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Joint Interference Mitigation with Channel Estimated in Underwater Acoustic System

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Abstract

Orthogonal frequency-division-multiplexing (OFDM) has been actively pursued in underwater acoustic communications. However, OFDM is extremely sensitive to carrier frequency offset (CFO) which cause significant problems. The utilization and performance of traditional channel estimation are poor. Although underwater acoustic communication was recently the focus of research, literatures on joint interference mitigation with channel estimation using compressed sensing (CS) in underwater acoustic OFDM has been very limited. In the Cyclic prefix orthogonal frequency division multiplexing (CP-OFDM) system, a novel method was proposed which combined interference mitigation with channel estimation based on compressed sensing (CS). We propose a two-step approach which contains resampling and high-resolution uniform residual Doppler compensation by null subcarriers. After Doppler compensation, exploiting sparsity in the delay-Doppler domain, CS-based channel estimation is adopted for an increase in spectral efficiency though a reduction of the number of pilot symbols. Simulations and tank experiments have verified the better performance of the receiver based on the joint interference and CS channel estimation. The proposed CFO compensation method can reduce the BER. The Joint interference estimation with CS-based channel estimation has better performance than that with LS channel estimation.

Keywords: *underwater acoustic communication, OFDM, CFO estimation, CS-based channel estimation*

1. Introduction

Multicarrier modulation in the form of orthogonal frequency division multiplexing (OFDM) is a promising alternative for underwater acoustic communication due to its ability in mitigating multipath effects. The OFDM system decreases the inter-symbol interference by increasing the symbol interval. It also can convert a frequency selective channel into a set of parallel frequency flat sub-channels, and simplify the receiver equalization, as proposed equalization methods in [1-2] are complicated.In order to maximize the elimination of the inter-symbol interference, the cyclic prfix ofdm (CP-OFDM) is used in the system.

However, OFDM is extremely sensitive to synchronization errors, especially the carrier frequency offset (CFO), which is induced by oscillator discrepancies between the transmitter and receiver and/or Doppler shifts. The CFO destroys the orthogonality among sub-carreirs, and degrades the bit error rate (BER) performance severely. In the underwater acoustic (UWA) channels, the impulse response of the communication channel is sparse. However, the conventional channel estimation method does not make full use of the priori knowledge that the transmission channel is is inherently sparse. So the accuracy and Validity of channel estimation is not high enough.

CFO estimation for OFDM is an active area of research. The extensive literature on CFO estimation for OFDM can be categorized as data-aided schemes (e.g., [3]) and non-data-aided (blind) schemes that only rely on the received OFDM symbols [4]. The CFO acquisition range is extended in [5] by using one training symbol with more than two identical parts. A blind method along with a maximum likelihood CFO estimator based on two consecutive and identical received blocks, where the estimable CFO is restricted to less than half of the

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subcarrier spacing. Existing CFO compensation typically entails transmission of a known preamble prior to the data, which can be easily detected by the receiver with the correlator. Existing preambles in underwater telemetry are almost exclusively based on Chirp signals. CFO estimators that exploit the cyclic prefix are summarized in [6], which also presents a minimum variance unbiased CFO estimator. The receiver deploys correlators to Compute time compression ratio and resample signals. This method is lacking in real-time underwater acoustic communication. The remaining CFO still threatens data transmission. In order to more accurately estimate the CFO, combined a bank of self-correlators and two linearly-frequency-modulated (LFM) signals [7] to compensate for the Doppler shift. Based on the compensated signal and channel' sparsity in the delay-Doppler domain, we estimate the channel paths using orthogonal matching pursuit (OMP) [8-9].

