<table>
<thead>
<tr>
<th>Section</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.1</td>
<td>Introduction</td>
<td>30</td>
</tr>
<tr>
<td>3.2</td>
<td>Nakagami-(m) Fading Channel</td>
<td>30</td>
</tr>
<tr>
<td>3.3</td>
<td>Performance of Basic Diversity Combining Schemes</td>
<td>31</td>
</tr>
<tr>
<td>3.3.1</td>
<td>SC Scheme with (m = 0.5)</td>
<td>33</td>
</tr>
<tr>
<td>3.3.2</td>
<td>MRC Scheme with (m = 0.5)</td>
<td>35</td>
</tr>
<tr>
<td>3.3.3</td>
<td>EGC Scheme with (m = 0.5)</td>
<td>36</td>
</tr>
<tr>
<td>3.3.4</td>
<td>Average BER of MPSK Scheme</td>
<td>37</td>
</tr>
<tr>
<td>3.4</td>
<td>Numerical Computational Results</td>
<td>40</td>
</tr>
<tr>
<td>3.5</td>
<td>MUD in Nakagami-(m) Fading Channel</td>
<td>50</td>
</tr>
<tr>
<td>3.5.1</td>
<td>Synchronous Case</td>
<td>50</td>
</tr>
<tr>
<td>3.5.2</td>
<td>Average BER of (M)-decorrelator</td>
<td>51</td>
</tr>
<tr>
<td>3.6</td>
<td>Simulation Results</td>
<td>52</td>
</tr>
<tr>
<td>3.7</td>
<td>Summary</td>
<td>63</td>
</tr>
</tbody>
</table>
Chapter 3

Multiuser Detection for DS-CDMA Systems over Nakagami-m Fading Channels

3.1 INTRODUCTION

CDMA transmissions are made over multipath fading channels and it is of interest to design receivers that take fading behavior of the channel into account. In this chapter, the problem of data detection in non-Gaussian impulsive noise channels with the proposed influence function is considered. MUD for DS-CDMA systems with MRC receive diversity over Nakagami-m fading channels in a non-Gaussian noise environment is also presented in this chapter. Performance of an M-estimation based decorrelating detector (M-decorrelator) that detects BPSK signals, is studied by deriving a closed-form expression for average BER over a single-path Nakagami-m fading channel with and without code mismatch.

In a cellular mobile communication system, a transmission channel is a propagation path over which radio signals travel from a BS to a MS (forward link) or from a MS to BS (reverse link) [60-62]. Typical cellular mobile communication channel can be a simple line-of-sight (LOS) transmission channel or, for a very complicated one, which may be blocked by different obstacles like trees, mountains, buildings, vehicles etc. Moreover, due to the relative motion of mobiles and other radio propagation media with respect to the BS, the received signals often exhibit rapid random fluctuations. Therefore, cellular mobile communication channels are represented using statistical models that are described widely in the literature [18, 19, 38, 61]. Multipath fading is the rapid and random variation of received signal’s power due to constructive and destructive addition of multipath signal components at the receiver front-end. This degrades the performance of any mobile wireless communication system. In this thesis, the Nakagami-m fading channel model is considered to study the performance of M-decorrelator.

3.2 NAKAGAMI-m FADING CHANNEL

The Nakagami-m multi-path fading channel model has been widely used in the literature for analyzing wireless mobile communication systems [27] and it received considerable attention as it can provide a good fit to measured data in different multi-
path fading environments [21] like Rayleigh, log-normal or Ricean fading channels. Nakagami-\( m \) fading is frequently considered in many practical communication systems such as cellular mobile communications systems [22]. Recently, [20] considered the Nakagami-0.5 fading channel model as a special case of the Nakagami-\( m \) channel (with fading severity parameter, \( m = 0.5 \)) and showed that the worst-case of Nakagami-\( m \) fading model is \( m = 0.5 \). The performance of wireless communication system on Nakagami-0.5 fading channels has importance, as it evaluates the system in a worst-case scenario and this analysis becomes more crucial when a high level of quality of service (QoS) is required [27]. The performance analysis of basic linear diversity combining techniques is considered in [20] where the average BER of BPSK modulation scheme with dual branch and \( D \)-branch diversity are derived. The calculation of average outage duration of MRC over independent and identically distributed (i.i.d) Rayleigh and Rice fading channels is studied in [63] by deriving closed-form expressions using the classical PDF-based approach and the characteristic function (CF)-based approach. The performance of a multi-hop amplify-and-forward relaying on Nakagami-0.5 fading channels is evaluated in [26] by deriving closed-form expressions for the outage probability (OP), the average symbol error probability, and the ergodic capacity.

3.3 PERFORMANCE OF BASIC DIVERSITY COMBINING SCHEMES

Diversity combining is a technique that is being used to combat the effects of fading. In diversity combining several copies of the transmitted signal, received by different antennas that are spatially separated, are combined to increase the SNR. The three diversity combining schemes, SC, EGC and MRC, are the basic diversity combining schemes that have been studied widely in the literature [18, 19, 20, 64].

System outages occur when the instantaneous quality of the wireless network measured in terms of a parameter of interest fails to achieve the desired target value. The OP is one of the important performance measures of wireless communication systems. The OP is defined as the event that the instantaneous received SNR, \( \gamma \), is below a predefined threshold, \( \gamma_{th} \). Therefore, the OP, \( P_{out} \), can be expressed as [26, 64-66]
\[ P_{\text{out}} = P(\gamma_i < \gamma_{\text{th}}) = \int_0^{\gamma_{\text{th}}} f_{\Lambda}(\gamma) d\gamma \]  \hspace{1cm} (3.1)

where \( P(\cdot) \) is the probability of an event and \( f_{\Lambda}(\cdot) \) is the PDF of random variable \( \Lambda \). Another parameter, outage duration (OD) is the average time the received SNR is below the threshold and is given by [65]

\[ T_o = \frac{P(\gamma < \gamma_{\text{th}})}{L_{\gamma_o}} = \frac{P_{\text{out}}}{L_{\gamma_o}} \]  \hspace{1cm} (3.2)

where \( L_{\gamma_o} \) is the level crossing rate (LCR) at which the instantaneous SNR crosses the pre-defined SNR-threshold.

