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 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 compromised. 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 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

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 \( k \)th receiver of array with \( N_r \) elements can be expressed as:
\[
    r_{k,n} = (x_n \otimes h_{k,n}) + w_{k,n}, k = 1 \ldots N_r
\]  
where \( n \) are time indices, respectively. \( h_{k,n} \) is the \( n \)th gain in \( k \)th receiver channel impulse response, \( x_n \) is the QPSK symbol generated by transmitter, \( w_{k,n} \) is additive noise in \( k \)th 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.
\[
    z_n = \sum_{k=1}^{N_r} r_{k,n} \otimes \hat{h}_{k,-n}^\ast
\]  
The estimation of \( k \)th 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
\[
    z_n = \sum_{k=1}^{N_r} (r_{k,n} \otimes \hat{h}_{k,-n}^\ast)
    = \sum_{k=1}^{N_r} ((h_{k,n} \otimes \hat{h}_{k,-n}^\ast) \otimes x_n) + w_{k,n} \otimes \hat{h}_{k,-n}^\ast
    = \sum_{k=1}^{N_r} (\{ (h_{k,n} \otimes \hat{h}_{k,-n}^\ast) \otimes x_n \} + \sum_{k=1}^{N_r} w_{k,n} \otimes \hat{h}_{k,-n}^\ast)
    = \{ q_n \otimes x_n \} + \zeta_n
\]
where \( q_n \) is the autocorrelation of the channel impulse response summed over \( N_r \) channels, \( \zeta_n \) 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 transmitters, the received probe signal at the \( k \)th receiver element can be expressed as
\[
d_{n,k} = p_n \otimes h_{k,n} + w_{k,n}
\]
The known probe signal \( p_n \) can be used for correlating with the received probe signal \( d_{n,k} \), so \( h_{k,n} \) can be obtained by
\[
    \hat{h}_{k,n} = p_{n}^\ast \otimes d_{n,k}
    = (p_{n}^\ast \otimes p_{n}^\ast) \otimes h_{k,n} + p_{n}^\ast \otimes w_{k,n}
    = \lambda_n \otimes h_{k,n} + p_{n}^\ast \otimes w_{k,n}
\]
where \( \lambda_n = p_{n}^\ast \otimes p_{n}^\ast \) is the autocorrelation of the LFM signal which is not a perfect unit impulse response in practical 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 \cite{6}. The truncation for channel impulse in this scheme is took into consideration. As the time reversal processing in Section II, \( z_n \) can be developed into matrix form:
\[
    z_n = Q_n x_n + \zeta_n
\]
where
\[
    x_n = [x_{n-K_f-M_f}, \ldots, x_{n+K_p+M_p}]^T
    \zeta_n = [\zeta_{n-K_f}, \ldots, \zeta_{n+K_p}]^T
    z_n = [z_{n-K_f}, \ldots, z_{n+K_p}]^T
    q_n = [q_{n-M_f}, \ldots, q_{n+M_p}]^T
    Q_n = \begin{bmatrix}
        q_n-K_f & 0 & 0 \\
        0 & \ddots & 0 \\
        0 & 0 & q_n+K_p \\
    \end{bmatrix}
\]
The observation window of \( z_n \) contains \( K_f+K_p+1 \) received symbols. The channel length of \( q_n \) 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 \( q_n \) 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 \( q_n \) for main power is truncated to implement the channel estimated based turbo equalizer to make the complexity low. \( q_n \) are truncated into \( \tilde{g}_n \) as:
\[
    \tilde{g}_n = [q_{n-L_f}, \ldots, q_{n,L_p}] (L_p + L_f < < M_p + M_f)
\]
As the operation for truncating the \( q_n \) increases the noise power by \( \sigma_{\text{trunc}}^2 = (q_n q_n^H - q_n q_n^H) \sigma_{xx}^2 \). So noise correlation matrix can be expressed as:
\[
    \mathbf{R} = \sigma_{\text{trunc}}^2 \mathbf{I}_M + E (\zeta_n \zeta_n^H), M = M_p + M_f + 1
\]
the matrix \( \mathbf{I}_M \) is the \( i \times i \) identity matrix.