In this paper, we treat the channel as having a common Doppler scaling factor on all propagation paths [5], and apply a two-step approach which containing non-uniform Doppler compensation via resampling and high-resolution uniform compensation of the residual Doppler to mitigate the Doppler effect. After Doppler compensation, CS-based channel estimation is adopted by exploiting sparsity in the delay-Doppler domain [10]. The CS-based channel estimation allows for an increase in spectral efficiency through a reduction of the number of pilot symbols. We consider the joint effect of CFO and channel estimation for OFDM modulation. We study two methods based on different channel estimation methods, one is Least Squares (LS), and the other is CS-based estimation. Simulation results validated the effectiveness of the presented methods. It is shown that the joint interference estimation with CS-based channel estimation has better performance than that with LS channel estimation.

We have tested the proposed method with real data from an experiment at Key Laboratory of Underwater Acoustic Communication and Marine Information Technology, Ministry of Education, China, Dec. 28, 2012. Using two linearly-frequency modulated (LFM) waveforms as preamble, the proposed method divides subcarriers into groups, where each group of 1024 subcarriers consists of 672 data subcarriers, 256 pilot subcarriers,and 96 carefully spaced null subcarriers. The proposed practical receiver algorithms rely on the preamble of a packet consisting of multiple OFDM blocks to estimate the resampling factor, the null subcarriers to facilitate high-resolution residual Doppler compensation, and the pilot subcarriers for channel estimation.

2. Research Method and Realization

Consider a CP-OFDM system with the OFDM symbol duration T and the cyclic prefixed length T_{cp}. The subcarrier spacing is $K \times K$. Assume that there are a total of K subcarriers, located at:

$$f_n = f_c + n\Delta f, n = -K/2, ..., K/2-1 \tag{1}$$

Let $s[n]$ denote the information symbol to be transmitted on the nth subcarrier. The baseband CP-OFDM signal is:

$$s(t) = \begin{cases} \sum_{n=0}^{N-1} s[n]q(t)e^{j\frac{2\pi nt}{T}} & t \in [0, T+T_{CP}] \\ 0 & t \notin [0, T+Tcp] \end{cases} \tag{2}$$

Where $q(t)$ is a pulse shaping window, describing the Cyclic prefix-padding operation, $T^{'} = T + T_{CP}$.

$$q(t) = \begin{cases} 1 & t \in [0, T+T_{CP}] \\ 0 & elsewhere. \end{cases} \tag{3}$$

Sampling s(t) at the rate of T/N , and define $t=\frac{mT_S}{N}$. We obtain:

$$s_m = s\ (\mathrm{m}\frac{T_S}{N})=\sum_{\mathrm{n}=0}^{N-1} s[n]\ e^{j2\pi m\frac{n}{N}} (0 \le m \le N-1) \tag{4}$$

Here, the expression of (4) is same as the discrete inverse Fourier transform'expression, i.e. $s[n]$ complete discrete inverse Fourier transform (IDFT). At the receiver, s_m can be obtained through the discrete Fourier (DFT).

$$s[n]=\sum_{m=0}^{N-1} s_m e^{-j\frac{2\pi nm}{N}} \ (0 \le n \le N-1) \tag{5}$$

Performing the CP-OFDM modulation, the set that contains all the data sub-carriers indexes is denoted by Γ_D , the set that contains all the pilot sub-carriers indexes is denoted by Γ_P and the set that contains all the null subcarrier indexes is denoted by Γz , which satisfy $\Gamma_D \cup \Gamma_P \cup \Gamma_Z = [-N/2,...N/2-1]$, $\Gamma_A = \Gamma_D \cup \Gamma_P$. So the passband received signal is:

$$x(t)= \mathrm{Re}\{[\sum_{n\in\Gamma_A} S[n]e^{j2\pi n\Delta f t}]e^{j2\pi f_c t}\}, t \in [0,T^{'}] \tag{6}$$

