CHAPTER 7

SPREAD SPECTRUM MIMO RECEIVERS

7.1 INTRODUCTION

This chapter investigates the performance analysis and comparison of full chip and half chip rate of Noncoherent (NC) and Differentially Coherent (DC) code acquisition scheme in MIMO (Multiple Input-Multiple Output) assisted by Direct Sequence Spread Spectrum (DSSS) wireless system when communicated over uncorrelated Raleigh channel. A novel approach for half chip is proposed and is proved to better choice comparing to full chip rate system. Four schemes are investigated, namely SISO with full chip rate, SISO with Half chip rate, MIMO with full chip rate and MIMO with Half chip rate by varying the code acquisition technique. The performance metrics considered are namely Bit Error Rate and Mean Acquisition Time. Viterbi (1995) has discussed the principles of spread spectrum communication.

7.2 CONVENTIONAL DS-CDMA RECEIVER MODEL

In CDMA scheme, a locally generated code runs at a much higher rate than the data to be transmitted. In this method, the modulated waveform is spread second time in such a way to produce an expanded wide band signal whose bandwidth is greater than the available bandwidth. This signal does not interfere with other signals. Such an expansion is achieved using second modulation. Many potential advantages are achieved over conventional
systems such as improved interference rejection, high resolution ranging, secured communication and increased capacity and a better spectral efficiency.

Figure 7.1 shows the amplitude spectrum before and after spreading. Popovic and et al (1999) has presented analysis of the basic criteria for the selection of spreading sequences for the multicarrier CDMA systems. Some well-known classes of sequences, such as Walsh, Gold, orthogonal Gold, and Zadoff-Chu sequences, as well as Legandre and Golay complementary sequences, were considered. Park (1999) has performed numerical evaluations to examine the effects of decision thresholds, post detection integration, and fading rate. Also a double-dwell serial search technique for cellular Code-Division Multiple-Access (CDMA) networks was developed. Suwansantisuk and Win (2007), have proposed a novel a search technique and derived closed-form expressions for both the minimum and maximum Mean Acquisition times (MAT) and the conditions for achieving these limits. Claude Amours and et al (2009) have discussed about the Spreading Code assignment Strategies for MIMO-CDMA Systems Operating in frequency-selective channels. Lee et al (2001) presented scrambling techniques for CDMA Communications.

In Direct Sequence Spread Spectrum (DSSS), the spread signal is obtained by multiplying the information signal directly with a wideband PN signal. The information rate is \( R = \frac{1}{T_b} \) bits per sec, where ‘\( T_b \)’ is one bit interval. The PN signal rate is \( \frac{1}{T_c} \) ‘chips’ per sec where ‘\( T_c \)’ is one chip duration. One bit interval occupies \( N \) chips, ie

\[ N = \frac{T_b}{T_c} \quad (7.1) \]
There are many issues in the conventional DS-CDMA systems which are as given below:

- If all users transmit almost with identical power, then beyond certain point, increase of power by every user will not reduce the bit error rate.

- If the user transmits with widely different power, then the conventional receiver allows the signal from the powerful user by suppressing the signal of weaker user.

- Detection of desired signal is limited by inherent suppression capacity of the system

In a DS-CDMA, the data stream is multiplied with the unique spreading sequence and then it is transmitted. This multiplication operation serves to spread the bandwidth of the information signal into wider bandwidth occupied by PN signal. The receiver has the knowledge of the spreading sequence and can demodulate the transmitted data. The demodulation of the signal is performed by multiplying the received signal with a replica of the waveform generated by the PN code sequence generator at the receiver, which is synchronized to the PN code in the received signal.
Synchronous DS-CDMA and Asynchronous DS-CDMA are the two major classifications of DS-CDMA systems. In a Synchronous DS-CDMA, all the user transmissions are synchronized at the chip level, whereas in an Asynchronous DS-CDMA system, transmissions from different users arrive at the receiver at different time. The synchronous system model is best suited for mobile stations because the received signal at the mobile stations is synchronous. The asynchronous system model is well suited for base stations because the transmitted signals from the mobile stations are asynchronous.

In DS-CDMA spread spectrum systems, the chip rate is much greater than the flat fading bandwidth of the channel and the spreading codes are designed to provide very low correlation between the successive chips. Therefore multiple versions of the transmitted signals are available at the receiver. The transmitted signals are subject to multipath fading, due to multiple propagation paths between the transmitter and the receiver, because of the differences in the path delays and phases of signals received through multiple reflected paths. Multipath Fading introduces

- Delay-spread due to power delay.
- Doppler-spread due to time varying nature of the channel causing the relative motion between the mobile and base station.
- Fade amplitude statistics.

7.3 MULTIPLE ACCESS INTERFERENCE (MAI)

Channels shared by multiple users are frequently damaged by MAI, in which users’ transmissions interfere with each other. In DS-CDMA, (where users are assigned signaling waveforms that are known to have slight correlation) the MAI becomes substantial when the number of interferers or
their power relatively increases. The conventional CDMA system treats each user as a separate signal, with the other users considered as either interference or noise. This assumption produced the conventional receiver, which is a Matched Filter (MF) receiver equivalent to optimal receiver in the single user Additive White Gaussian Noise (AWGN) communication channel.

