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MULTIPLE-INPUT HYBRID FREQUENCY-TIME DOMAIN ADAPTIVE DECISION FEEDBACK EQUALIZATION FOR UNDERWATER ACOUSTIC DIGITAL COMMUNICATIONS

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Abstract

In this paper, a multiple-input hybrid frequency-time domain decision feedback equalizer is proposed to deal with the intersymbol interference in underwater acoustic communications. To track the time-varying underwater acoustic channel, a block-based adaptive algorithm is considered. In addition, we apply the overlap-and-save technique at the receiver. This technique makes it possible to remove the overhead due to the transmission of cyclic prefix over each block of data, such as in usual frequency-domain equalization systems. The second advantage of The OS technique is the possibility to change the block and equalizer sizes. The transmission scheme is validated in the Atlantic Ocean over a distance up to 2 kilometers in shallow water with high data rate single carrier QPSK communications (10 kbps). The receiver includes an efficient timing recovery tracking scheme followed by equalization. The results of simulations and experiments are evaluated in terms of bit error rate (BER) and mean square error (MSE).

1 INTRODUCTION

The underwater acoustic (UWA) channel is characterized by a high frequency selectivity resulting in intersymbol interference (ISI). UWA channel is also time-varying yielding Doppler spread [1]. Single carrier equalization techniques are necessary at the receiver side to compensate this ISI. The decision feedback equalizers (DFE) offer better performance than linear equalizers (LE) [2]. The frequency-domain equalization (FDE) is an interesting alternative to time-domain equalization (TDE) [3]. The gain is twofold: reduce the computational complexity and simplify the equalizer parameters setting [4]. However, when implementing the DFE in the block-based frequency-domain, a causality problem appears [5]. A hybrid frequency-time domain implementation, where the feedforward filter is implemented in the frequency-domain on a block-by-block basis and the feedback filter is implemented in the time-domain on a symbol-by-symbol basis, allows to overcome this causality problem [6]. It is worth noting that the DFE incurs error propagation when incorrect decisions are taken, which mainly occurs at low signal to noise (SNR) ratios [2]. The use of multiple-input allows to achieve a higher equivalent SNR at the equalizer output and prevents the equalizer from error propagation phenomenon when the SNR is low [7]. The equalizer is designed using the minimum mean square error (MMSE) criterion based on adaptive processing [9]. When the channel impulse response (CIR) changes within a block duration, that is the case in long transmissions and time-varying channels, adaptive processing enables the equalizer to track the time-variation in the channel. The equalizer taps are iteratively updated with a decision directed mode. The advantage compared to channel estimation-based equalization [8] is that the transmission of preambles, to re-estimate the channel whenever it varies, is avoided. A training sequence is sent at the beginning of the transmission to allow equalizer convergence to the optimal response. Motivated by the work in [10], we use the overlapping techniques such as the overlap-and-save (OS), which avoids the spectral efficiency loss due to the insertion of cyclic-prefix (CP) intervals in CP-based FDE.

In this paper, we propose a multiple-input hybrid frequency-time domain adaptive DFE (HD-OS-ADFE) where a common feedback filter is used for all the inputs. Second order phase locked loops (PLL) are

introduced in order to compensate the residual frequency offsets within each input and are optimized jointly with the equalizers. The receiver includes also a channel decoder. The results of simulations and experiments are given in terms of BER and mean square error (MSE). The HD-OS-ADFE is tested over real time-variant underwater acoustic channels in the Atlantic Ocean.

2 RECEIVER DESIGN

We consider an adaptive DFE for an all-digital single-carrier receiver in a time-varying channel environment. The DFE is implemented on a hybrid frequency-time domain block basis. The information data are encoded using a convolutional code, interleaved and passed through a QPSK modulator. Each set of $m = \log_2(M)$ coded data is associated with an M-ary complex symbol d_n by a Gray mapping. The resulting data stream is transmitted through an underwater time varying channel. Let $s(t)$ be the transmitted signal:

$$s(t) = \Re\left\{ \sum_{n=-\infty}^{+\infty} d_n g(t - nT) e^{j(2\pi f_c t + \psi)} \right\} \quad (1)$$

where f_c is the carrier frequency, ψ is the carrier phase uncertainty, $1/T$ is the symbol rate with T the symbol duration. $\{d_n\}$ are the QPSK symbols with variance σ_d^2 . $g(t)$ is a pulse-shaping filter.

