

Non-Uniform Doppler Compensation for Zero-Padded OFDM over Fast-Varying Underwater Acoustic Channels

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Abstract—Underwater acoustic channels are wideband in nature due to the fact that the signal bandwidth is not negligible with respect to the center frequency. OFDM transmissions over UWA channels encounter frequency-dependent Doppler drifts that destroy the orthogonality among OFDM subcarriers. In this paper, we propose a two-step approach to mitigating the frequency-dependent Doppler drifts in zero-padded OFDM transmissions over fast-varying channels: (1) non-uniform Doppler compensation via resampling that converts a “wideband” problem into a “narrowband” problem; and (2) high-resolution uniform compensation on the residual Doppler. Based on block-by-block processing, our receiver does not rely on channel dependence across OFDM blocks, and is thus desirable for fast-varying UWA channels. We test our receiver with data from a shallow water experiment at Buzzards Bay, Massachusetts. Our receiver achieves excellent performance even when the transmitter and the receivers have a relative speed up to 10 knots, where the Doppler drifts are several times larger than the OFDM subcarrier spacing.

I. INTRODUCTION

The success of multicarrier modulation in the form of OFDM in radio channels motivates its use in underwater acoustic communications; see e.g., [1]–[3]. However, underwater acoustic (UWA) channels are far more challenging than their radio counterparts, preventing direct application of OFDM detection methods developed for radio channels, and requiring a careful receiver design. Recently, there has been an increased interest in underwater OFDM communication, including [4] on a low-complexity adaptive OFDM receiver, [5] on a pilot-tone based block-by-block receiver, and [6] on a non-coherent OFDM receiver based on on-off-keying.

In this paper, we adopt zero-padded OFDM [7] for underwater acoustic communications. The performance of a conventional ZP-OFDM receiver is severely limited by the intercarrier interference (ICI) due to fast channel variations within each OFDM symbol. Furthermore, the UWA channel is wideband in nature due to the fact that the signal bandwidth is not negligible with respect to the center frequency. The resulting frequency-dependent Doppler drifts render existing

ICI reduction techniques used in radio channels not effective. We propose a two-step approach to mitigating the frequency-dependent Doppler drifts in zero-padded OFDM transmissions over fast-varying underwater acoustic channels: (1) non-uniform Doppler compensation via resampling that converts a “wideband” problem into a “narrowband” problem; and (2) high-resolution uniform compensation on the residual Doppler for best ICI reduction.

Our practical receiver algorithms rely on the preamble and the postamble of a packet consisting of multiple OFDM blocks to estimate the resampling factor, null subcarriers to facilitate high-resolution residual Doppler compensation, and pilot subcarriers for channel estimation. Based on block-by-block processing, our coherent receiver does not rely on channel dependence across OFDM blocks, and is thus effective for fast-varying underwater acoustic channels. To verify our approach, we have conducted an experiment in shallow water at Mudhole, Buzzards Bay, MA, on Dec. 15, 2006. The transmitter was moving from 600 meters towards the receiver with a varying speed between 3 knots to 10 knots. Even when the Doppler drifts are several times larger than the OFDM subcarrier spacing, the experimental results show that the proposed receiver achieves excellent performance. (With a bandwidth of 12kHz, the data rate was from 10.5 kbps to 14.5 kbps for different settings when no channel coding is used.) The results in this paper suggest that OFDM is an appealing option for high-rate underwater acoustic communications over fast-varying channels.

The rest of the paper is organized as follows. In Section II, the performance of a conventional receiver is discussed. The approach to mitigating the Doppler effects is presented in Section III, and the receiver algorithms are presented in Section IV. In Section V, we specify the experimental signal design, and in Section VI we report on the receiver performance. We conclude in Section VII.

II. ZERO-PADDED OFDM TRANSMISSION

Let T denote the OFDM duration and T_g the guard interval. The total OFDM block duration is $T' = T + T_g$. The frequency spacing is $\Delta f = 1/T$. The k th subcarrier is at frequency

$$f_k = f_c + k\Delta f, \quad k = -K/2, \dots, K/2 - 1, \quad (1)$$

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where f_c is the carrier frequency and K subcarriers are used so that the bandwidth is $B = K\Delta f$.

