# RF Mixer design, principles

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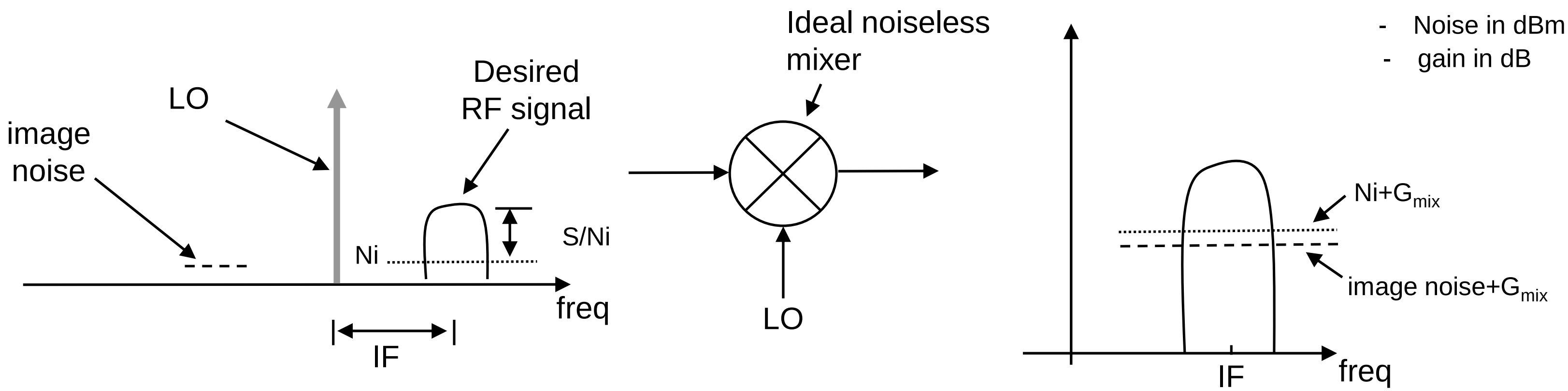
# SSB vs. DSB NF definition:




Because of the image problem, a receive mixer down converts both desired and the image bands to IF frequency. This means folding the noise at the image frequency on top of the desired band at IF.

Therefore, the total noise at IF is as follows:

1. The noise at desired RF band down converted to IF
2. the noise at image RF band down converted to IF
3. The noise added by the mixer noisy circuit itself.

There are several definitions to express the mixer noise figure due to the issue of image noise folding. We will consider one definition at a time.

## Single-side band (SSB) NF definition:

The single-side band NF definition assumes that there is no signal at the image frequency except the source noise. This definition is useful in finite IF receiver architectures. The NF is the degradation of S/N at mixer output. Therefore, one can write:

$$S_{out} = S_d G_{mix\_d}$$

$$N_{out} = N_d G_{mix\_d} + N_{im} G_{mix\_im} + N_{mix\_d} G_{mix\_d} + N_{mix\_im} G_{mix\_im}$$

where $S_d$ is the desired signal, $N_d$ is the noise in the desired band, $N_{im}$ is the noise in the image band, $G_{mix\_d}$ and $G_{mix\_im}$ is the mixer gain at both desired and image frequencies, respectively. $N_{mix\_d}$ and $N_{mix\_im}$ are the desired and image band noise due the mixer circuit itself referred to its input. To simplify analysis, we will assume $N_d = N_{im}$ , $N_{mix\_d} = N_{mix\_im}$ and $G_{mix\_d} = G_{mix\_im}$, therefore,

$$N_{out} = 2N_d G_{mix\_d} + 2N_{mix} G_{mix\_d}$$

$$\Rightarrow \frac{S_{out}}{N_{out}} = \frac{S_d}{2N_d + 2N_{mix}} = \frac{S_d}{N_d}\left(\frac{1}{2 + \frac{2N_{mix}}{N_d}}\right)$$

$$\Rightarrow F_{SSB} = 2 + \frac{2N_{mix}}{N_d}$$

As seen from the SSB noise figure equation, if the mixer is noiseless ($N_{mix}$=0), the mixer SSB NF is **3dB** because of the image noise folding. It is important to know that this definition is the one used by microwave mixer designers for years. It is also the definition used in SpectreRF simulator.

## The IEEE definition of Single-Side band (SSB) NF:

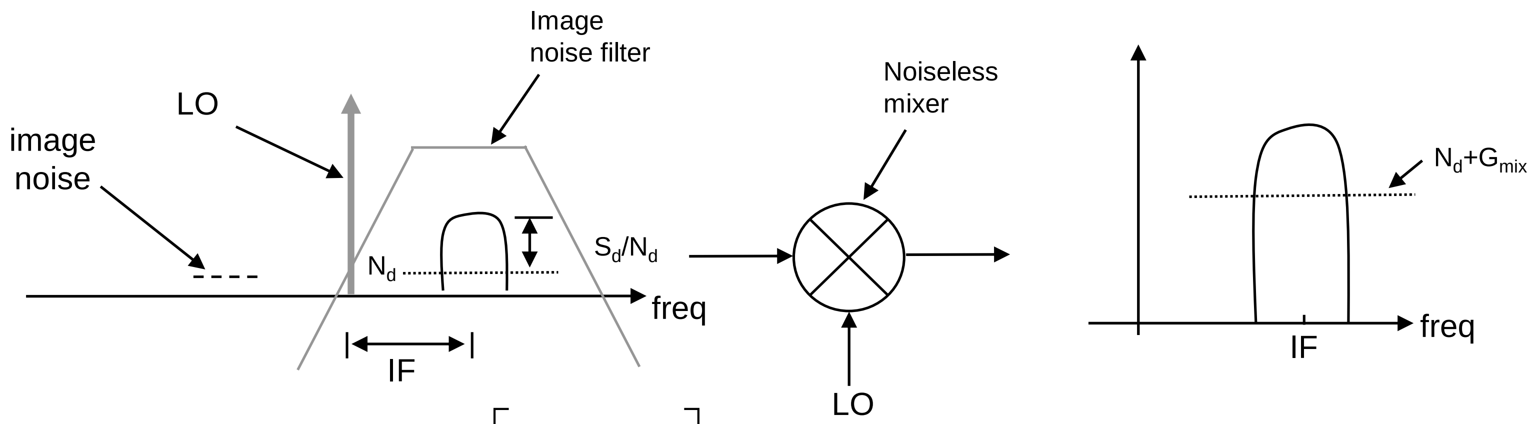
The IEEE has a slightly different definition for SSB NF. It argues that the mixer should not be “penalized” by the image source noise folding. The only image noise folding that is allowed to count towards calculating the mixer SSB NF is that which is due to the mixer circuitry itself. The input image noise should not be counted. This definition is useful in finite IF receive architectures in which there is either a sharp bandpass image filter to attenuate both image and image noise to a negligible level, or if there is an image-rejection baseband scheme that removes image signal and its noise. As a result, one can write

