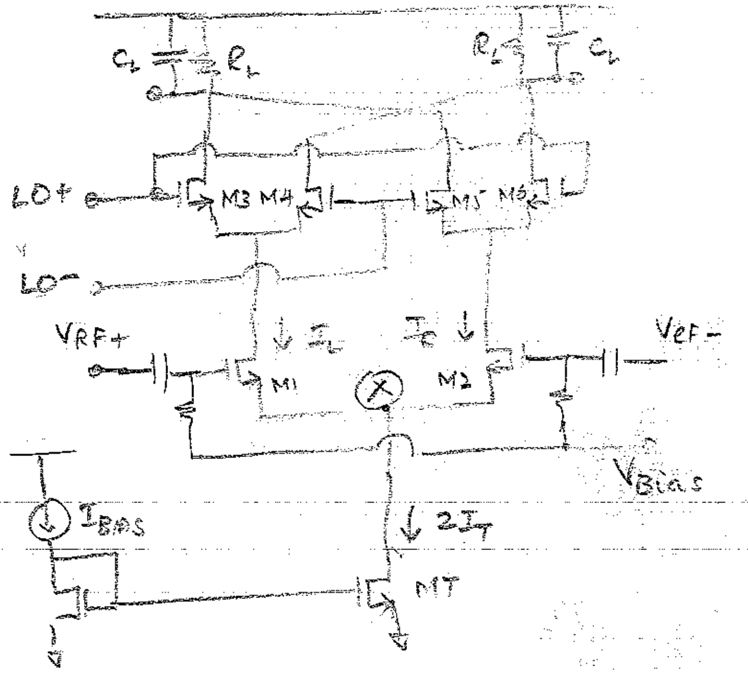
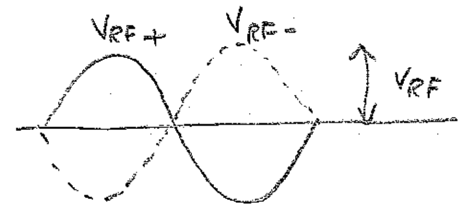
Gilbert-cell double balanced mixer:



- Two cross-connected single balanced mixers.
- The differential pair formed by M1, M2 and MT acts as a simple transconductor.
- M3, M4, M5 and M6 form a Gilbert quad.

- The LO inputs are made large enough to commutate the currents I_L (and I_R) periodically from M3 and M4 (and M5 and M6)
- The LO must not be made just large enough to switch I_L and I_R completely, but not so large as to drive M3/M4/M5 and M6 out of saturation (ie. they should either be in cut-off or saturation)
- For small signal input



$$I_L = I_T + g_m V_{RF} \cos\omega_{RF} t$$

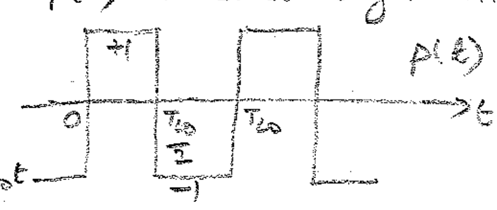
$$I_R = I_T - g_m V_{RF} \cos\omega_{RF} t$$

Differential output current

$$I_{out} = (I_3 + I_5) - (I_4 + I_6) = (I_3 - I_4) + (I_5 - I_6)$$

$$= I_L\, p(t) - I_R\, p(t)$$

$p(t) \rightarrow$ switching function



$$\Rightarrow I_{out} = 2 g_m V_{RF} \cos\omega_{RF} t\, p(t)$$

$$= 2 g_m V_{RF} \cos\omega_{RF} t \cdot \frac{4}{\pi}\left[\sin\omega_{LO} t + \frac{1}{3}\sin 3\omega_{LO} t + \dots\right]$$

$$\Rightarrow I_{out} = \frac{8 g_m V_{RF}}{\pi} \Big[\sin(\omega_{LO} - \omega_{RF})t + \sin(\omega_{LO} + \omega_{RF})t + \frac{1}{3}\sin(3\omega_{LO} - \omega_{RF})t + \frac{1}{3}\sin(3\omega_{LO} + \omega_{RF})t + \cdots \Big]$$

Conversion gain $G_C = \dfrac{8 g_m R_L V_{RF} / \pi}{2 V_{RF}} = \dfrac{4}{\pi} g_m R_L$

(We need to filter off un-desired components)

- LO-to-IF port isolation much better in double-balanced mixers. Port isolation is limited by mismatch between ($M1$ and $M2$) and ($M3$, $M4$, $M5$, $M6$).

