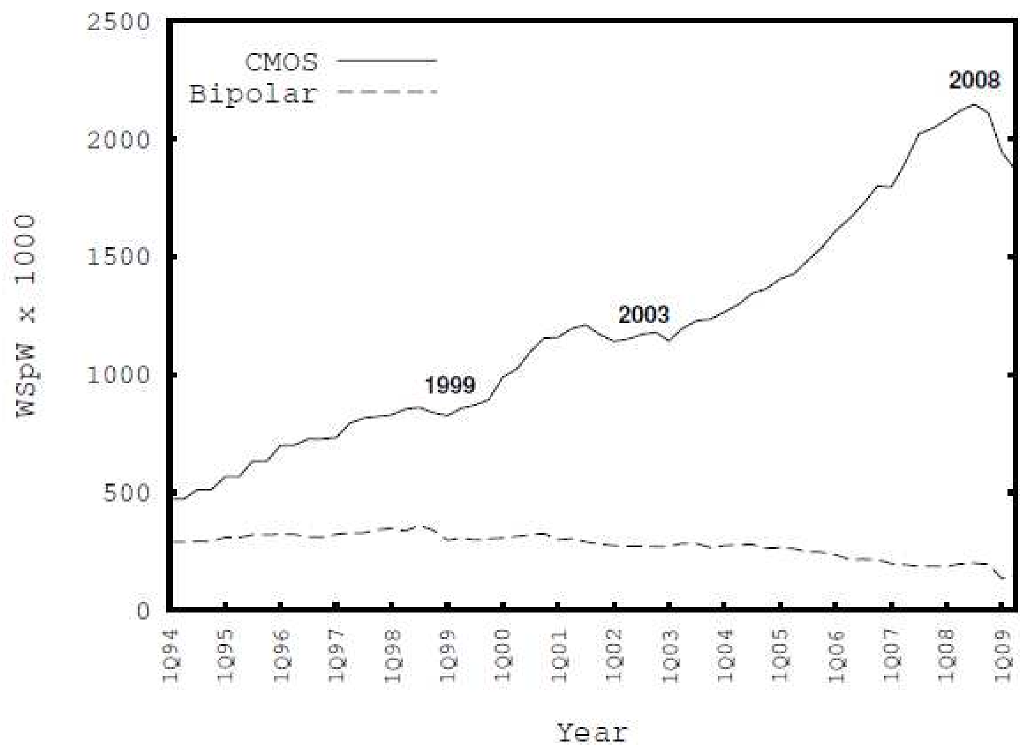
RF-CMOS

LNA & PA

Outline

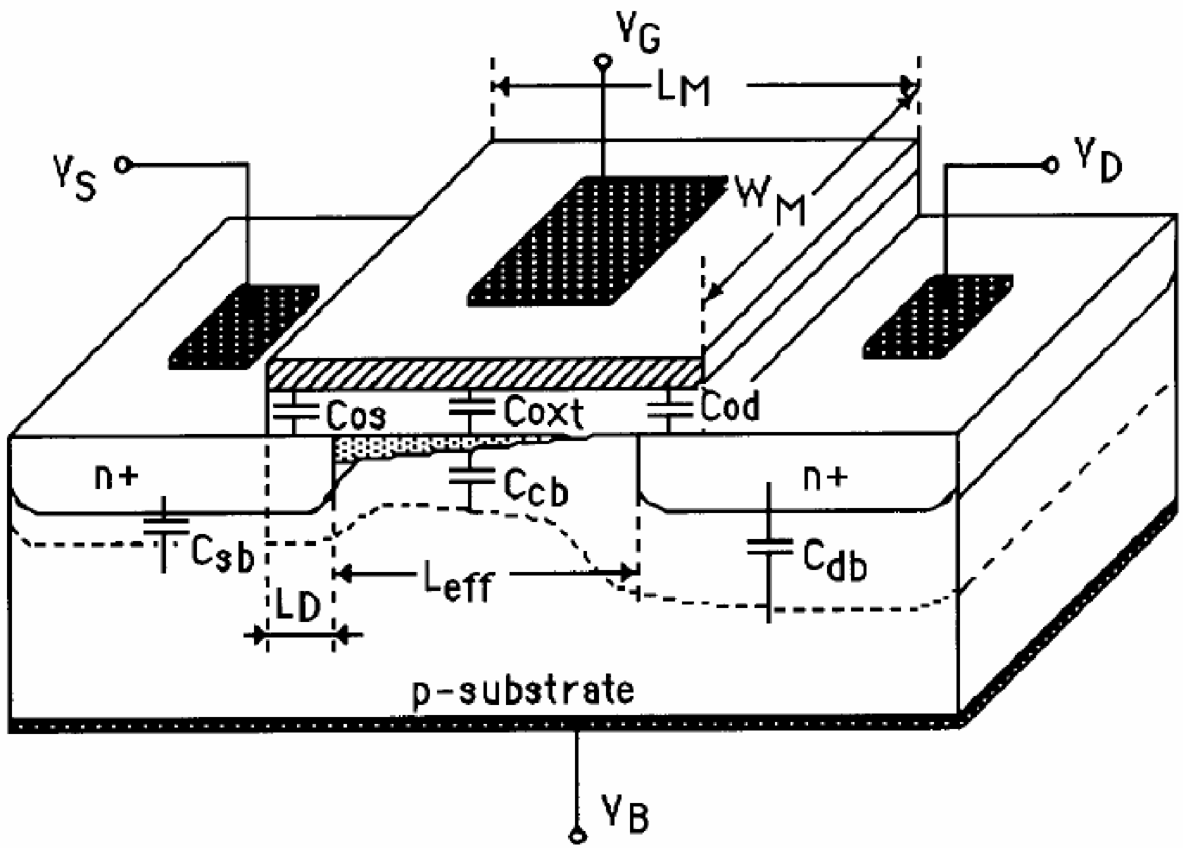
1. RF-CMOS
   – Driving forces
   – MOS RF model (capacitance)
   – f_T and deep-submicron
   – Saturation Velocity
   – NQS
2. Brief Review of basic concepts
   – Sensitivity
   – Noise Figure
   – “NonLinearity” related: P1db, IP3
3. Transceiver Overview
4. Low Noise Amplifier
   – Why?
   – Discrete towards IC implementation
   – Common topologies and analysis
   – Example of published implementations
5. Power Amplifier
   – Type of operations (A, B, ...,F) and characterization Parameters

RF-CMOS

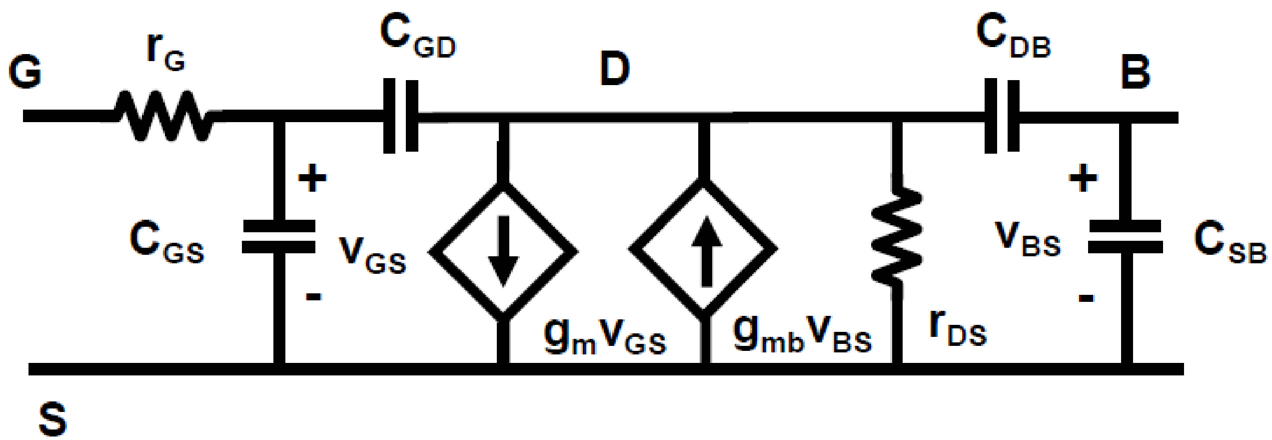


The evolution of CPU capacity according to the Moore's Law, has pushed the RF part to be integrated in CMOS.

MOS Transistor



MOS Transistor Capacitances

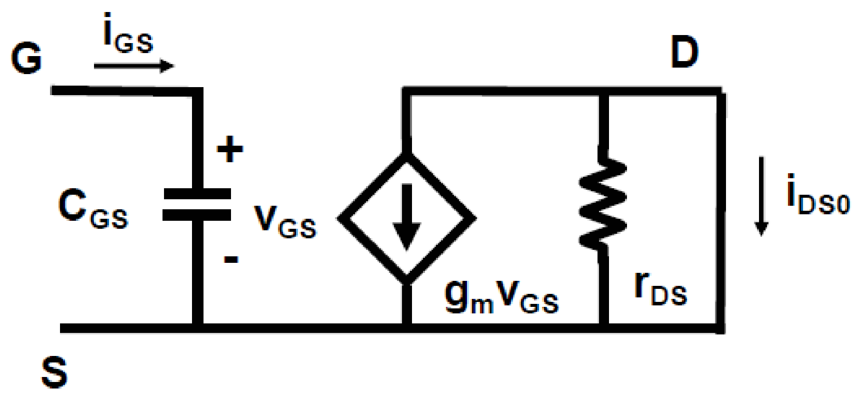


$$C_{GS} \approx \frac{2}{3} WLC_{ox} \quad \approx 2W \ \ \mathrm{fF}/\mu\mathrm{m \ for} \ L_{min}$$

$$L_{min}C_{ox} \approx L_{min}\frac{\varepsilon_{ox}}{t_{ox}} \approx 50\ \varepsilon_{ox} \approx 2\ \mathrm{fF}/\mu\mathrm{m}$$

