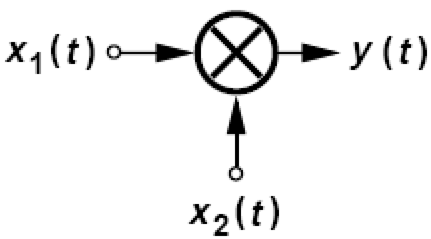
Radio-Frequency IC Design

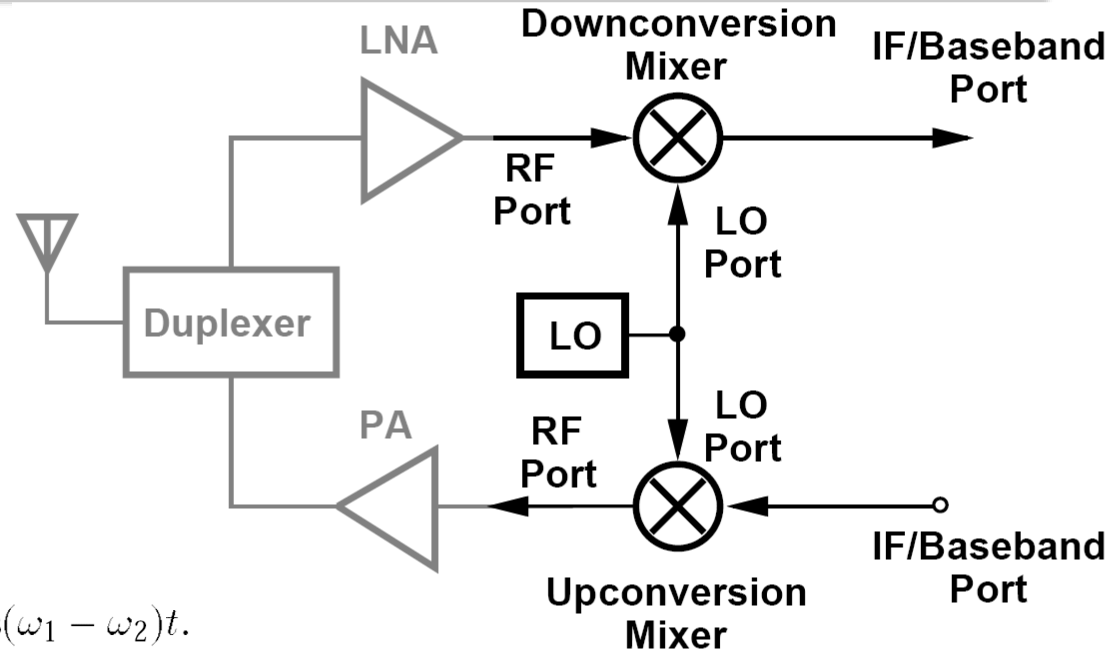
Lecture 9:
Receivers & Mixers

ELEC 404

Acknowledgement: *RF Microelectronics*, B. Razavi

Mixers

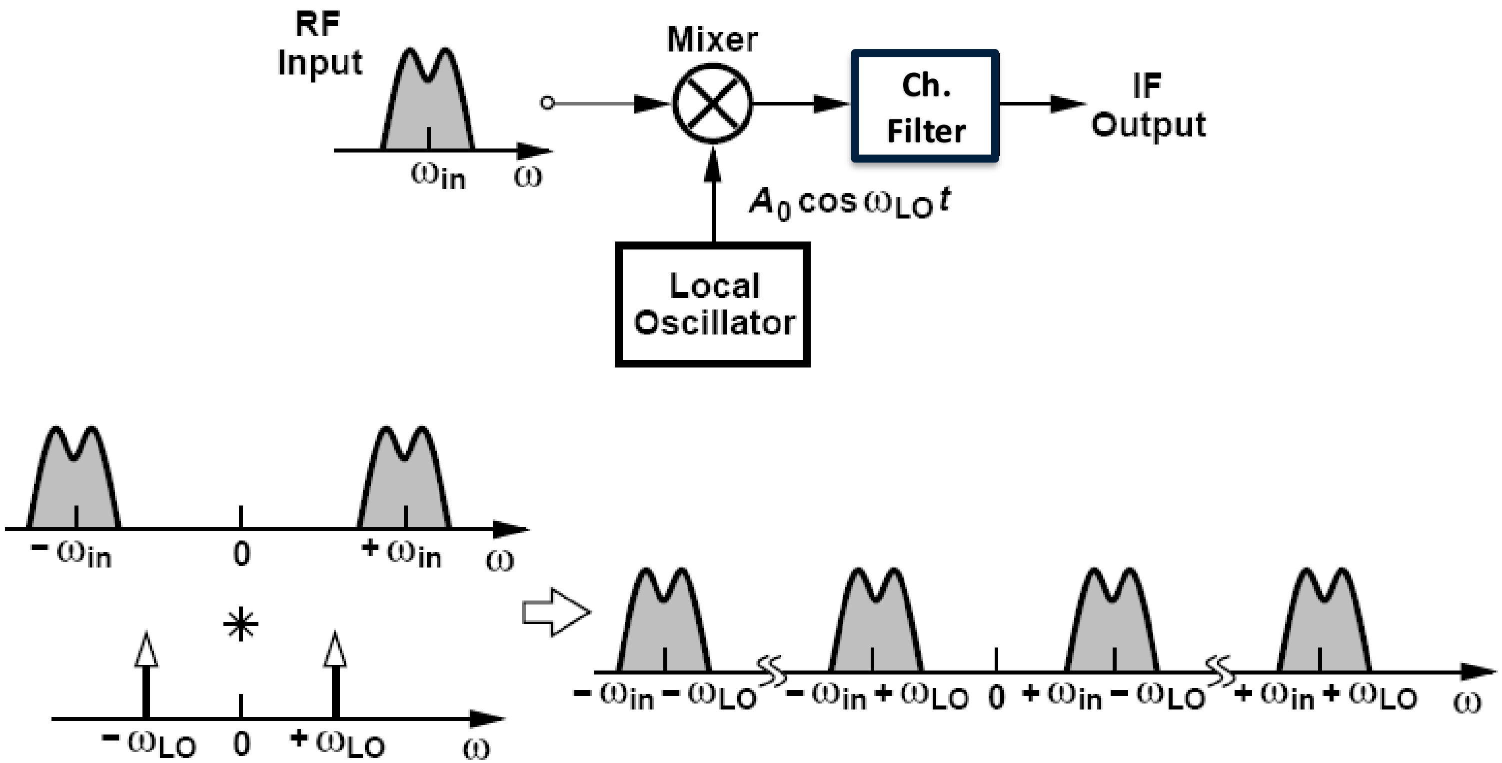




$$
\begin{aligned}
y(t) &= k(A_1 \cos \omega_1 t)(A_2 \cos \omega_2 t) \\
&= \frac{kA_1A_2}{2} \cos(\omega_1 + \omega_2)t + \frac{kA_1A_2}{2} \cos(\omega_1 - \omega_2)t.
\end{aligned}
$$

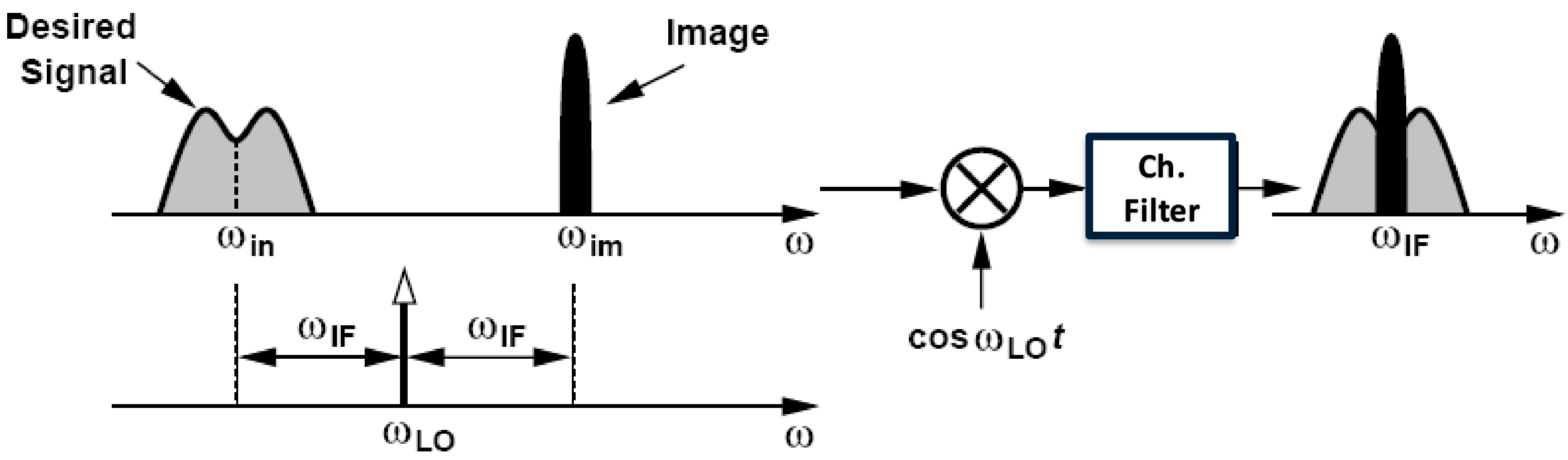
- **Perform frequency translation by multiplying two waveforms (and possibly their harmonics).**
- **Nonlinear LO port.**
- **Linear RF port to minimize compression and/or intermodulation.**
- **Implementation – Active/passive, single-balanced/double-balanced**
- **Voltage <u>Conversion</u> Gain = $[kA_1A_2/2]/A_1 = kA_2/2$**

Basic Heterodyne Receivers



- **Nonzero IF (Intermediate Frequency) → $\omega_{LO} \neq \omega_{in}$**
- **Due to its high noise, the downconversion mixer is preceded by an LNA**

