



**SPACE IT UP/LISAR-DES**

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# **Development of an SF-CW Radar for Subsurface Ice Detection on the Moon**

**Lunar Advanced Radar for Subsurface Investigation**

**“LISAR”**

## **System Design Definition**

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## 1 SCIENTIFIC CONTEXT

Design and development of a GPR for a micro-rover aimed at searching for water ice in the lunar subsurface, down to a nominal depth of 5m.

Recent studies based on the analysis of data acquired by the GPR (Lunar Penetrating Radar) from the Chang'E-4 mission have made possible the estimation of the following parameters:

- Propagation speed in the lunar subsurface:

$$v = 0.16 \frac{m}{ns} \rightarrow \epsilon_{r1} = 3.0 \text{ (Dielectric permittivity of lunar regolith)}$$

- Dielectric loss factor of the lunar soil:

$$\tan\delta = (5 \pm 2) \times 10^{-3}$$

The value of  $\tan\delta$  allows for the estimation of the two-way attenuation of the radar signal in the lunar subsurface using the following equation:

$$L_{SS/m} = 1,8 \cdot 10^{-7} f_c \sqrt{\epsilon_{r1}} \cdot \tan\delta = 1.0dB/m \div 2.0dB/m \text{ (eq. 1.1)}$$

Considering the worst-case scenario, the following value was selected for defining the instrument link budget:

$$L_{SS/m} = 2dB/m \text{ ( Two way travel time )}$$

Finally, the dielectric permittivity of ice mixed with regolith has been estimated in:

$$\epsilon_{r2} = 5 \text{ (Mixture of ice and regolith)}$$

## 2 SCIENTIFIC REQUIREMENTS

The main objective of the LISAR mission is the search for water ice within the first 5 meters of depth, with a range resolution of approximately 15cm in a vacuum (8.6 cm in the medium).

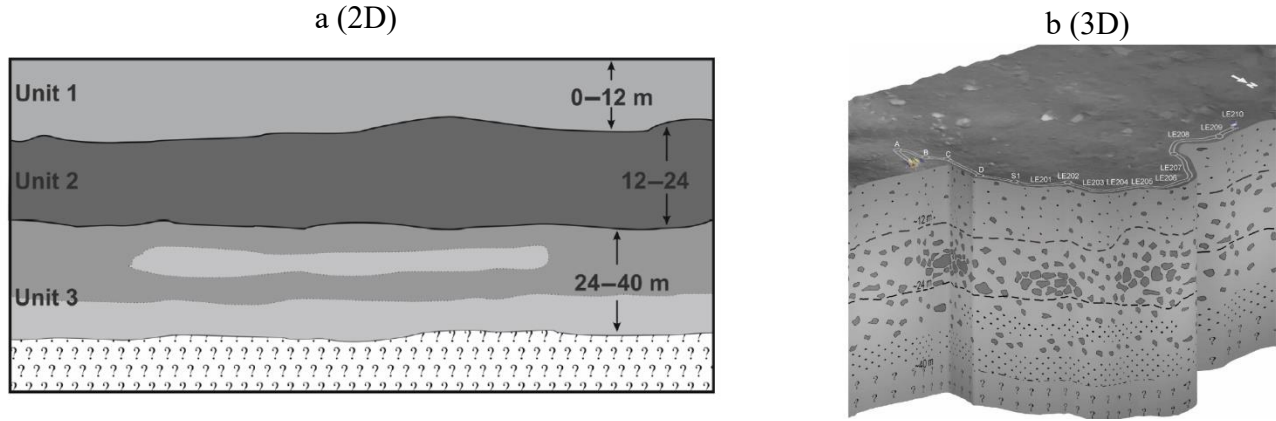


**Figure.2.1** Lunar geological profile, estimated with Chang'E-4 data.

From the following equations, it is possible to estimate the required bandwidth to achieve the mission's scientific objective.

$$R_d = \frac{c \cdot \tau_c}{2 \cdot \sqrt{\epsilon_{r1}}} = \frac{c}{2 \cdot B_c \cdot \sqrt{\epsilon_{r1}}} \quad (eq. 2.1)$$

By imposing a resolution of 15cm in a vacuum (8.6 cm in the lunar subsurface), an RF bandwidth of:

$$B_c = \frac{c}{2 \cdot R_d \cdot \sqrt{\epsilon_{r1}}} \rightarrow B_c = 1GHz \quad (eq. 2.2)$$

In addition to the identification of extended targets, such as ice deposits, the GPR must be capable of detecting boulders ranging from 15 to 30cm in size. This requirement contributes to the definition of the instrument's carrier frequency, which must have a wavelength of approximately 30cm:

$$f_c = \frac{c}{\lambda_c} \rightarrow f_c = 1GHz \quad (\lambda_c = 30cm) \quad (eq. 2.3)$$



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An additional criterion for defining the value of the carrier frequency is to consider the ratio between the physical size of the antenna and the signal attenuation in the medium. Low carrier frequencies are preferable to reduce attenuation but require larger antennas. On the other hand, high carrier frequencies allow for smaller antenna sizes, but at the cost of greater attenuation. A good compromise between these two aspects is represented by a carrier frequency of 1 GHz.

The fractional bandwidth of the system is:

$$B_{\text{frac}} = \frac{f_H - f_L}{f_c} = \frac{B}{f_c} = 1 \quad (\text{eq. 2.4})$$

$B_{\text{frac}}[0 \div 2]$ . Optimal values are:  $B_{\text{frac}} \leq 0.5$

Systems with a fractional bandwidth greater than 0.5 require complex architectures and design solutions to ensure consistent performance across the entire band. This applies both to the radar electronics and, even more significantly, to the antennas, whose behaviour in terms of input impedance, radiation pattern, and polarization varies significantly with frequency.



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### 3 ANTENNA REQUIREMENT

To simplify the problem, it is assumed that all the energy radiated by the antenna is transmitted exclusively into the underlying medium, without generating any reflection from the lunar surface. In other words, it is assumed that both the transmitting and receiving antennas are immersed in a medium with the same permittivity as the regolith.

An important aspect is that, since the radar system will be installed on a micro-rover (small dimensions, around 60cmx60cmx40cm), the antennas will be very close to the lunar surface, at a distance of about 10-20 cm. This creates a strong interference zone (blind spot) in the subsurface, approximately  $1,5\lambda$  (~45 cm), where it will be difficult to detect echoes.

Currently, the only defined requirement for the antenna is the operational band of 1GHz, centered around a carrier frequency of GHz, with a gain of 4/7dB. The antenna must operate in an environment with a dielectric constant  $\epsilon_{r1} = 3$  (typical of lunar regolith) and an input impedance of  $Z = 50 \Omega$ .

## 4 RECEIVER DYNAMICS REQUIREMENTS

To simplify the estimation of the main operational parameters of the instrument, it is assumed to operate in uniform media, consisting exclusively of smooth and parallel surfaces. Appropriate margins, applied to the various system parameters, will allow the instrument's operation to be adapted to more complex and realistic scenarios.

One of the most critical aspects of the GPR is the correct estimation of the required dynamic range to ensure that the receiver handles the received echoes properly. The most intense signal is generated by the direct coupling between the transmitting and receiving antennas, while the weakest signal is the echo generated by the reflection from the layer at the maximum detectable depth of 5 meters, as illustrated in Figure 4.1.

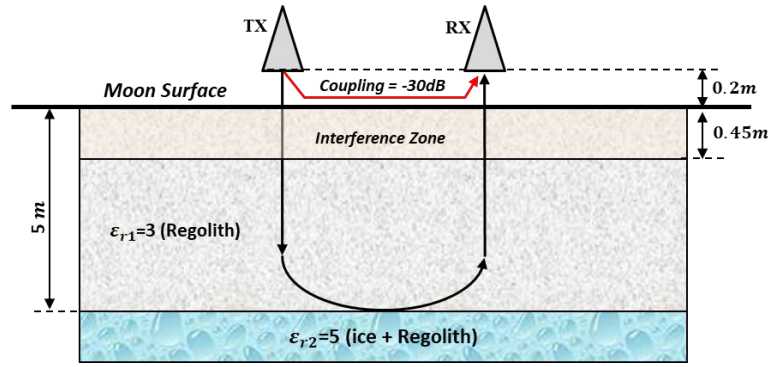


**Figure.4.1** Nominal observation geometry

In order to define the power levels of the received echoes, it is necessary to define the various levels of attenuation in the lunar subsurface.

For the model of radio wave propagation in the lunar subsurface, the approximation described in the work of Porcello et al. has been adopted, valid for extended, smooth, and parallel surfaces, as described by the following relationship:

$$P_{RX} = P_{TX} \cdot G_{Ant}^2 \cdot \left( \frac{c}{f_c \cdot 8 \cdot \pi \cdot Z} \right)^2 \cdot |R_{SS}|^2 \cdot \frac{1}{L_{SOIL}} \quad (eq. 4.1)$$

$$R_{SS} = \frac{\sqrt{\epsilon_{r1}} - \sqrt{\epsilon_{r2}}}{\sqrt{\epsilon_{r1}} + \sqrt{\epsilon_{r2}}} \quad (eq. 4.2)$$

$$L_{Soil} = L_{SS} \cdot Z \quad (eq. 4.3)$$



By expanding the attenuation term of the radar's operating environment, as follows:

$$EnvAtt = \left( \frac{c}{f_c \cdot 8 \cdot \pi \cdot Z} \right)^2 \cdot |R_{SS}|^2 \cdot \frac{1}{L_{SOIL}} \quad (eq. 4.4)$$

And by substituting it into equation 4.1, the following is obtained:

$$P_{RX} = P_{TX} \cdot G_{Ant}^2 \cdot EnvAtt \quad (eq. 4.5)$$

Where:

|                   |                                                                                |
|-------------------|--------------------------------------------------------------------------------|
| $P_{TX}$ :        | Power of the signal at the antenna input (Transmitted Power)                   |
| $P_{RX}$ :        | Power of the signal at the output of the antenna (Received Power)              |
| $f_c$ :           | Carrier frequency                                                              |
| $G_{Ant}$ :       | Gain of the TX and RX antennas (it is assumed that the antennas are identical) |
| $Z$ :             | Penetration depth                                                              |
| $R_{SS}$ :        | Reflection coefficient between the regolith and the ice layer                  |
| $L_{SOIL}$ :      | Total attenuation factor of the subsurface at depth Z                          |
| $L_{SS}$ :        | Two-way attenuation in the regolith                                            |
| $\epsilon_{r1}$ : | Dielectric permittivity of lunar regolith                                      |
| $\epsilon_{r2}$ : | Dielectric permittivity of lunar regolith mixed with water ice                 |

The value of  $EnvAtt$  at a depth of 5 meters is:

$$EnvAtt_{5m} = 80dB$$

The useful dynamic range of the signals, which we will call "Instrument Dynamic," is given by the following equation

$$INS\_DYN = P_{RX,Max,dB} - P_{RX,min,dB} \quad (eq. 4.6)$$

$P_{RX,Max}$  is the maximum power that the instrument must be able to handle without saturating, generated by the antennas coupling, which is undesirable but unfortunately often inevitable.

Meanwhile,  $P_{RX,min}$  is the minimum power of the return echo at the maximum investigation depth of 5m. Below, these two terms are specified:



$$P_{RX,Max,dB} = P_{TX,dB} - AC_{dB} \quad (eq. 4.7)$$

$$P_{RX,min,dB} = P_{TX,dB} + 2 \cdot G_{Ant,dB} - EnvAtt_{dB}(5m) \quad (eq. 4.8)$$

Substituting equations 4.7 and 4.8 into equation 4.6 and assuming the following values:

$$\begin{aligned} G_{Ant} &= 7dB \\ AC &= 30dB \end{aligned}$$

It is obtained:

$$INS\_DYN_{dB} = P_{RX,Max,dB} - P_{RX,min,dB} = -AC_{dB} - 2 \cdot G_{Ant,dB} + EnvAtt_{5m} = 36dB \quad (eq. 4.9)$$

$$INS\_DYN_{@5m} = 36dB$$

The estimated 36dB dynamic range of the signals must be compatible with the receiver's dynamic range. Excluding, for now, the contribution of the intrinsic noise from the receiver's electronic circuits, it is calculated using the following equation:

$$REC\_DYN = P_{in\_FULL\_SCALE} - P_{in\_MIN} \quad (eq. 4.10)$$

Assuming a receiver with a nominal input peak-to-peak voltage of  $V_{pp} = 1V_{olt}$  across a  $100 \Omega$  load and 14-bit ADC quantization, the following is obtained:

$$P_{in\_FULL\_SCALE} = \frac{V_p^2}{2 \cdot R} = 1dBm \quad (eq. 4.11)$$

$$P_{in\_MIN} = \frac{d_{Vp}^2}{2 \cdot R} = \frac{V_{PP}^2}{(2^{N_{bit}})^2 \cdot 2 \cdot R} = -77dBm \quad (eq. 4.12)$$

By substituting the values from equations (eq.4.1.11) and (eq.4.1.12) into (eq.4.1.10), the theoretical dynamics is given by:

$$REC\_DYN = 78dB$$

However from the board specific and from the literature the minimum detectable signal (ADC limited) is -55dBm (SNR=0dB) or -49dBm (SNR=6dB).

