



Télécom Paris  
Département Communications et Électronique

TELECOM201

PRACTICAL WORKS

Hardware design and sizing of the uplink modules of a  
Cloud Radio Access Network

Chadi JABBOUR – Germain PHAM

# 1 Introduction

## 1.1 About

In this lab, we will focus on the design of the uplink modules of the Cloud Radio Access Network that will be studied in details during the second semester project. This lab will be as a consequence a preparation for the work that you will have to do in the AMS&RF section of the project (TELECOM205). So feel free to play with the codes to test your ideas.

## 1.2 Reminder of the context

We consider a Cloud Radio Access Network (C-RAN) composed of a set of nodes that may transmit information wirelessly to a basestation. This basestation is connected via an optical fiber to a remote server which corresponds to the final destination as described in Fig. 1.

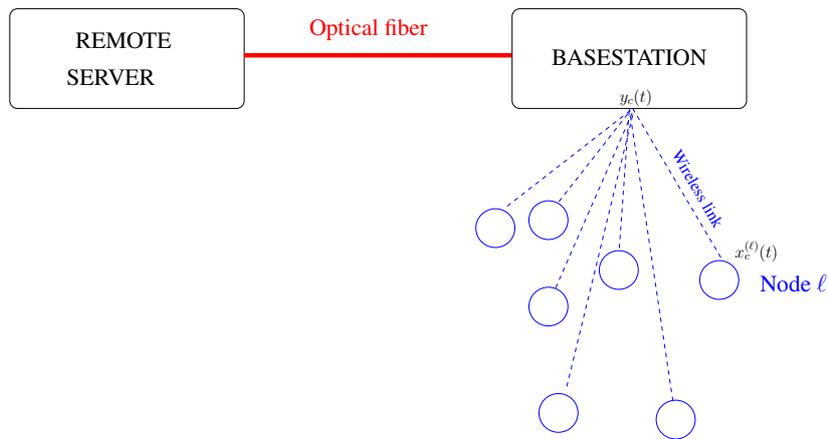


FIGURE 1 – C-RAN system model

As mentioned earlier, in this lab, we will just focus on the hardware implementation of the transmitter (TX) of the node and the receiver (RX) of the basestation. For the sake of simplicity, we will consider only one node in the framework of this lab. The multi-access aspects will be studied during the project. The useful signal has an RF bandwidth of 20 MHz and is centered at 2.4 GHz. The TX and RX are matched with a  $50 \Omega$  resistance. The analysis will be carried out for a temperature of 290 K.

## 1.3 Organization of the lab

The lab will be divided into two sections, a first one focusing on the node TX and a second one focusing on the basestation RX.

## 1.4 Simulation environment

In order to properly execute the lab scripts, you should start Matlab/Octave in the `scripts` directory. In the contrary case, handy graphical enhancements will be disabled.

# 2 Transmitter (TX)

The block diagram of the TX system is shown in Fig. 2. It is an Inphase/Quadrature (IQ) Homodyne architecture. The I-path and Q-path are both composed of a Digital to Analogue Converter (DAC) followed by a low pass filter. The I-path and the Q-path will

process respectively the real and the imaginary part of the baseband signal and therefore, each path will process a (unilateral) bandwidth of 10 MHz. The outputs of the two paths are upmixed to the central frequency of 2.4 GHz and combined (i.e. summed). After this reconstruction, the RF bandwidth will have a (bilateral) RF bandwidth of 20 MHz. A power amplifier (PA) follows the mixer/combiner to amplify the signal to the desired output power of 20 dBm.

