Oversampled OFDM Detector for MIMO Underwater Acoustic Communications

Jun Tao[†], Jingxian Wu[*], Yahong Rosa Zheng[†], and Chengshan Xiao[†]

[†]Dept. of Electrical & Computer Eng., Missouri University of Science & Technology, Rolla, MO 65409
[*] Dept. of Electrical Engineering, University of Arkansas, Fayetteville, AR 72701, USA

***Abstract*— In this paper, an oversampled orthogonal frequency division multiplexing (OFDM) detector is proposed for multiple-input multiple-output (MIMO) underwater acoustic (UWA) communications. With simple time-domain oversampling and frequency-domain signal processing, the new oversampled OFDM (OOFDM) detector can achieve multipath diversity that is unavailable in a conventional uncoded OFDM system, while maintaining the similar complexity as the conventional OFDM system. The new OOFDM detector can be directly applied to the received OFDM signals without changing the structure of the OFDM transmitter. The proposed MIMO OOFDM detector has been tested by field trial data collected in the SPACE08 experiment launched near Martha's Vineyard, Edgartown, MA, in 2008. Experimental results show that the oversampled OFDM detector outperforms the conventional OFDM detector, and an oversampling factor of 2 achieves the best tradeoff between the performance gain and the computational complexity.**

I. Introduction

One of the main challenges in underwater acoustic (UWA) communications is the extremely long delay spread of the channel, which introduces severe inter-symbol interference (ISI). The time-domain equalization of such a highly dispersive channel demands an equalizer with a large number of taps, resulting in very high complexity. Multicarrier modulation, such as orthogonal frequency division multiplexing (OFDM), can achieve low-complexity equalization in the frequency domain by transmitting narrow-band signals over mutually orthogonal subcarriers [1]. However, compared to single-carrier systems, uncoded OFDM systems can not achieve multipath diversity due to the frequency-nonselective nature of the subcarriers used in multicarrier communication.

Several linear precoding/decoding techniques have been proposed to achieve multipath diversity in an OFDM system [2], [3], at the cost of intercarrier interference (ICI) and an increased computational complexity. A more common (also much simpler) method resorts to the time-domain oversampling operation, which leads to the fractional-sampling (FS) OFDM (FS-OFDM) system [4] or the oversampled OFDM (OOFDM) system [5]. By sampling the received signal at a rate higher than the symbol rate, multipath diversity that is unavailable in conventional uncoded OFDM systems, can be achieved in OOFDM systems. With an oversampling factor of $\eta > 1$, where η is an integer, each modulation symbol is equivalently transmitted over η subcarriers. Therefore, extra diversity gain can be obtained by performing linear combining in the frequency domain. The results in [4] and [5] are obtained under the assumption of perfect channel state information at the receiver side, and they are only applicable to single-input single-output (SISO) systems.

This paper proposes a zero-padding (ZP) OOFDM system for multiple-input multiple-output (MIMO) UWA communications with practical channel estimation. Compared to OOFDM systems using cyclic prefix (CP) [5], the ZP OOFDM requires less transmission power due to the absence of the CP. An overlap-add (OLA) method is used in combination with the ZP OOFDM detector to ensure the orthogonality among the subcarriers. The OOFDM detector is applied to process the undersea trial data collected in the SPACE08 experiment launched near Martha's Vineyard, Edgartown, MA, in 2008. In the new detector, the channel estimation is performed in the time domain and the symbol detection is operated in the frequency domain, both with oversampled signals. The oversampling operation enables better channel estimation in the time domain, and additional multipath diversity in the frequency domain. Both of these two factors contribute to performance improvement over conventional OFDM systems, as demonstrated through the experiment results. The detection performance increases with the oversampling factor η, with the largest gain obtained from $\eta = 1$ (no oversampling) to $\eta = 2$.

The remainder of the paper is organized as follows. In Section II, the continuous-time MIMO OFDM system using ZP is described. Section III presents the MIMO OOFDM detector structure, which includes both time-domain oversampling and frequency-domain symbol detection. Experimental results are presented in Section IV, and Section V concludes the paper.

