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HERENCSÁR, N.; KOTON, J.; VRBA, K.; LATTENBERG, I.

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New voltage-mode universal filter and sinusoidal oscillator using only single DBTA

Norbert Herencsár*, Jaroslav Koton, Kamil Vrba and Ivo Lattenberg

Department of Telecommunications, Faculty of Electrical Engineering and Communication, Brno University of Technology, Purkynova 118, 612 00 Brno, Czech Republic

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In this paper, a new voltage-mode second-order universal frequency filter and sinusoidal oscillator using only single differential-input buffered and transconductance amplifier (DBTA) is presented. Proposed voltage-mode filter structure using single DBTA and four passive elements can provide all standard filter functions, i.e. low-, band-, high-pass, band-stop, and all-pass without changing the circuit topology and enables independent control of the quality factor Q using single passive element. The circuit requires the minimal number of active and passive elements with no conditions for component matching. By slight modification of the proposed filter structure, the new DBTA-based sinusoidal oscillator is easily obtained. The oscillation condition and the oscillation frequency are independently adjustable by different virtually grounded passive elements. The proposed sinusoidal oscillator employs only grounded capacitors. The passive and active sensitivities of all the proposed circuit configurations are low. PSPICE simulations using a BJT realization of DBTA and experimental results based on commercially available amplifiers OPA860 and MAX436 are included, which prove the workability of the proposed circuits.

Keywords: active filter; analogue signal processing; universal frequency filter; differential-input buffered and transconductance amplifier (DBTA); voltage-mode; sinusoidal oscillator

1. Introduction

Frequency filters and sinusoidal oscillators are linear electric circuits (Toumazou *et al.* 1990) that are used in wide area of electronics and also are the basic building blocks in analogue signal processing. The analogue frequency filters are the most often used as anti-aliasing video filters in the analogue sections of high-speed data communication systems defined by ITU BT 601 standard (Uygur and Kuntman 2007) or for signal processing in wireless LANs described by IEEE 802.11 standard (Lo *et al.* 2009), in IF (intermediate frequency) receiver stages of the GSM cellular telephones (Fabre *et al.* 1997, Ibrahim *et al.* 2005), in receiver baseband blocks of modern radio systems (Rudell *et al.* 1997, Toker *et al.* 2001), in hard-drive communication interfaces (Laber and Gray 1993), measurement systems (Vainio and Ovaska 1997), automotive industry (Ferri and Guerrini 2003), or in piezoresistive pressure sensors (Samitier *et al.* 1998). Oscillators also represent an important unit in many telecommunication, instrumentation and control systems (Holzel 1993, Ahmed *et al.* 1997,

*Corresponding author. Email: herencsn@feec.vutbr.cz

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Cam *et al.* 1998, Cicekoglu and Toker 1999, Soliman 1999, Khan and Khwaja 2000, Abuelma'atti and Al-Absi 2005, Horng *et al.* 2005).

The presented structures of active frequency filters and oscillators are often employing current conveyors (CCs), where the second-generation current conveyor (CCII) (Sedra and Smith 1970) is the most popular. The CCII is the basic block of many other active elements. Here, the current-feedback operational amplifier (CFOA) (Evans 1988, Svoboda *et al.* 1991, Fabre 1992, Liu 1995, Soliman 1996) that is a combination of the CCII and voltage follower (VF) (Sedra 1972) or the modified CFOA (MCFOA) that is the interconnection of the plus-type and minus-type CCIIs (Yuce and Minaei 2008) can be mentioned. Later, the inverting second-generation CC (ICCII) as a missing building block in analogue signal processing techniques has been introduced (Awad and Soliman 1999). By the combination of CCII and ICCII the dual-X second-generation CC (DXCCII) (Zeki and Toker 2002) for the tuneable continuous-time filter design has been built. Recently, further research has focused on CCs with variable current and/or voltage gains such as electronically tuneable second-generation current conveyor (ECCII) (Minaei *et al.* 2006), variable gain current conveyor (VGCCII) (Yuce *et al.* 2008), or voltage and current gain second generation current conveyor (VCG-CCII) (De Marcellis *et al.* 2009).

Using the duality principle, the voltage conveyor (VC) has been presented (Dostal and Pospisil 1982a). As in the theory of current conveyors, also here the first- and second-generation VCs (VCI, VCII, IVCI, and IVCII) were described (Dostal and Pospisil 1982b, Novotny and Vrba 2003, Minarcik and Vrba 2006). The best known is the plus-type differential current voltage conveyor (DCVC+) (Salama and Soliman 1999) that is more often labelled as the current differencing buffered amplifier (CDBA) (Acar and Ozoguz 1999). By the modification of the CDBA or replacement of the VF by the operational transconductance amplifier (OTA) (Geiger and Sanchez-Sinencio 1985) the differential-input current feedback amplifier (DCFA) (Zeki *et al.* 2001), current differencing transconductance amplifier (CDTA) (Biolek 2003), and current follower transconductance amplifier (CFTA) (Herencsar *et al.* 2008a, Herencsar *et al.* 2008b) or inverted current follower transconductance amplifier (ICFTA) (Herencsar *et al.* 2008c) have been presented.

