

École polytechnique de Louvain

Characterization of a FMCW harmonic radar system for Asian hornet tracking

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Abstract

With the recent spread of Asian hornets across Europe, the development of technologies enabling nest localization has become an active area of research. Existing solutions being costly and ecologically problematic, a solution based on an FMCW harmonic radar and a low-cost harmonic tag is developed in this work. A radar transmitting a signal at 22.85 dBm with a receiving noise figure of 3.95 dB has been designed and tested in laboratory. A processing gain of 43.01 dB achieved through Range-Doppler map computation yields a theoretical maximum range of 11.5 m. In practice, a maximum range of 3.5 m was reached in laboratory, the gap being mainly attributed to signal impurity, harmonic leakage of the fundamental frequency, and an underestimation of the noise floor. These results validate the concept and identify concrete directions for future improvements toward a field-deployable system.

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Chapter 1

Introduction

1.1 Context

Asian hornets are a species of hornet that was accidentally introduced in Europe in the early 21st century as a result of human activity. The population of Asian hornets kept growing in all European country, including Belgium, and soon became an invasive species. [5] Those insects cause substantial ecological problems, mainly because they are bees predators and the decline in bee population affects the ecosystem and the agriculture. Today, beekeepers and civilians can try to get rid of Asian hornets on their own by trapping them or locating nests with naked eyes. Another effective known method is to find a hornet and to track it back to its nest. Once the nest is found, it can be safely destroyed. Locating and destroying nests is therefore considered an efficient long-term solution to reduce the damage caused by this invasive species.



Figure 1: Asian hornet nest in dense woods (from Portsmouth & District Beekeeping Association [1])

The Asian hornet builds several nests during its life, most of which are located in trees. When those trees are in dense woods or far from populated areas, it becomes difficult to distinguish it with the naked eye, and technology becomes needed. Commercially available systems consist of a tracker/tag positioned on the back of an Asian hornet or hung to it, and a receiver.

1.II State of the art

Since the hornets invasion, several technologies have already been designed. One already commercially available and operational system is proposed by Robor Electronics Nature [6], which relies on an active transmitter attached to the Asian hornet and one or three antennas receivers. A signal is transmitted by the tag on the hornet, and the receiver(s) can be put on a pylon, a Unmanned Aerial Vehicle (UAV) like a drone, a car, or can simply be a phone with a downloaded app. The hornet is tracked until the nest is found, which is then destroyed. The tag attached to the hornet and examples of receivers are shown in Fig. 2. Tags are reusable and must be recovered from the hornets after use. To do so, hornets are lured repeatedly to a bait station over several days, and those returning multiple times are selected for tagging. Once the nest has been located, the tagged hornet returns to the bait, allowing the tag to be removed from its body. Even with this technique, those tags are still expensive. The price of the whole kit is 2900€ [7], and the transmitter, required to be a low-weight and small-size active tracker, is a really high technology costing around a hundred euros each. This price can be dissuasive for a lot of land owners, and the ecological impact of potentially losing high-technology tags in nature is not negligible.



(a)



(b)



(c)

Figure 2: Robor electronics technology to track an Asian hornet back to its nest. (a) Example of a portable receiver. (b) Example of three portable receivers. (c) Tag positioned on an Asian hornet

The efficiency of a tracking technology of Asian hornets can be evaluated with three criteria:

1. **Weight and size of the tag.** Studies showed that an Asian hornet can carry maximum 0.25 g until its behavior becomes disturbed [8]. The average size of an Asian hornet is 2cm, and according to how the tag is placed, it should be designed to avoid interfering with the flight.
2. **Power of the signal and the maximum detectable range.** In a tracking system, maximizing the received power is desirable, which can be achieved either by boosting the transmitted power or by improving the reception capabilities. The Signal to Noise Ratio (SNR) at the reception must be high enough for the processing algorithm to be able to locate the target, it means that the power of the useful signal must be sufficiently higher than the noise. The useful signal power received will decrease with the distance between the tag and the receiver. The desired maximum detectable range vary according to the technology used, but it is defined by the power of the signal. With a static receiver, the system must be able to detect the signal at approximately 2 km, as it is as far an Asian hornet tends to fly from its nest [9]. For a moving receiver on a UAV the range can be lower, but it must be high enough to avoid disturbing the hornet's behavior by following it too closely. The exact range needed for a UAV-based system is hard to define, and practical tests would be needed to estimate it.
3. **Frequency of the signal.** The frequency chosen for the modulation of the signals must be in a range with high legal power limitations and low interferences with other technologies.

Among the technologies already studied in research papers, several are promising, like the Radio-Telemetry tracking or the harmonic radars.

The Radio-Telemetry tracking is a technology involving an active transmitter placed on insects and transmitting a signal. The receiver can evaluate the distance of the insect thanks to the Received Signal Strength Indicator (RSSI). Some tags have been designed with a weight and a size low enough to be put on an Asian hornet, but the lack of accuracy of the radio telemetry tracking is still a main disadvantage [10]. Also, the fact that it necessitates an active tag does not solve the problems of the existing technologies.

The FMCW (Frequency Modulated Continuous Wave) harmonic radar tracking is another technology that can be used with a passive tag placed on the hornet, and a radar on a UAV. This alternative has been studied for several years as part of the tracking of Asian hornets. A typical harmonic tag is made of a diode and a wire which has the advantage to be compact, low-cost, and low-weight. Several papers were written about this technology, and it seems promising. That is why this thesis is focused on the design of a harmonic radar tracking system.

1.III Purposes of this thesis

In this thesis, a complete explanation of the FMCW technology is first presented, to get a knowledge of how it works. The mathematics behind it are the basis of the limitations and the main concerns when conceiving a FMCW harmonic radar.

The current developments in tracking technology using radar and harmonic tags are reviewed, then a ready-to-use FMCW harmonic radar and its associated tag is characterized. After that, a complete laboratory test setup can be designed, and limitations and potential improvements are evaluated.

The theoretical performance of the system are also investigated by accounting all signal gain and losses in the telecommunication system.

Finally, experimental validation is carried out by processing the radar output to assess the system functionality together with obtaining a first real-world evaluation of the target localization performance.

Chapter 2

FMCW working principle and application

2.I Introduction to the FMCW technology

In a FMCW harmonic radar, the transmitted signal follows a rising sawtooth frequency pattern, as illustrated in Fig. 3. This signal is transmitted with a frequency $f(t)$ that sweeps depending on time, and the harmonic tag re-emits this signal at the harmonic frequency. Distance estimation relies on a straightforward principle: the signal received at time t corresponds to the frequency transmitted at time $t - \Delta t$. The frequency difference between the transmitted and received signals, done comparing the signals RX and LO, is therefore directly proportional to the round-trip travel time of the signal, and thus to the target distance.

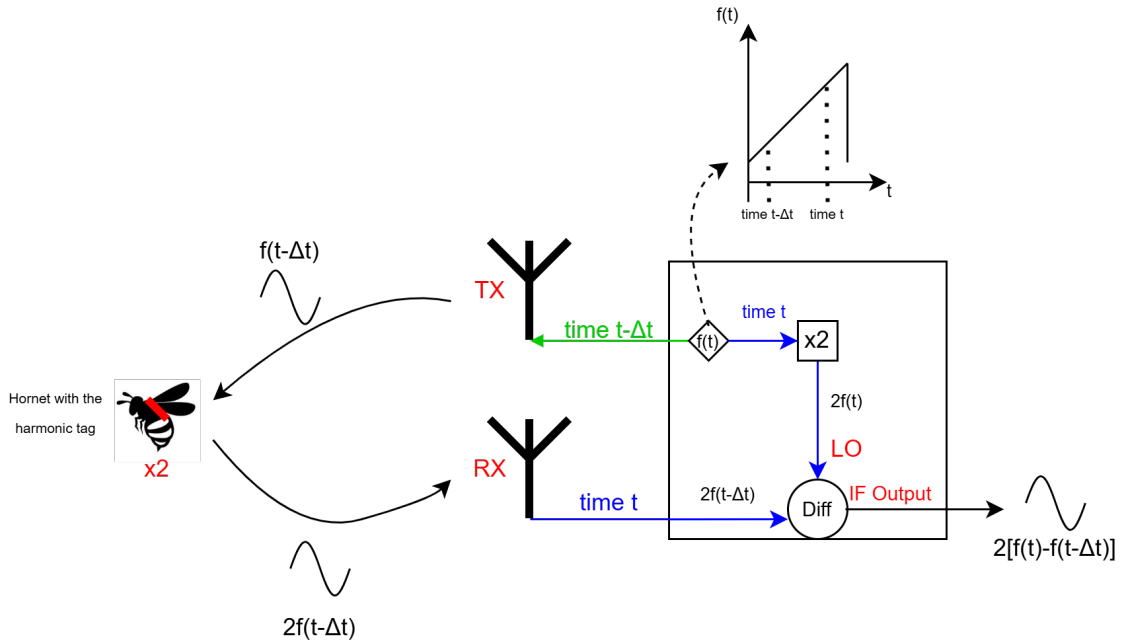


Figure 3: Explanation schematic of the working principle of a FMCW harmonic radar system

The speed of the target can also be profiled thanks to the Doppler effect. A radar signal reflected by a moving target experiences a small frequency shift upon reception, a phenomenon caused by the Doppler effect. While sending a signal at a certain frequency to a target with a speed v , the frequency seen by this target gets a shift of $\Delta f_D = \frac{vf}{c}$. In this application, a first Doppler frequency shift appears when the signal reaches the tag at the fundamental, and this shift is multiplied by two when this signal is emitted at the harmonic frequency by the tag. Then, a new Doppler frequency shift appears at the harmonic when the signal is received by the radar. The total Doppler frequency shift equals:

$$\Delta f_D = \frac{2vf_0}{c} + \frac{vf_{harmonic}}{c} = \frac{2vf_{harmonic}}{c} \quad (1)$$

2.II Mathematics of an FMCW radar

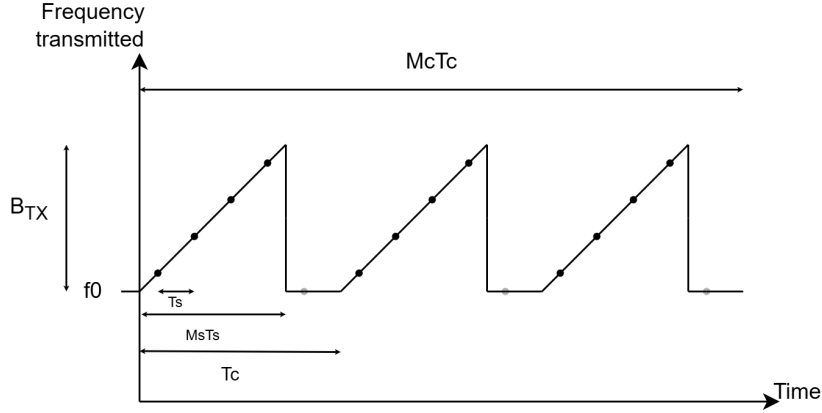


Figure 4: FMCW signal with parameters

The parameters shown in Fig. 4 define the limits of the FMCW radar. The bandwidth B defines the range of frequencies occupied by the transmitted signal starting from f_0 . A sawtooth waveform contains M_s samples of duration T_s , and a signal of duration T_c , called a *chirp*, consists of one sawtooth ramp plus a constant-frequency idle time. This idle time prevents inter-chirp mixing at the receiver and eliminates transient artifacts arising from the oscillator's abrupt frequency reset. The range is encoded in the short-time axis (within a single chirp of time $M_s T_s$), while velocity is encoded in the long-time axis (across M_c chirps, an acquisition of time $M_c T_c$), as the speed variation is producing a frequency shift too slow to be seen in one single chirp.

For a distance d_{target} between the target and the antennas, a time Δt is taken by the signal to propagate to the target and come back to the receiver antenna:

$$\Delta t = \frac{2 * d_{target}}{c} \quad (2)$$

With c the speed of light or of a radio frequency wave, equal to $3e8$. During this time Δt , the frequency of the transmitted signal has risen of:

$$\Delta f_{TX} = \frac{\Delta t}{MsTs} * B_{TX} \quad (3)$$

This equation is valid when the speed equals 0 and no Doppler effect is accounted. This is a valid hypothesis in a short time acquisition, when the speed variation is negligible. The received signal is compared with a LO signal, but this comparison is done at the harmonic frequency. The tag on the hornet emits a signal at twice the frequency transmitted, and the LO frequency is accordingly doubled in the RF circuit before comparison. The bandwidth B_{LO} is then the double of the fundamental one, and the difference in frequency measured at the output of the radar in the IF signal follows the same formula as in equation 3, with the bandwidth doubled.

$$f_{IF} = \frac{\Delta t}{MsTs} * 2B_{TX} \quad (4)$$

The range of the target is computed by combining equation 2 and 4.

$$d_{target} = \frac{f_{IF} c MsTs}{4B_{TX}} \quad (5)$$

The limits of the range measurement with FMCW signal are the resolution and the maximum range/speed measurable. The resolution of the frequency f_{IF} for a distance measurement, thus the resolution of an FFT computation on the short-time axis, is equal to $\frac{1}{MsTs}$, which transposed with the expression from the equation 4 gives a range resolution of:

$$\Delta d = \frac{f_{IF} * c * MsTs}{4B_{TX}} = \frac{c}{4B_{TX}} \quad (6)$$

For a speed measurement, the resolution of the FFT on the long-time axis becomes $\Delta f_{Doppler} = \frac{1}{McT_c}$, which gives the following speed resolution:

$$\Delta v = \frac{c}{4f_0 McT_c} \quad (7)$$

The maximum distance is limited by the Shannon-Nyquist theorem, that asserts that the maximum frequency measured cannot be higher than half of the sampling frequency to avoid aliasing. In this case, it means that $f_{IF,max} = \frac{1}{2Ts}$, transposed to the range it makes a maximum range of:

$$d_{max} = \frac{cMs}{8B_{TX}} \quad (8)$$

It is the same principle for the maximum speed limit, but on a longer sampling period of T_c . The maximum $\Delta f_{Doppler}$ is equal to $\frac{1}{2T_c}$, which makes a maximum speed of:

$$v_{max} = \frac{c}{4T_c f_{harmonic}} = \frac{c}{8T_c f_0} \quad (9)$$

Range Resolution	Speed Resolution	Maximum Range	Maximum Speed
$\frac{c}{4B_{TX}}$	$\frac{c}{4f_0 M_c T_c}$	$\frac{c * M_s}{8B_{TX}}$	$\frac{c}{8f_0 T_c}$

Table 1: Resolution and maximum measurable values of a FMCW harmonic radar system

Actually, the domains obtained after those computations could be improved. In the previous developments, the signal was considered as Real, with a symmetric spectrum in frequency. The measurable domains was limited to positive frequencies, i.e. $[0; d_{max}]$ for range and $[0; v_{max}]$ for velocity. If the signal was complex, the negative frequencies could be exploited, which would extend the domains. For this reason, most FMCW radar systems use the complex form of the received signal $I + jQ$ to improve the performances. This option will not be applied yet in the system, but it is an interesting feature to add in further researches.

2.III Choice of FMCW parameters

As said in the overviews of the radar and the tag, the bandwidth for the transmission signal is $[2.9; 2.975 \text{ GHz}]$ which gives a bandwidth of 75 MHz. This bandwidth implies a range resolution of **1 meter**, which is small enough for the application. The other parameters can be chosen following the model of a paper written with the New-Zealand Forest Research Institute developing a prototype of a FMCW harmonic radar [5]. The authors proposed a chirp time T_c of **1 ms** with a time of sawtooth $M_s T_s$ of **0.1 ms**. Estimating the maximum range needed for a Asian hornet track of 40 meters and a speed resolution of 0.2 m/s, it gives a sample duration of **1.25 μ s** and a number of chirps acquisition of **130**.

f_0 , GHz	2.9
Bandwidth B_{TX} , MHz	75
T_s , μ s	1.25
M_s	80
T_c , ms	1
M_c	130

Table 2: Example of transmission parameters for a FMCW harmonic radar system in a context of Asian hornet tracking

2.IV Post processing method and software

To work with an FMCW radar, it is common to generate a ramp voltage that, once injected in a voltage-controlled oscillator, is transformed into a ramp in frequency. The generation of such a signal and the recording of the output of the radar must be synchronized to be able to process the output correctly.

A software able to control the signal generation and signal measurement of a device plugged

into the radar to extract the range and speed components had already been created by Gilles Monnoyer de Galand de Carnières. This software was used to record the structured data at the output of the radar, and a python code was written to post-process the data and obtain the estimated range and speed. The software is called "Picoradar", it controls a device called a picoscope to generate the FMCW chirp with the asked parameters while measuring the IF signal at the output of the radar. Thanks to the software, both the source and the measurement of the picoscope are synchronized, to compute the FFT of a single chirp to extract the range component, and the FFT of M_c chirps to extract the speed component.

One frame of the signal recorded has the following matrix shape:

$$[M_c, M_{s2}]$$

With M_{s2} the number of samples in one whole chirp, the sawtooth wave-form and the idle time. Each element of the matrix is a sample of the received signal. One line of length M_{s2} corresponds to the signal received during one chirp, when the M_c lines correspond to the M_c chirps. The matrix of one frame must only contain the useful part of the signal, so the part with the ramp frequency. Each line of the matrix is cropped on the M_s first samples to only keep the useful samples.

$$x[i, j] = Matrix[i, j] \forall j \in [0, M_s]$$

The matrix, called x , has the shape shown in Fig. 5.

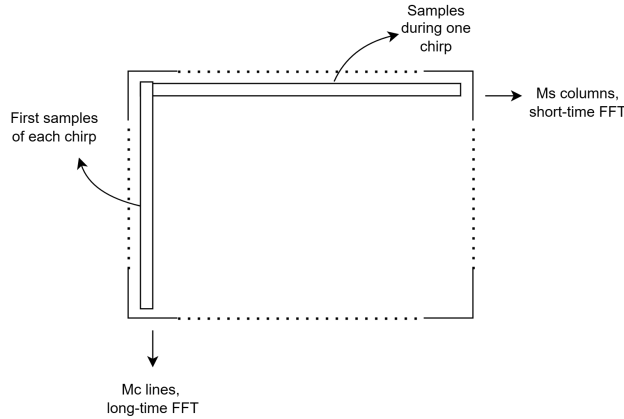


Figure 5: Structure of the data contained in a single radar output frame

To compute the short-time FFT, the FFT of each line of the matrix x is done before calculating the quadratic mean, as shown in these equations:

$$X_{range}[i, k] = \sum_{j=0}^{M_s-1} x[i, j] e^{\frac{2\pi k j}{M_s}} \quad (10)$$

$$S_{range}[k] = \sqrt{\sum_{i=0}^{M_c-1} |X[i, k]|^2} \quad (11)$$

And the long-time FFT formula is the following:

$$X_{speed}[k, j] = \sum_{i=0}^{M_c-1} x[i, j] e^{\frac{2\pi k i}{M_c}} \quad (12)$$

$$S_{speed}[k] = \sqrt{\sum_{j=0}^{M_s-1} |X[k, j]|^2} \quad (13)$$

The algorithm is thus a 1D FFT followed by a root mean square (RMS). The complete code of the computation is shown in the Appendix.

The processing gain of such an algorithm can be computed: for the range profile, it is an FFT on a signal of length M_s and a quadratic mean of M_c measurements. It gives a processing gain, in dB, of $10\log(M_s) + 10\log(\sqrt{M_c})$ [11]. The parameters are exchanged for the speed-profile.

A Range-Doppler map can also be displayed: a 2D FFT shown in a graph with range on one axis and velocity on the other, using a color scale to represent the intensity of detected targets. An example of a Range-Doppler map is shown in Fig. 6.

