Performance Analysis of Active Inductor Based Tunable Band Pass Filter for Multiband RF Front end

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Abstract—**This paper presents an active inductor based low power, low noise tunable RF band pass filter suitable for multiband RF front end circuits. The designed active inductor and RF band pass filter are simulated in 180nm CMOS process using Synopsys tool. The resonators of the top coupled topology of the band pass filter are replaced with the tunable active inductors. The center frequency and the transfer gain can be varied through the controllable current source. The center frequency of the band pass filter is tuned from 2.4 GHz to 12.4 GHz with S21 values varies from 5 dB to 23 dB respectively. The noise figure ranges from 6.8 dB to 11.9dB for the entire tuning range of the band pass filter with less power consumption of 2.9mW.**

Keyword-Active inductor, Quality factor, Centre frequency tuning, PMOS cascode pair, tuning range, tunable RF band pass filter, Multiband RF front end

I. Introduction

The great demand for multiband wireless communication systems is increasing the requirement of integrated CMOS products for high performance RF front end circuits. Using CMOS process, the RF systems can be realized with low power consumption, high frequency range (GHz) and high reliability. The block diagram of the Mulitband RF front end is shown in Fig. 1. An analog RF band pass filter is an essential block of RF front end to select the band of interest of the received signal in the entire spectrum. The designing of the band pass filter for tuning to various center frequencies benefits hardware reuse for multi band applications. Low noise figure, low power consumption, tuning to the different center frequencies and better linearity are some of the major challenges in the design of multiband RF band pass filter.

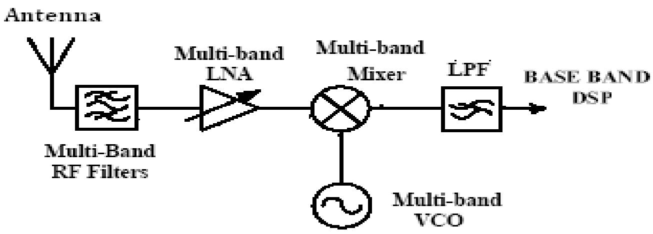

Fig. 1. Block Diagram of Multiband RF Front End

Most of the band pass filter topologies used in thin film technology is of the coupled resonator type. The advantage of coupled resonator filter is that they do not require a wide range of inductance values and are often realized using the same inductance for all resonators. The top coupled topology shown in Fig. 2 is more suitable to attenuate strong blocking signals in the cellular communications bands. However, in this type of filter the resonators operate in a single ended mode [1]. The resonators of the filter can be tuned to operate at the various centre frequencies, thus it can be suitable for RF multiband operation.

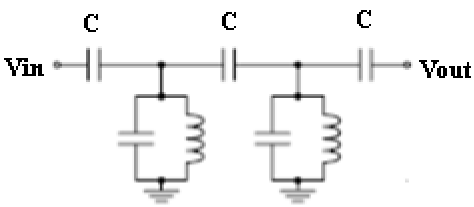

Fig. 2. Top Coupled topology of Band Pass Filter

The conventional method of tuning the center frequency of the resonator is done by using varactors [2] But The tuning range of the varactors are limited to 10 – 20%, that is not sufficient for mulitstandard applications [3]. Most of the RF band pass filters are implemented using on-chip spiral inductors [4]. But it is difficult to realize it for larger inductance values, high quality factor and smaller chip area [5]. Controllable or tunable active inductors provide many advantages in RF circuits. It reduces silicon footprint of a design by providing multiple inductor values from the same circuit element. And also it provides large inductance value with high resonance frequency, high quality factor, small chip area and wide range of tuning ability [6-7].

The task is to design a low power top coupled topology of active inductor based RF band pass filter tuning to the various center frequencies with better gain suitable for multiband RF front end.