To mitigate potential nonlinearities in the receiver's electronic components, it is recommended not to use the full theoretical dynamic range. By assuming a 3 dB margin below the ADC's saturation point.

The effective dynamic range of the instrument becomes:

$$\text{BOARD\_DYNAMIC} = 53\text{dB}$$

Figure 4.2 illustrates the dynamic range of the signals involved (36dB in yellow), which the instrument must handle correctly, in relation to the theoretical dynamic range of the receiver (53dB in grey).

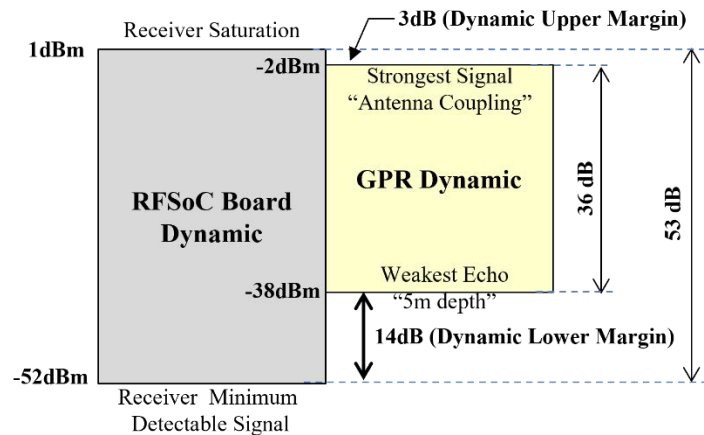


**Figure.4.2** Theoretical Dynamics of the Instrument



## 5 LINK BUDGET DEFINITION

In this section, the power levels of the various devices will be defined to ensure that the radar dynamic range of 36 dB will be well-suited with the receiver's dynamic range of 53dB.

Specifically, the goal is to match the maximum signal power generated by antenna coupling with the receiver's maximum input power limit, while maintaining a 3dB safety margin, as illustrated in Figure 4.2. This ensures that the dynamic range of the GPR is properly aligned with that of the receiver.

The input power to the ADC is given by the following equation:

$$P_{R,dBm} = P_{0,dBm} - LOSS_{TX\ CH,dB} - AC_{dB} - LOSS_{RX\ CH,dB} + G_{TX,dB} + G_{TR,dB} \quad (eq. 5.1)$$

where:

|                    |                                                                         |
|--------------------|-------------------------------------------------------------------------|
| $P_{0,dBm}$        | Power at the output of the DAC                                          |
| $P_{R,dBm}$        | Power input to the ADC                                                  |
| $LOSS_{TX\ CH,dB}$ | Total losses of the transmitting chain (positive sign indicates losses) |
| $LOSS_{RX\ CH,dB}$ | Total losses of the receiving chain (positive sign indicates losses)    |
| $AC_{dB}$          | Coupling between the two antennas (positive sign indicates coupling)    |
| $G_{TX,dB}$        | Gain of the TX amplifier.                                               |
| $G_{RX,dB}$        | Gain of the RX amplifier.                                               |

By expressing the equation in terms of the overall gain:

$$G_{TX,dB} + G_{RX,dB} = P_{R,dBm} - P_{0,dBm} + LOSS_{TX\ CH,dB} + AC_{dB} + LOSS_{RX\ CH,dB} \quad (eq. 5.2)$$

Assuming the following values as a first approximation:

$$\begin{aligned} P_R &= -2dBm \\ LOSS_{TX\ CHAIN} &= 2.25dB \\ LOSS_{RX\ CHAIN} &= 10.25dB \\ P_0 &= -9.5dBm \end{aligned}$$

The equation (eq.5.2) simplifies to:

$$G_{TX,dB} + G_{RX,dB} = -2dBm + 9.5dBm + 2.25dB + 30dB + 10.25dB = 50dB \quad (eq. 5.3)$$

The overall gain of the amplifiers is:

$$G_{Ampli} = 50dB$$

The estimated gain of 50dB will be distributed between a 25dB amplifier, characterized by a low noise figure installed at the output of the RX antenna, and an amplification stage of 25dB installed in the TX chain, as shown in the high-level scheme in Paragraph 6.8.

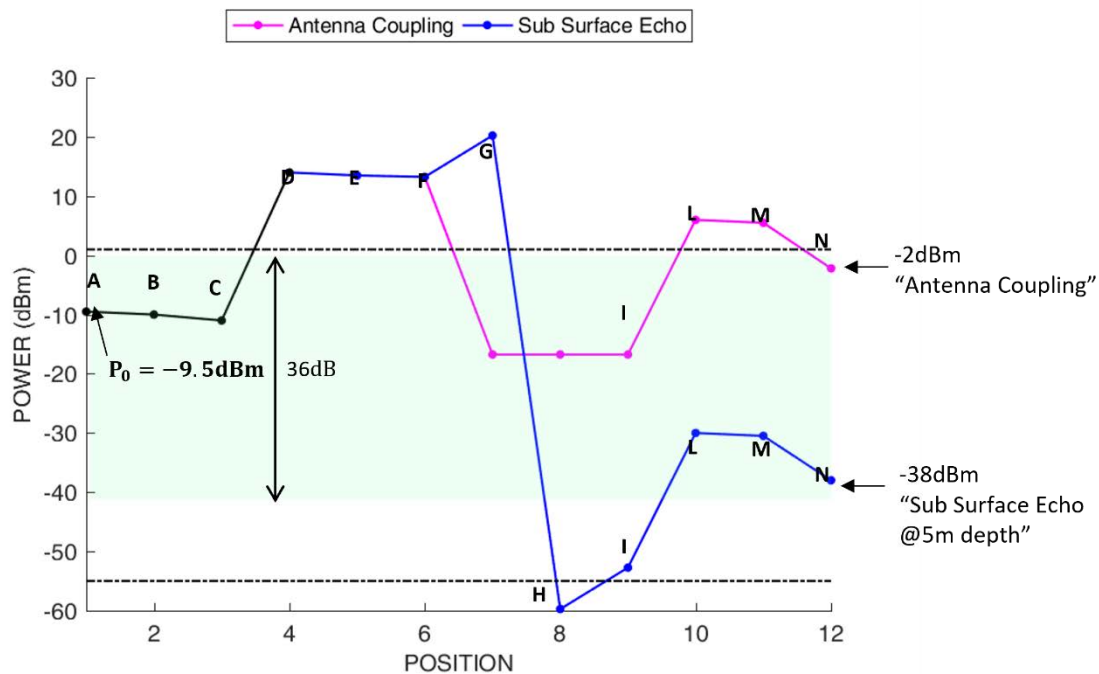


**Figure.5.1** Main power levels at the key points of the instrument.



## 6 RADAR SF-CW

The Stepped Frequency Continuous Wave (SF-CW) radars operate in the frequency domain. Unlike pulse radar systems, they do not directly generate the pulse in the time domain but instead synthesize it by transmitting its spectral components. The bandwidth of the synthetic pulse is defined as follows:

$$B_c = F_{MAX} - F_{MIN} \quad (eq. 6.1)$$

The range resolution in the vacuum is given by the following equation:

$$R_{d,vacuum} = \frac{c}{2 \cdot B_c} = 15cm \quad (eq. 6.2)$$

The range resolution in the medium is given by the following equation:

$$R_{d,medium} = \frac{c}{2 \cdot B_c \cdot \sqrt{\epsilon_r}} \sim 8.6cm \quad (eq. 6.3)$$

The radar generates and transmits a finite number of distinct sinusoids (tones) with constant amplitude and duration  $T_{tone}$ , spaced by discrete frequency steps  $\Delta_f$ , da  $F_{MIN}$  a  $F_{MAX}$ .

The duration of  $T_{tone}$  must be sufficiently long to ensure that the frequency change occurs only after the signal has travelled the necessary time to reach the target at the maximum detectable range and return to the receiver, along with other factors that will be discussed in the following paragraphs:

$$Z_{Max} = \frac{c \cdot T_{tone}}{2 \cdot \sqrt{\epsilon_r}} \quad (eq. 6.4)$$

Considering a wide margin for the unambiguous distance, for example of 85m in the medium (150m in a vacuum), this helps limit the effect of the Hermitian image, which is the unwanted echo positioned at the unambiguous distance minus the maximum expected delay.

$$T_{tone} \gg \frac{Z_{Max} \cdot 2 \cdot \sqrt{\epsilon_r}}{c} \rightarrow T_{tone} > 1\mu s \quad (eq. 6.5)$$

Below is shown the waveform of the transmitted signal and the main parameters of the SF-CW radar.

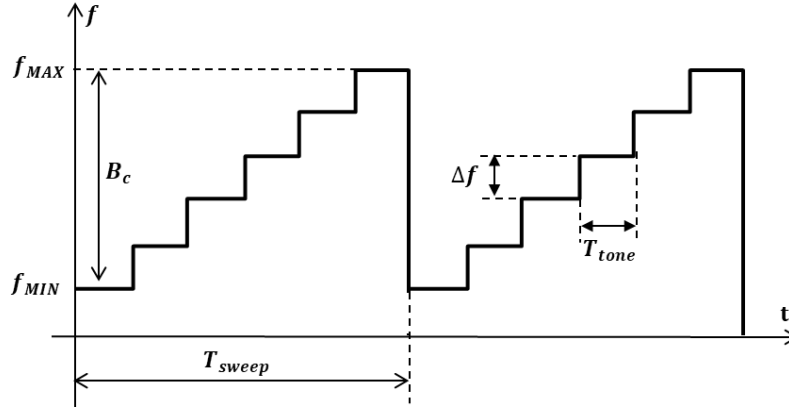


**Figure 6.1** Transmitted waveform of the SF-CW radar.

The receiver demodulates the collected signals by mixing it with a copy of the transmitted signal, sampling each tone in both amplitude and phase. Then, using the inverse Fourier transform, the data is converted from the frequency domain to the time domain, reconstructing the range profile. This output can then be compared to that of pulse radars. The time-domain response, obtained through pulse synthesis, is a sinc( $t$ ) function, with its peak positioned at the time delay  $\tau$ .

In SF-CW radars, the unambiguous range depends on the frequency step  $\Delta f$ . It is evident that, for higher unambiguous ranges, the frequency step will be small.