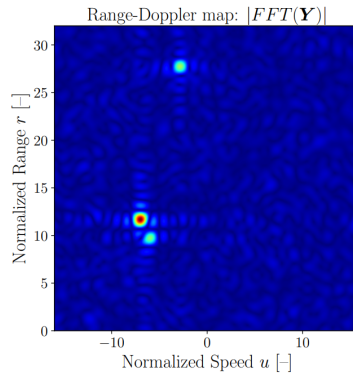


Figure 6: Example of a Range-Doppler Map [2]

Some post-processing methods can improve the quality of the results:

- **DC removal on the range profile (short-time FFT).** The DC short-time removal has to be done for several reasons. It is common for a Radio Frequency radar to have a

DC offset at the output. It comes from the RF components, from the coupling between the components on the board, or from the coupling between the antennas. The different sources are explored in this thesis, and they usually can not be completely erased. By removing the short-time DC component, those DC offsets are attenuated, but it also implies to delete the potential targets that would be at a null distance of the radar. In this application, it is not a problem, so this method can be used.

- **DC removal on the speed profile (long-time FFT).** This feature is applied for the same reasons as the DC short-time removal, with a stronger effect. It also deletes every false targets that has a null speed, so this feature cannot be used in the future for the system. Indeed, as the radar would be moving on a UAV at a speed close to the Asian hornet's speed, the risk to hide the hornet's signal because of the null-speed targets deletion is too high.
- **Background noise removal.** By recording the output signal when there are not any target in front of the antennas, it enables to subtract this background to the measured signal to delete the results that are not caused by the target's range and speed measurement, such as environmental reflections or static hardware artifacts. This feature does not affect the measurements and can therefore be used in the processing of the radar output.

Chapter 3

Study of the FMCW harmonic radar system

3.I Advantages of a FMCW harmonic radar system

The study of a system with a FMCW harmonic radar and a passive tag has been the subject of several papers. It enables to get a trade-off between the three criteria presented in the introduction, i.e. the shape of the tag, the maximum detectable range and the frequency bandwidth. Tags for harmonic radar systems have already been designed with a low-weight and a low-cost, and their size can vary between the applications.

The detection range depends of the efficiency of the system, and the estimation necessitates to compute a link budget on the chain, as it is developed in the next section. A potential range of 40 meters was estimated with the prototype of a FMCW harmonic radar in the previous paper [5]. To reach this range, the biggest challenge in this system is the power of the received signal. A sufficient power received should give a SNR high enough at the reception for the algorithm of detection to work. The Signal to Noise Ratio (SNR) is the ratio between the power of the useful signal and the power of the noise. A useful signal with a power lower than the noise cannot be interpreted, and the target cannot be located. In this thesis, the minimum SNR considered after the processing chain is 0 dB. With a lower SNR, the target cannot be located.

Finally, the frequency chosen are the same as in the previous works. As the aim of this chapter is to combine the radar and the tag designed by Hugo Damme ([12], 2026) and Frédéric Druppel ([3], 2026) in their Master thesis, the same frequencies are chosen, hence 2.9-2.975 GHz for the transmission and 5.8-5.95 GHz for the reception.

In this chapter, the radar and the tag designed by Hugo Damme and Frédéric Druppel are then studied and their performances will be measured. The starting point of the assembly of the system is to know the behavior and the limitations of the components already designed. After that, the determining points of the received power are established by studying the link budget computation of a FMCW harmonic radar system

3.II Overview of the tag of the harmonic radar tracking system

Frédéric Druppel studied the tag of a harmonic radar tracking system in his master thesis [3]. It describes a new tag design, composed of a diode and an inductance but carefully manufactured to get a conversion efficiency as high as possible. A Conversion Efficiency (CE) is the difference between the power transmitted by the tag at the harmonic frequency and the power of the signal receives by the tag at the fundamental frequency. A harmonic tag has the specificity to emit a signal with a power proportional to the square of its received signal. The best tag produced in this thesis is shown in Fig. 7.

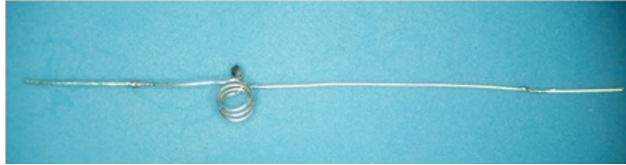


Figure 7: Tag used in the system, from the Master Thesis of Frédéric Druppel [3]

This tag is made of a diode and coupled inductors to increase the non-linearity of the diode and the power of the harmonic transmitted. The conversion efficiency of this tag was computed to have a knowledge of the power loss in the tag according to its received power. Though, to get the function $CE(PowerReceivedTag)$, an estimation of the power loss on a wave propagation must be made. The equation of free-space path loss is used, given for a distance d :

$$Path_loss(d)[dB] = 10 * \log\left(\left(\frac{4\pi d}{\lambda}\right)^2\right) \quad (14)$$

With λ the wave-length of the wave. This estimation must then be verified to have a valid conversion efficiency estimation.

3.II.A Method

The tag was placed in an anechoic chamber: a highly specialized room designed to completely absorb reflections of either sound or electromagnetic waves. This room enables to only measure the signal emitted by the tag without interferences. The tag was in a vertical position facing two horn antennas WBH218HS, one transmitting at 2.9 GHz and the other receiving at 5.8 GHz. The gain of those antennas is 13 dBi at 2.9 GHz and 17 dBi at 5.8 GHz. There were 3 meters between the tag and each antenna, and 90 cm between the two antennas. The transmitted signal was controlled by a N5183B Signal Generator and the measurement of the power received was made with a N9042B Spectrum Analyzer. GPIB and USB connections were used to synchronize the two devices and to record the measurement on a computer using the python library "pyvisa".

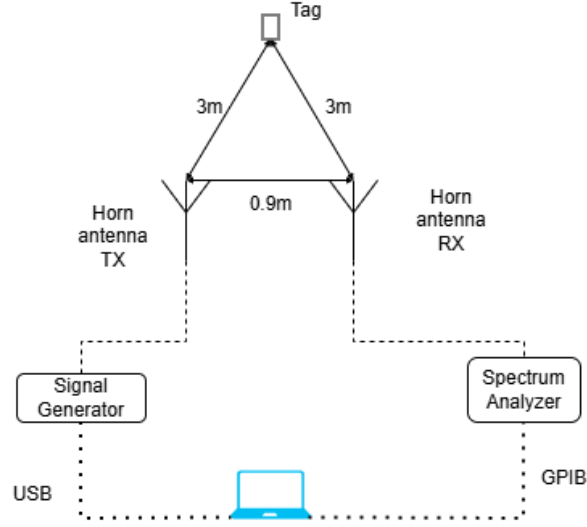


Figure 8: Connection schematic for the tag conversion efficiency measurement

To obtain the conversion efficiency of the tag according to the power received, the power transmitted by the signal generator was swept from 6 dBm to 19 dBm. The following formula was used to deduce the Conversion Efficiency from the power received at the Spectrum Analyzer.

$$P_{emitted_tag}[dB] = P_{received_SA}[dBm] - Gain_{RX_antenna}[dBi] + Path_Loss_{harmonic}(d)[dB] - 30 \quad (15)$$

$$P_{received_tag}[dB] = P_{emitted_SG}[dBm] + Gain_{TX_antenna}[dBi] - Path_Loss_{fundamental}(d)[dB] - 30 \quad (16)$$

$$CE[dB] = P_{emitted_tag}[dB] - P_{received_tag}[dB] \quad (17)$$

To check the validity of the path loss estimation, the same setup was used, for a constant power generated of 19 dBm and three different distances between the antennas and the tag: [1.8, 2.43, 3] meters.

3.II.B Results

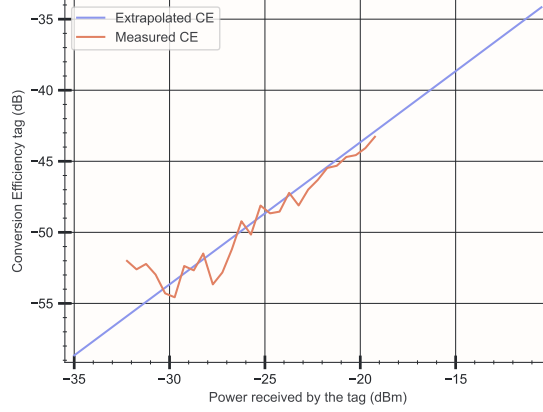


Figure 9: Conversion efficiency of the tag according to the power received by the tag

The graph in Fig. 9 represents the loss of power between the signal transmitted by the tag at 5.8 GHz and the power it received at 2.9 GHz. As this efficiency depends on the pins, the diode and the inductor, it is simpler in this thesis to consider the tag as a "black box" that emits a signal with a certain loss according to its received power. An extrapolation has been made for a bigger range of power received by the tag based on the hypothesis that a tag made of a diode and an inductance has a power output at the harmonic frequency proportional to the square of the power received at the fundamental frequency. In a logarithmic scale, it means that the slope of the line is 2. Given this slope, the extrapolation was fixed with the least squares method.

The reliability of the free-space path loss formula has been verified comparing the power received after going through the tag for different distances, and the expected power that should be received according to the estimation of the Conversion Efficiency of the tag based on the free-space path loss equation. This comparison is shown in Fig. 10. The measurements are close to the estimation, which proves that the Conversion Efficiency computation is reliable.

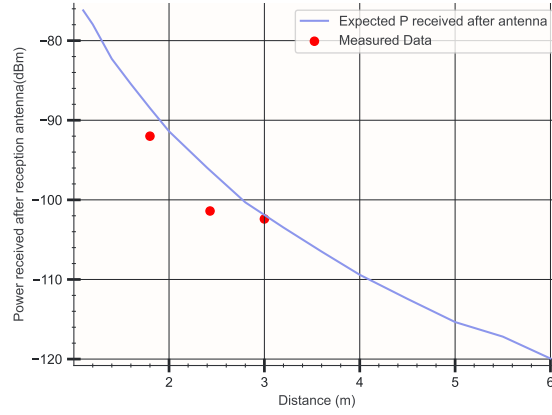


Figure 10: Path Loss validity

3.III Overview of the first version of the harmonic radar

The schematic of the first version of the radar designed by Hugo Damme is shown in Fig. 11.

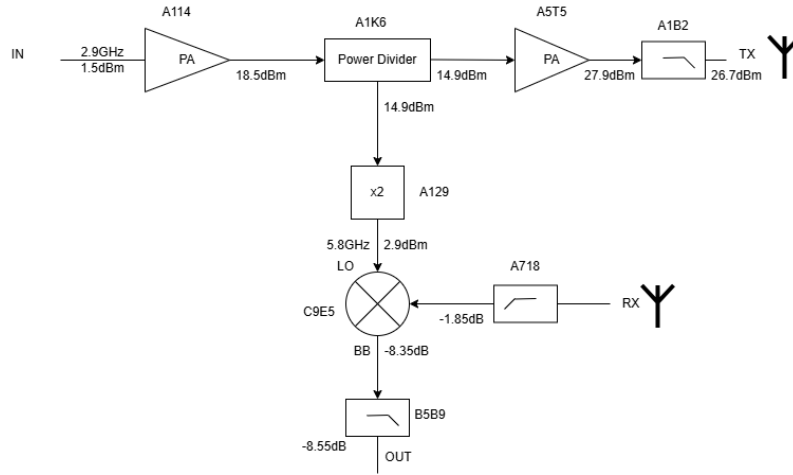


Figure 11: Schematic of the radar with expected power / power gain at each output of the blocks, designed by Hugo Damme

This radar was assembled with the Quantic X- microwave technology, that enables to get a compact and low-loss system. The code associated to each RF component corresponds to the name of the element in the Quantic X-Microwave products list [13]. The represented gains are sourced from the data sheets, though actual gain values may exhibit slight deviations. The radar works with a signal input of 1.5 dBm at 2.9 GHz, that is amplified and goes through a power divider to be on one side transmitted through an antenna (signal TX) and on the other side given as the reference signal to the reception chain(signal LO). The LO signal goes through a frequency multiplier to obtain the harmonic of the transmitted signal. Then, in the reception chain, the LO signal and the received signal at the RX antenna are mixed to get

the difference in frequency between LO and RX. This operating mode is intended for FMCW technology, but this version of the radar currently does not transmit a FMCW signal. At this point, it is expected from this circuit to efficiently transmit the signal, and to get an output representing the difference in frequency between RX and TX.

A power supply is needed for the two amplifiers of this radar. They are both supplied by an LDO regulator to provide a steady voltage, and those LDOs are supplied with a 7V voltage source.

An antenna was also designed by Hugo Damme. It was named a "cake antenna" and it is the layering of two hat antennas, shown in Fig. 12.

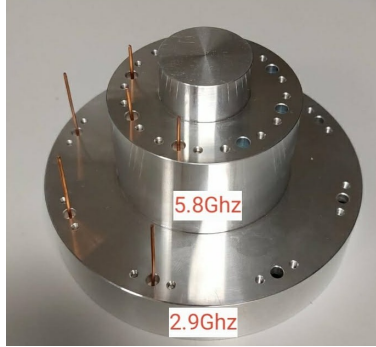


Figure 12: Cake antenna designed by Hugo Damme

The lowest and biggest part of the antenna is the emission part at 2.9 GHz, and the highest and smallest part is for the reception at 5.8 GHz. The reflection parameter of the antenna is expected to be low enough to transmit a sufficient part of the power of the captured signal.

3.III.A Method

To observe the TX output, a N5242B Vector Network Analyzer (VNA) was used to sweep the power of the input signal at PIN from -20 dBm to 1.5 dBm and to measure the power TX at the 2.9 GHz frequency. When observing TX, 50 Ω loads were put at the RX connector and the IF connector to avoid reflection or losses.

As explained in the presentation of the radar, the proper operation of the output signal can be verified by comparing the frequencies transmitted and received. To observe the IF output, the same VNA was used as a constant 1.5 dBm signal source at 2.9 GHz in the input IN, a sweep power source at 5.81 GHz in the input RX, sweeping from -60 dBm to -30 dBm, and a measurement of the output OUT at 10MHz. 10 MHz is the difference in frequency of the received signal (5.81 GHz) and the harmonic of the transmitted signal carried in LO (5.8 GHz). This frequency is high for a FMCW application, but this was the lowest frequency that the VNA could measure, and it is low enough for the low pass filter in the reception chain to not decrease the power from more than 1 dB according to its data sheet.

Fig. 13 shows the connections for TX and IF measurements. To get relevant results, the sources were calibrated to include the losses in the cables. Then the power losses in the measurement cable were measured to add them in the results computations.

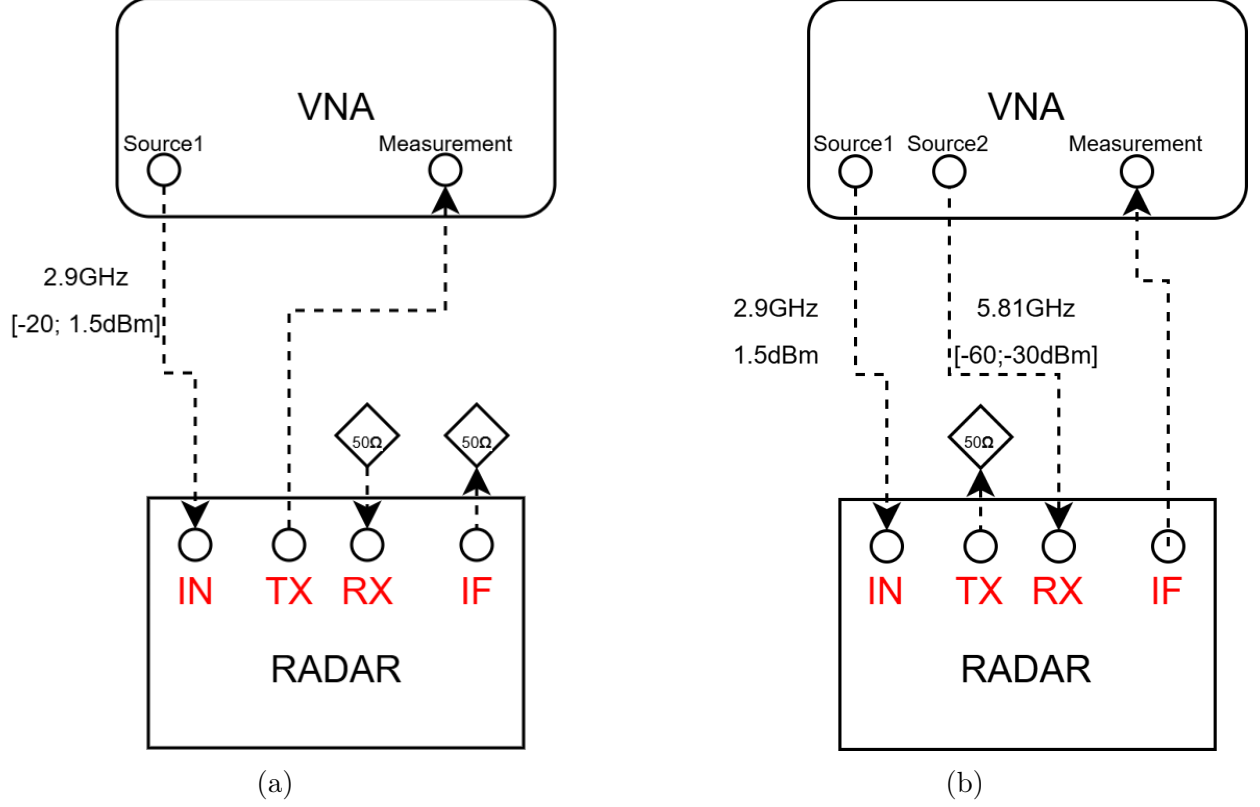


Figure 13: (a) Connection schematic for TX measurement with a VNA. (b) Connection schematic for IF measurement with a VNA.

A N9917A VNA FieldFox was used to measure the S11 parameter of both parts of the antenna. S-parameter measurement requires prior VNA calibration, in which the cable is included and the antenna port is set as the reference plane. Once it is done, the port of the VNA is connected at the input of the antenna to compute the S11 parameter.

3.III.B Results

Fig. 14 shows the power transmitted by the radar according to the power IN. At 1.5 dBm, the optimal power at the input, the magnitude of TX can be seen starting to stop increasing. It is because of the power amplifier after the power divider, A5T5, that has a P1dB of 28 dBm. It means that if the power amplifier has 28 dBm of output, the gain given at the signal is 1 dB lower than the maximum gain. That is the limitation of amplifiers, after that point non-linearities and a lower gain start to be observed. The output magnitude is 29 dBm for 1.5 dBm at the input, which is close to the expected TX power. In practice, higher transmit power levels would be legally permissible, as 2.9 GHz falls within a frequency bandwidth allocated to radio localization and military radars [14].

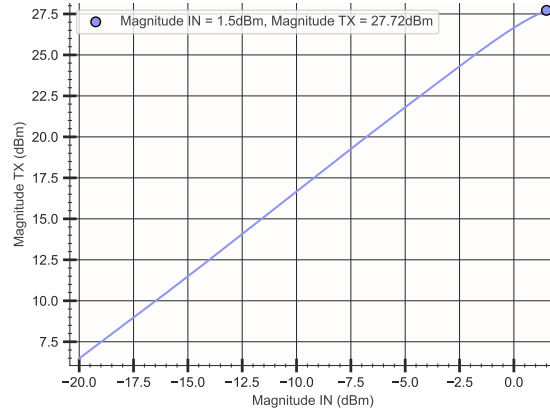


Figure 14: TX power with a sweep power on signal IN

The IF magnitude at the output of the system is shown in Fig. 15. The choice for the RX power was made according to the expected RX power in practice and the measurement noise of the VNA. At the lowest power inputs, the curve is more noisy than at higher power, because the magnitude of the useful signal is close to the noise level, so -60 dBm was the lowest TX power giving reliable measurements. Highest power than -30 dBm is useless as it is not expected to receive a power that high because of the transmission power, the path losses and the conversion efficiency of the tag.

The average loss in the reception chain is -8.84 dB, which is close to the expected value of -8.55 dB, demonstrating good agreement between measurement and expectation.

