An Integrated SAW-less Narrowband RF Front-end

Mohamed Abouzied, Hattem Osman, Ahmed N. Mohieldin, Ahmed Emira, Ahmed Soliman
Department of Electrical Engineering, Cairo University, Giza 12613, Egypt
Email: aemira@ieee.org

Abstract— **In this paper an integrated SAW-less narrowband RF front-end for direct conversion wireless receivers is presented. The analysis of the feedback system shows a shift of the center frequency f_{RX} for the overall RF bandpass filter from its nominal value f_{LO}. The proposed architecture incorporates a notch filter at $2f_{LO}$ to insure that there is no shift in f_{RX}. The design has been implemented in 65nm CMOS process. It consumes 44mA from a single 1.2V supply. Simulation results show a rejection of more than 15dB in a bandwidth of +/-500MHz around 2GHz due to the additional feedback loop. The theoretical and simulation results are in close agreement.**

I. INTRODUCTION

High frequency RF bandpass filters with narrow fractional bandwidths are required in wireless communication receivers for band selection and image rejection. RF surface acoustic wave (SAW) filters have been widely used for RF band selection due to their accuracy and high selectivity. However, SAW filters are expensive, bulky, and lossy elements. Furthermore, the fact that SAW filters are not programmable makes them unattractive in software defined radios (SDRs) [1]. Several attempts can be found in the literature to eliminate the off-chip passive filters and obtain a fully integrated solution which will reduce power consumption, area, cost, and need of impedance matching. This becomes more important for today's multi-standard and multi-band receivers. Monolithic implementations of active LC filters have been proposed in low cost CMOS technology [2]. Due to the limited quality factor of on-chip inductors, Q-enhancement has been used to design narrowband RF filters. However, this has been plagued with high power consumption, poor selectivity, linearity and noise performance.

The concept of blocker filtering using a receiver translational loop was introduced in [3]. It employs a feed-forward frequency translational loop driven by the receiver local oscillator (RX LO). This technique suffers from limited blocker rejection due to mismatches between two inherently different RF paths which do not track across process and temperature variations. Another blocker filtering technique that uses either a feedforward or a feedback frequency translational path driven by the transmitter local oscillator (TX LO) was introduced in [4]. This technique is only useful for rejecting blockers generated by self transmitter signals in single-chip full duplex systems such as WCDMA. A negative feedback blocker filtering technique using a frequency translational path driven by the receiver local oscillator was introduced in [5]. The rejection in the feedback technique does not depend on gain matching. It has been shown [6] that the feedback technique is more robust to I/Q mismatch effects than its feedforward counterpart. In this paper, the system of the negative feedback blocker cancellation loop is analyzed. It will be analytically proved that the presence of a signal at $2f_{LO}$ after down-conversion results in limited blocker rejection and more seriously, in a significant frequency shift in the closed-loop filter characteristic. It will be shown that tradeoffs between blocker rejection, frequency shift, and stability render this implementation unpractical.

The need for an alternative architecture to implement a monolithic tunable RF filter with high blocker rejection remains as the key towards the realization of a monolithic universal SDR receiver. A new architecture will be presented to decouple the blocker rejection, frequency shift, and stability by placing a notch filter at $2f_{LO}$ at the IF node.

II. SYSTEM ANALYSIS

A. Architecture

The proposed narrowband direct conversion front-end with $2f_{LO}$ notch is shown in Fig. 1. The incoming desired signal is at frequency f_{LO} associated with the blocker signal. Both are amplified by the LNA and then down-converted to baseband by the receiver LO frequency f_{LO}. The LNA is composed of a transconductance g_m and an output LC tank $Z(s)$. The output of the down-conversion mixer contains two frequency components; one at baseband and the other at $2f_{LO}$. A high-pass filter HPF is then used to remove the desired baseband signal component while the blocker passes through. The blocker is then up-converted and subtracted at the output of g_m of the LNA.

