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Low Power Ultra Wideband Up-Conversion Mixer in CMOS 0.13 μm Technology

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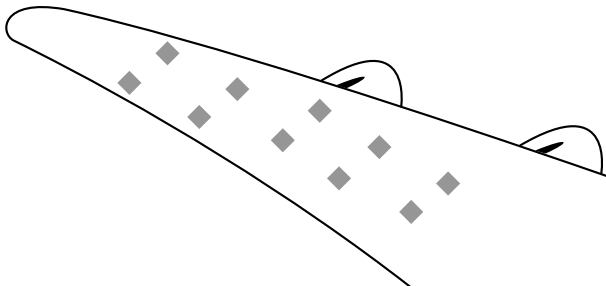
Abstract—This paper presents a low power, UWB up-conversion mixer, designed on CMOS 0.13 μm technology. This circuit is based on a double balanced Gilbert topology, which uses a feedback loop in its transconductance stage; in order to match the IF input impedance over a 10-510 MHz frequency range. This mixer exhibits a 0 dB conversion gain, a 7 dB noise factor, a DC power in the range of 4.2 mW and can output an RF signal over the 6-8.5 GHz frequency range.

I. INTRODUCTION

This paper reports the design of a low power UWB up-conversion mixer, operating in the ISM UWB frequencies bands from 6 up to 8.5 GHz [1]. This work contributes at the design of an UWB-OFDM transceiver for wireless sensor network (WSN) for instrumentation measurements.



(a)



(b)

Figure 1: WSN applications for in flight aircrafts testing and ground satellite testing

This wireless sensor network intend to measure pressure and temperature in real time for aero-space applications [2]. There are two main applications: satellite structure test (Figure 1.a) and in flight aircraft testing (Figure 1.b).

The first application is the ground test of the satellite structure before launching by using strain gauges or thermal sensors. For the second application, the in flight aircraft testing, the most critical test is the measurement of pressure around the wings to validate the designed computer model before industrialization,

This wireless sensor network has to replace the wired system used in theses applications. The wired sensor measurement system presents strong drawbacks in term of weight and cost especially for aircraft due to aircraft immobilization for cable routing.

The main features of the new wireless sensor network are:

- high number of measurements points (more than 400 points); so an significantly increase in the number of used sensors
- High data rate
- Low error rate

The most challenge for this system is the high data rate, thus the nodes are self-powered and cannot introduce sleeping mode to minimize power consumption. To respond to this challenge we propose an UWB transmitter using MB-OFDM [2] and an RF front end between 6 GHz and 8.5 GHz. The architecture of the RF front-end transceiver is described in Figure 2.

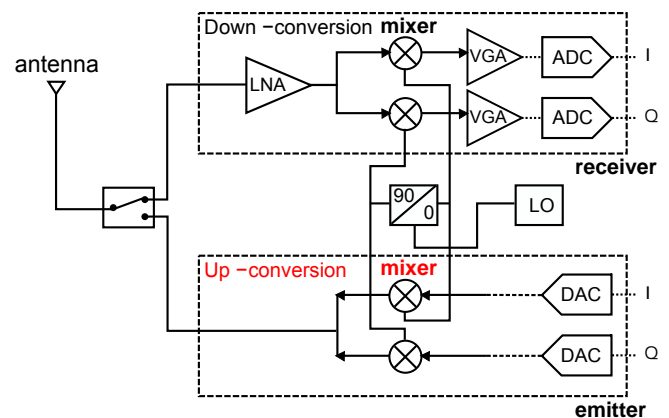


Figure 2: Transceiver Architecture

Transmitter (Tx) and receiver (Rx) channels are timely multiplexed by using an RF switch. Tx and Rx channels are driven by one quadrate local frequency synthesizer. The down-conversion mixers and VGA (variable gain amplifier) will provide I and Q base-band signals on Rx. Up-conversion mixers will provide RF signal on Tx channel, without the using of a power amplifier. Due to the usage of OFDM modulation technique, base-band signals spectrums of this transceiver, will be spread from 10 to 510 MHz.