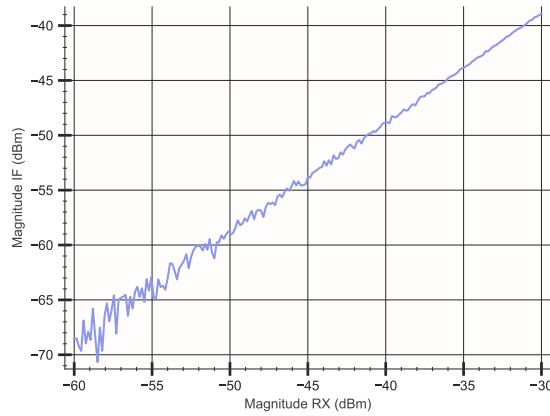


Figure 15: Magnitude of the output IF with 1.5 dBm power IN and sweep power RX

In this case of full-working system, the current consumption is **0.47 Ampere** which, with the 7V voltage source, gives a consumption of **3.29 Watts**.

Finally, the matching of the antenna is essential for a good operation of the radar. If there is some power reflection into the antenna, the radar cannot operate at full capacity.

The S11 parameters for the two parts of the antenna were computed at the corresponding frequency. A good S11 parameter is very low, around -10 dB, to deliver a good part of the available power through the antenna [15]. Those parameters are shown in Fig. 16.

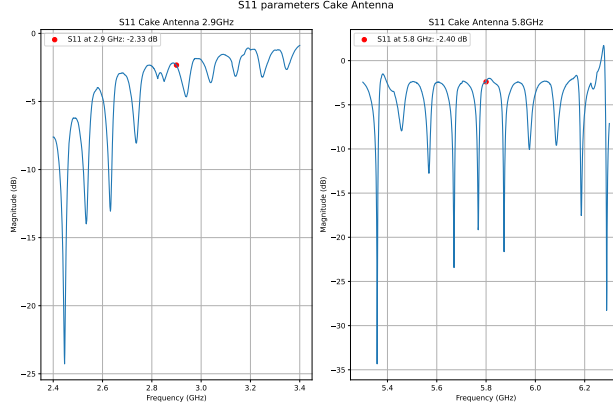


Figure 16: S11 parameters of the cake antenna for each frequency

The Fig. shows bad S11 parameters at both frequencies. It even shows that it does not seem matched at those frequencies, as a matching would induce a dip in the S11 parameter curve around the frequency. This result is consistent with the thesis of Hugo Damme, who mentioned that the antenna was not matched at 50Ω at the needed frequencies. To improve the system, other antennas must be used to avoid decreasing the performances.

3.IV Theory of the link budget computation

A link budget is the estimation of the power received at the output of the radar while adding every losses and gain in the chain. It is essential to compute it to be able to estimate the maximum range of the system, and to know if the power received is high enough for the processing algorithm to be able to locate the target.

For a FMCW harmonic radar, the gains and losses are the following:

- **The transmitted power:** The transmitted power is the most important step to obtain a good SNR at the end of the chain. Harmonic tags have the particularity of emitting a power proportional to the square of their received power, which means that doubling the transmitted power increases the received power by a factor of four. This transmitted power depends entirely on the hardware and the power amplifiers. However, power amplifiers have limited output power before exhibiting non-linearities, making it not straightforward to increase the transmitted power.
- **TX Antenna gain**
- **Path Loss forward:** The same equation of the free space path loss as in the overview of the tag is used with the fundamental frequency. This loss is proportional to the square of the distance.

- **Conversion Loss of the tag:** It was improved by the previous works, further progress are not studied in this thesis.
- **Path loss backward:** It is the same principle as the path loss forward but at the harmonic frequency, which is two times bigger. The path loss is proportional to the square of the frequency, so this loss is two times bigger than the forward loss, in dB.
- **RX Antenna gain**
- **Receiver chain gain:** Amplifying the received signal is a straightforward way to increase the signal level at the processing stage. It is less efficient than increasing the emitted power but easier, as adding a low noise amplifier would increase the SNR, and there are less risks to exceed the maximum output power of the amplifier as the received power is low.
- **Processing gain:** As presented in the FMCW theory, the processing method is an FFT of the output signal. Depending of the representation: Range profile, Speed profile or a Range-Doppler map, this processing gain varies.

Chapter 4

Improvements and characterization of the FMCW harmonic radar system

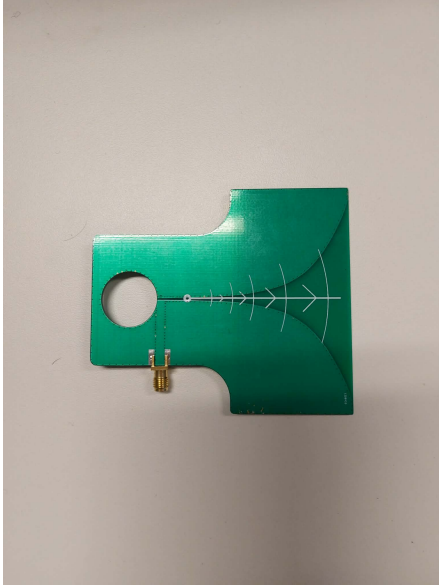
To complete the FMCW harmonic radar system from the existing hardware, several improvements and additional functionalities are required. This chapter covers the steps taken to bring the system to a fully operational and characterized state.

As explained in the first overview of the system, the antenna originally designed in the master thesis of Hugo Damme is not efficient and cannot be utilized. Thus the first elements developed in this section are the replacement antennas.

The radar also needs to be improved and completed for the tracking system to work. The possibility to transmit an FMCW signal must be added, and some improvements and measurements must be done to be able to obtain tracking results.

4.I Antennas selection

The only specification needed for the antennas were to be matched at 2.9 GHz and 5.8 GHz. The Ultra Wide Band (UWB) "Vivaldi Antennas" were used, purchased on Amazon [16], as shown in Fig. 17(a). They are supposed to be adapted to transmit and receive a signal between 1.4 and 10.5 GHz with a gain of 7 dBi. Those antennas are compact but efficient, and can be adapted to hang on an UAV. For example, a 3D printing has been done to comfortably use the antennas with a fixed position, shown in Fig. 17(b). A distance of 17 cm separates the antennas, and the prototype could be hanged to an UAV for further experiences.



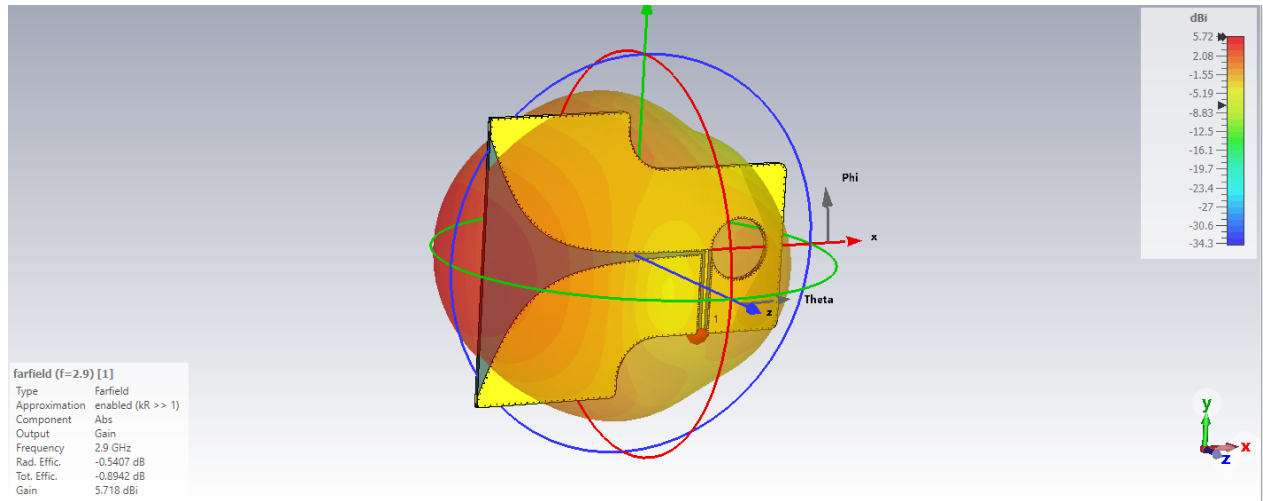
(a)



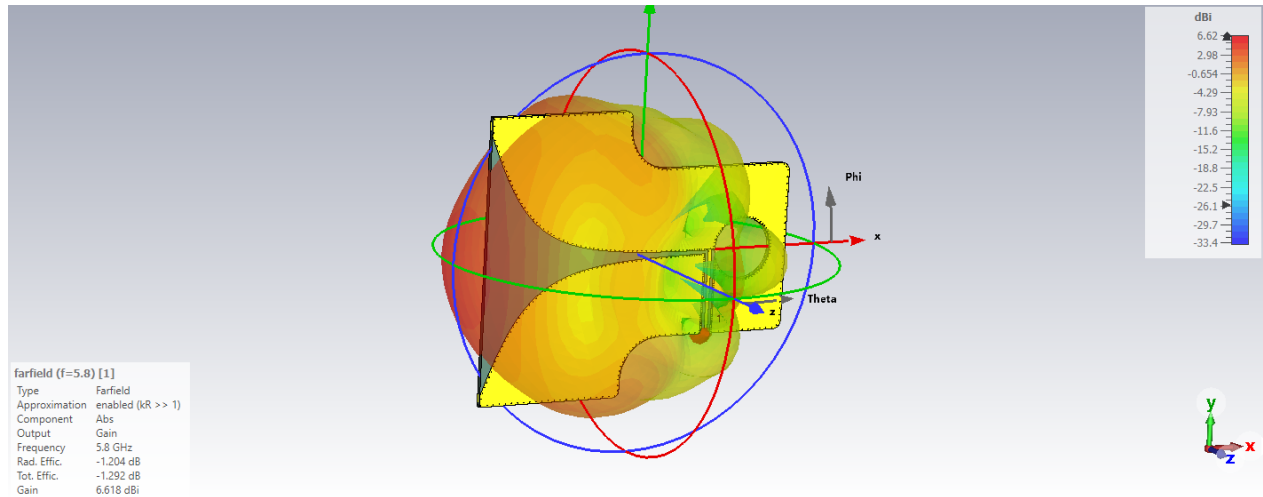
(b)

Figure 17: (a) Vivaldi Antenna. (b) 3D printing that supports the two antennas.

To get a complete knowledge about the antennas, it would have been ideal to measure the complete radiation pattern of the antenna. It consists in a series of measurement in laboratory done in an anechoic chamber where the antenna is put on a turning table with a known-characteristics antenna in front of it. For a signal generated and transmitted by the Antenna Under Test, the reference antenna captures what is transmitted which gives the gain coefficient of the Antenna Under Test for every direction of the turning table. Unfortunately, due to the computational and storage demands of this process, practical limitations were encountered in the laboratory and it has not been possible to do so. Instead, the modeling of the antenna has been done and the far-field gains have been simulated on the software *CST simulation*, which gave an estimation of the real behavior of the antenna. 3D pictures of the CST simulation with the far-field gain at 2.9 GHz and 5.8 GHz are shown in Fig. 18, while the cut planes are given in Fig. 19 and 20.



(a)



(b)

Figure 18: Modeling of the far-field gain of the Vivaldi antenna at (a) 2.9 GHz and (b) 5.8 GHz

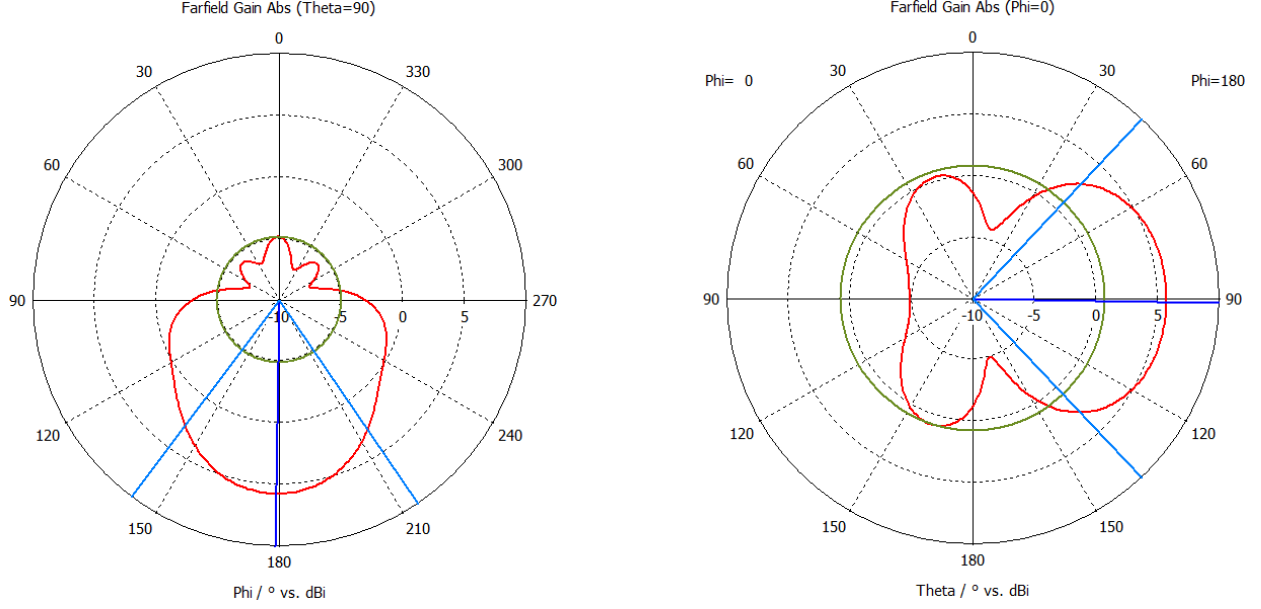


Figure 19: Simulation results of the far-field gain of the Vivaldi antenna at 2.9 GHz

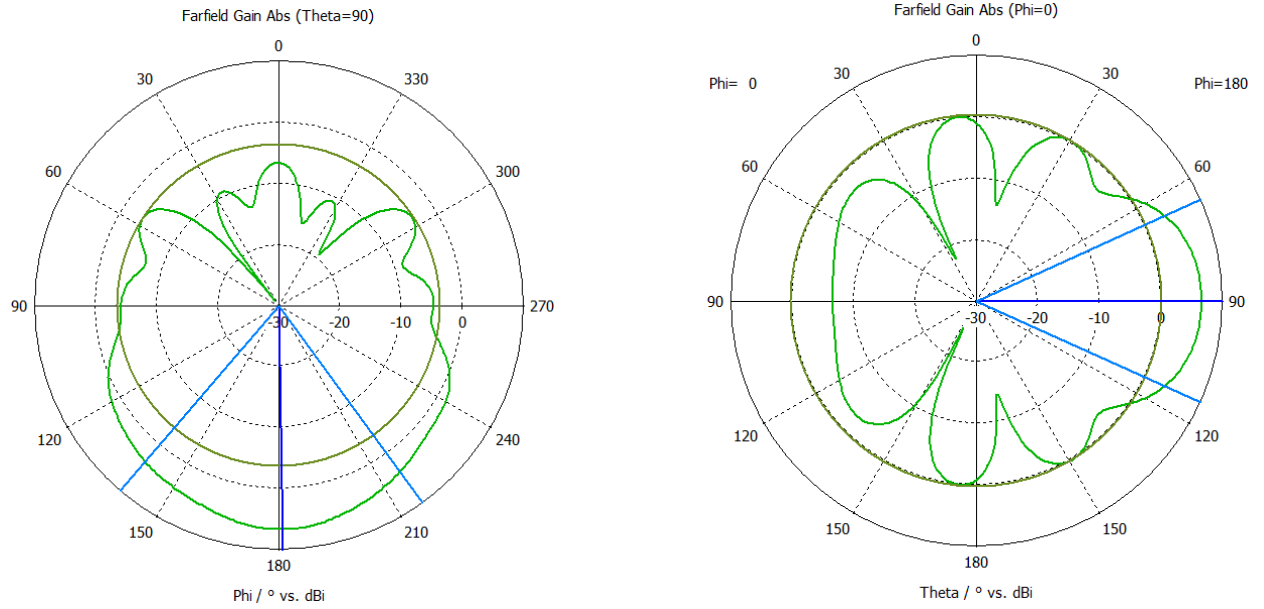


Figure 20: Simulation results of the far-field gain of the Vivaldi antenna at 5.8 GHz

Those results give the gain of the main lobe but also its half-power beam-width which reflects the coverage area of the antenna.

Frequency [GHz]	Gain [dBi]	Vertical half-power beam-width [°]	Horizontal half-power beam-width [°]
2.9	5.72	71	93.5
5.8	6.62	76.6	48.6

Table 3: Estimated characteristics of the antenna

The gain results are satisfying for a small antenna adapted to an embedded circuit. They are close to gains of patch antennas that are often used in those scenarios. To obtain a higher gain, a trade-off between the directivity of the antenna and the gain should be made, which means get a smaller coverage area for a higher gain. It is not necessarily a good option, the tracking system operates on a short range, so the half-power beam-width must be large enough to not have a problematic blind spot. The horizontal half-power beam-width at 5.8 GHz can be noticed as much smaller than at 2.9 GHz. A 48.6° horizontal beam-width might not be sufficient for the application. Another antenna for the 5.8 GHz could be found with a better coverage area.

The S11 parameter of the antenna must also be sufficiently low as explained in the study of the FMCW harmonic radar system. Its simulation is shown in Fig. 21. As mentioned, a S11 parameter lower than -10 dB is typically a good performance for an antenna. The results show that this antenna is indeed an ultra wide band antenna, with a low S11 parameter for frequencies between 1.4 GHz and 10 GHz, and lower than -10 dB at the two working frequencies.

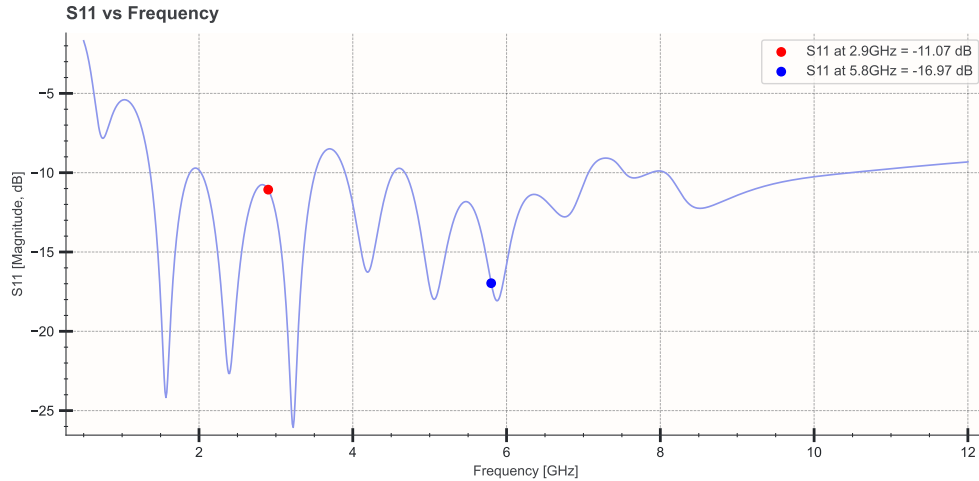


Figure 21: Simulation of the S11 parameter of the Vivaldi antenna

Ultimately, the antennas performances are satisfying for the application and they can be used for the remainder of this thesis. For further improvements, it could be interesting to choose an antenna optimized only for a 2.9 GHz frequency at the transmission, and only for 5.8 GHz at the reception. As the efficiency of the system can decrease sorely with the parasitic harmonic transmitted and the parasitic fundamental received, it would probably improve the performances. The impact of the parasitic signals is developed later in this thesis.

4.II Radar improvements

4.II.A Low Noise Amplifier in the reception chain

As seen in the overview, the reception chain does not yet include any amplification stage. Amplifying the received signal is essential, however the choice of amplifier is critical: a standard amplifier would introduce significant noise, degrading the SNR at the radar output. A Low Noise Amplifier (LNA) is therefore required, as it is specifically designed to amplify weak signals while adding minimal noise to the chain. The LNA chosen in the Quantic X-Microwave is the XM-A9T6-0404D, an amplifier with a gain of 20 dB and a noise figure around 2 dB between 50 MHz and 8000 MHz. It is then a relevant choice of LNA to amplify the 5.8 GHz signal in the reception chain. Fig. 22 shows the reception chain schematic with the added LNA that is supplied with a LDO XM-A2J9-0404D, the cumulative signal gain is reported at each stage of the receiver chain.