Let us consider one ZP-OFDM block. Let $d[k]$ denote the information symbol to be transmitted on the k th subcarrier. We introduce null subcarriers in the transmission. The set of active subcarriers \mathcal{S}_A and the set of null subcarriers \mathcal{S}_N satisfy $\mathcal{S}_A \cup \mathcal{S}_N = \{-K/2, \dots, K/2 - 1\}$. The transmitted signal in passband is

$$s(t) = \text{Re} \left\{ \left[\sum_{k \in \mathcal{S}_A} d[k] e^{j2\pi k \Delta f t} g(t) \right] e^{j2\pi f_c t} \right\}, t \in [0, T + T_g], \quad (2)$$

where $g(t)$ describes the zero-padding operation as

$$g(t) = \begin{cases} 1, & t \in [0, T] \\ 0, & t \in [T, T + T_g]. \end{cases} \quad (3)$$

We consider a multipath underwater channel that has the impulse response as

$$c(t, \tau) = \sum_p A_p(t) \delta(\tau - \tau_p(t)), \quad (4)$$

where $A_p(t)$ is the path amplitude and $\tau_p(t)$ is the time-varying path delay. For ease of presentation, we assume that all paths have similar Doppler rate

$$\tau_p(t) \approx \tau_p - at, \quad (5)$$

and the path gains $A_p(t)$ and the Doppler rate are constant over the block duration T' .

The received signal in passband is then

$$\tilde{y}(t) = \text{Re} \left\{ \sum_p A_p \left[\sum_{k \in \mathcal{S}_A} d[k] e^{j2\pi k \Delta f (t+at-\tau_p)} g(t+at-\tau_p) \right] \cdot e^{j2\pi f_c (t+at-\tau_p)} \right\} + \tilde{n}(t), \quad (6)$$

where $\tilde{n}(t)$ is the additive noise. Converting $\tilde{y}(t)$ to its baseband version $y(t)$, we have

$$y(t) = \sum_{k \in \mathcal{S}_A} \left\{ d[k] e^{j2\pi k \Delta f t} e^{j2\pi a f_k t} \cdot \left[\sum_p A_p e^{-j2\pi f_k \tau_p} g(t+at-\tau_p) \right] \right\} + n(t), \quad (7)$$

where $n(t)$ is the additive noise in baseband. We observe from (7) two effects:

- (i) the signal from each path is scaled in duration, from T to $T/(1+a)$;
- (ii) each subcarrier experiences a Doppler shift $e^{j2\pi a f_k t}$, which depends on the frequency of each subcarrier. Since the bandwidth of OFDM is comparable to the center frequency, the Doppler shifts on different OFDM subcarriers differ considerably; i.e., the narrowband assumption does not hold true.

We now present the performance of a conventional OFDM receiver that does not perform any Doppler compensation [7]. Overlapping and adding of the received signal followed by

FFT processing leads to the output of the demodulator in the m -th subchannel as

$$y_m = \frac{1}{T} \int_0^T [y(t) + y(t+T)] e^{-j2\pi m \Delta f t} dt. \quad (8)$$

Substituting (7) into (8) and assuming that T_g is larger than the channel delay spread, we obtain

$$y_m = C \left(\frac{f_m}{1+a} \right) \sum_{k \in \mathcal{S}_A} d[k] \rho_{m,k} + n_m \quad (9)$$

where we define

$$C(f) = \sum_p A_p e^{-j2\pi f \tau_p}, \quad \alpha_{m,k} = \frac{(m-k) + a f_k / \Delta f}{1+a}, \quad (10)$$

$$\rho_{m,k} = \frac{1}{1+a} e^{j\alpha_{m,k}} \text{sinc}(\alpha_{m,k}). \quad (11)$$

The desired signal in y_m is $C(f_m/(1+a))\rho_{m,m}d[m]$, and the rest is the ICI pulse additive noise. The signal to interference-plus-noise ratio is

$$\gamma_m = \frac{|\rho_{m,m}|^2 \sigma_d^2}{\sigma_v^2 / |C(f_m/(1+a))|^2 + \sum_{k \neq m} |\rho_{m,k}|^2 \sigma_d^2}, \quad (12)$$

where σ_v^2 is the noise variance and $\sigma_d^2 = E[|d[m]|^2]$. Note that γ_m has a floor which does not depends on the channel frequency response when σ_v^2 goes to zero.

