Synchronization and Doppler scale estimation with dual PN padding TDS-OFDM for underwater acoustic communication

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***Abstract*—We propose a joint synchronization and Doppler scale estimation method with time domain synchronous orthogonal frequency division multiplexing (TDS-OFDM) for underwater acoustic communication. Dual frequency-domain pseudo-random noise (PN) sequences serve as training sequences for synchronization and channel estimation. Frame synchronization, carrier frequency offset (CFO) estimation and Doppler scale estimation are obtained simultaneously by two-dimensional auto-correlations of received signals. Theoretical analyses and computer simulations show that the proposed method is suitable for synchronization and Doppler scale estimation, even in the presence of long delay spread and significant Doppler effects.**

I. Introduction

Underwater acoustic (UWA) channels present many challenges for reliable high data-rate communication. Channels are characterized by long delay spread and significant Doppler effects, which qualifies as doubly selective (time-selective and frequency-selective) channel [1]. The long channel delay spread leads to severe inter-symbol interference (ISI) in single-carrier transmissions. Multicarrier approaches like orthogonal frequency division multiplexing (OFDM) can equalize the channel at low complexity, but significant Doppler effects lead to severe inter-carrier interference (ICI).

Training-based channel estimation methods are commonly used in wireless communication. Time domain synchronous orthogonal frequency division multiplexing (TDS-OFDM) outperforms the traditional ZP-OFDM and CP-OFDM in terms of spectrum efficiency [2]. Dual frequency-domain pseudo-random noise (PN) sequences are adopted as training sequences as well as guard intervals for the multipath channel, which improves accuracy of channel estimation [3]. TDS-OFDM is introduced to underwater acoustic communication with compressed sensing channel estimation in [4].

Synchronization and Doppler scale estimation are essential tasks in underwater acoustic communication. There are three typical parts at the receivers, i.e. frame synchronization, carrier frequency offset (CFO) estimation and Doppler scale estimation. The frame and sampling frequency synchronization method is applicable to block transmission systems with fixed training sequence [5]. A joint synchronization method based on cyclic property achieves frame, sampling frequency and carrier frequency synchronization [6]. Instead of CFO estimation, the rotating phases caused by multiple CFOs is estimated in [7]. A method for detection, synchronization and Doppler scale estimation for underwater acoustic communication is proposed in [8], which eliminates the dual LFM signals, and uses two identical OFDM symbols as a cyclic prefix.

This paper propose a joint synchronization and Doppler scale estimation method for underwater acoustic communication. Frame synchronization, carrier frequency offset estimation and Doppler scale estimation are implemented simultaneously. Dual frequency-domain PN sequences serve as training sequences. Two-dimensional auto-correlations of received signals are calculated to implement the proposed method. Phase compensation is adopted to remove the rotating phases from received data, based on the results of CFO estimation.

II. System model

Each PN sequence is $\mathbf{s} = [s[0], s[1], \cdots, s[N-1]]^T$, where N is the length of PN sequence. The transmitted signals with TDS-OFDM structure are shown in Fig. 1. Shaded areas represent preambles, which are N-point inverse fourier transforms of $\mathbf{s}$, i.e. $\mathbf{c}_1 = \mathbf{c}_2 = \mathrm{ifft}(\mathbf{s})$. Duration of each preamble is $2T_0$, and the i-th frame starts from t_i. The payload in the i-th frame is $\mathbf{d}^{(i)}$. As we concentrate on synchronization and Doppler scale estimation, the payloads are not discussed in this paper.

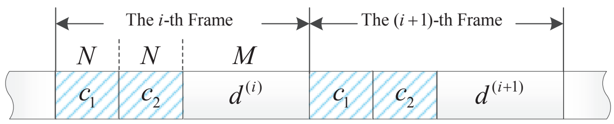

Fig. 1. Transmitted signal with TDS-OFDM structure.