Basic Heterodyne RX: Image Aliasing

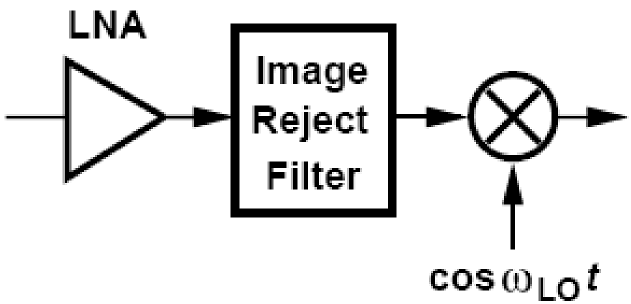


$$A\cos(\omega_{LO}-\omega_{in})t = A\cos\omega_{IF}t = A\cos(\omega_{im}-\omega_{LO})t$$

$$\omega_{im} = \omega_{in} + 2\omega_{IF} = 2\omega_{LO} - \omega_{in}$$

- **Two spectra located symmetrically around ω_{LO} downconverted to IF**
- **Image cannot be removed through filtering at the IF output**
- **Image rejection ratio: dB(P_{IF} desired/P_{IF} image)**

Image Rejection



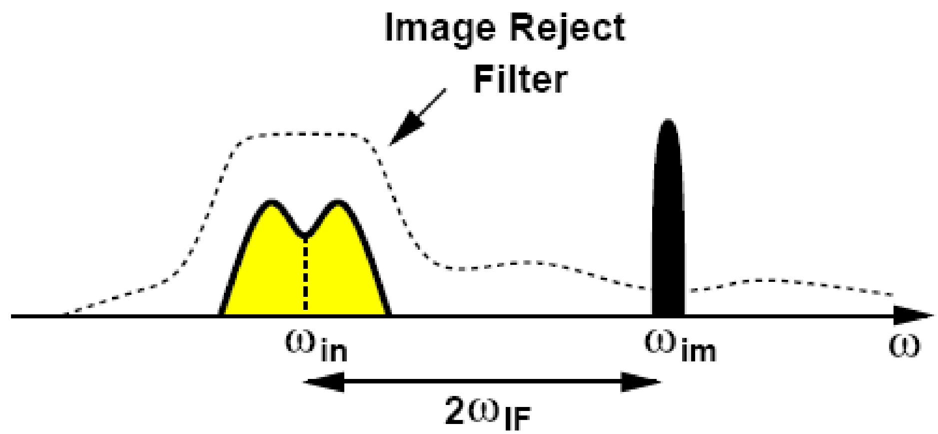
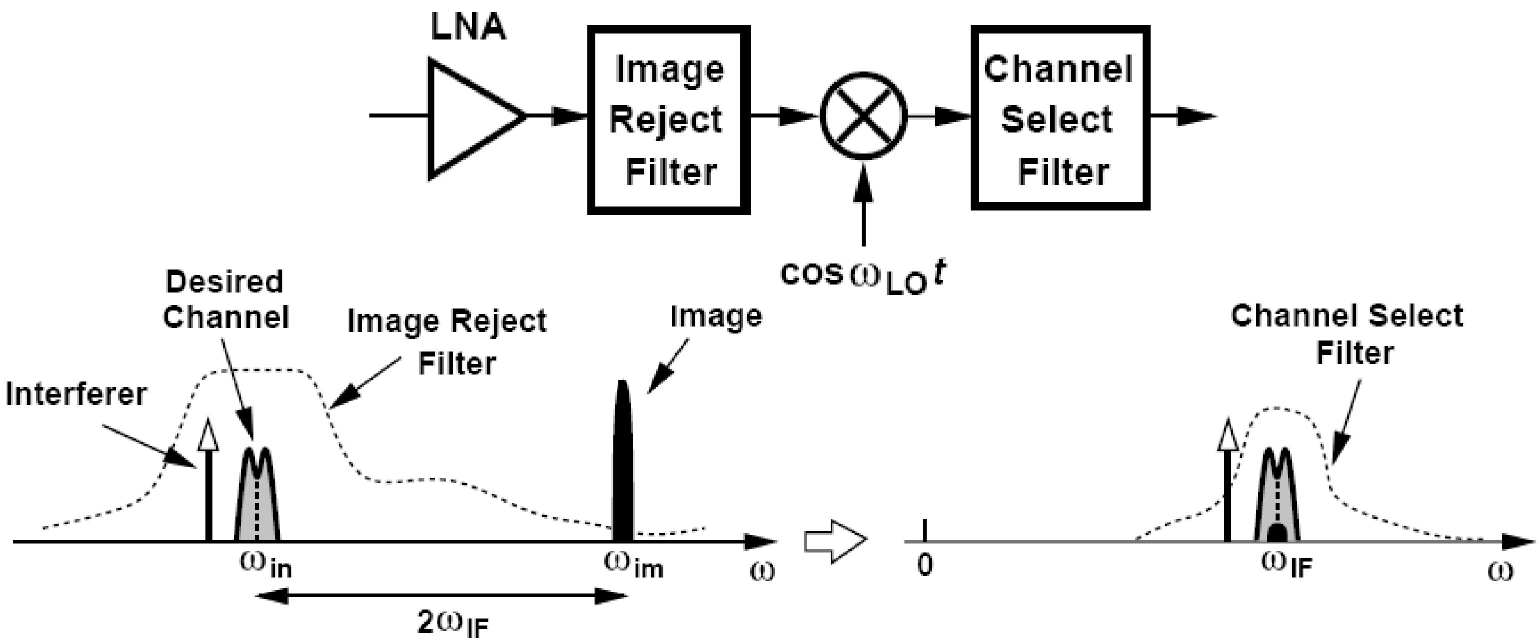


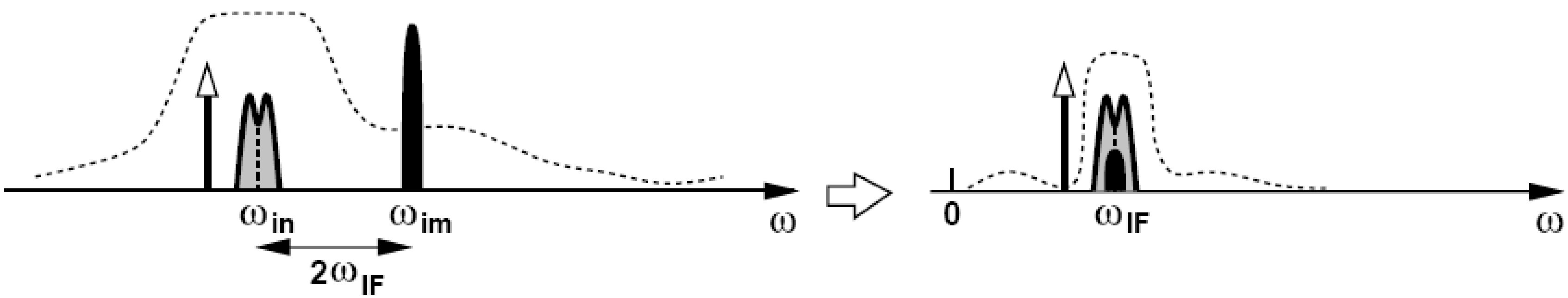
Image-reject filter for high image rejection

- **Typically between LNA & mixer**
  - **LNA gain lowers the filter's contribution to the RX NF**
- **Linearity and selectivity requirements have dictated passive, off-chip implementations**

Image Rejection vs. Channel Selection

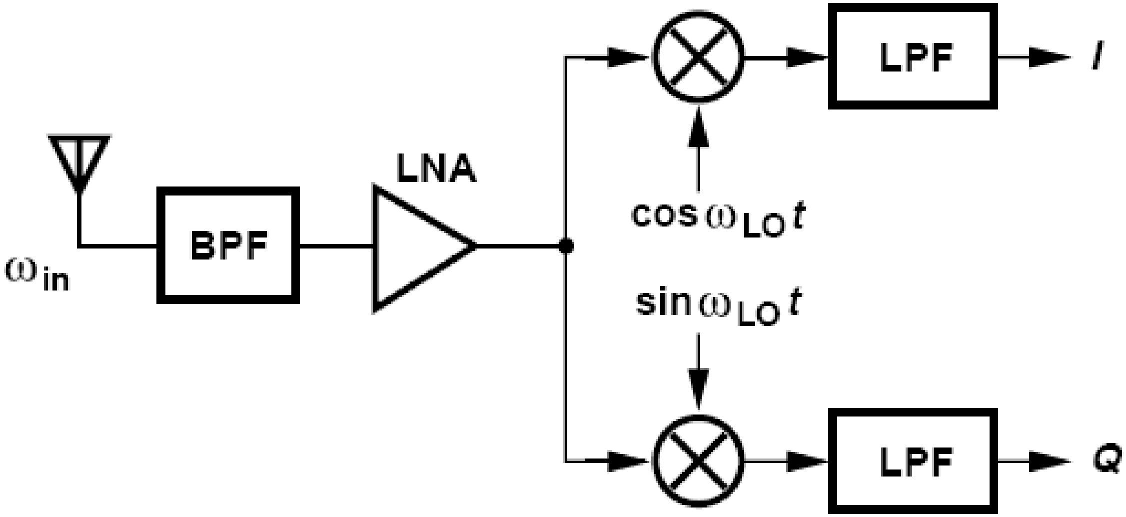


- A high IF allows substantial image-rejection but poor channel selection for given Q_{CSF} .



- A low IF helps with the channel selection and suppression of in-band interferers.

