ON THE TUNING PERFORMANCE OF AN ACTIVE RF BANDPASS FILTER

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Abstract: The paper describes a new principle of independent frequency and quality factor tuning suitable for simulated inductors based active RF filters. The method makes use of two dc decoupled negative resistances in order to avoid frequency and quality factor inter–dependency. A differential architecture is studied and several considerations regarding the tuning capability and stability behavior are presented. The target application is a multi–standard active bandpass filter envisaging wireless transceivers.

Key words: *simulated inductor, CMOS RF filters, negative resistance, independent tuning.*

I. INTRODUCTION

On–chip inductors are currently used in practical implementation of LC passive filters [1-3]. However, they suffer from low quality factor values. Even when extra negative resistances are used to improve their performances they still remain unpractical for RF applications.

The practical implementation of band-pass filters for wireless RF is based on SAW filters which are passive, external and large sized and have a very good frequency response [4-5]. However their limitations make them unsuitable for reconfigurable multi–standard systems. Furthermore, in SDR architectures, the problem of implementing a reconfigurable RF filter still remains.

In contrast to passive implementation, active ones gain more interest today due to their performances regarding size, quality factor and tuning possibilities. There are two points of view concerning active implementations described below.

The first way of implementing active filters for telecommunication applications consists in using active gm-C filters. A detailed description of such filters is given in [6] where advantages as well as limitations are presented. Since these filters hardly achieve higher frequencies than 2 GHz, they are not a suitable choice in implementing reconfigurable filters for wireless applications (Bluetooth, WLAN).

The second possibility is based on gyrator theory, consisting in the implementation of transistor only active filters based on simulated inductors. Several configurations of simulated inductors have been proposed in the literature, grounded and floating configurations [7-11]. Since most of them do not allow an independent frequency and quality factor tuning, thus being unsuitable for reconfigurable systems, the implementation of independent tuning is compulsory, a principle being presented in this paper.

Since only one second order band-pass filter does not satisfy the filter requirements for GSM and wireless standards regarding bandwidth and attenuation, higher order structures should be used [12-14]. In any case, extra RF amplifiers may be needed in order to adjust the output power level of each band-pass cell as a certain power is lost due to the presence of buffers and matching networks. However this is the case of conventional transceivers architectures.

This paper presents an RF band-pass filter designed for frequencies up to 2.4 GHz, high enough to cover most wireless standards and allowing independent tuning of the central frequency and quality factor. The circuit could be used to implement higher order active filters for multi–standard applications.

II. THE CORE TOPOLOGY

As already mentioned above, a high interest has been shown to simulated inductors based high frequency filters, most of them envisaging wireless multi–standard transceivers. All structures aim at keeping the number of transistors and sizes as small as possible, in order to avoid parasitic capacitors, low output resistances and noise effects, all these degrading the performances. There are many active inductor architectures presented in the literature, one of them, used in this paper, being shown in Figure 1 together with its equivalent circuit.

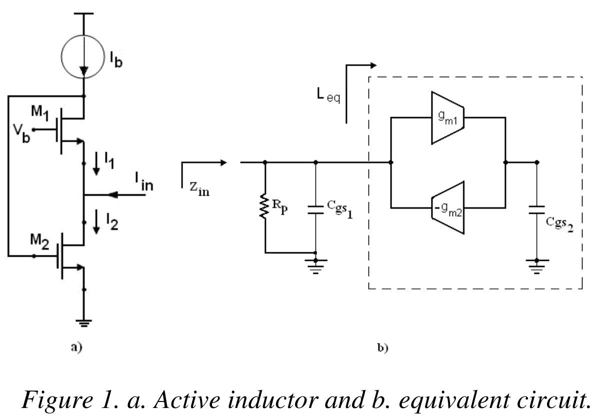

Figure 1. a. Active inductor and b. equivalent circuit.

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The active inductor consists of two NMOS transistors, connected M_1 in common gate and M_2 in common source configuration. The load capacitor of the gyrator which emulates the inductor behavior is C_{gs2} while C_{gs1} represents the parasitic capacitor of the simulated inductor. In the following, only the C_{gs} capacitors will be considered as C_{gd} are too small for being taken into account. Moreover, there is a parallel parasitic resistance whose value depends on the output resistances of the transistors. The small signal equivalent models for the simulated inductor are given in Figure 2.

