Passive Weather Radar Theory

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Chapter 1

Project Purpose and Scope

The primary objectives of this project is to build and instrument using GNURadio and an Ettus LLC USRP in order to evaluate the former and gain experience on both. A Passive Weather Radar has been chosen as the vehicle to these ends because it is expansive enough to exercise GNURadio and it has some chance of success.

The literature suggests that frequencies below L band reflect well from precipitation. With the introduction of Telstra's "Next G" cell phone net work a number of wide band signals from moderately powerful transmitters using 800 MHz to 900 MHz is available. These signals have the desirable characteristic of mimicking bandwidth limited white noise.

There is also some suggestion that air current eddies may be detectable at about 900 MHz using Doppler signals.

So the secondary objectives are to investigate the possibility of using the Telstra signals for examining these phenomenon with no particular expectation of success or failure.

Chapter 2

Introduction

2.1 Definition

Passive Radar is a system that has many of the same functionality as an active radar system but achieves its goals without transmitting any radio frequency energy. As a substitute for the energy transmitted by an active radar a passive radar relies upon remote transmitters illuminating the area of interest.

Weather radar relies on Rayleigh scattering of radio frequency energy by water droplets. Existing active weather radars are typically pulsed Doppler radars using a wavelength of 10cm (S band).

2.2 Basic Active Radar Theory

Active radar works by transmitting a short pulse through a directional antennae and receiving any subsequent reflections via the same antenna. The antenna is then rotated to point in a new direction and the process repeated. Figure 2.1 shows a typical transmission and reception of an active radar.

The range aperture in the radial axis is a function of the transmit pulse width and the aperture in the circumferential axis is a function of the beam width formed by the directional antenna. The transmit function can be considered an approximation to a Dirac function $\delta(t)$ in time and the closer the approximation to the ideal the better the range resolution. However, resolution of Doppler values requires a wider transmit pulse so there is a design trade off to be made here.

In practice the antenna is made to turn at a constant rate and the transmitter is made to transmit at a constant repetition rate. The returned radio frequency energy is converted to baseband where signal processing occurs and is then presented to a PPI display for viewing or an A/D converter for further processing (in which case the analog signal processing is, typically, minimal).

In addition to signals of interest the return is typically contaminated with unwanted artifacts such as ground clutter. A large amount of this noise may be removed by using Doppler processing in order to remove those echoes having a

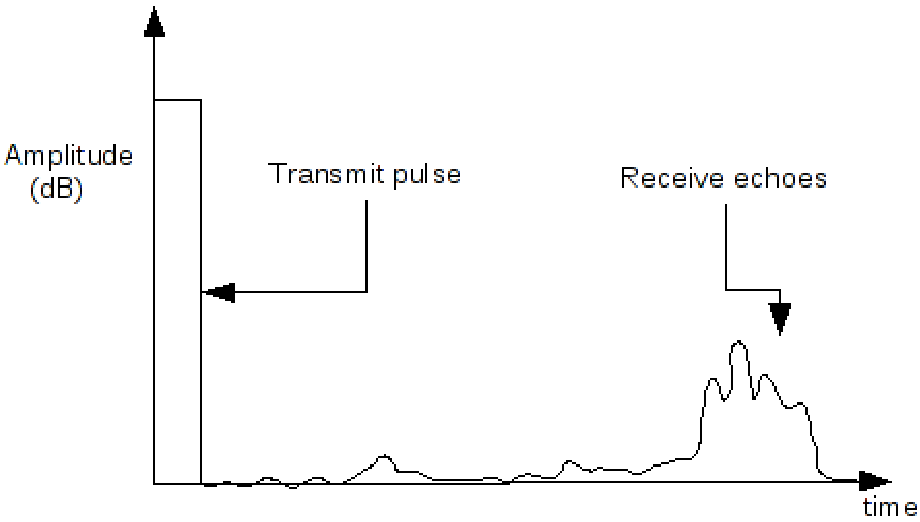

Figure 2.1: Time series of an active radar signal

Figure 2.2: Marine radar antenna consisting of a rotating slotted wave guide

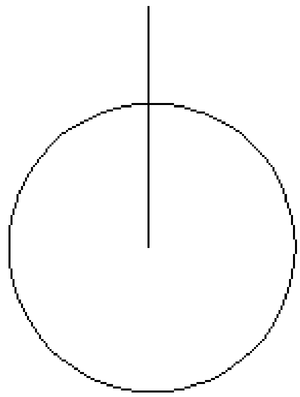

Figure 2.3: Geometry of an active radar system in the horizontal plane

velocity less than some minimum value. In addition the velocity of the returns has significant value in the context of a weather radar. Of course only velocities along the radial axis can be observed.

The geometry of an active radar system when the transmitter and the receiver antennas are co located (as they invariably are) is that of radials intersecting co located circles.

2.3 Basic Passive Radar Theory

Because passive radar relies upon transmitters of opportunity the receiver and signal processing must be designed to suit the available illumination. However the transmitter and receiver are very unlikely to be co located so the geometry is possibly not going to be the same as the in the case of an active radar. In general a passive radar will be made to receive the transmission signal and the delayed reflected signals in order to identify targets of interest. As in the case of the active radar, and knowing the frequency and pseudo phase of the transmitter, Doppler processing may be employed. In general geometrical and receiver dynamic range considerations will require the more than one transmitter be used.

There is an analog between active and passive radars with respect to the range and Doppler resolutions including the trade off between the precision of the two.

There are two realizable implementations of a passive radar. One involves a directional beam pattern for the receiver which can be made to point in a certain direction. Each transmitter forms an elliptical isovalue with the transmitter and receiver as foci. Another measures the difference in time (and, hence, distance) between the direct arrival of the transmitter signal and the reflected signal in order to form an ellipse with foci at the transmitter and receiver upon which the target must lie. In this case the available degrees of freedom is equal to the number of transmitters.

Chapter 3

Design Decisions

3.1 Decisions on the Measurement Method

The most basic decisions to be made are those concerning the measurement method. These decisions will drive other decisions such as the number of receivers required and the processing chains.

A key decision is the illumination source. It is proposed to use the Telstra Next G "Upper 800 MHz Band" signals. These signals emanate from Telstra's cell phone towers. They are Direct Spread Spectrum signal of about 10 MHz wide in the 880 MHz to 890 MHz band. This bandwidth will yield a resolution in the order of 30 meters. It may be possible to supplement illumination from this band with illumination from Telstra's "Lower 800 MHz Band" signals. These occupy the band from 810 MHz to 820 MHz.

Another key decision is to answer the question, "should the direction from the receiver to the target

be measured and, if so, how"? Measuring the angle imposes difficulties but yields an extra degree of freedom so that it will be possible to get target estimates from just one transmitter. However, it may be necessary to drop this requirement and fall back on a time difference or pseudo time zero scheme.

3.1.1 Direction Measuring

3.1.1.1 Mechanical

Measuring the direction by mechanical means involves building a parabolic dish antenna mounted on a turntable. This is difficult to implement, complex, possibly unreliable and potentially dangerous to operate. This solution is rejected out of hand on these grounds.

