Performance analysis of optimum receivers for fast frequency-hopped systems over composite interference and fading channels with imperfect side information

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**NANYANG
TECHNOLOGICAL
UNIVERSITY**

**PERFORMANCE ANALYSIS OF OPTIMUM
RECEIVERS FOR FAST FREQUENCY-HOPPED
SYSTEMS OVER COMPOSITE INTERFERENCE
AND FADING CHANNELS WITH IMPERFECT
SIDE INFORMATION**

LE LY MINH DUY
School of Electrical and Electronic Engineering

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A thesis submitted to the Nanyang Technological University
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List of Abbreviations

AWGN	Additive white Gaussian noise
BER	Bit-error rate
BFSK	Binary frequency-shift keying
BPF	Bandpass filter
CDF	Cumulative distribution function
CDMA	Code-division multiple-access
CF	Characteristic function
CR	Clipper receiver
DS	Direct-sequence
EGC	Equal-gain combining
FH	Frequency hopping
FFH	Fast frequency hopping
FFT	Fast Fourier transform

HDMV	Hard-decision-majority-vote
LOS	Line-of-sight
LCR	Linear-combining receiver
MA	Multiple-access
MAI	Multiple-access interference
MFSK	M -ary frequency-shift keying
ML	Maximum-likelihood
MTI	Multitone interference
NCBFSK	Non-coherent binary frequency-shift keying
NED	Normalized envelope detection
OFDM	Orthogonal frequency-division multiplexing
OS-NED	Order statistics-NED
PAN	Personal area network
PBNI	Partial-band noise interference
PCR	Product-combining receiver
PDF	Probability density function
PN	Pseudo-noise
PSD	Power spectral density
RF	Radio frequency

RSS	Reduced rank sum
SER	Symbol-error rate
SIR	Signal-to-interference ratio
SNR	Signal-to-noise ratio
SS	Spread spectrum
WLAN	Wireless local area network

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List of Symbols

σ_ω^2	variance of Gaussian noise
σ_P^2	variance of PBNI
ν_P	power ratio of PBNI with respect to the total interference power
ν_μ	power ratio of MTI with respect to the total interference power
r_{ml}	square-law detector output
y_m	final-decision variable of the m th branch
B	bandwidth of each FH slot
E_b	bit energy
E_s	symbol energy
L	diversity order
M	modulation level
N	total number of non-overlapping FH bands
N_0	one-sided power spectral density of AWGN
N_J	equivalent power spectral density of both PBNI and MTI
P_b	average bit-error rate
P_s	average symbol error probability
P_J	total interference power
P_P	total interference power of PBNI
P_μ	total interference power of MTI

$P_{\mu T}$	power of each single interfering tone
Q	number of MTI tones
T_b	bit duration
T_h	hopping duration
T_s	symbol duration
W_{ss}	total spread-spectrum bandwidth

Summary

A fast frequency-hopped (FFH) system has an efficient anti-interference capability and is robust against fading. The performance of FFH systems is mainly affected by the intentional or unintentional interference, additive white Gaussian noise, timing and frequency offsets, as well as fading channels. The interference can be mitigated or avoided by hopping the carrier frequency of the transmitted signal over a wide bandwidth in a pseudorandom pattern. Fading channels can cause a significant degradation in the performance of a communication system. The concept of diversity methods for fading channels was therefore introduced in the literature. Diversity methods have been shown to help overcome the performance degradation caused by fading channels.

In practice, the two commonly encountered interference models against the FFH systems are partial-band noise interference (PBNI) and multitone interference (MTI). Hence, it is essential to investigate the composite effect of PBNI and MTI on FFH systems over fading channels. The analytical bit-error rate expressions are derived for FFH M -ary frequency-shift-keying (MFSK) maximum-likelihood (ML) receivers under the composite effect of MTI and PBNI over either frequency-nonselective Rician fading or frequency-selective Rayleigh fading channels. The ML structures are also proposed and analyzed. The numerical results show that the proposed systems can effectively suppress the composite interference effect over fading channels.

In practical systems, it is well known that there always exists time and frequency mismatch between the transmitter and the receiver. Based on the log-likelihood ratio criterion, a ML receiver for FFH/MFSK systems is proposed to combat the composite effect of PBNI and MTI with the presence of timing and frequency offsets over frequency-nonselective Rayleigh fading channels. Analytical bit-error rate (BER) expression for the proposed receiver is derived and validated by simulations. Comparing with other existing diversity-combining receivers such as the linear-combining and

product-combining receivers, the proposed receiver achieves the best performance at the expense of system complexity. In addition, the sensitivity of system performance to imperfect side information is also examined.

The effect of multiple-access interference (MAI) on the performance of FFH multiple-access (MA) systems is investigated. The analysis of FFH-MA systems over frequency-selective Rayleigh fading channels is first performed, followed by the frequency-selective Rician fading. For the case of frequency-selective Rayleigh fading, the exact closed-form expression of the characteristic function of the ML decision variable will be derived. For the case of frequency-selective Rician fading, we derive the probability density function (pdf) of the ML decision variable. Following that, the discrete convolution approach will be applied to obtain the pdf of the final decision statistics. Our analysis shows that the proposed ML receivers can suppress the MAI more effectively than the conventional receivers, i.e., up to two orders of magnitude in the BER performance improvement.

Chapter 1

Introduction

1.1 Background

Spread spectrum (SS) has been proposed as an effective technique to combat the deleterious effects of fading and interference. SS techniques can be used to suppress interference, provide low probability of interception, accommodate fading and multipath channels, and to provide multiple-access (MA) capabilities [1]. In a SS system, the spreading waveform is controlled by a pseudo-noise (PN) sequence, which is a binary sequence that is apparently random but that can be reproduced deterministically by intended users. There are primarily two types of SS techniques, i.e., direct sequence (DS) and frequency hopping (FH). Direct-sequence SS modulates the original signal with a very wideband signal relative to the data bandwidth to achieve a spreading of the spectrum [2]. FH spread spectrum system divides the total available bandwidth into N channels and hops between these channels according to a PN code known to both the modulator and demodulator. The noncoherent M -ary frequency-shift-keying (MFSK) modulation scheme is often used for FH systems because of the difficulty of maintaining phase coherence between consecutive hops in FH systems, the simplicity of the MFSK receiver, and its insensitivity to rapid phase variation in the received signal [3], [4].

FH systems have certain advantages over DS systems such as low probability of detection and interception, reduced nearfar problem, less stringent power control, better resistance to interference, not requiring a contiguous and much lower portable power consumption [5]–[7]. FH systems may be categorized as slow hopping, where there are multiple data symbols transmitted per hop, or fast hopping, where there are multiple hops per data symbol. In fast FH (FFH) systems, the hopping speed is greater than the source information rate, and for each bit of information the modulator sends many signals at different frequencies. If the diversity order of FFH systems is increased, the signal is more redundant and resistance to perturbations is improved. In slow FH (SFH) systems, the hopping speed is lower than the source information bit rate and the modulator behaves like a narrow-band modulator at each hop [8]. The main difference is that with SFH we lose blocks of information while with FFH the errors are uncorrelated from one bit to another. However, the main advantage of SFH is that it uses simpler modems than FFH [8].

Fading effects and interference can cause severe degradation in performance of a communication system. When the channel is in deep fade, the signal power can be significantly attenuated and this causes errors in detection. Diversity-combining techniques are employed to provide the receiver with multiple replicas of the transmitted signal that have experienced independent fading channels and thus can reduce the depth of the fade and/or the fade duration [9]. The sources of diversity are frequency diversity, time diversity, spatial diversity and interference diversity. Frequency diversity is the simultaneous use of multiple frequencies to transmit information. Time diversity provides transmitting replicas with enough separation in time to allow signal decorrelation, whereas space diversity uses two or more antennas that are sufficiently spaced to receive the same information [10].

Interference may be classified as intentional or unintentional interference. The main

difference between intentional and unintentional interference is that the former's goal is to jam the transmitted signal, whereas the latter is a consequence of collisions with signals coming from different users in the same hopping frequency [11]. Intentional interference is used to prevent reliable communications between two or more parties [11] and is also called jamming. The commonly encountered interfering waveforms are wide-band noise interference, partial-band noise interference (PBNI), single-tone interference, multitone interference (MTI), pulse-noise interference and repeat-back interference [9]. Among them, PBNI and MTI have been shown to be very harmful to FFH systems [12]. For FH multiple-access (FH-MA) systems, the desired signal is affected by unintentional interference, which is called multiple-access interference (MAI). This interference is caused by simultaneous transmission of the reference user and non-reference users in the same frequency band.

1.2 Motivation

Over the years, FH-SS has found a number of applications in both military and commercial communications. FH-SS is originally used in military communications and is employed for the single channel ground and airborne radio system (SINCGARS) and Have Quick system [13]. SINCGARS uses 25 kHz channels in the very high frequency (VHF) band, from 30 to 87.975 MHz and has frequency hopping modes. The frequency-hopping mode hops 111 times a second [14]. For commercial applications, this technique is mainly adopted in the license-free industrial, scientific and medical (ISM) bands, wireless local area network (WLAN) based on IEEE 802.11 standard, Bluetooth personal area network (PAN) and satellite communications [4], [15]–[16]. Bluetooth operates in the range of 2400–2483.5 MHz (including guard bands) and usually performs 1600 hops per second. The transmitted data are divided into packets and each packet is transmitted on one of the 79 designated Bluetooth channels with a bandwidth of 1 MHz [17].

Besides that, FH-SS technology has been selected for installations designed to cover big areas, where a big number of collocated systems is required and where the use of directional antennas in order to minimize environmental influence is impossible. Typical applications for FH-SS include cellular deployments for fixed broadband wireless access, where the use of direct-sequence SS is virtually impossible because of its limitations [18]. In recent years, a combined FH-SS system with noncoherent MFSK is considered as an attractive approach to construct wireless ad hoc mobile networks [4], [19]. FH has also been adopted in multiband orthogonal frequency-division multiplexing ultra wideband (MB-OFDM UWB) systems [20]. Furthermore, FH is employed in the FH/MFSK wireless sensor network for assisting the fusion center to distinguish the source events and for reducing the correlation among the signals transmitted by the local sensors, so that the fusion detection can benefit from both the space diversity and the frequency diversity [21]. In [22], the authors proposed a novel smart grid communication system based on the FH multiple-access technology. FH is employed in the smart meter network to avoid possible attacks and challenges like jamming, multiple access, etc.

Frequency-hopping technique is commonly used for secure communications such as military applications due to its anti-jamming capability, low probability of detection, resistance to the near-far problem and robustness against fading [6], [23]–[26]. FH will cause different signals to interfere with the carrier at different times. This property is called interference diversity [27]. Interference diversity refers to the statistical averaging of cochannel interference from other cells. It implies that the same interferer will not collide with a time slot on every hop and the effect is to replace worst-case interference with average interference [28], [29]. FFH systems employ both time and frequency diversities and therefore can better combat fading and interference effects [30]. In practice, the performance of FFH systems is mainly affected by the intentional or unintentional interference, additive white Gaussian noise (AWGN), timing

and frequency offsets, as well as fading channels. Diversity methods have been shown to help overcome the performance degradation of fading channels [9]. Many types of diversity-combining receivers have been investigated for FFH systems, such as linear-combining receiver (LCR) [31]–[35], clipper receiver (CR) [36]–[40], noise-normalizing receiver [37], [38], [41]–[43], self-normalizing receiver [44]–[46], product-combining receiver (PCR) [16], [47], [48], ratio-statistic combining receiver [49], and maximum-likelihood (ML) receiver [50]–[54]. ML receivers have been shown to exhibit very good bit-error rate (BER) performance for FFH systems over fading channels. This gives us the motivation to propose and investigate the ML structures for FFH/MFSK systems under different environment conditions over various fading channels.

1.3 Objectives

The objective of this thesis is to study the BER performance of various diversity-combining techniques for FFH/MFSK systems with different types of interference over various types of fading environments. The ML diversity-combining method is the main focus in this work since it has been shown to provide very good BER performance as compared with other existing conventional receivers. Note that the ML receiver structures are different under different conditions. They depend on system parameters, channel fading, forms of interference considered, frequency offset, timing offset, etc. Mathematical expressions for the BER performance of the proposed systems are derived and numerical results based on both simulations and theoretical expressions are presented in order to study the BER performance of FFH systems. In practical FH communication systems, perfect knowledge of the time epoch and the received carrier frequency is not available and there always exist timing and frequency offsets. Therefore in this work, the practical issues of timing and frequency offsets are also considered. In addition, the robustness of the proposed ML receivers with respect to imperfect

estimation of side information has been investigated.

1.4 Major Contributions of the Thesis

The major contributions of this thesis can be summarized as follows. The ML receiver structure is different for different scenerio (system parameters, fading channel, form of interference considered in the system, and the presence of timing and frequency offsets). For the system under composite effect of MTI and PBNI, based on the log-likelihood ratio criterion, we derive the theoretical expressions for the ML receiver over either frequency-nonselective Rician or frequency-selective Rayleigh fading channels. For the case of frequency-nonselective Rician fading, one of the existing approximations is required to derive the BER expressions as the computation of the logarithm of the modified Bessel function is non-linear in nature. A suboptimum ML receiver structure is therefore proposed and analyzed. Following that, an optimum ML receiver is derived to combat the interference composite effects in a frequency-selective Rayleigh fading environment. Under various interference and fading conditions, the proposed ML receivers are found to be able to suppress the composite interference effects effectively. In addition, we study the composite effects of MTI and PBNI on an FFH/MFSK system for a more practical scenario where timing and frequency offsets are present. In practical systems, there is always time and frequency mismatch between the transmitter and the receiver, which can cause severe degradation to system performance. We first consider a frequency-nonselective Rayleigh fading channel and then extend our study to a frequency-selective Rician fading channel. Numerical results demonstrate that the proposed receivers are able to suppress the MTI more effectively than the PBNI. Furthermore, analytical BER expressions of various diversity-combining receivers, including the ML receiver, linear combining receiver and product combining receiver, are derived for a synchronous FFH-MA system employing MFSK modulation over frequency-selective

Rayleigh fading channels. The study is then extended to a frequency-selective Rician fading channel. The analytical BER expressions are validated by simulations. Performance comparisons among several commonly-used receivers show that the proposed ML receiver gives the best BER performance at the expense of system complexity for different combinations of parameters over various fading conditions. Moreover, the sensitivity of system performance to imperfect side information is also examined and the proposed ML receiver is shown to be robust against the inaccuracy in estimation of the required side information.

1.5 Organization

This thesis is divided into six chapters. Chapter 1 contains a brief introduction to FFH systems, diversity techniques for multipath fading channels and commonly encountered interference models against FFH systems. We also introduce the motivation behind this research and the objectives to be achieved.

In Chapter 2, some theoretical backgrounds are provided. We describe the receiver models, including the maximum-likelihood receiver, linear-combining receiver, product-combining receiver, self-normalizing receiver and order statistics-normalized envelope detection receiver, as well as the interference models, which will be used in the analyses. The literature survey on various diversity-combining techniques for suppressing different types of interference in FFH/MFSK systems over various fading channels are also presented.

In Chapter 3, we investigate the composite effect of PBNI and MTI on the FFH systems employing MFSK modulation over either frequency-nonselective Rician or frequency-selective Rayleigh fading channels. Mathematical expressions for the BER performance of the systems are derived based on the log-likelihood ratio criterion. Numerical results based on both simulations and the theoretical expressions are presented

to study the system performance. The system performance is examined under various interference and fading conditions.

In Chapter 4, the impacts of timing and frequency offsets, as well as the composite effects of MTI and PBNI on BER performance of FFH ML receivers are studied. The sensitivity of system performance to imperfect side information is examined. In addition, simulation results of the LCR, PCR, OS-NED and self-normalizing receivers are also provided to compare with the performance of the ML receivers.

In Chapter 5, we first derive analytical BER expressions of various diversity-combining receivers for a synchronous FFH-MA system over frequency-selective Rayleigh fading channels. Following that, the BER expressions for the ML receiver over frequency-selective Rician fading are also derived. In addition, the BER results with imperfect side information are provided to examine their effects on the system performance.

Finally, in Chapter 6, we conclude the achievements of the research, as well as provide some possible directions for future work to be done.

Chapter 2

Theoretical Background and Literature Survey

2.1 System Model

In this section, we introduce the transmitter and receiver models, as well as the diversity-combining receivers, which will be used in the subsequent analyses.

The transmitter block diagram of an FFH/MFSK system is depicted in Fig. 2.1. The binary data streams are first converted into M -ary data streams by an M -ary encoder. Following that, the conventional MFSK modulator selects one of the M baseband frequencies f_m based on the input symbol, where $f_m = m/T_h$ ($m = 1, 2, \dots, M$), $T_h = T_s/L$ denotes the hopping duration and T_s is the symbol duration. Here M represents the modulation order and L is the diversity order. In FH communication systems, the total available channel bandwidth W_{ss} is subdivided into N frequency-hopped bands, as depicted in Fig. 2.2. In FFH systems, the carrier hops multiple times during the transmission of a data symbol [55]. For FFH systems employing MFSK modulation, each FH band consists of M FH slots. The bandwidth of each FH slot is $B = 1/T_h$ Hz and thus the total spread-spectrum bandwidth is given by $W_{ss} = N \times M \times B$.

For a fair comparison, when generating the subsequent numerical results, we fix the total SS bandwidth and set the total number of FH bands to be $N = 1000$ for the binary frequency-shift-keying (BFSK) modulation case without diversity ($L = 1$). The following table shows the values of N used in the thesis:

Table 2.1. Some values of N used in the thesis.

M	L	N
2	1	1000
2	3	333
4	3	333
4	5	200
8	3	250

The output of the MFSK modulator is mixed with a frequency synthesizer, which is controlled by a PN code generator, to be pseudo-randomly hopped to one of the N non-overlapping FH bands. Following that, the resultant signal is passed through a bandpass filter (BPF) with bandwidth W_{ss} and then translated by a radio frequency (RF) oscillator for transmission.

The block diagram of an FFH/MFSK receiver employing different diversity-combining techniques is shown in Fig. 2.3. The desired signal is assumed to be corrupted by interference and AWGN. In this thesis, the fading is assumed to be constant over one hop, and independent from hop to hop and hence is similar to block fading channel (BFC) model. The BFC is characterized by the fact that the noise severity remains constant in blocks of consecutive transmitted symbols but is independent from block to block [29], [56]. This assumption makes our channel models mathematically tractable. The receiver consists of a frequency dehopper, a bank of $M = 2^k$ square-law detectors, the diversity-combining processing and the decision making device. We assume that the received signal is perfectly dehopped by a frequency synthesizer, which is controlled

Figure 2.1. An FFH/MFSK transmitter.

Figure 2.2. An illustration of FH bands and FH slots.

by a PN code generator. Following that, the dehopped signal is passed through a BPF and then detected by a bank of M square-law detectors. Following that, the square-law detector outputs r_{ml} are processed by different diversity-combining techniques, denoted as the function $F(\cdot)$ in Fig. 2.3, where $l = 1, 2, \dots, L$. These resultant outputs z_{ml} are then combined over L diversity receptions to form M decision statistics y_m , for $m = 1, 2, \dots, M$. Finally, the symbol corresponding to the largest combiner output is selected as the decision.

Figure 2.3. Block diagram of an FFH/MFSK diversity-combining receiver.

2.1.1 Maximum-Likelihood Receiver

The function $F(\cdot)$ of the ML receiver can be expressed as [53]

$$z_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|S)}{p_{r_{ml}}(r_{ml}|NS)} \right), \quad (2.1)$$

where $p_{r_{ml}}(r_{ml}|S)$ represents the probability density function (pdf) of the square-law detector output when the desired signal is present and $p_{r_{ml}}(r_{ml}|NS)$ denotes the corresponding pdf when the signal is not present. The resultant random variables z_{ml} are

then summed over L diversity receptions to produce the final decision statistic as

$$y_m = \sum_{l=1}^L z_{ml} \quad (2.2)$$

2.1.2 Linear-Combining Receiver

For the LCR, the function $F(\cdot)$, as shown in Fig. 2.3, is given by

$$z_{ml} = F(r_{ml}) = r_{ml}. \quad (2.3)$$

2.1.3 Product-Combining Receiver

For the PCR, the square-law detector outputs r_{ml} are multiplied over the L hops to form the variable

$$w_m = \prod_{l=1}^L r_{ml}. \quad (2.4)$$

Since it is mathematically intractable, to derive the pdf of the PCR output, we can convert from multiplication into addition operations by using the natural logarithm function [57]. Since all the square-law detector outputs r_{ml} are positive, we can take the natural logarithm on both sides of (2.4) without affecting the final decision. Thus, the function $F(\cdot)$ for the PCR as shown in Fig. 2.3 is given by

$$z_{ml} = F(r_{ml}) = \ln(r_{ml}). \quad (2.5)$$

2.1.4 Self-Normalizing Receiver

For the self-normalizing receiver (also called normalized envelope detection (NED)), the square-law detector output of each hop is normalized by the sum of the outputs of all the M square-law detectors before the L hops receptions are combined. The function

$F(\cdot)$ for the self-normalizing receiver is given by

$$z_{ml} = F(r_{ml}) = \frac{r_{ml}}{\sum_{m=1}^M r_{ml}}. \quad (2.6)$$

2.1.5 Order Statistics-NED Receiver

For the order statistics-NED (OS-NED) diversity-combining receiver, it is a combination of order statistics and NED combining, in which the L outputs of each of the M square-law detectors are first ranked in the ascending order according to the magnitude as [58], [59]

$$r_{lm1}^{(r)} \leq r_{lm2}^{(r)} \leq \dots \leq r_{lmL}^{(r)}, \quad (2.7)$$

where $r_{lmi}^{(r)}$ denotes the ranked square-law detector outputs. Note that $i = 1, \dots, L$ is the rank of that particular output in the ordered list. Following that, the self-normalization operation is then performed on the matrix of sorted values.

2.2 Interference Models

In the following, we introduce some commonly encountered interference models that will be investigated in this thesis.

2.2.1 Partial-Band Noise Interference

The PBNI uniformly distributes its finite interference power over a fraction $0 < \gamma \leq 1$ of the total SS bandwidth W_{ss} . Thus, each hopping symbol will be interfered by PBNI with probability γ , while the probability that the FH band does not contain PBNI is $(1 - \gamma)$ [31]. The power of the PBNI in an FH slot of bandwidth B Hz is given by $\sigma_P^2 = N_P B / \gamma$, where $N_P = P_P / W_{ss}$ is the average PSD over entire spread-spectrum bandwidth [45]. PBNI places noise-interference energy across multiple channels in the

spectrum used by the targets. These channels may or may not be contiguous. It is assumed that the PBNI is contiguous and thus all of the MFSK tones are interfered if the interfering signal is present in that FH band [48]. Note that PBNI is particularly effective against a FH SS system because the desired signal will hop in and out of the interfering band and can be seriously degraded when it is in the interfering band.

2.2.2 Multitone Interference

MTI is normally classified as either band MTI or independent MTI. For the band MTI, each affected FH band contains the same number of interfering tones, whereas for the independent MTI model, an affected FH band may be interfered with as few as one tone and as many as M tones [60]. For a fixed total interference power, the single-tone per band MTI is considered in our analysis since it results in the poorest system performance as compared with other types of MTI [35], [46]. The strategy of MTI is to distribute its total interference power P_μ over Q equal-power interfering tones spreading randomly over W_{ss} . Then the power of each single interfering tone is $P_{\mu_T} = P_\mu/Q$. For the single-tone per-band MTI, the number of MTI tones Q is always less than or equal to the number of FH bands. An equivalent PSD of the MTI can be defined as $N_\mu = P_\mu/W_{ss}$. The probability that an FH band contains an interfering tone is Q/N , and the probability that an FH band does not contain an interfering tone is $1 - Q/N$. In Chapters 3 and 4 of the thesis, where the composite effects of PBNI and MTI are considered, the worst-case BER performance of the proposed ML receivers is presented. For each SIR level, we numerically search through the number of MTI tones Q and the PBNI fraction γ that give the worst-case BER.

2.2.3 Multiple-Access Interference

The increasing demand for bandwidth in mobile commercial services has fueled the search for higher capacity wireless systems. This search has created a commercial interest in SS systems which offer significant capacity improvement in MA systems [6]. One important multiple-access scheme is the FFH-MA system. The interference results from simultaneous transmission is called MAI. This interference occurs when the reference user and non-reference users transmit their signals at the same frequency band. The MAI over one hop can be expressed as

$$J(t) = \sum_{k=1}^{K_l} \sqrt{2}a_k \cos(2\pi f_k t + \theta_k), \quad (2.8)$$

where K_l is the number of interfering users, $\sqrt{2}a_k$ and θ_k denote the amplitude and the baseband frequency of the k th interfering user, respectively.

2.3 Literature Survey

In this section, we provide a literature survey on diversity-combining techniques for suppressing different types of interference in FFH/MFSK systems over various fading channels.

2.3.1 PBNI Mitigation

In the literature, several diversity-combining techniques have been proposed for mitigating the effects of PBNI on FFH system performance [16], [31], [32], [36]–[39], [41]–[45], [47], [48], [50], [61], [62]. The simplest diversity-combining receiver, namely the LCR, was proposed in [31] for an FFH/BFSK system in the presence of PBNI over non-fading channels. Both exact and approximate (multiple bound-parameter Cher-

noff bound) solutions were obtained by making use of the statistical independence of the hopping symbols under two separate assumptions, i.e., with and without thermal noise. The authors concluded that for this type of receiver, when bit energy is fixed and the diversity level L is increased, the system performance is degraded because of the dominance of increased noncoherent combining loss for higher values of L . In [32], the throughput performance of an FFH/MFSK LCR was studied with a fixed hop rate over a non-faded environment. In [32], both the PBNI and MTI were considered in the analysis and it has been shown that for fixed data rate systems, the worst-case MTI is more effective against the FFH/MFSK LCR than the worst-case PBNI for $M > 2$ and the opposite is true for $M = 2$. To mitigate the PBNI effect, the error performance of FFH systems employing the noise-normalizing method, also referred to as adaptive gain control (AGC), was analyzed in [37], [38], and [41]–[43] over different types of fading channels. The noise-normalization combining receiver has a “noise-only” channel for noise power measurement, which is used to normalize the output of each detector before L receptions are combined [41]. This normalization can prevent jammed hops from dominating in the decision process and therefore it improves the system performance [42]. In [37], the BFSK modulation scheme was considered in a non-faded environment and was then extended to the MFSK case in [38]. In [41] and [43], BER expressions for an FFH noise-normalization combining receiver were derived over frequency-nonselective Rician and Nakagami- m fading channels, respectively. In [37], [38], [41], and [43], the performance of the noise-normalization combining receiver for the case of perfect noise power measurement has been studied. However, in practice, it is not possible to have perfect measurement. In [42], Robertson *et al.* investigated the effect of inaccurate noise power measurement on the system performance. This receiver is called a receiver with non-ideal adaptive gain control. It was concluded that the adaptive gain control receiver is relatively insensitive to noise power measurement error and any negative

impact on system performance can be completely counter-measured by making a very simple modification, i.e., including a positive bias in the noise power measurement circuitry. In contrast with the LCR, the noise-normalization combining receiver has been shown to have diversity gains for BER improvement.

For the clipper receiver, the output for each diversity reception is passed through a clipping process to limit the effects of strong PBNI [36]. It has been investigated in [36]–[39]. This receiver requires the side information of signal output level in order to set the clipping threshold [36]. An alternative implementation approach using fast Fourier transform (FFT) was proposed in [39]. The receiver in this case is called the FFT-based clipper receiver. The motivation for deriving this receiver is that the FFT technique can achieve guaranteed accuracy and perfect reproducibility. It is also more flexible as compared with analog signal processing [39]. The FFT-based clipper receiver has been shown to achieve comparable system performance as compared with that of the square-law clipper receiver. Both the noise-normalized and clipper receivers require some knowledge of system parameters, i.e., the noise power and the signal output voltage, respectively. As a result, the self-normalization receiver which does not require measurement of noise or signal power is therefore introduced in [44]. In [44], the authors considered the BFSK modulation scheme over a non-fading channel. The upper bound on the probability of bit error for an FFH/MFSK system employing self-normalizing method was derived and analyzed in [45] over a frequency-nonselective Rician fading channel. It has been demonstrated in [45] that using higher modulation orders M can improve the system performance and this improvement is more significant for the case of Rician fading as compared to the case of Rayleigh fading.

Viswanathan and Taghizadeh proposed a new scheme of diversity combining, namely the PCR, for an FFH/BFSK system over non-fading channels [47]. The major advantage of this receiver is that it has comparable performance with that of the clipper

receiver, while at the same time it does not require any side information to be implemented. In [16], the performance of an FFH/BFSK PCR over Rayleigh fading channels was analyzed. Based on the theory of the product and quotient of random variables, the authors derived the exact closed-form BER expressions in terms of Meijer's G function and infinite series. The performance of an FFH/MFSK PCR over Rayleigh fading channels was analyzed in [48] using the Melin-transform-based technique. The pdf and cumulative distribution function (cdf) expressions of the PCR outputs were derived in a semi-closed form and the system performance was evaluated. Keller and Pursley proposed two diversity-combining schemes based on ratio statistics for mitigating PBNI effects in FFH systems [61], [62]. These receivers do not require side information. Based on the likelihood ratio, the ML receiver was proposed in [50] for an FFH/MFSK system to reject PBNI in a non-faded environment. Side information on noise variance is required to implement this receiver. The BER results presented in [50] were obtained via simulation.