$$S_{out} = S_d G_{mix\_d}$$

$$N_{out} = N_d G_{mix\_d} + N_{mix\_d} G_{mix\_d} + N_{mix\_im} G_{mix\_im}$$

To simplify analysis, we will assume $N_d = N_{im}$ and $G_{mix\_d} = G_{mix\_im}$, therefore,

$$N_{out} = N_d G_{mix\_d} + 2 N_{mix} G_{mix\_d}$$

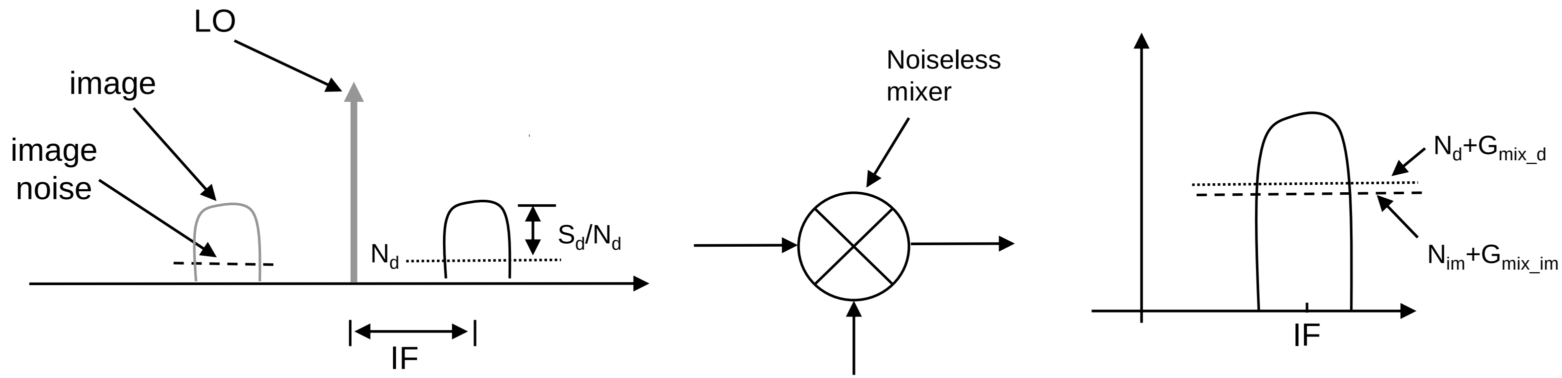



$$\Rightarrow \frac{S_{out}}{N_{out}} = \frac{S_d}{N_d + 2N_{mix}} = \frac{S_d}{N_d}\left[\frac{1}{1+\frac{2N_{mix}}{N_d}}\right]$$

$$\Rightarrow F_{SSB\_IEEE} = 1 + \frac{2N_{mix}}{N_d}$$

As seen from the IEEE SSB equation, if the mixer is noiseless, the mixer SSB NF is actually 0dB, the spirit behind the new definition.

# The Double-Side band (DSB) NF:




The double-side band NF definition assumes that the image band contains both noise and an image signal identical to the desired band signal. This definition is useful in direct-conversion receiver where the image is the signal itself. Therefore, one can write:

$$S_{out} = S_d G_{mix\_d} + S_{im} G_{mix\_im}$$

$$N_{out} = N_d G_{mix\_d} + N_{im} G_{mix\_im} + N_{mix\_d} G_{mix\_d} + N_{mix\_im} G_{mix\_im}$$

where $S_{im}$ is the image signal.

To simplify analysis, we will assume $S_d = S_{im}$ , $N_d = N_{im}$ , $N_{mix\_d} = N_{mix\_im}$ and $G_{mix\_d} = G_{mix\_im}$

therefore,

$$N_{out} = 2N_d G_{mix\_d} + 2N_{mix} G_{mix\_d}$$

Therefore:

$$\Rightarrow \frac{S_{out}}{N_{out}} = \frac{2S_d}{2N_d + 2N_{mix}} = \frac{S_d}{N_d}\left[\frac{1}{1+\frac{N_{mix}}{N_d}}\right]$$

$$\Rightarrow F_{DSB} = 1 + \frac{N_{mix}}{N_d}$$

It can be seen that the difference between the SSB NF and DSB NF is exactly 3dB. However, with the SSB_IEEE NF, the difference to DSB NF is not exactly 3dB. In fact the difference between the SSB_IEEE and the SSB NF approaches 3dB as the mixer NF is very high. The SSB_IEEE noise factor can be related to that of the DSB as:

