ECEN 665

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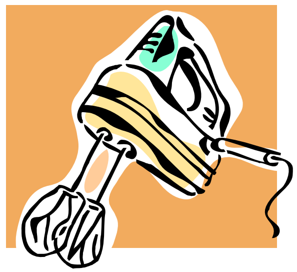
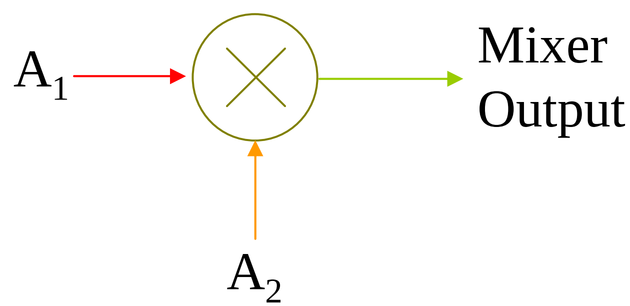


Mixer



Analog and Mixed-Signal Center, TAMU

What Devices Perform Frequency Translation?

- Linear, time-invariant systems can not generate spectral component not present in the input.
- Mixer must be non-linear or time-variant system.
- Historically, a lot of devices are being tried as mixers: electrolytic cells, magnetic ribbons, brain tissues, rusty scissors, vacuum tubes and transistors.

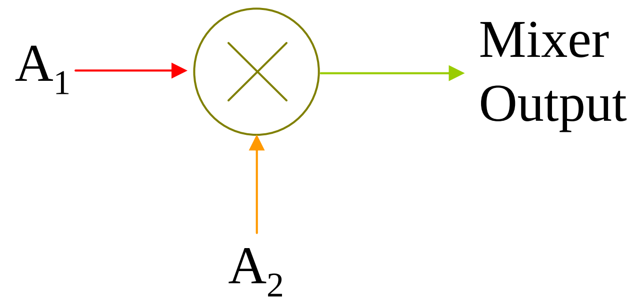
Virtually any nonlinear elements can be used as mixer. Some nonlinearities work better and more practical.

Mathematical Model

Most mixer implementations use some kind of multiplication of two signals in time domain:

RF×LO→ IF (down conversion)

IF×LO → RF (up conversion)



$$\left(A_1 \cos\omega_1 t\right)\times\left(A_2 \cos\omega_2 t\right)=\frac{A_1A_2}{2}\cos(\omega_1-\omega_2)t+\frac{A_1A_2}{2}\cos(\omega_1+\omega_2)t$$

- Up conversion filters out $\omega_1 - \omega_2$ component.
- Down conversion filter out $\omega_1 + \omega_2$ component.

Mixer Metrics

In order to evaluate the performance of mixers, several metrics are defined:

- ❑ Conversion gain/loss
- ❑ Noise figure
- ❑ Port isolations
- ❑ Linearity
- ❑ Power consumption

Mixer Metrics (cont'd)

Conversion Gain

Conversion gain or loss is the ratio of the desired IF output (voltage or power) to the RF input signal value (voltage or power). More specifically:

$$\text{Voltage Conversion Gain} = \frac{\text{r.m.s. voltage of the IF signal}}{\text{r.m.s. voltage of the RF signal}}$$

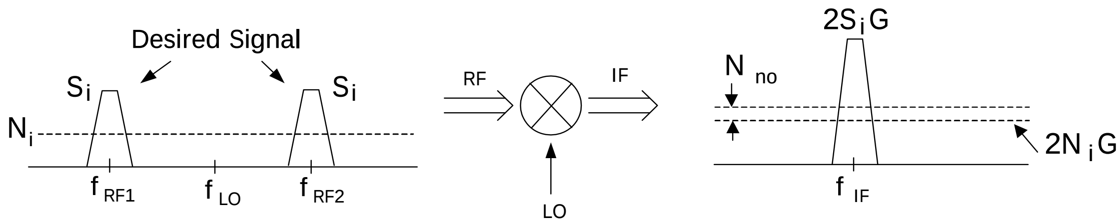
$$\text{Power Conversion Gain} = G_c = \frac{\text{IF power delivered to the load}}{\text{Available power from the source}}$$

The power gain definition is actually transducer power gain.

Mixer Metrics (cont’d)

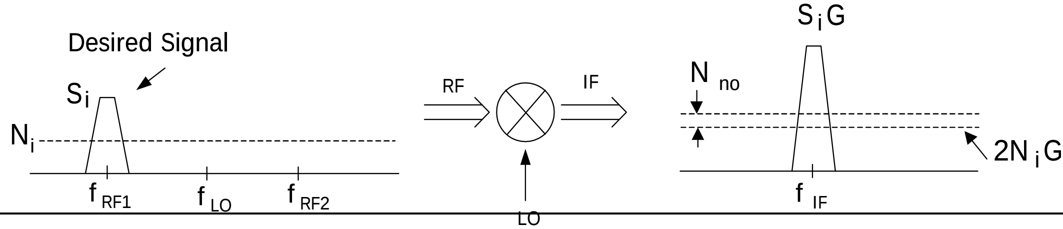
Mixer’s noise figure

- Double side-band (DSB)



$$F_{DSB} = 1 + \frac{N_{no}}{2N_i G}$$

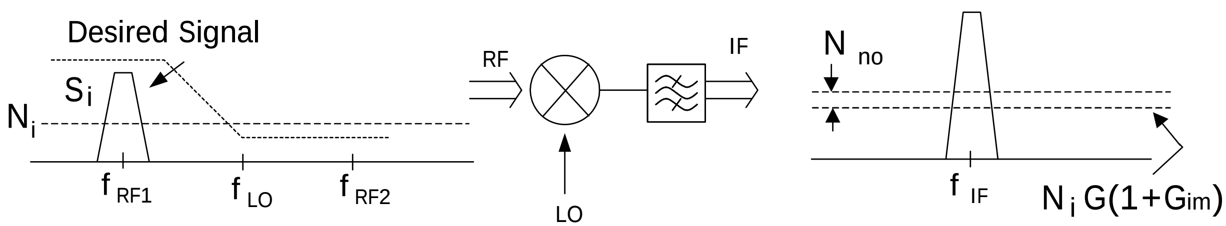
- Single side-band (SSB)



$$F_{SSB} = 2 + \frac{N_{no}}{N_i G}$$

$$F_{SSB} = 2F_{DSB}$$

- Single side-band with image rejection filtering



$$F'_{SSB} = F_{SSB} + \frac{G_{im}}{G} - 1$$

$$F'_{SSB} < 2F_{DSB}$$

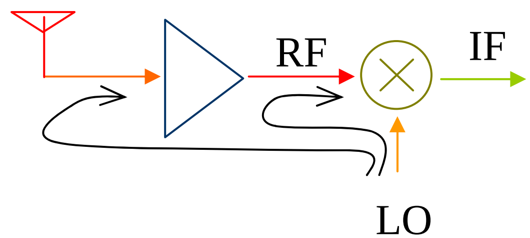
Where Nno, Ni, and G are the output noise power, the input noise power, and the gain of the system, respectively.

Mixer Metrics (cont'd)

Port isolation

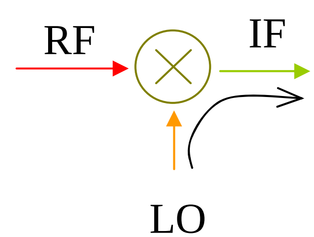
- LO-to-RF leakage, which will mix with LO again, causes self-mixing problem in direct conversion. Due to the nonzero reverse gain of LNA, the LO leakage may even reach the antenna through the LNA

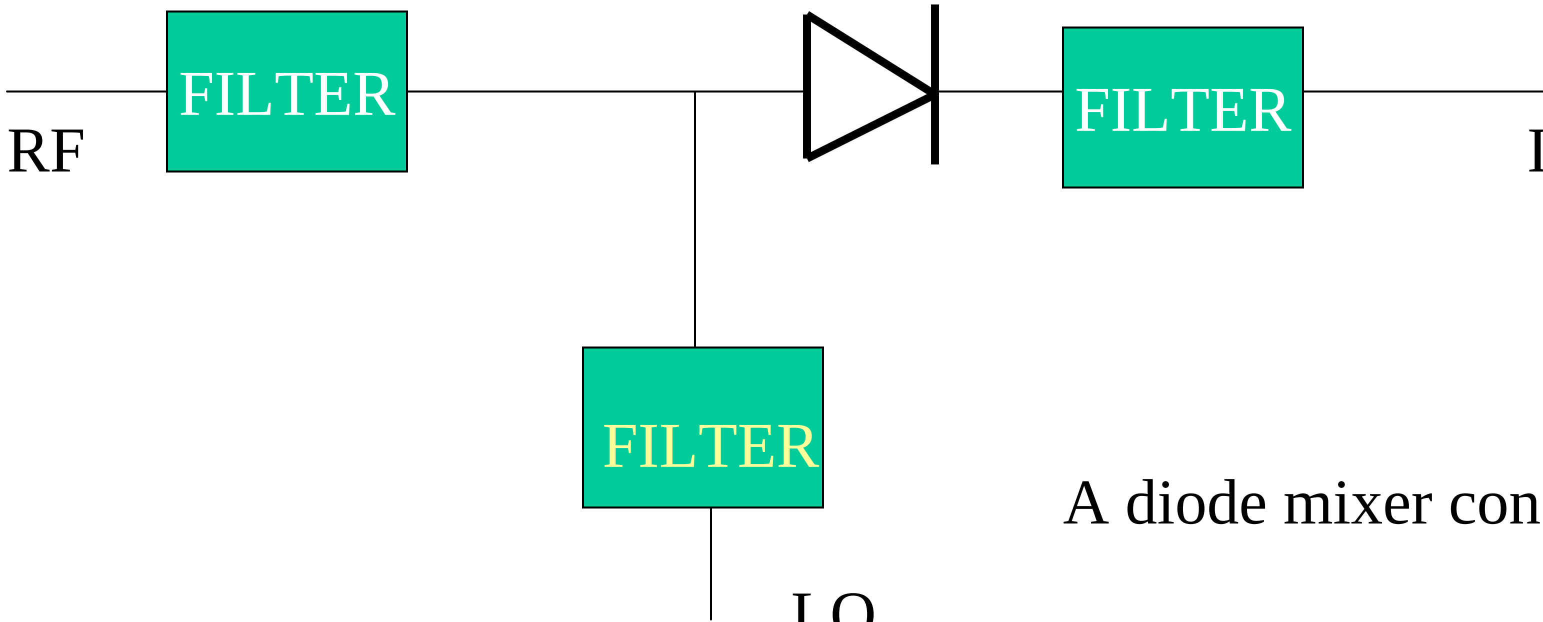
- LO-to-IF feed through may cause desensitization of the blocks following the mixer. (recall that the LO power is usually greater than that of the desired IF signal.)



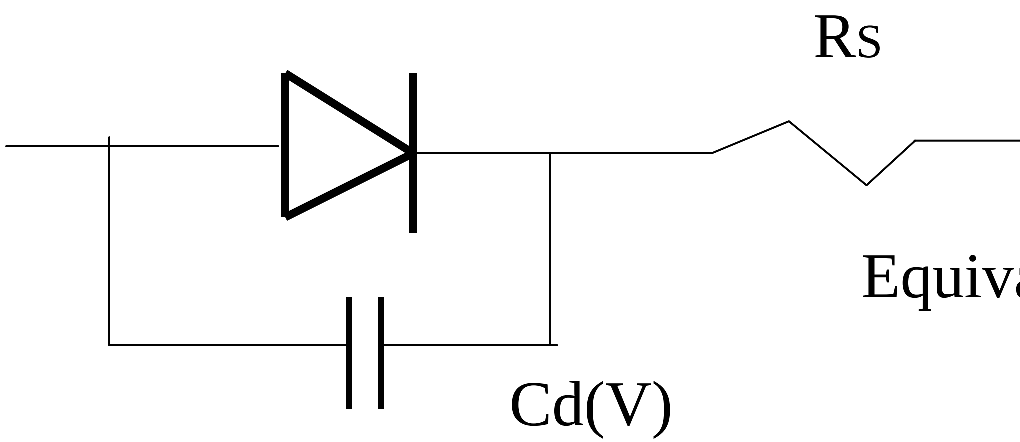
- RF-to-LO feed through allows interferers and spurs present in the RF signal interact with the LO.

- RF-to-IF feed through may cause problems in direct conversion architecture due to the low-frequency even-order inter-modulation product.





A diode mixer concept



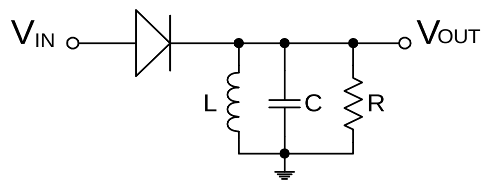
Equivalent Circuit

Mixer Topologies

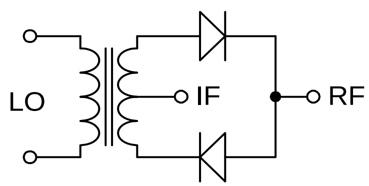
- Discrete implementations:
  - Single-diode and diode-ring mixers
  - Schottky barrier diode is preferred to regular diode due to its low junction capacitance and low series resistor
- IC implementations:
  - MOSFET passive mixer
  - Gilbert-cell based mixer
  - Harmonic mixer

Passive mixer topologies

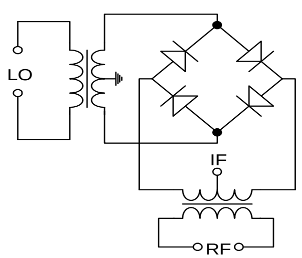
- Single-diode:



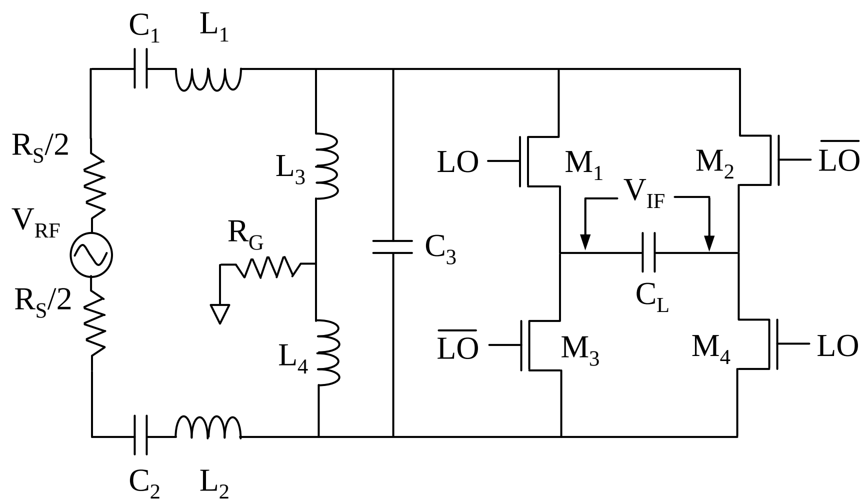
- Single-balanced:



- Double-balanced:



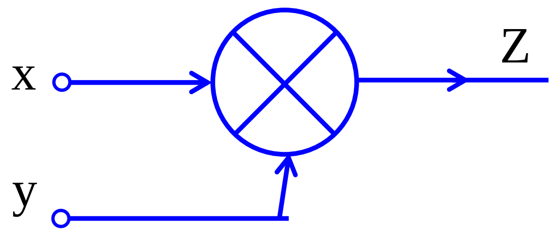
- CMOS Passive Mixer



SSB noise figure of a mixer is 3 dB higher than the DSB noise figure if the signal and image bands experience equal gains at the RF port of a mixer.

Multipliers

What is a multiplier?



$Z = Kxy$

Where

x and y are the input signals, Z is the output and K is a constant with suitable dimensions.

How do you obtain a multiplier?

