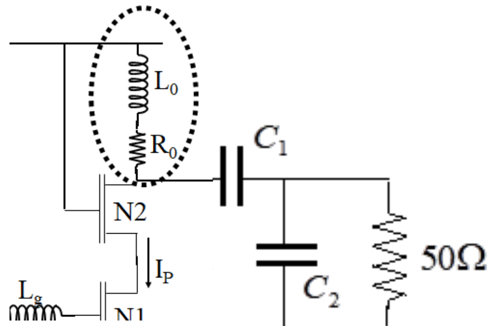
RF Basis



Exercise 1: It is aimed in this exercise to match an LNA to 50 Ω. The LNA load is an inductor of 2.12 nH having a quality factor Q_L=10. A narrow-band capacitive matching network, called up-stepper network, is thus necessary. We consider quasi-ideal capacitors. Working frequency is 3 GHz.

I-1. Transform the inductance series circuit towards a // one. Calculate the // resistance to be matched.

I-2. Transform the matching network loaded with 50 Ω into another network having only // elements.

I-3. Calculate the values of C_1 and C_2 in order to get matching at f_0=3GHz.

Exercise 2: We consider a LNA followed by a Mixer. The mixer has an IIP3(V_{peak}) = 1V and the LNA an $IIP3_{dBm}(50\Omega) = -6dBm$. Calculate the voltage gain of the LNA so as the whole receiver (LNA+Mixer) has an $IIP3_{dBm}(50\Omega) = -10dBm$. We recall the relationship between the power of a sine wave and its peak value is : $P = V^2/2R$.

Exercics 3 : We consider a RF receiver with the input matched to 50Ω and having a total noise figure of 8dB and an IIP3(V_{peak})=56mV.

1/ Demonstrate that $P_{si_{max}}(dBm) = \frac{2P_{IIP3}(dBm) + P_{NFi}(dBm)}{3}$

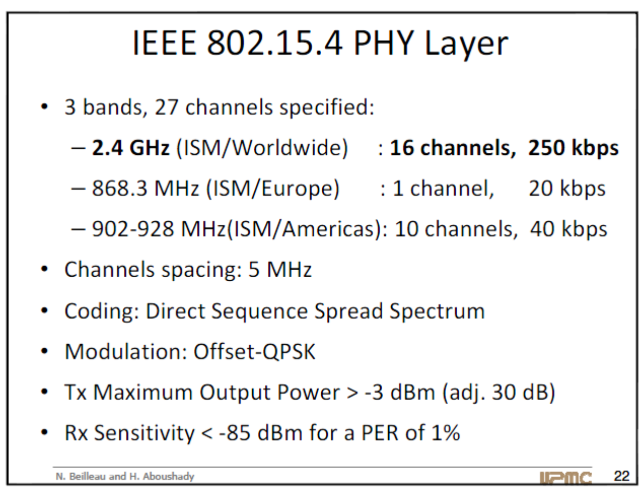
2/Calculate the dynamic and the SFDR of the receiver for 10dB output SNR and 100KHz bandwidth.

Exercice 4: We consider a RF receiver with the input matched to 50Ω and having a total noise figure Ftot = 8dB and an IIP3tot(dBm) = -13dBm.

1/ Show that the Noise Figure of a device is always greater than one (include the noise added by the device (V_{na})).

2/ Evaluate the DSP of the added noise by the receiver if the voltage gain of the receiver is 30dB.

3/ The receiver is based on a LNA followed by a mixer. The LNA has 20dB voltage gain and IIP3(V_{peak})=100mV. What is the Mixer IIP3 so that the whole receiver IIP3 is –13dBm ?



Exercice 5 : A receiver for the ZigBee Standard is based on an ideal external 50Ω filter, a LNA having a voltage gain G_{LNA}=15dB a noise figure F_{LNA} and an $IIP3_{LNA}$. The LNA is followed by an I.Q mixer having a noise figure $F_{IQ\text{-}MIX}$=21dB and an $IIP3_{DEMOD}(V_{peak})$=0.6 V_{peak}

1 System Distance Range: From the standard requirement, (emitted power, sensitivity, …) give the maximal distance range (@2.4GHz) if a fading margin of 10 dB is considered. Note that the loss in free space is given by:

$$Att_{dB} = 20.\log\left(\frac{4\pi d}{\lambda}\right)$$

2 Receiver Noise Factor: The chosen demodulator needs a 14dB output SNR to achieve a 0.01 PER. What is the requirement on the receiver noise figure in this context?

3 Receiver IIP3: A 45 dB dynamic is required. Give the receiver IIP3 in dBm.

4 LNA requirements:

a) Give the LNA noise figure if we suppose that the LNA and the DEMOD are matched together.

b) The IIP3 of the DEMOD following the LNA is 0.6 Vpeak. Gives the IIP3 of the LNA in dBm.

Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr
Grenoble INP
TIMA lab and PHELMA engineering School

Low Noise Amplifier LNA

Parameters of the Design Kit (CMOS RF 0,35 µm)

NMOS : K_n=80µA/V^2 ; k_{en}=0,03µm/V ; Vt_{n0}=0,57V ; $\gamma = 2$
C_{OX}=5.10^{-3}pF/µm^2 ; C_{db}=C_{sb}=1.10^{-3}pF/µm and L_{min}=0,35µm
NMOS current in saturation region: $I_d = K_n \frac{W}{L}(V_{gs} - Vt_n)^2$

I- Preliminary Study : RF receiver

The figure1 gives an heterodyne RX architecture used to recover a narrow band signal centered at f_0=2GHz. The Rx is based on a LNA having a voltage gain $G_1 = 10$, a noise factor F_1 and an input 3rd order intermodulation point $IIP3_1$. An image rejection filter follows the LNA. This filter is not integrated and is based on passive LC devices which are quasi-ideal. The losses in the frequency band are represented by a voltage gain $G_2 = 0,8$. A mixer follows the filter which achieves a noise factor of F_3=10 and an IIP3(V)=2Vc.

Due to the filter which is not integrated, all the devices are matched to 50Ω.

