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High Pass Filter and Bandpass Filter Using Voltage Differencing Buffered Amplifier

Ridouane Hamdaouy[#1*], Boussetta Mostapha[#2], Khadija Slaoui[#3]
[#]University Sidi Mohamed Ben Abdellah, LESSI Laboratory, Department of Physics Faculty of Sciences, Dhar El Mehrez B.P. 1796, 30003 Fez-Atlas, Morocco
Email address: [1]ridouane.hamdaouy@usmba.ac.ma, [2]mostapha.boussetta@usmba.ac.ma, [3]slaoui.khadija@usmba.ac.ma

Abstract**—***A novel filter is proposed in this work. Voltage differencing buffered amplifier (VDBA) as an active component is used in the implementation of the proposed filter. Also, one capacitor and one resistor are employed in the proposed filter as passive circuit elements. On the other hand, voltage differencing buffered amplifier (VDBA) can be easily constructed by using a CMOS transistor. One of the main advantages of the proposed filter is the feature of low output impedance resulting in easy cascadability with other current Mode circuits. Moreover, the quality factor of the proposed filter can be easily adjusted by changing value of only one of the resistors without disturbing its angular resonance frequency. Nonetheless, it requires a single resistive matching condition for proper circuit operation and a unity gain non inverting amplifier for highpass/bandpass filter responses. A number of simulation results are achieved by using 0.13 µm TSMC13RF technology parameters with 3.3 V DC power supply voltages. Power consumption of the proposed filter is approximately found as 1.9432 mW through SPECTRE simulations. Furthermore, experimental test results are included to confirm the theory.*

Keywords**—** *Voltage Differencing Buffered Amplifier (VDBA), inductance simulator, lossy inductor, electronically tunable, highpass/bandpass filter responses.*

I. INTRODUCTION

Recently, actively simulated lossy inductor has become an important research issue, since it can be applied in various areas like active filter design, sinusoidal oscillator design, and parasitic element cancellations. In advanced integrated circuit technology, it encourages the design of synthetic inductance simulators, which can be employed to replace the bulky physical inductors in passive filters. In recent years, a number of topologies for realizing lossy inductance simulator based on a single active component have been developed [1]-[10]. Although the presented circuits of [1]-[10] use only one active component to realize grounded lossy inductors, but they still require three to four passive components, and also a floating capacitor for their realizations.

Nowadays, modern electronic active building blocks are gaining importance in analog signal processing applications and designs. In [11], modern day active components have been reviewed and discussed. One of them is the circuit principle called as VDBA (voltage differencing buffered amplifier). Its several applications, such as active filters, sinusoidal oscillators and immittance function simulators, were also introduced to demonstrate its usefulness and versatile [12]-[13].

The major purpose of this study is to present a grounded parallel inductance simulator employing single VDBA. The simulator requires a minimum number of active and passive components, i.e. one VDBA, one grounded capacitor, and one floating resistor, as well as no component matching conditions are necessary. The quality factor (Q) of the proposed filter circuit can be controlled orthogonally by changing value of only one resistor without disturbing its resonance frequency (fo). A number of simulation results are carried out by using 0.13 µm CMOS technology parameters with +3.3 V DC power supply voltages. Power dissipation of the proposed filter is nearly found as 1.9432 mW through SPECTRE simulations. Moreover, experimental test results are given to verify the theory.

The equivalent inductance value (Leq) of the realized simulator can be tuned electronically through the transconductance gain (gm) of the VDBA, without influencing the equivalent resistance value (Req). The performance of the proposed simulator is demonstrated on the current-mode highpass/bandpass filter. The simulation results based on TSMC13RF 0.13-µm CMOS technology demonstrate the feasibility of the designed circuit and its filter application.

This paper is organized as follows: After the first section, basic operation of voltage differencing buffered amplifier is introduced in section II. In Section III, proposed grounded lossy inductance simulator circuit and the parasitic impedance effects on the proposed filter are investigated is introduced in section IV. The simulation in Sections V and application to filter design is introduced in section VI. Section VII concludes the paper.