The MAI term is often modeled as Gaussian in various analyses and the increase in MAI due to other user transmission are viewed as an increased noise level. The conventional DS-CDMA receiver consists of a bank of ‘K’ matched filters, and each matches to different users’ spreading sequence, as shown in Figure 7.2.

![Figure 7.2 DS-CDMA system](image-url)
The conventional receiver is optimal for certain applications. The general block conventional DS-CDMA receiver is as shown in Figure 7.3 However; it has few basic problems as given below:

- If all users transmit with approximately the same power, then beyond some point, increasing the transmitted power of every user will not decrease the Bit Error Rate (BER).

- If the users transmit with widely different powers, the conventional receiver allows the signal from powerful user by suppressing signal of weaker user, referred to as the ‘near-far effect’.

- The detection of the desired signal is limited by the inherent interference suppression capacity of the system.

![Figure 7.3 Conventional DS-CDMA Receiver](image)
7.4 MULTI CARRIER DS-CDMA

Corazza et al (2004) discussed the theoretical aspects of DS-CDMA code acquisition in the presence of correlated fading. With the substantial increase of internet users and with the development of new services, high-speed access in the future generations of wireless systems is an important requirement. Consequently, broadband systems with bandwidths much wider than that of the 3rd-generation systems are required for meeting future requirements.

The multiple-access scheme is constituted by Frequency-Hopping (FH) based multicarrier DS-CDMA (FH/MC DS-CDMA), where the entire bandwidth of future systems can be divided into a number of sub-bands and each sub-band can be assigned a sub-carrier.

According to the existing service requirements, the set of legitimate sub-carriers can be distributed in line with the instantaneous information rate requirements. FH techniques are employed for each user, in order to occupy the whole system bandwidth and to efficiently utilize the system's frequency resources. Specifically, slow FH, fast FH or adaptive FH techniques can be utilized depending on the system's design and the state-of-the-art. In FH/MC DS-CDMA systems the sub-bands are not required to be of equal bandwidth. Hence existing 2nd- and 3rd-generation CDMA systems can be supported using one or more sub-carriers, consequently simplifying the frequency resource management and efficiently utilizing the entire bandwidth available.
7.5 NONCOHERENT AND DIFFERENTIALLY COHERENT DETECTION

7.5.1 Non Coherent Code Detection

Consider a pair of orthogonal basis functions that span the 2D signal space as given by

\[ \phi_1(t) = \sqrt{\frac{2}{T}} \cos(2\pi ft)\text{rect}(t / T) \]  
\[ \phi_2(t) = \sqrt{\frac{2}{T}} \cos(2\pi ft)\text{rect}(t / T) \] 

Such that

\[ S_i(t) = S_{i1}(t) + S_{i2}(t) \]  

In a non-coherent receiver, no assumption of the carrier phase is made by the receiver. While, communication is possible without knowing the carrier phase, it will be shown in this section that there is penalty in the required \( \frac{E_b}{N_0} \) in comparison to coherent receivers which are given below:

1. Tracking the incoming signal's carrier frequency and synchronizing the receiver to it requires additional hardware which usually involves analog Components which are relatively expensive. Also the Digital Signal Processing (DSP) portion of the carrier tracking loop will consume power. However, since DSP Application Specific Integrated Circuits (ASICs) are becoming very power efficient, this is less of a concern.
2. No information about the content of career signal at the receiver.

3. It is very difficult to track the carrier phase. An example can be cited as in the case of IS95 CDMA reverse link, many simultaneous superimposed signals from many unsynchronized sources, exist.

4. There are many cases where the phase fluctuates randomly in such a manner that it is not practical to track it.

Figure 7.4 shows the architecture of non coherent receiver. It consist of matched filter and post detection processing unit. In the non-coherent receiver two locally generated basis functions that span the 2D signal space as given in equation 7.5 and 7.6 are assumed.
\[
\phi_{r_2}(t) = \sqrt{\frac{2}{T}} \cos(2\pi f t) \text{rect}(t / T) \quad (7.5)
\]

\[
\phi_{r_2}(t) = \sqrt{\frac{2}{T}} \cos(2\pi f t) \text{rect}(t / T) \quad (7.6)
\]

Where \( \theta(t) \) is a random phase that is not known or tracked by the receiver. The subscript ‘\( r_i \)’ denotes the receiver basis functions with \( i=1,2,3,\ldots \) which are rotated in phase relative to the transmitter basis functions. Note that this random phase can even be a function of time representing some uncertainty in the carrier frequency also. Following assumptions are made for carrying out the analysis.

1. The fluctuations in \( \theta(t) \) are insignificant relative to the symbol period \( T \).

2. The receiver synchronizer does track the symbol periods accurately such that matched filter sampling is assumed to be at the appropriate instance in time relative to the transmitted signal. This is a very reasonable assumption unless the SNR is really poor such that the receiver loses synchronization.

This architecture is similar to both the non-coherent and the coherent receiver types. The difference is that coherent receivers track and lock onto the carrier phase of the incoming signal, \( x(t) \), such that the locally generated basis functions are synchronized in phase. A non-coherent receiver makes no attempt to synchronize the basis functions to the incoming phase of \( x(t) \). Non-coherent receivers basically fall into three categories:

1. Energy detector determines if there is a signal present or not?

OOK (On Off Keying)
2. Determines basis functions and alphabet signals are orthogonal are not, irrespective of carrier phase.

