The Union of Time Reversal and Turbo Equalization On Underwater Acoustic Communication

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***Abstract*—A receiver scheme combing time reversal processing with turbo equalization (TE) is presented for underwater acoustic communication. Time reversal processing is a means of refocusing signal that has been spread in time domain, which significantly mitigates the inter-symbol interference caused by multipath underwater environments. Turbo equalization (TE) based on channel estimation performs well against residual inter-symbol interferences after the time reversal processing. This scheme has a low complexity with multiple elements. Due to the time reversal processing, the number of equalizer taps in this scheme is 15 percents of that in direct adaptive turbo equalizer without performance degradation.**

I. Introduction

The underwater acoustic channel has characteristics of extended multipath spread and rapidly changing . Phase coherent communication is limited by the multipath fading and the intersymbol interference (ISI) deteriorated by acoustic propagation in a temporal variant wave guide. High rate phase coherent underwater communications have traditionally relied on adaptive equalization methods and spatial diversity to overcome the intersymbol interference (ISI) caused by multipath propagation.

Multichannel decision feedback equalizer (McDFE) plus carrier-phase tracking is used to remove ISI of the channel and exploit the spatial diversity, which is proposed by Stojanovic in early 1990s [1]. Since the equalizer and the decoder are separated in the conventional McDFE, it is inevitable that the performance of the receiver is comprimised. As discussed in [4], turbo equalizers being applied with multichannel combining due to the harsh channel conditions in underwater acoustic channels. The soft information is been exchanged between the equalizer and decoder by the interleaver and de-interleaver in iterations. There are two different suboptimal turbo equalizer structures applied in underwater acoustic communications: turbo equalizer based on channel estimation (CE-TE) [5] and turbo equalizer based on direct adaptive equalizer (DA-TE) [6]. CE-TE is more complicated with the increased equalizer complexity caused by matrix inverse during iterations,especially for the long time spread of channel and multi-array processing. Because of lower complexity, DA-TE is more widely used in phase coherent communications. DA-TE updates the coefficients of the equalizer based on the step size parameter for convergence. For practical application, the large step-size to accelerate the convergence could make the performance of the multi-channel direct adaptive equalizer(McDA-TE) worse than CE-TE.

Time reversal combining (TRC) is the practical application of time reversal mirror. TRC can make use of the spatial diversity and filter the each of the multichannel by the information of each channel [7]. Moreover, TRC makes full use of the energy of the received signal and compress the time spread caused by multipath. If the number of the channels is limited, there is still residual ISI after the TRC operation [8]. These kinds of feature limit the application of TRC .

Motivated by the respective advantages and limitations of the methods of turbo equalizer and TRC, combining CE-TE with TRC can be a very potential technique. TRC compresses the long time spread of the received signals and combines multichannel signal. Then the practical CE-TE eliminates the residual ISI after the TRC operation. This kind of scheme can be more efficient and robust.

The rest of this paper is organized as follows. In Section II, time reversal is briefly described and the processing of the received signal is described in detail. In Section III, channel estimation based turbo equalizations is briefly reviewed. In Section IV, the conditions of experiments will be introduced. Moreover, the comparison of the different turbo equalizer and the result of the experiments will be stated. In the last Section, some conclusions are discussed.

II. Time Reversal Combining

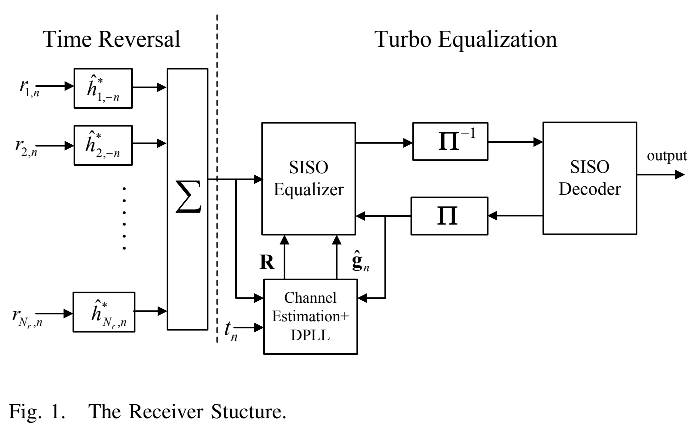

Fig. 1. The Receiver Stucture.