3.1.1.2 Electronic

There are at least two electronic methods to estimate direction which may be employed

3.1.1.3 Phased Array

A phased array consists of a matrix of regularly spaced simple antennae on a plane surface. Each antenna is connected to its own receiver. The advantage of a phased array is a super abundance of data and, if the matrix is more than one row, the ability to measure elevations and direction in the horizontal plane. This disadvantage is that it requires a lot of receivers and is expensive. This solution is rejected because of the cost.

3.1.1.4 Phase Interferometry

If two simple antennae are place on a plane separated by a distance with a wave front incident at a certain angle then the path difference between the target is a function of that angle and is represented by a phase difference between the antennae. For unambiguous results the distance should be less than, or equal to, one half the wavelength of the incident signal. However, two antennae located so close together will almost certainly mutually couple and distort the phase measurement of each. A possible solution to this problem is to measure the phase difference not of the 889 Mhz carrier frequency but the 10 MHz chip frequency used to directly spread the carrier. In this case the half wavelength spacing is about 15 meters.

Elevation data can be measured in a similar manner by placing a third antenna in the vertical plane. As with the Phased Array approach Phased Interferometry requires one receiver per antenna.

The supper abundance of data that exists in the Phased Array does not exist in this arrangement. Phase Interferometry is sensitive to poor signal to noise.

Phase Interferometry provides angle information at a relatively cheap hardware cost.

3.2 Omni-Directional Measuring

For this solution a minimum of one antenna and receiver are required. The tracking geometry is an ellipse intersected by a line per transmitter. A minimum of two transmitters are required for an unambiguous position determination.

Elevation determination will probably not be realistically achievable unless a Phase Interferometry is implemented in the vertical plane. In addition, due to the non regular geometry the 30 meter resolution from a 10 MHz bandwidth will not be achievable making higher bandwidth more important.

This solution is one of the the cheapest of solutions in terms of hardware.

Chapter 4

Signal Processing

4.1 Major Signal Processing Blocks

4.1.0.5 Receiver and Baseband Processing

This block is wholly implemented in the USRP using the FPGA and analog hardware. As can be seen from figure 4.2 this block is a conventional single conversion receiver. The real and imaginary values produced by the mixer are digitized by the analog to digital converter in order to preserve the phase information.

4.1.1 Phase Interferometer Processing

This block develops a stream of $\begin{bmatrix} S & \Theta \end{bmatrix}$.

4.1.2 Adaptive Filtering

The purpose of the Adaptive Filtering process is to remove the transmitter signal from the input. It uses an input from the Reference Tracking block as a reference of the signal to be removed.

As can be seen from figure 4.3 the filter is a standard FIR type. The Update Algorithm is made to update the tap coefficients in order to filter out the transmitter signal.

4.1.3 Reference Tracking

The purpose of the Reference Tracker is to track all transmitters of interest and, possibly, reconstruct transmitter signals if necessary. Transmitters need to be tracked in order to develop time zero signals and to present references to the Adaptive Filter and the Cross Correlation blocks.

4.1.4 Cross Correlation

The purpose of this block is to perform conventional cross correlations between transmitter reference signals and the filtered received echoes. The cross corre-

