ON A RF BANDPASS FILTER TUNING METHOD

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An optimised differential topology of an RF active filter and its performances are presented. It makes use of two negative resistances in order to independently tune the central frequency and the quality factor. The differential active filter is designed in a 0.18 µm CMOS process, allows a maximum wide range of tunability between 1.6 and 2.6 GHz and exhibits high quality factor values with a maximum power consumption of 0.6 mW. The principle seems to be sufficiently general to be used for other active filters single–ended or differential topologies based on simulated inductors.

1. INTRODUCTION

The increasing demand for low cost and high integration level of RF front–end made CMOS technology to be an important option, a critical issue being represented by the bandpass filters. At present all RF transceivers use SAW (surface acoustic wave) filters which have good frequency performances but are bulky and give no possibility of full on–chip integration. The architecture of a multistandard transceiver currently used in mobile devices includes at least 10 SAW filters to cover all mobile standards. This has a great impact on the final market price and the customer behaviour when buying mobile terminals. Although many passive solutions are available on the market [1–2], reconfigurable RF preselective filters are rather difficult to implement. To fulfil the frequency specifications higher order bandpass filters have to be used for their implementation.

Many solutions of passive LC bandpass filters have been proposed in the literature [3–6], most of them making use of negative resistances in order to increase the inductors quality factor. Closely to this technology trend and research efforts, another trend which gains more and more attention is represented by RF simulated inductors. Beside classical simulated inductors, based on the gyrator theory and most suitable for low frequency applications, RF simulated inductors

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make use of transistors only with no external capacitors. In this case, instead of external capacitors, the parasitic transistor capacitors are used to simulate the active inductor. Many active RF inductors topologies have been proposed in the literature [7–14].

A supplementary negative resistance is used to increase the quality factor for both active and passive inductors. Active inductors based RF bandpass filters prove to be a suitable solution for multistandard systems in which case the frequency and quality factor must be independently adjusted in order to fulfil the specifications. However, the tuning process is only nearly orthogonal for the method proposed in [8]. In all cases, separate tuning of either frequency or quality factor has a negative effect upon the other parameter, decreasing its value due to the direct dc connection of negative resistances to the active filter and the inter–dependency between these two parameters.

To overcome the mentioned difficulties, the principle of independent frequency and quality factor tuning has been proposed for single-ended topology [15] and differential one [16]. In this paper, an inductorless bandpass filter is designed for the bands 1.6–2.6 GHz with good tuning performances. As simulations reveals, this principle proves to be more suitable for implementation concerning area efficiency, current consumption and tuning capabilities. In order to fulfil the RF filtering specifications a cascade of minimum two such bandpass filters should be used.

2. **CIRCUIT PRINCIPLE**

As shown above, a large number of active inductor topologies have been introduced in the literature. All structures aim at keeping the number of transistors as small as possible due to their noise effect and parasitic resistances which degrade the quality factor of the active inductor.

The inductor used in the filter structure contains two transistors and one current source, as shown in Fig. 1. Its main advantage consists in the ease of frequency tuning possibility, allowing the possibility to work in the GHz range. The inductor presented in Fig. 1a 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 R_s, which is determined by the finite output resistances of transistors, especially that of M_1. An equivalent small signal circuit is shown in Fig. 1b.

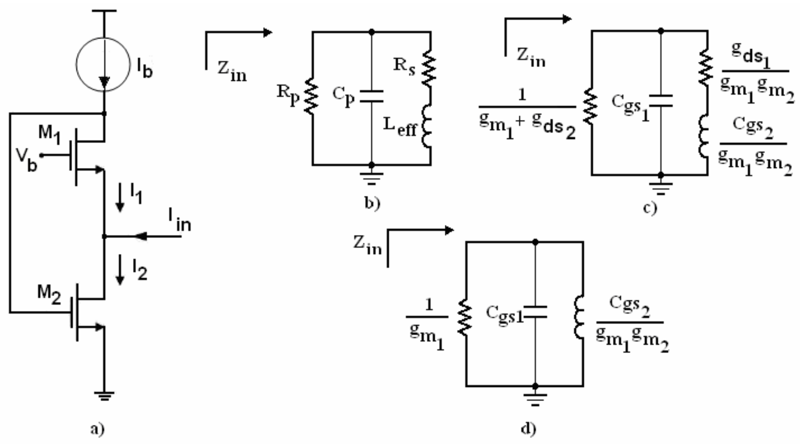

Fig. 1 – Active inductor a), its equivalent circuit b), and two particular cases c) and d).

