MIMO TDS-OFDM for Underwater Acoustic Communication with Turbo Equalization

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***Abstract*—Time domain synchronous orthogonal frequency division multiplexing (OFDM) is a novel OFDM transmission scheme which utilizes time-domain sequences as the guard intervals and as the training sequences to enhance the overall data efficiency. By adopting turbo equalization and data-aided channel re-estimation, the MIMO TDS-OFDM system is applied to underwater acoustic (UWA) communications and achieves better error performance than zero-padded OFDM with comparable sizes of guard interval and pilots. The proposed MIMO TDS-OFDM scheme achieves high performance and high data efficiency, and keeps low complexity. Its performance is demonstrated by a simulation and an underwater pool experiment.**

***Index Terms*—Underwater acoustic communication, Multiple-input multiple-output (MIMO), Time-domain synchronous orthogonal frequency division multiplexing (TDS-OFDM), Turbo equalization, Decision-aided channel estimation**

I. Introduction

Multicarrier transmission such as orthogonal frequency division multiplexing (OFDM) has been applied to underwater acoustic (UWA) communications [1][2][3], since OFDM systems naturally perform frequency domain equalization with low complexity, and can handle the long delay spread in underwater acoustic environment. In these proposed schemes, zero padding (ZP) OFDM and cyclic-prefix (CP) OFDM are conventionally adopted, which suffer from large overhead. ZP-OFDM and CP-OFDM use padded zeros and cyclic prefix as the guard interval. Some of the subcarriers are also used as pilots for channel estimation. Therefore in these two schemes, guard intervals and the pilot subcarriers are both transmission overhead. Large time variation is another main challenge in combating underwater channels. As a result, accurate synchronization is important [4]. For ZP-OFDM and CP-OFDM, additional time-domain sequences are usually inserted once every several blocks, to ensure precise estimation of frame starts.

Following the work of single-input single-output (SISO) TDS-OFDM in [5], we extend the TDS-OFDM to multiple-input multiple-output (MIMO) in this paper. In the proposed TDS-OFDM UWA communication system with one training sequence to further exploit the capability, the channels are estimated by the training sequence and re-estimated by decision-aided method. Turbo equalization with soft information exchange is adopted to demodulate the symbols.

To evaluate the proposed system, a simulation with MIMO TDS-OFDM and MIMO ZP-OFDM is applied. In the simulation, the proposed scheme shows high performance, high data efficiency and low complexity, compared with the ZP-OFDM counterpart. Besides, a pool experiment result is also presented to address the effectiveness of TDS-OFDM in actual underwater acoustic channels.

This paper is organized as follows. In Section II, the MIMO TDS-OFDM system model for underwater acoustic communication is introduced. In Section III, the receiver algorithm of the scheme is thoroughly demonstrated, including synchronization, channel estimation and equalization. Section IV shows the simulation, and Section V demonstrates results of the pool experiment.

II. Frame and Data Structure

Consider an $N \times M$ MIMO TDS-OFDM underwater acoustic communication system, where N is the number of transmit transducers and M is the number of hydrophones. Bit streams are independently coded, interleaved, modulated and transmitted on different transducers. A typical signaling process is illustrated in Fig. 1.

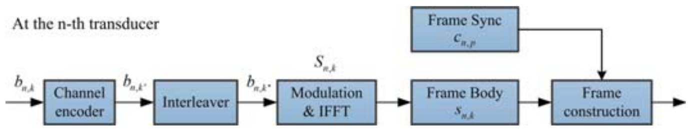

Fig. 1. MIMO TDS-OFDM signaling at the n-th transducer.

For the n-th transmit branch, the information bit, encoded bit, interleaved bit are represented by $b_{n,k}$, $b_{n,k'}$, $b_{n,k''}$, respectively. The modulated frequency and time domain symbols are denoted by $S_{n,k}$ and $s_{n,k}$.

The OFDM length is indicated by K, so the index of subcarriers is $k = 0, 1, ..., K-1$. A $1/2$ convolutional encoder with generator polynomial $[G_1, G_2] = [17, 13]_{oct}$ and a random interleaver ($\mathbf{\Pi}$) are used in the system. In bit to symbol modulation, QAM modulators map every $\log_2^Q$ interleaved bits into complex frequency domain symbols $\mathcal{S} = \{\chi_q\}_{q=0}^{Q-1}$ with constellation size Q.

