BANDWIDTH EXTENSION OF ACTIVE INDUCTOR

Mostapha Boussetta, Ridouane Hamdaouy, Khadija Slaoui

Abstract—This paper deals with the design of a microwave active inductor using 0.13 µm CMOS technology. The proposed circuit configuration is based on the Gyrator-C topology. Such circuits are mainly made out of simple current mirrors requiring essentially low power consumption. Compared to typical spiral inductors, circuits of active inductors provide small tuning inductance values, high quality factor and tiny chip area. Several simulations have been carried out by means of SPECTRE software leading in the configuration of the expected high standard performances. The simulated inductor’s parameters validates the proposed design process and allows several microwave applications to be implemented as band pass filter.

Index Terms—active inductor, CMOS, Transconductance, current mirror, Gyrator-C, 0.13µm technologie, band pass filter.

I. INTRODUCTION:
Due to the scaling of the transistor, the cut-off frequency of the 0.13-um transistor reaches tens of gigahertz. for this reason, we exploit the high frequency attribute of the transistor to design the RLC passive filter[1].The Sio2 insulator of the mos transistor replaces the poly -capacitance to obtain the circuit with small-size, low-cost, and high integration. Through the specter simulation, the RLC bandpass filter is exploited around 200MHz -300MHz which is the if filter design specification of modern wireless mobile phones. The center frequency f0 and the Q-factor are tuned by varying two independent dc bias.

Active inductors are based on the gyrator principle. A symbol representing such inductors is shown in Figure .1. The forward path transconductor G, has a positive transconductance while the transonductor in the feedback path, -G, has a negative transconductance. Inductance is simulated by the capacitance C. For this reason, the synthesized inductor is called a gyrator-C active inductor with an inductance value \[ \frac{C}{(G*G)}. \]

For gyrator based simulated inductors the inductance value can be changed either by modifying the load capacitors (varactors, switched-capacitors) or by changing the gyration resistance. However in a filter configuration based on gyrator simulated inductor, changing its value usually leads to changes in its central frequency and Q-factor. The same behavior occurs in the case of many transistor simulated inductors RF filter.

II. ACTIVE INDUCTOR:
One type of active inductor shown in (Figure.2) were proposed in Wu’s paper [3]. In Figure .2 the input signal at the drain of M1 generates a current \( gm2 \) vi ,that charges Cgs1through the feedback path. The voltage vx then provides for the current \( gm1v_{gs1} \) \( (V_{gs1}=V_X) \) and generates an inductive loading effect.

Figure.2: Schematic diagram of a gyrator based active inductor.

On one hand, a positive transconductor is obtained by means of a common gate configuration. On the other hand, a negative transconductor. Cgs stands for C in the circuit model. Single ended active inductor circuit illustrating both the model (figure 1) and the real structure (figure 2). Cgs stands for C in the circuit model.

On one hand, a positive transconductor is obtained by means of a common gate configuration. On the other hand, a common source stage leads to a negative transconductor.
Figure 2: Wu active inductor.

Another way to look at this is that at low frequencies the impedance at the input node is low due to the shunt feedback as the frequency increases, the current starts to flow through the gate-source capacitance and causes the loop gain to reduce. Hence, the impedance increases as the frequency increases, emulating the impedance of an inductor [4]. The active inductor shown in Figure 2 is chosen for this paper because it works better at higher frequencies and has superior noise performance. The small-signal model of such an the active inductor is shown in Figure 3.

Figure 3: Small-Signal Model of an Active Inductor in high frequency.

Assuming an ideal current source, the input admittance of the active inductor can be written
\[ Y_{in} = g_{o1} + \left( g_{m_1} \right) \left( g_{m_2} + g_{o_2} + sCgd_1 \right) \]
\[ \frac{sCgd_1 + sCgs_1 + g_{o_2} + sCgd_2}{sCgd_1 + sCgs_1 + g_{o_2} + sCgd_2} \]
\[ + \frac{g_{o_2} + g_{o_2} + sCgd_1}{sCgd_1 + sCgs_1 + g_{o_2} + sCgd_2} \]
\[ + sCgs_2 - \frac{sCgd_2 - g_{o_2} - sCgd_1}{sCgd_1 + sCgs_1 + g_{o_2} + sCgd_2} \]

where \( g_m \) and \( g_o \) are the transconductance and output conductance of the transistor. The body effect of \( M_2 \), \( g_{mb_2} \), can be included in \( g_{m_2} \), or it can be eliminated using a triple well structure. \( C_{dd_3} \) and \( g_{o_3} \) are zero for an ideal current source and \( C_{ds_1} \) and \( C_{ds_2} \) are neglected since they are likely to be small when the transistors are saturated. From its approximations, \( Y_{in} \) becomes in forms \( Y_{in1} \) The equivalent circuit is shown in Figure 4.