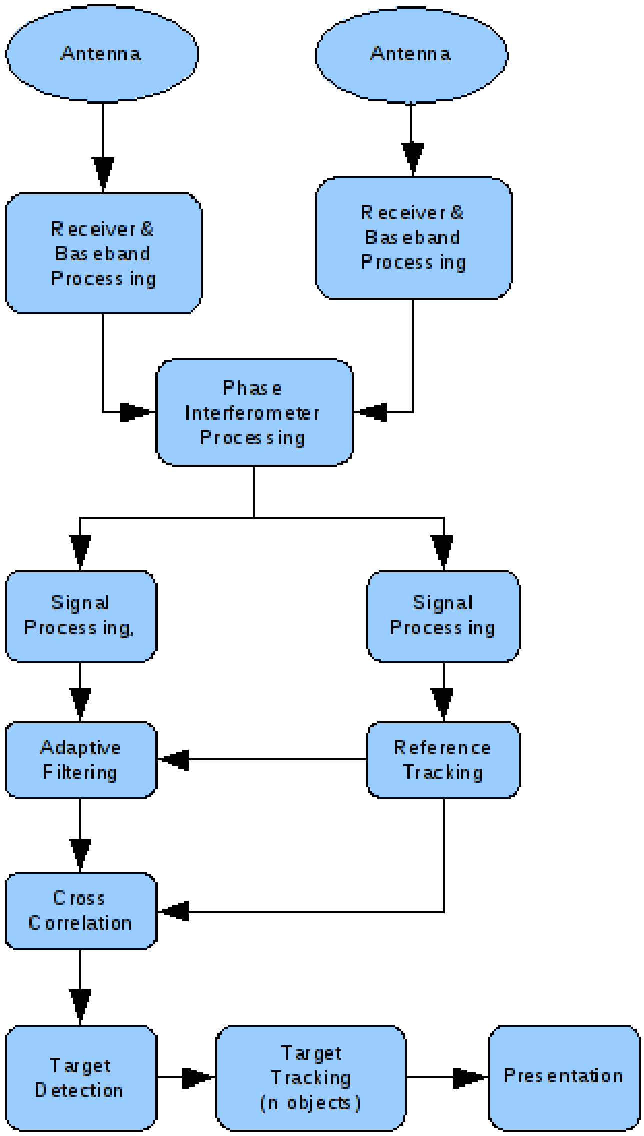

Figure 4.1: Signal Processing Block Diagram

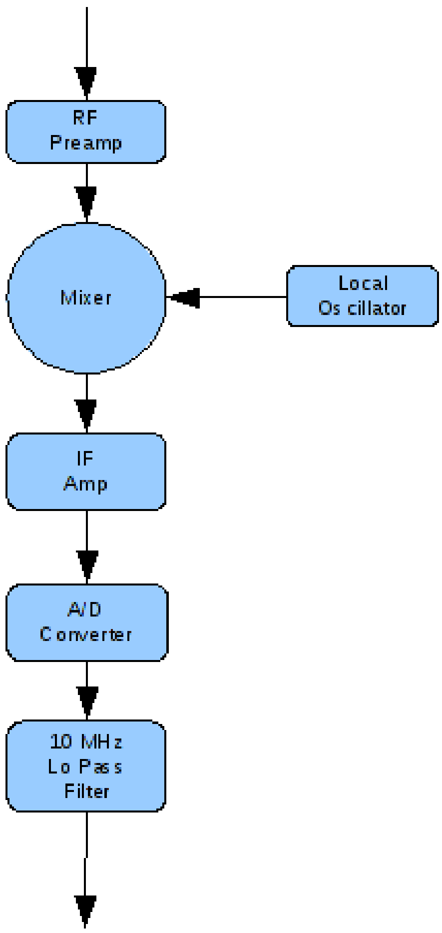

Figure 4.2: Receiver and Baseband Processing

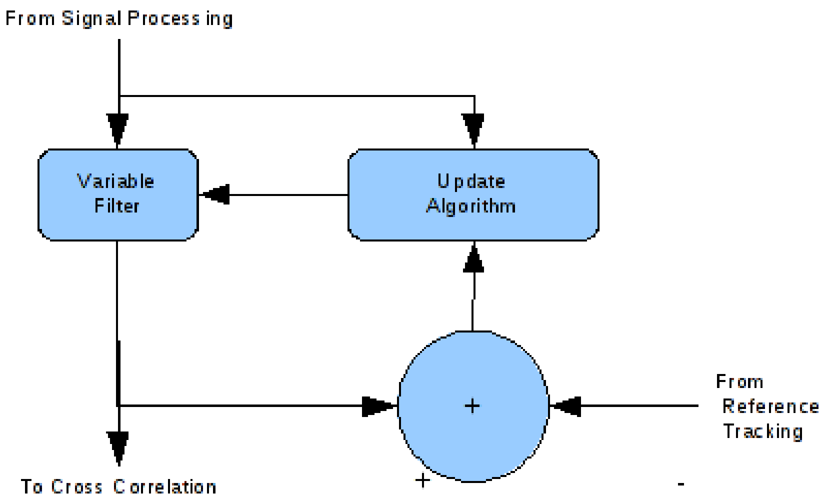

Figure 4.3: Adaptive Filtering Block

lator acts as a matched filter and will provide processing gain. The intention is to aim for 50 dB of processing gain in this procedure.

Estimates of bistatic range and Doppler velocity will be developed in this block.

4.1.5 Target Selection

The purpose of this block is to plot the received correlated echoes using the appropriate geometry onto a regular grid and to use the accumulated result to detect targets. Target detection will use the Constant False Alarm Rate (CFAR) algorithm. A user probability setting will be implemented.

4.1.6 Target Tracking

State information for targets is developed by the Target Tracking block. Each target will be represented by an object which implements a Kalman Filter.

Chapter 5

Implementation

In order to detect valid signals from the clutter it is important to maximize the processing gain and SNR. This chapter investigates the Matched Filter which is the block which will achieve this.

In this paper range refers to the sum of the distances from the transmitter to the target and from the target to the receiver and velocity refers to the rate of change to the range.

5.1 Matched Filter Receivers

The output of the Matched Filter $M(\tau, \phi)$ is:

$$M(\tau, \phi) = \frac{1}{T} \int_0^T x_r(t) x_t^*(t + \tau) e^{-i\phi t} dt \tag{5.1}$$

where
T = the radar integration time, a constant, in seconds
$x_r(t)$ = the received signal
$x_t^*(t + \tau)$ = the complex conjugate transmitted signal
τ = the delay induced by the sum of the ranges from the transmitter to the receiver via the target
ϕ = the Doppler frequency induced by the sum of the velocities from the transmitter to the receiver via the target.

In practice a matched filter is usually implemented digitally in which case $x_r(t)$ and $x_t(t)$ are converted into digital signals $x_r[n]$ and $x_t[n]$ where $x[n] = x(nT_s)$ and T_s is the constant sample period. In discrete form the output is:

$$M(\tau, \phi) = \frac{1}{N} \sum_{n=1}^{N} x_r[n] x_t^*[n + \tau] e^{-i\phi n} \tag{5.2}$$

where
N = the number of samples in the integration time ($T_s N = T$)

It is evident from equations 5.1 and 5.2 that the expression for a matched filter is the Fourier Transform in ϕ of the correlation between the received signal and the transmitted signal delayed by time τ.

If the noise corrupting the received signal is additive white Gaussian the following strategy will give the optimal target detection:

- From the received signal $x_r(t)$, and a copy of the transmitted signal, $x_t(t)$, calculate $M(\tau, \phi)$.
- Calculate the magnitude, $|M(\tau, \phi)|^2$.
- Compare this quantity to some threshold, l. A target is indicated by $|M(\tau, \phi)|^2 > l$ at time delay τ and Doppler shift ϕ.

5.2 The Ambiguity Function

Passive Radars are reliant on transmissions of opportunity and the maximum information that may be extracted from a signal from a target is encapsulated in these transmissions. It is useful to characterize the available transmissions. This is the purpose of the ambiguity function $\chi(\tau, \phi)$.

$$\chi(\tau, \phi) = \frac{1}{T}\int_0^T x(t)x^*(t+\tau)e^{-i\phi t}dt \tag{5.3}$$

As can be seen from equation 5.3 the ambiguity function describes how correlated a transmitter signal is with itself delayed by τ and shifted by Doppler ϕ. It is also known as the time frequency autocorrelation function.

The discrete form of equation 5.3 is:

$$\chi(\tau, \phi) = \frac{1}{N}\sum_{n=1}^{N} x[n]x^*[n+\tau]e^{-i\phi n} \tag{5.4}$$

It has been recognized that the ambiguity function of a transmission signal should have certain properties in order to be of use. These are due to factors such as the transmitted signal processing positive finite energy and that the signal cannot occupy a point in time and frequency. The most important of these properties are:

- The maximum magnitude of the ambiguity function is a maximum at the origin. That is,

$$|\chi(0,0)| \geq |\chi(\tau, \phi)| \tag{5.5}$$

- The magnitude of the ambiguity function is symmetric about the line $\tau = \phi$. That is,

$$|\chi(\tau, \phi)| = |\chi(-\tau, -\phi)| \tag{5.6}$$

- $|\chi(0,0)|^2|$ is equal to the the amount of energy of the transmitted signal in the integration time. That is,

$$|\chi(0,0)|^2 = \int_0^T |x(t)|^2 dt \tag{5.7}$$

- The volume under the surface described by $|\chi(\tau,\phi)|^2$ is equal to the energy in the transmitted signal. That is,

$$\int_0^T \int_0^T |\chi(\tau,\phi)|^2 d\tau d\phi = \left[\int_0^T |x(t)|^2 dt \right]^2 \tag{5.8}$$

These necessary properties limit the possible ambiguity functions and, so, the possible transmitter signals. Any transmitter signal available is certainly going to be a compromise between the above properties.

5.3 Resolution

In order to optimally detect a target which produces time delay, τ, and Doppler shift, ϕ, it is necessary to maximize the value of $|\chi(0,0)|^2$. Also, in order to minimize the detection of false targets it is necessary to minimize the value $|\chi(\tau,\phi)|^2$ for $(\tau,\phi) \neq (0,0)$. Therefore the ideal transmitter signal is one that yields the ambiguity function so that:

$$|\chi_{ideal}(\tau,\phi)|^2 = \begin{cases} 1 & \text{if } (\tau,\phi) = (0,0) \\ 0 & \text{if } (\tau,\phi) \neq (0,0) \end{cases} \tag{5.9}$$

This is the Dirac function in τ and ϕ, $\delta(\tau,\phi)$, and is not realizable in practice (and is not a function).