According to the sampling theorem, by synthesizing a pulse with a frequency step  $\Delta f$ , the target's response in the range profile (time domain) appears centered around the time delay and repeats at intervals of  $\Delta t$ , with:

$$\Delta t = \frac{1}{\Delta f} \quad (eq. 6.6)$$

Consequently, the unambiguous range is related to the frequency sampling step according to the following relation:

$$Z_{Max} = \frac{c \cdot \Delta t}{2} = \frac{c}{2 \cdot \sqrt{\epsilon_r} \cdot \Delta f} \quad (eq. 6.7)$$

From Equation 6.7, it is possible to derive the upper limit of the frequency step. By performing the calculation for an unambiguous distance of 85m in the medium (150m in a vacuum), the following limit is obtained:

$$\Delta f \leq \frac{c}{2 \cdot \sqrt{\epsilon_r} \cdot Z_{Max}} \rightarrow \Delta f = \mathbf{1MHz} \quad (eq. 6.8)$$

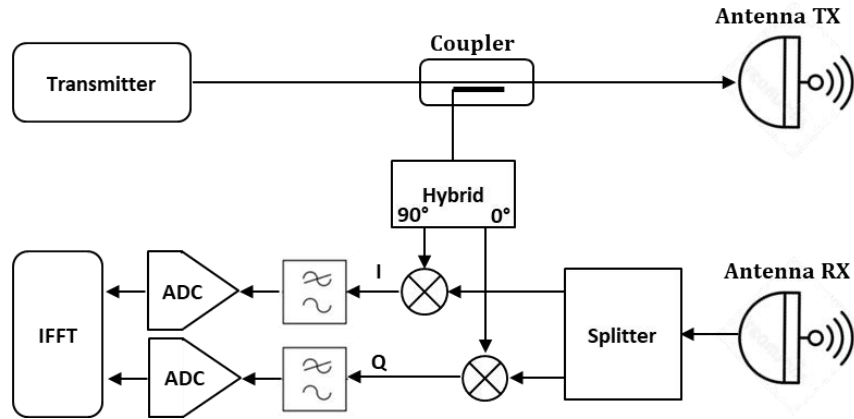


**Figure 6.2** General architecture for the Radar SW-CW

Another important aspect to evaluate is the side lobes of the strongest echoes, which can completely mask the weaker echoes. To limit this effect, common weighting functions are applied in the frequency domain, specifically before the IFFT operation that synthesizes the pulse, i.e., right after the ADC. For this type of GPR, the Chebyshev window seems to be the most suitable.

## 6.1 TRANSMITTER CHARACTERISTICS

A percentage deviation from the reference frequency results in the same percentage, change in: system's bandwidth, unambiguous range, range profile.

The SF-CW synthesizer must generate all the necessary frequencies to sample the system's bandwidth with constant frequency steps (Figure 6.1.1). However, regardless of the technology used, it is not possible to perform an instantaneous frequency change (frequency hopping). As shown in Figure 6.1.1, each transition requires a lock time ( $T_{lock}$ ), which does not contribute to the radar signal acquisition.

The integration time  $T_{int}$ , or the useful interval before the next frequency switch, affects the receiver's dynamic range. A longer integration reduces noise and increases the system's dynamic range due to the IFFT gain. The sum of  $T_{lock}$  e  $T_{int}$  defines the tone duration  $T_{tone}$  which, together with the delay time  $\tau$ , determines the system's temporal efficiency:

$$\eta = \frac{T_{int}}{T_{tone} + \tau} = \frac{T_{int}}{T_{lock} + T_{int} + \tau} \quad (eq. 6.1.1)$$

In GPR applications,  $\tau$  is generally negligible compared to  $T_{tone}$ , so the temporal efficiency depends solely on  $T_{int}$ . As  $T_{lock}$  increases, the efficiency tends to zero, limiting the scanning speed. Consequently, in high-speed acquisition systems,  $T_{lock}$  becomes a critical constraint.  $T_{int}$  also affects performance: while it improves the dynamic range, it also lengthens the scanning duration. The time required to complete an entire sweep  $T_{sweep}$  is given by:

$$T_{sweep} = N_{tones} \cdot T_{tone} = N_{tones} \cdot (T_{lock} + T_{int}) \quad (eq. 6.1.2)$$

A longer integration brings the temporal efficiency closer to one and improves the dynamic range, but it reduces the scanning speed.

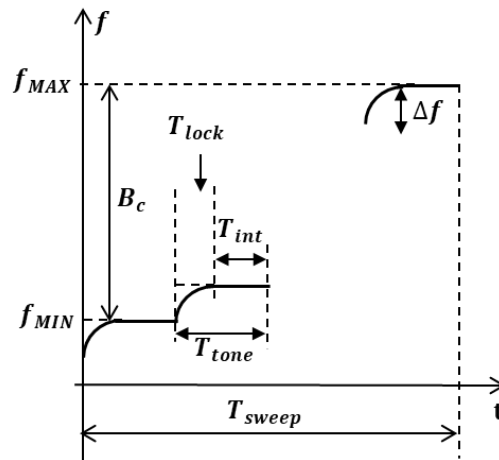


**Figure 6.1.1** Transients in the transmitted waveform



In the SF-CW transmitter, the harmonic distortion introduced by non-linear elements is particularly significant due to the high power levels (on the order of a few dBm). As a result, in addition to the desired tones  $[f_0:f_N]$ , the transmitter will also generate unwanted harmonic contributions  $[kf_0:kf_N]$ , with “k” being a positive integer. These terms, captured by the receiver, will lead to intermodulation products due to the demodulation process, so it is advisable to filter the signal before transmitting it through the antenna.

After low-pass filtering, the most significant contributions will come from the difference terms of the same factor “k”. This implies that both in-phase and quadrature components will contain the fundamental contribution, proportional to the target's delay time, as well as the respective harmonic contributions. The result is a distorted range profile containing both the fundamental pulse and its harmonic pulses, attenuated according to the same reduction factor present in the transmitter's harmonics.

From a radar perspective, this effect can be explained by considering that the harmonic components have a wavelength of  $[\lambda_i/k]$  for each frequency  $[f_i]$ , causing each target to appear at a distance “k” times greater than the actual position. Specifically, the second, third, and fourth harmonics have attenuations of approximately: 30, 50, and 70 dB compared to the fundamental.

The analysis conducted highlights the importance of strict control over harmonic distortion to avoid the formation of false targets in the range profile. However, the use of a superheterodyne structure and preselector filters can make the system much more robust against these issues.



## 6.2 RECEIVER CHARACTERISTICS

In the front-end and demodulator, harmonic distortion and phase and amplitude imbalance will inevitably be present. Analysing these phenomena is crucial to understand their impact on the system and the range profile.

Even a slight imbalance between the IQ components, typically around  $1^\circ$  in phase and less than 1dB in amplitude, can generate artefacts in the range profile. This is because the demodulated signal is not purely complex, but includes an unwanted real component.

In a single target scenario, the range profile will show the main impulse centered on the target's flight time, but also an artefact impulse positioned at a distance equal to the difference between the unambiguous range and the target's flight time. This artefact is known as the Hermitian image. Below are some suggestions to mitigate the problem:

- **Extend the unambiguous range:** The most immediate solution is to increase the unambiguous range. During the design phase, it is common practice to consider at least double the ambiguity compared to the maximum visible distance (determined by the system's dynamics). This way, the Hermitian image will be positioned at a non-relevant distance in the range profile.
- **Correct the imbalance via software:** A more complex alternative involves compensating for the imbalance through software. This technique relies on measuring, in the laboratory, the misalignment of the demodulator and subsequently correcting the acquired components by adding terms to adjust amplitude and phase.
- **Switch to a superheterodyne receiver:** Another solution, particularly effective when combined with software compensation, is the use of a superheterodyne receiver instead of a homodyne receiver, thereby improving the management of artefacts. This shields the system from harmonic distortion in the front-end but does not prevent distortion that may occur in the intermediate frequency section. Additionally, it reduces the occurrence of the Hermitian image, as it allows the demodulator to operate at a constant frequency, minimizing imbalances. Any remaining residues can be compensated in post-processing



### 6.3 DESIGN OF THE TRANSMITTED SF-CW SIGNAL

The number of tones to be generated is calculated using the following equation, given that:  $B_c = 1\text{GHz}$ :

$$N_{tones} = \frac{B_c}{\Delta_f} + 1 = 1001 \text{ (eq. 6.3.1)}$$

To make the inverse Fourier transform more efficient, the nearest power of 2 to the calculated value (1001) is chosen.

$$N_{tones} = 1024$$

With 1024 tones the frequency step becomes:

$$\Delta_f = 977\text{KHz}$$

A higher number of tones increases the unambiguous range and shifts potential Hermitian replicas to non-relevant range bins. Additionally, it improves the integration gain of the IFFT, enhancing the effective dynamic range and penetration depth. However, a large number of tones slows down the instrument. In applications such as subsurface lunar ice detection, scanning speed is secondary to resolution and depth of analysis.

The duration of  $T_{tone}$  is a key factor in reducing sweep times, which is crucial in high-speed scanning systems that require short sweeps while maintaining the same number of generated tones.

The value of  $T_{tone}$  depends on three main factors:

1. **Transmitter settling time**, which is related to the electronics used for tone generation.
2. **Delay time**, determined by the need to acquire stationary data only after receiving the echo from the deepest target. However, in GPR systems, the limited penetration depth makes this effect negligible.
3. **Integration time**, the most critical parameter, as it directly affects the radar's dynamic range. A longer integration time increases the IFFT gain in reconstructing the synthetic impulse, enhancing system sensitivity up to the noise floor of the instrument, beyond which further increases provide no additional benefit.

The transmitted signal will be:

$$P(t) = \sum_{i=0}^{N_f-1} e^{j2\pi(f_{MIN}+i\Delta f)t} \cdot \text{rect}\left(\frac{t - \frac{T_{tone}}{2} - iT_{tone}}{T_{tone}}\right) \quad (eq. 6.3.2)$$

If the transmitted signal is reflected by a single target with a reflection coefficient  $\Gamma$ , and neglecting the amplitude scaling given by the radar equation, the received signal  $s(t)$  can be approximated as a time-delayed copy of the transmitted signal.

$$S(t) = \Gamma \cdot P(t - \tau) = \sum_{i=0}^{N_f-1} \Gamma e^{j2\pi(f_{MIN}+i\Delta f)(t-\tau)} \cdot \text{rect}\left(\frac{t - \frac{T_{tone}}{2} - iT_{tone} - \tau}{T_{tone}}\right) \quad (eq. 6.3.3)$$

In baseband, after vector demodulation and low-pass filtering, the complex envelope of the in-phase and quadrature components can be expressed as:

$$S_b(t) = \sum_{i=0}^{N_f-1} \Gamma e^{-j2\pi(f_{MIN}+i\Delta f)\tau} \cdot \text{rect}\left(\frac{t - \frac{T_{tone}}{2} - iT_{tone}}{T_{tone}}\right) \quad (eq. 6.3.4)$$

The equation describes two sinusoids with a 90° phase shift, forming a stepped sinusoid with discrete voltage levels. These levels originate from radar measurements for each transmitted tone, with amplitude and phase determined by the target's reflection coefficient. The frequency of the sinusoid is proportional to the signal's delay  $\tau$  within the medium under investigation. To clarify the relationship between the frequency of the IQ components and the delay time, the following figure illustrates in black the transmitted sweep and in red the sweep reflected from a target at a distance  $\tau$ .

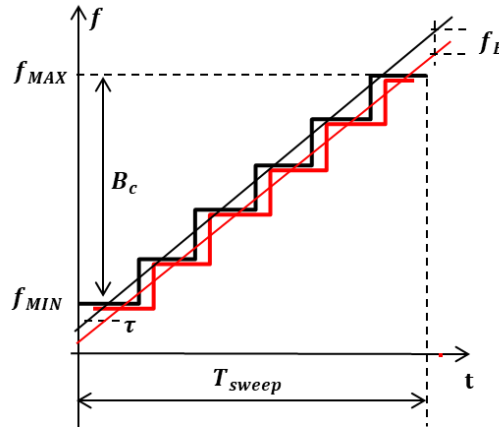


**Figure 6.3.1.** Transmitted and Received Signal in SF-CW Radar



Sampling the received signal at the centre of each tone, a linear ramp can be plotted that represents the transmitted and received sweeps. This allows highlighting the frequency difference  $f_B$  between the two sweeps, and, using the similarity of the triangles, the corresponding relation can be expressed as follows:

$$\frac{f_B}{B_c} = \frac{\tau}{T_{SWEEP}} \quad (eq. 6.3.5)$$

The frequency of the IQ components can be derived by substituting the time delay with the target distance.

$$f_B = B_c \cdot \frac{2 \cdot Z_{Max} \cdot \sqrt{\epsilon_r}}{c \cdot T_{sweep}} \quad (eq. 6.3.6)$$

Sampling the signal (eq. 6.3.4) at the central point of each tone, the resulting vector can be expressed through the following relation:

$$s[k] = \Gamma e^{-j2\pi(f_{MIN} + k\Delta f) \cdot \tau} \quad (eq. 6.3.7) \quad k = 0, 1, 2, \dots, N_f - 1 \quad (eq. 6.3.8)$$

Writing the equation as a function of the k-th transmitted frequency, we have:

$$S(f_k) = \Gamma e^{-j2\pi f_k \cdot \tau} \quad (eq. 6.3.9) \quad f_k = (f_{MIN} + k\Delta f) \quad (eq. 6.3.10)$$

Applying the inverse Fourier transform to the signal, the synthetic response in the time domain is obtained. This is expressed as a function of the central frequency  $f_c$  of the bandwidth  $B_c$ , according to the following relationship:

$$\hat{S}(t) = IDFT\{S(f_k)\} = \Gamma B_c \frac{\sin(\pi B_c(t - \tau))}{\pi B_c(t - \tau)} \cdot e^{j2\pi f_c(t - \tau)} \quad (eq. 6.3.11)$$

$$f_c = \frac{f_{MIN} + f_{MAX}}{2} \quad (eq. 6.3.12)$$

This synthetic impulse is a sinc(t) function, proportional to the bandwidth  $B_c$  of the transmitted signal and centered on the delay time.

