Blind Receivers for Underwater Acoustic Communication: An Example of a Non-Co-operative Bistatic Transmission System

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ABSTRACT

Underwater acoustic (UWA) communication systems represent a special class of submersed bistatic transmission systems, where the receiver is designed to recover the information that have been transmitted by a spatially separated transmitter. Although this task is in contrast to bistatic sonar applications, that aim to estimate certain channel features, the basic system setups are akin to each other and the transmission conditions are identical. Therefore, signal processing strategies that have been developed in one field might be inspiring for the other. In this paper a blind multichannel receiver for UWA communication is introduced that does not rely on any assistance provided by the transmitter. Furthermore, an extension of the proposed structure will be given which allows the inclusion of possibly available information of the channel physics to enhance the receivers' robustness with respect to the acoustic propagation conditions within the medium.

1.0 INTRODUCTION

Underwater acoustic links enable wireles and hence spatial flexible data exchanges that are for example needed for data retrieval from remotely operating autonomous underwater vehicles, for ocean monitoring, and for communication between submerged systems. The increasing need for transmitting large amounts of data in short periods of time over distances up to several kilometers has attracted considerable research work during the past decades to enable high-bit rate untethered data transmission. However, the desired high data rates are in contrast to the transmission conditions induced by the underwater telemetry channel, which generally exhibits time-varying multipath propagation leading to a dispersion of the transmitted signal in time. Movements of transmitter and receiver or of the medium itself cause Doppler shifts leading to an additional dispersion of the received signal in frequency domain. Furthermore, the available bandwidth is severly limited due to frequency dependent attenuation of the underwater medium. Therefore, phase-coherent receivers should be employed, since they allow for an efficient use of the available bandwidth. In the first half of the 1990s Stojanovic et. al. proposed a coherent receiver structure that can cope with the afore mentioned distortions of the transmitted signal [18],[19]. It employs joint synchronization and adaptive channel equalization together with proper training, i.e. transmission of data sequences that are a priori known to the receiver. However, in a highly time-varying environment the training procedure has to be performed periodically, which can reduce the effective data throughput considerably. In contrast to this, so called blind receiver structures do not rely on any explicit co-operation by the transmitter, offering the potential of a more efficient use of the maximum available data rate. It

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furthermore allows for a non-co-operative interception of acoustic data links. If knowledge of the channel physics is available, it can be incorporated into the receiver structure to improve the performance of such blind receivers [24], [16], [17].

The paper is organized as follows. After a short introduction to the properties of underwater acoustic channels and a discussion of their impact on the communication problem, a spatio-temporal receiver architecture that allows for timing estimation, carrier-phase recovery, and mitigation of multipath induced intersymbol interference (ISI) without explicit training is proposed. For ISI compensation linear filters will be employed with their coefficients being adapted to the transmission conditions using the constant modulus algorithm (CMA). To improve the receivers performance, model-based pre-processing steps will be introduced that utilize knowledge of the oceanographic parameters of the surrounding sea area. The proposed blind receiver is then verified by processing experimental communication data gathered in the North Sea during the ROBLINKS 1999 experiments [22]. The obtained results proof the applicability of the developed structure, and show the possible performance enhancement using additional information of the channel physics.

2.0 UNDERWATER ACOUSTIC CHANNEL

Underwater channels act like an acoustic wave guide with its boundaries given by the sea surface and the bottom. Although the propagation of acoustic waves through the underwater medium has already been discussed in detail in standard literature like [21], [7], and [3], the transmission conditions and effects that are most adverse to UWA communications will be summarized in the following.

1.1 Refraction of acoustic waves

The ocean itself is generally an inhomogeneous medium, where the speed of sound c_w in the water layer varies basically with depth. Since sound waves are refracted to regions with slower speed of sound their propagation in the underwater channel is determined by the depth dependence of c_w, which is termed as the sound speed profile (SSP). As a result, the formation of caustic zones, where the acoustic field is bundled, and acoustic shadow zones, where no sound energy can be detected, is possible along the vertical water column for a fixed horizontal distance between transmitter and receiver.

