Tunable Active Grounded Lossless and Lossy Inductance Simulators with Single Grounded Capacitor Using VDBAs

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Abstract

In this paper, three active-C synthetic grounded inductance simulator circuits are presented, which realize tunable lossless and lossy series and parallel R-L type inductances. Each of which employs two voltage differencing buffered amplifiers (VDBAs) as active components, and a single grounded capacitor as a passive component. In all the proposed circuits, the simulated equivalent resistance and inductance values can be adjusted electronically through the transconductance gains of the VDBAs. They also do not require any critical component matching conditions and cancellation constraints. Detail non-ideal analysis including transfer errors of the VDBA has been analyzed. For circuit performance
verification and comparison, some application examples are given together with computer simulation results by PSPICE program.

**Keywords**: Voltage Differencing Buffered Amplifier (VDBA), lossless inductor, lossy inductor, grounded inductance simulator, electronically tunable circuit.

**1. Introduction**

Inductors are one of the most essential circuit components of analog and mixed-signal integrated circuits and systems. Particularly, they are used to realize a variety of active networks, such as active ladder filters, impedance matching circuitry and chaotic oscillators [1]. However, in modern integrated circuit design, the integration of high-valued inductors is a fundamental problem, because of the occupied silicon area. In addition, their values cannot adjust easily after fabrication. Owing to these restrictions, numerous synthetic simulated inductors using various kinds of active elements have been created instead of large-valued physical inductors [2-16]. These simulators can be classified based on whether they realize a grounded or floating inductor configuration or whether they simulate a lossless or lossy type of inductor. In the class of grounded inductance simulators, they have simpler structure than floating ones. Further, from these simulators, some of them are related with simulation of grounded lossless inductor [2-8]. Moreover, several specific configurations for simulating series and parallel R-L type lossy inductor are also available in [9-16]. Among these simulators, the circuits in [4, 6, 10] require two active devices and three or more passive elements for simulating grounded lossless and parallel R-L type lossy inductors. In the realizations at [2-3, 5, 7, 9, 11-15], only one active device and at least three passive elements are employed to simulate grounded inductances. In addition to the circuits of [2-4, 6-16], they are realized with ungrounded resistors, which are impediments in integrated circuit (IC)
fabrication. Most of them also include a floating capacitor [2-3, 6-8, 10-14]. It is well-known that the use of only grounded capacitors particularly makes the circuits suitable for fully IC design [17]. Although the lossless grounded inductor realizations discussed in [8, 16] employ only one active component, they still contain two floating passive components for their realizations. Moreover, the synthetic inductors developed in [2-7, 9-14] do not possess the property of electronic tunable performance. Some of above mentioned simulators also suffer from the need of component-matching constraints [2-3, 5-7, 12], where some capacitors or resistors are needed either to satisfy certain conditions and/or to have identical values. In literature, the external adjustability of circuits without requirement any critical component-matching choices has considerable advantages for IC design [18-19].

Recently, the voltage differencing buffered amplifier (VDBA), which is a novel introduced versatile active building block, has been introduced and also applied in design and synthesis of diverse analog circuit functions and applications [20-21]. The VDBA essentially combines the electronic tuning feature of the classical transconductance amplifier together with the voltage following property of the output voltage buffer. In comparison to the existing active elements, the VDBA-based circuit does not need any additional resistors for voltage-to-current conversion in some voltage-mode application circuits as that use in current differencing buffered amplifier-based circuit [22]. With respect to existing transconductance-based active devices such as, current differencing transconductance amplifier (CDTA), current conveyor transconductance amplifier (CCTA), differential voltage CCTA (DV-CCTA), the VDBA device permits low-impedance voltage output, which is directly cascaddable in voltage-mode operation owing to the effect of load capacitance or resistance changing can be avoided [20].

To demonstrate the utility as well as the workability of the VDBA device, three canonical grounded inductance simulator circuits with a single grounded capacitor are
proposed. Out of the three proposed circuits, the first circuit realizes a lossless or pure inductor, while two circuits realize series and parallel lossy inductors. In all the proposed inductor realizations, only two VDBAs and one grounded capacitor are employed. The simulated equivalent resistance and inductance values can be tuned electronically through adjusting the bias currents of the VDBAs. Computer simulations based on TSMC 0.25-μm CMOS real process parameters are performed to validate the circuit operation. In addition to demonstrate the utility of the proposed simulator circuits, some illustrative applications are also given.