In this section, closed-form expressions for OP and average OD (AOD) are derived to analyze the performance of basic receive-diversity combining schemes in the Nakagami-0.5 fading channels. This case has great practical importance as a worst-case fading scenario [20].

Consider a wireless radio communication system with \( D \)-branch receive-antenna diversity operating at SNR per bit \( \gamma \) in every branch. The instantaneous SNR is related to the received signal envelope as \( \gamma_i = (E_b/N_o)X_i^2 \) \((i = 1, 2, \ldots, D)\) where \( E_b \) is the total transmitted signal energy per bit in all the receive-diversity branches and \( N_o/2 \) is the ambient noise power spectral density. Assuming that the received signal envelope \( X \) is Nakagami-\( m \) distributed, the PDF is given by [38]

\[ f_X(x) = 2 \left( \frac{m}{\Omega} \right)^{mD} \frac{x^{2mD-1}}{\Gamma(mD)} \exp\left( -\frac{m}{\Omega} x^2 \right) \]  \hspace{1cm} (3.3)

where \( m \) is the Nakagami fading parameter that determines the severity of the fading, \( \Gamma(\cdot) \) is the gamma function [67, 68] and \( \Omega = E[X^2] \) is the total average multipath received signal power for a single channel. In a wireless mobile communication channel, the exponentially decaying multipath intensity profile (MIP) follows the relation [38, 68]

\[ \Omega = \Omega_o \exp(-\delta) \]  \hspace{1cm} (3.4)

where \( \Omega_o \) is the initial path strength of a channel and \( \delta \) is the power decay factor. The PDF of \( X \) is for different values of \( m \) (with \( \Omega = 1 \)) is shown in Fig: 3.1.
Figure 3.1 The Nakagami distribution for different values of fading parameter $m$ with $\Omega = 1$.

To study the performance of the system under worst-case fading scenario (with $m = 0.5$), it is assumed that the received signal envelope $X$ is Nakagami-$m$ (with $m = 0.5$) distributed, the PDF and the cumulative distribution function (CDF) of $X$ are given, respectively, by [20]

$$f_X(x) = \frac{2}{\sqrt{2\pi\Omega}} \exp\left(-\frac{x^2}{2\Omega}\right), \quad x \geq 0$$  \hspace{1cm} (3.5)

and

$$F_X(x) = \int_0^x f(\xi) d\xi = \text{erf}\left(\frac{x}{\sqrt{2\Omega}}\right), \quad x \geq 0$$  \hspace{1cm} (3.6)

where $\text{erf}(z) = \frac{2}{\sqrt{\pi}} \int_0^z e^{-t^2} dt$ is the error function.

### 3.3.1 SC Scheme with $m = 0.5$

The SC receive-diversity system consists of $D$ receive-antennas and the combiner only selects the particular receive-antenna’s output with the largest SNR as the received signal [19]. That is,
\[ X = \max \{ X_1, \ldots, X_D \} \quad (3.7) \]

where \( X_i (i = 1, \ldots, D) \) are assumed to be i.i.d (i.e., Nakagami-0.5 distributed). The PDF of \( X \) for \( D \)-branch SC receive-diversity is given by [20]

\[
f_X^{SC}(x) = \frac{2De^{x^2/2\Omega}}{\sqrt{2\pi\Omega}} \left[ \text{erf} \left( \frac{x}{\sqrt{2\Omega}} \right) \right]^{D-1}, \ x \geq 0. \quad (3.8)
\]

Substituting (3.8) in (3.1) and using the result [67]

\[
\int_0^\infty \left[ \text{erf} \left( \frac{x}{\sqrt{2a}} \right) \right]^{D-1} e^{-\frac{x^2}{2a}} \, dx = \frac{1}{D} \left( \frac{\pi}{2} \right)^{\frac{D}{2}} \left[ \text{erf} \left( \frac{\lambda}{\sqrt{2a}} \right) \right]^D
\]

the OP for \( D \)-branch SC receive-diversity scheme can be derived as

\[
P_o^{SC} = \left[ \text{erf} \left( \frac{x_{th}}{\sqrt{2\Omega}} \right) \right]^{D} \quad (3.10)
\]

where \( x_{th} \) is the minimum required threshold value of received signal strength at the output of receive-diversity combiner. By using the relation [63]

\[
\gamma_{th} = \left( \frac{E_b}{N_o} \right) x_{th}^2 \quad (3.11)
\]

and defining the ratio \( \mu = \frac{\gamma_{th}}{\gamma} \) with \( \gamma = (E_b/N_o)\Omega \), (3.10) can be written as

\[
P_o^{SC} = \left[ \text{erf} \left( \frac{\sqrt{\gamma_{th}/E_b}}{\sqrt{2\Omega}} \right) \right]^{D} = \left[ \text{erf} \left( \frac{\mu}{\sqrt{2}} \right) \right]^{D}.
\]

Let \( L_{A_x} \) be the rate at which the received signal envelope \( x \) crosses the level \( A_x \) in the positive-going direction. The parameter \( L_{A_x} \) depends on the rate \( \dot{x} \) at which the level \( x \) is crossed and the probability that \( x = A_x \) is given in terms of the joint PDF as [65]

\[
L_{x_{\tau}} = \int_0^\infty \dot{x} f_{x,\dot{x}}(x_{th}, \dot{x}) \, d\dot{x} = \frac{\dot{\sigma}}{\sqrt{2\pi}} f(x_{th})
\]

where \( \dot{x} \) is the time derivative of \( x \), \( \dot{\sigma} = \sqrt{2\pi f_m} \sigma \) with \( \sigma^2 = \Omega \), \( A = x_{th} \) and \( f_m \) is the maximum Doppler frequency. Using (3.8) and (3.13), the LCR of the SC receive-diversity system can be derived as
\[ L_{\gamma_o}^{SC} = \sqrt{2} f_m D \left[ \text{erf} \left( \frac{\gamma_{lh}}{2\Omega} \frac{E_b}{N_o} \right) \right]^{D-1} \exp \left( -\frac{\gamma_{lh}}{2\Omega} \frac{E_b}{N_o} \right) \]. \tag{3.14}