II. DESIGN OF RF BAND PASS FILTER

A. Circuit Diagram Description

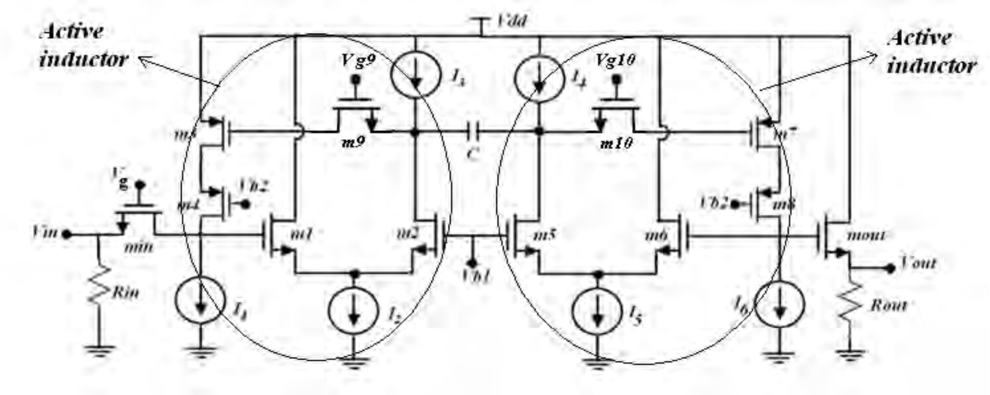

Fig. 3 Top Coupled Active inductor based RF Band Pass Filter

The second order RF band pass filter of Fig 3 is based on the active inductor topology. It consists two resonators (m1, m2, m3, m4, m9 and m5, m6, m7, m8, m10) which are made up of the active inductors, coupled through the capacitance C [7]. Min (W/L=1/0.18 in μm) is the common gate transistor, is used as the input stage for input matching. A source follower stage mout (W/L=1/0.18 in μm), is used as an output buffer stage for output matching and to reduce the loading effect. Rin and Rout are the input and output resistances which are selected to be 1kΩ. The current sources are realized using single MOS transistor current sources to make the active inductor compact and to operate at low voltage.

The circuit diagram of a single ended active inductor which is acting as a resonator in band pass filter configuration is shown in Fig. 4. It is based on Gyrator topology [6]. The transistor sizes (W/L in μm) of Fig. 4 are m1 (1 /0.18), m2 (1.8/0.18), m3 (2/0.18), m4 (3.3/0.18) and m5 (1/0.18). The gate bias voltages are kept as Vb1=0.2V, Vb2=137mV and Vgs=0.3V. It consists of differential pair m1 and m2 which represents the positive transconductor G_{m1} between the input (node 1) and the output (node 3). The cascode pair m3 and m4 represents the negative transconductor $-G_{m2}$ between the input (node 3) and the output (node 1). Thus the G_{m1} and $-G_{m2}$ forms the gyrator which converts the parasitic capacitance C_3 at node 3 to an equivalent inductance $L_{eq} = C_3/G_{m1}G_{m2}$. The passive equivalent circuit is as shown in Fig. 6.

The active inductor uses PMOS cascode structure as negative transconductor, reduces the output conductance to reduce series resistance of the inductor to compensate the inductor loss. Also, the cascode structure provides frequency range expansion by lowering the lower bound of the frequency range, thus increases the inductive bandwidth. The p-channel transistors are preferred for cascode structure as they have low noise and they can be placed in separate n-wells, thus eliminating the non- linear body effect [8]. Thereby, it enhances the quality factor of the active inductor. It also achieves high resonance frequency and better inductive bandwidth. To further improve the quality factor, series resistance R_s has to be reduced. This can be done by adding the transistor m5 between the positive transconductor and the negative transconductor of the active inductor.

