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An underwater acoustic direct sequence spread spectrum communication system using dual spread spectrum code*

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Abstract: With the goal of achieving high stability and reliability to support underwater point-to-point communications and code division multiple access (CDMA) based underwater networks, a direct sequence spread spectrum based underwater acoustic communication system using dual spread spectrum code is proposed. To solve the contradictions between the information data rate and the accuracy of Doppler estimation, channel estimation, and frame synchronization, a data frame structure based on dual spread spectrum code is designed. A long spread spectrum code is used as the training sequence, which can be used for data frame detection and synchronization, Doppler estimation, and channel estimation. A short spread spectrum code is used to modulate the effective information data. A delay cross-correlation algorithm is used for Doppler estimation, and a correlation algorithm is used for channel estimation. For underwater networking, each user is assigned a different pair of spread spectrum codes. Simulation results show that the system has a good anti-multipath, anti-interference, and anti-Doppler performance, the bit error rate can be smaller than 10^{-6} when the signal-to-noise ratio is larger than −10 dB, the data rate can be as high as 355 bits/s, and the system can be used in the downlink of CDMA based networks.

Key words: Underwater acoustic communication; Direct sequence spread spectrum; Doppler estimation and compensation; Channel estimation and equalization; Gold code; Single carrier; Code division multiple access

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1 Introduction

An underwater acoustic channel has the characteristics of a strong multipath effect, frequency-selective fading, high noise interference, low sound propagation velocity, and a narrow bandwidth. Because of the low sound propagation velocity, the Doppler frequency shift caused by the relative motion among platforms or seawater movements cannot be ignored. Because of the multipath effect, signal superposition at different levels of the receiver may cause serious inter symbol interference (ISI). Because of the frequency-selective fading and high noise interference, underwater acoustic communication failures may frequently occur (Chitre et al., 2004). Therefore, it is difficult and challenging to build stable and reliable underwater acoustic communication systems.

Currently, research on underwater acoustic communication technology is divided mainly into two categories: multi-carrier underwater acoustic (MC-UWA) communication technology, such as orthogonal frequency division multiplexing (OFDM) aiming to provide a high communication rate for a relatively short distance channel or a high quality

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channel, and single-carrier underwater acoustic (SC-UWA) communication technology which aims to provide stable and reliable communication for a relatively long distance channel. By dividing the channel frequency band into several sub-bands, OFDM uses the orthogonality of the sub-carriers to transmit data. Although it can greatly improve the system communication rate, this approach may lead to a high peak-to-average-power ratio (PAPR) of the system and high environmental requirements. As a result, the OFDM technology is suitable for applications involving relatively short distance communication and relatively good quality channel communication (Rojo and Stojanovic, 2010). Because it uses a single carrier to transmit data, SC-UWA communication has a low PAPR. Furthermore, with the introduction of the spread spectrum technique, the SC-UWA system has a strong capability to anti-multipath, anti-interference, and anti-frequency-selective fading (Yang and Yang, 2008). Thus, the SC-UWA communication technology with the spread spectrum technique is one of the solutions for long distance stable and reliable communication in a complex environment (Sozer et al., 1999; Yoshizawa et al., 2017). Moreover, the direct sequence spread spectrum (DS/SS) scheme is a good approach for implementing code division multiple access (CDMA) based underwater acoustic networks (Yang, 2016).

For most DS/SS underwater acoustic communication systems, the information data are modulated by a pseudo random sequence, such as the gold sequence, the m sequence, and the chaotic sequence (Heidari-Bateni and McGillem, 1994; Palmese et al., 2007; Liu et al., 2017). A wideband signal is sent to the channel after carrier modulation. For a binary sequence, such as the gold sequence and the m sequence, the received signal is demodulated by a correlation operation of the received signal and a copy of the pseudo random sequence used by the transmitter (Huang and Sang, 2007; Zhang et al., 2007). For a non-binary sequence, such as the chaotic sequence, dual unscented Kalman filtering is used to recover the transmitted signal (Luca et al., 2005). Moreover, the chaotic sequence can be used in a secure communication system (Shu et al., 2015). To improve the communication rate, the M-ary code shift keying scheme (Sun et al., 2014) uses different spread sequences to modulate the different information data blocks in a point-to-point communication system and recovers the original data by using different sequences to de-spread the received signal. DS/SS technology not only supports point-to-point communication, but also is an effective method for CDMA based networks (Liu et al., 2009). At the transmitter, different users' data are spread by different spread spectrum sequences. At the receiver, different receiver structures are designed to distinguish each user's data. A set of correlators are used to distinguish each user's data (He et al., 2011). Aiming at the multichannel detection problem in CDMA communication, the symbol decision feedback (SDF) receiver and the chip hypothesis feedback (CHF) receiver were proposed by Stojanovic and Freitag (2007), and the results showed that these two receivers can well support a CDMA communication. Different spread spectrum schemes have been studied to support multiusers (Kumar and Kumar, 2015). The results showed that, compared with the Walsh-Hadamard (WH) code scheme, the carrier interferometry (CI) code scheme has a lower bit error rate (BER) and a better performance in supporting multiusers.

