Voltage Differencing Buffered/Inverted Amplifiers and Their Applications for Signal Generation

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Abstract. *This paper presents some interesting new applications in the field of analog signal processing focused on signal generation. A novel modifications of recently developed and studied family of active elements, called voltage differencing buffered amplifier (VDBA) and voltage differencing inverted buffered amplifier (VDIBA) are discussed. Our attention is focused on simple application of active elements like dual output VDBAs (DO-VDBAs) and fully balanced VDBAs (FB-VDBAs), where one or two z terminals and always voltage outputs of both polarities are present. The last modification of VDBA allows additional electronic control of voltage gain in frame of active element except standard transconductance control. Discussed active elements were used to build very simple multiphase oscillators with minimal complexity as a simple non-tunable alternative to classical conceptions utilizing lossy integrators in phase-shifted loop. Linearly tunable quadrature differential mode (balanced) oscillator or balanced simple triangle and square wave generator were chosen as other useful examples. Features of proposed circuits are discussed and selected examples verified and evaluated by computer simulations with appropriate low-voltage TSMC 0.18 μm CMOS technology models.*

Keywords

Analog signal processing, low-voltage circuit, voltage differencing inverted buffered amplifier, VDIBA, voltage differencing buffered amplifier, VDBA, multiphase oscillators, quadrature oscillator, triangle and square wave generator, differential output responses, electronic control.

1. Introduction

There are many active elements in the field of analog signal processing, however new ideas in this area help to provide further improvements in order to obtain more effective and interesting circuits. Plenty of novel active elements were introduced by Biolek et al. [1]. Nevertheless, many of them are only hypothetical elements and offer further research mainly from practical point of view. Main aim of this paper is to show simple alternatives allowing multiphase generation, simple quadrature generation and square and triangle wave differential mode generation with help of modifications of novel active elements known as voltage differencing buffered amplifier (VDBA) [1]-[4] or voltage differencing inverted amplifier (VDIBA) [5]. Presented active elements and their modifications allow interesting utilization and design of more profitable or more challenging application (differential mode operation, advanced electronic control, etc.) in comparison to classical VDBA or VDIBA elements.

This paper is divided to three parts. The first part deals with behavioral principles of used active elements. The second part discusses possible applications in analog signal generation (there are three areas: multiphase oscillators, quadrature/differential mode oscillator and simple functional generator). The third part deals with possible structures of discussed active elements that are mostly used in presented designs, their behavior and simulation results, and features of selected of proposed applications (multiphase oscillator and functional generator).

Several approaches to design of multiphase oscillators are available in the open literature. The first way uses integrators or similar selective sections in the loop [6]-[8]. Classical integrator phase shifted loops [6]-[8] were key circuits for generation signals with several phase shifts between them for many years. However, such circuits require many lossy integrators or selective sections (it depends on the number of multiplicand of basic shift - π/2, π/4 or π/6 typically). Such structures were based on classical simple active elements like current conveyors (CC) [6], [8], current amplifiers [7], etc. However, complexity and number of sections required for such type of oscillators is high. It means at least 4 sections (4 capacitors and 4 active elements) for phase shifts 45, 90, 135 and 180 degrees for example in classical loop way of synthesis. The second possibility of construction of multiphase oscillators (unfortunately number of phases is more limited) utilizes all-pass sections (for example [9]-[11]). Here simple (operational amplifiers for example [9]) or more complex active elements like current differencing transconductance amplifiers (CDTAs) have been employed [10]. These already studied circuits usually allow quadrature phase shifts (or multiple of $\pi/2$) [11]. However, they usually do not allow nonstandard phase shifts such as 45, 90, 135 and 180 degrees simultaneously in comparison to classical types employing phase shifted integrator loops. The last used approach is based on combination of active all-pass section with lossy or lossless integrators or differentiators or similar sections. This approach is also in field of our interest in design of multiphase oscillator used in this paper. Unfortunately, our study below shows that not all published oscillator structures have been analyzed carefully in the past and many other circuits with similar features probably exist in the literature. In addition, the analysis of possible outputs of oscillator circuits or analysis of the relation between them is also not provided in many papers. In the following paragraphs we summarized the most important features of approaches used in recent literature and also by us:

a) Phase shifted loops with integrators or similar selective sections and typical examples

Advantages: tunability; easy synthesis; easy CO (condition of oscillation) control by gain of amplifier cascaded in the loop; even and odd phases of $\pi/2$, $\pi/4$ or $\pi/6$.

Disadvantages: many sections (one section = the lowest available phase shift) for required number of phase shifts; higher power consumption (many active elements); in some cases simultaneously matched control of parameter of each elementary transfer in each section - complication of FO (frequency of oscillation) and CO control (some types are controllable by simultaneous changes of each time constant [7], some require special matching conditions [6], [8]).

Abuelmaatti et al. [6] proposed multiphase current-mode oscillator employing controllable current conveyors in lossy integrators. Loop structure utilizes two capacitors per section and FO, CO control require matching condition between capacitor values. Voltage-mode multiphase oscillator based on classical operational amplifiers was presented by Gift [9] where all-pass sections with adjustable time constants (by R and C values only) were used for construction. However, number of passive elements seems to be very high for large number of phase shifts. Souliotis et al. utilized current amplifiers in cascade of lossy integrators creating current-mode multiphase oscillator [7]. Adjusting of FO is quite simple by bias current of active elements. Kumngern et al. [8] also built their current-mode oscillator with lossy integrators utilizing current conveyors, where intrinsic resistance of the x-terminal and gain between z and x terminal simultaneously is possible. However, matching between capacitors is also required for realization of higher number of phases and therefore accuracy of such matching condition influences also accuracy of the phase shift.

b) All-pass section in the loop based oscillators and typical examples

Advantages: simpler circuits; lower number of passive and active elements (in many cases two sections are sufficient [10]); lower power consumption.

Disadvantages: complicated tunability or CO control and matching in parameter values is required in many cases; accuracy of relation (transfers) between available outputs for accurate phase shifts required; phase shifts are available as multiple of $\pi/2$ almost in all cases.

Keskin et al. [10] published the perfect example of oscillator based on two all-pass sections employing two CDTAs and 6 passive elements that serve for CO and FO control. The oscillator produces output signals also in form of currents. The discussed circuit provides quadrature phase shifts only. Similarly, Songsuwankit et al. [11] also deal with phase shifter-based (all-pass section consists of three OTAs and one floating capacitor) oscillator design. Here, arbitrary setting of phase shift is possible, but the proposed oscillator provides only two outputs.

c) All-pass sections in combination with integrators or integrators/differentiators in simple loops and typical examples

Advantages: similar to the previous group; produced phase shifts are available similarly as in phase shifted loop oscillators (it depends on construction of particular circuit); CO control independent on FO in some cases.

Disadvantages: similar to the previous discussion; if FO is controllable by specific parameter it influences phase or at least amplitude relations.

Keawon et al. [12] proposed an oscillator employing single current controlled current differencing transconductance amplifier (CCCDTA), where two capacitors are required. Control of FO and CO is established by intrinsic resistance and tranconductance control. Output signals are in form of currents. Jaikla et al. [13] designed a circuit utilizing single CDTA and three passive elements. The independent CO and FO control is not possible and their adjustment is possible only by passive elements (also floating capacitor). Quadrature oscillator produces signals in form of currents. Also Pandey et al. [14] presented a circuit, which produces currents in quadrature phase shift with two CDTAs and two passive elements, where control is given by transconductances. Herencsar et al. [5] also discussed an oscillator, where one simple all-pass section and lossy integrator employing two VDIBA elements and three passive elements with control of CO were used. Keskin et al. [15] also presented an oscillator producing voltage signals, which is based on two current differencing buffered amplifiers (CDBAs) in which 8 passive elements are required and independent control of CO is difficult. Songkla et al. [16] presented an oscillator, where three current controlled current conveyors (controlled by intrinsic resistance) of second generation and two passive elements generate current output signals. Minaei et al. [17] utilized differential voltage current conveyor (DVCC) and the circuit requires three DVCCs, four passive elements, and produces voltage output responses. All discussed examples provide quadrature outputs. None of them show production of phase shifts such as 45, 90, 135, 180 degrees, which is the main contribution of our circuits.

