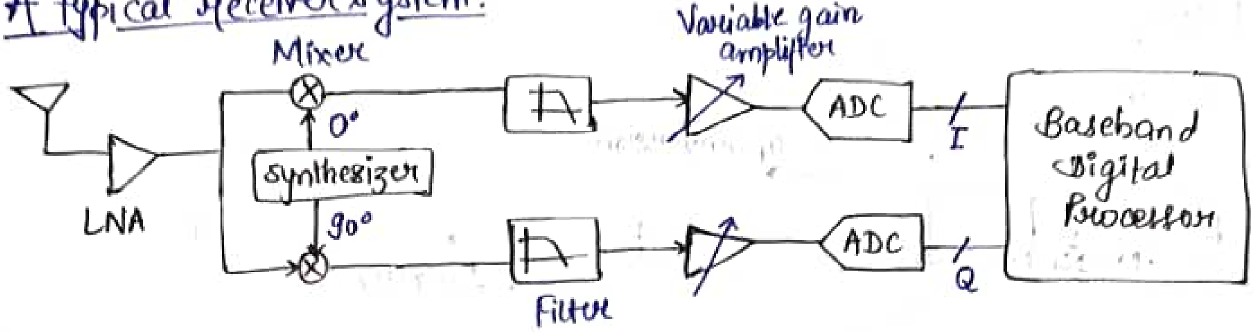
15-01-2025

R.F. and Integrated Circuit

A typical receiver system:



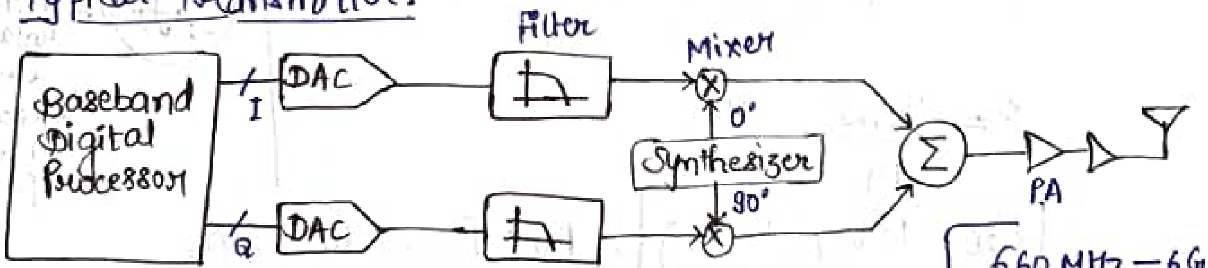
→ CMOS IC design (RFIC)
↳ Technology node: CMOS
↳ Cost effective, size effective.
↳ Extremely good for digital but not much for analog & RF.
↳ Beats other technology in terms of frequency of operation.

Discrete design ⇝ Not part of course
↳ PCB design with RF components

MMIC design ⇝ Not part of course (Monolithic Microwave IC)

Main focus of course: CMOS SoC
→ Synthesiser (LO, PLL,...), Mixers
→ LNA (Low Noise Amplifier)
→ Filters (RF filters)
→ Variable amplifier
→ Digital comm's system.

Typical Transmitter:



660 MHz – 6 GHz : 5G Band
660 M – 3G | 3G – 6 GHz
Midband | High band
2G, 3G : Requires 1W–2W RF out
4G, 5G : Required only 100–200 mW
↳ CMOS PA can be used.

→ Matching in Audio amplifier, etc.: Not considered
↳ Audio freq. ≃ 20 KHz → λ much larger than length of wire used

Need of Matching for High frequency:
(When λ < length of transmission lines)

In RFIC design, on a chip,
lines are very narrow (~ mm) << λsignal (1 GHz ↔ 30 cm = λ)

(phase difference is insignificant b/w the nodes of wire)

↳ Matching not necessary in RFIC design ⇒ saves a lot of time and work

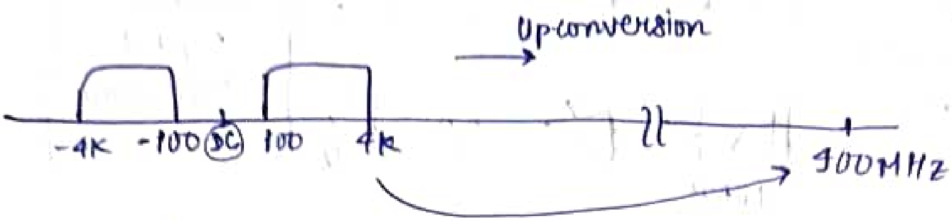
↳ Main difference b/w RFIC & Traditional RF design

17-01-2024

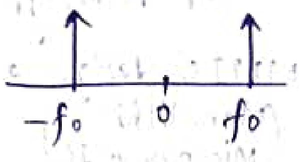
RF ($\sim$100 MHz) : Due ~~to~~ small antenna size, we use high frequency.

for $f = 1$ GHz,

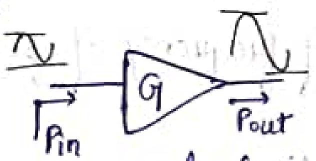
$\lambda \approx 30$cm



$A\cos(2\pi f_0 t)$



dBm Unit:



Voltage gain: G

Power gain: P

$20\log(G)$

In analog circuit, we talk about voltage gain, not power gain.

$$P_{dB} = 10\log P$$

$$= 10\log\left(\frac{V_{out}}{V_{in}}\right)^2$$

$$= 20\log G$$

$f \rightarrow$ vector rotating both dirn

$\rightarrow$ Reason for +ve & -ve frequencies

$$P = \frac{V_{rms}^2}{R_L} = \frac{V_{pk}^2}{2R_L}$$

$$\boxed{P_{dB} = 10\log\left(\frac{P_W}{1\text{mW}}\right)}$$

, P_W: Power in Watts

1 mW $\Leftrightarrow$ 0 dBm

1 W $\Leftrightarrow$ +30 dBm

} Transmitter design

1 μW $\Leftrightarrow$ -30 dBm

1 nW $\Leftrightarrow$ -60 dBm

1 pW $\Leftrightarrow$ -90 dBm

1 fW $\Leftrightarrow$ -120 dBm

} Receiver design

$\downarrow$ noise

Transmit Power:

For $R_L = 50\,\Omega$

$1\,mW \longrightarrow V_{PK} = 316\,mV$

$100\,mW \longrightarrow V_{PK} = 3.16\,V$

$1\,W \longrightarrow V_{PK} = 10\,V$

$$\left[\begin{aligned} V_{PK}^2 &= 1\,mW \times 50 \times 2 \\ &= 100 \times 10^{-3} \\ &= 10^{-1} \\ \Rightarrow V_{PK} &= 0.316\,V \\ &= 316\,mV \end{aligned}\right]$$

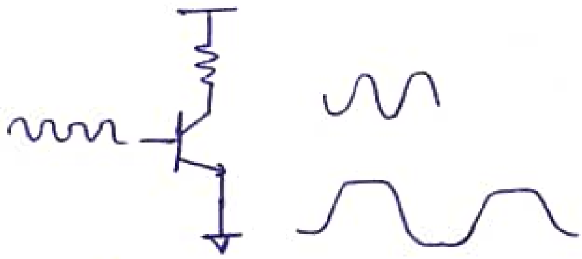
CMOS V_{DD}:

65 nm	1.2 V
45 nm	1.0 V
28 nm	0.91
16 nm	0.0109

$\Rightarrow$ Transmit $V_{PK} > V_{DD}$

$\hookrightarrow$ can push the device to non-linearity when V_{pk} becomes comparable to V_{DD}.

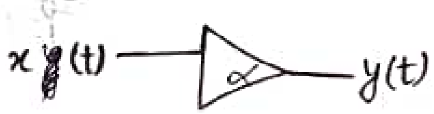
CE Amplifier:



$\longrightarrow$ clipping / distortion

Every amplifier exhibits non-linear behaviour.

Polynomial Model of Non-Linearity



$y(t) = \alpha\, x(t)$ [Linear assumption]

$y(t) = \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t) + \ldots \longrightarrow$ Memoryless polynomial

$\hookrightarrow$ Good model for memoryless amplifiers

$x(t) = A\cos\omega_0 t$

$$y(t) = \alpha_1 A\cos\omega_0 t + \alpha_2 A^2 \cos^2\omega_0 t + \alpha_3 A^3 \cos^3\omega_0 t + \ldots$$

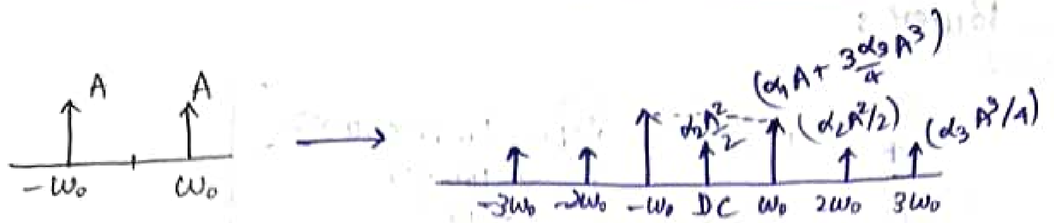
$$= \alpha_1 A\cos\omega_0 t + \alpha_2 A^2\left(\frac{1+\cos 2\omega_0 t}{2}\right) + \alpha_3 A^3 (3\cos\omega_0 t + \cos 3\omega_0 t)\frac{1}{4}$$

$$= \alpha_1 A\cos\omega_0 t + \frac{3}{4}\alpha_3 A^3 \cos\omega_0 t$$

$+ \frac{\alpha_2 A^2}{2}\cos 2\omega_0 t$ [2nd harmonic]

$+ \frac{\alpha_3 A^3}{4}\cos 3\omega_0 t$ [3rd harmonic]

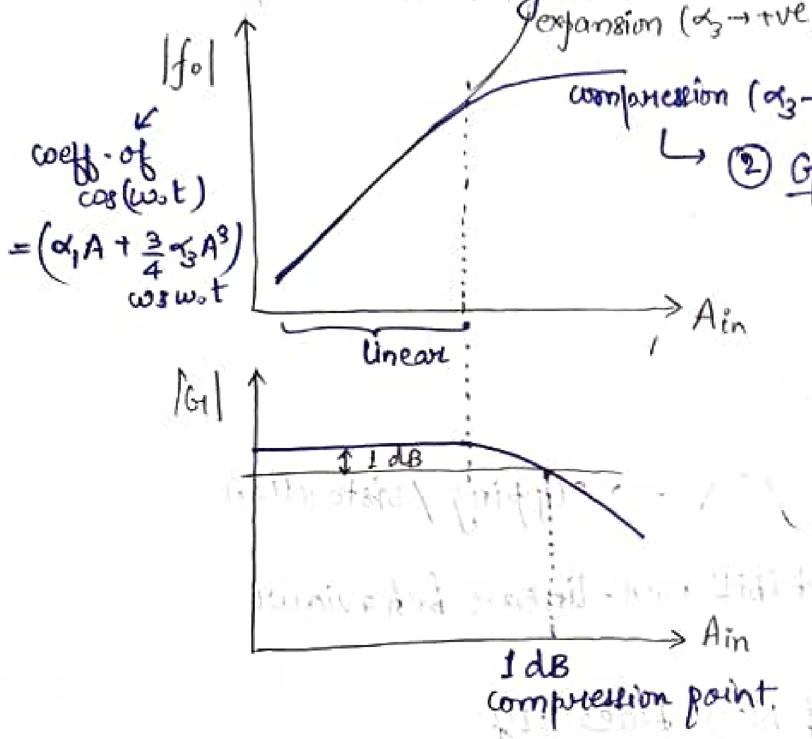
$+ \alpha_2 A^2/2$ [DC]



Effects:

① Generation of harmonics: Can lead to o/p signal in freq. out of band.

f_0: $\underbrace{\alpha_1 A \cos \omega_0 t}_{\text{desired}} + \underbrace{\frac{3}{4}\alpha_3 A^3 \cos \omega_0 t}_{\text{Extra gain}}$ [α_3: usually $\ominus$ve]



② Gain Compression

21-01-2025

1-dB Compression Point:

$y = \alpha_1 x + \alpha_2 x^2 + \alpha_3 x^3 + \ldots$

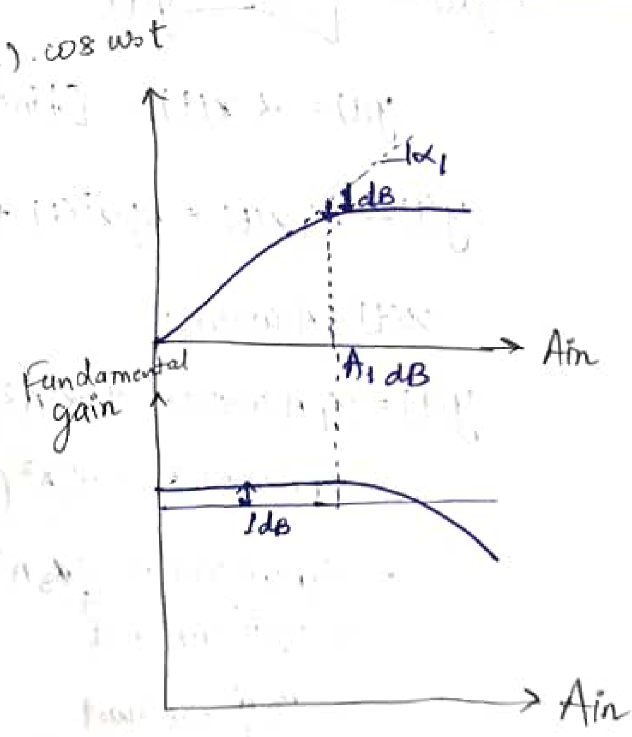
$x[t] = A_{in} \cos \omega_0 t$

Fundamental term:

$\left[\alpha_1 A_{in} + \frac{3}{4}\alpha_3 A_{in}^3\right] \cos \omega_0 t$

Typically $\alpha_3 < 0$

$\approx \alpha_1 A_{in}$ (linear)



$$V_{out}\Big|_{\omega_0} = \left[\alpha_1 A_{in} + \frac{3}{4}\alpha_3 A_{in}^3\right]; \quad V_{in} = A_{in}\cos\omega_0 t$$

$$\text{Gain} = \alpha_1 + \frac{3}{4}\alpha_3 A_{in}^2$$

$\dfrac{V_{out}|_{\omega_0}}{V_{in}}$ (at 1 dB compression)

$$= 20\log\left(\alpha_1 + \frac{3}{4}\alpha_3 A_{1dB}^2\right) = 20\log\alpha_1 - 1\text{ dB}$$

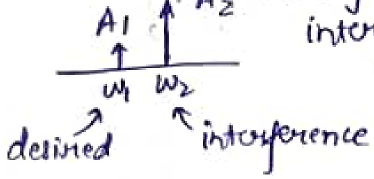
$$\Rightarrow 1\text{dB} = 20\log\left(\frac{\alpha_1}{\alpha_1 + \frac{3}{4}\alpha_3 A_{1dB}^2}\right)$$

$$A_{1dB} = \sqrt{0.145\left|\frac{\alpha_1}{\alpha_3}\right|}$$

(input-referred 1dB) → point at which gain reduces by factor of 10%.

$x_1(t) = A_1\cos\omega_1 t$; $A_2(t) = A_2\cos\omega_2 t$

$A_2 >> A_1$



A_2 → very large interfering signal (blocker)

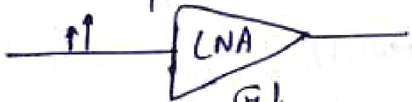
ω_1, ω_2 : very close
↳ cannot be filtered

$$V_{out} = \alpha_1(A_1\cos\omega_1 t + A_2\cos\omega_2 t) + \alpha_3(A_1\cos\omega_1 t + A_2\cos\omega_2 t)^3$$

frequencies at fundamental:

$$\approx \alpha_1(A_1\cos\omega_1 t + A_2\cos\omega_2 t) + \frac{3}{4}\alpha_3 A_2^3\cos\omega_2 t$$

Due to interference,



$G\downarrow$
$\Downarrow$
$SNR\downarrow$ $\Rightarrow$ $\boxed{\text{Jamming}}$

↳ Receiver Desensitization

Cross Modulation

$x_1(t) = A_1 \cos\omega_1 t$

$x_2(t) = A_2 \, (1 + \underset{\text{mod. index}}{m} \underbrace{\cos\omega_m t}_{\text{message signal}}) \cos\omega_2 t$ [Amplitude Modulated] ↳ interferer (strong)

$(A_1 \ll A_2)$ $\qquad (\omega_2 - \omega_m)\;(\omega_2 + \omega_m)$

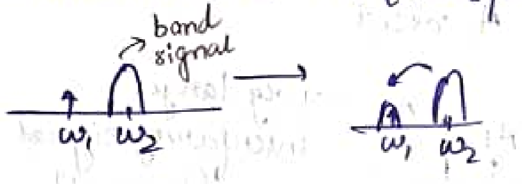
$y(t) = \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t) + \dots$

fundamental term:

$y(t) = \alpha_1 (x_1(t) + x_2(t)) + \frac{3}{4}\alpha_3 \left(x_1(t) + x_2(t) \right)^3$

$\cancel{x_1^3(t)} + \underbrace{x_2^3(t)}_{\approx 0} + \underbrace{3\cancel{x_1^2(t)x_2(t)}}_{\approx 0} + \underbrace{3x_1(t)x_2^2(t)}_{\omega_1}$

Modulation @ ω_2 → transferred to ω_1 ⇝ Undesired



→ Cross Modulation

OFDM

[22-01-2025]

Intermodulation

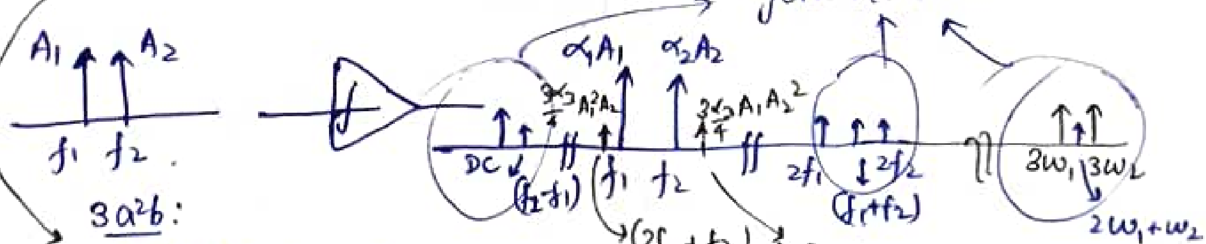
$x(t) = A_1 \cos\omega_1 t + A_2 \cos\omega_2 t$

$y(t) = \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t) + \dots$

$= \alpha_1 (A_1 \cos\omega_1 t + A_2 \cos\omega_2 t)$

$+ \alpha_2 (\underset{(DC + 2\omega_1)}{A_1 \cos\omega_1 t} + \underset{(DC + 2\omega_2)}{A_2 \cos\omega_2 t})^2$ ⇝ $(\omega_1 \pm \omega_2)$

$+ \alpha_3 (\underset{(\omega_1, 3\omega_1)}{A_1 \cos\omega_1 t} + \underset{(\omega_2, 3\omega_2)}{A_2 \cos\omega_2 t})^3$ $\quad [a^3 + 3a^2 b + 3ab^2 + b^3]$



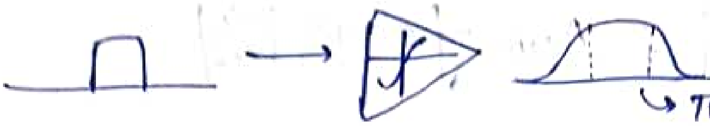
$3a^2 b$:

$3\alpha_3 A_1^2 \cos^2\omega_1 t \cdot A_2 \cos\omega_2 t$

$= 3\alpha_3 A_1^2 A_2 \, (2\cos 2\omega_1 t - 1) \cdot \cos\omega_2 t$

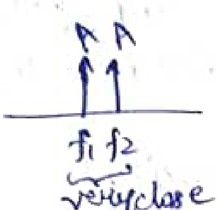
$= \frac{3}{4}\alpha_3 A_1^2 A_2 \left[\cos(2\omega_1 \pm \omega_2)t + 2\cos\omega_2 t \right]$

$\left.\begin{array}{l} 2f_1 - f_2 \\ 2f_2 - f_1 \end{array}\right\}$ IM (Intermodulation) tones

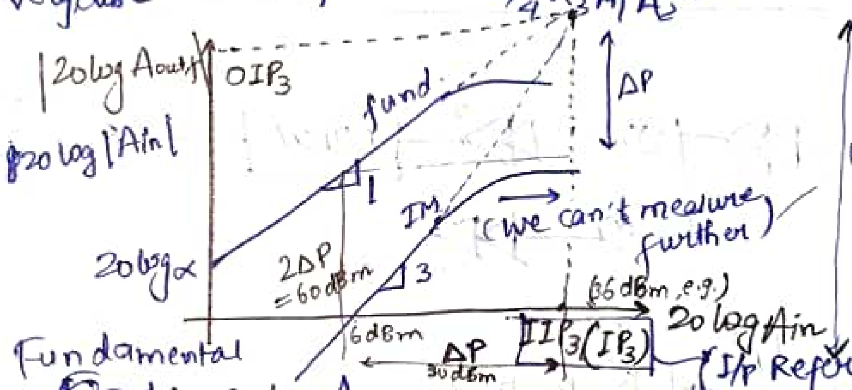


$f_1 = 1\ GHz$
$f_2 = 1.01\ GHz$
$2f_1 - f_2 = 990\ MHz$
$2f_2 - f_1 = 1.02\ GHz$

"<u>Two tone test</u>": Give two tones as i/p to amplifier and measure IM tones.



IM: $\frac{3}{4}\alpha_3 A_1^2 A_2 \longrightarrow \frac{3}{4}\alpha_3 A^3$
$\frac{3}{4}\alpha_3 A_1 A_2^2 \nearrow$



→ We can't go around this
↳ Burns the LNA
→ Help us compare 2 different products
→ Interpolated point
→ cannot be measured directly
→ Not physically achievable

(I/P Refered 3rd order intercept point)

Fundamental: $20\log A_{out,f} = \underbrace{20\log A_{in}}_{slope = 1} + 20\log\alpha$

IM: $20\log A_{out,f} = 20\log\left(\left|\frac{3}{4}\alpha_3 A^3\right|\right)$ ~~+ 20 log Ain~~

$= 60\log\left(\sqrt[3]{\tfrac{3}{4}\alpha_3}\,A\right) + 60\log A_{in}$ ~~+ 20 log Ain~~

<u>IM Suppression</u>: $20\log\left(\frac{A_f}{A_{IM}}\right)$

$\left(20\log A_f - 20\log A_{IM}\right)$
↳ slope of 2

IM suppression $= 2\left(20\log IIP_3 - 20\log A_{in}\right)$

$\boxed{24-01-2025}$

Eg
$\left.\begin{array}{l} IIP_3 = +30\ dBm \\ P_{in} = -20\ dBm \end{array}\right\}$ 50 dB
$P_{IM} = -120\ dBm$

$\left.\begin{array}{l} IIP_3 = +20\ dBm \\ P_{in} = -20\ dBm \end{array}\right\}$ 40 dB.
$P_{IM} = -100\ dBm$

→ $IIP_3 \uparrow \Rightarrow$ Linearity $\uparrow$ (better)

Feeding IIP_3, burns the device.

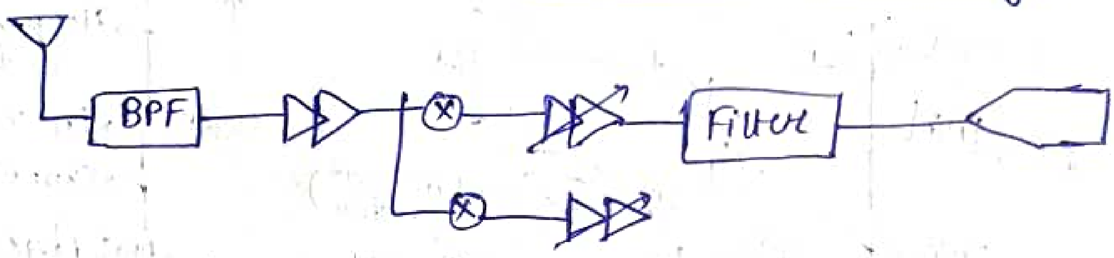
$\alpha_1 A_{in} = \frac{9}{4}\alpha_3 A_{in}^3$ (linear & 3rd order IM tone terms become equal at IIP_3 point)

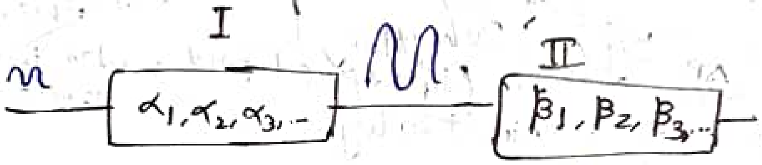
$$\Rightarrow \boxed{A_{IIP_3} = \sqrt{\frac{4}{3}\left|\frac{\alpha_1}{\alpha_3}\right|}} \quad ; \quad A_{1dB} = \sqrt{0.145\left|\frac{\alpha_1}{\alpha_3}\right|}$$

$$20\log\left(\frac{A_{IIP_3}}{A_{1dB}}\right) = 9.6\,dB \approx 10\,dB$$

⇝ IIP_3 point is ~10 dB higher than 1-dB comp. point. (for no other non-linearity)

Non-linear terms cascaded (Cascaded Non-linearity)



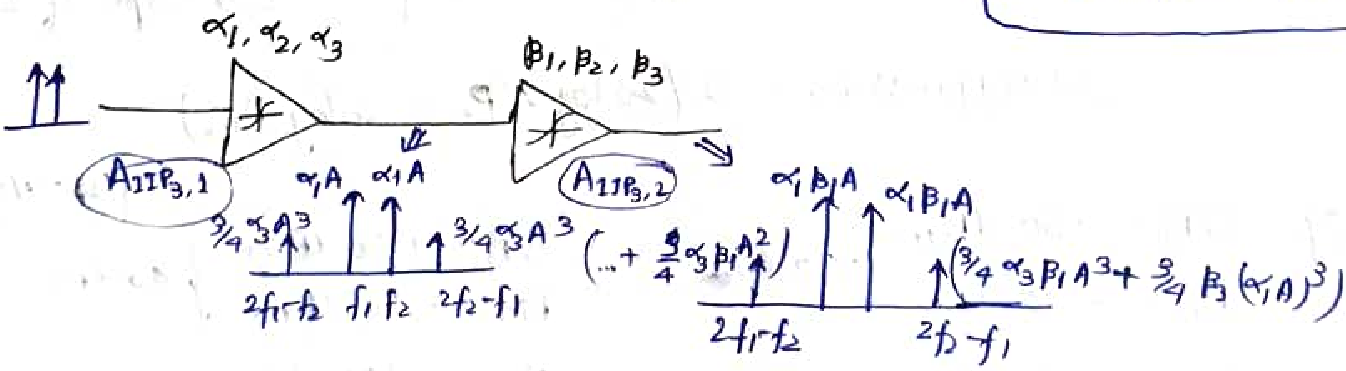


↳ Large signal to begin with
↳ Produce more non-linearity

For same (II),

(I) : $G_1 > G_2$
will produce more non-linearity

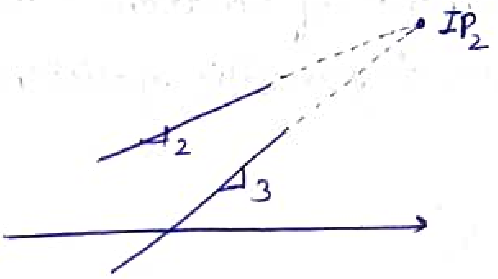
> **Book:** RF Microelectronics 2nd ed. by Behzad Razavi



2^{nd} order distortion:

$$\alpha_2 (A\cos\omega_1 t + A\cos\omega_2 t)^2$$

$$= \alpha_2 \Big[\underbrace{A^2\cos^2\omega_1 t}_{2\omega_1 \ \ DC} + \underbrace{A^2\cos^2\omega_2 t}_{2\omega_2 \ \ DC} + \underbrace{2A^2\cos\omega_1 t\cdot\cos\omega_2 t}_{\omega_1 \pm \omega_2} \Big]$$



$$\rightarrow \boxed{\frac{1}{A_{IIP_{3,total}}^2} = \frac{1}{A_{IIP_{3,1}}^2} + \frac{\alpha_1^2}{A_{IIP_{3,2}}^2}} + \left(\frac{\alpha_1^2 \beta_1^2}{A_{IIP_{3,3}}^2}\right)$$

28-01-2025

Noise

↳ Random signal generated by any electronic device.

Thermal Noise:

↳ Anything having non-zero real resistance causes noise, due to brownian movement of electrons (above 0 K).

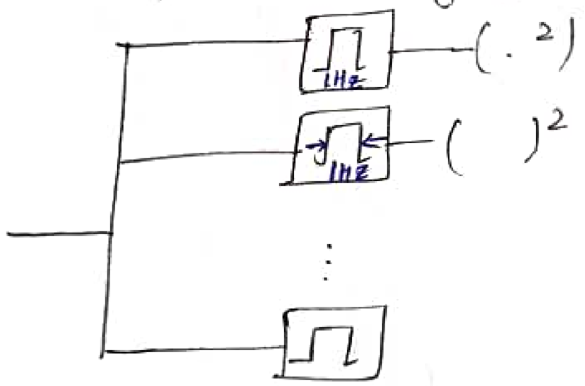
→ Inductors & capacitors, due to parasitic resistances, generate noise.

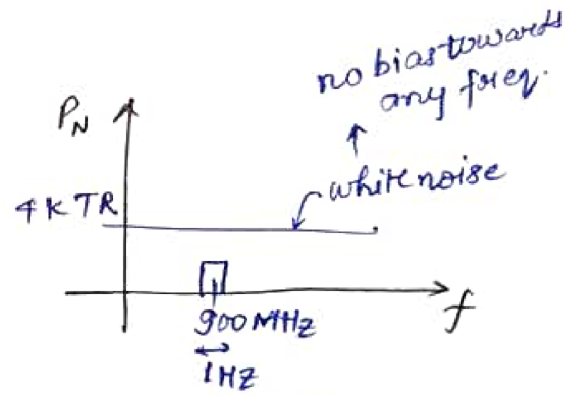
Average (noise), $\mu = 0$

Std. deviation, or variance, σ^2

↳ gives power in the noise

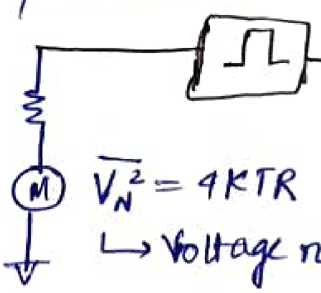
Power Spectral Density (PSD):





↳ PSD represent Power of 1Hz BW at that freq.

→ Resistor produces noise.



Total integrated noise: $4KTR\,|H(f)|^2$

$4KTR\,(f_2 - f_1)$

$\overline{V_N^2} = 4KTR$ ⇝ white noise (V^2/Hz)

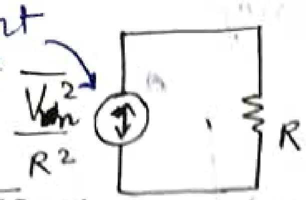
↳ Voltage noise spectral density

$X(f) \longrightarrow \boxed{H(f)} \longrightarrow Y(f) = X(f)\,|H(f)|^2$

↳ Any random variable

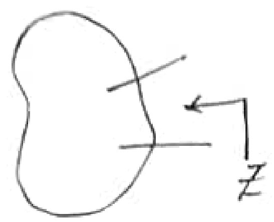
↳ power quantities

Current noise



$\overline{I_n^2} = \frac{4KT}{R}$ (A^2/Hz)

→



Noise through a port of any network, $\overline{V_n^2} = 4KT\, Re\{Z\}$.

→

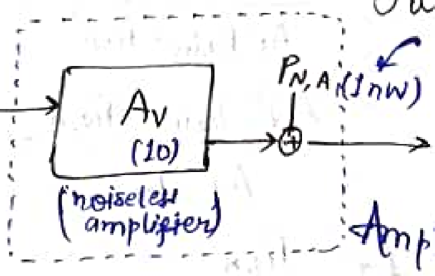
$$P_{rad} = \frac{V_{Tx}^2}{R_{Rad.}}$$ → Radiated power

Noise in the antenna,

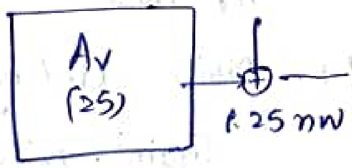
$$\overline{V_{n,\,ant.}^2} = 4KT R_{rad}.$$

Output Referred Noise

→ $P_{s_{in}} + P_{N,in}$



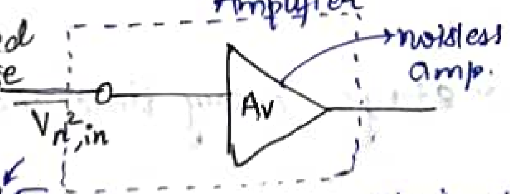
$$\frac{A_v^2 \cdot P_s}{A_v^2 \cdot P_{N,in} + P_{N,A}}$$



→ less degradation of SNR

↳ Signal amplification is much more than the marginal noise added.

Input Referred Noise



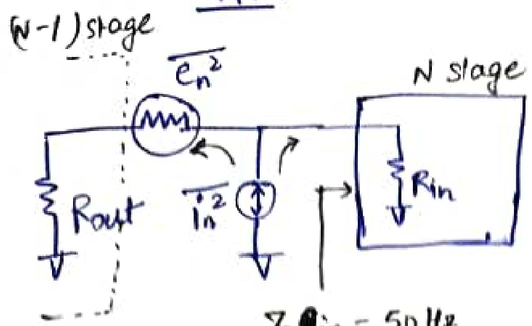
→ Not suitable for RF.

