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A High IIP2 and IIP3 CMOS Down-Conversion Active Mixer

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***Abstract*— Two linearization techniques for double-balanced active mixers are introduced in this paper. The proposed techniques are based on cancelling the second- and third-order kernels of the transconductance stage output current. The second-order kernel is cancelled by adding a second-order intermodulation (IM2) current with an opposite sign generated by injecting a first-order term to the transconductance stage and also attenuating the third-order kernel of the transconductance current significantly by an improved form of the IM2 injection method. The proposed mixer has been designed for IEEE 802.11 applications with input frequency and output bandwidth equal to 2.4 GHz and 20 MHz, respectively, and simulated using a 90 nm RF-CMOS technology. Spectre-RF simulation results show that the proposed techniques have simultaneously improved IIP2 and IIP3 by approximately 26.7 dB and 21.4 dB, respectively, while only a 0.2 mA extra current is drawn from a 1.2 V power supply. In addition, the proposed techniques have no effect on the other parameters of the mixer such as the noise figure and conversion gain.**

Keywords*— *IIP2, IM2, IIP3, IM3, linearity, CMOS active mixers, direct conversion receivers.

I. INTRODUCTION

CMOS direct-conversion receivers (DCR) are more attractive rather than other receivers because of their high integration level, low cost, and simplicity of baseband circuitry and have been widely used in modern wireless terminals. However, the even-order nonlinearities and flicker noise are the challenging issues in designing these kinds of receivers affecting their linearity and noise figure [1-3]. Two most important parameters indicating the linearity in DCRs are IIP2 and IIP3 [4]. The most dominant source of nonlinearity and flicker noise in a receiver is the mixer.

The conventional double-balanced Gilbert mixer as shown in Fig. 1(a), because of its high port-to-port isolation, has widely been used as a down-converter in DCRs. In a perfectly balanced mixer, even-order nonlinearities generated by transistors nonlinearity would not appear at the output. However, in practice due to the existence of mismatch between the local oscillator (LO) signals, load resistances and switching transistors, even-order nonlinearities appear at the signal path [5]. The even-order nonlinearities can be attenuated using fully differential topologies and symmetric layouts; however, the required performance especially for cellular phone applications cannot be met. Therefore, in order to meet the required IIP2 performance, for cellular phone applications, it is important to use on-chip techniques such as analog techniques or calibration circuits [5]. Many calibration techniques have been presented to address IIP2 problem. However, they are complex and power hungry [6, 7]. Using analog techniques and optimizing the mixer for IIP2 is another approach [5, 8, 9]

Several pre-distortion, multiple gated transistor (MGTR) and feed-forward techniques have been presented to cancel the third-order transconductance [10-12]. MGTR and pre-distortion techniques, being two special cases of the feed-forward technique, use an auxiliary transistor (AT) in parallel with the main transistor (MT) that needs to be properly sized and biased. The third-order derivatives of AT and MT are out of phase with each other, and a as result, the third-order distortion is cancelled. However, due to the interaction between the second-order intermodulation and the input signal, caused by the intrinsic feedback structure of the transistor, the IIP3 improvement is limited, while the noise figure is increased due to using the extra transistor.

In this paper, two linearization techniques based on cancelling the second- and the third-order kernels of the transconductance stage output current are introduced. The third-order intermodulation is attenuated significantly by employing an improved IM2 injection method. The transconductance stage is a fully-differential configuration that filters out the common-mode current terms and includes an inherently poor IM2 component. However, employing the IM2 injection circuit in this stage, the total output IM2 current is increased. In the last method, an IM2 current with an equal magnitude and opposite phase is generated by injecting a first-order current to the transconductance stage. The very same current is subtracted from the output current of this stage resulting in the cancellation of the IM2 current of the transconductor.

The paper is organized as follows. The proposed linearization techniques are explained in Sect. II based on the Volterra linearity analysis. The Spectre-RF simulation results of the proposed mixer using a 90 nm RF-CMOS technology are given in Sect. III which is followed by the conclusion in Sect. IV.