The scheme of Receiver is plotted in Fig.1. The channel with intersymbol interference can be equivalent to finite impulse response (FIR) filter. Assuming perfect synchronization, demodulation, symbol spaced sampling and Doppler frequency shift compensation, the received signal in kth receiver of array with N_r elements can be expressed as:

$$r_{k,n} = (x_n \otimes h_{k,n}) + w_{k,n}, k = 1 \cdots N_r \quad (1)$$

where n are time indices, respectively. $h_{k,n}$ is the nth gain in kth receiver channel impulse response, x_n is the QPSK symbol generated by transmitter, $w_{k,n}$ is additive noise in kth receiver element. After TRC, the sequence z_n as (2) can be obtained, which can be viewed as the transmitted symbols passed through a simplified channel with small time spread.

$$z_n = \sum_{k=1}^{N_r} r_{k,n} \otimes \hat{h}^*_{k,-n} \quad (2)$$

The estimation of kth receiver channel impulse response is denoted as $\hat{h}_{k,n}$, so the time reversal combine the N_r receivers to produce the signal z_n can be expressed as

$$\begin{aligned} z_n &= \sum_{k=1}^{N_r} (r_{k,n} \otimes \hat{h}^*_{k,-n}) \\ &= \sum_{k=1}^{N_r} ((h_{k,n} \otimes \hat{h}^*_{k,-n} \otimes x_n) + w_{k,n} \otimes \hat{h}^*_{k,-n}) \\ &= \sum_{k=1}^{N_r} ((h_{k,n} \otimes \hat{h}^*_{k,-n}) \otimes x_n) + \sum_{k=1}^{N_r} w_{k,n} \otimes \hat{h}^*_{k,-n} \\ &= (q_n \otimes x_n) + \zeta_n \end{aligned} \quad (3)$$

where q_n is the autocorrelation of the channel impulse response summed over N_r channels, ζ_n is the sum of the filtered noise of N_r channels.

The key problem of the time reversal operation is how to get the channel impulse response. In practical application, passive time reversal is used for convenience. As the auto-correlation of the the linear frequency modulated signal LFM, which is often used for synchronization, can be approximated as unit impulse response, it also can be used to get the impulse response of each channel. For the probe signal p_n transmitted from the transimitter, the received probe signal at the kth receiver element can be expressed as $d_{n,k}$:

$$d_{n,k} = p_n \otimes h_{k,n} + w_{k,n}$$

The known probe signal p_n can be used for correlating with the received probe siganl d_n, so $\hat{h}_{k,n}$ can be obtained by

$$\begin{aligned} \hat{h}_{k,n} &= p^*_{-n} \otimes d_{n,k} \\ &= (p_n \otimes p^*_{-n}) \otimes h_{k,n} + p^*_{-n} \otimes w_{k,n} \\ &= \lambda_n \otimes h_{k,n} + p^*_{-n} \otimes w_{k,n} \end{aligned} \quad (4)$$

where $\lambda_n = p_n \otimes p^*_{-n}$ is the autocorrelation of the LFM signal which is not a perfect unit impulse response in practial application, but as its very little time spread it does not deteriorate the performance of the TRC.

In practical experiments, each frame contains a LFM signal used for synchronization and time reversal. Since the TRC can not get the single channel with the ideal unit impulse response, which is limited by the low gain of the spatial diversity and environment noise. And actually the signal of the single channel still suffer residual ISI. It is important to depend on a equalizer to eliminate the residual ISI. Since the time reversal operation reduced the ISI and exploit the gain of spatial diversity, the structure of the receiver and especially the length of the equalizer is simplified.

TRC is an application of the principle of sound field reciprocity. The processing of time reversal can be viewed as matched filters. In classic turbo equalizer, matched filters are realized by feedforward filters .