$5nV/\sqrt{Hz}$
or
$25 \times 10^{-18}\ V^2/Hz$ (PSD)

→ Noise is added in series:

$$A_v^2 \cdot P_s + A_v^2 \cdot P_{n,in} + A_v^2 \overline{V_{n,in}^2}$$

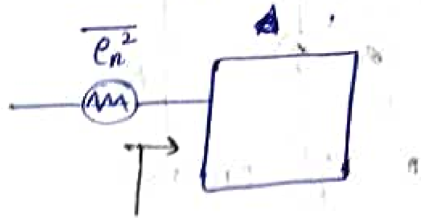
RF:



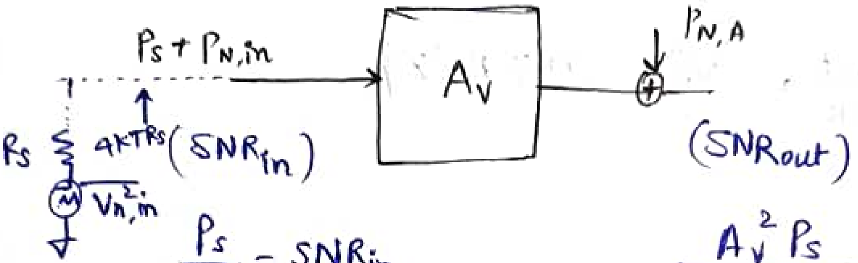
$Z_{in} = 50\Omega$
(not ∞)

↳ Noise voltage as well as noise current : Model

Analog:



$R_{in} = \infty$



$$\frac{P_s}{P_{N,in}} = SNR_{in} \qquad \frac{A_V^2 P_s}{A_V^2 P_{N,in} + P_{N,A}} = SNR_{out}$$

Noise figure, $NF = \dfrac{SNR_{in}}{SNR_{out}}$

$$= \frac{\dfrac{P_s}{P_{N,in}}}{\dfrac{A_V^2 P_s}{A_V^2 P_{N,in} + P_{N,A}}} = \frac{(A_V^2 P_{N,in} + P_{N,A})\cancel{P_s}}{(A_V^2 \cancel{P_s}) \cdot P_{N,in}}$$

$$= \frac{A_V^2 \cdot P_{N,in} + P_{N,A}}{A_V^2 \cdot P_{N,in}} = \frac{\text{Total noise power @ o/p}}{\text{Noise @ o/p due i/p noise}}$$

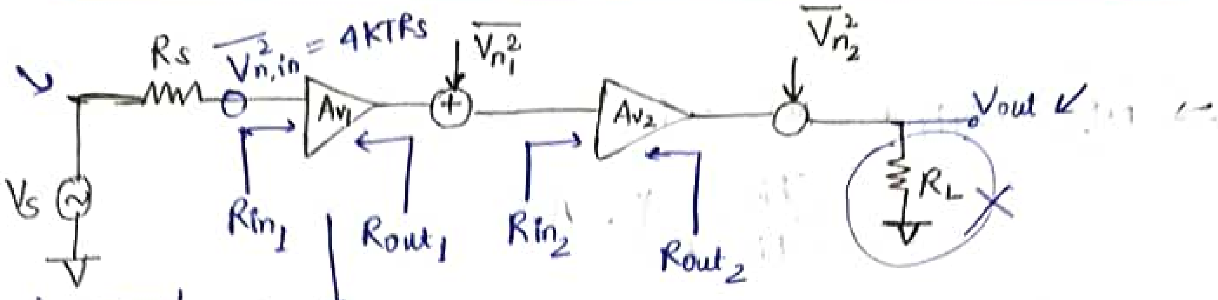
$$NF = 1 + \frac{P_{N,A}}{A_V^2 \cdot P_{N,in}}$$

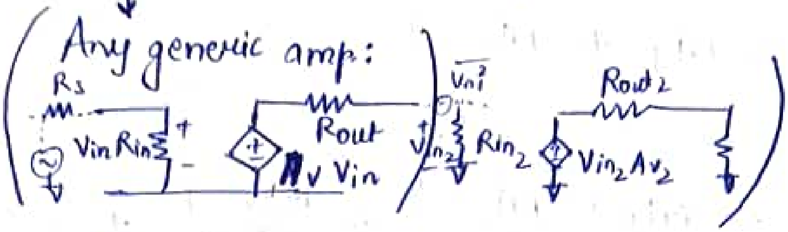
$$= 1 + \frac{\text{Noise added by amplifier}}{\text{Gain}^2 \cdot \text{Noise @ i/p}}$$

$\boxed{29\text{-}01\text{-}2025}$

→ $NF_{dB} = 10 \log\left(\dfrac{SNR_{in}}{SNR_{out}}\right)$

→ SNR degrades stage after stage, every time we add an electrical component.





$$V_{out,s} = V_s \underbrace{\frac{R_{in_1}}{R_{in_1}+R_s}}_{\substack{\text{Impedance}\\ \text{voltage}\\ \text{division}}} \cdot \underbrace{A_{v_1}}_{(\text{Amp.})} \cdot \frac{R_{in_2}}{R_{in_2}+R_{out_1}} \cdot A_{v_2} \cdot \cancel{\left(\frac{R_L}{R_L+R_{out_2}}\right)}$$

A_0 ⇝ overall gain (from ↘ to ↙)

$$NF = 1 + \frac{\text{Noise added}}{\text{Gain}^2 \cdot \overline{V_{n,in}^2}}$$

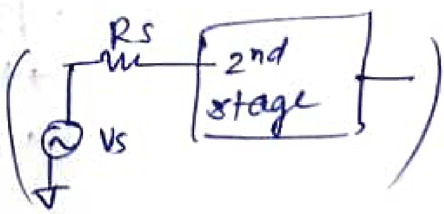
$$\overline{V_{out,n}^2} = 4KTR_s \cdot A_0^2 + \underbrace{\overline{V_{n_1}^2}\left(\frac{R_{in_2}}{R_{in_2}+R_{out_1}}\right)^2 A_{v_2}^2 + \overline{V_{n_2}^2}}_{\text{noise added by the system}}$$

$$\overline{V_{n,added}^2} = \overline{V_{n_1}^2} \cdot \left(\frac{R_{in_2}}{R_{in_2}+R_{out_1}}\right)^2 A_{v_2}^2 + \overline{V_{n_2}^2}$$

$$\therefore NF = 1 + \frac{\overline{V_{n_1}^2}\cancel{\left(\frac{R_{in_2}}{R_{in_2}+R_{out_1}}\right)^2 \cdot A_v^2} + \overline{V_{n_2}^2}}{\left(\frac{R_{in_1}}{R_{in_1}+R_s}\right)^2 A_{v_1}^2 \cancel{\left(\frac{R_{in_2}}{R_{in_2}+R_{out_1}}\right)^2 A_v^2} \cdot 4KTR_s \quad A_0^2 \cdot 4KTR_s}$$

$$\left[\frac{\overline{V_{n_1}^2}}{\left(\frac{R_{in}}{R_{in}+R_s}\right)^2 \cdot A_{v_1}^2 \cdot 4KTR_s} = \underbrace{NF_1}_{\text{for 1st stage}} - 1\right]$$

$$\left[\underbrace{NF_2}_{\text{NF of 2nd stage in isolation}} = 1 + \frac{\overline{V_{n_2}^2}}{\left(\frac{R_{in_2}}{R_{in_2}+R_{out_1}}\right)^2 A_{v_2}^2 \cdot 4KTR_s}\right] \longrightarrow$$



$$\Rightarrow NF_{tot.} = NF_1 + \frac{NF_2 - 1}{\left(\frac{R_{in}}{R_{in}+R_s}\right)^2 \cdot A_{v_1}^2}$$

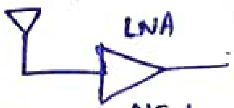
$$= NF_1 + \frac{NF_2 - 1}{A_1}$$

In general,

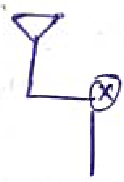
$$\boxed{NF_{total} = \underbrace{NF_1}_{\text{directly appears as NF}} + \frac{NF_2 - 1}{A_1} + \frac{NF_3 - 1}{A_1 A_2} + \dots}$$

(NF ↳ not log ↳ regular scale)

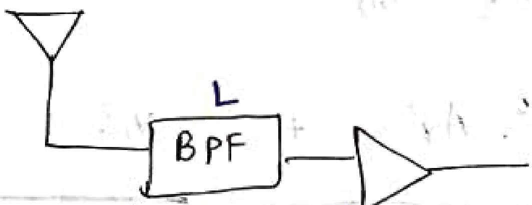
directly appears as NF
↳ Very critical
⇓
Low Noise Amplifier (LNA)



We want { NF↓, A↑ (for 1st stage)



→ Mixer typically not added first
↳ High NF
↳ Added after 2 stages typically



↳ Made of:
SAW / BAW (Surface/Bulk Acoustic Wave)
(not capacitors, T/L)
↳ To avoid interference (avoiding jamming, sensitization, etc.)
↳ For fixed band comms.
↳ Passive filter.
↳ Has loss (=NF)
↳ 1st stage

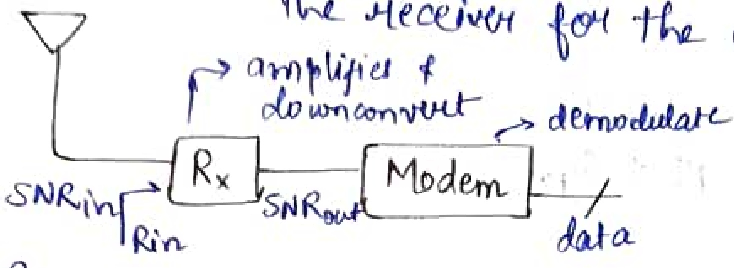
$$NF = \overset{\text{loss element (BPF)}}{NF_L} + (NF_{LNA}-1)\,\overset{\text{Loss} = 1/G_1}{L} + \frac{(NF_{LNA}-1)\,L}{A_{LNA}}$$

Some antennas have attached LNA to avoid loss (as 1st stage).

→ Every loss element (cable, etc.) has to be modelled.

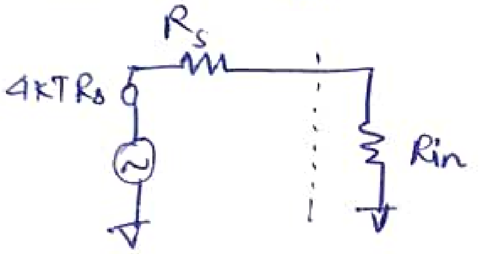
04-02-2025

Sensitivity → Minm power that has to be given to the i/p of the receiver for the communication.



$\frac{P_{Rx}}{P_{N,in}} = SNR_{in}$

0110001 → mod ↓ up



$R_{in} = R_s$

$$P_{N,in} = 4KTR_s \left(\frac{1}{2}\right)^2 \cdot \frac{1}{R_{in}} = KT$$

$$NF = \frac{SNR_{in}}{SNR_{out}} = \frac{P_{Rx}/KT}{SNR_{out}}$$

$$P_{Rx} = KT \cdot NF \cdot SNR_{out}$$

$$P_{Rx,min} = KT \cdot NF \cdot SNR_{min} \cdot \times BW$$ → to get total integrated noise

$$P_{sens,\,dBm} = 10\log\left(\frac{KT}{1mW}\right) + 10\log NF + 10\log SNR_{min} + 10\log BW$$

$$= -174\ dBm/Hz + NF_{dB} + SNR_{min,dB} + BW_{dB}$$

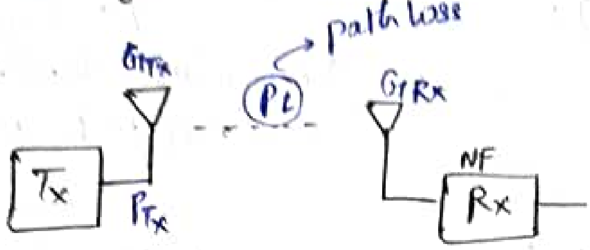
Eg. $NF_{dB} = 7\ dB$

$BW = 100\ KHz$

$SNR_{min} = 10\ dB.$

$$\therefore P_{SNR,dBm} = -174 + 50 + 7 + 10$$
$$= -107\ dBm$$

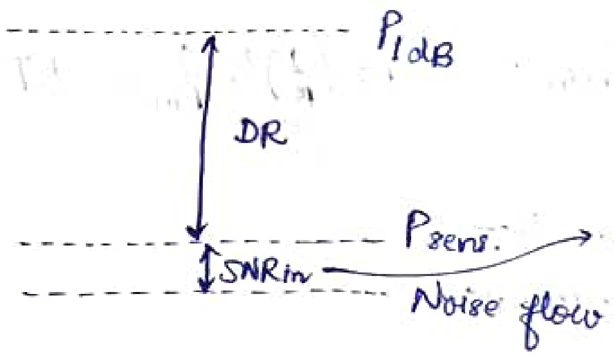
Link Budget



$$P_{\text{rad.}} = P_{Tx} + G_{Tx}$$

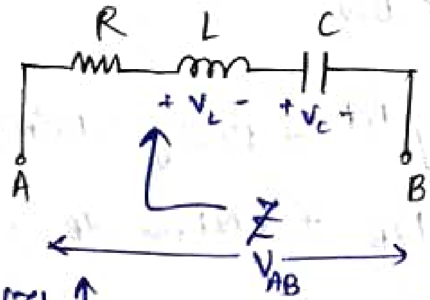
$$P_{Rx} = P_{Tx} + G_{Tx} + \underbrace{PL}_{-ve} + G_{Rx}$$

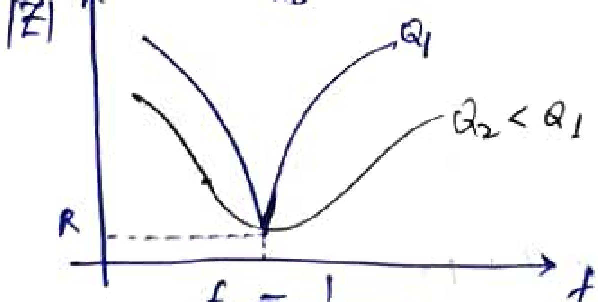
Dynamic Range



$P_{sens.} \rightarrow -174\ dBm/Hz + NF_{dB} + BW_{dB}$

RLC Circuit:





$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

$$|j\omega_o L| = \left|\frac{1}{j\omega_o C}\right|$$

$$V_L \Big|_{\omega=\omega_0} = \frac{V_{AB}}{R} \cdot \omega_0 L$$

$$= V_{AB}\left(\frac{\omega_0 L}{R}\right)$$

$$= V_{AB} \cdot \underbrace{Q}_{\text{"quality factor"}}$$

$$V_C \Big|_{\omega=\omega_0} = \frac{V_{AB}}{R} \cdot \frac{1}{\omega_0 C}$$

$$= V_{AB} \cdot \frac{1}{\omega_0 RC}$$

$$= V_{AB} \cdot Q$$

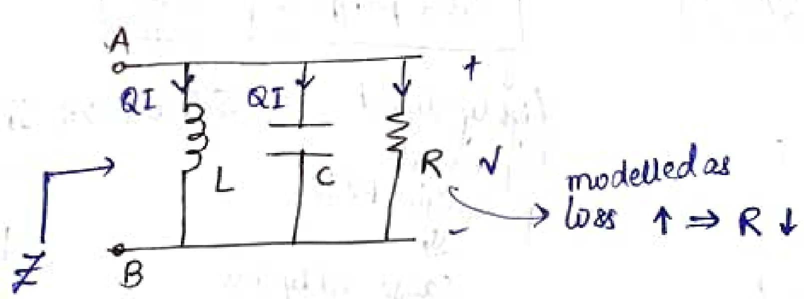
$$Q = \frac{\omega_0 L}{R} = \frac{1}{\omega_0 RC}$$

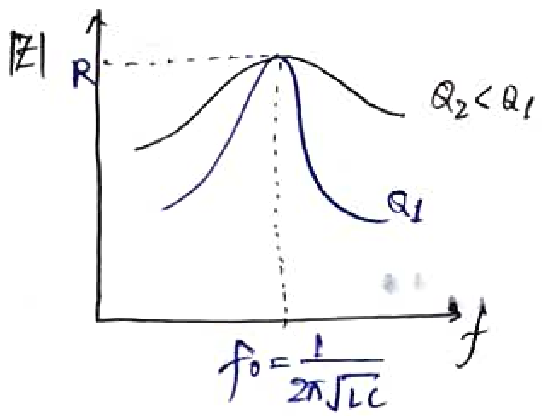
$R \downarrow \Rightarrow Q \uparrow$

$\downarrow$

loss

Parallel RLC Network:



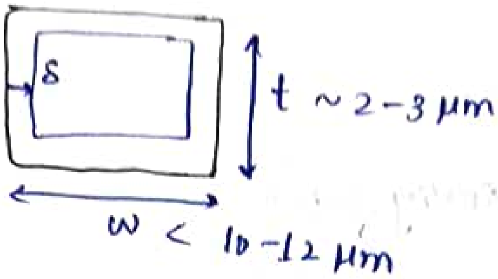


$$Q = \frac{R}{\omega_0 L} = \omega_0 RC$$

$R \downarrow \Rightarrow Q \downarrow \Rightarrow \text{loss} \uparrow$

05-02-2025

Conductor: Metal in the inductor

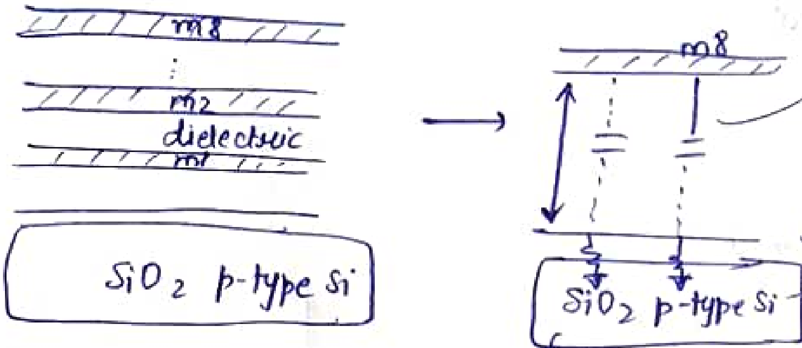


Losses:

① Ohmic loss
+
Skin effect

$R = \rho \cdot \frac{l}{wt}$

② Eddy current loss



→ Capacitive coupling

No metals in b/w inductor & Si substrate

→ Buried oxide layer → to reduce eddy current

highly doped to prevent parasitic BJTs
⇓
cause eddy loss

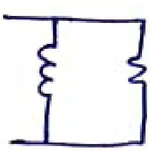
Silicon on Insulator (SOI)

FD SOI

Modelling the loss:

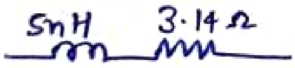
① 

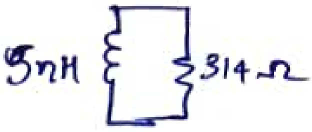
$Q = \frac{\omega_0 L}{R}$

② 

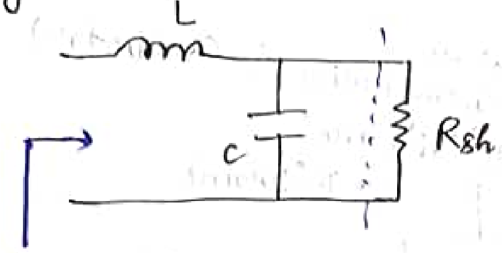
$Q = \frac{R}{\omega_0 L}$

Eg. $f = 1\,GHz$
$L = 5\,nH$
$Q = 10$





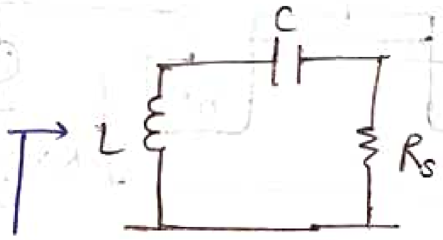
Matching:



$$SL + \underbrace{\frac{1}{SC} \parallel R_{sh}}$$

$$= \frac{R_{sh}}{1+j\omega R_{sh} C} \times \frac{1-j\omega R_{sh} C}{1-j\omega R_{sh} C}$$

$$= \frac{R_{sh}}{1+\omega^2 R_{sh}^2 C^2} - \frac{j\omega R_{sh}^2 C}{1+\omega^2 R_{sh}^2 C^2}$$

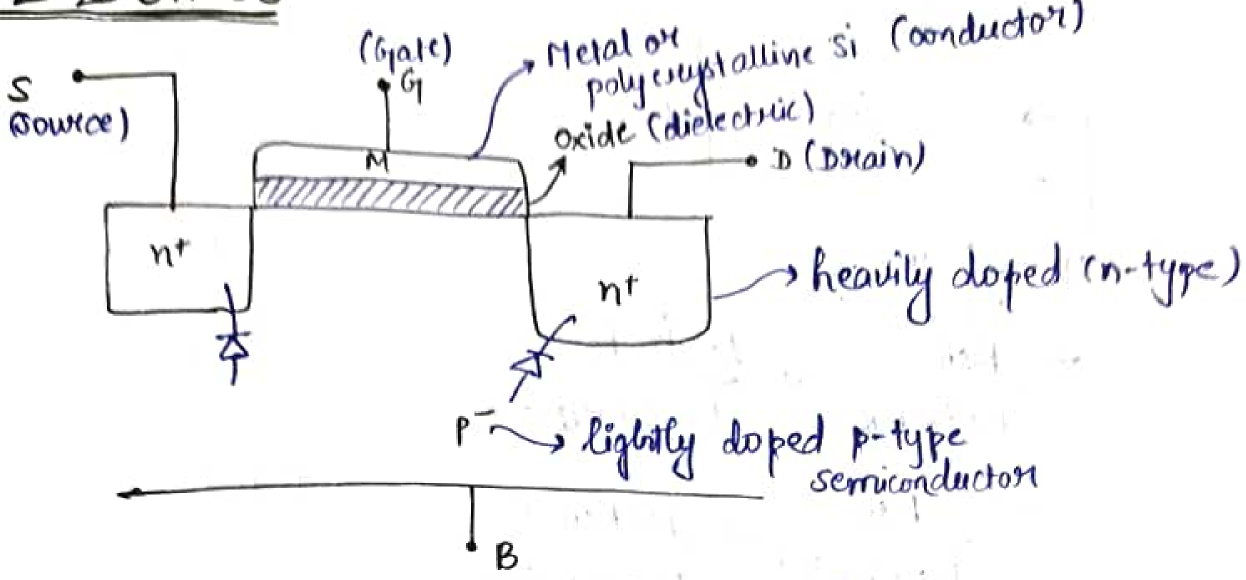


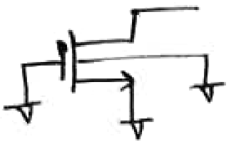
$$R_T = (1+Q^2)R_s$$

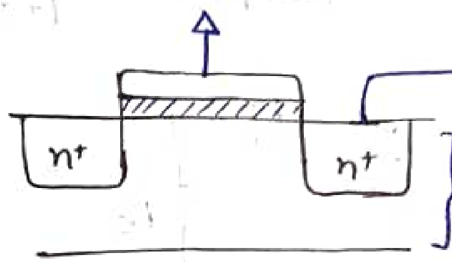
$$R_s = \frac{R_{sh}}{1+Q^2}$$

07-02-2025

MOS Devices

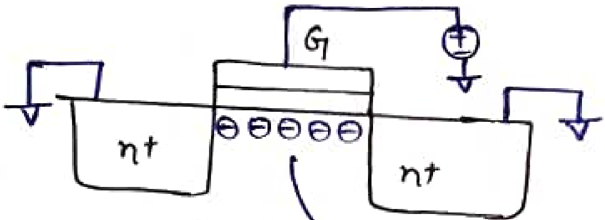






Reverse biased
→ no current flows

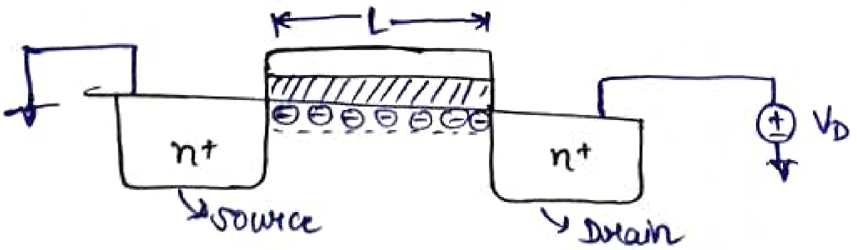
Put some voltage to gate.



→ Due to gate voltage, ⊖ve particles form a layer.

→ Inversion layer (sufficient width of $e^{\ominus}$)
↳ forms a channel
↳ Gives a way for $e^{\ominus}$s to move from n^+ to n^+, if there is a potential difference b/w D & S.

<u>Threshold voltage</u>: Min^m voltage required to setup the channel.



Higher potential : Drain
Lower potential : Source

$e^- : S \rightarrow D$

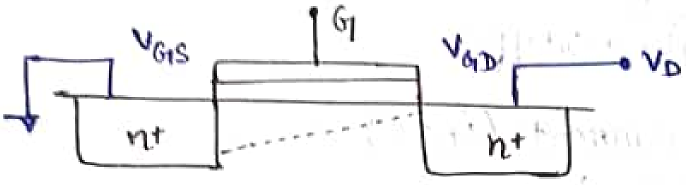
$I : D \rightarrow S$

$V_{GS} > V_{TH}$
$V_{DS} > 0$ $\Big\} \Rightarrow$ current flow

$$\boxed{I_D = \mu_n C_{ox} \left(\frac{W}{L}\right)(V_{GS} - V_{TH}) V_{DS}}$$ → Triode/Linear Region

(Small V_{DS})

μ_n: Mobility
C_{ox}: Oxide capacitance / unit area
L: Length of the channel (length e^- has to flow through from S to D)
W: width of the channel (going inside the plane of paper)



$V_{GS} > V_{GD} > V_{TH}$

As $V_D \uparrow$, $V_{GD} \downarrow$ and approaches to $V_{TH} \Rightarrow$ channel starts decreasing on drain side.

After $V_{GD} < V_{TH}$, channel breaks on drain side but still active on source side $\Rightarrow e^-$s able to cross drain side.
(as in BJT from base to collector)

⇓

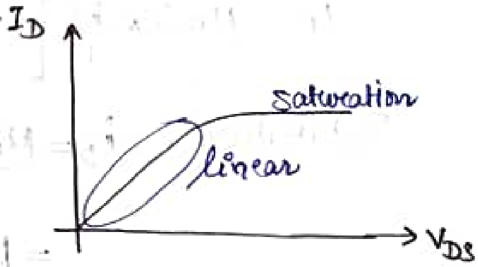
But current : constant

⇓

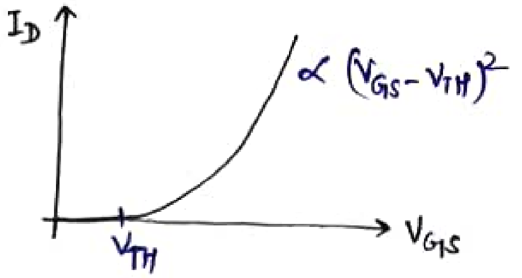
Saturation Region

$$\boxed{I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_{TH})^2}$$

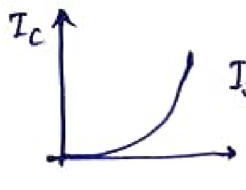
For saturation, $\boxed{V_{DS} = V_{GS} - V_{TH}}$



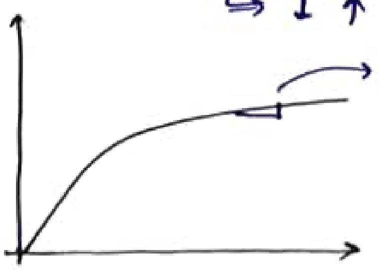
$$I_D = \mu_n C_{ox}\left(\frac{W}{L}\right)\left[(V_{GS} - V_{TH}) V_{DS} - \frac{V_{DS}^2}{2}\right]$$ (large V_{DS})



BJT

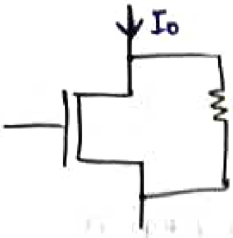


As channel length $\downarrow$
↳ $I \uparrow$ $\Big\}$ channel length modulation



$\left.\frac{\partial I_D}{\partial V_{DS}}\right|_{V_{GS}} = g_{ds}$ (conductance)

$$r_o = \frac{1}{g_{ds}} = \left(\frac{\partial I_{ds}}{\partial V_{ds}}\right)^{-1}$$

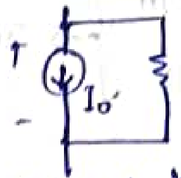


$r_o = \frac{1}{g_m}$

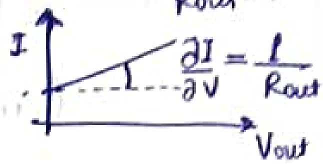
$r_o = \frac{1}{\lambda I_D}$ $\left(\lambda = \frac{\Delta L}{L}\right)$ ← Channel-length modulation coefficient

$L \downarrow \Rightarrow r_o \downarrow \Rightarrow$ slope $\uparrow$

Current source:



$I = I_o + \frac{V_{out}}{R_{out}}$



Subthreshold: Region where current flows before the formation of channel.

$(V_{GS} < V_{TH}) \rightarrow$ cutoff

Cutoff Mode: Drain current $(i_D = 0)$

$(V_{GS} < V_{TH})$

Linear Mode: Linear drain current

- $V_{DS} \rightarrow$ small

$$i_D = \mu_n C_{ox} \frac{W}{L} \left[(V_{GS} - V_{TH}) V_{DS} \right]$$

- Large V_{DS}

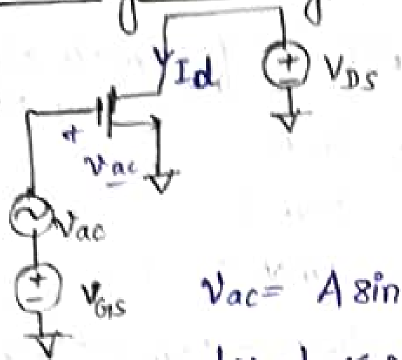
$$i_D = \mu_n C_{ox} \frac{W}{L} \left[(V_{GS} - V_{TH}) V_{DS} - \frac{V_{DS}^2}{2} \right]$$

Saturation: $i_D = \frac{\mu_n C_{ox}}{2} \frac{W}{L} (V_{GS} - V_T)^2$

$$= \frac{\mu_n C_{ox}}{2} \frac{W}{L} V_{DS}^2.$$

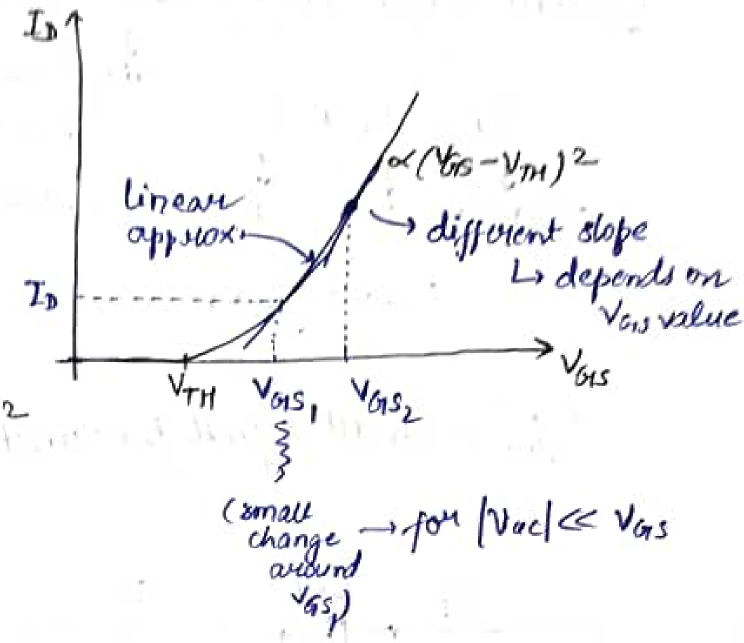
Small Signal Analysis

11-02-2025

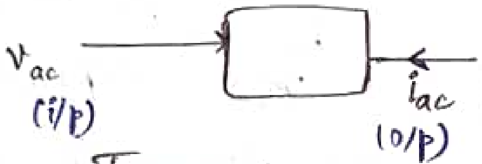


$v_{ac} = A \sin \omega t$

$|v_{ac}| \ll V_{GS}$



$$I_d = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} + v_{ac} - V_{TH})^2$$



Transconductance (g_m) → $\left(\frac{\text{Output}}{\text{Input}}\right)$ has a unit of conductance, & the current results around different terminal

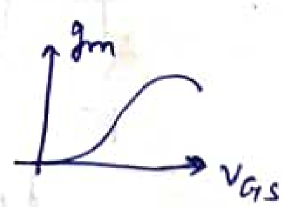
$$I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_{TH})^2$$

$$\Rightarrow \boxed{\begin{aligned} \frac{\partial I_D}{\partial V_{GS}} &= \mu_n C_{ox}\left(\frac{W}{L}\right)(V_{GS} - V_{TH}) \\ &= \frac{2 I_D}{(V_{GS} - V_{TH})} \\ &= \sqrt{2 I_D \mu_n C_{ox}\left(\frac{W}{L}\right)} \end{aligned}} \; \Bigg\} \; g_m$$

We would want a particular 'g_m' specification for LNA.

