

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, 0.13 μm technologie, band pass filter.

I. INTRODUCTION:

Due to the scaling of the transistor, the cut-off frequency of the 0.13- μm 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 SiO_2 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 f_0 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 transconductor 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 [2] $C / (G^*G)$.

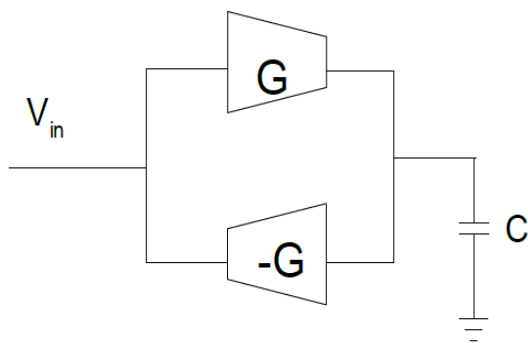


Figure.1: Schematic diagram of a gyrator based active inductor.

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 M_1 generates a current $gm_2 v_i$, that charges C_{gs1} through the feedback path. The voltage v_x then provides for the current $gm_1 v_{gs1}$ ($V_{gs1}=V_x$) and generates an inductive loading effect.

Single ended active inductor circuit illustrating both the model (figure 1) and the real structure (figure 2). C_{gs} 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.

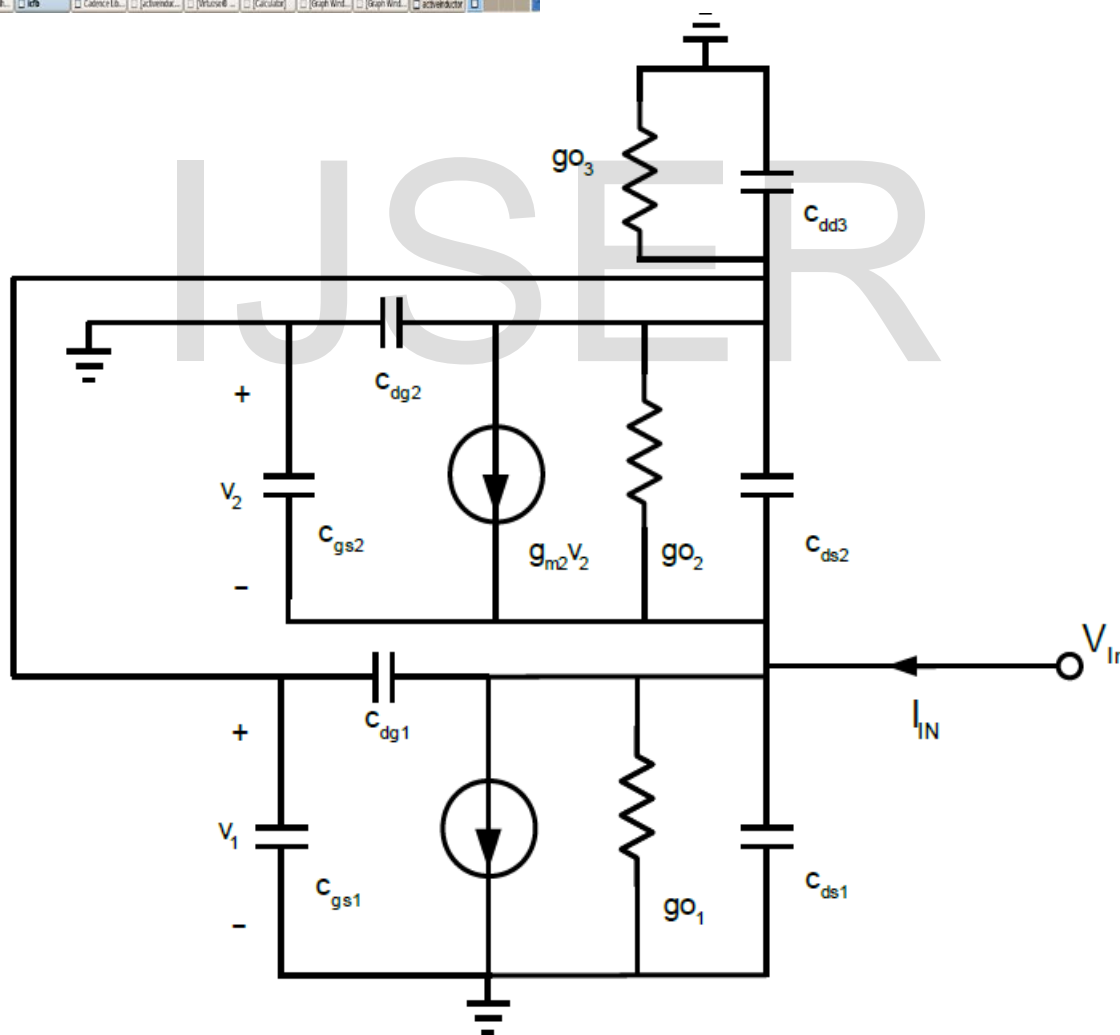


Figure .3: Small-Signal Model of an Active Inductor in high frequency.

Assuming an ideal current source, the input admittance of the active inductor can be written

$$Y_{in} = g_{o_1} + \frac{(g_{m_1})(g_{m_2} + g_{o_2} + sCgd_1)}{sCgd_1 + sCgs_1 + g_{o_2} + sCgd_2} + g_{m_2} + g_{o_2} + sCgd_1 + sCgs_2 - \frac{(g_{o_2})(g_{m_2} + g_{o_2} + sCgd_1)}{sCgd_1 + sCgs_1 + g_{o_2} + sCgd_2}$$

where gm and go are the transconductance and output conductance of the transistor. The body effect of M2, gmb2, can be included in gm2, or it can be eliminated using a triple well structure. Cdd3 and go3 are zero for an ideal current source and Cds1 and Cds2 are neglected since they are likely to be small when the transistors are saturated. From its approximations, Yin becomes in forms Yin1 The equivalent circuit is shown in Figure 4.

