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Joint time-frequency domain cyclostationarity-based approach to blind estimation of OFDM transmission parameters

Zhuo Sun*, Ruzhe Liu and Wenbo Wang

Abstract

Blind estimation of the transmission parameters of orthogonal frequency division multiplexing (OFDM) signal is an important issue for various civilian and military applications. This paper presents a new cyclostationarity-based approach to blind estimation of OFDM signal parameters. Analytical expression for extracting transmission parameters from second-order cyclic cumulant (CC) of OFDM signal is derived first, with the consideration of the effects of time dispersion and consistent estimator errors on CC. The approach exploits the cyclostationarity of OFDM both in time delay domain and cyclic frequency domain; a statistic hypothesis testing framework was formulated to determine the threshold for detecting the presence of cycles of OFDM signal. The simulation results reveal that the proposed approach is robust for both previous synchronization and channel condition concurrently.

Keywords: OFDM, Blind parameter estimation, Cyclostationarity, Second-order cyclic cumulant

1. Introduction

Blind estimation of communication signal parameters plays a key role in various civilian and military applications, such as cognitive radio (CR) and noncooperative communication [1,2]. The cognitive radio aims at improving the spectrum utilization by allowing the secondary user to sense and use the licensed spectrum, which needs to have the intelligence of dynamically changing its parameters. Therefore, it is necessary for the blind parameter estimation for CR to efficiently demodulate the signals in hostile wireless channel. In noncooperative communication, hostile communication signal must be detected, estimated, and recognized by its transmission parameters, which are vital for decisions involving electronic warfare operations and other tactical actions. Over the recent decades, orthogonal frequency division multiplexing (OFDM) has been widely employed in many wireless communication systems, owing to the advantages of high spectral efficiency and robustness against channel fading. OFDM is also a good alternative transmission scheme for underwater communication that both remedies the problem of ISI and provides low complexity solutions. For the OFDM signal, the recognized parameters include the symbol periodic, subcarrier spacing, subcarrier number, and cyclic prefix (CP) length.

The cyclostationarity properties of OFDM signal was widely exploited for various purposes: signal detection, classification [3-6], synchronization [7-9], and parameter estimation [10-13], in which the blind estimation of OFDM parameters have been investigated only in recent years, and these literatures generally achieved the estimation parameters by finding the cycles present in the cyclic statistics of OFDM signal. The algorithm proposed in [10] is based on the signal empirical distribution function. In [11], the cyclic correlation of OFDM signal is used to estimate the sampling frequency and CP length. A blind parameter estimation algorithm exploits the cyclostationary property of OFDM while considering the effect of time-dispersive channel [12]. The work in [13] pursues more accurate blindly estimated results by processing the candidate set of cyclic frequencies with average idea. However, in all of these literatures, the detection procedure is overly empirical and not rigorous for estimation accuracy, e.g., by checking the cycle presence based on a predefined threshold. In addition, the exploitation of cyclic statistics only applied in single domain (time or frequency) does not make full use of the cyclostationarity properties of OFDM signal [14].

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This paper presents a cyclostationarity-based approach to the blind estimation of OFDM transmission parameters, which derives the estimated parameters from both time (delay) and frequency domain using the second-order cyclic cumulant (CC) in a similar way as [14]. It is shown that detection and estimation in two domains can improve the accuracy probability of estimating evidently. To ensure the preciseness of the approach, a statistic hypothesis testing was developed to determine the threshold for detecting the presence of cycles of OFDM signal, by which the theoretical relation between the threshold value and constant false alarm rate (CFAR) was derived. Another contribution of this paper is that it derived the analytical expression of second-order CC of OFDM signal in both additive white Gaussian noise (AWGN) and multipath channel, and consequently, the condition of estimating the parameters from second-order CC are achieved with the consideration of the effects of time dispersion and consistent estimator errors.

The paper is organized as follows: we first give the OFDM signal model in AWGN and multipath channels in Section 2. In Section 3, the analytical expression of second-order CC of OFDM signal and the condition of estimating the parameters are derived. The proposed cyclostationarity-based approach to the blind estimation of OFDM transmission parameters is given in Section 4. Simulation results are presented in Section 5, and finally, conclusions are drawn in Section 6.

