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Symbol-by-symbol detection scheme for IEEE 802.15.4 MPSK receiver

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Abstract: This paper proposes a symbol-by-symbol detection scheme for IEEE 802.15.4 M-ary phase shift keying (MPSK) receiver. The chip sample is first processed by the correlation operation, and then the traditional symbol-by-symbol detection scheme without carrier frequency offset is directly configured. Simulations are developed to verify the performance. The simulation results show that the detection performance of the proposed scheme can satisfy the requirements, and the complexity is also simple.
Keywords: Symbol-by-symbol detection; MPSK receiver; IEEE 802.15.4; Wireless sensor networks.

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1. INTRODUCTION

With the widespread application of new generation information and communication technologies such as the Internet of Things, cloud computing, and big data, smart city projects have developed rapidly in recent years. It has penetrated into all aspects of people's lives, greatly satisfying modern people's convenience, The pursuit of fast and high-quality life [1-3]. The application of smart cities is inseparable from the support of information and data reliability. In this work, attention is paid to the reliability of IEEE 802.15.4 *M*-ary phase shift keying (MPSK) receiver [4]. Specifically, a noncoherent detection scheme for MPSK physical layer (PHY) is studied. At present, many documents study binary phase shift keying (BPSK) and quadrature phase shift keying (QPSK) [5,6], and our focus is on MPSK, which is due to the higher data transmission rate of MPSK.

MPSK PHY is required in the 315/430/780MHz frequency band [4,7]. In the IEEE 802.15.4 MPSK receiver, generally speaking, since carrier synchronization can be well implemented at the receiving end [8], the coherent detection scheme has good performance. However, it is not suitable for battery-powered wireless sensor networks. This is because its performance depends entirely on the perfect estimation of the carrier phase offset (CPO), which is obviously a highly complex and energy-intensive process [9]. Therefore, the incoherent detection scheme that does not require carrier synchronization at the receiving end is an attractive choice in high-performance wireless sensor networks (WSNs) [10].

In the literature, many non-coherent detection schemes for 802.15.4 MPSK receivers have been proposed. An incoherent sequence detection scheme is recommended in [11], which can withstand carrier frequency offset (CFO) of up to 5%. However, since the complexity of the receiver is closely related to the transmission time, its actual performance is limited by the front end of the matched

filter. In the Internet of Things communication, due to Doppler shift and/or oscillator changes of the transmitter and receiver, there is inevitably a deviation between the transmitted and received carrier signals. J. Y. Oh et al. pointed out that the initial phase mismatch and CFO will cause a significant performance degradation. Therefore, it is necessary to implement CFO estimation and compensation on the receiver before demodulation [12]. Specifically, a phase locked loop (PLL) method for compensating CFO is considered in [13], where the synthesizer is composed of analog devices that are highly sensitive to temperature and humidity. However, it is difficult to adjust the corresponding frequency of the analog device, which is not applicable in actual operation. Considering the deployment of appropriate auxiliary preamble sequences to deal with unknown frequency offsets, J. H. Do et al. proposed a coherent detection-based scheme with good performance [14]. However, this scheme still has a high degree of complexity, which is mainly due to multiplication and trigonometric operations. In addition, in [15] and [16], cross-correlation and auto-correlation operations are used to synchronize the receiver, which is actually a complicated and energy-consuming process. Therefore, there is an urgent need to develop a compatible and low-complexity CFO estimation scheme for the MPSK incoherent detection scheme in IEEE 802.15.4 WSN.

In this research, we propose a detection scheme suitable for IEEE 802.15.4 MPSK receiver. The main contributions are as follows. Previous studies assumed that the amplitude of the signal is not affected by the channel, that is, fading is not considered. In this article, we have considered a more general situation and studied receivers in slow fading channels and purely additive white Gaussian noise channels. The detection characteristics of the receiver are studied from many aspects.

The rest of this article is organized as follows. Section 2 focuses on the signal model of the fading channel. Section 3 describes detection scheme based on traditional coherence. Section 4 discusses the numerical results. Finally, some conclusions are provided in Section 5.

2. SIGNAL MODEL

According to the IEEE802.15.4c protocol [7], the MPSK physical layer uses 16-element cyclic shift element orthogonal modulation technology. The specific data modulation process is recommended in [7]. In particular, the analysis in this work is all based on the fading channel (16, 4)-DSSS system, and each data symbol is extended to 16 chips. For more details on the mapping rules, please refer to Table II in [17].