III. TURBO EQUALIZATION BASED ON CHANNEL ESTIMATION

In this section, CE-TE techniques will be briefly reviewed . The CE-TE is in the same principle with that of [6]. The truncation for chanel impulse in this scheme is took into consideration. As the time reversal processing in Section II, (3) can be developed into matrix form:

$$\mathbf{z}_n = \mathbf{Q}_n \mathbf{x}_n + \boldsymbol{\zeta}_n \quad (5)$$

where

$$\mathbf{x}_n = [x_{n-K_f-M_f}, \cdots, x_{n+K_p+M_p}]^T$$

$$\boldsymbol{\zeta}_n = [\zeta_{n-K_f}, \cdots, \zeta_{n+K_p}]^T$$

$$\mathbf{z}_n = [z_{n-K_f}, \cdots, z_{n+K_p}]^T$$

$$\mathbf{q}_n = [q_{n-M_f}, \cdots, q_{n+M_p}]^T$$

$$\mathbf{Q}_n = \begin{bmatrix} \mathbf{q}_{n-K_f} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \ddots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{q}_{n+K_p} \end{bmatrix}$$

The observation window of $\mathbf{z}_n$ contains $K_f + K_p + 1$ received symbols. The channel length of $\mathbf{q}_n$ is assumed to be at most $M = M_p + M_f + 1$,where M_f is the length of the precursor and M_p is that of the postcursor response.

In practical operation, the length of $\mathbf{q}_n$ is very long. As plotted in Fig.4 in Section IV, the taps of the main power are concentrated in a limited length. So the $\mathbf{q}_n$ for main power is truncated to implement the channel estimated based turbo equalizer to make the complexity low. $\mathbf{q}_n$ are truncated into $\mathbf{g}_n$ as:

$$\mathbf{g}_n = [q_{n,-L_f}, \cdots, q_{n,L_p}](L_p + L_f << M_p + M_f)$$

As the operation for truncating the $\mathbf{q}_n$ increases the noise power by $\sigma^2_{trunc} = (\mathbf{q}_n\mathbf{q}_n^H - \mathbf{g}_n\mathbf{g}_n^H)\sigma^2_{xx}$, . So noise correlation matrix can be expressed as:

$$\mathbf{R} = \sigma^2_{trunc}\mathbf{I}_M + E(\boldsymbol{\zeta}_n\boldsymbol{\zeta}_n^H), M = M_p + M_f + 1$$

the matrix $\mathbf{I}_i$ is the $i \times i$ identity matrix.

Without perfect knowledge of the channel matrix $\mathbf{g_n}$, the inperfect estimated channel impluse response $\hat{\mathbf{g}}_n$, channel estimation error can be stated as $\mathbf{e}_n = \mathbf{g}_n - \hat{\mathbf{g}}_n$. The MMSE estimate of the symbol $\hat{x}_n$ is given by:

$$\begin{aligned} \hat{x}_n &= ((\hat{\mathbf{g}}_n \boldsymbol{\Sigma}_n \hat{\mathbf{g}}_n^H + \mathbf{R} + E[\mathbf{e_n}\mathbf{e}_n^H])^{-1}\mathbf{s}_n)^H \mathbf{u}_n \\ \mathbf{u}_n &= \mathbf{y}_n - \hat{\mathbf{g}}_n \bar{\mathbf{x}}_n \\ \mathbf{s}_n &= \hat{\mathbf{g}}_n [\mathbf{0}_{1\times(K_f+M_f)}, 1, \mathbf{0}_{1\times(K_p+M_p)}]^T \end{aligned} \quad (6)$$

where

$$\bar{\mathbf{x}}_n = [\bar{x}_{n-K_f-M_f}, \cdots, \bar{x}_{n-1}, 0, \bar{x}_{n+1}, \cdots, \bar{x}_{n+K_p+M_p}]^T$$

$$\boldsymbol{\Sigma}_n = diag([v_{n-K_f-M_f}, \cdots, v_{n-1}, 1, v_{n+1}, \cdots, v_{n+K_p+M_p}])$$

$\bar{x}_n$ is the prior mean and variance of the symbol x_n. The SISO linear equalizer convert the soft information from the SISO decoder into v_n and $\bar{x}_n$:

$$\begin{aligned} \bar{x}_n &= \mathrm{E}_L\{x_n\} = \sum_{s \in \mathcal{S}} s \cdot \Pr_L\{x_n = s\} \\ v_n &= \mathrm{Cov}_L\{x_n, x_n\} \\ &= \left(\sum_{s \in \mathcal{S}} s s^* \cdot \Pr_L\{x_n = s\} \right) - \bar{x}_n \bar{x}_n^* \end{aligned} \tag{7}$$