2.3.2 MTI Mitigation

In this section, we consider the mitigation of MTI in FFH systems. Robertson and Shelly studied the effects of both band MTI and independent MTI on the performance of FFH/MFSK systems without diversity-combining over a frequency-nonselective Rician fading channel [60]. It has been shown that the band MTI gives poorer system performance. In [33], the author derived BER expressions of an FFH/MFSK LCR in terms of an exponentially tight Chernoff bound over non-fading channels. Based on the circularly symmetric function theory, Fourier-Bessel series, and Fourier series, the expressions of pdf and cdf of the final decision statistics for an FFH/MFSK LCR were derived in [34] over a non-faded environment. The performance of the LCR over a frequency-nonselective Rician fading channel was investigated in [35]. The authors also

analyzed the impact of fading of the desired signal and the interfering tones on system performance.

In [40], the effect of band MTI on the FFH systems employing CR was investigated over non-fading channels. It has been shown in [40] that BER performance of the CR can be significantly improved with appropriate choice of the clipping level. In [63], based on pattern recognition concepts, the authors proposed new diversity-combining techniques, namely gradient diversity combining, to mitigate the effect of MTI for FFH systems over non-fading channels. In [46] and [49], the authors investigated the effect of MTI on performance of FFH systems employing diversity-combining techniques that do not require side information such as self-normalization receiver [46] and ratio-statistic combining receiver [49]. The BER performance of the self-normalized receiver in the presence of band MTI was presented in [46] over a frequency-nonselective Rician fading channel. The performance comparison of the self-normalized receiver under MTI and PBNI demonstrated that MTI is more harmful than PBNI for the self-normalized receiver. In [49], the authors examined the FFH/BFSK ratio-statistic combining receiver for suppressing band MTI. Based on the circularly symmetric signal theory, the closed-form expressions of pdf and cdf of the ratio-statistic output have been derived. It has been shown that the performance of this receiver is sensitive to the fading effect on the desired signal, but it is insensitive to the fading effect on the interfering tones.

ML receivers have been studied in the presence of band MTI over non-faded, frequency-nonselective Rician and frequency-selective Rayleigh fading channels in [51], [53] and [52], respectively. The corresponding ML receiver structures were also proposed and analyzed. Due to the intractability of the analytical expression for the probability of bit error, some approximations have been adopted to derive the suboptimum ML receivers in [51] and [52]. Note that in [53], the term “suboptimum” refers to the two ML receivers derived separately for the case of Rayleigh and Rician fading channels. In [51], the sub-

optimum ML receivers have been shown to outperform the LCR, the self-normalized receiver and the PCR. However, to implement these suboptimum ML receivers, the side information of the signal-tone amplitude, the jamming-tone amplitude, the resultant signal-plus-jamming tone amplitude, and the variance of Gaussian noise is required. The noise variance can be estimated from a noise-only channel for noise power measurement [37], and the tone amplitudes (or powers) can be extracted from an energy detector [64]. In [53], it has been shown that the proposed suboptimum ML receivers significantly outperform the other existing receivers over fading channels. In [52], the multipath frequency-selective Rayleigh fading channel was modeled as a correlated two-path channel model within one hopping duration, i.e., all possible paths are summed up and then split into two main clumps for forming these two paths. In [12], the author proposed a log-likelihood combining metric which requires only easy-to-obtain system parameters, such as signal-to-noise ratio (SNR), signal-to-interference ratio (SIR), energy of signal, and jamming ratio. The MTI and frequency-nonselective Rayleigh fading were considered in [12]. An upper bound on the bit error rate of the ML receiver using the proposed metric was also presented. It was found that the ML receiver employing the proposed metric significantly outperforms the other existing diversity-combining receivers, i.e., the LCR, PCR, CR and self-normalization receiver, and the performance improvement is most prominent for low SIR, which is of practical interest. Based on the property that the multipath components of the signal and the jamming tones can be combined with AWGN, the compact and exact symbol-error rate expressions for both slow and fast FH/MFSK systems employing LCR under independent MTI over Rician fading channels have been presented in [65]. The system performance for independent MTI and single-tone per band MTI was compared and it has been concluded that the single-tone per band MTI gives worse BER performance. In Sections 2.3.1 and 2.3.2, the literature survey on various diversity-combining techniques for FFH communica-

tion systems has been presented over various fading channels. These techniques have been proposed for mitigating the effect of only one type of interference (i.e., either PBNI or MTI) on the performance of FFH systems. However, in practice, PBNI and MTI may coexist concurrently in FFH systems. In addition, in the literature, most of performance analyses are based on the assumption that there is no timing and frequency offsets in the systems. In practical systems, there always exists some mismatch between the transmitter and the receiver. Therefore, it is important to find a diversity-combining receiver that can effectively suppress the composite effects of PBNI and MTI with or without timing and frequency offsets. This gives us motivation to propose the ML receiver structures as presented in Chapters 3 and 4 of this thesis. The proposed receivers have been shown to be more effective in suppressing the composite interference effects, as well as the timing and frequency offsets, than that of the conventional diversity-combining receivers.

2.3.3 Mitigation of MAI

The performance of an synchronous FH-MA system employing optimal diversity rules was investigated in [54] with either noncoherent or differentially coherent detection over frequency-nonsselective Rayleigh fading channels. Since the characteristic function (CF) expression derived in [54] is not in closed form, the obtained optimum nonlinearity was approximated by a soft limiter. Based upon the Gauss quadrature rule, Agusti presented a calculation method to evaluate performance of asynchronous FH-MA systems over either non-fading or frequency-nonsselective Rayleigh fading channels [66]. In [67], the authors investigated the performance of FFH-MA/BFSK systems employing the equal-gain diversity-combining technique over a frequency-selective Rayleigh fading environment. The performance of an FFH-MA/MFSK system for Rayleigh, Rician and log-normal multipath fading over a factory environment was studied in [68]. In [68], the

root-mean-square multipath delay spread for the factory environment is in the range of 100 to 200 nanoseconds and the approximate BER expressions were derived based on Stirling's formula. Assuming that the number of active users in the system is known, the optimal diversity-combining rule for a synchronous FH-MA system over a frequency-nonselective Rician fading channel was derived in [69]. Besides that, the authors also proposed three suboptimal receivers to replace the soft-limiter-like nonlinearity with a multi-level quantizer and analyzed their performance. In [70], the performance of FFH-MA/MFSK systems was examined over Rician, shadowed Rician and Nakagami fading channels. The authors also evaluated the system capacity in terms of the maximum number of users versus the BER for each fading case. For Rician fading, it was found in [70] that the capacity of the system with a Rician factor of 2 dB is reduced by 13% as compared to the capacity for the nonfading case. For shadowed Rician fading, the authors considered three shadowing scenarios, i.e., light, average, and heavy. The heavy shadowing scenario has exactly the same performance as that for the Rayleigh fading case. On the other hand, the light and average shadowing scenarios provide only a slight decrease in the capacity as compared with the nonfading case. For Nakagami fading, it was found that the system capacity decreases drastically by nearly 50% as compared to the nonfading case when the Nakagami fading parameter is equal to 0.5.

In [58], the authors investigated the performance of a synchronous FFH-MA/MFSK system employing various diversity-combining methods, i.e., the hard-decision majority-vote (HDMV), NED, OS-NED and reduced rank sum (RRS), over a frequency-nonselective Rayleigh fading environment. For the HDMV, the output of the detector is assigned a value of 1, if it exceeds a pre-set threshold, otherwise it will be assigned a value of 0 [59]. Since the HDMV combining method removes a significant portion of information available in the $L \times M$ matrix, it provides the worst system performance as compared to the other methods presented in [58]. The RRS combining method uses the ranks of the

values in the $L \times M$ matrix to make a decision rather than using actual values, whereas in the NED receiver, the output of each detector is normalized by the sum of the outputs of all M detectors [58], [59]. Among these four diversity-combining techniques, the OS-NED has been shown to have the best system performance [58]. The BER performance of synchronous FFH-MA/MFSK systems with various diversity-combining receivers, such as the HDMV, the soft limiting, the self-normalization, the PCR and the OS-NED, was examined and compared in [59] over Nakagami- m fading channels. It has been shown in [59] that despite of inferior performance in comparison to the optimum soft limiting combining receiver, the PCR and the OS-NED receiver do not require side information and hence constitute attractive design options. The performance of an asynchronous FFH-MA/MFSK system with hard-limited LCR was examined in [71] over a frequency-nonselective Rayleigh fading channel. The BER expressions for evaluating system performance are simple since one-dimensional numerical integrations were used in the analysis. The diversity of the FFH-MA system depends on the diversity order L and the total number of frequency bands in the system and thus these two parameters should be selected properly according to the number of users in order to obtain the best system capacity [71].

Chapter 3

Effects of Composite PBNI and MTI on FFH/MFSK Systems over Fading Channels

3.1 Overview

In this chapter, the composite effects of PBNI and MTI on performance of FFH/MFSK systems employing ML receivers are investigated. For FFH systems, the main sources of degradation in BER performance are the unintentional and intentional interference, as well as the fading effects. Among many types of intentional interference strategies, PBNI and MTI have been shown to be the most harmful against FFH communication systems [12], [52]. Although sometimes an interferer may want to concentrate its limited power in one interfering form, it is also possible for the interferer to divide its power between the two above-mentioned principal interfering forms in order to determine which strategy (MTI only, PBNI only or the composite interference effect) causes the most degradation to the communicator's system performance. Thus, it is important to

study the composite interference effect on the performance of FFH systems in order to determine the most effective interference strategy from the interferer's perspective.

3.2 Performance Study over Frequency-Nonselective Fading Channels

In frequency-nonselective fading, the coherence bandwidth of the channel is larger than the bandwidth of the signal. Thus, all frequency components of the signal will experience the same fading magnitude. For FFH systems, the bandwidth of each frequency slot B is equal to $L \times R_s$, where R_s is the symbol rate. Hence, for FFH systems with small diversity order L and low data rates R_s , the received signal will experience frequency-nonselective fading.

3.2.1 BER Performance Analysis

In this section, BER expressions of a suboptimum ML receiver with the composite interference effect are derived for FFH/MFSK systems over a frequency-nonselective Rician fading channel. We assume that the received signal is corrupted by AWGN, MTI and PBNI. The dehopped signal in the l th hop, for $l = 1, 2, \dots, L$, is given by

$$r_l(t) = s_l(t) + q_{ml}J_{\mu,l}(t) + q_l n_P(t) + \omega(t), \quad (3.1)$$

where $s_l(t)$ denotes the desired signal. The terms $J_{\mu,l}(t)$, $n_P(t)$ and $\omega(t)$ denote the MTI, PBNI and AWGN, respectively. The thermal noise has a variance of $N_0 B$ where N_0 is the one-sided PSD level.

In practice, there are a few approaches to model the received signal that experiences Rician fading channels. In [72], the received signal is assumed to consist of several specu-

lar components plus several Rayleigh fading components. Another approach is to model the Rician fading channel as the sum of a dominant line-of-sight (LOS) component and a Rayleigh fading component [73]. Based on the Rician fading model presented in [73], the desired signal $s_l(t)$ can be rewritten as

$$s_l(t) = \sqrt{2}\alpha_S \cos(2\pi f_m t + \phi_{S,l}) + \sqrt{2}|a_{S,l}| \cos(2\pi f_m t + \theta_{S,l}), \text{ for } m = 1, 2, \dots, M \quad (3.2)$$

where $\sqrt{2}\alpha_S$ and $\phi_{S,l}$ represent the amplitude and the random phase of the LOS component of the desired signal, respectively. The terms $\sqrt{2}|a_{S,l}|$ and $\theta_{S,l}$ are the amplitude and the random phase of the Rayleigh fading component, respectively. Note that f_m is the baseband frequency of the desired signal.

In (3.1), $q_l = 1$ or 0 is defined as the PBNI state, whereas $q_{ml} = 1$ or 0 indicates the MTI state of the m th branch in the l th hop. The MTI $J_{\mu,l}(t)$ is modeled the same way as that of the desired signal with parameters $\alpha_S, a_{S,l}, f_m, \phi_{S,l}$ and $\theta_{S,l}$ replaced by $\alpha_\mu, a_{\mu,l}, f_\chi, \phi_{\mu,l}$ and $\theta_{\mu,l}$, respectively. Each MTI tone in the l th hop is given by

$$J_{\mu,l}(t) = \sqrt{2}\alpha_\mu \cos(2\pi f_\chi t + \phi_{\mu,l}) + \sqrt{2}|a_{\mu,l}| \cos(2\pi f_\chi t + \theta_{\mu,l}), \text{ for } \chi = 1, 2, \dots, M. \quad (3.3)$$

We define the corresponding signal-to-interference power ratio as

$$\text{SIR} = E_b/N_J \quad (3.4)$$

where N_J is the equivalent PSD of both PBNI and MTI. The total interference power is $P_J = P_P + P_\mu$. We also define the power ratios of PBNI and MTI with respect to the total interference power as $\nu_P = P_P/P_J$ and $\nu_\mu = P_\mu/P_J$, respectively.

In our analysis, we assume that the signal and the MTI tone experience independent fading [35]. The channel for each hop is modeled as a frequency-nonselective slow

fading Rician process. In other words, we assume that the signal bandwidth and the interference bandwidth are much smaller than the coherence bandwidth of the channel and that the hop rate is much greater than the Doppler spread of the channel [74]. The total average power of the signal $P_S = \alpha_S^2 + 2\sigma_S^2$ and the average interference tone power $P_\mu = \alpha_\mu^2 + 2\sigma_\mu^2$ are assumed to remain constant from hop to hop. The parameters α_S^2 and $2\sigma_S^2$ represent the power levels of the direct and diffused components of the desired signal, respectively. Similarly, α_μ^2 and $2\sigma_\mu^2$ are the power levels of the direct and diffused components of the MTI tone, respectively. The Rician fading ratios $F_S = \alpha_S^2/2\sigma_S^2$ and $F_\mu = \alpha_\mu^2/2\sigma_\mu^2$ determine different levels of fading conditions, e.g., essentially no fading ($F_S = 1000$ or $F_\mu = 1000$), Rician fading ($F_S = 10$ or $F_\mu = 10$) and Rayleigh fading ($F_S = 0$ or $F_\mu = 0$).

Without loss of generality, we assume that the desired signal is present in the first branch of the receiver with frequency f_1 . The decision random variable y_m , for $m = 1, 2, \dots, M$, can be expressed as

$$y_m = \sum_{l=1}^L z_{ml} \quad (3.5)$$

where the random variable z_{ml} for the suboptimum ML receiver is given by

$$z_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \phi, f_m = f_1)}{p_{r_{ml}}(r_{ml}|q_{ml}, q_l, f_m \neq f_1)} \right). \quad (3.6)$$

In (3.6), the numerator of the natural logarithm function represents the pdf of the square-law detector output when the desired signal is present and the denominator denotes the corresponding pdf when the signal is not present. The square-law detector output r_{ml} can be expressed as

$$r_{ml} = r_{I,ml}^2 + r_{Q,ml}^2 \quad (3.7)$$

where

$$r_{I,ml} = \frac{\sqrt{2}}{T_h} \int_{(l-1)T_h}^{lT_h} r_l(t) \cos(2\pi f_m t) dt$$

and

$$r_{Q,ml} = \frac{\sqrt{2}}{T_h} \int_{(l-1)T_h}^{lT_h} r_l(t) \sin(2\pi f_m t) dt$$

are the corresponding in-phase and quadrature-phase components, respectively [1].

After some mathematical manipulations, we can obtain the expressions of the in-phase and quadrature-phase components as

$$\begin{aligned} r_{I,ml} = & \alpha_S \cos(\phi_{S,l}) \delta(f_m - f_1) + |a_{S,l}| \cos(\theta_{S,l}) \delta(f_m - f_1) \\ & + q_{ml} \alpha_\mu \cos(\phi_{\mu,l}) + q_{ml} |a_{\mu,l}| \cos(\theta_{\mu,l}) + n_{I,l} \end{aligned} \quad (3.8)$$

and

$$\begin{aligned} r_{Q,ml} = & -\alpha_S \sin(\phi_{S,l}) \delta(f_m - f_1) - |a_{S,l}| \sin(\theta_{S,l}) \delta(f_m - f_1) \\ & - q_{ml} \alpha_\mu \sin(\phi_{\mu,l}) - q_{ml} |a_{\mu,l}| \sin(\theta_{\mu,l}) - n_{Q,l}, \end{aligned} \quad (3.9)$$

respectively. After some mathematical manipulations, we obtain the following expressions:

$$u_{I,ml} = \frac{A_{ml}}{\sqrt{2}} + x_{S,l} \sigma_S \delta(f_m - f_1) + q_{ml} x_{\mu,l} \sigma_\mu + n_{I,l} \quad (3.10)$$

and

$$u_{Q,ml} = \frac{A_{ml}}{\sqrt{2}} + y_{S,l} \sigma_S \delta(f_m - f_1) + q_{ml} y_{\mu,l} \sigma_\mu + n_{Q,l}, \quad (3.11)$$

where

$$A_{ml}^2 = \alpha_S^2 \delta(f_m - f_1) + q_{ml}^2 \alpha_\mu^2 + 2q_{ml} \alpha_S \alpha_\mu \cos(\phi_{S,l} - \phi_{\mu,l}) \delta(f_m - f_1). \quad (3.12)$$

Note that $\delta(\cdot)$ denotes the Dirac delta function. Furthermore, the variables $x_{S,l}$, $y_{S,l}$, $x_{\mu,l}$ and $y_{\mu,l}$ are independently Gaussian distributed with zero mean and variance of 1, whereas $n_{I,l}$ and $n_{Q,l}$ are independently Gaussian distributed with zero mean and variance of $(\sigma_\omega^2 + q_l \sigma_P^2)/2$. The terms $u_{I,ml}$ and $u_{Q,ml}$ in (3.10) and (3.11) are related to $r_{I,ml}$ and $r_{Q,ml}$ in (3.8) and (3.9) by the following expression

$$u_{I,ml}^2 + u_{Q,ml}^2 = r_{I,ml}^2 + r_{Q,ml}^2. \quad (3.13)$$

The random variables $u_{I,ml}$ and $u_{Q,ml}$ are Gaussian distributed with a mean value of $A_{ml}/\sqrt{2}$ and a common variance given by

$$\sigma_{ml}^2 = \sigma_S^2 \delta(f_m - f_1) + q_{ml} \sigma_\mu^2 + \frac{\sigma_\omega^2}{2} + \frac{q_l \sigma_P^2}{2}. \quad (3.14)$$

Hence, $r_{ml} = r_{I,ml}^2 + r_{Q,ml}^2 = u_{I,ml}^2 + u_{Q,ml}^2$ is a non-central chi-squared random variable with two degrees of freedom and A_{ml}^2 is its noncentrality parameter [74].

The pdf of the square-law detector output r_{ml} can be expressed as [74]

$$p_{r_{ml}}(r_{ml}|q_{ml}, q_l, f_m, \phi) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml} + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{A_{ml}\sqrt{r_{ml}}}{\sigma_{ml}^2}\right) U(r_{ml}). \quad (3.15)$$

From (3.15), when the desired signal is present, we have

$$p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \phi, f_m = f_1) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml} + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) I_0\left(\frac{A_{ml,S}\sqrt{r_{ml}}}{\sigma_{ml,S}^2}\right) U(r_{ml}), \quad (3.16)$$

where

$$\sigma_{ml,S}^2 = \sigma_S^2 + q_{ml} \sigma_\mu^2 + \frac{\sigma_\omega^2}{2} + \frac{q_l \sigma_P^2}{2} \quad (3.17)$$

and

$$A_{ml,S} = \sqrt{\alpha_S^2 + q_{ml}\alpha_\mu^2 + 2q_{ml}\alpha_S\alpha_\mu \cos \phi}. \quad (3.18)$$

Note that $\phi = \phi_{s,l} - \phi_{\mu,l}$ is the phase difference between the signal tone and the MTI tone, which can be modeled as a random variable uniformly distributed over $[0, 2\pi]$.

For the non-signal branch, the pdf of the square-law detector output r_{ml} can be written as

$$p_{r_{ml}}(r_{ml}|q_{ml}, q_l, f_m \neq f_1) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml} + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) I_0\left(\frac{A_{ml,NS}\sqrt{r_{ml}}}{\sigma_{ml,NS}^2}\right) U(r_{ml}) \quad (3.19)$$

where

$$\sigma_{ml,NS}^2 = q_{ml}\sigma_\mu^2 + \frac{\sigma_\omega^2}{2} + \frac{q_l\sigma_P^2}{2} \quad (3.20)$$

and

$$A_{ml,NS} = q_{ml}\alpha_\mu. \quad (3.21)$$

Substituting (3.16) and (3.19) into (3.6), we obtain

$$z_{ml} = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2}\right) r_{ml} + \ln\left(\frac{I_0\left(\frac{A_{ml,S}\sqrt{r_{ml}}}{\sigma_{ml,S}^2}\right)}{I_0\left(\frac{A_{ml,NS}\sqrt{r_{ml}}}{\sigma_{ml,NS}^2}\right)}\right) + \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2} + \ln\left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2}\right). \quad (3.22)$$

Note that the ML receiver structure presented in (3.22) is optimum. Nevertheless, the computation of the function $\ln I_0(\cdot)$ in (3.22) is nonlinear in nature and thus the analytical BER expressions for the optimum ML receiver is mathematically intractable.

Hence, for some practical SNR ranges, i.e., $\text{SNR} > 6$ dB, the approximation $\ln I_0(x) = x$ is used to simplify the expression of z_{ml} as [51], [52]

$$z_{ml} = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \left(\frac{A_{ml,S}}{\sigma_{ml,S}^2} - \frac{A_{ml,NS}}{\sigma_{ml,NS}^2} \right) \sqrt{r_{ml}} + \lambda_{ml} \quad (3.23)$$

where

$$\lambda_{ml} = \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2} + \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right). \quad (3.24)$$

After some mathematical manipulations, we get

$$z_{ml} = \zeta_{ml} (\sqrt{r_{ml}} + \beta_{ml})^2 + \gamma_{ml} \quad (3.25)$$

where

$$\zeta_{ml} = \frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2}, \quad (3.26)$$

$$\beta_{ml} = \frac{A_{ml,NS}\sigma_{ml,S}^2 - A_{ml,S}\sigma_{ml,NS}^2}{\sigma_{ml,NS}^2 - \sigma_{ml,S}^2} \quad (3.27)$$

and

$$\gamma_{ml} = \lambda_{ml} - \zeta_{ml}\beta_{ml}^2. \quad (3.28)$$

From (3.25), it can be observed that the proposed suboptimum ML receiver requires the side information on the noise variance; the power level of MTI; the PBNI variance; and the Rician fading factor. The Gaussian noise variance can be obtained from a “noise-only” channel for noise power measurement [37], whereas the power level of MTI, and the PBNI variance can be extracted from an energy detector [64]. The Rician fading factor can be obtained by using moment-based estimation method [75] or ML estimation method [76]. The moment-based estimator in [75] uses the first and the second order moments of the fading envelope. In [76], the authors proposed the ML estimator using

samples of both the fading envelope and the fading phase.

The conditional pdf of the suboptimum ML receiver output z_{ml} , when signal is present or not present, can be obtained as [74]

$$\begin{aligned}
p_{z_{ml}}(z_{ml}|q_{ml}, q_l, \phi, f_m = f_1) \\
= \frac{\psi_{ml}}{2\sigma_{ml,S}^2 \sqrt{\zeta_{ml}} (z_{ml} - \gamma_{ml})} \exp\left(-\frac{\psi_{ml}^2 + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) I_0\left(\frac{A_{ml,S}\psi_{ml}}{\sigma_{ml,S}^2}\right) U(\psi_{ml}) \\
- \frac{\xi_{ml}}{2\sigma_{ml,S}^2 \sqrt{\zeta_{ml}} (z_{ml} - \gamma_{ml})} \exp\left(-\frac{\xi_{ml}^2 + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) I_0\left(\frac{A_{ml,S}\xi_{ml}}{\sigma_{ml,S}^2}\right) U(-\xi_{ml}) \quad (3.29)
\end{aligned}$$

and

$$\begin{aligned}
p_{z_{ml}}(z_{ml}|q_{ml}, q_l, \phi, f_m \neq f_1) \\
= \frac{\psi_{ml}}{2\sigma_{ml,NS}^2 \sqrt{\zeta_{ml}} (z_{ml} - \gamma_{ml})} \exp\left(-\frac{\psi_{ml}^2 + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) I_0\left(\frac{A_{ml,NS}\psi_{ml}}{\sigma_{ml,NS}^2}\right) U(\psi_{ml}) \\
- \frac{\xi_{ml}}{2\sigma_{ml,NS}^2 \sqrt{\zeta_{ml}} (z_{ml} - \gamma_{ml})} \exp\left(-\frac{\xi_{ml}^2 + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) I_0\left(\frac{A_{ml,NS}\xi_{ml}}{\sigma_{ml,NS}^2}\right) U(-\xi_{ml}) \quad (3.30)
\end{aligned}$$

respectively, where

$$\psi_{ml} = \sqrt{\frac{z_{ml} - \gamma_{ml}}{\zeta_{ml}}} - \beta_{ml} \quad (3.31)$$

and

$$\xi_{ml} = \sqrt{\frac{z_{ml} - \gamma_{ml}}{\zeta_{ml}}} + \beta_{ml}. \quad (3.32)$$

In order to obtain the unconditional pdf on parameter ϕ , we use the following approximation based on Taylor series expansion in central differences [77]

$$p_{z_{ml}}(z_{ml}|q_{ml}, q_l, f_m) \approx \frac{1}{3} \sum_{k=-1}^1 p\left(z_{ml}|q_{ml}, q_l, f_m, \cos\phi = \frac{\sqrt{3}k}{2}\right). \quad (3.33)$$

The final decision statistics are given by

$$\hat{y}_m = \sum_{l=1}^L \hat{z}_{ml} . \quad (3.34)$$

By applying the quantization approach, we can obtain the pdf of the discrete random variable $\hat{z}_{ml}$. Following that, the pdf of the final decision statistics $\hat{y}_m$ can be expressed as

$$p_{\hat{y}_1}(\hat{y}_1 | \mathbf{K}_\mu, \mathbf{K}_\mathbf{P}) = \left[p_{\hat{z}_{1l}}(\hat{z}_{1l} | q_{1l}, q_l, f_m = f_1) \right]^{\otimes L} \quad (3.35)$$

and

$$p_{\hat{y}_k}(\hat{y}_k | \mathbf{K}_\mu, \mathbf{K}_\mathbf{P}) = \left[p_{\hat{z}_{kl}}(\hat{z}_{kl} | q_{kl}, q_l, f_m \neq f_1) \right]^{\otimes L} \quad (3.36)$$

for $k = 2, 3, \dots, M$, where the superscript $\otimes L$ denotes the $(L - 1)$ -fold discrete convolution operation; $\mathbf{K}_\mu$ and $\mathbf{K}_\mathbf{P}$ represent the MTI and PBNI distribution patterns, respectively, given by

$$\mathbf{K}_\mu = \begin{pmatrix} q_{11} & q_{12} & \dots & q_{1L} \\ q_{21} & q_{22} & \dots & q_{2L} \\ \vdots & \vdots & \ddots & \vdots \\ q_{M1} & q_{M2} & \dots & q_{ML} \end{pmatrix} \quad (3.37)$$

and

$$\mathbf{K}_\mathbf{P} = (q_1, q_2, \dots, q_L) . \quad (3.38)$$

For orthogonal signaling, the average BER is given by [74]

$$P_b = \frac{M}{2(M - 1)} P_s . \quad (3.39)$$

The symbol error probability, P_s , is obtained by averaging the conditional probability

of symbol error over all the possible combinations of $\mathbf{K}_\mu$ and $\mathbf{K}_\mathbf{P}$, that is

$$P_s = \sum_{\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}} P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}) P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}}. \quad (3.40)$$

Note that $P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P})$ is the probability of both MTI and PBNI patterns, given by

$$\begin{aligned} P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}) &= \gamma^{\sum_{l=1}^L q_l} (1 - \gamma)^{L - \sum_{l=1}^L q_l} \left(\frac{Q}{MN} \right)^{\sum_{l=1}^L \sum_{m=1}^M q_{ml}} \\ &\times \left(1 - \frac{Q}{N} \right)^{L - \sum_{l=1}^L \sum_{m=1}^M q_{ml}} \end{aligned} \quad (3.41)$$

and $P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}}$ is the conditional symbol error probability, conditioned on the interference patterns described by matrices $\mathbf{K}_\mu$ and $\mathbf{K}_\mathbf{P}$, which is given by

$$\begin{aligned} P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}} &= 1 - \sum_{g_1 = -GL/2}^{GL/2} p_{\hat{y}_1|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}} \left(\frac{2g_1 C}{G} \right) \left[\sum_{g_k = -GL/2}^{g_1} p_{\hat{y}_k|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}} \left(\frac{2g_k C}{G} \right) \right]^{M-1}, \\ &\text{for } k = 2, 3, \dots, M, \end{aligned} \quad (3.42)$$

where C denotes the upper limit of quantization range and G is the number of steps used in the quantization process.