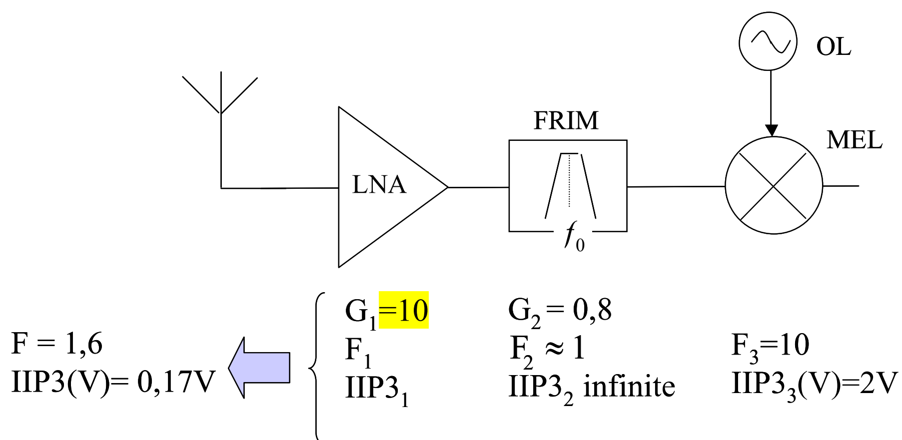


Figure 1

We want to define the requirement of the LNA to have the Rx performances so as : F_{tot} =1,6 and $IIP3_{tot}$=0,17V_{peak} or 0, 17$V_{crête}$ in French.

I-1 Why the filter IIP3 is quasi-infinite and the noise factor equal to 1 ?
I-2 Give the value of F_1 and $IIP3_1$ of the LNA
Answers : $IIP3_1 \approx 0,23Vc$ *et* $F_1 \approx 1,46$
I-3 The signal bandwidth is 100 kHz, calculate the sensitivity P_{se_min}(dBm) and the dynamic (in dB) of the Rx if we target an output $SNR_O = 10dB$

I-4 Demonstrate that the input impedance of a degenerated LNA is: $Z_i = \left(L_s p + \frac{1}{C_{gs} p}\right) + \left(\frac{gm_N . L_s}{C_{gs}}\right)$

II : Sizing of the LNA

The topology given in the figure 2 represents a cascode LNA based on two NMOS N_1 and N_2 with a RLC load. The inductors L_G and L_S with the input capacitor C_{gs} (N_1) achieves an input network having an input impedance $Z_e = 50\Omega$ at f_0 =2GHz. The output network (L_L, C_1, C_2) achieves at f_0 the impedance matching between the input impedance of the filter (50Ω) and the output load of the amplifier (RL).

Based on the preliminary study the LNA requirements are V_{dd}=3V ; $Z_E = Z_s = 50\Omega$; F=1.46 ; $IIP3(V) = 0.23V_{peak}$ and $G_v = \frac{\delta V_s}{\delta V_e}$

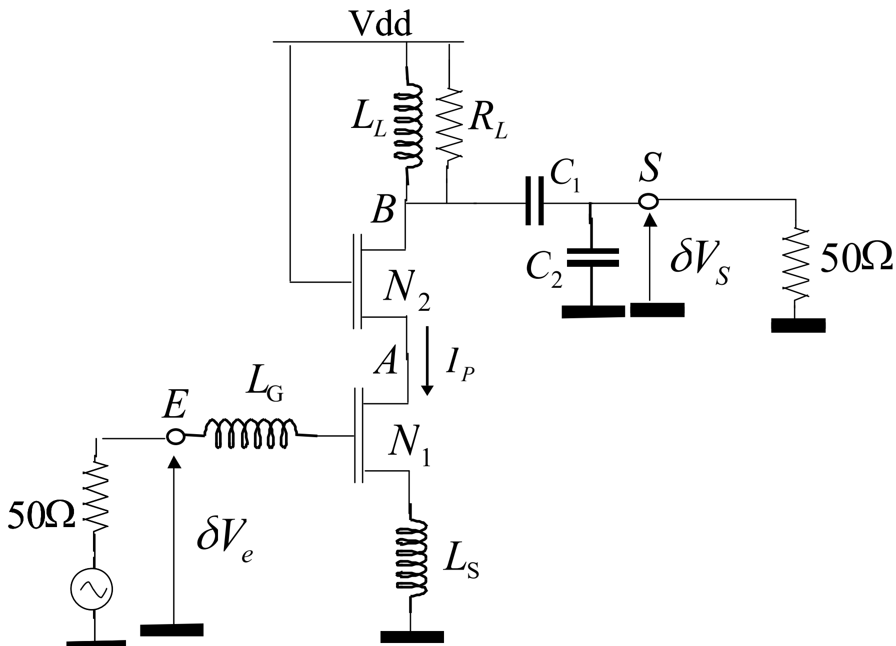


Figure 2

We assume that the non-linearities (IIP3) are due to the transconductance of the MOS N_1 and that only the thermal noise of the channel affects the noise figure as it is given in the following relationships:

$IIP3(V) = 0{,}64\sqrt{(V_{gs} - V_{tn})}\frac{1}{Q_e}$ *(1)* and $F = 1 + \frac{2}{50.gm}\frac{1}{Q_e^2}$ *(2) with Q_e the Q factor of the input network and gm the transconductance of N_1.*

II - 1 Based on the requirements and the equations (1) et (2), give a relationship between Q_e and $(V_{gs} - V_{tn})$ from one side and between Q_e and the $gm(N_1)$ from the other side.

The inductor L_L at the output resonates at f_0 with the equivalent capacitor of the network

C_1-C_2.

II – 2 We set $L_L = 5nH$. Calculate the value of R_L if we want the load (R-L-C) at the node B to have a Q = 20 at f_0.
Remark : *To make the calculus we consider the real part of the impedance seen at the node B given by the network (C_1-C_2) which achieved the impedance matching between R_L and* 50Ω.

II - 3 By considering that the matching network (*C_1-C_2*) is lossless, express the ratio $\frac{\delta V_S}{\delta V(B)}$ at f_0.

II - 4 Then, give an expression of the voltage gain $G_V = \frac{\delta V_S}{\delta V_E}$ as a function of Q_e and $gm(N_1)$. With the results of the question I-1, calculate the values of Q_e and $gm(N_1)$.