where $\mathcal{S}$ denotes the map of the signal constellation and can be computed by the soft information with the constellation map. The linear equalizer subtracted the ISI mean computed by the from received signal and the residual signal is equalized by the linear filter. In [3], it is introduced in detail that the principle of generation of the coefficients of the linear equalizers by channel impulse response and the prior variances. $\mathbf{R} + E[\mathbf{e_n}\mathbf{e_n^H}]$ can be estimated by

$$\mathbf{R} + E[\mathbf{e_n}\mathbf{e_n^H}] \approx \frac{1}{M_t} \sum_{n=1}^{M_t} (\mathbf{z}_n - \hat{\mathbf{g}}_n \mathbf{x}_n)(\mathbf{z}_n - \hat{\mathbf{g}}_n \mathbf{x}_n)^H \tag{8}$$

from the channel estimator in training modes, M_t is the window length for mean value. When channel estimator is in detection modes, $\mathbf{x}_n$ is changed into $\bar{\mathbf{x}}_n$. The same algorithm as [3] is used to obtain the extrinsc LLR for all bits.

IV. Experimental Results

A. The environment of the experiments

The experiments are conducted at QianDao Lake in Zhejiang Province, China. As shown in Fig. 2, the fixed 6-elements array receiver and a moving transmitter to conduct the communication experiments. The receiver array is spanning depth from 14.8 m to 18.4 m. The Transmitter node drifts slowly with the boats, the experiments of communications at the distance of from 1.5km to 3km are conducted. Fig. 2 also shows the sound speed profile (SSP) at the receiver array. And the parameter setup of the experiments is shown in Table I. As shown in Fig. 3, 14 frames each epoch are transmitted, the chirp signal of LFM is at the head of the each frame for synchronization. There is protect interval with a length of 32ms between the LFM signal and the QPSK signals.The frequency range of the LFM signal is 5kHz-11kHz which can cover the signal bandwidth. This LFM signal is used to obtain the channel impulse response for time reversal operation. The sample rates of the receiver 64k samples/second.

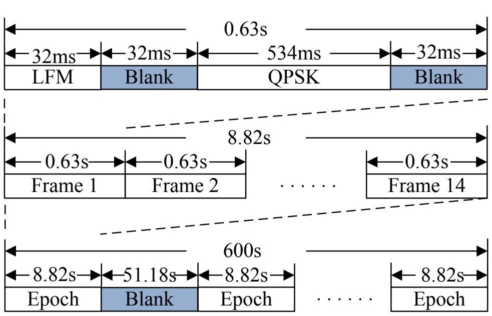

Fig. 3. The Frame Stucture.

The channel impulse response of experiments is shown in Fig.4 , only the first channel response is been plotto investigate time varying channel in a epoch. The channel response varies slowly in 8 seconds, but the severe intersymbol-interference caused by multipath challenges the design of the equalizer. Fig. 4a shows that the time spread caused by multipath can last 10ms or even longer. The 10ms time spread challenges the stucture and the convergence of the linear equalizer. The choice for step size in McDA-TE is a trade-off between the convergence of in training period and performance of the equalizer.

TABLE I. The Experiment Parameters setup

Parameter	Description
Channel Code	Turbo Code(1/2 Rate)
Interleaver	pseduo random interleaver
Modulation Mode	QPSK(Gray Mapping)
Carrier Frequency	8kHz
Symbol Rate	4k Symbols per second
Frame Length(Data+Train symbols)	2136(1936+200)

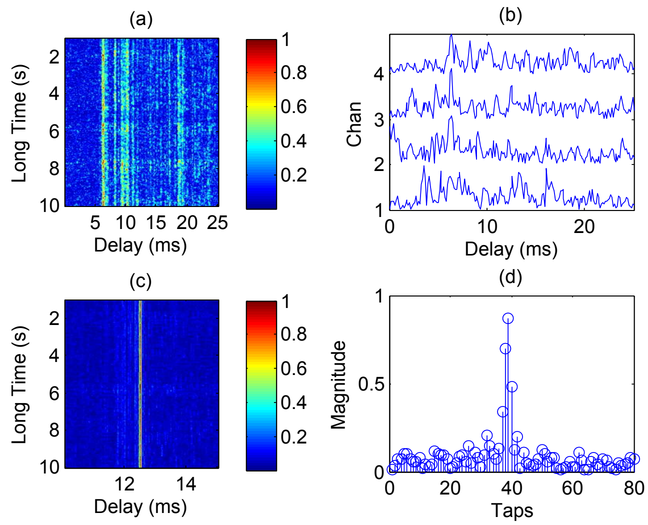

Fig. 4. The channel impulse response (CIR) estimated at the receivers. (a) is for the variation of CIR with the distance of 1.5km to 3km for a single array. (b) is the CIR of comparsion of 4 channels. (c) is for the variation of CIR with the time reversal channel combining 1-4 channel ,and (d) is the main part of the CIR after TRC.

