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ANALYSIS OF LO LEAKAGE DUE TO LO MISMATCH IN CMOS GILBERT MIXER FOR DIRECT CONVERSION APPLICATION

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ABSTRACT

The DC offset caused by the LO-RF feedthrough in the mixer of a direct conversion receiver is a serious problem. The LO leakage caused by the LO mismatch in double-balanced CMOS Gilbert mixers is examined and analyzed in four cases, no LO mismatch, LO phase mismatch only, LO amplitude mismatch only and the combination of the two mismatches. The relation between the mismatch and the LO leakage is derived. Simulation results shown are consistent with the analysis.

1. INTRODUCTION

In recent years, the direct conversion architecture has increasingly gained more attention as a possible solution for a single-chip radio due to its low power, low complexity and easy integrating properties. However, a number of issues which do not exist or are not serious in the heterodyne architecture become important in the homodyne architecture, such as DC offset, I/Q mismatch, even order distortion and so on. Among these the DC offset generated by self-mixing is the most critical. Approaches to remove the offset have so far mostly been focused on three methods. For modulation formats that have no or little spectral power at DC, AC coupling at the mixer output, or at some downstream stage, can be used to remove the offset [1, 2]. This traditional method requires large capacitor values that are not realizable on-chip. The second common approach is the use of baseband analog and/or digital signal processing (DSP) techniques for offset estimation and cancellation [3-5]. It is a widely used method, but increases the complexity and cost of receivers. Another approach is to use a DC-offset-free mixing topology, such as harmonic mixers [6-8]. This method is not common in the current wireless industry.

The DC offset is caused by the LO-RF leakage that is induced by the substrate coupling and the asymmetry of the mixer. Much research has been done on DC offset cancellation techniques. Nevertheless, no detailed analysis of the LO-RF leakage has been reported in the literature. In this paper, the theoretical analysis of the common mode LO leakage caused by the LO mismatch in a double-balanced CMOS Gilbert mixer is introduced. Simulation results shown are consistent with the analysis.

2. ANALYSIS OF LO LEAKAGE

To better understand the origin of DC offset, consider the received signal path shown in Fig. 1. First, the isolation between the LO port and the inputs of the mixer is not perfect, and a finite amount of feedthrough, which is known as LO leakage, exists from the LO port to the other mixer inputs. The leakage signal is reflected back at the outputs of LNA (ignoring the feedthrough from the LNA outputs to its inputs) and now mixed with the original LO signal, thus producing a DC component at the output of the mixer. So the DC offset can be expressed by Eq. (1).

$$DC\,Offset = LO\,Leakage \times \beta \times G \qquad (1)$$

where β is the reflection factor and G is the mixer gain.

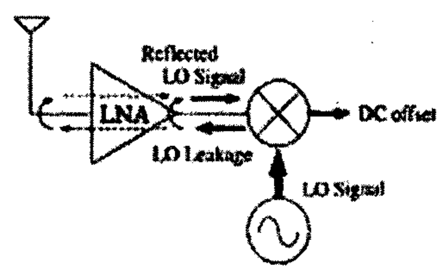

Figure 1. LO signal Self-mixing

The LO leakage arises from many factors, such as substrate coupling, asymmetry of the circuit and so on. For a double-balanced Gilbert mixer, as shown in Fig. 2, if the two differential LO signals mismatch, it will also induce the LO leakage and DC offset. The LO leakage can be obtained by disconnecting RF signals and only

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injecting LO signals to the mixer. In the following subsections, the LO leakage will be discussed in four cases. They are no LO mismatch, LO phase mismatch only, LO amplitude mismatch only and the combination of the two mismatch.

