Application of the active inductor circuit (AI) based band pass Filter (BPF)

Hamdaouy Ridouane, Boussetta Mostapha, Khadija Slaoui

Abstract— **A novel active inductor approach, which can improve the quality-factor, was presented in this report. A current reuse active inductor circuit topology was proposed, which can substantially improve its equivalent inductance. This active inductor was implemented by using a 0.13um- CMOS technology, which gives a maximum quality-factor of 1.45 with a 40 nH at 2.259GHz. An optimized differential topology of an RF active filter and its performances are presented. It make use of negative resistance and fully differential in order to independently tune the central frequency and the quality factor with a minimum power consumption of 1.98 mW. The principle seems to be sufficiently general to be used for other active filters single–ended or differential topologies based on simulated inductors.**

Index Terms— **A Active Inductor, Active Inductor, CMOS, Q factor, band pass filter.**

I. INTRODUCTION

Active inductors have gained great interests in low-power applications over the last decade in an effort to reduce the need for large chip area required to fabricate spiral inductors and transformers. Inductive impedances are essential in high speed application due to their ability to improve bandwidth and boost gain, and to perform impedance matching and frequency selection.

Analog circuit design is a challenging field. At the same time designing an analog circuit with reduced area or smaller size is further challenging. The technology is also racing towards nm range. For the reasons maintaining cost to market ratio is a bigger challenge.

Integrated embedded passive components can save the circuit real estate and are critical for the circuit performance of RF circuit. Integrated passive components improve package and Multi-chip Module (MCM) efficiency, enhance electrical and high frequency performance by reducing the parasitic, and eliminate surface mount assembly procedure which in turn improves the yield and reliability due to the reduced solder joint failures.

At high frequencies, particularly radio frequencies (RF), inductors have higher resistance and other losses. In addition to causing power loss, in resonant circuits this can reduce the Q factor of the circuit, broadening the bandwidth. In RF inductors, which are mostly air core types, specialized construction techniques are used to minimize these losses. To reduce parasitic capacitance and proximity effect, RF coils are constructed to avoid having many turns lying close together, parallel to one another. Which are mostly air core types, specialized construction techniques are used to minimize these losses. The windings of RF coils are often limited to a single layer, and the turns are spaced apart. To reduce resistance due to skin effect, in high-power inductors such as those used in transmitters the windings are sometimes made of a metal strip or tubing which has a larger surface area, and the surface is silver-plated. An on-chip passive inductor presents major disadvantages such as large silicon area, limited inductance value and quality factor. Most of the time, the inductor will be a major factor in determining the total chip area [1] where higher inductance values imply larger area consumption. Furthermore, their values are not precise even if the technology is well-characterized [2, 3].

The increasing popularity and growth of wireless communications has inevitably boosted research in the field of radio-frequency integrated circuit (RFIC) design, especially in CMOS technology due to the shrinking of sizes and low cost availability of the process. The inductor, an essential component in RF design, finds use in many blocks such as oscillators, filters, phase shifters, low noise amplifiers, impedance matching circuitry, biasing, etc. [4,5,2.1].however their implementation still remains to be a challenging task in CMOS.A more area-efficient alternative to realize on-chip inductors is to use active inductors. Active inductors are particularly suitable for implementing shunt peaking in current-mode logic (CML) circuits due to both the low-Q requirement and low-swing nature of such circuits.

The active inductors offer much less area consumption independent of the desired inductance value, high quality factors and tunability. Although the noise performance and dynamic range will be degraded, it can be maintained at low enough levels for many applications [6, 7]. Nonetheless, active inductors have several drawbacks including: higher power consumption for biasing the transistors, relatively low power handling capability since the active inductors do not have the energy storage properties of physical inductors, higher noise figure owing to active components and limited bandwidth due to the active devices. Additionally, one side of the active inductor being always held at a constant DC bias restricts the possible applications. All these make active inductors difficult to integrate.

II. THEORY

A gyrator consists of two back-to-back connected transconductors. [8] When one port of the gyrator is connected to a capacitor, as shown in Fig.1, the network is called the gyrator-C network.

