INSA INSTITUT NATIONAL DES SCIENCES APPLIQUÉES TOULOUSE

Communication – Protocols for connected objects

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PRACTICAL WORK REPORT

Introduction to Software-Defined Radio (SDR)

ANALOG DEMODULATION OF SIGNALS USING GNURadio

INNOVATIVE SMART SYSTEMS

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Introduction

During this project, we focused on the reception of real communication signals (e.g. broadcasting, aeronautical communications, etc.). A specific part of the wide frequency range of radio spectrum is allocated to each signal with regards to the targeted applications and constraints.

The complete description of the allocations of radiofrequency bands are available at https://en.wikipedia.org/wiki/Electromagnetic_spectrum/. We have seen that a single receiver can record almost all these signals and then process them numerically using GNURadio, in order to recover all the information.

Firstly, in this report, we demonstrate that, in the case of narrowband signals, the IQ transceivers (In-phase/Quadrature) allow transmissions with frequency transportation without altering data. Secondly, we report our work on FM broadcasting, where our aim was to demodulate it in real time. Thirdly, the VOLMET AM broadcasting have been treated, again in real time.

Ultimately, a short project around the OSI physical layer of an Internet-of-Things (IoT) technology have been achieved in autonomy.

1 Presentation of the data acquisition device: In-phase/Quadrature Software-Defined Radio transceiver

1.1 Context: about Software-Defined Radio

The "Software-Defined Radio" (SRD) refers to a type of radiofrequency transceiver in which most of the processing is done digitally, whether it is the transmission or the reception of data. For a reception, the received signal is digitized through an ADC, then processed (for instance: filtered, decimated, demodulated, decoded, etc.). For the transmission, the data to transmit are processed (for instance: coded, modulated, etc.), then converted through a DAC and sent to the RF front-end.

Thus, thanks to a sufficiently generic hardware (wide frequency range, high sampling frequency, etc.), it is possible to easily develop many applications by only changing the software part of the transceiver (e.g. cellular telephony, digital television, etc.). Moreover, many applications can operate at the same time on the same hardware (e.g. simultaneous received signals—with different technologies, frequencies, modulations and/or encodings—can be processed concurrently, as well as other signals that can be transmitted).

In addition, it is possible to update the software without stopping the hardware.

This modular approach is not in consensus with the current trend of creating specific electronic circuits with a view to maximum integration. For instance, in your smartphone there are at least six independent transmission channels: GSM (2G), UMTS (3G), LTE (4G), Wi-Fi (IEEE 802.11), Bluetooth (802.15.1), NFC (ISO/IEC 14443), etc. Each of these technologies has a specific integrated circuit and antenna. Thus, at each evolution of the technology, the hardware must be changed. Soon, the software-defined radio will mutualize these devices in order to save space and energy and allow reusing and evolving of the system.

1.2 Universal Software Radio Peripheral

Materials

During our labs, we used a National Instruments USRP-2900 software defined radio transceiver, which is connected to a computer via a Gigabit Ethernet connection (Fig. 1). The receiver is composed of two stages (Fig. 2): the first allows the transposition of the signals around the zero frequency by heterodyning and the decomposition in in-phase and quadrature signals—like the carrier frequency has to be known, this receiver is said coherent—and the second carries out the analog to digital conversion (CAN) (sampling at a T_e period, then quantization scalar uniform over 12 bits).

Figure 1: National Instruments USRP-2920 transceiver

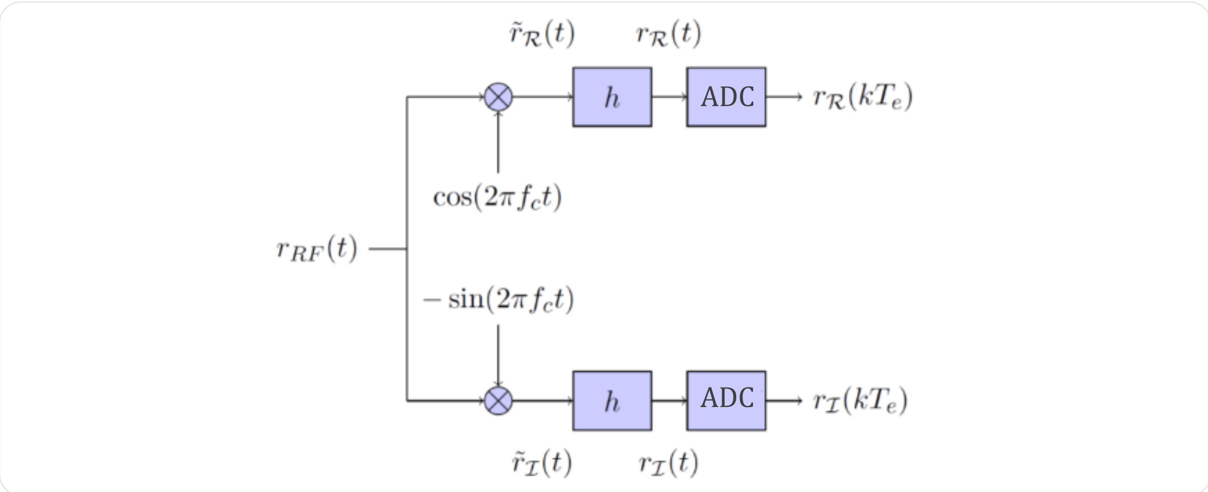


Figure 2: Block diagram of the receiver

To understand the structure of the receiver, we can note a communication signal s_{RF} transmitted around a carrier frequency f_0, with $A(t) \in \mathbb{R}$ the envelope and $\varphi(t) \in \mathbb{R}$ the phase, as:

$$s_{RF}(t) = A(t)\cos\big(2\pi f_0 t + \varphi(t)\big) \tag{1}$$

Through the following questions, we explain how this architecture can achieve the goal of the USRP, and what are the parameters required, to fill this architecture: which f_c frequency? which h filter? and what T_e sampling period for the Analog Digital Converter?

Question 1

According to Fig. 2, considering that the received signal is like the transmitted one ($r_{RF}(t) = s_{RF}(t)$), using (2) and trigonometric formula below), express the signal $\tilde{r}_R(t)$ and $\tilde{r}_I(t)$, function of $s_R(t)$, $s_I(t)$, f_0 and , f_c.

Considering that the received signal is like the transmitted one $\left(r_{RF}(t) = s_{RF}(t)\right)$ and that the bandwidth is a least twice smaller than the frequency we focus on $\left(\frac{B}{2} < f_0\right)$, we have:

$$s_{RF}(t) = A(t)\cos\left(2\pi f_0 t + \varphi(t)\right) \quad (1)$$

$$s_{RF}(t) = A(t)\cos(2\pi f_0 t)\cos\left(\varphi(t)\right) - A(t)\sin(2\pi f_0 t)\sin\left(\varphi(t)\right)\left(\varphi(t)\right)$$

$$\begin{cases} s_R = A(t)\cos(\varphi) \\ s_I = A(t)\sin(\varphi) \end{cases}$$

$$s_{RF}(t) = r_{RF}(t) = s_R(t)\cos(2\pi f_0 t) - s_I(t)\sin(2\pi f_0 t) \quad (2)$$

Thus, we can express the signal $\tilde{r}_R(t)$ and $\tilde{r}_I(t)$, function of $s_R(t)$, $s_I(t)$, f_0 and , f_c:

$$\tilde{r}_R(t) = r_{RF}(t)\cos(2\pi f_c t)$$
$$\tilde{r}_R(t) = s_R(t)\cos(2\pi f_0 t\,)\cos(2\pi f_c t) - s_I\sin(2\pi f_0 t)\cos(2\pi f_c t)$$

$$\tilde{r}_R(t) = \frac{s_R(t)}{2}\left(\cos\left(2\pi t(f_0 - f_c)\right) + \cos\left(2\pi t(f_0 + f_c)\right)\right) - \frac{s_I(t)}{2}\left(\sin\left(2\pi t(f_0 - f_c)\right) + \sin\left(2\pi t(f_0 + f_c)\right)\right)$$

$$\tilde{r}_I(t) = r_{RF}(t)\sin(2\pi f_c t)$$
$$\tilde{r}_I(t) = s_R(t)\cos(2\pi f_0 t\,)\sin(2\pi f_c t) - s_I\sin(2\pi f_0 t)\sin(2\pi f_c t)$$

$$\tilde{r}_I(t) = \frac{s_R(t)}{2}\left(\cos\left(2\pi t(f_0 - f_c)\right) - \cos\left(2\pi t(f_0 + f_c)\right)\right) - \frac{s_I(t)}{2}\left(\sin\left(2\pi t(f_0 - f_c)\right) - \sin\left(2\pi t(f_0 + f_c)\right)\right)$$

Question 2

If we take ($f_c = f_0$)—translation in the baseband by heterodyning—what should be the characteristics of the h filters to get $r_R(t) = s_R(t)$ and $r_I(t) = s_I(t)$? You have to make a representation of $\tilde{R}_R(f)$ and $\tilde{R}_I(f)$, after using the Fourier transforms. If you want to work in the time domain, a hypothesis must be clearly stated and checked.