2. Voltage Differencing Buffered Amplifier (VDBA)

As a symbolic notation represented in Fig.1, the VDBA is an active building block with four terminals, namely p, n, z and w [20-21]. The defining operation of the VDBA can be expressed by the following equation:

\[
\begin{bmatrix}
i_p \\
i_n \\
i_z \\
v_w
\end{bmatrix} = \begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
\alpha g_m & -\alpha g_m & 0 & 0 \\
0 & 0 & \beta & 0
\end{bmatrix}
\begin{bmatrix}
v_p \\
v_n \\
v_z \\
i_w
\end{bmatrix},
\]

where \( g_m \) is the transconductance value, and \( \alpha \) and \( \beta \) are the non-ideal transfer gains of the VDBA, respectively. Note that, in ideal case, the values of \( \alpha \) and \( \beta \) are equal to unity.

3. Proposed Inductance Simulators

The proposed circuits for simulating the grounded lossless and lossy inductors are shown in Fig.2-4. It is readily seen that each of the proposed circuits consists of solely two VDBAs and a single grounded capacitor without the requirement of any external resistors. Under ideal conditions (\( \alpha = \beta = 1 \)), straightforward analysis of the circuits in Fig.2-4 yields their input impedances to be expressed as:
\[ Z_{in1} = \frac{sC_L}{g_{m1}g_{m2}} = sL_{eq1} \]

for the circuit of Fig.2, \((2)\)

\[ Z_{in2} = \frac{1}{g_{m1}} + \frac{sC_g}{g_{m1}g_{m2}} = R_{eq2} + sL_{eq2} \]

for the circuit of Fig.3, \((3)\)

and

\[ Z_{in3} = \frac{sC_p}{g_{m1}g_{m2} + sC_p g_{m1}} = R_{eq3} // sL_{eq3} \]

for the circuit of Fig.4. \((4)\)

where \(g_{mi} (i = 1, 2)\) stands for the parameter \(g_m\) associated with the \(i^{th}\) VDBA. Therefore, the proposed circuit of Fig.2 simulates the lossless inductor, while the circuits of Fig.3 and 4 realize the series and parallel R-L type lossy inductors, respectively. In addition, all the proposed circuits do not require any critical element-matching constraints and cancellation choices for the desired realizations. Performance comparison of the previously similar grounded inductance simulators [2-16] and the proposed VDBA-based simulators is summarized in Table 1.

### 3.1. Simulated Lossless Inductor

From Eq.(2), the simulated lossless equivalent inductance of Fig.2 is realized as: \(L_{eq1} = C_L/g_{m1}g_{m2}\). It can be deduced that the realized equivalent inductance value can be tuned electronically through \(g_{m1}\) and/or \(g_{m2}\) of the corresponding VDBAs. Sensitivity analysis shows that all various sensitivity coefficients of the simulated lossless inductor with respect to active and passive elements are found as:

\[ S_{L_{eq1}}^{C_L} = 1 \quad , \quad S_{g_{m1}}^{L_{eq1}} = S_{g_{m2}}^{L_{eq1}} = -1 \]

\((5)\)

Recall that sensitivity is the percentage that the dependent variable, in this case the \(L_{eq1}\) value, changes relative to the independent variables (\(g_{m1}, g_{m2}\) and \(C_L\)). From Eq.(5), the sensitivity relations are identical except for the sign. It is noted that the \(L_{eq1}\) sensitivity with respect to \(C_L\) is positive. As the positive sign implies, when \(C_L\) increases, the \(L_{eq1}\) also increases. On the other hand, when \(g_{m1}\) increases, the \(L_{eq1}\) decreases, which is expected since the sensitivity
to $g_{m1}$ is negative. This means that the value of $L_{eq1}$ should change by very nearly 1% for
every 1% deviation in either $C_L$ or $g_{mi}$ under these conditions. For example, if $g_{m1}$ or $g_{m2}$
increases by 1%, then $L_{eq1}$ becomes a 1% reduction. Similarly, increase $C_L$ by 1% results in
$L_{eq1}$ being an increase of 1%.