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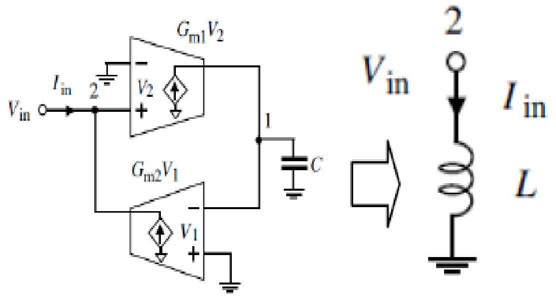

Figure .1. Lossless singel-ended gyrator-C active inductors.

A gyrator-C network is said to be lossless when both the input and output impedances of the transconductors of the network are infinite and the transconductances of the transconductors are constant. Consider the lossless gyrator-C network shown in Fig.1 The admittance looking into port 2 of the gyrator-C network is given by

$$Y_{in}=\frac{I_{in}}{V_{in}} = \frac{1}{\frac{sC_{GS}}{g_{m1}\cdot g_{m2}}} \quad . \qquad (1)$$

Eq.1 indicates that port 2 of the gyrator-C network behaves as a single-ended lossless inductor with its inductance given by,

$$L=\frac{C_{gs}}{g_{m1}g_{m2}} \qquad (2)$$

Gyrator-C networks can therefore be used to synthesize inductors. These synthesized inductors are called gyrator-C active inductors. The inductance of gyrator-C active inductor is directly proportional to the load capacitance C and inversely proportional to the product of the transconductances of the transconductors of the gyrator. Also, the gyrator-C network is inductive over the entire frequency spectrum. Although the transconductors of gyrator-C networks can be configured in various ways, the constraint that the synthesized inductors should have a large frequency range, a low level of power consumption, and a small silicon area requires that these transconductors be configured as simple as possible. Fig.2 shows the simplified schematic of the basic transconductors that are widely used in the configuration of gyrator-C active inductors.

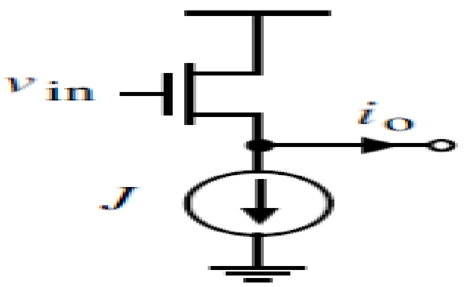

Figure 2 Simplified schematic of basic transconductors.

When either the input or the output impedances of the transconductors of gyrator-C networks are finite, the synthesized inductors are no longer lossless [9]. Also, the gyrator-C networks are inductive only in a specific frequency range.Consider the gyrator-C network shown in Fig.3 where Go1 and Go2 denote the total conductances at nodes 1 and 2, respectively. Note Go1 is due to the finite output impedance of transconductor1 and the finite input impedance of transconductor2. To simplify analysis, we continue to assume that the transconductances of the transconductors are constant.

The admittance looking into port 2 of the gyrator-C network is obtained from,

$$Yin=sC_{gs2}+GO2+\frac{1}{\frac{sC_{gs1}}{g_{m1}g_{m2}}+\frac{GO1}{g_{m1}g_{m2}}} \qquad (3)$$

Eq3 can be represented by the RLC networks shown in Fig.3

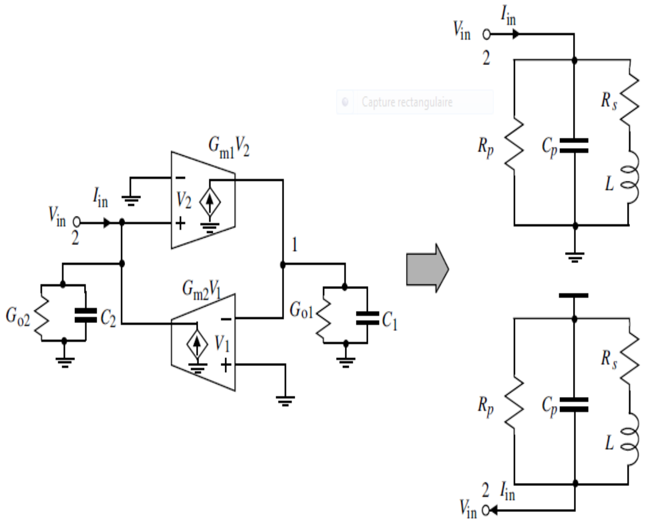

Figure .3: lossy single-ended gyrator-C active inductor.