If the propagation conditions are unknown a priori or subject to rapid change it is beneficial to use an array of receiving hydrophones that sample the water column vertically, and to process the sensor outputs jointly in a suitable way.

1.2 Reflection of acoustic waves

Besides the previously described refraction, sound waves emitted by the transmitter can also be reflected at the surface and the seafloor, where -especially in shallow water environments- several bounces between the boundaries can occur. Therefore, several propagation paths can exist between the transmitter and the receiving sensors, introducing different time delays to the transmitted signal.

For transmission of digital signals using pulse amplitude modulation the dispersion in time domain leads to an overlap of successive transmitted pulses at the receiver positions which is known as intersymbol interference (ISI). To combat the multipath induced ISI a so called equalization step has to be implemented in order to restore the originally transmitted information from the received data.

1.3 Propagation losses

During their propagation through the underwater medium acoustic waves are attenuated due to geometrical spreading losses as well as due to absorption effects that lead to an attenuation, which is increasing with frequency. Therefore, either the maximum bridgeable distance between the transmitter and the receiver is restricted for a given bandwidth of the transmit signal, or the usable bandwidth is limited by the intended transmission distance. In [6] typical examples of useful spectral widths of the transmit signal are given in dependence on the assumed desired distance.

Since in digital communications the available data rate is determined by the signal bandwidth [13], the afore mentioned bandwidth limitations translate into according restrictions of the data throughput for UWA communication systems.

1.4 Doppler

Due to the slow propagation velocity c_w of acoustic waves within the water medium, relative motion between transmitter and receiver as well as motion of the medium itself can cause non negligible Doppler shifts. For a typical value of c_w =1500 m/s this frequency deviation will be in the order of 0.35 Hz / (kn kHz) [1].

Frequency shifts of a signal in passband domain translate into phase shifts in the corresponding baseband representation of the signal. In case of modulation schemes that encode the information in the absolute phase of the transmit signal the receiver must be able to compensate frequency and/or phase deviations.

1.5 Noise

Underwater acoustic communication links can be affected by different noise sources in the water medium. Man-made noise, e.g. caused by shipping traffic, occurs predominantly in coastal areas. It is usually of low frequency and often localizable. Therefore, these disturbances can be suppressed by proper spatial and/or spectral filtering. In contrast to this, interferences of natural cause, e.g. induced by atmospheric influences like impact of rain and wind on the sea surface, usually exhibit a broad spectrum and can be received from almost any direction. However, the spectral power density of this kind of noise is approximately decreasing with 20 dB per decade within the frequency band that is of technical interest for UWA communication [6], [7].

In case of long distance communication and/or high sea states the signal-to-noise ratio (SNR) at the receiver can be low, making the signal-detection and -decoding difficult. Provided that noise contributions are independent between the sensors, the problem of low SNR situations can be enhanced by using a multichannel receiver.

3.0 BLIND RECEIVER

The limited available bandwidth of underwater acoustic channels motivated the development of phase-coherent transmission systems for UWA communication, since they allow for an efficient spectral use of the channel [9]. However, the needed receivers must be able to compensate the dispersion of the transmit signal in time and frequency domain by employing suitable signal processing steps. In 1993 Stojanovic et. al. proposed a coherent receiver structure that can cope with the channel induced distortion of the transmitted signal by performing joint synchronization and equalization [18]. The adaptation of the receiver parameters to current channel conditions was realized by evaluating training data, i.e. data sequences that are known a priori to the receiver. Due to possibly rapid time varying channel conditions the training process has to be accomplished periodically, hence reducing the effective data throughput. To bypass this burden so called blind signal processing methods that do not rely on knowledge about the transmitted information can be implemented for adaptation of the receivers subsystems. In this paper we propose a digital multichannel receiver structure that utilizes untrained adaptive algorithms for timing recovery, equalization of channel induced signal distortions, and compensation of phase deviations.

