Interference suppression for fast frequency-hopped communication systems over fading channels

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INTERFERENCE SUPPRESSION FOR FAST FREQUENCY-HOPPED COMMUNICATION SYSTEMS OVER FADING CHANNELS

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Summary

In this thesis, we consider interference suppression for fast frequency-hopped (FFH) communication systems over fading channels. Three types of interference, namely, partial-band noise jamming (PBNJ), multitone jamming (MTJ) and multiple-access interference (MAI), are of interest. To suppress the effect of PBNJ, we propose a maximum-likelihood (ML) receiver for frequency-selective Rayleigh-fading channels and derive the corresponding analytical bit-error rate (BER) expressions. We then focus on a more general frequency-selective Rician-fading channel, in which an optimum ML receiver structure is proposed. In addition, two suboptimum ML receivers and their corresponding analytical BER expressions are presented. Furthermore, the linear-combining and product-combining receivers are also investigated under the same channel conditions.

To mitigate the composite effect of MTJ and PBNJ, we present analytical BER expressions for an ML receiver under frequency non-selective Rayleigh-fading channel conditions. Following that, frequency-selective Rayleigh-fading channels are considered. System performance of the linear-combining, product-combining and ML receivers is studied and compared. We then investigate the effects of timing and frequency offsets on system performance under frequency non-selective Rician-fading channel conditions. Numerical results show that the FFH system is very sensitive to the timing and frequency offsets.

Lastly, we focus on the suppression of MAI. An optimum ML receiver and two suboptimum ML receivers, namely, the ML-A and ML-B receivers, are proposed for synchronous systems over frequency non-selective Rician-fading channels. Analytical BER expressions of these two proposed suboptimum ML receivers are presented. In addition, we study the system performance of the linear-combining, product-combining and clipper receivers, and compare their BER performance with the proposed optimum and two suboptimum ML receivers. Performance comparisons show that the proposed optimum ML receiver has the best BER performance, and the ML-A receiver outperforms the other four receivers.

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List of Symbols

σ_w^2	variance of Gaussian noise
σ_P^2	variance of partial-band noise jamming
P_b	average bit-error rate
P_s	average symbol error rate
q_l	PBNJ jamming state of the l th hop
q_{ml}	MTJ jamming state of the m th branch and l th hop
r_{ml}	square-law detector output
y_m	final-decision variable of the m th branch
E_b	bit energy
E_s	symbol energy
F	specular-to-diffused power ratio
L	diversity order
M	modulation level
N	total number of non-overlapping frequency-hopped band
N_0	one-sided power spectral density of AWGN
P_P	total jamming power of PBNJ
P_M	total jamming power of MTJ
Q	number of MTJ tones
T	clipping threshold for the clipper receiver
T_b	bit duration

T_h	hop duration
T_s	symbol duration
W_{ss}	total spread-spectrum bandwidth

Chapter 1

Introduction

1.1 Motivation

Spread spectrum (SS) is a modulation technique that energy generated in a certain bandwidth is deliberately spread over a wider bandwidth W_{ss} , which is much greater than the information rate R in bit/s [1]-[3]. SS techniques, first developed for military applications in the United States, possess the capability of jamming protection to provide secure communication. In addition, the military advantages of SS communication systems over conventional non-spread spectrum systems can also be illustrated in terms of a communication range extension capability [4]-[5]. Moreover, by using SS techniques, we can also obtain the following advantages [6]-[9]:

- Low probability of detection;
- Low probability of interception;
- Low probability of exploitation;
- Anti-jamming;

- Multiple-access operation;
- High degree of ranging precision.

One example of military applications of SS techniques is the global positioning system (GPS), as SS signals of GPS can provide high degree of ranging precision and have the capability of jamming protection [6]. Nowadays, the SS technology has also been applied for many commercial applications. One example of commercial applications is the Interim Standard 95, which is a code-division multiple-access standard pioneered by Qualcomm for cellular and personal communication systems (PCS) [9]. Interestingly, GPS has also been launched for commercial applications nowadays to benefit all personal communication users. In addition to those applications, SS communication systems are widely used in signal hiding, packet radio networks, data transmission, etc. Therefore, SS is an essential technology for mobile, cellular and PCS, and its application to commercial domain seems unlimited [6].

A wideband SS signal can be generated by modulating the data modulated carrier a second time using a very wideband spreading signal. The spreading modulation can be in the form of a phase modulation or a rapid changing of the carrier frequency [10]. A direct-sequence (DS) SS signal is the resultant signal of using a phase modulation. When the spectrum spreading is accomplished by a rapid changing of the carrier frequency, the resultant signal is a frequency-hopped (FH) SS signal. Both DS/SS and FH/SS are commonly used SS techniques. For an FH/SS system, the carrier frequency remains unchanged within one hop duration and varies from hop to hop. The changes of carrier frequencies depend on a frequency synthesizer, which is controlled by a pseudo-noise (PN) code generator. There are two types of FH/SS signals, namely, fast FH (FFH) and slow FH (SFH) signals [11]. If there is

more than one hop transmitted for each information symbol, the resultant signal is an FFH signal. The hopping duration is defined as $T_h = T_s/L$, where T_s is the symbol duration and $L \geq 1$ is the diversity order. If one or more than one information symbols are transmitted in each hopping duration T_h , where $T_h = LT_s$, the resultant signal is an SFH signal. In this thesis, we are interested in an FFH system, as it can provide both time and frequency diversities to better counteract the effects of channel fading and interference, and enjoy the advantage of having a “coding gain” [12].

The issues of interference rejection in SS systems have become more and more important [9]. The growth in wireless communication makes it necessary to utilize the spectrum more efficiently, and the increasing sharing of spectrum results in a higher likelihood of users interfering one another. Interference-rejection techniques can increase cellular capacity by mitigating both co-channel and adjacent channel interference in cellular systems. Moreover, interference-rejection techniques can also reduce the cross-interference among different wireless systems that share the same frequency (i.e., wireless local area networks and wireless personal area networks) and improve their performance [13], [14]. Furthermore, interference-rejection techniques can help to facilitate compatibility during transitions between the old and new communication technologies. The code-division multiple-access (CDMA) overlay or coexistence with the global system for mobile communication results in interference, which limits the performance of the CDMA system. In military applications, the most obvious application of interference-rejection techniques is to suppress the effects of intentional jamming generated by hostile parties. In commercial applications, interference-rejection techniques are mainly used to mitigate the effects of unintentional multiple-access interference (MAI) in order to improve system performance. Therefore, it is very essential to study the effectiveness of

various interference-rejection techniques for SS systems.

The most commonly encountered interference waveforms are multitone jamming (MTJ), partial-band noise jamming (PBNJ), wideband noise jamming, follower jamming, pulse jamming, smart jamming, MAI, etc [9]. It is essential for a communication designer to understand how to combat these interference waveforms when processing SS signals. In this thesis, we will focus on the suppression of MTJ, PBNJ and MAI. MTJ and PBNJ normally exist in military communication systems, as they are catagued as intensional jamming waveforms that come from hostile parties. PBNJ may also exist in commercial systems, such as in cognitive radio systems [15]. MAI normally exists in commercial communication systems, as it comes from other users that share the same spectrum.

Channel fading is also a major degradation factor for signal transmission [16], [17]. In practice, both frequency non-selective and frequency-selective fading channels are commonly encountered. In this thesis, we will consider the effects of these two fading models on system performance.

In practical applications, perfect synchronization between FH/SS transmitters and receivers is not easy to be achieved. Specifically, an FH/SS receiver does not know the exact time instant at which a signal changes its frequency. Instead, the receiver will use its own estimated time instant to de-hop the received signals. Therefore, there exists some mismatch between the estimated and true values. Moreover, the receiver does not have the perfect knowledge of the frequencies, and thus there will also be some frequency offsets with respect to the true values. Those timing and frequency offsets could distort the orthogonality properties among the baseband frequency branches and severely degrade system performance. Therefore, the effects of timing and frequency offsets on FH/SS system performance will also be studied

in this thesis.

1.2 Objectives and Major Contributions

The objective of this thesis is to theoretically analyze bit-error rate (BER) performance of FFH M -ary frequency-shift keying (FFH/MFSK) communication systems over fading channels against different types of interference. Specifically, we will derive the analytical BER expressions for FFH/MFSK systems over both frequency-selective Rayleigh-fading and Rician-fading channels with PBNJ and additive white Gaussian noise (AWGN). We will also study FFH/MFSK system performance under both frequency non-selective and frequency-selective Rayleigh-fading channel conditions with the composite effect of MTJ and PBNJ. In addition, the effects of timing and frequency offsets on FFH/MFSK system performance are investigated for frequency non-selective Rician-fading channels with both MTJ and PBNJ. Moreover, to suppress the effect of MAI, the maximum-likelihood (ML), linear-combining, product-combining, and clipper receivers for synchronous systems over frequency non-selective Rician-fading channels with AWGN are investigated.

The major contributions of this thesis are summarized as follows. To suppress the effect of PBNJ, an optimum ML receiver for FFH/MFSK systems over frequency-selective Rayleigh-fading channels is proposed and its corresponding analytical BER expressions are also presented. We also extend our study to frequency-selective Rician-fading channels. We first propose an optimum ML receivers and two suboptimum ML receivers, namely, the ML-I and ML-II receivers. We then present the analytical BER expressions for these two suboptimum ML receivers, as well as the linear-combining and product-combining receivers. Numerical results show that the proposed optimum ML receiver has the best BER performance, and the proposed

ML-I receiver outperforms the other three receivers. Furthermore, under frequency non-selective Rayleigh-fading channel conditions and the composite effect of MTJ and PBNJ, an optimum ML receiver and its corresponding analytical BER expressions are presented. It is observed that PBNJ is more harmful to the system than that of MTJ when the jamming power is high and moderate. Therefore, we conclude that the proposed ML receiver can better counteract MTJ than PBNJ at high and moderate jamming power levels. We also extend the analysis to a frequency-selective Rayleigh-fading channel. After that, we study the effects of timing and frequency offsets on FFH/MFSK system performance over frequency non-selective Rician-fading channels with the joint effects of MTJ and PBNJ. Numerical results show that there is an optimum diversity order, which gives the best BER performance. It is also observed that for the linear-combining receiver under both frequency non-selective Rician-fading and Rayleigh-fading channel conditions, the optimum diversity order increases as the timing and frequency offsets increase; however, the reverse is true for the product-combining receiver. To suppress the effect of MAI, we propose an optimum and two suboptimum ML receiver structures for synchronous FFH/MA systems over frequency non-selective Rician-fading channels, and derive their corresponding BER expressions. In addition, the linear-combining, product-combining and clipper receivers are investigated and compared. Our analysis shows that the optimum ML receiver is superior to the other five receivers.

1.3 Organization of the Thesis

This thesis is organized as follows. In Chapter 2, FFH/MFSK transmitter and receiver models, different diversity-combining techniques, fading-channel models, and various types of interference will be introduced, and the literature survey will

be also presented.

In Chapter 3, we focus on the suppression of PBNJ. An optimum ML receiver structure for FFH/MFSK systems over frequency-selective Rayleigh-fading channels is proposed. We subsequently derive its corresponding analytical BER expressions, which are validated by simulation results. Following that, we consider a frequency-selective Rician-fading channel. We first propose an optimum and two suboptimum ML receivers and then present the analytical BER expressions for these two suboptimum ML receivers, as well as the linear-combining and product-combining receivers.

The suppression of composite effect of MTJ and PBNJ is considered in Chapter 4. First, we propose an optimum ML receiver structure for frequency non-selective Rayleigh-fading channels. After that, we present the analytical BER expressions of the linear-combining, product-combining and ML receivers for FFH/MFSK systems over frequency-selective Rayleigh-fading channels. BER results of these three receivers are also compared. Lastly, the effects of timing and frequency offsets on the performance of FFH/MFSK systems over frequency non-selective Rician-fading channels in the presence of MTJ and PBNJ are investigated.

In Chapter 5, we consider the suppression of MAI under frequency non-selective Rician-fading channels for synchronous FFH/MA systems with MFSK modulation. We propose an optimum and two suboptimum ML receiver structures, and then derive the analytical BER expressions for these two suboptimum ML receivers. We also investigate the BER performance of three conventional receivers, namely, the linear-combining, product-combining, and clipper receivers.

Finally, conclusions and future work are given in Chapter 6.

Chapter 2

FFH/MFSK System Model and Literature Survey

In this chapter, we will introduce FFH/MFSK transmitter and receiver models, fading channel models and various types of interference. The literature survey for each part is also provided.

2.1 Transmitter Model

The block diagram of a typical FFH/MFSK transmitter is shown in Fig. 2.1. The input binary data bits with bit energy E_b are first encoded into M -ary sequence by a binary-to- M -ary encoder. The M -ary symbol energy is $E_s = (\log_2 M)E_b$. After that, each M -ary symbol is fed into a conventional MFSK modulator, which selects one of the M baseband frequencies, f_n , according to each symbol of the M -ary sequence. Note that $f_n = n/T_h$, $n = 1, 2, \dots, M$, and $T_h = T_s/L$ is the hopping duration, $T_s = (\log_2 M)T_b$ is the symbol duration, T_b is the bit duration and L

is the number of hops per symbol. Note that L is also called the diversity order. The MFSK modulator output is then mixed with a frequency synthesizer that is controlled by a PN code generator to pseudo-randomly hop over N non-overlapping FH bands. Following that, the output signal is fed into a band-pass filter (BPF) with bandwidth W_{ss} and subsequently frequency-translated by a radio-frequency (RF) oscillator before transmission. Therefore, the transmitted signal within one hop duration can be expressed as

$$T(t) = \sqrt{2E_s/LT_h} \cos(2\pi(f_c + f_n)t + \varphi) \quad (2.1)$$

where f_c is the carrier frequency and φ is the random phase. For fair comparisons, both the total SS bandwidth W_{ss} and the bit duration T_b are assumed to be fixed. Therefore, the number of non-overlapping FH bands, N , for an M -ary case with diversity order of L is [18]

$$N = \left\lfloor \frac{W_{ss} \times T_b \times \log_2 M}{M \times L} \right\rfloor \quad (2.2)$$

where $\lfloor \cdot \rfloor$ denotes the integer part of the argument.

2.2 Receiver Model

A typical receiver structure of an FFH/MFSK system is shown in Fig. 2.2. The received signal is first down-converted by a RF oscillator, and then passed through a BPF with bandwidth W_{ss} . This output is then de-hopped by a frequency synthesizer, which is controlled by a PN code generator. After the de-hopping process, the de-hopped signal $r(t)$ is fed into a square-law detector, and its output r_{ml} (for

Figure 2.1: Transmitter block diagram of an FFH/MFSK system.

Figure 2.2: Receiver block diagram of an FFH/MFSK system.

$m = 1, 2, \dots, M; l = 1, 2, \dots, L$) can be expressed as [1]

$$r_{ml} = \left| \frac{\sqrt{2}}{T_h} \int_{(l-1)T_h}^{lT_h} r(t) \exp[i2\pi f_m t] dt \right|^2 \quad (2.3)$$

where $|\cdot|$ denotes the absolute value of the argument. Following that, various diversity-combining techniques will be employed to process the square-law detector output r_{ml} , and this process is denoted as the function $F(\cdot)$ as shown in Fig. 2.2. After that, the L -hop receptions are summed before making a final decision. For some diversity-combining techniques, side information is required. In the literature, the most commonly used diversity-combining receivers are linear-combining receiver [19], product-combining receiver [20], clipper receiver [21], self-normalizing receiver [22], ratio-threshold test diversity-combining receiver [23], order statistics normalizing receiver [24], hard-limited linear-combining receiver [25], maximum-likelihood receiver [26], etc. In the subsequent sections, we will discuss some of these typical diversity-combining receivers in detail.

2.2.1 Linear-Combining Receiver

For the linear-combining receiver, the function $F(\cdot)$ shown in Fig. 2.2 produces the same output as the input and there is no side information required. Hence,

$$z_{ml} = F(r_{ml}) = r_{ml} \quad (2.4)$$

and by summing over all the L -hop receptions, we obtain the final-decision variable y_m as

$$y_m = \sum_{l=1}^L z_{ml}. \quad (2.5)$$

BER performance of the linear-combining receiver has been studied in [19] and [27]-[34]. Lee, French and Miller [19] studied the performance of a FFH binary FSK (BFSK) system against PBNJ under non-fading channel conditions. By

using the Taylor-series expansion in central difference and the characteristic function approach, analytical BER expressions for FFH/BFSK systems over non-fading channels with MTJ and AWGN were presented in [27], which can also be applied to higher diversity orders. The analysis of [27] was then extended to frequency non-selective Rayleigh-fading and Rician-fading channels in [28] and [29], respectively. BER performance for both FFH/BFSK and SFH/MFSK systems over frequency non-selective Rician-fading channels with MTJ and AWGN was studied in [30]. Wang and Jiang [31] extended the analysis to a more general FFH/MFSK system over frequency non-selective Rician-fading channels and presented its corresponding analytical BER expressions. The upper bound on BER performance of a synchronous system in the presence of MAI was presented in [32]. In [33], the performance of an asynchronous system over frequency non-selective Rayleigh-fading channels with MAI was analyzed. BER performance for FFH/MA systems over frequency non-selective Rician-fading channels was evaluated in [34], where the Reed-Solomon coding method was used to encode the input data.

2.2.2 Product-Combining Receiver

For the product-combining receiver, the final-decision variable $\bar{y}_m$ is obtained by multiplying all the L -hop receptions as

$$\bar{y}_m = \prod_{l=1}^L r_{ml}. \quad (2.6)$$

Since each hop reception in (2.6) is positive, taking the natural logarithm on both sides will not affect the final decision. Therefore, for the product-combining receiver,

the function $F(\cdot)$ shown in Fig. 2.2 is given by

$$z_{ml} = F(r_{ml}) = \ln(r_{ml}). \quad (2.7)$$

Side information is not required for the product-combining receiver. BER performance of the product-combining receiver for FFH systems can be found in [20] and [35]-[43]. The product-combining receiver was first proposed in [20] to combat PBNJ for FFH/BFSK systems over non-fading channels. In [35], the authors also investigated this type of receivers in conjunction with a fast Fourier transform (FFT) processor under the same channel conditions as those in [20]. The product-combining receiver was also shown to be effective in suppressing the effect of PBNJ under various fading channel conditions in [36]-[41]. Analytical BER expressions for FFH/BFSK systems in the presence of MTJ over non-fading and frequency non-selective Rician-fading channels were presented in [42] and [43], respectively.

2.2.3 Clipper Receiver

For the clipper receiver, the function $F(\cdot)$ shown in Fig. 2.2 performs a clipping operation to produce a new variable given by

$$z_{ml} = F(r_{ml}) = \begin{cases} r_{ml}, & \text{if } r_{ml} \leq T \\ T, & \text{if } r_{ml} > T \end{cases} \quad (2.8)$$

where T is the clipping threshold. The clipper receiver requires the side information on signal power level in order to set an appropriate threshold level.

System performance of the clipper receiver can be found in [21] and [44]-[49]. The clipper receiver was first introduced in [21] to mitigate the effect of PBNJ for

FFH/MFSK systems under non-fading channel conditions. In conjunction with an FFT processor, analytical BER expressions were presented in [44] for FFH/BFSK systems in the presence of PBNJ. The performance degradation of FFH systems due to MTJ was investigated in [45]-[47], where non-fading channels were considered. In [48], the authors extended the analysis to a frequency non-selective Rician-fading channel. The clipper receiver was employed to suppress the effect of MAI for synchronous systems over frequency non-selective Rician-fading channels in [49], in which the probability density function (pdf) expression of the square-law detector output was obtained by using Fourier-Bessel series and Hankel characteristic function. Hence, the corresponding characteristic functions of the square-law detector output are quite complex.

2.2.4 Maximum-Likelihood Receiver

The ML receiver can be obtained based on the *maximum a posteriori probability* (MAP) criterion [1], which is to select the signal corresponding to the largest posterior probability $P(f_m|\mathbf{r})$, where f_m is the baseband frequency, $\mathbf{r} = (\mathbf{r}_1, \mathbf{r}_2, \dots, \mathbf{r}_L)$ and $\mathbf{r}_l = (r_{1l}, r_{2l}, \dots, r_{Ml})$. According to Bayes' rule [1] and under the assumption that all signals are equally probable, the MAP criterion is equivalent to finding the signal that maximizes the probability $P(\mathbf{r}|f_m)$. Since all the L -hop receptions are independent of each other, we can obtain

$$p_{\mathbf{r}}(\mathbf{r}|f_m) = \prod_{l=1}^L p_{\mathbf{r}_l}(\mathbf{r}_l|f_m) \quad (2.9)$$

where $p_{\mathbf{r}}(\cdot|f_m)$ and $p_{\mathbf{r}_l}(\cdot|f_m)$ denote the conditional pdf of $\mathbf{r}$ and $\mathbf{r}_l$, respectively. Within one hop duration, all the signals in M branches are independent of each

other. We thus have

$$\begin{aligned}
 p_{\mathbf{r}_l}(\mathbf{r}_l|f_m) &= \prod_{m=1}^M p_{r_{ml}}(r_{ml}|f_m) \\
 &= p_{r_{ml}}(r_{ml}|\text{S}) \times \prod_{p=1, p \neq m}^M p_{r_{pl}}(r_{pl}|\text{NS}) \\
 &= \frac{p_{r_{ml}}(r_{ml}|\text{S})}{p_{r_{ml}}(r_{ml}|\text{NS})} \times \prod_{p=1}^M p_{r_{pl}}(r_{pl}|\text{NS})
 \end{aligned} \tag{2.10}$$

where S and NS denote the signal and non-signal branches, respectively. The final decision will not be affected if we divide both sides of (2.10) by $\prod_{p=1}^M p_{r_{pl}}(r_{pl}|\text{NS})$. Following that, by combining (2.9) and (2.10), the final-decision variable can be obtained as

$$\bar{y}_m = \prod_{l=1}^L \frac{p_{r_{ml}}(r_{ml}|\text{S})}{p_{r_{ml}}(r_{ml}|\text{NS})}. \tag{2.11}$$

Taking the natural logarithm on both sides of (2.11), we obtain

$$y_m = \sum_{l=1}^L z_{ml} = \sum_{l=1}^L \ln \left(\frac{p_{r_{ml}}(r_{ml}|\text{S})}{p_{r_{ml}}(r_{ml}|\text{NS})} \right). \tag{2.12}$$

To implement the ML receiver, side information on noise variance, power level of MTJ, PBNJ variance, etc., is required. The ML receiver was first proposed in [26] to combat PBNJ under non-fading channel conditions. In [50], the authors derived an optimum ML receiver and proposed two suboptimum ML receivers for large and small signal-to-noise ratio (SNR) for FFH/BFSK systems with MTJ under non-fading environments. An optimum ML receiver was also proposed in [51] to mitigate the effect of MTJ for frequency non-selective Rayleigh-fading channels. However, only simulation results were presented. In [18], the authors extended the analysis of [50] and [51] to a more general FFH/MFSK system and derived two

suboptimum ML receivers for frequency non-selective Rayleigh-fading and Rician-fading channels. However, the proposed suboptimum ML receivers are complex and the exact pdf expressions of the decision variables of these two suboptimum ML receivers cannot be obtained. Instead, a numerical approach is used to obtain the pdf values of these two receivers in order to obtain the average BER results. To combat MAI, the ML receiver was also investigated in [52]-[54]. In [52], the ML receiver for synchronous systems over frequency non-selective Rayleigh-fading channels was studied. In [53], the authors investigated the ML receiver for an asynchronous system over frequency non-selective Rician-fading channels. As the pdf expressions of the square-law detector output are obtained based on the Hankel characteristic function in [53], the proposed ML receiver structure is complex and not in closed form. Hence, a suboptimum ML receiver was proposed and analyzed instead. An optimum ML receiver was proposed in [54] for asynchronous systems over frequency non-selective Rayleigh-fading channels with MAI.

