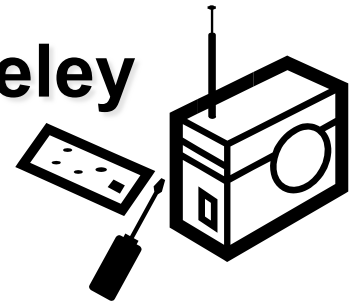# EECS 242:
Mixer Noise and Design
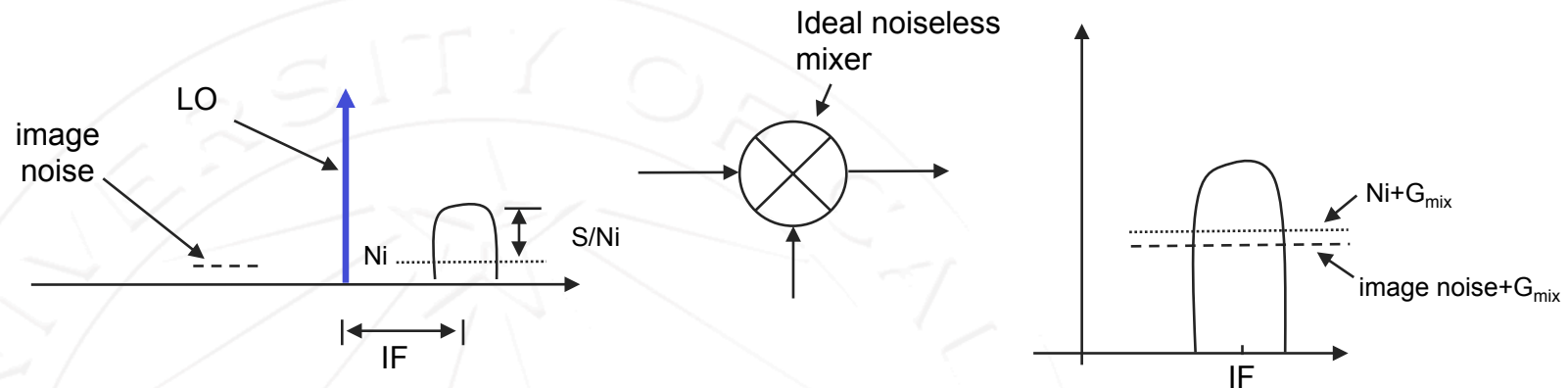
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SSB vs. DSB NF definition:

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.

Single-side band (SSB) NF

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 architectures, where the image signal is suppressed by an image filter before reaching the mixer. 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,

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

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

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

IEEE Noise Definition

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. Therefore, the IEEE SSB NF assumes there is a sharp bandpass filter that passes the desired band with the source noise and knocks down the image noise to the negligible level. As a result, one can write
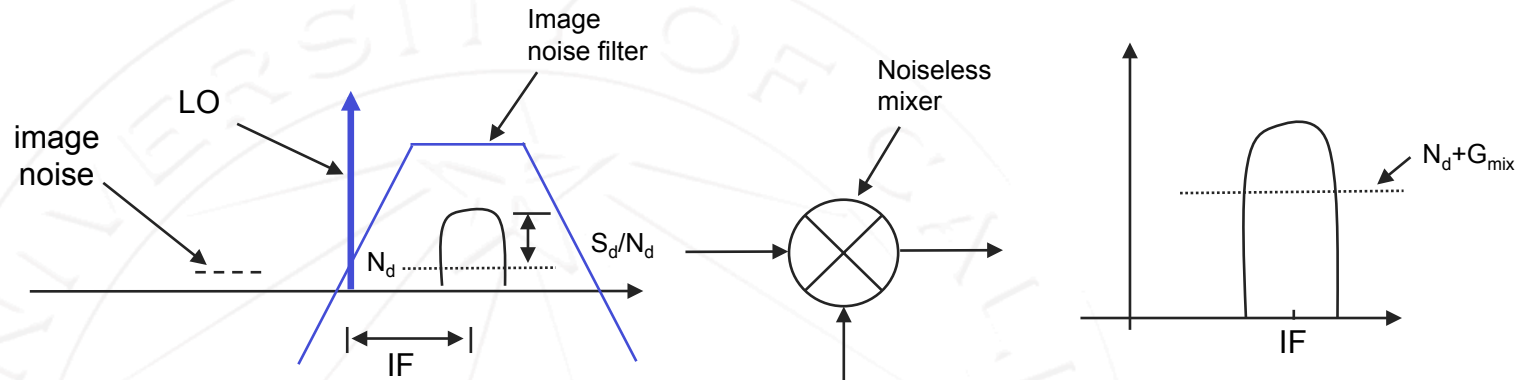
$$S_{out} = S_d G_{mix\_d}$$

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

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

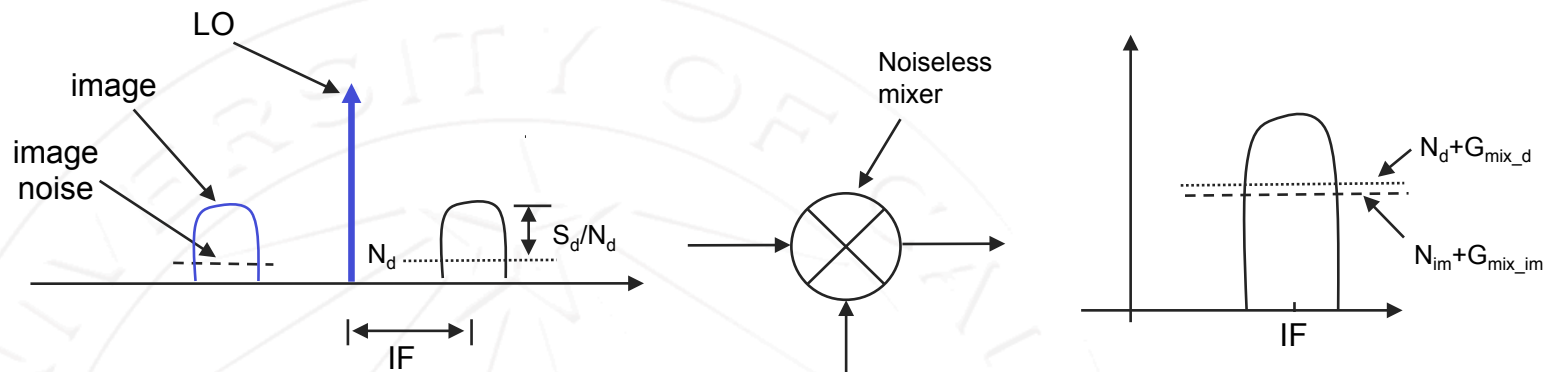$$N_{out} = N_d G_{mix\_d} + 2N_{mix} G_{mix\_d}$$

IEEE Noise Def (cont)

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

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

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.

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,

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

DSB NF (cont)

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, the difference 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:

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

Periodically Time-Varying Systems
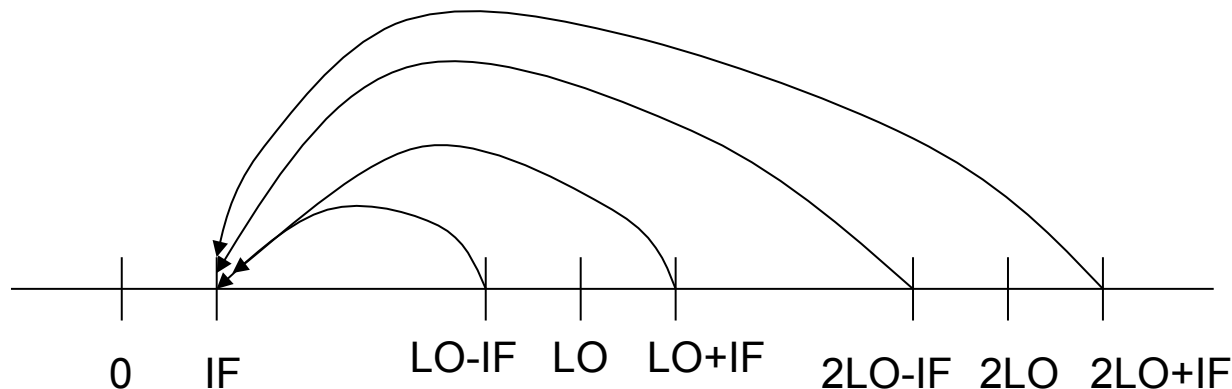
Suppose the AC signal at output is given by

$$I_{o1} = P_1(t) \cdot i_s$$

Let X(f) be the spectrum of i_s

Y(f) be the spectrum of I_{o1}

$$Y(f) = \sum_{n=-\infty}^{\infty} P_{1,n} X(f - nf_{LO})$$

If we fix the output frequency at f_{IF}, then we see energy from multiple bands folding into output spectrum

Mixer Conversion Gain

$$Y(f_{IF}) = \sum_{n=-\infty}^{\infty} P_{1,n} X(f_{IF} - n f_{LO})$$

n=+1	$f_{LO}-f_{IF}$
n=-1	$f_{LO}+f_{IF}$
n=+2	$2f_{LO}-f_{IF}$

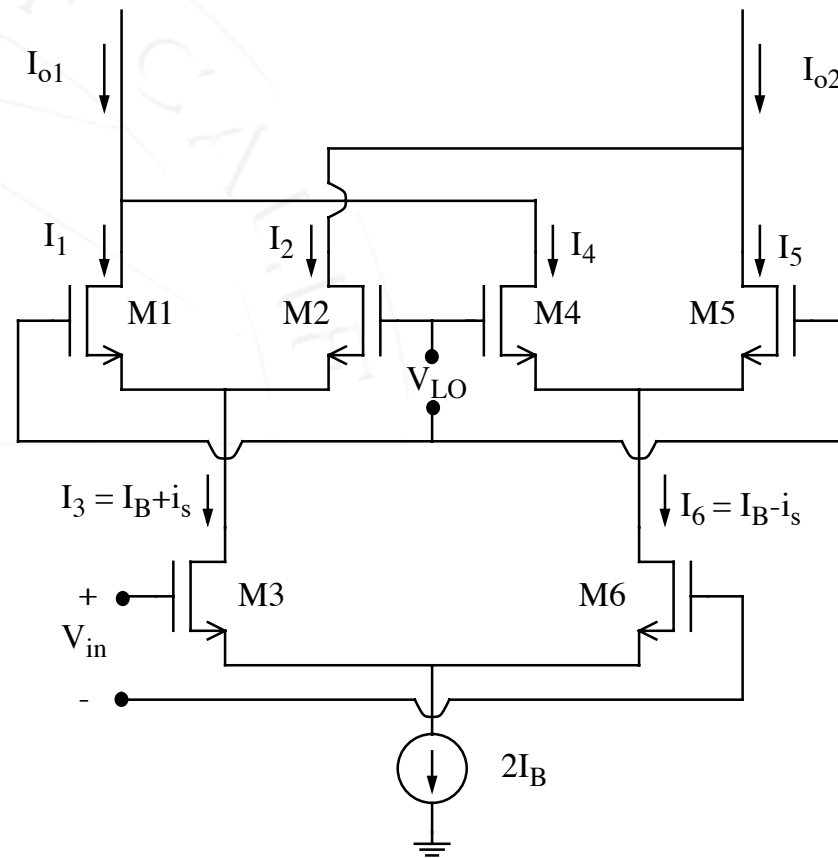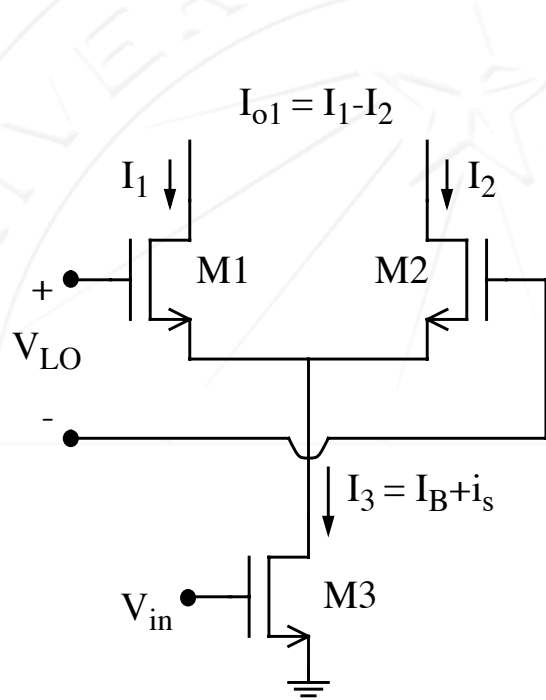$$= \sum_{n=-\infty}^{\infty} P_{1,n} X(n f_{LO} - f_{IF})$$

The coefficient $P_{1,\pm k}$ then represents the conversion gain from frequency $k \cdot f_{LO} \pm f_{IF}$ to f_{IF}.