The signal has a frequency of f_0, an amplitude of A_0 and a bandwidth of B. Therefore, its spectrum looks like this (Fig. 3):

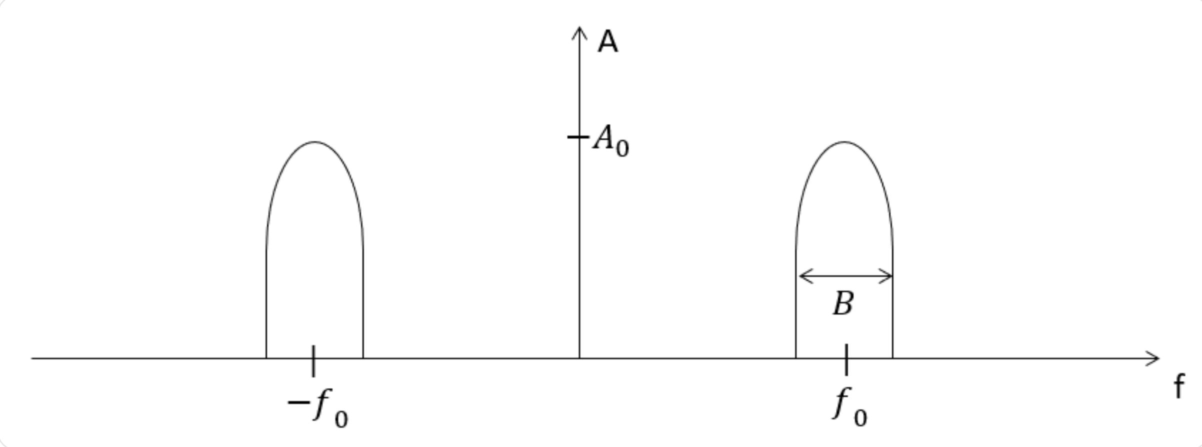


Figure 3: Spectrum of the received signal r_{RF}

Usually, this signal cannot be converted by a standard CAN. If we take the Wi-Fi for example, we would need a CAN with a data frequency of at least $5GHz$: this highly expensive.

To get the signal, we need to translate the frequency near $0Hz$—bringing back the signal in its baseband. We need to shift the frequency by itself to center it around $0Hz$, which can be achieved by taking $(f_c = f_0)$. We thus obtain $f_c + f_0 = 2f_0$ and $f_c + f_0 = 0$, so we can now deduce from (2) that:

$$\tilde{r}_R(t) = \frac{s_R(t)}{2}(1 + \cos(4\pi f_0 t)) - \frac{s_I(t)}{2}\sin(4\pi f_0 t)$$
$$\tilde{r}_I(t) = \frac{s_I(t)}{2}(1 - \cos(4\pi f_0 t)) - \frac{s_R(t)}{2}\sin(4\pi f_0 t)$$

After applying the Fourier transform, we get:

$$\begin{cases} \tilde{R}_R(f) = \frac{S_R(f)}{4} * (2\delta(f) + \delta(f + 2f_0) + \delta(f - 2f0)) - \frac{S_I(f)}{4} * j(\delta(f + 2f_0) - \delta(f - 2f_0)) \\ \tilde{R}_I(f) = \frac{S_I(f)}{4} * (2\delta(f) - \delta(f + 2f_0) - \delta(f - 2f0)) - \frac{S_R(f)}{4} * j(\delta(f + 2f_0) - \delta(f - 2f_0)) \end{cases}$$

$$\Leftrightarrow \begin{cases} \tilde{R}_R(f) = \frac{1}{4}(2S_R(f) + S_R(f + 2f_0) + S_R(f - 2f_0) - jS_I(f + 2f_0) + jS_I(f - 2f_0)) \\ \tilde{R}_I(f) = \frac{1}{4}(2S_I(f) - S_I(f + 2f_0) - S_I(f - 2f_0) - jS_R(f + 2f_0) + jS_R(f - 2f_0)) \end{cases}$$

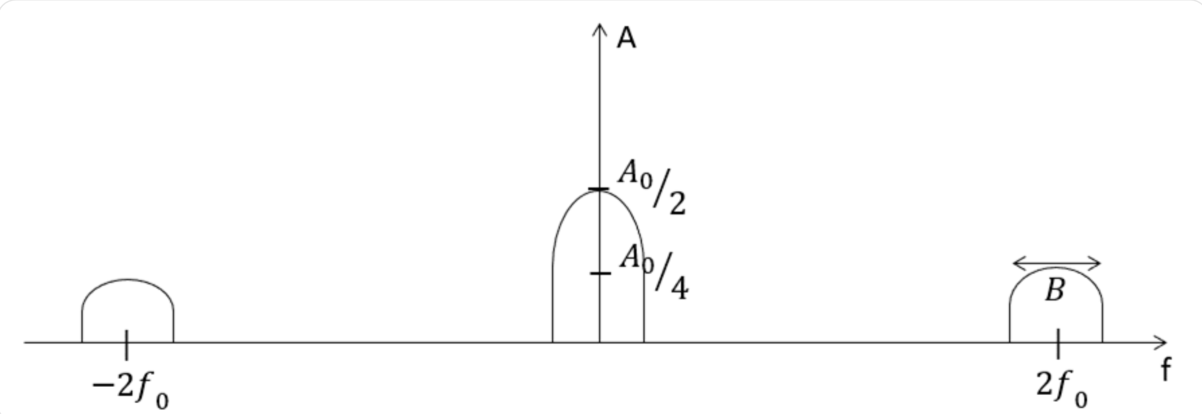


Figure 4: Spectral representation of $|\tilde{R}_R| = |\tilde{R}_I|$, the received signal after its first stage in the USRP

To retrieve the original signal, from this spectral representation, the h filter should thus be a low pass filter, and have a f_{cutoff} cutoff frequency taller than half of the bandwidth (to avoid filtering the payload of the signal), and smaller than twice f_0, minus half of the bandwidth, as written below. Also, the signal needs to be amplified, and require a gain equals to 2 to retrieve the original amplitude.

$$\textit{Charecteristics of the h filters}$$

$$\begin{cases} \frac{BW}{2} < f_{cutoff} < f_0 - \frac{BW}{2} \\ \text{gain} = 2 \end{cases}$$

Question 3

Can the receiver presented in Fig. 2 work with wide-band signals? Explain.

In narrow-band signals, $f_0 > \frac{B}{2}$: it corresponds to the spectrum displayed above (Fig. 4), and the receiver can work.

However, with a wide-band signal, $f_0 < \frac{B}{2}$. Therefore, as a result, we would obtain:

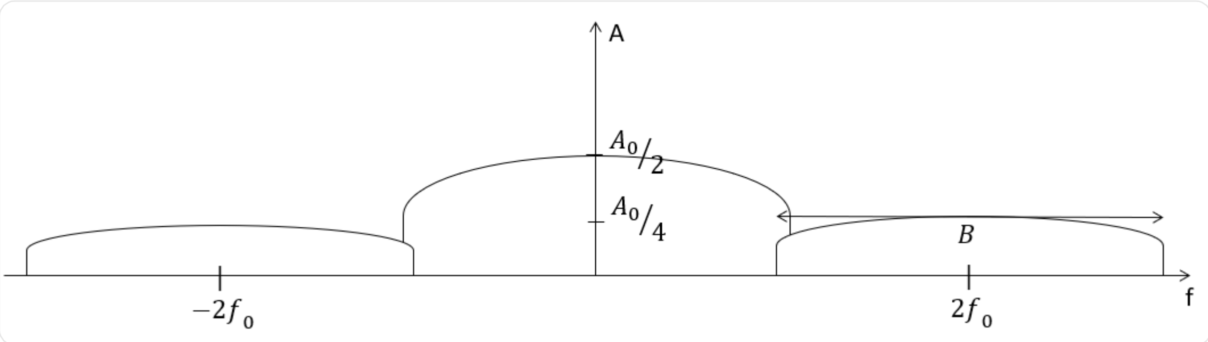


Figure 5 Spectral representation of a wide-band signal with $f_0 < \frac{B}{2}$

With the graph above, we can see that an IQ receiver cannot work in the case of a wide band: it needs non-causal filters, which is impossible to do analogically.

Question 4

How must the sampling period T_e be chosen in order to recover $r_R(t), t \in \mathbb{R}$ from $r_R(k \times T_e), k \in \mathbb{Z}$?

To determine the sampling period T_e, we apply the Shannon-Nyquist theorem: $f_e > 2f_{max}$

Since $f_{max} = \frac{BW}{2}$, we have $f_e > \frac{2BW}{2} \Leftrightarrow f_e > BW$

$$\Leftrightarrow \qquad T_e < \frac{1}{BW}$$

Question 5

Why do not we interchange the stages of frequency transposition and analog to digital conversion?