III. MITIGATING THE DOPPLER EFFECT

We propose a two-step approach to mitigating the frequency-dependent Doppler drifts due to fast-varying underwater acoustic channels:

1. Non-uniform Doppler compensation via resampling. This step converts a ‘‘wideband’’ problem into a ‘‘narrowband’’ problem.
2. High-resolution uniform compensation on residual Doppler by modeling it as induced by carrier frequency offset (CFO). This step fine-tunes the CFO term corresponding to a ‘‘narrowband’’ model for best ICI reduction.

For convenience, let us present these steps using baseband signals. On the first step, we resample the received waveform $y(t)$ with a resampling factor b :

$$z(t) = y \left(\frac{t}{1+b} \right). \quad (13)$$

Resampling has two effects: (1) it rescales the waveform, and (2) it introduces a frequency-dependent Doppler compensation. With $y(t)$ in (7), we have

$$z(t) = e^{j2\pi f_c t \frac{1+a}{1+b}} \sum_{k \in \mathcal{S}_A} \left\{ d[k] e^{j2\pi k \Delta f \frac{1+a}{1+b} t} \left[\sum_p A_p e^{-j2\pi f_k \tau_p} g \left(\frac{1+a}{1+b} t - \tau_p \right) \right] \right\}. \quad (14)$$

The target is to make

$$\frac{1+a}{1+b} \quad (15)$$

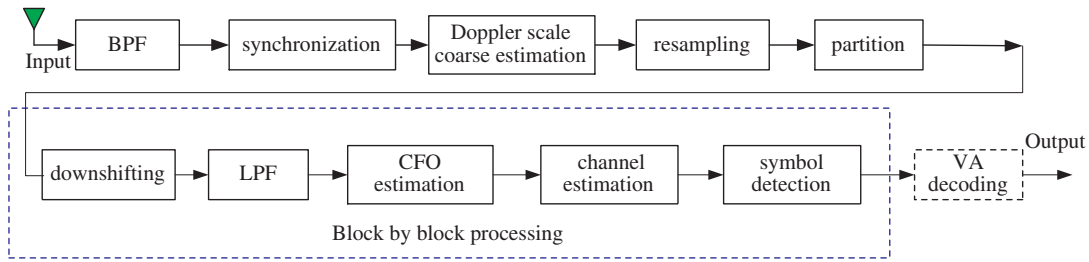


Fig. 1. Block diagram of the receiver

as close as possible to one. After resampling, we have

$$z(t) \approx e^{j2\pi f_c t \frac{a}{1+b}} \sum_k d[k] e^{j2\pi k \Delta f t} \left[\sum_p A_p e^{-j2\pi f_k \tau_p} g(t - \tau_p) \right] \quad (16)$$

The Doppler effect becomes almost the same for all subcarriers. Hence, a wideband OFDM system is converted into a narrowband OFDM system with a single equivalent CFO as

$$\epsilon = \frac{a}{1+b} f_c. \quad (17)$$

Compensating for CFO in $z(t)$, we obtain

$$e^{-j2\pi \epsilon t} z(t) = \sum_{k \in \mathcal{S}_A} d[k] e^{j2\pi k \Delta f t} \left[\sum_p A_p e^{-j2\pi f_k \tau_p} g(t - \tau_p) \right], \quad (18)$$

which leads to ICI-free reception as the channel is time-invariant. De-scaling and de-rotation of the received signal restore the orthogonality of the subcarriers of ZP-OFDM.

In practice, the scale factor b and the CFO ϵ need to be determined from the received data. They can be estimated either separately or jointly. In the next section, we will develop practical algorithms for Doppler scale and CFO estimation.

IV. RECEIVER ALGORITHMS

The received signal is directly sampled and all processing is performed on discrete-time signal. Fig. 1 depicts the receiver processing. Many steps in the receiver diagram are self-explanatory. We next present several key modules.