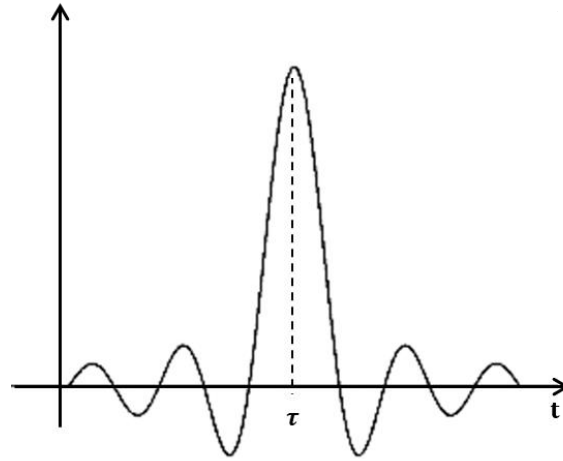


**Figure 6.3.2** SF-CW range profile centered at delay  $\tau$  after IFFT

## 6.4 CONSIDERATIONS ON HOMODYNE AND HETERODYNE DEMODULATION

The homodyne architecture, while keeping the hardware complexity limited, is susceptible to undesirable effects such as harmonic distortion and phase and amplitude imbalance. Moving from a homodyne receiver to a superheterodyne receiver increases the complexity of the RF front end, but leads to improvements in both the demodulator imbalance and harmonic distortion. Essentially, this involves modifying the receiver architecture by introducing an intermediate frequency stage  $f_{IF}$  in the front end. This stage consists of a wideband mixer and a narrowband filter centered on  $f_{IF}$ . The conversion occurs by mixing the received tones  $f_0 \dots f_{N-1}$  with tones that are frequency-shifted by a constant value,  $f_{IF}$ , relative to the received ones, as shown in Figure 6.4.1.

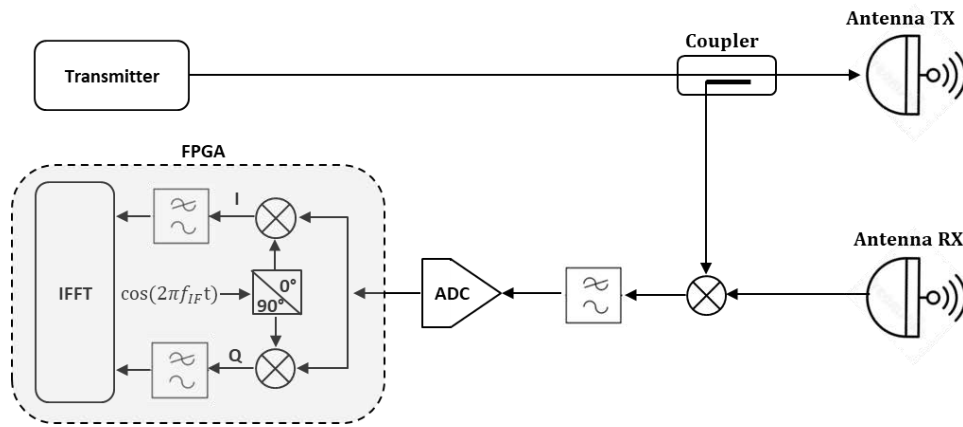


Figure 6.4.1. Superheterodyne Demodulation Architecture

The first advantage of this solution concerns the demodulator: the mixing process converts all the received tones to the intermediate frequency  $f_{IF}$ , allowing the demodulator to operate at a constant frequency. Although there is still a slight imbalance in this case, a significant reduction in Hermitian distortion is achieved compared to the homodyne architecture.

Moreover, there are also benefits in terms of harmonic distortion in the front end. In fact, with this architecture, the only signals remaining downstream of the IF filter are the useful ones, expressed as  $(f_i + f_{IF}) - f_i$ , while higher-order signals, with the formation of  $k$  replicas at a distance  $k$  times that of the target, are automatically eliminated. As a result, the superheterodyne structure protects against nonlinearities present in the front end and at the intermediate frequency mixer port, preventing the generation of artefacts in the range profile.

In summary, the superheterodyne receiver protects against harmonic distortions in the front end, but not those that may arise in the intermediate frequency section. Additionally, it reduces the occurrence of Hermitian image distortion, as it allows the demodulator to operate at a constant frequency, minimizing imbalances. Any residual distortions can be compensated for during the post-processing stage.

It has been decided to use an Intermediate Frequency of  $f_{IF} = 2.5\text{MHz}$  so the mixing in the mixer will not be done with a mirrored copy of the transmitted signal but with a frequency-shifted replica by a constant factor of 2.5MHz. The beat frequency will range from (0.5GHz : 1.5GHz) to (0.5GHz+2.5MHz : 1.5GHz+2.5MHz). The IF bandwidth will be a function of the number of transmitted tones (1024) and the duration of the tone, as shown in the following figure:

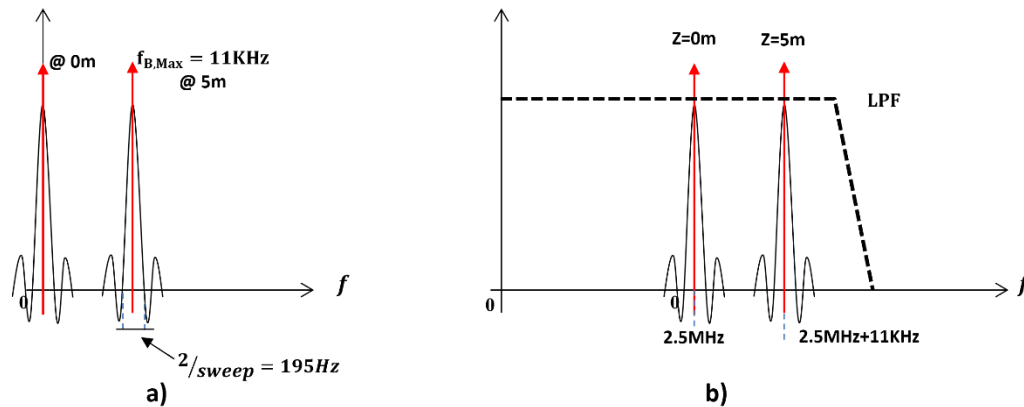


**Figure 6.4.2** Bandwidth occupation after demodulation. (a) Homodyne demodulation. (b) Superheterodyne demodulation

## 6.5 RECEIVER NOISE FIGURE ESTIMATION

This section defines the maximum acceptable noise figure  $F_N$  for the receiver to meet the required  $SNR_{CMP}$  after processing (FFT Gain).

The SNR is imposed as requirement on a weaker signal, that is the one from 5m depth, as following:

$$SNR_{CMP,dB} = P_{SS,5m,dB} - N_{pk,dB} (eq. 6.3.1)$$

$N_{pk,dB}$ : The pick noise level

$P_{SS,5m,dB}$ : Sub-surface reflection at depth of 5m

Making explicit for the noise, we obtain the following:

$$N_{pk,dB} = P_{SS,5m,dB} - SNR_{CMP,dB} (eq. 6.3.2)$$

After processing, the power of the signal is increased by the FFT compression factor, as reported by the following equation:

$$N_{pk,dB} = (P_{SS,5m,dB} + RC_{dB}) - SNR_{cmp,dB} (eq. 6.3.3)$$

$$RC_{dB} = 10 \cdot \log_{10} N_{tones} = 30dB$$

The conversion from effective noise to peak-to-peak noise is a simple multiplication by 6.6. In our case, the conversion is reversed, meaning the effective noise value is obtained by dividing the peak noise by 6.6 or, equivalently, by subtracting 16dB (since  $20 \log_{10}(6.6) = 16dB$ ) from the peak noise value expressed on a logarithmic scale.

$$N_{pk} = 6.6 \cdot N_{RMS} (eq. 6.3.4)$$

$$N_{RMS,dB} = N_{pk,dB} - 20 \cdot \log_{10} 6.6 (eq. 6.3.5)$$

$$N_{RMS,dB} = N_{pk,dB} - 16 (eq. 6.3.6)$$

$$N_{RMS,dB} = (P_{SS,5m,dB} + RC_{dB}) - SNR_{CMP,dB} - 16 (eq. 6.3.7)$$

The RMS noise power level, can be expressed as following:

$$N_{RMS} = K \cdot T \cdot B_N \cdot NF (eq. 6.3.8)$$



$B_N$  is the equivalent noise bandwidth, determined by the low-pass filter placed after the intermediate frequency mixer.

$$(K \cdot T)_{dB} + B_{N,dB} + NF_{dB} = (P_{SS,5m,dB} + RC_{dB}) - SNR_{CMP,dB} - 16 \text{ (eq. 6.3.9)}$$

In an SF-CW radar, the equivalent noise bandwidth is determined by the low-pass filter placed after the intermediate frequency mixer. For a first-order filter with a time constant  $\tau$  and a cut-off frequency  $f_t$ , such that, The equivalent noise bandwidth at intermediate frequency is given by:

$$B_N = \frac{\pi}{2} \cdot f_{t,IF} \text{ (eq. 6.3.10)}$$

Since we will not use a homodyne demodulator but a superheterodyne intermediate frequency demodulator, with  $f_{IF} = 2.5MHz$ , and considering that the beat frequency typically lies in the kHz range, we have chosen a low-pass filter after the mixer with a cut-off frequency of approximately  $f_{t,IF} = 5MHz$ .

Where  $f_t$  represent the cut-off frequency of the low pass filter after the Mixer. The equation (6.3.9) becomes:

$$(K \cdot T)_{dB} + \left(\frac{\pi}{2}\right)_{dB} + f_{t,dB} + NF_{dB} = (P_{SS,5m,dB} + RC_{dB}) - SNR_{cmp,dB} - 16 \text{ (eq. 6.3.11)}$$

Expressing  $P_{SS,5m,dB}$  in dBm we obtain:

$$(K \cdot T)_{dB} + \left(\frac{\pi}{2}\right)_{dB} + f_{t,dB} + NF_{dB} = (P_{SS,5m,dBm} - 30 + RC_{dB}) - SNR_{CMP,dB} - 16 \text{ (eq. 6.3.12)}$$

It now possible to calculate the noise figure:

$$NF_{dB} = P_{SS,5m,dBm} - 30 + RC_{dB} - SNR_{CMP,dB} - 16 - (K \cdot T)_{dB} - \left(\frac{\pi}{2}\right)_{dB} - f_{t,dB} \text{ (eq. 6.3.13)}$$



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$$NF_{dB} = P_{SS,5m,dBm} - 46 + RC_{dB} - SNR_{Cmp,dB} - \left( \frac{K \cdot T \cdot \pi}{2} \right)_{dB} - f_{t,dB} (eq. 6.3.14)$$

Assuming the following values:

$$\begin{aligned} P_{SS,5m,dBm} &= -70dBm \\ RC_{dB} &= 30dB \\ SNR_{Cmp,dB} &= 15dB \\ T &= 290K \\ f_{t,dB} &= 69dB \\ K &= 1.38e-23 \text{ (The Boltzmann constant)} \\ T_e &= 290K \text{ (Receiver noise temperature)} \end{aligned}$$

$$NF_{dB} = 52dB$$

This seemingly absurd value indicates that the receiver's noise figure will not negatively impact the instrument's dynamic range. This is primarily due to the relatively shallow investigation depths (only a few meters below the surface) and the high value of the integration gain of the IFFT.