General conclusions from above discussions are that tunability of cascaded phase shifted loop is better, but complexity is several times higher. Therefore, simpler structures that utilize another design approaches are more interesting in some cases. Therefore, our intention in this paper is to design simpler solutions that have not been developed so far.

The next part of our contribution deals with quadrature oscillator design. Important quadrature oscillators with independent CO and FO adjusting, grounded capacitors, and voltage output responses are compared in the following text.

Many from recently investigated structures use passive elements for control of CO and FO, therefore their replacement by electronically adjustable equivalents is necessary. Soliman [18] utilizes current conveyors (two or three) and 5-6 passive elements in solution of an oscillator, where differential output signals are not easily available. Oscillator presented by Herencsar et al. based on generalized current follower transconductance amplifier (GCFTA) and voltage buffer employs two active and four passive elements. However, it does not provide linear control of FO (produced amplitude is dependent on tuning process) [19]. Tuning is realized by changes of passive elements. Similar type of FO and CO control is used in works [20]-[22] as well. Gupta et al. [20] developed an oscillator based on current and voltage followers as active elements (2-4 in the proposed circuits) and 5 passive elements. Amplitude of generated signal was influenced by FO adjusting without possibility of linear control. Oscillator with differential output signals and quadrature phase shift is proposed by Biolkova et al. in [21]. The circuit is based on two dual output current inverter buffered amplifiers (DO-CIBAs) and 5-6 passive elements. The FO control without influence on produced amplitudes is linear, but simultaneous change of floating resistor values is necessary. Lahiri [22] proposed an oscillator employing three current feedback amplifiers (CFAs) and 6 passive elements, where adjusting of resistor values is the only way how to control CO and FO. Lahiri et al. [23] also proposed another oscillator based on single current conveyor transconductance amplifier (CCTA) and four passive elements, where electronic control is realized by g_m. Nevertheless, generated amplitude is influenced by tuning of FO and dependence of control is not linear. The same features were achieved in solution presented in [24], where two CDTAs and three passive elements were used. Rodriguez-Vazquez et al. [25] presented an oscillator based on 3-4 transconductors (OTAs) and two capacitors. Linear control of FO without influence on generated amplitudes is possible electronically by transconductances. The oscillator in [26] utilizes special configuration of three specially modified CFAs and five passive elements, which allows linear control of FO without impact on generated amplitudes. Digital control of such application has been also discussed in the past, for example by Alzaher et al. [27], where 3 or 5 active elements (current amplifiers and voltage buffers) and 6 passive elements were used and linear control of FO without impact on generated amplitudes is allowed. Interesting digitally controllable solution was presented by Biolek et al. [28], where two active elements - so called z-copy controlled gain current differencing buffered amplifiers (ZC-CG-CDBAs), five passive elements are used and allow linear control of FO. Several solutions of quadrature oscillators, where control procedure was focused on current and voltage gain adjusting in frame of active elements called controlled gain current follower differential output buffered amplifiers (CG-CFDOBA), controlled gain current inverter buffered amplifiers (CG-CIBAs) and controlled gain current amplified voltage amplifier (CG-CVA), are discussed in [29] (five passive and two active elements are used in the proposed oscillators). Some of them allow differential output responses and linear control of FO. Galan et al. [30] utilizes four dual output OTAs (g_m control) and four capacitors to achieve fully differential output oscillator. The first note about utilization of FB-VDBA in differential (balanced) quadrature was discussed by Bajer et al. [31]. The oscillator consists of two FB-VDBAs, two resistors, four grounded capacitors and allows linear electronic control of FO and independent control of CO (unfortunately by a floating resistor).

Oscillator structure presented in this contribution seems to be very economical in comparison to above discussed circuits, because only single DO-VDBA and dual output controlled gain voltage differencing buffered voltage amplifier (DO-CG-VDBVA) and three passive elements (only one of them is floating) are required. The proposed oscillator provides differential outputs, linear control of FO by simultaneous adjusting of both g_m, and CO control by adjustable voltage gain. In some above discussed solutions utilizing other type of active elements the potential possibility to obtain voltage or even voltage differential (balanced) outputs also exists, but additional current to voltage conversion or voltage buffering/inversion (out of active element) is required. However, if it is possible, than these discussed oscillators have other drawbacks such as control by passive elements [21] or high number of active elements [30].

Principle of triangle and square generator (called functional) is known very well. There were interesting attempts to build very simple generator from active elements like basic operational amplifiers, current conveyors, etc. [1]. However, these simple circuits have some drawbacks (many passive elements) and lack of electronic controllability. The following discussion deals with typical examples.

Biolek et al. [32] introduced an interesting circuit with single CDTA, 3 resistors and one capacitor which provides FO in range of MHz tunable by resistor value. De Marcellis et al. [33] proposed generator employing 2 CCIIs, 6 resistors and one capacitor. Control of FO is also possible by adjusting of resistor value. Chien et al. [34] presented solution based on two differential voltage current conveyors (DVCCs) [1], three resistors and control of duty cycle is also allowed. Almashary et al. [35] presented a generator, where 2 CCIIs, 3 resistors, and 2 capacitors are required. Tunability of FO is possible by value of resistor. Pal et al. [36] employed two CCIIs, three resistors and floating capacitor in his approach. Two current feedback operational amplifiers (CFOAs) [1] with two capacitors and two resistor utilized generator presented by Saque et al. [37]. Minaei et al. [38] introduced similar approach based on CFOAs and DVCCs. Two operational transresistance amplifiers (OTRAs), three resistors and one floating capacitor based generator is shown by Lo et al. [39].

Electronically controllable active elements allow better performance in these types of generators. Works [40]-[42] brought key information for design of generators with transconductance (OTA) sections (controlled by DC bias currents) and special comparators with hysteresis (so called Schmitt trigger [40]), for example). Our presented topology of triangular and square wave generator is based on similar principles. Kim et al. [41] and Chung et al. [42] utilized 3 OTA sections, one capacitor and two resistors, similarly Siripruchynanun et al. [43].

Several generators were designed with current-mode outputs. Kumbun et al. [44] employed two multiple outputs through transconductance amplifiers (MO-CTTAs) only to realize adjustable generator. Two multiple output current controlled current differencing amplifiers (MO-CCCDTAs) were used for design of generator by Silapan et al. [45] and Sristakul et al. [46]. Silapan et al. [45] published a very excellent work (fundamentally very similar to our solution), where two controllable multi-output CDTA (MO-CCCDTA) elements, and one capacitor is sufficient. In fact, there were used two independently adjustable OTA sections in each MO-CCCDTA and it allows to "integrate" resistor inside of the active element. The generator is designed with current output responses.

In our topology only two OTA sections, one capacitor and one resistor are sufficient (reasons for the second are explained in a specific chapter - Schmitt comparator in our contribution is adapted for differential output purposes). Differential output means two times higher output signals and immunity to common mode disturbances, which is really important in modern low-voltage CMOS technologies. The resulting conclusion from hitherto published works is clear - to the best of the authors' knowledge no generator (with electronic control of repeating frequency and duty cycle) for differential (balanced output) signal generation that is simpler exists in the literature and many presented single-ended solutions seem to be quite complicated and use also floating passive elements ([33], [36] for example).

