Equalizer for Video Rate Transmission in Multipath Underwater Communications

Benoît Geller, Vittorio Capellano, Jean-Marc Brossier, Abderrahman Essebbar, and Geneviève Jourdain

Abstract—**This paper concerns the equalization problem of an underwater high rate transmission system. Because the channel delay spread of an horizontal link is large compared with the data duration, we have developed an adaptive equalizer in order to minimize the calculus burden of the moving receiver. The performance of this equalizer on synthetic and real channels is discussed.**

I. Introduction

UNDERWATER transmission systems have experienced an incredible growth in the recent years. Powerful algorithms were developed [1], [2] to cope with the time-variant dispersive constraints of the underwater acoustic channel. New needs like local area networks (LAN's) [3] or higher information rate [4] are technical challenges with a transmission channel such as the underwater one. A typical application of a high information rate link is a sea bottom imaging system on the continental shelf (depth less than 300 m) from an underwater vehicle to a surface vessel, at a typical range of one nautical mile [5]. This corresponds to a horizontal transmission with high multipath. The low rate (1 kb/s) down link from the surface ship has already been studied [6] but the problems of the intersymbol interference is even more crucial at the high rate (up to 150 kb/s) uplink from the underwater vehicle. To counter it, the combination [7] of an adaptive beamforming [8] and an adaptive equalizer [9], [10] was considered.

This article addresses the equalization process of the underwater high rate transmission system. The problems are as follows. The combination of the high data rate and the multipath due to the bottom and the surface reflections [5] can generate intersymbol interference up to 30 symbols long. Furthermore, the equalizer must handle slow variations as well as abrupt changes in the channel characteristics (the transmitter as well as the receiver are likely to move). An additional problem is that classical transversal filters need too many coefficients to update for real time transmission (over than 100 taps, see Section II). We have developed an algorithm in order to reduce this number of coefficients and achieve real-time processing. Section II describes how to choose this limited number of coefficients. Section III recalls how to update the equalizer coefficients adaptively and also how to cope with phase recovery. This equalizer is then tested over two sets of real data that were captured during experiments conducted at Blyth Harbour. Section IV describes the performance of such an equalizer with limited number of taps over real channels. Conclusions are finally summarized in Section V. This work was performed at the CEPHAG for the project AIDA of the EC research program MAST2.

II. A Transversal Structure

An equalizer is a well-known filter [9] used to reduce the effect of intersymbol interference. The need to choose a better equalizer structure than the classical transversal one is motivated by the real-time high-rate processing constraint; classical transversal filters need too many coefficients to update for real-time transmission: ordinarily, the number of taps required equals approximately four times the ratio of the channel delay spread divided by the data duration. The study of the channel impulse response then allows us to predict that one would need over than a hundred taps to update, each tap requiring at least two complex multiplications and additions for the easiest to implement adaptive algorithm: the stochastic gradient. Let us illustrate this numerically; the stochastic gradient algorithm is the smallest calculus burden algorithm as it requires $2N$ complex multiplications per cycle where N is the number of coefficients of the equalizer. The coefficients are updated with a classical gradient algorithm

$$H_k = H_{k-1} + \mu Y_k^* e_k, \tag{1}$$

where * stands for the complex conjugate, $Y_k = [y_k, \cdots, y_{k-N+1}]^T$ is the observation vector (at time k, denotes the transposition), H_k is the equalizer coefficients vector, e_k is the error between the equalizer output and the decided signal.

Still, for a video transmission rate of 150 kb/s and for the simplest two paths channel (17 is a typical channel reflected path delay, predicted with a ray tracing method [5]), the transfer function is

$$T(z) = 1 + Az^{-17}. \tag{2}$$

In that case one would make $2 * 4 * 17 * 150.10^3 = 21.10^6$ complex multiplications and additions per second for the very simplest algorithm. The implementation of a Viterbi algorithm would neither be possible for real-time processing with such an underwater channel because of the exponential increase of the calculus burden with the channel delay spread. Other

