### **Proceedings of Meetings on Acoustics**

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Manuscript Number:	POMA-D-17-00125R1
Full Title:	Cyclic-feature based Doppler scale estimation for orthogonal frequency-division multiplexing (OFDM) signals over doubly selective underwater acoustic channels
Article Type:	ASA Meeting Paper
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Section/Category:	Signal Processing in Acoustics



**Proceedings of Meetings on Acoustics** 

Volume 30

http://acousticalsociety.org/



# **Acoustics** `17 Boston



**173rd Meeting of Acoustical Society of America and 8th Forum Acusticum** Boston, Massachusetts 25-29 June 2017

## Signal Processing in Acoustics: Paper 5aSPb11

# Cyclic-feature based Doppler scale estimation for orthogonal frequency-division multiplexing (OFDM) signals over doubly selective underwater acoustic channels

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Doppler scale estimation for underwater acoustic (UWA) communications is important yet challenging. Most Doppler scale estimators rely on training data or specially designed packet structures. These methods have fundamental limitations in transmission rate and spectral efficiency. Different from these methods, this paper presents a Doppler scale estimation approach exploiting the cyclostationarity of orthogonal frequency-division multiplexing (OFDM) signals. We analyze the cyclic features of cyclic prefix (CP) OFDM signals over doubly selective underwater channels in order to demonstrate the relationship between the cyclic features and the Doppler scale. Comparing with CP-OFDM signals, zero padding (ZP) OFDM signals have no redundant information contained within each block. Therefore, the proposed estimation method turns to be more significant for ZP-OFDM systems, which is usually used for underwater communication to save energy. Simulation results validate our theoretical analysis and the performance of the proposed Doppler scale estimator.

Published by the Acoustical Society of America



#### 1. INTRODUCTION

Underwater acoustic (UWA) communications enable underwater wireless networks to be applied in various applications, such as oceanographic research, pollution early-warning, disaster prevention, and military systems.<sup>1</sup> However, a major challenge in UWA communications is to combat doubly selective (time and frequency selective) channels.<sup>2</sup> Orthogonal frequency-division multiplexing (OFDM) transmission scheme is an attractive solution for reliable high data rate UWA communications. The OFDM transmission model has its own merits in efficient implementation (IFFT/FFT modulator/demodulator) and frequency domain equalization. However, the subcarrier orthogonality in OFDM is extremely sensitive to frequency offsets which may be introduced by Doppler distortion due to transmitter/receiver/medium motion. In underwater environment, the low sound propagation speed (1500 m/s) and relative high platform speed result in the Doppler scale several times larger than the one in radio transmission. In addition, underwater channels are wideband in nature because the signal bandwidth is not negligible compared to the carrier frequency.<sup>3</sup> For these reasons, in UWA OFDM communication systems, Doppler distortions on different subcarriers differ considerably. This frequency-dependent Doppler distortion causes strong intercarrier interference (ICI) if it is not corrected accurately.<sup>4</sup>

Methods for Doppler scale estimation can be divided into three categories.<sup>5</sup> The first category is to insert Doppler-insensitive waveforms which are known to the receiver.<sup>6</sup> Typical Doppler-insensitive waveforms include linear frequency-modulated (LFM) waveform and hyperbolic frequency-modulated waveform.

The second one is with training data assisted. In a typical study,<sup>4</sup> coarse estimation of the Doppler scale is obtained by cross correlating the received signal with the known preamble and postamble. Another work considers a preamble that consists of two identical OFDM symbols preceded by a cyclic prefix (CP), while the receiver uses a bank of parallel self-correlators.<sup>7</sup> The concept of the second order stationary statistics is employed for the Doppler scale estimation in UWA communications.<sup>8,9</sup> However, in these works, the receiver needs to transmit *m*-sequences, which is long in order to guarantee a good performance.

