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Design and Implementation of a Multichannel Radio-Frequency Receiver for Passive Radar Applications

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Design and Implementation of a Multichannel Radio-Frequency Receiver for Passive Radar Applications

Theory and Practical Implementation of a Four-Channel Radio Frequency Receiver for an FM-Based Passive Radar

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ملخص:

الإستعمال المكثف للموجات الكهرومغناطيسية في مجال الإتصالات اللاسلكية يتيح خيار إستعمالها في ما يسمى الرادار السلبي. حيث أن إستقبال هذه الموجات وتحويلها إلى إشارات إلكترونية عبر مستقبل خاص يمكننا من إستخراج عدة معلومات حول الأجسام التي تعكس هذه الموجات كموقعها وسرعتها. الهدف من هذا العمل هو دراسة وإنجاز جهاز إستقبال متعدد القنوات يعمل على النطاق الترددي للراديو لإستعماله في مجال الرادار.

كلمات مفتاحية: مستقبل ترددات راديو, مستقبل متعدد القنوات, مضخم منخفض التشويش, محول رقمي.

Resumé

La bande de fréquences radio est largement utilisée dans les systèmes de télécommunication actuels. Ceci suggère une possible application dans les radars passifs. Le but est de recuperer cette énergie électromagnétique et de la convertir en un signal électrique pour un traitement ultérieur. L'energie réfléchit à partir des cibles, nous permet d'extraire des informations (telles que : la position, vitesse, etc.) à propos de ces dernières, après l'application des techniques de traitement de signal. Ce projet vise à concevoir et mettre en œuvre un tel récepteur dans la bande FM ainsi que de présenter la théorie de ces systèmes.

Mots-clés : Recepteur RF, Recepteur Multivoie, Radar Passif, FM, LNA, Mixer, Mélangeur, Oscillateur Local, ADC.

Abstract

Electromagnetic radiation in the radio frequency band is used extensively in today's telecommunication systems. Its quasi-omnipresence suggests a possible application in passive radars. The goal is to acquire this electromagnetic energy and convert it to an electrical signal for later processing. Information regarding the target (such as position, speed, etc.) which is reflecting or transmitting this energy can be extracted after applying adequate signal processing techniques. This project aims to design and implement such a receiver in the FM-Band as well as present the underlying theory of such systems.

Keywords: RF Receiver, Multichannel Receiver, Passive Radar, FM, LNA, Mixer, Local Oscillator, ADC.

Acknowledgements

We would like to offer our sincerest gratitude to our supervisors, **Dr. Mourad ADNANE** and **Pr. Adel BELOUCHRANI** who supported us throughout this thesis with their patience and knowledge while allowing us a room to work in our own way. Many thanks for the many opportunities they gave us to increase our knowledge and for providing us with invaluable guidance and advice in all different aspects. A sincere gratitude and appreciation for making themselves available to us on a personal level; one simply could not wish for better or friendlier supervisors.

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Finally, we would like to thank the professors of the electronics department for their dedication and for transferring their valuable knowledge.

Brahim SAADI | Ala Eddine EL-MOUETS
Algiers, June 2015

Dedication

I would like to dedicate this modest work to my family and to my many friends.

A special feeling of gratitude to my loving parents, Abdallah and Khalissa whose support is beyond description. To my sister Meriem and to my brother Mouatez.

To my university friends: Farouk, Tariq, Hacene, Samir, Djalil and to my old friends which are beyond numbering.

Brahim SAADI, Algiers, June 2015

I dedicate this work to my beloved parents...

To my brothers and sisters and to all my family...

To all my friends.

Ala Eddine EL-MOUETS, Algiers, June 2015

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List of Abbreviations

RF Radio Frequency
FM Frequency Modulation
LNA Low Noise Amplifier
IF Intermediate Frequency
FM Frequency Modulation
AM Amplitude Modulation
PM Phase Modulation
ADC Analog Digital Converter
ADS Advanced Design System
DC Direct Current
DSP Digital Signal Processor/Processing
IP3 Third order Interception Point
OIP3 Output Third order Interception Point
IQ In phase and Quadrature
NF Noise Figure
NI National Instrument
LO Local Oscillator
IM InterModulation
SNR Signal to Noise Ratio
SMD Surface Mount Device
VSWR Voltage Standing Wave Ratio
FFT Fast Fourier Transform
VHF Very High Frequency
UHF Ultra High Frequency

Introduction

Analog design is arguably the hardest sub-discipline in electronics engineering. It has been termed "Black Magic"[1] in a reference to the unpredictable system behavior and the daunting complex nonlinear effects that one has to deal with. Going up in frequency only exacerbates matters and increases the challenge for the designer where more effects are to be taken into account. This is why RF (Radio-Frequency) design is one of the toughest subjects to be dealt with. Moreover RF design is somewhat different from low frequency designs where circuit theory is applicable. In this realm, quantities like voltages and currents, which are taken for granted at low frequencies, are no longer easily measurable as they exhibit wave-like characteristics.

The goal of our project is to design and implement a multi-channel RF receiver in the FM band. The receiver is going to be used for passive radar applications and angle of arrival estimation.

The first chapter is an introduction to RF receivers and their basic building blocks, the problems of RF reception and figures of merit in such systems and each composing stage.

In the second chapter, our design is discussed and elaborated and our implementation presented. We will be concerned with the design of every stage and their corresponding circuit schematics and block diagrams. Finally, in the third chapter, experimental results are shown and interpreted and a final conclusion is drawn.

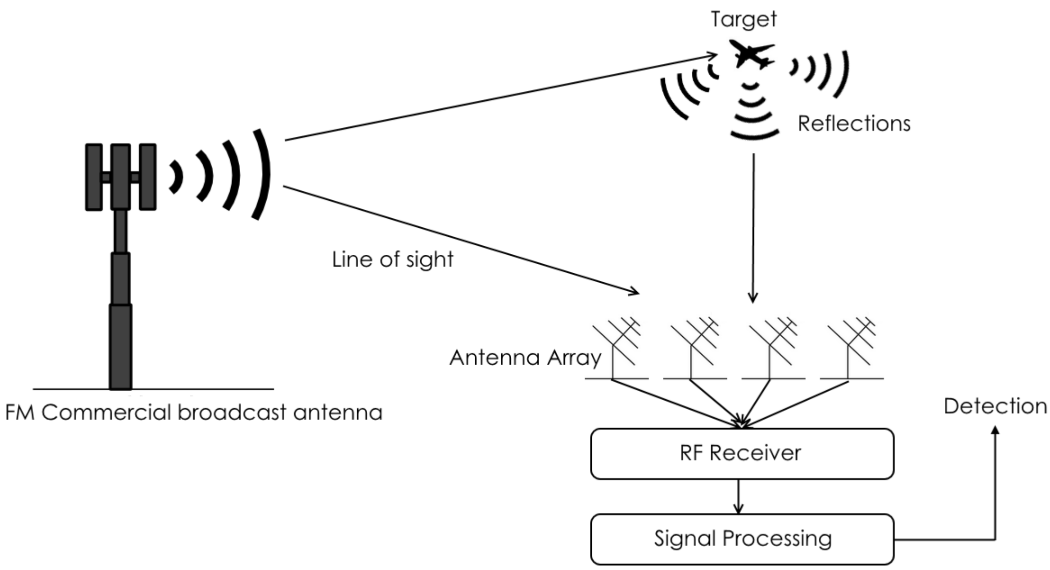
Passive Radars

In this brief section we introduce the basic principles behind passive radars.

Passive radar is a type of radars that does not emit any signal. Instead, it takes advantage of commercial transmitters to discover its environment and estimate or detect targets.

FM broadcast antennas are everywhere and are emitting power continuously. Some of this power will reflect off objects of interest such as aircrafts as shown in the next figure.

[1] IEEE SOCC Conference, Taiwan 2007 | Mike Santarini's blog (09/18/07) at EDN Magazine



Principle of passive radar illustrated

This reflected power along with the main original signal will arrive at an array of antennas through different paths. By applying adequate digital signal processing methods, a passive radar can extract valuable information from these signals such as the position and the speed of the target reflecting this power.

Roughly speaking, this information is extracted by measuring the time difference of arrival between the the line of sight signal (arriving directly from the antenna) and the the ones arriving via reflections of targets.

Passive radar is still an immature system. It is quickly gaining ground on the field since it offers many advantage such as low cost and east of deployment. It also have the advantage of being stealthy, since it does not emit any power in contrast to conventional radar.

In this project, we will be interested in designing and implementing a receiver that is mainly destined for a passive radar application.

1

RF Receivers

Radio communication is about transmitting information from one place to another without the use of wires or optical fibers. The basic building blocks of radio communication systems are the *transmitter* and the *receiver*. Radio communication takes place when a transmitter sends an electromagnetic wave to a receiver through a channel (air, free space, etc.). The receiver then demodulates the signal and recovers the information impressed on the transmitted signal.

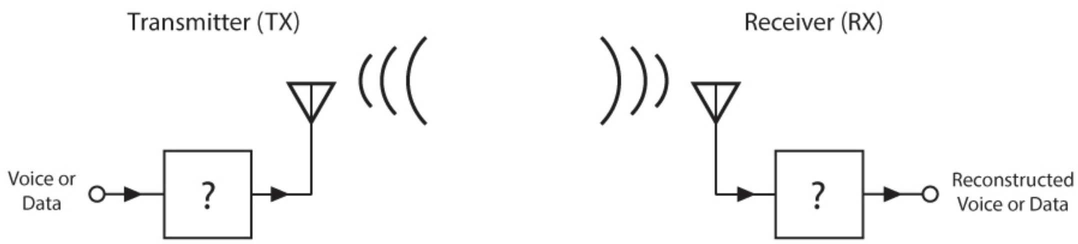


Figure 1.1: Simple view of RF communication

RF Receivers' job is to gather energy (in the form of electromagnetic waves) from its surroundings and convert this energy to a form in which it could be comprehended by the receiver's user.
"Functionally, the receiver must perform a number of tasks to accomplish this miracle. These tasks include at least:

1. Transducer (antennas, etc.) matching
2. Selection of desired signals
3. Rejection of undesired signals
4. Amplification by very large factors
5. Demodulation
6. Error detection and/or correction

7. Received information conditioning and output"[1]

Above all these tasks, a radio receiver must perform one basic and fundamental function: *to distinguish signals from noise*. Noise is the most limiting factor in communication systems, there would be no need for communication theory and communications engineering in a noiseless universe.

1.1 Signals, Noise, and Reception

The notion of noise covers both man-made and naturally occurring disturbances in the electromagnetic spectrum. For instance, man-made signals coming from various sources may end up in the band of the spectrum we are interested in. Whereas an easily observed noise signal in the 18-30 MHz range emanating from planet Jupiter[2] is an example of naturally occurring disturbances.

In communication systems, the signal of interest is a usually a modulated sine wave called a *carrier*. The modulated carrier propagates as an electromagnetic wave through a channel (air, free space, etc.) carrying information with it. Examples of modulation schemes are: AM, FM, PM, QAM, SSB, etc.

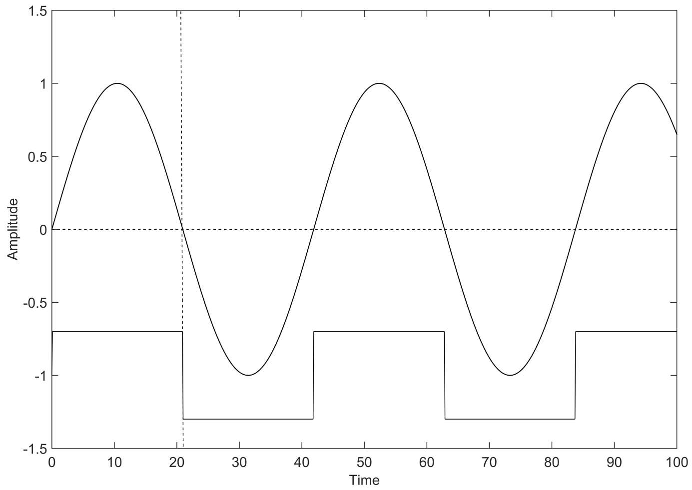
Noise, on the other hand, may be a random signal that carries no useful information. An example of noise is the "hiss" sound heard between radio stations. The spectrum of such signals is usually Gaussian ("white noise") or pseudo-gaussian ("pink noise")[2].

Figure 1.2(a) shows a pure and unmodulated sine wave and its corresponding zero crossings. The crossings happen at exact mathematically deterministic times that are T_0 seconds apart (where T_0 is the period of the sine waveform).

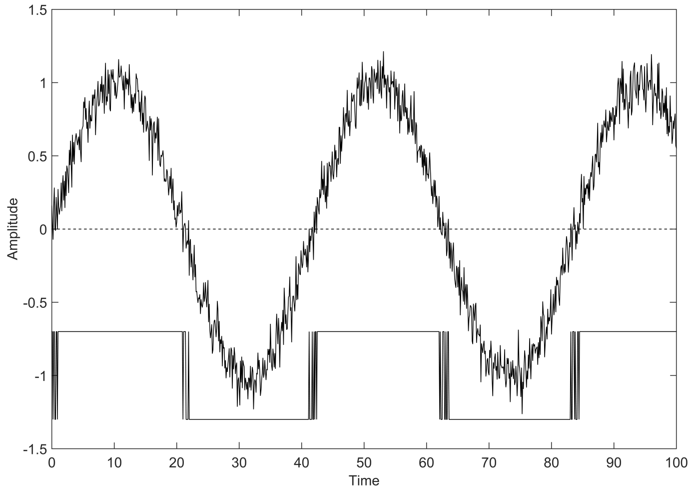
In Figure 1.2(b) a noisy sine wave is generated by creating a pure sine wave and adding a white gaussian noise to it using Matlab. Since the noise amplitude is unpredictable (but can be described statistically), the zero crossings are no longer deterministic. An uncertainty in the period (and hence the frequency) of the sine wave is added as shown in Figure 1.3.

We can readily see how this may cause several undesirable effects. For example, when discussing mixers, we can use a sine wave to control a switch. When the instantaneous value of the waveform is greater than zero, the switch will be open. When the waveform is less than zero, the switch will be closed. The uncertainty in the zero crossings translates to the switch oscillating between the closed and open positions several times instead of a single transition in the case of a pure sinusoid.

Another more trivial effect of noise is to completely obscure the signal of interest. If the signal is weak and the noise is strong enough, the signal may drown in noise and can never be recovered.



(a) Pure sine wave



(b) Noisy sine wave

Figure 1.2: Sine wave zero crossings

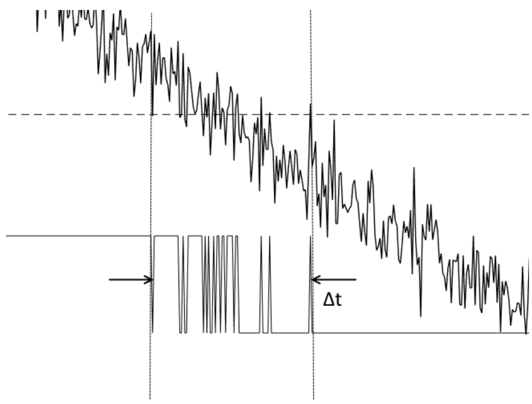


Figure 1.3: Zero crossings of a noisy sine wave

1.1.1 Thermal Noise

Every electronic system (even a single simple resistor) will generate thermal noise as long as its temperature is above absolute zero ($-273.16\ °C$). The designer's goal is to minimize the noise added by the system, so that weak signals of interest are not obscured by it.

Thermal noise (also known as "Johnson–Nyquist noise") is generated by the random thermal agitation of electrons in an electrical conductor regardless of the applied voltage. It is present in all electronic components. For example, the noise power present in a resistor is [2]

$$P_N = KTBR \quad \text{watts} \tag{1.1}$$

where
K is Boltzmann's constant (1.38×10^{-23} Joules/K);
T is the temperature in Kelvin (K);
R is the resistance in ohms (Ω);
B is the bandwidth in hertz (Hz).

1.1.2 The Reception Problem

The signal in Figure 1.2(b) is easily detectable because the signal amplitude is much higher than the noise amplitude. Detection becomes difficult when the signal strength is comparable to the average noise power level.

We use the Signal to Noise Ratio (SNR) to quantify how strong a signal is to the noise floor. The signal-to-noise ratio (SNR) of a receiver system tells us something about the detectability of the signal. The SNR is expressed in decibels as follows

$$SNR = 10 \log \left[\frac{P_s}{P_n} \right] \text{dB} \tag{1.2}$$

where
P_s is the signal power;
P_n is the noise power.

We can see that there are two ways to improve the SNR: either increase the signal power or decrease the noise. Successful systems do both, but they must be done carefully.

One way to achieve this is to use a preamplifier ahead of the receiver antenna terminals. The amplifier will amplify noise from outside (received through the antenna) and the desired signal equally so the SNR is still the same. The problem is that the amplifier produces noise of its own, which causes the SNR to degrade. The key here is to use a very low-noise amplifier (LNA) for the preamplifier. Using an LNA for the preamplifier may actually reduce the noise figure of the receiver system[2]. We will discuss LNAs in subsequent sections.

The noise performance of a receiver can be assessed in three different ways: noise factor, noise figure, and equivalent noise temperature.

Noise Factor

The noise factor of a system is the ratio of the output noise power P_{NO} to the input noise power P_{NI} at a standardized temperature.

$$F_N = \frac{P_{NO}}{P_{NI}} \tag{1.3}$$

Noise Figure

The noise figure is just the noise factor F_N converted to decibel notation:

$$NF = 10\log(F_N) \tag{1.4}$$

Equivalent Noise Temperature

As noted before, noise power is proportional to the temperature of the system. This suggests another way of expressing the noise of a system. Noise temperature T_e is the temperature at which an ideal resistor would produce the same noise level of the system. For ideal resistors T_e is equal to its actual temperature. But for real systems, it is just a *virtual* temperature used to describe noisy systems. T_e is related to the noise factor F_N by

$$T_e = (F_N - 1)T_0 \tag{1.5}$$

where T_0 is the standardized room temperature (T_0 = 290 K).

Noise in Cascaded Amplifiers

In a multistage system, the input noise and the signal of interest are equally amplified by a factor of G_n in each stage. In addition, each stage (having its own noise factor F_N) will generally add some noise of its own. The overall F_N of the system can be expressed as: (Friis' formula)

$$F_N = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + ... + \frac{F_n - 1}{G_1 G_2 ... G_{n-1}} \tag{1.6}$$

where
F_N is the overall Noise Factor;
F_1 is the Noise Factor of stage 1;
F_n is the Noise Factor of stage n;
G_1 is the gain of stage 1;
G_{n-1} is the gain of stage n-1

This is an important result. Given that the gains $(G_1, G_2 ... G_{n-1})$ are high enough (which is almost always the case), the Noise Factor F_N of the overall system is dominated by F_1. If the first stage has a low noise factor and a good enough gain, then the overall noise performance of the system will be close to the performance of

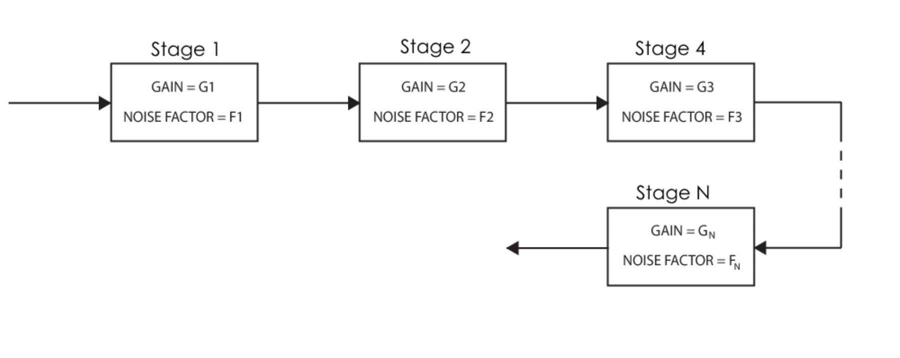


Figure 1.4: Cascaded system noise figure

the first stage. This is why we usually use an LNA (low-noise amplifier) in the first stage of the reception chain.

1.2 Radio Receivers Architectures

The signals received by the antenna must be conditioned and often converted down in frequency in order to extract information from them through demodulation or to be processed digitally. For that, several architectures can be implemented such as the classical heterodyne and homodyne receivers. Less common architectures such as "undersampling receivers" or even bandpass-sampling receivers are now gaining ground thanks to the ever evolving capabilities of Analog to Digital Converters (ADCs).

1.2.1 The Heterodyne Receiver

The super heterodyne receiver is the most common type of radio receivers. Analyzing its principle of operation will help us introduce other types of receivers.

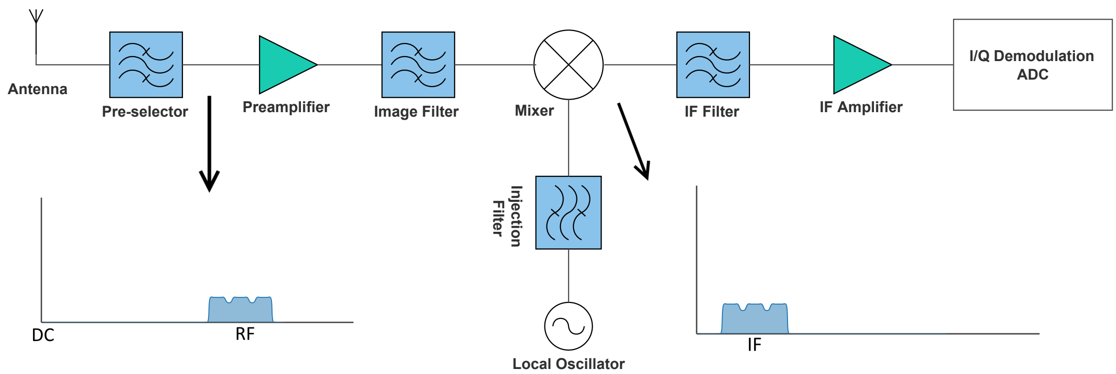


Figure 1.5: Heterodyne receiver block diagram

Several signal processing operations such as channel-selection filtering prove very difficult to implement at high carrier frequencies. Thus a scheme to "shift" or "translate" the desired signal/channel to a lower frequency (called intermediate frequency or IF) has been devised to alleviate such complications. Filtering, for example, can be performed with a much more reasonable Q factor at lower frequencies.