Doppler estimation and compensation is an important part of high-performance underwater acoustic communication. It includes mainly two technical methods. One is the data-aided Doppler estimation method, which involves an additional part to the data frame, such as a chirp signal, a single frequency signal, and a pseudo random sequence. Based on a certain tolerance of Doppler shift, the chirp signal was used for Doppler frequency shift estimation by transmitting 'up' and 'down' chirp signals to calculate the time shift (Diamant et al., 2012); the results showed that a quadratic-frequency modulation (QFM) chirp signal is more resilient to Doppler shift than a linear frequency-modulation (LFM) chirp signal. The traditional Doppler factor estimation method involves a cross-correlation between the received aid data and the local copy of the aid data (Sharif et al., 2000), which may sometimes fail to detect the correlation peak or have a peak fracture. To solve this problem, He et al. (2010) proposed a delay cross-correlation Doppler factor estimation method that uses the received aid data to perform a delay cross-correlation with the received aid data. The data-aided methods may reduce the effective data rate. The other method is the none-data-aided blind Doppler estimation.

Cotae and Yang (2010) proposed a cyclostationary blind Doppler estimation method that uses the *m*-sequence and the properties of the spectrum correlation function (SCF) to conduct Doppler estimation. Simulation results showed that it could estimate the Doppler frequency shift well. Several blind Doppler estimation methods were compared by Cotae et al. (2014). The results showed that the double modulation technique, combining the LFM and *m*-sequence for message signal modulation, could better minimize the phase shift. The none-data-aided methods have high implementation complexity.

Channel estimation and equalization, which is divided mainly into time-domain equalization and frequency-domain equalization, is another important part of the high-performance underwater acoustic communication. Tao et al. (2008) proposed a time-domain receiver scheme for multiple-input multiple-output (MIMO) underwater acoustic communication, using time-domain least-square (LS) and linear minimum mean-square error (LMMSE) to conduct a channel estimation and minimum mean square error (MMSE) criterion to conduct channel equalization. Zheng et al. (2007) proposed a single-carrier frequency-domain equalization (FDE) scheme, employing a small training signal block for initial channel estimation in the frequency domain and a linear MMSE criterion to conduct channel equalization. Zheng et al. (2015) studied time-domain turbo equalization and frequency-domain turbo equalization. Their results showed that the time-domain scheme has a better performance with higher complexity. Chen et al. (2017) studied frequency-domain turbo equalization based on a single-carrier MIMO underwater acoustic communication system. Simulation results showed that, even using 16-quadrature amplitude modulation (QAM) modulation, the system could still correctly demodulate the received signal. He et al. (2012) proposed a hybrid time-frequency-domain equalization method based on the pseudo-noise (PN) sequence for single-carrier communication. After channel estimation, the time-domain signal is converted to the frequency domain to conduct frequency-domain equalization, and then the frequency-domain signal is converted to the time-domain to conduct adaptive time-domain decision feedback equalization. Simulation and experimental results showed that the proposed scheme has good performance in eliminating multipath interference.

In summary, a large number of studies on underwater acoustic DS/SS systems have been carried out. Different signal processing algorithms have been applied in these systems. For DS/SS systems, the larger the spread spectrum gain, the better the system BER performance. However, the larger the spread spectrum gain, the lower the system communication rate. The larger the length of the spreading code used for correlation operations, the better the performance of Doppler estimation, channel estimation, and frame synchronization. Therefore, in terms of spread spectrum code length, the system communication rate is in conflict with the performance of Doppler estimation, channel estimation, and frame synchronization. This problem has not been adequately considered in previous studies. In this study, aiming at high stability and reliability to support underwater point-to-point communications and CDMA based underwater networks, we propose a direct sequence spread spectrum based single-carrier underwater acoustic communication system using dual spread spectrum code (SC-UWA/DSSC). To solve the contradiction between the information data rate and the accuracy of Doppler estimation, channel estimation, and frame synchronization, a data frame structure based on dual spread spectrum code is designed. A long spread spectrum code is used as the training sequence that can be used for data frame detection and synchronization, Doppler estimation and compensation, and channel estimation and equalization. A short spread spectrum code is used to modulate the effective information data, for Doppler estimation and compensation, a delay cross-correlation method based on the training sequence is used to estimate the Doppler factor, and then resampling is used to compensate for the Doppler frequency shift. For channel estimation and equalization, based on the training sequence, time-domain channel impulse response estimation is first conducted, and then the MMSE algorithm is used for channel equalization. For underwater CDMA based networking, each user is assigned a different pair of spread spectrum codes.

