

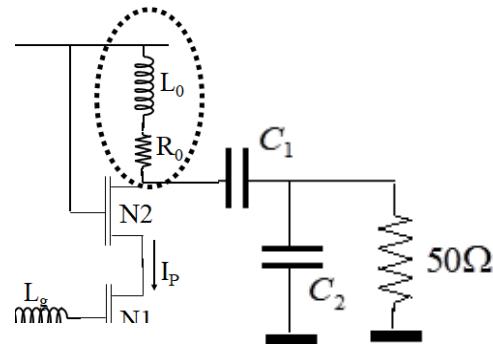
Matching-System

Exercise 1: It is aimed in this exercise to match an LNA to 50Ω . 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 50Ω into another network having only // elements.

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



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

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

$$P_{s_{\text{limax}}}(\text{dBm}) = \frac{2P_{\text{IIP3}}(\text{dBm}) + P_{\text{NF}}(\text{dBm})}{3}$$

1/ Demonstrate that

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

Exercise 4: We consider a RF receiver with the input matched to 50Ω and having a total noise figure $F_{\text{tot}} = 8\text{dB}$ and an $\text{IIP3}_{\text{tot}}(\text{dBm}) = -13\text{dBm}$.

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 $\text{IIP3(V}_{\text{peak}}\text{)}=100\text{mV}$. 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 $G_{\text{LNA}}=15\text{dB}$ a noise figure F_{LNA} and an IIP3_{LNA} . The LNA is followed by an I.Q mixer having a noise figure $F_{\text{IQ-MIX}}=21\text{dB}$ and an $\text{IIP3}_{\text{DEMOD}}(\text{V}_{\text{peak}})=0.6\text{ V}_{\text{peak}}$

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:

$$\text{Att}_{\text{dB}} = 20 \cdot \log \left(\frac{4\pi d}{\lambda} \right)$$

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 $0.6\text{ V}_{\text{peak}}$. Gives the IIP3 of the LNA in dBm .

IEEE 802.15.4 PHY Layer

- 3 bands, 27 channels specified:
 - **2.4 GHz (ISM/Worldwide)** : **16 channels, 250 kbps**
 - 868.3 MHz (ISM/Europe) : 1 channel, 20 kbps
 - 902-928 MHz (ISM/Americas): 10 channels, 40 kbps
- Channels spacing: 5 MHz
- Coding: Direct Sequence Spread Spectrum
- Modulation: Offset-QPSK
- Tx Maximum Output Power > -3 dBm (adj. 30 dB)
- Rx Sensitivity < -85 dBm for a PER of 1%

Low Noise Amplifier LNA

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

NMOS : $K_n = 80 \mu\text{A}/\text{V}^2$; $k_{en} = 0,03 \mu\text{m}/\text{V}$; $V_{t,n} = 0,57\text{V}$; $\gamma = 2$
 $C_{ox} = 5.10^{-3} \text{pF}/\mu\text{m}^2$; $C_{db} = C_{sb} = 1.10^{-3} \text{pF}/\mu\text{m}$ and $L_{min} = 0,35\mu\text{m}$

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

I- Preliminary Study : RF receiver

The figure 1 gives an heterodyne RX architecture used to recover a narrow band signal centered at $f_0 = 2\text{GHz}$. 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 IIP3₁. 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 .