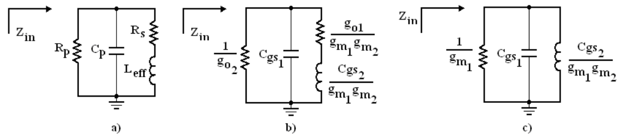

Figure 2. Small signal equivalents with parasitic for the simulated inductors.

As it can be noticed from the figure above, the simulated inductor can be considered as an ideal inductor having the inductance value $L_{eq} \approx \dfrac{C_{gs2}}{g_{m1}g_{m2}}$ with its series parasitic resistance expressed by $r_S = \dfrac{g_{o1}}{g_{m1}g_{m2}}$ all in parallel to a parasitic capacitor C_{gs1} and a positive resistance R_p. Consequently, when the output resistances are negligible, which is the case when small current biasing is used, the circuit approaches very close to an ideal one (Figure 2c). When designing the active inductor, for a certain parasitic capacitor C_{gs2}, fixed by a proper sizing of transistors, the transconductances g_{m1} and g_{m2} are continuously changed in order to obtain the required value for L_{eq}. The equivalent circuit corresponding to an RLC parallel tank, is characterized by a resonance frequency f_0 and a quality factor Q_0, expressed by relations (1) and (2) where the output resistances were neglected for the sake of simplicity.

$$\omega_0^2 = \frac{g_{m1}g_{m2}}{C_{gs1}C_{gs2}} \tag{1}$$

$$Q = \sqrt{\frac{g_{m1}C_{gs2}}{g_{m2}C_{gs1}}} \tag{2}$$

From the circuits shown in Figure 2 and the simplified relations (1) and (2) the influence of the transistor parameters g_{m1}. g_{m2}, C_{gs1}, C_{gs2} on the main parameters of the equivalent circuit are reflected by the following relations where ↑ means "increases" and → means "leads to":

g_{m1}↑ → high f_0, high Q_0, low R_S, low L_{eq}
g_{m2}↑ → high f_0, low Q_0, low R_S, low L_{eq} (3)
C_{gs1}↑ → low f_0, low Q_0
C_{gs2}↑ → low f_0, high Q_0

From the above relations it is apparent that, in order to obtain higher frequency ranges, higher transconductance values are needed, a fact that has a positive influence regarding the series resistance minimization. In order to reach higher Q values only g_{m1} must be increased noticeably. A high product value $g_{m1}g_{m2}$ corresponding to a higher resonance frequency, has a positive effect upon the inductance value since the inductor is supposed to be used for RF applications where small values are required. To further increase the quality factor, C_{gs2} might be increased too, fact that has a negative influence on the inductance value. The output resistances have been neglected in this discussion since the possibility of achieving higher frequencies, up to 3GHz and higher, with small biasing currents, still exists if the transistors are enough saturated.

The parasitic capacitor C_{gs1} is as small as C_{gs2}, i.e., several fF, small enough to be neglected when using this inductor in a parallel configuration in order to design higher orders for the RF band-pass filters. This is the case of pseudo–passive filter implementation, where a capacitance of several pF is used in parallel to the active inductor in order to obtain the desired resonant frequencies for the implemented filter reported in [10] and presented in Figure 3. Higher order filters for RF band-pass filters can be obtained using this method, with very small size compared to passive implementation.

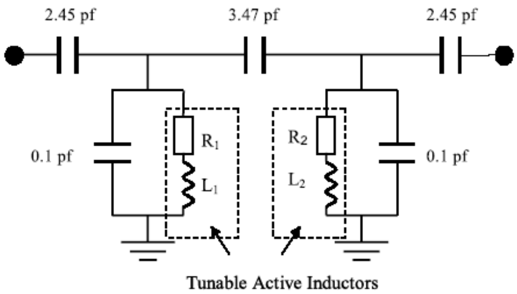

Figure 3. 5th Order CMOS RF Bandpass Filter with simulated inductor [10]

Another method of implementing higher order active RF band-pass filters consists in cascading 2nd order sections, directly or using capacitors, as shown in [12-13].

Regarding the possibility of quality factor improvement using a negative resistance it can be implemented with a cross–coupled pair of transistors and exhibits a small influence on the central frequency of the filter. If the required negative resistance is small enough, for example several kΩ's, higher parasitic capacitances are added to the filter and lower resonance frequencies are obtained. Furthermore, the use of negative resistance is the main reason of implementing differential topologies for the simulated inductors [9], [11–13], since single–ended negative resistances are difficult to implement. Moreover, the differential topology has a better response with respect to noise and linearity.