II. THE PROPOSED LINEARIZED MIXER

A conventional double-balanced Gilbert cell mixer is shown in Fig. 1(a). This mixer consists of three stages: transconductance stage, switching stage and load stage. With assuming the switching stage is ideal, the transistors of the transconductance

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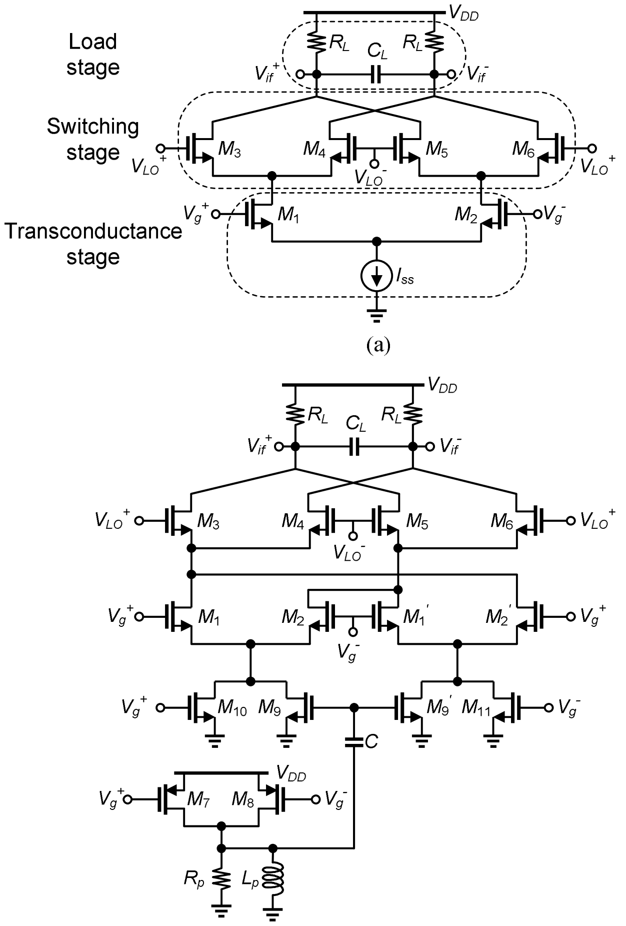

(a)

(b)

Fig. 1. (a) Conventional mixer and (b) proposed mixer.

stage are the most dominant sources of nonlinear current in a mixer. The second- and the third-order intermodulations are the most important parts of the nonlinear current that limit the IIP2 and IIP3 of the mixer. In order to improve these parameters, two new techniques are introduced to cancel IM2 and IM3 currents of the transconductance stage. To investigate the mechanism of above-mentioned methods, the Volterra series analysis is employed.

A. IM3 Canceling

A significant portion of the distortion in an active mixer is due to the non-ideality in the transistors of the transconductance stage. The small-signal drain current of a transistor can be expressed by Taylor series expansion as follows:

$$i_d = g_m(v_g - v_s) + g'_m(v_g - v_s)^2 + g''_m(v_g - v_s)^3 + \ldots \qquad (1)$$

where, g_m, g'_m and g''_m are the transconductance, the first and the second derivatives of the transconductance, respectively. If the source signal has a second-order component of the gate signal or vice versa, another third-order intermodulation term is likely to appear in the drain current equation, in addition to the intrinsic third-order intermodulation component arising from g''_m. This term is called the interaction caused by the gate signal multiplying by the source signal. In the transconductance stage of a differential structure as one shown in Fig. 1(a), the intrinsic and interaction terms sum up due to the opposite signs of g'_m and g''_m, and as a result, the total third-order intermodulation term is increased. Changing the sign of the interaction term not only decreases the IM3 current but also may cancel it. The IM2 injection method externally generates and applies the second-order component to the transconductance stage [13]. This method is a version of the interaction linearization method.

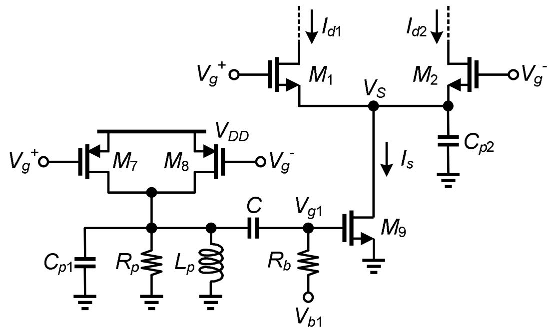

Fig. 2. The circuit realization of the improved form of the IM2 injection method.