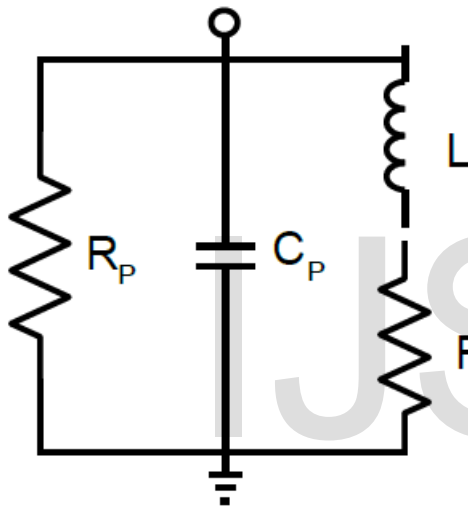


Figure.4:Equivalent circuit of the active inductor.

$$Y_{in1} = \frac{(g_{m_1})(g_{m_2})}{sCgs_1 + g_{o_2}} + g_{m_2} + sCgs_2 - \frac{(g_{o_2})(g_{m_2})}{sCgs_1 + g_{o_2}}$$

$$Y_{in1} \cong \frac{(g_{m_1})(g_{m_2})}{sCgs_1 + g_{o_2}} + g_{m_2} + sCgs_2$$

The resonance frequency f0 is measured when maximum input impedance Zin as shown in the figure.5 thus this resonance frequency f0 is equal to its value 5.495GHz, but factor quality Q is measured by the following method $Q = \omega_0 / (\Delta\omega)$ with Bandwidth $\Delta\omega$ is equal to

With the parameters of the equivalent circuit approximately given by: $C_p = Cgs_2$, $R_p = \frac{1}{gm_2}$

$$L_s = \frac{Cgs_1}{(gm_1)(gm_2)}, R_s = \frac{g_{o_2}}{(gm_1)(gm_2)}$$

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 $gm_{2,1}$ 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:

$$\omega_0 \cong \frac{1}{\sqrt{LC_p}} \cong \sqrt{\frac{gm_1 gm_2}{Cgs_1 Cgs_2}} \cong \sqrt{\omega_{t1} \omega_{t2}}$$

$$Q \cong \frac{R_p}{\sqrt{L}} = \sqrt{\frac{Cgs_2 gm_1}{gm_2 Cgs_1}} \cong \sqrt{\omega_{t1} / \omega_{t2}}, \omega_z \cong 0.$$

Where ω_{t1} and ω_{t2} 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 M2, M1 are size (w/l), respectively (350n/2u, finger=1), (235n/2u, finger=43), bias current $I_1 = 130\mu A$, $V_{DD} = 3.3V$ and $V_b = v_{dc} = 1.5V$.

$(\Delta\omega) = \omega_2 - \omega_1$ 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.

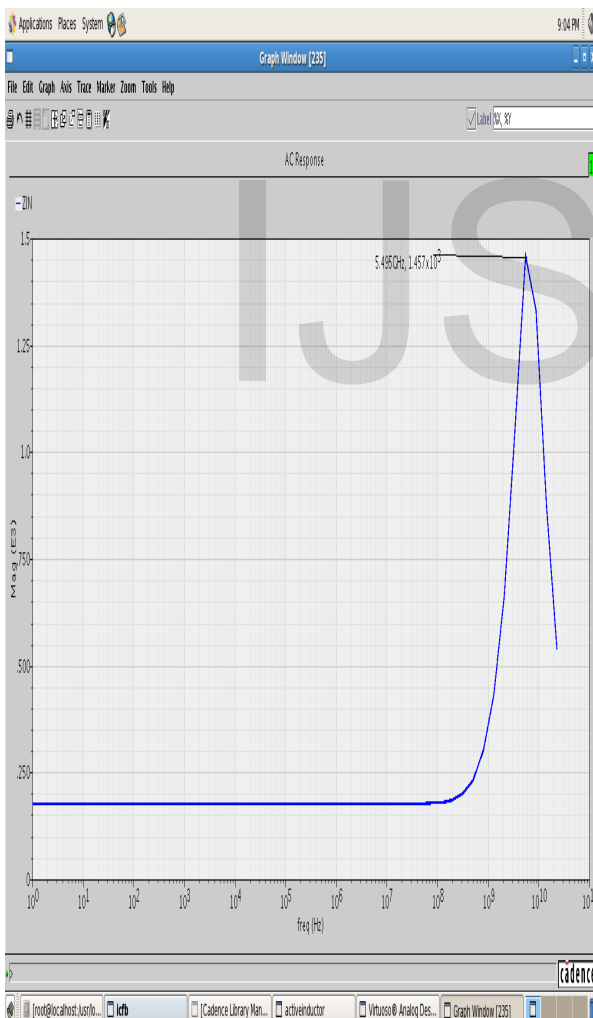
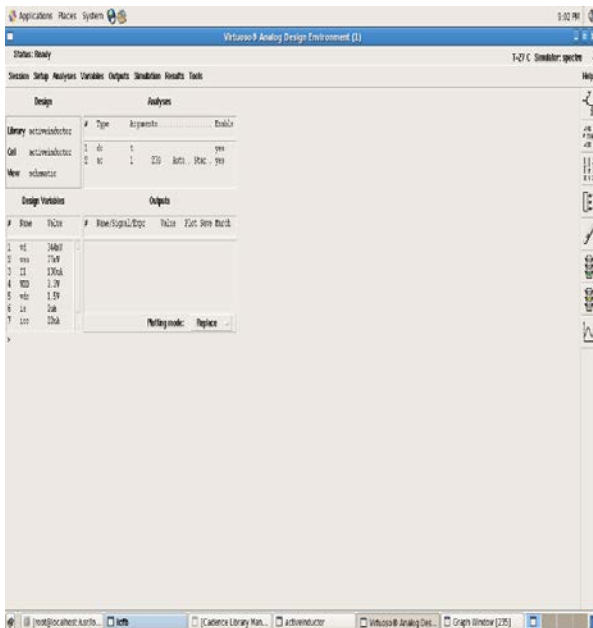


Figure .5: Simulation the Zin.