The impulse response of the time reversal single channel, which is obtained by combining channel 1-4 , is also shown in Fig. 4a. Comparing these two figures, the impulse response for the single channel is more stable and more easily estimated. The time spread of the ISI is only about 1ms, the main signal power in a small interval about 1ms. It can be concluded that the TRC can compress the long time spread ISI obviously and make full use of received power of each channel. With this property, TRC could simplify the structure of the equalizer.

B. Experiment Analysis

To process the real data received in experiments, the parameters of two scheme of equalizers are stated as Table II. The number of the taps of equalizers and channel estimators are been $1/2$ fractionally spaced. The number of the taps of channel impulse used for time reveral is 200, which is also fractionally spaced. For 4 channels being processed, the number of equalizer taps and that of channel estimator taps

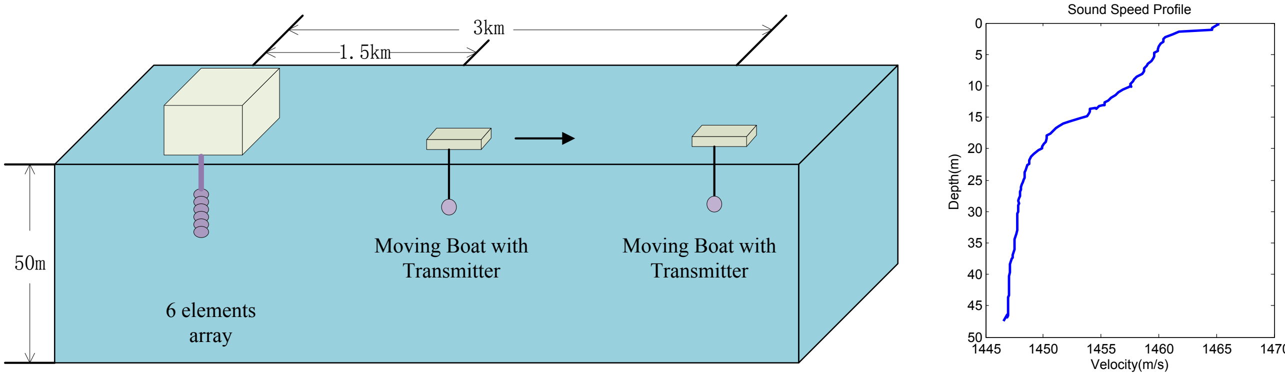

Fig. 2. The environment of the experiment and sound speed profile at the receiver.

TABLE II. THE EQUALIZER PARAMETERS SETUP

Parameter	McDA-TE	TR-CE-TE
Equalizer Order	Feedforward: $30 \times 2 \times N_r$ Feedback: 20	Linear Equalizer:10×2 Channel Estimator:8×2
Number of Channels	4-6	4-6
Number of Iterations	Turbo Equalizer:5 Turbo Decoder:2	Turbo Equalizer:5 Turbo Decoder:2

in TR-CE-TE are only 10×2 and 8×2, while equalizer taps in McDA-TE is about $30 \times 2 \times 4 + 20$. It is easy to figure out that the number of equalizer taps in this scheme is about 15 percents of that in McDA-TE without performance degradation.