More importantly, without IF translation, channelized systems would not exist. One would need to make a dedicated system (including Filters, Amps, etc.) for a certain one and only one channel. With IF translation, one single system can be tuned to many channels. One only needs to translate the spectrum of the desired channel to the IF of the system, which is pretty straightforward and use the same circuitry for all channels.

We start describing the blocks of such an architecture. Referring to Figure 1.5 we start from left to right:

- The antenna feeds the received signal to a filter that serves as a preselector. The output of the antenna may be very small, often only a few microvolts, thus a very large amplification factor is required throughout the chain of the receiver.

- The preselector has three functions: to limit the bandwidth of the received signal (and define the receiver's noise bandwidth B_N) before amplification and mixing to minimize intermodulation distortion and prevent the saturation of the preamplifier; attenuate spurious responses (image frequency, etc.); attenuate the local oscillator leaking back into the antenna.

- The RF amplifier (also: premplifier, LNA) then amplifies the desired signal without adding significant noise to it. It has been shown earlier in equation 1.6 that this stage almost solely dictates the noise figure of the whole receiver so it is crucial that this amplifier adds as little noise as possible.

- The role of the second filter is to attenuate spurious response frequencies and attenuate the noise at the image frequency originating from the RF amplifier. This is is why it is called "the image filter".

- The signal is then passed to a mixer where it is multiplied with a local oscillator. The frequency of the local oscillator is offset from the RF signal frequency by an amount equal to the desired IF frequency. The result of this operation is the shifting of the signal spectrum from RF to a lower intermediate frequency (IF) where several signal processing operations are more easily feasible. The frequency of this first LO is usually variable in order to achieve channel tuning.

- The injection filter attenuates wideband noise around the LO frequency as well as the harmonics that may degrade the mixer's performance and cause spurious responses in the receiver.

- The IF filter is tuned to the difference between the RF frequency and the local oscillator's frequency and will pass only the narrow-band signal centered around IF (usually a single channel).

$$f_{IF} = f_{RF} - f_{LO}$$

This filter suppresses the intermodulation products (IM products) [1] of the mixer and attenuates the second image frequency for the next mixing process.

- After channel selection around the IF, the signal is amplified one final time in order for it to be sensible by the ADC or the demodulator at the last stage. This stage usually employs a high gain amplifier.
- The signal is then fed to a demodulator. This stage recovers the information imprinted on the carrier signal and converts it to a format that is exploitable by the user of the receiver (audio, video, digital data, etc.). Oftentimes, the demodulator is another mixing stage that converts the IF signal down to baseband.

Challenges that must be dealt with when implementing this architecture are the image and spurious signal control and power dissipation especially in off-chip filters which are cumbersome and power consuming [3].

1.2.2 The Zero-IF Receiver

This architecture can be thought of as a special case of the heterodyne receiver. In the previous section, the RF signal has been down-converted to an IF frequency by judiciously choosing the LO frequency ($f_{IF} = f_{RF} - f_{LO}$). A Zero-IF receiver employs the most obvious choice of the LO: $f_{LO} = f_{RF}$. This scheme shifts the RF spectrum right to DC (0 Hz). It is indeed the natural approach to convert a signal from RF to baseband. The zero-IF architecture is also called direct-conversion and when the local oscillator is synchronized in phase with the incoming carrier frequency, the receiver is called a homodyne receiver.

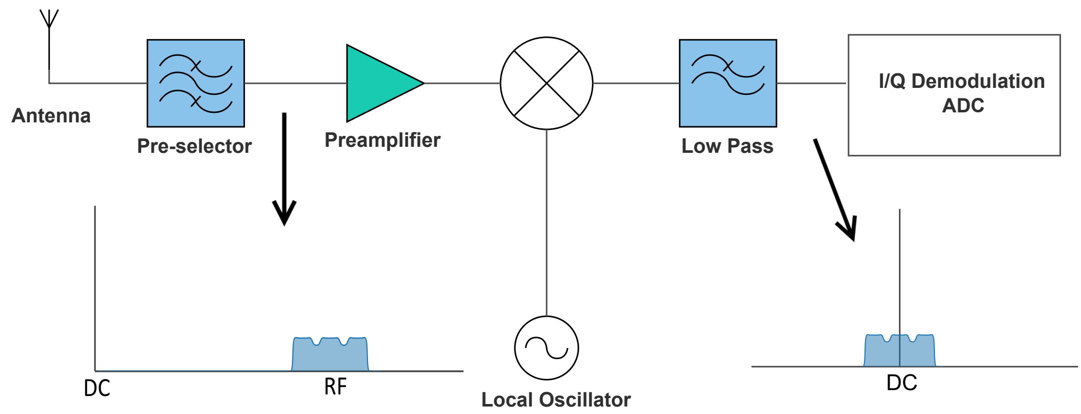


Figure 1.6: Zero-IF receiver block diagram

The Zero-IF scheme simplifies the architecture of the receiver and reduces both the stage and component count but several other practical challenges arise.

[1] Intermodulation products are spurious frequency components that are the result of multiple frequency components at the input of a system being multiplied together due to nonlinearities.

LO Leakage, Self-Mixing and DC-Offset

In practical realizations, the isolation between the LO and the RF input of the mixer is not perfect (i.e finite). The LO can therefore leak into the previous stages of the receiver chain. The effect is exacerbated since the LO is a relatively high power signal.

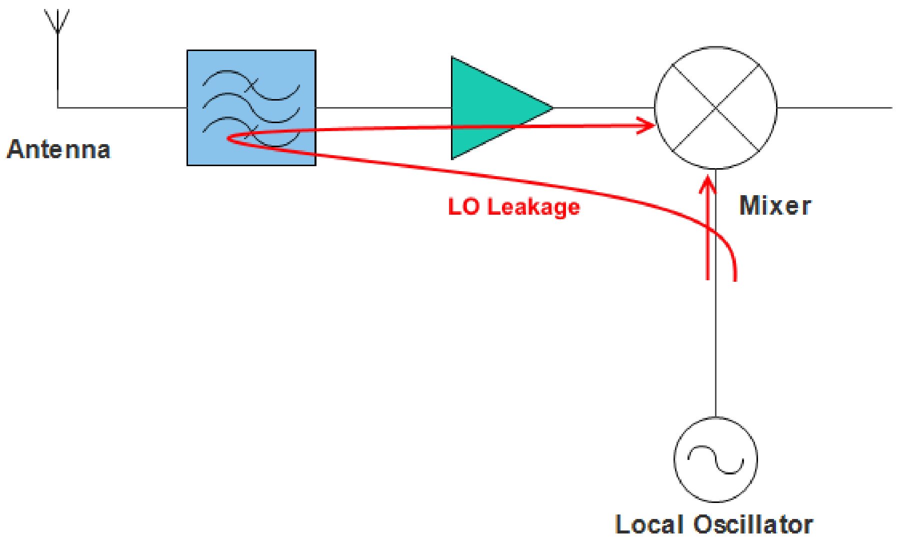


Figure 1.7: Depiction of LO Leakage

The leakage LO signal now present at the input of the LNA and the mixer is amplified and mixed with itself. This self-mixing produces a large DC component since $f_{IF} = f_{LO} - f_{LO} = 0$ Hz. This DC component may be very powerful, stronger than the desired signal even. Which leads to SNR degradation or the saturation of the amplifier. The problem is exacerbated if the LO finds its way to the antenna and gets radiated and reflected off objects back to the receiver. This produces a varying DC-offset depending on the objects around the antenna [4].

A possible remedy for this problem is AC-coupling through a bypass capacitor (the corner frequency should be around a few kHz) or a programmable DC-offset cancellation loop [3].

Flicker Noise

Flicker noise is a type of electronic noise with a $1/f$ or pink power density spectrum. It occurs in virtually all electronic devices since it is caused by material defects such as impurities in conductive channels. The flicker noise is characterized by a corner frequency f_c that is around a few kHz for BJT transistors but can go as high as a few GHz in MOSFETs. Thus the effect is more pronounced in MOS devices [5].

Since the downconverted spectrum is located around 0 Hz the $1/f$ noise may cause problems for the signal and results in SNR degradation.

If these challenges are circumvented or are unimportant for a specific application then the advantage of the Zero-IF receiver is a very simplified architecture and a much more relaxed filtering. The filtering can also be done digitally or using active filters to further reduce the stages needed.

1.2.3 The Low-IF Receiver

As the name suggests, in a low-IF receiver the RF signal is mixed down to a low (non-zero) intermediate frequency f_{IF}. The down-mixed signal can now be digitized directly from this low-IF and mixed down to baseband (DC) through subsequent digital signal processing. This architecture has many of the desirable properties the zero-IF offers but also avoids the previous challenges such as DC-offset and $1/f$ noise. This technique is widely used in tiny FM receivers and mobile phones.

1.2.4 Undersampling Receivers

The receiver is now simplified to a preselector filter, an amplifier, and an ADC. To convert down the RF signal to baseband without a mixer, a technique called *undersampling* is employed.

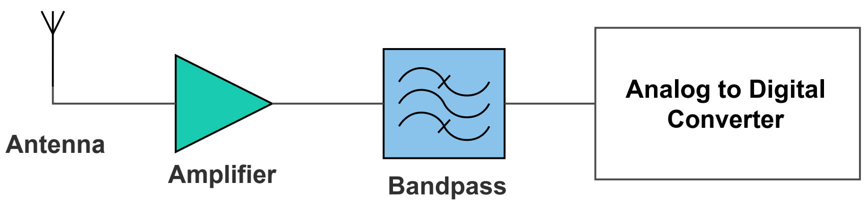


Figure 1.8: Undersampling receiver block diagram

The filter and the amplifier are used to do the exact same functions as in the previous architectures. Namely, band-limit the signal, reject unwanted signals and amplify the desired signal. The ADC (and the subsequent digital signal processing) now does the rest of the job. This is done by cleverly choosing the sampling rate of the ADC.

Undersampling and Oversampling

To reconstruct the signal being sampled without loss of information, the sampling rate must fulfill the Nyquist criterion. The latter is often quoted as: "The sampling frequency/rate must be at least twice the highest frequency of the signal being sampled". This means that for an IF-sampling application at 70 MHz, the sampling rate must be at least 145 MHz (145 MSPS). If the signal only occupies a mere 5 MHz around the IF-frequency as shown in Figure 1.9 then sampling at 145 MSPS is wasteful because for passband signals, the Nyquist criterion specifies that the sampling rate be at least twice the signal bandwidth. Therefore in the previous case, a sampling rate of 10 MSPS would have been enough. Sampling at higher rates is called *oversampling* and results in SNR improvement [6].

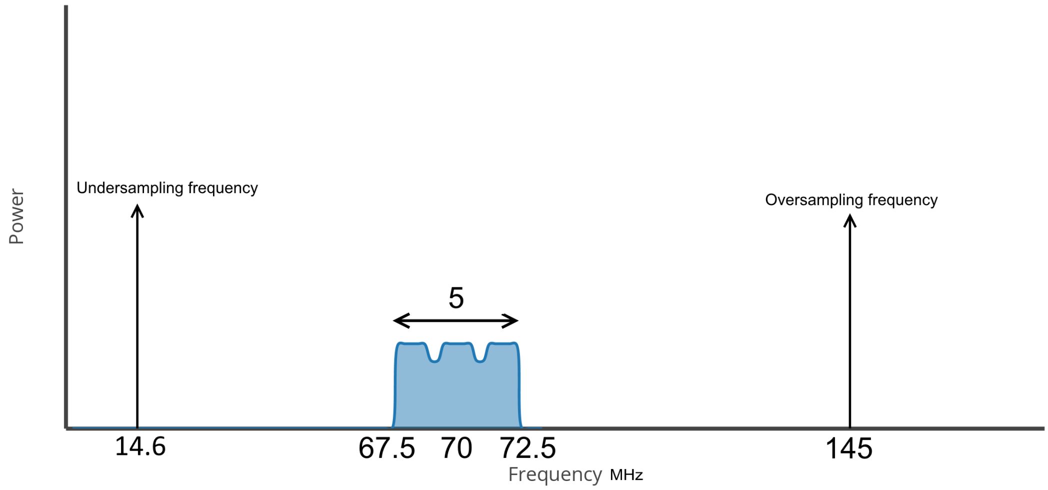


Figure 1.9: Bandpass IF signal and examples of sampling frequencies

In contrast to oversampling, undersampling as the name suggests, is sampling the signal at a rate that is lower than the Nyquist frequency. The rate must always be higher than the signal bandwidth and has to satisfy the following inequality:

$$\frac{2f_H}{n} \leq f_s \leq \frac{2f_L}{n-1} \tag{1.7}$$

where n satisfies $1 \leq n \leq \left\lfloor \frac{f_H}{f_H - f_L} \right\rfloor$;

f_H and f_L are the highest and lowest frequency components of the signal respectively. Solving the inequality for different values of n produces several bands of "allowed undersampling frequencies".

Following our previous example, $1 \leq n \leq \left\lfloor \frac{72.5}{5} \right\rfloor = 14$,

choosing $n = 10$ gives us the interval $\frac{2 \times 72.5}{10} = 14.5 \leq f_s \leq 15 = \frac{2 \times 67.5}{9}$. A frequency in this range will allow us to undersample the signal and recover the full information even though we are sampling it at much less than its highest frequency (72.5 MHz). Notice that the sampling rate is also higher than twice the signal bandwidth (2×5 MHz) thus we are doing *both* undersampling and oversampling. The two are not mutually exclusive since one is defined with respect to the signal frequency and the other with respect to its bandwidth.

Taking Advantage of Aliasing

When an analog signal is sampled, its spectrum is replicated (in the frequency domain) at integer multiples of the sampling rate $\pm nf_s$ where $n = 0, 1, 2...$
The effect is shown for both a baseband and a bandpass signal in Figure 1.10. In the latter case, we can see that we now have a version (alias) of the signal's spectrum residing around DC (first Nyquist zone). After applying a digital filter we recover only the replica of interest. Thus, virtually mixing down the signal from the RF to baseband. This is a case where aliasing is not undesirable but rather used intentionally to our advantage.

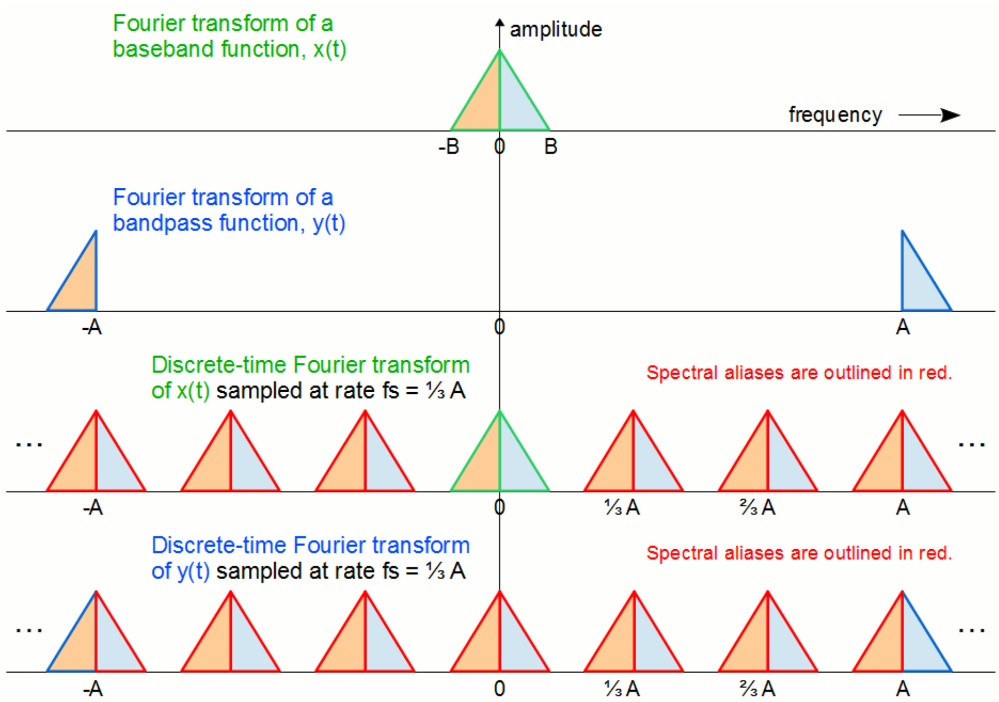


Figure 1.10: Effect of sampling on baseband and bandpass signals

Undersampling receivers prove very efficient in terms of component count as well as power consumption. But the issue of clock jitter is now amplified since we are sampling higher frequency signals (demanding higher slew rate) [7].

1.3 Basic Receiver Blocks

We now move to discuss the basic building blocs of a receiver and their figures of merit.

1.3.1 Antenna

An antenna is an electrical device which converts electric power into electromagnetic waves and vice versa (in the case of reception). Thus providing an interface between the receiver and free space.

The receiving antenna intercepts the electromagnetic waves radiated from the transmitting antenna. When these waves impinge upon the receiving antenna, they induce a small voltage in it. This voltage causes a weak current to flow, which contains the same frequency (and information imprinted on the carrier) as the original current in the transmitting antenna.

Receiving antennas are often overlooked and do not get much of the designer's attention as much as they do when designing a transmitter. Especially in broadcast applications where transmitting power is in the order of MegaWatts. Furthermore, using a poor antenna will result in a power loss of a few dB. Compared to the 100 dB medium loss, it does not seem to matter much. That is why many receivers may have no more than a poorly matched piece of wire as an antenna [1].

An antenna has many properties to consider in the design. The most important ones being: polarization, bandwidth, gain, directivity, equivalent temperature noise.

Polarization: Antenna polarity is an important consideration in a receiver. The receiver's antenna should be of the same polarity as the transmitter for best reception. A vertically polarized receiving antenna will not receive any power from a horizontally polarized transmitting antenna and will lose 3 dB of power in the case of a circularly polarized transmitter [1].

Gain: An antenna that transmits the same amount of power evenly, or receives with exactly the same capability, in all directions is called an "isotropic" antenna. This antenna cannot exist in the real world. All real antennas have some gain. This simply means that they do not transmit or receive exactly the same in all directions. Gains is the effective power transmitted by an antenna compared to the power transmitted by an isotropic antenna. Gain is expressed in dBi (the reference is the isotropic model).

Directivity, or the antenna pattern, is the graphical representation of the radiation properties of the antenna as a function of space. That is, the antenna's pattern describes how the antenna radiates energy out into space (or how it receives energy). Directivity can reduce noise. An antenna does this by being more sensitive in the direction or polarization of desired signals, and less sensitive in the direction or at the polarization of noise.

The *bandwidth* of an antenna is the band of frequencies for which the performance of the antenna is acceptable, in other words, the properties and performance of the antenna (gain, directivity, impedance, etc.) must be within specs.

1.3.2 Filters

The energy captured by the antenna contains not only the signal of interest but a multitude of signals from different sources as well as noise. Simply put, a filter is a two-port network that will pass only a desirable band of frequencies (where the signal of interest resides) and severely attenuate the signals that are out of the band of interest.

Filters are implemented using many different technologies and they can be categorized in so many ways. We will only discuss a certain category of filters. Analog passive filters have been around for a long time. Their theory and design are well established in the literature.

First let's introduce some filter terminology.

- *The Passband* refers to a band of frequencies that a filter will pass with minimum to no attenuation. The frequencies within the passband usually contain the signals of interest.

- *The stopband* refers to the band of frequencies that the filter will attenuate. Frequencies in the stopband are not signals of interest and can cause problems if they propagate further into a system. Hence the necessity to suppress them.

- *The transition band* is the band of frequencies between a filter's passband and its stopband. Ideally, we would like a filter to transition immediately from its passband to its stopband but all practical filters have a transition band.

- *The ripple* of a filter is a measure of the "flatness" of its magnitude response in the passband.

- *The initial roll-off* is a measure of how fast the filter's magnitude response drops off in the transition band. The more amplitude ripple we allow in the filter's passband, the steeper the initial roll-off of the filter.
- *The ultimate roll-off* rate is how fast the magnitude response drops off far from the passband and the transition band. This is proportional to the order of the filter. The higher the order, the higher component count and the complexity of the filter.
- *Re-entrant Response* is a realization problem, it is an undesirable response in the stopband that cannot be predicted mathematically.
- *Insertion loss* in dB (IL_{dB}) is the minimum attenuation in the passband. Ideally the attenuation is 0 dB but for real practical filters it can be a few dB. This is due to the imperfect components that make up the filter (inductor DC resistance, etc.)

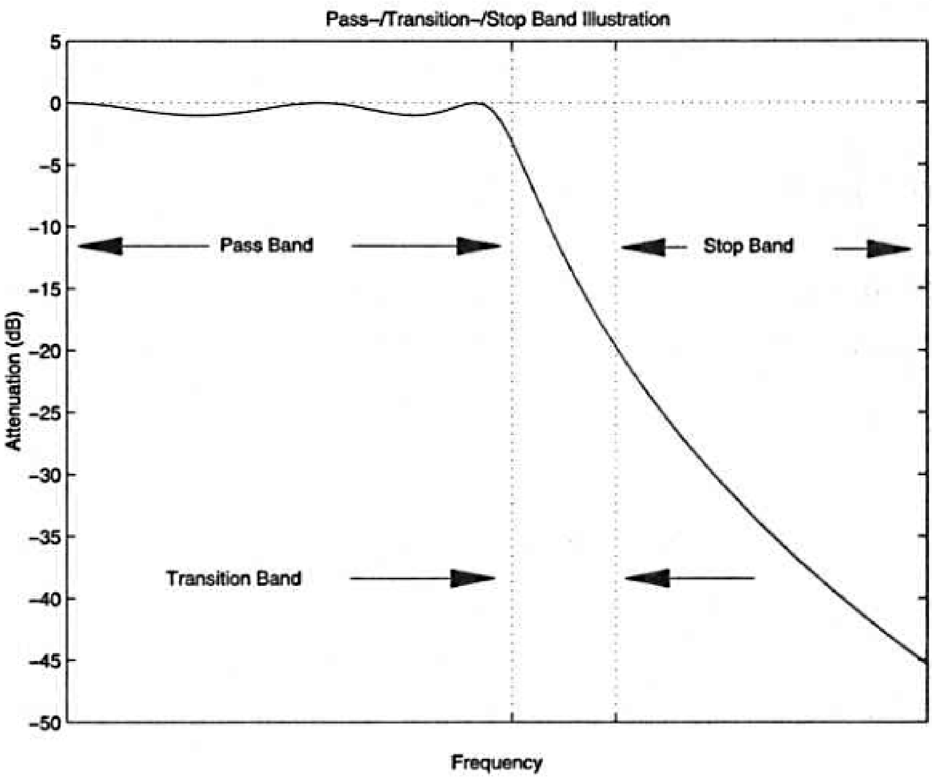


Figure 1.11: Regions of a low-pass filter response

Several filter types or classes exist. The most popular ones being: Butterworth, Chebychev, and the Elliptic Filter. Each filter type has a different magnitude and phase response (for the same desired passband). Each class of filters optimizes a certain property or measure by making trade-offs.

