A high order statistics based multipath interference detection method

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Abstract
Multipath interference commonly exists in wireless communication, navigation and radar systems, which may cause severe signal fading and bit error rate (BER) performance degradation. Therefore the detection of multipath interference is urgently needed. This paper proposes a high order statistics (HOS) based multipath interference detection method, since the HOS of the received signal show distinct difference when multipath interference exists. Moreover, the generalized theoretical values of the moments and cumulants, which release the demand for prior knowledge of timing information and symbol period, are deduced in this paper. It makes the proposed HOS features based detection method robust in blind environment. Computer simulations are performed to verify the proposed method.

1 INTRODUCTION
Multipath interference commonly exists in wireless communication, navigation and radar systems affected by obstacle reflection [1–3] as shown in Figure 1. And the reflection path signal has a relative time delay corresponding to the main path signal. For wireless communication systems, inter-symbol interference emerges and severely degrades bit-error rate performance of the receiver [5] when the relative time delay is larger than the symbol period [4]. Moreover, multipath interference is the main error source in global navigation systems and radar systems [6–12]. The detection of multipath interference can benefit the design of the receiver and provide prior information for antenna selection in antenna arraying systems. Therefore, there is an urgent need for the detection of multipath interference.

Some literatures have studied the problem of multipath interference detection, and the most commonly used methods can be divided into two types, the multipath parameters estimation based method [13–15] and the feature-based method [16–18]. The multipath parameters estimation based method detects multipath interference by estimating the number, time delays, amplitudes and phase lags of paths. Sequential maximum likelihood (SML) algorithm proposed by Sahmoudi et al. [13] uses the scalar maximum likelihood (ML) algorithm to estimate path parameters in sequence, and estimates the number of paths by comparing signal to noise ratio (SNR) with light of sight to residual signal power ratio. Modified sequential maximum likelihood (MSML) algorithm proposed by Sokhandan et al. [14] first estimates the strongest signal components, then refines all the estimated path parameters, then removes the influence of the estimated signal components on the correlation function, and finally calculates a normalized likelihood ratio according to the parameters of the detected paths. If the likelihood ratio exceed a certain threshold, a new path is considered to exist. Parameters of all paths are estimated by repeating the above operations. Krach et al. [15] used twofold marginalized Bayesian filter for multipath estimation. The algorithm fixes the number of signal paths, and introduces path activity, a vector of discrete...
variables with value of \{0, 1\}, to represent the existence of paths. The vector is in a discrete finite state space, which can be optimally estimated by grid based filter. These methods estimate the multipath channel comprehensively, and they have low computational efficiency when only the existence of multipath interference is needed to be detected. In addition, these methods require prior information of the channel model and are not suitable for low signal to noise ratio (SNR) or fast changing channels. The feature based method realizes multipath detection by observing the signal characteristics affected by multipath interference. Strode et al. [16] detected multipath interference by analysing carrier to noise ratio (CNR) fluctuations at different frequencies. The principle is that the phase delay between the reflection path and the direct path is determined by the frequency, and phase delays on different frequencies have constructive or destructive effects on the CNR, which is a feature reflecting the multipath interference. The method obtains the detection threshold according to the CNR under weak multipath interference environment. To ensure reliability, CNRs on three frequencies are used for detection. Špánik et al. [17] adopted CNRs on two frequencies to detect multipath interference, which enables the applicability for GPS and GLONASS systems. Considering the multipath error in the pseudo range measurement of GNSS signal, Beitler et al. [18] adopted the code-minus-carrier delta range (CMCD) feature, which means the difference of delta ranges derived by the code and the carrier. When multipath does not exist, CMCD is equal to the noise process of receiver. When multipath exists, the statistical characteristics of CMCD change. By observing the statistical characteristics of CMCD, multipath interference can be detected. Fluctuation of the CNR is generally considered in feature based methods, but it is difficult to detect in slow changing channels, which limits the applicability of the methods.

In this paper, we first establish the multipath interference model, obtain the expression of the received signal in the multipath interference environment, and analyse the influence of multipath interference on BER through theoretical derivation and simulation. Then, based on the moments and cumulants of the received signal, two HOS features are constructed. For BPSK, MPSK and QAM modulations, we derive the theoretical values of the HOS features, and verify their correctness through Monte-Carlo simulation. Moreover, the simulation results indicate that the HOS features show distinct difference in the multipath interference environment, and our proposed multipath detection method based on the HOS features has a high correct detection probability.