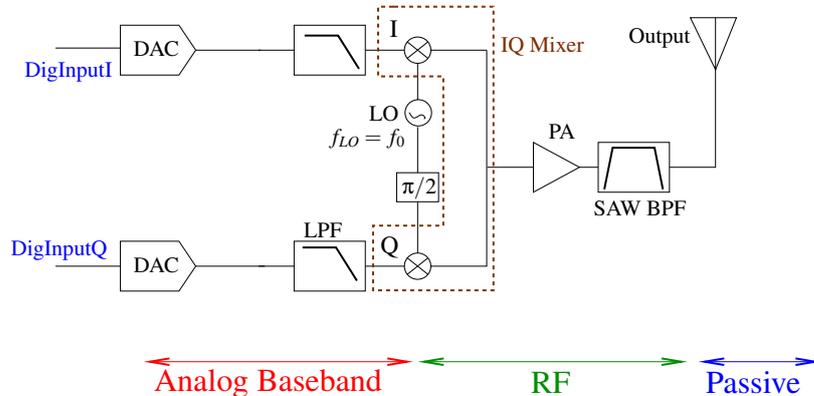


FIGURE 2 – TX block diagram

## 2.1 DAC

The baseband DAC function is to convert the discrete time digital input issued from the digital signal processor to a continuous time wave. Since Octave is a discrete time simulator, emulating a continuous time wave will be realized by using a very high speed clock `continuousTimeSamplingRate`. As mentioned earlier, the useful bandwidth at the DAC input is 10 MHz. The DAC will be clocked at a frequency of 30 MHz to allow some room for the filtering.

To implement the conversion from discrete time to continuous time, a large diversity of approaches exist. In the framework of the lab, we will focus on two :

- The zero padding : The continuous time output signal is equal to the discrete time input signal at instants  $n \times T_s^{DAC}$ , where  $T_s^{DAC}$  is the sampling period of the DAC and is equal to zero otherwise.
- The zero-order hold : The continuous time output signal is set to the discrete time input signal at instant  $n \times T_s^{DAC}$  and is kept constant till the instant  $(n + 1)T_s^{DAC}$  on which the next sample arrives.

**Question 2.1** Execute the file `DAC_TB.m`. Analyze the spectrums of the zero padding DAC and the zero-order hold DAC.

**Question 2.2** Give an analytical explanation for the zero-order hold DAC filtering behavior.

**Question 2.3** Change the number of bits of the DAC  $N_{bits}^{DAC}$  to 4, 8, 12 and 16. Analyze the impact of this change on the performance.

For the remaining of the lab, we will keep  $N_{bits}^{DAC} = 12$  and a zero-order hold DAC.

## 2.2 Analogue baseband filter

To attenuate the signal replicas, a low pass filter is needed. Based on a budgeting between the different errors and the required performance, we decide to set the inband attenuation  $A_{max}$  to 1 dB and to attenuate all the replicas ( $f > 20$  MHz) by at least 30 dB.

**Question 2.4** Draw by hand on your report the filter attenuation template on which should be specified  $f_1$ ,  $f_2$ ,  $A_{min}$  and  $A_{max}$ .

**Question 2.5** Calculate the order needed for a filter with a Butterworth approximation to satisfy these requirements.

**Question 2.6** Load the script `ButterworthFilterSpecifications.m` and set  $f_1$ ,  $f_2$ ,  $A_{min}$  and  $A_{max}$  to the values determined in the previous questions. Compare the calculated order to the script results.

**Question 2.7** Load the script `TX_BasebandChain.m`. Set `TXBB_Fil_Fcut` and `TXBB_Fil_Order` to the values determined in the previous question. For this simulation, the worst case scenario will be considered, i.e. a sine wave at 10 MHz. Pick up the power difference between the input signal and the first replica at 20 MHz and make sure that it is higher than the targeted value.

### 2.3 Up-conversion Mixer

An IQ mixer achieves the signal reconstruction and the frequency translation by multiplying the I and Q baseband signals by respectively an inphase local oscillator signal and a shifted version by  $\pi/2$ . To be more precise, the mixer is modeled by :

- a multiplication by  $\cos(2\pi f_{lo}t)$  for the *in-phase* signal;
- a multiplication by  $\cos(2\pi f_{lo}t + \frac{\pi}{2})$  for the *in-quadrature* signal;

Moreover, we consider a baseband (complex) sine wave :

$$x(t) = Ae^{j2\pi f_1 t}. \quad (1)$$

We separate the real and imaginary part of  $x(t)$  to obtain the I and Q channels as follows :

$$\begin{cases} x^I(t) &= \Re\{x(t)\} = A \cos(2\pi f_1 t) \\ x^Q(t) &= \Im\{x(t)\} = A \sin(2\pi f_1 t) \end{cases} \quad (2)$$

**Question 2.8** Calculate the expression of the output of combiner-mixer and explain its operation.

**Question 2.9** Load the script `TX_Chain_woPA.m` and set `TXBB_Fil_Fcut` and `TXBB_Fil_Order` to the values determined in the previous subsection. Analyze the output and compare to the result of the previous question.