Fig. 2 shows the transconductance stage along with the regenerating and injecting circuit. M_1 and M_2 form the transconductance stage. M_7, M_8, M_9, R_p and L_p make the IM2 injection circuit and C_{p1} and C_{p2} are the total parasitic capacitances at the drain(s) of M_7 (and M_8) and the source(s) of M_1 (and M_2), respectively. In this circuit, M_7 and M_8 convert the input signal voltage to a nonlinear current and since the drains of these transistors are connected together, their even-order nonlinear terms are summed up and the odd-order terms are cancelled because of their opposite signs. This current then is injected to the sources of M_1 and M_2 while flowing from the *RLC* network (R_p, L_p and C_{p1}) and being amplified by M_9. The injected current changes the sign of the interaction term in the drain currents of M_1 and M_2. In order to cancel or more attenuate the IM3 current, the magnitude and the phase of this current must be controlled. In this paper, by adding L_p to the IM2 circuit, the phase of the injected current can more be controlled and the IM3 current will be attenuated significantly. To more investigate the behavior of the IM3 cancellation mechanism, the circuit in Fig. 2 is analyzed using the Volterra series. This analysis can provide equations that indicate how the nonlinearities of the mixer can interact in a way to improve the overall linearity performance. The drain current of M_1 in Fig. 2 can be expressed as follows:

$$I_{d1} = A_1(\omega)\circ V_g + A_2(\omega_1,\omega_2)\circ V_g^2 + A_3(\omega_1,\omega_2,\omega_3)\circ V_g^3 + \ldots \qquad (2)$$

where, A_1, A_2 and A_3 are the first-, second- and third-order kernels of I_{d1}, respectively. I_{d1} will be obtained as follows [9]:

$$\begin{aligned} I_{d1} = &-(g_{m1}N_1)I_s + (g_{m1})V_g - (2g'_{m1}N_1)V_g I_s \\ &+ (g'_{m1} - 2g_{m1}K_2)V_g^2 + (g''_{m1} - 4g'_{m1}K_2)V_g^3 \end{aligned} \qquad (3)$$

where, N_1 and K_2 can be calculated as below [9]:

$$N_1(\omega) = -\frac{r_{ds}}{1 + 2r_{ds}g_{m1} + j\omega r_{ds}C_{p2}} \tag{4}$$

$$K_2(\pm\omega_1, \mp\omega_2) = \frac{r_{ds}g'_{m1}}{1 + j(\pm\omega_1 \mp \omega_2)r_{ds}C_{p1} + 2r_{ds}g_{m1}} \tag{5}$$

where r_{ds}, g_{m1} and g'_{m1} are the output resistance of M_9, transconductance of M_1 and the first derivative transconductance of M_1. I_s can be expressed versus V_g as below:

$$\begin{aligned} I_s &= X_2(\pm\omega_1, \mp\omega_2) \circ V_g^2 + \ldots = g_{m9} V_{g1} + \ldots \\ &= \underbrace{\frac{2g_{m9}g'_{m7}R_pL_p j(\pm\omega_1, \mp\omega_2)}{-(\pm\omega_1, \mp\omega_2)^2 R_pL_pC_{p1} + j(\pm\omega_1, \mp\omega_2)L_p + R_p}}_{X_2} V_g^2 + \ldots \end{aligned} \tag{6}$$

By equating (2) and (3), A_3 will be obtained as follows:

$$A_3(\omega_1, \omega_2, \omega_3) = g''_{m1} - 4g'_{m1}K_2 - 2g'_{m1}N_1X_2 \tag{7}$$

In (7), the third term originates from the injected low frequency current (I_s). As can be seen from (7), tuning the magnitude and phase of H_2 by selecting a proper value for L_p, one can attenuate A_3 and improve IIP3 according to the following equation:

$$A_{IIP3} = \sqrt{\frac{4}{3}\frac{A_1(\omega)}{A_3(\omega_1, \omega_2, \omega_3)}} \tag{8}$$

Equation (7) states that A_3 can be cancelled resulting in an infinite A_{IIP3} provided that the following equation to be held:

$$X_2 = \frac{g''_{m1} - 4g'_{m1}K_2}{2g'_{m1}N_1} \tag{9}$$

B. IM2 Canceling

In the IM2 injection method, the injected IM2 current produces an interaction term out of phase with the intrinsic term resulting in the attenuation of the IM3 current and improving IIP3. However, it can be shown that the injected current increases the IM2 current of the transconductor.