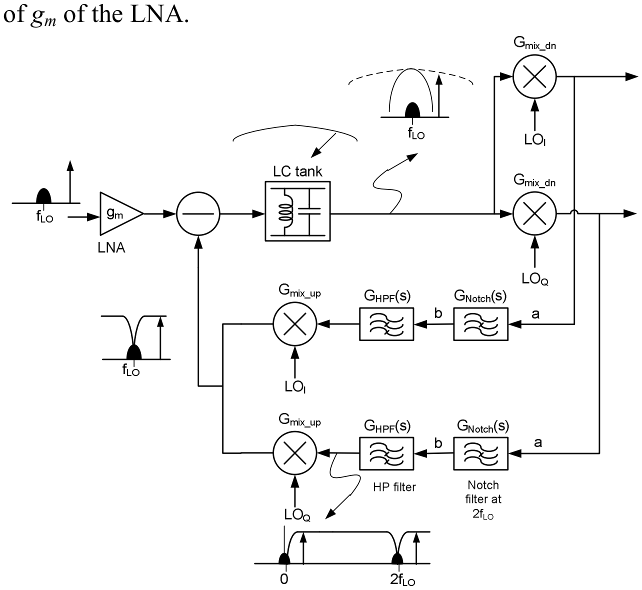

Figure 1 Proposed narrowband RF front-end architecture with $2f_{LO}$ notch

This work has been supported by the Science and Technology Development Fund (STDF), Egypt

Note that the $2f_{LO}$ component will pass through the HPF limiting the gain of the feedback loop for the desired signal, and therefore, limits the rejection. In the system proposed in [5], the $2f_{LO}$ component is rejected by the roll off due to the IF bandwidth f_{IF} at the output of the down-conversion mixer. A small f_{IF} is needed for good rejection at $2f_{LO}$ which results in a limited bandwidth of the feedback loop. Thus loop regulation cannot be maintained after f_{IF} resulting in a poor far-away rejection as shown in Fig. 2, where f_{HP} is the bandwidth of the HPF and $|LG|$ is the open loop gain and is given by:

$$|LG| = |Z(j\omega_{LO})| \times G_{mix\_dn} \times G_{mix\_up} \quad (1)$$

In the feedback loop, the loop gain is almost zero at frequencies close to f_{LO} due to the action of the HPF, thus the desired signal is not affected by the feedback loop and the closed loop gain around f_{LO} is approximately given by $g_{m,LNA}Z(j\omega_{LO})$. On the other hand, as we depart away from f_{LO} the loop gain is set large which scales the closed loop gain by the loop gain and thus effectively rejects the out of band blockers by approximately 20log|1+LG|. Thus the blocker rejection ratio can be increased by increasing the loop gain |LG|. The resulting closed loop response is a narrowband bandpass filter whose -3dB bandwidth equals $2\frac{f_{HP}}{|LG|}$ as shown in Fig. 2.

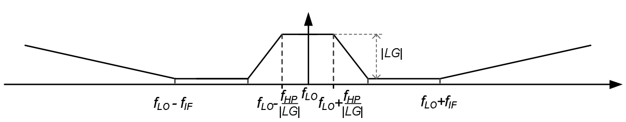

Figure 2 Closed loop filter response

As will be shown shortly, the center frequency of the closed loop response is shifted from f_{LO}, which can cause a loss of attenuation in the band of interest. In our proposal, a notch filter $G_{Notch}(s)$ is used to remove the signal component at $2f_{LO}$. Thus the IF bandwidth can be maximized to avoid the degradation of the far-away rejection. It will be also shown that using the notch filter will avoid the shift of the narrowband RF filter's center frequency.

B. Transfer function

The closed loop transfer function of the narrowband bandpass RF filter is given by:

$$G(s) = g_{m,LNA}\frac{Z(s)}{1 + Z(s) \times H(s)} \quad (2)$$

where H(s) is the transfer function of the translational loop [2]. The impulse response of the translational loop is given by $h(t) = h_{BB}(t).\cos(w_{lo}t)$, where $h_{BB}(t)$ is the impulse response of the baseband filter. That maps to the following frequency response in the s-domain:

$$H(s) = \frac{1}{2}[H_{BB}(s + jw_{lo}) + H_{BB}(s - jw_{lo})] \quad (3)$$

Equation (3) is simply the translation of the baseband HPF to the receiver LO frequency $\pm f_{LO}$. Thus the choice of the baseband filter as a HPF results in an RF filter with notch response around f_{LO}. For a first-order HPF combined with the effect of the finite IF bandwidth at the output of the down-conversion mixer, we have:

$$H_{BB}(s) = \frac{s}{s + \omega_{HP}}.\frac{\omega_{IF}}{s + \omega_{IF}} \quad (4)$$

