CHAPTER – 3
ACTIVE INDUCTANCE SIMULATION

The content and results of the following papers have been reported in this chapter.


3.1 INTRODUCTION

Inductors find application in areas such as filter design, oscillator design, phase shifters and parasitic element cancellation. But realization of a spiral inductor in an integrated circuit has some drawbacks in terms of the usage of space, weight, cost and tunability. This resulted in design of inductorless active RC filters, oscillators etc., thus making active inductance simulation an important research topic in active network synthesis. The simulated active inductors only mimic some properties of the real inductors and cannot replace them in all possible applications of inductors. Despite this limitation, their usages are widespread in analog design which leads the analog designers to explore this area to the extent possible.

This chapter deals with the realization of both lossy and lossless active grounded inductance simulation using OTRA. The chapter begins with the design description of lossy grounded inductance which is followed by design details concerning lossless grounded inductance simulation. A brief account of existing literature on OTRA based grounded inductance simulation is presented before describing the proposed work.

3.2 LITERATURE REVIEW ON INDUCTANCE SIMULATION

Review of earlier work suggests that both lossy and lossless grounded inductor topologies using OTRA have been reported in open literature. A number of grounded parallel lossy inductor topologies using single OTRA were proposed in [59], [60], which can realize L parallel with R, (-L) parallel with R, (-L) parallel with (±R), (±L) parallel with R; and L and C parallel with (±R). Two topologies which simulate grounded parallel lossy inductor using two OTRAs were proposed in [62]. These topologies realized L parallel with R and (-L) parallel with R. A lossless grounded inductor using two OTRAs was also proposed in [62] apart from lossy inductor topologies. A lossless grounded inductor topology using two OTRAs was presented in [61] and this inductor was used to design an LC oscillator. A lossless grounded inductance simulator using an OTRA and a buffer [63] came up recently in 2011. This topology could also simulate grounded frequency-dependent negative-resistance (FDNR). Actively simulated negative inductors play an important role in
cancellation/compensation of parasitic inductances and can also be used in microwave circuits for impedance matching. A single OTRA based generic grounded negative inductance simulator was proposed in [64] from which four different topologies could be deduced.

### 3.3 LOSSY GROUNDED INDUCTOR

This section presents the design of proposed lossy grounded inductor which can realize inductance types ($\pm L$) parallel with $R$. Two application example of the proposed lossy inductor are also discussed.

#### 3.3.1 Proposed Circuit

![Fig. 3.1 Lossy grounded inductor.](image)

The proposed grounded immittance simulator is shown in Fig. 3.1. Using routine analysis of the circuit the expression for input admittance $Y_{in}$ can be written as

$$Y_{in} = G_1 + G_2 - \frac{G_1 G_2}{s(C_1 - C_2)}$$

(3.1)

which represents an impedance of type $L_{eq} \parallel R_{eq}$, where
\[
R_{eq} = \frac{1}{g_1 + g_2}, \quad L_{eq} = \frac{(C_1 - C_2)}{g_1 g_2} \quad \text{and} \quad R_l = \frac{1}{g_l}
\] (3.2)

Proper choice of \(C_1\) and \(C_2\) may result in realization of inductance type (+L) parallel with R or inductance type (-L) parallel with R.

### 3.3.2 Nonideality Analysis

In analysis so far, ideal characteristics of the OTRA have been considered. However, the effect of the parasitics needs to be taken into consideration for performing nonideality analysis. Using the nonideal model of OTRA as shown in Fig. 2.8, the circuit of Fig. 3.1 can be redrawn as shown in Fig. 3.2.

![Fig. 3.2 Nonideal model of the proposed lossy inductor.](image-url)
Nodal equations at input node and nodes 1 and 2 can respectively be written as

\[
\frac{(v_{in} - v_1)}{R_1} = i_{in} + \frac{(v_o - v_{in})}{R_2} \tag{3.3}
\]

\[
\frac{(v_{in} - v_1)}{R_1} + (v_o - v_1)sC_2 = \frac{v_1}{R_{x1}} \tag{3.4}
\]

\[
(v_o - v_2) sC_1 = \frac{v_2}{R_{x1} || R_{x2}} \tag{3.5}
\]

