

A Highly Linear SAW-less CMOS Receiver Using a Mixer with Embedded Tx Filtering for CDMA

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Abstract- An embedded filtering passive receiver mixer is used to overcome transmitter power leakage without the use of a SAW filter. The receiver IC exhibits more than +60dBm of Rx IIP₂, 2.4dB Rx NF, and +77dB of Triple Beat (TB) with 45MHz transmit leakage at 900MHz Rx frequency while consuming only 18mA from a 2.1V supply.

I. INTRODUCTION

Zero-IF RF front-end architectures [1] are attractive for cellular systems due to lower cost and Bill-of-Material (BOM). But the external SAW filter after the LNA stage has been an essential component of cellular systems for several reasons. The two main reasons for using an external SAW filter in full duplex communication systems, like CDMA and WCDMA, are Triple Beat (TB) and IIP₂ performance requirements at the transmit (Tx) offset frequency. Tx power can leak to the LNA despite the duplexer isolation between the Rx and Tx band. The typical duplexer isolation is 55dB in the CDMA Cellular band, and the maximum Tx power can be as high as +27dBm, resulting in -28dBm of Tx power at the Rx input port. This strong Tx power can cause well-known cross-modulation distortion (XMD) [2].

This distortion problem is usually dominated by the LNA performance, because Tx rejection by an external SAW filter reduces the mixer TB requirement. More importantly, the reduced Tx power at the mixer input reduces the IIP₂ performance concern at the Tx offset frequency, since the second-order distortion at the Tx frequency offset can increase the noise floor of the receiver in a Zero-IF system.

A SAW-less receiver system is desirable since it eliminates the SAW filter as well as the external matching components. There have been several efforts to implement a SAW-less CDMA receiver. A Tx canceller was reported in [3], which used an LMS adaptive filter requiring up/down conversion mixers and a low-pass filter in the loop. This method suffers from several performance problems. First, the NF of the receiver is degraded due to the operation of the LMS loop. Second, the rejection varies depending on the group delay of the external matching network. Third, the TB performance of the overall system can be degraded due to the LMS loop.

An on-chip Tx reject band-pass filter using bond-wire inductors is reported for a WCDMA system in [4]. This method has benefits of saving area compare to an on-chip inductor and increasing the selectivity of the filter due to the high-Q of the bond-wire. But this method may have limited feasibility in real production due to bond wire variations

In this paper, an embedded filtering receiver is introduced.

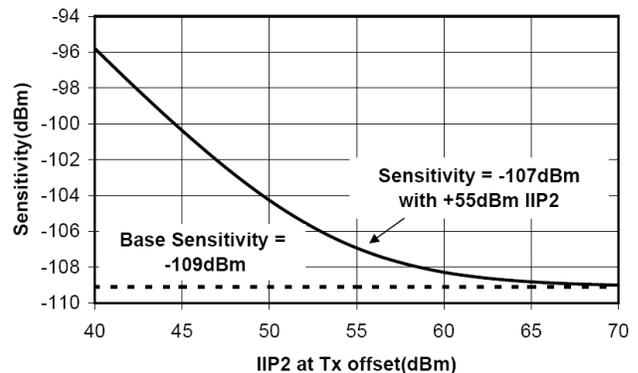


Fig. 1. IIP₂ performance vs sensitivity.

The system does not require a SAW filter and obtains an additional 15dB of Tx rejection; it shows more than +60dBm of receiver IIP₂ at Tx offset and +77dB TB. The double side-band (DSB) NF is 2.4dB and the total current consumption is 18mA from 2.1V supply.

II. REQUIREMENT OF SAW-LESS RECEIVER FOR CDMA

A. Sensitivity Specification

As explained in Section I, the IIP₂ performance at the Tx frequency offset degrades the sensitivity. The trade-off between IIP₂ and sensitivity with -28dBm of Tx leakage power is shown in Fig. 1. The relationship can be defined as

$$Sensitivity = 10 \cdot \log_{10} \left[10^{\frac{IM_{2,Tx}}{10}} + \kappa \cdot T \cdot B \cdot 10^3 \cdot 10^{\frac{NF_{ant}}{10}} \right] + CNR \quad (1)$$

where, κ is Boltzmann's Constant, T is absolute temperature, B is signal bandwidth (1.23MHz in CDMA), CNR is Carrier to Noise Ratio, the NF_{ant} is the noise figure referred to the antenna, and $IM_{2,Tx}$ is Tx IM_2 power in the Rx band. As can be seen from Fig. 1, the sensitivity is highly dependent on the Tx IIP₂ performance, and a +55dBm Tx IIP₂ is required for a 2dB sensitivity degradation. A +55dBm receiver IIP₂ - without some form of Tx rejection - is extremely challenging due to the high LNA gain at the Tx frequency.