2. Controllable Voltage Differencing Buffered Amplifiers

Conceptions of VDBA/VDIBA [1]-[5] and their behavioral model are discussed in this chapter. Advantageous differences from classical VDBA [1]-[4] or VDIBA approach based on OTA and inverter section [5] are discussed and explained. Classical VDBA/VDIBA employs high-impedance voltage differencing input terminals (in this paper labeled as *p* and *n*), auxiliary high-impedance *z* terminal, and low-impedance output of voltage buffer/inverter noted as +*w*/-*w*. Modified conception uses transconductance section with one output polarity (one *z* terminal) and cascade of two voltage inverters. Therefore, both inverting (-*w*) and direct voltage outputs (+*w*) are available. However, some applications require two auxiliary terminals *z* for special purposes of circuit synthesis. Internal structure of FB-VDBA [2] is more complicated, because different transconductance section with current mirrors is required. Advanced VDBA/VDIBA modification allows several types of electronic control as will be shown in more details.

2.1 DO-VDBA

The first possible modification of VDBA/VDIBA [1]-[6] is created very simply by additional voltage inverter. Therefore, we have now also direct voltage buffered output, which is very useful for differential mode signal operations. Symbol and behavioral model is shown in Fig. 1.

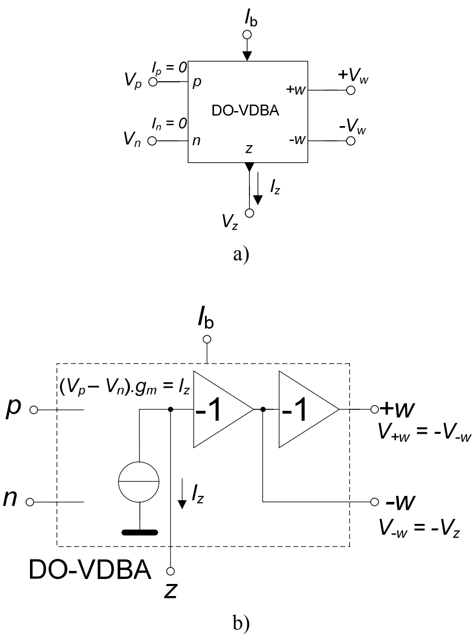


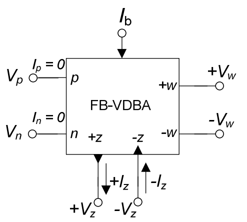
Fig. 1. Dual output voltage differencing buffered amplifier (DO-VDBA): a) symbol, b) behavioral model.

We called this element dual output voltage differencing buffered amplifier (DO-VDBA) in accordance to [1]. The main principle is obvious from behavioral model (see Fig. 1b). Control of transconductance is possible by external biasing (I_b). Relation between terminals can be written in hybrid matrix as follows:

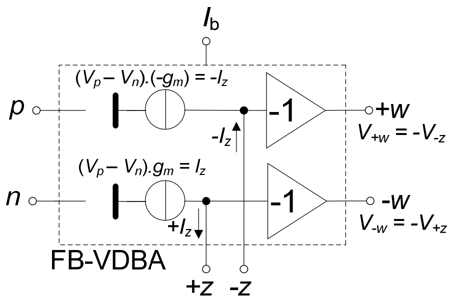
$$\begin{bmatrix} I_p \\ I_n \\ I_z \\ V_{+w} \\ V_{-w} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ g_m & -g_m & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & -1 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} V_p \\ V_n \\ V_z \\ I_{+w} \\ I_{-w} \end{bmatrix}. \quad (1)$$

2.2 FB-VDBA

In some cases of circuit synthesis two auxiliary z terminals (both polarities for example) are required. A very useful and easily obtainable version [2], [3] is shown in Fig. 2.



a)



b)

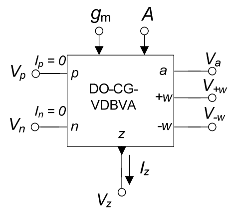
Fig. 2. Fully balanced voltage differencing buffered amplifier (FB-VDBA): a) symbol, b) behavioral model.

$$\begin{bmatrix} I_p \\ I_n \\ I_{+z} \\ I_{-z} \\ V_{+w} \\ V_{-w} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ g_m & -g_m & 0 & 0 & 0 & 0 \\ -g_m & g_m & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 & 0 & 0 \\ 0 & 0 & -1 & 0 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} V_p \\ V_n \\ V_{+z} \\ V_{-z} \\ I_{+w} \\ I_{-w} \end{bmatrix}. \quad (2)$$

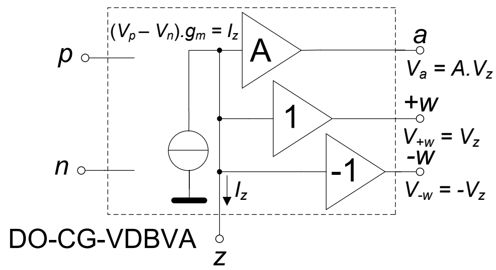
Matrix description is very similar to (1), but main difference is in additional equation for the second output current from the second z terminal, see (2).

2.3 DO-CG-VDBVA

Previous types of VDBAs (DO-VDBA and FB-VDBA) [1]-[6] allow only one possibility of electronic control. However, in many applications various possibilities of controls are required. As an example, sinusoidal oscillators can be mentioned, where one controllable parameter serves for control of oscillation condition and the second one adjusts oscillation frequency. Therefore, introduced modification contains two adjustable parameters. These parameters can be obtained by additional voltage amplifier together with buffer/inverter in behavioral structure or by replacement of inverter/buffer by this controllable voltage amplifier. This active element received a typical name consisting of main features typical for classical VDBA or VDIBA [1]-[6] but considering also controllability of voltage gain (A). Symbol and behavioral model of the dual output controlled gain voltage differencing buffered voltage amplifier (DO-CG-VDBVA) is depicted in Fig. 3.



a)



b)

Fig. 3. Dual output controlled gain voltage differencing buffered voltage amplifier: a) symbol, b) behavioral model.

Matrix equations also include a new parameter - adjustable voltage gain as follows:

$$\begin{bmatrix} I_p \\ I_n \\ I_z \\ V_a \\ V_{+w} \\ V_{-w} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ g_m & -g_m & 0 & 0 & 0 & 0 \\ 0 & 0 & A & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & -1 & 0 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} V_p \\ V_n \\ V_z \\ I_a \\ I_{+w} \\ I_{-w} \end{bmatrix}, \quad (3)$$

from which it is obvious that direct voltage buffering can be achieved by a simple way shown in Fig. 1 (two inverters in cascade).

3. Applications for Signal Generation

The discussed active elements can be used in interesting applications in analog signal processing and mainly in signal generation. Several types of sinusoidal oscillators and functional (triangle and square wave) generators based on the discussed active elements are presented in this section.

3.1 Multiphase Harmonic Oscillator SRCO

So-called single resistance controllable oscillators (SRCO) [20] were very popular for many years due to their simplicity (one-two active elements maximally) and simultaneously independent control of CO and FO by resistor values. The basic conception employs lossy integrator (R_1, C_1), one DO-VDBA and special impedance constructed from -R_2 and C_2) in an interesting circuit configuration (Fig. 4).

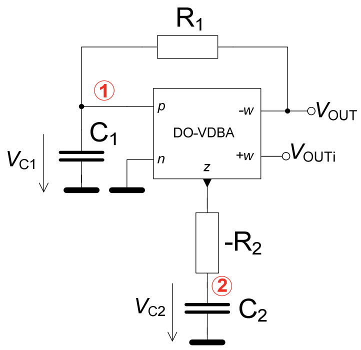


Fig. 4. Simple multiphase oscillator using single DO-VDBA.