The design of transceiver, with such a value of bandwidth, requires novel topologies of RF circuits in order to reach a high transmission quality (low distortion), which is a critical point of high speed telecommunications. In such an OFDM transmitter the up-conversion mixers, must be able to shift a 10 MHz up to 510 MHz signal around the 6 GHz to 8.5 GHz microwaves frequencies, with a low distortion. Due to components parasitic, and thus impedance mismatching effect, this condition is not obvious over a large frequency.

This paper reports the design of such an up-conversion mixer, which can handle OFDM signals without these distortion effects by using a simple feedback loop device.

II. CIRCUIT DESIGN

The base-band signal, to transmit over the RF channel, will be provided by DAC device (see Figure 2). Due to parasitic behavior, capacitive effects appear on input impedance of transistor, which are responsible of impedance mismatching at high frequencies. Such phenomenon will be encountered on the design of up-conversion mixers requiring a very high bandwidth. This will result in amplitude and phase distortion of base-band signal. The amplitude distortion phenomenon will decrease the signal to noise ratio, on high frequencies of base-band signal's spectrum. This phenomenon can not be removed by digital filtering. This will result in an increase of mixer noise factor and thus a degradation of radio budget link. For these reasons, it's important to reach good impedance matching on IF (intermediate frequency) input of an up-conversion mixer, for an efficient OFDM transmission.

Due to the important value of bandwidth required on IF mixer's input (10-510 MHz), the usage of impedance matching devices based on lumped elements, is not possible. Our investigations on mixer design have shown that this problem can be overcome by using a feedback device in order to obtain good impedance matching over a large frequency band.

The operating principle and the quantitative analysis of this feedback device are discussed in the following paragraphs. The Figure 3 shows a simplified schematic and its electrical small signal model of the transconductance stage of the mixer with the feedback device.

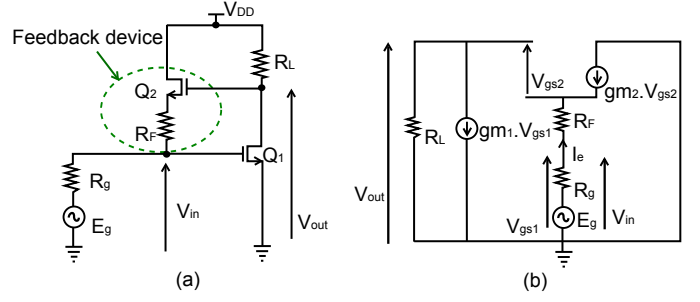


Figure 3: (a) Simplified schematic of the transconductance stage, (b) Small signal model of the transconductance stage

From Figure 3, we can extract the gain formula of the transconductance stage. This formula is reported below:

$$Av = K.H(j\omega)$$

With K , which is a constant factor against frequencies variations done by the following equation:

$$K = -gm_{01}.R_L \cdot \frac{1 + R_F.gm_{02}}{1 + R_F.gm_{02} + [gm_{02}.Rg(1 + R_L.gm_{01})]}$$

And H , which is the frequency transfer function :

$$H(j\omega) = \frac{\left(1 + j \frac{\omega}{\omega_{02} \cdot (1 + R_F.gm_{02})}\right)}{1 + j \frac{\omega}{\omega_{c1}} + j \frac{\omega}{\omega_{c2}} + j^2 \frac{\omega^2}{\omega_{c3}^2}}$$

gm_{01} and gm_{02} , are respectively Q_1 and Q_2 MOS transconductance. K is the static gain and ω_{c1} , ω_{c2} , ω_{c3} are the cut-off frequencies. The input impedance formula is given by:

$$Z_{in} = \frac{Eg}{Ie} = Rg + Z_{in1}, \quad Z_{in1} = K.H(j\omega) \quad (1)$$

with:

$$K = \frac{1}{gm_{02}} \cdot \left(\frac{1 + R_F.gm_{02}}{1 + R_L.gm_{01}} \right) \text{ and } H(j\omega) = \left(1 + j \frac{\omega}{\omega_{01}} \right)$$

From equation (1), we can extract an equivalent circuit of mixer's input stage and its input impedance. The input impedance of mixer can be modeled by a resistor R_0 in series with an inductance L_0 .