$$F_{SSB\_IEEE} = 2F_{DSB} - 1$$

# Single-balanced v.s. double-balanced mixers:

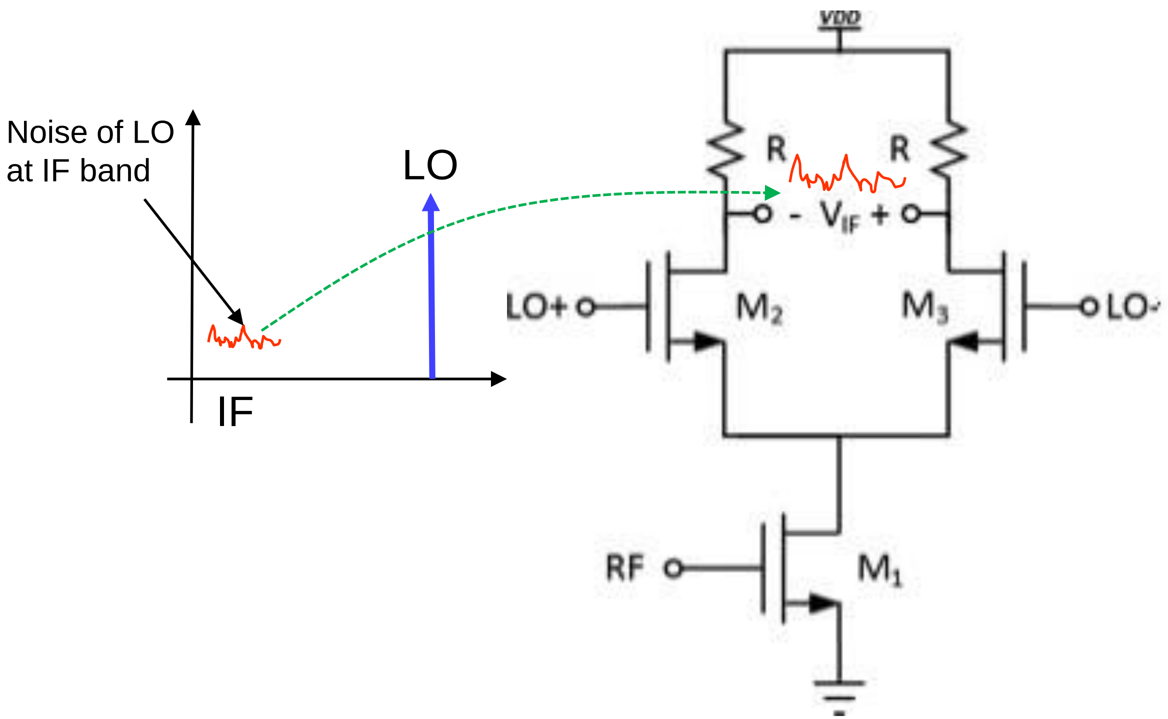
The word “single-balanced” means that one port of the mixer is balanced and can be bipolar in sign, whereas the other port is single-ended and is unipolar. With this arrangement, the signal at the unipolar port can be written as an RF signal riding on top of a DC component, where the sum of the RF signal plus DC carries always one sign. The bipolar signal on the balanced port (usually the LO) can be written as a simple LO signal as follows:

$$V_{RF} = A_{dc} + \cos(\omega_{RF} t)$$

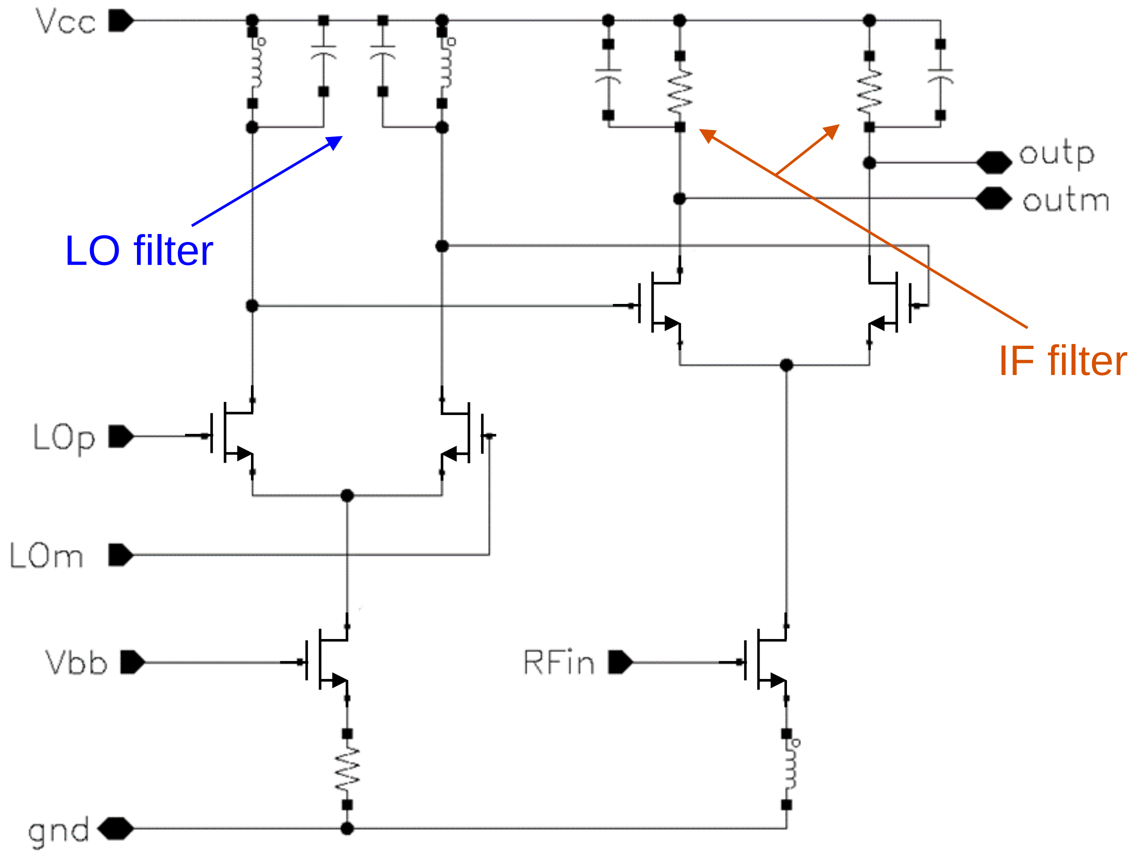
$$V_{LO} = \cos(\omega_{LO} t)$$

$$V_{IF} = V_{RF} \times V_{LO} = A_{dc} \cos(\omega_{LO} t) + \frac{1}{2}\cos(\omega_{IF})t + \frac{1}{2}\cos(\omega_{RF} + \omega_{LO})t$$

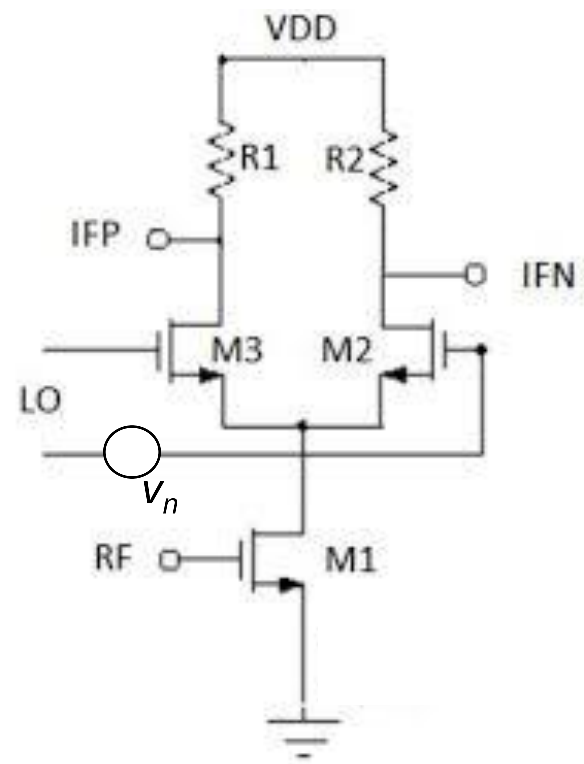
it is apparent that the LO signal shows up at the output of a single-balanced mixer, even in the absence of any RF signal. This generally has the disadvantage of increasing the composite signal swing at the output, limiting the available headroom, unless the LO signal is properly filtered. Another disadvantage is the sensitivity to common-mode noise and interference as will be shown next. However, the advantage of single-balanced mixers is their simplicity and generally low power.