3. Differential non coherent receivers estimate the carrier phase of present symbol pulse and use it to decode the difference in carrier phase relative to the next symbol pulse.

7.5.2 Correct detection and false alarm probability of non coherent code Acquisition Scheme

For comparison, the NC counterpart of the previously described DC scheme is characterized here, where the final decision variable of the \( l^{th} \) path is given by

\[
Z_{k(l)}^{\text{DC}} = \sum_{m=1}^{M} \sum_{n=1}^{N} \frac{1}{\sqrt{2}} \left( \sqrt{\frac{4E_s}{N_0 P}} S_k(l, m, n) + I_k(l, m, n) \right)^2
\]

(7.7)

where \( \| \cdot \| \) represents the Euclidian norm of the complex valued argument and the factor of \( \frac{1}{\sqrt{2}} \) is employed to normalize the noise variance. The NC decision variable \( z_{k(l), l}^{\text{NC}} \) has exactly the same statistical behavior as \( X_{k(l)} \).

Described \( S_{k(l, m, n)} \) becomes deterministic while \( I_{k(l, m, n)} \) is the complex valued Additive White Gaussian Noise (AWGN) having zero means and variances of \( \sigma^2=2 \) for both their real and imaginary parts. Finally, the probability of correct detection corresponding to \( x = 1 \) for the \( l^{th} \) path is obtained as

\[
\frac{D(l)}{\text{NC}} = e^{-\frac{1}{N_0 P} \sum_{k=0}^{MNL} (\theta / \mu_0)^k} \frac{1}{k!}
\]

(7.8)
Where \( k \) = False locking penalty factor.

\[ \theta = \text{threshold value} \]

While the false alarm probability in the context of a \( H_0 \) hypothesis is expressed as

\[
P_{\text{NC}} = e^{-\mu_0} \sum_{k=0}^{\Omega} \left( \frac{\theta}{\mu_0} \right)^k
\]

Where

\[
\mu_x = \left( 2 + \frac{1}{\lambda_x} \right) \text{ and } \lambda_x \text{ is either } \frac{2N}{P} \left( \frac{E_c}{I_0} \right) \text{ for the hypothesis of the desired signal being present } (x = 1) \text{ or } \frac{2N}{P} \left( \frac{E_c}{I_0} \right) \text{ for it being absent } (x = 0). \]

\( P_f = \text{False alarm probability} \)

### 7.5.3 Differentially Coherent (DC) Detection

Zamabizadeh et al (1997) discussed a differential coherent PN code acquisition CDMA receiver. Ikai et al (1999) discussed a new acquisition scheme of DS/SS signal with transmitter and receiver diversity. Shin et al (2003) presented a differentially coherent combining for doubled well code acquisition in DS-CDMA systems. A differentially coherent detection scheme employed in MF-based PN code acquisition has been proposed and well investigated previously by Jia-Chin Lin (2002). In this section, the Full-Period Cross correlation (FPC) model will be further simplified to be deterministic for certain types of PN sequences. The “ideal” shape of their autocorrelation functions with side lobes, the same as those of
maximum length shift registration sequences, will be assumed. An MF-based and a fixed-sample-size PN code acquisition technique with differentially coherent detection is investigated by means of statistical analyze. It is verified that the proposed techniques can provide lower error probabilities and more rapid acquisition.

In conventional DC combining, correlations on two consecutive portions of the incoming signal are evaluated and the decision variable is obtained by taking the real part of the product of these two correlations as shown in the Figure 7.5. In this, the phase of the second correlation is used to compensate the phase of the first one. There will be low noise amplification possible due to the fact that noise terms are independent. Differential combining is effective as long as the hypothesis of constant phase on the two subsequent correlations holds; degradations are expected in the presence of a time-varying phase.
7.6 SYSTEM DESCRIPTION

For system description the following assumptions are made:

1. A finite-length tapped delay line channel model generates ‘L’ Rayleigh-faded multi-path signals, each arriving with a time delay $\tau_i$ having a tap spacing of one chip duration where $l = L$ is the number of multi-path components.

2. The Rayleigh fading is sufficiently slow for the faded envelope to remain constant over ‘$\tau_D$’ chip intervals, but fast enough so that $\tau_D$ consecutive chip segments may be
considered essentially independently faded, where ‘\( \tau_D \)’ is the coherent integration interval that is used.

3. The Neyman-Pearson criterion is adopted, which leads to a Constant False-Alarm Rate (CFAR). This is because as a benefit of normalizing the received signal by the background noise variance estimate, threading channel’s attenuation no longer affects the outcome of the hypothesis test, regardless whether the desired signal is present or absent. The resultant scenario and the related test become reminiscent of an AWGN scenario. Consequently, in line with the findings of the mobile channel only affects the correct detection probability.