III. CONSTANT Q 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 $I_{in,max}$ and $I_{d2} = I_{in,max}$ 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

$$[6]: Q_m(\omega_0) = \frac{1}{I_{max} - I_{min}} \int_{I_{min}}^{I_{max}} Q(\omega_0, I) dI.$$

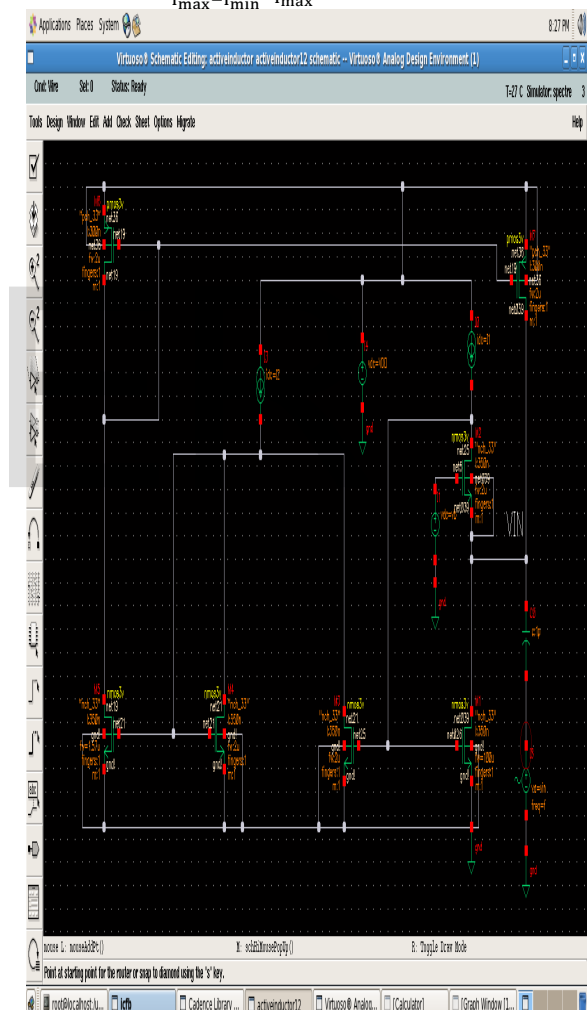


Figure.6: Constant Q active inductor

Where I_{max} and I_{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. CL_i 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 (I_1, I_2) 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_{in} in the

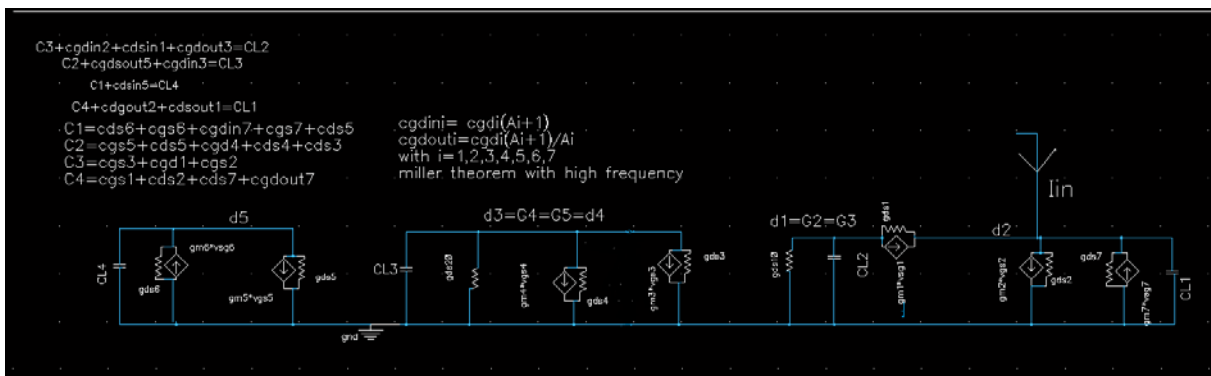


Fig .7: shows small signal analysis of active inductor.

following form :

$$Y_{in} = \left[g_{ds1} - g_{ds1} * \left(\frac{(g_{m1} + g_{ds1})}{(g_{ds1} + g_{ds10} + Scl2)} \right) + g_{m1} \right] + [(sCL1 + g_{ds7} + g_{ds2})] + \frac{((g_{m7} * g_{m5})((g_{m3} * (g_{m1} + g_{ds1}))))}{((Scl4 + g_{ds6} + g_{ds5} + g_{m6})((sCL3 + g_{ds20} + g_{ds4} + g_{ds3} + g_{m4}) * (g_{ds1} + g_{ds10} + Scl2)))} + g_{m2} * \frac{(g_{m1} + g_{ds1})}{(g_{ds1} + g_{ds10} + Scl2)}$$

We can write input admittance Y_{in} in form:

$Y_{in} = B + A$; with

$$A = \frac{((g_{m7} * g_{m5})((g_{m3} * (g_{m1} + g_{ds1}))) * (g_{ds1} + g_{ds10} + Scl2) + \dots)}{((Scl4 + g_{ds6} + g_{ds5} + g_{m6})((sCL3 + g_{ds20} + g_{ds4} + g_{ds3} + g_{m4}) * (g_{ds1} + g_{ds10} + Scl2))) * (g_{ds1} + g_{ds10} + Scl2)}$$

and

$$B = \frac{g_{ds1} * (g_{ds1} + g_{ds10} + Scl2) - g_{ds1} * (g_{m1} + g_{ds1}) + (g_{ds1} + g_{ds10} + Scl2) * g_{m1}}{(g_{ds1} + g_{ds10} + Scl2) * [(sCL1 + g_{ds7} + g_{ds2}) * (g_{ds1} + g_{ds10} + Scl2)]}$$

The input impedance of an active inductor is given as:

$$Z_{in} = 1/Y_{in}$$

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 ($g_m \gg |s c_{gs}|$).