II. BASIC OPERATION OF VOLTAGE DIFFERENCING BUFFERED AMPLIFIER (VDBA)

In this study, schematic symbol of the VDBA is realized by using transconductance amplifier as an input stage, and the unity-gain voltage buffer as an output stage. Fig. 1 shows the VDBA circuit symbol, and it can be defined by the following matrix equation given in [11]-[13]:

$$\begin{bmatrix} i_p \\ i_n \\ i_z \\ v_w \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ gm & -gm & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} v_p \\ v_n \\ v_z \\ i_w \end{bmatrix} \quad (1)$$

High frequency active inductor it is known that the mobility of NMOS is around four to five times greater than PMOS in submicron CMOS technology. Although the P channel transistors help to reduce noise and nonlinearity in the circuit, the high frequency performance of active inductor is

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possible only by using NMOS transistors as shown in Fig. 2. That increases the operating frequency significantly. Less number of transistors proves the small chip area and low power consumption of active inductor circuit but the higher value of parasitic series resistance and lower value of parallel resistance value provide the lower value of quality factor.

In above expression, g_m is the small-signal transconductance gain of the VDBA. In general, the value of g_m is electronically controllable by a supplied bias current/voltage, which lends electronic controllability to design circuit parameters. From equation (1), the differential input voltage between the terminals p and n (v_p-v_n) is converted to a current at the z terminal (iz) by a g_m-parameter. The voltage across the z terminal (v_z) is then conveyed to the output voltage at the w terminal (v_w).

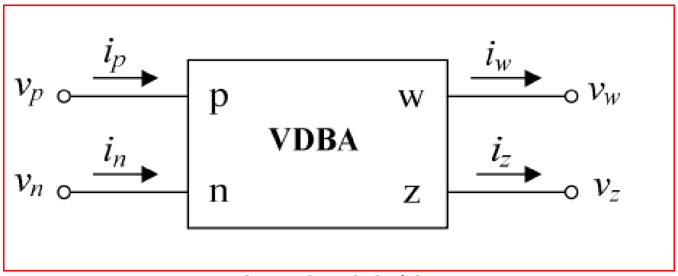

Fig. 1. Electrical symbol of the VDBA.

Fig. 2 shows the possible realization of the VDBA using CMOS technology. Groups of transistors M1-M2 and M5-M6 function as source-coupled pairs and current mirrors M3-M4 and M7-M8, which act as active loads. The source follower M9 forms a current follower in order to provide low-output impedance at the terminal w. Assume for the moment that M1-M2 as well as M5-M6 are well matched, the transconductance gain of this VDBA can be given by :

$$g_m = \sqrt{\mu C_{ox} \frac{W}{L} I_{B1}} \tag{2}$$

where μ is the mobility of the carriers, C_{ox} is the gate-oxide capacitance per unit area, W is the effective channel width, and L is the effective channel length.

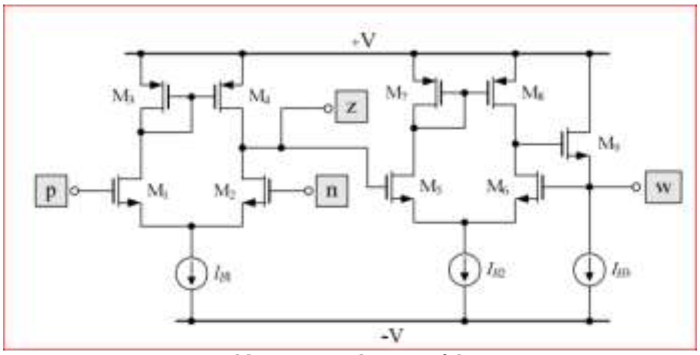

Fig. 2. Possible CMOS realization of the VDBA.

Fig. 3 shows the necessary connection among n, p, w and z terminals of the VDBA to form a Proposed grounded lossy parallel-type inductance simulator circuit.