$$R_0 = \frac{1}{gm_{02}} \cdot \left(\frac{1 + R_F.gm_{02}}{1 + R_L.gm_{01}} \right) \text{ and } L_0 = \frac{R_0}{\omega_{01}}$$

From these formulas, real and imaginary parts of input impedance can be simplified to the following expressions:

$$R \approx \frac{1}{gm_{02}} \text{ and } L \approx \frac{1}{\omega_{01}gm_{02}}$$

R depends on the transconductance value (gm_{02}) of MOS devices used in the mixer's feedback loop. L depends on the cut-off frequency ω_{01} of MOS devices used in the transconductance input stage of mixer (Q1 transistor). To get 50- Ω value input impedance, we have to fix the bias current of MOS devices (Q2 transistor) used in the feedback, in the way to obtain:

$$\frac{1}{gm_{02}} = 50\Omega$$

In the same time, the bias current of the transconductance transistor (Q1 transistor) has to have a high cut-off frequency ω_{01} to decrease the inductance effect.

This transconductance stage, with its feedback device, drives a switching stage in order to mix the IF and local oscillator (LO) signals. The Figure 4 depicts the topology of our mixer. The mixer is supplied by a 1.8 V voltage and shifts an OFDM base-band signal (10-510 MHz) from DC toward the ISM frequencies bands (6-8.5 GHz). This mixer was designed in a 0.13 μ m CMOS technology. Its MOS devices exhibit a cut-off frequency in the range of 110 GHz.

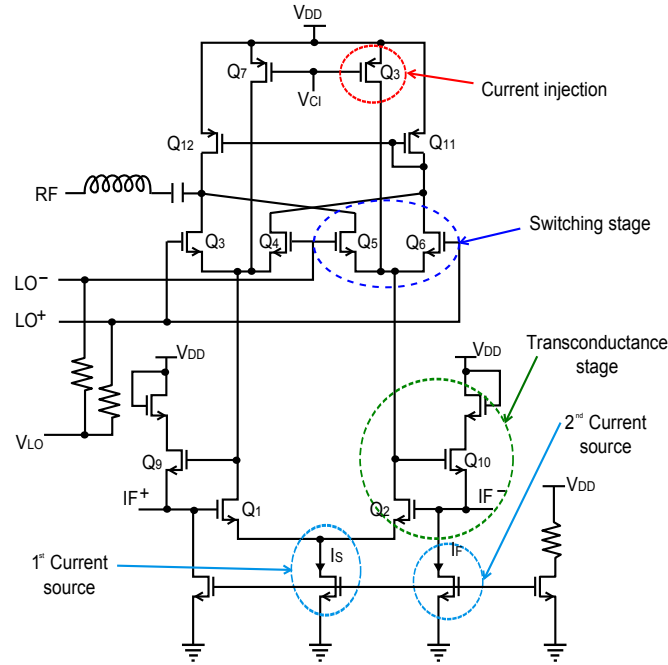


Figure 4: Schematic of the mixer

A transconductance stage (Q1-Q2) drives the switching stage (Q3-Q8) by using an IF differential input signal. A feedback loop (Q9-Q10) is added to the transconductance stage in order to match input impedance over a 10-510 MHz frequency range.

Transistors Q7 and Q8 are used as current injection source to increase conversion gain [3]. The principle is described in Figure 5. As shown, a current injection source (I_{CR}) is added to the basic core of the mixer, at the drain of Q1 and Q2.