Equations (3.3), (3.4) and (3.5) can be solved, assuming \(R_{x1} = R_{x2} = R_x\) and using the relation \(R_x \ll R_{z1}\), resulting in

\[
i_{in} = (G_1 + G_2) v_{in} - \frac{G_1 G_2 (1 + sC_1 R_x (1 + R_2))}{s(C_1 - C_2)} v_{in} \tag{3.6}
\]

From (3.6) the input admittance can be computed to be

\[
Y_{in}(s)\bigg|_{in} = (G_1 + G_2) - \frac{G_1 G_2}{s(C_1 - C_2)} (1 + sC_1 R_x (1 + R_2)) \tag{3.7}
\]

### 3.3.3 Simulation Results

The functionality of the proposed simulated lossy inductor is verified through SPICE simulation using CFOA based realization of OTRA as discussed in section 2.4.1. Impedance magnitude response of the simulated lossy inductor having \(L_{eq} = 1\,\text{mH}\) and \(R_{eq} = 500\,\Omega\), realized using component values as \(C_1 = 2\,\text{nF}\), \(C_2 = 1\,\text{nF}\), and \(R_1 = R_2 = 1\,\text{K}\Omega\) is given in Fig. 3.3(a). Impedance magnitude response of another instance of lossy inductor with \(L_{eq} = 1\,\mu\text{H}\) and \(R_{eq} \approx 10\,\Omega\), implemented through \(C_1 = 2\,\text{nF}\), \(C_2 = 1\,\text{nF}\), \(R_1 = 100\,\Omega\) and \(R_2 = 10\,\Omega\); is depicted in Fig. 3.3(b). It is observed that the impedance value remains within ± 10% of the theoretically calculated value in the frequency range of 100 Hz – 100 KHz for 1 mH. For inductance value of 1 \(\mu\text{H}\) the frequency range is found to be 10 KHz - 1.5 MHz. It indicates that the frequency range, over which the inductance value remains almost constant, decreases with increasing value of simulated inductance.
3.3.4 Signal Processing Applications

To show the application of the proposed immittance simulator a current mode filter, giving

Fig 3.3 Impedance magnitude responses (a) $L_{eq} = 1\text{mH}, R_{eq} = 500\Omega$,
(b) $L_{eq} = 1\mu\text{H}, R_{eq} = 10\Omega$. 
high pass and band pass responses, and an LC oscillator are designed. The workability of these applications is verified through SPICE simulation.

### 3.3.4.1 Current Mode Filter

The designed current mode filter configuration is shown in Fig. 3.4. The simulated lossy inductor replaces the parallel R L circuit. Current transfer functions can be written as

\[
\frac{i_R}{i_n} = \frac{s(G_{eq}+G)}{s^2 + s\left(\frac{G+G_{eq}}{C}\right) + \frac{1}{cL_{eq}}} \tag{3.8}
\]

\[
\frac{i_C}{i_n} = \frac{s^2}{s^2 + s\left(\frac{G+G_{eq}}{C}\right) + \frac{1}{cL_{eq}}} \tag{3.9}
\]

where \(G_{eq} = 1/ R_{eq}\) and \(G = 1/ R\). The filter functions can be characterized by following parameters

\[
\omega_0 = \frac{1}{\sqrt{cL_{eq}}}, \quad Q_0 = \frac{1}{(G+G_{eq})\sqrt{L_{eq}}} \tag{3.10}
\]

![Fig. 3.4 Current Mode Filter.](image)

To verify the functionality of the proposed current mode filter a design having \(f_0 = 79.6\) KHz and \(Q_0 = 0.25\) is developed, for which the capacitance value \(C = 1\) nF is chosen and accordingly \(L_{eq}\) is computed to be 4 mH and \(R = 1\) KΩ. For realizing \(L_{eq} = 4\) mH, the values of capacitive components are chosen as \(C_1 = 2\) nF and \(C_2 = 1\) nF and the resistive
components are calculated as $R_1 = R_2 = 2 \, \text{K}\Omega$. The simulated frequency responses for the current mode BP and HP filters are shown in Fig. 3.5 and are found to be in close agreement with the theoretical predictions.