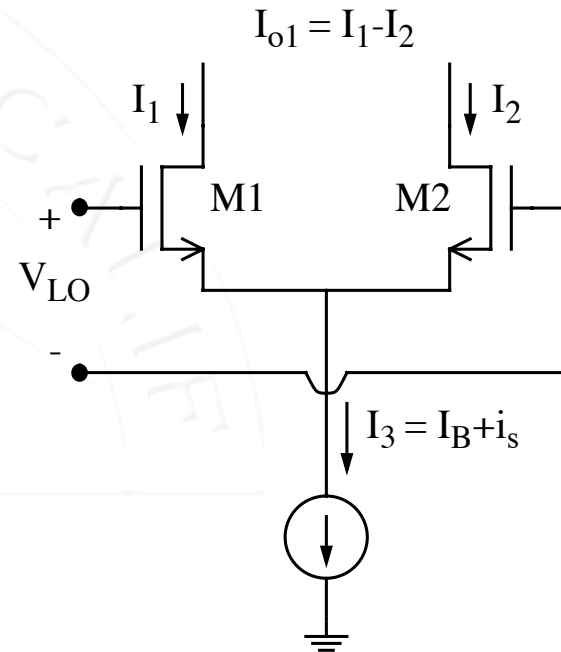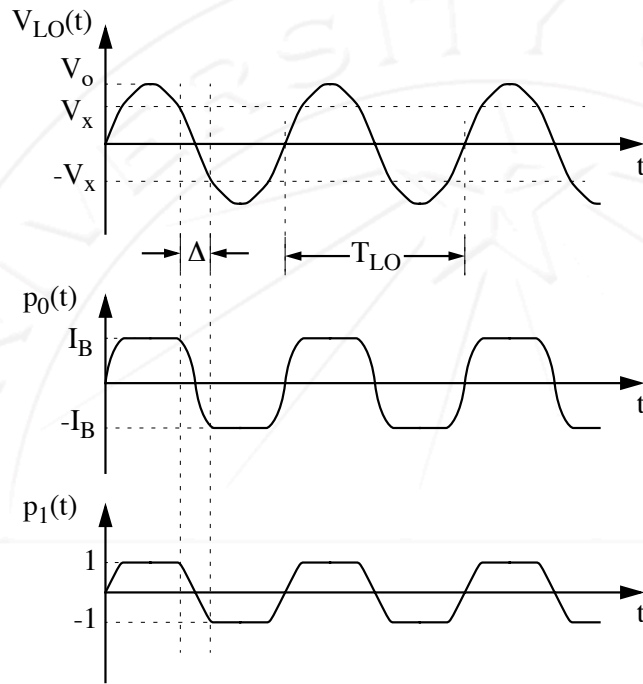If we assume hard switching, then $P_1(t)$ is a square wave and the coefficients are

$$P_{1,2k-1} = \frac{2}{\pi(2k-1)}$$

Single and Double Bal Mixer

Core Noise Model

$$I_{o1} = p_0(t) + p_1(t) \cdot i_s.$$

$$p_1(t) = \frac{g_{m1}(t) - g_{m2}(t)}{g_{m1}(t) + g_{m2}(t)}$$

Noise from Gm Stage

$$y_{n_3} = n_3(t) \cdot p_1(t)$$

$n_3(t)$ WSS

PSD $\quad s_{n3}(f)$

A cyclostationary process is a random process whose statistics are periodic functions of time

The PSD of a cyclostationary process is given by S(f,t)
If we measure a cyclostationary process over a bandwidth < 1/period, we observe $\overline{S(f,t)}$, a stationary process

Example:

White noise

$$y(t) = n(t) \cdot p(t) \longleftarrow \text{deter periodic func.}$$

$$S_{n3}^o(f) = \sum_{n=-\infty}^{\infty} |p_{1,n}|^2 \, S_{n3}(f - n f_{LO}).$$
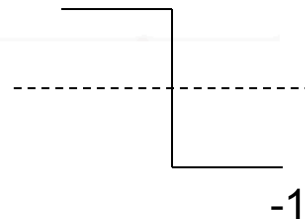
Gm Stage Noise (cont)

If $S_{n3}(t) = N_{n3}$ (white), then

$$S_{n3}^o(f) = N_{n3} \cdot \sum_{n=-\infty}^{\infty} |p_{1,n}|^2 = \alpha N_{n3}$$

$$\alpha = \sum_{n=-\infty}^{\infty} |p_{1,n}|^2 = \frac{1}{T_{LO}} \int_{-\infty}^{\infty} (p_1(t))^2 \, dt$$
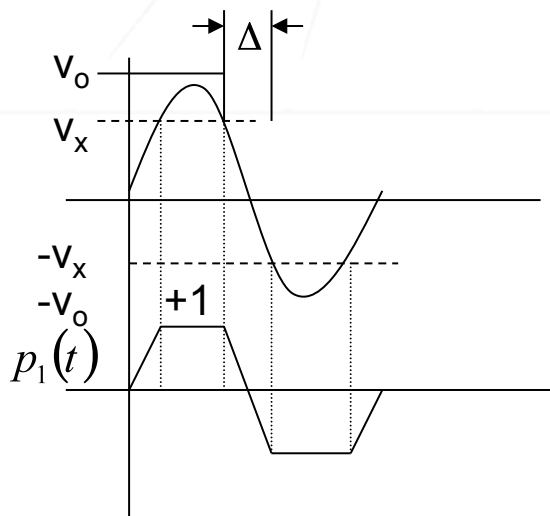
For a square wave $p_1(t)$,

$\alpha = 1$

For first sidebands ($f_{LO} \pm f_{IF}$), this noise accounts for

$2 \times \left(\dfrac{2}{\pi} \right)^2 \approx 81\%$ of noise.

Gm Stage Noise (cont)

$(3f_{LO} \pm f_{IF})$ accounts for $2 \times \left(\dfrac{2}{3\pi}\right)^2 \approx 7\%$

The remaining harmonics account for the rest ~ 10%

Assume $p_1(t)$ is a straight line during period Δ when both devices on

$$\alpha = 1 - \frac{4}{3}\Delta f_{LO}$$

$$\pi \Delta f_{LO} = \sin^{-1}\left(\frac{v_x}{v_o}\right)$$

Single Balanced Mixer Noise

$$S_{n3}(f) = 4kT\left(R_s + r_{g3} + \frac{\gamma}{g_{m3}}\right) \cdot \underbrace{g_{m3}^2}_{G_{m3}^2} \cdot \alpha$$

Noise due to M3 (transconductor)

Input referred noise voltage

G_{m3}^2 for M3 with degeneration
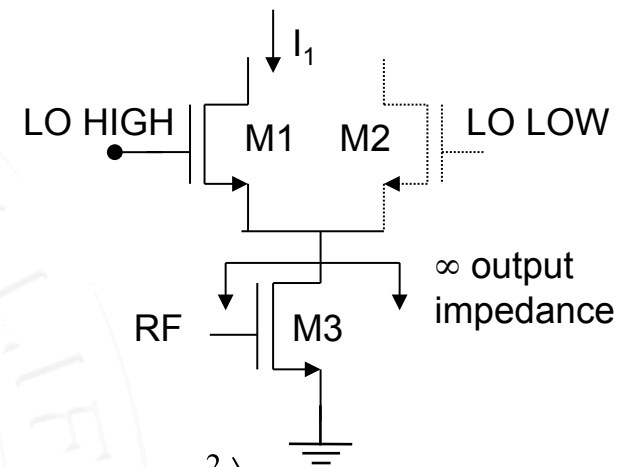
Time variance of problem

Gilbert Cell: $\quad S_{n3}(f) = 4kT\left(R_s + 2r_{gs} + \frac{2\gamma}{g_{m3}}\right)G_{m3}^2 \cdot \alpha$

Thermal noise of switching pair

When we switch hard, the noise is due to M3 only (transconductor) (neglecting capcitance and output impedance of M3)

Noise of Switching Pair

Hard switching is good for low noise output. When both M1 & M2 on, during the switching period Δ, then the noise PSD at output is

$$S_{I_1} = 4kT\gamma \left(\frac{1}{g_{m1}} \left(\frac{g_{m1}}{1 + \frac{g_{m1}}{g_{m2}}} \right)^2 + \frac{1}{g_{m2}} \left(\frac{g_{m2}}{1 + \frac{g_{m2}}{g_{m1}}} \right)^2 \right)$$

$$= 4kT\gamma \, \frac{g_{m1} \cdot g_{m2}}{g_{m1} + g_{m2}}$$

Noise of Switching Pair (cont)

$$I_3 = I_1 + I_2 \qquad I_{o1,n} = 2I_{1,n} \implies \quad \text{4X power}$$

$$S_{n1,2}(f,t) = 16kT\gamma \frac{g_{m1}g_{m2}}{g_{m1} + g_{m2}} = 8kT\gamma G(t)$$

$$G(t) = \frac{2g_{m1}g_{m2}}{g_{m1} + g_{m2}}$$

Total transconductance of diff pair for LO port to differential output current

$$S_{n1,2}(f) = \overline{S_{n1,2}(f,t)} = 8kT\gamma \left(\frac{1}{T_{LO}} \int_0^{T_{LO}} G(t)\,dt \right)$$

$$= 8kT\gamma \overline{G}$$

Noise of Switching Pair (cont)

$$v_{LO} = v_o \sin\left(2\pi f_{LO} t\right)$$

$$\overline{G} = \frac{1}{\pi v_o} \int_{-v_x}^{v_x} G(v_{LO}) \underbrace{\frac{1}{\sqrt{1 - \left(\dfrac{v_{LO}}{v_o}\right)^2}}}_{\approx 1} dv_{LO}$$

$$\approx \frac{1}{\pi v_o} \int_{-v_x}^{v_x} \frac{dI_{o1}}{dv_{LO}} \cdot dv_{LO} = \frac{2I_B}{\pi v_o}$$

$$S_{n1,2}(f) \approx \frac{16kT\gamma}{\pi}\left(\frac{I_B}{v_o}\right)$$
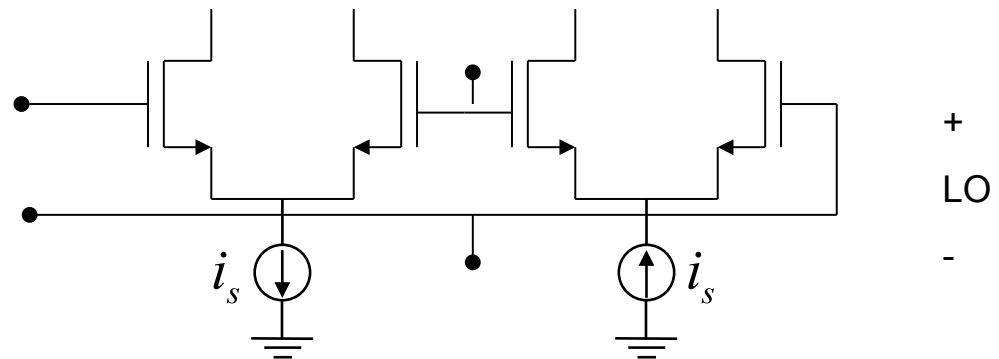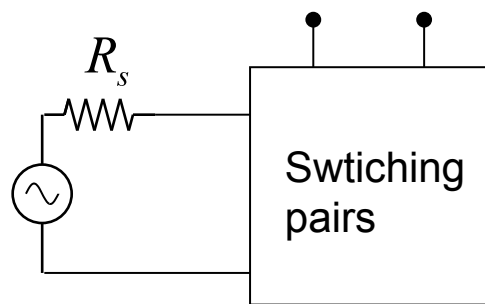
← LO amplitude

Noise from LO Port

$$Y_{nLO} = G(t) \cdot n_{LO}(t)$$

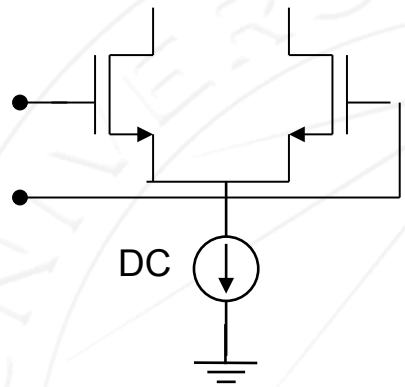$$S_{nLO}(t) = N_{LO} \sum_{-\infty}^{\infty} |G_n|^2 = \overline{G^2} N_{LO}$$

$$S_{n,LO}(f) = 4kT \big(4r_{g1}\big) \overline{G^2}$$

2 switching pairs

R_s

Swtiching pairs

i_s i_s

+

LO

-

Flicker Noise

DC $\rightarrow$ LO in upconversion from f_{LO} and all odd harmonics

Flicker noise of switching transferred to output by multiplication by G(t)
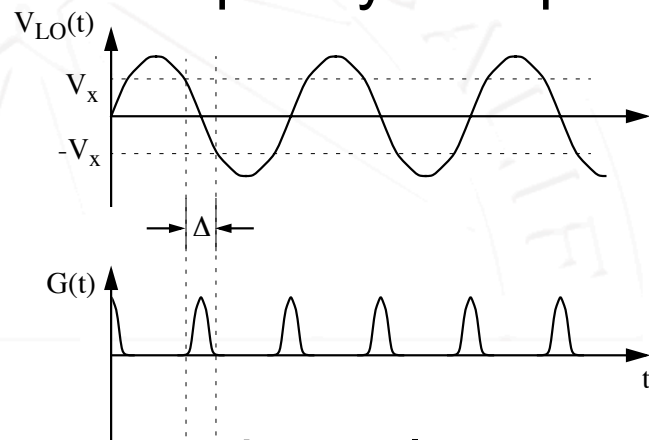
G(t) has period $\dfrac{T_{LO}}{2}$

G(t) Fourier series contains only even-order harmonics of the LO

$\Rightarrow$ Flicker noise from the switching pair will appear at output at DC but <u>NOT</u> at f_{LO}.

