A 0.23mm^2 free coil ZigBee receiver based on a bond-wire self-oscillating mixer

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Abstract— **A low-IF very compact low power quadrature receiver for ZigBee applications is presented. The receiver saves area and power with a quadrature self oscillating mixer based on high Q bond-wire inductors. The prototype, integrated in CMOS 90nm, provides 76dB of maximum voltage gain, with a 10dB noise figure, an IIP3 of -13dBm and a phase noise of -124dBc/Hz @ 3.5MHz with an active area of only 0.23mm^2 and a power consumption of 3.6mW (including the baseband complex filter).**

I. INTRODUCTION

Wireless sensor networks and ZigBee systems consist in a spatial distribution of autonomous short-range transceivers to monitor and control environment and/or devices. The large number of units present in the network relaxes the sensitivity of the single receiver but, at the same time, demands a low cost solution to increase the density of elements and thus the system flexibility [1-3]. According to this, the performance of the single transceiver are exchanged with the possibility of designing a long-lasting and cheap device.

In RF front-ends power and area-saving requirements trade-off with each other, since an inductor-free approach results in a cheaper design, while the use of resonant loads can guarantee high power efficiency. This trade-off disappears when integrated inductors are replaced by bond-wires, which guarantee at the same time a high quality factor and a small area. Although bond-wire inductors are not extensively used in large-volume product, due to concerns about their reproducibility, in some case this technique leads to commercially viable solutions (especially in the case of LC oscillators where the inductance spread can be compensated electronically with a sufficient varactor tuning range [4]).

In this work, bond-wire inductors were used in a quadrature self oscillating mixer derived from the LMV cell [5-6], thereby minimizing the active area of the receiver. It will be shown that a very high quality factor offers a more efficient current distribution among the RF building blocks but can increase losses or amplitude/phase mismatches in the front-end transfer function. An analytical description of the phenomenon was developed, resulting in new design optimization to minimize these effects.

II. RX ARCHITECTURE AND BOND-WIRE SELF OSCILLATING MIXER

The starting point for the receiver design is the LMV cell in Fig.1. Low power consumption and small area are obtained sharing the bias current between LNA, Mixer and VCO [5]. The power efficiency of this structure is paid in terms of flexibility since the stacking limits the possibility to optimize the performance of each block. However, when the optimal bias currents of the VCO and the LNA are comparable, this technique is extremely advantageous because matching, RF signal amplification and down-conversion can be realized without any extra power consumption [5].

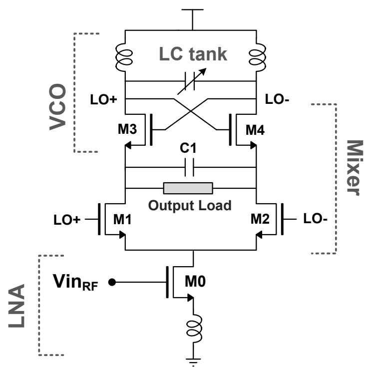

Fig. 1. Current sharing in the LMV cell (bias circuits not shown)

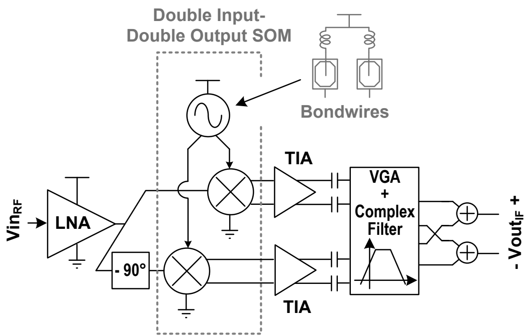

Fig. 2. Proposed receiver with bond-wire double input- double output SOM

When bond-wire inductances are used in the LMV cell instead of integrated coils, there is an immediate reduction of the active area. On the contrary, the total power consumption remains approximately the same, since the bias current cannot be reduced without degrading the LNA performance. In this case the stacking of the LNA appears as a limit in the minimum current consumption, without taking advantages of the use of high-Q inductances in the oscillator tank.