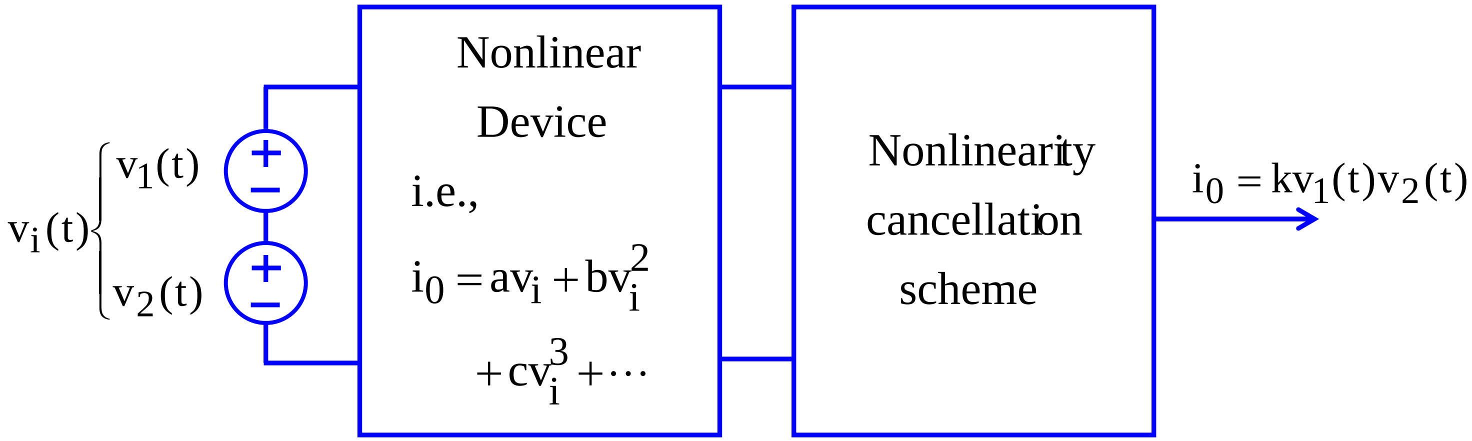


Fig. 1 Basic idea of multiplier

Transconductance-Mode Multiplier

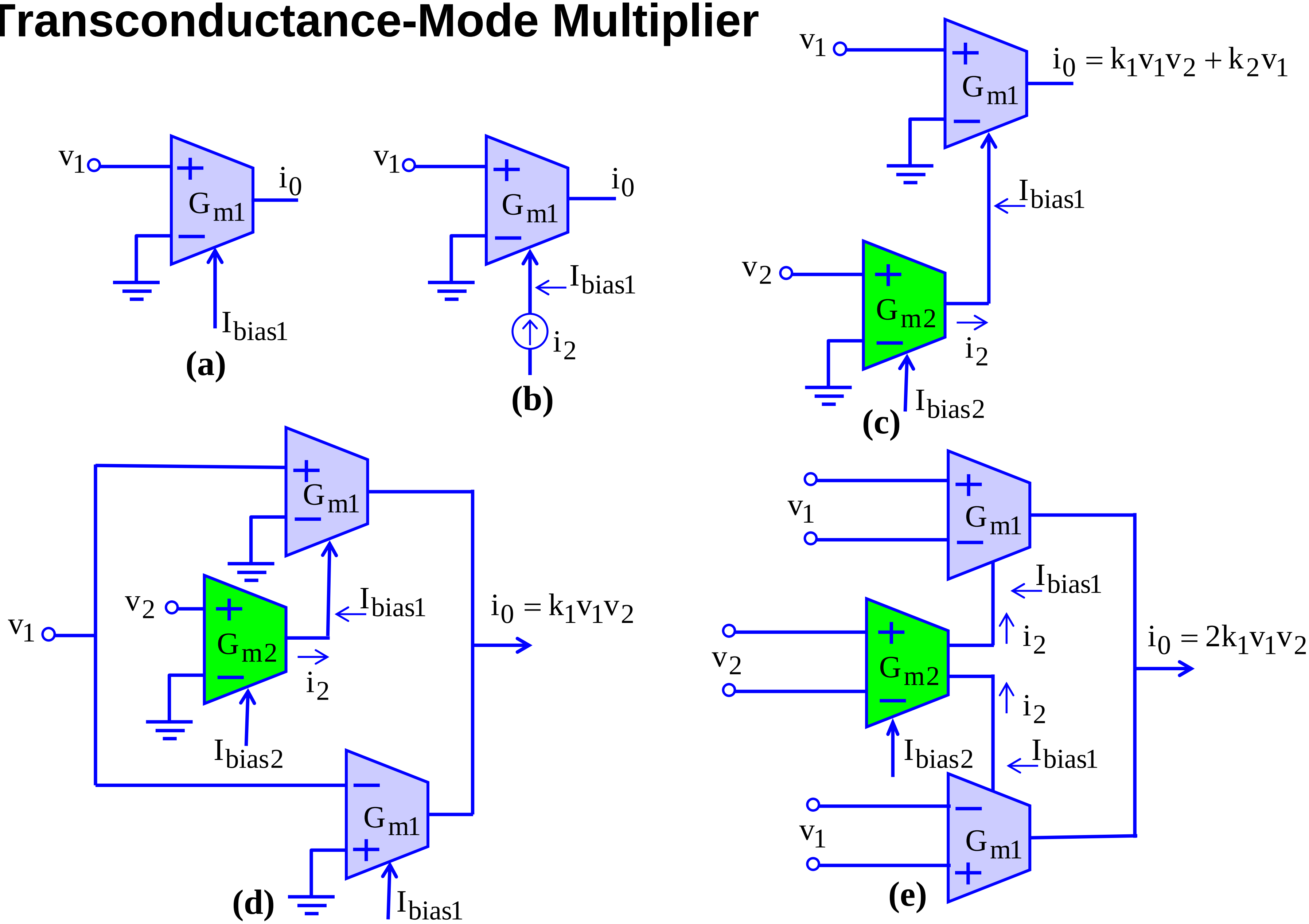


Fig. 2 Multiplication operation using programmable transconductor

How Does It Work?

$$i_0 = G_{m1} v_1 \tag{1}$$

where

$$G_{m1} = G_{m1}(I_{bias1}) \tag{2a}$$

For a bipolar transconductor, G_{m1} becomes

$$G_{m1} = \frac{I_{bias1}}{2V_t} \tag{2b}$$

$$i_0(t) = G_{m1} v_1 = \frac{I_{bias1} + G_{m2} v_2(t)}{2V_t} v_1(t) \tag{3a}$$

$$i_0(t) = \frac{G_{m2} v_1(t) v_2(t)}{2V_t} + \frac{I_{bias1}}{2V_t} v_1(t) = \frac{I_{bias2} v_1(t) v_2(t)}{2V_t 2V_t} + \frac{I_{bias1} v_1(t)}{2V_t} \tag{3b}$$

or

$$i_0(t) = k_1 v_1(t) v_2(t) + k_2 v_1(t) \tag{3c}$$

Thus, $i_0(t)$ represents the multiplication of two signals $v_1(t)$ and $v_2(t)$, and an unwanted component, $k_2 v_1(t)$. This component can be eliminated as shown in Fig. 2(d). Better cancellation is achieved when the third transconductor (G_{m2}) becomes a fully differential transconductor, and v_1 and v_2 are fully differential inputs as illustrated in Fig. 2(e).

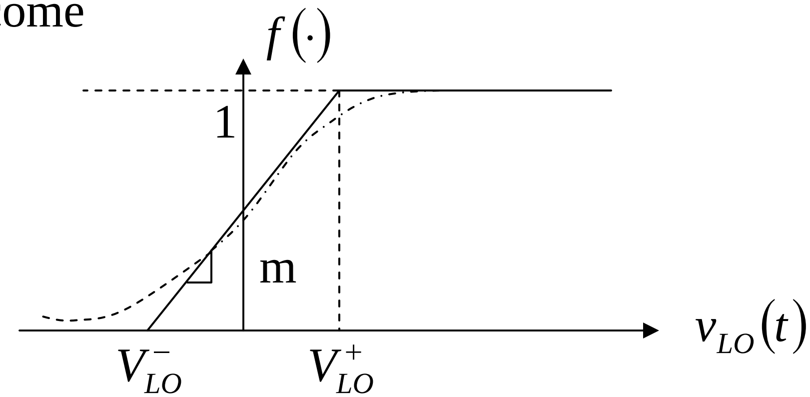
$$i_0(t) = 2k_1 v_1(t) v_2(t) \tag{4}$$

ACTIVE MIXER

$$i_o = k_1 v_1 v_2$$

For a mixer, the variables for the multiplier become

$$i_o = i_{IF}(t)$$
$$v_1 = v_{RF}(t)$$
$$k_1 = G_o$$
$$v_2 = f\left(v_{LO}(t)\right)$$



A Sigmoidal Function

Thus

$$i_{IF}(t) = v_{RF}(t) G_o f\left[v_{LO}(t)\right] \quad ; \quad v_{LO}(t) = A_{LO} \cos \omega_{LO} t$$

We have two important cases:

$$f\left(V_{LO}(t)\right) = \begin{cases} 1 \quad \text{for} \;\; v_{LO} > \mathrm{V}_{\mathrm{LO}}^+ \\ 0 \quad \text{for} \;\; v_{LO} < V_{LO}^- \\ \text{-----------------} \\ \dfrac{\mathrm{v}_{\mathrm{LO}}(t) - V_{LO}^-}{V_{LO}^+ - V_{LO}^-} \;\; \text{or} \; m v_{LO}(t) \;\; \text{for} \;\; V_{LO}^- < v_{LO}(t) < V_{LO}^+ \end{cases}$$

Assume $A_{LO} < V_{LO}^{+}, \left|V_{LO}^{-}\right|$ then

$$i_{IF}(t) = v_{RF}(t) G_o A_{LO} \cos\omega_{LO} t$$

For $v_{RF}(t) = A_{RF} \cos\omega_{RF} t$, $i_{IF}(t)$ yields

$$i_{IF}(t) = G_o A_{LO} A_{RF} \cos\omega_{RF} t \cdot \cos\omega_{LO} t$$

$$i_{IF}(t) = \frac{G_o A_{LO} A_{RF}}{2} \left\{ \cos(\omega_{RF} - \omega_{LO})t + \cos(\omega_{RF} + \omega_{LO})t \right\}$$

After filtering $(\omega_{RF} + \omega_{LO})$;

$$i_{IF} = G_C A_{RF} \cos(\omega_{RF} - \omega_{LO})t \quad ; \quad G_C = \frac{G_o A_{LO}}{2}$$

This case corresponds to a multiplier where i_{IF} (t) is a function of A_{LO}

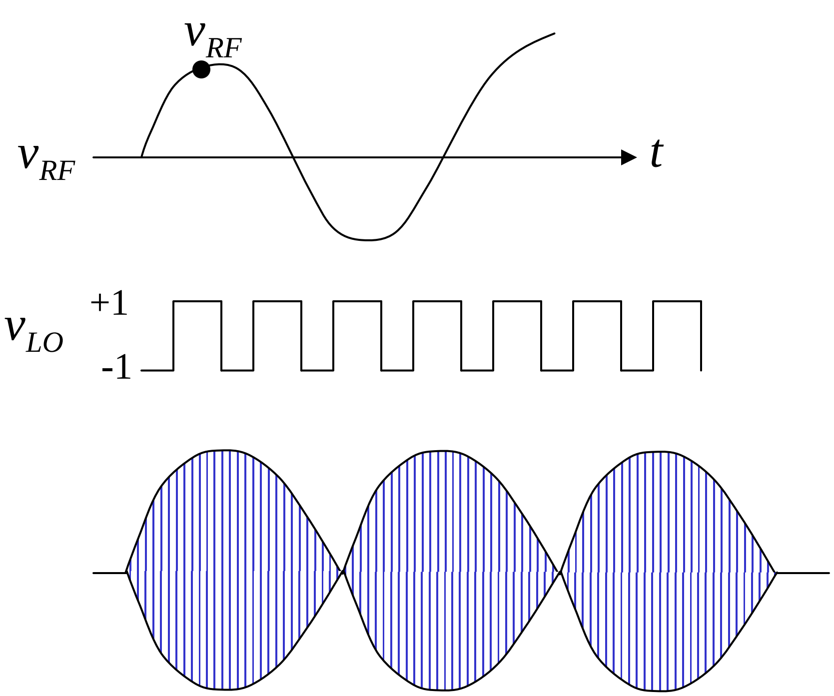
The conversion gain is given by $\dfrac{P_{out}(\omega = \omega_{IF})}{P_{in}(\omega = \omega_{RF})}$.

For this case, the maximum output noise occurs in comparison of large driving A_{LO}. That is the case to be discussed next.

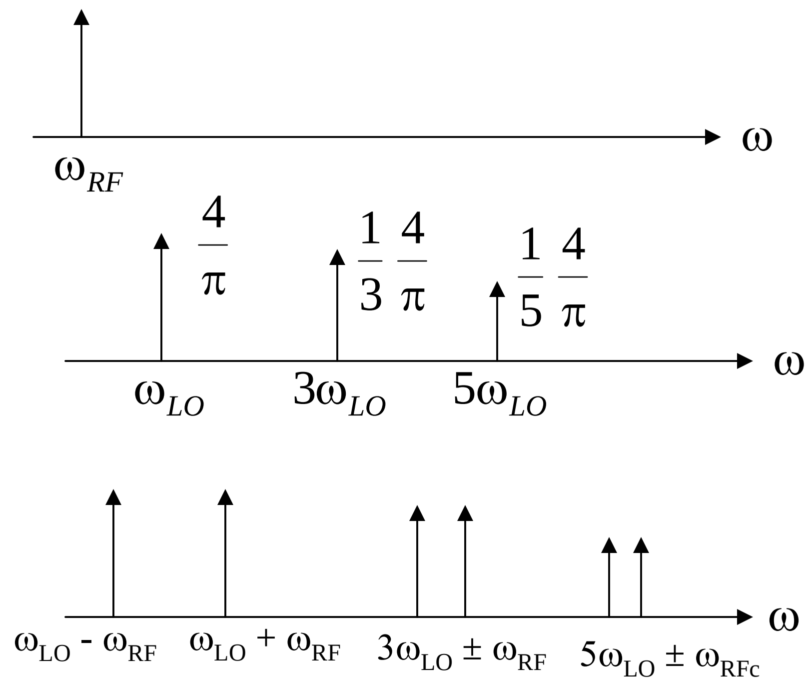
Now let us focus on the case of large A_{LO}, that is

$A_{LO} > V_{LO}^{+}, \left| -V_{LO}^{-} \right|$ and recall

$$\mathrm{f}\left(\mathrm{v}_{\mathrm{LO}}(t)\right) = \begin{cases} 1 & \text{for} \quad v_{LO}(t) > V_{LO}^{+} \\ 0(-1) & \text{for} \quad v_{LO}(t) < V_{LO}^{-} \end{cases}$$



Time-Domain



To handle mathematically the product $v_{RF}(t)$ $v_{LO}(t)$ we resort to fourier series to express $v_{LO}(t)$ and its product with $v_{RF}(t)$.

$$F\left[v_{LO}(t)\right] = \frac{4}{\pi} \sum_{n=1,3,5,\cdots}^{\infty} \frac{1}{n} \sin n\,\omega_o t$$

Then

$$i_{IF}(t) = G_o v_{RF}(t) v_{LO}(t) = G_o A_{RF} \cos \omega_{RF} t \left[\frac{4}{\pi} \sin \omega_{LO} t + \frac{4}{3\pi} \sin 3\omega_{LO} t + \right.$$
$$\left. \frac{4}{5\pi} \sin 5\omega_{LO} t + \frac{4}{7\pi} \sin 7\omega_{LO} t + \cdots \right]$$

Let us focus on the term

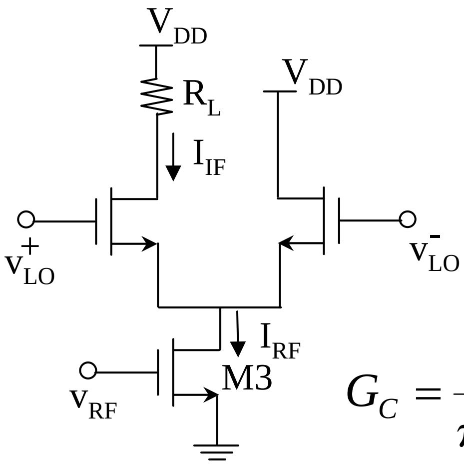
$$G_o A_{RF} \cos \omega_{RP} t \cdot \frac{4}{\pi} \sin \omega_{LO} t = \frac{G_o A_{RF}\, 4}{\pi} \frac{1}{2} \left[\sin(\omega_{LO} + \omega_{RF}) t + \sin(\omega_{RF} - \omega_{LO}) t \right]$$

Using a suitable filter one can, ideally, end with

$$i_{IF}(t) \cong \frac{2 G_o A_{RF}}{\pi} \sin(\omega_{RF} - \omega_{LO}) t = \frac{2 G_o A_{RF}}{\pi} \sin \omega_{IF} t$$

$$G_C = \frac{2 G_o}{\pi}$$

MIXER IMPLEMENTATIONS



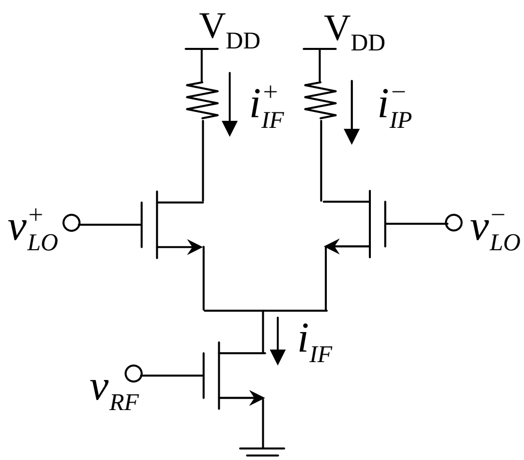
$$G_C = \frac{1}{\pi\left(g_{m3} R_L\right)^{-1}}$$

Unbalanced Mixer
(see Figs 2(b) & (c))

$$v_{IF}(t) = i_{IF}(t) R_L$$

$$v_{IF}(t) = A_{RF} \cos \omega_{RF} t \cdot G_o \left[\frac{1}{2} + \sum_{n=1}^{\infty} \frac{\sin \frac{n\pi}{2}}{\frac{n\pi}{2}} \cos n\omega_o t \right]$$

• Undesirable $v_{RF}(t)$ feedthrough.