$$C_{GD} = WC_{gdo}$$

MOS Transistor f_T



$$i_{GS} = v_{GS} C_{GS} s$$

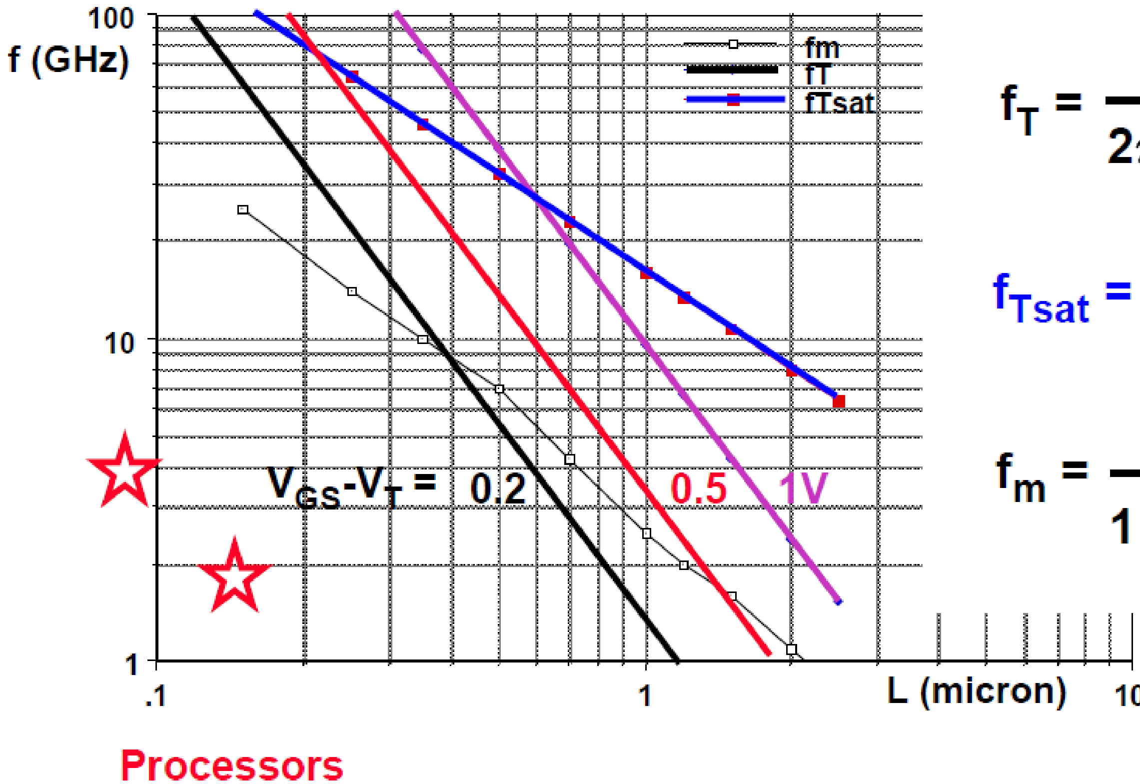
$$i_{DS} = g_m v_{GS}$$

$$C_{GS} = \frac{2}{3} WLC_{ox} \quad g_m = 2K' \frac{W}{L} (V_{GS}-V_T) \quad K' = \frac{\mu C_{ox}}{2n}$$

$$\boxed{f_T = \frac{g_m}{2\pi C_{GS}} = \frac{1}{2\pi} \frac{3}{2n} \frac{\mu}{L^2} (V_{GS}-V_T)} \quad \text{or} \quad \boxed{\approx \frac{v_{sat}}{2\pi L}}$$

$$f_{max} \approx \sqrt{f_T / 8\pi r_G C_{GD}}$$

MOS Transistor f_T



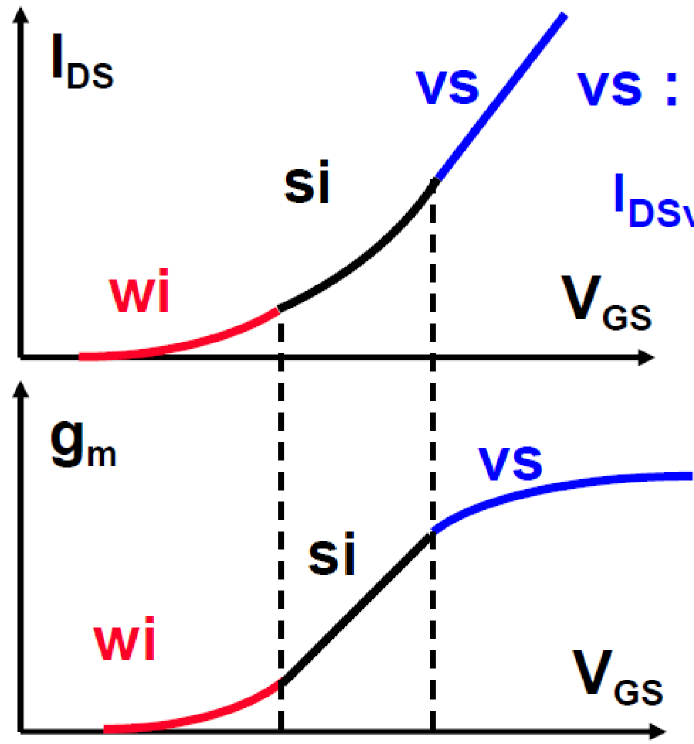
$$f_T = \frac{\mu}{2\pi L^2}\underbrace{(V_{GS}-V_T)}_{0.2\ \ldots\ 1\ V}$$

$$f_{Tsat} = \frac{v_{sat}}{2\pi L}$$

$$f_m = \frac{f_T}{1+\alpha_{BD}}$$

$$\alpha_{BD} \approx \frac{C_{BD}}{C_{ox}}$$

Velocity Saturation



vs : velocity saturation

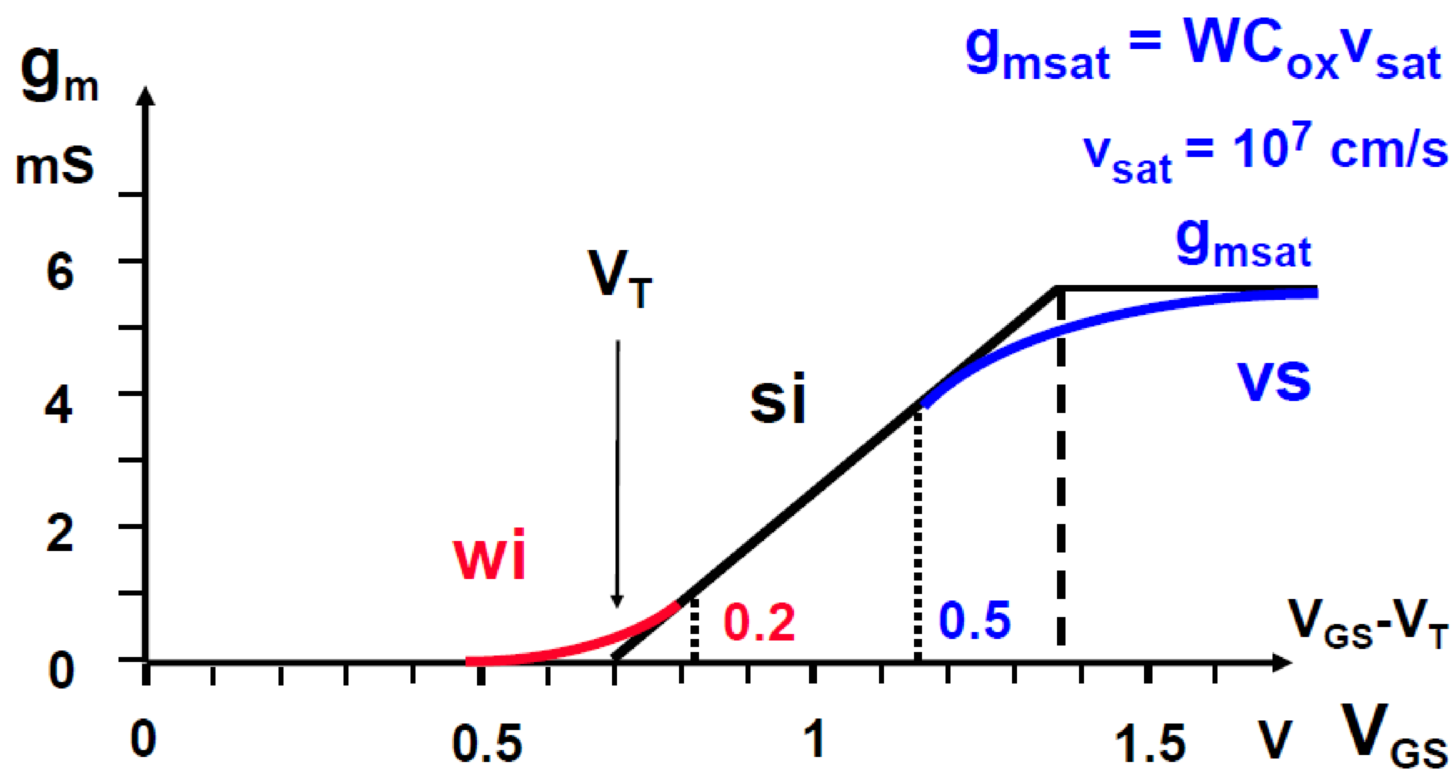
$I_{DSvs} = WC_{ox}v_{sat}\,(V_{GS}-V_T)$

$v_{sat} \approx 10^7$ cm/s

$g_{msat} = WC_{ox}v_{sat}$

is absolute max. !

Velocity Saturation



Velocity Saturation

[large V_{GS}]

$$I_{DS} = \frac{K'_n \frac{W}{L}(V_{GS}-V_T)^2}{1+\theta(V_{GS}-V_T)} \approx \frac{K'_n}{\theta}\frac{W}{L}(V_{GS}-V_T)$$

$$g_{msat} \approx 2K'_n \frac{W}{L}(V_{GS}-V_T)^2 \frac{1+\frac{\theta}{2}(V_{GS}-V_T)}{[1+\theta(V_{GS}-V_T)]^2} \approx \frac{K'_n}{\theta}\frac{W}{L}$$

$$= WC_{ox}v_{sat}$$

$$\boxed{\theta L = \frac{\mu}{2n}\frac{1}{v_{sat}}} = \frac{1}{E_c}$$

E_c is the vertical critical field !

$\theta L \approx 0.2\ \mu m/V$: For $L = 0.13\ \mu m$ $\theta \approx 1.6\ V^{-1}$

MOS Transistor

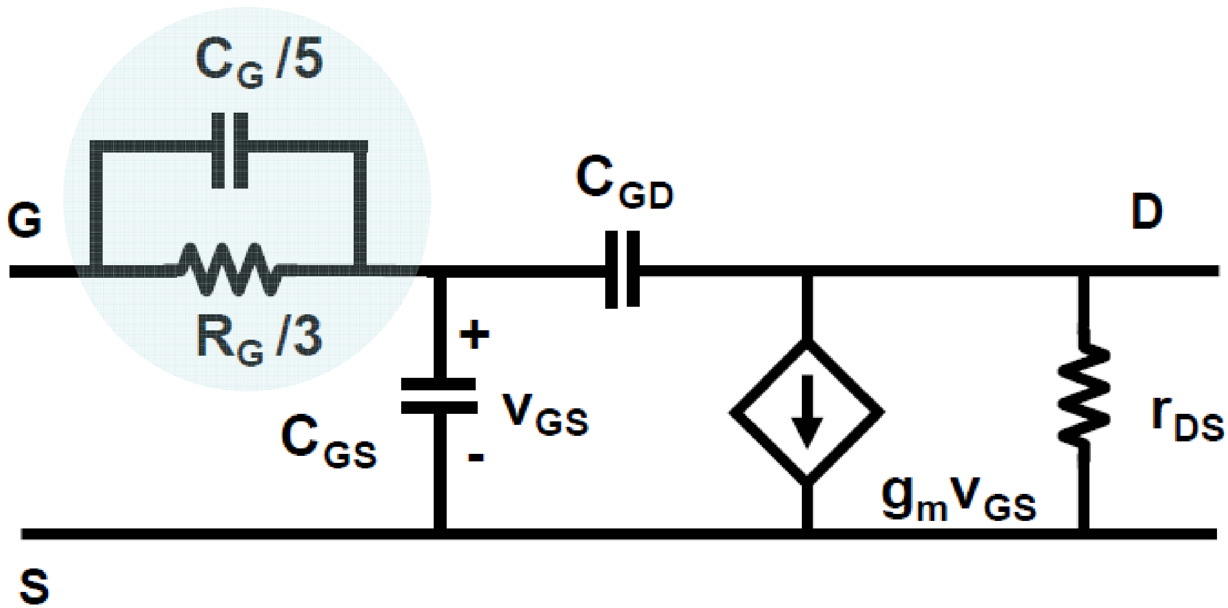
Summary :