For this reason, the receiver architecture reported in Fig.2 was adopted, where the LNA does not share the bias current with the mixers and the VCO. In this case, the

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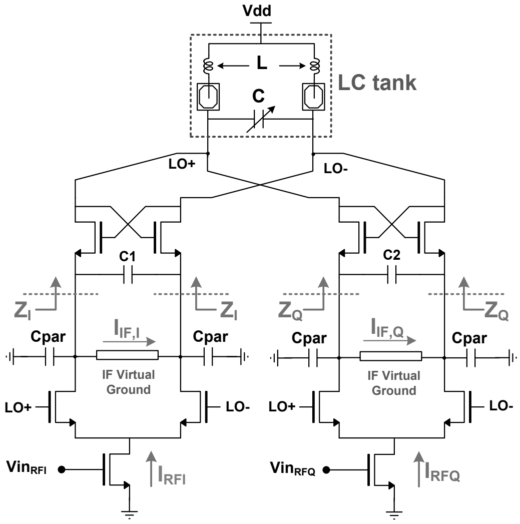

Fig. 3. Bond-wire double input- double output SOM

quadrature is realized on the RF signal path, while the down conversion is performed using the double input-double output self oscillating mixer (SOM) reported in Fig.3. The working principle of this SOM is the same of the LMV cell [5-6] where the oscillation is sustained through a positive feedback closed at RF by capacitor C1 and C2. The RF quadrature signals are injected in the mixer by two transconductors, down-converted and collected into the virtual grounds of two trans-impedance amplifiers (TIA).

The SOM in Fig.3 is particularly suitable to be used with bond-wires because requires only a couple of inductors and the capacitive load at the tank is minimal (merging mixers and VCO), maximizing the tuning range for a given varactor.

III. EFFECT OF HIGH Q TANK ON SOM TRANSFER FUNCTION

The tank sharing realized in the double input-double output SOM reported in Fig.3, introduces an amplitude/phase error between the I and Q paths, proportional to the quality factor of the inductors used. This phenomenon was investigated in order to minimize its effects maximizing the benefits provided by the use of bond-wires.

A. Origin of Mismatches and Losses

The working principle of the double input-double output SOM is identical to the current-mode LMV cell. The main losses derive from the current division at RF between the common mode capacitors C_{par} and the LC tank impedance reflected at the IF outputs (Z_I and Z_Q in Fig.3) [5]. As for the single LMV cell in [5], the Z_I and Z_Q of the double input-double output SOM were evaluated as a function of the common mode and of the differential parts of the tank impedances (Z_{tankCM} and $Z_{tankDIFF}$ in Fig.4):

$$\begin{cases} Z_I(\omega) = \frac{Z_{tankCM}(\omega)}{2}(1-j) \\ Z_Q(\omega) = \frac{Z_{tankCM}(\omega)}{2}(1+j) \end{cases} \quad (1)$$

Since the tank is shared between I and Q paths, the two impedances Z_I and Z_Q are complex conjugated and thus, in addition to a current loss, they produce a phase/amplitude mismatch. However this effect can be minimized

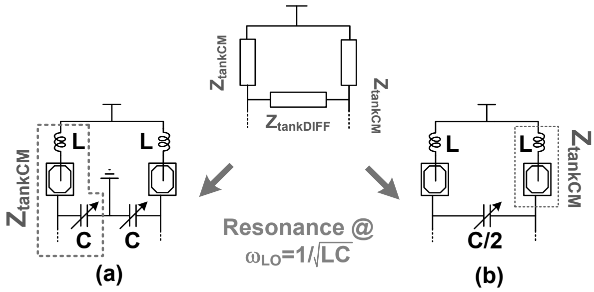

Fig. 4. VCO tank configurations: (a) resonance for common mode and differential signals, (b) resonance only for differential signal

considering that only the common mode component Z_{tankCM} appears at the IF outputs of the SOM.

B. Tank design strategy

In the previous solutions reported in literature [5-6], the use of integrated coils (with relative low Q) has limited to a negligible level the impact of the tank impedance reflection on the LMV cell transfer function. In this case, the use of bond-wires requires a more careful design of the resonant tank to minimize the losses in the presence of such a high Q resonator.