Therefore, using (3.2) and (3.14), the OD of selection diversity combiner can be expressed as \( T_o^{SC} = P_o^{SC} / L_{\gamma_o}^{SC} \) and the AOD is

\[ \overline{T}_o^{SC} = f_m T_o^{SC} = \frac{1}{\sqrt{2D}} \text{erf} \left( \frac{\gamma_{lh}}{2\Omega} \frac{E_b}{N_o} \right) \exp \left( \frac{\gamma_{lh}}{2\Omega} \frac{E_b}{N_o} \right) \]. \tag{3.15}

### 3.3.2 MRC Scheme with \( m = 0.5 \)

In MRC system, the output of the diversity combiner is a weighted sum of all receive-diversity branch signals. All these received signals are co-phased and the receive-diversity combiner output is given by \([18, 19, 20, 64, 66]\)

\[ X = \sum_{i=1}^{L} X_i^2. \tag{3.16} \]

The PDF of the received signal envelope \( X \) in MRC system is given by \([20]\)

\[ f_X^{MRC}(x) = \frac{2}{\Gamma \left( \frac{D}{2} \right)} \left( \frac{1}{2\Omega} \right)^{D/2} x^{D-1} \exp \left( -\frac{x^2}{2\Omega} \right), \quad x \geq 0. \tag{3.17} \]

Using (3.1), (3.11), (3.17) and \([68, (3.381.1)]\) the OP of MRC can be derived as

\[ P_o^{MRC} = \frac{1}{\Gamma \left( \frac{D}{2} \right)} \Gamma \left( \frac{D}{2} - \frac{\gamma_{lh}}{2\Omega} \frac{E_b}{N_o} \right) = \frac{\Gamma \left( \frac{D}{2}, \frac{\mu}{2} \right)}{\Gamma \left( \frac{D}{2} \right)} \]. \tag{3.18}

where \( \Gamma(\cdot, \cdot) \) is the incomplete gamma function \([68, 69]\). Now, using (3.13) and (3.17), the LCR of MRC scheme can be expressed as
\[
\begin{align*}
I_{o}^{MRC}^{\gamma_{a}} &= \frac{2f_{m}\sqrt{\pi\Omega}}{\Gamma(D/2)} \left(\frac{1}{2\Omega}\right)^{D/2} e^{-\frac{\gamma_{a}}{2\Omega N_{o}}} \left(\frac{\theta_{h} / E_{b}}{N_{o}}\right)^{-D-1} \\
&= \frac{2\sqrt{2\pi}f_{m}}{\Gamma(D/2)} \left(\frac{\mu}{2}\right)^{D-1} \exp\left(-\frac{\mu}{2}\right) \\
\end{align*}
\]

Then, combining (3.19) with (3.18) and using (3.2), the AOD of MRC system can be derived as

\[
\bar{T}_{o}^{MRC} = f_{m}T_{o}^{MRC} = \frac{\gamma(D/2) \theta_{h} / E_{b}}{2\sqrt{2\Omega} \left(\frac{1}{2\Omega}\right)^{D/2} \left(\frac{\theta_{h} / E_{b}}{N_{o}}\right)^{-D-1}} \left[\exp\left(\frac{\gamma_{a}}{2\Omega N_{o}}\right)\right] \\
= \frac{\gamma(D/2) \mu}{\sqrt{2\pi} \left(\frac{\mu}{2}\right)^{D-1} \exp\left(\frac{\mu}{2}\right)} \\
\]

where \( T_{o}^{MRC} = D_{MRC}^{MRC} / \gamma_{a} \).

The MRC diversity scheme is the optimum among linear diversity combining schemes, in the sense that it produces the maximum possible value of instantaneous output SNR. But, MRC system requires knowledge of time-varying SNR on each branch, which can be difficult to measure in practice [64].

### 3.3.3 EGC Scheme with \( m = 0.5 \)

The EGC receive-diversity combining scheme co-phases the signals from all the branches with equal weighting [64]. The PDF of received signal envelope \( X \) in \( D \)-branch EGC receive-diversity is given by [20]

\[
F_{X}^{EGC}(x) = \frac{2D}{\sqrt{\pi\Omega}} \left[ \text{erf}\left(\frac{x}{\sqrt{2D\Omega}}\right)\right]^{D-1} \exp\left(-\frac{x^2}{2D\Omega}\right), \ x \geq 0. \quad (3.21)
\]

Substituting (3.21) in (3.1) and using the result [67]

\[
\int_{0}^{\lambda} \text{erf}\left(\frac{x}{\sqrt{2Da}}\right) e^{-\frac{x^2}{2Da}} dx = \sqrt{\frac{\pi\Omega}{2D}} \left[ \text{erf}\left(\frac{\lambda}{\sqrt{2La}}\right)\right]^{D} \quad (3.22)
\]
the OP of EGC scheme can be derived as

$$P_o^{EGC} = \left[ \text{erf}\left( \frac{\gamma \theta / E_b}{\sqrt{2D\Omega / N_o}} \right) \right]^D = \left[ \text{erf}\left( \frac{\mu}{\sqrt{2D}} \right) \right]^D.$$  \hfill (3.23)

Then, using (3.11) and (3.21), the LCR of EGC scheme can be derived as

$$L_{\gamma_o}^{EGC} = \sqrt{2D} f_m E_r^{D-1} e^{- \frac{\gamma_o}{2D\Omega / N_o}} \text{erf}\left( \frac{\mu}{\sqrt{2D}} \right)^D \exp\left( -\frac{\mu}{2D} \right).$$  \hfill (3.24)

By substituting (3.23) and (3.24) in (3.2), the OD of EGC can be expressed as

$$T_o^{EGC} = \frac{P_o^{EGC}}{L_{\gamma_o}^{EGC}}$$

and the AOD is

$$\bar{T}_o^{EGC} = f_m T_o^{EGC} = \frac{1}{\sqrt{2D} f_m} \text{erf}\left( \frac{\mu}{\sqrt{2D}} \right) \exp\left( \frac{\mu}{2D} \right).$$  \hfill (3.25)

### 3.3.4 Average BER of M-ary PSK Scheme

This section discusses MRC diversity scheme over Nakagami-$m$ fading channels by deriving the expressions for average BER of MPSK during system outage.