The main contributions of this paper are as follows:

1. The HOS features are adopted to detect multipath interference because the HOS features of the received signal show distinct difference when multipath interference exists.
2. The generalized theoretical values of the moments and the cumulants are deduced.
3. The HOS features release the requirement for prior knowledge of timing information and symbol periods and thus make our proposed HOS features based detection method robust in blind environment.

This paper is organized as follows. Section 2 theoretically describes the system model and analyses the BER performance degradation caused by multipath interference. Section 3 deduces the generalized theoretical values of HOS and introduces the HOS based multipath interference detection method. Section 4 summarizes the proposed method.

2 | MULTIPATH INTERFERENCE MODEL AND ITS IMPACT ON BER PERFORMANCE

2.1 | Multipath interference model

Considering reflection path signals have different carrier phases and delays corresponding to the main path signal, the received signal through a multipath fading channel can be modelled as follows [19].

\[
r(t) = b_1 e^{j(2\pi f_c t + \theta_1)} x(t - \tau_1) + \sum_{i=2}^{I+1} b_i e^{j(2\pi f_i t + \theta_i)} x(t - \tau_i) + v(t)
\]  

where \(b_1\) is the amplitude of the main path signal and \(b_i\) is the amplitude of the \(i\)th reflection path signal; \(f_c\) is the carrier frequency; \(\theta_i\) is the carrier phase of the main path signal and \(\theta_i\) is the carrier phase of the \(i\)th reflection path signal; \(\tau_1\) and \(\tau_i\) represent the time delay of the main path signal and the \(i\)th

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**FIGURE 1** Multipath interference in wireless communication, navigation and radar systems
reflection path signal, respectively; \( I \) represents the number of the reflection paths and is assumed to be 1 for simplicity in this paper; \( n(\ell) \) is the zero-mean additive white Gaussian noise with variance \( \sigma^2 \); \( x(\ell) \) is the baseband modulation signal as

\[
x(\ell) = \sum_k a_k g(\ell - kT)
\]

where \( a_k \) represents the \( k \)th modulated symbol of the transmitted signal; \( T \) is the symbol period; \( g(\ell) \) is the impulse response of raised cosine roll-off filter. For simplicity, the part of impulse response \( g(\ell) \) exceeding \([-2.5T, 2.5T]\) is neglected and assumed as 0, which is too weak to influence the baseband modulation signal.

Assuming the receiver is sampled at the sampling rate \( 1/T_s \). The over-sampled data can be written as

\[
r_\ell = b_1 e^{j(2\pi f_s kT)} x_{1,\ell} + \sum_{i=2}^{I+1} b_i e^{j(2\pi f_s kT)} x_{i,\ell} + n_\ell
\]

where \( r_\ell, x_{1,\ell}, x_{i,\ell}, n_\ell \) are the sampling points of \( r(\ell T), x(\ell T - \tau_1), x(\ell T - \tau_2), n(\ell T) \), respectively.

### 2.2 Impact of multipath interference on BER performance

For simplicity, the carrier synchronization of the received signal \( n(\ell) \) is supposed to be perfect and only the reflection path signal with maximum energy is taken into consideration, other weaker reflection path signals and the noise \( n(\ell) \) are neglected. Therefore the baseband signal \( r_\ell(\ell) \) after carrier synchronization can be described as

\[
r_\ell(\ell) = \sum_k a_k g(\ell - kT) + b_2 \sum_k a_k g(\ell - \Delta \tau_{21} - kT)
\]

where \( b_2 \) is the amplitude ratio between the reflection path signal and the main path signal, i.e. \( b_2 = b_2/b_1 \); \( \Delta \tau_{21} \) is the delay difference between the reflection path signal and the main path signal, i.e. \( \Delta \tau_{21} = \tau_2 - \tau_1 \); \( a_k \) is assumed to be BPSK modulated and takes 1 or -1 with equal probability for simplicity.