Thus substituting with (4) in (3), we get:

$$H(s) = \omega_{IF}\frac{(\omega_{HP} + \omega_{IF})(s^2 + \omega_{LO}^2) + s(s^2 + \omega_{LO}^2 + \omega_{IF}\omega_{HP})}{[(s + \omega_{HP})^2 + \omega_{LO}^2].[(s + \omega_{IF})^2 + \omega_{LO}^2]} \quad (5)$$

The center frequency of the resulting bandpass response is the same as the center frequency of the notch filter in the feedback path. Thus it is desirable that H(s) exhibits a deep notch centered at f_{LO} to avoid any in-band attenuation. It should be noted that the choice of a large IF bandwidth, in order to obtain good far-away rejection, will cause a shallow notch response.

The center frequency of the notch response is obtained by a study of the minima of the numerator *N(s)* of the transfer function in (5). It can be shown that the frequency at which $|N(j\omega)|$ is minimum is approximately given by:

$$\omega_{RX} \approx \sqrt{\omega_{LO}^2 + \omega_{IF}\omega_{HP}} \approx \omega_{LO}(1 + \frac{\omega_{IF}\omega_{HP}}{2\omega_{LO}^2}) \quad (6)$$

Thus the notch as well as the closed loop bandpass response is not centered at f_{LO}. In order that this shift does not affect the narrowband filtering, it is required that the gain at f_{LO} does not drop so much from it nominal value at f_{RX}. So f_{LO} must be kept within the pass band of the closed loop response or equivalently:

$$\frac{f_{IF}f_{HP}}{2f_{LO}} \ll \frac{f_{HP}}{|LG|} \quad \Rightarrow \quad f_{IF} \ll \frac{2f_{LO}}{|LG|} \quad (7)$$

Thus the IF bandwidth must be kept small and a Q-enhanced LNA LC tank must be used to account for the far-away rejection degradation [5].

C. 2 f_{LO} notch filter

We propose to include a notch filter at $2f_{LO}$ (between nodes a and b) as shown in Fig. 1. Fig. 3 shows the implementation of the notch filter. It uses a lowpass filter LPF in another frequency translational loop driven by $2f_{LO}$ [7]. This will relax the trade-off between in-band attenuation due to center frequency shift and far-away rejection degradation due to limited IF bandwidth.

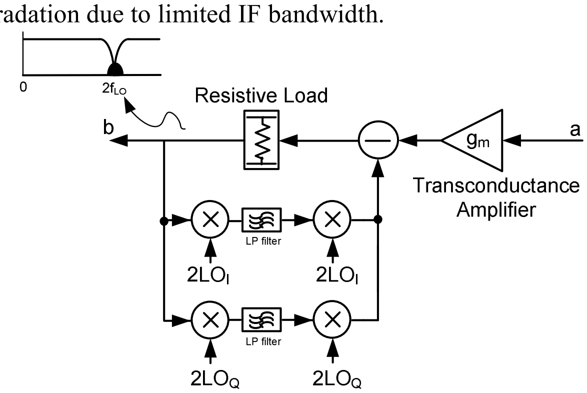

Figure 3 Implementation of the notch filter at 2 f_{LO}

Fig. 4(A) shows the baseband filter response with the notch at $2f_{LO}$. The RF notch filter response is shown in Fig.

4(B). It is clear that the translation of the baseband filter to f_{LO} will result in a notch response around f_{LO}. On the other hand the translation of the baseband filter to -f_{LO} will still result in a notch response around f_{LO} due to the existence of the 2f_{LO} notch.

Thus the design strategy of our system is to increase the IF bandwidth as needed for stability (discussed in next section) without the need of a Q-enhanced LNA LC tank. This will result in a system more suitable for receiver applications with a wide frequency range such as SDR.

