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A linearization technique for active mixers in zero-IF receivers with inherent balun

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Abstract: A new technique with two different structures for linearization of active mixers in zero-IF receivers is presented. The proposed technique improves the IIP2 more than 20 dB by removing the transconductance common mode current at the mixer's output and IIP3 more than 5 dB in comparison with a basic Gilbert-cell mixer at equal conversion gain. Also, this technique allows us to use a single-ended input in a fully differential structure without needing a bulky and noisy balun. These improvements are achieved at the cost of 0.8 dB increase in NF and also 0.6 mA more current dissipation in one structure.

Keywords: Intermodulation, Zero-IF, Gilbert-cell, IIP2, IIP3, NF

Classification: Integrated circuits

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1 Introduction

Mixer's linearity and 1/f noise determine the total linearity and 1/f noise in zero-IF receivers [1]. In zero-IF receivers the 2nd and 3rd order intermodulations are the most important parts of nonlinearity at the output. Even order intermodulations do not appear at the output when mixer is perfectly balanced. But, in practice, there are some mismatch between the corresponding elements used in the circuit and loads and hence IM2 appears at the output.

The Gilbert-cell is the most popular active mixer which is widely used in zero-IF receivers. In the differential pair of the Gilbert-cell, the current of each transistor can be written as sum of the common mode (CM) and differential mode terms. In down-conversion mixers, it is desirable to remove the CM currents because of two reasons. Firstly, it makes an offset voltage at the output of the mixer and causes the next stages to go in saturation or cutoff mode. Secondly, the CM of IM2 current is larger than its differential part as shown by the following equation [1]:

$$\frac{I_{IM2,Diff}}{I_{IM2,CM}} = \frac{3\sigma_{V_{TH}}}{2(v_{GS} - v_{TH})} << 1 \tag{1}$$

where $\sigma_{V_{TH}}$ is the standard deviation of the threshold voltage. By assuming the standard deviation of the threshold voltage to be 2 mV and the overdrive voltage as $(v_{GS} - v_{TH})$ =100 mV, this ratio is 0.03 indicating that the CM current is the dominant part.

The idea in this paper is to remove the CM current in transconductance stage of the mixer to achieve a high IIP2. When we use the single-ended input, this circuit removes the CM current in input RF frequency. Consequently, it is equivalent to a differential input signal. Usually it is done by a balun which converts the single-ended path to a differential one. It means that the proposed circuit makes us needless of using a bulky and noisy balun.

2 IM2 generation of input transconductance stage and switches

The IM2 product of transconductance (Gm) stage in down-conversion mixers after passing the switching stage is up-converted. So, at the output of the mixer, the IM2 term is not important when the circuit is fully balanced. But, due to the mismatch between the switches and loads the IM2 term leaks to the output. The leakage gain by which the IM2 term in the transconductance current appears at the output has been analyzed in [2]. The CM of IM2 current ($I_{IM2,CM}$) is leaked to the output and becomes differential and hence it is added to the differential IM2 output current as shown by the following equation [3]:

$$I^2_{IM2,out} = M^2 I^2_{IM2,Diff} + [M^2 + (\frac{\sigma_R}{R_L})^2] I^2_{IM2,CM} \tag{2}$$

where $I_{IM2,diff}, R_L, \sigma_R$, and M are the differential IM2 current in the mixer output, the mixer's load, standard deviation of mismatch in the load and low frequency leakage of the switching transistors, respectively.

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3 Proposed mixer structures

The current of each transistor in the differential transconductance stage can be decomposed into a CM and a differential mode current. As shown in Fig. 1 (a), the input stage of the proposed mixer is decomposed into two differential pairs. The first pair which is the input pair (M1, M2), makes positive I_{CM} current and the second pair (M3, M4), which is called the quasi pair in this paper, makes a I_{CM} equal to the other pair but with the opposite sign. By summing these two outcoming currents, the CM terms are subtracted from each other, but differential mode currents which have the same signs are added together. The roll of Mq1-Mq4 is to transfer the v_{gs} of input pair to the quasi pair by the opposite sign to cause the quasi pair to generate the I_{CM} by the opposite sign of the input pair. This method is called the common mode remover (CMR). According to Fig. 1 (a) it can be written:

$$i_{O1} = i_{M1} + i_{M3} = (i_{CM} + i_{Diff}) + (-i_{CM} + i_{Diff}) = 2i_{Diff} \qquad (3)$$

where i_{O1} and i_{O2} are the output differential current of Gm stage of the mixer. It is seen that the CM current is removed from the output current.