A. Doppler scale estimation

Doppler scale coarse estimation is based on the preamble and postamble of a data packet¹. The packet structure is shown in Fig. 2. This idea has been used in [8] for single carrier transmissions. Via synchronization with the preamble and postamble, the receiver estimates the time duration of a packet as T_{rx} . The time duration of this packet at the transmitter side is T_{tx} . By comparing T_{rx} with T_{tx} , the receiver infers how the received signal has been compressed or dilated by the channel:

$$T_{rx} = (1 + a)T_{tx} \quad \Rightarrow \quad \hat{a} = \frac{T_{rx}}{T_{tx}} - 1. \quad (19)$$

¹This pre- and post-amble approach requires a nearly constant moving speed during the transmission of a packet. Speed changes within a packet might considerably deteriorate the receiver performance of such an approach.

The Doppler scale factor is related to the relative speed v between the transmitter and the receiver by $a = v/c$, where c is the speed of sound. Hence, the speed estimate is

$$\hat{v} = c \cdot \hat{a}. \quad (20)$$

The receiver then resamples the packet using a resampling factor $b = \hat{a}$. The resampling operation introduces frequency-dependent Doppler offsets: at the k th subcarrier, the Doppler offset is

$$f_{\text{Doppler}} = \hat{a} f_k. \quad (21)$$

B. CFO estimation

We use null subcarriers to facilitate the finding of the CFO. We collect $K + L$ samples after the resampling operation for each OFDM block into a vector $\mathbf{z} = [z(0), \dots, z(K + L - 1)]^T$, where L is the channel length in discrete-time. Let $(\cdot)^T$ and $(\cdot)^H$ denote transpose and Hermitian transpose, respectively. Define a $(K + L) \times 1$ vector as $\mathbf{f}_m = [1, e^{j2\pi m/K}, \dots, e^{j2\pi m(K+L-1)/K}]^T$. Define a $(K + L) \times (K + L)$ diagonal matrix as $\mathbf{\Gamma}(\epsilon) = \text{diag}(1, e^{j2\pi T_c \epsilon}, \dots, e^{j2\pi T_c (K+L-1)\epsilon})$, where $T_c = T/K$ is the time interval for each sample. The energy on the null subcarriers is used as the cost function

$$J(\epsilon) = \sum_{m \in \mathcal{S}_N} |\mathbf{f}_m^H \mathbf{\Gamma}^H(\epsilon) \mathbf{z}|^2. \quad (22)$$

If the receiver compensates the data samples with the correct CFO before FFT operation, the null subcarriers will not see the ICI spilled over from neighboring data subcarriers. Hence, an estimate of ϵ can be found through

$$\hat{\epsilon} = \arg \min_{\epsilon} J(\epsilon), \quad (23)$$

which can be solved via one-dimensional search on ϵ . This high-resolution algorithm corresponds to the MUSIC-like algorithm proposed in [9] for OFDM with cyclic prefix.

C. Pilot-tone based channel estimation

After resampling and CFO compensation, the ICI is greatly reduced. We use equi-spaced pilot tones for channel estimation, as in [5].

TABLE I
INPUT DATA STRUCTURE AND ACHIEVED BIT RATES

K	input bits or symbols (N_d)	# of active subcarriers (K_a)	# of null subcarriers (K_n)	# of blocks in a packet (N_b)	raw bit rates over $B = 12$ kHz $2K_a/(T + T_g)$	bit rates excluding $K/4$ pilot tones (uncoded) $2(K_a - K/4)/(T + T_g)$
512	30976	484	28	64	14.30 kbps	10.52 kbps
1024	30976	968	56	32	17.55 kbps	12.90 kbps
2048	30976	1936	112	16	19.79 kbps	14.55 kbps

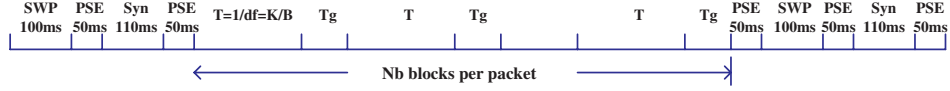


Fig. 2. Each packet consists of preamble, N_b OFDM blocks, and postamble.

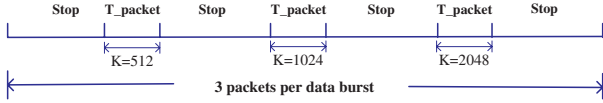


Fig. 3. Each data burst consists of three packets, with $K = 512$, $K = 1024$, and $K = 2048$, respectively.