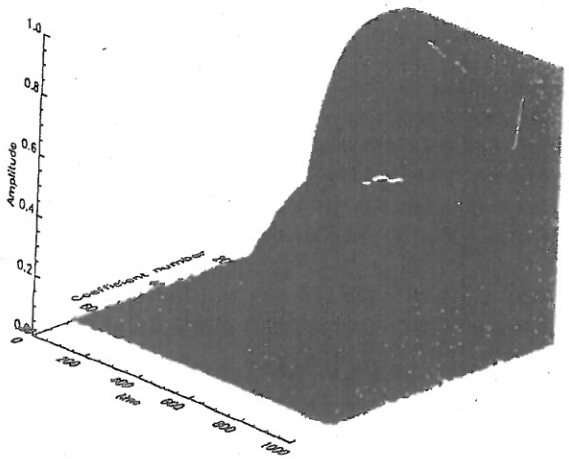


Fig. 1. Transversal coefficients for a two path channel.

filter structures than the transversal one have to be used with extreme care because there are often channel configurations that cause the equalizer transfer function to diverge [5]. Some decision feedback equalizers [14] have been used in a 6 kb/s horizontal transmission in a pool. We thus turn back to the transversal structure and try to reduce its number of coefficients to update.

We now see the principles underlying the fact that the number of taps of the transversal structure can be reduced. Let us take the simplest two paths channel described by (2). Although the number of taps to update is large, one readily sees that most of these coefficients are negligible (see Fig. 1).

In fact, the significant taps (peaks of Fig. 1) correspond decreasingly to the multiples of the channel delay i.e.,

$$H(z) = \sum_i h_i z^{-i} \approx h_{17}z^{-17} + h_{34}z^{-34} + h_{51}z^{-51} + h_{68}z^{-68}.$$

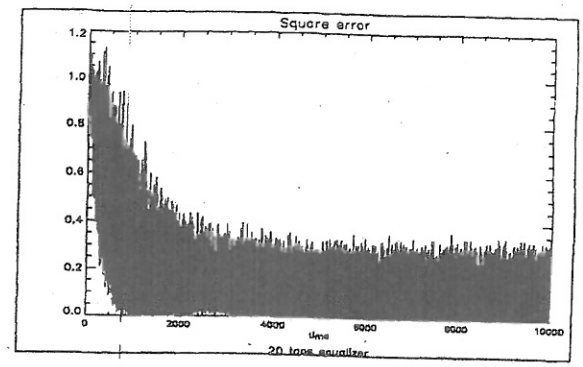
This is logical because the equalizer tries to insure (by assuming high SNR)

$$T(z)H(z) \approx 1 \tag{3}$$

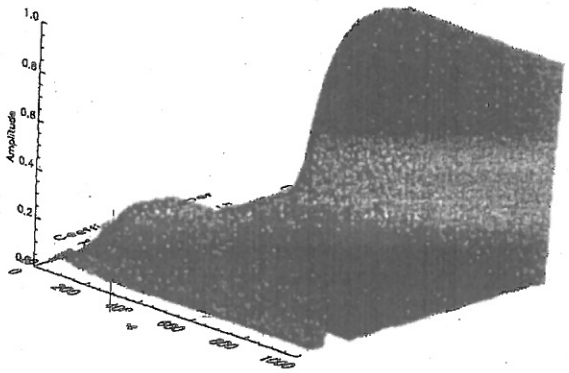
where $T(z)$ and $H(z)$ are polynomials.

Furthermore, if one does not take enough coefficients in his transversal filter [for instance $N = 20$ in the case of (2)], the error between the actual and the predicted output does not converge as shown by Fig. 2(a) because the degree of $H(z)$ is too small to obtain (3); but it does not diverge either, and still the coefficient corresponding to the channel delay is more important than the other ones [see h_{17} of Fig. 2(b) where one only takes 20 coefficients in the case described by (2)].

We thus propose the following procedure to make a sixteen coefficients transversal equalizer [4], [10]. We periodically (e.g., every T_{work} equaling 500 data durations) sweep across the coefficient indexes corresponding to possible channel delays (i.e., here 30-data duration). At the end of a time period T_{work}, we select, for instance, the two-largest transversal coefficients h_i and we decide that for the next T_{work} duration,



(a)



(b)

Fig. 2. (a) Error when N is too small. (b) Transversal coefficients for a three path channel.

we shall still update these largest weights but also coefficients whose indexes are multiples of the indexes of the largest coefficients indexes; (in the previous example, if h_{17} was the largest coefficient at time T_{work}, we shall update h_{17}, h_{34}, h_{51}, and h_{68} between time $(iT_{\text{work}}+1)$ and time $(i+1)T_{\text{work}}$). The remaining coefficients to be updated correspond to indexes within the possible channel delays, that were not updated during the last T_{work} duration. Consequently, if new propagation rays corresponding to those indexes appear at the receiver, or if there are path delay changes, the corresponding equalizer coefficients will rise and the equalizer is suited to nonstationarities.