We comments on the preceding results :

■ When the input and output conductances of the transconductors are considered, the gyrator-C network behaves as a lossy inductor. Rp should be maximized while Rs should be minimized to low the ohmic loss. The finite input and output impedances of the transconductors of the gyrator-C network, however, have no effect on the inductance of the active inductor.

■The resonant frequency of the RLC network of the active inductor is given by

$$\omega_0 \cong \frac{1}{\sqrt{LC_p}} \cong \sqrt{\frac{g_{m1}\, g_{m2}}{C_{gs1}\, C_{gs2}}} \cong \sqrt{\omega_{t1}\omega_{t2}}. \qquad (4)$$

Where

$\omega_{t1,2}$ =($g_{m1,2}$)/Cgs1,2 ; is the cut-off frequency of the transconductors. ωo is the self-resonant frequency of the gyrator-C active inductor. This self-resonant frequency is typically the maximum frequency at which the active inductor operates.

The finite input and output impedances of the transconductors constituting active inductors result in a finite quality factor Q. Q-enhancement techniques that can offset the detrimental effect of Rp and Rs should be employed to boost the quality factor of the active inductors. In 1988,89 a broad-band microwave active inductor has been proposed by Hara[9]. Inductances of 3.0 to 0.4nH and 7.7 to 0.9 nH are obtained at frequencies ranging up to 7.6 GHz and 5.5 GHz, respectively This active inductor consists of a cascode FET with a feedback resistor. The two FETs are obtained with 150 um &75 um gate width. Their areas are less than 0.15 mm2 even at high inductance value.

In 2000, a new circuit configuration of VHF CMOS transistor-only active inductor which allows very high frequency operation under low power supply voltages (<2 V) is proposed[10]. Gain enhancement techniques are applied to reduce the inductor losses achieving high-Q and wide operating bandwidth. HSPICE simulations using process parameters from a 0.35-µm CMOS technology show that the proposed floating active inductor operates under a single 1.5-V power supply voltage, exhibiting a self-resonant frequency of 2.5 GHz and a quality factor greater than 120 (phase errors <0.50) over the operating frequencies extending from 500 MHz to

1 GHz In 2002, a cascode-grounded active inductor circuit topology with a feedback resistance was proposed, which can substantially improve its equivalent inductance and quality-factor [11]. This feedback resistance active inductor was implemented by using a 0.18um 1P6M CMOS technology, which demonstrates a maximum quality-factor of 70 with a 5.7-nH inductance at 1.55 GHz.

III. PROPOSED ACTIVE INDUCTOR

A current reuse high-Q [11, 12,13] active inductor is introduced by Wu which avoids the use of expensive external or on-chip inductors. In this work we propose an active inductor based on Wu's current reuse active inductor . The schematic of Wu's active inductor is shown below in figure 4. The inductor presented in Fig. 4 is obtained based on the gyrator theory, where the load capacitor is represented by the parasitic capacitor C_{gs2} . Due to the presence of parasitic capacitor C_{gs1}, the circuit will have a self–resonance frequency, determined by the effective value of the inductor and parasitic capacitor C_{gs1} The quality factor for this circuit is limited by the series resistance Rs, which is determined by the finite output resistances of transistors, especially that of M_1.

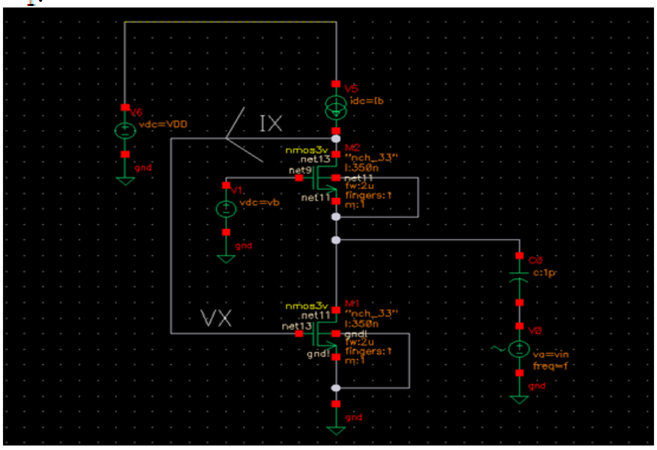

Figure .4: Active inductor architecture used in this work.