Direct-Conversion Receivers



- Absence of an image greatly simplifies the design process
- Channel selection is performed by on-chip LPF
- Mixing spurs are considerably reduced in number
- Quadrature phases to prevent self-corruption of signal (asymmetrically-modulated) upon downconversion

Specifications: Noise, Linearity, Gain

LNA Gain and Mixer Noise/Linearity tradeoff

- Input noise of mixer divided by the LNA gain when referred to the RX input
- But IP_3 of the mixer also scaled down by the LNA gain

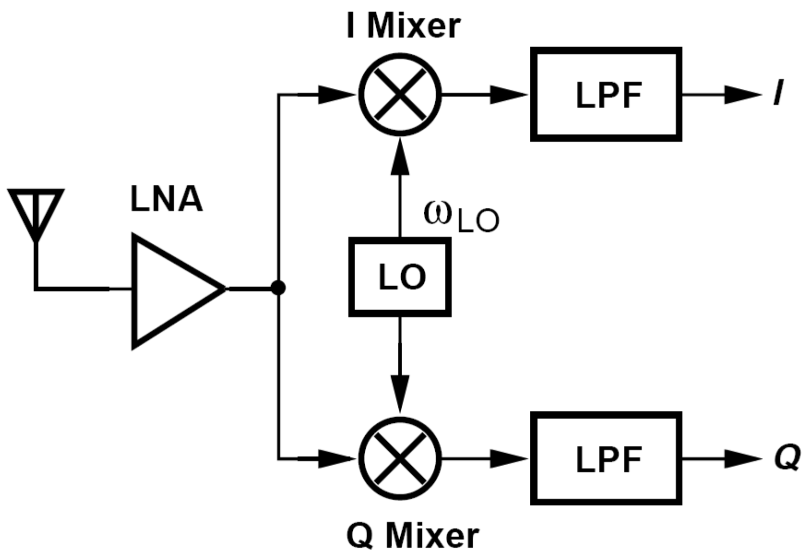
$$F \approx F_{LNA} + \frac{F_{Mixer} - 1}{A_{PLNA}} + \frac{F_{substg} - 1}{A_{PLNA}A_{PMixer}}$$

$$\frac{1}{IIP3} \approx \frac{1}{IIP3_{LNA}} + \frac{G_{LNA}}{IIP3_{Mixer}} + \frac{G_{LNA}G_{Mixer}}{IIP3_{substg}}$$

Mixer Gain and RX Noise/Linearity tradeoff

- High mixer gain important in suppressing noise from subsequent stages
- Linearity of mixer critical for RX linearity
- Both high gain and high linearity difficult to attain in Mixer
  - ~ <10dB gain while retaining linearity at low Vdd

NF of DCR

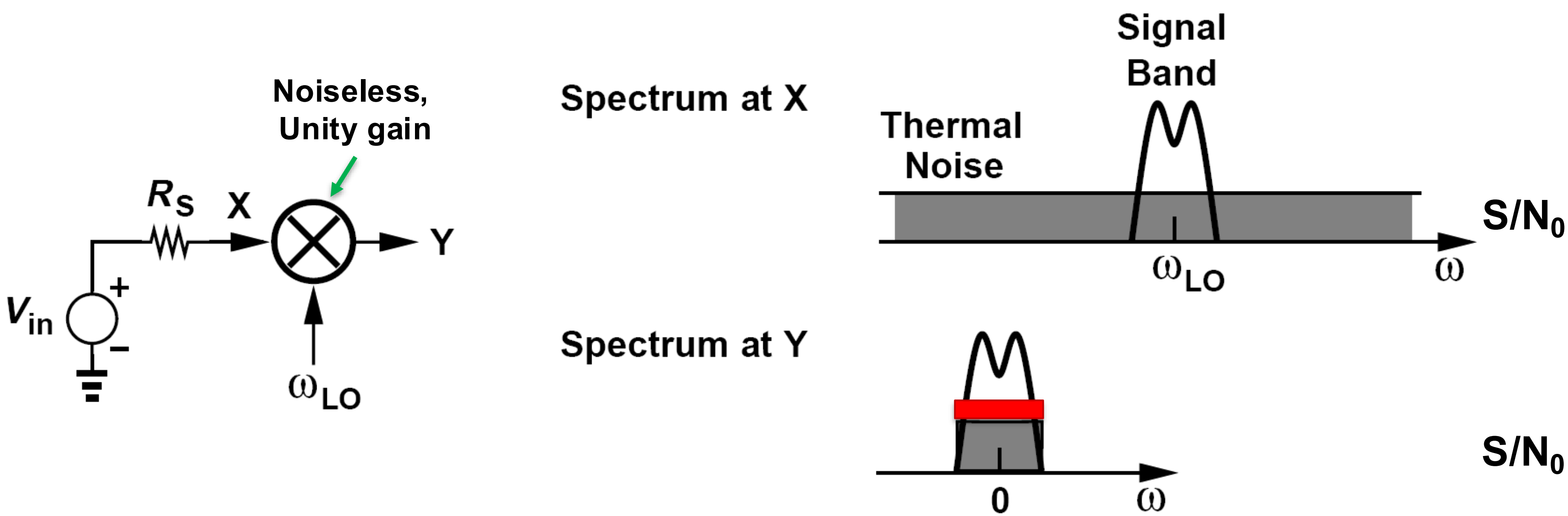


$$F = \frac{\mathrm{SNR}_{in}}{\mathrm{SNR}_{I}} = \frac{\mathrm{SNR}_{in}}{\mathrm{SNR}_{Q}}$$

- **SNR in the final combined output would serve as a more accurate measure of the noise performance, but it depends on the modulation scheme.**
- **Double Sideband (DSB) or Single Sideband (SSB)?**

Mixer Noise Figures: DSB NF

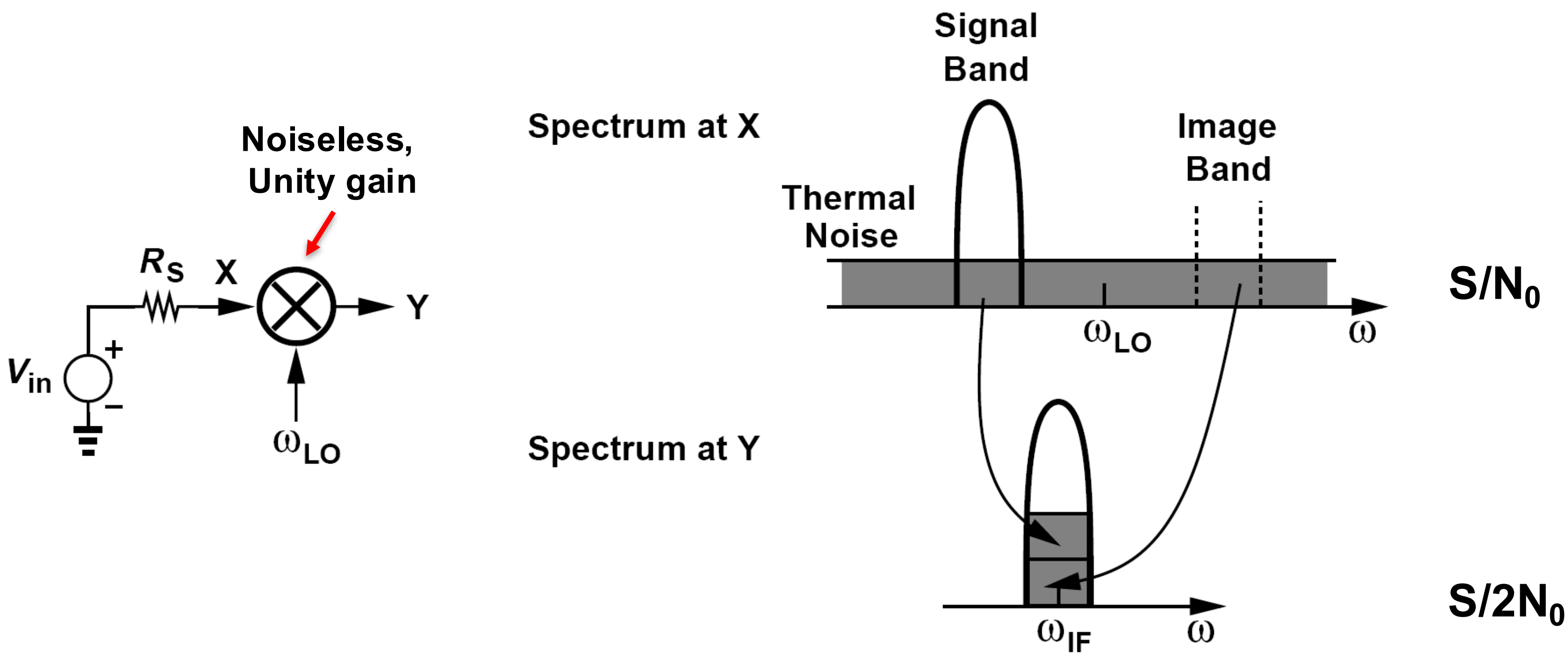
Mixer in a DCR implementation



- Signal & noise in the signal band translated to the BB → $SNR_{op} = SNR_{ip}$
- NF of the noiseless mixer = 0dB
- For a noisy mixer, NF calculated this way gives the DSB NF = DSB-NF_{mixer}

Mixer Noise Figures: SSB NF

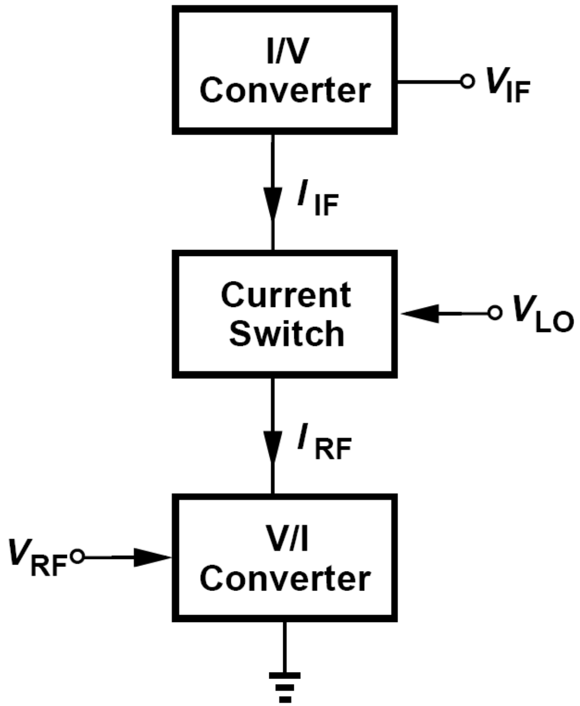
Same mixer in Heterodyne implementation, exhibiting a flat frequency response @ RF Port from the image band to the signal band.