TABLE 1-4 EXPRESSIONS OF I_{DS}, g_m AND g_m/I_{DS} FOR MOST

	I_{DS}	g_m	$\frac{g_m}{I_{DS}} = f(v_{GS}-V_T)$	$\frac{g_m}{I_{DS}} = f(I_{DS})$
wi	$I_{D0}\frac{W}{L}\exp\left(\frac{v_{GS}}{nkT/q}\right)$ (1-25*a*)	$\frac{I_{D0}}{nkT/q}\frac{W}{L}\exp\left(\frac{v_{GS}}{nkT/q}\right)$ (1-25*b*)	$\frac{1}{nkT/q}$ (1-26*b*)	$\frac{1}{nkT/q}$ (1-26*b*)
ws			$(v_{GS}-V_T)_{ws} = 2n\frac{kT}{q}$	$I_{DSws} = \frac{\text{KP}}{2n}\frac{W}{L}\left(2n\frac{kT}{q}\right)^2$
si	$\frac{\text{KP}}{2n}\frac{W}{L}(v_{GS}-V_T)^2$ (1-18*c*)	$2\frac{\text{KP}}{2n}\frac{W}{L}(v_{GS}-V_T)$ (1-22*a*)	$\frac{2}{v_{GS}-V_T}$ (1-26*a*)	$2\sqrt{\frac{\text{KP}}{2n}\frac{W}{L}\frac{1}{I_{DS}}}$ (1-26*a*)
sv			$(v_{GS}-V_T)_{sv} = \frac{2nLC_{\text{ox}}v_{\text{sat}}}{\text{KP}}$	$I_{DSsv} = \frac{2WLC_{\text{ox}}^2 v_{\text{sat}}^2}{\text{KP}/2n}$
vs	$WC_{\text{ox}}v_{\text{sat}}(v_{GS}-V_T)$ (1-38*b*)	$WC_{\text{ox}}v_{sat}$ (1-39)	$\frac{1}{v_{GS}-V_T}$	$\frac{WC_{\text{ox}}v_{\text{sat}}}{I_{DS}}$

Ref.: Laker, Sansen : Design of analog …, MacGrawHill 1994; Table 1-4

MOS Transistor at RF

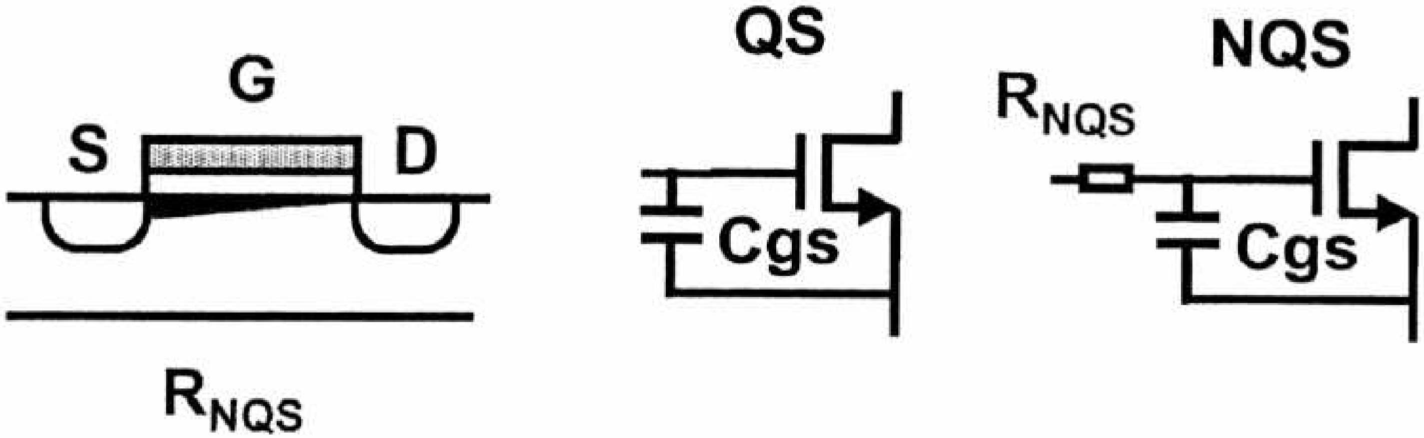


$$C_G = C_{GS} + C_{GD}$$

Ref. Tin, Tr. CAD, April 1998, 372

Ref. Sansen, etal, ACD, XDSL, RFMOS models, Kluwer 1999

Non-Quasi Static Model



$$R_{NQS} = \frac{1}{5\,g_m}$$

(Y. Tsividis, “The MOS Transistor...”, Oxford Press)

MOS Transistor

$$I_{DS} = K'_n \frac{W}{L} (V_{GS}-V_T)^2$$

$V_{GS}-V_T \approx 0.2$ V

$K'_n \approx 100\ \mu A/V^2$
$K'_p \approx 40\ \mu A/V^2$

$$g_m = 2K'_n \frac{W}{L} (V_{GS}-V_T) = 2\sqrt{K'_n \frac{W}{L} I_{DS}} = \frac{2\, I_{DS}}{V_{GS}-V_T}$$

$$r_{DS} = r_o = \frac{V_E L}{I_{DS}}$$

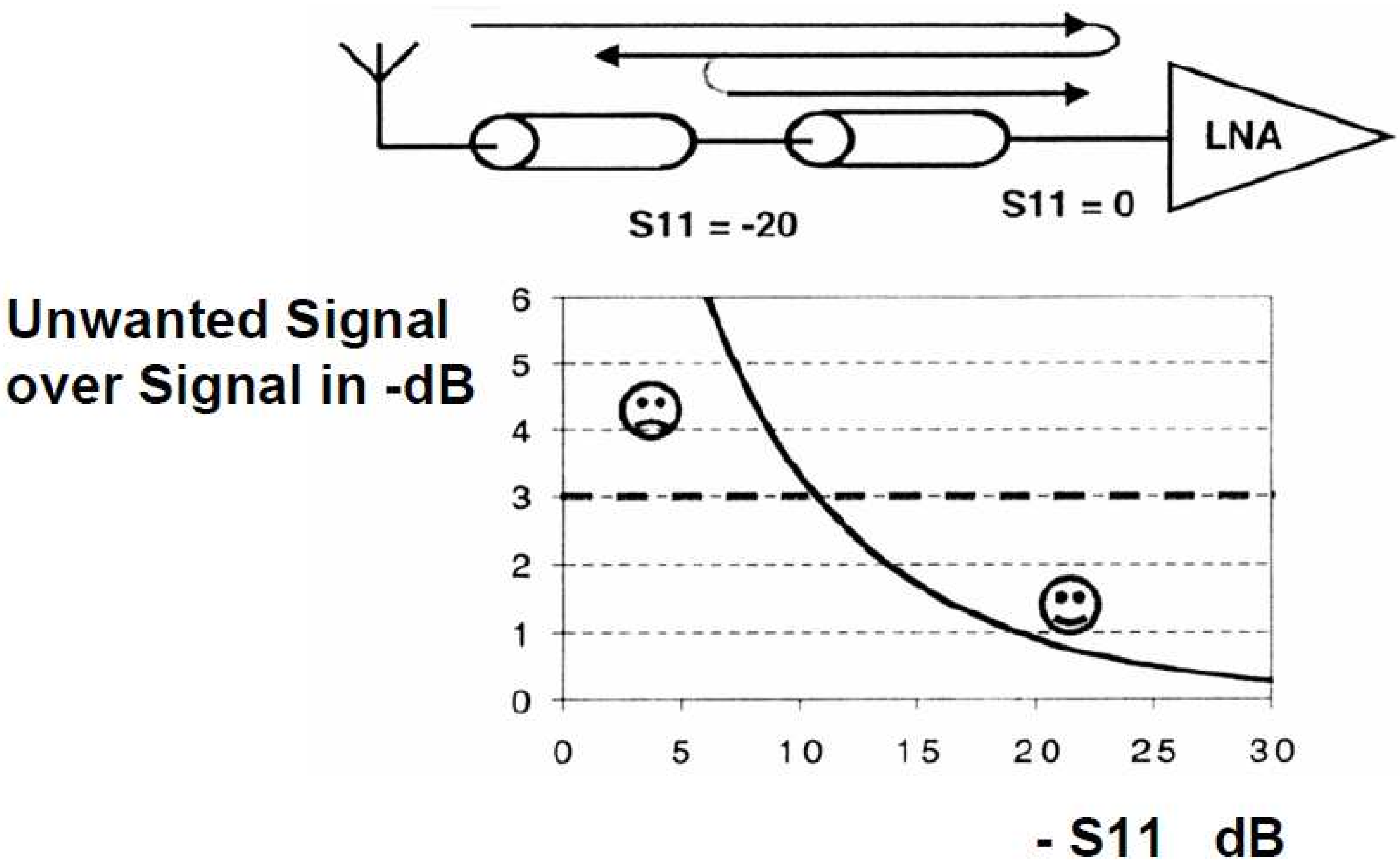
$V_{En} \approx 5\ V/\mu mL$ $V_{Ep} \approx 8\ V/\mu mL$

$v_{sat} = 10^7$ cm/s

$$f_T = \frac{1}{2\pi} \frac{3}{2n} \frac{\mu}{L^2} (V_{GS}-V_T) \quad \text{or now} \approx \frac{v_{sat}}{2\pi L}$$

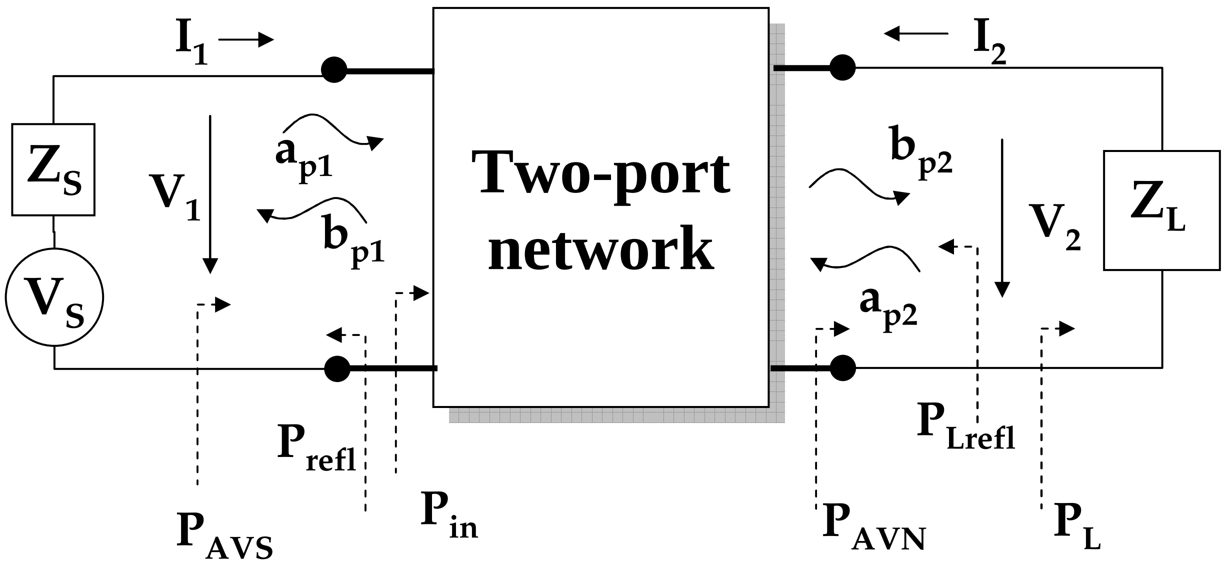
Brief Overview of Basic Concepts

Transmission line effects



Basic Concepts

Power gains from S-parameters



$$G_P = \frac{P_L}{P_{in}} = \frac{1}{1-|\rho_{in}|^2}|S_{21}|^2\frac{1-|\rho_L|^2}{|1-S_{22}\rho_L|^2}$$

$$G_A = \frac{P_{AVN}}{P_{AVS}} = G_T\frac{P_{AVN}}{P_L} = \frac{1-|\rho_S|^2}{|1-S_{11}\rho_S|^2}|S_{21}|^2\frac{1}{1-|\rho_{out}|^2}$$

$$G_T = \frac{P_L}{P_{AVS}} = G_P\frac{P_{in}}{P_{AVS}} = \frac{1-|\rho_s|^2}{|1-S_{11}\rho_s|^2}|S_{21}|^2\frac{1-|\rho_L|^2}{|1-\rho_{out}\rho_L|^2}$$