Characteristic equation of the discussed oscillator has form:

$$s^2+\frac{(1-g_mR_2)}{R_1C_1}s+\frac{g_m}{R_1C_1C_2}=0, \tag{4}$$

where CO is controllable by -R_2 value and FO by value of R_1. Therefore, specifications for SRCO type of oscillator are fulfilled. Practical utilization of floating resistance -R_2 is different, because Fig. 4 serves only for theoretical and simple explanation of the principle. Controllability of -R_2 value in order to ensure soft CO control is not very comfortable. Therefore, we implemented the second active element to replace inconvenient floating negative passive element from another FB-VDBA. Transconductance control now realizes CO adjusting by a simple electronic way. Replacement of the impedance connected to z terminal of DO-VDBA is shown in Fig. 5. The principle of the design of the basic (Fig. 4) and modified circuit is clear from the diagram in Fig. 5. Two blocks with specific transfers (lossy integrator and special synthetic function) between current and voltage (transadmitance Y_T and transimpedance Z_T) were used for synthesis of the discussed circuit. Impedance of serial R_2C_2 combination in Fig. 5 has form:

$$Z_{INP\_RC}=\frac{1-sR_2C_2}{sC_2}, \tag{5}$$

and from equivalent circuit resulting impedance:

$$Z_{INP\_EQ}=\frac{1-s\dfrac{C_2}{g_{m2}}}{sC_2}. \tag{6}$$

The improved characteristic equation is now:

$$s^2+\frac{g_{m2}-g_{m1}}{R_1C_1g_{m2}}s+\frac{g_{m1}}{R_1C_1C_2}=0 \tag{7}$$

where CO and FO can be easily obtained as:

$$g_{m2}\le g_{m1},\quad \omega_0=\sqrt{\frac{g_{m1}}{R_1C_1C_2}}. \tag{8), (9}$$

We can determine transfers between outputs of the oscillator and get relations between produced amplitudes and phase shifts respectively as:

$$\frac{V_{OUT\_0}}{V_{C1}}=\frac{g_{m1}\left(-1+s\dfrac{C_2}{g_{m2}}\right)}{sC_2}, \tag{10}$$

$$\frac{V_{OUT\_0}}{V_{C2}}=-\frac{1+sC_1R_1}{1-\dfrac{g_{m1}}{g_{m2}}+sC_1R_1}, \tag{11}$$

$$\frac{V_{C2}}{V_{C1}}=-\frac{g_{m1}}{sC_2}. \tag{12}$$

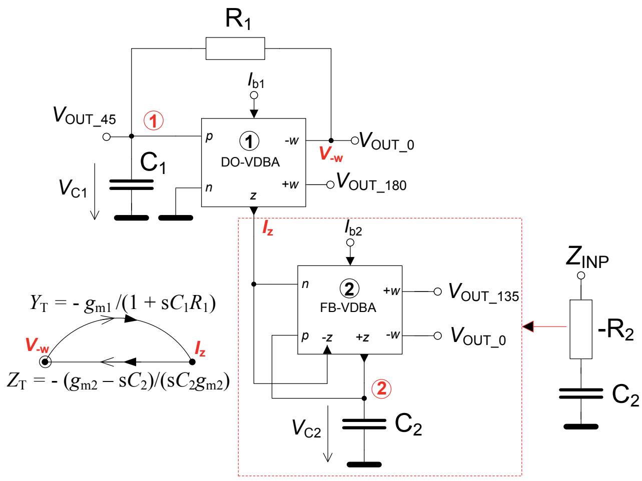


Fig. 5. Modification of the oscillator in Fig. 4 with electronically controllable CO.

We can simplify these equations considering fulfilled CO (8) and equality of $C_1 = C_2 = C$ and determine relations as:

$$\frac{V_{OUT\_0}}{V_{C1}}=\frac{g_{m1}(-1+sC)}{sC}, \tag{13}$$

$$V_{OUT\_0}=\sqrt{1+g_{m1}R_1}\,e^{\sqrt{g_{m1}R_1}\,j}V_{C1}, \tag{14}$$

$$\varphi_{VOUT\_0\_C1} = \mathrm{arctg}\left(\sqrt{g_{m1}R_1}\right), \tag{15}$$

which leads to 45° phase shift in case of equality $g_{m1} = 1/R_1$. The next relation has form:

$$\frac{V_{OUT\_0}}{V_{C2}} = -\frac{1+sCR_1}{sCR_1}, \tag{16}$$

$$V_{OUT\_0} = -\sqrt{1+\frac{1}{g_{m1}R_1}}\,\mathrm{e}^{-\frac{1}{\sqrt{g_{m1}R_1}}j}V_{C2}, \tag{17}$$

$$\varphi_{VOUT\_0\_C2} = 180 + \mathrm{arctg}\left(-\frac{1}{\sqrt{g_{m1}R_1}}\right), \tag{18}$$

which leads to 135° for fulfilled $g_{m1} = 1/R_1$. The last relation that we can find between voltages across capacitors is:

$$\frac{V_{C2}}{V_{C1}} = -\frac{g_{m1}}{sC}, \tag{19}$$

$$V_{C2} = \sqrt{g_m R_1}\,e^{-\frac{\pi}{2}j}V_{C1}, \tag{20}$$

$$\varphi_{C2\_C1} = \mathrm{arctg}\left(\frac{\sqrt{g_m R_1}}{0}\right) = \frac{\pi}{2} \tag{21}$$

where only amplitude relation depends on R_1 value (FO control). We can find out that the proposed oscillator produces signals with phase shifts 45, 90, 135 and 180 degrees for fulfilled CO and $g_{m1} = 1/R_1$ as conclusion of this analysis.

This type of oscillators is not very suitable for FO adjusting, if multiphase outputs are also required. Any change of FO leads to inequality of $g_{m1} = 1/R_1$ through R_1 and causes disturbance of phase and amplitude proportions in the circuit. Only phase relation between V_{C1} and V_{C2} keeps preserved (but not amplitude). Additional voltage buffering of V_{OUT_45} is also a complication for practical utilization. However, obtaining of four-phase outputs is possible by quite a simple way (similarly as in [5], where one disadvantage remains – floating capacitor) without necessity of many active elements in comparison to [6]-[8], for example. Therefore, the presented solution offers some benefits and improvements.

3.2 Multiphase Oscillator - Special Requirements

The previous type of the oscillator requires synthetic replacement of floating negative resistance. Therefore, in Fig. 6 another solution with similar phase shifts like the above discussed circuit is introduced, but usage of floating negative resistance is not required. This oscillator uses conversion ($V \to I$, $I \to V$) transfer sections in the loop based on only lossy integrators and additional positive feedback in comparison to the previous type (diagram in Fig. 6). The second DO-VDBA$_2$ with R_3 serves as a subtracter, which allows to obtain required outputs with appropriate phase shifts. Characteristic equation, CO and elementary FO have forms:

$$s^2 + \frac{R_2C_2 + R_1C_1(1-R_2g_{m1})}{R_1R_2C_1C_2}s + \frac{1}{R_1R_2C_1C_2} = 0, \tag{22}$$

$$g_{m1} \geq \frac{R_1C_1 + R_2C_2}{R_1R_2C_1}, \quad \omega_0 = \sqrt{\frac{1}{R_1R_2C_1C_2}}. \tag{23), (24}$$

Relations between generated signals are:

$$\frac{V_{C1}}{V_{C2}} = \frac{1}{1+sC_1R_1}, \tag{25}$$

$$\frac{V_{OUT\_135}}{V_{C1}} = R_3g_{m2}\left(\frac{1+sC_2R_2}{-R_2g_{m1}+1+sC_2R_2}\right), \tag{26}$$

$$\frac{V_{OUT\_135}}{V_{C2}} = -R_3g_{m2}\left(\frac{sC_1R_1}{1+sC_1R_1}\right), \tag{27}$$

which leads to (considering $C_1 = C_2 = C$, fulfilled CO: $R_2g_{m1} = 2$ and $R_3g_{m2} = 1$):

$$V_{C1} = \frac{\sqrt{2}}{2}\mathrm{e}^{-\frac{\pi}{4}j}V_{C2}, \tag{28}$$

$$V_{OUT\_135} = \mathrm{e}^{-\frac{\pi}{2}j}V_{C1}, \tag{29}$$

$$V_{OUT\_135} = -\frac{\sqrt{2}}{2}\mathrm{e}^{-\frac{\pi}{4}j}V_{C2} = \frac{\sqrt{2}}{2}\mathrm{e}^{-\frac{3\pi}{4}j}V_{C2}. \tag{30}$$

Simple solution of both DO-VDBAs (only one positive z terminal) is sufficient in this modification of the oscillator in comparison to the previous type.