The matched filter output of the \( m \)th symbol, represented by \( z_m \), can be deduced according to Equation (4).

\[
\Delta \tau_{21} < T : z_m = a_m + b_2 (g(-2T + \Delta \tau_{21}) a_{m-2} + g(-T + \Delta \tau_{21}) a_{m-1} + g(T + \Delta \tau_{21}) a_{m+1} + g(2T + \Delta \tau_{21}) a_{m+2})
\]

\[
T \leq \Delta \tau_{21} < 2T : z_m = a_m + b_2 (g(-2T + \Delta \tau_{21}) a_{m-2} + g(-T + \Delta \tau_{21}) a_{m-1} + g(T + \Delta \tau_{21}) a_{m+1} + g(2T + \Delta \tau_{21}) a_{m+2})
\]

\[
2T \leq \Delta \tau_{21} < 3T : z_m = a_m + b_2 (g(-2T + \Delta \tau_{21}) a_{m-2} + g(-T + \Delta \tau_{21}) a_{m-1} + g(T + \Delta \tau_{21}) a_{m+1} + g(2T + \Delta \tau_{21}) a_{m+2})
\]

where \( m \) is the index of the transmitted symbol. Therefore, we have

\[
\Delta \tau_{21} < (n + 1) T : z_m = a_m + b_2 (g(-2T + \Delta \tau_{21}) a_{m-2n} + g(-T + \Delta \tau_{21}) a_{m-2n+1} + g(T + \Delta \tau_{21}) a_{m+1+n} + g(2T + \Delta \tau_{21}) a_{m+2+n})
\]

where \( n \) is the count of symbol period \( T \).

The probability density function of the matched filter output \( z \) with consideration of the noise \( n(\ell) \) is shown as follows:

\[
p(z) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{(z - \mu)^2}{2\sigma^2}}
\]

Therefore, the bit error rate \( P_e \) considering multipath interference can be deduced as

\[
P_e = \frac{1}{2} P(-1|1) + \frac{1}{2} P(1|1)
\]

\[
= \frac{1}{2} \int_{-\infty}^{0} \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{(z - \mu)^2}{2\sigma^2}} dz + \frac{1}{2} \int_{0}^{\infty} \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{(z - \mu)^2}{2\sigma^2}} dz
\]

\[
= \frac{1}{2\sqrt{\pi}} \int_{-\infty}^{0} e^{-t^2} dt + \frac{1}{2\sqrt{\pi}} \int_{0}^{\infty} e^{-t^2} dt
\]

where \( \mu_{m,1} \) and \( \mu_{m,-1} \) represent the matched filter output \( z_m \) when \( a_m \) is 1 and -1, respectively.

Monte-Carlo simulations are carried out to prove the above derivation. The simulation parameters are set as follows: the baud rate \( 1/T \) is 5 Mbps; the carrier frequency \( f_s \) is 10 MHz; the sampling rate \( F_s \) is 50 MHz; the amplitude \( b_1 \) of the main path signal is normalized as 1 V and the amplitude \( b_2 \) of the reflection path signal is 0.1 V; the delay difference \( \Delta \tau_{21} \) varies from 0.1T to 5T; the \( E_{i0}/N_0 \) of the received signal is set as 0, 2 and 4 dB, respectively; the timing synchronization is assumed to be perfect during the recovery of the transmitted symbols from the received signal. The BER performance is shown in Figure 2. The solid horizontal lines represent the theoretical BER without multipath interference under different SNRs. It is clear that the simulation values coincide with the theoretical values. And the BER performance is degraded by the multipath interference only when the delay difference \( \Delta \tau_{21} \) between the reflection path signal and the main path signal is large. If the delay difference
\[ \Delta \tau_{21} \] is smaller than \( T \), the multipath interference may benefit the recovery of the transmitted symbols.

### 3 THE MULTIPATH INTERFERENCE DETECTION METHOD BASED ON HOS FEATURES

In this section, the theoretical values of the moments and cumulants influenced by multipath interference are first investigated. The traditional calculation method of the moments and cumulants is based on the timing synchronization sampled sequence which removes the inter-symbol interference from the over-sampled data [20–24]. That means the moments and cumulants are difficult to be obtained when the SNR is low. Therefore, more generalized theoretical values of the moments and cumulants of the over-sampled data are deduced. Instead of only using the optimal sample per symbol for calculation, all the samples per symbol are used. Then the HOS based multipath interference detection method is introduced.