When the pass-band digital received signal is centered on a relatively low carrier frequency, an all-digital reception can be applied [11]. Since the received signal is band-limited, an oversampling is performed at the rate $1/T_s$ where T_s is chosen to fulfill the sampling theorem conditions. The down-conversion is performed digitally and the optimum sampling is realized from a sample rate converter. The resulting all-digital receiver is depicted in Fig. 1.

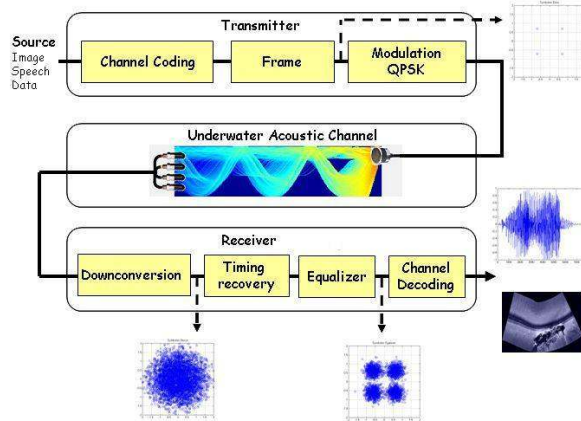


Figure 1: All-digital receiver

In wide-band transmissions, such as underwater acoustic communications [1, 12], the Doppler effect introduces a scaling of the symbol period that must be taken into account in the design of the timing recovery scheme. To restore the transmitted signal, the optimum sample time depends not only on the propagation delay observed at the antenna, but also on a common Doppler shift function of the relative speed of the platforms and the propagation wave velocity [13]. At the receiver side, the optimum sampling time is unknown and must be estimated. Initial compensation of common Doppler shift is performed, based on the estimation of the relative velocity measured using a short preamble [14] transmitted at the beginning. Note that the preamble is also used to realize the transmission detection and frame synchronization. Then, a non data aided (NDA) timing recovery tracking scheme is designed which takes into account the residual Doppler shift due to the platform movements and the channel delay observed at the antenna [15].

After demodulation, sampling, Doppler compensation and timing recovery, the received signal is modeled as output of a discrete time channel where each output $j; j = 1, \dots, N_R$ is corrupted by an additive,

zero-mean, white Gaussian noise (AWGN) $w_n^{(j)}$ with variance $\sigma_{(j)}^2$. The n -th sample received by the j -th antenna can be written as follows,

$$r_n^{(j)} = \sum_{l=0}^{L-1} h_{n,l}^{(j)} d_{n-l} + w_n^{(j)} \quad (2)$$

where $h_{n,l}^{(j)}$ are the L coefficients of the multi-path time-varying channel corresponding to the j^{th} antenna.

3 HYBRID FREQUENCY-TIME DOMAIN ADAPTIVE DFE

FDE is an interesting alternative to time-domain equalization for single carrier transmission systems. Using fast Fourier transform (FFT), the FDE is performed over a block of N samples by multiplying term by term the N samples by the N equalizer taps. The performance is equivalent to that of the TDE. FDE has low computational complexity than TDE depending on the channel delay spread in terms of symbol duration.

A hybrid frequency-time domain OS adaptive decision feedback equalizer (HD-OS-ADFE) is proposed in this paper. The feedforward filter is implemented in the frequency-domain on a block-by-block basis whereas the feedback filter is implemented in the time-domain on a symbol-by-symbol basis, as depicted in Fig. 2. The OS technique is applied to the block frequency-domain part in order to avoid the overhead resulting from the transmission of CP intervals [10]. We denote N_{FF} the length of the feedforward filter and N_{FB} the length of the feedback filter.