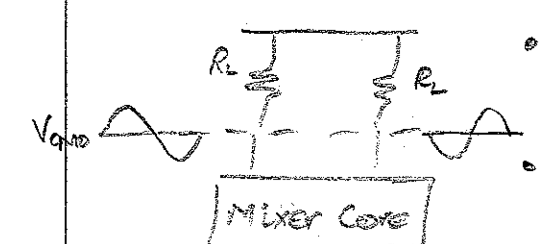
  Sources of random mismatch:
  - ΔW, ΔL, minimized by careful layout, ultimately limited by accuracy of photolithography
  - ΔV_t mismatch
  - C_{ox} mismatch (due to variations in t_{ox} and ε_{ox})
- Can routinely achieve 40+ dB of port isolation.

Noise in current commutating mixers:

Ref: H. Darabi and A. Abidi, "Noise in RF-CMOS Mixers: A Simple Physical Model", IEEE Journal of Solid-State Circuits, Vol. 35, No. 1, Jan 2000, pp. 15 – 25

(a) Load Noise:

(i) Resistive loads:

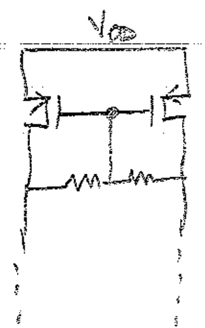


- Thermal noise from resistors competes with output signal.
- No $1/f$ noise in typical resistors available in CMOS (i.e. poly-si or metal)

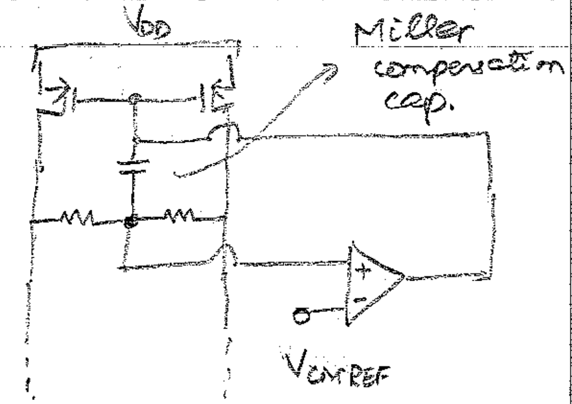
The disadvantage with resistive loads is the loss of voltage headroom: The common-mode voltage at the output is set by the bias current in the mixer core; $V_{CMO} = V_{DD} - I_T R_L$ (total tail current $= 2I_T$)

But $G_C \propto R_L \Rightarrow$ want large R_L for high gain, but V_{CMO} decreases which degrades voltage headroom available for the mixer core:

(ii) <u>PMOS loads:</u>



or



Somewhat better, but headroom still limited by V_{GS} of PMOS devices

$V_{CMO} \leq V_{DD} - V_{GS}$

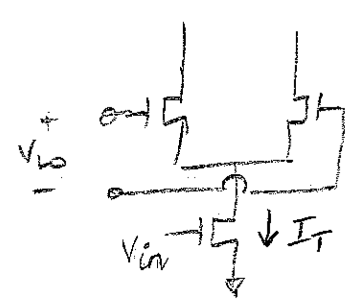
Common-mode feedback loop ⇒ Best headroom

$V_{CMO} \leq V_{DD} - V_{DSAT}$

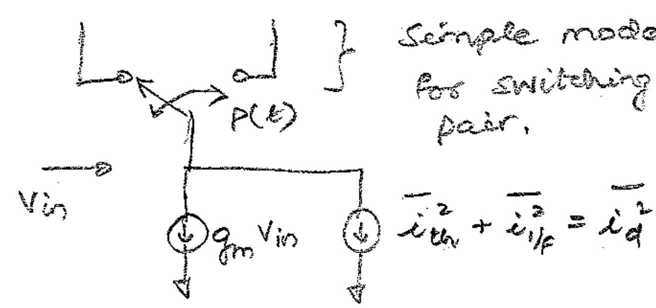
- Thermal noise of PMOS competes with output signal
- $1/f$ noise is important in direct conversion or low-IF receivers. ⇒ use PMOS loads with large area.

