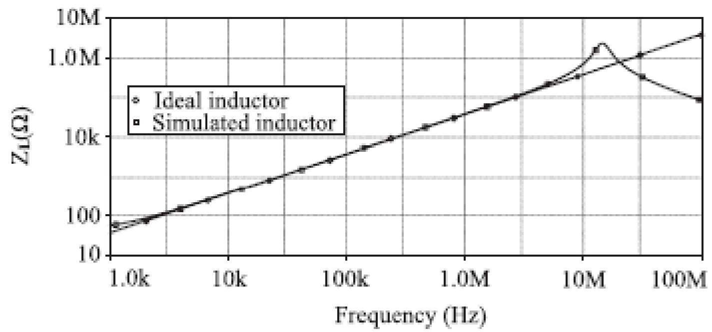

Fig. 17. The impedance values relative to frequency of the simulator.

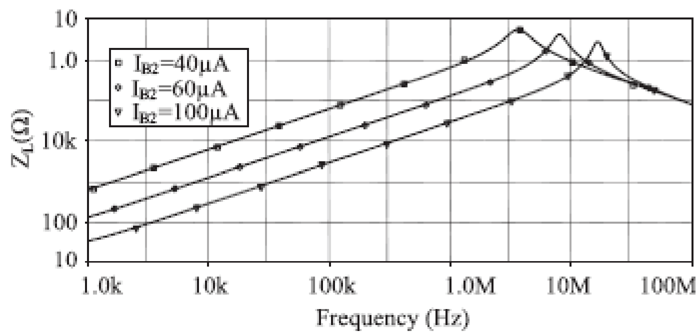

Fig. 18. The impedance values relative to frequency of the simulators with different I_{B2}.

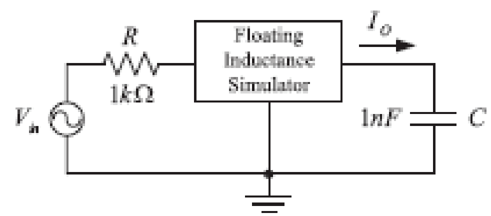

Fig. 19.. Series RLC resonant circuit.

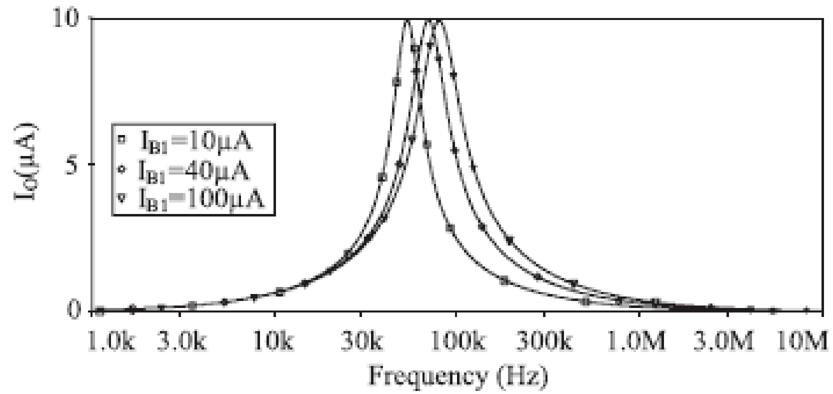

Fig. 20. Simulated current characteristics of the resonant circuit in Fig. 24 when I_{B2} is varied.

Application Examples – Quadrature Oscillator

- Quadrature Oscillator consists of one multiple output CCCDTA (MO-CCCDTA) and 2 grounded capacitors.

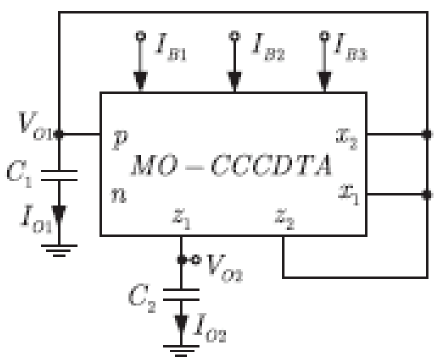

Fig. 21. Quadrature oscillator using MO-CCCDTA.

MO-CCCDTA.

$$I_{z1} = I_{z2} = I_p - I_n,$$

$$I_{x1} = g_{m1} V_{z1}$$

$$I_{x2} = g_{m2} V_{z2},$$

$$s^2 + s\frac{1}{C_1}\left(\frac{2}{R_P} - g_{m2}\right) + \frac{g_{m1}}{C_1 C_2 R_P} = 0.$$

condition of oscillation

$$\frac{2}{R_P} = g_{m2}, \quad \Rightarrow \quad I_{B1} = 0.25 I_{B3},$$

oscillation frequency

$$\omega_0 = \sqrt{\frac{g_{m1}}{C_1 C_2 R_P}} = \left(\frac{\beta_n \sqrt{8 I_{B1} I_{B2}}}{C_1 C_2}\right)^{1/2}$$

Application Examples – Quadrature Oscillator

- The oscillation frequency w_o can be controlled by bias current. It should be remarked that the condition of oscillation can be tuned by I_{B3} without affecting the oscillation frequency which can be adjusted by I_{B1} and I_{B2}. Furthermore, the quadrature sinusoidal signals can be obtained at I_{O1} and I_{O2} or V_{O1} and V_{O2}

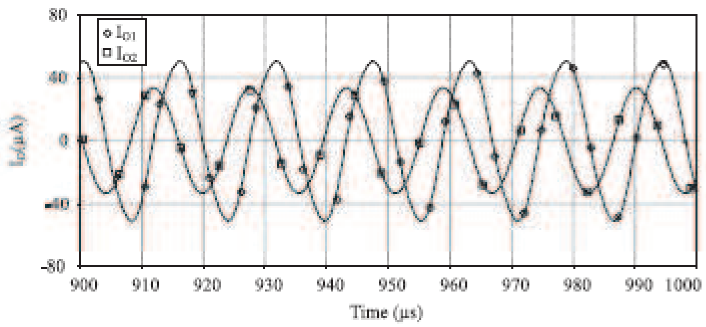

Fig. 22. Transient responses in current-mode signals.

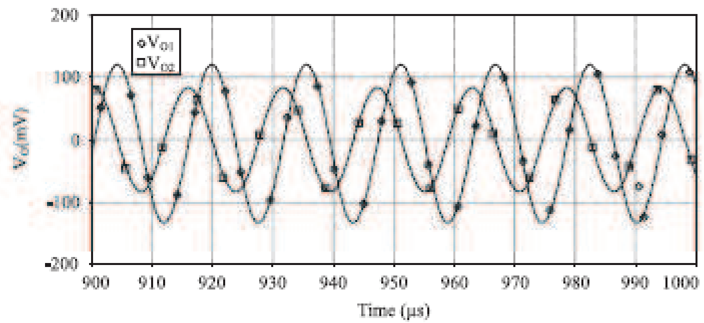

Fig. 23. Transient responses in voltage-mode signals.

Reference Article

- Montree Siripruchyanun,Winai Jaikla, CMOS current-controlled current differencing transconductance amplifier and applications to analog signal processing, Int. J. Electron. Commun. (AEÜ) 62 (2008) 277 – 287.

THANK YOU...