More significantly if the transmitter signal approximates bandwidth limited white noise as DSS does then when N is large (in normalized form):

$$|\chi(\tau,\phi)|^2 = \begin{cases} 1 & \text{if } (\tau,\phi) = (0,0) \\ \frac{1}{TB} & \text{if } (\tau,\phi) \neq (0,0) \end{cases} \tag{5.10}$$

where
T = the radar integration time in seconds
B = the signal bandwidth in Hz

Equation 5.10 resembles an autocorrelation function squared. That is the best realizable ambiguity function has a spike at the origin and a surrounding floor. It follows that any signal having an amplitude less than $1/TB$ of the strongest signal cannot be detected. This relationship has significance when receiving signal having a wide range of energy amplitudes such as the direct signal from the transmitter and the targets. For instance with an integration time of 1 second and a bandwidth of 10 MHz the Matched Filter has a dynamic range of $|\chi(\tau,\phi)|^2 = 10\log\left(\frac{1}{1*10^6}\right) = -70dB$ with respect to the strongest signal.

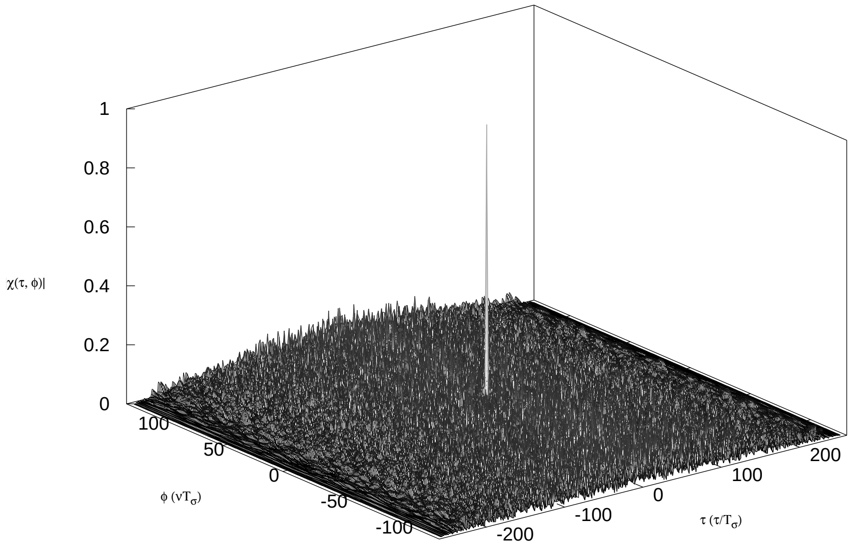

Figure 5.1: Ambiguity function of white Gaussian noise

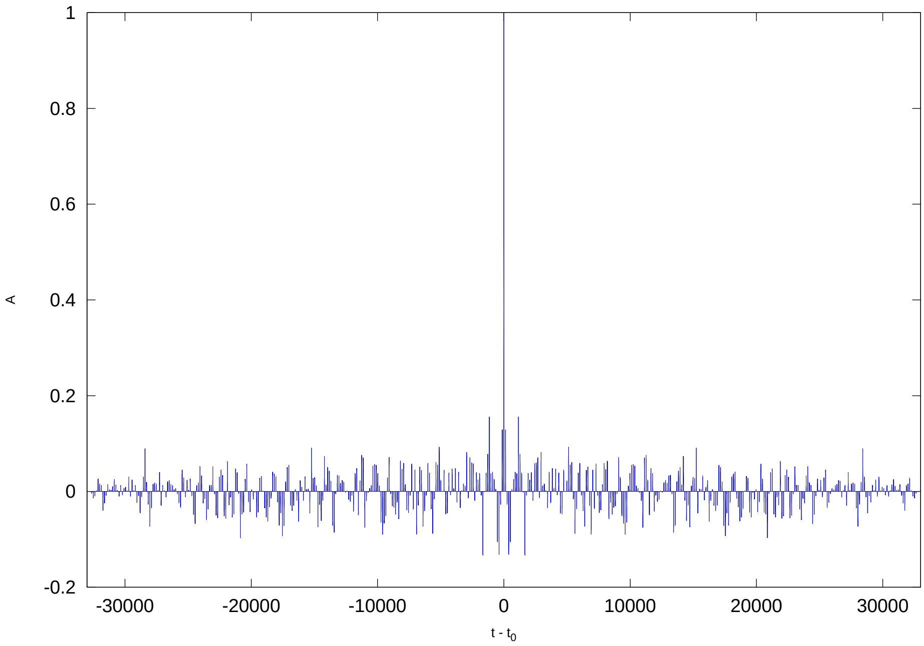

Figure 5.2: white Gaussian autocorrelation

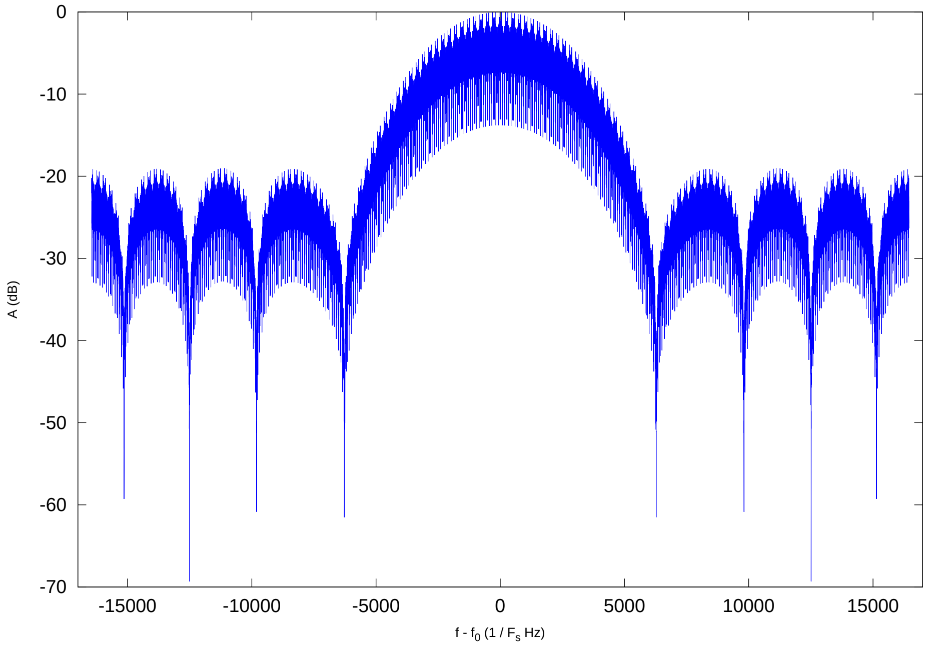

Figure 5.3: white Gaussian noise envelope

Figure 5.1 shows a typical ambiguity function of a white Gaussian noise source. Notice the narrow spike at the origin and the surrounding noise floor. The amplitude ratio between the spike and the floor is improved by increasing the time bandwidth product. Figures 5.2 and 5.3 show the autocorrelation function and power or energy spectrum for the white noise source associated with 5.1. Note the good autocorrelation characteristic and the $sinc^2()$ shape of the power function, both of which are typical of a source of this type.