II. System Model

Consider an $N \times M$ MIMO-OFDM UWA communication system, where N and M are the numbers of transmit transducers and receive hydrophones, respectively. The information bit stream to be transmitted is demultiplexed into N substreams, which are then spatially multiplexed onto the N transducers. The data stream at each transducer is independently encoded, modulated and then transmitted through a P-subcarrier OFDM scheme. The transmitted time-domain OFDM samples on the n-th transducer can be expressed as

$$x_{n,k} = \frac{1}{\sqrt{P}} \sum_{p=0}^{P-1} S_{n,p} e^{j2\pi \frac{kp}{P}}, \quad (1)$$

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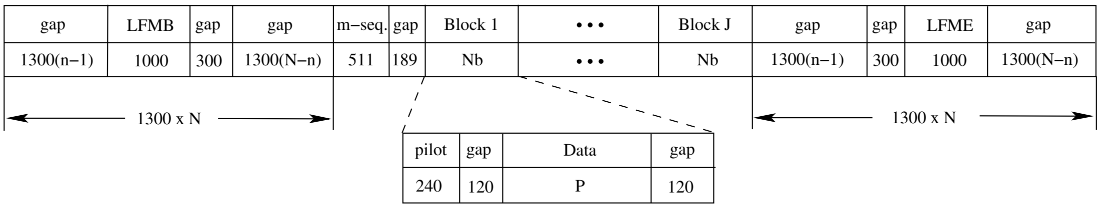

Fig. 1. Structure of transmitted packet from the n-th transducer

where $S_{n,p} \in \mathcal{S}$ is the modulated symbol transmitted by the n-th transducer on the p-th subcarrier, $\mathcal{S}$ is the modulation constellation set, and $x_{n,k}$ is the k-th time-domain sample of the OFDM symbol, for $k = 0, 1, \cdots, P-1$.

During the SPACE08 experiment for multicarrier MIMO UWA communication, the time-domain samples were organized in packets, with the packet structure on the n-th transducer shown in Fig. 1. From the figure, auxiliary signals including two linear frequency modulation (LFM) signals named LFMB and LFME and one m-sequence, are inserted in the packet. The functions of the LFM signals and the m-sequence, can be found in [6]. Every packet carries J blocks. Each block consists of one OFDM symbol which includes P time-domain data samples, one pilot block of size $N_p = 240$, and two $N_g = 120$ ZP blocks as prefix and postfix (denoted as "gap" in the figure). The pilot block is used for channel estimation, and the ZP blocks are used to avoid inter-symbol interference.

The time-domain samples are passed through a transmit filter and transmitted through the UWA channel. At the m-th hydrophone, the signal at the output of the receive filter can be expressed by

$$y_m(t) = \sum_{n=1}^{N} \sum_{k=-\infty}^{\infty} x_{n,k} h_{m,n}(t, t - kT_s) + w_m(t) \tag{2}$$

where $h_{m,n}(t, \tau)$ is the time-varying continuous-time composite channel impulse response (CIR) between the n-th transducer and the m-th hydrophone, and it incorporates the effects of the transmit filter, the physical UWA channel, and the receive filter, T_s is the sample period of the time-domain sequence $x_{n,k}$, and $w_m(t)$ denotes the random noise process on the m-th hydrophone.

A conventional OFDM system will sample $y_m(t)$ with a sampling period T_s and then convert the time-domain samples into the frequency domain with a P-point discrete Fourier transform (DFT). We will show in the next section that a better performance can be achieved with the OOFDM detector that performs oversampling on $y_m(t)$.

III. Proposed Oversampled OFDM Detector

The block diagram of the OOFDM detector is shown in Fig. 2. From the figure, the received analog signals from the M hydrophones are first passed through an oversampler before detection. The upsampled samples are then delivered to the channel estimator for channel estimation and to the OOFDM detector for symbol detection. The OOFDM detector outputs N detected bit streams, which are decoded by the N channel decoders.