Consider an $D$-branch diversity receiver in frequency-nonselective, slowly fading Nakagami-$m$ channels with SNR per bit on $i$th branch as $\gamma_i, i=1, 2, \ldots, D$. The instantaneous output SNR per bit for pre-detection MRC over Nakagami-$m$ fading channels is given by [38]

$$\gamma_b = \frac{E_b}{N_o} \sum_{i=1}^{D} \alpha_i^2 = \sum_{i=1}^{D} \gamma_i.$$  \hfill (3.26)

The instantaneous output SNR per bit $\gamma_b$ is gamma distributed with PDF

$$f_r(\gamma_b) = \left( \frac{m}{\bar{\gamma}} \right)^m \gamma_b^{mD-1} \exp\left( -\frac{m\gamma_b}{\bar{\gamma}} \right), \ m \geq 0.5$$ \hfill (3.27)

where $\bar{\gamma} = (E_b / N_o) \mathbb{E}[\alpha_i^2]$ is the average output received SNR per bit for a channel.
During the system outages $\gamma_b \leq \gamma_{th}$ and the error performance of the system depends on the SNR regime $0 \leq \gamma_b \leq \gamma_{th}$, where $\gamma_{th}$ is minimum acceptable threshold value. The average BER during system outage can be represented as [65]

$$
\overline{P}_b(\gamma_{th}) = \frac{1}{\gamma_{th}} \int_0^{\gamma_{th}} P_b(\gamma_b) f(\gamma_b) d\gamma_b
$$

(3.28)

where $P_b(\gamma_b)$ is the conditional BER in AWGN channel with SNR $\gamma_b$, which can be approximated, for MPSK modulation scheme with coherent detection, by [17]

$$
P_b(\gamma_b) = \frac{2}{k} Q\left(2\gamma_b k \sin^2\left(\frac{\pi}{\mathcal{M}}\right)\right)
$$

(3.29)

where $k = \log_2 \mathcal{M}$ is the number of bits per symbol, $\mathcal{M}$ is the number of possible symbols. Applying the limit $\gamma_{th} \to \infty$, the overall BER, $\overline{P}_b$ can be represented as [65]

$$
\overline{P}_b(\gamma_{th}) = \overline{P}_b
$$

(3.30)

Upper-bound approximation of $Q$-function, which is the Chernoff bound, given by [19]

$$
Q(x) \leq \frac{1}{2} \exp\left(-\frac{x^2}{2}\right)
$$

(3.31)

can be used to represent the conditional probability (3.29) as

$$
P^u_b(\gamma_b) = \frac{1}{k} \exp\left(-k\gamma_b \sin^2\left(\frac{\pi}{\mathcal{M}}\right)\right) = c_1 \exp(-c_2\gamma_b)
$$

(3.32)

where the superscript $u$ represents the upper-bound approximation, $c_1 = \frac{1}{k}$ and $c_2 = k \sin^2\left(\frac{\pi}{\mathcal{M}}\right)$. Average BER of MPSK modulation can be obtained by substituting (3.3) and (3.32) in (3.28) as [65]

$$
\overline{P}^u_b(\gamma_{th}) = c_1 \left(\frac{m}{\overline{f}}\right)^{mD} \frac{1}{\Gamma(mD)} \int_0^{\gamma_{th}} \gamma_b^{mD-1} \exp(-\beta\gamma_b) d\gamma_b
$$

(3.33)
where $\beta = c_2 + \frac{m}{\gamma}$. By using the result given by [68]

$$\int_0^\infty x^n \exp(-\zeta x) dx = \frac{n!}{\zeta^{n+1}}$$  \hspace{1cm} (3.34)

and (3.30), the integral in (3.33) can be computed to obtain the expression of overall average BER with upper-bound approximation (3.31) as [65]

$$P_b^u = c_1 \left( \frac{m}{\gamma} \right)^{mD} \frac{1}{\Gamma(mD)} \frac{(mD-1)!}{\beta^{mL}} = c_1 \left( \frac{m}{m + c_2 \gamma} \right)^{mD}.$$  \hspace{1cm} (3.35)

Now, a new and very simple upper-bound on $Q(x)$ given by [70]

$$Q(x) \leq \frac{1}{50} \exp(-x^2) + \frac{1}{2(x+1)} \exp(-x^2/2)$$  \hspace{1cm} (3.36)

is used to find the average BER of MPSK system. The superscript $nu$ is used to indicate the new upper-bound. Using (3.36), the conditional BER (3.29) can be represented as

$$P_b^{nu} = \frac{c_1}{25} \exp(-2c_2 \gamma_b) + \frac{c_1}{1 + \sqrt{2c_2 \gamma_b}} \exp(-c_2 \gamma_b).$$  \hspace{1cm} (3.37)

Then the average BER can be obtained by inserting (3.3) and (3.37) in (3.28) as

$$P_b^{nu} = I_1^{nu} + I_2^{nu}$$  \hspace{1cm} (3.38)

where

$$I_1^{nu} = \frac{c_1}{25} \left( \frac{m}{\gamma} \right)^{mD} \frac{1}{\Gamma(mD)} \int_0^{\gamma_b} \gamma_b^{mD-1} \exp(-\rho \gamma_b) \ d\gamma_b$$  \hspace{1cm} (3.39)

with $\rho = 2c_2 + \frac{m}{\gamma}$, and

$$I_2^{nu} = c_1 \left( \frac{m}{\gamma} \right)^{mD} \frac{1}{\Gamma(mD)} \int_0^{\gamma_b} \gamma_b^{mD-1} \exp(-\beta \gamma_b) \ d\gamma_b.$$  \hspace{1cm} (3.40)

Applying the limit $\gamma_b \to \infty$ and using (3.34), the equation (3.39) evaluates to
\[ I_1^u = \frac{c_1}{25} \left( \frac{m}{m + 2c_2 \gamma} \right)^{mD}. \]  