## 6.6 DEFINITION OF TONES DURATION

By setting the equivalent noise bandwidth for the video filter on the in-phase and quadrature components (and thus for the receiver), it allows the calculation of the instrument's temporal efficiency and, consequently, setting the duration of each acquisition. Still assuming a first-order filter for the filter after the MIXER, we can estimate the settling time in 5 time constants  $\tau$ , which in this case corresponds to a time of approximately:

$$f_t = \frac{1}{2 \cdot \pi \cdot \tau} \quad (\text{eq. 6.6.1})$$

Considering the cut-off frequency of 2.75 MHz, we obtain:

$$\tau = \frac{1}{2 \cdot \pi \cdot f_t} = 58\text{ns} \quad (\text{eq. 6.6.2})$$

Five times  $\tau$  corresponds to 0.3  $\mu\text{s}$ , which would be the  $T_{lock}$ . Considering a good compromise between the acquisition duration and the integration time, a temporal efficiency of 0.8 has been set.

$$\eta = \frac{T_{int}}{T_{lock} + T_{int} + \tau} \approx \frac{T_{int}}{T_{lock} + T_{int}} = 0.8 \quad (\text{eq. 6.6.3})$$

$$T_{int} > \frac{\eta}{1 - \eta} \cdot (T_{lock} + \tau) \quad (\text{eq. 6.6.4})$$

Considering  $\tau = 115\text{ns}$  (target at 10m of depth)

$$T_{int} > 1.6\mu\text{s}$$

Considering an adequate margin, the following value for the  $T_{tone}$  will be chosen:

$$T_{tone} = 10\mu\text{s}$$

It is therefore possible to calculate the duration of the Sweep using the following relation:

$$T_{sweep} = N_{tones} \cdot T_{tone} \rightarrow T_{sweep} = 10.24\text{ms} \quad (\text{eq. 6.6.5})$$

The expected Maximum beat frequency is:

$$f_{B,Max(5m)} = 5.6\text{ KHz}$$

## 6.7 ELECTRONIC SYSTEM CALIBRATION

Radar calibration is necessary because the frequency response of electronic systems varies in phase and amplitude, affecting the acquired data. Since radar samples contain both the response of the scenario and that of the instrument, it is essential to correct these effects in order to obtain an accurate characterization of the analysed environment.

This is possible by including a calibration unit, selectable via appropriate RF switches, as shown in Figure 6.7.1.

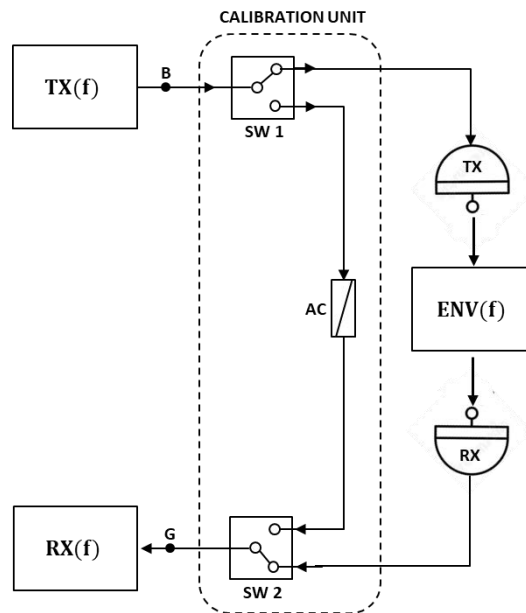


**Figure 6.7.1** SF-CW Radar Calibration Unit

From an analytical point of view, the entire system can be described by the following mathematical functions:

- $TX(f)$  Transmitter frequency response (depends on the electronics)
- $RX(f)$  Receiver frequency response (depends on the electronics)
- $AC$  Attenuator (can be considered constant with the frequency)
- $ENV(f)$  Scenario frequency response



By keeping the switches on the antennas, the entire system response is obtained:

$$\mathbf{ACQ}(f) = TX(f) \cdot ENV(f) \cdot RX(f) \text{ (eq. 6.7.1)}$$

Switching on the calibrator, the frequency response obtained from this configuration is described by the following function:

$$\mathbf{CAL}(f) = TX(f) \cdot AC \cdot RX(f) \text{ (eq. 6.7.2)}$$

Combining the two equations described above, it can be demonstrated that the calibration algorithm described in the following equation corrects the effect of the transmitter and receiver frequency responses, except for an amplitude scaling equal to the attenuation value on the calibration path.

$$\frac{ACQ(f)}{CAL(f)} = \frac{TX(f) \cdot ENV(f) \cdot RX(f)}{TX(f) \cdot AC \cdot RX(f)} = \frac{ENV(f)}{AC} \text{ (eq. 6.7.3)}$$

$$\frac{ACQ(f)}{CAL(f)} = \frac{ENV(f)}{AC} \text{ (eq. 6.7.4)}$$

This correction significantly improves the output of the ground-penetrating radar, however, it is not perfect because it does not take into account the effect of the cables connecting the antennas to the calibration switches, nor does it consider the effect of the antennas. This is because both the antennas and the cables are external to the calibration path. It will be also necessary to estimate the delays within the instrument, to be estimated with special test to be conducted with the complete instrument in a controlled environment.

## 6.8 LABORATORY TEST

This paragraph aims to describe the tests conducted in a controlled environment, essential for the initial characterization of the instrument's performance, without the antennas connected. Figure 6.8.1 illustrates the architecture of the TEST UNIT block.

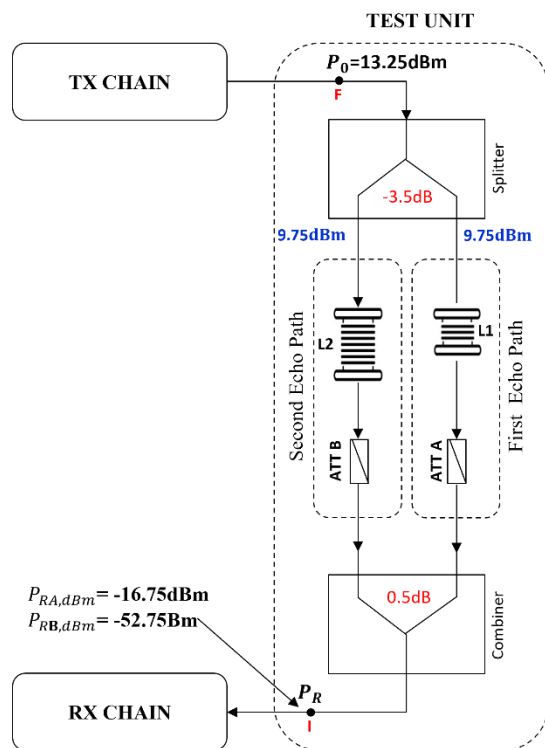


**Figure 6.8.1** SF-CW Laboratory Test Configuration

Depending on the configuration of the two attenuators (ATT\_A and ATT\_B) and the lengths of the cables (L1 and L2) used to simulate echo delays, the following three tests will be conducted:

Assuming:

$CAtt_{\text{dB/m}} = 0.65\text{dB/m}$  Estimation of the RG58 cable loss, to be validated in laboratory

$SLoss_{\text{dB}} = 3.5\text{dB}$  Splitter loss, 0.5dB insertion loss and value in dB of the division by two of the Power

$CLoss_{\text{dB}} = 0.5\text{dB}$  Combiner Insertion loss



### Test 1: Full Dynamic Performance Test

The purpose of this test is to evaluate the instrument's performance at 36dB of dynamic range, measured as the difference between the maximum signal generated by antenna coupling and the echo reflected by a subsurface discontinuity located at a depth of 5 meters.

The length of cable L1 is set to 0m to simulate the antenna coupling, while the length of cable L2 is set to 10m, that is the double of the nominal depth of 5m, to simulate the delay of the Echo.

the attenuators: ATT\_A and ATT\_B are calculated by the following equations, setting  $P_{RA,dBm} = -16.75dBm$  (power level at the receiver due to the antennas coupling) and  $P_{RB,dBm} = -52.75dBm$  (power level at the receiver due to the echo coming from 5m depth).

$$Att_{A/B_{dB}} = P_{0,dBm} - S_{Loss_{dB}} - L_{2m} \cdot C_{Att_{dB/m}} - C_{Loss_{dB}} - P_{RB,dBm} (eq. 6.8.1)$$

### Test 2: Instrument Resolution Assessment

The aim of this test is to verify that the instrument is able to distinguish two objects at a mutual distance of 15cm in a vacuum. The attenuators could be considered identical, while it is important to have a difference of 15cm between cable L1 and cable L2

The following table summarize the configuration of the attenuators and cables length of the three tests described above

| Device      | Test 1 | Test 2                         |
|-------------|--------|--------------------------------|
| L1          | 0      | L2-L1 = 15cm                   |
| L2          | 10     |                                |
| $AttA_{dB}$ | 26     | $AttA_{dB} = AttB_{dB} = 40dB$ |
| $AttB_{dB}$ | 54     |                                |

**Table 6.8.1** Devices configuration for the laboratory tests

## 7 RADAR FM-CW

This paragraph, outlines the radar design, based on Frequency Modulated Continuous Wave (FM-CW), which essentially consists of the beat between the received signal and a copy of the transmitted one. The beat frequency is proportional to the delay time  $\tau$  of the received echo, as shown in the following Figure 7.1

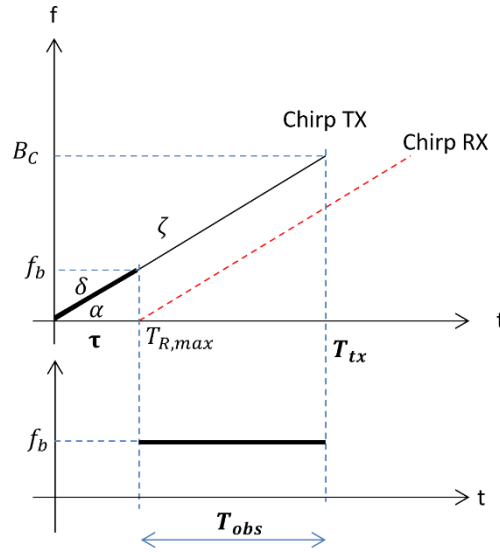


**Figure 7.1** FM-CW Radar

The radar resolution depends on the transmitted bandwidth. To achieve a resolution of 8.6cm in the luna subsoil ( $\epsilon_{r1} = 3$ ), or 15cm in a vacuum, the required bandwidth can be obtained from the following equation:

$$B_c = \frac{c}{2 \cdot R_d \cdot \sqrt{\epsilon_{r1}}} \rightarrow B_c = \mathbf{1GHz} \text{ (eq. 7.1)}$$

Also, for the FM-CW radar, the resolution does not depend on the duration of the transmitted pulse, but only on the transmitted bandwidth. However, the pulse duration must be much greater than the maximum expected echo delay, as highlighted by the following relation:

$$T_{R,max} = \frac{2 \cdot Z_{max} \cdot \sqrt{\epsilon_{r1}}}{c} = 58ns \rightarrow T_{tx} \gg 58ns \text{ (eq. 7.2)}$$

Having assumed:

$Z_{max} = 5\text{m}$                       Maximum penetration depth  
 $\epsilon_{r1} = 3$                               Dielectric permittivity of the mars sub-soil

Another important parameter, to take into account during the development of a FM-CW radar, is the slope (S) which the signal generator must be able to support, in order to modulate the required bandwidth. From Figure 7.1 it is easy to demonstrate the following relationship between the slope S and the other keys parameters:

$$S = \frac{B_c}{T_{tx}} = \frac{f_b}{\tau} \text{ (eq. 7.3)}$$

Where:

$B_c$ :                      Radar Band  
 $R_d$ :                      Range resolution  
 $f_b$ :                      Beat Frequency  
 $\tau$ :                        Time delay of the received echo  
 $T_{tx}$ :                    Duration of the transmitted signal  
 $\epsilon_{r1}$ :                    Dielectric permittivity of the mars sub-soil

The slope of the RFSoc (4x2) board can be configured within the following range:

$$S = 10^{13}: 10^{14} \left[ \frac{\text{Hz}}{\text{sec}} \right] = 10\text{MHz}/\mu\text{s} \div 100\text{MHz}/\mu\text{s}$$

To avoid excessive stress on the board, the lower value will be used for this project:

$$S = 10^{13} \left[ \frac{\text{Hz}}{\text{sec}} \right] = \mathbf{10\text{MHz}/\mu\text{s}}$$

Consequently, the duration of the transmitted signal is:

$$T_{tx} = \frac{B_c}{S} \rightarrow \mathbf{T_{tx} = 100\mu\text{s}} \text{ (eq. 7.4)}$$