The performance of this 16 coefficients transversal equalizer is displayed in Table I for a binary phase shift keying (BPSK) modulation. It works as well as a classical transversal equalizer (with, e.g., 70 coefficients). This means that one should have a signal-to-noise ratio equal at least to 15 dB in order to recover the data correctly.

In acoustic underwater transmission, one often has to cope with several paths. In the case of AIDA [5], one expects a direct path, a path reflected at the sea surface and one at the sea bottom. Nevertheless, the number of significant transversal coefficients is still small; effectively, in order to insure (3), the polynomial $H(z)$ needs significant coefficients at degrees

TABLE I
A ATTENUATION OF (2)

A (eq 2)	SNR (dB)	MSE (N=70)	MSE (N=16)
0.4	30	$5.0\ 10^{-3}$	$9.1\ 10^{-3}$
0.4	20	$3.7\ 10^{-2}$	$8.3\ 10^{-2}$
0.4	10	0.24	$9.5\ 10^{-2}$
0.4	3	0.57	0.21
0.8	30	$3.5\ 10^{-2}$	$5.2\ 10^{-2}$
0.8	20	0.1	0.1
0.8	10	0.35	0.15
0.8	3	0.76	0.23
0.99	30	$8\ 10^{-2}$	0.11
0.99	20	0.25	0.13
0.99	10	0.48	0.16
0.99	3	0.8	0.27

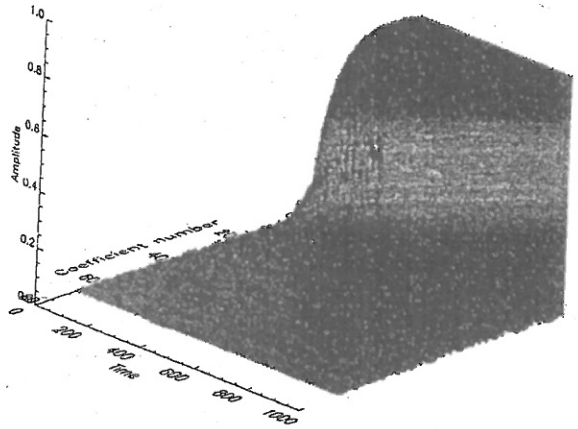


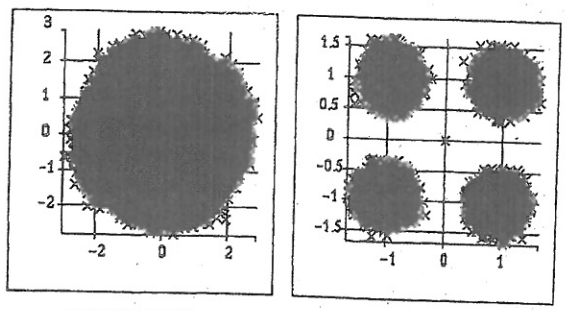
Fig. 3. Transversal coefficients for a three paths channel.

multiples of the channel path delays (in data duration) and at degrees that are sums of the previous degrees (multiplication of two polynomials)(see Fig. 3). This 16-weight transversal filter has been tested in the case of a three path channel as well as real chann els and we shall display that it works as well as the classical transversal filter.

A way of achieving higher information rate is to use a multilevel modulation. The 16-weight equalizer has also been tested in the case of a 4-PSK signal transmitted over a three paths channel. As shown by Fig. 4, the original data is perfectly recovered as well.

III. EQUALIZATION, PHASE RECOVERY, AND SELF-OPTIMIZED ALGORITHM

An additional trouble is the phase recovery of the received signal, particularly in the case of a Doppler drift due for instance to the relative motion between the transmitter and the receiver. If no attempt to recover the phase is made, the previous equalizer cannot work properly. As a Costas loop does not work well in our multipath context [5], we have used an equalizer that jointly tries to cancel the intersymbol interference and provides a phase estimator [11]. This transversal



RECEIVED SIGNAL (a) EQUALIZER OUTPUT (b)

Fig. 4. 4-PSK constellations (a) before and (b) after equalization.

filter builds the following output

$$c_k = Y_k^T H_{k-1} e^{-i\varphi_{k-1}}$$

where $Y_k = [y_k, \cdots, y_{k-N+1}]^T$ is always the observation vector (at time k), H_k is the equalizer coefficients vector (only 16 selected weights among the N coordinates are nonzero), and φ_k is the receiver phase.