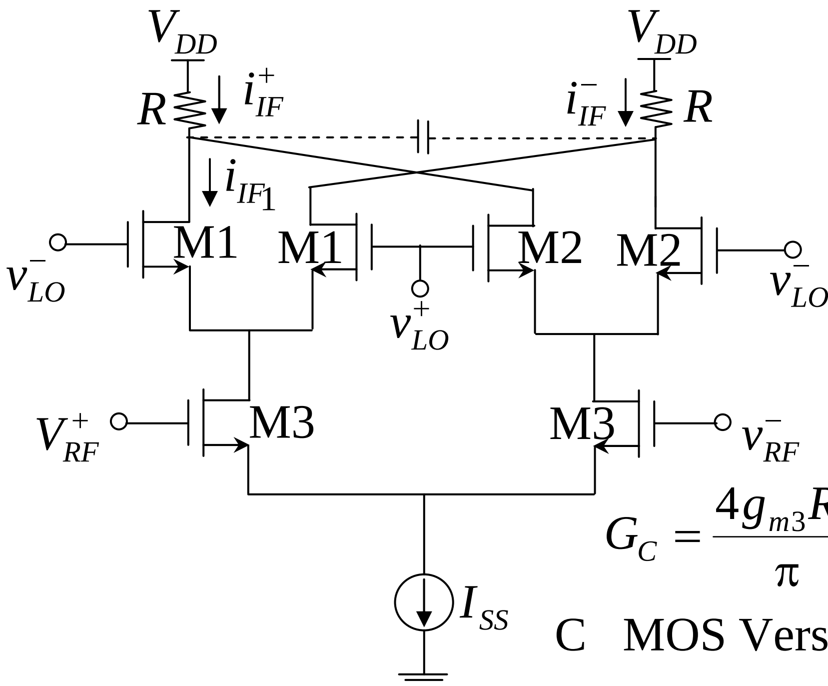


Single Balanced Mixer
(see Fig 2(d))

$$v_{IF}(t) = A_{RF} \cos \omega_{RF} t \cdot 2G_o \sum_{n=1}^{\infty} \frac{\sin \frac{n\pi}{2}}{\frac{n\pi}{2}} \cos n\omega_o t$$

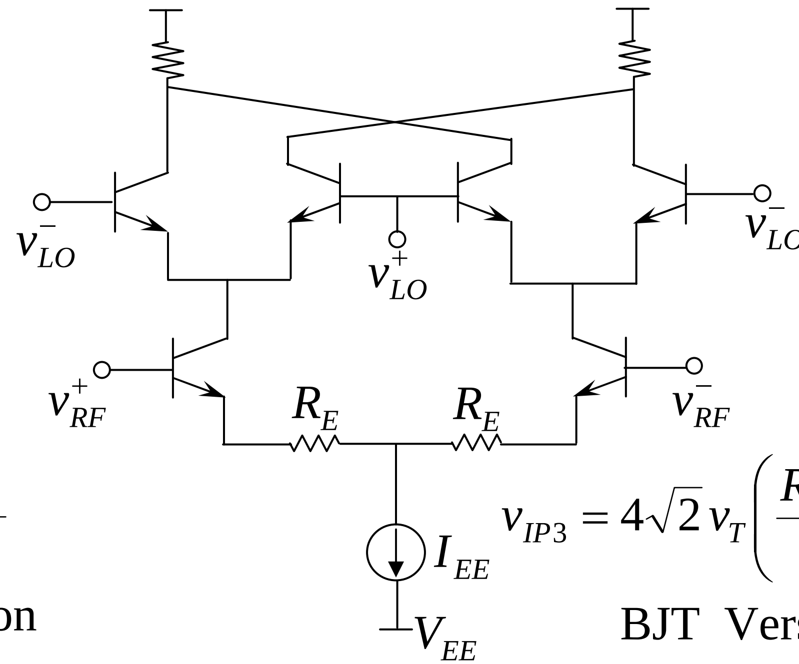
• No even harmonics

• Undesirable $v_{LO}(t)$ feedthourgh



$$G_C = \frac{4 g_{m3} R_L}{\pi}$$

C MOS Version

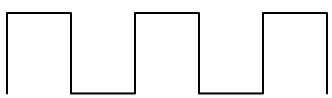


$$v_{IP3} = 4\sqrt{2}\, v_T \left(\frac{R_E + r_e}{r_e} \right)^{3/2}$$

BJT Version

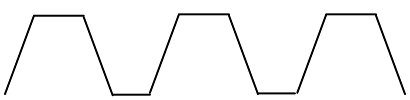
Double Balanced Mixer

The conversion gain with a differential load C_L is



Ideal Square
Wave Mixing

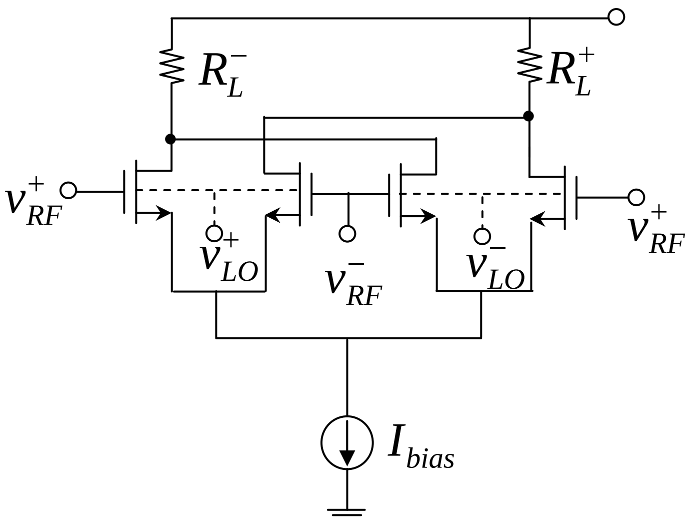
$$G_V = \frac{2}{\pi} \frac{g_m R}{1 + s2RC_L}$$



Non-Ideal
Mixing Functions

$$G_V = \frac{2}{\pi} \frac{g_m R}{1 + s2RC_L} \frac{sin(tr/T)}{(tr/T)}$$

Bulk-Driven Mixer

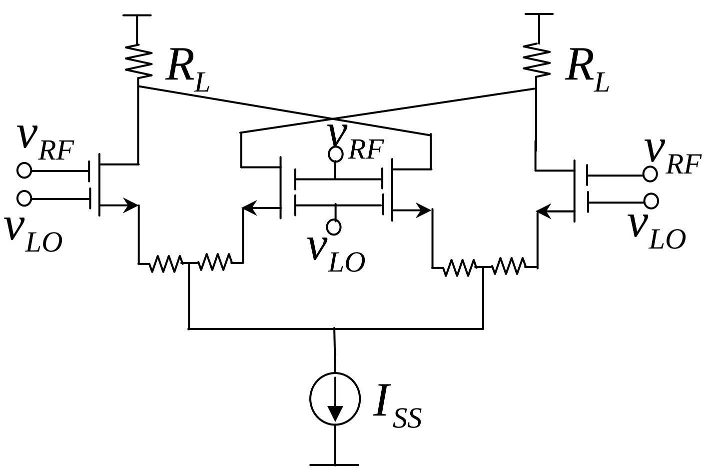


Pluses

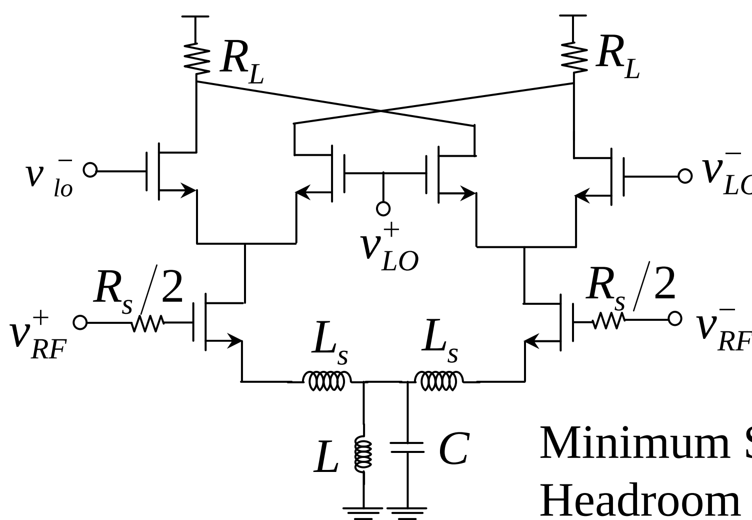
- Low Power Consumption
- Low Power Supply
- Good Conversion Gain
- NF Reasonable
- Acceptable IIP_3 (poor)

Minus

- High A_{LO} (power)
- Low Gain Compression



A Possible Floating Gate Mixer



Minimum Supply-Headroom DB Mixer

Distortion, Low Frequency Analysis of MOS Gilbert Mixer

It can be proved that one can express

$$v_{RF}(t) = \sqrt{\frac{I_{SS_n}}{2}}\left[\frac{2}{I_{SS_n}}\left(a_1 v_{RF} + a_2 v_{RF}^2 + a_3 v_{RF}^3 + \cdots\right) + \frac{1}{8}\left(\frac{2}{I_{SS_n}}\right)^3 \left(a_1 v_{RF} + a_2 v_{RF}^2 + a_3 v_{RF}^3 + \cdots\right)\right]$$

$$I_{SS_n} = \frac{2I_{SS}}{k} = \frac{2I_{SS}}{\mu_o cox(W/L)}$$

Solving for these coefficients, it can be obtained the following:

$$a_1^2 = \frac{I_{SS_n}}{2} \quad ; \quad a_2 = 0 \quad ; \quad a_3 = -\frac{1}{8}\left(\frac{2}{I_{SS_n}}\right)^2 a_1^3$$

Thus

$$HD_3 = \frac{1}{4}\frac{a_3}{a_1}A_{RF}^2 = \frac{1}{32}\frac{\mu_o cox(W/L)_3}{I_{SS}}A_{RF}^2$$

$$IM3 = 3HD_3$$

Assuming we are interested that the amplitude of the inteference can be denoted as A_{INT}. Then

$$IM3 = \frac{3A_{INT}^2}{32\left(V_{GS_3} - V_T\right)^2}$$

Furthermore, it can be shown that

$$IM3 = \frac{A_{INT}^2}{A_{IP3}^2}$$

then

$$A_{IP3}^2 = \frac{A_{INT}^2}{IM3} = \frac{32\left(V_{GS3} - V_T\right)^2 A_{int\,erference}^2}{3A_{int\,erference}^2} = \frac{32}{3}\left(V_{GS3} - V_t\right)^2$$

Example

$$V_{GS3} - V_T = 0.4V \text{ , } A_{RF} = A_{INT} = 0.212V_p \quad or \quad -3.472\,dBm$$

Assuming the $v_{LO}(t)$ is not large enough to force switching.

Then

$$IM_3 = \frac{3A_{INT}^2}{32\left(V_{GS3} - V_T\right)^2} \Rightarrow -31.6dB$$

$$A_{IP3} = 1.3064V_p = 2 \times 1.3064 \times 0.3536\,V_{rms}$$

$$IIP_3\Big|_{dBm} = P_i\Big|_{dBm} - \frac{IM_3 1_{dB}}{2} = -3.472 + 15.8 = 12.33dBm$$

Mixer Noise

- v_{LO} not switching case
  - LO switches behave like a regular Differential pair
  - Maximum output noise contribution

- V_{LO} switching case
  - LO switches behave like a Cascode transistor
  - Minimal output noise contribution.

Noise factor for a mixer is given by

$$F = \frac{N_{o\,tot}(\omega_{IF})}{N_{o(source)}(\omega_{IF})}$$

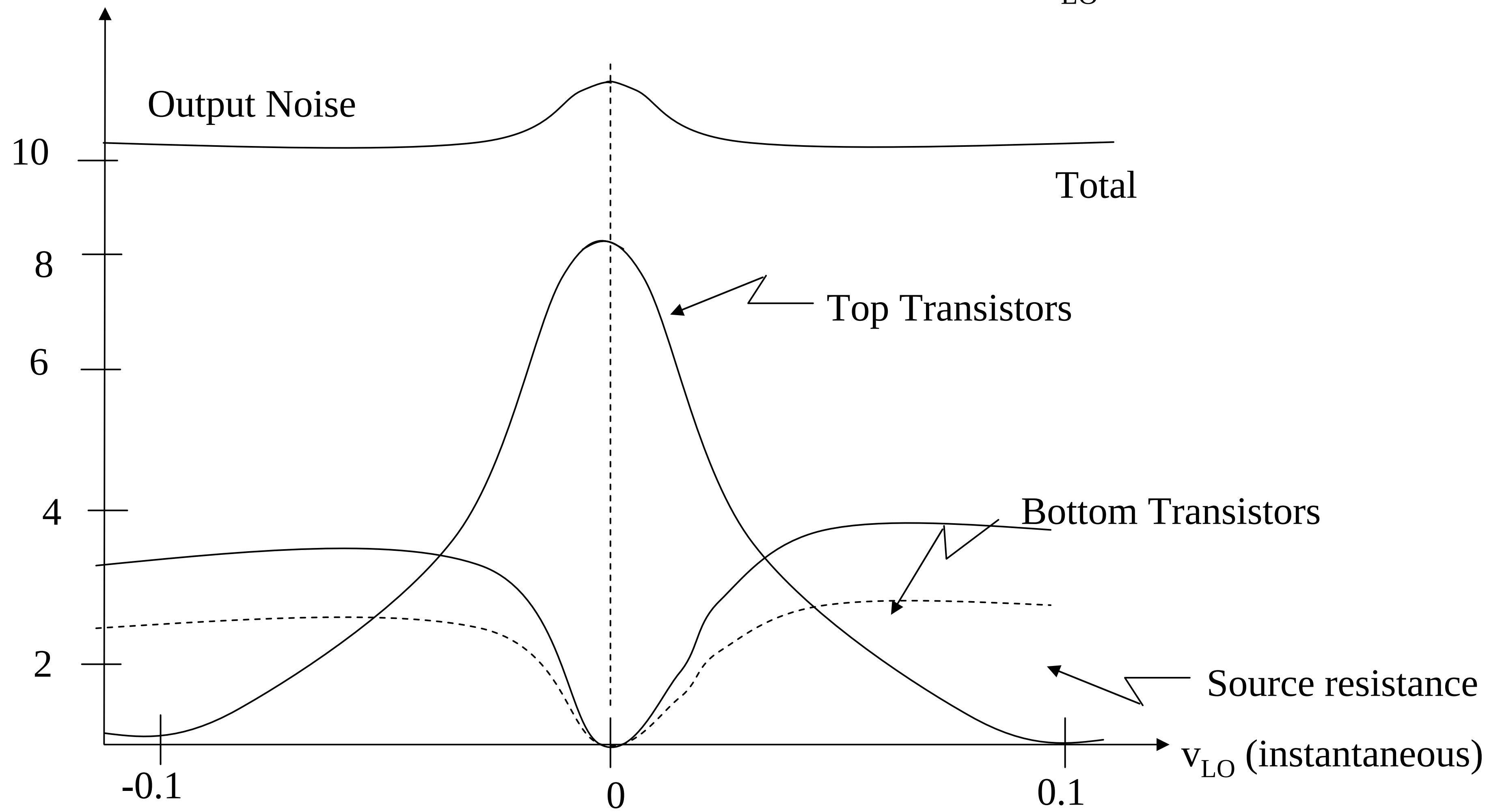
where

$$N_{o(source)}DSB = N_{o(source)}SSB + 3dB$$

and

$$NF_{DSB} \triangleq NF_{SSB} - 3dB$$

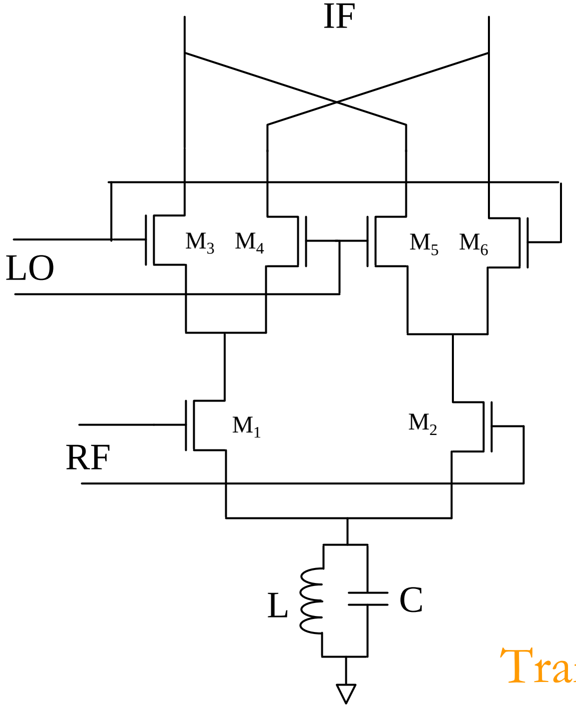
TYPICAL BJT MIXER NOISE AT VARIOUS v_{LO} LEVELS



s/n around $v_{LO} = 0$ is very poor (gain is also low)