↳ Depends upon (↑)

- $\left(\frac{W}{L}\right)$, for fixed I_D. ↑
- V_{GS}, for fixed I_D. ↓ →
- $g_m \propto \sqrt{I_D}$



Small signal Model (Low frequency) : Valid only for small signals (not DC)

Low frequency → didn't consider the capacitances

$I_d = I_D + i_{ac}$

I_d → composite, I_D → DC, i_{ac} → ac (convention)

$= I_D + g_m v_{ac}$

'g_m' : small signal parameter

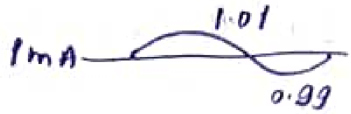
Eg. $I_D = 1mA$

$g_m = 10 mS$

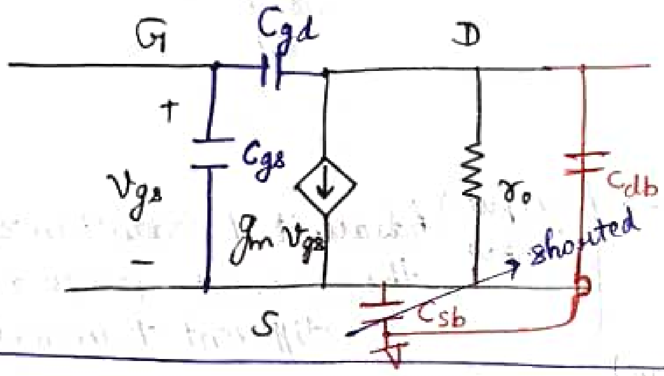
$v_{ac} = 1mV \sin\omega t$

$I_d = 1mA + 10mS \times 1mV \sin\omega t$

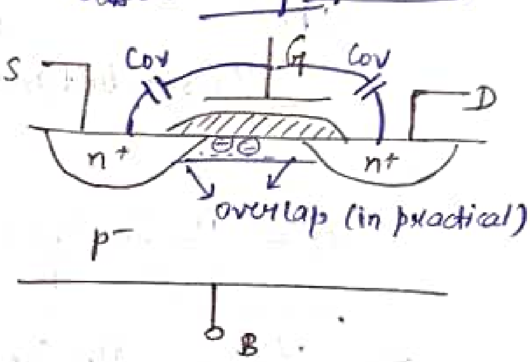
$= 1mA + 10\mu A \sin\omega t$



$I_D = g_m \cdot V_{GS}$ X



Capacitances:



C_{ov} : overlap capacitance
↳ exist even if device is off
↳ due to overlapping of oxide

(i) $V_{GS} < V_{TH}$ ($V_{GS} = 0$)

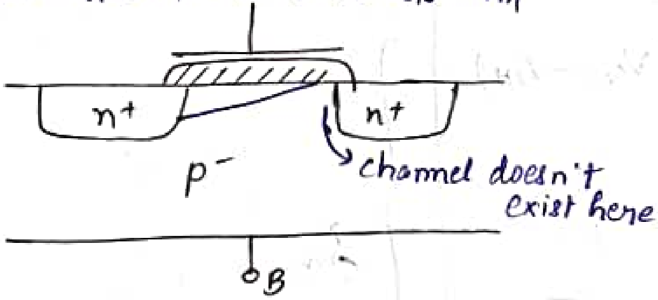
(ii) $V_{GS} > V_{TH}$ & $V_{DS} < V_{GS} - V_{TH}$
↳ channel exists & in triode region

Triode: $C_{tot.} = WL\,C_{ox}$

$C_{gs} = C_{ov} + \underbrace{\frac{1}{2} WL\,C_{ox}}_{\text{budgeted half for source}} = C_{gd}$

(iii) $V_{GS} > V_{TH}$ & $V_{DS} \geq V_{GS} - V_{TH}$ (saturation)

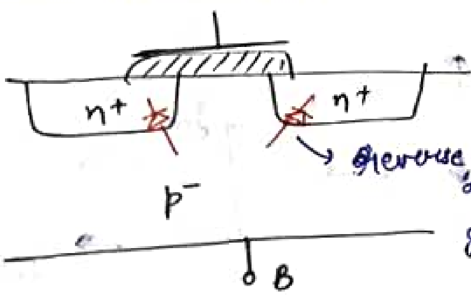
(iii) $V_{GS} > V_{TH}$ & $V_{DS} \geq V_{GS} - V_{TH}$

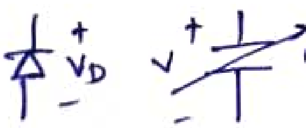


$C_{gs} = \underbrace{\frac{2}{3} WL\,C_{ox}}_{\text{(experimental)}} + C_{ov}$

$C_{gd} = C_{ov}$ (as channel ~~go~~ doesn't exist)

Other capacitances:



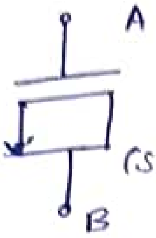


$C = f(V)$

MOS capacitance : MOSFETs have their own capacitances

↳ can be used as capacitor

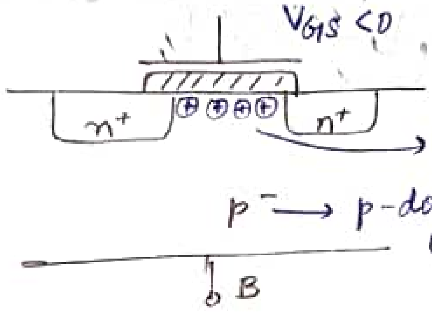
↳ Voltage-dependendent capacitance



(S & D shorted)

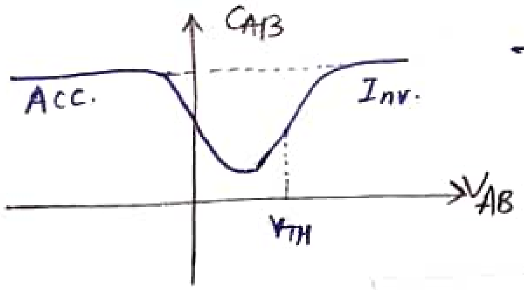
$V_{AB} > V_{TH}$

$C_{tot.} = 2C_{ov} + WLC_{ox}$



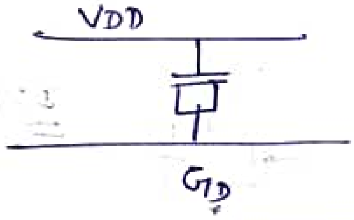
→ accumulation $\Rightarrow V_{AB} < 0$

$p^- \longrightarrow$ p-doped (⊕ve/hole carrier)



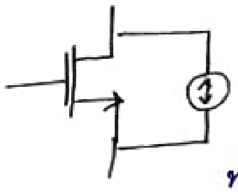
→ Voltage-dependent

↳ Used in supply decoupling capacitor



12-02-2025

Noise in a MOSFET



$\overline{i_{nd}^2} = 4KT\gamma g_m$

↓ noise drain current (thermal noise)

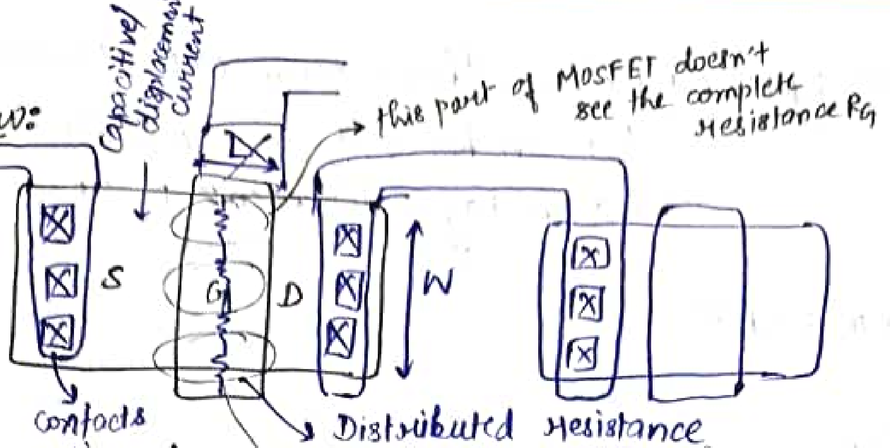
↳ white

γ: Noise coefficient

long channel : $\gamma \sim 2/3$

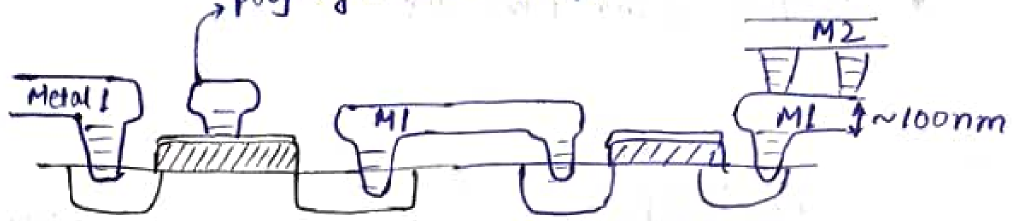
Short channel : $\gamma \sim 1$

Top view:



Cross-sectional view

polycrystalline Si (poly)

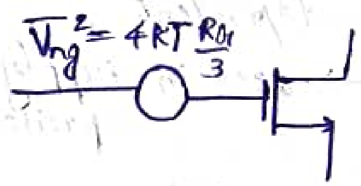


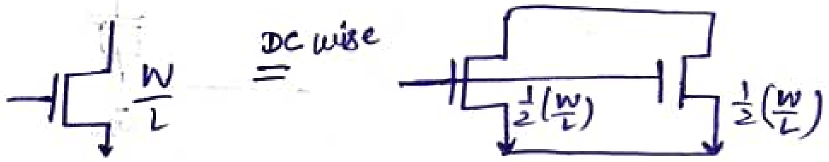
R_G : total gate resistance

Effective gate resistance $= \frac{R_G}{3}$

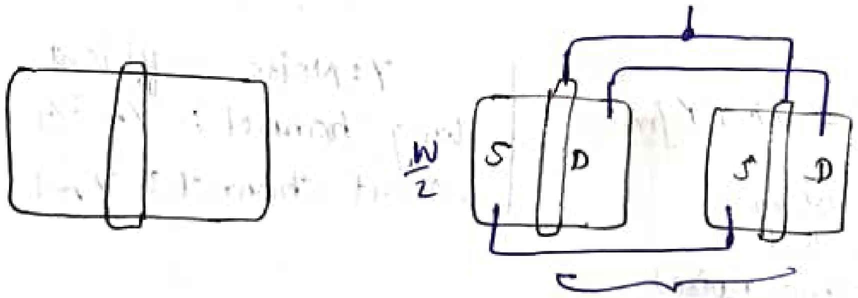
Model:

$\overline{V_{ng}^2} = 4KT\frac{R_G}{3}$





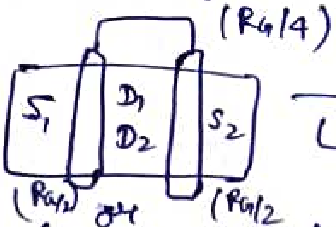
(S, D, G shorted)



↓ Merge

$\overline{V_{ng}^2} = 4KT\frac{R_G}{12} \leftarrow$

(×4 ↓)



or (sources merged)

→ No. of fingers = 2

↳ Total drain area has reduced.

↳ $C_{db}\downarrow$, $C_{sb}\downarrow$

→ 4 fingers

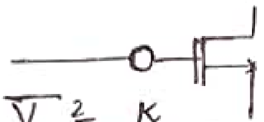
→ $\overline{V_{ng}^2} = 4KT \frac{R_G}{48}$ (x16↓)

→ Noise figure ↓

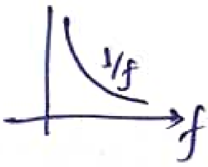
→ Gate noise can be handled by adding no. of fingers.
↳ can be managed.
↳ Routing complexity↑

Flicker Noise

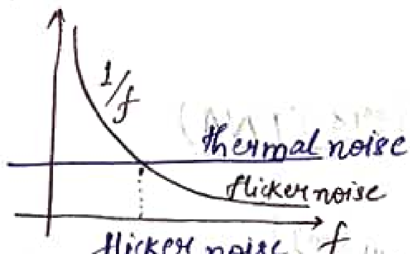
↳ source: Dangling bonds capturing charges in the (gate) oxide layer.



$\overline{V_{nf}^2} = \frac{K}{WLC_{ox}f}$

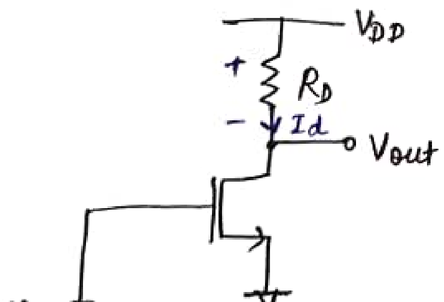


$\propto \frac{1}{f}$ → Not considerable for RF. ↳ considered in oscillator, and in analog circuits.



$\sim 10-100$ MHz

→ No perfect DC
↳ sampled at finite freq.
↳ observing longer ⇒ flicker noise ↑



Amplifier (saturation region)

$V_B \to I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_B - V_{TH})^2 \to V_{out}$

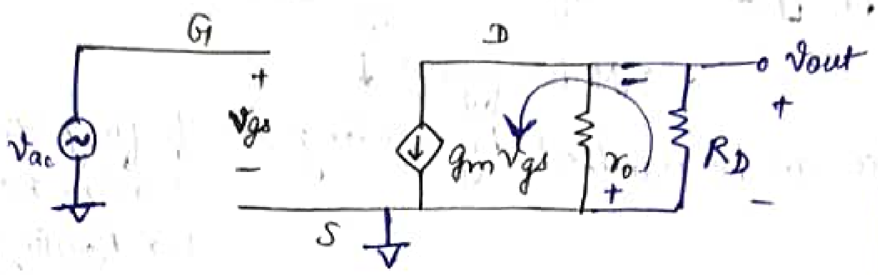
$I_d = I_D + g_m V_{ac}$

$$V_{out} = V_{DD} - I_d R_D$$
$$= V_{DD} - (I_D + g_m V_{ac}) R_D$$
$$= \underbrace{V_{DD} - I_D R_D}_{DC} - \underbrace{g_m V_{ac} R_D}_{ac}$$

$$A_V = \frac{-g_m V_{ac} R_D}{V_{ac}} = -g_m R_D$$

Small-signal equivalent circuit:

Set all DC source = 0



V_{DD} → does not change its value however much current is drawn.
↳ ac ground

$$v_{out} = -g_m v_{gs}(r_o \| R_D)$$

$$v_{gs} = v_{ac}$$

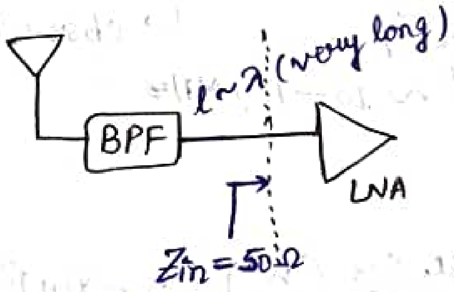
$$A_v = \frac{v_{out}}{v_{ac}} = \frac{-g_m v_{ac}(r_o \| R_D)}{v_{ac}}$$

$$= -g_m (r_o \| R_D)$$

$$\approx -g_m R_D$$

14-02-2025

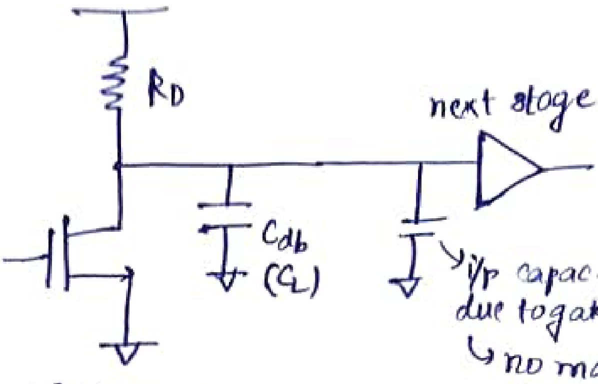
Low Noise Amplifiers (LNA)



(Input matching required for LNA, otherwise reflections)

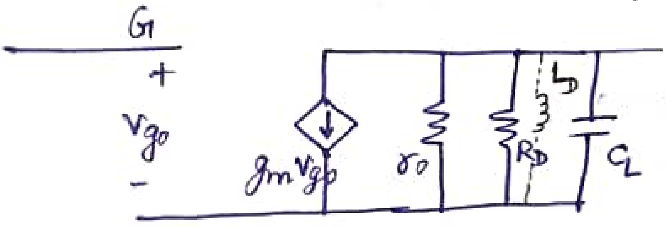
- → Input matching
- → Noise figure
- → Gain
- → Power consumption
- → Stability ⇝ Amplifier should not end up being oscillator ⇝ Risk greater in RF ckt. due to unwanted feedback path

→ Linearity

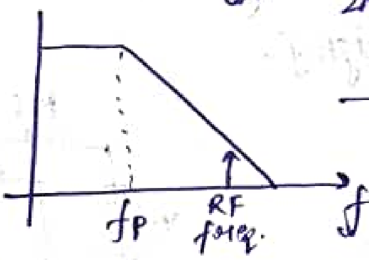


(Common source)

↳ i/p capac. due to gate
↳ no matching required inside chip
↳ ~~not~~ b/c no significant length.

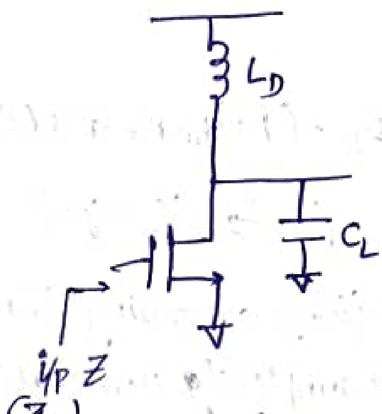


$$f_P = \frac{1}{2\pi (R_D \| r_o) C_L}$$



→ Pole location in low frequent RF freq.
↳ Gets filtered out
↳ No significant gain.

⇓

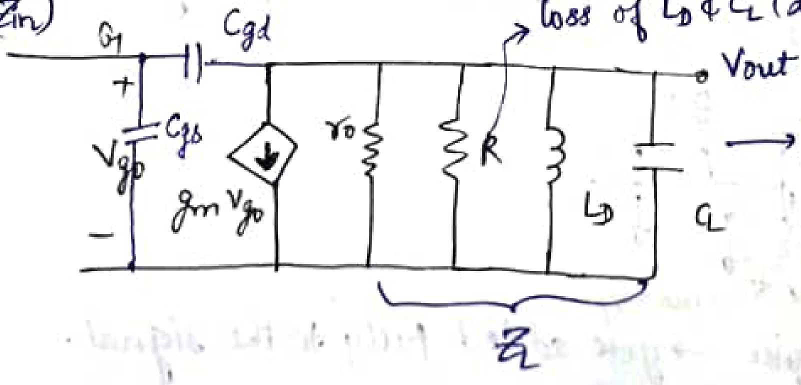


$$f_o = \frac{1}{2\pi\sqrt{L_D C_L}}$$

→ Tune for f_o

→ ~~Gain~~ Pole location in RF region
↳ Gain limited by r_o, (significant gain), ideally ∞.

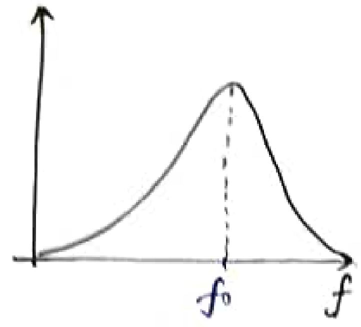
loss of L_D & C_L (due to shunt R)



→ Parallel RLC n/w.

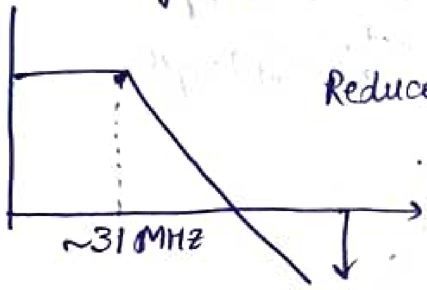
$V_{out} = -g_m v_{gs} \cdot Z_L$

$A_v = -g_m Z_L$



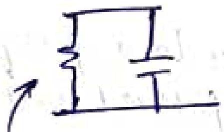
Eg. $R_D = 5K\Omega$
$C_L = 1\,pF$
$\Rightarrow f_P \approx 31\ MHz$

$A_v = g_m R_D$

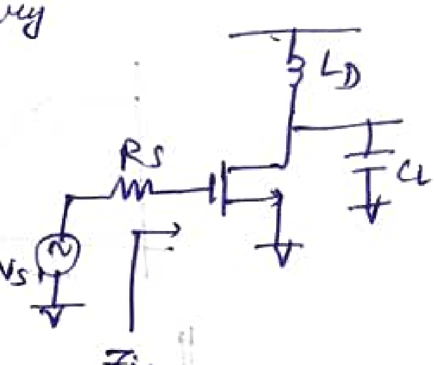


Reduce $R_D \downarrow \Rightarrow$ But gain $\downarrow$.

Input impedance $\rightarrow$ $\underbrace{\text{real}}$ + imaginary
low for high f.



Real part due to R at o/p



$$Z_{in} = \underbrace{Re\{\ \}} + j\{\ \}$$

↳ can become ⊖ve

beyond $\omega_1^2 = \dfrac{R_S C_L + g_m L_D - (1+g_m R_S) R_S (C_L + C_{gd})}{g_m L_D^2 (C_L + C_{gd})}$

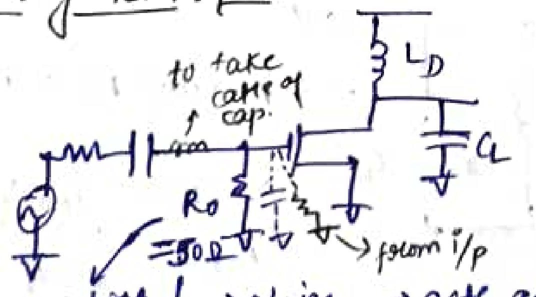
⇓

Oscillation

↳ Due to ⊖ve $Re\{Z_{in}\}$ cancelling the loss.

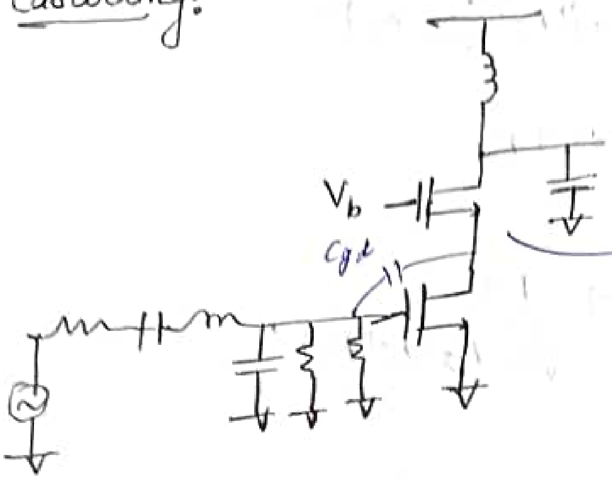
↳ oscillates at completely unrelated freq.

<u>Matching Technique</u>:



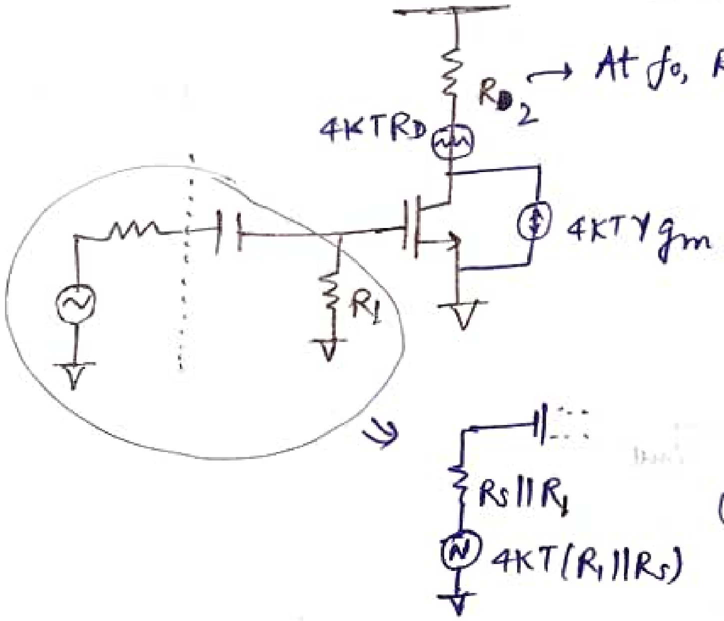
for matching ↳ Noise → gets added fully to the signal.
↳ straightforward

Cascoding:



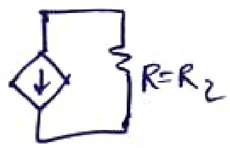
Cascode: Provide one level of isolation

At f_0, RLC ckt. reduced to R



$(R_S = R_1)$

$r_0 >> R_2$



Voltage o/p noise power spectral density,

$$\overline{V^2_{n,out}} = 4KTR_2 + 4KT\gamma\, g_m \cdot \underbrace{R_2^{2}}_{\text{due to power quantity}} + 4KT(R_1 \| R_S)(g_m R_2)^2$$

$$NF = \frac{\text{Total noise @ o/p}}{(\text{Gain})^2 \cdot 4KTR_S}$$

$$\text{gain} = \left(\frac{R_1}{R_1 + R_S}\right) g_m \cdot R_2$$

$$NF = \frac{4KTR_2 + 4KT\gamma g_m R_2^2 + 4KT(R_1 \| R_s)(g_m \cdot R_2)^2}{\left(\frac{R_1}{R_1+R_s}\right)^2 (g_m R_2)^2 \cdot 4KTR_s}$$

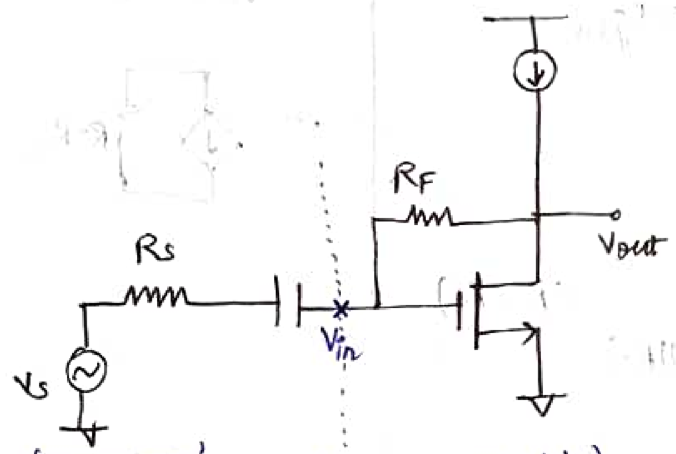
$$= \frac{R_s + R_1}{R_1} + \frac{\gamma R_s}{g_m (R_s \| R_1)^2} + \frac{R_s}{g_m^2 R_2 (R_s \| R_s)^2}$$

> 2

$\therefore$ NF > 3dB. ⇝ high.
↳ Not used

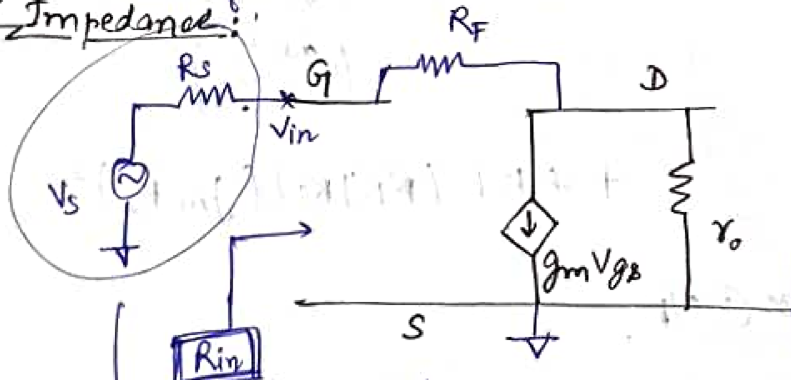
Resistive Feedback LNA

28-02-2025



Model of source (Thevenin eqv. ckt.)

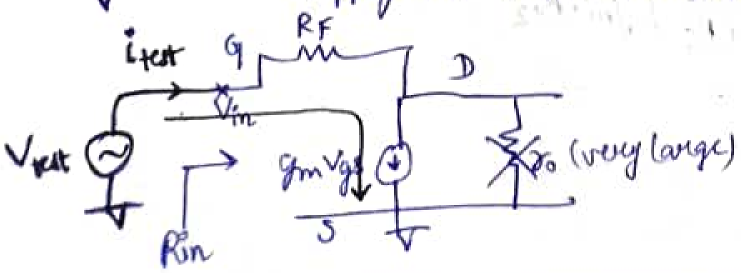
I/P Impedance:



(Capacitance not shown for simplification)

Active equipments (one with supply) should not be given DC, otherwise it may get damaged.

To get i/p impedance (of any circuit):
delete previous circuit,
supply test V and calculate test I.

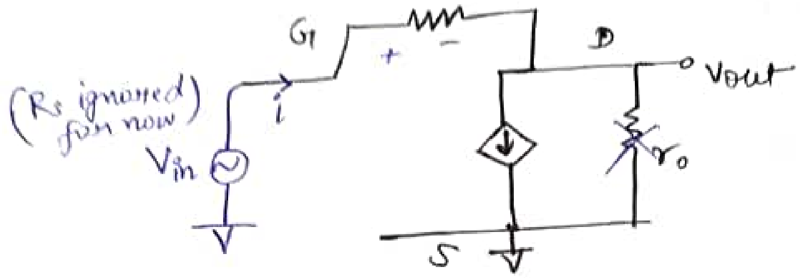


$$i_{test} = g_m V_{in} = g_m V_{test}$$

$$\therefore \boxed{R_{in} = \frac{V_{test}}{i_{test}} = \frac{1}{g_m}}$$

To get $R_{in} = 50\Omega$,
$g_m = 20mS$.

Gain:

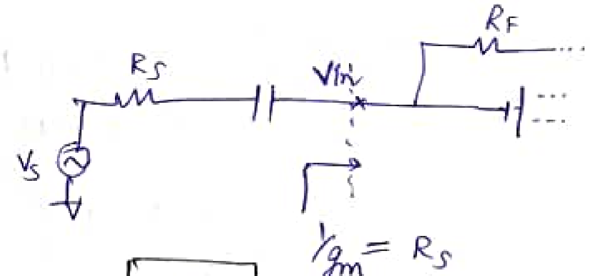


$$V_{out} = V_{in} - iR_F$$
$$= V_{in} - g_m V_{in} R_F = V_{in}(1 - g_m R_F)$$

$$\frac{V_{out}}{V_{in}} = 1 - g_m R_F$$

$$= \underbrace{1 - \frac{R_F^{\sim 100-1000\Omega}}{R_{S\,\sim 10\Omega}}}_{\to\ \ominus ve} \quad \left(\text{To ensure } R_S = \frac{1}{g_m} = 50\Omega\right)$$

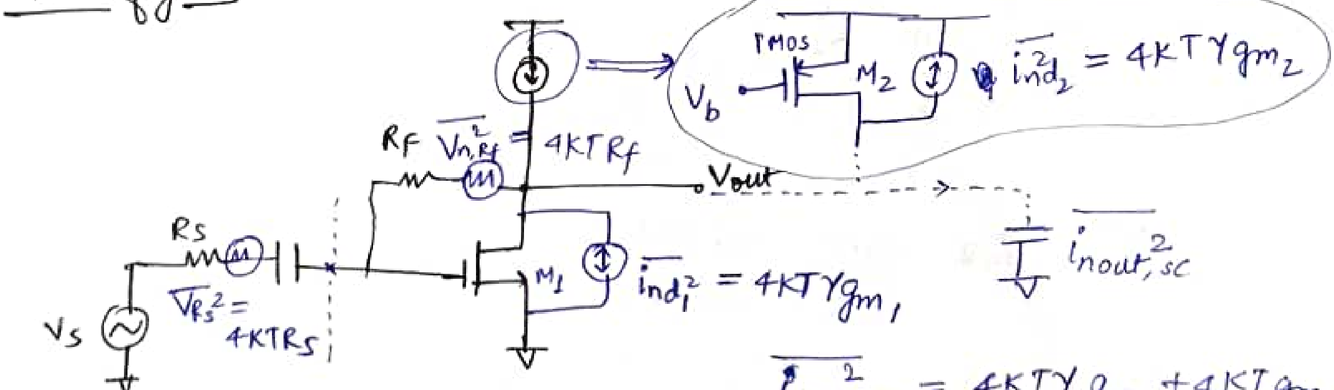
$$\boxed{\text{gain} \approx -\frac{R_F}{R_S} \approx -g_m R_F}$$



$$\frac{V_{in}}{V_s} = \frac{1}{2}, \text{ if } R_S = \frac{1}{g_m}$$

overall gain: $\frac{V_{out}}{V_S} = \frac{1}{2}\left(1 - \frac{R_F}{R_S}\right) \approx \boxed{\frac{-R_F}{2R_S}}$

Noise figure:



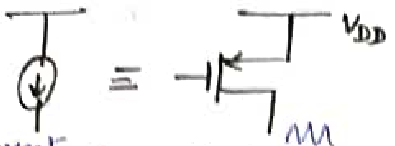
$$\overline{i_{nout,sc}^2} = 4KT\gamma g_{m_1} + 4KTg_{m_2} + \frac{4KT}{R_F}$$

(current noise figure of ckt. when load is ∞)

[V_s ignored]

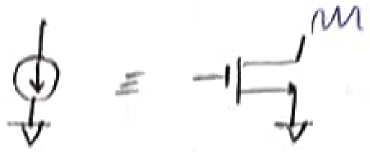
$$NF = 1 + \frac{\text{Noise added}}{\text{Gain}^2 \cdot 4KTR_S}$$

Current source:



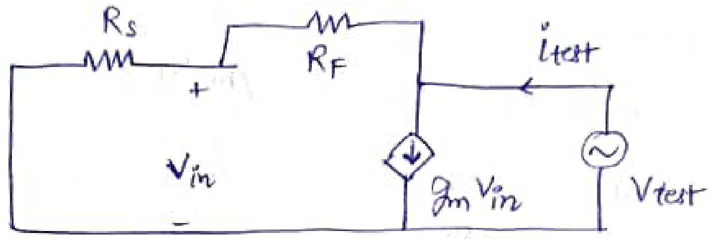
current shouldn't change $\rightarrow$ (V_{gs} shouldn't change)

Current sink:



$$\overline{V_{n,out}^2} = \overline{i_{n,out,sc}^2} \cdot R_{out}^2$$

To find R_{out}:



$$i_{test} = g_m V_{in} + \frac{V_{test}}{R_F + R_S}$$

$$V_{in} = V_{test} \cdot \frac{R_S}{R_S + R_F}$$

$$i_{test} = \frac{g_m \cdot V_{test} \cdot R_S}{R_S + R_F} + \frac{V_{test}}{R_F + R_S}$$

$$= V_{test}\left(\frac{g_m R_S + 1}{R_S + R_F}\right)$$

$$\therefore\ R_{out} = \frac{V_{test}}{i_{test}} = \frac{R_F + R_S}{1 + g_m R_S}$$

19-02-2025

and $g_m = \frac{1}{R_S}$

$$\Rightarrow \quad R_{out} = \frac{R_S + R_F}{2}$$

$$NF = 1 + \frac{\text{noise added}}{(\text{Gain})^2 \cdot 4KTR_S}$$

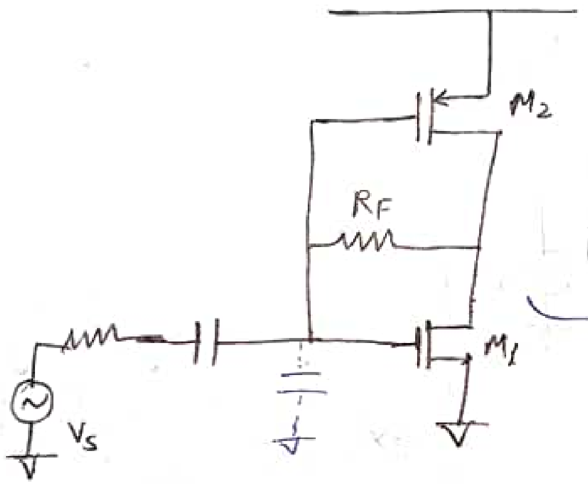
$$\overline{V_{n,out}^2} = 4KT\gamma(g_{m1} + g_{m2}) \cdot R_{out}^2 + 4KTR_F$$

$$\Rightarrow NF = 1 + \frac{\cancel{4K}T\gamma(g_{m1} + g_{m2}) \cdot (R_S + R_F)^2}{\left(\frac{R_F}{2R_S}\right)^2 \times \cancel{4}\cancel{K}\cancel{T}R_S \times \cancel{4}} + \frac{\cancel{4KT}R_F}{\left(\frac{R_F}{2R_S}\right)^2 \cancel{4KT}R_S}$$

$$= 1 + \frac{\gamma(g_{m_1}+g_{m_2})\,\cancel{(R_S+R_F)}^2}{\cancel{R_F^2}/R_S} + \frac{1}{R_F/R_S} \quad \left(\because R_S+R_F \approx R_F\right)$$

$$= \underbrace{1 + \underbrace{\frac{4R_S}{R_F}}_{\text{very small}} + \underbrace{\gamma(g_{m_1}+g_{m_2})\,R_S}}_{\text{little over 2 (better than the prev.)}}$$

Modification:



(Active load)

→ CMOS (inverter) with resistive feedback

$$R_{in} = \frac{1}{g_{m_1}+g_{m_2}}$$

For lower technology (short channel length), parasitic resistance is low
⇓
$S_{11} < -10\,dB$

For inv,

$$NF = 1 + \frac{4R_S}{R_F} + \gamma$$

5G:

600M — 900M — 2.7G — 6GHz

LB — MB — UHB

divided into chunks ~ 40 MHZ (allocated to each user)

20MHZ:
- 10M → around (2.5GHZ)
- 5M → (720MHZ) } LB (interband) } each handled by separate LNAs
- 5M → (830MHZ) }

⇒ Carrier Aggregation

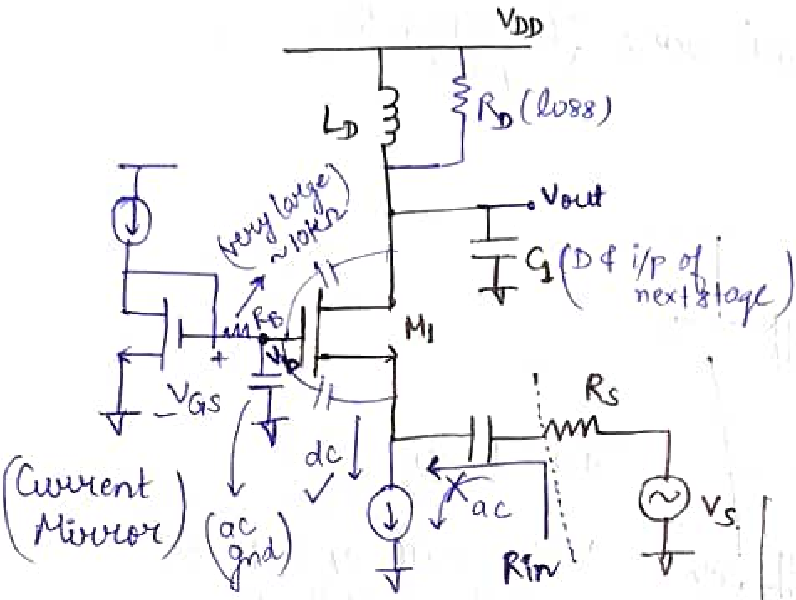
↳ Interaband and interband CA

(Interaband: within one band (eg, LB))

Inducator : Huge size compared to this design.

