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A Weak-Inversion Low-Power Active Mixer for 2.4 GHz ISM Band Applications

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Abstract—This work describes a fully integrated active mixer designed in 0.18 μm CMOS technology for 2.4 GHz Industrial, Scientific and Medical (ISM) band applications. The mixer uses a source driven local oscillator (LO) to integrate the RF-LO buffering with the active mixer, and weakly inverted input MOSFETs to improve the gain and noise performance. The double-balanced quadrature mixer has a measured conversion gain of 32 dB with a double side-band (DSB) noise figure (NF) of 8.5 dB at 30 MHz intermediate frequency (IF) while consuming only 0.56 mA of current from a 1.8 V supply.

Index Terms—CMOS integrated circuits, front-end, monolithic microwave integrated circuit (MMIC) mixers, subthreshold.

I. INTRODUCTION

DOWN-CONVERSION mixers are a key building block in any modern wireless receiver. Typically, active mixers are favored over their passive counterparts for their superior conversion gain and port to port isolation. Unfortunately, active mixers generally consume a significant portion of the receiver's overall power. Furthermore, active mixers introduce flicker noise into the receiver that can degrade the systems dynamic range. In general, if a transistor's device length is kept constant, its input-referred flicker noise is inversely related to power consumption [1]. This makes it difficult to realize a low-power active mixer with a sufficiently low flicker noise corner frequency.

In this letter, we present a CMOS active mixer that uses a source-driven local oscillator (LO) that doubles as an LO buffer to save significant power consumption. The RF input transistors are sized so that they operate in the weak inversion region when conducting current. This results in an improvement in their transconductance [2]. High conversion gain is achieved by using active PMOS loads as opposed to resistive loads. Coupled with the matched input impedance, the high conversion gain allows the design to be used without an additional low-noise amplifier (LNA). Details of the design are presented in Section II. Measured results and comparison with state-of-the-art active mixers are presented in Section III.

II. PROPOSED ACTIVE MIXER DESIGN

A. LO Buffering

An LO buffer normally consists of a common-source differential pair with inductive loads [3], [4]. Such a buffer could con-

Fig. 1. Current reuse active mixer and LO buffer including both I and Q paths.

sume current in the range of a milliamp [5], which proves costly when considering low-power receivers which can use a total current consumption of less than 10 mA [4]–[7]. An LO buffer helps both in increasing the drive strength of the LO signal and also in providing necessary isolation between the RF signal path and the PLL. RF-LO leakage can lead to VCO pulling [8].

In this work, the LO buffering has been combined with the active mixer in a current reuse style. The schematic of the fully-balanced IQ mixer is shown in Fig. 1. The LO buffer consists of common source transistors $N_9 - N_{12}$. $N_1 - N_8$ form the switching transistors. Because a transistor's g_m only depends weakly on the drain source voltage for a MOSFET, LO buffering can be achieved without using too much voltage headroom. Furthermore, this topology obviates the need for large on-chip inductors normally used in LO buffers.

A Gilbert Cell provides some RF-LO isolation due to the balanced architecture. However, any mismatch in device sizes can cause RF-LO leakage through the gate source capacitances of the switching transistors. In the proposed design, the balance of the switching stage provides some isolation, and additional isolation is provided by the LO buffering transistors. The RF-LO leakage paths for the proposed mixer and a typical Gilbert Cell mixer are illustrated in Fig. 2. For the proposed design, N_2 and N_3 provide isolation due to their balance, while N_9 provides additional isolation. In the standard Gilbert cell, N_3 and N_4 provide isolation due to the differential operation, but no additional isolation exists. Simulations suggest that the proposed integrated LO buffer and active mixer can provide over 20 dB of additional RF-LO isolation over a standard Gilbert Cell mixer averting the need for separate LO buffering. In this work, the LO buffer operates as a class AB amplifier as a tradeoff between efficiency and linearity [9]. A highly nonlinear LO buffer will result in more harmonic mixing and therefore high noise folding.

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Fig. 2. RF-LO leakage path for (a) the proposed mixer and (b) a Gilbert Cell.

Fig. 3. Illustration of LO leakage and usable current for the proposed mixer.

B. Weak Inversion Biasing

Overall, the proposed mixer operates somewhat differently from a typical Gilbert Cell mixer. In our case, the transistors are arranged in pairs which can be switched on and off. When conducting, a transistor pair acts as a differential amplifier providing a transconductance, g_m , from the input to the output. The switching transistors are sized to ensure that the transistors are weak-inversion biased during peak current conduction. This improves the g_m to I_{DS} ratio.