The preamble in baseband can be written as

$$x(t) = \sum_{k=0}^{N-1} s[k] e^{j2\pi \frac{k}{T_0} t}, \ t \in [t_i, t_i + 2T_0). \tag{1}$$

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The corresponding passband signal is

$$\tilde{x}(t) = \mathrm{Re}\left\{\sum_{k=0}^{N-1} s[k]e^{j2\pi\left(\frac{k}{T_0}+f_c\right)t}\right\},\ t \in [t_i, t_i + 2T_0) \quad (2)$$

where f_c is the carrier frequency at the transmitter.

Channel impulse respond of a doubly selective channel can be described as

$$h(\tau, t) = \sum_p A_p(t)\delta(\tau - \tau_p(t)) \quad (3)$$

where $A_p(t)$ is the amplitude of path p, and $\tau_p(t)$ is the path delay. As in [8], the following two assumptions are adopted for underwater acoustic channels.

A1 All paths have a similar Doppler scaling factor a such that

$$\tau_p(t) = \tau_p - at. \quad (4)$$

A2 The path delay τ_p, path amplitude A_p, and Doppler scaling factor a are constant during the preamble duration $2T_0$.

The passband received signal through channel described above can be expressed as

$$\tilde{y}(t) = \mathrm{Re}\left\{\sum_{k=0}^{N-1} s[k]H_k e^{j2\pi\left(\frac{k}{T_0}+f_c\right)(1+a)t}\right\} + \tilde{w}(t), \quad t \in [t_i, t_i + \frac{2T_0}{1+a}) \quad (5)$$

where $\tilde{w}(t)$ is the additional white Gaussian noise, and the channel transfer function is defined as

$$H_k = \sum_p A_p e^{-j2\pi\left(\frac{k}{T_0}+f_c\right)\tau_p}. \quad (6)$$

Converting to baseband, the received signal can be written as

$$y(t) = e^{j\Omega t}\sum_{k=0}^{N-1} s[k]H_k e^{j2\pi\frac{k}{T_0}(1+a)t} + w(t) \quad (7)$$

where $\Omega = 2\pi(af_c + f_{cr} - f_c)$ is the carrier frequency offset (CFO), and f_{cr} is the carrier frequency at the receiver. Repetition pattern persists in the baseband received signal as

$$y(t + T_1) = e^{j\Omega T_1}y(t), \quad t \in [t_i, t_i + T_1) \quad (8)$$

where the repetition interval at the receiver is $T_1 = \frac{T_0}{1+a}$.

Sampled at the receiver, the baseband received signals are obtained as

$$y[n] = y(t)|_{t=t_i+nT_s} \quad (9)$$

where $T_s = \frac{T_0}{N_F}$ is the sampling interval, and $F = \frac{N_F}{N}$ is the oversampling factor. The differences between repetition interval at the transmitter T_0 and repetition interval at the receiver T_1 cause a relative time shift θ. Then T_1 can be rewritten as

$$T_1 = T_0 + \theta T_s = N_F T_s + (\beta + \mu)T_s \quad (10)$$

where β is the integer part of θ, and $\mu \in [-0.5, 0.5)$ is the fractional part of θ.

III. The Proposed Algorithm

The amplitude and phase of two-dimensional auto-correlations $R[n, l]$ are adopted to implement frame synchronization, carrier frequency offset estimation and Doppler scale estimation.

The two-dimensional auto-correlations are calculated as

$$R[n, l] = \sum_{m=0}^{N_F+l-1} y^*[n + m]y[n + m + N_F + l] \quad (11)$$

where

$$y[n + N_F] = y\left(t_i + nT_s + T_1 - \theta T_s\right). \quad (12)$$

From (8), we obtain

$$y[n + N_F] = e^{j\Omega T_1}y[n] * h_I(-\theta T_s) \quad (13)$$

where $h_I(-\tau_h)$ is the impulse response of ideal interpolator with delay τ_h [9]. Thus the correlations can be rewritten as

$$R[n, l] = e^{j\Omega T_1}G[n, l] * h_I(-\theta T_s) \quad (14)$$

where

$$G[n, l] = \sum_{m=0}^{N_F+l-1} y^*[n + m]y[n + m + l] \quad (15)$$

are auto-correlations of baseband received signal.