III. THE INDEPENDENT TUNING PRINCIPLE

In this section, the principle of independent tuning is presented together with a study of the quality factor capability and stability problem.

The principle of independent tuning using negative resistances only has been introduced for the first time in [15] for a single–ended topology and in [16] for a differential one. The method consists in introducing a second negative resistance to the second node and decoupling both of them. In this way, by keeping constant one and varying the second, an independent frequency and quality factor tuning is possible.

Three active inductor topologies with ideal negative resistance are presented in Figure 4. The corresponding input impedances are studied in the followings and presented in relations (4–7) where (4) represents the general form of the transimpedance function. In these relations, C_1 and C_2 are the parasitic capacitors while g_1 and g_2 are the output transistor conductances. The output resistances for transistors are taken into account as well and the possibility of independent tuning is described next.

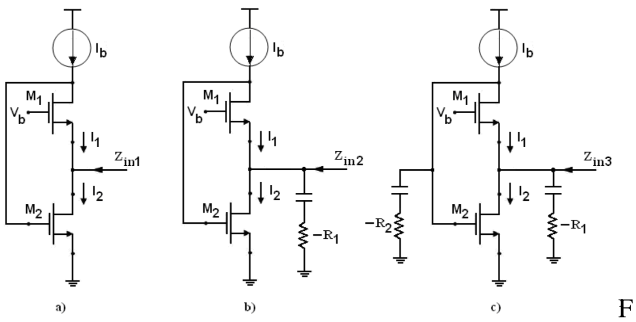

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Figure 4. a. Basic structure, b. Active inductor with negative resistance, c. Active inductor with two negative resistances.

$$Z_{in} = \frac{b_1 s + b_0}{s^2 + 2\alpha s + \omega_0^2} \tag{4}$$

$$\begin{cases} b_1 = 1/C_1; \quad b_0 = \dfrac{g_1}{C_1 C_2} \\ 2\alpha = \dfrac{g_1 C_2 + g_1 C_1 + g_2 C_2 + g_{m1} C_2}{C_1 C_2} \\ \omega_0^2 = \dfrac{g_{m1} g_{m2} + g_1 g_2 + g_1 g_{m2}}{C_1 C_2} \end{cases} \tag{5}$$

$$\begin{cases} b_1 = 1/C_1; \quad b_0 = \dfrac{g_1}{C_1 C_2} \\ 2\alpha = \dfrac{g_1 C_2 + g_1 C_1 + g_2 C_2 + g_{m1} C_2 - G_1 C_2}{C_1 C_2} \\ \omega_0^2 = \dfrac{g_{m1} g_{m2} + g_1 g_2 + g_1 g_{m2} - G_1 g_1}{C_1 C_2} \end{cases} \tag{6}$$

$$\begin{cases} b_1 = 1/C_1; \quad b_0 = \dfrac{g_1 - G_2}{C_1 C_2} \\ 2\alpha = \dfrac{g_1 C_2 + g_1 C_1 + g_2 C_2 + g_{m1} C_2 - G_1 C_2 - G_2 C_1}{C_1 C_2} \\ \omega_0^2 = \dfrac{g_{m1} g_{m2} + g_1 g_2 + g_1 g_{m2} + G_1 G_2 - g_1 G_1 - g_{m1} G_2 - g_2 G_2 - g_1 G_2}{C_1 C_2} \end{cases} \tag{7}$$

From the above relations several conclusions regarding the frequency behavior of the active inductor when using negative resistance can be drawn. These are presented below.

First of all, the second order filter has the central frequency and quality factor dependent on each other. Any change of a transconductance value for changing one parameter will change the second as well.

When comparing relations (5) and (6) we notice a significant decrease of the 3–dB bandwidth when a negative resistance is introduced. Since the parasitic capacitors C_1 and C_2 are almost equal, the term G_1 will compensate the sum $g_{m1}+2g_1+g_2$. It is obvious that by designing the circuit in such manner that C_1 is higher than C_2, a higher value for the negative resistance is required so smaller power consumption is needed. Consequently, a corresponding increase of the quality factor is obtained in this manner. When changing the quality factor, the center frequency decreases too but with a small amount. The problem of changing the center frequency due to the change of the quality factor still remains. Furthermore, in this short analysis, the small signal circuit model was studied. In practice the use of dc coupled negative resistances has a supplementary negative effect on the frequency stability and design stage. In Figure 5 an active band-pass filter based on this inductor topology and using a dc coupled negative resistance is presented [13], together with its frequency response when tuning the quality factor.