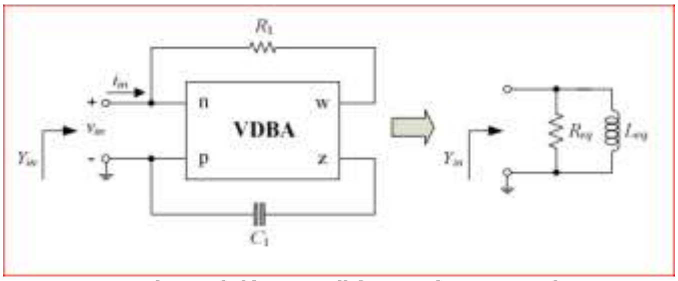

Fig. 3. Proposed grounded lossy parallel-type inductance simulator circuit, and its equivalent behavior.

III. PROPOSED GROUNDED LOSSY INDUCTANCE SIMULATOR CIRCUIT

Fig. 3 shows the proposed actively simulated R-L parallel impedance function. The simulator contains only one VDBA as an active element together with one grounded capacitor C_1 and one floating resistor R_1 as external passive components. Using equation (1) and deriving the configuration of Fig. 3, its admittance Y_{in} is realized of value:

$$Y_{in} = \frac{i_n}{v_n} = \frac{1}{R_{eq}} + \frac{1}{SL_{eq}} = \frac{1}{R_1} + \frac{g_m}{SR_1C_1} \tag{3}$$

which represents the parallel connection of equivalent resistance (Req) and equivalent inductance (Leq) as:

$$R_{eq} = R_1 \tag{4}$$

and

$$L_{eq} = \frac{R_1C_1}{g_m} \tag{5}$$

It should be noted from above expressions that the proposed circuit can simulate parallel R and L impedance. The realized values of the proposed simulator circuit do not require any element-matching condition. Since the gm-value of the VDBA directly depends on the external biasing current, the simulated Leq value is electronically tunable.

IV. NON-IDEAL ANALYSIS AND SENSITIVITY PERFORMANCE

For non-ideal case, the voltage-current relations of the VDBA can be rewritten as:

$$\begin{bmatrix} i_p \\ i_n \\ i_z \\ v_w \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ \alpha gm & -\alpha gm & 0 & 0 \\ 0 & 0 & \beta & 0 \end{bmatrix} \begin{bmatrix} v_p \\ v_n \\ v_z \\ i_w \end{bmatrix} \tag{6}$$

where $\alpha = (1 - \varepsilon_{gm})$ and $\beta = (1 - \varepsilon_v)$. Also, $|\varepsilon_{gm}| << 1$ denotes the transconductance inaccuracy, and $|\varepsilon v| << 1$ denote the voltage tracking error from terminal z to terminal w, respectively. Taking into consider the non-ideal properties of the VDBA on the performance of the realized simulator in Fig. 3, the equivalent non-ideal equivalent resistance and inductance are found as, respectively:

$$R_{eq} = R_1 \tag{7}$$

$$L_{eq} = \frac{R_1C_1}{\alpha\beta g_m} \tag{8}$$

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It may be pointed out that the non-ideal L_{eq}-value slightly deviates from its ideal value. It may be pointed out that the non-ideal L_{eq}-value slightly deviates from its ideal value. However, this small deviation can be compensated by properly tuning the value of g_m of the VDBA. The effect of the deviations in active and passive component values is determined by evaluating sensitivity coefficients, which are found to be:

$$S_{\alpha}^{R_{eq}} = S_{\beta}^{R_{eq}} = S_{g_m}^{R_{eq}} = 0 \tag{9}$$

$$S_{R_1}^{R_{eq}} = 1 \quad , \quad S_{C_1}^{R_{eq}} = 0 \tag{10}$$

$$S_{\alpha}^{L_{eq}} = S_{\beta}^{L_{eq}} = S_{g_m}^{L_{eq}} = -1 \tag{11}$$

$$S_{R_1}^{L_{eq}} = S_{C_1}^{L_{eq}} = 1 \tag{12}$$

From equations (9)-(12), it can be clearly seen that all the sensitivities of the various parameters of the proposed inductance simulator are no more than unity. It can be seen that inductance has no passive component matching constraints and low sensitivity. And the inductance is free from passive component matching requirements. It is evident that the value of the inductance can be controlled electronically by adjusting the bias current of VDBA .It is also prove that the sensitivities do not depend upon the active and passive component values.