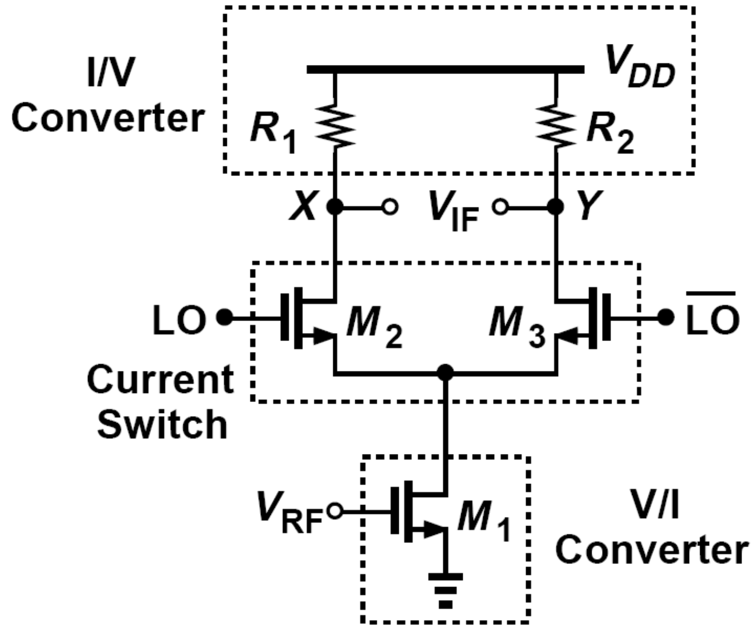


- $SNR_{op} = SNR_{ip}/2$
- NF of a noiseless mixer is 3dB!
- For a noisy mixer, NF calculated this way gives the "single-sideband" (SSB) NF
- $SSB\text{-}NF_{mixer}$ = 3dB + $DSB\text{-}NF_{mixer}$

Active Downconversion Mixers

- Conversion and gain in one stage.
  - Convert RF voltage → RF current,
  - "commutate" (steer) the RF current by the LO, and
  - convert IF current → IF voltage.



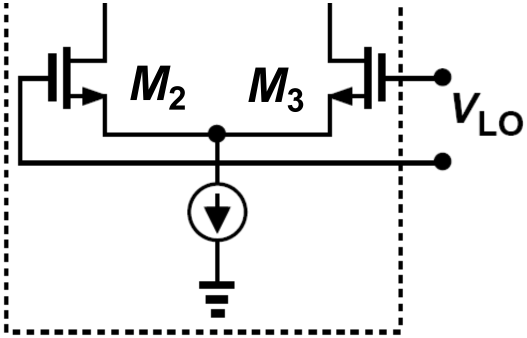


- M_2 and M_3 "switching pair"
- Switching pair needs *large* LO swings for complete steering from one side to another, but not too large (or rail-to-rail).

Single Balanced Mixer – I

$$i_{out,\,diff} = sgn(cos(\omega_{LO}t))[I_{DC} + I_{RF}cos(\omega_{RF}t)]$$

$$sgn\left(cos(\omega_{LO}t)\right) = \frac{4}{\pi}\left(cos(\omega_{LO}t) + \frac{1}{3}cos(3\omega_{LO}t) + \frac{1}{5}cos(5\omega_{LO}t) + \cdots\right)$$



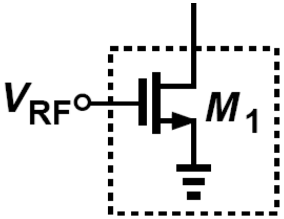
Single Balanced Mixer – II

$$i_{out,\,diff} = \frac{4}{\pi}\left(cos(\omega_{LO}t) + \frac{1}{3}cos(3\omega_{LO}t) + \cdots\right)I_{DC} +$$

$$\frac{4}{\pi}\left(cos(\omega_{LO}t)cos(\omega_{RF}t) + \frac{1}{3}cos(3\omega_{LO}t)cos(\omega_{RF}t) + \cdots\right)I_{RF}$$

- **Poor LO-IF isolation**
- **Conversion Gain (Transconductance):**

RF transconductors – CS



$$I_D = \frac{\mu COX \frac{W}{L}}{2} \frac{(VGS - VTH)^2}{1 + \left(\frac{\mu_0}{2vsatL} + \theta\right)(V_{GS} - VTH)}$$

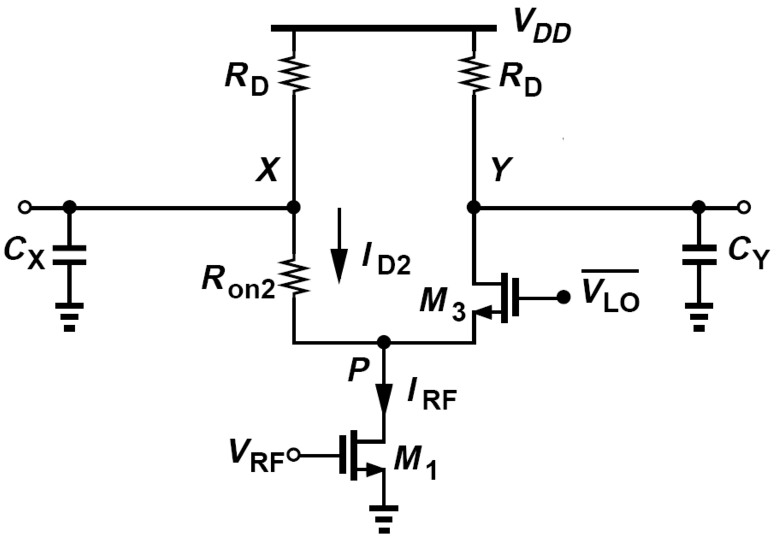
$$A_{IIP3} = \sqrt{\frac{\frac{8}{3}(V_{GS0} - VTH)}{\frac{\mu_0}{2vsatL} + \theta} - 4(VGS_0 - VTH)^2}$$

Linearity

- The input transistor imposes a direct trade-off:

$$IIP3 \propto V_{GS} - VTH \qquad G_{c,\,max} \propto \frac{1}{V_{GS} - VTH} \qquad \overline{V_{n,in}^2} \propto \frac{1}{g_m} \propto \frac{V_{GS} - VTH}{I_D}$$

- Linearity also degrades if the switching transistors enter the triode region → LO swings cannot be arbitrarily large.
  - If M_2 and M_3 operate in saturation, I_{RF} splits in the ratio of g_{m2}/g_{m3}, and independent of V_X and V_Y
  - If M_2 is in triode while M_3 is still ON and in saturation, I_{D2} is a function of V_X, leading to signal-dependent current division between M_2 and M_3.



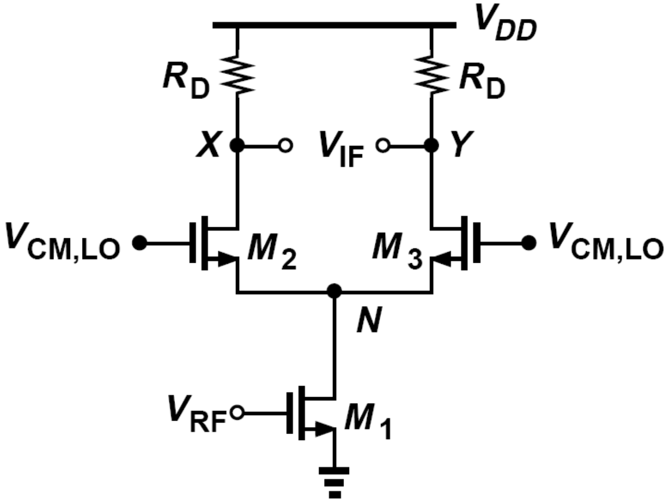
Equilibrium Overdrive of a Diffamp

Overdrive volt. seen by M_2 and M_3 when both of them carry a current of $I_1/2$

- **For square law devices,**

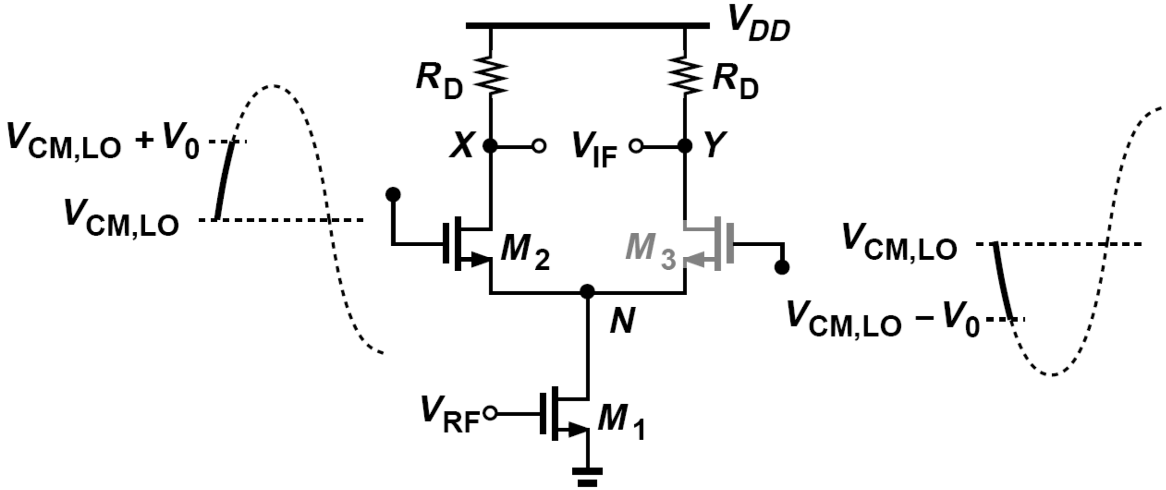
$$\frac{I_1}{2} = \frac{\beta_2 (VGS_2 - VTH_2)_{eq}^2}{2}$$