3.2.2 Numerical Results and Discussions

In this section, the worst-case BER performance of the proposed suboptimum ML receiver under the composite effect of MTI and PBNI over Rician fading channels is presented. In the context of this thesis, we use the term BER for both analytical and simulation results. Similar definition has been adopted in some existing research works such as [4], [11], [48], [52]. Figure 3.1 presents the worst-case BER with Rician-faded signal ($F_S = 10$), Rician-faded MTI tone ($F_\mu = 10$), $\nu_\mu = 0.5$, and various diversity

Figure 3.1. Worst-case analytical and simulated BER results of FFH suboptimum ML receiver with $\text{SNR} = 13.35$ dB, $M = 4$, $\nu_\mu = 0.5$, $F_\mu = 10$ and $F_S = 10$ for various L .

levels L . The close match between the analytical and simulation results validates the theoretical analysis presented in Section 3.2.1. It can also be observed from Fig. 3.1 that for $5 < \text{SIR} < 15$ dB, the optimum diversity order is 3, while at moderate-to-high SIR regions (15-50 dB), the optimum diversity order is 5. This phenomenon can be explained as follows: Diversity can reduce the depth of fades and/or the fade duration by supplying receiver with multiple replicas of the transmitted signal that have undergone independent fading channels [9]. However, the use of diversity can

Figure 3.2. Worst-case BER performance of FFH suboptimum ML receiver with $\text{SNR} = 16$ dB, $M = 4$, $L = 3$, different interference conditions and different fading channels.

lead to the noncoherent diversity-combining loss [74]. Thus, there exists an optimum diversity level for the proposed system. This is the tradeoff between gain achieved due to interference countermeasure and loss due to higher non-coherent diversity-combining loss. For strong-to-moderate interference conditions ($5 < \text{SIR} < 25$ dB), the use of diversity ($L = 3$ or 5) improves the system performance significantly, i.e., up to one order of magnitude in the improvement in BER performance. Hence, we use the diversity order $L = 3$ and $L = 5$ for the subsequent results of the proposed system.

In Fig. 3.2, we present the BER performance of the proposed system over Rayleigh

Figure 3.3. Worst-case BER performance of FFH suboptimum ML receiver with $\text{SNR} = 13.35$ dB, $M = 4$, $L = 3$ and various ν_μ with Rician-faded MTI ($F_\mu = 10$) for different fading conditions of signal tone.

and Rician fading channels. The desired signal and the MTI tone are assumed to experience the same type of fading ($F_S = F_\mu$). When the interference power is high, we observe that the BER result under the case of $\nu_\mu = 1$ is the best. The improvement in BER performance can be as large as three orders of magnitude. This shows that the proposed suboptimum ML receiver can suppress the MTI more effectively as compared to that of the PBNI. This is because each MTI tone is basically a sinusoidal waveform and it is easier for the proposed ML receiver to estimate and mitigate the MTI as

Figure 3.4. Worst-case BER performance of FFH suboptimum ML receiver with $\text{SNR} = 13.35$ dB, $L = 3$, $\nu_\mu = 0.5$ with Rician-faded MTI ($F_\mu = 10$) and Rician-faded signal ($F_S = 10$) for various M .

compared to that of the PBNI which is Gaussian noise in nature.

In Fig. 3.3, we fix the MTI Rician fading factor to be $F_\mu = 10$ and investigate the system performance with various F_S levels. It is shown that the proposed suboptimum ML receiver gives the best performance when the desired signal does not undergo fading ($F_S = 1000$). From Figs. 3.2 and 3.3, we observe that the system performance is not very sensitive to the types of MTI fading. On the other hand, as can be seen in Figs. 3.2 and 3.3, when the desired signal undergoes Rayleigh fading ($F_S = 0$), the BER is the

worst as compared with Rician fading ($F_S = 10$) or no fading ($F_S = 1000$) cases. For all different fading cases, when the interference power is high, we observe that the BER curve of the proposed system under the composite effect lies between the BER curves under one type of interference only ($\nu_\mu = 0$ or 1). It can also be observed that the BER curves under MTI only are almost flat throughout the range of SIR levels. This means that the proposed suboptimum ML receiver can almost completely suppress the MTI. In all the above figures, each BER curve has an error floor for large SIR level. For instance, from Fig. 3.3, for the case of SNR = 13.35 dB, $M = 4$, $L = 3$ and $F_S = F_\mu = 10$, this error floor is equal to 1.8×10^{-5} . This phenomenon is due to the fact that when SIR is large, the effect of the interference (MTI or PBNI or both) on the system performance is negligible and the error is mainly caused by AWGN.

The worst-case BER performance of the proposed suboptimum ML receiver and the LCR for various modulation orders of $M = 2, 4, 8$ and 16 is presented in Fig. 3.4. It can be seen that for the proposed suboptimum ML receiver, a larger modulation order M improves the system performance significantly as compared with the linear-combining receiver. Note that this system improvement is consistent with the BER performance of MFSK system with noncoherent detection shown in [74]. Since $N = \lfloor \frac{W_{ss}T_b \log_2 M}{ML} \rfloor$, where $\lfloor x \rfloor$ denotes the largest integer $< x$, we observe that when the modulation order M increases, the number of frequency-hop bands N decreases. This increases the probability that the MTI tone may be present in the FH band of the desired signal branch. Fortunately, when MTI is present, our proposed suboptimum ML receiver can still effectively suppress the MTI as compared to the LCR. Moreover, it can also be observed that the interference suppression of the proposed suboptimum ML receiver is better than the LCR for all modulation orders at all SIR levels.

In Fig. 3.5, we compare the BER performance of our proposed suboptimum ML receiver with some conventional diversity-combining receivers, namely, the linear-combining

Figure 3.5. Worst-case BER performance comparisons among various receivers with SNR = 13.35 dB, $M = 4$, $L = 5$, and various ν_μ over Rician fading channels ($F_S = F_\mu = 10$).

and product-combining receivers. The BER results of these two conventional receivers are obtained based on computer simulation. Both the desired signal and MTI tone are assumed to undergo Rician fading ($F_S = F_\mu = 10$). In addition, we set $M = 4$, $L = 5$ and $\nu_\mu = 0, 1$ and 0.5 . From Fig. 3.5, we observe that the proposed suboptimum ML receiver outperforms the linear-combining and product-combining receivers. This improvement in BER performance is more significant under low-to-moderate SIR levels, which are of practical interest. However, it should be noted that the BER improvement of the proposed suboptimum ML receiver is achieved at the expense of higher complex-

ity. The proposed suboptimum ML receiver requires more side information, i.e., noise variance, PBNI variance, Rician fading factors and the power level of MTI.

3.3 Performance Study over Frequency-Selective Fading Channels

In this section, we investigate the composite effect of both PBNI and MTI on an FFH/MFSK ML receiver over frequency-selective Rayleigh fading channels. Both the desired signal and the MTI are assumed to experience independent frequency-selective Rayleigh fading. In this thesis, to simplify the derivations and the analyses, we can sum up all possible non-line-of-sight paths and then split them into two main clumps to model the Rayleigh frequency-selective fading channel or the Rayleigh component of the frequency-selective Rician-fading channel as the two-path model [52], [78], [79]. Besides [52], [78] and [79], the used frequency-selective model is also widely adopted in research works of [80]–[82] to model frequency-selective fading channels due to its analytical simplicity. In [80], a new scheme of fractional-bit differential detection of minimum-shift-keying was proposed in order to reduce timing fluctuation of eye patterns, which is found to be a cause of abnormally long burst errors in a frequency-selective fading channel. In [81], the authors presented two methods for the computation of the error probability of M-ary differential phase-shift-keying with differential phase detection in a two-path Rayleigh channel taking into account intersymbol, cochannel, and adjacent channel interference as well as Doppler frequency shift. The effect of frequency-selective multipath fading using two-path model on the performance of noncoherent delay-locked loops for spread spectrum systems was investigated in [82].

3.3.1 BER Performance Analysis

The desired signal $s_l(t)$ in (3.1) can be expressed as

$$s_l(t) = \sqrt{2}|a_{s_1,l}|\cos(2\pi f_m t + \theta_{s_1,l}) + \sqrt{2}|a_{s_2,l}|\cos(2\pi f_m(t - \tau_s) + \theta_{s_2,l}), \quad (3.43)$$

where $\sqrt{2}|a_{s_1,l}|$ and $\sqrt{2}|a_{s_2,l}|$ are the two main correlated Rayleigh amplitudes; $\theta_{s_1,l}$ and $\theta_{s_2,l}$ represent their corresponding phases in the l th hop, which are uniformly distributed over $[0, 2\pi]$. Note that τ_s denotes the time delay between two paths and is assumed to be uniformly distributed over $(0, T_h]$. Similarly, the MTI $J_{\mu,l}(t)$ is modeled the same way as that of the desired signal with parameters $a_{s_1,l}, a_{s_2,l}, f_m, \theta_{s_1,l}, \theta_{s_2,l}$ and τ_s replaced by $a_{\mu_1,l}, a_{\mu_2,l}, f_\chi, \theta_{\mu_1,l}, \theta_{\mu_2,l}$ and τ_μ , respectively. It is given by

$$J_{\mu,l}(t) = \sqrt{2}|a_{\mu_1,l}|\cos(2\pi f_\chi t + \theta_{\mu_1,l}) + \sqrt{2}|a_{\mu_2,l}|\cos(2\pi f_\chi(t - \tau_\mu) + \theta_{\mu_2,l}),$$

for $\chi = 1, 2, \dots, M$. (3.44)

The two main correlated Rayleigh amplitudes of the desired signal are generated from two uncorrelated complex Gaussian-fading amplitudes with unit variance, $b_{s_1,l} = x_{s_1,l} + jy_{s_1,l}$ and $b_{s_2,l} = x_{s_2,l} + jy_{s_2,l}$ [83]. Note that $x_{s_1,l}, x_{s_2,l}, y_{s_1,l}$ and $y_{s_2,l}$ are independent Gaussian random variables with zero mean and variance of one. Hence, $s_l(t)$ can be rewritten as

$$s_l(t) = \sqrt{2}\sigma_{s_1}|x_{s_1,l} + jy_{s_1,l}|\cos(2\pi f_m t + \theta_{s_1,l})$$

$$+ \sqrt{2}\sigma_{s_2}\left[\lambda_s^2|x_{s_1,l} + jy_{s_1,l}|^2 + (1 - \lambda_s^2)|x_{s_2,l} + jy_{s_2,l}|^2\right. \\ \left.+ 2\lambda_s\sqrt{1 - \lambda_s^2}(x_{s_1,l}x_{s_2,l} + y_{s_1,l}y_{s_2,l})\right]^{1/2}\cos(2\pi f_m(t - \tau_s) + \theta_{s_2,l}) \quad (3.45)$$

where $\varepsilon_s = \sigma_{s_2}^2/\sigma_{s_1}^2$ denotes the power ratio of the two main correlated paths of the

desired signal and λ_s is a function of the cross-correlation coefficient of two main paths, given by

$$\lambda_s^2 \cong 8 \left[\sqrt{(4/\pi - 1) \rho_s + 1} - 1 \right]. \quad (3.46)$$

Note that $2\sigma_{s_1}^2$ and $2\sigma_{s_2}^2$ are the diffused power levels of the first and second main paths of the desired signal, respectively. Similarly, the MTI $J_{\mu,l}(t)$ is given by

$$\begin{aligned} J_{\mu,l}(t) = & \sqrt{2}\sigma_{\mu_1}|x_{\mu_1,l} + jy_{\mu_1,l}|\cos(2\pi f_\chi t + \theta_{\mu_1,l}) + \sqrt{2}\sigma_{\mu_2} \left[\lambda_\mu^2 |x_{\mu_1,l} + jy_{\mu_1,l}|^2 \right. \\ & \left. + (1 - \lambda_\mu^2) |x_{\mu_2,l} + jy_{\mu_2,l}|^2 + 2\lambda_\mu \sqrt{1 - \lambda_\mu^2} (x_{\mu_1,l}x_{\mu_2,l} + y_{\mu_1,l}y_{\mu_2,l}) \right]^{1/2} \\ & \times \cos(2\pi f_\chi(t - \tau_\mu) + \theta_{\mu_2,l}), \end{aligned} \quad (3.47)$$

where $2\sigma_{\mu_1}^2$ and $2\sigma_{\mu_2}^2$ are the diffused power levels of the first and second main paths of the MTI, respectively. The power ratio of the two main paths of the MTI is defined as $\varepsilon_\mu = \sigma_{\mu_2}^2/\sigma_{\mu_1}^2$ and λ_μ^2 is given by

$$\lambda_\mu^2 \cong 8 \left[\sqrt{(4/\pi - 1) \rho_\mu + 1} - 1 \right], \quad (3.48)$$

where ρ_μ is the cross-correlation coefficient of the two main paths of the MTI.

The random variable z_{ml} shown in (3.5) for the ML receiver can be expressed as

$$z_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \theta, f_m = f_1)}{p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \theta, f_m \neq f_1)} \right). \quad (3.49)$$

In (3.49), the numerator of the natural logarithm function represents the pdf of the square-law detector output when the desired signal is present and the denominator denotes the corresponding pdf when the signal is not present. After some mathematical manipulations, the expressions of the in-phase and quadrature-phase components can

be obtained as

$$\begin{aligned}
r_{I,ml} = & \sigma_{s_1} |b_{s_1,l}| \cos(\theta_{s_1,l}) \delta(f_m - f_1) + \sigma_{s_2} \left[\lambda_s^2 |b_{s_1,l}|^2 + (1 - \lambda_s^2) |b_{s_2,l}|^2 \right. \\
& \left. + \lambda_s \sqrt{1 - \lambda_s^2} (x_{s_1,l} x_{s_2,l} + y_{s_1,l} y_{s_2,l}) \right]^{1/2} \cos(\theta_{s_2,l} - 2\pi f_m \tau_s) \delta(f_m - f_1) \\
& + q_{ml} \sigma_{\mu_1} |b_{\mu_1,l}| \cos(\theta_{\mu_1,l}) + q_{ml} \sigma_{\mu_2} \left[\lambda_\mu^2 |b_{\mu_1,l}|^2 + (1 - \lambda_\mu^2) |b_{\mu_2,l}|^2 \right. \\
& \left. + \lambda_\mu \sqrt{1 - \lambda_\mu^2} (x_{\mu_1,l} x_{\mu_2,l} + y_{\mu_1,l} y_{\mu_2,l}) \right]^{1/2} \cos(\theta_{\mu_2,l} - 2\pi f_\chi \tau_\mu) + n_{I,l} \quad (3.50)
\end{aligned}$$

and

$$\begin{aligned}
r_{Q,ml} = & -\sigma_{s_1} |b_{s_1,l}| \sin(\theta_{s_1,l}) \delta(f_m - f_1) - \sigma_{s_2} \left[\lambda_s^2 |b_{s_1,l}|^2 + (1 - \lambda_s^2) |b_{s_2,l}|^2 \right. \\
& \left. + \lambda_s \sqrt{1 - \lambda_s^2} (x_{s_1,l} x_{s_2,l} + y_{s_1,l} y_{s_2,l}) \right]^{1/2} \sin(\theta_{s_2,l} - 2\pi f_m \tau_s) \delta(f_m - f_1) \\
& - q_{ml} \sigma_{\mu_1} |b_{\mu_1,l}| \sin(\theta_{\mu_1,l}) - q_{ml} \sigma_{\mu_2} \left[\lambda_\mu^2 |b_{\mu_1,l}|^2 + (1 - \lambda_\mu^2) |b_{\mu_2,l}|^2 \right. \\
& \left. + \lambda_\mu \sqrt{1 - \lambda_\mu^2} (x_{\mu_1,l} x_{\mu_2,l} + y_{\mu_1,l} y_{\mu_2,l}) \right]^{1/2} \sin(\theta_{\mu_2,l} - 2\pi f_\chi \tau_\mu) - n_{Q,l}, \quad (3.51)
\end{aligned}$$

respectively. Note that $n_{I,l}$ and $n_{Q,l}$ are independently Gaussian distributed with zero mean and variance of $(\sigma_\omega^2 + q_l \sigma_P^2)/2$.

Applying the similar technique used in [52] and after some modifications, we obtain

$$\begin{aligned}
u_{I,ml} = & \left[x_{s_1,l} \sigma_{s_1} \sqrt{1 + \lambda_s^2 \varepsilon_s + 2\lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l}} + x_{s_2,l} \sigma_{s_2} \sqrt{1 - \lambda_s^2} \right] \delta(f_m - f_1) \\
& + q_{ml} x_{\mu_1,l} \sigma_{\mu_1} \sqrt{1 + \lambda_\mu^2 \varepsilon_\mu + 2\lambda_\mu \sqrt{\varepsilon_\mu} \cos \theta_{\mu,l}} + q_{ml} x_{\mu_2,l} \sigma_{\mu_2} \sqrt{1 - \lambda_\mu^2} + n_{I,l} \quad (3.52)
\end{aligned}$$

and

$$u_{Q,ml} = - \left[y_{s_1,l} \sigma_{s_1} \sqrt{1 + \lambda_s^2 \varepsilon_s + 2 \lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l}} + y_{s_2,l} \sigma_{s_2} \sqrt{1 - \lambda_s^2} \right] \delta(f_m - f_1) \\ - q_{ml} y_{\mu_1,l} \sigma_{\mu_1} \sqrt{1 + \lambda_\mu^2 \varepsilon_\mu + 2 \lambda_\mu \sqrt{\varepsilon_\mu} \cos \theta_{\mu,l}} - q_{ml} y_{\mu_2,l} \sigma_{\mu_2} \sqrt{1 - \lambda_\mu^2} - n_{Q,l} \quad (3.53)$$

where

$$r_{ml} = r_{I,ml}^2 + r_{Q,ml}^2 = u_{I,ml}^2 + u_{Q,ml}^2,$$

$$\theta_{s,l} = \theta_{s_1,l} - \theta_{s_2,l} + 2\pi f_m \tau_s,$$

and

$$\theta_{\mu,l} = \theta_{\mu_1,l} - \theta_{\mu_2,l} + 2\pi f_\chi \tau_\mu.$$

The random variables $u_{I,ml}$ and $u_{Q,ml}$ are Gaussian distributed with zero mean and a common variance given by

$$\sigma_{ml}^2 = (\sigma_{s_1}^2 + 2\sigma_{s_1}^2 \lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l} + \sigma_{s_2}^2) \delta(f_m - f_1) \\ + q_{ml} (\sigma_{\mu_1}^2 + 2\sigma_{\mu_1}^2 \lambda_\mu \sqrt{\varepsilon_\mu} \cos \theta_{\mu,l} + \sigma_{\mu_2}^2) + \frac{q_l \sigma_P^2}{2} + \frac{\sigma_\omega^2}{2}. \quad (3.54)$$

Hence, the square-law detector output r_{ml} is a central chi-squared random variable with two degrees of freedom [74]. Its corresponding pdf expression is given by

$$p_{r_{ml}}(r_{ml} | q_{ml}, q_l, f_m, \cos \theta_{s,l}, \cos \theta_{\mu,l}) = \frac{1}{2\sigma_{ml}^2} \exp \left(-\frac{r_{ml}}{2\sigma_{ml}^2} \right) U(r_{ml}). \quad (3.55)$$

From (3.55), the conditional pdf expressions of the square-law detector outputs r_{ml} ,

when the signal is present or not present, respectively, can be expressed as

$$p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \cos\theta_{s,l}, \cos\theta_{\mu,l}, f_m = f_1) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,S}^2}\right) U(r_{ml}) \quad (3.56)$$

and

$$p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \cos\theta_{\mu,l}, f_m \neq f_1) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,NS}^2}\right) U(r_{ml}). \quad (3.57)$$

Note that

$$\sigma_{ml,S}^2 = \sigma_{s_1}^2 + 2\sigma_{s_1}^2 \lambda_s \sqrt{\varepsilon_s} \cos\theta_{s,l} + \sigma_{s_2}^2 + q_{ml}(\sigma_{\mu_1}^2 + 2\sigma_{\mu_1}^2 \lambda_\mu \sqrt{\varepsilon_\mu} \cos\theta_{\mu,l} + \sigma_{\mu_2}^2) + \frac{q_l \sigma_P^2}{2} + \frac{\sigma_\omega^2}{2} \quad (3.58)$$

and

$$\sigma_{ml,NS}^2 = q_{ml}(\sigma_{\mu_1}^2 + 2\sigma_{\mu_1}^2 \lambda_\mu \sqrt{\varepsilon_\mu} \cos\theta_{\mu,l} + \sigma_{\mu_2}^2) + \frac{q_l \sigma_P^2}{2} + \frac{\sigma_\omega^2}{2}. \quad (3.59)$$

From (3.49), we have

$$z_{ml} = \ln\left(\frac{p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \theta, f_m = f_1)}{p_{r_{ml}}(r_{ml}|q_{ml}, q_l, \theta, f_m \neq f_1)}\right) = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2}\right) r_{ml} + \ln\left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2}\right). \quad (3.60)$$

Let $\alpha_{ml} = \frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2}$ and $\beta_{ml} = \ln\left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2}\right)$, (3.60) can be rewritten as

$$z_{ml} = \alpha_{ml} r_{ml} + \beta_{ml}. \quad (3.61)$$

The conditional pdf of the ML receiver output z_{ml} can be derived as [74]

$$p_{z_{ml}}(z_{ml}|q_{ml}, q_l, f_m, \cos\theta_{s,l}, \cos\theta_{\mu,l}) = \frac{1}{2\alpha_{ml}\sigma_{ml}^2} \exp\left(-\frac{z_{ml} - \beta_{ml}}{2\alpha_{ml}\sigma_{ml}^2}\right) U(z_{ml} - \beta_{ml}). \quad (3.62)$$

The corresponding conditional characteristic function of z_{ml} can be obtained as [74]

$$\psi_{z_{ml}}(jv|q_{ml}, q_l, f_m, \cos\theta_{s,l}, \cos\theta_{\mu,l}) = \frac{\exp(jv\beta_{ml})}{1 - j2v\alpha_{ml}\sigma_{ml}^2}. \quad (3.63)$$

We use the following approximation based on Taylor series expansion in central differences to obtain the characteristic function of z_{ml} , unconditioned on parameters $\theta_{s,l}$ and $\theta_{\mu,l}$ [77]

$$\psi_{z_{ml}}(jv|q_{ml}, q_l, f_m, \cos\theta_{\mu,l}) \approx \frac{1}{3} \sum_{k=-1}^1 \psi_{z_{ml}}\left(jv/q_{ml}, q_l, f_m, \cos\theta_{s,l} = \frac{\sqrt{3}k}{2}, \cos\theta_{\mu,l}\right) \quad (3.64)$$

and

$$\psi_{z_{ml}}(jv|q_{ml}, q_l, f_m) \approx \frac{1}{3} \sum_{b=-1}^1 \psi_{z_{ml}}\left(jv|q_{ml}, q_l, f_m, \cos\theta_{\mu,l} = \frac{\sqrt{3}b}{2}\right). \quad (3.65)$$

The characteristic function of the final decision variable y_m is given by

$$\psi_{y_m}(jv|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_m) = \prod_{l=1}^L \psi_{z_{ml}}(jv|q_{ml}, q_l, f_m), \quad (3.66)$$

where $\mathbf{K}_\mu$ and $\mathbf{K}_\mathbf{P}$ are the MTI and PBNI interference distribution patterns, as defined in (3.37) and (3.38), respectively. The corresponding pdf of the final decision variable y_m can be evaluated by performing the inverse Fourier transform [74]

$$p_{y_m}(y_m|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_m) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \psi_{y_m}(jv|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_m) e^{-jvy_m} dv. \quad (3.67)$$

Following that, the average BER can be computed using the expressions as shown in (3.39)-(3.41). For this case, the conditional symbol error probability $P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}}$ used in

Figure 3.6. Worst-case analytical and simulated BER results of the FFH/MFSK ML receiver with $M = 4$, $\nu_\mu = \nu_P = 0.5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_s = \rho_\mu = 0.7$ and various L .

(3.40) can be expressed as

$$\begin{aligned}
 P_{s|\mathbf{K}_\mu, \mathbf{K}_P} &= 1 - \int_{-\infty}^{\infty} \left[\int_{-\infty}^{y_1} p_{y_2}(y_2|\mathbf{K}_\mu, \mathbf{K}_P) dy_2 \times \cdots \times \int_{-\infty}^{y_1} p_{y_M}(y_M|\mathbf{K}_\mu, \mathbf{K}_P) dy_M \right] \\
 &\quad \times p_{y_1}(y_1|\mathbf{K}_\mu, \mathbf{K}_P) dy_1
 \end{aligned} \tag{3.68}$$

Figure 3.7. Worst-case BER performance of the FFH/MFSK ML receiver with $M = 4$, $L = 5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_s = \rho_\mu = 0.7$ and various ν_μ and ν_P .

3.3.2 Numerical Results and Discussions

In this section, the worst-case BER performance of the proposed ML receiver under the composite effect of MTI and PBNI over frequency-selective Rayleigh channels is provided. The SNR is fixed at $E_b/N_0 = 13.35$ dB, which corresponds to a BER of 10^{-5} for an ideal BFSK system under nonfading and unjammed conditions [31]. Figure 3.6 shows both the analytical and simulated BER curves of the proposed ML receiver with

Figure 3.8. Worst-case BER performance of the FFH/MFSK ML receiver with $M = 4$, $L = 5$, $\nu_\mu = \nu_P = 0.5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_\mu = 0.7$ and various ρ_s .

$M = 4$, $\nu_\mu = \nu_P = 0.5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_s = \rho_\mu = 0.7$ and diversity levels of $L = 1, 3, 5, 7, 19$ and 31 . The close match in both analytical and simulation results validates the analytical expressions derived in Section 3.3.1. Furthermore, we observe from Fig. 3.6 that there is an optimum diversity level for the proposed ML receiver, i.e., $L = 19$ at the SNR of $E_b/N_0 = 13.35$ dB. This is because although the FFH system with higher diversity orders L can effectively counteract interference and fading, increasing L also leads to the noncoherent combining loss [74].

In Fig. 3.7, the worst-case system performance of the proposed ML receiver with $M = 4$, $L = 5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_s = \rho_\mu = 0.7$ is provided for various interference conditions ($\nu_\mu = 0, 0.5, 0.75$ and 1). We observe that the BER curve of the proposed ML receiver under the composite effect lies between the BER curves under one type of interference only. When the power ratio of MTI with respect to the total interference power ν_μ is increased, the system performance improves significantly. This shows that the proposed system can suppress the MTI more effectively as compared to that of the PBNI. The worst-case BER results of the proposed ML receiver for various cross-correlation coefficients of the desired signal ρ_s are presented in Fig. 3.8. We observe that the system with a lower ρ_s gives better BER performance. However, it should be noted that the difference is not significant for strong-to-moderate interference condition ($0 < \text{SIR} < 15$ dB), which is of practical interest. Thus, we can conclude that the performance of the proposed system under the composite effect of PBNI and MTI is not very sensitive to the cross-correlation coefficient of the desired signal.

In Fig. 3.9, we compare the BER performance of the proposed system for various combinations of the power ratios of the two main correlated paths of the desired signal ε_s and of the MTI ε_μ . These results show that the system performance is more sensitive to the parameter ε_s than the parameter ε_μ . It is observed that the system performance degrades as ε_s increases. However, the difference in system performance is not significant, i.e., only up to half an order of magnitude. In Fig. 3.10, we fix the cross-correlation coefficient of the desired signal of $\rho_s = 0.5$ and compare the BER performance of the proposed ML receiver for different values of ρ_μ . It can be observed that the BER curves are almost the same. This shows that the correlation characteristic on fading of the MTI does not affect the system performance of the proposed system.

Figure 3.11 shows the system performance comparisons of the proposed ML receiver with some conventional diversity-combining receivers, namely, the linear-combining and

Figure 3.9. Worst-case BER performance of FFH/MFSK ML receiver with $M = 4$, $L = 3$, $\nu_\mu = \nu_P = 0.5$, $\rho_s = \rho_\mu = 0.7$ and various ϵ_s and ϵ_μ .

product-combining receivers. The worst-case BER results of these two conventional receivers are obtained via simulation. The system parameters are set to be $M = 8$, $L = 3$, $\nu_\mu = \nu_P = 0.5$, $\epsilon_s = \epsilon_\mu = 1$, and $\rho_s = \rho_\mu = 0.5$. From this figure, we observe that for the practical SIR range of interest ($0 < \text{SIR} < 20$ dB), the LCR is outperformed by the PCR. Furthermore, for all SIR levels, the proposed ML receiver gives the best BER performance under the composite interference effect, and the improvement by up to one order of magnitude in BER performance can be observed.

Figure 3.10. Worst-case BER performance of FFH/MFSK ML receiver with $M = 8$, $L = 3$, $\nu_\mu = \nu_P = 0.5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_s = 0.5$ and various ρ_μ .

3.4 Conclusion

In this chapter, the analytical BER expressions have been derived for FFH/MFSK ML receivers under the composite effect of MTI and PBNI over either frequency-nonselective or frequency-selective fading channels. These expressions have been validated by the simulation results. Also, the corresponding ML structures have been proposed and analyzed. Under different types of fading conditions, it has been shown that the proposed

Figure 3.11. Worst-case BER performance comparisons among various receivers with $M = 8$, $L = 3$, $\nu_\mu = \nu_P = 0.5$, $\varepsilon_s = \varepsilon_\mu = 1$, $\rho_s = \rho_\mu = 0.5$.

