Complementary Code Keying Modulation and Frequency Domain Equalization for Single Carrier Underwater Acoustic Communications

Xialin Jiang, Wei Su*, and En Cheng

***Abstract*—**In this paper, we investigate the single carrier modulation and detection schemes for medium range underwater acoustic communication with severe inter-symbol interference (ISI) and ocean ambient noise. Complementary code keying (CCK), a variation of M-ary bi-orthogonal keying, is used as the modulation scheme. The CCK modulation with good auto-correlation and cross-correlation properties provides a strong tolerance to ISI caused by multipath distortion and supports high data rate transmission. In addition, we propose the low complexity method of frequency domain equalization applied to single carrier system. Two types of channel estimation algorithms, least squares (LS) and matching pursuit (MP), are also implemented. We present performance analysis on bit error rate (BER) for different detection schemes. The ocean experiment results show that a reliable communication over 10 kilometers underwater transmission is achieved with a data rate at 4 kbit/s in 4 kHz of bandwidth.

***Keywords*—**Channel estimation, complementary code keying (CCK), frequency domain equalization, underwater acoustic communication.

I. Introduction

OVER the past few decades, there have been a number of activities in the area of underwater acoustic communications. An underwater acoustic channel is characterized as a multipath propagation, frequency-selective fading, limited available bandwidth, and ambient ocean acoustic noise channel [1]. These characteristics result in that high data rates reliable transmission in the underwater acoustic channel is very challenging [2]. It is generally recognized as one of the most difficult digital communication channel. In underwater acoustic communication, the long delay spread, which cause inter-symbol interference (ISI), may spoil tens or even hundreds of symbols. Thus, some sort of anti-ISI technique is needed to combat the channel distortion. To improve performance of data detection and obtain the high data rates communications over the band-limited underwater acoustic channels, orthogonal frequency-division multiplex (OFDM) has been applied as effective anti-multipath technique to underwater acoustic communications [3][4]. With the very special structure, the OFDM system can efficiently combat with ISI and distortion caused by multipath propagation. Additionally, another key feature that has made OFDM system popular in communication systems is that it allows a very low complexity frequency domain equalization (FDE) method via Discrete Fourier Transform (DFT) algorithm. However, high peak-to-average-power ration (PAPR) and sensitive to carrier frequency offset (CFO), which causes inter carrier interference (ICI), are major challenges with OFDM system.

An alternative to the OFDM approach, more traditional single carrier (SC) system with advanced modulation and detection techniques has comparable performance and low complexity. SC system uses a single carrier, instead of multi-carriers in OFDM system. Therefore, the peek-to-average power is much lower for single carrier system compared with OFDM system.

In the realm of single carrier system, research work has been active on modulation and detection techniques. In this paper, we design a single-carrier underwater acoustic communication system using complementary code keying (CCK) modulation [5][6][7] and frequency domain equalization techniques [8][9][10]. CCK modulation technique, which is adopted by IEEE 802.11b standard, with good auto-correlation and cross-correlation properties, provides a strong tolerance to multipath distortion and supports high data rates transmission. Transmissions over long memory underwater acoustic channel, time domain equalization technique has the disadvantages of high computational complexity. Compared to time domain equalization, single carrier frequency domain equalizers (SC-FDEs) can effectively reduce the computational complexity by converting the received signals from time domain to frequency domain. Similar to OFDM, the frequency domain equalization in SC system is conducted block-by-block via DFT and inverse DFT (IDFT) operations. Hence, SC-FDE system achieves the similar complexity and performance as OFDM system.

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The rest of the paper is organized as follows. In section II, we describe the system model of an underwater acoustic transmission system. In section III, we present the SC-FDE system employing the CCK modulation technique. We present experimental results and discussions in section IV. Finally, we draw conclusions in section V.

II. System Model

In this section, we present the system model for underwater acoustic communications. At the receiver end, the passband receive signals are generally down-converted to the equivalent baseband signals. The baseband models are usually used for further signal processing algorithms, because they are more mathematically tractable than passband signals.

In underwater acoustic communications systems, the transmitted signals typically propagate different paths from transmitter to receiver. This arises from sound reflection at the water surface, bottom and other objects, and sound refraction in the water. It is commonly known as multipath propagation. At the receiver, the multipath propagation results in filtering effect on the transmitted signals. In this paper, we assume down-conversion, carrier synchronization, and timing recovery are operated flawlessly, such that the underwater acoustic channel can be approximated by a discrete time equivalent baseband model, as shown in Fig. 1, where the underwater acoustic channel is modelled as a time-invariant FIR (finite impulse response) filter. While the underwater acoustic channel may actually be time-varying, if the variations are much slower than the data rate, the underwater acoustic channel can be viewed as time-invariant for each DFT block time scale.