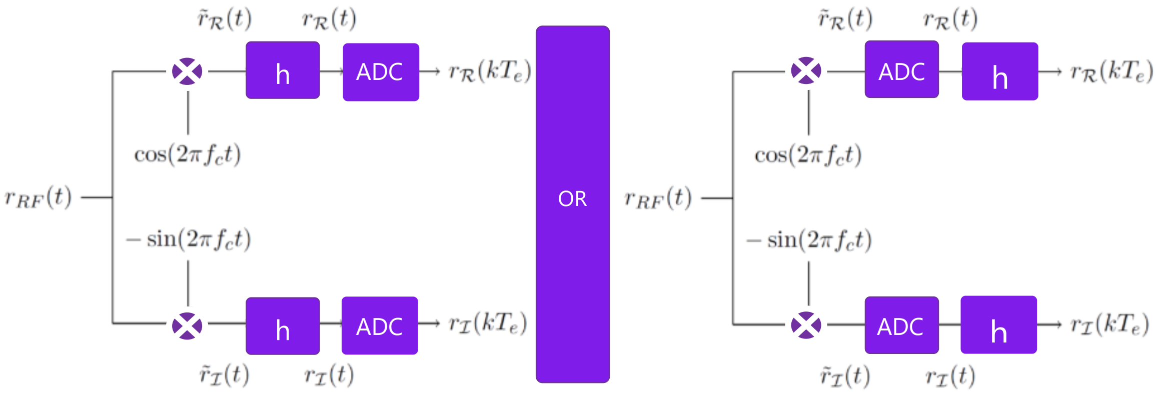


Figure 6. Interchanging the stages of frequency transposition and digital conversion

We do not interchange the stages of frequency transposition and analog to digital conversion in order to lower the frequency needed for the ADC. In fact, for high frequency signal such as WIFI (2,4 GHz), regarding Shanon-Nyquist criteria, the ADC would need to sample at 5 GHz if there is no frequency transposition before $\left(f_e > 2\left(f_0 + \frac{BW}{2}\right)\right)$

Through those five first questions, we have already proved that this circuit (Fig. 2) can extract the signal of a specific frequency and then re-build it, if working in the narrow-band.

We indeed have demonstrated that two real signals $s_R(t)$ and $s_I(t)$ can be transmitted on the carrier frequency f_0 and then be perfectly recovered thanks to a well-adapted IQ receiver. Next, we will ignore the operation of transposition on the carrier frequency to study the transmission system. We define two equivalent models for $s_{RF}(t)$ based on the principle that all real signals have a symmetrical Hermitian Fourier transform. Thereby, in the case of narrowband signals—as presented in Fig. 7 below—the representation in positive frequencies (or negative) is sufficient and it is possible to hide the value of the carrier frequency f_0.

Question 6

Supposing a real narrow-band signal:

$$\begin{aligned} s_{RF}(t) &= A(t)\cos\big(2\pi f_0 t + \varphi(t)\big) \\ &= s_R(t)\cos(2\pi f_0 t) - s_I \sin(2\pi f_0 t), \qquad t \in \mathbb{R} \end{aligned}$$

Express—in frequential then in temporal—its analytic signal and its complex envelop in function of f_0, knowing that $S_{RF}(f) = {S_{RF}}^*(-f)$.

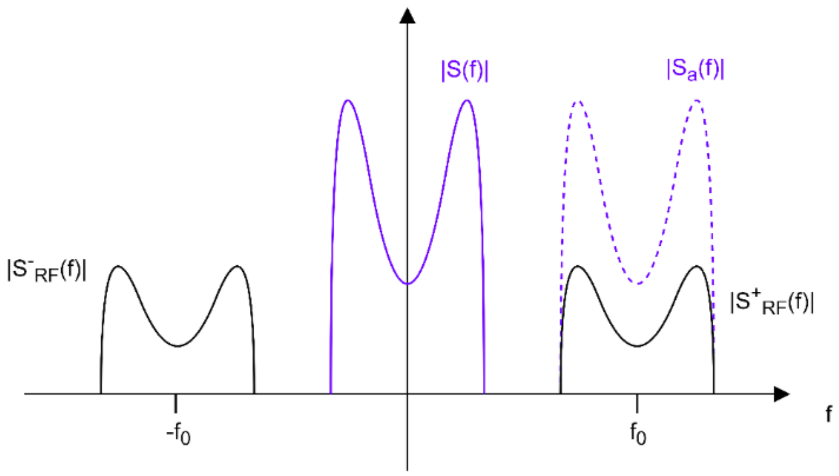


Figure 7: Spectral representations of a real narrow-band signal $|S(f)|$, of its analytic signal $|S_a(f)|$ and of its complex envelope $|S^+{}_{RF}(f)|$

In the frequency domain:

$$S_{RF}(f) = \frac{S_R(f)}{2} * \left(\delta(f - f_0) + \delta(f + f_0)\right) - \frac{S_I(f)}{2} * \left(\delta(f - f_0) + \delta(f + f_0)\right)$$

$$S_{RF}(f) = \frac{1}{2}\left(S_R(f - f_0) + S_R(f + f_0) + jS_I(f - f_0) - jS_I(f + f_0)\right)$$

So $$S_a(f) = \begin{cases} 2 \times S_{RF}^{+}(f), & \text{if } f > 0 \\ 0, & \text{if } f < 0 \end{cases}$$

$$\Leftrightarrow \quad S_a(f) = S_R(f - f_0) + jS_I(f - f_0), \quad \forall f > 0$$

Switching the narrow-band signal and its analytic signal to a temporal representation:

$$S_a(f) = \left(S_R(f) + jS_I(f)\right) * \delta(f - f_0), \forall f > 0$$

So $$s_a(t) = \left(s_R(t) + j.s_I(t)\right) \times e^{j2\pi f_0 t}, \quad \forall t$$

$$S(f) = S_a(f + f_0) = S_R(f) + j.S_I(f)$$

So $$s(t) = s_R(t) + j.s_I(t), \quad \forall t$$

Concerning the use of URSP in practice

An USPR can convert lots of protocols. Instead of having N hardware for N protocols it allows to have one hardware for all protocols. The aim of USPR is 5G for interprotocol connectivity.

2. Reception of frequency modulation (FM) broadcasting

The Very High Frequency (VHF) band extends from $30MHz$ to $300MHz$. The radiofrequency propagation in this band and for terrestrial communications is nearly geometrical. We focus on the sub-band between $87.5MHz$ and $108MHz$, which is now dedicated to FM broadcasting (https://en.wikipedia.org/wiki/FM_broadcasting/). The different channels are spaced by at least $100kHz$ so that it is theoretically possible to have 203 simultaneous broadcasting stations. In practice, a particular transmitter broadcasts far fewer stations, allowing the reuse of the frequency channels in the frame of a cellular planning and guaranteeing a good isolation between channels.

The FM broadcasting recording file has been obtained thanks to the acquisition system introduced in the first part. It was recorded at Toulouse in 2015, using a center frequency of $99.5MHz$ and a sampling frequency of $F_e = 1.5MHz$. The transmitter is probably located at the Pech David observatory, near Rangueil Hospital.

The objective of this first exercise is to learn as much as possible about the content of this recording and in particular to restore the audio content. To meet this goal, we will use the GNURadio development environment and precisely the GRC tool with its graphical interface similar to Matlab/Simulink one.

2.1 Frequency analysis of the recording

To exploit the recording, we implement the chain presented below in Fig. 7.

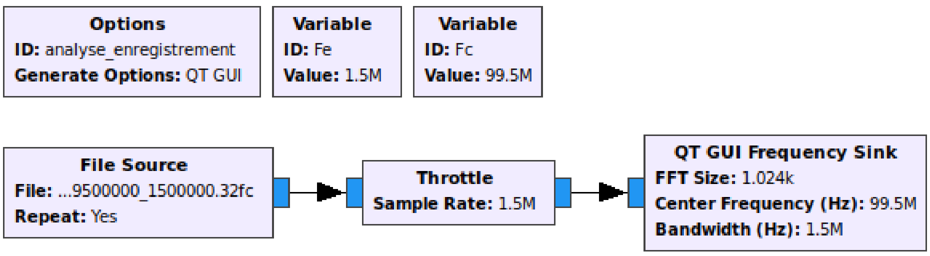


Figure 8: Frequency analysis processing chain on GNU Radio

Question 7

Present the role of each block used in the processing chain.

Here are the different blocks used:

- **File Source** to extract the signal raw from the file
- **Throttle** to limit the data rate of the non-hardware source block
- **QT GUI Frequency Sink** to display signals in frequency

Question 8

Specify the values of the variables in the characteristics function of the recording file.

The source file comes from a signal centered at $99.5M\ Hz$, called f_c. It was shifted to $0\ Hz$ then sampled with a frequency of $1.5M\ Hz$, called F_e.

To analyze this file, we throttle the sample rate to the value of F_e. Then we display the signal centered at f_c with a range of F_e.

Question 9

How many frequency channels—to be noted L—do you observe? According to the allocation of frequencies in the FM band near Toulouse.

We can distinctly observe two frequencies: RFM $(99.1MHz)$ & Skyrock $(100.0MHz)$. We can guess a third one, at $99.5MHz$: Nostalgie, but the signal is very weak.