V. SIMULATION RESULTS AND DISCUSSIONS

The performance of the proposed simulator circuit in Fig. 3 has been evaluated by SPECTRE simulation. The CMOS VDBA shown in Fig. 2 with +V = 3.3V, -V=0, IB2 = 63μA and IB3 = 7μA is used in all the simulations. Also, the SPETRE simulations are performed with TSMC13RF 0.13-μm CMOS process parameters. The aspect ratios of the MOS transistors are given in Table I using 0.13 um MOSFET.

TABLE I. The aspect ratio of the mos transistors in fig. 2.

Transistor	W/L (μm/μm)
M1-M2 , M5-M6	0.55/0.35
M3 , M4	0.15/20
M7 , M8	100/0.3
M9	100/0.35

As an example, the proposed inductance simulator of Fig. 3 is realized with IB1 = 518.859 μA, C_1 = 0.1nF and R_1 = 1 kΩ, which results in total power consumption of 1.9432 mW. Fig. 4 shows simulated results of the input voltage (v_{in}) and input current (i_{in}) waveforms with 100 mV (peak) input voltage at f = 100 kHz. As can be measured from the result, there is a 6.156° phase difference between v_{in} and i_{in}, which demonstrates that the circuit works as a lossy inductance simulator. With the same component setting, the simulated frequency responses for the input impedance Z_{in} of the proposed circuit in Fig. 3 comparing with the ideal responses are also plotted in Fig. 5.

To show the adjustability of Req, the external resistor R1 has been changed to the values of 0.9 kΩ, 2 kΩ, and 4 kΩ, resulting in Req = 0.9 kΩ, 2 kΩ, and 4 kΩ, respectively. The impedancefrequency characteristics of the proposed inductance simulator for various R_1 values are shown in Fig. 6. It can be observed that the Req values can be adjusted precisely by changing R_1. The electronic variation of the equivalent inductance value Leq with the VDBA biasing current IB1, obtained by simulation, is shown in Fig. 7. When the biasing current IB1 was changed through 418 μA, 500 μA, and 502 μA, the value of Leq also changed through roughly 9.398 nH , 3.343uH and 3.046 uH, respectively.

