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A Blind Side Information Detection Method for Partial Transmitted Sequence Peak-to-Average Power Reduction Scheme in OFDM Underwater Acoustic Communication System

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ABSTRACT In this paper, we propose a blind side information detection scheme for partial transmitted sequence (PTS) peak-to-average power ratio reduction (PAPR) method in underwater acoustic (UWA) orthogonal frequency division multiplexing (OFDM) communication systems. The proposed scheme employs a corresponding table between the distributions of pilot tones and the phase rotation candidates to obtain a pseudo-optimum PTS PAPR reduction performance and a dramatic reduction of the total calculation cost. Meanwhile, predefined distributions of pilot tones are used to identify the phase rotation factor combination that has been selected and used at the transmitter. The spectral efficiency and data rate remain unaffected as the proposed PTS technique requires no additional pilot tones except for the tones used for channel estimation. Due to the sparse characteristic of the underwater acoustic channel, compressed sensing is adopted to complete the channel estimation and the phase rotation detection. The basis pursuit denoising algorithm employed here reduces the number of pilot tones as compared with linear channel estimation method. Simulation results show that the proposed scheme has better PAPR reduction performance compared to the conventional PTS scheme and the performance gap increases with the number of subblocks. Simulation and field experimental results demonstrate that the proposed scheme can differentiate the phase rotation factor. Therefore, the quality of the UWA OFDM communication system is significantly enhanced.

INDEX TERMS Orthogonal frequency division multiplexing (OFDM), peak-to-average power ratio (PAPR) reduction, partial transmitted sequence (PTS), side information, underwater acoustic communication (UAC).

I. INTRODUCTION

Low carrier frequency, strong multi-path interferences and high noise level make the shallow underwater acoustic (UWA) channel the most difficult wireless communication channel. Orthogonal frequency division multiplexing (OFDM) is an attractive technique for high-rate data transmission due to the minimizing effect over frequency-selective fading channels [1], therefore, OFDM has been extensively used in high-rate UAC system to overcome the frequency-selective fading [2]–[4]. However, there are some drawbacks which limit the use of OFDM in underwater acoustic communication (UAC) systems. One of which is the higher peak-to-average power ratio (PAPR) than the single carrier communication systems [5]. High PAPR requires a large linear region of the amplifier, in order to avoid distortion and results in an increase in the cost of the amplifier. Due to the limited emission power of the UWA transducer, high PAPR of the OFDM signal is a serious limitation for the effective utilization of the transmit power, reducing the transmission distance of the communication signal and the received signal-to-noise ratio (SNR). Therefore, an efficient PAPR reduction algorithm is highly desirable for the power restricted and high ambient background noise UAC system.

OFDM PAPR reduction has been a subject of interest since the last decade. Numerous methods have been proposed, which can be sorted into three main types, including clipping, coding and probabilistic (scrambling) techniques [6]–[8]. The clipping technique is the simplest to implement among these three kinds of techniques, which employs clipping or nonlinear saturation around the peaks to reduce the PAPR. However, it may cause in-band and out-band interferences because of destroying the orthogonality among the subcarriers. This is hardly acceptable for the UAC, which requires highly qualified channel coding techniques to ensure the BER performance because of the complex channel characters. The coding method causes no distortion and creates no out-of-band radiation; however, it suffers from low bandwidth efficiency as the code rate is reduced, making it unsuitable for UWA communication.

The key idea of the probability techniques is to scramble an input data block of the OFDM symbols and transmit the one with the minimum PAPR so that the probability of occurrence of high PAPR can be reduced. This kind of technique does not suffer from the out-of-band power, so the BER performance of the communication system will not decrease. Among them, SLM and PTS are two promising techniques because there is no distortion in the transmitted signal and can significantly improve the statistics of the PAPR [9], [10]. However, the main drawbacks of conventional SLM and PTS techniques are high computational complexity and transmission of several side information (SI) bits. If the receiver wants to obtain the sending information by demodulation and decoding, the additional SI on the phase factors must also be transferred from the transmitter to the receiver, which leads to a loss in spectral efficiency and data throughput. The symbols carrying the side information needs to be received and decoded first, and the remaining symbols can be de-rotated and decoded according to the decoded side information for every symbol afterwards. The symbols carrying SI will always be added at the very end of all the transmitted symbols or at the end of each frame. Therefore, the real-time performance of the communication system will be lost. Furthermore, the BER performance of the UWA OFDM systems will possibly be ruined since any error in the detection of side information can damage the entire data symbols.