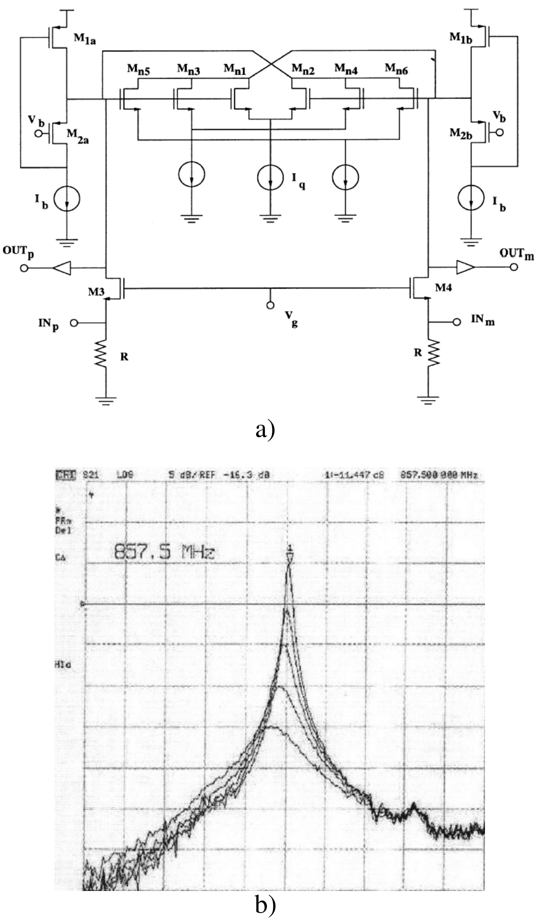


Figure 5. a. Second order CMOS active inductor filter b. Quality factor tuning – measurements for the filter

For the circuit presented in Figure 5a, a change of the negative resistance value will determine a corresponding biasing change for transistors M_{1a} and M_{1b} so that the frequency behavior is actually different of that described by relations (5–7) due to the change of the transconductances in parallel to the negative resistance. The second negative effect is that the circuits need readjustments of biasing voltages and currents after any change for Q tuning. Moreover this limits the maximum frequency tuning range for a certain biasing of the active inductor while keeping the transistors in saturation. There are two solutions to solve this problem: either by finding a suitable configuration in which the negative resistance is connected to the same node as the current source of the active inductor or by decoupling the negative resistance so that the transistors biasing is no more dependent on the negative resistance value. For both cases, the problem of a small frequency deviation still remains. For the decoupling solution, the decoupling means actually the use of two supplementary current sources for the cross–coupled pair, fact which determines a small increase of current consumption in order to obtain the same center frequency. The sacrifice of current consumption (and thus the current–reuse concept) for certain frequency stability is compulsory in order to fulfill the stringent requirements imposed by different telecommunications standards.

From relations (5) and (7) a higher increase of the quality factor can be noticed comparing to the above case this representing a significant advantage. The same effect is noticed on the center frequency which has a larger deviation than the previous case when a single negative resistance was used.

The main advantage of using two decoupled negative resistances consists in the possibility of implementing independent frequency and quality factor tuning. In this case, two possibilities of implementing independent tuning exist, called as "right tuning" and "left tuning":

$$\begin{cases} G_2 = g_1, \quad G_1 = \textit{any value}, \quad \text{"}\textit{right tuning}\text{"} \\ G_1 = g_{m1} + g_1 + g_2, \quad G_2 = \textit{any value}, \quad \text{"}\textit{left tuning}\text{"} \end{cases} \tag{8}$$

Regarding the power consumption, both methods are equivalent. This fact is proved by relations (9) – right tuning and (10) – left tuning.

$$2\alpha\big|_{G_2=g_1} = \frac{g_{m1} + g_1 + g_2 - G_1}{C_1} \tag{9}$$

$$2\alpha\big|_{G_1=g_{m1}+g_1+g_2} = \frac{g_1 - G_2}{C_2} \tag{10}$$

From the stability point of view, in order to keep the filter stable relation (11) below must be always fulfilled. In this case, relation (12) is general true and a fixed power consumption for the negative resistances is determined by the filter biasing.

$$2\alpha > 0 \tag{11}$$

$$G_1 + G_2 < g_{m1} + 2g_1 + g_2 \tag{12}$$

An interesting fact is that by using two negative resistances one can interchange the capacitors in relation (2). This can be helpful in the cases when the parasitic capacitances have different values, in which case, the use of "right" or "left" tuning can assure the highest quality factor value.