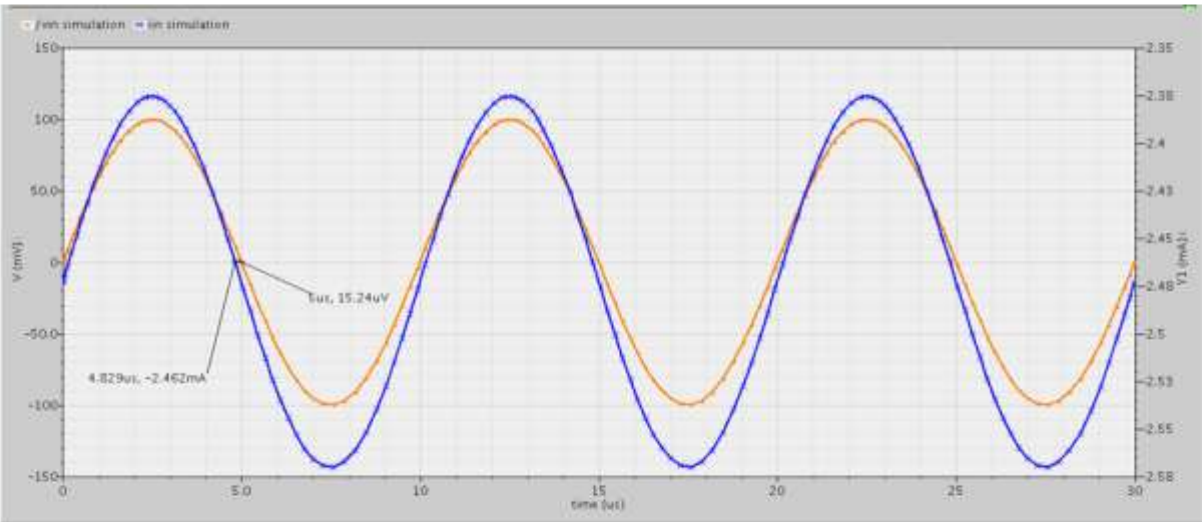

Fig. 4. Simulated time-domain voltage and current responses of fig. 3.

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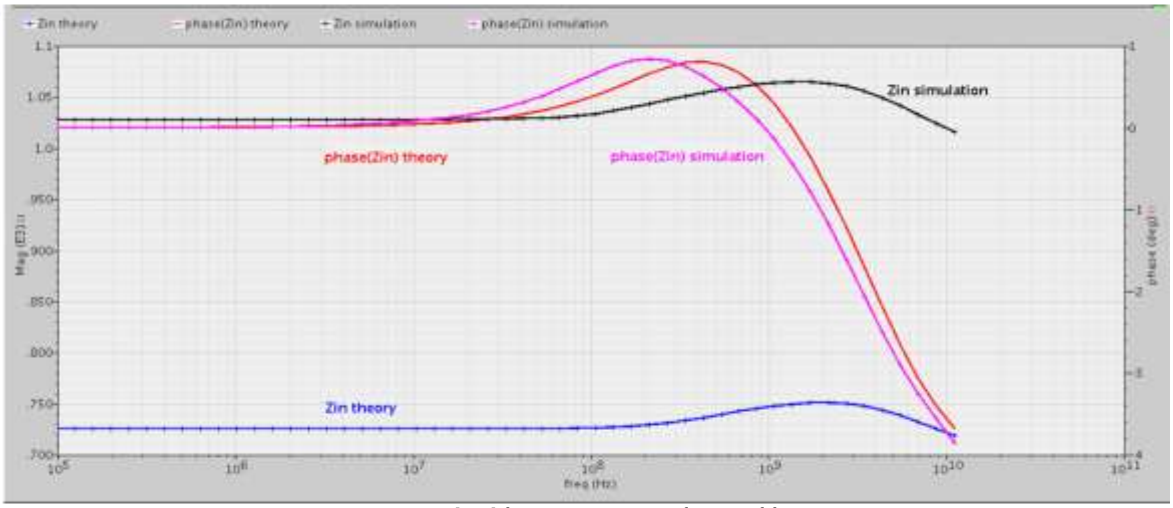

Fig. 5. Simulated frequency responses for Zin of fig. 3.