The received signal of the MIMO-aided DS-CDMA DL over the multi-path Rayleigh fading channel considered may be expressed as

\[
r(t) = \sum_{l=1}^{L} \sum_{m=1}^{M} \sum_{n=1}^{N} \alpha_{(l,m,n)} \sqrt{\frac{E_c}{P_t}} C(t + dT_c + \eta) \cos(t + dT_c + \eta)e^{j2\pi f_0(t + dT_c + \eta)} + I_{k(l,m,n)}(t)\]

(7.10)

Where

- \( M \) = Number of transmit antennas,
- \( N \) = Number of receive antenna
- \( \alpha_{(l,m,n)} \) = Complex-valued envelope of the \((l,m, n)^{th}\) signal path obeying a Rayleigh magnitude distribution and a uniform phase distribution,
- \( E_c \) = Pilot signal energy per PN code chip,
- \( C(t) \) = Common PN sequence having a cell-specific code-phase offset and is the code phase offset with respect to the phase of the local code,
- \( T_c \) = Chip duration,
\[ w_m(t) = \text{Specific Walsh code assigned to the } m^{th} \text{ transmit antenna}, \]
\[ f = \text{Carrier frequency} \]
\[ \varphi = \text{Carrier phase of a specific user’s modulator.} \]

\[ I_{k(l,m,n)}(t) \] is the complex valued AWGN having a double-sided power spectral density of \( I_0 \) at the \((l,m,n)^{th}\) path. Here the total allocated power is equally shared by the ‘\( M \)’ transmit antennas. Figure 7.6 shows 'N' receiver structure employing both the DC and the NC receiver’s schematic designed for proposed code acquisition scheme, using

**Figure 7.6 MIMO system**
transmit/receive antennas, where the timing hypothesis test is carried out for binary spreading. The NC module generates its decision variable by accumulating M·N number of independently faded signals observed over a given time interval. In the DC scheme, instead of squaring the summed energy as suggested by the procedures outlined, the channel’s output samples accumulated over a full spreading code period are multiplied by the conjugate of the $N_c$-chip-delayed samples.

Yang and Hanzo (2001), have analyzed that analyzes the Mean Acquisition Time (MAT) performance of the serial search acquisition methods with multiple timing states. When analyzing the MAT performance of the Single Dwell Serial Search (SDSS) technique, where the NC receiver structure of Figure 7.6 is used in the search mode constituting the SDSS scheme, the decision statistics, $Z_{tot}$ generated by the NC module are compared to the threshold $\theta_1$, which is optimized for a specific $Ec/I_0$. This completes the single-step search-mode of the SDSS scheme. By contrast, the Double Dwell Serial Search (DDSS) technique of Figure 7.7 calls for a two-step process. When the desired user’s tentative code phase was found in the search mode of the DDSS scheme, the verification mode is immediately activated. The verification mode may use either the DC or the NC modules, in order to confirm that the correct code-phase is indeed the one identified in the search mode. Only the NC scheme is used in search mode.
Received Signal

![Flowchart of Receiver Flow Chart](image)

**Figure 7.7 Receiver flow chart**
The DC scheme is excluded from the search mode, because it requires further processing carried out within the DC module and hence the complexity may be minimized by limiting the employment of the DC scheme to the verification mode.

More explicitly, the NC scheme is employed in two consecutive decision processes, namely first in finding and then confirming the correct code phase in order to improve the reliability of SDSS, which results in the DDSS acquisition scheme. \( \theta_1 \) and \( \theta_2 \) represent the acquisition thresholds of the search and verification mode, respectively. Furthermore, \( Z_1 \) and \( Z_2 \) denote the decision variables of the search and verification mode, respectively.

\( Z_1 \) is compared to \( \theta_1 \) and if it exceeds the threshold and \( Z_2 \) generated by either DC or NC module is compared to \( \theta_2 \). If successful code acquisition is confirmed, then the code tracking loop is enabled. Otherwise, the acquisition system goes back to the search-stage, until the correct code and its phase are found. Specifically, SDSS employing both DC and NC schemes as well as DDSS exploiting both DC and NC arrangements are invoked in the verification mode.

7.7 ANALYSIS OF DIFFERENTIALLY COHERENT AND NON-COHERENT CODE ACQUISITION

A decision variable is generated by accumulating M.N number of independently faded received signals observed over a time interval for improving the correct detection probability in the mobile channel imposing both fading and poor Signal to Interference Noise Ratio (SINR) conditions. Here formulating the final decision variable is neglected, which is readily derived from the procedures proposed in the context of the DC receiver structure. The final DC decision variable of the \( l^{th} \) path may be written as
\[
Z_{\text{DC}}^{(k)} = \sum_{m=1}^{M} \sum_{n=1}^{N} \left( \sqrt[4]{\frac{4E}{N^2}} S_{k(m,n)} + W_{1,k(l,m,n)} \right)^2 + W_{2,k(l,m,n)}^2 + \sum_{m=1}^{M} \sum_{n=1}^{N} \left( W_{2,k(l,m,n)}^2 + W_{4,k(l,m,n)}^2 \right)
\]  

(7.11)

where \( k = k^{th} \) chip’s sampling instant \( S_{k(l,m,n)} \) = a deterministic value, which depends on whether a signal is present or absent. Furthermore, the definition of \( W_{1,k(l,m,n)}, W_{2,k(l,m,n)}, W_{3,k(l,m,n)} \) and \( W_{4,k(l,m,n)} \) are mutually independent Gaussian random variables having zero means and unit variances. Rewriting equation 7.12 in short form,

\[
X_{k(l)} = \sum_{m=1}^{M} \sum_{n=1}^{N} \left( \sqrt[4]{\frac{4E}{N^2}} S_{k(m,n)} + W_{1,k(l,m,n)} \right)^2 + W_{2,k(l,m,n)}^2 \]  