![Figure 3.5 Frequency Response of current mode filter.](image)

3.3.4.2 Realization of an LC Oscillator

An LC oscillator is designed using new simulated inductor and is shown in Fig. 3.6. From the routine analysis of the circuit various current equations can be written as

\[ I_p = \frac{v_1}{R_1}, \quad I_n = \frac{v_o}{R_2} \]  \hspace{1cm} (3.11)

Therefore $v_1$ can be expressed as

\[ v_1 = \frac{R_1}{R_2} v_o \]  \hspace{1cm} (3.12)

Nodal equation at node 1 can be written as

\[ \frac{v_1}{R_1} + \frac{v_1}{Z} + \frac{(v_1 - v_o)}{R_3} = 0 \]  \hspace{1cm} (3.13)

where $Z = X_C \parallel X_{Leq} \parallel R_{eq} = \frac{R_{eq}L_{eq}}{s^2C R_{eq}L_{eq} + sL_{eq} + R_{eq}}$  \hspace{1cm} (3.14)

Substituting $v_1$ and $Z$ from (3.12) and (3.14) respectively in (3.13) results
\[ s^2 + \frac{s}{CR_{eq}} + \frac{1}{CL_{eq}} = \frac{R_2 - (R_1 + R_3)}{CR_1 R_3} \]  

(3.15)

From (3.15) the condition of oscillation (CO) and frequency of oscillation (FO) can be expressed as

**CO:** \[ R_2 = (R_1 + R_3) \]  

(3.16)

**FO:** \[ \omega_0 = \frac{1}{\sqrt{CL_{eq}}} \]  

(3.17)

![Fig. 3.6 LC oscillator using realized lossy inductor.](image)

Fig. 3.6 LC oscillator using realized lossy inductor.

A typical simulation for element values \( R_1 = 4 \, \text{K}\Omega \), \( R_2 = 8 \, \text{K}\Omega \), \( R_3 = 4 \, \text{K}\Omega \), \( R_{eq} = 16 \, \text{K}\Omega \), \( L_{eq} = 0.64 \, \text{mH} \) and \( C = 10 \, \text{pF} \) is shown in Fig. 3.7. The simulated \( f_0 \) is observed to be 1.92 MHz as against the theoretical value of 1.99 MHz and are in close agreement.

![Fig. 3.7 Simulation result of LC oscillator.](image)
3.4 LOSSLESS GROUNDED INDUCTANCE SIMULATION

This section deals with OTRA based lossless grounded active inductance simulation and describes two different design topologies involving (i) two OTRAs and (ii) single OTRA.

3.4.1 Two OTRA Based Circuits

Two different topologies of lossless grounded inductor using two OTRAs, in addition to already existing structures in the literature [61], [62] are proposed in this section and are shown in Fig. 3.8 (a) and (b) respectively. The input admittance of the circuit of Fig. 3.8 (a) can be computed as

\[ Y_{in}(s) = (G_1 + G_3 + G_5 - \frac{G_1 G_5}{G_4} ) + \frac{G_2 G_3 G_5}{sC_1 G_4} \]  

(3.18)

The input admittance \( Y_{in} \) will be purely inductive if following condition is met

\[ G_1 + G_3 + G_5 = \frac{G_1 G_5}{G_4} \]  

(3.19)

Similarly for inductance topology of Fig. 3.8 (b) input admittance is given by

\[ Y_{in}(s) = (G_1 + G_2 + G_4 - \frac{G_2 G_4}{G_3} ) + \frac{G_1 G_2 G_5}{sC_1 G_3} \]  

(3.20)

It will be purely inductive provided

\[ G_1 + G_2 + G_4 = \frac{G_2 G_4}{G_3} \]  

(3.21)

The equivalent inductance values along with conditions are summarized in Table 3.1. It is clear from Table 3.1 that for both the topologies the inductance value can be controlled independent of condition of realization.
Fig. 3.8 Lossless grounded inductors. (a) Topology-I. (b) Topology-II.
Table 3.1: Inductors realized by the topologies shown in Fig. 3.8.