ML receivers can effectively suppress the composite effect of MTI and PBNI. Although the proposed ML receivers has been shown to significantly outperform the LCR and the PCR, they require side information to implement and this increases their complexity. In the next chapter, we will investigate the impact of both timing and frequency offsets on FFH systems under the composite interference effect, and derive the corresponding ML receiver structures.

Chapter 4

Impact of Timing and Frequency Offsets plus Composite Interference on FFH/MFSK Systems over Fading Channels

4.1 Overview

In this chapter, we study the composite effects of MTI and PBNI on an FFH/MFSK system for a more practical scenario where both timing and frequency offsets are present. In practical systems, it is well-known that there always exists time and frequency mismatch between the transmitter and the receiver. Analytical BER expressions for the ML receivers, as opposed to conventional diversity-combining receivers, are derived. The practical issues of timing and frequency offsets are considered. In addition, the sensitivity of system performance to imperfect side information is also examined. The analysis of FFH systems over frequency-nonselective Rayleigh fading channels is first

performed, followed by the frequency-selective Rician fading.

4.2 Performance Study over Frequency-Nonselective Fading Channels

4.2.1 BER Performance Analysis

In this analysis, it is assumed that the signal is transmitted through a frequency-nonselective Rayleigh fading channel and the received signal is contaminated by MTI, PBNI and AWGN simultaneously. The power ratios of the PBNI and the MTI with respect to the total interference power are defined as $\nu_P = P_P/P_J$ and $\nu_\mu = 1 - \nu_P = P_\mu/P_J$, respectively. In FH communication systems, imperfect time and frequency synchronization could significantly degrade BER performance. In our analysis, we consider the timing and frequency offsets for the MTI and the desired signal. For the PBNI, only timing offset is considered as PBNI is the noise interference. It is assumed that both interference and signal experience the same timing and frequency offsets. The offsets of the time epoch and received carrier frequency are denoted as ΔT and Δf , respectively. It is also assumed that the desired signal and the MTI undergo independent Rayleigh fading, as shown in Fig. 2.3. The dehopped signal of the l th hop in the

interval $[(l-1)T_h + \Delta T, lT_h + \Delta T]$ can be expressed as [84]

$$r_l(t) = \begin{cases} \sqrt{2}|a_{S,l}|\cos[2\pi(f_n + \Delta f)t + \phi_{S,l}] + q_l J_P(t) + \omega(t) \\ \quad + q_{\mu_1 l} \sqrt{2}|a_{\mu_1,l}|\cos[2\pi(f_{\mu_1} + \Delta f)t + \phi_{\mu_1,l}], \\ \quad \text{for } (l-1)T_h + \Delta T \leq t \leq lT_h \\ g_{\mu_2 l} \sqrt{2}|a_{\mu_2,l}|\cos[2\pi(f_{\mu_2} + \Delta f)t + \phi_{\mu_2,l}] \\ \quad + q'_l J_P(t) + \omega(t), \\ \quad \text{for } lT_h \leq t \leq lT_h + \Delta T \end{cases}$$

for $n = 1, 2, \dots, M$, (4.1)

where $\sqrt{2}|a_{S,l}|\cos[2\pi(f_n + \Delta f)t + \phi_{S,l}]$ denotes the desired signal, $\sqrt{2}|a_{\mu_1,l}|\cos[2\pi(f_{\mu_1} + \Delta f)t + \phi_{\mu_1,l}]$ and $\sqrt{2}|a_{\mu_2,l}|\cos[2\pi(f_{\mu_2} + \Delta f)t + \phi_{\mu_2,l}]$ denote the MTI tones in the time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$, respectively. Note that $J_P(t)$ represents the PBNI, and $\omega(t)$ is the noise term due to AWGN.

In (4.1), the terms $q_{\mu_1 l}$ and $g_{\mu_2 l}$ (either 1 or 0) indicate whether the l th hop is interfered by the MTI or not in the time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$, respectively. Similarly, the terms q_l and q'_l (either 1 or 0) denote the PBNI states in the time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$, respectively. The terms $\sqrt{2}|a_{S,l}|$, f_n and $\phi_{S,l}$ represent faded amplitude, baseband frequency and random phase of the desired signal, respectively. Similar notations for $(\sqrt{2}|a_{\mu_1,l}|, f_{\mu_1}, \phi_{\mu_1,l})$ and $(\sqrt{2}|a_{\mu_2,l}|, f_{\mu_2}, \phi_{\mu_2,l})$ are adopted for the MTI in time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$, respectively. The random variables $\phi_{S,l}$, $\phi_{\mu_1,l}$, $\phi_{\mu_2,l}$ are assumed to be uniformly distributed over $[0, 2\pi]$. We also assume that the average power of the desired user signal $P_S = 2\sigma_S^2$ and the average power of the MTI $P_\mu = 2\sigma_\mu^2$ remain constant from hop to hop. Note that $2\sigma_S^2$ and $2\sigma_\mu^2$ are the diffused power levels of the desired signal and MTI, respectively.

In order to derive the pdf of the square-law detector output r_{ml} , we express the faded

amplitudes $|a_{S,l}|$, $|a_{\mu_1,l}|$ and $|a_{\mu_2,l}|$ as $|a_{S,l}| = \sigma_S |x_{S,l} + jy_{S,l}|$, $|a_{\mu_1,l}| = \sigma_\mu |x_{\mu_1,l} + jy_{\mu_1,l}|$ and $|a_{\mu_2,l}| = \sigma_\mu |x_{\mu_2,l} + jy_{\mu_2,l}|$, respectively. Note that $x_{S,l}$, $y_{S,l}$, $x_{\mu_1,l}$, $y_{\mu_1,l}$, $x_{\mu_2,l}$ and $y_{\mu_2,l}$ are independent Gaussian random variables with zero mean and unit variance. The square-law detector output r_{ml} including the composite effect of MTI, PBNI and thermal noise is given in (3.7). Note that due to the timing offset, the in-phase and quadrature-phase components $r_{I,ml}$ and $r_{Q,ml}$ are given by

$$r_{I,ml} = \frac{\sqrt{2}}{T_h} \int_{(l-1)T_h+\Delta T}^{lT_h+\Delta T} r_l(t) \cos(2\pi f_m t) dt \quad (4.2)$$

and

$$r_{Q,ml} = \frac{\sqrt{2}}{T_h} \int_{(l-1)T_h+\Delta T}^{lT_h+\Delta T} r_l(t) \sin(2\pi f_m t) dt, \quad (4.3)$$

respectively. After some mathematical manipulations, the expressions of the in-phase and quadrature-phase components can be obtained as

$$\begin{aligned} r_{I,ml} = & x_{S,l} \sigma_S (1 - \chi) \text{sinc}\{(1 - \chi) [(f_n - f_m)T_h + \xi]\} \\ & + q_{\mu_1 l} x_{\mu_1,l} \sigma_\mu (1 - \chi) \text{sinc}\{(1 - \chi) [(f_{\mu_1} - f_m)T_h + \xi]\} \\ & + g_{\mu_2 l} x_{\mu_2,l} \sigma_\mu \chi \text{sinc}\{\chi [(f_{\mu_2} - f_m)T_h + \xi]\} + n_{I,l} \end{aligned} \quad (4.4)$$

and

$$\begin{aligned} r_{Q,ml} = & -y_{S,l} \sigma_S (1 - \chi) \text{sinc}\{(1 - \chi) [(f_n - f_m)T_h + \xi]\} \\ & - q_{\mu_1 l} y_{\mu_1,l} \sigma_\mu (1 - \chi) \text{sinc}\{(1 - \chi) [(f_{\mu_1} - f_m)T_h + \xi]\} \\ & - g_{\mu_2 l} y_{\mu_2,l} \sigma_\mu \chi \text{sinc}\{\chi [(f_{\mu_2} - f_m)T_h + \xi]\} - n_{Q,l}, \end{aligned} \quad (4.5)$$

respectively. In (4.4) and (4.5), $n_{I,l}$ and $n_{Q,l}$ represent independent zero-mean Gaussian random variables with variance $[q_l (1 - \chi) + q'_l \chi]^2 \sigma_P^2 / 2 + \sigma_\omega^2 / 2$, where σ_ω^2 is the variance

of AWGN. Note that $\chi = \Delta T/T_h$ and $\xi = \Delta f \times T_h$ are the normalized timing and frequency offsets [85]. The derived in-phase and quadrature-phase components in (4.4) and (4.5) are Gaussian random variables with zero mean and a common variance given by

$$\begin{aligned}\sigma_{ml}^2 &= \sigma_S^2(1-\chi)^2 \text{sinc}^2\{(1-\chi)[(f_n - f_m)T_h + \xi]\} \\ &+ q_{\mu_1 l} \sigma_\mu^2(1-\chi)^2 \text{sinc}^2\{(1-\chi)[(f_{\mu_1} - f_m)T_h + \xi]\} \\ &+ g_{\mu_2 l} \sigma_\mu^2 \chi^2 \text{sinc}^2\{\chi[(f_{\mu_2} - f_m)T_h + \xi]\} + \frac{[q_l(1-\chi) + q'_l \chi]^2}{2} \sigma_P^2 + \frac{\sigma_\omega^2}{2}.\end{aligned}\quad (4.6)$$

According to [74, p. 41-42], the distribution of square-law detector output r_{ml} is central chi-squared with two degrees of freedom. The corresponding conditional pdf expression of r_{ml} can be expressed as

$$p_{r_{ml}}(r_{ml}|f_n, q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml}^2}\right) U(r_{ml}). \quad (4.7)$$

The resultant output of the ML receiver can be expressed as [53]

$$z_{ml} = \ln\left(\frac{p_{r_{ml}}(r_{ml}|q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l, S)}{p_{r_{ml}}(r_{ml}|q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l, NS)}\right), \quad (4.8)$$

where S represents the signal hop and NS denotes the non-signal hop. From (4.7), we have

$$p_{r_{ml}}(r_{ml}|q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l, S) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,S}^2}\right) U(r_{ml}) \quad (4.9)$$

and

$$p_{r_{ml}}(r_{ml}|q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l, NS) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,NS}^2}\right) U(r_{ml}), \quad (4.10)$$

where

$$\begin{aligned}\sigma_{ml,S}^2 &= \sigma_S^2(1-\chi)^2 \text{sinc}^2[(1-\chi)\xi] \\ &+ q_{\mu_1 l} \sigma_\mu^2 (1-\chi)^2 \text{sinc}^2\{(1-\chi)[(f_{\mu_1} - f_m)T_h + \xi]\} \\ &+ g_{\mu_2 l} \sigma_\mu^2 \chi^2 \text{sinc}^2\{\chi[(f_{\mu_2} - f_m)T_h + \xi]\} + \frac{[q_l(1-\chi) + q'_l \chi]^2}{2} \sigma_P^2 + \frac{\sigma_\omega^2}{2}\end{aligned}\quad (4.11)$$

and

$$\begin{aligned}\sigma_{ml,NS}^2 &= q_{\mu_1 l} \sigma_\mu^2 (1-\chi)^2 \text{sinc}^2\{(1-\chi)[(f_{\mu_1} - f_m)T_h + \xi]\} \\ &+ g_{\mu_2 l} \sigma_\mu^2 \chi^2 \text{sinc}^2\{\chi[(f_{\mu_2} - f_m)T_h + \xi]\} + \frac{[q_l(1-\chi) + q'_l \chi]^2}{2} \sigma_P^2 + \frac{\sigma_\omega^2}{2}.\end{aligned}\quad (4.12)$$

Following that, by substituting (4.9) and (4.10) into (4.8), the ML receiver output z_{ml} can be obtained as

$$z_{ml} = \alpha_{ml} r_{ml} + \beta_{ml}, \quad (4.13)$$

where

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

and

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

The corresponding pdf expressions of z_{ml} can be derived as

$$p_{z_{ml}}(z_{ml} | f_n, q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l) = \frac{1}{2\alpha_{ml}\sigma_{ml}^2} \exp\left(-\frac{z_{ml} - \beta_{ml}}{2\alpha_{ml}\sigma_{ml}^2}\right) U(z_{ml} - \beta_{ml}). \quad (4.16)$$

After some algebraic manipulations, the CF of z_{ml} can be shown to be

$$\psi_{z_{ml}}(jv | f_n, q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l) = \frac{\exp(jv\beta_{ml})}{1 - j2v\alpha_{ml}\sigma_{ml}^2}. \quad (4.17)$$

Following that, the CF of the final decision statistics $y_m = \sum_{l=1}^L z_{ml}$ is given by

$$\psi_{y_m}(jv|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n) = \prod_{l=1}^L \psi_{z_{ml}}(jv|f_n, q_{\mu_1 l}, g_{\mu_2 l}, q_l, q'_l), \quad (4.18)$$

where $\mathbf{K}_\mathbf{P}$ and $\mathbf{K}_\mu$ are the PBNI and MTI distribution patterns, respectively, given by

$$\mathbf{K}_\mathbf{P} = \begin{pmatrix} q_1 & q_2 & \dots & q_L \\ q'_1 & q'_2 & \dots & q'_L \end{pmatrix},$$

and

$$\mathbf{K}_\mu = \begin{pmatrix} q_{11} & q_{12} & \dots & q_{1L} & g_{11} & g_{12} & \dots & g_{1L} \\ q_{21} & q_{22} & \dots & q_{2L} & g_{21} & g_{22} & \dots & g_{2L} \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\ q_{M1} & q_{M2} & \dots & q_{ML} & g_{M1} & g_{M2} & \dots & g_{ML} \end{pmatrix}.$$

The pdf of the final decision statistic y_m can be obtained by performing the inverse Fourier transform

$$p_{y_m}(y_m|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \psi_{y_m}(jv|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n) e^{-jvy_m} dv. \quad (4.19)$$

Following that, the average BER can be obtained as

$$P_b = \frac{1}{2(M-1)} \sum_{n=1}^M P_{s|f_n}, \quad (4.20)$$

where $P_{s|f_n}$ is the conditional probability of symbol error, conditioned on f_n . To obtain $P_{s|f_n}$, we average the conditional probability of symbol error $P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n}$ over all the possible combinations of $\mathbf{K}_\mu$ and $\mathbf{K}_\mathbf{P}$, that is

$$P_{s|f_n} = \sum_{\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}} P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}) P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n}, \quad (4.21)$$

Figure 4.1. Proposed FFH/MFSK ML receiver structure.

where $P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P})$ represents the probability of both MTI and PBNI interference pattern, given by [84]

$$\begin{aligned}
P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}) &= \left(\frac{Q}{MN} \right)^{\sum_{l=1}^L \sum_{m=1}^M q_{ml}} \left(1 - \frac{Q}{N} \right)^{L - \sum_{l=1}^L \sum_{m=1}^M q_{ml}} \\
&\times \prod_{l=1}^L \left[\left(\frac{Q - q_{m(l+1)}}{M(N-1)} \right)^{\sum_{m=1}^M g_{ml}} \left(1 - \frac{Q - q_{m(l+1)}}{N-1} \right)^{1 - \sum_{m=1}^M g_{ml}} \right] \\
&\times \gamma^{\sum_{l=1}^L q_l} (1 - \gamma)^{L - \sum_{l=1}^L q_l} \prod_{l=1}^L \left[\left(\frac{\gamma - q_{l+1}/N}{1 - 1/N} \right)^{q'_l} \left(1 - \frac{\gamma - q_{l+1}/N}{1 - 1/N} \right)^{1 - q'_l} \right]. \quad (4.22)
\end{aligned}$$

Note that $P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n}$ denotes the conditional symbol error probability, conditioned on

the interference pattern described by matrices $\mathbf{K}_\mu$ and $\mathbf{K}_P$, which can be obtained by

$$\begin{aligned}
P_{s|\mathbf{K}_\mu, \mathbf{K}_P, f_n} &= 1 - \int_{-\infty}^{\infty} \left[\int_{-\infty}^{y_n} p_{y_1}(y_1|\mathbf{K}_\mu, \mathbf{K}_P, f_n) dy_1 \times \dots \right. \\
&\times \int_{-\infty}^{y_n} p_{y_{n-1}}(y_{n-1}|\mathbf{K}_\mu, \mathbf{K}_P, f_n) dy_{n-1} \int_{-\infty}^{y_n} p_{y_{n+1}}(y_{n+1}|\mathbf{K}_\mu, \mathbf{K}_P, f_n) dy_{n+1} \times \dots \\
&\times \left. \int_{-\infty}^{y_n} p_{y_M}(y_M|\mathbf{K}_\mu, \mathbf{K}_P, f_n) dy_M \right] p_{y_n}(y_n|\mathbf{K}_\mu, \mathbf{K}_P, f_n) dy_n. \tag{4.23}
\end{aligned}$$

The block diagram of the proposed ML combining block bounded by dashed line in Fig. 2.3 is depicted in Fig. 4.1. This diagram is obtained from (4.13). The proposed ML receiver structure derived in (4.13) is simple, but it requires side information on the noise variance, PBNI variance, power level of the MTI, as well as timing and frequency offsets. The frequency offset can be obtained from the nondata-aided carrier frequency estimator [86], whereas the timing offset can be estimated with the aid of the ML epoch-estimation scheme presented in [87], which is based on the multiple-hop combining of single-hop autocorrelation (SHAC) power sums. The power level of MTI, and the PBNI variance can be extracted from an energy detector [64], whereas the noise variance can be obtained from a “noise-only” channel for noise power measurement [37].

4.2.2 Numerical Results and Discussions

In this section, we present the worst-case BER results of the proposed ML receiver under the composite effect of MTI and PBNI with the presence of timing and frequency offsets. We also compare BER performance of the ML receiver with that of the conventional diversity-combining receivers, namely, the linear-combining, product-combining, self-normalizing and OS-NED receivers.

In Fig. 4.2, both analytical and simulated BER results of the proposed ML receiver over Rayleigh fading channels are provided when both the timing and frequency offsets are present. The power ratios of the MTI and PBNI are set to be $\nu_\mu = 0.5$ and $\nu_P = 0.5$,

Figure 4.2. Analytical and simulated worst-case BER results of the ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 16$ dB, and $\nu_\mu = \nu_P = 0.5$.

respectively. The close match in both results validates the analytical BER expressions derived in Section 4.2.1. This figure also shows sensitivity of BER performance to the combined effect of timing and frequency offsets when the desired signal is corrupted by both MTI and PBNI. It can be observed that this combined effect significantly degrades the system performance, i.e., up to one order of magnitude. In this figure, we also include simulated BER results of the linear-combining receiver with $\chi = \xi = 0$ to show the effectiveness of the proposed ML receiver in suppressing the composite

Figure 4.3. BER performance of the ML receiver with $M = 4$, $\text{SNR} = 13.35$ dB, and $\nu_\mu = \nu_P = 0.5$ with different diversity orders L .

interference effect. Comparing BER curves of the ML receiver with $\chi = \xi = 0.1$ and that of the LCR with $\chi = \xi = 0$, we can observe that the former with timing and frequency offsets of $\chi = \xi = 0.1$ is even better than the latter without these two offsets for moderate-to-high interference power regions ($0 \leq \text{SIR} \leq 25$ dB), which are of practical interest. When the combined effect of timing and frequency offsets becomes more severe ($\chi = 0.1, \xi = 0.2$ or $\chi = \xi = 0.2$), performance of the ML receiver is still slightly better than that of the LCR for strong interference condition.

Figure 4.4. Worst-case BER results of the ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 16$ dB, and $\nu_\mu = \nu_P = 0.5$.

Figure 4.3 shows the benefits of using diversity for the FFH/MFSK ML receiver with the presence of composite interference effect ($\nu_\mu = \nu_P = 0.5$) when timing and frequency offsets are present ($\chi = \xi = 0.1$) and absent ($\chi = \xi = 0$). When the diversity level L is increased, FFH systems can more effectively countermeasure the interference and fading effects. However, at the same time, system performance degrades due to the noncoherent diversity-combining loss [74]. Therefore, there exists an optimum diversity order for the proposed ML receiver. When $\chi = \xi = 0.1$, this optimum diversity order

Figure 4.5. Worst-case performance comparisons among various receivers with $M = 4$, $L = 3$, $\text{SNR} = 20$ dB, with different interference conditions.

is observed to be $L = 11$. It can also be observed that there exists an error floor for large SIR level for each BER curve. For the case of $L = 5$ and $\chi = \xi = 0.1$, this error floor is observed to be 3.5×10^{-3} . This phenomenon is due to the fact that when SIR is large, the effect of the interference (MTI or PBNI or both) on the system performance is negligible and the error is mainly caused by Gaussian noise. Figure 4.4 demonstrates whether performance of the proposed ML receiver is sensitive to either the timing or frequency offset. In general, increasing the timing or frequency offset degrades the

Figure 4.6. Worst-case BER performance comparisons among various diversity-combining receivers with $\text{SNR} = 20$ dB, $M = 4$, $L = 3$, and $\chi = \xi = 0.1$, with different interference conditions.

system performance. This is because these two offsets cause the loss of orthogonality. In addition, the second source of degradation in system performance due to timing offset is that the imperfect timing synchronization leads to the signal attenuation in the detector matched to the incoming frequency [85]. It can also be observed that when the normalized offset (χ and ξ) is less than 0.3, the effect of timing offset on the BER performance is more severe than that of frequency offset. However, the opposite observation is true when the normalized offset is greater than 0.3.

In Fig. 4.5, we compare BER performance of the ML receiver with that of the

Figure 4.7. Worst-case BER results of the ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 16$ dB, $\nu_\mu = 1$, and $\nu_P = 0$.

linear-combining and product-combining receivers under different interference conditions ($\nu_\mu = 0, 0.5$ and 1). Both the timing and frequency offsets are present for the case of $\chi = \xi = 0.1$. It should be noted that BER results of the two conventional diversity-combining receivers are obtained via simulation. When the interference power is high, it can be observed that the BER curve under the composite interference effect lies between the BER curves under one type of interference only. Moreover, BER performance of the ML receiver with $\nu_\mu = 1$ is obviously better than that with $\nu_\mu = 0$. Thus, it can

Figure 4.8. Worst-case BER results of the ML receiver with $L = 5$, $\text{SNR} = 16$ dB, $\nu_\mu = 1$, $\nu_P = 0$, and $\chi = \xi = 0.1$, with different modulation orders M .

be concluded that the proposed ML receiver is more effective in mitigating MTI than PBNI. As expected, we can observe from Fig. 4.5 that the proposed ML receiver always outperforms the conventional receivers for all interference conditions. The improvement in system performance achieved by using this ML scheme can be as large as two orders of magnitude when the desired signal is corrupted by MTI only. Hence, it is interesting to examine the BER performance of the ML receiver with $\nu_\mu = 1$. In Fig. 4.6, we compare the worst-case BER performance of the proposed ML receiver with that of the self-normalizing and OS-NED receivers with the presence of both timing and frequency

Figure 4.9. Worst-case BER results of the ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 16$ dB, $\nu_\mu = 1$, $\nu_P = 0$, and imperfect side information on parameters χ , ξ .

offsets. The results for OS-NED and self-normalizing receivers are obtained via simulations. It can also be observed that the ML receiver also outperforms the OS-NED and self-normalizing receivers.

In Fig. 4.7, we investigate the combined effect of timing and frequency offsets on BER performance of the proposed ML receiver when the desired signal is contaminated by MTI only. The system parameters are set to be $\text{SNR} = 16$ dB, $M = 4$, $L = 3$, $\nu_\mu = 1$, $\nu_P = 0$. We can observe that the BER curves are almost flat throughout the

range of SIR levels for all offset values. It can be concluded that the ML receiver can almost completely suppress the MTI regardless of whether offsets are present or not. The BER results of the proposed ML receiver under MTI only with various modulation orders M ($M = 2, 4, 6, 8, 16$ and 32) are provided in Fig. 4.8 when both timing and frequency offsets are present. From this figure, we observe the proposed receiver can achieve remarkable improvement in system performance by increasing M . Lastly, we study the sensitivity of system performance to imperfect side information in Fig. 4.9. Numerical results with imperfect side information are obtained via simulation. From Fig. 4.9, it confirms that performance of the proposed ML receiver is robust against inaccuracy in estimation of the required side information.

4.3 Performance Study over Frequency-Selective Fading Channels

In a practical environment, most wireless channels are frequency selective [88], [89]. Besides that, for FH systems, the bandwidth of each frequency slot is comparable to the coherence bandwidth of the multipath channel [90], [91]. Hence, for FFH systems with large diversity order and high data rates, the received signal is subject to frequency-selective fading. The objective of this section is to study both the composite effect of PBNI and MTI, and the impact of the timing and frequency offsets on the performance of an FFH system employing MFSK modulation over frequency-selective Rician fading channels.

4.3.1 BER Performance Analysis

Assuming that the received signal is corrupted by PBNI, MTI and AWGN concurrently, the dehopped signal over the interval $[(l-1)T_h + \Delta T, lT_h + \Delta T]$ of the l th hop can be

expressed as [84]

$$r_l(t) = \begin{cases} s_l(t) + q_{\alpha l} J_\mu(t) + q_l J_P(t) + \omega(t), & \text{for } (l-1)T_h + \Delta T \leq t \leq lT_h \\ q'_{\beta l} J'_\mu(t) + q'_l J_P(t) + \omega(t), & \text{for } lT_h \leq t \leq lT_h + \Delta T \end{cases} \quad (4.24)$$

where $s_l(t)$ denotes the desired signal, and $J_\mu(t)$ and $J'_\mu(t)$ are the MTI tones in the time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$, respectively. The terms $q_{\alpha l}$ and $q'_{\beta l}$ (1 or 0) indicate whether the l th hop is interfered by the MTI or not in the time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$, respectively.

Based on the Rician fading model presented in [73], a frequency-selective Rician fading channel can be modeled as the sum of the dominant LOS component and the Rayleigh component. In addition, to simplify the derivations and the analyses, we can model the Rayleigh component as the correlated two-path channel model by summing up all possible non-line-of-sight (NLOS) paths and then splitting them into two main clumps for forming these two paths [52], [78], [79]. The two-path model is also widely adopted in the research works of [80]–[82] to model frequency-selective fading channels due to its analytical simplicity. Therefore, the term $s_l(t)$ in (4.24) can be expressed as

$$s_l(t) = \sqrt{2}A_S \cos[2\pi(f_n + \Delta f)t + \phi_{S,l}] + \sqrt{2}|a_{S_1,l}| \cos[2\pi(f_n + \Delta f)t + \theta_{S_1,l}] \\ + \sqrt{2}|a_{S_2,l}| \cos[2\pi(f_n + \Delta f)(t - \tau_S) + \theta_{S_2,l}], \text{ for } n = 1, 2, \dots, M, \quad (4.25)$$

where $\sqrt{2}A_S$ and $\phi_{S,l}$ denote the amplitude and phase of the LOS component of the desired signal, respectively. Note that the terms $J_P(t)$, $\omega(t)$, f_n , q_l and q'_l have the same definitions as those in Section 4.2.1. The terms $\sqrt{2}|a_{S_1,l}|$ and $\sqrt{2}|a_{S_2,l}|$ are the two main correlated Rayleigh amplitudes of the desired signal, whereas $\theta_{S_1,l}$ and $\theta_{S_2,l}$ represent their corresponding random phases. Note that τ_S represents the time delay between the two main correlated Rayleigh paths, which is assumed to be uniformly distributed

over $(0, T_h]$.