Several proposals suggested different phase rotation schemes, such as concentric circle mapping, extended constellation or cyclically shifting, to reduce the complexity and remove the SI [11]–[13]. However, all these schemes caused BER performance distortion at different levels. An SI free PTS scheme relying on a preamble was proposed in [14], based on the assumption that the channel stays still in the adjacent symbols. The results of channel estimation using the preamble were not accurate enough for reliable UWA communication system. In literature [15], each SI index was associated with a particular set of locations inside the data block at which the modulation symbols have been extended. In the receiver, an SI detection block attempted to determine the locations of the extended symbols. Literature [16] made the pilot tones partitioning and scrambling exactly as what the data tones do, and then the channel estimation can be conducted without de-rotation, which means no side information was needed. The only drawback was the existence of channel estimation error on the edge of every subblock. It was overcome by adding another two pilot tones at the beginning and the end of every subblock, however, the data rate was reduced [17]. Paper [18] embedded the SI identifying rotating vectors into the alternative signal sequences by giving identifiable phase offset to the elements of each rotating vector. Papers [19] and [20] put only one pilot in every subblock as a symbolic of the phase rotation vector. However, channel estimation cannot be accomplished when the number of the subblocks is rather too small, even by adopting compressed sensing.

We are motivated to deal with the two drawbacks of the conventional PTS (C-PTS) technique: computational complexity and side information transmission. In this paper, a pseudo-optimum PTS scheme is proposed to reduce the computational complexity of the C-PTS when dealing with the selection of the OFDM signal with the minimum PAPR. It also modifies the PAPR reduction performance because of the random selection of the phase rotation vectors, while the C-PTS fixes the phase rotation vectors only in the set $\{1, -1\}$. The detector can distinguish the selected candidate through the sparse characteristic of the underwater acoustic channel. Compressed sensing [21] has been adopted to complete the channel estimation and detection of the phase rotation vectors index. The proposed PTS technique avoids any data rate loss as no side information transmission nor any additional pilot tones are required except for the tones used for channel estimation. Our scheme reduces the complexity and significantly enhances the quality and the frequency efficiency of the OFDM communication system, thus making it most suitable for UWA communication.

II. DEFINITION OF THE PAPR AND THE CONVENTIONAL PTS SCHEME

A. DEFINITION OF PEAK-TO-AVERAGE POWER RATIO

In OFDM communication system, PAPR is the ratio of the maximum power and the average power of the complex passband signal $s(t)$ [22].

$$\mathrm{PAPR}\left\{\widetilde{s}(t)\right\} = \frac{\max\left|Re(\widetilde{s}(t)e^{j2\pi f_c t})\right|^2}{E\left\{\left|Re(\widetilde{s}(t)e^{j2\pi f_c t})\right|^2\right\}} = \frac{\max|s(t)|^2}{E\left\{|s(t)|^2\right\}} \quad (1)$$

where $\widetilde{s}(t)$ represent the complex baseband signal, while $s(t)$ is the complex passband signal.

In general, Complementary cumulative distribution function (CCDF) is utilized to measure the PAPR distribution of OFDM symbols. The function is given as

$$\begin{aligned}\tilde{F}_{Z_{\max}}(z) &= P\left(Z_{\max} > z\right) = 1 - P\left(Z_{\max} \leq z\right)\\ &= 1 - \left(1 - e^{-z^2}\right)^N. \end{aligned} \quad (2)$$

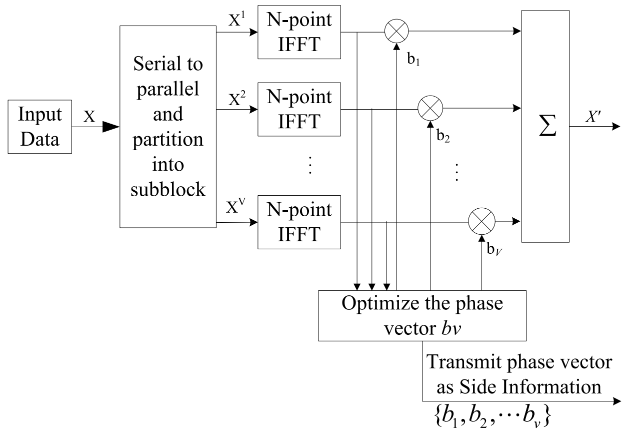

FIGURE 1. The framework of C-PTS technique for PAPR reduction.

B. THE CONVENTIONAL PARTIAL TRANSMIT SEQUENCE SCHEME

The C-PTS technique partitions an input data block into V disjoint subblocks and combines them together after applying phase rotation (scramble) to each subblock. Then minimize the PAPR of OFDM symbol by optimizing the phase vector. The framework of C-PTS technique is shown in Fig.1.

Input data block $X = [X_0, X_1, \cdots, X_{N-1}]^T$ is partitioned into V disjoint subblocks as follows:

$$X = [X^0, X^1, X^2, \cdots X^{V-1}]^T, \tag{3}$$

where X^v denotes the subblocks that are consecutively located and are of equal size. Each subblock rotates its phase independently by multiplying the corresponding complex phase factor $b_v = e^{j\phi_v}, v = 1, 2, \ldots, V$, then merge all the subblocks to obtain

$$X' = \sum_{v=1}^{V} b_v X_v. \tag{4}$$

The phase vector b_v, also known as the side information, is required to be sent to the receiver in order to recover the transmitted data correctly. Subsequently taking its IFFT to get

$$x = IFFT\{\sum_{v=1}^{V} b_v X_v\} = \sum_{v=1}^{V} b_v \bullet IFFT\{X_v\} = \sum_{v=1}^{V} b_v x_v. \tag{5}$$