2.2.5 Self-Normalizing Receiver

For the self-normalizing receiver, the decision variable z_{ml} is obtained by normalizing the square-law detector output r_{ml} by the sum of all the M branch outputs. The function $F(\cdot)$ shown in Fig. 2.2 is given by

$$z_{ml} = F(r_{ml}) = \frac{r_{ml}}{\sum_{m=1}^M r_{ml}}. \quad (2.13)$$

For the self-normalizing receiver, no side information is required. Performance studies of the self-normalizing receiver under different jamming and interference conditions have been shown in [22] and [55]-[61]. BER performance of FFH/BFSK systems over non-fading channels with PBNJ was studied in [22] and [55]. This analysis

was extended to frequency non-selective Rician-fading channels for FFH/BFSK and FFH/MFSK systems in [56] and [57], respectively. With the presence of MTJ, analytical BER expressions for non-fading and frequency non-selective Rician-fading channels were presented in [58] and [59], respectively. The analysis was later extended to a more general FFH/MFSK system over frequency non-selective Rician-fading channels in [60]. In [61], the effect of MAI and Nakagami- m fading channels on BER performance of the self-normalizing receiver was investigated.

2.2.6 Order Statistics Normalizing Receiver

For the order statistics normalizing receiver, the square-law detector outputs r_{ml} for each branch of all the L -hop receptions are first sorted in an ascending order as

$$\hat{r}_{m1} \leq \hat{r}_{m2} \leq \cdots \leq \hat{r}_{m(L-1)} \leq \hat{r}_{mL} \quad (2.14)$$

where $\hat{r}_{m1}$ and $\hat{r}_{mL}$ are the smallest and largest of $\hat{r}_{m1}$, respectively. Following that, similar to the self-normalizing receiver, the normalizing operation is performed to produce the decision variable as

$$z_{ml} = \frac{\hat{r}_{ml}}{\sum_{m=1}^M \hat{r}_{ml}}. \quad (2.15)$$

Analytical BER expressions of the order statistics normalizing receiver were presented in [24] for FFH/MFSK systems over frequency non-selective Rayleigh-fading channels with PBNJ. In [62] and [63], the authors studied the system performance of this type of receivers in the presence of MTJ. The order statistics normalizing receiver was also investigated in [64] for FFH/MFSK systems over Rayleigh-fading channels with MAI.

2.3 Fading Channel Models

In wireless communications systems, signals travel from a transmitter to a receiver over multiple reflective paths. This phenomenon causes fluctuations in the received signal's amplitude, phase, and angle of arrival, giving rise to the terminology multi-path fading [65]-[67]. Fading channels can be frequency non-selective or frequency-selective, depending on the relationship between the bandwidth used by a given transmitted signal and the coherence bandwidth of the fading channel [23]. If the bandwidth used by the given transmitted signal is much smaller than the coherence bandwidth of the fading channel, the fading is frequency non-selective; otherwise, it is frequency-selective [23]. In this section, we will introduce both fading channel models in details.

2.3.1 Frequency Non-Selective Fading Channel Models

Under frequency non-selective fading channels, the received signal, which consists of multiple reflective paths, appears to arrive at the receiver via a single fading path [1]. The received signal's amplitude can be modeled as a frequency non-selective Rayleigh-fading or Rician-fading random variable [1].

A) Frequency Non-Selective Rayleigh-Fading Channel Model

Under Rayleigh-fading channels, there is no significant line-of-sight (LOS) component for the received signal, and the continuous wave complex baseband voltage, $\tilde{V}$, as seen by the receiver is [68]

$$\tilde{V} = X + iY \quad (2.16)$$

Figure 2.3: Plots of pdf curves for Rayleigh-faded random variable with different values of σ .

where X and Y are independent Gaussian random variables with zero mean and common variance σ^2 , and $i = \sqrt{-1}$ is the complex operator. Its magnitude, i.e., $r = \sqrt{X^2 + Y^2}$, has Rayleigh distribution and this type of fading is referred to as Rayleigh fading with Rayleigh pdf expression denoted as [65]

$$p(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right), & \text{for } r \geq 0 \\ 0, & \text{otherwise.} \end{cases} \quad (2.17)$$

Figure 2.3 shows the plots of pdf curves for Rayleigh-faded random variable r with different values of σ . Performance studies of various FFH systems over frequency non-selective Rayleigh-fading channels were presented in [25], [28], [32], [40]-[41], and [51]. BER performance of FFH/BFSK systems employing a linear-combining

receiver with MTJ and AWGN was studied in [28], where the characteristic function approach was used to derive the analytical BER expressions for higher diversity orders. Analytical BER expressions of an FFH/MFSK product-combining receiver were presented in [40] under jamming-free conditions. The analysis of [40] was then extended to mitigate the effect of PBNJ in [41]. The ML receiver was proposed in [51] for FFH/BFSK systems in the presence of MTJ. To suppress the effect of MAI, the hard-limited diversity-combining receiver was investigated in [25] for asynchronous systems. In [32], the upper bounds on BER performance of the hard-limited and linear-combining receivers for synchronous FFH/MA communication systems were presented.

B) Frequency Non-Selective Rician-Fading Channel Model

When there exists an LOS component for the received signal, the continuous wave complex baseband voltage, $\tilde{V}$, as seen by the receiver is [68]

$$\tilde{V} = A \exp(i\theta) + X + iY \quad (2.18)$$

where A is the amplitude of the LOS component, and θ is its corresponding phase. Therefore, its magnitude has a Rician distribution and this type of fading is referred to as Rician fading with the Rician pdf expression denoted as [65]

$$p(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2+A^2}{2\sigma^2}\right) I_0\left(\frac{rA}{\sigma^2}\right), & \text{for } r \geq 0 \text{ and } A \geq 0 \\ 0, & \text{otherwise} \end{cases} \quad (2.19)$$

where $I_0(\cdot)$ is the zeroth-order modified Bessel function of the first kind. Moreover, Rician-fading channels are often described in terms of a specular-to-diffused power ratio F , which is defined as $F = A^2/(2\sigma^2)$ [65]. When $F = 0$, i.e., $A = 0$, Rician

Figure 2.4: Plots of pdf curves for Rician-faded random variable with unit average power level and various F .

distribution is the same as Rayleigh distribution. For $F \gg 1$, e.g., $F = 1000$, Rician distribution converges to a Gaussian distribution.

Figure 2.4 shows the plots of pdf curves for a Rician-faded random variable with unit average power level and various F . Analytical BER expressions of the linear-combining and clipper receivers for FFH/BFSK systems over frequency non-selective Rician-fading channels with MTJ and AWGN were presented in [29], in which the uniform-quantization approach was adopted. The product-combining receiver was investigated in [69] for FFH/BFSK systems in the presence of PBNJ, where both the signal and jamming are assumed to experience independent Rician fading. In [59], the authors studied the system performance of the self-normalizing receiver over frequency non-selective Rician-fading channels with MTJ. Analytical

BER expressions of the linear-combining, product-combining and clipper receivers for FFH/MA systems over frequency non-selective Rician-fading channels with MAI were presented in [49], where the Fourier-Bessel series and exponential characteristic function approaches were adopted.

2.3.2 Frequency-Selective Fading Channel Models

Under frequency-selective fading channels, the received signal consists of multiple versions of the transmitted waveforms. For FFH communication systems, there is a distinction made between frequency-selective and highly frequency-selective fading channels [70]. In a frequency-selective fading channel, the maximum time delay that may occur along the communication link is assumed to be less than the hopping duration, which is smaller than the symbol duration; while in a highly frequency-selective fading channel, the maximum time delay is assumed to be greater than the hopping duration [70]. In this thesis, we focus on frequency-selective fading channels only, where the maximum time delay is smaller than the hopping duration [70]-[72]. From [72], the multipath frequency-selective fading channel is modeled as a correlated two-path channel model within one hopping duration. That is, we sum up all possible paths and realize them into two main clumps to form these two paths. Though this correlated two-path channel model is an approximated model, it provides some insights to the behavior of FFH systems under frequency-selective fading channel conditions. Based on this correlated two-path channel model, both frequency-selective Rayleigh-fading and Rician-fading channels will be considered in this thesis, and detailed descriptions on how to generate the two correlated Rayleigh-fading and Rician-fading envelopes can be found in [72] and [73], respectively.

In this thesis, it is also assumed that each diversity reception experiences in-

dependent fading, and the smallest spacing between the frequency slots that are used by the diversity transmissions of the same symbol is larger than the coherence bandwidth of the channel [23]. In an FFH system, though the smallest spacing between adjacent frequency slots might not be greater than the coherence bandwidth of the fading channel, it is not likely that the diversity transmissions for the same symbol will be sent over adjacent frequency slots. In fact, it is possible to design a hopping pattern to avoid this [23]. Therefore, the assumption that each diversity reception fades independently is justifiable. Moreover, each diversity reception may experience frequency non-selective fading or frequency-selective fading, and both scenarios will be considered in this thesis.

2.4 Interference Models

In this section, some commonly encountered interference models will be introduced.

2.4.1 Additive White Gaussian Noise

Additive white Gaussian noise is the background noise with a constant power spectral density (PSD) level over the whole frequency band. AWGN is modeled as a zero-mean Gaussian process with one-sided PSD of N_0 . The AWGN model is mathematically tractable and does not undergo any fading, frequency selectivity and dispersion.

2.4.2 Partial-Band Noise Jamming

PBNJ is also modeled as a Gaussian process with zero mean and constant equivalent one-sided PSD level of $N_P = P_P/W_{ss}$, where P_P is the total power of the PBNJ. The total power is uniformly distributed over only a fraction γ ($0 < \gamma \leq 1$) of the total SS bandwidth W_{ss} . Thus it results in a high PSD level of N_P/γ in that particular bandwidth where PBNJ is present [7].

BER performance of the linear-combining receiver was studied in [19] to combat the effect of PBNJ. The product-combining receiver was first proposed in [20] for FFH/BFSK systems under non-fading channel conditions with PBNJ. This analysis was then extended to frequency non-selective Rayleigh-fading and Rician-fading channels in [37] and [38], respectively. Assuming that both the signal and jamming experience independent frequency non-selective Rician-fading, analytical BER expressions of the product-combining receiver have been presented in [69] based on the uniform-quantization approach. BER performance of the ratio statistic combining receiver with PBNJ was investigated in [74]. The ML diversity-combining receiver was proposed in [26] to mitigate the effect of PBNJ.

2.4.3 Multitone Jamming

Multitone jamming is another major type of intentional jamming. The total jamming power P_M is assumed to be distributed over Q equal-power jamming tones with each having a power of $P_J = P_M/Q$, and the equivalent one-sided PSD level is defined as $N_M = P_M/W_{ss}$. The jammer is assumed to have the complete knowledge about the transmission system except it does not know the exact FH band to which the signal is hopped [3], [75]. The jamming strategy is to choose the number of jamming tones, Q , such that it can cause the worst-case system performance.

There are two types of MTJ, namely, the single-tone per-band MTJ and independent MTJ. For the single-tone per-band MTJ, there is at most one jamming tone at each FH band; while for the independent MTJ, the jamming tone can be present at any branch of the FH band. BER performance for FFH/BFSK systems with either single-tone per-band MTJ or independent MTJ over non-fading and frequency non-selective Rician-fading channels has been presented in [42] and [43], respectively. In this thesis, we will focus on the single-tone per-band MTJ only, as it results in the poorest system performance comparing to other distributions of MTJ under non-fading and low signal-to-jamming ratio (SJR) conditions [60], [76]. Each MTJ tone can be expressed as

$$J(t) = \sqrt{2}a_J \cos(2\pi f_\alpha t + \phi_J), \quad \text{for } \alpha = 1, 2, \dots, M \quad (2.20)$$

where $\sqrt{2}a_J$, f_α and ϕ_J are the amplitude, frequency and phase of the jamming tone, respectively. The random phase ϕ_J is uniformly distributed over $[-\pi, \pi]$. The linear-combining receiver was investigated in [27] for FFH/BFSK systems in the presence of MTJ, in which the Taylor series expansion in central difference and the characteristic function approach were used to derive the compact analytical BER expressions for higher diversity orders. In [29], the performance of FFH/BFSK systems employing the linear-combining and clipper receivers were investigated for frequency non-selective Rician-fading channels with MTJ. BER performance of the self-normalizing receiver over frequency non-selective Rician-fading channels with MTJ was studied in [60]. An optimum ML receiver was proposed in [50] to combat MTJ under non-fading channel conditions. In addition, two suboptimum ML receivers were also proposed and analyzed for high and low SNRs, respectively. In [77], the authors studied the ML receiver for a more general FFH/MFSK system under non-fading channel conditions with MTJ and derived the corresponding BER

expressions. Moreover, the product-combining and clipper receivers were also investigated and compared with the ML receiver.

2.4.4 Multiple-Access Interference

When several users transmit their signals at the same frequency band, wrong decision may be made at the receiver due to interference caused by other users, which is called MAI. The MAI can be expressed as

$$I(t) = \sum_{k=1}^C \sqrt{2}a_k \cos(2\pi f_k t + \phi_k) \quad (2.21)$$

where C is the total number of interfering users present in this hop, $\sqrt{2}a_k$ and f_k are the amplitude and baseband frequency of the k th user, respectively, and the phase ϕ_k is uniformly distributed over $[-\pi, \pi]$.

The effects of MAI on various FH/MA systems have been studied in [32], [49], [52], and [79]-[82]. Yue [32] investigated the hard-limited and linear diversity-combining receivers for synchronous FFH/MA communication systems over frequency non-selective Rayleigh-fading channels. The upper bounds on BER performance of these two receivers were also presented. The ML receiver for synchronous FFH/MA systems with MFSK and differential phase-shift keying schemes over frequency non-selective Rayleigh-fading channels were presented in [52]. Solaiman, Glavieux and Hillion [78] investigated the equal-gain diversity improvement for FFH/MA systems under Rayleigh-fading environments. The BER expressions of an asynchronous system over non-fading channels were derived in [79], and the derived BER expressions were only valid for a two-user system. When there were more than two users in the system, only the upper bound and lower bound on the error probabilities were pre-

sented. Under frequency non-selective Rician-fading channel conditions, analytical BER expressions of three receivers, i.e., the linear-combining, product-combining and clipper receivers, were presented for synchronous and asynchronous FFH/MA systems in [49] and [80], respectively. The Gaussian quadrature rule and a moment generation algorithm were adopted in [81] to calculate the error probability for asynchronous FH/MA systems. An optimum ML receiver structure was proposed in [82] for synchronous FH/MA systems. However, the proposed ML receiver structure was not in closed form. Therefore, three suboptimum ML receivers were proposed and analyzed instead.

Chapter 3

Suppression of PBNJ for FFH/MFSK Systems over Frequency-Selective Fading Channels

In this chapter, an ML receiver for FFH/MFSK systems under frequency-selective Rayleigh-fading channel conditions with PBNJ is proposed, and its corresponding theoretical BER expressions are also derived. After that, we extend the analysis to a more general frequency-selective Rician-fading channel with PBNJ. We first propose an optimum and two suboptimum ML receivers, and then present the analytical BER expressions for these two suboptimum ML receivers and two conventional receivers, namely, the linear-combining and product-combining receivers. Performance comparisons among these receivers are also presented.

3.1 Introduction

Performance studies of various receivers for FFH systems in the presence of PBNJ have been investigated by many researchers [19]-[21], [26], [36], [56]-[57], [69], [83]. BER performance of the linear-combining and product-combining receivers was investigated in [19] and [20], respectively, under non-fading channel conditions. Based on the characteristic-function approach, analytical BER expressions of the product-combining receiver over frequency non-selective Rayleigh-fading channels were presented in [36]. Assuming that both the signal and PBNJ experience independent Rician-fading, the product-combining and linear-combining receivers were investigated in [69] and [83], respectively, where the uniform-quantization approach was used. Performance study of a FFH/MFSK system employing the clipper receiver was presented in [21]. The self-normalizing receiver for FFH/BFSK and FFH/MFSK systems was analyzed in [56] and [57], respectively, over frequency non-selective Rician-fading channels. In [26], the ML receiver was proposed to suppress the effect of PBNJ for FFH/BFSK systems under non-fading channel conditions.

In this chapter, we focus on suppression of PBNJ for various FFH/MFSK receivers over frequency-selective fading channels. In Section 3.2, we propose an optimum ML receiver over frequency-selective Rayleigh-fading channels. The corresponding analytical BER expressions are also derived. In Section 3.3, we will focus on a more general frequency-selective Rician-fading channel. An optimum ML receiver is proposed. In addition, two suboptimum ML receivers and their corresponding analytical BER expressions are presented. Furthermore, the linear-combining and product-combining receivers are also investigated and compared with the proposed optimum and suboptimum ML receivers.

3.2 Maximum-Likelihood Receiver over Frequency-Selective Rayleigh-Fading Channels

3.2.1 Bit-Error Rate Analysis

The transmitter and receiver structures of FFH systems have been shown in Figs. 2.1 and 2.2, respectively. The frequency synthesizer at the receiver side is assumed to be perfectly synchronized with that at the transmitter side. Since the multipath frequency-selective fading channel is modeled as a correlated two-path channel model within one hopping duration [72], the de-hopped signal $r(t)$ can be expressed as

$$r(t) = \sqrt{2}|a_{s_1}| \cos(2\pi f_n t + \phi_{s_1}) + \sqrt{2}|a_{s_2}| \cos(2\pi f_n(t - \tau) + \phi_{s_2}) + w(t) + q_l n_J(t) \quad (3.1)$$

where $\sqrt{2}|a_{s_1}|$ and $\sqrt{2}|a_{s_2}|$ denote the two main correlated Rayleigh amplitudes, f_n is the baseband frequency of the desired signal, ϕ_{s_1} and ϕ_{s_2} are the random signal phases uniformly distributed over $[-\pi, \pi]$. Note that, under the condition of frequency non-selective fading channels, the transmitted signal will suffer from the same attenuation instead of at least two different attenuations within one hopping duration under frequency-selective fading channel conditions, and the channel model will reduce into a one-path channel model [72]. As frequency-selective fading channels are of interest, the time delay, τ , between the two correlated paths is assumed to be smaller than the hopping duration [70] and it is uniformly distributed over $[0, T_h]$. The term $w(t)$ is the Gaussian noise with zero mean and variance $\sigma_w^2 = N_0 B$, where N_0 is the one-sided PSD level of the noise and $B = 1/T_h$. In addition, $n_J(t)$ denotes the PBNJ with the equivalent one-sided PSD level denoted as $N_P = P_P/W_{ss}$, and

variance $\sigma_P^2 = BN_J/\gamma$, where P_P is the total power of PBNJ. The parameter γ ($0 < \gamma \leq 1$) denotes the fraction of the total SS bandwidth W_{ss} being jammed by PBNJ. The term $q_l = 1$ or 0 is an indicator function, denoting whether the l th hop is jammed or not. The total variance of AWGN and PBNJ is expressed as

$$\sigma_T^2 = \sigma_w^2 + q_l \sigma_P^2. \quad (3.2)$$

It is assumed that the PBNJ is present simultaneously and stationary at all the M branches for the entire hop duration or not at all, and each hop is jammed by PBNJ independently [84]. Note that the two correlated Rayleigh amplitudes, $|a_{s_1}|$ and $|a_{s_2}|$, can be expressed as [72]

$$|a_{s_1}| = \sigma_{s_1} |c_1 + id_1| \quad (3.3)$$

and

$$|a_{s_2}| = \sigma_{s_2} [\lambda^2 |c_1 + id_1|^2 + (1 - \lambda^2) |c_2 + id_2|^2 + 2\lambda\sqrt{1 - \lambda^2}(c_1c_2 + d_1d_2)]^{1/2} \quad (3.4)$$

respectively, where c_1 , c_2 , d_1 , and d_2 are independent Gaussian random variables with zero mean and unit variance, and $i = \sqrt{-1}$ is the complex operator. The power ratio of the two paths is defined as $\epsilon = \sigma_{s_2}^2/\sigma_{s_1}^2$ ($0 < \epsilon \leq 1$), where $2\sigma_{s_1}^2$ and $2\sigma_{s_2}^2$ are the diffused power levels of the first and second paths, respectively. The parameter λ is a function of the cross-correlation coefficient ρ of the two paths, and it is given by [85]

$$\lambda \cong \sqrt{8[\sqrt{(4/\pi - 1)\rho + 1} - 1]}. \quad (3.5)$$

The de-hopped signal $r(t)$ is detected by a square-law detector and the output is denoted as r_{ml} . Following that, the ML diversity-combining technique is used to combine the L -hop receptions before making the final decision.

The ML decision variable is defined as [18], [72]

$$\bar{y}_m = \prod_{l=1}^L \frac{p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_l, S)}{p_{r_{ml}}(r_{ml}|q_l, NS)} \quad (3.6)$$

where $\phi_{ss_{nl}} = \phi_{s_1} - \phi_{s_2} + 2\pi f_n \tau$ is the phase difference between the two main correlated paths. Taking the natural logarithm on both sides, we obtain

$$y_m = \sum_{l=1}^L z_{ml} \quad (3.7)$$

where

$$z_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_l, S)}{p_{r_{ml}}(r_{ml}|q_l, NS)} \right). \quad (3.8)$$

From [72] and after some simplifications, we can obtain the conditional pdf expressions $p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_l, S)$ and $p_{r_{ml}}(r_{ml}|q_l, NS)$ as

$$p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_l, S) = \frac{1}{2(\sigma_p^2 + \sigma_T^2)} \exp \left(-\frac{r_{ml}}{2(\sigma_p^2 + \sigma_T^2)} \right) U(r_{ml}) \quad (3.9)$$

and

$$p_{r_{ml}}(r_{ml}|q_l, NS) = \frac{1}{2\sigma_T^2} \exp \left(-\frac{r_{ml}}{2\sigma_T^2} \right) U(r_{ml}) \quad (3.10)$$

respectively, where $\sigma_p^2 = [1 + 2\lambda\sqrt{\epsilon} \cos(\phi_{ss_{nl}})]\sigma_{s_1}^2 + \sigma_{s_2}^2$, and $U(\cdot)$ is the unit step function. Substituting (3.9) and (3.10) into (3.8), we obtain the ML decision variable

given by

$$z_{ml} = A_{ml}r_{ml} + B_{ml} \quad (3.11)$$

where $A_{ml} = \frac{\sigma_p^2}{2\sigma_T^2(\sigma_p^2 + \sigma_T^2)}$ and $B_{ml} = \ln\left(\frac{\sigma_T^2}{\sigma_p^2 + \sigma_T^2}\right)$. To implement the proposed ML receiver, side information on noise and jamming variance is required. In practice, the Gaussian noise variance can be estimated from a “noise-only” channel for noise power measurement [84], and the jamming variance can be extracted from an energy detector [86]. Without loss of generality, we assume that the desired signal with frequency f_1 was transmitted. The conditional pdf of z_{ml} is given by [87]

$$p_{z_{ml}}(z_{ml}|q_l, \cos(\phi_{ss_{nl}})) = \frac{1}{2A_{ml}\sigma_{ml}^2} \exp\left(-\frac{z_{ml} - B_{ml}}{2A_{ml}\sigma_{ml}^2}\right) U\left(\frac{z_{ml} - B_{ml}}{A_{ml}}\right) \quad (3.12)$$

where

$$\sigma_{ml}^2 = \begin{cases} \sigma_p^2 + \sigma_T^2, & \text{for } m = 1, \\ \sigma_T^2, & \text{for } m = 2, 3, \dots, M. \end{cases} \quad (3.13)$$

Applying the approximation based on the Taylor series expansion in central differences [88] (detailed derivation is shown in Appendix), we obtain

$$p_{z_{ml}}(z_{ml}|q_l) \approx \frac{1}{3} \sum_{b=-1}^1 p_{z_{ml}}\left(z_{ml}|q_l, \cos(\phi_{ss_{nl}}) = \frac{\sqrt{3}b}{2}\right). \quad (3.14)$$

The corresponding characteristic function is given by [87]

$$\Phi_{z_{ml}}(iv|q_l) = \int_{-\infty}^{\infty} p_{z_{ml}}(z|q_l) e^{ivz} dz. \quad (3.15)$$

Therefore, we obtain the characteristic function of z_{ml} as

$$\Phi_{z_{ml}}(iv|q_l) \approx \frac{1}{3} \sum_{b=-1}^{b=1} \frac{\exp(ivB_{ml})}{1 - i2vA_{ml}\sigma_{ml}^2} \Big|_{\cos(\phi_{ss_{nl}}) = \frac{\sqrt{3}b}{2}}. \quad (3.16)$$

The characteristic function of the final-decision variable y_m is computed from

$$\Phi_{y_m}(iv|j) = [\Phi_{z_{ml}}(iv|q_l = 1)]^j \times [\Phi_{z_{ml}}(iv|q_l = 0)]^{L-j} \quad (3.17)$$

where $j = \sum_{l=1}^L q_l$ denotes the number of hops being jammed. The corresponding pdf can be obtained as [87]

$$p_{y_m}(y_m|j) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \Phi_{y_m}(iv|j) e^{-ivy_m} dv. \quad (3.18)$$

The conditional probability of symbol error is

$$P_s(j) = 1 - \int_{-\infty}^{\infty} p_{y_1}(y_1|j) \left[\int_{-\infty}^{y_1} p_{y_m}(y_m|j) dy_m \right]^{M-1} dy_1. \quad (3.19)$$

The average symbol error probability is given by

$$P_s = \sum_{j=0}^L \binom{L}{j} \gamma^j (1 - \gamma)^{L-j} P_s(j). \quad (3.20)$$

Finally, the average BER is [1]

$$P_b = \frac{M}{2(M-1)} P_s. \quad (3.21)$$

Figure 3.1: Worst-case BER results of the ML receiver under different diversity orders with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$ and $\rho = 0.5$.