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

$$A \cong \frac{[(g_{m6} * g_{m4}) * (g_{m2} * (g_{m1})) + g_{m2} * (g_{m1}) * (g_{m6}) * (g_{m4})]}{(g_{m6}) * (g_{m4}) * (g_{ds1} + 2s c_{gs2})}, \quad B \cong \frac{(g_{m1}) * (g_{ds10} + 2s c_{gs2}) + [(s c_{gs1} + g_{ds7} + g_{ds2}) * (g_{ds1} + 2s c_{gs2})]}{(g_{ds1} + 2s c_{gs2})}$$

$$Z_{in} = \frac{(g_{ds1} + 2s c_{gs2})}{[(g_{m1}) * (g_{ds10} + 2s c_{gs2}) + [(s c_{gs1} + g_{ds7} + g_{ds2}) * (g_{ds1} + 2s c_{gs2})] + 2g_{m2} * (g_{m1})]} = \frac{(g_{ds1} + 2s c_{gs2})}{don}$$

$$s^2 * c_{gs1} * c_{gs2} + s * (g_{m1} * 2c_{gs2} + 2 * (g_{ds7} + g_{ds2}) * c_{gs2} + g_{ds} * c_{gs1}) + g_{ds10} * g_{m1} + (g_{ds7} + g_{ds2}) * (g_{ds1}) + 2g_{m2} * (g_{m1}) = don$$

$$Z_{in} = \frac{(g_{ds1} + 2s c_{gs2})}{s^2 * c_{gs1} * c_{gs2} + s * (g_{m1} * 2c_{gs2} + 2 * (g_{ds7} + g_{ds2}) * c_{gs2} + g_{ds1} * c_{gs1}) + g_{ds10} * g_{m1} + (g_{ds7} + g_{ds2}) * (g_{ds1}) + 2g_{m2} * (g_{m1})}$$

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

$$Z_{in} = \frac{\frac{(g_{ds1} + 2s c_{gs2})}{c_{gs1} * c_{gs2}}}{s^2 + s * \left[\frac{g_{m1} * 2c_{gs2} + 2 * (g_{ds7} + g_{ds2}) * c_{gs2} + g_{ds1} * c_{gs1}}{(c_{gs1} * c_{gs2})} \right] + \left[\frac{(g_{ds7} + g_{ds2}) * (g_{ds1}) + 2g_{m2} * (g_{m1}) + g_{ds10} * g_{m1}}{(c_{gs1} * c_{gs2})} \right]}$$

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

$$Z_{in} \cong \frac{\frac{(2s c_{gs2})}{c_{gs1} * c_{gs2}}}{s^2 + s * \left[\frac{g_{m1} * 2c_{gs2} + 2 * (g_{ds7} + g_{ds2}) * c_{gs2}}{(c_{gs1} * c_{gs2})} \right] + \left[\frac{(2g_{m2} * (g_{m1}) + g_{ds10} * g_{m1})}{(c_{gs1} * c_{gs2})} \right]}$$

$$Z_{in} \cong \frac{\frac{(2s c_{gs2})}{c_{gs1} * c_{gs2}}}{s^2 + s * \left[\frac{\omega_0}{Q} \right] + [\omega_0^2]}$$

$$\text{with } \omega_0^2 = \frac{(2g_{m2} * (g_{m1}) + g_{ds10} * g_{m1})}{(c_{gs1} * c_{gs2})}, \quad \frac{\omega_0}{Q} = \frac{g_{m1} * 2c_{gs2} + 2 * (g_{ds7} + g_{ds2}) * c_{gs2}}{(c_{gs1} * c_{gs2})}$$

$$Q(\omega_0) = \frac{(c_{gs1} * c_{gs2})}{g_{m1} * 2c_{gs2} + 2 * (g_{ds7} + g_{ds2}) * c_{gs2}} * \sqrt{\frac{(2g_{m2} * (g_{m1}) + g_{ds10} * g_{m1})}{(c_{gs1} * c_{gs2})}}$$

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 (I_2) 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.

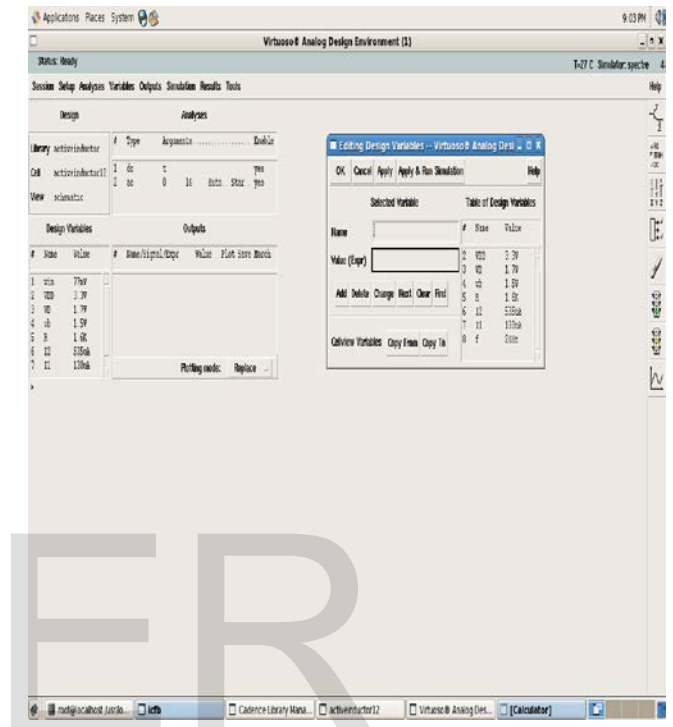
Table1. Transistor parameters

Transistors	W [um]	L [um]	Multiplicity
M1(Nmos)	2u	350n	1
M2(Nmos)	100u	350n	1
M3(Nmos)	2u	350n	1
M4(Nmos)	2u	350n	1
M5(Nmos)	1.57u	350n	1
M6(Pmos)	2u	350n	1
M7(Pmos)	2u	350n	1

Table 2. Component values

Component	Value
VDD	3.3V
I1	130uA
I2	215uA-535uA
Vb	1.5V

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:

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

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.