(b) <u>Noise from transconductor stage:</u>

(i) consider a single-balanced mixer:



⇒

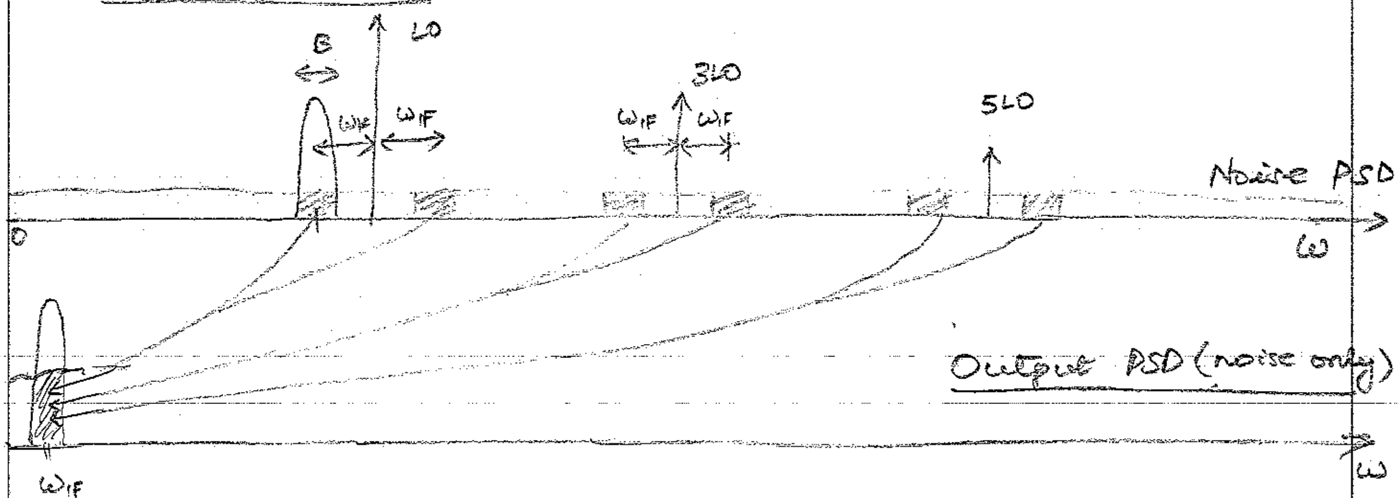


Simple model for switching pair.

- $\overline{i_d^2}$ gets amplitude modulated by $p(t)$

⇒ Need to find noise frequency bands which fold down to the IF output after downconversion