(See Ref [0], Manolis Terrovitis, JSSC)

Mixer Design Issues

Single Balanced Mixer
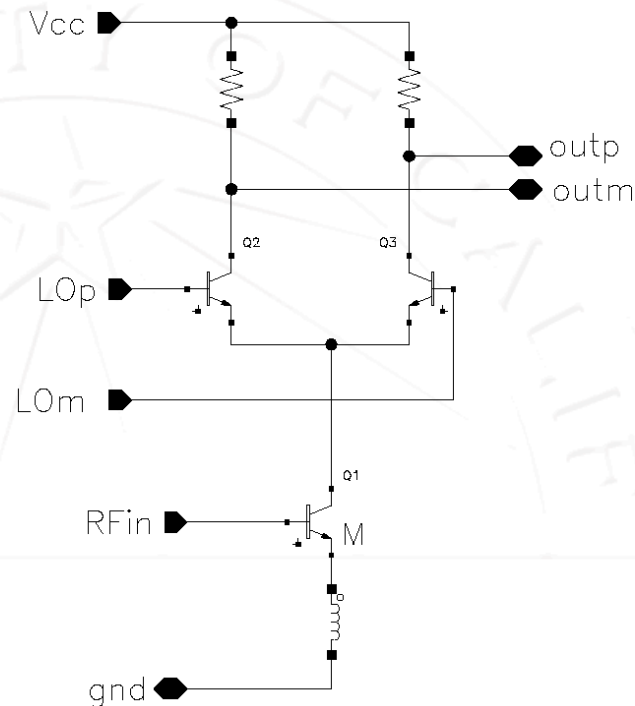
The above circuit is an example of a single-balanced mixer. The diff-pair Q2-Q3 behave like a differential amplifier to the LO signal. Q1 acts in this case as a current source in the absence of the RF signal. With this arrangement, any signal, or noise, occupying the IF band at the LO port will get amplified and transferred right to the mixer output causing significant degradation to the mixer NF.

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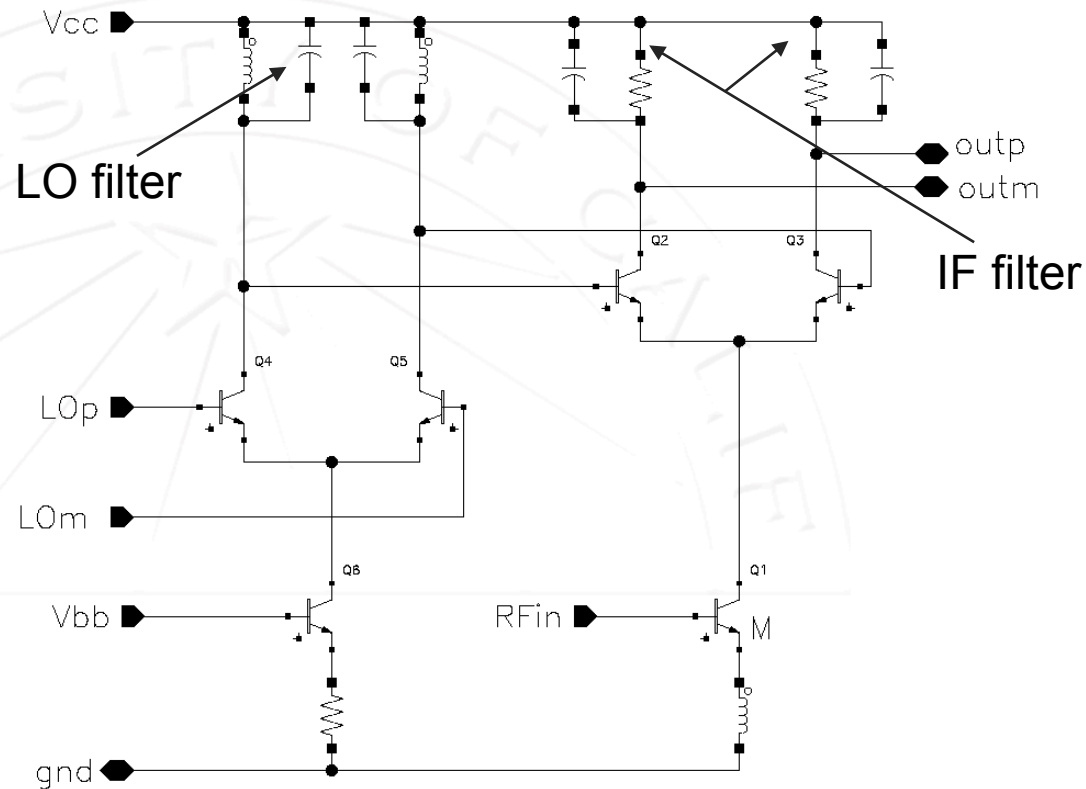
To reduce the sensitivity to LO noise, a bandpass filter is placed at the LO port to filter out any LO noise at the IF band, as shown above. It is important here to note that the effect of the LO noise can be highly suppressed if the LO signal is an ideal square wave with zero rise and fall time. This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

LO Port Noise
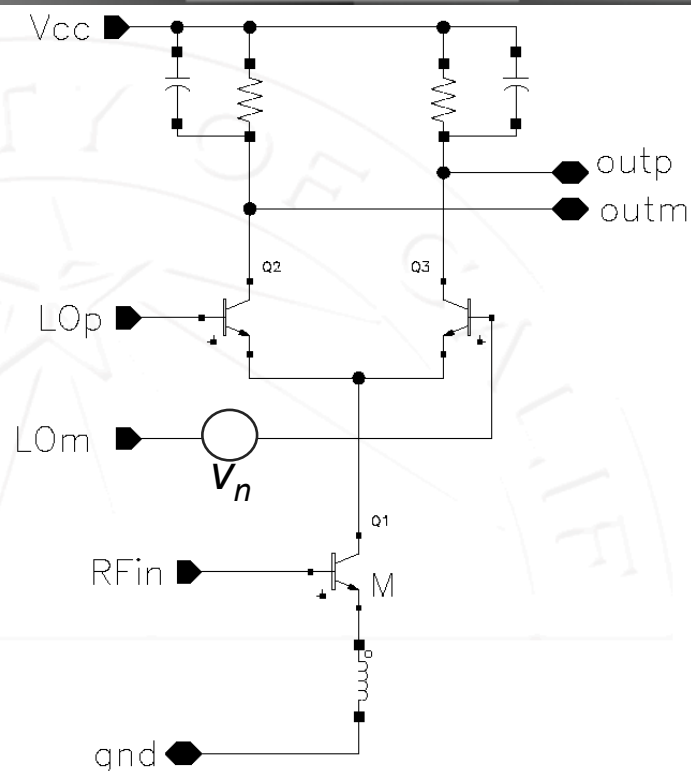
The LO noise can be modeled as a voltage source in series with the base of the diff-pair Q2-Q3. If the LO signal is an ideal square wave from a zero source impedance with a large voltage swing, the time over which both devices are on at the same time is zero. Therefore the impact of the LO noise is eliminated. However, if the LO signal has a finite rise and fall time, both Q2 and Q3 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 Mixer
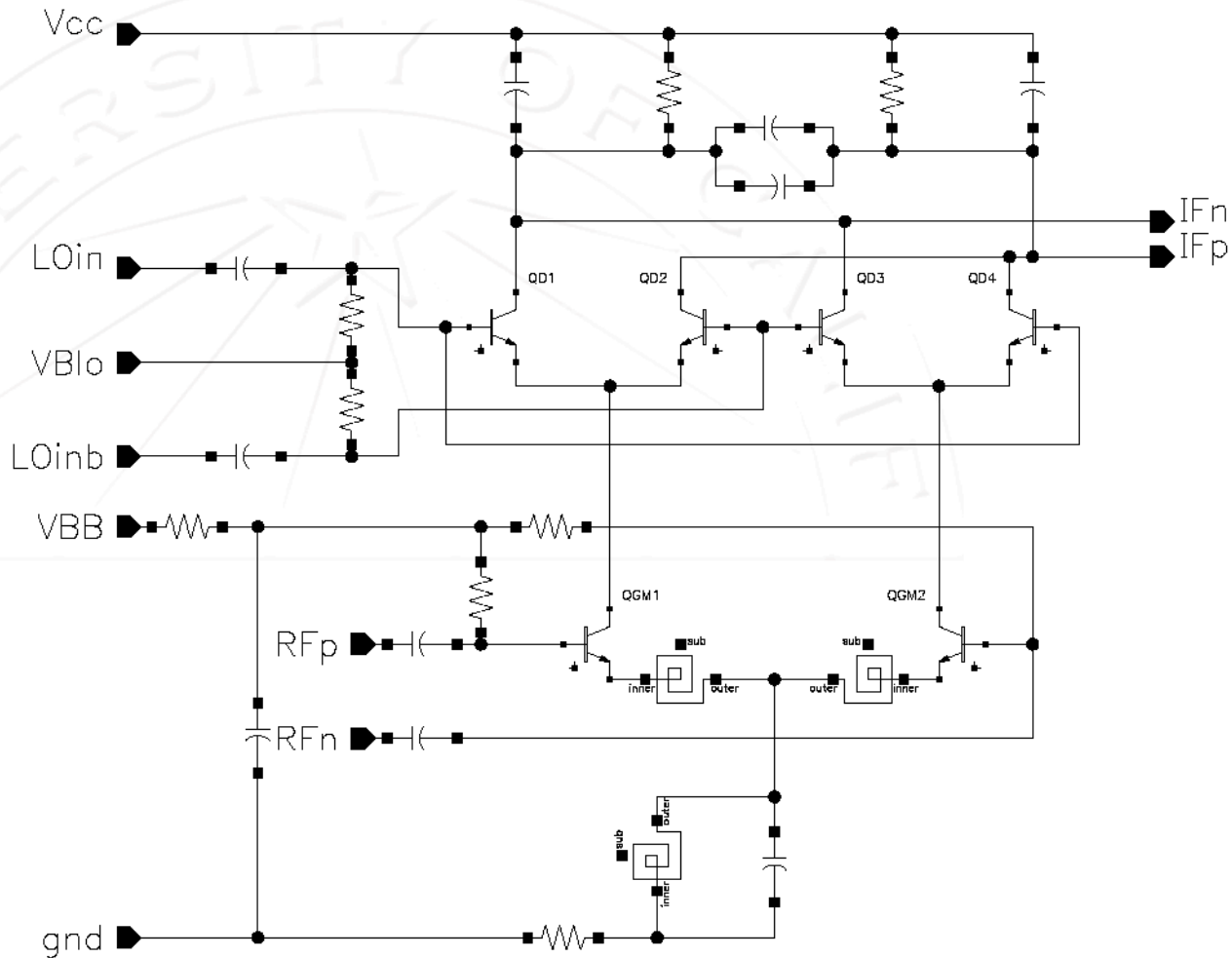
Advantages of Gilbert Mixer

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 600mVpp for bipolar)

5. Adequate NF in the range of ~6-12dB SSB, depending on gain and IP3.

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

Mixer 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 $\sim R_L/Z_E$, where R_L is the differential load resistor and Z_E 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 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_E 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.

$$Gain_{mix} = \frac{2}{\pi}\frac{R_L}{Z_E}$$

This result assumes that for the input Gm diff pair, $1/g_m << Z_L$, and that the device $f_T >> f$. Of course, the gain equation is simplified, but it gives some insight.

Gilbert Cell noise

The noise in the Gilbert cell mixer is divided between the input Gm cell, the 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 CE NF. However, usually the noise of the Gm cell is a bit compromised by increasing the emitter degeneration in order to increase the mixer IP3. 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 1.5~2dB.

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.

Mixer Quad Noise
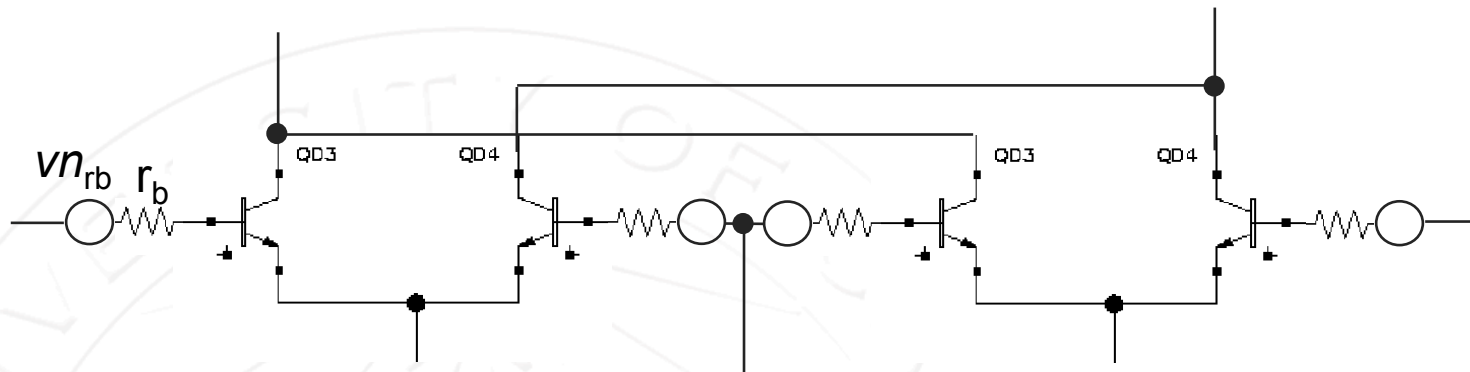
The noise of the quad is at its maximum when each of the quad devices is conducting an equal amount of current. This is because when QD3-QD4 devices are ON at the same time, they act as a simple diff-pair amplifying each other's noise, which is uncorrelated. This noise adds up at the output severely degrading the S/N. Since the quad diff-pairs are not degenerated, the amplifying gain of this noise is quite large. 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.