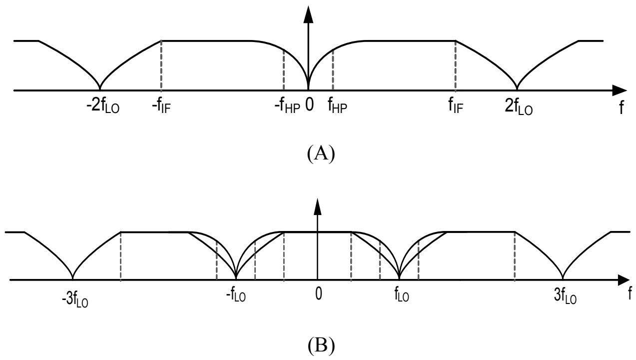

Figure 4 (A) Baseband filter response (B) RF notch filter response

D. Stability analysis

The Bode plot of the open loop gain is presented in Fig. 5. It is assumed that the center frequency of the LC tank is tuned at f_{LO} and the bandwidth of the LC tank represents the dominant pole of the open loop.

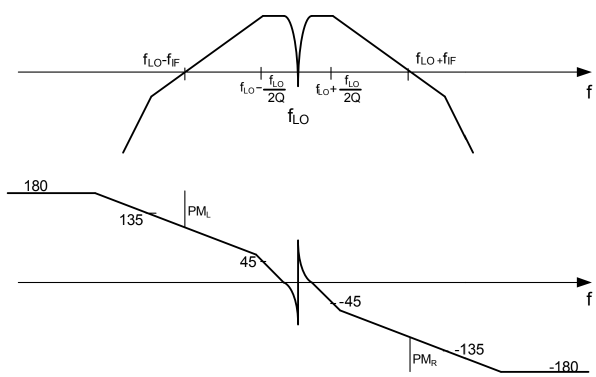

Figure 5 Bode plot of open loop gain

So, in order to achieve phase margin PM better than 45^0, the second pole representing the IF bandwidth must be larger than the gain bandwidth product or

$$f_{IF} > |LG|.\frac{f_{LO}}{2Q} \tag{8}$$

which in turn results in the following condition for the quality factor of the LC tank:

$$|LG|.\frac{f_{LO}}{2Q} < f_{IF} \ll \frac{2f_{LO}}{|LG|} \quad \Rightarrow \quad Q \gg \frac{|LG|^2}{4} \tag{9}$$

Thus to achieve 20dB rejection for the blockers close to the receiver LO frequency, the LG is set to be 20dB at which the quality factor of the LC tank must be much larger than 25 thus a Q-enhancement technique must be used. This will result in degradation in the dynamic range (DR) that is proportional to the quantity of Q-enhancement [8].

Fig. 6 shows the PM versus the IF bandwidth set by the output resistance of the down conversion mixer and the input capacitance of the transconductance cell of the 2*flo* loop (see Fig. 7).

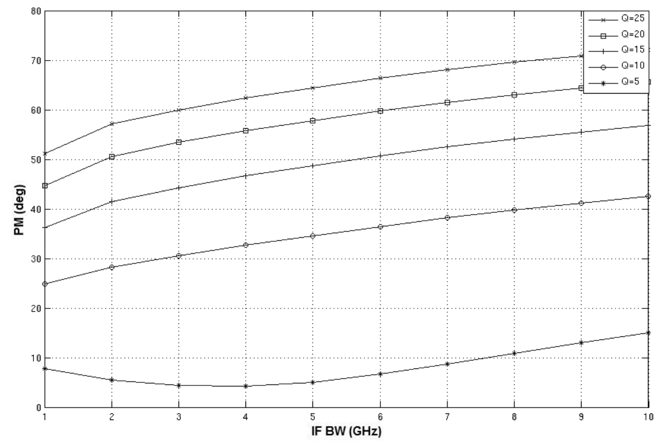

Figure 6 PM versus IF BW for different Q of LNA's LC tank

III. Circuit Implementation

A. Main feedback loop

The design of the integrated SAW-less narrowband RF front-end has been implemented in 65nm CMOS process. The circuit implementation of the main feedback path is shown in Fig. 7.