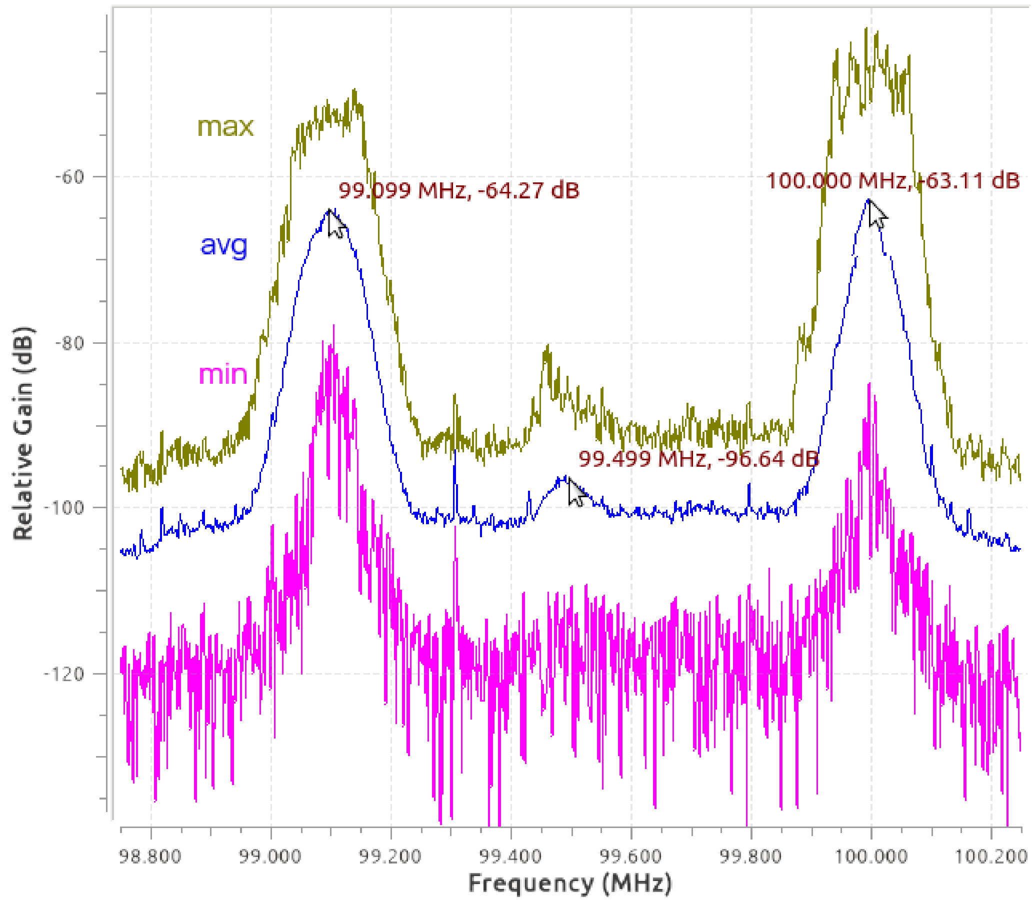


Figure 9: Spectrum (FFT) of the raw signal

Question 10

What is the measured signal-to-noise ratio in decibel? Do you think that is enough to be able to demodulate the signal?

$$SNR_{dB} = 10 \times \log\left(\frac{P_{signal}}{P_{noise}}\right)$$
$$\text{with } P = 10^{\frac{P_{dB}}{10}} \text{ and } P_{dB} = 10 \times \log_{10}(P)$$

$$SNR_{Skyrock} = 10 \times \log\left(\frac{10^{-6.4}}{10^{-10}}\right) = 36\ dB$$
$$SNR_{RFM} = 10 \times \log\left(\frac{10^{-6.3}}{10^{-10}}\right) = 37\ dB$$
$$SNR_{Nostalgie} = 10 \times \log\left(\frac{10^{-9.6}}{10^{-10}}\right) = 4\ dB$$

Skyrock & RFM can easily be demodulated, according that signal is almost 10^4 times greater than the noise. Nostalgie has a SNR of 2.5, so during the demodulation most of the signal will be lost.

Question 11

What is the approximate bandwidth of a channel?

If we consider the actual 3dB-bandwidth for the average signal of RFM and Skyrock, we find a very short bandwidth (respectively $80kHz$ and $40kHz$). Considering it from the maximum signal we get, we can define a window of $140\ kHz$.

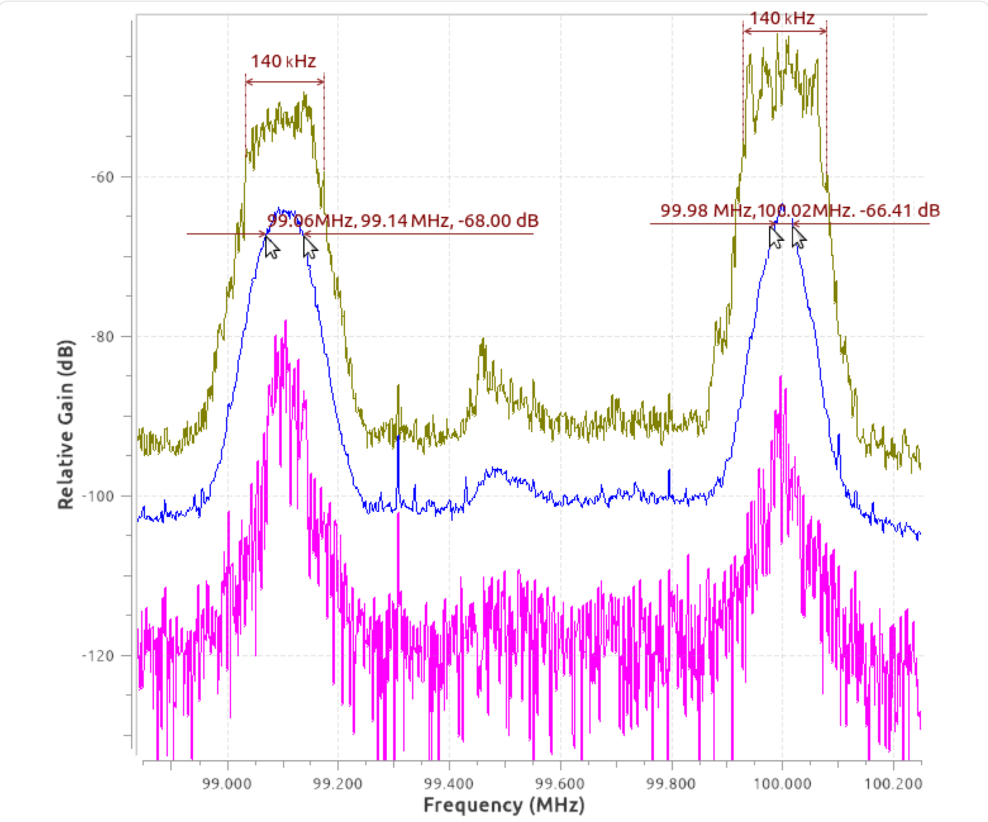


Figure 10: Bandwidth of the signals

2.2 Channel extraction by frequency transposition and low-pass filtering

During the previous step, several RF broadcasting stations have been identified. Now, we want to receive each one separately, using a new processing chain based on the previous one. To do this, we propose a two stages reception described as:

1. Frequency transposition, to center the useful signal in terms of frequencies
2. Low pass filtering, to attenuate the out-of-band noise.

Noting $r[k]$ the complex envelope (regarding f_c) of the recorded signal, sampled at the frequency F_e, the frequency transposition for a quantity $f_l, l \in \{1; ...; L\}$ can be write:

$$r_l[k] = r[k]e^{-j\times 2\pi\frac{f_1}{F_e}k} \quad (8)$$

To do this frequency transposition through GRC, we use the next blocks:

- **Signal Source** — to generate the complex exponential
- **QT GUI Range** — to define dynamically f_l during the software execution
- **Multiply** — to make the product (8)

This time, we center the visualization on the null frequency (**QT GUI Frequency Sink** > Center Frequency: 99.5 MHz).

Question 12

What are the frequency offsets needed to center each channel?

To estimate the frequency offset needed, we made a system shown below (Fig. 11) where a slider is available on the interface to offset the signal accordingly.

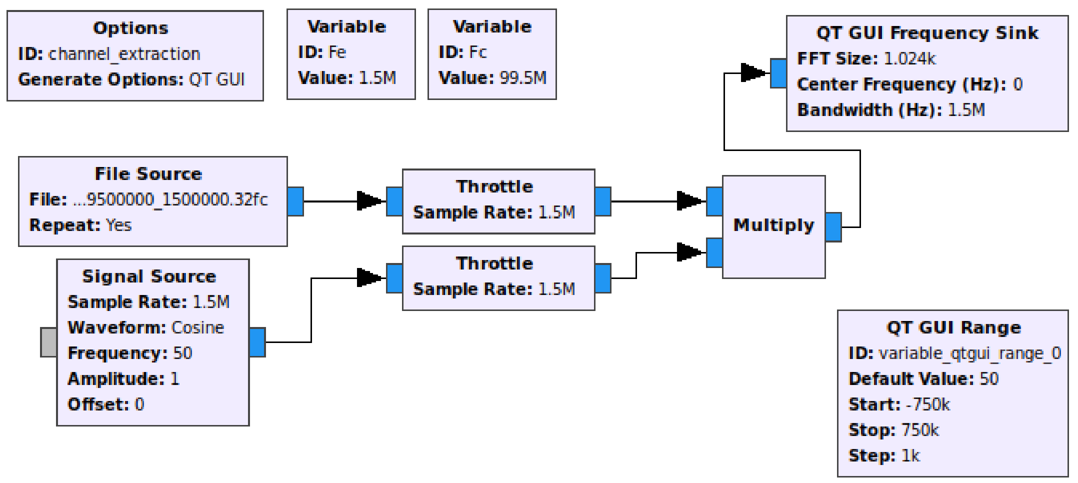


Figure 11: Introduction of the frequency transpostion in the frequency analysis processing chain

We estimated that RFM needs a frequency correction of $-500\,kHz$. Nostalgie needs a correction of $10kHz$ and Skyrock needs a shift of $400\,kHz$.

Question 13

What happens if the frequency offset is higher than the sampling frequency F_e?