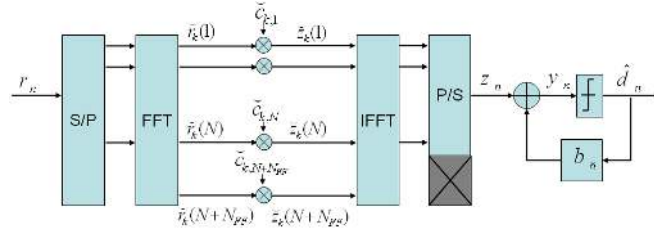


Figure 2: Hybrid frequency-time domain overlap-and-save DFE (HD-OS-DFE)

\mathbf{F} and \mathbf{F}^{-1} are the $(N + N_{FF}) \times (N + N_{FF})$ discrete Fourier transform (DFT) matrix and the inverse DFT matrix, respectively. The elements of \mathbf{F} are defined as follows,

$$F(s, q) = e^{-j2\pi sq / (N + N_{FF})} \quad (3)$$

After conversion to time-domain using \mathbf{F}^{-1} , the feedforward filter output block is given as follows,

$$\mathbf{z}_k = \text{Last } N \text{ terms of } \left\{ \mathbf{F}^{-1} \check{\mathbf{R}}_k \check{\mathbf{c}}_k \right\} \quad (4)$$

where k is the block index, $\check{\mathbf{R}}_k$ is a diagonal matrix containing the k^{th} frequency-domain overlapped input data block \mathbf{r}_k ,

$$\check{\mathbf{R}}_k = \text{diag}(\mathbf{F} \mathbf{r}_k) \quad (5)$$

\mathbf{r}_k is the time-domain overlapped input block containing the last N_{FF} elements from the previous data block and the N elements of the current data block, according to the OS principle,

$$\mathbf{r}_k = [r_{kN-N_{FF}}, \dots, r_{kN-1}, r_{kN}, \dots, r_{kN+N-1}]^T \quad (6)$$

and finally, $\check{\mathbf{c}}_k$ is the frequency-domain feedforward filter vector containing the N_{FF} significant taps augmented by N zeros according to the OS principle,

$$\check{\mathbf{c}}_k = \mathbf{F} [c_k, 0, \dots, 0]^T \quad (7)$$

\mathbf{c}_k is the k^{th} time-domain coefficient vector of the feedforward filter,

$$\mathbf{c}_k = [c_{k,1}, \dots, c_{k,l}, \dots, c_{k,N_{FF}}] \quad (8)$$

After parallel-to-serial conversion, data symbols are detected and filtered symbol-by-symbol by the feedback filter in the time-domain to remove the residual a-priori interference part, as given in (9).

$$y_n = z_n - \sum_{l=1}^{N_{FB}} b_{k,l} \hat{d}_{n-l} \quad (9)$$

where $b_{k,l}$ is the l^{th} component of the k^{th} time-domain feedback filter coefficient vector

$$\mathbf{b}_k = [b_{k,1}, \dots, b_{k,l}, \dots, b_{k,N_{FB}}] \quad (10)$$

Using the minimum mean square error (MMSE) criterion, the feedforward and feedback filters taps are designed based on the least mean square (LMS) adaptive algorithm, where the equalizers taps are iteratively updated. This enables to track the time-variation in the channel in a decision directed mode. However, a training sequence must be sent at the beginning of the transmission to allow the equalizer to converge to the optimal solution. Each symbol error is evaluated as a difference between the equalizer output and the corresponding decision as follows,

$$e_n = y_n - \hat{d}(y_n) \quad (11)$$

where $\hat{d}(y_n)$ is the decision taken from y_n . The block error vector is defined as,

$$\mathbf{e}_k = [e_{kN}, e_{kN+1}, \dots, e_{kN+N-1}]^T \quad (12)$$

and the corresponding frequency-domain error block $\check{\mathbf{e}}_k$ is given as follows,

$$\check{\mathbf{e}}_k = \mathbf{F} \mathbf{e}_k \quad (13)$$

Both feedforward and feedback filters are updated block by block according to the hypothesis of slow time-varying channel within the block duration. In other words, the filters taps are kept fixed during the block interval and are updated at the next block iteration. The update equations are given as follows,