In the proposed technique, the total common-mode IM2 current component is cancelled and IIP2 is increased by subtracting an IM2 current component from the transconductance stage output current, according to equation (10):

$$A_{IIP2} = \frac{G_{m,RF}v_{RF}^2}{\sqrt{L^2\left(I_{IM2,diff}^2 + I_{IM2,CM}^2\right) + \left(\dfrac{\sigma_R}{R_L}\right)I_{IM2,CM}^2}} \tag{10}$$

where G_m, L, R, σ_R, $I_{IM2,diff}$ and $I_{IM2,CM}$ are the mixer input transconductance, low-frequency leakage of the switching transistors, load resistance, standard deviation of mismatch between load resistances, common-mode and differential components of the output IM2 current, respectively.

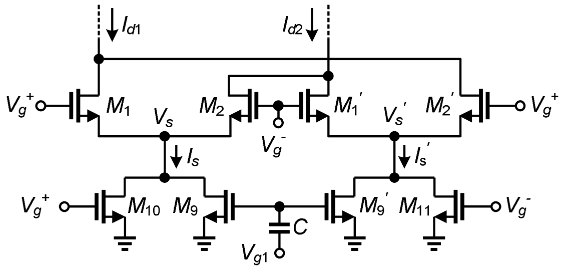

Fig. 3. IM2 and IM3 cancellation circuit.

The circuit realization of this technique is shown in Fig. 3. Although there are not any odd-order terms at the source voltage of the input transistors in a fully-differential structure, a first-order term has been created at the source voltage of M_1 and M_2 through adding M_{10} (M'_{10}) in the proposed circuit. This term produces a second-order term with the same magnitude and opposite phase in the drain currents of M_1 and M_2. This term is caused by g'_m of M_1 and M_2 and cancels the second-order kernel of the transconductance stage output current.

In order to find A_2 and A_3 cancelling conditions, the circuit is analyzed using Volterra series. Here, I_s can be expressed as below:

$$I_s = H_1(\omega) \circ V_g + H_2(\omega_1, \omega_2) \circ V_g^2 + H_3(\omega_1, \omega_2, \omega_3) \circ V_g^3 + \ldots \tag{11}$$

It can be shown that the second- and third-order kernels of the transconductance stage output current will be cancelled if the following conditions to be satisfied:

$$\left|H_1(\omega)\right| = g_{m10} = \left|\frac{1}{N_1(\omega)}\right|\sqrt{\left|\frac{-2g'_{m1}g''_{m1} + 2g'^2_{m1}}{3g_{m1}g''_{m1} - 2g'^2_{m1}}\right|} \tag{12}$$

$$\begin{aligned} H_2(\pm\omega_1, \mp\omega_2) &= g'_{m10} + g_{m9}X_2(\pm\omega_1, \mp\omega_2) \\ &= \frac{-2g'_{m1}g''_{m1} + 4N_1(\pm\omega_1 \mp \omega_2)g_{m1}g'_{m1}g''_{m1}}{N_1(\pm\omega_1 \mp \omega_2)\left(3g_{m1}g''_{m1} - 2g'^2_{m1}\right)} \end{aligned} \tag{13}$$

As it is seen, the condition in (12) can be met by choosing a proper aspect ratio for M_{10}. On the other hand, the condition in (13) can be satisfied by changing the magnitude and phase of the IM2 injected signal. Since, the noise of M_7-M_{11} (and M'_9) and fundamental currents of M_{10}-M_{11} appear as the common-mode at the output, these transistors do not have any effect in the noise figure and conversion gain of the mixer.