A Popular mixer topology: IC implementation

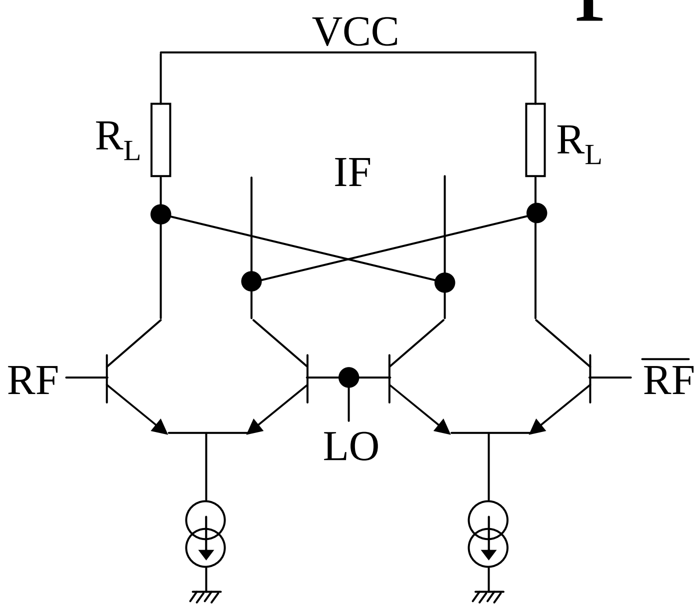
Gilbert-Cell based mixer



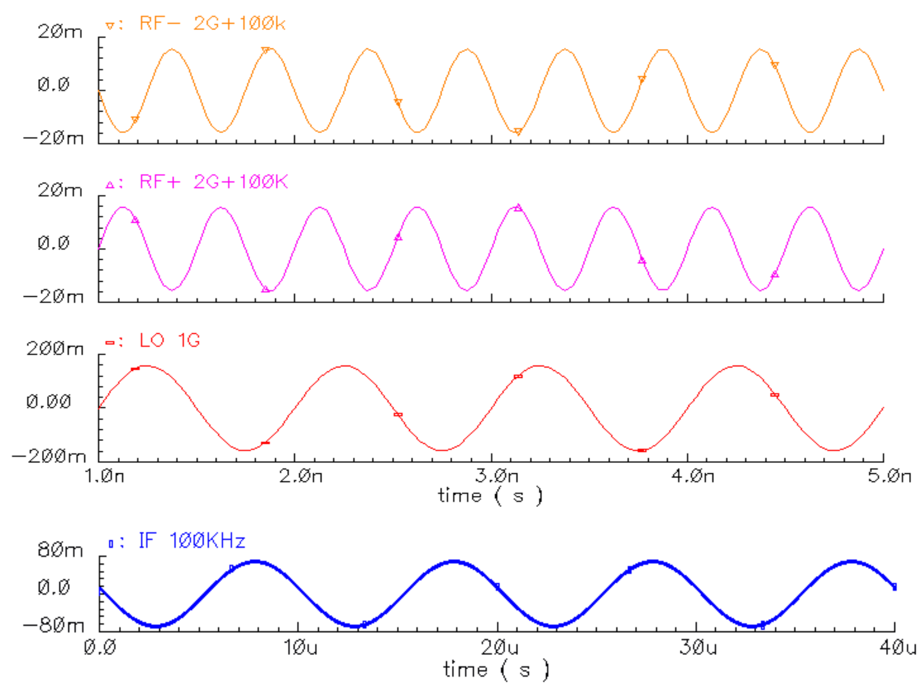
- Single or Double-balanced mixer
- Good LO-IF isolation (40dB~60dB)
- M1, M2 work as V-I converter
- M3~M6 work as current commuting switches
- LC tank: zero-headroom
- Low IF noise figure problem

Transconductance conversion gain: $$G_c = \frac{2}{\pi} g_m$$

Harmonic mixing: Bipolar implementation



Sub-harmonic mixer



- Two emitter-coupled BJT pairs work as two limiters.
- The small RF signal will modulate the zero crossing point of the relatively large LO signal.

Design Considerations for Gilbert-cell Mixer

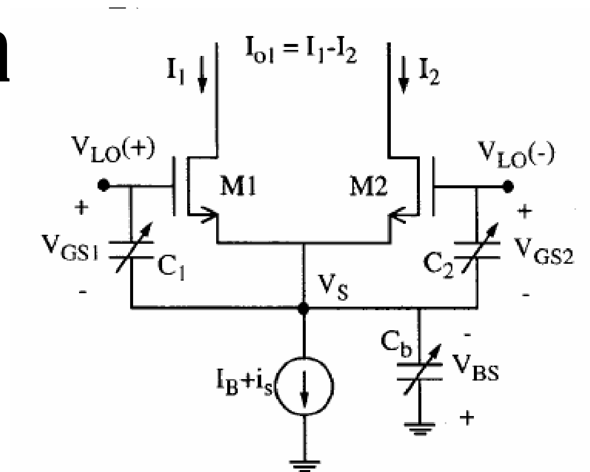
Noise contributions

- Load noise
- V-I converter noise
  - whit noise: up/down conversion
  - flicker noise: up conversion
- Switch pair noise
  - High frequency noise of switches or coming together with LO signal
  - flicker noise modulates switching point
  - Charging and discharging the parasitic capacitance at the common-source of differential pair makes flicker noise appearing at output

H. Darabi, A.A. Abidi, *"**Noise in RF-CMOS mixers: a simple physical model**,"* *IEEE JSSC Vol. 35 Issue: 1 , Jan. 2000 pp. 15 -25*

Design Considerations for Gilbert-cell Mixer (con

Linearity



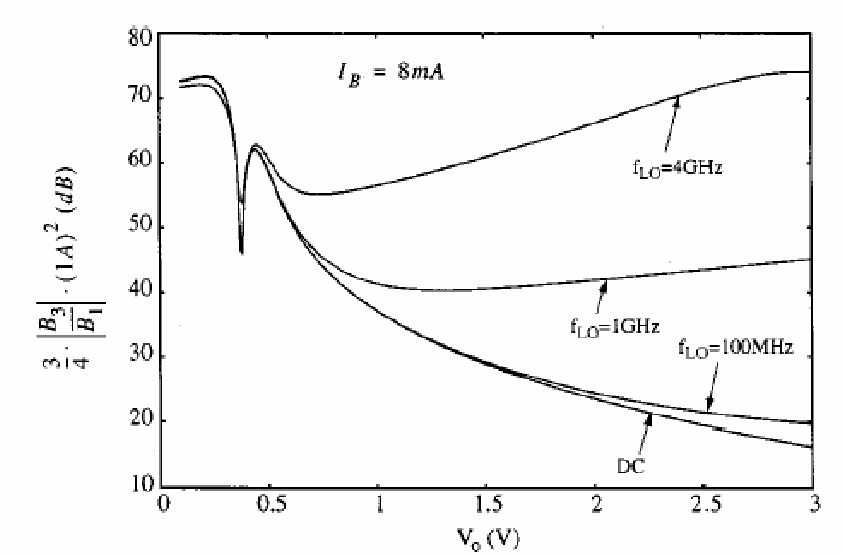
- V-I converter

The arguments of LNA linearity apply here

- Switch pair

Low frequency: higher LO swing better linearity

High frequency: Optimal LO swing



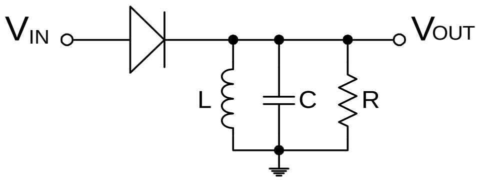
M. T. Terrovitis, R. G. Meyer, “Intermodulation distortion in current-commutating CMOS mixers,” IEEE JSSC, Vol. 35 Issue 10 , Oct. 2000 pp. 1461 -1473

Mixer topologies for Narrowband Applications

- Discrete implementations:
  - Single-diode and diode-ring mixers
- IC implementations:
  - MOSFET passive mixer
  - Gilbert-cell based mixer
  - Harmonic mixer

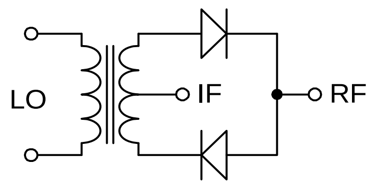
Diode Mixers

Single-diode:



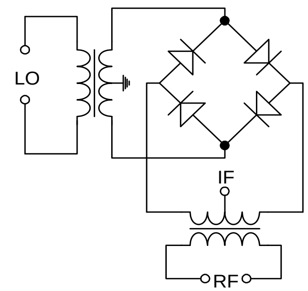
The single-diode mixer is the simplest and oldest passive mixer. The output RLC tank is tuned to the desired IF, and input is the sum of RF, LO and DC bias.This mixer can not provide any isolation and conversion gain. However, at very high frequency (millimeter-wave band) this kind of mixer is extremely useful.

Single-balanced:



LO is large enough to make the diodes work as switches, regardless of the level of RF signal. When the diodes are on, RF and IF are connected together, so the RF-IF isolation is poor. But the RF signal is common-mode for the transformer, so the RF-LO isolation is excellent.

Double-balanced:

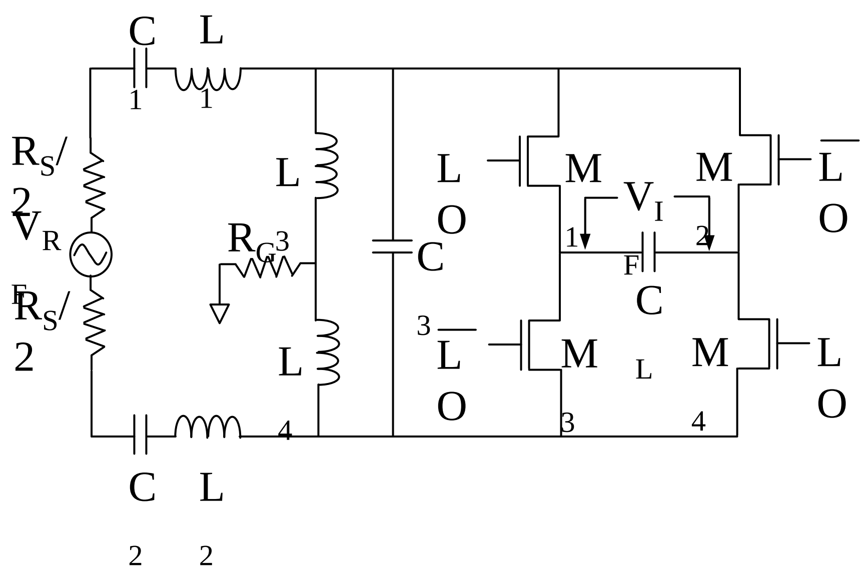


Due to the symmetry of the circuit, isolations between each pair of ports are excellent, mainly limited by the device matching. The diode mixer is pretty much linear and the upper limit of the dynamic range is constrained by diode break-down.

Typically, double-balanced mixers can achieve conversion loss of around 6dB, isolation of at least 30dB.

CMOS Passive Mixer

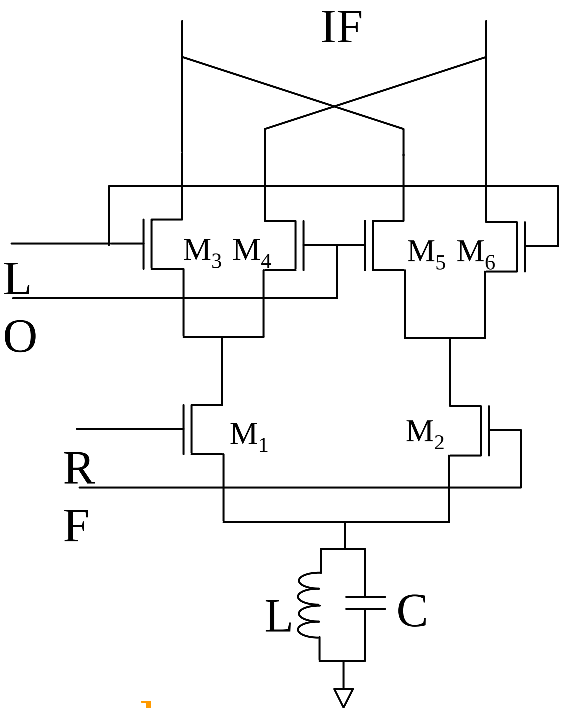
Input LC network provide matching and filtering. R1 sets the input common-mode level.



Due to the matching network, voltage conversion gain can be greater than 1. Noise figure and IIP3 are strong functions of LO drive level.

MOSFET M1~M4 are working as switches and are driven by LO in anti-phase. Only one diagonal pair of transistors is conducting at any given time. When M1 and M4 are on, V_{IF} equals V_{RF}, and when M2 and M3 are on, V_{IF} equals $-V_{RF}$. So it is equivalent to observe that the mixer multiplies the RF signal by square wave whose amplitude is alternating between +1 and –1 and whose frequency is that of LO.

Gilbert-Cell Based Mixer



This is a double-balanced mixer. Good LO-IF isolation (40dB~60dB) can be achieved due to the symmetry. M1 and M2 work as V-I converter and M3~M6 are driven by large enough LO, working as current commuting switches. LC tank is to create zero-headroom AC current source. If the power supply voltage is not a limitation factor, the LC tank can be replace with a transistor working as current source.