2. Signal model

The OFDM signal is transmitted over wireless channel and impacted by additive white Gaussian noise, timing offset, and carrier frequency offset; the received continuous time baseband signal can be represented as

$$r(t) = ae^{j\theta}e^{j2\pi\Delta f_c t}\sum_{k=0}^{K-1}\sum_{l=-\infty}^{+\infty} s_{k,l}e^{j2\pi k\Delta f_k(t-lT-\varepsilon T)}g(t-lT-\varepsilon T) + w(t), \tag{1}$$

where a is the power factor, K is the number of subcarriers of the OFDM signal, and Δf_K is the spacing of two adjacent subcarriers. T is the OFDM symbol period, $T = T_{cp} + T_u, T_u = \Delta f_K$ is the useful symbol duration, and T_{cp} is the cyclic prefix duration. Δf_c is the carrier frequency offset, and ε is the time offset introduced by channel and the inconsistent between transmitter and receiver. $g(t)$ is the resulting pulse shape with the transmitting root-raised cosine windowing function and the receive low-pass filter. The independent identically distributed symbol ($s_{k,l}$) is drawn from a M-ary PSK or M-ary QAM modulated signal constellation and transmitted on the kth subcarriers over the l symbol period, with the zero mean and the variance σ^2. As a result, there is

$$E(s_{k.l}) = E\left(s^*_{k,l}\right) = E\left(s_{k.l}s_{k'l'}\right) = 0, E\left(s_{k.l}s^*_{k,l}\right) = \sigma_s^2\delta\left[k-k'\right]\delta\left[l-l'\right]. \tag{2}$$

Under the multipath fading channel, the OFDM signal at the receiver consists of multiple branches accompanied with channel coefficient $h(\xi_m)$ at time ξ_m, $m = 1,\ldots\ldots M$:

$$r(t) = ae^{j\theta}e^{j2\pi\Delta f_c t}\sum_{m=0}^{M}\sum_{k=0}^{K-1}\sum_{l=-\infty}^{+\infty} s_{k,l}e^{j2\pi k\Delta f_k(t-\xi_m-lT-\varepsilon T)}h(\xi_m) \times g(t-\xi_m-lT-\varepsilon T) + w(t). \tag{3}$$

A discrete-time baseband OFDM signal, $r(u)$, can be obtained by oversampling $r(t)$ at a sampling frequency $f_s = \rho K/T_u$, with ρ as the oversampling factor per subcarrier.

3. Second-order cyclostationarity of OFDM signal

3.1. Derivation for second-order cyclic cumulant

The cyclostationarity of OFDM signals has been proved in previous literature [7], which could be introduced by CP, symbol, or pilot. In the following, we derived the analytical expressions for the second-order cyclostationarity features of OFDM signal, which is specifically used by the blind parameter estimation approach proposed in Section 3. Basically, the higher-order cumulant needs the more computation complexity; hence, we only take the second-order cyclic cumulant of OFDM signal as our reference target, on which transmission parameter extraction executes under both AWGN and multipath channels.

According to the definition of the cumulant in [15,16], the second-order time-varying cumulant of OFDM signal in (1) is expressed as

$$\begin{aligned} c_r(t;\tau) &= E[r(t+\tau/2)r^*(t-\tau/2)] \\ &= a^2 c_{s,2,1}e^{-j2\pi\Delta f_c\tau}\sum_{k=0}^{K-1}e^{j2\pi k\Delta f_k\tau} \\ &\quad \times g(t)g^*(t+\tau)\otimes\sum_{l=-\infty}^{\infty}\delta(t-lT-\varepsilon T) + c_w(t;\tau), \end{aligned} \tag{4}$$

where $\otimes$ represents convolution operation and $\delta(t)$ is the Dirac delta function. $c_{s,2,1}$ represents the second-order cumulant of the signal constellation, $c_{s,2,1} = E[s_{k,l}s_{k,l+\tau/T}]$. $c_w(t;\tau)$ is the second-order cumulant of Gauss noise. Due to the wide-sense stationary of the noise, the cumulant of the noise does not depend on time; for simplicity, we omit the noise term in the following analysis.