As shown in Fig. 2, TDS-OFDM uses time-domain training sequences as the guard interval instead of padding zeros or inserting cyclic prefix. TDS-OFDM signals transmit in frames,

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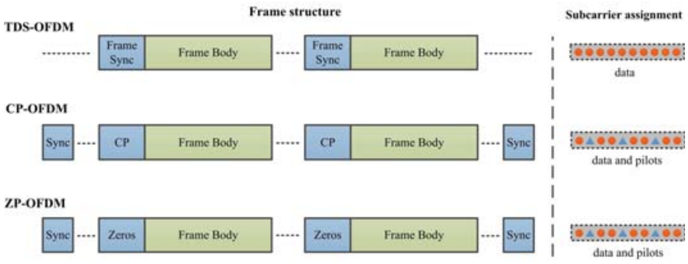

Fig. 2. Comparison of TDS-/CP-/ZP-OFDM.

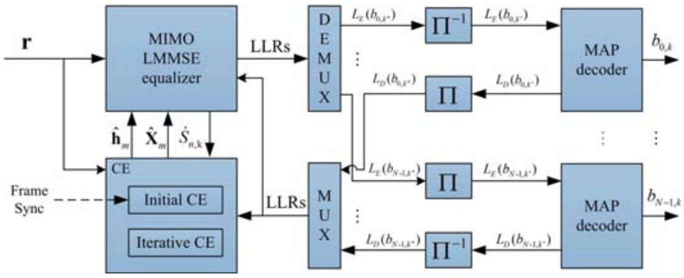

Fig. 3. Turbo equalization work flow of TDS-OFDM

so we name the training sequence as Frame Sync and the OFDM data part as Frame Body. For underwater acoustic ZP- and CP-OFDM systems, synchronization sequences such as linear frequency modulation (LFM) signals are usually necessary. However, TDS-OFDM system doesn't rely on extra sequences to achieve robust synchronization. Instead, the Frame Sync can be used for synchronization. So the overall spectral efficiency of TDS-OFDM is higher in underwater applications. Moreover, Frame Sync is also used for channel estimation, thus the spectral efficiency of TDS-OFDM is even higher. As there is no guard interval between Frame Sync and Frame Body, the inter-symbol-interference should be subtly handled in TDS-OFDM systems. We denote the Frame Sync from the n-th transmit branch by $\{c_{n,p}\}_{p=0}^{P-1}$, with the sequence length P.

A MIMO TDS-OFDM receiver first detects and segments the baseband signal into frames. Then channel estimation and OFDM reconstruction are performed simultaneously. The estimated channel and the reconstructed OFDM symbols are fed into a turbo equalizer, while the channel estimation is adaptively refined by a decision-aided re-estimator. Details of the algorithms are shown in the next section.

III. MIMO TDS-OFDM Receiver Algorithms

A. System Model

Fig. 3 depicts the iterative structure of MIMO TDS-OFDM receiver, where $\mathbf{x}_m^i$ denotes the i-th received and reconstructed Frame Body at the m-th hydrophone, $L_E(b_{n,k''})$ and $L_D(b_{n,k'})$ denote the extrinsic bit log likelihood ratio (LLR) from the equalizer and the decoder, respectively. The receiver adopts a MIMO-OFDM LMMSE equalizer [3] to calculate $L_E(b_{n,k''})$, with the Frame Body, the estimated channel and the *a priori* LLRs as the input. The output LLRs are then de-interleaved and fed into the channel decoder as its *a priori* information. At the same time, the detected symbols are fed into a sparse channel re-estimator to refine the initial channel estimation. The channel decoder works on maximum *a posteriori* (MAP) principle, and outputs the LLRs $L_D(b_{n,k'})$, which are then interleaved and regrouped to provide the *a priori* information for the equalizer in the next iteration of turbo equalization.