To reducing implementation complexity, the correlations in (11) can be computed recursively as

$$\begin{aligned} R[n + 1, l] =& R[n, l] - y^*[n]y[n + N_F + l] \\ &+ y^*[n + N_F + l]y[n + 2N_F + 2l]. \end{aligned} \quad (16)$$

A. Frame synchronization

The 2D-peak position of $R[n, l]$ is obtained as

$$[n_p, l_p] = \arg\max_{n,l} |R[n, l]| \quad (17)$$

where n_p represents the start position of preamble, and $l_p = \hat{N} - N_F$ is the difference between the sampling preamble length N_F and estimation of sampling preamble length $\hat{N}$ at the receiver.

B. Doppler scale estimation

The integer part β is estimated from 2D-peak position as

$$\hat{\beta} = l_p. \quad (18)$$

Assume $R(l) = R[n_p, l]$. Let $R'(l)$ be the derivative of $R(l)$ at l. Theoretically, the fractional part μ is estimated as

$$\hat{\mu} = \frac{R'(l_p) - R'(l_{\max})}{R'(l_p - 1) - R'(l_p)} = \frac{R'(l_p)}{R'(l_p - 1) - R'(l_p)} \quad (19)$$

where l_{max} is the peak position of $|R(l)|$, and $R'(l_{max}) = 0$. With the discrete time signal, we can estimate μ as

$$\hat{\mu} = \frac{R_d[l_p]}{R_d[l_p - 1] - R_d[l_p]} \quad (20)$$

where numerical difference $R_d[l]$ measures the asymmetry of $R[n_p, l]$ as

$$R_d[l] = \frac{|R[n_p, l+1]| - |R[n_p, l-1]|}{2}. \quad (21)$$

The Doppler scaling factor changes duration of preamble from T_0 to T_1. Thus, the Doppler scaling factor can be expressed as

$$\hat{a} = -\frac{\hat{\beta} + \hat{\mu}}{N_F + \hat{\beta} + \hat{\mu}}. \quad (22)$$

C. Carrier frequency offset estimation

The phase of two-dimensional auto-correlations can be expressed as

$$\arg\{R[n, l]\} = \lfloor \Omega T_1 + \arg\{G[n, l] * h_I(-\theta T_s)\} \rfloor_{2\pi} \quad (23)$$

where $\lfloor \cdot \rfloor_{2\pi}$ is the modulo operator within $[-\pi, \pi)$. Since the received signal is oversampled by factor F, the angle of $G(n_p, l)$ can be approximated as

$$\arg\{G(n_p, l)\} = \Omega T_s l, \quad l \in [-F, F]. \quad (24)$$

Thus, we obtain

$$\arg\{R(n_p, l)\} = \lfloor \Omega T_1 + \Omega T_s l - \Omega T_s(\beta + \mu) \rfloor_{2\pi} \quad (25)$$

The CFO can be estimated as

$$\hat{\Omega}_p = \frac{\arg\{R(n_p, l_p)\}}{(N_F + l_p) T_s}. \quad (26)$$

Since N_F is much larger than l_p, the range of $\hat{\Omega}_p$ can be approximated as $[-\frac{\pi}{N_F T_s}, \frac{\pi}{N_F T_s})$.

IV. Simulation Results

To confirm the theoretical results, we tested the performance of the proposed method in flat fading channel and in multipath channel. Doppler spread of the flat fading channel was 0.2 Hz. Channel impulse respond of the multipath channel is shown in Table I. The length of PN sequence was 512, and 8-times oversampling was adopted. For each frame, 3780-symbol data were modulated with QPSK, while the symbol rate was 10.0 Ksymbol/s. As the Doppler scaling factor can not be simulated directly, baseband received signals were generated by resampling and convolution with channel impulse respond, according to (7) and (9). All the values of carrier frequency offset were normalized to parts per million (ppm). Doppler scaling factors were also expressed in ppm.