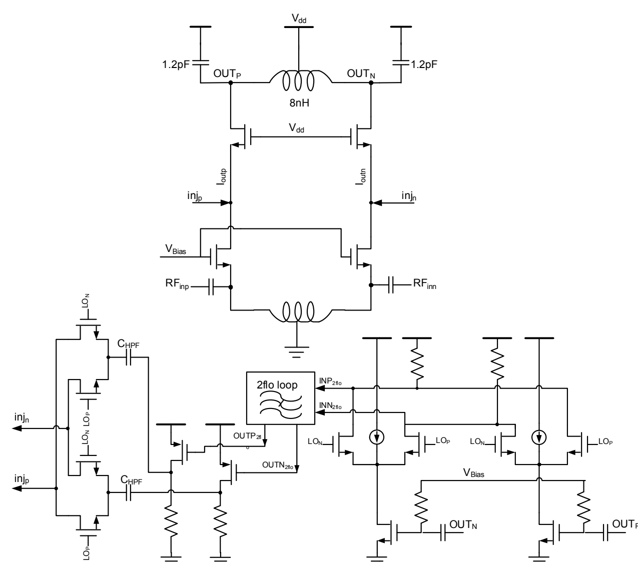

Figure 7 Circuit implementation of the main feedback loop (for clarity, only I branch is shown)

The LNA is a cascode common-gate LNA (CCGLNA) topology for wide-band input impedance matching to the antenna which commonly needs $Z_{in} = 50\Omega$ and for the CCGLNA $Z_{in} \sim 1/g_{m_i}$ $(g_{m_i} = 1/50)$ over a wide-band frequency range. The input voltage is converted to current with the transconductance input stage then; the feedback current from the loop is subtracted from it at the cascode node of the CCGLNA and the resultant current flows in the LC tank of the CCGLNA. The nominal value of the LC tank is set to a center frequency of 2GHz. The quality factor of the LC tank is chosen to be 20. This together with IF bandwidth of 2.5GHz would result in PM of more than 50^0 as shown in Fig. 6. The output voltage at the LC tank is down converted by a current-bleeding Gilbert mixer which is used to decouple the trade-off between output swing and gain.

A transconductance stage is used to obtain a current signal that is AC coupled with a coupling capacitor which acts as a current-mode HPF such that its cut-off frequency is controlled by C_{HPF}. The output current of that HPF is fed to a passive current-mode mixer. The up-converted current signal is fedback to the cascode node and subtracted from the current of the transconductance input stage.

B. $2\, f_{LO}$ notch filter feedback loop

The integrated notch filter that used to reject the up-converted sideband at $2f_{LO}$is shown in Fig. 8. It consists of a wide-band cascode amplifier with the cascode node as the injection node of the feedback current. The output voltage is down converted to baseband, lowpass filtered and the cut-off frequency of the LPF is controlled by C_{LPF}. Then, the up-converted current is injected to the cascode node and subtracted from the output of the transconductance input stage of the wide-band amplifier.

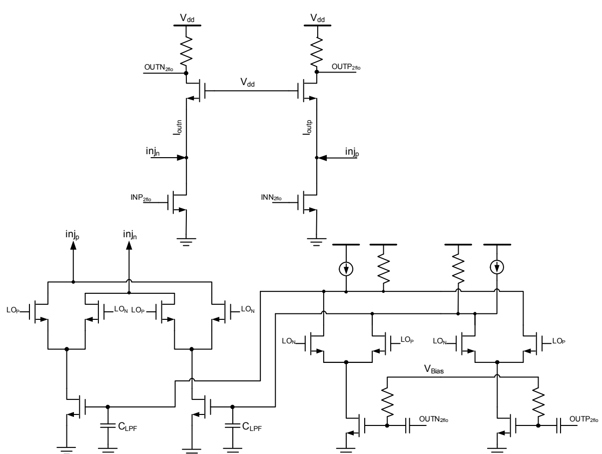

Figure 8 Circuit implementation of the $2f_{LO}$ notch filter (I branch only)

IV. Simulation Results

The closed loop feedback system with the $2f_{LO}$notch filter has been simulated in Spectre RF. Fig. 9 shows the closed loop narrowband bandpass filter. The rejection is more than 15dB in a bandwidth of +/-500MHz around 2GHz. This has been achieved without using Q-enhancement. The noise figure (NF) of the overall system is 6.2dB at 2GHz as shown in Fig. 10. The NF without the feedback loop activated, i.e., LNA and mixer in open loop, is 4.4dB. The degradation in the NF is comparable to the degradation expected due to the losses of a SAW filter. The system consumes 44mA (including the $2f_{LO}$ feedback loop) and operates from a single 1.2V supply.