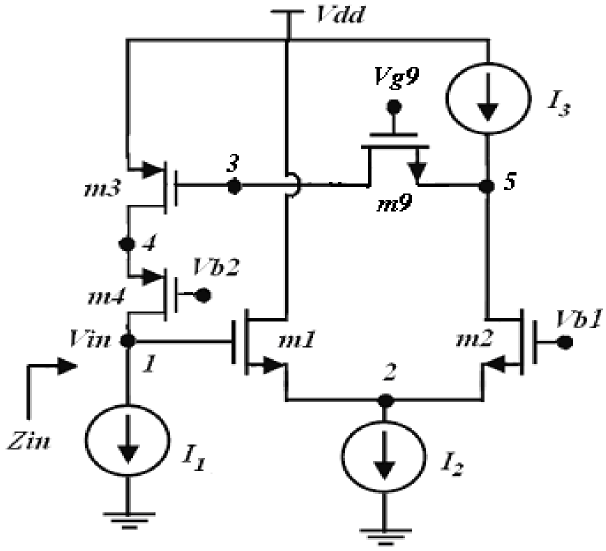

Fig. 4. Circuit diagram of the Active Inductor

The small signal equivalent circuit of the active inductor is shown in Fig. 5

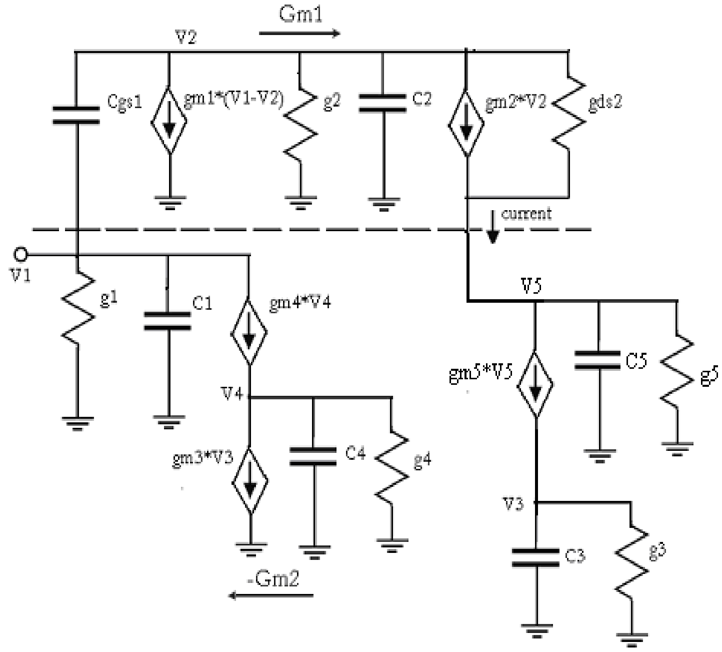

Fig. 5. Small signal equivalent circuit of the active inductor

The equivalent input impedance Z_{in}, of the active inductor can be obtained from the small signal analysis circuit shown in Fig. 5. gm_{1-5} are the transconductances of M_{1-5}, C_{1-5}, g_{1-5} are the total parasitic capacitances and conductance at nodes 1-5.V_{1-5} are the node voltages of nodes 1-5.

$$Y_{in}(s) = g_1 + sC_1 + \frac{gm_5 gm_4 gm_3 gm_2 gm_1}{(G+sC_2)(g_3+sC_3)(gm_4+g_4+sC_4)(gm_5+g_5+sC_5)} \quad (1)$$

where $G = gm_1 + gm_2 + g_2$.

The equation (1) is simplified and using the equivalence $s^3 = -\omega^2 s$ [4], it is given as,

$$Z_{in}(s) \approx \frac{\frac{sg_4g_5}{C_1} + \frac{g_3\,g_4g_5}{C_1C_3}}{s^2 + sXg_4g_5 + Y} \qquad (2)$$

where X and Y are,

$$X = \frac{g_1}{C_1} + \frac{C_2g_1g_3}{GC_1C_3} + \frac{g_3}{C_3} - \frac{\omega^2 C_2}{G}$$

$$Y = \frac{gm_5gm_4gm_3gm_2gm_1 + Gg_1g_3\,[gm_4 + g_4][gm_5 + g_5]}{GC_1C_3}$$

The format of Z_{in} shows that it is equivalent to a RLC network, as shown in Fig. 4. The‘s’ term in the numerator of equation (2), indicates the equivalent inductance and the real term indicates a resistor in series with the inductor.

From equation (2), L_{eq} and R_s can be written as,

$$L_{eq} = \frac{C_3Gg_4g_5}{gm_5gm_4gm_3gm_2gm_1 + Gg_1g_3\,[gm_4 + g_4][gm_5 + g_5]} \qquad (3)$$

$$R_s = \frac{Gg_3g_4g_5}{gm_5gm_4gm_3gm_2gm_1 + Gg_1g_3\,[gm_4 + g_4][gm_5 + g_5]} \qquad (4)$$

The parallel capacitance $C_p = C_1$ and the parallel resistance $R_p = 1/g_2$.