The VCO load has to guarantee a DC path for the bias of the SOM and a differential resonant impedance to set the proper oscillation frequency. Two different configuration, reported in Fig.4, were considered. In the solution of Fig.4.a the tank resonates both for common mode and differential signals and from (1) it has

$$\begin{cases} Z_I(\omega_{LO}) = \frac{\omega_{LO} LQ}{2}(1-j) \\ Z_Q(\omega_{LO}) = \frac{\omega_{LO} LQ}{2}(1+j) \end{cases} \quad (2)$$

where ω_{LO} and Q are the resonance frequency and the quality factor of the resonator.

The configuration in Fig.4.b resonates only for differential components while the common mode impedance at ω_{LO} is given by $Z_{tankCM} \approx j\omega_{LO}L$ leading to

$$\begin{cases} Z_I(\omega_{LO}) = \frac{\omega_{LO} L}{2}(1+j) \\ Z_Q(\omega_{LO}) = \frac{\omega_{LO} L}{2}(j-1) \end{cases} \quad (3)$$

Notice that in this case, the impedances reflected are smaller than (2) and independent from the quality factor.

C. Amplitude and Phase Errors

The two load configurations where compared in terms of amplitude/phase errors introduced in the double input-double output SOM. This was realized evaluating the transfer function for both cases using the same approach proposed in [5]. The down-converted current becomes:

$$\begin{cases} I_{IF,I} = \frac{1}{\pi}\frac{2+(1+j)\omega_{RF} C_{par} Z_{tankCM}(\omega_{RF})}{1+(1+j)\omega_{RF} C_{par} Z_{tankCM}(\omega_{RF})} I_{RFI}(\omega_{RF}-\omega_{LO}) \\ I_{IF,Q} = \frac{1}{\pi}\frac{2+(j-1)\omega_{RF} C_{par} Z_{tankCM}(\omega_{RF})}{1+(j-1)\omega_{RF} C_{par} Z_{tankCM}(\omega_{RF})} I_{RFQ}(\omega_{RF}-\omega_{LO}) \end{cases} \quad (4)$$

where Z_{tankCM} depends on the load configuration used. In particular in the case of a differential resonator, (4) becomes

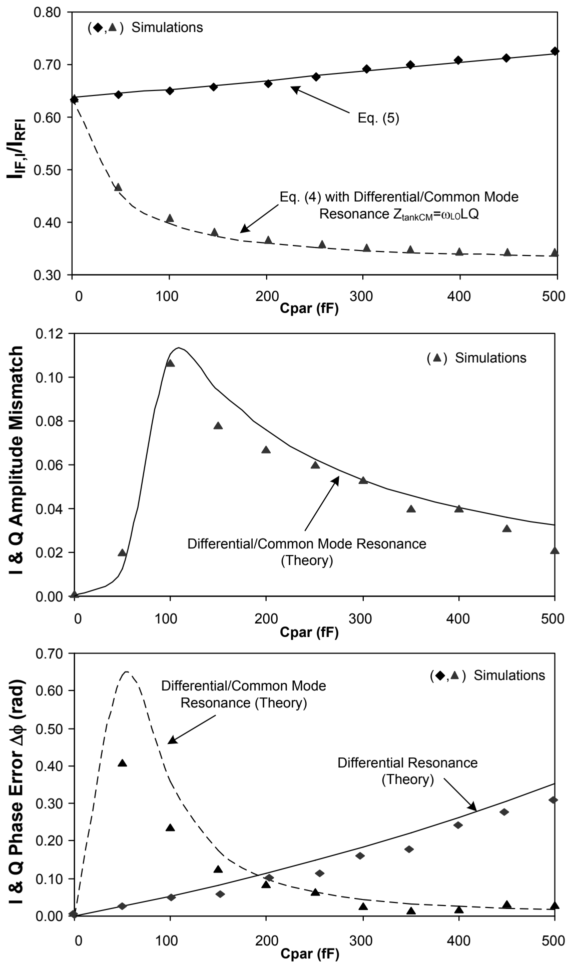

Fig. 5. Effect of VCO tank configuration. Theory vs. Simulations (f_{LO}=2.45GHz, Q=40, L=2nH).