Referring to figure 4.1 it can be seen that the input to the Cross Correlation Block is from the Adaptive Filtering Block. It is proposed that this latter block in conjunction with the Reference Tracking Block will attenuate the transmitter signal with respect to the echoes and so increase the dynamic range of the Cross Correlation Block. Further it will probably be necessary to shield the antennae in the direction of the transmitter in order to attenuate this signal even further although the advantage of this measure is to maximize the dynamic range of the analog portion of the receiver in order not to overload it.

5.4 Implementation Problems

Typically Passive Radar processing is done digitally and, in fact, cannot, now days, be done practically done using analog techniques. The digital nature of the processing imposes its own set of problems.

Digital implementation of the Matched Filter is given by equation 5.2. The signals $x_r[n]$ and $x_t[n]$ are received and sampled at discrete points in time nT_s

where T_s is the constant sampling period. Therefore τ is only defined at periods T_s seconds. Likewise ϕ is only defined at multiples of $1/(NT_s)$ Hz. These two factors limit the resolutions of τ and ϕ and these limits are a function of T_s and N. This corresponds to resolutions of cT_s meters and $c/(NT_s f_c)$ Hz where c is the speed of light and f_c is the center frequency of the transmission.

Increasing the sampling frequency, F_s, and NT_s decrease the granularity of the discrete representation of $M(\tau, \phi)$ in the respective domains. However, it is pointless to decrease these granularities such that the peak at the origin occupies a large number of (τ, ϕ) cells. In this way T_s and N are tied to the function of merit, the ambiguity function $\chi(\tau, \phi)$, of the transmitter's signal.

5.5 Radar Equation

The power P_τ returning to the receiving antenna is given by the radar equation:

$$P_\tau = \frac{P_t G_t A_\tau \sigma F^4}{(4\pi^2)\, R_t^2 R_\tau^2} \tag{5.11}$$

where
P_t= transmitter power
G_t= gain of the transmitting antenna
A_τ= effective aperture of the receiving antenna
σ= radar cross section, or scattering coefficient, of the target
F= pattern propagation factor
R_t= distance between the transmitter and the target
R_τ= distance between the target and the receiver

For atmospheric targets the pattern propagation factor will consist mainly of the path loss. The path loss in rural areas L_O(in dB) is given by the Hata Model for Open Areas:

$$L_O = L_U - 4.78\,(\log(f))^2 + 18.33\log(f) - 40.94 \tag{5.12}$$

where
L_U= path loss in urban areas in dB
f= frequency in MHz

In turn the path loss in urban areas L_U is given by the Hata Model for Urban Areas:

$$L_U = 69.55 + 26.16\log(f) - 13.82\log(h_t) - C_H + (44.9 - 6.55\log(h_t))\log(R_t + R_\tau) \tag{5.13}$$

where the antenna height correction factor C_H is given by:

$$C_H = 0.8 + (1.1\log(f) - 0.7)\,h_\tau - 1.56\log(f) \tag{5.14}$$

where
f= transmission frequency in MHz
h_t= height of transmitter antenna

h_τ= height of receiver antenna
R_t= distance between the transmitter and the target
R_τ= distance between the target and the receiver

A simpler approximation for the path loss, L, between two isotropic antennae is:

$$L = 20log\left(\frac{4\pi\left(R_t + R_\tau\right)}{\lambda}\right) \tag{5.15}$$

where
R_t= distance between the transmitter and the target in meters
R_τ= distance between the target and the receiver in meters
λ= wavelength of the transmission in meters

5.6 Hardware

5.6.1 Universal Software Radio Peripheral (USRP)

The software radio hardware will be the Ettus Research USRP (also known as the USRP1) with two DBSRX2 receiver daughter modules each connected to a VERT900 whip antenna.

The connection from the URSP to the host computer is effected using a USB2 link.

5.6.2 USB Interface Bandwidth

Due to the limited bandwidth of the USB 2.0 data channel and the relatively small capacity of the on board USRP data buffers, there is a limit to the number of receive samples that can be transfered from the USRP to the host computer per unit time. In fact the baseband bandwidth is limited by the USB data rate.

An effective baseband bandwidth may be synthesized by making a series of observations closely spaced in time at various frequency offsets.

The achievable data transfer rate on a USB 2.0 link is 53 MB/s. Each sample consists of an 16 bit I and a 16 bit Q value. Therefore the USB has a capacity of approximately 13 MS/s, or a baseband bandwidth of 6.5 MHz.

5.6.3 USRP FPGA

The USRP uses an Altera *Cyclone EP1C12Q240C8* FPGA to control the various operations. The relevant characteristics of this device are given in table 5.1. As can be seen this device has very limited logic capabilities and memory. In addition the lack of multipliers severely limit the use of this device for signal processing.