3.2.2 Numerical Results and Discussions

In this section, we present some numerical results for the proposed ML receiver. The worst-case BER results are considered, where it is numerically searched over the parameter γ to maximize the BER results for each E_b/N_P level.

In Fig. 3.1, the worst-case analytical and simulation BER results of the proposed ML receiver are presented under different diversity orders with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$ and $\rho = 0.5$. The close match between the analytical and simulation results validates the BER expressions derived in Section 3.2.1. Moreover, the close match in both results also shows that the error incurred from the Taylor-series approximation on the BER performance is insignificant. It is also observed that for $E_b/N_P > 10$ dB, the system performance gradually improves as the diversity order increases initially. This is because the system with higher diversity orders can better

Figure 3.2: Performance comparisons of the worst-case BER results for the ML receiver under different power ratios and cross-correlation coefficients with $E_b/N_0 = 13.35$ dB, $L = 3$ and $M = 32$.

combat jamming and fading effects. However, a higher diversity order also leads to higher non-coherent diversity-combining loss. Hence, by numerically searching for every value of L , we can find an optimum diversity order, i.e., $L = 23$ for this case. The optimum diversity order depends on E_b/N_0 values and modulation levels M . We can numerically search for the optimum diversity orders under other conditions, though only results for $E_b/N_0 = 13.35$ dB and $M = 32$ are provided. Moreover, it is also observed that there is an error floor at high E_b/N_P values for each diversity order. This is because at high E_b/N_P values, the error is mainly caused by AWGN. Under the condition that E_b/N_0 is fixed, the error rate is also fixed, which leads to the error floors at high E_b/N_P values for each diversity order.

Figure 3.2 illustrates the worst-case BER results under different power ratios ϵ and cross-correlation coefficients ρ . We can observe from Fig. 3.2 that a lower ϵ leads to better BER performance. This is because the multipath interference can degrade

the system performance. Since this interference comes from the second path, as ϵ decreases, the power level of the second path decreases, which results in less severe interference. Moreover, it is shown that better BER performance can be obtained when ρ becomes lower as these two paths become less correlated.

3.3 Various FFH/MFSK Receivers over Frequency-Selective Rician-Fading Channels

In this section, we will focus on a more general frequency-selective Rician-fading channel. Analytical BER expressions of the ML, linear-combining and product-combining receivers will be presented.

3.3.1 Proposed ML and Suboptimum ML Receiver Structures

The de-hopped signal $r(t)$ is expressed as

$$r(t) = \sqrt{2}|a_{s_1}|\cos(2\pi f_n t + \phi_{s_1}) + \sqrt{2}|a_{s_2}|\cos(2\pi f_n(t - \tau) + \phi_{s_2}) + w(t) + q_l n_J(t). \quad (3.22)$$

Except for the two correlated amplitudes, $|a_{s_1}|$ and $|a_{s_2}|$, all the other parameters in (3.22) have the same definitions as those in Section 3.2. The models of a_{s_1} and a_{s_2} in Section 3.2 are only applicable to frequency-selective Rayleigh-fading channels. Under the condition of frequency-selective Rician-fading channels, there exists an LOS component for the received signal, and it only exists in the first path [89]. The two main correlated amplitudes for frequency-selective Rician-fading channels can

be modeled as [73]

$$a_{s_1} = (\sigma_{s_1} \sqrt{1-\rho} c_1 + \sigma_{s_1} \sqrt{\rho} c_0 + m_1) + i(\sigma_{s_1} \sqrt{1-\rho} d_1 + \sigma_{s_1} \sqrt{\rho} d_0 + m_2) \quad (3.23)$$

and

$$a_{s_2} = (\sigma_{s_2} \sqrt{1-\rho} c_2 + \sigma_{s_2} \sqrt{\rho} c_0) + i(\sigma_{s_2} \sqrt{1-\rho} d_2 + \sigma_{s_2} \sqrt{\rho} d_0) \quad (3.24)$$

respectively, where c_0, c_1, c_2, d_0, d_1 and d_2 are independent Gaussian random variables with zero mean and unit variance. The terms m_1 and m_2 are real values, and the sum of squares of those two terms is the power level of the LOS component, i.e., $a_s^2 = m_1^2 + m_2^2$. Note that the models of a_{s_1} and a_{s_2} , as shown in (3.23) and (3.24), are more general, which makes the frequency-selective Rayleigh-fading condition a special case. More discussions on this will be presented later. In addition, we also define a specular-to-total diffused power ratio as $F = a_s^2 / (2\sigma_{s_1}^2 + 2\sigma_{s_2}^2)$. The de-hopped signal is then fed into a square-law detector. After some mathematical simplifications, the square-law detector output r_{ml} is obtained as

$$\begin{aligned} r_{ml} = & \left[c_0 \sqrt{\sigma_{s_1}^2 \rho + \sigma_{s_2}^2 \rho + 2\sigma_{s_1} \sigma_{s_2} \rho \cos(\phi_{ss_{nl}})} + c_1 \sigma_{s_1} \sqrt{1-\rho} + c_2 \sigma_{s_2} \sqrt{1-\rho} \right. \\ & \left. + n_I + \frac{a_s}{\sqrt{2}} \right]^2 + \left[d_0 \sqrt{\sigma_{s_1}^2 \rho + \sigma_{s_2}^2 \rho + 2\sigma_{s_1} \sigma_{s_2} \rho \cos(\phi_{ss_{nl}})} + d_1 \sigma_{s_1} \sqrt{1-\rho} \right. \\ & \left. + d_2 \sigma_{s_2} \sqrt{1-\rho} + n_Q + \frac{a_s}{\sqrt{2}} \right]^2 \end{aligned} \quad (3.25)$$

where n_I and n_Q are Gaussian distributed with zero mean and variance σ_T^2 . Applying the similar technique shown in [72] and after some mathematical simplifications, the

pdf expressions of r_{ml} can be obtained as

$$p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_l, S) = \frac{1}{2(\sigma_p^2 + \sigma_T^2)} \exp\left(-\frac{r_{ml} + a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right) I_0\left(\frac{a_s\sqrt{r_{ml}}}{\sigma_p^2 + \sigma_T^2}\right) U(r_{ml}) \quad (3.26)$$

and

$$p_{r_{ml}}(r_{ml}|q_l, NS) = \frac{1}{2\sigma_T^2} \exp\left(-\frac{r_{ml}}{2\sigma_T^2}\right) U(r_{ml}) \quad (3.27)$$

respectively, where $\sigma_p^2 = \sigma_{s_1}^2 + 2\rho\sigma_{s_1}\sigma_{s_2}\cos(\phi_{ss_{nl}}) + \sigma_{s_2}^2$, and $I_0(\cdot)$ represents the zeroth-order modified Bessel function of the first kind. The ML decision variable is defined as [18], [72]

$$w_{ml} = \ln\left(\frac{p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_l, S)}{p_{r_{ml}}(r_{ml}|q_l, NS)}\right) \quad (3.28)$$

and by substituting (3.26) and (3.27) into (3.28), the ML decision variable is obtained as

$$w_{ml} = \left(\frac{1}{2\sigma_T^2} - \frac{1}{2(\sigma_p^2 + \sigma_T^2)}\right) r_{ml} + \ln\left[I_0\left(\frac{a_s\sqrt{r_{ml}}}{\sigma_p^2 + \sigma_T^2}\right)\right] + \ln\left(\frac{\sigma_T^2}{\sigma_p^2 + \sigma_T^2}\right) - \frac{a_s^2}{2(\sigma_p^2 + \sigma_T^2)}. \quad (3.29)$$

Side information is also required to implement the proposed ML receiver structure shown in (3.29). Specifically, the side information on noise and jamming variance is required. Due to the nonlinear operation of the natural logarithm of the modified Bessel function, it is mathematically intractable to derive the analytical BER expressions for the ML receiver. Therefore, we propose two suboptimum ML receivers

based on the following approximation [26]

$$\ln[I_0(x)] \approx \begin{cases} x, & \text{if } x > 4 \\ x^2/4, & \text{if } 0 < x \leq 4. \end{cases} \quad (3.30)$$

When the signal power and specular-to-total diffused power ratio are relatively high, we can adopt the first approximation of (3.30) to obtain

$$z_{ml} = \beta_{ml}(\sqrt{r_{ml}} + \Theta_{ml})^2 + U_{ml} \quad (3.31)$$

where

$$\beta_{ml} = \frac{\sigma_p^2}{2\sigma_T^2(\sigma_p^2 + \sigma_T^2)}, \quad (3.32)$$

$$\Theta_{ml} = \frac{a_s \sigma_T^2}{\sigma_p^2}, \quad (3.33)$$

and

$$U_{ml} = \ln\left(\frac{\sigma_T^2}{\sigma_p^2 + \sigma_T^2}\right) - \frac{a_s^2}{2\sigma_p^2}. \quad (3.34)$$

In the following analysis, this type of receiver is called the ML-I receiver.

When the signal power and specular-to-total diffused power ratio are relatively low, the second approximation of (3.30) can be used to obtain

$$x_{ml} = \Psi_{ml} r_{ml} + \zeta_{ml} \quad (3.35)$$

where

$$\Psi_{ml} = \frac{a_s^2}{4(\sigma_p^2 + \sigma_T^2)^2} + \frac{\sigma_p^2}{2\sigma_T^2(\sigma_p^2 + \sigma_T^2)} \quad (3.36)$$

and

$$\zeta_{ml} = \ln \left(\frac{\sigma_T^2}{\sigma_p^2 + \sigma_T^2} \right) - \frac{a_s^2}{2(\sigma_p^2 + \sigma_T^2)}. \quad (3.37)$$

This type of receiver is called the ML-II receiver in the subsequent analysis. Comparing the two suboptimum ML receivers proposed in (3.31) and (3.35) with the optimum one in (3.29), it can be observed that the two suboptimum ML receiver structures are simpler than that of the optimum one, though the same side information is still required.

3.3.2 Performance Analysis of Various Receivers

In this section, we derive the analytical BER expressions of the proposed ML-I and ML-II receivers. Two conventional receivers, namely, the linear-combining and product-combining receivers are also investigated.

A) ML-I Receiver

The conditional pdf expressions of z_{1l} and z_{pl} , for $p = 2, 3, \dots, M$, can be expressed as [87]

$$p_{z_{1l}}(z_{1l} | \cos(\phi_{ss_{1l}}, q_l)) = \frac{\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}}{2(\sigma_p^2 + \sigma_T^2)\sqrt{\beta_{1l}(z_{1l} - U_{1l})}} \exp\left(-\frac{\left(\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}\right)^2 + a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right) \\ \times I_0\left(\frac{a_s\left(\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}\right)}{\sigma_p^2 + \sigma_T^2}\right) U\left(\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}\right) \quad (3.38)$$

and

$$p_{z_{pl}}(z_{pl} | \cos(\phi_{ss_{1l}}, q_l)) = \frac{\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}}{2\sigma_T^2\sqrt{\beta_{pl}(z_{pl} - U_{pl})}} \exp\left(-\frac{\left(\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}\right)^2}{2\sigma_T^2}\right) \\ \times U\left(\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}\right) \quad (3.39)$$

respectively. We then apply the approximation of Taylor series expansion in central differences to obtain the pdf expression of z_{ml} as [88]

$$p_{z_{ml}}(z_{ml} | q_l) \approx \frac{1}{3} \sum_{b=-1}^1 p_{z_{ml}}\left(z_{ml} | \cos(\phi_{ss_{1l}}) = \frac{\sqrt{3}}{2}b, q_l\right). \quad (3.40)$$

As it is very difficult to obtain the corresponding characteristic function, we adopt a G -level uniform-quantization approach to approximate the decision variable z_{ml} , and denote its output as $\tilde{z}_{ml}$. The pdf expression of the discrete random variable $\tilde{z}_{ml}$ is expressed as

$$p_{\tilde{z}_{ml}}(\tilde{z}_{ml} | q_l) = \sum_{g=-G/2}^{G/2} V_{z_{ml}}(g | q_l) \delta(\tilde{z}_{ml} - a_g) \quad (3.41)$$

where $a_g = 2gH/G$, for $g = -G/2, -G/2+1, \dots, G/2$, and H is set at a substantially high level such that the probability of z_{ml} exceeding this level is negligible. The expression of $V_{z_{ml}}(g|q_l)$ is given by

$$V_{z_{ml}}(g|q_l) = \int_{a_g-H/G}^{a_g+H/G} p_{z_{ml}}(z_{ml}|q_l) dz_{ml}. \quad (3.42)$$

The pdf expression of the final-decision variable, $y_m = \sum_{l=1}^L \tilde{z}_{ml}$, is obtained as

$$p_{y_m}(y_m|j) = [p_{\tilde{z}_{ml}}(\tilde{z}_{ml}|q_l = 1)]^{\otimes j} \otimes [p_{\tilde{z}_{ml}}(\tilde{z}_{ml}|q_l = 0)]^{\otimes (L-j)} \quad (3.43)$$

where $j = \sum_{l=1}^L q_l$, and $\otimes$ means the discrete convolution operation and the superscript $(\otimes j)$ means the $(j-1)$ -fold discrete convolution. The conditional probability of symbol error is given by

$$P_s(j) = 1 - \sum_{g_1=-GL/2}^{GL/2} p_{y_1}\left(\frac{2g_1H}{G} \middle| j\right) \left[\sum_{g_p=-GL/2}^{g_1} p_{y_p}\left(\frac{2g_pH}{G} \middle| j\right) \right]^{M-1}. \quad (3.44)$$

After that, we can obtain the average probability of symbol error as

$$P_s = \sum_{j=0}^L \binom{L}{j} \gamma^j (1-\gamma)^{L-j} P_s(j). \quad (3.45)$$

Finally, the average BER is [1]

$$P_b = \frac{M}{2(M-1)} P_s. \quad (3.46)$$

B) ML-II Receiver

The conditional pdf expressions of x_{1l} and x_{pl} , for $p = 2, 3, \dots, M$, can be expressed as

$$p_{x_{1l}}(x_{1l} | \cos(\phi_{ss_{1l}}), q_l) = \frac{1}{2\Psi_{1l}(\sigma_p^2 + \sigma_T^2)} \exp\left(-\frac{\frac{x_{1l}-\zeta_{1l}}{\Psi_{1l}} + a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right) I_0\left(\frac{a_s \sqrt{\frac{x_{1l}-\zeta_{1l}}{\Psi_{1l}}}}{\sigma_p^2 + \sigma_T^2}\right) \times U\left(\frac{x_{1l} - \zeta_{1l}}{\Psi_{1l}}\right) \quad (3.47)$$

and

$$p_{x_{pl}}(x_{pl} | \cos(\phi_{ss_{1l}}), q_l) = \frac{1}{2\Psi_{pl}\sigma_T^2} \exp\left(-\frac{x_{pl} - \zeta_{pl}}{2\Psi_{pl}\sigma_T^2}\right) U\left(\frac{x_{pl} - \zeta_{pl}}{\Psi_{pl}}\right) \quad (3.48)$$

respectively. Using the same approximation of (3.40), we obtain the conditional pdf expression. The corresponding characteristic functions are given by [90]

$$\Phi_{x_{1l}}(iv|q_l) \approx \frac{1}{3} \sum_{b=-1}^{b=1} \frac{\exp(iv\zeta_{1l})}{1 - i2v\Psi_{1l}(\sigma_p^2 + \sigma_T^2)} \exp\left(\frac{iv\Psi_{1l}a_s^2}{1 - i2v\Psi_{1l}(\sigma_p^2 + \sigma_T^2)}\right) \Bigg|_{\cos(\phi_{ss_{1l}})=\frac{\sqrt{3}b}{2}} \quad (3.49)$$

and

$$\Phi_{x_{pl}}(iv|q_l) \approx \frac{1}{3} \sum_{b=-1}^{b=1} \frac{\exp(iv\zeta_{pl})}{1 - i2v\Psi_{pl}\sigma_T^2} \Bigg|_{\cos(\phi_{ss_{1l}})=\frac{\sqrt{3}b}{2}} \quad (3.50)$$

respectively. After that, the characteristic function of the final-decision variable for the ML-II receiver is

$$\Phi_{y_m}(iv|j) = [\Phi_{x_{m_l}}(iv|q_l = 1)]^j \times [\Phi_{x_{m_l}}(iv|q_l = 0)]^{L-j}. \quad (3.51)$$

The pdf expressions of y_m can be obtained by performing the inverse transformation of its characteristic function. The conditional probability of symbol error is given by

$$P_s(j) = 1 - \int_{-\infty}^{\infty} p_{y_1}(y_1|j) \left[\int_{-\infty}^{y_1} p_{y_p}(y_p|j) dy_p \right]^{M-1} dy_1. \quad (3.52)$$

Substituting (3.52) into (3.45) and (3.46), we can obtain the average BER expressions of the proposed ML-II receiver.

C) Linear-Combining Receiver

For the linear-combining receiver, the function $F(\cdot)$ shown in Fig. 2.2 is to produce the same output as the input, and the conditional pdf expressions of z_{1l} and z_{pl} can be expressed as

$$p_{z_{1l}}(z_{1l}|q_l) \approx \frac{1}{3} \sum_{b=-1}^{b=1} \frac{1}{2(\sigma_p^2 + \sigma_T^2)} \exp\left(-\frac{z_{1l} + a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right) I_0\left(\frac{a_s \sqrt{z_{1l}}}{\sigma_p^2 + \sigma_T^2}\right) U(z_{1l}) \Bigg|_{\cos(\phi_{ss1l})=\frac{\sqrt{3}b}{2}} \quad (3.53)$$

and

$$p_{z_{pl}}(z_{pl}|q_l) = \frac{1}{2\sigma_T^2} \exp\left(-\frac{z_{pl}}{2\sigma_T^2}\right) U(z_{pl}) \quad (3.54)$$

respectively. Their corresponding characteristic functions are [90]

$$\Phi_{z_{1l}}(iv|q_l) \approx \frac{1}{3} \sum_{b=-1}^{b=1} \frac{1}{1 - i2v(\sigma_p^2 + \sigma_T^2)} \exp\left(\frac{iva_s^2}{1 - i2v(\sigma_p^2 + \sigma_T^2)}\right) \Bigg|_{\cos(\phi_{ss1l})=\frac{\sqrt{3}b}{2}} \quad (3.55)$$

and

$$\Phi_{z_{pl}}(iv|q_l) = \frac{1}{1 - i2v\sigma_T^2} \quad (3.56)$$

respectively. We can then follow the same steps shown in (3.51), (3.52), (3.45) and (3.46) to obtain the average BER expressions of the linear-combining receiver.

D) Product-Combining Receiver

For the product-combining receiver, the function $F(\cdot)$ shown in Fig. 2.2 is to produce

$$z_{ml} = \ln(r_{ml}). \quad (3.57)$$

Therefore, the pdf expressions of z_{1l} and z_{pl} are [87]

$$p_{z_{1l}}(z_{1l}|q_l) \approx \frac{1}{3} \sum_{b=-1}^{b=1} \frac{e^{z_{1l}}}{2(\sigma_p^2 + \sigma_T^2)} \exp\left(-\frac{e^{z_{1l}} + a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right) I_0\left(\frac{a_s \sqrt{e^{z_{1l}}}}{\sigma_p^2 + \sigma_T^2}\right) \Bigg|_{\cos(\phi_{ss1l})=\frac{\sqrt{3}b}{2}} \quad (3.58)$$

and

$$p_{z_{pl}}(x_{pl}|q_l) = \frac{e^{z_{pl}}}{2\sigma_T^2} \exp\left(-\frac{e^{z_{pl}}}{2\sigma_T^2}\right) \quad (3.59)$$

respectively. Their corresponding characteristic functions are given by

$$\begin{aligned} \Phi_{z_{1l}}(iv|q_l) &\approx \frac{1}{3} \sum_{b=-1}^{b=1} [2(\sigma_p^2 + \sigma_T^2)]^{iv} \exp\left(-\frac{a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right) \\ &\times \sum_{a=0}^{\infty} \frac{\Gamma(iv + 1 + a)}{(a!)^2} \left(\frac{a_s^2}{2(\sigma_p^2 + \sigma_T^2)}\right)^a \Bigg|_{\cos(\phi_{ss1l})=\frac{\sqrt{3}b}{2}} \end{aligned} \quad (3.60)$$

and

$$\Phi_{z_{pl}}(iv|q_l) = (2\sigma_T^2)^{iv} \Gamma(iv + 1), \quad (3.61)$$

respectively, where $\Gamma(\cdot)$ is the gamma function. Though the index a in (3.60) runs from zero to infinity, to numerically calculate (3.60), a good approximation can be achieved by just summing over a small number of terms. After that, following the similar steps shown in (3.51), (3.52), (3.45) and (3.46), we can obtain the analytical BER expressions of the product-combining receiver.

3.3.3 Numerical Results and Discussions

In this section, numerical results will be presented for the proposed optimum and suboptimum ML receivers, as well as the linear-combining and product-combining receivers.

Figure 3.3 shows the worst-case analytical and simulation results for the proposed ML-I, ML-II, linear-combining and product-combining receivers. In addition, simulation results of the proposed ML receiver are also presented in this figure. It is observed that both the analytical and simulation results match well with each other, which validates the analytical BER expressions derived in Section 3.3.2. Moreover, the close match in both results also shows that the error incurred from the Taylor-series approximation on the BER performance is insignificant. Furthermore, performance comparisons among these receivers show that the proposed ML receiver has the best performance, and the proposed ML-I receiver outperforms the proposed ML-II, linear-combining and product-combining receivers. When the jamming power is relatively strong and moderate (i.e., $E_b/N_P < 27$ dB), the proposed ML-II receiver is superior to the linear-combining and product-combining receivers.

This is due to the fact that the ML-II receiver can use the side information of the system (i.e., the jamming information) to counteract the jamming. At high E_b/N_P levels, the ML-II receiver is outperformed by the linear-combining receiver, as side information that the ML-II receiver relies on is not that helpful. Comparing the BER performance of the linear-combining and product-combining receivers, the BER performance of the product-combining receiver is better than that of the linear-combining receiver at strong and moderate jamming conditions. This is because when the jamming power is relatively strong, more PBNJ is present and it can do more harm to the linear-combining receiver because it can place a lot more interference power in the final decision statistics through the addition operation. For the product-combining receiver, the corresponding effect on the final decision statistics is limited due to the nature of the multiplication operation. On the other hand, the linear-combining receiver outperforms the product-combining receiver when jamming is weak, as the error is mainly caused by AWGN. Moreover, it is also observed that there is an error floor at high E_b/N_P values for each receiver. This is because at high E_b/N_P values, the error is mainly caused by AWGN. Under the condition that E_b/N_0 is fixed, the error rate is also fixed, which leads to the error floors at high E_b/N_P values for each receiver.

