EVALUATION OF POLARIZATION DIVERSITY TECHNIQUE FOR M2M WIRELESS COMMUNICATION IN STATIC ENVIRONMENT

A Project Report

submitted by

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THESIS CERTIFICATE

This is to certify that the thesis titled **EVALUATION OF POLARIZATION DIVERSITY TECHNIQUE FOR M2M WIRELESS COMMUNICATION IN STATIC ENVIRONMENT**, submitted by **KONDURI SRI RAM**, to the Indian Institute of Technology, Madras, for the award of the degree of **MASTER OF TECHNOLOGY**, is a bona fide record of the research work done by him under our supervision. The contents of this thesis, in full or in parts, have not been submitted to any other Institute or University for the award of any degree or diploma.

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ABSTRACT

Polarization angle diversity technique proposed by Hitachi for reliable M2M communication was studied and an experimental set up was proposed, measurements were obtained. This thesis explains the offered solution under formal signal processing language used by communication engineers. Also the architecture of transmitters and receivers built using USRPs for carrying out propagation studies in environment similar to the intended application.

Multiple Universal Software Radio Peripherals (USRPs) were used to set up the transmitter, the receiver and custom antenna were set up to evaluate circular polarization angle diversity. Measurements were obtained with and without obstructions in both (1) lab environment with reflective walls and (2) workshop environment with a lot of machinery.

This thesis describes the mathematical modelling of the problem, the experimental set up and how the measurements were done. The MATLAB programs developed for the transmitter and receiver and the signal processing steps carried out are given thoroughly. It also gives detailed understanding regarding the implications of the obtained results on the reliability and feasibility of the proposed diversity technique.

The Project Team

The Hitachi team that took part in the experiment had the following members:

1. Dr. Ken Takei, chief designer and overall incharge

2. Mr. Aono, Engineer

3. Mr. Kawakami, Engineer

The IITM team consisted of the following:

1. Prof. Radhakrishna Ganti

2. Prof. David Koilpillai

3. Prof. Devendra Jalihal

4. Mr. Saikiran Valluri, Student

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CHAPTER 1

INTRODUCTION

Even to this day, reliable machine to machine(M2M) communication in industries with large electromagnetic scatterers is established using wired communication. Wireless communication cannot be relied upon due to the highly non-line-of-sight multipath propagation channel in industries. Also the environment in industries is static and is of extremely slow fading type. If the signal is in deep fade it continues to be in deep fade for a very long time. But wired communication has certain disadvantages regarding security of the equipment as the wiring used can be intentionally or unintentionally get damaged.

A radio architecture was proposed by the Hitachi laboratories, Japan, for wireless monitoring and control of infrastructure equipment in industries.The proposed architecture is useful for reliable machine-to-machine(M2M) communication using non-line-of-sight waves which are reflected by electromagnetic scatterers.The proposed radio transmits the signal, which is expanded by orthogonal codes, on different wireless paths using different polarisation directions at different times.This approach results in good resistance against outer interferences and unexpected interruption in highly electromagnetic scattered environment, which enables it to achieve highly-reliable wireless communication,which is inevitable in industries. Also this approach converts the very slow fading channel to fast fading channel so that the signal is relatively not in deep fades resulting in low BERs.

This technique of polarisation angle diversity, proposed by the Hitachi for reliable M2M communications was studied theoretically. Then a contract was established between the Hitachi and Indian Institute of Technology, Madras team for the evaluation of reliability and feasibility of the technique in wireless M2M communication.

1.1 Organization of the thesis

Hitachi has proposed a new radio with circular polarization as an effective solution for M-2-M communication for process industry such as oil refinery, thermal power plants and cement plants. The thesis describes the following works carried out in the project.

In Chapter 2 : Understanding the Hitachi solution in formal signal processing language used by communications engineers, the mathematical modelling of the problem statement is explained.

In Chapter 3 : Building transmitters and receivers using USRPs and carry out propagation studies in environment similar to process industry are described and also samples of data collected are also shown.

In Chapter 4 : Developing transmitter and receiver algorithms in MATLAB and brief description of the code is given.

In Chapter 5 : The problems faced during several phases of project and the solutions to overcome them. The problems of Timing Offset, IQ Imbalance, Carrier Frequency Offset, Frequency aliasing due to uncontrolled bandwidth are explained.

In Chapter 6 : Analysis of the data collected is explained. In this chapter channel estimation, timing offset with SNR, Chip error rate curves are explained.

CHAPTER 2

Mathematical Analysis of Proposed Method

The main important aspect is that the transmitter and the receiver should be in perfect frequency synchronization for better results. In order to conduct the experiments for transmission and reception of data Universal Software Radio Peripherals (USRPs) are used. Special antennas were built at the receiver to go with the USRPs.

The transmitter was implemented using two USRPs synchronized by an exernal cable, as shown in Fig. 2.1

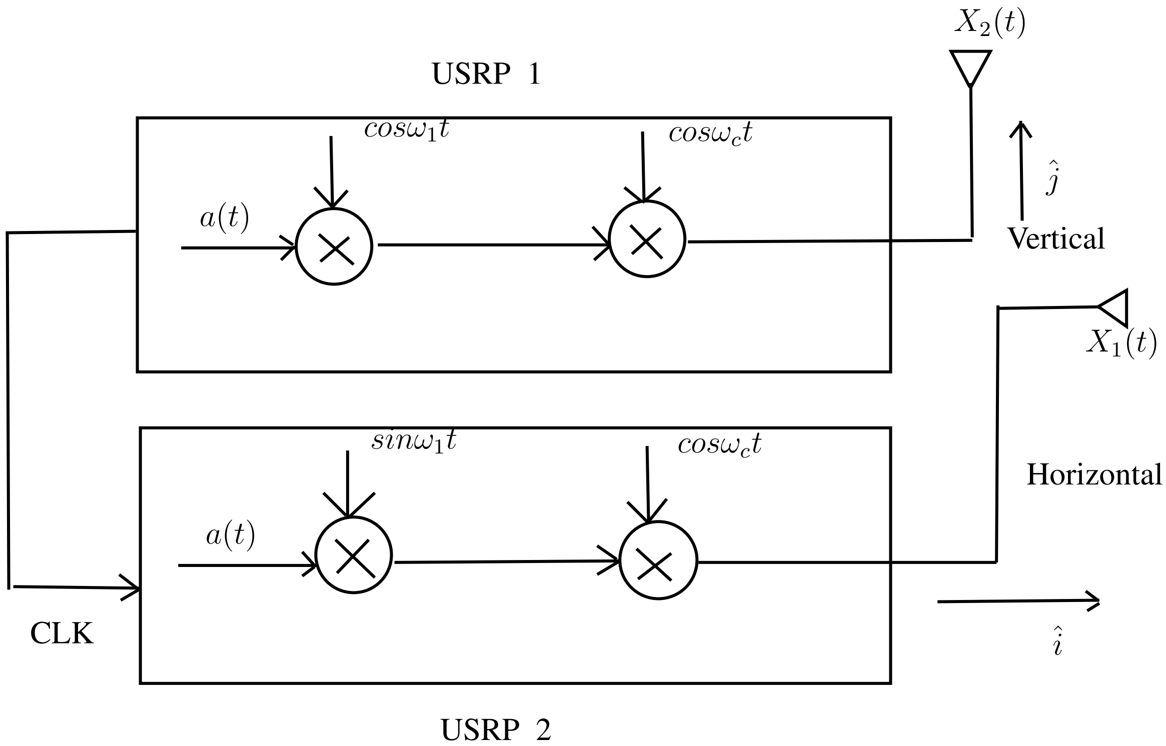

Figure 2.1: Realization of the transmitter using two USRPs synchronized externally.