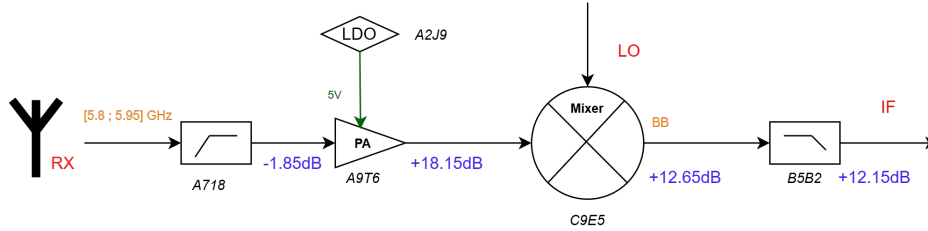


Figure 22: Reception chain of the radar with a A976 Power Amplifier

The overall NF of this receiver chain can be computed with the noise figure and the gain of each blocks. Passive elements have a noise figure equal to their conversion loss, which means it is equal to the opposite of their gain G in dB.

RF elements	Noise Figure[dB]	Gain [dB]
Filter A718	1.85	-1.85
LNA A976	2	20
Mixer C9E5	6.5	-6.5
B5B2	0.2	-0.2

The computation of the noise figure of a chain of n blocks is made with the following formula, with F_n the linear value of the noise figure of the n^{th} block ($10^{NF_n/10}$) and its gain G_n in a linear form too.

$$F_{tot} = F_1 + \frac{F_2 - 1}{G_1} + \dots + \frac{F_n - 1}{G_1 * G_2 * \dots * G_{n-1}} \quad (18)$$

The RX chain achieves a total of **3.95 dB** of noise figure (NF when expressed in dB) with a total gain of **12.15 dB**.

Fig. 23 checks the IF output magnitude according to the RX power at the input with the same method as in the radar overview. A VNA sweeps the input RX power at 5.81 GHz

and measures the output IF at 10 MHz which is the difference between the harmonic of TX (5.8 GHz) and RX. A comparison with the previous results without the LNA is also put on the graph. The mean gain is 12.19 dB, which is close to the expected RX chain gain and satisfying. The comparison with the previous results shows that the gain has improved of 21.03 dB.

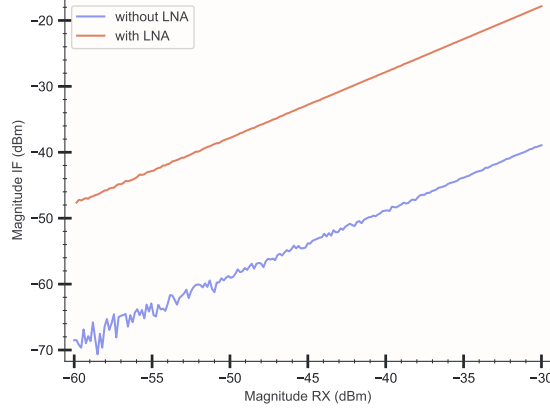


Figure 23: Magnitude of the IF output of the radar according to the RX signal magnitude

4.II.B Voltage Controlled Oscillator (VCO)

To obtain an FMCW radar, a VCO was added to generate the input signal. It is an electronic oscillator whose output frequency is controlled by an input control voltage. A linear sawtooth ramp is used as the control voltage of the VCO, it is the most common way to generate an FMCW signal. In the Quantic-X Microwave catalog, several VCO blocks were available. Because of the limited time, a VCO available in the provider stock was selected: the XM-C187-0809D. Its characteristics are the following:

Frequencies [GHz]	[2.12 ; 3.36]
Input Voltage [V]	[0.5 ; 18]
Pout typ [dBm]	9 dBm

Table 4: VCO C187 characteristics

The frequency bandwidth of interest ranges from 2.9 GHz to 2.975 GHz. There exists some VCO that would be more centered on the used bandwidth, which would give a better accuracy of the output and a lower input voltage. Indeed, as explained in Chapter 2 about the FMCW application, the goal is to use a picoscope to generate and measure the signals of the radar. The picoscope 3000 Series 3403D used in laboratory has a maximum output voltage of 2 V, which is too low to produce a 2.9 GHz signal with this VCO. A DC amplifier circuit has to be placed between the picoscope output and the VCO input to obtain the proper frequency, which adds imprecision to the system. However, this VCO has the benefit of having two output signals at the same frequency. This simplifies the circuit, as the power

divider is not needed anymore.

A new transmitting chain was designed for the radar to obtain the highest TX power while keeping the LO power in an optimum range for the performances of the mixer. The data sheet of the mixer shows that the best LO power is 4 dBm, and performance deteriorates sharply away from the optimum. For a LO power of 1 dBm or 7 dBm, the conversion loss and the isolation of the ports of the mixer become deficient. Fig. 24 shows the schematic of the transmitting circuit. The amplifier A114 has been placed in the LO branch, followed by an attenuator to get a LO power close to 4 dBm. The TX power lost almost 4 dB compared to the previous circuit, which could be improved by finding better power amplifiers, or by adding a low-gain amplifier after the VCO output in order for the A5T5 power amplifier to operate closer to its maximum linear output power, i.e., its P1dB of 28 dBm. The LDO powering the VCO requires a 14 V input to deliver a stable 12 V output. For simplicity, this 14 V supply was used for all other LDOs as well, rather than an other 7 V supply. It has no impact on their performance, as it does not exceed their maximum input voltage.

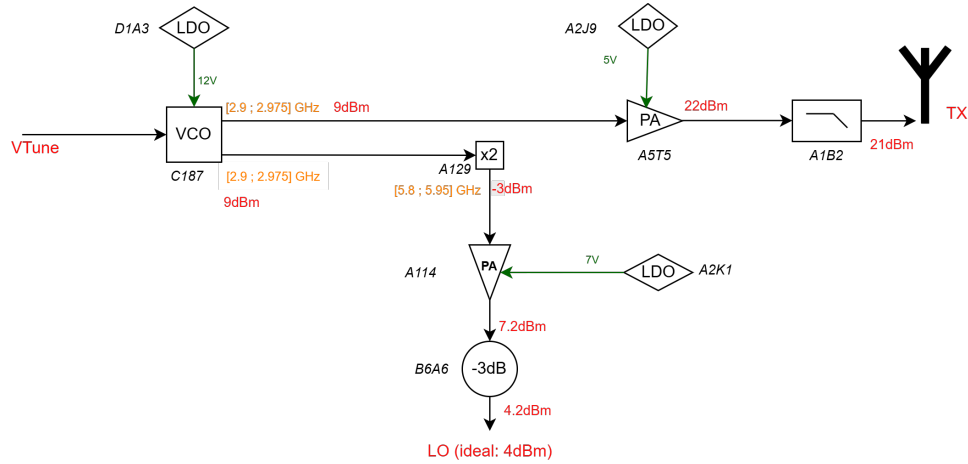


Figure 24: Transmitter circuit schematic of the radar after adapting to the VCO

Differential Amplifier and Tune voltage Management

As mentioned in the data sheet of the VCO, the tune voltage needed for an output of 2.9 GHz is supposed to be around 9V which is much higher than the 2 V picoscope output. To be able to use the picoscope to get a saw-tooth voltage ramp, a differential amplifier as used. The schematic of such an amplifier preceded by a follower amplifier is shown in Fig. 25. A follower amplifier was also added to prevent input voltage offsets because of the internal resistance of the picoscope. A follower amplifier preserves the integrity of the voltage thanks to the high input impedance of the amplifier and its low output impedance.

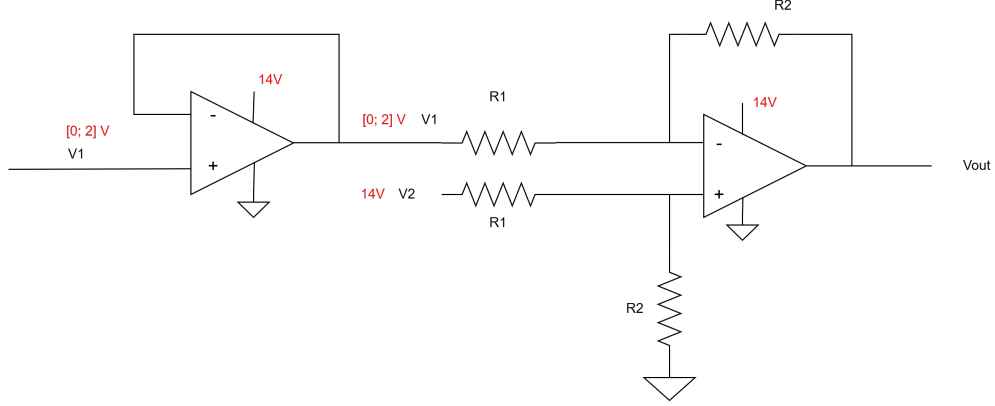


Figure 25: DC amplifier circuit made of a follower and a differential amplifier

The following equation describes the output voltage of the circuit:

$$V_{out} = \frac{R_2}{R_1} * (V_2 - V_1) = \frac{R_2}{R_1} * (14 - V_1) \quad (19)$$

Again, voltage supply of the amplifiers and the offset voltage V_2 in the circuit were fixed at 14V to simplify the setup and use the same supply than for the radar. An important point of this circuit is that it is an inverting amplifier, which leads to the inversion of the ramp. For an increasing voltage V_1 coming from the picoscope, the output voltage is decreasing.

To be able to adapt the resistance values to the non-idealities of the setup, potentiometer of 10 k Ω were used for the duets R_1 and R_2 , which involves that $R_1 + R_2 = 10k\Omega$. To choose their values, the equation (19) was used with the following characteristics of the VCO V_T (Tune Voltage) and with a margin considered for the picoscope. Even though it can generate a voltage from -2 V to 2 V, a safety margin on the maximum voltage $V_{max,pico}$ and on the minimum output voltage $V_{min,pico}$ is taken.

V_T VCO (2.9 GHz output)	V_T VCO (2.975 GHz output)	$V_{max,pico}$	$V_{min,pico}$
8.96 V	9.72 V	1.6 V	-1.6 V

Table 5: Equation parameters for the choice of R_1 and R_2

$$\begin{aligned}
8.96 &= \frac{R_2}{R_1}(14 - 1.6) \\
\hookrightarrow \frac{R_2}{R_1} &= 0.72 \\
9.72 &= 0.72(14 - V_{min}) \\
\hookrightarrow V_{min} &= 0.5V > V_{min,pico} \\
\frac{R_2}{(10 - R_2)} &= 0.72 \\
\hookrightarrow R_2 &= 4.19k\Omega \\
\hookrightarrow R_1 &= 5.81k\Omega
\end{aligned}$$

At first the circuit was mounted on a breadboard and supplied with alligator clips on banana jacks, which is not stable and trustful. On top of that the VCO behavior varies with temperature, as mentioned in its data sheet. So after configuring the potentiometers to obtain R1 and R2, they had to be slightly adapted to obtain the required output frequencies. For the future, to put the amplifier on a PCB and to keep a constant temperature for the radar would improve the stability of the system.

With the connections shown in the schematic shown in Fig. 26, the output voltage of the picoscope is amplified by the differential amplifier before going into the tuned voltage of the VCO. The transmitting circuit is functional.

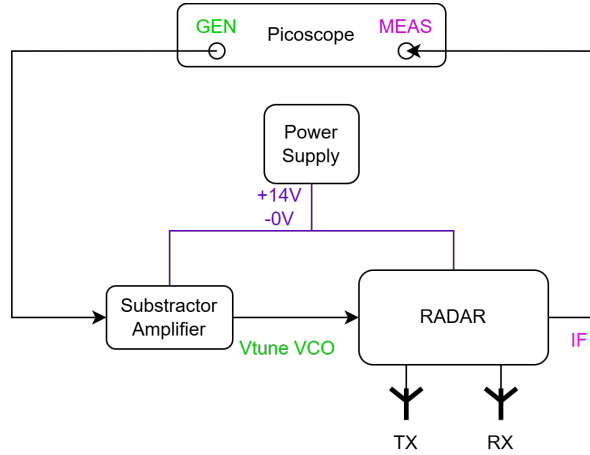


Figure 26: Connection schematic for the Tune Voltage management

A power budget is established to know the exact power levels at each stage of the transmitting circuit and verify that the measured values are consistent with theoretical predictions. It consists of assembling the circuit block after block and measuring the power output at each stage. This measurement showed that the frequency multiplier has lower losses than expected, 1.14 dB of difference. Though, because of the lack of space on the Quantic X-

Microwave board, a small and rigid cable was added between the multiplier and the amplifier, which causes a 0.36 dB loss, so the final LO power is still within an acceptable range to ensure proper operation of the mixer. The power budget is shown in Fig. 27. The transmitted power is higher of 1.85 dB to the one expected which is a good thing for the application.

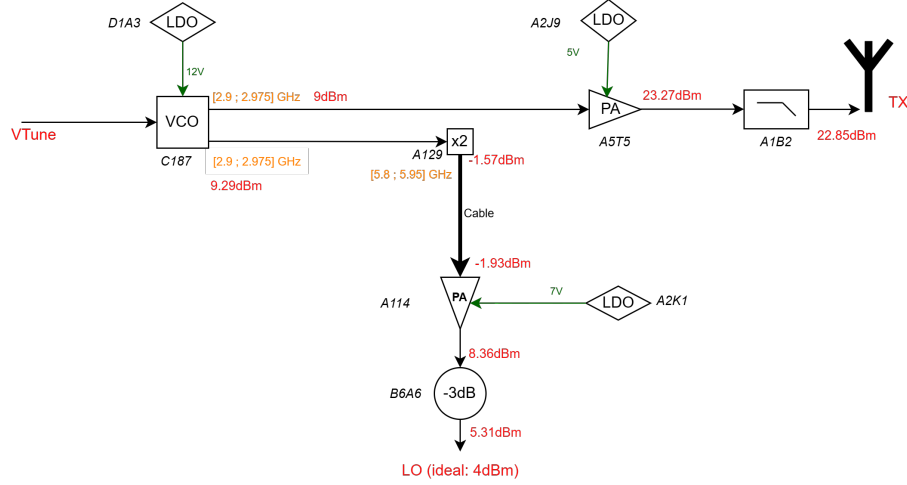


Figure 27: Power budget of the TX chain of the radar with the VCO

The LO signal spectrum is shown in Fig. 28 with a magnitude of 5.6 dBm which is close enough from the optimal magnitude of 4 dBm.

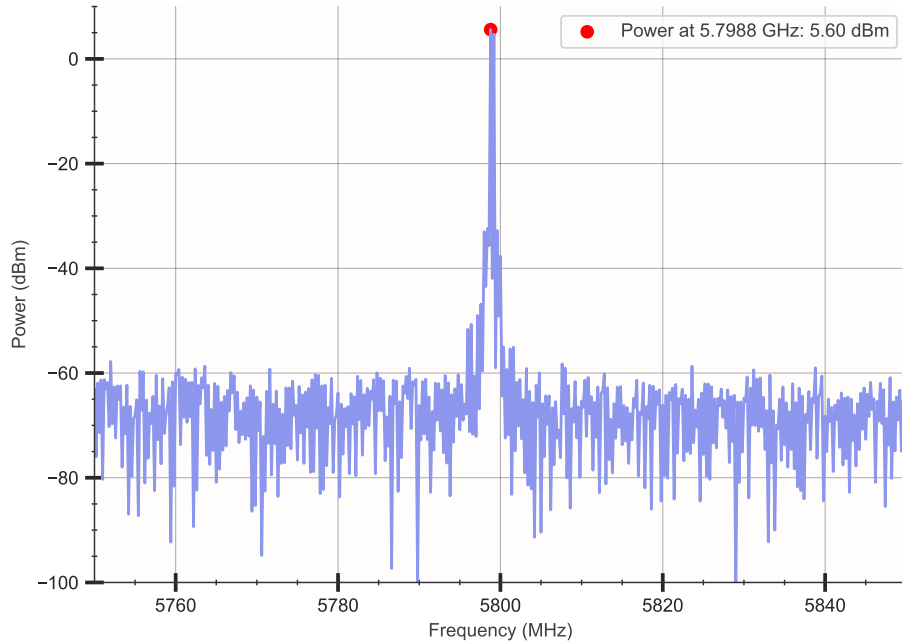


Figure 28: Spectrum LO signal at 5.8 GHz in the configuration with the VCO

VCO temperature variation

As mentioned in the data sheet of the VCO, significant frequency variations can be observed with increasing circuit board temperature. The frequency output of the VCO is extremely volatile in a range up to approximately 4 MHz. Depending on the temperature or the position of the supply cables, the frequency can vary. Fig. 29 shows a comparison of two TX signal measurements taken with the same setup but with a board temperature difference of approximately 10 degrees between the two measurements. The resulting frequency shift between the two temperatures is **3.6 MHz**. The temperature dependency could not be characterized..

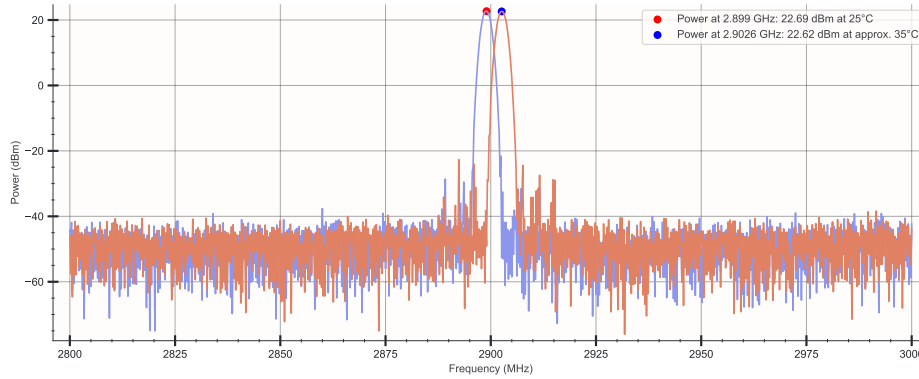


Figure 29: TX signal for two temperatures of the board, same setup

For this reason, for the remainder of this thesis, the frequencies of the signals shown are not exactly at 2.9 GHz or 5.8 GHz. The potential problems that could occur are reviewed in the next section about the curve non-linearity of the VCO.

VCO Tuning curve non-linearity

The efficiency of an FMCW radar is correlated to the linearity of its VCO. In the post processing of the IF signal, it is considered that the frequency ramp is perfectly linear with respect to the tuning voltage. The linearity was verified by recording, for several control voltage given to the VCO, the frequency of the signal at its output. This analysis was done connecting the system setup in the same way as presented earlier (setup shown in Fig. 26), and connecting a N9042B Spectrum Analyzer at the output RX and a picoscope 3000 Series to control the VCO. The voltage step was 0.05 V and the frequency was measured over 33 points with a resolution of 0.1 MHz. This is the minimum resolution of the spectrum analyzer. For those measurements, it can take some time to measure each frequency output according to the voltage control, and temperature variations could affect the results. The system was therefore powered on for a sufficient amount of time beforehand to allow it to warm up and minimize temperature differences during the measurement.

The purpose of this measurement is to assess whether the VCO gain remains linear across the tuning bandwidth, or exhibits slope variation. The slope variation is the common default

of VCOs, but is commonly observed on edges of the output bandwidth. As the VCO is employed in the center of its bandwidth, there are most probably not significant non-linearity. Fig. 30 shows in blue dots the measurements, and the red line is the least square approximation of the line. The linearity of the measurement is satisfying, the system can be employed considering the frequency ramp as linear. The mean absolute error between measurement and the approximation is **0.164 MHz**. This result is close to the resolution of the spectrum analyzer, thus it could come from the lack of precision of the measurement instead of the non-linearity of the VCO.