III. SIMULATION RESULTS

The complete schematic of the proposed mixer is shown in Fig. 1(b) and it has been simulated using a 90 nm CMOS process with Spectre-RF along with the conventional active mixer. It was designed for 2.4 GHz input signal frequency. A local oscillator with +3 dBm power drives the switching transistors. The IIP3 and IIP2 simulations have been done by applying a two-tone test with 5 MHz spacing. The simulation results of the proposed and the conventional mixers are illustrated in Figs. 5, 6. These figures indicate that the proposed mixer has +23.2 dBm IIP3 and +87.2 dBm IIP2 which are improved by approximately 21.4 and 26.7 dB, respectively, compared to the conventional mixer. The conversion gain and noise figure of the proposed mixer are the same as the conventional one. Table I summarizes the simulation results and compares the proposed mixer with several state-of-the-art linearized mixers [8, 9, 14] with the following figure-of-merit (FoM) [15]:

$$FoM = 10\log\left(\frac{10^{\frac{CG(dB)}{20}} \times 10^{\frac{IIP3(dBm)-10}{20}}}{10^{\frac{NF(dB)}{10}} \times P(mW) \times V_{DD}(V)}\right) \qquad (14)$$

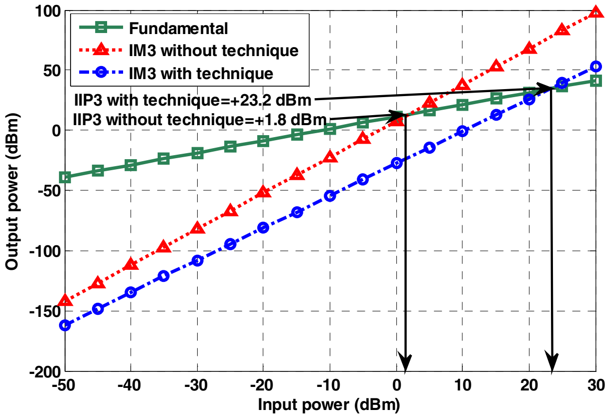

Fig. 4. Simulated IIP3 of the proposed and conventional mixers.

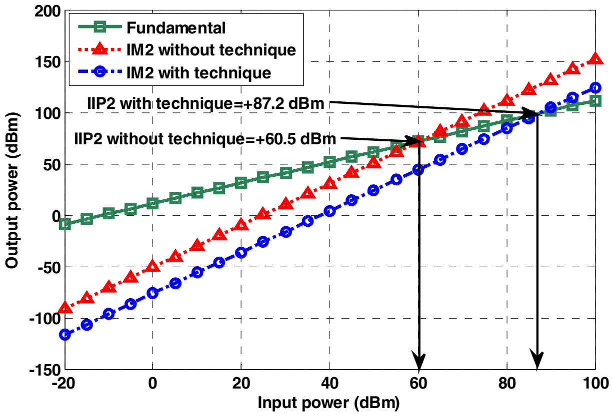

Fig. 5. Simulated IIP2 of the proposed and conventional mixers.

TABLE I. SIMULATION RESULTS SUMMARY.

Parameters	Proposed Mixer	Conv. Mixer	[9]	[14]	[8]*
Frequency (GHz)	2.4	2.4	2.1	2.1	2.1
IIP3 (dBm)	+23.2	+1.8	+15	+6	+10
IIP2 (dBm)	+87.2	+60.5	+93	+90	+82
Conversion Gain (dB)	12.3	12.3	15	12	16
Average NF (dB)	11	11	14	17.5	18.5
Power supply (V)	1.2	1.2	1.8	1	1.8
Power (mW)	2.87	2.68	8	6	7.2
Process (nm)	90	90	180	65	180
FoM	-3.62	-14	-15.3	-21.3	-21.6

* Measurement result

IV. CONCLUSION

In this paper, a linearized mixer along with its Volterra series analysis has been introduced in a 90 nm RF-CMOS technology. In order to improve the linearity, two new techniques are introduced. The first technique removes the IM2 current and improves the IIP2 by about 26.7 dB and the second one is the improved form of the IM2 injection method which attenuates the IM3 current significantly and improves the IIP3 by nearly 21.4 dB. Since, the proposed techniques do not have any influence on the noise figure and input impedance of the mixer; they can be used in low noise amplifiers as well.

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