<table>
<thead>
<tr>
<th>Figure</th>
<th>Condition</th>
<th>$L_{eq}$</th>
<th>Non Interactive Control</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fig. 2(a)</td>
<td>$g_1 + g_3 + g_5 = \frac{g_1g_5}{g_4}$</td>
<td>$\frac{c_1g_4}{g_2g_3g_5}$</td>
<td>Independent Control on condition through $g_1$ and on value through $g_2$.</td>
</tr>
<tr>
<td>Fig. 2(b)</td>
<td>$g_1 + g_2 + g_4 = \frac{g_2g_4}{g_3}$</td>
<td>$\frac{c_1g_3}{g_1g_2g_5}$</td>
<td>Independent Control on condition through $g_4$ and on value through $g_5$.</td>
</tr>
</tbody>
</table>

3.4.1.1 Simulation Results

Two inductor instances of value $L_{eq} = 100 \mu H$ and $L_{eq} = 10 \mu H$ are simulated using inductor topologies of Fig. 3.8 (a) and (b) respectively. To obtain these inductance values the component values are chosen as $R_1 = R_2 = R_3 = R_5 = 1 \, K\Omega$, $R_4 = 3 \, K\Omega$, $C_1 = 300 \, pF$ for $L_{eq} = 100 \mu H$ and $R_1 = R_2 = R_4 = R_5 = 1 \, K\Omega$, $R_3 = 3 \, K\Omega$, $C_1 = 30 \, pF$ for $L_{eq} = 10 \mu H$ respectively. The CMOS schematic of OTRA shown in Fig. 2.9 is used for simulation. The ideal and simulated impedance magnitude responses of 100 $\mu H$ and 10 $\mu H$ inductors are shown in Fig 3.9 (a) and (b) respectively. The inset depicts the enlarged view of magnitude response in lower frequency range. It is observed that the inductance instance of 100 $\mu H$ remains well within the ± 10% of the designed value in the frequency range of 3 KHz - 1.2 MHz whereas that of 10 $\mu H$ in the frequency range of 10 KHz - 1.2 MHz.

3.4.1.2 Applications

In this section some applications of the proposed topologies have been presented. Both the topologies may be used for constructing filter and oscillator circuits.
Fig. 3.9 Impedance magnitude response (a) 100 µH. (b) 10 µH.
3.4.1.2.1 High Pass Filter

A high pass filter, as shown in Fig. 3.10 (a), can be constructed using proposed inductors.

![Simulated Inductance based HPF.](image)

Fig. 3.10 (a) Simulated Inductance based HPF. (b) Frequency response of HPF.

The transfer function for high pass response is obtained as

\[
\frac{V_o}{V_{in}} = \frac{s^2 + \frac{s}{CR} + \frac{1}{LeqC}}{s^2 + \frac{s}{CR} + \frac{1}{LeqC}}
\]

(3.22)

and is characterized by

\[
\omega_0 = \frac{1}{\sqrt{C Leq}} \quad \text{and} \quad Q_0 = R \sqrt{\frac{C}{Leq}}
\]

(3.23)

The functionality of the HPF is verified by designing a filter having lower cutoff frequency of 503.2 KHz and \(Q_0 = 1\). The component values for the design are chosen as \(Leq = 0.1 \text{ mH}\) and \(C = 1 \text{ nF}\). The value of resistor \(R\) is computed to be 300 \(\Omega\). Inductor topology of Fig. 3.8(a) is
used for realizing the grounded inductor of value 0.1 mH, with component values as 
\[ C_1 = 300 \text{ pF and } R_1 = R_2 = R_3 = R_5 = 1 \text{ K}\Omega, \ R_4 = 3 \text{ K}\Omega. \] 
The simulated frequency response of the HPF using SPICE is depicted in Fig. 3.10(b). Simulated value of lower cut off frequency is obtained as 505 KHz which is in close agreement to the theoretical value of 503.2 KHz.

### 3.4.1.2.2 Band Pass Filter

The proposed inductor topologies may also be used to obtain band pass response using the circuit given in Fig. 3.11(a). Using routine analysis the transfer function for band pass response can be obtained as

\[
\frac{v_o}{v_{in}} = \frac{s}{s^2 + \frac{s}{CR} + \frac{1}{L_{eq}C}}
\]

(3.24)

where \[ \omega_0 = \frac{1}{\sqrt{CL_{eq}}} \] and \[ Q_0 = \frac{C}{L_{eq}} \]

(3.25)

![Fig. 3.11(a) Simulated Inductance based BPF. (b) Frequency response of BPF.](image)

(a) (b)
This theoretical proposition is verified through simulations using the topology of Fig. 3.8 (b). A BPF is designed having center frequency of 503.29 KHz. The component values are computed as $R = 1 \, \text{K}\Omega$ and $C = 1 \, \text{nF}$ for a chosen value of $L_{eq} = 0.1 \, \text{mH}$. The value of $L_{eq} = 0.1 \, \text{mH}$ is obtained using component values of $R_1 = R_2 = R_4 = R_5 = 1 \, \text{K}\Omega$, $R_3 = 3 \, \text{K}\Omega$, and $C_1 = 300 \, \text{pF}$. The simulated frequency response of the BPF is depicted in Fig. 3.11(b). It may be observed that the simulated and theoretical responses closely follow each other.