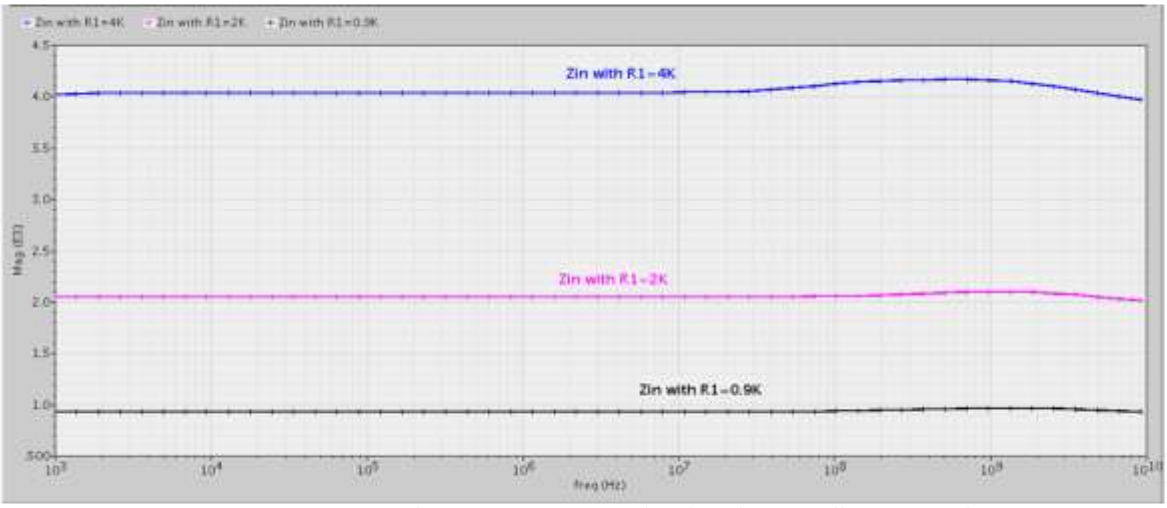

Fig. 6. Simulated impedance-frequency characteristics of the fig. 3 for three different values of R1.

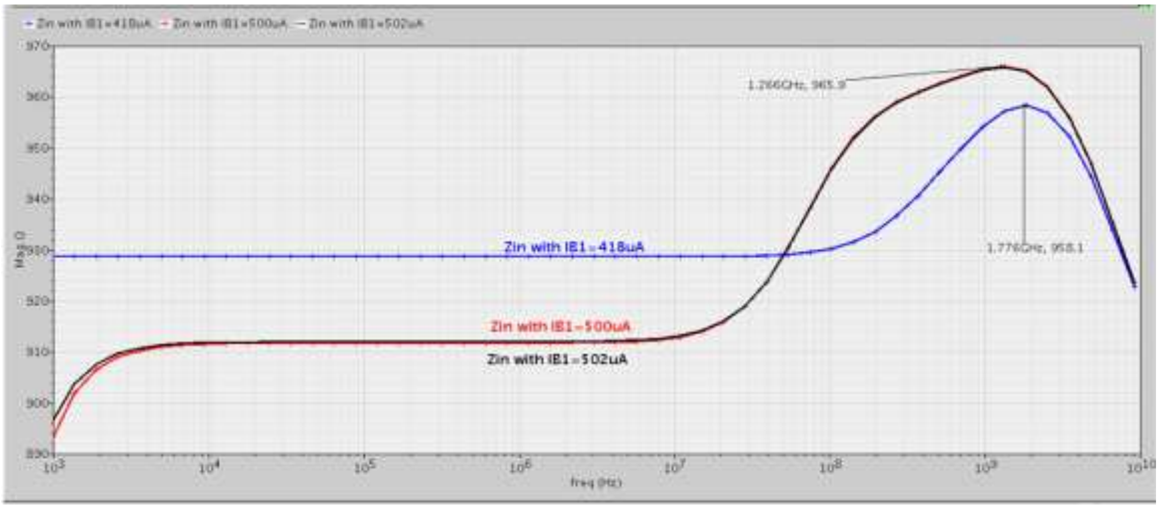

Fig. 7. Simulated impedance-frequency characteristics of the fig. 3 for three different values of IB1.

VI. Application to Filter Design

To demonstrate the performances of the proposed circuits, the loss inductance simulator in Fig. 3 is used to construct a second-order filter as shown in Fig. 8. It can be possible to make further analysis on the current transfer functions of the proposed filter when the passive component matching condition is taken into account. The current transfer functions

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of the proposed filter can be expressed with respect to the input voltage selections. An application example is the realization of a current-mode highpass/bandpass filter shown in Fig. 8 [6], obtained by using the proposed simulator of Fig. 3 in place of the passive parallel R-L branch. The current transfer functions of the proposed filter with and without the considering any matching conditions can be expressed as:

$$\frac{I_{HP}(s)}{I_{in}(s)} = \frac{S^2}{S^2 + s\left(\frac{1}{R_{eq}C_2} + \frac{1}{R_2 C_2}\right) + \frac{1}{L_{eq}C_2}} \quad (13)$$

$$\frac{I_{BP}(s)}{I_{in}(s)} = \frac{\left(\frac{1}{R_{eq}C_2} + \frac{1}{R_2 C_2}\right)\left(\frac{R_{eq}}{R_{eq} + C_2}\right)S}{S^2 + s\left(\frac{1}{R_{eq}C_2} + \frac{1}{R_2 C_2}\right) + \frac{1}{L_{eq}C_2}} \quad (14)$$

The equations in (13) and (14) can also be used to determine f_0 and Q of the proposed filter.