The Butterworth filter is a type of filters designed to have a flat magnitude response in the passband. Which is why it is also referred to as a maximally flat magnitude filter.

The Chebychev filter aims for a steeper initial roll-off by sacrificing the flatness of the passband. The more ripple we allow in the passband, the steeper the initial-roll-off.

The Elliptic filter trades off ultimate roll-off for a steeper transition band roll-off. This is due to the fact that its transfer function contains zeros in addition to poles.

The Bessel filter is designed to have a good transient response and superior phase response (as linear as possible) sacrificing the magnitude response (slower transition from passband to stopband).

A quick comparison in figures 1.13 and 1.14 shows that Butterworth, Chebychev and Bessel filters all have the same ultimate roll-off (30 dB/octave) since they all

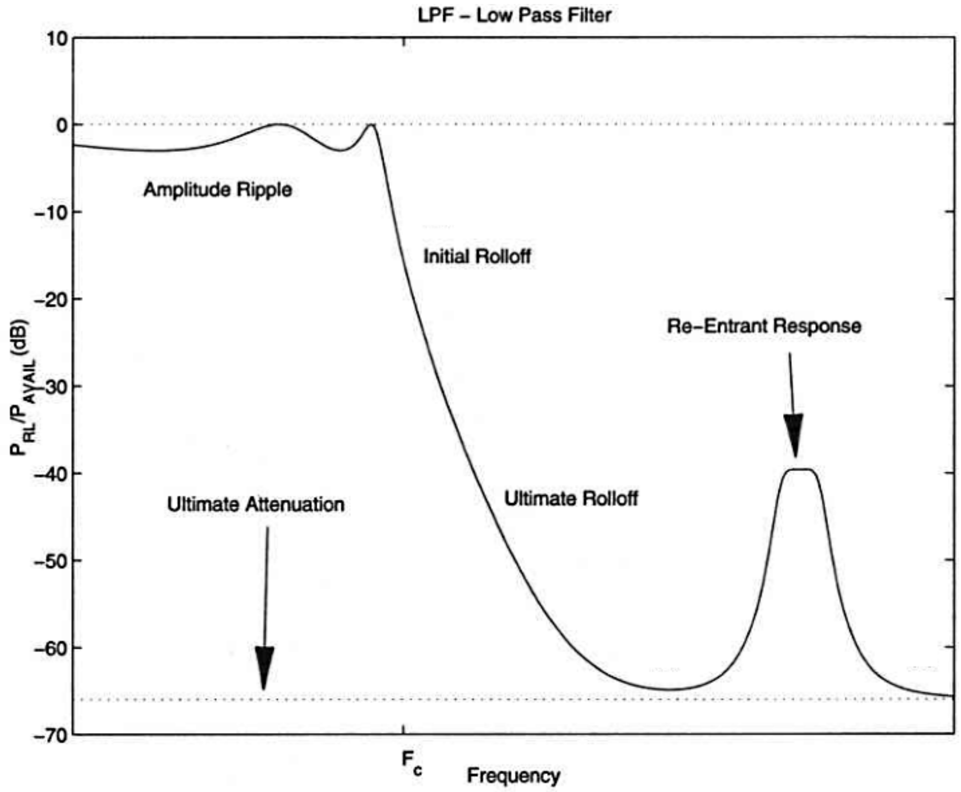


Figure 1.12: Filter terminology

have the same order. We can also see that although the elliptic filter has a smaller ultimate roll-off (6 dB/octave), it provides the most attenuation immediately above the filter's passband (steep initial roll-off).

Figure 1.15 shows the passband performance of the four aforementioned filters, it can be seen that the Butterworth filter has a perfectly flat magnitude response while Chebychev and the Elliptic filter both contain ripples that could be traded-off for more initial roll-off. The Butterworth performs in a middle-of-the-road fashion. It is not quite as good as the Chebychev, but better than Bessel. The Bessel poor magnitude response is compensated by an excellent phase response as shown in Figure 1.16. Here, group delay is plotted against frequency. As it can be seen, the Bessel has a perfectly flat (constant) group delay for all frequencies within the passband.

Depending on the application, choosing the right filter class is crucial for the receiver performance. The filter is often required to perfectly preserve the signal of interest. But effects such as uneven group delay and non-flat magnitude response cause undesirable distortions to the signal. Oftentimes a trade-off between the two is to be made. Nevertheless, several equalization techniques exist to reverse the undesirable effects of the filter on the signal of interest.

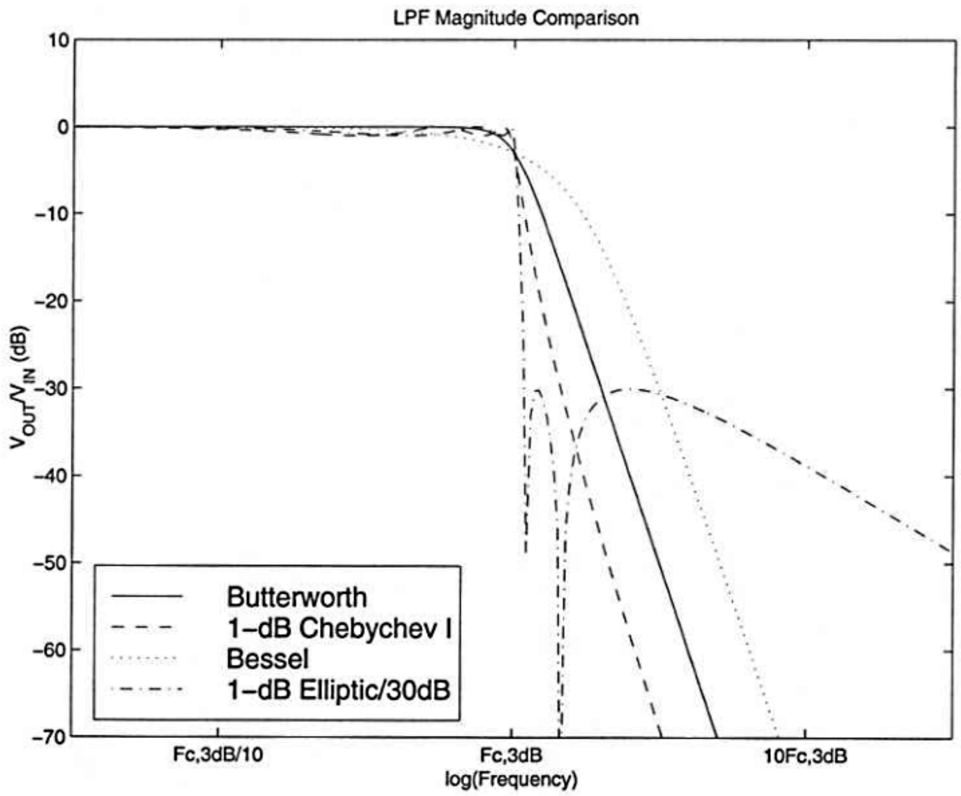


Figure 1.13: Comparison of four low-pass filters

1.3.3 Mixers

A mixer shifts or translates signals from one center frequency to another while keeping the modulation (and therefore the information) intact. Mixing is required for many reasons: Filter realizations are often hard or impossible in certain bands of frequencies. Frequency assignments are done by government agencies such as the FCC which places many restrictions on which frequency band one is allowed to transmit/receive. Antenna size is another reason since its physical dimensions depend strongly on frequency, antennas tend to get large and bulky at lower frequencies.

A mixer is a nonlinear device that performs frequency translation by multiplying two signals. There are two inputs to the mixer, the RF port and the LO (Local Oscillator) port. The desired output is the IF (Intermediate Frequency). The mixer contains a device (diodes or amplifiers, or any device with a nonlinear transfer function) that multiplies the RF signal by the LO signal. The product of these two sinusoids can be decomposed into a sinusoid whose frequency is the sum $f_{RF}+f_{LO}$ of the RF and LO frequencies and another having the difference frequency $f_{RF}-f_{LO}$. One of these is the desired frequency-shifted IF.

Consider two signals, one is the modulated carrier (carrying our information) V_{RF} and the other is the local oscillator V_{LO}, a pure sinusoid. These signals are written as:

$$V_{RF}(t) = A(t)\cos[\omega_{RF}t + \phi(t)]$$

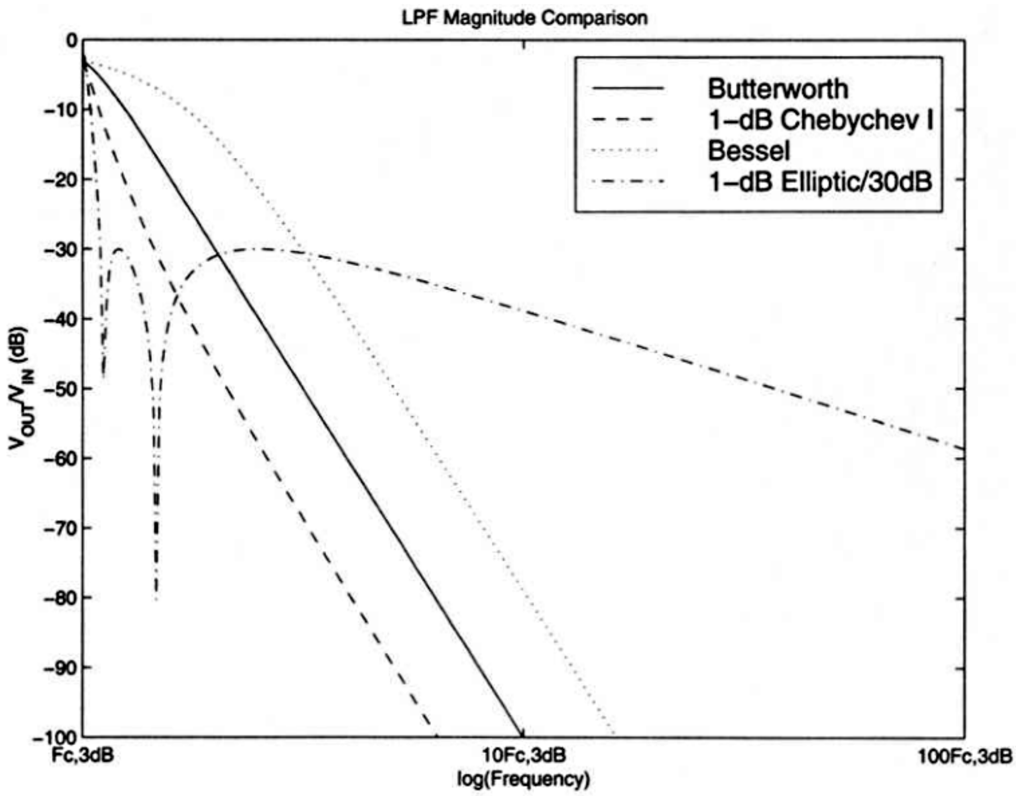


Figure 1.14: Ultimate roll-off of four low-pass filters

$$V_{LO}(t) = \cos(\omega_{LO} t)$$

The multiplication of these two signals produces V_{IF}:

$$\begin{aligned} V_{IF}(t) &= V_{RF}(t)V_{LO}(t) \\ &= A(t)\cos[\omega_{RF}t + \phi(t)]\cos(\omega_{LO}t) \\ &= \frac{A(t)}{2}\cos[(\omega_{RF} + \omega_{LO})t + \phi(t))] \\ &+ \frac{A(t)}{2}\cos[(\omega_{RF} - \omega_{LO})t + \phi(t))] \end{aligned} \tag{1.8}$$

We can see that the multiplication process produces two versions of the RF signal, one centered around the sum frequency $f_{IF,SUM} = f_{RF} + f_{LO}$ and the other around the difference $f_{IF,DIFF} = f_{RF} - f_{LO}$. Either of them can be recovered through filtering. We are interested in recovering the difference frequency $f_{IF,DIFF}$ in receivers and the process is called "downconversion" since we are translating the signal to a lower center frequency.
The equation describing this frequency translation is:

$$f_{IF} = |f_{RF} \pm f_{LO}| \tag{1.9}$$

Figure 1.18 shows the conversion process in the frequency domain.

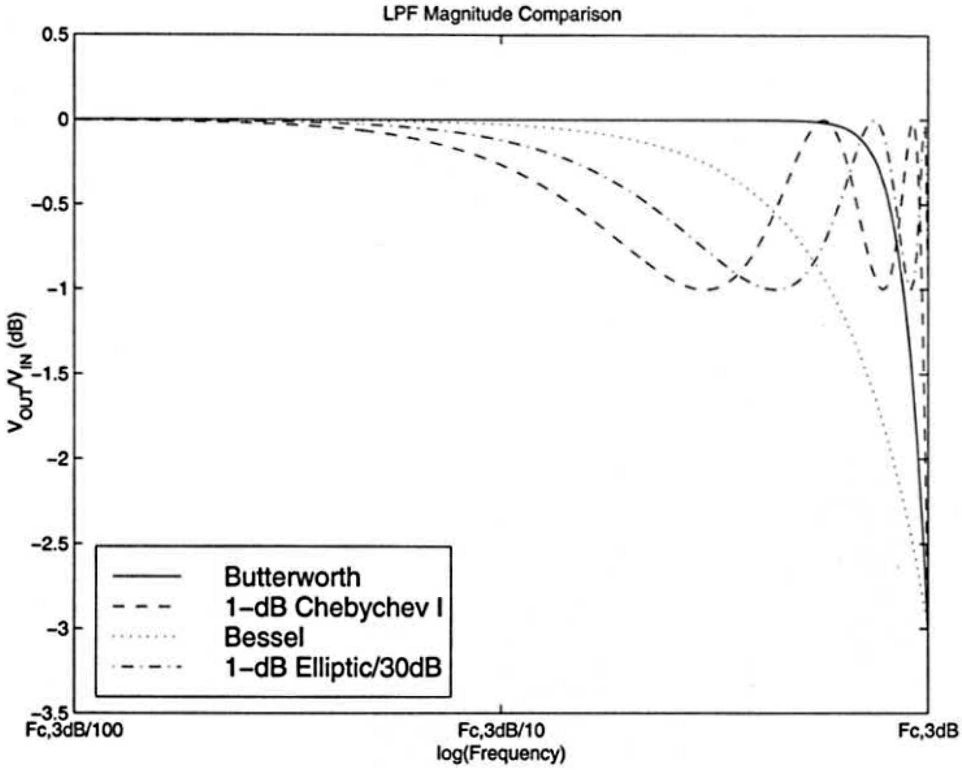


Figure 1.15: Passband performance of four low-pass filters

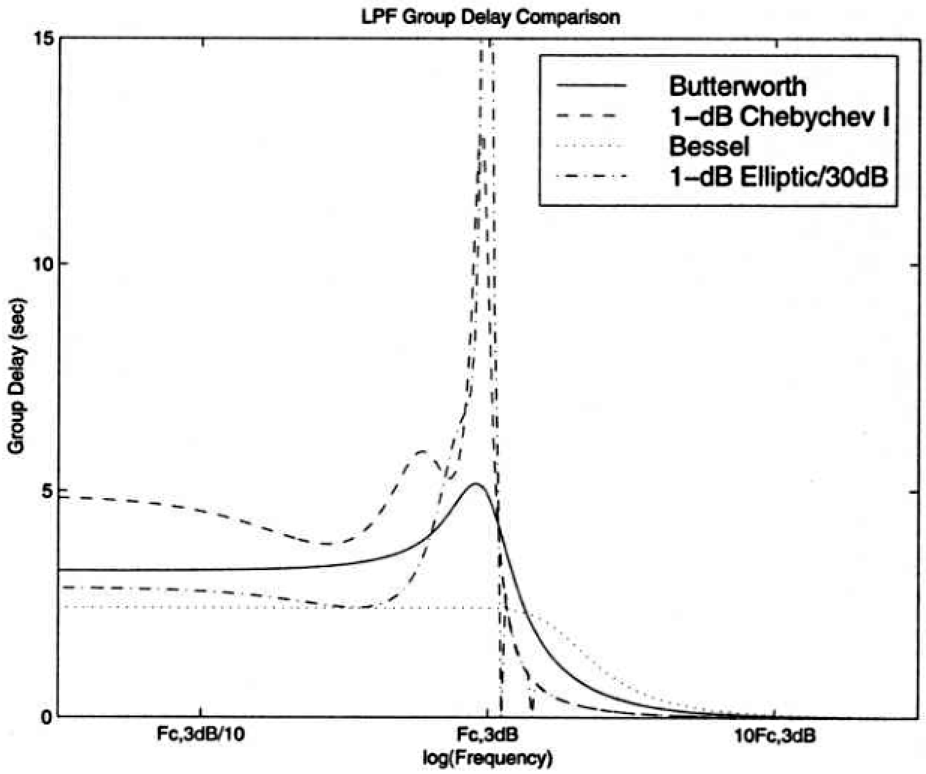


Figure 1.16: Group delay comparison of four low-pass filters

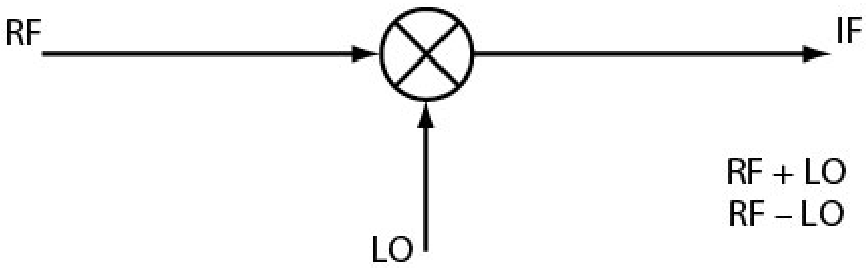


Figure 1.17: Mixer symbol

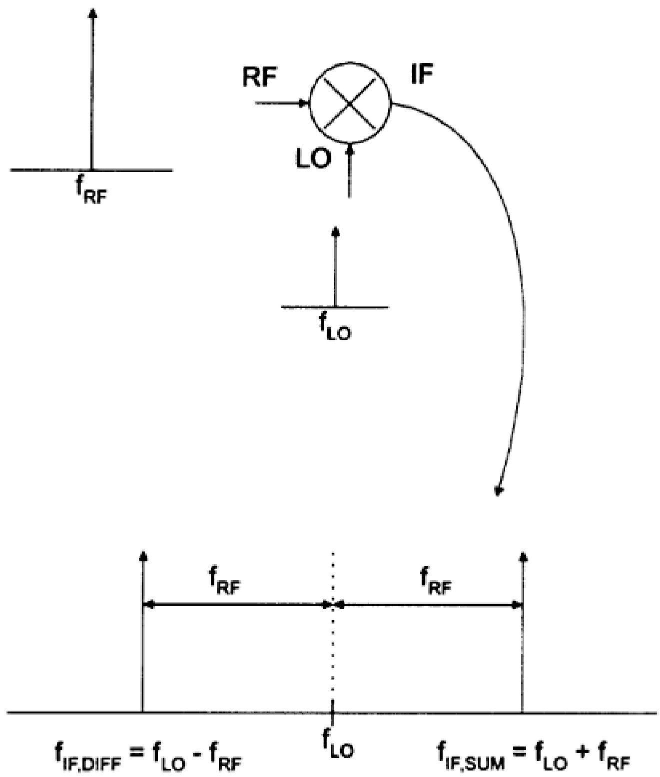


Figure 1.18: Frequency conversion spectrum

Equation 1.9 can be expressed as three separate equations:
Knowing the LO and the RF we can solve for the IF:

$$f_{IF} = f_{LO} \pm f_{RF} \tag{1.10}$$

If we know the LO and IF frequencies, we can find the two possible RFs, which will be converted to the IF by the mixer:

$$f_{RF} = f_{LO} \pm f_{IF} \tag{1.11}$$

If we know the RF and the desired IF, we can solve the two possible LOs.

$$f_{LO} = f_{RF} \pm f_{IF} \tag{1.12}$$

Figure 1.19 depicts the previous equations in the frequency domain.