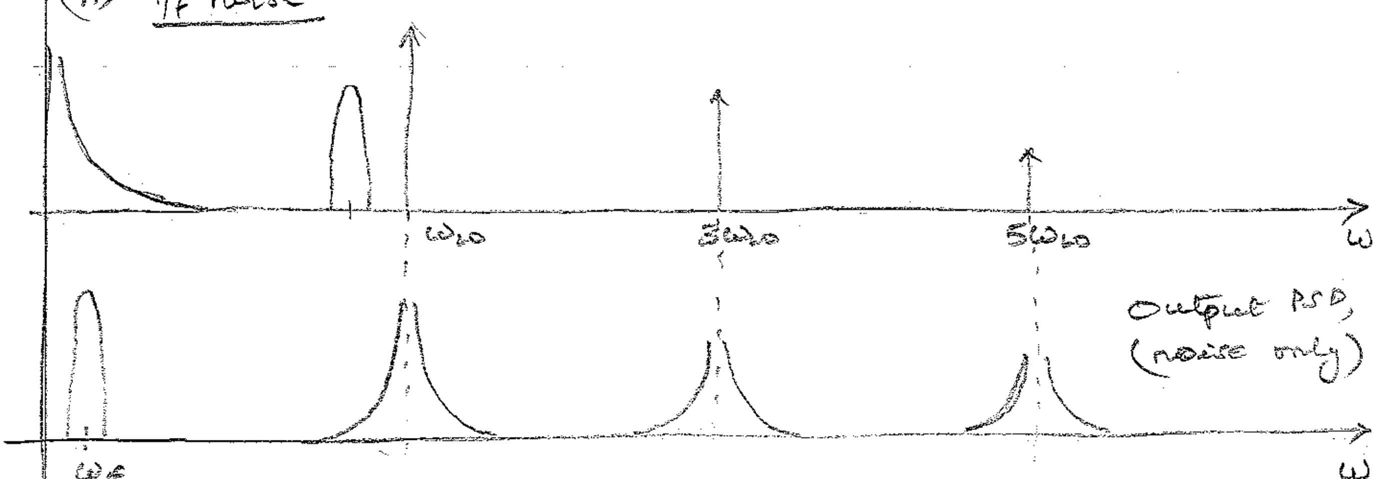
(i) <u>Thermal noise:</u>



Approximate mean-square noise contribution

$$\overline{v^2}_{o,gm} = \underbrace{\frac{4kT\gamma}{g_m}}_{PSD} \times \underbrace{\left(\frac{2}{\pi} g_m R_L\right)^2}_{G_c^2} \times \underbrace{\left(1+\frac{1}{3^2}+\frac{1}{5^2}+\cdots\right)}_{=\pi^2/4}$$

(ii) <u>1/f noise</u>

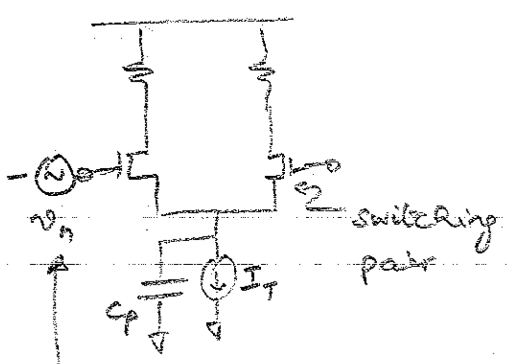


- <u>Downconversion mixer:</u> 1/f noise from g_m-stage <u>does not</u> have a big impact on o/p noise. However, if the switching pair has mismatches, some 1/f noise remains at Baseband and degrades the signal at the output
- <u>Upconversion mixer</u>: Since input signal and 1/f noise both get upconverted, 1/f noise from the g_m-stage can have a <u>significant</u> impact. → manage 1/f noise by using large g_m-devices

(e) <u>Noise from switching pair:</u>

— Relatively complicated to analyze. See above ref. for details

— Switching devices do not contribute noise when tail current is completely switched.

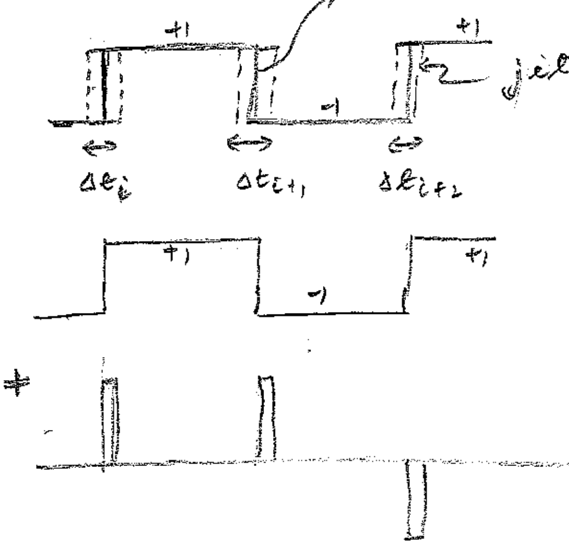


single input-referred noise source captures effect of noise from both devices.

— If C_p is large, switching devices contribute noise to the output even when completely switched.

— Both devices contribute noise directly during the switching phase

In the simple model above, v_n modulates the switching function:



→ Approximate the real switching function as the sum of the ideal switching function and a train of pulses.

- The height of each pulse is constant, but its sign can be +1 or −1
- The width of each pulse $\Delta t = \frac{V_n(t_i)}{S}$ where S is the