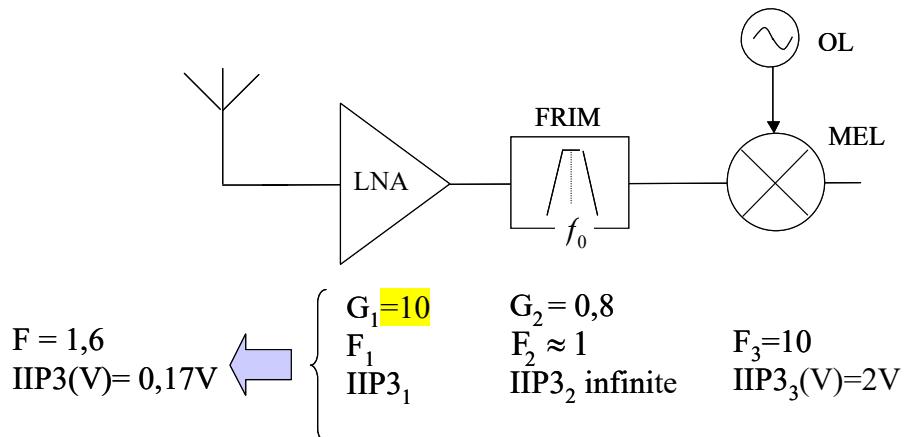


Figure 1

We want to define the requirement of the LNA to have the Rx performances so as : $F_{tot} = 1,6$ and $\text{IIP3}_{tot} = 0,17\text{V}_{peak}$ or $0,17\text{V}_{crête}$ 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 : $\text{IIP3}_1 \approx 0,23\text{Vc}$ et $F_1 \approx 1,46$

I-3 The signal bandwidth is 100 kHz, calculate the sensitivity $P_{se_min}(\text{dBm})$ and the dynamic (in dB) of the Rx if we target an output $\text{SNR}_o = 10\text{dB}$

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

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 $f_0 = 2\text{GHz}$. 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 (R_L).

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 $G_v = \frac{\delta V_s}{\delta V_e}$

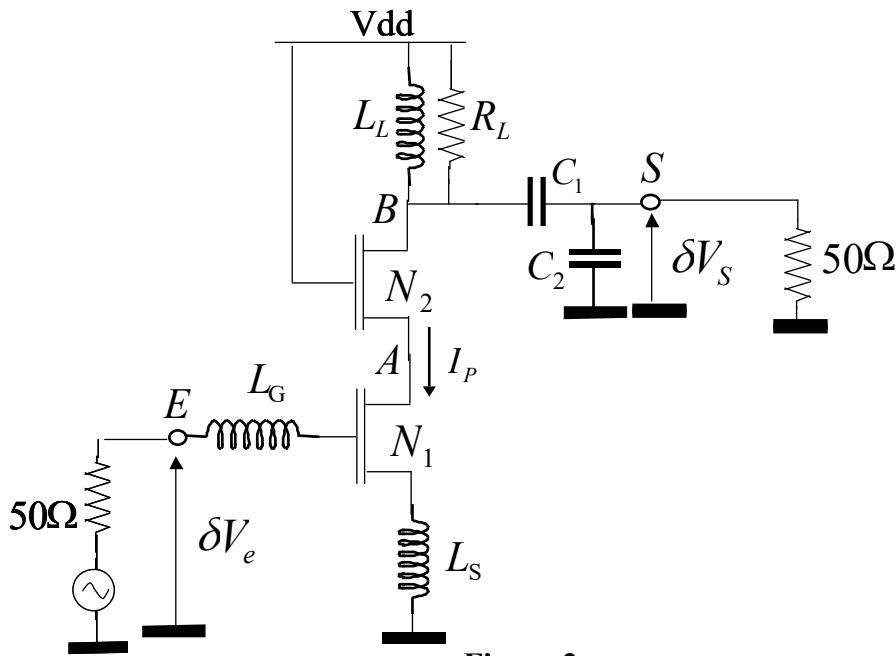


Figure 2

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:

$$IIP3(V) = 0.64\sqrt{(V_{gs} - V_{tn})}\frac{1}{Q_e} \quad (1) \quad \text{and} \quad F = 1 + \frac{2}{50.gm}\frac{1}{Q_e^2} \quad (2) \quad \text{with } Q_e \text{ the } Q \text{ factor of the}$$

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

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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.