B. Synchronization

Correlating the received signal with the Frame Sync of transmitter n, the output of the correlator is

$$R_{n,m,k} = \sum_{p=0}^{P-1} r_{m,p} c_{n,k-p}^{*} \tag{1}$$

where $r_{m,p}$ denotes the received signal at the m-th hydrophone, and $R_{n,m,k}$ is the correlation output of $r_{m,p}$ with the Frame Sync of the n-th transmit unit. A peak detector then continuously searches through the correlation amplitude sequence $\{|R_{n,m,k}|\}$ for the possible synchronization peaks. As the sequence is 4 times over-sampled, several points surrounding the peak would hold high amplitude, a differentiator is used to locate the precise synchronization point,

$$\begin{aligned} d_{n,m,k} = {} & \|R_{n,m,k-1}\| - \|R_{n,m,k+1}\| + \\ & 2(\|R_{n,m,k-2}\| - \|R_{n,m,k+2}\|) \end{aligned} \tag{2}$$

If $\|R_{n,m,\bar{k}}\| \geq T_h$, $d_{n,m,\bar{k}-1} < 0$ and $d_{n,m,\bar{k}} \leq 0$ hold at time instant $\bar{k}$, the synchronizer would declare that the correlation peak is at time instant $k'_{n,m}$, which is given as

$$k'_{n,m} = \begin{cases} \bar{k}-1 & \text{if } \|R_{n,m,\bar{k}}\| > \|R_{n,m,\bar{k}-1}\| \\ \bar{k} & \text{else} \end{cases} \tag{3}$$

$\{k'_{n,m}\}_{m=0}^{M-1}$ composes the set of peaks from M receive branches. If the relative peak positions are stable in consecutive frames, the synchronized state is confirmed.

C. Iterative Decision Aided Channel Estimation

The proposed MIMO TDS-OFDM system has two modes in channel estimation: the initial estimation mode using the Frame Sync, and the decision-aided mode using the detected symbols. A unique stage of Frame Body reconstruction is necessary in TDS-OFDM, after the initial channel estimation, in order to remove the interference of neighboring Frame Sync sequences from the Frame Body.

1) Initial Channel Estimation: TDS-OFDM utilizes the Frame Sync as the training sequence to estimate the channel of current frame. Ignoring the interference from the Frame Body of the last frame, the received Frame Sync at the m-th hydrophone is expressed as

$$r_{m,k} = \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} c_{n,k-l} h_{n,m,l} + w_k \tag{4}$$

where L is the channel length, $h_{n,m,l}$ is the channel impulse response (CIR) from transmitter n to receiver m, and w_k is the time domain additive white Gaussian noise (AWGN). Although $r_{m,k}$ is the superposition of all transmitted signals at receiver m, the cross-correlation between $\{c_{n_1,p}\}_{p=0}^{P-1}$ and $c_{n_2,p}{}_{p=0}^{P-1}$ with $n_1 \neq n_2$ is $1/\sqrt{P}$, when Zadoff-Chu sequence is used as Frame Sync and unit energy is preserved [6]. So time domain convoluting $\{r_{m,k}\}$ with $\{c_{n_1,p}\}_{p=0}^{P-1}$ leads to the roughly estimated channel $\{\hat{h}^{tc}_{n,m,l}\}_{l=0}^{L-1}$,

$$\begin{aligned}\hat{h}^{tc}_{n,m,l} &= \frac{1}{K}\sum_{p=0}^{P-1} c^*(l-p)r(p) \\ &= h_{n,m,l} + \frac{N}{\sqrt{P}} + w_l\end{aligned} \tag{5}$$

As there is a constant amplitude offset in the convoluted sequence $\hat{h}^{tc}_{n,m,l}$, low-amplitude channel taps are no longer reliable. Therefore, an minimum mean square error (MMSE) filter is then used to suppress the low amplitude taps, $\{\tilde{h}^{tc}_{n,m,l}\}_{l=0}^{L-1} = \{\hat{h}^{tc}_{n,m,l}\}_{l=0}^{L-1} * \{h_{mmse}\}$. $\{\tilde{h}^{tc}_{n,m,l}\}_{l=0}^{L-1}$ is the initial channel estimation that the system used for Frame Body reconstruction and the initial equalization.

2) Frame Body Reconstruction: As shown in Fig. 4, multipath channel causes ISI from Frame Sync into Frame Body and ISI from Frame Body into the next frame's Frame Sync. Therefore, Frame Body reconstruction is a crucial process for robust reception. Assume the receiver is to decode the i-th frame, the ISI from the i-th Frame Sync and from the $i+1$-th Frame Sync are reconstructed using the Frame Sync and the estimated channel.