3.1 Multichannel Receiver Structure

3.1.1 Received Signal

In the following we consider a receiver that jointly processes the output of C receiving sensors in order to recover the information bearing symbol sequence a_n that has been transmitted using a linear modulation scheme. On the transmitters side, the symbols are drawn every T seconds from a finite alphabet, and are modeled as independent and identically distributed random variables. The resulting time continuous baseband signal received by the cth sensor of the array is then given by

$$r_c(t) = \sum_{n=-\infty}^{\infty} a_n h_c(t - nT - \varepsilon_c T)\, \mathrm{e}^{j\varphi_c(t)} + w_c(t)\,, \quad c = 1, \ldots, C \tag{1}$$

where $h_c(t)$ represents the impulse response of the transmission channel between the transmitter and the cth sensor, including the influence of transmit filters. This possibly time varying channel can be modeled as a cascade of a system with impulse response $h_{c,\mathrm{det}}(t)$, that is considering the deterministic transmission behaviour, and a system with impulse response $h_{c,\mathrm{stoch}}(t)$ representing stochastic fluctuations, yielding

$$h_c(t) = h_{c,\mathrm{det}}(t) * h_{c,\mathrm{stoch}}(t)\,. \tag{2}$$

Since the local clocks of transmitter and receiver are operating asynchronously, an offset $\varepsilon_c T$ between the timing grid of the transmitter and the receiver has to be taken into account. The phase term $\varphi_c(t)$ reflects phase offsets that can for example be caused by possible deviations between the carrier of the received signal and the mixing frequency used in the receiver front-end. Finally, $w_c(t)$ represents additive noise which is assumed to be independent of the symbol sequence a_n.

3.1.2 Time Discrete Baseband Multichannel Receiver

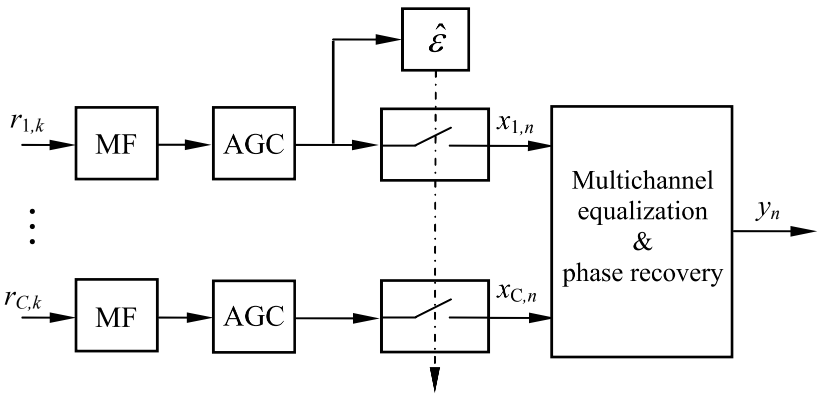

Figure 1: Multichannel receiver (front end, timing synchronization, and equalization)

Figure 1 shows the time discrete realization of the baseband processing part of the proposed receiver structure. The input series $r_{c,k}$ in the cth channel arise from sampling the signal (1) with Q times the underlying symbol rate $1/T$, i.e.

$$r_{c,k} = r_c\left(k\frac{T}{Q}\right). \tag{3}$$

At first, $r_{c,k}$ is applied to a filter matched to the transmit pulse to enhance the signal-to-noise ratio. Following this step, an automatic gain control (AGC) is utilized to ensure the same mean power level in all C channels. Although this is not strictly necessary, it can improve the succeeding signal processing steps. Therefore, a modified version of the AGC proposed in [10] was implemented in each physical channel of the proposed receiver.

For compensation of the misalignment between the transmitters and receivers timing grid, the resulting AGC output $z_{c,k} = z_c(kT/Q)$ is applied to an asynchronous sampling rate conversion, which transforms the *asynchronously* sampled signal into a signal $x_{c,n} = x(nT/P)$, which is *synchronously* oversampled by a factor P with respect to the nominal symbol rate $1/T$. In the considered receiver structure, this is realized by applying $z_{c,k}$ to interpolation and decimation steps that are adapted based on estimates $\hat{\varepsilon}_c$ of the timing error. Due to the assumed non co-operation of the transmitter, this property has to be derived from $z_{c,k}$ itself. Since neither the multipath influence nor phase deviations have been compensated at this stage, the utilized estimation procedure has to be robust against these kinds of signal distortions. Suitable methods can be found in [11], which have successfully been adapted to the proposed receiver structure for processing offset quadriphase shift keying (OQPSK) signals [23] and minimum shift keying (MSK) signals [25]. Since the implementation details of the timing synchronization are beyond the scope of this paper, the reader is referred to the according publications. If the timing errors ε_c can expected to be of same magnitude for all channels, it is sufficient to derive the estimate $\hat{\varepsilon}_c$ from only one channel, and to use it for timing correction in all channels. Finally, the C P-times oversampled signals $x_{c,k}$ are used as inputs to a signal processing unit which jointly performs multichannel equalization and phase-error compensation. The corresponding processing steps are adapted blindly as will be explained in section 3.3.