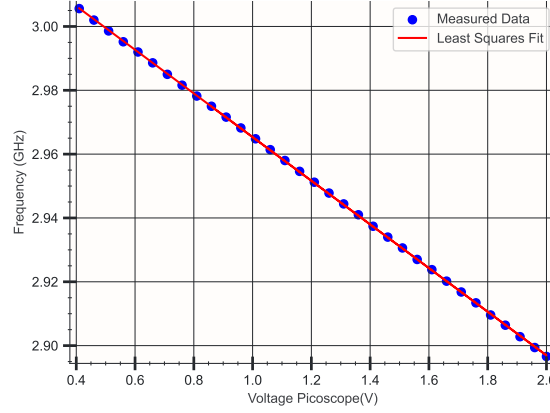


Figure 30: Linearity verification of the VCO

The least square approximation gave the equation of the line:

$$F_{VCO}[MHz] = -68.04 * (V_T[V]) + 3033.74$$

With F_{VCO} the frequency of the VCO output signal. This equation is used for the generation of the FMCW signal by the picoscope, and for the interpretation of the output signal.

Finally, the behavior variation of the VCO with the temperature could seem like a problem in the tracking operation. It can be considered that it does not affect the performances, as the temperature variation is slower than one acquisition. During one acquisition, the ramp can safely be considered perfectly linear. And when the frequency is different than the one expected in the Least Square Approximation, it should not lead to a problem in the tracking process, as the range evaluation is based on the difference in frequency between the signal emitted and received. As long as the frequency ramp is linear during one acquisition, the processing is accurate.

4.III Characterization

As a summary of the previous steps, the complete radar schematic at this step is shown in Fig. 31(a), with the connection schematic to supply and tune the radar in Fig. 31(b). This system must be characterized to reveal potential limitations and to configure a laboratory setup for measurements.

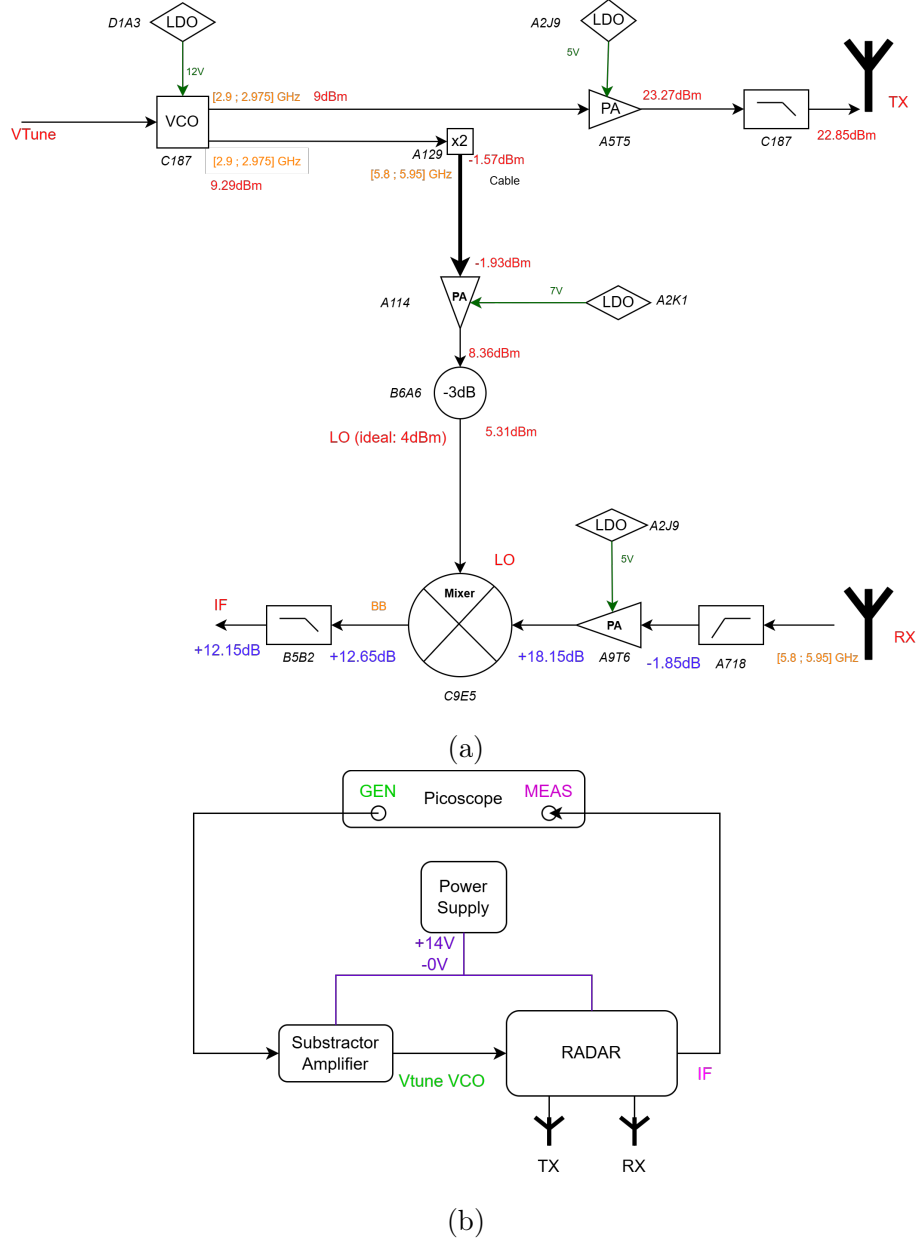


Figure 31: Setup after improvements. (a) Schematic of the improved radar chain, with added VCO, LNA and the power budget of the transmission chain (b) Connection schematic to supply and control the radar

4.III.A Harmonic content and signal integrity

The signal integrity of a harmonic radar is the key of an efficient and stable system. An important step of the characterization is to be aware of the non-idealities of the radar to interpret the potential problems that are going to be encountered, and to try and correct the non-idealities.

The transmitted signal must be in the band of 2.9 GHz and its harmonic component must be as low as possible. The stability of an FMCW harmonic radar depends on the integrity of the transmitted signal. The core purpose of using a harmonic radar, i.e to compute the localization at the harmonic frequency, is to avoid interferences with other targets than the tag of interest. If the transmitted signal has a harmonic component, it can be reflected by obstacles and be captured by the RX antenna, which causes interferences. For the same reason, the LO signal mixed to the RX signal in the reception chain must be in the 5.8 GHz band without any 2.9 GHz component. A 2.9 GHz LO component would mix with the 2.9 GHz received and it would cause interferences.

Finally, another parasitic signal is the one coming from the coupling between the components of the radar on the board. Indeed, it is common for a Radio Frequency circuit to have some coupling caused by the proximity between the RF elements. The problematic one would be the 2.9 GHz coming into the RX received signal. This non-ideality would rather cause a DC component, the mixing between two equal frequencies.

Even if the processing enables to correct the non-idealities of the radar by deleting the DC component or subtracting the background noise, those are not perfect processing steps and it does not correct entirely the results.

Method

First, the LO signal was measured by connecting the whole connection setup but disconnecting, in the radar, the LO output from the reception chain. A SMA connector was placed instead of the mixer to measure the LO signal with a N9042B Spectrum Analyzer, and the RX output was loaded by 50Ω . The oscilloscope was generating the tune voltage amplified by the Differential Amplifier to have a 5.8 GHz LO signal.

A high pass filter had to be added in the LO chain, its S21 parameter was measured to verify its efficiency in this case. Its S21 parameter shows the gain at the output of the filter according to the input frequency. It was measured with a N9917A VNA FieldFox, calibrated to get the S21 of the filter without including the cables connected to the VNA.

The transmitted signal harmonic component was measured while still connecting the whole setup, with 50Ω loads at the RX and IF connectors to avoid parasitic reflections. The N9042B Spectrum Analyzer was connected to the TX connector. The maximum input power of the SA is 30 dBm, and the expected power level of TX was around 23 dBm, thus no attenuator was needed.

Finally, to measure the coupling component in the received signal, the mixer was replaced by two SMA connectors, one on the RF entry to measure its power with the Spectrum Analyzer, and another on the LO entry to load it at 50Ω . A 50Ω load was put on the RX radar input to measure only the coupling component, and connected the rest of the setup to tune and

supply the radar.

The corresponding simplified schematics at the radar levels can be seen in Fig. 32(a) for the LO measurement, Fig. 32(b) for the RX measurement, and Fig. 32(c) for the coupling measurement. Those are simplified schematics, the arrows hide the different blocks that were showed in previous sections, as the aim of those schematics is to understand which signals were measured and the technique used.

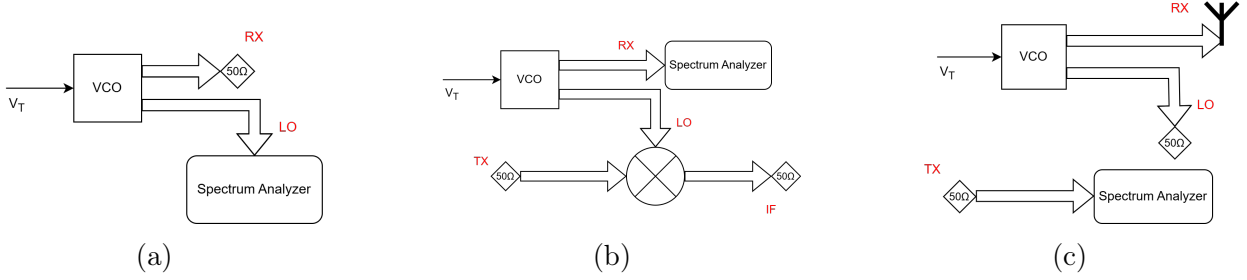


Figure 32: Simplified radar schematic for the impurity measurements. (a) LO measurement. (b) TX measurement. (c) Coupling at the RF entry of the mixer measurement.

Results

The LO measurement of the 2.9 GHz component gave the result in Fig. 33. The magnitude is of -14.7 dBm, in comparison the 5.8 GHz component of the LO signal has a magnitude of 5.31 dBm. This 2.9 GHz spurious component is significantly high relative to the expected RF input level at the mixer of around -80 dBm. The filter after the RX antenna does not completely erase the 2.9 GHz component, and the risk of mixing with the 2.9 GHz LO component is high. This default would make the radar see any reflective target, not only the tag emitting the 5.8 GHz signal.

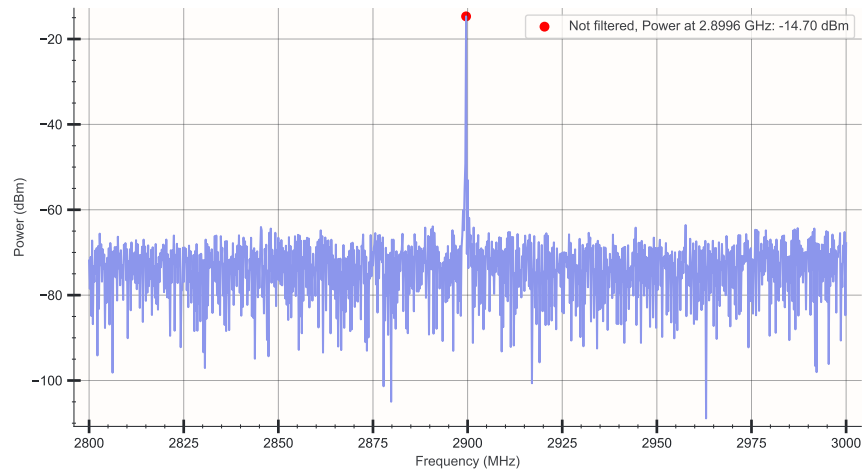


Figure 33: 2.9 GHz component of LO

This 2.9 GHz component can have multiple sources. First, the frequency multiplier used

in the LO chain decreases the fundamental frequency magnitude but do not erase it. Also, the coupling between the TX chain with a high 2.9 GHz power and the LO output can have a significant impact on the signal.

To decrease the sensitivity of the radar to other targets than the tag, a Mini-circuits VHF 3800+ high pass filter was added to the LO signal chain. The S21 parameter of the filter is shown in Fig. 34. It should decrease the 2.9 GHz magnitude of 30.65 dB while decreasing the 5.8 GHz component magnitude of only 0.5 dB.

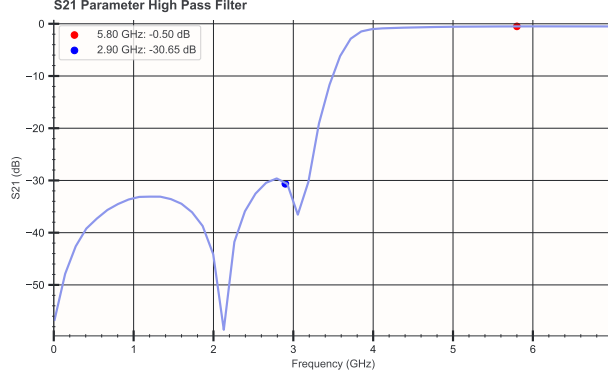


Figure 34: S21 parameter of the high pass filter added in the LO chain

The filtered LO component at 2.9 GHz is shown in Fig. 35. It has been decreased of 31.52 dB, which is close to the expected attenuation given by the S21 parameter of the filter. That is a satisfying attenuation for a filter, but a more integrated solution, such as an X-Microwave filter mounted directly on the board, would be preferable in a future design. The schematic of the transmitting chain with the high pass filter in the LO chain is shown in Fig. 36, with the power budget measured the same way as in the improvements section. The 5.8 GHz component has a 0.39 dB loss caused by the filter, and it actually improves the LO signal magnitude as it gets closer to 4 dBm. The parasitic component in the LO signal has now a 51.4 dB power difference with the useful component.

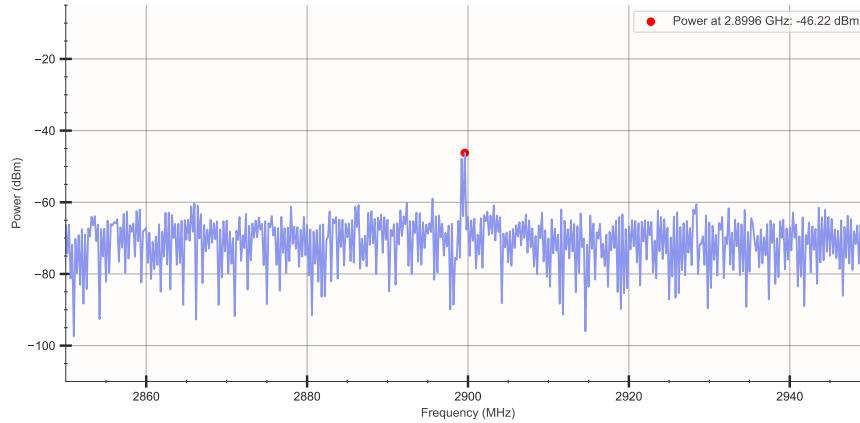


Figure 35: 2.9 GHz component of LO after filtering

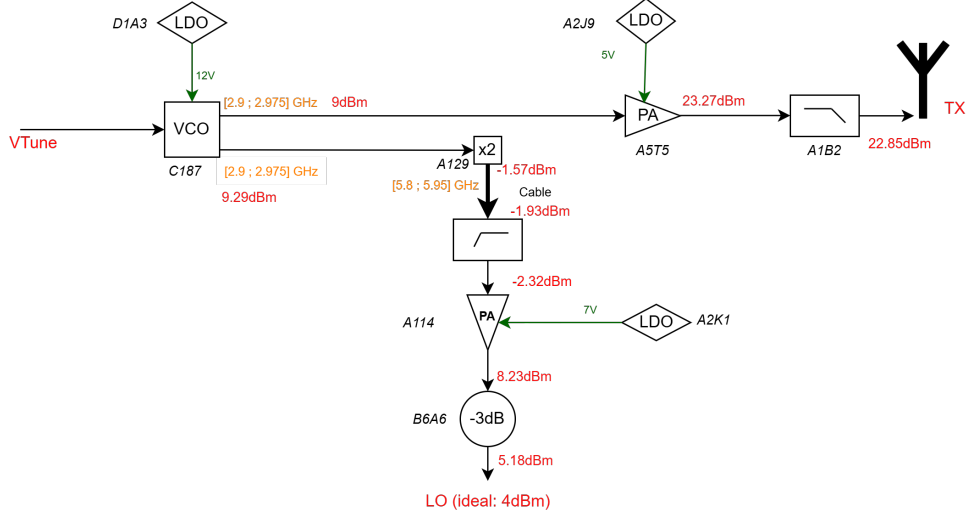


Figure 36: Transmitting chain schematic of the radar with the high pass filter in the LO chain

The TX harmonic component is shown in Fig. 37. The magnitude of -60.85 dBm may appear negligible, as the difference with the fundamental component with a magnitude of 22.85 dBm is around 80 dB. However, the relevant power difference is between the 5.8 GHz transmitted at TX that could be captured the RX antenna and the 5.8 GHz emitted by the tag that is the useful signal. Those two signals could be erroneously combined in the reception chain and cause interferences and a wrong target localization.

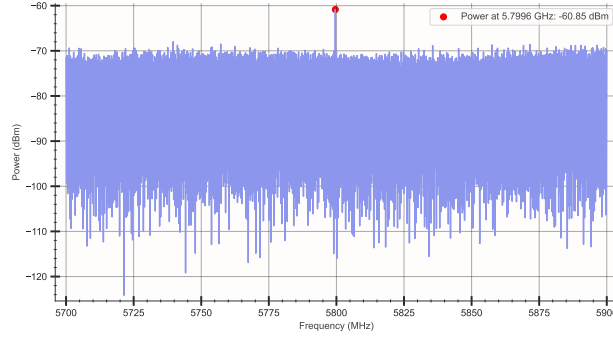


Figure 37: Harmonic component of the transmitted signal

Finally, the 2.9 GHz and 5.8 GHz components at the RF entry of the mixer while connecting a 50Ω load at RX are shown in Fig. 38. Both couplings are high compared to the magnitude of the useful signal expected at this point of the circuit. This coupling should only increase the DC component of the IF signal, as it is at the same frequency than the LO signal. This DC component is still parasitic in the output of the radar, so it would be better to decrease this coupling in future researches by isolating the different parts of the circuit via RF shielding enclosures.

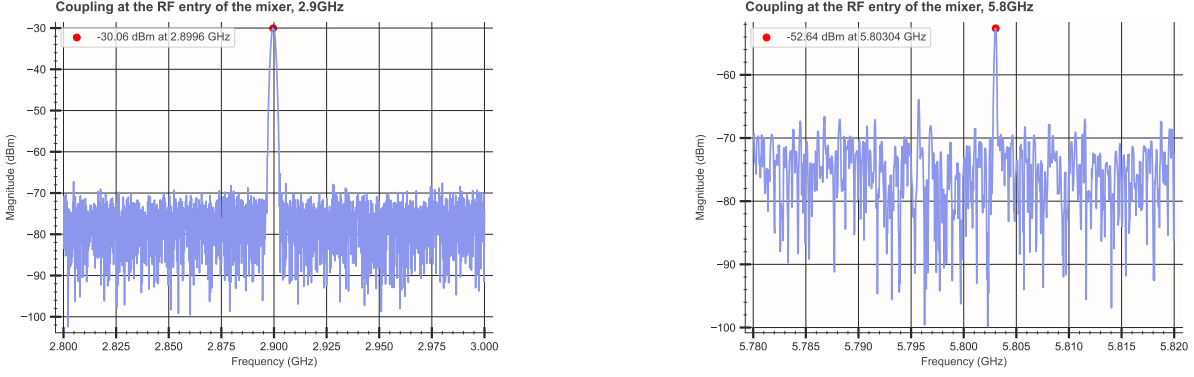


Figure 38: Coupling at the RF entry of the mixer

To conclude the analysis of the signals integrity, the limitations that can be expected in the performance evaluation of the system are the interferences and the DC offset. The interferences can be caused either by the fundamental component of the transmitted signal reflected and captured in the receiving chain, or by the harmonic component transmitted and reflected by obstacles. The DC offset is caused by the coupling in the circuit, and by the non-idealities of the RF elements.

4.III.B Summary of the final radar

After the improvements and the manipulations, Fig. 39, 40, and 41 show the final schematic and pictures of the radar setup with the antennas, the voltage supply and the VCO control with a picoscope.

