A Highly Linear Mixer with Inherent Balun Using A New Technique to Remove Common Mode Currents

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***Abstract*—A highly linear active mixer is introduced for IEEE 802.11 applications. It uses a new technique to remove the common mode currents in all frequencies. In zero-IF receivers, IIP2 is the most important parameter which shows the output's IM2 nonlinearity. In this paper, IIP2 improves more than 17 dBm by removing its common mode current. Also, the new technique, improves IIP3 more than 5dBm in comparison with a simple Gilbert-Cell mixer at equal conversion gain. These are achieved at the cost of 0.8 dB increase in NF and 0.6 mA more current dissipation. Simulations are performed in a 90 nm RF-CMOS technology by HSPICE-RF.**

I. INTRODUCTION

Mixer's linearity and 1/f noise, are the dominant parts of linearity and 1/f noise of the receivers [1-3]. The most important parts of linearity in zero-IF receivers are 2nd and 3rd order inter-modulations. In a perfectly balanced mixer, even order inter-modulations do not appear at the output. In practice, there is a mismatch between the switches and loads and the IM2 appears at the signal path.

Gilbert-Cell is the most popular active mixer which is used in zero-IF receivers. As shown in Fig. 1, the current of each transistor in differential pair can be written as sum of the common mode and differential currents. In down-conversion mixers, it is desirable to remove the common mode currents because of two reasons. First, it makes an offset voltage at the output of the mixer and it causes the next stage to go in saturation or cutoff mode. The other is that the common mode of IM2 current is more important than its differential part as shown in the following equation [2]:

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

where σ_{Vth} is the standard deviation of the threshold voltage.

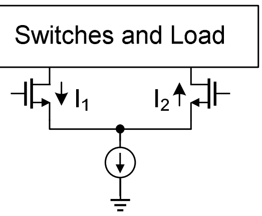

Figure 1. Tranconductance stage of a simple Gilbert-Cell

By assuming the standard deviation of the threshold voltage is about 2 mV and v_{GS}-v_{TH}=100 mV, this ratio is 0.03 which is small and shows that common mode current is the dominant one [4]. Many references such as [4-6] worked on removing the IM2 common mode current.

So, the idea is to remove the common mode current in transconductance stage of the mixer, which gives a high IIP2. Removing the common mode current in all frequencies has another special result which is occur in fundamental frequency. We can use the single-ended input and because the circuit removes the common mode current, it is equivalent to use a differential signal as the input. It means that this circuit makes us needless of using bulky and noisy balun to convert the single path signal to the differential one.

To make this circuit suitable for IEEE 802.11 applications, the frequency and bandwidth are choose 2.4 GHz and 20 MHz respectively. The technology is a 90 nm CMOS with 1 V power supply.

The paper is organized as follows. The role of switches and transconductance stage in IM2 generation will be explained in Sect. II. Section III will explain the proposed structure that removes the CM currents. Also in this part, the conversion gain and noise of the proposed structure will be presented. Simulation results are given in section IV. Finally, the paper is concluded in Sect. V.

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II. ROLE OF INPUT TRANSCONDUCTANCE STAGE AND SWITCHES IN IM2 GENERATION

The IM2 product of Gm stage in down-conversion mixers after passing switching stage is up-converted. So, at the output it is not important. But, mismatch of the switches and loads cause the IM2 term leaks to the output. If we model the mismatches of the switches with an offset voltage in series with the gate of one FET of a pair, the switching waveform can be decomposed into a symmetric square wave at ω_{LO} and a train of error pulses at $2\omega_{LO}$. The average of the error pulses determines the leakage gain L by which the IM2 in the transconductor current appears at the baseband output of the switching core [3]:

$$L = \frac{2V_{OS}}{S \times T_{LO}} \tag{2}$$

where S is the slope of the single-ended LO during the transition and T_{LO} is its period.[7]

In a perfectly tuned switching stage, the IM2 of the mixer is also usually upper than the IEEE 802.11's requirement, because of the large common mode IM2 current. These CM currents ($I_{IM2,CM}$) are leaked to the output and because of the switches and load mismatches become differential and added to differential IM2 output current [8], as it is shown in the following equation:

$$I_{IM2,out}^2 = M^2 I_{IM2,diff}^2 + \left(M^2 + \left(\frac{\sigma_R}{R_L} \right)^2 \right) I_{IM2,CM}^2 \tag{3}$$

in which $I^2_{IM2,diff}$, R_L, σ_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.[4]

If $I_{IM2,CM}$ is eliminated almost completely, the mixer IIP2 is enough for standards like UMTS, GSM and IEEE 802.11. To address the mentioned level of linearity, a new technique is proposed in the next section that easily can be optimized for different standards.