Mixer Load Noise
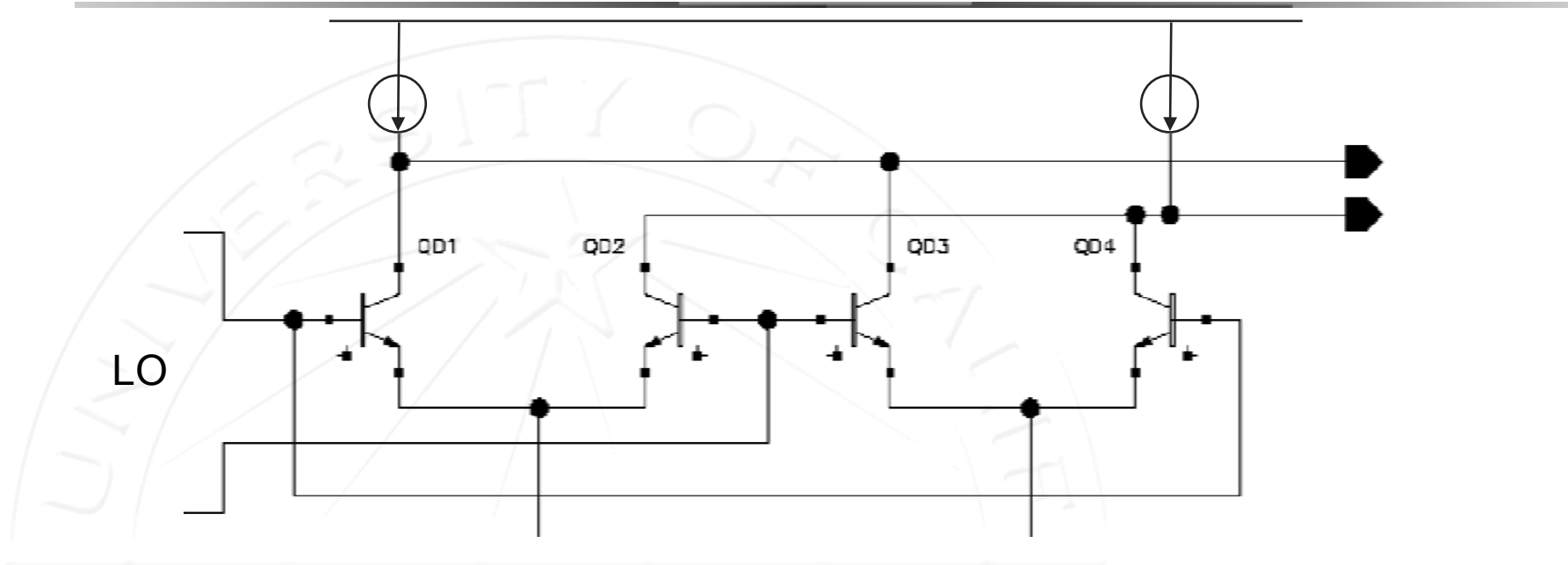
The mixer load also contributes to the overall NF. In low or zero IF receivers, the load is a simple resistor, whose noise contributes to the mixer NF. In some cases the mixer is configured as an OTA (operational transconductance amplifier) with the mixer output being current. This means the mixer is 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.

Gilbert Cell Linearity

IP3:

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 excellent 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].

Mixer IP2

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:

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

Mixer IM2

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:

$$V_i = V_1 \cos(\omega_1 t) + V_2 \cos(\omega_2 t)$$

$g_{m2}(+V_i/2)^2$ $g_{m2}(-V_i/2)^2$

$+V_i/2$ $-V_i/2$
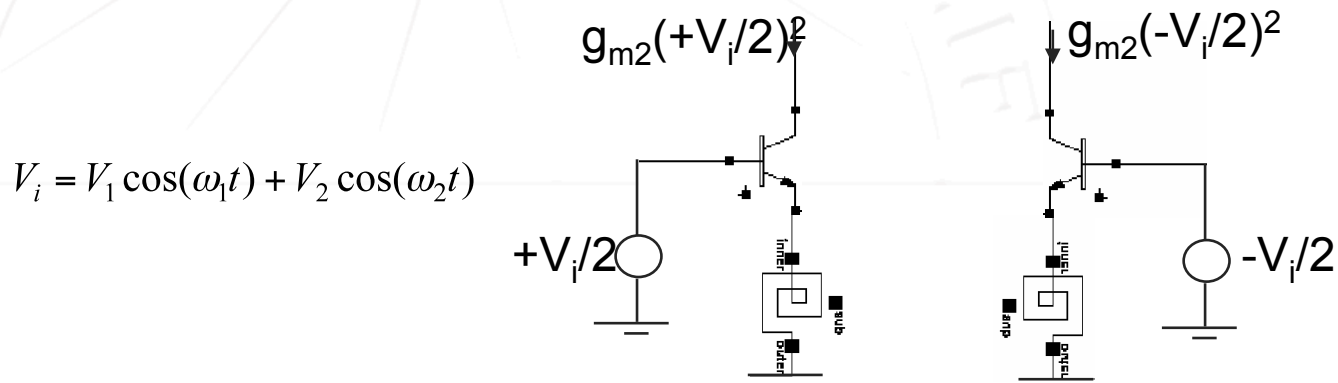
Any mismatch in the Gm stage half circuit (due to circuit components or layout) results in a finite differential IM2 at the output. This low frequency IM2 gets upconverted by the LO when it reaches the quad and so won't appear at the IF output of the mixer. However, any DC offset in the LO quad will result in this IM2 differential component to appear at the mixer output. This can be verified as follows.

Gilbert Quad DC Offset

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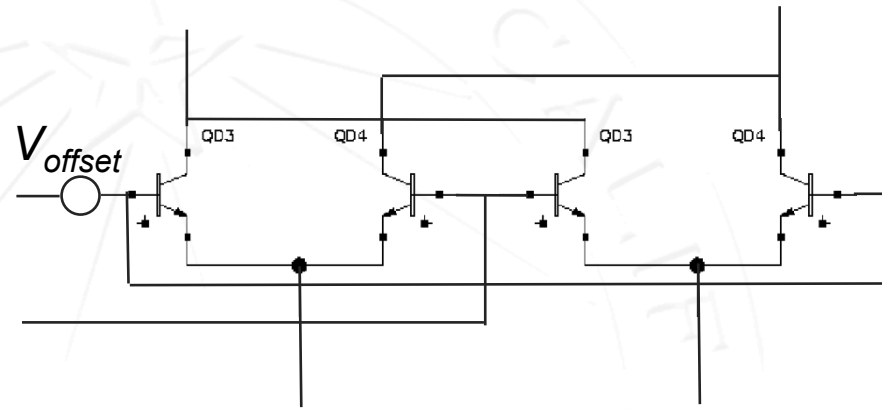
$$V_{LO} = V_{offset} + A\cos(\omega_{LO}t)$$

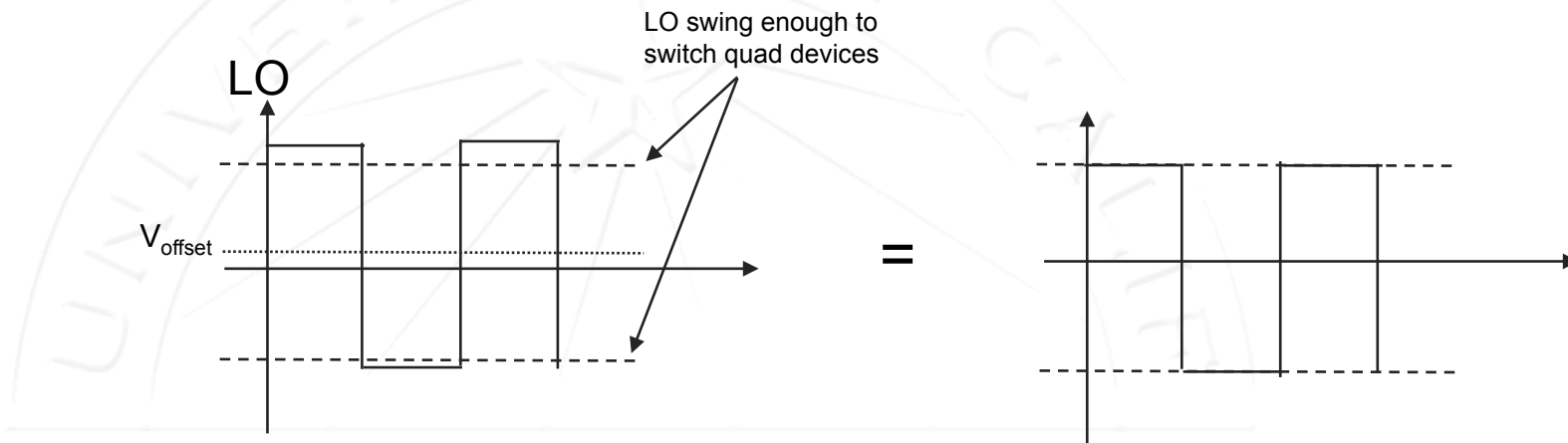with the IM2 signal coming out of the Gm cell being :

$$V_{Gm-IM2} = g_{m2}V_1V_2\cos(\omega_1 - \omega_2)t$$

the IF output then is (after lowpass filter):

$$V_{IF} = V_{LO} \times V_{Gm-IM2} = V_{offset}g_{m2}V_1V_2\cos(\omega_1 - \omega_2)t$$

Reducing IM2

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 size.

● Use low bias resistors at the base to minimize offsets due beta mismatch (for bipolar design)

● 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)

● Use 50% duty cycle LO generation scheme

Duty Cycle Mismatch

- IM2 due LO duty cycle error:

LO

V_{DC}
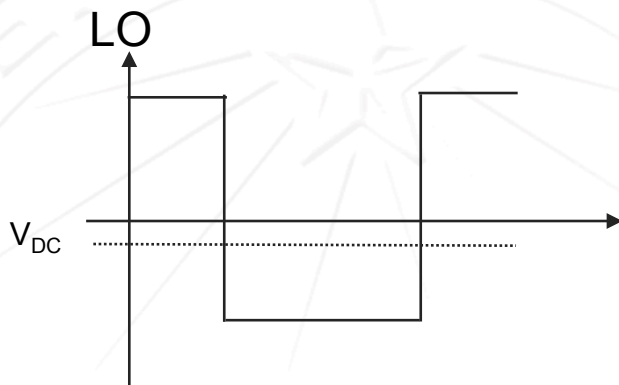
if LO does not have an exact 50% duty cycle, some IP2 degradation can occur. This is because a none 50% duty results in an effective DC offset of the LO signal applied to the quad. This DC offset will not be different from the offset due to the quad circuitry itself in providing a path of the IM2 generated within the Gm cell to find its way to the mixer output.

IM2 Generated By Quad

RF → ⊗ BB

LO

Load mismatch can seriously degrade IP2, even if everything else is perfect
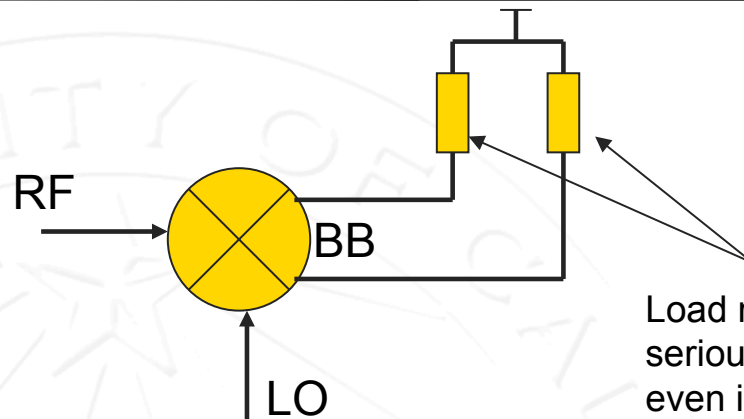
● IM2 generated within the quad itself:

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.
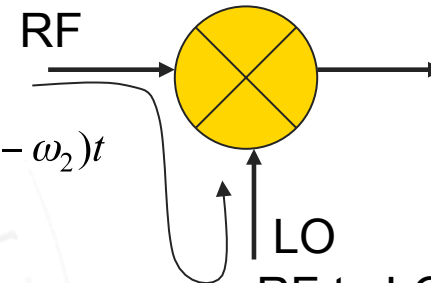
IM2 Due to RF→LO Leakage

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.