The alternative approach which is called the current reuse common mode remover (CRCMR) is shown in Fig. 1 (b). In CRCMR, the transistor M3 should provide the CM current equal to the sum of CM currents of M1 and Mq1 and also M4 should provide the CM current equal to the sum of M2 and Mq2 to eliminate the CM current at the output.

In CMR technique, the currents of Mq1 and Mq2 are wasted without participating in the mixer's conversion gain. But, in CRCMR, these currents are reused to achieve more gain. So, at the equal conversion gain, NF, and linearity, the CMR scheme dissipates more current than CRCMR one. But, the required high CM voltage level for LO port of CRCMR is undesirable. As is seen in Fig. 1 (b), the LO CM voltage should be at least $2v_{GS} + 2v_{od}$.

In CM small-signal analysis to calculate $I_{IM2,CM}$, the source nodes of M1 and Mq1 are high impedance nodes. So, the mismatch between M1 and M2 and between Mq1 and Mq2 and also the different output resistance of the tails of input and intermediate pairs makes the v_{gs} of M1 and Mq1 to be non-equal. Putting a capacitor between the tails of input and intermediate pairs solves this problem and hence we will have a better matching between the $I_{IM2,CM}$ of these two pairs. In Figs. 1 (a) and 1 (b), $v_{gs,q1} = v_{gs,q3}$. It can be seen that $v_{s,q3} = v_{g4}$ and $v_{g,q3} = 0$ for small-signal analysis but for CM analysis v_{s4} is not grounded. With the added capacitor between the source of M4 and ground, we will have $v_{gs,q3} = -v_{gs4}$ and as a result $I_{IM2,q3} = I_{IM2,1} = -I_{IM2,4}$ in CM small-signal analysis. As a conclusion, to have equal $I_{IM2,CM}$ in input and quasi pairs, in CM analysis, M1 (M2) and Mq1 (Mq2) should have equal v_{gs} and also v_{gs} in M4 (M3) and Mq3 (Mq4) should be equal with the opposite sign. This is performed by placing two capacitors in the tails of input and quasi pairs in Figs. 1 (a) and 1 (b).

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3.1 Small-signal analysis

By assuming $g_{m1} = g_{m2} = g_{m3} = g_{m4}$, and also $i_{out+}, i_{out-}, i_1, i_2, i_3$ and i_4 to be the small-signal part of i_{O1}, i_{O2} and drain current of M1-M4 transistors, respectively, the small-signal output current can be obtained as:

$$i_{out} = i_{out+} - i_{out-} = (i_1 + i_3) - (i_2 + i_4) = g_m v_{in} - (-g_m v_{in}) = 2g_m v_{in} \quad (4)$$

Therefore, the conversion gain of the proposed CMR mixer is similar to that of the basic Gilbert-cell. Four intermediate transistors between two differential pairs of Gm stage (Mq1, Mq2, Mq3, Mq4), transport $-v_{in}$ to the quasi pair. The small-signal gain for CRCMR shown in Fig. 1 (b) is like to CMR in Fig. 1 (a) instead of the added gm of Mq1, Mq2 to the coefficient of v_{in} in the output. Hence, it allows us to use a smaller g_m for M1 and M2.

3.2 IM3 attenuation

The coefficient of third order nonlinearity of the Gm stage is the second order derivative of the transconductance which is derived from the Tailor series as:

$$i_{DC} = I_{DC} + g_m v_{gs} + \frac{g'_m}{2!} v_{gs}^2 + \frac{g''_m}{3!} v_{gs}^3 + \ldots \quad (5)$$

Plot of the g''_m versus to the v_{GS} shows that at some values of v_{GS} the g''_m is negative and at some values of v_{GS} it is positive [4]. So, biasing the input pair and the quasi pair in Gm stage at different voltages attenuates the IM3 term because the output current of them is added and hence g''_m becomes smaller resulting in an improvement in IIP3.

3.3 Noise analysis

In order to compare the circuit noise of CMR and CRCMR with the basic Gilbert-cell, firstly it is worth to mention that only the thermal noise of the Gm stage is important and we can neglect the 1/f noise of these transistors. Secondly, the total thermal noise effect of input and quasi pairs in the proposed mixers are equal to the thermal noise contribution of two input transistors in the basic Gilbert-cell. So, the only added noise is the thermal noise contribution of four intermediate transistors which passing through the quasi pair in Figs. 1 (a) and (b). Therefore, to estimate the added noise, it is enough to calculate the noise effect of four intermediate transistors.