Assuming that the MTI also experiences frequency-selective Rician fading, the two MTI tones in the time intervals $[(l-1)T_h + \Delta T, lT_h]$ and $[lT_h, lT_h + \Delta T]$ can be modeled in the same way as that of the desired signal by replacing the parameters $(\sqrt{2}A_S, \phi_{S,l}, \sqrt{2}|a_{S1,l}|, \sqrt{2}|a_{S2,l}|, \theta_{S1,l}, \theta_{S2,l}, f_n, \tau_S)$ by $(\sqrt{2}A_\mu, \phi_{\mu,l}, \sqrt{2}|a_{\mu1,l}|, \sqrt{2}|a_{\mu2,l}|, \theta_{\mu1,l}, \theta_{\mu2,l}, f_\alpha, \tau_\mu)$ and $(A_\mu, \phi'_{\mu,l}, \sqrt{2}|a'_{\mu1,l}|, \sqrt{2}|a'_{\mu2,l}|, \theta'_{\mu1,l}, \theta'_{\mu2,l}, f_\beta, \tau_\mu)$, respectively. Note that τ_μ denotes the difference between the two-path delays of the desired signal and MTI. Therefore, their corresponding expressions can be shown as

$$\begin{aligned} J_\mu(t) = & \sqrt{2}A_\mu \cos[2\pi(f_\alpha + \Delta f)t + \phi_{\mu,l}] + \sqrt{2}|a_{\mu1,l}| \cos[2\pi(f_\alpha + \Delta f)t + \theta_{\mu1,l}] \\ & + \sqrt{2}|a_{\mu2,l}| \cos[2\pi(f_\alpha + \Delta f)(t - \tau_\mu) + \theta_{\mu2,l}], \text{ for } \alpha = 1, 2, \dots, M, \end{aligned} \quad (4.26)$$

and

$$\begin{aligned} J'_\mu(t) = & \sqrt{2}A_\mu \cos[2\pi(f_\beta + \Delta f)t + \phi'_{\mu,l}] + \sqrt{2}|a'_{\mu1,l}| \cos[2\pi(f_\beta + \Delta f)t + \theta'_{\mu1,l}] \\ & + \sqrt{2}|a'_{\mu2,l}| \cos[2\pi(f_\beta + \Delta f)(t - \tau_\mu) + \theta'_{\mu2,l}], \text{ for } \beta = 1, 2, \dots, M. \end{aligned} \quad (4.27)$$

It is also assumed that the random phases $\phi_{S,l}, \theta_{S1,l}, \theta_{S2,l}, \phi_{\mu,l}, \theta_{\mu1,l}, \theta_{\mu2,l}, \phi'_{\mu,l}, \theta'_{\mu1,l}$ and $\theta'_{\mu2,l}$ are uniformly distributed over $[0, 2\pi]$. According to [83], we can generate the correlated Rayleigh amplitudes $a_{S1,l}$ and $a_{S2,l}$ from the uncorrelated complex Gaussian-fading amplitudes $b_{S1,l} = x_{S1,l} + jy_{S1,l}$ and $b_{S2,l} = x_{S2,l} + jy_{S2,l}$; $a_{\mu1,l}$ and $a_{\mu2,l}$ from $b_{\mu1,l} = x_{\mu1,l} + jy_{\mu1,l}$ and $b_{\mu2,l} = x_{\mu2,l} + jy_{\mu2,l}$; $a'_{\mu1,l}$ and $a'_{\mu2,l}$ from $b'_{\mu1,l} = x'_{\mu1,l} + jy'_{\mu1,l}$ and $b'_{\mu2,l} = x'_{\mu2,l} + jy'_{\mu2,l}$. Note that $x_{S1,l}, y_{S1,l}, x_{S2,l}, y_{S2,l}, x_{\mu1,l}, y_{\mu1,l}, x_{\mu2,l}, y_{\mu2,l}, x'_{\mu1,l}, y'_{\mu1,l}, x'_{\mu2,l}$ and $y'_{\mu2,l}$ are independent Gaussian random variables with zero mean and unit variance. Thus, $|a_{S1,l}|, |a_{S2,l}|, |a_{\mu1,l}|, |a_{\mu2,l}|, |a'_{\mu1,l}|$ and $|a'_{\mu2,l}|$ can be expressed as [52]

$$|a_{S1,l}| = \sigma_{S1} |x_{S1,l} + jy_{S1,l}|, \quad (4.28)$$

$$|a_{S_2,l}| = \sigma_{S_2} \left[\lambda_S^2 |x_{S_1,l} + jy_{S_1,l}|^2 + (1 - \lambda_S^2) |x_{S_2,l} + jy_{S_2,l}|^2 + 2\lambda_S \sqrt{1 - \lambda_S^2} (x_{S_1,l}x_{S_2,l} + y_{S_1,l}y_{S_2,l}) \right]^{1/2}, \quad (4.29)$$

$$|a_{\mu_1,l}| = \sigma_{\mu_1} |x_{\mu_1,l} + jy_{\mu_1,l}|, \quad (4.30)$$

$$|a_{\mu_2,l}| = \sigma_{\mu_2} \left[\lambda_\mu^2 |x_{\mu_1,l} + jy_{\mu_1,l}|^2 + (1 - \lambda_\mu^2) |x_{\mu_2,l} + jy_{\mu_2,l}|^2 + 2\lambda_\mu \sqrt{1 - \lambda_\mu^2} (x_{\mu_1,l}x_{\mu_2,l} + y_{\mu_1,l}y_{\mu_2,l}) \right]^{1/2}, \quad (4.31)$$

$$|a'_{\mu_1,l}| = \sigma_{\mu_1} |x'_{\mu_1,l} + jy'_{\mu_1,l}|, \quad (4.32)$$

and

$$|a'_{\mu_2,l}| = \sigma_{\mu_2} \left[\lambda_\mu^2 |x'_{\mu_1,l} + jy'_{\mu_1,l}|^2 + (1 - \lambda_\mu^2) |x'_{\mu_2,l} + jy'_{\mu_2,l}|^2 + 2\lambda_\mu \sqrt{1 - \lambda_\mu^2} (x'_{\mu_1,l}x'_{\mu_2,l} + y'_{\mu_1,l}y'_{\mu_2,l}) \right]^{1/2}, \quad (4.33)$$

where $2\sigma_{S_1}^2$ and $2\sigma_{S_2}^2$ denote the diffused power levels of the two correlated paths of the desired signal; $2\sigma_{\mu_1}^2$ and $2\sigma_{\mu_2}^2$ are the diffused power levels of the first and the second correlated paths of the MTI; $\lambda_S \cong \sqrt{8 \left(\sqrt{(4/\pi - 1) \rho_s + 1} - 1 \right)}$ and $\lambda_\mu \cong$

$\sqrt{8 \left(\sqrt{(4/\pi - 1) \rho_\mu + 1} - 1 \right)}$ are the functions of the cross-correlation coefficients (ρ_S and ρ_μ) of the two main correlated paths of the desired signal and the MTI, respectively.

From the intentional interferer's viewpoint, the maximum BER value is described as "optimum" rather than "worst case". In addition, from the viewpoint of the interferer, the interfering strategy or the values of ν_μ and ν_P can be chosen to cause the maximum degradation to the system performance. Thus, it is also important to study the performance of FFH systems under the composite interference effect in order to determine the most effective interfering strategy from the intentional interferer's perspective. The Rician fading factors of the desired signal $F_S = A_S^2 / (2\sigma_{S_1}^2 + 2\sigma_{S_2}^2)$ and the MTI $F_\mu = A_\mu^2 / (2\sigma_{\mu_1}^2 + 2\sigma_{\mu_2}^2)$ determine different levels of fading conditions, e.g., Rayleigh fading ($F_S = 0$ or $F_\mu = 0$), Rician fading ($F_S = 10$ or $F_\mu = 10$) and essentially no fading ($F_S = 1000$ or $F_\mu = 1000$). Note that A_S^2 and A_μ^2 represent the power levels of the LOS component of the desired signal and the MTI, respectively. Finally, we also define the power ratios of the two main correlated diffused paths of the desired signal and of the MTI as $\varepsilon_S = \sigma_{S_2}^2 / \sigma_{S_1}^2$ and $\varepsilon_\mu = \sigma_{\mu_2}^2 / \sigma_{\mu_1}^2$, respectively.

We can derive the ML receiver structure by using the following expression [78]:

$$z_{ml} = \ln \left[\frac{p_{r_{ml}}(r_{ml} | q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \boldsymbol{\psi}_S, \boldsymbol{\psi}_\mu, S)}{p_{r_{ml}}(r_{ml} | q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \boldsymbol{\psi}_\mu, NS)} \right], \quad (4.34)$$

where $\boldsymbol{\psi}_S = [\phi_{S,l}, \theta_{S_1,l}, \theta_{S_2,l}]$, $\boldsymbol{\psi}_\mu = [\phi_{\mu l}, \phi'_{\mu l}, \theta_{\mu_1,l}, \theta_{\mu_2,l}, \theta'_{\mu_1,l}, \theta'_{\mu_2,l}]$. Note that the numerator of the logarithm function is the pdf of the square-law detector output of the signal branch and the denominator is the corresponding pdf of the non-signal branch.

After some mathematical manipulations, the in-phase and quadrature-phase components of r_{ml} , as defined in (4.2) and (4.3), can be derived as

$$\begin{aligned}
r_{I,ml} = & x_{S_1,l}\sigma_{S_1}u_1\sqrt{1+\lambda_S^2\varepsilon_S+u_2}+x_{S_2,l}\sigma_{S_2}u_1\sqrt{1-\lambda_S^2}+q_{\alpha l}x_{\mu_1,l}\sigma_{\mu_1}u_3\sqrt{1+\lambda_\mu^2\varepsilon_\mu+u_4} \\
& + q_{\alpha l}x_{\mu_2,l}\sigma_{\mu_2}u_3\sqrt{1-\lambda_\mu^2}+q'_{\beta l}x'_{\mu_1,l}\sigma_{\mu_1}u_5\sqrt{1+\lambda_\mu^2\varepsilon_\mu+u_6} \\
& + q'_{\beta l}x'_{\mu_2,l}\sigma_{\mu_2}u_5\sqrt{1-\lambda_\mu^2}+\frac{A_{ml}}{\sqrt{2}}+n_{I,l}
\end{aligned} \tag{4.35}$$

and

$$\begin{aligned}
r_{Q,ml} = & -y_{S_1,l}\sigma_{S_1}u_1\sqrt{1+\lambda_S^2\varepsilon_S+u_2}-y_{S_2,l}\sigma_{S_2}u_1\sqrt{1-\lambda_S^2}-q_{\alpha l}y_{\mu_1,l}\sigma_{\mu_1}u_3\sqrt{1+\lambda_\mu^2\varepsilon_\mu+u_4} \\
& - q_{\alpha l}y_{\mu_2,l}\sigma_{\mu_2}u_3\sqrt{1-\lambda_\mu^2}-q'_{\beta l}y'_{\mu_1,l}\sigma_{\mu_1}u_5\sqrt{1+\lambda_\mu^2\varepsilon_\mu+u_6} \\
& - q'_{\beta l}y'_{\mu_2,l}\sigma_{\mu_2}u_5\sqrt{1-\lambda_\mu^2}-\frac{A_{ml}}{\sqrt{2}}-n_{Q,l},
\end{aligned} \tag{4.36}$$

where

$$u_1 = (1-\chi)\text{sinc}\{(1-\chi)[(f_n-f_m)T_h+\xi]\},$$

$$u_2 = 2\lambda_S\sqrt{\varepsilon_S}\cos\left\{\frac{2\pi\tau_S[(f_n-f_m)T_h+\xi]}{T_h}+\theta_{S_1,l}-\theta_{S_2,l}\right\},$$

$$u_3 = (1-\chi)\text{sinc}\{(1-\chi)[(f_\alpha-f_m)T_h+\xi]\},$$

$$u_4 = 2\lambda_\mu\sqrt{\varepsilon_\mu}\cos\left\{\frac{2\pi\tau_\mu[(f_\alpha-f_m)T_h+\xi]}{T_h}+\theta_{\mu_1,l}-\theta_{\mu_2,l}\right\},$$

$$u_5 = \chi\text{sinc}\{\chi[(f_\beta-f_m)T_h+\xi]\},$$

and

$$u_6 = 2\lambda_\mu\sqrt{\varepsilon_\mu}\cos\left\{\frac{2\pi\tau_\mu[(f_\beta-f_m)T_h+\xi]}{T_h}+\theta'_{\mu_1,l}-\theta'_{\mu_2,l}\right\}.$$

Note that $n_{I,l}$ and $n_{Q,l}$ are independent Gaussian random variables with zero mean and variance of $[q_l(1-\chi) + q'_l\chi]^2 \sigma_P^2/2 + \sigma_\omega^2/2$; and the parameter A_{ml} is determined as

$$A_{ml}^2 = \left(A_S \cos \Phi_1 u_1 + q_{\alpha l} A_\mu \cos \Phi_2 u_3 + q'_{\beta l} A_\mu \cos \Phi_3 u_5 \right)^2 + \left(-A_S \sin \Phi_1 u_1 - q_{\alpha l} A_\mu \sin \Phi_2 u_3 - q'_{\beta l} A_\mu \sin \Phi_3 u_5 \right)^2, \quad (4.37)$$

where $\Phi_1 = \pi(\chi+1)[(f_n - f_m)T_h + \xi] + \phi_{S,l}$, $\Phi_2 = \pi(\chi+1)[(f_\alpha - f_m)T_h + \xi] + \phi_{\mu,l}$ and $\Phi_3 = \pi(\chi+2)[(f_\beta - f_m)T_h + \xi] + \phi'_{\mu,l}$. Both the derived random variables $r_{I,ml}$ and $r_{Q,ml}$ in (4.35) and (4.36) are Gaussian distributed with mean $A_{ml}/\sqrt{2}$ and common variance of

$$\sigma_{ml}^2 = \sigma_{S_1}^2 u_1^2 (1 + \varepsilon_S + u_2) + q_{\alpha l} \sigma_{\mu_1}^2 u_3^2 (1 + \varepsilon_\mu + u_4) + q'_{\beta l} \sigma_{\mu_1}^2 u_5^2 (1 + \varepsilon_\mu + u_6) + \frac{[q_l(1-\chi) + q'_l\chi]^2}{2} \sigma_P^2 + \frac{\sigma_\omega^2}{2}. \quad (4.38)$$

According to [74], the sum of squares of two statistically independent Gaussian random variables with non-zero mean and common variance has a non-central chi-square distribution. Thus r_{ml} is non-central chi-square distributed with two degrees of freedom and its corresponding pdf can be expressed as [74]

$$p_{r_{ml}}(r_{ml} | q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \psi_S, \psi_\mu, f_n) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml} + A_{ml}^2}{2\sigma_{ml}^2}\right) \times I_0\left(\frac{A_{ml}\sqrt{r_{ml}}}{\sigma_{ml}^2}\right) U(r_{ml}). \quad (4.39)$$

The parameter A_{ml}^2 given by (4.37) is called the noncentrality parameter of the non-central chi-square distribution. From (4.34) and (4.39), we can derive the suboptimum ML receiver structure as

$$z_{ml} = B_{ml} (\sqrt{r_{ml}} + C_{ml})^2 + D_{ml} \quad (4.40)$$

where

$$B_{ml} = \frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2},$$

$$C_{ml} = \frac{A_{ml,NS}\sigma_{ml,S}^2 - A_{ml,S}\sigma_{ml,NS}^2}{\sigma_{ml,NS}^2 - \sigma_{ml,S}^2},$$

and

$$D_{ml} = E_{ml} - B_{ml}C_{ml}^2.$$

The expressions for $p_{r_{ml}}(r_{ml}|q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \psi_S, \psi_\mu, S)$, $p_{r_{ml}}(r_{ml}|q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \psi_\mu, NS)$, E_{ml} , $\sigma_{ml,S}^2$, $\sigma_{ml,NS}^2$, $A_{ml,S}^2$ and $A_{ml,NS}^2$, and the detailed derivation of the suboptimum ML receiver structure can be found in Appendix A.

By performing some mathematical manipulations with the aid of [74], the pdf of the suboptimum ML receiver output can be derived as

$$\begin{aligned} p_{z_{ml}}(z_{ml}|q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \psi_S, \psi_\mu, f_n) \\ = \frac{G_{ml}}{2\sigma_{ml}^2 \sqrt{B_{ml}(z_{ml} - D_{ml})}} \exp\left(-\frac{G_{ml}^2 + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{A_{ml}G_{ml}}{\sigma_{ml}^2}\right) U(G_{ml}) \\ - \frac{H_{ml}}{2\sigma_{ml}^2 \sqrt{B_{ml}(z_{ml} - D_{ml})}} \exp\left(-\frac{H_{ml}^2 + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{A_{ml}H_{ml}}{\sigma_{ml}^2}\right) U(-H_{ml}), \quad (4.41) \end{aligned}$$

where $G_{ml} = \sqrt{\frac{z_{ml} - D_{ml}}{B_{ml}}} - C_{ml}$ and $H_{ml} = \sqrt{\frac{z_{ml} - D_{ml}}{B_{ml}}} + C_{ml}$.

By averaging the conditional symbol error probability $P_{s|\mathbf{K}_\mu, \mathbf{K}_P, f_n}$ over $\mathbf{K}_P$ and $\mathbf{K}_\mu$, we can obtain the average symbol error probability as given in (4.21). Note that the matrices $\mathbf{K}_P$ and $\mathbf{K}_\mu$ represent the PBNI and MTI distribution patterns, which are given by (4.19) and

$$\mathbf{K}_\mu = \begin{pmatrix} q_{11} & q_{12} & \cdots & q_{1L} & q'_{11} & q'_{12} & \cdots & q'_{1L} \\ q_{21} & q_{22} & \cdots & q_{2L} & q'_{21} & q'_{22} & \cdots & q'_{2L} \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\ q_{M1} & q_{M2} & \cdots & q_{ML} & q'_{M1} & q'_{M2} & \cdots & q'_{ML} \end{pmatrix}, \quad (4.42)$$

respectively. The conditional symbol error probability $P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n}$ in (4.21) can be expressed as

$$\begin{aligned} P_{s|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n} &= 1 - \sum_{v_1=-VL/2}^{v_n} p_{\hat{y}_1|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n} \left(\frac{2v_1 R}{V} \right) \times \cdots \\ &\times \sum_{v_{n-1}=-VL/2}^{v_n} p_{\hat{y}_{n-1}|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n} \left(\frac{2v_{n-1} R}{V} \right) \times \sum_{v_n=-VL/2}^{VL/2} p_{\hat{y}_n|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n} \left(\frac{2v_n R}{V} \right) \\ &\times \sum_{v_{n+1}=-VL/2}^{v_n} p_{\hat{y}_{n+1}|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n} \left(\frac{2v_{n+1} R}{V} \right) \times \cdots \times \sum_{v_M=-VL/2}^{v_n} p_{\hat{y}_M|\mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n} \left(\frac{2v_M R}{V} \right) \end{aligned} \quad (4.43)$$

where R and V are the number of quantization steps and the upper limit of quantization range, respectively. Note that $\hat{y}_m = \sum_{l=1}^L \hat{z}_{ml}$ denotes the final decision statistics, where $\hat{z}_{ml}$ is the V -level uniform quantization approximation of z_{ml} . The corresponding pdf expressions of $\hat{y}_m$ are given by

$$p_{\hat{y}_m}(\hat{y}_m | \mathbf{K}_\mu, \mathbf{K}_\mathbf{P}, f_n) = \left[p_{\hat{z}_m}(\hat{z}_m | q_{\alpha l}, q'_{\beta l}, q_l, q'_l, f_n) \right]^{\otimes L}. \quad (4.44)$$

Here the subscript l of $p_{z_{ml}}(z_{ml} | q_{\alpha l}, q'_{\beta l}, q_l, q'_l, f_n)$ is ignored because all the L hops are independent of each other. Note that the superscript $\otimes L$ represents the $(L-1)$ -fold discrete convolution operation. The pdf $p_{z_{ml}}(z_{ml} | q_{\alpha l}, q'_{\beta l}, q_l, q'_l, f_n)$ can be obtained by taking the expectation of (4.41) over $\psi_\mathbf{S}$ and ψ_μ . Note that the term $P(\mathbf{K}_\mu, \mathbf{K}_\mathbf{P})$ in (4.21) is the probability of PBNI and MTI patterns, which is given by (4.22), except

Figure 4.10. Analytical and simulated worst-case BER results of the suboptimum ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $\chi = \xi = 0.1$, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.5$, $\nu_\mu = \nu_P = 0.5$, and different Rician fading factors F_S , F_μ .

that the term g_{ml} is replaced by q'_{ml} . Finally, the average BER can be obtained by following (4.20). The derived BER expressions will allow system engineers to optimize the performance and design the FFH ML receiver in a fast and efficient way to combat the composite interference effects, as well as the timing and frequency offsets.

Figure 4.11. Worst-case BER performance of the suboptimum ML receiver with $M = 4$, SNR = 13.35 dB, $\chi = \xi = 0.1$, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.5$, $F_S = F_\mu = 3$, $\nu_\mu = \nu_P = 0.5$, and different diversity orders L .

4.3.2 Numerical Results and Discussions

In Fig. 4.10, we present both the analytical and simulated worst-case BER results of the proposed suboptimum ML receiver under various fading conditions when both timing and frequency offsets are present ($\chi = \xi = 0.1$). Both the MTI and PBNI coexist in the system ($\nu_\mu = \nu_P = 0.5$). From this figure, it can be observed that the frequency-selective Rayleigh fading case ($F_S = F_\mu = 0$) causes the worst degradation in the system performance as compared with the frequency-selective Rician fading cases ($F \neq 0$).

This can be explained as follows. The Rayleigh fading channel is a special case of the Rician fading channel when there are multiple indirect paths and there is no dominant LOS propagation between the transmitter and the receiver. Due to the absence of the LOS component, the Rayleigh fading channel will degrade the system error probability performance more than that of the Rician fading channel. Both the analytical and simulation results match well with each other, which validates the analytical BER expressions derived in the previous section. Thus, only analytical results are presented in Figs. 4.11–4.19. In Fig. 4.11, the worst-case BER results of the suboptimum ML receiver are provided for different diversity levels L . For the low-to-moderate SIR range, which is of practical interest, we can observe that there exists an optimum diversity level, i.e., $L = 5$, whereas for the $\text{SIR} > 25$ dB region, the optimum diversity level is $L = 11$.

In Fig. 4.12, we present the worst-case BER results and demonstrate the composite effect of PBNI and MTI on the system performance of the suboptimum ML receiver with the presence of the timing and frequency offsets ($\chi = \xi = 0.1$). For both cases of frequency-selective Rician and Rayleigh fading channels, the BER curves under the composite effect are observed to lie between the ones under one type of interference only. It can also be observed that the system performance under the MTI only is the best and its BER curve is almost flat throughout the SIR range. This demonstrates that the proposed suboptimum ML receiver is very effective in suppressing the MTI. Figure 4.13 illustrates the effect of the timing and frequency offsets on the proposed suboptimum ML receiver performance. Generally, when either the timing offset or the frequency offset increases, the system performance degrades significantly. This is due to the loss of orthogonality caused by those offsets. Moreover, the imperfect timing synchronization also causes the signal attenuation at the detector matched to the incoming frequency [85]. The degradation in performance due to the timing offset

Figure 4.12. Worst-case BER performance of the suboptimum ML receiver with $M = 4$, $L = 3$, SNR = 13.35 dB, $\chi = \xi = 0.1$, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.5$, different Rician fading factors F_S , F_μ , and various interference power ratios ν_μ , ν_P .

is observed to be more severe than that of the frequency offset when the normalized value of those offsets is less than 0.3. The opposite is true for larger normalized values.

Figure 4.14 shows the sensitivity of the system to the frequency-selectivity of the Rician fading channel, which is represented by the parameters ε_S , ε_μ , ρ_S and ρ_μ . The desired signal is corrupted by both the PBNI and the MTI ($\nu_\mu = \nu_P = 0.5$) with the presence of timing and frequency offsets ($\chi \neq 0$, $\xi \neq 0$). From this figure, we can

Figure 4.13. Worst-case BER results of the suboptimum ML receiver with $M = 4$, $\text{SNR} = 13.35$ dB, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.7$, $F_S = F_\mu = 3$, $\nu_\mu = 1$, $\nu_P = 0$, and different timing and frequency offsets χ , ξ .

conclude that the BER performance of the proposed receiver and LCR is not very sensitive to the frequency-selectivity of the Rician fading channel.

In Fig. 4.15, we provide the worst-case BER results of the optimum ML receiver, which are obtained by using (A.7). Those results are obtained via simulation. The BER results of the suboptimum and optimum ML receivers with the same system parameters ($\chi = \xi = 0.1$, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.9$) are quite close to each other. This confirms

Figure 4.14. Worst-case BER results of the suboptimum ML and linear-combining receivers with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $F_S = F_\mu = 5$, $\nu_\mu = \nu_P = 0.5$, and different values of χ , ξ , ε_S , ε_μ , ρ_S and ρ_μ .

that the approximation $\ln I_0(x) = x$ adopted in the analysis is reasonable. Besides that, the BER performance of the LCR for the case of $\chi = \xi = 0$, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.1$ is also presented in Fig. 4.15. Note that the LCR has lower BER results than the proposed ML receiver in high SIR condition as the BER curve of the LCR was obtained without timing and frequency offsets ($\chi = \xi = 0$), whereas the BER curves of the proposed ML receiver were obtained with timing and frequency offsets ($\chi = \xi = 0.1$). From this figure, we can observe that for strong and moderate interference ($0 \text{ dB} \leq \text{SIR}$

Figure 4.15. Worst-case BER results of the suboptimum ML, optimum ML and linear-combining receivers with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $F_S = F_\mu = 5$, $\nu_\mu = \nu_P = 0.5$, and different values of χ , ξ , ε_S , ε_μ , ρ_S and ρ_μ .

≤ 28 dB), which is of practical interest, the ML receiver with $\chi = \xi = 0.1$ significantly outperforms the LCR with $\chi = \xi = 0$. This shows that when timing and frequency offsets are present (specifically, $\chi = \xi = 0.1$), the proposed suboptimum ML receiver can suppress the composite interference effect even more effectively as compared to that of the LCR with perfect compensation of its timing and frequency offsets.

In Fig. 4.16, the worst-case BER results of various diversity-combining receivers are provided. Note that the results of the LCR and the PCR are obtained via simulations.

Figure 4.16. Worst-case BER performance comparison of various diversity-combining receivers with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $F_S = F_\mu = 3$, $\chi = \xi = 0.1$, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.5$, and various interference power ratios ν_μ , ν_P .

It can be observed that the proposed suboptimum ML receiver outperforms these two conventional receivers. The improvement in performance is very significant for the case of MTI only, i.e., up to two orders of magnitude. This is because the suboptimum ML receiver has knowledge of some side information, such as the noise variance; the power level, delay and frequency of MTI; the PBNI variance; the Rician fading factor; and the timing and frequency offsets, which can help to better counteract the interference and

Figure 4.17. Worst-case BER results of the suboptimum ML receiver with $M = 16$, $L = 5$, $\text{SNR} = 13.35$ dB, $\chi = \xi = 0.1$, $\nu_\mu = 1$, and $\nu_P = 0$ over frequency-selective Rayleigh fading channels with different values of ρ_S and ρ_μ .

fading effects. This is also the tradeoff between system performance and complexity. In this figure, the BER results with imperfect side information on χ are also included to confirm that the suboptimum ML receiver performance is robust against inaccurate estimation of χ . The robustness of the system performance to other required side information has also been examined and it has been observed that in general, the performance is insensitive to those side information.

Figure 4.18. Worst-case BER results of the ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 20$ dB, $\varepsilon_S = \varepsilon_\mu = 1$, $\rho_S = \rho_\mu = 0.5$, $\nu_\mu = 1$, and $\nu_P = 0$ over frequency-selective Rayleigh fading channels with various combinations of χ and ξ .

The worst-case BER results of the suboptimum ML receiver and LCR with various cross-correlation coefficients over frequency-selective Rayleigh fading channels are provided in Fig. 4.17. Note that by setting the Rician fading factors to be $F_S = 0$ and $F_\mu = 0$, we can use the pdf expressions derived in Section 4.3.1 to investigate the performance of the FFH/MFSK ML receiver under the frequency-selective Rayleigh fading channel as a special case. From this figure, it can be observed that the frequency-

Figure 4.19. Worst-case BER results of the ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 20$ dB, $\nu_\mu = \nu_P = 0.5$, and $\chi = \xi = 0.1$ over frequency-selective Rayleigh fading channels with different values of ϵ_S , ϵ_μ , ρ_S and ρ_μ .

selectivity of the Rayleigh fading affects the system performance of both the LCR and the ML receiver. It can also be observed that when the cross-correlation coefficients (ρ_S and ρ_μ) are increased, the system performance is degraded because the two main correlated diffused paths become more correlated. This observation for the ML receiver is consistent with the results presented in [52] under MTI only without the timing and frequency offsets.

Figure 4.18 demonstrates the impact of timing and frequency offsets on the worst-case BER performance of the proposed suboptimum ML receiver over frequency-selective Rayleigh fading channels. When the normalized values of timing and frequency offsets are less than 0.3, the performance of the suboptimum ML receiver is shown to be more sensitive to the timing offset than that of the frequency offset. It can be also observed that both these offsets degrade the performance very significantly. In this figure, the BER curves under MTI only are almost flat throughout the SIR range. This shows that the proposed suboptimum ML receiver is very effective in suppressing the MTI. In Fig. 4.19, the worst-case BER results of the suboptimum ML receiver with different values of ε_S , ε_μ , ρ_S and ρ_μ are provided. Both MTI and PBNI are present in the system ($\nu_\mu = \nu_P = 0.5$). We observe that the performance of this receiver is more sensitive to the cross-correlation coefficients (ρ_S and ρ_μ) than that of the power ratios of the two main correlated diffused paths (ε_S and ε_μ).

4.4 Conclusion

The composite effect of MTI and PBNI, as well as timing and frequency offsets on the performance of an FFH/MFSK system employing ML diversity-combining method has been studied over both frequency-nonselective and frequency-selective fading channels. The ML receivers have been proposed by using the log-likelihood ratio criterion. The analytical BER expressions for the proposed receivers have been derived and validated by simulations. Numerical results have demonstrated that the proposed suboptimum ML receiver is more effective in mitigating the composite interference effect and the effect of timing and frequency offsets on the system performance as compared with other existing receivers. Although the proposed receiver requires some side information to implement and this increases its complexity, it has been shown to be robust against the inaccuracy in estimation of those side information. Up to this point, only single-

user FFH systems have been studied. In the next chapter, FFH-MA systems will be investigated.