Without perfect knowledge of the channel matrix \( \mathbf{g}_n \), the imperfect estimated channel impulse response \( \tilde{g}_n \), channel estimation error can be stated as \( \epsilon_n = g_n - \tilde{g}_n \). The MMSE estimate of the symbol \( \hat{x}_n \) is given by:
\[
    \hat{x}_n = ((\tilde{g}_n \Sigma_n \tilde{g}_n^H + \mathbf{R} + E (\epsilon_n \epsilon_n^H))^{-1} \mathbf{s}_n \) \mathbf{u}_n
    \mathbf{u}_n = \mathbf{y}_n - \tilde{g}_n \mathbf{x}_n
    \mathbf{s}_n = \tilde{g}_n [0_{1 \times (K_f+M_f)}, 1, 0_{1 \times (K_p+M_p)}]^T
\]
where
\[
    \mathbf{x}_n = [\bar{x}_{n-K_f-M_f}, \ldots, \bar{x}_{n-1}, 0, \bar{x}_{n+1}, \ldots, \bar{x}_{n+K_p+M_p}]^T
    \Sigma_n = \text{diag}([\mu_{n-K_f-M_f}, \ldots, \mu_{n-1}, 1, \mu_{n+1}, \ldots, \mu_{n+K_p+M_p}])
\]
\( \bar{x}_n \) 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 \( \bar{x}_n \):

\[
\bar{x}_n = \mathbb{E}_L \{ x_n \} = \sum_{s \in S} s \cdot \Pr_L \{ x_n = s \}
\]

\[
v_n = \text{Cov}_L \{ x_n, \bar{x}_n \}
= \left( \sum_{s \in S} ss^* \cdot \Pr_L \{ x_n = s \} \right) - \bar{x}_n \bar{x}_n^*
\]

where \( S \) 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.

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

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 \( \bar{x}_n \). The same algorithm as [3] is used to obtain the extrinsic 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 depth from 14.8 m to 18.4 m. 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 depth from 14.8 m to 18.4 m. The experiments of communications 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.

#### 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 reversal is 200, which is also fractionally spaced. For 4 channels being processed, the number of equalizer taps and that of channel estimator taps

<table>
<thead>
<tr>
<th>FRAME LENGTH</th>
<th>DIFFERENT SYMBOL RATES</th>
</tr>
</thead>
<tbody>
<tr>
<td>DATA+TRAIN</td>
<td>2136 (1936+200)</td>
</tr>
</tbody>
</table>

\( \frac{8}{12} \) is used to obtain the extrinsic LLR for all bits.

![Fig. 3. The Frame Structure.](image)

![Fig. 4.](image)
TABLE II. THE EQUALIZER PARAMETERS SETUP

<table>
<thead>
<tr>
<th>Parameter</th>
<th>McDA-TE</th>
<th>TR-CE-TE</th>
</tr>
</thead>
<tbody>
<tr>
<td>Equalizer Order</td>
<td>Feedforward: $30 \times 2 \times N_r$</td>
<td>Linear Equalizer: $10 \times 2$</td>
</tr>
<tr>
<td></td>
<td>Feedback: 20</td>
<td>Channel Estimator: $8 \times 2$</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>4-6</td>
<td>4-6</td>
</tr>
<tr>
<td>Number of Iterations</td>
<td>Turbo Equalizer: 5</td>
<td>Turbo Equalizer: 5</td>
</tr>
<tr>
<td></td>
<td>Turbo Decoder: 2</td>
<td>Turbo Decoder: 2</td>
</tr>
</tbody>
</table>

in TR-CE-TE are only $10 \times 2$ and $8 \times 2$, 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.

In Fig. 5, the convergence behavior is been plotted. As the turbo code is powerful enough to provide soft information, turbo equalization could achieve 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.
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.

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References