Thus, f_0 and Q are respectively given in the following equations:

$$f_0 = \frac{1}{2\pi}\frac{1}{\sqrt{L_{eq}C_2}} \quad (15)$$

$$Q = \left(\frac{1}{R_{eq}C_2} + \frac{1}{R_2 C_2}\right)^{-1}\frac{1}{\sqrt{L_{eq}C_2}} \quad (16)$$

It is seen from the equations in (13) and (14) that f_0 and Q can be controlled orthogonally by the value of the resistor R2.

The passive component sensitivities with respect to f_0 and Q are evaluated as follows:

$$S_{R_2}^{Q} = -1 \ , \ S_{C_2}^{Q} = 1/2 \quad (17a)$$

$$S_{R_2}^{f_0} = 0 \ , \ S_{C_2}^{f_0} = -1/2 \quad (17b)$$

It is observed from equations (17) that the values of passive component sensitivities with respect to f_0 and Q are no more than unity in magnitude.

The designed filter of Fig. 8 is simulated with R2 = 11kΩ, C1 = C2 = 0.1 nF, R1 = 1 kΩ, and IB1 = 502 μA. The above designed element values lead to obtain R_{eq} = 1 KΩ in parallel with L_{eq} = 3.046 uH, which results in a natural angular frequency f_0 ≅ 263.6 MHz, and a quality factor Q = 0.25. The simulated frequency responses of the filter are shown in Fig. 9.

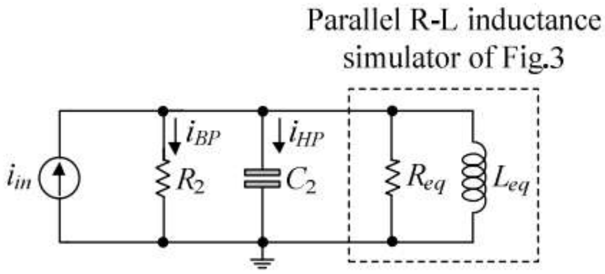
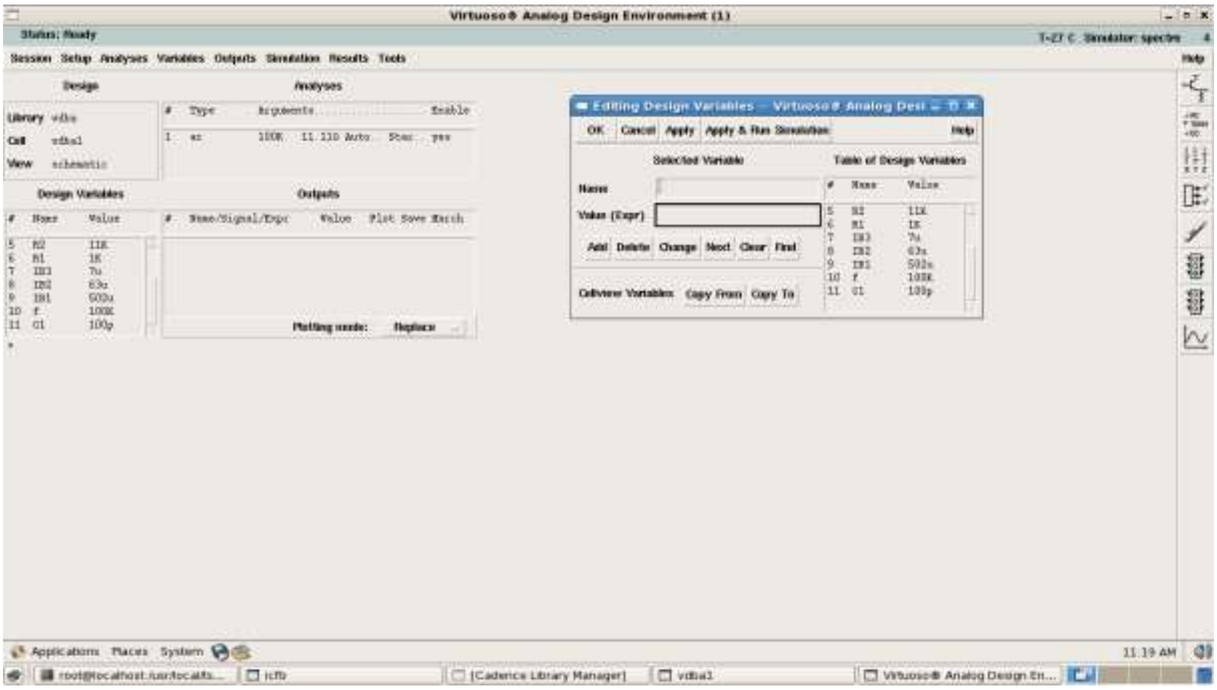