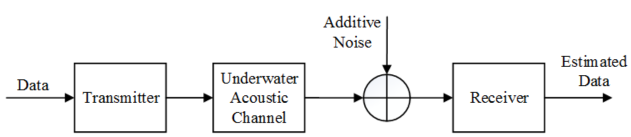

Fig. 1 representation of an underwater acoustic transmission system

The transmit-receive equivalent baseband signals relation in the time domain can be expressed as:

$$y = h * x + w \ , \tag{1}$$

Where $*$ denotes the convolution operation, y is the time-domain channel received signal, h is the channel impulse response, x is the time-domain transmitted signal, and w is the additive noise. In an SC-FDE system, the time-domain received signal is converted to frequency domain via DFT in the receiver end.

III. SC-FDE System Employing CCK Modulation

In this section, we propose the CCK modulation technique that works in conjunction with frequency domain equalization for a single carrier underwater acoustic communication system.

A. SC-FDE System Structure

The proposed single carrier system structure with CCK modulation and frequency domain equalization is depicted in Fig. 2.

The stream of input data is partitioned into 8-bit block through 1:8 serial-to-parallel (S/P) conversion. Each 8-bit block is mapped into eight Quadrature Phase Shift Keying (QPSK) chip symbols by employing CCK modulation. After the eight QPSK symbols undergo 8:1 parallel-to-serial (P/S) conversion, the serial symbols are partitioned again into the block of N symbols, where N is the size of DFT. Each block is cyclically extended by copying the last Ncp symbols to the beginning of the block. After N:1 P/S conversion, the two identity training sequences as pilot signal, which are known to the receiver, are inserted in the front of data blocks. The length of training sequence has to be set at least equal to the length of channel impulse response, and it may be equal to or less than the length of a data block N, and it is often set to the power of 2. In this paper, two golden sequences are used as pilots. The golden sequence has attractive periodic autocorrelation property. The first golden sequence acts as cyclic prefix (CP) that absorbs inter-symbol interference, and the second golden sequence is used for channel estimation. After through the pulse-shaping filtering, the baseband signals are formed; then the baseband signals are modulated to the desired frequency by multiplying the carrier signals. A chirp signal is inserted in front of the packet as a preamble, which is used for time synchronization. The packet structure is depicted in Fig. 3. The digital-to-analog (D/A) conversion and transducers produces the acoustic signals, which are transmitted over underwater acoustic channels.

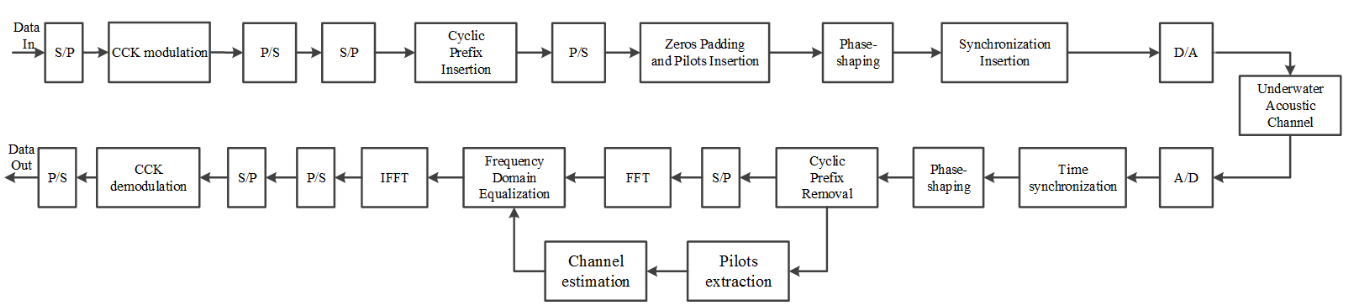

Fig. 2 the block diagram of an SC-FDE system employing CCK modulation

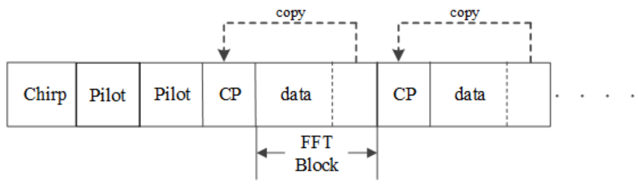

Fig. 3 packet structure of SC-FDE.