S-Parameters

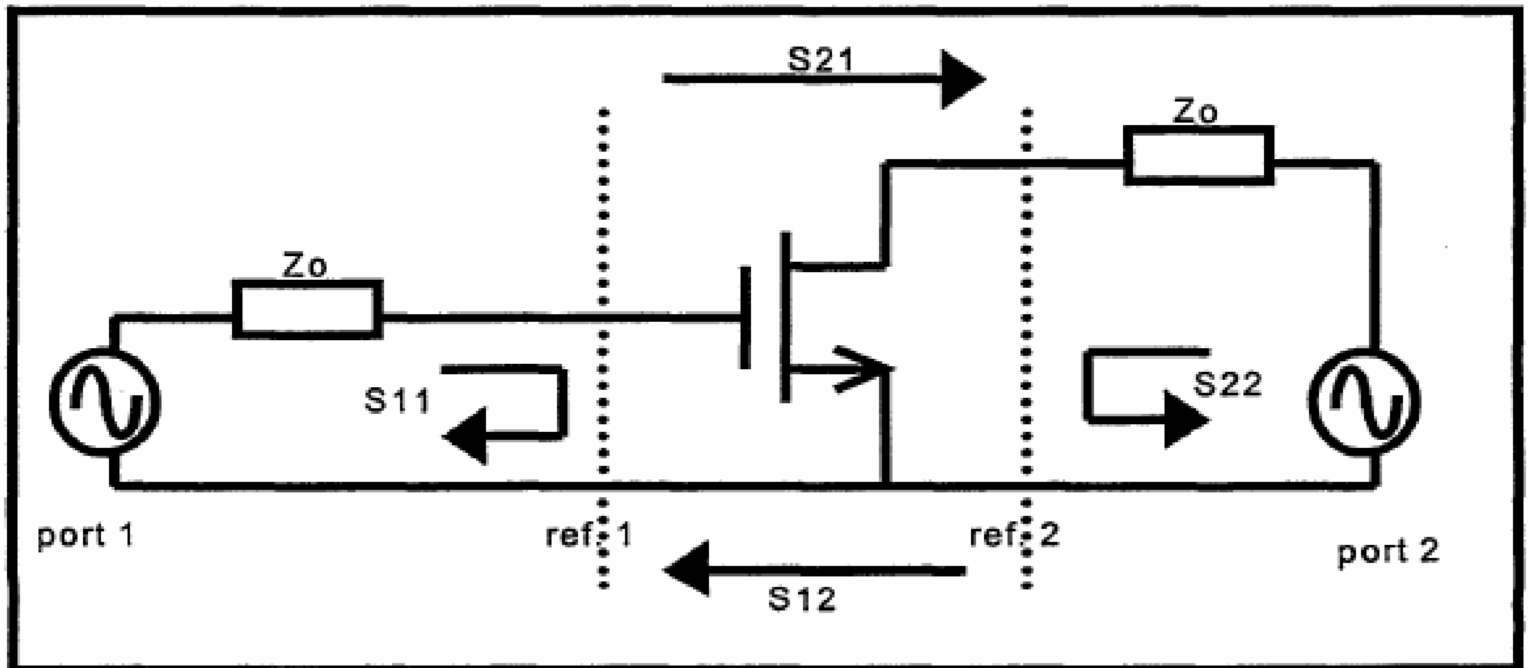
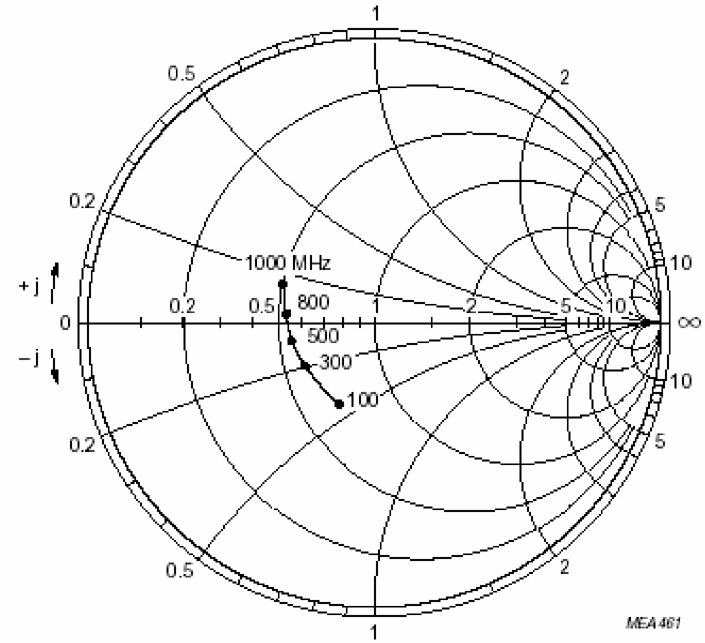


Figure1: Representation of S-parameters for a two port active device

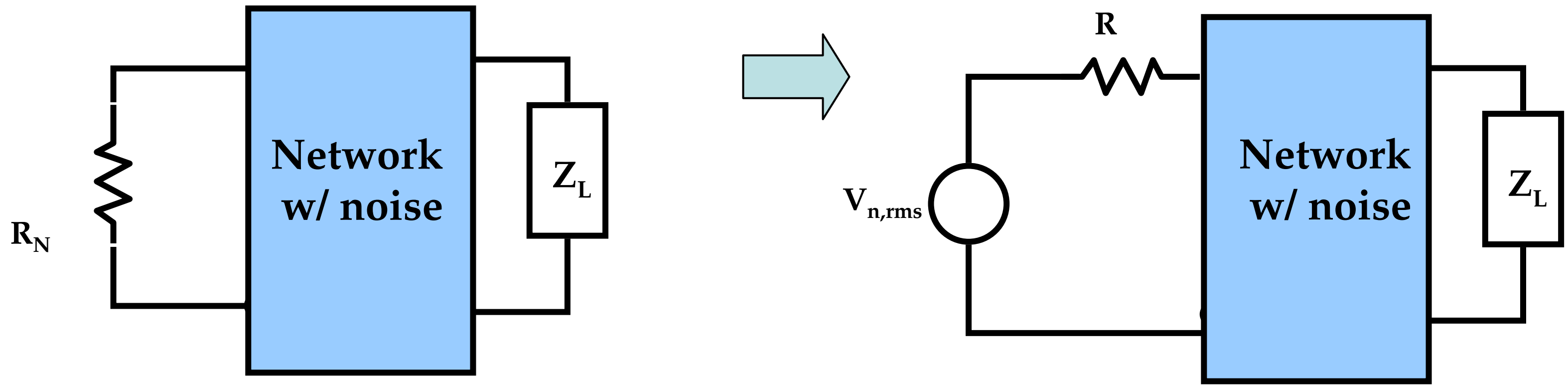
S-Parameter: S_{11}



I_C = 30 mA; V_{CE} = 5 V; Z_o = 50 Ω; T_{amb} = 25 °C.

Fig.8 Common emitter input reflection coefficient (S_{11}).

Noise Figure



The noise figure of a microwave amplifier is defined as the ratio of the total available noise power at the output of the amplifier to the available noise power at output due to thermal noise from the input termination R at T = 290°K.

$$F = \frac{P_{No}}{P_{Ni} G_A} = \frac{P_{No}}{P_{Ni} \frac{P_{So}}{P_{Si}}} = \frac{P_{Si} / P_{Ni}}{P_{So} / P_{No}} = \frac{(SNR)_i}{(SNR)_o}$$

Noise Figure

Total output noise power given by the sum of the noise generated by the amplifier and input noise power afected by the amplifier power gain.

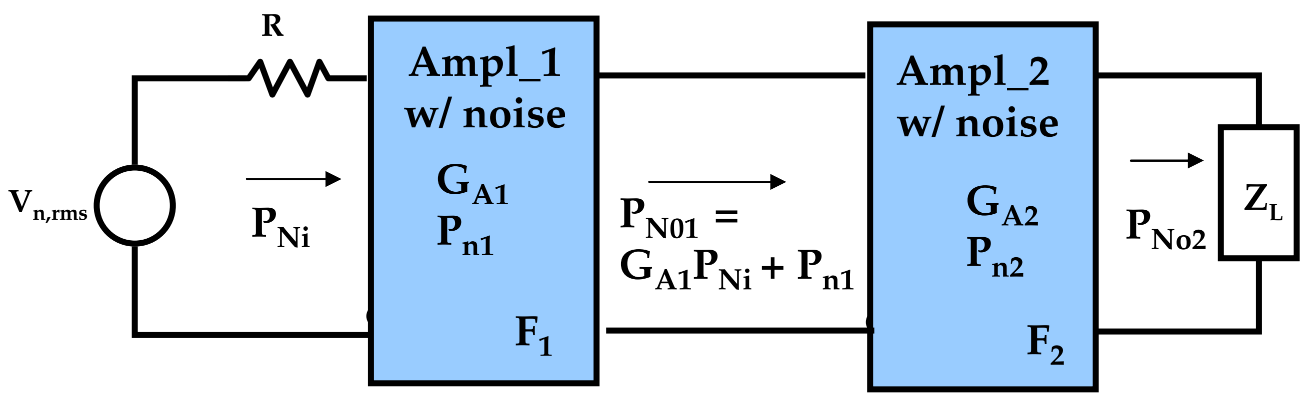
$$F = \frac{P_{No}}{P_{Ni} G_A} = \frac{P_{Ni} G_A + P_n}{P_{Ni} G_A} = 1 + \frac{P_n}{P_{Ni} G_A}$$

$F \geq 1$; when F=1 the amplifier does not add noise

P_{No} – Available noise power at the output
P_{Ni} – Available noise power at the input
P_{Si} – Available signal power at the input
P_{So} – Available signal power at the output
P_n – Noise power added by the amplifier