$$(VGS_2 - VTH_2)_{eq} = \sqrt{\frac{I_1}{\beta}}$$



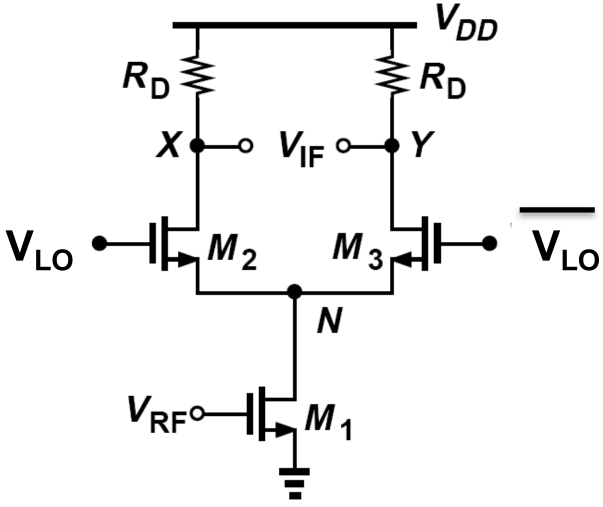
- **The diffamp steers most of its tail current if**

$$V_0 = \frac{\sqrt{2}(V_{GS2} - VTH_2)_{eq}}{2}$$



Conversion Gain vs. Volt. Headroom

$$A_v = \frac{2}{\pi} g_{m1} R_D$$



- G_{m1} is limited by current budget and IP3

$$g_{m1} = \frac{2ID_1}{V_{GS} - VTH}$$

- R_D is limited by max allowable DC voltage across it.

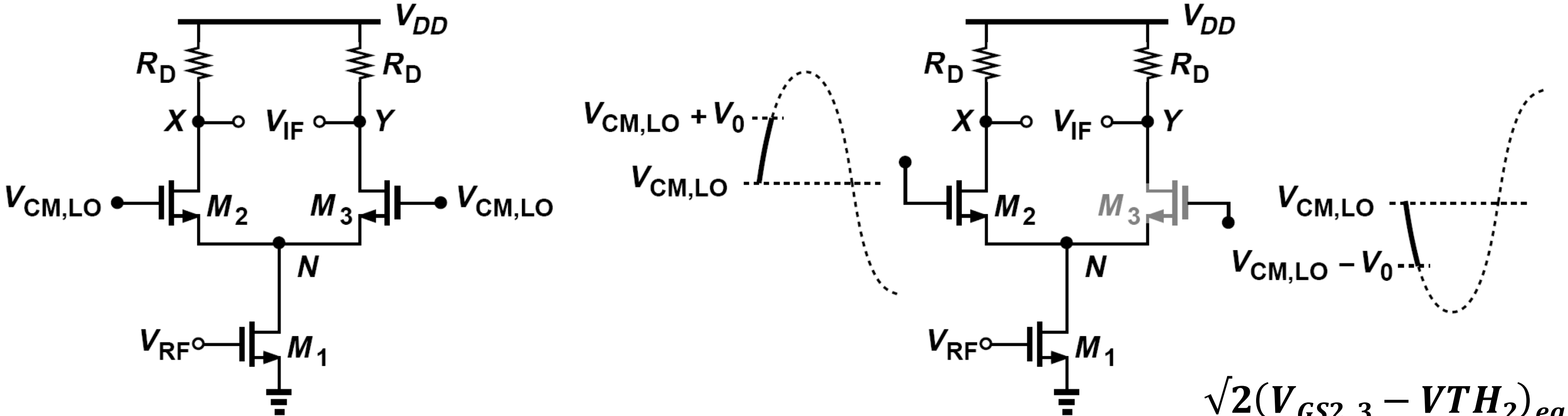
$$R_{D\,max} = \frac{V_{R\,max}}{I_{D1}/2}$$

$$V_{R\,max} = V_{DD} - \left[V_{GS1} - VTH_1 + \left(1 + \frac{1}{\sqrt{2}}\right)(V_{GS2} - VTH_2)_{eq}\right]$$

$$A_{V,max} = \frac{2}{\pi} g_{m1} R_{D,max} = \frac{8}{\pi} \frac{V_{R,max}}{V_{GS1} - V_{TH1}}$$

- Low V_{DD} severely limits the conversion gain of active mixers.

$R_{d,max}$ [→ $V_{x,min}$ or $V_{y,min}$]



$$V_0 = \frac{\sqrt{2}(V_{GS2,3} - VTH_2)_{eq}}{2}$$

If M_1 is at the edge of saturation, then $V_N \geq V_{GS1} - V_{TH1}$

$$V_{CM,LO} - V_{GS2,3} \geq V_{GS1} - V_{TH1}$$

For M_3 to turn off and M_2 to remain in saturation up to this point, $V_{D2} \geq V_{G2} - V_{TH2}$

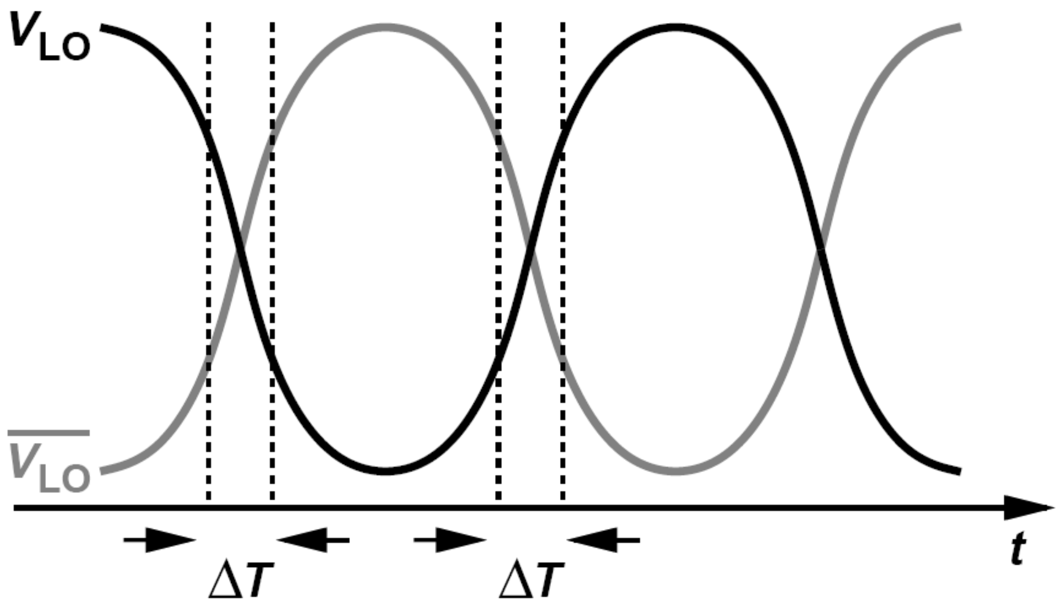
$$V_{X,min} = V_{CM,LO} + \frac{\sqrt{2}}{2}(V_{GS2,3} - V_{TH2}) - V_{TH2}$$

$$= V_{GS1} - V_{TH1} + \left(1 + \frac{\sqrt{2}}{2}\right)(V_{GS2,3} - V_{TH2})$$

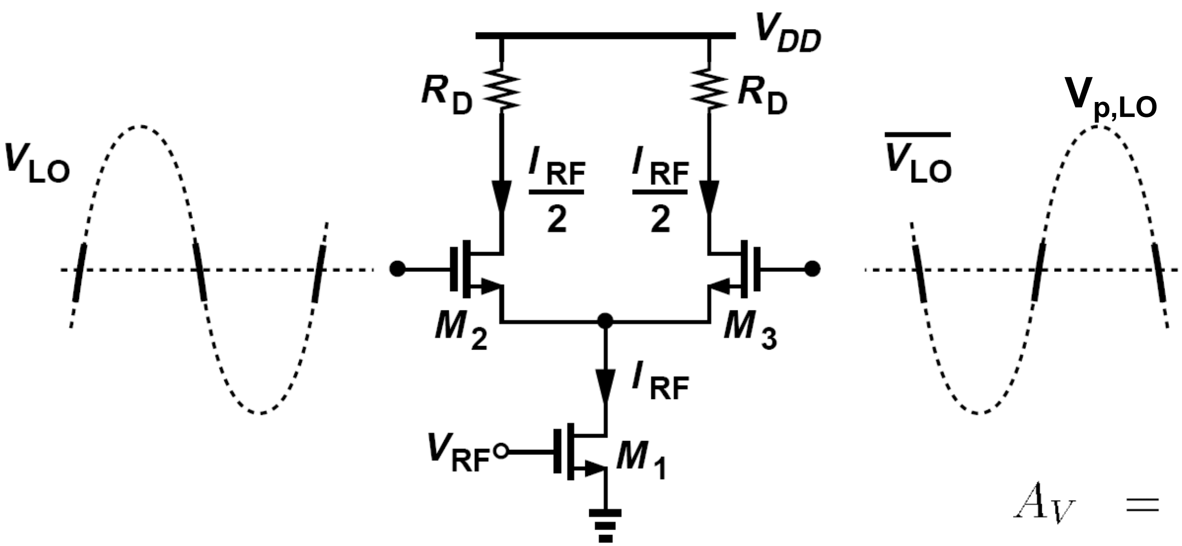
$V_{R,max} = V_{DD} - V_{X,min}$

Ideal LO Waveform

- Ideal LO → Square wave to ensure abrupt switching & max. conversion gain
- @ high RF, LO → inevitably sinusoids
- Downconversion of interferers located at the LO harmonics is a serious issue in broadband RX.



