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Improved Fully Differential Low Power Active Filter

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ABSTRACT

This paper relates the new topology and simulations of a fully differential CMOS active filter for mm wave band applications. The advantages of the differential topology over the single ended one are discussed and the quality factor is tuned to insure application requirements, including narrow bandwidth and high selectivity due to a differential negative resistance that reuses the filter's current. Using this topology enables independent tuning of the quality factor and low power consumption while compensating the resistive loss of the filter. Very high filter performance was obtained with the simulated active inductor based active filter that was designed using CMOS 0.35 µm technology from AMS foundry and that resonates at 30 GHz with a high quality factor of Q > 500.

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1. INTRODUCTION

The great revolution in wireless and satellite communications requires the implementation of high performance On-chip tunable systems with a high quality factor and low power consumption which remain some of the most challenging stages in transceiver conception. Narrow bandwidth and high quality factor can be obtained by SAW filters which present the advantages of an out-of-band rejection and flexibility bandwidth control [1], however they have a large size and can't be tuned in frequency easily. Active inductor based filters reported in literature [2]-[3], have many advantages such as frequency tuning ease, high quality factors and on-chip integration, however the transistor's parasitic effects impact the linearity and stability of systems which calls for accurate noise and linearization studies [4]. These filters are based on the gyrator principle that transforms a capacitance connected to its input to an inductance without the use of passive inductors. Since their creation by HARA [5] many topologies were reported, but still cannot reach the very high quality factors of passive off-chip circuits. The major constraint is the frequency tuning independently of the quality factor tuning. The quality factor can be enhanced by using a negative resistance that compensates the resistive loss which changes the centre frequency of the filter as well. The purpose is to propose a topology that allows very high quality factor, low power consumption and narrow bandwidth.

According to the spectrum specifications for frequencies around 1.4 GHz used in fixed satellite services [6], the requirements are a very narrow bandwidth (< 2MHz) and a high quality factor. In this paper, the proposed circuit design is using two reciprocal active filters joined by a BALUN; the resistive serial and parallel loss are compensated using a negative resistance which is designed to perform independent quality factor tuning without the variation of centre frequency. It is first intended to operate in the L band, then it will be improved in order to reach higher frequencies namely the millimetre wave band around 30 GHz for 5G applications. The new topology uses a varactor that allows frequency tuning for a range up to 16 GHz with a constant bandwidth of 40 MHz. Active matching networks were added with properly biased active components to achieve better performances, namely out-of-band rejection, stability and low noise figure.

2. DIFFERENTIAL FILTER DESIGN

The fully differential active filter shown in Figure 1 consists of two reciprocal active inductors made of transistors M1 and M2 connected in feedback based on HARA topology[5]. The active inductor transforms the gate-to-source capacitance Cgs of transistor M1 to an inductance replacing classical spiral inductors. Each active inductor is equivalent to an RLC resonator as seen from input impedance of the active inductor composed of M1 and M2 (1). The series resistance Rs, the parallel inductor Lp, the parallel resistance Rp and capacitance Cp, form an RLC resonator that resonates at 1.43 GHz where gdsi and gmi i(1,2) are respectively the output conductance and the transconductance of transistor i.

$$\mathrm{Zin} = \underbrace{\frac{\mathrm{gds1}}{\mathrm{gm1}\times\mathrm{gm2}}}_{\mathrm{Rs}} + \underbrace{\frac{\mathrm{p\,Cgs2}}{gm1\times gm2}}_{\mathrm{Lp}} + \underbrace{\frac{1}{\mathrm{gm1}}}_{\mathrm{Rp}} + \underbrace{\frac{1}{\mathrm{p\,Cgs1}}}_{\mathrm{Cp}} \qquad (1)$$

