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HAL Id: hal-02276898

https://hal.archives-ouvertes.fr/hal-02276898

Submitted on 11 Jun 2021

Adaptive Frequency-Domain Equalization for Underwater Acoustic Communications

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Abstract—In this paper, frequency-domain equalization approach is proposed to deal with the intersymbol interference in underwater acoustic communications. To track the time-varying underwater acoustic channel, an adaptive algorithm is considered using overlap-and-save method. 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. We compare both schemes through simulations to select the best one. 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).

I. INTRODUCTION

Underwater acoustic (UWA) communication systems suffer from the frequency selectivity of the multipath channel, which causes inter-symbol interference (ISI). UWA channel is also time-variant yielding Doppler spread. Often Orthogonal frequency division multiplex (OFDM) and equalization techniques are used at the receiver to compensate for the ISI [1]. OFDM system can avoid the inter-block interference (IBI) using guard period (e.g. cyclic prefix) inserted between consecutive OFDM symbols. However, in UWA communications, OFDM is very sensitive to Doppler spread, which induces inter carrier interference (ICI) [1], [2].

In this paper, we focus on single carrier (SC) transmissions. So, time-domain equalization (TDE) and frequency-domain equalization (FDE) are applied. FDE and TDE have the same theoretical performance with lower computational complexity for the FDE if the channel has a large delay spread compared to the symbol duration. In fact, FDE became recently an important alternative to the TDE in systems such as the 4th mobile generation long term evolution (LTE). Applied to single carrier systems in order to deal with the ISI, equalization generally requires a preliminary estimation of the channel carried out from training sequences preceding each data block [3], [4]. This assumes a time-invariant channel over a block duration and a channel estimation for each block. Another solution is to make a tracking of the time-varying channel using an adaptive processing method based on the least mean square algorithm (LMS). An adaptive frequency-domain equalizer (AFDE) is considered in this paper. For usual FDE-

based systems, each data block of length N is preceded by a cyclic prefix (CP) [2]. At the receiver, IBI can be removed by discarding the corresponding CP symbols. However, the transmission of CP generates a spectral efficiency loss and additional treatment is required at the transmitter to insert CP at the beginning of each block. We propose to use a technique avoiding the need of CP insertion. It applies the sectioning and overlapping methods such as overlap and save/add (OS/OA) [5], which enable to remove the overhead due to the transmission of CP. Another important advantage of the proposed method is that the block length is selected at the receiver according to the channel time-variation over the block duration. To these ends, we apply overlap-and-save technique to the adaptive equalizer (OS-AFDE). The receiver includes an efficient timing recovery tracking scheme to compensate for the Doppler effect induced by the platform movements, followed by OS-AFDE. The OS-AFDE is tested over real time-variant underwater acoustic channels in the Atlantic Ocean.

In section II, we describe the receiver design. In section III, we introduce a frequency domain equalizer based on cyclic prefix in the first part. In the second part, a frequency domain equalizer based on an overlapping method without CP transmission is presented using an adaptive algorithm. Some simulation results to make performance comparison between the approaches with and without CP are provided in section IV. Experimental results are presented in section V where a multiple input receiver scheme is considered. Finally, concluding remarks with some perspectives are given in the last section.

II. RECEIVER DESIGN

We consider an adaptive FDE for an all-digital single-carrier receiver in a time-varying channel environment. 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:

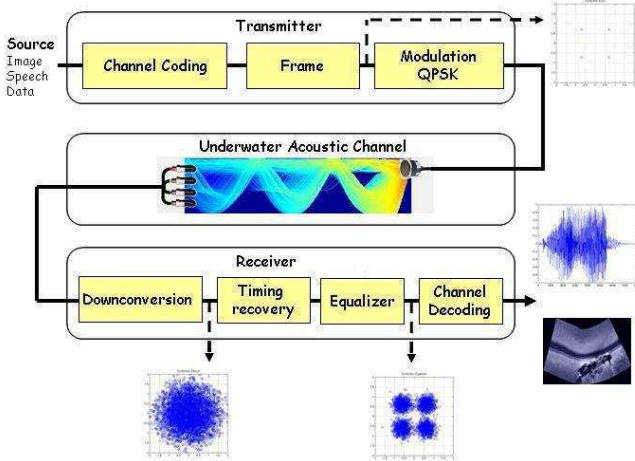$$s(t) = \Re e\{ \sum_{n=-\infty}^{+\infty} d_n g(t - nT) e^{j(2\pi f_c t + \psi)} \} \qquad (1)$$

Fig. 1. All-digital receiver

Fig. 2. multiple input receiver design

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 transmitted symbols with variance σ_d^2 and $g(t)$ 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 [6]. 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 downconversion is performed digitally and the optimum sampling is realized from a sample rate converter. The resulting all-digital receiver is depicted in Fig. 1.