Figure 4: Equivalent circuit of the active inductor.

With the parameters of the equivalent circuit approximately given by:
\[ C_p = C_{gs_2}, \quad R_p = \frac{1}{g_{m_2}} \]
\[ L_s = \frac{C_{gs_1}}{(g_{m_1})(g_{m_2})}, \quad R_s = \frac{g_{o_2}}{(g_{m_1})(g_{m_2})} \]

There is clear note that all these elements are frequency independent and tunable via specific parameters. So, the equivalent inductance \( L_s \) of the circuit is tunable via the transconductance \( g_{m_2} \) with respect to the biasing current \( I_{bias} \), while the shunt conductance \( C_p \) stands for the active inductor losses. \( C_p \) is a parasitic element and \( R_s \) is a series loss resistance.

Assuming that all transistor output impedances are infinite, the relations for the resonant frequency and quality factor are given respectively by:
\[ \omega_0 \equiv \frac{1}{\sqrt{L_s C_p}} \approx \frac{g_{m_1} g_{m_2}}{C_{gs_1} C_{gs_2}} \approx \sqrt{\omega_{t_1} \omega_{t_2}} \]
\[ Q \equiv \frac{R_p}{\sqrt{C_p}} \approx \sqrt{\frac{g_{o_2} g_{m_1}}{g_{m_2} C_{gs_1}}} \approx \sqrt{\omega_{t_1} / \omega_{t_2}} \quad \omega_2 \equiv 0. \]

Where \( \omega_{t_1} \) and \( \omega_{t_2} \) are the transit (unity - gain) frequencies of \( M_1 \) and \( M_2 \) respectively. The impedance, \( Z_{in} \), of a component or a circuit is defined as, \( Z_{in} = R + jX \) where \( R \) is the resistance, \( j \) is the imaginary unit, and \( X \) is the reactance and simulation the \( Z_{in} \) as shown in figure 5. The circuit has the response of a second order bandpass filter hence, the resonant frequency occurs as the imaginary portion of the impedance is equal to zero. The circuit behaves like an inductor below the resonant frequency, and a capacitor above the resonant frequency. The real portion of the impedance peaks at the resonant frequency; hence, this is a parallel RLC circuit. This active inductor is simulated by spectre in CADENCE. Their simulation results as shown in figs. 5 with two transistor \( M_2, M_1 \) are size \((w/l), \) respectively \((350n/2u, \text{finger}=1), \)
\((235n/2u, \text{finger}=43), \) bias current \((I=130uA, VDD=3.3V \text{ and } Vb=vdc=1.5V)\). \((\Delta \omega) = \omega_2 - \omega_1 \) then changing \((\Delta \omega) \) the changes \( Q \) but quality factor \( Q \) is equal to 0.6 from simulation as shown in figure 5. Thus for the next study based on the quality factor constant active inductor.
As proposed in [5] active inductor whose quality factor is constant irrespective of signal swing is shown in Fig.6. The transistors M2-M3, M4-M5, and M6-M7 forms a simple current mirrors having current gain B1, B2 and B3 respectively. Transistors M1 & M2 constitutes WU’s active inductor. I2 is set at maximum input current swing \( I_{in,\text{max}} \) and \( I_{d2}=I_{in,\text{max}} \) with \( I_{d2} \) is the drain current of the transistors M2. Thus in presence largely varying input current quality factor of active inductor will remain approximately unchanged.

From the large-signal analysis, it is shown that the self-resonant frequency and quality factor all vary with signal swing. If the mean quality factor is defined to be [6]:

\[
Q_m(\omega_0) = \frac{1}{\int_{I_{\text{min}}}^{I_{\text{max}}}} Q(\omega_0, I) dI.
\]

Figure 5: Simulation the Zin.