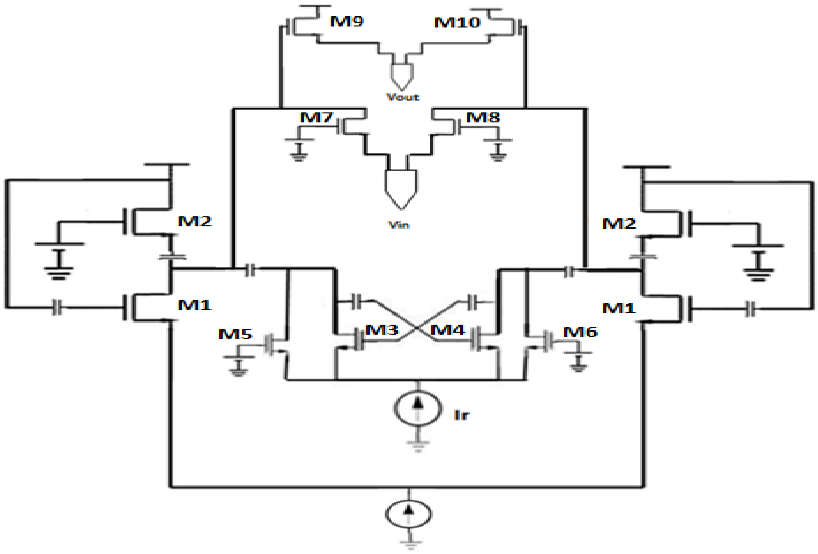


Figure 1. Fully differential CMOS active filter topology

The inductive effect is obtained below the centre frequency due to R and L and the capacitive effect above it due to C. The real part of the impedance reaches the peak at the resonant frequency. Transistors M3-M6 form a differential negative resistor at the input of the active inductors that cancels parasitic loss caused by Rs and Rp (conductance gds1 and transconductances gm1 and gm2). These cross-coupled transistors generate the current necessary to compensate the current loss in the active inductor. An increase of current in the gate of M3 will increase the drain current of M3 and at the same time increase the drain current of M5 and gate current of M4 which will raise the drain current of M4 and M6 and gate of M3 accordingly. Therefore a positive feedback loop is set up. The value of the negative resistor can be controlled by current Ir or voltage Vgs of the transistors M3 and M4, which will subsequently boost the quality factor of the filter.

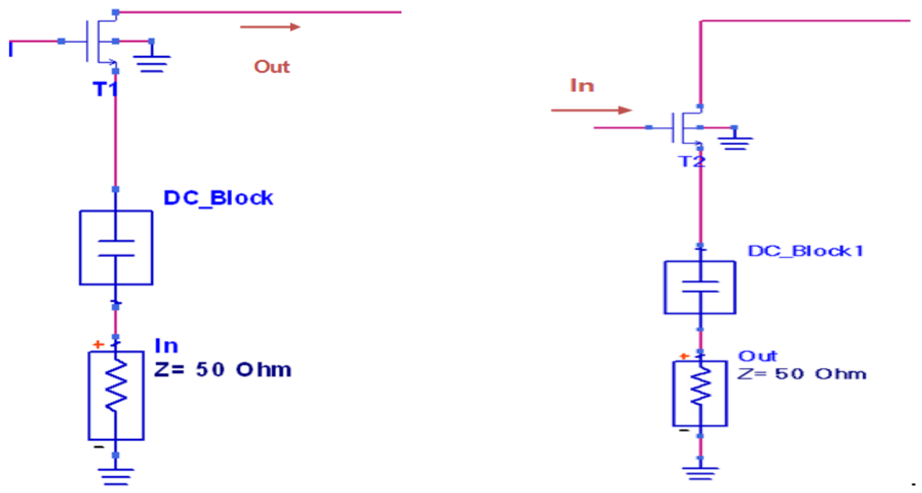
M7 and M8 are used in the input stage. They are common gate mounted and biased to have the lowest input impedance possible. pMOS transistors can also be used at the input stage due to their unity gain. The transistor used (Figure 2(a)) has a low input impedance (1) and a very high output impedance. At low frequencies, the input and output admittances of the input stage can be approximated to (2)-(3):

$$Ye \approx gm + j.w.Cgs \qquad (2)$$

$$Ys \approx gds \qquad (3)$$

On the output side, common drain transistors M9-M10 are utilized. The common drain transistor (Figure 2 (b)) has infinite input impedance and output impedance of the order of 1/gm. The optimization of the input and output buffers consists in choosing an optimum operating point in the saturation region in order to reduce the sensitivity of the frequency caused by a variation of the load.