It is useful to consider the effect of the VDBA non-idealities on the realized inductor.
Taking into account the non-ideal gains $\alpha$ and $\beta$, the equivalent inductance for the lossless
simulator circuit shown in Fig.2 is modified as:

$$L_{eq1} = \frac{C_L}{\alpha_1 \alpha_2 \beta_1 \beta_2 g_{m1} g_{m2}}, \quad (6)$$

where $\alpha_i$ and $\beta_i$ represent the non-ideal gains $\alpha$ and $\beta$ of the related VDBA. This implies that
the VDBA non-ideal gains have directly affected on the value of $L_{eq1}$. However, the
transconductances $g_{mi}$ of the devices can be fine-tuned to compensate the deviation in the $L_{eq1}$
value of the proposed simulator circuit.

3.2. Simulated Series R-L Type Lossy Inductor

From Eq.(3), it can be observed that a grounded series R-L impedance with equivalent
series resistance $R_{eq2} = 1/g_{m1}$ and equivalent series inductance $L_{eq2} = C_S/g_{m1}g_{m2}$ is realized
from the simulator of Fig.3. It is worth mentioning that the value of $R_{eq2}$ value can be
adjusted electronically by $g_{m1}$, while the value of $L_{eq2}$ can be changed without affecting $R_{eq2}$
by $g_{m2}$. Moreover, the simulator of Fig.3 can be used as a lossless inductor in the operating
frequency region of $\omega >> R_{eq2}/L_{eq2} = g_{m2}/C_S$. Also from Eq.(3), normalized active and passive
sensitivities of $R_{eq2}$ and $L_{eq2}$ are derived and obtained as unity in magnitude. The quality
factor ($Q_S$) of this lossy inductor can be found as:

$$Q_S = \frac{\omega L_{eq2}}{R_{eq2}} = \frac{\omega C_S}{g_{m2}}, \quad (7)$$
If the aforementioned non-ideal gains are taken into consider, the non-ideal equivalent resistance and inductance of the series lossy inductor in Fig.3 are evaluated as follows:

\[ R_{eq2} = \frac{1}{\alpha_2 \beta_2 g_{m1}}, \quad (8) \]

and

\[ L_{eq2} = \frac{C_s}{\alpha_2 \beta_1 \beta_2 g_{m1} g_{m2}} \cdot (9) \]

Note that the small deviation on \( R_{eq2} \) value can sufficiently be minimized by changing \( g_{m1} \), while the error in \( L_{eq2} \) value can be compensated by tuning \( g_{m2} \).

**3.3. Simulated Parallel R-L Type Lossy Inductor**

From Eq.(4), the simulated equivalent input admittance for the circuit of Fig.4 is obtained as:

\[ Y_{in3} = \frac{1}{Z_{in3}} = \frac{1}{R_{eq3}} + \frac{1}{sL_{eq3}} = g_{m1} + \frac{g_{m1} g_{m2}}{s C_p} \cdot (10) \]

It should be recalled that the circuit of Fig.4 simulates a parallel R-L impedance with the equivalent parallel resistance and inductance values of \( R_{eq3} = 1/g_{m1} \) and \( L_{eq3} = C_p/g_{m1} g_{m2} \). One can find that \( R_{eq3} \) is tunable electronically by means of \( g_{m1} \), while \( L_{eq3} \) is adjustable independently through a single transconductance \( g_{m2} \). Also from Eq.(10), the proposed circuit of Fig.4 can work as a grounded lossless inductor for \( \omega << R_{eq3}/L_{eq3} = g_{m2}/C_p \). The various sensitivities of the simulated \( R_{eq3} \) and \( L_{eq3} \) with respect to active and passive components are also analyzed, and obtained that the resultant sensitivity values are all no more than unity in magnitude. For this parallel inductor, the quality factor \((Q_p)\) can be given by:

\[ Q_p = \frac{R_{eq3}}{\omega L_{eq3}} = \frac{g_{m2}}{\omega C_p} \cdot (11) \]
Taking into account the non-ideal gains $\alpha$ and $\beta$ for the circuit in Fig.4, one can obtain the modified $R_{eq3}$ and $L_{eq3}$ as, respectively,