3.2 Signal Model

3.2.1 Time discrete signal model

In this section we define the simplified time discrete signal model that is utilized throughout the paper. All signals are considered in baseband domain, whereas perfect timing synchronization is assumed. Although the implemented receiver operates on data that are sampled with a multiple of the nominal symbol rate $1/T$, the processing chain will be considered completely in its baud-spaced representation since this allows for a compact description of the signal relations. To further ease the comprehension, the treatment of the signal relations is at first restricted to the problem of "single input-single output" transmission.

Let a_n represent the transmitted sequence of information bearing symbols that has already been introduced in section 3.1.1, while the channel between the transmitter and the cth sensor is described by a linear system which is assumed to have an impulse response $h_{c,n}$ of finite length M. According to (1), disturbances of the channel output are considered by an additive noise process $w_{c,n}$, yielding

$$x_{c,n} = a_n * h_{c,n} + w_{c,n} = \sum_{l=0}^{M-1} h_{c,l}a_{n-l} + w_{c,n} \tag{4}$$

as the signal received by the cth channel.

3.2.2 Matrix-Vector Notation

For a corresponding representation of the signal model in matrix-vector notation let $\mathbf{a}_n = \left(a_n, \ldots, a_{n-(L+M-2)}\right)'$ be a vector consisting of the $L + M - 1$ recent samples of the transmit symbol sequence, while $\mathbf{w}_{c,n} = \left(w_{c,n}, \ldots, w_{c,n-(L-1)}\right)'$ and $\mathbf{x}_{c,n} = \left(x_{c,n}, \ldots, x_{c,n-(L-1)}\right)'$ contain L samples of the received signal $x_{c,n}$ and the noise disturbance $w_{c,n}$ respectively. The signal relations in (4) can then be described as

$$\mathbf{x}_{c,n} = \mathbf{H}_c \mathbf{a}_n + \mathbf{w}_{c,n}, \tag{5}$$

where

$$\mathbf{H}_c = \begin{pmatrix} h_{c,0} & \cdots & \cdots & h_{c,M-1} & 0 & \cdots & 0 \\ & \ddots & & & \ddots & & \\ & & \ddots & & & \ddots & \\ 0 & \cdots & 0 & h_{c,0} & \cdots & \cdots & h_{c,M-1} \end{pmatrix} \tag{6}$$

represents the $L \times (L + M - 1)$ convolution matrix of the cth channel.

3.3 Blind Multichannel Equalization Using the Constant Modulus Algorithm

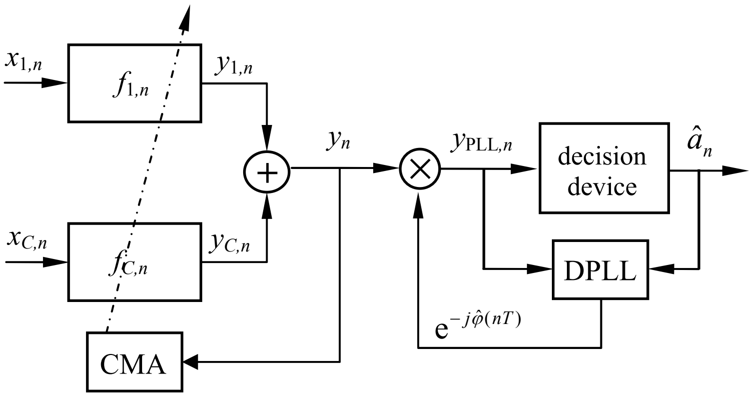

Figure 2: Linear multichannel equalization and phase correction

In order to recover the originally transmitted information sequence a_n from the received data, we now consider a joint processing of the C received signals utilizing the structure shown in Figure 2.