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

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

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

RF

LO

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 jammers, all of which appear at the mixer output IF port. In some systems the jammers 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 jammer 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:

IM2 due to Jammers

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

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

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

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

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

● IM2 due LNA:

The jammers 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 of hammers. 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 (usually it is mixer and baseband filter limited).
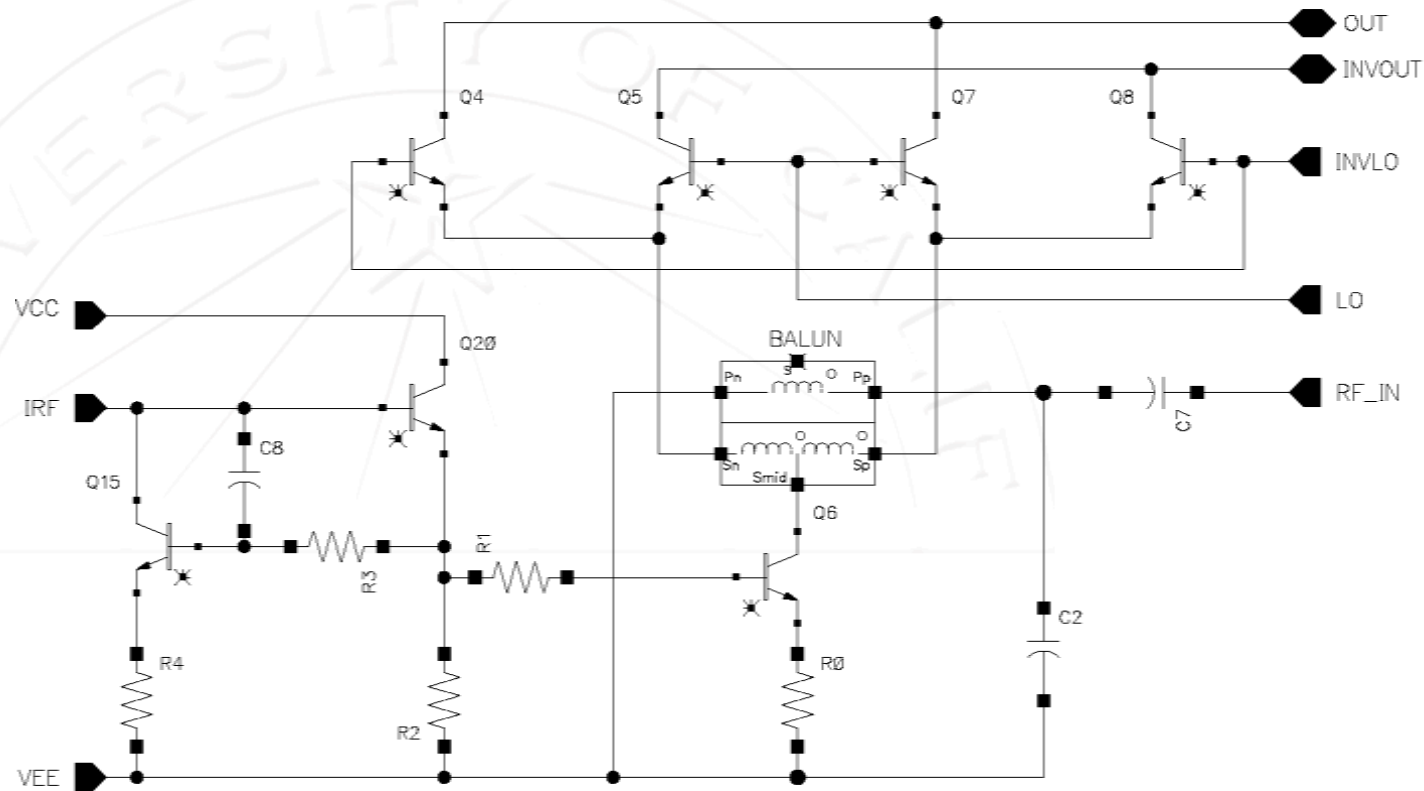
RF Mixer Design II
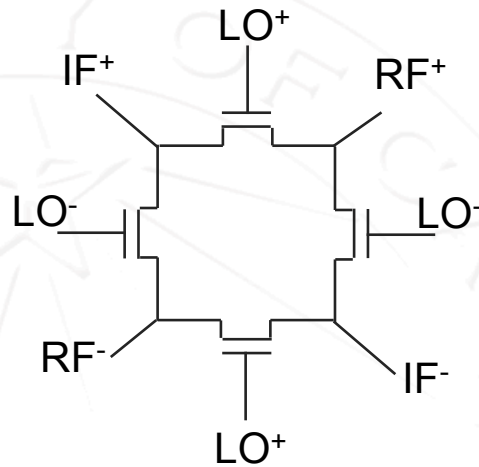
High IIP3 Gm Cell

The passive Gm stage results in high IP3 for very low current (1mA or so). The NF, however, is relatively large, 12dB or so because of the balun loss as well as "unshielded" quad shot noise from the input.

BiCMOS Mixer

CMOS Gm cell has a high IP3 for a given current compared to bipolar with small or even no degeneration. The bipolar quad will ensure low 1/f noise at the output and fast switching for good IP2 and IP3. Note that the Gm cell 1/f noise gets upconverted by the quad away from the baseband output. Such mixer can have an IP3 of +10dBm with <4mA of current at a gain of 10dB.
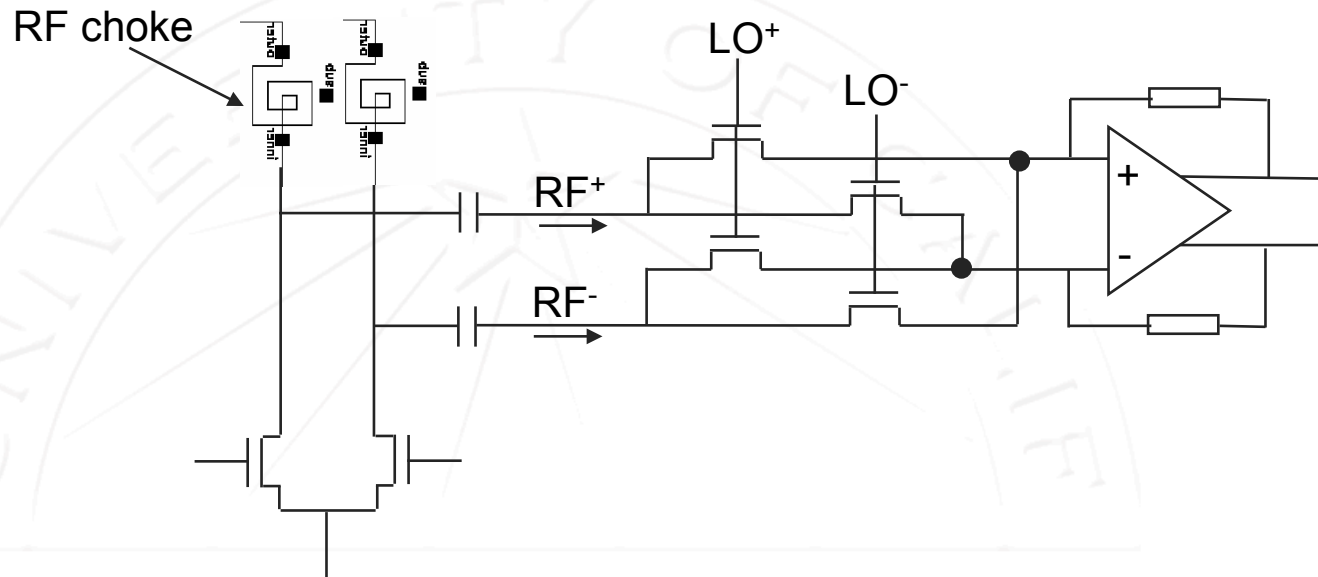
Passive MOS Commutator

The advantages of the passive MOS commutating mixer:
- very good linearity
- zero current consumption
- no 1/f noise (no DC current)
- small area

The disadvantages
- no gain, rather it has loss ranging from 4~6dB
- large required LO drive, almost rail to rail (power consumption!)

Passive Current Commutator

- LNA output RF current is fed into the passive mixer
- no DC current in passive mixer results in low 1/f mixer noise
- virtual ground of opamp improves overall linearity since mixer output (and associated nonlinear parasitic caps) does not swing in such configuration.
- mixer device sizing for min loss and acceptable LO drive.
- opamp is designed and device sized for best 1/f noise corner

Passive Mixer Filtering Effect
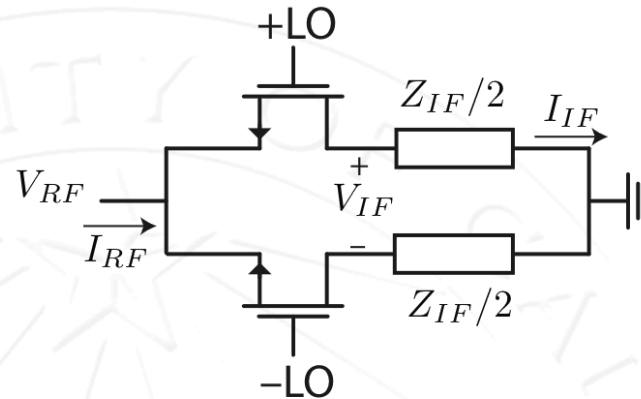
- The RF current is modulated to a baseband current given by

$$v_{IF}(t) = \left(\frac{2}{\pi} i_{RF}(t) \cos \omega_{LO} t \right) \star Z_{IF}(t)$$

- Due to the lack of isolation in a passive mixer, the RF side of the switch is affected by the IF voltage. The RF side sees the voltage drop across the switch and a modulated baseband waveform (multiplied by +/-1)

$$v_{RF}(t) = R_{sw} i_{RF}(t) + \frac{4}{\pi^2} \cos \omega_{LO} t \left(i_{RF}(t) \cos \omega_{LO} t \star Z_{IF}(t) \right)$$

Passive Filtering (cont)

$$
\begin{aligned}
V_{RF}(s) \;=\;\; & R_{sw} I_{RF}(s) + \\
& \frac{1}{\pi^2} \left(\, I_{RF}(s) Z_{IF}(s + j\omega_{LO}) + \right. \\
& I_{RF}(s) Z_{IF}(s - j\omega_{LO}) + \\
& I_{RF}(s - 2j\omega_{LO}) Z_{IF}(s - j\omega_{LO}) + \\
& \left. I_{RF}(s + 2j\omega_{LO}) Z_{IF}(s + j\omega_{LO}) \,\right)
\end{aligned}
$$

- We see that the baseband filter response is itself frequency translated and converted to a high Q bandpass characteristic at the RF port

- Thus the baseband filter can attenuate out of band blockers at the RF port itself. This leads to a high IIP3 for these style of mixers.

- Note that the reverse isolation led to this result. In an active mixer, the Gm stage is isolated from the IF load and no frequency translation of the load takes place.
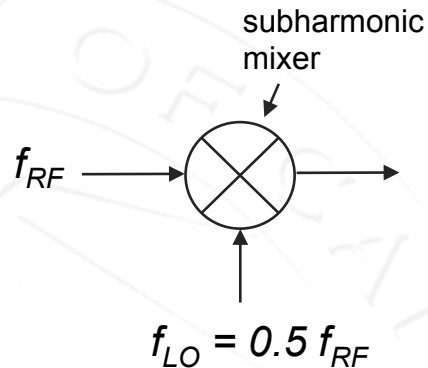
Op-Amp Noise Amplification
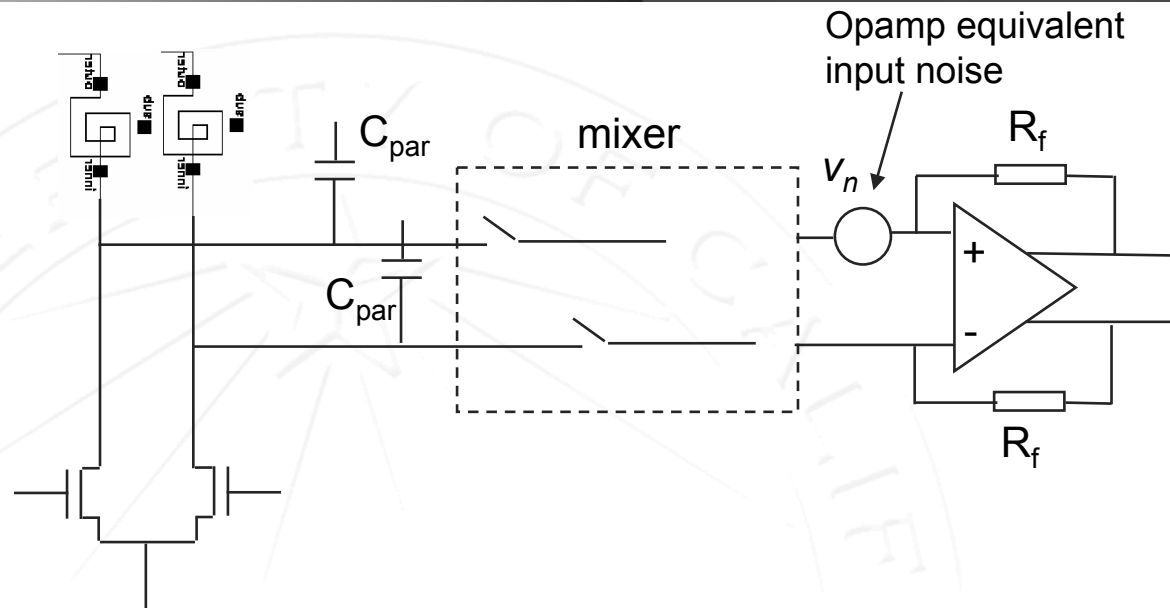
- the parasitic capacitor at the mixer input (due to mixer itself, LNA out, or layout) results in an effective switched-capacitor resistor due to the mixer switching action. The value of this resistor is

$$R_{par} = \frac{1}{4\pi f_{LO} C_{par}}$$

- the noise of the opamp gets gained up to the output by: $v_{n\_opamp\_out} = \left(1 + \frac{R_f}{R_{par}}\right) v_n$

- to minimize this noise, the LNA inductive load must be designed to resonate with all parasitic capacitors at the mixer input to provide high impedance.