Performance comparisons between the proposed ML-I and ML-II receivers under different specular-to-total diffused power ratios are shown in Fig. 3.4. It is observed that, in general, a higher specular-to-total diffused power ratio F leads to better BER performance, and this observation is consistent with that in [10] for the cases under frequency non-selective Rician-fading channel conditions. Furthermore, the proposed ML-I receiver is superior to the proposed ML-II receiver under the conditions of high F values (such as $F = 5$ & $F = 10$) and $E_b/N_P > 7$ dB, and the BER gap between the ML-I and ML-II receivers increases as the F value increases. This

Figure 3.3: Performance comparisons among various receivers with $E_b/N_0 = 10$ dB, $F = 5$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$ and $L = 3$.

is due to the fact that the ML-I receiver is obtained by using the approximation of $\ln[I_0(x)] \approx x$, for $x > 4$; while the ML-II receiver is derived with the approximation of $\ln[I_0(x)] \approx x^2/4$, for $0 < x \leq 4$. At a high F level, the approximation that used to derived the ML-I receiver is more accurate than that used to derive the ML-II receiver. On the other hand, under the condition of low F values and E_b/N_P levels, the BER performance of the proposed ML-II receiver is found to be slightly better than that of the proposed ML-I receiver. In general, if $\frac{a_s \sqrt{r_{ml}}}{\sigma_p^2 + \sigma_T^2}$ is greater than 4 (i.e., at high E_b/N_0 and F values), the ML-I receiver will outperform the ML-II receiver; while if $\frac{a_s \sqrt{r_{ml}}}{\sigma_p^2 + \sigma_T^2}$ is smaller than 4 and greater than 0, the ML-II receiver will be better. When $F = 0$, i.e., the LOS component does not exist, it is found that the BER curves of the proposed ML-I and ML-II receivers converge to each other. This is because there is no LOS component (i.e., $a_s = 0$) at $F = 0$, and the approximation of (3.30) is not required as the term $\ln[I_0(x)]$ does not exist. Therefore, the ML

Figure 3.4: Performance comparisons between the proposed ML-I and ML-II receivers with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$, $\rho = 0.73$, $L = 3$, and various F values.

receiver structure in (3.29) is simplified to $w_{ml} = \frac{\sigma_p^2}{2\sigma_T^2(\sigma_p^2 + \sigma_T^2)} r_{ml} + \ln\left(\frac{\sigma_T^2}{\sigma_p^2 + \sigma_T^2}\right)$. The same expression can also be obtained by substituting $a_s = 0$ into (3.31) and (3.35). Thus, when $F = 0$, i.e., under the frequency-selective Rayleigh-fading channels, the proposed ML and suboptimum ML-I and ML-II receivers are the same.

Figure 3.5 shows BER results of the ML receiver over frequency-selective Rayleigh-fading channels with $L = 1, 3$ and 5 . Those BER results are obtained from the ML receiver structure proposed in (3.29) with $a_s = 0$. Comparing the BER performance in Fig. 3.5 with that in Fig. 3.1, we can see that those BER curves coincide with each other respectively. This demonstrates that the ML receiver proposed in Section 3.3 is more general.

BER results of the proposed ML-I, ML-II, linear-combining and product-combining receivers under different diversity orders are shown in Figs. 3.6-3.9, respectively.

Figure 3.5: BER results of the ML receiver with $E_b/N_0 = 13.35$ dB, $F = 0$, $M = 32$, $\epsilon = 1$, $\rho = 0.73$, and different diversity orders.

It is observed that there is an optimum diversity order for each receiver. This is because for FFH/MFSK systems, a higher diversity order can counteract the jamming and fading effects more efficiently; however, it also introduces more non-coherent diversity-combining loss. The optimum diversity order depends on E_b/N_0 and E_b/N_P values, and M -ary levels. We can numerically search for the optimum diversity orders at different conditions, though only the results for $M = 32$, $E_b/N_0 = 13.35$ dB and $E_b/N_P = 35$ dB are provided.

Figure 3.10 shows the BER results of the proposed ML-I receiver under different power ratios and cross-correlation coefficients. It is found that a lower power ratio results in better BER performance. This is because the power level of the second path drops as the power ratio decreases, resulting in less interference. Furthermore, we observe that system performance improves as the cross-correlation coefficient decreases due to the fact that these two paths are less correlated. The

Figure 3.6: BER results of the ML-I receiver with $E_b/N_0 = 13.35$ dB, $F = 10$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

above observations are also applicable to the proposed ML-II, linear-combining and product-combining receivers shown in Figs. 3.11-3.13, respectively.

Figure 3.7: BER results of the ML-II receiver with $E_b/N_0 = 13.35$ dB, $F = 10$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

Figure 3.8: BER results of the linear-combining receiver with $E_b/N_0 = 13.35$ dB, $F = 10$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

Figure 3.9: BER results of the product-combining receiver with $E_b/N_0 = 13.35$ dB, $F = 10$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

Figure 3.10: Performance comparisons under different power ratios and cross-correlation coefficients for the ML-I receiver with $E_b/N_0 = 16$ dB, $L = 3$, $F = 5$ and $M = 32$.

Figure 3.11: Performance comparisons under different power ratios and cross-correlation coefficients for the ML-II receiver with $E_b/N_0 = 16dB$, $L = 3$, $F = 5$ and $M = 32$.

Figure 3.12: Performance comparisons under different power ratios and cross-correlation coefficients for the linear-combining receiver with $E_b/N_0 = 16dB$, $L = 3$, $F = 5$ and $M = 32$.

Figure 3.13: Performance comparisons under different power ratios and cross-correlation coefficients for the product-combining receiver with $E_b/N_0 = 16dB$, $L = 3$, $F = 5$ and $M = 32$.

3.4 Conclusions

In this chapter, various FFH/MFSK receivers have been investigated over frequency-selective fading channels with PBNJ. The ML receiver has been proposed for frequency-selective Rayleigh-fading channels. In addition, we have also extended the analysis to a more general frequency-selective Rician-fading channel. An optimum and two suboptimum ML receivers have been proposed. We have also derived the analytical BER expressions for these two suboptimum ML receivers, as well as the linear-combining and product-combining receivers. Numerical results have shown that the proposed optimum ML receiver gives the best BER performance, and the proposed ML-I receiver outperforms the other three receivers. It is also found that there is an optimum diversity order for each receiver. Finally, the effects of power ratio and cross-correlation coefficient on system performance have also been studied.

Chapter 4

Performance Study of FFH/MFSK Systems with Composite Effect of MTJ and PBNJ over Fading Channels

In this chapter, we focus on FFH/MFSK system performance with the composite effects of MTJ and PBNJ. An ML receiver and its corresponding analytical BER expressions for frequency non-selective Rayleigh-fading channels are presented in Section 4.2. Numerical results show that the BER performance for the case when both MTJ and PBNJ are present is bounded by the two extreme cases, i.e., either MTJ or PBNJ is present. Following that, we extend the analysis to a frequency-selective Rayleigh-fading channel in Section 4.3. Lastly, in Section 4.4, we study the effects of timing and frequency offsets on system performance over frequency non-selective Rician-fading channels. The effects of some parameters, i.e., timing offset, frequency offset, and optimum diversity order, are also discussed.

4.1 Introduction

Performance studies of conventional and ML receivers with either MTJ or PBNJ have been investigated by many researchers. Based on the characteristic-function approach, analytical BER expressions for a FFH/BFSK system employing the linear-combining receiver over non-fading channels with MTJ were presented in [27]. BER performance of the linear-combining receiver was also studied in [91] for FH/BFSK systems with MTJ, where one symbol per hop was considered. The linear-combining and clipper receivers were investigated in [29] for FFH/BFSK systems over frequency non-selective Rician-fading channels with MTJ, where the uniform-quantization approach was used to derive the analytical BER expressions for higher diversity orders. In [31], the authors investigated the linear-combining receiver for a more general FFH/MFSK system over frequency non-selective Rician-fading channels with MTJ. The product-combining receiver was investigated in [38] under frequency non-selective Rician-fading channel conditions with PBNJ, in which an L -level integration was used to obtain the conditional BER expressions, where L is the number of hops per symbol. BER performance of the FFH/BFSK product-combining receiver over non-fading channels and frequency non-selective Rician-fading channels with MTJ was studied in [42] and [43], respectively. Performance comparisons among several diversity-combining techniques were presented in [92] for FFH/MFSK systems over frequency non-selective Rayleigh-fading channels with MTJ and AWGN. The ML receiver was proposed to counteract PBNJ and MTJ under non-fading channel conditions in [26] and [50], respectively. Wu [72] proposed an optimum and a suboptimum ML receivers for FFH/BFSK systems against MTJ over frequency-selective Rayleigh-fading channels.

From the above literature survey, it can be seen that only one type of jam-

ming waveform (i.e., either MTJ or PBNJ) was considered in those works, and FFH/MFSK communication systems may be contaminated by the joint effect of MTJ and PBNJ in practice, as jamming may come from different sources [8]. Therefore, it is important to study system performance when both MTJ and PBNJ are present concurrently. Hence, in this chapter, we will focus on the suppression of those two types of jamming waveforms. Specifically, in Section 4.2, we will propose and analyze an optimum ML receiver for FFH/MFSK systems over frequency non-selective Rayleigh-fading channels with the composite effect of MTJ and PBNJ. Single-tone per-band MTJ is considered, as it results in the poorest system performance comparing to other distributions of MTJ under non-fading and low SJR conditions [60], [76]. Following that, the ML, linear-combining and product-combining receivers are investigated for a more general frequency-selective Rayleigh-fading channel with MTJ and PBNJ in Section 4.3.

In the literature, most of performance studies are based on the assumption that the receiver has the perfect knowledge of hopping time and hopping frequency, i.e., the receiver knows the exact time instant at which a signal changes its frequency, and there is no frequency offset caused by the imperfect estimation of the hopping frequency. In practical systems, such perfect knowledge is not easy to be obtained, and there exists some mismatch between the estimated and true values. Those timing and frequency offsets could distort the orthogonality properties among the M branches and severely degrade the BER performance, as the energy of the signal branch is reduced and part of the energy infiltrates into the $(M - 1)$ adjacent branches. Therefore, it is essential to study the effects of timing and frequency offsets on BER performance of FFH/MFSK systems. Very few pioneer works have investigated the effects of the timing and frequency offsets on system performance [93], [94]. Hinedi, Simon and Raphaeli [93] studied the performance of a conventional

MFSK system under non-fading conditions with timing and frequency offsets. In [94], the authors also analyzed those two effects for FFH/MFSK systems with the hard-limited combining receiver under MAI over frequency non-selective Rayleigh-fading channels. Therefore, in Section 4.4, we will explore the effects of timing and frequency offsets on FFH/MFSK system performance under frequency non-selective Rician-fading channel conditions. The composite effect of MTJ and PBNJ are also considered. FFH/MFSK systems employing both linear-combining and product-combining receivers are investigated and compared.

4.2 ML Receiver over Frequency Non-Selective Rayleigh-Fading Channels

4.2.1 System Description

The transmitter and receiver structures have been presented in Chapter 2. The frequency synthesizer at the receiver side is assumed to be perfectly synchronized with that at the transmitter side. The de-hopped signal is

$$r(t) = \sqrt{2}|a_s| \cos(2\pi f_n t + \phi_s) + q_{\alpha l} J(t) + q_l n_J(t) + w(t) \quad (4.1)$$

where $\sqrt{2}|a_s|$, f_n and ϕ_s denote the Rayleigh-faded amplitude, baseband frequency and random phase of the desired signal, respectively. The phase ϕ_s is assumed to be uniformly distributed over $[-\pi, \pi]$. $J(t)$ is the multitone jammer, $n_J(t)$ represents PBNJ with total power of P_P , equivalent one-sided PSD $N_P = P_P/W_{ss}$, and variance $\sigma_P^2 = BN_P/\gamma$, and $w(t)$ is AWGN with zero mean and variance $\sigma_w^2 = N_0B$. The term $q_{\alpha l} = 1$ or 0 is a parameter indicating whether the α th branch of the

l th hop is jammed by MTJ or not. Similarly, the indicating function $q_l = 1$ or 0 indicates whether the l th hop is jammed by PBNJ or not. Each MTJ tone can be expressed as

$$J(t) = \sqrt{2}|a_J| \cos(2\pi f_\alpha t + \theta_J), \quad \text{for } \alpha = 1, 2, \dots, M \quad (4.2)$$

where $\sqrt{2}|a_J|$, f_α , and θ_J are the amplitude, jamming frequency and random phase of MTJ, respectively. Note that the terms a_s in (4.1) and a_J in (4.2) can be respectively expressed as $a_s = \sigma_s(d_s + i e_s)$ and $a_J = \sigma_M(d_J + i e_J)$, where d_s , e_s , d_J and e_J are independent real-valued Gaussian random variables with zero mean and unit variance. The diffused power levels of the signal and MTJ are $2\sigma_s^2$ and $2\sigma_M^2$, respectively. It is assumed that the signal, MTJ and PBNJ are independent of each other. The MTJ is assumed to have a total power of P_M distributed over Q equal-power jamming tones with each tone having a power level of $P_J = P_M/Q$, and the equivalent one-sided PSD level is defined as $N_M = P_M/W_{ss}$. The total jamming power is defined as $P_T = P_M + P_P$ with the equivalent PSD level denoted as $N_J = P_T/W_{ss}$. The power ratios of MTJ and PBNJ with respect to the total jamming power are, respectively, defined as $\varepsilon_M = P_M/P_T$ and $\varepsilon_P = 1 - \varepsilon_M = P_P/P_T$, i.e., MTJ only ($\varepsilon_M = 1$ and $\varepsilon_P = 0$), PBNJ only ($\varepsilon_M = 0$ and $\varepsilon_P = 1$), both MTJ and PBNJ present with equal power ($\varepsilon_M = 0.5$ and $\varepsilon_P = 0.5$), etc.

4.2.2 Performance Analysis of the ML Receiver

After some mathematical simplifications, the in-phase component $\eta_{I,ml}$ and quadrature-phase component $\eta_{Q,ml}$ of the square-law detector output, $r_{ml} = \eta_{I,ml}^2 + \eta_{Q,ml}^2$, can

be obtained as

$$\eta_{I,ml} = d_s \sigma_s \delta(f_n - f_m) + q_{\alpha l} d_J \sigma_M \delta(f_\alpha - f_m) + n_I \quad (4.3)$$

and

$$\eta_{Q,ml} = -e_s \sigma_s \delta(f_n - f_m) - q_{\alpha l} e_J \sigma_M \delta(f_\alpha - f_m) - n_Q \quad (4.4)$$

respectively, where the terms n_I and n_Q are independent Gaussian distributed with zero mean and variance $\sigma_T^2 = \sigma_w^2 + q_l \sigma_P^2$, and the function $\delta(\cdot)$ is defined as

$$\delta(x) = \begin{cases} 1, & \text{if } x = 0, \\ 0, & \text{if } x \neq 0. \end{cases} \quad (4.5)$$

Both the in-phase and quadrature-phase components shown in (4.3) and (4.4) are independent Gaussian distributed with zero mean and a common variance

$$\sigma_{ml}^2 = \sigma_s^2 \delta(f_n - f_m) + q_{\alpha l} \sigma_M^2 \delta(f_\alpha - f_m) + \sigma_T^2. \quad (4.6)$$

Therefore, the square-law detector output r_{ml} has a Rayleigh distribution with the conditional pdf expression denoted as [1]

$$p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, f_n) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml}^2}\right) U(r_{ml}). \quad (4.7)$$

Hence, the conditional pdf expressions at the signal and non-signal branches are expressed as

$$p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, S) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,S}^2}\right) U(r_{ml}) \quad (4.8)$$

and

$$p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, \text{NS}) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,NS}^2}\right) U(r_{ml}) \quad (4.9)$$

respectively, where

$$\sigma_{ml,S}^2 = \sigma_s^2 + q_{\alpha l} \sigma_M^2 \delta(f_{\alpha} - f_m) + \sigma_T^2 \quad (4.10)$$

and

$$\sigma_{ml,NS}^2 = q_{\alpha l} \sigma_M^2 \delta(f_{\alpha} - f_m) + \sigma_T^2. \quad (4.11)$$

The ML decision variable is given by [18]

$$z_{ml} = \ln\left(\frac{p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, S)}{p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, \text{NS})}\right). \quad (4.12)$$

Substituting (4.8) and (4.9) into (4.12), we can obtain the ML decision variable as

$$z_{ml} = A_{ml} r_{ml} + B_{ml} \quad (4.13)$$

where

$$A_{ml} = \frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \quad (4.14)$$

and

$$B_{ml} = \ln\left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2}\right). \quad (4.15)$$

The final-decision variable can be obtained by summing all the L -hop receptions, which is denoted as $y_m = \sum_{l=1}^L z_{ml}$. To implement the proposed ML receiver, side information is required. Specifically, noise variance, PBNJ variance and the power level of MTJ are required. In practice, the Gaussian noise variance can be estimated from a “noise-only” channel for noise power measurement [84], and the PBNJ variance and power level of MTJ can be extracted from an energy detector [86]. The conditional pdf expression of the decision variable z_{ml} is obtained as

$$p_{z_{ml}}(z_{ml}|q_{\alpha l}, q_l) = \frac{1}{2A_{ml}\sigma_{ml}^2} \exp\left(-\frac{z_{ml} - B_{ml}}{2A_{ml}\sigma_{ml}^2}\right) U\left(\frac{z_{ml} - B_{ml}}{A_{ml}}\right). \quad (4.16)$$

The corresponding characteristic function is given by

$$\Phi_{z_{ml}}(iv|q_{\alpha l}, q_l) = \frac{\exp(ivB_{ml})}{1 - i2vA_{ml}\sigma_{ml}^2}. \quad (4.17)$$

Following that, the characteristic function of the final-decision variable is

$$\Phi_{y_m}(iv|\mathbf{J}, \mathbf{q}) = \prod_{l=1}^L \Phi_{z_{ml}}(iv|q_{\alpha l}, q_l) \quad (4.18)$$

where $\mathbf{J} = (\mathbf{q}_1, \mathbf{q}_2, \dots, \mathbf{q}_l, \dots, \mathbf{q}_L)$ with $\mathbf{q}_l = (q_{1l}, q_{2l}, \dots, q_{Ml})^T$, $\mathbf{q} = (q_1, q_2, \dots, q_L)^T$, and $(\cdot)^T$ means the transpose operation. Note that as the single-tone per-band MTJ is considered in this thesis, there is at most one element of $\mathbf{q}_l$ equal to one. Performing the inverse transformation of $\Phi_{y_m}(iv|\mathbf{J}, \mathbf{q})$, we can obtain the conditional pdf expression of y_m . Without loss of the generality, we assume that the signal with frequency f_1 is transmitted. Therefore, the conditional symbol error probability is

$$P_s(\mathbf{J}, \mathbf{q}) = 1 - \int_{-\infty}^{\infty} \left[\int_{-\infty}^{y_1} p_{y_2}(y_2|\mathbf{J}, \mathbf{q}) dy_2 \times \int_{-\infty}^{y_1} p_{y_3}(y_3|\mathbf{J}, \mathbf{q}) dy_3 \times \dots \times \int_{-\infty}^{y_1} p_{y_M}(y_M|\mathbf{J}, \mathbf{q}) dy_M \right] p_{y_1}(y_1|\mathbf{J}, \mathbf{q}) dy_1. \quad (4.19)$$

We can use quantization and summation to replace the multi-dimensional integration to evaluate (4.19). Finally, the average BER is [1]

$$P_b = \frac{M}{2(M-1)} P_s \quad (4.20)$$

with the average probability of symbol error given by

$$P_s = \sum_{\text{all } \mathbf{J}, \mathbf{q}} P_s(\mathbf{J}, \mathbf{q}) \left[\prod_{l=1}^L \left(\frac{Q}{MN} \right)^{\sum_{m=1}^M q_{ml}} \left(1 - \frac{Q}{N} \right)^{1 - \sum_{m=1}^M q_{ml}} \right] \times \left[\gamma^{\sum_{i=1}^L q_i} (1 - \gamma)^{L - \sum_{i=1}^L q_i} \right]. \quad (4.21)$$

To evaluate (4.21), we should list all the possible combinations of $\mathbf{J}$ and $\mathbf{q}$ and then add all the cases together.

4.2.3 Numerical Results and Discussions

Numerical results of the proposed ML receiver will be presented in this section. The total number of FH bands is set to be 1000 for the binary case without any diversity. A numerical search is performed over the parameters Q and γ to obtain the corresponding worst-case BER results for each E_b/N_J level.

The worst-case analytical and simulation BER results of the proposed ML receiver with $E_b/N_0 = 16$ dB, $M = 4$, $\varepsilon_M = 0.5$, $\varepsilon_P = 0.5$, and $L = 1, 3$, and 5 are shown in Fig. 4.1. From Fig. 4.1, we can observe that the analytical BER results, obtained from the theoretical BER expressions derived in Section 4.2.2, match well with the simulation results. This close match validates the analytical BER expressions. Moreover, it is also observed that there is a diversity gain for higher diversity orders, as higher diversity orders can improve the system performance

Figure 4.1: Worst-case analytical and simulation results for the ML receiver with $E_b/N_0 = 16$ dB, $M = 4$, $\varepsilon_M = 0.5$, $\varepsilon_P = 0.5$, and different diversity orders.

and better counteract the fading and jamming effects. However, for FFH/MFSK communication systems with non-coherent square-law detection, the non-coherent diversity-combining loss also increases as the diversity order increases [77]. Therefore, there exists an optimum diversity order, at which the BER performance is the best. Some values of the optimum diversity orders for the ML receiver are shown in Table 4.1. From this table, it is observed that, in general, the optimum diversity order increases as the M level increases. This is due to the fact that the number of the non-overlapping FH bands decreases as the value of M increases. Therefore, the signal tone is more likely to be jammed. Hence, the system requires a higher diversity order to suppress MTJ and PBNJ at a higher value of M . Furthermore, when E_b/N_0 increases, the optimum diversity order also becomes larger as there is less non-coherent diversity-combining loss.

Figure 4.2: Performance comparisons of ML and linear-combining receivers with $E_b/N_0 = 16$ dB, $M = 4$, $L = 3$, and different jamming conditions.

Performance comparisons between the proposed ML receiver and the linear-combining receiver under different jamming conditions are shown in Fig. 4.2. Numerical results of the linear-combining receiver under either PBNJ or MTJ are obtained from [19] and [28], respectively. It is observed that the proposed ML receiver outperforms the linear-combining receiver when either MTJ or PBNJ is present. When there is only one type of jamming waveform (i.e., $\varepsilon_M = 1$ or $\varepsilon_P = 1$), it is found that PBNJ is more harmful to the proposed ML receiver than that of MTJ when the jamming power is high and moderate (i.e., $E_b/N_J < 25$ dB), as the proposed ML receiver has the exact information at which frequency slot MTJ is present. At high E_b/N_J levels (i.e., $E_b/N_J > 25$ dB), the jamming power is low and the probability of error is mainly caused by AWGN. Thus, the BER results are similar under either MTJ or PBNJ condition. Therefore, we can conclude that the proposed ML receiver can better counteract MTJ than PBNJ at high and moderate

Table 4.1: Optimum diversity orders of the ML receiver with $\varepsilon_M = 0.5$, $\varepsilon_P = 0.5$, and $E_b/N_J = 25$ dB.

ML receiver	$M = 2$	$M = 4$	$M = 8$
$E_b/N_0 = 6$ dB	$L = 3$	$L = 3$	$L = 3$
$E_b/N_0 = 10$ dB	$L = 3$	$L = 5$	$L = 7$
$E_b/N_0 = 16$ dB	$L = 11$	$L = 21$	$L = 31$

jamming power levels. When both MTJ and PBNJ are present concurrently, it is found that the BER performance is bounded by those two extreme cases.

4.3 FFH/MFSK Systems over Frequency-Selective Rayleigh-Fading Channels

In Section 4.2, we have studied the composite effect of MTJ and PBNJ on FFH/MFSK system performance over frequency non-selective Rayleigh-fading channels. In this section, we extend the analysis to a frequency-selective Rayleigh-fading channel.