2 SC-UWA/DSSC system structure design

The structure of the proposed SC-UWA/DSSC system includes a transmitter and a receiver (Fig. 1). In the transmitter, each user's data are framed after

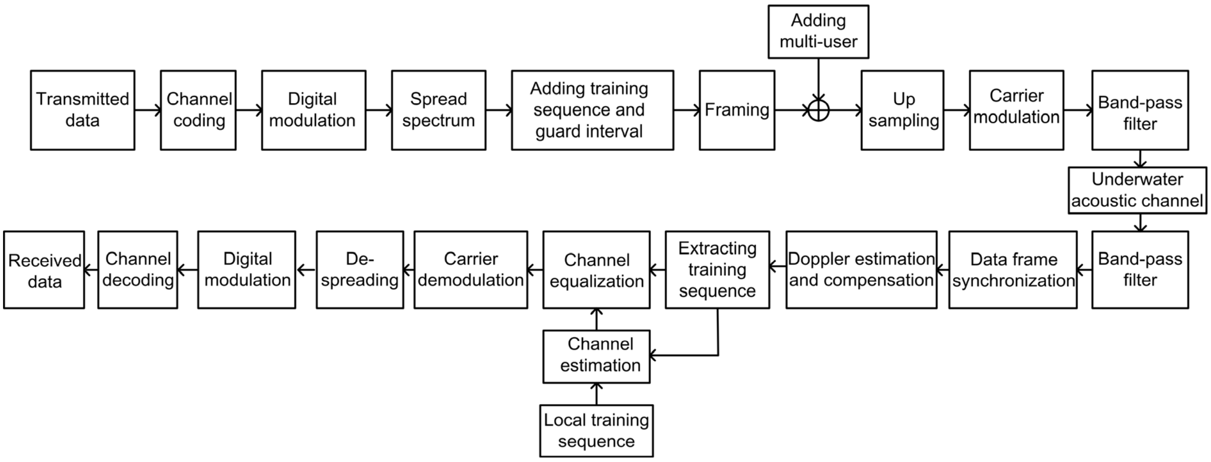

Fig. 1 SC-UWA/DSSC system structure

channel coding, digital modulation, spreading of the spectrum, and adding the training sequence and guard interval. After the multiuser signals are superimposed on the baseband, the mixed-signal is modulated by the carrier after being up sampled. Then the frequency band signal is sent to the underwater channel. The receiver continuously detects the received signal. The received signal is filtered by a band-pass filter. Data frame synchronization is then performed, and the Doppler factor is estimated. The Doppler shift is compensated for by resampling. Subsequently, the received training sequence is extracted to conduct channel estimation with the local training sequence. The channel estimation information is used to perform channel equalization. Finally, the transmitted data are recovered through carrier demodulation, de-spreading, digital demodulation, and channel decoding.

A typical application scenario of the SC-UWA/DSSC is a point-to-point communication in a shallow sea environment or a multiuser downlink communication of an underwater acoustic communication network in a deep sea environment (Fig. 2). As shown in Fig. 1, the multiuser signal is superimposed after baseband framing, which is a single-transmitting and multi-receiving communication scheme designed for the downlink channel from the base station to the sub-node in underwater acoustic networks. The data of different sub-nodes are superimposed by the different spread spectrum codes at the base station, and the sub-nodes recover each dataset according to its respective spread spectrum code (Fig. 2a).

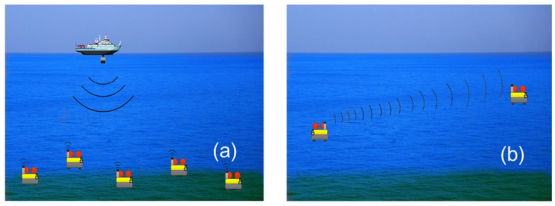

Fig. 2 An application scenario of the SC-UWA/DSSC system: (a) underwater acoustic CDMA based networks; (b) point-to-point communications

3 Dual spread spectrum code based data frame structure design

Data frame of the SC-UWA/DSSC system is composed of three identical training sequence segments and two effective data segments (Fig. 3). A zero sequence is used as the guard interval before and after each effective data segment. The training sequence uses a long gold code, and the effective data segment is spread spectrum modulated with a short gold code. For multiuser applications, each user has a different pair of spread spectrum codes including a training sequence and a spread spectrum code sequence. Different users' data are superimposed at the baseband after the data framing operation. Because of the multipath effect, the time for the signal to reach the receiver along each path is different, leading to transmitted signal superposition at the receiver and causing serious ISI. The guard interval is used to eliminate the ISI effect. Its duration is longer than the maximum multipath time delay. The guard interval can also be used as auxiliary data of the training sequence for channel estimation. The training sequences have four functions. The first training sequence is used for data frame detection and synchronization, and the first and the last training sequences are used for Doppler estimation and compensation. The training sequence before each effective dataset is used for channel estimation and equalization and for distinguishing different users in the underwater acoustic CDMA based networks.