The goal is to accurately recover the single sinusoidal signal observed in the fading channel. Assuming ideal carrier synchronization at the receiver on the fading channel, without losing generality, but also to simplify the discussion, follow the signal model introduced in [16] with some minor changes. Specifically, for the xth symbol period $E[x]$, the complex base-band receive chip sequence is as follows:

$$r_{x,m} = h_{x,m} s_{x,m} e^{j\left(\omega_{x,m} m N \mathrm{T}_c + \theta_{x,m}\right)} + \eta_{x,m}, 1 \le m \le M \tag{1}$$

Where $h_{x,m}$ represents multiplicative decline, $s_{x,m}$ represents the m MPSK modulation chip in the xth symbol interval, j is an imaginary unit, $\omega_{x,m}=2\pi f_{x,m}$ represents the carrier frequency offset in radians, $f_{x,m}$ represents the residual carrier frequency offset in Hz, $\theta_{x,m}$ represents the carrier phase

offset in radians, T_c represents the spreading chip period, $\eta_{x,m}$ is a discrete, cyclic symmetric, complex Gaussian random variable with zero mean and variance $\sigma_{x,m}^2$, $M=16$ represents the length of the pseudo-random sequence [7].

We assume that the CFO, CPO, and noise terms are unknown and random at the receiver, but are constant for some specified time period. In other words, we make a piece-wise constant approximation to these unknown parameters [17], which is $h_{x,m}=h, \omega_{x,m}=\omega, \theta_{x,m}=\theta$. In addition, the receiver does not have any auxiliary information about the CPO, that is to say, θ is evenly distributed in the interval $(-\pi,\pi)$. The CFO f follows a symmetrical triangular distribution [17]. The multiplicative noise term h follows Rayleigh distribution. Finally, we believe that h, f and θ are statistically independent from $\eta_{x,m}$. Unless otherwise specified, this article assumes that all interference parameters of the receiver are unknown.

3. SYMBOL-BY-SYMBOL DETECTION SCHEME

The concrete realization process of the proposed scheme is as follows.

First, the delay difference calculation is performed on the complex base-band reception sequence and the PN sequence corresponding to the preamble, respectively. We can get the complex expressions $P_{x,m-N}$ and $Q_{0,m-N}$ as follows:

$$P_{x,m-N}=r_{x,m}r^*{}_{x,m-N}=|h|^2 s_{x,m}s^*{}_{x,m-N}e^{j\omega NT_c}+\eta(1), 1\le x\le J_1, N+1\le m\le L_1 \tag{2}$$

$$Q_{0,m-N}=|h|^2 s_{0,m}s^*_{0,m-N}, N+1\le m\le L_2 \tag{3}$$

Here, J_1 represents the symbol length of the preamble, and $1\le J_1\le J$, where $J=8$ is the preamble length threshold. L_2 represents the sample number, and $N+1\le L_2\le K$. $\{s_{0,k}\}$ represents the complex form spreading sequence of the preamble "0000". $\eta(1)$ represents the integrated additive noise component. Note that unless otherwise stated, this article will consider the maximum preamble length, i.e. $J_1=8$.

Using the correlation between the preamble and the PN sequence, the frequency offset estimate is obtained as:

$$\begin{aligned} f_{est} &= \frac{1}{J(L_2-N)}\sum_{x=1}^{J}\sum_{m=N+1}^{L_2}P_{x,m-N}Q^*_{0,m-N} \\ &= \frac{1}{J(L_2-N)}\sum_{x=1}^{J}\sum_{m=N+1}^{L_2}\left[|h|^2 s_{x,m}s^*{}_{x,m-N}e^{j\omega NT_c}+\eta(1)\right]\left[|h|^2 s_{0,m}s^*_{0,m-N}\right]^* \\ &= \frac{1}{J(L_2-N)}\sum_{x=1}^{J}\sum_{m=N+1}^{L_2}\left[|h|^4 s_{x,m}s^*{}_{x,m-N}s^*{}_{0,m}s_{0,m-N}e^{j\omega NT_c}+\eta(1)s^*{}_{0,m}s_{0,m-N}\right] \\ &= h_1e^{j\omega NT_c}+\eta(2) \end{aligned} \tag{4}$$

Assume that the received sampling power is 1. The superscript $*$ in formula (8) represents the complex conjugate operation, and h_1 represents the comprehensive multiplicative noise. $\eta(2)$ is the comprehensive additive noise term.

Then use the frequency offset estimator f_{est}^{*} in equation (8) to perform post-compensation operation on the detection data to obtain the output of the combiner of the detection scheme as follows:

$$V_{x,y} = \mathrm{Re}\left(\sum_{N=1}^{3} \sum_{m=N+1}^{L_1} P_{x,m-N} Q_{y,m-N}^{*} f_{est}^{*} \right) \tag{5}$$

Finally, select the maximum value of 16 symbols as the output of the maximum selector:

$$V_x = \arg\max_{0 \le y \le 15} \left\{ V_{x,y} \right\} \tag{6}$$

After de-mapping, the final bit output data $E[m]$ can be obtained. Based on the traditional coherent detection scheme, the correlation between the preamble and the PN sequence is used to obtain the frequency offset estimator, and the post-compensation operation is performed on the detection data to improve the traditional non-coherent detection scheme. However, it is still a high-complexity receiver, which is mainly due to the complexity of the delay differential operation and related operations of the estimator f_{est}. After a simple comparison, we found that the main difference between this work and [14] is that the transmitter of [14] does not consider the effects of multiplicative noise and phase offset on detection.