Chapter 5

Effect of MAI on FFH/MFSK

Systems over Frequency-Selective Fading Channels

5.1 Overview

When the transmitted signal bandwidth is greater than the coherence bandwidth of the channel, the frequency components of the transmitted signal with frequency separation exceeding the coherence bandwidth are subject to different gains and phase shifts. In such a case, the channel is said to be frequency-selective [92]. Frequency-selective multipath fading is commonly encountered in urban and indoor environments and is a significant source of potential degradation in a wideband mobile communication system [93]. In [67], the authors studied the performance of FFH-MA/BFSK systems employing equal-gain diversity-combining technique over frequency-selective Rayleigh fading channels. In [54], the performance of both MFSK and differentially phase-shift keying systems using the ML method has been analyzed in a frequency-selective

Rayleigh fading environment. It should be noted that the characteristic function derived in [54] is not a closed-form expression and thus the optimum nonlinearity was approximated by a soft limiter. In this chapter, we derive simple analytical BER expressions of various diversity-combining receivers for an FFH-MA system employing MFSK modulation scheme over frequency-selective Rayleigh fading channels. A ML receiver structure is proposed and analyzed. Following that, the analytical BER expressions for the suboptimum ML FFH-MA receiver will be derived over frequency-selective Rician fading channels. It is assumed that all the active users are chip-synchronous [69]. In mobile radio, this is the case for base-to-user transmission. We will first derive the pdf of the ML decision variable. Next, the discrete convolution approach will be applied to obtain the pdf of the final decision statistics. Moreover, the robustness of the proposed suboptimum ML receiver against the inaccuracy in estimation of the required side information is also examined.

5.2 Performance Study over Frequency-Selective Rayleigh Fading Channels

5.2.1 PDF Expression of the Square-Law Detector Outputs

Let K be the number of active users operating in this FFH-MA/MFSK system. It is assumed that all the signals arrive at a receiver with the same power strength [69]. The square-law detector outputs r_{ml} are processed by different diversity-combining techniques, denoted as the function $F(\cdot)$ in Fig. 2.3, to produce the resultant outputs z_{ml} . Without loss of generality, we assume that the first user is the desired user and there are K_l interfering users in the l th hop, where $K_l < K$. We further assume that the desired signal with the frequency f_1 is present at the first branch of the receiver. The desired signal is contaminated by MAI and AWGN and the dehopped signal in the

l th hop can be expressed as

$$r_l(t) = s_l(t) + \sum_{k=1}^{K_l} J_k(t) + \omega(t), \quad (5.1)$$

where $s_l(t)$ denotes the desired signal, $\sum_{k=1}^{K_l} J_k(t)$ is the MAI and $\omega(t)$ represents the noise term due to AWGN. Unless otherwise stated, we use the symbol k , for $k = 1, 2, \dots, K_l$ to denote the k th interfering user in the subsequent analysis. Following [52], the multipath frequency-selective Rayleigh fading channel is modeled as a correlated two-path channel with one hopping duration T_h . In our analysis, we assume that all the users experience mutually independent frequency-selective Rayleigh fading [35]. Thus we can rewrite (5.1) as

$$\begin{aligned} r_l(t) = & \sqrt{2}|a_{s_1,l}|\cos(2\pi f_1 t + \theta_{s_1,l}) + \sqrt{2}|a_{s_2,l}|\cos(2\pi f_1(t - \tau_s) + \theta_{s_2,l}) \\ & + \sum_{k=1}^{K_l} \left[\sqrt{2}|a_{k_1,l}|\cos(2\pi f_{k,l} t + \theta_{k_1,l}) + \sqrt{2}|a_{k_2,l}|\cos(2\pi f_{k,l}(t - \tau_k) + \theta_{k_2,l}) \right] + \omega(t) \end{aligned} \quad (5.2)$$

where $\sqrt{2}|a_{s_1,l}|$ and $\sqrt{2}|a_{s_2,l}|$ denote the two main correlated Rayleigh amplitudes of the desired signal; $\sqrt{2}|a_{k_1,l}|$ and $\sqrt{2}|a_{k_2,l}|$ are the two main correlated Rayleigh amplitudes of the k th interfering user signal. In (5.2), f_1 and $f_{k,l}$ are the baseband frequencies of the desired signal and the k th interfering user signal, respectively. The terms $\theta_{s_1,l}$, $\theta_{s_2,l}$ and τ_s denote the random phases of the two paths of the desired signal in the l th hop and the time delay between the two correlated paths of the desired signal, respectively. Similar notations ($\theta_{k_1,l}$, $\theta_{k_2,l}$ and τ_k) are used for the k th interfering user signal. The random variables $\theta_{s_1,l}$, $\theta_{s_2,l}$, $\theta_{k_1,l}$, $\theta_{k_2,l}$ are assumed to be uniformly distributed over $[0, 2\pi]$ and τ_s , τ_k to be uniformly distributed over $(0, T_h]$.

We define the power ratios of the two main correlated paths of the desired signal and the k th interfering user signal as $\varepsilon_s = \sigma_{s_2}^2/\sigma_{s_1}^2$ and $\varepsilon_k = \sigma_{k_2}^2/\sigma_{k_1}^2$, respectively. Note

that $\sigma_{s_1}^2$ and $\sigma_{s_2}^2$ are the diffused power levels of the first and second main paths of the desired signal, respectively; $\sigma_{k_1}^2$ and $\sigma_{k_2}^2$ denote the diffused power levels of the first and second main paths of the k th interfering user signal, respectively. Next, the two main correlated Rayleigh amplitudes of the desired signal can be generated from the two uncorrelated complex Gaussian-fading amplitudes with unit variance, $b_{s_1,l} = x_{s_1,l} + jy_{s_1,l}$ and $b_{s_2,l} = x_{s_2,l} + jy_{s_2,l}$ [83]. Similarly, the two main correlated Rayleigh amplitudes of the k th interfering user signal are generated from two random variables, i.e, $b_{k_1,l} = x_{k_1,l} + jy_{k_1,l}$ and $b_{k_2,l} = x_{k_2,l} + jy_{k_2,l}$. Following that, $|a_{s_1,l}|$, $|a_{s_2,l}|$, $|a_{k_1,l}|$, and $|a_{k_2,l}|$ can be rewritten as [52]

$$|a_{s_1,l}| = \sigma_{s_1} |x_{s_1,l} + jy_{s_1,l}|, \quad (5.3)$$

$$|a_{s_2,l}| = \sigma_{s_2} \left[\lambda_s^2 |x_{s_1,l} + jy_{s_1,l}|^2 + (1 - \lambda_s^2) |x_{s_2,l} + jy_{s_2,l}|^2 + 2\lambda_s \sqrt{1 - \lambda_s^2} (x_{s_1,l}x_{s_2,l} + y_{s_1,l}y_{s_2,l}) \right]^{1/2}, \quad (5.4)$$

$$|a_{k_1,l}| = \sigma_{k_1} |x_{k_1,l} + jy_{k_1,l}|, \quad (5.5)$$

and

$$|a_{k_2,l}| = \sigma_{k_2} \left[\lambda_k^2 |x_{k_1,l} + jy_{k_1,l}|^2 + (1 - \lambda_k^2) |x_{k_2,l} + jy_{k_2,l}|^2 + 2\lambda_k \sqrt{1 - \lambda_k^2} (x_{k_1,l}x_{k_2,l} + y_{k_1,l}y_{k_2,l}) \right]^{1/2}, \quad (5.6)$$

respectively. Note that $x_{s_1,l}, x_{s_2,l}, y_{s_1,l}, y_{s_2,l}, x_{k_1,l}, x_{k_2,l}, y_{k_1,l}$ and $y_{k_2,l}$ are independent

Gaussian random variables with zero mean and variance of one; λ_s and λ_k are functions of the cross-correlation coefficients (ρ_s and ρ_k) of the two main paths of the desired signal and the k th interfering user signal, given by (3.46) and

$$\lambda_k^2 \cong 8 \left[\sqrt{(4/\pi - 1) \rho_k + 1} - 1 \right], \quad (5.7)$$

respectively.

The square-law detector output r_{ml} including the effects of MAI and receiver noise is given in (3.7). After some mathematical manipulations, the expressions of the in-phase and quadrature-phase components can be obtained as

$$\begin{aligned} r_{I,ml} = & \sigma_{s1} |b_{s1,l}| \cos(\theta_{s1,l}) \delta(f_m - f_1) + \sigma_{s2} \left[\lambda_s^2 |b_{s1,l}|^2 + (1 - \lambda_s^2) |b_{s2,l}|^2 \right. \\ & \left. + \lambda_s \sqrt{1 - \lambda_s^2} (x_{s1,l} x_{s2,l} + y_{s1,l} y_{s2,l}) \right]^{1/2} \cos(\theta_{s2,l} - 2\pi f_m \tau_s) \delta(f_m - f_1) \\ & + \sum_{k=1}^{K_l} \sigma_{k1} |b_{k1,l}| \cos(\theta_{k1,l}) \delta(f_{k,l} - f_m) + \sum_{k=1}^{K_l} \sigma_{k2} \left[\lambda_k^2 |b_{k1,l}|^2 + (1 - \lambda_k^2) |b_{k2,l}|^2 \right. \\ & \left. + \lambda_k \sqrt{1 - \lambda_k^2} (x_{k1,l} x_{k2,l} + y_{k1,l} y_{k2,l}) \right]^{1/2} \cos(\theta_{k2,l} - 2\pi f_m \tau_k) \delta(f_{k,l} - f_m) + n_{I,l} \end{aligned} \quad (5.8)$$

and

$$\begin{aligned} r_{Q,ml} = & -\sigma_{s1} |b_{s1,l}| \sin(\theta_{s1,l}) \delta(f_m - f_1) - \sigma_{s2} \left[\lambda_s^2 |b_{s1,l}|^2 + (1 - \lambda_s^2) |b_{s2,l}|^2 \right. \\ & \left. + \lambda_s \sqrt{1 - \lambda_s^2} (x_{s1,l} x_{s2,l} + y_{s1,l} y_{s2,l}) \right]^{1/2} \sin(\theta_{s2,l} - 2\pi f_m \tau_s) \delta(f_m - f_1) \\ & - \sum_{k=1}^{K_l} \sigma_{k1} |b_{k1,l}| \sin(\theta_{k1,l}) \delta(f_{k,l} - f_m) - \sum_{k=1}^{K_l} \sigma_{k2} \left[\lambda_k^2 |b_{k1,l}|^2 + (1 - \lambda_k^2) |b_{k2,l}|^2 \right. \\ & \left. + \lambda_k \sqrt{1 - \lambda_k^2} (x_{k1,l} x_{k2,l} + y_{k1,l} y_{k2,l}) \right]^{1/2} \sin(\theta_{k2,l} - 2\pi f_m \tau_k) \delta(f_{k,l} - f_m) - n_{Q,l}, \end{aligned} \quad (5.9)$$

respectively. After some mathematical manipulations, we can obtain the following expressions

$$\begin{aligned}
u_{I,ml} = & \left[x_{s_1,l} \sigma_{s_1} \sqrt{1 + \lambda_s^2 \varepsilon_s + 2\lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l}} + x_{s_2,l} \sigma_{s_2} \sqrt{1 - \lambda_s^2} \right] \delta(f_m - f_1) \\
& + \sum_{k=1}^{K_l} x_{k_1,l} \sigma_{k_1} \sqrt{1 + \lambda_k^2 \varepsilon_k + 2\lambda_k \sqrt{\varepsilon_k} \cos \theta_{k,l}} \delta(f_{k,l} - f_m) \\
& + \sum_{k=1}^{K_l} x_{k_2,l} \sigma_{k_2} \sqrt{1 - \lambda_k^2} \delta(f_{k,l} - f_m) + n_{I,l}
\end{aligned} \tag{5.10}$$

and

$$\begin{aligned}
u_{Q,ml} = & - \left[y_{s_1,l} \sigma_{s_1} \sqrt{1 + \lambda_s^2 \varepsilon_s + 2\lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l}} + y_{s_2,l} \sigma_{s_2} \sqrt{1 - \lambda_s^2} \right] \delta(f_m - f_1) \\
& - \sum_{k=1}^{K_l} y_{k_1,l} \sigma_{k_1} \sqrt{1 + \lambda_k^2 \varepsilon_k + 2\lambda_k \sqrt{\varepsilon_k} \cos \theta_{k,l}} \delta(f_{k,l} - f_m) \\
& - \sum_{k=1}^{K_l} y_{k_2,l} \sigma_{k_2} \sqrt{1 - \lambda_k^2} \delta(f_{k,l} - f_m) - n_{Q,l}.
\end{aligned} \tag{5.11}$$

Note that

$$u_{I,ml}^2 + u_{Q,ml}^2 = r_{I,ml}^2 + r_{Q,ml}^2, \tag{5.12}$$

$$\theta_{s,l} = \theta_{s_1,l} - \theta_{s_2,l} + 2\pi f_m \tau_s \tag{5.13}$$

and

$$\theta_{k,l} = \theta_{k_1,l} - \theta_{k_2,l} + 2\pi f_m \tau_k. \tag{5.14}$$

The random variables $u_{I,ml}$ and $u_{Q,ml}$ in (5.10) and (5.11) are Gaussian distributed

with zero mean and a common variance given by

$$\begin{aligned} \sigma_{ml}^2 = & \left(\sigma_{s_1}^2 + 2\sigma_{s_1}^2 \lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l} + \sigma_{s_2}^2 \right) \delta(f_m - f_1) \\ & + \sum_{k=1}^{K_l} \left(\sigma_{k_1}^2 + 2\sigma_{k_1}^2 \lambda_k \sqrt{\varepsilon_k} \cos \theta_{k,l} + \sigma_{k_2}^2 \right) \delta(f_{k,l} - f_m) + \frac{\sigma_\omega^2}{2}. \end{aligned} \quad (5.15)$$

According to [74], the square-law detector output r_{ml} is central chi-squared distributed with two degrees of freedom and its corresponding conditional pdf expression can be expressed as

$$p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml}^2}\right) U(r_{ml}), \quad (5.16)$$

where the vector $\boldsymbol{\theta}_l = [\theta_{s,l}, \theta_{1,l}, \dots, \theta_{K_l,l}]$ represents the phases of all active users' signals in the l th hop, as defined in (5.13) and (5.14). The vector $\mathbf{f}_l = [f_{1,l}, f_{2,l}, \dots, f_{K_l,l}]$ represents the frequencies of all the K_l interfering users' signals in the l th hop.

5.2.2 Performance Analysis of the ML Receiver

The function $F(\cdot)$ of the ML receiver can be expressed as [53]

$$z_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l, f_m = f_1)}{p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l, f_m \neq f_1)} \right). \quad (5.17)$$

where $p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l, f_m = f_1)$ represents the pdf of the square-law detector output when the desired signal is present and $p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l, f_m \neq f_1)$ denotes the corresponding pdf when the signal is not present. From (5.16), we obtain

$$p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l, f_m = f_1) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,S}^2}\right) U(r_{ml}) \quad (5.18)$$

and

$$p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l, f_m \neq f_1) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml}}{2\sigma_{ml,NS}^2}\right) U(r_{ml}), \quad (5.19)$$

where

$$\begin{aligned} \sigma_{ml,S}^2 &= \sigma_{s_1}^2 + 2\sigma_{s_1}^2 \lambda_s \sqrt{\varepsilon_s} \cos\theta_{s,l} + \sigma_{s_2}^2 \\ &+ \sum_{k=1}^{K_l} (\sigma_{k_1}^2 + 2\sigma_{k_1}^2 \lambda_k \sqrt{\varepsilon_k} \cos\theta_{k,l} + \sigma_{k_2}^2) \delta(f_{k,l} - f_m) + \frac{\sigma_\omega^2}{2} \end{aligned} \quad (5.20)$$

and

$$\sigma_{ml,NS}^2 = \sum_{k=1}^{K_l} (\sigma_{k_1}^2 + 2\sigma_{k_1}^2 \lambda_k \sqrt{\varepsilon_k} \cos\theta_{k,l} + \sigma_{k_2}^2) \delta(f_{k,l} - f_m) + \frac{\sigma_\omega^2}{2}. \quad (5.21)$$

Following that, by substituting (5.18) and (5.19) into (5.17), we obtain the ML decision variable z_{ml} as

$$z_{ml} = \alpha_{ml} r_{ml} + \beta_{ml}, \quad (5.22)$$

where $\alpha_{ml} = \frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2}$ and $\beta_{ml} = \ln\left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2}\right)$. The proposed ML receiver structure derived in (5.22) is simple but requires the side information on the noise variance σ_ω^2 and the frequencies $f_{k,l}$. The Gaussian noise variance can be estimated from a “noise-only” channel for noise power measurement [37], and the frequencies $f_{k,l}$ can be extracted from energy-detector [64]. From (5.16) and (5.22), the corresponding conditional pdf of z_{ml} can be obtained as

$$p_{z_{ml}}(z_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l) = \frac{1}{2\alpha_{ml}\sigma_{ml}^2} \exp\left(-\frac{z_{ml} - \beta_{ml}}{2\alpha_{ml}\sigma_{ml}^2}\right) U(z_{ml} - \beta_{ml}). \quad (5.23)$$

The corresponding CF expression of z_{ml} can be derived as

$$\psi_{z_{ml}}(jv|\boldsymbol{\theta}_l, \mathbf{f}_l) = \frac{\exp(jv\beta_{ml})}{1 - j2v\alpha_{ml}\sigma_{ml}^2}. \quad (5.24)$$

Following that, by taking expectation of $\psi_{z_{ml}}(jv|\boldsymbol{\theta}_l, \mathbf{f}_l)$ over $\boldsymbol{\theta}_l$ and $\mathbf{f}_l$, we obtain the unconditional CF of z_{ml} as

$$\psi_{z_{ml}}(jv) = E_{\boldsymbol{\theta}_l, \mathbf{f}_l} [\psi_{z_{ml}}(jv|\boldsymbol{\theta}_l, \mathbf{f}_l)]. \quad (5.25)$$

Since all the L hops are independent of each other, the subscript l can be ignored and the CF of the final decision statistics $y_m = \sum_{l=1}^L z_{ml}$ can be obtained as [94]

$$\psi_{y_m}(jv) = [\psi_{z_m}(jv)]^L. \quad (5.26)$$

Note that the CF of y_m in (5.26) can be obtained for higher diversity level L without extra computational complexity.

The corresponding pdf of y_m can be obtained by performing the inverse Fourier transform given by

$$p_{y_m}(y_m) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \psi_{y_m}(jv) e^{-jvy_m} dv. \quad (5.27)$$

Following that, the average BER P_b can be obtained by using (3.39), where P_s denotes the probability of symbol error, which is given by

$$P_s = 1 - \int_{-\infty}^{\infty} \left[\int_{-\infty}^{y_1} p_{y_2}(y_2) dy_2 \times \cdots \times \int_{-\infty}^{y_1} p_{y_M}(y_M) dy_M \right] \times p_{y_1}(y_1) dy_1 \quad (5.28)$$

5.2.3 Performance Analysis of the LCR

The function $F(\cdot)$ of the LCR is given in (2.3). The conditional pdf of z_{ml} can be obtained from (5.16) by replacing the variable r_{ml} by the variable z_{ml} . The corresponding

conditional CF of z_{ml} can be expressed as [74]

$$\psi_{z_{ml}}(jv|\boldsymbol{\theta}_l, \mathbf{f}_l) = \frac{1}{1 - j2v\sigma_{ml}^2}. \quad (5.29)$$

In order to obtain the average BER expression for the LCR, we follow the same steps shown in (5.25)–(5.28).

5.2.4 Performance Analysis of the PCR

For the PCR, the function $F(\cdot)$ is given in (2.5). The pdf of the decision variable z_{ml} can be obtained as

$$p_{z_{ml}}(z_{ml}|\boldsymbol{\theta}_l, \mathbf{f}_l) = p_{r_{ml}}(e^{z_{ml}}|\boldsymbol{\theta}_l, \mathbf{f}_l) e^{z_{ml}} = \frac{e^{z_{ml}}}{2\sigma_{ml}^2} \exp\left(-\frac{e^{z_{ml}}}{2\sigma_{ml}^2}\right). \quad (5.30)$$

With the aid of [95], the corresponding CF can be derived as

$$\psi_{z_{ml}}(jv|\boldsymbol{\theta}_l, \mathbf{f}_l) = (2\sigma_{ml}^2)^{jv} \Gamma(1 + jv). \quad (5.31)$$

Finally, we can follow (5.25)–(5.28) to obtain the average BER expression for the PCR.

5.2.5 Numerical Results and Discussions

In this section, we present the performance comparison of various diversity-combining receivers, namely, the LCR, the PCR and the ML receiver for the FFH-MA/MFSK system over frequency-selective Rayleigh fading channels. For subsequent numerical results, the BER values are plotted against the SNR and the number of users K . These BER curves allow engineers to determine how many users can be accommodated in the system based on the required BER values. In Fig. 5.1, both the analytical and simulated BER results of the proposed ML receiver and the two conventional

Figure 5.1. Analytical and simulated BER results of various receivers with $M = 4$, $L = 5$, $K = 10$, $\varepsilon_s = \varepsilon_k = 1$, and $\rho_s = \rho_k = 0.2$.

receivers are presented. The analytical results for these receivers are obtained based on the analytical expressions derived in Sections 5.2.2–5.2.4. The close match in both theoretical and simulation results validates the analytical expressions derived in the previous sections. We observe that the proposed ML receiver gives the best BER performance as compared with the two conventional receivers. For the low-to-moderate SNR values, the PCR is the worst while the LCR gives the worst system performance

Figure 5.2. BER performance of the ML receiver with $M = 4$, $\text{SNR} = 16$ dB, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.9$, and various diversity orders L .

for the high SNR region.

Figures 5.2, 5.3 and 5.4 show the BER results of the proposed ML receiver, the LCR and the PCR, respectively. The BER results are plotted against the number of users K for different diversity levels L . The parameters of the system are set to be $M = 4$, $\text{SNR} = 16$ dB, $\varepsilon_s = \varepsilon_k = 1$ and $\rho_s = \rho_k = 0.9$. From Figs. 5.2 and 5.4, it can be observed that there exists an optimum diversity level for the ML receiver and the PCR, i.e., $L = 15$ for the case of the ML receiver with $K \geq 56$ and $L = 11$ for the case of the PCR with $K \geq 45$. On the other hand, we observe that for the case of the LCR, the

Figure 5.3. BER performance of the LCR with $M = 4$, $\text{SNR} = 16$ dB, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.9$, and various diversity orders L .

further we increase the diversity order L , the better the BER performance is obtained. However, the incremental improvement will diminish eventually.

In Figs. 5.5 and 5.6, we compare the system performance of the proposed ML receiver with the two conventional receivers for different modulation orders M . The BER performance is plotted against the SNR and the number of users K in Figs. 5.5 and 5.6, respectively. From these two figures, we observe that for the cases of the ML receiver and the PCR, increasing the modulation order M improves the system performance significantly. This is consistent with the results for the MFSK system under AGWN

Figure 5.4. BER performance of the PCR with $M = 4$, $\text{SNR} = 16$ dB, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.9$, and various diversity orders L .

channels presented in [74]. However, for the case of the LCR, a larger M does not ensure the system performance improvement. It can also be observed that the system performance improvement is obtained by increasing M for all the values of SNR and the number of users K in the cases of the proposed ML receiver and the PCR. Moreover, for the ML receiver, this improvement is very significant (up to one order of magnitude). From these two figures, we can conclude that the improvement in system performance depends on many parameters of the system itself, i.e., the SNR, the number of users operating in the system, the diversity order L , the power ratios ε_s , ε_k and

Figure 5.5. Performance comparisons of various receivers with $L = 3$, $K = 10$, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.7$, and various modulation orders M .

the cross-correlation coefficients ρ_s, ρ_k .

Figures 5.7 and 5.8 present the system performance comparison of the three diversity-combining receivers for different values of $\varepsilon_s, \varepsilon_k$ and ρ_s, ρ_k , respectively. The BER values are plotted against K . The system performance under the fading channel with larger values of the power ratios ε_s and ε_k defined in Section 5.2.1 is slightly worse than that of the smaller values of ε_s and ε_k . It should be noted that if we set $\varepsilon_s = 0$ and

Figure 5.6. Performance comparisons of various receivers with $L = 5$, $\text{SNR} = 13.35$ dB, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.5$, and various modulation orders M .

$\varepsilon_k = 0$, the channel described in Section 5.2.1 becomes frequency-nonselective Rayleigh fading. In other words, by setting $\varepsilon_s = 0$ and $\varepsilon_k = 0$, the analytical BER expressions derived in Sections 5.2.2–5.2.4 can be applied to study the system performance under frequency-nonselective Rayleigh fading channels as a special case. Three BER curves with circular markers in Fig. 5.7 are the examples of the system performance of various diversity-combining receivers under frequency-nonselective Rayleigh fading channels.

Figure 5.7. Performance comparisons of various receivers with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $\rho_s = \rho_k = 0.5$, and various power ratios ϵ_s , ϵ_k .

Comparing to the power ratios ϵ_s and ϵ_k , increasing the values of the cross-correlation coefficients ρ_s and ρ_k degrades the system performance more significantly. This observation is consistent with the results shown in Fig. 3 of [52] for the BFSK system with MTI over frequency-selective Rayleigh fading channels. In Figs. 5.1 and 5.5-5.8, we observe that the proposed ML receiver always outperforms the two conventional receivers for all different parameter values of the investigated system and the fading

Figure 5.8. Performance comparisons of various receivers with $M = 4$, $L = 5$, $\text{SNR} = 13.35$ dB, $\varepsilon_s = \varepsilon_k = 1$, and various cross-correlation coefficients ρ_s, ρ_k .

channels. The improvement in performance can be as large as two orders of magnitude. This demonstrates that the proposed ML receiver is more effective in suppressing the MAI as compared with the linear-combining and product-combining receivers. However, the proposed ML receiver requires more side information to counteract the MAI more efficiently and this is the tradeoff between the system performance and complexity.

5.3 Performance Study over Frequency-Selective Rician Fading Channels

5.3.1 BER Performance Analysis

A Rician fading channel can be modeled as the sum of a dominant LOS component and a Rayleigh fading component [73]. The frequency-selective Rayleigh fading component can be modeled as a correlated two-path model within one hopping duration T_h by following [52]. Thus (5.1) can be rewritten as

$$\begin{aligned}
 r_l(t) = & \sqrt{2}A_s \cos(2\pi f_n t + \phi_{s,l}) + \sqrt{2}|a_{s1,l}| \cos(2\pi f_n t + \theta_{s1,l}) \\
 & + \sqrt{2}|a_{s2,l}| \cos(2\pi f_n(t - \tau_s) + \theta_{s2,l}) + \sum_{k=1}^{K_l} \left[\sqrt{2}A_k \cos(2\pi f_{k,l} t + \phi_{k,l}) \right. \\
 & \left. + \sqrt{2}|a_{k1,l}| \cos(2\pi f_{k,l} t + \theta_{k1,l}) + \sqrt{2}|a_{k2,l}| \cos(2\pi f_{k,l}(t - \tau_k) + \theta_{k2,l}) \right] + \omega(t),
 \end{aligned} \tag{5.32}$$

for $n = 1, 2, \dots, M$, where $\sqrt{2}A_s$ and $\sqrt{2}A_k$ represent the amplitudes of the LOS of the desired signal and the k th interfering user signal, respectively. Note that the terms $\sqrt{2}|a_{s1,l}|$, $\sqrt{2}|a_{s2,l}|$, $\sqrt{2}|a_{k1,l}|$, $\sqrt{2}|a_{k2,l}|$, f_n , $f_{k,l}$, $\theta_{s1,l}$, $\theta_{s2,l}$, τ_s , $\theta_{k1,l}$, $\theta_{k2,l}$, and τ_k have the same definitions as those in Section 5.2. The terms $\phi_{s,l}$ and $\phi_{k,l}$ represent the random phases of the LOS of the desired signal and the k th interfering user signal in the l th hop, respectively. The random variables $\phi_{s,l}$ and $\phi_{k,l}$ are assumed to be uniformly distributed over $[0, 2\pi]$.

It is assumed that the total average power of the desired user signal $P_S = A_s^2 + 2\sigma_{s1}^2 + 2\sigma_{s2}^2$ and the average power of all the interfering users' signals $P_k = A_k^2 + 2\sigma_{k1}^2 + 2\sigma_{k2}^2$, for $k = 1, 2, \dots, K_l$, remain constant from hop to hop. The parameters A_s^2 and A_k^2 represent the power levels of the LOS component of the desired user signal and the

k th interfering user's signal, respectively. The specular-to-total diffused power ratios $F_s = A_s^2 / (2\sigma_{s_1}^2 + 2\sigma_{s_2}^2)$ and $F_k = A_k^2 / (2\sigma_{k_1}^2 + 2\sigma_{k_2}^2)$ determine different levels of fading conditions, e.g., essentially no fading ($F_s = 1000$ or $F_k = 1000$), Rician fading ($F_s = 10$ or $F_k = 10$) and Rayleigh fading ($F_s = 0$ or $F_k = 0$).