100μs is much greater than 58ns, so the condition of equation 7.2 is fully satisfied. The transmitted signal waveform will therefore have the shape shown in Figure 7.2

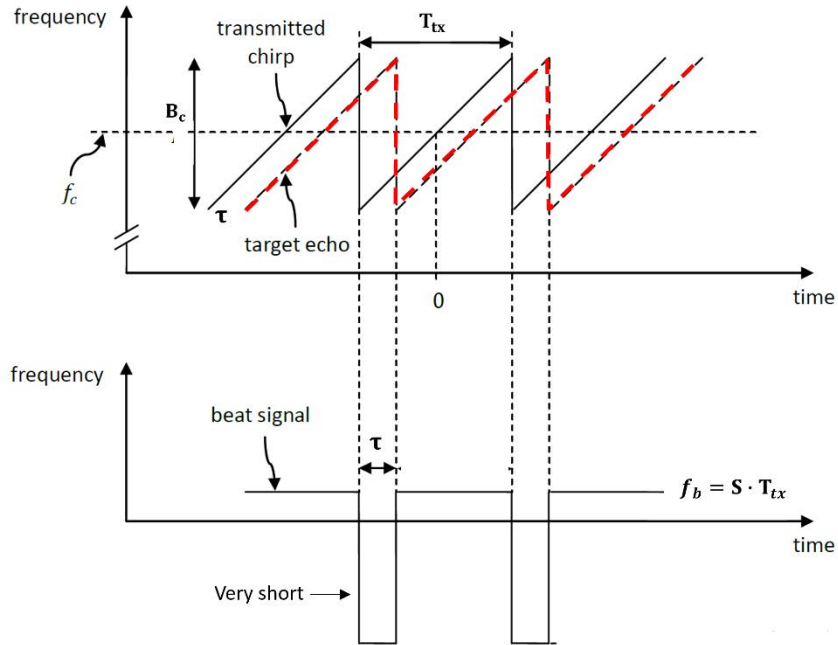


Figure 7.2 FM-CW Radar

The beat frequency, at the output of the mixer can be calculated by equation 7.3:

$$f_b = \frac{\tau}{T_{tx}} \cdot B_c \text{ (eq. 7.5)}$$

The maximum beat frequency, occurs for the echo coming from the maximum depth of 5m and given by the following equation:

$$f_{b,sub-surf} = \left( \frac{2 \cdot Z_{max} \cdot \sqrt{\epsilon_{r1}}}{c} \right) \cdot \frac{B_c}{T_{tx}} \rightarrow f_{b,max} = 0.6 \text{ MHz (eq. 7.6)}$$

The beat frequency due to antenna coupling is zero, since the time delay is negligible

$$f_{b,coupling} = 0 \text{ MHz}$$

The frequency increment  $\Delta f$  at each resolution step is given by the following equation:

$$\Delta f = \frac{2 \cdot R_d \cdot B_c \cdot \sqrt{\epsilon_{r1}}}{T_{tx} \cdot c} \rightarrow \Delta f = 10 \text{ KHz (eq. 7.7)}$$

Where  $R_d$  is the resolution in the regolith which is 8.6cm.

All these considerations are illustrated in Figure 7.3(a). However, the beat frequency is not an exact copy of the transmitted signal, but a version shifted in frequency by a certain value  $f_{IF}$ , as shown in Figure 7.3(b) and explained in the previous sections.

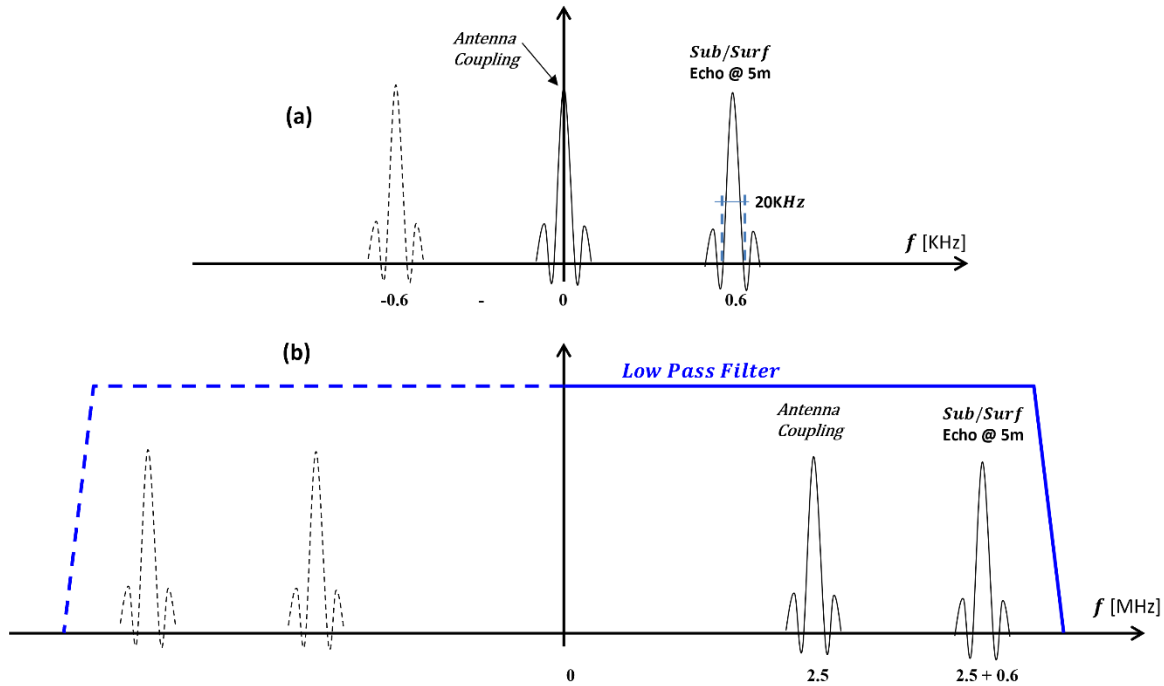


Figure 7.3 FM-CW Radar

After the demodulation process, carried out by the FPGA at the digital level, the frequencies return to their original positions, as shown in Figure 7.3(a)

### BEAT FREQUENCY AT THE MIXER OUTPUT

The following figures shows the Fourier Transform of a simple rectangle that represents the finite duration of the sinusoid (beat frequency) which maximum duration is:

$$T = T_{obs} = T_{tx} - \tau = 100\mu s - 58ns \sim 100\mu s \text{ (eq. 7.8)}$$

The bandwidth occupancy, limited to the first two nulls of the Fourier transform, is:

$$B_{base} = \frac{2}{T_{obs}} \rightarrow B_{base} = 20KHz \text{ (eq. 7.9)}$$



**Figure 7.4** Fourier Transform of a rectangle



## 8 INSTRUMENT SUMMARY TABLE

| Summary Table                   |              |        |                                                                     |
|---------------------------------|--------------|--------|---------------------------------------------------------------------|
| Parameter                       | Value        | UM     | Description                                                         |
| <b>ENVIRONMENT</b>              |              |        |                                                                     |
| $\tan\delta$                    | 0.003: 0.007 |        | Dielectric loss factor of the lunar soil                            |
| $\epsilon_{r1}$                 | 3            |        | Dielectric permittivity of lunar regolith                           |
| $\epsilon_{r2}$                 | 5            |        | Dielectric permittivity of the mixture: Regolith & Ice              |
| $L_{SS/m}$                      | 2            | dB     | Two way travel time attenuation in the lunar Sub-Soil               |
| $EnvAtt_{5m}$                   | 80           | dB     | Total environment attenuation at depth of 5m                        |
| <b>INSTRUMENT REQUIREMENT</b>   |              |        |                                                                     |
| $R_d/vacum$                     | 15           | cm     | Vertical resolution in vacuum                                       |
| $R_{d,\epsilon_{r1}=3}$         | 8.6          | cm     | Vertical resolution in the medium                                   |
| $INS\_DYN_{@5m}$                | 36           | dB     | Instrument Dynamic for a 5m Detection Range                         |
| <b>GENERAL RADAR PARAMETERS</b> |              |        |                                                                     |
| $f_c$                           | 1(30cm)      | GHz    | Carrier frequency                                                   |
| $G_{Ant}$                       | 7            | dB     | Antennas Gain (TBC)                                                 |
| AC                              | 30           | dB     | Antennas Coupling (TBC)                                             |
| $LOSS_{TX CHAIN}$               | 2.25         | dB     | Total losses of the transmitting chain (TBC)                        |
| $LOSS_{RX CHAIN}$               | 10.25        | dB     | Total losses of the receiving chain (TBC)                           |
| $G_{Ampli}$                     | 50           | dB     | Total gain of the instrument to ensure the desired performance.     |
| $N_{0,RMS}$                     | -90          | dBm    | RMS noise level of the Receiver                                     |
| $N_{F,Max}$                     | 52           | dB     | Maximum Noise Figure acceptable by the receiver                     |
| <b>SF-CW PARAMETERS</b>         |              |        |                                                                     |
| $R_{NA/vacum}$                  | 150          | m      | Unambiguous range in a vacuum, for radar design                     |
| $R_{NA/\epsilon_{r1}=3}$        | 86           | m      | Unambiguous range in the medium, for radar design                   |
| $\Delta f$                      | 977          | KHz    | Radar Frequency Steps                                               |
| $N_{toni}$                      | 1024         |        | Number of tones in the SF-CW radar                                  |
| $f_{IF}$                        | 2.5          | MHz    | Intermediate frequency of the instrument                            |
| $f_{t,IF}$                      | 5            | MHz    | Cut-off frequency of the intermediate frequency pass filter         |
| $T_{tone}$                      | 10           | us     | Single tone duration                                                |
| $T_{sweep}$                     | 10.24        | ms     | Sweep Duration                                                      |
| <b>FM-CW PARAMETERS</b>         |              |        |                                                                     |
| S                               | 10           | MHz/us | RFSoc(4x2) Slope                                                    |
| $T_{tx}$                        | 100          | us     | Duration of the transmitted Chirp                                   |
| $f_{b,max}$                     | 0.6          | MHz    | Maximum beat frequency, in base band, due to echo from 5m depth     |
| $\Delta f$                      | 10           | KHz    | Frequency increment at each resolution step of 8.6cm in the subsoil |
| $f_{IF}$                        | 2.5          | MHz    | Intermediate frequency of the instrument                            |