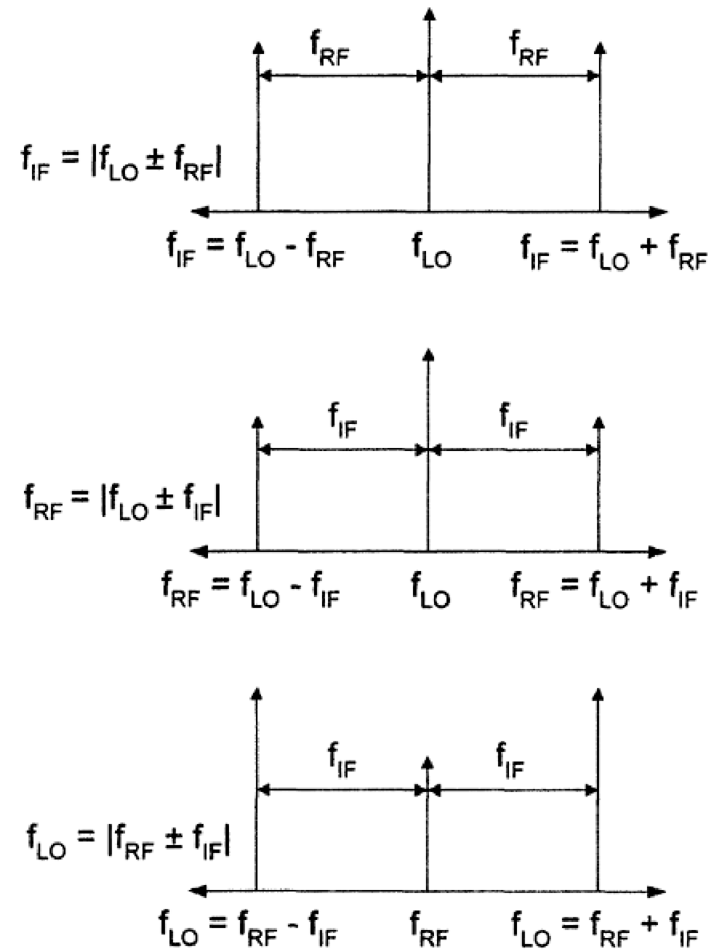


Figure 1.19: Graphical depiction of frequency translation equations

Conversion Loss

When a mixer converts a signal from RF to IF, the signal at the IF port is usually weaker. This is due to the mixer's conversion loss. In other words, conversion loss (or gain in the case of many active converters) is a measure of how efficiently a mixer converts energy from the input frequency to the output frequency. It is defined as the ratio of the power at the output frequency to the power at the input frequency with a given LO power. A specified LO power is necessary because conversion loss varies with the level of the LO, as the impedance of the mixer diode changes.

Interport Isolation

Due to realization effects (device capacitance, etc.), mixers suffer from unwanted coupling (leakage) from one port to another. This occurs without any frequency translation.

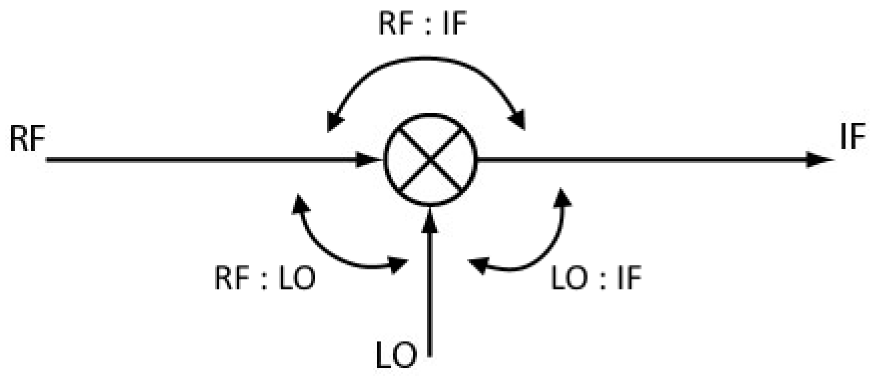


Figure 1.20: Mixer Port-to-Port isolation

Port to Port isolation is the amount of attenuation a signal endures when leaking from one port to another (say from the LO to the RF port). Usually the LO signal is much more powerful than the RF signal, that is why we are mostly concerned with the LO : RF isolation rather than the RF : LO. It is defined as:

$$Isol_{LO:RF} = \frac{P_{LO} \text{ present in the RF port at frequency } f_{LO}}{P_{LO} \text{ entering the LO port}}$$

Where, P_{LO} is the LO power.
$Isol_{Port_X:Port_Y}$ is defined in the same way.

Image Frequency

As stated by equation 1.11 and Figure 1.19 the mixer will convert down both frequencies above and below f_{LO} by f_{IF}, that is $f_{LO} + f_{IF}$ and $f_{LO} - f_{IF}$. Thus any signal at these frequencies will be converted down to the IF frequency. Imagine that our RF signal is centered at $f_{LO} + IF$ and another signal is present at $f_{LO} - IF$. The mixer will convert both of these signals down to the same IF center frequency. This problem is referred to as *the image frequency problem* and is illustrated in Figure 1.21.

Another problem resulting from this mixer property is the *image noise* where noise present in the image frequency region is converted down and added to our desired signal, which results in SNR reduction. To solve this problem, an image filter is added before the mixer to suppress any signal present in the image frequency.

Spurs

During the previous sections it has been assumed that the mixer is an ideal multiplier that will produce the sum and difference frequencies at its output. However, due to practical considerations the mixer will also generate other products called spurs at frequencies given by:

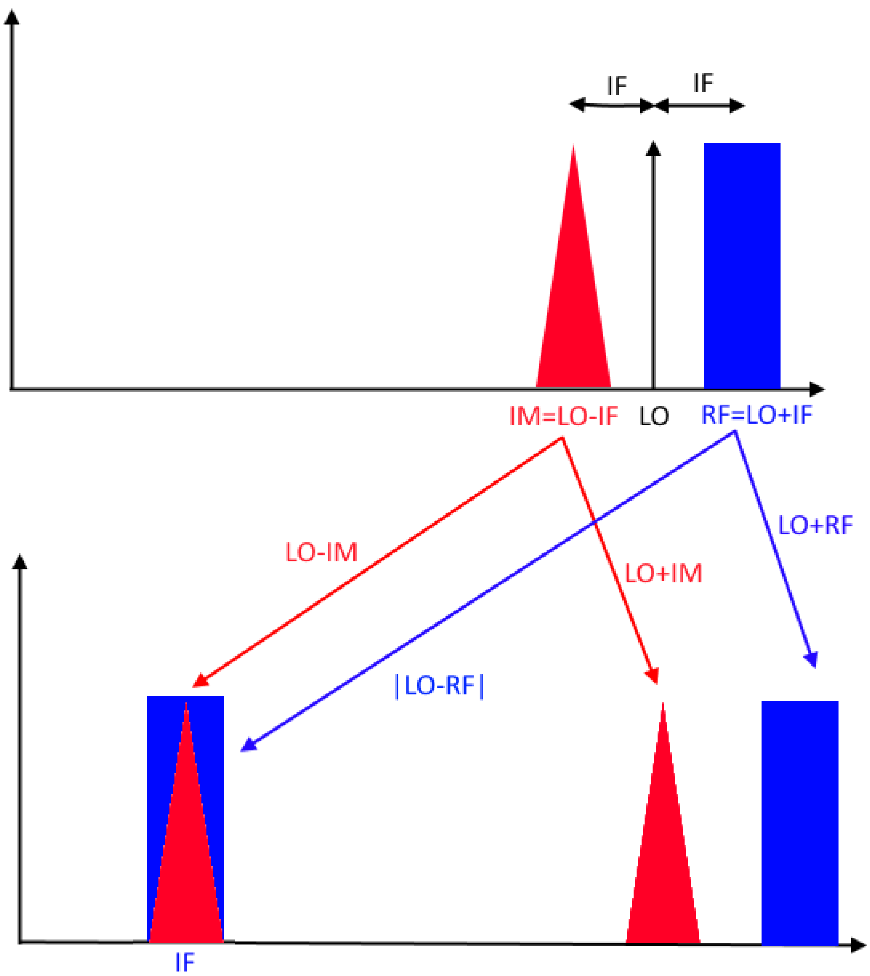


Figure 1.21: Illustration of the image frequency problem

$$|mf_{LO} \pm nf_{RF}| \ for \ n, m = 0, 1, 2, 3, ...$$

These frequency components are usually weaker than the first sum and difference ($m = 1, n = 1$) but may have a considerable spurious response. Mixer datasheets usually provide what is called *mixer spur tables* which provide information about the frequency and power of the output spurs. Figure 1.22 shows how "rich" is the output spectrum of a practical mixer. To solve the spurious response problem, a judicious filtering must be performed right after mixing.

1.3.4 Low-Noise Amplifiers

A low-noise amplifier (or LNA) is a type of amplifiers used to amplify very weak signals without adding much noise. The LNA is usually placed at the front-end of a receiver to constitute the first stage. Having a low noise figure, the LNA is crucial to the receiver's overall noise figure since it is largely dominated by the first stage as demonstrated in a previous section (equation 1.6).

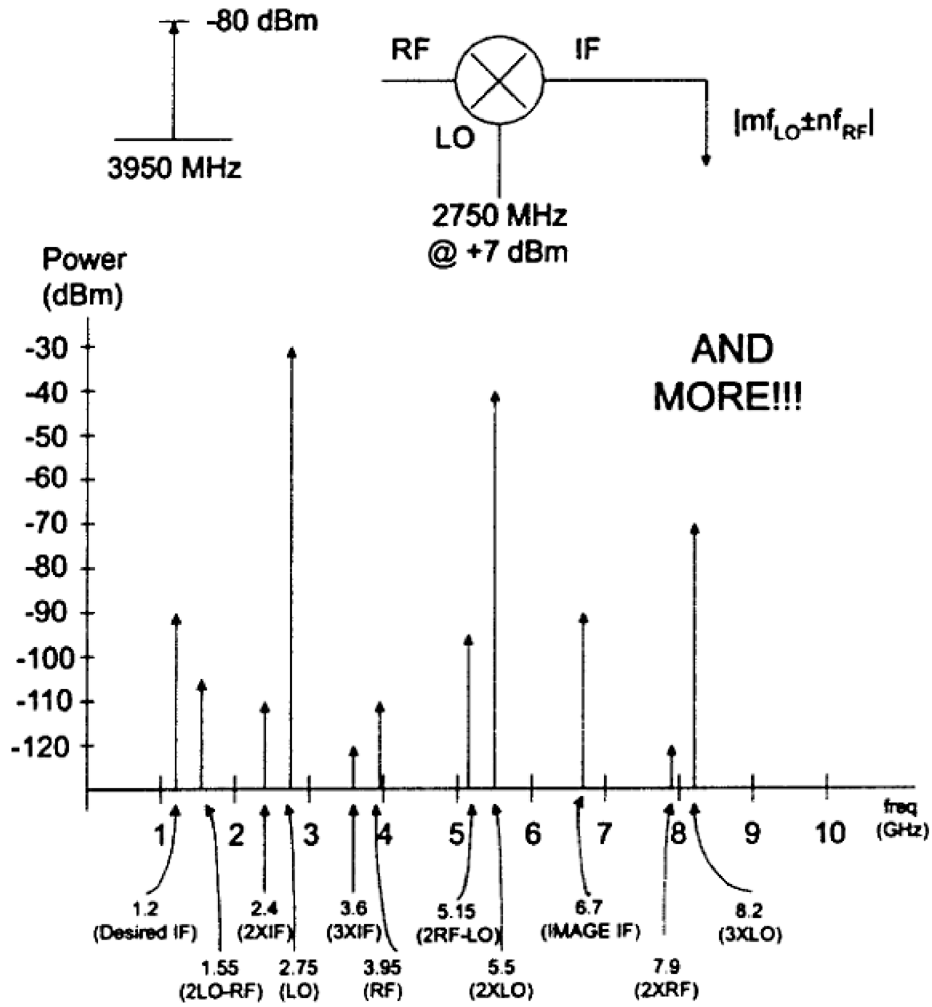


Figure 1.22: Output of a practical mixer

Friis' equation shows that the effect of noise from subsequent stages is reduced by the gain of the LNA while the noise of the LNA itself is injected directly into the reception chain. Thus the necessity to minimize the LNA's noise figure and maximize the gain.

Nonlinearity is the main issue along with noise figure for an amplifier. Figure 1.23 shows a typical curve of output voltage plotted against input voltage of a device. Ideally the curve would be a straight line extending indefinitely (linear), but, practically, it will have some curvature and eventually saturate. We can represent a curve such as this by a Taylor series:

$$V_{out} = a_0 + a_1 V_{in} + a_2 V_{in}^2 + a_3 V_{in}^3 + ... \tag{1.13}$$

Due to the higher order products in the transfer characteristics, many more frequency components will appear at the output of the device that are not at the input. For example if two frequencies f_1 and f_2 are fed to the amplifier then frequency components given by: $|mf_1 \pm nf_2|$ (where m and n are integers) will be present at the

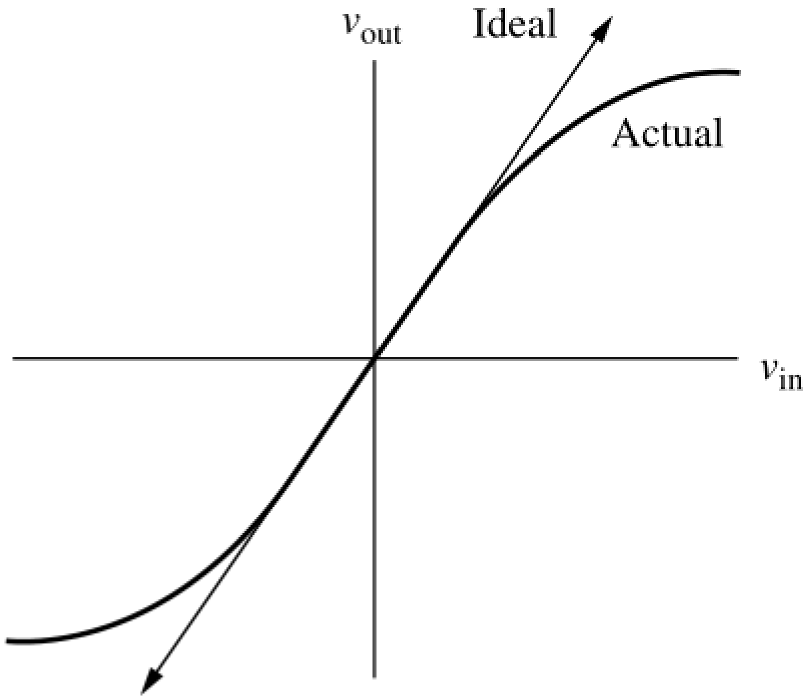


Figure 1.23: Input-Output characteristic of an ideal and practical device

output. These terms derive directly from the nonlinearity of the transfer function and this result can be derived by applying two cosines with frequencies f_1 and f_2 and using a few trigonometric identities after expanding the powers. The order of a term is defined as $m + n$. For example, "$2f_1 + f_2$" is a third-order term while "$2f_2$" is a second order term. We are often interested in the power of third-order terms because they are closer to the desired output in frequency and are harder to filter out.

It is very important to operate the amplifier in its linear region to avoid any spurious response. Two important parameters should be considered in this regard.

Compression Point P_{1dB}

As shown in Figure 1.23, the gain is not constant for all input power levels. As the input power levels increase beyond a certain point gain loss occurs as the amplifier goes into saturation. The 1 dB compression point (P_{1dB} often expressed in dBm) indicates the power level that causes the gain to drop by 1 dB from its ideal value. This is illustrated by Figure 1.24. It is important to restrict input power to a level below the compression point to avoid distortion.

This characteristic also applies to mixers and other nonlinear devices. Although passive devices have no "gain" we keep the same notation to be consistent.

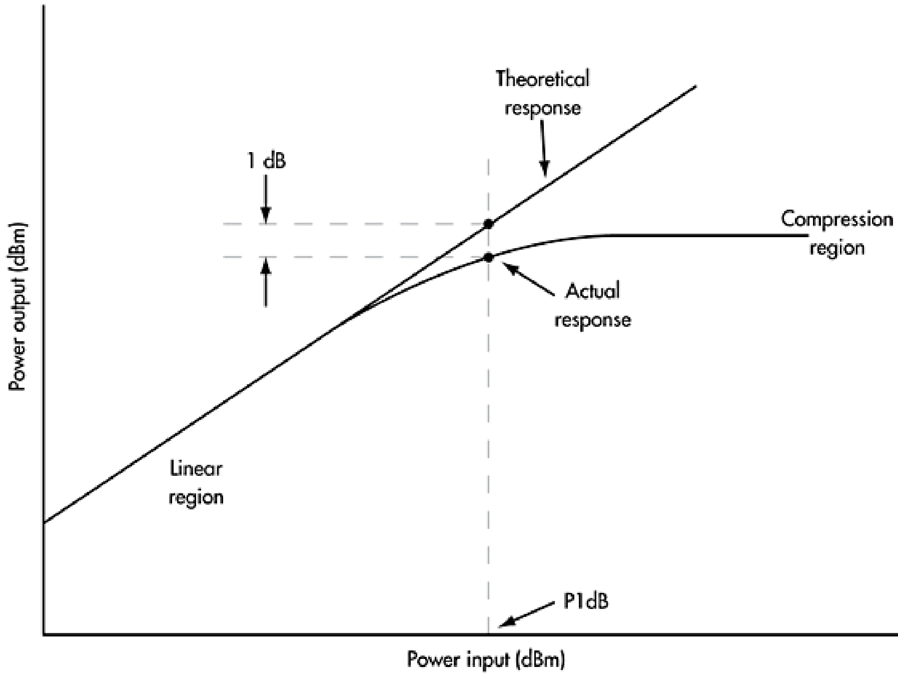


Figure 1.24: Figure illustrating gain compression and the 1 dB compression point.

Third-Order Intercept

When a device enters its nonlinear region, it begins to produce harmonics of the input signal (frequency components that are integer multiples of the input frequency nf_{in}). Luckily, these harmonics are easy to filter out since they are at least an octave higher than the fundamental (the desired output signal f_{in}) and they are usually out of the amplifier bandwidth.

Problems arise when we have multiple input frequencies. Due to the nonlinearity of the device, intermodulation products are produced at the output. These are the sum and differences of the input frequencies *and* their harmonics. These products can be very close to the desired output signal and well into the amplifier bandwidth. Therefore they cannot be filtered out easily and will ultimately become interfering signals to the main signals to be amplified. We are mostly concerned with the third-order products ($mf_1 + nf_2$, $m + n = 3$) because they are the closest to the main signal as shown in Figure 1.25.

The third-order intercept point (TOI or IP3) is obtained graphically by plotting the output power versus the input power on a logarithmic scale. Two curves are drawn, one for the linearly amplified signal, one for the third-order product. On a logarithmic scale, the function x^n translates into a line with a slope of n. Therefore, the linearly amplified signal will exhibit a slope of 1. And a third-order nonlinear product will exhibit a slope of 3.

The point where the curves intersect is the third-order intercept point. As shown in Figure 1.26 it can be read from the input or output power axis, leading to the input IP3 or the output IP3 (IIP3 and OIP3). The IP3 is a theoretical point

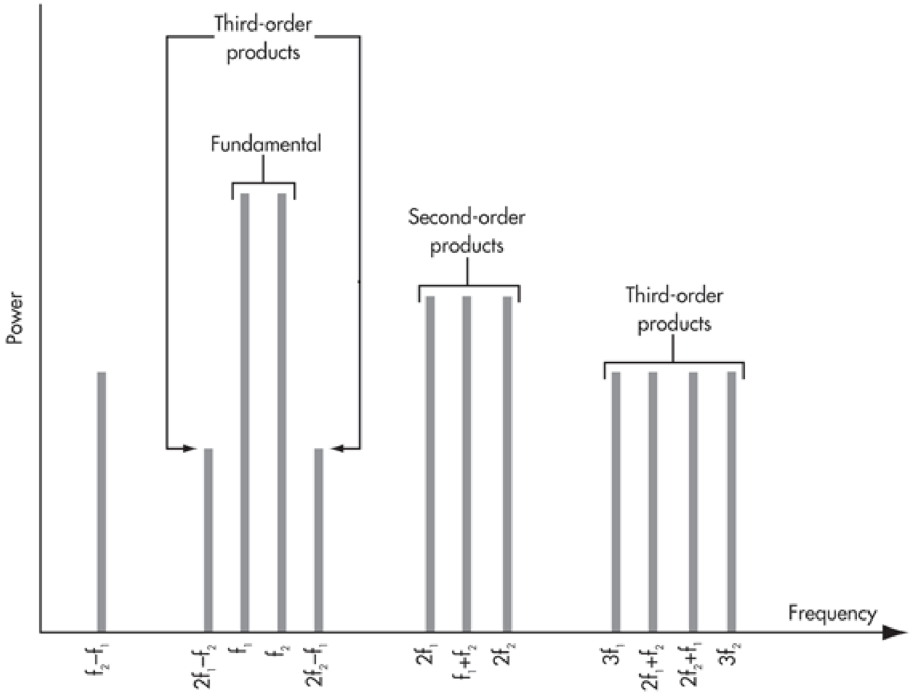


Figure 1.25: Intermodulation products of a nonlinear device

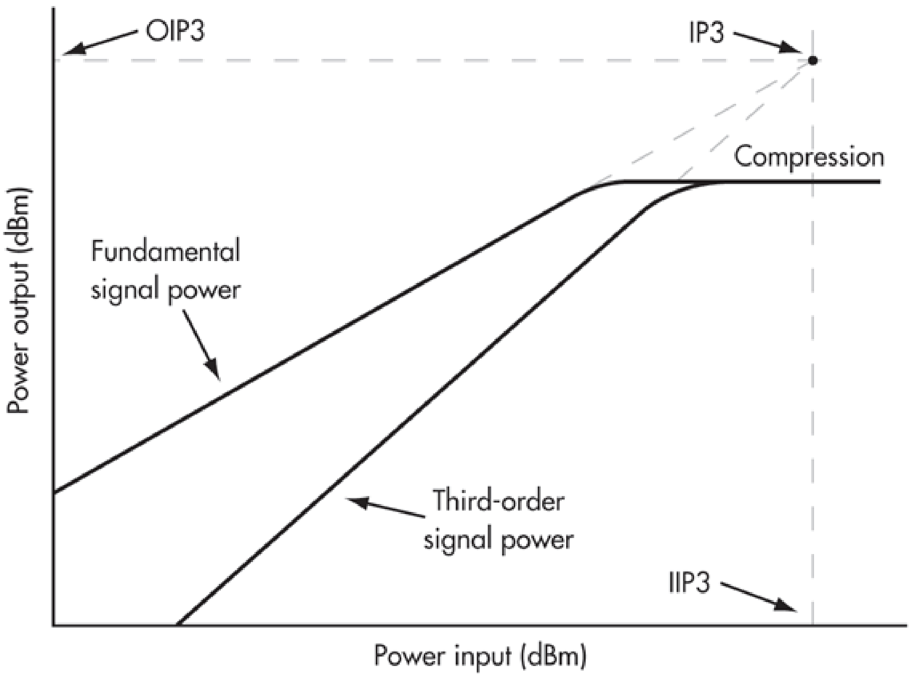


Figure 1.26: Input and Output third-order intercept points

that is never achieved in practice. However, it is useful in determining the linearity condition of an amplifier. *The higher the output at the intercept, the better the linearity and the lower the intermodulation distortion.* The IP3 value essentially indicates how large a signal the amplifier can process before intermodulation distortion occurs.