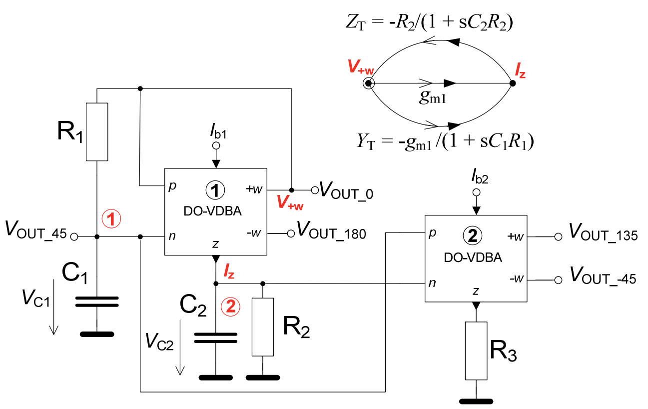


Fig. 6. Modification of the proposed oscillator without necessity of employing floating negative resistance.

3.3 Adjustable Harmonic Quadrature Oscillator

The solution presented in this section offers more benefits in comparison to the previous multiphase type, where FO adjusting was limited or not possible due to request of four multiphase outputs (45, 90, 135, 180 degrees). The main requirement of the intended synthesis is the design of fully electronically controllable (CO and linear control of FO) oscillator with multiphase purposes (90, 180, 270 degrees) or quadrature differential (balanced) mode oscillator. The presented circuit consists of lossy and loss-less integrators in one loop complemented by negative resistance controllable by voltage gain. This application (Fig. 7) is the perfect example of utilization of the DO-CG-VDBVA. Loss-less controllable (by g_{m2}) voltage integrator forms the first part (C_2 and DO-VDBA) of the oscillator. The second important part is the lossy integrator (controllable by g_{m1}) together with negative resistance supplementary simulating circuit (R_1, C_1 and DO-CG-VDBVA), where negative resistance is adjustable by voltage gain A_1. Characteristic equation, CO and FO have the following forms:

$$s^2 + \frac{(1-A_1)}{R_1 C_1} s + \frac{g_{m1} g_{m2}}{C_1 C_2} = 0, \quad (31)$$

$$A_1 \geq 1, \quad \omega_0 = \sqrt{\frac{g_{m1} g_{m2}}{C_1 C_2}}. \quad (32), (33)$$

Relative sensitivities of oscillation frequency on parameters in (33) achieve typical values (± 0.5). Linear control of FO is ensured by simultaneous adjusting of g_{m1} and g_{m2} ($g_{m1} = g_{m2}$) and independent control of CO by A_1. The oscillator provides low impedance at each of the outputs, therefore, easy connection to low-resistance loads is allowed. The relation between produced signals across the capacitors has form:

$$\frac{V_{C2}}{V_{C1}} = -\frac{g_{m2}}{sC_2} = j\sqrt{\frac{g_{m2} C_1}{g_{m1} C_2}}, \quad (34)$$

which means quadrature phase shift and equal amplitudes during the tuning process (FO) at all outputs in case of simultaneous control of both g_m and when capacitors have equal values. Single-ended operation mode allows to obtain the oscillator with unchangeable amplitudes providing four phase shifts (multiples of π/2). Differential operation mode has benefit of two-times higher output amplitudes, but in quadrature form only.

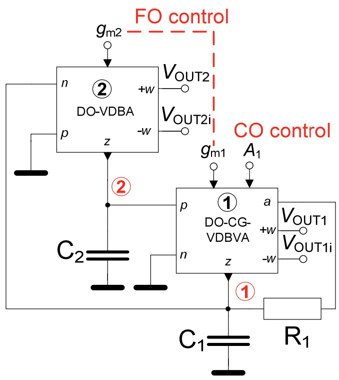


Fig. 7. Fully controllable multiphase/quadrature differential mode (balanced) oscillator.

3.4 Triangle and Square Wave Generator

The last presented useful and quite simple application of the discussed active elements is a differential output (balanced) triangle and square wave generator. Explanations of processes in the function of the generator are provided by charging of a capacitor and switching (turn-over) at reference levels. The loss-less integrator and comparator with hysteresis (Schmitt trigger) in the feedback loop are main building blocks of these types of non-harmonic signal sources [32]-[46]. We show possibilities how to build loss-less integrator with VDBA elements (partial block of the oscillator in Fig. 7).

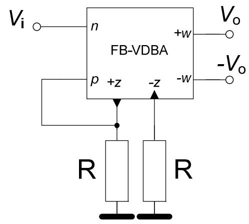


Fig. 8. Schmitt comparator with hysteresis employing FB-VDBA.

For construction of a dual-output Schmitt comparator (Fig. 8) one FB-VDBA with two *z* terminals was employed. Theoretically, the DO-VDBA (one *z* terminal) is also possible for construction of a comparator (we can save one resistor). However, the FB-VDBA (with current multiplying performed by current mirrors in internal topology - we will present it in experimental/simulation part of this paper) allows higher gain of the whole voltage feedback system, which is the key factor (in low-power and low-supply voltage technology) for precise flip-over process of the comparator. Quality of the comparator has a direct impact on accuracy of oscillation/repeating frequency (also noted by abbreviation FO) and accuracy of reference levels.

The following equation is valid between the input and output voltage of FB-VDBA in the comparator (Fig. 8):

$$(\mp V_{o\_sat}) = \frac{g_m R}{g_m R - 1}(\pm V_i), \quad (35)$$

which leads to

$$\mp V_{o\_sat} = \pm V_i, \quad (36)$$

if validity of $g_m R >> 1$ is ensured. We found referencing value (V_{ref}) for very high gains (given by transconductance g_m and resistor R) which is necessary for turnover of the output of the comparator from a high output voltage level to a low output voltage level respectively in (36). The output voltage (in ideal case also reference threshold voltage) is given by maximum output current $\pm I_{+z\_max}$, which means maximum of positive or negative output saturation: $\pm V_{+z} = \pm R.I_{+z\_max}$. Dynamical characteristic of the comparator has two comparative reference voltages determined as $\pm R.I_{+z\_max}$, thanks to positive feedback from voltage across R and very high gain ($g_m R$).

Fig. 9 shows the complete generator, where the loss-less integrator employs the first type of DO-VDBA and the comparator presents the same circuit as we discussed in Fig. 8.

Linear charging of the capacitor C starts at the negative reference voltage level ($-V_{C\_max}$) and is given by $+I_{C\_max} = +I_{z1\_max} + I_d = k_1.I_{b1} + n.k_1.I_{b1}$, where constant k represents current gain in the internal topology of VDBA (multiplying by current mirrors), I_d is auxiliary controlled DC current and n is the ratio between I_d and I_{b1} ($I_d = n.I_{b1}$). The input voltage linear range of DO-VDBA is very small for higher g_m and the slope of the DC transfer characteristic very sharp.