The phase vector b_v should be chosen so that the PAPR of the OFDM symbols can be minimized, which can be shown as

$$\{\tilde{b}_1, \tilde{b}_2, \cdots, \tilde{b}_v\} = \arg\min \left(\max_{n=0,1,\cdots N-1} \left| \sum_{v=1}^{V} b_v x_v \right|^2 \right). \tag{6}$$

Theoretically, the value of b_v can be chosen as any possible value belonging to $[0, 2\pi)$, making sure the PAPR is minimum. As the set of allowed phase factor $b = \{e^{j2\pi i/U} | i = 0, 1, \ldots U - 1\}$, U^{V-1} sets of phase factors should be searched to find the optimum set of phase vector. Therefore, the search complexity increases exponentially with the number of subblocks. Such a huge amount of search computation is a heavy burden for the UAC system. So, some researchers fixed the phase factor to a certain set of elements to reduce the search complexity. A common way is to choose b_v from the set $\{1, -1\}$ [23], however, the fixed set will reduce the PAPR reduction performance. In this paper, a pseudo-optimum phase rotation candidate selection scheme is proposed to overcome this problem. Another disadvantage of the C-PTS scheme is the transmission of the Side Information (SI), which causes the loss of data rate, therefore a blind SI detection scheme has also been proposed to avoid the transmission of SI.

TABLE 1. Example of pilot tones distribution table.

Index	Pilot tones Position	Phase Rotation Candidate
$m = 1, 2, \cdots, M$	$\mathbf{Po}^m = \{i_0^m, i_1^m, \cdots, i_{N_p-1}^m\}$	$b_v^m = \{e^{j2\pi i_1^m/U}, e^{j2\pi i_2^m/U}, \cdots e^{j2\pi i_V^m/U}\}$
1	$\mathbf{Po}^1 = \{3, 6, \ldots, 1023\}$	$e^{j2\pi i_1^1/U}, e^{j2\pi i_2^1/U}, \cdots e^{j2\pi i_V^1/U}$
2	$\mathbf{Po}^2 = \{4, 5, \ldots, 1024\}$	$e^{j2\pi i_1^2/U}, e^{j2\pi i_2^2/U}, \cdots e^{j2\pi i_V^2/U}$
…	…	…
M	$\mathbf{Po}^M = \{1, 13, \ldots, 1018\}$	$e^{j2\pi i_1^M/U}, e^{j2\pi i_2^M/U}, \cdots e^{j2\pi i_V^M/U}$

III. PRINCIPLE OF THE PROPOSED-PTS SCHEME

A. PSEUDO-OPTIMUM PHASE ROTATION CANDIDATE SELECTION SCHEME

The pilot tones distribution table is designed as follows:

1) Assume V denotes the number of the subblocks.
2) Define M as the total number of the pilot tones distributions, $\mathbf{Po}^m$ is the position of the comb pilot tones. $\mathbf{Po}^m$ can be non-uniform because of the employment of sparse channel estimation.
3) The decision the candidate vectors of the phase rotations are randomly selected from the set of U^{V-1} vectors. The length of the vector equals the number of subblocks which is defined in 1), and the number of the candidate vectors equals the number of pilot tones distributions in 2).

The example of pilot tones distribution is shown in Table. 1.

Take one OFDM symbol as an example to explain the principle of Pseudo-optimum phase rotation candidate selection. The input data block X_D is partitioned into V disjoint subblocks as $X_D = [X_D^0, X_D^1, \cdots, X_D^V]$, using the pseudo-random partitioning schemes. The modified data on the non-pilot tones after multiplexing the M corresponding complex phase factor b_v^m can be written as follows:

$$\mathbf{X}_D^m = \sum_{v=1}^{V} b_v^m X_{vD} \tag{7}$$

Then the OFDM symbol with the comb pilot X_P in the frequency domain can be represented as:

$$X^m(k) = aX_D^m(k) + bX_P(k), \tag{8}$$

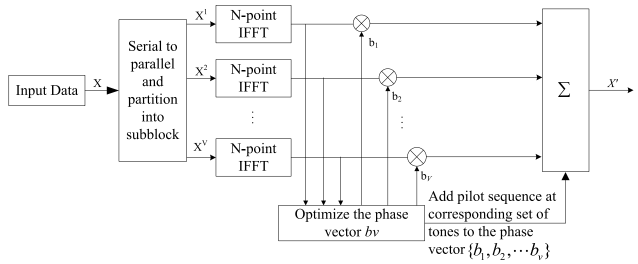

FIGURE 2. The framework of the proposed-PTS technique for PAPR reduction in underwater acoustic OFDM system.