A real challenge in this case remains to find the conditions to be fulfilled in order to ensure that by a full–change of the negative resistance, while keeping the other one constant, one can reach the highest possible quality factor with the circuit still stable. This is imposed by the fact there are cases in which, as a function of the filter biasing, the circuit begins to oscillate before reaching the highest Q_0 value. Also a comparison between the quality factor performances for "right" and "left" tuning should be considered. To do this, the quality factor is computed for both cases and is given in (13) and (14).

$$Q_1 = \sqrt{\frac{C_1(g_{m1}g_{m2} + g_1 g_{m2} - g_1 g_2 - g_1^2)}{C_2(g_{m1} + g_1 + g_2 - G_1)^2}} \tag{13}$$

$$Q_2 = \sqrt{\frac{C_2(g_{m1}g_{m2} + g_1 g_{m2} - g_1 g_2 - g_1^2)}{C_1(g_1 - G_2)^2}} \tag{14}$$

From the above relations it is obvious that there are three cases to study.

a) $C_1 = C_2$, in this case the tuning performances are identical for "right" and "left" tuning.
b) $C_1 > C_2$, in this case the "right" tuning offers best tuning performances, the circuit is quite stable while obtaining the best quality factor; using the "left" tuning, the circuit will pass into the instability region, low quality factors being achieved.
c) $C_2 > C_1$, in this case the "left" tuning offers best tuning performances, highest quality factor is achieved while the circuit is quite stable; the "right" tuning is useless.

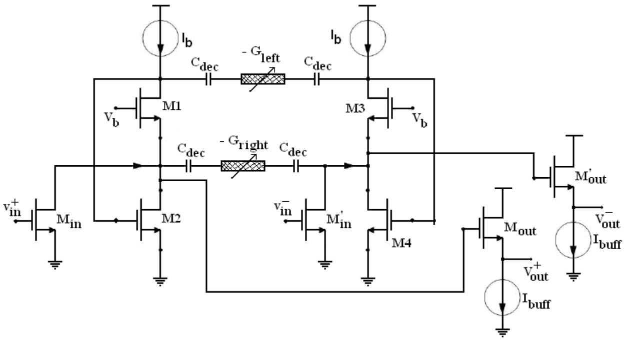
Another problem concerns the technology parameters scatter. In fact, the sizes of the transistors will never be exactly the designed ones. In this case, the use of varactors, connected to both nodes, can be used to achieve the right ratio for parasitic capacitors and thus a good independent tuning being possible. The use of varactors compensate also the impossibility of obtaining any Q value at any frequency due to the change of the ratio, this being an important issue.

The main drawback for this tuning principle consists in the precision required to implement the algorithm. The problem is that of fulfilling with a good precision the constraints imposed by the relations (8). In this case a very well stabilized negative resistance is needed to implement this principle so very good voltage references are required.

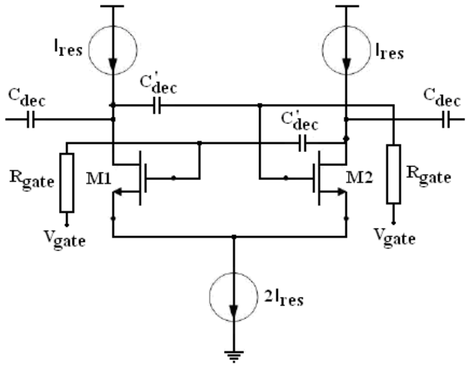
IV. IMPLEMENTATION AND SIMULATION RESULTS

The circuit used in the simulations, based on the above studied active inductor is given in Figure 6a and the negative resistance used for the filter is shown in Figure 6b.

The circuit is implemented as differential configuration, uses two decoupled negative resistances, has two input buffers which convert the input signal in a current signal (most used signals are voltages) and two output buffers to get the output voltage signal. The output buffers have a small influence on the frequency response.



a)



b)

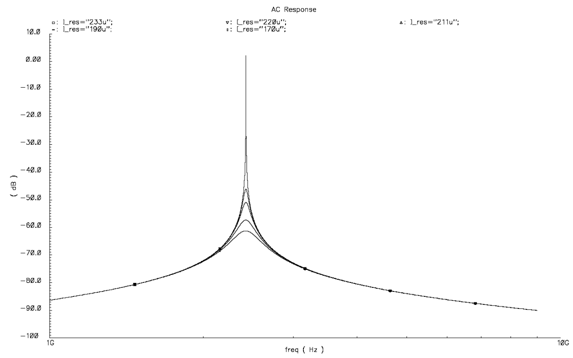
Figure 6 a. Practical implementation of the active RF band-pass filter.b. DC decoupled negative resistance.