III. CONSTANT Q ACTIVE INDUCTOR

Figure 6: Constant Q active inductor
Where \( I_{\text{max}} \) and \( I_{\text{min}} \) are the maximum and minimum of the reuse-current bias, respectively. Fig 7 shows small signal analysis of active inductor shown in Fig.6. In order to simplify the small signal analysis, the bulk and corresponding source leads are connected and All transistors in saturation. CLi capacity represent an approximation of a more complex capacitance structure in four terminal transistor model with \( i=(1,2,3,4) \) the number of input capacities and outputs by the theorem method miller. The rest of the parameters are typical MOS components: output conductances \( g_{dsi} \) and transconductances \( g_{mi} \) with \( i \) the numbers of the transistors. All transistors are biased using currents sources (I1, I2) but used the couples of the mirror current ((M5, M4), (M7, M6), (M3, M2)) in use to construct a constant quality factor. From the small signal analysis of a transistor in high frequency we had to replace it in each transistor of active inductor to obtain the overall schema equivalent ultimately in the theorem KCL on global nodes of figure then to calculate input admittance \( Y_{\text{in}} \) in the following form:

\[
Y_{\text{in}} = \frac{g_{ds1} - g_{ds1}}{\left( g_{ds1} + g_{ds10} + S_{cl2} \right)} + \left[ (S_{cl1} + g_{ds7} + g_{ds2}) \right] + \left\{ (g_{m7} \ast g_{m5}) \left( (g_{m3} \ast (g_{m1} + g_{ds1})) \right) + \left( (S_{cl4} + g_{ds6} + g_{ds5} + g_{m6}) \left( (S_{cl3} + g_{ds20} + g_{ds4} + g_{ds3} + g_{m4}) \ast (g_{ds1} + g_{ds10} + S_{cl2}) \right) \right) \right\} \]

We can write input admittance \( Y_{\text{in}} \) in form:

\[
Y_{\text{in}} = B + A \quad \text{with}
\]

\[
A = \left[ \frac{(g_{m7} \ast g_{m5}) \left( (g_{m3} \ast (g_{m1} + g_{ds1})) \right)}{(g_{ds1} + g_{ds10} + S_{cl2})} \right] + \left( \frac{g_{m2} \left( g_{m1} + g_{ds1} \right) \left( (S_{cl4} + g_{ds6} + g_{ds5} + g_{m6}) \left( (S_{cl3} + g_{ds20} + g_{ds4} + g_{ds3} + g_{m4}) \ast (g_{ds1} + g_{ds10} + S_{cl2}) \right) \right)}{(g_{ds1} + g_{ds10} + S_{cl2})} \right]
\]

and

\[
B = \frac{\left( g_{ds1} \ast g_{ds10} + S_{cl2} \right) - g_{ds1} \left( g_{m1} + g_{ds1} \right) + \left( g_{ds1} \ast g_{ds10} + S_{cl2} \right) \ast g_{m1} + \left( S_{cl1} + g_{ds7} + g_{ds2} \right) \ast \left( g_{ds1} + g_{ds10} + S_{cl2} \right)}{(g_{ds1} + g_{ds10} + S_{cl2})}
\]

The input impedance of an active inductor is given as:

\[
Z_{\text{in}} = \frac{1}{Y_{\text{in}}}
\]

By the following assumptions we can approximate the input admittance:
*For all the transistors in strong inversion, by applying such transistors in saturation regime with $v_{ds} \geq v_{dsat}$, $g_m \geq 10g_{ds}$, and $C_{gs} \geq (c_{gd}, c_{ds})$ with $v_{dsat} = (v_{gs} - v_{th0})$.

*Assumes parasitic capacities at high frequency are equal for all pairs of current mirrors where $c_{gs3} = c_{gs2}, c_{gs5} = c_{gs4}, c_{gs7} = c_{gs6}$.

* operating frequency is much lower than transistor’s cutoff frequency ($g_m \gg |s_{cgs}|$).

*For $g_{ds} \gg g_{ds10}$, $g_{m6} = g_{m7}, g_{m5} = g_{m4}$ and $g_{m2} = g_{m3}$ then

$$A \equiv \frac{\left(\frac{g_{m6} \cdot g_{m4}}{\left(\left(g_{m2} \cdot g_{m1}\right) \cdot g_{m6} \cdot \left(g_{m2} + g_{m1}\right) \cdot \left(g_{m4} \cdot g_{m1}\right) \cdot \left(g_{m2} + g_{m1}\right) \cdot \left(g_{m4} \cdot g_{m1}\right)\right)}\right)}}{\left(\left(g_{m1} \cdot g_{m6} \cdot g_{m4}\right) \cdot \left(g_{m2} \cdot g_{m1}\right) \cdot \left(g_{m4} \cdot g_{m1}\right)\right)}$$

$$B \equiv \frac{\left(\frac{g_{m1} \cdot g_{m6} \cdot g_{m4}\right)}}{\left(\left(g_{m1} \cdot g_{m6} \cdot g_{m4}\right) \cdot \left(g_{m2} \cdot g_{m1}\right) \cdot \left(g_{m4} \cdot g_{m1}\right)\right)}$$

$$Z_{in} = \frac{\left(s^2 \cdot c_{gs1} \cdot c_{gs2} + s \cdot (g_{m1} + 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds} \cdot c_{gs1}) + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}) + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}}{s^2 + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1})}$$