After some mathematical manipulations, the expressions of the in-phase and quadrature-phase components can be obtained as

$$\begin{aligned}
r_{I,ml} = & \left[x_{s_1,l} \sigma_{s_1} \sqrt{1 + \lambda_s^2 \varepsilon_s + 2\lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l}} + x_{s_2,l} \sigma_{s_2} \sqrt{1 - \lambda_s^2} \right] \delta(f_m - f_n) \\
& + \sum_{k=1}^{K_l} x_{k_1,l} \sigma_{k_1} \sqrt{1 + \lambda_k^2 \varepsilon_k + 2\lambda_k \sqrt{\varepsilon_k} \cos \theta_{k,l}} \delta(f_{k,l} - f_m) \\
& + \sum_{k=1}^{K_l} x_{k_2,l} \sigma_{k_2} \sqrt{1 - \lambda_k^2} \delta(f_{k,l} - f_m) + \frac{A_{ml}}{\sqrt{2}} + n_{I,l}
\end{aligned} \tag{5.33}$$

and

$$\begin{aligned}
r_{Q,ml} = & - \left[y_{s_1,l} \sigma_{s_1} \sqrt{1 + \lambda_s^2 \varepsilon_s + 2\lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l}} + y_{s_2,l} \sigma_{s_2} \sqrt{1 - \lambda_s^2} \right] \delta(f_m - f_n) \\
& - \sum_{k=1}^{K_l} y_{k_1,l} \sigma_{k_1} \sqrt{1 + \lambda_k^2 \varepsilon_k + 2\lambda_k \sqrt{\varepsilon_k} \cos \theta_{k,l}} \delta(f_{k,l} - f_m) \\
& - \sum_{k=1}^{K_l} y_{k_2,l} \sigma_{k_2} \sqrt{1 - \lambda_k^2} \delta(f_{k,l} - f_m) - \frac{A_{ml}}{\sqrt{2}} - n_{Q,l},
\end{aligned} \tag{5.34}$$

where

$$\theta_{s,l} = \theta_{s_1,l} - \theta_{s_2,l} + 2\pi f_m \tau_s \tag{5.35}$$

and

$$\theta_{k,l} = \theta_{k_1,l} - \theta_{k_2,l} + 2\pi f_m \tau_k. \tag{5.36}$$

In (5.33) and (5.34), the parameter A_{ml} is given by

$$A_{ml}^2 = A_s^2 \delta(f_m - f_n) + \sum_{k=1}^{K_l} A_k^2 \delta(f_{k,l} - f_m) + \sum_{k=1}^{K_l} 2A_s A_k \cos \phi_{sk,l} \delta(f_m - f_n) \\ \times \delta(f_{k,l} - f_m) + \sum_{i=1}^{K_l} \sum_{j=1(i \neq j)}^{K_l} A_i A_j \cos \phi_{ij,l} \delta(f_{i,l} - f_m) \delta(f_{j,l} - f_m), \quad (5.37)$$

where

$$\phi_{sk,l} = \phi_{s,l} - \phi_{k,l} \quad (5.38)$$

and

$$\phi_{ij,l} = \phi_{i,l} - \phi_{j,l}. \quad (5.39)$$

The derived in-phase component $r_{I,ml}$ and quadrature-phase component $r_{Q,ml}$ are Gaussian distributed with a mean value of $A_{ml}/\sqrt{2}$ and a common variance given by

$$\sigma_{ml}^2 = (\sigma_{s_1}^2 + 2\sigma_{s_1}^2 \lambda_s \sqrt{\varepsilon_s} \cos \theta_{s,l} + \sigma_{s_2}^2) \delta(f_m - f_n) + \frac{\sigma_\omega^2}{2} \\ + \sum_{k=1}^{K_l} (\sigma_{k_1}^2 + 2\sigma_{k_1}^2 \lambda_k \sqrt{\varepsilon_k} \cos \theta_{k,l} + \sigma_{k_2}^2) \delta(f_{k,l} - f_m). \quad (5.40)$$

According to [74], the square-law detector output r_{ml} is noncentral chi-squared distributed with two degrees of freedom and A_{ml}^2 is its noncentrality parameter. The corresponding conditional pdf expression of r_{ml} can be expressed as

$$p_{r_{ml}}(r_{ml} | \boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l) = \frac{1}{2\sigma_{ml}^2} \exp\left(-\frac{r_{ml} + A_{ml}^2}{2\sigma_{ml}^2}\right) I_0\left(\frac{A_{ml}\sqrt{r_{ml}}}{\sigma_{ml}^2}\right) U(r_{ml}), \quad (5.41)$$

where the vectors $\boldsymbol{\theta}_l = [\theta_{s,l}, \theta_{1,l}, \dots, \theta_{K_l,l}]$ and $\mathbf{f}_l = [f_{1,l}, f_{2,l}, \dots, f_{K_l,l}]$ have the same definitions as given in Section 5.2, whereas the vector $\boldsymbol{\phi}_l = \left[\phi_{s1,l}, \phi_{s2,l}, \dots, \phi_{sK_l,l}, \phi_{12,l}, \dots, \phi_{1K_l,l}, \right. \\ \left. \phi_{21,l}, \phi_{23,l}, \dots, \phi_{2K_l,l}, \dots, \phi_{K_l1,l}, \dots, \phi_{K_lK_l-1,l} \right]$ represent the phases of all active users' signals in the l th hop, as defined in (5.38)-(5.39).

Based on the pdf expression derived in (5.41) and the ML criteria, the suboptimum ML receiver structure can be derived as

$$z_{ml} = D_{ml} (\sqrt{r_{ml}} + B_{ml})^2 + E_{ml} \quad (5.42)$$

where

$$D_{ml} = \frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2}, \quad (5.43)$$

$$B_{ml} = \frac{A_{ml,NS}\sigma_{ml,S}^2 - A_{ml,S}\sigma_{ml,NS}^2}{\sigma_{ml,NS}^2 - \sigma_{ml,S}^2} \quad (5.44)$$

and

$$E_{ml} = C_{ml} - D_{ml}B_{ml}^2. \quad (5.45)$$

The detailed derivation is provided in Appendix B. With the aid of [74], the conditional pdf of z_{ml} , when signal is present or not present, can be derived as

$$\begin{aligned} & p_{z_{ml}}(z_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m = f_n) \\ &= \frac{O_{ml}}{2\sigma_{ml,S}^2 \sqrt{D_{ml}(z_{ml} - E_{ml})}} \exp\left(-\frac{O_{ml}^2 + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) I_0\left(\frac{A_{ml,S}O_{ml}}{\sigma_{ml,S}^2}\right) U(O_{ml}) \\ &- \frac{Q_{ml}}{2\sigma_{ml,S}^2 \sqrt{D_{ml}(z_{ml} - E_{ml})}} \exp\left(-\frac{Q_{ml}^2 + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) I_0\left(\frac{A_{ml,S}Q_{ml}}{\sigma_{ml,S}^2}\right) U(-Q_{ml}) \end{aligned} \quad (5.46)$$

and

$$\begin{aligned}
p_{z_{ml}}(z_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m \neq f_n) \\
= \frac{O_{ml}}{2\sigma_{ml,NS}^2 \sqrt{D_{ml}(z_{ml} - E_{ml})}} \exp\left(-\frac{O_{ml}^2 + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) I_0\left(\frac{A_{ml,NS}O_{ml}}{\sigma_{ml,NS}^2}\right) U(O_{ml}) \\
- \frac{Q_{ml}}{2\sigma_{ml,NS}^2 \sqrt{D_{ml}(z_{ml} - E_{ml})}} \exp\left(-\frac{Q_{ml}^2 + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) I_0\left(\frac{A_{ml,NS}Q_{ml}}{\sigma_{ml,NS}^2}\right) U(-Q_{ml}),
\end{aligned} \tag{5.47}$$

respectively, where $O_{ml} = \sqrt{\frac{z_{ml} - E_{ml}}{D_{ml}}} - B_{ml}$ and $Q_{ml} = \sqrt{\frac{z_{ml} - E_{ml}}{D_{ml}}} + B_{ml}$.

Following that, by taking the expectation of $p_{z_{ml}}(z_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m = f_n)$ and $p_{z_{ml}}(z_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m \neq f_n)$ over $\boldsymbol{\theta}_l$, $\boldsymbol{\phi}_l$ and $\mathbf{f}_l$, we obtain the corresponding unconditional pdf expressions of z_{ml} as

$$p_{z_{ml}}(z_{ml}|f_m = f_n) = E_{\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l} [p_{z_{ml}}(z_{ml}/\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m = f_n)] \tag{5.48}$$

and

$$p_{z_{ml}}(z_{ml}|f_m \neq f_n) = E_{\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l} [p_{z_{ml}}(z_{ml}/\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m \neq f_n)], \tag{5.49}$$

respectively. By adopting the quantization approach, we can obtain the pdf of the discrete random variable $\hat{z}_{ml}$. Since all the L hops are independent of each other, the subscript l can be ignored and the pdf expressions of the final decision statistics $\hat{y}_m = \sum_{l=1}^L \hat{z}_{ml}$ can be obtained as [94]

$$p_{\hat{y}_m}^{\wedge}(\hat{y}_m | f_m = f_n) = \left[p_{\hat{z}_m}^{\wedge}(\hat{z}_m / f_m = f_n) \right]^{\otimes L} \tag{5.50}$$

and

$$p_{\hat{y}_m}^{\wedge}(\hat{y}_m | f_m \neq f_n) = \left[p_{\hat{z}_m}^{\wedge}(\hat{z}_m / f_m \neq f_n) \right]^{\otimes L}, \tag{5.51}$$

where the superscript $\otimes L$ denotes the $(L-1)$ -fold discrete convolution operation. Without loss of generality, we assume that the desired signal is present in the first branch of the receiver with frequency f_1 . Following that, the average BER can be computed using (3.39), where the probability of symbol error P_s is given by

$$P_s = 1 - \sum_{g_1=-GL/2}^{GL/2} p_{y_1}^{\wedge} \left(\frac{2g_1C}{G} \right) \left[\sum_{g_k=-GL/2}^{g_1} p_{y_k}^{\wedge} \left(\frac{2g_kC}{G} \right) \right]^{M-1},$$

for $k = 2, 3, \dots, M$. (5.52)

Note that C denotes the upper limit of quantization range and G is the number of steps used in the quantization process. Specifically, to obtain the BER results based on (5.52), we first obtain the conditional pdf of the ML decision variable z_{ml} . Then we take the expectation of the conditional pdf over $\boldsymbol{\theta}_l$, $\boldsymbol{\phi}_l$ and $\mathbf{f}_l$ to get the corresponding unconditional pdf of z_{ml} . Following that, the discrete convolution approach is adopted to obtain the pdf of the final statistics. Finally, the pdf of the final statistics is used for (5.52). The analytical BER expressions will allow system engineers to optimize the performance and design the FFH-MA ML receiver in a fast and efficient way to suppress the MAI over frequency-selective fading channels.

5.3.2 Numerical Results and Discussions

In this section, we present the BER performance of the suboptimum ML receiver for the FFH-MA/MFSK system over a frequency-selective Rician fading channel. In Fig. 5.9, both the analytical and simulated BER results of the proposed suboptimum ML receiver are presented. The BER results are obtained with different specular-to-total diffused power ratios F_s and F_k . The close match in both theoretical and simulation results validates the analytical expressions derived in Section 5.3.1. From Fig. 5.9, it can be observed that a larger specular-to-total diffused power ratio leads to better BER

Figure 5.9. Analytical and simulated BER results of the suboptimum ML receiver with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $\varepsilon_s = \varepsilon_k = 1$, and $\rho_s = \rho_k = 0.5$.

performance. It should be noted that the analytical BER results of the suboptimum ML receiver are obtained based on the BER expressions derived in Section 5.3.1. In order to implement the proposed suboptimum ML receiver, side information is required. Therefore, in this figure, the BER results with the imperfect side information are included to examine their effects on the system performance. The results with imperfect

Figure 5.10. BER performance of the suboptimum ML receiver with $M = 8$, $\text{SNR} = 13.35$ dB, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.7$, $F_s = F_k = 2$, and various diversity orders L .

side information are obtained via simulation. It can be observed that these BER curves are close to the BER curves with perfect side information. This demonstrates that the proposed ML receivers are robust against the inaccuracy in estimation of the required side information.

Figure 5.10 shows BER results of the suboptimum ML receiver with different diversity levels L over frequency-selective Rician fading channels. It can be observed that there exists an optimum diversity level for the suboptimum ML receiver, i.e., $L = 9$ with $K > 40$. From this figure, we also observe that the improvement in system per-

Figure 5.11. BER performance of the suboptimum ML receiver with $M = 4$, $L = 5$, SNR = 13.35 dB, and various values of ε_s , ε_k , and ρ_s , ρ_k .

formance by using diversity decreases as the number of users increases. This is because the noncoherent diversity-combining loss is larger when K increases.

In Figs. 5.11 and 5.12, the BER results of the suboptimum ML receiver with different power ratios ε_s , ε_k and cross-correlation coefficients ρ_s , ρ_k are provided under various types of fading. The BER results against K and SNR are presented in Figs. 5.11 and 5.12, respectively. From these two figures, it can be observed that when all users' signals experience frequency-selective Rician fading, the system performance is not very sensitive to the channel parameters ε_s , ε_k , ρ_s and ρ_k , i.e, the frequency-selectivity of the fading channel. On the other hand, for the case of Rayleigh fading ($F_s = F_k = 0$),

Figure 5.12. BER performance of the suboptimum ML receiver with $M = 4$, $L = 3$, $K = 30$, and various values of ϵ_s , ϵ_k , and ρ_s , ρ_k .

the system performance under the fading channel with larger values of the power ratios ϵ_s and ϵ_k is slightly worse than that of the smaller values of ϵ_s and ϵ_k . Comparing to the power ratios ϵ_s and ϵ_k , increasing the values of the cross-correlation coefficients ρ_s and ρ_k will degrade the system performance more significantly.

Figure 5.13 demonstrates the benefits of increasing the modulation order M for the proposed suboptimum FFH/MFSK ML receiver over frequency-selective Rician fading channels. The system parameters are set to be $K = 30$, $L = 3$, $\epsilon_s = \epsilon_k = 1$, and

Figure 5.13. BER performance of the suboptimum ML receiver with $L = 3$, $K = 30$, $F_s = F_k = 2$, $\varepsilon_s = \varepsilon_k = 1$, $\rho_s = \rho_k = 0.8$, and various modulation orders M .

$\rho_s = \rho_k = 0.8$. As expected, it is observed that the BER performance of the suboptimum ML receiver improves as M is increased. However, it should be noted that increasing the modulation order also leads to higher system complexity. From this figure, we also observe that the improvement in system performance becomes more significant for high SNR region. This is because the lower noise power causes less impact on the system performance.

In Figs. 5.14 and 5.15, we compare the BER performance of the proposed suboptimum ML receiver with two conventional diversity-combining receivers, namely, the

Figure 5.14. BER performance comparisons of various diversity-combining receivers with $M = 4$, $L = 3$, $\text{SNR} = 13.35$ dB, $\varepsilon_s = \varepsilon_k = 1$, and $\rho_s = \rho_k = 0.7$.

LCR and the PCR for different specular-to-total diffused power ratios (F_s, F_k) . The BER results of the two conventional receivers are obtained via simulation. These figures show that the proposed suboptimum ML receiver outperforms the two conventional receivers over frequency-selective Rician fading channels. The improvement in performance between the suboptimum ML receiver and the two conventional receivers can be as large as four orders of magnitude. We observe that this improvement becomes more significant when the number of users K increases. This is because although a higher

Figure 5.15. BER performance comparisons of various diversity-combining receivers with $M = 4$, $L = 3$, $K = 30$, $\varepsilon_s = \varepsilon_k = 1$, and $\rho_s = \rho_k = 0.9$.

K leads to higher MAI power, the suboptimum ML receiver has the perfect MAI information to suppress the MAI more efficiently. From Fig. 5.15, it can also be observed that the gap in system performance between the proposed suboptimum ML receiver and the two conventional receivers increases as the SNR increases. This is because the proposed receiver can better mitigate the MAI when the Gaussian noise power is low (high SNR).

5.4 Conclusion

In this chapter, we have derived the analytical BER expressions of a synchronous FFH-MA/MFSK system with the ML receivers over frequency-selective fading channels. For the case of frequency-selective Rayleigh fading, a simple pdf expression for the output of the square-law detector has been derived. Based on this pdf expression, the ML receiver structure has been proposed, and the corresponding analytical BER expressions for the ML receiver, the LCR and the PCR have been derived. For the case of frequency-selective Rician fading, the pdf of the ML decision variable has been derived and the discrete convolution approach has been used to obtain the pdf of the final decision statistics. To implement the proposed ML receivers, the side information is required. However, it has been shown that the proposed ML receivers are not sensitive to the imperfect side information. Our analysis, validated by simulation results, shows that the proposed ML receivers can suppress the MAI more effectively than the conventional receivers, i.e., improvement by up to four orders of magnitude in performance can be observed.

Chapter 6

Conclusions and Future Work

6.1 Concluding Remarks

In this thesis, we focused on exploiting and evaluating the potentials of employing maximum-likelihood diversity-combining methods in improving the BER performance for FFH/MFSK communication systems over fading channels. The composite effects of MTI and PBNI, as well as the MAI have been considered in the thesis. For MTI, we focused on the single-tone per band MTI since it was reported in [46] that this type of MTI causes the most damage to FFH systems as compared with other types of MTI. In addition, we have also investigated the impact of timing and frequency offsets on the performance of FFH/MFSK systems. Both frequency-nonselective and frequency-selective fading channels have been studied in the thesis. Furthermore, the robustness of the ML receivers against imperfect side information has also been examined.

Analytical BER expressions have been derived for an FFH/MFSK system using the log-likelihood ratio criterion under the composite effect of MTI and PBNI over frequency-nonselective Rician fading channels. Due to the fact that the computation of the logarithm of the modified Bessel function is nonlinear in nature, one of the existing approximations was adopted to derive the BER expressions for the analyzed system. A

suboptimum ML receiver structure was therefore proposed and analyzed. The analytical results were compared with simulation results. It has been shown that the analysis can be applied to study the system performance under frequency-nonselective Rayleigh fading channel as a special case. Under various interference and fading conditions, the proposed suboptimum ML receiver was found to be able to suppress the composite interference effectively. The BER performance of the proposed suboptimum ML receiver has been shown to significantly outperform some conventional diversity-combining receivers. Next, the composite effects of both PBNI and MTI on an FFH/MFSK ML receiver over a frequency-selective Rayleigh fading channel have been examined. Both the desired signal and the MTI tones were assumed to undergo independent frequency-selective fading. The analytical BER expressions have been derived and validated by computer simulation. Our analysis shows that the proposed system can effectively suppress the composite effect of PBNI and MTI. The performance of the proposed ML receiver has also been shown to be better than that of the linear-combining receiver and product-combining receiver under this composite effect.

Following that, we studied the composite effects of MTI and PBNI on an FFH/MFSK system for a more practical scenario where both timing and frequency offsets are present. Based on the characteristic function approach, analytical BER expressions for a ML receiver, as opposed to conventional diversity-combining receivers, have been derived over frequency-nonselective Rayleigh fading channels. BER performance of the proposed ML receiver was compared with various conventional diversity-combining techniques, such as the LCR, PCR, self-normalizing and OS-NED receivers. Among these receivers, the proposed ML receiver has been found to achieve the best performance at the expense of system complexity. In addition, a suboptimum ML receiver for FFH/MFSK systems has been proposed to combat the composite interference effect, as well as the impact of timing and frequency offsets over a frequency-selective Rician fading environment.

It has been shown that the proposed ML and suboptimum ML receivers are more effective in suppressing MTI than PBNI. Moreover, the BER curves under MTI only are almost flat throughout the range of SIR levels for all offset values. It has been concluded that the proposed receivers can almost completely suppress the MTI regardless of whether offsets are present or not. Furthermore, it has been found that when timing and frequency offsets are present, the proposed ML and suboptimum ML receivers can suppress the composite interference effect even more effectively as compared to that of the LCR with perfect compensation of its timing and frequency offsets. Although these proposed receivers require side information, it has been shown that their BER performance is not sensitive to inaccuracy in estimation of the required side information. In addition, the practical issues of timing and frequency offsets have also been considered.

Lastly, we have derived a simple pdf expression for the output of the square-law detector of a synchronous FFH-MA/MFSK system employing various diversity-combining techniques under frequency-selective Rayleigh fading channels. Based on this pdf expression, we first derived the ML receiver structure. Following that, we derived the analytical BER expressions for the ML, linear-combining and product-combining receivers. It has been shown that these BER expressions can be used to investigate the FFH-MA/MFSK system over frequency-nonselective Rayleigh fading channels as a special case. It was assumed that all the signals arrive at a receiver with the same power strength and all the users operating in the system undergo independent frequency-selective fading. Next, a suboptimum ML receiver structure has been proposed for an FFH-MA communication system over a frequency-selective Rician fading environment. The pdf of the suboptimum ML decision variable has been derived and the discrete convolution approach has been used to obtain the pdf expressions of the final decision statistics. Numerical BER results have shown that the proposed suboptimum ML receiver can improve the system performance significantly. Moreover, the robust-

ness of the suboptimum ML receiver against imperfect side information has also been examined. For FFH-MA systems, these proposed receivers have been found to give significant advantages over other existing diversity-combining receivers in mitigating MAI over frequency-selective fading channels, i.e., improvement by up to four orders of magnitude in BER performance can be observed. In general, all the numerical results obtained from the derived analytical BER expressions are in good agreement with the simulation results. In the thesis, the verification is made by simulating at bit level the entire system. The proposed FFH/MFSK ML receivers can more effectively countermeasure the interference and fading effects as the diversity order is increased. However, increasing the diversity order also increases noncoherent diversity-combining loss in the receiver and thus there always exists an optimum diversity order.

6.2 Future Work

There are several communication links for which the conventional additive Gaussian noise assumption is inadequate and the impulsive noise is therefore introduced [96]. It is interesting to study the performance of FFH systems employing different diversity-combining techniques under impulsive noise and interference. Furthermore, the effect of phase noise [97] on performance of various FFH receivers is also an interesting topic to be studied. Next, most analyses on diversity-combining techniques assume that the diversity branches are independent of each other. It may be also essential to investigate the performance of FFH diversity-combining receivers in which the diversity branches are correlated.

In FFH communication systems, because of the combination of L receptions to produce the final decision statistics, the pdf expressions of the final decision statistics usually involve with multiple levels of convolution. It is interesting to study some innovative ways to derive closed-form expressions for the pdf of final decision statistics.

Next, ML receivers are optimum receivers, and they provide the best system performance as compared to other conventional receivers. However, these ML receivers always require some side information to operate and this may lead to complex implementation. Therefore, more studies are required to find simpler non-parametric diversity-combining schemes, which can achieve comparable system performance to the optimum receivers. Besides, to further improve system performance, some cascaded diversity-combining receivers for FFH/MFSK systems may be considered. Thus far, the composite effects of MTI and PBNI on system performance were only considered for a single-user FFH system. In practice, the desired signal in FFH-MA systems may also be contaminated by both MTI and PBNI. Thus, research work in this direction is of great practical interests as well.

Author's Publications

Journal Publications

1. L.-M.-D. Le, K. C. Teh, and K. H. Li, "Performance analysis of a suboptimum fast frequency-hopped/ M -ary frequency-shift-keying maximum-likelihood receiver over Rician-fading channels with composite effects of partial-band noise jamming and multitone jamming," *IET Commun.*, vol. 6, no. 13, pp. 1903-1911, Sep. 2012.
2. L.-M.-D. Le, K. C. Teh, and K. H. Li, "Maximum-likelihood receivers for synchronous fast frequency-hopped multiple-access M -ary frequency-shift-keying systems over frequency-selective Rician-fading channels," *IET Commun.*, vol. 6, no. 17, pp. 2870-2878, Nov. 2012.
3. L.-M.-D. Le, K. C. Teh, and K. H. Li, "Jamming rejection using FFH/MFSK ML receiver over fading channels with the presence of timing and frequency offsets," *IEEE Trans. Inf. Forensics Security*, vol. 8, no. 7, pp. 1195-1200, Jul. 2013.
4. L.-M.-D. Le, K. C. Teh, and K. H. Li, "A survey on diversity-combining techniques for interference suppression in FFH systems," submitted for publication, 2014.

Conference Publications

1. L.-M.-D. Le, K. C. Teh, and K. H. Li, "Effects of composite PBNJ and MTJ on FFH/MFSK systems with maximum-likelihood receiver over frequency-selective fading channels," in *Proc. IEEE ICICS*, 2011, pp. 1-5.
2. L.-M.-D. Le, K. C. Teh, and K. H. Li, "Performance analysis of synchronous FFH-MA systems with ML receiver over frequency-selective Rayleigh-fading channels," in *Proc. IEEE ICCS*, 2012, pp. 508-512.

3. L.-M.-D. Le and K. C. Teh, “Maximum-likelihood FFH/MFSK receiver with MTJ and PBNJ over frequency-selective Rayleigh fading channels plus timing and frequency offsets,” in *Proc. IEEE ICICS*, 2013, pp. 1-5.

Bibliography

- [1] D. J. Torrieri, *Principles of Spread-Spectrum Communication Systems*. New York: Springer, 2005.
- [2] D. L. Schilling, L. B. Milstein, R. L. Pickholtz, M. Kullback, and F. Miller, “Spread spectrum for commercial communications,” *IEEE Commun. Mag.*, vol. 29, no. 4, pp. 66–79, Apr. 1991.
- [3] P. Flikkema, “Spread-spectrum techniques for wireless communication,” *IEEE Signal Process. Mag.*, vol. 14, no. 3, pp. 26–36, May 1997.
- [4] A. J. Al-Dweik and B. S. Sharif, “Exact performance analysis of synchronous FH-MFSK wireless networks,” *IEEE Trans. Veh. Technol.*, vol. 58, no. 7, pp. 3771–3776, Sep. 2009.
- [5] R. L. Pickholtz, L. B. Milstein, and D. L. Schilling, “Spread spectrum for mobile communications,” *IEEE Trans. Veh. Technol.*, vol. 40, no. 2, pp. 313–322, May 1991.
- [6] C. Jiang and J. Wang, “A novel receiver for FHMA systems,” *IEEE Trans. Veh. Technol.*, vol. 51, no. 5, pp. 1128–1137, Sep. 2002.
- [7] S. M. Wu, G. C. Yang, C. Y. Chang, and W. C. Kwong, “A two-level FH-CDMA scheme for wireless communication systems over fading channels,” *IEEE Trans. Commun.*, vol. 59, no. 1, pp. 38–44, Jan. 2011.

- [8] D. Verhulst, M. Mouly, and J. Szpirglas, “Slow frequency hopping multiple access for digital cellular radiotelephone,” *IEEE J. Sel. Areas Commun.*, vol. 2, no. 4, pp. 563–574, Jul 1984.
- [9] R. L. Peterson, R. E. Ziemer, and D. E. Borth, *Introduction to Spread-Spectrum Communications*. Prentice-Hall, 1995.
- [10] F. Gomez-Cuba, R. Asorey-Cacheda, and F. J. Gonzalez-Castano, “A survey of cooperative diversity for wireless networks,” *IEEE Commun. Surveys Tutorials*, vol. 14, no. 5, pp. 822–835, Third Quarter 2012.
- [11] Y. Maghsoodi and A. Al-Dweik, “Error-rate analysis of FHSS networks using exact envelope characteristic functions of sums of stochastic signals,” *IEEE Trans. Veh. Technol.*, vol. 57, no. 2, pp. 974–985, Mar. 2008.
- [12] Y.-S. Liu, “Diversity-combining and error-correction coding for FFH/MFSK systems over Rayleigh fading channels under multitone jamming,” *IEEE Trans. Wireless Commun.*, vol. 11, no. 2, pp. 771–779, Feb. 2012.
- [13] X. Tan and J. Shea, “An EM approach to multiple-access interference mitigation in asynchronous slow FHSS systems,” *IEEE Trans. Wireless Commun.*, vol. 7, no. 7, pp. 2661–2670, Jul. 2008.
- [14] F. A. Caceres and B. E. Swearingin, “An analysis of 802.11b and 802.16 technologies as part of the tactical internet,” Master’s Thesis, Naval Postgraduate School, 2005.
- [15] H. Zhao and Q. Wang, “On frequency hop synchronization in multipath Rayleigh fading,” *IEEE Trans. Veh. Technol.*, vol. 47, no. 3, pp. 1049–1065, Aug. 1998.

- [16] G. Huo and M.-S. Alouini, "Another look at the BER performance of FFH/BFSK with product combining over partial-band jammed Rayleigh-fading channels," *IEEE Trans. Veh. Technol.*, vol. 50, no. 5, pp. 1203–1215, Sep. 2001.
- [17] H. Joh and I. Ryoo, "A hybrid Wi-Fi P2P with bluetooth low energy for optimizing smart device's communication property," *Peer-to-Peer Netw. Appl.*, pp. 1–11, Apr. 2014.
- [18] J. Kongtep, "Adaptive frequency hopping scheme for wireless distribution system (WDS) in WLAN," *Int. J. Comput. Commun. Eng.*, vol. 1, no. 4, pp. 486–489, Nov. 2012.
- [19] S. K. Sarkar, T. G. Basavaraju, and C. Puttamadappa, *Ad Hoc Mobile Wireless Networks*. New York: Auerbach, 2008.
- [20] Y.-C. Chen, "Minimax and genralized likelihood ratio test for noncoherent FH/BFSK in band multitone jamming," in *Proc. IEEE Int. Conf. Acoust. Speech Signal Process. (ICASSP)*, Kyoto, Japan, 2012, pp. 3033–3036.
- [21] F. Yang and L.-L. Yang, "Frequency-hopping/ M -ary frequency-shift keying wireless sensor network monitoring multiple source events," in *Proc. IEEE Veh. Technol. Conf. (VTC)*, Yokohama, Japan, 2012, pp. 1–5.
- [22] Q. Zeng, H. Li, Z. Zhang, and D. Peng, "A frequency-hopping based communication infrastructure for wireless metering in smart grid," in *Proc. IEEE Annu. Conf. Sci. Syst. (CISS)*, Baltimore, MD, USA, 2011, pp. 1–6.
- [23] L. Xiao, H. Dai, and P. Ning, "Jamming-resistant collaborative broadcast using uncoordinated frequency hopping," *IEEE Trans. Inf. Forensics Security*, vol. 7, no. 1, pp. 297–309, Feb. 2012.