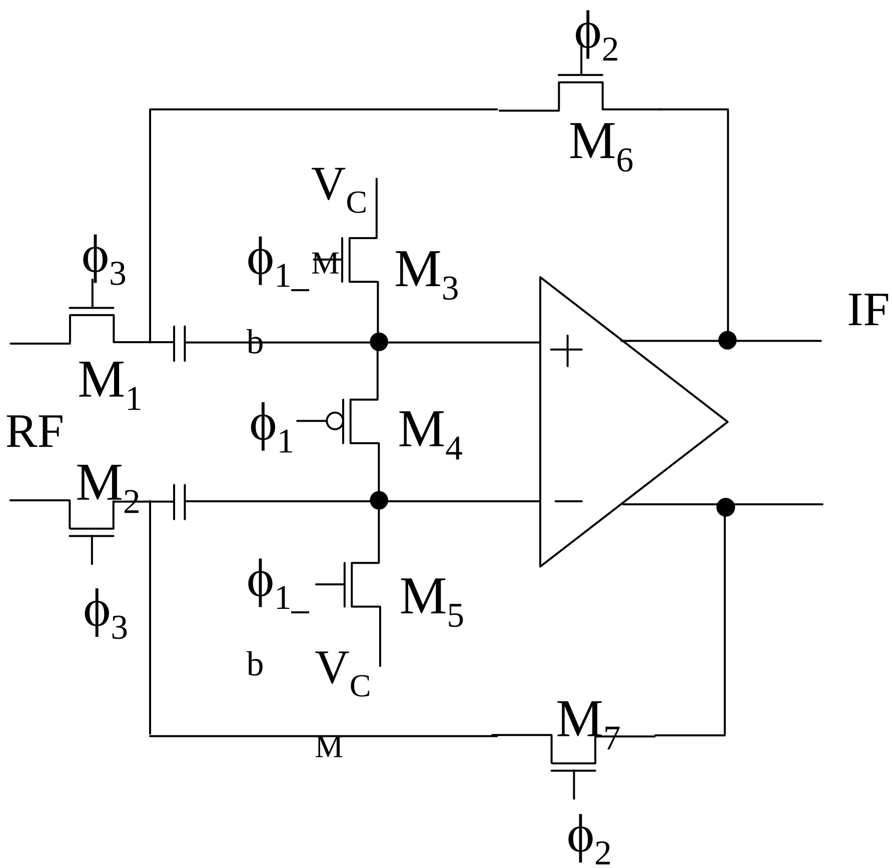
The linearity of the mixer is limited by the linearity of the V-I converter. For low IF, the noise figure is limited by the flicker noise of the current switches and for higher IF, the noise figure is limited by the thermal noise of the circuit.

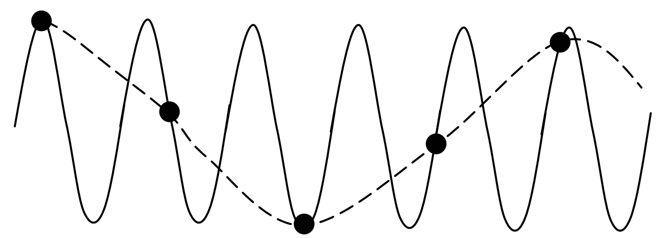
Transconductance conversion gain:

$$G_c = \frac{2}{\pi} g_m$$

Additional linearization techniques can be applied to improve the linearity of the mixer.

Sub-sampling Mixer





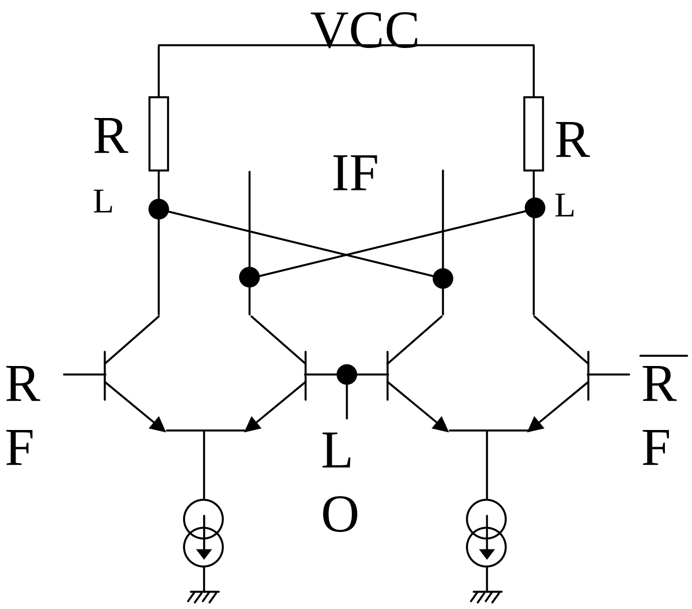
Properly designed track-and-hold circuit works as sub-sampling mixer

- The sampler must have good time resolution. So the clock's absolute time jitter must be a tiny fraction of the carrier period.
- Noise folding make the mixer present large noise figure.
- The linearity of the mixer is high

Harmonic Mixer (I)

- Harmonic mixer has low self-mixing DC offset. It is very attractive for direct conversion application.
- The RF single will mix with the second harmonic of the LO. So the LO can run at half rate, which make the VCO design easier.
- Because of the harmonic mixing, conversion gain is usually small (several

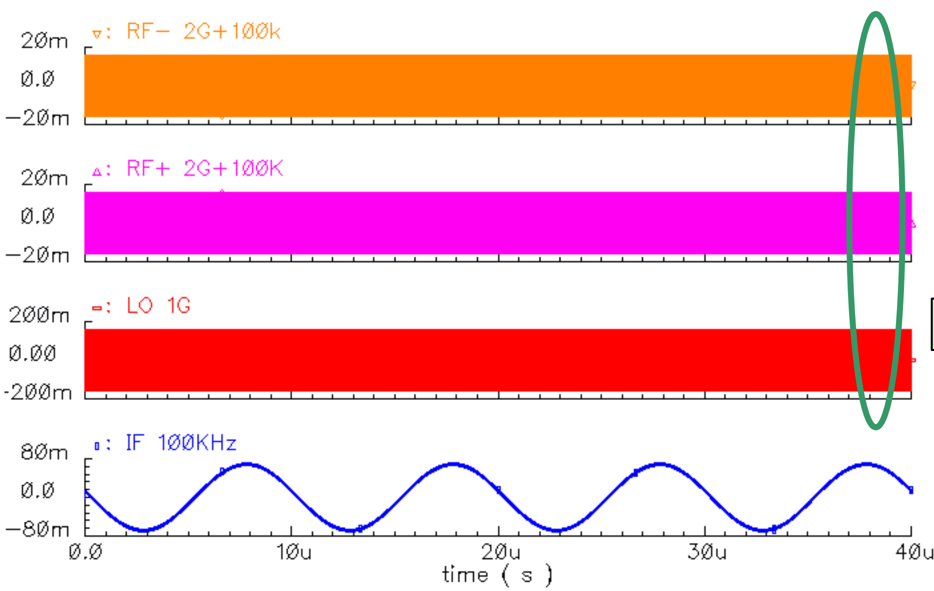
Harmonic Mixer (II)

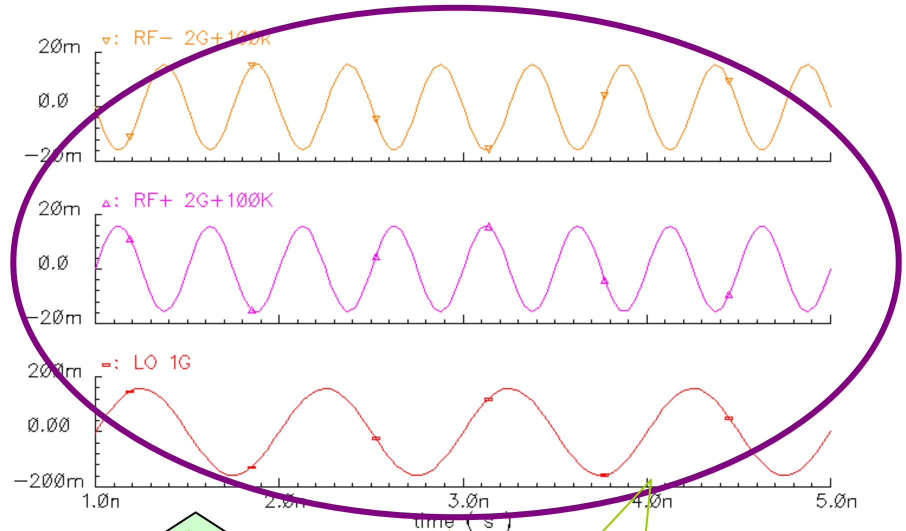


- Two emitter-coupled BJT pairs work as two limiters. The odd symmetry of their transfer function suppress even order distortion including LO self-mixing.
- The small RF signal will modulate the zero crossing point of the relatively large LO signal. The output of the mixer is rectangular wave in the pulse width modulation fashion, a low pass filter will demodulate the

Harmonic Mixer (III)

Simulated waveforms of the harmonic mixer:





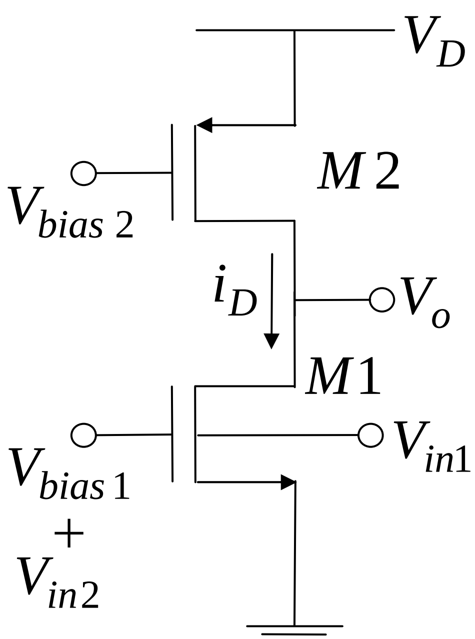
LO @ 1GHz

RF @ 2GHz+100kHz

IF @ 100kHz

A Bulk-Driven, Gate-Driven, Multiplier

Calculate the input impedance looking at the bulk, and the output voltage V_o expression of the simple bulk-driven amplifier. Note that the circuit has two small signal input signals. V_{in1} is injected to the bulk and V_{in2} to the gate. Also determine an expression for i_D .



For low frequency

$$\left. Z_{in} \right|_{@\,bulk} = \infty$$

For high frequency

$$\left. Z_{in} \right|_{@\,bulk} = \frac{1}{g_{mb_1} + g_{01} + g_{02}}$$

The drain current containing the DC and small signal is

$$i_d = \frac{1}{2} K_p \frac{w}{l} \left[v_{GS} - \left(V_{T_D} + \gamma^1 \left[\sqrt{2\phi_f} \right] \right) \right]^2$$

$$i_D = \frac{1}{2} K_p \frac{w}{L} \left[v_{in_2} + V_{bias1} - V_{T_D} - \gamma \sqrt{2\phi_F - v_{in_1}} + \sqrt{2\phi_F}\,\gamma^1 \right]^2$$

We will use the following approximation for the square root containing V_{in1} by means of a Taylor Series:

$$\sqrt{a - x} \cong \sqrt{a} - \frac{x}{2\sqrt{a}} + \cdots \qquad ; \quad a = 2\phi_F \quad \& \quad x = v_{in1}$$

$$i_D = \frac{1}{2} Kp \frac{w}{L}\left[\left(v_{bias_1} - V_{To} + \gamma\sqrt{2\phi_F}\right) + v_{in2} - \gamma\sqrt{2\phi_F} + \frac{\gamma\, v_{in1}}{2\sqrt{2\phi_f}}\right]^2$$

$$i_D = \frac{1}{2} Kp \frac{w}{L}\left[\left(v_{bias_1} - V_{To}\right) + \left(v_{in2} + \frac{\gamma\, v_{in1}}{2\sqrt{2\phi_F}}\right)\right]^2 = \frac{1}{2} Kp \frac{w}{L}\left[\left(v_{bias_1} - V_{To}\right)^2 + \cdots\right]$$

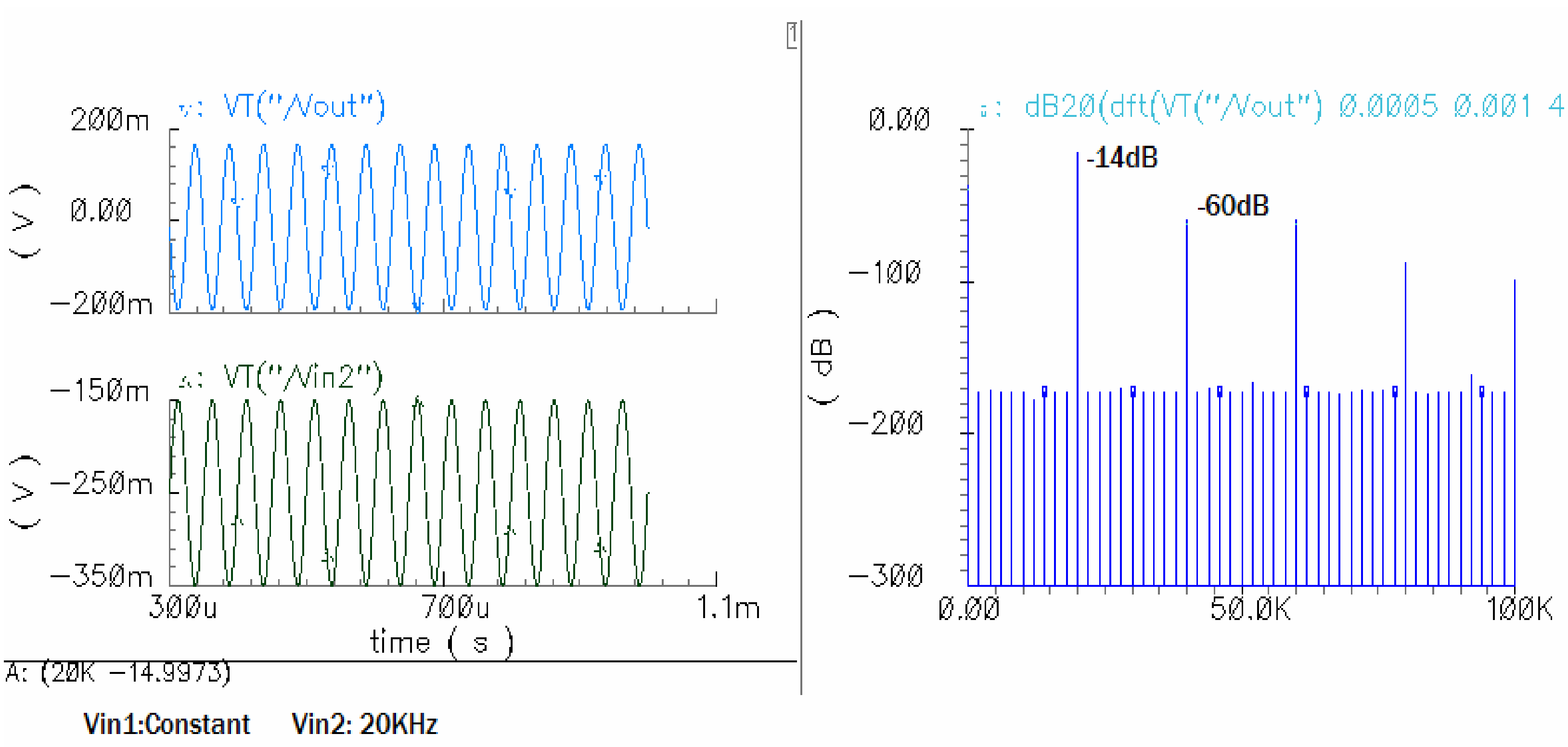
$$i_D = I_D + i_d$$

$$i_D = \frac{1}{2} Kp \frac{w}{L}\left[v_{in_2}^2 + \frac{\gamma\, v_{in1} v_{in2}}{\sqrt{2\phi_F}} + \frac{\gamma^2}{\gamma\,\phi_F} v_{in_1}^2 + \left(v_{bias_1} - V_{To}\right) v_{in_2} + 2\left(v_{bias_1} - V_{To}\right)\frac{\gamma\, v_{in1}}{2\sqrt{2\phi_F}}\right]$$

$$i_d = \frac{1}{2} Kp \frac{w}{L}\left[\frac{\left(V_{bias_1} - V_{To}\right)\gamma}{\sqrt{2\phi_F}} v_{in_1} + 2\left(V_{bias_1} - V_{To}\right) v_{in_2} + \frac{\gamma\, v_{in_1} v_{in_2}}{\sqrt{2\phi_F}} + \frac{\gamma^2}{\gamma\,\phi_F} v_{in_1}^2 + v_{in_2}^2\right]$$