An approximate expression for input admittance Yin(s)=1/Zin(s) can be expressed as:

$$Y_{in} = sC_{gs2} + g_{ds1} + g_{m2} + \frac{(g_{m1} - g_{ds2})(g_{m2})}{g_{oi} + g_{ds2} + sC_{gs1}} \qquad (5)$$

As expected, the input admittance shows that it is equivalent to an RLC network, as shown in Fig. 3.

The element values can be determined as follows.

$$L_S = \frac{C_{gs1}}{(g_{m1} - g_{ds2})(g_{m2})} ; \qquad (6)$$

$$R_s = \frac{g_{oi} + g_{ds2}}{(g_{m1} - g_{ds2})(g_{m2})} ; \qquad (7)$$

$$R_p = \frac{1}{g_{ds1} + g_{m2}} ; \qquad (8)$$

$$C_p = C_{gs2} ; \qquad (9)$$

$s = j\omega$

$$Y_{in} = sC_p + Y_p + \frac{1}{R_s + sL_s} \qquad (10)$$

$$Z_{in} = \frac{R_s/C_pL_s + s/C_p}{s^2 + s\frac{(C_pR_s + Y_pL_s)}{C_pL_s} + \frac{(Y_pR_s + 1)}{C_pL_s}} \qquad (11)$$

$$\omega_0 = \sqrt{\frac{(Y_pR_s + 1)}{C_pL_s}} ; \qquad (12)$$

$$\frac{\omega_0}{Q} = \frac{(C_pR_s + Y_pL_s)}{C_pL_s} . \qquad (13)$$

$$Q = \frac{C_pL_s}{C_pR_s + Y_pL_s} \times \sqrt{\frac{(Y_pR_s + 1)}{C_pL_s}} ; \qquad (14)$$

$$\omega_z = R_s/L_s \qquad (15)$$

In most cases, due to the presence of the parasitic capacitor C_{gs1}, the active inductor is seen as a band pass filter with its resonant frequency.

The inductance L, the parasitic series resistance Rs and parasitic parallel resistance Rp are all functions of gm1 and gm2, which are determined by the channel current of M1and M2. The quality factor of Wu's active inductors can be estimated by neglecting the effect of Rs and only focusing on Rp as Rp is small.

$$Q = \frac{R_p}{\omega 0 * L_p} \qquad (16)$$

IV. SIMULATION RESULTS:

The active inductor the width of the transistors and value of Ib but the values of Ib is chosen to ptimize the quality factor of the inductor as W1 = 2um & W2 = 2 um, l1=l2 = 350nm, Vb=1.5V. The input bias current Ib is varied from 25uA to 95uA. We get an inductive range of 0.2053nH to 0.616nH at 543.1-812.8 MHz. The frequency response is as shown in fig.7.At frequency of 2.259GHz we get QMAX factor of 1.43 for Ib=95uA as shown in fig.8.but the fiqure.6 shown the variation the inductance L(0.8152nH-0.5133nH) versus the frequency f(449.2MHz-813.2MHz)respectively with Ib=95uA and Figure .9 shown tuning the quality factor(Q) versus frequency response f of the active inductor with Several biases current values Ib.

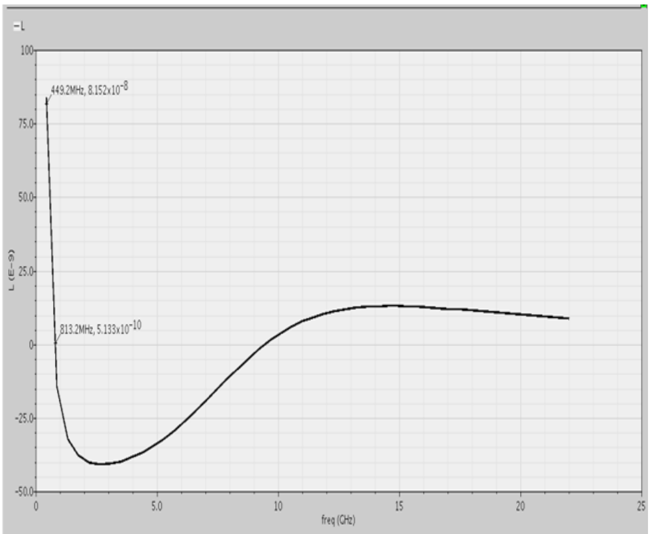

Figure .6: Inductance versus frequency response f of the active inductor with Ib=95Ua.