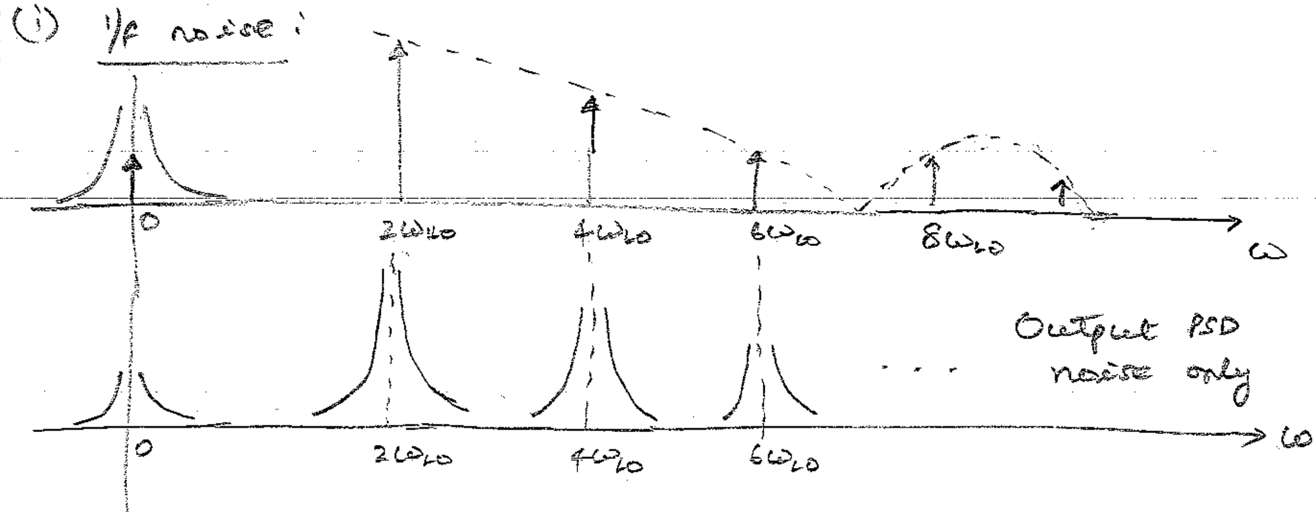
slope of the edge.

- The pulse rate is $2\omega_{LO}$

⇒ Noise from switching devices get modulated by even LO harmonics (ie $2\omega_{LO}$, $4\omega_{LO}$, etc. and possibly DC)

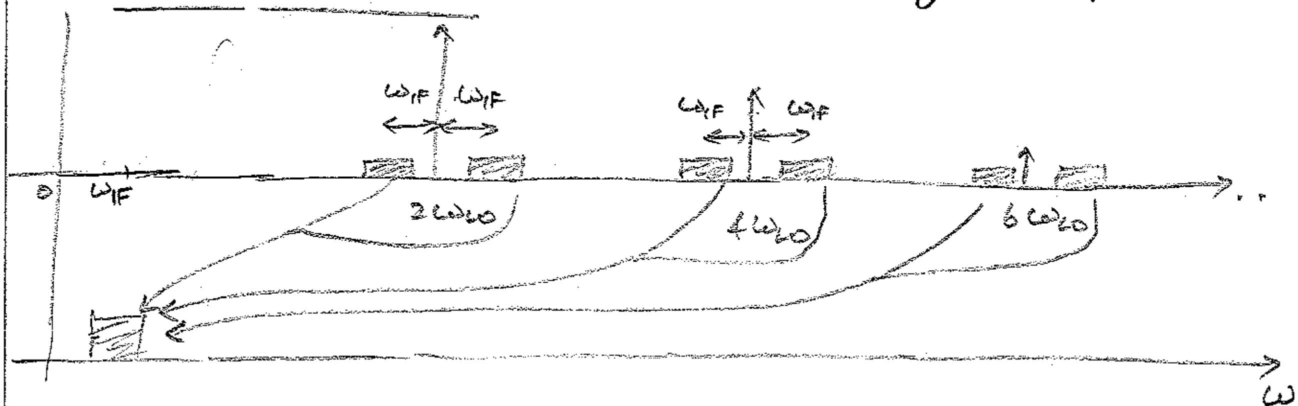
In cartoon form:

(i) <u>1/f noise</u>:



- If C_p is neglected, 1/f noise can be minimized by extremely fast edge rates.
  ⇒ Want small LO devices along with square-wave LO drive.

(ii) <u>Thermal noise</u>: Also modulated by $2\omega_{LO}$ pulse train



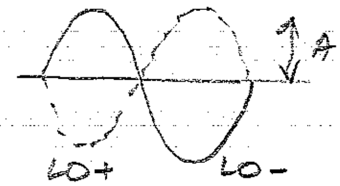
Mean-square voltage noise at output due to thermal noise in switches is:

$$\overline{v^2}_{oSW,th} = 4kT\gamma \frac{I_T}{\pi A}$$

where I_T = tail (bias current)

A = amplitude of LO →

(assuming sine-wave LO)



⇒ Thermal noise contribution from switching devices does not depend on their size
→ depends only on bias current & LO amplitude.