In the receiver, the signals at the output of the underwater acoustic channel undergo transducer/hydrophone conversion, analog-to-digital (A/D) conversion, and frequency down conversion. The time synchronization finds the start of the transmitted signal, and produces a sequence of noisy symbols whose length is equal to the transmitted data block. The CP of each data block is discarded. The rest of the block is sent to the DFT which converts the time domain symbols into frequency domain. The Frequency domain equalization combats with inter-symbol interference. The output results of frequency domain equalization are converted back to the time domain. Finally, symbols-to-bits and data decisions are made after CCK demodulation.

B. CCK Modulation

Complementary Code Keying is developed by Intersil and Lucent Technologies and is adopted as the modulation scheme of the IEEE 802.11 Standard for high-rate spread-spectrum communications (5.5 and 11 Mbps) at 2.4 GHz band. In this paper, the CCK modulation scheme we use is based on IEEE 802.11 11Mbps mode.

CCK is an M-ary orthogonal keying modulation method, where one of the M unique (almost orthogonal) signal code words are chosen carefully for transmission. The CCK has a code length of eight. The eight-chip CCK code words are defined as follows:

$$C = e^{j\varphi_1}\left\{e^{j(\varphi_2+\varphi_3+\varphi_4)}, e^{j(\varphi_3+\varphi_4)}, e^{j(\varphi_2+\varphi_4)}, -e^{j\varphi_4}, e^{j(\varphi_2+\varphi_3)}, e^{j\varphi_3}, -e^{j\varphi_2}, 1\right\} \quad (2)$$

where C is the code word. As shown in (2), the CCK code word is determined by the four phase parameters φ_1, φ_2, φ_3, φ_4 and, where φ_1 is contained in all eight chips, φ_2 is contained in all the odd chips, φ_3 is contained in all odd pairs of chips, and φ_4 is contained in all odd quads of chips. To begin, the information bits are partitioned into 8-bit block as $(d_0, d_1, \ldots, d_7)$. The eight bits encode the four phase parameters according to the scheme shown in Table I.

Table I. phase parameters encoding

dibits	phase parameter
d_0d_1	φ_1
d_2d_3	φ_2
d_4d_5	φ_3
d_6d_7	φ_4

The first two bits d_0 and d_1 encode the phase φ_1 based on differential quadrature phase shift keying (DQPSK) modulation as specified in Table II, i.e., the phase φ_1 is relative to the phase φ_1 in the previous symbol.

Table II DQPSK encoding

d_0d_1	φ_1 of Even symbols	φ_1 of Odd symbols
00	0	π
01	$\pi/2$	$-\pi/2$
11	π	0
10	$-\pi/2$	$\pi/2$

The dibits (d_2 d_3), (d_4 d_5), and (d_6 d_7) encode the phase φ_2, φ_3, and φ_4 based on QPSK modulation, respectively, as shown in Table III.

Table III QPSK encoding.

dibit (d_id_{i+1})	phase
00	0
01	$\pi/2$
10	π
11	$-\pi/2$

In CCK modulation, it takes 8 information bits to define each code word of an 8 complex chips. The 8 chips are QPSK symbols. Hence, there have 4^8=65536 possible code words for 8 information bits. Actually, the CCK modulation choses 2^8=256 code words from 65536 possible code words pool. With the large set of good codes to pick, the CCK modulation provides a strong tolerance to multipath distortion due to the large coding distance.

C. Frequency Domain Equalization

The frequency domain equalization in single carrier system makes use of the property of DFT to design a block data structure. To perform the frequency domain equalization, the stream of transmit data symbols is partitioned into the block size of *N* symbols. In generally, a guard interval of a length equal or greater to the duration of the channel impulse response is inserted at the beginning of each block. Moreover, the guard interval is often the repetition of last *Ncp* symbols of each block. It is so called cyclic prefix (CP). It acts as a guard interval that eliminates the inter-symbol interference (ISI) from the previous symbols, but also plays a key role in the task of frequency domain equalization. Because of the choice of the CP, the linear convolution between transmitted signal and the channel impulse response can be modelled as circular convolution. Consequently, frequency domain equalization can be performed by taking the DFT of each block of received signal, applying equalizer to remove the ISI effects in the frequency domain, and converting the results back to the time domain via IDFT.