$$\begin{cases} I_{IF,I} = \frac{1}{\pi}\frac{2+(j-1)\omega_{LO}^2 C_{par} L}{1+(j-1)\omega_{LO}^2 C_{par} L}\, I_{RFI}(\omega_{RF}-\omega_{LO}) \\ I_{IF,Q} = \frac{1}{\pi}\frac{2-(j+1)\omega_{LO}^2 C_{par} L}{1-(j+1)\omega_{LO}^2 C_{par} L}\, I_{RFQ}(\omega_{RF}-\omega_{LO}) \end{cases} \tag{5}$$

where the dependency on the quality factor disappears and the amplitude mismatch between the IF outputs is zero. Notice that, since the conversion gain of the single LMV cell in current mode depends on the tank impedance reflection too, it can be improved compared to the solution reported in [5] adopting the differential mode resonator here proposed.

The simulated and calculated gain and amplitude/phase error for input frequencies close to ω_{LO} are reported in Fig.5 and confirm the superior immunity to C_{par} when the tank resonates only differentially. For the differential configuration, the gain can be even greater than $2/\pi$ due to the reactive nature of the impedance reflected at the IF outputs.

IV. RECEIVER DESIGN

The receiver in Fig.2 was tailored to ZigBee application and for this reason a low-IF architecture at 2 MHz was chosen. This approach is particularly suitable for a low-power, low-cost solution, and can guarantee a greater immunity to flicker noise than a direct-conversion one.

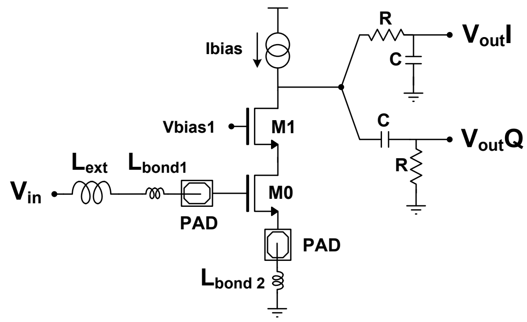

Fig.6 LNA input matching and quadrature generation

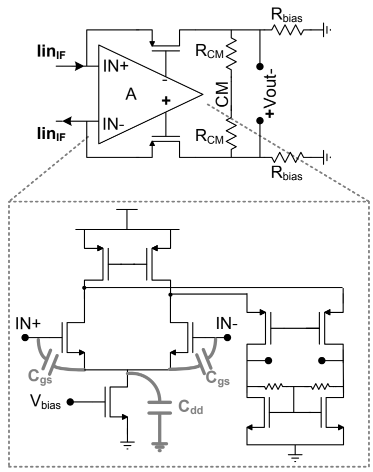

Fig.7 Virtual ground circuit details

After the down-conversion and the first voltage amplification, the signal is AC coupled and filtered by a fully differential three-stage variable gain complex gm-C filter [6]. At the output of the 3rd stage the two paths are finally combined for image rejection.

A. Quadrature Generation and LNA input matching

The low noise amplifier schematic is reported in Fig.6. The input matching is realized through a series resonance and a real impedance synthesized by a bond-wire inductive degeneration (L_{bond2}). Even if a moderate deviation of the real part from the nominal value can guarantee a S11 below -10dB, the high Q resonant network requires an external inductor L_{ext} to center the frequency of operation and to compensate the variation of $L_{bond1,2}$. Moreover, the narrow-band input matching network filters out blockers close to the double of the VCO oscillation frequency, avoiding any injection locking phenomena [5].

Contrary to the previous work [6], the LNA removal from the stack allows to realize a less noisy impedance matching since the quadrature is generated at its output over an RC-CR load. This simple way to generate quadrature is suitable just for narrow band applications, since the 90° phase difference and the amplitude matching are assured only around the cut-off frequency $1/(2\pi RC)$. Due to the relaxed specs of the ZigBee, the amplitude/phase mismatches remain acceptable in all the frequency range required by the standard. The network has to be finally dimensioned trading off between minimum noise contribution and area occupation.