The transmitted signal on the two antennas are modelled as $X_1(t) = a(t) \sin w_1 t \cos w_c t \ \hat{i}$ and $X_2(t) = a(t) \cos w_1 t \cos w_c t \ \hat{j}$. In our experiment w_1 and w_c correspond to 50 kHz and 460 MHz respectively. $a(t)$ was formed using a 8×4 block code C shown below.

$$C = \begin{bmatrix} -1 & -1 & 1 & 1 & 1 & -1 & 1 & -1 \\ -1 & 1 & -1 & -1 & 1 & 1 & 1 & -1 \\ 1 & 1 & -1 & 1 & -1 & -1 & 1 & -1 \\ 1 & 1 & 1 & -1 & 1 & -1 & -1 & -1 \end{bmatrix},$$

where C_1, C_2 C_3 and C_4 refer to the four rows of C. $a(t)$ is formed by putting the codes C_1 to C_4 in one half cycle of w_1 and repeating it in the other half. One cycle of w_1 signal spans 20 μs and thus, $a(t)$ will have a bandwidth of 3.2 MHz

The transmitted signal experiences narrowband fading due to scattering in the presence of obstruction and the signal is received using vertical and horizontal antenna as shown in Fig. 2.2

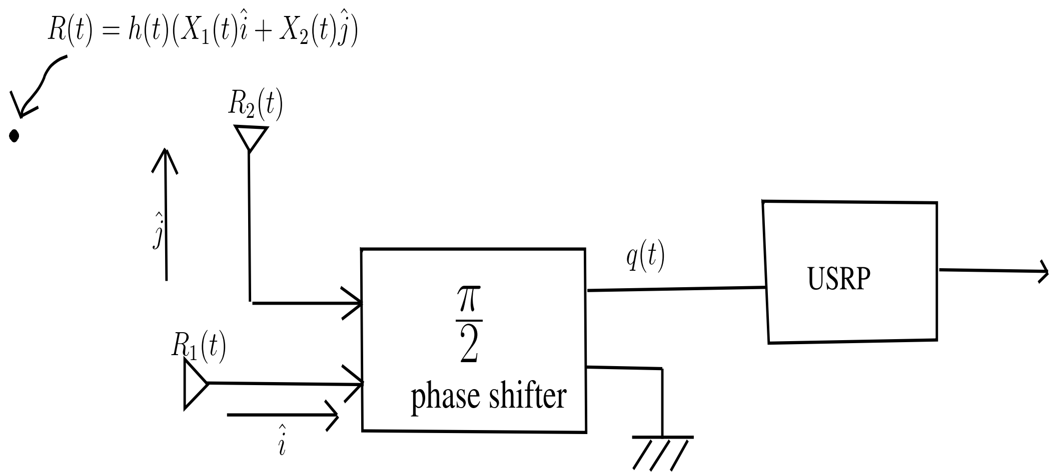

Figure 2.2: Realization of the transmitter using two USRPs synchronized externally.

The two signals add up in space and the receiver signal is given as

$$\begin{aligned}R(t) &= h(t)(X_1(t)\,\hat{i} + X_2(t)\,\hat{j})\\ &= R_1(t)\,\hat{i} + R_2(t)\,\hat{j}\\ &= h(t)(a(t)\sin w_1 t\cos w_c t\,\hat{i} + a(t)\cos w_1 t\cos w_c t\,\hat{j})\end{aligned}$$

From R(t) only the $\hat{i}$ component is received by horizontal antenna, only the $\hat{j}$ component is received by vertical antenna. When $R_1(t)$ and $R_2(t)$ are passed through the $\frac{\pi}{2}$ phase shifter $R_2(t)$ experiences 0^o phase shift and the $R_1(t)$ experiences 90^o phase shift. Therefore

$$\begin{aligned}q(t) &= phase\ shift\ of\ 90^o(R_1(t)) + R_2(t)\\ &= phase\ shift\ of\ 90^o(h(t)X_1(t)) + h(t)X_2(t)\\ &= -h(t)a(t)\sin w_1 t\sin w_c t\ + h(t)a(t)\cos w_1 t\cos w_c t\\ &= h(t)a(t)(\cos w_1 t\cos w_c t - \sin w_1 t\sin w_c t)\\ &= h(t)a(t)cos((w_c + w_1)t)\end{aligned}$$

When this signal is passed through the USRP the signal is converted to base band and we obtain $h(t)a(t)e^{j\omega_1 t}$.

CHAPTER 3

Experimental Set up and Measurements

The experiment was first carried out in the *project discusion room* (room CSD 309) whose walls were covered with aluminum foil that made the walls act like a reflecting surface and an obstruction was fashioned out of the discussion board using aluminum foil. This is shown below in fig 3.1.

Figure 3.1: The experimental set up in room CSD 309

The Machines Laboratory in the Eelctrical Engineering department was selected as the most suitable place to carry out the final measurements as this laboratory resembles a typical process industry. The laboratory is 40 m long and has a number of metal objects which function as obstructions that are so typical of a process industry. The transmitter and receiver were set up in the Machines Laboratory. Figure 3.2 shows the setup.

Figure 3.2: The set up in the Machines Laboratory

3.1 Preparatory steps before recording the measurement data

The transmitter and the receiver USRPs run from their clocks and hence are not synchronized. In order to check whether the system is capable of transmitting and receiving, both ends were fed a 10 MHz generated from a common source, an RF signal generator. Since both ends are synchronized, it is easy to compare received data with the transmitted sequence.

3.2 Measurements in room CSD 309

The discussion room (CSD 309) was modified to have perfect reflecting wall as described above. Here the transmitter and receiver were placed 3 meters apart. A MATLAB program generates the transmitting sequence and converts the samples to .bin files needed for sending it through USRPs.

The antennas used are of 400 MHz ratings and the center frequency at which data is transmitted is 460 MHz. The transmitter has a horizontally polarized antenna and vertically polarized antenna. At the receiver for testing purpose in CSD-309 only one antenna is used whose orientations was varied from 0^0 to 90^0 in steps of 30^0. As predicted, it is noticed that when receiver antenna is oriented at 0^0 only the data from horizontally polarized antenna is received and similarly at 90^0 receiver antenna orientation data from only vertically polarized antenna is received.

Also a hybrid antenna is built which has both horizontal antenna and vertical antenna and data is received with this antenna also at receiver.

Figure 3.3: Hand made hybrid antenna

A number of different scenarios described below were realized and in each case data was recorded with and without the obstruction.

Without Obstruction Date Conducted March 5, 2014.

1. Single transmit antenna at horizontal orientation, Single receive antenna at $0^0, 30^0, 45^0, 60^0, 90^0$ orientations.
2. Two transmit antennas (One horizontal and other vertical), One receive antenna at $0^0, 45^0, 90^0$ orientations.
3. Two transmit antennas (One horizontal and other vertical), hybrid antenna at receiver (One horizontal and one vertical).

The transmitter and receiver USRPs cannot have exactly same carrier frequency of 460 MHz. So there is a problem of Carrier Frequency Offset(CFO). For the purpose of testing, an RF frequency generator is used to give external clock to the USRPs to ensure synchronization of the USRPs. With this setup, from March 6, 2014 to March 10, 2014, following scenarios were setup.

Without Obstruction Two transmit antennas (One horizontal and other vertical), One receive antenna at $0^0, 45^0, 90^0$.

With Obstruction The following experiments are done at different obstruction spacings, *near* (0.5 m from receiver), *mid-range* (2.5 m from receiver), *far* (4 m from reciever).

1. Single transmit horizontal antenna, Single receive antenna at $0^0, 90^0$.
2. Two transmit (horizontal and vertical) and One receive antenna at $0^0, 90^0$
3. Two transmit (horizontal and vertical) and hybrid antenna.

The following scenarios were setup without using the external synchronizing source and repeated With and without obstruction.

- Two transmit (horizontal and vertical) and hybrid antenna with the obstacle placed at *near*, *mid-range* and *far* from the receiver.