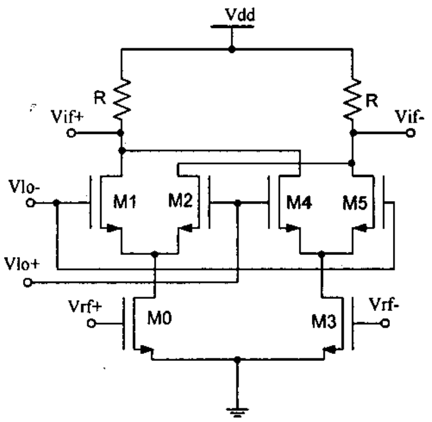

Figure 2. Double-balanced mixer

2.1. Assumptions

Before starting the analysis of the common mode LO leakage, some reasonable assumptions should be made to simplify the problem. Without loss of generality, consider the single-balanced mixer in Fig. 3. Firstly, it is assumed that the circuit switches sharply, that is, a small LO voltage change causes the current of V-I converter (M0) to completely switch from one side of the differential pair to the other. Secondly, the small signal current caused by the RF input is considered as negligible comparing to the M0 biasing current. Therefore only this biasing current is switched by the LO signals. Thirdly, C_{gb} and C_{sb} in the small signal equivalent circuit of M0 are ignored to further simplify the results. C_{gb} is normally ignored when the transistor works in the saturation region [9]. C_{sb} is very small because the source and body of M0 are tied together.

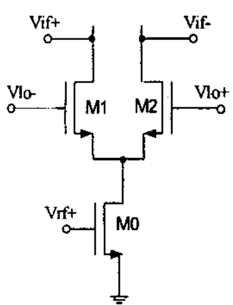

Figure 3. Single-balanced mixer

2.2. No LO mismatch

First consider the M0 biasing current. In order to reflect the behavior of sub-micrometer transistors, short channel modulation effects should be taken into account. An acceptable MOSFET model in the saturation region is

$$I_{bias} = K(V_{GS0} - V_{th})^2[1 + \lambda(V_{DS0} - V_{eff0})] \quad (2)$$

where V_{GS0} and V_{eff0} are the gate-source voltage and saturation voltage of M0 in Fig. 3, V_{th} is the threshold voltage, K is the current factor. λ is the channel length modulation coefficient. V_{DS0} is the drain-source voltage of M0, that is, the common source voltage of the switching pair. So it can be written as

$$V_{DS0} = V_{LObias} + V_{LO-} - V_{eff1} - V_{th} \quad (3)$$

where V_{LObias} is the biasing voltage of the switching pair, $V_{LO-} = A\sin\omega_{LO}t$ is the LO signal injected to M1. V_{eff1} is the saturation voltage of M1. Substitute Eq. (3) into Eq. (2), and rearrange it. We get

$$I_{bias} = I_0[1 + \lambda(V_{LO-} + C)] = (1 + \lambda C)I_0 + \lambda I_0 V_{LO-} \quad (4)$$

in which

$$I_0 = K(V_{GS0} - V_{th})^2$$

$$C = V_{LObias} - V_{eff0} - V_{eff1} - V_{th}$$

Note the M0 biasing current is not a constant current now, but a function of V_{LO-}.

Now let us get each drain current of the switching pair. For positive values of the LO voltage, M2 switches ON and M1 switches OFF, and a current equal to I_{bias} appears in the right branch. In the next half period, the current switches to the left branch. Then, each drain current of the switching pair should be equal to I_{bias} times a square-wave at the frequency of ω_{LO} with no DC value,

$$S(t) = \frac{1}{2} + \sum_{n=1}^{\infty} \frac{2}{n\pi} \sin n\omega_{LO}t, \quad (5)$$

where n is an odd integer.

$$\begin{aligned} I_{D1} &= I_{bias} \times S(t) \\ &= (1 + \lambda C)I_0 \times (\frac{1}{2} + \frac{2}{\pi}\sin\omega_{LO}t + \frac{2}{3\pi}\sin 3\omega_{LO}t \\ &\quad + \frac{2}{5\pi}\sin 5\omega_{LO}t + \cdots\cdots) + \frac{\lambda I_0 A}{\pi} + \frac{\lambda I_0 A}{2}\sin\omega_{LO}t \\ &\quad - \frac{2\lambda I_0 A}{\pi}(\frac{1}{1\times 3}\cos 2\omega_{LO}t + \frac{1}{3\times 5}\cos 4\omega_{LO}t + \cdots\cdots) \end{aligned} \quad (6)$$

In the same way, we get I_{D2}.

$$\begin{aligned} I_{D2} &= I_{bias} \times S(t - \frac{T_{LO}}{2}) \\ &= (1 + \lambda C)I_0 \times (\frac{1}{2} - \frac{2}{\pi}\sin\omega_{LO}t - \frac{2}{3\pi}\sin 3\omega_{LO}t \\ &\quad - \frac{2}{5\pi}\sin 5\omega_{LO}t - \cdots\cdots) + \frac{\lambda I_0 A}{\pi} - \frac{\lambda I_0 A}{2}\sin\omega_{LO}t \\ &\quad - \frac{2\lambda I_0 A}{\pi}(\frac{1}{1\times 3}\cos 2\omega_{LO}t + \frac{1}{3\times 5}\cos 4\omega_{LO}t + \cdots) \end{aligned} \quad (7)$$