(7.12a)

and

\[
Y_{k(l)} = \sum_{m=1}^{M} \sum_{n=1}^{N} \left( W_{2,k(l,m,n)}^2 + W_{4,k(l,m,n)}^2 \right) \]  

(7.12b)

Substituting and rewriting, the final decision variable of Eq (7.12) is obtained as

\[
Z_{\text{DC}}^{(k)} = X_{k(l)} - Y_{k(l)} = \sum_{m=1}^{M} \sum_{n=1}^{N} X_{k(l,m,n)} - \sum_{m=1}^{M} \sum_{n=1}^{N} Y_{k(l,m,n)}
\]  

(7.13)

where \( X_{k(l)} \) obeys a non central chi-square Probability Density Function (PDF) with \( 2MN \) degrees of freedom and its non centrality parameter \( \lambda_x \) is either \( \frac{4N}{P} \left( \frac{E}{I_0} \right) \), when the desired signal is deemed to be present \( (x = 1) \) or \( \frac{4N}{P} \left( \frac{E}{I_0} \right) \), when it is deemed to be absent \( (x = 0) \).
effects of both timing errors and the total frequency mismatches are covered by the definition of \( \frac{E_c}{I_0} \). \( \frac{E_c}{I_0} \) is defined mathematically (ie),
\[
\frac{E_c}{I_0} = \frac{E_c}{I_0} \cdot \sin^2 \left( \frac{\tau}{T_c} \right) \cdot \sin^2 \left( N \Delta f T_c \right),
\]
where the second term of the definition is the square of the autocorrelation function imposed on the timing error \( \tau \), the third term of the definition is the signal energy reduction expressed as a function of the total frequency mismatch, \( \Delta f \), after the squaring operation and \( 'N' \) represents the number of chips accumulated over the duration of \( \tau_p \). Finally, \( Y_{k(l)} \) is centrally chi-square distributed with 2MN degrees of freedom. It is also worth noting that the outputs of the squaring operation invoked both the in-phase and the quadrature branches are modeled as squares of Gaussian random variables, respectively.

Accordingly, the decision variable \( X_{k(l,m,n)} \) of each path, obeys a non-central chi-square Probability Density Functions (PDF) with two degrees of freedom, but \( Y_{k(l,m,n)} \) is centrally chi-square distributed with two degrees of freedom.

PDFs are given below:

\[
f_{X_{k(l,m,n)}}(z|H_s) = \frac{1}{2} e^{-\left(z^2 + x^2\hat{x}^2\right) / 2} J_0(z\hat{x})
\]

(7.14)

and

\[
f_{Y_{k(l,m,n)}}(z|H_s) = \frac{1}{2} e^{-\left(z^2 \right) / 2}
\]

(7.15)

These respectively, where \( z \geq 0 \), \( x = \{0\} \) or \( \{1\} \), \( I_0(.) \) is the Zero order modified Bessel function of the first kind. The PDF of the desired
user’s signal at the output of the acquisition scheme is conditioned on the presence of the desired signal in \( f_{x_k(l,m,n)}(z | H_x) \), when communicating over an uncorrelated Rayleigh channel. In this case \( E_c \) is multiplied by the square of the Rayleigh distributed fading amplitude, \( \beta \), which has a chi-square distribution with two degrees of freedom: \( f(\beta) = \frac{e^{-\beta/2\sigma^2}}{\sigma^2} \), where \( \sigma^2 \) is the variance of the constituent Gaussian distribution. Then the average pilot signal energy 'Ec' per PN code chip can be expressed as \( \overline{E_c} = \beta E_c = \sigma^2 E_c \).

Therefore first the PDF \( f_{x_k(l,m,n)}(z | H_x, \beta) \) corresponding to \( \beta \) conditioned on the hypothesis of the desired signal being transmitted over an Additive White Gaussian Noise channel having this specific SINR is weighted by the probability of occurrence \( f(\beta) \) of encountering \( \beta \), as quantified by the PDF. The resultant product is then averaged over its legitimate range of \(-\infty \) to \( \sim \infty \), yielding:

\[
\begin{align*}
  f_{x_k(l,m,n)}(z | H_x) &= \int_{-\infty}^{\infty} f(\beta).f_{x_k(l,m,n)}(z | H_x, \beta)d\beta \\
  &= \int_{0}^{\infty} \left( \frac{e^{-\beta/2\sigma^2}}{\sigma^2} \right) \left( \frac{e^{4z^2 \beta^2}}{2} \right) I_0\left( \sqrt{\beta \lambda_z z} \right) d\beta \\
  &= \frac{e^{-\beta(2+\lambda_z \sigma^2)}}{\left(2 + \lambda_z \sigma^2\right)} \\
  &= \frac{e^{-z(2+\lambda_z \sigma^2)}}{\left(2 + \lambda_z \sigma^2\right)}
\end{align*}
\]

where the corresponding non centrality parameter of \( \lambda = \lambda_z \sigma^2 \) is either

\[
\frac{4N}{P} \left( \frac{E_c}{I_0} \right)
\]

when the desired signal is deemed to be present \( (x = 1) \) or

\[
\frac{4N}{P} \left( \frac{E_c}{I_0} \right)
\]
when it is deemed to be absent \((x = 0)\). Similarly to the definition of \((E_e / I_o)\), \((E_e / I_o)\) defined as \(\left(\frac{E_e}{I_o}\right) = (E_e / I_o) \sin^2 \left(\frac{T}{T_c}\right) \sin^2 (N \Delta f T_c)\). For notational convenience, a new biased non-centrality parameter \(\mu_x = \left(2 + \bar{\lambda}_x\right)\) is defined. Finally, the PDF of \(X_{k(l,m,n)}\) conditioned on the presence of the desired signal is in the form of:

\[ f_{X_{k(l,m,n)}}(z \mid H_x) = \frac{1}{\mu_x} e^{-\frac{z}{\mu_x}} \] (7.19)