We consider the case that a block of signal $\mathbf{x} = [x_1, x_2, ..., x_N]^T$ is passed through the underwater acoustic channel with the length of M impulse response $\mathbf{h} = [h_1, h_2, ..., h_M]^T$, where T represents the matrix transpose. It is assumed that the size of block N is larger than the length of channel impulse response M. The noise signal is $\mathbf{w} = [w_1, w_2, ..., w_N]^T$. The received signal $\mathbf{y} = [y_1, y_2, ..., y_N]^T$ is obtained by convoluting $\mathbf{x}$ and the vector of the channel impulse response, $\mathbf{h}$. It is written in a matrix form as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w}, \tag{3}$$

where $\mathbf{H}$ is the time-domain channel matrix, it is a N-by-N circular square matrix whose first column is given by the channel impulse response $\mathbf{h}$ with a number of zeros padding to the end to extend its length to the block size of N.

$$\mathbf{H} = \begin{pmatrix} h_1 & 0 & \cdots & 0 & h_M & h_{M-1} & h_{M-2} & \cdots & h_2 \\ h_2 & h_1 & \ddots & 0 & 0 & h_M & h_{M-1} & \cdots & h_3 \\ h_3 & h_2 & h_1 & \ddots & \ddots & 0 & h_M & \cdots & h_{M-1} \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & 0 & \cdots & h_M & h_{M-1} & h_{M-2} & h_{M-3} & \cdots & h_1 \end{pmatrix} \tag{4}$$

According to the feature of circular matrices, every circular matrix $\mathbf{H}$ can be diagonalized by the DFT matrix. The N-by-N DFT matrix defined as

$$\mathbf{F} = \begin{pmatrix} 1 & 1 & 1 & \cdots & 1 \\ 1 & e^{-j\frac{2\pi}{N}} & e^{-j\frac{4\pi}{N}} & \cdots & e^{-j\frac{2\pi(N-1)}{N}} \\ 1 & e^{-j\frac{4\pi}{N}} & e^{-j\frac{8\pi}{N}} & \cdots & e^{-j\frac{4\pi(N-1)}{N}} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 1 & e^{-j\frac{2\pi(N-1)}{N}} & e^{-j\frac{4\pi(N-1)}{N}} & \cdots & e^{-j\frac{2\pi(N-1)^2}{N}} \end{pmatrix} \tag{5}$$

Applying the DFT on either side of (4), we get the

$$\mathbf{\Lambda} = \mathbf{F}\mathbf{H}\mathbf{F}^{-1}, \tag{6}$$

where $\mathbf{F}^{-1}$ is the inverse discrete Fourier transform, and $\mathbf{\Lambda}$ is the diagonal matrix whose eigenvalues are also given by taking the fast Fourier transform of the first column of $\mathbf{H}$.

We define the frequency transformed vectors as

$$\mathbf{Y} = \mathbf{F}\mathbf{y}, \ \ \mathbf{X} = \mathbf{F}\mathbf{x}, \ \ \mathbf{W} = \mathbf{F}\mathbf{w}, \tag{7}$$

The frequency domain transmit-receive relation is derived from the time domain transmit receive relation by applying the discrete Fourier transform on both side of the (3):

$$\mathbf{F}\mathbf{y} = \mathbf{F}\mathbf{H}\mathbf{F}^{-1}\mathbf{F}\mathbf{x} + \mathbf{F}\mathbf{w}, \tag{8}$$

$$\mathbf{Y} = \mathbf{\Lambda}\mathbf{X} + \mathbf{W}, \tag{9}$$

For a time invariant frequency selective channel with severe ISI, the channel matrix in the frequency domain $\mathbf{\Lambda}$ is diagonal. The entries of the diagonal matrix $\mathbf{\Lambda}$ can be denoted the channel attenuations in the frequency domain. The trivial equalization is done by pointwise division of the receive signal by the corresponding channel attenuations in the frequency domain. This method of equalization is known as zero-forcing (ZF) equalization [11][12]. Without any further statistical information about transmitted signal and channel, the zero-forcing equalization is the solution for the least squares problem. The frequency domain of zero-forcing equalization can be expressed as:

$$\mathbf{x} = \mathbf{F}^{-1}\mathbf{X} = \mathbf{F}^{-1}\mathbf{\Lambda}^{-1}\mathbf{Y}, \tag{10}$$