3.3 Measurement in the Machines Laboratory

The measurements in the Machines Laboratory was carried out on March 12, 2014. The Machines lab has a length of 40 m. The receiver was located at a range of 30, 15 and 5 m from the transmitter. At each of these locations the transmit power was varied over two gains of 7.5 and 12.5 dB, and the received signals were stored. This set up too uses two transmit antennas and at receiver a hybrid antenna.

1. In Line Of Sight (LOS) for transmitter and receiver at 5 m, 15 m and 30 m measurements were performed.
2. The measurements were repeated in NLOS situations when there is no direct path to receiver.

The following are the sample data collected. 3.6

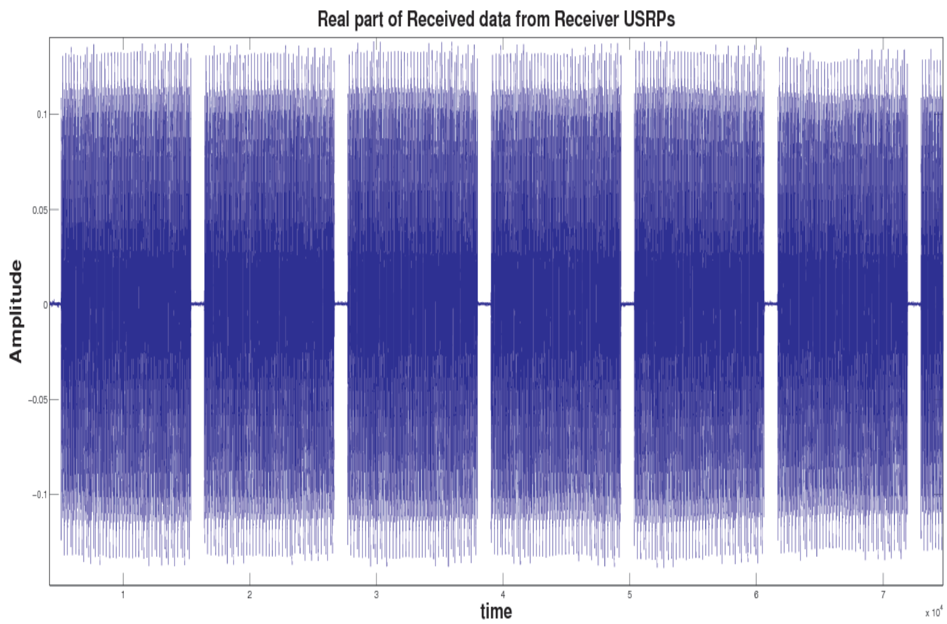

Figure 3.4: Real part of received data.

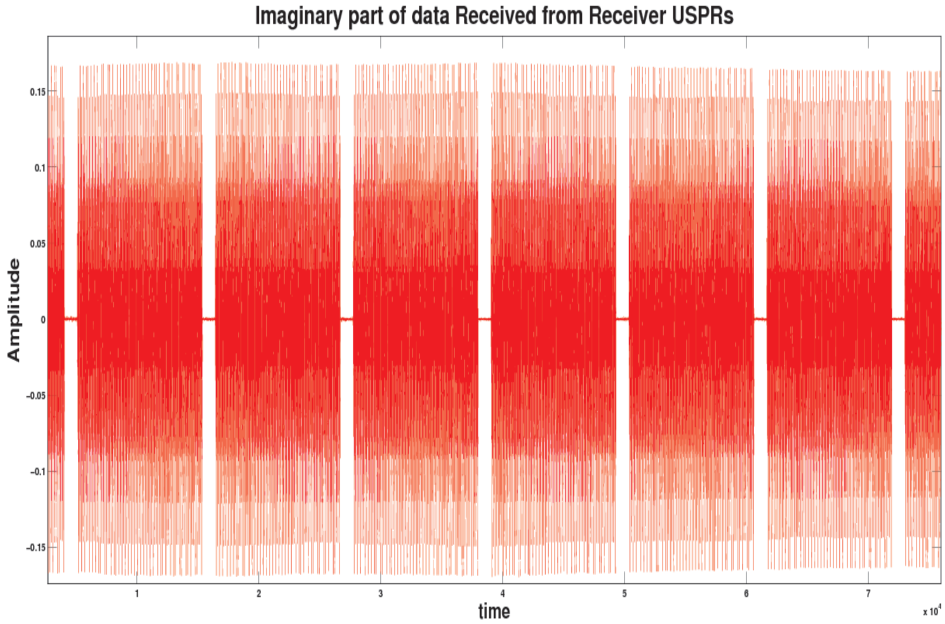

Figure 3.5: Imaginary part of received data.

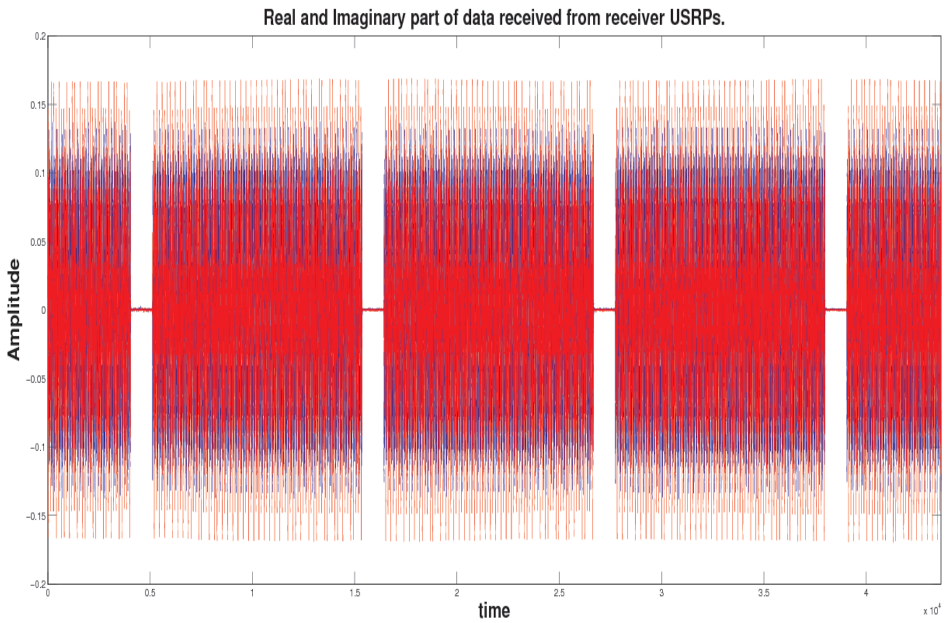

Figure 3.6: Real and Imaginary part of received data together.

CHAPTER 4

Transmitter and Receiver Block diagrams

4.1 Transmitter

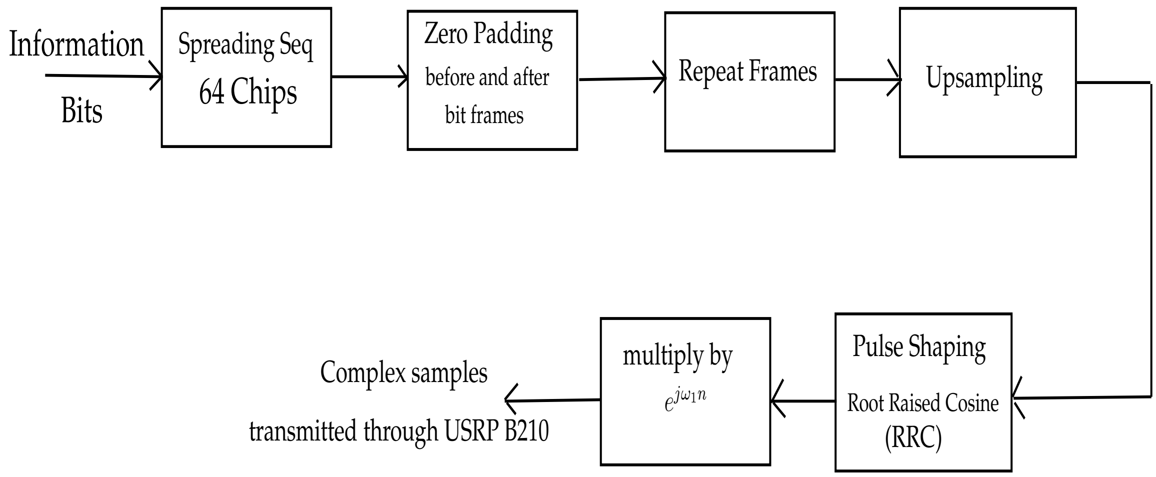

Figure 4.1: Block Diagram of the transmitter

The above block diagram describes all the functions carried out in transmitter. In our case there are two antennas transmitting at the transmitter. The connection to horizontal antenna sends $a(t)\sin(\omega_1 t)$ and the connection to vertical antenna sends $a(t)\cos(\omega_1 t)$. Then before sending to antennas the USRPs upconvert the signals in horizontal and vertical paths and transmit them.