B. Single Tone De-sense (STD) Specification

When the Tx power leaks to the Rx port, and a jammer is present in the vicinity of the Rx band, cross-modulation distortion (XMD) becomes the key determiner of the linearity and phase noise requirements of the receiver. The receiver needs to have +8dBm of effective IIP₃ to avoid corrupting the STD test defined in the CDMA standard [5].

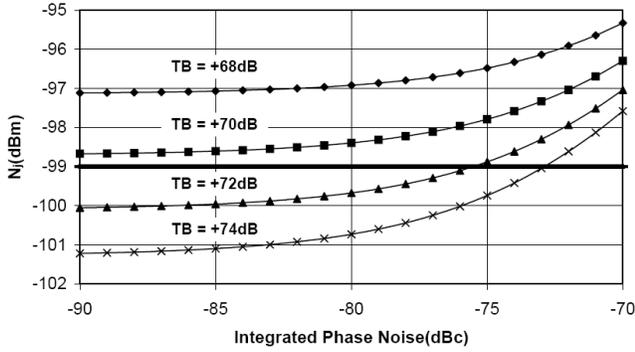


Fig. 2. Trade-off between phase noise and TB performance.

With a SAW filter between the LNA and mixer, the STD mainly depends on the linearity of the LNA, since the SAW filter will reject the Tx leakage by 35dB. On the other hand, a SAW-less receiver places an additional linearity burden on the mixer and the following stages. In addition, the phase noise requirement of the VCO at the jammer offset becomes extraordinarily difficult to meet because of reciprocal mixing. The receiver noise floor with jammer can be expressed as

$$N_j = 10 \cdot \log_{10} \left[10^{\frac{(P_j - TB)/10}{10}} + 10^{\frac{(P_j + P_{phase})/10}{10}} + \kappa \cdot T \cdot B \cdot 10^3 \cdot 10^{NF_{int}/10} \right] \quad (2)$$

where, P_j is jammer power at antenna (in dBm), P_{phase} is phase noise integrated over signal bandwidth with the center frequency at jammer offset (in dBc), and N_j is noise floor with jammer. The TB can be measured by applying three tones, two Tx tones and one jammer tone. Then, the resulting XMD tone will appear at the base-band. The TB is defined as the difference between jammer tone and the XMD tone. The trade-off between phase noise and TB performance is shown in Fig. 2. It shows the phase noise requirement for -99dBm of N_j with various TB performances at -30dBm of jammer power [5]. The noise floor with jammer can never meet -99dBm if the TB is +68dB. But the phase noise can be relaxed to -75dBc with +72dB of TB.

III. SAW-LESS RECEIVER DESIGN

A. CMOS LNA Design

Due to the high IIP₃ requirement of the LNA, the conventional source degenerated LNA is not suitable. There are various ways of designing a highly linear LNA, such as the modified derivative super-position method (MDS) [6] and the active post-distortion method (APD) [7]. In this design, the APD method is chosen. Using this method, the complexity of the bias circuitry and the related input parasitic capacitance can be reduced. The simplified schematic of the LNA is shown in Fig. 3. M_1 , M_2 , M_5 , and M_6 form the main signal path while M_3 and M_4 act as IM₃ cancellers.

The LNA adopts a differential architecture. A differential LNA provides several advantages compared to its single-ended counterpart. First, there is no need for an active or passive balun to connect the doubly-balanced mixer after the LNA. An active balun will cause additional current consumption and linearity degradation and a passive balun

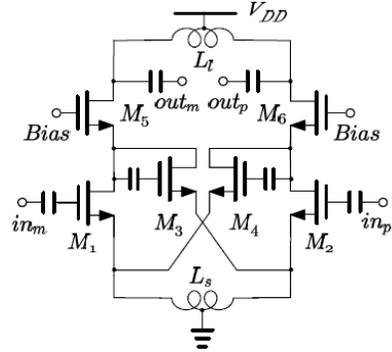


Fig. 3. Simplified schematic of differential APD LNA.

will create area and noise figure penalties due to the passive circuit losses. The differential design requires more external input matching components, but the overall reduction of external components is still significant. The detailed analysis of the linearity and NF of the APD method in a differential LNA is described in [8].