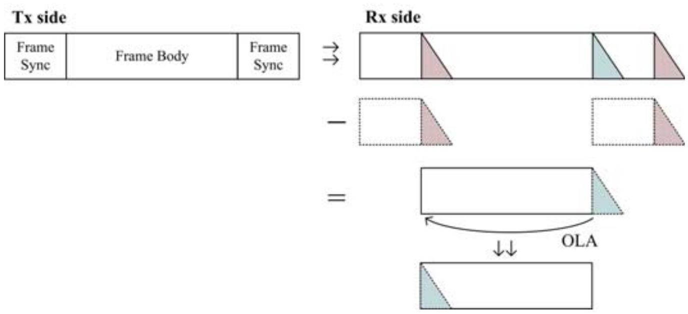

Fig. 4. Frame reconstruction of TDS-OFDM

Denote the initial channel estimation of the i-th frame by $\{\tilde{h}^i_{n,m,l}\}_{l=0}^{L-1}$, and the corresponding ISI-free received sequence of the i-th Frame by $\{y^i_{n,m,k}\}_{k=0}^{P+L-1}$, which is instantly estimated as the convolution of $\{\tilde{h}^i_{n,m,l}\}_{l=0}^{L-1}$ and $\{c_{n,p}\}_{p=0}^{P-1}$, denoted by $\{\hat{y}^i_{n,m,k}\}_{k=0}^{P+L-1}$. Extend the expression in (4), $\{r^i_{m,k}\}_{k=0}^{P+K+L-1}$ is used to represent the i-th received frame, including the Frame Sync and the Frame Body. Subtracting the interference from the received sequence, we get

$$\hat{x}^i_{m,k} = \begin{cases} r^i_{m,k+P} - \sum_{m=0}^{M-1}\hat{y}^i_{n,m,k+P} & (0 \leq k < L-1) \\ r^i_{m,k+P} & (L-1 \leq k < K) \\ r^i_{m,k+P} - \sum_{m=0}^{M-1}\hat{y}^{i+1}_{n,m,k-K} & (K \leq k < P+K) \end{cases}$$

as in ZP-OFDM, overlap-adding technique is then applied to reconstruct the circular convoluted received signal

$$\tilde{x}^i_{m,k} = \begin{cases} \hat{x}^i_{m,k} + \hat{x}^i_{m,k+K} & 0 \leq k < L-1 \\ \hat{x}^i_{m,k} & L-1 \leq k < K \end{cases}$$

A K-point FFT ($\mathbf{F}_K$) converts $\{\tilde{x}^i_{m,k}\}$ to frequency domain,

$$\tilde{\mathbf{X}}^i_m = \mathbf{F}_K\tilde{\mathbf{x}}^i_m \tag{6}$$

as the input of the equalizer.

3) Decision-aided Channel Re-estimation: After the frequency domain symbols are equalized, the soft symbol information from the equalizer provides more accurate estimation of the frequency domain symbol on each subcarrier. With the selected subcarrier set $\mathcal{S}^{(t)}$, we only need part of the equalized symbols as the observed vector $\tilde{\mathbf{X}}_{\mathcal{S}^{(t)}}$ for channel re-estimation, which is the a vector composed of the symbols indexed by $\mathcal{S}^{(t)}$, and the observation matrix is generated by the corresponding row selection on original observation matrix $\mathcal{D}$.

As there may be not enough pilots for conventional channel estimation, a sparse channel based re-estimation is applied. The decision aided principle is referred to [7]. In our approach, the sparse channel estimation performs in two stages. In the first stage, the coarse path delay taps are obtained from the initial sparse estimation using Frame Sync. Then a second-step priori aided compressive sampling matching pursuit (PA-CoSaMP) estimator is applied, which uses the path delay as the a priori information.

The re-estimation algorithm is list in Algorithm. 1. It's noted that the initial channel sparsity $\mathbb{S}_0$ is the support set of the pruned coarse channel estimation by Frame Sync, and the expected sparsity level is chosen to be $\mathbb{S} = \mathbb{S}_0 + 2N$, so that the initially falsely excluded channel taps due to noise effect in Frame Sync may be recovered in the second stage PA-CoSaMP estimation.