4. NUMERICAL RESULTS AND DISCUSSION

In this section, we evaluate the bit error rate (BER), symbol error rate (SER), and packet error rate (PER) performance of various detection schemes. Note that in the simulation, the PPDU is set to 22 bytes. For different detectors, the transmitter will repeatedly send random data packets to the detector for detection until enough error data packets are collected. We choose the maximum 780MHz frequency band as the carrier frequency, 786MHz. The detailed simulation parameters in this work are shown in Table 1.

Table 1: Parameters used in the simulation

Parameter	Detailed description
Channel condition	Fading channel or Complex AWGN channel
Power of the complex noise	1/SNR
Payload length of the PPDU (bits)	176
Carrier frequency (MHz)	786
Chip rate (chip/s)	1×106
CPO θ (rads)	Uniform distribution in (-π, π)
CFO f (ppm)	Symmetric triangular distribution in (-80, 80)
Timing synchronization	Perfect
Detection scheme	Noncoherent
PN length K	16
preamble length J_1	Maximum

In the fading channel, we compared the BER, SER and PER performances of the proposed scheme and the optimal noncoherent detection scheme. In the transmission of each data packet, CFO f was considered to be symmetrical triangular distribution from -80 to 80ppm, and CPO θ follows an uniform distribution in interval (-π, π). As shown in Figure 1, with the increase of SNR, the

performance of BER, SER and PER of our proposed scheme also improves. Compared with the optimal noncoherent scheme, the complexity of the proposed scheme is greatly reduced, while there is a small performance degradation. That is to say, in the normalized slow fading channel, the complexity of our detector is reduced, but no large performance degradation is observed.

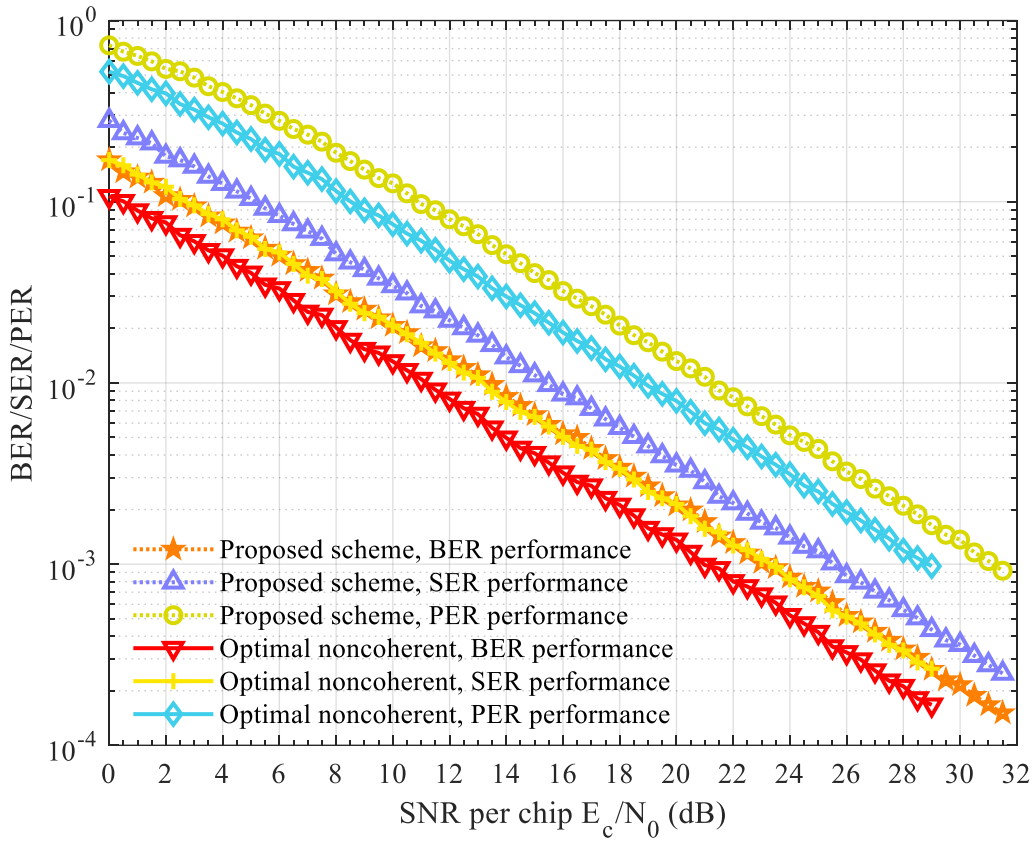

Fig. 1 Detection performance under fading channel.

5. CONCLUSIONS AND FUTURE WORK

For the IEEE 802.15.4 MPSK receiver, we proposed a friendly and efficient noncoherent detection scheme to estimate and compensate CFO by using the preamble-assisted idea. Experimental results show that, compared with the optimal noncoherent scheme, our proposed low-complexity receiver in implementation does not sacrifice much in terms of detection performance, which can meet the requirements of IEEE 802.15.4 WSN. Therefore, our improved detection scheme is more attractive to MSDD solution of choice for WSNs, whose representative application field is new Smart City.

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