Based on the idea of the "universal" active element (Carlosena *et al.* 1994) the universal current conveyor (UCC) was designed and developed (Becvar *et al.* 2000, Cajka *et al.* 2004, Herencsar and Vrba 2007, Jerabek and Vrba 2009), and produced by AMI Semiconductor Czech, Ltd., (now ON Semiconductor Czech Republic, Ltd.) in the CMOS 0.35 μm technology under the designation UCC-N1B 0520. Through suitable interconnection or grounding of the terminals, the UCC enables realization of all types and generations of CCs with single current input X (Becvar *et al.* 2000). On the basis of the UCC, the universal voltage conveyor (UVC) was designed (Novotny and Vrba 2003, Cajka and Vrba 2004, Minarcik and Vrba 2008, Koton *et al.* 2009) and produced under the designation UVC-N1C 0520. The realizable generations and types of VCs using the UVC were shown by Minarcik and Vrba (2006).

There is still a need to develop new filter and oscillator structures with new active elements that offer new advantages. In this paper, the application possibilities of the differential-input buffered and transconductance amplifier (DBTA) (Herencsar *et al.* 2009) for the design of frequency filter and oscillator are presented. The proposed voltage-mode (VM) filter employs only single DBTA and four passive elements. All standard filter functions without changing the circuit topology can be realized. Furthermore, independent control of the quality factor Q via single passive element is enabled. The presented circuit requires minimal number of active and passive elements with no conditions for component matching. By a simple modification of the proposed filter structure, novel DBTA-based sinusoidal oscillator is obtained. In the oscillator structure all passive elements are grounded, which makes the circuit attractive for VLSI. The oscillation condition and the oscillation frequency are independently adjustable by the passive elements. PSPICE simulations and experimental results based on commercially available amplifiers OPA860 and MAX436 are included to verify the theoretical conclusions of the proposed circuits.

2. Description of the DBTA

The schematic symbol and internal structure of the DBTA (differential-input buffered and transconductance amplifier) (Herencsar *et al.* 2009) is shown in Figure 1. It has low-impedance current inputs p, n and high-impedance voltage input y. The difference of the i_p and i_n currents flows from auxiliary terminal z. The voltage v_z on this terminal is transferred into output terminal w using the VF and also transformed into current using the transconductance g_m, which flows into output terminal x. Relations between the individual terminals of the DBTA can be described by following hybrid matrix:

$$\begin{bmatrix} v_\mathrm{p} \\ v_\mathrm{n} \\ i_\mathrm{y} \\ i_\mathrm{z} \\ v_\mathrm{w} \\ i_\mathrm{x} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 1 & -1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & \pm g_\mathrm{m} & 0 & 0 \end{bmatrix} \begin{bmatrix} i_\mathrm{p} \\ i_\mathrm{n} \\ v_\mathrm{y} \\ v_\mathrm{z} \\ i_\mathrm{w} \\ v_\mathrm{x} \end{bmatrix}. \tag{1}$$

The possible bipolar implementation of the DBTA is shown in Figure 2. The differential-input stage is formed by transistors Q_1–Q_{29}, transistors Q_{30}–Q_{35} form the VF, and the OTA consists of transistors Q_{36}–Q_{39}. In the design the transistor model parameters NR100N (NPN) and PR100N (PNP) of bipolar arrays ALA400 from AT&T were used (Frey 1993). Bias current $I_\mathrm{O} = 400\ \mu$A has been chosen. The transconductance g_m of DBTA can be set by current $I_\mathrm{B} = 2g_\mathrm{m}V_\mathrm{T}$, where V_T is the thermal voltage (approximately 26mV at 27°C). The parasitic elements of the proposed DBTA in Figure 2 have been computed using PSPICE simulation program. The parasitic elements in Figure 3 are computed as $C_\mathrm{p1} = C_\mathrm{n1} = C_\mathrm{w1} = 3.08$ pF, $C_\mathrm{p2} = 2.73$ pF, $C_\mathrm{n2} = 2.64$ pF, $C_\mathrm{y} = 4.41$ pF, $C_\mathrm{z} = 2.04$ pF, $C_\mathrm{w2} = 1.65$ pF, $C_\mathrm{x} = 1.06$ pF, $L_\mathrm{p} = L_\mathrm{n} = 119.91$ nH, $L_\mathrm{w} = 112.03$ nH, $R_\mathrm{p1} = R_\mathrm{n1} = 2.84$ kΩ, $R_\mathrm{p2} = R_\mathrm{n2} = R_\mathrm{w2} = 39.34$ Ω, $R_\mathrm{p3} = R_\mathrm{n3} = 205.67$ Ω, $R_\mathrm{y} = 34.17$ kΩ, $R_\mathrm{z} = 228.21$ kΩ, $R_\mathrm{w1} = 1.09$ kΩ, $R_\mathrm{w3} = 281.37$ Ω, $R_\mathrm{x} = 1.49$ MΩ. The maximum values of terminal voltages and terminal currents without producing significant distortion are computed as ±365 mV and ±605 μA, respectively. The dc voltage gains $\beta_\mathrm{p} \cong \beta_\mathrm{n} \cong 0.961$ and $\gamma \cong 0.962$ with bandwidths $f_{\beta\mathrm{p}} \cong f_{\beta\mathrm{n}} \cong 381.1$ MHz and $f_\gamma \cong 417.1$ MHz. The dc current gains $\alpha_\mathrm{p} \cong \alpha_\mathrm{n} \cong 0.989$ with bandwidths $f_{\alpha\mathrm{p}} \cong 170.9$ MHz and $f_{\alpha\mathrm{n}} \cong 173.9$ MHz. The transconductance $g_\mathrm{m} \cong 0.951$ mS with the bandwidth $f_{g\mathrm{m}} \cong 85.8$ MHz.