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Sub-harmonic Mixer Design

subharmonic
mixer

$$f_{RF} \longrightarrow \bigotimes \longrightarrow$$

$$f_{LO} = 0.5\, f_{RF}$$

The subharmonic mixer is driven by an LO signal that is an integer fraction, or subharmonic, of the desired LO frequency. For example, if the RF signal is 2GHz, and the desired LO is 2GHz for direct conversion, a subharmonic mixer will be driven by a 1GHz LO signal. The advantages are:

1. Lower LO re-radiation through the antenna (LO leakage)

2. lower LO self mixing (lower DC offset at IF)

3. Relaxed requirement on the device switching speed.

4. Lower LO buffer current

25% LO Waveform

The idea is to generate an LO signal that is half the desired frequency, yet is rich with a second harmonic, then somehow use the second harmonic, along with the RF, to get desired IF.

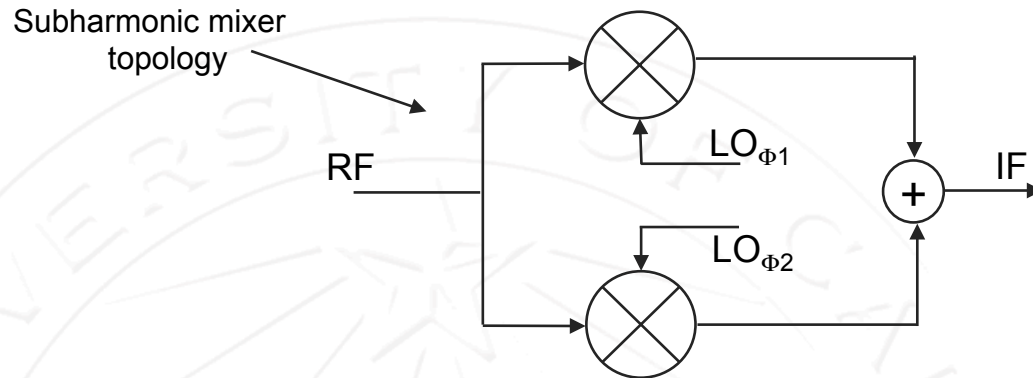A 50% duty cycle LO has no second harmonic. However a 25% duty cycle LO has.
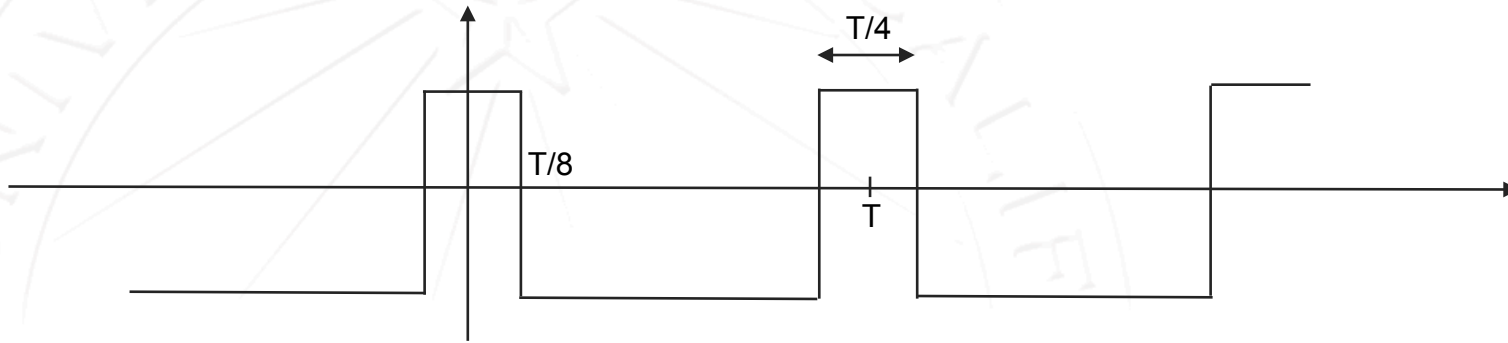
The Fourier series expansion of the above square wave is:

$$V_{\Phi 1} = \frac{1}{\pi} \left[\sqrt{2} \cos(\omega_0 t) + \cos(2\omega_0 t) + \frac{\sqrt{2}}{3} \cos(3\omega_0 t) - \frac{\sqrt{2}}{5} \cos(5\omega_0 t) + \ldots \right]$$

Where ω_0 is the subharmonic LO frequency = $2\pi/T$, which in this case is half the desired LO frequency.

Sub-Harmonic Mixer

The subharmonic mixer topology uses two identical mixers excited by two phases of the 25% duty cycles. The RF signal is multiplied by these two delayed 25% LO signals and the IF is added in phase at the output, as shown in the above figure.

The two LO signals, $V_{\Phi 1}$ and $V_{\Phi 2}$, are T/2 delayed relative to each other.

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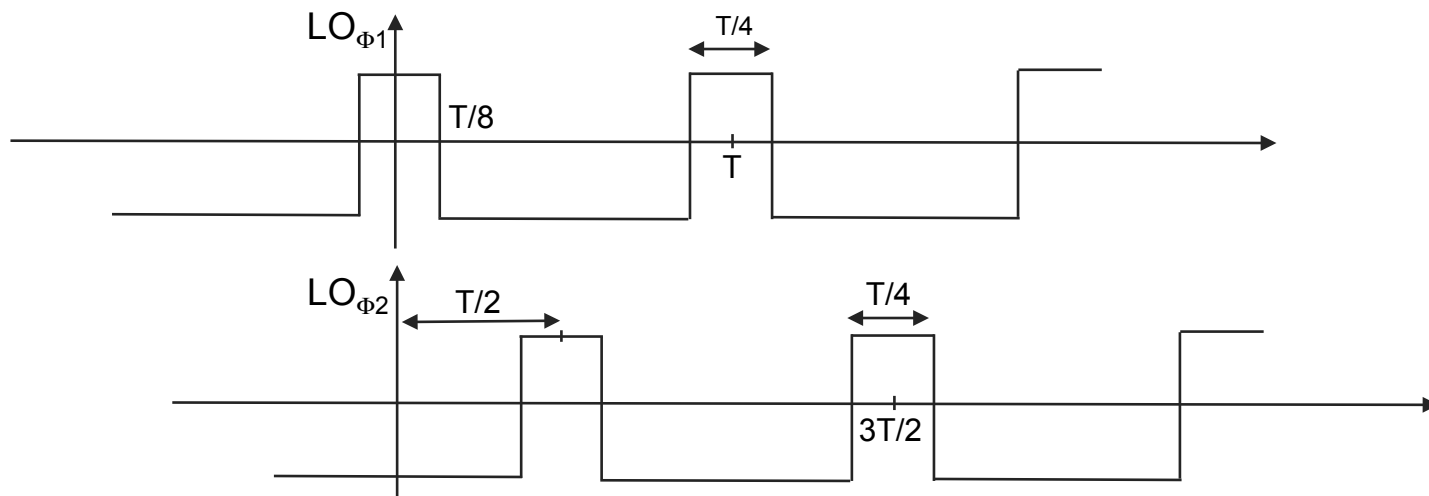
Sub-Harmonic Waveforms

$$V_{\Phi 2} = \frac{1}{\pi}\left[-\sqrt{2}\cos(\omega_0 t) + \cos(2\omega_0 t) - \frac{\sqrt{2}}{3}\cos(3\omega_0 t) + \frac{\sqrt{2}}{5}\cos(5\omega_0 t) + ...\right]$$

Looking at the combined IF output of the subharmonic mixer we can write:

$$V_{IF} = V_{RF}V_{\Phi 1} + V_{RF}V_{\Phi 2} = V_{RF}\left[\frac{2}{\pi}\cos(2\omega_0 t)\right]$$

As seen, the resulting IF output is the product of the RF signal and an "effective" LO that has twice the subharmonic LO frequency, which is in fact the desired LO.

The above analysis was for only one mixer. The question then, how can one build quadrature subharmonic mixer for both I and Q channels?

The answer comes in the way to generate a $\sin(2\omega_0 t)$ LO signal using similar 25% duty cycle signals as follows.

I/Q Sub-Harmonic Mixer
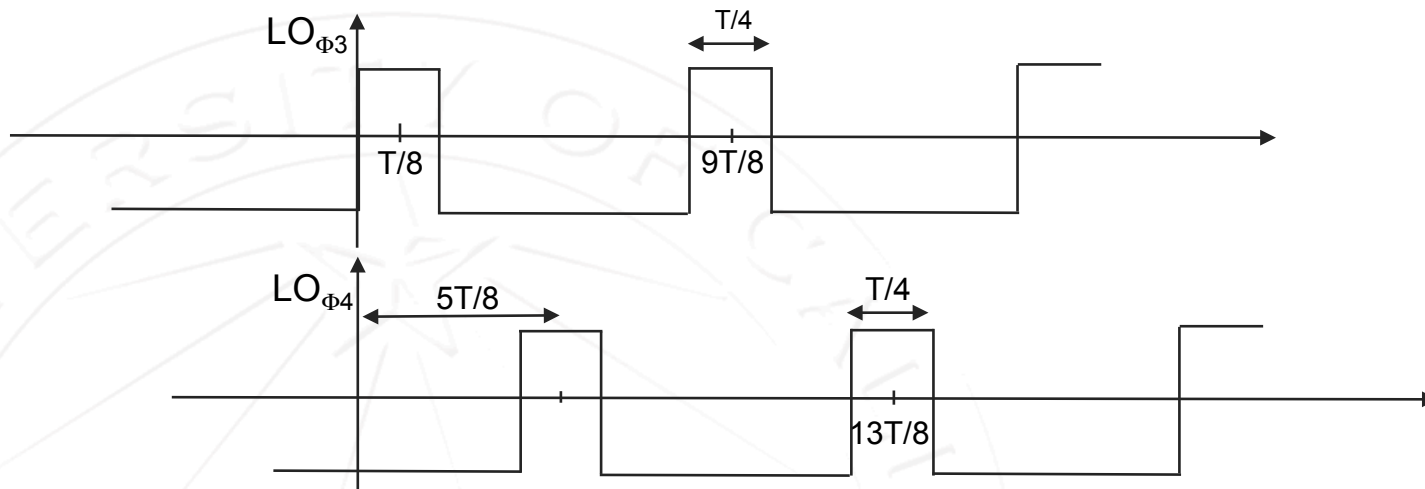
The LO signal $LO_{\Phi 3}$ is delayed T/8 relative to $LO_{\Phi 1}$. $LO_{\Phi 4}$ is delayed by T/2 relative to $LO_{\Phi 3}$.