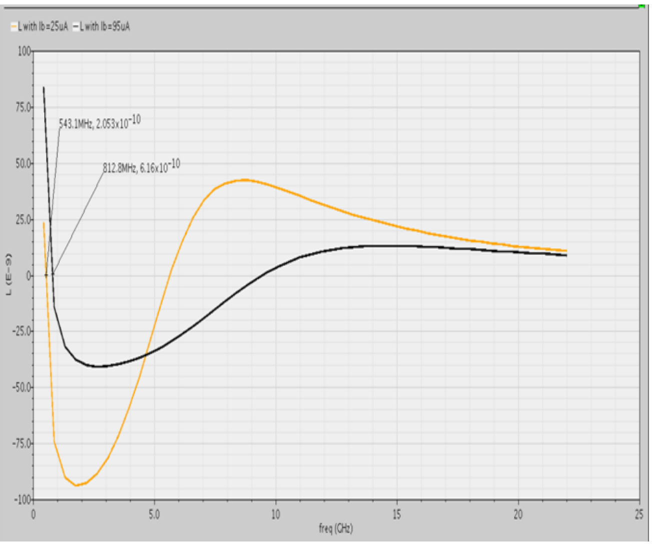

Figure .7:Inductance (L) versus frequency response (f) of the active inductor with Several biases current values Ib

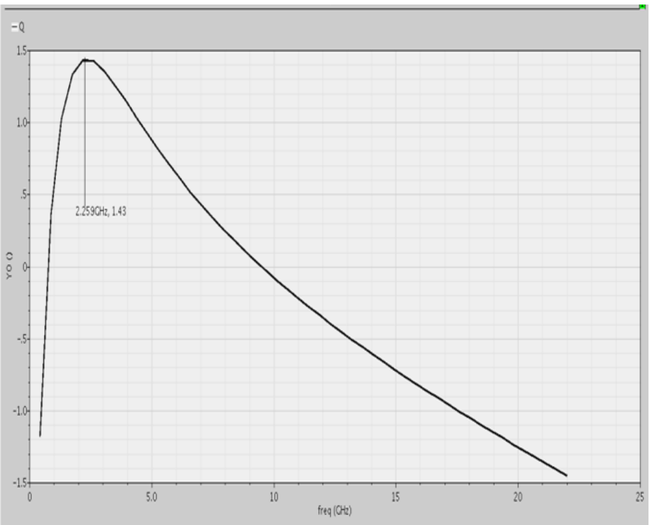

Figure .8:quality factor(Q) versus frequency response (f) of the active inductor.

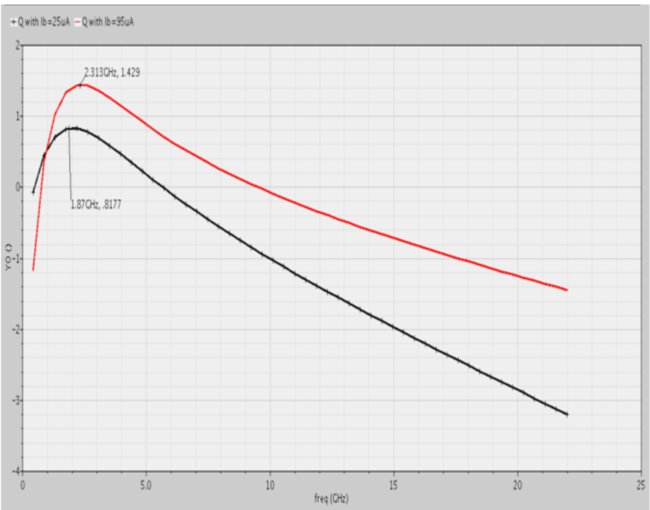

Figure .9:Quality factor(Q) versus frequency response f of the active inductor with Several biases current values Ib.

Where ω_{t2} and ω_{t1} are th transit (unit gain) frequency of M1 and M2, respectively.Thus, for a high- frequency active filter, both transistors need to be biased for a high unity -gain frequency.however, if M1 has a high unity-gain frequency , the selectivity of the filter will suffer.

It should be noted that because Rs and Rp are frequency-dependent, Rcom should be designed in such a way that a total resistance cancellation is achieved across the frequency range of the active inductor. It should also be noted that although the negative resistor compensation technique is widely used to improve the quality factor of spiral inductors, a total compensation in this case is difficult to achieve. This is because an active negative resistor is used to cancel out the largely skin-effect induced parasitic series resistance of spirals.