The negative resistance is AC_coupled to the active inductor to reduce linearity issues. All currents and voltages were optimized to attain the frequency and selectivity required. To realize the differential function at both ends of the filter, a BALUN (Balanced to Unbalanced voltage) is employed [7]-[8]. The ADS (Advanced Design System) component BALUN1 (Figure 3) maps between a grounded referred signal (Vin for input and Vout for output) and a balanced couple of signals (signals coming from both inputs and outputs of the active inductors).



(a) (b)

Figure 2. (a) Input stage (Common gate transistor) (b) output stage (Common drain transistor)

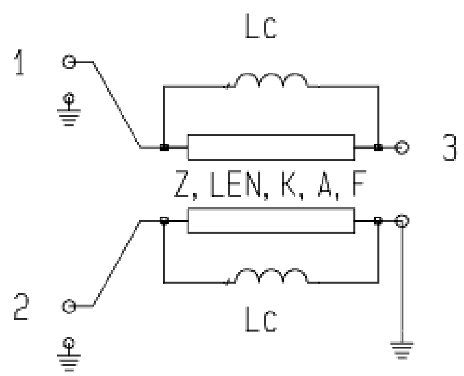


Figure 3. BALUN circuit

It is composed of two LC tanks whose role is to make the operation cited in (4) and (5). Where V1 and V2 are the input voltages of each active inductor, while Vin is the overall input of the filter.

$$\mathrm{Iin} = \frac{\mathrm{I1} - \mathrm{I2}}{2} \tag{4}$$

$$\mathrm{Vin} = \mathrm{V1} - \mathrm{V2} \tag{5}$$

Where V1, V2 and I1, I2 are input voltages and input currents of both active inductors (M1, M2). They can be calculated from the small signal circuit (Figure 4) taking into account the gate-to-source capacitance Cgs, the transconductance Gm and the output conductance Gds of transistors M1, M2. From the small signal study of the circuit, the expression of the input voltage can be given in (6):

$$V1 = \frac{Vds1\ (Gds1 + Cgs2.p)}{Gm1 + Gds1} \tag{6}$$

As can be deducted from (6), V1 is inversely proportional to the transconductance Gm of the transistor, thus, the filter presents more loss as Gm increases [9].

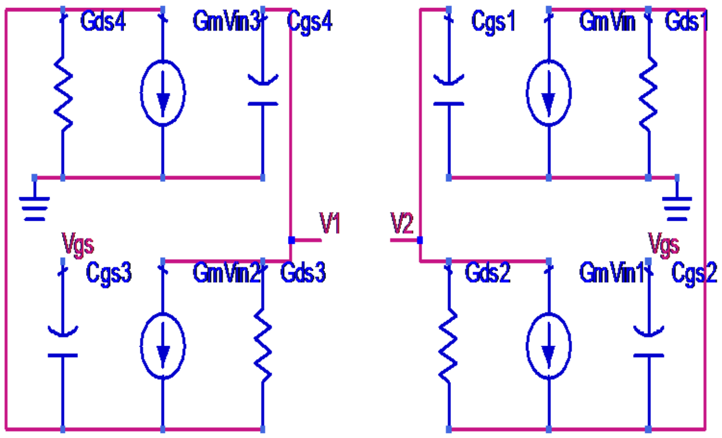


Figure 4. Small signal equivalent circuit of the differential active inductor

3. RESULTS AND DISCUSSION

Simulation of the power gain of the active filter shows that it reaches the resonance at 1.43 GHz (Figure 5).