In wide-band transmissions, such as underwater acoustic communications [7], [8], the Doppler effect introduces a scaling of the symbol period that must be taken into account in the design of the timing recovery scheme [9], [10]. 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 [9], [10]. Because the receiver is all-digital, the optimum sampling time does not necessary concur with an integer multiple of T_s and a sample rate conversion is thus required. The sample rate conversion uses interpolation, filtering and decimation [6].

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 [11] transmitted at the beginning of the transmission. 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 [12], [13].

After demodulation, sampling, Doppler compensation and timing recovery, the received signal is modeled as output of a discrete time channel where each output $j; j = 1, ..., N_R$ is corrupted by an additive, zero-mean, white Gaussian noise
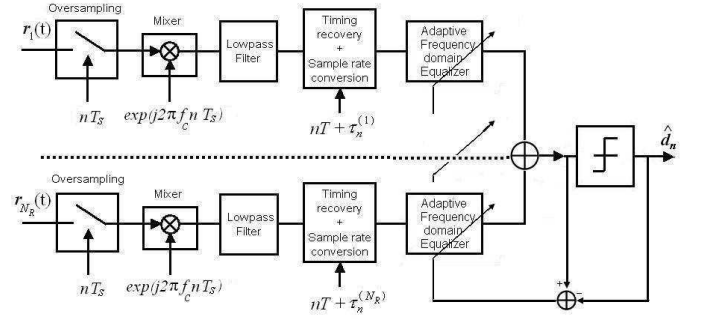
(AWGN) $w_n^{(j)}$ with variance $\sigma_{(j)}^2$. The n-th sample received by the j-th antenna (Fig. 2) can be written as follows,

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

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

III. FREQUENCY-DOMAIN EQUALIZER

Frequency-domain equalization is an interesting alternative to time-domain equalization for single carrier transmission systems. The equalization is performed over a block of N samples by multiplying term by term these N samples by the N coefficients of the equalizer using an efficient fast Fourier transform (FFT). This permits to obtain performance comparable to the TDE one and low computational complexity if the channel has a large delay spread compared to the symbol duration. Fig. 3 illustrates the principle of this equalizer.
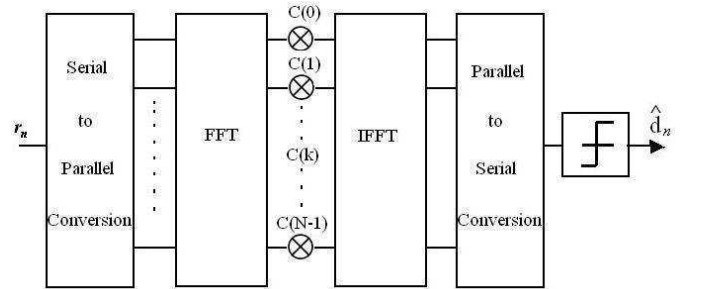

Fig. 3. Frequency-domain equalizer principle

A. Cyclic prefix FDE

At the transmitter, for each block of N symbols, a cyclic prefix that is the repetition of its last N_{CP} symbols is inserted at its beginning. This produces a circular convolution between the transmitted signal and the channel impulse response (CIR). To avoid inter block interference (IBI), the CP length must be chosen greater than or equal to the length of interference caused by the multipath time-varying channel ($N_{CP} \geq L$ [14]. At the receiver, the CP intervals are discarded and the observations are converted into blocks of N symbols.
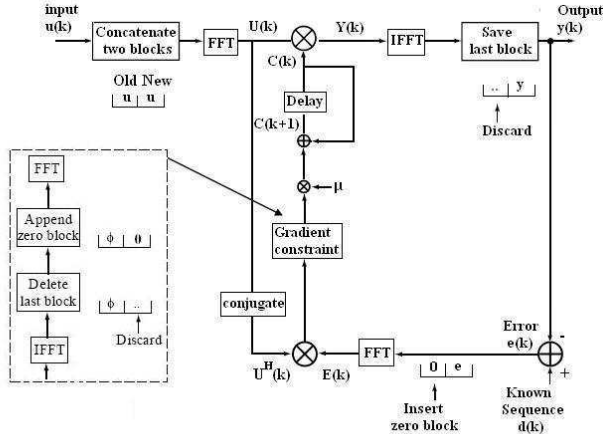
Fig. 4. OS AFDE equalizer