4.2 Receiver

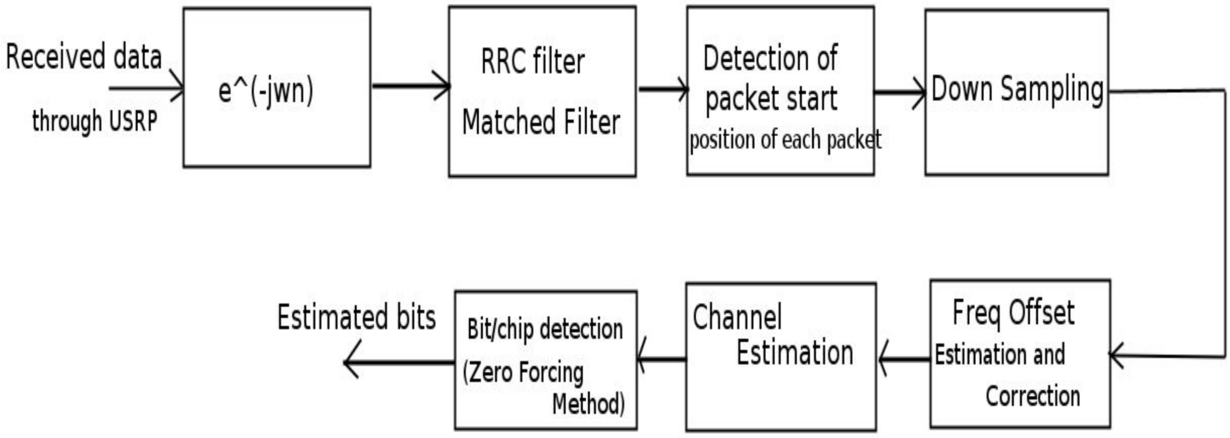

Figure 4.2: Block Diagram of the Receiver

The above block diagram shows the steps undertook in the receiver after the data is optained from the antennas. First the effect of $e^{j\omega n}$ is removed since it is used for generating circular polarization in antennas. Then matched filter is used which is RRC. For detection of data packets zero padding is done before and start of each frame. So by looking at the received data approximately the instant of start of frame is easily known. Then down sampling is done to get correct number of received samples. At this stage the data should be only real. But we see complex data because of CFO between the USRPs. So CFO is removed using unwrapping algorithm. Then channel is assumed to be narrow band fading. First the channel is estimated taking some preamble bits. With this channel the chips(data) is found using Zero forcing method. The the chip error rate is calculated.

CHAPTER 5

Problems faced during the project and the appropriate steps taken

The main problems are as follows:

1. When the two transmitter USRPs are connected by a *MIMO cable*, one of the USRPs serves as the master clock. However, this does not ensure that the rising edge of the data points are synchronized. There is indeed some offset. This timing offset is to be estimated and compensated.
2. The RF section of the USRP exhibits some I-Q imbalance which again needs to be estimated for each USRP and an appropriate pre distortion needs to be carried.
3. The Carrier Frequency offset between two USRPs also will effect the performance. This has to be taken care externally or within algorithm.
4. Frequency aliasing and Inter Symbol Interference due to large bandwidth of the data transmitted.

5.1 Timing Offset

When the two transmitter USRPs are connected by a *MIMO cable*, one of the USRPs serves as the master clock. However, this does not ensure that the rising edge of the data points are synchronized. There is indeed some offset. The two signals transmitted by USRPs are $X_1(t)$ and $X_2(t)$. If they have the timing offset then the transmitted signals are as shown.

$$\mathrm{X}_1(t) = a(t)\cos w_1 t \cos w_c t$$
$$\mathrm{X}_2(t) = a(t-\tau)\sin w_1 (t-\tau) \cos w_c t$$

where τ is the timing offset between the rising edges of data samples of the two Transmitter USRPs. This eventually leads to non-synchronisation between the USRP outputs.

In order to overcome this timing offset problem the transmitter USRPs are changed from USRP-N210 to USRP-B210 where can send the data directly without MIMO cable and enabling the two ports of USRP and transmitting data using C code. Due to this there is precise timing of both the transmitter antennas.

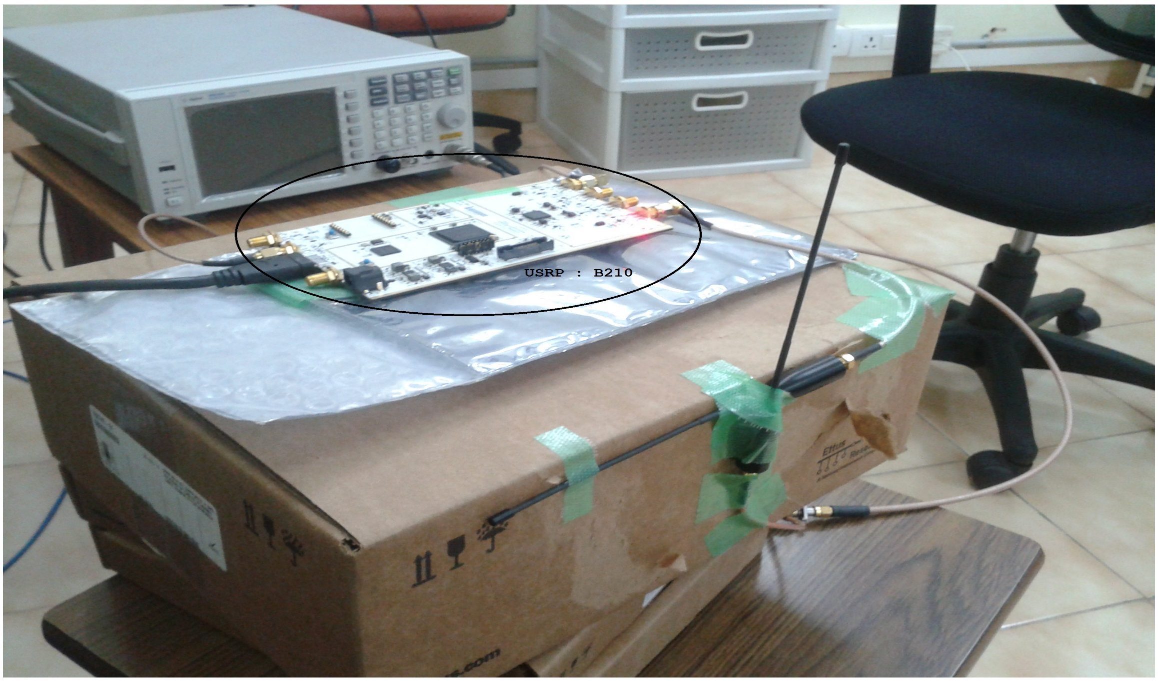

Figure 5.1: USRP B210 and the transmitting antennas

5.2 IQ Imbalance

5.2.1 Introduction

This is a severe problem in all Analog direct conversion receivers. Consider the conversion of single tone signal at RF to base band. Ideally the I and Q outputs at the receiver are

$$I(t) = cos(\omega t)$$
$$Q(t) = sin(\omega t)$$

respectively. ω is the base band frequency of the tone.With no loss of generality, we have normalized the magnitude to unity and the phase to zero, as these quantities are not relevant to this discussion. In contrast, a realistic direct conversion receiver produces:

$$I(t) = \alpha cos(\omega t) + \beta_I$$
$$Q(t) = sin(\omega t + \phi) + \beta_Q$$

where ϕ is the phase error, which we have assigned to the 'Q' path, α is the magnitude error, which we have assigned to the 'I' path, and β_I and β_Q are the DC biases associated with each path. The allocation of error mechanisms to paths is completely arbitrary and implies no loss of generality.