(3.41)

Similarly, using the lower-bound approximation given by [70]

\[ n^l Q(x) \geq \frac{1}{12} \exp(-x^2) + \frac{1}{\sqrt{2\pi} (x+1)} \exp(-x^2/2) \tag{3.42} \]

the average BER can be expressed as

\[ \bar{P}_b^n(l') = I_1^l + I_2^l \tag{3.43} \]

where the superscript \( nl \) is used to indicate the new lower-bound on Gaussian \( Q \)-function

\[ I_1^l = \frac{c_1}{6} \left( \frac{m}{\gamma} \right)^{mD} \frac{1}{\Gamma(mD)} \int_0^{\gamma} \gamma^{mD-1} \exp(-\rho \gamma_b) \, d\gamma_b \tag{3.44} \]

and

\[ I_2^l = \frac{2c_1}{\sqrt{2\pi}} \left( \frac{m}{\gamma} \right)^{mD} \frac{1}{\Gamma(mD)} \int_0^{\gamma} \frac{\gamma^{mD-1}}{1 + \sqrt{2c_2 \gamma_b}} \exp(-\beta \gamma_b) \, d\gamma_b. \tag{3.45} \]

Using (3.30) and (3.34), expression (3.44) reduces to

\[ I_1^l = \frac{c_1}{6} \left( \frac{m}{m + 2c_2 \gamma} \right)^{mD}. \tag{3.46} \]

The integrals in (3.40) and (3.45) can be evaluated using numerical integration.

### 3.4 NUMERICAL COMPUTATIONAL RESULTS

This section presents numerical computational results obtained from the expressions that have been derived for OP, LCR and AOD of the basic receive-diversity combining schemes over Nakagami-0.5 fading channels. The effects of exponentially decaying MIP (given by the power decay factor, \( \delta \)) and the diversity order, \( D \) are studied.

In Fig: 3.2, the OP of SC, MRC and EGC receive-diversity schemes is plotted as a function of \( \mu \). The OP of SC scheme is 0.7101 at \( \mu = 2 \) for dual branch \( (D = 2) \).
diversity and it is reduced to 0.5043 when the diversity order increased to 4. In case of MRC scheme, the OP reduced from 0.6321 to 0.2642 when $D$ is increased from 2 to 4. But, for EGC scheme, the OP is increased to 0.9814 from 0.9111 when $D$ is increased from 2 to 4. It is also observed that when $\gamma_h \rightarrow \bar{\gamma}$ (i.e., $\mu \rightarrow 1$), the OP of MRC scheme is 0.3935 for $D = 2$ and 0.0902 for $D = 4$ as shown in Table 3.1. This indicates that the outage performance of MRC scheme is optimum, under worst-case ($m = 0.5$) fading condition, when compared to SC and EGC schemes.

Table 3.1 The OP of SC, MRC and EGC Diversity Schemes.

<table>
<thead>
<tr>
<th>Diversity Scheme</th>
<th>$\mu = 1$</th>
<th>$\mu = 2$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$D = 2$</td>
<td>$D = 4$</td>
</tr>
<tr>
<td>SC</td>
<td>0.4661</td>
<td>0.2172</td>
</tr>
<tr>
<td>MRC</td>
<td>0.3935</td>
<td>0.0902</td>
</tr>
<tr>
<td>EGC</td>
<td>0.7101</td>
<td>0.8300</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Diversity Scheme</th>
<th>$\mu = 1$</th>
<th>$\mu = 2$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$D = 2$</td>
<td>$D = 4$</td>
</tr>
<tr>
<td>SC</td>
<td>1.5918</td>
<td>3.1836</td>
</tr>
<tr>
<td>MRC</td>
<td>0.3660</td>
<td>0.1678</td>
</tr>
<tr>
<td>EGC</td>
<td>0.1774</td>
<td>0.0562</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Diversity Scheme</th>
<th>$\mu = 1$</th>
<th>$\mu = 2$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$D = 2$</td>
<td>$D = 4$</td>
</tr>
<tr>
<td>SC</td>
<td>0.0010</td>
<td>2.2019e-05</td>
</tr>
<tr>
<td>MRC</td>
<td>8.2402e-04</td>
<td>6.8080e-06</td>
</tr>
<tr>
<td>EGC</td>
<td>5.2466e-04</td>
<td>1.3813e-06</td>
</tr>
</tbody>
</table>

Table 3.3 The OP of SC, MRC and EGC Diversity Schemes with exponentially decaying MIP

<table>
<thead>
<tr>
<th>Diversity Scheme</th>
<th>$E_b/N_o = 0$ dB, $\gamma_h = -20$ dB</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$\delta = 0.5$</td>
</tr>
<tr>
<td></td>
<td>$D = 2$</td>
</tr>
<tr>
<td>SC</td>
<td>0.0010</td>
</tr>
<tr>
<td>MRC</td>
<td>8.2402e-04</td>
</tr>
<tr>
<td>EGC</td>
<td>5.2466e-04</td>
</tr>
</tbody>
</table>

Table 3.4 The AOD of SC, MRC and EGC Diversity Schemes with exponentially decaying MIP
In Fig: 3.3, the normalized LCR \( \frac{L_{\alpha_b}}{f_m} \) of the three basic diversity combining schemes is plotted as a function of \( \mu \). When \( D = 1 \), the normalized LCR of three combining schemes is the same and decreases exponentially with \( \mu \).

Fig: 3.4 shows the AOD of three diversity schemes with \( D = 2 \) and \( D = 4 \). The AOD decreases with increase in diversity order for SC and MRC schemes whereas it increases for EGC scheme. The AOD increases with increase in \( \mu \) as shown in Table 3.2.

The OP of SC, MRC and EGC schemes is plotted as a function of \( E_b/N_o \) in Fig: 3.5. Observe that the OP increases when the power decay factor increases from 0 to 1. Note that this variation is significantly less when compared to the variation of OP with increase in diversity order as shown in Table 3.3.