RF Mixer

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

NMOS : $K_n = 80 \mu\text{A}/\text{V}^2$; $k_{en} = 0,03 \mu\text{m}/\text{V}$; $V_{t,n} = 0,57\text{V}$; $\gamma = 2$
 $C_{ox} = 5.10^{-3} \text{pF}/\mu\text{m}^2$; $C_{db} = C_{sb} = 1.10^{-3} \text{pF}/\mu\text{m}$ and $L_{min} = 0,35\mu\text{m}$

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

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?

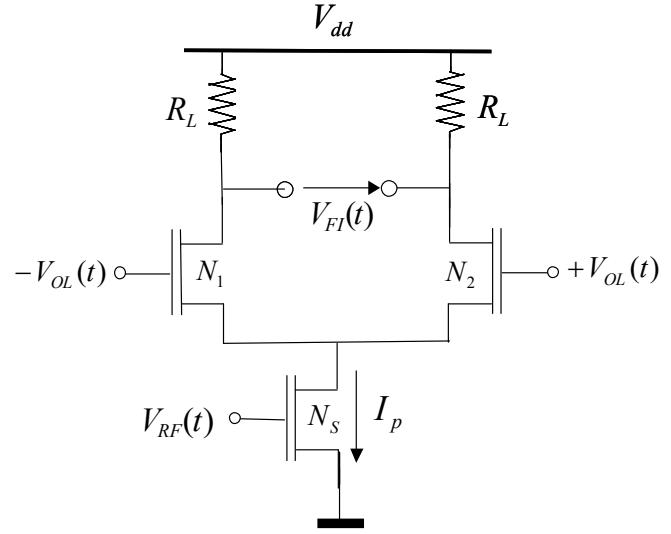
II – Active Mixer Design

We consider the active mixer described below. It is based on a transconductor (NS) a differential switch (N_1 and N_2) and two loads (R_L).

II-1 Knowing the expression of a modulated signal and of the local oscillator signal are:

$v_{rf}(t) = V_r \cos(\omega_{RF}t + \varphi(t))$ and $v_{ol}(t) = B \sin(\omega_{LO}t)$, gives the expression of the conversion gain G_C as a function of $gm(N_s)$ and R_L . With $R_L = 500\Omega$, calculate the $gm(N_s)$ value for $G_C = 15\text{dB}$.

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 $0.43\text{V}_{\text{peak}}$. Then give the ratio $\left(\frac{W}{L}\right)$ of N_s



The mixer is biased so that the N_s 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 $\left(\frac{W}{L}\right)_{N_{1,2}}$ and then give W .

II-4 By considering that the impedance at the N_s 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 .

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II-6 Propose a solution to increase the $gm(N_s)$ 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:

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

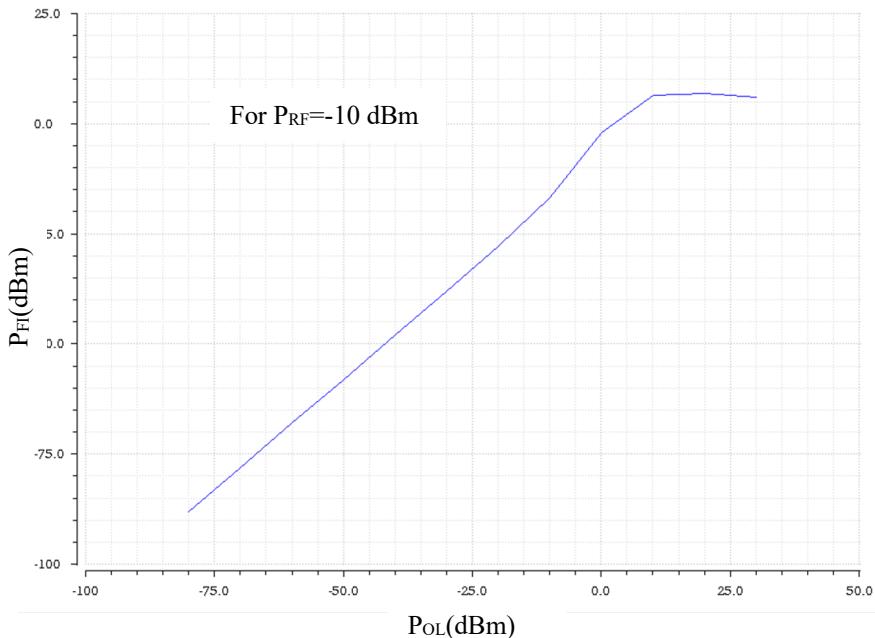
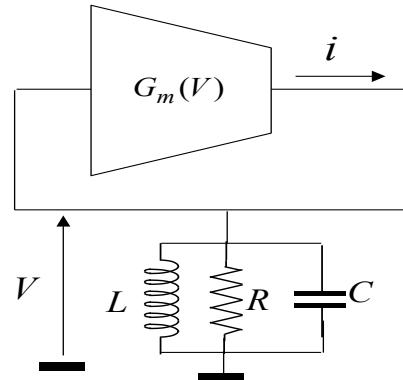
1 Give the oscillation conditions and give the value of G_{m0} in order to have an output magnitude of 1V.