The channel estimator performs time-domain channel estimation with the assistance of the pilot symbols as described in [6]. The detailed operations of the OOFDM detector, including the time-domain oversampling and the frequency-domain symbol detection, are described in the following two subsections.

A. Time-Domain Oversampling with an Overlap-Add Method

For the oversampler with a sampling period $T = \frac{T_s}{\eta}$, where the oversampling factor η is an integer larger than 1, the time-domain samples corresponding to the m-th hydrophone can be expressed by

$$y_{m,k} = \sum_{n=1}^{N} \sum_{l=0}^{\eta L-1} h_{m,n}(k,l) \tilde{x}_{n,k-l} + w_{m,k} \tag{3}$$

where ηL is the length of the T-spaced CIR, $y_{m,k} = y_m(kT)$, $h_{m,n}(k,l) = h_{m,n}(kT, lT)$, $w_{m,k} = w_m(kT)$, and

$$\tilde{x}_{n,k} = \begin{cases} x_{n,\frac{k}{\eta}}, & \frac{k}{\eta} \text{ is integer,} \\ 0, & \text{otherwise.} \end{cases} \tag{4}$$

It is assumed here that the channel length is an integer multiple of the oversampling factor η. This assumption can always be met by padding zeros at the end of the channel. In addition, the channel can be treated as quasi-static within one OFDM block, *i.e.*, $h_{m,n}(k,l) \approx h_{m,n}(l)$ for $k \in [-\eta(N_p+N_g), \eta(P+N_g)-1]$. This approximation is verified by the experimental data obtained from the SPACE08 experiment.

After the removal of the pilot symbols and the prefixed ZP samples, we denote the time-domain samples within one OFDM block (including both the data samples and the postfixed ZP samples) as $\{y_{m,k}\}_{k=0}^{\eta(P+N_g)-1}$. Due to the oversampling, the original $(P+N_g)$ samples with sample period T_s are increased to $\eta(P+N_g)$ samples with sample period T.

Since ZP instead of CP is used during the transmission, the received time-domain signal can no longer be expressed as the cyclic convolution between the transmitted signal and the channel, thus DFT cannot be directly applied to the samples. To achieve an equivalent cyclic convolution expression, we resort to the OLA method, where the last $\eta L - 1$

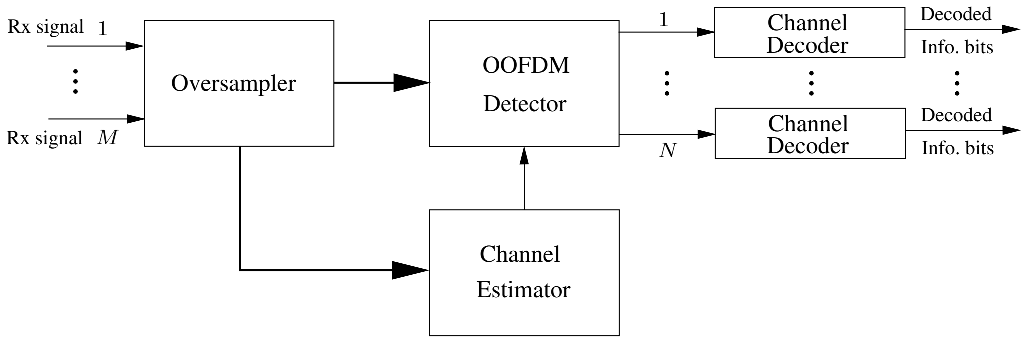

Fig. 2. Oversampled OFDM detector

samples, $\{y_{m,k}\}_{k=\eta P}^{\eta P+\eta L-2}$, are added to the first $\eta L-1$ samples, $\{y_{m,k}\}_{k=0}^{\eta L-2}$, as $\tilde{y}_{m,k}=y_{m,k}+y_{m,\eta P+k}$, for $k=0,1,\cdots,\eta L-2$. After OLA, the following system model is obtained