B. Embedded Filtering Passive Mixer Design

In a SAW-less receiver, the mixer becomes the critical component in terms of the linearity performance of the receiver.

The mixer is based on a doubly-balanced passive approach. A passive mixer gives better linearity and NF performance, especially in narrow-band communication systems [9]. The flicker noise (1/f) of the mixer can corrupt the integrated noise, but the passive mixer will not introduce significant flicker noise, since there is no dc current.

The impedance seen at the LNA output is quite high, due to the relatively high impedance of the inductor at the LNA output. The low impedance of the mixer input ensures that the RF current flows to the mixer input with minimal loss. The IM₂ component generated in the LNA can be blocked by a coupling capacitor between the LNA output and the mixer input. Hence, there will be almost no voltage signal swing at the LNA output due to the low impedance provided by the mixer. The nonlinearity of the mixer plays a role after the Tx signal is down-converted.

The problem of using a conventional passive mixer with a transimpedance amplifier (TIA) at the output is illustrated in Fig. 4. The finite opamp gain-bandwidth causes a large signal swing at the TIA input and introduces nonlinearity from both

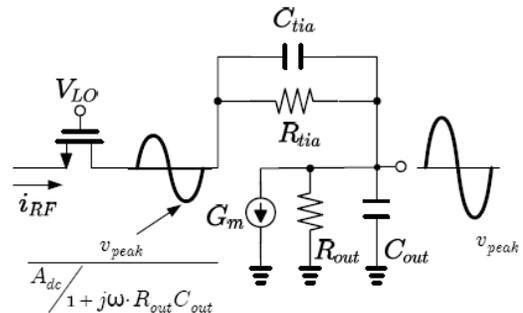


Fig. 4. The problem of using a conventional passive mixer and a TIA with a large Tx leakage signal.

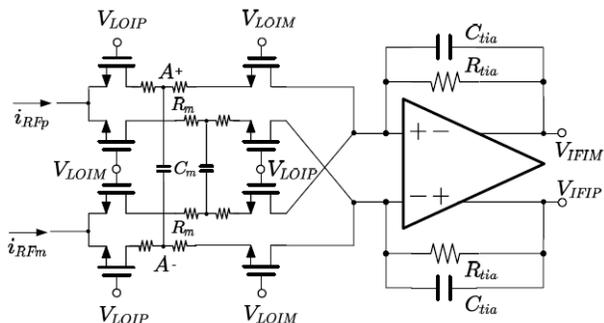


Fig. 5. Simplified schematic of embedded filtering passive mixer.

the Mixer and the TIA. Even if the TIA has a high gain-bandwidth product, it has to provide tremendous current into C_{tia} to avoid introducing nonlinearity caused by incomplete capacitor charging.

The simplified schematic of the embedded filtering passive Mixer (EFP Mixer), which avoids this problem, is shown in Fig. 5. The EFP Mixer has two switches connected in series, and each switch is controlled by opposite phases of the local oscillator (LO) signal. In between the switches is a series RC filter, which determines the bandwidth of the filtering. The RF current flows into the first set of switches, and is down-converted as a nominal passive mixer. The current at node A and the voltage across C_m , when the positive LO signal (V_{LOIP}) is high, can be expressed by

$$i_A(t) \approx g_{m,LNA} \cdot A_{rf} \cdot e^{-\frac{1}{\tau} \cdot 2f_{LO}} \cdot \left[\cos(\omega_{rf}t) + \frac{2}{3\pi} \cdot \cos(2\omega_{rf}t) - \frac{2}{15\pi} \cdot \cos(4\omega_{rf}t) \dots \right] \cdot LO(t) \quad (3)$$

$$v_{C_m} = \frac{1}{C_m} \cdot \int i_A(t) dt \cdot \left[1 + \left(\frac{f_{offset}}{f_{3dB}} \right)^2 \right]^{-\frac{1}{2}} \quad (4)$$

where, $\tau = R_m C_m$, $f_{3dB} = (2\pi\tau)^{-1}$, A_{rf} is the amplitude of the RF input voltage, and $g_{m,LNA}$ is the transconductance of the LNA. This voltage generates a current into the transimpedance amplifier, inversely proportional to R_m , when the negative LO signal (V_{LOIM}) is high, and the integration in (4) provides a low-pass filtering action.