$$R_{eq3} = \frac{1}{\alpha_1\beta_1g_{m1}}, \quad (12)$$

and

$$L_{eq3} = \frac{C_p}{\alpha_1\alpha_2\beta_1\beta_2g_{m1}g_{m2}}. \quad (13)$$

### 4. Simulations, Discussions and Applications

The validity of all the proposed grounded inductance simulators in Fig.2-4 has been evaluated using PSPICE simulation program. To achieve simulations, the VDBA is derived from the schematic CMOS implementation illustrated in Fig.5 with symmetrical supply voltages of $\pm0.75V$ and biasing currents $I_A$ of 50 $\mu$A [23]. CMOS transistors in Fig.5 are simulated with the real process parameters of 0.25-$\mu$m CMOS technology from TSMC. Transistor sizes ($W/L$) are given in Table 2. From internal structure of Fig.5, let us assume that M1-M2 of the source couple are well matched and the current mirror M3-M4 has unity current transfer gain. Therefore, the value of transconductance $g_m$ of the VDBA can be determined by: [23]

$$g_m = \sqrt{KI_B} \quad (14)$$

where $I_B$ is the external bias current of the VDBA, $K = C_{ox} W/L$ is the transconductance coefficient, $\mu$ is carrier mobility, $C_{ox}$ is the gate oxide capacitance per unit area, and $W$ and $L$ are the channel width and length of M1 and M2 devices, respectively. Eq.(14) clearly indicates that the $g_m$ value can be adjusted electronically with adjustment of $I_B$.

#### 4.1 Proposed Lossless Inductor in Fig.2

The proposed grounded lossless inductance simulator circuit shown in Fig.2 is simulated with the following component values: $g_m = g_{m1} = g_{m2} \cong 1.50$ mA/V and $C_L = 50$ pF.
For a given transconductance value, the corresponding bias currents calculated from Eq.(14) are equal to $I_B = I_{B1} = I_{B2} = 90 \, \mu A$. Also from Eq.(2), the simulated equivalent inductance will be obtained as: $L_{eq1} = 22.22 \, \mu H$. Fig.6 shows the simulated transient waveforms of the voltage $v_{in1}$ and the current $i_{in1}$ through the proposed lossless inductor of Fig.2, for a sinusoidal input signal with 50 mV peak value at $f = 2 \, MHz$. From the results of Fig.6, the phase difference between $v_{in1}$ and $i_{in1}$ is equal to $83.5^\circ$ lagging, where its ideal value is $90^\circ$ lagging. Total power consumption of the circuit is measured as: $0.653 \, mW$. On the other hand, the ideal and simulated frequency characteristics of the input impedance $Z_{in1}$ of the proposed lossless inductor with the same component setting are shown in Fig.7. It is shown that, for the acceptable phase error less than 8%, the operating frequency range of the simulator is approximately $100 \, kHz$ up to $70 \, MHz$. Furthermore, to illustrate the electronic tuning property of the synthetic inductor, the impedance magnitude frequency characteristics with variation of the equivalent inductance value by tuning $I_B$ as: $50 \, \mu A$ (for $L_{eq1} = 40 \, \mu H$), $150 \, \mu A$ (for $L_{eq1} = 13.3 \, \mu H$) and $300 \, \mu A$ (for $L_{eq1} = 6.7 \, \mu H$) are given in Fig.8. The resulting characteristics prove that the variation of $L_{eq1}$ can be performed electronically by changing the control current $I_B$.

The functionality of the simulated lossless inductor of Fig.2 has been verified by using it in the realization of a tunable voltage-mode bandpass (BP) and highpass (HP) filter shown in Fig.9 [4]. The voltage transfer functions are given by:

$$BP = \frac{V_{out1}(s)}{V_{in1}(s)} = \frac{\left(\frac{\omega_o}{Q}\right)s}{s^2 + \left(\frac{\omega_o}{Q}\right)s + \omega_o^2},$$

(15)

and

$$HP = \frac{V_{out2}(s)}{V_{in2}(s)} = \frac{s^2}{s^2 + \left(\frac{\omega_o}{Q}\right)s + \omega_o^2},$$

(16)
where \( \omega_0 \) and \( Q \) are the natural angular frequency and the quality factor of the filter, respectively. From Eq.(15) and (16), the characteristics \( \omega_0 \) and \( Q \) of the BP and HP voltage responses are equal to:

\[
\omega_0 = 2\pi f_o = \frac{1}{\sqrt{L_{eq1}C_1}},
\]

and

\[
Q = R_1 \sqrt{\frac{C_1}{L_{eq1}}}. \tag{18}
\]

As an example, the values of the passive elements used in the realized filter are taken as follows: \( R_1 = 200 \, \Omega \) and \( C_1 = 1 \, \text{nF} \). The gain-frequency responses of the BP and HP filters obtained from theoretical and simulation with variation in \( I_B \) are depicted in Fig.10. For \( I_B = 50 \, \mu\text{A}, 150 \, \mu\text{A} \) and \( 300 \, \mu\text{A} \), the filter parameters are calculated as: \( f_o = 0.8 \, \text{MHz}, 1.4 \, \text{MHz} \) and \( 2 \, \text{MHz} \), and \( Q = 1, 1.73 \) and \( 2.45 \), respectively.