The above circuit is an example of a single-balanced mixer. The diff-pair M2-M3 behave like a differential amplifier to the LO signal. M1 acts in this case as a current source in the absence of the RF signal. With this arrangement, any signal, or noise, occupying the IF band at the LO port will get amplified and transferred right to the mixer output on top of desired IF signal causing significant degradation to the mixer NF.




To reduce the sensitivity to LO noise, a bandpass filter is placed at the LO port to filter out any LO noise at the IF band, as shown above. It is important here to note that the effect of the LO noise can be highly suppressed if the LO signal is an ideal square wave with zero rise and fall time. This can be shown as follows




The LO noise can be modeled as a voltage source in series with the base of the diff-pair M2-M3. If the LO signal is an ideal square wave from a zero source impedance with a large voltage swing, the time over which both devices are on at the same time is zero. Therefore the impact of the LO noise is eliminated. However, if the LO signal has a finite rise and fall time, both M2 and M3 will be on at the same time, during LO transitions, acting like an amplifier to the LO noise as well as to their own noise, degrading the overall mixer NF.

# Double-balanced mixers:

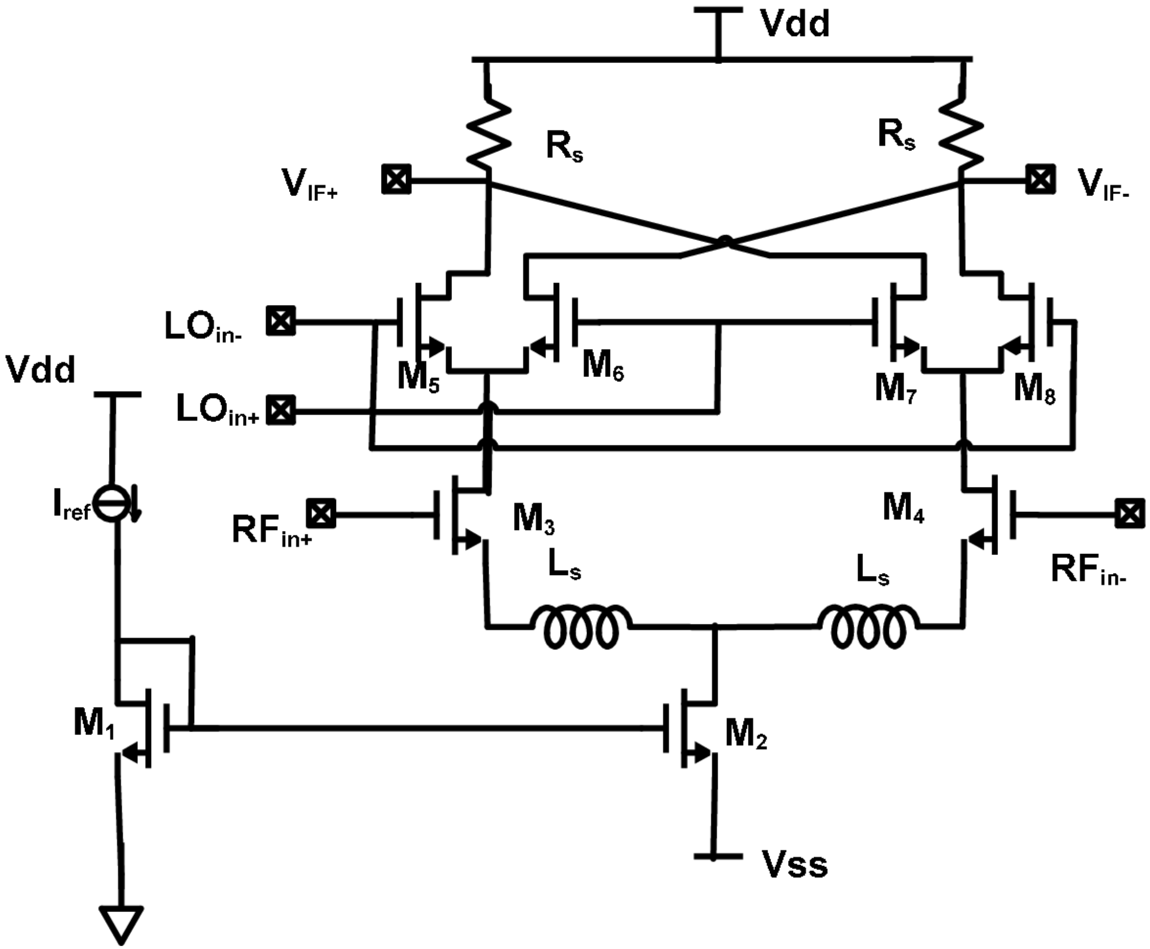
The word "double-balanced" means both mixer inputs (LO and RF) ports are bipolar in sign. Double-balanced mixers have the advantage of suppressing the LO signal at the IF output port. This is because the RF signal is bipolar with zero mean, therefore LO*RF will not result in an LO component at the output.

$$V_{RF} = \cos(\omega_{RF} t)$$
$$V_{LO} = \cos(\omega_{LO} t)$$
$$V_{RF} \times V_{LO} = 0.5\cos(\omega_{IF})t + 0.5\cos(\omega_{RF} + \omega_{LO})t$$

This also means that the noise in the IF band of the LO signal will get upconverted by the RF signal, preventing any additional S/N degradation. This can be seen as follows. Let us assume the IF band noise at the LO is modeled as a tone. Thus, one can write

$$V_{RF} = \cos(\omega_{RF} t)$$
$$V_{LO} = \cos(\omega_{LO} t) + v_n \cos(\omega_{IF} t)$$
$$V_{RF} \times V_{LO} = 0.5\cos(\omega_{IF})t + 0.5\cos(\omega_{RF} + \omega_{LO})t + 0.5 v_n \cos(\omega_{RF} + \omega_{IF})t + 0.5 v_n \cos(\omega_{RF} - \omega_{IF})t$$
$$\Rightarrow LPF(V_{RF} \times V_{LO}) = 0.5\cos(\omega_{IF})t$$