The fourier series expansion of the above square waves is:

$$V_{\Phi 3} = \frac{1}{\pi}\left[\sqrt{2}\cos(\omega_0 t - \frac{\pi}{2}) + \sin(2\omega_0 t) + \frac{\sqrt{2}}{3}\sin(3\omega_0 t - \frac{\pi}{2}) + \frac{\sqrt{2}}{5}\cos(5\omega_0 t - \frac{\pi}{2}) + ...\right]$$
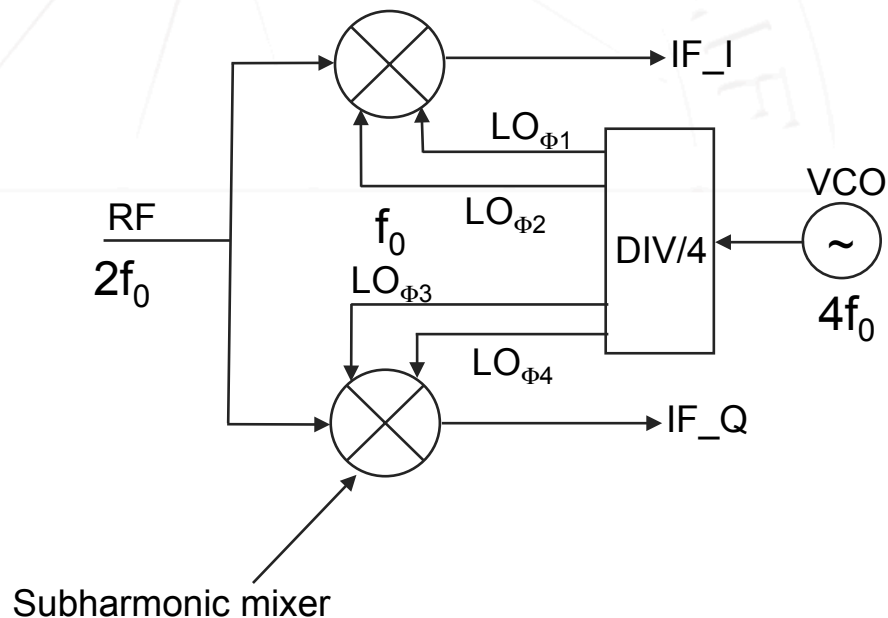
$$V_{\Phi 4} = \frac{1}{\pi}\left[-\sqrt{2}\cos(\omega_0 t - \frac{\pi}{2}) + \sin(2\omega_0 t) - \frac{\sqrt{2}}{3}\sin(3\omega_0 t - \frac{\pi}{2}) - \frac{\sqrt{2}}{5}\cos(5\omega_0 t - \frac{\pi}{2}) + ...\right]$$
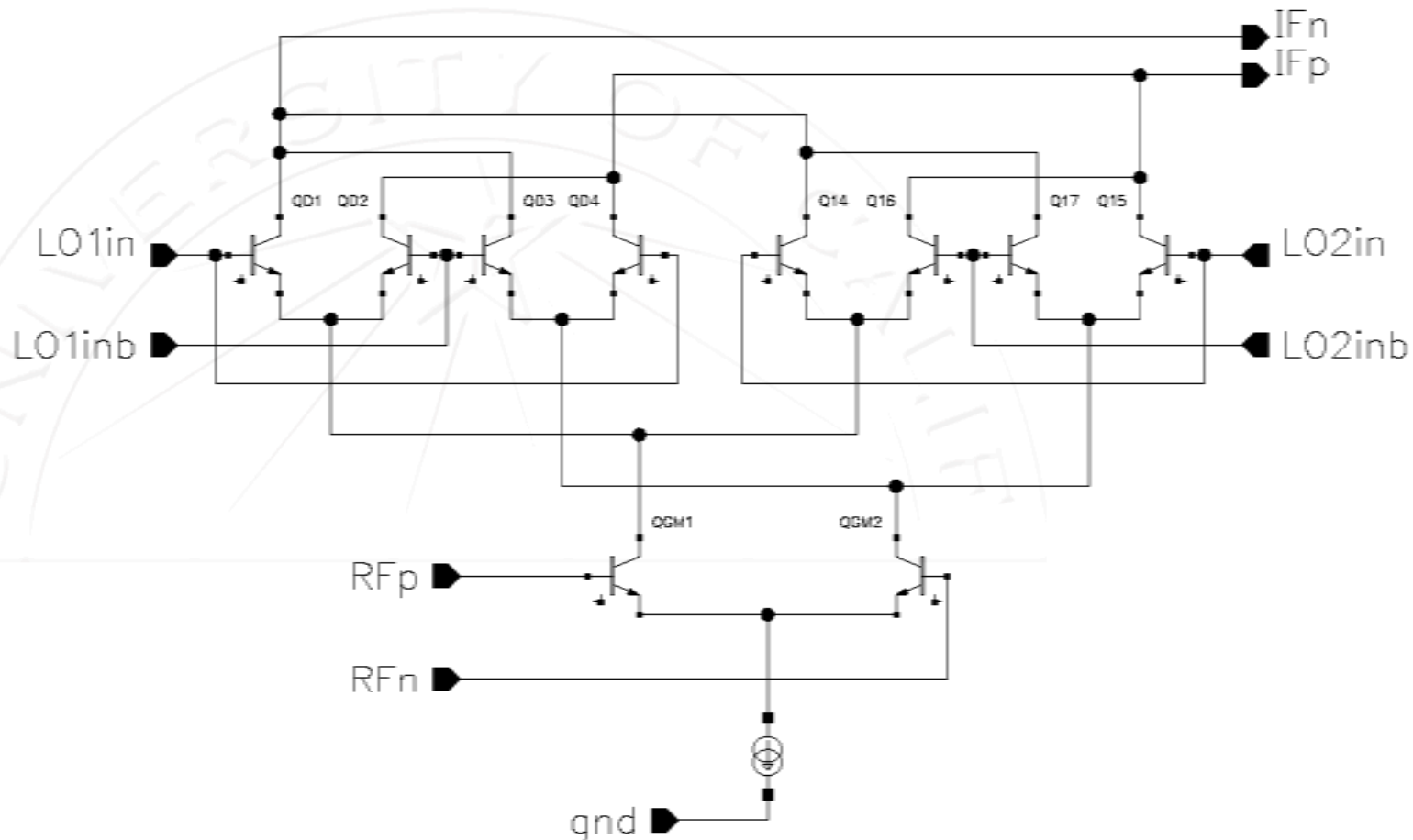
Generation of LO

Looking at the combined IF output of the subharmonic mixer we can write:

$$V_{IF\_Q} = V_{RF}V_{\Phi 3} + V_{RF}V_{\Phi 4} = V_{RF}\left[\frac{2}{\pi}\sin(2\omega_0 t)\right]$$

In order to generate all 4 phases of the subharmonic LO signals, a divide by 4 prescalar needs to be used. This means the VCO needs to run at 4X the subharmonic LO, 2X the RF signal for direct conversion.
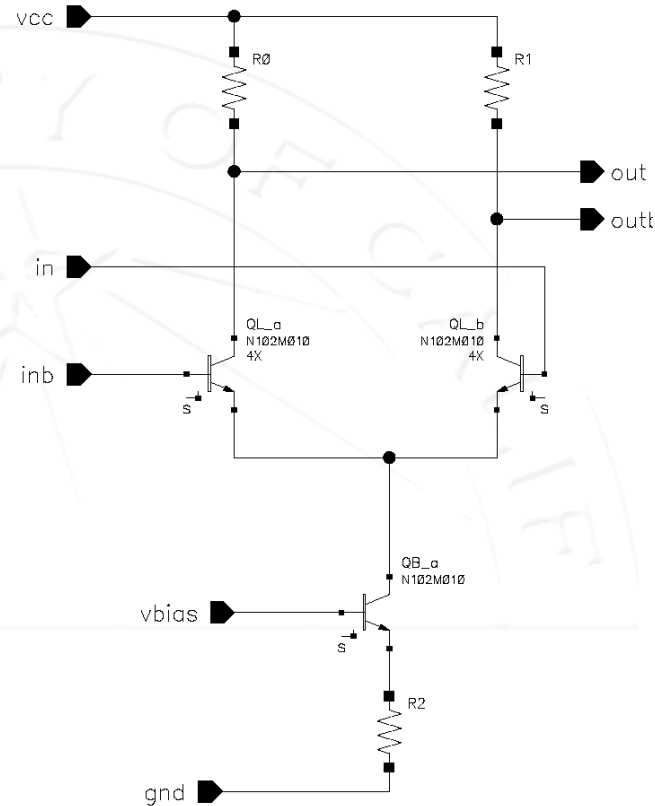

Subharmonic mixer

Sub-Harmonic Realization

The above is a simplified implementation of one subharmonic mixer. Since a subharmonic mixer is two mixers in one driven by the same RF input but has two different LO phases, a single common Gm stage is used.

LO Buffers

LO buffers:

It is evident by now that the LO signal "shape" highly impacts the mixer linearity and noise. The goal is to achieve as close to a square wave LO as possible. The above is a simple way to "square" an LO signal. However, the rise and fall time is severely limited by the RC time constant of the LO squaring circuit load and the mixer input quad capacitance.

Emitter Followers in LO Buffer
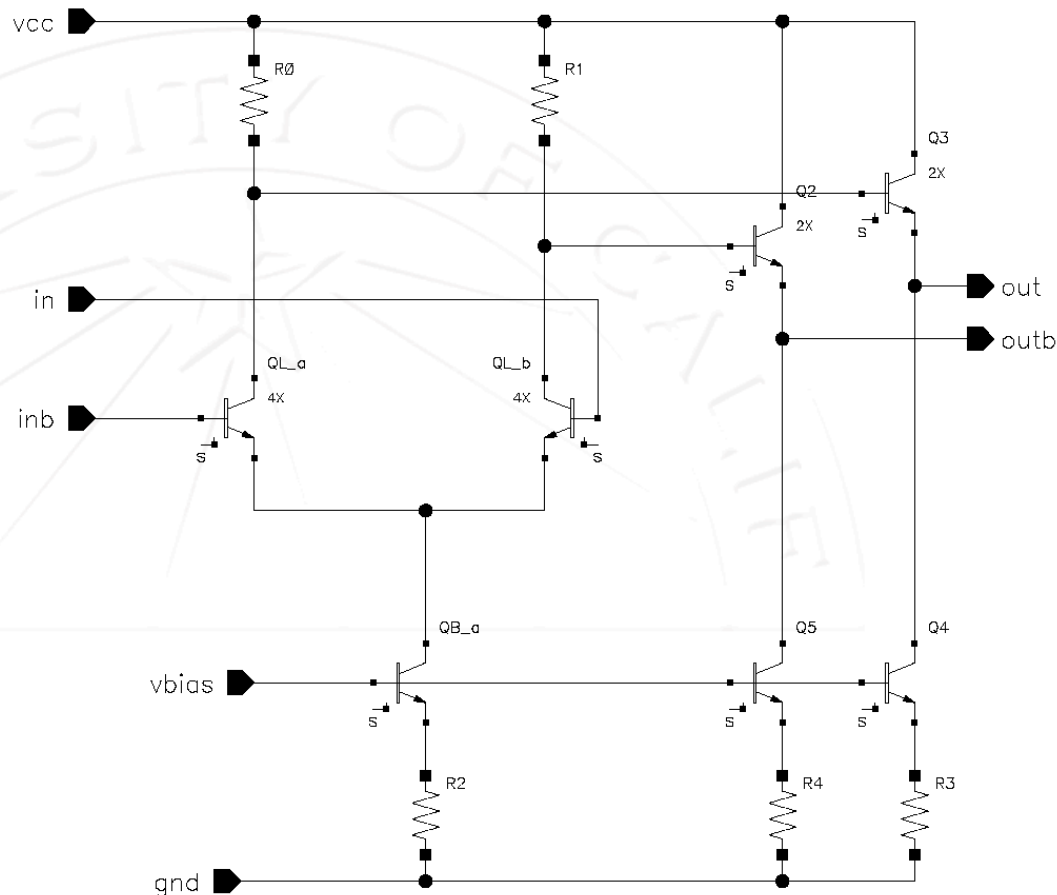
An emitter follower following the limiter circuit provides a low-impedance to drive the large mixer quad devices. However, the fall time will be limited by the emitter follower current source in discharging the quad capacitance. This means to achieve sharp fall time, more current needs to be used to bias the emitter followers.
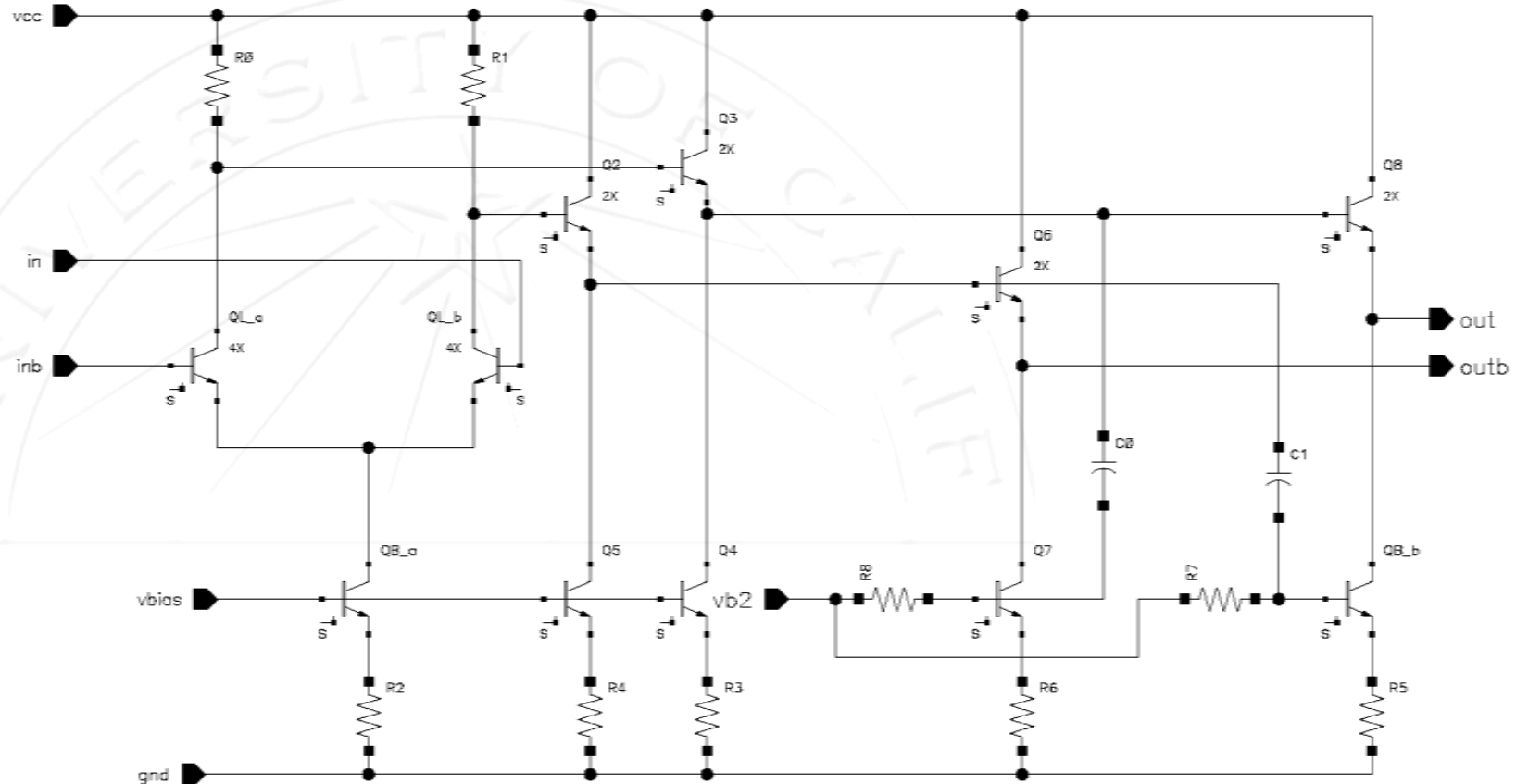
Faster Emitter Followers
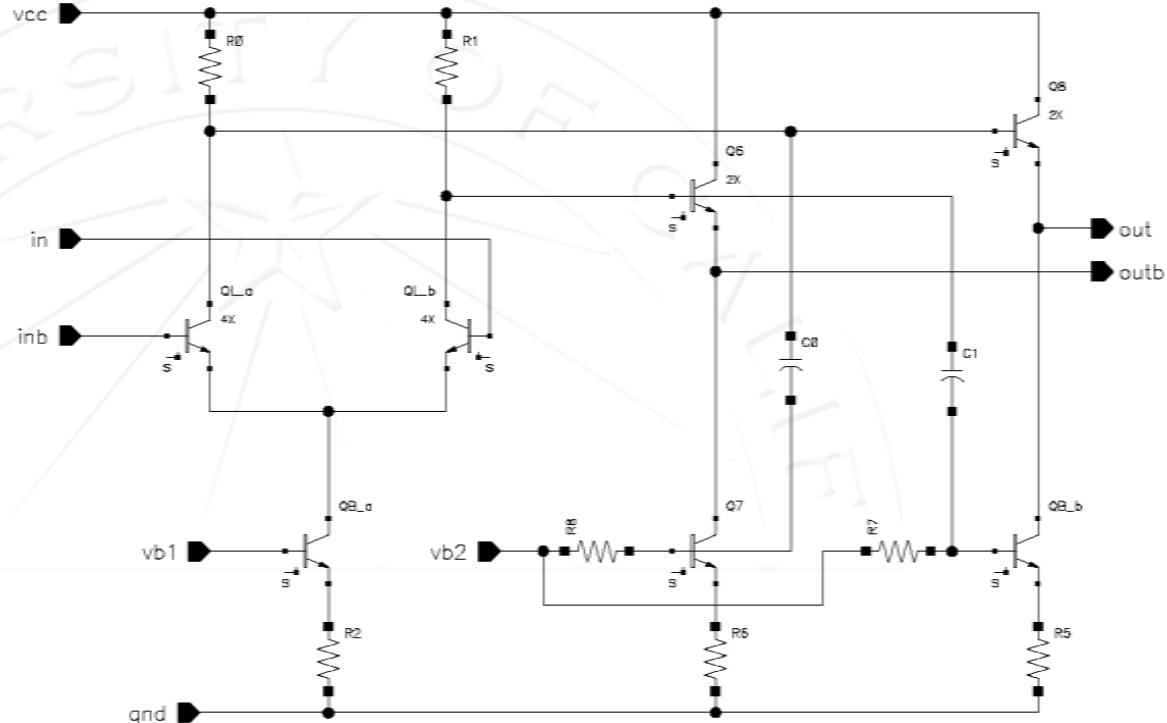
In order to reduce power consumption and make the LO buffer more efficient, some push-pull technique is required. Most bipolar technologies lack fast pnp devices for class AB buffers. Therefore an all-npn push-pull buffer is used as shown above. The out of phase signal is capacitively coupled to the base of the current source to increase its current during falling edge. The time constant of the load R and the coupling cap C limits both rise and fall time. A typical value is ~5V/ns for a 50GHz bipolar technology.