The filter coefficients are updated in order to minimize the mean-square error (mse)

$$J(H, \varphi) = E\{|e_k|^2\}$$

where $e_k = a_k - c_k(H, \varphi)$, c_k is the equalizer output, and a_k is the decided symbol at the output of the equalizer.

Once more, the stochastic gradient is able to reach the minimum of J. The derivation of J with respect to H and φ leads to the following gradients:

$$\begin{cases} \nabla_H |e_k|^2 = -2Y_k^* e^{i\varphi} e_k \\ \nabla_\varphi |e_k|^2 = -2\mathrm{Im}(e_k^* c_k) \end{cases}.$$

This leads to an algorithm that jointly estimates the 16 coefficients of the equalizer and the phase

$$\begin{cases} H_k = H_{k-1} + \mu Y_k^* e^{i\varphi_{k-1}} e_k \\ \varphi_k = \varphi_{k-1} + \gamma \mathrm{Im}(e_k^* c_k) \end{cases} \tag{4}$$

with $e_k = a_k - Y_k^T H_{k-1} e^{-i\varphi_{k-1}}$.

This joint Algorithm (4) has been tested over multipath channels experiencing a global Doppler drift like the following:

$$T(z) = (1 + Az^{-17} + Bz^{-28}) e^{-i\phi_k}$$

where ϕ_k represents a linear phase drift. For phase drifts up to 10^{-3} rad/symbol, such a joint equalizer gives results with no significant difference with Table I. It has also been tested over real channels, as we shall see in Section IV.

Finally, in a nonstationary context and when no *a priori* knowledge is available about the channel stationarities, the practical determination of the stepsizes γ and μ can be difficult. Self-adaptivity of those stepsizes is thus a highly desirable property [12], [13].

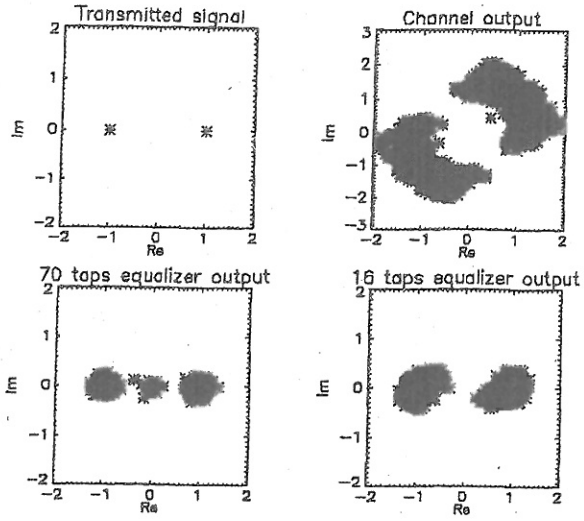


Fig. 5. Blyth3 constellations (over 10 000 symbols).

Let us describe an optimization scheme for the stepsizes μ and γ.

The steady-state (asymptotic) mse depends on the two stepsizes μ and γ, it can be written

$$J(\mu,\gamma) = \lim_{k\to\infty} E\left|a_k - Y_k^T H_{k-1} e^{-i\varphi_{k-1}}\right|^2. \tag{5}$$

This mse is evaluated on the algorithm trajectory. Our aim is to minimize (5) with respect to μ and γ under the constraints given by (4). Equation (4) define a relationship between φ_k and y_k (γ being constant) and a relationship between H_k and y_k (μ being constant). Let us use the notations

$$J(Y_k, a_k, H_{k-1}, \varphi_{k-1}; \mu, \gamma) = \left|a_k - Y_k^T H_{k-1} e^{-i\varphi_{k-1}}\right|^2$$

$$G_k = \frac{\partial H_k}{\partial \mu} \quad \text{and} \quad F_k = \frac{\partial \varphi_k}{\partial \gamma}.$$