# Double-balanced mixer topologies: Gilbert cell mixer

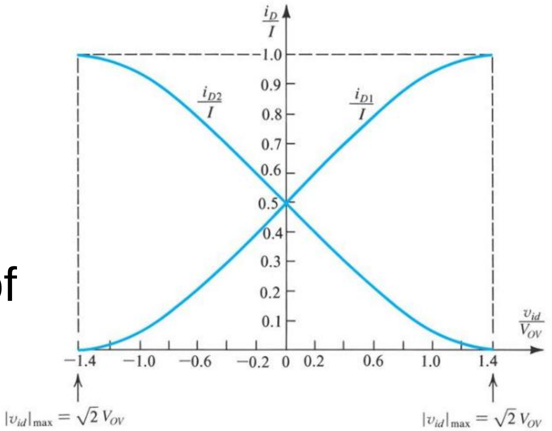



Current-source M2 can be replaced by either a choke or an LC-tank for better headroom and better noise

The Gilbert cell mixer is widely used in downconverter mixers because of the following:

1. Very good port isolation. This proven to be crucial to achieve good IP2 as will be seen later
2. They provide decent gain
3. Can easily be integrated on-chip
4. Moderate LO drive is needed (typically Vpp of

3~4 x Vdsat of switching quad transistors)




1. Adequate NF in the range of ~6-12dB SSB, depending on gain and IIP3.

In the following, we will discuss the performance of this Gilbert-cell topology and the some optimization techniques.

# Gain:

The mixer voltage gain is the ratio of the output voltage signal at the IF frequency to the RF input voltage signal. If the LO is at either peak, one pair of the quad transistors is completely off, while the other acts like a cascode device. In this case, one can think of the mixer as an amplifier with gain of ~$R_L/Z_{SS}$, where $R_L$ is the differential load resistor and $Z_{SS}$ is the degeneration impedance of the input diff-pair, which is really noting but a Gm stage.

Now, let us assume the LO signal is a perfect square wave normalized to its peak value (which is large enough to turn on completely one quad or the other). The LO signal can then be written as:

$$V_{LO} = \frac{2}{\pi}\cos(\omega_{LO}t) + \frac{1}{\pi}\cos(3\omega_{LO}t) + \frac{1}{6\pi}\cos(5\omega_{LO}t) + .....$$

Therefore, the gain $R_L/Z_{SS}$ will be modulated by the LO signal, since the IF signal is the product of RF*LO. Therefore, the gain will suffer a loss of $2/\pi$ or 3.9dB.

$$Gain_{mix} \approx \frac{2}{\pi}\frac{R_L}{Z_{SS}}$$

This result assumes that for the input Gm diff pair, $1/g_m \ll Z_{SS}$, and that the device $f_T \gg f$. Of course, the gain equation is simplified, but it gives some insight.

**noise:**

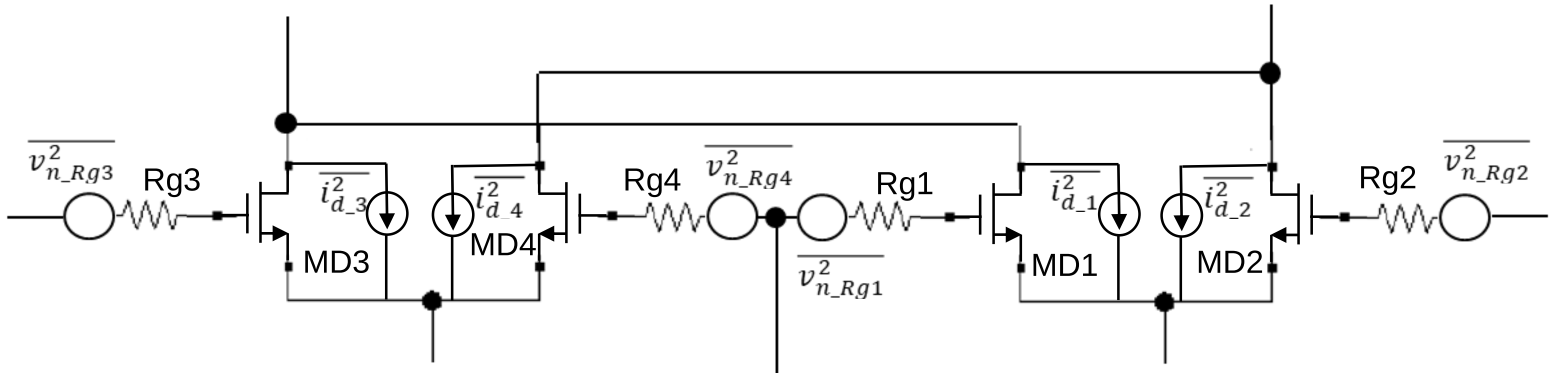
The noise in the Gilbert cell mixer is divided between the input Gm cell, the switching quad and the mixer load. We will addressed each section separately.

Noise in the Gm cell:

The noise in the Gm cell follows more or less the NF optimization theory and technique described in the single-ended CS NF optimization lecture. However, usually the noise of the Gm cell is a bit compromised by increasing the source degeneration in order to increase the mixer IIP3. This is because the mixer linearity has more impact on the overall Rx performance than its NF (to an extent). The NF contribution of the Gm cell to the overall mixer NF is roughly 2~3dB.