4.3.1 Bit-Error Rate Analysis for Various Receivers

When the transmitted signal experiences frequency-selective Rayleigh fading, the de-hopped signal is expressed as

$$\begin{aligned}
 r(t) = & \sqrt{2}|a_{s_1}| \cos(2\pi f_n t + \phi_{s_1}) + \sqrt{2}|a_{s_2}| \cos(2\pi f_n(t - \tau) + \phi_{s_2}) \\
 & + q_{\alpha l}J(t) + q_l n_J(t) + w(t).
 \end{aligned} \tag{4.22}$$

The terms $J(t)$, $n(t)$, $w(t)$, $q_{\alpha l}$ and q_l have the same definitions as those in Section 4.2, and the other terms have the same definitions as those in (3.1). It has been shown in [29] and [60] that the system performance is not sensitive to the types of MTJ fading conditions. Hence, for mathematical convenience, we assume that the MTJ experiences frequency non-selective Rayleigh fading. The pdf expression of a_J is given by [1]

$$p_{a_J}(a_J) = \frac{a_J}{\sigma_M^2} \exp\left(-\frac{a_J^2}{2\sigma_M^2}\right) U(a_J) \quad (4.23)$$

where $2\sigma_M^2$ is the power level of the jamming tone. The de-hopped signal $r(t)$ is fed into a square-law detector, and the output is denoted as r_{ml} . After that, various diversity-combining techniques can be used to combine the L -hop receptions before making a final decision. We consider the ML, linear-combining and product-combining receivers in this section.

The conditional pdf expression of r_{ml} is [72]

$$p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_{\alpha l}, q_l, S, a_J) = \frac{1}{2\sigma_p^2} \exp\left(-\frac{r_{ml} + q_{\alpha l}2a_J^2}{2\sigma_p^2}\right) I_0\left(\frac{q_{\alpha l}a_J\sqrt{2r_{ml}}}{\sigma_p^2}\right) \times U(r_{ml}) \quad (4.24)$$

where $\sigma_p^2 = [1 + 2\lambda\sqrt{\epsilon}\cos(\phi_{ss_{nl}})]\sigma_{s_1}^2 + \sigma_{s_2}^2 + \sigma_T^2$. Integrating (4.24) over the pdf of a_J in (4.23), we obtain the conditional pdf expressions of r_{ml} in the signal and non-signal branches as

$$p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_{\alpha l}, q_l, S) = \frac{1}{2(\sigma_p^2 + q_{\alpha l}2\sigma_M^2)} \exp\left(-\frac{r_{ml}}{2(\sigma_p^2 + q_{\alpha l}2\sigma_M^2)}\right) U(r_{ml}) \quad (4.25)$$

and

$$p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, \text{NS}) = \frac{1}{2(\sigma_T^2 + q_{\alpha l}2\sigma_M^2)} \exp\left(-\frac{r_{ml}}{2(\sigma_T^2 + q_{ml}2\sigma_M^2)}\right) U(r_{ml}) \quad (4.26)$$

respectively. The ML decision variable is defined as [18], [72]

$$z_{ml} = \ln\left(\frac{p_{r_{ml}}(r_{ml}|\phi_{ss_{nl}}, q_{\alpha l}, q_l, \text{S})}{p_{r_{ml}}(r_{ml}|q_{\alpha l}, q_l, \text{NS})}\right). \quad (4.27)$$

Substituting (4.25) and (4.26) into (4.27), we obtain the ML decision variable as

$$z_{ml} = A_{ml}r_{ml} + B_{ml} \quad (4.28)$$

where

$$A_{ml} = \frac{\sigma_p^2 - \sigma_T^2}{2(\sigma_T^2 + q_{\alpha l}2\sigma_M^2)(\sigma_p^2 + q_{\alpha l}2\sigma_M^2)} \quad (4.29)$$

and

$$B_{ml} = \ln\left(\frac{\sigma_T^2 + q_{\alpha l}2\sigma_M^2}{\sigma_p^2 + q_{\alpha l}2\sigma_M^2}\right). \quad (4.30)$$

Note that side information is also required to implement the ML receiver. In the following parts, analytical BER expressions will be derived for the ML, linear-combining and product-combining receivers.

A) ML Receiver

The conditional pdf expression of z_{ml} is given by [87]

$$p_{z_{ml}}(z_{ml} | \cos(\phi_{ss_{nl}}), q_{\alpha l}, q_l) = \frac{1}{2A_{ml}\sigma_{ml}^2} \exp\left(-\frac{z_{ml} - B_{ml}}{2A_{ml}\sigma_{ml}^2}\right) U\left(\frac{z_{ml} - B_{ml}}{A_{ml}}\right) \quad (4.31)$$

where

$$\sigma_{ml}^2 = \begin{cases} \sigma_p^2 + q_{ml}2\sigma_M^2, & \text{for } m = 1, \\ \sigma_T^2 + q_{ml}2\sigma_M^2, & \text{for } m = 2, 3, \dots, M. \end{cases} \quad (4.32)$$

Applying the approximation based on Taylor series expansion in central differences, we obtain the pdf expression of z_{ml} as [88]

$$p_{z_{ml}}(z_{ml} | q_{\alpha l}, q_l) \approx \frac{1}{3} \sum_{b=-1}^1 p_{z_{ml}}\left(z_{ml} | \cos(\phi_{ss_{nl}}) = \frac{\sqrt{3}}{2}b, q_{\alpha l}, q_l\right). \quad (4.33)$$

The corresponding characteristic function is [87]

$$\Phi_{z_{ml}}(iv | q_{\alpha l}, q_l) \approx \frac{1}{3} \sum_{b=-1}^1 \frac{e^{ivB_{ml}}}{1 - i2vA_{ml}\sigma_{ml}^2} \Big|_{\cos(\phi_{ss_{nl}}) = \frac{\sqrt{3}}{2}b}. \quad (4.34)$$

After that, we can follow the similar steps shown in (4.18)-(4.20) to obtain the average BER expressions of the proposed ML receiver.

B) Linear-Combining Receiver

For the linear-combining receiver, the decision variable is $z_{ml} = F(r_{ml}) = r_{ml}$. The pdf expressions of z_{1l} and z_{pl} , for $p = 2, 3, \dots, M$, can be obtained as

$$p_{z_{1l}}(z_{1l}|q_{1l}, q_l) \approx \frac{1}{3} \sum_{b=-1}^1 \frac{1}{2\sigma_{1l}^2} \exp\left(-\frac{z_{1l}}{2\sigma_{1l}^2}\right) U(z_{1l}) \Big|_{\cos(\phi_{ss_{1l}})=\frac{\sqrt{3}}{2}b} \quad (4.35)$$

and

$$p_{z_{pl}}(z_{pl}|q_{pl}, q_l) = \frac{1}{2\sigma_{pl}^2} \exp\left(-\frac{z_{pl}}{2\sigma_{pl}^2}\right) U(z_{pl}) \quad (4.36)$$

respectively. The corresponding characteristic functions are

$$\Phi_{z_{1l}}(iv|q_{1l}, q_l) \approx \frac{1}{3} \sum_{b=-1}^1 \frac{1}{1 - 2iv\sigma_{1l}^2} \Big|_{\cos(\phi_{ss_{1l}})=\frac{\sqrt{3}}{2}b} \quad (4.37)$$

and

$$\Phi_{z_{pl}}(iv|q_{pl}, q_l) = \frac{1}{1 - 2iv\sigma_{pl}^2} \quad (4.38)$$

respectively. Following the similar analysis shown in (4.18) to (4.20), we can obtain the average BER expressions of the linear-combining receiver.

C) Product-Combining Receiver

The decision variable for the product-combining receiver is $z_{ml} = \ln(r_{ml})$, and the pdf expressions of z_{1l} and z_{pl} , for $p = 2, 3, \dots, M$, are [87]

$$p_{z_{1l}}(z_{1l}|q_{1l}, q_l) \approx \frac{1}{3} \sum_{b=-1}^1 \frac{e^{z_{1l}}}{2\sigma_{1l}^2} \exp\left(-\frac{e^{z_{1l}}}{2\sigma_{1l}^2}\right) \Big|_{\cos(\phi_{ss_{1l}})=\frac{\sqrt{3}}{2}b} \quad (4.39)$$

and

$$p_{z_{pl}}(z_{pl}|q_{pl}, q_l) = \frac{e^{z_{pl}}}{2\sigma_{pl}^2} \exp\left(-\frac{e^{z_{pl}}}{2\sigma_{pl}^2}\right) \quad (4.40)$$

respectively. Their corresponding characteristic functions can be obtained as [90]

$$\Phi_{z_{1l}}(iv|q_{1l}, q_l) \approx \frac{1}{3} \sum_{b=-1}^1 [2\sigma_{1l}^2]^{iv} \Gamma(1+iv) \Big|_{\cos(\phi_{ss_{1l}})=\frac{\sqrt{3}}{2}b} \quad (4.41)$$

and

$$\Phi_{z_{pl}}(iv|q_{pl}, q_l) = [2\sigma_{pl}^2]^{iv} \Gamma(1+iv) \quad (4.42)$$

respectively. Following (4.18) to (4.20), we can obtain the average BER expressions of the product-combining receiver.

4.3.2 Numerical Results and Discussions

In this section, BER results for the proposed ML receiver, as well as the linear-combining and product-combining receivers over frequency-selective Rayleigh-fading channels with MTJ and PBNJ are presented.

In Fig. 4.3, analytical and simulation results for the proposed ML, linear-combining, and product-combining receivers are presented with $E_b/N_0 = 13.35$ dB, $L = 5$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, $\varepsilon_M = 0.5$, and $\varepsilon_P = 0.5$. Analytical results are obtained from the BER expressions derived. The close match between the analytical and simulation results validates the BER expressions derived in Section 4.3.1. It is observed that the ML receiver outperforms the linear-combining and product-combining receivers. For example, at $E_b/N_J = 20$ dB, the proposed ML re-

Figure 4.3: Worst-case analytical and simulation BER results for the conventional and ML receivers with $E_b/N_0 = 13.35$ dB, $L = 5$, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, $\varepsilon_M = 0.5$ and $\varepsilon_P = 0.5$

ceiver outperforms the linear-combining receiver by more than one decade in terms of BER performance. However, it should be noted that the ML receiver requires some side information. Therefore, there is a tradeoff between the performance and system complexity. Performance comparisons between the product-combining and linear-combining receivers show that the product-combining receiver is significantly better than that of the linear-combining receiver for $E_b/N_J \leq 29$ dB. This is because when the jamming power is relatively strong, more jamming is present and it can do more harm to the linear-combining receiver because it can place a lot more interference power in the final decision statistics through the addition operation. For the product-combining receiver, the corresponding effect on the final decision statistics is limited due to the nature of the multiplication operation. When the jamming power is relatively weak, i.e., $E_b/N_J > 29$ dB, the product-combining re-

ceiver is outperformed by the linear-combining receiver, as the error is mainly caused by AWGN. Hence, we can conclude that the product-combining receiver can better counteract the jamming and fading effects at relatively low and moderate E_b/N_J levels than that of the linear-combining receiver.

BER results of the proposed ML, linear-combining and product-combining receivers under different jamming conditions are shown in Figs. 4.4-4.6, respectively, with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and various diversity orders. It is observed that, in general, the BER results for the case when both MTJ and PBNJ are present concurrently are bounded by the two extreme cases, where either MTJ or PBNJ is present. Furthermore, it is also observed from Fig. 4.4 that when the jamming power is strong (i.e., $E_b/N_J < 20$ dB), the BER performance of the proposed ML receiver with MTJ is better than that with PBNJ. This is because the ML receiver can suppress the tone jamming more effectively. At low jamming power ranges, such as $E_b/N_J > 20$ dB, the performance of the proposed ML receiver under either MTJ or PBNJ condition is comparable with each other. This is because under this condition, the probability of error is mainly caused by AWGN. Therefore, it can be concluded that the ML receiver can better counteract MTJ than PBNJ when jamming power is strong. While for the linear-combining and product-combining receivers shown in Figs. 4.5-4.6, respectively, MTJ is more harmful to the system than that of PBNJ when the jamming power is relatively strong.

Figures 4.7 and 4.8 compare the BER results of these three receivers under different power ratios and cross-correlation coefficients, respectively. As shown in Fig. 4.7, the BER performance improves as the power ratio becomes smaller. This is because the power level of the second path is low when the power ratio is small, which results in less interference. From Fig. 4.8, we can observe that a lower value of the cross-correlation coefficient gives better BER performance, as those two paths

Figure 4.4: BER performance of the proposed ML receiver under different jamming conditions with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

become less correlated.

In Figs. 4.9-4.11, we show the BER curves for the proposed ML, product-combining and linear-combining receivers with various diversity orders, respectively. For the proposed ML and product-combining receivers shown in Figs. 4.9 and 4.10, respectively, we can observe that the system performance improves as the diversity order increases initially. However, there is an optimum diversity order, which gives the best BER performance. At $E_b/N_J > 25$ dB, the corresponding optimum diversity order is equal to $L = 23$ and $L = 33$ for the ML and product-combining receivers, respectively. When the diversity order is higher than the optimum diversity order, the system performance degrades as the diversity order increases. The reason for this phenomenon is that for FFH/MFSK systems with non-coherent detection, the system with a higher diversity order can counteract the jamming and fading effects

Figure 4.5: BER performance of the linear-combining receiver under different jamming conditions with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

more efficiently. However, at the same time, the non-coherent diversity-combining loss increases as the diversity order increases. Hence, there exists an optimum diversity order for a given E_b/N_J level. For the linear-combining receiver shown in Fig. 4.11, there is generally no diversity gain for $E_b/N_J \leq 30$ dB. However, when $E_b/N_J > 30$ dB, we observe that there is also an optimum diversity order with $L = 21$.

Figure 4.6: BER performance of the product-combining receiver under different jamming conditions with $E_b/N_0 = 13.35$ dB, $M = 32$, $\epsilon = 1$, $\rho = 0.5$, and different diversity orders.

Figure 4.7: Performance comparisons among the three receivers under different power ratio with $E_b/N_0 = 13.35$ dB, $L = 3$, $M = 32$, $\epsilon_M = 1$, $\epsilon_P = 0$ and $\rho = 0.5$.

Figure 4.8: Performance comparisons among the three receivers under different cross-correlation coefficient with $E_b/N_0 = 13.35$ dB, $L = 3$, $M = 32$, $\varepsilon_M = 1$, $\varepsilon_P = 0$ and $\epsilon = 1$.

Figure 4.9: BER performance of the proposed ML receiver under different diversity levels with $E_b/N_0 = 13.35$ dB, $M = 32$, $\varepsilon_M = 1$, $\varepsilon_P = 0$, $\epsilon = 1$ and $\rho = 0.5$.

Figure 4.10: BER performance of the product-combining receiver under different diversity levels with $E_b/N_0 = 13.35$ dB, $M = 32$, $\varepsilon_M = 1$, $\varepsilon_P = 0$, $\epsilon = 1$ and $\rho = 0.5$.

Figure 4.11: BER performance of the linear-combining receiver under different diversity levels with $E_b/N_0 = 13.35$ dB, $M = 32$, $\varepsilon_M = 1$, $\varepsilon_P = 0$, $\epsilon = 1$ and $\rho = 0.5$.

4.4 Effects of Timing and Frequency Offsets on FFH/MFSK System Performance

In this section, we will focus on the effects of timing and frequency offsets on performance of FFH systems with both MTJ and PBNJ.

4.4.1 System Model

The transmitter and receiver structures have been presented in Chapter 2. The hopping pattern of the desired signal is assumed to be first-order Markov with two consecutive hopping frequencies being different [95]. It is assumed that the receiver does not have the perfect knowledge of the hopping time and hopping frequency of the received signal. Therefore, when the exact time instant at which the signal changes its frequency is not known precisely, the receiver will use its own imperfect estimated epoch, which is offset from the true value by ΔT . It is assumed that the timing offset ΔT is less than one hopping duration [94]. Under this condition, if the timing offset is delayed by ΔT with respect to the true value, the de-hopped signal $r(t)$ of the l th hop in the interval $(\Delta T + (l - 1)T_h, \Delta T + lT_h)$ can be expressed as

$$r(t) = \begin{cases} s(t) + w(t) + q_{\alpha l}J_1(t) + q_l n_J(t), & \Delta T + (l - 1)T_h \leq t < lT_h \\ w(t) + q'_{\beta l}J_2(t) + q'_l n_J(t), & lT_h \leq t \leq \Delta T + lT_h \end{cases} \quad (4.43)$$

where $s(t)$ is the desired baseband signal, $w(t)$ is AWGN with equivalent one-sided PSD of N_0 , $J_1(t)$ and $J_2(t)$ represent MTJ, and $n_J(t)$ is PBNJ. The term $q_{\alpha l} = 1$ or 0 is a parameter indicating whether the α th branch of the l th hop is jammed by MTJ or not in the time duration $(\Delta T + (l - 1)T_h, lT_h)$, and $q'_{\beta l} = 1$ or 0 denotes whether the β th branch of the l th hop is jammed by MTJ or not in the time duration

$(lT_h, \Delta T + lT_h)$. Similarly, the indicators $q_l = 1$ or 0 and $q'_l = 1$ or 0 denote whether the hop is jammed by PBNJ or not in the time duration $(\Delta T + (l-1)T_h, lT_h)$ and $(lT_h, \Delta T + lT_h)$, respectively. The desired signal is assumed to pass through a frequency non-selective Rician-fading channel, which can be modeled as the sum of an LOS component and a Rayleigh-fading component [10], [96]. Hence, after the de-hopping process, the desired baseband signal $s(t)$ is expressed as

$$s(t) = \sqrt{2}A_l \cos(2\pi(f_n + \Delta f)t + \phi_l) + \sqrt{2}|a_l| \cos(2\pi(f_n + \Delta f)t + \theta_l) \quad (4.44)$$

where $\sqrt{2}A_l$ and $\sqrt{2}|a_l|$ denote the amplitudes of the LOS and Rayleigh-faded components, respectively, f_n is the baseband frequency, and Δf denotes the frequency offset due to the imperfect estimate of the hopping frequency. It is assumed that the frequency offset, Δf , remain unchanged over one symbol duration, and it is also assumed to be less than the minimum frequency separation between any two adjacent branches [94]. The two random phases, ϕ_l and θ_l , are assumed to be uniformly distributed over $[-\pi, \pi]$. Note that the term a_l can be expressed as $a_l = \sigma_s(d_l + i e_l)$, where d_l and e_l are independent real-valued Gaussian random variables with zero mean and unit variance. The diffused power level of the signal is $2\sigma_s^2$, and the specular-to-diffused power ratio of the signal is defined as $F_S = A_l^2/(2\sigma_s^2)$. We also assume that both the desired signal and MTJ experience independent frequency non-selective Rician-fading. Therefore, the jamming tones $J_1(t)$ and $J_2(t)$ shown in (4.43) can be expressed as

$$J_1(t) = \sqrt{2}A_{J_1l} \cos(2\pi(f_\alpha + \Delta f)t + \phi_{J_1l}) + \sqrt{2}|a_{J_1l}| \cos(2\pi(f_\alpha + \Delta f)t + \theta_{J_1l}),$$

for $\alpha = 1, 2, \dots, M$ (4.45)

and

$$J_2(t) = \sqrt{2}A_{J_2l} \cos(2\pi(f_\beta + \Delta f)t + \phi_{J_2l}) + \sqrt{2}|a_{J_2l}| \cos(2\pi(f_\beta + \Delta f)t + \theta_{J_2l}),$$

$$\text{for } \beta = 1, 2, \dots, M \quad (4.46)$$

respectively. Note that all the terms in (4.45) and (4.46) have the same meanings as their counterparts in (4.44), except that they are for MTJ. The terms a_{J_1l} and a_{J_2l} can also be expressed as $a_{J_1l} = \sigma_M(d_{J_1l} + i e_{J_1l})$ and $a_{J_2l} = \sigma_M(d_{J_2l} + i e_{J_2l})$, respectively, where d_{J_1l} , e_{J_1l} , d_{J_2l} and e_{J_2l} are independent real-valued Gaussian random variables with zero mean and unit variance, and the diffused power level of MTJ is $2\sigma_M^2$. The specular-to-diffused power ratio of MTJ is defined as $F_J = A_{J_1l}^2/(2\sigma_M^2)$.

4.4.2 Error Probability Analysis

In this section, analytical BER expressions of the linear-combining and product-combining receivers will be derived for FFH/MFSK systems with timing and frequency offsets and the composite effect of MTJ and PBNJ. The de-hopped signal will be fed into the square-law detector. After some mathematical simplifications, the in-phase component $\eta_{I,ml}$ and quadrature-phase component $\eta_{Q,ml}$ of the square-law

detector output r_{ml} are obtained as

$$\begin{aligned}
 \eta_{I,ml} = & A_l(1 - \Delta T_N) \cos(\pi(1 + \Delta T_N)(\Delta F_N + (f_n - f_m)T_h) + \phi_l) \\
 & \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_n - f_m)T_h)) \\
 & + \frac{\sigma_s d_l (\sin(2\pi(\Delta F_N + (f_n - f_m)T_h)) - \sin(2\pi\Delta T_N(\Delta F_N + (f_n - f_m)T_h)))}{2\pi(\Delta F_N + (f_n - f_m)T_h)} \\
 & + \frac{\sigma_s e_l (\cos(2\pi(\Delta F_N + (f_n - f_m)T_h)) - \cos(2\pi\Delta T_N(\Delta F_N + (f_n - f_m)T_h)))}{2\pi(\Delta F_N + (f_n - f_m)T_h)} \\
 & + q_{\alpha l} \left[A_{J_1 l}(1 - \Delta T_N) \cos(\pi(1 + \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h) + \phi_{J_1 l}) \right. \\
 & \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h)) \\
 & + \frac{\sigma_M d_{J_1 l} (\sin(2\pi(\Delta F_N + (f_\alpha - f_m)T_h)) - \sin(2\pi\Delta T_N(\Delta F_N + (f_\alpha - f_m)T_h)))}{2\pi(\Delta F_N + (f_\alpha - f_m)T_h)} \\
 & \left. + \frac{\sigma_M e_{J_1 l} (\cos(2\pi(\Delta F_N + (f_\alpha - f_m)T_h)) - \cos(2\pi\Delta T_N(\Delta F_N + (f_\alpha - f_m)T_h)))}{2\pi(\Delta F_N + (f_\alpha - f_m)T_h)} \right] \\
 & + q'_{\beta l} \left[A_{J_2 l} \Delta T_N \cos(\pi(2 + \Delta T_N)(\Delta F_N + (f_\beta - f_m)T_h) + \phi_{J_2 l}) \right. \\
 & \times \text{sinc}(\Delta T_N(\Delta F_N + (f_\beta - f_m)T_h)) \\
 & + \frac{\sigma_M d_{J_2 l} (\sin(2\pi(1 + \Delta T_N)(\Delta F_N + (f_\beta - f_m)T_h)) - \sin(2\pi(\Delta F_N + (f_\alpha - f_m)T_h)))}{2\pi(\Delta F_N + (f_\beta - f_m)T_h)} \\
 & \left. + \frac{\sigma_M e_{J_2 l} (\cos(2\pi(1 + \Delta T_N)(\Delta F_N + (f_\beta - f_m)T_h)) - \cos(2\pi(\Delta F_N + (f_\beta - f_m)T_h)))}{2\pi(\Delta F_N + (f_\beta - f_m)T_h)} \right] \\
 & + n_I
 \end{aligned} \tag{4.47}$$

and

$$\begin{aligned}
 \eta_{Q,ml} = & -A_l(1 - \Delta T_N) \sin(\pi(1 + \Delta T_N)(\Delta F_N + (f_n - f_m)T_h) + \phi_l) \\
 & \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_n - f_m)T_h)) \\
 & + \frac{\sigma_s d_l (\cos(2\pi(\Delta F_N + (f_n - f_m)T_h)) - \cos(2\pi\Delta T_N(\Delta F_N + (f_n - f_m)T_h)))}{2\pi(\Delta F_N + (f_n - f_m)T_h)} \\
 & - \frac{\sigma_s e_l (\sin(2\pi(\Delta F_N + (f_n - f_m)T_h)) - \sin(2\pi\Delta T_N(\Delta F_N + (f_n - f_m)T_h)))}{2\pi(\Delta F_N + (f_n - f_m)T_h)} \\
 & - q_{\alpha l} \left[A_{J_{1l}}(1 - \Delta T_N) \sin(\pi(1 + \Delta T_N)(\Delta F_N + (f_{\alpha} - f_m)T_h) + \phi_{J_{1l}}) \right. \\
 & \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_{\alpha} - f_m)T_h)) \\
 & - \frac{\sigma_M d_{J_{1l}} (\cos(2\pi(\Delta F_N + (f_{\alpha} - f_m)T_h)) - \cos(2\pi\Delta T_N(\Delta F_N + (f_{\alpha} - f_m)T_h)))}{2\pi(\Delta F_N + (f_{\alpha} - f_m)T_h)} \\
 & \left. + \frac{\sigma_M e_{J_{1l}} (\sin(2\pi(\Delta F_N + (f_{\alpha} - f_m)T_h)) - \sin(2\pi\Delta T_N(\Delta F_N + (f_{\alpha} - f_m)T_h)))}{2\pi(\Delta F_N + (f_{\alpha} - f_m)T_h)} \right] \\
 & - q'_{\beta l} \left[A_{J_{2l}}\Delta T_N \sin(\pi(2 + \Delta T_N)(\Delta F_N + (f_{\beta} - f_m)T_h) + \phi_{J_{2l}}) \right. \\
 & \times \text{sinc}(\Delta T_N(\Delta F_N + (f_{\beta} - f_m)T_h)) \\
 & - \frac{\sigma_M d_{J_{2l}} (\cos(2\pi(1 + \Delta T_N)(\Delta F_N + (f_{\beta} - f_m)T_h)) - \cos(2\pi(\Delta F_N + (f_{\alpha} - f_m)T_h)))}{2\pi(\Delta F_N + (f_{\beta} - f_m)T_h)} \\
 & \left. + \frac{\sigma_M e_{J_{2l}} (\sin(2\pi(1 + \Delta T_N)(\Delta F_N + (f_{\beta} - f_m)T_h)) - \sin(2\pi(\Delta F_N + (f_{\beta} - f_m)T_h)))}{2\pi(\Delta F_N + (f_{\beta} - f_m)T_h)} \right] \\
 & - n_Q
 \end{aligned} \tag{4.48}$$

respectively. Note that $\Delta T_N = \Delta T/T_h$ and $\Delta F_N = \Delta f T_h$ represent the normalized timing and frequency offsets, respectively. The function $\text{sinc}(x)$ is defined as $\text{sinc}(x) = \sin(\pi x)/(\pi x)$ [10], and n_I and n_Q are the Gaussian distributed random variables with zero mean and variance $\sigma_T^2 = \sigma_w^2 + (q_l(1 - \Delta T_N) + q'_l \Delta T_N)^2 \sigma_P^2$. Both the in-phase component $\eta_{I,ml}$ and quadrature-phase component $\eta_{Q,ml}$ are independent