The normalized LCR of three linear diversity combining schemes is plotted as a function of \( E_b/N_o \) in Fig: 3.6. The LCR decreases with increase in \( D \) whenever \( E_b/N_o \) greater than \( \gamma_{th} = -10 \) dB. In Fig: 3.7, the AOD of SC, MRC and EGC diversity schemes is plotted as a function of \( E_b/N_o \) with \( \gamma_{th} = -20 \) dB and \( \Omega_o = 10 \) dB. It is observed that the duration of system under outage increases with increase in power decay factor from 0 to 1 and the outage duration increases when the diversity order is increased as shown in Table 3.4.

In Nakagami-0.5 fading channel, system performance is better when the diversity order is increased. That is, the multipath fading effects of the channel can be mitigated by employing the receive diversity combining techniques even under severe fading conditions.
Figure 3.2  The OP of SC, MRC and EGC receive diversity combining schemes.

Figure 3.3  The normalized LCR of SC, MRC and EGC receive diversity combining schemes.
Figure 3.4  The AOD of SC, MRC and EGC receive diversity combining schemes.

Figure 3.5  The OP of receive diversity schemes (a) SC, (b) MRC and (c) EGC with $\gamma_{th} = -20$ dB and $\Omega_o = 10$ dB.
Figure 3.6  The normalized LCR of receive diversity schemes (a) SC, (b) MRC and (c) EGC with $\gamma_{th} = -10$ dB and $\Omega_o = 10$ dB.

Figure 3.7  The AOD of receive diversity schemes (a) SC, (b) MRC and (c) EGC with $\gamma_{th} = -20$ dB and $\Omega_o = 10$ dB.
Figure 3.8  The OP of SC with varying power decay factor $\delta = 0, 0.5, 1$ and diversity order $D = 2, 4$ ($\Omega_o=10$ dB, $E_b/N_o = 5$ dB).

Figure 3.9  The OP of MRC with varying power decay factor $\delta = 0, 0.5, 1$ and diversity order $D = 2, 4$ ($\Omega_o=10$ dB, $E_b/N_o = 5$ dB).
Figure 3.10 The OP of EGC with varying power decay factor $\delta = 0, 0.5, 1$ and diversity order $D = 2, 4$ ($\Omega_o=10$ dB, $E_b/N_o = 5$ dB).

Now, the results obtained by computing the expressions (3.35), (3.38) and (3.43) are presented. A closed-form solution to the integrals in (3.40) and (3.45) does not exist and these integrals can be computed numerically. We have used the MATLAB® function quadgk, which numerically evaluates the integral by adaptive Gauss-Kronrod quadrature. For the comparison purpose, we have plotted (3.35), (3.38) and (3.43) in Fig: 3.11 to Fig: 3.14 for different values of fading parameter and the diversity order. In Fig: 3.11 and Fig: 3.13, the average BER of binary PSK (BPSK) is plotted as a function of average input SNR per bit $\gamma$ for various values of fading parameter $m$ with no diversity ($D=1$) and with diversity ($D=2$) respectively. It is clear that the average BER improves with an increase in $D$. Similarly, in Fig: 3.12 and Fig: 3.14, the average BER for 8-ary PSK is plotted which improves when $D$ increases from 1 to 2.
Figure 3.11 Average BER versus average SNR per bit \( \bar{\gamma} \) for BPSK coherent modulation scheme with diversity order \( D = 1 \) and fading parameter \( m = 1, 2, 10 \).

Figure 3.12 Average BER versus average SNR per bit \( \bar{\gamma} \) for 8-ary PSK coherent modulation scheme with diversity order \( D = 1 \) and fading parameter \( m = 1, 2, 10 \).
Figure 3.13 Average BER versus average SNR per bit $\bar{f}$ for BPSK coherent modulation scheme with diversity order $D = 2$ and fading parameter $m = 1, 2, 10$.

Figure 3.14 Average BER versus average SNR per bit $\bar{f}$ for 8-ary PSK coherent modulation scheme with diversity order $D = 2$ and fading parameter $m = 1, 2, 10$. 
3.5 MUD IN NAKAGAMI-\(m\) FADING CHANNEL

From the outage analysis of basic diversity combining schemes, it is found that the MRC scheme is optimum. The performance of \(M\)-decorrelator over Nakagami-\(m\) fading channel is presented in this section.

3.5.1 Synchronous Case

An \(L\)-user synchronous DS-CDMA system signaling through frequency-
nonselective, slowly fading channels is considered. The received signal during \(i^{th}\) symbol interval is given by [23]

\[
r(t) = \sum_{i=0}^{\infty} \sum_{l=1}^{L} \sqrt{\frac{2E_b}{T_s}} \alpha_l(i) e^{j\theta_l(i)} b_l(i) s_i(t - iT_s) + n(t)
\]

(3.47)

where \(\alpha_l[i]\) is the fading gain of the \(l^{th}\) user’s channel during the \(i^{th}\) symbol interval, \(b_l[i]\) is the \(i^{th}\) bit of the \(l^{th}\) user, \(s_i(t)\), \(\int_0^T s_i^2(t)dt = 1\), is the spreading waveform of the \(l^{th}\) user and \(n(t)\) is assumed as a zero-mean complex non-Gaussian noise with PDF (2.7). It is assumed that the \(l^{th}\) user employs BPSK modulation to transmit the data bits \(b_l \in [-1,1]\) with equal probability and a symbol rate \(1/T\). Here, it is also assumed that the signal of each user arrives at the receiver through an independent, single-path fading channel.

The received signal \(r(t)\) is passed through a matched filter bank and its output at the \(i^{th}\) sampling instant can be represented as a column vector of length \(L\) as

\[
r[i] = RW[i]b[i] + n[i]
\]

(3.48)

where \(R\) is the signature cross-correlation matrix with elements \(\rho_{lm} = \int_0^T s_l(t)s_m(t)dt\), \((l, m = 1,2,\ldots L)\), with unity diagonal elements, \(b\) is the data vector with components \(b_l\), and the vector \(n\) contains the corresponding samples of the noise process. The channel matrix \(W[i]\) is the diagonal matrix with diagonal elements \(W_{lj} = \sqrt{E_bC_t(i)} > 0\) with \(C_t(i) = \alpha_l(i)e^{-j\theta_l(i)}\). When the channel is assumed as a slowly fading channel, \(C_t(i)\) can be modeled as a constant over a symbol period \(T\).
and the phase $\phi_i(i)$ can be estimated from the received signal. Assuming the Nakagami-$m$ fading channel, $\alpha_i[i]$ are independent and identically distributed Nakagami random variables with PDF (3.3). Over Nakagami-$m$ flat fading channel, $W_i$ is a Nakagami random variable.