D. Turbo Equalization

The MIMO-OFDM detector adopts frequency domain linear minimum mean square (LMMSE) turbo equalization. In the t-th turbo iteration, $\hat{\mathbf{h}}^{(t)}_m$ is used. The initial estimation by Frame Sync is $\hat{\mathbf{h}}^{tc}_m = \hat{\mathbf{h}}^{(0)}_m$ for the 0-th iteration. The turbo equalization follows a similar procedure as in [3]. The major difference is that the channel is updated in each iteration by re-estimation, and that the OFDM reconstruction is performed with the updated with the updated channel. The algorithm of the iterative channel estimation and turbo equalization for MIMO-TDS-OFDM is organized in Fig. 3 and Algorithm 2.

E. Computational complexity analysis

The computational complexity of the proposed TDS-OFDM is listed in Table I in terms of the number of multiplications for one frame. Compared with the counterpart of ZP-OFDM which also adopts turbo equalization and channel re-estimation, TDS-OFDM endures almost no complexity enhancement (assume that TDS-OFDM and ZP-OFDM has similar number of pilots P).

Algorithm 1 Decision-aided sparse channel re-estimation.

Require:

1) Noisy observation vector $\mathbf{v} \triangleq \tilde{\mathbf{X}}_{\mathcal{S}^{(t)}}$
2) Observation matrix $\mathbf{\Phi} \triangleq \mathcal{D}_{\mathcal{S}^{(t)}}$
3) Initial path delay set T_0 with its sparsity $\mathbb{S}_0$, expected channel sparsity $\mathbb{S}$.

Ensure: The $\mathbb{S}$-sparse CIR estimation $\hat{\mathbf{h}}_m^{(t)}$.

$\mathbf{r} \leftarrow \mathbf{v} - \mathbf{\Phi}_{\mathrm{T}^0}\mathbf{\Phi}_{\mathrm{T}^0}^{\dagger}\mathbf{v}$;
$k \leftarrow 0$;
while $k < \mathbb{S} - \mathbb{S}_0$ **do**
  $k \leftarrow k + 1$;
  $\mathbf{p} \leftarrow \mathbf{\Phi}^H \mathbf{r}$;
  $\tilde{\mathrm{T}}_k \leftarrow \mathrm{T}_{k-1} \cup \sup(\mathbf{p}_{\mathbb{S}-\mathbb{S}_0})$;
  $\mathbf{q} \leftarrow \mathbf{\Phi}_{\tilde{\mathrm{T}}_k}^{\dagger}\mathbf{v}$;
  $\mathrm{T}_k \leftarrow \sup(\mathbf{q}_{\mathbb{S}})$;
  $\mathbf{r} \leftarrow \mathbf{v} - \mathbf{\Phi}_{\mathrm{T}_k}\mathbf{\Phi}_{\mathrm{T}_k}^{\dagger}\mathbf{v}$;
end while
Actual path delay acquisition: $\mathrm{T} \leftarrow \mathrm{T}_k$;
Final least squares (LS) estimation:

$$\hat{\mathbf{h}}_m^{(t)}|_{\mathrm{T}} = \mathbf{\Phi}_{\mathrm{T}}^{\dagger}\mathbf{v} = \left(\mathbf{\Phi}_{\mathrm{T}}^H \mathbf{\Phi}_{\mathrm{T}}\right)^{-1} \mathbf{\Phi}_{\mathrm{T}}^H \mathbf{v}. \quad (7)$$

Algorithm 2 Iterative turbo equalization with channel re-estimation.