Gaussian distributed with mean values of

$$\begin{aligned}
 m_{\eta_{I,ml}} = & A_l(1 - \Delta T_N) \cos(\pi(1 + \Delta T_N)(\Delta F_N + (f_n - f_m)T_h) + \phi_l) \\
 & \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_n - f_m)T_h)) \\
 & + q_{\alpha l} \left[A_{J_1 l}(1 - \Delta T_N) \cos(\pi(1 + \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h) + \phi_{J_1 l}) \right. \\
 & \left. \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h)) \right] \\
 & + q'_{\beta l} \left[A_{J_2 l} \Delta T_N \cos(\pi(2 + \Delta T_N)(\Delta F_N + (f_\beta - f_m)T_h) + \phi_{J_2 l}) \right. \\
 & \left. \times \text{sinc}(\Delta T_N(\Delta F_N + (f_\beta - f_m)T_h)) \right] \quad (4.49)
 \end{aligned}$$

and

$$\begin{aligned}
 m_{\eta_{Q,ml}} = & -A_l(1 - \Delta T_N) \sin(\pi(1 + \Delta T_N)(\Delta F_N + (f_n - f_m)T_h) + \phi_l) \\
 & \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_n - f_m)T_h)) \\
 & - q_{\alpha l} \left[A_{J_1 l}(1 - \Delta T_N) \sin(\pi(1 + \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h) + \phi_{J_1 l}) \right. \\
 & \left. \times \text{sinc}((1 - \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h)) \right] \\
 & - q'_{\beta l} \left[A_{J_2 l} \Delta T_N \sin(\pi(2 + \Delta T_N)(\Delta F_N + (f_\beta - f_m)T_h) + \phi_{J_2 l}) \right. \\
 & \left. \times \text{sinc}(\Delta T_N(\Delta F_N + (f_\beta - f_m)T_h)) \right] \quad (4.50)
 \end{aligned}$$

respectively, and a common variance

$$\begin{aligned}
 \sigma_{ml}^2 = & \sigma_s^2(1 - \Delta T_N)^2 \text{sinc}^2((1 - \Delta T_N)(\Delta F_N + (f_n - f_m)T_h)) \\
 & + q_{\alpha l} [\sigma_M^2(1 - \Delta T_N)^2 \text{sinc}^2((1 - \Delta T_N)(\Delta F_N + (f_\alpha - f_m)T_h))] \\
 & + q'_{\beta l} [\sigma_M^2 \Delta T_N^2 \text{sinc}^2(\Delta T_N(\Delta F_N + (f_\beta - f_m)T_h))] + \sigma_T^2. \quad (4.51)
 \end{aligned}$$

Therefore, the square-law detector output r_{ml} has a noncentral chi-square distribution with two degrees of freedom and noncentrality parameter denoted as $A_{ml}^2 = m_{\eta_{I,ml}}^2 + m_{\eta_{Q,ml}}^2$ [1]. The conditional pdf expression of r_{ml} is expressed as [1]

$$p_{r_{ml}}(r_{ml}|f_n, q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \phi_l, \phi_{J_1 l}, \phi_{J_2 l}) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml} + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{\sqrt{r_{ml}} A_{ml}}{\sigma_{ml}^2}\right) \times U(r_{ml}). \quad (4.52)$$

Note that (4.51) and (4.52) are also valid for the case when the timing offset is advanced by ΔT .

A) Linear-Combining Receiver

The corresponding conditional characteristic function for the linear-combining receiver is

$$\Phi_{z_{ml}}(iv|f_n, q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \phi_l, \phi_{J_1 l}, \phi_{J_2 l}) = \frac{1}{1 - 2iv\sigma_{ml}^2} \exp\left(\frac{ivA_{ml}^2}{1 - 2iv\sigma_{ml}^2}\right). \quad (4.53)$$

The final-decision variable for the linear-combining receiver is $y_m = \sum_{l=1}^L z_{ml}$, and its corresponding characteristic function can be obtained as

$$\Phi_{y_m}(iv|f_n, \mathbf{J}_1, \mathbf{J}_2, \mathbf{\Phi}) = \prod_{l=1}^L \Phi_{z_{ml}}(iv|f_n, q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \phi_l, \phi_{J_1 l}, \phi_{J_2 l}) \quad (4.54)$$

where $\mathbf{J}_1 = (\mathbf{q}_1, \mathbf{q}_2, \dots, \mathbf{q}_L, \mathbf{q}'_1, \mathbf{q}'_2, \dots, \mathbf{q}'_L)$ with $\mathbf{q}_l = (q_{1l}, q_{2l}, \dots, q_{Ml})^T$ and $\mathbf{q}'_l = (q'_{1l}, q'_{2l}, \dots, q'_{Ml})^T$. Note that as the single-tone per-band MTJ is considered in this thesis, there is at most one element of $\mathbf{q}_l$ or $\mathbf{q}'_l$ equal to one. Similarly, $\mathbf{J}_2 = (\mathbf{q}, \mathbf{q}')$, where $\mathbf{q} = (q_1, q_2, \dots, q_L)^T$ and $\mathbf{q}' = (q'_1, q'_2, \dots, q'_L)^T$, and $\mathbf{\Phi} = (\Phi_1, \Phi_2, \Phi_3)$ with $\Phi_1 = (\phi_1, \phi_2, \dots, \phi_L)^T$, $\Phi_2 = (\phi_{J_1 1}, \phi_{J_1 2}, \dots, \phi_{J_1 L})^T$, and

$\Phi_3 = (\phi_{J_21}, \phi_{J_22}, \dots, \phi_{J_2L})^T$. Taking the inverse transformation of the characteristic function $\Phi_{y_m}(iv|f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi)$, we obtain the pdf expression of the final-decision variable. The conditional probability of symbol error is

$$\begin{aligned} P_s(f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi) &= 1 - \int_0^\infty \left[\int_0^{y_n} p_{y_1}(y_1|f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi) dy_1 \times \dots \times \int_0^{y_n} p_{y_{n-1}}(y_{n-1}|f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi) dy_{n-1} \right. \\ &\quad \times \int_0^{y_n} p_{y_{n+1}}(y_{n+1}|f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi) dy_{n+1} \times \dots \times \int_0^{y_n} p_{y_M}(y_M|f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi) dy_M \\ &\quad \left. \times p_{y_n}(y_n|f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi) dy_n \right] \end{aligned} \quad (4.55)$$

Taking the expectation over Φ , we obtain

$$P_s(f_n, \mathbf{J}_1, \mathbf{J}_2) = E_\Phi \{P_s(f_n, \mathbf{J}_1, \mathbf{J}_2, \Phi)\}. \quad (4.56)$$

Finally, the average BER is [1]

$$P_b = \frac{1}{2(M-1)} \sum_{n=1}^M P_s(f_n) \quad (4.57)$$

with the average probability of symbol error given by

$$\begin{aligned} P_s(f_n) &= \sum_{\text{all } \mathbf{J}_1, \mathbf{J}_2} P_s(f_n, \mathbf{J}_1, \mathbf{J}_2) \left[\prod_{l=1}^L \left(\frac{Q}{MN} \right)^{\sum_{m=1}^M q_{ml}} \left(1 - \frac{Q}{N} \right)^{1 - \sum_{m=1}^M q_{ml}} \right] \\ &\quad \times \left[\prod_{l=1}^L \left(\frac{Q - q_{m(l+1)}}{M(N-1)} \right)^{\sum_{m=1}^M q'_{ml}} \left(1 - \frac{Q - q_{m(l+1)}}{N-1} \right)^{1 - \sum_{m=1}^M q'_{ml}} \right] \\ &\quad \times \left[\gamma^{\sum_{l=1}^L q_l} (1 - \gamma)^{L - \sum_{l=1}^L q_l} \left[\prod_{l=1}^L \left(\frac{\gamma - q_{l+1}/N}{1 - 1/N} \right)^{q'_l} \left(1 - \frac{\gamma - q_{l+1}/N}{1 - 1/N} \right)^{1 - q'_l} \right] \right]. \end{aligned} \quad (4.58)$$

To numerically evaluate (4.58), we should consider all the possible combinations of $\mathbf{J}_1$ and $\mathbf{J}_2$.

B) Product-Combining Receiver

The decision variable for the product-combining receiver is $z_{ml} = \ln(r_{ml})$, and its pdf expression is given by

$$p_{z_{ml}}(z_{ml}|f_n, q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \phi_l, \phi_{J_1 l}, \phi_{J_2 l}) = \frac{e^{z_{ml}}}{2\sigma_{ml}^2} \exp\left(-\frac{e^{z_{ml}} + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{\sqrt{e^{z_{ml}}} A_{ml}}{\sigma_{ml}^2}\right). \quad (4.59)$$

Following that, we obtain the conditional characteristic function of z_{ml} as [90]

$$\begin{aligned} \Phi_{z_{ml}}(iv|f_n, q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \phi_l, \phi_{J_1 l}, \phi_{J_2 l}) \\ = (2\sigma_{ml}^2)^{iv} \exp\left(-\frac{A_{ml}^2}{2\sigma_{ml}^2}\right) \sum_{a=0}^{\infty} \frac{\Gamma(iv + 1 + a)}{(a!)^2} \left(\frac{A_{ml}^2}{2\sigma_{ml}^2}\right)^a. \end{aligned} \quad (4.60)$$

Finally, following the similar steps shown in (4.54)-(4.57), we can obtain the average BER expression of the product-combining receiver.

4.4.3 Numerical Results and Discussions

In this section, we present BER results for FFH/MFSK systems with timing and frequency offsets over frequency non-selective Rician-fading channels when both MTJ and PBNJ coexist.

The worst-case analytical and simulation BER results for the linear-combining and product-combining receivers with $E_b/N_0 = 20$ dB, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 0.5$, $\varepsilon_P = 0.5$, $\Delta F_N = 0.2$ and $\Delta T_N = 0.2$ are shown in Fig. 4.12. The close

Figure 4.12: Worst-case analytical and simulation results for the linear-combining and product-combining receivers with $E_b/N_0 = 20$ dB, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 0.5$, $\varepsilon_P = 0.5$, $\Delta F_N = 0.2$ and $\Delta T_N = 0.2$.

match between the analytical and simulation results validates the analytical BER expressions derived in Section 4.4.2. It is also observed that the product-combining receiver outperforms the linear-combining receiver at low E_b/N_J levels (i.e., $E_b/N_J < 18$). This is because when the jamming power is relatively strong, more jamming is present and it can do more harm to the linear-combining receiver because it can place a lot more interference power in the final decision statistics through the addition operation. For the product-combining receiver, the corresponding effect on the final decision statistics is limited due to the nature of the multiplication operation. On the other hand, the linear-combining receiver outperforms the product-combining receiver when jamming power is low, as the error is mainly caused by AWGN.

Figures 4.13 and 4.14 illustrate the BER performance of the linear-combining and product-combining receivers, respectively, over Rician-fading channels with different

Figure 4.13: BER performance of the linear-combining receiver with $E_b/N_0 = 12$ dB, $M = 4$, $F_S = 10$ and $L = 3$.

timing and frequency offsets, and different jamming power ratios. It is observed from Fig. 4.13 that for the linear-combining receiver, when only one type of jamming waveform is present in the system (i.e., either MTJ or PBNJ), MTJ is more harmful to the system than that of PBNJ. This trend of mismatch in performance is abated as the timing and frequency offsets increase due to the fact that the effects of timing and frequency offsets on system performance dominate. At high and moderate jamming powers (i.e., $E_b/N_J < 35$ dB), the system performance under MTJ is worse than that of PBNJ. On the other hand, when E_b/N_J level is high (i.e., $E_b/N_J > 35$ dB), the jamming power is low and the probability of error is mainly caused by AWGN. Thus, the BER results under either MTJ or PBNJ condition are similar. When the product-combining receiver is employed, we observe from Fig. 4.14 that the performance under MTJ only is slightly better than that under PBNJ only, which is opposite to that when the linear-combining receiver is employed. However, the

Figure 4.14: BER performance of the product-combining receiver with $E_b/N_0 = 12$ dB, $M = 4$, $F_S = 10$ and $L = 3$.

trend of mismatch is the same as that of the linear-combining receiver. When both MTJ and PBNJ are present, it is found that the BER performance is bounded by the two extreme cases when either MTJ or PBNJ is present.

In Fig. 4.15, we present the effect of timing offset only on the BER performance of both linear-combining and product-combining receivers. It is found that, in general, larger timing offset results in worse BER performance for both receivers. This is because a portion of signal energy is lost at the detector output when there is a timing offset. Moreover, the timing offset also distorts the orthogonality properties among the M frequency branches, which in turn introduces cross interference to the signal branch. As the timing offset increases, the energy loss of the signal branch and the cross interference increase. Therefore, the BER performance worsens as the timing offset increases. Comparing between the BER performance of both receivers, it is observed that the product-combining receiver is more sensitive to the timing

Figure 4.15: Performance comparisons between the linear-combining and product-combining receivers under different normalized timing offsets with $E_b/N_0 = 12$ dB, $L = 3$, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 1$, $\varepsilon_P = 0$ and $\Delta F_N = 0$.

offset than that of the linear-combining receiver. For example, to obtain a BER value of 2×10^{-3} at $\Delta T_N = 0.1$, around 1 dB more in E_b/N_J is required for the linear-combining receiver as compared to the one without any timing offset, whereas the product-combining receiver requires 8 dB more. Moreover, it is also found that when the jamming power is high (i.e., $E_b/N_J < 25$ dB), the BER performance of the linear-combining receiver is not quite sensitive to the timing offset under the condition that the normalized timing offset ΔT_N is smaller than 0.2. This is because under this condition, the jamming effect becomes dominant. Therefore, the timing offset has less effect on the BER performance for the linear-combining receiver. When the jamming power is weak, we can observe that there is an error floor as the effect of timing offset becomes dominant. For large timing offsets, the BER performance is poor for both receivers.

Figure 4.16: Performance comparisons between the linear-combining and product-combining receivers under different normalized frequency offsets with $E_b/N_0 = 12$ dB, $L = 3$, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 1$, $\varepsilon_P = 0$ and $\Delta T_N = 0$.

In Fig. 4.16, we show the effect of frequency offset only on the BER performance of both receivers. From this figure, we can observe that higher frequency offset leads to worse BER performance for both receivers. When the frequency offset is low, i.e., $\Delta F_N = 0.1$, both receivers are not sensitive to the frequency offset. When $\Delta F_N > 0.1$, it is observed that the product-combining receiver is more sensitive to the frequency offset as compared to the linear-combining receiver.

The effects of either timing or frequency offset on the BER performance of the linear-combining receiver are shown in Fig. 4.17, and the corresponding results for the product-combining receiver are shown in Fig. 4.18. It is observed that both systems are more sensitive to the timing offset as compared to the frequency offset when the normalized offset value is less than 0.3. However, the reverse is true when the normalized offset value is greater than 0.4.

Figure 4.17: Performance comparisons under either timing or frequency offset alone for the linear-combining receiver with $E_b/N_0 = 12$ dB, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 1$, $\varepsilon_P = 0$ and $L = 3$

Figure 4.18: Performance comparisons under either timing or frequency offset alone for the product-combining receiver with $E_b/N_0 = 12$ dB, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 1$, $\varepsilon_P = 0$ and $L = 3$

Figure 4.19: Performance comparisons between the linear-combining and product-combining receivers under both timing and frequency offsets with $E_b/N_0 = 12$ dB, $M = 4$, $F_S = 10$, $F_J = 10$, $\varepsilon_M = 1$, $\varepsilon_P = 0$ and $L = 3$.

Figure 4.19 shows the effects of both timing and frequency offsets on the BER performance of the linear-combining and product-combining receivers. Similar to Figs. 4.15 and 4.17, it is also observed that higher timing and frequency offsets result in worse BER performance. Moreover, by comparing the results in Figs. 4.15, 4.17 and 4.19, we observe that the degradation of BER due to both timing and frequency offsets is larger than the degradation due to either timing or frequency offset alone.

Tables 4.2 and 4.3 show some typical optimum diversity orders for the linear-combining receiver with different normalized timing and frequency offsets under Rician-fading ($F = 10$) and Rayleigh-fading ($F = 0$) conditions, respectively. Tables 4.4 and 4.5 show the corresponding results for the product-combining receiver. From Tables 4.2 and 4.3, it is found that for the linear-combining receiver, the optimum

Table 4.2: Optimum diversity orders of the linear-combining receiver under different normalized offsets with $E_b/N_0 = 13.35$ dB, $M = 4$, $F_S = 10$, $\varepsilon_M = 0$, $\varepsilon_P = 1$ and $E_b/N_J = 20$ dB.

Normalized offsets	$\Delta F_N = 0$	$\Delta F_N = 0.1$	$\Delta F_N = 0.2$	$\Delta F_N = 0.3$
$\Delta T_N = 0$	$L = 1$	$L = 1$	$L = 1$	$L = 2$
$\Delta T_N = 0.1$	$L = 1$	$L = 1$	$L = 2$	$L = 3$
$\Delta T_N = 0.2$	$L = 2$	$L = 2$	$L = 3$	$L = 4$
$\Delta T_N = 0.3$	$L = 2$	$L = 3$	$L = 4$	$L = 4$

diversity order increases as the normalized timing or frequency offset increases under both Rician-fading and Rayleigh-fading environments. This is due to the fact that for the linear-combining receiver with timing and frequency offsets, higher diversity orders can counteract those effects more efficiently. In addition, we also find that under the same timing and frequency offset condition, the optimum diversity order for the linear-combining receiver under Rayleigh-fading channels is higher than that under Rician-fading channels. This is because a higher diversity order is required to counteract the more severe fading effects. When the product-combining receiver is employed, it is interesting to see that the optimum diversity order decreases as the normalized timing or frequency offset increases. This observation is just the opposite to that of the linear-combining receiver. This is because for the product-combining receiver with timing and frequency offsets, the non-coherent diversity-combining loss will gradually dominate as the offset value increases. Lastly, similar to the linear-combining receiver, we observe from Tables 4.4 and 4.5 that the optimum diversity order for the product-combining receiver is also higher under Rayleigh-fading channels as compared to that under Rician-fading channels.

Table 4.3: Optimum diversity orders of the linear-combining receiver under different normalized offsets with $E_b/N_0 = 13.35$ dB, $M = 4$, $F_S = 0$, $\varepsilon_M = 0$, $\varepsilon_P = 1$ and $E_b/N_J = 20$ dB.

Normalized offsets	$\Delta F_N = 0$	$\Delta F_N = 0.1$	$\Delta F_N = 0.2$	$\Delta F_N = 0.3$
$\Delta T_N = 0$	$L = 3$	$L = 3$	$L = 3$	$L = 5$
$\Delta T_N = 0.1$	$L = 4$	$L = 6$	$L = 6$	$L = 8$
$\Delta T_N = 0.2$	$L = 5$	$L = 6$	$L = 7$	$L = 9$
$\Delta T_N = 0.3$	$L = 6$	$L = 7$	$L = 8$	$L = 10$

Table 4.4: Optimum diversity orders of the product-combining receiver under different normalized offsets with $E_b/N_0 = 13.35$ dB, $M = 4$, $F_S = 10$, $\varepsilon_M = 0$, $\varepsilon_P = 1$ and $E_b/N_J = 20$ dB.

Normalized offsets	$\Delta F_N = 0$	$\Delta F_N = 0.1$	$\Delta F_N = 0.2$	$\Delta F_N = 0.3$
$\Delta T_N = 0$	$L = 5$	$L = 5$	$L = 5$	$L = 5$
$\Delta T_N = 0.1$	$L = 5$	$L = 5$	$L = 4$	$L = 4$
$\Delta T_N = 0.2$	$L = 4$	$L = 4$	$L = 3$	$L = 3$
$\Delta T_N = 0.3$	$L = 3$	$L = 3$	$L = 3$	$L = 3$

Table 4.5: Optimum diversity orders of the product-combining receiver under different normalized offsets with $E_b/N_0 = 13.35$ dB, $M = 4$, $F_S = 0$, $\varepsilon_M = 0$, $\varepsilon_P = 1$ and $E_b/N_J = 20$ dB.

Normalized offsets	$\Delta F_N = 0$	$\Delta F_N = 0.1$	$\Delta F_N = 0.2$	$\Delta F_N = 0.3$
$\Delta T_N = 0$	$L = 10$	$L = 8$	$L = 7$	$L = 7$
$\Delta T_N = 0.1$	$L = 10$	$L = 8$	$L = 6$	$L = 6$
$\Delta T_N = 0.2$	$L = 8$	$L = 7$	$L = 6$	$L = 6$
$\Delta T_N = 0.3$	$L = 7$	$L = 6$	$L = 5$	$L = 5$

4.5 Conclusions

In this chapter, the composite effect of MTJ and PBNJ on FFH/MFSK system performance has been investigated. We have proposed an optimum ML receiver for FFH/MFSK systems over frequency non-selective Rayleigh-fading channels when both MTJ and PBNJ are present concurrently. Numerical results have shown that the proposed ML receiver can better counteract MTJ than PBNJ. When both types of jamming waveforms are present, the BER performance is bounded by the two extreme cases, where either MTJ or PBNJ is present. In addition, we have also extended the analysis to a frequency-selective Rayleigh-fading channel. Three types of receivers, namely, the ML, linear-combining and product-combining receivers have been studied. Lastly, we have investigated the effects of timing and frequency offsets on FFH/MFSK system performance over frequency non-selective Rician-fading channels with both MTJ and PBNJ. New analytical BER expressions are derived for both linear-combining and product-combining receivers. It is observed that FFH/MFSK systems are sensitive to the timing and frequency offsets. The product-combining receiver is also found to be more sensitive to the timing

and frequency offsets than the linear-combining receiver. Moreover, MTJ is more harmful to the system performance than PBNJ for the linear-combining receiver; while PBNJ is more harmful to the product-combining receiver under frequency non-selective Rician-fading channel conditions. The optimum diversity order of the linear-combining receiver increases as timing and frequency offsets increase, whereas the reverse is true for the product-combining receiver.