#### 3.1 Theoretical value of the HOS

The moments \( M_{pq} \) and cumulants \( C_{pq} \) of the received signal \( r(t) \) represent the shape of the distribution of signal constellation. They can be obtained by calculating the expectation of \( \{ (r_k)^{p-q}(r_k^*)^q \} \) [20, 21].

\[
M_{pq} = E \left\{ (r_k)^{p-q}(r_k^*)^q \right\} \\
C_{20} = M_{20} = E \left\{ (r_k)^2 \right\} \\
C_{21} = M_{21} = E \left\{ r_k r_k^* \right\} 
\]

(11)

For simplicity, the additive noise \( v_k \) is neglected in the derivation of the generalized theoretical values of the HOS features and the number of the reflection path is assumed to be 1. Thus the received signal \( r(t) \) can be revised as

\[
r(t) = h_1 e^{j(\omega_1 t + \Theta_1)} x(t) + h_2 e^{j(\omega_1 t + \Theta_2)} x(t - \Delta \tau_{21})
\]

\[ C_{42} = M_{42} - |M_{20}|^2 - 2M_{21}^2 \]

\[ = E \left\{ (r_k r_k^*)^2 \right\} - E^2 \left\{ (r_k)^2 \right\} - 2E^2 \left\{ r_k r_k^* \right\} \]

\[ C_{63} = M_{63} - 9C_{42}C_{21} - 6C_{21}^3 \]

\[ = E \left\{ (r_k r_k^*)^3 \right\} - 9E \left\{ (r_k r_k^*)^2 \right\} E \left\{ r_k r_k^* \right\} \]

\[ + 9E^2 \left\{ (r_k)^2 \right\} E \left\{ r_k r_k^* \right\} + 12E^3 \left\{ r_k r_k^* \right\} \]

(12)

where \( (\cdot)^* \) represents taking the conjugate; \( E\{\cdot\} \) represents the expectation operation (In the following derivation, the expectation is replaced by ergodic average).

According to Equations (11) and (12), the signal power has an impact on the moments and cumulants. To remove this impact, some HOS features are constructed by the cumulants for signal detection and classification [21], as HOS features characterize the distribution shape of the received signal constellation with normalized signal power. And here we construct two HOS features as

\[
F_1 = \begin{vmatrix} C_{63} \end{vmatrix} \begin{vmatrix} C_{42} \end{vmatrix} \begin{vmatrix} C_{21} \end{vmatrix} \]

(13)

\[
F_2 \begin{vmatrix} C_{42} \end{vmatrix} \begin{vmatrix} C_{21} \end{vmatrix} \]

(14)
\[
\begin{align*}
&= b_1 e^{j(\omega_1 t + \theta_1)} x(t) \ast \left[ 1 + b_2 e^{j(\omega_2 t - \theta_2)} \delta(t - \Delta t_2) \right] \\
&= b_1 e^{j(\omega_1 t + \theta_1)} a(t) \ast \tilde{g}(t) \\
&= b_1 e^{j(\omega_1 t + \theta_1)} \tilde{x}(t)
\end{align*}
\]

where \( \omega = 2\pi f_c \); \( \tilde{g}(t) = g(t) \ast \left[ 1 + b_2 e^{j(\omega_2 t - \theta_2)} \delta(t - \Delta t_2) \right] \); \( \tilde{x}(t) = a(t) \ast \tilde{g}(t) \); * represents convolution operation. According to Equation (15), Equations (11) and (12) can be further expanded as

\[
\begin{align*}
M_{20} &= E \left\{ h_1^2 /2\omega_1 kt + \theta_1 x_k^2 \right\} \approx 0 \\
M_{21} &= b_2^2 E \left\{ \tilde{x}_k x_k^2 \right\} \\
M_{42} &= b_1^4 E \left\{ \left( \tilde{x}_k x_k^2 \right)^2 \right\} \\
M_{63} &= b_1^6 E \left\{ \left( \tilde{x}_k x_k^2 \right)^3 \right\}
\end{align*}
\]

\[
C_{20} \approx 0 \\
C_{21} = b_2^2 E \left\{ \tilde{x}_k x_k^2 \right\} \\
C_{42} = b_1^4 E \left\{ \left( \tilde{x}_k x_k^2 \right)^2 \right\} - 2b_1^6 E \left\{ \tilde{x}_k x_k^2 \right\} \\
C_{63} = b_1^6 E \left\{ \left( \tilde{x}_k x_k^2 \right)^3 \right\} - 9b_1^8 E \left\{ \left( \tilde{x}_k x_k^2 \right)^2 \right\} E \left\{ \tilde{x}_k x_k^2 \right\} + 12b_1^{10} E \left\{ \tilde{x}_k x_k^2 \right\}
\]