Correcting β_I and β_Q is very simple. For example, β_I is simply the mean of I (t) over an integer number of periods. Given this estimate, the correction is simply a matter of subtracting β_I from the 'I' path signal. The process is the same for the 'Q' path. Then, we are left with:

$$I(t) = \alpha cos(\omega t)$$
$$Q(t) = sin(\omega t + \phi)$$

This can be rewritten in matrix form as:

$$\begin{pmatrix} I''(t) \\ Q''(t) \end{pmatrix} = \begin{pmatrix} \alpha & 0 \\ \sin(\phi) & \cos(\phi) \end{pmatrix} \begin{pmatrix} I(t) \\ Q(t) \end{pmatrix}$$

Thus we find that the correction is :

$$\begin{pmatrix} I(t) \\ Q(t) \end{pmatrix} = \begin{pmatrix} \alpha^{-1} & 0 \\ \alpha^{-1} tan(\phi) & sec(\phi) \end{pmatrix} \begin{pmatrix} I''(t) \\ Q''(t) \end{pmatrix}$$

Thus, it remains only to findα and ϕ. To find α, define :

$$< I''(t) I''(t) >= \alpha^2 < cos^2(\omega t) >= \alpha^2 < \frac{1}{2} + \frac{1}{2}\cos(2\omega t) >= \frac{1}{2}\alpha^2 \tag{5.1}$$

and similar analysis shows that

$$< I''(t) Q''(t) >= \frac{1}{2}\alpha^2 \sin(\phi) \tag{5.2}$$

Thus the above equations are used to estimate and compensate the IQ imbalance.

5.2.2 IQ Imbalance for the data transmitted using USRPs

The RF section of a USRP has some Inphase-Quadraturephase(I-Q) imbalance which results in distortion of the complex samples as illustrated.Let us suppose the USRP is for implementation of receiver and the passband signal received is:

$$q(t) = h(t)a(t)\cos\psi(t)\cos(w_c t + \theta) - h(t)a(t)\sin\psi(t)\sin(w_c t + \theta)$$

The expected equivalent baseband signal $\tilde{q}(t)$ as USRP output is given by

$$\tilde{q}(t) = h(t)a(t)e^{j\psi(t)} = h(t)a(t)e^{j(w_1 t + \phi(t))}$$

so that,

$$Inphase[\tilde{q}(t)] = h(t)a(t)cos(w_1 t + \phi(t))$$

and

$$Quadraturephase[\tilde{q}(t)] = h(t)a(t)sin(w_1 t + \phi(t))$$

but the actual received baseband signal components due to I-Q imbalance is:

$$Inphase[\tilde{q}(t)] = \alpha\ h(t)a(t)cos(w_1 t + \phi(t))$$

and

$$Quadraturephase[\tilde{q}(t)] = h(t)a(t)sin(w_1 t + \phi(t) + \psi),$$

for some $\alpha\ \epsilon\ (-1, \infty)$.

In order to prevent this distortion,I-Q imbalance needs to be appropriately estimated for α and ψ and compensated for.

5.2.3 IQ Imbalance Compensation Using Block in GNU Radio Companion

However in GNU radio companion this can be done by a block called 'IQ Bal Optimize'. This will estimate the IQ imbalance factor and removes it.

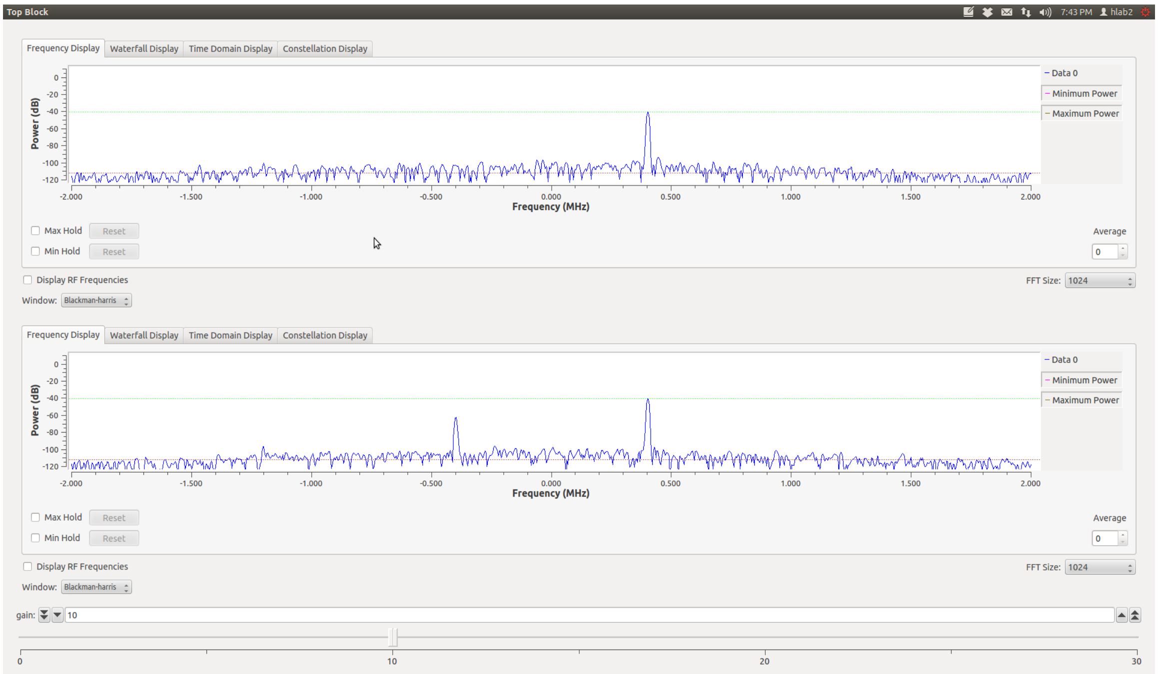

Figure 5.2: IQ imbalance after and before compensation

In the above we can observe that when IQ imbalance is not compensated an extra tone appears. When compensated with the block in GNU Radio Companion only the desired tone i.e the carrier is only present.

5.3 Carrier Frequency offset

There is significant carrier frequency offset(CFO) between the transmitter and receiver USRPs, which needs to be compensated.Due to this, the resultant received baseband signal is:

$$\tilde{q}(t) = h(t)a(t)e^{j(w_1 t + w_o t)},$$

where w_o is the carrier frequency offset. This carrier frequency offset is observed that it remains nearly same over large data.