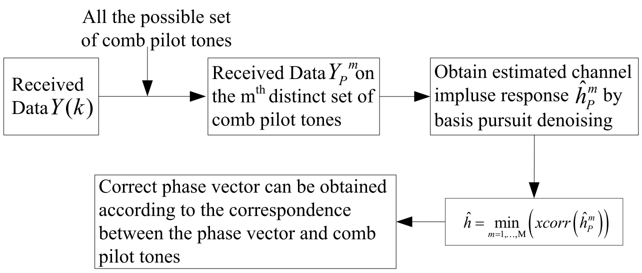

FIGURE 3. The process of the blind phase vector detector for the proposed-PTS technique at the receiver.

where

$$\begin{aligned} a &= 1, \quad b = 0 \quad k \notin \mathbf{Po}^m \\ a &= 0, \quad b = 1 \quad k \in \mathbf{Po}^m. \end{aligned} \tag{9}$$

Subsequently, IFFT is taken to generate the M OFDM signals in time domain. Finally, the signal with the lowest PAPR $\tilde{x} = x^{\tilde{m}}$ can be chosen as the transmitted signal. The framework of the proposed PTS technique using the pseudo-optimum phase rotation candidate selection scheme is shown in Fig. 2.

B. BLIND PHASE VECTOR DETECTOR

The process of the blind phase vector detector at the receiver for the proposed PTS technique is illustrated in Fig. 3.

After taking the fast Fourier transform (FFT) operation at the receiver, the received signal in the frequency domain can be given as

$$Y(k) = X(k)H(k) + W(k), \tag{10}$$

where $H(k)$ denotes the frequency response of the underwater acoustic channel and $W(k)$ represents the white Gaussian noise in the frequency domain. Denote $\hat{H}^m(k)$ as the estimation of $H(k)$, since the value of the pilot tones and all the M pilot tones distributions $\mathbf{Po}^m$ are known as Y_P^m, the sampling value of the underwater acoustic channel frequency response $\hat{H}_P^m$ can be achieved by

$$\hat{H}_P^m = \frac{Y_P^m}{X_P}. \tag{11}$$

Now we adopt the Sparse Reconstruction by Separable Approximation (SpaRSA) algorithm based on Basis Pursuit De-Noising (BPDN) to estimate the underwater acoustic channel [24]. Compared with other algorithms, sparse signals of BP have the ultimate global solution, so the standard $\ell_2 - \ell_1$ case is adopted. At the same time, we consider the effect of noise, which belongs to the observation values, and manipulate the balance between the sparse signals and residual by adjusting a regularization parameter. Then the SpaRSA algorithm is defined by the following process.

- Initializing

1) $y = Y_P^m$ stands for the received signal on the pilot tones, the transmitted pilot can be written as $A = X_P$, and the noise variance is σ^2.
2) Denote constant $I = 5$ and $\tau = 0.01$, factor $\zeta = 0.2$, $\eta > 1$ and variable $\alpha_0 = 1$. Regularization parameter $\lambda = 0.1 * \sigma^2 * \left\| A^T y \right\|_\infty$, where $(\bullet)^T$ denotes the matrix transposition.
3) Iteration times $t = 0$, Inner iteration times $i = 0$, and the Initial value of Channel Impulse Response (CIR) $h_{BP}^0 = 0$

- Step 1: $y^0 = y$.
- Step 2: $\lambda_t = \max\{\zeta \left\| A^T y^t \right\|_\infty, \lambda\}$.
- Step 3: choose h_{BP}^{t+1}.

1) Calculate h_{BP}^{i+1}:

$$x^{i+1} \in \arg\min_z \frac{1}{2} \|z - u_i\|_2^2 + \frac{\lambda_t}{\alpha_i} c(z) = soft(u_i, \frac{\lambda_t}{\alpha_i})$$
$$i = 0, 1, 2 \ldots$$
$$u_i = h_{BP}^i - \frac{1}{\alpha_i} \nabla f(h_{BP}^i) = h_{BP}^i - \frac{1}{\alpha_i} A^T (A h_{BP}^i - y)$$

Where $soft(a, b) = \frac{\max\{|a| - b, 0\}}{\max\{|a| - b, 0\} + b} a$.

2) If $\phi(h_{BP}^{i+1}) \le \max_{i = \max(t - I, 0), \ldots, t} \phi(h_{BP}^i) + \frac{\tau}{2} \alpha_i \left\| h_{BP}^{i+1} - h_{BP}^i \right\|_2^2$, continue, else, $\alpha_i = \eta \alpha_i$, and turn to (1).
Where $\phi(h) = \frac{1}{2} \left\| y^t - X_P h \right\|_2^2 + \lambda \|h\|_1$.
3) Update α_{i+1}, $\alpha_{i+1} = \frac{\left\| A(h_{BP}^{i+1} - h_{BP}^i) \right\|_2^2}{\left\| h_{BP}^{i+1} - h_{BP}^i \right\|_2^2}$
4) If $\frac{\left| \phi(h_{BP}^{i+1}) - \phi(h_{BP}^i) \right|}{\phi(h_{BP}^i)} \le \varepsilon$, continue to step 4, else, $i = i + 1$, turn to (1).

- Step 4: $y^{t+1} = y - A h_{BP}^{t+1}$.