$$\tilde{\mathbf{y}}_m=\sum_{n=1}^{N}\bar{\mathbf{h}}_{m,n}\mathbf{x}_n+\tilde{\mathbf{w}}_m \tag{5}$$

where $\tilde{\mathbf{y}}_m=[\tilde{y}_{m,0},\cdots,\tilde{y}_{m,\eta L-2},y_{m,\eta L-1},\cdots,y_{m,\eta P-1}]^t\in\mathcal{C}^{\eta P\times 1}$, $\mathbf{x}_n=[x_{n,0},x_{n,1},\cdots,x_{n,L-1},\cdots,x_{n,P-1}]^t\in\mathcal{C}^{P\times 1}$, $\tilde{\mathbf{w}}_m=[\tilde{w}_{m,0},\cdots,\tilde{w}_{m,\eta L-2},w_{m,\eta L-1},\cdots,w_{m,\eta P-1}]^t\in\mathcal{C}^{\eta P\times 1}$ with $\tilde{w}_{m,k}=w_{m,k}+w_{m,\eta P+k}$, for $k=0,1,\cdots,\eta L-2$. The operation $(\cdot)^t$ denotes matrix transpose. The channel matrix, $\bar{\mathbf{h}}_{m,n}\in\mathcal{C}^{\eta P\times P}$, can be expressed by

$$\bar{\mathbf{h}}_{m,n}=[\mathbf{h}_{m,n}^{1},\mathbf{h}_{m,n}^{\eta+1},\mathbf{h}_{m,n}^{2\eta+1},\cdots,\mathbf{h}_{m,n}^{(P-1)\eta+1}] \tag{6}$$

where $\mathbf{h}_{m,n}^{1}=[h_{m,n}(0),h_{m,n}(1),\cdots,h_{m,n}(\eta L-2),h_{m,n}(\eta L-1),0,\cdots,0]^t\in\mathcal{C}^{\eta P\times 1}$, and $\mathbf{h}_{m,n}^{j}$ is obtained by cyclic shifting $\mathbf{h}_{m,n}^{1}$ downwards for $j-1$ positions. With the OLA method, the equivalent channel matrix $\bar{\mathbf{h}}_{m,n}$ is expressed as a submatrix of a circulant matrix. Such a circulant structure simplifies the frequency-domain symbol detection as described in the next subsection.

B. Frequency-Domain Symbol Detection

Due to the oversampling in the time domain, there are ηP time-domain samples at the receiver after the OLA operation. Performing the normalized ηP-point DFT on both sides of (5), leads to

$$\mathbf{Y}_m=\sum_{n=1}^{N}\bar{\mathbf{H}}_{m,n}\mathbf{S}_n+\mathbf{W}_m \tag{7}$$

where $\mathbf{Y}_m=\mathbf{F}_{\eta P}\tilde{\mathbf{y}}_m$, $\bar{\mathbf{H}}_{m,n}=\mathbf{F}_{\eta P}\bar{\mathbf{h}}_{m,n}\mathbf{F}_P^h\in\mathcal{C}^{\eta P\times P}$, $\mathbf{S}_n=\mathbf{F}_P\mathbf{x}_n$, $\mathbf{W}_m=\mathbf{F}_{\eta P}\tilde{\mathbf{w}}_m$, and $(\cdot)^h$ denotes the matrix Hermitian. The matrices $\mathbf{F}_{\eta P}\in\mathcal{C}^{\eta P\times\eta P}$ and $\mathbf{F}_P\in\mathcal{C}^{P\times P}$ are normalized ηP-point and P-point DFT matrices with their (p,q)-th entry being $(\mathbf{F}_{\eta P})_{p,q}=\frac{1}{\sqrt{\eta P}}e^{-j\frac{2\pi(p-1)(q-1)}{\eta P}}$ and $(\mathbf{F}_P)_{p,q}=\frac{1}{\sqrt{P}}e^{-j\frac{2\pi(p-1)(q-1)}{P}}$, respectively. The channel matrix $\bar{\mathbf{H}}_{m,n}$ can be represented as