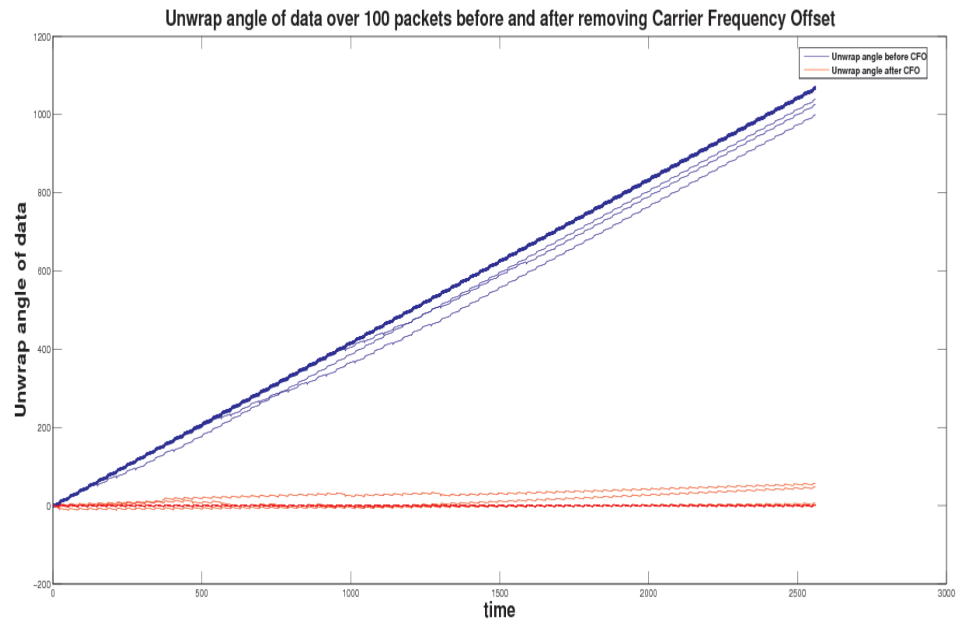

Figure 5.3: Unwrap angle of data before and after removing Carrier Frequency Offset

The above plot shows the unwrap angle of data over several packets before correcting for CFO and after correcting for CFO.

5.4 Frequency aliasing due to uncontrolled bandwidth

The 'chips' forming the block codes were used without any pulse shaping which resulted in uncontrolled bandwidth,which was greater than the sampling rate used at transmitter.When the same sampling rate was used at the receiver to extract the complex samples,there was frequency domain aliasing.Thus, the bandwidth of the signal transmitted needs to be reduced.

In order to constrain the bandwidth, pulse shaping of the transmitting data is required. So Root Raised Cosine(RRC) is used as pulse shaping signal at transmitter and also used at receiver as a matched filter.

CHAPTER 6

Analysis of the Data Collected and Results

6.1 Fade Coefficients seen by polarised and single antenna receivers

In order to remove the dependency of the results on the accuracy of the synchronization algorithms at the receiver, it was decided to externally synchronize the transmitter and receiver by RF frequency generator. This allows us to focus on understanding the behaviour of polarized antennas under fading. With these changes effected, the experiment was repeated in the machines lab.

The data was recorded and analysed. The averaged $|h(t)|^2$(where $h(t)$ is the narrowband channel fade coefficient) obtained after down converting and down sampling, compensating for $e^{jw_1 t}$ and squaring, is plotted as a function of 64 chips.The averaging is done over 100 packets (each packet containing 40 bits i.e. 2560 chips).

Figures 6.1 to 6.4 compare the average fade coefficient $|h(t)|^2$ for the two scenarios, the case that uses the polarization angle diversity(polarised),shown in blue colour, and the case that uses single antenna both at the transmitter and the receiver,shown in red, and the comparison is done for both Line-of-sight(LOS) and non line-of-sight(NLOS) metallically obstructed propagation path scenarios.The x-axis shows the chip position number out of 64 chips forming a bit.

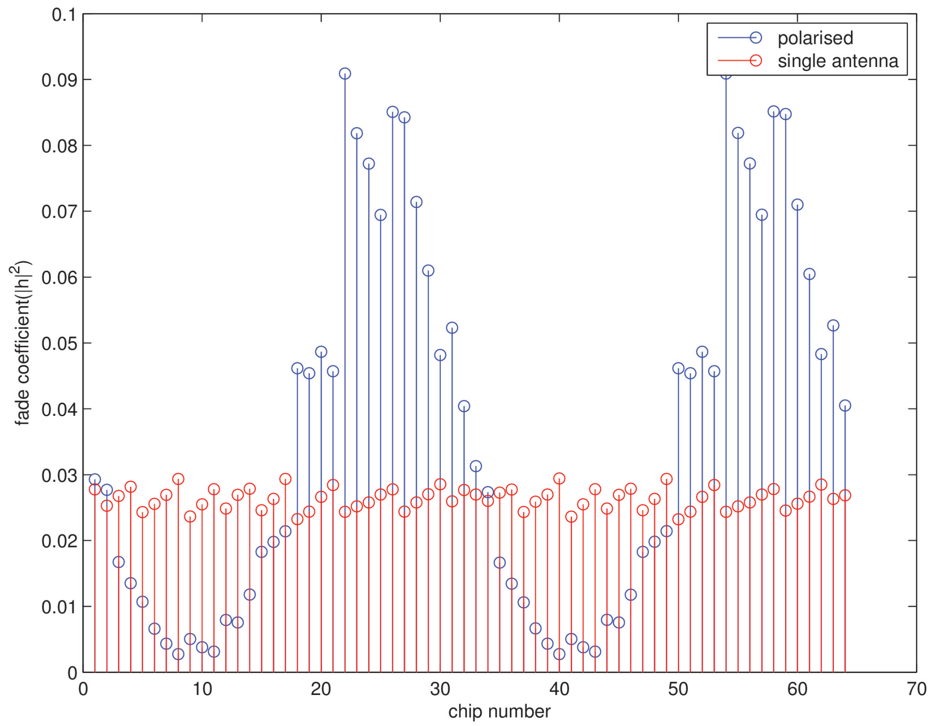

Figure 6.1: Comparison of fade coefficients as seen by polarised angle diversity and single antenna transceivers under an LOS scenario.

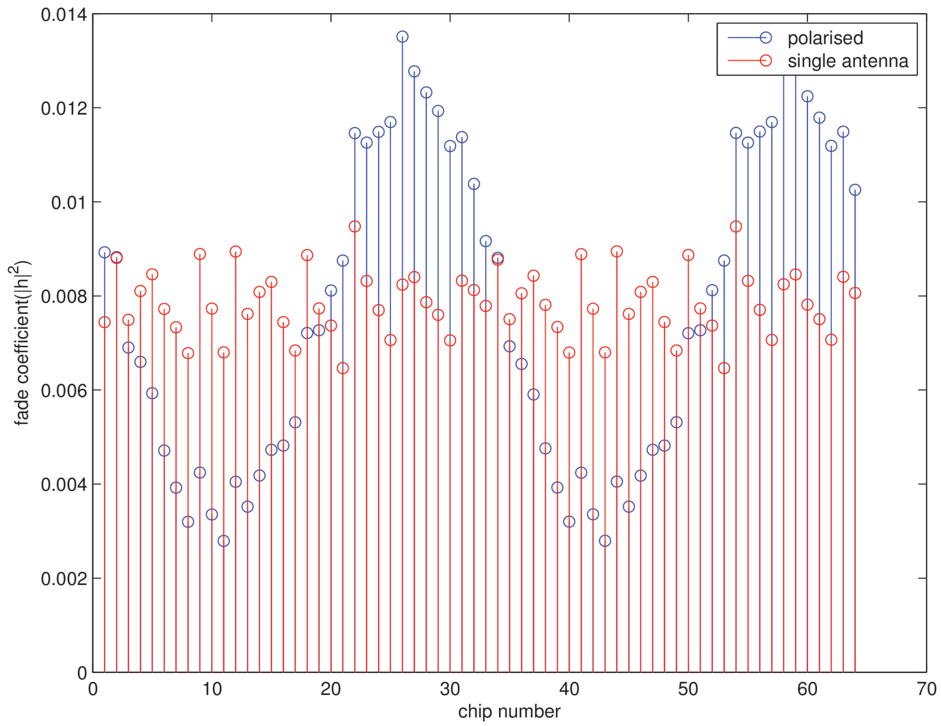

Figure 6.2: Comparison of fade coefficients as seen by polarised angle diversity and single antenna transceivers under another LOS scenario.