⇒ Want large LO drive.

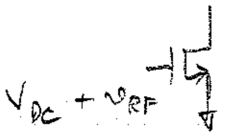
Total noise for single-balanced mixer with resistive loads:

$$\boxed{\overline{v^2_{n,out}} = 8kTR_L\left(1+\gamma\frac{R_L I}{\pi A}+\gamma\frac{g_m R_L}{2}\right)}$$

Linearity of Gilbert mixer:

Overall linearity limited by:

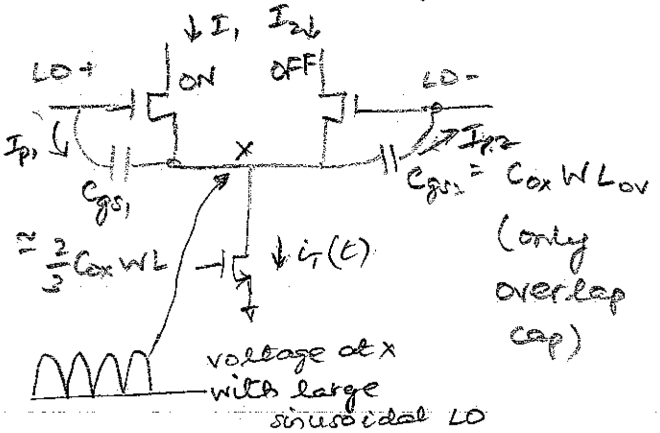
(i) linearity of transconductor:



$$I = I_{DC} + g_{m0} v_{RF} + g_{m1} v_{RF}^2 + g_{m2} v_{RF}^3 + \ldots$$

More on transconductor linearization techniques later.

(ii) Non-linearity due to switches:



- Gate-source caps are vastly different between on & OFF states
- If $C_{gs1} \neq C_{gs2}$ then $I_{p1} \neq I_{p2}$ ⇒ difference in displacement current flows into node x

KCL @ x: $\quad I_1 + I_{p1} = I_{p2} + i_T$

$$\Rightarrow \quad I_1 = i_T + I_{p2} - I_{p1}$$

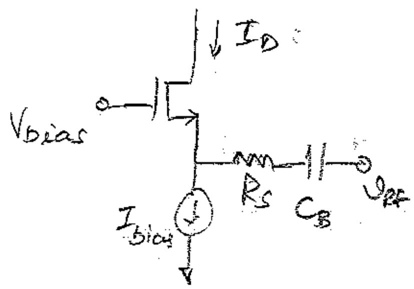
$$\simeq i_T - C_{gs1} \frac{d(V_{LO} - V_x)}{dt}$$

⇒ I_1 is modulated by the LO

→ sharp LO edges & high LO amplitudes hurt linearity!!
(exact opposite of LO drive requirements for low-noise)

More on transconductors:

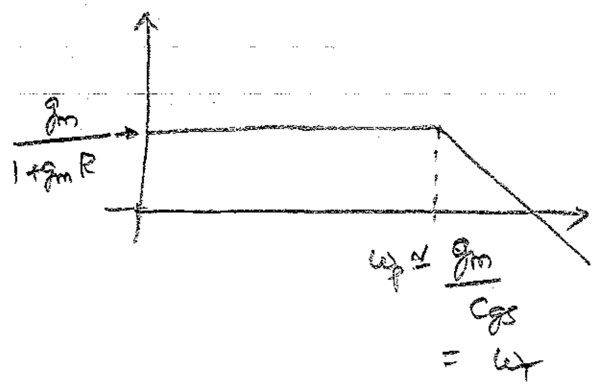
1) Common gate:



- C_B = blocking cap, short circuit at high frequencies
- Neglect back-gate effects