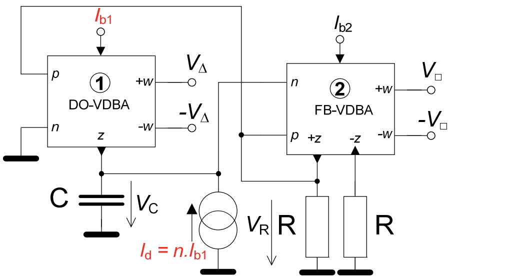


Fig. 9. Differential mode triangle and square wave generator employing DO-VDBA and FB-VDBA.

Therefore, the output currents are given by one from both saturation corners limited by bias current I_b as $\pm I_{z1\_max} = \pm k_1.I_{b1}$ (DO-VDBA) and $\pm I_{z2\_max} = \pm k_2.I_{b2}$ (FB-VDBA), see Fig. 11a and Fig. 13a. Discharging that starts at $+V_{C\_max}$ is given similarly, because $-I_{C\_max} = -I_{z1\_max} + I_d$. Both time intervals per one signal period are obtained from:

$$\Delta V_C = \frac{I_{z1\_max} + I_d}{C} T_1, \quad (37)$$

$$\Delta V_C = \frac{-\left(-I_{z1\_max} + I_d\right)}{C} T_2. \quad (38)$$

The distance between the negative and positive threshold values ($-V_{C\_max} = +RI_{z2\_max}$ and $+V_{C\_max} = -R.I_{z2\_max}$) can be expressed as:

$$\Delta V_C = \Delta V_R = +V_{C\_max} - (-V_{C\_max}) = 2V_{C\_max}, \quad (39)$$

$$2V_{R\_max} = +RI_{z2\_max} - (-RI_{z2\_max}). \quad (40)$$

Both time intervals are stated as:

$$T_1 = \frac{\Delta V_C C}{I_{z1\_max} + I_d} = \frac{2RCk_2 I_{b2}}{k_1 I_{b1} + n.k_1 I_{b1}}, \quad (41)$$

$$T_2 = \frac{\Delta V_C C}{I_{z1\_max} - I_d} = \frac{2RCk_2 I_{b2}}{k_1 I_{b1} - n.k_1 I_{b1}}. \quad (42)$$

The repeating period and frequency have forms:

$$T = T_1 + T_2 = \frac{4RCk_1 I_{b1} k_2 I_{b2}}{(k_1 I_{b1} + n.k_1 I_{b1})(k_1 I_{b1} - n.k_1 I_{b1})}, \quad (43)$$

$$f_0 = \frac{k_1 I_{b1}(1+n)(1-n)}{4RCk_2 I_{b2}} = \frac{k_1 I_{b1}(1-D)D}{RCk_2 I_{b2}}. \quad (44)$$

It is obvious that current I_d is DC component, which shifts linear trace (triangular signal), i.e. offset. It influences the duty cycle of the produced wave as:

$$D = \frac{T_1}{T} = \frac{1}{2}\left(1 - \frac{I_d}{I_{b1}}\right) = \frac{1}{2}(1-n). \quad (45)$$

The maximal theoretical limits of I_d are given by $\pm I_{b1}$ ($n = 1$ for $D = 0\%$ and $n = -1$ for $D = 100\%$). Therefore, change of the polarity of I_d is required. The repeating frequency can be controlled independently with respect to the duty cycle if ratio $I_d/I_{b1} = n$ is kept strictly constant while frequency f_0 is tuned by I_{b1}.

4. Simulation Results

4.1 Possible CMOS Implementations of Selected DO-VDBA Solutions

We designed internal topologies of DO-VDBA and FB-VDBA to demonstrate functionality and practical features of the designed applications. Models of TSMC 0.18 µm CMOS technology parameters [47] were used for our simulations. The first topology uses classical transconductance section with active load (PMOS mirror) and two simple voltage inverters, see Fig. 10. Lower gain of one transconductance section (higher gain requires cascading) and only one z terminal are the main disadvantages of this solution. Nevertheless, it is a useful solution in a specific situation and power consumption is lower in comparison to the second solution (it will be discussed later). PSpice analyses provided the following results ($I_b = 50$ µA, $V_{CC} = \pm 1.2$ V): $g_m = 500$ µS, $R_z \approx 170$ kΩ, $C_z \approx 0.35$ pF, $R_{\pm w} \approx 53$ Ω, voltage gain between z and $+w$ has value 0.926 and gain between z and $-w$ is 0.962. Differential input resistance of OTA section is high - at 1 MHz has values higher than 1.1 MΩ (parasitic capacitance ≈ 0.27 pF). Adjusting of I_b influences R_z value, for $I_b = 200$ µA R_z value is approx. 46 kΩ (value decreases with higher I_b).

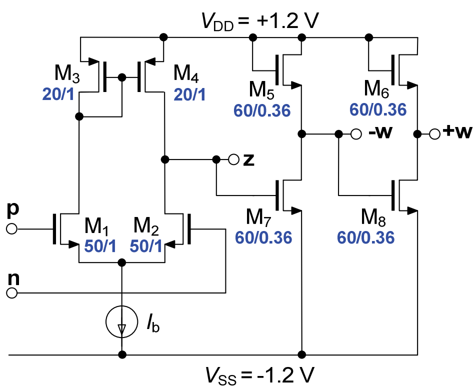
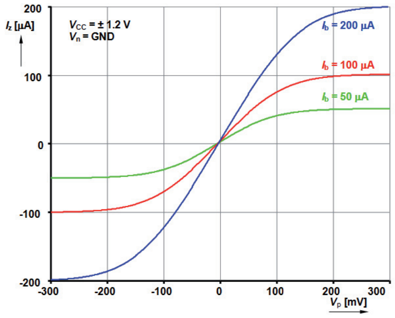
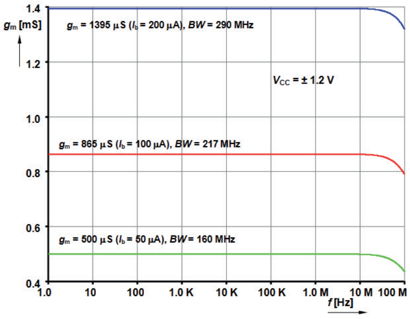


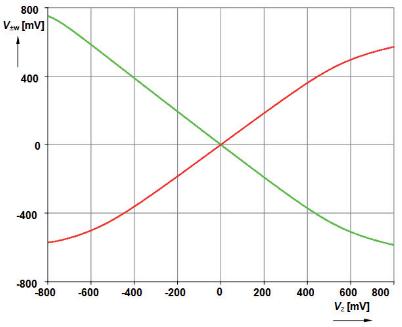
Fig. 10. DO-VDBA with one z terminal.



a)



b)



c)

Fig. 11. Selected features of the proposed DO-VDBA: a) DC transfer characteristic of OTA section, b) dependence of g_m on frequency, c) DC transfer characteristics of inverter/buffer.

Adjusting of I_b in the range of 5 - 200 µA causes changes of g_m from 60 µS to 1.4 mS. Some of the simulation results are documented graphically in Fig. 11.