III. PROPOSED STRUCTURE

The current of each transistor of a differential transconductance stage can be decomposed into a common mode and a differential current. As shown in Fig. 2, the input stage of a Gilbert-Cell is decomposed in to two differential pairs. The first pair which is the input pair, makes positive I_{CM} current and the second pair, which is called the quasi differential pair here, makes I_{CM} equal to the other pair but with the opposite sign. By summing these two out coming currents, the CM currents are subtracted from each other, but differential currents which are with the same sign are gathered with each other. According to Fig. 2 the following equations can be written:

$$i_{O1} = i_1 + i_3 = (i_{CM} + i_{Diff}) + (-i_{CM} + i_{Diff}) = 2i_{Diff} \tag{4}$$

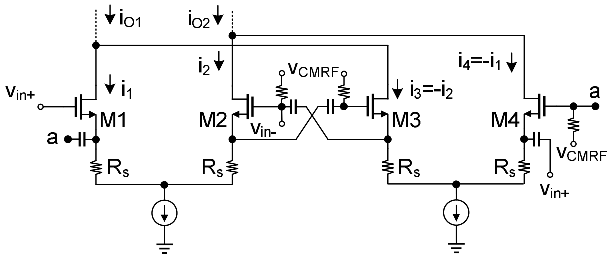

Figure 2. Schematic of CM Remover technique

$$i_{O2} = i_2 + i_4 = (i_{CM} - i_{Diff}) + (-i_{CM} - i_{Diff}) = -2i_{Diff} \tag{5}$$

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.

A. Small-Signal Analysis to Estimate the Gain

By supposing that $g_{m1} = g_{m2} = g_{m3} = g_{m4}$, as shown in Fig. 2, the small-signal output current can be computed as:

$$i_{out+} = i_1 + i_3 = \frac{g_m}{1 + g_m R_S} v_{in} \tag{6}$$

$$i_{out-} = i_2 + i_4 = -\frac{g_m}{1 + g_m R_S} v_{in} \tag{7}$$

which are the same small-signal currents gained for simple Gilbert-Cell. So in the new scheme, the conversion gain of Gilbert-Cell is constant and it is not affected.

B. Final Scheme of the Proposed Structure

The scheme in Fig. 2 has a problem which is the low input impedance. It can be easily calculated:

$$R_{in} \simeq R_S \,||\, \frac{1}{g_m} = \frac{R_S}{1 + R_S g_m} \tag{8}$$

where g_m is the transconductance of each of the transistors in Fig. 2.

In the final circuit which is shown in Fig. 3, the problem is removed by adding 4 intermediate transistors between two differential pairs of Gm stage. Like above small signal analysis, it can be shown that the gain is equal to the gain of a simple Gilbert-Cell. So when it is used as Gm stage of a Gilbert-Cell mixer, it has no attenuation in conversion gain. Also like before it can prove easily that the common mode currents are removed from the output current.

C. IM3 attenuation

The coefficient of third order nonlinearity of the Gm stage is the second order derivative of transconductance which is proved from tailor series:

$$i_{DC} = I_{DC} + g_m v_{gs} + \frac{g_m^{'}}{2!} v_{gs}^2 + \frac{g_m^{''}}{3!} v_{gs}^3 + ... \quad (9)$$

Plot of the g''_m respect 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 [9]. So biasing the input pair and the quasi pair in Gm stage at different voltages, causes the attenuation at the IM3 because the output current of them is gathered and the g''_m can become small and results in an improvement in IIP3.

D. Output Noise in Compare with the Simple Gilbert-Cell

To compare the added output noise with the simple Gilbert-Cell, first it is reminded that only the thermal noise of the Gm stage is important and we dissemble the 1/f noise of these transistors. Second, it should be considered that the total thermal noise of two input transistors and also two transistors of quasi stage are equal to the thermal noise of two input transistors of input stage in simple Gilbert-Cell. So the only added noise is the thermal noise of 4 intermediate transistors between input transistors and the quasi differential transistors. Now, to estimate the added noise, it is enough to calculate the noise of 4 intermediate transistors.