### 3.4.1.2.3 LC Oscillator

An LC Oscillator can also be realized using the proposed inductor topologies. Fig. 3.12 (a) shows the schematic of an LC oscillator using topology of Fig. 3.8 (a) for which the condition of oscillation (CO) and frequency of oscillation (FO) can be computed as

$$\text{CO: } G_1 + G_3 + G_5 = \frac{G_1 G_5}{G_4} \quad (3.26)$$

$$\text{FO: } \omega_0 = \frac{1}{\sqrt{L_{eq} C}} = \sqrt{\frac{R_4}{C C_1 R_2 R_3 R_5}} \quad (3.27)$$
Figure 3.12(b) show simulated output of oscillator for $L_{eq} = 0.1$ mH and $C = 300$ pF. Simulated frequency of oscillation is found to be 860 KHz as against the calculated value of 876.2 KHz with % error of 1.85%.

### 3.4.1.3 Experimental Verification

The functionality of proposed inductor topologies is also confirmed experimentally. The OTRA is realized using two CFOAs (IC AD844AN) for experimental work. The HPF of Fig. 3.10 (a) is prototyped with $R = 680$ $\Omega$, $C = 1$ nF and $L_{eq} = 1$ mH. Inductor topology of Fig 3.8 (a) is used with component values $R_1 = R_2 = R_3 = R_5 = 1$ K$\Omega$, $R_1 = 3$ K$\Omega$ and $C_1 = 3$ nF, to simulate $L_{eq} = 1$ mH. Theoretical, simulated and experimental frequency responses are shown in Fig. 3.13 and it is observed that the experimental response is almost in agreement with the theoretical and simulated responses.

The oscillator circuit of Fig. 3.12(a) is also tested experimentally. The output waveform observed on oscilloscope is shown in Fig. 3.14. The observed frequency of oscillation is found to be 872.5 KHz, which is in close agreement to theoretically calculated value of 876.2 KHz.
In this section a grounded simulated inductor topology based on a single OTRA is proposed. It provides non interactive control between inductance value and realizabilty condition. The proposed lossless grounded inductor is shown in Fig. 3.15. From routine analysis of the circuit the input admittance $Y_{in}(s)$ can be expressed as

$$Y_{in}(s) = (3G - \frac{C_1G}{C_2}) + \frac{G^2}{sC_2}, \text{ where } G = 1/R$$  \hspace{1cm} (3.28)

The admittance $Y_{in}$ will be purely inductive if following condition is met
\[ C_1 = 3C_2 \]  
The inductance value that results is
\[ L_{eq} = \frac{C_2}{G^2} = C_2 R^2 \]

Thus inductance value can be adjusted by appropriate selection of R without disturbing the condition of realization of inductor, as it does not depend on resistance R. It may be noted that the floating resistors can be implemented using one-port active MOS resistor architecture [76], [77] wherein the resistance values can be adjusted by simply changing the dc control voltage thus making inductor value electronically tunable.

### 3.4.2.1 Nonideality Analysis

In this section the effect of finite transresistance gain on inductor is considered and for high frequency applications, passive compensation is employed. Taking the effect of nonideality of OTRA into account (3.28) modifies to

\[ Y_{in}(s)|_{in} = 2G + sC_1 - \frac{(sC_1-G)(sC_2+G)\varepsilon uc(s)}{sC_2} \]  
\[ (3.31) \]
where $\varepsilon_{uc}(s) = \frac{1}{1+(1 + \frac{sC_2}{g}) \frac{C_p}{C_2}}$ \hspace{1cm} (3.32)

is uncompensated error function.

\[ Y_{in}(s)|_{n_c} = 2G + sC_1 - \frac{(sC_1-G)(sC_2+G)}{sC_2} \varepsilon_c(s) \hspace{1cm} (3.33) \]

where $\varepsilon_c(s) = \frac{1}{1+(sC_p-Y)\frac{(G+sC_2)}{Gsc_2}}$ \hspace{1cm} (3.34)

is compensated error function.