The stochastic scheme for stepsizes optimization is

$$\gamma_k = \gamma_{k-1} - \alpha \frac{\partial J}{\partial \gamma} \tag{6}$$

$$\mu_k = \mu_{k-1} - \beta \frac{\partial J}{\partial \mu}. \tag{7}$$

The phase of filter H_k is defined using a constraint on the phase of one of the taps of vector H_k; this constraint enables us to assume that the phase estimation and the estimated equalizer are uncoupled. So we assume that H_k does not depend on γ and that φ_k does not depend on μ. Denoting ξ the variable being equal to μ or γ, the derivative of J with respect to ξ is given by

$$-\frac{1}{2}\frac{\partial J}{\partial \xi} = \mathrm{Re}\left(Y_k^T \frac{\partial H}{\partial \xi} e^{-i\varphi} e_k^*\right) + \mathrm{Im}(a_k^* c_k)\frac{\partial \varphi}{\partial \xi}.$$

Thus, using a constraint on the maximum and on the minimum value of $\gamma(\gamma_{\max}, \gamma_{\min})$, and $\mu(\mu_{\max}, \mu_{\min})$, the

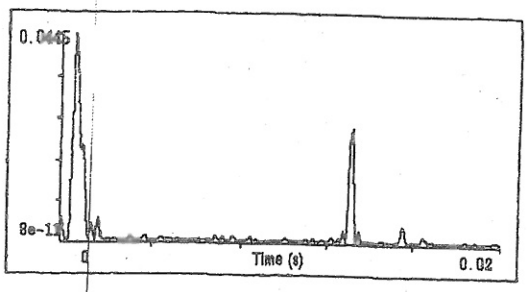


Fig. 6. Impulse response of the Blyth5 channel.

stepsizes are updated according to

$$\gamma_k = \left[\gamma_{k-1} - \alpha F_{k-1}\mathrm{Im}(a_k^* c_k)\right]_{\gamma_{\min}}^{\gamma_{\max}} \tag{8}$$

$$\mu_k = \left[\mu_{k-1} - \beta \mathrm{Re}\left(Y_k^T G_{k-1} e^{-i\varphi_{k-1}} e_k^*\right)\right]_{\mu_{\min}}^{\mu_{\max}} \tag{9}$$

with the notation

$$[x]_a^b = \begin{cases} x, & \text{for } x \in [a,b] \\ a, & \text{for } x < a \\ b, & \text{for } x > b \end{cases}. \tag{10}$$

According to our practical experiments, the "second-order" stepsizes α and β can be chosen within a very large range without loss of performance. In practice, constraints like (10) are to be used on γ and μ in order to avoid instabilities.

To update the derivatives G_k and F_k we derive (4) with respect to μ and γ [15]. This leads to the algorithm described by (11)–(18) which minimizes the mse with respect to γ and μ. It is given by

$$c_k = Y_k^T H_{k-1} e^{-i\varphi_{k-1}} \tag{11}$$

$$e_k = a_k - c_k \tag{12}$$

$$\varphi_k = \varphi_{k-1} + \gamma_{k-1}\mathrm{Im}(e_k^* c_k) \tag{13}$$

$$\gamma_k = \left[\gamma_{k-1} - \alpha F_{k-1}\mathrm{Im}(a_k^* c_k)\right]_{\gamma_{\min}}^{\gamma_{\max}} \tag{14}$$

$$F_k = \mathrm{Im}(a_k^* c_k) + (1 - \gamma_k \mathrm{Re}(a_k^* c_k))F_{k-1} \tag{15}$$

$$H_k = H_{k-1} + \mu_{k-1} Y_k^* e^{i\varphi_{k-1}} e_k \tag{16}$$

$$\mu_k = \left[\mu_{k-1} - \beta \mathrm{Re}\left(Y_k^T G_{k-1} e^{-i\varphi_{k-1}} e_k^*\right)\right]_{\mu_{\min}}^{\mu_{\max}} \tag{17}$$

$$G_k = \left(I - \mu_k Y_k^* Y_k^T\right) G_{k-1} + Y_k^* e^{i\varphi_k} e_k. \tag{18}$$

In fact, this algorithm exhibits good performance in a lot of practical contexts. A theoretical study of this kind of adaptive scheme is given by Kushner in [15]. Sometimes, only one of the two stepsizes needs to be adapted, the other one being chosen before running the algorithm. In such a situation, one can omit some of the previous equations [(14), (15), or (17), (18)].