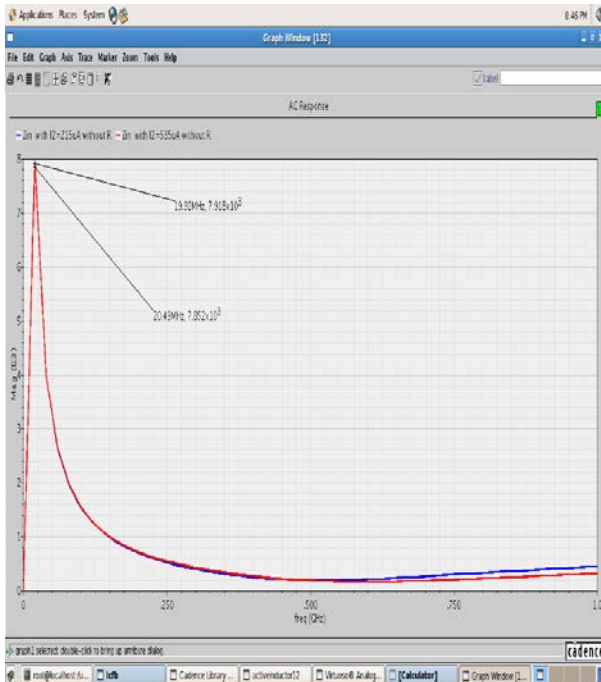


Figure.8: Simulation of input voltage Z_{in} as a function of frequency space f with several current values I_2 .

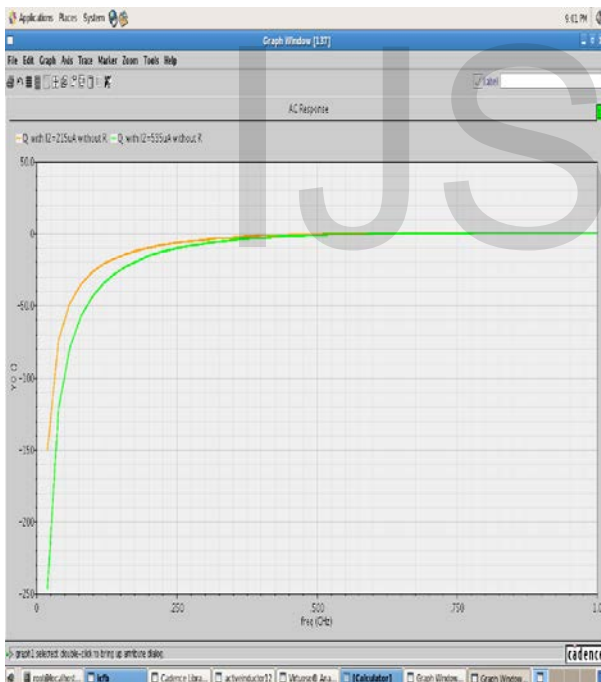


Figure.9: Simulation of Constant quality factor (Q) as a function of frequency space (f) with several current values I_2 .

IV.RESISTIVECOMPENSATION 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 M_2 and M_3 . A resistor is introduced in basic WU's active inductor circuit which is constituted by M_1 and M_2 . Fig. 11 shows small signal equivalent in high frequency of modified circuit of active inductor.

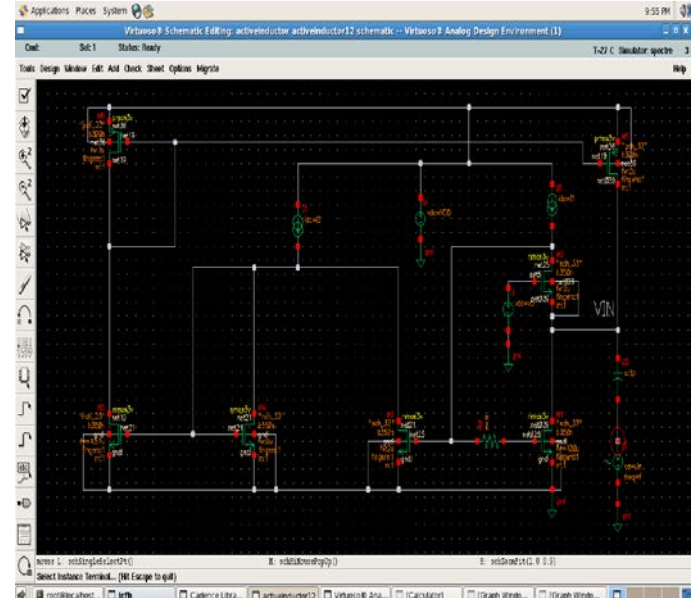


Figure.10.Modified active inductor with compensation resistor.

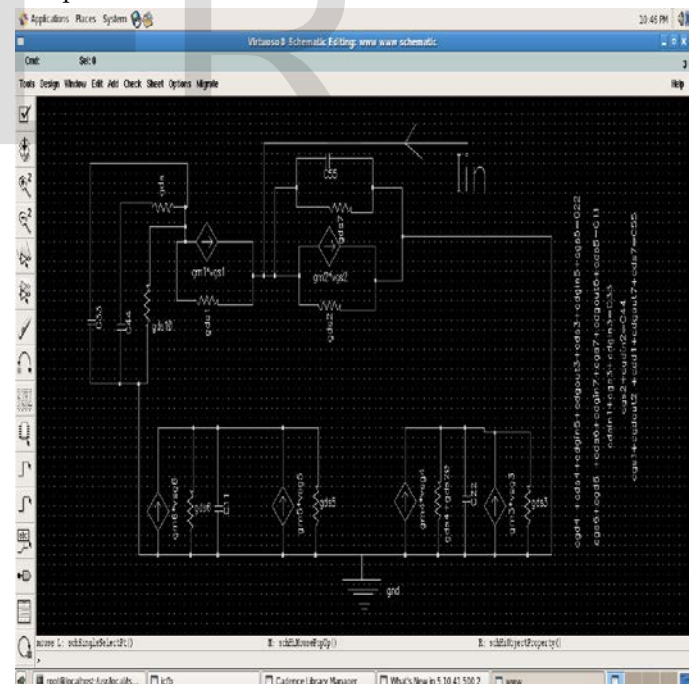


Fig.11. Small signal equivalent of modified active inductor.