By taking $Y = sC_p$, $\varepsilon_c(s)$ reduces to 1, which makes (3.33) same as (3.28). The effect of single pole model of $R_m$ can thus be eliminated by connecting a single capacitor having value $C_p$ in place of $Y$ as shown in Fig. 3.16.
3.4.2.2 Simulation Results

The performance of the simulated inductor is evaluated with SPICE simulation using CMOS schematic of OTRA shown in Fig. 2.9. Impedance magnitude responses of simulated and ideal inductors for $L_{eq} = 10 \, \mu\text{H}$ and $1\text{mH}$ are given in Fig. 3.17(a) and (b) respectively. Component values for $L_{eq} = 10 \, \mu\text{H}$ are chosen as $R = 1 \, \text{K}\Omega$, $C_1 = 30 \, \text{pF}$ and $C_2 = 10 \, \text{pF}$ and those for $L_{eq} = 1 \, \text{mH}$ are $R = 10 \, \text{K}\Omega$, $C_1 = 30 \, \text{pF}$ and $C_2 = 10 \, \text{pF}$. Inductance value remains within $\pm 10\%$ in the frequency range of 8 KHz - 5.0 MHz for $10 \, \mu\text{H}$ whereas for $1 \, \text{mH}$ the frequency range is found to be 200 Hz - 2.5 MHz. It indicates that the frequency range, over which the inductance value remains almost constant, decreases with increasing value of simulated inductance. The dynamic range of the proposed active inductor is simulated to be 71 dB. Total power consumption of the proposed inductor is simulated to be 0.809 mW.

3.4.2.3 Applications

In this section the workability of the proposed grounded inductor has been illustrated by realizing a BP filter and an LC oscillator.

3.4.2.3.1 Band Pass Filter

A BP filter, as shown in Fig. 3.18 (a), is constructed using proposed inductor and using routine analysis the transfer function can be obtained as

$$\frac{v_o}{v_{in}} = \frac{s}{s^2 + \frac{s}{C_B R_B} + \frac{1}{L_{eq} C_B}}$$  \hspace{1cm} (3.35)

where $\omega_0 = \frac{1}{\sqrt{L_{eq} C_B}} = \frac{1}{R \sqrt{C_2 C_B}}$, $Q_0 = R_B \frac{C_B}{\sqrt{L_{eq}}} = \frac{R_B}{R} \frac{C_B}{C_2}$  \hspace{1cm} (3.36)

This suggests that the $Q_0$ can be independently controlled by varying $R_B$ without affecting the centre frequency $\omega_0$. Passive sensitivities for $\omega_0$ and $Q_0$ can be calculated as follows

$$S_{C_2}^{\omega_0} = S_{C_B}^{\omega_0} = -\frac{1}{2}, S_R^{\omega_0} = -1, S_{R_B}^{Q_0} = 1, S_{R}^{Q_0} = -1, S_C^{Q_0} = \frac{1}{2}, S_C^{Q_0} = -\frac{1}{2}$$  \hspace{1cm} (3.37)
Fig. 3.17 Impedance magnitude response (a) $L_{eq} = 10 \, \mu H$ (b) $L_{eq} = 1 \, mH$. 
To see the correctness of the theoretical proposition, the BP filter of Fig. 3.18 (a) is designed having a centre frequency of 1.59 MHz. The component values are computed as $R_B = 1 \, \text{K} \Omega$, $C_B = 1 \, \text{nF}$ and $L_{eq} = 10 \, \mu\text{H}$. The $L_{eq} = 10 \, \mu\text{H}$ is obtained by choosing $R = 1 \, \text{K} \Omega$, $C_1 = 30 \, \text{pF}$ and $C_2 = 10 \, \text{pF}$. The simulated frequency response of the filter using CMOS OTRA of Fig. 2.9 is depicted in Fig. 3.18 (b). The simulated results are in close agreement with the theoretical predictions. Figure 3.18 (c) shows the frequency response of the BPF, for different $Q_0$ values as 0.5, 5, 10 and 20, for which value of $R_B$ is chosen as 50 $\\Omega$, 100 $\\Omega$, 1 K$\\Omega$, and 2 K$\\Omega$ respectively while keeping $C_B = 1 \, \text{nF}$ and $L_{eq} = 10 \, \mu\text{H}$ as constant.