$\Rightarrow$ G_m = effective transconductance

$$G_m = \frac{i_D}{v_{RF}} = \frac{g_m}{1 + g_m R + sC_{gs} R}$$

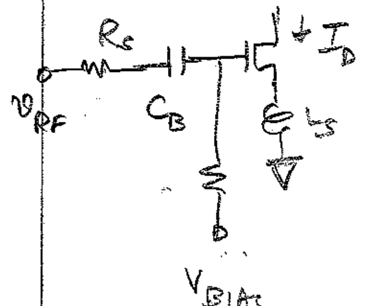


$\Rightarrow$ Very wideband stage

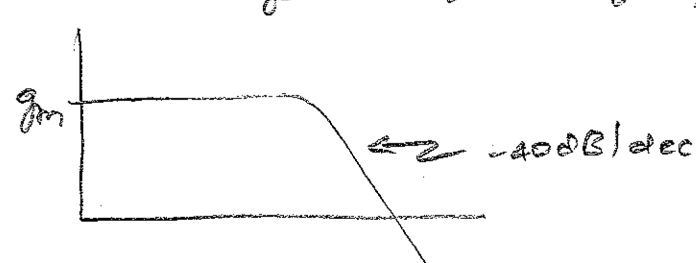
- Useful if matching is required at mixer input
- good linearity because R_s gives source degeneration
- Can use without a degeneration resistor (i.e. $R_s = 0$)

2) Common-source stage:

– Use inductive degeneration for linearity (better than resistor).



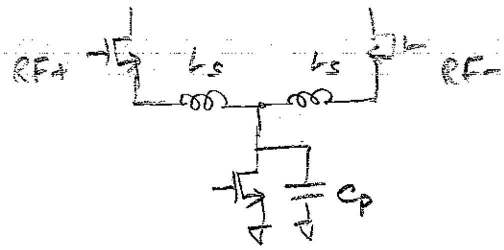
$$G_m = \frac{g_m}{s^2 L_s C_{gs} + s(g_m L_s + C_{gs} R_s) + 1}$$



- Some high-frequency filtering at input.
- Matching to 50Ω not required if input comes from on-chip LNA

3) Differential transconductors:

(i)



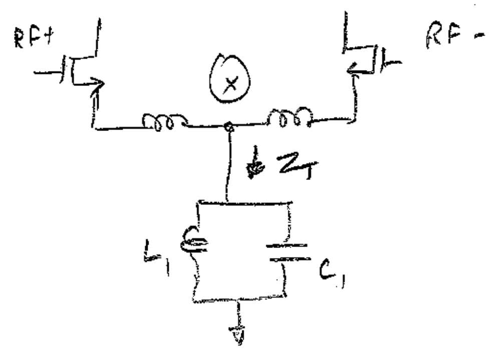
- Fully differential g_m-stage, degeneration optional.
- Good CMRR at low frequencies
- CMRR becomes poor at high frequencies due to C_p.

- Large-signal transfer characteristic is (for $L_s = 0$)

$$I_d = g_m v_{id}\sqrt{1-\left(\frac{v_{id}}{2V_{dsat}}\right)^2} \equiv v_{id}\left[g_{m0} + g_{m1}v_{id} + g_{m3}v_{id}^3 + g_{m5}v_{id}^5 + \dots\right]$$

- no even harmonics
- significant 3rd order non-linearity
  - ⇒ 3rd. order intermodulation
- tail current source uses up voltage headroom.

(ii)

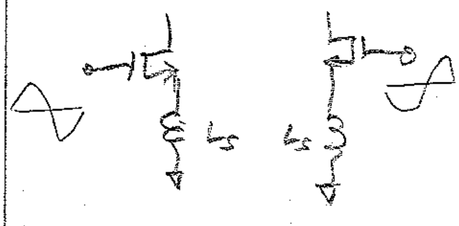


- Voltage at node x has a number of even order harmonics (for a balanced input)
- For high CMRR, want high Z_T at some frequency of interest (e.g $2\omega_{RF}$)

- LC tank effectively creates a <u>zero-headroom AC current source</u>.

(iii) <u>Balanced C-S transconductor</u>:

(or pseudo-differential)



- No current source ⇒ zero CMRR at all frequencies (i.e CMRR is obtained through the perfectly balance)