Gain with gradual LO transitions

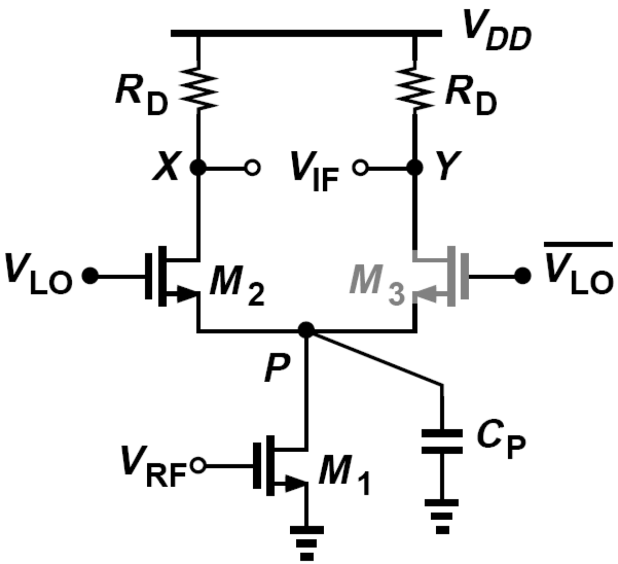


$$V_0 = \frac{\sqrt{2}(V_{GS2,3} - VTH_2)}{2}$$

$$\begin{aligned} A_V &= \frac{2}{\pi} g_{m1} R_D \left(1 - \frac{2\Delta T}{T_{LO}}\right) \\ &= \frac{2}{\pi} g_{m1} R_D \left[1 - \frac{(V_{GS} - V_{TH})_{eq}}{5\pi V_{p,LO}}\right] \end{aligned}$$

- **When M_2 and M_3 are near equilibrium, the RF current produced by M_1 is split approximately equally between them, thus appearing as a CM current and yielding little conversion gain for that period of time (ΔT).**
- **Lower LO swing → higher ΔT and lower gain**

Gain with Parasitic Cap



With abrupt LO edges, M_2 is on and M_3 is off →

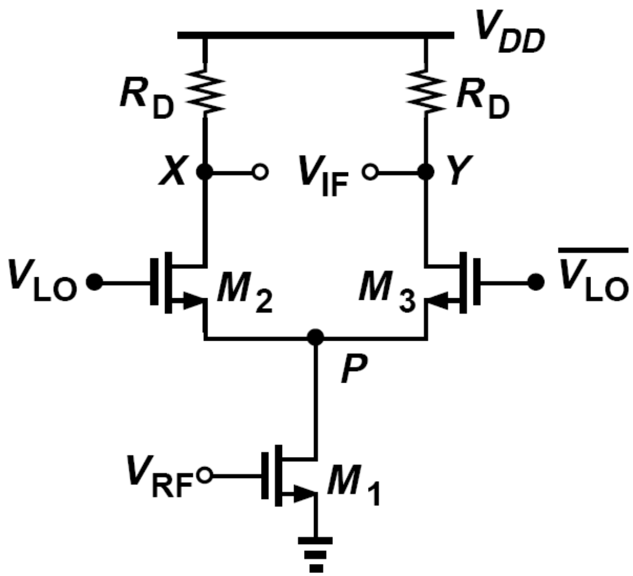
$$C_P = C_{DB1} + C_{GS2} + C_{GS3} + C_{SB2} + C_{SB3}.$$

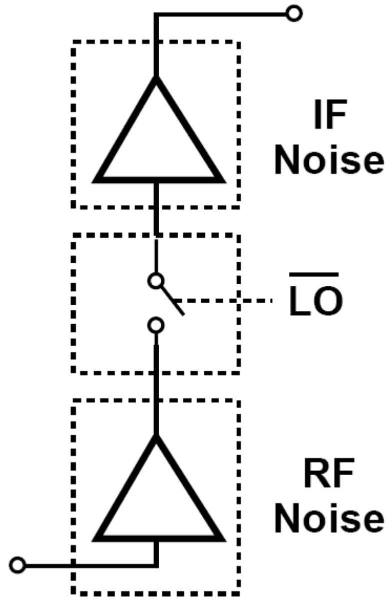
- **The RF current produced by M_1 is split between C_P and $1/g_{m2}$.**

$$A_{V,max} = \frac{2}{\pi} g_{m1} R_D \left[1 - \frac{(V_{GS} - V_{TH})_{eq}}{5\pi V_{P,LO}}\right] \frac{g_{m2}}{\sqrt{C_P^2 \omega^2 + g_{m2}^2}}.$$

Noise in Active Mixers: Analysis

- The noise components of interest lie in the RF range before downconversion and in the IF range after downconversion.





Thermal	Flicker (active load)
Thermal	Flicker
Thermal	

- Total noise power @ IF o/p
  = Σ Power[magnitude of each noise source x its conversion gain to the IF o/p]
- Total noise power @ RF i/p
  = Total noise power @ IF o/p ÷ [overall mixer conversion gain]^2

Flicker Noise Referred to I/P

$$V_{n,out}(f) = \left(\frac{I_{SS}R_D}{\pi Vp_{,LO}} + 2fLOCP\right)V_{n2}(f)$$

Considering flicker noise of M_3 also, ignoring C_p and dividing by conversion gain,

$$V_{n,in}(f) = \frac{\sqrt{2}I_{SS}}{2g_{m1}V_{p,LO}}V_{n2}(f)$$

$$= \frac{\sqrt{2}(V_{GS}-V_{TH})_1}{4V_{p,LO}}V_{n2}(f)$$

- **$V_{n2}(f)$ is typically very large because M_2 and M_3 are relatively small**
- **For active load, possibly much higher flicker noise**

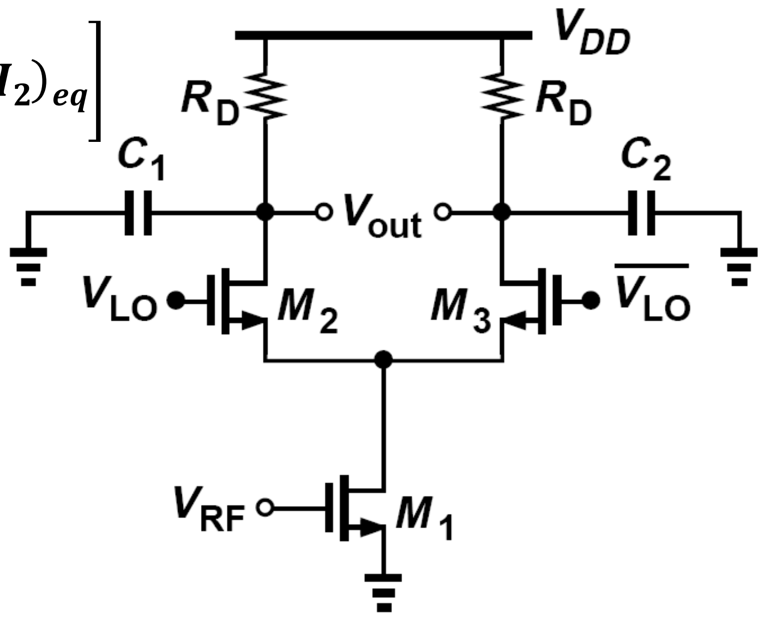
Single Balanced Mixer Design

$f_{RF} = f_{LO}$ = 6GHz, 65nm CMOS with I_{bias} = 2mA from V_{DD} = 1.2V. Assume a single-ended sinusoidal $V_{LO,0\text{-}pk}$ = 400mV.

Assign an overdrive volt. of 300mV to M_1, and 150mV to M_2 and M_3 (in equilibrium).

$$V_{R\,max} = V_{DD} - \left[V_{GS1} - VTH_1 + \left(1 + \frac{1}{\sqrt{2}}\right)(V_{GS2} - VTH_2)_{eq}\right]$$

= 600mV for each R_D.
With total I_{bias} = 2mA, conservatively R_D = 500 Ω.



$$\frac{2\Delta T}{T_{LO}} = \frac{(V_{GS2} - VTH_2)_{eq}}{5\pi V p_{,LO}}$$ is small. (Good!)

- Large V_{LO} → no degradation of conversion gain.
- However, very large V_{LO} can push switching transistor in triode.

Mixer Design – II

Transistor simulation curves for overdrive and I_D → W_1 =15μm, $W_{2,3}$ =20μm, g_{m1} =12.75mS.
C_1 = C_2 = 2pF to suppress LO component at o/p (prevent compression at the o/p).

$$\begin{aligned} A_v &= \frac{2}{\pi} g_{m1} R_D \\ &= 4.1\,(= 12.3\text{ dB}) \end{aligned}$$

$$\begin{aligned} \overline{V_{n,in}^2} &= \pi^2 kT \left(\frac{\gamma}{g_{m1}} + \frac{2}{g_{m1}^2 R_D} \right) \\ &= 4.21 \times 10^{-18}\ \text{V}^2/\text{Hz}, \end{aligned}$$

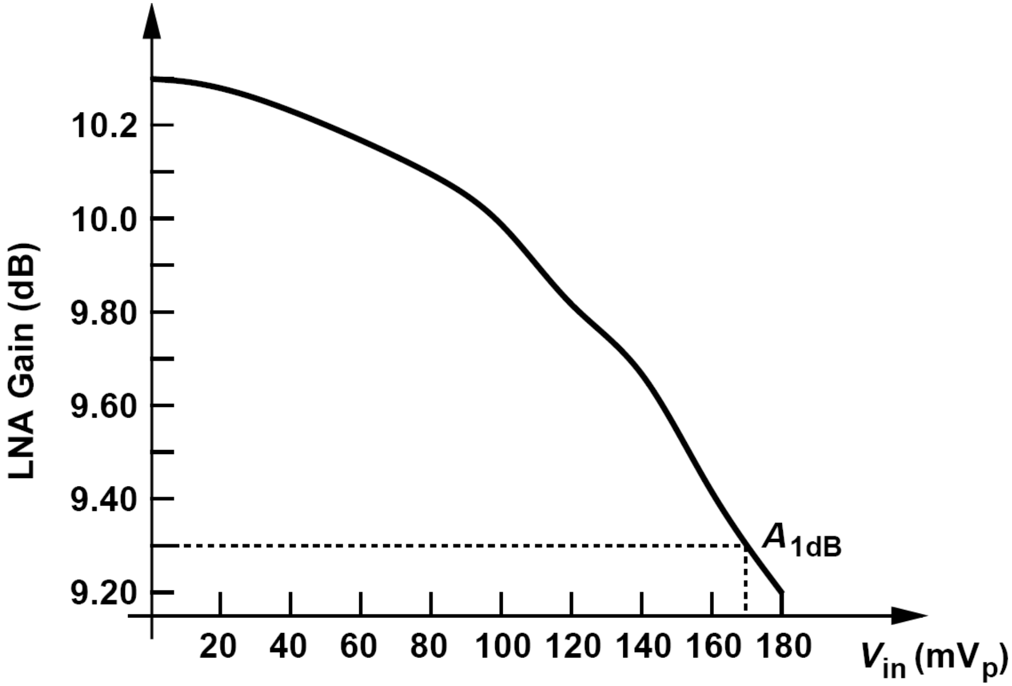
Single-sideband NF with respect to R_S = 50Ω:

$$\begin{aligned} \text{NF}_{SSB} &= 1 + \frac{\overline{V_{n,in}^2}}{4kTR_S} \\ &= 6.1\,(= 7.84\text{ dB}) \end{aligned}$$

The double-sideband NF is 3 dB less.