These measures also apply for any weakly nonlinear device such as general purpose amplifiers and mixers and should be considered in every stage of the receiver.

1.3.5 Local Oscillators

The local oscillator provides a signal with a stable frequency for mixing operations. Any signal that is mixed from one frequency to another by a local oscillator inherits "the faults" of that local oscillator [8]. So it is very important that an oscillator has a good performance for the overall operation of the receiver.

Ideally, an oscillator would provide a pure noiseless sine wave which corresponds to an impulse of zero width in the frequency domain. Unfortunately, all practical oscillators are far from ideal and their spectrum contains harmonics and subharmonics that are remnants of the oscillator realization. Figure 1.27 shows a the spectrum of a real oscillator. It can be seen that the oscillator is offset from its nominal frequency F_0 and its spectrum contains many subharmonics.

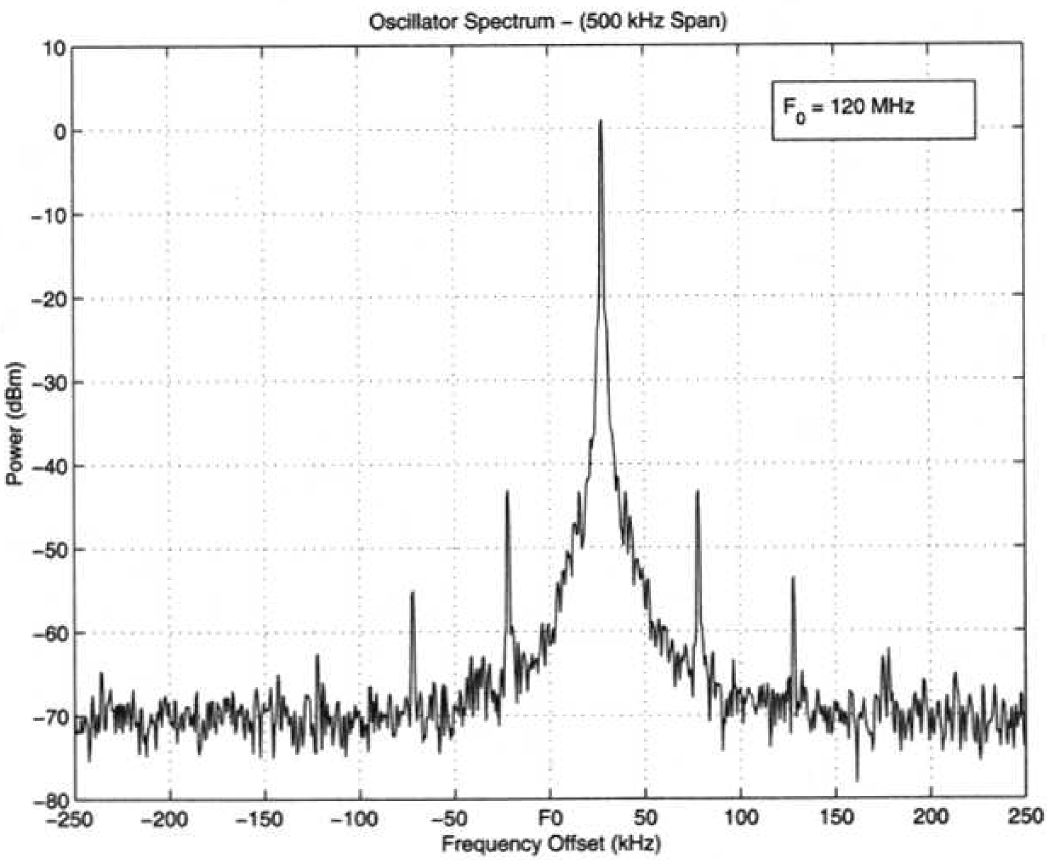


Figure 1.27: Spectrum of a real oscillator

A practical oscillator exhibits many undesirable effects mainly concerning its frequency stability. Changes in the oscillator's frequency over time periods greater than one second are referred to as drift. While frequency changes over time periods less than a second or so are called phase noise [8].

Phase noise is measured in dBc/Hz at a given offset from the carrier. This is the noise power in a 1 Hz bandwidth at the offset from the carrier. For example, -80 dBc/Hz at an offset of 10 kHz means that the noise power density at 10 kHz away from the carrier frequency (or the desired LO frequency) is 80 dB lower than the power of the carrier (the LO).

Oscillator phase noise can limit the ultimate SNR of any signal processed by a receiver. It can cause unwanted signals to mask wanted signals. In radar systems, receiver phase noise helps clutter mask the targets we want to see [8]. Thus considerable care should be taken while choosing an oscillator for the receiver.

2

Design and Implementation

In this chapter, we discuss the design and implementation of a multi-channel (four channels) FM receiver and show experimental results after assessing the performance of the receiver.

The design process was highly iterative. Basically we had to review and redesign almost every single stage every time we added or modified something.

Our receiver's main objective is to be able to tune in to the same FM-band on its four channels and sample them simultaneously for further processing. This requires all four sub-circuits to operate with the same oscillator and that the signal paths be identical to minimize phases errors. This implies that a great deal of symmetry must be devised in the layout of the circuit board.

The receiver consists of four circuits on one board operating under one oscillator output and the same clock for sampling and digitization. This board will be connected to four antennas through coaxial cables and to a digitizer through an interface that will be discussed later.

2.1 Choice of Architecture

The low-IF architecture offers a compromise between the stringy filtering requirements and image rejection of the heterodyne receiver and the practical issues such as DC-offset of the zero-IF receiver. Moreover, its component count and overall complexity is almost identical to the zero-IF receiver if we do IF sampling. Which is going to be the case as explained in the following sections.

The four sub-circuits will all employ a low-IF architecture that will translate the signal of interest (the desired FM-channel) to a low frequency IF where the ADC is able to sample it. The type of ADCs available and the data rate of the current general purpose protocols restrict us to a few MS/s. Furthermore, this data rate is to be divided by 4 since we are doing simultaneous sampling. This means that we can only sample signals whose bandwidths do not exceed a few hundred kilohertz. Therefore, we will not be able to sample more than one FM channel at a time.

Moreover, the signal of interest (the desired FM channel) must be translated to a somewhat very low IF for the ADC to be able to sample it. We have chosen the IF to be 75 kHz which is enough to translate one channel from DC to 150 kHz. This channel will be directly sampled by the ADC.

Figure 2.1 shows the basic blocks of the receiver. These blocks will be expanded and their technical details developed as well as their interconnections and interfacing in the next sections.

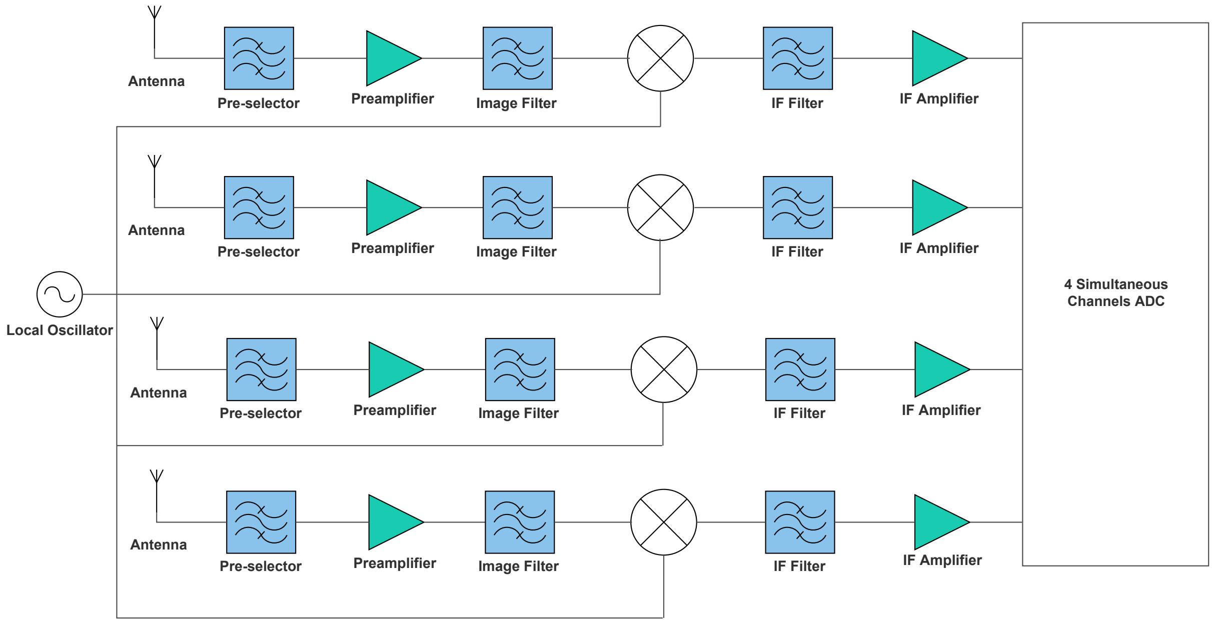


Figure 2.1: Simplified block diagram of the multi-channel receiver

In addition, we devised an undersampling scheme (not shown in the figure) in which we will be using a powerful digitizer (60 MS/s) to undersample half the FM bandwidth (10 MHz).

The FM band extends from 88 to 108 MHz. We are interested in capturing one of the channels at the lower end of the FM bandwidth and translate it to a few hundred kilohertz. After inspecting the list of all FM transmitters and the corresponding frequencies in the region as well as their transmitting power and distance; we have chosen "Chaîne 3" which is transmitting from Chréa at 10 kW of power and a center frequency of 88.4 MHz [1]

2.2 Antenna

The antenna in our receiver is going to be a quarter wavelength monopole mounted on a half wavelength ground.

[1] http://www.elahcene.co.uk/algeria/chrea.htm

The monopole's radiation pattern is shown in Figure 2.2. It shows that when placed vertically, the monopole has a maximum radiation power at about 25° above the horizon which is desirable for radar systems.

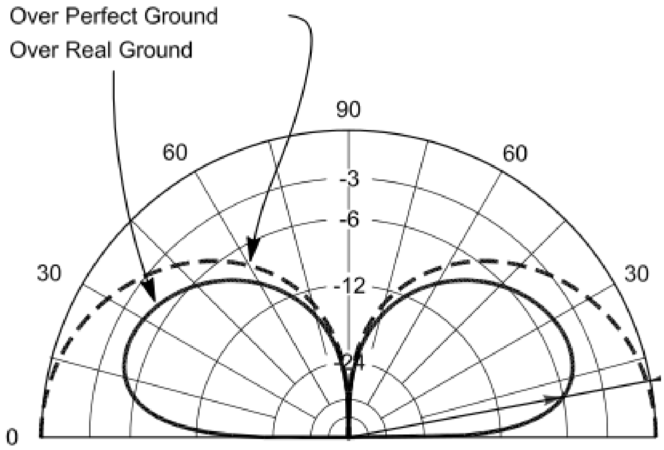


Figure 2.2: Quarter-wavelength Monopole Radiation Pattern

The conductors of the antenna are stranded copper wires to minimize losses due to skin effect. The monopole is 0.8 meter ($\lambda/4$) long mounted on $\lambda/4$ radial grounds. The length takes into account the velocity factor of the insulated copper wire which is given in the literatures as approximately 0.93.

We have made a PVC structure to accommodate the conductors and to maintain their position and protect them from winds and moisture to avoid static charge accumulation, this structure will also contain the preamplifier module which is connected through a coaxial cable. The feedline base containing the preamlifier module is shielded using aluminum foil. The 10 meter long cable is then connected to the main board containing the rest of the components.

The antenna must be connected to a DC ground either by the structure of RF filter or by a separate RF choke or a high-valued resistor. Any static charge accumulated on the antenna will translate to a high voltage in the internal receiver components due to the low value of capacitors used at RF frequencies. This is why many pieces of equipment can be seen with a choke soldered to the antenna terminal [9]. We used a 10k Ω metallic film resistor to form a DC path to ground.

2.3 Pre-selector Filter

The pre-selector must perform all the operations discussed in the previous chapter. This filter should also pass the signal of interest with minimum attenuation. The signal of interest is the Chaîne 3 which occupies a 200 kHz band around a center of

88.4 MHz.

We decided to preselect only the first half (or less) of the FM band since it will halve the noise bandwidth. Moreover, sampling the whole FM band provides little interest in our application.

Therefore, the preselector's passband should be from 88 to 98 MHz (or less). Practical realizations with minimum component count will help decide the upper cutoff frequency. We used Agilent's Advanced Design System (ADS) [2] filter design tool to produce the preliminary filter architecture and to simulate the results.

The filter type does not really matter in our application since we do not intend to recover the signal perfectly. Magnitude and group delay flatness are not key parameters here. We only need the same response to be experienced by the signal in all four channels. Therefore, we are only interested in the *reproducibility* of the filter in all the circuits since this will introduce minimal phase error.

Since the filter type is unimportant, we are going to choose whichever one that produces the most simple topology in terms of component count and the one whose component values are close to standard values found in the market. Component count is very important here because we are going to make many instances of this filter later on. After tuning and substituting standard component values we get the 2^{nd} order Butterworth filter in Figure 2.3.

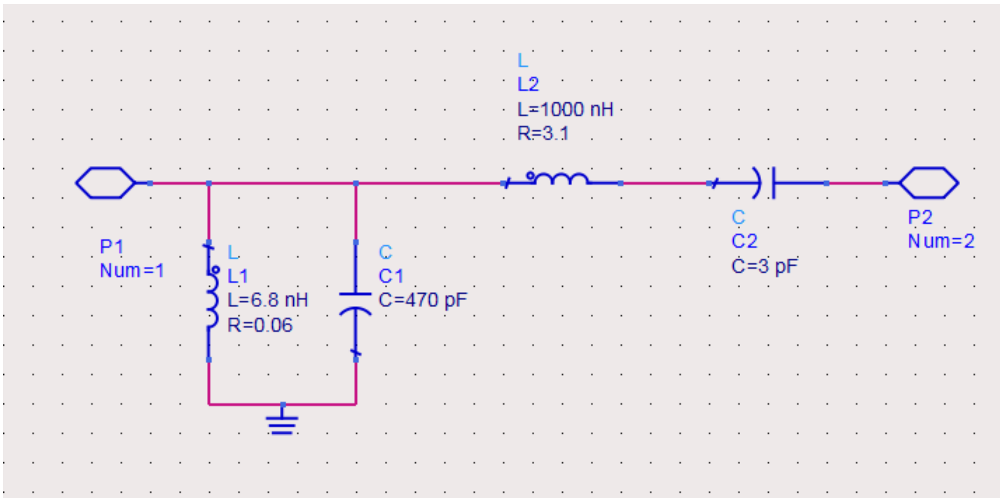


Figure 2.3: Preselector circuit

One more advantage of this topology is that we do not need to add a coupling capacitor to the filter in order to preserve the bias of the preamlifier which comes right afterwards.

[2]http://www.keysight.com/find/eesof-ads

Simulation using *real* component parameters shows that the filter has an insertion loss a little bit less than 2 dB in the passband (Figure 2.4). The group delay is also shown in Figure 2.5.

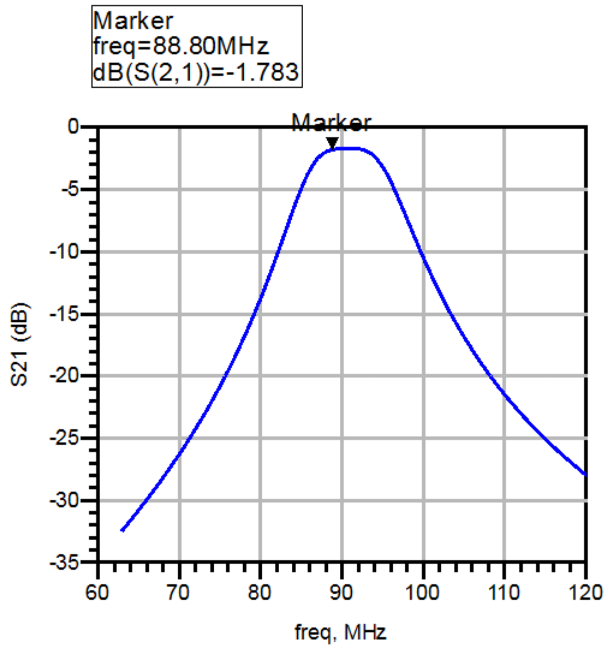


Figure 2.4: Preselector magnitude response

All components of this filter are SMD components which have a far better performance than their through-hole counterparts.

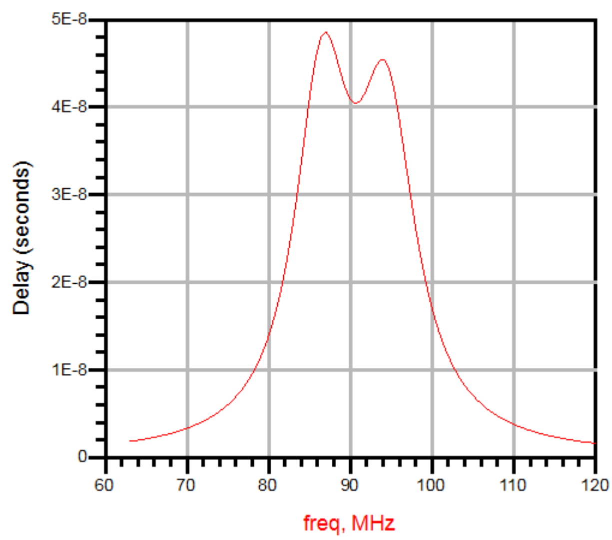


Figure 2.5: Preselector group delay

2.4 Preamplifier

For the preamplifier we use an LNA as discussed in the preamplifier section. The most relevant parameter here is the LNA's noise figure.

The PSA-5455

We are going to use an ultra low noise amplifier with a noise figure of 0.8 dB from MiniCircuits.

The PSA-5455 is internally matched to 50 Ω and provides a gain of 23 dB in the FM band as shown in Figure 2.6.

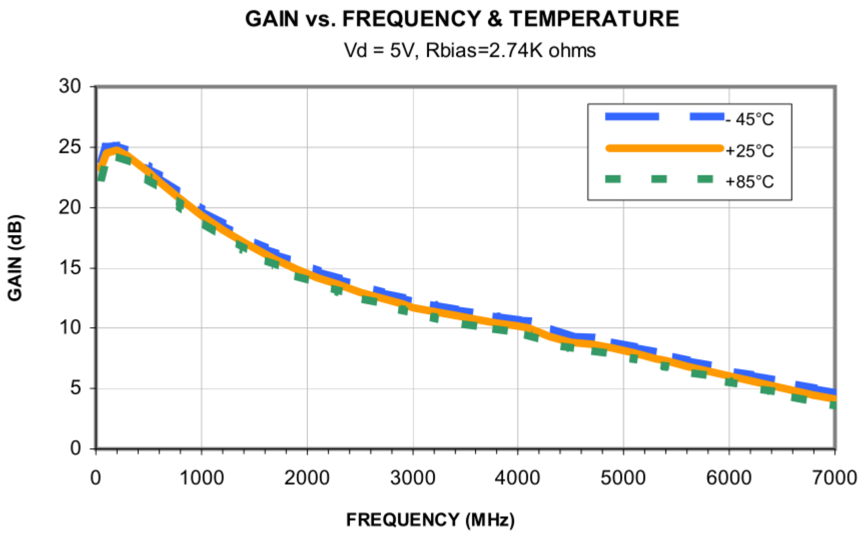


Figure 2.6: Gain vs. Frequency for the PSA-5455

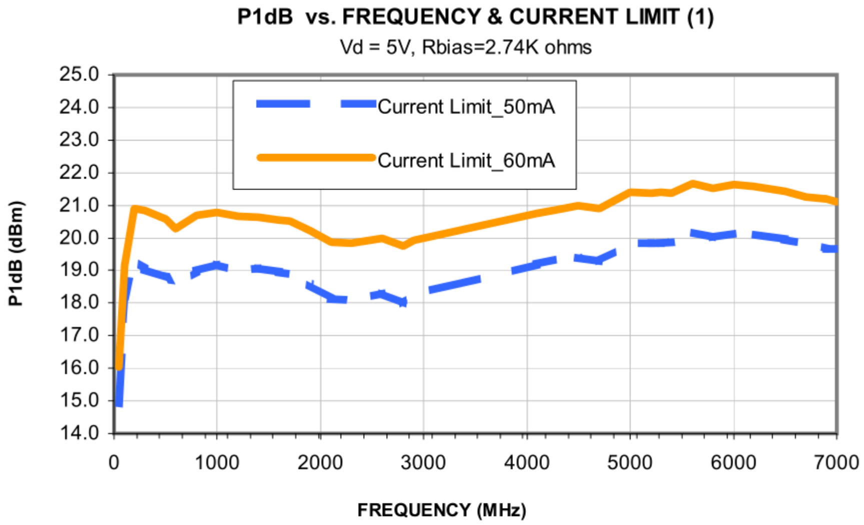
Figure 2.7 shows that the Output 1 dB compression point is around 15 dBm which is way higher than our expected power output (-89 to -29 dBm).

Figure 2.8 shows that the Output IP3 is 27 dBm at worst. This is way beyond our expected power level at the output of the LNA.

We conclude that the LNA is adequate for our application and will produce minimal spurious responses and noise.

Biasing and Interface of the PSA-5455

To function within the aforementioned specs, the LNA must be properly biased. For that, we use the recommended biasing circuit provided by the manufacturer.