Push-Pull LO Buffer

An improved push-pull buffer uses emitter followers to shield push-pull coupling cap from the limiter load. This circuit can achieve real sharp rise and fall time in the range of 10~15V/ns for 50GHz bipolar technology. A typical LO swing for a bipolar Gilbert cell quad is in the range of 600mVpp.

Active Push-Pull Technique
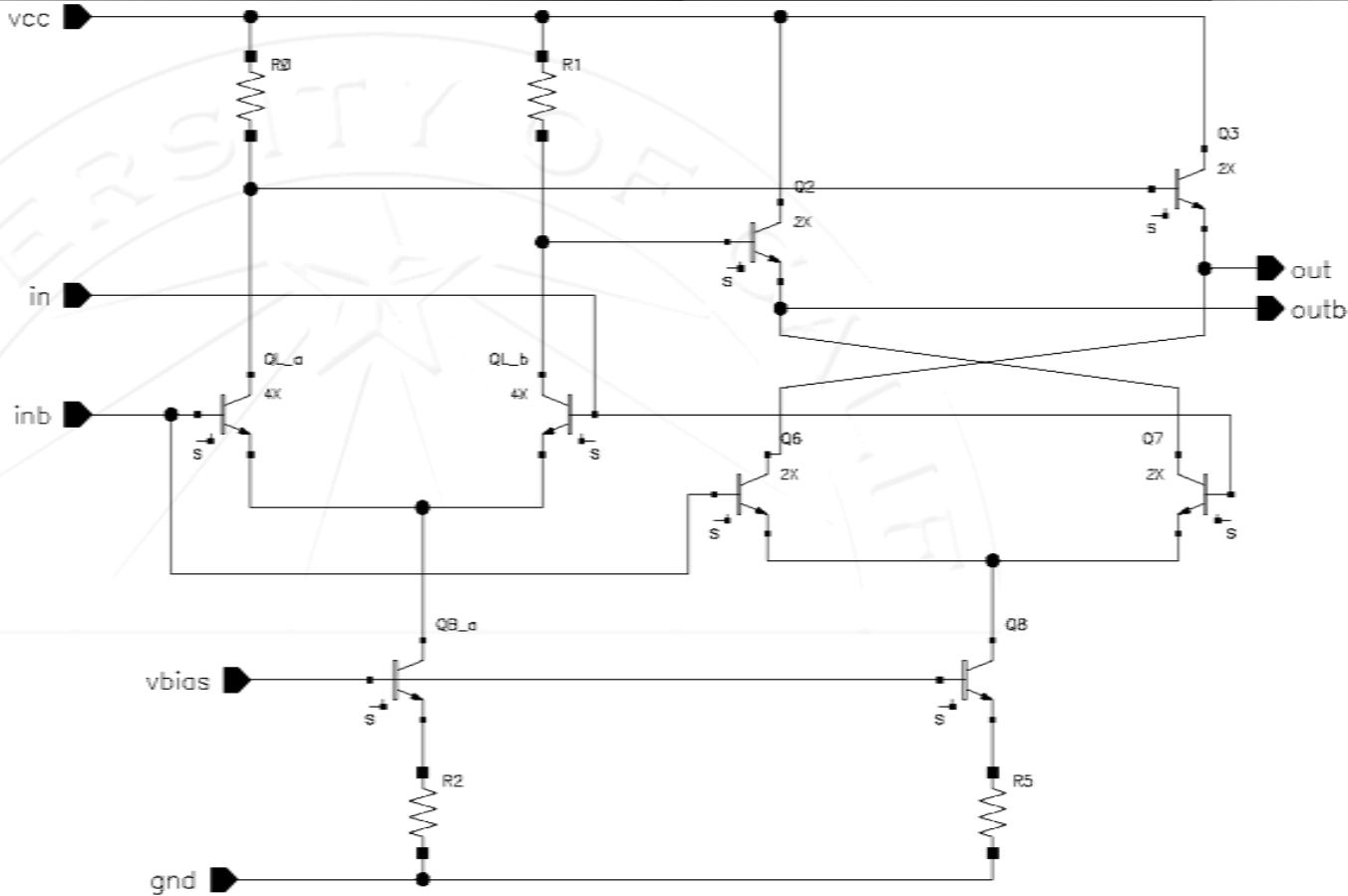
An active push-pull can also be used. The principle is to move the DC current from the "rising" branch to the "falling" branch, where it is needed the most. Note that this circuit is prone to oscillation If not carefully designed. This is because the 1/gm load of the push-pull gets larger as the current is steered to one side.

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Bleed Current in Active Push/Pull
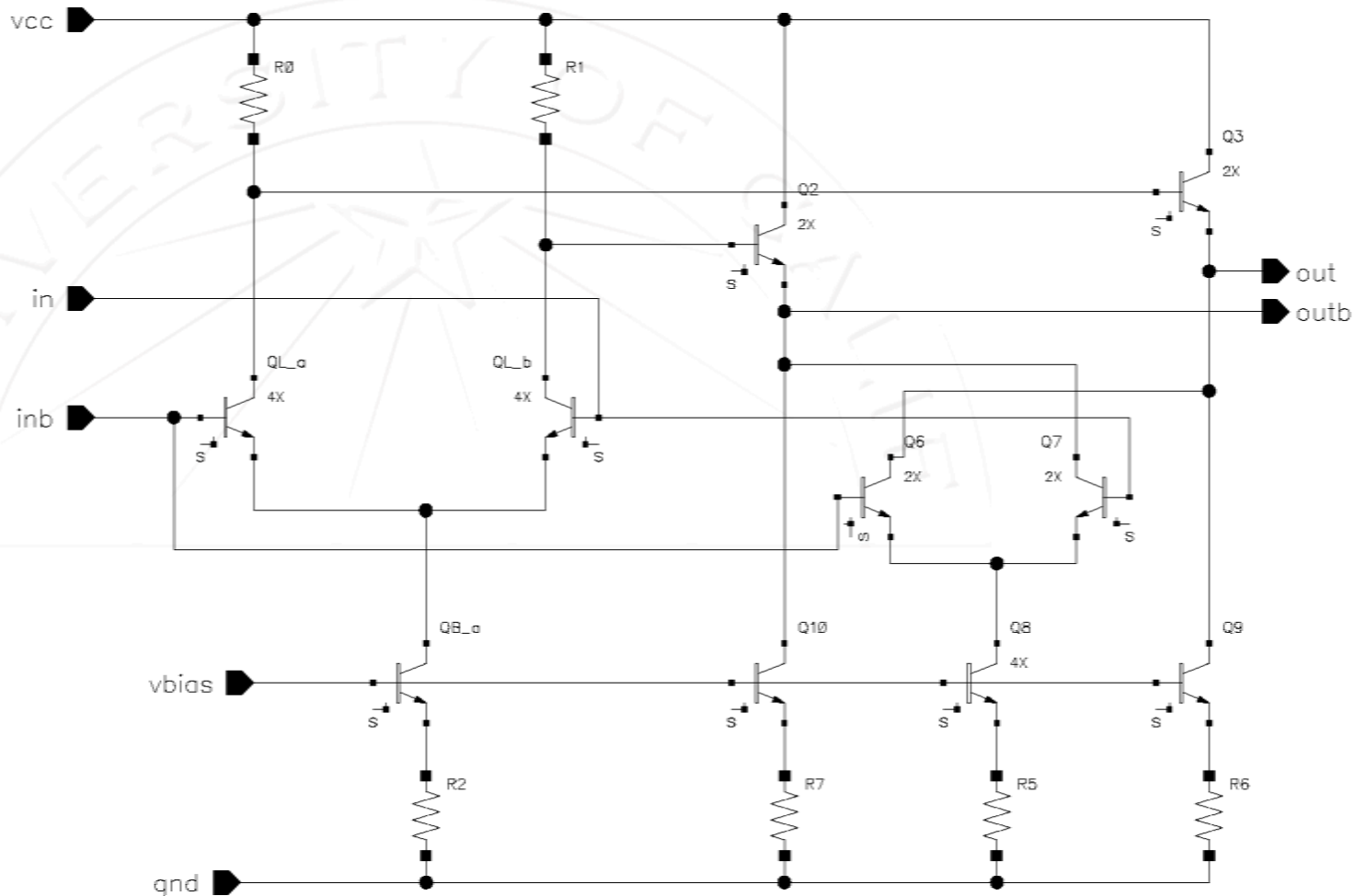
A bleed current can be used to limit the 1/gm load of the active push-pull.

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