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

$$Y_{in} = g_{m1} + g_{ds1} - g_{ds1} * a_1 / b_1 + g_{m2} * (a_{11} / b_{11}) * A + g_{ds2} + sC_{55} + g_{ds7}$$

$$\begin{aligned}
 a1 &= (sc44+gds) (gm1+gds1), \\
 b1 &= (gds10+sC33+gds1) * (sc44+gds)+(sc44*gds). \\
 a11 &= (gm1 + gds1) * (Sc44 + gds). \\
 b11 &= (gds10 + sC33 + gds1) * (sc44 + gds) + (sc44 * gds). \\
 A &= (gds10+sC33)*(sc44+gds)/ (sc44*gds). \\
 \text{For the Transistors in regime the saturation we assume that } gm > 10gds \text{ and } cgs > (cgs, cgd) \text{ then The parameters included in Yin can be approximated to the following new parameters:} \\
 c44 &\cong cgs2, C33 \cong cgs3, C55 \cong cgs1. \\
 b1 &\cong s^2 \quad cgs2 \\
 cgs3+s(gds10*cgs2+gds1*cgs2+cgs3*gds+cgs2*gds) &+ (gds10+gds1)*(gds). \\
 a1 &\cong (scgs2+gds) * (gm1). \\
 a11 &\cong (gm1) * (scgs2 + gds). \\
 b11 &\cong s^2cgs3*cgs2+s * ((gds10 + gds1) * cgs2+ cgs3 * gds)+ (gds10 + gds1) * (gds). \\
 A &\cong (s^2 *cgs3*(cgs2)+s*(gds*cgs3+cgs2*gds10)+gds*gds10)/ (scgs2*gds). \\
 Yin &\cong gm1- gds1*a1/b1+ gm2*a11/b11*A + scgs1 \\
 b1 &\cong [s^2 cgs2 * cgs3 + s * (gds10 * cgs2 + gds1 * cgs2 + cgs3 * gds + cgs2 * gds) + (gds10 + gds1) * (gds).] \cong \alpha11 * s^2 + \alpha12 * s + \alpha13 \text{ with } \alpha11 = cgs2 * cgs3, \alpha12 = (gds10 * cgs2 + gds1 * cgs2 + cgs3 * gds + cgs2 * gds), \alpha13 = (gds10 + gds1) * (gds) \\
 a1 &\cong [(scgs2 + gds) (gm1)] \cong \alpha1 * s + \alpha2 \text{ with } \alpha1 = (cgs2) * (gm1), \alpha2=(gds) * (gm1). \\
 a1/b1 &\cong (\alpha1 * s + \alpha2)/(\alpha11 * s^2 + \alpha12 * s + \alpha13). \\
 a11 &\cong (gm1) * (Scgs2 + gds) \cong \beta1 * s + \beta2 \text{ with } \beta1 = gm1 * cgs2, \beta2 = gm1 * gds. \\
 b11 &\cong s^2cgs3*cgs2+s * ((gds10 + gds1) * cgs2+ cgs3 * gds)+ (gds10 + gds1) * (gds) \cong s^2 * \beta11+s * \beta12+\beta13 \text{ with } \beta11 = cgs3*cgs2, \beta12 = ((gds10 + gds1) * cgs2+ cgs3 * gds), \beta13 = (gds10 + gds1) * (gds). \\
 a11/b11 &\cong (\beta1 * s + \beta2)/(s^2 * \beta11 + s * \beta12 + \beta13). \\
 A &\cong (s^2 *cgs3* cgs2+s*(gds*cgs3+cgs2*gds10)+(gds *gds10))/ (scgs2*gds). \\
 &\cong (s^2 \gamma1+ s*\gamma2 + \gamma3) / (s*\gamma11), \text{ with } \gamma1=(cgs3*cgs2), \gamma2=(gds *cgs3+cgs2*gds10) , \gamma3=gds*gds10, \gamma11 =cgs2*gds. \\
 \text{then} \\
 Yin &\cong yp- gds1*a1/b1+ gm2*(a11/b11)*A. \\
 \text{with } (yp = gm1 + gds1 + gds2 + sC55 + gds7) \\
 &\cong \delta1+ s* \delta2- gds1 *(\alpha1 * s + \alpha2)/(\alpha11 * s^2 + \alpha12 * s + \alpha13) + [gm2 * (\beta1 * s + \beta2)]/(s^2 * \beta11 + s * \beta12 + \beta13)]*[(s^2 \gamma1 + s * \gamma2 + \gamma3) / (s * \gamma11)]. \text{with } \delta1 = gm1 + gds1 + gds2 + gds7, \delta2 = C55. \\
 \text{We pose} \\
 don &= (\alpha11 * s^2 + \alpha12 * s + \alpha13)*(s^2 * \beta11 + s * \beta12 + \beta13) * (s * \gamma11) \text{ then}
 \end{aligned}$$

$$Yin=1/Zin= \frac{(\delta1+ s* \delta2)*don- [gds1*(\alpha1*s+ \alpha2)*(s^2*\beta11+s*\beta12+\beta13)*(s*\gamma11)]+ [gm2*(\beta1*s+ \beta2)*(\alpha11*s^2+\alpha12*s+\alpha13)]*(s^2 \gamma1+ s* \gamma2 + \gamma3)}{don}$$

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.