II - 5 Calculate the bias voltage of $(V_{gs} - V_{tn})$ and the bias current I_p.

II - 6 Calculate the $\left(\frac{W}{L}\right)$ ration of N_1. With $L = L_{\min} = 0{,}35\mu m$, calculate the value of $C_{gs}(N_1)$.
Then give the value of C_0 that must be added in parallel with $C_{gs}(N_1)$ to obtain the Q_e calculated in II-4.

II – 7 Calculate the transition frequency f_T of the MOS (we consider C_0 in parallel with C_{gs} (N_1)). Then, give the values of L_G and L_S to be matched to $Z_e = 50\Omega$ at f_0.

II – 8 Calculate the value of the capacitance that should present the (*C_1 - C_2*) network at node B in order to have this capacitance resonating with inductance L_L at frequency f_0. Then deduce the capacitance values C_1 and C_2 that enable matching between R_L and the 50Ω resistance.

II – 9 N_2 has identical dimensions to N_1's. With the help of the results presented during lecture, and by considering that the impedance brought at node A is mainly due to capacitance $C_{gs}(N_2)$, calculate the noise ratio (in %) added by transistor N_2 at the output as compared to the noise that would be generated if the source of N_2 were directly connected to ground.

PHELMA EMT 2A / SEI-SyRF 3A / WICS
RFIC, TD3

RF Mixer

Parameters of the Design Kit (CMOS RF 0,35 µm)

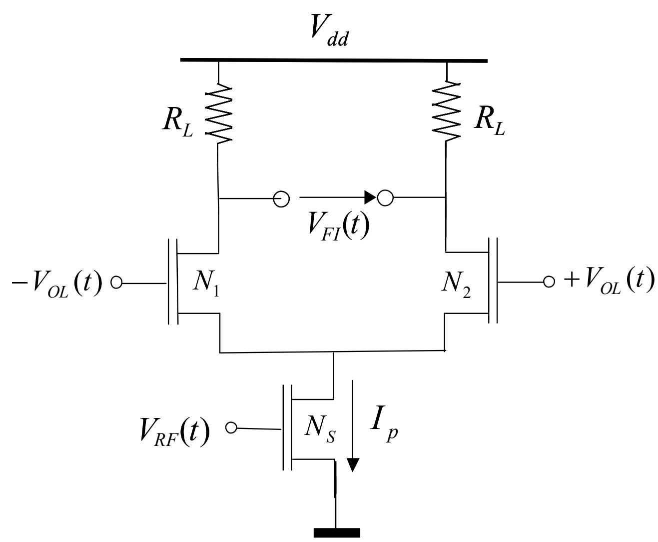
NMOS : K_n=80µA/V^2 ; k_{en}=0,03µm/V ; Vt_{n0}=0,57V ; $\gamma = 2$
C_{OX}=5.10^{-3}pF/µm^2 ; C_{db}=C_{sb}=1.10^{-3}pF/µm and L_{min}=0,35µm

NMOS current in saturation region: $I_d = K_n \dfrac{W}{L}(V_{gs} - Vt_n)^2$

I - Questions related to the course

I-1 In a mixer, why the peak voltage of the quasi-sine wave from the LO should be as large as possible ?

I-2 Considering integration issues, what are the advantage of an homodyne structure?



II – Active Mixer Design

We consider the active mixer described below. It is based on a transconductor (NS) a differential switch (N1 and N2) and two loads (RL).

II-1 Knowing the expression of a modulated signal and of the local oscillator signal are:
$v_{rf}(t) = V.\cos(\omega_{RF}t + \varphi(t))$ and $v_{ol}(t) = B\sin(\omega_{LO}t)$, gives the expression of the conversion gain Gc as a function of $gm(N_S)$ and R_L. With R_L=500Ω, calculate the $gm(N_S)$ value for $G_C = 15dB$.

II-2 We consider that for the used technology the IIP3 of the mixer is given by: $IIP3(V) = 0.64\sqrt{(V_{gs} - V_{tn})}_{N_S}$. Calculate the bias current I_p so as the IIP3 is 0.43V_{peak}. Then give the ratio $\left(\dfrac{W}{L}\right)$ of N_S

The mixer is biased so that the Ns is at the limit of the saturation region during the transition state. We will assume that during this state the applied voltage at the LO input is 1.5V.

II-3 Give the ratio $\dfrac{W}{L}(N_{1,2})$ and then give W.

II-4 By considering that the impedance at the Ns input is very high, give the NF as a function of the added noise during the transition state ($e_{n\_tran}^2$) and the switched state ($e_{n\_sw}^2$). Calculate the noise contribution in each state and the noise factor. The duty cycling is assumed to be 20%.

II-5 Give an expression of the noise factor as a function of the different *gm*.

Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr

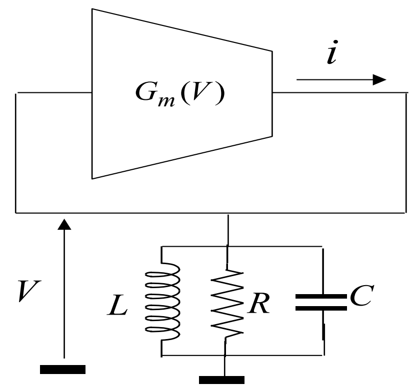
Grenoble INP
TIMA lab and PHELMA engineering School

II-6 Propose a solution to increase the *gm(Ns)* without increasing the noise contribution of N(1,2).