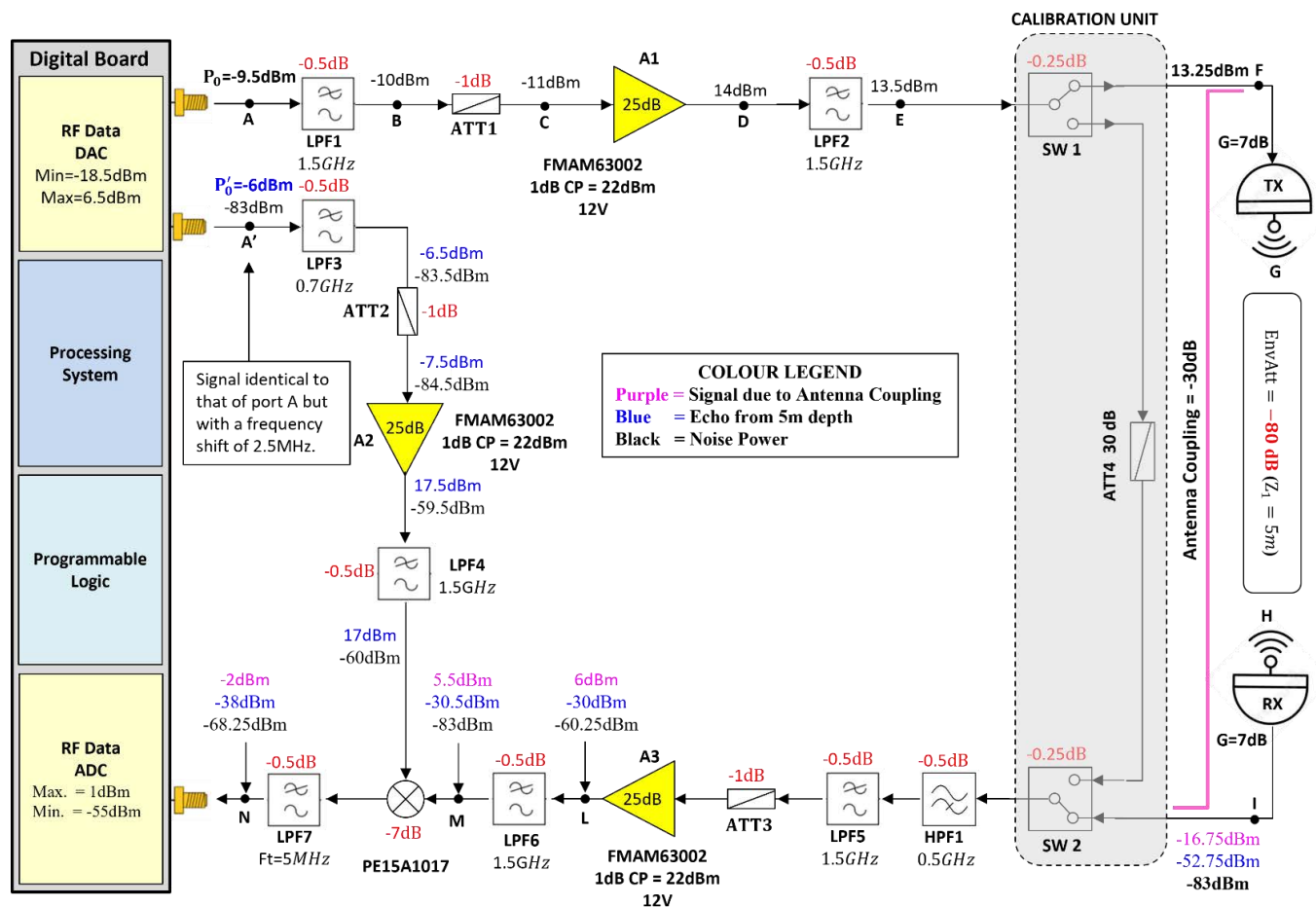
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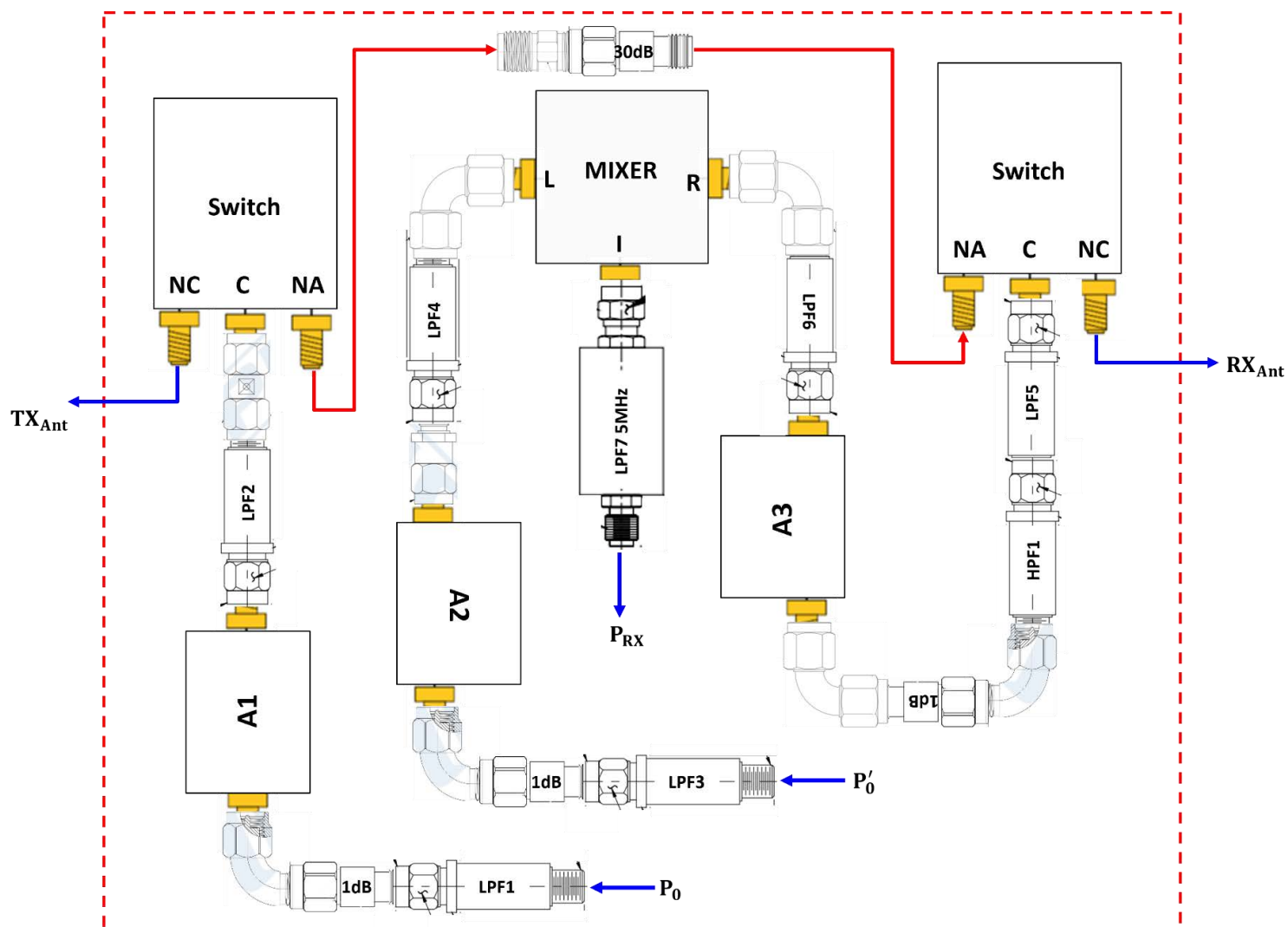
### DIGITAL BOARD CONFIGURATION (TBC)

|                       |      |      |                                                                   |
|-----------------------|------|------|-------------------------------------------------------------------|
| $V_{pp}$              | 1    | Volt | Nominal input voltage of the ADC (TBV)                            |
| $N_{bit}$             | 14   |      | Number of bits of the ADC conversion                              |
| $P_{in\_FULL\_SCALE}$ | 1    | dBm  | Full-scale input power of the ADC                                 |
| $P_{in\_MIN}$         | -55  | dBm  | Minimum input power of the ADC                                    |
| $P_{R,ADC,Max}$       | -2   | dBm  | Maximum input power to the ADC due to Antenna Coupling            |
| $P_{R,ADC,5m}$        | -38  | dBm  | Maximum input power to the ADC from the 5m depth of discontinuity |
| $P_0$                 | -9.5 | dBm  | Nominal power provided by the DAC (TBC)                           |

## 9 RADAR SW-CW HIGH LEVEL ELECTRONICS



## 9.1 INSTRUMENT PRELIMINARY LAYOUT





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## 10 SELECTION OF COMPONENTS

### AMPLIFIER: 10 dB

| Model              | Distributore          | Costo € | Peso gr | Band          | NF dB | G dB | G flat dB | Reverse Isol. S12 dB | Direct. dB | VSWR In | VSWR Out | @1 dB CP dBm | 3rd dBm | V volt | I mA | Pwr watt |
|--------------------|-----------------------|---------|---------|---------------|-------|------|-----------|----------------------|------------|---------|----------|--------------|---------|--------|------|----------|
| ZX60-3011+         | mini-circuits         | 190     | 23      | 0.4GHz:3GHz   | 1.75  | 14.8 | +/-1      | NA                   | 10.3       | 1.67    | 1.66     | 21.48        | 31.33   | 12     | 120  | 1.4      |
| SLNA-030-15-25-SMA | Fairview Microwave    | 133     | ?       | 0.5GHz : 3GHz | 2:3   | 14.5 | +/-0.4    | NA                   | NA         | 1.93    | 1.93     | 20           | 35      | 5:12   | 90   | 0.45     |
| ABL0200-10-1420    | WENTEQ Microwave Corp | 394     | NA      | 0.2GHz : 2GHz | 2     | 14   | +/-0.75   | 20                   | NA         | 1.5     | 1.5      | 21           | 40      | 12     | 90   | 1        |
| ZX60-P103LN+       | Mini-circuits         | 114\$   | 23      | 50MHz : 3GHz  | 0.55  | 15   | NA        | NA                   | 6.3        | 1.6     | 1.5      | 22           | 40      | 5      | 95   | 0.5      |

### AMPLIFIER: 20 dB

| Model           | Distributore          | Costo € | Peso gr | Band          | NF dB   | G dB  | G flat dB | Reverse Isol. S12 dB | Direct. dB | VSWR In | VSWR Out | @1 dB dBm | 3rd dBm | V volt | I mA  | Pwr watt |
|-----------------|-----------------------|---------|---------|---------------|---------|-------|-----------|----------------------|------------|---------|----------|-----------|---------|--------|-------|----------|
| ZX60-3018G-S+   | mini-circuits         | 126     | 23      | 20MHz:3GHz    | 2.64    | 21.83 | NA        | NA                   | 3.03       | 1.31    | 1.39     | 12.76     | 25.66   | 12     | 34/45 | 0.41     |
| PE15A63016      | Pasternack            | 1040    | NA      | 20MHz : 3GHz  | 1.7/2   | 25    | +/-0.75   | 40/45                | NA         | 1.5/1.8 | 1.6/1.8  | 12/13     | 20/23   | 12     | 75    | 0.9      |
| PE15A1008       | Pasternack            | 835     | 20      | 20MHz : 3GHz  | 1.6/1.9 | 25    | +/-0.75   | 40/50                | NA         | 1.5/1.8 | 1.5/1.8  | 12/13.5   | 20/25   | 12     | 75    | 0.9      |
| R10M06GSA       | Rf-Lambda             | TBD     | 37      | 0.1GHz : 6GHz | 1.2/1.6 | 20.5  | +/-1      | 35                   | NA         | 1.5/1.8 | 1.8/2.0  | 21/23.5   | 33      | 12     | 160   | 1.9      |
| RLNA10M04GE     | Rf-Lambda             | TBD     | NA      | 0.1GHz : 4GHz | 1.2     | 25    | +/-1      | 50                   | NA         | 2.0     | 2.0      | 23        | 33      | 12     | 420   | 5        |
| FMAM1057        | Fairview Microwave    | 885     | 39      | 20MHz:1.5GHz  | 1.5     | 20    | +/-1.2    | 24                   | NA         | 1.5/2.0 | 1.5/2.0  | 17        | 32/35   | 12     | 70    | 0.84     |
| FMAM1068        | Fairview Microwave    | 1130    | 22      | 0.5GHz : 4GHz | 0.7/0.9 | 24    | +/-2.5    | 33                   | NA         | 1.8/2.5 | 1.2/2.0  | 17        | 32/34   | 12     | 65    | 0.78     |
| ABL0200-02-2015 | WENTEQ Microwave Corp | 419     | NA      | 40MHz : 2GHz  | 1.5/2.0 | 20    | +/-1      | 28                   | NA         | 2.0/2.5 | 2.0/2.5  | 24/26     | 39/45   | 12     | 240   | 2.88     |



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## SPLITTER/COMBINER

| Model                  | Distributore    | Costo € | Peso gr | Band           | Los S1 dB | Los S2 dB | Ampli Bala.. dB | Phase Bala. (deg) | Isolation dB | VSWR-S dB | VSWR-1 dB | VSWR-2 dB |
|------------------------|-----------------|---------|---------|----------------|-----------|-----------|-----------------|-------------------|--------------|-----------|-----------|-----------|
| ZESC-2-11+             | mini-circuits   | 94      | 22      | 20MHz : 2GHz   | 3.55      | 3.60      | 0.04            | 0.07              | 19.87        | 1.75      | 1.56      | 1.56      |
| ZFSC-2-372+            | mini-circuits   | 125     | 70      | 10MHz : 3.7GHz | 3.39      | 3.37      | 0.02            | 0.02              | 18.71        | 1.53      | 1.32      | 1.32      |
| ZFSC-2-2500+           | mini-circuits   | 110     | 70      | 10MHz : 2.5GHz | 3.58      | 3.57      | 0.01            | 0.3               | 19.28        | 1.71      | 1.52      | 1.52      |
| ZX10-2-20-S+           | mini-circuits   | 39      | 22      | 200MHz : 2GHz  | 3.62      | 3.56      | 0.06            | 0.41              | 19.84        | 1.38      | 1.05      | 1.10      |
| D204 (1W) Only Divider | Clear Microwave | 252     | NA      | DC: 4GHz       | 0.2       | 0.2       | +/-0.05         | +/-0.2            | NA           | 1.05      | 1.05      | 1.05      |
| D25003 Divider/comb    | Clear Microwave | 544     | NA      | 0.5GHz : 3GHz  | 0.4       | 0.4       | +/-0.03         | +/-0.3            | 23           | 1.15      | 1.15      | 1.15      |
|                        |                 |         |         |                |           |           |                 |                   |              |           |           |           |