To calculate the added noise in Figs. 1 (a) and (b), the noise of Mq1 and Mq3 are calculated like the noise of Mq2 and Mq4. Therefore, just the calculation of added noise of Mq2 and Mq4 are given here. According to Fig. 1 (c) the thermal noise effect of Mq2 on the output current of the Gm stage is as:

$$i_{n,outq2} = i_{out+} - i_{out-} = g_{m3}(v_{q+} - v_{q-}) = -\frac{g_{m3}}{g_{mq4}} i_n \quad (6)$$

where g_{m3} is the transconductance of the quasi pair transistors.

According to Fig. 1 (d) it is seen that all of the thermal noise current of Mq4 will passes through itself since $1/g_{mq4} << r_{dsq2}$. It results in:

$$i_{n,outq4} = i_{out+} - i_{out-} = g_{m3}(v_{q+} - v_{q-}) = \frac{g_{m3}}{g_{mq4}} i_n \quad (7)$$

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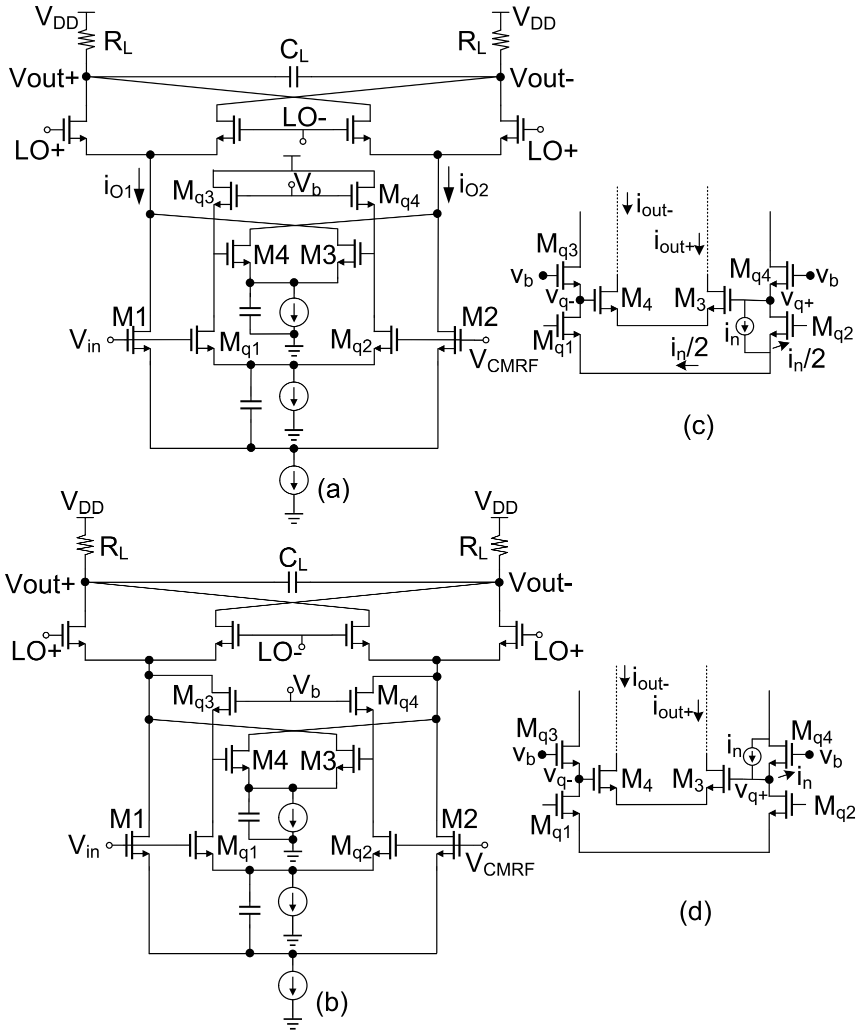

Fig. 1. The proposed (a) CMR and (b) CRCMR mixers, noise effect of (c) Mq2 and (d) Mq4.

In relations (7) and (8), in is the root square density of thermal noise of Mq2 and Mq4. It is concluded that:

$$
\begin{aligned}
i^2_{n,out} &= i^2_{n,outq1} + i^2_{n,outq2} + i^2_{n,out3} + i^2_{n,out4} \\
&= 4(\frac{g_{m3}}{g_{mq4}} i_n)^2 = 4(\frac{g_{m3}}{g_{mq4}})4kT\gamma g_{mq4} = 4\frac{g^2_{m3}}{g_{mq4}}4kT\gamma \qquad (8)
\end{aligned}
$$

From the above equations it is seen that we can minimize the effect of noise of four intermediate transistors by choosing a proper g_m for these transistors.