$$\check{\mathbf{c}}_{k+1} = \check{\mathbf{c}}_k - \mu_c \mathbf{F} \mathbf{G} \mathbf{F}^{-1} \check{\mathbf{R}}_k^H \check{\mathbf{e}}_k \quad (14)$$

for the feedforward filter, where \mathbf{G} is an $(N + N_{FF}) \times (N + N_{FF})$ time-domain constraint matrix also allowing, according to the OS principle, to discard and insert elements when performing FFT operations,

$$\mathbf{G} = \begin{pmatrix} \mathbf{I}_{N_{FF}} & \mathbf{0}_N \\ \mathbf{0}_N & \mathbf{0}_{N_{FF}} \end{pmatrix}, \quad (15)$$

and

$$\mathbf{b}_{k+1} = \mathbf{b}_k + \mu_b \mathbf{D}_k^H \mathbf{e}_k \quad (16)$$

for the feedback filter. μ_c and μ_b are the adaptation step sizes of the feedforward filter and the feedback filter, respectively.

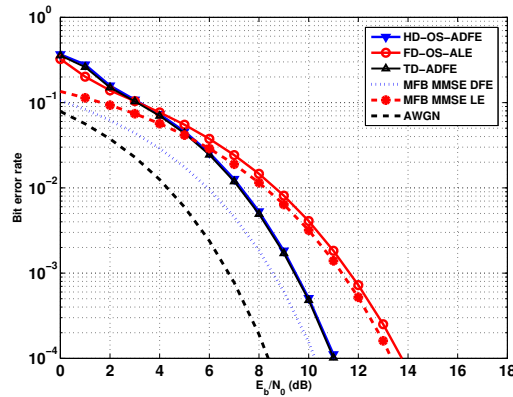


Figure 3: Bit error rate vs. E_b/N_0 in Porat channel model, number of transmitted blocks: 400, QPSK ; training period: 80 blocks

4 SIMULATION RESULTS

The performance are given in terms of bit error rate (BER) as a function of the signal to noise ratio E_b/N_0 , where E_b is the average energy per data bit and N_0 is the one-sided noise power spectral density. The HD-OS-ADFE is simulated over a medium frequency-selective channel model (Porat) [2]. The transmitted signal is QPSK modulated. The HD-OS-ADFE is compared with the following equalizers: the matched filter bound (MFB) MMSE DFE that is an ideal equalizer with infinite length and perfectly known CIR, the matched filter bound (MFB) MMSE linear equalizer, the time-domain adaptive DFE (TD-ADFE) and the frequency-domain OS adaptive linear equalizer (FD-OS-ALE). A theoretical BER for additive white Gaussian noise (AWGN) channel is also included as a reference.

As shown in Fig. 3, the linear equalizer (FD-OS-ALE) is less efficient than the HD-OS-ADFE. The HD-OS-ADFE and the TD-ADFE perform the same. While the HD-OS-ADFE achieves a BER of about 10^{-4} for $E_b/N_0 = 11$ dB, the FD-OS-ALE achieves the same BER with a gap of 3 dB compared to the HD-OS-ADFE. The HD-OS-ADFE and the TD-ADFE become as close to the lower theoretical MFB-MMSE-DFE of Porat channel model as the SNR increases.

5 MULTIPLE-INPUT HD-OS-ADFE

A multiple-input HD-OS-ADFE is proposed using a common time-domain feedback filter for all the inputs as depicted in Fig. 4, using N_R receive antennas. Second order phase locked loops (PLL) follow feedforward filter outputs in order to compensate the residual frequency offsets at each input. A joint optimization with the equalizer is performed, based on the same error minimisation.