The voltage supply of the radar is set at 14V and the power consumption, with antennas and an FMCW tune, is 7.42W. With multiple voltage source, some LDOs could be supplied at a lower voltage for the same current consumption, which would decrease the power consumption

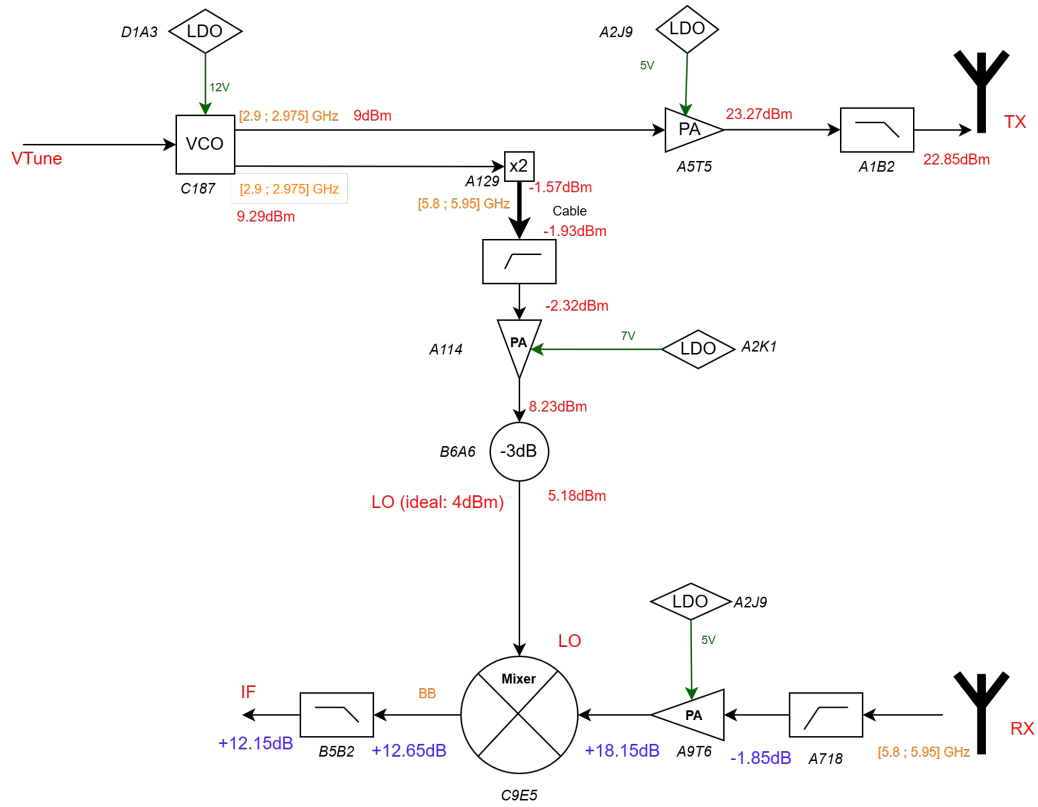


Figure 39: Final schematic of the radar

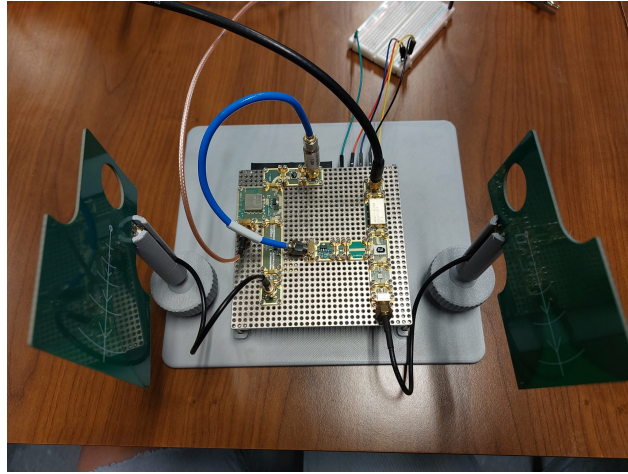


Figure 40: Picture of the radar mounted with the Quantic X-Microwave technology, with the antennas, the voltage supply and voltage tuning

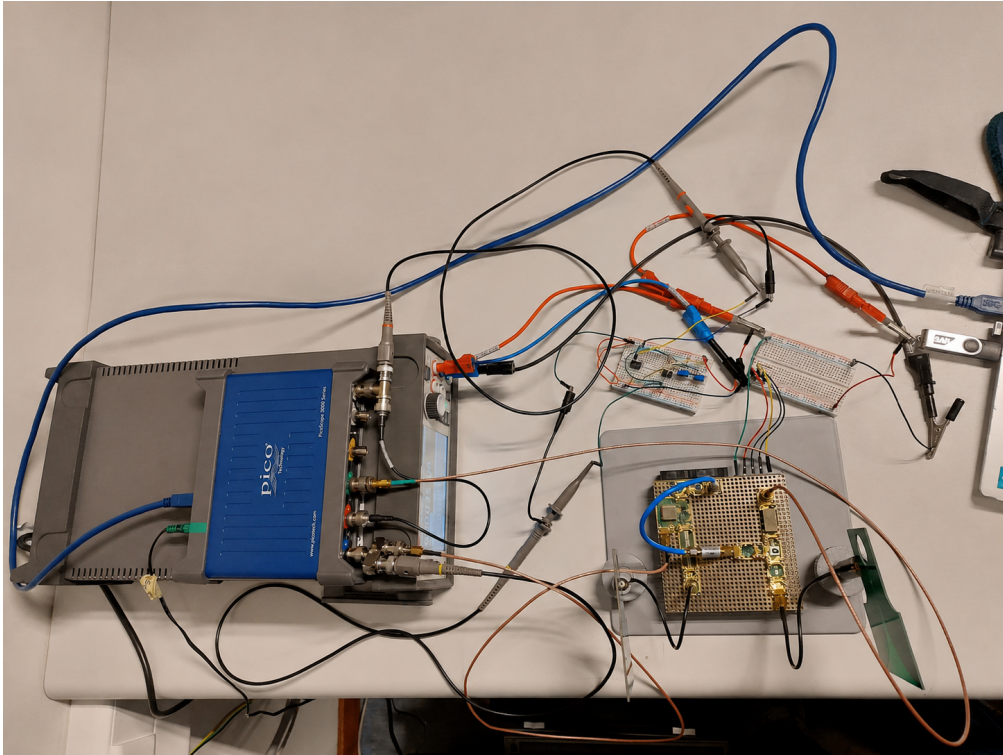


Figure 41: Complete measurement setup in laboratory

Chapter 5

Global system performance evaluation

Firstly, this chapter presents the expected performances of the improved system, through a link budget computation. The radiation pattern of the tag is then measured to evaluate the potential limitations in a real environment. Finally, the radar is tested in laboratory to identify the improvements that could be applied in the future. For this last step, some measurements were made in an anechoic chamber to eliminate reflections from the environment.

5.I Expected performances from link budget computation

Knowing the output power of the radar, the antennas gain and the gain in the reception chain, the maximum range at which the radar can locate the target can be evaluated. A link budget is computed using the measured conversion efficiency of the tag and the path loss estimation.

The hypothesis in this thesis is that the minimum Signal to Noise Ratio (SNR) needed after the signal processing step is 0 dB. This implies that, for a given noise, the ratio between the useful signal power and the noise must be bigger than 1. Thus, this SNR sets the minimum received signal power that the radar is capable of detecting and interpreting, called the sensitivity $P_{RX,sens}$ of the radar. As this SNR is desired at the end of the reception chain after the signal processing, the formula of the sensitivity must include the gain in the reception chain G_{RX} and the processing gain G_{Proc} .

$$P_{RX,sens}[dBm] = SNR_{min}[dB] + Noise_level[dB] - G_{Proc}[dB] - G_{RX}[dB] + 30 \quad (20)$$

The different required values can be computed. The noise can be estimated, considering that it is mainly composed of the thermal noise N_{Th} and the reception chain noise N_{RX} . It can be approximated with the following formula, using the reception chain noise computed in the previous section: $N_{RX} = 3.95dB$.

$$Noise_level[dB] = N_{Th}[dB] + N_{RX}[dB] \quad (21)$$

$$N_{Th}[dB] = 10\log(k_B * Band * Temperature) \quad (22)$$

The thermal noise depends on the bandwidth at the output, and the reception chain ends with a low pass filter that has a bandwidth from DC to 11 MHz. Thus, a value of 11 MHz for the output bandwidth is used, which gives a noise level of **-129.57 dB**. The processing gain for the 1D FFT to obtain the range profile has been given as:

$$Gain_{Proc,range-profile} = 10\log(M_s) + 10\log(\sqrt{M_c}) \quad (23)$$

With a short-time FFT on M_s samples and a quadratic mean on the M_c chirps. With the parameters chosen in Chapter 2 developing an FMCW radar, $M_s = 80$ and $M_c = 130$, it gives a gain of **29.6 dB**. Though it will be explained in the performance evaluation through measurements that those parameters are not ideal for short range and short speed laboratory tests. Thus, the parameters that are used in the laboratory measurements and compared to the expected performances are different: $M_s = 200$ and $M_c = 100$. The same parameters must be used for the estimated performances and the link budget, they give a processing gain of **33.01 dB**. Finally, $G_{RX} = 12.15dB$ from the previous section.

The $P_{RX,sens}$ obtained is the following:

$$P_{RX,sens}[dBm] = 0 - 129.57 - 33.01 - 12.15 + 30 = -141.32$$

Once this limit defined, the curve of the power received by the radar according to the distance is computed in Fig. 42 with its intersection with the sensitivity of the radar.

$$\begin{aligned} P_{RX}(d)[dBm] = & P_{TX}[dBm] + G_{RX,antenna}[dBi] - PathLoss_{f0}(d)[dB] - CE_{tag}(d)[dB] \\ & - PathLoss_{f_harmonic}(d)[dB] + G_{TX,antenna}[dBi] \end{aligned}$$

A maximum range of 7.9 meters is obtained; this is the maximum range for which the power of the signal in the range profile is higher than the noise.

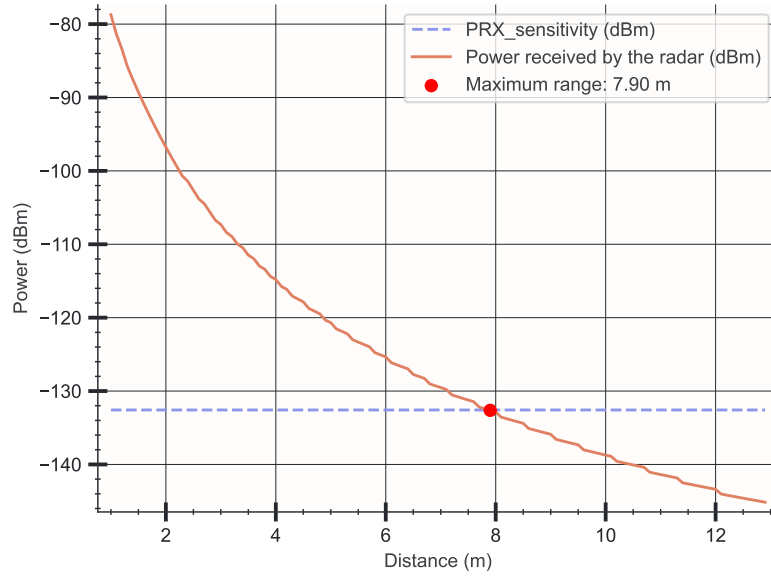


Figure 42: Estimated power budget after the RX antenna according to the distance. Power RX sensitivity for a range profile processing.

The link budget at a distance of 7.5 meters is shown in Fig. 43.

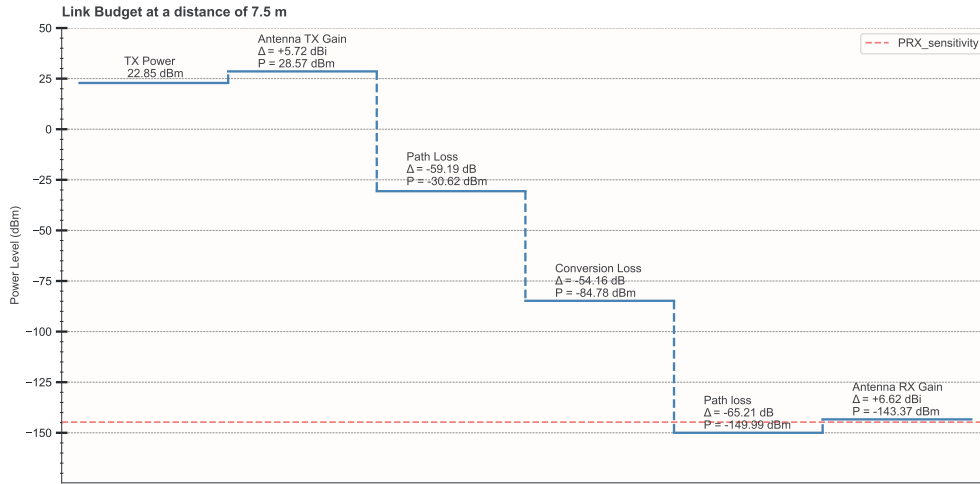


Figure 43: Link Budget of the telecommunication chain for a distance of 7.5 meters. Power RX sensitivity for a range profile processing.

This link budget has been done with the hypothesis that the target can be observed for a SNR higher than 0 dB in the range profile. However, if it is desired to have a SNR higher than 0 dB after a speed profile processing, the processing gain is different, as it is a long-time FFT on M_c chirps and a quadratic mean on M_s samples. The processing gain becomes $10\log(M_c) + 10\log(\sqrt{M_s})$ which makes **31.5 dB** and a maximum range of **7.4 m**.

The third processing method is the Range-Doppler map, which is a 2D FFT of dimension $M_s \times M_c$, the processing gain becomes $10\log(M_s * M_c)$ which equals **43.01 dB** and a maximum range of **11.5 m**. This is the optimum processing gain, which is usually used in the FMCW radars. The link budget for this last case is shown in Fig. 45.

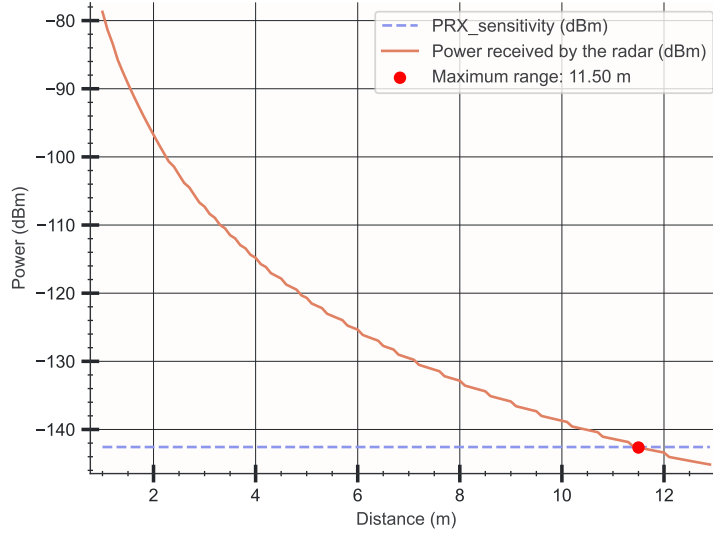


Figure 44: Estimated power budget after the RX antenna according to the distance. Power RX sensitivity for a Range-Doppler map processing.

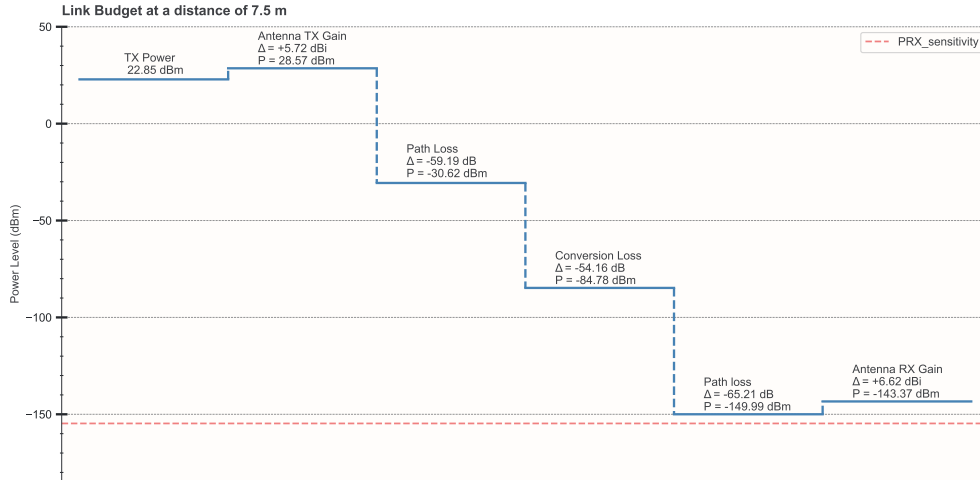


Figure 45: Link Budget for a distance of 7.5 m. Power RX sensitivity for a Range-Doppler map processing.

The quantification noise of the ADC in the picoscope was neglected, which is a valid approximation if the Signal to Quantification Noise Ratio (SQNR) is significantly higher

than the SNR of the reception chain. With the 16 bits resolution of the quantification, the SQNR for a sinusoidal signal can be computed with the following formula:

$$SQNR[dB] = 6.02n + 1.76dB = 98.08 \quad (24)$$

In comparison, for a range of 2 m, the power of the signal at the end of the reception chain is approximately -90.7 dBm (Fig. 42), or -120.7 dB. The power of the noise is -129.57 dB, which gives a SNR of 8.87 dB. This SNR is 89.21 dB lower than the SQNR, so the quantification noise can be neglected. Even for a small range, it is a negligible noise compared to the thermal noise.

5.II Radiation pattern of the tag

As presented in the overview of the tag, its conversion efficiency has already been computed in one specific position. Though in the application of hornet tracking with a UAV, the hornet will most likely change its orientation and direction, and the tag will not always be in the optimal position. To evaluate the waste in power caused by the variation of orientation, Fig. 46 shows the tag in a spherical coordinate system and a picture of how the tag would be positioned on the hornet. The picture comes from an article developing the same technology of hornet tracking with a harmonic radar and a tag [4].

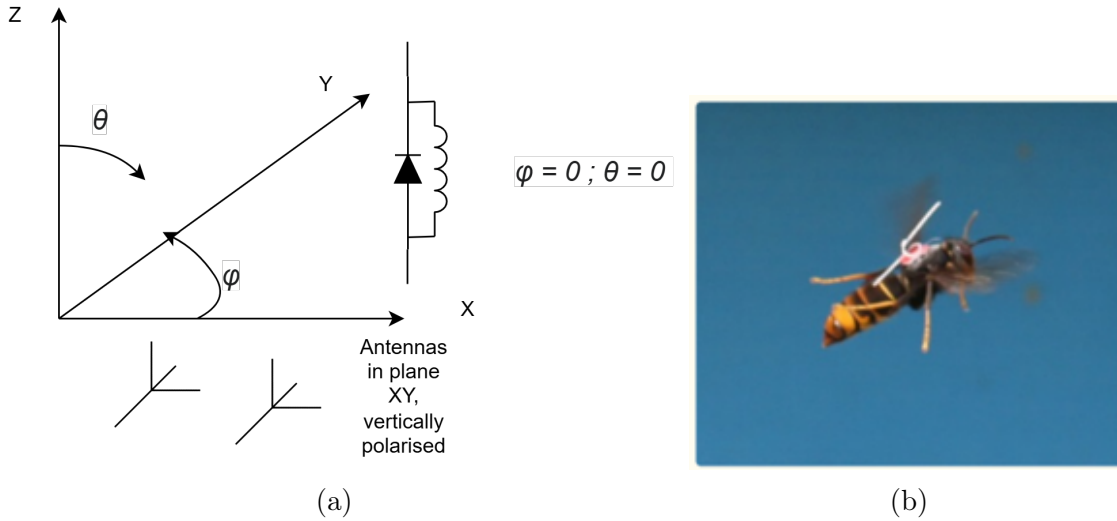


Figure 46: Tag representation. (a) Spherical coordinate system considered of the tag. (b) Illustration of the position of the tag on the hornet [4].