Output Power @ 1 dB compression (P1dB) (2)	0.05		14.9		dBm
	0.5		18.8		
	1.0		19.2		
	2.0		18.5		
	3.0		18.3		
	4.0		19.1		

Figure 2.7: Output 1 dB compression point vs. Frequency for the PSA-5455

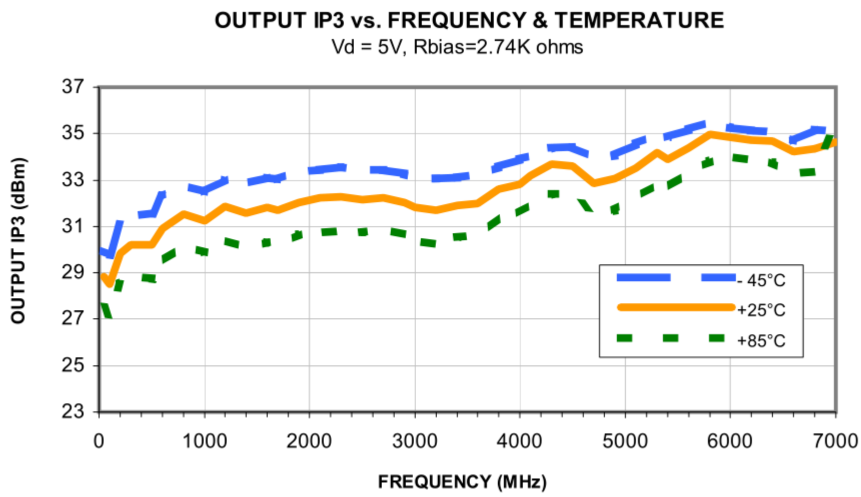


Figure 2.8: Output IP3 vs. Frequency for the PSA-5455

Figure 2.9 shows the bias circuit as well as the connection to the preceding preselector. The filter is set to see a 50 Ω on its terminals so there is no need for matching since the LNA is internally matched to 50 Ω.

A doubly terminated filter should be injected after the LNA to stop any spurious responses from propagating further into the reception chain. This filter also attenuates the image signal and prevents it from getting downconverted to the IF. This is the image filter. Our previous preselector filter is adequate for this operation as it will suppress any harmonics or third order IM products as well as the image frequency which will be in the band from 77.325 – 88.325 MHz.

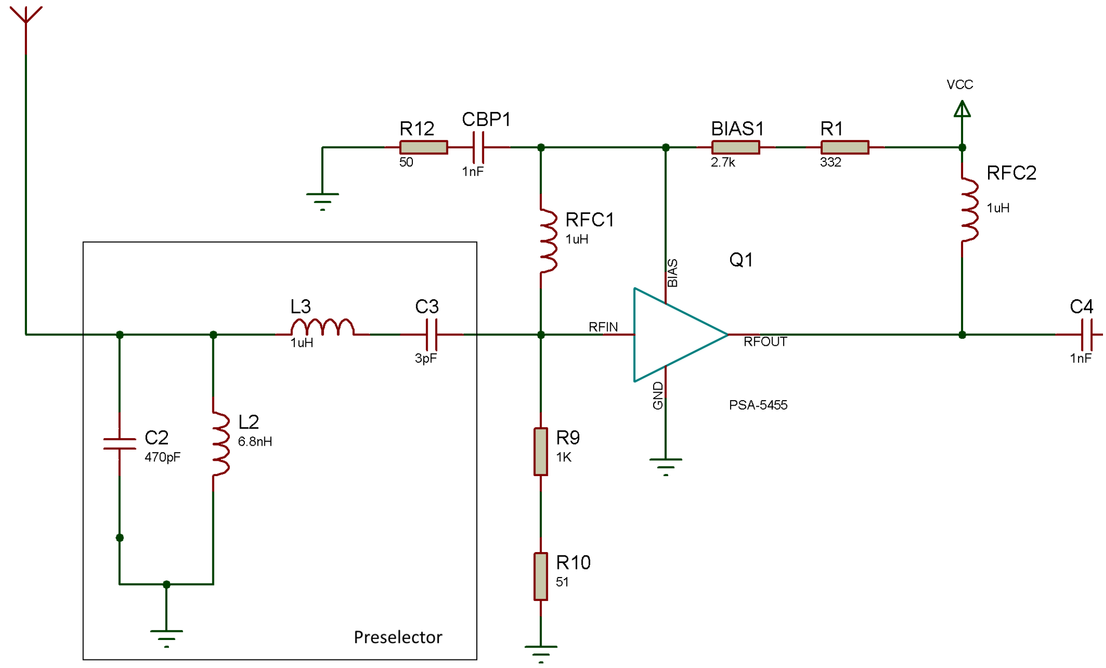


Figure 2.9: Circuit schematic of the PSA-5455 bias and the preceding preselector

2.5 Mixer

For the mixer, we use a double-balanced passive diode mixer from MiniCircuits. The SBL-1 is a military grade mixer operating from 1 to 500 MHz with an excellent 5.6 dB conversion loss.

Figure 2.10 shows the mixer's conversion loss vs IF at an LO power level of 7 dBm.

The SBL-1 also provides great isolation between ports. Figure 2.11 shows that we get about a 55 dB of isolation between LO:RF (since our LO will be around 80 MHz). Figure 2.12 shows that we get about the same isolation for the LO:IF port.

The mixer's mismatch to a 50 Ω load is shown through the "VSWR vs Frequency" plot in Figure 2.13.

The mixer exhibits a very complicated impedance behavior at its ports. A mixer cannot be modeled as a static load. In fact, a mixer is both a load and a source. It can be easily shown that all three mixer ports act as sources for the intermodulation distortion intrinsic to the diodes [10]. A problem arises when a mixer port is reactively terminated with, for example, a bandpass filter (**which is our case [the image filter]**). Acting as a source, the mixer produces the 2LO - RF product which exits the RF port and enters the RF filter. Most commonly, this product is outside the passband of the filter and will hence be reflected back into

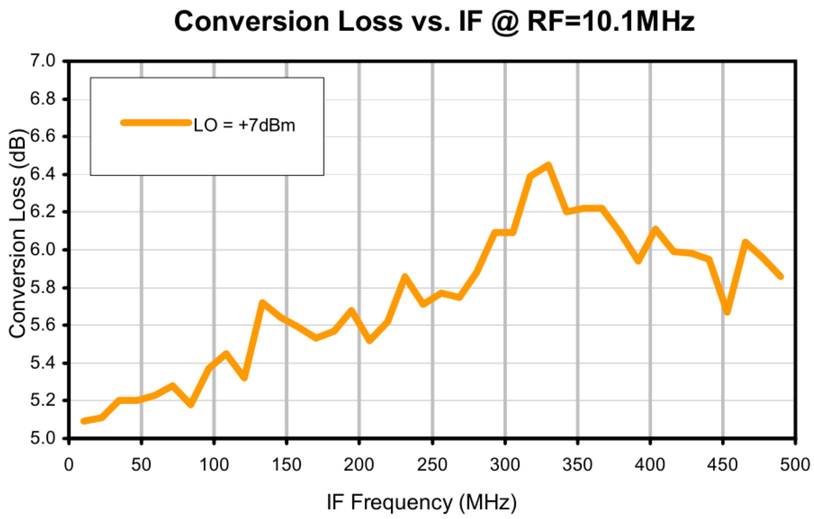


Figure 2.10: Mixer's conversion loss

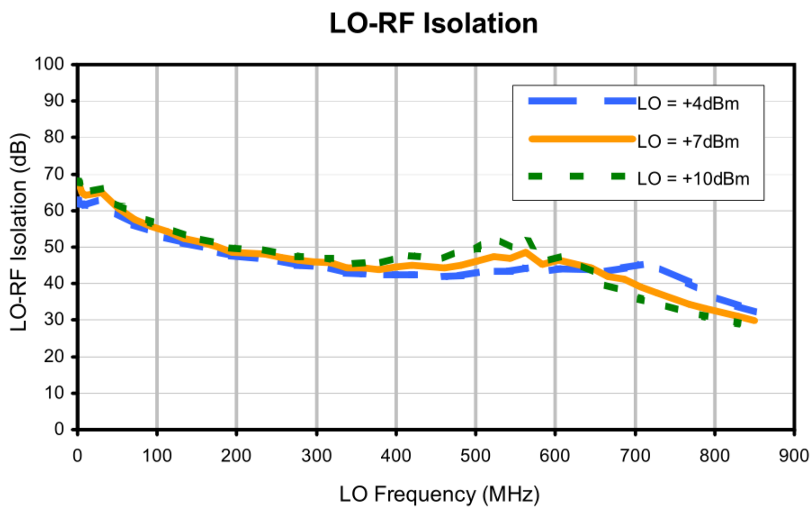


Figure 2.11: LO:RF isolation vs LO frequency

the mixer and will be downconverted to the IF frequency [11].

To remedy these issues, a non-reflective match such as a 50-Ω-terminated diplexer should be used to present a 50 Ω load for the mixer [11]. As a cheaper alternative, we used a resistive attenuator instead. This attenuator will also alleviate the sensitivity of the mixer to the load which is a very cheap and efficient alternative to the expensive "termination-insensitive mixers" [12]. This will also improve the VSWR since the attenuator will absorb the reflections. Moreover, the preceding image filter and the IF filter are very sensitive to terminations, so an attenuator is very well needed to present a stiff wideband 50 Ω load to them.

We use a 4 dB resistive matching attenuator which corresponds to a 2×4 dB $= 8$ dB minimum return loss. The T-pad is shown in Figure 2.14. This attenuator will present the varying impedance of the mixer as a stiff resistive load that is ap-

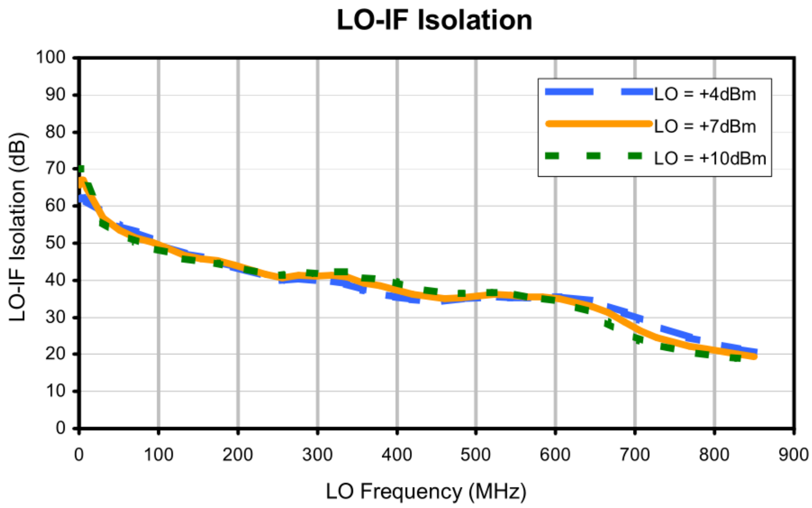


Figure 2.12: LO:IF isolation vs LO frequency

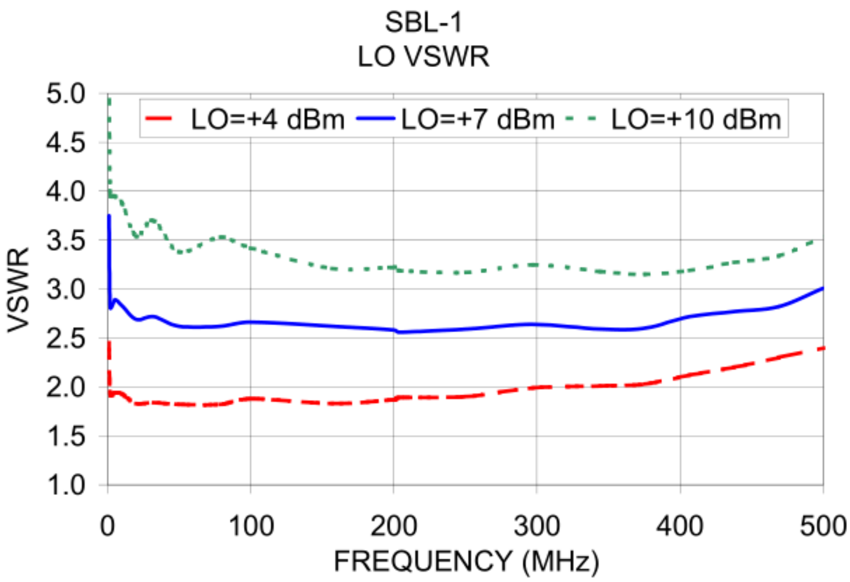


Figure 2.13: SLB-1 VSWR on the LO port

proximately 50 Ω. This is very important since we are going to place filters before and after the mixer and filters are very sensitive to terminations.

Figure 2.15 shows the mixer stage. We will discuss the LO connection to the mixer in the Oscillator section.

Since we are going to be using a coaxial cable that has a characteristic impedance of 75 Ω and both the LNA module and the Mainboard (Mixer) have a 50 Ω impedance at their ports, there will be some loss due to the mismatch. For that, we will use two LC networks to match the system. Figure 2.16 shows the used circuits and the results after running a simulation using Multisim. The figure shows that the network analyzer sees a 50 Ω impedance at the input and the output of the two

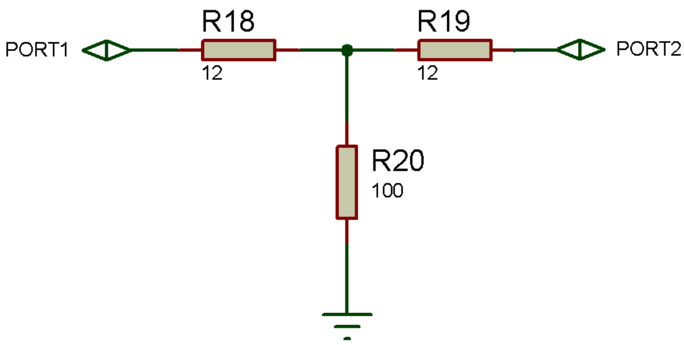


Figure 2.14: 4 dB T-pad attenuator

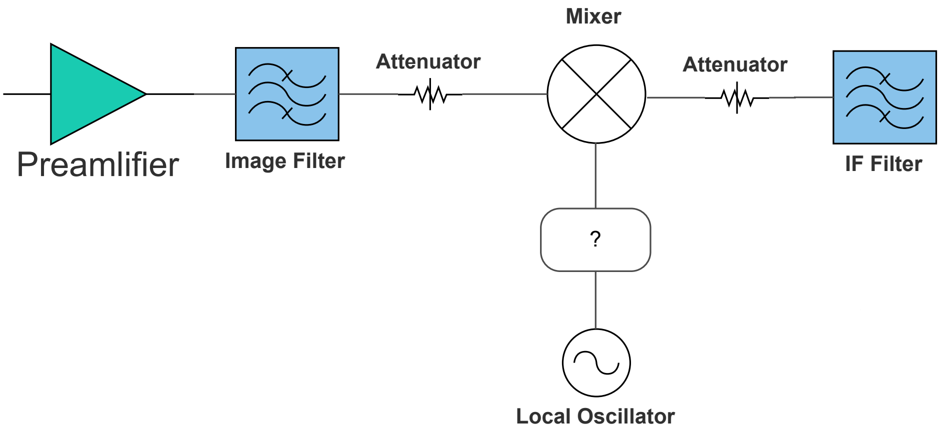


Figure 2.15: Mixer stage block diagram

port network. This is only valid in the band of frequencies we are interested in as reactive matching is frequency dependent.

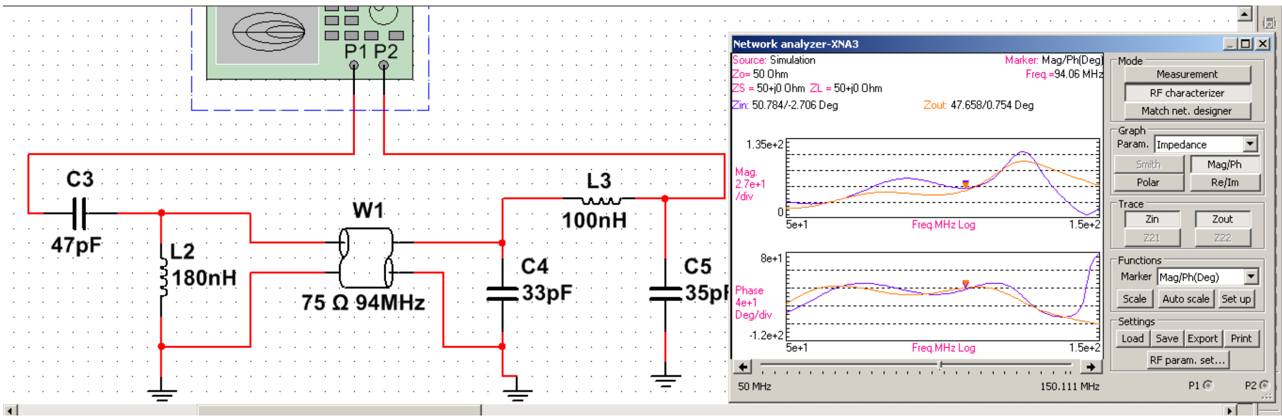


Figure 2.16: Simulation of the matching network

2.6 Oscillator

Special care should be taken when designing the oscillator stage since any inherent artifacts or noise in the oscillator will be produced in the downconverted signal and will thus have a big impact on the SNR of the whole system.

For a high stability oscillator output we use an RF frequency synthesizer. The Si4123 contains three PLLs (Phase Locked Loop) with integrated VCOs (Voltage Controlled Oscillator) to produce two outputs:

- The RF output ranges from 750 MHz to 1.8 GHz.
- The IF output ranges from 62.5 MHz to 1 GHz.

We are going to use the latter output for our LO. These outputs and their frequencies are controlled (programmable) through a three-wire serial interface. This is done through a microcontroller.

Setting up the PLL

To setup the PLL we inspect the datasheet for the required connections and for the needed external components.

The PLL requires an external clock on the XIN pin. For that, we are using an HCMOS/TTL 10 MHz clock generator with a 50 ppm frequency stability. The clock output is connected to the pin through a 560 pF capacitor. This same clock will drive the ADC later on. This connection, along with other connections and external components, is shown in Figure 2.17.

The PLL is connected to a microcontroller through a three-wire serial interface. The connection is not shown in the figure. This will allow us to program the required frequency to be synthesized with a step of 1 kHz.

Oscillator - Mixer Interface

The mixer requires an LO power level of 4 - 10 dBm to function within the advertised specs (which corresponds to 2.5 - 10 mW). The Si4123 provides a typical power level of -4 dBm for a 50 Ω load. Moreover, the same LO drives all four mixers which multiplies the previous levels by four. Hence the need for an amplifier to provide some gain to the LO signal before being fed to the four mixers.

Before feeding the LO signal to the amplifier we need to filter the output of the LO to prevent any harmonics from getting amplified and propagating to the reception chain. Another filter (injection filter) should be injected after the amplified LO to remove any spurious frequencies originating from the amplifier's nonlinear

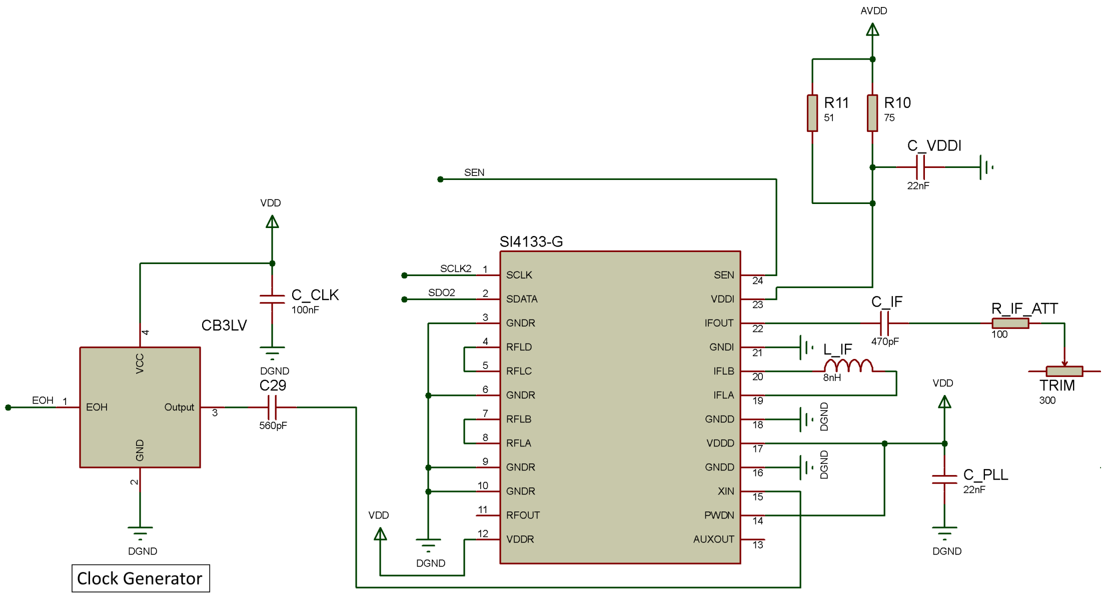


Figure 2.17: PLL Setup

effects. Figure 2.18 shows the amplification and filtering of the LO. An attenuator has been added to control the amplification factor since the output of the PLL is not precisely known a priori (in the datasheet it says -8 to 0 dBm).

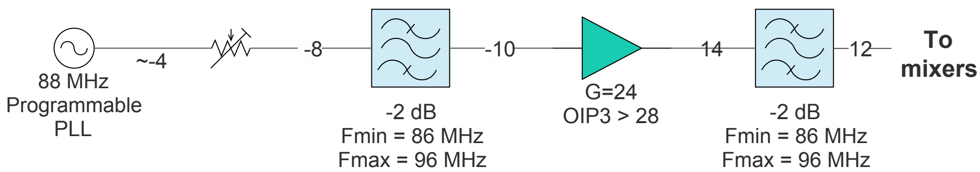


Figure 2.18: LO amplification

For the amplifier, we use Analog Devices' ADL-5545 which has a fixed gain of 24 dB from 30 MHz to 6 GHz. A P1dB compression point of 16 dBm and an OIP3 of 36 dBm. Our expected output power level is around 12 dBm so we are somewhat below the spurious range. The amplifier is internally matched to 50 Ω which allows it to be directly connected to the filters on both its input and output ports.

For the filters, we employ our previously designed preselector which will do the job of "purifying" our 88.3 MHz signal (our LO signal). The circuit schematic of the amplifier stage is shown in Figure 2.19.