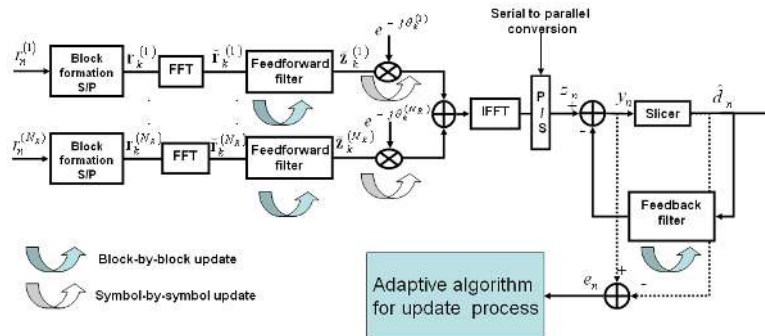


Figure 4: Multiple-input hybrid frequency-time domain adaptive DFE using one feedback filter

6 EXPERIMENTAL RESULTS

The HD-OS-ADFE is validated over real time-variant underwater acoustic channel in the Atlantic Ocean. Experimental sea trials were carried out in October 2006 in the site "bay of Brest", France, by DGA/GESMA. During these trials, transmission was carried out in a shallow water environment with a depth of about 10 to 30 meters. A transmission was performed over a distance of 500 meters and the carrier frequency was equal to $f_c = 35kHz$. The coded bit rate was about $10kbits/s$. The training period was performed during 1 second. The transmitter was submerged and fixed at a buoy. Real-time text sentences were transmitted to a ship over a distance of 500 meters. The transmit power was set to 180 dB ref $1\mu Pa$ at 1 meter and the relative transmitter to receiver velocity was about $v = 0.5$ m/s. A multiple-input receiver scheme is considered using four hydrophones equally spaced at 25 cm. Four second order phase locked loops (PLL) were optimized jointly with the equalizers in order to iteratively compensate the residual frequency offsets in each input. Performance are presented in terms of measured bit error rate (BER) and measured mean square error (MSE) without taking into account the training period. The proposed structure is compared with a multiple-input frequency-domain adaptive time-domain (LMS) linear equalizer, which is also tested in the same conditions using the same receiver design. Two scenarios are provided to validate the proposed equalizer.

6.1 SCENARIO A: LOW ISI

In the first scenario (scenario A) of the experiment, the channel selectivity is medium and the ISI is relatively low. The estimated UWA channel impulse response is given in Fig. 5. The channel in this scenario is referred as channel A.

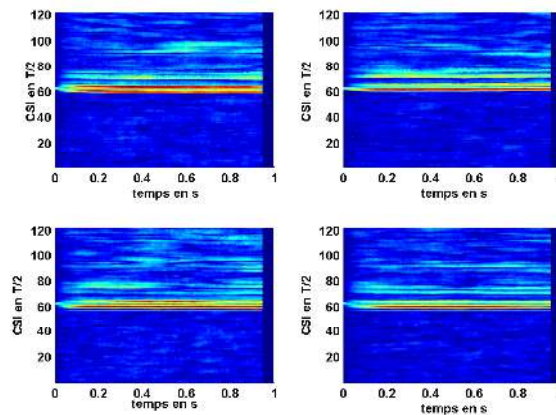


Figure 5: Estimated UWA channel impulse response channel A

In Fig. 6, we show the time evolution of the MSE. The block-length was fixed to 32 and the feedforward and feedback filter sizes were fixed to 32 and 40, respectively. The HD-OS-ADFE performs slightly better than the FD-OS-ALE, achieving a MSE value of about -9 dB. The evaluated BER after channel decoding was equal to 0 during the transmission for both HD-OS-ADFE and FD-OS-ALE.

6.2 SCENARIO B: HIGH ISI

In the second scenario (scenario B), the ISI was higher. The estimated UWA channel impulse response is shown in Fig. 7. The channel is referred as channel B. This channel contains more multipath than channel A and has a higher delay spread. The Block-length was fixed to 32 and the feedforward and feedback filter sizes were fixed to 32 and 40, respectively. The estimated SNR is lower. The use of multiple hydrophone was of great interest to achieve a higher equivalent SNR at the equalizer output.