Noise Figure

Noise figure of cascade of amplifiers



$$F = \frac{P_{No2}}{P_{Ni}G_{A1}G_{A2}} = \frac{G_{A2}\left(P_{Ni}G_{A1} + P_{n1}\right) + P_{n2}}{P_{Ni}G_{A1}G_{A2}} = F_1 + \frac{F_2 - 1}{G_{A1}}$$

General result: $$F = F_1 + \frac{F_2 - 1}{G_{A1}} + \frac{F_3 - 1}{G_{A1}G_{A2}} + \ldots$$

Noise Figure

From the noise analysis of a two-port network, it is possible to observe the inpact of the source admittance seen by the device

$$F = F_{\min} + \frac{r_n}{g_s}\left|y_s - y_{opt}\right|^2$$

$$r_n = \frac{R_n}{Z_0}$$ Equivalent normalized noise resistance

$$y_s = g_s + jb_s$$ Normalized source admittance

$$y_{opt} = g_{opt} + jb_{opt}$$ Normalized source admittance necessary to obtain $F=F_{\min}$

$$F = F_{\min} + \frac{4r_n\left|\rho_s - \rho_{opt}\right|^2}{\left(1-\left|\rho_s\right|^2\right)\left|1+\rho_{opt}\right|^2}$$

Parameters
r_n, ρ_{opt}, F_{min},
given by the manufacturer

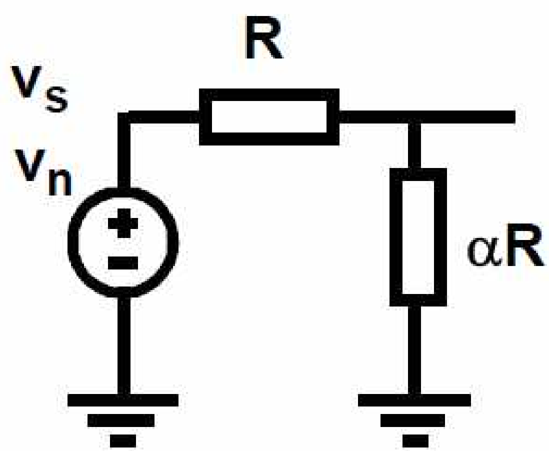
Noise Figure

$$Vn = \sqrt{4RkTB}$$

$$Vs = \sqrt{4R.S}$$

$$NF = \frac{\textit{total output noise power}}{\textit{output noise due to input source}}$$

$$NF = \frac{4kTB[\frac{\alpha}{(1+\alpha)^2} + \frac{1}{(1+\alpha)^2}]}{4kTB\frac{\alpha}{(1+\alpha)^2}}$$



$$NF = \frac{1+\alpha}{\alpha}$$

Resistive match : α = 1 : NF = 3 dB

Sensitivity

$$P_{in,\min}\Big|_{dBm} = P_{RS}\Big|_{dBm} + NF\Big|_{dB} + SNR_{\min}\Big|_{dB} + 10\log \underset{\text{Hz}}{B}$$

(~ -174dBm/Hz)

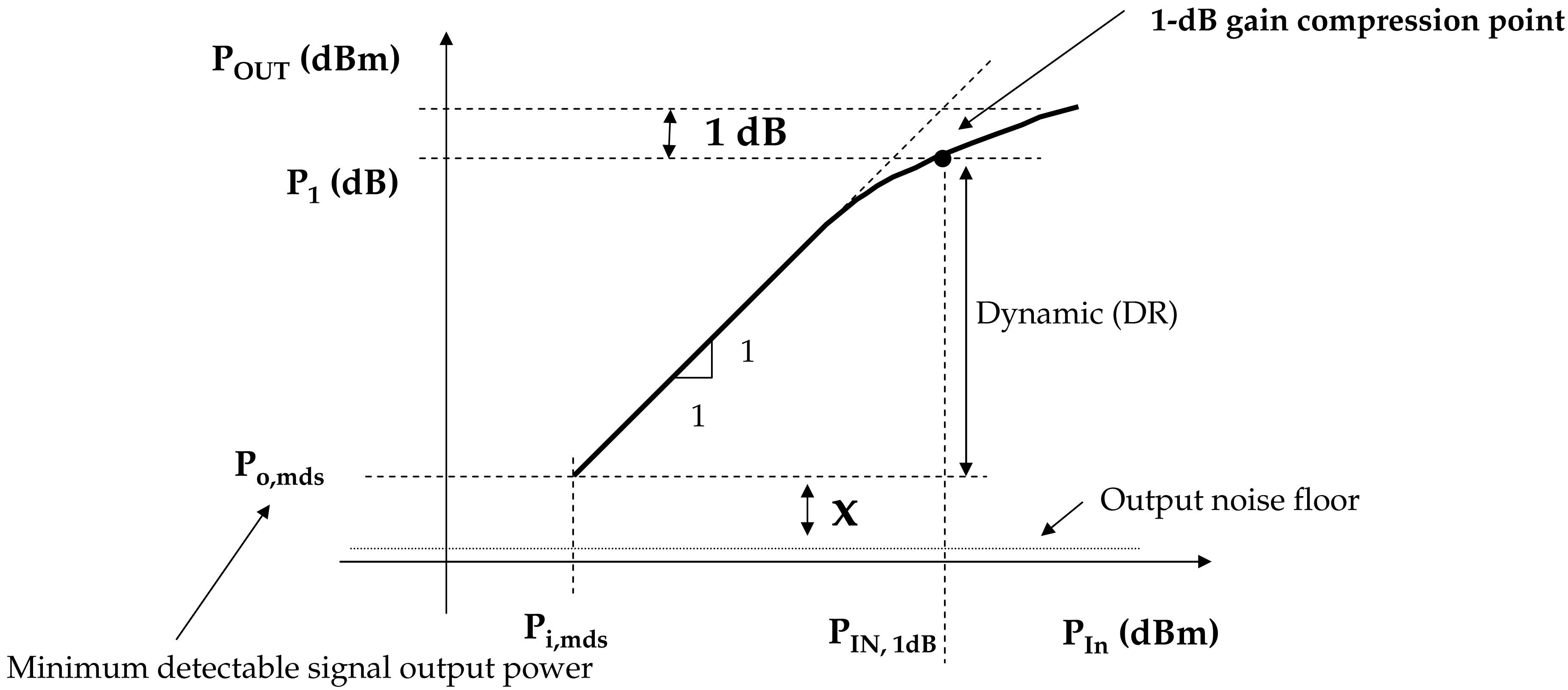
e.g. : In WCDAM, B= 3.84 MHz, SNR=4dB, and for a sensitivity of -107dBm

→ NF~ 7-9dB

1dB Compression point

Nonlinear characterization : 1-dB gain compression point

The 1-dB gain compression point (G1dB) is defined as the power gain where the nonlinearities of the transistor reduce the power gain by 1dB over the small-signal power gain



Dynamic Range

Nonlinear characterization : 1-dB gain compression point

$$G_{1dB}(dB) = G_o(dB) - 1$$

Small signal power gain

$$G_{1dB}(dB) = \frac{P_{1dB}}{P_{IN,1dB}}$$

$$P_{1dB}(dBm) = P_{IN,1dB}(dBm) + G_{1dB}(dB)$$

Dynamic range

$$DR = P_{1dB} - P_{0,mds} \qquad (dB)$$

$$P_{o,mds}(dBm) = P_{i,mds}(dBm) + G_A(dB)$$

Detection margin, 3dB.

$$P_{o,mds}(dBm) = \underbrace{-174dBm + 10\log B}_{\text{Noise power in B}} + F(dB) + X(dB) + G_A(dB)$$

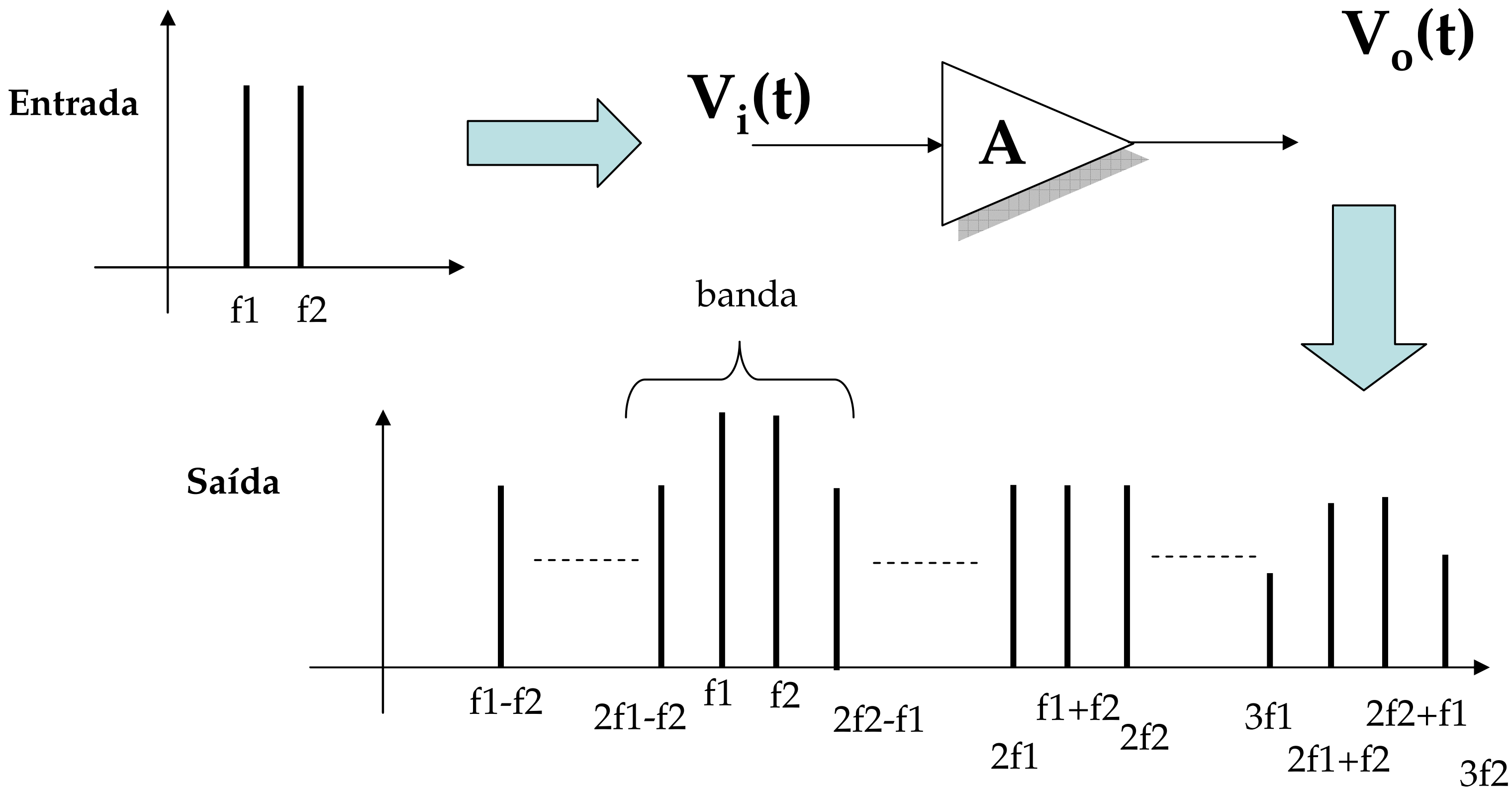
Amplifier noise figure

Available gain.