- [24] D. Angelosante, G. B. Giannakis, and N. D. Sidiropoulos, “Estimating multiple frequency-hopping signal parameters via sparse linear regression,” *IEEE Trans. Signal Process.*, vol. 58, no. 10, pp. 5044–5056, Oct. 2010.
- [25] A. Valyrakis, E. E. Tsakonas, N. D. Sidiropoulos, and A. Swami, “Stochastic modeling and particle filtering algorithms for tracking a frequency-hopped signal,” *IEEE Trans. Signal Process.*, vol. 57, no. 8, pp. 3108–3118, Jul. 2009.
- [26] D. J. Torrieri, “Mobile frequency-hopping CDMA systems,” *IEEE Trans. Commun.*, vol. 48, no. 8, pp. 1318–1327, Aug. 2000.
- [27] H. Olofsson, J. Naslund, and A. Skold, “Interference diversity gain in frequency hopping GSM,” in *Proc. IEEE Veh. Technol. Conf. (VTC)*, Chicago, IL, USA, 1995, pp. 102–106.
- [28] S. Chennakeshu, A. Hassan, J. Anderson, and B. Gudmundson, “Capacity analysis of a TDMA-based slow-frequency-hopped cellular system,” *IEEE Trans. Veh. Technol.*, vol. 45, no. 3, pp. 531–542, Aug 1996.
- [29] M. Chiani, “Error probability for block codes over channels with block interference,” *IEEE Trans. Inf. Theory*, vol. 44, no. 7, pp. 2998–3008, Nov 1998.
- [30] Y. T. Su and L.-D. Jeng, “Antijam capability analysis of RS-coded slow frequency-hopped systems,” *IEEE Trans. Commun.*, vol. 48, no. 2, pp. 270–281, Feb. 2000.
- [31] J. S. Lee, R. H. French, and L. E. Miller, “Probability of error analyses of a BFSK frequency-hopping system with diversity under partial-band jamming interference—Part I: Performance of square-law linear combining soft decision receivers,” *IEEE Trans. Commun.*, vol. 32, no. 12, pp. 1243–1250, Dec. 1984.

- [32] Q. Wang, T. A. Gulliver, V. K. Bhargava, and E. B. Felstead, "Performance of fast frequency hopped noncoherent MFSK with a fixed hop rate under worst case jamming," *IEEE Trans. Commun.*, vol. 38, no. 10, pp. 1786–1798, Oct. 1990.
- [33] B. Levitt, "FH/MFSK performance in multitone jamming," *IEEE J. Sel. Areas Commun.*, vol. 3, no. 5, pp. 627–643, Sep. 1985.
- [34] R. E. Ezers, E. B. Felstead, T. A. Gulliver, and J. S. Wight, "An analytical method for linear combining with application to FFH NCFSK receivers," *IEEE J. Sel. Areas Commun.*, vol. 11, no. 3, pp. 454–464, Apr. 1993.
- [35] J. Wang and C. Jiang, "Analytical study of FFH systems with square-law diversity combining in the presence of multitone interference," *IEEE Trans. Commun.*, vol. 48, no. 7, pp. 1188–1196, Jul. 2000.
- [36] C. M. Keller and M. B. Pursley, "Clipped diversity combining for channels with partial-band interference-Part I: Clipped-linear combining," *IEEE Trans. Commun.*, vol. 35, no. 12, pp. 1320–1328, Dec. 1987.
- [37] J. S. Lee, L. E. Miller, and Y. K. Kim, "Probability of error analyses of a BFSK frequency-hopping system with diversity under partial-band jamming interference-Part II: Performance of square-law nonlinear combining soft decision receivers," *IEEE Trans. Commun.*, vol. 32, no. 12, pp. 1243–1250, Dec. 1984.
- [38] J. S. Lee, L. E. Miller, and R. H. French, "The analyses of uncoded performances for certain ECCM receiver design strategies for multihops/symbol FH/MFSK waveforms," *IEEE J. Sel. Areas Commun.*, vol. SAC-3, no. 5, pp. 611–621, Sep. 1985.
- [39] K. C. Teh, A. C. Kot, and K. H. Li, "Partial-band jammer suppression in FFH spread-spectrum system using FFT," *IEEE Trans. Veh. Technol.*, vol. 48, no. 2, pp. 478–486, Mar. 1999.

- [40] J.-J. Chang and L.-S. Lee, "An exact performance analysis of the clipped diversity combining receiver for FH/MFSK systems against a band multitone jammer," *IEEE Trans. Commun.*, vol. 42, no. 2/3/4, pp. 700–710, Feb./Mar./Apr. 1994.
- [41] R. C. Robertson and T. T. Ha, "Error probabilities of fast frequency-hopped MFSK with noise-normalization combining in a fading channel with partial-band interference," *IEEE Trans. Commun.*, vol. 40, no. 2, pp. 404–412, Feb. 1992.
- [42] R. C. Robertson, H. Iwasaki, and M. Kragh, "Performance of a fast frequency-hopped noncoherent MFSK receiver with nonideal adaptive gain control," *IEEE Trans. Commun.*, vol. 46, no. 1, pp. 104–114, Jan. 1998.
- [43] L.-P. Zhu, Y. Yao, and Y.-S. Zhu, "Antijam performance of FFH/BFSK with noise-normalization combining in a Nakagami-m fading channel with partial-band interference," *IEEE Commun. Lett.*, vol. 10, no. 6, pp. 429–431, Jun. 2006.
- [44] L. E. Miller, J. S. Lee, and A. P. Kadrach, "Probability of error analyses of a BFSK frequency-hopping system with diversity under partial-band jamming interference-Part III: Performance of a square-law self-normalizing soft decision receiver," *IEEE Trans. Commun.*, vol. 34, no. 7, pp. 669–675, Jul. 1986.
- [45] R. C. Robertson and K. Y. Lee, "Performance of fast frequency-hopped MFSK receivers with linear and self-normalization combining in a Rician fading channel with partial-band interference," *IEEE J. Sel. Areas Commun.*, vol. 10, no. 4, pp. 731–741, May 1992.
- [46] C. Jiang and J. Wang, "Performance analysis of FFH/MFSK receivers with self-normalization combining in the presence of multitone jamming," *IEEE Trans. Veh. Technol.*, vol. 51, no. 5, pp. 1120–1127, Sep. 2002.

- [47] R. Viswanathan and K. Taghizadeh, "Diversity combining in FH/BFSK systems to combat partial band jamming," *IEEE Trans. Commun.*, vol. 36, no. 9, pp. 1062–1069, Sep. 1988.
- [48] S. Ahmed, L.-L. Yang, and L. Hanzo, "Mellin-transform-based performance analysis of FFH M -ary FSK using product combining for combatting partial-band noise jamming," *IEEE Trans. Veh. Technol.*, vol. 57, no. 5, pp. 2757–2765, Sep. 2008.
- [49] Y.-S. Shen and S.-L. Su, "Performance analysis of an FFH/BFSK receiver with ratio-statistic combining in a fading channel with multitone interference," *IEEE Trans. Commun.*, vol. 51, no. 10, pp. 1643–1648, Oct. 2003.
- [50] G. Li, Q. Wang, V. Bhargava, and L. Mason, "Maximum-likelihood diversity combining in partial-band noise," *IEEE Trans. Commun.*, vol. 46, no. 12, pp. 1569–1574, Dec. 1998.
- [51] K. C. Teh, A. C. Kot, and K. H. Li, "Performance study of a maximum-likelihood receiver for FFH/BFSK systems with multitone jamming," *IEEE Trans. Commun.*, vol. 47, no. 5, pp. 766–772, May 1999.
- [52] T.-M. Wu, "A suboptimal maximum-likelihood receiver for FFH/BFSK systems with multitone jamming over frequency-selective Rayleigh-fading channels," *IEEE Trans. Veh. Technol.*, vol. 57, no. 2, pp. 1316–1322, Mar. 2008.
- [53] Y. Han and K. C. Teh, "Performance study of suboptimum maximum-likelihood receivers for FFH/MFSK systems with multitone jamming over fading channels," *IEEE Trans. Veh. Technol.*, vol. 54, no. 1, pp. 82–90, Jan. 2005.
- [54] O. C. Yue, "Maximum likelihood combining for noncoherent and differentially coherent frequency-hopping multiple-access systems," *IEEE Trans. Inf. Theory*, vol. 28, no. 4, pp. 631–639, Jul. 1982.

- [55] T. C. Royster IV, "Current designs for fast frequency hopping with MFSK," in *Proc. IEEE Military Commun. Conf.*, Baltimore, MD, USA, 2011, pp. 543–548.
- [56] R. McEliece and W. Stark, "Channels with block interference," *IEEE Trans. Inf. Theory*, vol. 30, no. 1, pp. 44–53, Jan 1984.
- [57] C. L. Chang and T. M. Tu, "Performance analysis of FFH/BFSK product-combining receiver with partial-band jamming over independent Rician fading channels," *IEEE Trans. Wireless Commun.*, vol. 4, no. 6, pp. 2629–2635, Nov. 2005.
- [58] E. D. Spilchen and T. A. Gulliver, "Diversity combining for M -ary fast frequency hopping in a multiple-access Rayleigh fading channel," in *Proc. IEEE Military Commun. Conf.*, McLean, VA, USA, 1996, pp. 893–897.
- [59] S. Ahmed, L.-L. Yang, and L. Hanzo, "Diversity combining for fast frequency hopping multiple access systems subjected to Nakagami- m fading," in *Proc. IEE Int. Conf. 3G Beyond*, Washington, DC, USA, 2005, pp. 235–239.
- [60] R. C. Robertson and J. F. Sheltry, "Multiple tone interference of frequency-hopped noncoherent MFSK signals transmitted over Ricean fading channels," *IEEE Trans. Commun.*, vol. 44, no. 7, pp. 867–875, Jul. 1996.
- [61] C. M. Keller and M. B. Pursley, "Diversity combining for channels with fading and partial-band interference," *IEEE J. Sel. Areas Commun.*, vol. 5, no. 2, pp. 248–260, Feb. 1987.
- [62] C. M. Keller and M. B. Pursley, "Clipped diversity combining for channels with partial-band interference-Part II: Ratio-statistic combining," *IEEE Trans. Commun.*, vol. 37, no. 2, pp. 145–151, Feb. 1989.

- [63] T. A. Gulliver and H. H. S. Yep, "Gradient diversity combining for fast frequency-hopped MFSK spread-spectrum communications," in *Proc. IEEE Military Commun. Conf.*, McLean, VA, USA, 1996, pp. 75–79.
- [64] D. J. Torrieri, *Principles of Secure Communication Systems*, 2nd ed. Boston: Artech House, 1992.
- [65] Z. Yu, T. T. Tjhung, and C. C. Chai, "Independent multitone jamming of FH/MFSK in Rician channels," *IEEE Trans. Commun.*, vol. 49, no. 11, pp. 2006–2015, Nov. 2001.
- [66] R. Agusti, "On the performance analysis of asynchronous FH-SSMA communications," *IEEE Trans. Commun.*, vol. 31, no. 5, pp. 488–499, May 1989.
- [67] B. Solaiman, A. Glavieux, and A. Hillion, "Equal gain diversity improvement in fast frequency hopping spread spectrum multiple-access (FFH-SSMA) communications over Rayleigh fading channels," *IEEE J. Sel. Areas Commun.*, vol. 1, no. 1, pp. 140–147, Jan. 1989.
- [68] P. Yegani and C. McGillem, "FH-MFSK multiple-access communications systems performance in the factory environment," *IEEE Trans. Veh. Technol.*, vol. 42, no. 2, pp. 148–155, May 1993.
- [69] C. P. Hung and Y. T. Su, "Diversity combining considerations for incoherent frequency hopping multiple access systems," *IEEE J. Sel. Areas Commun.*, vol. 13, no. 2, pp. 333–344, Feb. 1995.
- [70] U. Svasti-Xuto, Q. Wang, and V. K. Bhargava, "Capacity of an FH-SSMA system in different fading environments," *IEEE Trans. Veh. Technol.*, vol. 47, no. 1, pp. 75–83, Feb. 1998.

- [71] J. Joo, H. Kang, K. Kim, and K. Kim, "Analysis of asynchronous FFH-MA systems with a hard-limited combining in Rayleigh fading," *IEEE Commun. Lett.*, vol. 9, no. 4, pp. 295–297, Apr. 2005.
- [72] J. Wang and L. B. Milstein, "CDMA overlay situations for microcellular mobile communications," *IEEE Trans. Commun.*, vol. 43, no. 2, pp. 603–614, 1995.
- [73] R. E. Ziemer and R. L. Peterson, *Introduction to Digital Communication*, 2nd ed. Upper Saddle River, New Jersey: Prentice Hall, 2001.
- [74] J. G. Proakis, *Digital Communications*, 4th ed. New York: McGraw-Hill, 2001.
- [75] C. Tepedelenlioglu, A. Abdi, and G. B. Giannakis, "The Ricean K factor: Estimation and performance analysis," *IEEE Trans. Wireless Commun.*, vol. 2, no. 4, pp. 799–810, Jul. 2003.
- [76] Y. Chen and N. C. Beaulieu, "Maximum likelihood estimation of the K factor in Ricean fading channels," *IEEE Commun. Lett.*, vol. 9, no. 12, pp. 1040–1042, Dec. 2005.
- [77] J. M. Holtzman, "On using perturbation analysis to do sensitivity analysis: Derivatives versus differences," *IEEE Trans. Automat. Contr.*, vol. 37, no. 2, pp. 243–247, Feb. 1992.
- [78] T.-M. Wu and P.-C. Hung, "Maximum-likelihood receivers for FFH/BFSK systems with multitone jamming over frequency-selective Rayleigh fading channels," in *Proc. IEEE International Conf. Commun.*, Glasgow, Scotland, UK, 2007, pp. 815–820.
- [79] T.-M. Wu and T.-H. Tsai, "Digital code tracking loops over frequency-selective fading channels," in *Proc. IEEE International Conf. Commun.*, Glasgow, Scotland, UK, 2007, pp. 5246–5251.

- [80] S. Ariyavisitakul, S. Yoshida, F. Ikegami, and T. Takeuchi, "Fractional-bit differential detection of MSK: A scheme to avoid outages due to frequency-selective fading," *IEEE Trans. Veh. Technol.*, vol. 36, no. 1, pp. 36–42, Feb. 1987.
- [81] I. Korn, "Differential phase shift keying in two-path Rayleigh channel with adjacent channel interference," *IEEE Trans. Veh. Technol.*, vol. 40, no. 2, pp. 461–471, May 1991.
- [82] W.-H. Sheen and G. L. Stuber, "Effects of multipath fading on delay-locked loops for spread spectrum systems," *IEEE Trans. Commun.*, vol. 42, no. 2/3/4, pp. 1947–1956, Feb./Mar./Apr. 1994.
- [83] N. C. Beaulieu, "Generation of correlated Rayleigh fading envelopes," *IEEE Commun. Lett.*, vol. 3, no. 6, pp. 172–174, Jun. 1999.
- [84] 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 multitone jamming and partial-band noise jamming," *IET Commun.*, vol. 4, no. 10, pp. 1153–1163, Jul. 2010.
- [85] S. Hinedi, M. Simon, and D. Raphaeli, "The performance of noncoherent orthogonal M -FSK in the presence of timing and frequency errors," *IEEE Trans Commun.*, vol. 43, no. 2/3/4, pp. 922–933, Feb./Mar./Apr. 1995.
- [86] Y. Wang, P. Ciblat, E. Serpedin, and P. Loubaton, "Performance analysis of a class of nondata-aided frequency offset and symbol timing estimators for flat-fading channels," *IEEE Trans. Signal Process.*, vol. 50, no. 9, pp. 2295–2305, Sep. 2002.
- [87] C.-D. Chung and A. Polydoros, "Parameter estimation of random FH signals using autocorrelation techniques," *IEEE Trans Commun.*, vol. 43, no. 2/3/4, pp. 1097–1106, Feb./Mar./Apr. 1995.

- [88] R. Schober, L. H. J. Lampe, W. H. Gerstacker, and S. Pasupathy, "Modulation diversity for frequency-selective fading channels," *IEEE Trans. Inf. Theory*, vol. 49, no. 9, pp. 2268–2276, Sep. 2003.
- [89] A. M. Tulino, G. Caire, S. Shamai, and S. Verdú, "Capacity of channels with frequency-selective and time-selective fading," *IEEE Trans. Inf. Theory*, vol. 56, no. 3, pp. 1187–1215, Mar 2010.
- [90] P. Hande, L. Tong, and A. Swami, "Multipath delay estimation for frequency hopping systems," *J. VLSI Signal Process.*, vol. 30, no. 1-3, pp. 163–178, 2002.
- [91] P. Hande, L. Tong, and A. Swami, "Channel estimation for frequency hopping systems," in *Proc. IEEE MILCOM*, vol. 2, Atlantic City, New Jersey, US, 1999, pp. 1323–1327.
- [92] E. Biglieri, J. Proakis, and S. Shamai, "Fading channels: Information-theoretic and communications aspects," *IEEE Trans. Inf. Theory*, vol. 44, pp. 2619–2692, Oct. 1998.
- [93] X. Gui and T. S. Ng, "Performance of asynchronous orthogonal multicarrier CDMA system in frequency selective fading channel," *IEEE Trans. Commun.*, vol. 47, pp. 1084–1091, Jul. 1999.
- [94] Y. Han and K. C. Teh, "Maximum-likelihood receiver with side information for asynchronous FFH-MA/MFSK systems over Rayleigh fading channels," *IEEE Commun. Lett.*, vol. 10, pp. 435–437, Jun. 2006.
- [95] I. S. Gradshteyn and I. M. Ryzhik, *Tables of Integrals, Series and Products*, 7th ed. New York: Academic, 2007.

- [96] G. A. Tsihrintzis and C. L. Nikias, “Performance of optimum and suboptimum receivers in the presence of impulsive noise modeled as an alpha-stable process,” *IEEE Trans. Commun.*, vol. 43, no. 2/3/4, pp. 904–914, Feb./Mar./Apr. 1995.
- [97] H.-G. Ryu, Y. Li, and J.-S. Park, “Effects of frequency instability caused by phase noise on the performance of the fast FH communication system,” *IEEE Trans. Veh. Technol.*, vol. 53, no. 5, pp. 1626–1632, Sep. 2004.
- [98] R. N. McDonough and A. D. Whalen, *Detection of Signals in Noise*, 2nd ed. New York: Academic Press, 1995.

Appendix A

In this Appendix, we provide the detailed derivation for the suboptimum ML receiver structure as shown in (4.40). From (4.39), the pdf expressions of r_{ml} for the signal and non-signal branches can be obtained as

$$p_{r_{ml}}(r_{ml}|q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \boldsymbol{\psi}_S, \boldsymbol{\psi}_\mu, S) = \frac{1}{2\sigma_{ml,S}^2} \exp\left(-\frac{r_{ml} + A_{ml,S}^2}{2\sigma_{ml,S}^2}\right) \times I_0\left(\frac{A_{ml,S}\sqrt{r_{ml}}}{\sigma_{ml,S}^2}\right) U(r_{ml}) \quad (\text{A.1})$$

and

$$p_{r_{ml}}(r_{ml}|q_{\alpha l}, q'_{\beta l}, q_l, q'_l, \boldsymbol{\psi}_\mu, NS) = \frac{1}{2\sigma_{ml,NS}^2} \exp\left(-\frac{r_{ml} + A_{ml,NS}^2}{2\sigma_{ml,NS}^2}\right) \times I_0\left(\frac{A_{ml,NS}\sqrt{r_{ml}}}{\sigma_{ml,NS}^2}\right) U(r_{ml}), \quad (\text{A.2})$$

respectively, where

$$\begin{aligned} \sigma_{ml,S}^2 = & \sigma_{S_1}^2 (1 - \chi)^2 \text{sinc}^2[(1 - \chi)\xi] \left[1 + 2\lambda_S \sqrt{\varepsilon_S} \cos\left(\frac{2\pi\tau_S\xi}{T_h} + \theta_{S_1,l} - \theta_{S_2,l}\right) + \varepsilon_S \right] \\ & + q_{\alpha l} \sigma_{\mu_1}^2 u_3^2 (1 + \varepsilon_\mu + u_4) + q'_{\beta l} \sigma_{\mu_1}^2 u_5^2 (1 + \varepsilon_\mu + u_6) + \frac{[q_l(1 - \chi) + q'_l\chi]^2}{2} \sigma_P^2 + \frac{\sigma_\omega^2}{2}, \end{aligned} \quad (\text{A.3})$$

$$\begin{aligned}\sigma_{ml,NS}^2 &= q_{\alpha l} \sigma_{\mu_1}^2 u_3^2 (1 + \varepsilon_\mu + u_4) + q'_{\beta l} \sigma_{\mu_1}^2 u_5^2 (1 + \varepsilon_\mu + u_6) \\ &\quad + \frac{[q_l (1 - \chi) + q'_l \chi]^2}{2} \sigma_P^2 + \frac{\sigma_\omega^2}{2},\end{aligned}\tag{A.4}$$

$$\begin{aligned}A_{ml,S}^2 &= A_S^2 (1 - \chi)^2 \text{sinc}^2 [(1 - \chi)\xi] + q_{\alpha l} A_\mu^2 u_3^2 + q'_{\beta l} A_\mu^2 u_5^2 + 2q_{\alpha l} A_S A_\mu (1 - \chi) \\ &\quad \times \text{sinc} [(1 - \chi)\xi] u_3 \cos (\Phi_1 - \Phi_2) + 2q'_{\beta l} A_S A_\mu (1 - \chi) \text{sinc} [(1 - \chi)\xi] \\ &\quad \times u_5 \cos (\Phi_1 - \Phi_3) + 2q_{\alpha l} q'_{\beta l} A_\mu^2 u_3 u_5 \cos (\Phi_2 - \Phi_3),\end{aligned}\tag{A.5}$$

and

$$A_{ml,NS}^2 = q_{\alpha l} A_\mu^2 u_3^2 + q'_{\beta l} A_\mu^2 u_5^2 + 2q_{\alpha l} q'_{\beta l} A_\mu^2 u_3 u_5 \cos (\Phi_2 - \Phi_3).\tag{A.6}$$

By substituting (A.1) and (A.2) into (4.34), z_{ml} can be expressed as

$$\begin{aligned}z_{ml} &= \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \ln \left[I_0 \left(\frac{A_{ml,S} \sqrt{r_{ml}}}{\sigma_{ml,S}^2} \right) \right] \\ &\quad - \ln \left[I_0 \left(\frac{A_{ml,NS} \sqrt{r_{ml}}}{\sigma_{ml,NS}^2} \right) \right] + E_{ml},\end{aligned}\tag{A.7}$$

where

$$E_{ml} = \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2} + \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right).\tag{A.8}$$

Note that the computation of $\ln [I_0(\cdot)]$ is nonlinear in nature [98]. Following [52], we adopt the approximation $\ln [I_0(x)] = x$, when $x > 4$, to simplify (A.7) to

$$z_{ml} = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \left(\frac{A_{ml,S}}{\sigma_{ml,S}^2} - \frac{A_{ml,NS}}{\sigma_{ml,NS}^2} \right) \sqrt{r_{ml}} + E_{ml}. \quad (\text{A.9})$$

Following that, we further simplify (A.9) and obtain the suboptimum ML receiver structure as shown in (4.40).

Appendix B

In this Appendix, the detailed derivation of (5.42) is presented. The function $F(\cdot)$ of the ML receiver can be expressed as [78]

$$z_{ml} = \ln \left(\frac{p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m = f_n)}{p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m \neq f_n)} \right), \quad (\text{B.1})$$

where $p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m = f_n)$ represents the pdf of the square-law detector output when the desired signal is present and $p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m \neq f_n)$ denotes the corresponding pdf when the signal is not present. From (5.41), we have

$$\begin{aligned} p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m = f_n) &= \frac{1}{2\sigma_{ml,S}^2} \exp \left(-\frac{r_{ml} + A_{ml,S}^2}{2\sigma_{ml,S}^2} \right) \\ &\times I_0 \left(\frac{A_{ml,S} \sqrt{r_{ml}}}{\sigma_{ml,S}^2} \right) U(r_{ml}) \end{aligned} \quad (\text{B.2})$$

and

$$\begin{aligned} p_{r_{ml}}(r_{ml}|\boldsymbol{\theta}_l, \boldsymbol{\phi}_l, \mathbf{f}_l, f_m \neq f_n) &= \frac{1}{2\sigma_{ml,NS}^2} \exp \left(-\frac{r_{ml} + A_{ml,NS}^2}{2\sigma_{ml,NS}^2} \right) \\ &\times I_0 \left(\frac{A_{ml,NS} \sqrt{r_{ml}}}{\sigma_{ml,NS}^2} \right) U(r_{ml}), \end{aligned} \quad (\text{B.3})$$

where

$$\begin{aligned}
A_{ml,S}^2 = & A_s^2 + \sum_{k=1}^{K_l} A_k^2 \delta(f_{k,l} - f_m) + \sum_{k=1}^{K_l} 2A_s A_k \cos \phi_{sk,l} \delta(f_{k,l} - f_m) \\
& + \sum_{i=1}^{K_l} \sum_{j=1(i \neq j)}^{K_l} A_i A_j \cos \phi_{ij,l} \delta(f_{i,l} - f_m) \delta(f_{j,l} - f_m), \tag{B.4}
\end{aligned}$$

$$A_{ml,NS}^2 = \sum_{k=1}^{K_l} A_k^2 \delta(f_{k,l} - f_m) + \sum_{i=1}^{K_l} \sum_{j=1(i \neq j)}^{K_l} A_i A_j \cos \phi_{ij,l} \delta(f_{i,l} - f_m) \delta(f_{j,l} - f_m), \tag{B.5}$$

$$\sigma_{ml,S}^2 = \sigma_{s_1}^2 + 2\sigma_{s_1}^2 \lambda_s \sqrt{\epsilon_s} \cos \theta_{s,l} + \sigma_{s_2}^2 + \sum_{k=1}^{K_l} (\sigma_{k_1}^2 + 2\sigma_{k_1}^2 \lambda_k \sqrt{\epsilon_k} \cos \theta_{k,l} + \sigma_{k_2}^2) \delta(f_{k,l} - f_m) + \frac{\sigma_\omega^2}{2}, \tag{B.6}$$

and

$$\sigma_{ml,NS}^2 = \sum_{k=1}^{K_l} (\sigma_{k_1}^2 + 2\sigma_{k_1}^2 \lambda_k \sqrt{\epsilon_k} \cos \theta_{k,l} + \sigma_{k_2}^2) \delta(f_{k,l} - f_m) + \frac{\sigma_\omega^2}{2}. \tag{B.7}$$

Following that, the ML decision variable z_{ml} can be obtained by substituting (B.2) and (B.3) into (B.1) as

$$\begin{aligned}
z_{ml} = & \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \ln \left[I_0 \left(\frac{A_{ml,S} \sqrt{r_{ml}}}{\sigma_{ml,S}^2} \right) \right] - \ln \left[I_0 \left(\frac{A_{ml,NS} \sqrt{r_{ml}}}{\sigma_{ml,NS}^2} \right) \right] \\
& + \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2} + \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right). \tag{B.8}
\end{aligned}$$

Since the computation of $\ln(I_0(\cdot))$ is nonlinear in nature, by following [52], we adopt

the approximation $\ln I_0(x) = x$, for $x > 4$, to simplify (B.8) to

$$z_{ml} = \left(\frac{1}{2\sigma_{ml,NS}^2} - \frac{1}{2\sigma_{ml,S}^2} \right) r_{ml} + \left(\frac{A_{ml,S}}{\sigma_{ml,S}^2} - \frac{A_{ml,NS}}{\sigma_{ml,NS}^2} \right) \sqrt{r_{ml}} + C_{ml} \quad (\text{B.9})$$

where

$$C_{ml} = \frac{A_{ml,NS}^2}{2\sigma_{ml,NS}^2} - \frac{A_{ml,S}^2}{2\sigma_{ml,S}^2} + \ln \left(\frac{\sigma_{ml,NS}^2}{\sigma_{ml,S}^2} \right). \quad (\text{B.10})$$

The above approximation is applicable when the signal-to-noise ratio is high ($\text{SNR} > 6$ dB), which is of practical use [51]. In order to derive the pdf expressions of the suboptimum ML receiver output z_{ml} from (B.2) and (B.3), we perform some mathematical manipulations to simplify (B.9) to

$$z_{ml} = D_{ml} (\sqrt{r_{ml}} + B_{ml})^2 + E_{ml}. \quad (\text{B.11})$$