Two-dimensional auto-correlations of 8-times oversampled received signal in multipath channel are shown in Fig. 2, where l varied from -20 to 20. The auto-correlations on discrete time dimension $R(n, l_p)$ and on correlation lag dimension $R(n_p, l)$ are shown in Fig. 3, respectively. The Doppler scaling factors were set to 0 ppm and 500 ppm for comparison. The 2D-peak position on the correlation lag dimension was shifted by two samples when the Doppler scaling factor was 500 ppm.

Simulations of CFO estimation are shown in the top sub-figure of Fig. 4, where the Doppler scaling factor was 500 ppm. The bottom subfigure of Fig. 4 illustrated MSEs of CFO estimation with different SNR. The simulation results were quite close to ideal values, and the accuracy of CFO estimation was acceptable.

Doppler scale estimations are shown in Fig. 5, where carrier frequency offset was 25 ppm. The simulation results had the same tendency with ideal values. However, the differences between simulation results and ideal values were unignorable.

Simulation results of bit error rate (BER) without channel encoding are shown in Fig. 6, where the Doppler scaling factor was 0 ppm, and carrier frequency offset was 25 ppm. LS

TABLE I
THE MULTIPATH CHANNEL

Multipath	Delay Time (ms)	Relative Amplitude (dB)
1	0	0
2	0.3	-12
3	3.3	-4
4	4.4	-7
5	9.5	-15
6	12.7	-22

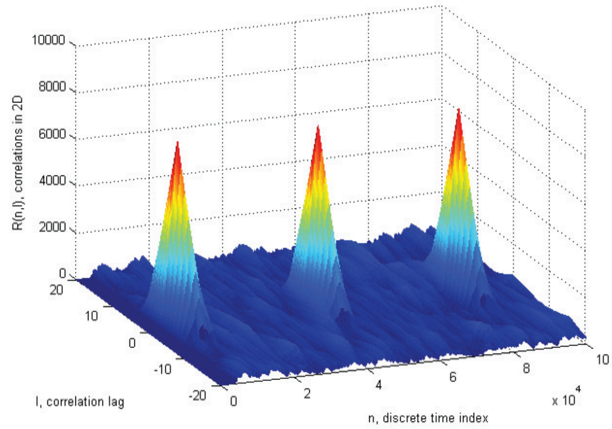

Fig. 2. Two-dimensional auto-correlations of the received signal in multipath channel, where the Doppler scaling factor was 0 ppm.

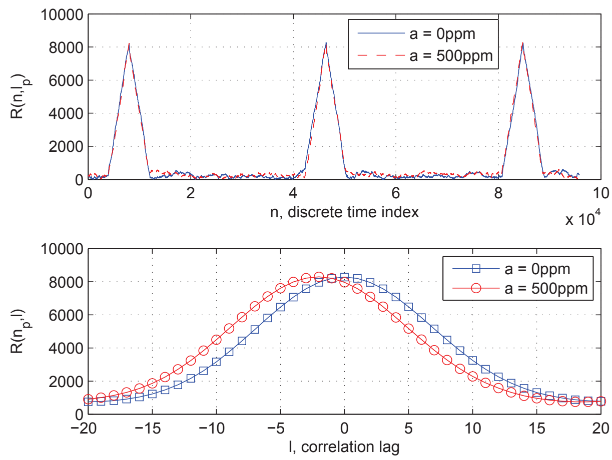

Fig. 3. Correlations result on discrete time dimension (top) and correlation lag dimension (bottom) in multipath channel. The Doppler scaling factors were set to 0 ppm and 500 ppm for comparison.