Computing the expression for the input impedance (taking into account the output resistance for transistors) and simplifying it we find that the input admittance $Y_{\text{in}}(s) = 1/Z_{\text{in}}(s)$ is given by:

$$Y(s) = s \cdot C_{gs1} + g_{ds2} + \frac{(g_{ds1} + g_{m1})(g_{m2} + s \cdot C_{gs2})}{g_{ds1} + s \cdot C_{gs2}}. \tag{1}$$

The equivalent circuit can be represented by a parallel circuit consisting of the parasitic capacitor C_{gs1}, the output resistance of transistor M_2 and another term which is influenced by the M_1 output resistance. Regarding the effect of g_{ds1}, there are two cases. The first corresponds to $g_{ds1} \neq 0$, meaning that the output conductance is high enough and should be taken into account – $Y(s)$ can be expressed as:

$$Y(s) = s \cdot C_{gs1} + g_{ds2} + g_{m1} + \frac{g_{m1} g_{m2}}{g_{ds1} + s \cdot C_{gs2}}. \tag{2}$$

It is worth noticing that in this case the input impedance consists of a parasitic capacitor (C_{gs1}), a parallel resistance (r_{ds2}) and an inductor $\left(L_{eff} = \dfrac{C_{gs2}}{g_{m1} g_{m2}} \right)$ with a series parasitic resistance $\left(r_s = \dfrac{g_{ds1}}{g_{m1} g_{m2}} \right)$. The equivalent circuit is represented in Fig. 1c. It can be seen that the inductor quality factor strongly depends on the output resistance of transistor M_1. When the current supply increases in order to raise the resonance frequency both g_{ds1} and g_{ds2} increase. This is the reason of using a negative resistance, whose value slightly decreases the series resistance.

In the second case, when $g_{ds1} = 0$, $Y(s)$ is expressed by (3), where the output conductance g_{ds2} was also neglected.

$$Y(s) = s \cdot C_{gs1} + g_{m1} + \frac{g_{m1} g_{m2}}{s \cdot C_{gs2}} . \tag{3}$$

For this case, the equivalent small signal circuit for the active inductor is given in Fig. 1d. It is worth noticing that a good inductor can be obtained in this case with no series parasitic resistance.

In conclusion, the main condition to design a good simulated inductor consists in keeping the output resistances for both transistors as high as possible, fact almost impossible in practice, especially for high frequencies where high current values are needed.

Neglecting the transistor output resistance, the relations for the resonant frequency and quality factor are given respectively by:

$$\omega_0 = \sqrt{\frac{g_{m1} g_{m2}}{C_{gs1} C_{gs2}}} , \quad Q = \sqrt{\frac{g_{m2} C_{gs1}}{g_{m1} C_{gs2}}} . \tag{4}$$

It is obvious that it is impossible to change one of the above parameters while keeping unchanged the other one. This has a negative effect on the tuning principle as, in order to be used as a multi-standard filter, there is a need to keep the frequency response unchanged at any frequency of interest. This can be overcome by the independent tuning principle for frequency and quality factor which is reviewed in the following.

3. **INDEPENDENT TUNING PROTOTYPE**

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 compulsory 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.

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. For the differential version of the active inductor topology proposed in [13], any change of the negative resistance value changes the current through M2 and thus has a negative impact upon the resonant frequency as well. By decoupling the negative resistance the tuning process is greatly improved but the interdependence problem still remains.

In [15] an independent frequency and quality factor tuning principle has been proposed and represented in Fig. 2 as well, proving to be a promising solution in implementing CMOS reconfigurable RF filters. Thus, by adding a second negative resistance to the active filter, decoupling them and setting the right values for both negative resistances, the interdependence between those two parameters is eliminated and thus an independent tuning for both parameters becomes possible. Further details regarding this principle are presented in [15].