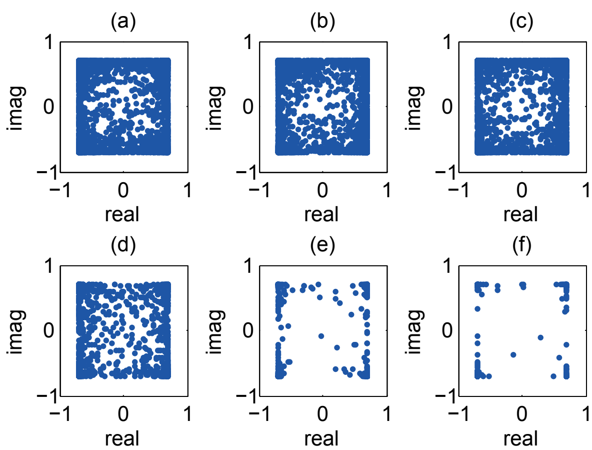

Fig. 5. The Constellation recoverd from the soft information output of SISO decoder with signal of 4 channels used for equalization. (a), (b), (c) are the constellation result of TR-CE-TE at 1st,2nd,5th iteation,(d), (e), (f) are the constellation result of McDA-TE at 1st,2nd,5th iteation.

In Fig.5, the convegence behavior is been plotted. As the turbo code is powerful enough to provide soft information, turbo equalization could acheive error free even the output constellation of the turbo equalizer is hard to distinguish. So using the LLR output of TR-CE-TE is better to investigate the performance comparison. For 4 channels being processed, TR-CE-TE utilizes more performance gain with iterations than McDA-TE.

In Fig.6, the BER performance of the two scheme are shown for in these experiments with different number of channels to be used for equalizer. Data in epoch 1-3,epoch 4-7,epoch 8-10 and epoch 11-14 are came from the experiments conducted at the distance of 1.5km, 2km, 2.5km, 3km. In Fig.6, TR-CE-TE can decode successfully with 5 iterations in 6 channels(channel 1-6) combined in all epochs, McDA-TE could only successfully decode in a part of epochs. When the number of the channels decrease to 5(channel 1-5), TR-CE-TE can decode successfully in most epochs. While the McDA-TE can not decode in almost all epochs successfully. Moreover, with 4 channels processed, TR-CE-TE can be successful only in part of the epochs while McDA-TE fails in all epochs. The lower performance of McDA-TE can be viewed as the result of the suboptimal choice for step size.

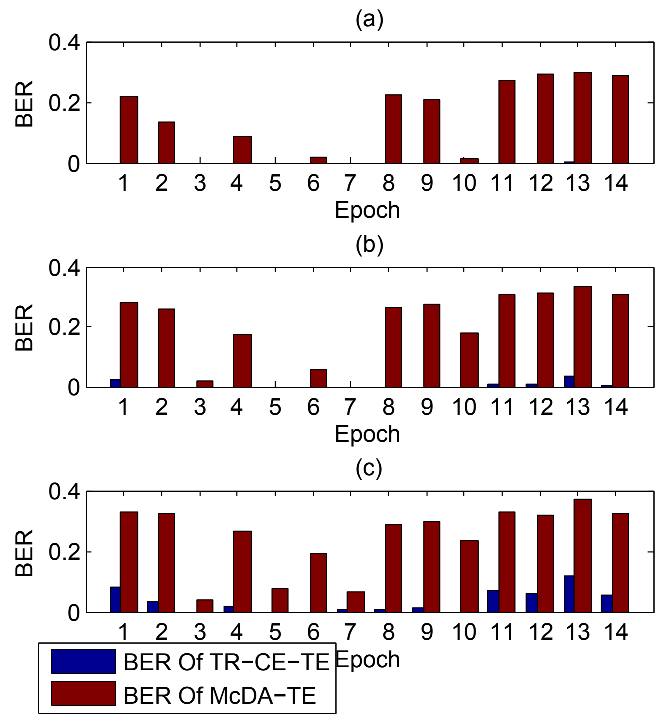

Fig. 6. The BER performance of the TR-CE-TE and McDA-TE in experiments . (a) is BER performance of two scheme of equalizers with 6 channels (channel 1-6) used. (b) is that with 5 channels (channel 1-5) used,and (c) is that with 4 channels (channel 1-4) used.

V. Conclusion

A receiver scheme combining TRC with CE-TE is been proposed for phase coherent underwater communication. The TRC compresses the ISI induced by multipath propagation in ocean environments. CE-TE eliminates residual ISI with a reasonable complexity. In the field experiment, proposed scheme can be realized with error-free performance at an 3km range in 6 channels combined. The experimental results provide further evidences that the proposed scheme can be a quite promising option for realizing high-speed, highly reliable data communication in a complex underwater acoustic channel.

Acknowledgment

This work was supported by Chinese National 863 projects 2002AA401004, 2009AA093301 and 2009AA093601.

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