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# Forward-Backward Turbo Equalization for Underwater Communication Systems

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**Abstract**—This paper proposes a novel equalization algorithm for underwater communication. It performs iteratively adaptive equalization and channel decoding in order to reduce intersymbol interference (ISI). In particular, our proposed forward-backward turbo equalizer uses two sparse feedforward-feedback equalizers operating in opposite time directions, each sparse equalizer performing jointly phase synchronization to enhance the overall performance. Our simulations, both on synthetic and real underwater channels validate the effectiveness of the proposed scheme as they show a considerable gain compared to the conventional adaptive turbo equalization algorithm.

**Index Terms**—Turbo equalization, channel decoding, underwater acoustic communication.

## I. INTRODUCTION

The underwater communication is very challenging due to the severe characteristics of underwater channels: echos, limited bandwidth, time-frequency selectivity, long multipath effect, Doppler and high ambient noise. All these factors severely impact the communication system performance [1]–[3]. To face these issues, various equalization techniques have been proposed in the literature [4]–[6]. In particular at the time as the discovery of turbo codes, turbo equalization was proposed [1]; however, the exponential complexity of this algorithm considerably limits its use in practice. To tackle this issue, several alternative turbo equalizers were proposed with much lower computational complexity; in particular, a linear Minimum Mean Square Error (MMSE) turbo equalization scheme was investigated in [2], [3]. In [7] a linear adaptive turbo equalization proposal uses a feedforward-feedback filters and the coefficients of the equalizer are obtained in an adaptive way. A sparse turbo equalizer is studied in [8] for underwater acoustic systems, and a fully adaptive turbo equalization technique is proposed in [9]. However, over severe underwater channels, the number of required iterations to reach an acceptable system performance may be very high leading to a considerable computational complexity; the convergence of such algorithm is not always possible in particular at low Signal-to-Noise Ratio (SNR) and over fast varying channels [10].

In this paper, we propose a novel turbo equalizer. It consists of two feedforward-feedback equalizers operating in opposite time direction. Both equalizers are sparse to reduce the computational complexity, and they perform jointly phase

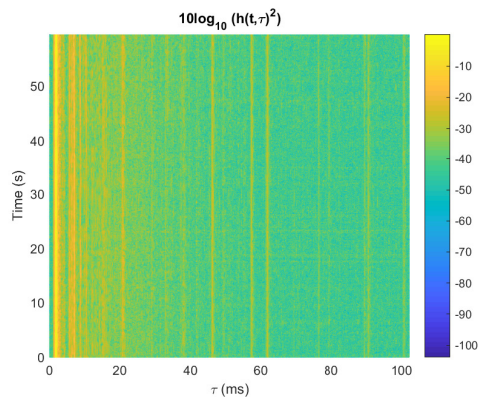


Fig. 1. The multi-path BCH1 channel.

synchronization to better track phase variations, as a small phase error may lead to a severe Bit Error Rate (BER) degradation. Our proposal is tested over both synthetic time-varying channels and some real underwater channels using the “WATERMARK” platform [11]; [11] includes a library of impulse responses measured in Norway, France, and Hawaii. For instance, Fig.1 shows the impulse response of BCH1 measured off the commercial harbor of Brest, France.

The paper is organized as follows. Section II describes the system model. Section III presents the Forward-Backward equalizer structure, and details the process to update the equalizer coefficients. Some numerical results obtained over real underwater channels and over discrete time-varying channels are displayed in section IV. Conclusions are finally drawn in Section V.

## II. SYSTEM MODEL

The transmission scheme is displayed in Fig.2. First, the information bits are channel encoded [12], [13] and then interleaved to enhance the code diversity. It is worth mentioning that the diversity could inherently be further enhanced by using rotated constellation with no bandwidth loss [14]–[19]. The interleaved bits are then transformed into a serie of modulated symbols  $s_k$ . A known training sequence is added at the head and tail of the symbol block. The frame is then transmitted through a doubly-selective channel.

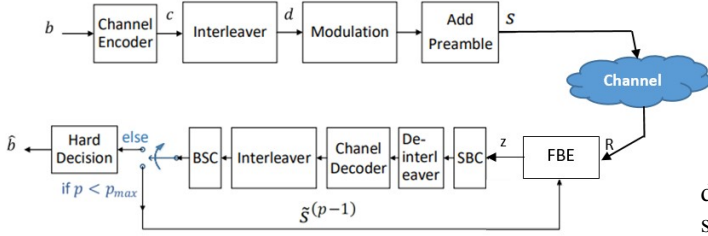


Fig. 2. System model.

The received symbols can be written as:

$$r_k = \sum_l h(l; k) s_{k-l} + w_k, \quad (1)$$

where  $h(l; k)$  is a channel impulse response tap, and  $w_k$  is an additive white Gaussian noise at time index  $k$ .