### 3.5.2 Average BER of $M$-decorrelator

A closed-form expression for average BER of $M$-decorrelator is derived by averaging the asymptotic BER, (2.19), over the PDF of Nakagami random variable. Substituting $\alpha_i = |W_i|$, the conditional BER, (2.19), for user 1 can be expressed, using the relation between $Q(\cdot)$ and complementary error function $\text{erfc}(\cdot)$, as [68]

\[
P_e^1 = Q\left(\frac{a_1}{\sqrt{\left[ \mathbf{R}^{-1} \right]_{11}}}\right) = \frac{1}{2} \text{erfc}\left(\frac{a_1}{\sqrt{2\left[ \mathbf{R}^{-1} \right]_{11}}}\right).
\]  

(3.49)

Substituting (3.3) and (3.49) in (3.28), the average BER can be expressed as

\[
\overline{P}_e^1 = \left(\frac{m}{\Omega} \right)^{mD} \frac{1}{\Gamma(mD)} \int_0^{\infty} \alpha_1^{2mD-1} e^{\frac{m\alpha_1^2}{\Omega}} \text{erfc}\left(\frac{a_1}{\sqrt{2\left[ \mathbf{R}^{-1} \right]_{11}}}\right) da_1.
\]  

(3.50)

Substituting $\xi^2 = \frac{m}{\Omega} \alpha_1^2$ in the integral $I_1$ of (3.50), we get

\[
\overline{P}_e^1 = \left(\frac{\Omega}{2m} \right)^{mD} \int_0^{\infty} \xi^{2mD-1} e^{-\frac{\xi^2}{2}} \text{erfc}\left(\frac{\xi}{\sigma \sqrt{2}}\right) d\xi.
\]  

(3.51)

where $\sigma = \sqrt{\frac{2m}{\Omega} \left[ \mathbf{R}^{-1} \right]_{11}}$. From the properties of $\text{erfc}(\cdot)$, integral in (19) can be expressed, for integer values of $mD$, as

\[
\int_0^{\infty} \xi^{2mD-1} e^{-\frac{\xi^2}{2}} \text{erfc}\left(\frac{\xi}{\sigma \sqrt{2}}\right) d\xi = (mD-1)! \sum_{j=0}^{mD-1} \left( \begin{array}{c} mD-1+j \\ j \end{array} \right) \left[ 1 + (\sigma^2 + 1)^{-\frac{j}{2}} \right]^j.
\]  

(3.52)

Therefore, the average BER of $M$-decorrelator, for integer values of $mD$, can be derived as
\[
P_e = \left( \frac{1}{2} \right)^{mD} F^{mD} \cdot \sum_{j=0}^{mD-1} 2^{-j} \binom{mD-1+j}{j} \cdot G^j
\]  
(3.53)

where \( \binom{mD-1+j}{j} \) is the Binomial coefficient,

\[
F = 1 - (\sigma^2 + 1)^{-l/2}
\]  
(3.54)

and

\[
G = 1 + (\sigma^2 + 1)^{-l/2}.
\]  
(3.55)

Now, by assuming that the \( \alpha_i(i) \) are i.i.d. Nakagami-0.5 random variables with PDF (3.17), the average BER for \( M \)-decorrelator over single path Nakagami-0.5 fading channel can be derived as

\[
\bar{P}_e = \left( \frac{1}{2} \right)^{D/2} F^{D/2} \cdot \sum_{j=0}^{D-1} 2^{-j} \binom{D-1+j}{j} \cdot G^j; D = 2, 4, ...
\]  
(3.56)

3.6 SIMULATION RESULTS

In this section, simulation results for the average BER are presented by computing (3.53) and (3.56) for different values of Nakagami fading parameter and for different order of diversity with \( \Omega_o = 10 \) dB, \( \delta = 0.1, 0.3 \) and 0.9.

Performance of \( M \)-decorrelator with different influence functions is shown in Fig: 3.15, Fig: 3.16 and Fig: 3.17. In Fig: 3.15, Fig: 3.16 and Fig: 3.17, the performance of four decorrelating detectors is studied by plotting the average BER versus the SNR related to the user #1 under the assumption of perfect power control of a synchronous DS-CDMA system with six users (\( L = 6 \)) and a processing gain of 31 (\( N = 31 \)). The noise distribution parameters are \( \varepsilon = 0, 0.01, 0.1 \) & \( \kappa = 100 \) and \( mD = 1, \ \Omega_o = 10 \) dB, \( \delta = 0, 0.9 \). Similarly, in Fig: 3.18, Fig: 3.19 and Fig: 3.20, performance is studied for noise distribution parameters \( \varepsilon = 0, 0.01, 0.1 \) & \( \kappa = 100 \)
and $mD = 2, 3$, $\Omega_0 = 10$ dB, $\delta = 0.3$. The simulation results reveal the effect of Nakagami fading parameter and power decay factor. However, the proposed $M$-estimator outperforms the linear decorrelating detector and minimax decorrelating detector (both with Huber and Hampel estimators), even in highly impulsive noise. Moreover, this performance gain increases as the SNR increases and also with increased diversity order.