When the passband signal $x(t)$ goes through the channel $h(t,\tau)=\sum_{p=1}^{N_{pa}} A_P \delta(\tau-[\tau_p - at])$, where $A_p(t)$ and $\tau_p(t)$ denote the amplitude and delay of the P_{th} path, we receive:

$$\begin{aligned}\tilde{r}(t) &= \mathrm{Re}\{\sum_p A_p[\sum_{n\in\Gamma_A} S[n]e^{j2\pi n\Delta f(\tau+at-\tau_p)}]e^{j2\pi f_c(t+at-\tau_p)}\} + i(n) + v(n) \\ &= \mathrm{Re}\{\sum_{n\in\Gamma_A}\{S[n]e^{j2\pi(1+a)(f_c+\frac{n}{T})t}\sum_p A_p e^{-j2\pi(f_c+n\Delta f)\tau_p}\}\} + i(n) + v(n)\end{aligned} \tag{7}$$

Seen from above, the Doppler shifts can cause strong interference. At the receiving end, we must eliminate their influence. Receiver use a two-step approach to mitigate the Doppler effect, and a CS-based method to estimate channel impulse noise. Two methods were used for CFO and channel estimation, one is separate CFO and channel estimation, and the other is the CS joint estimation.

2.1. Doppler scale estimation

The signal design assumes that the channel have a common Doppler scaling factor on all propagation paths, and apply a two-step approach which containing non-uniform Doppler compensation via resampling and high-resolution uniform compensation of the residual Doppler to mitigate the Doppler effect. Null subcarriers are used to facilitate Doppler estimation. The compensation steps are shown in Figure 1.

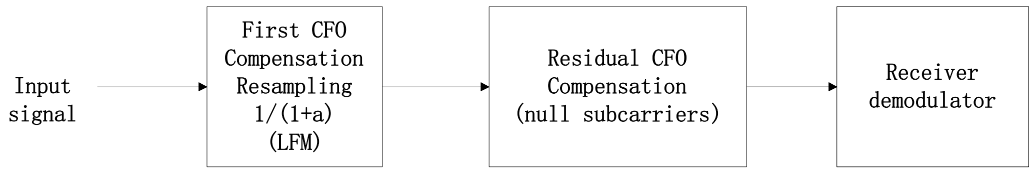

Figure 1. CFO Estimations and Compensation Algorithm Flow Chart

We use a bank of self-correlators proposed by Bayan S.Sharif [11] to estimation resampling factor $\tilde{a}$, with resampling, the baseband received signal is:

$$y'(t) = \tilde{r}\ (\frac{t}{1+\tilde{a}}) \tag{8}$$

$$\begin{aligned} y'(t) &= \mathrm{Re}\{\sum_{n\in\Gamma_A}\{S[n]e^{j2\pi\frac{1+a}{1+\tilde{a}}(f_c+\frac{n}{T})t}\sum_p A_p e^{-j2\pi(f_c+n\Delta f)\tau_p}\}\}+i(t)+v(t) \\ &= \mathrm{Re}\{\underbrace{e^{j2\pi\frac{1+a}{1+\tilde{a}}f_c t}\sum_{n\in\Gamma_A}\{S[n]e^{j2\pi n\Delta f\frac{1+a}{1+\tilde{a}}t}\sum_p A_p e^{-j2\pi f_n\tau_p}}_{A}\}\}+i(t)+v(t) \end{aligned} \tag{9}$$

With $y'(t) = \mathrm{Re}\{y(t)e^{j2\pi f_c t}\}$ the baseband signal $y(t)$ is:

$$\begin{aligned} y(t) &= Ae^{-j2\pi f_c t} \\ &= e^{j2\pi\frac{a-\tilde{a}}{1+\tilde{a}}f_c t}\sum_{n\in\Gamma_A}\{S[n]e^{j2\pi n\Delta f\frac{1+a}{1+\tilde{a}}t}[\sum_p A_p e^{-j2\pi f_n\tau_p}]\}+i(t)+v(t) \end{aligned} \tag{10}$$