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 $P_{fRF} = -10 \text{ dBm}$.

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 ($f_{Block} = f_{RF} + 1 \text{ MHz}$). Draw a figure representing the several harmonics due to the channel, the blocker and the oscillator.

4 Calculate the phase noise in dBC $L(1 \text{ MHz})$ that should be achieved by the oscillator at the frequency $f_{OL} + 1 \text{ MHz}$ 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.



Power amplifier

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

NMOS : $K_n = 80 \mu\text{A}/\text{V}^2$; $k_{en} = 0,03 \mu\text{m}/\text{V}$; $V_{t,n} = 0,57\text{V}$; $\gamma = 2$
 $C_{ox} = 5 \cdot 10^{-3} \text{ pF}/\mu\text{m}^2$; $C_{db} = C_{sb} = 1 \cdot 10^{-3} \text{ pF}/\mu\text{m}$ and $L_{min} = 0,35 \mu\text{m}$

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

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

$$f_0 = \frac{\omega_0}{2\pi} = 2\text{GHz}.$$

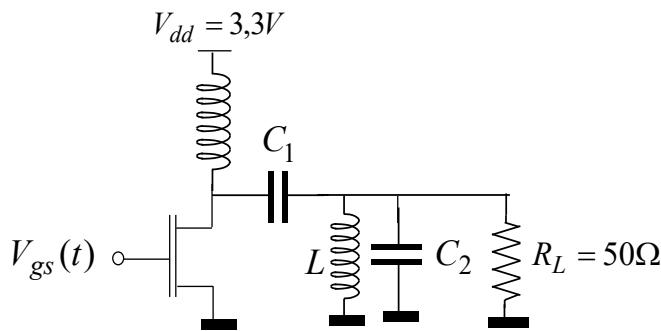


Figure 1

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 $V_{gs}(t)$, $V_{ds}(t)$, $I_{ds}(t)$. 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 $P_{ac_max} = 30\text{dBm}$.

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

II- Class B:

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II-1 Draw $V_{gs}(t)$ and $V_{ds}(t)$, then draw $I_{ds}(t)$ 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 $P_{\max} = 30 \text{ dBm}$.

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.

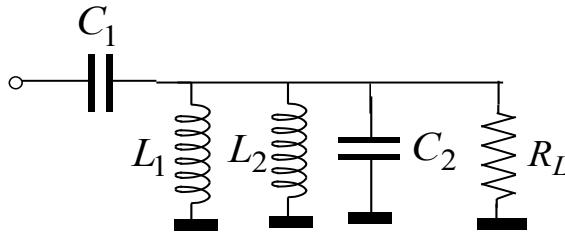


Figure 2a

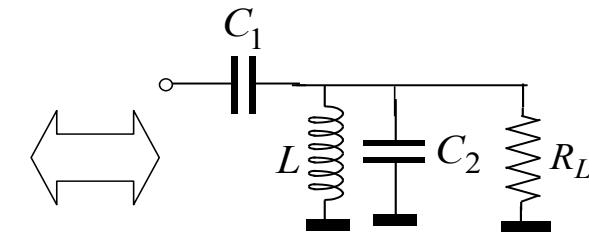


Figure 2b

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.

