Mixers II

As we have seen before mixers* play an important role in communication systems. On the transmitter side they perform up-conversion and on the receiver side they down-convert. We will focus mainly on down-conversion mixers.

In this section, we will discuss:

• mixer fundamentals

• mixer metrics

• mixer implementations in Bipolar and CMOS

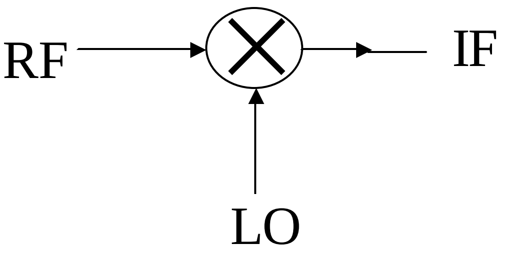
* This Part 2 material on mixers is mainly contributed by Dr. S. H. K. Embabi's previous notes

Mixer Fundamentals

A mixer is expected to produce signals with spectral components which do not exist in the input signal (a 900MHz mixed with a 800MHz produces components at 100MHz and 1700MHz). Only ***nonlinear*** or ***time-varying*** devices can perform such a function.

Most mixer implementations use some kind of a *multiplication* of two signals, the signal to be down-converted (*RFI* or *IF*) and the signal whose frequency determines the output frequency (LO). If we assume that

$$V_{RF}\cos(w_{RF}t)\times V_{LO}\cos(w_{LO}t)$$
$$=\frac{V_{RF}V_{LO}}{2}\left[\cos(w_{RF}-w_{LO})t+\cos(w_{RF}+w_{LO})t\right]$$



The desired spectral component can be acquired by filtering the sun term. The amplitude of the product (*IF*) is proportional to the *RF* signal if the amplitude of the LO is constant.

Mixer Metrics

Conversion Gain

The conversion gain may be measured as ***voltage*** or ***power gain***. The *voltage conversion gain* of a mixer is defined as follows:

$$Voltage\ Conversion\ Ganin = \frac{rms\ voltage\ of\ the\ IF\ signal}{rms\ voltage\ of\ the\ RF\ signal}$$

The *power conversion gain* of a mixer is defined as follows:

$$Power Conversion\,(transducer)\,Gain = \frac{the\ IF\ power\ delivered\ to\ the\ load}{rms\ available\ RF\ power\ from\ the\ source}$$

If the load and source impedance are equal, the voltage and power conversion gains are equal in dB. In the case where the mixer is driven by a filter (image reject filter) the input impedance of the mixer must provide proper termination (50Ω) to the filter otherwise the filter will exhibit ripples. The mixer's output sees the IF filter, which may be passive. The input impedance of such a passive filter is typically not equal to 50Ω. Therefore, the source impedance is different than the load impedance the voltage and power conversion gain would be different.

Port-to-Port Isolation

Signals may leak through different mechanisms from one port to the other.

We may have:

- LO-to-RF leakage, which causes self-mixing (problem for zero-IF). The LO leakage may even reach the antenna through the LNA (recall the LNA has nonzero return "gain").

- RF-to-LO feedthrough allows interferers and spurs present in the RF signal to interact with the LO.

- LO-to-IF feedthrough may cause desensitization of consequent blocks (remember that the LO power may be greater than that of the desired IF signal).

- RF-to-IF feedthrough causes problems in some architectures such as zero-IF because of the leakage of low-frequency even-order intermod. Products (even-order distortion).

Either the mixer's isolation must be good enough or the surrounding components must be tolerant to leakage.

Mixer Implementations

As mentioned before, mixers use multiplication to achieve frequency translation. Multiplication can be implemented either directly, using a multiplier circuit, or indirectly, using a nonlinear circuit.

Let us consider the second option first.

Mixers using Nonlinear Circuits

Consider a nonlinear circuit with the following input-output relation:

$$v_{out} = a_0 + a_1 v_{in} + a_2 v_{in}^2 + a_3 v_{in}^3 + a_4 v_{in}^4 + \cdots$$

If v_{in} is the sum of the *RF* and the *LO* signals, the output would have spectral components at:

- DC (resulting from the *even*-order nonlinear terms)

much smaller than IF and can thus be filtered out

- Harmonics of the *RF* ($m w_{RF}$) and harmonics of the LO ($m w_{LO}$)

much higher than IF and is filtered out (example: RF=900MHz and LO=830MHz, hence IF=70MHz, while harmonics of RF are at above 900MHz and those of the LO are above 800 MHz)

- Intermodulation Products $\left(p w_{RF} \pm q_{LO}\right)$. The desired IF is intermod. Product with $q=p=1$. The rest are undesired. Unfortunately, some of them are too close to the desired IF to be selectively suppressed through filtering. One needs to estimate the power of the intermod spurs for a certain mixer nonlinearity, i.e. given the IP_3 and 1-dB_c how much power will the spurs have? There are programs which can do such estimation. This would help in designing or chosing the proper mixer and/or the filter following the mixer.

Mixer using "Square-Law" Circuits

Next we will consider nonlinear circuits with second-order nonlinearity which are used to implement mixers. Assuming that the input is:

$$v_{in} = v_{RF}\cos(w_{RF}t) + v_{LO}\cos(w_{LO}t)$$

The output of the square-law circuit is given by:

$$\begin{aligned} v_o &= a_1[v_{RF}\cos(w_{RF}t) + v_{LO}\cos(w_{LO}t)] && \textit{Fundamental} \\ &+ a_2\left[\{v_{RF}\cos(w_{RF}t)\}^2 + \{v_{LO}\cos(w_{LO}t)\}^2\right] && \textit{SquareTerms} \\ &+ 2a_2 v_{RF}[\{\cos(w_{RF}t)\}\{\cos(w_{LO}t)\}] && \textit{Cross}\,\mathrm{Pr}\,\textit{oduct} \end{aligned}$$

The square terms produce the DC and 2nd harmonics spectral components, while the cross product term can be rewritten as:

$$v_{crossprod.} = \alpha_2 v_{RF} v_{LO}\left[cos(\omega_{RF} - {}_{LO})t + cos(\omega_{RF} + \omega_{LO})t\right]$$

After filtering the sum term, the DC and the harmonics, we are let with the desired difference term. It is obvious that the undesired signal are distant from the IF and hence can be easily filter out. *This makes the square-law circuits more useful than circuits with higher order nonlinearity.*

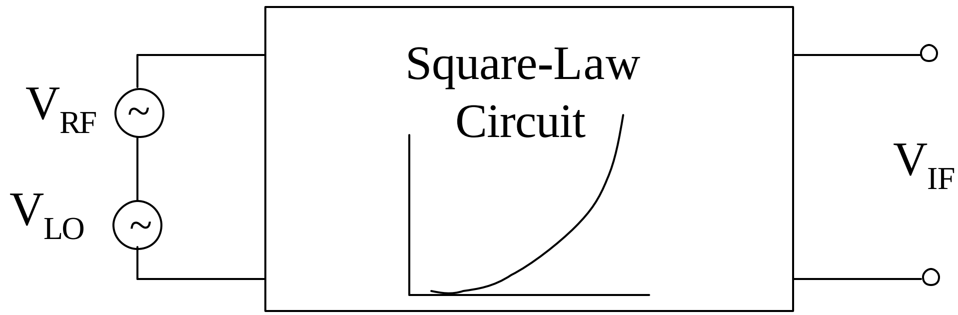
Since the amplitude of the LO is constant (v_{LO}=*constant*). The output will be proportional to the RF signal and is given by:

$$v_o = a_2 v_{RF} v_{LO} \left[\cos(w_{RF} - w_{LO})t\right]$$

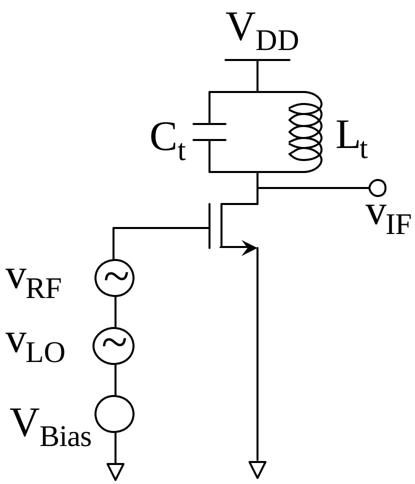
The conversion given can now be easily derived as

$$G_c = \frac{a_2 v_{RF} v_{LO}}{v_{RF}} = a_2 v_{LO}$$

Notice that mixer with nonlinear circuits can be implemented using ***two-port networks***.

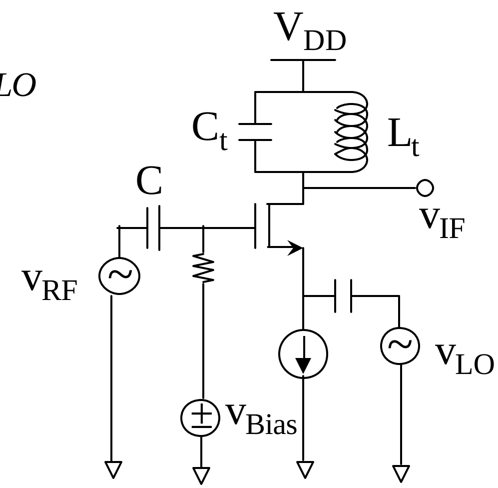


A good approximation for a square-law device is a MOSFET (with long enough channel). Two examples of MOSFET based mixers are shown below. It can be shown that the (voltage) conversion gain is



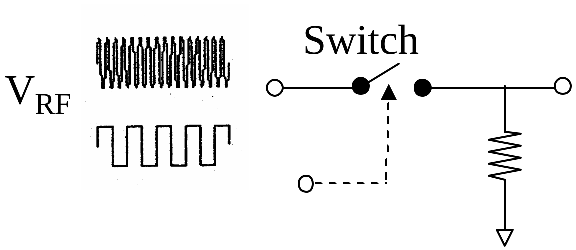
$$G_c = \frac{m_n C_{ox} W}{2L} v_{LO}$$

$$\alpha_2 = \frac{\mu_n C_{ox} \omega}{2L}$$

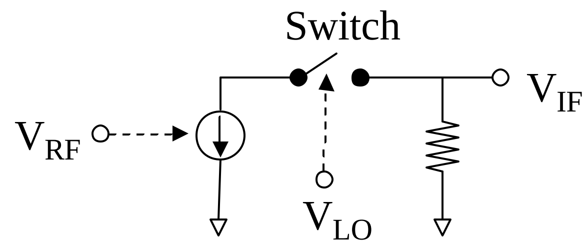


Passive vs Active Mixers

Passive mixers provide no power gain (but may have either current ***or*** voltage gain), instead they have conversion loss.



Passive Mixer

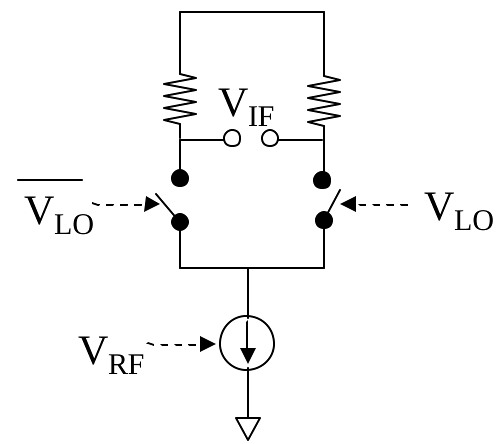


Active Mixer

Active mixers provide power gain using active devices.

The advantage of active mixers is that its gain reduces the contribution of the noise of the subsequent blocks.

The advantages of passive mixers over active mixers is the superior linearity and speed.