**Question 2.10** Modify the amplitude of the Q-path input signal  $A_Q$  to 0.99 and 0.9. Analyze the impact of this modification on the mixer output spectrum. Give a theoretical explanation to the observed results based on the analysis performed in Q.2.8 and considering :

$$\begin{cases} x^I(t) &= A_I \cos(2\pi f_1 t) \\ x^Q(t) &= A_Q \sin(2\pi f_1 t) \end{cases} \quad (3)$$

### 2.4 Power amplifier

The power amplifier is the last block of the TX chain before the RF passive components<sup>1</sup>. Its function is to amplify the signal to the value set by the communication protocol, 20 dBm in our case. The main constraint in a PA is the non linearity. As a matter of fact, since it is processing a signal with a high power, the impact of non linearity becomes very critical. In practice, there are two metrics that are widely used to quantify PA non linearity : the third

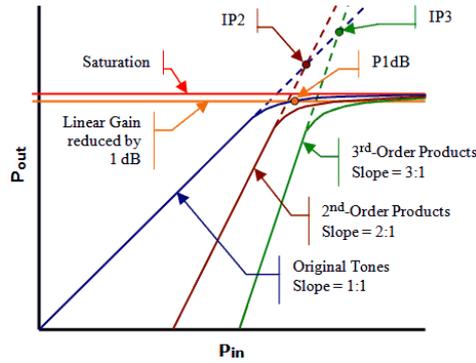


FIGURE 3 – IIP3 and P1dB

order intercept point (IIP3) and the 1 dB compression point (P1dB). Fig. 3 explains these two metrics.

PA non linearity causes two unwanted effects : a modification of the useful signal and also a signal leakage to adjacent channels. There are several metrics to measure the non linearity impact such as the Adjacent Channel Power Ratio (ACPR) or the Error Vector Magnitude (EVM).

In this lab, we will use a two-tone simulation and measure the IM3 which is the ratio between the power of one input tone and the highest intermodulation term which arises at  $2f_1 - f_2$  and  $2f_2 - f_1$ , where  $f_1$  and  $f_2$  are the frequencies of the two input tones. We require in our application that our IM3 should be higher than 40 dB.

The available PAs are listed in Table 1. They are modeled by the following polynomial model :

$$y(t) = \alpha_1 x(t) + \alpha_3 x(t)^3 \quad (\alpha_3 < 0 \text{ in practice}) \quad (4)$$

The power efficiency of a PA can be approximated by the ratio between the transmitted power and the consumed power.

	IIP3	P1dB	Gain	Efficiency	Power Consumption
PA1	35 dBm	24.6 dBm	14 dB	20 %	0.5 W
PA2	30 dBm	19.6 dBm	14 dB	30 %	0.33 W
PA3	20 dBm	9.6 dBm	14 dB	40 %	0.25 W

TABLE 1 – PAs performances

**Question 2.11** Load the script *TX\_Chain.m*, set *TXBB\_Fil\_Fcut* and *TXBB\_Fil\_Order* to the values determined earlier. Test the TX chain with the three PAs by changing *PA\_model*. Pick up the value of the IM3 in each case.

**Question 2.12** Determine which PA(s) could be used for our application. Among this list, pick the one that you would like to use and explain why.

### 3 Receiver (RX)

Now we focus on the basestation RX part. For sake of simplicity, we will assume that the channel only causes an attenuation of the signal due to the distance. In our scenario, the attenuation variation from the PA output to the Low Noise Amplifier input varies from 50 dB

1. RF passive filter, Duplexer and/or antenna

to 110 dB. Therefore the receiver must cope with the power variation and the noise level at its the antenna output.

The diagram of the considered receiver is shown in Fig. 4. It is also an Inphase/Quadrature (IQ) Homodyne architecture. The signal is first filtered by an RF passive filter that will filter far out-of-band interferers. Then the signal, is downconverted by an IQ mixer where the signal is separated into two paths. The I-path and Q-path are both composed of a low pass Anti-Aliasing Filter and a Analog to Digital Converter (ADC). At the ADC output, the signal should have an SNR higher than 2 dB.