IV. Experimental Results

An adaptive 16-tap equalizer performing jointly a phase estimation was tested at Blyth Harbour. For the Blyth3 experiment, the transmitter and the receiver were put 1 and 5 m deep, respectively, at a distance of 300 m, whereas for the Blyth5 experiment they were put 10 m deep at a distance of 100 m. A 50-kHz frequency carrier was modulated by a 10 kb/s BPSK signal. The direct path signal-to-noise ratio was approximately

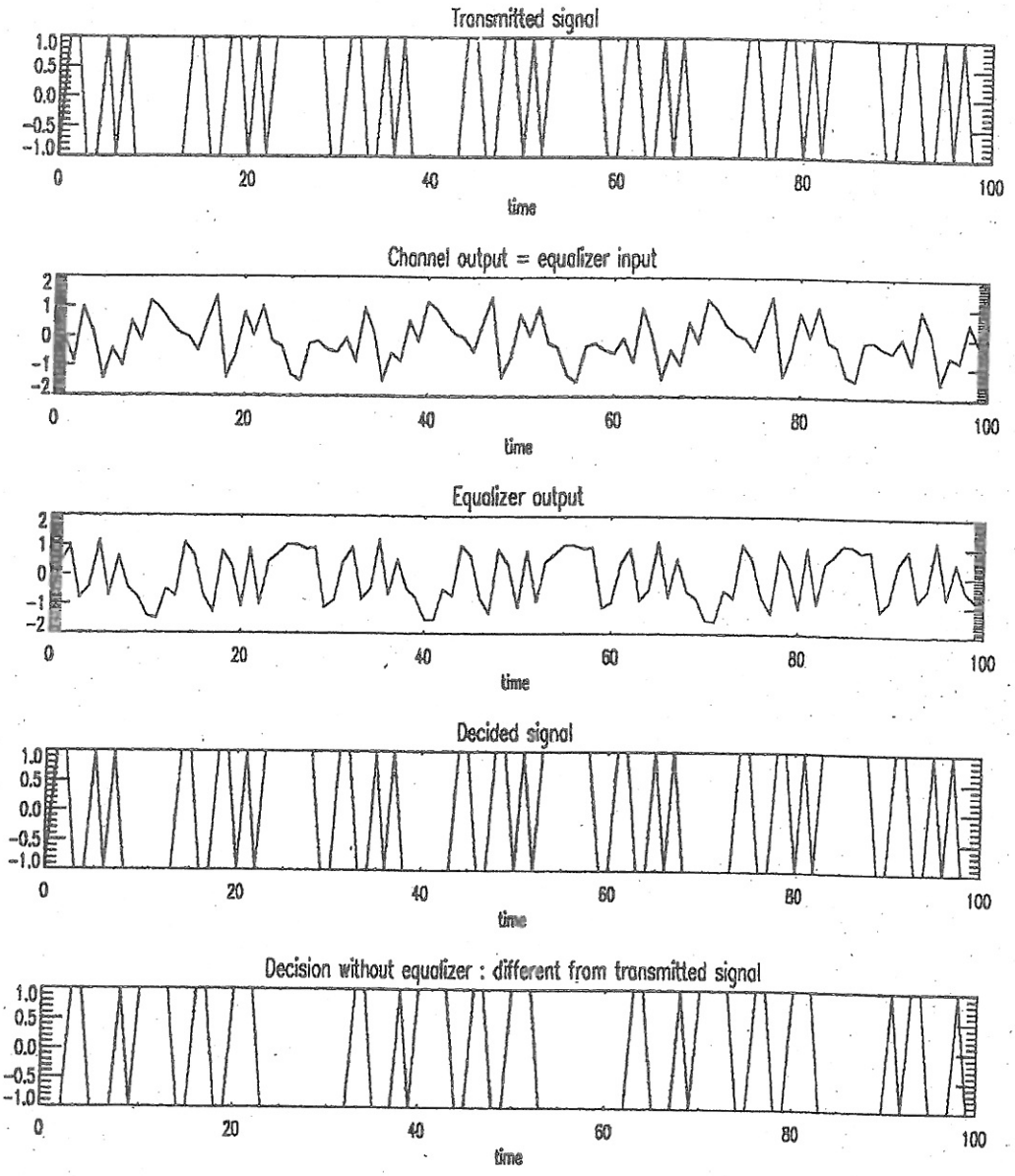


Fig. 7. Blyth5 signals.