### 4.2 Proposed Series Lossy Inductor in Fig.3

For the simulation of \( R_{eq2} = 666 \, \Omega \) and \( L_{eq2} = 22.21 \, \mu\text{H} \) by using the grounded series lossy inductor in Fig.3, the designed active and passive components taken from Eq.(14) and (3) are: \( I_B = I_{B1} = I_{B2} = 90 \, \mu\text{A} \) and \( C_S = 50 \, \text{pF} \). Fig.11 shows the simulated input waveforms through the simulator when applying a 5-MHz sinusoidal signal with 50 mV peak. It appears that the phase angle of the current \( i_{in2} \) lags that of the voltage \( v_{in2} \) by 48°, where the theoretically predicted value is equal to 46°. Fig.12(a) illustrates the magnitude-frequency characteristics of \( Z_{in2} \) for \( I_{B2} = 90 \, \mu\text{A} \) and various values of \( I_{B1} \) (i.e. 50 \( \mu\text{A}, 150 \, \mu\text{A} \) and 400 \( \mu\text{A} \)). According to Eq.(3), the corresponding \( R_{eq2} \) and \( L_{eq2} \) are obtained as: \( (R_{eq2} = 894 \, \Omega \) and \( L_{eq2} = 30 \, \mu\text{H}) \), \( (R_{eq2} = 516 \, \Omega \) and \( L_{eq2} = 17 \, \mu\text{H}) \), and \( (R_{eq2} = 316 \, \Omega \) and \( L_{eq2} = 10.5 \, \mu\text{H}) \), respectively. On the other hand, to show the independent electronic tuning of the \( L_{eq2} \) value, the circuit component values are chosen as: \( C_S = 50 \, \text{pF}, I_{B1} = 90 \, \mu\text{A} \) and \( I_{B2} \) changing from 50
μA, 150 μA to 400 μA, which lead to $R_{eq2} = 666 \, \Omega$ and $L_{eq2}$ altering from 30 μH, 17 μH to 10.5 μH. Both of the theoretical and simulated results are given in Fig.12(b). As it is readily seen from these characteristics, the $R_{eq2}$ value is electronically tunable through $I_{B1}$ ($g_m1$), while the $L_{eq2}$ value is orthogonally adjustable through $I_{B2}$ ($g_m2$).

To further evaluate the behavior of the proposed series lossy inductor in Fig.3, it is tested on an active RLC current-mode lowpass (LP) filter given in Fig.13. The active and passive component values used for Fig.13 are as follows: $I_{B1} = 90 \, \mu A$, $C_S = 50 \, \text{pF}$ and $C_2 = 47 \, \text{pF}$. Fig.14 shows the simulated frequency responses of the filter with three different values of $I_{B2}$. As a result, the achieved cut-off frequency and the quality factor of this filter are about $f_o = 1/2\pi(L_{eq2}C_2)^{1/2} \cong 4.25 \, \text{MHz}$, 5.60 MHz, 7.15 MHz and $Q = (1/R_{eq2})(L_{eq2}/C_2)^{1/2} \cong 1.2$, 0.9 and 0.7, respectively. In addition, the total harmonic distortion (THD) variations of $i_{LP}$ in Fig.13 against an applied sinusoidal input current amplitude at $f_o = 4.25 \, \text{MHz}$ are given in Fig.15.