4 Simulation results

To verify the usefulness of the proposed mixer structures, simulation results are provided. The simulations are performed using a 90 nm CMOS with HSPICE-RF. To have a fair comparison, the basic Gilbert-cell was also simulated. To calculate the IIP2, an intentional 2% mismatch was considered in resistors and for transistors, the random mismatch is given by the Gaussian distribution with a standard deviation ($\frac{\Delta W}{W}$) equal to 0.373/(W*L*M) which

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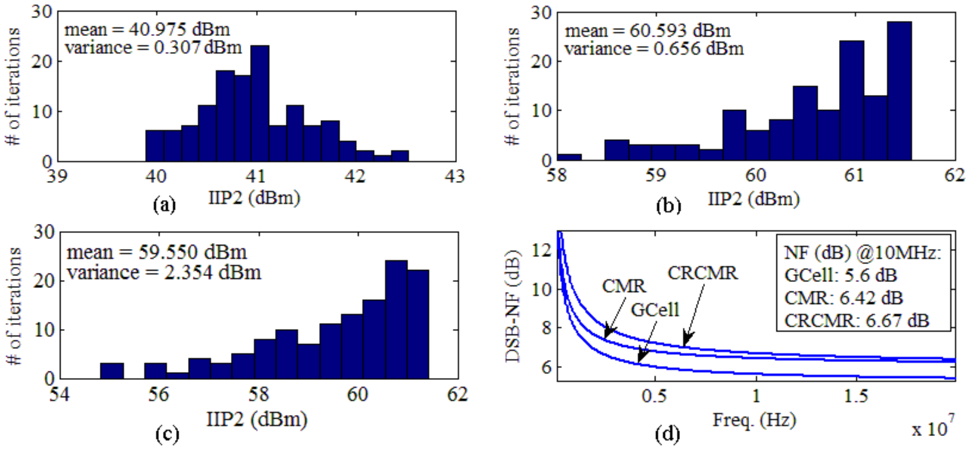

Fig. 2. Monte-Carlo simulation results of IIP2 in (a) basic Gilbert-cell, (b) CMR, and (c) CRCMR mixers, and (d) simulated DSB-NF.

Table I. Simulation results summary.

Parameter	Gilbert cell	CMR	CRCMR
Frequency	2.4 GHz / 20 MHz		
IIP2 (mean)	40.97 dBm	60.59 dBm	59.55 dBm
IIP3	−20.64 dBm	−14.78 dBm	−15.68 dBm
Conversion Gain	9.21 dB	9.28 dB	9.08 dB
DSB-NF (@10 MHz)	5.6 dB	6.42 dB	6.67 dB
Power Dissipation	2.82 mW	4.06 mW	3.26 mW
Power Supply Voltage	1.2 V		
Technology	90 nm		

W, L and M are width, length and number of fingers respectively. 130 iterations for Monte-Carlo simulations were performed and the results are shown in Figs. 2 (a), 2 (b) and 2 (c). Table I summarizes the simulation results. According to Table I, the mean IIP2 is improved more than 19 dB and the IIP3 is improved about 5 dB for both structures i.e. CMR and CRCMR compared to the basic Gilbert-cell. The conversion gain is equal in basic Gilbert-cell and the proposed techniques to provide a fair comparison in linearity. According to Fig. 2 (d), the DSB-NF of the proposed techniques in comparison with the Gilbert-cell is slightly increased as theoretically expected. It is worth mentioning that several techniques have been proposed to improve the linearity of active mixers in zero-IF receivers. But, most of them improve just one of the linearity parameters, i.e. IIP2 or IIP3. In addition, some of them use an inductor which occupies a large silicon area and also has radiation effects on other parts of the receiver [3, 4, 5]. In the proposed techniques, both of the linearity parameters have been improved without needing any inductor. Besides, the need for a bulky balun is eliminated which allows us to use a single-ended LNA in the previous stage further to save the power.

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5 Conclusions

Two active mixers with improved linearity were proposed. The common mode currents of transconductance stage are removed by using a quasi differential pair making a significant improvement in both IIP2 and IIP3. Besides, the value of IIP3 is increased since the proposed technique attenuates the g_m''. Another benefit of this method is to convert the single-ended input signal to a differential one avoiding the need for a bulky and noisy balun before the fully differential mixer when the LNA is single-ended.

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