$$\bar{\mathbf{H}}_{m,n}=[(\bar{\mathbf{H}}_{m,n}^{0})^t,(\bar{\mathbf{H}}_{m,n}^{1})^t,\cdots,(\bar{\mathbf{H}}_{m,n}^{\eta-1})^t]^t \tag{8}$$

where $\bar{\mathbf{H}}_{m,n}^{i}$ is a $P\times P$ diagonal matrix with its p-th diagonal element being

$$\begin{aligned}(\bar{\mathbf{H}}_{m,n}^{i})_{p,p} &\triangleq \bar{H}_{m,n}^{i}(p)\\ &= \frac{1}{\sqrt{\eta}}\sum_{l=0}^{\eta L-1}h_{m,n}(l)e^{-j\frac{2\pi(iP+p-1)l}{\eta P}}\end{aligned} \tag{9}$$

With the frequency-domain system representation given in (7), the symbol, $S_{n,p}$, can be considered as transmitted over η subcarriers with the channel coefficients $\{\bar{H}_{m,n}^{i}(p)\}_{i=0}^{\eta-1}$, and there is no interference among the subcarriers thanks to the diagonal structure of $\bar{\mathbf{H}}_{m,n}^{i}$. Therefore, with oversampling at the receiver, the OFDM system with P subcarriers at the transmitter is equivalently converted to a system with ηP subcarriers at the receiver, and each data symbol is equivalently transmitted over η subcarriers. Multipath diversity is thus achieved.

Since there is no interference among the symbols transmitted on different subcarriers, the system in (7) can be equivalently represented as P parallel subsystems, as

$$\mathbf{Y}_m^p=\bar{\mathbf{H}}_m^p\mathbf{S}^p+\mathbf{W}_m^p,\ \text{for } p=1,2,\cdots,P \tag{10}$$

where $\mathbf{Y}_m^p=[Y_{m,p},Y_{m,P+p},\cdots,Y_{m,(\eta-1)P+p}]^t\in\mathcal{C}^{\eta\times 1}$, $\mathbf{W}_m^p=[W_{m,p},W_{m,P+p},\cdots,W_{m,(\eta-1)P+p}]^t\in\mathcal{C}^{\eta\times 1}$, $\mathbf{S}^p=[S_{1,p},S_{2,p},\cdots,S_{N,p}]^t\in\mathcal{C}^{N\times 1}$, and the channel matrix $\bar{\mathbf{H}}_m^p\in\mathcal{C}^{\eta\times N}$ is defined as

$$\bar{\mathbf{H}}_m^p=\begin{bmatrix}\bar{H}_{m,1}^{0}(p) & \bar{H}_{m,2}^{0}(p) & \cdots & \bar{H}_{m,N}^{0}(p)\\ \bar{H}_{m,1}^{1}(p) & \bar{H}_{m,2}^{1}(p) & \cdots & \bar{H}_{m,N}^{1}(p)\\ \vdots & \vdots & \cdots & \vdots\\ \bar{H}_{m,1}^{\eta-1}(p) & \bar{H}_{m,2}^{\eta-1}(p) & \cdots & \bar{H}_{m,N}^{\eta-1}(p)\end{bmatrix}. \tag{11}$$

Stacking the frequency-domain signals from all the M hydrophones, $\{\mathbf{Y}_m^p\}_{m=1}^{M}$, into a column vector leads to

$$\mathbf{Y}^p=\bar{\mathbf{H}}^p\mathbf{S}^p+\mathbf{W}^p,\ \text{for } p=1,2,\cdots,P \tag{12}$$

where $\mathbf{Y}^p=[(\mathbf{Y}_1^p)^t,(\mathbf{Y}_2^p)^t,\cdots,(\mathbf{Y}_M^p)^t]^t\in\mathcal{C}^{\eta M\times 1}$, $\mathbf{W}^p=[(\mathbf{W}_1^p)^t,(\mathbf{W}_2^p)^t,\cdots,(\mathbf{W}_M^p)^t]^t\in\mathcal{C}^{\eta M\times 1}$, and $\bar{\mathbf{H}}^p=[(\bar{\mathbf{H}}_1^p)^t,(\bar{\mathbf{H}}_1^p)^t,\cdots,(\bar{\mathbf{H}}_M^p)^t]^t\in\mathcal{C}^{\eta M\times N}$.