Require:

1) Number of turbo iterations M_{turbo}
2) Number of frames M_f
3) Frame Sync sequences $\{c_{n,p}\}_{p=0}^{P-1}$

$m_f \leftarrow 1$
while $m_f \leq M_f$ **do**
  $t \leftarrow 0$
  while $t \leq M_{turbo}$ **do**
    if $t = 0$ **then**
      $\hat{\mathbf{h}}_m^{(0)}(0 \leq m \leq M-1) \leftarrow$ TD channel estimation based on received
      Frame Sync and $\{c_{n,p}\}_{p=0}^{P-1}$;
      $L_D^{(0)}(b_{n,k''}) \leftarrow 0$;
      $\tilde{\mathbf{X}}^k \leftarrow$ ISI cancelation, OLA, and DFT.
    else
      $\mathcal{S}^{(t)} \leftarrow$ Selection based on $L_D^{(t)}(b_{n,k''})$;
      $\hat{\mathbf{h}}_m^{(0)}(0 \leq m \leq M-1) \leftarrow$ FD channel estimation based on $\{\dot{S}_{n,k}^{(t)} | k \in \mathcal{S}^{(t)}\}$.
    end if
    $L_E^{(t)}(b_{n,k''}) \leftarrow$ Frequency equalization based on $\tilde{\mathbf{X}}^k$ and $L_D^{(t)}(b_{n,k''})$;
    $L_E^{(t)}(b_{n,k'}) \leftarrow \mathbf{\Pi}(L_E^{(t)}(b_{n,k''}))$;
    $L_D^{(t)}(b_{n,k}), L_D^{(t)}(b_{n,k'}) \leftarrow$ MAP decoding $L_E^{(t)}(b_{n,k'})$;
    $L_D^{(t+1)}(b_{n,k''}) \leftarrow \mathbf{\Pi}^{-1}(L_D^{(t)}(b_{n,k'}))$;
    $t \leftarrow t + 1$.
  end while
  output bits $\leftarrow$ Hard decision on $L_D^{(t)}(b_{n,k})$.
end while

TABLE I
COMPLEX MULTIPLICATIONS OF TDS- AND ZP-OFDM PER FRAME

Step	TDS-OFDM	ZP-OFDM
$Synchronization$	$4N_r(K+P)$	$4N_r(K+P)$
$Initial-CE$	0	$N_tN_rP^2$
$Recontruction$	$2N_tN_r(P+L)$	$N_tN_r(P+L)$
$Equalizatoin$	$(3N_tN_r^3+2^Q)K$	$(3N_tN_r^3+2^Q)(K-P)$
$Re-CE$	$N_tN_rK^2$	$N_tN_rK^2$

IV. SIMULATION

The proposed TDS-OFDM scheme is based on a turbo receiver with iterative equalization and channel re-estimation. The performance of the scheme is firstly tested by a simulation system using channels from an underwater experiment. The effectiveness is addressed by comparison with ZP-OFDM receiver, in comparable spectral efficiency. The results on QPSK reception show that the proposed is superior to the ZP-OFDM counterpart.

In the simulation, there are 2 transmitters and 12 receivers for both TDS-OFDM and ZP-OFDM with QPSK modulation. The length the OFDM data part is $K = 1024$, $K = 2048$ and $K = 4096$. For TDS-OFDM, the length of Frame Sync is 240; for ZP-OFDM, the length of padded zeros is 120, and the number of pilots is chosen to be 170. As a result, the overall overhead of ZP-OFDM is slightly higher than that of TDS-OFDM: the pilot overhead of TDS-OFDM is 19% when $K = 1024$, 10.5% when $K = 2048$ and 5.5% when $K = 4096$, and that of ZP-OFDM is 22.6% when $K = 1024$, 11.8% when $K = 2048$ and 6% when $K = 4096$.

ZP-OFDM scheme is implemented for performance comparison. As there is no time-domain training sequences in ZP-OFDM, frequency domain pilots are pre-determined and used for initial channel estimation. Besides, turbo equalization and iterative data-aided channel re-estimation is also applied in the ZP-OFDM receiver.

The BER performance with different OFDM block length is listed in Table II. In the simulation when $K = 1024$, the BER of TDS-OFDM is below 10^{-3}, the BER of ZP-OFDM is slightly above 10^{-2}. When $K = 2048$, we used channels which were in better condition than that when $K = 1024$, so both TDS-OFDM and ZP-OFDM achieve BER lower than 10^{-4}, with TDS-OFDM slightly better than ZP-OFDM. When $K = 4096$, the BER is below 10^{-3}. For both TDS-OFDM and ZP-OFDM, the channel re-estimation and turbo equalization improves the bit error performance, which is obviously seen from the bit error degradation from iteration 1 to iteration 3. Although the BER of good packets is better when the OFDM data length gets larger, the bit error performance of bad packets gets much worse. The BER distribution of frames is shown in Fig. 5.