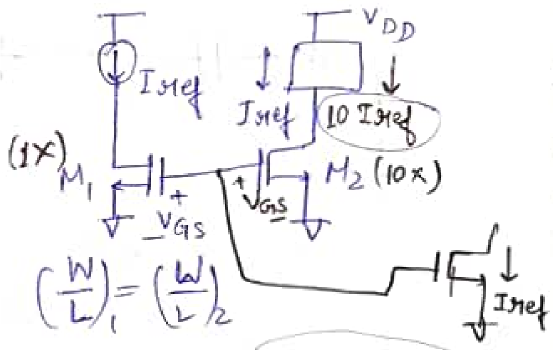
signal cancellation (dip) due to interference from two towers.
↳ Devices with two LNAs → main and diversity.
(switch if dip)

Common Gate LNA



i/p → S
O/p → D
ac GND → G

Feeding voltage division of V_{DD} to G will not work, as manufactory large no. of devices give mismatch of resistors (for division).



$\left(\frac{W}{L}\right)_1 = \left(\frac{W}{L}\right)_2$

$10\left(\frac{W}{L}\right)_1 = \left(\frac{W}{L}\right)_2$

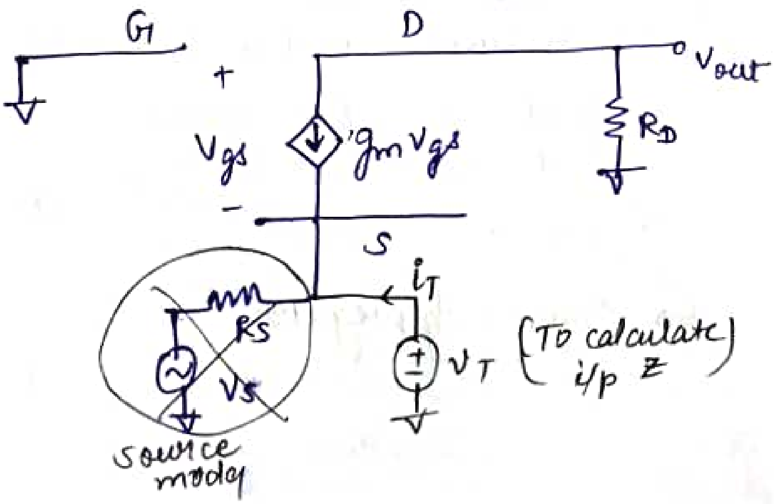
$I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_{TH})^2$

$$f_D = \frac{1}{2\pi\sqrt{L_D C_1}}$$

To resonate at high f,
if $C_1 \downarrow \Rightarrow L_D \uparrow\uparrow$
OR add some capacitance

25-02-2025

Small signal model at resonance : L_D & C_1 cancel.



(To calculate i/p Z)

① Input impedance

$V_{gs} = -V_T$

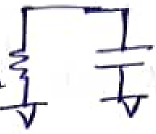
$i_T = -g_m V_{gs}$

$= + g_m V_T$

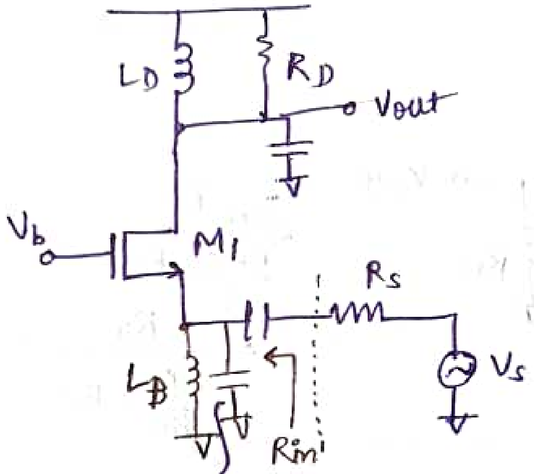
$$\boxed{R_{in} = \frac{1}{g_m}} = \frac{V_T}{i_T}$$

[If $g_m = 20\,mS \Rightarrow Re\{R_{in}\} = 50\,\Omega$]

↳ capacitances are not considered.

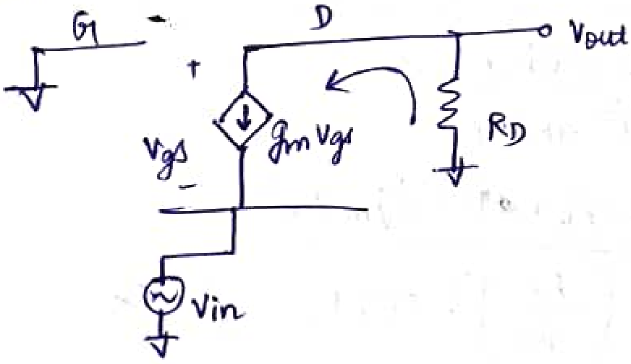


$Z_{in} = \frac{1}{g_m}$ $(C_{gs} + C_{bs} + \dots)$



(device cap + pad cap + routing cap)

② Gain



$V_{gs} = -V_{in}$

$V_{out} = -g_m V_{gs} R_D$

$= g_m V_{in} R_D$

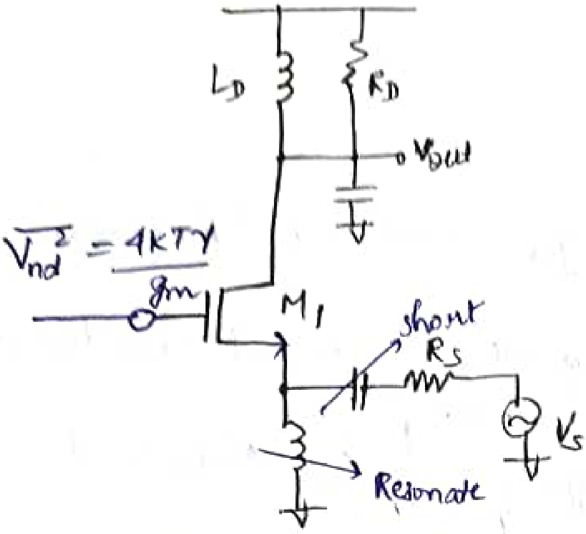
$\frac{V_{out}}{V_{in}} = g_m R_D$,

$g_m = 1/R_s$

$\therefore \frac{V_{out}}{V_{in}} = \frac{R_D}{R_s}$

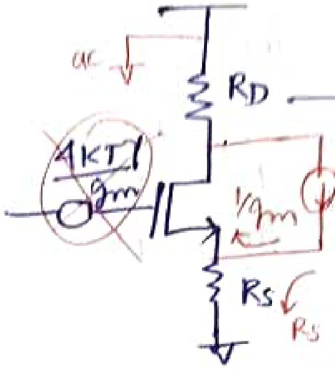
$$\therefore \boxed{\frac{V_{out}}{V_{in}} = A_V = \frac{R_D}{2R_s}}$$ (Overall gain)

③ Noise



(For resonating & perfectly matched)

$\equiv$

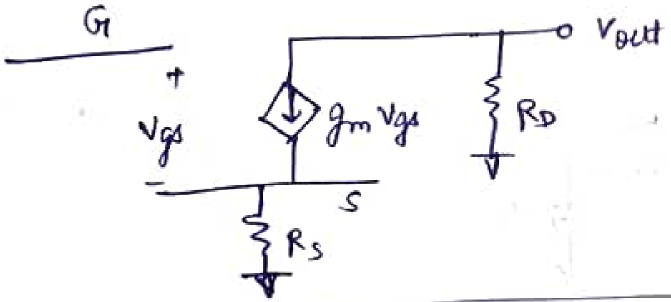


$R_D \longrightarrow$ Common source with resistive degeneration

$(R_s = 1/g_m)$

(For current source)

$4KT\gamma g_m \frac{1}{4} R_D^2 = KT\gamma g_m R_D^2$



$$\frac{g_m R_D}{1+g_m R_s} = \frac{R_D}{\frac{1}{g_m} + R_s} = \frac{R_D}{2R_s}$$

$$\therefore \boxed{\overline{V_{n,out,add}^2} = 4KTR_D + \underbrace{\frac{4KT\gamma}{g_m}\left(\frac{R_D}{2R_s}\right)^2}_{(KT\gamma g_m R_D^2)}}$$

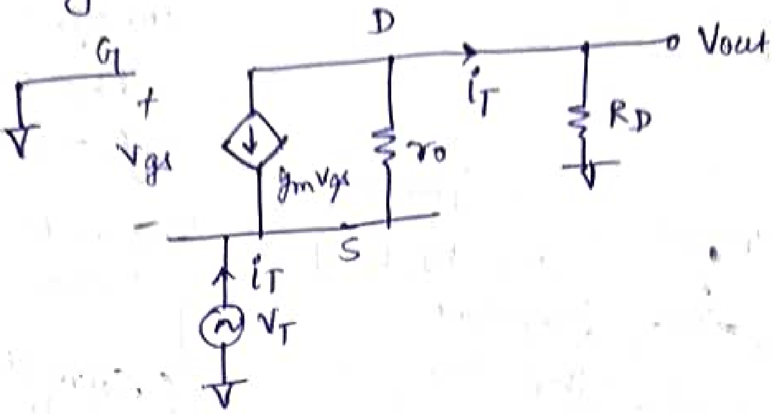
④ $$NF = 1 + \frac{\text{Noise added}}{\text{gain}^2 \cdot 4KTR_s}$$

$$= 1 + \frac{4KTR_D + 4KT\gamma g_m R_D^2}{\left(\frac{R_D}{2R_s}\right)^2 \cdot 4KTR_s}$$

$$= 1 + \frac{4 + \gamma g_m R_D}{\left(\frac{R_D}{R_s}\right)(4)}$$

$$= \boxed{1 + \gamma + 4\frac{R_s}{R_D}} \qquad \left(g_m = \frac{1}{R_s}\right)$$

Considering r_o,



$V_{gs} = -V_T$

$i_T = -g_m V_{gs} + i_{r_o}$

$\Rightarrow i_{r_o} = \dfrac{V_T - V_{out}}{r_o} = \dfrac{V_T - i_T R_D}{r_o}$

$i_T = g_m V_T + \dfrac{V_T - i_T R_D}{r_o}$

$i_T\left(1 + \dfrac{R_D}{r_o}\right) = V_T\left(g_m + \dfrac{1}{r_o}\right)$

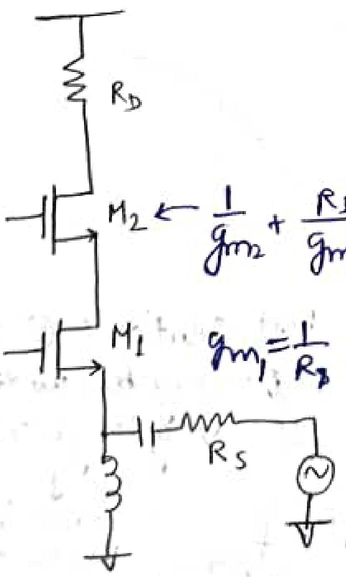
$\therefore R_{in} = \dfrac{V_T}{i_T} = \dfrac{1 + \frac{R_D}{r_o}}{g_m + \frac{1}{r_o}} = \dfrac{R_D + r_o}{1 + g_m r_o} \approx \dfrac{R_D + r_o}{g_m r_o}$

$\approx \boxed{\underbrace{\dfrac{1}{g_m} + \dfrac{R_D}{g_m r_o}}_{> 1/g_m}}$

Intrinsic gain (limit)

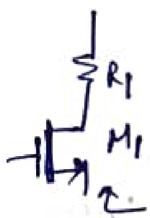
$\hookrightarrow$ determined by R_D

28-02-2025



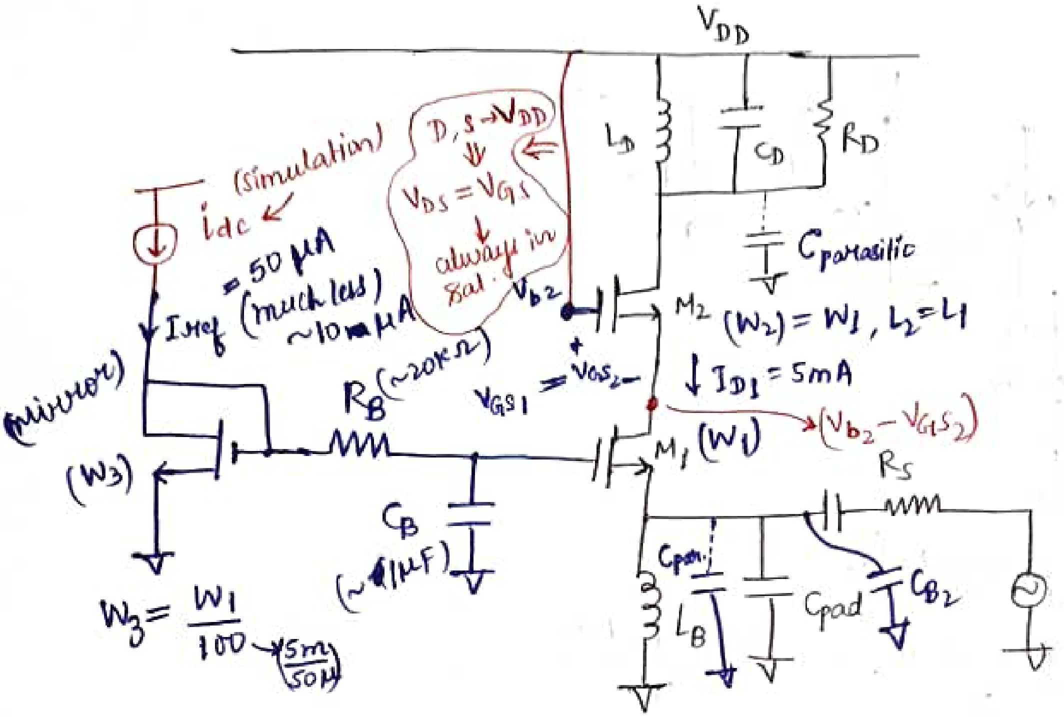
$M_2 \leftarrow \dfrac{1}{g_{m_2}} + \dfrac{R_D}{g_{m_2} + r_{o2}} = R_1$

$g_{m_1} = \dfrac{1}{R_S}$



$R_{in} = \dfrac{1}{g_{m_1}} + \cancel{\dfrac{1/g_{m_2}}{g_{m_1} r_{o1}}}^{\approx 0} + \cancel{\dfrac{R_D}{(g_{m_1} r_{o1})(g_{m_2} r_{o2})}}^{\approx 0}$

$\approx \dfrac{1}{g_{m_1}}$



Design:

$R_{in} = 50\,\Omega$

$f_0 = 2.4$ GHz

Tech.node = 28nm

↳ $V_{DD} = 1$ or 0.9 V

Saturation Mode:

$V_{DS} > V_{GS} - V_{TH}$

Simulation gives $V_{DS,sat}$. For $V_{DS} > V_{DS,sat}$, device is in sat.

$$\frac{1}{g_{m_1}} = 50\,\Omega \Rightarrow g_{m_1} = 20\,mS$$

[# First simulate DC]

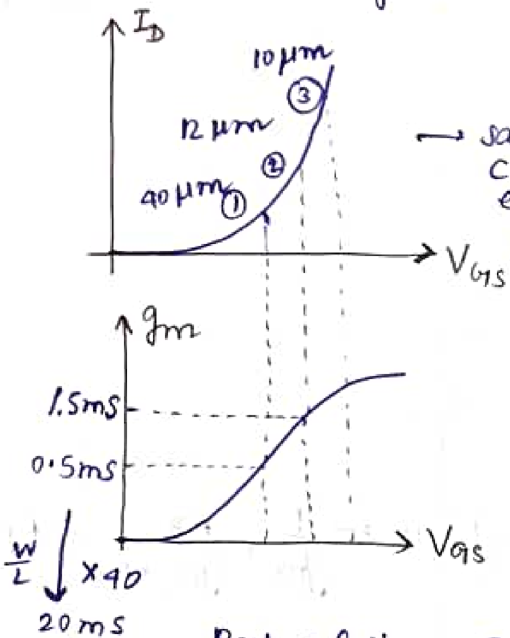
↳ To check if ckt. is alive.

Decide W_1, L_1, V_{bias}

- $L = L_{min}$ → To reduce the parasitic capacitances (Take lowest channel length)
- $W_0 = 1\,\mu m$ (easier!)

  $L_0 = L_{min}$

  $\frac{W_0}{L_0} = \frac{1\,\mu m}{L_{min}}$ ← start with



→ save as CSV / excel sheet.

$$g_m = \frac{\partial I_D}{\partial V_{GS}}$$

$$g_m = \mu_n C_{ox} \left(\frac{W}{L}\right)(V_{GS} - V_{th})$$

$$\propto \frac{W}{L}$$

Best solution: ② large

①: very ~~less~~ ~~width~~ (device)

③: very large current

↓

Take little lower than the saturated value.

↓

~80% of sat. value.

- $f_0 = \frac{1}{2\pi\sqrt{LC}}$ ⇒ For small C, L_B has to be large

  ↓ 2.4 GHz

  ⇓

  Replace inductor by transistor as current source

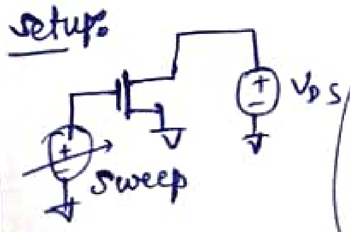
  OR

  Add an extra capacitor along with inductor. (C_{B_2})

can be pushed offchip.

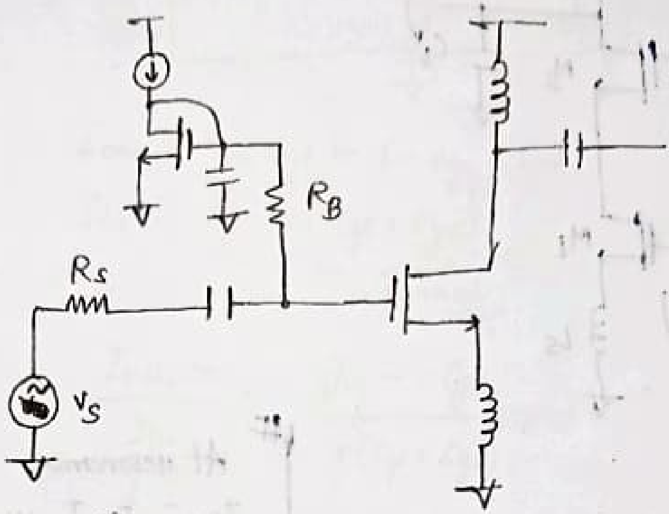
L_D → cannot be offchip.

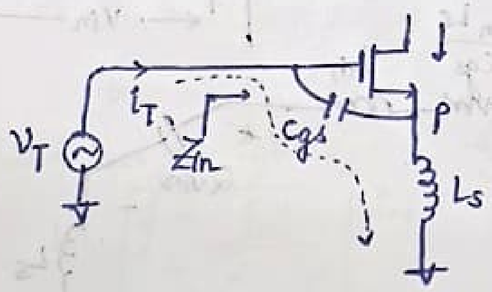
Setup



18-03-2025

Common Source with Inductive Degeneration LNA





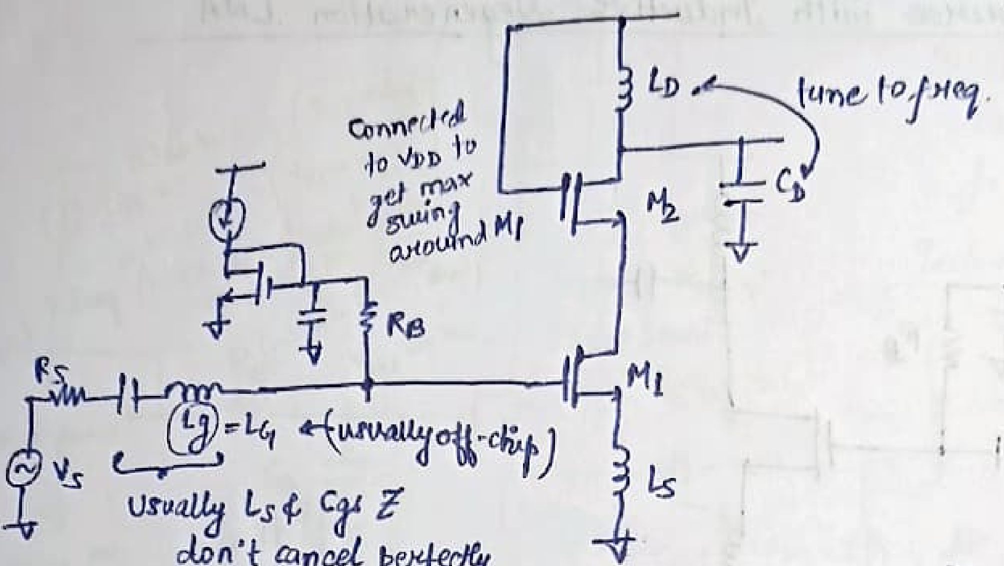
$$v_P = sL_S\,(i_T + g_m v_{gs})$$

$$v_{gs} = \frac{i_T}{sC_{gs}}$$

$$v_T = \underbrace{\frac{i_T}{sC_{gs}}}_{v_{gs}} + sL_S\left(i_T + \frac{g_m i_T}{sC_{gs}}\right)$$

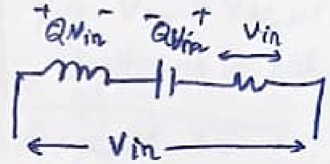
$$\therefore Z_{in} = \frac{v_T}{i_T} = \underbrace{\underbrace{\frac{1}{sC_{gs}} + sL_S}_{\substack{\text{Obvious}\\(C+L)}} + \underbrace{\frac{g_m L_S}{C_{gs}}}_{\text{Real part}}}_{\text{3 components in series}}$$

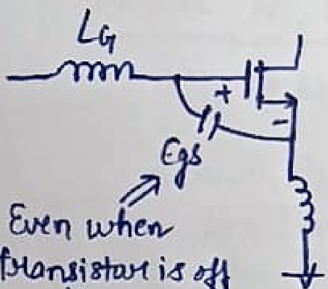
$$Re\{Z_{in}\} = \frac{g_m L_S}{C_{gs}}.$$



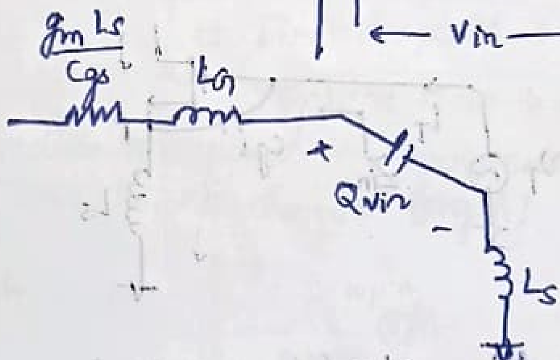
Usually L_S & C_{gs} Z don't cancel perfectly
⇓
Introduce L_g to make $Im\{Z\}=0$, $Re\{Z\}=50\,\Omega$.

At resonance,





$\equiv$



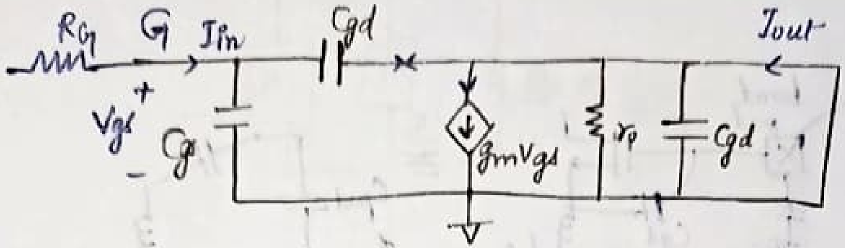
Even when transistor is off
⇓
Provides extra gain ⇒ overall good gain

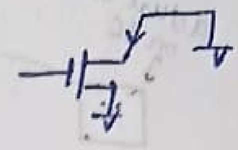
#
$$\left[j\omega (L_S + L_G) - \frac{j}{\omega C_{gs}} \right] = 0 \quad \text{(at resonance)}$$

$$\omega (L_S + L_G) = \frac{1}{\omega C_{gs}}$$

$$\omega^2 = \frac{1}{C_{gs}(L_S + L_G)} \Rightarrow \omega = \frac{1}{\sqrt{C_{gs}(L_S+L_G)}}$$

f_T : Transition frequency (for any active device)
= frequency @ short-circuit current gain = 1 [Best current gain]





$$I_{out,sc} = g_m V_{gs} + (-V_{gs} \cdot s C_{gd})$$

$$I_{in} = V_{gs} \cdot s\underbrace{(C_{gs} + C_{gd})}_{\text{in parallel}}$$

$$\underbrace{\frac{I_{out,sc}}{I_{in}}}_{\text{s.c. current gain}} = \frac{g_m - sC_{gd}}{s(C_{gs}+C_{gd})} = 1$$

$$\Rightarrow g_m = s(C_{gs} + 2C_{gd})$$

$$\omega_T = \frac{g_m}{C_{gs} + 2C_{gd}} \approx \frac{g_m}{C_{gs}}$$

$(\approx C_{ov} + \frac{2}{3} WLC_{ox}) = C_{bulk} \leftarrow C_{gs}$; $2C_{gd} \to C_{ov}$

$$\therefore \boxed{f = \frac{g_m}{2\pi C_{gs}}}$$

→ Beyond f_T (and around f_T), the device is useless.