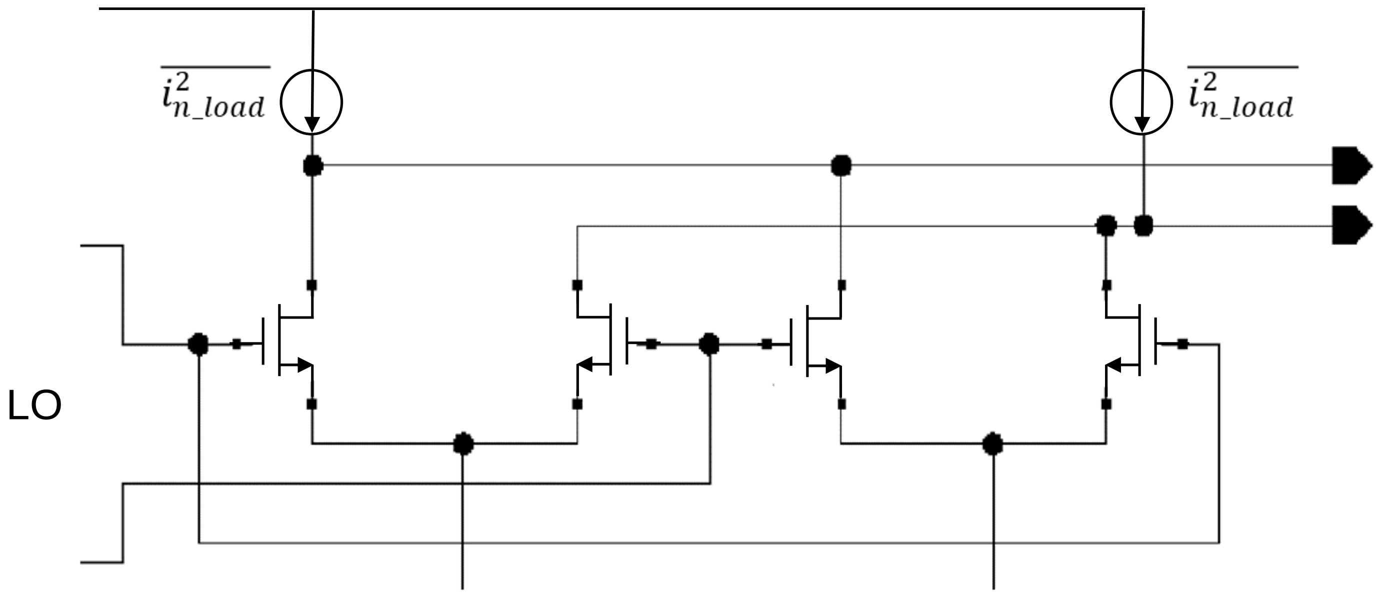
Noise due to mixer quad:

The noise in the quad has two parts. The first part is simply the folding of the image noise coming from the Gm cell, as well as the input source. This happens even if the mixer quad is totally noiseless. The second source of mixer quad noise is the one due to the quad noisy devices. It is important to shed more light on the quad noise as follows.




**Switching qud**

- The noise of the quad is at its maximum when each of the quad devices is conducting an equal amount of current. This is because when MD3-MD4 devices are ON at the same time, they act as a simple diff-pair amplifying each other's noise (for example gate resistance noise), which is uncorrelated. This noise adds up at the output severely degrading the S/N. Since the quad diff-pairs are not degenerated, the amplifying gain of this noise is quite large.
- When both transistors are on (during transition), their drain noise at IF frequency also leaks to output. Outside transition, ON device acts as cascode → its noise does not appear at output
- The time when both quad diff-pair are partially on is during the LO transition. This means in order to suppress the quad noise, the LO transition (rise and fall time) must be as sharp as possible and close to an ideal square signal as possible.




<u>Noise due to mixer load:</u>
The mixer load also contributes to the overall NF. In low or zero IF receivers, the load is a simple resistor, whose noise contributes to the mixer NF. In some cases the mixer is configured as an OTA (operational transconductance amplifier) with the mixer output being current. This means the mixer is designed to have active loads with high output impedance. The noise of such active load can be significant if not designed properly. Note that the load noise is referred back to input by dividing it over the mixer Gm, so the larger the input diff-pair Gm is, the less the load noise contribution becomes.

# Linearity of a Gilbert-cell mixer:

<u>IP3:</u>

The mixer IP3 is usually limited by the input Gm cell. The 3rd order linearity of the Gm cell depends on the amount of degeneration used, the type of degeneration (inductive or resistive) and the bias current. The reader is referred to [1] for analysis of mixer Gm linearity. Inductive degeneration is widely used for its low noise and the higher linearity it provides compared to resistive or capacitive degeneration. In highly degenerated input Gm cells with relatively large bias current, the mixer linearity will be limited by both Gm stage as well as the mixer quad. The 3rd order nonlinearity of the mixer quad is discussed in detail in [2].

## IP2:

The mixer IP2 is one of the most important spec for low-IF or direct conversion receivers. In fact, in some systems like CDMA, mixer IP2 is the limiting spec in making a direct conversion CDMA receiver even feasible. The IP2 of the Gilbert-cell mixer relies on circuit symmetry as well as LO drive duty cycle as will be discussed in detail next.

- IM2 generated within the Gm stage:

Due to finite linearity of the Gm cell, its transfer function can be written as:

$$I_{out} = G_m V_{in} = g_{m1} V_i + g_{m2} V_i^2 + g_{m3} V_i^3 + ..$$

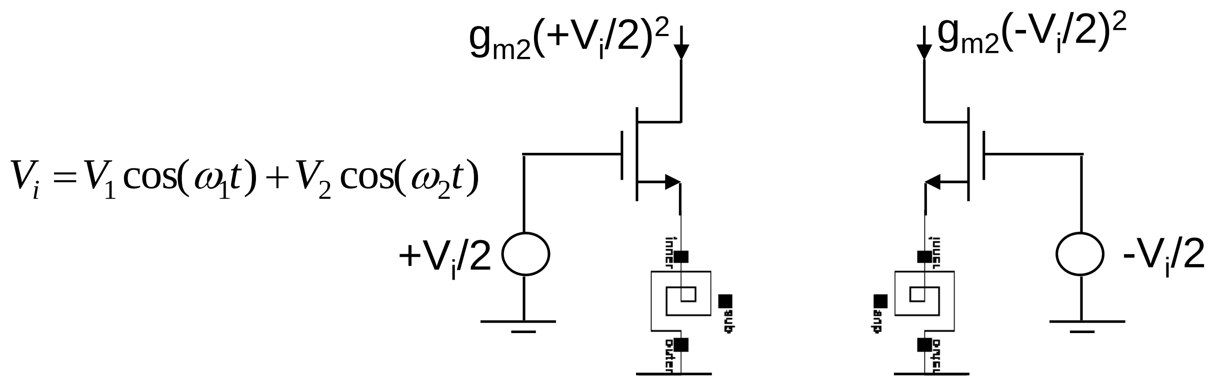
Let us assume the inputs to the mixer are:

$$V_i = V_1 \cos(\omega_1 t) + V_2 \cos(\omega_2 t) \quad \text{where} \;\; \omega_1 - \omega_2 \leq \omega_{IF}$$