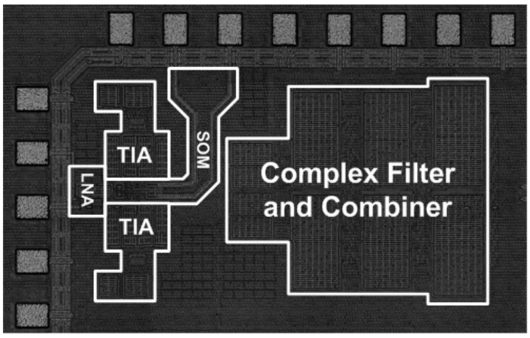

Fig.8. Chip Micrograph

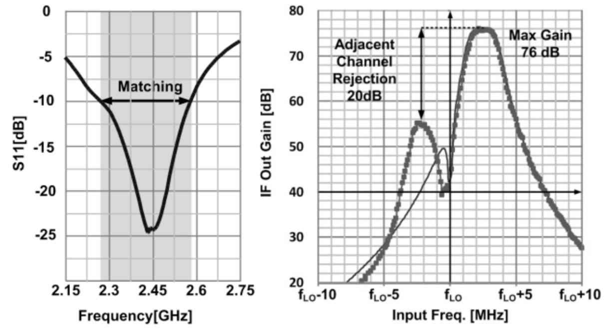

Fig.9. S11 and IF Gain Profile Measurements (f_{LO}=2.45 GHz)

B. *Virtual ground design details*

The virtual ground is provided by a trans-impedance-amplifier in a gain boosted cascode configuration (Fig.7). The differential low impedance, synthesized over a large bandwidth (around 10MHz), limits current losses and ensures high linearity in the presence of large interferers.

In the amplifier design, the most critical element is the input differential pair since its input capacitance contributes to the C_{par} and affects the conversion gain. In particular, using a differential pair input stage, the common mode capacitance is minimized being the series of $2C_{gs}$ and C_{dd} (Fig.7).

V. MEASUREMENTS RESULTS

The ZigBee receiver has been realized in a 90nm CMOS technology. Fig.8 shows the chip micrograph. The use of bond-wire inductors has minimized the area of the RF part (only 0.03 mm^2) making the baseband section dominant (0.20mm^2). The circuits draws 3mA from a power supply of 1.2V.

The external inductor L_{ext} (Fig.6) allows to centre the input resonance, resulting in a good S11 (Fig.9), while the varactor in the tank sets the proper oscillating frequency of the SOM (f_{LO}=2.45 GHz). Fig.9 shows also the IF gain profile with a maximum in band gain of 76dB (from 1MHz to 3MHz). A 20dB image rejection, is obtained without any calibration, giving a safe margin from a target spec of 4dB [2]. The spurious energy at the image frequency is due primarily to the error in the quadrature generation in the RC-CR filter, while the phase shift introduced in the I & Q SOMs was minimized through the use of a differential resonant tank.

In Table I the prototype is compared to the complete ZigBee receivers present in literature. The noise figure averaged over the band from 1MHz to 3MHz is around 10dB while the IIP3 is -13dBm. This results in a spurious free dynamic range of 54.5dB that is comparable to the state of art. Furthermore, compared to the previous solution reported in [6], the total area is 45% less thanks to a reduction of 80% of the RF Front-End part, keeping constant the power consumption.

VI. CONCLUSIONS

In this work bond-wires inductors were successfully introduced in the LMV cell producing very compact self oscillating mixer. A deep analysis on the mechanisms of loss and amplitude/phase mismatches has shown that the use of a differential VCO tank makes the cell conversion gain independent from the resonator quality factor. As a result a very compact receiver for ZigBee application based on bond-wires inductors has been designed minimizing the active area.

TABLE I
MEASUREMENTS AND COMPARISON

	This work	**[1]**	**[2]**	**[6]**
Gain (dB)	76	-	-	75
NF (dB)	10	24.7	5.7	9
IIP3 (dBm)	-13	-4.5	-16	-12.5
SFDR (dB)	**54.5**	50.3	55.3	55.5
PN @ 3.5MHz (dBc/Hz)	-124	-	-	-116
Power diss. (mW)	**3.6**	15	17	3.6
Integrated inductors	**0**	6	4	1
Area (mm^2)	**0.23**	2.1	0.8	0.35
Vdd (V)	1.2	1.8	1.8	1.2
Technology (μm)	0.09	0.18	0.18	0.09

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