→ f_T should be much higher (~5-10 times) than the frequency of interest.

$$\therefore Re\{Z_{in}\} = \frac{g_m L_s}{C_{gs}} = \omega_T L_s$$

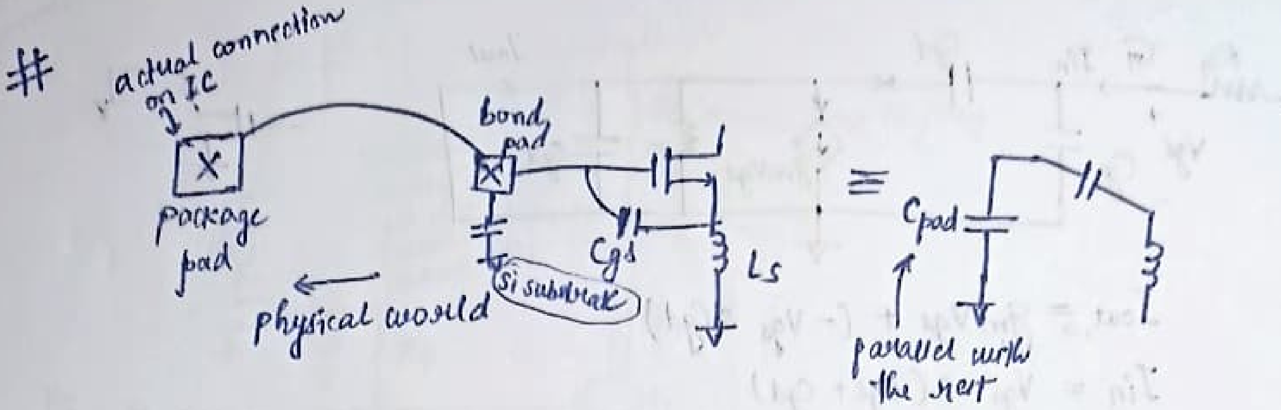
f_{max} : frequency @ power gain = 1.

$$f_{max} \approx \frac{f_T}{2\sqrt{R_G\left(g_m \frac{C_{gd}}{C_{gg}}\right) + (R_G + R_{ch} + R_s)\, g_{ds}}}$$

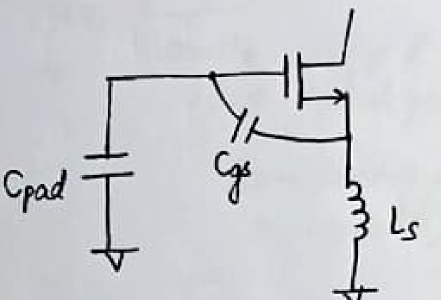
C_{gg} → total cap. at gate

R_{ch} → physical channel resistance

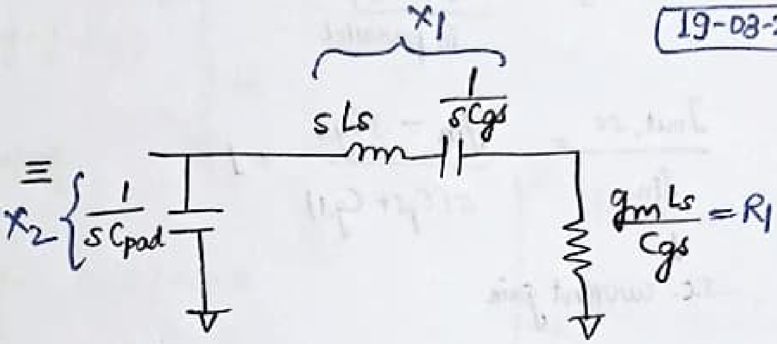
#



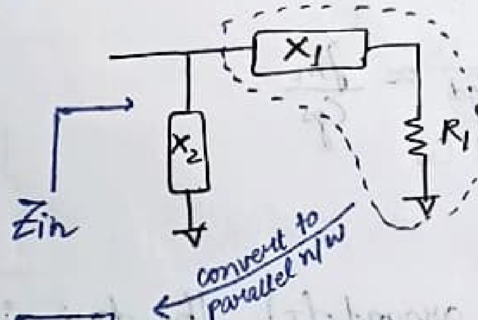
19-08-2025



$\equiv$



$\equiv$

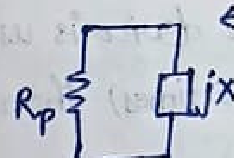


$$X_1 = j\left(\omega L_s - \frac{1}{\omega C_{gs}}\right)$$

$$X_2 = \frac{1}{j\omega C_{pad}}$$

$$R_1 = \frac{g_m L_s}{C_{gs}}$$

$(R_1 + jX_1) =$



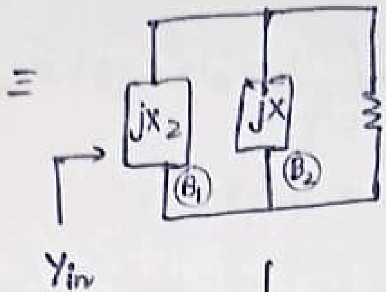
$$= \frac{R_p \cdot jX}{R_p + jX} = R_1 + jX_1$$

$$\frac{jX R_p}{(R_p + jX)} \cdot \frac{(R_p - jX)}{(R_p - jX)} = \frac{X^2 R_p}{R_p^2 + X^2} + j\frac{R_p^2 X}{R_p^2 + X^2}$$

$$\therefore \frac{X^2 R_p}{R_p^2 + X^2} = R_1,$$

$$\frac{X R_p^2}{R_p^2 + X^2} = X_1$$

$\equiv$



$R_P = \frac{1}{G_P}$

$B_1 = \frac{1}{jX_2}$,
$B_2 = \frac{1}{jX}$

$\downarrow$

$$= G_P + j(B_1 + B_2)$$

$$Z_{in} = \frac{1}{G_P + j(B_1+B_2)} \cdot \frac{G_P - j(B_1+B_2)}{G_P - j(B_1+B_2)}$$

$$\Rightarrow \mathrm{Re}\{Z_{in}\} = \frac{G_P}{G_P^2 + (B_1+B_2)^2} = \frac{1/R_P}{\left(\frac{1}{R_P}\right)^2 + \left(\frac{1}{X_1} + \frac{1}{X_2}\right)^2} \approx \left(\frac{X_2}{X_1+X_2}\right)^2 \cdot R_1$$

$$X_1 = j\Big(\underset{\substack{\text{very small} \\ (\sim 100\,pH)}}{\omega L_s} - \frac{1}{\omega C_{gs}}\Big) \sim \frac{1}{\omega C_{gs}}$$

$$\sim \Big(10^9 \times 10^{-12} - \underbrace{\frac{1}{10^9 \times 10^{-15}}}_{\text{dominate}}\Big)$$

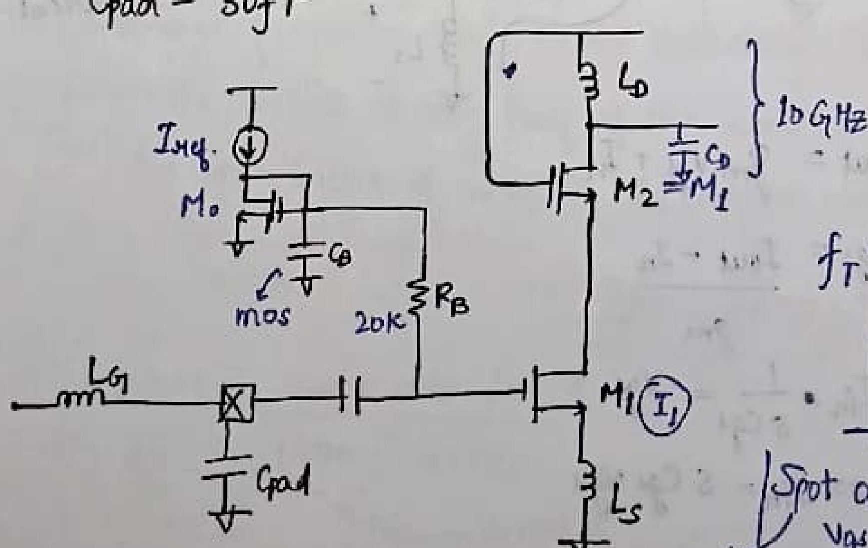
$$\therefore R_{in} = \left(\frac{C_{gs}}{C_{gs} + C_{pad}}\right)^2 \cdot \frac{g_m L_s}{C_{gs}}$$

Eg. $f_0 = 10\,GHz$

65nm tech $\Rightarrow L_{min} = 60\,nm$

$I_{ref} = 0.5\,mA$

$C_{pad} = 50\,fF$



Spot a suitable pt. for V_{gs}, I_D, $\left(\frac{W}{L}\right)$ for perfect f_T with appropriate current/power budget

[Choose / Fix V_{gs}, $I_D \left(\frac{W}{L}\right)$, ω_T, g_m, C_{gs}]

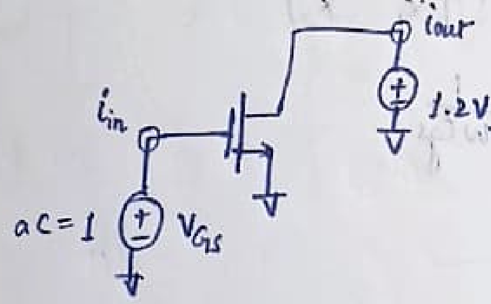
$$\underset{(50\,\Omega)}{R_{in}} = \left(\frac{C_{gs}}{C_{gs} + C_{gd}}\right)^2 \omega_T L_s \Rightarrow L_s \checkmark$$

L_G : Off-chip

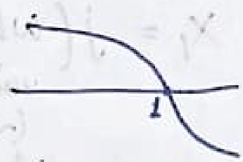
↳ To cancel the imaginary part of $\underset{50\,\Omega}{R_1} - \underbrace{jX_1}$

↳ $|j\omega L_G| = |jX_1|$

#. Measuring f_T :

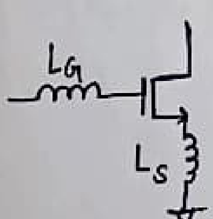


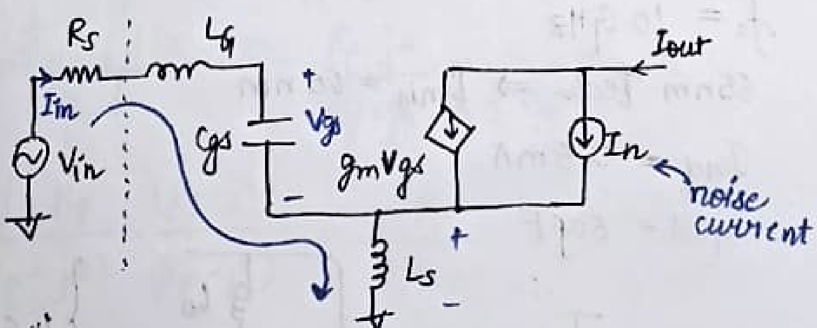
→ Measure $\frac{i_{out}}{i_{in}}$



21-03-2025

Noise Figure – CS with Inductor Degeneration





$$I_{out} = g_m V_{gs} + I_n$$

$$V_{gs} = \frac{I_{out} - I_n}{g_m}$$

$$I_{in} * \frac{1}{sC_{gs}} = V_{gs}$$

$$\Rightarrow I_{in} = s\, C_{gs} V_{gs}$$

$$V_{in} = \left(R_s + sL_G + \frac{1}{sC_{gs}}\right) sC_{gs}V_{gs} + sL_s\left(sC_{gs}V_{gs} + I_{out}\right)$$

$$= V_{gs}\left[sR_sC_{gs} + s^2(L_G+L_s)C_{gs} + 1\right] + sL_s I_{out}$$

$$= \frac{I_{out} - I_n}{g_m}\left[sR_sC_{gs} + \underbrace{s^2(L_G+L_s)C_{gs}}_{-1} + \cancel{1}\right] + sL_s I_{out}$$

At $\omega = \omega_0 = \dfrac{1}{\sqrt{(L_G+L_s)C_{gs}}}$ and $s = j\omega$; and $\dfrac{g_m L_s}{C_{gs}} = R_s$. (matching)

$$\Rightarrow V_{in} = \frac{I_{out} - I_n}{g_m}(s g_m L_s) + sL_s I_{out}$$

$$\Rightarrow V_{in} = I_{out}(2sL_s) - I_n sL_s$$

$$\Rightarrow \left|\frac{I_{out}}{V_{in}}\right| = \left|\frac{1}{2j\omega L_s}\right| = \frac{1}{2\omega_0 L_s} = \boxed{\left(\frac{\omega_T}{\omega_0}\right)\cdot\frac{1}{2R_s}} \quad (I_n = 0)\ (V_{in}, I\ out)$$

(Effective Transconductance)

→ Farther away from the f_T, higher the transconductance and hence the gain.

For $V_{in} = 0$

$$\Rightarrow I_n sL_s = I_{out}(2sL_s)$$

$$\Rightarrow \boxed{I_{out} = \frac{I_n}{2}}$$

→ At resonant frequency and perfect matching, output current from the circuit is only half of the noise current.

↳ Contribution of noise in the output current.

$$\therefore \overline{I^2_{out,n}} = \frac{\overline{I_n^2}}{4} = \frac{4KT\gamma g_m}{4} = \boxed{KT\gamma g_m}$$

$$\therefore NF = 1 + \frac{\text{noise added}}{\underbrace{(\text{gain})^2}_{\text{transconductance gain}} \cdot 4KTR_s}$$

$$= 1 + \frac{kT\gamma g_m}{\left[\left(\frac{\omega_T}{\omega_0}\right)\frac{1}{2R_s}\right]^2 \cdot 4kTR_s}$$

$$= 1 + \left(\frac{\omega_0}{\omega_T}\right)^2 \frac{kT\gamma\, g_m \cdot (4R_s^2)}{4kTR_s}$$

$$\therefore \boxed{NF = 1 + \left(\frac{\omega_0}{\omega_T}\right)^2 \gamma g_m R_s}$$

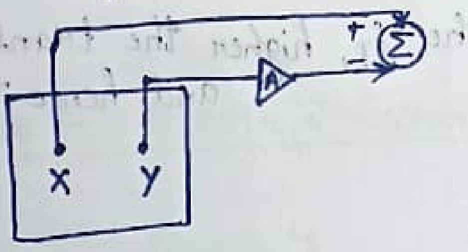
→ Moving away from ω_T, we have very good NF.

→ Choose design ~~such that dim~~ of dimensions, bias, current such that f_T is maximum (very high).

Noise Cancelling LNAs

$$NF = 1 + \left(\cancel{\begin{matrix}\text{due to} \\ \text{i/p transistor}\end{matrix}}\right) + \left(\begin{matrix}\text{due to} \\ \text{o/p load}\end{matrix}\right)$$

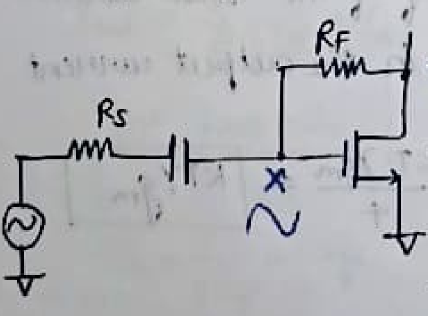
to cancel



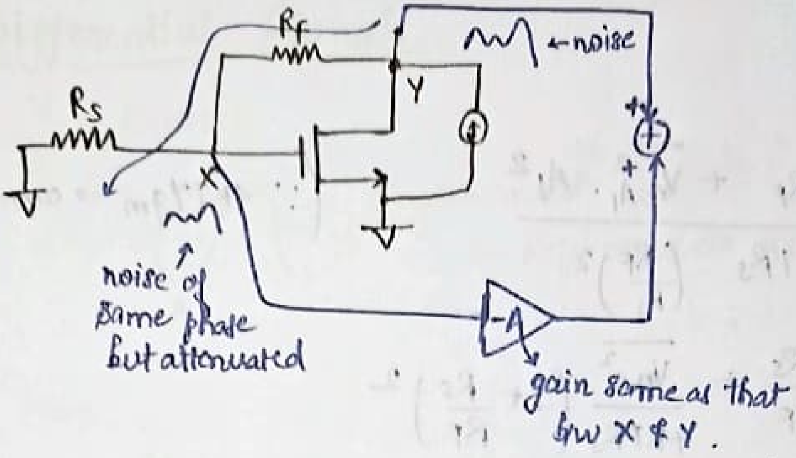
Identify two points in the circuit such that

signal → out of phase

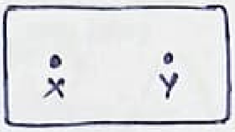
noise → in phase



∽ ↔ Inversion. → (not exactly 180° inversion for very high freq.



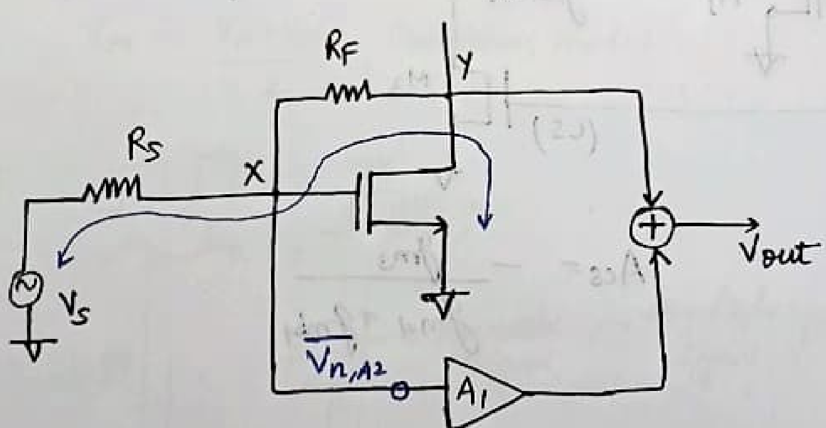
25-03-2025



Signal : Out of phase
Noise : In phase

(Signal) $\boxed{\dfrac{V_{YS}}{V_{XS}} = 1 - g_m R_F = 1 - \dfrac{R_F}{R_S} \approx -\dfrac{R_F}{R_S}}$

((Noise) : $V_S = 0$)



$$V_{XN} = V_{YN} \cdot \frac{R_S}{R_S + R_F}$$

$$\Rightarrow \frac{V_{YN}}{V_{XN}} = 1 + \frac{R_F}{R_S}$$

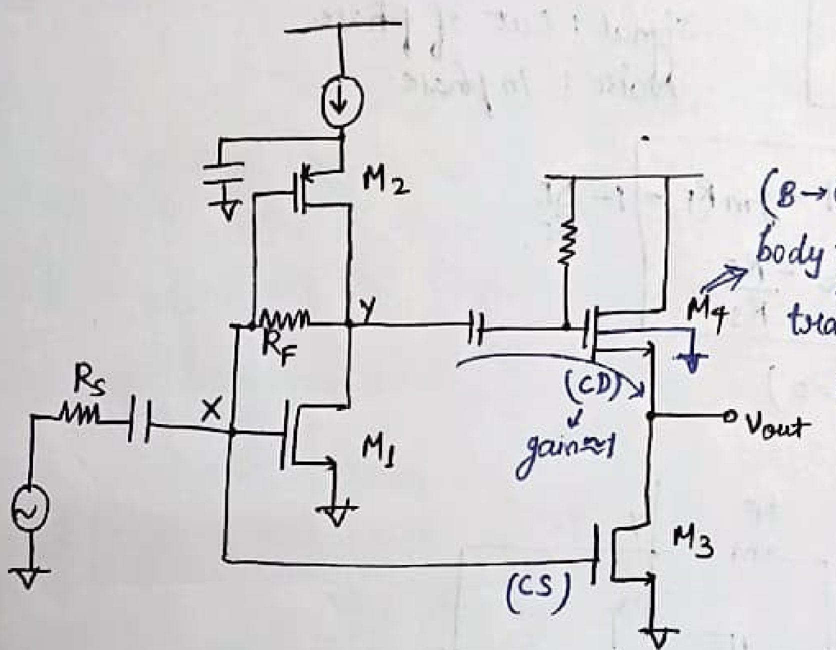
$$\therefore A_1 = -\left(1 + \frac{R_F}{R_S}\right)$$ [To cancel the noise]

$$\frac{V_{out}}{V_X} = \left(1 - \frac{R_F}{R_S}\right) - \left(1 + \frac{R_F}{R_S}\right) = -\frac{2R_F}{R_S}$$

$$\frac{V_{out}}{V_S} = \frac{-R_F}{R_S}$$

$$NF = 1 + \frac{4KTR_F + \overline{V_{n,A_1}^2} \cdot A_1^2}{4KTR_S \cdot \left(\frac{R_F}{R_S}\right)^2} \qquad \left(\because 4KT\gamma g_m \rightarrow \text{cancelled}\right)$$

$$= 1 + \cancel{\frac{R_F}{R_S}}\ \underbrace{\frac{R_S}{R_F}}_{\text{small}} + \frac{\overline{V_{n,A_1}^2}}{4KTR_S}\left(1 + \frac{R_S}{R_F}\right)^2$$



(B → GND, S ↛ GND)

body to source potential (≠0)

↓

transconductance changes

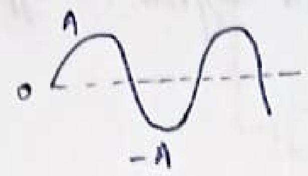
↓

overall g_m : $g_m V_g + g_{mb} \cdot V_{bs}$

$$A_{CS} = -\frac{g_{m3}}{g_{m4} + g_{mb4}}$$

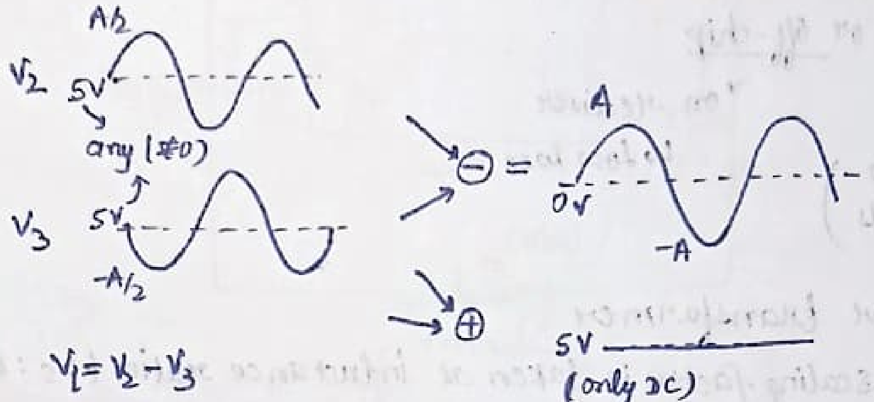
Differential Signals

$V_1 = A\sin\omega t$



single-ended

↳ Referenced at zero.



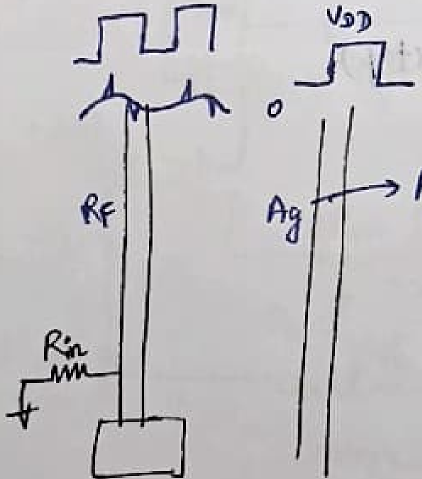
$V_1 = V_2 - V_3$

For V_A, V_B,

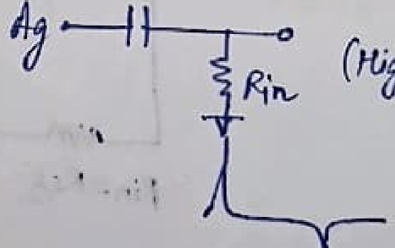
$V_d = V_A - V_B$ (differential)

$V_{CM} = \dfrac{V_A + V_B}{2}$ (common mode)

#



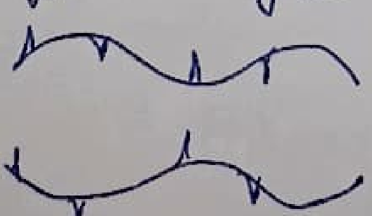
Ag → Aggresive signal → Any high signal (eg. clock).



(High pass filter)

→ corruption cannot be removed

For differential signals



Both signal corrupted by small amount

↳ Gets eliminated

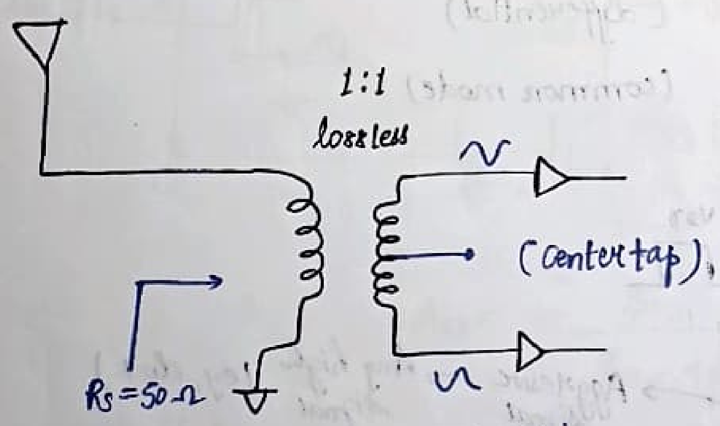
Balun

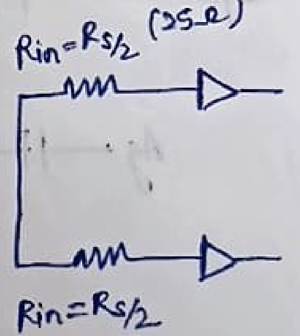
↳ Transformer that converts single-ended signal to differential signal.

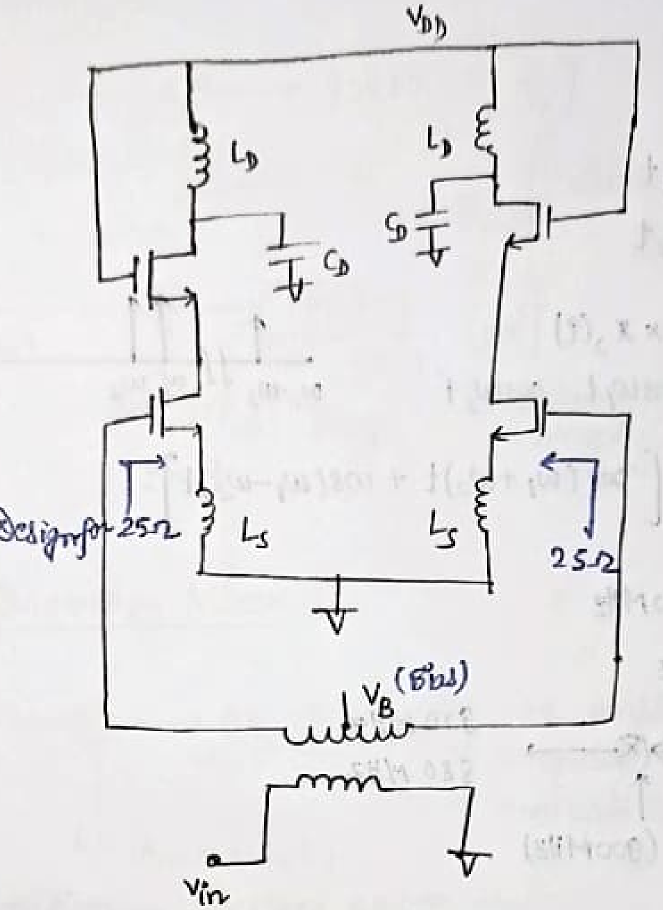
Balanced – **Unbalanced**
(Differential) (Single-ended)

↳ On-chip or off-chip
- On-chip → High loss (used for large no. receive signals)
- Off-chip → on receiver
  ↳ Low loss

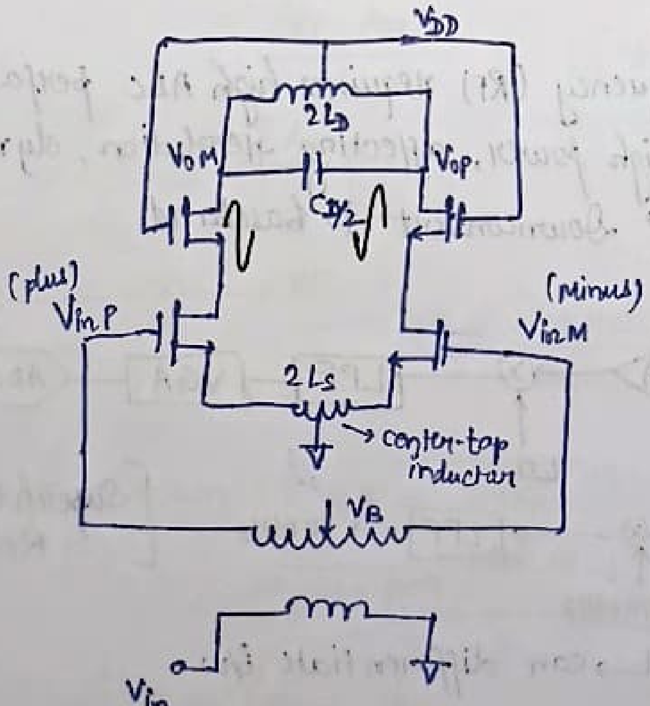
↳ Planar transformer
  ↳ scaling factor is taken as inductance ratio ($L_s : L_p$), not turns ratio.







⇓



2 LNAs
↳ double power consumption
↳ 2-times matching required

→ Double output swing

26-03-2025

MIXERS

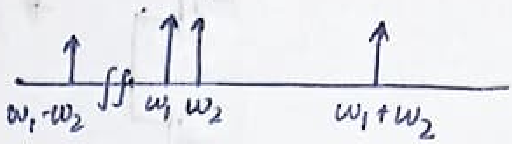
$x_1(t) = A_1 \cos\omega_1 t$

$x_2(t) = A_2 \cos\omega_2 t$

$x_{out}(t) = x_1(t) \times x_2(t)$

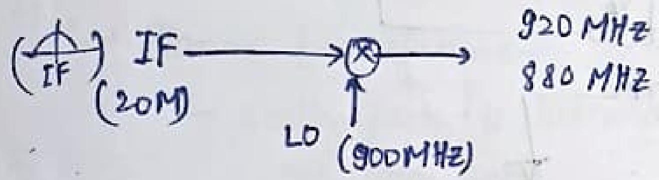
$= A_1 A_2 \cos\omega_1 t \cdot \cos\omega_2 t$

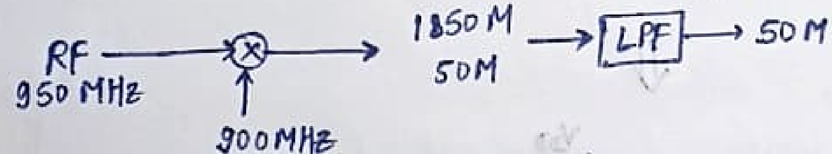
$= \frac{A_1 A_2}{2}\left[\cos(\omega_1+\omega_2)t + \cos(\omega_1-\omega_2)t\right]$



$\omega_1 = IF \approx 20\,MHz$

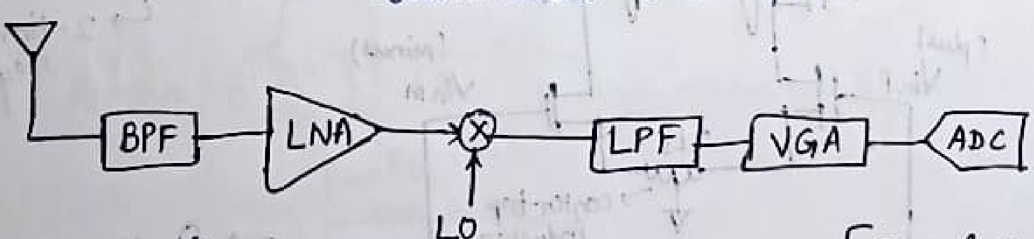
$\omega_2 = 900\,MHZ$





→ Digitizing high frequency (RF) requires high ADC performance and thus high power, affecting resolution, dynamic range.

↳ Downconvert to baseband



[Superheterodyne Receiver]

Close-by freq.

↳ can differentiate b/w 950 & 850 MHZ.

↳ Produce image frequency.

↳ use image reject filter.

$f_{LO} = f_{RF}$ — LO — 920 MHz — $\otimes$ → 920M [Tx]

[Direct conversion OR Zero IF]

RF (920M) → $\otimes$ → LPF → (DC) → Baseband centered at zero. [Rx]

$f_{LO} = f_{RF}$ (920M) "Image"

Down Conversion Mixer

RF ($A_{RF}\cos\omega_{RF}t$) → $\otimes$ → BB ; LO ($A_{LO}\cdot\cos\omega_{RF}t$)

→ Not an amplifier, as i/p & o/p are at completely different frequencies (in amplifier, i/p & o/p freq. are same)

$$\text{Conversion gain} = \frac{\text{Desired BB/IF o/p}}{\text{Input RF}}$$

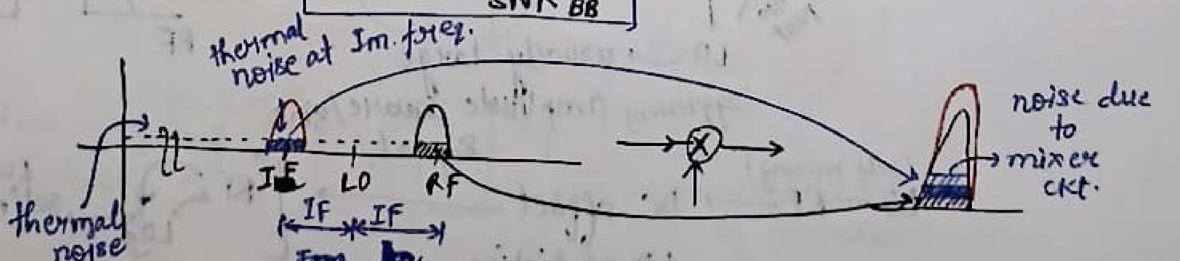
$$= \frac{A_{RF}\cdot A_{LO}/2}{A_{RF}} \qquad \left(\text{o/p}: \frac{A_{RF}A_{LO}}{2}\left[\cos(0)t + \cos(2\omega_{RF}t)\right]\right)$$

$$= \boxed{\frac{A_{LO}}{2}}$$

BB → $\otimes$ → RF ; LO

$$CG = \frac{\text{Desired RF o/p}}{\text{Input BB/IF}}$$

Downconversion : $\boxed{NF = \frac{SNR_{RF}}{SNR_{BB}}}$



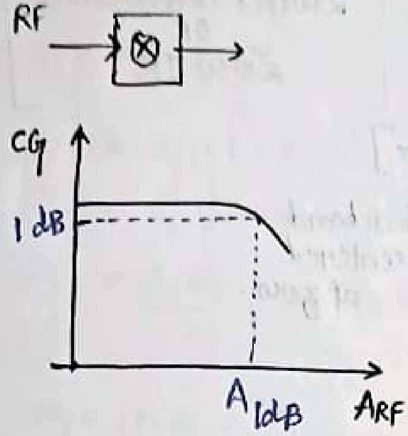
DSB NF → double signal power
(Double sideband NF)

NF → SSB NF
(Single sideband NF)
↳ Don't take signal at IF.