When applying the Fourier transform to the second-order time-varying cumulant in (4), we can obtain

$$\Im\{c_r(t;\tau)\} = a^2 c_{s,2,1} e^{-j2\pi\Delta f_c \tau} \int_{-\infty}^{\infty} [\sum_{k=0}^{K-1} e^{j2\pi k\Delta f_k \tau} \times g(t) g^*(t+\tau) \otimes \sum_{l=-\infty}^{\infty} \delta(t-lT-\varepsilon T)] e^{-j2\pi\beta t} dt. \tag{5}$$

Using the property of Fourier transform for the Dirac delta function: $\Im\left\{\sum_l \delta(t-lT)\right\} = T^{-1}\sum_l \delta\left(\beta - lT^{-1}\right)$, we can rewrite (5) into

$$\Im\{c_r(t;\tau)\} = a^2 c_{s,2,1} T^{-1} e^{-j2\pi\Delta f_c \tau} e^{-j2\pi\beta\varepsilon T} \sum_{k=0}^{K-1} e^{j2\pi k\Delta f_k \tau} \int_{-\infty}^{\infty} g(t) g^*(t+\tau) e^{-j2\pi\beta t} dt \times \sum_l \delta\left(\beta - lT^{-1}\right). \tag{6}$$

On the other hand, the second-order CC is defined by Fourier series expansion of the second-order time-varying cumulant when the latter is viewed as an almost periodic function of time [15,16]:

$$c_r(t;\tau) = \sum_{\beta\in\kappa_{2,1}} C_r(\beta;\tau) e^{j2\pi\beta t}. \tag{7}$$

According to (7), we can derive the second-order CC at cyclic frequency (CF) β and delay τ by taking the inverse Fourier transform on (6):

$$C_r(\beta;\tau) = a^2 c_{s,2,1} T^{-1} e^{-j2\pi\Delta f_c \tau} e^{-j2\pi\beta\varepsilon T} \sum_{k=0}^{K-1} e^{j2\pi k\Delta f_k \tau} \int_{-\infty}^{\infty} g(t) g^*(t+\tau) e^{-j2\pi\beta t} dt. \tag{8}$$

The set of CFs is given by $\kappa_{2,1}^{\mathrm{OFDM}} = \{\beta | C_r(\beta,\tau) \neq 0\}$, and it can be proved by [4]:

$$\kappa_{2,1}^{\mathrm{OFDM}} = \left\{\beta : \beta = lT^{-1}, l = 1,2,3...\right\}. \tag{9}$$

For a discrete-time OFDM signal $r(u)$, its second-order CC is given by

$$C_r(\beta;\tau) = C_r\left(\beta f_s; \tau f_s^{-1}\right). \tag{10}$$

By substituting (8) into (10), the second-order CC for the discrete-time OFDM signal at CF β and delay τ and the set of CFs can be respectively obtained:

$$C_r(\beta;\tau) = a^2 c_{s,2,1} D^{-1} e^{-j2\pi\beta\varepsilon D} e^{-j\frac{2\pi}{\rho K}\Delta f_c T_u \tau} \times \Gamma_K(\tau) \Lambda(\tau) \tag{11}$$

and

$$\kappa_{2,1}^{\mathrm{OFDM}} = \{\beta \in [-1/2, 1/2) | \beta = lD^{-1}, l = 1,2,3...\}, \tag{12}$$

where $\Gamma_K(\tau) = e^{j\pi(K-1)\tau/\rho K} \sin(\pi\tau)/\sin(\pi\tau/\rho K)$, $\Lambda(\tau) = \sum_{u=-\infty}^{\infty} g(u) g^*(u+\tau) e^{-j2\pi\beta u}$, and $D = \rho K(1 + T_{\mathrm{cp}}\ T_u^{-1})$ represents the number of samples over an OFDM symbol.

Using the same derivation process, we can extend the result of the derivation of second-order CCs of OFDM signal in AWGN channel to multipath channel. Due to the limitation of the manuscript, we present the second-order CCs of OFDM signal in multipath channel directly:

$$C_r(\beta;\tau) = a^2 C_{s,2,1} D^{-1} e^{-j2\pi\beta\varepsilon\rho} e^{-j\frac{2\pi}{\rho K}\Delta f_c f_s^{-1}\tau} \Gamma_K(\tau) \times \sum_{n=-\infty}^{\infty} \sum_{m_1=1}^{M} h(\xi_{m_1}) g(u-\xi_{m_1}) \sum_{m_2=1}^{M} h^*(\xi_{m_2}) \times g^*(u-\xi_{m_2}+\tau) e^{-j2\pi\beta n}. \tag{13}$$

Similarly, the result in (13) omits the term for noise cumulant.

3.2. Analytical expression of CC properties for parameter extraction

This section will elaborate on the analysis on the CC properties in both delay and CF domain in order to efficiently extract and estimate the transmission parameters.