Step 5: If $\lambda_t = \lambda$, end the iteration, $\hat{h} = h_{BP}^{t+1}$ is the output; else $t = t + 1$, go again to Step 2. As there is M pilot tones distributions, all the steps of SpaRSA algorithm need to be repeated M times to obtain all the possible CIR $\hat{h}^m$, among them there exists only one match $\hat{h} = \min_{m=1,\ldots,M} \left(xcorr \left(\hat{h}_{BP}^m \right) \right)$, and it is the real CIR of the underwater acoustic channel. According to the pilot tones distribution table, the number of selected phase rotation vector at the transmitter can be found.

IV. NUMERICAL SIMULATION AND FIELD EXPERIMENT RESULTS

A. SIMULATION PARAMETERS

The main parameters of the UWA OFDM system are given in Table 2. The total number of OFDM symbols is 10^6,

TABLE 2. Simulation parameters of the underwater acoustic OFDM system.

Parameters	Value
Transmission Frequency Band	6000-12000 Hz
Bandwidth	6000 Hz
Sampling Frequency	48 kHz
Number of Subcarriers	1024
Subcarrier Bandwidth	5.859375Hz
Number of Bits per Subcarrier	2(QPSK modulation)
Guard Interval of the cyclic prefix	43 ms
Symbol Duration	170.6 ms
Pilot Spacing	4
Code Rate	1/2
Subblock Number	2&4&8

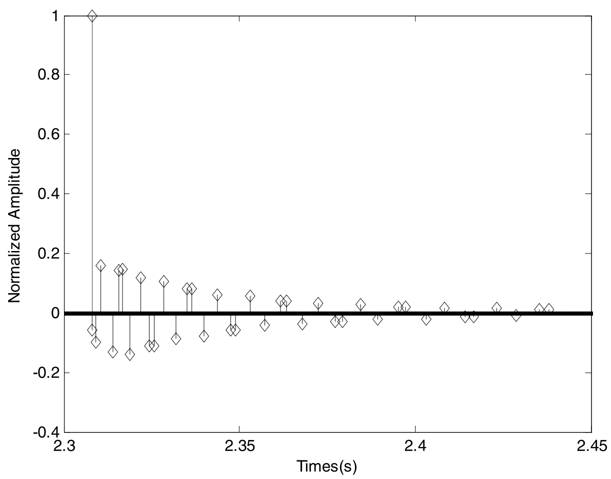

FIGURE 4. The impulse response of the simulation channel.

which are randomly generated. The generator matrix of the convolutional code is [171,131], and the subblock partitioning scheme is pseudo-random in order to get the best PAPR reduction performance.

A sparse channel generated from channel simulation software is adopted in the simulation to evaluate the system BER performance. It is assumed that the depths of the transducer and hydrophone are 20m and 9m respectively. The distance between the transducer and hydrophone is 3150m, and the average depth of the sea is 55m. Velocity gradient distribution is set as a surface channel. The impulse response of the simulation channel is illustrated in the Fig.4. It can be seen that the direct sound intensity is higher than any other multipath, and the number of multipath is a lot while the intensity of each path is muted.

In these simulations, C-PTS scheme given in [23] is chosen as a control scheme, which phase rotation vectors are selected in the set of $\{1, -1\}$ and fixed $b_v = 1$ for $v = 1$. Therefore, the number of possible phase rotation candidates is 2^{V-1}, while it is M in the proposed pseudo-optimum phase

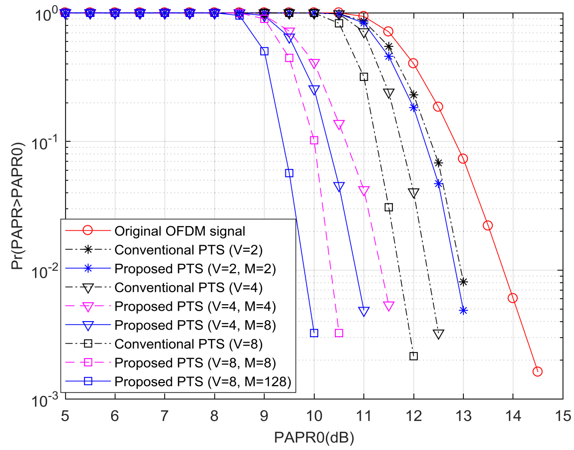

FIGURE 5. PAPR distribution of the original signal, the C-PTS technique and proposed PTS technique with a various number of subblocks.

rotation candidate selection scheme. Since the computational complexity is mainly due to the calculation of Equation (5), the number of computations for Equation (5) Com_{Num} is used as a measure of the computational complexity. The number of computation Equation (5) equals to the subblock number $\mathrm{Com}_{Num} = V$ in C-PTS scheme while it is $\mathrm{Com}_{Num} = M$ in the proposed PTS technique. The computational complexity of the proposed PTS technique and the C-PTS scheme is the same when $M = V$, however, the number of phase rotation candidates is $V/2^{V-1}$ times to the C-PTS scheme. The computational complexity of the proposed PTS technique is $2^{V-1}/V$ times to the C-PTS scheme when $M = 2^{V-1}$, while the number of phase rotation candidates is identical to the C-PTS scheme. Therefore, M is set to V and 2^{V-1} for different values of the subblock number V in simulations of the proposed-PTS technique.