The FB-VDBA has a different construction (Fig. 12). Function of the second type (FB-VDBA) is practically similar, only internal topology is slightly complicated due to necessity of z terminals of both polarities and k-times higher gain of transconductance section and mirrors of FB-VDBA. Many parameters of the second type of DO-VDBA are practically identical to the previously discussed type (parameters of voltage inverters). Simulations provided the following results (I_b = 50 µA, V_{CC} = ±1.2V): g_m = 1030 µS, $R_{\pm w} \approx 130$ kΩ, $C_{\pm z} \approx 0.38$ pF. The differential input resistance of OTA section has higher values than 0.64 MΩ at 1 MHz (parasitic capacitance ≈ 0.25 pF). Adjusting of g_m from 132 µS to 2.76 mS was verified (I_b between 5 and 200 µA). The highest tested I_b = 200 µA decreases $R_{\pm z}$ value to 53 kΩ approximately. Selected results are documented in Fig. 13.

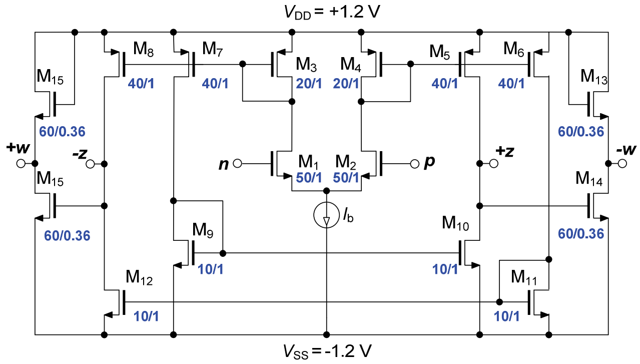
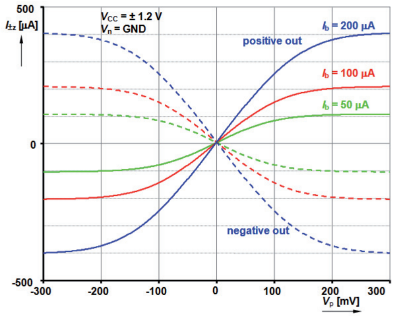
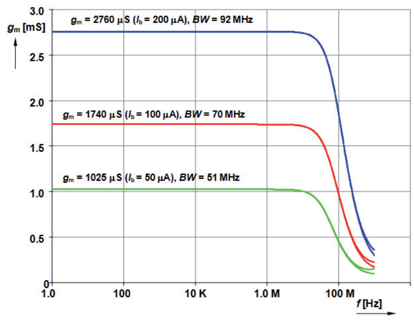


Fig. 12. FB-VDBA with z terminals of both polarities.



a)



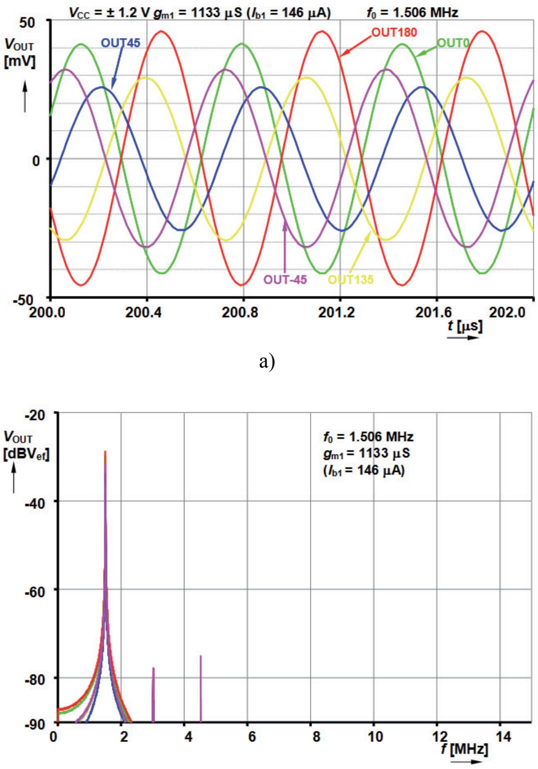
b)

Fig. 13. Selected features of FB-VDBA: a) DC transfer characteristic of OTA section, b) dependence of g_m on frequency.

4.2 Detailed PSpice Analysis of Some Proposed Applications

Multiphase oscillator - type without necessity of floating negative resistance

Oscillator from Fig. 6 we designed for operation at f_0 = 1.539 MHz with parameters: $R_1 = R_2 = R = 2.2$ kΩ, $C_1 = C_2 = C = 47$ pF, R_3 = 1 kΩ, and g_{m2} = 1 mS. Obtained results of simulation are shown in Fig. 14. Voltage gain of subtracting point created by DO-VDBA$_2$ and R_3 is approximately equal to 1. Real value of g_{m1} necessary for start of oscillation was increased to 1.133 mS (I_{b1} = =146 µA). Oscillation frequency 1.506 MHz was obtained, which is very close to the ideal value (error about 2%). Achieved total harmonic distortion (THD) was 0.40%, 0.39%, 0.26%, 0.81% and 0.81% for all outputs namely: V_{OUT0}, V_{OUT180}, V_{OUT45}, V_{OUT135}, V_{OUT-45}.



a)

b)

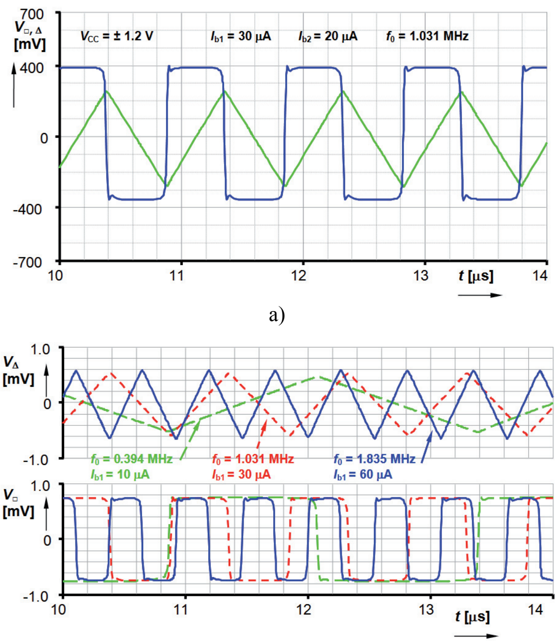
Fig. 14. Results of simulation of the oscillator from Fig. 6: a) transient responses, b) frequency spectrum.

Functional generator

We require operation in hundreds of kHz, therefore parameters of the design (considering solution shown in Fig. 9) are the following: $C = 22$ pF, $R = 10$ kΩ, $I_{b2} = 20$ µA, $k_2 = 2$ and suitable range of maximal output current (given by $k_1.I_{b1}$, where $k_1 = 1$) adjusting between 5 µA and 100 µA. Duty cycle was set to 50% ($n = 0$). Load resistances $R_L = 1$ kΩ were connected at the output ports (w) in simulations. We used FB-VDBA based comparator. This solution has one disadvantage, which is the necessity of the second resistor. However, such OTA section with mirrors in FB-VDBA structure has higher available voltage gain (bias mirroring with gain $k_2 = 2$, see Fig. 12) than simpler OTA section in the DO-VDBA. Of course, there is possibility to use simpler DO-VDBA, but validity of (36) is highly influenced by insufficient gain of the comparator in low-power solution. In fact, we are discussing a specific design for our requirements in the following text. Quality of the comparator directly influences oscillation frequency (threshold referencing voltage for changing of output polarity). The transconductance of OTA section of FB-VDBA has value 470 µS for $I_{b2} = 20$ µA. This value is quite low for sufficient voltage amplification in the comparator (it is better than for the same solution with DO-VDBA). We expected equality between input threshold voltage and output voltage of the comparator. Unfortunately, gain is not sufficient and therefore we achieved only $V_{\square} \approx 1.3V_{\Delta}$. In addition this gain is not valid in the whole DC transfer characteristic due to its nonlinearity in frame of the OTA section as part of the FB-VDBA based comparator. AC analysis allows finding small-signal g_m for ideal calculation (35) of the comparator behavior. Nevertheless, g_m has a lower value than in the origin of the characteristic (around 0) in case of comparator in overturn corners (reference voltages) of DC transfer characteristic. The new equation for repeating frequency considering the above discussed inequality of input (reference) and output voltage of the comparator and $D = 50\%$ has the following form:

$$f_{0\_\exp} = \frac{I_{b1}}{4C\dfrac{V_R}{\left(\dfrac{g_{m2}R}{g_{m2}R-1}\right)}} = \frac{I_{b1}}{\dfrac{4RCk_2I_{b2}}{\left(\dfrac{g_{m2}R}{g_{m2}R-1}\right)}} \cong \frac{I_{b1}}{\dfrac{8RCI_{b2}}{1.3}}. \quad (46)$$