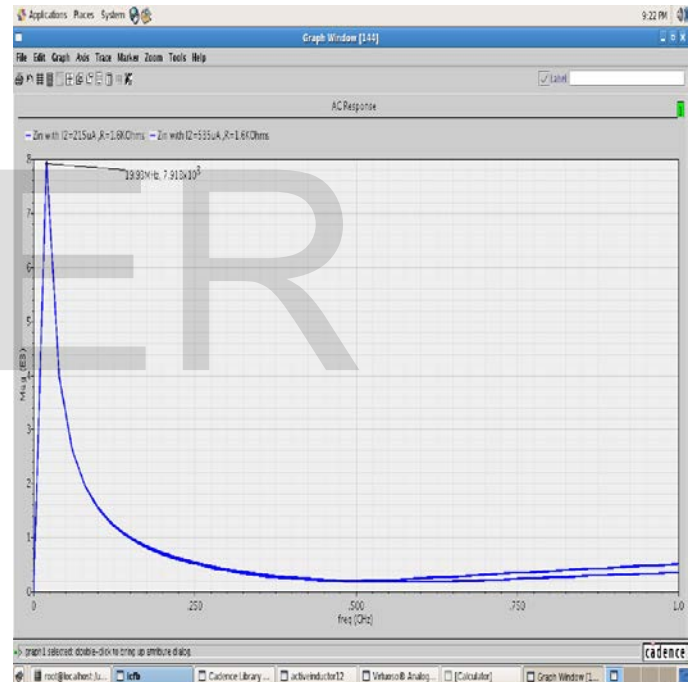


Figure.12: Simulation of input voltage Zin in variation of frequency f with several values of sources current I2 between 215uA and 535uA for resistance negative equal its value R=1.6Kohms.

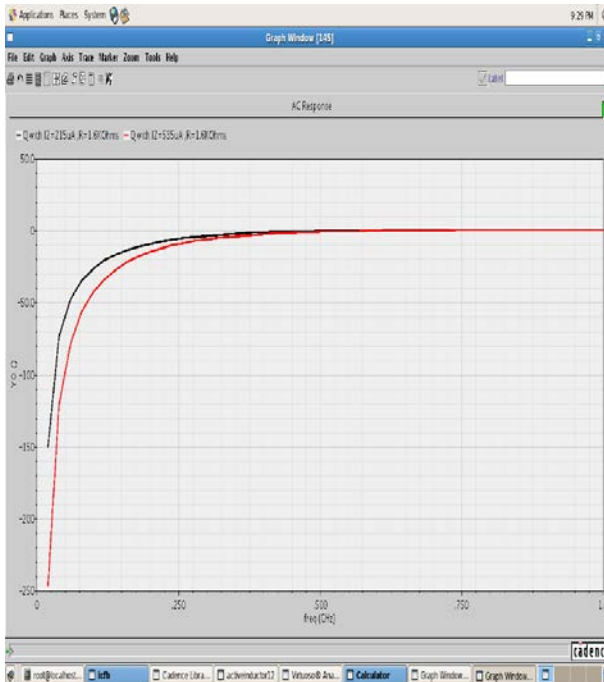


Figure.13: Simulation of input voltage Q in variation of frequency f with several values of sources current I_2 between 215uA and 535uA for resistance negative equal its value $R=1.6Kohms$. The lower the value of the Q factor the wider is the bandwidth of the filter and consequently the higher the Q 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 M_{in} converts input voltage to current. This current is applied to active inductor and output is taken at transistor M_{out} , which is an output buffer used to drive resistive loads.

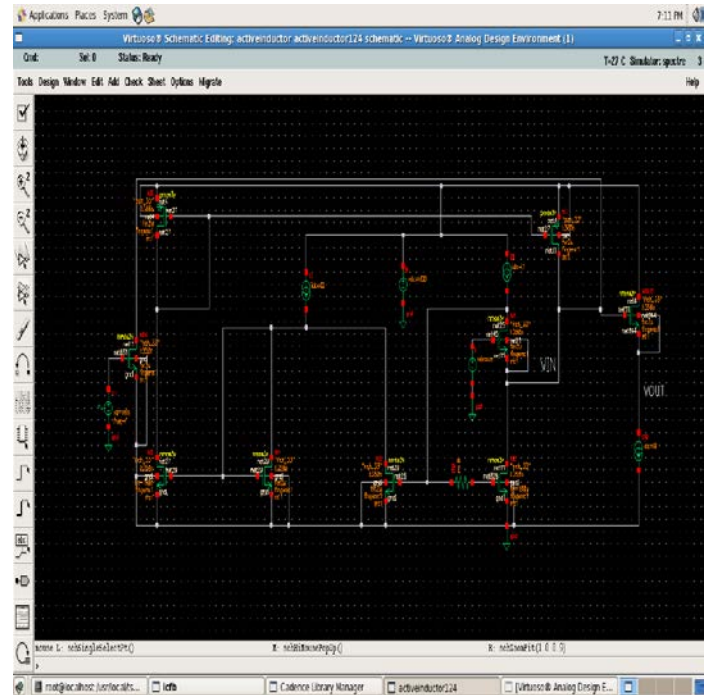
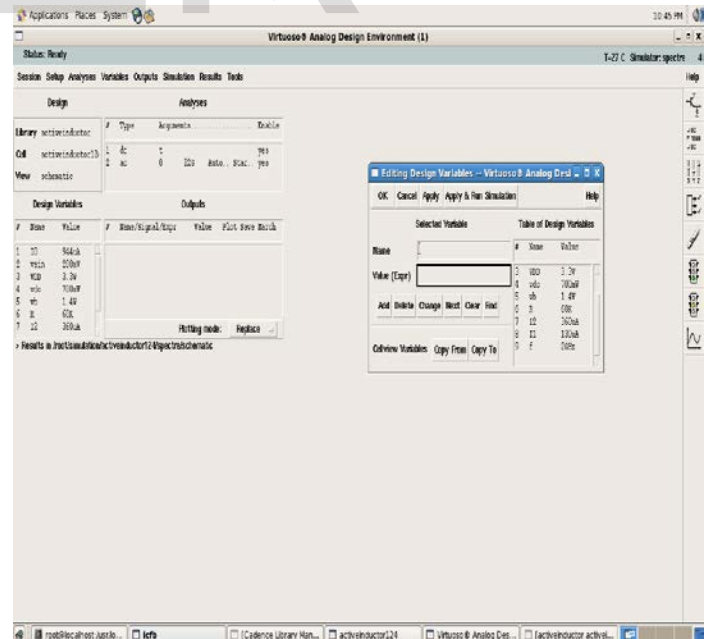