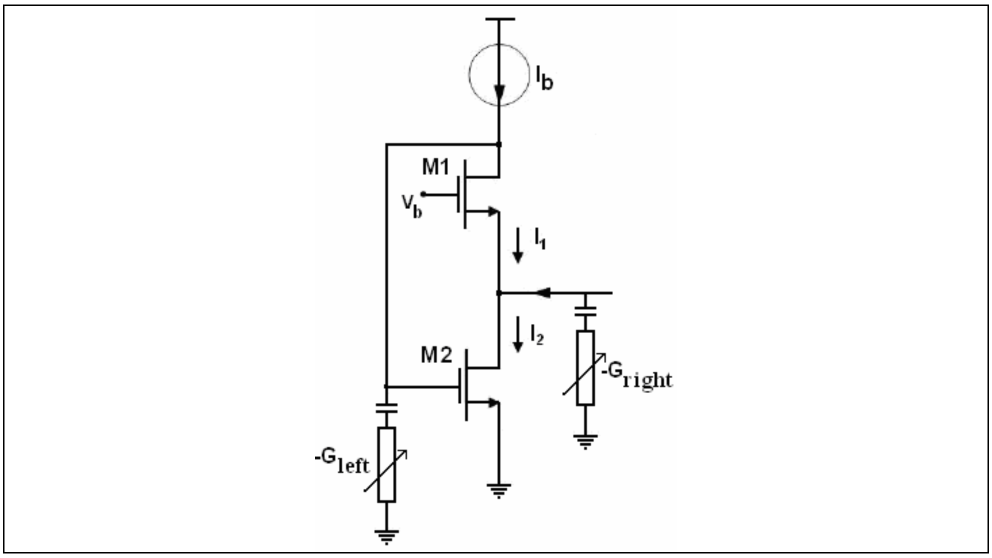

Fig. 2 – Independent tuning principle illustrated for a simulated inductor based active filter.

The circuit implemented in this paper follows the general architecture shown in Fig. 3.

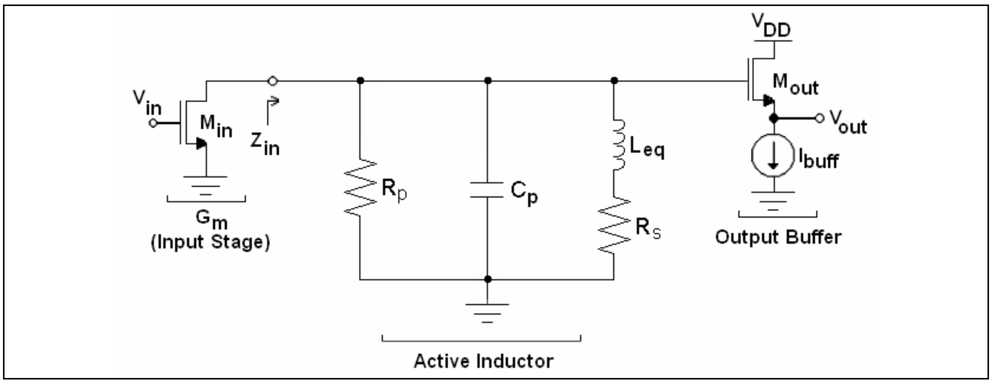

Fig. 3 – Active inductor based implemented bandpass filter.

For the circuit shown above, the input transistor M_{in} converts the input voltage v_{in} 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 C_{gs} 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. 4.