To calculate the added noise, in Fig. 4, the noise of M_{q1} and M_{q3} are calculated like the noise of M_{q2} and M_{q4}, so just the calculation of noise of M_{q2} and M_{q4} are brought here. To calculate the noise of M_{q2} consider the Fig. 4(a). Here is the effect of noise of M_{q2} on the output current of the Gm stage:

$$i_{out} = i_{out+} - i_{out-} = g_{m3}(v_{q+} - v_{q-}) = -\frac{g_{m3}}{g_{mq4}} i_n \quad (10)$$

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

According to Fig. 4(b), it is seen that all of noise current of M_{q4} will passes through itself because $1/g_{mq4} << r_{dsq2}$. It results in:

$$i_{out} = i_{out+} - i_{out-} = g_{m3}(v_{q+} - v_{q-}) = \frac{g_{m3}}{g_{mq4}} i_n \quad (11)$$

In equations 10 and 11, i_n^2 is the power of thermal noise of M_{q2} and M_{q4}. It is concluded that:

$$\left(\frac{g_{m3}}{g_{mq4}} i_n\right)^2 = \left(\frac{g_{m3}}{g_{mq4}}\right)^2 4kT\gamma g_{mq4} = \frac{g_{m3}^2}{g_{mq4}} 4kT\gamma \quad (12)$$

From the above equation it can be concluded that if $g_{mq4}=g_{m3}$ is chosen, then the effect of noise of M_{q2} and M_{q4} is equal M_3. So the effect of the noise of the 4 intermediate transistors is equal to triple the output noise of the quasi differential pair or multiply the output noise of the transconductance stage by 1.5. It means that at the worst case – gm of input and quasi differential pairs are equal and g_{mq4} is equal to g_{m3} and ignoring the noise of the switching transistors – the 4 intermediate transistors can increase the output noise up to 1.7 dB.

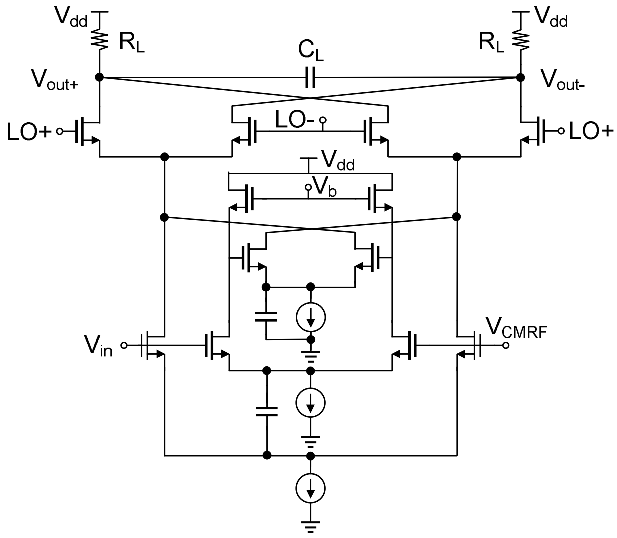

Figure 3. Schematic of the Proposed Mixer

IV. SIMULATION RESULTS

The new technique is applied to a simple Gilbert-Cell as shown in Fig. 3 and is simulated by HSPICE-RF and the results are compared with the simple Gilbert-Cell in Fig. 5. To calculate the IIP2 and IIP3, the Monte-Carlo simulations are performed as shown in Figs. 6 and 7, and the worst case is brought here for comparison. Table I shows the comparison results. According to Table I, IIP2 improved more than 17 dBm and IIP3 improved more than 5 dBm.

To have a well comparison in linearity, it is assumed that the conversion gain is equal in simple Gilbert-Cell and the proposed technique as shown in Fig. 8. According to Fig. 9, the DSB NF of the proposed technique in comparison with the simple Gilbert-Cell is increased insignificantly. As brought in Table. I, at 10 MHz DSB NF is increased about 0.74 dB.

Techniques such as [4-6, 10] improve just one of the linearity parameters, 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. In the proposed technique, both of the linearity parameters has been improved; besides, the need for a bulky balun is eliminated which allows us to use a single-ended LNA in the previous stage, further saving the power.