By factorization of the terms of impedance dominators we obtain $Z_{in}$:

$$Z_{in} = \frac{\left(s^2 \cdot c_{gs1} \cdot c_{gs2} + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}) + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}}{s^2 + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1})}$$

For negligible conductance $g_{ds1}$ the input impedance can be written as:

$$Z_{in} \cong \frac{\left(s^2 \cdot c_{gs1} \cdot c_{gs2} + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}) + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}}{s^2 + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1})}$$

$$Z_{in} \cong \frac{\left(s^2 \cdot c_{gs1} \cdot c_{gs2} + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}) + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}}{s^2 + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1})}$$

$$Z_{in} \cong \frac{\left(s^2 \cdot c_{gs1} \cdot c_{gs2} + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}) + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1}}{s^2 + s \cdot (g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2} + g_{ds10} \cdot g_{m1} + (g_{ds} + g_{ds2}) \cdot c_{gs2} + g_{ds1} \cdot c_{gs1})}$$

with $\omega_0^2 = \frac{\left(2g_{m2} \cdot (g_{m1}) + g_{ds10} \cdot g_{m1}\right)}{\left(c_{gs1} \cdot c_{gs2}\right)}$, $\omega_0 = \frac{g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2}}{\left(c_{gs1} \cdot c_{gs2}\right)}$,

$$Q(\omega_0) = \frac{\left(c_{gs1} \cdot c_{gs2}\right)}{\left(g_{m1} \cdot 2c_{gs2} + 2 \cdot (g_{ds1} + g_{ds2}) \cdot c_{gs2}\right)} \cdot \sqrt{\left(2g_{m2} \cdot (g_{m1}) + g_{ds10} \cdot g_{m1}\right) / \left((c_{gs1} \cdot c_{gs2})\right)}$$

From the above equations for the transconductances $M1$ and $M2$ of the transistors are constant, it is necessary to give a constant quality factor ($Q$) whatever the values of the capacities parasitic.

Using a library for tsmc13rf 0.13um technology in cadence, create the schematic shown in figure 6. This is an $Q$-CONSTANT QUALITY FACTOR ACTIVE INDUCTOR based on source current controlled (I2) which
is designed to operate Q constant. The Transistor parameters and component values are shown in tables 1-2. The active inductor is formed by transistors M1, M2, M3, M4, M5, M6 and M7.

Table 1. Transistor parameters

<table>
<thead>
<tr>
<th>Transistors</th>
<th>W [um]</th>
<th>L [um]</th>
<th>Multiplicity</th>
</tr>
</thead>
<tbody>
<tr>
<td>M1(Nmos)</td>
<td>2u</td>
<td>350n</td>
<td>1</td>
</tr>
<tr>
<td>M2(Nmos)</td>
<td>100u</td>
<td>350n</td>
<td>1</td>
</tr>
<tr>
<td>M3(Nmos)</td>
<td>2u</td>
<td>350n</td>
<td>1</td>
</tr>
<tr>
<td>M4(Nmos)</td>
<td>2u</td>
<td>350n</td>
<td>1</td>
</tr>
<tr>
<td>M5(Nmos)</td>
<td>1.57u</td>
<td>350n</td>
<td>1</td>
</tr>
<tr>
<td>M6(Pmos)</td>
<td>2u</td>
<td>350n</td>
<td>1</td>
</tr>
<tr>
<td>M7(Pmos)</td>
<td>2u</td>
<td>350n</td>
<td>1</td>
</tr>
</tbody>
</table>

Table 2. Component values

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>VDD</td>
<td>3.3V</td>
</tr>
<tr>
<td>I1</td>
<td>130uA</td>
</tr>
<tr>
<td>I2</td>
<td>215uA-535uA</td>
</tr>
<tr>
<td>Vb</td>
<td>1.5V</td>
</tr>
</tbody>
</table>

To observe the behavior of the active inductor with respect to the tuning current I2, a parametric simulation can be used. Set the control current I2 as a variable and sweep in the range shown in figure 8 and figure 9. Keep one, the analysis ac simulations active. The simulation results (input waveform impedance and quality factor Q) will show multiple curves. To obtain parametric plot like the one shown in figure 9, use the calculator and select the curves with the button ‘wave’.