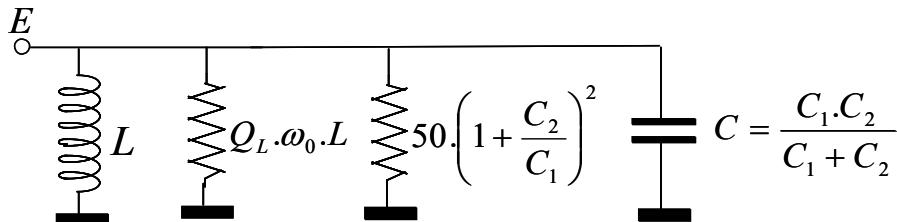
(L, C_1, C_2) network consist in integrated components on silicon. The network losses are represented by a resistance $R_s = 2\Omega$ in series with C_1 .

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 .

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 :

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

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 :

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

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

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Solution ex. 4

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

L-2 $IIP3_{\text{recept.}}(V) = 71 \text{mV}$ soit $\frac{1}{(71 \cdot 10^{-3})^2} = \frac{1}{(100 \cdot 10^{-3})^2} + \frac{10^2}{(IIP3_{\text{mel.}})^2}$ soit : $IIP3_{\text{mel.}} = 1V$

Solution ex. 5 :

1 :

$$P_r = P_e - Att - F$$

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

$$d = 38.8m$$

2 :

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

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

3 :

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

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

4 :

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

$$IIP3_{LNA} = \sqrt{\frac{1}{\frac{1}{IIP3_{tot}^2} - \frac{G_{vLNA}^2}{IIP3_{MIX}^2}}} = 0,05 V_{\text{pic}} = 196 mV_{\text{crête}}$$

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TD2 – LNA - Correction

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

I-2 On a $\frac{1}{(\text{IIP3})^2} = \frac{1}{(\text{IIP3}_1)^2} + \frac{\text{G}_1^2}{(\text{IIP3}_2)^2} + \frac{\text{G}_1^2 \cdot \text{G}_2^2}{(\text{IIP3}_3)^2}$ avec $\text{IIP3}_2 \approx \infty$

Soit : $\frac{1}{(\text{IIP3}_1)^2} = \frac{1}{(\text{IIP3})^2} - \frac{\text{G}_{v,1}^2 \cdot \text{G}_{v,2}^2}{(\text{IIP3}_3)^2} = \frac{1}{(\text{IIP3}_1)^2} = \frac{1}{(0,17)^2} - \frac{64}{(2)^2}$, soit : $\text{IIP3}_1 = 0,23\text{V}$

On a $F = F_1 + \frac{(F_2 - 1)}{\text{G}_1^2} + \frac{(F_3 - 1)}{\text{G}_1^2 \cdot \text{G}_2^2}$ avec $F_2 \approx 1$, soit $F_1 = F - \frac{(F_3 - 1)}{\text{G}_1^2 \cdot \text{G}_2^2} = 1,6 - \frac{9}{64} = 1,46$

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

avec $\text{Pe}_{B(\text{dBm})} = 10 \log(1,6) - 174 + 10 \log(10^5) = -122 \text{ dBm}$ et $\text{IIP3}_{(\text{dBm})} = 10 \log \frac{(0,17)^2}{2,50 \cdot 10^{-3}} \approx -5,4 \text{ dBm}$

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

II- Etude de l'amplificateur faible bruit

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

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

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

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II - 4 On a $G_V = \frac{\delta V_S}{\delta V_E} = \frac{\delta V_S}{\delta V_B} \frac{\delta V_B}{\delta V_E} = Q_e \cdot gm(N_1) \cdot \frac{R_L}{2} \frac{\delta V_S}{\delta V_B} = 175 \cdot Q_e \cdot gm(N_1) = 10$