The third category exploits null subcarriers. In some studies, the null subcarriers are exploited to perform carrier frequency offset (CFO) estimation.<sup>4,7</sup> Similarly, the null subcarriers can also facilitate the Doppler scale estimation.<sup>5,10</sup> The total energy of null subcarriers is used as the cost function which has an extremum for an accurate estimation of the Doppler scale. In the recent work, the cost function is constructed through discrete Fourier transform (DFT) which has less computational complexity than the resampling method.<sup>11</sup> However, the null subcarriers based methods have a low spectral efficiency.

Different from the previous work, this paper investigates the Doppler scale estimation in OFDM communication systems for doubly selective underwater channels, without using the training data or relying on the special designed packet structure. We show the relationship between the Doppler scale and the cyclic features of the received signal. Based on the theoretical analysis, we find that the Doppler scale can be estimated from the extremums of the cyclic autocorrelation function (CAF) of the received signal. Note that the derivation in this work is based on the CP-OFDM,<sup>12</sup> however, the extension to ZP-OFDM<sup>12</sup> is straightforward. For the CP-OFDM, we propose the cyclic feature-based Doppler scale estimation method (CFDE) and the CP redundant information-based Doppler scale estimation method (CFDE). Simulation results show that the CFDE performs much better than the CPDE when using three or more OFDM blocks. For the ZP-OFDM, there is no CP information to exploit and the CPDE cannot be used. Fortunately, the cyclic feature of ZP-OFDM signals still exists such that the CFDE can also be used. This is a key advantage of the proposed method using cyclostationarity properties of received OFDM signals.

The rest of the paper is organized as follows. The system model is presented in Section 2. Section 3 presents the OFDM signal cyclostationarity in the doubly selective underwater channels and the proposed Doppler scale estimator. Simulation results are exhibited in Section 4, and we conclude the paper in Section 5.

#### 2. SYSTEM MODEL

Consider an OFDM system with N subcarriers, where N is an integer power of two. An efficient OFDM implementation is obtained by means of the inverse fast Fourier transform (IFFT). To avoid interblock interference, a CP is inserted in the beginning of each OFDM block. The transmitted signal in the passband is then given by

$$\tilde{s}(t) = \operatorname{Re}\left\{\frac{1}{\sqrt{N}}\sum_{m=-\infty}^{+\infty} \left[\sum_{k=0}^{N-1} X_{k,m} e^{j2\pi\Delta fk(t-T_g-mT)}g(t-mT)\right] e^{j2\pi f_c t}\right\},\tag{1}$$

where  $X_{k,m}$  denotes the kth data-modulated subcarrier in the mth OFDM symbol. The duration of an OFDM block  $T = T_N + T_g$ , where  $T_N$  denotes the OFDM symbol duration and  $T_g$  is the length of the CP. The frequency spacing is  $\Delta f = 1/T_N$ . g(t) is a raised cosine shaping filter with the duration of T.  $f_c$  is the carrier frequency.

The signal s(t) passes through a doubly selective underwater channel, which can be described as

$$h(t,\zeta) = \sum_{p} A_{p}(t)\delta(\zeta - \zeta_{p} + \gamma t), \qquad (2)$$

where  $A_p(t)$  and  $\zeta_p$  are the time-varying channel gain and path delay of the *p*-th path, respectively. The Doppler scale  $\gamma$  is to be estimated. We adopt three assumptions as follows:<sup>4</sup>

A1) All paths have the same Doppler scaling factor. Note that this is an approximation due to the fact that different multipaths could have different Doppler scaling factors;

A2) The path delays  $\zeta_p$ , channel gains  $A_p$ , and the Doppler scaling factor  $\gamma$  are constant over the block duration T;

A3) Time synchronization has been achieved.