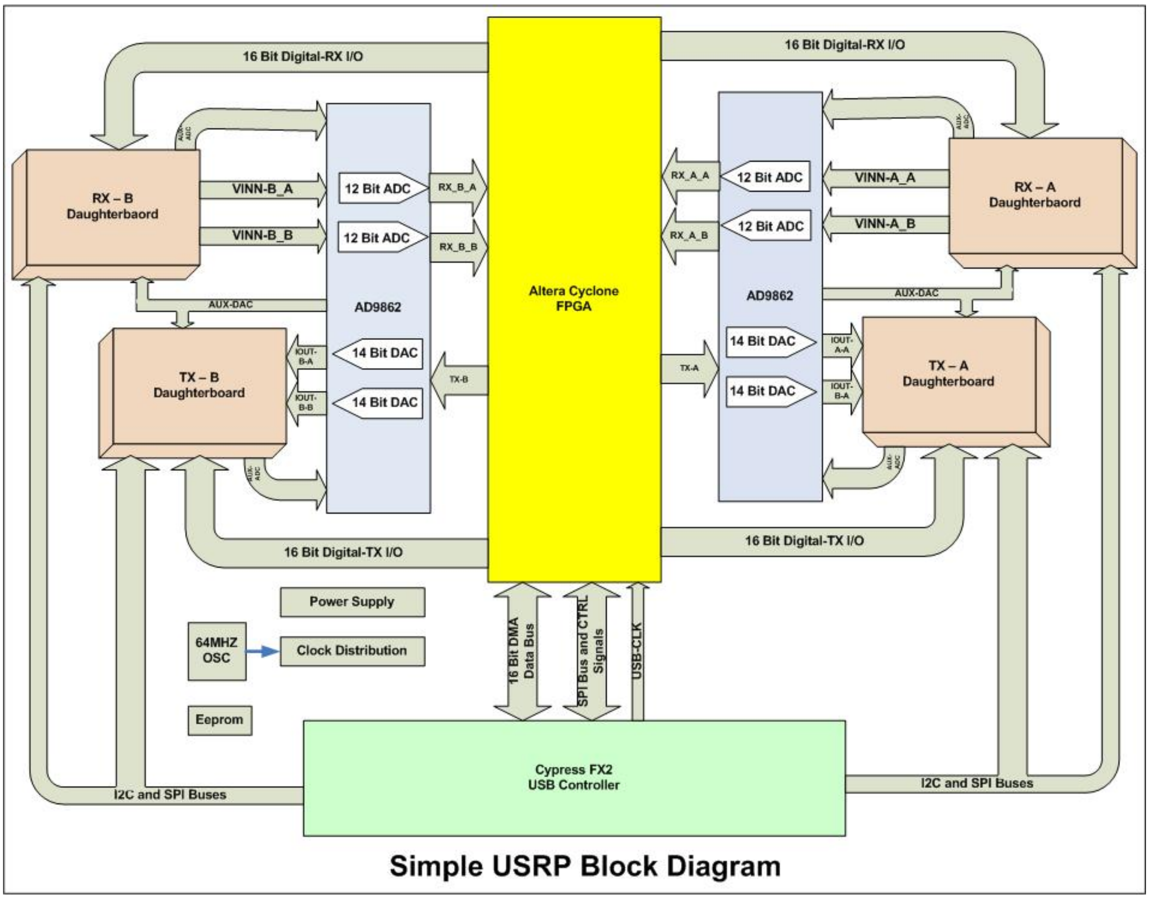

Figure 5.4: Simplified USRP Block Diagram

Logic elements	12060
Memory bits	239616
Phase Locked Loops	2
Multipliers	0

Table 5.1: Significant FPGA characteristics

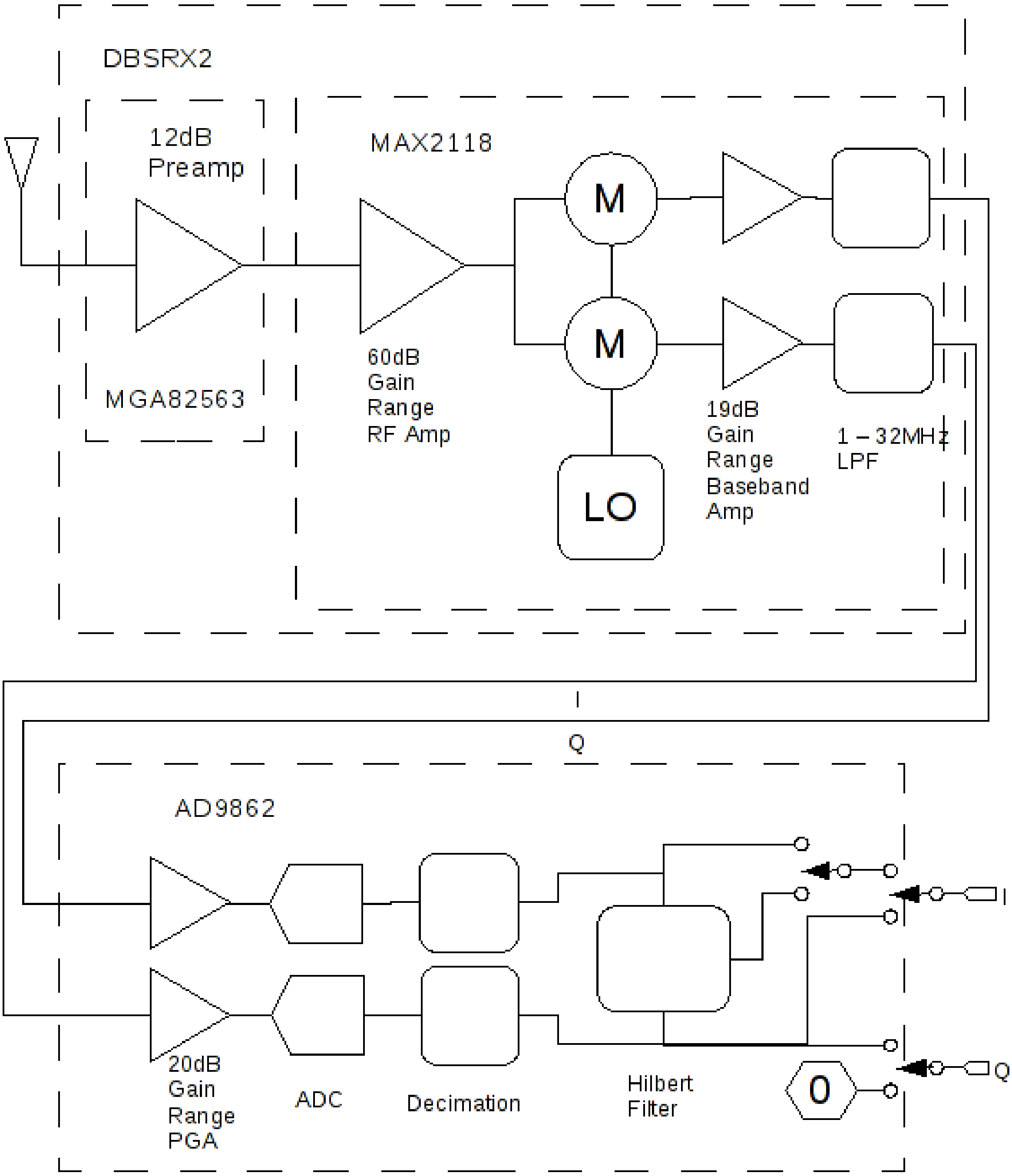

Figure 5.5: Simplified Signal Path Elements of One Receive Channel

5.7 Design Constrains

5.7.1 USB Channel Bandwidth

As noted in 5.6.2 the USB data bandwidth is about 13 MS/s which limits the baseband width to about 6.5 MHz. This encourages the design to move as much processing to the USRP FPGA in order to reduce the demands on the USB channel. If this proves impossible it may be necessary to synthesize a baseband bandwidth by making observations closely spaced in time with different local oscillator values.

5.7.2 Clock Stability

The USRP master clock provides the reference for all local oscillators in the reception chain. Therefore variations in the frequency of this clock have a potential impact on the baseband signal, especially any Doppler processing of that signal. The specified instability of the master clock is 20 ppm. Therefore the master clock may vary up to 1280 Hz from its nominal 64 MHz.

In this application it is not the absolute clock stability that is so important but its variation with respect to the transmitter carrier frequency. This fact suggests a possible solution to master clock stability, namely using the received direct transmission as a frequency reference in order to correct the distortions in the baseband signal.

5.7.3 FPGA

As noted in section 5.6.3 the FPGA has limited capability but it is desirable to move as much of the front end processing that might normally be done on the host computer into the USRP as possible with a view to reducing USB bandwidth. As a matter of priority a surgery will be made of the Verilog code to see which of it may be discarded in this application and whether discarding any code will allow processing to be moved to the FPGA.

Chapter 6

Design in the Enviroment

6.1 Transmitters of Opurtunity

6.1.1 Characteristics of Telstra Next G Transmitters

It is proposed to use transmitters that are components of Telstra's NextG" cell phone network. This network is widely deployed in WA. Typically there is one transmitter in each town which services the surrounding district. In addition transmitters are often placed in non urban situations as in fill. Also, in urban areas, there are a number of "segmented" transmitters usually driving sector antennae that are tilted in the vertical plane towards the ground.