Next, we have to devise a method to equally divide the output power of the LO between the four mixers.

Power dividers are needed to equally distribute the LO power to the four mixers. While Wilkinson dividers have the advantage of high interport-isolation and lower

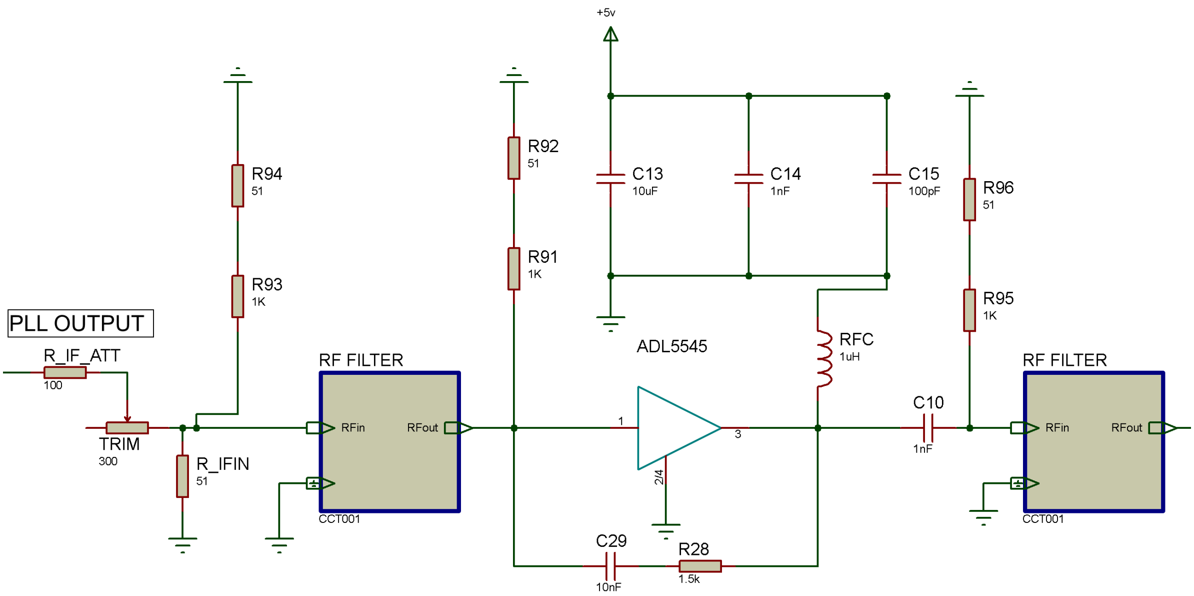


Figure 2.19: LO amplification circuit diagram

losses than their resistive counterparts, they are practically limited to a few hundred MHz in their lower frequency range while resistive power dividers are wideband and can reach DC.

Since we need to present the mixers' LO input with a 50 Ω wideband impedance, we cannot use a Wilkinson divider because of its bandlimited match. Moreover, it is not easy to preserve the phase of the oscillator in all three ports of the divider which is crucial in our application.

With this in mind, we use a 1/4 resistive power divider as shown in Figure 2.20 where RL1,2,3 along with the impedance of the network analyzer represent the load impedances of the mixers. The divider is made of 27 Ω high quality 1 % metal film resistors. The same simulation gives that the Total Power Gain is about -11 dB.

To assess the performance of this divider we add the injection filter and check if the impedance seen by the mixer is always 50 Ω. Figure 2.21 shows that indeed the divider presents a very stiff 50 Ω impedance for all mixers both in the passband and the stopband of the filters.

2.7 IF Stage

The IF amplifier is an OP Amp used in an inverting configuration. This configuration allows us to have a variable gain by using a potentiometer as a feedback resistor. This will increase the dynamic range of the receiver.

The IF stage is shown in Figure 2.22 where we are using a low noise, precision

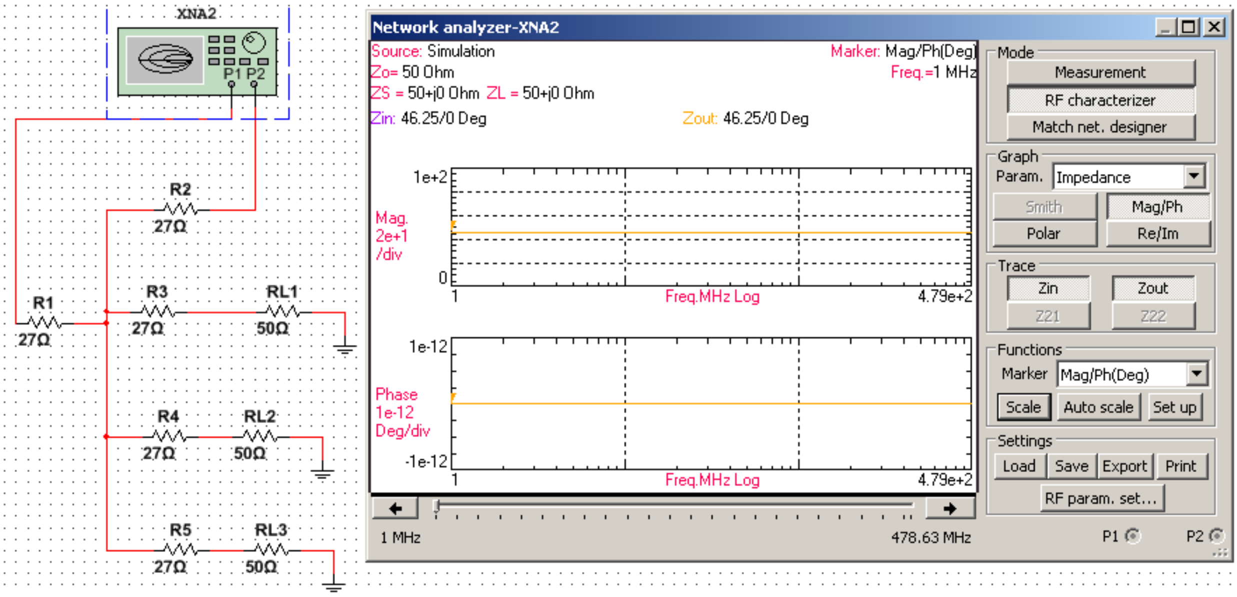


Figure 2.20: Resistive power divider impedance characteristic

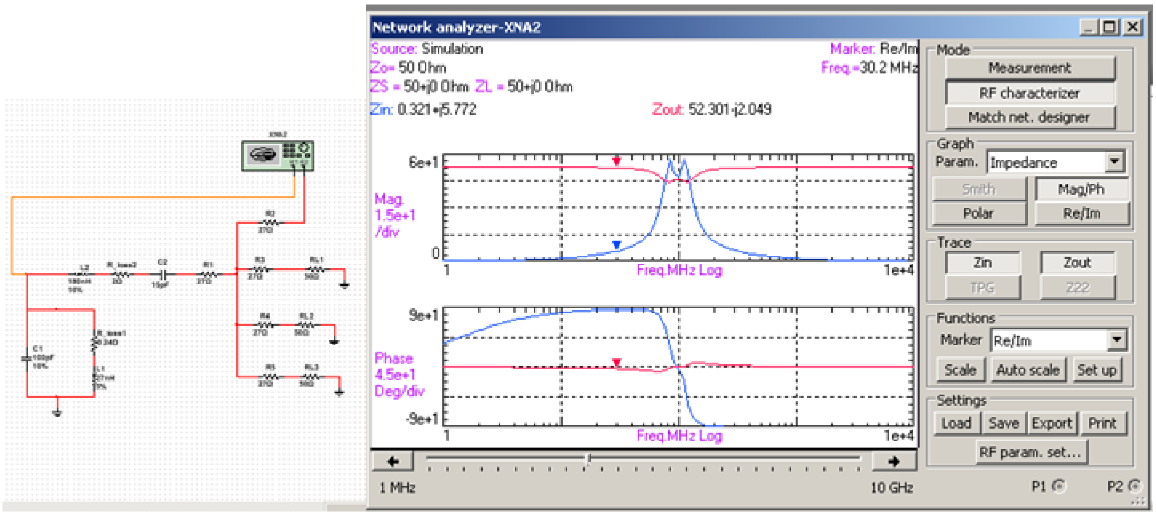


Figure 2.21: Resistive power divider with injection filter performance

operational amplifier. The OP27 has an input noise of 3 $nV/\sqrt{Hz}$ and an 8 MHz gain bandwidth as well as an excellent CMRR = 126 dB.

Since the downconverted channel is from DC to 150 kHz, we are going to need a low pass filter covering this frequency band. The filter topology and magnitude response are shown in Figure 2.23 and 2.24 respectively.

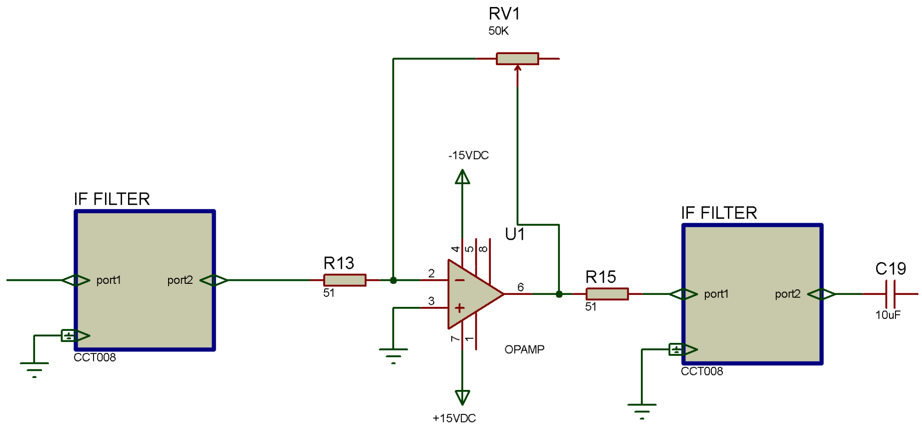


Figure 2.22: IF Stage diagram

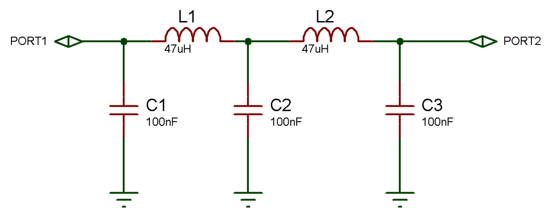


Figure 2.23: IF Filter Circuit

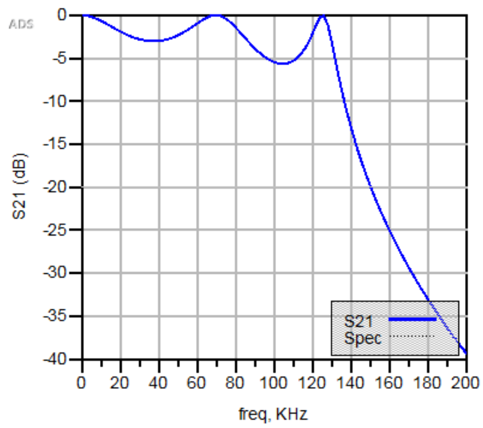


Figure 2.24: IF Filter Magnitude Response

2.8 ADC and Acquisition

Using dsPIC

As noted earlier, it is crucial to sample all four channels simultaneously. Thus the need for perfectly synchronized ADCs or ones that offer simultaneous sampling. We are going to go with the latter choice and make use of the dsPIC's four simultaneous channels. **dsP**IC is a microcontroller that has **d**igital **s**ignal **p**rocessing capabilities.

The dsPIC33FJ64GP802 can sample four channels simultaneously at a 10-bit resolution with a maximum sampling frequency of 280 kHz.

The dsPIC then transfers the sampled channels through SPI to a Raspberry PI. The latter is connected to a computer through ethernet for later processing.

Using a digitizer

We also have a professional digitizer at our disposal with the following characteristics:

- Model: NI PXI-5105
- Sampling rate = 60 MS/s
- 12 Bit resolution
- 8 Simultaneous sampling channels
- Input impedance: 50 Ω or 1 MΩ (software selectable)
- Analog bandwidth $\simeq$ 60 MHz

The digitizer is connected to a computer through a high speed ExpressCard that is capable of transferring such data rates. Figure 2.25 shows the digitizer connected to our four antennas through SMB cables.

Figure 2.25: The NI PXI-5105 digitizer

At the computer side, we are using NI-Scope to visualize all four channels simultaneously. The software also allows us to perform several useful operations on the signals such as an FFT. It also provides a means to select many functionalities of the digitizer such as changing the input impedance, sampling rate, and activating/deactivating the built-in anti-aliasing filter.

2.9 Power Supply

The power of the system is drawn from two transformers, one for analog, and the other for digital components.

In our receiver, we will be needing many voltages for the different components we are using. For example, the OP27 requires a +15 V and -15 V power supply, the PSA-5455 requires a 5V, and the PLL requires 3.3 V. For that we use adequate voltage regulators and the corresponding filtering capacitors:

- S111733PI - 3.3V.
- The 7815 for the +15V and the 7915 for the -15V rail.
- The 7805 for the +15V rail.

3
Experimental Results

In this chapter, we present the experimental results that we have obtained after probing and testing the receiver at each stage.

3.1 Antenna

To test the antenna, we make use of a spectrum analyzer and a coaxial cable. We could not find a 50 Ω cable so we are using a 75 Ω one instead. Figure 3.1 shows the final antenna in its PVC structure.

Figure 3.1: Final antenna

The radiation resistance of a quarter-wavelength monopole antenna is determined theoretically to be about 36 Ω. But that could be different from the actual value since this impedance depends on the surrounding environment which is uncontrolled. Moreover, we have an array of four antennas which means that there will be coupling effects that will change the impedance of the antenna. Therefore the input impedance is to be determined by sophisticated equipments such as a network analyzer.

We connect out antenna to the spectrum analyzer using the coax and we immediately see the spectrum from around DC to 1 GHz. Figure 3.2 shows the the results. We can clearly see:

- The FM band from 88 to 108 MHz.
- The VHF TV band from 203 - 233 MHz.
- The UHF TV band from 634 to 700 MHz.

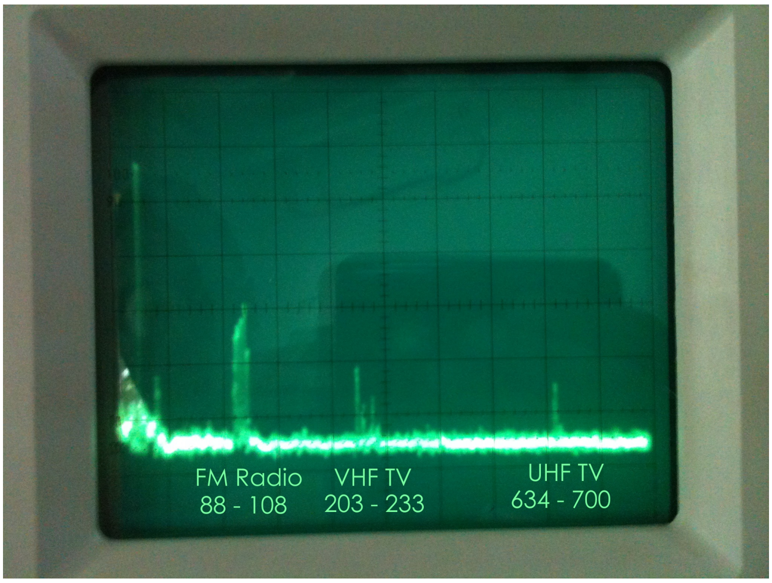


Figure 3.2: Spectrum of received signal using the monopole antenna

If we zoom into the FM band we can clearly see the different channels being broadcast in Figure 3.3.

The power level at the FM band ranges from -75 to -68 dBm inside the building of the laboratory.

3.2 Preselector Filter

To test the filter we connect its input to the receiving antenna and its output to the spectrum analyzer.

Figure 3.4 shows that the filter has severely attenuated all frequencies outside the FM band and that they are at least 20 dB weaker than the signal of interest. We observe that there is some remnant power at higher frequencies (around 700 MHz) which may be due to the imperfections of the filter or a re-entrant response.

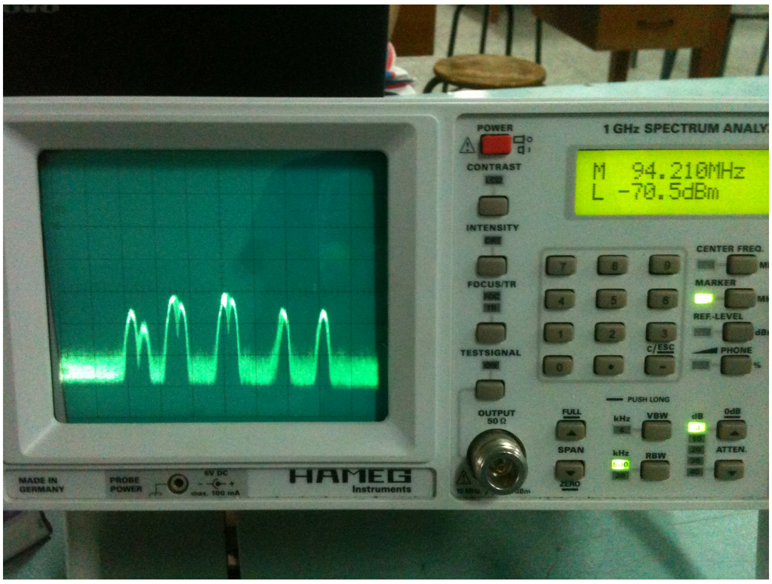


Figure 3.3: The FM band and the individual channels

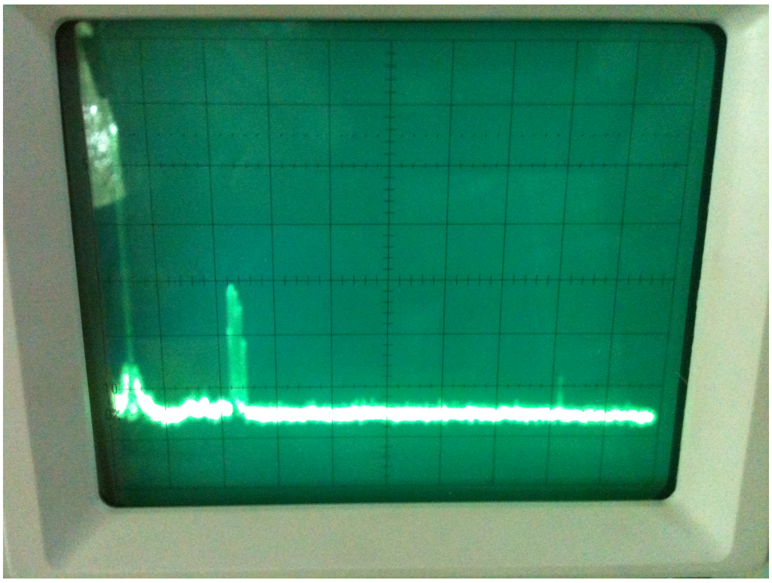


Figure 3.4: Spectrum of received signal after filtering

3.3 LNA Module

The final circuit of the LNA module is shown in Figure 3.5. It has been implemented on a small printed circuit board that contains SMD components only. The printed circuit board is small in size and could fit in the base of the antenna. Power supply is provided through an external cable.

The signal is visualized before and after amplification in figures 3.6 and 3.7. It can be seen that there is a gain of at least 22 dB (the levels fluctuate) after passing through the LNA module. It should be noted that the noise floor has been raised by the amplifier but the SNR is in the order of 40 dB.

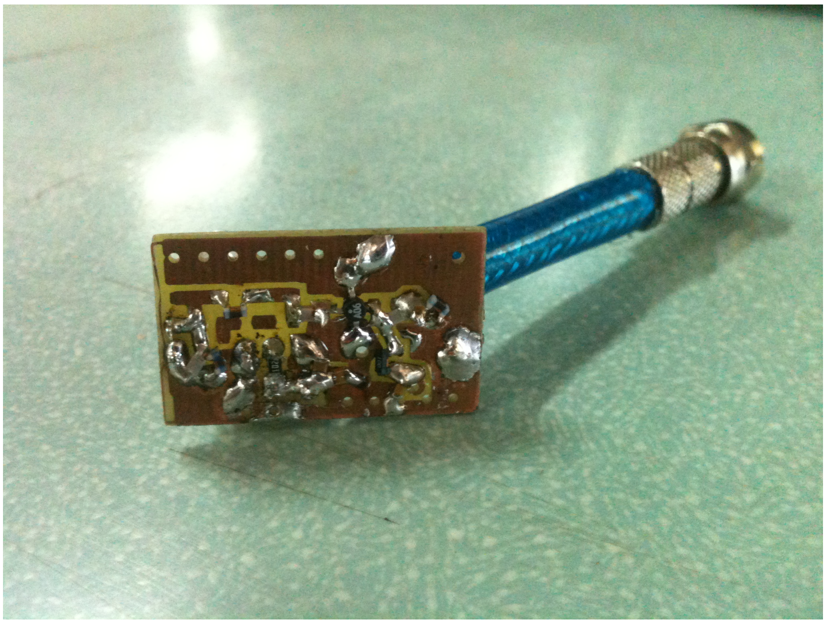


Figure 3.5: LNA Module final circuit

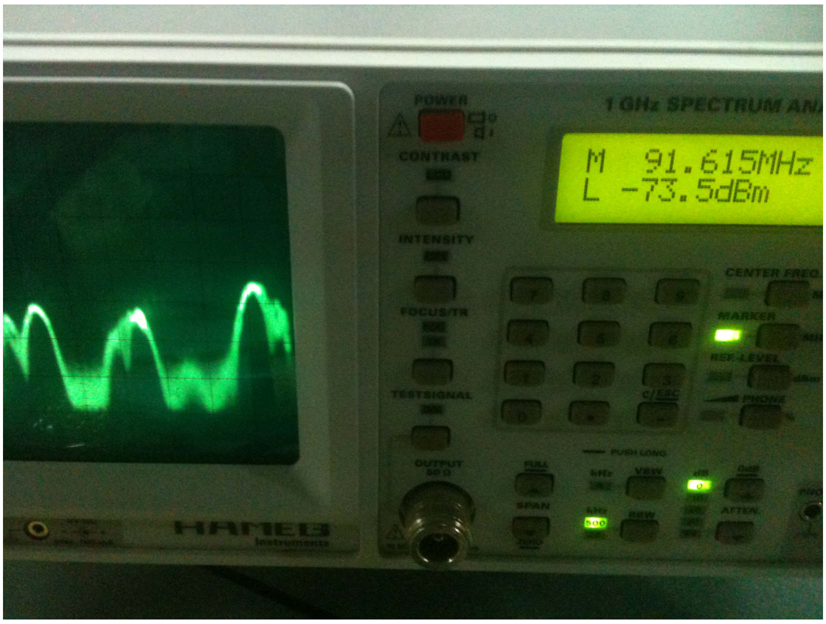


Figure 3.6: Spectrum and power of the received signal before amplification (for reference)

To conclude, the LNA module has been a success. It amplified only the band of interest and the filter attenuated all undesirable signals. Furthermore the matching network solved the mismatch between the module and the coax cable and the next stage (the spectrum analyzer in this case).