The received signal in passband can be derived as

$$\tilde{r}(t) = \tilde{s}(t) \otimes h(t,\zeta) = \sum_{p} A_{p}\tilde{s}\left((1+\gamma)t - \zeta_{p}\right)$$

$$= \operatorname{Re}\left\{\sum_{p} A_{p}\left\{\frac{1}{\sqrt{N}}\sum_{m=-\infty}^{+\infty}\left[\sum_{k=0}^{N-1} X_{k,m}e^{j2\pi\Delta fk((1+\gamma)t - \zeta_{p} - T_{g} - mT)}\right] \times g\left((1+\gamma)t - \zeta_{p} - mT\right)\right]e^{j2\pi f_{c}\left((1+\gamma)t - \zeta_{p}\right)}\right\}\right\} + \tilde{n}(t), \qquad (3)$$

where  $\otimes$  is the convolution operator and  $\tilde{n}(t)$  is the complex additive white Gaussian noise (AWGN). The received signal in baseband r(t) is obtained by down-conversion, and the local carrier frequency is assumed to be  $f_c + \Delta f_c$ , where  $\Delta f_c$  is the CFO. Using the relationship between the baseband and passband signal, i.e.  $\tilde{r}(t) = \operatorname{Re} \{r(t)e^{j2\pi(f_c+\Delta f_c)t}\}, r(t)$  can be written as

$$r(t) = \frac{1}{\sqrt{N}} e^{-j2\pi\Delta f_c t} \sum_{m=-\infty}^{+\infty} \sum_{k=0}^{N-1} \sum_{p} X_{k,m} A_p e^{-j2\pi f_k \zeta_p} e^{j2\pi f_k \gamma t} e^{j2\pi\Delta f k (t - T_g - mT)} \times g((1+\gamma)t - \zeta_p - mT) + n(t),$$
(4)

where  $f_k = f_c + k\Delta f$ , and n(t) is the baseband version of  $\tilde{n}(t)$ . From (4), we observe that the k-th subcarrier experiences a frequency shift  $\gamma f_k$ . Resampling is an effective methodology to handle the time-scale change in UWA communications. However, resampling requires the resampling parameter  $\gamma$  to be estimated first. The focus of this paper is to determine the resampling parameter  $\gamma$ .

#### 3. CYCLOSTATIONARITY BASED DOPPLER ESTIMATION

It is well known that most of communication signals are cyclostationary. The CAF provides a comprehensive means of signal detection, parameter estimation and classification.<sup>13–15</sup> With the received signal given by (4), we shall next show that the signal r(t) is called second-order cyclostationarity,<sup>16</sup> i.e., its time varying autocorrelation function is a periodic function of time. For simplicity, the noise is ignored and the autocorrelation function of r(t) is given by

$$R_{rr}(t,\tau) = E\left[r(t)r^{*}(t+\tau)\right] = \frac{\sigma_{X}^{2}e^{j2\pi\Delta f_{c}\tau}}{N} \sum_{k=0}^{N-1} e^{-j2\pi f_{k}(1+\gamma)\tau} \sum_{p} \sigma_{p}^{2}R_{gg}(t,\tau,\zeta_{p}),$$
(5)

where  $R_{gg}(t,\tau,\zeta_p) = \sum_{n=-\infty}^{+\infty} g((1+\gamma)t - \zeta_p - nT)g((1+\gamma)(t+\tau) - \zeta_p - nT))$ , and we assume the mul-

tipath channel gain is uncorrelated and its second moment is  $E\left[|A_p|^2\right] = \sigma_p^2$ . In addition, we assume the data symbols are taken from a finite complex alphabet constellation. The symbols are uncorrelated, and their second moment is  $\sigma_X^2$ . We argue that the order of the constellation has no effects on the irrelevance between the symbols. Discussions about this are given in the Appendix A. It is easy to verify that  $\sum_p \sigma_p^2 R_{gg}(t,\tau,\zeta_p) = \sum_p \sigma_p^2 R_{gg}(t+\frac{T}{1+\gamma},\tau,\zeta_p)$ . Thus  $R_{rr}(t,\tau) = R_{rr}(t+\frac{T}{1+\gamma},\tau)$  which implies that the received OFDM signal went through a doubly selective underwater channel exhibits the second-order cyclostationarity with cycle frequencies (CFs)  $\{\omega_m|\omega_m = iT_r^{-1}, i \in Z\}$ , where  $T_r = \frac{T}{1+\gamma}$ . Based on this, the CAF of r(t) can be calculated as<sup>17</sup>

$$R_{rr}^{\omega}(\tau) = \lim_{T_r \to \infty} \frac{1}{T_r} \int_{-T_r/2}^{T_r/2} R_{rr}(t,\tau) e^{-j2\pi\omega t} dt$$
  
$$= \frac{\sigma_X^2 e^{-j2\pi\Delta f_c \tau}}{NT_r} \sum_{k=0}^{N-1} e^{-j2\pi f_k(1+\gamma)\tau} \sum_p \sigma_p^2 \int_{-\infty}^{+\infty} R_{gg}(t,\tau,\zeta_p) e^{-j2\pi\omega t} dt.$$
(6)