Since some of these blocks have already been studied in the previous section or in previous labs, they will be touched on superficially here in order to focus on the other blocks of the chain.

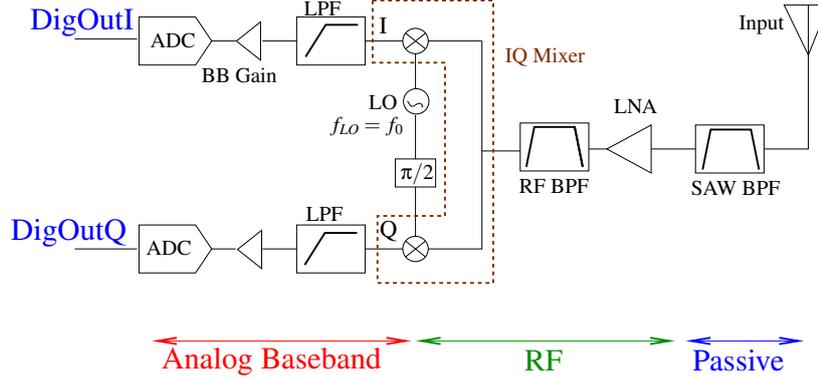


FIGURE 4 – RX block diagram

### 3.1 General Specifications

**Question 3.1** Calculate the noise power integrated in the 20 MHz useful bandwidth at the receiver input.

**Question 3.2** Calculate the power range of the received signal and determine the range of the SNR at the receiver input.

**Question 3.3** Determine the value of the highest NF allowed for our receiver.

The ADC full scale is set  $\pm 1$  V.

**Question 3.4** Calculate the receiver gain that allows to profit optimally from the ADC Dynamic Range (DR) without saturating it. You can assume that the signal is a sine wave for this calculation.

### 3.2 Low noise amplifier

After RF filtering, the signal is amplified by the LNA. We set the LNA contribution to the overall noise to 15%.

**Question 3.5** Determine the SNR at the LNA output and deduce its NF. You can use the following results demonstrated in the appendix.

$$(SNR_{LNA}^{lin})^{-1} = (SNR_{in}^{lin})^{-1} + 0.15(SNR_{out}^{lin})^{-1} \quad (5)$$

The baseband gain is set to 20 dB.

**Question 3.6** Calculate the maximum allowed gain for the LNA.

We will set the the LNA gain 5 dB lower than this value to avoid saturation due to interferers.

**Question 3.7** Load the script `LNA_TB.m`, set the `LNA_NF` and `LNA_Gain` to the calculated values. Compare the obtained SNR and NF to the calculated ones.

### 3.3 RX baseband

The channel filtering is realized by an analog anti-aliasing filter. The design approach of this filter is similar to the TX baseband filter. The design of those filters is skipped for focusing the ADC. The anti-aliasing filter will be a 6<sup>th</sup> order Butterworth with a cutoff frequency of 11.1 MHz. The filter is followed by a baseband amplifier whose gain is set to 20 dB. The digitization of the signal is performed by an ADC whose sampling frequency is equal to  $F_s^{ADC} = 30$  MHz. This ADC (similarly to DAC) induces perturbations due to quantization over  $N_{\text{bits,ADC}}$ .

**Question 3.8** Write the expression that links the signal to quantization noise ratio (SQNR) of an ADC to its input amplitude, its full scale and its number of bits.

10% of the overall noise budget is allocated to the ADC. This requires that the ADC SQNR should be higher than 12 dB for all the input values.

**Question 3.9** Calculate the number of bits needed for the ADC to cover all the dynamic range of the input signal. To do so, you have to identify the minimum signal amplitude at the ADC input.

**Question 3.10** Load the script `RX_TB.m`, set the LNA parameters and the ADC parameters to the calculated values. Pick up the values of the SNR for the maximum and minimum input powers and make sure that we meet the desired specifications.