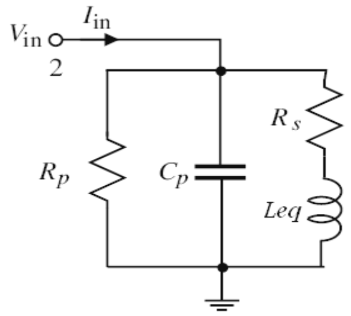

Fig. 6. Passive equivalent circuit of the Active inductor

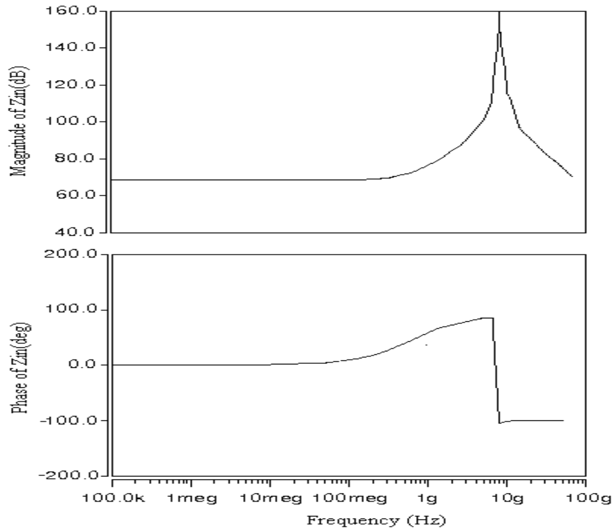

Fig. 7. Simulated Frequency Response of Input Impedance

The circuit is simulated using Synopsys HSPICE simulator in 180nm CMOS technology. The small signal parameters gm_1 = 156µS, gm_2 = 566µS, gm_3 = 250µS, gm_4 = 1.3mS, gm_5 = 60.5mS, g_1= 22µS, g_2= 80µS, g_3= 669µS, g_4= 94µS, g_5= 45µS, C_1=270.66aF, C_2=621.3aF, C_3=2.35fF, C_4= 3.45fF,C_5= 592.78aF and G= 802µS are found from the operating points.The simulated frequency response of Z_{in} of the active inductor is shown in Fig. 7. The magnitude of Z_{in} is nearly 160dB and the phase change is from +90◦ to -90◦. The magnitude response shows that it has real term and imaginary term. It is constant at 55dB up to 6.45 MHz which is equivalent to the real term. From the small signal parameters, the real term is calculated to be 123Ω, which is the series resistance Rs of the active inductor. The response is increased from 6.45 MHz to 6.3 GHz which is equivalent to the imaginary term, the equivalent inductance L_{eq}. The value of inductance ranges from 5nH to 550nH. Since it has less series resistance, the inductor loss is reduced.

The resonance frequency ω_o of the active inductor is given as,

$$\omega_0 = \sqrt{\frac{gm_5 gm_4 gm_3 gm_2 gm_1 + Gg_1 g_3 \left[gm_4 + g_4\right]\left[gm_5 + g_5\right]}{GC_1C_3}} \qquad (5)$$

The center frequency f_o is tuned through the controllable current source I_3 of Fig 4. Fig. 8 shows the tuning of the active inductor for various center frequencies. The controllable current source I_3 is varied between 35µA and 55 µA for tuning the center frequency of the active inductor. The designed active inductor has wide tuning range of 3.99 GHz to 16 GHz. For better gain Vgs has to be fixed to the different voltage for various center frequencies.