Under the assumption of no aliasing,<sup>17</sup> the corresponding CAF for the discrete time signal  $r(n) = r(t)|_{t=nT_s}$  is given as

$$Q_{rr}^{\alpha}(\eta) = R_{rr}^{\omega}(\tau)|_{\omega = \alpha T_s^{-1}, \tau = \eta T_s} = \frac{\sigma_X^2}{N T_r T_s^{-1}} e^{-j2\pi \Delta f_c \eta T_s - j2\pi f_c \eta \frac{1+\gamma}{\rho N} - j\pi \eta \frac{N-1}{\rho N}(1+\gamma)} \sum_p \sigma_p^2 G(\alpha, \eta, \xi_p) \Psi_N(\gamma, \eta),$$
(7)

where  $T_s = (\rho N \Delta f)^{-1}$  is the sampling period, with  $\rho$  as the oversampling factor. Also, we define  $G(\alpha, \eta, \xi_p) = \sum_n \bar{R}_{gg}(n, \eta, \xi_p) e^{-j2\pi\alpha n}$  as the Fourier series of  $\bar{R}_{gg}(n, \eta, \xi_p)$ , and  $\bar{R}_{gg}(n, \eta, \xi_p)$  is the discrete version of  $R_{gg}(t, \tau, \zeta_p)$ . The corresponding CFs satisfy  $\{\alpha_m\} = \{\alpha | Q_{rr}^{\alpha}(\eta) \neq 0, \alpha = \omega T_s, \eta = \tau T_s^{-1}\}$ . In addition,  $\Psi_N(\gamma, \eta) = \frac{\sin[\pi((1+\gamma)\eta)/\rho]}{\sin[\pi((1+\gamma)\eta)/\rho N]}$  comes from the following equation

$$\sum_{k=0}^{N-1} e^{\frac{-j2\pi k(1+\gamma)\eta}{\rho N}} = \frac{\sin[\pi(1+\gamma)\eta/\rho]}{\sin[\pi(1+\gamma)\tau/\rho N]} e^{\frac{-j\pi(N-1)(1+\gamma)\eta}{\rho N}}.$$
(8)

We are now ready to provide an estimator for the Doppler scale *a* based on (7). Recall that  $\bar{R}_{gg}(n,\eta,\xi_p)$  is  $\left|\frac{T}{(1+\gamma)T_s}\right|$ -periodic in *n*, which suggests that the magnitude of  $G(\alpha,\eta,\xi_p)$  is non-zero only for CFs. As a

result, the magnitude of  $Q_{rr}^{\alpha}(\eta)$  is non-zero only for CFs.  $\Delta f_c$  on the phase will not affect the the magnitude of  $Q_{rr}^{\alpha}(\eta)$ . In addition, for a given CF, an non-zero magnitude peak can be obtained at delays  $\pm \left\lfloor \frac{\rho N}{1+\gamma} \right\rfloor$ . This is due to the cyclic prefix of OFDM block and the factor  $\Psi_N(\gamma, \eta)$ . The proposed Doppler scaling factor estimator is now summarized as follows:

(1) Compute the CAF based on M samples of the received signal via

$$\hat{Q}_{rr}^{\alpha}(\eta) = \frac{1}{M} \sum_{n=0}^{M-1} r(n) r^*(n+\eta) e^{-j2\pi\alpha n}.$$
(9)

(2) For the CFDE method, we find the first cyclic frequency (i = 1) satisfies

$$\hat{\alpha}_m = \arg\max_{\substack{T_N(1+\gamma_{\min})\\T\rho N} \le \alpha \le \frac{T_N(1+\gamma_{\max})}{T\rho N}} \left\{ \left| \hat{Q}_{rr}^{\alpha}(0) \right| \right\},\tag{10}$$

then compute the estimated Doppler scaling factor via  $\hat{\gamma} = \frac{T\rho N\hat{\alpha}_m}{T_N} - 1$ .