At this time not a lot is known about how the transmitters are modulated. However, from the license assignments the bandwidth is 9.4 MHz, or less, and the emission designator is 9M40W7WEC. The first four alpha-numeric characters of the emission designator signify the bandwidth, in this case 9.40 MHz. The next three alpha-numeric characters signify the modulation, baseband and data types, in this case unspecified modulation, multichannel digital and unspecified data content.

It is expected that the transport access methods is CDMA using DSSS. In this case it is reasonable to assume that the transmitter signal will be modulated by some maximum length sequence, similar to those generated by linear feedback shift register generators, so that the signal is a Pseudo Random Noise (PRN) stream. In that case the chip rate is probably about 9 MHz.

In rural areas Telstra seems to operate their transmitters on a center frequency of 884.8 MHz (lower 880.1 MHz and upper 889.5 MHz). No doubt each station uses a different spreading code so that cell phones may discriminate between the several stations that may be within RF range.

The details of the spreading code use to produce the PRN stream is not known but, for this application, it is no necessary to have this knowledge. It is necessary that we faithfully receive each transmitter so that it can be used as a reference for correlating received echoes. Referring to figure 4.1 this is the main function of the Reference Tracking block.

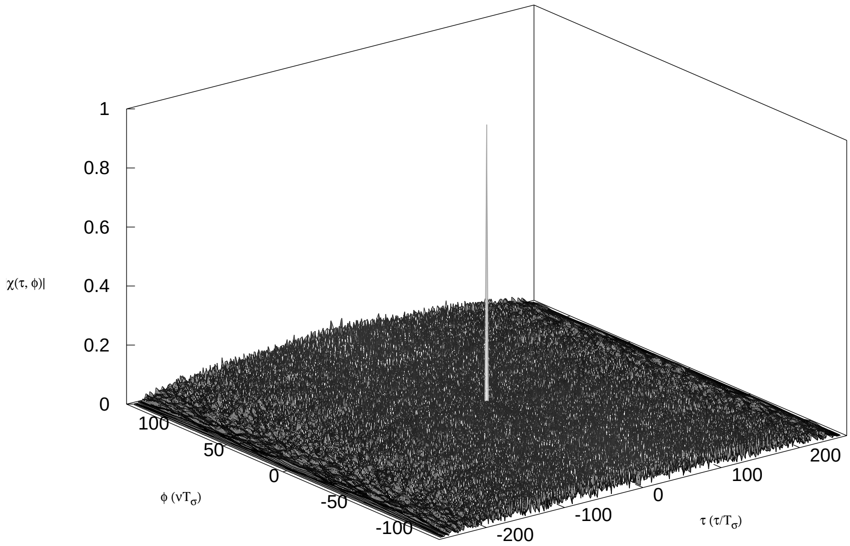

Figure 6.1: M-Sequence ambiguity function

In section 5.3 it was asserted that, from SNR and resolution point of views, white Gaussian noise is the best achievable transmitter signal. However, for any number of reasons, white Gaussian noise is not used as the spreading signal. Instead a PRN stream is used in order to produce a signal having equivalent properties, as seen in figure 6.1, 6.2 and 6.3.

6.1.2 Narrogin, WA

Table 6.1 lists some of the transmitters in the Narrogin area.

6.1.3 Melville, WA

Table 6.2 lists some of the transmitters in the Melville area.

6.2 Problem of Limited Baseband Bandwidth

As mention in 5.7.1 in the default configuration the USB link limits the baseband bandwidth to about 6.5 MHz. From 6.1.1 it can be seen that the required baseband bandwidth is about 9.5 MHz. Possible solutions are:

- Decimate the data in order to reduce the baseband bandwidth to about 6.5 MHz.
- Transfer 8 bit samples across the USB link.
- Transfer 12 bit samples across the USB link.