In this coordinate system, the radiation pattern of the tag is computed according to the variation θ and ϕ . Following the common aviation practice, the radiation pattern is computed on the "Pitch" rotation, the "Yaw" rotation and the "Roll" rotation. Their description following the angles of the system is done in the following table:

"Pitch"	$\theta = [0;360^\circ] \ \phi = 90^\circ$	The tilting of the hornet's nose up or down
"Yaw"	$\theta = [0;360^\circ] \ \phi = 0^\circ$	The twisting of the hornet around its normal axis
"Roll"	$\theta = 0^\circ \ \phi = [0;360^\circ]$	The tilting of the hornet from side to side.

Table 6: Rotational axis explanation

5.II.A Method

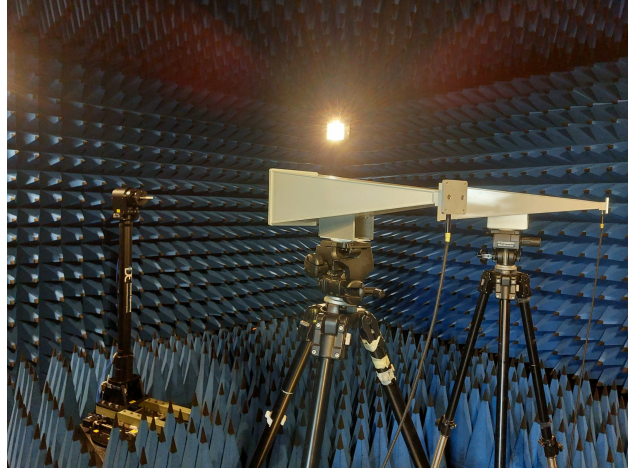


Figure 47: Setup for the tag radiation pattern computation in the anechoic chamber.

A setup similar to the one used to compute the conversion efficiency of the tag was assembled to measure its radiation pattern. The tag was placed in a anechoic chamber facing the same two horn antennas WBH218HS (gain of 13 dBi at 2.9 GHz and 17 dBi at 5.8 GHz). There were 1.8 meters between the tag and each antenna, and 90 cm between the two antennas. The picture in Fig. 47 shows the setup in the anechoic chamber. A N5242B VNA produced the 2.9 GHz signal sent to the first horn antenna at 12 dBm, with a gain of 13 dBi provided by the antenna. The signal received by the other antenna at 5.8 GHz, with a gain of 17 dBi, is measured at the receiver port of the VNA. The magnitude of the power received is extracted, and the conversion efficiency of the tag is computed with the same hypothesis than in the previous overview of the tag.

5.II.B Results

It is important to notice that the numerical values that are observed are not relevant for the application, as it is a specific case of 25 dBm (12 dBm at the VNA output + antenna gain) transmitted at a distance of 1.8 m from the tag. What is interesting is the shape of the patterns or how well the radar is going to be able to track the hornet if it changes its direction. Fig. 48 shows the variation of conversion efficiency according to the rotations. The pitch pattern and the yaw pattern were made with diode in the back, the inductors towards the antennas.

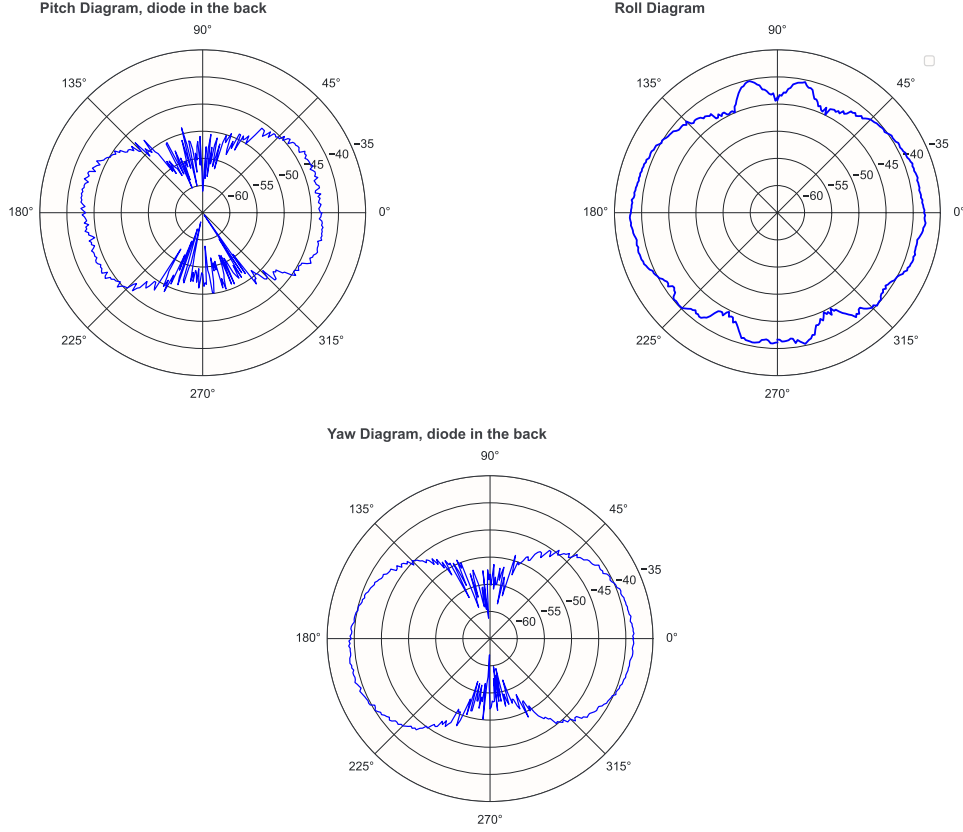


Figure 48: Conversion efficiency patterns of the tag according to the axis of rotation

As could be expected, the roll rotation has the lowest impact on the efficiency as this axis of rotation does not include any blind spot. The lowest difference with the maximum is a 7 dB loss of efficiency at 64° . The 3 dB angles can be observed, corresponding to the angles, in degree, for which the conversion efficiency is not lower than 3 dB below the highest gain. The yaw pattern has a 3 dB angle of 70° and the pitch pattern has a 3 dB angle of 90° . This is consistent with the expected behavior in the intended application, as it is not expected from a hornet to have significant rotation in those directions.

Those patterns also show the instability of the tag. Indeed, the 0° conversion efficiency should be the same for the three patterns, as it corresponds to the same position of the tag. Though a significant difference is visible, probably caused by the manipulation of the tag between the takes. For further application, an important step would be to protect it in a resin, foam, or substrate support.

5.III Radar performance evaluation through distance and speed measurements

With a complete radar, efficient antennas, and a categorized tag, the picoradar software can be used to generate the FMCW tune voltage of the radar and measure its output to locate the tag positioned in front of the antennas.

The tracking of a target by a radar is commonly done by measuring its speed and its range. The speed profile and the range profile of the measurements are computed, and the Range-Doppler maps are shown. Different measurements were done to evaluate the issues caused by the non-idealities or by the limited transmitted power.

5.III.A Method

As explained in the study of the FMCW technology, the range profile is found by computing the FFT of each chirp of length M_s , followed by a quadratic mean between the M_c chirps. The structure of one acquisition of the output of the radar is shown in Fig. 49 as in the FMCW technology chapter.

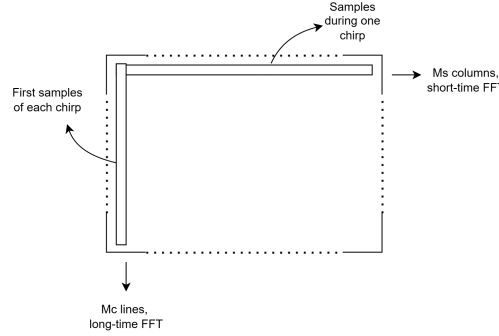


Figure 49: Structure of the data contained in a single radar output frame

To delete every null-range target, the mean value of every line of the matrix is removed. To delete the null-speed targets, the mean value of every column is removed. The code used to plot the graphs shown in the results can be found in the Appendix.

Deleting the null-range targets is always useful and not a limitation in the application, as the hornet should not be at a null distance of the UAV. However, deleting the null-speed targets is indeed a limitation, the hornet must not be invisible to the UAV in the cases where it stops or it has the same speed as the UAV.

During measurements, the reliability of the system was greater when deleting the null-speed target. As the aim of the thesis is to observe the limits of the system, it is interesting to show data processed removing the null-speed targets and without removing them.

For the laboratory analysis of the system, the best choices of FMCW parameters were not the same as those for an actual real-life hornet tracking. The current system is only efficient for a small range, the time of acquisition and the power consumption is not a limit in laboratory, and a high processing gain would be preferred. Thus, other parameters were chosen for the characterization of the system. High M_s and M_c parameters improve the processing gain of the FFT. As a reminder, the processing gain of the range computation is $10\log(M_s) + 10\log(\sqrt{M_c})$ and it is the same formula for the speed computation by switching M_s and M_c . The parameters $M_s = 200$ and $M_c = 100$ are then chosen. By selecting $T_c = 0.4ms$ and $T_s = 1\mu s$, the duty cycle is 50% and the speed resolution is 0.65 m/s. A low

speed resolution is better as high speeds can not be reached in a laboratory and at a small range. Finally, the only way to improve the range resolution was to increase the bandwidth ($\Delta d = \frac{c}{4B_{TX}}$). The radar system is nominally specified to operate over the 2.9–2.975 GHz bandwidth. However, the operating bandwidth has been slightly extended in this work, up to 2.995 GHz. This choice is justified as every component of the radar is expected to behave identically at 2.995 GHz as at 2.975 GHz, and the tag has been validated up to 3 GHz in the original thesis [3]. This bandwidth gives a range resolution of 0.79 m. Table 7 shows the final parameters. The time of acquisition is relatively long: 4 ms for the speed analysis and 200 μ s for the range analysis. Some frames of the recording could contain some artifacts, due to the change of speed during a single acquisition. In the thesis, representative and artifact-free frames are presented.

f_0 , GHz	2.9
Bandwidth B, MHz	95
T_s , μ s	1
M_s	200
T_c , ms	0.4
M_c	100

Table 7: Transmission parameters for tests of the FMCW harmonic radar in laboratory

Finally, several test were done to evaluate the efficiency of the system. The setup presented in the characterization chapter (Chapter 4) was used to power and tune the radar. A ramp in voltage was generated by a picoscope 3000 Series 3403D to produce the FMCW signal at the output of the VCO. The software Picoradar was used to synchronize the generated voltage and the measured output of the radar. The tag, in front of the radar, was placed in a vertical position thanks to a 3D printed structure filled with foam to maintain it without having too much of an impact on the RF behavior.

Processing Comparison - Anechoic chamber

As explained in the chapter 2 that described the FMCW post-processing steps, the results can be improved by deleting the short-time DC component, the long-time DC component, or the background noise. The background noise is usually subtracted in radar post processing, so this operation is done in every results shown in this thesis. However, three processing configurations are considered:

- Keeping entirely the DC component
- Removing the short-time DC component (which hides the null-range targets)
- Removing the short-time and long-time DC components (which hides the null-range and null-speed targets)

For this experiment, the whole setup of the radar was placed in an anechoic chamber in front of the tag, and the tag was moved toward and away from the radar, centered at a 2-meter distance. The setup was placed on a isolated block, surrounded by foam to avoid reflections of the emitted signal.

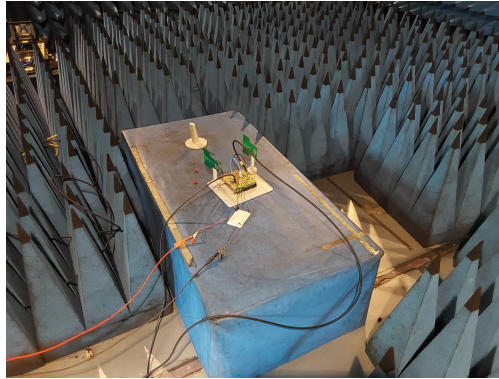


Figure 50: Measurement setup in the anechoic chamber to test the performances of the system

The output signal of the radar was recorded while moving the tag toward and away from the radar.

Interferences - Anechoic chamber

As presented in the section 4.III.A about the integrity of the signals, there is a significant risk to have interferences in the system between the signal coming from the tag and the signal reflected by other targets in the environment. To try and evaluate the impact of those interferences, a comparison in the anechoic room has been done between the localization of the moving tag and the localization of a moving hand. Both recordings were done at a distance of two meters, with the tag moving toward and away from the radar.

Range Measurement - Anechoic chamber

The accuracy of the range measurement was also measured in the anechoic chamber, by using the same setup without moving the tag. It was positioned at exactly 2 meters from the antennas of the radar.

Maximum range - Common room

The approximate maximum measurable range was measured in a laboratory where enough space would be available. Range measurements were conducted up to 3.5 meters, which was found to be the approximate maximum detectable range of the system under the current experimental conditions, as target detection became unreliable beyond this distance. The measurements were done with the tag moving toward and away from the radar.

5.III.B Results

The plots that are shown in this section are cropped for ranges lower than 10 meters and speeds between 0 and 5m/s to get a better view of the accuracy. The Range-Doppler maps are shown with the same cropping but a speed between -5 and 5m/s. This difference of axis can be explained by the type of processing. For the speed-profile, a 1D FFT is computed on a real signal, so the FFT is symmetrical according to the 0 frequency axis. In this case, no difference can be seen between the positive and negative speeds: for a given speed v measured, the opposite speed $-v$ will be seen on the profile because of the symmetry. For the range profile, the domain observed goes from 0 to d_{max} , so a distance d measured will make appear a symmetric distance at $d_{max} - d$.

For the Range-Doppler map, a 2D FFT is computed on a real signal. A symmetry also occurs, but it is a central symmetry because of the two dimensions. A measured target of distance d and speed v will make appear a symmetrical target at distance $d_{max} - d$ and speed $-v$. This enables us, by cropping the ranges to 10 meters instead of d_{max} , to see negative speed targets.

Processing Comparison - Anechoic chamber

Fig. 51 shows the three range profiles computed with the three post-processing methods on one Frame of the recording. Even if the deletion of null-range targets improves precision, there are still false targets seen at lower ranges. The curve with the deletion of null-range and null-speed targets shows a magnitude of the peak, seen around 2 and a half meters, almost 5 times higher than the noise in mV.

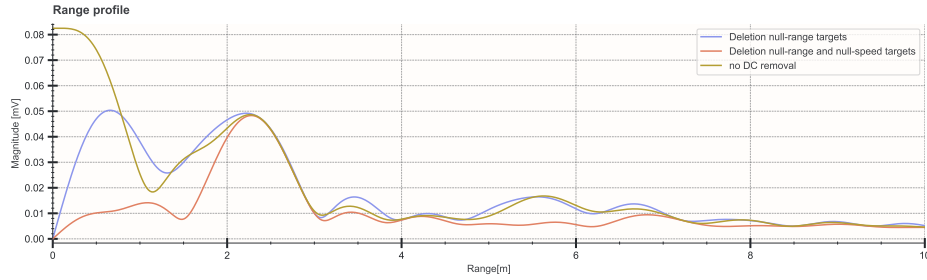


Figure 51: Range profile of a moving target at 2 meters from the radar

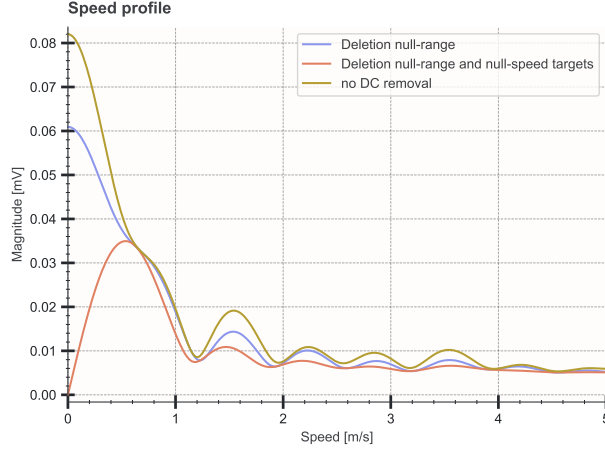


Figure 52: Speed profile of a moving target at 2 meters from the radar

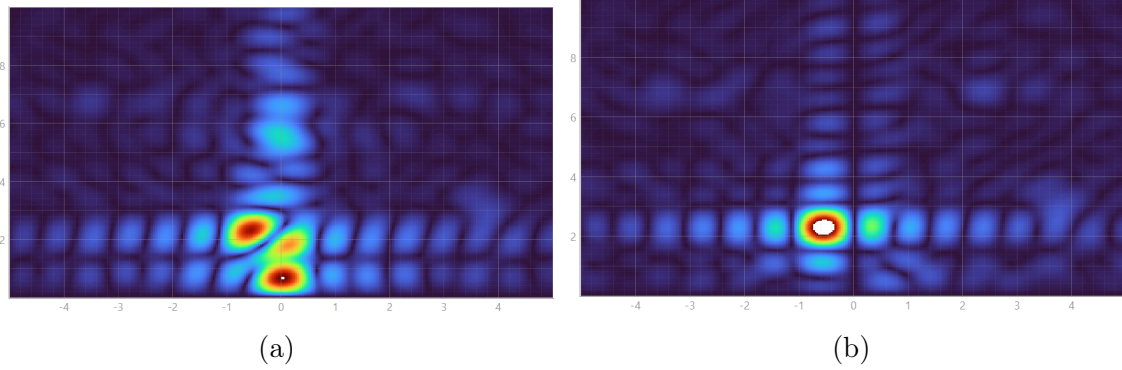


Figure 53: (a) Normalized Range-Doppler Map of a moving target at 2 meters, removing the short-time DC component. (b) Normalized Range-Doppler Map of a moving target at 2 meters, removing the short-time and the long-time DC component.

The speed profile contains a high peak at low speeds when removing only the null-range targets, deleting the null-speed targets enables to only have a peak at the speed of the target. The tag was manually displaced at an estimated speed of 1 m/s, so the measurement is plausible.

The Range-Doppler map clearly shows the difference of processing. The removal of the null-speed targets deletes the parasitic detection, and the localization succeeded. Without deleting null-speed targets, two potential locations appear.

Interferences - Anechoic room

To measure the level of interference that can appear in the radar, Fig. 54 and 55 show the difference between a tag localization and a hand localization thanks to a range profile and a speed profile. Those profiles were obtained with a post processing that deletes the null-range and null-speed targets to be able to observe clearly the difference between the two

measurements.

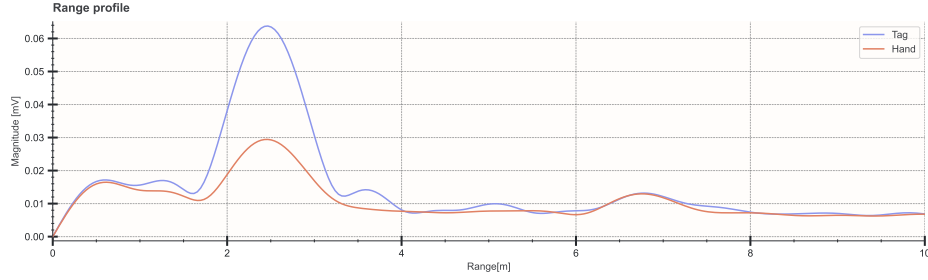


Figure 54: Comparison of the range profiles obtained moving a tag or with a hand

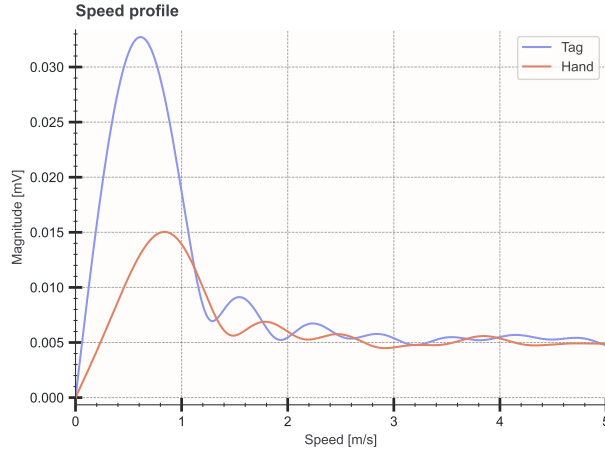


Figure 55: Comparison of the speed profiles obtained moving a tag or with a hand

Unfortunately, the radar is sensitive to a hand moving in front of it, and it can locate it. This means that the radar can not only track the tag positioned on the back of a hornet, but also the other obstacles it could meet on the way. However, for a range of 2 meters between the target and the radar, there is approximately a 3.5 dB difference (a ratio of 2.2) between the tag peak and the hand peak in the range profile and in the speed profile. So if the tag and other obstacles are observed, the tag is the main target and it can still be followed.