V. VERIFICATION BY POOL EXPERIMENT

In order to further evaluate the effectiveness of TDS-OFDM in underwater acoustic communication by means except for simulation mentioned in the last section, the result of a recent applied pool experiment is presented. In the experiment, an

TABLE II
BER PERFORMANCE OF TDS-OFDM AND ZP-OFDM

K	CE iteration	TDS-OFDM	ZP-OFDM
1024	1	0.03922526	0.174291992
	3	0.000427246	0.029284668
2048	1	0.090189616	0.191805753
	3	0.010843913	0.057228782
4096	1	0.136088053	0.220681895
	3	0.040690104	0.064113451

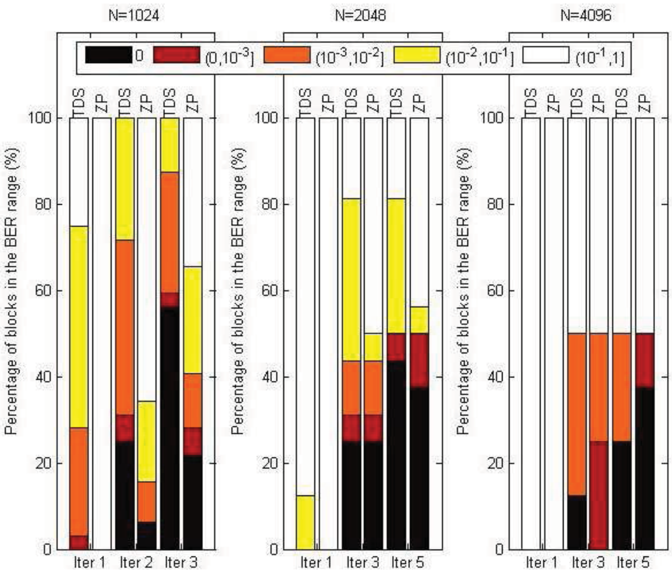

Fig. 5. Percentage of frames in BER ranges.

LDPC-coded MIMO TDS-OFDM system of bandwidth 6 kHz was working on a 2-transmitter 4-receiver setup.

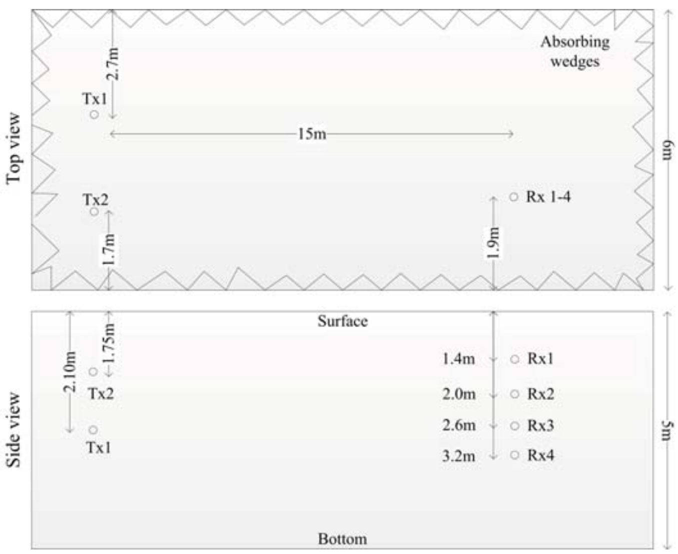

Fig. 6. Pool experiment setup.

The experiment setup is illustrated in Fig. 6. Tx 1 and Tx 2 are 1.6 m apart horizontally and 0.35 m apart vertically. Both transducers were let down with a rope at about 2 m below the surface. At the receiver side, 4 receiver units are vertically placed from 1.4 m to 3.2 m deep into water. With the pool walls and pool ends exempted from reflection by sound-absorbing wedges, the experiment assimilated the environment in short distance shallow water communication with multipaths from only the surface and the bottom. As depicted in Fig. 7, the channels have strong multipaths. The main cause of the multipaths is the reflections at the surface and at the pool bottom.