Chapter 5

Error Probability Analysis for Synchronous FFH/MA Systems over Fading Channels

In this chapter, we first derive a simple pdf expression of the square-law detector output for synchronous FFH/MA communication systems with MFSK modulation over frequency non-selective Rician-fading channels. Next, an optimum ML receiver is proposed based on the pdf expression derived. In addition, we propose two sub-optimum ML receivers, namely, the ML-A and ML-B receivers, and derive their corresponding analytical BER expressions, which are validated by simulation results. Furthermore, the linear-combining, product-combining and clipper receivers are also investigated and compared with the proposed optimum and suboptimum ML receivers.

5.1 Introduction

In the previous two chapters, we have investigated the effects of MTJ and PBNJ on FFH/MFSK system performance. In this chapter, we will focus on mitigating the effect of MAI. The effects of MAI on various FH/MA systems have been studied in [32], [49], [52], [78], [81], and [97]-[99]. Yue [32] studied the hard-limited and linear-combining receivers for synchronous FFH/MA communication systems over Rayleigh-fading channels. The upper bounds on BER performance of those two receivers were also presented. The ML receiver for synchronous FFH/MA systems with MFSK and differential phase-shift keying schemes over Rayleigh-fading channels were investigated in [52], in which the union upper bound on BER performance for the ML receiver was also presented. In [49], the authors investigated the linear-combining, product-combining and clipper receivers for FFH/MA systems over Rician-fading channels. The pdf expression of the square-law detector output was obtained by using Fourier-Bessel series and Hankel characteristic function, and the corresponding characteristic functions derived for the square-law detector output are quite complex to compute the corresponding analytical BER results. Moreover, the pdf expression presented in [49] cannot be used to derive the ML receiver structure. Solaiman, Glavieux and Hillion [78] investigated the equal-gain diversity improvement for FFH/MA systems under Rayleigh-fading environments. The Gaussian quadrature rule and a moment generation algorithm were adopted in [81] to calculate the error probability for asynchronous FH/MA systems. The symbol error probabilities for both synchronous and asynchronous FH/MA systems over Rayleigh-fading channels via semi-analytical Monte Carlo simulation were presented in [97], where only one symbol per hop was considered. In [98], the author presented an accurate approximation and tight upper bound on BER performance for an asynchronous FH/MA system with binary modulation. When the modulation level M is

greater than 2, only the tight upper bound was shown. Using the Stirling's formula, the average BER expressions were derived in [99] for synchronous FH/MA systems over Rayleigh-fading, Rician-fading and logNormal fading channels.

In this chapter, we will first derive a simple pdf expression of the square-law detector output of synchronous FFH/MA systems with MFSK modulation over a frequency non-selective Rician-fading channel. Then, based on this pdf expression, an optimum ML receiver is proposed. In addition, we also propose two suboptimum ML receivers and derive their corresponding analytical BER expressions, which are validated by simulation results. Furthermore, analytical BER expressions of the linear-combining, product-combining and clipper receivers are also derived. The derived analytical BER expressions can help to provide more insights to the behaviour of FFH/MA systems in very low BER range, where simulations are too time consuming to be performed.

5.2 System Model

In this chapter, we consider an FFH/MA system that operates in a mobile satellite environment characterized by frequency non-selective Rician fading. It is assumed that all the K active users in the synchronous system fade independently and employ random FH patterns [49], [82]. Without loss of generality, we also assume that the first user is the desired user, and a single-user detector is performed to detect the desired user's signal.

At the transmitter side, the binary data bits are first encoded into M -ary sequence by a binary-to- M -ary encoder. After that, each M -ary symbol is fed into an MFSK modulator, which selects one of the M baseband frequency according to

the input symbol. The output of the MFSK modulator is mixed with a frequency synthesizer to be pseudo-randomly hopped over one of the N non-overlapping FH bands according to each individual user's hopping pattern. Following that, all the K active users' signals are then passed through a BPF with bandwidth W_{ss} , and up-converted to a radio frequency before transmission.

At the front end of a typical FFH/MFSK receiver, the received signal, consisting of the desired signal, MAI and AWGN, is down-converted, band-pass filtered and de-hopped. The received signal is assumed to pass through a frequency non-selective Rician-fading channel, which can be modeled as a sum of a dominant LOS component and a Rayleigh-fading component [10]. Hence, the de-hopped signal $r(t)$ of the l th hop is expressed as

$$\begin{aligned} r(t) = & \sqrt{2}A_{sl} \cos(2\pi f_n t + \phi_{sl}) + \sqrt{2}|a_{sl}| \cos(2\pi f_n t + \theta_{sl}) \\ & + \sum_{k=1}^{C_l} \left(\sqrt{2}A_{kl} \cos(2\pi f_{kl} t + \phi_{kl}) + \sqrt{2}|a_{kl}| \cos(2\pi f_{kl} t + \theta_{kl}) \right) + w(t) \end{aligned} \quad (5.1)$$

where $\sqrt{2}A_{sl}$ and $\sqrt{2}|a_{sl}|$ denote the amplitudes of the LOS and Rayleigh-fading components, and f_n is the baseband MFSK modulation frequency of the desired user. The terms ϕ_{sl} and θ_{sl} denote the phase angles of the LOS and diffused components of the desired user, respectively, and they are assumed to be uniformly distributed over $[-\pi, \pi]$ [100], [101]. The parameter C_l denotes the total number of interfering users present in the l th hop. Similarly, the terms $\sqrt{2}A_{kl}$, $\sqrt{2}|a_{kl}|$, f_{kl} , ϕ_{kl} and θ_{kl} have the corresponding meanings as $\sqrt{2}A_{sl}$, $\sqrt{2}|a_{sl}|$, f_n , ϕ_{sl} and θ_{sl} , except that they are for the k -th interfering user in the l th hop. The random noise $w(t)$ is a Gaussian process with zero mean and variance $\sigma_w^2 = N_0 B$. Note that the terms a_{sl} and a_{kl} can be expressed as $a_{sl} = \sigma_s(d_{sl} + i e_{sl})$ and $a_{kl} = \sigma_k(d_{kl} + i e_{kl})$, respectively, where

d_{sl} , e_{sl} , d_{kl} and e_{kl} are the independent real-valued Gaussian random variables with zero mean and unit variance. The diffused power levels of the desired user and the k -th interfering user are denoted as $2\sigma_s^2$ and $2\sigma_k^2$, respectively. It is assumed that all the active users have the same power levels [49], [82], and we define a common specular-to-diffused power ratio as $F = A_{sl}^2/2\sigma_s^2$. In addition, we define a vector $\Phi_l = (\phi_{sl}, \phi_{1l}, \phi_{2l}, \dots, \phi_{C_l l})$ to represent the phase angles of all active users' LOS components in the l th hop.

Following that, the square-law detection is performed to the de-hopped signal, and its output is denoted as r_{ml} . Finally, various diversity-combining techniques will be used to combine the L -hop receptions before making a final decision.

5.3 Proposed ML and Suboptimum ML Receiver Structures

The in-phase component $\eta_{I,ml}$ and quadrature-phase component $\eta_{Q,ml}$ of the square-law detector output r_{ml} can be obtained as

$$\begin{aligned} \eta_{I,ml} = & A_{sl} \cos(\phi_{sl})\delta(f_n - f_m) + d_{sl}\sigma_s\delta(f_n - f_m) + \sum_{k=1}^{C_l} A_{kl} \cos(\phi_{kl})\delta(f_{kl} - f_m) \\ & + \sum_{k=1}^{C_l} d_{kl}\sigma_k\delta(f_{kl} - f_m) + n_I \end{aligned} \quad (5.2)$$

and

$$\begin{aligned} \eta_{Q,ml} = & -A_{sl} \sin(\phi_{sl})\delta(f_n - f_m) - e_{sl}\sigma_s\delta(f_n - f_m) - \sum_{k=1}^{C_l} A_{kl} \sin(\phi_{kl})\delta(f_{kl} - f_m) \\ & - \sum_{k=1}^{C_l} e_{kl}\sigma_k\delta(f_{kl} - f_m) - n_Q \end{aligned} \quad (5.3)$$

respectively, where n_I and n_Q are Gaussian distributed with zero mean and variance σ_w^2 . Both the in-phase component $\eta_{I,ml}$ and quadrature-phase component $\eta_{Q,ml}$ are independent Gaussian distributed with mean values of [1]

$$m_{\eta_{I,ml}} = A_{sl} \cos(\phi_{sl}) \delta(f_n - f_m) + \sum_{k=1}^{C_l} A_{kl} \cos(\phi_{kl}) \delta(f_{kl} - f_m) \quad (5.4)$$

and

$$m_{\eta_{Q,ml}} = -A_{sl} \sin(\phi_{sl}) \delta(f_n - f_m) - \sum_{k=1}^{C_l} A_{kl} \sin(\phi_{kl}) \delta(f_{kl} - f_m) \quad (5.5)$$

respectively, and a common variance

$$\sigma_{ml}^2 = \sigma_s^2 \delta(f_n - f_m) + \sum_{k=1}^{C_l} \sigma_k^2 \delta(f_{kl} - f_m) + \sigma_w^2. \quad (5.6)$$

Therefore, the square-law detector output r_{ml} has a noncentral chi-square distribution with two degrees of freedom and noncentrality parameter denoted as $A_{ml}^2 = m_{\eta_{I,ml}}^2 + m_{\eta_{Q,ml}}^2$ [1]. Hence, we obtain the conditional pdf expression of r_{ml} as

$$p_{r_{ml}}(r_{ml}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml} + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{\sqrt{r_{ml}} A_{ml}}{\sigma_{ml}^2}\right) U(r_{ml}) \quad (5.7)$$

where $\mathbf{f}_l = (f_{1l}, f_{2l}, \dots, f_{C_l l})$ is a vector to represent the baseband frequencies of all the C_l interfering users in the l th hop. The conditional pdf expressions of the signal and non-signal branches are expressed as

$$p_{r_{ml}}(r_{ml}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l, S) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml} + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) I_0\left(\frac{\sqrt{r_{ml}} A_{ml,S}}{\sigma_{ml,S}^2}\right) U(r_{ml}) \quad (5.8)$$

and

$$p_{r_{ml}}(r_{ml}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l, \text{NS}) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml} + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) \times I_0\left(\frac{\sqrt{r_{ml}}A_{ml,NS}}{\sigma_{ml,NS}^2}\right) U(r_{ml}) \quad (5.9)$$

respectively, where

$$\sigma_{ml,S}^2 = \sigma_s^2 + \sum_{k=1}^{C_l} \sigma_k^2 \delta(f_{kl} - f_m) + \sigma_w^2, \quad (5.10)$$

$$A_{ml,S}^2 = \left[A_{sl} \cos(\phi_{sl}) + \sum_{k=1}^{C_l} A_{kl} \cos(\phi_{kl}) \delta(f_{kl} - f_m) \right]^2 + \left[A_{sl} \sin(\phi_{sl}) + \sum_{k=1}^{C_l} A_{kl} \sin(\phi_{kl}) \delta(f_{kl} - f_m) \right]^2, \quad (5.11)$$

$$\sigma_{ml,NS}^2 = \sum_{k=1}^{C_l} \sigma_k^2 \delta(f_{kl} - f_m) + \sigma_w^2, \quad (5.12)$$

and

$$A_{ml,NS}^2 = \left[\sum_{k=1}^{C_l} A_{kl} \cos(\phi_{kl}) \delta(f_{kl} - f_m) \right]^2 + \left[\sum_{k=1}^{C_l} A_{kl} \sin(\phi_{kl}) \delta(f_{kl} - f_m) \right]^2. \quad (5.13)$$

The decision variable of the ML receiver is given by [54]

$$w_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l, S)}{p_{r_{ml}}(r_{ml}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l, \text{NS})} \right). \quad (5.14)$$

Therefore, substituting (5.8) and (5.9) into (5.14), we can obtain the ML decision variable as

$$w_{ml} = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \ln \left[I_0 \left(\frac{\sqrt{r_{ml}} A_{ml,S}}{\sigma_{ml,S}^2} \right) \right] - \ln \left[I_0 \left(\frac{\sqrt{r_{ml}} A_{ml,NS}}{\sigma_{ml,NS}^2} \right) \right] + \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right) + \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2}. \quad (5.15)$$

The decision variables w_{ml} are then summed over all the L -hop receptions to obtain the final-decision variable, which is denoted as $y_m = \sum_{l=1}^L w_{ml}$. To implement the ML receiver, side information on noise variance and the power levels of the direct paths and the diffused paths of all the active users, is required. In practice, the noise variance can be estimated from a “noise-only” channel for noise power measurement [84], the power level can be extracted from an energy detector [11], and the specular-to-diffused power ratio can be obtained by moment-based estimation [102]. Because of the nonlinear operations of the natural logarithm of the modified Bessel function, it is mathematically intractable to derive the analytical BER expressions for the ML receiver shown in (5.15). Therefore, we apply the following approximation [26]

$$\ln[I_0(x)] \approx \begin{cases} x, & \text{if } x > 4 \\ x^2/4, & \text{if } 0 < x \leq 4 \end{cases} \quad (5.16)$$

to simplify the ML receiver structure, and derive two suboptimum ML receivers, namely, the ML-A and ML-B receivers. These two suboptimum ML receivers are simpler and easier to implement in practice, and the corresponding BER expressions can be also derived analytically. When the signal-to-noise ratio (E_b/N_0) and the specular-to-diffused power ratio are relatively high, we can adopt the first ap-

proximation of (5.16) to obtain

$$z_{ml} = \beta_{ml}(\sqrt{r_{ml}} + \Theta_{ml})^2 + U_{ml} \quad (5.17)$$

where

$$\beta_{ml} = \frac{\sigma_{ml,S}^2 - \sigma_{ml,NS}^2}{2\sigma_{ml,S}^2\sigma_{ml,NS}^2}, \quad (5.18)$$

$$\Theta_{ml} = \frac{\sigma_{ml,NS}^2 A_{ml,S} - \sigma_{ml,S}^2 A_{ml,NS}}{\sigma_{ml,S}^2 - \sigma_{ml,NS}^2}, \quad (5.19)$$

and

$$U_{ml} = \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right) - \frac{(A_{ml,S} - A_{ml,NS})^2}{2(\sigma_{ml,S}^2 - \sigma_{ml,NS}^2)}. \quad (5.20)$$

The ML receiver structure shown in (5.17) is called the ML-A receiver in the following analysis. When the E_b/N_0 value and the specular-to-diffused power ratio are relatively low, the second approximation of (5.16) can be used to obtain

$$x_{ml} = \Psi_{ml}r_{ml} + \zeta_{ml} \quad (5.21)$$

where

$$\Psi_{ml} = \frac{\sigma_{ml,S}^2 - \sigma_{ml,NS}^2}{2\sigma_{ml,S}^2\sigma_{ml,NS}^2} + \frac{A_{ml,S}^2}{4\sigma_{ml,S}^4} - \frac{A_{ml,NS}^2}{4\sigma_{ml,NS}^4} \quad (5.22)$$

and

$$\zeta_{ml} = \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right) + \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2}. \quad (5.23)$$

This type of receiver structure is called the ML-B receiver in the subsequent analysis.

5.4 Performance Analysis

In this section, analytical BER expressions of the ML-A, ML-B, linear-combining, product-combining and clipper receivers will be presented. Without loss of generality, we assume that the desired user with frequency f_1 is transmitted.

5.4.1 ML-A Receiver

For the ML-A receiver, the conditional pdf expressions of z_{1l} and z_{pl} , for $p = 2, 3, \dots, M$, can be expressed as

$$p_{z_{1l}}(z_{1l}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}}{2\sigma_{1l,S}^2\sqrt{\beta_{1l}(z_{1l}-U_{1l})}} \exp\left(-\frac{\left(\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}\right)^2 + A_{1l,S}^2}{2\sigma_{1l,S}^2}\right) \\ \times I_0\left(\frac{A_{1l,S}\left(\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}\right)}{\sigma_{1l,S}^2}\right) U\left(\sqrt{\frac{z_{1l}-U_{1l}}{\beta_{1l}}} - \Theta_{1l}\right) \quad (5.24)$$

and

$$p_{z_{pl}}(z_{pl}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}}{2\sigma_{pl,NS}^2\sqrt{\beta_{pl}(z_{pl}-U_{pl})}} \exp\left(-\frac{\left(\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}\right)^2 + A_{pl,NS}^2}{2\sigma_{pl,NS}^2}\right) \\ \times I_0\left(\frac{A_{pl,NS}\left(\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}\right)}{\sigma_{pl,NS}^2}\right) U\left(\sqrt{\frac{z_{pl}-U_{pl}}{\beta_{pl}}} - \Theta_{pl}\right) \quad (5.25)$$

respectively. For the convenience of mathematical analysis, a G -level uniform quantization is used to approximate the decision variable z_{ml} and we denote its output as $\tilde{z}_{ml}$. Therefore, the conditional pdf expression of the discrete random variable $\tilde{z}_{ml}$ is expressed as

$$p_{\tilde{z}_{ml}}(\tilde{z}_{ml}|C_l, \mathbf{f}_l, \Phi_l) = \sum_{g=-G/2}^{G/2} V_{z_{ml}}(g)\delta(\tilde{z}_{ml} - a_g) \quad (5.26)$$

where $a_g = 2gH/G$, for $g = -G/2, -G/2+1, \dots, G/2$, and H is set at a substantially high level such that the probability of z_{ml} exceeding this level can be neglected. The expression of $V_{z_{ml}}(g)$ is given by

$$V_{z_{ml}}(g) = \int_{a_g-H/G}^{a_g+H/G} p_{z_{ml}}(z_{ml}|C_l, \mathbf{f}_l, \Phi_l) dz_{ml}. \quad (5.27)$$

The final-decision variable for the ML-A receiver is $y_m = \sum_{l=1}^L \tilde{z}_{ml}$, and its corresponding pdf expression can be expressed as

$$\begin{aligned} p_{y_m}(y_m|\mathbf{C}, \mathbf{f}, \Phi) &= p_{\tilde{z}_{m1}}(\tilde{z}_{m1}|C_1, \mathbf{f}_1, \Phi_1) \otimes p_{\tilde{z}_{m2}}(\tilde{z}_{m2}|C_2, \mathbf{f}_2, \Phi_2) \\ &\quad \otimes \dots \otimes p_{\tilde{z}_{mL}}(\tilde{z}_{mL}|C_L, \mathbf{f}_L, \Phi_L) \end{aligned} \quad (5.28)$$

where $\mathbf{C} = (C_1, C_2, \dots, C_L)$, $\mathbf{f} = (\mathbf{f}_1, \mathbf{f}_2, \dots, \mathbf{f}_L)$, and $\Phi = (\Phi_1, \Phi_2, \dots, \Phi_L)$. The conditional probability of symbol error is

$$\begin{aligned} P_s(\mathbf{C}, \mathbf{f}, \Phi) &= 1 - \sum_{g_1=-GL/2}^{GL/2} \sum_{g_2=-GL/2}^{g_1} \dots \sum_{g_M=-GL/2}^{g_1} p_{y_1}\left(\frac{2g_1H}{G} \middle| \mathbf{C}, \mathbf{f}, \Phi\right) \\ &\quad \times p_{y_2}\left(\frac{2g_2H}{G} \middle| \mathbf{C}, \mathbf{f}, \Phi\right) \times \dots \times p_{y_M}\left(\frac{2g_MH}{G} \middle| \mathbf{C}, \mathbf{f}, \Phi\right). \end{aligned} \quad (5.29)$$

By taking the expectation over Φ , we obtain

$$P_s(\mathbf{C}, \mathbf{f}) = E_{\Phi} \{P_s(\mathbf{C}, \mathbf{f}, \Phi)\}. \quad (5.30)$$

Finally, the average BER is [1]

$$P_b = \frac{M}{2(M-1)} P_s \quad (5.31)$$

with the probability of symbol error given by

$$P_s = \prod_{l=1}^L \left[\sum_{C_l=0}^{K-1} \binom{K-1}{C_l} \left(\frac{1}{MQ} \right)^{C_l} \left(1 - \frac{1}{Q} \right)^{K-1-C_l} \right] \sum_{\text{all } \mathbf{f}} P_s(\mathbf{C}, \mathbf{f}). \quad (5.32)$$

5.4.2 ML-B Receiver

For the ML-B receiver, the pdf expressions of the decision variables x_{1l} and x_{pl} , for $p = 2, 3, \dots, M$, can be expressed as

$$\begin{aligned} p_{x_{1l}}(x_{1l}|C_l, \mathbf{f}_l, \Phi_l) &= \frac{1}{2\Psi_{1l}\sigma_{1l,S}^2} \exp \left(-\frac{\frac{x_{1l}-\zeta_{1l}}{\Psi_{1l}} + A_{1l,S}^2}{2\sigma_{1l,S}^2} \right) I_0 \left(\frac{A_{1l,S} \sqrt{\frac{x_{1l}-\zeta_{1l}}{\Psi_{1l}}}}{\sigma_{1l,S}^2} \right) \\ &\quad \times U \left(\frac{x_{1l} - \zeta_{1l}}{\Psi_{1l}} \right) \end{aligned} \quad (5.33)$$

and

$$\begin{aligned} p_{x_{pl}}(x_{pl}|C_l, \mathbf{f}_l, \Phi_l) &= \frac{1}{2\Psi_{pl}\sigma_{pl,NS}^2} \exp \left(-\frac{\frac{x_{pl}-\zeta_{pl}}{\Psi_{pl}} + A_{pl,NS}^2}{2\sigma_{pl,NS}^2} \right) I_0 \left(\frac{A_{pl,NS} \sqrt{\frac{x_{pl}-\zeta_{pl}}{\Psi_{pl}}}}{\sigma_{pl,NS}^2} \right) \\ &\quad \times U \left(\frac{x_{pl} - \zeta_{pl}}{\Psi_{pl}} \right) \end{aligned} \quad (5.34)$$

respectively. Following that, we obtain the corresponding characteristic functions as [90]

$$\Phi_{x_{1l}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{\exp(iv\zeta_{1l})}{1 - i2v\Psi_{1l}\sigma_{1l,S}^2} \exp\left(\frac{iv\Psi_{1l}A_{1l,S}^2}{1 - i2v\Psi_{1l}\sigma_{1l,S}^2}\right) \quad (5.35)$$

and

$$\Phi_{x_{pl}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{\exp(iv\zeta_{pl})}{1 - i2v\Psi_{pl}\sigma_{pl,NS}^2} \exp\left(\frac{iv\Psi_{pl}A_{pl,NS}^2}{1 - i2v\Psi_{pl}\sigma_{pl,NS}^2}\right) \quad \text{for } p = 2, 3, \dots, M \quad (5.36)$$

respectively. The characteristic function of the final-decision variable for the ML-B receiver is given by

$$\Phi_{y_m}(iv|\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) = \prod_{l=1}^L \Phi_{x_{ml}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l). \quad (5.37)$$

By performing the inverse transformation of (5.37), we obtain the pdf expression of the final-decision variable, which is denoted as $p_{y_m}(y_m|\mathbf{C}, \mathbf{f}, \mathbf{\Phi})$. The conditional probability of symbol error is

$$\begin{aligned} P_s(\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) &= 1 - \int_{-\infty}^{\infty} \left[\int_{-\infty}^{y_1} p_{y_2}(y_2|\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) dy_2 \times \int_{-\infty}^{y_1} p_{y_3}(y_3|\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) dy_3 \right. \\ &\quad \left. \times \dots \times \int_{-\infty}^{y_1} p_{y_M}(y_M|\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) dy_M \right] \times p_{y_1}(y_1|\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) dy_1. \end{aligned} \quad (5.38)$$

Note that we can use quantization and summation to replace the multi-dimensional integration to evaluate (5.38). After that, we follow the similar steps shown in (5.30)-(5.31) to obtain the average BER expressions of the proposed ML-B receiver.