From (11) and (13), it is easily concluded that $C_r(\beta;\tau) \neq 0$ only when $\beta = lT^{-1}$, with l as a positive integer. The conclusion implies that, in theory, the second-order CCs are discrete in cyclic frequency domain, which consists of a set of finite-strength additive components. However, in practice, because the specific cyclostationarity property is not known as *a priori*, the consistent estimator for $C_r(\beta;\tau)$ is used:

$$\hat{C}_r(\beta;\tau) = \frac{1}{T_0} \sum_{t=0}^{T_0} r(t) r^*(t+\tau) \exp(-j2\pi\beta t) \tag{14}$$

when $T_0 \Rightarrow +\infty$, the estimator is close to theoretical $C_r(\beta;\tau)$ asymptotically. However, because of the limitation of the sample amount, the CC magnitude is seldom exactly zero even if β is not a cyclic frequency.

In Figure 1, we demonstrate the simulating results of the second-order CC of OFDM signal verse delay and CF under AWGN channel, for which the following parameters are used: subcarriers number 128, CP ratio 1/4, and oversampling factor 4. The samples over 2,000 OFDM symbols are collected for formulating the result. It can be seen that in the frequency axis direction, the magnitude of CC is nonzero even if $\beta \neq 0$ and 512, which supports the conclusion mentioned above.

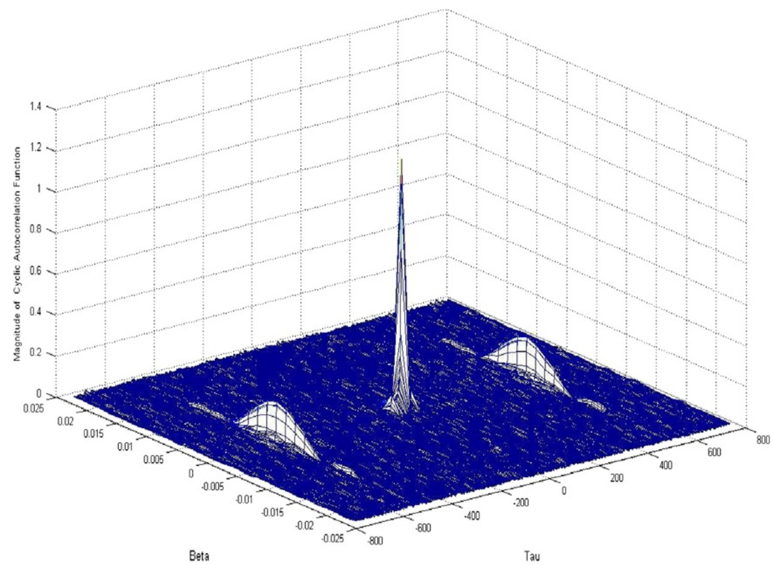

Figure 1 Second-order CC of OFDM signal versus delay and CF, AWGN channel.

Now let us discuss (13) in terms of the delay. As we know, the delay less than T_u/K OFDM signal in (1) does not produce the inter-subcarrier interference. If we assume the delay equals to integral multiple of T_u/K but not integral multiple of T_u, which can be represented as $\tau = mT_u/K$, m is a positive integer and $[m/K] \neq 0$. In this condition, the item $\sum_{k=0}^{K-1} e^{j2\pi k\Delta f_K \tau}$ in (13) can be rewritten as

$$\sum_{k=0}^{K-1} e^{j2\pi k\Delta f_k \tau} = \sum_{k=0}^{K-1} e^{j2\pi k\Delta f_k m/\left(K*\Delta f_k\right)} = \sum_{k=0}^{K-1} e^{j2\pi km/N} = 0. \tag{15}$$

When the delay is equivalent to the integral multiple of T_u, which is represented as $m \times T_u$, there is

$$\sum_{k=0}^{K-1} e^{j2\pi k\Delta f_k \tau} = \sum_{k=0}^{K-1} e^{j2\pi m T_u k\Delta f_k} = \sum_{k=0}^{K-1} e^{j2\pi km} \neq 0. \tag{16}$$

From (15) and (16), the property of second-order CC of OFDM signal in time domain is concluded to be equal to zero unless when the delay is $m \times T_u$. Note that the estimation error because of limitation of the sample number also results in (15) being almost not established in practice. Apart from that, the effect of time dispersion came from multipath propagation which also appeared around $\tau = m \times T_u$.