The decision variables, \(X_{k(l)}\) and \(Y_{k(l)}\) are constituted by the sum of M.N number of independent variables \(X_{k(l)} = \sum_{m=1}^{M} \sum_{n=1}^{N} X_{k(l,m,n)}\) and \(Y_{k(l)} = \sum_{m=1}^{M} \sum_{n=1}^{N} Y_{k(l,m,n)}\).

Both decision variables constitute independent Gamma variables, leading to:

\[ f_{X_{k(l)}}(z \mid H_x) = \frac{z^{(M.N-1)} e^{-\frac{z}{\mu_x}}}{\Gamma(M.N) \mu_x^{MN}} \] (7.20)

\[ f_{Y_{k(l)}}(z \mid H_x) = \frac{z^{(M.N-1)} e^{-\frac{z}{2}}}{\Gamma(M.N) 2^{MN}} \] (7.21)

where \(\Gamma(\cdot)\) is the Gamma function.

The probability of correct detection for the \(l^{th}\) path according to \(x = 1\), is expressed as:

\[ P_{D(l)}^{DC} = \int_{0}^{\infty} f_{X_{k(l)}}(z \mid H_x) dz, \hat{\theta} \neq 0 \] (7.22)

where \(\hat{\theta}\) = threshold value.

\(P_{D(l)}^{DC}\) = probability of correct detection for the \(l^{th}\) path.
The false alarm probability in the context of a $H_0$ hypothesis is expressed as

$$P_{F}^{DC} = \int_0^\infty f_{z_k|z}(z|H_0)dz, \theta \neq 0$$

(7.23)

Where

$$P_{F}^{DC} = \text{the false alarm probability.}$$

### 7.8 MEAN ACQUISITION TIME (MAT) ANALYSIS

The MAT is expressed with SI unit ‘seconds’. MAT formulas were provided for as single-antenna-aided serial search based code acquisition system. There is no distinction between a single-antenna-aided scheme and a multiple-antenna assisted one in terms of analyzing their MAT performance, except for deriving their correct detection and the false alarm probability based upon using number of transmit/receive antennas. MAT performance of both DC and acquisition schemes, employed in SDSS and DDSS are analyzed. Assume that in each chip duration ‘$T_c$’, ‘$\alpha$’ number of timing Hypotheses are tested, which are spaced by $T_c/\alpha$ is assumed. Hence the total uncertainty region is increased by a factor of ‘$\alpha$’. Moreover, when the ‘$L$’ multi-path signals arrive with time delays ‘$\tau_l$’ having a tap spacing of one chip duration, the relative frequency of the signal being present is increased $L$-fold. The required transfer functions are defined as follows:

The entire successful detection function $H_d(Z)$ encompasses all the branches of a state diagram, which lead to successful detection. Furthermore, $H_d(Z)$ indicates the absence of the desired user’s signal at the output of the acquisition scheme, while $H_m(Z)$ represents the overall miss
probability of a search run carried out across the entire uncertainty region. The related processes are detailed for SDSS and for DDSS. Then, it may be shown that the generalized expression derived for computing the MAT of the serial search based code acquisition scheme is given by

$$E[T_{ACQ}] = \frac{1}{H_D(1)} \left[ H_D'(1) + H_M'(1) + \left( v - 2\gamma L \right) \left( 1 - \frac{H_D'(1)}{2} \right) + \frac{1}{2} H_D(1) \right] . \tau_D$$

(7.24)

Where, $H_{\gamma(z)} \vert_{z=D}$, $M$ or $0$ is a derivative of $H_{\gamma(z)} \vert_{z=D}$, $M$ or $0$ and $\tau_D$ denotes either the dwell time three for the SDSS scenario or the dwell time of the search mode for the DDSS case. If the total number of states $\gamma$ is significantly higher than the number of $H_D$ states, the exact MAT formula of Eq (7.24) can be simplified as follows

$$E[T_{ACQ}] \approx \frac{(1 + H_M(1)) - H_D'(1))}{2(1 - H_M(1))) \cdot (v \cdot \tau_D)$$

(7.25)

Where, $E[T_{ACQ}]$ = MAT of the serial search based code acquisition scheme.