3. Proposed universal filter structure

The proposed VM second-order frequency filter employing single DBTA and four passive elements is shown in Figure 4. Even if all passive elements are shown as floating, which might be not attractive for integration (Bhusan and Newcomb 1967, Abuelma'atti *et al.* 1995), it should be mentioned that unused voltage inputs are always grounded, as described below.

The output voltages $\boldsymbol{V}_{o1}$ and $\boldsymbol{V}_{o2}$ of this circuit are given by the relations:

$$\boldsymbol{V}_{o1} = \frac{1}{\Delta}\left[G_1 g_{\mathrm{m}} \boldsymbol{V}_{i1} + \boldsymbol{s}C_2 G_2 \boldsymbol{V}_{i2} + \boldsymbol{s}^2 C_1 C_2 \boldsymbol{V}_{i3} - \boldsymbol{s}C_2 g_{\mathrm{m}} \boldsymbol{V}_{i4}\right], \tag{2}$$

$$\boldsymbol{V}_{o2} = \frac{1}{\Delta}\begin{bmatrix} -\left(\boldsymbol{s}C_1 G_1 + G_1 G_2\right)\boldsymbol{V}_{i1} + G_1 G_2 \boldsymbol{V}_{i2} + \\ +\boldsymbol{s}C_1 G_1 \boldsymbol{V}_{i3} + \left(\boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s}C_2 G_2\right)\boldsymbol{V}_{i4} \end{bmatrix}, \tag{3}$$

where

$$\Delta = \boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s}C_2 G_2 + G_1 g_{\mathrm{m}}. \tag{4}$$

For the proposed filter depending on the status of circuit input four voltages $\boldsymbol{V}_{i1}$, $\boldsymbol{V}_{i2}$, $\boldsymbol{V}_{i3}$ and $\boldsymbol{V}_{i4}$ numerous filter functions are obtained. Based on the output selected there are two cases show as presented below:

Case I. If the $\boldsymbol{V}_o = \boldsymbol{V}_{o1}$ is used as output, then from (2) the realizable transfer functions in voltage mode are:

(i) If $\boldsymbol{V}_{i2} = \boldsymbol{V}_{i3} = \boldsymbol{V}_{i4} = 0$ (grounded), a low-pass filter (LP1) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{i1}$;
(ii) If $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i3} = \boldsymbol{V}_{i4} = 0$ (grounded), a band-pass filter (BP1) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{i2}$;
(iii) If $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i2} = \boldsymbol{V}_{i4} = 0$ (grounded), a high-pass filter (HP1) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{i3}$;
(iv) If $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i2} = \boldsymbol{V}_{i3} = 0$ (grounded), a band-pass filter (BP2) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{i4}$;
(v) If $\boldsymbol{V}_{i2} = \boldsymbol{V}_{i4} = 0$ (grounded) and $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i3} = \boldsymbol{V}_{in}$, a band-stop filter (BS) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{in}$;
(vi) If $\boldsymbol{V}_{i1} = 0$ (grounded) and $\boldsymbol{V}_{i2} = \boldsymbol{V}_{i3} = \boldsymbol{V}_{i4} = \boldsymbol{V}_{in}$, an all-pass (AP) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{in}$.

In this case the proposed circuit is universal and can provide all standard types of filter functions, i.e. low-, band-, high-pass, band-stop, and an all-pass response without changing the circuit topology.

Case II. If the $\boldsymbol{V}_o = \boldsymbol{V}_{o2}$ is used as output, then from (3) the realizable transfer functions in voltage mode are:

(vii) If $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i3} = \boldsymbol{V}_{i4} = 0$ (grounded), a low-pass filter (LP2) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{i2}$;
(viii) If $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i2} = \boldsymbol{V}_{i4} = 0$ (grounded), a band-pass filter (BP3) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{i3}$;
(ix) If $\boldsymbol{V}_{i3} = 0$ (grounded) and $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i2} = \boldsymbol{V}_{i4} = \boldsymbol{V}_{in}$, a high-pass filter (HP2) can be obtained with $\boldsymbol{V}_o/\boldsymbol{V}_{in}$.

Thus, the circuit is multifunction and it is capable of realizing low-, band- and high-pass response without changing the circuit topology. In case of HP2 response the proposed circuit requires component matching conditions $C_1 = C_2$ and $G_1 = G_2$.