Because the LO signal current flowing through each switching pair switches at RF, current meant for the switching transistors can leak through the gate-source capacitance of the switching transistors. This is illustrated in Fig. 3 where I_{usable} is the current that flows through the channels of the switching transistors, and $I_{leakage}$ is current supplied by I_{buffer} that leaks through the parasitic capacitances. Note that the odd harmonic components of $I_{leakage}$ are cancelled at V_{X-} by the differential architecture. As an approximation, increasing the overdrive voltage of the switching transistors reduces their channel resistance without significantly increasing their gate-source capacitance. This implies that a higher overdrive voltage is needed to maximize the channel current of the switching transistors, I_{usable} .

However, it is well known that an improvement in g_m is yielded with reduced overdrive voltage, and g_m reaches its maximum value when the transistor is biased in the subthreshold region [2]. Our simulations confirm that this improvement in g_m exceeds the degradation in g_m due to the reduced usable current. As such, we used relatively large switching transistors so that they are only weakly inverted when on.

Fig. 4. Measured average conversion gain (five samples) versus IF frequency for both the I and Q paths over five different samples. LO power = -2 dBm, RF frequency = 2.5 GHz.

Fig. 5. Measured double sideband NF versus IF frequency for five different samples. LO power = -2 dBm, RF frequency = 2.45 GHz.

C. Input Matching

A 50 Ω LC matching network is added to the input and the impedance transformation yields a 15 dB voltage gain. The two inductors can be replaced by a single center-tapped inductor to reduce chip area. Unfortunately, center-tapped inductors are not available in our process design kit (PDK). A 15 dB voltage gain corresponds to an input quality factor (Q) of 5.62 which gives a bandwidth of over 430 MHz at a center frequency of 2.4 GHz. This is more than sufficient for 2.4 GHz ISM band applications.

III. MEASUREMENT AND DISCUSSION

The design was fabricated in the Chartered Semiconductor Manufacturing 0.18 μm RFCMOS technology and characterized using on-wafer probing. The fabricated design includes a poly-phase filter to split the LO into I and Q phases. Five random samples on the wafer were characterized.

The conversion gain versus IF frequency is shown in Fig. 4 for both the I and Q paths. The average conversion gain mismatch between I and Q at 30 MHz is 0.15 dB. The input bandwidth for the RF signal is over 400 MHz. The double sideband (DSB) NF versus IF frequency is shown in Fig. 5. The $1/f$ corner frequency is 24 MHz. The RF-LO and LO-RF isolations are shown in Fig. 6. The RF-LO isolation is sufficient to prevent LO pulling. The average current consumption was 560 μA from the 1.8 V supply. A comparison between this work and

Fig. 6. Average (five samples) RF-LO and LO-RF isolation with LO power = -2 dBm and RF Power = -20 dBm. For RF-LO isolation, LO frequency = 2.45 GHz.

TABLE I
ACTIVE MIXER COMPARISON TABLE

	This Work	[10]	[10]	[11]	[12]
Frequency (GHz)	2.4	2.4	2.4	2.4	2.4
Gain (dB)	32	32	33	15.7	15.7
DSB NF (dB)	8.5	9	4.5	18.3	~ 9.9
IF (MHz)	30	~ 30	~ 0.1	60	1
IIP ₃ (dBm)	-14.5	-13	-12	-9	1
DC Current (mA)	0.56	5.4	5.0	0.5	4.5
LO Power (dBm)	-2	-	-	-9	4^A
Technology	$0.18\ \mu\text{m}$ CMOS	$0.18\ \mu\text{m}$ CMOS	$0.18\ \mu\text{m}$ CMOS	$0.13\ \mu\text{m}$ CMOS	$0.18\ \mu\text{m}$ CMOS

^ANormalized to $50\ \Omega$

Fig. 7. Die photo showing the proposed active mixer including metal fill and output buffering.

the state-of-the-art in active mixers is shown in Table I. The low LO power of -2 dBm illustrates the buffering ability of the proposed design.

From the table, this work shows excellent tradeoff between NF and current consumption among the higher IF designs [10], [11]. Furthermore, by reusing the LO buffer current, additional current consumption is avoided. This is not apparent from the table since the other designs do not include LO buffering. From the table, [10] used an LNA and high current consumption allowing it to achieve the best NF. On the other hand, [11] and [12] had no input matching networks, and therefore no voltage gain due to the impedance transformation of the matching network, resulting in poor NF given their power consumption compared to other works.

Fig. 7 shows a die photo of the proposed active mixer including metal fill, and output buffers. The entire layout including signal pads occupies an area of $1\ \text{mm}^2$ while the area of the proposed mixer is $0.2\ \text{mm}^2$.

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