Since each resolvable path contributes two hypotheses and because the average correct detection probability associated with these two hypotheses is the same, the overall miss probabilities of both the SDSS and the DDSS schemes may be expressed as

$$H_M(1) = \prod_{i=1}^{L} \prod_{\zeta=1}^{\gamma}(1 - P_{D(i,\zeta)})^2$$

(7.26)

and
where \( PD (l, \zeta) \) represents the correct detection probability of the SDSS scheme and \( P_{D_{q=1}} |_{l-1} \), or 2 are the correct detection probability of both the search and the verification modes of the DDSS arrangements, respectively. The \( H_0'(l) \) values of the SDSS and DDSS schemes are expressed as

\[
H_0'(l) = 1 + K P_F 
\]  
(7.28)

and

\[
H_0'(l) = 1 + e P_{F_1} + K P_{F_1} P_{F_2}, 
\]  
(7.29)

where \( K \) denotes the false locking penalty factor expressed in terms of the number of chip intervals required by an auxiliary device for recognizing. The code-tracking loop is still unlocked and \( e \) represents the ratio defined as the dwell time for the verification mode over that for the search mode. ‘ \( P_F \) ’ is the false alarm probability of the SDSS scheme and \( P_{F_1} |_{l-1} \), represent the false alarm probability of both the search and the verification mode of the DDSS scheme, respectively.

### 7.9 SIMULATION RESULTS FOR THE PROPOSED SPREAD SPECTRUM MIMO RECEIVERS

Table 7.1 summarizes the simulation parameter used as input. Simulation is carried out using Matlab Communication tool box version 7 by varying the number of transmission path.
Table 7.1 Simulation parameters for proposed spread spectrum MIMO receivers

<table>
<thead>
<tr>
<th>Sl.No</th>
<th>System Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Band width</td>
<td>1.22 MHZs</td>
</tr>
<tr>
<td>2</td>
<td>Carrier frequency</td>
<td>1.89 GHz</td>
</tr>
<tr>
<td>3</td>
<td>Diversity</td>
<td>Both transmitter and receiver.</td>
</tr>
<tr>
<td>4</td>
<td>Clock drift</td>
<td>1000 HZs</td>
</tr>
<tr>
<td>5</td>
<td>Mobile speed</td>
<td>100 Km/Hr</td>
</tr>
<tr>
<td>6</td>
<td>No of Chips (SDSS)</td>
<td>256 Chips</td>
</tr>
<tr>
<td>7</td>
<td>No of Chips (DDSS)</td>
<td>256 Chips</td>
</tr>
<tr>
<td>8</td>
<td>Spreading factor of Walsh code</td>
<td>128</td>
</tr>
<tr>
<td>9</td>
<td>False Locking penalty factor</td>
<td>1000</td>
</tr>
</tbody>
</table>

ANALYSIS-1

Simulation analysis 1 deals with the SISO system employing full chip and half chip for DC and NC schemes. The performance curves in Figure 7.8 show the bit error rate of the Single Input Single Output (SISO) system using the full-chip and half chip with respect to the SINR (E_c/N_0) ratio. From the simulation result, it is observed that differentially coherent code acquisition is better method than the non-coherent code acquisition.
Figure 7.8 (a-c) Bit Error Rates (BER) of SISO (full-chip and half Chip)
In this SISO (full-chip and half) receiver, single path is considered and BPSK modulation is used. Figure 7.8 shows that the BER of both the differentially coherent code acquisition and non-coherent code acquisition for full chip and half chip rate system. From Figure 7.8 (a) at $E_c/N_o = 2.5$ db, DC shows a BER of $10^{-2.8}$ and NC scheme shows a BER of $10^{-1.5}$. Figure 7.5 (b) at $E_c/N_o = 2.5$ db, DC shows a BER of $10^{-3.0}$ and NC scheme shows a BER of $10^{-2.0}$. Hence the accuracy has been improved by incorporating half-chip version of the proposed model. From Table 7.2 following observation is made. Full chip DC scheme offers BER of $10^{-2.8}$ and NC scheme offers $10^{-1.5}$, hence DC scheme is better than NC Scheme. half chip DC scheme offers BER of $10^{-3.0}$ and NC scheme offers $10^{-2.0}$, hence DC scheme is better than NC Scheme. It is inferred that Performance of DC scheme is better than NC scheme.

Table 7.2 Performance comparison of full chip and half chip SISO system

<table>
<thead>
<tr>
<th>System</th>
<th>DC</th>
<th>NC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full chip</td>
<td>$10^{-2.8}$</td>
<td>$10^{-1.5}$</td>
</tr>
<tr>
<td>Half Chip</td>
<td>$10^{-3.0}$</td>
<td>$10^{-2.0}$</td>
</tr>
</tbody>
</table>
ANALYSIS-2:

a) $M^*N (1, 4) = 4$ paths  
b) $M^*N (2, 4) = 8$ paths  
c) $M^*N (4, 4) = 16$ paths  
d) $M^*N (8, 4) = 32$ paths

Figure 7.9 (a-d) BER Characteristics of MIMO (full-chip) using BPSK modulation for various transmission path
Table 7.3  Performance comparison of full chip using BPSK-MIMO at 
$E_b/N_0 = -5$ dB

<table>
<thead>
<tr>
<th>Transmission Path</th>
<th>DC</th>
<th>NC</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>$10^{-1.2}$</td>
<td>$10^{-0.8}$</td>
</tr>
<tr>
<td>8</td>
<td>$10^{-2.1}$</td>
<td>$10^{-1.1}$</td>
</tr>
<tr>
<td>16</td>
<td>$10^{-0.01}$</td>
<td>$10^{-1.5}$</td>
</tr>
<tr>
<td>32</td>
<td>$10^{-0.01}$</td>
<td>$10^{-2.0}$</td>
</tr>
</tbody>
</table>