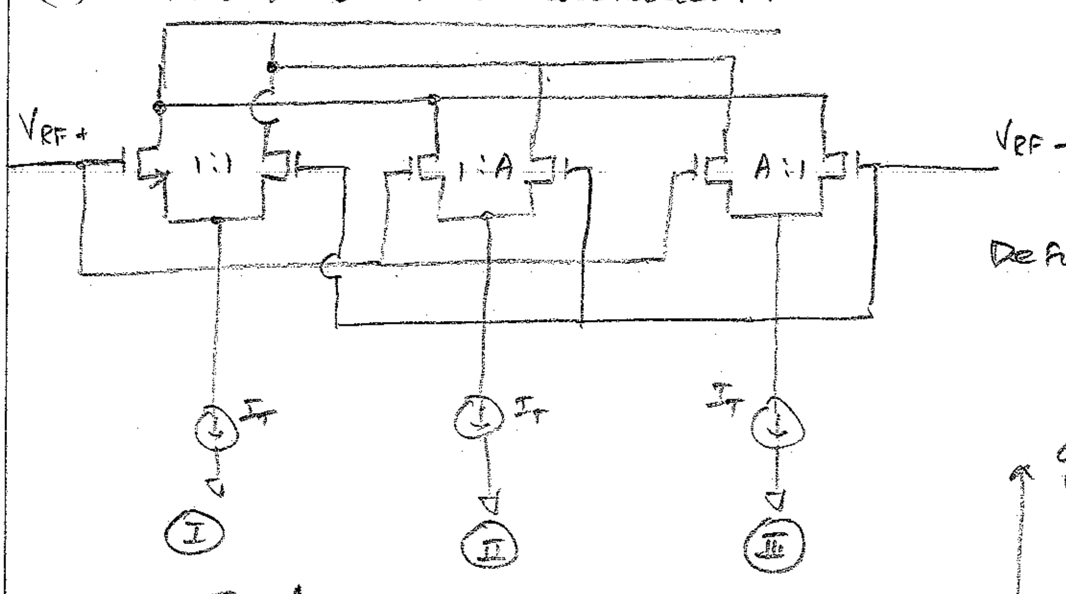
- If we make long-channel assumption,

$$I_d = \frac{K_0'}{2}\left(\frac{W}{L}\right)(V_{GS}-V_T)^2 \Rightarrow \text{no 3rd order component}$$

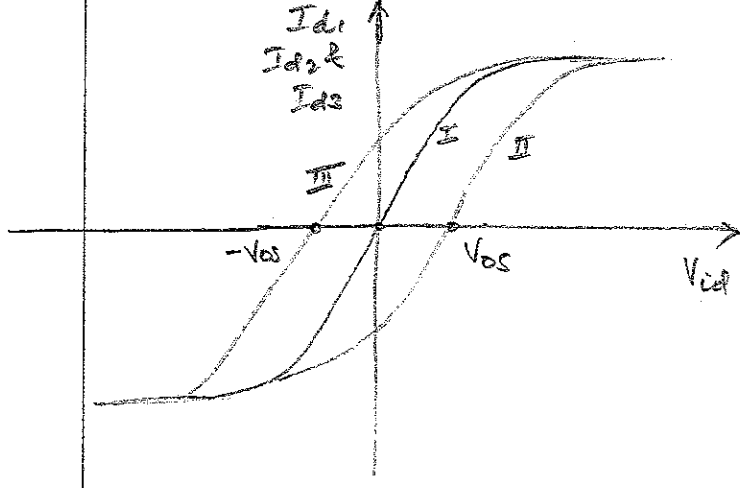
⇒ excellent IIP3

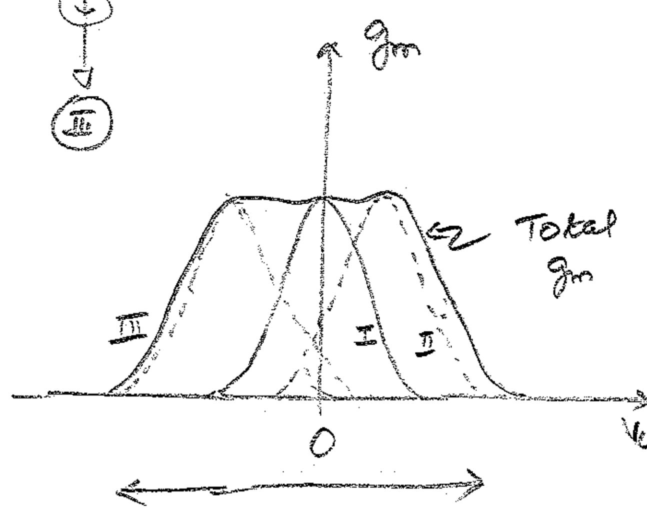
- Highest voltage headroom

(iv) <u>"Multi-tanh" transconductor</u>:



Define $V_{id} = V_{RF+} - V_{RF-}$





$g_{m,eff}$ flat over much wider input signal range ⇒ better linearity