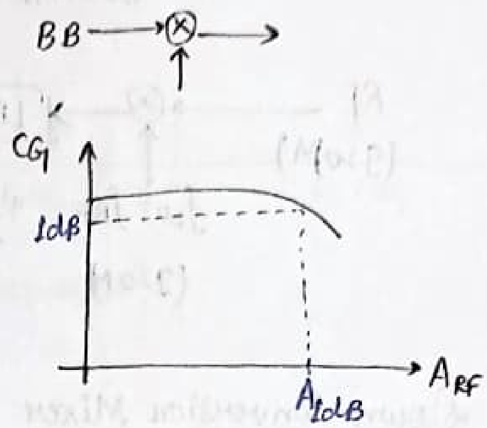
→ $\boxed{\text{DSB NF} < \text{SSB NF by 3dB}}$

1-dB ~~conversion~~ compression point of mixers:

Downconversion:

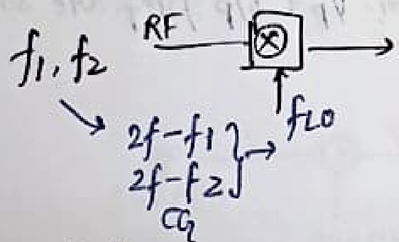


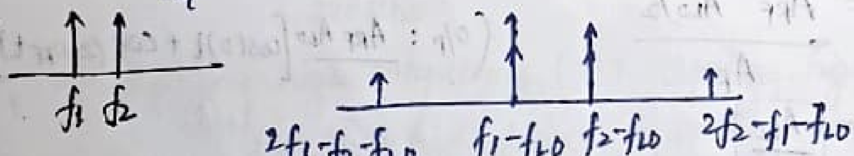
Upconversion:



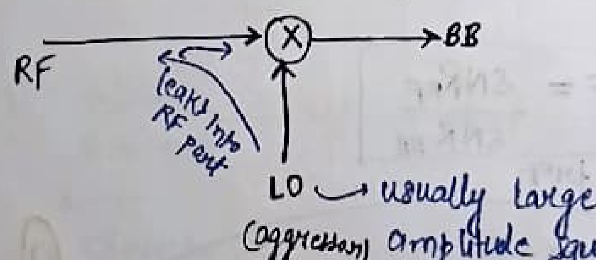
IIP3

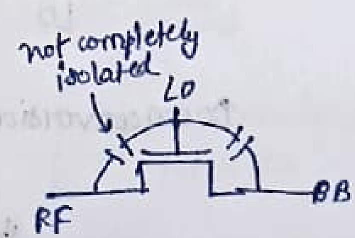
Downconversion:





Feed Through / Port Leakage





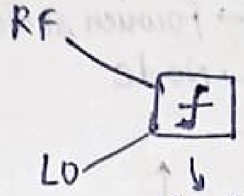
LO → usually large (aggressor) amplitude square/sine signal

(self mixing) LO — RF ⟶ DC offset ⟶ [RF ↶ LO ⊗ → LO selfmixing]

LO radiation

Oscillator pulling → strong RF pulls LO

Classical Microwave technique ~~way~~ of Mixing:



non-linear system
e.g., square

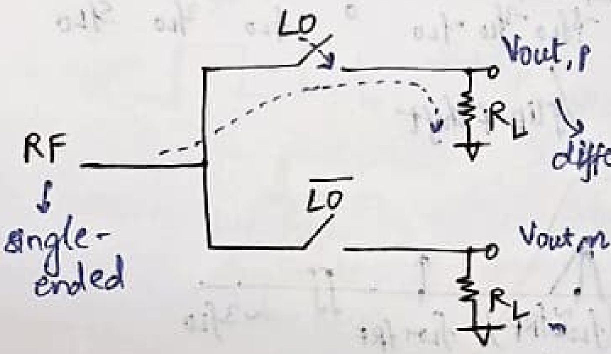
$\underbrace{x^2+y^2}_{\text{filter out}} + 2xy$

$RF \xrightarrow{\text{single ended}} \otimes \longrightarrow BB/IF$ (differential), with f_{LO} into $\otimes$

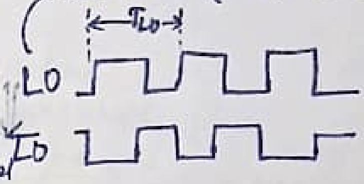
⇓

only one port is balanced
⇓
single balanced

Passive Mixers



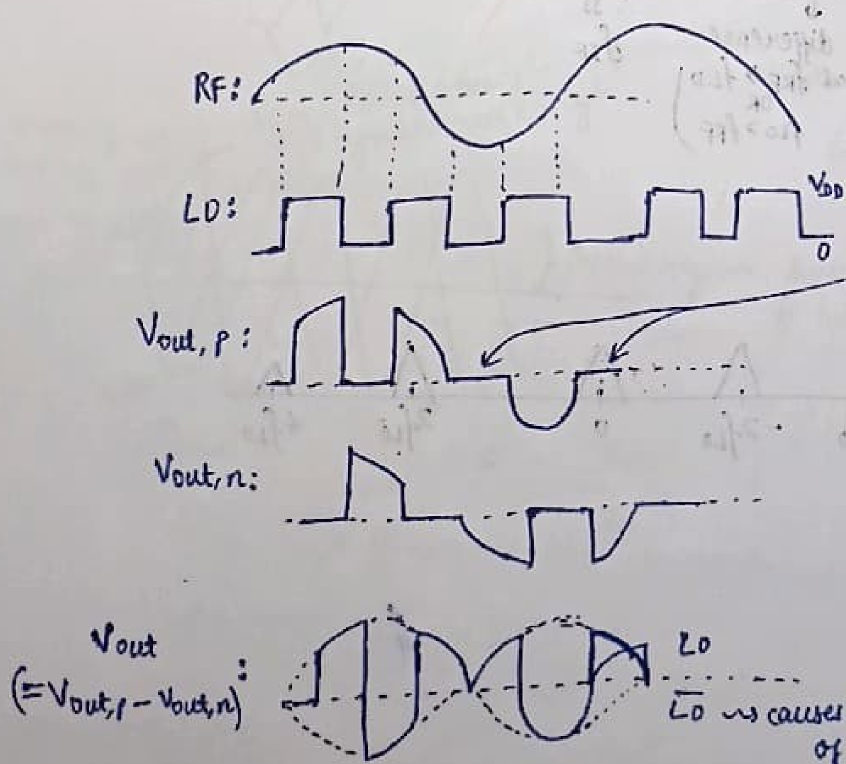
clock → square waves (or large amplitude sine waves)



→ Single Balanced Mixer

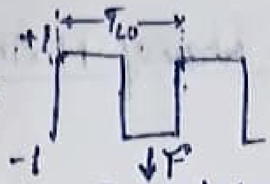
$CG = \frac{2/\pi}{2} = \frac{1}{\pi}$

↳ "Return-to-Zero" Mixer



LO
$\overline{LO}$ ⇝ causes inversion of sine wave

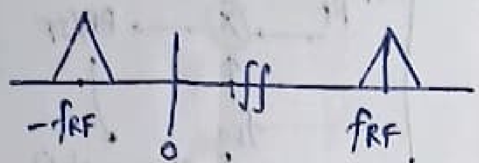
$V_{out} = V_{RF} \times$

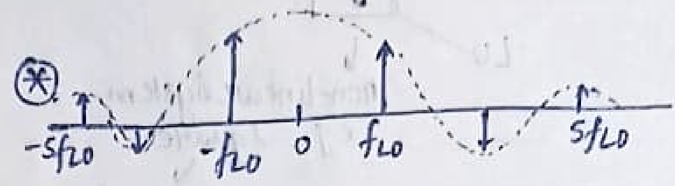


→ periodic → 50% duty cycle → only odd harmonics

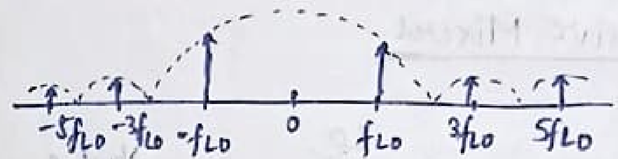
↳ Fourier series ✓

↳ No dc

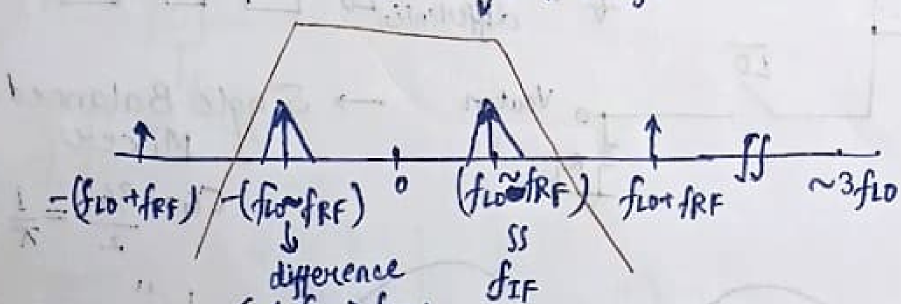




only look at the magnitude (not sign)



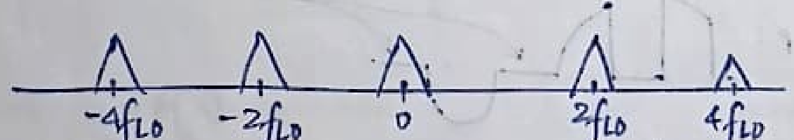
flip & shift



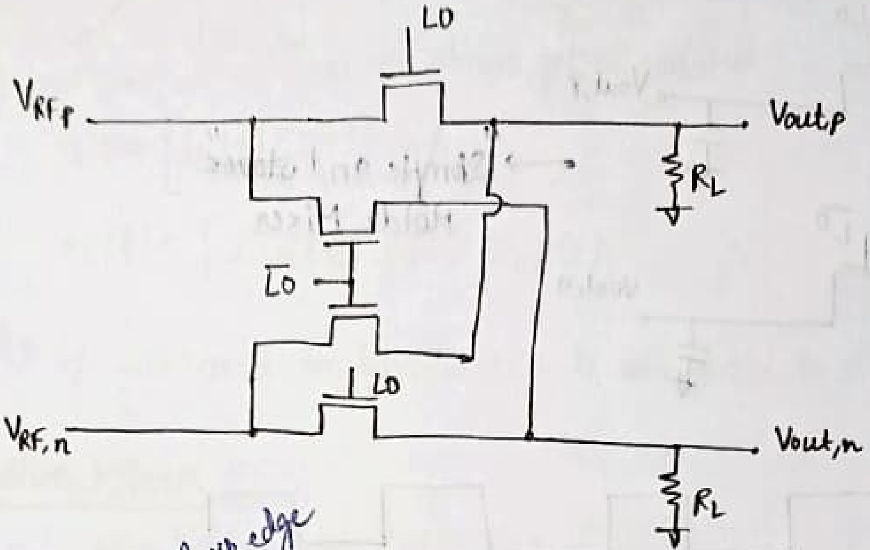
difference (as $f_{RF} > f_{LO}$ OR $f_{LO} > f_{RF}$)

f_{IF}

<u>$f_{LO} = f_{RF}$:</u>

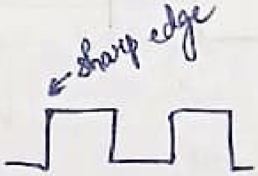


Double Balanced Passive Mixer



LO : $p \rightarrow p$
$n \rightarrow n$

$\overline{LO}$: $p \rightarrow n$
$n \rightarrow p$



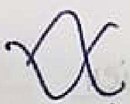
→ difficult to produce perfect square wave

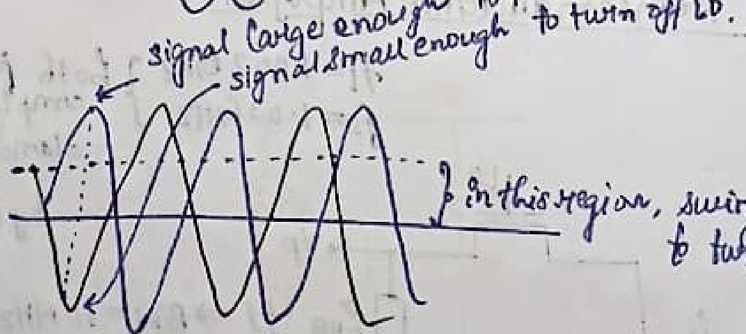
↳ Requires 3rd, 5th harmonic to get something like square wave ⇒ Requires 30 GHz, 50 GHz

↳ ~~needs in~~ difficult to get without inductive tuning

⇓

Use high amplitude sine wave

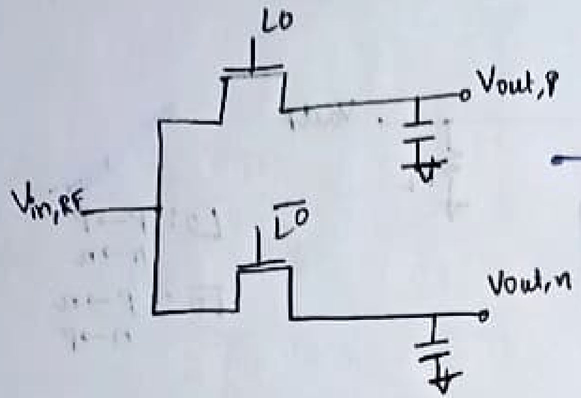




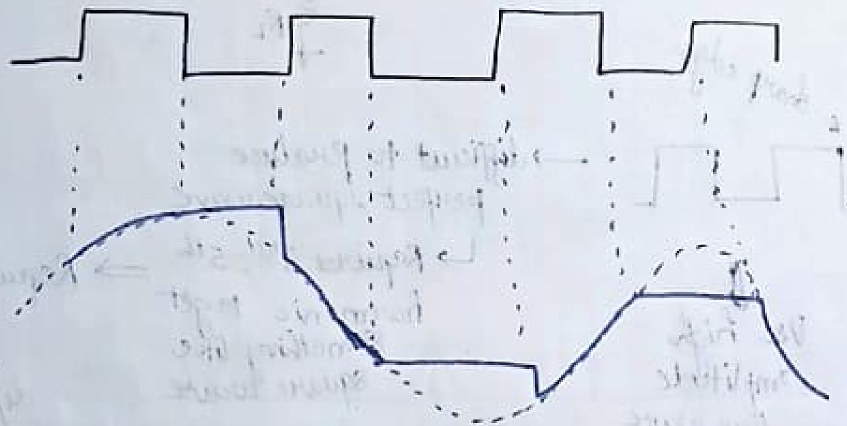
} in this region, swing is not sufficient to turn off $\overline{LO}$, or LO.

28-03-2025

Passive Mixers: Sampling Mixer



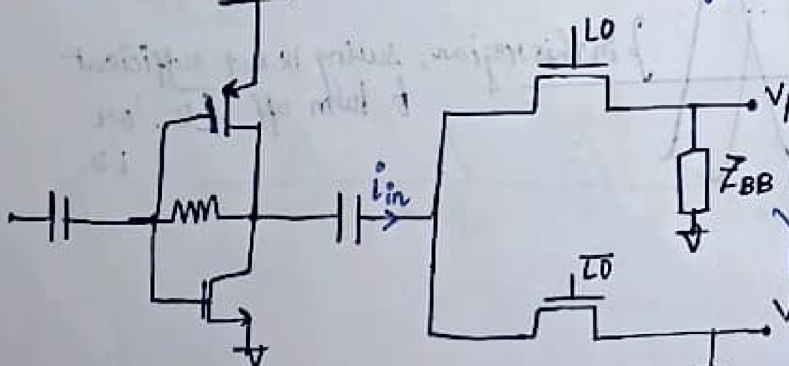
$\longrightarrow$ "Sample and ~~store~~ Hold Mixer"



LNTA: Low Noise Transconductance Amplifier

$f_{RF} = 1\,GHz$

$f_1 = 1.002\,GHz$, $f_2 = 1.03\,GHz$ } both passes & amplified by almost same by LNTA

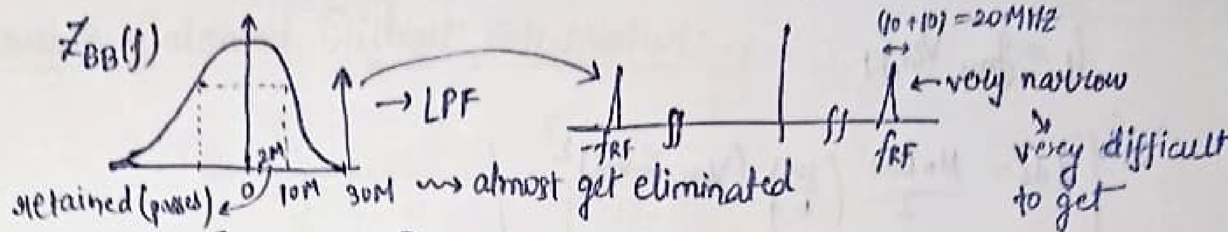


$Z_{BB}(f) \Rightarrow BW = 10\,MHz$ (LPF)

$\searrow$ baseband impedance

$Z_{BB}(f) \Longleftrightarrow h(t)$

$\Rightarrow s(t)$



$$V_1(t) = [i_{in}(t) \times S(t)] \circledast h(t)$$

$$V_1(f) = [I_{in}(f) \circledast S(f)]\, Z_{BB}(f).$$

→ As if baseband ~~bm~~ bandwidth is converted to RF.

Active Mixer

RF ↳ To get conversion gain

↓

$\boxed{V \to I}$ G_m (transconductance)

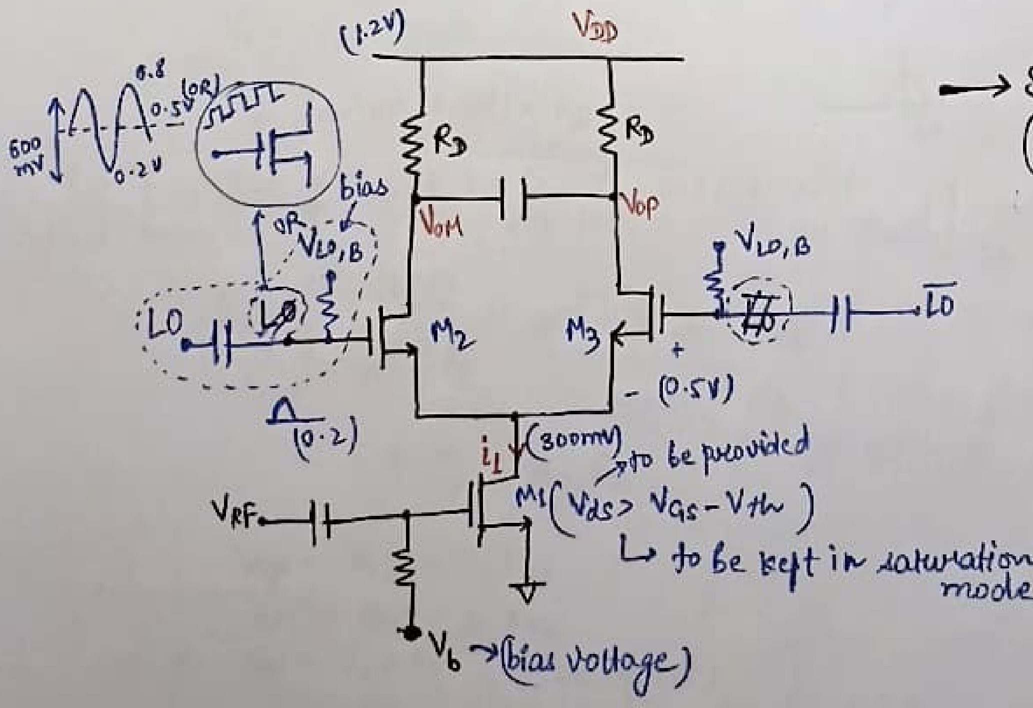
↓

~~I → V~~

Commutation (One phase of LO goes other, other phase gets inverted) ⇝ like taking passive filter, rotating it and fitting it in

↓

$\boxed{I - V}$



→ Single Balanced (Gilbert cell) Mixture

$i_1 = g_{m1} V_{in,RF}$

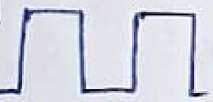
$$\left[\begin{array}{l} I_{d1} = \frac{\mu_n C_{ox}}{2} \left(\frac{W}{L}\right)_1 (V_{GS_1} - V_{th})^2 \\ V_{op} = V_{OM} = V_{DD} - \frac{I_{d1}}{2} R_D \end{array}\right]$$

$$V_o = i \times S(t)\, R_D$$

$$= g_m V_{in,RF} \cdot R_D \cdot S(t)$$

$$= g_m V_{in,RF}\, R_D \cdot \frac{4}{\pi} \cos \omega_{LO} t$$

$S(t) =$ +1 / -1

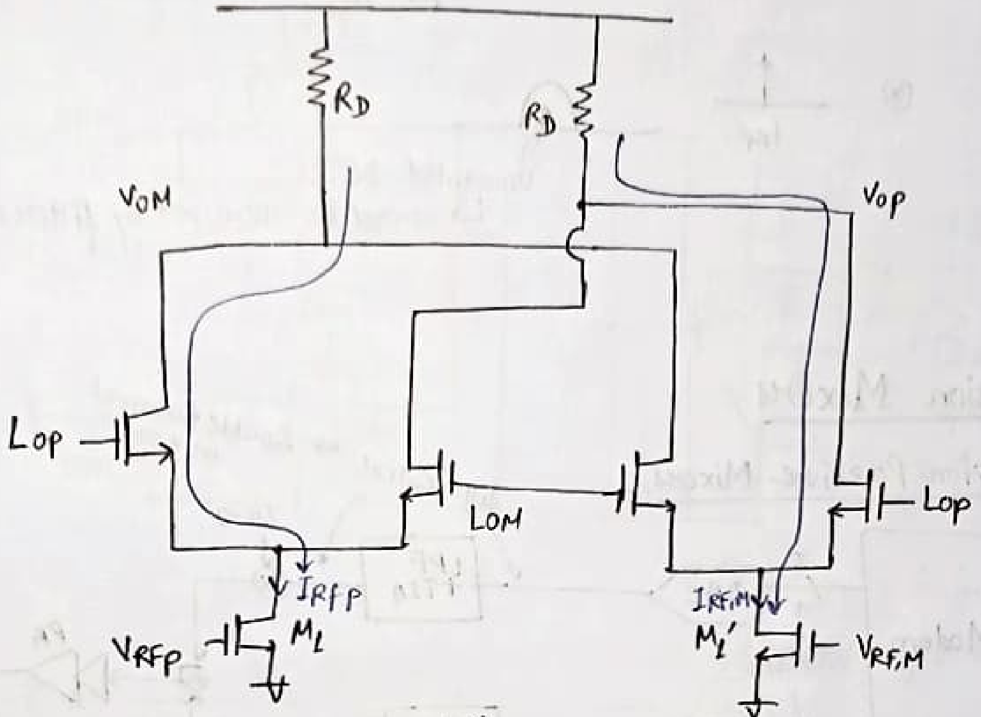


$$V_{in,RF} = A \cos \omega_{RF} t.$$

$$\therefore CG = \frac{2}{\pi} g_m R_D$$

01-04-2025

Double Balanced Gilbert cell Mixture



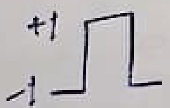
$I_0 + g_m V_{RFP} = I_{RFP}$

$V_{RF} = V_{RFP} - V_{RFM}$

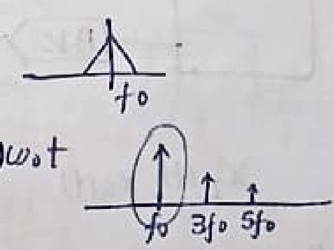
$I_{RF} = I_{RFP} - I_{RFM} = g_m (V_{Rf})$

$V_0 = V_{OP} - V_{OM}$

$= g_m V_{RF} \times S(t) \times R_D$



$= \frac{4}{\pi}\cos\omega_0 t + \sum_{n=2}^{\infty} \frac{4}{1 2\pi} \cos(2n-1)\omega_0 t$



(LPF)

$= g_m V_{RF} \frac{2}{\pi} R_D$

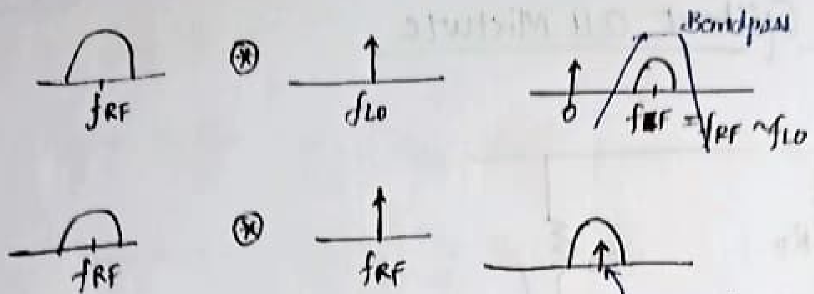
$\therefore$ C.G. $= \frac{2}{\pi} g_m R_D$

$V_{OP} = V_{DD} - I_0 R_{D1}$

$V_{OM} = V_{DD} - I_0 R_{D2}$

$V_0 = I_0 \Delta R_D$. (DC offset)

↳ due to mismatch b/w 2 R_D's (2 g_m's, 2 I_0's)

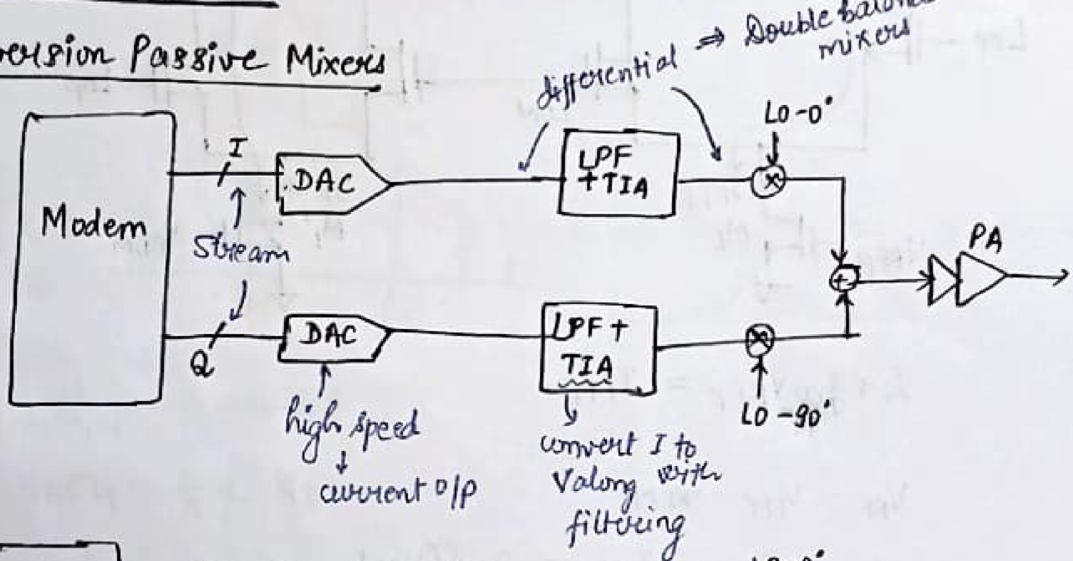


Unwanted DC
↳ cannot be removed by filters

Upconversion Mixers

Upconversion Passive Mixers

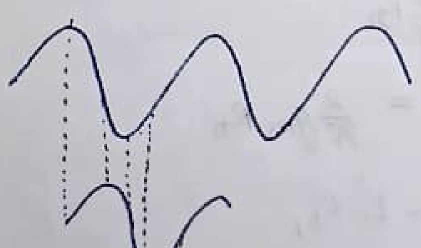
System :



$x_I(t)$, $x_Q(t)$

$V_{RF,out}(t) = x_I(t) \cos \omega_{LO} t + j x_Q(t) \sin \omega_{LO} t$

$\cos \omega_{LO} t$:

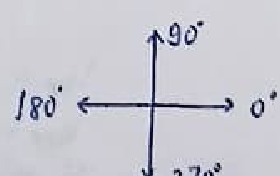


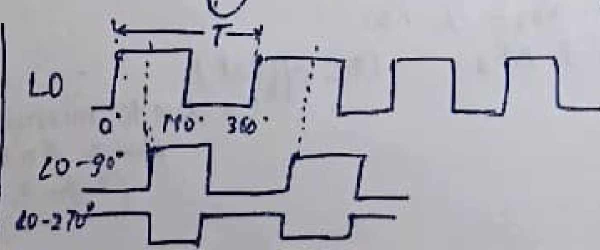
$(\cos \theta)$

$\theta = \frac{t}{T} \times 360°$

OR

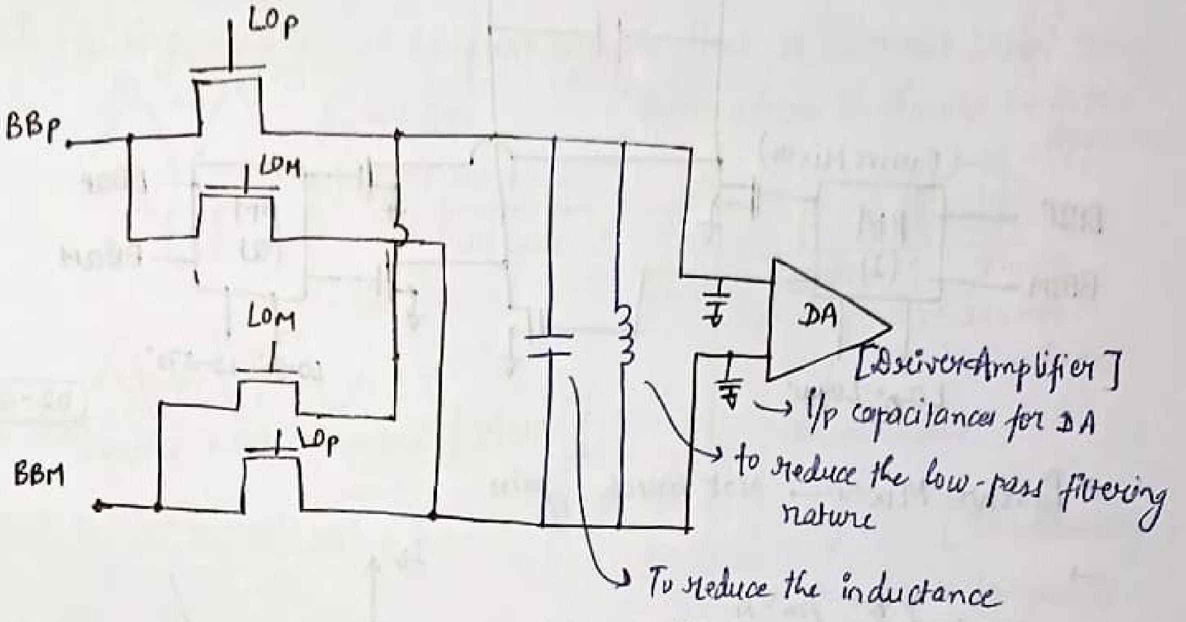
$\frac{t}{T} \times 2\pi$

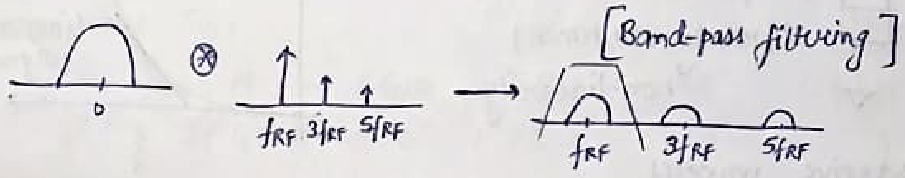




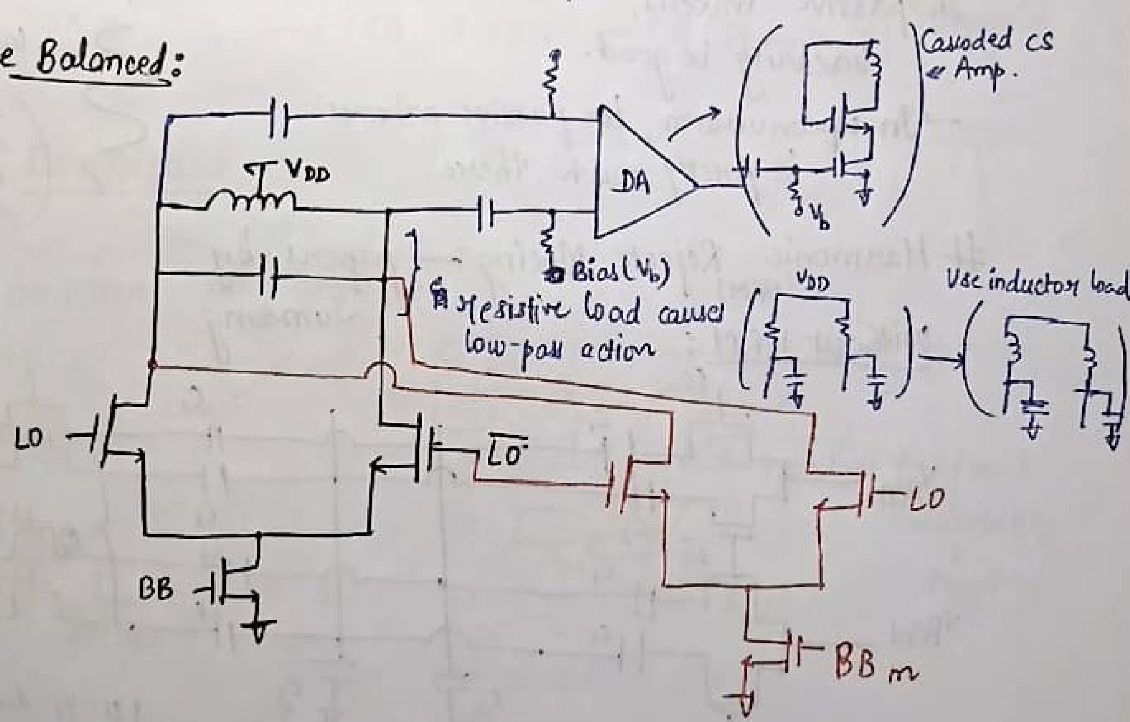
$90° \Rightarrow t = T/4$

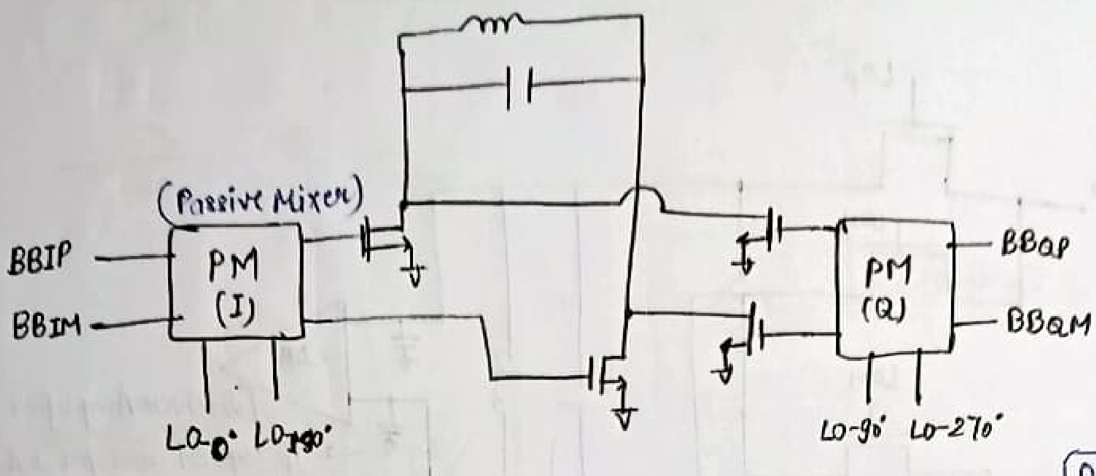
Upconversion Passive Mixers





Single Balanced:

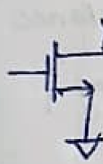




02-04-2025

→ Passive Mixer → Not much gain.

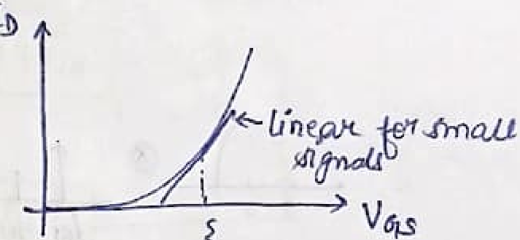
→



$i_D = g_m V_{RF}$

(transconductance)

→ non-linearity



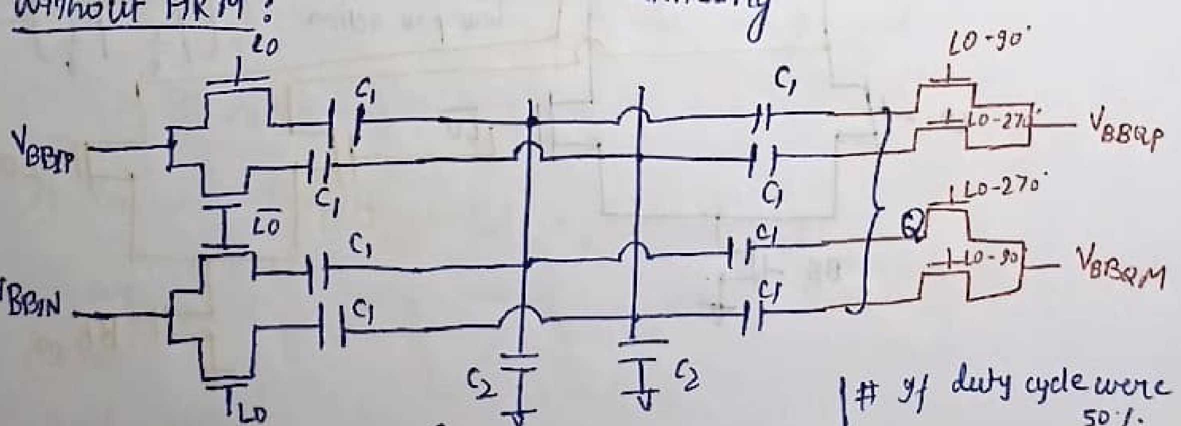
In passive mixers,
linearity is good.

→ In upconversion, ~~in~~ passive mixers is pretty much there.