For all filters the natural frequency ω_0, quality factor Q and bandwidth BW derived from (4) are:

$$\omega_0 = \sqrt{\frac{G_1 g_\mathrm{m}}{C_1 C_2}},\ Q = \frac{1}{G_2}\sqrt{\frac{C_1 G_1 g_\mathrm{m}}{C_2}},\ \mathrm{BW} = \frac{\omega_0}{Q} = \frac{G_2}{C_1}. \tag{5}$$

Note that the quality factor Q can be controlled independently of natural frequency ω_0 by G_2. By replacing appropriate conductor by FET-based voltage-controlled resistor (VCR) (Hribsek and Newcomb 1976, Senani 1994), the quality factor Q can be controlled electronically that is particular advantage of the proposed circuit. The natural frequency ω_0 can be independently adjusted from the bandwidth, by varying C_2, G_1 or g_m of the proposed frequency filter. Here, the appropriate capacitor can be replaced by a voltage-controlled capacitor (VCC) (Newcomb 1981, Lee *et al.* 2008) or by digitally-controlled varactor (DCV) (Chen *et al.* 2005) for electronical control of the natural frequency ω_0 independently from the bandwidth.

The relative sensitivities of the ω_0, Q and BW parameters of the designed circuit derived from (5) are:

$$S^{\omega_0}_{G_1,g_\mathrm{m}} = -S^{\omega_0}_{C_1,C_2} = \frac{1}{2}, \quad S^{\omega_0}_{G_2} = 0,$$

$$S^{Q}_{C_1,G_1,g_\mathrm{m}} = -S^{Q}_{C_2} = \frac{1}{2}, \quad S^{Q}_{G_2} = -1, \tag{6}$$

$$S^{\mathrm{BW}}_{G_2} = -S^{\mathrm{BW}}_{C_1} = 1, \quad S^{\mathrm{BW}}_{C_2,G_1,g_\mathrm{m}} = 0.$$

From the results it is evident that the sensitivities are low and not larger than unity of absolute value.

Taking into account the non-idealities of DBTA, the relationship of the terminal currents and voltages given in (1) can be rewritten as:

$$\begin{bmatrix} v_\mathrm{p} \\ v_\mathrm{n} \\ i_\mathrm{y} \\ i_\mathrm{z} \\ v_\mathrm{w} \\ i_\mathrm{x} \end{bmatrix} = \begin{bmatrix} 0 & 0 & \beta_\mathrm{p} & 0 & 0 & 0 \\ 0 & 0 & \beta_\mathrm{n} & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ \alpha_\mathrm{p} & -\alpha_\mathrm{n} & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & \gamma & 0 & 0 \\ 0 & 0 & 0 & \pm g_\mathrm{m} & 0 & 0 \end{bmatrix} \begin{bmatrix} i_\mathrm{p} \\ i_\mathrm{n} \\ v_\mathrm{y} \\ v_\mathrm{z} \\ i_\mathrm{w} \\ v_\mathrm{x} \end{bmatrix}, \tag{7}$$

where $\alpha_\mathrm{p} = 1 - \varepsilon_\mathrm{i}$, $\alpha_\mathrm{n} = 1 - \varepsilon_\mathrm{i}$ and ε_i ($|\varepsilon_\mathrm{i}| \ll 1$) are the current tracking errors from p and n terminals to z terminal, $\beta_\mathrm{p} = 1 - \varepsilon_\mathrm{v}$, $\beta_\mathrm{n} = 1 - \varepsilon_\mathrm{v}$ and ε_v ($|\varepsilon_\mathrm{v}| \ll 1$) are the voltage tracking errors from p and n terminals to z terminal and $\gamma = 1 - \varepsilon_\mathrm{v}$ and ε_v ($|\varepsilon_\mathrm{v}| \ll 1$) is the voltage tracking error from z terminal to w terminal of DBTA, respectively. The transconductance g_m of the OTA with the non-idealities can be assumed as (Tsukutani *et al.* 2006, Chen *et al.* 2008):

$$g_\mathrm{m} = \frac{g_\mathrm{m}\omega_g}{\boldsymbol{s} + \omega_g} \cong g_\mathrm{m}(1 - \mu\boldsymbol{s}), \tag{8}$$

where ω_g is the first-pole of the OTA and $\mu = 1/\omega_g$. Taking into account non-idealities of the DBTA mentioned above, the denominator of (2), (3) becomes:

$$\Delta = \boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s} C_2 G_2 \left(1 - \frac{\alpha_{\mathrm{n}}\beta_{\mathrm{n}} G_1 g_{\mathrm{m}} \mu}{C_2 G_2}\right) + \alpha_{\mathrm{n}}\beta_{\mathrm{n}} G_1 g_{\mathrm{m}}. \tag{9}$$

Due to the parasitic effect, the characteristic departs from the ideal responses. But, the parasitic effect can be made negligible satisfying the following condition:

$$\frac{\alpha_{\mathrm{n}}\beta_{\mathrm{n}} G_1 g_{\mathrm{m}} \mu}{C_2 G_2} \ll 1. \tag{10}$$

4. Proposed sinusoidal oscillator

From the filter structure in Figure 4, by setting $\boldsymbol{V}_{i1} = \boldsymbol{V}_{i2} = \boldsymbol{V}_{i3} = \boldsymbol{V}_{i4} = 0$ (grounded) and connecting the capacitor C_3 to terminal p of the DBTA, a single DBTA-based sinusoidal oscillator can be obtained as shown in Figure 5.