V. SIGNAL DESIGN FOR UNDERWATER EXPERIMENTS

Our transmitted signal is designed as follows. The bandwidth of our OFDM signal is $B = 12$ kHz, and the carrier frequency is $f_c = 27$ kHz. The transmitted OFDM signal occupies the frequency band of 21 to 33 kHz. We use zero-padded OFDM with a guard interval of $T_g = 25$ ms per OFDM symbol. We test three different settings for the number of subcarriers: $K = 512$, $K = 1024$, and $K = 2048$. We use rate $2/3$ convolutional coding (obtained by puncturing a rate $1/2$ code with polynomial (23,35)) and QPSK modulation. We let each packet have $N_d = 30976$ information bits. Hence, each packet will contain $N_b = N_d/K_a$ OFDM blocks, where K_a is the number of active carriers. For $K = 512, 1024, 2048$, each packet contains $N_b = 64, 32, 16$ OFDM blocks, respectively. These parameters are summarized in Table I.

Fig. 3 depicts one data burst that consists of three packets with $K = 512$, $K = 1024$, and $K = 2048$, respectively. During the experiments, the same data burst was transmitted multiple times while the transmitter was moving.

VI. EXPERIMENTAL RESULTS

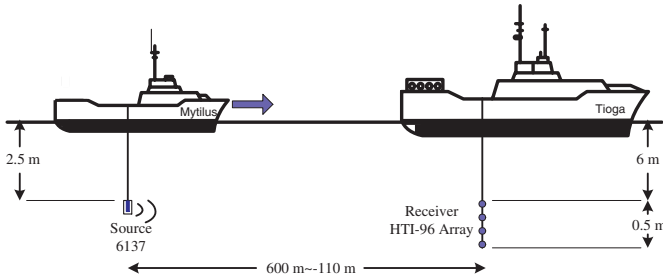


Fig. 4. The experiment setup for the Dec. 15, 2006 experiment

The WHOI acoustic communication group conducted the experiment on Dec. 15, 2006. The experiment location was at Mudhole, Buzzards Bay, MA. The transmitter was submerged at a depth of about 2.5 meters. The receiver was a four-element array (of length 0.5 m) submerged at a depth of about 6 meters. The transmitter was mounted on the arm of the vessel Mytilus, and the receiver array was mounted on the arm of the vessel Tioga. OFDM signals were transmitted while Mytilus was moving towards Tioga, starting at 600 m, passing by Tioga, and ending at about -100 m in distance. The configuration is shown in Fig. 4.

The data burst in Fig. 3 was transmitted multiple times when Mytilus was moving towards Tioga. The received signal was directly A/D converted. The receiver array has four elements, providing four parallel received data streams. The received signal observed on one element is shown in Fig. 5, which contains 7 data bursts (21 packets).

We directly observe that:

- 1) The received power is increasing before packet 19 and decreasing after packet 19. This is because Mytilus passed Tioga around that time. Hence, the transmitter was moving towards the receiver before packet 19 while moving away from the receiver after packet 19.
- 2) An increase in the noise level is observed around packet 19. This noise is from the Mytilus when it is very close to Tioga.

Simple data processing reveals the following:

- 3) The signal prior to packet 19 is compressed, which means that the transmitter was moving towards the receiver. The signals after that is dilated, which means that transmitter was moving away from the receiver. This agrees with the observation based on the received power.

We next present numerical results based on the receiver processing outlined in Section IV.

A. Doppler scale estimation

As shown in Fig. 5, we have 21 packets transmitted. We use the algorithm of Section IV-A to estimate the Doppler scale on the packet level. Hence, we have 21 estimated Doppler scales. Based on each Doppler scale, we estimate