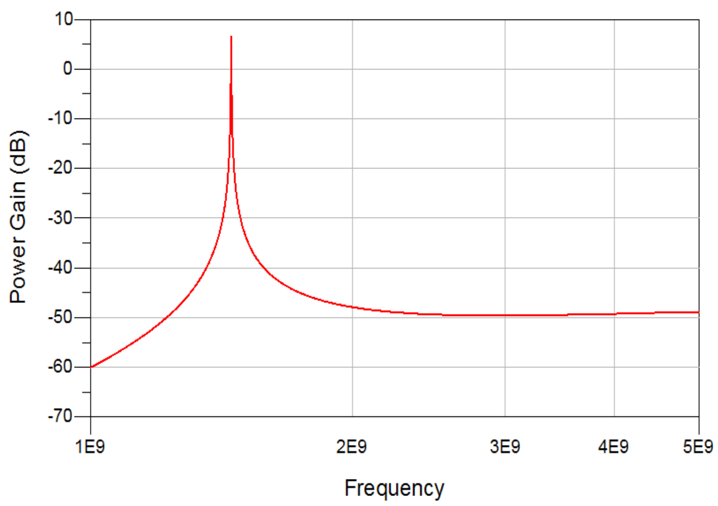


Figure 5. Gain of the differential active filter F=1.43 GHz, Q=1452

The quality factor of a single ended filter defines the selectivity of the filter; it is defined as (7).

$$Q = \frac{Imaginary\ (Zin)}{Real\ (Zin)} = \sqrt[2]{\frac{Gm2 \times Cgs1}{Gm1 \times Cgs2}} \tag{7}$$

For the 1.4 GHz frequency, widely used for point-to-point communications, the maximum bandwidth is 3.5 MHz. The simulated bandwidth of this active filter is 1MHz with a quality factor >1000. The quality factor of the filter can be tuned by varying the negative resistor current Ir or voltage Vgs of both transistors M3 and M4. The small signal circuit of the negative resistor is given in Figure 6 (neglecting M5 and M6).

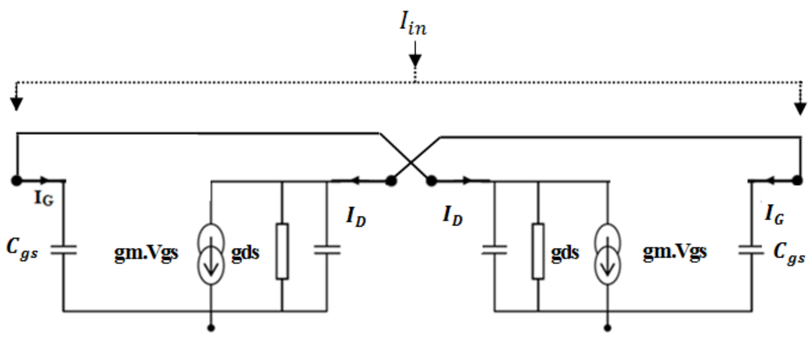


Figure 6. Small signal circuit of the negative resistor

From Figure 6 and considering a current source Iin connected between the two drains of the transistors, the input impedance of the negative resistor is:

$$Zin = \frac{Vin}{Iin} = \frac{1}{-(gm1+gm2)+(dgs1+gds2)+j.w.(\frac{1}{Cgs1}+\frac{1}{Cgs2})} \tag{8}$$

$$gm_{1,2} \gg gds_{1,2}$$

Zin becomes:

$$Zin = \frac{Vin}{Iin} = \frac{1}{-(gm1+gm2)+j.w.(\frac{1}{Cgs1}+\frac{1}{Cgs2})} \tag{9}$$

When the two transconductances are identical, the real part of the input impedance is given in (10):

$$\mathcal{Re}(Zin) = \frac{1}{-2gm} \tag{10}$$

The negative resistor can then be equivalent to the circuit of Figure 7:

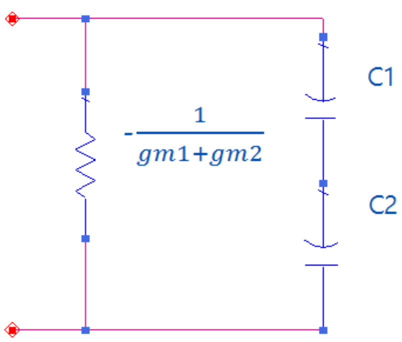


Figure 7. Equivalent negative resistor circuit

According to Equation (9) and Figure 7, the negative resistor introduces a parasitic capacitance which affects the resonance frequency. Therefore a variation of the quality factor will inevitably lead to a variation of the central frequency, hence the need to use a varactor to control the centre frequency tuning. This principle is discussed in another work [10].