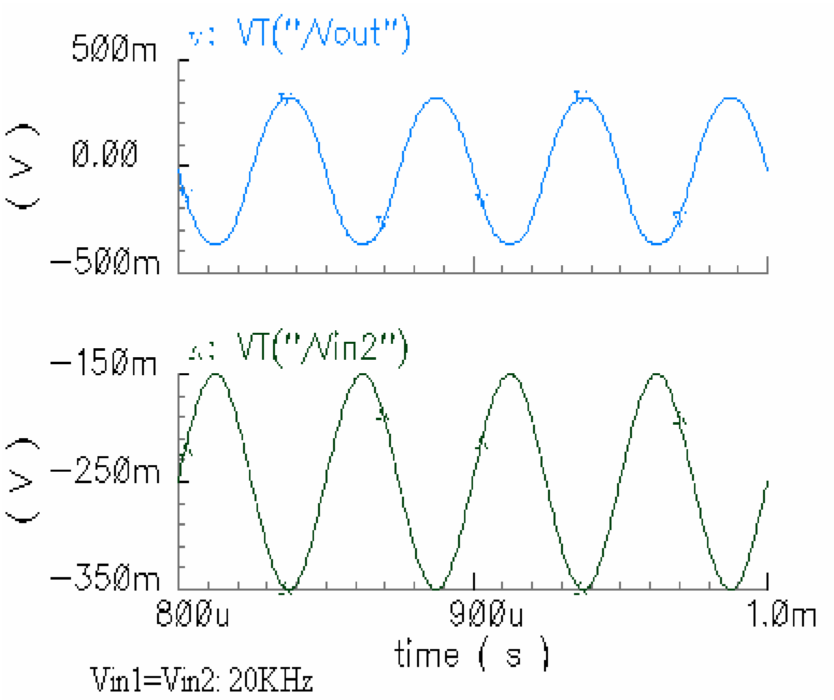
$$i_d = a_1 v_{in1} + a_2 v_{in2} + a_M v_{in1} v_{in2} + a_3 v_{in1}^2 + a_4 v_{in2}^2$$

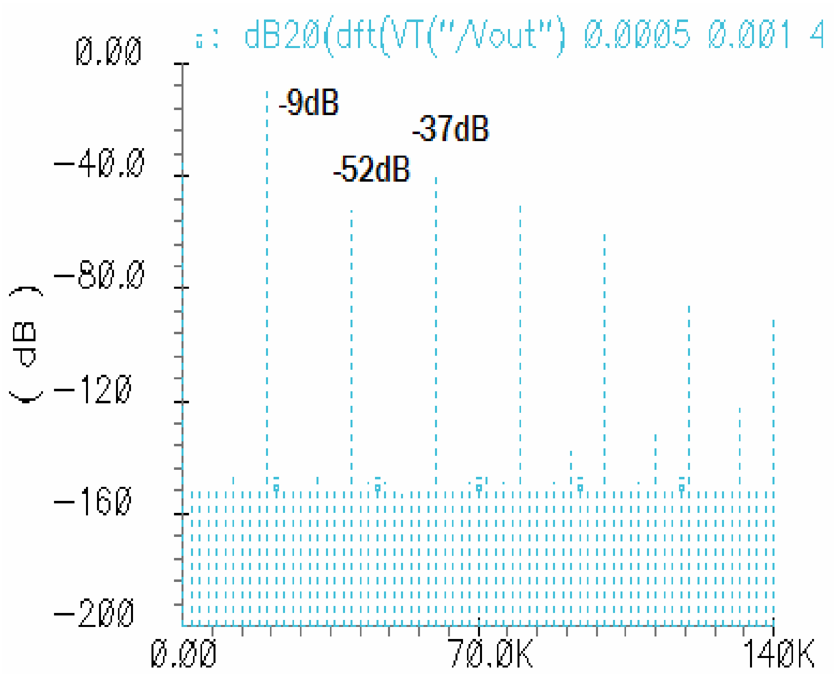
Simulation Results



Vin1: Constant Input, Vin2: 20KHz Sine Input

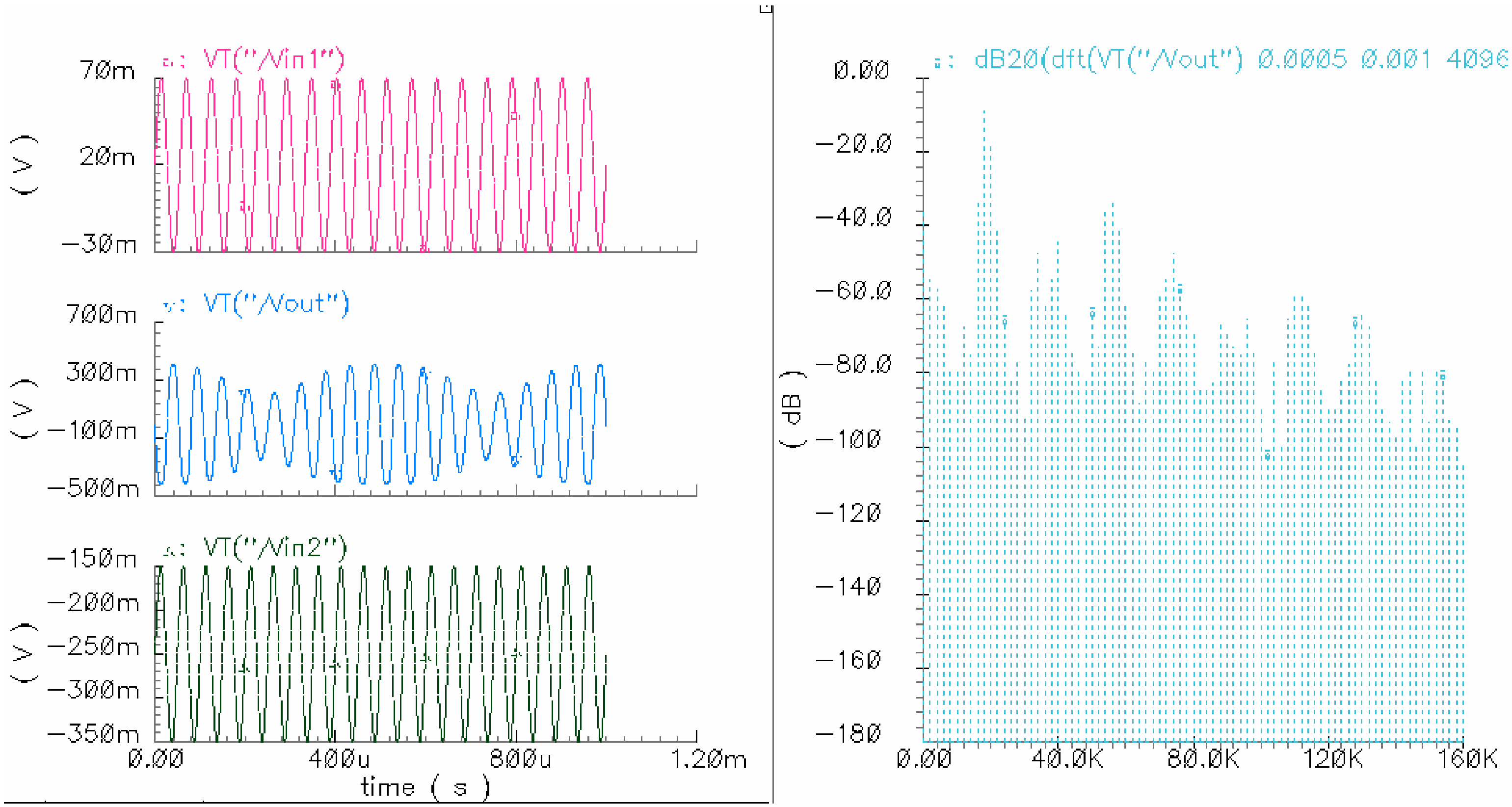
Simulation Results





Vin1, Vin2 are sine signal with same frequency

Simulation Bulk-Driven Multiplier Results

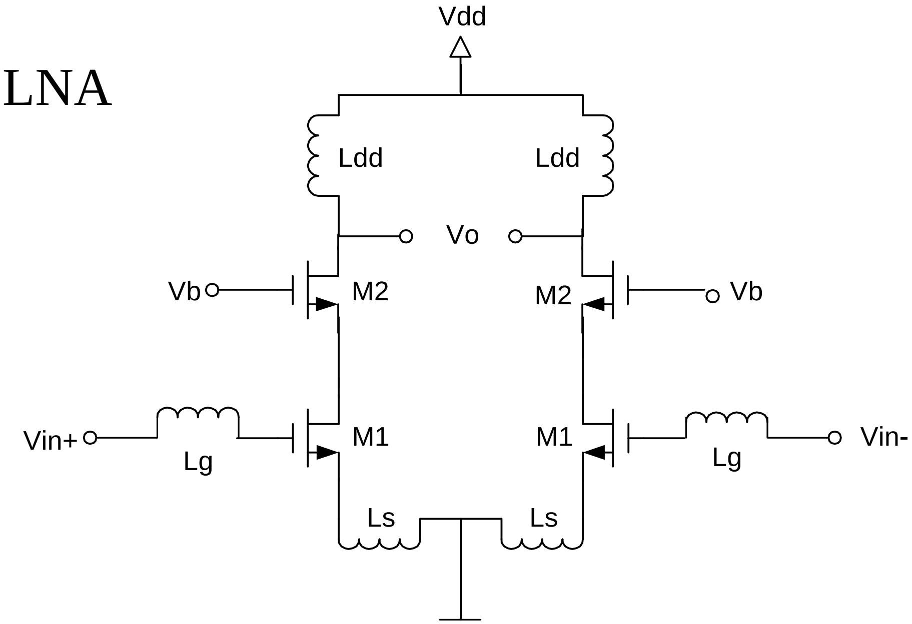


Vin1:18KHz Vin2: 20KHz

Vin1: 18KHz, Vin2: 20KHz sine wave, Vout: Modulated Output

RF Front-end for Bluetooth Low IF Receiver

LNA



Voltage Gain: 18dB

Noise Figure:2.6dB

IIP3: 0dBm

Current: 4.4mA

Supply: 3V

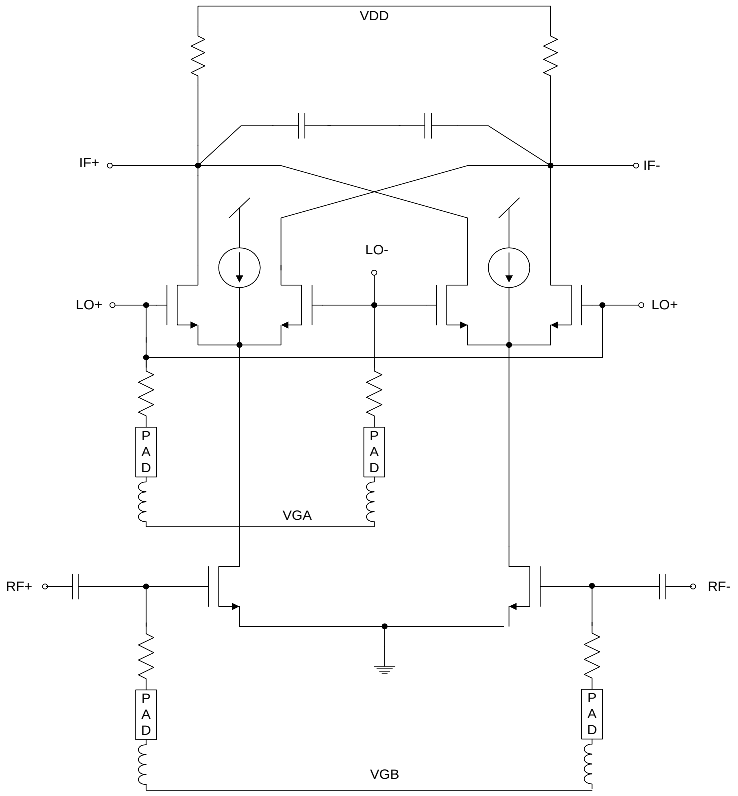
Gate NQS resistance Rgs=1/5gm

Gat induced noise modeled by gate NQS resistance $\overline{V}_g^{\,2} = 4kT\delta R_{gs}$

RF Front-end for Bluetooth Low IF Receiver

Mixer

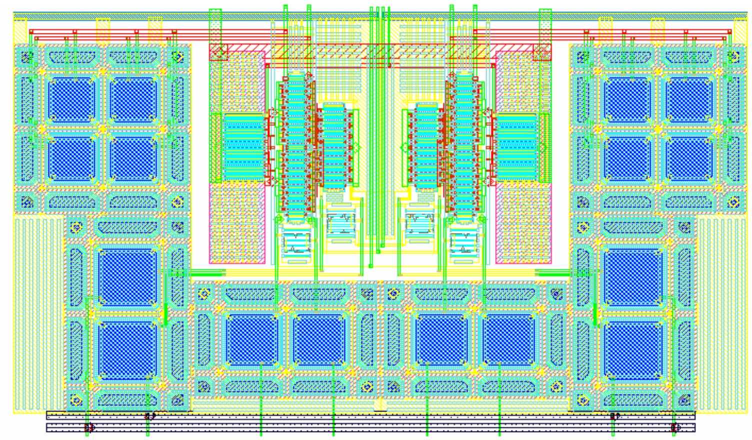
- Double balanced mixer
- Current injection to alleviate the trade off between the linearity and power supply voltage
- Voltage conversion gain: 26dB
- Noise figure: 12.4dB
- IIP3: -3dBm



RF Front-end for Bluetooth Low IF Receiver

Mixer layout considerations

- The length of the poly gate should be kept short enough to reduce the effect of gate resistance
- For the layout of poly-poly capacitor, if the bottom plate is floating, the parasitic capacitance from it to substrate should be considered. It is about ¼ of the nominal capacitance.

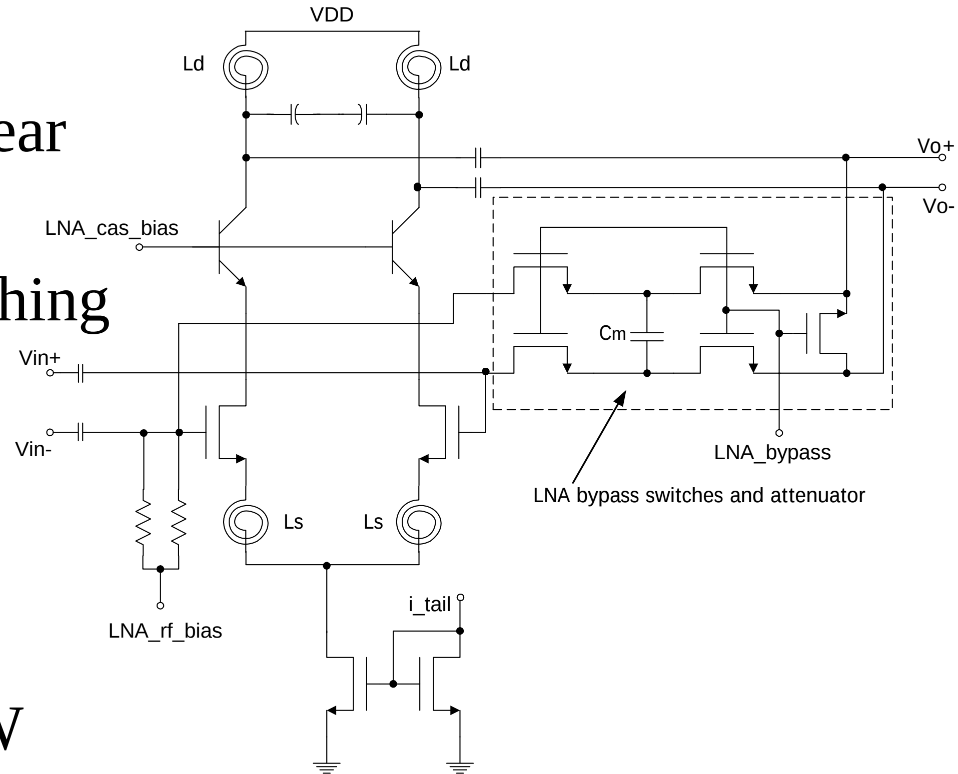


- De-coupling capacitor may be needed to prevent the circuits from oscillation
- Metal should be wide enough to carry large current. The current density allowed through metal is about 1mA/1μm
- Guarding rings are required around the circuits to provide isolation from other blocks