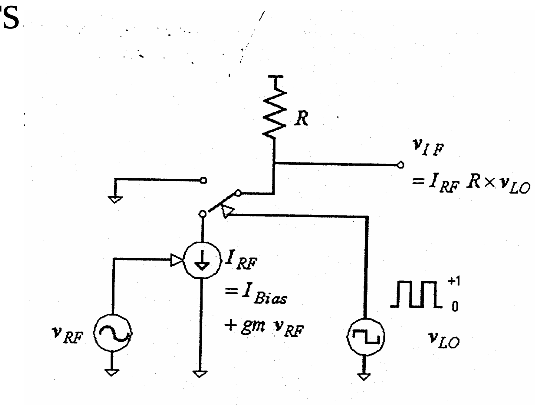


Example:
Active Mixer

Multiplier-Based Mixers

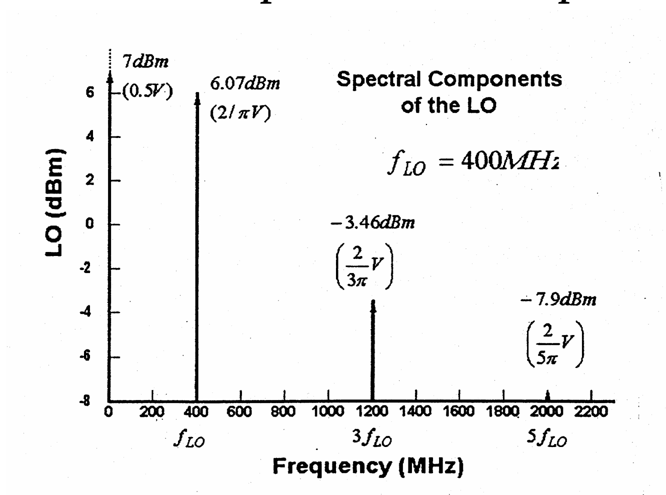
These can be classified as ***single-balanced and double-balanced*** mixers. Both implementations, which will be introduced next, are current mode. First, we will consider, single-ended and differential single-balanced mixers.

Unbalanced Mixers



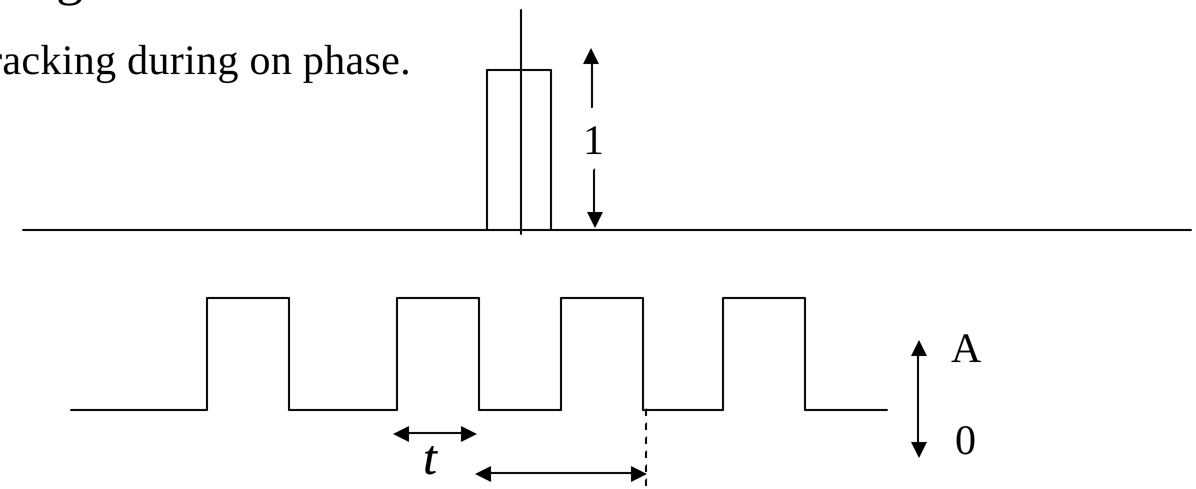
The RF signal is converted to a current signal using a transconductance. This current is switched to the IF port via a switch-wave signal. The output IF current is switching between I_{RF} and zero at the LO frequency. Therefore, it can be expressed, assuming 50% duty cycle, as:

$$I_{IF} = \left(I_{BIAS} + gm\, v_{RF}\right)$$
$$\times\left[0.5 + \frac{2}{p}\cos w_{LO}t + \frac{2}{3p}\cos 3w_{LO}t + \frac{2}{5p}\cos 5w_{LO}t + \cdots\right]$$

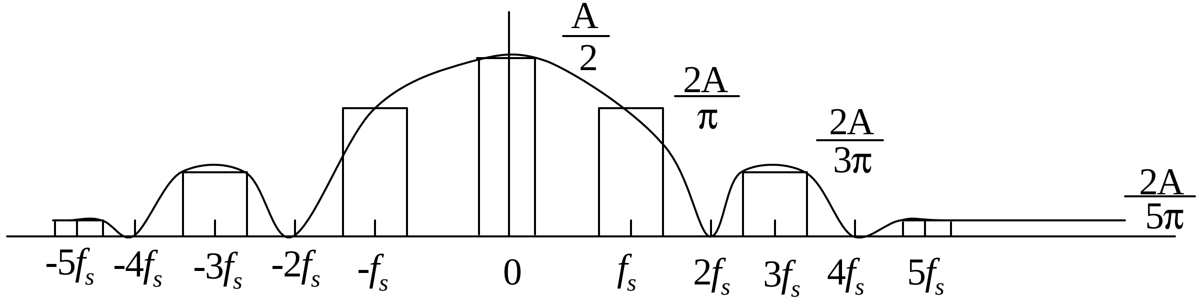


Natural Sampling

This is basically tracking during on phase.

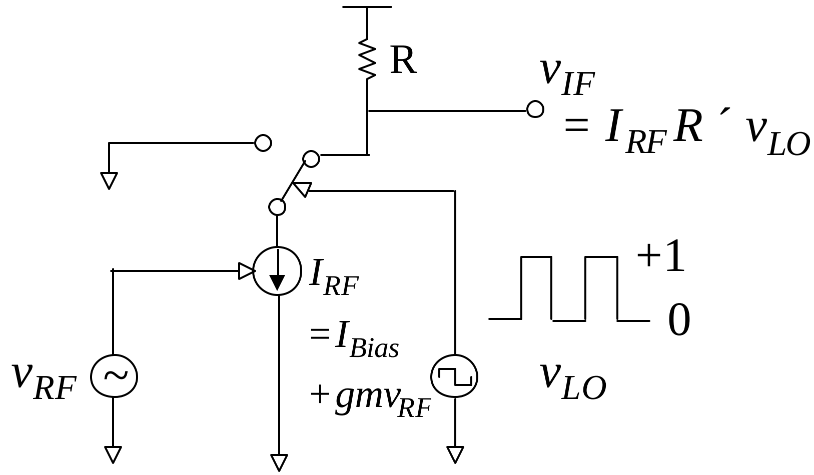


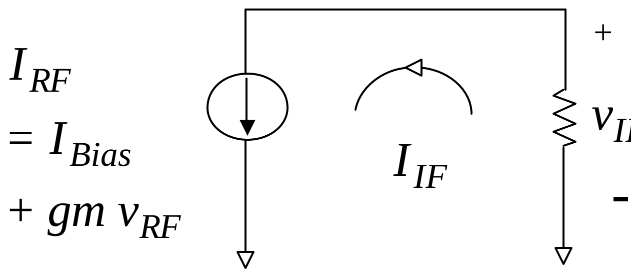
$$T = 2t = 1/f_s$$



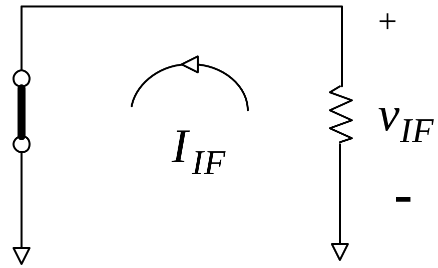
For a ***50% duty cycle*** the images will appear at the odd multiples of f_s, and the amplitude of the images is given by:

- A/2 f=0
- 2A(nπ) $f=nf_s$ where n=1,2,5,7, …





On Phase #1



On Phase #2

The current at the IF-port switches between I_{RF} and zero.

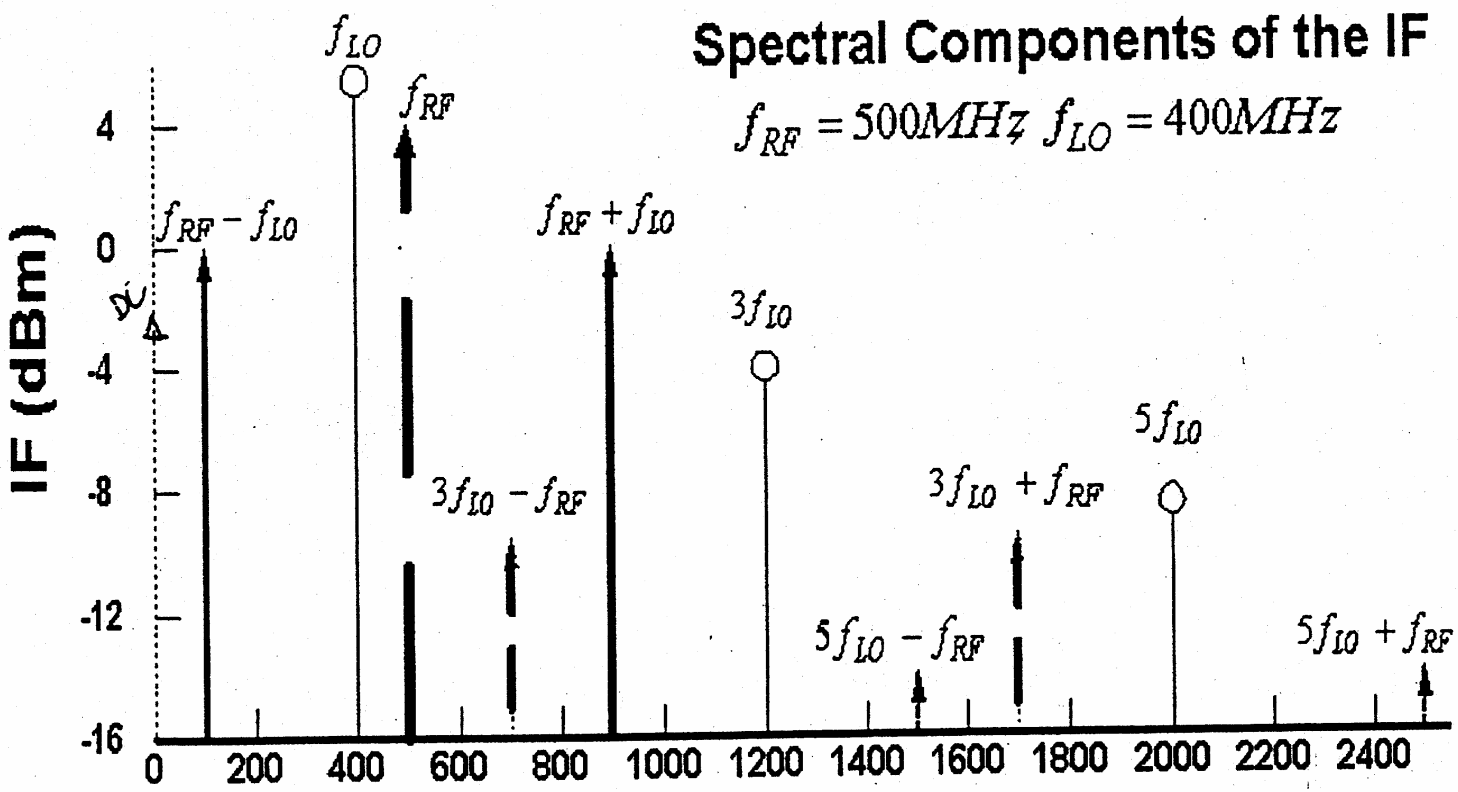
Assuming that $v_{RF} = V_{RF}\cos\omega_{RF}t$, the current at the IF port becomes

$$I_{IF} = \frac{I_{BIAS}}{2}\left[1 + \frac{4}{\pi}\cos\omega_{LO}t + \frac{4}{3\pi}\cos 3\omega_{LO}t + \frac{4}{5\pi}\cos 5\omega_{LO}t + \cdots\right] + \frac{gm}{2}V_{RF}\cos\omega_{RF}t$$

$$+ gmV_{RF}\begin{bmatrix} \frac{1}{\pi}\cos(\omega_{LO} - \omega_{RF})t + \frac{1}{\pi}\cos(\omega_{LO} + \omega_{RF})t \\ + \frac{1}{3\pi}\cos(3\omega_{LO} - \omega_{RF})t + \frac{1}{3\pi}\cos(3\omega_{LO} + \omega_{RF})t \\ + \frac{1}{5\pi}\cos(5\omega_{LO} - \omega_{RF})t + \frac{1}{5\pi}\cos(5\omega_{LO} - \omega_{RF})t\cdots \end{bmatrix}$$