(3) For the CPDE method, we first find the delay value  $\hat{\eta}_m$  satisfies

$$\hat{\eta}_m = \arg\max_{\substack{\rho N\\1+\gamma_{\max}} \le \eta \le \frac{\rho N}{1+\gamma_{\min}}} \left\{ \left| \hat{Q}^0_{rr}(\eta) \right| \right\},\tag{11}$$

then compute the estimated Doppler scaling factor via  $\hat{\gamma} = 1 - \frac{\rho N}{\hat{\eta}_m}$ . where  $\gamma_{\min}$  and  $\gamma_{\max}$  can be selected from empirical values. Since only the CAF information at zero cyclic frequency is used, this method is nothing else than a simple correlator that exploits the redundant information contained within CPs.

The resolution of the Doppler scale estimates obtained by both estimators is dependent upon the number of subcarriers (N) of the OFDM signal and the oversampling factor ( $\rho$ ), as  $\delta_a = 1/\rho N$ . A major consideration in the design of the estimator is the computational burden incurred as a result of the CAF computation when a high Doppler resolution is demanded. There exists an important trade off between the oversampling rate and the estimation performance which may be achieved.

#### 4. SIMULATION RESULTS

In this section, we provide simulation results demonstrating our analysis and the performance of the proposed estimator. The bandwidth of the OFDM signal is BW = 5 kHz, and the carrier frequency is  $f_c = 15$  kHz with a CFO  $\Delta f_c = 10$  Hz. In addition, 128-point FFT with oversampling factor  $\rho = 8$  and 64-QAM modulation are used for the simulated OFDM transmissions. The CP length is set to 1/4 of an OFDM block length. We consider a doubly selective channel with 5 channel tap gains and the exponentially decaying power-delay profile. The path delays at t = 0 are randomly distributed within [10, 25] samples, and the Doppler scaling factor is set to be  $\gamma = 0.001$ .

We first evaluate the CAFs of the CP-OFDM signal and the ZP-OFDM signal to illustrate the features of the cyclostationarity. Recall that the CP-OFDM signal exhibits the second-order cyclostationarity with CFs  $\left\{ i \frac{1+\gamma}{T}T_s \middle| i \in Z \right\}$ , and a local maximum of the magnitude of the CAF can be obtained at delay  $\frac{\rho N}{1+\gamma}$ . Different from the CP-OFDM signal, the ZP-OFDM signal only exhibits the cyclic feature, and no local maximum of the magnitude of the CAF over delays. These features are illustrated in Fig. 1 (a) and Fig. 1 (b), respectively. According to our parameter settings, and ignoring the quantization error, the CFs should be  $\{0.0007i | i \in Z\}$  and the local maximum of the magnitude of CAF ought to locate at  $\eta = 1023$ . As can be observed from Fig. 1, the simulation results are consistent with our analysis.



Figure 1: The estimated magnitude of the CAF of OFDM signals with respect to CF and delay over doubly selective channels for 20 dB SNR: (a) CP-OFDM; and (b) ZP-OFDM



Figure 2: The  $P_{ce}$  of the CPDE method and CFDE method versus SNR for different number of involved OFDM blocks.



Figure 3: The  $P_{ce}$  of the CFDE method versus SNR for different Doppler scaling factor under  $N_u = 4$ .