Substituting $V_i$ into the Gm equation, taking only the second order distortion into account yields:

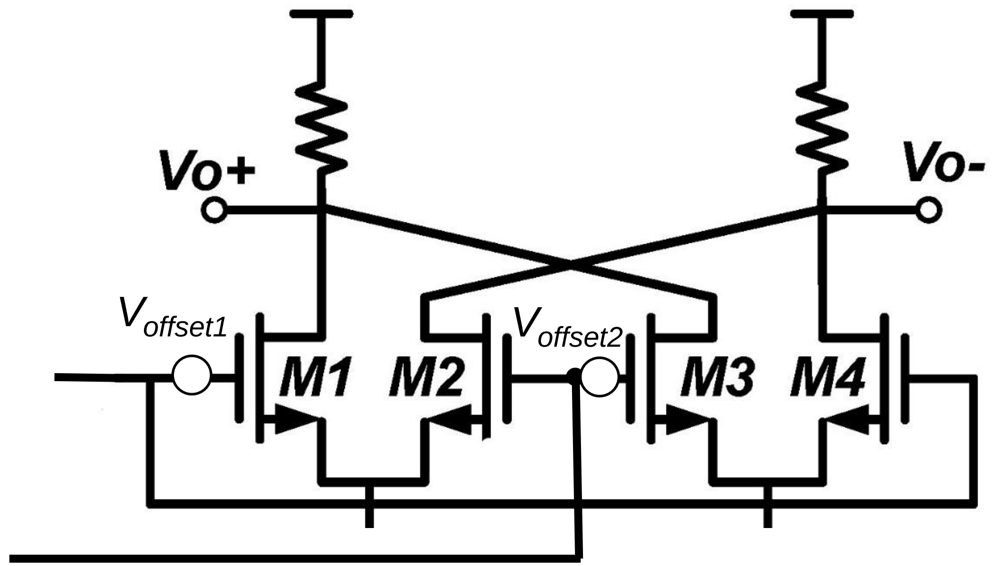
$$I_{out} = G_m V_{in} = g_{m2} V_1 V_2 \cos(\omega_1 - \omega_2)t + ..$$

Note that the IM2 generated in each half circuit of the Gm stage diff-pair are equal in both magnitude and sign, assuming perfect matching. This is because of the square exponent in the distortion equation. Therefore, the IM2 signal shows up as a common-mode signal at the output current of the Gm-stage, and so the differential value of the IM2 is zero, if the Gm cell is perfectly symmetrical. This can be seen in the following:




Any mismatch in the Gm stage half circuit (due to circuit components or layout) results in a finite differential IM2 at the output. This low frequency IM2 gets upconverted by the LO when it reaches the quad and so won't appear at the IF output of the mixer. However, any DC offset in the LO quad will result in this IM2 differential component to appear at the mixer output. This can be verified as follows.

Any DC offset in the quad can be modeled as a DC voltage source in series with an offset-free quad. This way, the LO signal driving the offset-free quad can be written as:




$$V_{LO} = V_{offset} + A\cos(\omega_{LO} t)$$

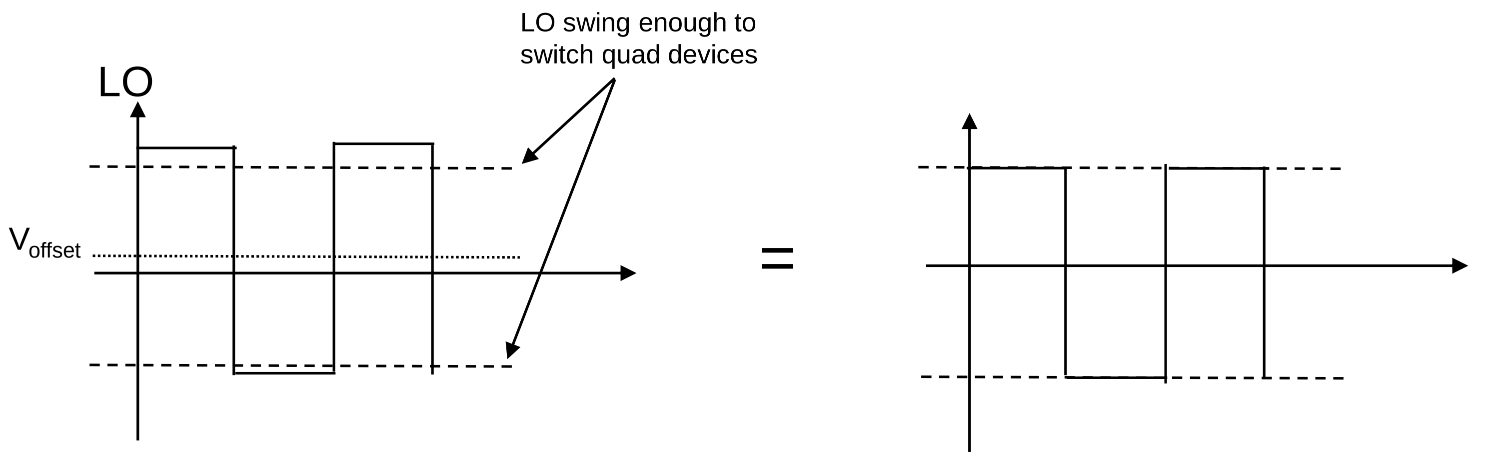
with the IM2 signal coming out of the Gm cell being:

$$V_{Gm-IM2} = g_{m2} V_1 V_2 \cos(\omega_1 - \omega_2)t$$

the IF output then is (after lowpass filter):

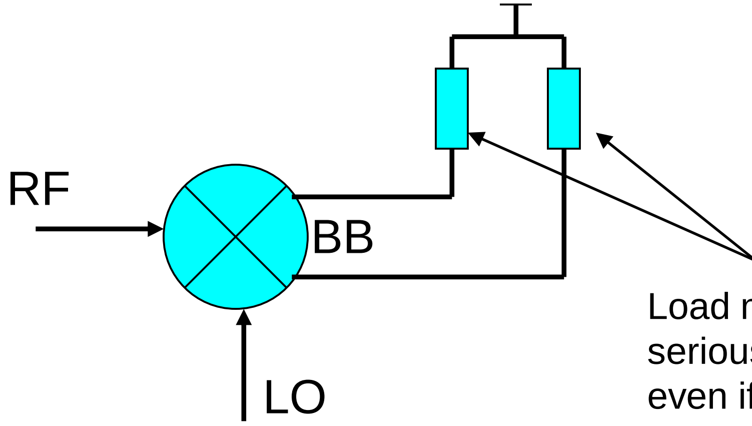
$$V_{IF} = V_{LO} \times V_{Gm-IM2} = V_{offset} g_{m2} V_1 V_2 \cos(\omega_1 - \omega_2)t$$

Note that the more the LO signal gets closer to an ideal square wave (with high enough amplitude), the less the quad DC offset impact on the mixer IP2 becomes.