The above expression shows several spectral components at :

- DC
- nw_{LO} where $n = 1,3,5\cdots$
- $nw_{LO} \pm w_{RF}$ where $n = 1,3,5,\cdots$
- w_{RF}



Frequency (MHz)

The desired signal is the component at w_{RF} - w_{LO}. Its amplitude is

$$I_{IF}\Big|_{w=w_{RF}-w_{LO}} = \frac{gm}{p} V_{RF}$$

The conversion transconductance is

$$G_c = \frac{gm}{p}$$

The disadvantages of the unbalanced topology are:

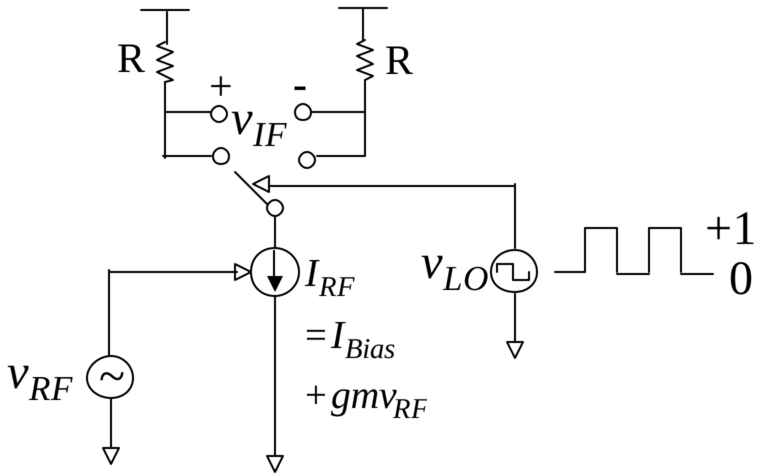
- LO-IF feedthrough,

  A spectra component at ω_{LO} appears at the IF-port. If the LO frequency is not far enough from the desired RF, it may be difficult to attenuate the LO component enough via filtering.

- RF-IF feedthrough (or *direct feedthrough*).

  An RF spectra components shows at the IF-port. Direct feedthrough worsens the NF of the mixer because it allows the noise at the RF-port at the desired IF frequency to leak to the IF-port.

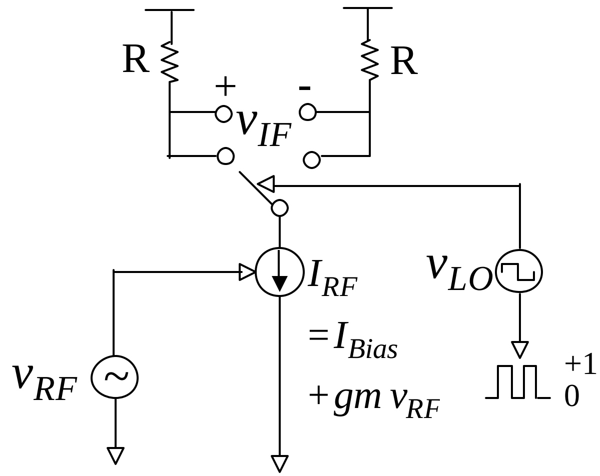
Single-Balanced Mixers

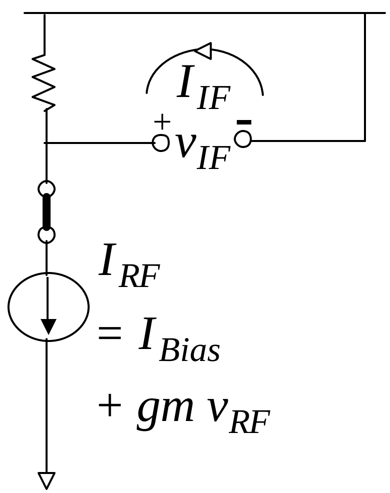


The mixer process is similar to that of unbalanced mixer, except that now the output IF current is switching between I_{IF} and $-I_{RF}$ at the LO frequency. Therefore, it can be expressed as:

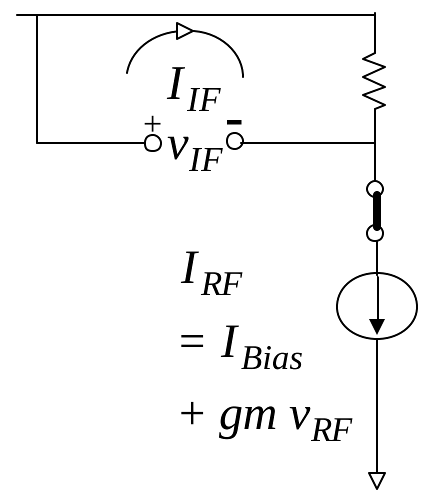
$$I_{IF} = \left(I_{BIAS} + gm\, v_{RF}\right) \times \left[\frac{4}{\pi} cos\, \omega_{LO} t + \frac{4}{3\pi} cos\, 3\omega_{LO} t + \frac{4}{5\pi} cos\, 5\omega_{LO} t + \cdots\right]$$

Notice that the difference between this equation and that of the unbalanced is the DC term (between the square brackets). In the single-balanced the IF current is switching between two "almost" equal levels with opposite signs, so the average is zero. The amplitude of the spectral components of the square wave (terms of the series) has doubled, again, because the IF current is switching between $+/-I_{RF}$. While, in the unbalanced mixer, the current swing is halved.





On Phase #1



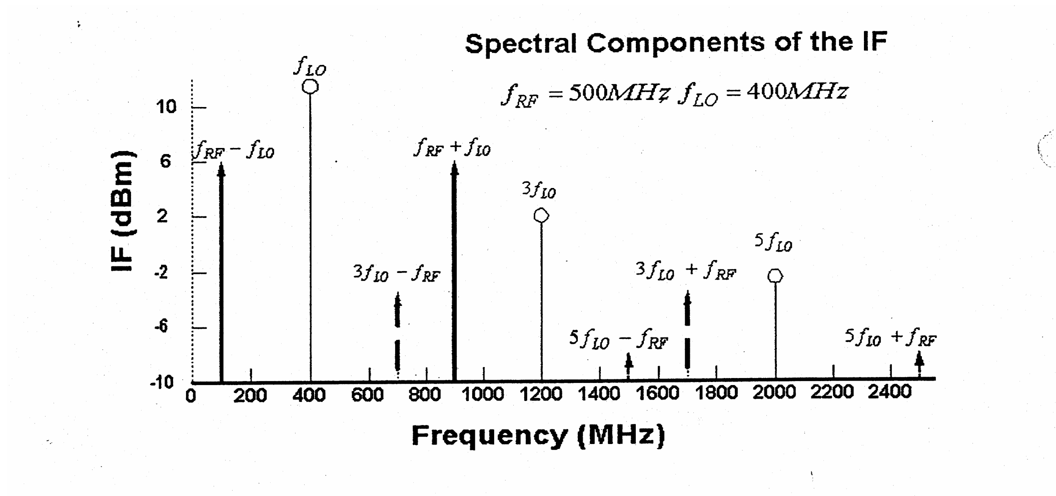
On Phase #2

On phase #1, $I_{IF} = I_{BIAS} + gm\ v_{RF}$, while, on the 2nd phase, the current is equal to that of phase #1 but with opposite sign.

Again, assuming that $v_{RF} = V_{RF} \cos \omega_{RF} t$, the current at the IF port becomes

$$I_{IF} = I_{BIAS}\left[2 + \frac{4}{\pi}\cos\omega_{LO}t + \frac{4}{3\pi}\cos 3\omega_{LO}t + \frac{4}{5\pi}\cos 5\omega_{LO}t + \cdots\right]$$

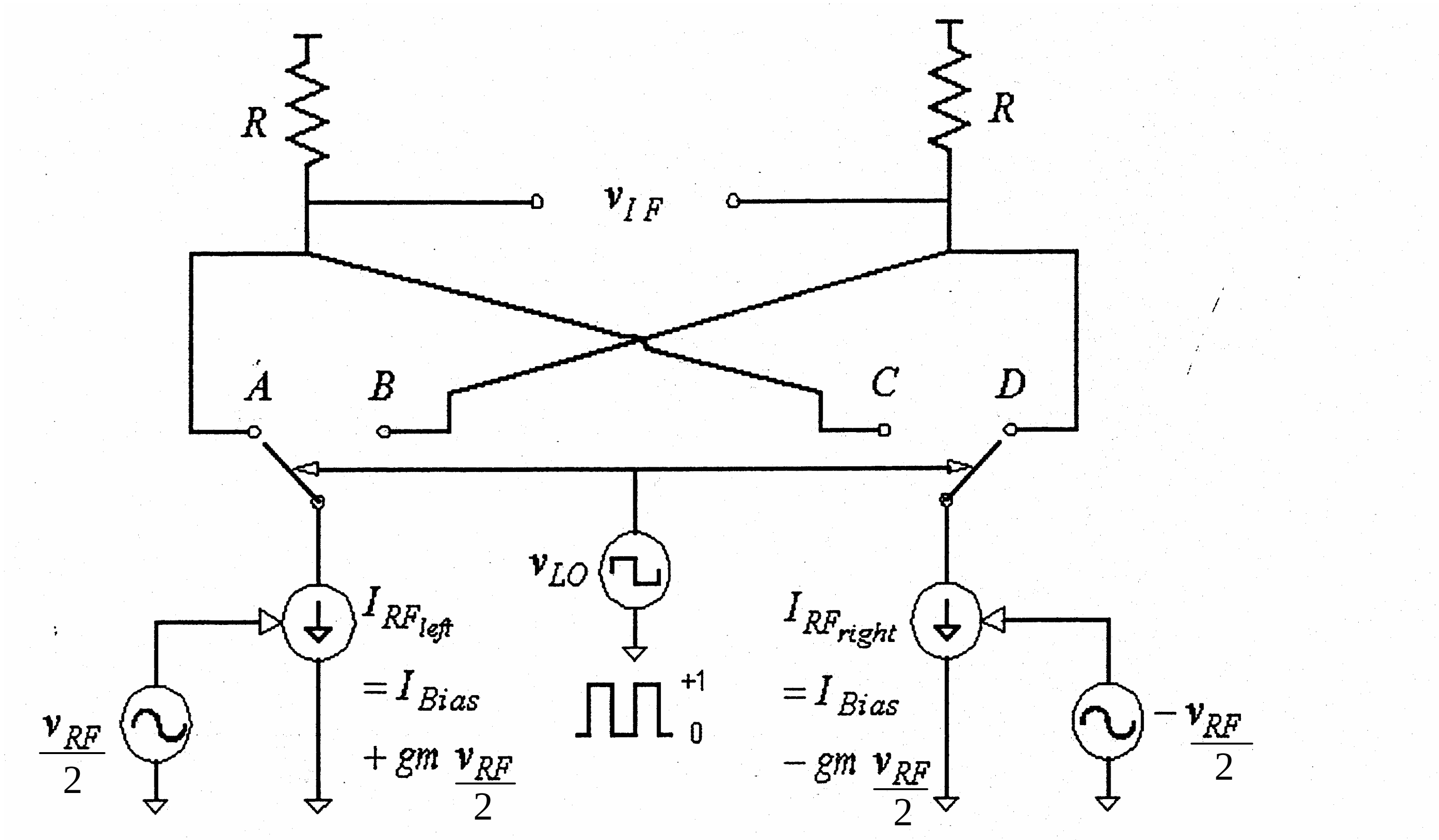
$$+ gmV_{RF}\left[\begin{array}{l} \frac{2}{\pi}\cos(\omega_{LO} - \omega_{RF})t + \frac{2}{\pi}\cos(\omega_{LO} + \omega_{RF})t \\ + \frac{2}{3\pi}\cos(3\omega_{LO} - \omega_{RF})t + \frac{2}{3\pi}\cos(3\omega_{LO} + \omega_{RF})t \\ + \frac{2}{5\pi}\cos(5\omega_{LO} - \omega_{RF})t + \frac{2}{5\pi}\cos(5\omega_{LO} -_{RF})t + \cdots \end{array}\right]$$



The spectral components at the IF port of the unbalanced single-balanced mixer are similar to those of the unbalanced mixers except that the RF component is suppressed (no RF to IF feedthrough. All other components are increased by 6dB. The conversion transconductance is:

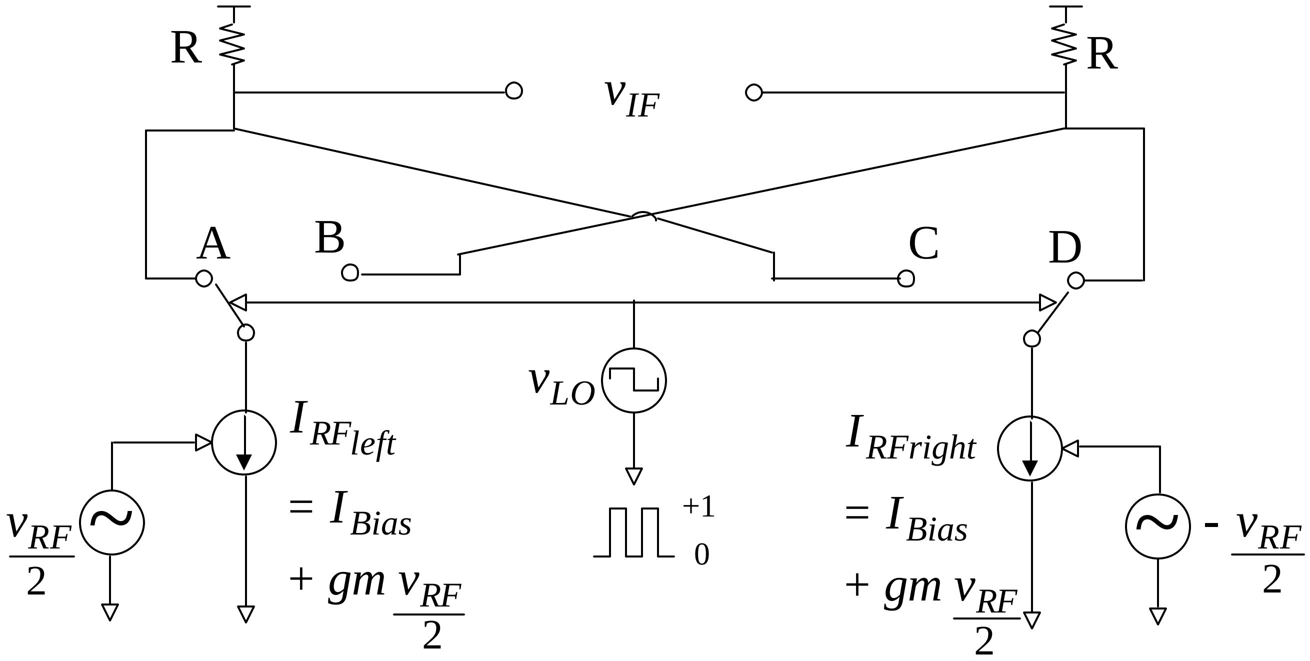
$$G_c = \frac{2gm}{p}$$

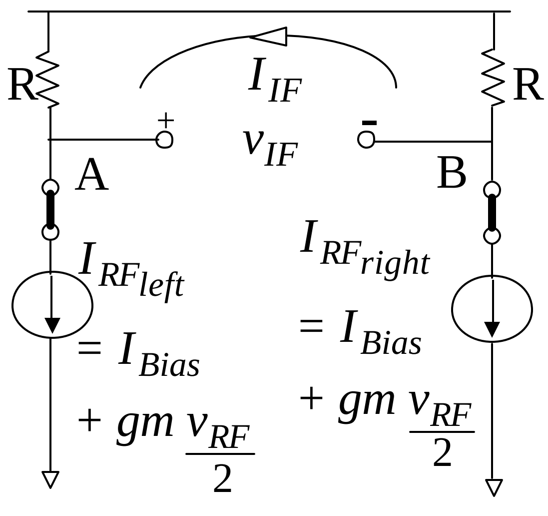
Double-Balanced Mixers



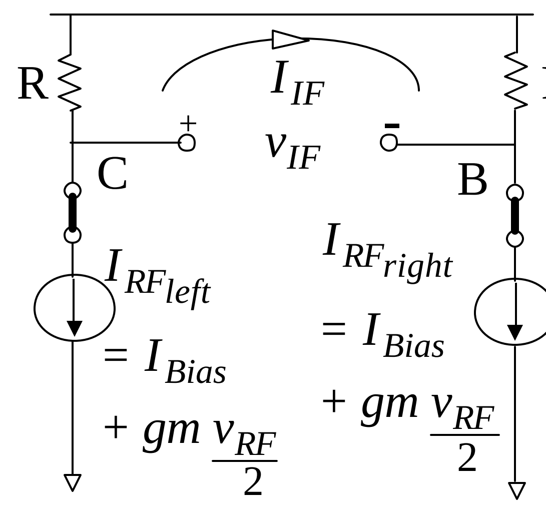
As shown in the figure, the double-balanced mixer uses two voltage dependent current sources. One is controlled by v_{RF} and the other is controlled by $-v_{RF}$. In additions, both current sources carry equal bias current. The current at the IF-port switches between ***gm*** $\boldsymbol{v_{RF}}$ (when the switches are connected to pints *A* and *D)* and ***–gm*** $\boldsymbol{v_{RF}}$ (when the switches are connected to points *B* and *C*). Note that the DC biasing current I_{BIAS} is canceled out since it flows through both load resistors at all times. The differential output voltage v_{IF} is thus not a function of the biasing current. The current at the IF-port can be expressed as:

$$I_{IF} = gmV_{RF}\begin{bmatrix} \frac{2}{\pi}cos(\omega_{LO}-\omega_{RF})t+\frac{2}{\pi}cos(\omega_{LO}+\omega_{RF})t \\ +\frac{2}{3\pi}cos(3\omega_{LO}-\omega_{RF})t+\frac{2}{3\pi}cos(3\omega_{LO}+\omega_{RF})t \\ +\frac{2}{5\pi}cos(5\omega_{LO}-\omega_{RF})t+\frac{2}{5\pi}cos(5\omega_{LO}-\omega_{RF})t+\cdots \end{bmatrix}$$





On Phase #1

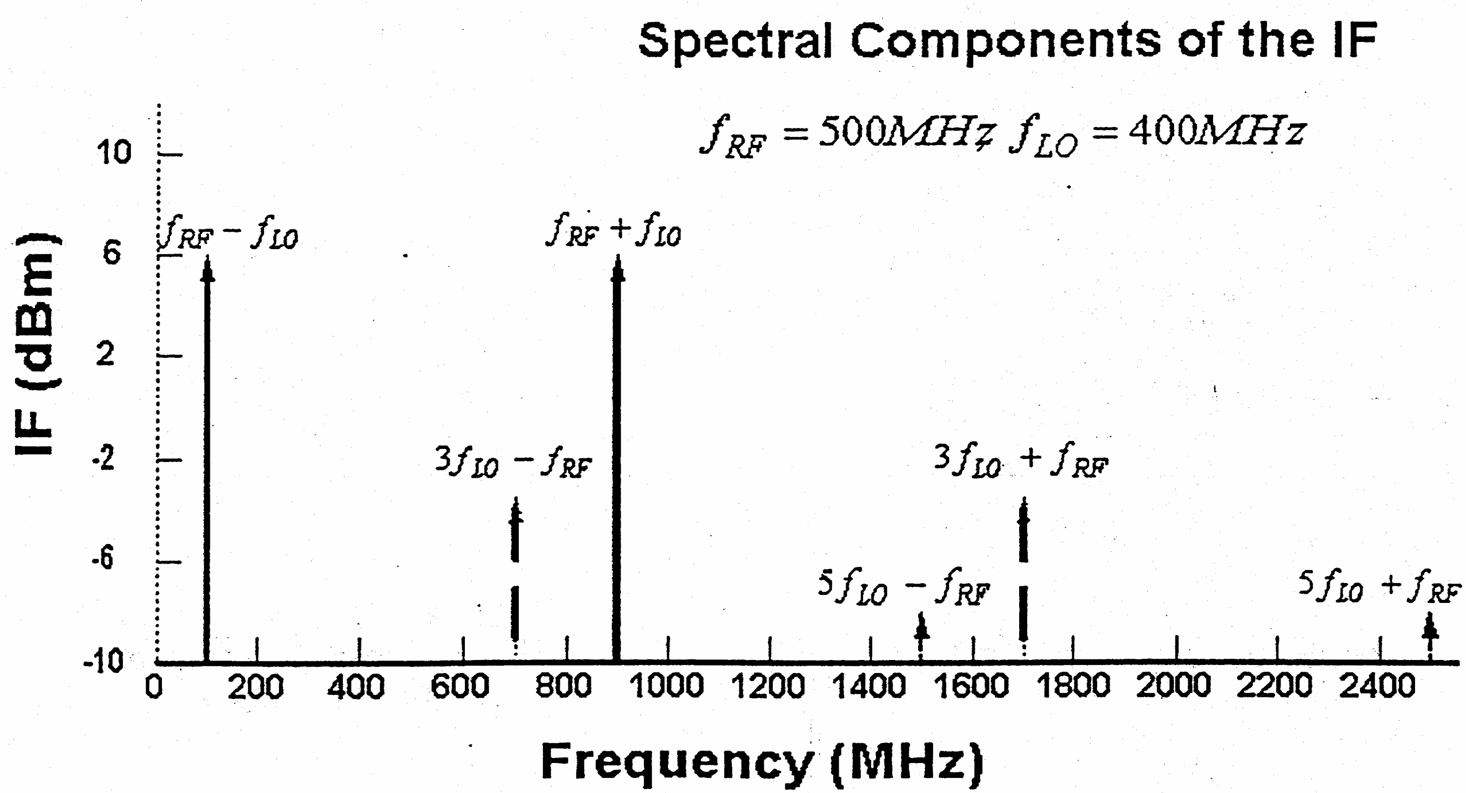


On Phase #2

From the above expression we can make the following observation :

- the double - balanced mixer has no sprectal components at the $n w_{LO}$ where $n = 1,3,5,\ldots$ This implies that the LO - IF feedthrough is significantly better than that of the single - balanced mixer (the LO - IF feedthrough is 40 to 60dB for IC realization).
- The RF - IF feedthrough is theoretically nonexistent.
- While the amplitude of the spectral components of the IF current are equal for both the double - balanced and differential single - balanced mixers, the differential voltage at the IF - port (v_{RF}) of the double - balanced mixer is twice that of the differential single - balanced mixer.
- The conversion transconductance of the double - balanced mixer equal to that of the differential single - balanced mixer.

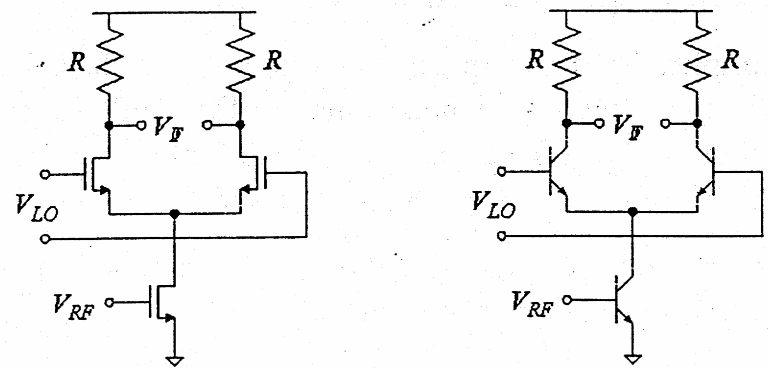
$$G_c = \frac{2gm}{p}$$



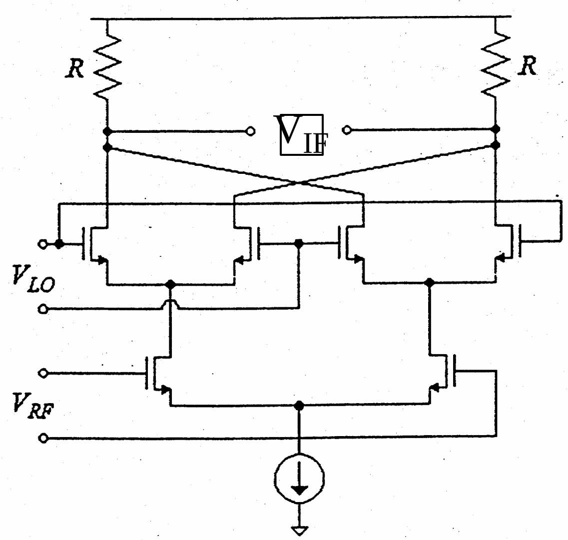
Spectral Components of the IF
$f_{RF} = 500MHz$ $f_{LO} = 400MHz$
$f_{RF} - f_{LO}$
$f_{RF} + f_{LO}$
$3f_{LO} - f_{RF}$
$3f_{LO} + f_{RF}$
$5f_{LO} - f_{RF}$
$5f_{LO} + f_{RF}$
IF (dBm)
10
6
2
-2
-6
-10
0
200
400
600
800
1000
1200
1400
1600
1800
2000
2200
2400
Frequency (MHz)

Implementation of Mixers

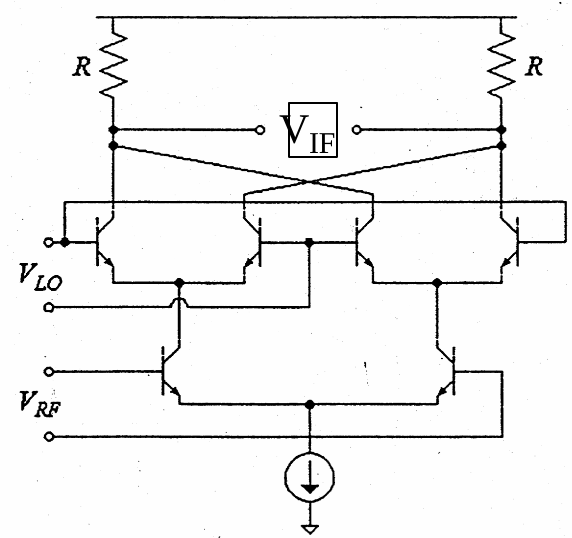
The most straightforward way to implement single- and double- balanced mixers is to use a current –mode (differential pairs).