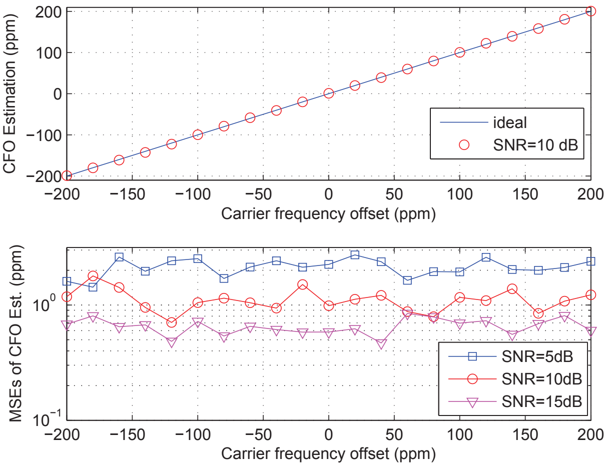

Fig. 4. CFO estimation results in multipath channel, where the Doppler scaling factor was 500 ppm, and SNR = 10 dB (top). CFO estimation MSEs with different SNR, where the Doppler scaling factor was 500 ppm (bottom).

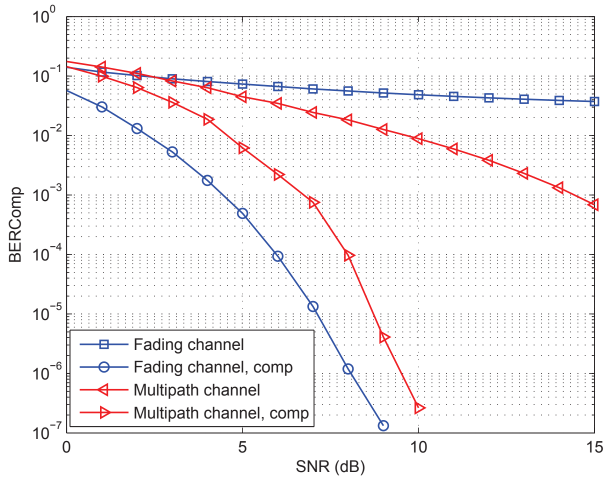

Fig. 6. BER performance in flat fading channel and in multipath channel, where the Doppler scaling factor was 0 ppm, and carrier frequency offset was 25 ppm.

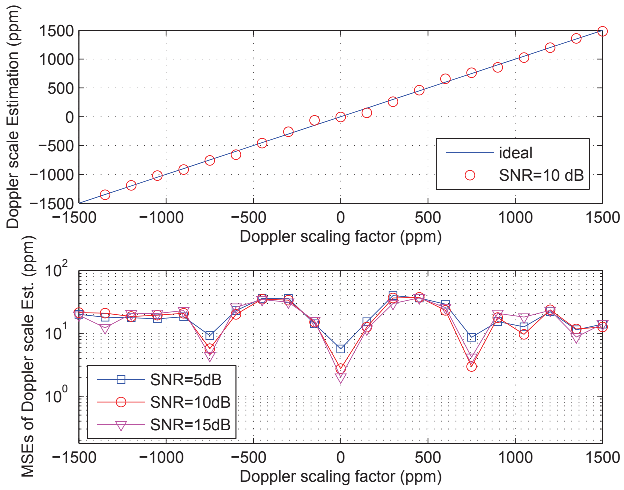

Fig. 5. Doppler scale estimation results in multipath channel, where CFO = 25 ppm, and SNR = 10 dB (top). MSEs of Doppler scale estimation with different SNR, where CFO = 25 ppm (bottom).

channel estimation and frequency equalization were adopted. With phase compensation, the BERs reduced greatly with the increase of SNR, both in flat fading channel and in multipath channel.

V. Conclusion

Based on dual PN padding TDS-OFDM, a joint synchronization and Doppler scale estimation method has been proposed for underwater acoustic communication. Frame synchronization, carrier frequency offset estimation, and Doppler scale estimation are obtained simultaneously. Simulation results show that the proposed method is suitable for synchronization and Doppler scale estimation in flat fading channel and in multipath channel.

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