Distortion

Nonlinear characterization : Harmonic distortion and intermodulation

→ A source of distortion in amplifier is that caused by intermodulation products.

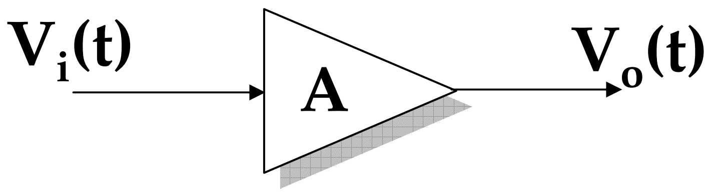


Distortion

Nonlinear characterization : Harmonic distortion and intermodulation

Applying two sinusoidal frequencies to the amplifier, Vi(t)

$$V_i(t) = \cos(2\pi f_1 t) + \cos(2\pi f_2 t)$$



$$V_o(t) = \underbrace{AV_i(t)}_{\text{components } f_1 \text{ and } f_2} + \underbrace{BV_i^2(t)}_{} + \underbrace{CV_i^3(t)}_{}$$

components f_1 and f_2

- DC
- 2nd harmonic : $2f_1$ e $2f_2$
- 2nd order intermodulation products: $f_{1\pm}\ f_2$

- f_1 and f_2
- 3rd harmonic : $3f_1$ e $3f_2$
- 3rd order intermodulation products: $2f_{1\pm}\ f_2$, $2f_{1\pm}\ f_2$

Distortion

Nonlinear characterization : Harmonic distortion and intermodulation

3rd order intermodultaion products: $2f_{1\pm} f_2$, $2f_{1\pm} f_2$

limits
- dynamic range
- amplifier's bandwidth

→ Compare them to the power of the fundamental, f_1.
Definition of Third-order intercept point P_{IP} (IP3), which corresponds to the interception between P_{f1} and P_{2f1-f2}

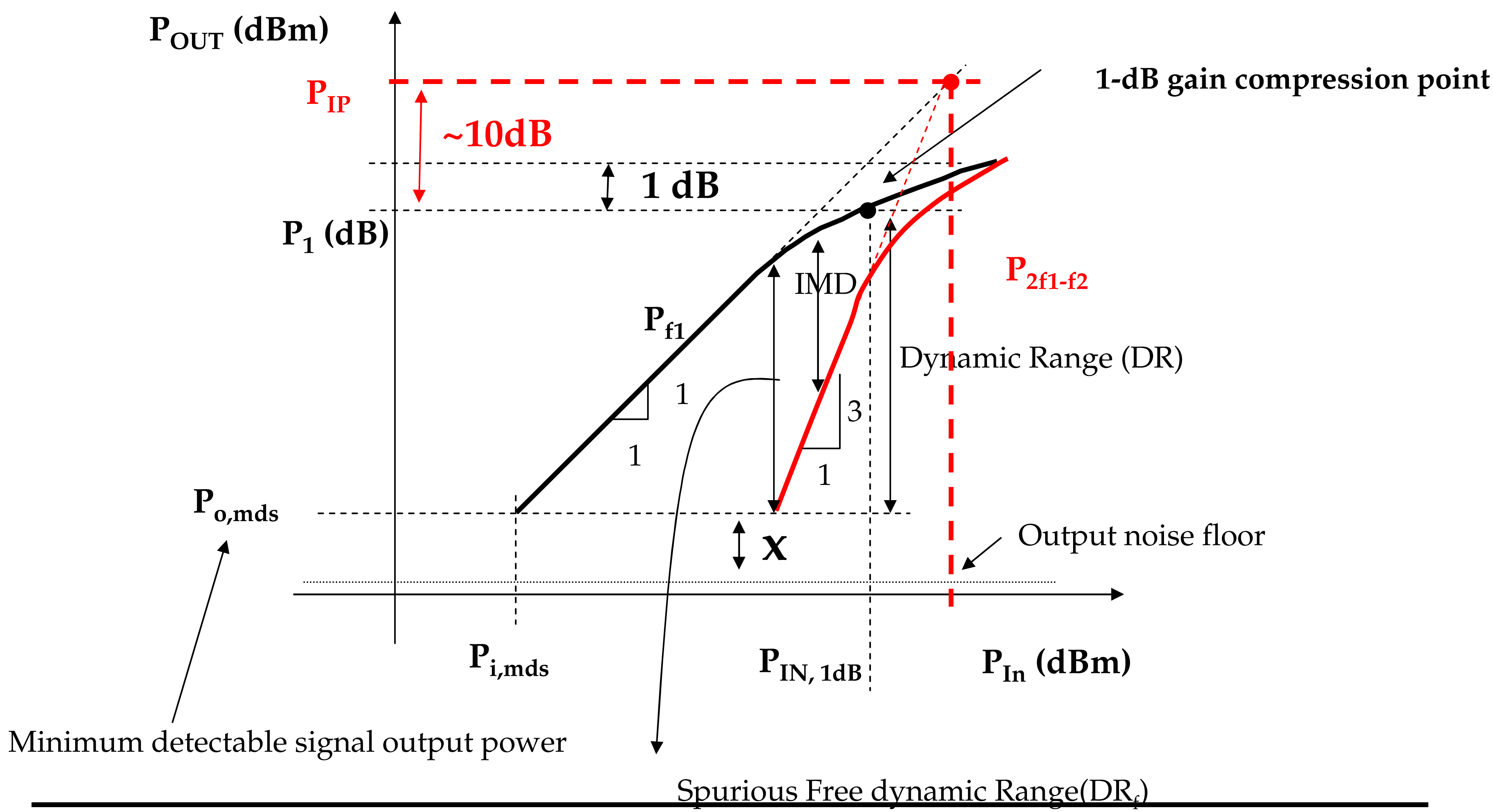
$$P_{IP}(dBm) = P_{1dB}(dBm) + 10$$

$$P_{2f1-f2} = 3P_{f1} - 2P_{IP}$$

$$P_{f1} - P_{2f1-f2} = \frac{2}{3}\left(P_{IP} - P_{2f1-f2}\right)$$

Distortion

Nonlinear characterization : Harmonic distortion and intermodulation



Distortion

Nonlinear characterization : Harmonic distortion and intermodulation

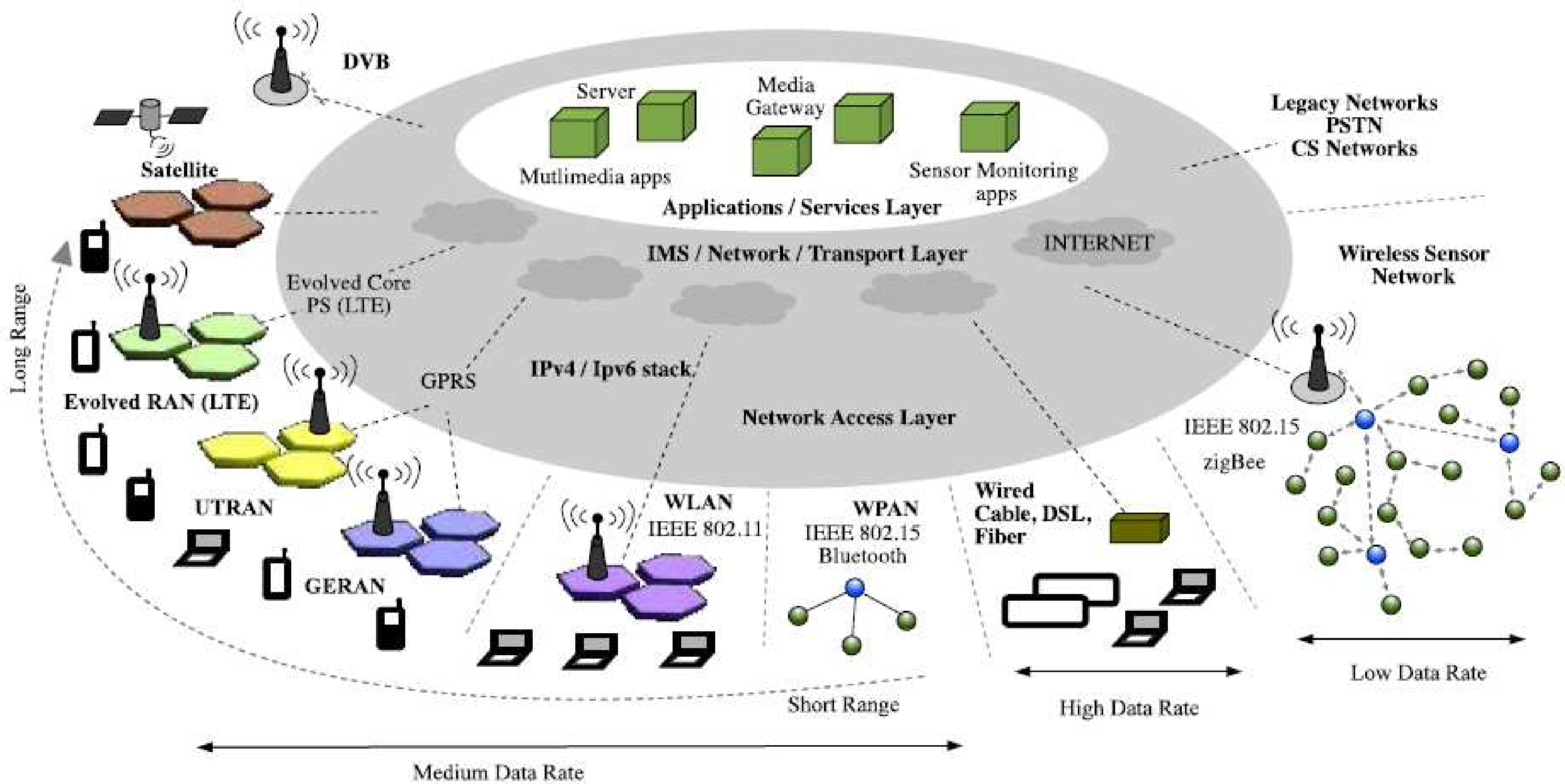
Spurious free dynamic range

Define-se DR_f of an amplifier is defined as the range [P_{f1}–P_{2f1-f2}] when P_{2f1-f2} is equal to the mininum detectable output signal