MOS and BJT Single-Balanced Mixers

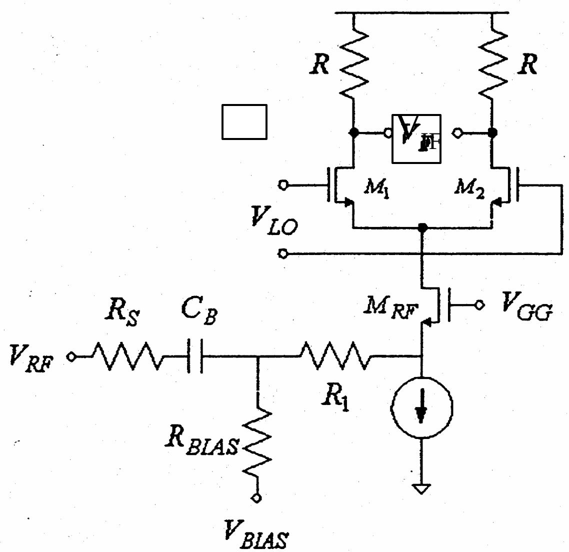


MOS Double-Balanced Mixer

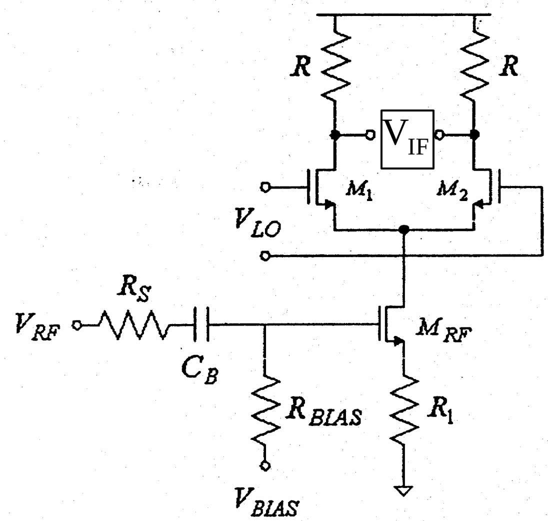


BJT Double-Balanced Mixer (Gilbert Multiplier)

As we have seen before the conversion transconductance and, hence, the conversion gain depends on the device transconductance of the BJT or MOSFET used to convert the RF voltage signal to a current signal. It is, therefore, very important to maintain a linear device transconductance to avoid distortion. Several techniques have been proposed and employed to achieve that. Some of the linearization technique will be introduced next.

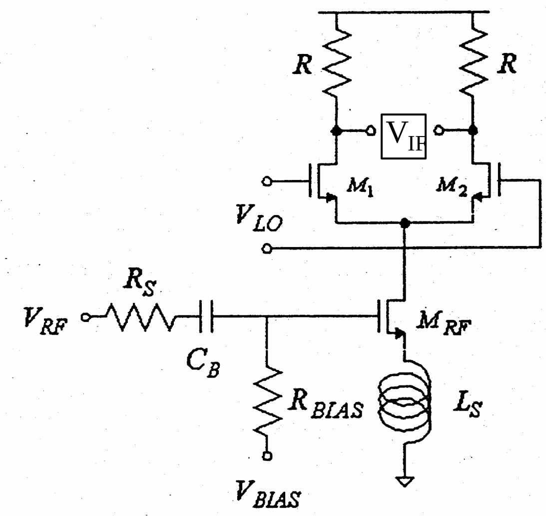


Common-Gate
Linearization



Common-Source
Linearization

Inductive Degeneration Linearization



- Common-Gate Linearization:

The transconductance of the common-gate at the RF port is

$$Gm = \frac{I_{M_{RF}}}{V_{RF}} = \frac{gm}{1+gmR_1} \cong \frac{1}{R_1}$$

The above approximation is possible if $gm\ R_1 >> 1$. This can be achieved since the matching is not determined by gm only. The transistor transconductance gm can be made large enough, while R_1 will be close to $R_s\left(R_s = R_1 + 1/gm\right)$.

The disadvantage is the additional noise due to R_1.

The conversion transconductance is $\mathbf{G_c = 2Gm/}\mathrm{p}$

- Common-Source Linearization:

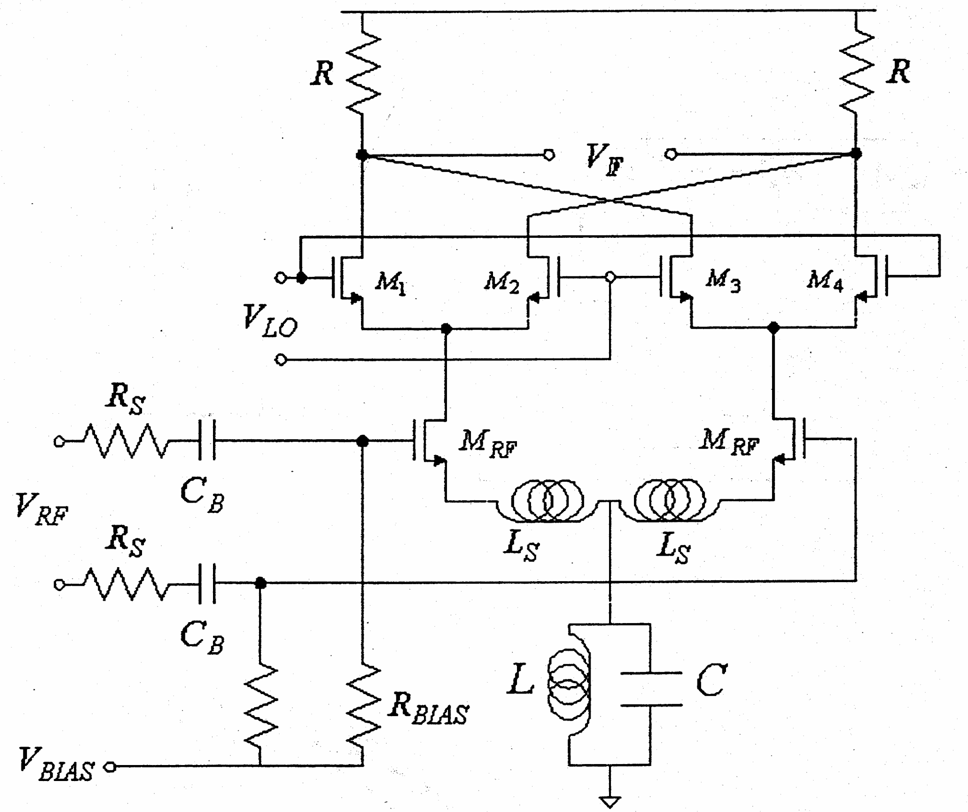
  The common-source will not match the source resistance which would be the output resistance of the image reject filter (50Ω) for superheterodyne receivers.

- Inductive Degeneration:

  This is the preferred topology since it allows for matching, linearization without additional noise. Furthermore, the DC drop across the inductor is zero which makes this topology suitable for low-voltage application. It can be shown that effective transconductance, at the frequency which corresponds to matching for max power transfer, is given by:

$$Gm = \frac{gm}{gm\sqrt{\dfrac{L_1}{C_{gs_{MRF}}}} + R_S\sqrt{\dfrac{C_{gs}}{L_1}}} = \sqrt{\frac{C_{gs_{MRF}}}{L_1}}$$

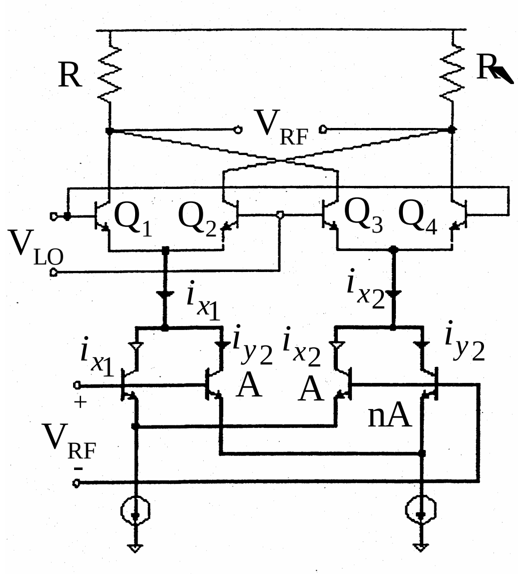
Double-Balanced Mixer

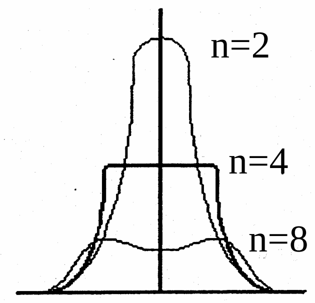


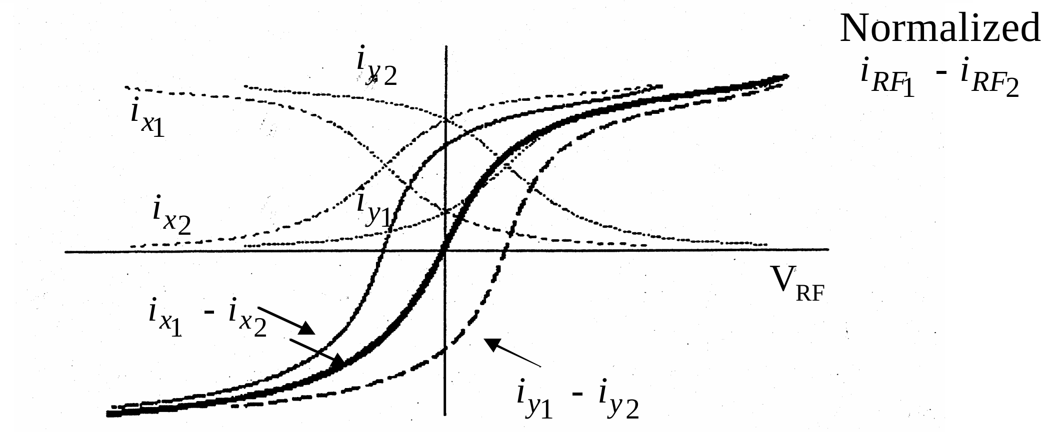
The inductive degeneration can be used in the double-balanced mixer, by adding source inductances. The biasing tail current is replaced by a tank whose resonance frequency is chosen to reject the common-mode components.

The conversion gain is similar to that of the single-balanced mixer.

Schmook's Linearization Multi-tanh Technique







$$i_{RF1} - i_{RF2} = i_{x1} + i_{y1} - i_{x2} - i_{y2}$$
$$= (i_{x1} - i_{x2}) + (i_{y1} - i_{y2})$$

Passive CMOS Mixers

The advantage of passive mixers is its low-power consumption. It also may have superior linearity compared to active mixers.