Fig. 8. Current-mode highpass/bandpass filter using the proposed inductance simulator obtained from fig. 3.



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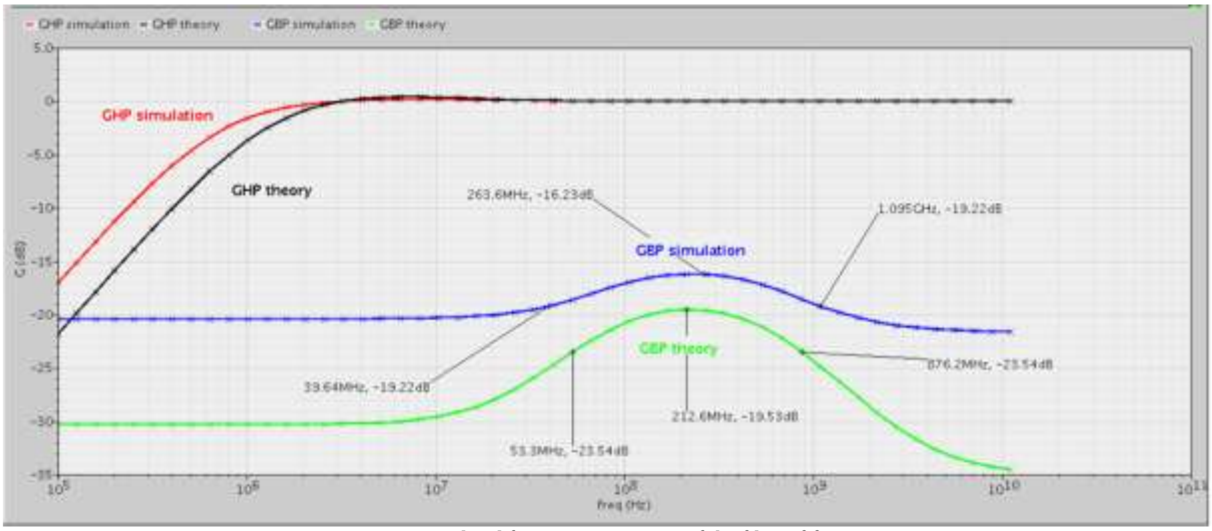

Fig. 9. Simulated frequency responses of the filter of fig. 8.

VII. CONCLUSION

A novel second-order filter composed of a VDBA as active element is proposed in this study. Due to the nature of the proposed filter, the active devices have of the terminals and can be easily constructed by employing VDBA. Moreover, the proposed filter employs one capacitor and one resistors as the passive elements. One of the main advantages of the proposed filter is to include a reduced number of active and passive circuit elements. The other important properties of the proposed filter are both to have low output impedance resulting in easy cascadability with other topologies and to provide orthogonal control of the angular resonance frequency and the quality factor.

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