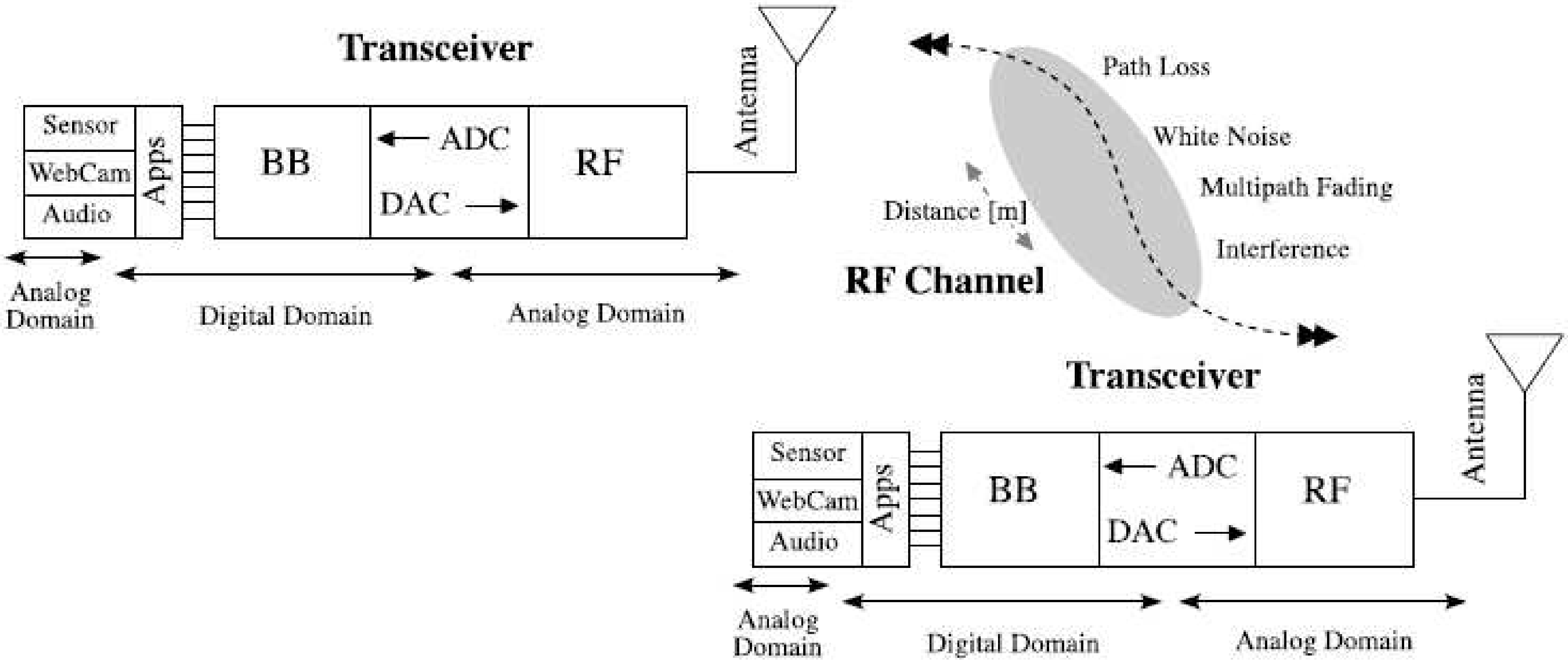
$$DR_f = \frac{2}{3}\left(P_{IP} - P_{o,mds}\right)$$

$$DR_f = \frac{2}{3}\left[P_{IP} + 174(dBm) - 10\log B - F(dB) - X(dB) - G_A(dB)\right]$$