If we exceed f_e, we see that there is a circular permutation of the signal, since the complex exponential has the periodicity $\frac{2\pi}{f_e}$.

Question 14

What are the low-pass filter parameters, as well as those of the frequency analyzer at the output of the filter?

Once a channel $r_l[k]$ frequency is centered, we must implement a low-pass filter by considering the channel bandwidth as defined in Part 2.1. Then we can obtain the signal noted $y_l[k]$.

To do this low pass filtering through GRC, we use the next block: **Low Pass Filter** where the cut-off frequency must ideally be half of the frequency bandwidth, according to what we have seen in the first part; to be sure we don't lose any part of the signal, we decided to take a larger cut-off frequency, at $100kHz$.

We define the transition as 10% of the cut-off frequency. After filtering, we can process a decimation with a 6 factor in order to lighten the computation load. After that, the sample rate will be divided by this same factor. Below is presented our new frequency analysis processing chain.

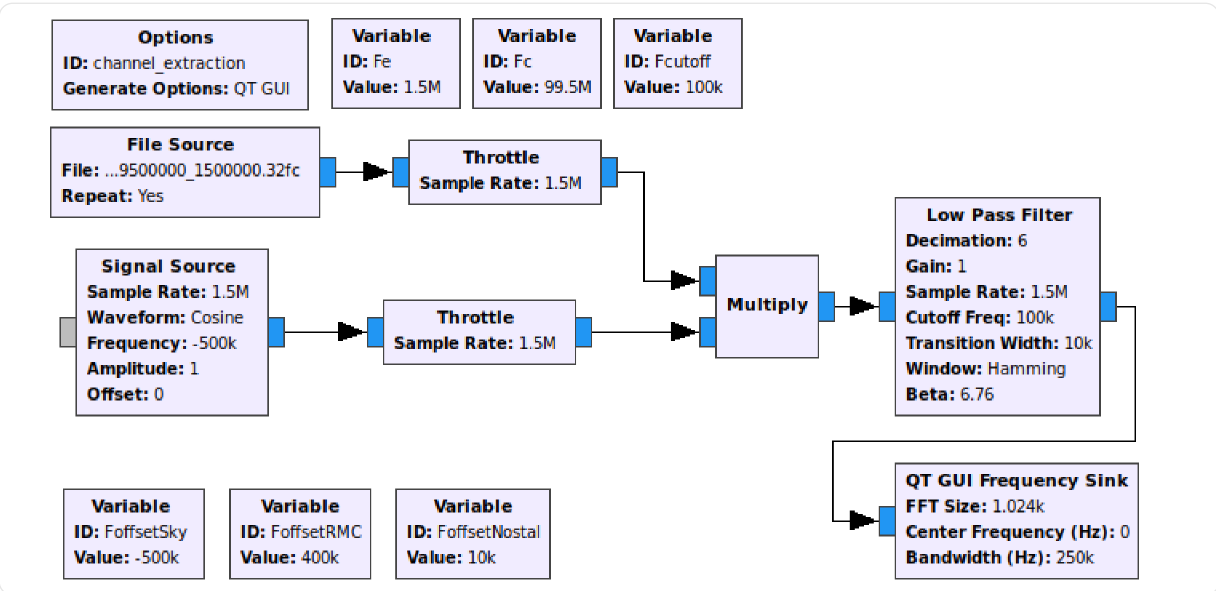
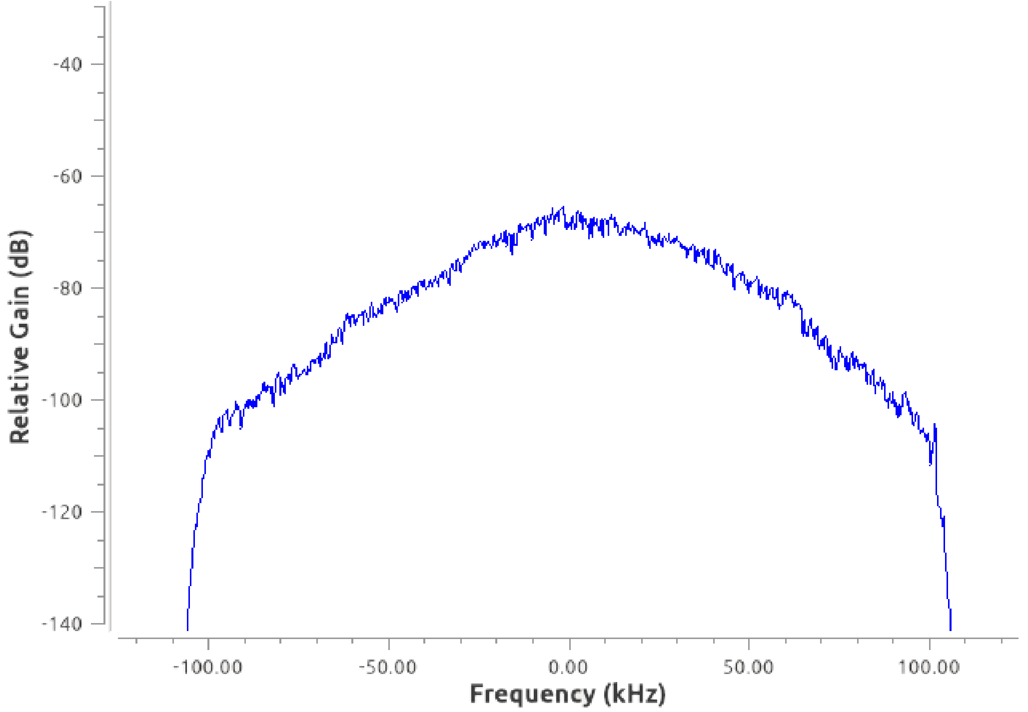


Figure 12: Introduction of the low-pass filter in the frequency analysis processing chain

Below is presented the result of the new processing chain. It's still not clear, but next stage of the frequency analysis process will allow us to retrieve the actual signal.



Firgure 13: Signal filtered

2.3 Frequency demodulation and restitution

In order to restore the content of each radio broadcasting station using a sound card, we have firstly to understand the modulation method used. The signal to be transmitted consists of two stereophonic channels $g(t) \in \mathbb{R}$ and $d(t) \in \mathbb{R}$, they are centered in frequency and have a maximum frequency of 15 kHz (or mono-lateral band). To ensure compatibility between monophonic receivers and stereophonic ones, these two channels are multiplexed to form the message:

$$\mathrm{m(t) = g(t) + d(t) + Asp \cdot \cos(2 \cdot \pi \cdot fsp \cdot t) + [g(t) - d(t)] \cdot \cos(2 \cdot \pi \cdot 2 \cdot fsp \cdot t)} \quad (9)$$

with $fsp = 19\ kHz$ a pilot carrier frequency with an amplitude of $Asp = 2$

According to the spectrum of the signal presented in Fig. 6, we note that the monophonic receiver consists of implementing a low-pass filter at 15 kHz. The stereophonic receiver (not discussed here) is required to amplitude demodulate the signal around 38 kHz and recombine it with the baseband signal to reconstruct the left and right channels.

In practice, the composite signal may comprise of frequencies greater than 53 kHz. This is notably the case for the radio data system (RDS) positioned around 57 kHz, which allows the transmission of digital information relating to the program (e.g. the name of the radio station, the title and the name of the artist of the played music, etc.). This service is not considered in the rest of the work.

The composite signal $m(t)$ is then frequency modulated, and the radiofrequency signal centered on $f0$ at the output of the transmitter is noted:

$$s_{RF}(t) = \mathrm{A} \cdot \cos(2 \cdot \pi \cdot \mathrm{f_0} \cdot \mathrm{t} + \frac{\Delta \mathrm{f}}{\max(|\mathrm{m(t)}|)} \cdot \int_{-\infty}^{t} \mathrm{m(u).\,du\,u}) \qquad (10)$$

...with Δf the maximum frequency excursion of the modulation, fixed at 75 kHz in the present case. We show that the frequency-modulated signal occupies an infinite band, but decreases rapidly, so that it can be approached via the Carson rule:

$$\mathrm{BFM} \approx 2 \cdot (\Delta \mathrm{f} + \mathrm{fm}) \qquad (11)$$
$$\text{with } f_m \text{ the maximum frequency of the composite signal } m(t).$$

Question 15

Using the Carson rule, check that the bandwidth of the channel measured in the previous part confirms the theory.