Another benefit of adding negative resistance is that the circuit can function over a wider range of frequencies since M1and M2 are only used to tune the centre frequency. Active inductor allowing to vary the parameters by processes variation then to implement in the pass band filter .Band pass filter: In a novel design approach, two active inductors are connected back-to-back through a coupling full differential and negative impedance convertor (NIC) are resulting in a wideband active BPF, as shown in Fig.10. The independent tuning of these inductors (using current Ib) allows one to obtain a circuit with amplitude response similar to filter characteristics. In an alternative approach, by controlling the voltage VIN0, the resonant peak of the BPF can not be tuned, as shown in fig.11. The frequency responses of the proposed BPF, terminated per both ends (input buffer (RS at source) and output buffer (RL at load)), as shown in Figure.12 this shows the characteristics of a filter.

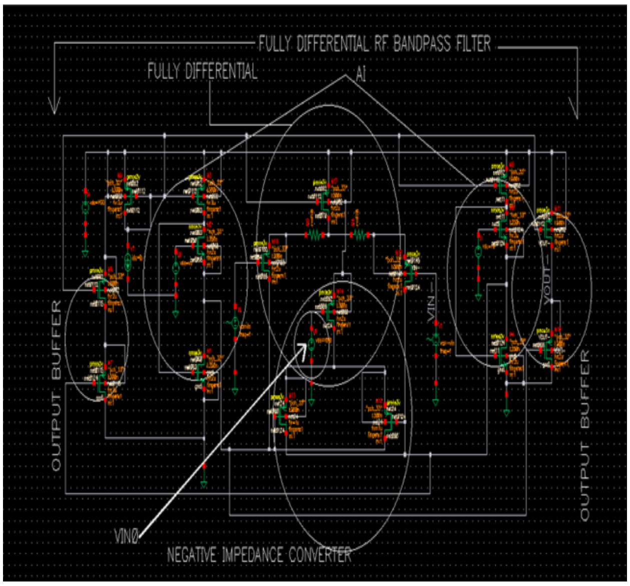

Figure.10: band pass filter(BPF).

A band pass filter passes the frequencies within a certain range and attenuates the other. A band pass filter can be synthesized by cascading a low pass and a high pass filter, where the cut-off frequency of the low pass filter is higher than the lower cut-off frequency of the high pass filter. Thus, a band pass filter requires at least two poles to shape its frequency response. By introducing a negative resistance in parallel to the active inductor, the frequency response of the active inductor is changed. On the other hand, the use of such a negative resistance is mandatory in designing RF tunable filters with a wide range of frequency and quality factor tuning. By setting suitable values for the negative resistance, an increase of the quality factor will be noticed while the resonant frequency will decrease.

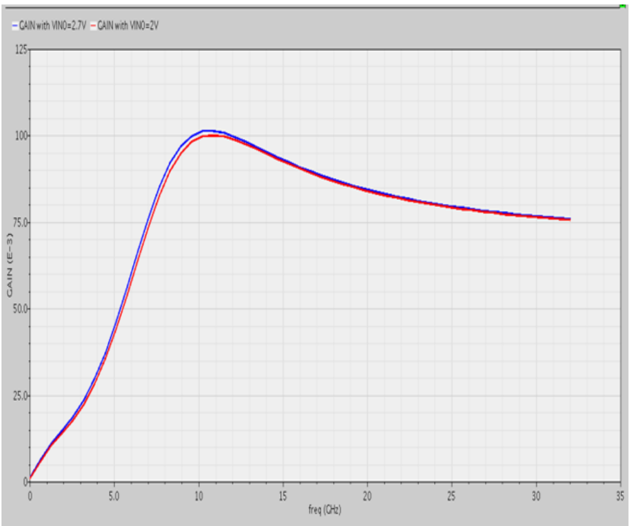

Figure.11: quality factor and resonance are invariant versus bias voltage VIN0.