Figure 2 depicts the simulation result under multipath channel with the same parameter setting. By comparing Figures 1 and 2, we can find that multipath propagation leads to time dispersion, which behaves that multiple nonzero values arise around the peak position. Although these values are usually lower than the peak value, these introduce the amount of interference for searching for the peak. In Section 4, we demonstrate how to eliminate these interferences in practice.

The analysis above employs the CC for continuous-time OFDM signal. For discrete-time OFDM signal, the derivation above and the conclusions for which the second-order CC still holds on are valid except that CF and delay should use the discrete expression. In other words, the $\tau = \pm\rho K$ only when $\beta = lD^{-1}$, and (16) establishes when $\tau = \pm\rho K$.

4. Joint delay and CF domain estimation for OFDM parameters

Based on the analysis of the second cyclic statistics of OFDM signal in Section 3, the number of subcarriers K can be estimated by searching the peak of second-order CC at delay domain, while the estimation of the duration of CP can utilize the discrete behaviors of CC in the cyclic frequency domain. Hence, the approach proposed for the joint blind estimation of the OFDM signal transmission parameters is comprised of two steps. We assume that the parameter estimation is applied after the process of signal detection, so the signal is already oversampled based on an estimate of the signal bandwidth. Moreover, we assume that sampling does not yield aliasing in the spectral and cyclic frequency domains. For the case of aliasing existing, it can be referred to [4] for an elimination method.

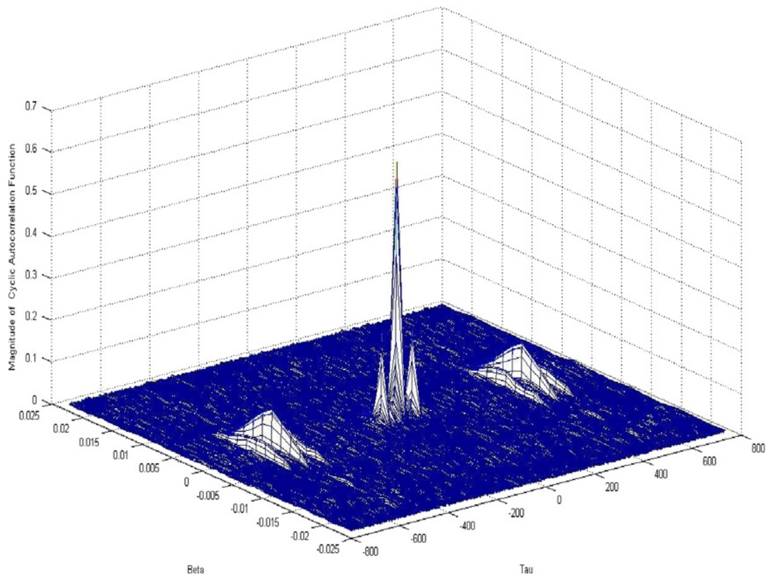

Figure 2 Second-order CC of OFDM signal versus delay and CF, multipath channel.

4.1. Estimation of subcarrier number *K* and symbol duration T_u

The number of subcarriers of the OFDM signal, K, is estimated by exploiting the presence of the CP-induced peak in the CC magnitude at delay ρK and zero CF.

On the condition of $\beta = 0$, the CC magnitude of the baseband-received OFDM signal is checked over the delay range $[\rho K_{\min}, \rho K_{\max}]$. Herein, $K_{\min}$ and $K_{\max}$ represent the minimum and maximum number of subcarriers to be estimated, respectively. The parameter $\rho K_{\max}$ should be chosen large enough to cover the largest candidate peak while avoiding introducing more amount of unnecessary calculation process. The parameter $\rho K_{\min}$ should be chosen far from $\tau = 0$ in order to reduce the influence caused by the channel time dispersion (See Figure 2), but larger $\rho K_{\min}$ leading to some small candidate peak cannot be covered. With respect to the consideration, we propose an approach to estimating the number of subcarrier K as follows.

Firstly, we should set a predefined threshold T_{th} for checking the presence of nonzero second-order CC at interest delay, and the determination of the threshold value will be derived in Section 4.3. Next, we search the candidate second-order CC for the amplitude value larger than the threshold over the $[\rho K_{\min}, \rho K_{\max}]$ delay range. Among the set of the discrete values larger than the threshold, we choose that with the largest delay value, depicted by τ_K. This delay value is exactly what we will utilize to extract the number of subcarriers K.