40 dB. We display here two typical results (Blyth3 and Blyth5). In both experiences, the impulse response is characterized by two main paths: a direct one and a surface reflected one with a delay of approximately 13-data duration (1.3 ms). This is typically the kind of impulse response expected for the high rate transmission near the continental shelf [5].

Furthermore, in the case of Blyth3, a BPSK modulated carrier was sent through a two paths channel experiencing a global Doppler drift (see the $\pi/2$ rotation over 10 000 received data displayed on the right upper quarter of Fig. 5). This signal was processed by an ordinary transversal filter and by a limited number of coefficients transversal filter (both equalizers jointly performed the phase recovery). The 16-tap equalizer made absolutely no error as illustrated by the lower right quarter of Fig. 5, where the dots around the transmitted $+1$ and -1 are distinctly separated. One also remarks that an ordinary transversal equalizer with all of its 70 coefficients made a few errors leading to a poorer bit error rate (BER).

During the Blyth5 experiment, the direct path was composed of two contributions, probably due to a reflection near the receiving antenna. The corresponding impulse response is displayed on Fig. 6 where one can remark the spread of the direct path due to the two contributions. The signals processed by the sixteen taps equalizer are displayed on Fig. 7. The 16-tap

equalizer correctly chose the largest coefficients of the ordinary transversal equalizer. Both equalizers made approximately 100 errors over 70 000 data. Without equalization, the BER would have been of 0.5. This clearly shows the need for an equalizer at the receiving end. BER of magnitude 10^{-3} are sufficient for video applications. Let us remark that such a BER allows the use of an error correcting code that could then improve even more the link quality of the fading channel [9].

V. Conclusion

An equalizer for high data rate underwater transmission was presented. It is fully adaptive and jointly recovers the phase. Furthermore, it is devoted to real-time processing thanks to its small number of updating taps, and allows video short-distance transmission. This equalizer was tested successfully over multipath underwater channels. Combined with an adaptive beamforming, this "sparse" equalizer should permit the realization of high rate local area networks.

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Benoît Geller was born in Paris in 1965. He received the engineer degree from ENSERB (Ecole Nationale Supérieure d'Electronique et de Radioélectricité de Bordeaux) in 1989, and the Ph.D. degree in signal processing from Institut National Polytechnique de Grenoble (INPG) in 1992.

He is with the University Paris XII as Maître de Conférences in Telecommunications. His research activities are digital communications and error correcting codes.

Vittorio Capellano received the engineer degree from ENSIEG (Ecole Nationale Supérieure des Ingénieurs Electriciens de Grenoble) in 1994, and is pursuing the Ph.D. degree at Institut National Polytechnique de Grenoble (INPG).

His research interests include adaptive algorithms and array processing applied to underwater communications.

Jean-Marc Brossier received the Ph.D. degree in signal processing from Institut National Polytechnique de Grenoble (INPG) in 1992.

He is with Ecole Nationale Supérieure des Ingénieurs Electriciens de Grenoble (ENSIEG) as Maître de Conférences in Signal Processing. His research activities are adaptive algorithms, array processing and higher order statistics with application to digital communications and Geophysics.

Abderrahman Essebbar was born in Morocco in 1964. He received the Ph.D. degree in signal processing from Institut National Polytechnique de Grenoble (INPG) in 1992 and is working as a postdoc in the CEPHAG Laboratory.

His research interests include digital communications and array processing with application in seismic and underwater acoustics.

Geneviève Jourdain (SM'93) received the engineer degree from ENSERG (Ecole Nationale Supérieure d'Electronique et de Radioélectricité de Grenoble) in 1966, and the Docteur es sciences degree in 1976 in signal processing.

She is a Professor at INPG (Institut National Polytechnique de Grenoble), where her teaching activities concern signal processing in a broad sense. Since 1989, she has been Head of CEPHAG (Centre d'Etude des Phénomènes Aléatoires et Géophysiques), a laboratory which studies signal processing, especially with the French Navy, for 25 years. Her research activities are active time and frequency characterization of targets and propagation channels, time delay estimation, optimal receivers for detection and communication through dispersive channels, underwater communications, and adaptive equalization.

Dr. Jourdain is involved in many Signal Processing Groups (GRETSI, GDR TSI, CNRS Commity, DRED, etc.).