Thereafter, frequency domain equalization is proceeded such as in Fig. 3. In fact, the FFT processing is equivalent to circular convolution in the time-domain [5], that agrees with the circular convolution produced in the channel. On the other hand, the insertion of a cyclic prefix involves overhead and spectral efficiency loss. In addition, the block size is defined at the transmitter and there is no possibility to change it at the receiver. The channel is assumed to be invariant during the transmission of each data block. In next subsection, we apply a technique avoiding the CP insertion that allows achieving the same performance with some advantages.

B. Overlap-and-save adaptive FDE

In this case, the transmitted signal does not contain CP and thus, a linear convolution is performed with the CIR. The overlap-and-save adaptive FDE (OS-AFDE) is a fast and efficient implementation of the block least mean square (BLMS) time domain equalization in the frequency domain [15]. The BLMS TDE is a generalized form of the LMS TDE. The equalizer taps are constant over a block duration and are updated from a block to another. To implement the BLMS TDE in the frequency domain, the FFT operations are combined with other treatments to be equivalent to a linear convolution in the time domain. These treatments enable to section and overlap the data and then retain a subset of the resulting output that matches with the linear convolution ones. The demonstration of this equivalence is detailed in [16].

The principle of the OS-AFDE is summarized in Fig. 4. The input data r_n are converted into blocks of N symbols. So, according to the overlap-and-save principle, the k-th time domain input, denoted as $\mathbf{u}(k)$, contains N samples from the current block and N samples from the previous block.

$$\mathbf{u}(k) = [r_{kN-N}, ..., r_{kN-1}, r_{kN}, ..., r_{kN+N-1}] \quad (3)$$

$\mathbf{U}(k)$ denotes the $2N$-FFT of $\mathbf{u}(k)$. It can be written as follows,

$$\mathbf{U}(k) = \mathbf{F}\mathbf{u}(k) \quad (4)$$

Where $\mathbf{F}$ is the $2N$ Fourier matrix. Let consider $\mathbf{c}(k)$ the N time-domain equalizer taps; the corresponding $\mathbf{C}(k)$ equalizer coefficient vector in the frequency domain is augmented by N zeros in to obtain an equivalent $2N$ frequency domain equalizer coefficients.

$$\mathbf{C}(k) = \mathbf{F}[\mathbf{c}(k) \; \mathbf{0}_N] \quad (5)$$

The equalization is then performed by multiplying term-by-term the $2N$ input data $\mathbf{U}(k)$ by the $2N$ equalizer vector $\mathbf{C}(k)$. The equalizer output in the time domain $\mathbf{y}(k)$ is then obtained using the inverse of Fourier matrix as follows,

$$\mathbf{y}(k) = \text{Last } N \text{ terms of } \mathbf{F}^{-1}(\mathbf{U}(k)\mathbf{C}(k)) \quad (6)$$

Only the last N samples are retained because the first N samples are the results of a circular convolution. For more details, readers can refer to [5] and [16]. According to the adaptive algorithm, the equalizer coefficient vector is updated at each block iteration as follows,

$$\mathbf{C}(k+1) = \mathbf{C}(k) + \mu\mathbf{U}^H(k)\mathbf{E}(k) \quad (7)$$

where the superscript H denotes a Hermitian transpose, μ is the adaptation step and $\mathbf{E}(k)$ is the k-th frequency-domain error vector between the k-th equalizer output $\mathbf{y}(k)$ and either the k-th original transmitted symbol sequence $\mathbf{d}(k)$ for the training mode (8) or the k-th hard decision taken from the equalizer output $\hat{\mathbf{d}}(k)$ for the tracking mode (9), made in the time domain. The training mode is proceeded during M blocks. The time-domain error vector is augmented by N zeros to obtain an equivalent of $2N$ samples [5]. The adaptation step parameter μ depends on the convergence rate of the algorithm.

$$\mathbf{E}(k) = \mathbf{F}[\mathbf{0}_N \; (\mathbf{y}(k) - \mathbf{d}(k))] \quad (8)$$

$$\mathbf{E}(k) = \mathbf{F}[\mathbf{0}_N \; (\mathbf{y}(k) - \hat{\mathbf{d}}(k))] \quad (9)$$

The gradient constraint mentioned in Fig. 4 permits to perform a linear correlation between the input signal and the frequency-domain error and ensures that the algorithm be an exact implementation of the block adaptive TDE [5].