In this case, the characteristic equation of the proposed configuration can be given by:

$$\boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s}(C_2 G_2 - C_3 g_{\mathrm{m}}) + G_1 g_{\mathrm{m}}. \tag{11}$$

The oscillation condition and oscillation frequency ω_0 of this circuit can be obtained as:

$$C_2 G_2 = C_3 g_{\mathrm{m}}, \tag{12}$$

and

$$\omega_0 = \sqrt{\frac{G_1 g_{\mathrm{m}}}{C_1 C_2}}. \tag{13}$$

It should be noted that the condition of oscillator (12) can be controlled through adjusting the value of the conductor G_2 and/or the capacitor C_3 without affecting the oscillation frequency ω_0 (13). Analogously, oscillation frequency ω_0 can be adjusted by controlling the value of the conductor G_1 and/or the capacitor C_1 without affecting the oscillation condition. Similarly, as the frequency filter, the oscillation condition and oscillation frequency of this solution might be adjusted using FET-based VCRs (Hribsek and Newcomb 1976, Senani 1994).

The active and passive sensitivities of the proposed oscillator are low, not larger than unity of absolute value and are obtained as:

$$S^{\omega_0}_{G_1,g_{\mathrm{m}}} = -S^{\omega_0}_{C_1,C_2} = \frac{1}{2}, \tag{14}$$

and

$$S^{\omega_0}_{G_2,C_3} = 0. \tag{15}$$

5. Simulation and measurement results

Using the bipolar implementation of the DBTA (Figure 2), the proposed universal filter structure (*Case I*) has been designed for characteristic frequency $f_0 \approx 1$ MHz and the quality factor of filters $Q = 1$, and simulated in PSPICE software. The following values have been chosen: $C_1 = C_2 = 150$ pF, $G_1 = G_2 = 1$ mS ($R_1 = R_2 = 1$ kΩ) and $g_\mathrm{m} = 1$ mS ($I_\mathrm{B} = 50$ μA). For the practical measurements the DBTA has been implemented by using commercially available amplifiers, as shown in Figure 6. The simulation and measurement results of the low- (LP1), band- (BP2), high-pass (HP1), band-stop (BS), and all-pass (AP) frequency filter working in voltage mode are shown in Figure 7 and Figure 8. Simulation and measurement results of the band-pass filter BP1 working in voltage mode are shown in Figure 9. Here, the possibility of adjusting the quality factor Q is demonstrated. For required values $Q = \{0.3; 1; 3; 10; 30\}$ the conductivity must be $G_2 = \{3.333; 1.000; 0.333; 0.100; 0.033\}$ mS ($R_2 = \{0.3; 1; 3; 10; 30\}$ kΩ). From the results it is evident that the results of the measurements are in agreement with the simulations. In the higher-frequency region the real properties of the OPA860, MAX436 amplifiers and parasitic capacities or inductances of the constructed prototypes begin to be more significant.

In order to confirm the above given theoretical analysis, the proposed DBTA-based sinusoidal oscillator in Figure 5 has also been simulated using PSPICE software. The DBTA model, using commercially available amplifiers OPA860 and MAX436, illustrated in Figure 6 has been used with the supply voltages of ±5 V. To obtain the sinusoidal output waveform with the oscillation frequency of $f_0 = \omega_0/2\pi \cong 15.92$ kHz, the following passive component values have been chosen: $G_1 = G_2 = 0.1$ mS ($R_1 = R_2 = 10$ kΩ), $g_\mathrm{m} = 0.1$ mS, $C_1 = C_2 = C_3 = 1$ nF where $G_2 \cong 93.46$ μS ($R_2 \cong 10.7$ kΩ) was designed to be smaller than g_m to ensure the oscillations would start. The simulated sinusoidal output $\boldsymbol{V}_o$ of the proposed oscillator is shown in Figure 10 (a). From the simulation results, the oscillation frequency of $f_0 \cong 15.88$ kHz is obtained, which is slightly lower than theoretical oscillation frequency. Figure 10 (b) shows the simulated frequency spectrum of $\boldsymbol{V}_o$. The total harmonic distortion (THD) is equal to 2.069 %, and the results are summarized in Table 1. The control of f_0 via G_1 without affecting the oscillation condition is shown in Figure 11.

6. Conclusion

In this paper, the application possibilities of the novel versatile active function block for analogue signal processing, namely the differential-input buffered and transconductance amplifier (DBTA), have been demonstrated in voltage-mode universal filter and sinusoidal oscillator design. The DBTA consists of differential-input stage, voltage follower and operational transconductance amplifier. The possible implementation of the DBTA by commercially available active elements and its possible internal bipolar structure have been shown. Proposed VM universal filter structure using only single DBTA and four passive elements provides all standard filter functions (LP, BP, HP, BS, and AP) without changing the circuit topology and enables independent control of the quality factor Q via single passive element. The circuit requires minimal number of active and passive elements with no conditions for component matching. Furthermore, a single DBTA-based sinusoidal oscillator can be easily obtained by a slight modification of the proposed single DBTA-based filter. Its oscillation condition and oscillation frequency can be controlled separately via different grounded conductor and/or capacitor. In addition, the sensitivities of all proposed circuits are low. From simulation and measurement results it is evident that the final solution corresponds to theoretical expectations and the DBTA appears as very useful active function block for analogue frequency filter and oscillator design.