### 3.4.2.3.2 LC Oscillator

Using the BPF of Fig. 3.18 an LC oscillator can be realized as shown in Fig. 3.19 (a), for which the characteristic equation is obtained as

$$s^2 + \frac{s((R_{o1}+R_{o2})-R_{o3})}{R_{o2}(R_{o2}+R_{o1})} + \frac{1}{L_{eq}C_{o1}} = 0 \quad (3.38)$$

From characteristic equation the CO and FO can be derived as

CO: \quad $(R_{o1} + R_{o2}) = R_{o3} \quad (3.39)$

FO: \quad $f_0 = \frac{1}{2\pi\sqrt{(L_{eq}C_{o1})}} \quad (3.40)$

The simulated output waveform of the oscillator topology of Fig. 3.19 (a) for component values of $R_{o1} = 4.6 \, \text{K} \Omega$, $R_{o2} = 400 \, \Omega$, $R_{o3} = 5 \, \text{K} \Omega$, $C_{o1} = 1 \, \text{nF}$ and $L_{eq} = 10 \, \mu\text{H}$ is shown in Fig. 3.19 (b) and the output frequency spectrum is depicted in Fig. 3.19 (c). The simulated frequency of oscillation being 1.53 MHz is in close agreement with the theoretically calculated value of 1.59 MHz. The $\%$ total harmonic distortion ($\%$THD) being 0.49$\%$, is a considerably low value.
Fig. 3.18 (a) BPF using single OTRA based simulated inductance. (b) Frequency response.

(c) BP response for different $Q_0$ values with $\omega_0 = 1.59$ MHz.
3.4.2.4 Experimental Verification

The functionality of the proposed grounded inductor is verified experimentally also. The commercial IC AD844AN is used to implement an OTRA as shown in Fig. 2.7 with a supply voltage of ± 5V. The BPF of Fig. 3.18 (a) is prototyped with $C_B = 1 \text{nF}$, $R_B = 1 \text{KΩ}$, and $L_{eq} = 100 \mu\text{H}$. The $L_{eq} = 100 \mu\text{H}$ is implemented with component values of $C_1 = 330 \text{pF}$, $C_2 = 100 \text{pF}$ and $R = 1 \text{KΩ}$. Theoretical, simulated (using macromodel of AD844) and experimental frequency responses are shown in Fig. 3.20. It is observed that the experimental cut off frequency (478 KHz) is close to theoretical (501 KHz) and simulated (493 KHz).
values. The oscillator circuit of Fig. 3.19 (a) is also tested experimentally for $C_{o1} = 1 \, \text{nF}$ and $L_{eq} = 5 \, \mu\text{H}$. The $L_{eq} = 5 \, \mu\text{H}$ is implemented with component values of $R = 680 \, \Omega$, $C_1 = 33 \, \text{pF}$ and $C_2 = 10 \, \text{pF}$. The output waveform observed on oscilloscope is shown in Fig. 3.21. The observed frequency of oscillation is found to be 2.47 MHz, as against theoretically calculated value of 2.3 MHz. The minor deviations in experimental results can be attributed to component tolerance of $\pm 10\%$, used for experiments.

Fig. 3.20 Ideal, simulated and experimental frequency responses of BPF of Fig. 3.18 (a).

Fig. 3.21 Experimental output of LC oscillator of Fig. 3.19 (a).
3.4.3. Comparison

Table 3.2 shows the comparison of the proposed OTRA based grounded inductors with the previously reported lossless grounded inductors [61] – [64]. The study of Table 3.2 reveals that topologies presented in [61], [62] and those proposed in section 3.4.1 use more number of active and passive components as compared to the proposed single OTRA based topology. In the most recently published topology [63] though the number of passive components is reduced by one yet it uses an extra active element (buffer) as compared to the proposed circuit. The topology in [64] uses same number of active components and an extra passive component and simulates negative grounded inductor.