IV. SIMULATION RESULTS

We assess the performance 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 energy per data bit and N_0 is the noise variance. In Fig. 5 and Fig. 6, we compare the BER performance of CP-AFDE and OS-AFDE simulated over two time-invariant channel models: a medium frequency selective channel (Porat) and a deep frequency selective channel (Proakis B). The transmitted signal is QPSK modulated. We also compare these algorithms with the following equalizers: theoretical minimum mean square error (MMSE) TDE and a CP-FDE with perfect channel estimation. A theoretical BER for additive white gaussian noise channel (AWGN) is also included as a reference. The adaptation step parameter is set to $\mu = 10^{-3}$. The block length is about $N = 64$ samples and the cyclic prefix is set to $N_{CP} = 16$ for the CP-AFDE algorithm. The performance is evaluated for 400 received blocks and the
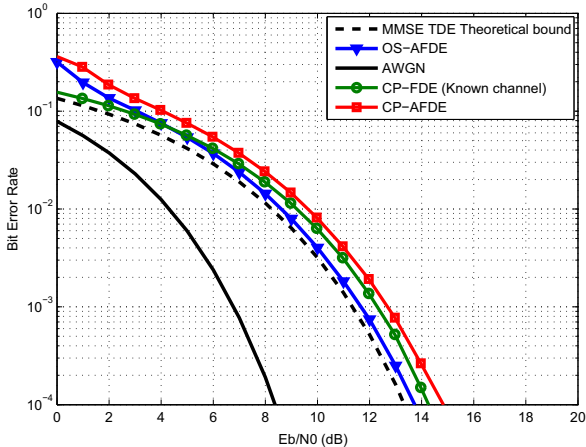
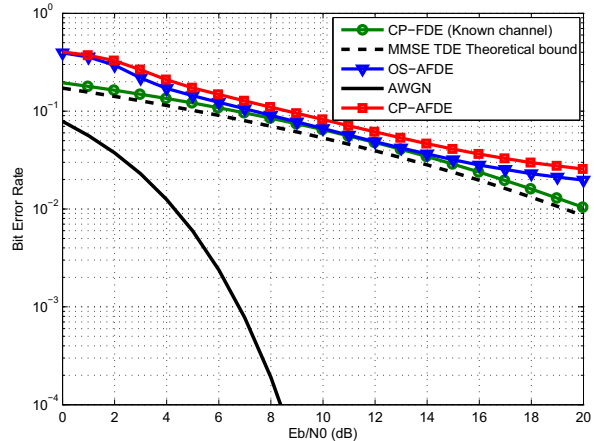Fig. 5. Bit error rate vs E_b/N_0 for Porat channel

Fig. 6. Bit error rate vs E_b/N_0 for Proakis B channel

BER is calculated over the tracking mode, without taking into account the $M = 80$ blocks of the training period. Note that the loss due the transmission of training sequence is not taken into account for the computation of the BER.

From Fig. 5, we see that a SNR of approximately 13.5 dB is required to achieve a BER of 10^{-4} for the OS-AFDE algorithm, not far from the theoretical limit (difference of about 0.4 dB). We also see a steady performance difference between OS-AFDE and CP-AFDE of about 1 dB for all SNR values. This difference results from the consideration of symbols transmitted during the guard period (cyclic prefix), which reduces the energy per bit that is available at the transmitter. Theoretically, there is a performance loss P_{loss} in decibel unit such as,

$$P_{loss} = 10log_{10}(N/(N_{CP} + N)) \qquad (10)$$

We also show that CP-FDE requires 0.5 dB more than the OS-AFDE to achieve the same BER of 10^{-4}. The reduced loss from 1 dB for CP-AFDE to 0.5 dB for the CP-FDE is due to the fact that we consider known channel response in CP-FDE when the OS-AFDE is an adaptive algorithm, which needs a training period to converge to the optimal equalizer coefficients.

For the deep selective channel model of Fig. 6, we achieve a BER of 3.10^{-2} with an SNR value of 15.5 dB, at about 1.3 dB from the theoretical bound. For the CP-AFDE, an increase in signal power of about 1 dB compared to the OS-AFDE is needed in order to reach the same BER.