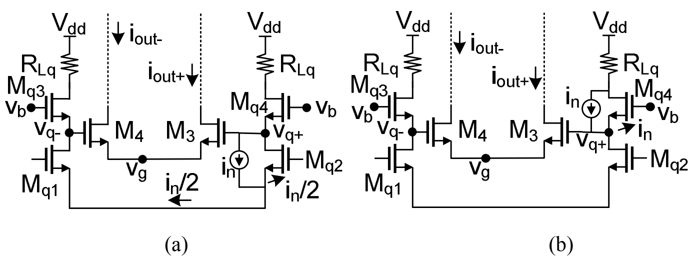

Figure 4. Proposed schematic for calculating: a. M_{q2}'s noise b. M_{q4}'s noise

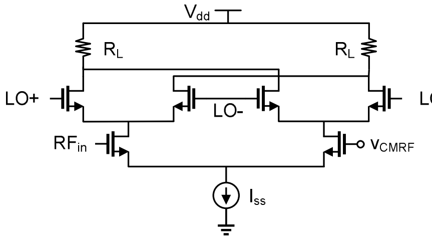

Figure 5. Simple Gilbert-Cell mixer

TABLE I. SIMULATION RESULT

	CM Remover	Gilbert Cell
Freq / BW	2.4 G / 20 MHz	2.4 G / 20 MHz
IIP2 (min)	46.74 dBm	29.484 dBm
IIP3 (min)	-16.318 dBm	-21.3445 dBm
CGain (Tran)	9.23 dB	9.03 dB
NF (@10 Meg)	6.37 dB	5.63 dB
Power	2.97mA / 1V	2.35mA / 1V
Technology	90 nm	90 nm

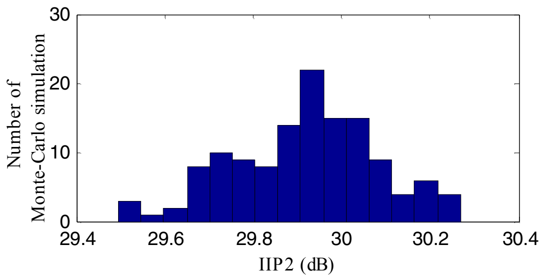

Figure 6. Monte-Carlo simulation results for simple Gilbert-Cell

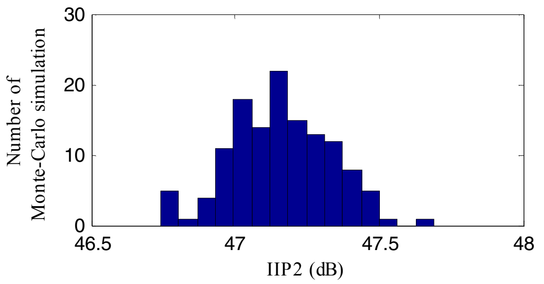

Figure 7. Monte-Carlo simulation results for the proposed technique

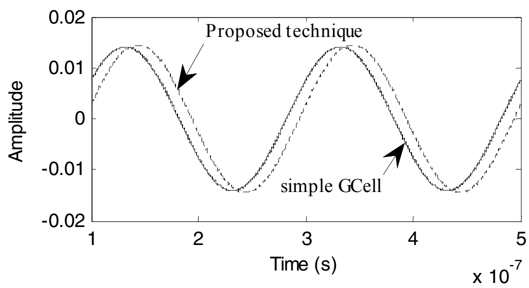

Figure 8. Comparison of conversion gain in Gilbert-Cell with the proposed technique

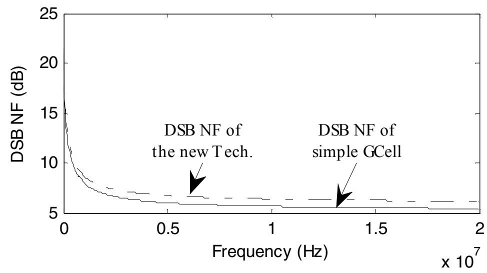

Figure 9. Comparison of DSB NF in Gilbert-Cell with the proposed technique

V. CONCLUSION

The proposed new technique is applied to the simple Gilbert-Cell. Simulation results show that the proposed mixer can be optimized for different standards like GSM, UMTS and IEEE 802.11. This method removes the CM currents in different frequencies. In Gilbert-Cell Gm stage, the CM IM2 current is the dominant one. So one advantage of removing CM currents is a highly increment in IIP2 of the mixer. The IIP3 is increased because removing its CM part and also because this technique attenuate the g_m''. Another benefit of this method is to convert the single signal path to a differential signal path avoiding the need for a bulky and noisy balun.

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