Assume \( \Delta t_2 / T_c = a \) and \( d = NL_d + p_d \) (where \( N \) is the sampling number per symbol period and \( L_d, p_d \) are positive integers), the discrete point \( \tilde{x}_k \) of \( \tilde{x}(kT_c) \) can be obtained as follows:

\[
\tilde{x}_k = a(k) \ast \tilde{g}(k)
\]

\[
= \sum_{m=0}^{2NL_d + d + 1} \tilde{g}(m) a(N(m - 1) + p - n + 1 + L_d) \\
+ \cdots + \tilde{g}(N(L_d + L_d + d + 1) + p - n + 1 + L_d)
\]

\[
E \left\{ (\tilde{x}_k \tilde{x}_k^*)^q \right\}
\]

To simplify Equation (18), the coefficient of each symbol \( a_k \) is replaced by \( \tilde{C}_{nk} \), and we have

\[
\tilde{x}_k = \tilde{C}_{nm-L_d-L_d-1} a_{m-L_d-1} + \cdots + \tilde{C}_{m+L_d} a_{m+L_d}
\]

Therefore the expectation of \( (\tilde{x}_k \tilde{x}_k^*)^q \) (\( q \leq 3 \)) when \( x_k \) is BPSK, MPSK (\( M \geq 4 \)) and QAM modulated can be deduced as Equations (20), (21) and (22) respectively:

\[
E \left\{ (\tilde{x}_k \tilde{x}_k^*)^q \right\} = \sum_{j=0}^{N} \left\{ \sum_{qL_d=0}^{qL_d} \tilde{C}_{nm-L_d-L_d-1}^{qL_d+L_d+2} \times \tilde{C}_{m+L_d}^{qL_d+L_d+2} \right\}^{qL_d+L_d+2} / N
\]

\[
E \left\{ (\tilde{x}_k \tilde{x}_k^*)^q \right\} = \left\{ \sum_{j=1}^{N} \left\{ \sum_{qL_d=0}^{qL_d} \tilde{C}_{nm-L_d-L_d-1}^{qL_d+L_d+2} \times \tilde{C}_{m+L_d}^{qL_d+L_d+2} \right\}^{qL_d+L_d+2} / N \right\}^{qL_d+L_d+2}
\]

\[
E \left\{ (\tilde{x}_k \tilde{x}_k^*)^q \right\} = E \left\{ [J(k) \ast \tilde{g}(k)]^2 + (Q(k) \ast \tilde{g}(k))^2 \right\}^{qL_d+L_d+2}
\]
FIGURE 3 The HOS feature $F_1$ of BPSK modulated signal as a function of the power ratio $(h_1/h_2)^2$.

![Graph showing the HOS feature $F_1$ of BPSK modulated signal as a function of the power ratio $(h_1/h_2)^2$.]

$F_1 = \frac{1}{N} \sum_1^{q_1} \sum_{q_2}^2 C_{q_2}^{q_1} \tilde{G}_{m-L_d-L_d}^{q_1} \times ... \times \tilde{G}_{m+L_d}^{2q_2+2} \times E[I_m^{q_1}] \times ... \times E[I_m^{2q_2+2}] / N$ (22b)

where $q_1, ..., q_{L_d+L_d+2}$ are positive integers; $C_{q_1}$ denotes binomial coefficient; $[.]$ represents floor rounding; $\%$ represents taking modulo operation; $I_m, Q_m$ are the in-phase and quadrature parts of the transmitted symbol $a_m$ when the signal is QAM modulated, i.e. $a_m = I_m + jQ_m$.

Therefore the theoretical HOS features $F_1$ and $F_2$ can be deduced according to Equations (17), (20)–(22) as

$F_1 = \frac{C_{q_1}^{2}}{C_{q_2}^{2}} = \left| \frac{E_3 - 9E_2E_1 + 12E_3^2}{E_2 - 2E_1} \right|^2$ (23)

$F_2 = \frac{C_{q_2}^{2}}{C_{q_2}^{2}} = \left| \frac{E_1 - 2E_1^2}{E_{1,1}} \right|^2$ (24)

where $E_q = E\{\tilde{X}_m^{q} \tilde{X}_m^{q}\}$.