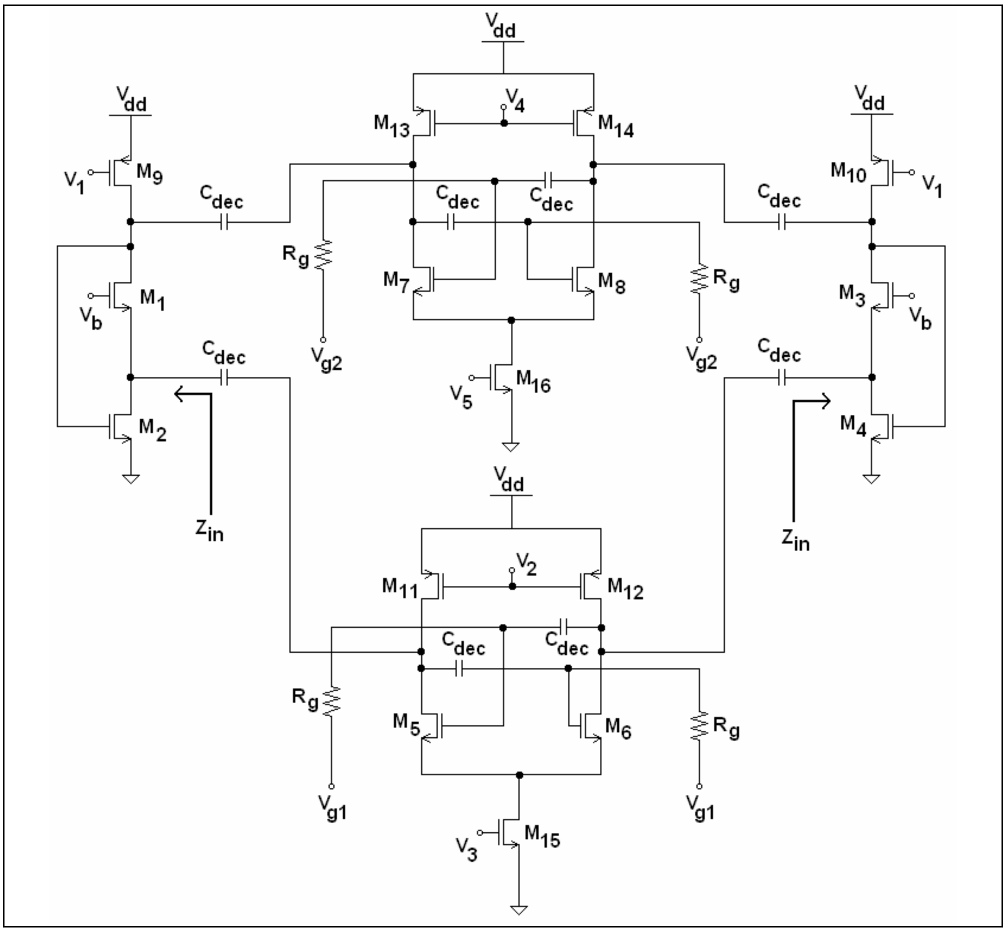

Fig. 4 – Differential active filter with independent tuning.

For the circuit presented in Fig. 4, the transistors M_1–M_4 simulate the active inductor, M_5–M_8 simulate the negative resistance while M_9–M_{16} are current sources.

The differential pairs M_5–M_6 and M_7–M_8 are AC cross-coupled by capacitors C_{dec} while the DC biasing circuit is decoupled by large resistors R_g. The effective value of the negative resistance is given by (5), where g_m and G_g are transconductances corresponding to the cross-coupled pair transistors and the supplementary equivalent gate conductance respectively.

$$R_{neg} = -\frac{2}{g_m - G_g}. \tag{5}$$

R_g might be implemented with off MOS transistors this being a more attractive tuning method since no extra current is required. The change in R_g value allows a finer Q factor tuning.

C_{dec} represents the decoupling capacitor and its main role is to decouple the negative resistance, thus reducing to zero the effect on the biasing of the active inductor. All decoupling capacitors are equal to 1 pF.

4. **SIMULATION RESULTS**

Simulations have shown that a “right” tuning (changing $-G_{\text{right}}$) allows higher quality factor values than the “left” tuning (changing $-G_{\text{left}}$) [17]. As mentioned in this reference a larger parasitic capacitance for transistor M_1 determines better performances with “right” tuning. A larger capacitance for M_1 means also larger size for this transistor which adds only benefits for the circuit concerning the noise behaviour. In consequence, all simulations regarding the independent tuning principle used only the “right” tuning.

The principle of independent frequency and quality factor tuning is illustrated in Figs. 5 and 6 where the circuit proposed in Fig. 4 has been studied and the frequency limits of the frequency tunability range are found while sufficiently high quality factor values are obtained.

Several considerations regarding this implementation are in order. Thus, as it has been noticed during the simulations, there is a frequency error given by the introduction of two real negative resistances. This deviation reaches a maximum value of at least 1 GHz compared to the case when ideal negative resistances were used. The parasitic capacitors of the negative resistances (C_{GS}) 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 the right negative resistance is 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.