To compare the performance of all the circuits reported in [61] – [64] with proposed work in terms of power consumption and frequency range of operation an inductance ($L_{eq}$) value of 1 mH was implemented using CFOA based OTRA realization with supply voltage of ± 10V. Table 3.3 records the power consumption of various circuits and the observations indicating the range of frequency in which the inductor value remains well within the ± 10% of the designed value. Similar observations with CMOS based OTRA realization as given in Fig. 2.9, with supply voltages of ± 1.5V are also listed in Table 3.3. It is observed that the proposed single OTRA based topology outperforms both in terms of frequency range and power consumption, except for lossless negative inductor [64]. The components required for implementing an inductance ($L_{eq}$) value of 1 mH are given in the last column of Table 3.3. This suggests that the proposed single OTRA based circuit consumes the optimum chip area in terms of active component count and also the area used in terms passive components is minimum as compared to all existing circuits except the one presented in [63].
Table 3.2: Comparison of proposed lossless grounded inductor topologies with previously reported work.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>No. of active components</th>
<th>No. of passive components</th>
<th>Passive component matching required</th>
<th>Type of simulated inductor</th>
<th>Non-interactive tuning of $L_{eq}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>[61]</td>
<td>Two OTRA</td>
<td>Single capacitor, Five resistors</td>
<td>Yes</td>
<td>Lossless inductor</td>
<td>Yes</td>
</tr>
<tr>
<td>[62]</td>
<td>Two OTRA</td>
<td>Single capacitor, Five resistors</td>
<td>Yes</td>
<td>Lossless Inductor</td>
<td>Yes</td>
</tr>
<tr>
<td>Proposed Work of section 3.4.1</td>
<td>Two OTRA</td>
<td>Single capacitor, Five resistors</td>
<td>Yes</td>
<td>Lossless Inductor</td>
<td>Yes</td>
</tr>
<tr>
<td>[63]</td>
<td>Single OTRA Two; An OTRA and a Buffer</td>
<td>Two capacitors, Two resistors</td>
<td>Yes</td>
<td>Lossless Inductor</td>
<td>Yes</td>
</tr>
<tr>
<td>[64]</td>
<td>Single OTRA</td>
<td>Single capacitor, Five resistors</td>
<td>Yes</td>
<td>Lossless negative Inductor</td>
<td>Yes</td>
</tr>
<tr>
<td>Proposed Single OTRA based topology</td>
<td>Single OTRA</td>
<td>Two capacitors, Three resistors</td>
<td>Yes</td>
<td>Lossless Inductor</td>
<td>Yes</td>
</tr>
</tbody>
</table>
Table 3.3: Useful frequency range and power consumption of 1 mH inductor.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>CFOA based OTRA realization</th>
<th>CMOS based OTRA realization</th>
<th>Components used for $L_{eq} = 1$ mH implementation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Frequency Range</td>
<td>Power consumption</td>
<td>Frequency Range</td>
</tr>
<tr>
<td>[61]</td>
<td>150Hz-100KHz</td>
<td>0.533W</td>
<td>200Hz-1MHz</td>
</tr>
<tr>
<td>[62]</td>
<td>150Hz-30KHz</td>
<td>0.553W</td>
<td>200Hz-1.2MHz</td>
</tr>
<tr>
<td>Proposed</td>
<td>200Hz-100KHz</td>
<td>0.53W</td>
<td>200Hz-1.2MHz</td>
</tr>
<tr>
<td>work of</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>section</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3.4.1</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>[63]</td>
<td>200Hz-100KHz</td>
<td>0.53W</td>
<td>300Hz-2MHz</td>
</tr>
<tr>
<td>[64]</td>
<td>250Hz-350KHz</td>
<td>0.262W</td>
<td>200Hz-3MHz</td>
</tr>
<tr>
<td>Proposed</td>
<td>200Hz-100KHz</td>
<td>0.26</td>
<td>200Hz-2.5MHz</td>
</tr>
</tbody>
</table>
3.5 CONCLUDING REMARKS

Following grounded parallel inductance simulation topologies are proposed in this chapter

1. Single OTRA based lossy inductance simulator
2. Two topologies of lossless inductance using two OTRAs
3. Single OTRA based lossless grounded inductor

In lossless inductance topologies the value of simulated inductance can be tuned independent of the condition of realization. The effect of nonidealities of OTRA on the proposed inductance’s performance has been analyzed and compensation methods for high frequency applications are also presented. Application examples such as filters and LC oscillators using various proposed topologies have been included to demonstrate their practical use. SPICE simulations and experimental results are included to verify the theoretical propositions. It is found that the results obtained are in close agreement with the ideal values. Hence it is expected that the proposed topologies will provide a design option to integrated circuit designers where grounded lossy/lossless inductor applications are required.