→ large signals ⇓ Not linear (due to inherent behaviour of the device)

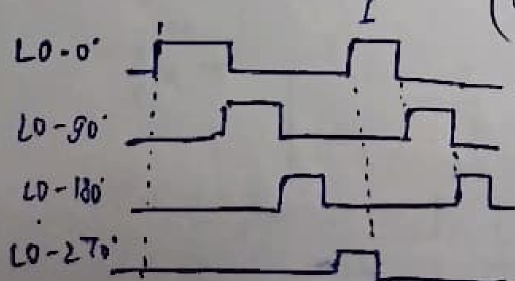
Harmonic Reject Mixing (HRM) — papers by Samsung

<u>Without HRM</u>:



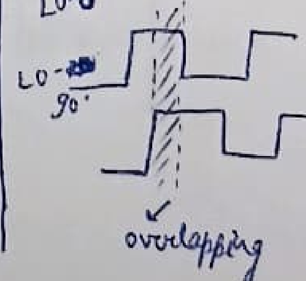
I

(Capacitance division ratio)



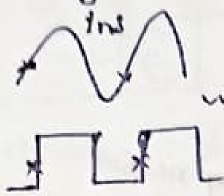
(25% duty cycle LO)

If duty cycle were 50%.



Synthesizer

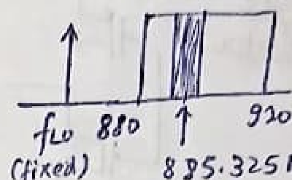
① Purity → should be pure enough that it does not drift away from where it should be in the spectrum



$f_0 = 1\,GHz$

⇝ perfect 1ns time period for all cycles

1.0001ns 0.998ns 1.003ns [Jitter]



↳ stay here

↓

Downconvert

↓

Digitize [Traditionally]

↳ No longer used

For. Mixers

(Square LO)

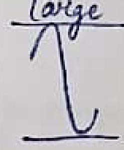
Commercial 5Gs:

560 M → 6 GHz : FR1 → 4G

24G → 40G : FR2 → 5G

\# 7G → 15G : FR3 → 6G (Upcoming)

Sinusoid LO ↓ Large



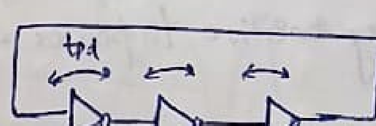
(RF) Oscillators

LC oscillators | Ring Oscillators

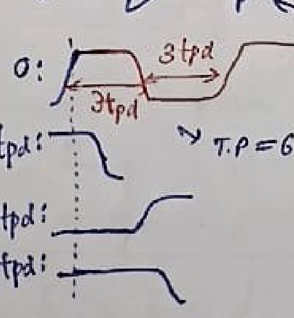
N stage RO:

$2N T_{pd} = T$

$\Rightarrow f_0 = \frac{1}{2N T_{pd}}$



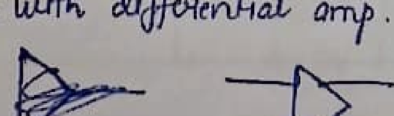
← inverters (odd no.) ↓ instability ↓ toggling



↘ T.P = 6 tpd

Even-no.-stage oscillator

↳ possible with differential amp.



RO

↳ Tiny ↳ Direct square wave o/p [Very Attractive]

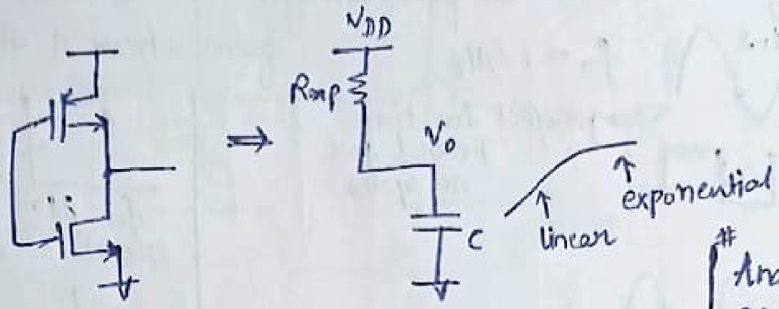
↳ All n-stages available.

RO → still not used in RF, due to:

① Zitter noise ↑

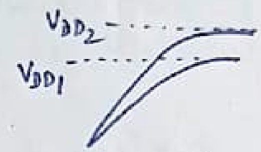
② Supply sensitive

<u>Single inverter</u>:



Analog V_{DD} → very noise
Digital VDD (DVDD)

$$V_0(t) = V_{DD}\left(1 - e^{-t/R_{on}C}\right)$$



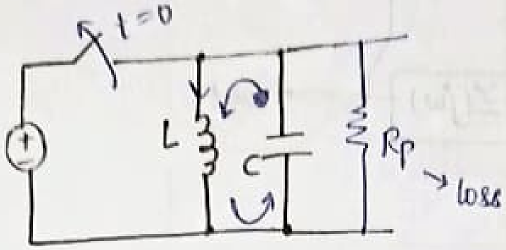
If V_{DD} changes ⇒ t_{pd} changes

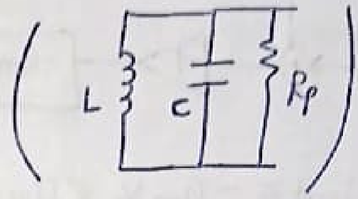
- Power supply induced Zitter (→ very small) ↔ Bad phase noise in freq. domain [Horrible Zitter]
  ↳ Due to noise in supply
  ⇓
  Changes tpd
  ⇓
  Changes f_0

Dynamic power : CV^2f ← to reduce power [Dynamic voltage & frequency scaling]
(reduce voltage) (reduce freq.)

→ RO → very sensitive to power supply.

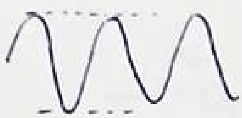
LC Oscillation





(L charges C and C charges L)

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$



→ same o/p. amplitude for no loss

→ decay if there is a loss R_p
~~(decay ↑ for R_p ↑)~~

$$Q = \frac{R_p}{\omega_0 L}$$

$Q\uparrow, R_p\uparrow$ → underdamped, prolonged riging

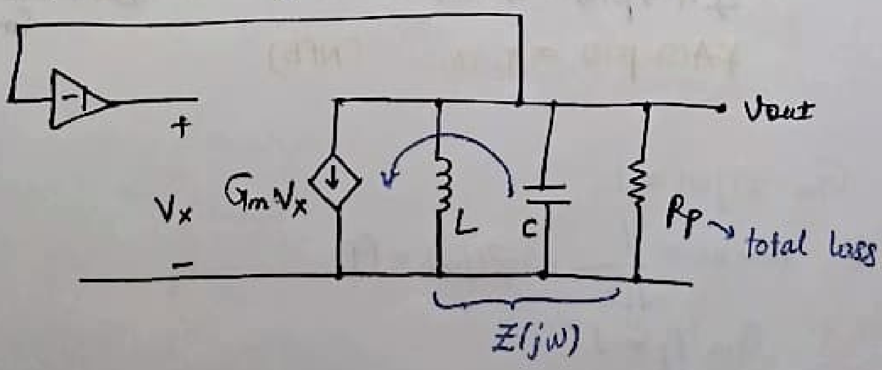
$Q\downarrow, R_p\downarrow$ → quick decay

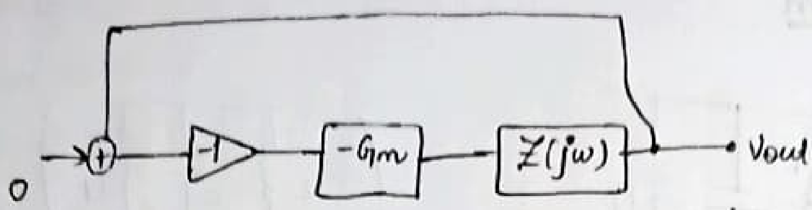
Sustained oscillation → if energy feed to the system to compensate for the loss

If more energy is given ⇒

↓
overdamped

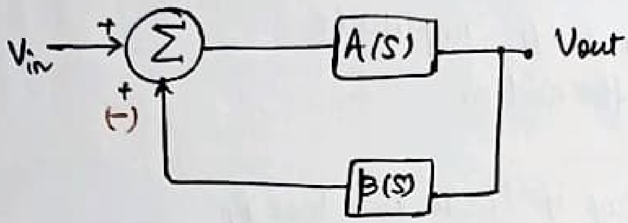
Feed energy → using transistor → +ve feedback





$$V_{out} = -G_m V_x Z(j\omega)$$

$$\text{Loop gain} = G_m \cdot Z(j\omega)$$



$$V_{out} = \left[V_{in} + \beta(s) \cdot V_{out}(s)\right] A(s)$$

$$V_{out}(s)\left[1 - A(s)\beta(s)\right] = V_{in} \cdot A(s)$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{A(s)}{1 - A(s)\beta(s)} \quad \text{(Positive feedback)}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{A(s)}{1 + A(s)\beta(s)} \quad \text{(Negative feedback)}$$

For the system to become oscillator,

$$D^r = 0 \Rightarrow TF \to \infty$$

$$A(s) \cdot \beta(s) = 1 \to \text{(PFB)}$$

$$A(s) \cdot \beta(s) = -1 \longrightarrow \text{NFB}$$

$$|A(s)\,\beta(s)| = 1$$

$$\measuredangle A(s) \cdot \beta(s) = 0, 2\pi, \ldots \quad \text{(PFB)}$$: "Barkhausen's Condition"

$$\measuredangle A(s) \cdot \beta(s) = \pi, 3\pi, \ldots \quad \text{(NFB)}$$

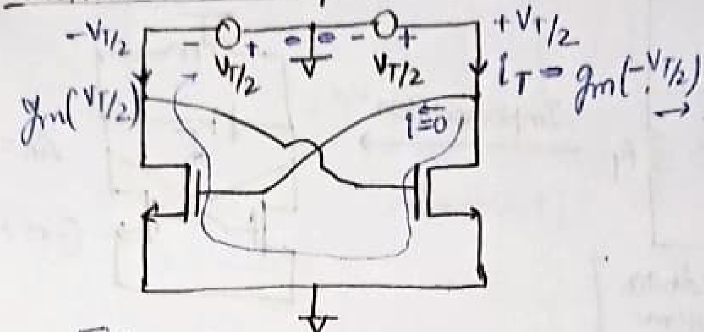
$$G_m \cdot Z(j\omega) = 1$$

At $\omega = \omega_0 = \frac{1}{\sqrt{LC}}$, $Z(j\omega) = R_P$

$$G_m \cdot R_p = 1$$

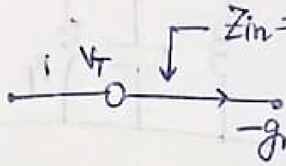
$$\Rightarrow \boxed{G_m = \frac{1}{R_p}} \longrightarrow$$ More energy requirement for more loss in the system

Differential cross-coupled Pair:



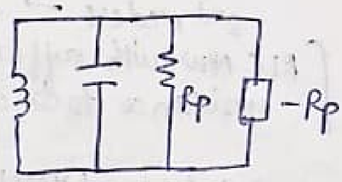
→ D & G are cross-connected

[Assumption: Transistors are already biased]

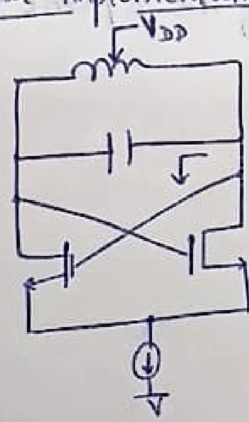


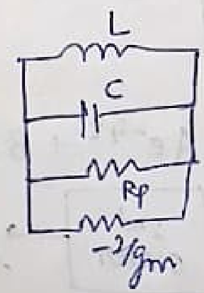
$$Z_{in} = \frac{V_T}{-g_m V_T/2} = -\frac{2}{g_m}$$

(negative impedance)



Actual implementation:

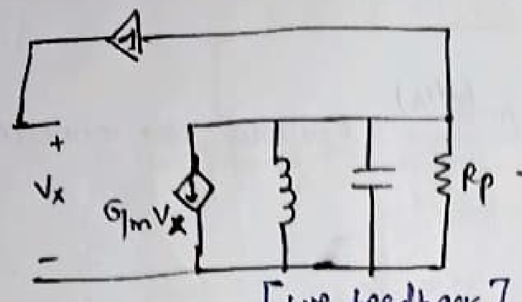




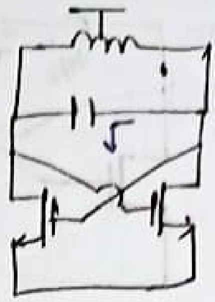
[If Rp & $-2/g_m$ cancel each other perfectly, we get L, C oscillation]

↳ cancels the loss of the system

04-04-2025



Implementation →

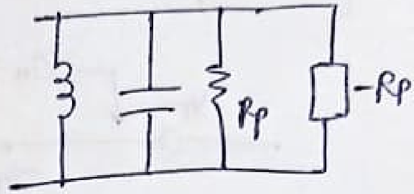


$Z_{in} = -\frac{2}{g_m}$
(−ve resistance)

[+ve feedback system]

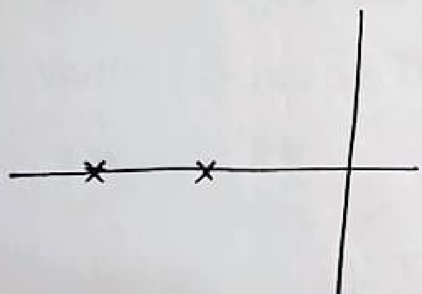
$G_m = \frac{1}{R_p}$

$G_m R_p < 1$: Dying oscillation
$G_m R_p > 1$: Exponentially increasing oscillation



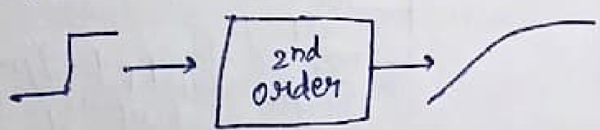
$G_m Z(j\omega) = 1$

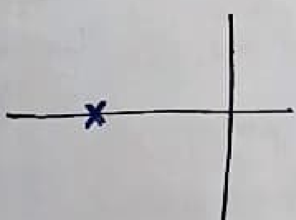
↓
2nd order → 2 poles
[RLC N/W with sufficiently large -ve resistance to cancel the loss]

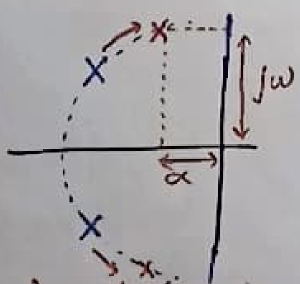


→ 2 poles on (real) LHP of the Z-plane

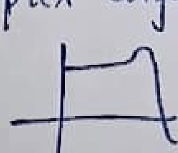
time-domain soln: $Ae^{-s_1 t} + Be^{-s_2 t}$





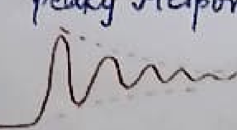


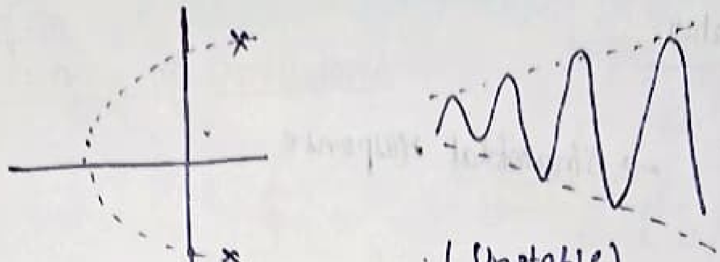
→ complex conjugate poles



peaky response

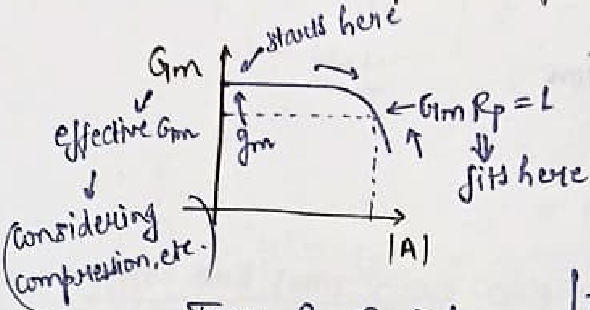
As poles move towards $j\omega$ axis
poles: $-\alpha \pm j\omega$
soln: $e^{(-\alpha \pm j\omega)t} \Rightarrow \alpha \downarrow$
⇓
slow dying

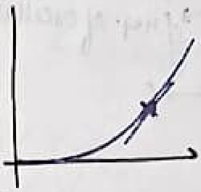




(Unstable)

<u>Undesirable properties of the device</u> → Non-linearity





$$\left[G_m = -\frac{2}{g_m}\right]$$

Take $G_m R_p > 1$ $\quad \left|\frac{2}{g_m}\right| > R_p$

↳ Oscillator → infinite response (no i/p required)

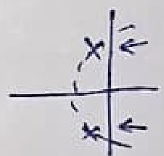
↳ starts with noise in the device

↳ Poles on RHP.

(Temp. variation ⇒) $G_m \downarrow$ ⇒ $R_p \downarrow$ ⇒ $G_m R_p$ becomes < 1

⇒ Signal amplitude ↓

 (oscillation ↓)



(poles move to LHP)

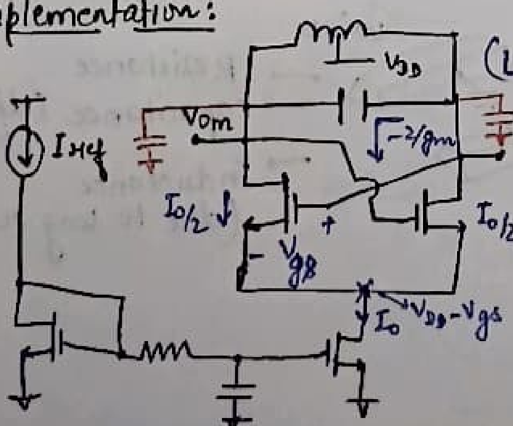
G_m tries to go to the linear region.

⇓

Fits to $G_m R_p = 1$ (due to feedback)

$\cancel{\left|\frac{2}{g_m}\right| \times R_p > 1}$ ~~(to start the oscillation)~~

<u>Implementation:</u>



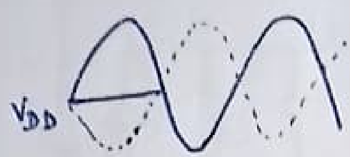
(LC tank) → made to oscillate

(extra capacitors) → oscillation freq. changes

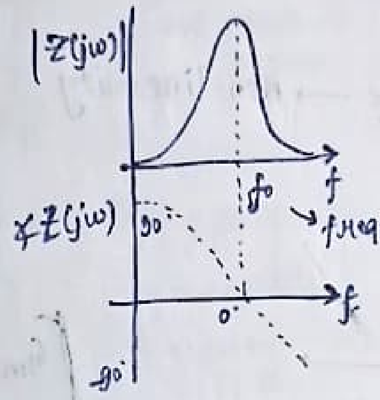
$$g_m = \sqrt{2 I_D \mu_n C_{ox}\left(\frac{W}{L}\right)}$$

$$\propto \sqrt{I_D}$$

Diff. Cross-Coupled Oscillator

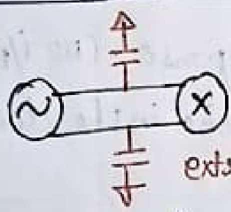


→ Sinusoidal Response



→ freq. of oscillation

Oscillator connected to mixer:



→ Never connect oscillator to any load directly

⇓

Add buffer

⇓

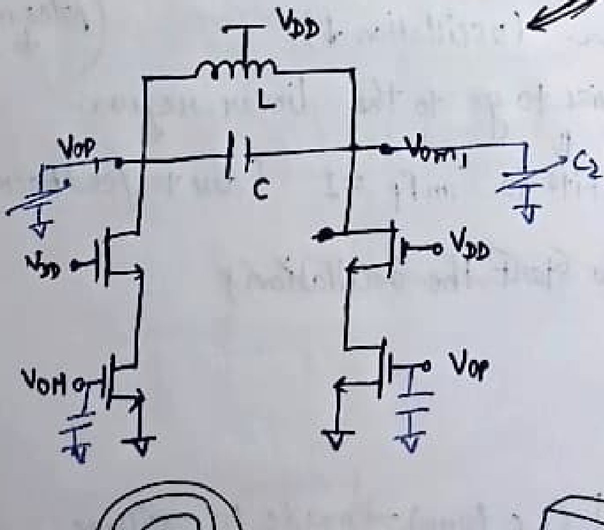
Always design oscillator with a buffer

(use cascoding)

↳ Can connect to anything.

↓

freq. shifts a/c to load



(tiny transistors)

wires

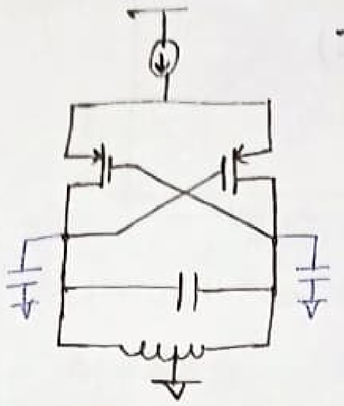
→ Resistance

→ capacitance (dielectric in b/w)

→ Inductance (due to long routing)

08-04-2025

Tuning of Oscillators



→ For PMOS:

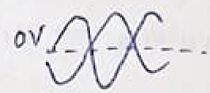
$\mu_p < \mu_e$ (mobility)

↓ requires larger width ↑

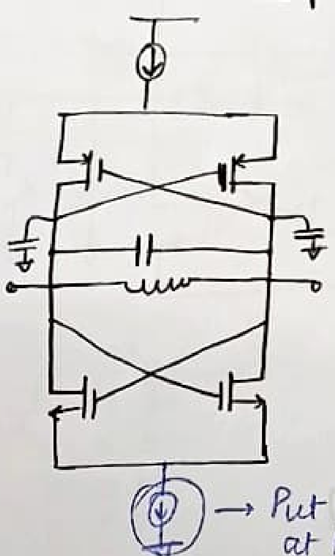
$g_m = \mu C_{ox} \left(\frac{W}{L}\right)(V_{GS} - V_{Th})$

↓ For same g_m, needs to burn more current.

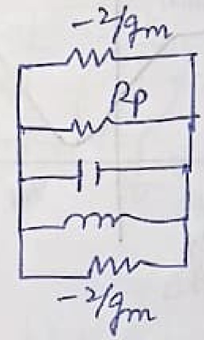
↳ More capacitance



NMOS and PMOS cross-coupled:



→ Put either at top or at bottom

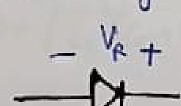


→ More -ve resistance

~~Digital~~ / Electronic Tuning: (Not manual)

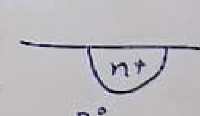
↳ Look for voltage / bits-controller capacitance.

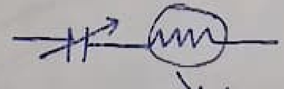
Varactor:



↳ R.B. diode

Diode:



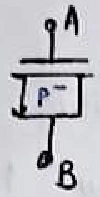


↘ loss of varactor gets added to the loss of tank

↘ Power consumption ↑

↳ Not a good solution → Use MOS cap.

MOS Capacitor (or MOS Varactor):



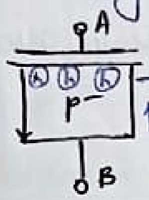
→ MOS (short S and D)

To turn ON,

$V_{AB} > V_{TH}$

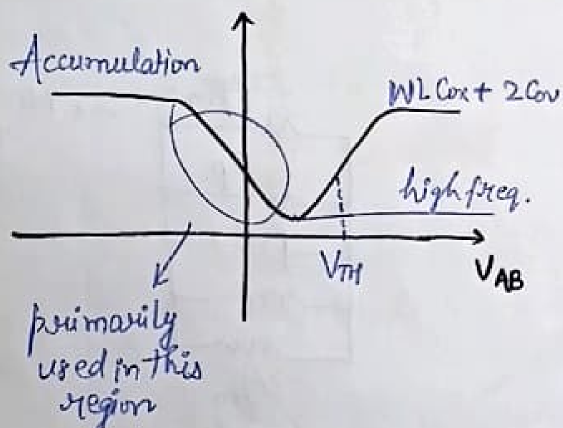
$WL\,C_{ox} + 2\,C_{ov}$

When ⊖ve voltage is applied,

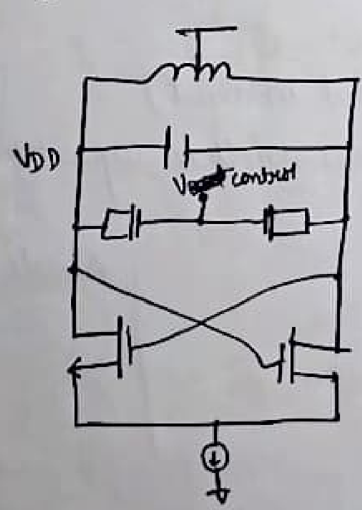


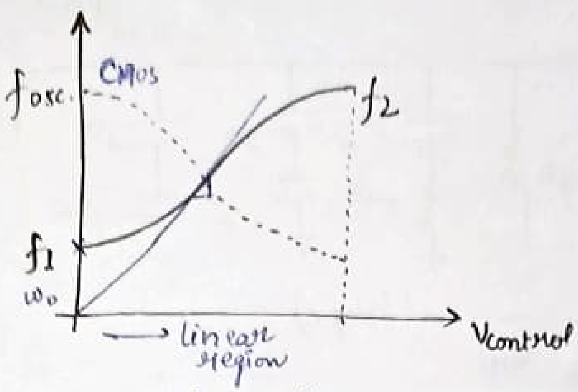
→ accumulation → same cap.

↑ holes



VCO (Voltage-Controlled Oscillator)

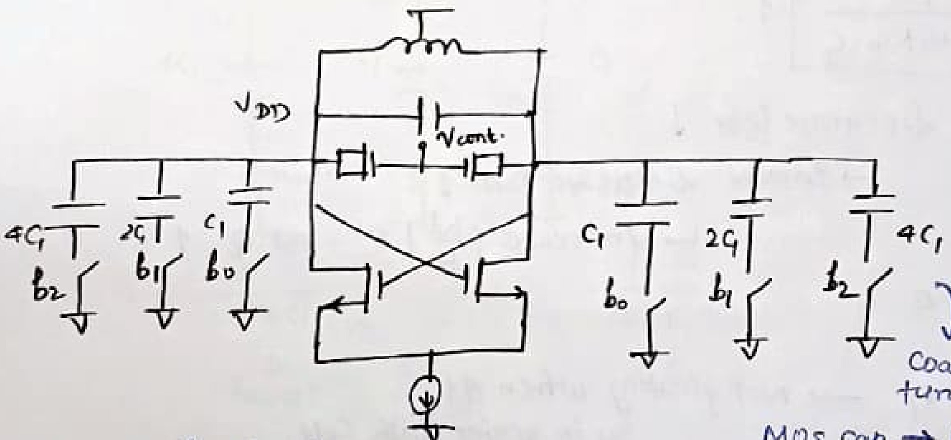




$$\omega_{out} = \omega_0 + K_{VCO} \cdot V_{cont.}$$

$$\Delta f = f_2 - f_1$$

Digital Tuning:

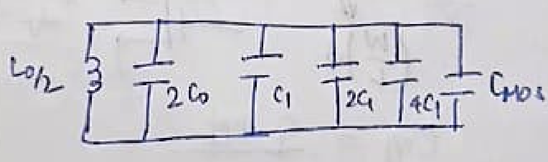


← Digitally-tunable bank of capacitors

Coarse tuning (digital bits) → Implement the switches through MOS

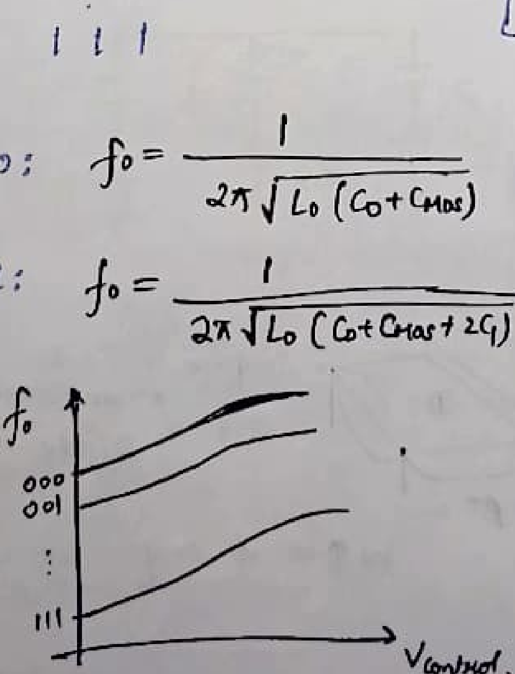
MOS cap. (Vcontrol) → Fine tuning

b_2	b_1	b_0
0	0	0
0	0	1
0	1	0
⋮		
1	1	1



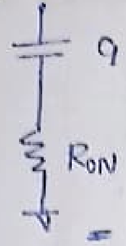
@ 000: $f_0 = \dfrac{1}{2\pi\sqrt{L_0(C_0 + C_{MOS})}}$

@ 001: $f_0 = \dfrac{1}{2\pi\sqrt{L_0(C_0 + C_{MOS} + 2C_1)}}$

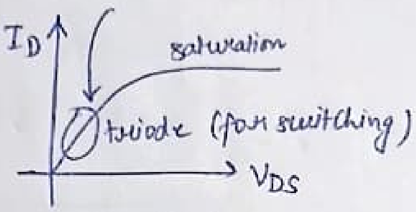


MOSFET switch:

ON: $b_0 = 1$



$$= \frac{1}{\mu_n C_{ox} \left(\frac{W}{L}\right)(V_{DD} - V_{TH})} = (1/\text{slope})$$



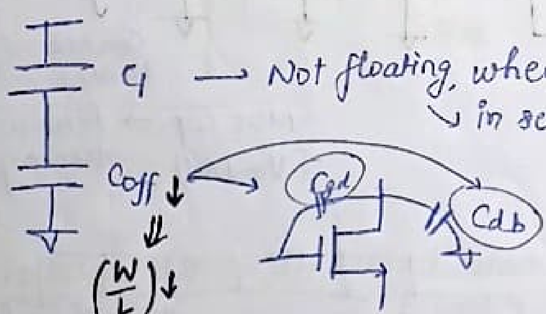
$$\boxed{Q = \frac{1}{\omega_0 R_{ON} \cdot C_l}} \uparrow$$

↳ Decrease loss ↓

↳ Decrease R_{ON} ↓

↳ Increase $\left(\frac{W}{L}\right)$ ↑ $\Rightarrow Q \uparrow$

OFF: $b_0 = 0$



C_l → Not floating, when off

↳ in series with C_{off}.

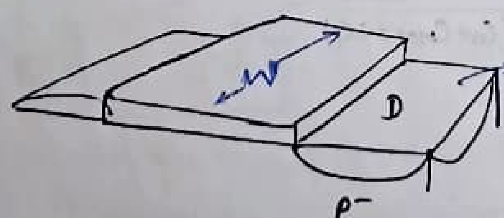
$0 \neq \frac{C_l C_{off}}{C_l + C_{off}} =$

$C_{off} \downarrow \Downarrow \left(\frac{W}{L}\right) \downarrow$

$$C_{eff} = \frac{C_l C_{off}}{C_l + C_{off}}$$

$\left(\frac{W}{L}\right)$ → needs tradeoff.

08-04-2025

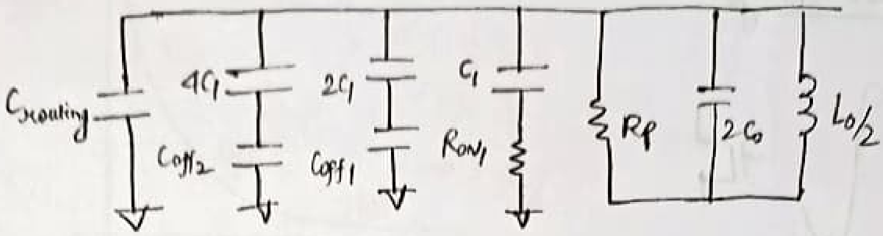


→ surface + 3 sides + underneath

⇒ Diode

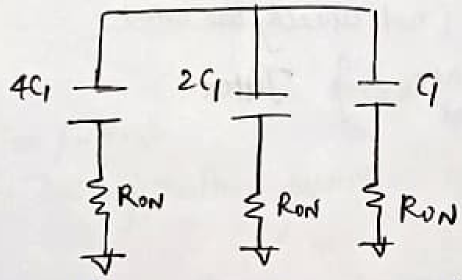
$W \uparrow \Rightarrow$ D surface area ↑

001



$$f_0 = \frac{1}{2\pi\sqrt{\frac{L_0}{2}\left[2C_0 + C_1 + \frac{2C_1 C_{off1}}{2C_1 + C_{off1}} + \frac{4C_1 C_{off2}}{4C_1 + C_{off2}} + C_{routing}\right]}}$$

To increase $f_0 \uparrow \Rightarrow 2C_0 \uparrow$



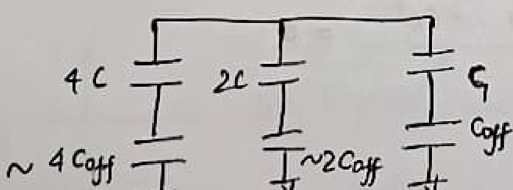
Q: $\frac{1}{\omega_0 \cdot 4C_1 \cdot R_{ON}}$ $\quad \frac{1}{\omega_0 \cdot 2C_1 \cdot R_{ON}}$ $\quad \frac{1}{\omega_0 \cdot C_1 \cdot R_{ON}}$

↓ lowest ↓ highest

$R_{ON} \rightarrow \frac{R_{ON}}{4}$ $\quad R_{ON} \rightarrow \frac{R_{ON}}{2}$ $\quad \Rightarrow$ Constant-Q Sizing

$4\times\left(\frac{W}{L}\right)$ $\quad 2\times\left(\frac{W}{L}\right)$ $\quad \left(\frac{W}{L}\right)_{switch}$

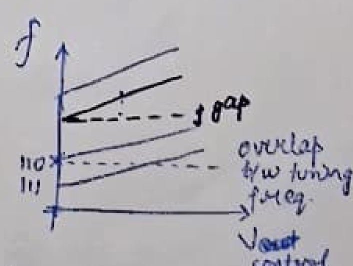
↳ Scale the MOS accordingly.