We want to verify Carson's rule:

$$\begin{aligned} B_{FM} &= 2(\Delta f + f_m) \simeq 256kHz \\ \Delta f \ &: \text{Max freq. excursion } (75k \text{ for FM}) \\ f_m \ &: \text{Max freq. of modulating signal} \end{aligned}$$

We thus find almost what we have in the previous part (almost 200 kHz). The signal is a signal with stereophonic sound, as can be seen here:

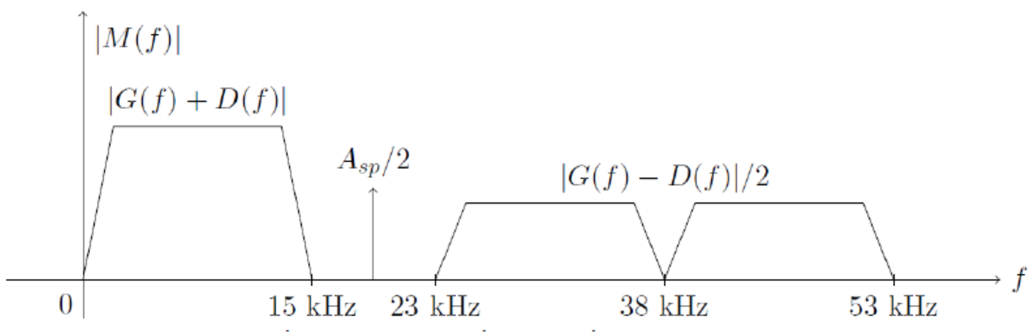


Figure 14: Stereophonic sound signal

Question 16

From the expression (10) of the transmitted signal and the affected processes until now (frequency transposition and low-pass filtering), show that the signals $y_l[k]$ can be noted:

$$y_l[k] = Ae^{j \times k_f \times \sum_{i=0}^{k} m[i]} + b[k]$$

Define the value of k_f, with $b[k]$ a complex noise term introduced by the propagation channel as well as by the transceiver itself.

$$(1) : s_{RF} = A(t)\cos\big(2\pi f_0 t + \varphi(t)\big),\ \forall t \in \mathbb{R}$$
$$(10) : s_{RF}(t) = A(t)\cos\left(2\pi f_0 t + \frac{\Delta f}{\max(|m(t)|)} \int_{-\infty}^{t} m(u)du\right)$$

We identify:

$$\varphi(t) = \frac{\Delta f}{\max(|m(t)|)} \int_{-\infty}^{t} m(u)du$$
$$\begin{cases} s_R(t) = A(t)\cos\big(\varphi(t)\big) \\ s_I(t) = A(t)\sin\big(\varphi(t)\big) \end{cases}$$

We remember from question 6 that:

$$s(t) = s_R(t) + js_I(t) = A(t)\cos\big(\varphi(t)\big) + jA(t)\sin\big(\varphi(t)\big)$$
$$s(t) = A(t)e^{j(\varphi(t))} = A(t)e^{j\frac{\Delta f}{\max(|m(t)|)}\int_{-\infty}^{t} m(u)du}$$

We discretize:

$$s[k] = A\left[\frac{k}{F_e}\right] e^{j\frac{\Delta f}{\max(|m(t)|)}\sum_{i=0}^{k} m(i)} + b[k]$$

For FM, $A\left[\frac{k}{F_e}\right]$ is constant, and equals A, so:

$$y_f[k] = Ae^{jk_f\sum_{i=0}^{k} m(i)} + b[k], \qquad \text{where } k_f = \frac{\Delta f}{\max(|m(t)|)}$$

Question 17

Plot the spectrum of the demodulated channel and compare with the Fig. 14.

We finally present the demodulated channel on the figure below:

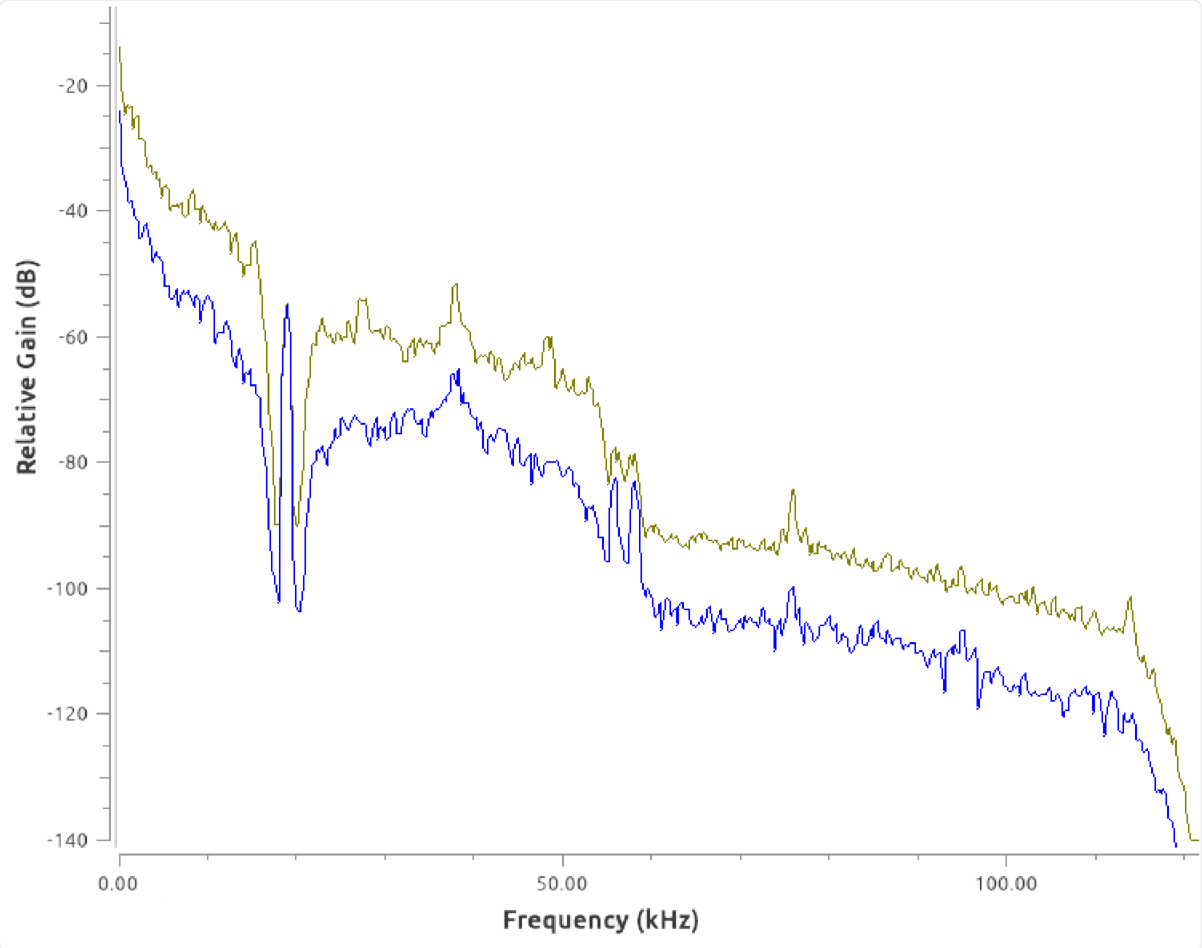


Figure 15: Spectrum of the demodulated signal

This is the spectrum of the demodulated signal for Skyrock radio at -500 kHz. In comparison with Figure 6, the following similarities can be noted:

- Carrier signal at 19 kHz
- Monophonic signal between -19 kHz and 19 kHz centred around zero
- Stereo phonic signal around 38 kHz with L and R signal
- And even the RDS (channel info) around 57 kHz. (19k > max=avg, "fixe")

18. Who won the Sam Smith album? What do we listen to on other stations?

As we listen to the radio, on the channel Skyrock, we clearly heard that it was actually Jordi who ended up winning the Sam Smith album.

2.4 Real time implementation with an USRP receiver

In this part, we replace the previously used recording file with a SDR transceiver in order to receive a signal in real time. We can save the new processing chain as "usrp_fm_receiver.grc" and connect an USRP on the computer. To do this, we propose the use of the following block, UHD USRP Source, to interface GRC with the driver of the USRP. As the USRP has frequency conversion stage, we do not have to care about the frequency transposition stage treated in Part. 2. 2

$$(16): \; s(t) = m(t) \pm j\mathcal{H}\{m(t)\}$$

$$\text{and using } \mathcal{H}\{m(f)\} = -j\; sgn(f)m(f)$$

$$S(f) = m(f) \pm sgn(f)m(f)$$

The $\pm$ depends if the left or right part of the signal is emitted by the station.

$$S_a(f) = m(f - f_0) \pm sgn(f - f_0) + sgn(f - f_0)m(f - f_0)$$

$$\begin{cases} \text{case } (+): S_a(f) = m(f - f_0) + sgn(+f - f_0)m(f - f_0) \\ \text{case } (-): S_a(f) = m(f - f_0) + sgn(-f + f_0)m(f - f_0) \end{cases}$$

$$(15): s_{RF}(t) = Re\left(s(t)e^{j2\pi f_0 t}\right)$$

$$\text{NB: } Re(c) = \frac{1}{2}[c + c^*]$$

$$|S_{RF}(f)| = \frac{1}{2}[|S_a(f)| + |S_a^*(f)|]$$

$$\begin{cases} \text{case } (+): |S_{RF}| = \frac{1}{2}[m(f - f_0) + sgn\,(+f - f_0) \times m(f - f_0) + m(-f - f_0) + sgn(-f - f_0) \times m(-f - f_0)] \\ \text{case } (-): |S_{RF}| = \frac{1}{2}[m(f - f_0) + sgn\,(-f + f_0) \times m(f - f_0) + m(-f - f_0) + sgn(+f + f_0) \times m(-f - f_0)] \end{cases}$$

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3. Reception of VOLMET messages in AM-SSB

The frequency band named High Frequency (HF) ranges from $3MHz$ to $30MHz$ .The propagation in this band is done by successive reflections on the ionosphere and the Earth's crust. The main advantage of this technique is to have intercontinental links with a reasonable power budget (several tens of Watts). In return, it is necessary to accommodate the multiple paths resulting from this propagation process, as well as the constant evolution of the ionosphere channel (function of time of the day, solar cycles, etc.).