From Table 7.4, it is inferred that as the number of transmission path is increased the BER decreases for DC scheme and for the transmission path 16 and 32 , the BER approaches zero. The same comparison is in the form of bar graph in Figure 7.10

Figure 7.10  Bar chart for the performance comparison of full chip Using BPSK –MIMO
a). $M^*N (1, 4) = 4$ paths

b). $M^*N (2, 4) = 8$ paths

c). $M^*N (4, 4) = 16$ path

d). $M^*N (8, 4) = 32$ path

Figure 7.11 (a-d) BER of MIMO (half-chip)- using BPSK modulation for various transmission Path
Table 7.4 Performance comparison of full chip Using BPSK-MIMO Half Chip at $E_b/N_0 = -5$ dB

<table>
<thead>
<tr>
<th>Transmission Path</th>
<th>DC</th>
<th>NC</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>$10^{-1.5}$</td>
<td>$10^{-0.9}$</td>
</tr>
<tr>
<td>8</td>
<td>$10^{-0.01}$</td>
<td>$10^{-1.5}$</td>
</tr>
<tr>
<td>16</td>
<td>$10^{-0.01}$</td>
<td>$10^{-1.6}$</td>
</tr>
<tr>
<td>32</td>
<td>$10^{-0.01}$</td>
<td>$10^{-1.7}$</td>
</tr>
</tbody>
</table>

From Table 7.4 it is inferred that as the number of transmission path is increased, the BER decreases for DC scheme and for the transmission path 8, 16 and 32 , the BER approaches zero. The same comparison is in the form of bar graph in Figure 7.12.

Figure 7.12 Bar chart for the performance comparison of half chip Using BPSK -MIMO
Simulation analysis 2 deals with the MIMO system employing full chip and half chip for DC and NC schemes. From the Figures 7.9 (a-d) it is observed that the BER for both DC and NC code acquisition is decreasing considerably. Especially in the figures 7.9 (c-d) the BER for the DC code acquisition is approaching zero. This shows that DC is better than NC and the BER is reduced with increase in the number of transmission path. From the figures 7.11 (a-d) it is observed that the BER for both DC and NC code acquisition is decreasing considerably. Especially in the Figures 7.11 (b-d) the BER for the DC code acquisition is approaching zero. This shows that DC is better than NC and the BER is reduced with increase in the number of transmission path.

Simulation analysis 3 deals with Mean acquisition Time analysis for both DC and NC schemes. The Figures 7.13 (a-d) shows the variation of the Mean Acquisition Time (MAT) with respect to the SINR(E_c/n_0) ratio. From Figure 7.13(a-d) observation it is clear that the MAT decreases with increase of SINR ratio. This is one of the favorable results for this system. The MAT decreases with increase of the paths or the transmit antenna.
ANALYSIS-3

(a) \(M\times N (1, 4) = 4\) paths
(b) \(M\times N (2, 4) = 8\) paths
(c) \(M\times N (4, 4) = 16\) path
(d) \(M\times N (8, 4) = 32\) path

Figure 7.13 (a-d) Achievable MAT Versus SINR per chip Characteristics
7.10 CONCLUSION

This chapter discussed about the code acquisition methods in spread spectrum MIMO receiver. This chapter has investigated about the performance analysis and comparison of full chip and half chip rate of Noncoherent (NC) and Differentially Coherent (DC) code acquisition scheme in MIMO assisted by Direct Sequence spread spectrum (DS-CDMA) and a novel approach MIMO half chip rate performance is compared. Table 7.5 shows performance comparison of full chip and half chip rate – MIMO system.

Table 7.5 Performance comparison of full chip and half chip rate – MIMO system

<table>
<thead>
<tr>
<th>Transmission Path</th>
<th>Full Chip Rate</th>
<th>Half Chip Rate</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>DC</td>
<td>NC</td>
</tr>
<tr>
<td>4</td>
<td>$10^{-1.2}$</td>
<td>$10^{-0.8}$</td>
</tr>
<tr>
<td>8</td>
<td>$10^{-2.1}$</td>
<td>$10^{-1.1}$</td>
</tr>
<tr>
<td>16</td>
<td>$10^{-0.01}$</td>
<td>$10^{-1.5}$</td>
</tr>
<tr>
<td>32</td>
<td>$10^{-0.01}$</td>
<td>$10^{-2.0}$</td>
</tr>
</tbody>
</table>
The following conclusions are derived for the analysis carried out:

(i) As the number of paths increased from 4, 8, 16, and 32, the BER is decreased for both DC code acquisition and NC code acquisition for both the half and full chip rate schemes.

(ii) DC performance is better than NC in both half chip and full chip for both SISO and MIMO.

(iii) Half chip method is best suited in terms of BER when compared to full chip for both DC and NC schemes.

(iv) As the number of paths increases the BER is decreased for both half chip and full chip.

(v) From Figure 17.13 (a-d) it is inferred that the mean acquisition time of DC is less than NC code acquisition.

(vi) As SINR ratio is increased, the mean acquisition time is reduced considerably for both DC and NC code acquisition methods.

(vii) For 32 paths, the MAT of DC code acquisition is nearing zero stating that the mean acquisition time is reduced as the number of transmission paths increases.

It is inferred that proposed half chip MIMO system is superior in performance comparing to full chip rate system. Also, DC performance is better than NC in both half chip and full chip rate schemes.