5.4.3 Linear-Combining Receiver

For the linear-combining receiver, the function $F(\cdot)$ shown in Fig. 2.2 is to produce the same output as the input. Hence,

$$z_{ml} = F(r_{ml}) = r_{ml}. \quad (5.39)$$

The conditional characteristic functions of z_{1l} and z_{pl} , for $p = 2, 3, \dots, M$, can be obtained as

$$\Phi_{z_{1l}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{1}{1 - 2iv\sigma_{1l,S}^2} \exp\left(\frac{ivA_{1l,S}^2}{1 - 2iv\sigma_{1l,S}^2}\right) \quad (5.40)$$

and

$$\Phi_{z_{pl}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{1}{1 - 2iv\sigma_{pl,NS}^2} \exp\left(\frac{ivA_{pl,NS}^2}{1 - 2iv\sigma_{pl,NS}^2}\right) \quad (5.41)$$

respectively. Following the similar steps shown in (5.37), (5.38), (5.30) and (5.31), analytical BER expressions of the linear-combining receiver can be obtained.

5.4.4 Product-Combining Receiver

For the product-combining receiver, the function $F(\cdot)$ shown in Fig. 2.2 is given by

$$z_{ml} = F(r_{ml}) = \ln(r_{ml}). \quad (5.42)$$

The conditional pdf expressions of z_{1l} and z_{pl} , for $p = 2, 3, \dots, M$, can be derived as [87]

$$p_{z_{1l}}(z_{1l}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{e^{z_{1l}}}{2\sigma_{1l,S}^2} \exp\left(-\frac{e^{z_{1l}} + A_{1l,S}^2}{2\sigma_{1l,S}^2}\right) I_0\left(\frac{\sqrt{e^{z_{1l}}} A_{1l,S}}{\sigma_{1l,S}^2}\right) \quad (5.43)$$

and

$$p_{z_{pl}}(z_{pl}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \frac{e^{z_{pl}}}{2\sigma_{pl,NS}^2} \exp\left(-\frac{e^{z_{pl}} + A_{pl,NS}^2}{2\sigma_{pl,NS}^2}\right) I_0\left(\frac{\sqrt{e^{z_{pl}}} A_{pl,NS}}{\sigma_{pl,NS}^2}\right) \quad (5.44)$$

respectively. After some mathematical manipulations, we obtain the corresponding conditional characteristic functions as [90]

$$\Phi_{z_{1l}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = (2\sigma_{1l,S}^2)^{iv} \exp\left(-\frac{A_{1l,S}^2}{2\sigma_{1l,S}^2}\right) \sum_{a=0}^{\infty} \frac{\Gamma(iv + 1 + a)}{(a!)^2} \left(\frac{A_{1l,S}^2}{2\sigma_{1l,S}^2}\right)^a \quad (5.45)$$

and

$$\Phi_{z_{pl}}(iv|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = (2\sigma_{pl,NS}^2)^{iv} \exp\left(-\frac{A_{pl,NS}^2}{2\sigma_{pl,NS}^2}\right) \sum_{a=0}^{\infty} \frac{\Gamma(iv + 1 + a)}{(a!)^2} \left(\frac{A_{pl,NS}^2}{2\sigma_{pl,NS}^2}\right)^a \quad \text{for } p = 2, 3, \dots, M \quad (5.46)$$

respectively. Though the index a in (5.45) and (5.46) runs from zero to infinity, to numerically calculate those two expressions, a good approximation can be achieved by just summing over a small number of terms. Following that, we follow the same steps shown in (5.37), (5.38), (5.30) and (5.31) to obtain the average BER expressions of the product-combining receiver.

5.4.5 Clipper Receiver

For the clipper receiver, the function $F(\cdot)$ shown in Fig. 2.2 performs a clipping operation to produce

$$z_{ml} = F(r_{ml}) = \begin{cases} r_{ml}, & \text{if } r_{ml} \leq T \\ T, & \text{if } r_{ml} > T \end{cases} \quad (5.47)$$

where T is the clipping threshold. For mathematical convenience, we use a G -level uniform quantizer to approximate the output z_{ml} as

$$\tilde{z}_{ml} = \begin{cases} a_g, & \text{if } a_g \leq z_{ml} < a_{g+1} \\ a_{G-1}, & \text{if } z_{ml} \geq a_{G-1} \end{cases} \quad (5.48)$$

where $a_g = gT/(G-1)$, and $g = 0, 1, 2, \dots, G-1$. The discrete pdf expression of $\tilde{z}_{ml}$ can be expressed as

$$p_{z_{ml}}(\tilde{z}_{ml}|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) = \sum_{g=0}^{G-1} D_{z_{ml}} \delta(\tilde{z}_{ml} - a_g) \quad (5.49)$$

where $D_{z_{ml}}$ is given by

$$D_{z_{ml}} = \begin{cases} \int_{a_g}^{a_{g+1}} p_{z_{ml}}(z|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) dz, & \text{if } g = 0, 1, 2, \dots, G-2 \\ \int_C^\infty p_{z_{ml}}(z|C_l, \mathbf{f}_l, \mathbf{\Phi}_l) dz, & \text{if } g = G-1. \end{cases} \quad (5.50)$$

To obtain the pdf expression of the final-decision variable $y_m = \sum_{l=1}^L \tilde{z}_{ml}$, we can perform the discrete convolution as

$$\begin{aligned} p_{y_m}(y_m|\mathbf{C}, \mathbf{f}, \mathbf{\Phi}) &= p_{\tilde{z}_{m1}}(\tilde{z}_{m1}|C_1, \mathbf{f}_1, \mathbf{\Phi}_1) \otimes p_{\tilde{z}_{m2}}(\tilde{z}_{m2}|C_2, \mathbf{f}_2, \mathbf{\Phi}_2) \otimes \dots \\ &\quad \otimes p_{\tilde{z}_{mL}}(\tilde{z}_{mL}|C_L, \mathbf{f}_L, \mathbf{\Phi}_L). \end{aligned} \quad (5.51)$$

The conditional probability of symbol error is given by

$$P_s(\mathbf{C}, \mathbf{f}, \Phi) = 1 - \sum_{g_1=0}^{L(G-1)} \sum_{g_2=0}^{g_1} \cdots \sum_{g_M=0}^{g_1} p_{y_1} \left(g_1 a_1 \middle| \mathbf{C}, \mathbf{f}, \Phi \right) \times p_{y_2} \left(g_2 a_1 \middle| \mathbf{C}, \mathbf{f}, \Phi \right) \\ \times \cdots \times p_{y_M} \left(g_M a_1 \middle| \mathbf{C}, \mathbf{f}, \Phi \right). \quad (5.52)$$

Following that, we can follow (5.30) and (5.31) to obtain the average BER expressions of the clipper receiver.

5.5 Numerical Results and Discussions

In this section, numerical results for the proposed ML, ML-A, ML-B, linear-combining, product-combining, and clipper receivers will be presented. For fair comparisons, the total SS bandwidth W_{ss} is fixed and the total number of FH bands is set to be 1000 for the binary case without any diversity. Figure 5.1 shows the analytical and simulation BER results for the proposed ML-A, ML-B, linear-combining, product-combining and clipper receivers over Rician-fading channels. Analytical results are obtained from the analytical BER expressions derived in Section 5.4. For the clipper receiver, the BER results are obtained when the optimum threshold level is used. This optimum threshold level is obtained by numerically searching for the value of T corresponding to the best BER result for each E_b/N_0 level. The close match between the analytical and simulation results validates the analytical BER expressions derived in Section 5.4. It is also observed that the BER performance of all the five receivers are very close to each other at low E_b/N_0 levels (i.e., $E_b/N_0 < 6$ dB). When E_b/N_0 level is high (i.e., $E_b/N_0 > 6$ dB), the proposed ML-A receiver has the best performance, and the clipper receiver outperforms the other three receivers. Furthermore, the product-combining receiver, which is slightly outperformed by the

Figure 5.1: Analytical and simulation results of various receivers with $L = 3$, $F = 5$, $M = 4$ and $K = 10$.

ML-B receiver, is superior to the linear-combining receiver at $E_b/N_0 > 11$ dB. Note that side information is required for the proposed ML-A, ML-B and clipper receivers. Therefore, there is a tradeoff between performance and system complexity. If side information is not available, the product-combining receiver, which is nonparametric, is a good choice.

Performance comparisons between the ML-A and ML-B receivers under various specular-to-diffused power ratios F are shown in Fig. 5.2. It is observed that, in general, higher specular-to-diffused power ratio leads to better BER performance. Furthermore, it is also found that at high E_b/N_0 levels (i.e., $E_b/N_0 > 3$ dB) and high F values (i.e., $F = 5$ and $F = 10$), the ML-A receiver is superior to the ML-B receiver, and the BER gap between the ML-A and ML-B receivers increases as the E_b/N_0 level increases. This is due to the fact that the ML-A receiver is obtained by using the approximation of $\ln[I_0(x)] \approx x$, for $x > 4$; while the ML-B

Figure 5.2: Performance comparisons between the ML-A and ML-B receivers with $L = 3$, $M = 4$, $K = 10$, and various F .

receiver is derived with the approximation of $\ln[I_0(x)] \approx x^2/4$, for $0 < x \leq 4$. At a high E_b/N_0 level, the approximation that used to derived the ML-A receiver is more accurate than that used to derive the ML-B receiver. On the other hand, under the conditions of low E_b/N_0 levels and small F values, the ML-B receiver slightly outperforms the ML-A receiver. This is because under this condition, the approximation, $\ln[I_0(x)] \approx x^2/4$, is more accurate. Moreover, under frequency non-selective Rayleigh-fading channels (i.e., $F = 0$), the BER curves of the ML-A and ML-B receivers coincide with each other. This is because under Rayleigh-fading channels, there is no LOS component. Thus, the approximation shown in (5.16) is no longer needed, and the random variable for the optimum ML receiver in (5.15) is simplified into $x_{ml} = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right)$, which can also be obtained by setting $F = 0$ into (5.17) and (5.21). Therefore, the ML-A and ML-B receivers turn out to be the same as the optimum ML receiver for the Rayleigh-fading case.

Figure 5.3: BER performance of the proposed ML-A receiver over Rician-fading channels ($F = 5$) with $M = 4$, $K = 10$, and different diversity orders.

Figure 5.4: BER performance of the proposed ML-B receiver over Rician-fading channels ($F = 5$) with $M = 4$, $K = 10$, and different diversity orders.

Figure 5.5: BER performance of the product-combining receiver over Rician-fading channels ($F = 5$) with $M = 4$, $K = 10$, and different diversity orders.

BER performance of the proposed ML-A, ML-B, linear-combining, product-combining and clipper receivers over Rician-fading channels with various diversity orders L are shown in Figs. 5.3-5.7, respectively. It is observed that at low E_b/N_0 levels (i.e., $E_b/N_0 < 6.5$ dB for the ML-A receiver and $E_b/N_0 < 9$ dB for the ML-B receiver), there is no diversity gains, i.e., the case with $L = 1$ gives the best BER performance. On the other hand, at high E_b/N_0 levels, we can observe some diversity gains. In addition, there is also an optimum diversity order (i.e., $L = 3$ at $E_b/N_0 = 8$ and 10 dB for the ML-A receiver and $L = 3$ at $E_b/N_0 = 10$ dB for the ML-B receiver) at each E_b/N_0 level. For FFH/MA systems, higher diversity orders can counteract the fading and MAI effects more effectively while introducing more non-coherent diversity-combining loss. Hence, there exists an optimum diversity order, which depends on how effective is the diversity-combining scheme in counteracting the fading and MAI effects.

Figure 5.6: BER performance of the linear-combining receiver over Rician-fading channels ($F = 5$) with $M = 4$, $K = 10$, and different diversity orders.

BER performance of these five receivers under various modulation levels M over Rician-fading channels are presented in Figs. 5.8-5.12, respectively. The BER results are plotted against the number of active users. It is found that a higher modulation level M results in better BER performance for all the five receivers, and this observation is consistent with that of conventional MFSK systems under AWGN channels [10].

Figure 5.13 shows the performance comparisons among the six receivers over Rician-fading channels. The BER results for the ML receiver are obtained via simulation, whereas the BER curves of the other five receivers are obtained based on the analytical BER expressions. From this figure, it is observed that the BER results increase as the number of users increases due to the effect of more severe MAI. It is also observed that the proposed ML receiver gives the best BER performance, and the proposed ML-A receiver significantly outperforms the other four receivers. The

Figure 5.7: BER performance of the clipper receiver over Rician-fading channels ($F = 5$) with $M = 4$, $K = 10$, and different diversity orders.

proposed ML-B receiver, which is outperformed by the clipper receiver, is superior to the linear-combining and product-combining receivers. Performance comparisons between the linear-combining and product-combining receivers show that the product-combining receiver gives reasonably good BER performance.

Performance comparisons among various receiver over Rayleigh-fading channels are shown in Fig. 5.14. From this figure, we can obtain similar observations as those in Fig. 5.13, except that the absolute values of the BER results are higher as the channel fading is more severe.

Figure 5.8: BER performance of the proposed ML-A receiver over Rician-fading channels ($F = 10$) with $E_b/N_0 = 12$ dB, $L = 3$, and various M .

Figure 5.9: BER performance of the proposed ML-B receiver over Rician-fading channels ($F = 10$) with $E_b/N_0 = 12$ dB, $L = 3$, and various M .

Figure 5.10: BER performance of the linear-combining receiver over Rician-fading channels ($F = 10$) with $E_b/N_0 = 12$ dB, $L = 3$, and various M .

Figure 5.11: BER performance of the product-combining receiver over Rician-fading channels ($F = 10$) with $E_b/N_0 = 12$ dB, $L = 3$, and various M .

Figure 5.12: BER performance of the clipper receiver over Rician-fading channels ($F = 10$) with $E_b/N_0 = 12$ dB, $L = 3$, and various M .

Figure 5.13: Performance comparisons among various receivers over Rician-fading channels ($F = 10$) with $M = 4$, $E_b/N_0 = 12$ dB and $L = 3$.

Figure 5.14: Performance comparisons among various receivers over Rayleigh-fading channels ($F = 0$) with $M = 4$, $E_b/N_0 = 12$ dB and $L = 3$.

5.6 Conclusions

In this chapter, an optimum ML receiver has been derived for synchronous FFH/MA systems with MFSK over frequency non-selective Rician-fading channels. In addition, we have proposed two suboptimum ML receivers and derived their analytical BER expressions. Moreover, the linear-combining, product-combining and clipper receivers are also investigated. Those analytical expressions have been validated by simulation results. Numerical results have shown that the optimum ML receiver gives the best BER performance and the proposed ML-A receiver outperforms the other four receivers. The proposed ML-B receiver, which is outperformed by the clipper receiver, is superior to the linear-combining and product-combining receivers. Since the proposed ML, ML-A, ML-B and clipper receivers require side information of the system, there is a tradeoff between performance and system complexity. If side information is not available, the product-combining receiver, which is nonparamet-

ric, is a good choice. In addition, the effects of some parameters, such as modulation level and diversity order, have also been studied.

Chapter 6

Conclusions and Recommendations

6.1 Conclusions

In this thesis, we have studied FFH/MFSK communication systems over fading channels in the presence of single-tone per-band MTJ, PBNJ, MAI and AWGN. Both frequency non-selective and frequency-selective fading channels have been considered. FFH/MFSK systems employing various diversity-combining techniques, including the linear-combining, product-combining, clipper and ML receivers, have been investigated under different channel and interference conditions.

To suppress the effect of PBNJ, we have proposed an optimum ML receiver for FFH/MFSK systems over a frequency-selective Rayleigh-fading channel and derived its corresponding analytical BER expressions. These BER expressions have been validated by simulation results. Moreover, the effects of power ratio and cross-correlation coefficient on system performance have also been studied. Numerical

results have shown that lower power ratio and cross-correlation coefficient lead to better BER performance. It has also been observed that higher diversity orders can better counteract PBNJ and fading effects more efficiently while introducing more non-coherent diversity-combining loss. Therefore, there exists an optimum diversity order, at which the BER performance is the best. Following that, we have extended the analysis to a more general frequency-selective Rician-fading channel. An optimum ML receiver has been proposed. In addition, two suboptimum ML receivers and their corresponding theoretical BER expressions have been presented. Furthermore, we have also investigated the linear-combining and product-combining receivers under the same channel conditions.

To mitigate the composite effect of MTJ and PBNJ, an optimum ML receiver for frequency non-selective Rayleigh-fading channels has been proposed. Its analytical BER expressions have also been derived and validated by simulation results. Performance comparisons have shown that the proposed ML receiver can better counteract the effect of MTJ than that of PBNJ. Under the condition when both MTJ and PBNJ are present concurrently, the BER results are bounded by the two extreme cases, where either MTJ or PBNJ is present. It is also observed that the optimum diversity order of the proposed ML receiver increases as the modulation level M and E_b/N_0 level increase. Following that, we have considered a frequency-selective Rayleigh-fading channel with MTJ and PBNJ. An optimum ML receiver structure and its corresponding analytical BER expressions have been presented. In addition, we have investigated the linear-combining and product-combining receivers under the same channel conditions. Numerical results have shown that the ML receiver has the best BER performance. The product-combining receiver outperforms the linear-combining receiver when the jamming power is relatively strong. Some parameters, such as power ratio, cross-correlation coefficient and diversity order, on system per-

formance have also been studied. Furthermore, we have investigated the effects of timing and frequency offsets on system performance under frequency non-selective Rician-fading channels with MTJ and PBNJ. New analytical BER expressions of the linear-combining and product-combining receivers have been derived. It has been observed that for the linear-combining receiver, MTJ is more harmful to the system than that of PBNJ, while the reverse is true for the product-combining receiver. Moreover, it has been found that the BER performance is severely degraded as the timing and frequency offsets increase. The product-combining receiver is more sensitive to the timing and frequency offsets than that of the linear-combining receiver. When the normalized offset value is less than 0.3, both receivers are more sensitive to the timing offset as compared to the frequency offset. However, the system is more sensitive to frequency offsets when the normalized offset value is greater than 0.4. It has also been observed that for the linear-combining receiver, the optimum diversity order increases as the timing and frequency offsets increase, whereas the reverse is true for the product-combining receiver.

A new pdf expression of the square-law detector output has been derived for synchronous FFH/MA systems with MFSK modulation over frequency non-selective Rician-fading channels. Based on this derived pdf expression, an optimum ML receiver has been proposed. In addition, two suboptimum ML receivers, namely, the ML-A and ML-B receivers, have been proposed for different channel conditions. Analytical BER expressions of these two suboptimum ML receivers, as well as the linear-combining, product-combining and clipper receivers, have been derived, and validated by simulation results. The derived analytical BER expressions can help to understand the behaviour of FFH/MA system performance in very low BER range, where simulations are too time consuming to be performed. Numerical results have shown that the proposed ML-A receiver is superior to the ML-B receiver and

the other three receivers, though it is slightly outperformed by the optimum ML receiver. The ML-B receiver, which is outperformed by the clipper receiver, performs better than the product-combining receiver, which in turn is superior to the linear-combining receiver at relatively high E_b/N_0 levels. Note that side information is required for the optimum ML, suboptimum ML and clipper receivers. Hence, there is a tradeoff between the performance and system complexity. If side information is not available, the product-combining receiver, which is nonparametric, is a good choice. The effects of some other parameters, i.e., diversity order and modulation level, on the BER performance have also been studied.

6.2 Recommendations for Future Work

In this thesis, we have studied the performance of FFH/MFSK systems with timing and frequency offsets. Since it has been shown that both timing and frequency offsets can severely degrade the system performance, it may be interesting to propose some solutions to countermeasure the effects of timing and frequency offsets. The approaches based on pilot signals [103]-[105] or parameter estimations [106]-[108] may be suitable to recover the system performance from those two effects. Moreover, we may also extend the Rayleigh-fading cases considered in Chapter 4 to Rician-fading ones in future works.

In this thesis, we have proposed the optimum ML receiver for synchronous FFH/MA systems over frequency non-selective Rician-fading channels. In wireless communications, asynchronous systems are also commonly encountered. It is important to derive and study the ML receiver for asynchronous FFH/MA systems over frequency non-selective Rician-fading channels.

Moreover, in this thesis, we only focus on frequency non-selective fading channels for synchronous FFH/MA systems. However, in practice, frequency-selective fading is commonly encountered in both urban and indoor environments and it is a major source of potential degradation in wideband mobile communications [17]. Therefore, it is also interesting to study FFH/MA system performance under frequency-selective fading channel. Moreover, the RAKE receivers are often used to solve the frequency-selective channel problem in the direct-sequence code-division multiple-access systems [1], and it can improve the system performance substantially. Hence, it is possible to employ the RAKE receiver for FFH/MA systems with the expectation that the BER performance with the RAKE receiver is better than that without the RAKE receiver. Lastly, FFH/MA systems employing various types of error-correction coding techniques are also of interests for future research work.

Appendix

Based on Taylor series expansion in central differences [88], we can obtain

$$p_{z_{ml}}(z_{ml}|q_l) \approx \frac{h^2 - \sigma^2}{h^2} p_{z_{ml}}(z_{ml}|\mu, q_l) + \sum_{m=0}^1 \frac{\sigma^2}{2h^2} p_{z_{ml}}(z_{ml}|\mu + (2m-1)h, q_l) \quad (\text{A.1})$$

where μ and σ^2 are the mean and variance of cosin function within the expression of $p_{z_{ml}}(z_{ml}|\cos(\phi_{ss_{1l}}), q_l)$. The parameter h is given by [88]

$$h = \frac{1}{\sigma} \sqrt{4\mu_3 \frac{p_{z_{ml}}^{[III]}(z_{ml}|\cos(\phi_{ss_{1l}}), q_l)}{p_{z_{ml}}^{[IV]}(z_{ml}|\cos(\phi_{ss_{1l}}), q_l)} + \mu_4} \quad (\text{A.2})$$

where μ_3 and μ_4 are the third- and fourth-order central moments of cosine function. The superscripts [III] and [IV] denote the third- and fourth-order derivatives, respectively. Note that $\mu = 0$, $\sigma^2 = 1/2$, $\mu_3 = 0$, and $\mu_4 = 3/8$ in this case. Therefore, (A.1) can be simplified to

$$p_{z_{ml}}(z_{ml}|q_l) \approx \frac{1}{3} \sum_{b=-1}^1 p_{z_{ml}}\left(z_{ml}|\cos(\phi_{ss_{1l}}) = \frac{\sqrt{3}}{2}b, q_l\right). \quad (\text{A.3})$$

Author's Publications

Journal Publications

1. J. Zhang, K. C. Teh, and K. H. Li, "Error probability analysis of FFH/MFSK receivers over frequency-selective Rician-fading channels with partial band noise jamming," *IEEE Trans. Commun.*, vol. 57, pp. 2880-2885, Oct. 2009.
2. J. Zhang, K. C. Teh, and K. H. Li, "Performance analysis of a maximum-likelihood FFH/MFSK receiver with partial-band-noise jamming over frequency-selective fading channels," *IEEE Commun. Lett.*, vol. 12, pp. 401-403, June 2008.
3. J. Zhang, K. C. Teh, and K. H. Li, "Performance study of fast frequency-hopped/M-ary frequency-shift keying systems with timing and frequency offsets over Rician-fading channels with both MTJ and PBNJ," *IET Commun.*, vol. 4, pp. 1153-1163, Apr. 2010.
4. J. Zhang, K. C. Teh, and K. H. Li, "Maximum-likelihood FFH/MFSK receiver over Rayleigh-fading channels with composite effects of MTJ and PBNJ," second revision submitted to *IEEE Trans. Commun.*

Conference Publications

1. J. Zhang, K. C. Teh, and K. H. Li, "Rejection of multitone jamming for FFH/MFSK spread-spectrum systems over frequency-selective Rayleigh-fading channels," in *Proc. IEEE Vehicular Technology Conference*, May 2008, pp. 688-692.
2. J. Zhang, K. C. Teh, and K. H. Li, "Suppression of partial-band-noise jamming for FFH/MFSK systems over frequency-selective Rician-fading channels," in

Proc. IEEE International Symposium on Information Theory and its Applications, Dec. 2008, pp. 1-4.

3. J. Zhang, K. C. Teh, and K. H. Li, "An optimum diversity-combining technique for FFH/MFSK communication systems over frequency-selective fading channels with multitone jamming," in *Proc. IEEE International Conference on Information, Communications and Signal Processing*, Dec. 2009, pp. 1-5.
4. J. Zhang, K. C. Teh, and K. H. Li, "Performance study in maximum-likelihood receiver for synchronous FFH/MA systems over Rician-fading channels," submitted to *Proc. IEEE International Conference on Signal Processing and Communication Systems*.

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