Front-end for BT/WiFi Receiver

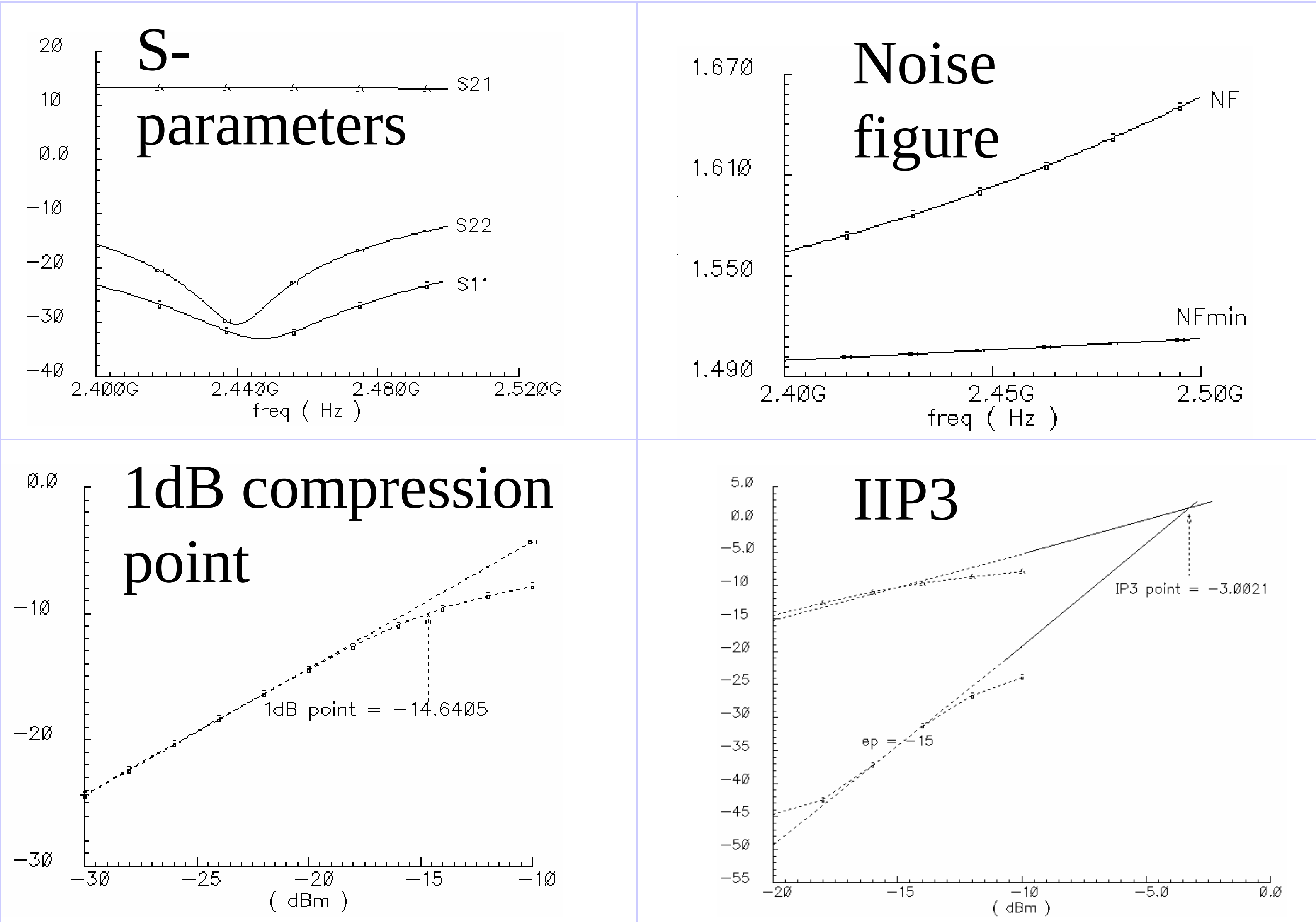
LNA

- Differential structure
- MOS transistor is more linear
- Inductor degeneration
- Cascoded BJT: better matching
- On-chip input matching
- Noise figure: 1.6dB
- Power/Voltage gain: 13dB
- Power consumption: 16mW
- NMOS attenuator for low gain(3dB)



Front-end for BT/WiFi Receiver

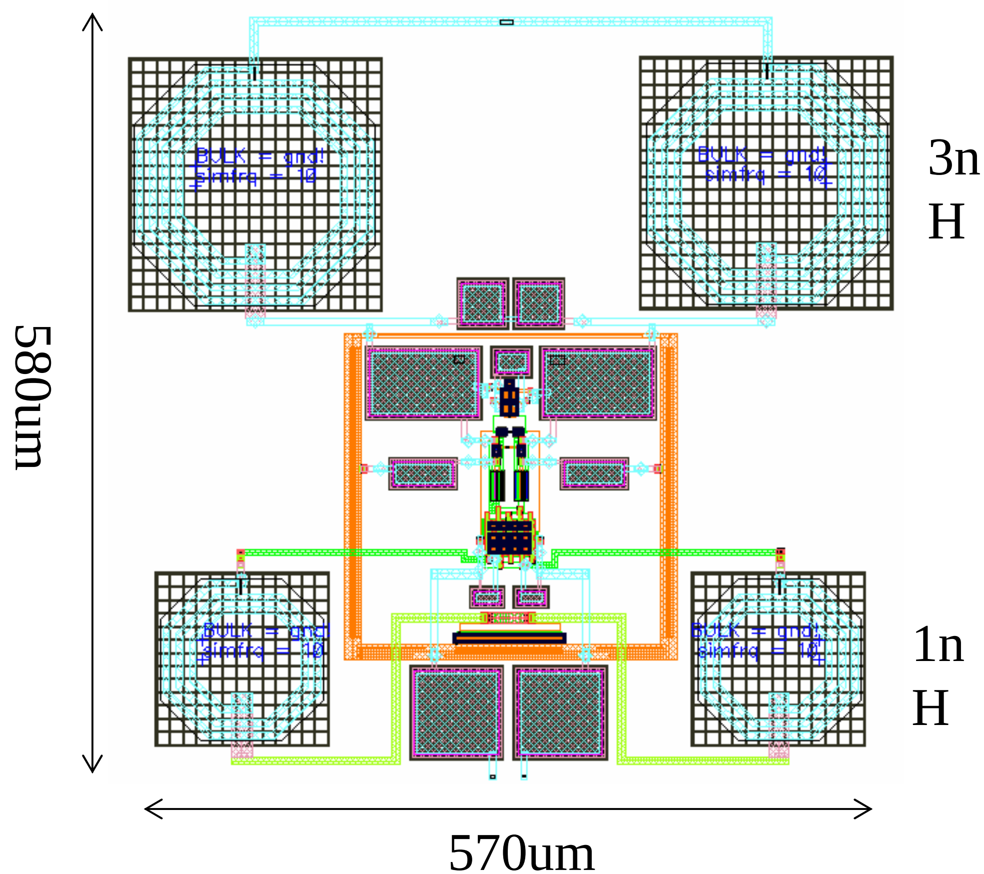
LNA simulation results



Front-end for BT/WiFi Receiver

LNA Layout

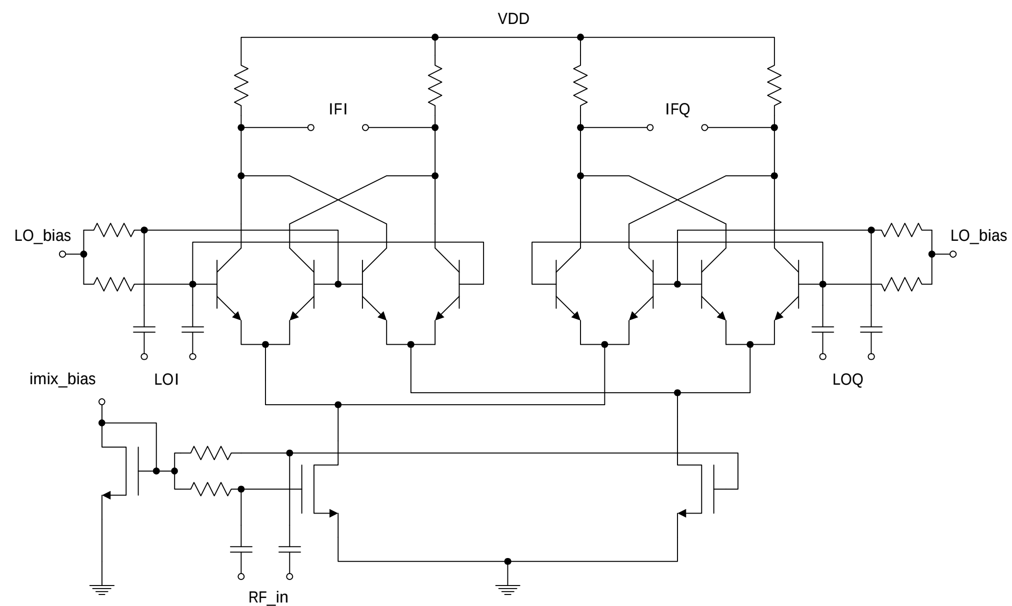
- Symmetrical layout
- Deep trench lattice under spiral inductor
- Inductors are placed far apart to avoid coupling
- Differential inputs are decoupled by GSGSG pattern



Front-end for BT/WiFi Receiver

I/Q Mixer

- Fully-differential structure to suppress common-mode noise
- I/Q branches share the same RF drive stage to achieve better matching between I and Q
- Bipolar current switches requires low LO drive power (-10dBm)
- Resistive loads have higher linearity and lower low-frequency noise compared to PMOS active load.