To compensate for channel induced distortions, each signal $x_{c,n}$ $(c=1,\ldots,C)$ is processed by a linear equalization filter with impulse response $f_{c,n}$ of finite length L, yielding the filter outputs

$$y_{c,n} = x_{c,n} * f_{c,n} = \sum_{l=0}^{L-1} f_{c,l} x_{c,n-l} = \mathbf{x}'_{c,n} \mathbf{f}_c, \tag{7}$$

where vector $\mathbf{f}_c = (f_{c,0}, \ldots, f_{c,L-1})'$ contains the coefficients of the equalization filter in the cth channel. The resulting C output signals $y_{c,n}$ are then coherently combined yielding the signal

$$y_n = \sum_{c=1}^{C} y_{c,n} = \sum_{c=1}^{C} \mathbf{x}'_{c,n}\mathbf{f}_c \tag{8}$$

from which the transmitted symbols will be estimated. Defining the stacked vectors $\mathbf{x}_n = (\mathbf{x}'_{1,n}, \ldots, \mathbf{x}'_{C,n})'$ and $\mathbf{f} = (\mathbf{f}'_1, \ldots, \mathbf{f}'_C)'$ equation (8) becomes

$$y_n = \mathbf{x}'_n\mathbf{f}\ . \tag{9}$$

It should be noted here that a coherent combination implies a compensation of phase errors in each channel separately, if the phase deviations of the channels do not correlate with each other. Otherwise it is sufficient to perform a phase recovery in the sum channel only [18], which is done in the proposed receiver structure.

For a joint blind adaptation of all filter equalizer coefficients gathered in vector **f** the constant modulus algorithm (CMA) [4], [20] is utilized that adapts **f** by minimizing the cost function

$$\begin{aligned} J_{\mathrm{CMA}}(\mathbf{f}) &= \frac{1}{4}\mathrm{E}\left(|y_n|^2 - \gamma\right)^2, \\ \gamma &= \frac{\mathrm{E}|a_n|^4}{\mathrm{E}|a_n|^2}, \end{aligned} \tag{10}$$

where E denotes the expectation operator. As becomes obvious from (10), the CMA statistically compares the modulus of the multichannel equalizer output with the assumed statistics of the transmit sequence a_n, which is independent from the particular transmitted information. The minimization procedure can be realized by applying a gradient descent approach to (10). Replacing the occurring expectation operation by its instantaneous estimate yields the equalizer coefficient update

$$\mathbf{f}_{n+1} = \mathbf{f}_n - \mu_{\mathrm{CMA}}\left(|y_n|^2 - \gamma\right)\mathbf{x}^*_n y_n, \tag{11}$$

where μ_{CMA} is a small positive step size parameter, and $\mathbf{f}_\mathrm{n}$ is the previously defined filter vector **f** at time instant n.

Since the CMA evaluates the modulus of y_n only, it is robust with respect to phase deviations. However, for the same reason it is not able to recover the original phase information. Therefore, the equalizer output has to be processed with a phase recovery unit, which is implemented as a first-order decision-directed phase-locked loop [11]. From the phase corrected equalizer output $y_{\mathrm{PLL},n}$ estimates $\hat{a}_n$ of the transmitted symbols are derived utilizing a decision device.

4.0 MODEL BASED RECEIVER IMPROVEMENT

As was described in section 2, the propagation of acoustic waves within the water medium is determined by oceanographic parameters of the underlying sea area. Therefore, some coarse knowledge of the channel physics and the source-receiver geometry can be used to estimate the deterministic channel behaviour $h_{c,\mathrm{det}}$