Regarding the negative resistance used in implementing the "right" tuning it has very small values (up to $-10k\Omega$) so high transconductances values are needed and biasing current too. The resistances R_{gate} are high enough to decrease the noise introduced by this negative resistance to the filter and can be implemented as transistors working in the triode region or in silicon (passive implementation lead to larger required area).

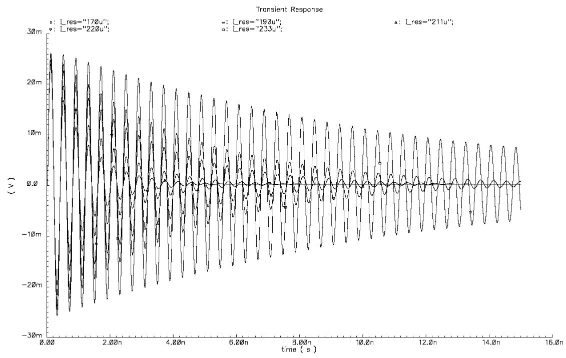
The advantage of using the differential topology over the single–ended one consists in the power consumption requirements. For a differential circuit, the required negative resistances are twice the values required in single–ended application so the transconductances of the cross–coupled transistors are halved and such the power consumption.

The total power consumption depends on the filter biasing, a higher current for active inductor requiring smaller negative resistances for compensation. In general, higher current values are required for higher frequencies ranges.

Two frequencies responses for the filter shown in Figure 6 are presented in Figure 7a and 8a and their transient response in Figure 7b and 8b.

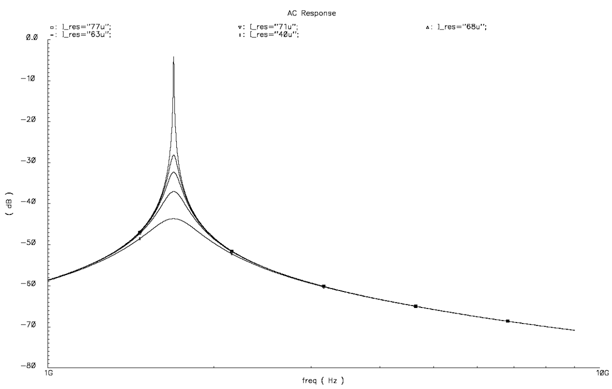


a)

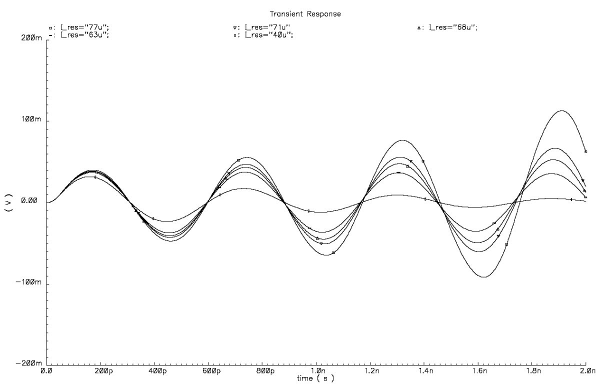


b)

Figure 7 a Quality factor tuning, f_0 = 2.4 GHz, b. Transient response, f_0 = 2.4 GHz



a)



b)

Figure 8 a. Quality factor tuning, f_0 = 1.7 GHz, b. Transient response, f_0 = 1.7 GHz

As it can be seen from Figure 8b, the circuit passes into the unstable regime, even though for higher frequencised the filter was stable. This is an example of the influence of parasitic capacitors on the frequency (low Q values) and transient response (instability). For both cases independent frequency and quality factor are implemented.

The circuit presented above has been designed at 2.4 GHz, can achieve very high quality factors while remaining in the stability region and the total current consumption is less than 0.5 mA from a 1.8 V supply. The active inductor uses approximately 0.1mA to implement the independent frequency and quality factor tuning.

V. CONCLUSION

A new tuning possibility for a differential active RF band-pass filter was presented. A detailed study of the tuning performances and stability was made. The filter based on this principle has been designed in UMC 0.18um technology. The simulations proved the filter capability of achieving high quality factor values with relatively low power consumption. Its small size and possibility of programmability make it suitable for multi–standard applications. Higher orders implementations are to be further investigated.

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