A steady state is achieved at MSE value of about -8 dB. Although the HD-OS-ADFE achieved a MSE which is slightly better than the FD-OS-ALE, the measured BER, after channel decoding, was equal

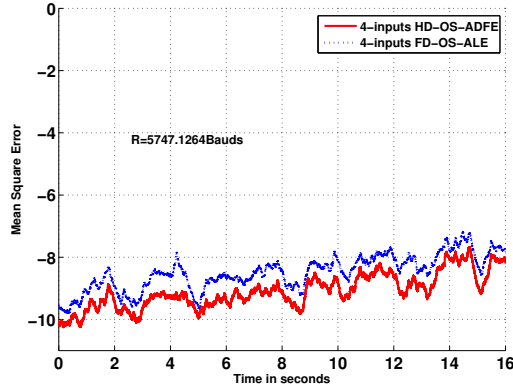


Figure 6: Mean square error, Channel A, $N_R = 4$, QPSK, 10 kbps, $d = 500$ meters, $\mu_c = \mu_b = 4 \cdot 10^{-3}$

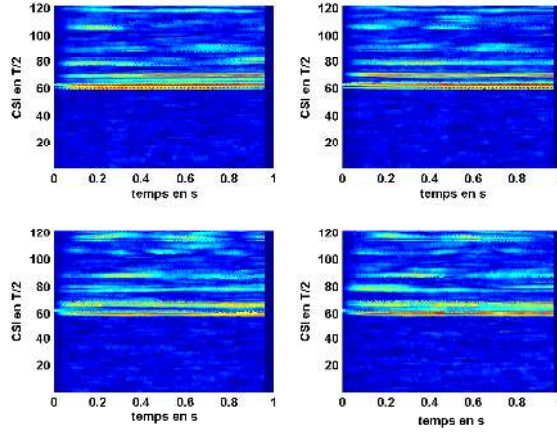


Figure 7: Estimated UWA channel impulse response channel B

to $2 \cdot 10^{-2}$ while the measured BER for the FD-OS-ALE was equal to $5.8 \cdot 10^{-3}$. This is mainly due to error propagation phenomenon. In fact, when MSE values are very high (i.e. around $-7/-6$ dB), the corresponding equivalent SNR is low and erroneous decisions are fed to the feedback filter. As a consequence, sets of neighbouring errors are formed and the channel decoder can not correct all errors. By setting a small value of the feedback filter adaptation step size, compared to the feedforward filter adaptation step size, the error propagation phenomenon can be limited and the ISI was completely cancelled after channel decoding (BER= 0) during the transmission time.

7 CONCLUSION

A hybrid frequency-time domain DFE is simulated and tested over a real UWA time-varying channel. We have first shown through simulations the advantages of the HD-OS-ADFE compared to the FD-OS-ALE in terms of BER performance and the gain in computational complexity vs. the time-domain DFE. Using multiple-input, the HD-OS-ADFE was able to cancel the ISI in the UWA channel. To prevent the equalizer output from residual frequency offsets, adaptive compensation is carried out using a PLL at each input, which is jointly optimized with the equalizer. Experimental results show the advantage of the proposed structure compared to the FD-OS-ALE when the ISI was relatively high.

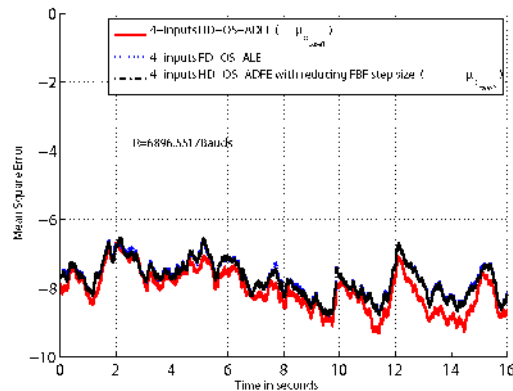


Figure 8: Mean square error, Channel B, $N_R = 4$, QPSK, 10 kbps, $d = 500$ meters, $\mu_c = 3 \cdot 10^{-3}$, $\mu_{b_{case1}} = 3 \cdot 10^{-3}$, $\mu_{b_{case2}} = 7 \cdot 10^{-5}$

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