At the receiver side, several signal processing algorithms are needed in order to correctly recover the transmitted frame such as time and frame synchronization [20]–[24]. In this paper, we design a novel equalization algorithm. The receiver performs jointly equalization, as described in section III, and channel decoding in an iterative manner. At the first iteration  $p = 1$ , the soft output of the decoder are unknown and the equalizer is a purely adaptive transversal filter. For the other iterations  $p > 1$ , it uses the received signal and the soft output of the decoder of the previous iteration to equalize the symbols with our proposed Forward-Backward equalizer (FBE). The receiver then computes the Log-Likelihood Ratio (LLR) values based on the equalized symbols and performs Bit-Interleaved Coded Modulation (BICM) decoding. These LLR values are used to estimate the transmitted symbol, and these symbols are used as an a priori in the next iteration. After  $p_{max}$  iterations, the decoder makes hard decisions on the LLR values to recover the information symbols.

### III. FORWARD-BACKWARD TURBO EQUALIZATION

Our proposed Forward-Backward equalizer (FBE) structure is displayed in Fig. 3, where the index ( $p$ ) is dropped for simplicity. It consists of two separate feedforward-feedbackward equalizers. The coefficients of these equalizers are updated over two separate known preambles: one at the head  $P1$  and the other at the tail  $P2$  (postamble in the forward time

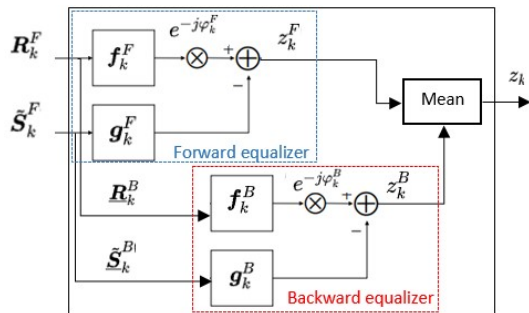


Fig. 3. The proposed Forward-backward equalizer structure.

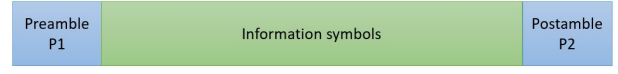


Fig. 4. The frame structure.

direction or preamble in the backward time direction) of a symbol block, (see Fig.4). The forward equalizer is composed of two filters  $f_k^F$  and  $g_k^F$  that are respectively fed by the output channel sequence  $\mathbf{R}_k^F = [r_{k+L_1}, \dots, r_{k-L_1}]^T$  and the symbol sequence  $\tilde{\mathbf{S}}_k^F = [\tilde{s}_{k+L_2}, \dots, \tilde{s}_{k-L_2}]^T$ , where  $\tilde{s}_k$  denotes either the known preamble or the soft estimated symbol obtained by the channel decoder at the previous iteration. Similarly, the Backward equalizer is composed of two filters  $f_k^B$  and  $g_k^B$  that are fed by  $\mathbf{R}_k^B = [r_{k-L_1}, \dots, r_{k+L_1}]^T$  and  $\tilde{\mathbf{S}}_k^B = [\tilde{s}_{k-L_2}, \dots, \tilde{s}_{k+L_2}]^T$ . The outputs of these equalizers at time instant  $k$  are given by:

$$\begin{cases} z_k^F = \mathbf{f}_k^{F^T} \mathbf{R}_k^F e^{-j\phi_k^F} - \mathbf{g}_k^{F^T} \tilde{\mathbf{S}}_k^F, \\ z_k^B = \mathbf{f}_k^{B^T} \mathbf{R}_k^B e^{-j\phi_k^B} - \mathbf{g}_k^{B^T} \tilde{\mathbf{S}}_k^B, \end{cases} \quad (2)$$

where  $\mathbf{f}_k^I = [f_{-L_1}^I(k), \dots, f_{L_1}^I(k)]^T$  (resp.  $\mathbf{g}_k^I = [g_{-L_2}^I(k), \dots, g_0^I(k) = 0, \dots, g_{L_2}^I(k)]^T$ ) is a transversal filter equalizer of length  $2L_1 + 1$  (resp.  $2L_2 + 1$ ), with  $I$  denoting being either F(Forward) or B(Backward), and  $\phi_k^I$  is the estimated phase error. The final output of our forward-backward equalizer is then obtained as:

$$z_k = (z_k^F + z_k^B) / 2. \quad (3)$$

The coefficients of the filters are updated in order to minimize the mean-square-error (MSE) criterion defined as:

$$MSE = \mathbb{E} (|z_k^I - \tilde{s}_k|^2), \quad (4)$$

The filter coefficients  $\mathbf{f}_k^I, \mathbf{g}_k^I$  and the estimated phase error  $\phi_k^I$ , for both the forward and backward equalizers, are updated as:

$$\begin{cases} f_{k+1}^I = f_k^I - \mu \nabla_{f_k^I} (|z_k^I - \tilde{s}_k|^2), \\ g_{k+1}^I = g_k^I - \mu \nabla_{g_k^I} (|z_k^I - \tilde{s}_k|^2), \\ \phi_{k+1}^I = \phi_k^I - \gamma \nabla_{\phi_k^I} (|z_k^I - \tilde{s}_k|^2), \end{cases} \quad (5)$$

where  $\nabla(X)$  is the gradient of  $X$  and  $\mu$  and  $\gamma$  are appropriate step-sizes.

In order to reduce complexity, once convergence is achieved and given the sparsity of the underwater channel, we zero out the filters coefficients whose energy is below a predefined threshold.

## IV. SIMULATION RESULTS

### A. Time-Varying Channels

In our simulations, user frames of 800 bits are coded with a conventional code of rate 1/2 with generator polynomials equal to (37,33), and we also used the QPSK constellation. In the case of a conventional turbo equalization, we used a training sequence of 200 symbols, and this sequence is divided in two when we simulate the performance with our proposed algorithm. The Feedforward (resp. Feedbackward) filter length

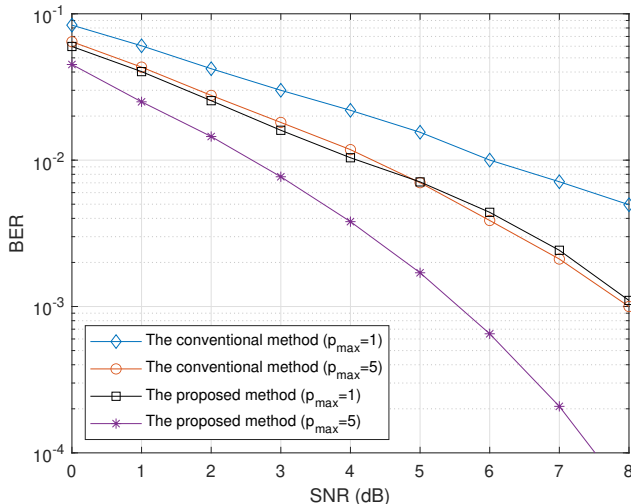


Fig. 5. BER comparison between the adaptive equalizer [8] and our proposal with  $B_d T = 3.10^{-4}$ .

is fixed to 21 (resp. to 11) and we set  $p_{max}$  to 5. We first tested our method over a synthetic time-varying channel with 3 taps, such that at time index  $k$  each of them are given by:

$$h_l(k) = \sqrt{\frac{P_l}{N}} \sum_{i=1}^N e^{j(2\pi(f_d \cos \xi_{l,i})kT + \psi_{l,i})}, \quad (6)$$

where  $f_d$  and  $P_l$  correspond to the maximum Doppler and to the mean power associated with the  $l$  path, respectively. Parameters  $\xi_{l,i}$  and  $\psi_{l,i}$  are uniform random variables over  $[0; 2\pi[$ . In our simulations  $N$  was fixed to 10. Generally the Doppler effect is characterized by the product of a Doppler band  $B_d = 2f_d T$  [7], where  $T$  is the symbol duration.

Fig. 5 (resp. Fig. 6) compares the BER performance between the conventional sparse turbo equalization algorithm [8] and our proposed algorithm for two Doppler bands. It can be observed in both figures that the performance obtained with both considered equalizers in terms of BER performance is considerably enhanced after 5 iterations compared with only one iteration. Moreover, our equalizer leads to a considerable gain of roughly 2dB compared to the conventional method for  $B_d T$  equals to  $3.10^{-4}$  after only one iteration, and we even obtain 2.5dB gain after 5 iterations.

Finally, Fig.6 shows that with a large Doppler effect only our method is able to reach an acceptable BER performance at a SNR equal to 10dB.

## B. WATERMARK

We also carried out simulations over two real UWA channels, one measured off the commercial harbor of Brest, (BCH1) and one off the shallow stretch of Oslofjorden (NOF1) over a distance of 750m. A 511 preamble was added at the beginning of the frame to allow frame detection and initial Doppler synchronization. The measurement parameters of these two channels are summarized in Table I.