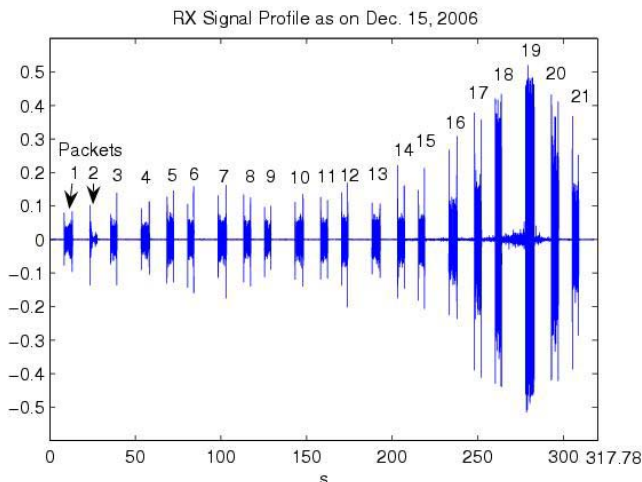


Fig. 5. The received signal

the relative speed between the transmitter and the receiver using (20) and a nominal sound speed of $c = 1500$ m/s. Note that the waveform compression/dilation introduces frequency-dependent Doppler shifts. We evaluate the Doppler shifts at the carrier frequency f_c (which is the same as the frequency f_0 of the 0th subcarrier). Table II summarizes the results for element 1.

We see from Table II that the Doppler shifts are much larger than the OFDM subcarrier spacing. For example, if $\hat{v} = -8.30$ knots, which means Mytilus was moving towards Tioga at such a speed, the Doppler shift is -76.98 Hz at $f_c = 27$ kHz, while the subcarrier spacing is only $\Delta f = 23.44, 11.72, 5.86$ Hz when $K = 512, 1024, 2048$. Re-scaling the waveform (even coarsely) is a necessary step to reduce the Doppler effect considerably, in a non-uniform fashion.

Table II also reveals how Mytilus was moving. At first, Mytilus was accelerating towards Tioga. When it was approaching Tioga, it slowed down but continued to move until it passed Tioga. During the time of transmitting packets 18 and 19, Mytilus was passing by Tioga, as the speed changed from a negative value to a positive value.

B. High-resolution residual Doppler estimation

The high-resolution CFO estimation is done on a block-by-block basis, as detailed in Section IV-B. Fig. 6 shows the CFO estimates for packets 5 and 17, respectively, where $K = 1024$ and each packet has 32 OFDM blocks. We observe that the CFO changes from block to block roughly continuously but cannot be regarded as constant. The CFO estimate is on the order of half of the subcarrier spacing. Without the CFO fine tuning, the receiver does not work.

C. Uncoded performance, single channel reception

Due to the large amount of blocks received on each of the four elements, we choose to demonstrate one set of results. In particular, we show the results for the $K = 512$ case, which corresponds to the packets 1, 4, 7, 10, 13, 16, 19. For the $K = 512$ case, each packet consists of 64 OFDM blocks.

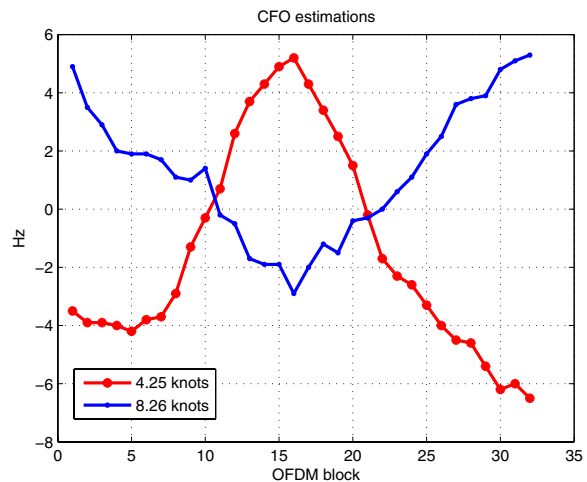


Fig. 6. The estimated residual Doppler is shown for two examples. One is for packet 5 with a relative speed of 4.25 knots, the other is for packet 17 with a relative speed of 8.26 knots. It can be seen that the CFO fluctuates rapidly from block to block.

We have the following observations on data demodulation.