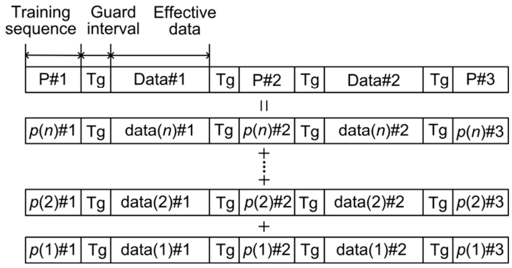

Fig. 3 Data frame structure of the SC-UWA/DSSC system

$p(n)$ represents the training sequence of the n^{th} user, and data(n) represents the effective data sequence of the n^{th} user which is spread by a unique spread spectrum code sequence

At the transmitter, assume that the user number is I, the spread spectrum code of the i^{th} user is $c_i(t)$, the data bit duration is T_{a}, and the chip duration of the spread spectrum code is T_{c}. Then the system spread spectrum factor can be expressed as

$$M = T_{\text{a}} / T_{\text{c}}. \tag{1}$$

Assume that the unit gate function is $g(t)$ and that $c_{i,m}$ is a binary sequence composed of +1 or −1 elements. The spread spectrum code of the i^{th} user can be expressed as

$$c_i(t) = \sum_{m=1}^{M} c_{i,m} g(t - mT_{\text{c}}). \tag{2}$$

Assume that the transmitted information data length is N and that the information data amplitude of the i^{th} user is $a_{i,n}$. The information data sequence of the i^{th} user $x_i(t)$ can be expressed as

$$x_i(t) = \sum_{n=1}^{N} a_{i,n} g(t - nT_{\text{a}}). \tag{3}$$

The transmitted information data sequence of the i^{th} user after spread spectrum can be expressed as

$$d_i(t) = x_i(t) c_i(t) = \sum_{n=1}^{N}\sum_{m=1}^{M} a_{i,n} c_{i,m} g(t - nmT_{\text{c}}). \tag{4}$$

Assume that the training sequence of the i^{th} user is $p_i(t)$, the length of $p_i(t)$ is L_{g}, and $p_{i,j}$ is a binary sequence composed of +1 or −1 elements. The data framing signal of the i^{th} user $f_i(t)$ can be expressed as

$$\begin{aligned} f_i(t) = d_i(t) + p_i(t) = &\sum_{n=1}^{N}\sum_{m=1}^{M} a_{i,n} c_{i,m} g(t - nmT_{\text{c}}) \\ &+ \sum_{j=1}^{L_g} p_{i,j} g(t - jT_{\text{c}}). \end{aligned} \tag{5}$$

For multiuser communication, the multiuser signal is superimposed on the baseband, and the baseband signal can be expressed as

$$\begin{aligned} F(t) = \sum_{i=1}^{I} f_i(t) = \sum_{i=1}^{I} \Bigg(&\sum_{n=1}^{N}\sum_{m=1}^{M} a_{i,n} c_{i,m} g(t - nmT_{\text{c}}) \\ &+ \sum_{j=1}^{L_{\text{g}}} p_{i,j} g(t - jT_{\text{c}}) \Bigg). \end{aligned} \tag{6}$$

The transmitted signal $s(t)$ can be expressed as

$$\begin{aligned} s(t) &= \text{Re}\left\{ F(t) \text{e}^{\text{j}2\pi f_{\text{c}} t} \right\} \\ &= \text{Re}\Bigg\{ \text{e}^{\text{j}2\pi f_{\text{c}} t} \sum_{i=1}^{I} \Bigg(\sum_{n=1}^{N}\sum_{m=1}^{M} a_{i,n} c_{i,m} g(t - nmT_{\text{c}}) \\ &\quad + \sum_{j=1}^{L_{\text{g}}} p_{i,j} g(t - jT_{\text{c}}) \Bigg) \Bigg\}. \end{aligned} \tag{7}$$

Unlike the radio channel, there is no standardized characterization of the underwater acoustic channels. In general, the frequency response of the p^{th} propagation path between the transmitter and receiver can be expressed as

$$H_p(f) = \frac{\varGamma_p}{\sqrt{A(l_p, f)}}, \tag{8}$$

where l_p is the length of the p^{th} propagation path, $\varGamma p$ the cumulative reflection coefficient along the p^{th} propagation path, and $\sqrt{A(l_p, f)}$ the propagation loss associated with this path. Each path of the acoustic channel can be considered as a low-pass filter, and the overall impulse response of the channel can be expressed as (Stojanovic and Freitag, 2009)