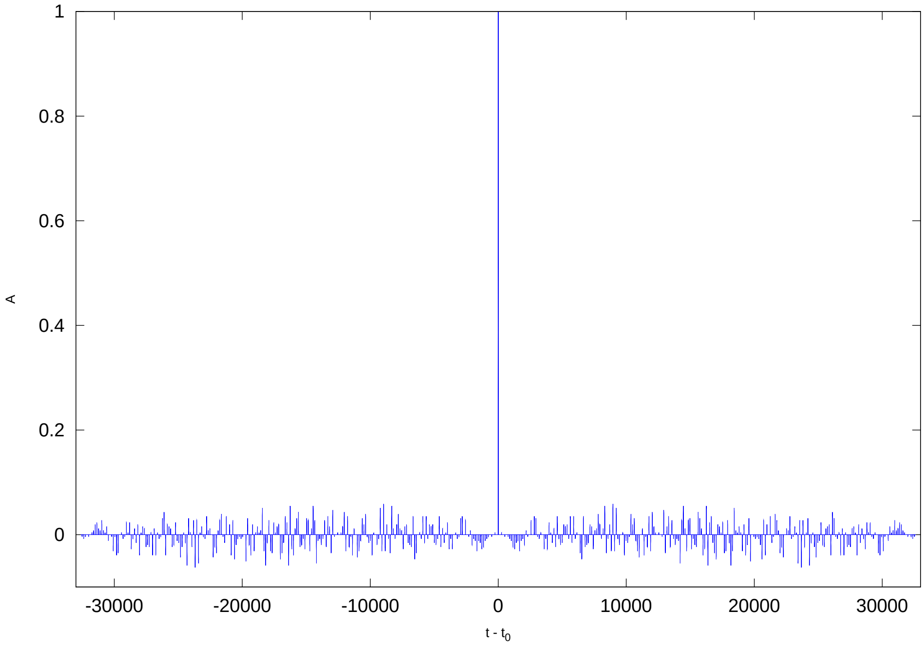

Figure 6.2: M-Sequence autocorrelation

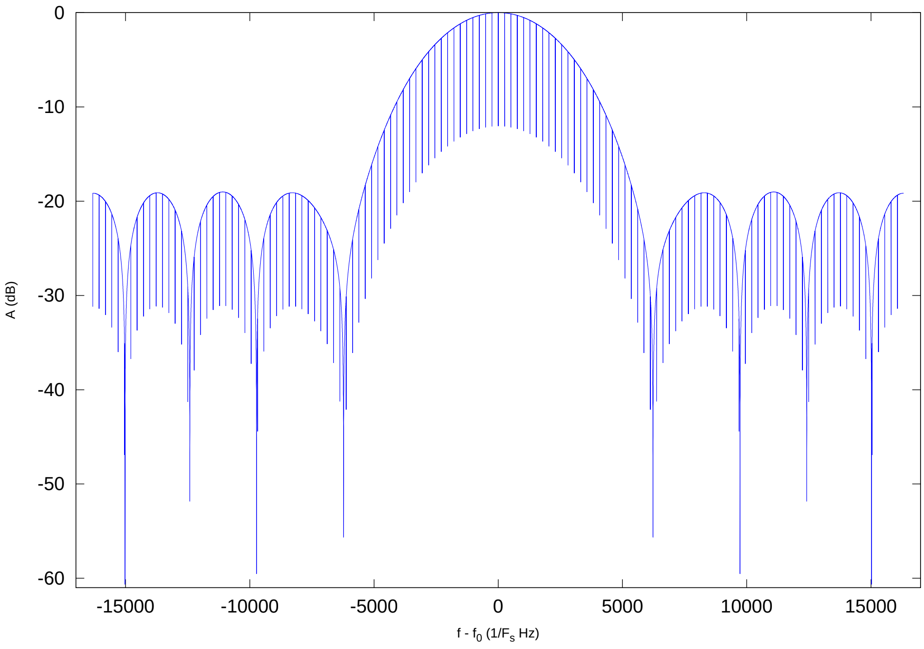

Figure 6.3: M-Sequence envelope

Site	EIRP	Position	AMG Zone 50
CALM Tower Williams Road Narrogin	34.6	S32.9383 E117.1575	514734 6355510
Telstra Radio Terminal Wanerie	41.1	S33.0278 E116.8447	485518 6345582
Western Power/Telstra Site Saunders Hill Wandering Rd Narrogin	32.4	S32.8942 E117.1592	514900 6360400
Telstra CDMA Pt, North Banister	39.9	S32.5714 E116.4333	446809 6396068
Telstra Radio Terminal, Mt Latham	45	S33.3153 E117.2769	625779 6313713
Telstra Site Woglin Hill	40	S32.6781 E116.6972	471624 6384336
Telstra Microwave Site Pingelly	40.1	S32.4681 E117.1014	509518 6407665

Table 6.1: Transmitters in the Narrogin, WA, area in frequency 884.8 MHz

Figure 6.4: Transmitter Locations in Narrogin Area

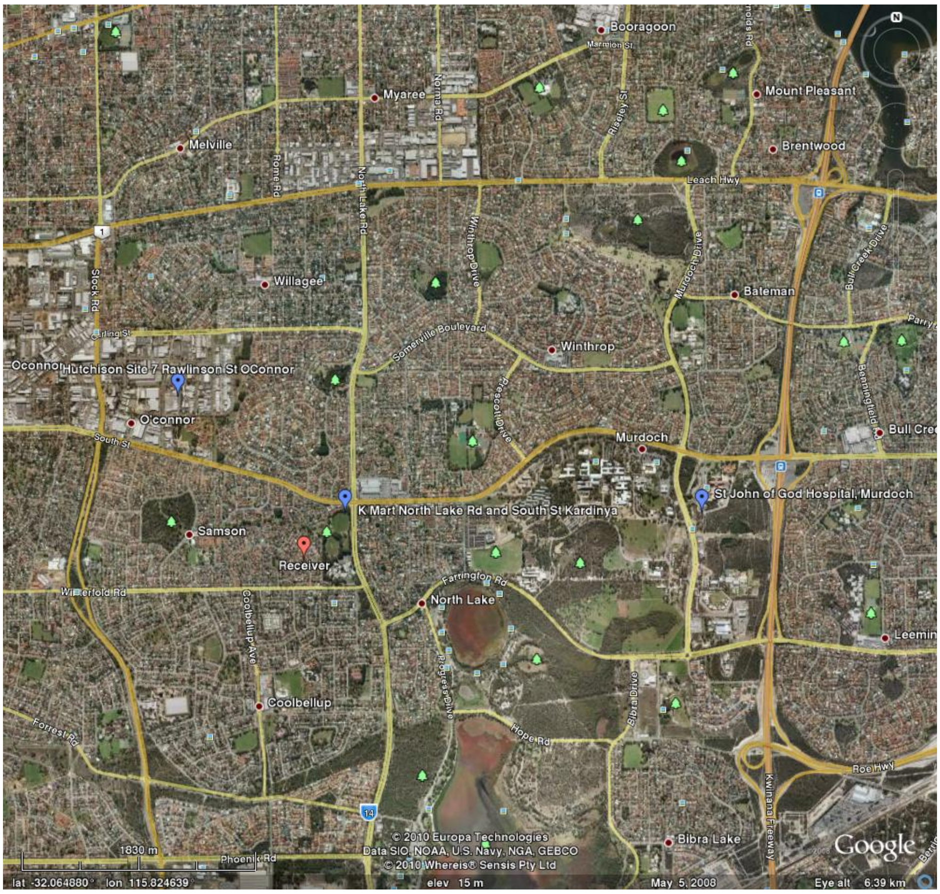

Figure 6.5: Transmitter Locations in Melville Area

Site	Frequency	EIRP	Position	AMG Zone 50
St John of God Hospital, Murdoch	884.8	39.5	S32.0689 E115.8439	390870 6451330
St John of God Hospital, Murdoch	2112.6	41.2	S32.0689 E115.8439	390870 6451330
St John of God Hospital, Murdoch	2162.4	40.4	S32.0689 E115.8439	390870 6451330
St John of God Hospital, Murdoch	2147.4	26.9	S32.0689 E115.8439	390870 6451330
K Mart North Lake Rd and South St Kardinya	884.8	40.3	S32.0686 E115.8142	388080 6451320
K Mart North Lake Rd and South St Kardinya	2112.6	44.1	S32.0686 E115.8142	388080 6451320
Hutchison Site 7 Rawlinson St OConnor	885.0	0	S32.0603 E115.8003	386753 6452212
Hutchison Site 7 Rawlinson St OConnor	2112.6	41.1	S32.0603 E115.8003	386753 6452212

Table 6.2: Transmitters in the Melville, WA, area

- "Chop" the data stream.

The first option will mean that the signal bandwidth will be much less that the expected bandwidth of the PRN signal. This will adversely effect the matched filter characteristics. Just as importantly the discrimination between the several transmitters relies on the ability to receive and track the PRN signal which will be adversely effected by the reduced bandwidth.

The second option will severely reduce the dynamic range and SNR of the system. This option may require changes to the FPGA and driver codes.

The third option shows promise. It is not sure why this is not done in any case. The receiver ADCs have 12 bit resolution. Packing these 12 bit samples would give a baseband bandwidth of about 9.31 MHz. For this application this bandwidth may be marginally sufficient. This option will require changes to the FPGA and driver codes.

The fourth option is probably the best. This option involves transmitting a block of time domain samples and then discarding a following block of samples. The size of the blocks would be such that the buffers in the FPGA and USB controller. The available buffering is about 4096 samples. From the signal processing point of view this is equivalent to multiplying the time domain signal by one corresponding to the samples transmitted and zero otherwise. As long as sample clock phase is maintained (i.e. the receiver software "knows" the number of samples dropped) then the bandwidth is equal to $F_s/2$. However the TB product is affected proportionally.

Chapter 7

Project Strategy

7.1 Driving Factors

There are a number of important factors that are unknown at this stage:

- The suitability of the chosen frequency (880 Mhz) for backscatter from clouds and rain
- The effect of target motion on transmitter chip rate and symbol rate and its effect on correlation
- The unknown nature of the propagation model
- The usefulness of the Telstra *Next G* transmitters as illuminators

7.2 Implications

In light of the above an initial receiver consisting of a USRP with one DBSRX2 module will be built first. This will allow an investigation of the above concerns. Additionally it will allow familiarization with the hardware and its connection to the software.