The zero-forcing equalization ignores the presence of additive noise. Its use may result in significant noise enhancement at those frequencies where channel frequency response has spectral nulls. To avoid such noise enhancement, an alternative is to adopt the minimum mean-squared error (MMSE) equalizer which minimizes the error between the transmitted symbols and the output of equalizer. The MMSE equalizer requires the second order statistical information about the noise signal and the transmitted signal, and provides a tradeoff between noise enhancement and ISI removal. The frequency domain of MMSE equalization can be expressed as

$$\mathbf{x} = \mathbf{F}^{-1}\mathbf{X} = \mathbf{F}^{-1}(\mathbf{\Lambda}\mathbf{\Lambda}^* + \frac{1}{SNR}\mathbf{I})^{-1}\mathbf{\Lambda}\mathbf{Y} \tag{11}$$

where I is an $N \times N$ identity matrix, and SNR is the signal-to-noise ratio at the input of MMSE equalizer.

Zero-forcing equalizer and MMSE equalizer are both linear equalizers and simple to implement, because the inversion of a diagonal matrix is easy to calculate. In addition, the DFT and IDFT operations can be implemented using the computationally efficient fast Fourier transform (FFT) and inverse FFT (IFFT) algorithms. However, there performance is not enough for underwater acoustic channels having spectral nulls. An obvious choice for underwater acoustic channels with sever ISI is a non-linear equalizer.

A decision feedback equalizer (DFE) is a nonlinear equalizer which employs previous decisions to estimate the current symbol with a symbol-by-symbol detector. The DFE consists of two filters. The first one is called a feedforward filter. The second one is called a feedback filter. Its input is the decision of the previous symbols from the detector. There are different variations of DFEs. In this paper, we focus on the hybrid time-frequency domain DFE approach, which uses frequency domain filtering only for the feedforward filter, and uses conventional time domain filtering for the feedback filter. The structure of a hybrid DFE is depicted in the Fig. 4.

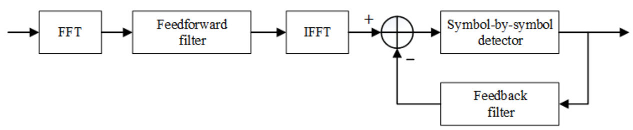

Fig. 4 the block diagram of a hybrid DFE

A block of N received ISI affected signal is transformed to the frequency domain via FFT operations, then passed to the feedforward filter. An IFFT is applied to the output of feedforward filter. Any trailing ISI is cancelled by the time domain feedback filter using previously decided symbols. The output of feedback filter is subtracted from the output of IFFT to form the estimated symbol, then the symbol-by-symbol detector maps the estimated symbol to the corresponding symbol in the signal constellation. In order to facilitate the design of the feedforward filter and the feedback filter coefficients, it is customary to assume that the detector gives correct decisions, and the error between estimated symbols and decisions is used to calculate the feedforward filter and the feedback filter coefficients. The feedforward filter and feedback filter coefficients can be optimized according to zero-forcing or MMSE criterion. The above mentioned zero-forcing equalizer or MMSE equalizer can be used for the feedforward filter and feedback filter, i.e., MMSE-DFE implies that the DFE coefficients are derived from MMSE criterion.

The DFE works better than linear equalizers assuming no error propagation, but when a symbol is incorrectly decided, the wrong decision will feedback and propagate during some symbol periods. Thus, the length of feedback filter has to be selected carefully.

D. Channel and Noise Variance Estimation

In order to design the equalizer coefficients as above, the channel impulse response and noise variance are needed to estimate. Therefore, two types of channel estimation algorithms, least squares (LS) and matching pursuit (MP), are implemented. As above mentioned, we use two golden sequences as pilots which are known to the receiver to perform channel estimation. The first golden sequence plays a role of the CP. The second golden sequence are used for channel and noise estimation. The LS channel estimation has a very low computational complexity. In addition to LS channel estimation, the MP channel estimation is also implemented. The MP algorithm is a sparse channel estimation algorithm. It is suitable for sparse underwater acoustic channels.