In the equivalent system representation in (12), each symbol is transmitted over η subcarriers, and the signal transmitted over each subcarrier is received by M hydrophones. Therefore, a hybrid frequency-space diversity of order ηM is achieved. When $\eta = 1$, the system degrades to a conventional MIMO OFDM system. Therefore, additional performance gain is collected by the OOFDM system with the extra multipath diversity that is not available in a conventional OFDM system.

Herein, the problem of estimating $\{\mathbf{S}_n\}_{n=1}^N$ based on the received signal $\{\mathbf{Y}_m\}_{m=1}^M$ is equivalent to estimating $\{\mathbf{S}^p\}_{p=1}^P$ from $\{\mathbf{Y}^p\}_{p=1}^P$ based on the P subsystems given in (12). For the p-th subsystem, applying the least square (LS) detection algorithm, we have

$$\hat{\mathbf{S}}^p = \left[(\bar{\mathbf{H}}^p)^h \bar{\mathbf{H}}^p\right]^{-1} (\bar{\mathbf{H}}^p)^h \mathbf{Y}^p \tag{13}$$

where $\eta M > N$ is assumed. Performing the above operation over all the P subsystems leads to the detection of $\{\mathbf{S}_n\}_{n=1}^N$.

IV. Experimental Results

The proposed MIMO OOFDM detection algorithm is applied to the trial data collected during the SPACE08 experiment. During the experiment, the sampling rate at the transmitter side was $f_s = \frac{1}{T_s} = 9.7656$ kilohertz (kHz), and the carrier frequency was $f_c = 13$ kHz. Three modulation schemes, QPSK, 8PSK and 16QAM, were adopted. The transmit equipment consisted of four transducers and the receive equipment consisted of twelve hydrophones. The transmission distances ranged from 200 meters to 1000 meters. The information bits on each transducer were convolutionally encoded then randomly interleaved before the symbol modulation. The experiment was performed with the number of subcarriers, P, being 1024, 2048, or 4096. OFDM symbols in the same packet have the same number of subcarriers. Within one packet, 8 OFDM symbols for each of the three modulation schemes are transmitted, resulting in $J = 24$ OFDM symbols per packet.

The signal at the output of the receive filter on each of the M hydrophones was recorded and stored for offline processing. Since the proposed MIMO OOFDM scheme only requires special processing over signals after the receive filter, it can be directly applied to the stored experiment data.

The length of the T_s-spaced equivalent discrete-time channel has been measured as $L = 100$ with the LFM signal, using the method given in [6]. For the detection of each OFDM symbol, the channel estimation is obtained with the pilot block prefixed the data samples. For an OOFDM system with $\eta = 2$, the estimated $\frac{T_s}{2}$-spaced equivalent discrete-time CIR for the channels from transducers 1 and 2 are shown in Figs. 3 and 4. With two times oversampling, the $\frac{T_s}{2}$-spaced discrete-time CIR has $\eta L = 200$ taps. The letters 'T' and 'H' in both figures denote the transducer and the hydrophone, respectively.

Table. I lists the number of bit errors detected in one packet with the conventional OFDM detector ($\eta = 1$), the OOFDM detector with $\eta = 2$, and the OOFDM detector with $\eta = 4$, respectively. The modulation scheme is QPSK. The number of subcarriers is $P = 1024$ and the transmission distance is 1000 meters. Five different system configurations with 2, 4, 6, 8, and 10 hydrophones have been studied. We have three observations from the table. First, the detection performance improves when the number of hydrophones used increases, for both conventional OFDM detector and the proposed OOFDM detector. Second, it is apparent from the results that the OOFDM detector with either $\eta = 2$ or $\eta = 4$ outperforms the conventional OFDM detector in all five system configurations. Third, the performance gain increases as the oversampling factor increases. The gain obtained from $\eta = 1$ (or conventional OFDM) to $\eta = 2$ is considerable, while it is relatively small when increasing η from 2 to 4. As a result, an oversampling factor of $\eta = 2$ provides the best tradeoff between complexity and performance gain.