If all of the second-order CC amplitudes over the $[\rho K_{\min}, \rho K_{\max}]$ are smaller than the threshold, we should only find the CC magnitude reaching a local maximum over the range; the corresponding delay value is used to calculate the number of subcarriers K. This seeking process can be formulated as:

$$\tau_K = \left\{ \begin{array}{l} \max\{\tau \in C_{r\text{OFDM}}(0;\tau) \geq D_{\text{th}}\}; \\ \qquad \exists\, \tau \in [\rho K_{\min}, \rho K_{\max}], \text{ satisfying } C_{r\text{OFDM}}(0;\tau) > T_{\text{th}} \\ \arg\max\limits_{\tau} \left\{ \left| c_{r_{\text{OFDM}}}(0;\tau)_{2,1} \right| \right\}; \\ \qquad C_{r\text{OFDM}}(0;\tau) < T_{\text{th}} \text{ for all } \tau \in [\rho K_{\min}, \rho K_{\max}] \end{array} \right\}. \tag{17}$$

Finally, let τ_K divided by the oversampling factor ρ; the nearest integer power of 2 to the result is the estimated subcarrier number K, which is represented as

$$K = \arg\min_{2^i} \left\{ \left| \tau_K/\rho - 2^i \right| \right\}, \quad i = 0, 1, 2.... \tag{18}$$

The duration of OFDM symbol is also calculated as $T_u = \rho K / f_s$ accordingly.

4.2. Estimation of CP duration and OFDM symbol period *T*

The duration of the CP, L, is estimated based on the CP-induced peaks in the CC magnitude at delay ρK and CFs other than zero. In order to estimate the CP length efficiently, we export the CC magnitude value at fixed delay τ_K and over a certain range of candidate CFs. This CF range should be initialized as small as possible to reduce the amount of calculation. The significant CC magnitude values are adopted only at a small number of CFs greater than zero, and the CF can be represented as $\beta = b\rho(K+L)^{-1}$, with b as a positive integer and CP as unknown. We set the maximum value of b as $b_{\max}$. When taking into account the condition $0 < L < K$, we can define that the CF range used to estimating the CP length is $[1/(2\rho K), b_{\max}(\rho K)]$, and the CF for which the CC magnitude reaches a local maximum, β_{cp} , is selected by

$$\beta_{\mathrm{cp}} = \arg\max_{\beta}\{|C_r(\beta;\tau_K)\}, \quad \beta\in\left(\frac{1}{2\rho K},\frac{b_{\max}}{\rho K}\right), \quad (19)$$

where K is estimated in Section 3.1. Accordingly, the CP duration is estimated as

$$L = \beta_{\mathrm{cp}} b/\rho - K, \quad (20)$$

where b is the minimum integer between 1 and $b_{\max}$ for which a positive value is obtained for CP. Note that although the maximum peak is achieved for $b = 1$ theoretically, another peak may be actually selected under both small observation intervals and low signal-to-noise ratio (SNR) conditions.

In the above analysis, we can see that the proposed algorithm requires neither carrier, waveform, and symbol-timing recovery nor estimation of noise and signal power. It means that the proposed algorithm is robust for both channel fading and synchronization results.

4.3. Determination of threshold value T_{th}

In the proposed approach to estimate the OFDM parameters, the extremely significant step is to exploit the presence of the CP-induced peak in the CC magnitude at delay ρK and zero CF. Basically, this decision-making problem can be formulated into the following hypothesis testing framework:

$$\begin{array}{ll} \mathrm{H}_0: & \tau_K/\rho \neq K \\ \mathrm{H}_1: & \tau_K/\rho = K. \end{array} \quad (21)$$

Because of the existence of noise and estimation error for CC, the CC amplitude is almost not a zero value even under the condition of (15). For this cause, it needs to preparedly set up a threshold for the test problem. However, the threshold relied on the experiment, and the observation (e.g., in [11-13]) is not rigorous, for which it lacks the relationship with the correct recognizing probability. In the following, by referring to the result in [17,18], we give a method to determine the threshold, by which the threshold value will be derived from the constant false alarm rate.