To assess the performance of the CPDE method and the CFDE method, we have conducted a Monte Carlo (MC) simulation with 200 trials. The metric named the probability of the correct estimation  $(P_{ce})$  is defined as the ratio of the number of the correct estimation to the total trials. A correct estimation means the difference between the estimate and the truth-value is within  $\pm \frac{1}{\rho N}$ . In Fig. 2, the  $P_{ce}$  of Doppler scaling factor versus signal-to-noise ratio (SNR) is plotted for different  $N_u$  which is defined as the number of involved OFDM blocks. As observed, for both  $N_u$  considered in this simulation, the  $P_{ce}$  of Doppler scaling factor starts to approach 100% when SNR reaches 8 dB. In addition, as  $N_u$  increases, the  $P_{ce}$ s of both methods are also improved. The reason for the improvement of CPDE is that the magnitude of the CAF at  $\alpha = 0$  and  $\eta = \eta_m$  is dependent upon the auto-correlation of CPs. As the number of OFDM blocks increases, a better detection of  $\eta_m$  is available for the Doppler estimator. The reason for the improvement of CFDE is that the cyclic feature of the signal becomes more obvious when the number of OFDM blocks increases. However, the improvement of the CFDE is much larger than the CPDE. Thus, it can be concluded that, for the CP-OFDM, the CFDE method reflects its advantage when more OFDM blocks ( $N_u > 3$ ) are considered. Also, we emphasis the significance of the CFDE method for ZP-OFDM signals, which has no

peak over delays and the CPDE becomes useless.

We now investigate the performance of the CFDE method for different Doppler scale factor. In Fig. 3, the  $P_{ce}$  versus SNR for different Doppler scale factor is presented. It is apparent from the figure that the proposed estimator is quite robust to Doppler scaling factor. In particular, with an increase of the Doppler scaling factor, the performance of the CFDE method gets better. The reason for this phenomenon is that larger Doppler scaling factor results in larger cyclic frequency, which provides the peaks being used to estimate  $\gamma$  further away from the main lobe of the CAF, making it easier to detect them.

#### 5. CONCLUSION

We have presented a novel approach for Doppler scale estimation for both CP-OFDM and ZP-OFDM signals applying to the doubly selective underwater acoustic channel. Our proposal is simple to implement and needs no training data or specially designed packet structure. The performance of the proposed estimator is verified through numerical simulations. It is shown that the CFDE performs better than the CPDE based on the redundant information contained within the CP under enough OFDM symbols (more than 3). Perhaps more significantly, CFDE can be used for ZP-OFDM signals as well, whereas CPDE cannot be used at all, because it has no CP to work with.

#### 6. APPENDIX

#### A. DISCUSSIONS ABOUT THE IRRELEVANCE BETWEEN DATA SYMBOLS

Defining an information sequence  $\{b_i = -1 \text{ or } 1, i = 1, 2, ...\}$ . We assume that these data bits are uncorrelated, and its first and second moment are  $E[b_i] = 0$  and  $E[|b_i|^2] = \sigma_b^2$ , respectively. Assuming *M*-ary quadrature amplitude modulation (QAM) is selected as the modulation scheme, then each data symbol contains  $k = log_2 M$  data bits and can be represented as  $X_m = b_m 2^{k-1} + ... + b_{m+k-1}$ . We shall discuss the irrelevance between data symbols  $X_m$ . It is easy to verify that  $E[X_m] = 0$ . The covariance between data symbol  $X_m$  and  $X_n$ , where  $|m - n| \ge k$ , can be calculated as

$$E[X_m X_n] = E\left[\left(b_m 2^{k-1} + \dots + b_{m+k-1}\right) \times \left(b_n 2^{k-1} + \dots + b_{n+k-1}\right)\right] = 0.$$
(12)

When m = n, the variance of the data symbol can be written as

$$E\left[\left|X_{m}\right|^{2}\right] = (2^{k-1} + \dots + 1)\sigma_{b}^{2}.$$
(13)

Base on the above analysis, it witnesses that the order of the modulation scheme has no effects on the irrelevance between the data symbols.

#### ACKNOWLEDGMENTS

This work is partly supported by the Major Research Development Program of Hunan Province of China (No. 2016JC2008), and by the National Nature Science Foundation of China under the grants 61604176, 61773264, 61633017, 61471237 and 61301223.

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