The overhead resulting from CP becomes increasingly high when the channel dispersion is large compared to the symbol duration. Another advantage of the OS-AFDE technique is that the block length N is chosen at the receiver, unlike to the CP-AFDE algorithm, when the insertion of the CP forces to select the block length at the transmitter. We can imagine an additional component at the receiver for setting N depending on the channel variability during the block.

V. EXPERIMENTAL RESULTS

A. Experiment Setup

The OS-AFDE algorithm is tested 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 to 2 kilometers and the carrier frequency was equal to $f_c = 35KHz$. The coded bit rate was about $10kbits/s$. The block length was fixed to $N = 32$ samples and the adaptation step was set to $\mu = 3.10^{-3}$. The training period was performed during 1 second. Two experimental sea trials were provided to validate the proposed equalizer.

In the first one (experiment A), real-time sonar images were transmitted from autonomous underwater vehicle (AUV) to a ship over a distance of 1.5 kilometers as depicted in Fig. 7. The transmit power was set to 180 dB ref μ Pa at 1 meter and the relative transmitter to receiver velocity was about $v = 1.4$ m/s.

For the second one (experiment B), 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 relative transmitter to receiver velocity was about $v = 0.5$ m/s.

B. Adaptive Multiple input FDE

The structure of the equalizer is represented in Fig. 8. A multiple input receiver scheme is considered using four hydrophones equally spaced at 25 cm (see Fig. 9). Four second order phase locked loops (PLL) were optimized jointly with the equalizers in order to compensate for the residual frequency offsets in each input.

Performance are presented in terms of measured mean square error (MSE) without taking into account the training period. A comparison with an adaptive time-domain (LMS) equalizer, which is also tested in the same conditions using
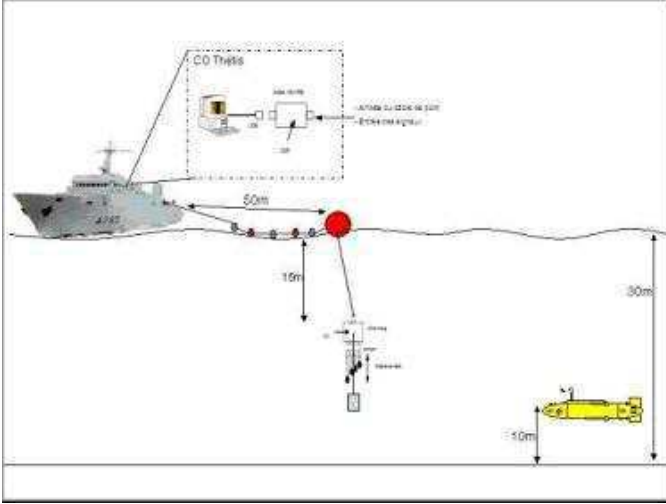
Fig. 7.   Experimental sea trial scheme
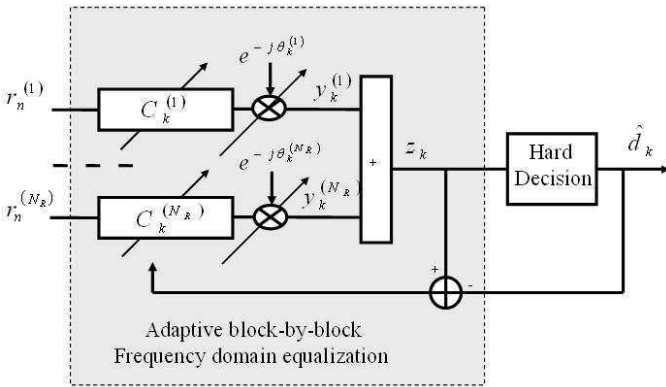

Fig. 8.   Multiple input receiver scheme for the OS-AFDE

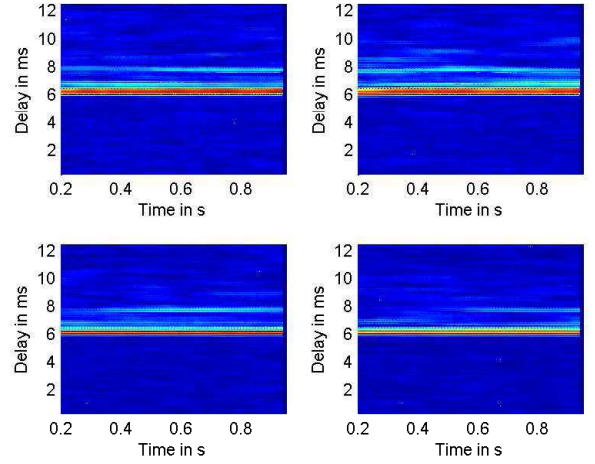Fig. 9.   The four hydrophones of the receiver