Similarly in [17], we firstly define a statistic testing function:

$$F_{\mathrm{st}} = \hat{C}_r(\Sigma_r)^{-1}\left(\hat{C}_r\right)^T, \quad (22)$$

where $\hat{C}_r$ is the consistent estimator of second-order CC given by (14), and Σ_r is the asymptotic covariance matrix of $\hat{C}_r$. $(\Sigma_r)^{-1}$ and $\left(\hat{C}_r\right)^T$ represent the matrix inverse of Σ_r and transpose of $\hat{C}_r$. Hence, the hypothesis testing problem in (21) can be reformulated as a CFAR-based testing problem:

$$\begin{array}{lll} \mathrm{H}_0: & \tau_K/\rho \neq K, & F_{st} \leq \eta\mathrm{th} \\ \mathrm{H}_1: & \tau_K/\rho = K, & F_{st} > \eta\mathrm{th}. \end{array} \quad (23)$$

ηth is to be determined as threshold for F_{st}. In order to design a decision strategy, the distribution of F_{st} is required to be derived. In our discussion, the difference between $\hat{C}_r$ and $C_r(\beta;\tau)$ is the estimation error:

$$\varepsilon_r = \hat{C}_r - C_r(\beta;\tau). \quad (24)$$

The distribution of ε_r is unknown, and it is close to zero when $T_0 \to \infty$. In [17], it is stated that $T_0\hat{C}_r$ is an asymptotical complex Gaussian variable, and $C_r(\beta;\tau)$ is nonrandom. Therefore, it can be proved that the independent distribution of $r(t)$, which is the distribution of

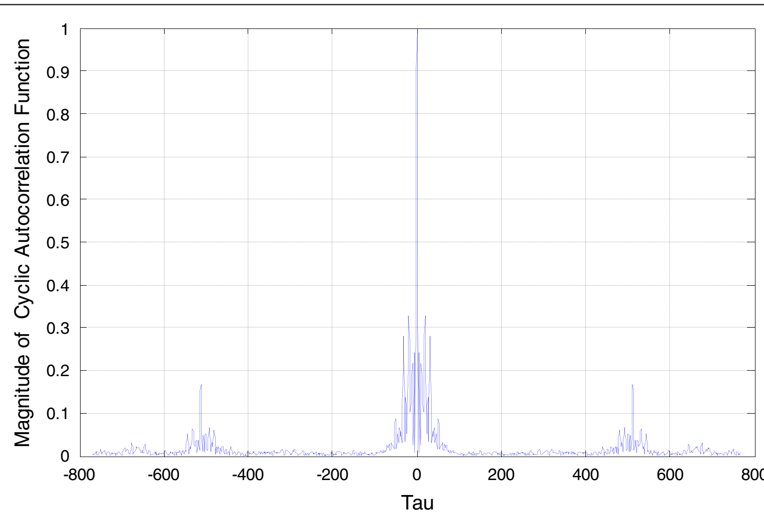

Figure 3 Second-order CC versus delay, at $\beta = 0$.

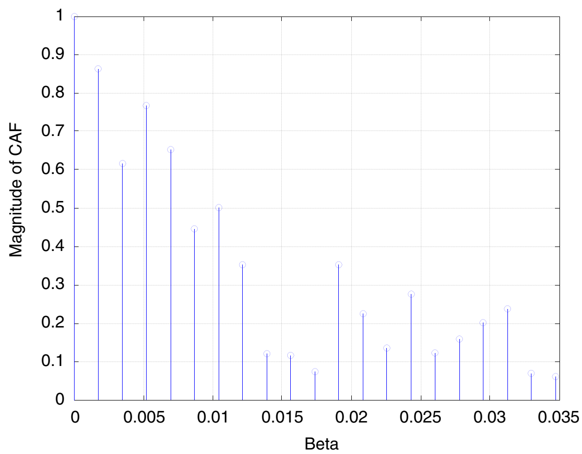

Figure 4 Second-order CC versus CF, at $\tau = \rho K$.

ε_n is asymptotical Gaussian with zero mean. Consequently, for hypothesis H_0 in (23), F_{st} can be viewed as a chi-square distribution variable with $2N$ degrees of freedom, where N is the number of collected samples with different delay used by outputting the consistent estimator in (14).

Hence, the constant false alarm rate in the testing of (23) is given by

$$P_f = \Pr\{F_{st} > \eta_{th} | H_0\}. \tag{25}$$

In other words, the threshold ηth is uniquely determined by the given CFAR probability P_f, by which the theoretical relation between the threshold value and constant false alarm rate was achieved.