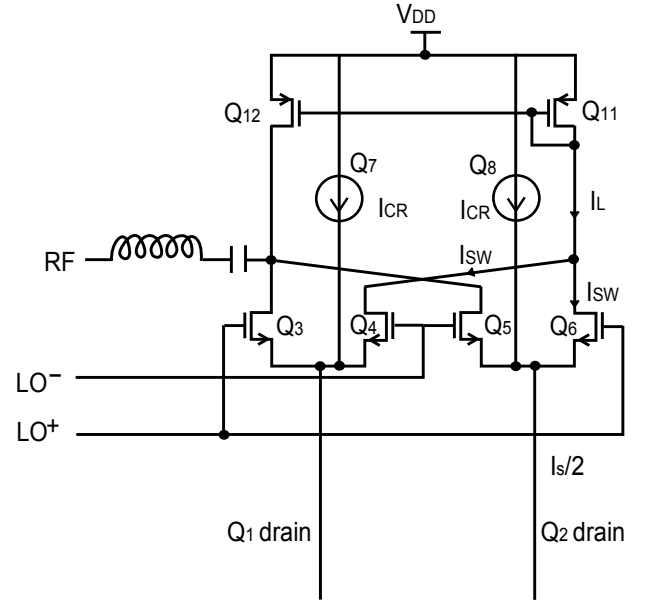


Figure 5: Schematic of the mixer

Using the current injection source, allow us to increase the biasing current of Q1 and Q2 transistors without modifying the drain current of the switching stage (Q3, Q4, Q5, Q6 transistors). This will improve linearity and conversion gain of the mixer at the same time.

The simulations of electrical behavior of this mixer, performed with Cadence Spectre software, are depicted on the next section of this paper.

III. SIMULATIONS RESULTS

The simulations conditions of this investigation are the following: the IF signal power is set to -20 dBm and LO power to -3 dBm.

The Figure 6 depicts the input return loss on IF input of mixer. Figure 7 reports the conversion gain versus power level of local oscillator, at a 6 GHz frequency. The conversion gain reaches a maximum of 0 dB for a local oscillator power between -4 and -3 dBm. We have chosen to work at a -3 dBm level, concerning the LO signal, for DC power consumption reasons, which are very important in our application.

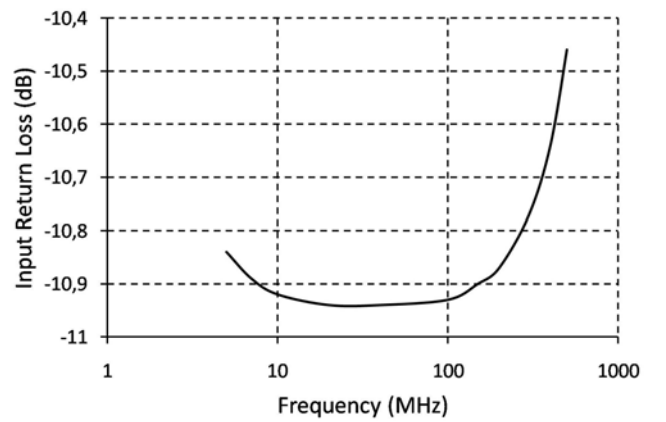


Figure 6: The input return loss

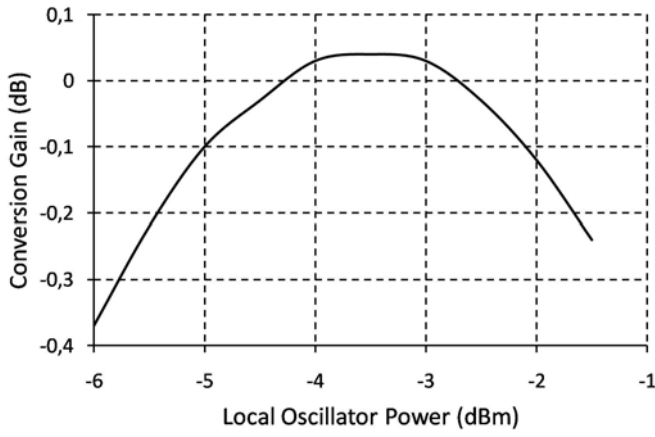


Figure 7: Conversion gain versus LO power at 6.5 GHz frequency with IF power of -20 dBm.

Reducing the required power level of the local signal to apply on a mixer is equivalent to reduce the required DC power by the local oscillator (a VCO), which will drive the mixer. This point is important for the development of a low DC power consumption OFDM transceiver.

We have simulated the conversion gain versus the RF frequency. This result is reported on Figure 8. During this simulation, the power level of the local oscillator was maintained to -3 dBm and the IF signal frequency was fixed to 500 MHz.