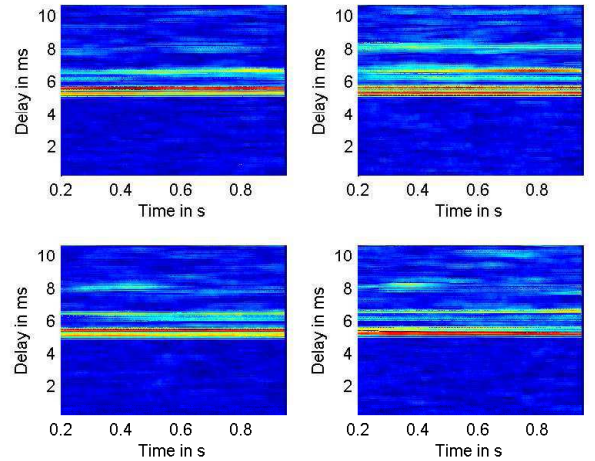Fig. 10.   Estimated UWA channel impulse response channel A

Fig. 11.   Estimated UWA channel impulse response channel B

the same receiver design, is made in order to validate the experimental result.

The estimated underwater acoustic channel impulse responses are given in Fig. 10 for the (experiment A) and in Fig. 11 for the (experiment B). Afterwards, the two channels are referred to channel A and channel B, respectively.

In Fig. 12, we show the MSE of the channel A. The OS-AFDE is almost equivalent to a classical adaptive time domain equalizer and achieving an MSE value of about -15 dB. For the channel B, which contains more multipath than channel A and a higher delay spread, the result is also satisfactory (see Fig. 13). A steady state is achieved at MSE near to -9 dB. The evaluated BER for the two experiments after the channel decoding was null over 10 seconds of transmission.

VI. CONCLUSION

An adaptive frequency domain equalizer have been simulated and tested over a real UWA time-varying channel.
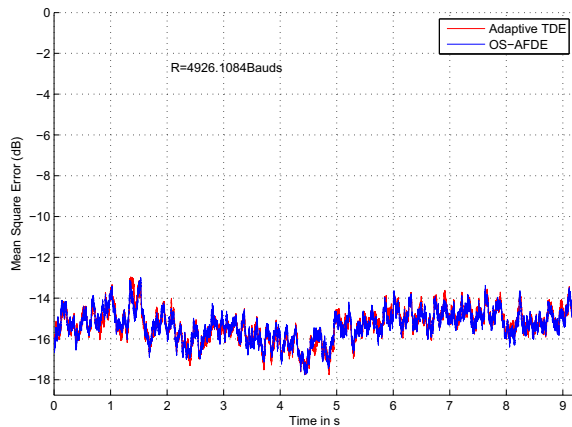
Fig. 12.   Mean square error, Channel A, $N_R = 4$, QPSK, 10 kbps, $d = 1.5$ kilometers
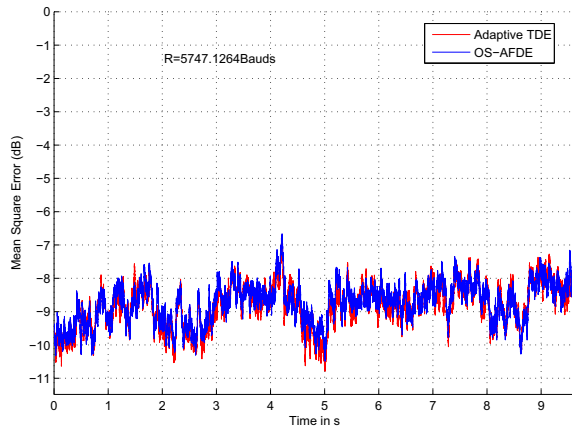

Fig. 13.   Mean square error, Channel B, $N_R = 4$, QPSK, 10 kbps, $d = 500$ meters

We have first shown through simulations the advantages of the OS-AFDE compared to the CP-AFDE. The OS-AFDE is then validated through real data transmission over an UWA channel. The performance is measured in terms of MSE and is compared to an adaptive time domain equalizer. To prevent the equalizer coefficient from residual frequency offsets, adaptive compensation is carried out on each input. A steady state of the MSE is achieved and is equivalent to adaptive TDE ones.

VII. ACKNOWLEDGMENT

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