Local Oscillator

The oscillator represented beside is studied. It is based on an LC oscillator and a transductor. The frequency resonance of the LC tank is 2GHz. It is based on a 2nH inductor having a Q factor of $Q_L = 10$. The $G_m(V)$ of the transconductor stage is non-linear and its behavior can be approximated by the following relationship:

$$i \approx G_{m0}.V + \frac{3}{4}\alpha_3 V^3 . \text{ (with } \alpha_3 = \text{- 2,66 mA/V}^3\text{)}$$

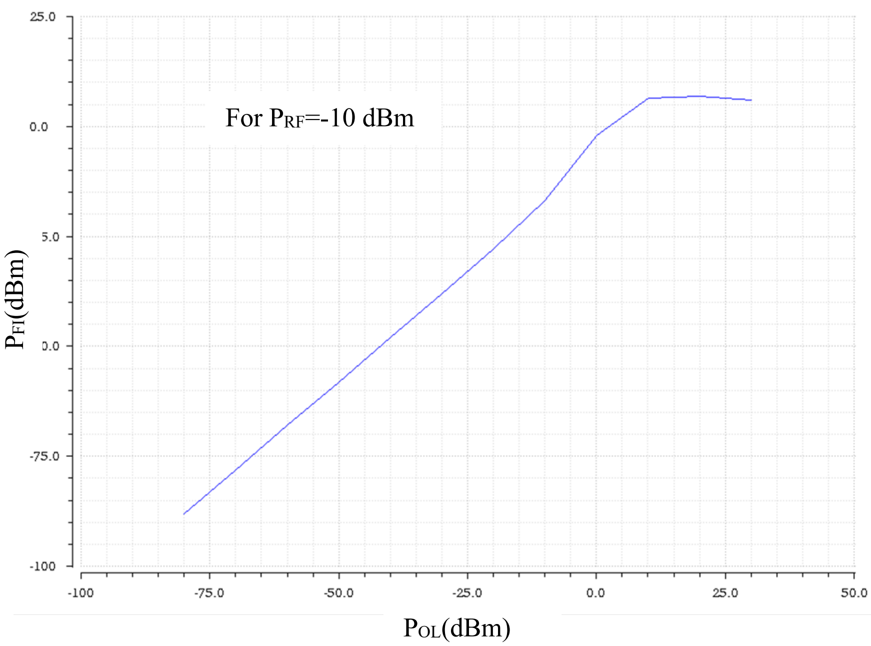


1 Give the oscillation conditions and give the value of G_{m0} in order to have an output magnitude of 1V.

This oscillator, connected to a mixer, is used to transpose a signal centered at f_{RF} to an intermediate frequency given by $f_{FI} = f_{OL} - f_{RF}$. We suppose that the oscillator achieves a phase noise centered at f_{OL}. The output power $P_{f_{FI}}(signal)$ as a function of the oscillator power $P_f\,(oscillateur)$ is given on the figure bellow for an input power of the RF signal of P_{fRF} = -10dBm.

2 Explain the mixer behavior and give $P_{FI}(P_{OL}, P_{RF})$ for the two operation regions observed on the curve. Give the gain for the different region. What is the optimal P_{OL}?

3 We suppose a channel to be received centered at f_{RF} and a blocker of 60dBc and centered at 1MHz from the channel ($f_{Block} = f_{RF} + 1MHz$). Draw a figure representing the several harmonics due to the channel, the blocker and the oscillator.



4 Calculate the phase noise in dBc $L(1MHz)$ that should be achieved by the oscillator at the frequency $f_{OL} + 1MHz$ if we want a SIR (signal to interferer ratio) equal to 40dB after mixing to f_{FI}. We suppose that the optimal P_{OL} is applied.

Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr

Grenoble INP
TIMA lab and PHELMA engineering School

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Power amplifier

Parameters of the Design Kit (CMOS RF 0,35 µm)

NMOS : K_n=80µA/V^2 ; k_{en}=0,03µm/V ; Vt_{n0}=0,57V ; $\gamma = 2$
C_{OX}=5.10^{-3}pF/µm^2 ; C_{db}=C_{sb}=1.10^{-3}pF/µm and L_{min}=0,35µm

NMOS current in saturation region: $I_d = K_n \frac{W}{L}(V_{gs} - Vt_n)^2$

This exercise aims at studying a power amplifier of class B with a supply voltage.
The power amplifier (fig1) is constituted of an NMOS, a choke inductance L_C of very high value (external to the circuit) and a matching network, also acting as a filtering network, (C_1, L, C_2), integrated in amount to load $R_L = 50\Omega$. Input RF signal is narrow band, centered around $f_0 = \frac{\omega_0}{2\pi} = 2GHz$.

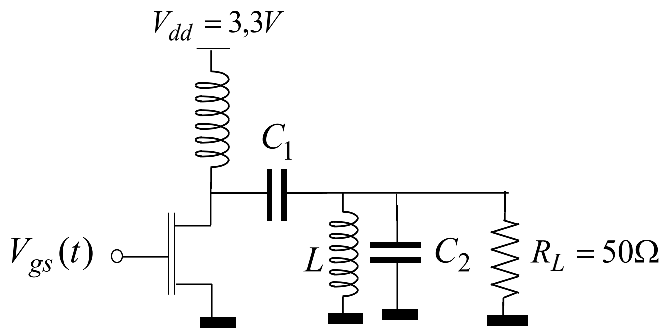


Figure 1

By considering, as a first approach, a lossless network (C_1, L, C_2).

I- Class A:

The PA is biased in A class A. For a maximal available power P_{max} delivered to load $R_L = 50\Omega$, the voltage between drain and source nodes is given by: $V_{ds}(t) = V_{dd} - V_{dd}\cos\omega_0 t$.

I-1 Draw Vgs(t), Vds(t), Ids(t). Give the min, max and mean values.

I-2 $V_{ds}(t) = V_{dd} - V_{dd}\cos\omega_0 t$. Calculate the real impedance value *R'* synthesized by the network (C_1, L, C_2, R_L) between the NMOS drain and source in order that $P_{ac\_max} = 30dBm$.

I-3. Give I_{DS0} to be at the center of the load line.

II- Class B:

II-1 Draw Vgs(t) and Vds(t), then draw Ids(t) and deduce the biasing conditions. Knowing $V_{ds}(t) = V_{dd} - V_{dd}\cos\omega_0 t$, Calculate the real impedance value R' synthesized by the network (C_1, L, C_2, R_L) between the NMOS drain and source in order that $P_{\max} = 30dBm$.

II-2 Give, for the maximum available power, the peak magnitude of the sinusoidal voltage $V_S(t)$ over resistance R_L.