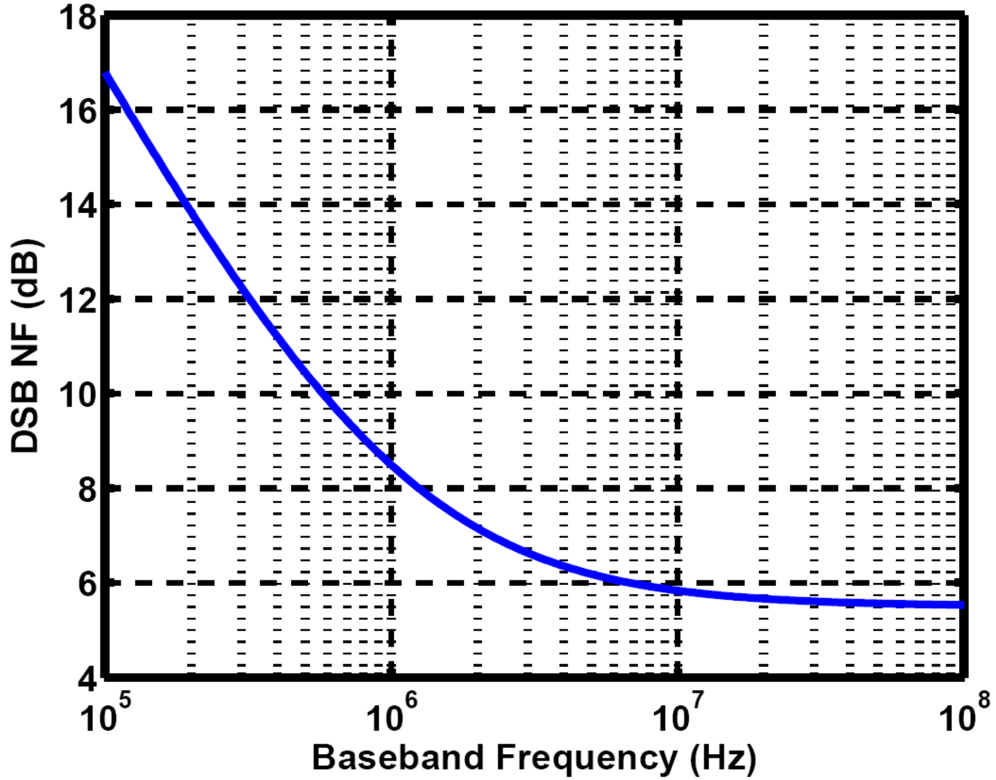
Mixer Design – III

- Simulated A_v = 10.3dB, (~ 2dB less than estimated).
- 1dB compression at $V_{in,p}$ = 170mV (-5.28dBm).
- Mixer compresses at i/p or o/p first? Reducing R_D by 5 and scaling the output voltage-swings proportionally, simulation shows that gain drops by 0.5dB only now at $V_{in,p}$ = 170mV → the output port (switches) reach compression first.



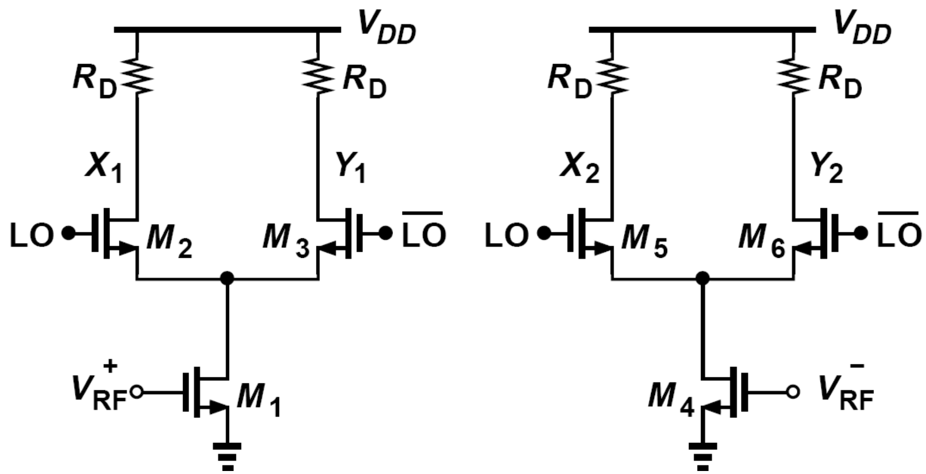
- IIP_3 = 711mV$_p$ (7dBm).
- IIP_3 = 12.3dB + P_{1dB} – perhaps due to higher order non-linearities.

Mixer Design – IV



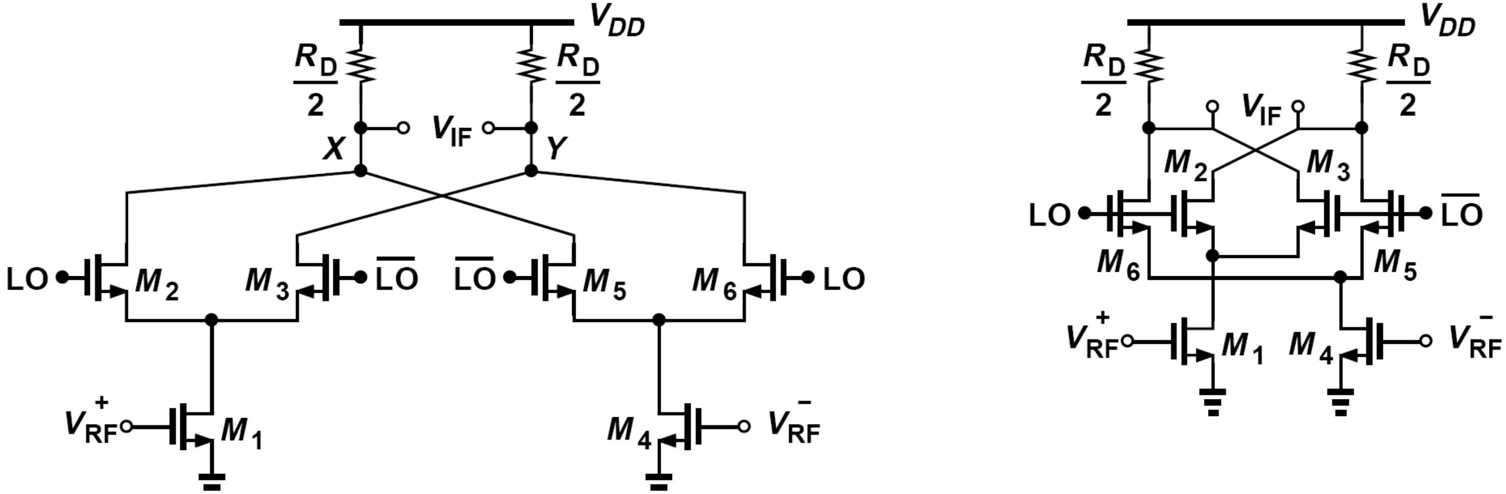
- Flicker noise heavily corrupts the baseband up to several MHz.
- NF @100MHz (thermal) = 5.5dB, ~ 0.7dB higher than predicted.