Next, we are interested by the frequency sub-band between $11.175\ MHz$ and $11.4\ MHz$, which is now reserved to the international aeronautic communications and in particular to VOL METEO service (VOLMET). This is a periodic broadcasting of meteorological information, using a single sideband amplitude modulation.

The VOLMET recording file has been obtained thanks to the acquisition system introduced in the first part. It was recorded at Toulouse in 2015, using a center frequency of $f_0 = 11.2965\ MHz$ and a sampling frequency of $F_e = 250kHz$. There is a single station in this record where the transmitter is located in the Royal Air Force air-base of St-Eval, United Kingdom.

The objective of this last part is to learn as much as possible of the content of this recording and to restore the phonic content.

4.1 Frequency analysis of the recording

Question 19

Plot the modulus of the discrete Fourier transform in decibels, between $f_0 - \frac{F_e}{2}$ and $f_0 + \frac{F_e}{2}$, by using *QT GUI Frequency Sink* block. Check with http://www.dxinfocentre.com/volmet.htm that the VOLMET station located in the Royal Air Force air-base of St-Eval, United Kingdom, is well observed at the expected frequency.

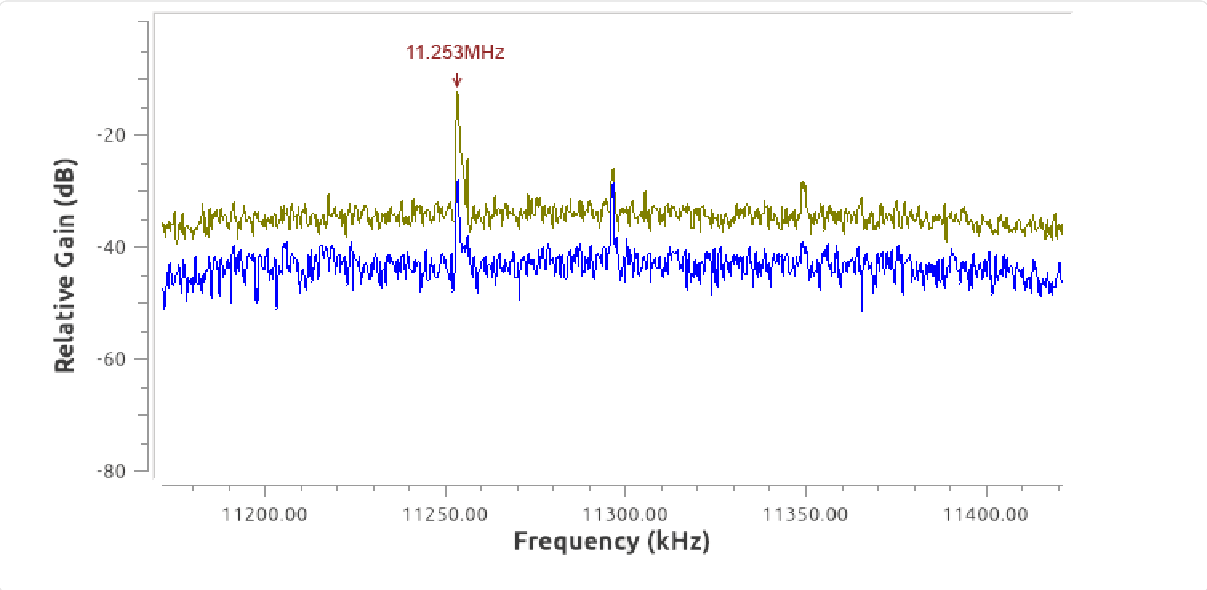

Figure 16: Fast-Fourier Transform of the signal between $f_0 - \frac{F_e}{2}$ and $f_0 + \frac{F_e}{2}$

We observed a pic at higher intensity at 11.253 MHz at shown above (Fig. 14). Online, the list of the VOLMET stations' frequencies are available. That way, we found that the signal was coming from a military station in Great Britain (Fig. 15).

11.3 MHz								
11.247	, 35	MTS	FLK	VIPER		-51 50 14	-58 28 14	IRREGULAR HOURS
11.253	Cont	MKL	GBR	MILITARY ONE		50 28 58	-5 00 00	
11.297	25 , 55	RLAP	RUS	ROSTOV	RR	47 15 12	39 49 02	DAY
11.318	00 , 30	UBB-2	RUS	SIVKAR	RR	61 38 17	50 31 49	DAY
	10 , 40	UNNN	RUS	NOVOSIBIRSK	RR	55 00 16	82 33 44	DAY
	15 , 45	RQCI	RUS	SAMARA	RR	53 11 00	49 46 00	DAY

Figure 17: VOLMET stations around 11.253 MHz from dxinfocentre.com/volmet.htm

4.2 Frequency transposition

To do this frequency transposition through GRC, as $r_1[k] = r[k] \times e^{j \times 2\pi \frac{f_1}{F_e} k}$, we use the next blocks:

- **Signal Source** to generate the complex exponential
- **QT GUI Range** to define dynamically f_1 during the software execution
- **Multiply** to make the product (8)

Question 20

What are the frequency offsets f_1 needed to center the channel with the maximum power at the null frequency? Plot the modulus of the discrete Fourier transform in decibels, between $f_0 - \frac{F_e}{2}$ and $f_0 + \frac{F_e}{2}$, by using *QT GUI Frequency Sink* block.

We made a variable to observe what offset is needed to center the signal. The variable is indicated at the top and can be changed by the slider of numeric values.

We estimated that way that an offset of 43.1 kHz is required as shown below (Fig 18).

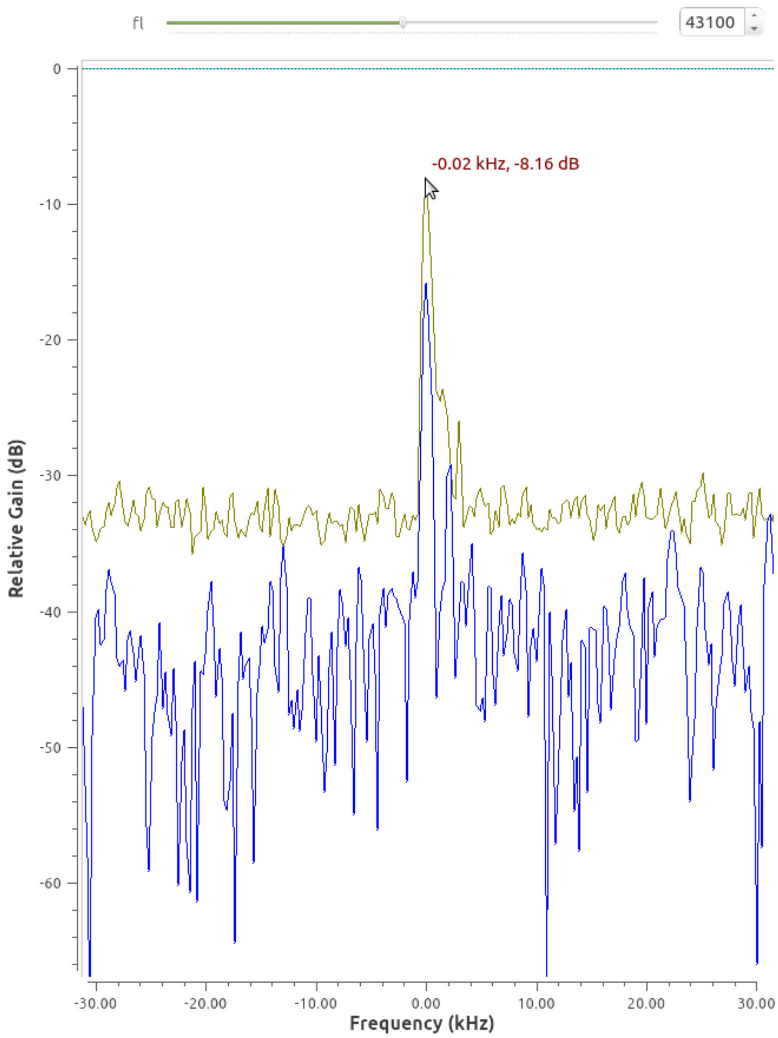

Figure 18: Spectrum of the signal shifted by an offset of 43.1 kHz to center the signal

4.3 Single sideband amplitude demodulation

The transmission method used by the VOLMET service is single-sideband amplitude modulation. We will first study the signal of the VOLMET service in order to apply the relevant filter.
At the beginning, we calculate the modulus of the VOLMET signal to identify the part we want to filter.