- 1) Without coding, our receiver is able to provide good performance.
- 2) The percentage of erroneously detected bits was zero when the moving speed is low (e.g., packet 1) or when the moving speed is very stable (e.g., packet 16).
- 3) Our receiver is able to handle a moving speed up to 9.04 knots.
- 4) There are several *consecutive* bad blocks in packets 19 and 20, which led to a high number of bit errors. The reason is that the TX was moving from 600m to the RX. At the time of transmitting packets 19 and 20, the TX was just passing by the RX. The Doppler frequencies may have some unexpected jitting when they were changing from negative to positive values. Also, the noise level was suddenly high during the passing. When the TX had passed by the RX and went away from it, which was the case in packet 21, everything worked again and the number of bit errors went to almost zero.

D. Coded performance, single channel reception

All the information bits have been coded by a rate $2/3$ convolutional code obtained by puncturing a rate $1/2$ code. To test the coded performance, we apply the Viterbi algorithm after the OFDM demodulation.

Most of the observed errors on the block level disappear. Corresponding to all cases listed in Table III, only blocks 23 to 33 of packet 19 have decoding errors, and the rest do not. The errors occur in a few consecutive blocks in packet 19, where the signal experienced a change from a negative Doppler to a positive Doppler and a suddenly increased noise. Once the Doppler becomes stable, our receiver has an acceptable performance. Thanks to the block-by-block processing, decoding errors in previous blocks have no impact on future blocks. This is desirable for fast-varying channels.

TABLE II
COARSE ESTIMATION OF RELATIVE SPEED AND DOPPLER SHIFTS FOR ELEMENT 1.

Packet	Doppler shift due to scaling at f_c (Hz)	Relative speed (knots)	Packet	Doppler shift due to scaling at f_c (Hz)	Relative speed (knots)
1	-17.34	-1.86	12	-41.79	-4.50
2	-42.49	-4.58	13	-42.45	-4.58
3	-41.87	-4.52	14	-64.04	-6.91
4	-40.29	-4.35	15	-76.98	-8.30
5	-39.37	-4.25	16	-83.95	-9.04
6	-39.69	-4.27	17	-76.68	-8.26
7	-41.91	-4.52	18	-73.34	-7.90
8	-41.62	-4.48	19	53.96	5.82
9	-40.34	-4.35	20	58.34	6.29
10	-39.68	-4.26	21	57.15	6.17
11	-40.60	-4.38			

TABLE III
UNCODED BIT ERROR RATE FOR $K = 512$, ELEMENT 4. (I)

Packet Block	1 (-1.86 knots)	4 (-4.35 knots)	7 (-4.52 knots)	10 (-4.26 knots)	13 (-4.58 knots)	16 (-9.04 knots)	19 (5.82 knots)
...
21	0	0	0.001	0.001	0.008	0	0.028
22	0	0	0.003	0.004	0.006	0	0.090
23	0	0	0.007	0.001	0	0	0.146
24	0	0	0.004	0	0.007	0	0.612
25	0	0	0.003	0.003	0.007	0	0.639
26	0	0	0.004	0.003	0.004	0	0.647
27	0	0	0.003	0.001	0	0	0.646
28	0	0	0.003	0.003	0.003	0	0.636
29	0	0	0	0.003	0.004	0	0.629
30	0	0	0	0.003	0.003	0	0.636
31	0	0	0.004	0.006	0.001	0	0.625
32	0	0	0.001	0.003	0	0	0.190
33	0	0	0.006	0.001	0.003	0	0.140
34	0	0	0.006	0.004	0.001	0	0.059
35	0	0	0.007	0.003	0.001	0	0.014
...
Average BER of 64 blocks	0	2.2×10^{-4}	2.5×10^{-3}	4.3×10^{-3}	1.4×10^{-3}	0	9.6×10^{-2}

VII. CONCLUSIONS

In this paper we investigated the application of zero-padded OFDM in fast-varying underwater acoustic channels. We proposed a two-step approach to mitigating the frequency-dependent Doppler drifts, namely non-uniform Doppler compensation via resampling, followed by high-resolution uniform compensation on the residual Doppler. We used null subcarriers to facilitate the Doppler compensation, and pilot subcarriers for channel estimation. Our receiver is based on block-by-block processing, bypassing the need of channel dependence across OFDM blocks, and is thus suitable for fast-varying underwater acoustic channels. We tested our methods in a shallow water experiment. Excellent performance was achieved even when the transmitter and the receivers had a relative speed up to 10 knots, where the Doppler drifts were several times larger than the OFDM subcarrier spacing.

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