II-3 For the PA to operate in class B, the drain current consists in half a sinus wave of magnitude I_m. Give the magnitude $I_{ds}[1]$ of the current fundamental component at 2GHz corresponding to output power $P_{\max}$. Deduce the value of I_m for $P_{\max}$.

II-4 Give the value of the current mean component $I_{ds}[0]$ for $P_{\max}$.

By considering the network of figure 2a equivalent to the initial network of figure 2b.

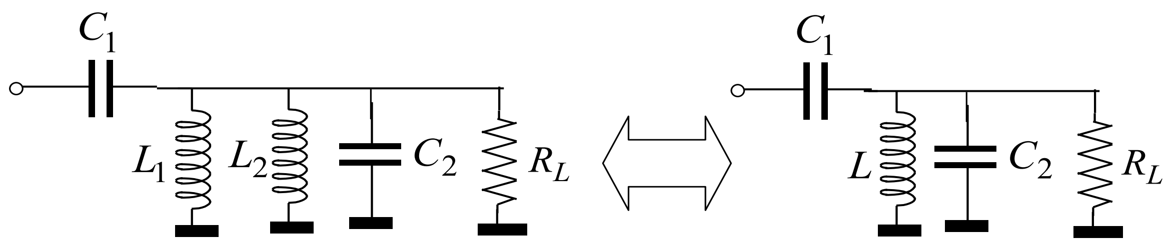


Figure 2a Figure 2b

II-5 The network (L_2, C_2, R_L), resonating at frequency f_0, enables to filter the residual harmonics of $V_S(t)$. Calculate the values of L_2 and C_2 for a filter quality factor $Q = 5$ at pulsation ω_0.

II-6 Calculate the values of network (L_1, C_1), resonating at frequency f_0, that enable to decrease the value of the R_L load down to the value R' already determined at question 1. Deduce the value of the inductance L of the initial network of figure 2b.

(L,C_1,C_2) network consist in integrated components on silicon. The network losses are represented by a resistance $R_s = 2\Omega$ in series with C1.

II-7 Calculate the new values of network *(L_1,C_1)* in order that the network show between the NMOS drain and source the same real impedance R' value than already calculated at question 1. Calculate the real power transmitted to the R_L load and the power amplifier efficiency η.

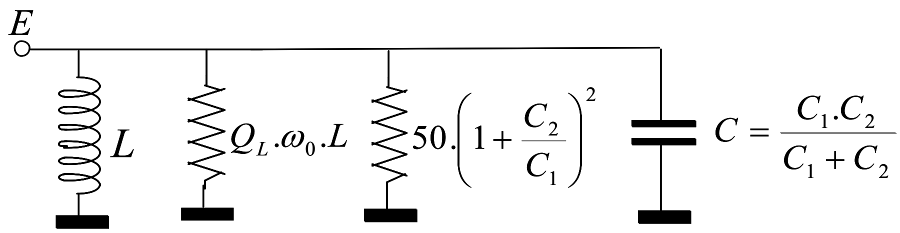
Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr

Grenoble INP
TIMA lab and PHELMA engineering School

TD1 – Matching-Système - Correction

Solution ex. 1

I-1.



I-2. Avec $L = 2{,}12nH$, $Q_L = 10$ et $f_0 = 3GHz = \frac{1}{2\pi.\sqrt{LC}}$ on obtient :

$C_1 = 2{,}05pF$ et $C_2 = 3{,}76pF$

Solution ex. 2

On a $IIP3_{dBm}(50\Omega) = -6dBm \Rightarrow IIP3(V) = 158mV$ pour le LNA

$IIP3_{dBm}(50\Omega) = -10dBm \Rightarrow IIP3(V) = 100mV$ pour le circuit total

On applique la relation $\frac{1}{[IIP3(V)]^2} = \frac{1}{[IIP3(V)_{LNA}]^2} + \frac{G_v^2}{[IIP3(V)_{mixer}]^2}$ soit $G_V = 7{,}7(17dB)$

Solution ex. 3

On a : +

$$(Dyn)_{dB} = P_{SI_{\max}} - P_{SI_{\min}} = \frac{2(P_{IIP3} - P_{NFI})}{3} - (S/B)_s$$

Calculons le plancher de bruit $P_{NFI}(dBm) = 8dB - 174dBm + 10\log(10^5) = -116dBm$.

On a $IIP3(V) = 56mV \Rightarrow IIP3(dBm) = P_{IIP3}(dBm) = -15dBm$ soit :

$$(Dyn)_{dB} = \frac{2.(116-15)}{3} - 10 \approx 56dB$$

$$(SFDR)_{dB} = P_{SI_{\max}} - P_{NFI} = \frac{2P_{IIP3}(dBm) - P_{NFI}(dBm)}{3} - P_{NFI}(dBm) = \frac{2}{3}(P_{IIP3}(dBm) - P_{NFI}(dBm)) = 67.7$$

Solution ex. 4

I -1 $NF = 10\log F = 10\log\left(1 + \frac{e_{ns}^2(recept.)}{\frac{1}{4}4KT50.G_V^2}\right)$, soit : $e_{ns}^2 = 1{,}15.10^{-15}V^2/Hz$

I -2 IIP3recept. (V) = 71mV soit $\frac{1}{(71.10^{-3})^2} = \frac{1}{(100.10^{-3})^2} + \frac{10^2}{(IIP3_{mel.})^2}$ soit : $IIP3_{mel} = 1V$

Solution ex. 5 :

1 :

$$P_r = P_e - Att - F$$

$$Att = 20.\log\left(\frac{4\pi d}{\lambda}\right) = P_e - F - P_r = 72dB$$

$$d = 38.8m$$

2 :

$$F = \frac{SNR_i}{SNR_o} \Rightarrow SNR_o = P_{si} - P_{ni} - F$$

$$F = S_e - \left(-174dBm + 10\log(5Mhz)\right) - SNR_o = -85 + 107 - 14 = 8$$

3 :

$$Dyn = \frac{2}{3}\left[IIP3 - P_{nfe}\right] - SNR_o$$

$$IIP3 = \frac{3}{2}\left[Dyn + SNR_o\right] + P_{nfe} = -10.5dBm$$

4 :

$$F_{tot} = F_{LNA} + \frac{F_{mix} - 1}{(Gv_{LNA})^2} \Rightarrow F_{LNA} = F_{tot} - \frac{F_{mix} - 1}{(Gv_{LNA})^2} = 3.8dB$$

$$IIP3_{LNA} = \sqrt{\frac{1}{\frac{1}{IIP3_{tot}^2} - \frac{G_{vLNA}^2}{IIP3_{MIX}^2}}}$$ =0,05Vpic=196mVcrête

Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr

TD2 – LNA - Correction

I- Etude préliminaire d'un système de réception en radiofréquence

I-2 On a $\frac{1}{(\mathrm{IIP3})^2} = \frac{1}{(\mathrm{IIP3}_1)^2} + \frac{G_1^2}{(\mathrm{IIP3}_2)^2} + \frac{G_1^2.G_2^2}{(\mathrm{IIP3}_3)^2}$ avec $\mathrm{IIP3}_2 \approx \infty$

Soit : $\frac{1}{(\mathrm{IIP3}_1)^2} = \frac{1}{(\mathrm{IIP3})^2} - \frac{G_{v,1}^2.G_{v,2}^2}{(\mathrm{IIP3}_3)^2} = \frac{1}{(\mathrm{IIP3}_1)^2} = \frac{1}{(0{,}17)^2} - \frac{64}{(2)^2}$, soit : $\mathrm{IIP3}_1 = 0{,}23\mathrm{V}$

On a $F = F_1 + \frac{(F_2-1)}{G_1^2} + \frac{(F_3-1)}{G_1^2.G_2^2}$ avec $F_2 \approx 1$, soit $F_1 = F - \frac{(F_3-1)}{G_1^2.G_2^2} = 1{,}6 - \frac{9}{64}$, soit : $F_1 \approx 1{,}46$

I-3 On a $\mathrm{Dyn}_{(\mathrm{dB})} = \frac{2\left(\mathrm{IIP3}_{(\mathrm{dBm})} - \mathrm{Pe}_{\mathrm{B(dBm)}}\right)}{3} - \left(\frac{S}{B}\right)_{\mathrm{S(dB)}}$

avec $Pe_{B(dBm)} = 10\log(1{,}6) - 174 + 10\log(10^5) = -122\,dBm$ et $IIP3_{(dBm)} = 10\log\frac{(0{,}17)^2}{2.50.10^{-3}} \approx -5{,}4dBm$

soit : $\mathrm{Dyn}_{(\mathrm{dB})} = \frac{2(-5{,}4+122)}{3} - 10 \approx 68dB$

II- Etude de l'amplificateur faible bruit

II-1 On a $IIP3(V) = 0{,}64\sqrt{(V_{gs}-V_{tn})}\frac{1}{Q_e}$, soit : $\sqrt{(V_{gs}-V_{tn})}\frac{1}{Q_e} = 0{,}36$

De même : $F = 1 + \frac{2}{50.gm}\frac{1}{Q_e^2}$, soit $gm.Q_e^2 \approx 0{,}087$ A/V

II – 2 L'impédance réelle ramenée entre B et la masse par le réseau d'adaptation est égale à l'impédance de source R_L.

On a donc $Q_S = \frac{R_L/2}{2\pi.L_S.f_0} = 20$ soit $R_L \approx 2{,}5K\Omega$

II – 3 On applique la loi de conservation de la puissance avec $\frac{(\delta V_B)^2}{2.R_L} = \frac{(\delta V_S)^2}{2.50}$ à f_0

Soit $\frac{(\delta V_S)}{(\delta V_B)} = \sqrt{\frac{50}{2500}} \approx 0{,}14$

II – 4 On a $G_V = \dfrac{\delta V_S}{\delta V_E} = \dfrac{\delta V_S}{\delta V_B}\dfrac{\delta V_B}{\delta V_E} = Q_e.gm(N_1).\dfrac{R_L}{2}\dfrac{\delta V_S}{\delta V_B} = 175.Q_e.gm(N_1) = 10$

soit : $Q_e.gm(N_1) = 0{,}057$. D'après II-1 : $gm(N_1).Q_e^2 \approx 0{,}087$, on en déduit : $Q_e \approx 1{,}5$

et $gm(N_1) = \dfrac{0{,}057}{1{,}5} \approx 38mA/V$

II – 5 On a d'après II-1 : $\sqrt{(V_{gs}-V_{tn})}\dfrac{1}{Q_e} = 0{,}36$, soit : $(V_{gs}-V_{tn}) \approx 0{,}3V$

On a aussi : $gm(N_1) = \dfrac{2I_P}{(V_{gs}-V_{tn})} = 38.10^{-3}$ soit : $I_P \approx 5{,}7mA$

II – 6 On a aussi : $gm(N_1) = 2\sqrt{K_n.\dfrac{W}{L}.I_p} = 38.10^{-3}$ soit : $\left(\dfrac{W}{L}.\right)_{N1}$ =1150=792

soit avec L = 0,35µm : W .~400µm ~277µm

On a $C_{GS}(N_1) \approx \dfrac{2}{3}C_{OX}.W.L$ ~0,47pF~0,32pF. Le coefficient Q_e à la résonance en entrée satisfait la relation $Q_e = \dfrac{1}{(C_0 + C_{GS}).2\pi.50.f_0}$, soit : $C_0 + C_{GS} \approx 1{,}06\,pF$ et donc :

C_0~0,6pF~0,74pF

II – 7 $f_T = \dfrac{gm(N_1)}{2\pi.(C_0 + C_{GS})} = 5{,}7\,GHz$

La condition d'adaptation en entrée s' écrit : $50 = L_S.2\pi.f_T$ soit $L_2 \approx 1{,}4nH$

La condition de résonance en entrée impose : $4.\pi^2\left(L_G + L_S\right)\left(C_0 + C_{GS}\right).f_0^2 = 1$

Soit $\left(L_G + L_S\right) \approx 6nH$, et donc : $L_1 \approx 4{,}6nH$

II – 8 D'après le cours, la capacité équivalente au réseau C_1-C_2 ramenée au point B est équivalente à ($C_1//C_2$). Cette capacité résonne à la fréquence f_0 avec L_S, ce qui implique la relation : $4.\pi^2(L_S)(C_1//C_2).f_0^2 = 1$ soit $(C_1//C_2) \approx 1{,}26\text{pF}$