We approximate $(1+a)/(1+\tilde{a}) = 1$ the baseband signal becomes:

$$y(t) \approx e^{j2\pi\frac{a-\tilde{a}}{1+\tilde{a}}f_c t}\sum_{n\in\Gamma_A}\{S[n]e^{j2\pi n\Delta f t}[\sum_p A_p e^{-j2\pi f_n\tau_p}]\}+i(t)+v(t) \tag{11}$$

$$\varepsilon = \frac{a-\tilde{a}}{1+\tilde{a}}f_c \tag{12}$$

Sampling $y(t)$ at a rate T/K, the discrete samples are:

$$y(n) \approx e^{j2\pi\varepsilon n}\sum_{k\in\Gamma_A}\{S[\mathrm{k}]e^{j2\pi kn/K}H[\mathrm{k}]\}+i(n)+v(n) \tag{13}$$

Then Define a selection matrix Θ that picks the frequency domain measurements on the null subcarriers out of all the K subcarriers. Define the following vectors $\mathrm{y} = [y[0],...,y[K-1]]^T$, $\mathrm{i} = [i[0],...,i[K-1]]^T$, $\mathrm{v} = [v[0],...,v[K-1]^T$ and a $K\times K$ diagonal matrix $D(\varepsilon) = diag(1, e^{j2\pi T_c\varepsilon},...,e^{j2\pi T_c\varepsilon(K-1)})$, where $T_c = T/N$.

$$\Lambda_H = \mathrm{diag}([H[0],...,H[K-1]]) \tag{14}$$

The matrix-vector representation of the channel input-output relationship is:

$$\mathrm{y{=}D}(\varepsilon)F^H\Lambda_H\mathrm{d{+}i{+}v} \tag{15}$$

Where F is the $K\times K$ Fourier transform matrix with the (p, q) entry $e^{-j2\pi pq/K}$. The energy of thenull sub-carriers is used as the cost function:

$$\gamma(\varepsilon) = \sum_{n\in\Gamma_Z}|\Theta F\ D^H(\varepsilon)y|^2 \tag{16}$$

Where $(.)^H$ denotes the complex conjugate transpose operation of a matrix. If the receiver compensates the data samples with the correct CFO, the null sub-carriers will not see the ICI spilled over from neighboring data sub-carriers. Hence, an estimate of ε can be found through H'.

Which can be solved via one-dimensional search for ε. This high-resolution algorithm corresponds to the MUSIC-like algorithm proposed in [12] for cyclic-prefixed OFDM. Hence, ICI-free reception is approximately achieved.

2.2. CS-based Channel Estimation

After CFO compensation, we presented CS based techniques for estimating doubly selective channels within CP-OFDM systems. Assuming that the channel has L taps in discrete time. We collect $2L$ samples after resampling for each OFDM block into a vector $y_p = [y_p(0), y_p(1), ..., y_p(2L-1)]$. The channel length can be inferred based on the synchronization output of the preamble, and its estimation does not need to be very accurate. We define a $2L \times 2L$ vector $X_P = diag\{d_p(0), d_p(1), ..., d_p(2L-1)\}$, and a $K \times K$ Fourier Transform matrix F. We construct the frequency domain received pilot signal as:

$$y_P = X_P S_p F h + S_p F \eta = X \Lambda h + F v \tag{17}$$

Where h is the channel impulsive response in the time domain, S_P is a selection matrix containing only one element equal to 1 per column, and η is the additive white gaussian noise. The value of selection matrix is specially formatted to correspond to pilot structure. Hence, h can be recovered using the following convex optimization program:

$$\min \| \tilde{h} \|_1 \quad subject\ to\ \ \| Y_P - X \Lambda h \|_2 \leq \varepsilon \tag{18}$$

Where $\| . \|_2$ denotes the 2-norm, and ε is chosen such that it bounds the amount of noise in the measurements. Hence, an estimate of channel impulsive response h can be found through:

$$\hat{h} = (\Lambda^H X^H X \Lambda)^{-1} \Lambda^H X^H Y_P \tag{19}$$

The orthogonal matching pursuit (OMP) can iteratively reconstruct the channel paths.