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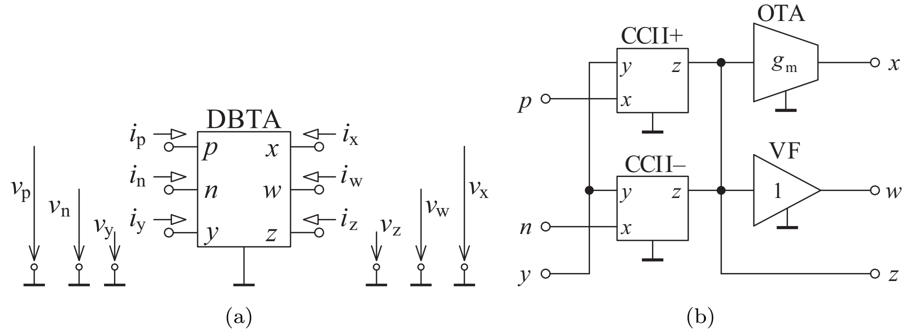


Figure 1. (a) Schematic symbol of DBTA, (b) internal structure of DBTA.

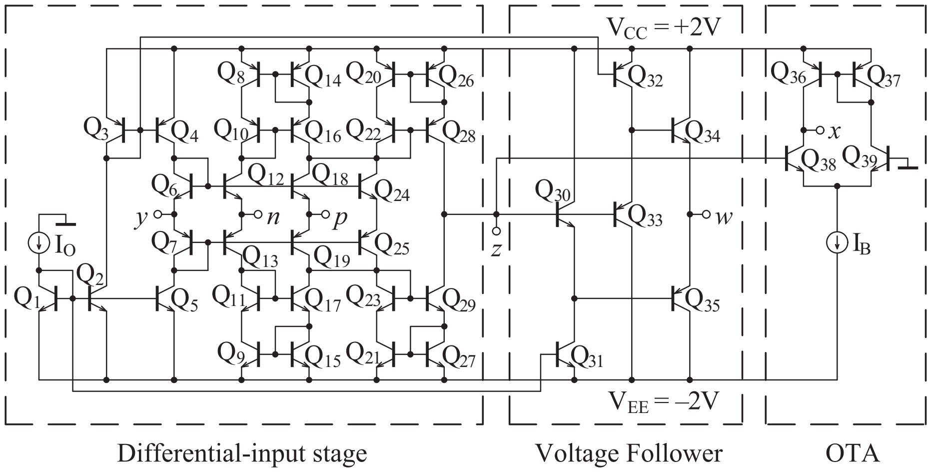


Figure 2. Bipolar implementation of the DBTA.

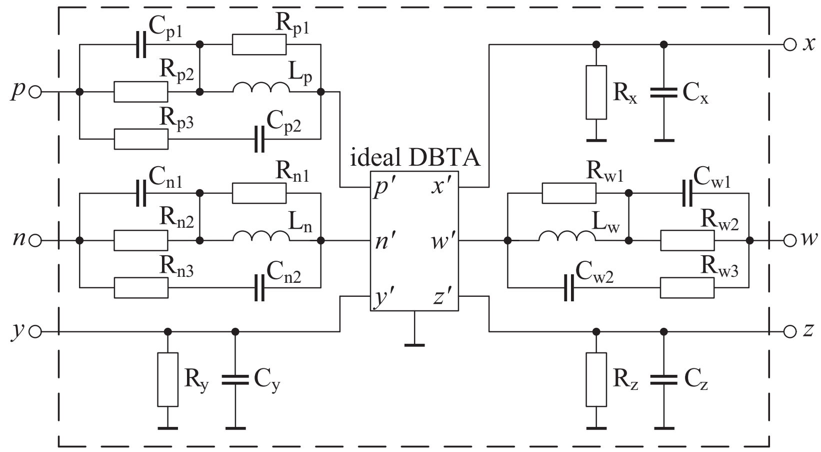


Figure 3. Model of the DBTA including parasitic elements.

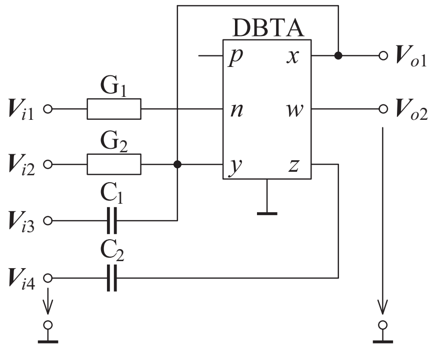


Figure 4. Proposed voltage-mode universal filter using single DBTA.

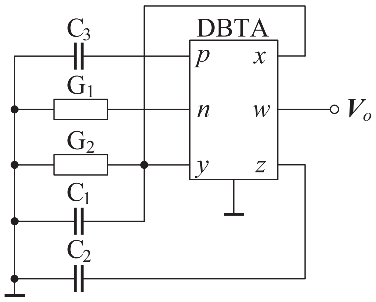


Figure 5. Proposed single DBTA-based sinusoidal oscillator.