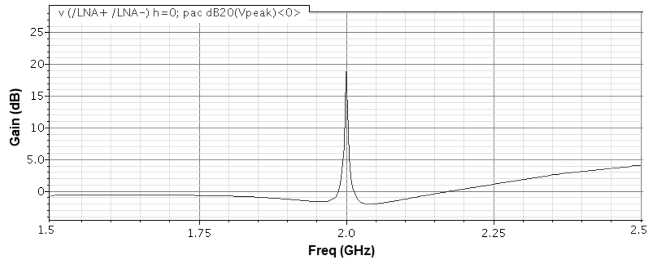

Figure 9 Simulated closed loop frequency response

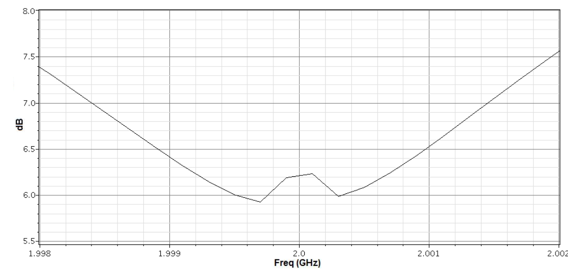

Figure 10 Simulated noise figure

V. Conclusion

A practical implementation of an integrated SAW-less direct conversion receiver front-end has been presented. The proposed architecture is less sensitive to mismatches unlike the feedforward architecture. We have mathematically proved that feedback translational loop results in frequency shifting the closed loop characteristic due to the finite attenuation of the $2f_{LO}$ at the IF node. The proposed architecture places a notch at $2f_{LO}$ at the IF node and therefore resolves the frequency shift issue and relaxes the tradeoff between frequency shift and stability. The proposed architecture has been simulated using TSMC 65nm CMOS process. The front-end consumes 44mA and has 6.2dB noise figure.

References

[1] Asad Abidi, “Evolution of a Software-Defi ned Radio Receiver’s RF Front-End,” Radio Frequency Integrated Circuits (RFIC) Symposium, pp. 27-30, 2006.

[2] A. N. Mohieldin, E. Sánchez-Sinencio, and J. Silva-Martínez, “A 2.7V, 1.8GHz, 4th Order Tunable LC Bandpass Filter Based on Emulation of Magnetically-Coupled Resonators,” IEEE J. Solid-State Circuits, vol. 38, no. 7, pp. 1172-1181, July 2003.

[3] H. Darabi, “A Blocker Filtering Technique for SAW-Less Wireless Receivers,” IEEE J. Solid-State Circuits, vol. 42, no. 12, pp. 2766–2773, December 2007.

[4] A. Safarian, A. Shameli, A. Rofougaran, M. Rofougaran, F. De Flaviis, “Integrated Blocker Filtering RF Front Ends,” Radio Frequency Integrated Circuits (RFIC) Symposium, pp. 13-16, June 2007.

[5] T. D. Werth, C. Schmits, and S. Heinen, “Active Feedback Interference Cancellation in RF Receiver Front-Ends,” Radio Frequency Integrated Circuits (RFIC) Symposium, pp. 379-382, June 2009.

[6] T. D. Werth, C. Schmits, and S. Heinen, “On I/Q-Mismatch in Active Interference Cancellation Schemes,” Midwest Symposium on Circuits and Systems (MWSCAS), pp. 991-994, August 2009.

[7] A. Mirzaei and H. Darabi, “A Low-Power WCDMA Transmitter With an Integrated Notch Filter,” IEEE J. Solid-State Circuits, vol. 43, no. 12, pp. 2868–2881, December 2008.

[8] A. N. Mohieldin, E. Sánchez-Sinencio, and J. Silva-Martínez, “Design Considerations of Bandpass LC Filters for RF Applications,” Proceedings of the IEEE Midwest Symposium on Circuits and Systems (MWSCAS), pp. 73-76, 2002.