3. Results and Analysis

We simulate a channel consisting of 10 discrete paths.The amplitudes are Rayleigh distributed with average power decreasing exponentially with the delay. The CP-OFDM signal parameters are listed in Table 1, three configurations are introduced to compare the performance of the proposed receiver with CFO.
Configuration 1: No CFO compensation, different channel estimators are adopted;
Configuration 2: With CFO compensation, the LS and OMP channel estimator are adopted;
Configuration 3: With CFO compensation,different compensation methods are adopted;

Table 1. Simulation Parameters

Parameter	value
SNR	0:1:26
FFT length	1024
Doppler shift	5
Channel length	1024
Symbol period	204.8ms
Symbol snumbers/frame	1
sub-carriers numbers Signal	1024
modulation type	QPSK
the cyclic prefix length	256

The bit-error-rate(BER) performance will be used as the performance metric. When applying the proposed receiver in configurations 1, LS and OMP channel estimation are introduced to compare the performance of the receiver which does not perform CFO cancellation. The simulated results are shown in Figure 2. It demonstrates CFO has a great impact on CP-OFDM system.The signal without CFO compensation led to poor performance. But the OMP is better than LS.

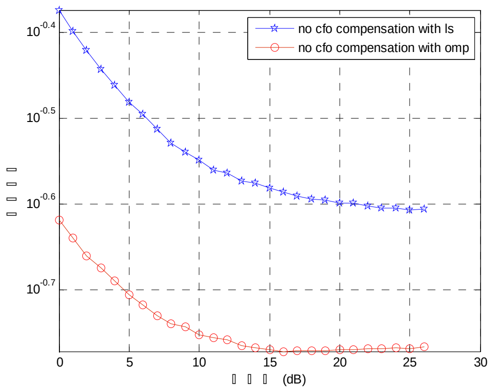

Figure 2. The Performance Comparison of Uncompensated CFO

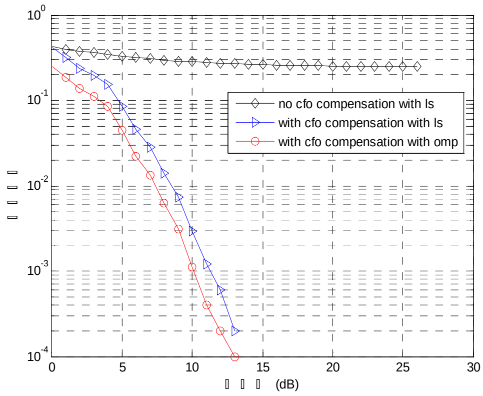

Figure 3. BER Performance for Different Channel Estimation Methods

The simulated results of configuration 2 are shown in Figure 3. It shows the BER performance with CFO compensation for CS and LS channel estimation methods, and the BER performance with LS channel estiamtion for CFO compensation or not. One can observe a performance gap between the receiver without CFO compensation and the proposed receiver which mitigates the CFO through estimation and subtraction.

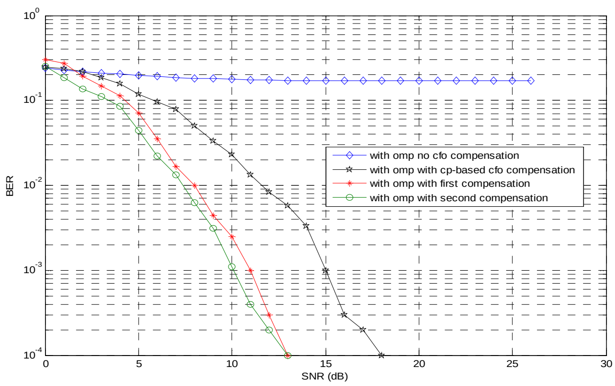

Figure 4. BER Performance for Different Compensation Schemes

Figure 4 compares the BER performance for different CFO compensation schemes based OMP channel estimation. It compares the first uncompensated CFO, initial compensation and secondary compensated error performance. As can be seen that increased levels of Doppler compensation, channel estimation performance improvement is more obvious. Comparing with the CFO estimation method that exploit the cyclic prefix, the proposed re-sampling and the second compensation method can get a better BER performance. We can also knows that the joint CFO estimation the CS channel estimation method can improve the error performance.