Figure.14: Implementing a band pass filter

This tunable active filters includes filters made up of tunable resonators whose losses are compensated. They do not show any gain. They can be made in the same way as a passive filter to which a loss compensation circuit is added in order to increase the quality factor, which leads to a decrease in linearity (due to the active part).



The bandwidth width deviates from the central frequency f_0 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).

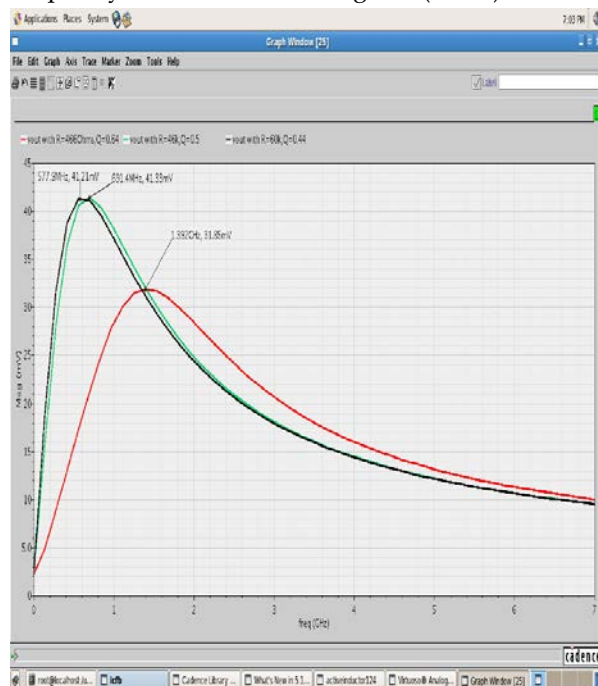


Figure.15: The bandwidth variation as a function of frequency f with several values of negative resistances R and bias current (I_2, I_1)

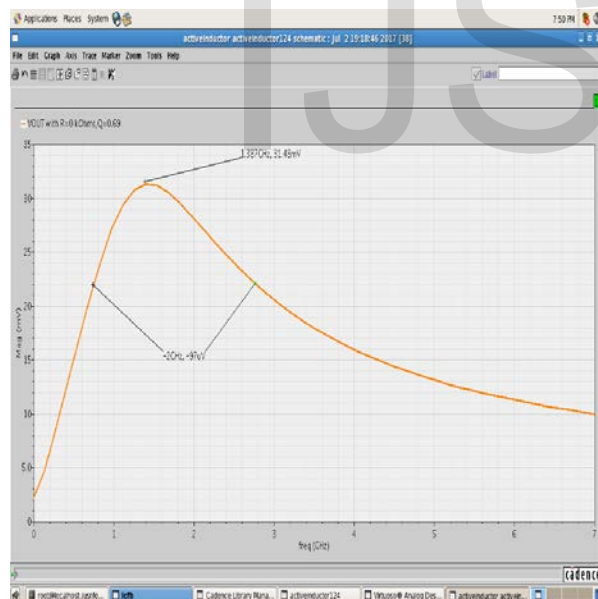


Figure.16: The bandwidth variation as a function of frequency f with constant values of current sources (I_2, I_1) without using the negative resistance. Compare the results of this work with the old results as shown at the table 3

Simulation results of bandpass filter at $0.13 \mu\text{m}$ technology are better than the $0.5 \mu\text{m}$ technology.

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

Table 3. Comparison of bandwidth extension and frequency.

Parameters	Without extension technique at $0.5 \mu\text{m}$ [14]	Resistive compensation at $0.5 \mu\text{m}$ [14]	Without extension technique at $0.13 \mu\text{m}$ in work	Resistive compensation at $0.13 \mu\text{m}$ in work
Frequency	122.462 MHz	148.59 MHz	1.387 GHz	1.392 GHz
Bandwidth	16 MHz	60 MHz	2GHz	2.175 GHz

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

- [1] Xiao, H., et al., (2004). "A radio-frequency CMOS active Inductor and its application in designing high-Q filters."
- [2] F. Yuan, CMOS Active Inductors and Transformers: Principle, Implementation, and Applications, Springer, 2008. 2004 IEEE International symposium on circuits and systems (ISCAS 2004), Vol.4, pp. 197-200.
- [3] Wu, Y. Ismail, M. Olsson, H. (2001). CMOS VHF/RF CCO Based on Active inductors. IEE Electronic Letters, 37(8), 472-473.
- [4] Y. Wu, et al "A novel cmos fully differential inductorless rf bandpass filter" ISCAS 2000 - IEEE International Symposium on Circuits and Systems, May 28-31, 2000
- [5] A. Tang. F. Yuan. E. Law (2009). "A new constant-Q CMOS Active inductor with applications to low-noise Oscillators." Analog integrated circuits and signal processing, Vol. 58, pp-77- 80.
- [6] A. Tang A F. Yuan A E. Law. A new constant-Q CMOS active inductor with applications to low-noise oscillators, Oct, 2008.
- [14] A. Neekhra and M. Gupta, "Bandwidth extension of a constant Q active," conference on computer and communication technology, pp. 311-313, 2010.