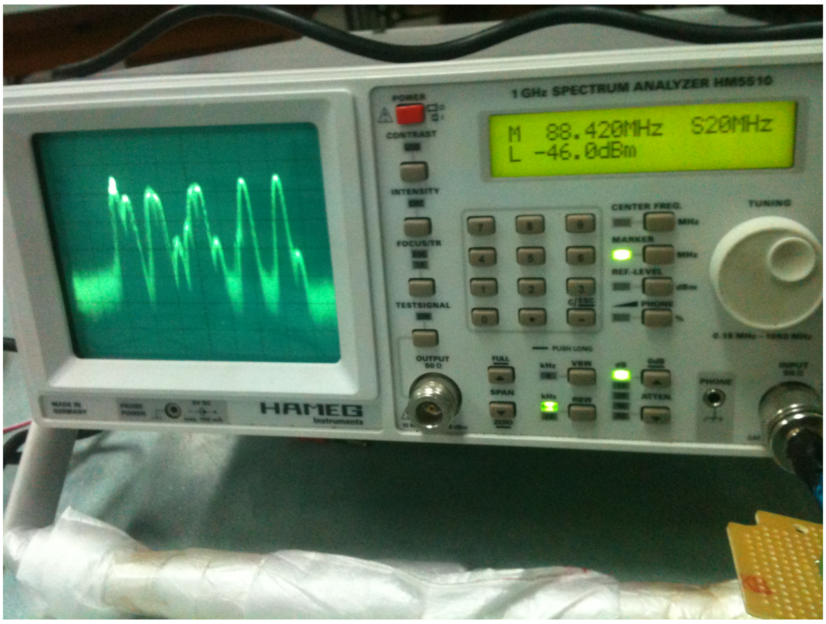


Figure 3.7: FM Band after amplification

3.4 Main Board

The mainboard contains all the remaining stages: a common local oscillator and four circuits representing the four channels of the receiver (mixer, IF filter, IF amplifier, etc.).

Unfortunately, the rest of the stages could not be assessed because of the extremely low quality of the printed board. This board was printed at a modest independent institution because there were no other options. After two attempts and extremely long waiting times, the board was deemed too “complex” to be printed correctly. The board has several issues:

- Joined and lumped traces
- Imprecise and missed holes
- Missed metalization of holes and therefore no VIA connectivity between the two layers
- Poor trace precision and routing

Nevertheless, we took time to do the arduous job of separating the joined traces by etching, connecting VIAs manually by adding wires that went through the board, rerouting certain parts of the circuit, etc. The results are shown in figures 3.8 and 3.9 where we soldered only one channel to test whether the circuit will produce any meaningful results but unfortunately it was not the case. Furthermore we were running out of time and we had to bet on a more promising alternative: the undersampling receiver.

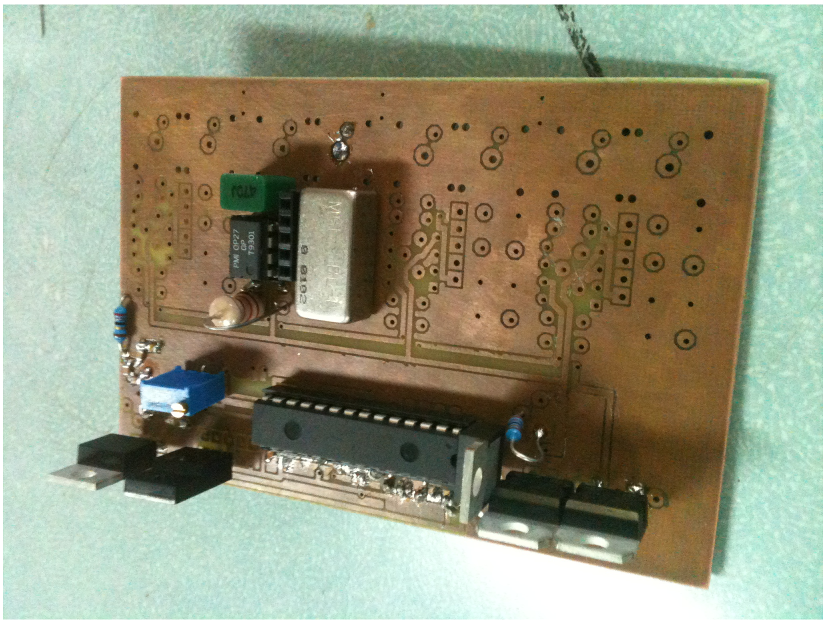


Figure 3.8: Mainboard final circuit - Top layer

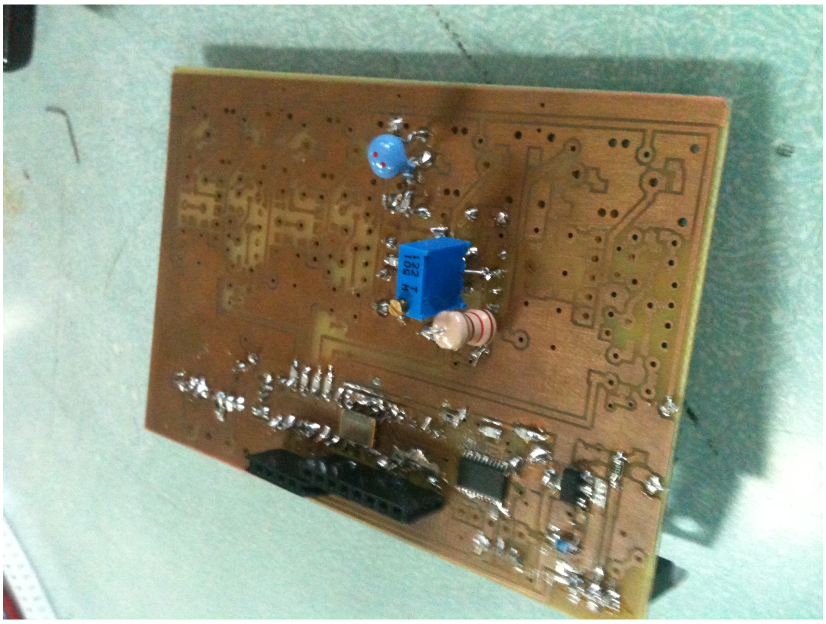


Figure 3.9: Mainboard final circuit - Bottom layer

We were only able to test the output of the clock generator for the PLL and its spectrum is shown in Figure 3.10. The fundamental frequency is 10 MHz as expected, and since the output is a square signal, we can see all the harmonics which are spaced at 10 MHz. Shown in this figure, from left to right: the -10 MHz component, the 60 Hz component (inherent in power supply), the fundamental 10 MHz, 20, 30, 40, ...

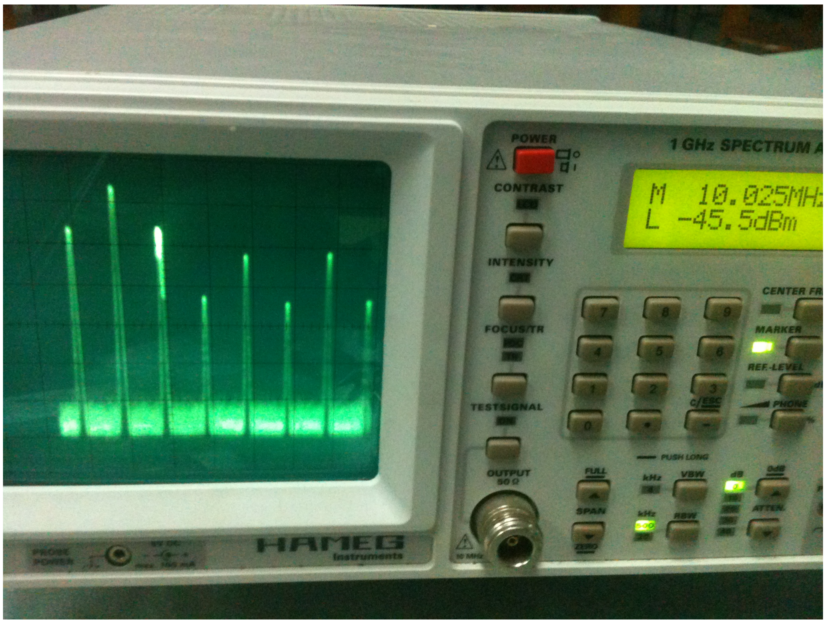


Figure 3.10: Spectrum of the clock signal generator

3.5 Undersampling Receiver

In this section we discuss the experimental results of an undersampling scheme in which we have used our previous LNA module to boost the signal level at the input of the NI PXI-5105 digitizer.

The LNA module is shown outside its feedline casing in Figure 3.11. This module will be connected to a perfboard containing a bias circuit to offset the signal to half its maximum range (because AC coupling is disabled on the digitizer). Then the perfboard is connected to the four channels of the digitizer as shown in Figure 3.12.

The advertised characteristics of the PXI-5105 is a 60 MHz analog bandwidth. But in order for our undersampling scheme to work, we should be able to pass signals at the FM band frequency (More than 88 MHz) thus a bandwidth of at least 100 MHz.

As it turns out, the exact frequency response of the digitizer shows (Figure 3.13) that it will allow the signal in the FM band to pass with a small attenuation. For example, the digitizer will allow a 90 Mhz signal to pass with a 3.5 dB attenuation which is very acceptable in our application as long as the signal power is still sensible by the ADC. Moreover, any attenuation can be equalized digitally after acquisition later on. It should be noted that in order for this to hold, the built in (24 MHz cutoff) anti-aliasing filter of the digitizer must be disabled.

After filtering the FM band and amplifying it, it goes practically unattenuated through the digitizer's sample and hold circuitry. According to the undersampling property we discussed in a previous section. The FM band will be replicated at

Figure 3.11: LNA module outside the shielded casing

Figure 3.12: Final perfboard and its connection to the digitizer

multiples of the sampling frequency which is 60 MHz. For example, the 88.4 Mhz channel will appear at 88.4 MHz $\pm$ 60 MHz = 28.4 MHz.

We use NI-Scope to visualize the spectrum of the digitized signals. Several peaks appear in the spectrum from 0 to 30 MHz. To show that these are indeed the peaks of the FM channel we used a commercial FM transmitter and we changed the frequency of emission as we visualize the received spectrum. One new peak appeared and shifted to the left or to the right when we increased or decreased the frequency of transmission. Which confirms that these were indeed the FM channels

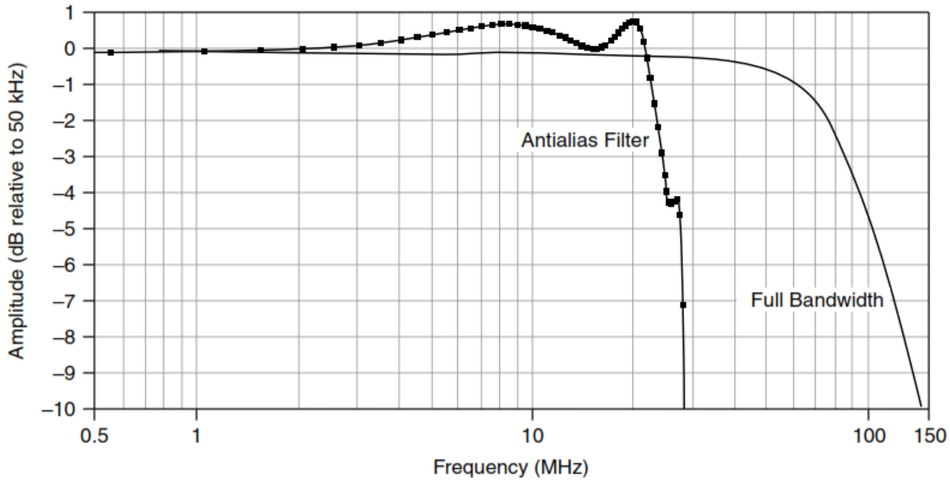


Figure 3.13: Frequency response of the NI PXI-5105

we were receiving.

Figure 3.14 shows clearly a single peak at 18 MHz which corresponded to 102 MHz on our transmitter. This peak has been shifted to -18 MHz (102 MHz - 2 * 60 MHz). This confirmed that the undersampling scheme is working. All what remains is to digitally filter the peak with a 200 kHz band-pass filter and downconvert it to baseband.

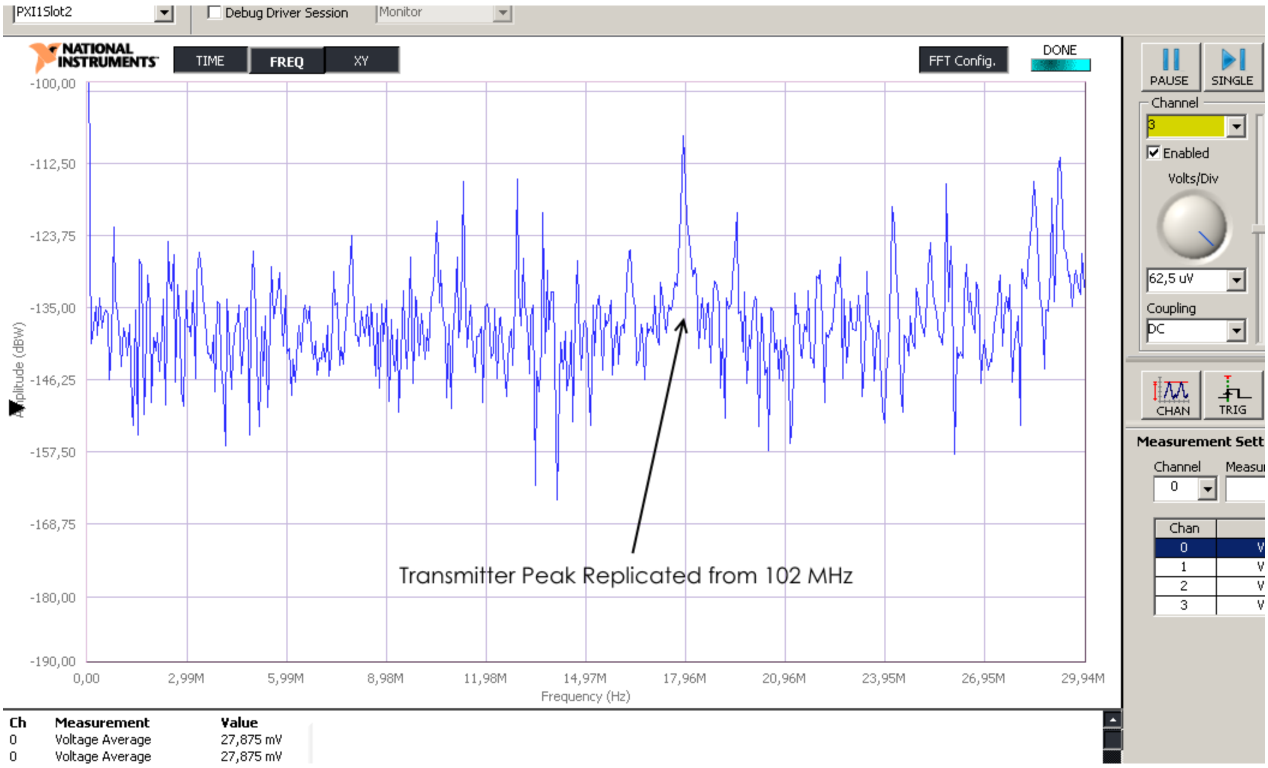


Figure 3.14: Digitized signal spectrum from 0 - 30 MHz

Figure 3.15 shows all four channels simultaneously. The channels are receiving the same signals (with minor phase difference) so the spectra are almost overlapping.

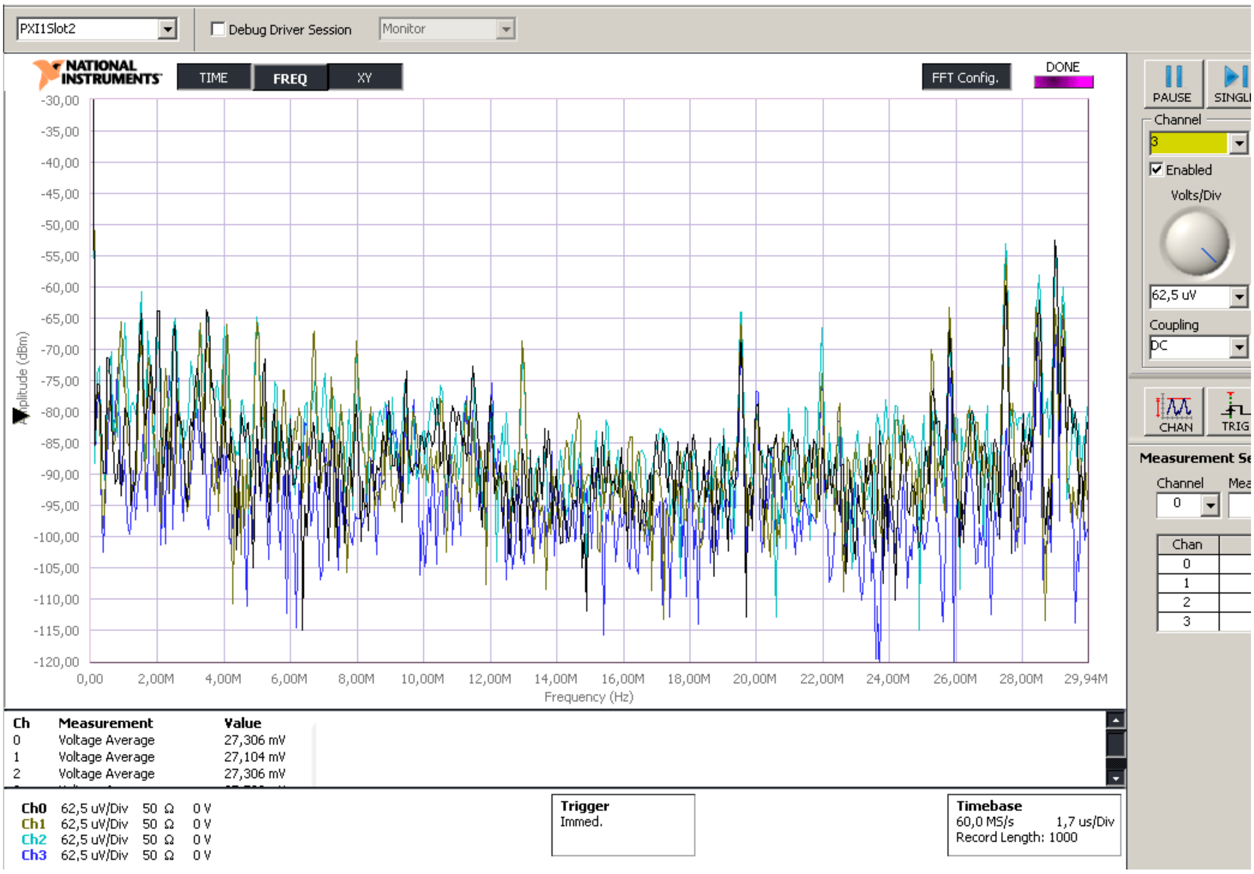


Figure 3.15: Spectra of the four digitized channels

In conclusion, the undersampling receiver works with minimalistic component count and it is clearly an advantage to carry out the rest of the processing digitally at virtually no cost.

Conclusion

This project has been a very brief introduction to the domain of RF design. A field of numerous and colorful applications and of an immense importance in telecommunications engineering.

The design of this multichannel RF receiver has been a very iterative process. We learned that it is often the case with systems engineering where there is a compromise and a trade-off to be made at almost every single stage and every single component.

This type of receivers is very demanding in terms of components and hardware and it is usually the field of very specific and specialized companies. Nevertheless, with extremely limited resources and little knowledge in the field, we have been able to implement a prototype with results that could be far superior with more adequate hardware. The final product is thus far from being final.

First, we designed an LNA module that goes in the feedline of a monopole antenna which we have also made ourselves. This module has been a success overall. We have had some problems with finding adequate cables and connectors and we have made adequate matching to avoid losses due to the mismatch.

Secondly, we designed a main board that contains all four signal paths. The circuit has been printed in an outside institution and the quality of the circuit was not promising. The board contained a mixture of analog and digital circuitry. For example we had a digitally controlled PLL and a microcontroller to make the acquisition. We have also made an alternative signal path to use a professional digitizer. Unfortunately, this part was not fruitful due to the poor printed board quality.

The results were more or less satisfying and we validated all of our work using a network analyzer where we visualized the spectrum of the signal at several stages.

We then shifter our focus to make try an undersampling scheme which was successful. Nevertheless, additional work is required to actually demodulate the signal or to perform phase splitting. The advantage here is that everything is digital which is fairly easy.

The project overall has been a successful experience in which we have been able to discover the extremely difficult problems that the system engineer has to deal with and the order of complexity of practical designs in this area.

We have also realized that practice, especially in this field, is extremely different from theory and perfect block diagrams. We have also learned that a very strong background and a great deal of practical experience is needed to work on such projects and we are glad that we have had experts on our side to help us.

Finally, we hope that this work would be a head-start for coming projects in this area and a brief introduction to RF design and the main challenges to be tackled.

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