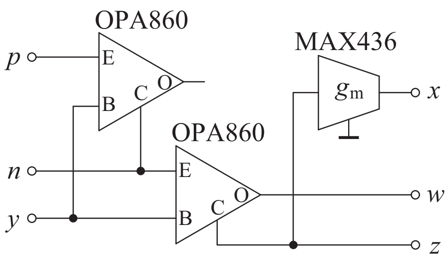
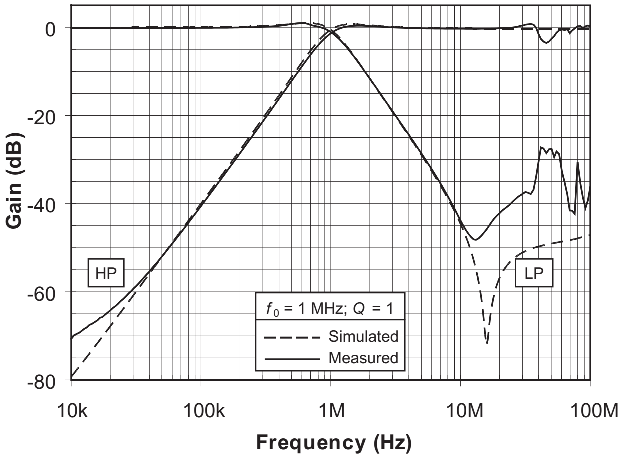


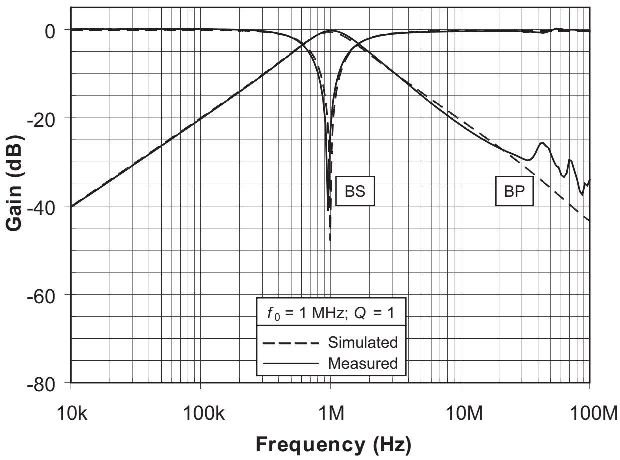
Figure 6. Possible realization of DBTA by commercially available amplifiers OPA860 and MAX436.

Table 1. Total harmonic distortion analysis of the DBTA-based oscillator in Figure 5.

Harmonic no.	Frequency (Hz)	Fourier component	Normalized component	Phase (Deg)	Normalized phase
1	1.592E+04	2.529E–01	1.000E+00	1.049E+02	0.000E+00
2	3.184E+04	3.450E–03	1.364E–02	1.142E+02	–9.565E+01
3	4.776E+04	1.149E–03	4.544E–03	–1.793E+02	–4.941E+02
4	6.368E+04	2.558E–03	1.012E–02	5.882E+01	–3.609E+02
5	7.960E+04	2.759E–03	1.091E–02	3.901E+01	–4.856E+02
DC component = 1.085017E–02					
Total harmonic distortion = 2.069141E+00 PERCENT					

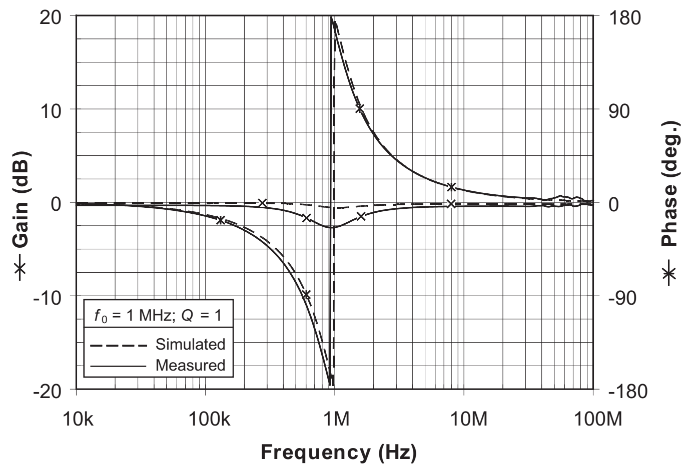


(a)

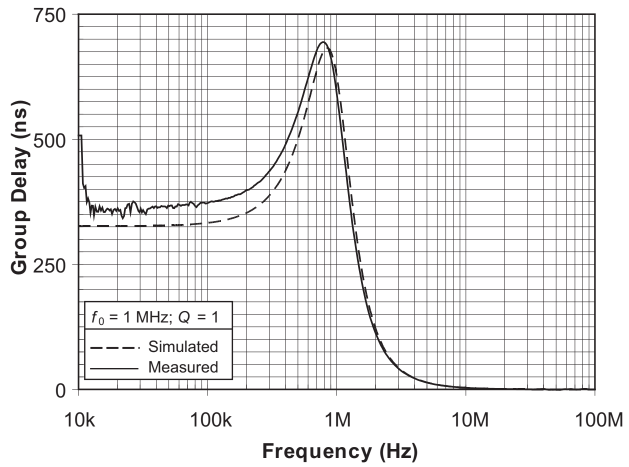


(b)

Figure 7. Simulated and measured frequency characteristics for: (a) LP1 and HP1, (b) BP2 and BS responses of the proposed circuit of Figure 4.