$$h(t)=\sum_{p} h_p(t-\tau_p), \tag{9}$$

where $h_p(t)$ is the inverse Fourier transform of $H_p(f)$ and τ_p is the path delay. Assume that the maximum multipath number is L_0, the Doppler frequency shift is invariable in one data frame duration, and the Doppler factor is Δ. The signal is disturbed by Gauss white noise $n(t)$ with average value 0 and variance σ^2. The received signal $r(t)$ can be expressed as

$$r(t)=\sum_{p=1}^{L_0} h_p s(t)[(1+\Delta)t-\tau_p]+n(t). \tag{10}$$

4 Doppler and channel estimation

4.1 Doppler estimation

The underwater sound velocity is about 1500 m/s. The Doppler frequency shift which may be caused by relative motion among platforms or seawater movements cannot be ignored. The Doppler frequency shift would cause a compression or extension of the data frame length. For the SC-UWA/DSSC system, the delay cross-correlation algorithm is used for Doppler factor estimation (He et al., 2010), and then the Doppler frequency shift is compensated for by resampling.

For Doppler factor estimation, assuming that T_{s} is the sampling period and ignoring the influence of noise and multipath, Eq. (10) can be rewritten as

$$r(nT_{\mathrm{s}})=s[(1+\Delta)nT_{\mathrm{s}}]. \tag{11}$$

The transmitted signal can be expressed as

$$s(nT_{\mathrm{s}})=r\left[\left(\frac{1}{1+\Delta}\right)nT_{\mathrm{s}}\right]. \tag{12}$$

Assume that the transmitted signal length is D and that the received signal length is D'. The Doppler factor Δ can be expressed as

$$\Delta=D/D'-1. \tag{13}$$

Assume that the initial sampling frequency of the received signal is f and that the resampling frequency is f'. To compensate for the Doppler frequency shift, the resampling frequency f' can be expressed as

$$f'=(1+\Delta)f. \tag{14}$$

Eq. (13) shows that the Doppler factor Δ can be calculated if D and D' are known. The transmitted signal length D is pre-defined. The traditional method for estimating D' is to perform cross-correlation between the local sequence and the received sequence. This method may cause correlation peak splitting or failure in detecting the correlation peak. To obtain the received frame length D', a data segment from the head of the received sequence is first obtained; the data segment may be 1–1.5 times the length of the modulated training sequence. Next, this data segment is used to perform the delay cross-correlation operation with the received sequence. There are three correlation peaks in the cross-correlation results. After determining the first and last correlation peak positions, the length between the two positions is the effective length D' of the received frame. Next, Eq. (13) is used to calculate the estimated Doppler factor Δ, and the calculated Δ is used to obtain the resampling frequency according to Eq. (14).

4.2 Channel estimation

Although the underwater acoustic spread spectrum communication has an anti-multipath interference ability, it cannot completely eliminate the multipath effect. Thus, it is necessary to conduct channel estimation and equalization. The first task is to estimate the channel impulse response (CIR). The cross-correlation channel estimation algorithm (Jiao et al., 2007) is used to estimate the CIR, and the linear MMSE equalization is used for channel equalization.

For channel estimation, the received training sequence should be extracted. First, perform sliding cross-correlation between the local copy of the training sequence and the received signals to obtain three correlation peaks. Second, extract the training sequence and effective data sequence according to each of the three correlation peaks. Third, use the training sequence and the local known training sequence to estimate the CIR. Finally, the channel estimation operation is conducted in the time domain.

Assuming that R_{pr} is the cross-correlation value between the received training sequence and the local training sequence and that R_{pp} is the auto-correlation value, R_{pr} can be expressed as

$$\begin{aligned} R_{\text{pr}} &= \sum_{j=1}^{L_g} p(j)\left[\sum_{l=1}^{L_0} h_l p(j+e-l) + n(j+e)\right] \\ &= \sum_{l=1}^{L_0} h_l \sum_{j=1}^{L_g} p(j)p(j+e-l) + \sum_{l=1}^{L_g} p(l)n(j+e) \\ &= \sum_{l=1}^{L_0} h_l R_{\text{pp}}(l) + R_{\text{pn}}. \end{aligned} \tag{15}$$

Through the multipath channel, the received training sequence becomes a superimposed signal of the transmitted signal through different delays and fading. R_{pp} reaches the maximum value when the received signal coincides with the local training sequence. From Eq. (15), it can be seen that the correlation value will exhibit L_0 peaks at locations with L_0 multipath and a smaller value at other locations. These smaller values can be considered as the noise effect. Thus, a threshold value should be set to eliminate these smaller values to achieve an accurate channel estimation, which can be expressed as

$$R_{\text{pr}} = \begin{cases} R_{\text{pr}}, & |R_{\text{pr}}| \geq \text{Threshold}, \\ 0, & |R_{\text{pr}}| < \text{Threshold}. \end{cases} \tag{16}$$

Using the correlation results, the reconstructed channel estimation can be calculated by Eq. (15). To facilitate the judgment, the above results can be normalized.