Multiplier-based mixers using switching can be best implemented in CMOS technology.

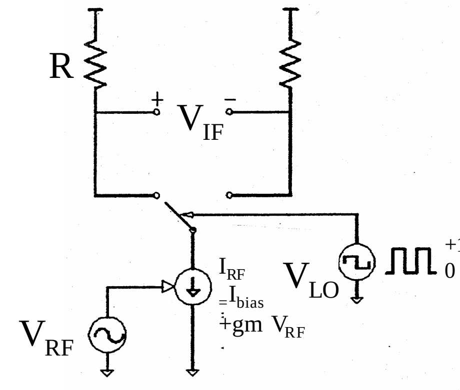
A simple differential single-balanced mixer is shown below.

The following observations are due:

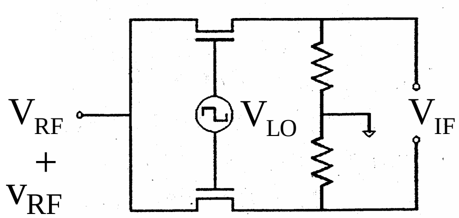
- If the swing of the LO voltage is large so that the gate-to-source overdrive is large enough the channel on-resistance is almost independent on the RF voltage signal. The mixer is, therefore, linear. If, however, the LO swing is small (which is the case for low-voltage applications) the linearity is degraded.

- The *conversion voltage gain* of this passive mixer is $2/\pi$ (-3.9dB) (assuming 50%-duty cycle square-wave LO voltage). As expected passive mixer have conversion gain less than unity.

- If the transistor width is increased to reduce the on-resistance, the capacitive LO-IF feedthrough increases.



Current-Mode Differential Single-Balanced Mixed



Passive CMOS Differential Single-Balanced Mixed

Double-Balanced Passive CMOS Mixers

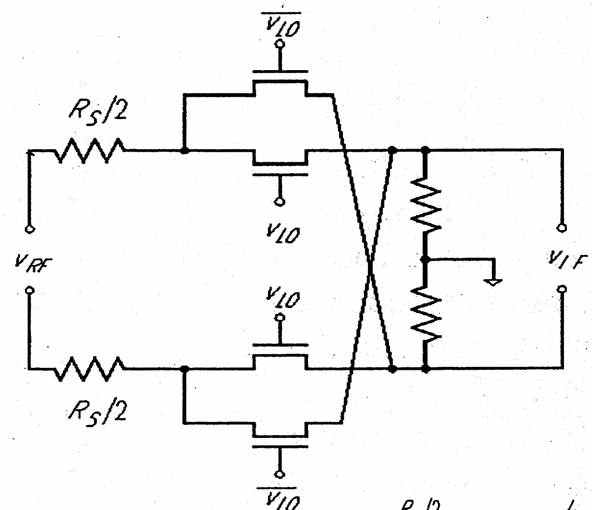
The advantage of passive mixers is its low-power consumption. It also may have superior linearity compared to active mixers.

Multiplier-based mixers using switching can be best implemented in CMOS technology.

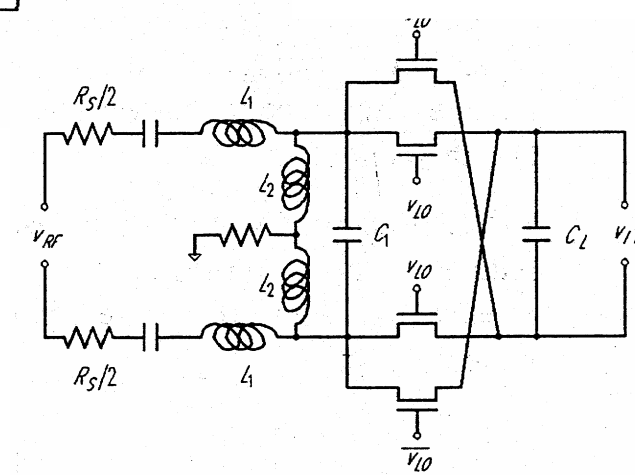
A simple double-balanced mixer is shown below.

$$\frac{v_{RF}}{2}+V_{RF}$$

$$\frac{-v_{RF}}{2}+V_{RF}$$

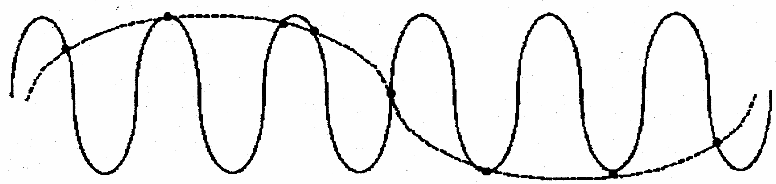


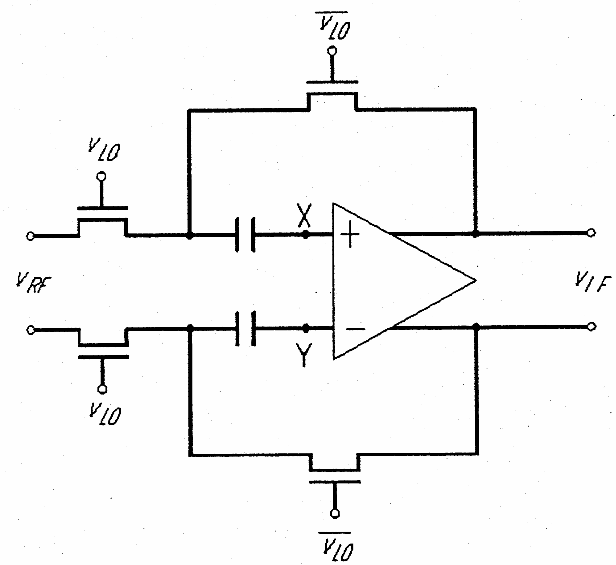
In the case of CMOS IC mixers, the load is typically capacitive. An 1-matching network (comprising of L_1 and $C_1$0. The parallel tank L_2 and C_1+C_L filters out the unwanted tones and out-of-band noise.



Subsampling Mixers

Subsampling the RF or IF signal is a way to downconvert and simultaneously sample for A/D operation. Subsampling RF is still too difficult to achieve in practice. The sampling rate has to be twice the information bandwidth, which is typically much smaller than the IF frequency.



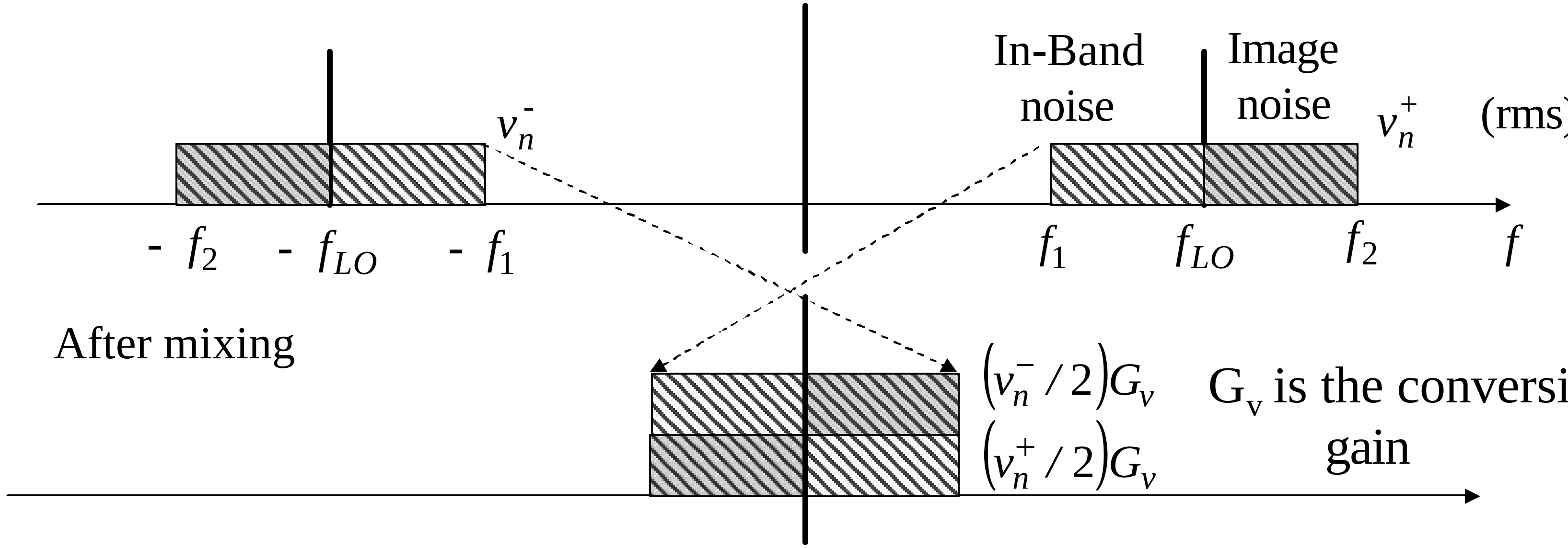


Advantage: is the high linearity

Disadvantages: Jitter noise and thermal noise folding

The overall dynamic range is still inferior compared to other mixers.

Frequency Translation of Noise



G_v is the conversion gain

v_n^+ and v_n^- are not correlated but are equal in magnitude.

$$N_i = v_n^{2+} = v_n^{2-} = v_n^2$$

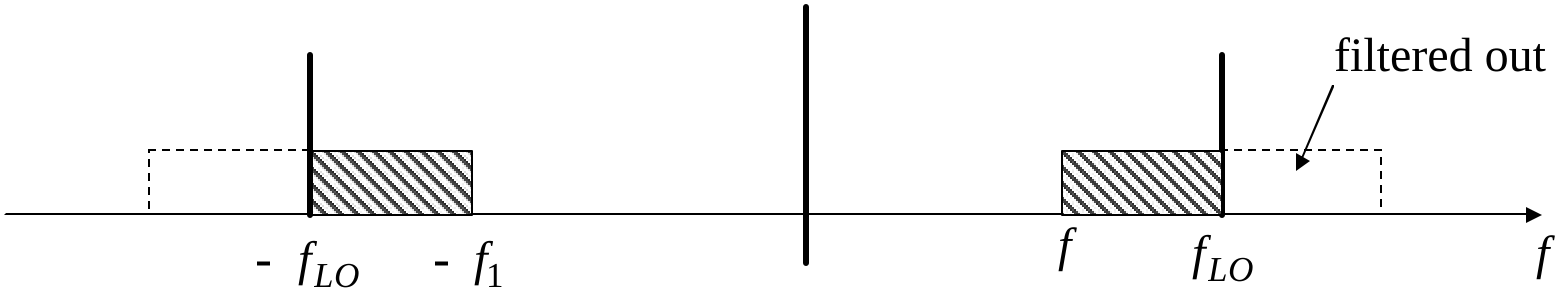
$$N_o = \overline{\left(\frac{v_n^+}{2} + \frac{v_n^-}{2}\right)^2} \; G_v^2$$

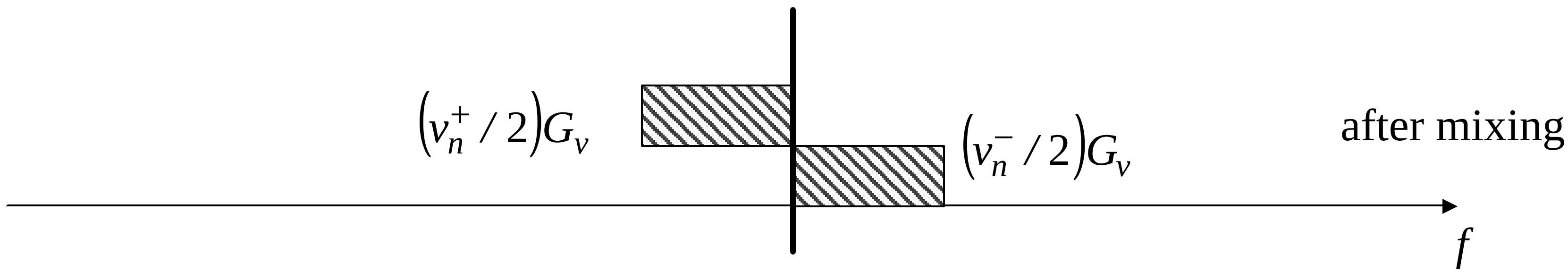
$$= \left(\frac{v_n^2}{2}\right)^2 + \left(\frac{v_n^-}{2}\right)^2 = \frac{v_n^2}{2} G_v^2$$

$$\therefore N_o = \frac{N_i G}{2}$$

(G is the power gain)