De plus, toujours d'après le cours, l'adaptation entre la résistance de 50Ω et R_L implique :

$50.\left(1 + \dfrac{C_2}{C_1}\right)^2 = R_L$ soit : $\dfrac{C_2}{C_1}$ =4=6. Avec la relation précédente on en déduit :

C_2 = 6,3pF=9pF et C_1=1,57pF

II – 9 D'après le cours le pourcentage N est donné par le relation : N=$\dfrac{\omega_0^2}{\omega_0^2+\omega_A^2}$ avec

$\omega_A = \dfrac{gm(N_2)}{C(A)} = \dfrac{gm(N_2)}{C_{GS}(N_2)} = \dfrac{gm(N_1)}{C_{GS}(N_1)} = \dfrac{38.10^{-3}}{0{,}47.10^{-12}} = 81\,rd/s$ soit N=0,06% , donc un bruit rajouté très faible.

TD3 – Mixer - Correction

II-1 **On a** $G_C = \frac{2}{\pi} gm(N_S).R_L$ **soit avec** $20\log G_C = 15$ **: G$_C$=5.6** $gm(N_S) = 17.6mA/V$

II-2 **On a** $\left(V_{gs} - V_{tn}\right) = \left(\frac{0,43}{0,64}\right)^2 = 0,45V$ **et** $gm(N_S) = \frac{2.I_p}{\left(V_{gs} - V_{tn}\right)} = 17.6mA/V$ **soit** $I_P = 4mA$

On a aussi la relation : $gm(N_S) = 2.\sqrt{K_n \frac{W}{L} I_p}$ ***soit :*** $\left(\frac{W}{L}\right)_{N_S} = \frac{gm^2}{4K_n.I_P} = 244$

II-3 LZA => Vds(Ns)=Vgs-Vtn=Vod=0.45:

$Id_0(N_{1,2})= Id_0(Ns)/2= Kn(W/L)(Vod(N1,2))^2$;

$Vod(N1,2) = (Vgs(N_{1,2})-Vtn) = (Vg-Vs-Vtn_0-0.2Vs)^2 = 1.5-0.45-0.57-0.2*0.45= 0.39$

=> $W/L=(Ids_0/2)/(Kn*0.39^2)=165$ et W=58

II-4 et II-5

$e_{ns\_ideal}^2=\alpha^2 G_c^2 4KTR_0$
Commuté : $e_{n\_com}^2=R_L^2.I_{nd}^2(Ns)+8KTR_L+ e_{ns\_ideal}^2= e_{na1}^2+ e_{ns\_ideal}^2$
Transitoire : $e_{na2}^2=2.R_L^2.I_{nd}^2(N_{1,2})+8KTR_L$

$$F = 1 + \frac{d_{\%} e_{na1}^2 + \left(1 - d_{\%}\right) e_{na1}^2}{\alpha^2 G_c^2 4KTR_0}$$

$$F = 1 + \frac{d_{\%}\left(R_L^2 I_{nd}^2(N_s) + 8KTR_L\right) + \left(1 - d_{\%}\right)\left(2R_L^2 I_{nd}^2(N_{1,2}) + 8KTR_L\right)}{\alpha^2 G_c^2 4KTR_0}$$

$$F = 1 + \frac{d_{\%}\left(R_L^2 4KT.\gamma.gm(N_s) + 4KTR_{ON} + 8KTR_L\right) + \left(1 - d_{\%}\right)\left(2R_L^2 4KT.\gamma.gm(N_{1,2}) + 8KTR_L\right)}{\alpha^2 G_c^2 4KTR_0}$$

$$F = 1 + \frac{d_{\%}\left(R_L^2.\gamma.gm(N_s) + 2R_L\right) + \left(1 - d_{\%}\right)\left(2R_L^2.\gamma.gm(N_{1,2}) + 2R_L\right)}{\alpha^2 \left(\frac{2.R_L.gm(N_s)}{\pi}\right)^2 R_0}$$

Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr
Grenoble INP
TIMA lab and PHELMA engineering School

soit : avec $\alpha = \dfrac{Z_{in}}{Z_{in}+50}$ **=1 car** Z_{in} **très élevée**

$$F = \frac{2{,}2\pi^2}{600.}\frac{1}{gm(N_s)} + \frac{2\pi^2}{200.R_L}\frac{1}{gm^2(N_s)}$$

soit : avec les valeurs de chaque coefficient : F=10,33 soit $F(dB) \approx 10$

II-6 Source de courant rajoutée en // au nœud A

TD4 – OL – Correction

III-1 A la résonance du réseau LC on à $V = R.i \approx R.G_{m0}.V - R\frac{3}{4}\gamma_3 V^3$ (1)

La valeur de R est déterminée à partir de $Q_L = \frac{R}{L.\omega_0} = 10 \Rightarrow R \approx 250\Omega$

La relation (1) donne donc : $1 = R.G_{m0} - R\frac{3}{4}\gamma_3.V^2$ soit avec V=1V et γ_3 = 2,66 mA/V³ $G_{m0} = 6mA/V$ au max.
La valeur min est donnée par Gm0.R>1 soit Gm0>1/R=4mS au min.