In Fig.5, the CCDF distribution of the original signal, the C-PTS technique and the proposed PTS technique are plotted for different numbers of the subblock. The outer curve is the original OFDM signal with fully interleaved input data. The two curves next to the outer one express the C-PTS technique and the proposed PTS technique with the subblock number $V = 2$. It is because that $M = 2^{V-1} = V = 2$ for $V = 2$, the computational complexity and the number of possible phase rotation candidates of the proposed PTS scheme are the same as the C-PTS scheme. The PAPR reduction performance is only a little bit improved. The three curves with the inverted triangle mark represent the case of the subblock number $V = 4$. The dash-dot one is the C-PTS scheme, the dash one is the proposed PTS technique with $M = V = 4$, and the solid one is the proposed PTS technique with $M = 2^{V-1} = 8$. The rest of the three curves with the square mark are the case of the subblock number $V = 8$ and the line type has the same meaning subblock number $V = 4$ case. It can be observed that the proposed PTS technique has approximately the same PAPR reduction performance as the conventional ones when $V = 2$, and it has about 0.3dB gains when $V = 4$. It can

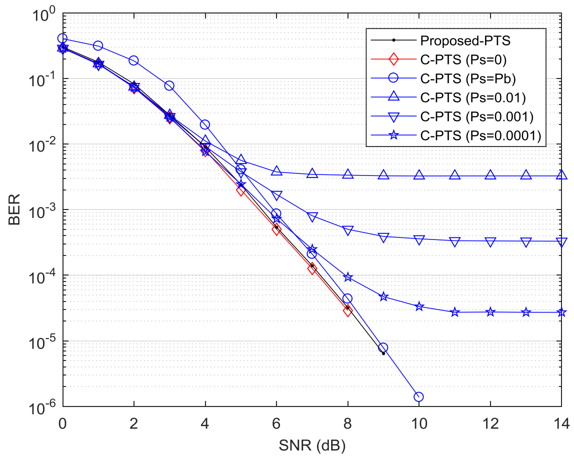

FIGURE 6. Comparisons of the overall BER performance between the proposed PTS technique and C-PTS technique for several error probabilities of the side information.

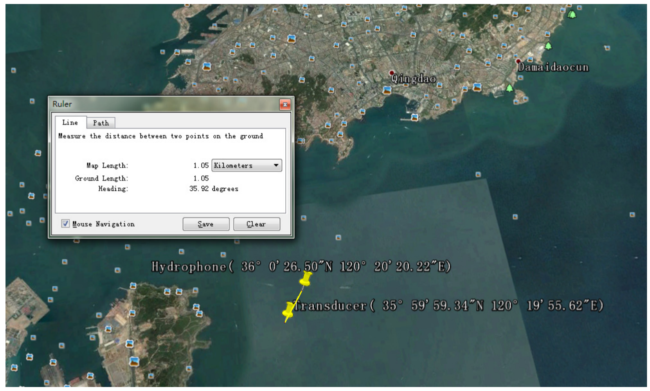

FIGURE 7. Location of the field experiment (Distance between the Transducer and the Hydrophone is about 1km).

be observed that the performance gap is about 1dB for $M = V = 4$ and it is almost 1.5dB for $M = V = 8$ when the computational complexity of the two techniques is the same. The performance difference is approximate 1.5 dB for $V = 4\&M = 8$, and it is about 2 dB for $V = 8\&M = 128$, when the number of phase rotation candidates are the same with the C-PTS scheme. It is the randomness brought by the choice of phase rotation vectors which makes the PAPR reduction performance improved. The PAPR reduction gain of pseudo-optimum PTS technique gradually increases with the increase in the subblock number. It also increases with the increase in the phase rotation candidate number when the subblock number stays the same.

When the signal is received, the receiver must recover the side information first to let the other symbols be detected correctly. The side information plays an important role in the overall system BER performance. P_s is defined as the error probability of the side information, which may control the overall system bit error probability P.

$$P = P_b(1 - P_s) + P_{b|\text{False}} \bullet P_s \tag{12}$$

P_b is the bit error probability of data symbol with perfect side information, and $P_{b|\text{False}}$ is the conditional bit error probability given that the side information is false. It can be considered as 50%.

Fig.6 compares the performance of UAC OFDM system utilizing the proposed PTS technique, in terms of BER, to the cases where the side information error probability at the receiver is chosen from the set$\{10^{-2}, 10^{-3}, 10^{-4}, P_b, 0\}$. It can be seen that if the side information error probability is at a high level, i.e. $P_s = 10^{-2}$, $P_s = 10^{-3}$, and $P_s = 10^{-4}$, there exists serious error floor. In C-PTS technique and most usual case, P_s can be regarded as equal asP_b. When $P_s = 0$ the overall BER is only affected by the bit error probability P_b. The gap between the two curves of $P_s = P_b$ and $P_s = 0$ reduces with the increase of SNR. We can see that the proposed PTS technique has approximately the identical BER performance as PTS with perfect side information. It is indicated that the blind side information detector can provide excellent side information to the receiver. As the error caused by mistranslating the side information can be avoided, the proposed scheme can significantly enhance the communication efficiency. The receiver of the C-PTS scheme needs to wait until the entire signal or the whole frame be received, to obtain and decode the SI. Then the other symbols which carry the transmitted data can be decoded. Thus, it needs a large number of storage resources and a long time to wait. However, the blind phase vector detector can recognize the phase vector automatically on a symbol by symbol basis. In other words, the proposed PTS scheme can detect the side information of a symbol only by the data in this symbol. There is no need to wait for the entire signal to be received so that the proposed scheme can guarantee the real-time performance of the UAC system.