### 4.3 Proposed Parallel Lossy Inductor in Fig.4

For the proposed grounded parallel R-L type lossy inductor in Fig.4, the component values chosen are $I_{B1} = I_{B2} = 90 \, \mu A$ and $C_P = 50 \, \text{pF}$, to obtain the ideal values of $R_{eq2} = 666 \, \Omega$ and $L_{eq2} = 22.21 \, \mu H$. The magnitude and phase characteristics of $Z_{m3}$ relative to frequency are depicted in Fig.16. The practical usability of the proposed parallel inductor is demonstrated on the voltage-mode HP filter shown in Fig.17. To design a filter with $f_o = \omega_0/2\pi(L_{eq3}C_3)^{1/2} \cong 4.67 \, \text{MHz}$, 6.14 MHz, 7.85 MHz, and $Q = R_{eq3}(C_3/L_{eq3})^{1/2} \cong 0.76$, 1, 1.28, it is realized by taking $C_3 = 40 \, \text{pF}$, $I_{B1} = 90 \, \mu A$ and $I_{B2} = 50 \, \mu A$, 150 μA, and 400 μA, which result in $R_{eq3} = 666 \, \Omega$ and $L_{eq3} = 30 \, \mu H$, 17 μH, and 10.5 μH, respectively. The simulated frequency responses of the realized HP filter with tuning $I_{B2}$ value are given in Fig.18.
5. Conclusions

Three configurations for simulating grounded lossless and lossy inductors are described. All the circuits employ two VDBAs and one grounded capacitor. The realized equivalent element values can be tuned electronically by adjusting the bias currents of the VDBAs. Moreover, all the proposed inductance simulator circuits do not need any critical constraints of component matching. The effects of VDBA non-ideal gains are examined. Several simulation results based on TSMC 0.25-μm CMOS technology are performed to verify the theoretical analysis. To demonstrate the practical workability of the proposed inductors, some examples of applications are also included.

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**Figure Captions:**

**Figure 1:** Symbolic representation of the VDBA.

**Figure 2:** Proposed grounded lossless inductance simulator.

**Figure 3:** Proposed grounded series-R-L type lossy inductance simulator.
Figure 4: Proposed grounded parallel R-L type lossy inductance simulator.

Figure 5: CMOS realization for the VDBA adopted from [23].
Figure 6: Simulated transient waveforms of $v_{in1}$ and $i_{in1}$ for the proposed lossless inductor of Fig. 2.

Figure 7: Theoretical and simulated frequency responses of $Z_{in1}$ for the proposed lossless inductor of Fig. 2.
Figure 8: Impedance magnitude-frequency responses of Fig.2 for three different values of $I_R$.

Figure 9: Tunable voltage-mode BP and HP filters constructed with the proposed lossless inductor in Fig.2.
Figure 10: Gain-frequency responses of the filter realization in Fig. 9 for three different values of $I_B$.

(a) BP responses  (b) HP responses
Figure 11: Simulated transient waveforms of $v_{in2}$ and $i_{in2}$ for the proposed series lossy inductor of Fig. 3.
Figure 12: Magnitude-frequency responses for $Z_{in2}$ of Fig. 3.

(a) with $I_{B1}$ tuning  
(b) with $I_{B2}$ tuning
Proposed series lossy inductor of Fig.3

Figure 13: Tunable current-mode LP filter realization using the proposed series lossy inductor of Fig.3.

![Proposed series lossy inductor of Fig.3](image)

Figure 14: Simulated frequency responses of LP filter in Fig.13.
Figure 15: THD variations of LP filter in Fig.13 against applied input signal amplitude.

Figure 16: Theoretical and simulated frequency responses of $Z_{in3}$ for the proposed parallel lossy inductor of Fig.4.
Figure 17: Tunable voltage-mode HP filter realization using the proposed parallel lossy inductor of Fig.4.

Figure 18: Simulated frequency responses of HP filter in Fig.17 for different $I_{B2}$ value.
### Table 1: Comparison of some exemplary grounded inductance simulators with the proposed VDBA-based simulators.