## Reducing IM2 mechanisms due to quad offset

- Minimize offsets in the quad by using optimum device size.
- Properly match the DC bias of LO switching quad by sharing that same DC bias node.
- Spend a lot of time optimizing the LO buffer to get the sharpest LO edges possible (5 to 10 V/ns is obtainable in modern technology)

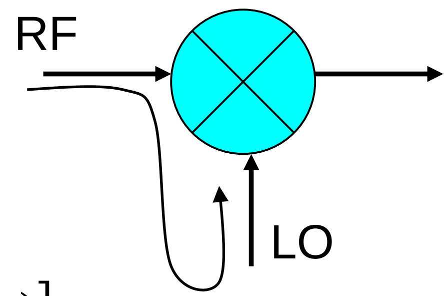



Load mismatch can seriously degrade IP2, even if everything else is perfect

● IM2 generated within the quad itself:
The output current of the Gm cell, carrying the two RF tones, passes through the quad, which acts as a current commutating stage. Being highly non-linear, the quad itself has a finite second order distortion, resulting in generating a low-frequency IM2 spur out of the two RF tones. Just like the Gm stage case, the IM2 spur is a common-mode signal if the quad circuit is perfectly symmetric. However, this IM2 generated within the quad, although common-mode, can show up as a finite differential mixer load matching signal at the output of the mixer if the mixer load is not perfectly matched. In fact, is one of the most important factor in achieving high mixer IP2.

● IM2 due RF to LO leakage:

If the isolation of mixer ports is not sufficient, appreciable amount of RF input power "leaks" to the LO port causing the RF signal to mix with itself.




$$V_{RF} = \cos(\omega_1 t) + \cos(\omega_2 t)$$

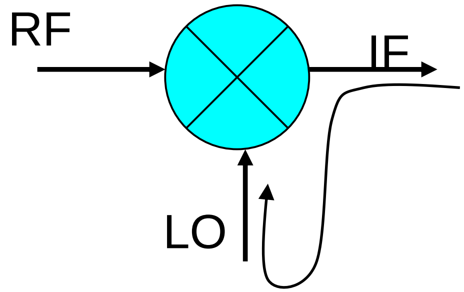
$$V_{RF\to LO} = V_{leak}\left[\cos(\omega_1 t) + \cos(\omega_2 t)\right]$$

$$V_{IF} = V_{RF} \times V_{RF\to LO} = V_{leak}\left[\cos(\omega_1 t) + \cos(\omega_2 t)\right]^2 = V_{leak}\cos(\omega_1 - \omega_2)t$$

Extreme care in mixer layout is necessary to ensure proper RF to LO isolation. Use all differential mixer topology (all mixer ports being differential)

● IM2 due baseband to LO leakage:

The mixer down converts desired signal as well as the close in blockers, all of which appear at the mixer output IF port. In some systems the blockers can be as high as 60dB above desired signal and can swing as high as 1Vpp at the mixer output. If portion of this baseband blocker signal at the mixer output leaks to the LO port, it will mix with the large LO signal within the quad itself and then mix with the RF signal, generating an inband IM2 spur as shown next:




$$V_{RF} = \cos(\omega_1 t) + \cos(\omega_2 t)$$

$$V_{IF} = \cos(\omega_1 - \omega_{LO})t + \cos(\omega_2 - \omega_{LO})t \quad \text{, where } \omega_1 - \omega_2 \text{ is close to desired } \omega_{\mathrm{IF}}$$

$$V_{IF \to LO} = V_{leak}\left[\cos(\omega_1 - \omega_{LO})t + \cos(\omega_2 - \omega_{LO})t\right]$$

$$V_{LO} \times V_{IF \to LO} = 0.5 V_{leak}\left[\cos(\omega_1 t) + \cos(\omega_2 t)\right]$$

$$V_{LO} \times V_{IF \to LO} \times V_{RF} = 0.5 A V_{leak}\left[\cos(\omega_1 t) + \cos(\omega_2 t)\right]^2 = A V_{leak} \cos(\omega_1 - \omega_2)t$$

- IM2 due LNA:

The blockers pass first through the LNA, which amplifies them before they reach the mixer input. Due to finite second order nonlinearity of the LNA, a low frequency IM2 appears at the LNA output as a result. If the LNA output is DC coupled to the mixer input, this IM2 will get amplified by the very large low-frequency gain of the mixer Gm cell, especially if the mixer is inductively generated. This IM2 will have a similar effect in degrading the mixer IP2 to that due mixer Gm itself. Therefore, it is essential to AC couple the LNA output to the mixer input to prevent such IM2 mechanism. Doing so will eliminate the LNA IP2 from affecting the receiver IP2 (which usually it is mixer and baseband filter limited).

## References:

[1] K. Fong, R. G. Meyer "High-Frequency Nonlinearity Analysis of Common-Emitter and Differential-Pair Transconductance Stages," IEEE JSSC, Vol. 33, No. 4, April 1998, pp. 548-555.

[2] R. G. Meyer, "Intermodulation in High-Frequency Bipolar Transistor Integrated-Circuit Mixers," IEEE JSSC, Vol. sc-21, No. 4, August 1986, pp. 534-537.

[3] D. Manstretta, M Brandolini and F. Svelto, "Second-Order Intermodulation Mechanisms in CMOS Downconverters," IEEE JSSC, Vol. 38, No. 3, March 2003.

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[5] K. Kivekas, A. Parssinen, K. Halonen, "Characterization of IIP2 and DC-Offsets in Transconductance Mixers," IEEE TCAS-II, Vol. 48, No. 11, November 2001, pp. 1028-1038.

[6] D. Coffing, E. Main, "Effects of Offsets on Bipolar Integrated Circuit Mixer Even-Order Distortion Terms," IEEE Transaction on Microwave Theory and Techniques, Vol. 49, No. 1, January 2001, pp. 23-30.

[7] Steven Maas, *Microwave Mixers*, Artech House Publishers; 2nd edition,1992, ISBN: 0890066051.