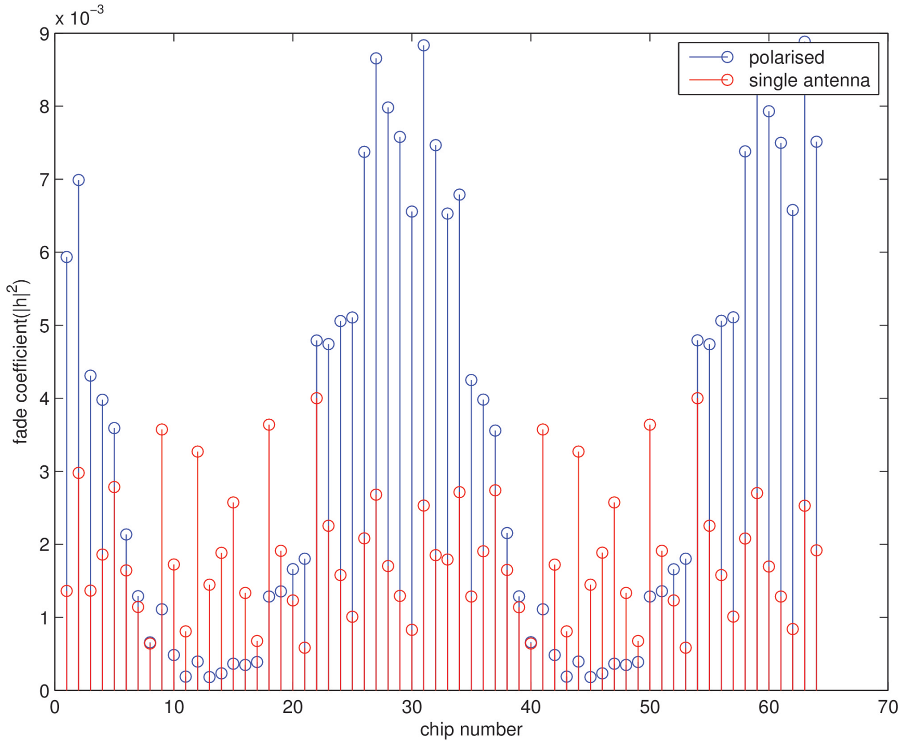

Figure 6.3: Comparison of fade coefficients as seen by polarised angle diversity and single antenna transceivers under an NLOS(non line-of-sight) scenario.

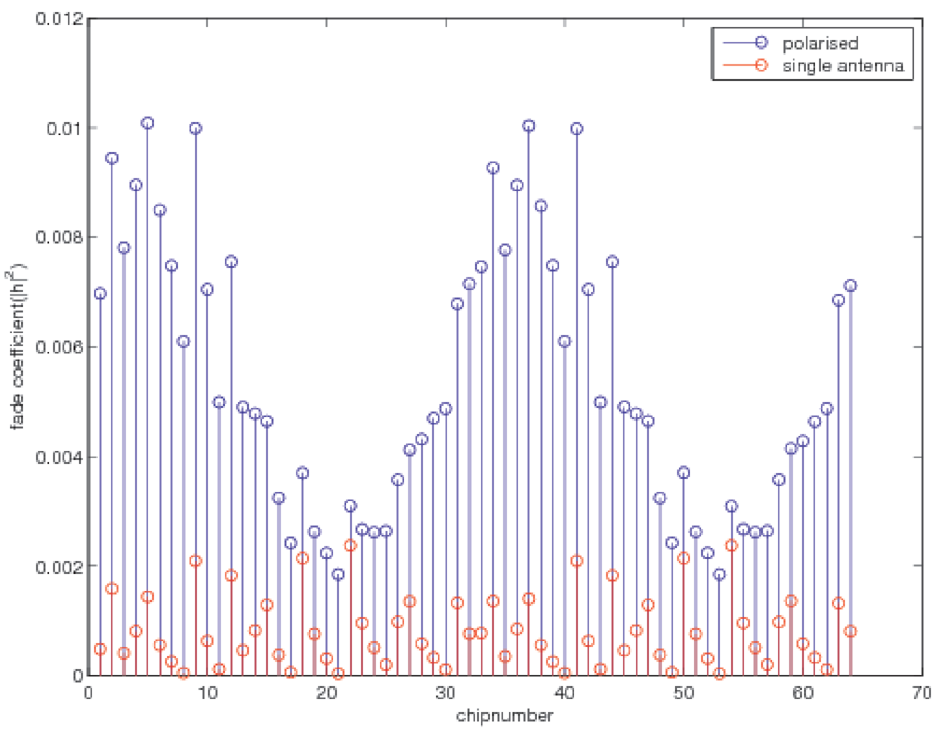

Figure 6.4: Comparison of fade coefficients as seen by polarised angle diversity and single antenna transceivers under another NLOS(non line-of-sight) scenario.

6.2 Channel Estimation

In order to estimate channel some of the starting bits of the transmitted bits are assumed to be known which act as training sequence. Using this training sequence the channel is estimated and eventually the chips and bits are recovered.

The received data after all the processing done at the receiver is

$$y(t) = h(t)a(t) + n(t)$$

where $h(t)$ is the fade coefficients,$a(t)$ is the actual signal and $n(t)$ is the noise. Using the training sequence $a(t)$ upto some time is assumed to be known. Therefore channel is estimated as

$$\hat{h}(t) = \frac{y(t)}{a(t)}$$

This approach is called '$ZeroForcingMethod$'.

Once $\hat{h}(t)$ is known $a(t)$ can be estimated as

$$\hat{a}(t)=\frac{y(t)}{\hat{h}(t)}$$

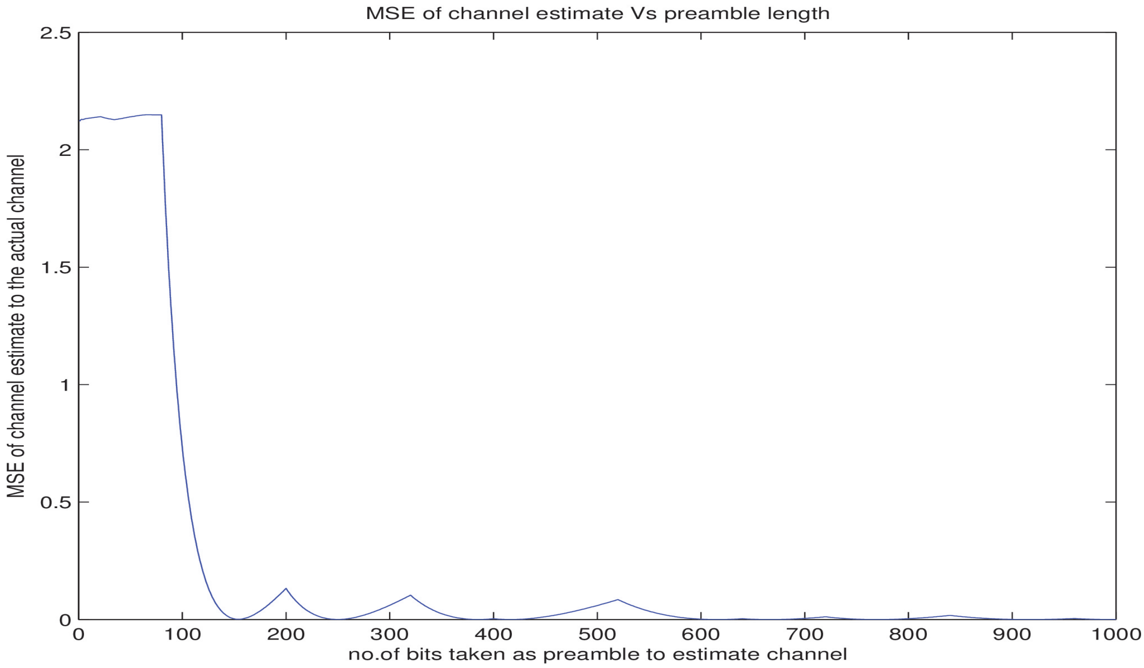

Figure 6.5: MSE between actual channel coefficients and estimated channel coefficients

6.3 Performance evaluation of receiver algorithm

The polarisation angle diversity transceiver algorithm was simulated in MATLAB.After obtaining the transmitter complex samples for a large number of frames,say 1000, by the transmitter algorithm mentioned earlier, the samples were subjected to complex additive white gaussian noise as shown in the equation below:

$$y[n] = s[n] + w[n],$$

where y[n] is the received sample,s[n] the transmitted sample and w[n] is the additive gaussian noise.The SNR of the simulated AWGN channel is varied from 0 to 20 dB.