The proposed design is also aimed to illustrate another important advantage, which must be pointed out, that is the quality factor Q of the circuit whose expression is given by:

$$Q = \frac{\text{IMA} [\text{Zin}]}{\text{REAL} [\text{Zin}]}$$

Where IMA [Zin] and REAL [Zin] are the imaginary and the real parts of the input impedance, respectively. They are functions of the transconductance $g_m$ of the circuit. As a result, the Q factor is tuned by means of the current Ibias leading to a change in the equivalent inductance L and the resistances.

As seen in the figures (8 and 9) shows that the quality factor Q and resonances frequency $f_0$ are constant, because we say a not tuning frequency resonance and factor quality by the variation of broad signal current I2 then these results shows that the independent quality factor of resonance frequency.
IV. RESISTIVE COMPENSATION TECHNIQUE

Resistive compensation technique is used to enhance the Bandwidth of active inductor. Usually, a method of improving frequency performance is to increase the DC biasing current. However this will trade off power consumption. Fig. 10 shows modified active inductor circuit. A compensation resistor R is connected between the gates of transistors M2 and M3. A resistor is introduced in basic WU’s active inductor circuit which is constituted by M1 and M2. Fig. 11 shows small signal equivalent in high frequency of modified circuit of active inductor.

From an equivalent small signal of modified active inductor, the admittance can be determined in the following form $Y_{in}$:

$$Y_{in} = \frac{I_{in}}{V_{in}} = g_m1 + g_{ds1} - g_{ds1} \frac{a1}{b1} + g_m2 \frac{a2}{b2} A + g_{ds2} + sC_{55} + g_{ds7}.$$
\[ a1 = (sc44+gds) (gm1+gds1), \]
\[ b1 = (gds10+sc33+gds1)* (sc44+gds)+(sc44*gd1), \]
\[ a11 = (gm1 + gds1) * (sc44 + gds), \]
\[ b11 = (gd10 + sc33 + gds1) * (sc44 + gds) + (sc44 + gds). \]
\[ A=(gd10+sc33)*(sc44+gds)/(sc44*gd1). \]

For the Transistors in regime the saturation we assume that \( gm > 10 \text{gds} \) and \( cgs > (\text{gds}, \text{egd}) \) then The parameters included in Yin can be approximated to the following new parameters: \( c44 \equiv cgs2, C33 \equiv cgs3, C55 \equiv cgs1. \)

\[ b1 \equiv s^2 * cgs2 \]
\[ *cgs3+s*(gds10*cgs2+gds1*cgs3+gdse+cgs2*gds)+((gd10+cgs11)*gd1) \]
\[ A \equiv (s^2) \equiv (cgs3)+(cgs2)+s^2*(gds10+cgs2+gds11)*gds/((gd10+cgs11)*gd1) \]

Yin\( \equiv gm1-\text{gd1}a11/b11+gm2a11/b11A+scgs1 \]
\[ b1 \equiv [s^2 * cgs2 * cgs3 * s + (gd10 * cgs2 + cgs1 * cgs2 + cgs3 * cgs2 + cgs2 * cgs + cgs2 + gds) + (gd10 + cgs1)] * (gd1) \]
\[ \equiv a11 * s^2 + a12 * s + a13 \] with \( a11 = \]
\[ cgs2 + cgs3, a12 = (gd10 + cgs2 + cgs1 + cgs2 + cgs3 + cgs2 + gds), a13 = (gd10 + cgs11) * (gd1) \]
\[ \equiv a1 \equiv [(scgs2 + cgs1) (gm1)] \equiv a1 * s + a2 \] with \( a1 = (cgs2) * (gm1), a2=(cgs1) * (gm1). \)
\[ a1/b1 \equiv (a1 * s + a2)/\alpha s^2 + a12 * s + a13. \]
\[ \equiv \alpha (gm1) * (scgs2 + cgs1) \equiv \beta1 * s + \beta2 \] with \( \beta1 = \alpha (gm1) * cgs2, \beta2 = \alpha (gm1) * cgs1. \)
\[ b11 = s^2 cgs3*cgs2+s=((gd10+gd11)*cgs2+cgs3*gdse+gdse)/(gd10+cgs1) \]
\[ \equiv s^2 * b11 + s * b12 + b13 \] with \( b11 = \]
\[ cgs3*cgs2, b12 = (gd10 + cgs11) * cgs2 + cgs3 * cgs1 \]
\[ b13 = (gd10 + cgs11) * (gd1). \]
\[ a11/b11 \equiv (\beta1 * s + \beta2)/(s^2 + \beta11 + s + \beta12 + \beta13). \]
\[ A \equiv (s^2) \equiv (cgs3) \equiv (cgs2+s)*(gds*e+cgs2+gds*gdse)/(sgs2+gdse). \]
\[ \equiv (s^2) \gamma 1 + s \gamma 2 + \gamma 3 \] / \( (s^2) \gamma 1, \) with \( \gamma 1=(cgs3+cgs2), \gamma 2=(gds*e+cgs3+cgs2+gdse), \gamma 3=gds*gdse, \]
\[ \gamma 11 = cgs2*cgs2. \]