III-2 Voir le cours + correction papier scannée

III-3 Voir le cours

III-4 La puissance du canal ramené à la fréquence f_{FI} s'écrit :
$P_{f_{FI}}(canal) = \mathrm{K}.\,P_{f_{RF}}(canal)\,.\,P_{f_{OL}}(\mathrm{VCO})$ Sylvain
La puissance du brouilleur (situé à f_{RF}+1MHz) ramené à la fréquence f_{FI} s'écrit :
$P_{f_{FI}}(B) = K.P_{f_{OL}+1MHz}(VCO).P(B)$ avec :
$P_{f_{OL}+1MHz}(VCO) = L(1MHz).P_{f_{OL}}(VCO)$
on a donc : $10\log\frac{P_{f_{FI}}(canal)}{P_{f_{FI}}(B)} = 10\log\frac{P(canal)}{L(1MHz).P(B)}$ =40dB

$$10\log\frac{P(canal)}{P(B)} - 10\log L(1MHz) = 40dB = -60dB - L_{dBc}(1MHz)$$
$$\Rightarrow L_{dBc}(1MHz) = -100dB$$

Prof. Sylvain BOURDEL
sylvain.bourdel@grenoble-inp.fr
Grenoble INP
TIMA lab and PHELMA engineering School

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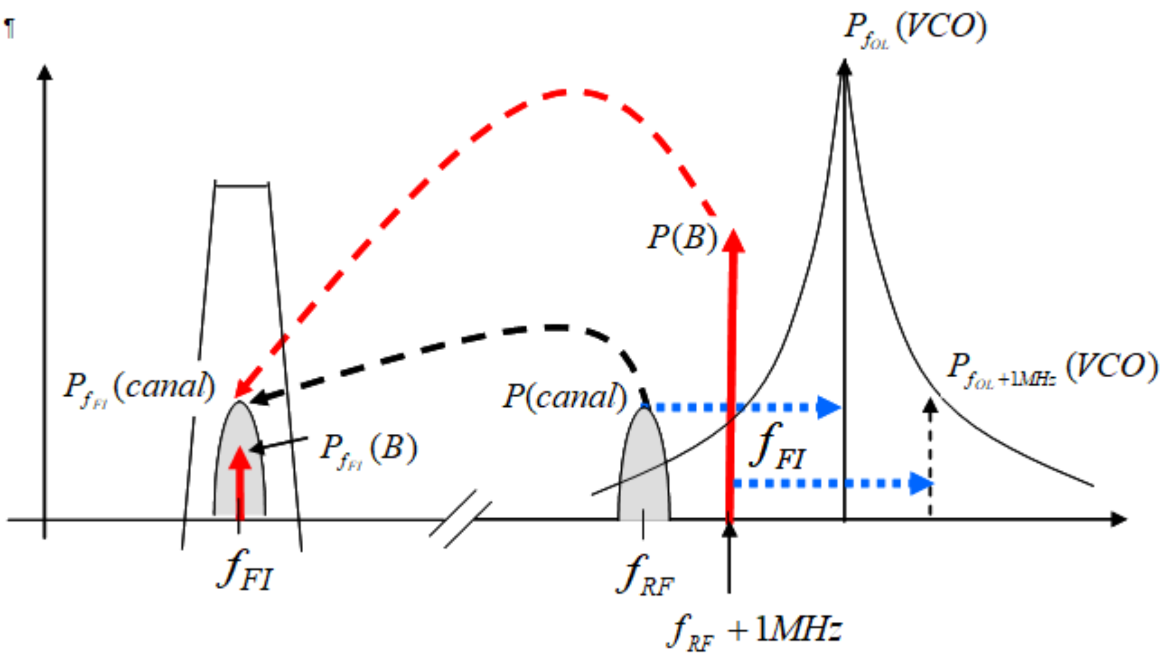
$P_{f_{OL}}(VCO)$
$P(B)$
$P_{f_{FI}}(canal)$
$P_{f_{FI}}(B)$
$P(canal)$
f_{FI}
$P_{f_{OL}+1MHz}(VCO)$
f_{FI}
f_{RF}
$f_{RF}+1MHz$

TD5 – PA - Eléments de correction

II-1 On a : $P_{\max} = \dfrac{Vdd^2}{2.R} = 10^{-3} \times 10^3 = 1 \Rightarrow R = 5{,}4\Omega$

II-2 $P_{\max} = \dfrac{V_s^2}{2 \times 50} \Rightarrow V_s = 10V$

II-3 On a : $\mathrm{I_{ds}}[1] = \dfrac{\mathrm{I_m}}{2.\pi}\left\|\theta - \sin\theta\right\| = \dfrac{\mathrm{I_m}}{2}$ en classe B.

On de plus : $P_{\max} = \dfrac{1}{2}.R.I_{ds}^2[1]$. Soit $I_{ds}[1] = 0{,}6A$ et $I_m = 1{,}2A$

II-4 $\mathrm{I_{ds}}[0] = \dfrac{\mathrm{I_m}}{\pi}\left\|\sin\dfrac{\theta}{2} - \dfrac{\theta}{2}\cos\dfrac{\theta}{2}\right\| = \dfrac{\mathrm{I_m}}{\pi}$ en classe B. Soit : $\mathrm{I_{ds}}[0] = 0{,}38A$

$\eta = \dfrac{\mathrm{I_{ds}}[1]}{2\mathrm{I_{ds}}[0]} = \dfrac{0{,}6}{0{,}76} = 0{,}78$ rendement théorique maximal en classe B

II-5 On a $Q = \dfrac{50}{L_2.\omega_0} = 5$ soit : $L_2 = 0{,}8nH$ et $C_2 = 8pF$

II-6 On a $Q = \sqrt{\dfrac{50}{5}} = 3 = \dfrac{50}{L_1.\omega_0}$ soit : L1=1,32 nH et $C_1 = 4{,}9pF$ soit L=L1//L2=0,5nH

II-7 Il faut que le réseau ramène une résistance $R' = 5{,}4 - 2 = 3{,}4\Omega$. Soit : $Q = \sqrt{\dfrac{50}{3{,}4}} = \dfrac{50}{L_1.\omega_0}$ soit : L'1=1nH et C'1=7,8pF=6,3pF et L'=0,44nH.

Soit V1 : la tension aux bornes de la résistance R'. On a $P_S' = \dfrac{V_1^2}{2.R'} = \dfrac{\left(\dfrac{3{,}4}{5{,}4}V_{dd}\right)^2}{6{,}8} = 0{,}63W$

Une autre solution plus rapide : le courant $I_{ds}[1]$ est le même donc : $\dfrac{P_S'}{P_s} = \dfrac{R'}{R} \Rightarrow P_S' = P_S\dfrac{3{,}4}{5{,}4} = 0{,}63W$.

Le courant Im est inchangé donc PDC aussi. Par conséquent $\eta = \dfrac{0{,}63}{0{,}38.3{,}3} = 0{,}5$

!