## HYBRID 90°

| Model          | Distributore | Costo € | Peso gr | Band          | Nominal Coup. dB | Loss dB | Isolation dB | VSWR    | Ampli Imbal dB | Phase Imbal deg |
|----------------|--------------|---------|---------|---------------|------------------|---------|--------------|---------|----------------|-----------------|
| RFHB03M03GVT   | RF-LAMBD     | 1495    | 231     | 0.3GHz : 3GHz | 3                | 0.5/0.8 | 20/21        | 1.2/1.3 | +/-0.4/0.75    | +/-3/5          |
| RFHB05M03GVT   | RF-LAMBD     | TBD     | 95      | 0.5GHz : 2GHz | 3                | 0.4/0.5 | 20/22        | 1.2 max | +/-0.4/0.5     | +/-3/4          |
| RFHB05M03GVT-H | RF-LAMBD     | TBD     | 122     | 0.5GHz : 2GHz | 3                | 0.4/0.5 | 18/20        | 1.25max | +/-0.4/0.5     | +/-3/4          |
|                |              |         |         |               |                  |         |              |         |                |                 |

## COUPLER

| Model         | Distributore       | Costo € | Peso gr | Band             | Nominal Coup. dB | Direct. dB | Loss dB | Freq. Sens. dB | VSWR primary | VSWR secondary |
|---------------|--------------------|---------|---------|------------------|------------------|------------|---------|----------------|--------------|----------------|
| RFDC4M3G20    | RF-LAMBD           | TBD     | 63      | 0.4GHz : 3GHz    | 20               | 20/23      | 0.3/0.4 | +/-0.7/1.0     | 1.1/1.2      | 1.1/1.2        |
| PE2208-20     | PASTERNAK          | 780     | 56      | 0.5GHz : 2GHz    | 20+/-1           | 23         | 0.4     | +/-0.75        | 1.2          | 1.2            |
| PE2242-20     | PASTERNAK          | 388     | 111     | 0.5GHz : 3GHz    | 20+/-1.5         | 18         | 0.5     | NA             | 1.3          | 1.3            |
| PE2CP003-20   | PASTERNAK          | 84      | 74      | 0.5GHz : 2GHz    | 20               | 20         | 0.4     | NA             | 1.2          | NA             |
| MC0512-20     | Fairview Microwave | 730     | 63      | 0.5GHz : 12.4GHz | 20+/-1.25        | 20         | 0.5     | +/-1           | 1.22         | 1.22           |
| ZSDC-0525-20S | ZYSEN Tech         | 189     |         | 0.5GHz : 2.5GHz  | 20+/-0.7         | 20         | 0.4     |                | 1.2          | 1.2            |



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## SWITCH

| Model        | Distributore       | Costo € | Peso gr | Band           | Switch t ns | Loss dB | Isolation dB | 1dB Comp P dBm | VSWR     | IP3 dBm | V volt | I mA | Pwr watt |
|--------------|--------------------|---------|---------|----------------|-------------|---------|--------------|----------------|----------|---------|--------|------|----------|
| PE71S1102    | Pasternack         | 4170    | 113     | DC:18GHz       | 50/100      | 0.8:1.3 | 40/45        | 40             | 1.15/1.3 | 55      | 5      | 100  | 0.5      |
| PE71S2048    | Pasternack         | 525     | 45      | 50MHz:1.5GHz   | 1.5/2us     | 1       | 60           | NA             | 1.2/1.5  | NA      | 5      | 1    | 5mW      |
| RFSP2TRDC04G | RF-LAMBD           | ?       | 17      | DC : 4GHz      | 100ns       | 1/1.2   | 43/48        |                | 1.3      | 48      | 5      | 50   | 0.25     |
| RFSP2TRDC12G | RF-LAMBD           | ?       | 45      | DC : 12GHz     | 150ns       | 1.2/1.5 | 30/35        | 44             | 1.6/1.8  | 50      | 5      | 100  | 0.5      |
| FMSW2048     | Fairview Microwave | 261     | 45      | 50MHz : 1.5GHz | 1.5/2us     | 1       | 60           | NA             | 1.2/1.5  | NA      | 5      | 1    | 5mW      |
|              |                    |         |         |                |             |         |              |                |          |         |        |      |          |

## MIXER DOBULE BALANCE

| Model            | Distributore          | Costo € | Peso gr | Band          | Conv loss dB | LO/RF Isol. dB | LO/IF Isol. dB | VSWR RF | VSWR IF | 3rd Ordd IP dBm | LO Input Pwr dBm | RF input pwer Max (dbm) |
|------------------|-----------------------|---------|---------|---------------|--------------|----------------|----------------|---------|---------|-----------------|------------------|-------------------------|
| PE86X1018        | Pasternak             | 462     | ?       | 5MHz:1.5GHz   | 7.5/9.3      | 23/40          | 10/30          | 2.0/2.5 | 1.5/2.0 | 12/15           | 7                | 1                       |
| PE86X1019        | Pasternak             | 650     | ?       | 40MHz:2.5GHz  | 8/9.5        | 22/37          | 20/35          | 2.1/2.5 | 1.5/2.0 | 23/26           | 13               | 9                       |
| PE86X1020        | Pasternak             | 620     | ?       | 1MHz:2.7GHz   | 7/9.5        | 18/39          | 16/36          | 2.1/2.5 | 1.5/2.0 | 17/20           | 10               | 6                       |
| PE86X1021        | Pasternak             | 755     | ?       | 1MHz:2.7GHz   | 7/9.5        | 17/38          | 16/31          | 2.1/2.5 | 1.5/2.0 | 25/27.8         | 17               | 13                      |
| PE86X1022        | Pasternak             | 535     | ?       | 5MHz : 2.5GHz | 8/9.5        | 18/33          | 17/28          | 2.0/2.5 | 1.5/2.0 | 15/18           | 13               | 9                       |
| PE86X1023        | Pasternak             | 580     | ?       | 5MHz:4.2GHz   | 9.8/11.8     | 15/29          | 17/26          | 2.0/2.5 | 1.5/2.0 | 14/17           | 13               | 9                       |
| FMMX1020         | Fairview Microwave    | 605     | 41      | 1MHz:2.7GHz   | 7/9.5        | 18/39          | 16/36          | 2.0/2.5 | 1.5/2.0 | 17/20           | 10               | 6                       |
| MBC10M2G 0-01-10 | WENTEQ Microwave Corp | 248     |         | 10MHz:2GHz    | 8            | 40             | 40             | NA      | NA      | 10              | 10               | 1 (1dB comp point)      |



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## LOW PASS FILTERS

| Model      | Distributore          | Costo<br>€ | Peso<br>gr | Band      | Insertion<br>Loss dB | Pass Band<br>Ripple dB | VSWR    | Rejection                                    | G. delay<br>ns | N section |
|------------|-----------------------|------------|------------|-----------|----------------------|------------------------|---------|----------------------------------------------|----------------|-----------|
| RLPF13G02  | RF-LAMBD              | ?          | 68         | DC-2GHZ   | 0.9/1.0              | 0.7/1.0                | 1.3/1.5 | 2.3-6GHz = -55dB                             | NA             | NA        |
| VLF-1500+  | Mini-circuits         | 31         | 10         | DC-1.5GHz | 0.5                  | NA                     | 1.2     | ?                                            | NA             | NA        |
| SLP-10.7+  | Mini-circuits         | 49         | 42         | DC-11MHz  | 0.32                 | NA                     | 1.7     | Return loss 18.8dB                           | 61             | NA        |
| FMFL021    | Fairview<br>Microwave | 535        | 22,45      | DC-2GHz   | 1                    | NA                     | 1.5     | 3GHz = 40dB<br>4GHz = 60dB                   | NA             | 5         |
| SF2405-SMA | Fairview<br>Microwave | 495        | ?          | DC-30MHz  | 1                    | NA                     | 1.5     | 45MHz = 40dB<br>60MHz = 60dB                 | NA             | 5         |
| SLP-2.5+   | Mini-circuits         | 52         | 42         | DC-2.5MHz | <1dB                 | NA                     | 1.7     | 3.8MHz : 5MHz = 20dB<br>5MHz : 200MHz = 40dB | 334ns          | 2?        |

## HIGH PASS FILTERS

| Model        | Distributore  | Costo<br>€ | Peso<br>gr | Band          | Insertion<br>Loss dB | Return<br>Loss dB | Pass Band<br>Ripple dB | VSWR | Rejection | G. delay<br>ns | N section |
|--------------|---------------|------------|------------|---------------|----------------------|-------------------|------------------------|------|-----------|----------------|-----------|
| SHP-500+     | Mini-circuits | 54         | 42         | 0.5GHz-3.2GHz | 0.3                  | 28.75             | NA                     | 1.9  | NA        | 0.74           | NA        |
| ZFHP-0R60-S+ | Mini-circuits | 50         | 70         | 0.6MHz-800MHz | <1dB                 | NA                | NA                     | 1.16 | NA        | NA             | 2?        |



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### 11 ESTIMATION OF POWER CONSUMPTION, COST AND WEIGHT

| <i>Device</i>    | <i>Description</i>    | <i>Voltage<br/>(Volt)</i> | <i>Current<br/>(Ampere)</i> | <i>Power<br/>Cons.<br/>(watt)</i> | <i>Weight<br/>(gr)</i> |
|------------------|-----------------------|---------------------------|-----------------------------|-----------------------------------|------------------------|
| Amplifier        | A1 (Tx Ampli)         | 12                        | 0.15                        | 1.8                               | 21                     |
| Amplifier        | A2 (Mixer Dive Ampli) | 12                        | 0.15                        | 1.8                               | 21                     |
| Amplifier        | A3 (Rx Ampli)         | 12                        | 0.15                        | 1.8                               | 21                     |
| Mixer            | Na                    | 0                         | 0                           | 0                                 | 41(tbc)                |
| Low Pass Filter  | LPF1                  | 0                         | 0                           | 0                                 | 22(tbc)                |
| Low Pass Filter  | LPF2                  | 0                         | 0                           | 0                                 | 22(tbc)                |
| Low Pass Filter  | LPF3                  | 0                         | 0                           | 0                                 | 22(tbc)                |
| Low Pass Filter  | LPF4                  | 0                         | 0                           | 0                                 | 22(tbc)                |
| Low Pass Filter  | LPF5                  | 0                         | 0                           | 0                                 | 22(tbc)                |
| Low Pass Filter  | LPF6                  | 0                         | 0                           | 0                                 | 22(tbc)                |
| Low Pass Filter  | LPF7                  | 0                         | 0                           | 0                                 | 42                     |
| High Pass Filter | HPF1                  | 0                         | 0                           | 0                                 | 42                     |
| RFSoc Board      | Digital Computer      | 12                        | 2.5                         | 30                                | 400(tbc)               |
| SWITCH           | Tx Switch             | 12                        | 0.25                        | 3                                 | 45                     |
| SWITCH           | Tx Switch             | 12                        | 0.25                        | 3                                 | 45                     |
| Straight Adapter | 3 Pieces              | 0                         | 0                           | 0                                 | 45(tbc)                |
| Curved Adapter   | 6 Pieces              | 0                         | 0                           | 0                                 | 90(tbc)                |
| Attenuator       | ATT1                  | 0                         | 0                           | 0                                 | 40(tbc)                |
| Attenuator       | ATT2                  | 0                         | 0                           | 0                                 | 40(tbc)                |
| Attenuator       | ATT3                  | 0                         | 0                           | 0                                 | 40(tbc)                |
| Attenuator       | ATT4                  | 0                         | 0                           | 0                                 | 40(tbc)                |
| <b>Total</b>     |                       | <b>12V</b>                | <b>3.45A</b>                | <b>41.4(35.4)W</b>                | <b>1105gr</b>          |