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

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

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

On a aussi : $gm(N_1) = \frac{2I_p}{(V_{gs} - V_{tn})} = 38 \cdot 10^{-3}$ soit : $I_p \approx 5,7mA$

II - 6 On a aussi : $gm(N_1) = 2\sqrt{K_n \cdot \frac{W}{L} \cdot I_p} = 38 \cdot 10^{-3}$ soit : $\left(\frac{W}{L}\right)_{N1} = 1150 = 792$

soit avec $L = 0,35\mu m$: $W \approx 400\mu m \approx 277\mu m$

On a $C_{GS}(N_1) \approx \frac{2}{3} C_{ox} \cdot W \cdot L \approx 0,47pF \approx 0,32pF$. Le coefficient Q_e à la résonance en entrée

satisfait la relation $Q_e = \frac{1}{(C_0 + C_{GS}) \cdot 2\pi \cdot 50 \cdot f_0}$, soit : $C_0 + C_{GS} \approx 1,06pF$ et donc :

$C_0 \approx 0,6pF \approx 0,74pF$

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

La condition d'adaptation en entrée s'écrit : $50 = L_s \cdot 2\pi \cdot f_T$ soit $L_s \approx 1,4nH$

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

Soit $(L_G + L_s) \approx 6nH$, 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 $(C_1//C_2) \approx 1,26pF$

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

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

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

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

$$\omega_A = \frac{gm(N_2)}{C(A)} = \frac{gm(N_2)}{C_{GS}(N_2)} = \frac{gm(N_1)}{C_{GS}(N_1)} = \frac{38 \cdot 10^{-3}}{0,47 \cdot 10^{-12}} = 81 \text{ rd/s}$$
 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 $(V_{gs} - V_{tn}) = \left(\frac{0,43}{0,64} \right)^2 = 0,45V$ et $gm(N_s) = \frac{2I_p}{(V_{gs} - V_{tn})} = 17.6mA/V$ soit $I_p = 4mA$

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

II-3 LZA => $V_{ds}(Ns) = V_{gs} - V_{tn} = V_{od} = 0.45$:

$$Id_0(N_{1,2}) = Id_0(N_s)/2 = Kn(W/L)(V_{od}(N_{1,2}))^2 ;$$

$$V_{od}(N_{1,2}) = (V_{gs}(N_{1,2}) - V_{tn}) = (V_g - V_s - V_{tn} - 0.2V_s)^2 = 1.5 - 0.45 - 0.57 - 0.2 * 0.45 = 0.39$$

$$\Rightarrow W/L = (Id_0/2)/(Kn * 0.39^2) = 165 \text{ et } W = 58$$

II-4 et II-5

$$e_{ns_ideal}^2 = \alpha^2 G_c^2 4KTR_0$$

$$Commuté : e_{n_com}^2 = R_L^2 I_{nd}^2(N_s) + 8KTR_L + e_{ns_ideal}^2 = e_{na1}^2 + e_{ns_ideal}^2$$

$$Transitoire : e_{na2}^2 = 2.R_L^2 I_{nd}^2(N_{1,2}) + 8KTR_L$$

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

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

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

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

soit : avec $\alpha = \frac{Z_{in}}{Z_{in} + 50} = 1$ 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 $\gamma_3 = 2,66$

mA/V^3 $G_{m0} = 6mA/V$ 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 :

$$P_{f_{FI}}(\text{canal}) = K.P_{f_{RF}}(\text{canal}).P_{f_{OL}}(\text{VCO}) \quad \text{Sylvain}$$