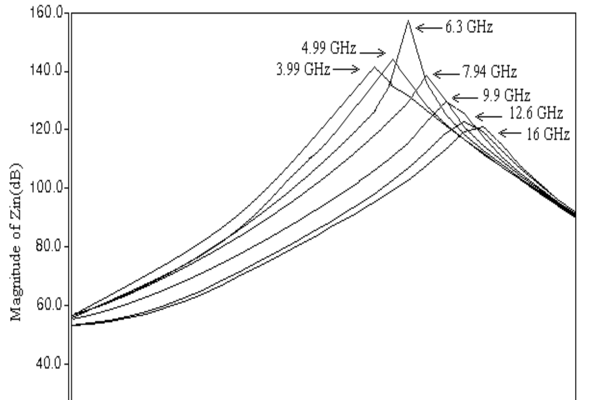

Fig. 8. Center Frequency Tuning of the Active Inductor

The quality factor Q_o of the active inductor at ω_o is given as

$$Q_0 = \frac{\sqrt{\dfrac{gm_5 gm_4 gm_3 gm_2 gm_1 + Gg_1 g_3 \left[gm_4 + g_4\right]\left[gm_5 + g_5\right]}{GC_1C_3}}}{\left[\dfrac{g_1}{C_1} + \dfrac{C_2 g_1 g_3}{GC_1C_3} + \dfrac{g_3}{C_3} - \dfrac{\omega^2 C_2}{G}\right] g_4 g_5} \qquad (6)$$

From the above equations, it has been noted that the transistor m5 enhances the loop gain and increases the quality factor of the active inductor by reducing the equivalent series resistance by the factor (gm_5+g_5). The transconductance of the m5 transistor is controlled through the Vgs supply at the gate of the transistor, which is turn varies the loop gain and the center frequency of the active inductor. Using the small signal parameters in the equations (5) and (6) the resonance frequency and quality factor are calculated to be f_o=6.3GHz and Q_o=723 respectively.

B. Simulation Results of Band Pass Filter

The bandpass filter of Fig. 3 is realized using two active inductors which are operated as two resonators of the top coupled topology as shown in Fig.2. The controllable current sources are used for tuning the gain and center frequencies of the band pass filter. The current source I_6 is used for tuning the center frequency and I_3 for varying the gain of the band pass filter.

The center frequency of the band pass filter is tuned through the controllable current source I_6 of Fig. 3. The center frequency ranges from 2.4 GHz to 12.4 GHz with the bandwidth ranges from 200MHz to 300MHz respectively. Fig. 9 shows the center frequency tuning of the band pass filter. Table 3 shows the experiment results of tuning of center frequency for different values of I_6 and corresponding Noise figure values.

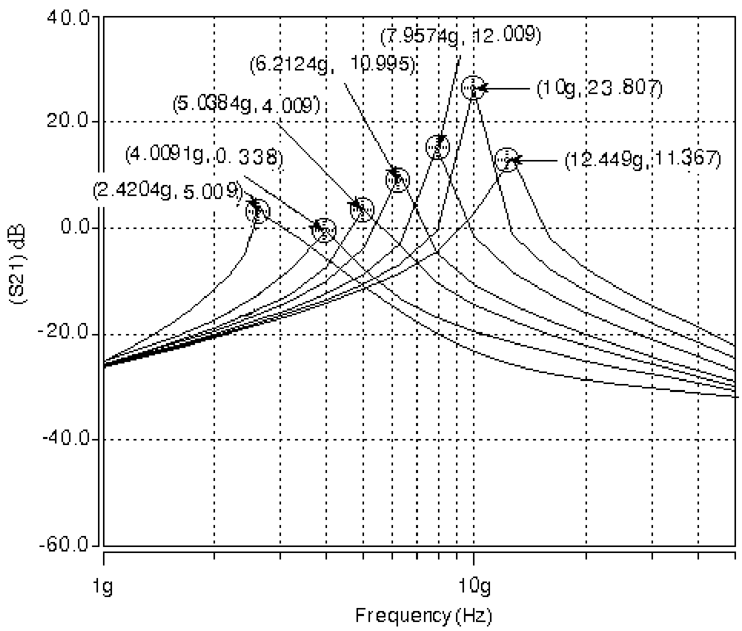

Fig. 9. Center Frequency Tuning of Band Pass Filter

The transfer gain S_{21} can be adjusted for different values through the current source I_3 of Fig. 3. Fig. 10 shows the tuning of the gain for 5GHz center frequency. The values of S21 ranges from -28dB to 18dB. Similarly, S21 can be adjusted for various centre frequencies through the current source I3.