then Yin\( \equiv \gamma p-gds1*a11/b11+gm2*(a11/b11)*A. \]

with \( (\gamma p = gm1 + gds1 + gds2 + c55 + gd7) \]
\[ \equiv \delta1 + s + \delta2 = d1s1 * (a1 * s + a2)/(a11 * s^2 + a12 * s + a13) \] + \[ [gm2*(\beta1 * s + \beta2)/(s^2 + b11 + s + b12 + b13)] * (s^2 + b1 + s * b12 + b1) \]
\[ \equiv (s^2) \gamma 1 + s \gamma 2 + \gamma 3 / (s^2) \gamma 1, \) with \( \delta1 = gm1 + gds1 + gds2 + c55, \delta2 = c55. \]

We pose \( \delta = a11 * s^2 + a12 * s + a13) * (s^2 + \beta11 + s + \beta12 + \beta13) \)

\( \equiv (s^2) \gamma 1, \) then

\[ \text{Yin}=1/Zin= \]
\[ (s^2) \gamma 1 \] \( (s^2) \gamma 2 \]
\[ \gamma 11, \] \( \gamma 11, \)
\[ \gamma 3, \] \( \gamma 3, \)
\[ (s^2) \gamma 1, \] \( (s^2) \gamma 1, \)
\[ \gamma 11. \] \( \gamma 11. \)

is admittance with negative resistance \( R=1/gds. \)

From an approximate input impedance calculation with a negative resistance, equivalent of the transfer function of a continuous and invariant linear dynamic system is stable if and only if the poles of its transfer function have strictly negative real parts. Equation Zin shows input impedance after the introduction of compensation resistor with new poles and zeros. The resistor delays the response but zero created by resistor also cancels the delay at the same time. The addition of zero makes system faster and more oscillatory as the zero moves in negative axis towards the origin. A maximum bandwidth is obtained when zero cancels low frequency pole as shown in the figures .12 and .13.

![Image](http://www.ijser.org)}
The lower the value of the Q factor the wider is the bandwidth of the filter and consequently the higher the Q factor the narrower and more “selective” is the filter.

Note that Q factor is a ratio and has no units.

V. BAND PASS FILTER DESIGN

A simple way of implementing a band pass filter is shown in Fig.14. Transconductance of Min converts input voltage to current. This current is applied to active inductor and output is taken at transistor Mout, which is an output buffer used to drive resistive loads.

The bandwidth width deviates from the central frequency f0 and the constant quality factor. The higher the resonance frequency is the greater large bandwidth, The filter is said to be selective.
when the bandwidth is small around the central frequency $f_0$ as shown in the figures (15, 16).

Simulation results of bandpass filter at 0.13 µm technology are better than the 0.5 µm technology.

With resistive compensation technique, frequency is 1.392 GHz with bandwidth 2GHz.

Table 3. Comparison of bandwidth extension and frequency.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Without extension technique at 0.5 µm [14]</th>
<th>Resistive compensation at 0.5µm [14]</th>
<th>Without extension technique at 0.13 µm in work</th>
<th>Resistive compensation at 0.13µm in work</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>122.46 MHz</td>
<td>148.59 MHz</td>
<td>1.387 GHz</td>
<td>1.392GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>16 MHz</td>
<td>60 MHz</td>
<td>2GHz</td>
<td>2.175GH z</td>
</tr>
</tbody>
</table>

VI. CONCLUSION

Bandwidth of band pass filter is enhanced by resistive compensation technique. The resonant circuit of band pass filter based on constant Q active inductor has been presented. The resonant frequency of modified circuit observed is 1.392GHz. Bandwidth is enhanced from its original value.

REFERENCES