5. Simulation results

In the simulation, the OFDM useful symbol duration K is set to 128, 256, 512, and 1,024, and the CP is set to 1/4, 1/8, 1/16, and 1/32. The OFDM subcarrier features a quadrature phase shift keying constellation and a bandwidth of 800 KHz. At the transmitter side, a root-raised cosine pulse shape with a roll-off factor of 0.025 is employed before the A/D converting. In addition, the oversampling frequency f_s is set to 3.2 MHz, Δf_K to 320KHz, and b_{max} to 4. All results are simulating under multipath fading channels.

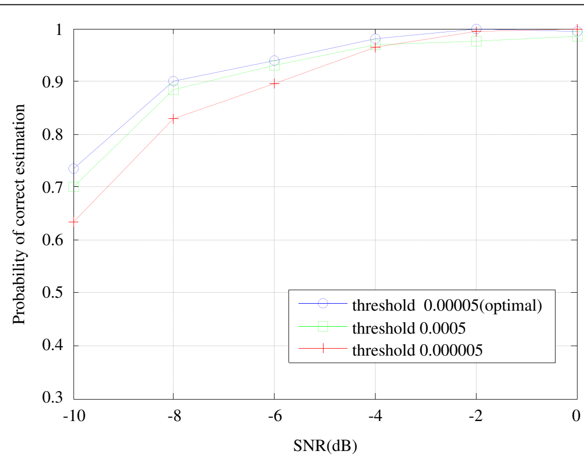

Figure 5 Probability of correct parameter estimation for different thresholds.

Table 1 Accuracy rate of estimation of subcarrier number *K*

K/CP	1/4	1/8	1/16	1/32
128	0.96	1.00	1.00	1.00
256	1.00	1.00	1.00	1.00
512	1.00	1.00	1.00	1.00
1,024	1.00	1.00	1.00	1.00

The magnitude of second-order CC of OFDM signal (K = 128, CP = 1/4, ρ = 4, 2,000 symbols) is plotted versus the delay and at CF = 0 in Figure 3. Apart from the maximum peak at τ = 0, there is another CP-induced peak appeared at about τ = 512, which is used in the proposed estimation method for estimating the subcarrier number K. Around the CC peak, the CC value is not zero due to the time dispersion, which introduces the interference for searching for the CC peak accurately. Furthermore, the CC magnitude estimate at delay ρK is depicted in Figure 4. Significant values can be seen around $\beta = 0$ and $\beta = lD^{-1}$.

Based on the threshold determination algorithm in Section 4.3, we can derive the optimal theoretical threshold value to minimize the constant false alarm rate. In order to evaluate the derivative result, we compare the correct estimation probability in terms of different threshold values: the theoretical optimal, larger and smaller than the optimal value, respectively. From Figure 5, we can conclude that the estimation with the theoretical optimal threshold can achieve the best correct probability, especially on low SNR condition, which proved the algorithm proposed in Section 4.3 in effect.

The simulation results for the blind parameter estimation under the vehicular B channel are presented in Tables 1 and 2. Table 1 shows the accuracy rate of estimation of subcarrier number K, and the accuracy rate of CP detection is depicted in Table 2. From Tables 1 and 2, we can see that the proposed algorithm can estimate the OFDM signal parameters precisely even at the condition of SNR = 0. However, we also should note that when the CP length is

Table 2 Accuracy rate of estimation of CP length

K/CP	1/4	1/8	1/16	1/32
128	1.00	1.00	-	-
256	1.00	1.00	1.00	-
512	1.00	1.00	1.00	0.82
1,024	1.00	1.00	0.99	0.98

too short, the cyclostationarity of OFDM signal will be affected significantly, and the algorithm performance will also decrease.

6. Conclusions

This paper has investigated the blind estimation of OFDM parameters based on the signal's cyclostationarity. Firstly, the possibility of extracting the transmission parameters from second-order CC is derived theoretically, which analyzes both the continuous and discrete signal models under the AWGN channel and multipath channel, respectively. The proposed estimation approach jointly exploits the delay and cyclic frequency domain properties of second-order CC and develops a CFAR-based statistic hypothesis testing framework to detect the presence of CP-introduced cyclostationarity. The approach only assumes that the OFDM signal have the cyclic prefix and does not require any priori condition, so it can be easily extended to most cases that need blind estimation of OFDM parameters.

Competing interests

The authors declare that they have no competing interests.

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