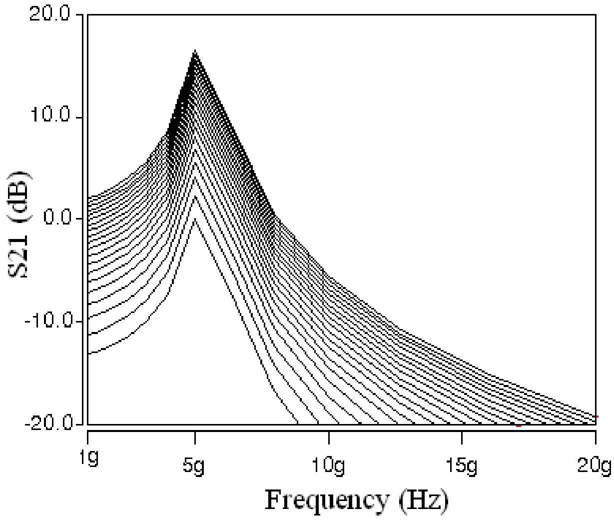

Fig. 10. Tuning of Gain S21

The reverse isolation S12 of the band pass filter for various centre frequencies have been simulated in Fig. 11. It has been noted that it has better reverse isolation gain for all the center frequencies and are less than -90dB.

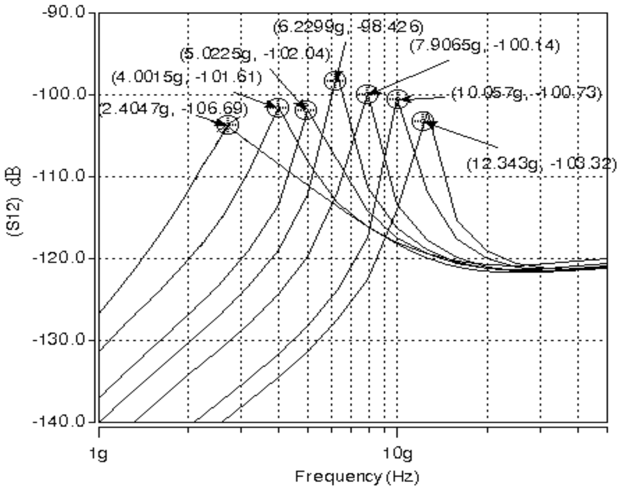

Fig. 11. Reverse Isolation S12 of Band Pass Filter for Various Center Frequencies

The simulated noise figure ranges from 6.8dB to 11.9dB for the entire tuning range 2.4 GHz to 12.3 GHz is shown in Fig. 12. The frequency spectrum for IIP3 calculation is shown in Fig. 13. The IIP3 is -3.5dBm which has been simulated for 1st and 3rd order frequency of 5GHz and 5.01GHz respectively as in Fig. 14.