FIGURE 4 The HOS feature $F_1$ of MPSK modulated signal as a function of the power ratio $(h_1/h_2)^2$.
To prove the above theoretical derivation, Monte-Carlo simulations are performed to obtain the simulation values of the HOS features. The simulation parameters are set as follows: the carrier frequency $f_c$ is 10 MHz; the baud rate $1/T$ is 1 Mbps; the sampling rate $F_s$ is 140 MHz; the delay difference $\Delta\tau_{21}$ is set as $1.5T, 2.5T, 3.5T$ and $5.5T$ respectively; the total symbol number used for simulation is $5 \times 10^4$. The simulation results are shown in Figures 3–8, revealing that the theoretical values highly match the simulation values. Moreover, the simulation results illustrate that the HOS features are influenced by the delay difference $\Delta\tau_{21}$ and the power ratio $h_1^2/h_2^2$. For MPSK and QAM modulated cases, the HOS features monotonically increase with the increasing power ratio and gradually become stable with the increasing delay difference. Due to the distinct difference of the HOS features with and without the existence of multipath interference, $F_1$ and $F_2$ can be adopted to detect the multipath interference in MPSK- and QAM-modulated cases.

However, Figure 3 reveals that the HOS feature $F_1$ does not increase monotonically with increasing power ratio in BPSK modulated case, which causes detection ambiguity, for the value of the HOS feature $F_1$ under some certain power ratios is the same as the value when multipath interference is weak or does not exist. Therefore the HOS feature $F_1$ is not applicable for BPSK modulated multipath interference detection. For the HOS feature $F_2$, Figure 6 reveals that the value of the HOS feature $F_2$ under some certain power ratios is larger when multipath interference is weak or does not exist than when multipath interference is strong. Thus, the HOS feature $F_2$ is suitable for BPSK-modulated multipath interference detection.
3.2 The HOS-based multipath interference detection method

The total block diagram of the HOS-based multipath interference detection method is shown in Figure 9. The HOS feature of the received signal is first extracted and then compared with the threshold $\eta$. If the HOS feature is larger than or equal to threshold $\eta$, multipath interference is determined to be non-existed or weak. Otherwise, multipath interference is determined to be existed. For the HOS features gradually become
stable with increasing delay difference $\Delta \tau_{21}$ and the BER performance of the received signal is degraded only when the delay difference $\Delta \tau_{21}$ is large and the power ratio $\frac{h_1}{h_2}$ is small, the threshold $\eta$ can be set as the HOS feature when the delay difference $\Delta \tau_{21}$ is $3T$ and the power ratio $\frac{h_1}{h_2}$ is 10.

In this section, computer simulations are carried out to illustrate the detection performance of the proposed method. To evaluate the detection performance, the correct detection probability $P_{CI}$ is counted in 500 times of simulations. The simulation parameters are set as follows: the carrier frequency $f_c$ is 10 MHz; the baud rate $1/T$ is 1 Mbps; the sampling rate $F_s$ is 140 MHz; the delay difference $\Delta \tau_{21}$ is $3.5T$; the total symbol number is $5 \times 10^4$. The results are shown in Figures 10 and 12. Figures 10 and 11 reveal that the proposed detection method performs better under a higher SNR and a larger symbol number. Moreover, Figure 12 reveals that the detection performance decreases with increasing power ratio under MPSK and QAM modulated cases. This is because the extracted HOS feature gets closer to the threshold $\eta$ when the power ratio is increasing. However, the detection performance increases with increasing power ratio under BPSK modulated case when the power ratio is small. This is because the HOS feature $F_2$ decreases with increasing power ratio when the power ratio is small, which makes the extracted HOS feature far from the threshold $\eta$.

4 | CONCLUSION

In order to detect multipath interference, more generalized theoretical values of the moments and cumulants are deduced based on the over-sampled data in this paper. Instead of using the timing synchronization points per symbol for calculation, all
the sampled points per symbol are taken into consideration in
the derivation of the generalized theoretical values. Therefore
it is more robust in blind environment for the release of the
demand for prior knowledge of timing information and symbol
period [20–24]. Moreover, a novel multipath interference detec-
tion method is proposed based on the derived generalized HOS
features. Computer simulations verify the feasibility of our pro-
posed method.

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