We already found the shape of S_{RF} before:

$$\begin{cases} \text{case } (+): |S_{RF}| = \frac{1}{2}[m(f-f_0) + sgn\,(+f-f_0) \times m(f-f_0) + m(-f-f_0) + sgn(-f-f_0) \times m(-f-f_0)] \\ \text{case } (-): |S_{RF}| = \frac{1}{2}[m(f-f_0) + sgn\,(-f+f_0) \times m(f-f_0) + m(-f-f_0) + sgn(+f+f_0) \times m(-f-f_0)] \end{cases}$$

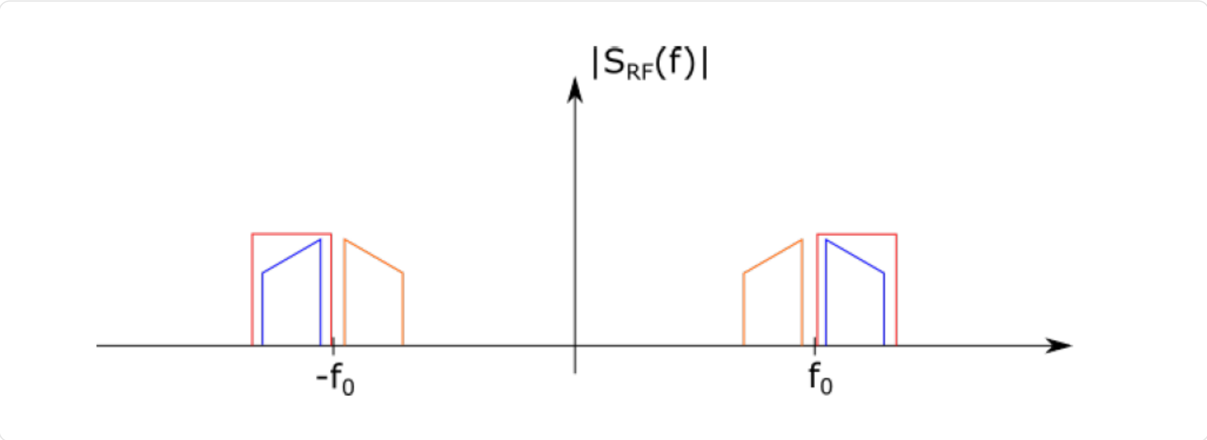

Figure 19: Half of the signal is received

We see that our bandwidth is 3 kHz (11.256 MHz - 11.253 MHz), so we only have half of the bandwidth ($\frac{B}{2}$), and we only have the positive part of the spectrum. We designed this filter, and here is its module and phase:

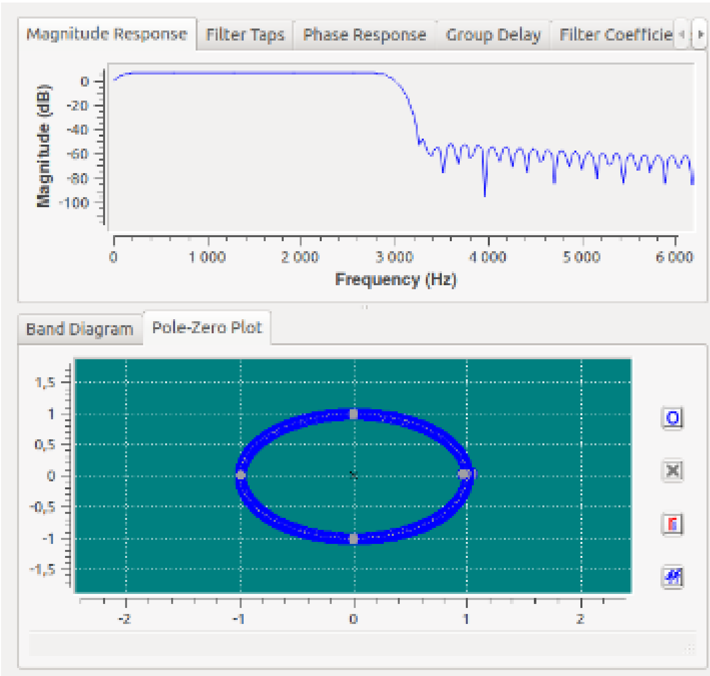

Figure 20: Module and phase of the pass-band filter

We now must filter the signal r1[k] with a bandpass filter to obtain as output y[k] which corresponds to the single upper sideband of the amplitude modulated signal (blue signal on the right side in figure 19).

The filter applied to the signal corresponds to the red filter to the right of figure 19, and is a bandpass filter acting between f0 and f0 + 3kHz.

We tried to recover the real signal in the form $\tilde{m}[k] = [k] + \tilde{b}[k]$:

$$r1[k] = m[k] + jH(M[k]) + b[k]$$

$$\tilde{m}[k] = Re(r1[k]) = m[k] + \tilde{b}[k] \; with \; \tilde{b}[k] = Re(b[k])$$

We have the following set-up:

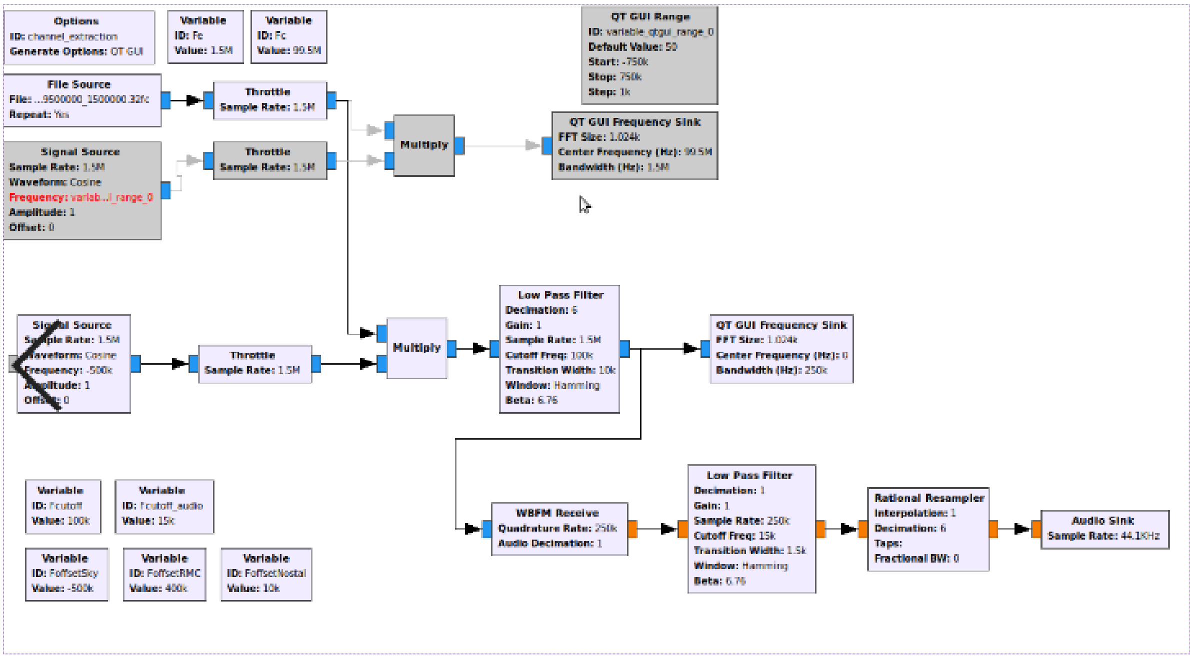

Figure 11: GNU Radio demodulation set-up

Eventually, we succeed to demodulate the signal from the military station in Great Britain

6 Conclusion

As a short conclusion, we can see that, through this practical work, we saw from the theorical point of view until the practice one how works—and how useful can be—SDR technology and the many possible applications of it, using tools like USRP and GRC, especially regarding the future of protocols of communication for IoT devices.

5 Help and glossary

5.1 Keywords

ADC Analog to Digital Converter
DAC Digital to Analog Converter
SDR Software Defined Radio
USRP Universal Software Radio Peripheral

5.2 Trigonometric formula

$$\cos(a) \times \cos(b) = \frac{\cos(a+b) + \cos(a-b)}{2}$$
$$\sin(a) \times \sin(b) = \frac{\cos(a-b) - \cos(a+b)}{2}$$
$$\sin(a) \times \cos(b) = \frac{\cos(a+b) + \sin(a-b)}{2}$$

5.3 Main Fourier transforms

$$\begin{aligned} s(t) &\Rightarrow S(f) \\ 1 &\Rightarrow \delta(f) \\ \cos(2\pi f_0 t) &\Rightarrow \frac{1}{2}[\delta(f+f_0) + \delta(f-f_0)] \\ \sin(2\pi f_0 t) &\Rightarrow \frac{j}{2}[\delta(f+f_0) - \delta(f-f_0)] \end{aligned}$$

$$\text{Multiplication} \Rightarrow \mathit{Convolution}$$

5.4 Nomenclature

$\boldsymbol{s_{RF}(t)}$	**transmitted signal in time domain**	
$s_R(t)$	**real** part ...	of the transmitted signal in time domain (in-phase)
$s_I(t)$	**imaginary** part ...	of the transmitted signal in time domain (quadrature)

$\boldsymbol{r_{RF}(t)}$	**received signal in time domain**	
$\tilde{r}_R(t)$	**real** part …	of the received signal in time domain
$\tilde{r}_I(t)$	**imaginary** part …	of the received signal in time domain
$r_R(t)$	real part …	of the received signal in time domain **after filtering**
$r_I(t)$	imaginary part …	of the received signal in time domain **after filtering**
$r_R(k.T_e)$	real part …	of the received signal in time domain **after filtering & sampling**
$r_I(k.T_e)$	imaginary part …	of the received signal in time domain **after filtering & Sampling**
$\boldsymbol{s_a(t)}$	**analytic signal**	
$\boldsymbol{s(t)}$	**complex envelop**	
$S_R(f)$	Fourier transform of real part of the received signal	
$S_I(f)$	Fourier transform of imaginary part of the received signal	
$\tilde{R}_R(f)$	Fourier transform of real part of the received signal, **after filtering**	
$\tilde{R}_I(f)$	Fourier transform of imaginary part of the received signal, **after filtering**	