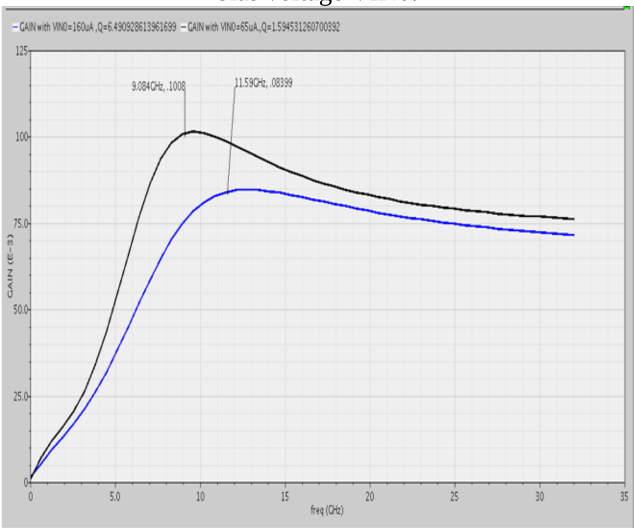

Figure.12: variants quality factor and resonance frequency versus Ib.

In all cases of proposed active inductors, the used negative resistances are dc coupled cross-coupled pairs of transistors, a fact that influences the transistor biasing and thus affects both filter parameters. By decoupling the negative resistance the tuning process is greatly improved but the interdependence problem still remains as shows fig.11 and fig .12.

For the circuit shown above, the input transistor M14 and M15 converts the input voltage ±vin to a current applied to the active inductor. This is the most effective way to change the gain of the filter. To keep the circuit as simple as possible a source follower is used as output buffer. If the input driver has no noticeable effect on the filter response, this is not the case for the output buffer, where the parasitic capacitance Cgs has a direct influence upon the resonant frequency. However this capacitor has a small value and the influence can be diminished by a proper design of the active filter. The real implemented circuit is given in Fig.10.

For the circuit presented in Fig. 10, the transistors (M1,M2 ,M5,M4) simulate the active inductor(AI), M12,M13 simulate the negative resistance ,the fully differential (M16,M15) is AC cross-coupled by capacitors while the DC biasing circuit is decoupled by large resistors RB, output buffer (M7,M8 ,M10,M11) while (M6,M3,M14 M9,M17) are current bias.

The principle of independent frequency and quality factor tuning is illustrated in Figs. 13 where the circuit proposed in Fig. 10 has been studied and the frequency limits of the frequency tunability range are found while sufficiently high quality factor values are obtained.

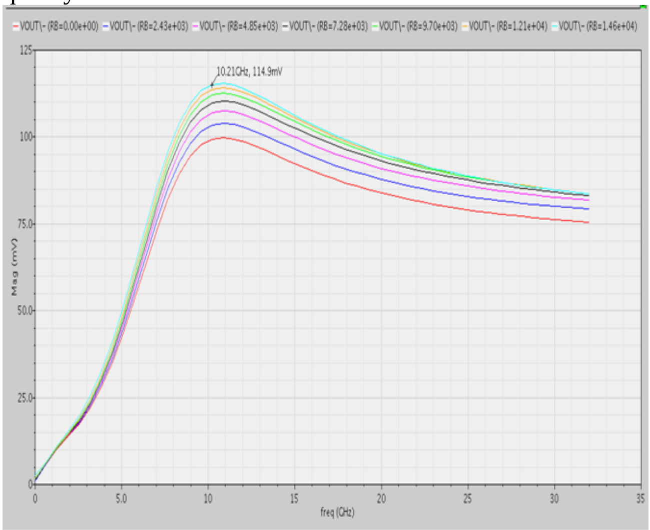

Fig. 13 – Independent tuning of Q, f0= 10.21 GHz.

The parasitic capacitors of the negative resistances (CGS) are responsible for the degradation of frequency response, these having a strong effect upon the resonant frequency of the active inductor. Since for a large tuning range a large negative resistance change for output resistance the output buffers required, the parasitic capacitors values are high as well. This drawback can be minimized by designing the active inductor at higher frequency values which means in fact higher power consumption.

An important advantage of the differential implementation consists in a more efficient use of the occupied area compared to the single-ended case. Furthermore, a negative resistance twice larger than in the single-ended case is needed and thus the power consumption is halved.

An important remark regarding the frequency tuning range regards the use of real current sources. The use of transistors instead of ideal current sources [14] decreases the frequency tuning range about 30% compared to the ideal case.

V. CONCLUSIONS

An independent tuning principle for resonant frequency and quality factor has been studied for a differential bandpass filter based on a simulated inductor in a 0.13µm CMOS process. The simulations show that the filter can achieve independent frequency and quality factor tuning with very low power consumption.

Due to its small size and programmability, it can be a potential candidate to multistandard communications systems in implementing higher order active bandpass filters..

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