Then the received samples are subjected to carrier frequency offset(Fo) at the receiver.The simulations are performed with Fo taking values as 4 KHz(which is worst case frequency offset for 5 ppm USRP at center frequency 400MHz),8 KHz(for 10 ppm USRP),12 KHz(15 ppm USRP) and 16 KHz(20 ppm USRP).Then, the chips are recovered using the receiver algorithm mentioned earlier.The mean square error of the estimated frequency offset,averaged over the 1000 frames used, is obtained by comparing the frequency offset, estimated by the technique mentioned in transmitter algorithm, with the actual frequency offset introduced.The plot for logarithm of mean square error of estimated carrier frequency offset(in radians/sample at sampling frequency 2MHz)for Signal to Noise ratio of the AWGN channel varying from 0 to 20 dB is shown below in fig.6.6

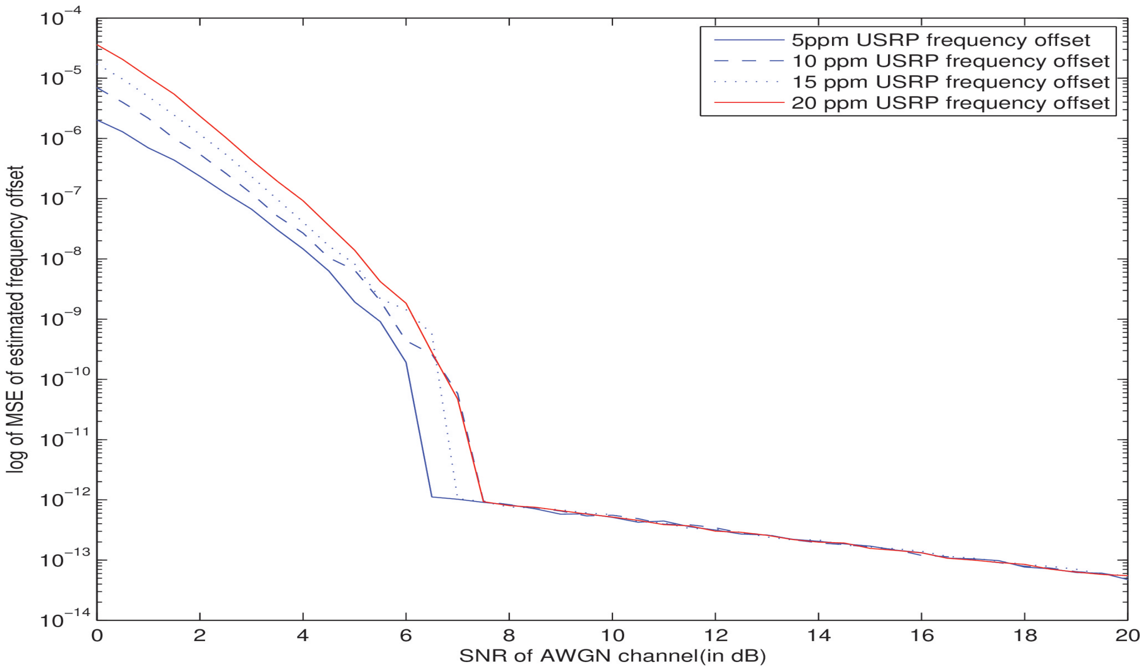

Figure 6.6: Logarithm of mean square error of estimated frequency offset(in radians/sample at sampling frequency 2MHz) vs Signal-to-Noise ratio(SNR) of AWGN channel(in dB) for worst case frequency offset for 5, 10, 15 and 20 ppm USRPs.

6.4 Chip Error Rate

As the data is obtained over many distances in the practical environment, with the increase in distance between the transmitter and receiver the chip error rate increases because of fading, attenuation, AWGN noise. In our case the noise is very less in the received data. So we have simulated AWGN in matlab and is added to the data with appropriate SNRs. Then channel estimation is done to know $h(n)$ and zero forcing equalizer is used to find the bits and finally chip error rates.

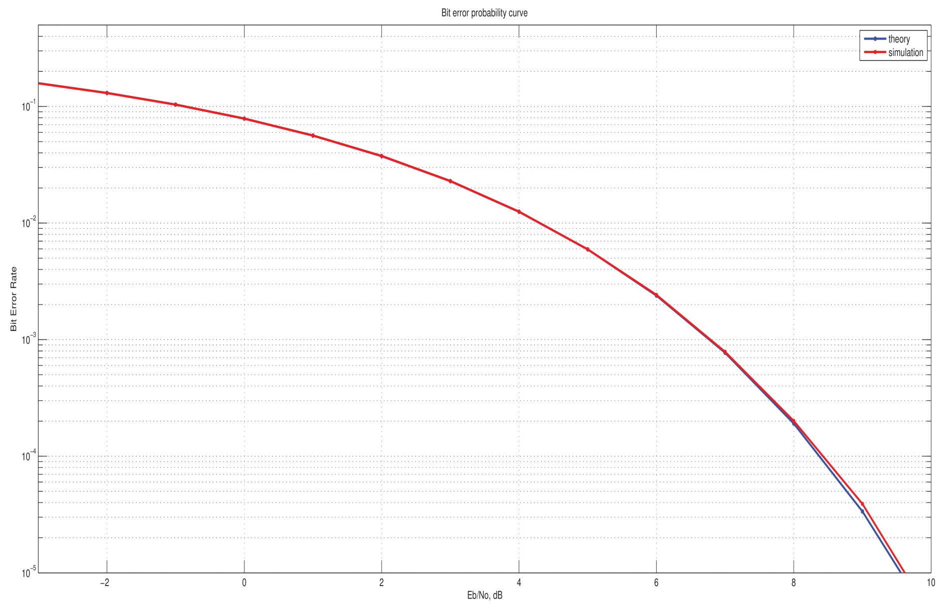

Figure 6.7: Chip Error Rate Vs SNR

6.5 Conclusion

1. The preliminary results of the experiment suggest that use of Hitachi polarized antennas results in shaping of the fading environment. These initial results shows circular polarization in action and we see that the fading seen by a particular chip is based on the polarization it experiences.

2. Frequency synchronization is very critical to the working of the system. Unlike other single-carrier systems, we deliberately multiply the signal with a complex exponential to induce circular polarization and this is of the order of few KHz. Hence the frequency offset should be much less than the KHz range for the system to operate correctly.

3. Pulse shaping with an RRC pulse and an appropriate bandwidth expansion factor is absolutely necessary for the system to work. Without pulse shaping, the receiver ADC would be required to operate at a much higher bandwidth with low-pass filtering would be required. This is mainly to prevent aliasing and ISI caused by the infinite bandwidth of the square pulses being transmitted.

4. With RRC pulse shaping, we would require at least twice the Nyquist sampling rate at the receiver so as to locate the optimal sampling time. These findings are confirmed by carrying out careful experiments at different distances between the transmitter and receiver, under both LOS and NLOS scenarios.