Maximal output voltage at terminal $+z_2$ (V_R) of FB-VDBA has amplitude value $V_R = V_{\square} = I_{b2}.k_2.R \approx 400$ mV. This value is also expected for output amplitude of square wave signal (in case of single-ended solution). Considering the practical inequality of (36) given by (35) means that input reference voltage (causing overturn of the comparator) is 1.3 times lower (approximately 300 mV) than $V_{\square}$. This value is expected for amplitude of triangle (symmetrical ramp) wave signal. All these presumptions expect unity-gain followers/inverters and symmetrical referencing voltage (it is given by quality of the comparator), in real case it can be slightly different (non-unity gains of followers/inverters causes changes of expected amplitudes). Repeating frequency f_0 is expected for selected parameters and $I_{b1} = 30$ µA and with help of (46) as $f_0 = 1.108$ MHz. Simulated value of generator with models of VDBAs discussed above was obtained as 1.031 MHz. Results are in Fig. 15.



a)

b)

Fig. 15. Simulated transient responses of the proposed generator: a) single-ended mode, b) example of electronic control of f_0 (differential-balanced output mode).

Differential mode of operation has advantage of two-times higher produced amplitude. Electronic adjusting of f_0 was verified between 211 kHz and 2.83 MHz (I_{b1} changed from 5 µA to 100 µA), see Fig. 16 where comparison of calculated f_0 (46) and simulation results are depicted.

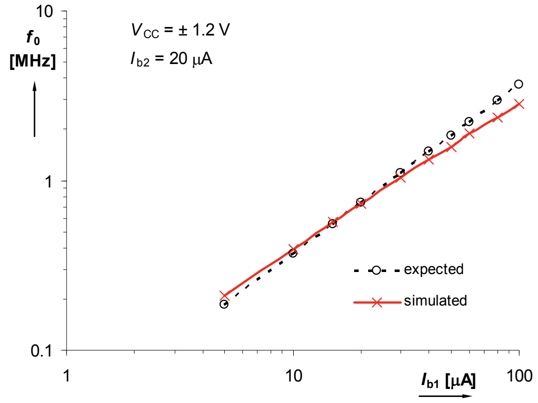


Fig. 16. Calculated and simulated dependence of f_0 on I_b.

An example of operation of the generator with $D = 24.5\%$ at $f_0 = 1$ MHz was provided. Expected f_0 can be expressed by the following equation considering a finite gain similarly as (46):

$$f_{0\_\exp} = \frac{I_{b1}(1-D)D}{2RCI_{b2}}\left(\frac{g_{m2}R-1}{g_{m2}R}\right). \qquad (47)$$

Analysis used the same values of the rest of parameters as in the previous case. The bias current $I_{b1} = 39$ µA was set in accordance to (47) for the above discussed assignment. Results in differential output mode are shown in Fig. 17 where possibility of adjusting of f_0 between two values is documented while D keeps unchangeable.

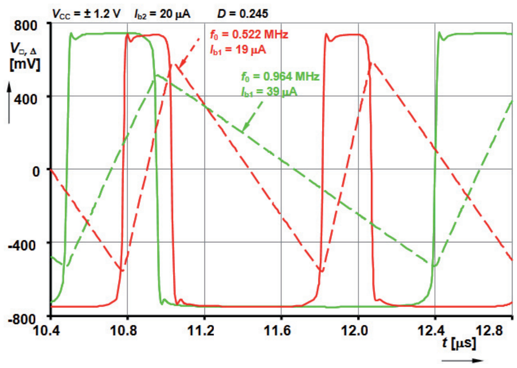


Fig. 17. Examples of variability f_0 in simulated transient responses with $D = 24.5$ %.

5. Experimental Results

The comparator in Fig. 8 was tested experimentally with behavioral model (Fig. 18) based on commercially available devices. All parameters and values are noted in Fig. 18. The circuit was tested for ramp-pulse excitation (±0.7 V, 1 kHz) and the results are in Fig. 19.

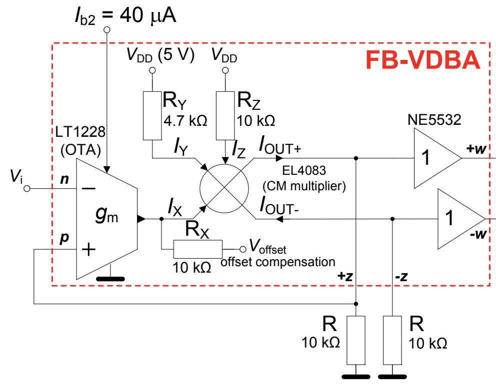


Fig. 18. Schmitt comparator employing behavioral model of FB-VDBA (based on commercially available devices).

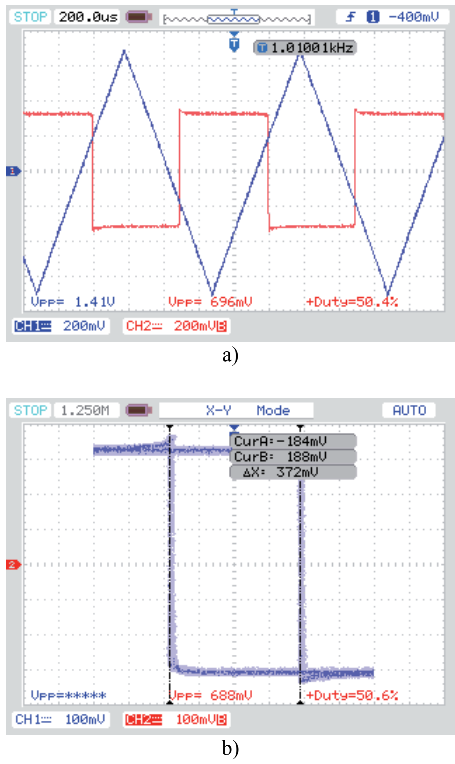


Fig. 19. Measurement results of the comparator: a) transient responses V_i (triangular), V_o, b) hysteresis characteristic of the comparator.

6. Conclusion

The designed active elements allow beneficial features in specific applications. They were intended for utilization in multiphase oscillators, quadrature oscillators (single-ended or differential mode) or triangle and square wave generators for example, where provided modifications of VDBA and VDIBA elements [1]-[5] with standard transconductance control or also additional electronic control of voltage gain in frame of active element were employed. Some of the presented circuits allow benefits of simplicity (multiphase oscillators for example) in comparison to classical design methods based on cascading lossy integrators or similar sections in phase shifted loop [6]-[8], where higher number of active and passive elements is necessary. Higher demands on precise accuracy of relations between generated outputs and necessity of several matching conditions is cost of these benefits. Nevertheless, matching conditions are also drawback of some more sophisticated phase-shifted solutions. Both introduced quadrature oscillators and triangle and square wave generator offer advantages of simplicity and differential output responses. The proposed circuits were verified by simulations at frequencies of several MHz and complemented by detailed discussions.

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