B. FIELD EXPERIMENT RESULTS

To evaluate the feasibility of the proposed PTS algorithm, a field experiment was conducted at the Huanghai Sea, Qingdao, China and the location is shown in Fig.7. During this experiment, both of the boats holding the transducer and the hydrophone remained in Anchorage. The depth of the transducer and receiver were 7m and 4m under the sea surface respectively. The Huanghai Sea was under state 3 conditions throughout the field experiment period. As the field experiment was carried out near the main vessel, the noise level is a bit high and the received signal to noise ratio is lower than it away from the vessel path.

The channel impulse responses (CIR) of the field channel as a function of time are presented in Fig.8 (a). At the beginning of each packet, a linear frequency modulated (LFM) signal is placed as probe signal for the purpose of symbol synchronization, and the estimation of the CIR for each packet can be obtained by matched filtering the received LFM data with the transmitted LFM signal. By stacking the measured CIRs against transmitted time, the channel variations on the time have been obtained [25]. It is noticed that the CIR has

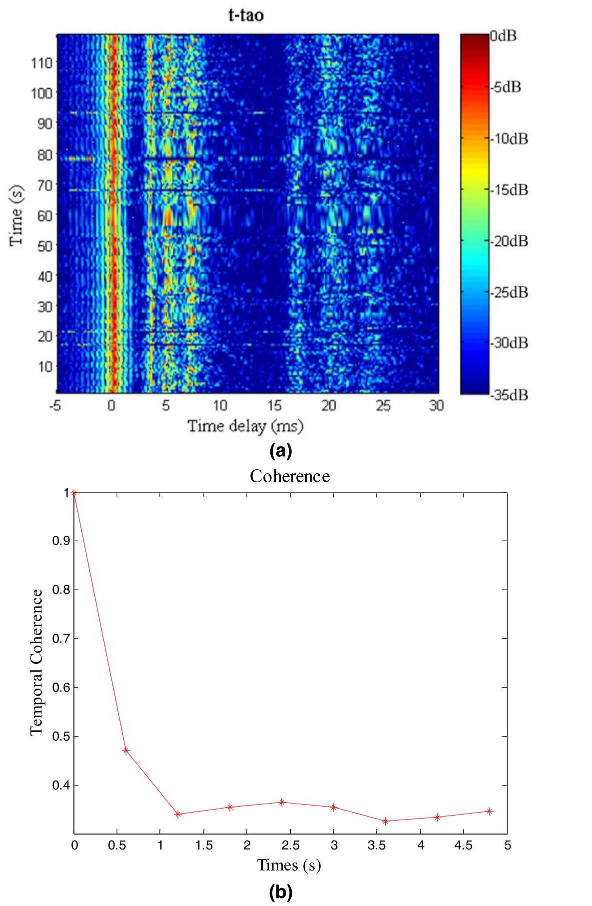

FIGURE 8. (a): The Channel Impulse Response of the field channel which is determined by LFM signals as a function of time. (b): The Temporal coherence of the LFM signal under a slight-sea condition at the Huanghai Sea.

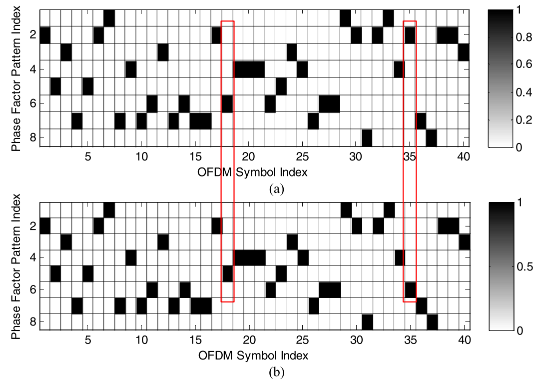

FIGURE 9. Experimental result of phase rotation candidate index detection using the C-PTS technique, which requires SI. (a): The original version of side information at the transmitter. (b): The translated version autonomous recognized by the blind phase vector detector.