We are interested in the common source current of the switching pair, so the sum of the I_{D1} and I_{D2} is

$$\begin{aligned} I_{common\_source} &= I_{D1} + I_{D2} \\ &= (1 + \lambda C)I_0 + \frac{2\lambda I_0 A}{\pi} \\ &\quad - \frac{4\lambda I_0 A}{\pi}(\frac{1}{1\times 3}\cos 2\omega_{LO}t + \frac{1}{3\times 5}\cos 4\omega_{LO}t + \cdots) \end{aligned} \quad (8)$$

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From above, we can see all the odd harmonics have been cancelled out including the fundamental frequency ω_{LO}. Therefore there are no fundamental frequency signals that can feedthrough to the RF port. So an ideal mixer has no LO leakage and DC offset.

2.3. LO phase mismatch only

Let us consider only the fundamental frequency term of I_{common_source}. Assume there is a mismatch of θ degrees between LO+ and LO–, then

$$\begin{aligned} I_{common\_source} &= I_{D1} + I_{D2} \\ &= D\sin(\omega_{LO}t + \frac{\theta}{2}) + D\sin(\omega_{LO}t + \pi - \frac{\theta}{2}) \\ &= 2D\cos\omega_{LO}t\sin\frac{\theta}{2} \approx D\theta\cos\omega_{LO}t \end{aligned} \tag{9}$$

where $D = [\frac{2(1+\lambda C)}{\pi} + \frac{\lambda A}{2}]I_0$ is the coefficient of fundamental frequency term of I_{D1} or I_{D2} and if $\theta/2$ is quite small. Hence if there is a small phase mismatch between LO+ and LO–, it will generate a small current of fundamental frequency which is proportional to the phase mismatch at the common source of the switching pair. This small current is injected into M0 from its drain and induces a voltage of fundamental frequency at the gate, which is considered to be the LO leakage. Now let us use the small signal equivalent circuit of MOS in Fig. 4 to obtain the relation between that small current and the LO leakage. Fig. 5 shows the small signal equivalent circuit of M0 in Fig. 3.

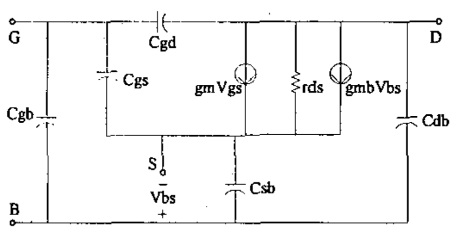

Figure 4. Small signal equivalent circuit of MOS

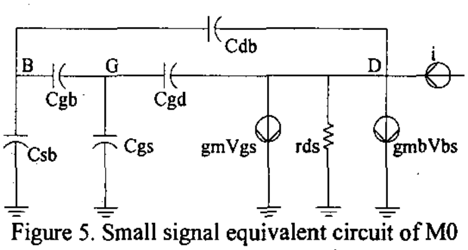

Figure 5. Small signal equivalent circuit of M0

After solving equations, we get

$$\begin{aligned} v_g &= \frac{i}{g_m + (g_{mb} + g_{ds})(1 + C_{gs}/C_{gd}) + sC_{gs}} \\ &\approx \frac{i}{g_m + (g_{mb} + g_{ds})(1 + C_{gs}/C_{gd})} \approx \frac{i}{g_m} \end{aligned} \tag{10}$$

where below 10GHz sC_{gs} is quite small comparing to the real part of the denominator, so it is ignored. As a result, LO leakage is proportional to the LO phase mismatch and its magnitude is approximately $D\theta/g_m$.

2.4. LO amplitude mismatch only

If there exists an amplitude mismatch ε between LO+ and LO–, the fundamental frequency term of common source current is

$$\begin{aligned} I_{common\_source} &= I_{D1} + I_{D2} \\ &= [\frac{2(1+\lambda C)}{\pi} + \frac{\lambda(A+\varepsilon/2)}{2}]I_0\sin\omega_{LO}t \\ &+ [\frac{2(1+\lambda C)}{\pi} + \frac{\lambda(A-\varepsilon/2)}{2}]I_0\sin(\omega_{LO}t + \pi) \\ &= \frac{\lambda I_0}{2}\varepsilon\sin\omega_{LO}t \end{aligned} \tag{11}$$

Hence if there is a small amplitude mismatch between LO+ and LO–, it will generate a small current of fundamental frequency which is proportional to the mismatch amplitude at the common source of the switching pair. Combining with Eq. (10), we can see that the LO leakage is proportional to the LO amplitude mismatch and its magnitude is approximately $\lambda I_0\varepsilon/(2g_m)$.