To obtain the estimation of the channel noise variance, we use the following approach. In this paper, the channel noise is modeled as additive Gaussian white with variance σ^2. With the transmitted pilot signal matrix $\mathbf{S}$, the estimated channel impulse response $\hat{\mathbf{h}}$ and the length N_p of received pilot signal $\mathbf{y}_p$, the estimation of noise variance can be obtained from

$$\hat{\sigma}^2 = \frac{1}{N_p}\left\|\mathbf{y}_p - \mathbf{S}\hat{\mathbf{h}}\right\|^2 \tag{12}$$

IV. Experiment Results and Discussions

In this section, we present the results and discussions of the proposed CCK modulation and SC-FDE solution for the shallow underwater acoustic channels. The experimental data is collected off the Coast of Xiamen City, Fujian, China. As shown in Fig. 5, the distance between transmitter and receiver is 10 kilometers. In the transmitter, one transducer is deployed 9 meters beneath the ocean surface. The depth of water is about 16 meters in the transmitter area. In the receiver, we arrange five transducers/hydrophones to receive the transmitted acoustic signals. Three transducers/hydrophones are deployed at the head of marine ship, equally spaced 3 meters on a vertical linear array, and the deepest one is 9 meters underwater. They are numbered as R1, R2, and R3 from the up to down.

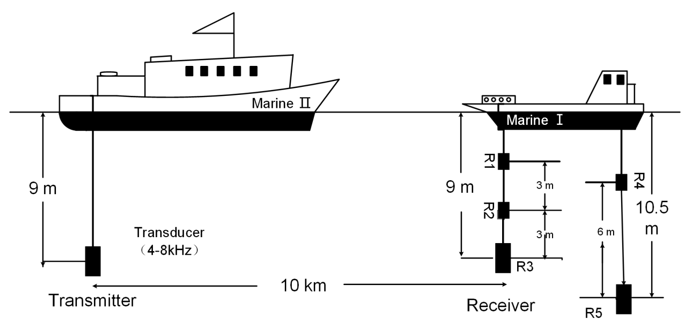

Fig. 5 the deployment of equipment

The other two transducers/hydrophones are deployed at the tail of marine ship, and also vertically arranged with 4.5 meters underwater and 10.5 meters underwater, respectively. They are numbered as R4 and R5. The depth of water is about 20 meters in the receiver area. The data is collected through NI DAQ (NI USB-6259) which is connected to a computer for off-line processing.

The experimental parameters are set as follows. The duration of every symbol is $Ts = 0.5$ ms. The sample rate is set to 60 kHz. The center frequency of carrier is 6 kHz. The frequency band range of transmitted signals are from 4 kHz to 8 kHz. The chirp signal lasts 2 seconds to make sure the accuracy of time synchronization. The training sequence is 512-bit golden sequence (appending a zero in the end) and is mapped to 256 QPSK symbols. 8192 information bits are mapped to QPSK symbols following CCK modulation technique. The data block size is $N = 1024$, and the length of CP is 256. The total number of data block sent is 8.

During the experiments, the received signals are recorded in files for offline processing. In Fig. 6, the magnitude of received signals are plotted. The power spectrum of the recorded signals is plotted in Fig. 7. It is clear that the received signal frequency components centers at the carrier frequency 6 kHz and spans from 4 kHz to 8 kHz due to the pulse shaping filter.

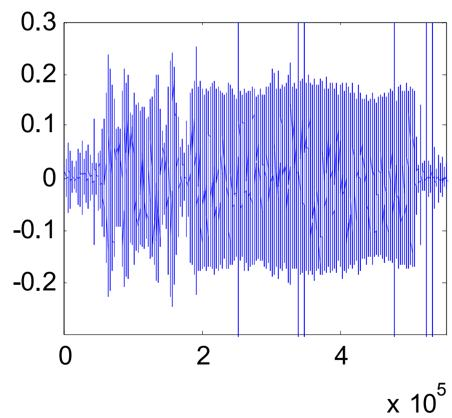

Fig.6 the received signals.

To recover the transmitted signals, the synchronization is necessary. The synchronization is obtained by correlating the received signals with a local chirp signal and the peak of correlation is the start of transmitted sequence.

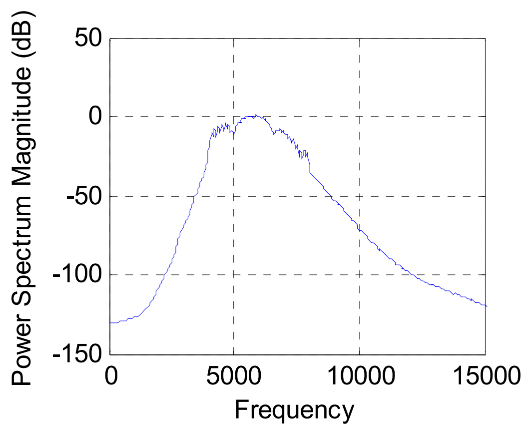

Fig. 7 the power spectrum magnitude of received signals.