Double-Balanced Mixer (Gilbert)



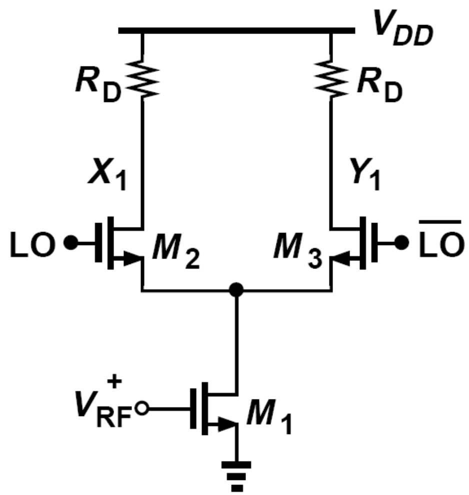
$V_{X1} = -V_{Y1} = -V_{X2} = V_{Y2}$

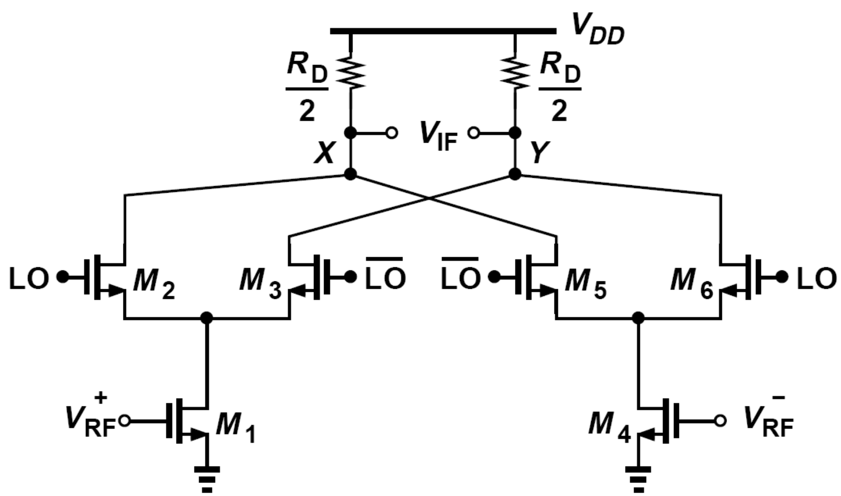
→ Short X1-Y2 and X2-Y1



I_{bias} of DBM = 2x I_{bias} of SBM

LO-IF Feedthrough: DBM vs. SBM





$I_{DC} + IRFcos\omega_{RF}t$ $\quad$ $I_{DC} + IRFcos\omega_{RF}t$ $\quad$ $I_{DC} - IRFcos\omega_{RF}t$

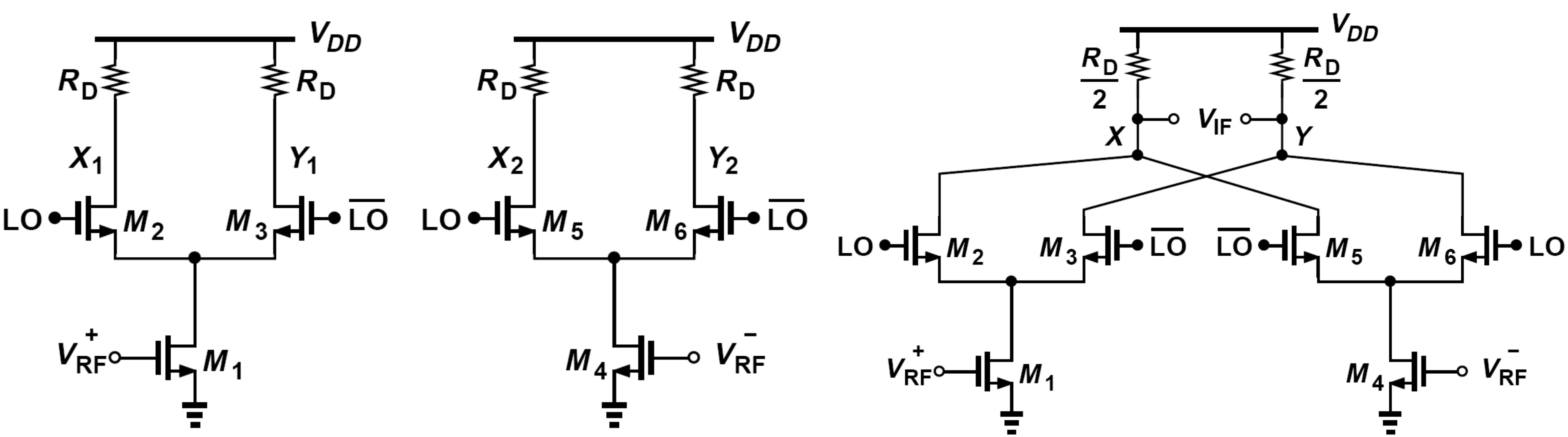
- SBM has LO (and its harmonics) feedthrough to IF because $I_{out,diff}$ has I_{DC} component mixing with LO (and its harmonics).
- DBM has no LO-IF feedthrough (ideally) because $I_{out,diff}$ has no I_{DC} component mixing with LO (and its harmonics).

Conversion Gain: SBM vs. DBM

Voltage conversion gain of SBM = $(V_{X1} - V_{Y1})/V_{RF}^{+}$.
But, shorting Y_2 to X_1, and X_2 to Y_1, these node voltages remain unchanged.
Differential voltage conversion gain of DBM is given by

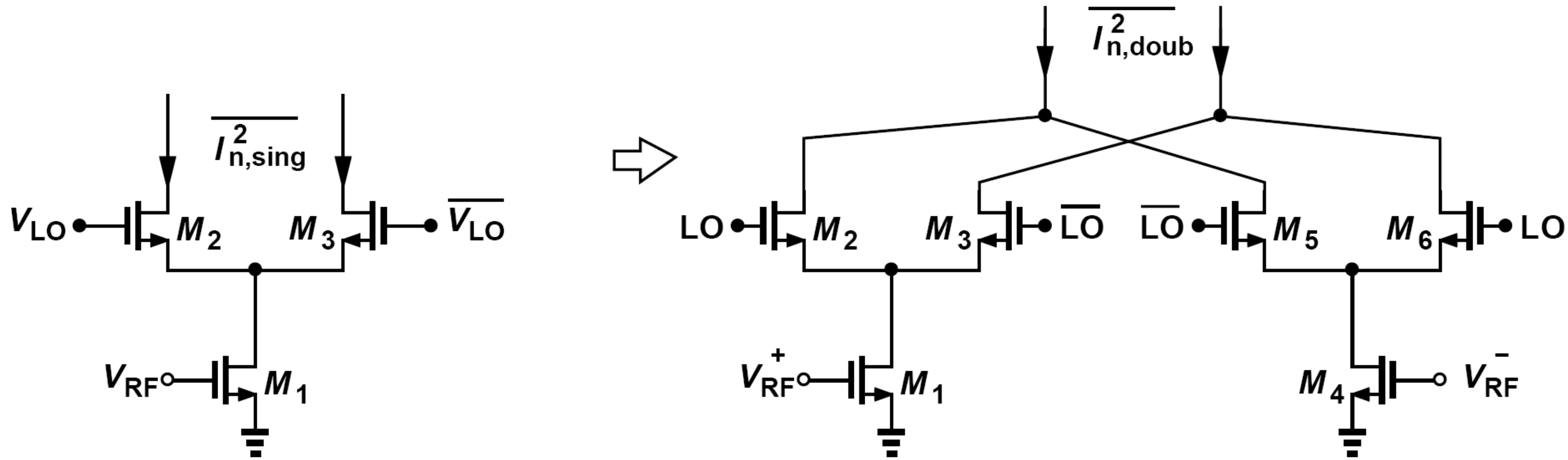
$$\frac{V_X - V_Y}{V_{RF}^{+} - V_{RF}^{-}} = \frac{V_{X1} - V_{Y1}}{2V_{RF}^{+}}$$

= ½ of SBM counterpart.
Cause: limited voltage headroom disallows a load resistance of R_D



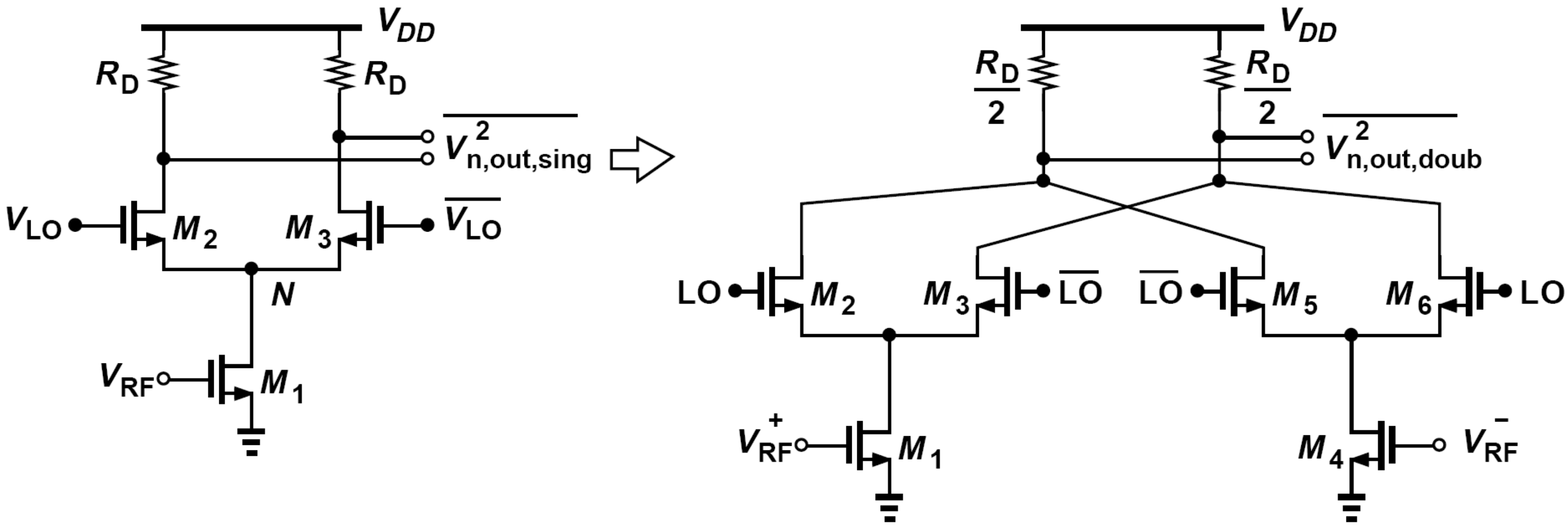
Noise: SBM vs. DBM

Assuming I_{bias} of DBM = 2x I_{bias} of SBM



$$\overline{I^2_{n,doub}} = 2\overline{I^2_{n,sing}}$$

Noise: SBM vs. DBM



O/P noise voltages, with load resistors differing by 2X

$$\overline{V^2_{n,out,sing}} = \overline{I^2_{n,sing}}(R_D)^2$$

$$\overline{V^2_{n,out,doub}} = \overline{I^2_{n,doub}}\left(\frac{R_D}{2}\right)^2$$

As voltage conversion gain of DBM = ½ of SBM, input-referred noise voltages are

$$\overline{V^2_{n,in,sing}} = \frac{1}{2}\overline{V^2_{n,in,doub}}$$