Assume Image Noise is Filtered



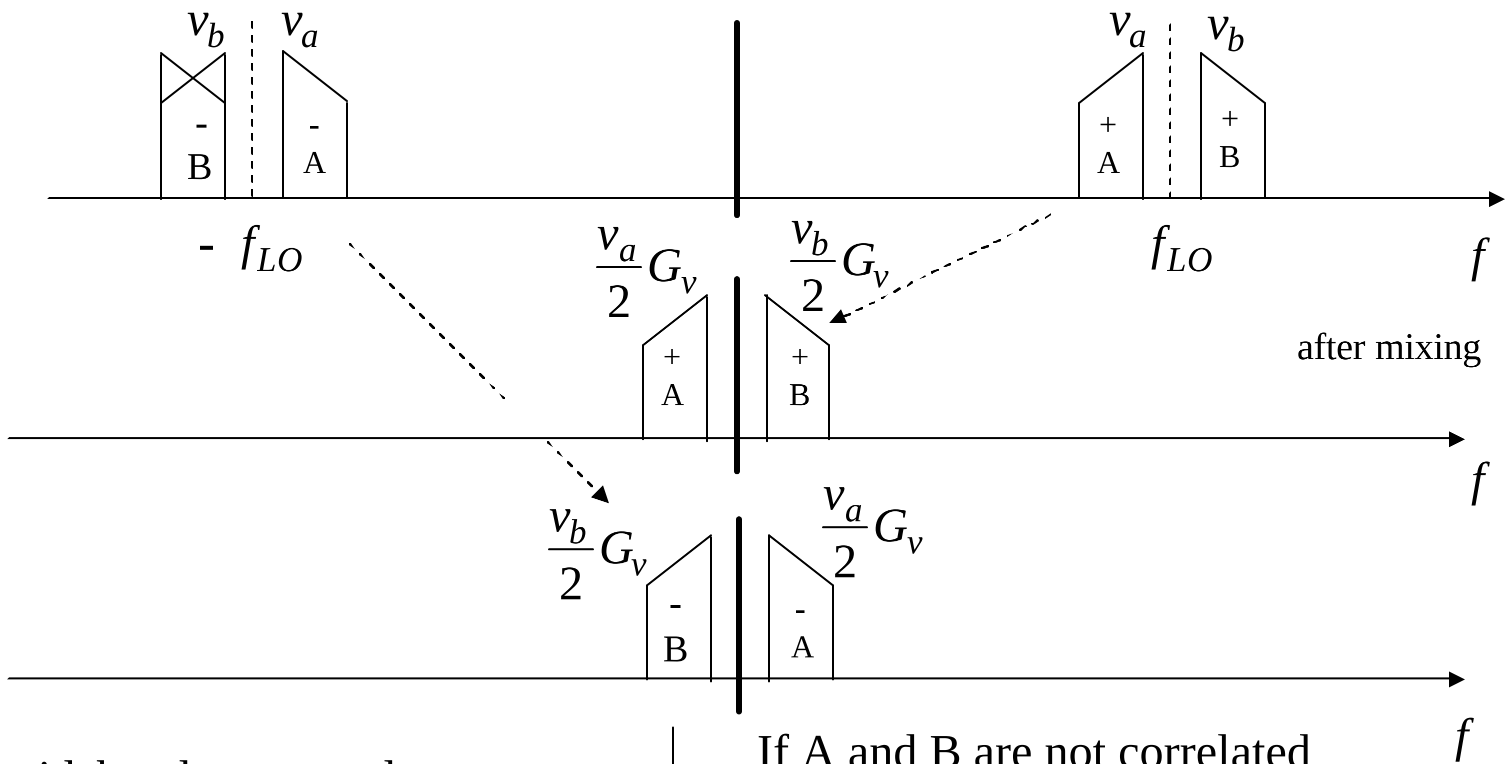


$$N_i = v_n^2$$

$$N_o = \left(\frac{v_n^-}{2}G_V\right)^2 = \left(\frac{V_n^2}{2}\right)G$$

$$N_o = \frac{N_i}{4}G$$

Frequency Translation of the Signal



DSB:

If both sidebands are equal

$$S_i = v_a^2 = v_b^2 = v^2$$

$$S_o\left(\frac{v}{2}\cdot G_v \times 2\right)^2 = S_i G$$

SSB: (assuming no B-band)

$$S_o = \left(\frac{v}{2}G_v\right)^2 = \frac{S_i}{4}G$$

If A and B are not correlated

$$S_o = \left(\frac{v_a}{2}G_v\right)^2 + \left(\frac{v_b}{2}G_v\right)^2$$

$$= \left(\frac{v_a^2}{4} + \frac{v_b^2}{4}\right)G$$

$$\cong \frac{S_i}{2}G$$

Noise Factor Definitions for Mixers

Case I: SSB NF

One sideband of the signal is down converted and assume that the noise in the image frequency range is not filtered.

Þ $$S_o = \frac{S_i}{4} G$$

$$N_o = \frac{N_i G}{2}$$

$$F_{ssB} = \frac{S_i}{N_i} \times \frac{\text{No Total}}{S_o} = \frac{S_i}{N_i} \times \frac{N_o + N_{no}}{S_o}$$

$$= \frac{S_i}{N_i} \times \frac{\frac{N_{iG}}{2} + N_{no}}{\frac{S_i G}{4}} = 2 + \frac{4 N_{no}}{G N_i}$$

Case II: DSB NF with S_{iA} and S_{iB} not correlated.

$$S_o = \frac{S_i}{2} G$$

$$N_o = \frac{N_i}{2} G$$

$$F_{DSB} = \frac{S_i}{N_i} \times \frac{N_o + N_{no}}{S_o}$$

$$= \frac{S_i}{N_i} \times \frac{\frac{N_i}{2} G + N_{no}}{\frac{S_i}{2} G} = 1 + \frac{2N_{no}}{GN_i}$$

Notice that

$$F_{DSB} = \frac{1}{2} F_{SSB}$$

DSB noise figure is 3dB less than the SSB noise figure.

Zero-IF is a typical example of a DSB case.

In practice, the image noise (in a SSB case) is filtered out using an IMR filter. The filter will not suppress the noise completely. Hence, we may assume that image noise will see a conversion gain of G_{imap} (= IMR less x G at image frequency). In such case

$$F_{SSB} = \frac{S_i}{N_i} \times \frac{N_o + N_{no}}{S_o} \qquad \left(S_o = \frac{S_i\, G}{4} \right)$$

$$N_o = \left(\frac{v_n^-}{2} G_v \right)^2 + \left(\frac{v_n^+}{2} G_{image} \right)^2 \qquad \overline{v_n^-}^2 = \overline{v_n^+}^2 = \overline{v_n}^2$$

$$= \frac{v_n^2}{4}\left(G + G_{image}\right) = \frac{N_i}{4}\left(G + G_{image}\right)$$

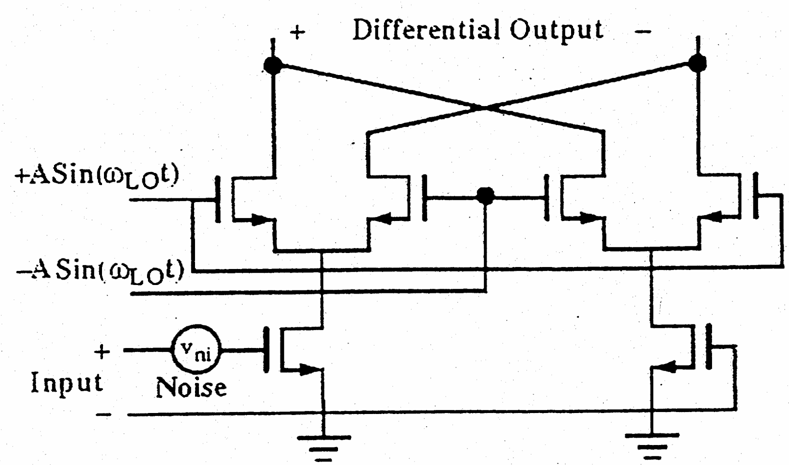
$$F_{SSB} = 1 + \frac{G_{image}}{G} + \frac{4N_{no}}{GN_i}$$

Since $\quad G_{image}/G < 1$

$$F_{SSB} \quad < 2F_{DSB}$$

Noise in RF-CMOS Mixers

Ref: H. Darabi et. al. JSSC, Jan. 2000



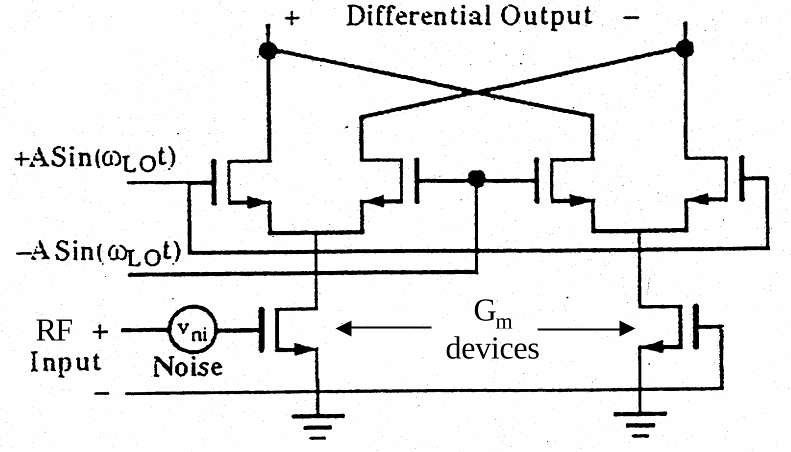
The noise is contributed by:

1. Loads
2. Transconductance FETs
3. Switches

Load Noise:

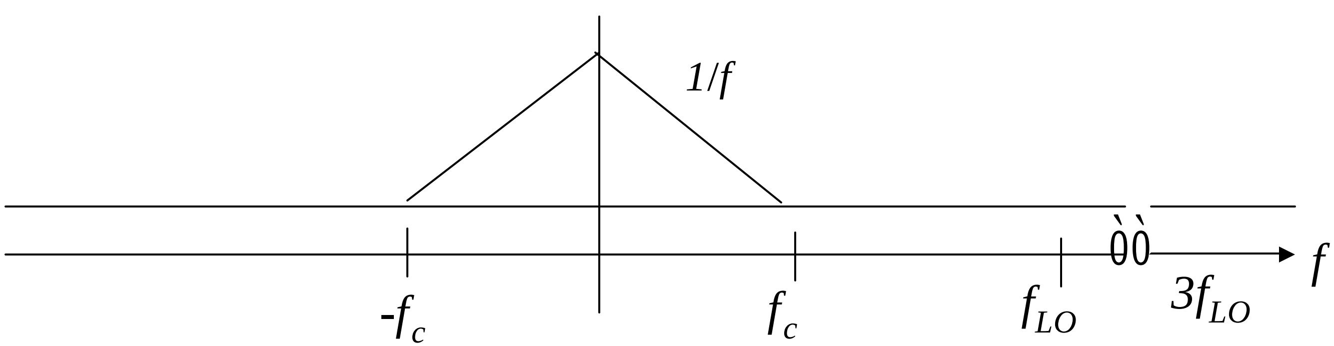
The noise of the load, especially the 1/f noise, will be critical for zero-IF down-conversion mixers.