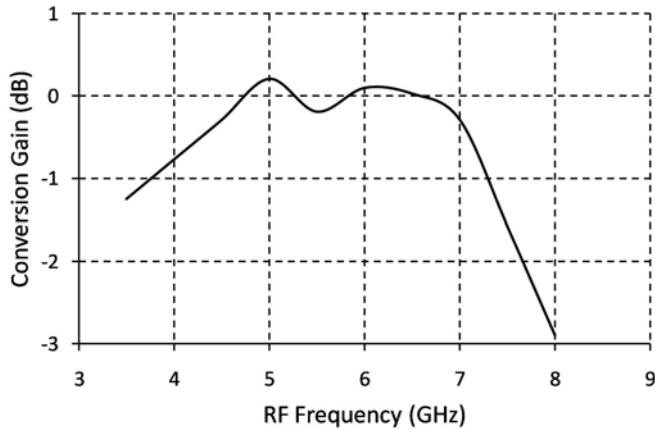


Figure 8: Conversion gain versus RF frequency with IF frequency of 500 MHz and IF power of -20 dBm, the LO power is -3 dBm.

Figure 9 report the simulation results of the noise factor of our mixer. The noise factor level is around of 7 dB.

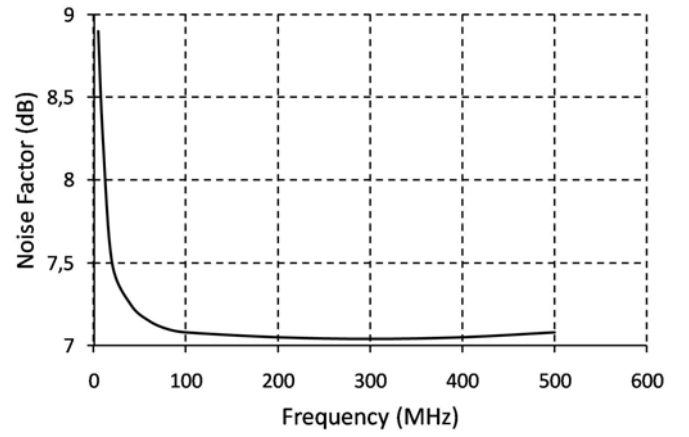


Figure 9: Noise factor in dB

In order to compare our results to the state of the art, we have reported several previously published data in the field of CMOS up-conversion mixers. These data are reported on table 1 as well as data of our design.

Because the available designs, don't have common characteristics (conversion gain, bandwidth, operating frequencies), we have used a Figure of merit in order to be able to compare each design with the others. Such a factor has been already defined in [4] described below

$$FoM(dB) = \frac{CG[dB].BW[GHz]}{P_{DC}[mW]}$$

In this factor of merit, the definition area is limited to positive values of conversion gain and the required power level of local signal, for efficient mixing, is not taken into account. For these reasons, we propose a modified definition of this Figure of merit. This new definition is given below:

$$FoM(dB) = 10 \log \left(\frac{10^{(Gc[dB]/20)} \cdot \frac{BW[GHz]}{1GHz}}{\frac{P_{DC}[mW]}{1mW} \cdot \frac{10^{(P_{Lo}[dBm]/10)}}{1mW}} \right)$$

where Gc is the conversion gain in dB, BW is the bandwidth of the RF output.

In Figure 10, we present the Figure of merit, versus the DC power consumption; for our work and all of the mixers listed on table 1. We can note that our mixer exhibits a good trade-off between DC power consumption and conversion gain. Our design has a DC power consumption of 4.2 mW and a conversion gain in the range of 0 dB.

	this work	[5]	[6]	[7]	[8]	[9]
Grid length (μm)	0.13	0.18	0.18	0.18	0.13	0.13
Die size (mm^2)	0.64	0,3	1,96*	0.69	0,47	3,16**
Conversion Gain (dB)	0.03	-6	-8	6.5	-2 – 0.7	1.1
Bandwidth (GHz)	6 – 8	1 – 11	3 – 5	3 – 5	22 – 29	3 – 11
Noise Factor (dB)	7	-	-	12-13	-	-
LO Power (dBm)	-3	3	-5	0	3	0
Supply (V)	1.8	1.8	1.8	1.0	1.2	1.2
DC Power (mW)	4.2	25	12.96	11	8	9.7
FoM	29.74	20.02	22.88	25.85	26.77	29.71
Topology	Gilbert	Folded - Inductor less	Gilbert	Gilbert	Dual gate	Gilbert