2.5. Both of LO phase and amplitude mismatch

In general case, the two differential LO signals have both phase mismatch θ and amplitude mismatch ε.

$$\begin{aligned} I_{common\_source} &= I_{D1} + I_{D2} \\ &= [\frac{2(1+\lambda C)}{\pi} + \frac{\lambda(A+\varepsilon/2)}{2}]I_0\sin(\omega_{LO}t + \theta/2) \\ &+ [\frac{2(1+\lambda C)}{\pi} + \frac{\lambda(A-\varepsilon/2)}{2}]I_0\sin(\omega_{LO}t + \pi - \theta/2) \\ &\approx [\frac{2(1+\lambda C)}{\pi} + \frac{\lambda A}{2}]I_0\theta\cos\omega_{LO}t + \frac{\lambda\varepsilon I_0}{2}\sin\omega_{LO}t \\ &= I_{phase\_mismatch} + I_{amplitude\_mismatch} \end{aligned} \tag{12}$$

It is found that when the phase mismatch and amplitude mismatch both occur, the common source current of fundamental frequency is just equal to the sum of the currents caused by each of them. So the superposition principle applies here. In this case, we get the LO leakage is

$$|V_{LOleakage\_total}|^2 = |V_{phase\_mismatch}|^2 + |V_{amplitude\_mismatch}|^2 \tag{13}$$

3. SIMULATION RESULTS AND CONCLUSIONS

In order to get the common mode LO leakage, a CMOS double-balanced Gilbert mixer operating at the centre frequency of 2.45GHz for direct conversion applications was designed using Cadence SpectreRF for the Chartered 0.18um CMOS Digital/Analog/RF process. Fig. 6 shows the relation of the total LO leakage to the amplitude mismatch for several fixed phase mismatches. It is found that when the phase mismatch is small, the total LO leakage is simply proportional to the amplitude mismatch. When the phase mismatch is large enough, the curve

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representing the relation between the total leakage and amplitude mismatch has been proved to be in the shape of hyperbola. It can be explained by Eq. (13). Fig. 7 shows the relation of the total LO leakage to the phase mismatch for several fixed amplitude mismatches. Similar conclusions can be found.

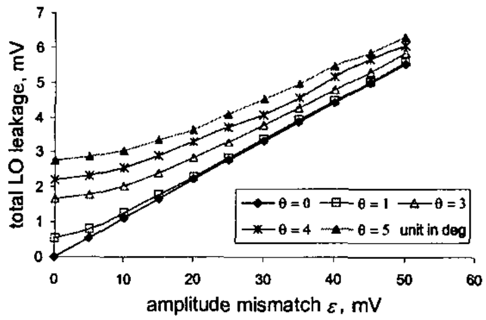

Figure 6. Total LO leakage vs. amplitude mismatch

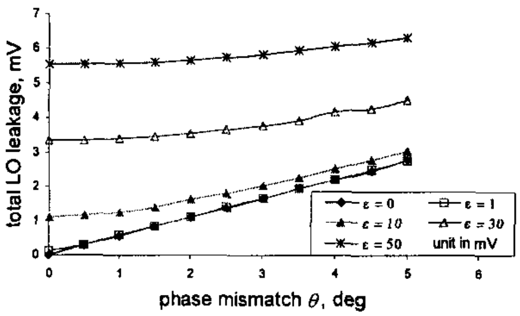

Figure 7. Total LO leakage vs. phase mismatch

This paper presents the theoretical analysis of LO leakage caused by the LO mismatch in double-balanced CMOS Gilbert mixers. It is found that if there is no LO mismatch, an ideal mixer will not have an LO leakage and DC offset. If a phase mismatch or amplitude mismatch exists, an LO leakage is proportional to the mismatch value. If both of them occur at the same time, the amplitude square of LO leakage is equal to the sum of amplitude square of each contribution. Simulation results are also shown. It can be seen that the LO mismatch could cause a large feedthrough and induce a DC offset.

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