Transconductance Noise:

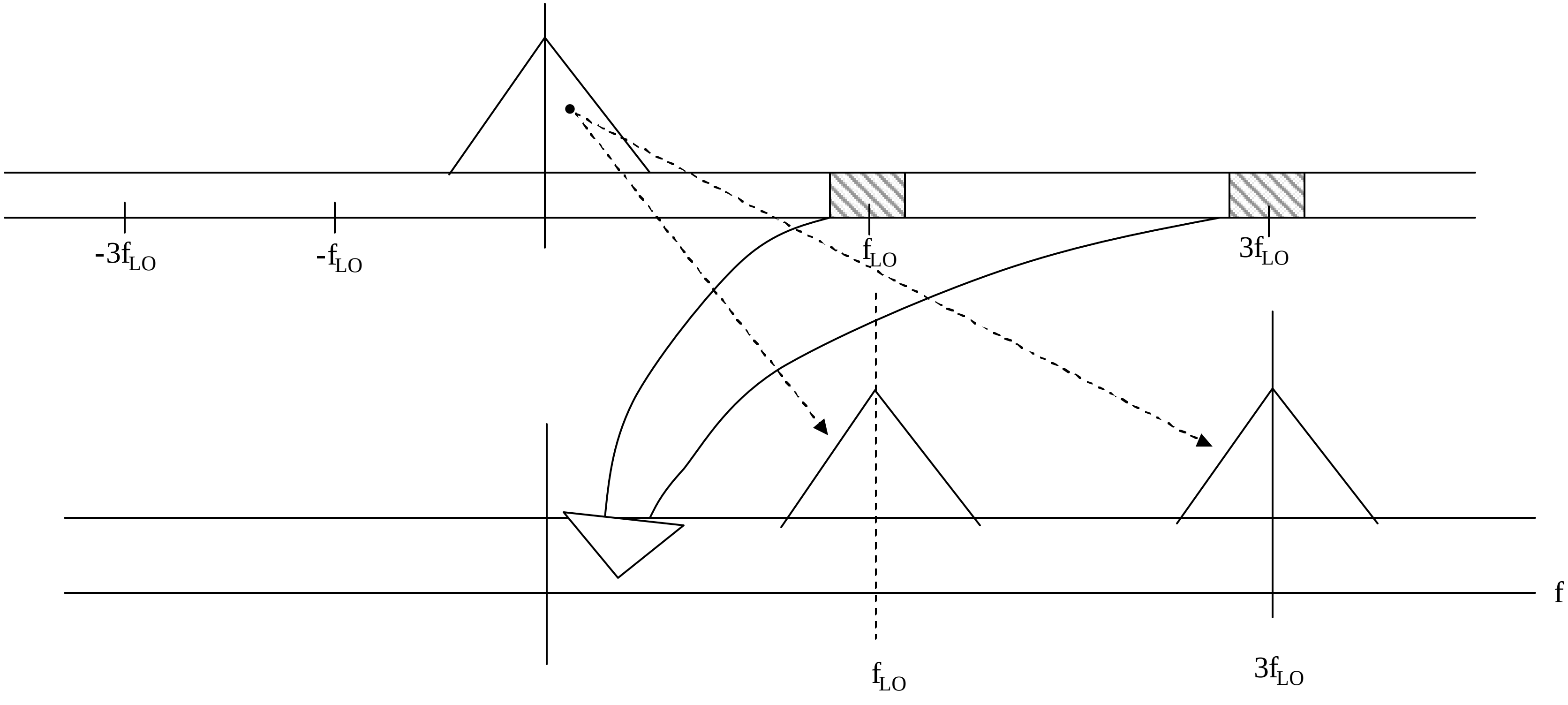


The input referred noise will be translated in frequency in a similar fashion as the RF-input signal.

The noise of the G_m MOSFETs consists of:
1/f noise and white noise



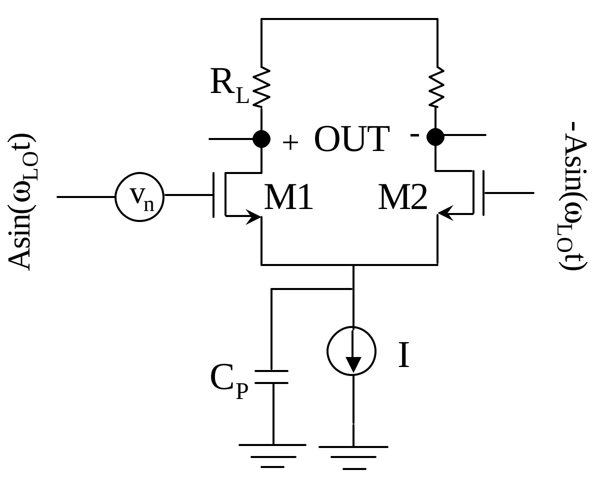
In the case of zero-IF, the white moise at f_{LO}, $3f_{LO}$, $5f_{LO}$, . . . (odd harmonics of LO) will be down converted to DC. The 1/f noise will be up converted to f_{LO} and its odd harmonics.



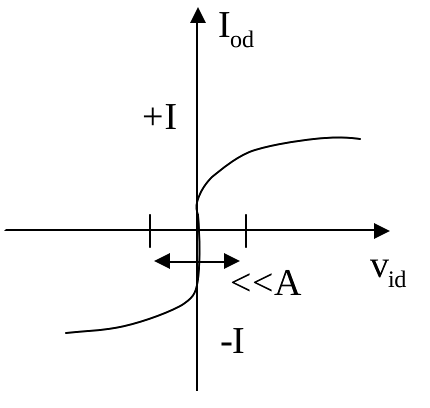
Direct Switch Noise

The following assumptions will be made:

- The 1/f noise of both switches M1 and M2 will be referred to the input of one of the devices. It is a slow varying signal compared to the LO signal.
- The g_m of the differential pair is high.



Single-balanced mixer with switch Noise modeled at gate.



Assumed switch *I-V* characteristic.

Assume that all input phase noise is reflected to one side of the differential pair and is given by:

$$v_{id} = v_n(t) + 2A\sin \omega_{LO} t$$

Each time the V_{id} crosses zero the i_{od} switches from –I to I or vice-versa.

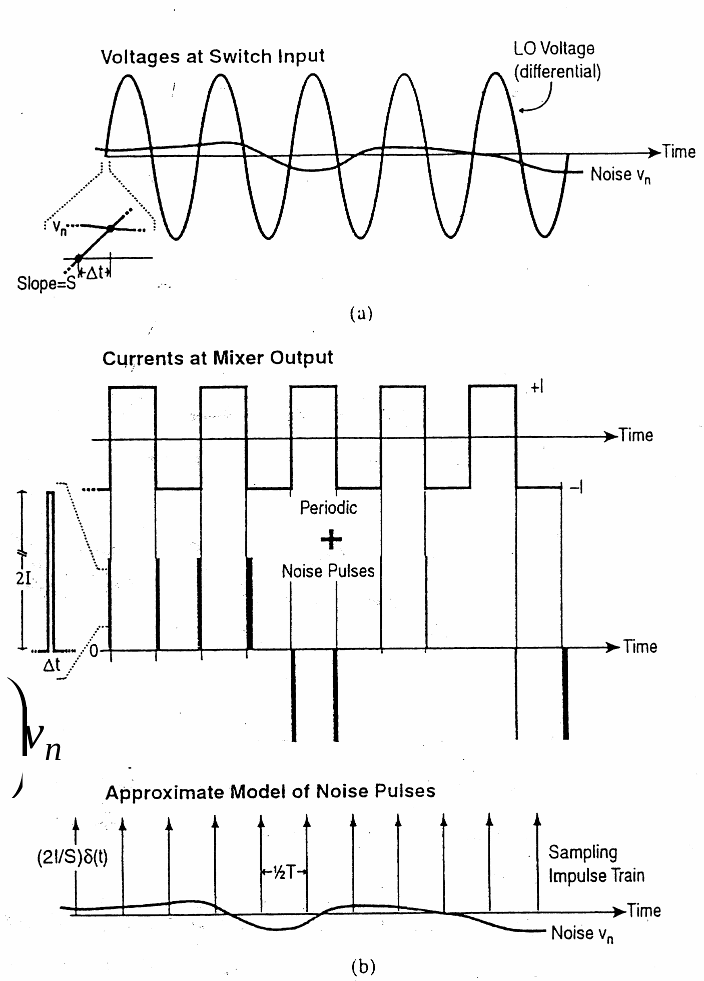
If $v_n(t)$ was ten the zero crossings of i_{od} would occur at exactly T/2 intervals (when $T=1/f_{LO}$).

If v_n (t) causes the zero crossings to shift right or left. The current waveform can be divided into two waveforms.

- One ideal 50% duty-cycle square wave (due to the ideal $\sin w_{LO\,t.}$
- Random current pulses due to $v_n(t)$. The width of each puse is $\Delta t = \dfrac{v_n(t)}{s}$

Slope of LO
@ zero 2 ωA

And its height = 2 I
Its frequency = 2 f_{LO}



$$2I_o t = \left(\frac{2I}{s}\right) v_n$$

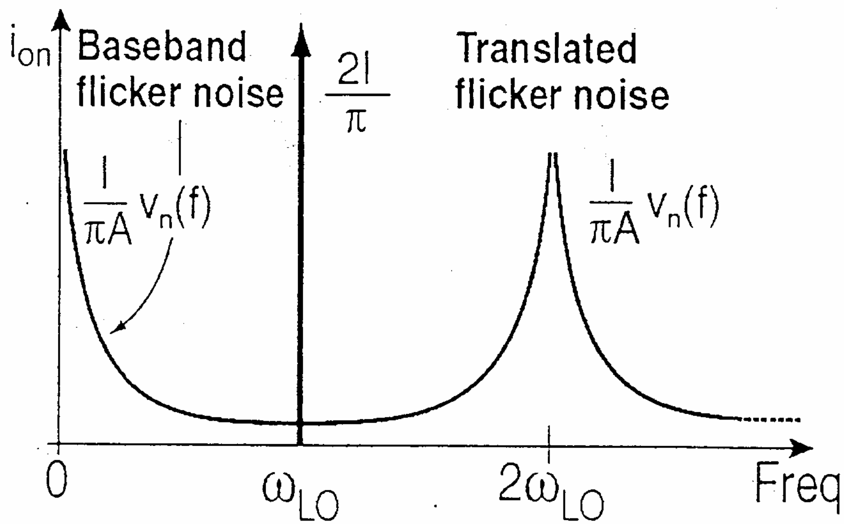
The current noise pulses have an averaging (over 1 T/2) of:

$$i_{o,n_{av}} = \frac{2I_o t}{T/2} = 2I\,\frac{v_n(t)}{s} \times \frac{2}{T} = \frac{I\,v_n}{\pi A}$$

2πf A x 2

The current noise pulses can be approximated by delta-function impulses (since $\Delta t \ll T$).

This is similar to a sampling process. Hence, the spectrum, due to the 1/f noise of the switches, at the output looks like:



Mixer output spectrum in presence of direct noise.

There are images (alias) of 1/f noise at DC, 2 f_{LO}, $4f_{LO}$, … (harmonics of $2f_{LO}$).

SNR due to the switch noise can be derived as follows:

$$SNR_1 = \frac{\text{signal}}{\text{noise}} = \frac{i_o}{i_{no}}$$

$$signal = \text{conversion } \; G_m \times V_{in} = \frac{2g_m}{\pi} V_{in}$$

$$SNR_1 = \frac{2g_m v_{in}}{\pi} \cdot \frac{1}{\frac{1}{\pi}\frac{IV_n}{A}} = \frac{2A}{V_{GS} - V_T} \cdot \frac{v_{in}}{v_n}$$

$$g_m = \frac{I}{V_{GS} - V_T} \quad \text{for short channel MOS}$$

⇨ This implies that SNRs can be improved by

— increasing A through LO swing

Trade off with frequency
— reducing V_{gs}-V_T (over drive of the switches)
— reducing N_n by increased WL of the switch.

For the double-balanced mixer, the previous analysis is pretty much the same.

The main differences are:

- No LO feedthrough
- v_n would be the noise of all four switches.

White Noise in Mixer Switches:

The approach used for 1/f noise (on pp 5-8) can be extended to the white noise.

By referring the white noise to the input of the switches as in the figure of pp 5.

The white noise results in train of pulses at $2f_{LO}$ (as shown below).

To simplify the analysis, the random current pulses is approximated by a train of rectangular pulses with a constant width (Ts) and a height of $2I/ST_s$. Then pulses sample the noise v_n. This imples that the white noise will be down converted to IF (or DC for zero – IF). The noise which will be down-converted, is at $2f_{LO}$ and its harmonics. This is a kind of subsampling.

Transconductor White Noise

The white noise of Gm device, when referred to the input, can be treated the same way as the (RF) input signal. Assuming a 50% duty cycle (square wave) LO, the white noise at f_{LO} + IF, $3f_{LO}$ + IF, $5f_{LO}$ + IF, … will all be down converted to IF .