La puissance du brouilleur (situé à $f_{RF} + 1MHz$) ramené à la fréquence f_{FI} s'écrit :

$$P_{f_{FI}}(B) = K.P_{f_{OL}+1MHz}(\text{VCO}).P(B) \text{ avec :}$$

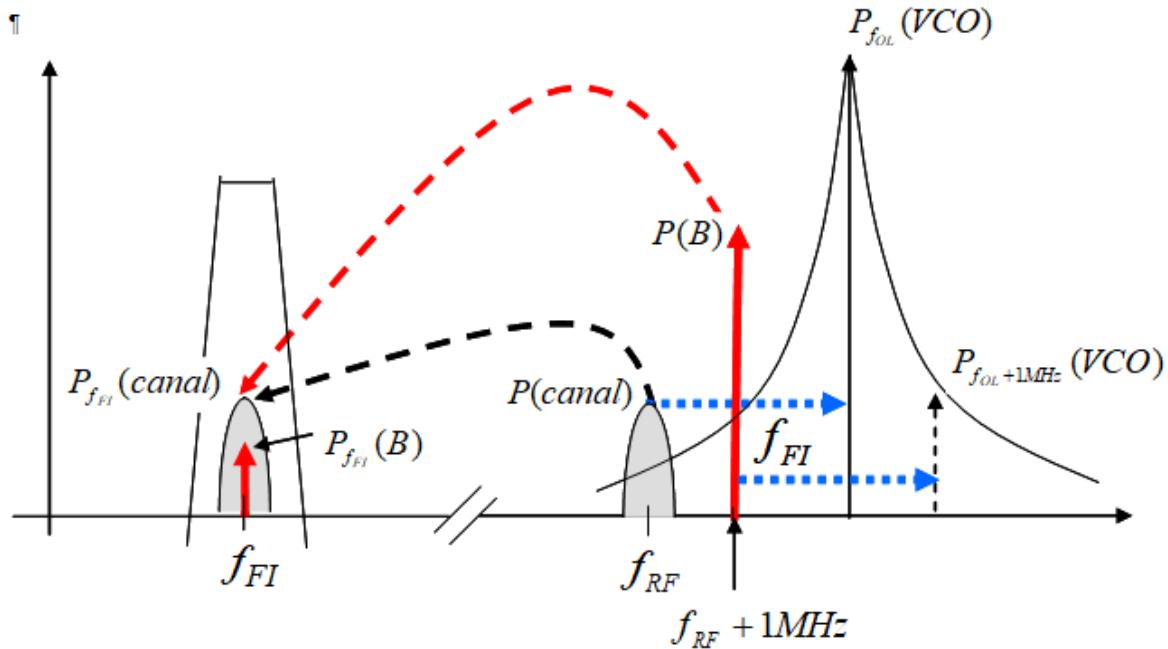
$$P_{f_{OL}+1MHz}(\text{VCO}) = L(1MHz).P_{f_{OL}}(\text{VCO})$$

$$\text{on a donc : } 10\log \frac{P_{f_{FI}}(\text{canal})}{P_{f_{FI}}(B)} = 10\log \frac{P(\text{canal})}{L(1MHz).P(B)} = 40\text{dB}$$

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

$$\Rightarrow L_{dBc}(1MHz) = -100\text{dB}$$

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TD5 – PA - Eléments de correction

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

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

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

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

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

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

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

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

II-7 Il faut que le réseau ramène une résistance $R' = 5,4 - 2 = 3,4\Omega$. Soit : $Q = \sqrt{\frac{50}{3,4}} = \frac{50}{L_1 \cdot \omega_0}$ soit :

$L'_1 = 1\text{nH}$ et $C'_1 = 7,8\text{pF} = 6,3\text{pF}$ et $L' = 0,44\text{nH}$.

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

Une autre solution plus rapide : le courant $I_{ds}[1]$ est le même donc :

$$\frac{P'_S}{P_s} = \frac{R'}{R} \Rightarrow P'_S = P_s \frac{3,4}{5,4} = 0,63W.$$

Le courant I_m est inchangé donc P_{DC} aussi. Par conséquent $\eta = \frac{0,63}{0,38 \cdot 3,3} = 0,5$

!