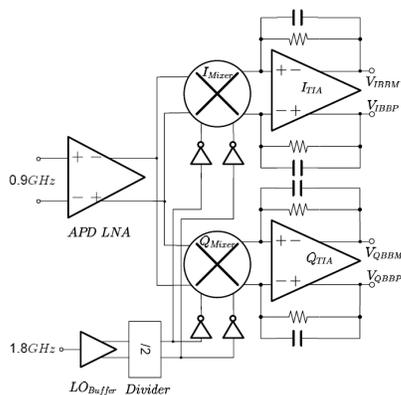


Fig. 6. Block Diagram of embedded filtering SAW-less receiver.

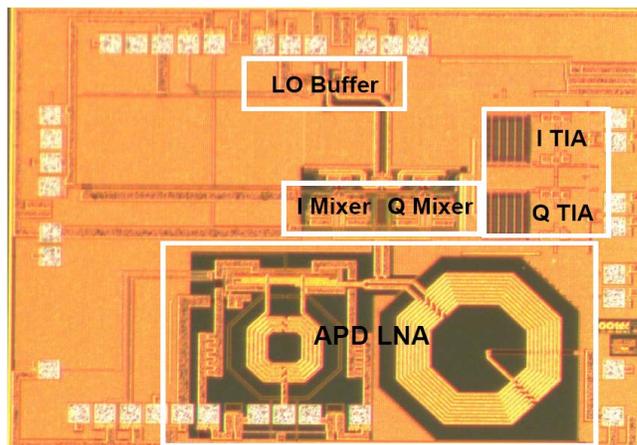


Fig. 7. Die-photograph of embedded filtering receiver.

IV. MEASUREMENT RESULTS AND DISCUSSION

The block diagram of the receiver is shown in Fig. 6 and a die photograph is shown in Fig. 7. It consists of a differential APD LNA, embedded filtering passive Mixer, TIA, LO buffer, divider, and input LO buffer. A conventional passive mixer version of the receiver was fabricated at the same time, in order to have a comparison with the EFP mixer.

The measured frequency responses for both receivers with the differing mixers are shown in Fig. 8. The measured gain performances are 42dB for EFP mixer and 44dB for conventional mixer, respectively. The rejection is normalized since there is 2dB of gain difference. The TIA provides 1.5MHz of 3dB cutoff frequency and the EFP mixer is designed to have an additional pole at 10MHz. As can be seen, the proposed mixer has 15dB more rejection at 45MHz offset, which is the Tx offset frequency in the CDMA CELL band. The performance of the conventional design illustrates the mixer limitation issue mentioned in Section III. Due to the finite opamp gain-bandwidth, the TIA cannot provide sufficient rejection at a high frequency offset.

The TB performance of the proposed receiver is shown in Fig. 9. Two Tx tones with -31dBm each at 45MHz offset and a jammer at 1MHz offset with -30dBm of power are applied. With the cancellation path on, the TB tone is at -65.8dBm, which implies 77.8dB of TB performance. With the APD cancellation path off, the TB is 65.3dB. This difference between cancellation ON and OFF implies two important facts. First, the APD method improves the TB by 12.5dB. Second, the TB variation with the cancellation ON and OFF means the mixer does not contribute any TB, otherwise the TB would not vary with the cancellation being turned ON and OFF, since it would be dominated by the mixer. Therefore, the additional rejection provided by EFP mixer dramatically improves system linearity performance.

The measured IIP₂ performance at the Tx offset frequency is shown in Fig. 10. The conventional passive Mixer starts showing a 4-to-1 nonlinearity, which is strongly nonlinear behavior, from -30dBm. The proposed EFP mixer, whereas, starts showing strongly nonlinear behavior above -24dBm of power.

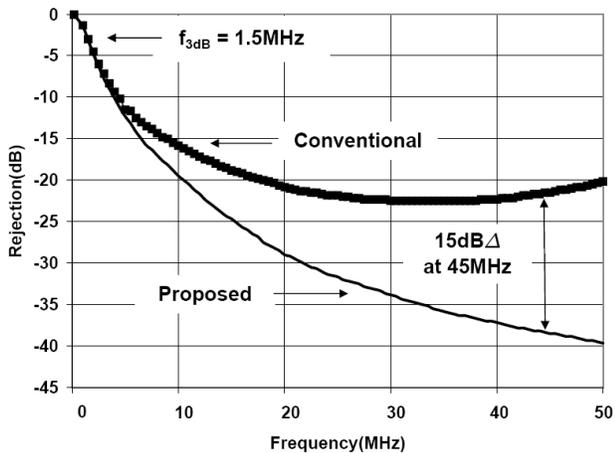


Fig. 8. Measured frequency response comparison.