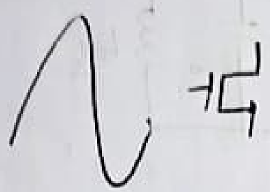
there should not be gap $\Rightarrow$ otherwise that f cannot be tuned

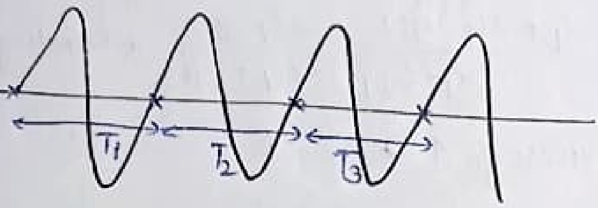
↳ there should be overlap

↳ Limited tuning range



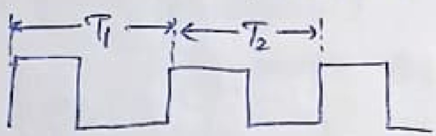
Phase Noise Analysis

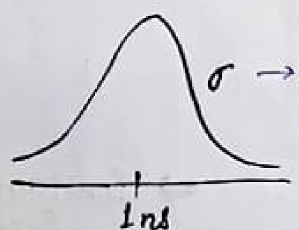




$T_1 = T_2 = T_3 = \ldots = T_N = 1\,ns$ (not usually the case)

Something like $1.0003\,ns$, $0.9990993\,ns$ } Jitter

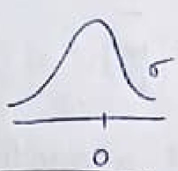




$\sigma \rightarrow$ std. deviation

$T_1 - \mu = \epsilon_1$
$T_2 - \mu = \epsilon_2$
$\vdots$

$\Rightarrow$



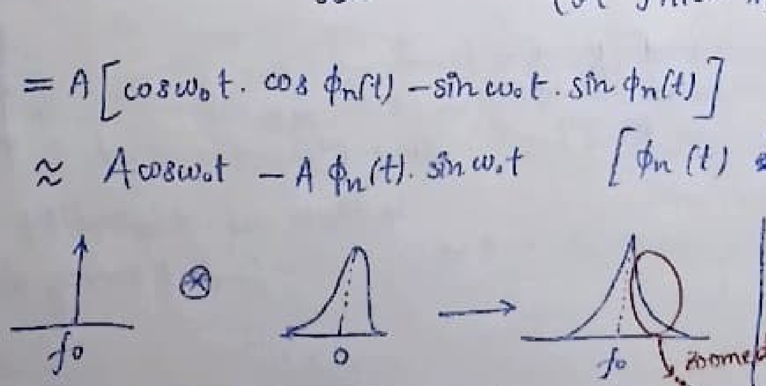
$$\text{RMS Jitter} = \sqrt{\frac{\epsilon_1^2 + \epsilon_2^2 + \ldots + \epsilon_N^2}{N}} = \sigma$$

↳ Noise in the phase:

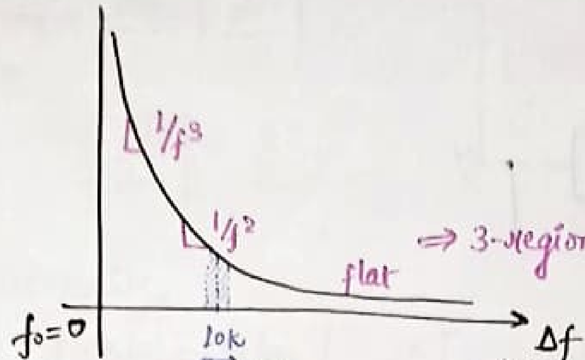
$A\cos(\omega_0 t + \phi_n(t)) \longrightarrow$ Phase noise (or Jitter in digital application)

$$= A\left[\cos\omega_0 t \cdot \cos\phi_n(t) - \sin\omega_0 t \cdot \sin\phi_n(t)\right]$$

$$\approx A\cos\omega_0 t - A\,\phi_n(t) \cdot \sin\omega_0 t \qquad [\phi_n(t) \ll 1]$$



Vector Spectrum Analyzer (VSA)
↳ To phase noise



⇒ 3-regions

1Hz B/W
↓
integrate the power in 1Hz B/W around 10 KHz.

freq. offset from carrier

80 dBc/Hz @ 10 KHz offset
w.r.t. fo
↳ phase noise no. in datasheet

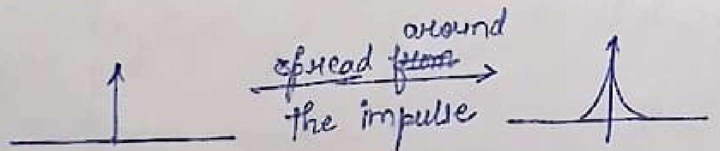
$$\underbrace{10\log\left(\frac{P_{10k}}{P_{f_0}}\right)}_{80\,\text{dBc}}$$

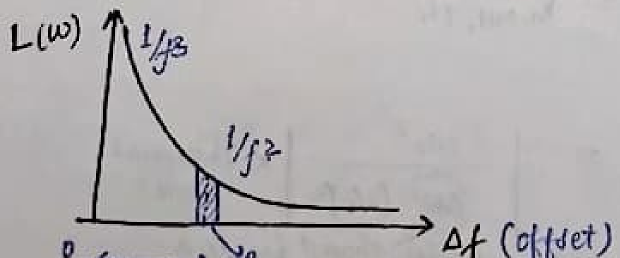
<u>Analysis : For project</u>

15-04-2025

DC → DC operating points
AC
s-parameter analysis → s-parameters
hb / pss analysis → P_{1dB}, IIP_3
hb ↓ harmonic balance (preferred)
pss ↓ periodic steady state

Phase Noise :

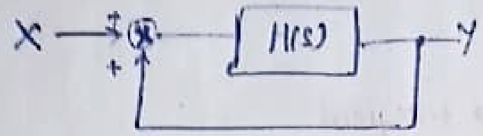




→ look at power in 1 Hz BW around the offset.

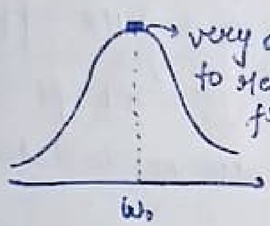
$10\log\left(\frac{P_1}{P_c}\right)$ [−90 dBc/Hz @ 1 KHz offset.]

Positive feedback system:



General TF:

$$\frac{Y}{X}(f) = \frac{H(f)}{1-H(f)}$$



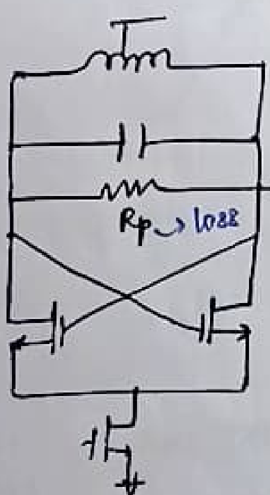
→ very close to resonance freq.

$$H(j(\omega_0+\Delta\omega)) = H(j\omega_0) + \frac{dH}{d\omega}\cdot\Delta\omega$$

$$H(j\omega_0) = 1$$

$$\left|\frac{dH}{d\omega}\cdot\Delta\omega\right| << 1.$$

$$\therefore \frac{Y}{X}\left(j(\omega_0+\Delta\omega)\right) = \frac{H(j\omega_0) + \frac{dH}{d\omega}\Delta\omega}{1 - H(j\omega_0) - \Delta\omega\frac{dH}{d\omega}} \approx \frac{-1}{\Delta\omega\cdot\frac{dH}{d\omega}}.$$



→ Noise is shaped by the transfer function of the system (+ve feedback)

$$S_y(f) = S_x(f)\cdot|H(f)|^2$$

$$\left|\frac{Y}{X}(j(\omega_0+\Delta\omega))\right|^2 \cdot \overline{V^2_{n,th}} = \overline{V^2_{n,out,th}}$$

$$\left|\frac{1}{\Delta\omega^2\left|\frac{dH}{d\omega}\right|^2}\right| \times \underline{\text{thermal noise}} = \left|\frac{\omega_0^2}{\Delta\omega^2 (2Q)^2}\right| \times \text{thermal noise}$$

thermal noise shaped by $1/f^2$

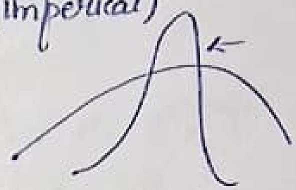
[$Z(j(\omega_0+\Delta\omega)) = j\frac{\omega L}{\frac{2\Delta\omega}{\omega_0}}$ → derived for series or parallel RLC

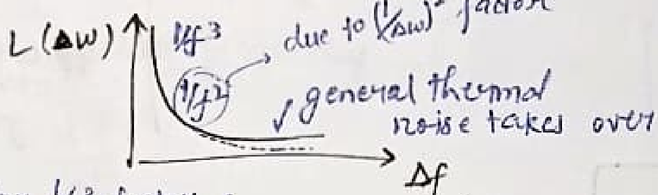
$$Q = \frac{\omega_0}{2}\left|\frac{dH}{d\omega}\right|$$

$$\Rightarrow \left|\frac{dH}{d\omega}\right| = \frac{2Q}{\omega_0}$$

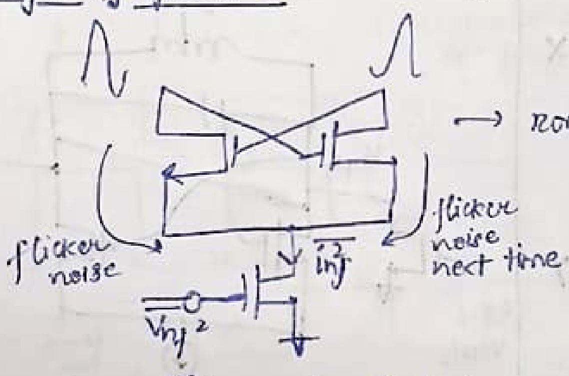
Phase noise,

$$L(\Delta\omega) = 10 \log\left[\frac{2KT}{P_{sig.}} \cdot \left(\frac{\omega_0}{2Q\Delta\omega}\right)^2\right]$$ (imperical)





Reason for $1/f^3$ factor :



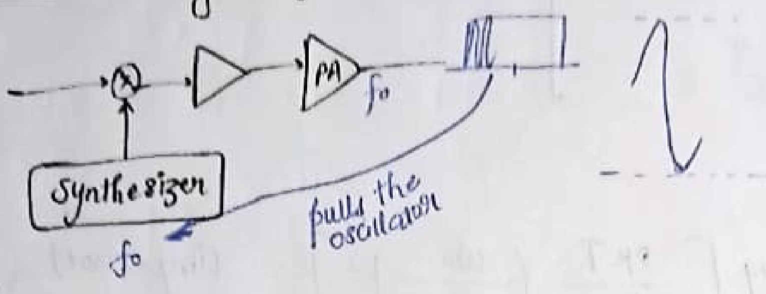
→ non-linear oscillator

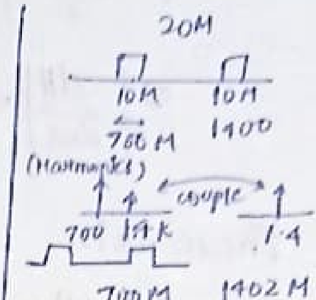
Including flicker noise factor,

$$L(\Delta\omega) = 10 \log\left[\frac{2FKT}{P_{sig}}\left(1 + \left(\frac{\omega_0}{2Q\Delta\omega}\right)^2\right)\left(1 + \frac{\Delta\omega_{1/f}}{\Delta\omega_0}\right)\right]$$

16-04-2025

Oscillator Pulling (Injection Pulling)

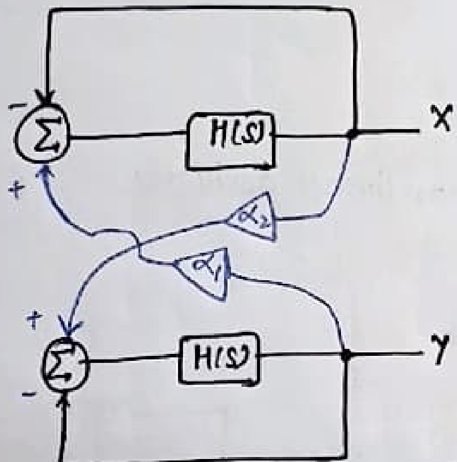




pulling ↓ was supposed to be 1.4K

\# we can lock bad oscillator with a good oscillator

Quadrature Coupling



[Two oscillators cross-coupled]

$$x = (\alpha_1 y - x)\, H(s) \;\}\times\; \alpha_2 x$$
$$y = (\alpha_2 x - y)\, H(s) \;\}\times\; \alpha_1 y$$

$$\begin{cases} \alpha_2 x^2 = (\alpha_1\alpha_2 xy - \alpha_2 x^2)\, H(s) \\ \alpha_1 y^2 = (\alpha_1\alpha_2 xy - \alpha_1 y^2)\, H(s) \end{cases}$$

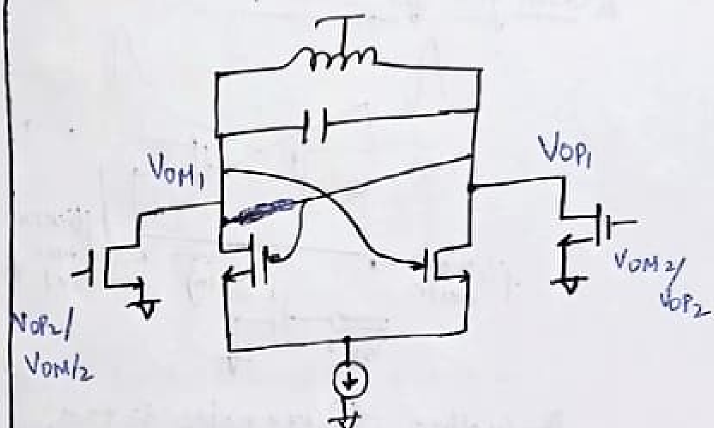
$$\Rightarrow \alpha_2 x^2 - \alpha_1 y^2 = -(\alpha_2 x^2 - \alpha_1 y^2)\, H(s)$$

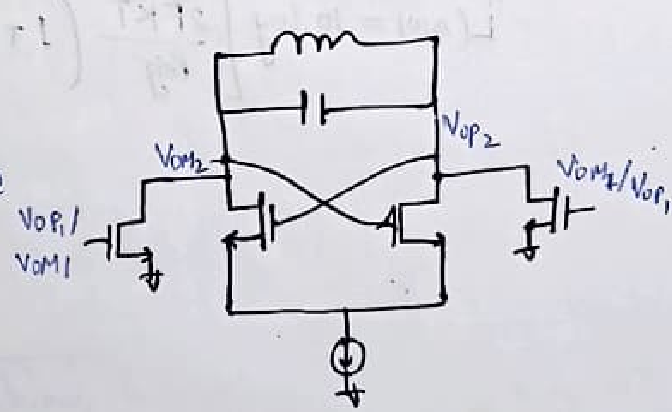
$$\Rightarrow (\alpha_2 x^2 - \alpha_1 y^2)\, [1 + H(s)] = 0$$

For $\alpha_2 x^2 - \alpha_1 y^2 = 0$ $\quad \underbrace{[1 + H(s)]}_{=0 \,\to\, \text{not possible}}$

$\Rightarrow \alpha_1 = \alpha_2$

$x = y$: Direct coupling





Coupling → Oscillates at same freq.

Cross-coupling

For $\alpha_1 = -\alpha_2$

$x^2 = -y^2$

$\Rightarrow x = \pm jy$ → 90° phase shift [one directly coupled, other cross-coupled]

↳ Quadrature coupling

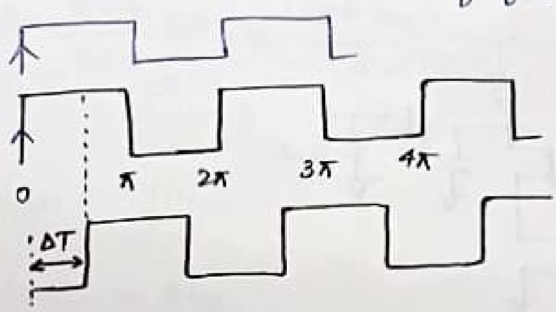
↳ Quadrature oscillator

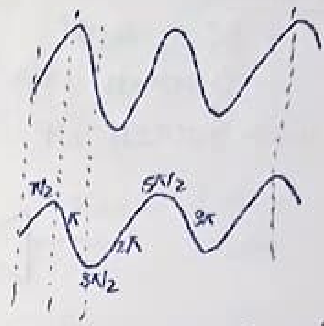
Phase Locked Loop

ω: angular frequency

$$\omega = \frac{d\phi}{dt}$$

$$\phi = \int \omega + K$$: accumulation of freq.





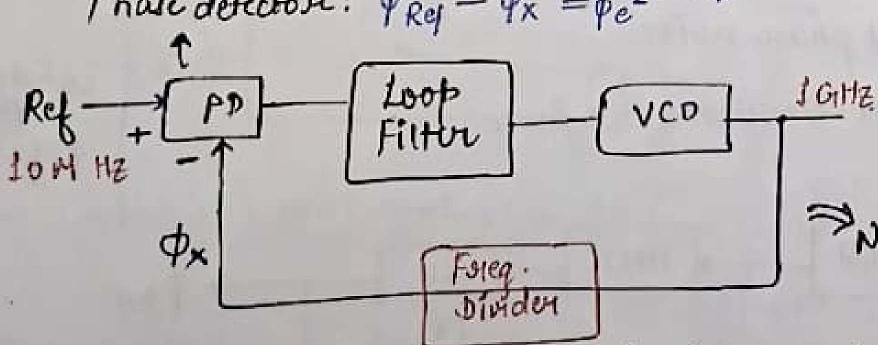
↳ Two signals locked

↳ same phase and frequency.

$$\frac{\Delta T}{T} \times 2\pi = \Delta\phi$$

Locked signals : { same freq. + same phase : rising edges are aligned.

Phase detector: $\phi_{Ref} - \phi_x = \phi_e \leftarrow V_{PD}$



Look at the phase error and ↑/↓ the freq.

↳ using VCO

⇒ Not actually used ⇒ use Freq. divider

↳ Integer-end divider ⇒ Integer-end synthesizer.

Fractional-end divider ⇒ Frac-end synthesizer

Crystal oscillator

↳ very precise but at low freq (Ref.)

22-04-2025

Project:

→ LO: DC/ superheterodyne

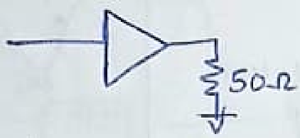
→ $\underbrace{\text{LNA}}_{\text{assignment}} + \underbrace{\text{mixer + LPF}}_{\text{project}}$ [Passive or active mixer]

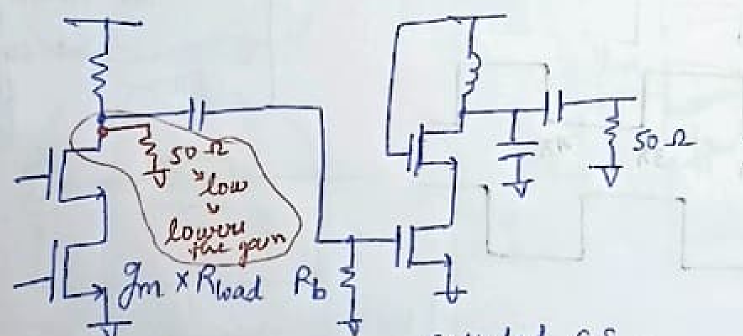
2.401 GHz ⟶ ⟶ 1 MHz

↑

$\omega = 2.4$ GHz

Toshow:
- DC results
- Overall gain
- Overall NF





↳ gives most of the gain

cascoded CS
↳ gives the marginal gain

⇒ overall good gain.

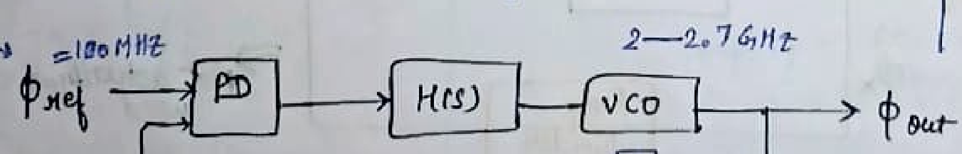
PLL:

Oscillator ⟶ bad phase noise

(Quartz) Crystal Oscillator ⟶ pure

$\omega = \frac{d\phi}{dt}$

$\phi = \int \omega dt$



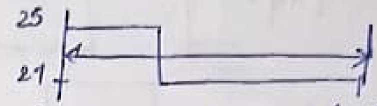
2.4 GHz : $N = \frac{2400}{100} = 24$ (division ratio)

↳ ÷ 3 → ÷ 2 → ÷ 2 → ÷ 2

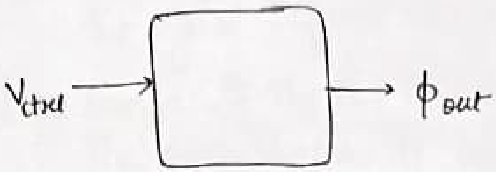
2.5 GHz : $N = 25$.

2.425 GHz : $N = 24.25$

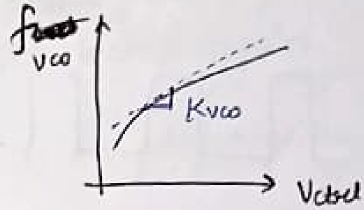
↳ 75% of time → $N = 24$
25% of time → $N = 25$



[Duty cycle decided by the decimal point]



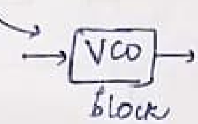
$$\omega_{vco} = 2\pi \cdot K_{vco} \cdot V_{ctrl}$$
$$f_{vco} = K_{vco} \cdot V_{ctrl}$$



$$K_{vco} = \frac{\partial f_{vco}}{\partial V_{ctrl}}$$

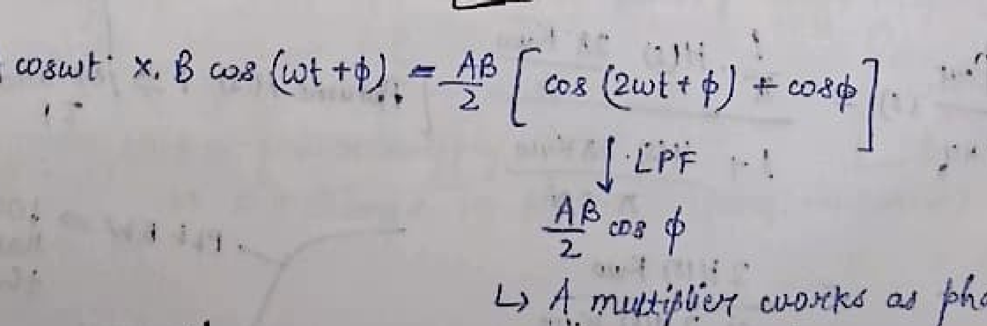
$$\phi_{out} = \int 2\pi \cdot K_{vco} \cdot V_{ctrl}$$

$$\Rightarrow \frac{\phi_{vco}}{V_{ctrl}}(s) = \frac{2\pi K_{vco}}{s}$$



$$f_{out} = \frac{f_{in}}{N}$$
$$\omega_{out} = \frac{\omega_{in}}{N}$$
$$\phi_{out} = \frac{1}{N}\int \omega_{in} = \frac{1}{N}\phi_{in}$$

Hence,

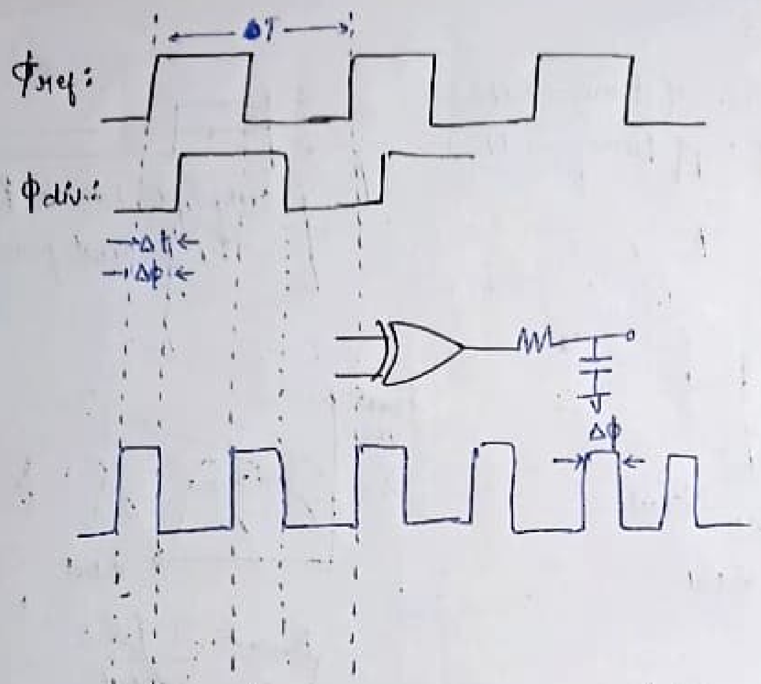


→ $A\cos\omega t \times B\cos(\omega t + \phi) = \frac{AB}{2}\left[\cos(2\omega t + \phi) + \cos\phi\right]$

$\downarrow$ LPF

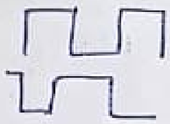
$\frac{AB}{2}\cos\phi$

↳ A multiplier works as phase detector
↳ cos of phase difference.
↳ Use XOR gate instead

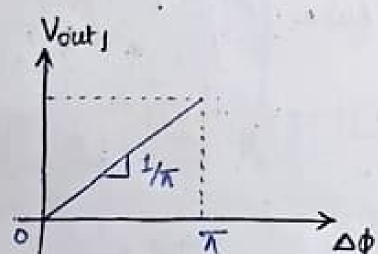


$\Delta\phi = \frac{\Delta T}{T} \times 2\pi$

0 : (for 0° phase diff)

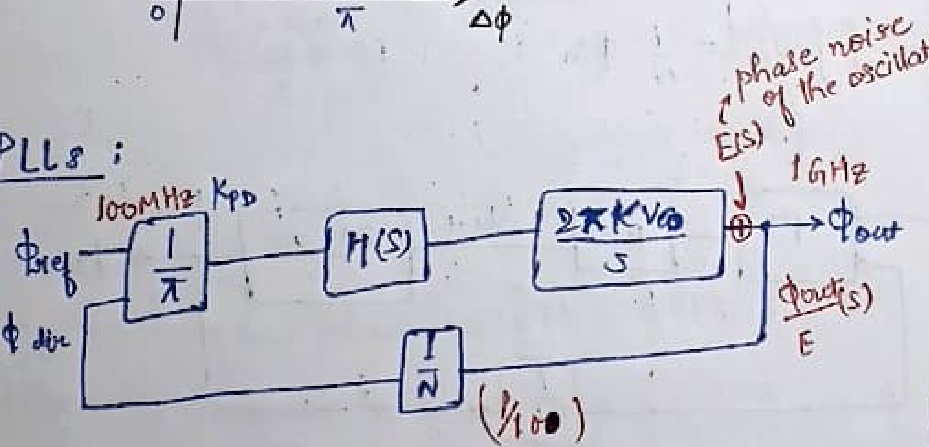


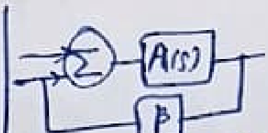
π : V_{DD} (for π phase diff)



25-04-2025

PLLs :





Loop gain $= \frac{A(s)}{1+A(s)\beta(s)}$

$$\frac{\phi_{out}}{\phi_{ref}}(s) = \frac{\frac{1}{\pi}\cdot H(s)\cdot\frac{2\pi K_{vco}}{s}}{1 + \frac{H(s)\cdot 2\pi K_{vco}}{\pi\cdot s\cdot N}}$$

[Assume $H(s)=1 \Rightarrow$ 1st order T.F.]

$$= \frac{2H(s)K_{vco}}{s + \frac{2H(s)\cdot K_{vco}}{N}}$$

$\Rightarrow$ Lock if the ratio is within the range of PLL

Low pass

Low pass: Jitter at the i/p appears at the o/p of PLL.

↳ we cannot have ϕ_{out} better than ϕ_{ref}.

<u>Capture Range</u>: Range of freq. that the PLL can capture and lock the signal before the i/p is given.

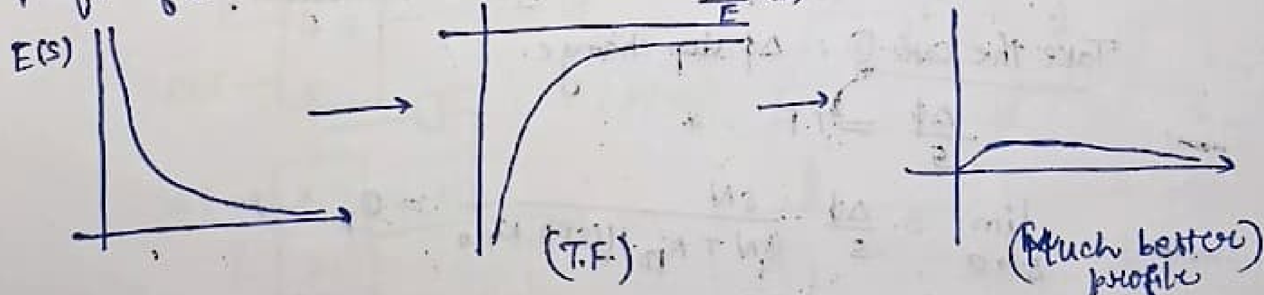
<u>Lock Range</u>: If the PLL is already locked and given a perturbation, the perturbation it can sustain to get locked again.

$$\phi_{out}(s) = \left(0 - \frac{\phi_{out}}{N}\right) \cdot \frac{1}{\pi} \cdot H(s) \cdot \frac{2\pi K_{vco}}{s} + E(s)$$ [superposition, $\phi_{ref} = 0$]

$$\Rightarrow \phi_{out}(s)\left[1 + \frac{H(s)\, 2\pi K_{vco}}{N\pi s}\right] = E(s)$$

$$\Rightarrow \frac{\phi_{out}}{E}(s) = \frac{sN}{sN + 2H(s) \cdot K_{vco}} \rightarrow \text{High-pass response}$$

Profile of E(s):



→ Anything put in PLL gets corrected

<u>Perturbations</u>: ↳ Present in all µC.

<u>Case-I</u>: $\Delta\omega$ steps [Loop is perturbed by ~~the~~ step change in freq.]

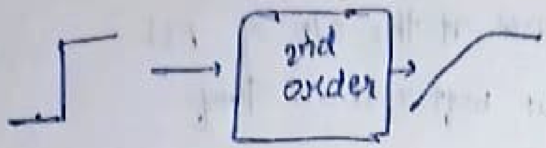
$$\frac{\Delta\omega}{s} \xrightarrow[\int]{\text{phase}} \frac{\Delta\phi}{s^2}$$

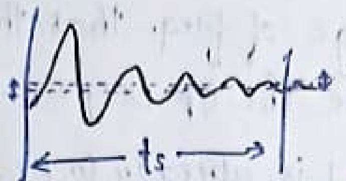
[$\Delta\omega$ should be within the BW range of PLL to be corrected]

<u>Case-II</u>: Glitch in power supply, etc.

↳ $\Delta\phi$ change in phase (freq. → same)

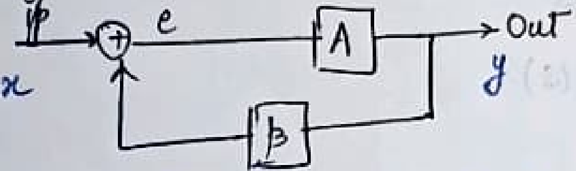
#





1% settling time (say) of PLL
↳ critical parameter (Regulated in ICs)

→



residual error: error (e) when the system has settled.

$$e = x - y\beta$$
$$= x - \frac{Ax\beta}{1+A\beta}$$
$$= x\left(\frac{1}{1+A\beta}\right)$$ → whether the error is zero once settled?

Take the case-②: $\Delta\phi$ ~~step~~ change.

$$\frac{\Delta\phi}{s} = \text{T.F.}$$

$$\lim_{s\to 0} s \cdot \frac{\Delta\phi}{s} \cdot \frac{sN}{sN + K_{PD} \cdot H(s) \cdot 2 \cdot K_{VCO}} = 0,$$

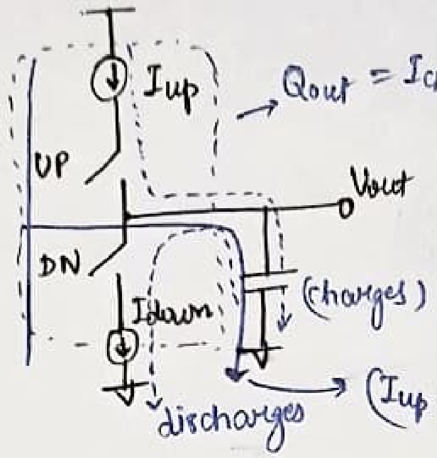
∴ Error → 0

Case-①: $\frac{\Delta\phi}{s^2}$ [Type-1 PLL]

$$\lim_{s\to 0} s \cdot \frac{\Delta\phi}{s^2} \cdot \frac{sN}{sN + K_{PD} \cdot H(s) \cdot 2\pi K_{VCO}} \neq 0$$

↳ Almost a residual error.
↳ Never able to correct.

Charge Pump Integrator



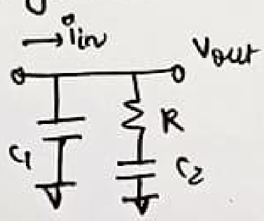
$\rightarrow Q_{out} = I_{cp} Q$

$$H(s) = I(s) \times \frac{1}{sC}$$

$$= (I_{up} - I_{down}) \frac{1}{sC}$$

$\rightarrow (I_{up} - I_{down})$ if both switches ON.

Lead/Lag Filter:



$$H(s) = \frac{1 + \frac{s}{\omega_z}}{s(C_1 + C_2)\left(1 + \frac{s}{\omega_p}\right)}$$

$$\omega_z = \frac{1}{R_1 C_2} \qquad \omega_p = \frac{C_1 + C_2}{R_1 C_1 C_2}$$

Phase-Frequency Detector