While using the single ended filter topology and as can be noticed from (7), the quality factor depends on the transconducance of the transistors M1 and M2, which is inversely proportional to the centre frequency of the filter, therefore any changing in the biasing current of the negative resistance to tune the quality factor of the filter will inevitably affect its centre frequency. Hence the advantage of using the differential topology instead. When both inputs of the differential filter are connected to the negative resistance, the need of the compensation in both ends is twice the need for single ended one, so the transconductances of transistors M3 and M4 are reduced thus low power consumption is needed. As a consequence, while tuning the quality factor of the filter, the centre frequency is slightly changed because of the low current. The simulated Q_tuning is shown in Figure 8 for a centre frequency of 1.4 GHz.

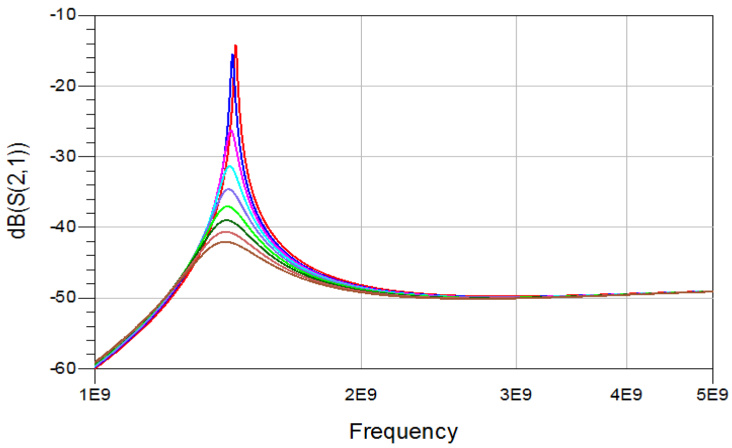


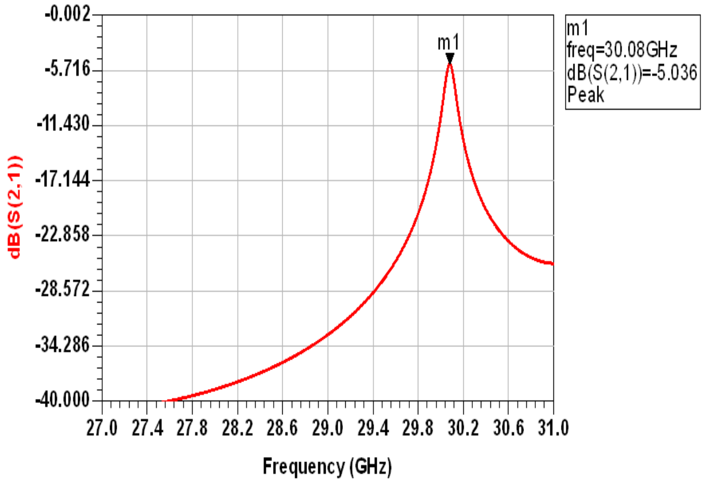
Figure 8. Q_tuning by varying current Ir of the negative resistor

The variation of the resistance is mainly used to control the filter bandwidth. Thus, for best selectivity, the lossy equivalent resistance of the circuit can be cancelled. Conversely, to widen the bandwidth, the value of this resistance must be increased. From the simulation results, the quality factor can be further improved to achieve very high selectivity of the bandpass filter. For applications in EHF band, the centre frequency of the filter may be tuned in order to operate in the millimetre band [11], by varying the polarization conditions and the values of the components of the circuit. A study of noise and stability of the differential filter is also necessary, insofar as this filter can operate correctly at very high frequencies.