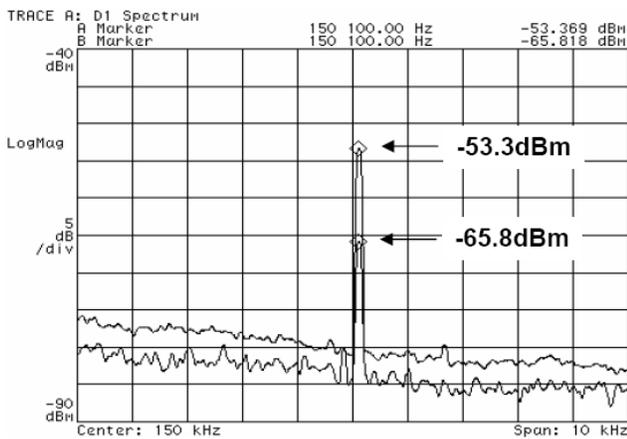


Fig. 9. Measured TB with and without cancellation path in APD LNA.

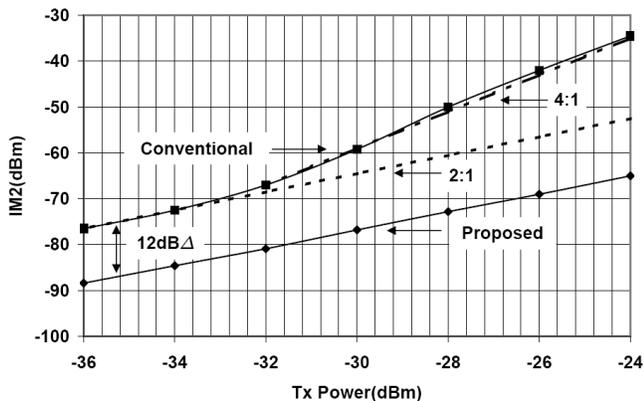


Fig. 10. Measured IIP2 performance comparison at Tx offset (45MHz).

The IIP₂ performance with the EFP mixer is +60dBm and +50dBm with the conventional passive mixer. The conventional mixer exhibits strongly (4:1) nonlinear behavior at -30dBm, which is a key power level in the specification. As a result, the IIP₂ performance of the conventional receiver is further degraded by 2dB.

The overall performance comparison is depicted in Table I. Each receiver is measured with the APD cancellation ON and

Table I: Receiver Performance Comparison

	Proposed Rx	Conventional Rx
V _{dd}	2.1V	2.1V
I _{dd}	18mA/17mA ¹	18mA/17mA
VSWR	<2:1	<2:1
Voltage Gain	42/44dB	44/46
Noise Figure	2.4/3.4dB	2.0/2.8dB
IIP ₂ at 45MHz	+60/+65dBm	+50/+55dBm
Triple Beat	+77.8/+65.3dB	+47/+47dB
Tx Rejection	37dB	22dB

¹ Cancellation ON/OFF

OFF. The Tx IIP₂ performance is important when there is no jammer present. The APD cancellation path can be turned OFF with this situation. The cancellation path needs to be ON only when the jammer is present. The EFP mixer has less gain and higher NF than a conventional mixer due to additional loss in the second set of switches.

The total power consumption for the signal path is 18mA, 14mA for the APD LNA and 4mA for the I/Q TIAs. The chip is fabricated in a 0.18μm CMOS process with 5 metal and 1 poly layer (5M1P). The total area is 2.25mm² including all the related pads and ESD circuitries.

V. CONCLUSION

A highly linear SAW-less receiver is introduced. The system performance trade-off with and without a SAW filter is explained. The high IIP₂ of the receiver at the Tx offset is essential to avoid corrupting the system's sensitivity performance, and a high TB is required to avoid STD performance degradation due to transmitter leakage. Thanks to the embedded filtering in the mixer and the APD method in the LNA, the required high linearity is achieved with low noise figure and power consumption.

ACKNOWLEDGMENTS

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