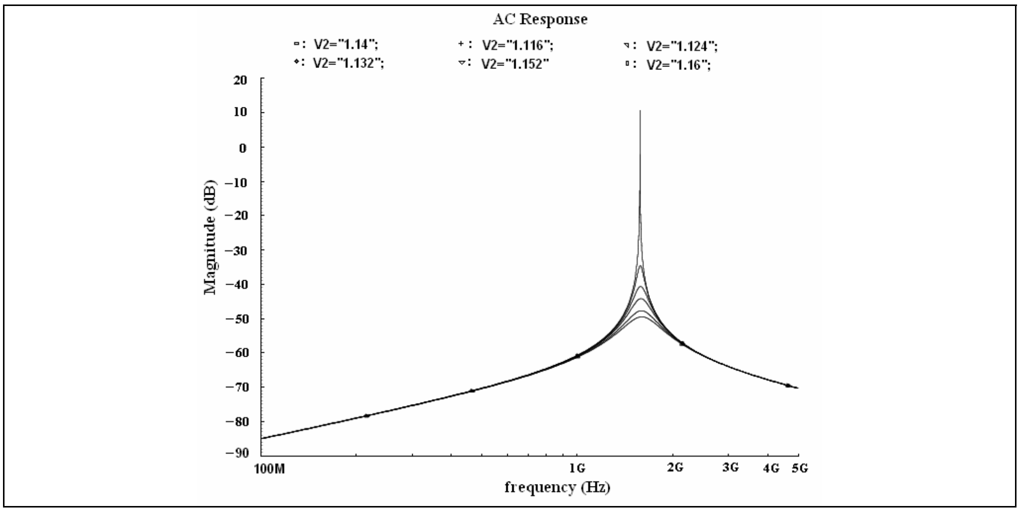

Fig. 5 – Independent tuning of Q, f_0= 1.6 GHz.

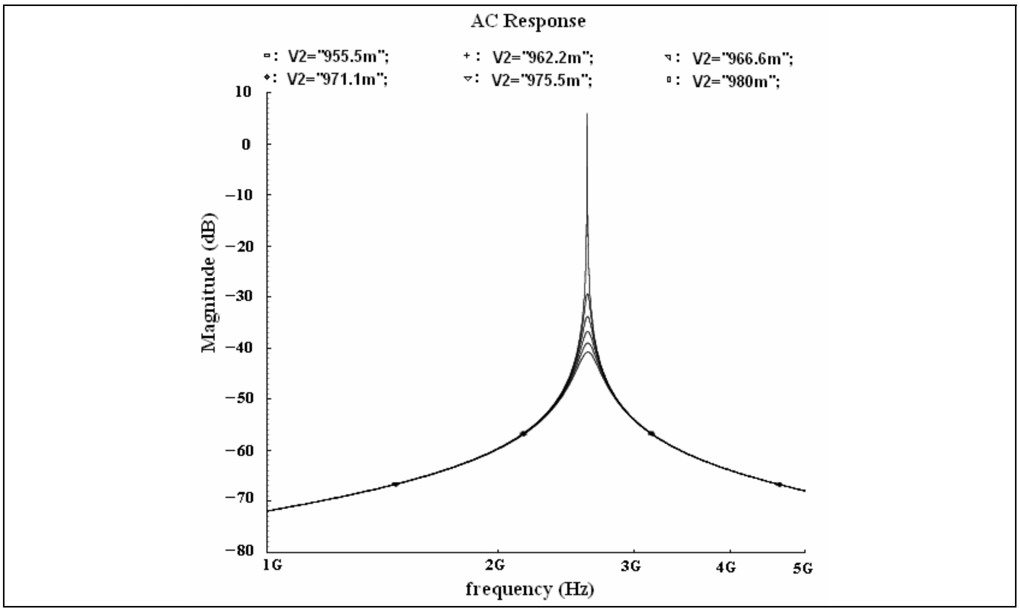

Fig. 6 – Independent tuning of Q, f_0= 2.6 GHz.

By introducing transistors instead of ideal current sources the equality that must be fulfilled for the independent tuning is given by (6):

$$g_{ds1} + g_{ds9} = g_{m7} - (g_{ds13} + g_{ds7} + G_g)\,. \tag{6}$$

Relation (6) shows that with a real circuit the condition of independent tuning is more difficult to implement than in the ideal case. Furthermore, this principle has its limitations since the values of parasitic capacitors are dependent on the voltage bias change through the transistor. Thus, for very poor quality factors a slight change in parasitic capacitors (C_{gs} and C_{gd}) appears since these transistors must support wide voltage biasing changes. This determines frequency deviations up to 10 MHz. However, high quality factor values are intended to be obtained so the parasitic capacitors could be considered constant.

Both negative resistances decrease the self-resonant frequency by their parasitic capacitance, especially the right resistance, the left imposing the frequency limits. This is the reason why any design must begin with the negative resistances since the active filter can be then designed to achieve any central frequency of interest.

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 [16] decreases the frequency tuning range about 30% compared to the ideal case.

Regarding the resistance power consumption, the fixed negative resistance (left negative resistance) has relative high values so that very small current is consumed (several µA). The right negative resistance has relative small values and higher current is consumed consequently.

The power consumption of the circuit depends strongly on the value of the right negative resistance and further on the central frequency. The current consumption obtained for this implementation is 0.16 mW at f_0=1.6 GHz and 0.6 mW at f_0=2.6 GHz, from a 1.8 V supply. The active inductor uses less than 100 µA at 2.6 GHz in order to implement the independent frequency and quality factor tuning.

5. 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 0.18µm 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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