| Reference | Type of inductor | No. of active element | No. of resistors grounded | No. of resistors floating | No. of capacitors grounded | No. of capacitors floating | Electronic Tunability | Matching condition | Supply voltages (V) | Technology |
|-----------|------------------|-----------------------|---------------------------|--------------------------|---------------------------|---------------------------|---------------------|-------------------|------------------|-------------|
| [2]       | Fig.2            | lossless              | 1                         | 1                        | 1                         | 0                         | 1                   | no                | ±2.5, -0.604, -0.25 | TSMC 0.35-μm |
| [3]       | Fig.1            | lossless              | 1                         | 1                        | 0                         | 1                         | no                  | yes               | ±1.5, +0.7        | TSMC 0.35-μm |
| [4]       | Fig.2            | lossless              | 3                         | 2                        | 1                         | 1                         | no                  | no                | ±5               | AD844       |
|           | Fig.3-4          |                      |                           |                          |                           |                           |                     |                   |                  |             |
| [5]       | Fig.2(a)-2(e)    | lossless              | 1                         | 1                        | 1                         | 0                         | no                  | yes               | ±1.5             | TSMC 0.35-μm |
| [6]       | Fig.1            | lossless              | 2                         | 1                        | 2                         | 0                         | 2                   | no                | ±15              | MC1458 op amp |
| [7]       | Fig.3(a)-3(b)    | lossless              | 1                         | 3                        | 0                         | 0                         | 1                   | no                | ±1.65, +0.45     | AMS 0.35-μm |
| [8]       | Fig.3            | lossless              | 1                         | 0                        | 1                         | 0                         | 1                   | yes               | ±5               | OPA860      |
| [9]       | Fig.2(a)-2(b)    | series/parallel lossy| 1                         | 1                        | 1                         | 1                         | 0                   | no                | ±2.5             | ALA400      |
|           | Fig.2(c)         | series/parallel lossy| 2                         | 0                        | 4                         | 0                         | 1                   | no                | ±10              | AD844       |
| [10]      | Fig.2(c)         | series/parallel lossy| 2                         | 0                        | 5                         | 0                         | 1                   | no                | ±15              | AD844       |
|           | Fig.2-5          | series/parallel lossy| 1                         | 1                        | 1                         | 0                         | 1                   | no                | ±15              | AD844       |
| [11]      | Fig.2(a)-2(c)    | series/parallel lossy| 1                         | 1                        | 1                         | 0                         | 1                   | no                | ±2.5, +1.44      | TSMC 0.35-μm |
|           | Fig.2(d)         | series/parallel lossy| 1                         | 3                        | 0                         | 0                         | 1                   | no                | ±0.75, +0.34     | IBM 0.13-μm |
| [12]      | Fig.3(a)-3(b)    | series/parallel lossy| 1                         | 1                        | 1                         | 0                         | 0                   | no                | ±10              | Bipolar B101 & B102 |
|           | Fig.4            | lossess               | 1                         | 1                        | 1                         | 0                         | 1                   | no                | ±10              | TSMC 0.25-μm |
|           | Fig.5(a)-5(c)    | series/parallel lossy| 1                         | 2                        | 0                         | 0                         | 1                   | no                | ±0.75            | TSMC 0.25-μm |
| [13]      | Fig.2            | parallel lossy        | 1                         | 0                        | 2                         | 0                         | 1                   | no                | ±10              | TSMC 0.35-μm |
| [14]      | Fig.2            | series/parallel lossy| 1                         | 2                        | 0                         | 0                         | 1                   | no                | ±10              | TSMC 0.35-μm |
|           | Fig.3(a)-3(c)    | series/parallel lossy| 1                         | 3                        | 0                         | 0                         | 1                   | no                | ±10              | TSMC 0.35-μm |
|           | Fig.4            | series/parallel lossy| 2                         | 0                        | 0                         | 1                         | 0                   | yes               | ±0.75            | TSMC 0.25-μm |

**Proposed circuits**

| Fig.2    | parallel lossy | 1 | 0 | 2 | 0 | 1 | yes | no | ±12 | AD844 |
| Fig.2    | parallel lossy | 1 | 0 | 2 | 0 | 0 | no  | no | Not reported | 0.5-μm |

**Electronic Tunability**

- Lossless: No tuning required.
- Series Lossy: Tuning required for series elements.
- Parallel Lossy: Tuning required for parallel elements.

**Supply Voltages (V)**

- ±2.5, -0.604, -0.25
- ±1.5, +0.7
- ±5
- ±15
- ±10
- ±12
- Not reported
Table 2: Dimensions of the CMOS transistors of the VDBA circuit in Fig.5.

| Transistors       | $W(\mu m)/L(\mu m)$ |
|-------------------|----------------------|
| $M_1 - M_2, M_5 - M_6$ | 25/0.25              |
| $M_3, M_7$        | 32/0.25              |
| $M_4, M_8$        | 35/0.25              |
| $M_9$             | 25/0.25              |