For completeness, an asynchronous DS-CDMA system with $L = 6$ and $N = 127$ is also considered. BER performance of the decorrelator for asynchronous-case is also presented through the simulation results in Fig: 3.21, Fig: 3.22 and Fig: 3.23, respectively, for $\varepsilon = 0$, $\varepsilon = 0.01$ and $\varepsilon = 0.1$. These computational results reveal that the increase in diversity order improves the detector performance in highly impulsive noise. Computational results also reveals that the new $M$-estimator outperforms the linear decorrelating detector and minimax decorrelating detector (both with Huber and Hampel estimators), even in highly impulsive noise under severe fading conditions of the channel. Moreover, this performance gain increases as the SNR increases. It is also clear that the proposed estimator performs well, for very heavy-tailed noise with little attendant increase in computational complexity, when compared with Huber and Hampel estimators. In Fig: 3.24, Fgiure 3.25 and Fig: 3.26, the BER performance of $M$-decorrelator is shown for $mD = 1, 2$ with fixed $\delta$ respectively, for $\varepsilon = 0$, $\varepsilon = 0.01$ and $\varepsilon = 0.1$. 
Figure 3.15  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in synchronous DS-CDMA system with AWGN, processing gain 31; $mD = 1$; $\Omega_o = 10$ dB, $\delta = 0$ (dotted), 0.9 (solid).

Figure 3.16  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 31, $mD = 1$; $\Omega_o = 10$ dB, $\delta = 0$ (dotted), 0.9 (solid).
Figure 3.17 Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) $M$-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 31, $mD = 1$; $\Omega_o = 10$ dB, $\delta = 0$ (dotted), 0.9 (solid).

Figure 3.18 Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) $M$-estimator in synchronous DS-CDMA system with AWGN, processing gain 31, $mD = 2$ (dotted), 3 (solid), $\Omega_o = 10$ dB, $\delta = 0.3$. 
Figure 3.19 Average BER performance for user #1 (of six users) for linear detector (LS), MD with $HU$, $HA$ and proposed ($PRO$) $M$-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 31, $mD = 2$ (dotted), 3 (solid), $\Omega_o = 10$ dB, $\delta=0.3$.

Figure 3.20 Average BER performance for user #1 (of six users) for linear detector (LS), MD with $HU$, $HA$ and proposed ($PRO$) $M$-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 31, $mD = 2$ (solid), 3 (dotted), $\Omega_o = 10$ dB, $\delta=0.3$. 
Figure 3.21  Average BER performance for user #1 (of six users) for linear detector (LS), MD with \textit{HU}, \textit{HA} and proposed (\textit{PRO}) $M$-estimator in asynchronous DS-CDMA system with AWGN, processing gain $127$, $mD = 1$, $\delta = 0$ (dotted), 0.9 (solid).

Figure 3.22  Average BER performance for user #1 (of six users) for linear detector (LS), MD with \textit{HU}, \textit{HA} and proposed (\textit{PRO}) $M$-estimator in asynchronous DS-CDMA system with moderate impulsive noise, processing gain $127$, $mD = 1$, $\delta = 0$ (dotted), 0.9 (solid).
Figure 3.23  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in asynchronous DS-CDMA system with highly impulsive noise, processing gain 127, $mD = 1$, $\delta = 0$ (dotted), 0.9 (solid).

Figure 3.24  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in asynchronous DS-CDMA system with AWGN, processing gain 127, $mD = 1$ (dotted), 2 (solid).
Figure 3.25  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in asynchronous DS-CDMA system with moderate impulsive noise, processing gain 127, $mD = 1$ (dotted), 2 (solid), $\Omega = 10$ dB, $\delta = 0.3$.

Figure 3.26  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in asynchronous DS-CDMA system with highly impulsive noise, processing gain 127, $mD = 1$ (dotted), 2 (solid), $\Omega = 10$ dB, $\delta = 0.3$. 
Figure 3.27  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in synchronous DS-CDMA system with AWGN, processing gain 127, $m = 0.5$, $\Omega = 10$ dB.

Figure 3.28  Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) M-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 127, $m = 0.5$, $\Omega = 10$ dB.
Figure 3.29 Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) $M$-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 127, $m = 0.5$, $\Omega = 10$ dB.

Figure 3.30 Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) $M$-estimator in asynchronous DS-CDMA system with AWGN, processing gain 127, $m = 0.5$, $\Omega = 10$ dB.
Figure 3.31 Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) $M$-estimator in asynchronous DS-CDMA system with moderate impulsive noise, processing gain 127, $m = 0.5$, $\Omega = 10$ dB.

Figure 3.32 Average BER performance for user #1 (of six users) for linear detector (LS), MD with HU, HA and proposed (PRO) $M$-estimator in asynchronous DS-CDMA system with highly impulsive noise, processing gain 127, $m = 0.5$, $\Omega = 10$ dB.
Now, the $M$-decorrelator’s performance is presented by evaluating (3.56) for $D = 2$ and 4 with different influence functions. In Fig: 3.27, Fig: 3.28 and Fig: 3.29, the BER performance of the detector is presented for synchronous CDMA system with different values of noise parameters. Finally, average BER versus SNR corresponding to the user 1 under perfect power control of an asynchronous system is shown. These simulation results show that the performance of proposed $M$-decorrelator is better compared to linear decorrelating detector and minimax decorrelating detector with Huber and Hampel estimators.

3.7 SUMMARY

The outage performance of three basic receive diversity combining schemes over Nakagami-$m$ fading channel is presented. Closed-form expressions for OP, LCR and AOD were derived for SC, MRC and EGC receive-diversity schemes for the Nakagami-$m$ fading channel to study a worst-case ($m = 0.5$) fading scenario. Effect of exponentially decaying MIP and the diversity order on the system performance was studied.

Robust multiuser detection for DS-CDMA system with MRC receive diversity over frequency-nonselective, slowly fading Nakagami-$m$ channels in a non-Gaussian environment is also presented. A closed-form expression for average probability of error of the decorrelating detector is derived for integer values of $mD$. A new $M$-estimator based robust multiuser detection technique is proposed, which significantly outperforms the linear decorrelating detector and minimax robust multiuser detector (with Huber and Hampel $M$-estimators) in non-Gaussian impulsive noise.

Simulation results show that the proposed robust multiuser detector offers significant performance gain over the linear multiuser detector and the minimax decorrelating detectors with Huber and Hampel $M$-estimator, in non-Gaussian noise with little attendant increase in the computational complexity. Effect of fading parameter, diversity order and power decay factor on the performance of decorrelator is also studied.