For the channel estimation, we adopt the conventional LS algorithm and MP algorithm. To make comparison, we list the bit error rates (BER) for LS and MP channel estimation in table IV. The hybrid MMSE-DFE is used for both two channel estimation algorithms.

Table IV BER results of channel estimations. Hybrid MMSE-DFE is used.

#Hydrophones	LS algorithm	MP algorithmta
R1	47%	3.87%
R2	0.78%	0.09%
R3	0.37%	0.05%
R4	34%	1.53%
R5	0.61%	0

The first column is the tag of five hydrophones/transducers. The LS channel estimation generates a slightly higher BER than the MP channel estimation with the data collected from the hydrophones R2, R3, and R5. With the low SNR conditions, such as data collected from the hydrophones R1 and R4, the data cannot recover correctly with LS channel estimation.

With the MP channel estimation and noise variance estimation, the three frequency domain equalization methods, ZF equalizer, MMSE equalizer and hybrid MMSE-DFE equalizer can be performed. To demonstrate the performance of equalization, we report BER results for the three frequency domain equalization methods in Table V. The MP channel estimation algorithm is used.

Table V BER results of equalization. MP channel estimation is used.

#Hydro-phones	ZF Equalizer	MMSE Equalizer	Hybrid MMSE-DFE
R1	45.96%	45.87%	3.87%
R2	5.04%	5.04%	0.09%
R3	4.76%	4.7%	0.05%
R4	38.92%	38.9%	1.53%
R5	3.2%	3.2%	0

With good channel condition, such as data collected from hydrophone R2, R3 and R4, the SC-FDE system with hybrid MMSE-DFE equalizer can achieve nearly zero error transmission. That means reliable underwater acoustic communications are obtained with data rate at 4 kbits/s in 4 kHz of bandwidth. The performance of hybrid MMSE-DFE equalizer is the best. The linear equalizers, Zero-force equalizer and MMSE equalizer, cannot eliminate channel distortion efficiently when channel has spectral nulls, such as the data collected from R1 and R4. To demonstrate the performance of equalization, the estimated QPSK symbols at the output of the three equalizers are demonstrated in Fig. 8. The data is collected from R2.

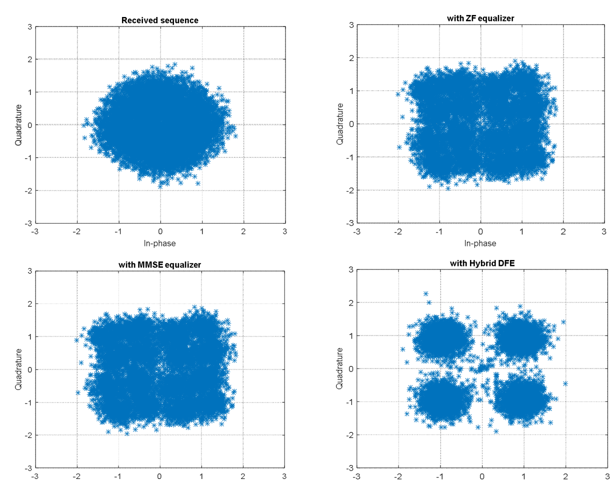

Fig. 8 demonstration of equalization with QPSK symbols

V. Conclusion

In this paper, a novel CCK modulation and frequency domain equalization scheme has been presented for single carrier underwater acoustic communication system. CCK modulation supports high data rate transmission and has a good auto-correlation property, which provides a strong tolerance to channel distortion. The performance of SC-FDE system is similar to that of OFDM, even with a long memory channel.

With the aid of cyclic prefix inserted between consecutive data blocks, a low complexity frequency domain equalization for single carrier system with CCK modulation and its performance is tested by processing the collected data in real-world underwater acoustic communications experiments. The practical ocean experimental results demonstrate that an SC-FDE system employing CCK modulation achieves a robust underwater acoustic communication. The bit error rate is below 10^{-4} with data rate at 4 kbits/s in 4 kHz of bandwidth.

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