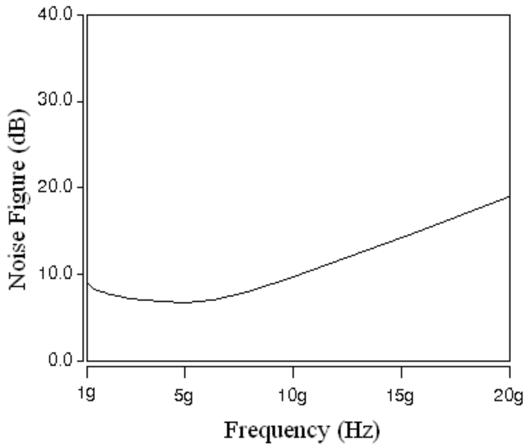

Fig. 12. Noise Figure of the band pass filter

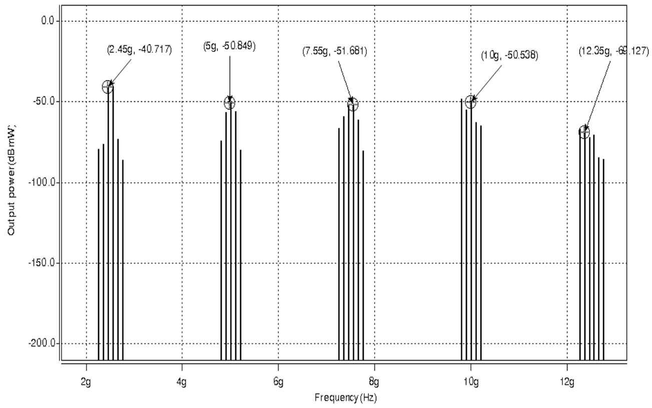

Fig. 13. Frequency Specturn of RF Band Pass Filter

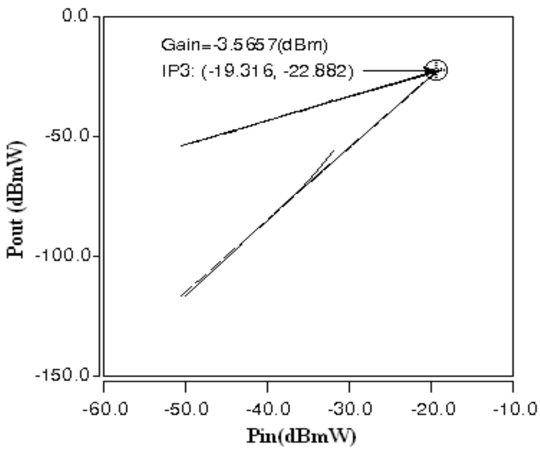

Fig. 14. IIP3 of the band pass filter

The measured gain of the band pass filter at 5GHz is shown in Fig. 15. When the current source I6 is set to 10µA by varying gate voltage of the controllable current source, the center frequency of the band pass filter is tuned to 5GHz.

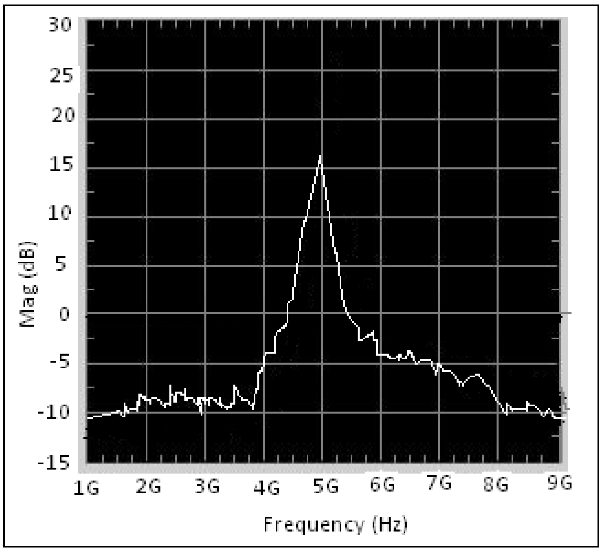

Fig. 15. Measured Gain of the Band Pass Filter at 5GHz.

The above simulation and the measured results shows that the band pass filter is tunable and consumes less power. Since the band pass filter is operated at 2.4GHz and 5GHz, it is suitable for WLAN RF front end. Table 1 compares the performances of the band pass filter with the reported works in the literatures [4, 9] of RF band pass filters. The comparison results show that the band pass filter features less power dissipation, low noise figure and better linearity.

Table 1: Comparison of Band Pass Filter Performances.

Parameter	Reference[4]	Reference[9]	This work
Technology	0.20 μm	0.18 μm	0.18μm
Filter order	2	2	2
ωo (GHz)	5.4	2.44	2.4 – 12.4 (tuning range)
P dis (mW)	4.4	10.8	2.9
Noise figure(dB)	25.6 (at 5.4GHz)	18 (at2. 45GHz)	6.8 – 11.9 (for tuning range)
IIP3(dBm)	-13.9	-	-3.5

III. CONCLUSION

A Tunable CMOS active inductor and RF band pass filter are simulated in a 180nm CMOS process. The simulation results of active inductor show that the circuit has wide inductive bandwidth and high resonance frequencies. The simulations results of RF band pass filter proves that it has better tuning of center frequencies, less noise and low power dissipation which makes it suitable for designing multiband RF front end circuits.

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