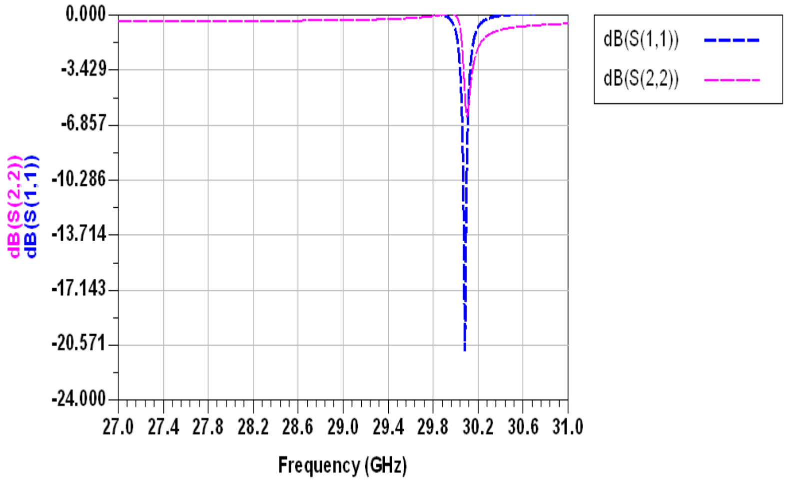
The optimization of polarization conditions of the differential active inductor as well as the variation of Vgs and Vds of input and output stages made it possible to reach the millimeter band beyond 30 GHz (Figure 9). Moreover, the use of the differential topology over the single ended one reduces considerably the current consumption; a very prevalent problem in RF design [12]. At the centre frequency of 30 GHz and for a bandwidth of 50 MHz, we find satisfactory S parameters: good input and output voltage reflection coefficients (-20 and -7 respectively). A good selectivity and acceptable insertion loss were also observed. Input and output matching network Regarding the evaluation of the stability of the filter, and as mentioned in Equation (11), the stability factor K is inversely proportional to the gain (S21) of the filter:

$$K = \frac{1+|\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{21}||S_{12}|} \qquad \text{with} \qquad \Delta = S_{11}S_{22} - S_{12}S_{21} \tag{11}$$

The stability factor must never be less than 1 for the filter to be unconditionally stable. Stability mainly depends on input and output matching networks [13]. The stability curve of the filter is given in Figure 10. It can be seen that the K factor remains greater than 1 over the entire frequency range.



(a)



(b)

Figure 9. S parameters (a) S21, (S11 and S22)

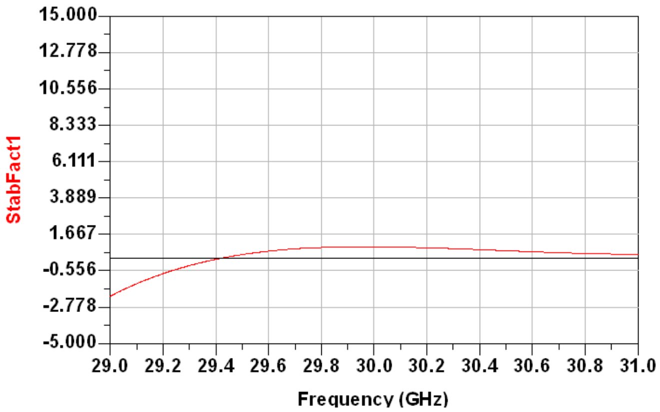


Figure 10. Stability performance

By comparing the performance of the differential topology with the single ended one, discussed in [10], the stability of the fully differential filter is observed over a wide band, unlike the single ended topology which varies abruptly around the central frequency. This feature is very necessary for frequency tunable applications that require a wide band of stability. In addition, by using differential voltage, differential mode reduces non-linear effects, which explains its enormous use in receivers. Finally, the particularity of this solution over others is that it can reach the millimeter band, thanks to the adaptation stages used and the negative resistance, while most active filters operate in UHF band [14]-[15], in addition to the improvement in energy consumption.

4. CONCLUSION

In this paper a fully differential active filter was described operating at 30 GHz for mm wave band applications. It has been demonstrated that active filters can reach very high quality factors due to gyrator topology and negative resistor principle. The use of the differential topology allows power economy and quality factor tuning without varying the centre frequency. Nevertheless, proper biasing conditions for all transistors are crucial to reach high selectivity and out_band rejection. The filter presents very good performances, frequency tuning ease and high frequency. The CMOS based filter offers a good selectivity and can have a larger bandwidth if required at the cost of gain loss. This filter may be used to design bi-band filters for applications requiring multiple frequencies.

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