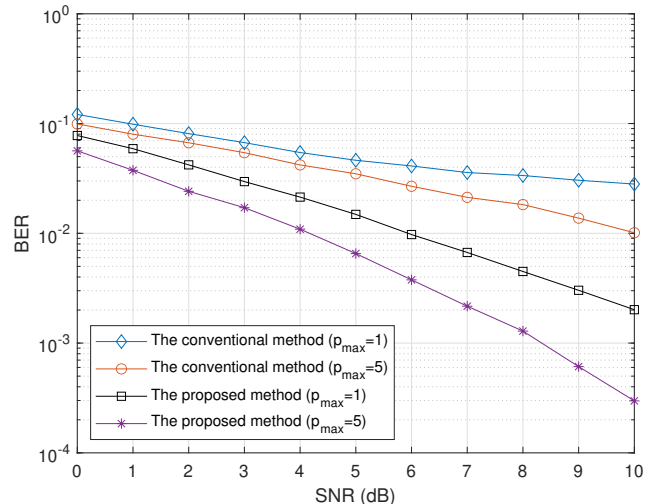


Fig. 6. BER comparison between the adaptive equalizer [8] and our proposal with  $B_d T = 10^{-3}$ .

TABLE I  
THE MEASUREMENT PARAMETERS.

Parameters	BCH1	NOF1
Time of the year	May	June
Water depth	20 m	10 m
-3dB freq. band	32.5-37.5 KHz	10-18 KHz
Duration	59.4 s	32.9s
Type	SIMO	SISO
Hydrophones	4	1

Fig. 7 (resp. Fig.8) compares the BER performance over the BCH1 channel (resp. NOF1 channel). We can observe again that our forward-backward equalization algorithm surpasses the conventional turbo equalization algorithm [8] by achieving a gain of 1 dB over the BCH1 channel (resp. 0.8 dB over the NOF1 channel) after the first iteration and 1.5 dB (resp. 1.2 dB) after the 5-th iteration. These results are coherent with figures 5 and 6 as the BCH1 channel is less stable than the NOF1 channel.

## V. CONCLUSION

In this paper, we proposed a novel forward-backward turbo equalizer for underwater communication systems. Our proposed scheme utilizes two sparse feedforward-feedback equalizers operating in opposite time directions with joint phase synchronization to mitigate the adverse effects of intersymbol interference (ISI) and phase errors in underwater channels. Simulation results, in particular over different real underwater channels, show that our proposal leads to a significant performance gain compared to conventional turbo equalization algorithm, specially over fast varying multipath channels. In the future, we would like to study the performance of our proposal with Faster-than-Nyquist signaling [25]–[29].

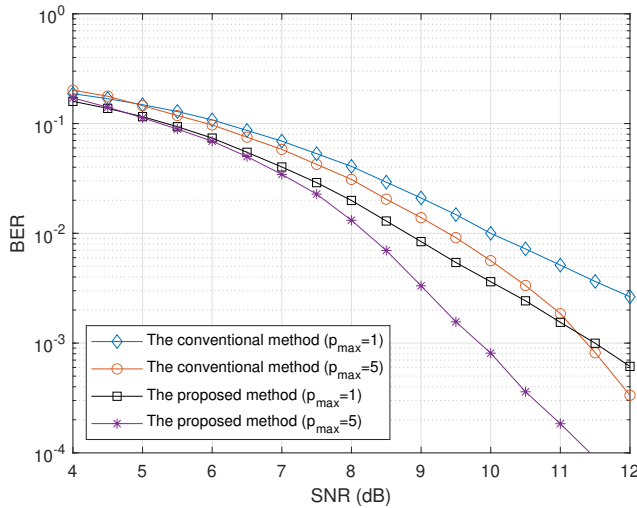


Fig. 7. BER comparison over the channel BCH1 [11].

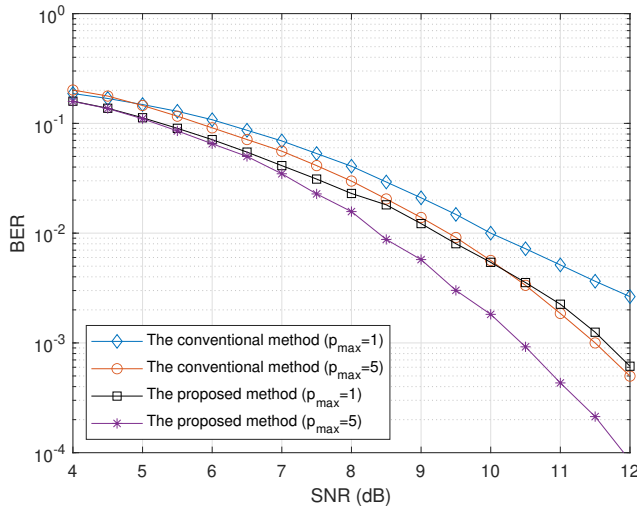


Fig. 8. BER comparison over the channel NOF1 [11].

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