Table. II lists the results for 8PSK modulation scheme, and all the other parameters are the same as the previous example. Similar observations are made as for Table. I, except that the number of bit errors increases due to a larger constellation size.

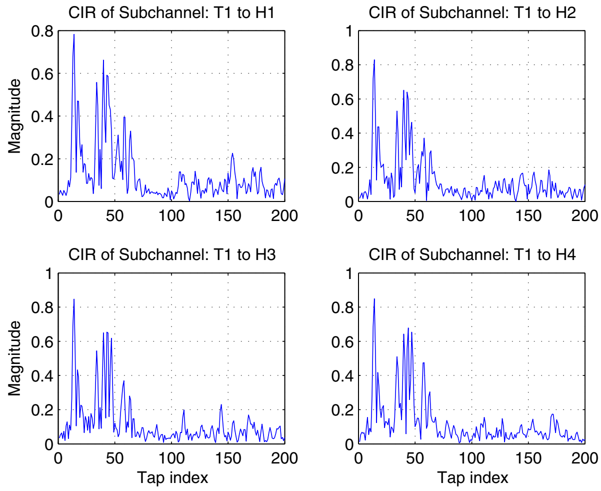

Fig. 3. Estimated UWA channels: $\eta = 2$, $L = 100$ (first transducer).

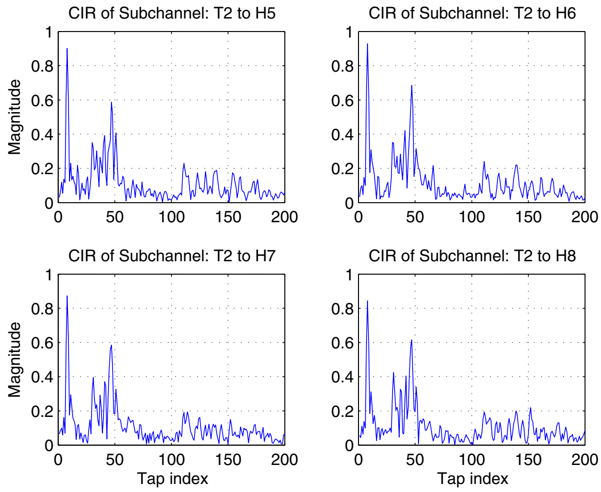

Fig. 4. Estimated UWA channels: $\eta = 2$, $L = 100$ (second transducer).

TABLE I

COMPARISON IN NUMBER OF DETECTION ERRORS (QPSK)

No. of channels \ Ovsp factor	1	2	4
2	3056	2847	2843
4	151	88	80
6	78	44	41
8	25	15	15
10	0	0	0

TABLE II

COMPARISON IN NUMBER OF DETECTION ERRORS (8PSK)

No. of channels \ Ovsp factor	1	2	4
2	8640	8474	8359
4	2638	2318	2304
6	823	639	618
8	621	493	492
10	375	295	264
12	355	231	230

V. CONCLUSION

Time-domain oversampling was investigated for MIMO OFDM underwater acoustic communications. The oversampling structure enabled multipath diversity in the frequency-domain and better channel estimation in the time-domain, thus achieved better detection performance compared to conventional OFDM systems. A time-domain OLA method was used in combination with the MIMO OOFDM detector to ensure the orthogonality among the subcarriers when ZP instead of CP is used at the transmitter. The proposed MIMO OOFDM detection scheme was applied to the experimental data collected from the SPACE08 undersea trial. Experimental results demonstrated that considerable performance gain was achieved with an oversampling factor of 2, which has the similar computational complexity as a conventional OFDM system.

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