Fig. 4 shows the channel estimation and equalization process. After re-synchronizing the received signal using Doppler estimation and compensation, channel estimation is conducted according to the above channel estimation principle.

Reconstruct the channel estimation value $\hat{h}$ according to the channel estimation information. Then transform the time-domain $\hat{h}$ to the frequency-domain $\hat{H}$ by fast Fourier transformation (FFT), and calculate the channel equalization coefficients by $W=\hat{H}^*/(|\hat{H}|^2+\sigma^2)$ according to the MMSE criterion (σ^2 denotes the noise influence). Channel equalization is conducted in the frequency domain. First, the serial time-domain signal r_n is converted to the parallel time-domain signal, and then r_n is converted to the frequency-domain R_n by FFT. Next, signal R_n is multiplied by the equalization coefficient W to obtain the compensated received signal Z_n, and the equalization coefficients are updated by the channel estimation values obtained from the training sequence. Then the frequency-domain equalization signal Z_n is converted to the time-domain signal z_n. Finally, the parallel time-domain signal is converted into the serial signal.

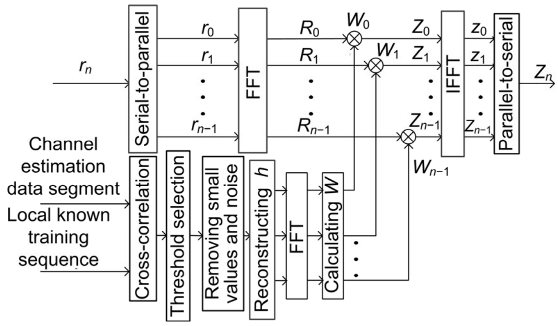

Fig. 4 Channel estimation and equalization process

5 Channel model

Fig. 5 shows the SC-UWA/DSSC simulation system based on Matlab, where the transmitter and receiver are the signal processing parts of the SC-UWA/DSSC system. The interpolation processing simulates the digital-to-analog (D/A) conversion and analog-to-digital (A/D) conversion modules of a real underwater acoustic communication system to ensure the independence of the sending and receiving processes. The part in the imaginary line block represents the transmission channel with Doppler, multipath, and additive noise. The Doppler factor denotes the Doppler shift between the transmitter and receiver. The propagation delay and signal fading factors denote the multipath transmission between the transmitter and the receiver.

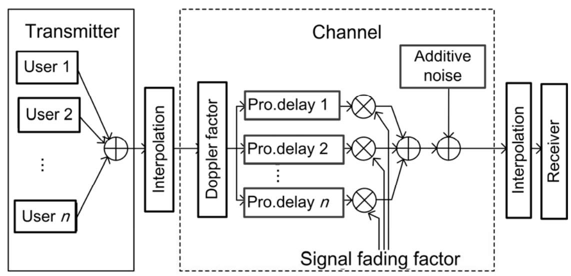

Fig. 5 SC-UWA/DSSC simulation system based on Matlab

Fig. 6 shows the channel models used in the simulation according to Eq. (9). Fig. 6a shows the channel model 1. It contains seven paths. The first path is the signal direct arrival path, and the maximum multipath delay is about 2.7 ms. Fig. 6b shows the channel model 2. It contains 12 paths. The first path is the signal direct arrival path, and the maximum multipath delay is about 27 ms.

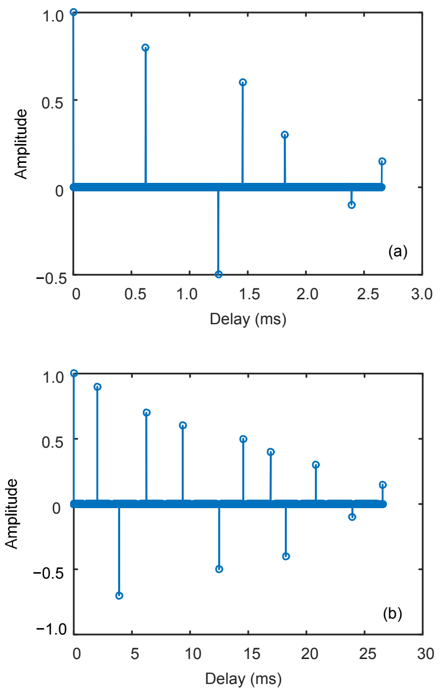

Fig. 6 Channel models: (a) channel model 1; (b) channel model 2

6 Simulation results and analysis

Table 1 shows the simulation parameters of the SC-UWA/DSSC system.