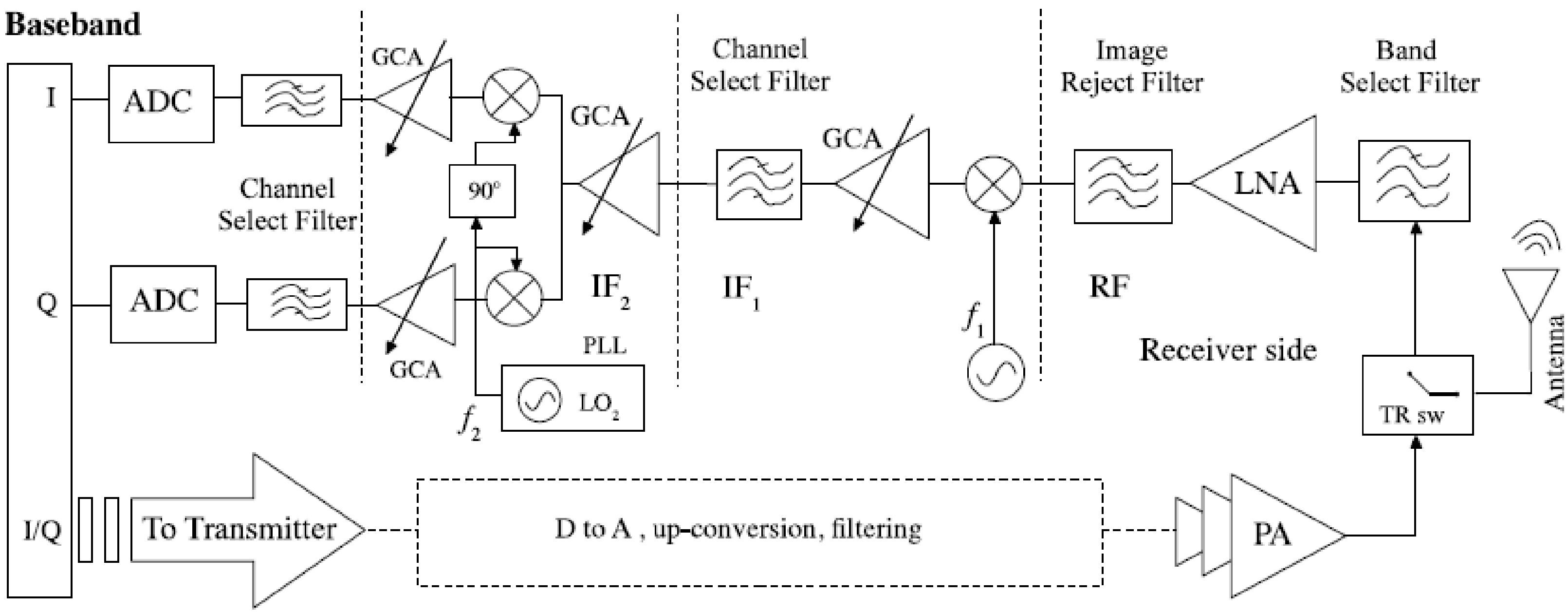
3- Transceiver Overview



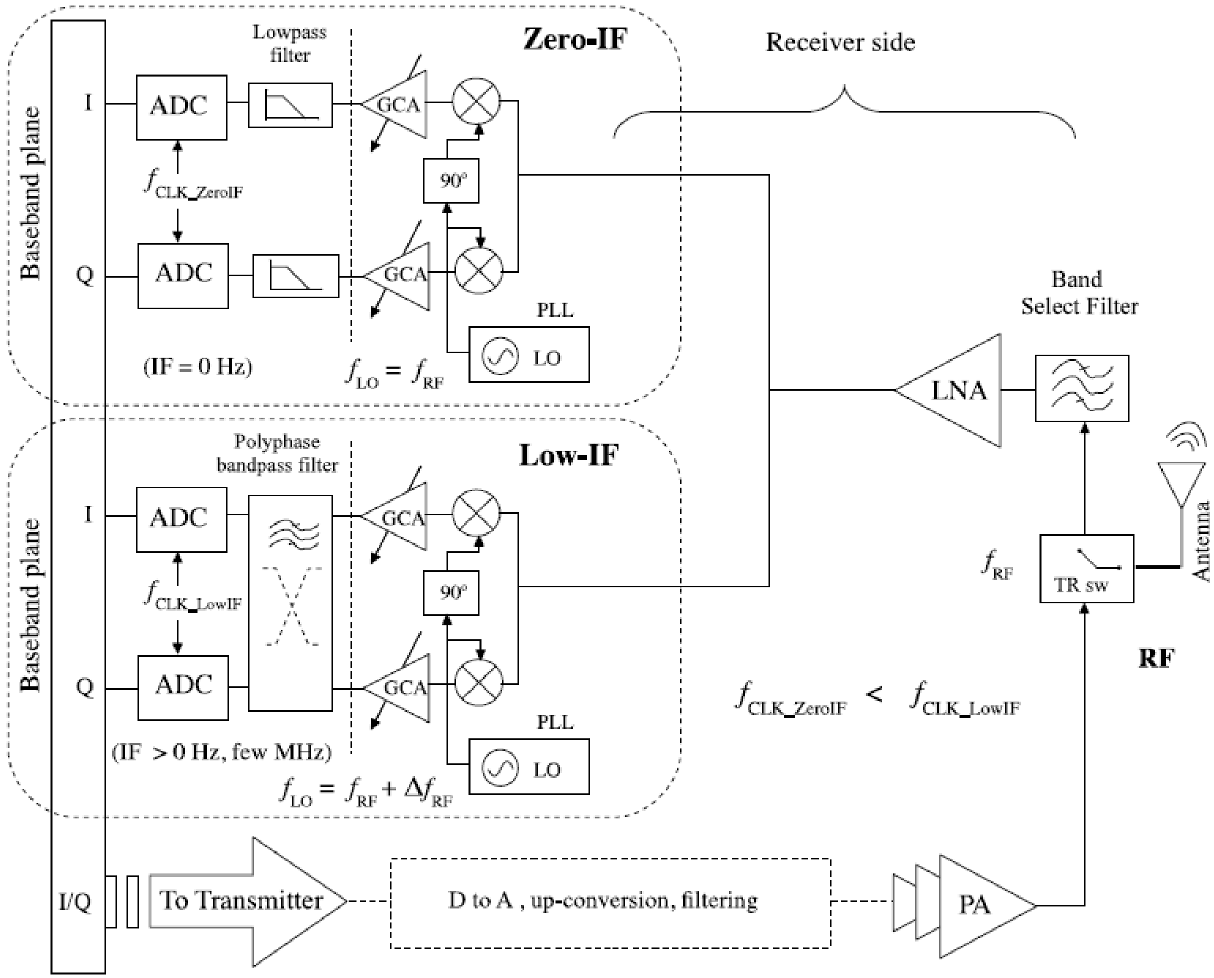
3- Transceiver Overview



Heterodyne based Transceiver



Zero-IF, Low-IF Transceiver



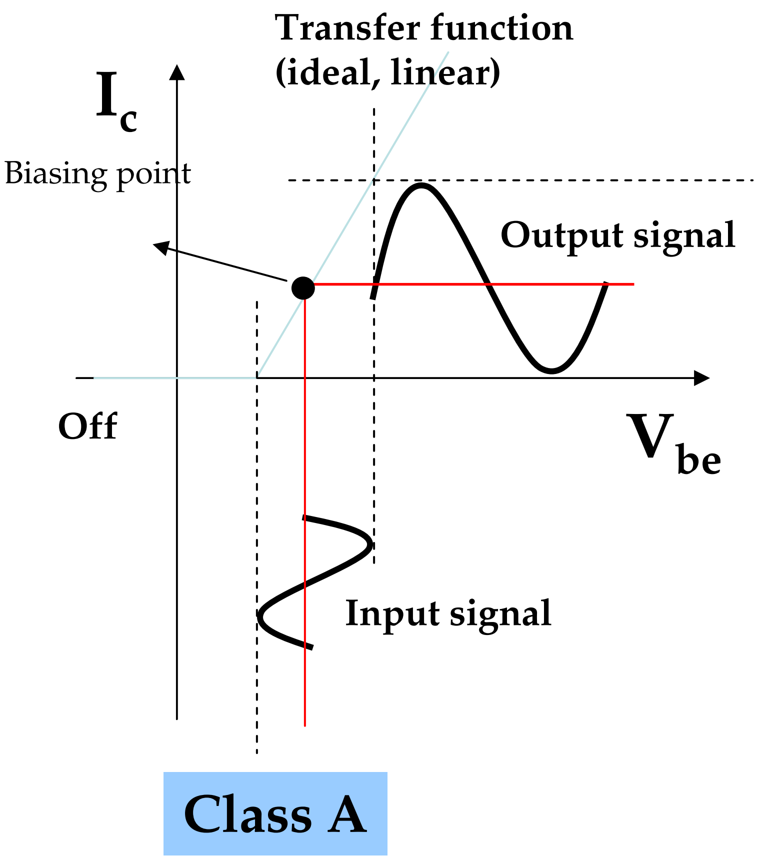
Low Noise Amplifier

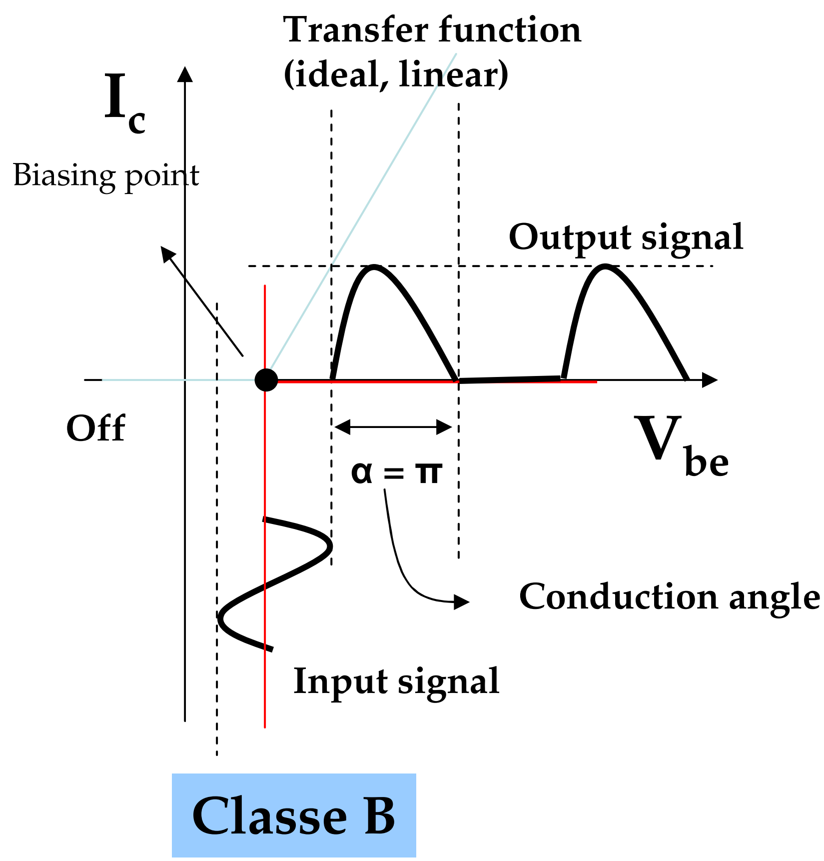
(
Chapter 23 of " Analog Design Essentials" by Willy Sansen, Springer)

Power Amplifiers

Class of operation of an amplifiwer

Based on the biasing point





Power Amplifiers

Class of operation of an amplifiwer

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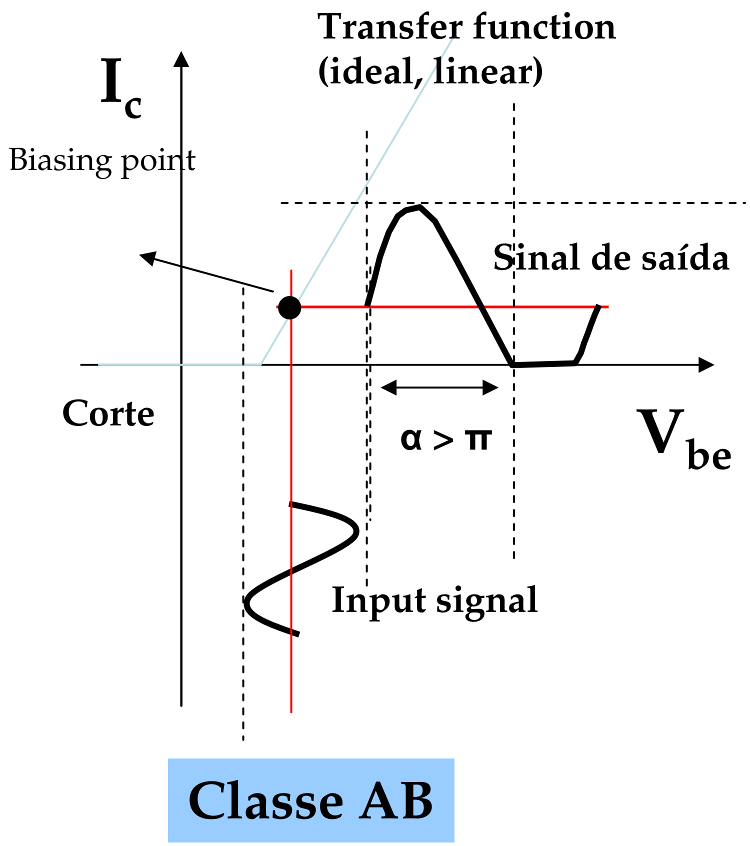
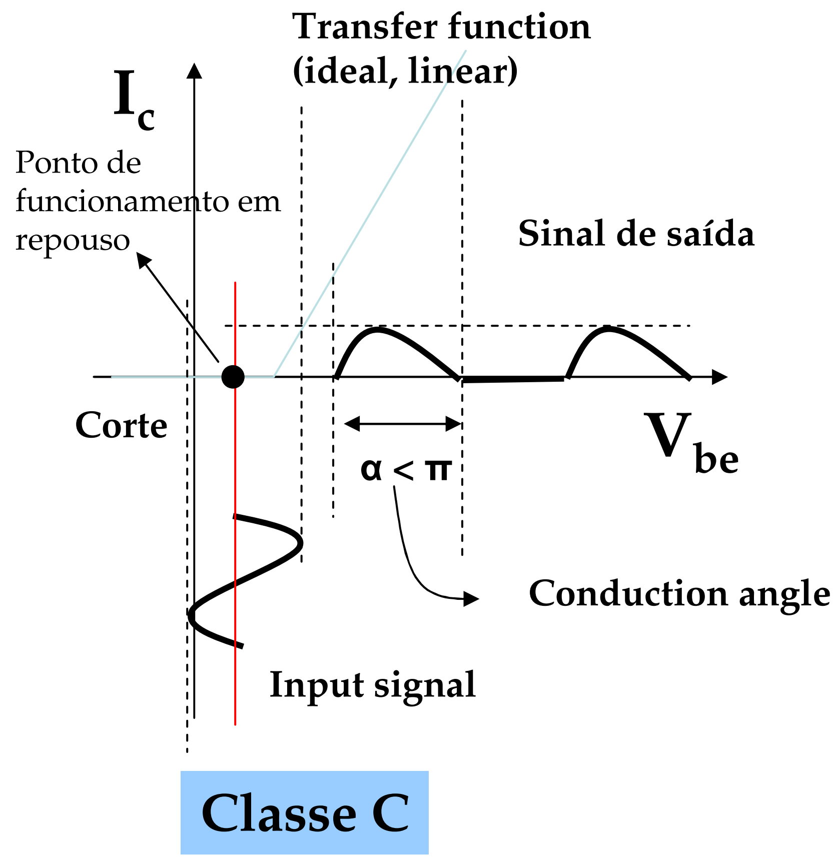




Power Amplifiers

Class of operation of an amplifiwer

Based on the biasing point

A polarização determina a classe de funcionamento em que o amplificador vai operar.

→ A diferenciação entre as diversas classes é baseada no ângulo de condução α, o qual indica a porção do ciclo do sinal em que a corrente está a fluir para a carga.

Classe A → $\alpha = 2\pi$
Classe B → $\alpha = \pi$
Classe AB → $\alpha =$ entre π e 2π
Classe C → $\alpha =$ entre 0 e π

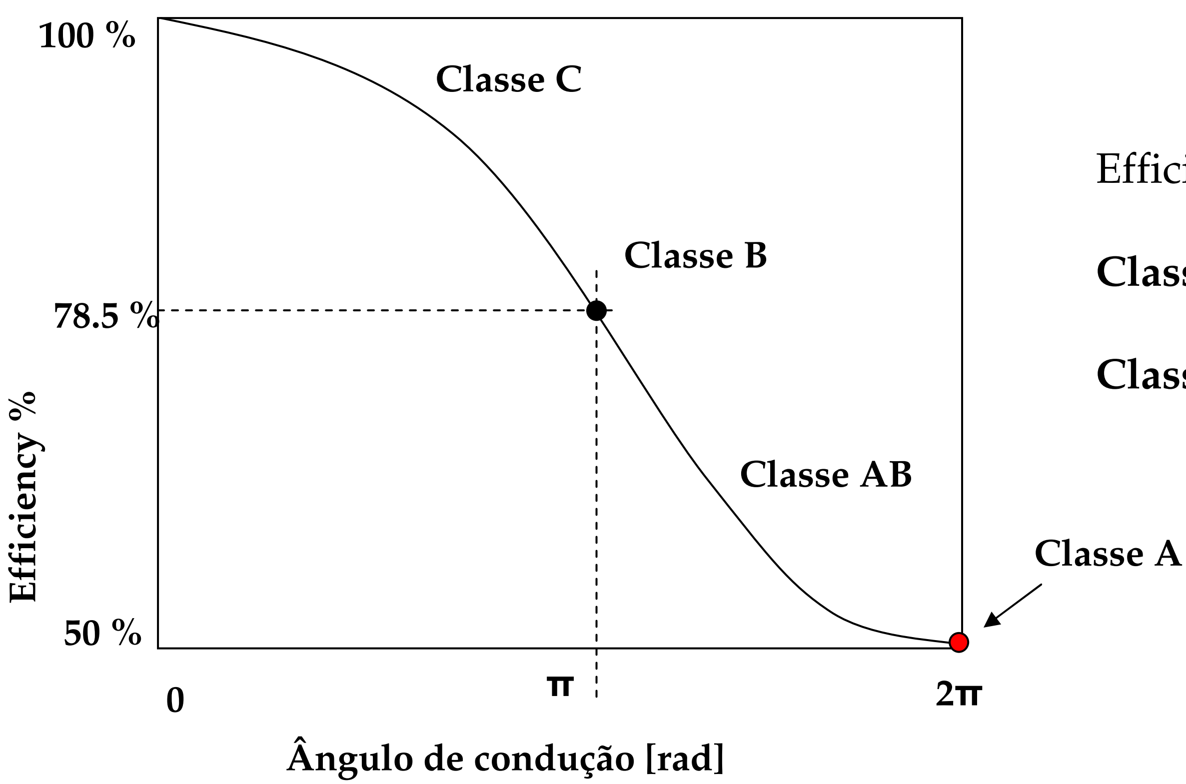
Rendimento

$$\eta = \frac{P_{RF}}{P_{Fonte}}\% = \frac{\alpha - \sin(\alpha)}{2\left[\alpha\cos\left(\alpha/2\right) - 2\sin\left(\alpha/2\right)\right]}$$

Power Amplifiers

Class of operation of an amplifiwer

Based on the biasing point



Efficiency

Class A → η=50%

Class B → η =78.5%