(a)



(b)

Figure 8. Simulated and measured frequency responses of the all-pass (AP) filter: (a) gain and phase responses, (b) group delay response.

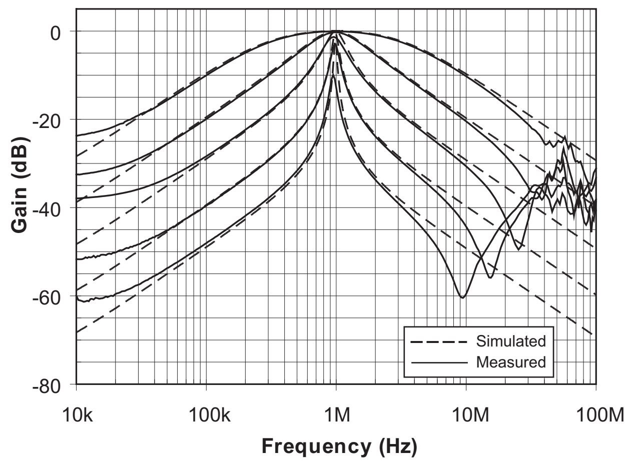
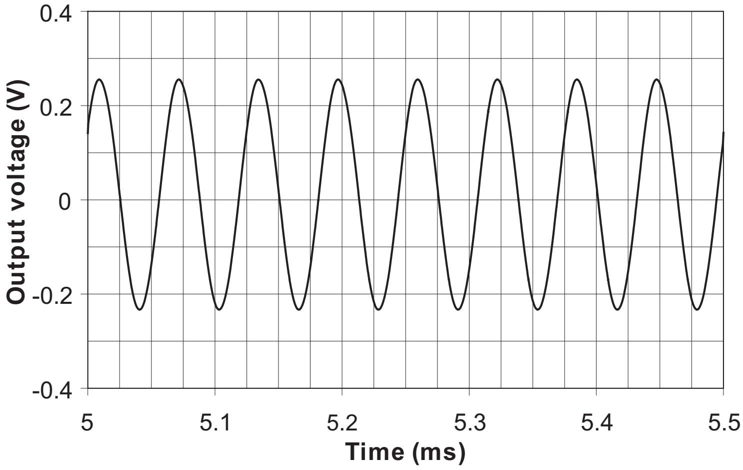
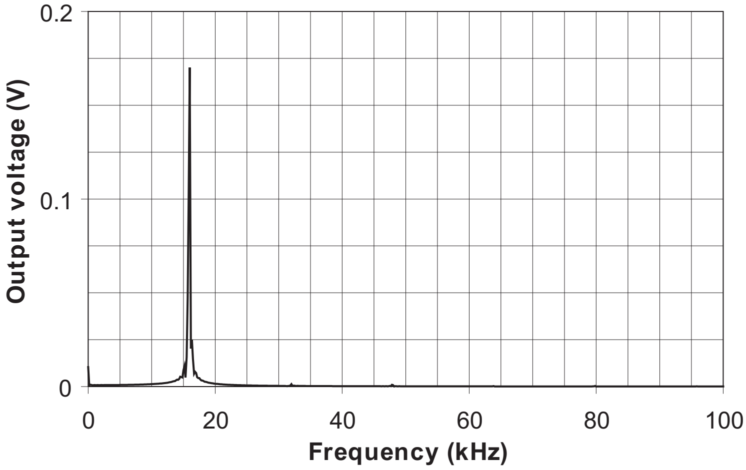


Figure 9. Simulated and measured results of proposed voltage-mode band-pass filter (BP1) for values of quality factor $Q = \{0.3;\ 1;\ 3;\ 10;\ 30\}$.



(a)



(b)

Figure 10. (a) The simulated waveform output and (b) the simulated frequency spectrum of the proposed oscillator in Figure 5.

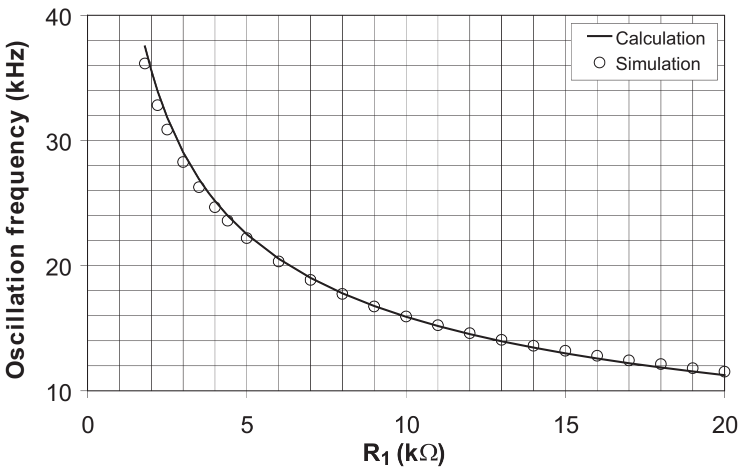


Figure 11. Calculation and simulation results of the oscillation frequencies of $\boldsymbol{V}_o$ by varying the value of the G_1.