6.1 Doppler estimation and compensation simulation results

Fig. 7 shows the Doppler estimation and compensation results at different velocities. The simulation condition includes channel model 2, a random relative velocity from −10–10 m/s for different data frames, and invariable velocity in one data frame. The simulation system does not involve channel coding or channel estimation and equalization. The system BER is about 0.5 without Doppler estimation and compensation. The system BER decreases obviously with Doppler estimation and compensation. A similar trend is seen between the fixed velocity and the random velocity curves. The results demonstrate that the proposed system can effectively estimate and compensate for the Doppler frequency shift.

Table 1 Simulation parameters of SC-UWA/DSSC

Parameter	Value	Parameter	Value
Sample frequency (kHz)	384	Mapping mode	BPSK
Bandwidth range (kHz)	18–30	Channel coding	LDPC
Carrier frequency (kHz)	24	Simulation time	5000
Baseband bandwidth (kHz)	6	Training sequence length	1023
Spread code length	31	Data length (bit)	1280
Relative velocity (m/s)	5	Tg length (bit)	1024
FFT length	333 824	Multiples of up sampling	16

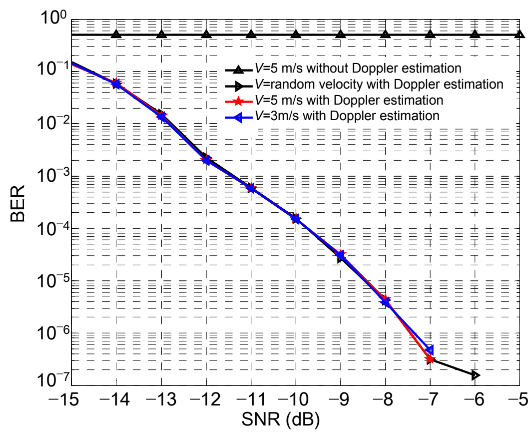

Fig. 7 Simulation results of Doppler estimation and compensation at different velocities

6.2 Channel estimation and equalization simulation results

Fig. 8 shows the channel estimation and equalization results. The simulation does not involve the Doppler effect or channel coding. For the same channel, the system BER with channel estimation and equalization can be 1–3 orders of magnitude lower than that without channel estimation and equalization. Compared with channel 1, channel 2 is worse, and the system BER increases by 1–3 orders of magnitude for the same SNR. The proposed system could recover the transmitted signal well at both channel 1 and channel 2 with channel estimation and equalization.

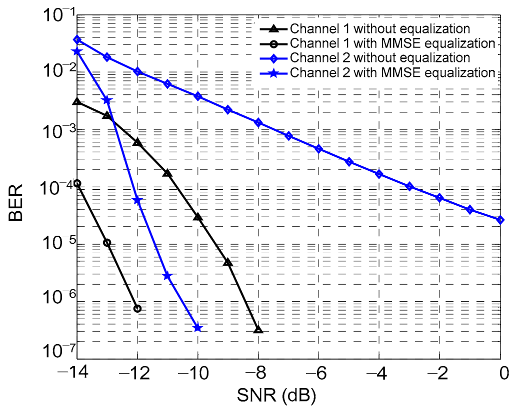

Fig. 8 Simulation results of channel estimation and equalization for different channels

Fig. 9 shows the received signal constellation graphs of the simulation results at channel 2. The received signal is the de-spreading signal. Figs. 9a and 9c are constellation diagrams with channel estimation and equalization when the SNR is −6 dB and −12 dB, respectively, and Figs. 9b and 9d are constellation diagrams without channel estimation and equalization. For the same SNR, the phase without channel estimation and equalization is more ambiguous than the phase with channel estimation and equalization.

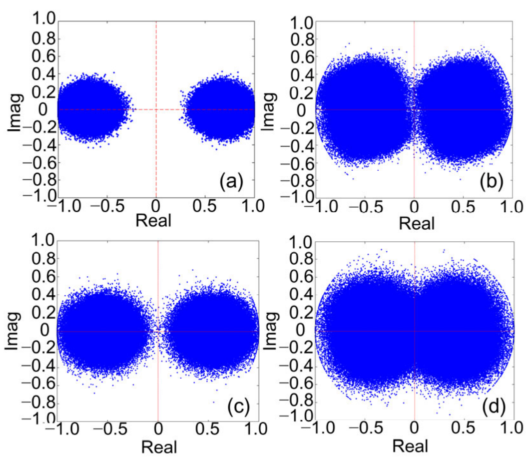

Fig. 9 Received signal constellation graphs: (a) SNR=−6 dB, MMSE equalization; (b) SNR=−6 dB, without equalization; (c) SNR=−12 dB, MMSE equalization; (d) SNR=−12 dB, without equalization

6.3 Point-to-point system performance simulation results

Fig. 10 shows the whole system's simulation results of point-to-point communication. The simulation condition is that the relative velocity is 5 m/s, and channel model 2 is used. It can be seen that the system BER can decrease by 1–3 orders of magnitude at the same SNR with channel equalization, and by 1–2 orders of magnitude at the same SNR with LDPC channel coding. The system BER can be less than 10^{-6} when the SNR is greater than −10 dB in an environment with multipath interference and Doppler frequency shift.