The experiment also shows excellent performance with TDS-OFDM, using QPSK modulation. The error bit number reaches zero for both transducers. The OFDM data block length was either 1024 or 2048, which almost does not alter the performance in the experiment. The average number of LDPC decoding iterations is 3.2 for Tx 1 and about 5.0 for Tx 2. We believe that with turbo equalization and channel re-estimation, MIMO TDS-OFDM is a promising scheme to achieve good performance with higher order modulation and in more sophisticated channels.

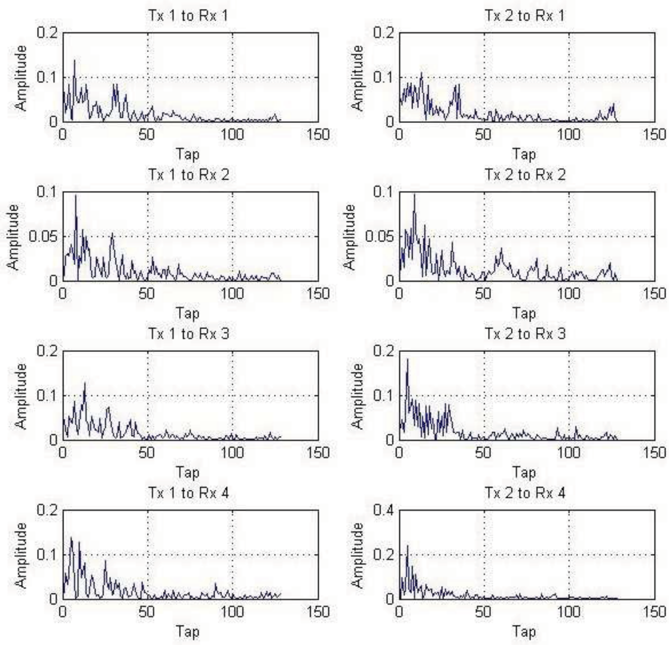

Fig. 7. Estimated channels in pool experiment.

VI. CONCLUSION

In this paper, a MIMO TDS-OFDM scheme based on turbo equalization and iterative channel re-estimation is proposed for underwater acoustic communication. The proposed scheme makes use of time domain training sequences for synchronization and channel estimation, therefore the spectral efficiency is greatly enhanced. Simulation and experimental results on QPSK transmission show that the proposed scheme achieves considerable bit error performance.

Several technical issues should be carefully considered in TDS-OFDM for further improvement, including the design of the Frame Sync sequences and reliable estimation of the time varying channel. Besides, STBC may be needed to allow more trustful pilot subcarrier selection in channel re-estimation.

VII. Acknowledgement

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References

[1] M. Stojanovic, "Low complexity OFDM detector for underwater acoustic channels", *MTS/IEEE OCEANS 2012*, Boston, MA, pp. 1-6, Sept. 2006.

[2] B. Li, S. Zhou, M. Stojanovic, L. Freitag, "Multicarrier communication over underwater acoustic channels with nonuniform Doppler shifts," *IEEE Journal of Oceanic Engineering*, vol. 33, no. 2, pp. 198-209. April 2008.

[3] J. Tao, and Y. R. Zheng, "Turbo Detection for MIMOCOFDM Underwater Acoustic Communications," *International journal of wireless information networks*, vol. 20, no. 1, pp. 27-38, 2013.

[4] M. Stojanovic, "Underwater acoustic communication channels: Propagation models and statistical characterization," *IEEE Communication Magazine*, Boston, MA, vol. 47, no. 1, pp. 84-89, Jan. 2009.

[5] J. Hao, Y. R. Zheng, J. Wang, and J. Song, "Dual PN Padding TDS-OFDM for Underwater Acoustic Communication," *MTS/IEEE OCEANS 2012*, Hampton Roads, VA, pp. 1-6, Oct. 23-27, 2012.

[6] D. Chu, "Polyphase codes wih good periodic correlation properties (Corresp.)," *IEEE Transactions on Information Theory*, vol. 18, no. 4, pp. 531-532, Jul. 1972.

[7] M. Mnster, and L. Hanzo, "Parallel-interference-cancellation-assisted decision-directed channel estimation for OFDM systems using multiple transmit antennas," *IEEE Transactions on Wireless Communications*, vol. 4, no. 5, Sept. 2005.