Fig. 56 shows the two Range-Doppler maps, in the case where a tag or a hand is located. The power difference between the target and noise is higher in the Range-Doppler map of the tag than the hand's, but the hand location is not negligible. In the application, it must be possible to notice when the tag is not found in the observed perimeter. In the current implementation, if the hornet is not seen, other obstacles can be confused with the hornet. Decreasing the interference in the radar would improve the results.

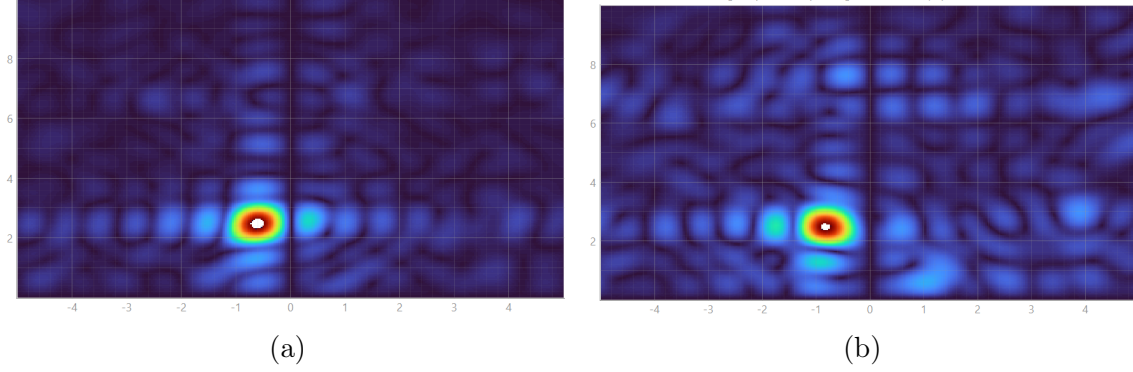


Figure 56: Range-Doppler maps (a) Result of the tag localization. (b) Result of the hand localization.

Range Measurement Accuracy - Anechoic chamber

Fig. 57 shows the range profile for two post processing methods, deleting the null-speed targets or not. Of course, as the tag was not moving, nothing is observed for the first processing. The second processing contains two peaks, the relevant one around 2 meters, and a peak at a low range. This parasitic peak is predictable, given the previous analysis comparing the different processing results. The speed profile is not relevant when removing the null-speed targets, as the target is static. The Range-Doppler map presents the static target and a parasitic null-speed target. When removing the null-speed targets, the Range-Doppler map is contains no discernible target, consisting entirely of noise. The measurement shows a defect in accuracy, giving a range of 2.775 meters instead of 2 meters. The processing is not accurate yet as the time loss in the cables is not taken into account. This is an improvement that should be done in the future.

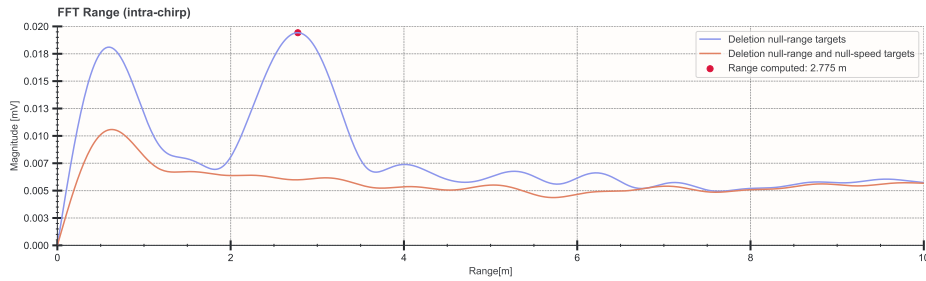


Figure 57: Range profile of a static target at 2 meters of the radar

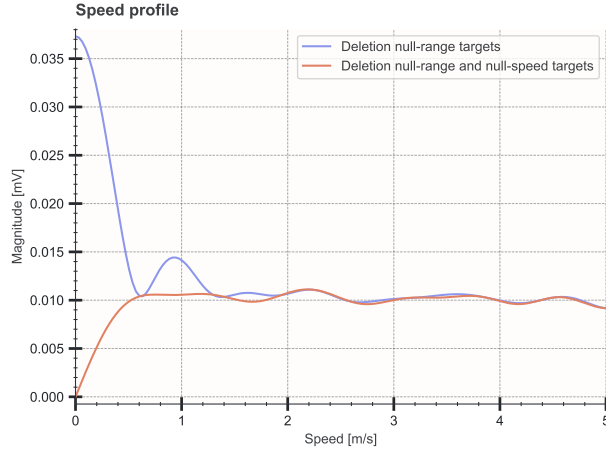


Figure 58: Speed profile of a static target at 2 meters of the radar

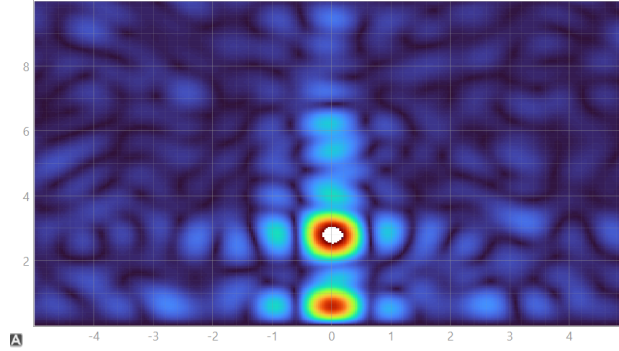


Figure 59: Normalized Range-Doppler Map of a static target at 2 meters, removing the null-range targets

Maximum range - Common room

The maximum range was assessed by progressively increasing the distance between the tag and the radar. Beyond 3.5 meters, the received signal magnitude approached the noise floor, and reliable target detection could no longer be achieved.

For a 3.5m distance, the only post processing that gives an interesting result is removing the null-range and null-speed targets. Not removing the null-speed targets while moving the tag at such a distance makes the results uncertain with random peaks. The limit of the tag is close to be reached, mostly because the measured magnitude approaches the noise. There is also a higher peak than the one caused by the target at a low range, that does not decrease the quality of the Range-Doppler map as the power of this first peak in the range profile is distributed along the speed axis. The second peak that corresponds to the target can still be observed, with an offset caused by the cable lengths.

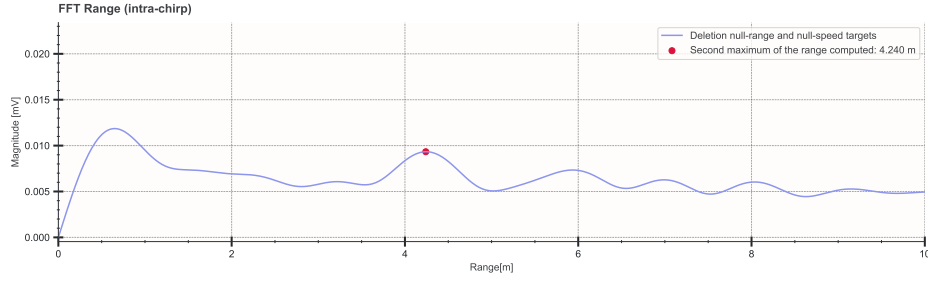


Figure 60: Range profile of a moving target at approximately 3.5m of the radar

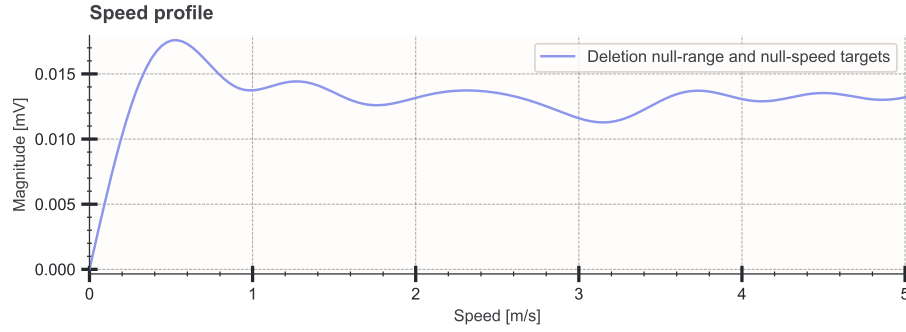


Figure 61: Speed profile of a moving target at approximately 3.5m of the radar

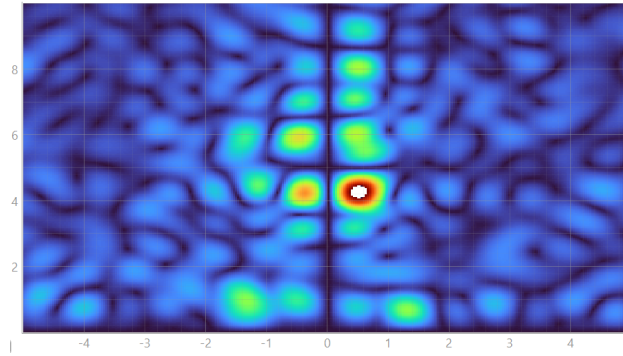


Figure 62: Normalized Range-Doppler Map of a moving target at 3.5 meters, removing the null-range and null-speed targets

5.IV Discussion of the results and area for improvement

Those measurements help to represent the potential of the system. They can be summarized to deduce the flaws and improvements that should be done in further researches.

5.IV.A DC offset

First of all, the most apparent shortcoming is the DC offset in the speed profile. It is extremely common in radars technology to contain a DC offset in the spectrum of the output, and it usually does not cause any problem as it can easily be removed from the spectrum, by subtracting the mean value on the recording. The issue in this case, however, is that null-speed targets must be observable: the radar would be on a drone, following an Asian hornet at a near-identical speed. The DC component that remains after subtracting the mean of the short-time measurements can either be decreased by insulating the radar to decrease the coupling, and thereby the DC offset.

5.IV.B Interferences

A more insidious issue is the interferences in the localization. It was a predictable problem, as the characterization of the radar highlighted that there was a high likelihood of interferences caused by other targets reflecting the fundamental frequency or the parasitic harmonic component of the transmitted signal. The interferences can be problematic if the tag is not in the observable range of the radar and it starts to locate any other obstacle. A solution could be to add a power threshold in the location algorithm, to only consider the target measured as the tag if the power of the signal is high enough, but the maximum range and the covered area should be decreased as the target signal would have to be above the imposed threshold.

The two sources of interference should be studied. First, the reflection of the 2.9 GHz transmitted signal captured by the RX antenna. To rigorously measure the interference, the mixer in the reception chain should be characterized in the subharmonic situation, to obtain the Conversion Loss of the 2.9 GHz component at the RF input according to its power and the power of the 2.9 GHz component at the LO input. The LO signal has a 5.8 GHz component at 5.18 dBm and a 2.9 GHz component at -46.22 dBm. By measuring the IF magnitude with a 2.9 GHz signal at the RF input of the mixer, the remaining amplitude of the parasitic output could be deduced. A Link Budget at 2.9 GHz could be done, with the path losses, the losses in the filter at the reception and in the mixer. If this parasitic output is too high, a narrow-band antenna specified to work at 5.8 GHz could improve the results.

The second interference, caused by the 5.8 GHz component in the transmitted signal, can be estimated with a Link Budget. Fig. 63 shows the Link Budget of a 5.8 GHz signal that would be reflected by an obstacle at 2 meters. The RX power sensitivity is also represented for a Range-Doppler map processing. In this graph, the received power caused by the transmitted harmonic is lower than the sensitivity limit of the receiver, so this is probably not the main cause of the interference. This component can be problematic with short range obstacles that would hide the target.

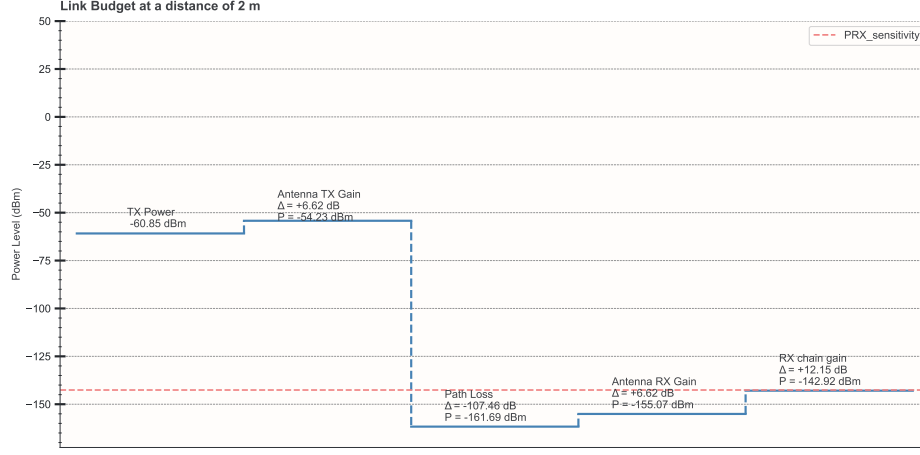


Figure 63: Link Budget at 2 meters of the 5.8 GHz transmitted signal component

5.IV.C Inaccuracy

A difference of 0.775 meters was noticed between the actual range and the one measured by the radar. The computation to obtain a distance according to the frequency of the output is made considering that the frequencies at the input of the mixers are equal to the frequencies of the transmitted and received signals. In reality, the time lost in the cables and in the RF circuit is not negligible. The actual formula to compute the distance from the frequency output should be:

$$d_{target} = \frac{\Delta f_{IFC} M_s T_S}{4B_{TX}} - d_{circuit}$$

5.IV.D Maximum range difference

There is a clear difference between the estimated maximum range and the maximum observable distance in the measurements. Even with the most efficient processing method, the target could not be seen further than approximately 3.5 meters. Firstly, this approximation is not rigorous as the tag had to be moving to be able to remove the null-speed targets in the processing, the range is not perfectly measured. Secondly, the noise computation done for the Link Budget is not actually complete. Noise can come from the non-idealities listed in the characterization of the radar, from the tag, or most importantly from the phase noise of the VCO. The phase noise comes from the short-term fluctuation of the frequency at the output of an oscillator. The VCO chosen in the radar is described as a "low Phase Noise VCO", but in this case with a extremely low power received, it can have a negative impact on the maximum range. The study of the phase noise in the system could be done in further researches.

To try and obtain the actual noise level, the output of the radar can be recorded while loading the RX and TX inputs at 50Ω. By measuring this output, every noise source would be included and a new approximation of the maximum range would be obtained. To measure the AC power of the noise from a voltage measurement in the temporal domain, the variance

of the noise σ^2 is extracted, which corresponds to V_{eff}^2 . Then, the power is computed from the internal resistance of the oscilloscope which is $1M\Omega$. The following expression gives the noise power:

$$P_{noise}[dBm] = 10\log\left(\frac{V_{eff}^2}{R_{oscilloscope}}\right) + 30 = 10\log\left(\frac{\sigma^2}{10^6}\right) + 30 \quad (25)$$

With a σ of 0.8 mV, the noise obtained is **-91.87 dBm** instead of the -99.57 dBm estimated. With this noise, the maximum range at which the Range-Doppler map shows a useful target with a power lower to the noise becomes 8.6m instead of the 11.5 m estimated. Also, the maximum range measurement was not made in an anechoic chamber because of the lack of space, so the environment caused reflections and noise in the background. A more in-depth study of the power budget of the signal measured by the oscilloscope must be done to obtain more complete results. Also, a study of the SNR in the oscilloscope measurements would help to get a better idea of the performances.

5.IV.E System improvements

In the global system, the tag could be improved by protecting it. The conversion efficiency of the tag varies because of the handling, as it is fragile and can be deformed. For further measurements, the tag should be put in a support to avoid behavior variations in the tests.

Also, a general change to improve the system would be to increase the transmitted power with an amplifier, at the cost of a higher energy consumption. It would be the most efficient way to increase the power received, thus the power of the output signal. The conversion efficiency of the tag being proportional to the square of its received power, an increase of the transmitted power is two times more efficient than increasing the reception gain. To transmit at a higher power level, the current amplifier can be changed for one with a bigger gain, or a powerful amplifier can be added between the TX chain and the antenna.

Chapter 6

Conclusion

The purpose of this thesis was to design a complete FMCW harmonic radar system operating in the 2.9 GHz/5.8 GHz frequency bands with the objective of tracking Asian hornets back to their nest. The use of a passive tag keeps the cost low and the weight minimal, but introduces the challenge to reach a sufficient range to make the system usable for real-world tracking in natural environments.

First, the FMCW technology was studied to establish the operating limits and select the optimal parameters for the application. The Picoradar software was reviewed to generate the FMCW signal, and the different processing methods available to locate the tag from the radar output were examined.

Building upon the work carried out in previous Master theses, the first version of the system was characterized, to obtain the function of the conversion efficiency of the tag, and to highlight the improvements required to obtain an operational system. The method to measure the maximum range was established, with the different steps of the link budget computation. By improving the radar and antenna, and implementing the missing components including the FMCW signal generation, the complete system was characterized. This revealed performance limitations caused by harmonic leakage and spectral impurity of the transmitted signal.

Finally, after evaluating the expected performances and studying the radiation pattern of the tag across multiple orientations, localization measurements were conducted both in an anechoic chamber and in a laboratory environment, allowing to assess the actual potential of the technology as well as the remaining areas for improvement.

The final radar transmits a signal at 22.85 dBm, while its reception chain provides 12.15 dB of gain with a noise figure of only 3.95 dB. A theoretical maximum range of 11.5 m. was estimated, but in practice, a 3.5 m maximum range was measured. This difference can be attributed to an underestimation of the noise floor, caused by the phase noise of the VCO or by the non-idealities of the radar. Additionally, a significant DC offset in range-speed profiles and interferences were identified as the main limitations of the system.

Nevertheless, further improvements could bring the system closer to field deployment: increasing the transmitted power would be the most efficient modification. Replacing the current antennas with models specified for the 5.8 GHz at the reception and 2.9 GHz at the

transmission would also reduce the interferences, and a better isolation of the RF circuit would attenuate the DC offset, allowing the detection of targets at null relative speed which is a critical requirement for a radar mounted on a drone tracking a hornet at near-identical speed.

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Appendix

Range-Speed Profile computation code:

```
1  FRAME = i          #frame observed
2  sig = data[FRAME]
3  sig = sig - background
4  y= sig.reshape(Mc, Ms2)[: , :Ms] * scale  # reshape, active samples selections, mV
   ↪ conversion
5
6  y = (y.T-np.mean(y, axis=1)).T  #Deletion of the null range targets
7  if (Delete_null_speed_range):
8      y= y-np.mean(y, axis=0) # Deletion of the null speed targets
9
10 # FFT Range (short-time : intra-chirp)
11 # 200 samples → zero-padding
12 NFFT_r      = 4096
13
14 Y_r          = np.fft.rfft(y , n=NFFT_r, axis=1)
15
16 freqs_r      = np.fft.rfftfreq(NFFT_r, d=1 / fs)
17
18 range_profile = np.sqrt((np.abs(Y_r)**2).mean(axis=0)) / math.sqrt(NFFT_r)  #
   ↪ Quadratic Mean and scaling in mV
19
20 # FFT speed (long-time : inter-chirp)
21 # 100 samples → zero-padding
22 NFFT_s      = 4096
23
24 Y_s          = np.fft.rfft(y , n=NFFT_s, axis=0)
25
26 freqs_s      = np.fft.rfftfreq(NFFT_d, d=Tc)
27
28 speed_profile = np.sqrt((np.abs(Y_s)**2).mean(axis=1)) / math.sqrt(NFFT_s)  #
   ↪ Quadratic Mean and scaling in mV
```

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