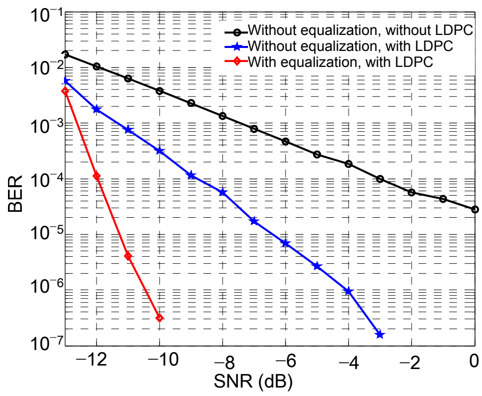

Fig. 10 Point-to-point system performance simulation results when the relative velocity is 5 m/s

6.4 Multiuser system performance simulation results

Fig. 11 shows the multiuser system simulation results. Channel model 1 is used. The BER increases as the user number increases at the same SNR. When the user number is less than three, the BER can be less than 10^{-6} when the SNR is greater than −1 dB. With four users, the system can still recover the transmitted signal at a higher SNR. With five users, the multiuser signal interference is larger and the system performance decreases by three orders of magnitude. The simulation results show that the system has some ability to support multiuser communication.

7 Comparison with related systems

Table 2 shows a performance comparison of the proposed SC-UWA/DSSC system with other DS/SS based SC-UWA communication systems. The comparison is based on the simulation results. We can obtain the following observations: (1) The proposed SC-UWA/DSSC has a lower system BER with a lower SNR and more multipath interference. This is the reason that the training sequence with a long spread spectrum code has the effective anti-multipath, anti-interference, and anti-Doppler abilities, and the LDPC channel decoding technique has a good BER correction performance. (2) SC-UWA/DSSC has a relatively high communication rate compared with the other systems, being as high as 355 bits/s. This demonstrates the effectiveness of the proposed dual spread spectrum code scheme.

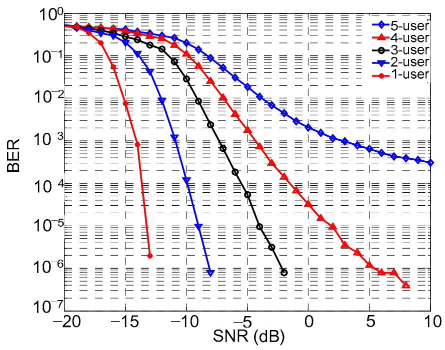

Fig. 11 Multiuser system performance simulation results

8 Conclusions

A DS/SS based single-carrier underwater acoustic communication system using dual spread spectrum code has been proposed. Moreover, a data frame structure based on dual spread code has been designed. To address the Doppler frequency shift and the multipath interference problem, a Doppler estimation and compensation method and a channel estimation and equalization method, both of which are based on a long spread spectrum code, have been introduced. Point-to-point communication simulation results showed that this system can estimate the Doppler factor and compensate for the Doppler effect accurately, and that the channel estimation and equalization algorithm based on training sequences can demodulate the correct signals, even with a strong multipath interference. The system BER can be smaller than 10^{-6} when the SNR is larger than −10 dB, even if there are as many as 12 paths with multipath interference. The multiuser communication simulation results showed that the proposed scheme can be used in the downlink channel of an underwater CDMA based networking system. Future study will involve conducting experiments to test the performance of the proposed system.

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Table 2 Comparison between the SC-UWA/DSSC system and other systems

Source	Spread code	Code length	Coding	Digital modulation	Band-width (kHz)	Carrier frequency (kHz)	Doppler	Multipath number	BER/SNR (dB)	Data rate (bit/s)
SC-UWA/DSSC	Gold	31	LDPC	BPSK	6	24	Yes	12	10^{-6}/−10	355
Sozer et al. (1999)	PN	25	–	BPSK	–	12	Yes	3	–	100
Palmese et al. (2007)	Gold	63	–	Chirp	3.5	–	Yes	Several	–	27
Huang and Sang (2007)	m	255/511/1023	RS	–	2.25	9	Yes	Several	–	8–35
Liu et al. (2009)	Gold	31	RA	QPSK	–	44	No	4	(10^{-5}–10^{-7})/−5	473
Yin et al. (2014)	m	63	–	CSK/BPSK/MFSK	4	6	Yes	5	10^{-2}/−5	317
Tang et al. (2016)	N-H	7/15	–	–	4	–	–	<10	10^{-4}/−5 10^{-4}/−7.5	428.8/ 272.7
Shu et al. (2017)	CMP	127	–	QPSK	–	–	Yes	2	10^{-3}/−5	100

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