* Dual band

** Full transmitter

Table 1: State of art of up-conversion mixers

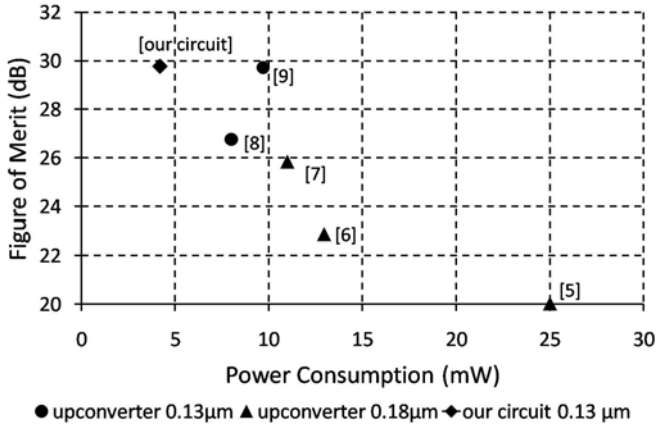


Figure 10: Figure of merit versus power consumption for our work and the state of the art samples

Concerning the power handling capabilities, our mixer is matched to the condition of our targeted architecture (not use of power amplifier in our transmitter architecture). Thus, our mixer must be able to handle the required power to transmit information over the RF channel.

The FCC allowed a maximum power spectral density radiation of -41.3dBm/MHz [10], at the antenna level. For a 500 MHz bandwidth, the maximum transmit-power is:

$$P_{\text{EMAX}} = -41.3 + 10 \log(500) = -14.3 \text{ dBm}$$

Our circuit exhibits an IIP1 in the range of -11.6 dBm, which satisfies the FCC condition, if we plan to connect directly antenna filter and multiplexing switch on our mixer's output.

IV. LAYOUT DESIGN

The layout of the mixer without the ground plane is presented in Figure 11. DRC test and LVS are used to check our circuit and we performed the back-annotation in order to include the extra interconnect parasitic elements.

The mixer is designed using UMC 0.13 μm RF CMOS process provided by United Microelectronics Corporation (UMC) The die size including RF-pads is 0.64 mm^2 .

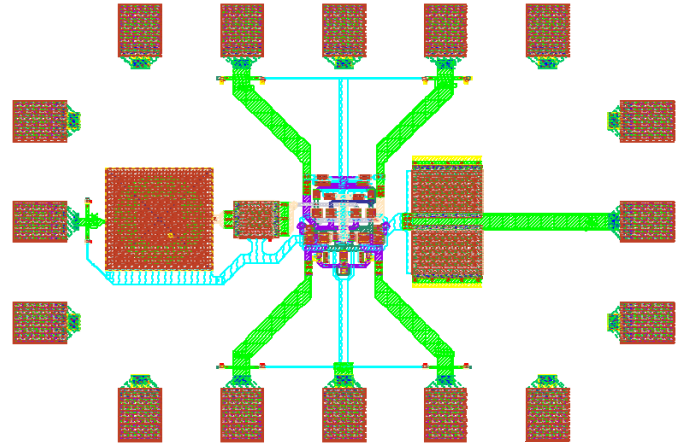


Figure 11: Layout view of the mixer without ground plane

V. CONCLUSIONS

In conclusion, we have presented a low DC power, ultra wide band, up-conversion mixer. This mixer consumes only 2.34 mA current under a DC voltage of 1.8 V, which is the best result in the state of the art. A new topology of mixer, based on the Gilbert structure, was proposed here to obtain a good input impedance matching condition over a large frequency range.

This condition is reached at a very low cost: the proposed feedback device requires a very low silicon surface and exhibits a very low DC power consumption. This matching impedance condition is important in order to remove any amplitude distortion effects, which induces a degradation of the signal to noise ratio.

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