4. Pool Experiment

The experiment was carried out at the experimental pool in Xiamen University. System specifications for the experiment is shown in Table 2. The transmitter and the receiver were both located at the depth of 0.8m below the surface and the distance between them is 7m. In orde to test the performance of the joint estimation of CFO and CS channel estimation. The sent machine move at a speed of 0.25m/s, based Doppler formula $f_v - f \approx f(\frac{v}{c})$ adds a Doppler shift, the transmission frequency f is 30kHz, the acoustic wave propagation velocity c of 1500m/s, Pool experiment results as shown in Table 3.

Table 2. Pool Experimental Parameters Set

Parameter	value
FFT length	1024
Data frames	20
Sampling frequency	96kHz
Carrier frequency	30kHz
Symbol period	343.33ms
Guard interval	85.33ms
sub-carriers numbers	1024
Signal modulation type	QPSK
Effective bandwidth	6kHz
Pilot type	Comb-type

Table 3. Underwater Experiment Results

No.	No CFO compensation	With CFO with LS estimation	With CFO with OMP estimation
1	0.4993	0.0002	0.0001
2	0.4992	0.0006	0.0002
3	0.5007	0.0004	0.0003
4	0.5029	0.0007	0.0002
5	0.5043	0.0064	0.0058
6	0.5013	0.0056	0.0001
7	0.4994	0.0049	0.0020
AVERAGE	0.5010	0.0027	0.0012

After CFO mitigation with traditional LS and OMP channel estimation, the average BER of the two cases is 0.27% and 0.12%,respectively. We can find the novel joint optimization of CFO estimation and CS-based channel estimation achieve better performance.

To further illustrate the feasibility of such a system, we conducted semi-physical simulation using the received signal to estimate the real-time underwater acoustic channel signal-to-noise ratio 19dB, then White Gaussian noise is added to the received to deteriorate the channel conditions gradually. So we obtain 20 SNR(0-19dB) then counduct 20 set of experiments, we can get the SNR-BER simulation graphics shown in Figure 5.

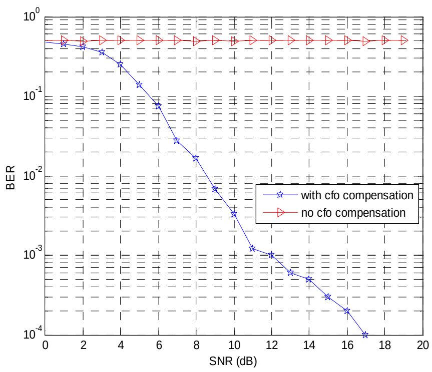

Figure 4. BER Performance for Different Signal to Noise Ratio

Seen from the figure, the system performance is poor with no CFO compensation. After CFO compensation, the system performance is greatly improved. With the increase of SNR, BER decreases; in underwater acoustic channel SNR range, using the two-step method to compensate for the CFO and co-OMP channel estimation, the system has good stability margin.

5. Conclusion

In this paper, we developed an interference compensation and mitigation receiver for CP-OFDM in underwater acoustic communications.Reciver uses the proposed two-step method to estimate the CFO. We simulate the BER performance of receiver before and after the CFO compensation. The simulation and experimental results show that the proposed CFO compensation method can reduce the BER. Receiver uses two algorithms LS and OMP to estimate the channel. Simulation and tank experiments verify the novel joint optimization of CFO estimation and CS-based channel estimation has better performance.

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