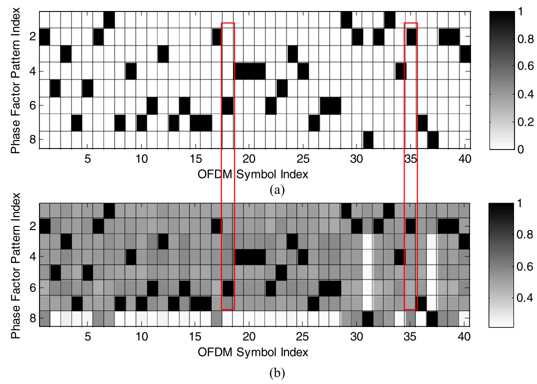

FIGURE 10. The blind detected results of phase rotation candidate index detection using the proposed PTS technique. (a): The original version of side information at the transmitter. (b): The translated version autonomous recognized by the blind phase vector detector.

seven paths, consisting direct sound, reflexed sound from the surface, reflexed sound from the bottom, reflexed sound from both the surface and the bottom, and the second reflexed sound. Each path fluctuates with respect to time under the sea state 3 condition and the existence of small waves cause the CIR vary temporally. The estimation of the temporal coherence of the channel can be obtained by using the measured CIR at different times as the reference signal as shown in Fig.8 (b). As the sea surface is not calm, the temporal coherence decreases rapidly, so the channel experienced by each symbol can be regarded as independent.

Fig.9 shows the experimental result of the phase rotation candidate index detection of the first 40 OFDM symbols in the transmitted signal utilizing the C-PTS technique which needs to transmit SI. The upper block is the original version of the rotation phase vector index at the transmitter, while the block below shows the decoded version of transmitted side information at the receiver. The black block means the selected or decoded weighting factor index. The symbol which transmits side information has the same mapping function, the same generator matrix of convolution code and the same modulation function with the data symbols, which means $P_s = P_b$. Received SNR is about 9.59dB in this experiment. The rotation phase vector index of the symbol in the red box is mistranslated because of the multi-path channel and the environmental noise (overall P_s is in this field experiment). This will cause decoding disaster because of the BER of those symbols with wrong side information is approximately 0.5, and the overall BER performance of the UAC OFDM system will be affected severely.

Fig. 10 shows the blind detected results of the phase rotation candidate index detection of the identical 40 OFDM symbols as C-PTS technique while exploiting the proposed PTS technique. The gradation value indicates the probability of the phase factor combination pattern index being the selected one at the transmitter. Fig.10 (a) is same as it is shown in Fig.9 (a), and Fig.10 (b) is the translated version recognized by blind side information detector autonomously at the receiver. It can be easily found that the phase factor combination pattern index of the symbol in red box has not been mistranslated. It is verified that the blind phase vector index detector in the proposed PTS technique can differentiate among all the permissible combinations of phase rotation

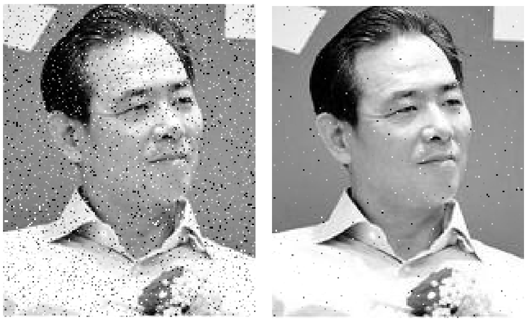

FIGURE 11. The figures restored by the received data after decoding in this field experiment. The left picture is the result of C-PTS technique which transmitted SI, and the right figure is the result of proposed-PTS technique which needs no SI.

factors, even if channel state information and side information remain unknown. The overall BER performance of the UWA OFDM communication system could achieve a reliable level.

In this field experiment, the original data comes from a grey-scale image, and the figures restored by the received data after decoding are shown in Fig. 11. It can be found that the existence of SI error makes the BER of C-PTS method almost a magnitude higher than the proposed PTS technique when comparing these two figures. It can be concluded that the proposed PTS technique, which requires no side information, can guarantee a reliable communication when the received SNR is about 9.5dB through a UWA multipath channel.

V. CONCLUSION

This paper proposes a novel PTS PAPR reduction technique which has pseudo-optimum PAPR reduction performance and requires no side information transmission in UAC OFDM system. The proposed PTS technique can significantly reduce the computational complexity by employing a corresponding table between the distributions of pilot tones and the phase rotation candidates. It also modifies the PAPR reduction performance because of the random selection of the phase rotation vectors, while the C-PTS fixed the phase rotation vectors only in the set $\{1, -1\}$. Due to the sparse characteristic of the underwater acoustic channel, compressed sensing is adopted to complete the channel estimation and the phase rotation vectors index detection. Then the proposed PTS technique requires no side information transmission and no additional pilot tones except for the tones used for channel estimation, therefore avoiding any data rate loss. As there is no need to wait for receiving the whole packet to obtain the SI, the proposed PTS scheme can guarantee the real-time performance of the UAC OFDM system. Simulation results show that the proposed scheme has better PAPR reduction performance as compared to the conventional PTS scheme and the performance gap increases with the number of subblocks. The BER performance of the PTS technique is proved to be approximately identical to the PTS with perfect side information, which indicates that the blind SI detector can provide excellent side information to the receiver. Field experimental results also demonstrate that the proposed scheme can differentiate the phase rotation factor, therefore the quality of the UAC OFDM system is significantly enhanced.

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