## 4 Whole chain

We will now study briefly a complete communication using the transmitter and receiver that we have designed. We will transmit an audio signal. The communication will be performed similarly to an amplitude modulation communication, i.e., the audio signal will be directly applied to the DAC and is played at the output of the ADC. The purpose of this part is to study the impact of the channel performance on the communication quality. The LNA IIP3 is set to  $-45$  dBm.

Load first the script `Play_OriginalAudio.m` to listen the original sound. Then, load the script `TXX.m` which simulates a complete communication.

**Question 4.1** Simulate the communication for channel attenuations of 50, 80 and 110 dB. Compare qualitatively the results and analyze the observed phenomenas.

**Question 4.2** What could be done to improve the performance of the system ?

## Appendix : Noise budgeting calculation

First, it is worth mentioning that when we talk about the contribution of a given block to the noise budget of a system (Receiver or transmitter), the contribution takes into account the gains or attenuations that were applied to the noise of this given block. To compare the contributions of the different blocks of a chain, two common ways are widely used : calculating

at the system output a.k.a. output referred calculation or at the system input a.k.a. input referred calculation. The two approaches are equivalent and give therefore the exact same results. In the following, we will do the calculation at the system output.

$$SNR_{in}^{lin} = \frac{P_{in}}{N_{Ant}}, \quad (6)$$

where  $SNR_{in}^{lin}$  is the SNR in linear at the receiver input.

$$SNR_{LNA}^{lin} = \frac{G_{p-LNA} \cdot P_{in}}{G_{p-LNA} \cdot N_{Ant} + N_{LNA}}, \quad (7)$$

where  $SNR_{LNA}^{lin}$  is the SNR in linear at the LNA output,  $G_{p-LNA}$  is the power gain of the LNA and  $N_{LNA}$  is the noise added by the LNA. The expression of this noise, as discussed in the introduction course, is given by :

$$N_{LNA} = (F_{LNA} - 1) \cdot G_{p-LNA} \cdot N_{Ant} \quad (8)$$

$$SNR_{out}^{lin} = \frac{G_{p-BB} \cdot G_{p-LNA} \cdot P_{in}}{G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + G_{p-BB} \cdot N_{LNA} + N_{other}}, \quad (9)$$

where  $SNR_{out}^{lin}$  is the SNR in linear at the receiver output,  $G_{p-BB}$  is the power gain of the rest of chain and  $N_{other}$  is the noise added by the remaining components of the chain (baseband amplifier, ADC, mixer ...).

We know that the contribution of the LNA to the overall noise budget is 15 % which could be expressed by the following expression :

$$\frac{G_{p-BB} \cdot N_{LNA}}{G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + G_{p-BB} \cdot N_{LNA} + N_{other}} = 0.15 \quad (10)$$

$$\implies N_{LNA} = \frac{\overbrace{0.15 (G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + G_{p-BB} \cdot N_{LNA} + N_{other})}^{\text{Overall Noise at the receiver output}}}{G_{p-BB}} \quad (11)$$

By replacing  $N_{LNA}$  in  $SNR_{LNA}^{lin}$  by its value computed in the previous equation, we obtain :

$$SNR_{LNA}^{lin} = \frac{G_{p-LNA} \cdot P_{in}}{G_{p-LNA} \cdot N_{Ant} + \frac{0.15(G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + G_{p-BB} \cdot N_{LNA} + N_{other})}{G_{p-BB}}} \quad (12)$$

$$SNR_{LNA}^{lin} = \frac{G_{p-BB} \cdot G_{p-LNA} \cdot P_{in}}{G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + 0.15(G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + G_{p-BB} \cdot N_{LNA} + N_{other})} \quad (13)$$

By re-arranging the terms,  $SNR_{LNA}^{lin}$  could be calculated as follows :

$$(SNR_{LNA}^{lin})^{-1} = \frac{G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant}}{G_{p-BB} \cdot G_{p-LNA} \cdot P_{in}} + 0.15 \frac{G_{p-BB} \cdot G_{p-LNA} \cdot N_{Ant} + G_{p-BB} \cdot N_{LNA} + N_{other}}{G_{p-BB} \cdot G_{p-LNA} \cdot P_{in}} \quad (14)$$

$$= (SNR_{in}^{lin})^{-1} + 0.15(SNR_{out}^{lin})^{-1} \quad (15)$$