IIP3: 2dBm

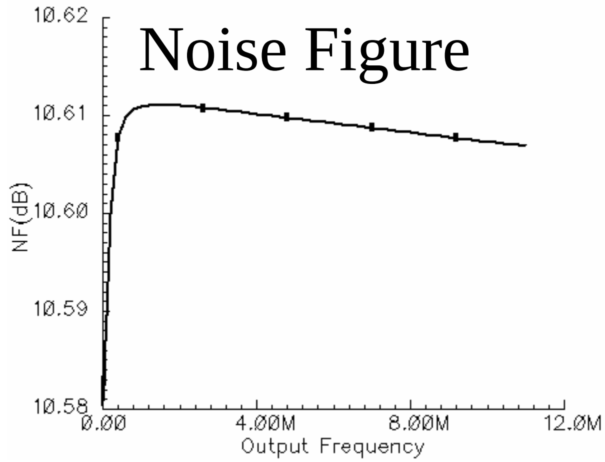
NF: 10.6dB

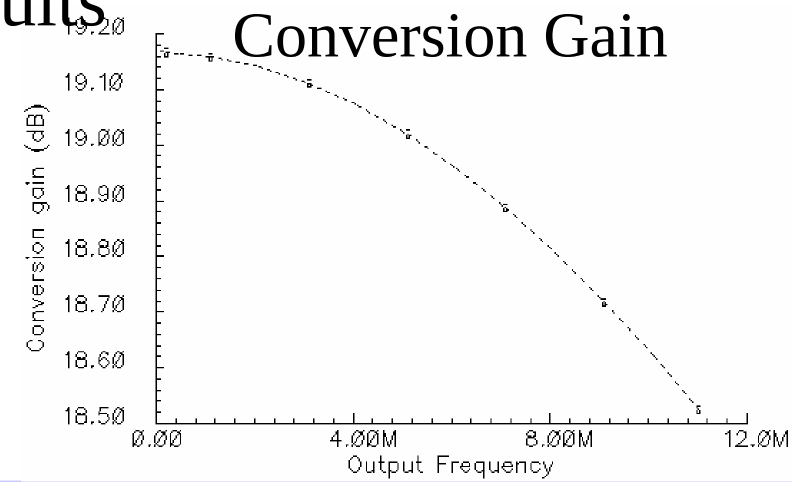
Gain: 18.5dB

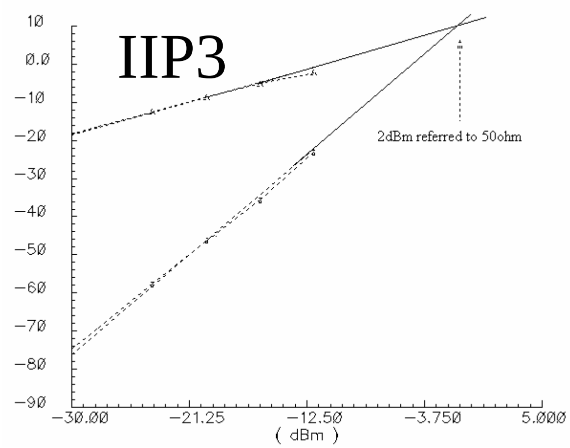
Power: 8.8mW

Front-end for BT/WiFi Receiver

I/Q Mixer Simulation Results

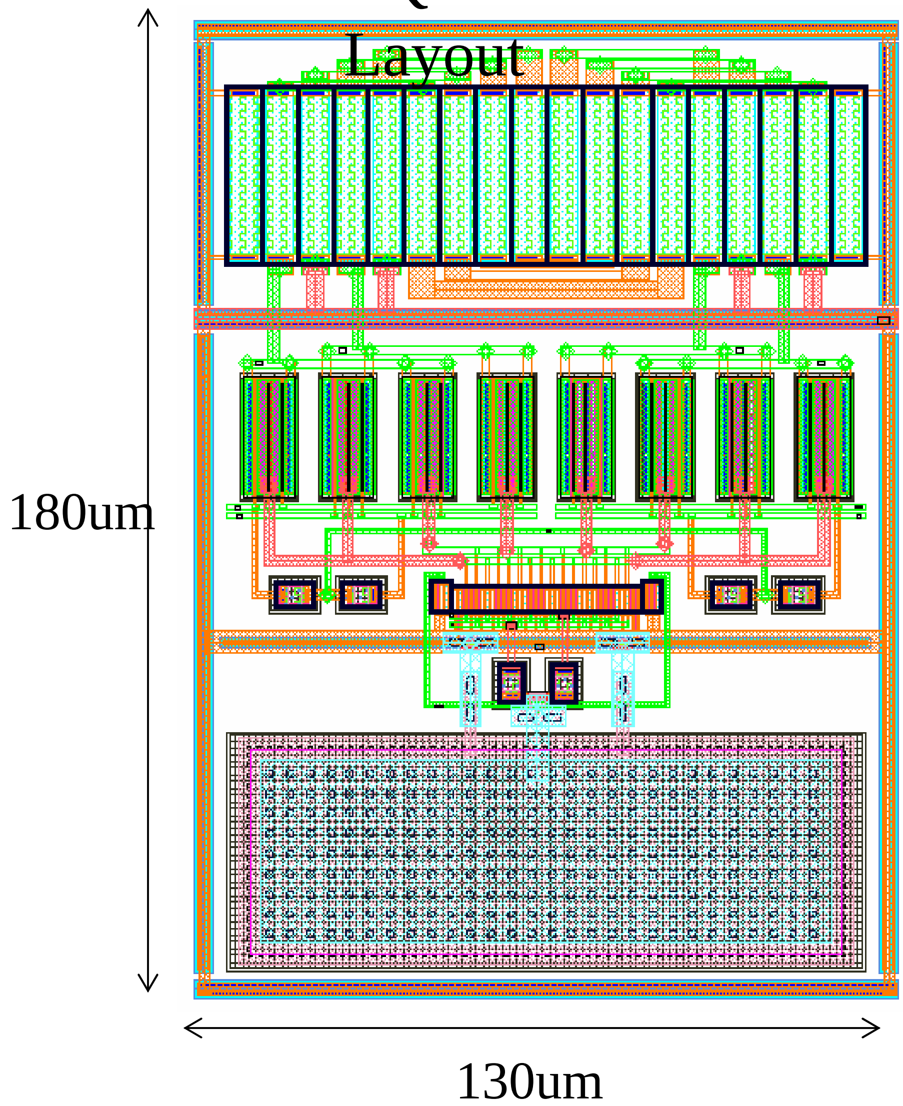






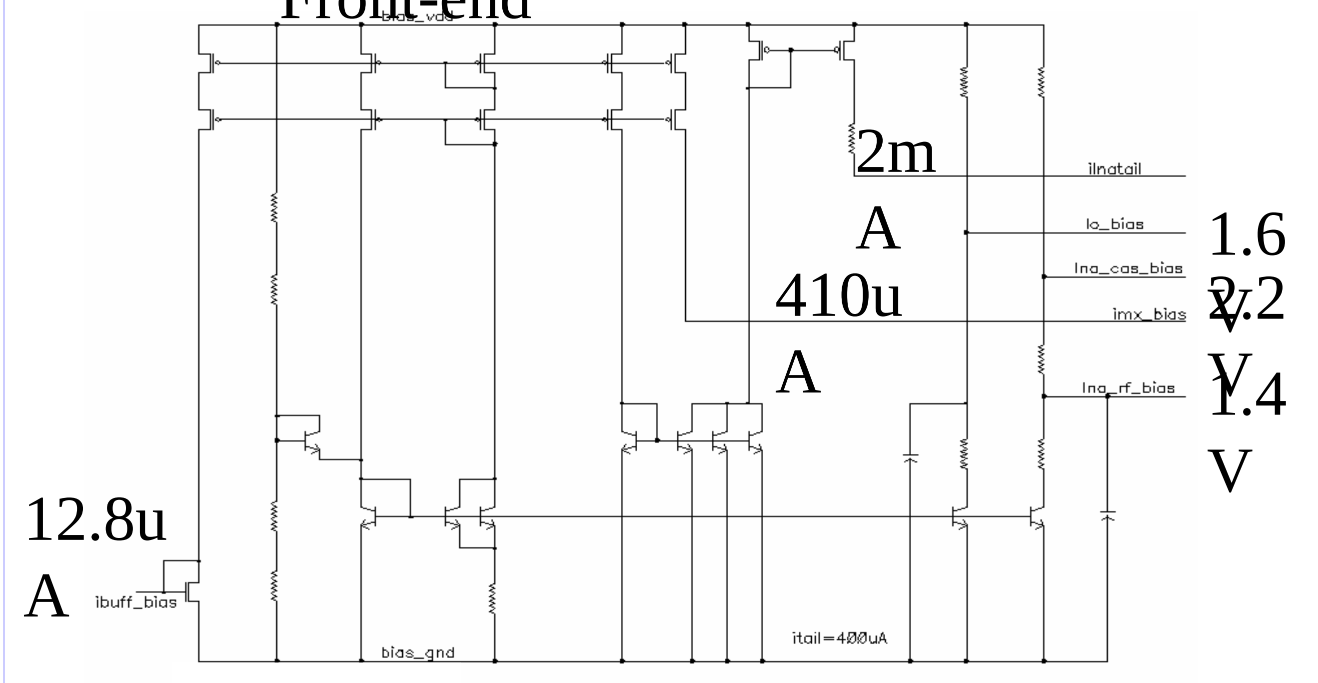
Front-end for BT/WiFi Receiver

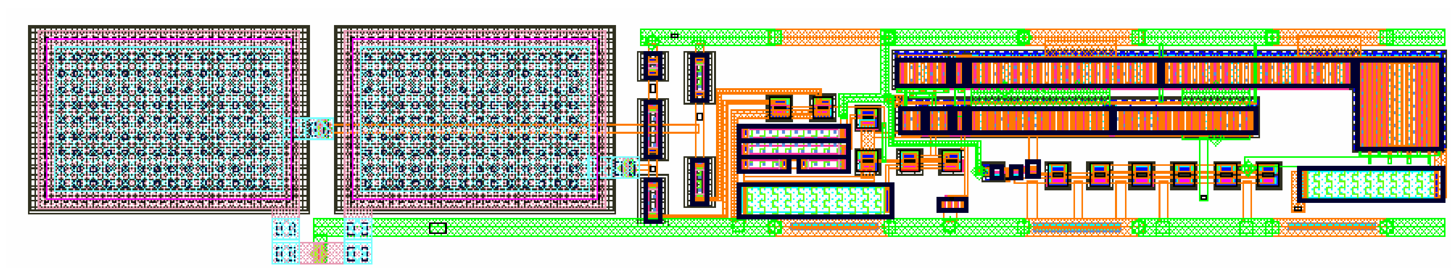
I/Q Mixer Layout



Front-end for BT/WiFi Receiver

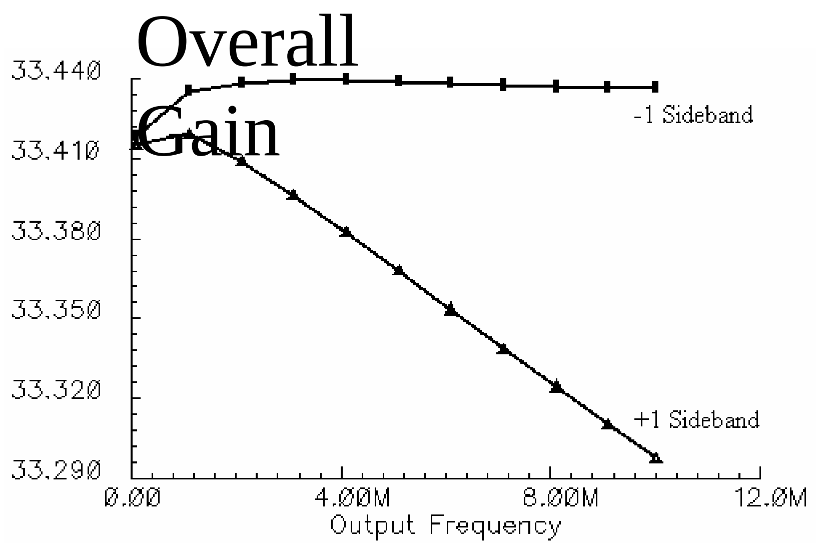
Biasing Circuits for the Front-end

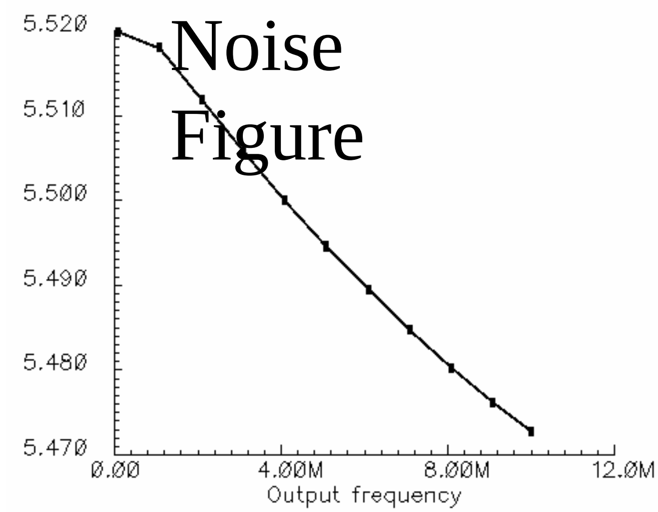




Front-end for BT/WiFi Receiver

Performance of the front-end





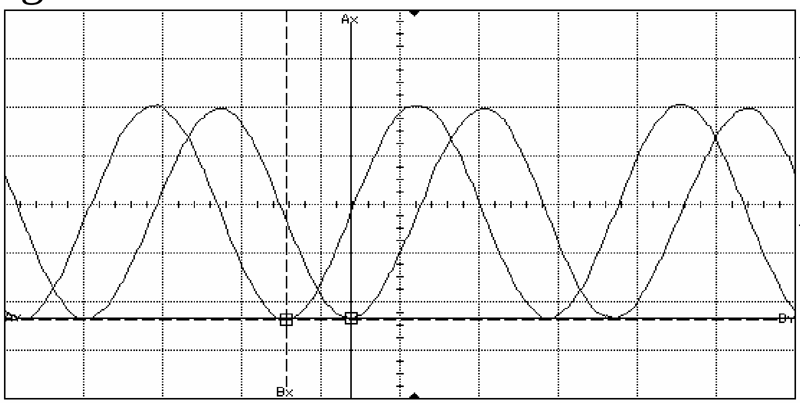
LO drive: -10dBm

Noise Figure: 5.5dB

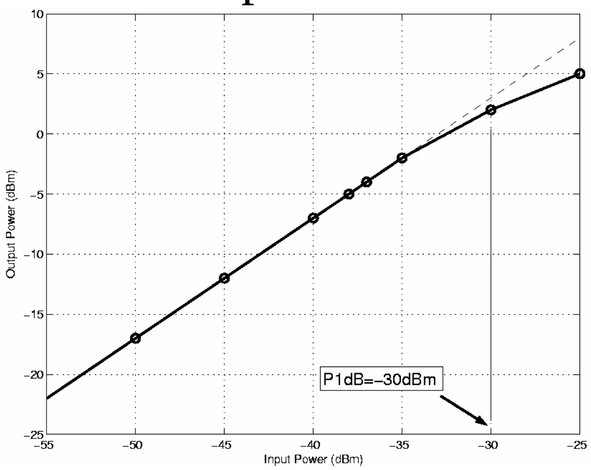
Overall gain: 33dB

Front-end for BT/WiFi Receiver

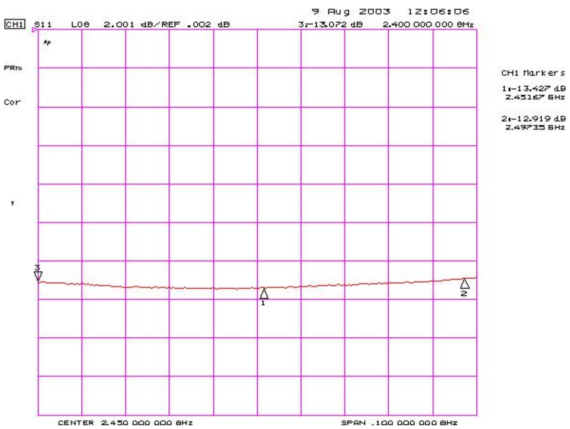
I/Q mismatch: amplitude 2.3% phase 1.7 degrees



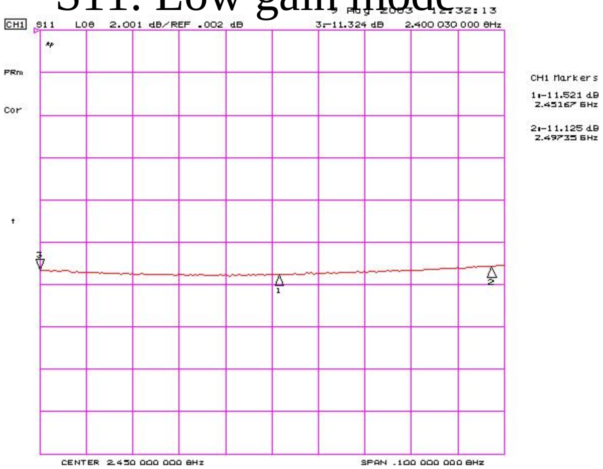
1dB compression



S11: High gain mode



S11: Low gain mode

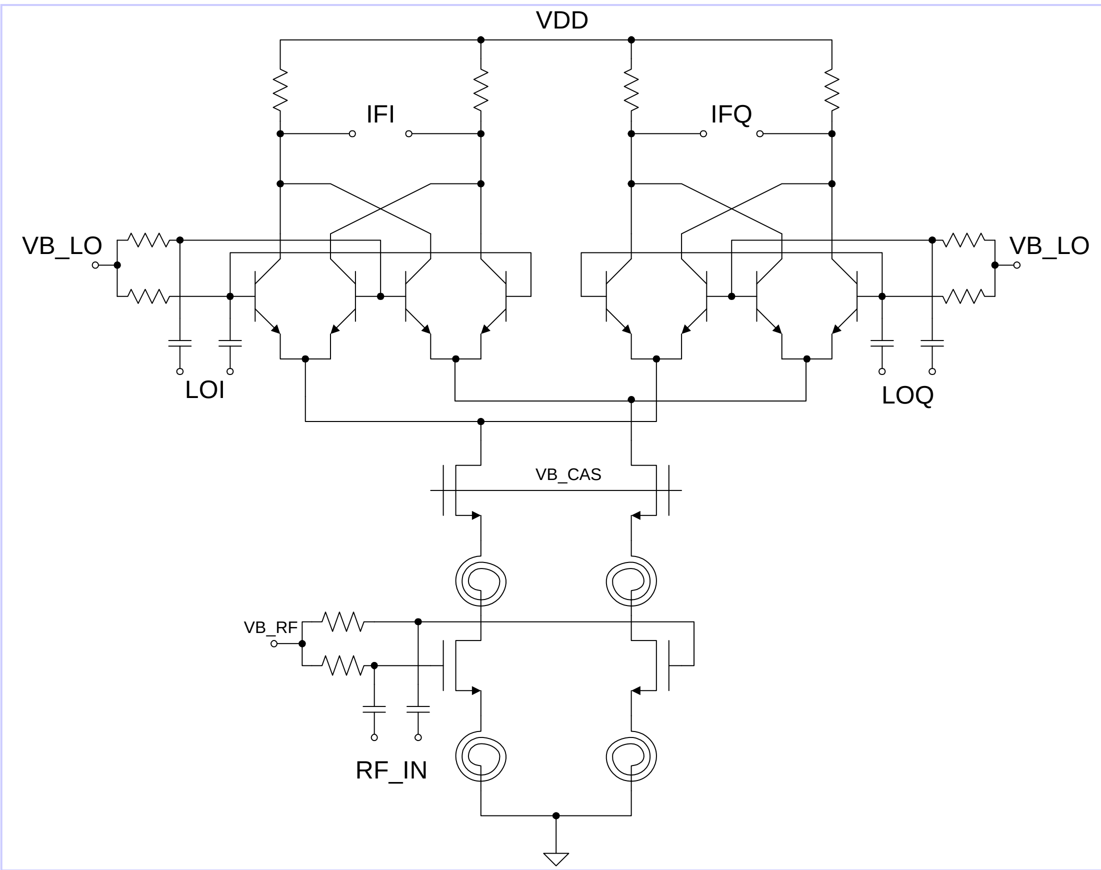


Circuit Implementation --- Merged LNA and Mixer (An alternative way)

Power: 28.5mW

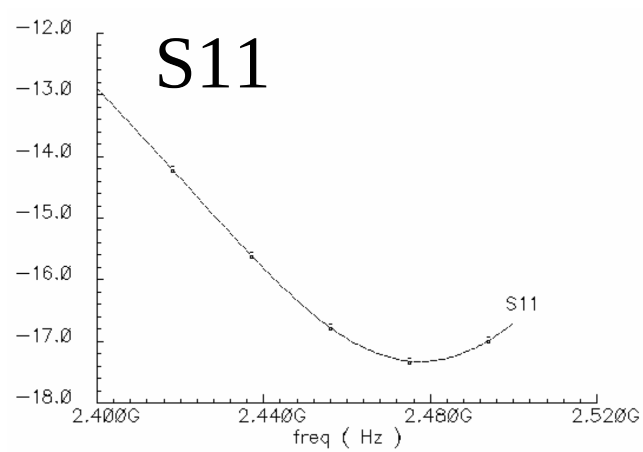
Noise Figure: 3.6dB

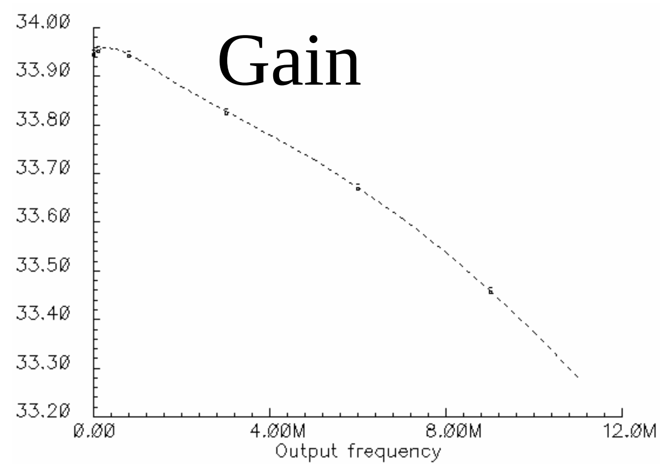
Gain: 33.4dB

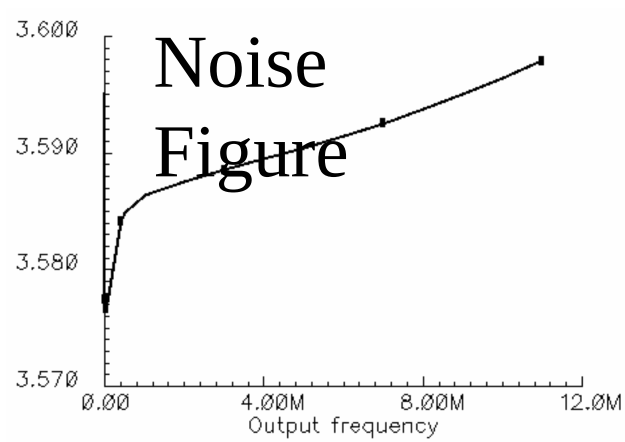


- Inter stage matching
- Inductive degeneration
- Current re-use
- Low power consumption

Simulation Results --- Merged LNA and Mixer







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4. G. Han, E. Sanchez-Sinencio, "CMOS transconductance multipliers: a tutorial ," *IEEE Transactions on Circuits and Systems II: Analog and Digital Signal Processing* , Volume: 45, Issue: 12 , pp. 1550 –1563, Dec 1998