by utilizing a numerical propagation model. It should be noted that this approach is not purely academic, since the needed physical parameters can either be determined by direct measurement and/or can be found in oceanographic databases that are already available for sonar performance prediction. Once the estimate $\hat{h}_{c,\mathrm{det}}$ is obtained it can be used in different ways to enhance the performance of underwater receivers. In order to improve the robustness of the introduced blind receiver with respect to the propagation conditions, we propose a pre-filtering of the received data prior to the equalization step, as indicated in Figure 3. The pre-filters in the receiving channels are designed utilizing the estimates $h_{c,\mathrm{det}}$ to absorb the deterministic component of the channel induced signal distortions, while the task of compensating the stochastic component remains to the adaptive equalizer. In the following, the utilized pre-filters are assumed to be linear FIR filters with a symbol spaced length of L_{pre}. According to the previously defined vector notation, the coefficients of the pre-filter in the cth receiving channel are gathered in the L_{pre}-dimensional vector $\tilde{\mathbf{f}}_c = \left(\tilde{f}_{c,0}, \ldots, \tilde{f}_{c,L_{\mathrm{pre}}-1}\right)'$.

In the following subsections two approaches for the specific design of $\tilde{\mathbf{f}}_c$ are shortly sketched and discussed.

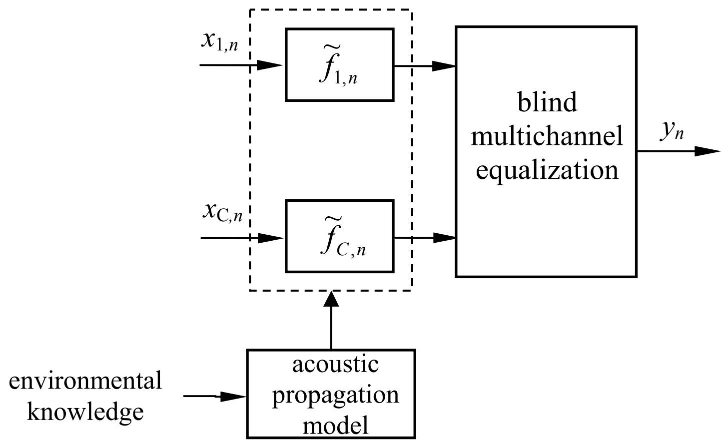

Figure 3: Incorporation of environmental knowledge utilizing model based pre-filtering

4.1 Channel Matched Filtering

Known impulse responses of the different physical transmission channels allow the implementation of the optimum diversity combiner (ODC) proposed by Balaban and Salz [2]. This receiver structure is a cascade of a passive phase conjugation (PPC) receiver [15], where in each channel the received signals are applied to a filter matched to the impulse response of the corresponding channel, and a single channel equalizer that processes the sum signal of the PPC filter bank to compensate for residual signal distortions. In [5] Gomes and Barroso utilized the ODC structure to incorporate knowledge of the channel physics into an underwater acoustic receiver by designing the PPC part using model based predictions of the acoustic field distribution. Assuming an ideal PPC receiver, the impulse response of the resulting transmission system between the transmitter and the output of each matched filter is given by the autocorrelation of the impulse response of the corresponding physical channel. Since the autocorrelation function consists on a distinct main peak and additional side lobes, whose properties are determined by the underlying channel characteristics, the outputs of the matched filter are in general still affected by an individual ISI situation. Therefore, we propose to replace the single channel equalization of the sum signal by a multichannel equalization of the matched filter outputs, which additionally allows for the absorption of errors due to possible mismatch of the model based designed pre-filter. This leads to the structure shown in Figure 3, where the coefficients of the pre-filter in the cth receiving channel is given by the vector

$$\tilde{\mathbf{f}}_c = \left(\hat{h}^*_{\det,c,M-1},\ldots,\hat{h}^*_{\det,c,0}\right), L_{\text{pre}} = M \,, \tag{12}$$

i.e. the time reversed complex conjugated version of the deterministic part of the channel impulse response. Stochastic fluctuations of the channel have to be compensated by the adaptive filters. A more detailed discussion of this approach can be found in [16].

4.2 Minimum Mean Squared Error Pre-Equalization

As explained in the previous section, pure filtering matched to the channel impulse response does not perfectly minimize signal distortions induced by the transmission channel. Therefore, we propose to design the pre-filter impulse response to minimize the mean squared error (MSE) $\mathrm{E}\left|\tilde{x}_{c,n} - a_{n-n_0}\right|^2$ between the output $\tilde{x}_{c,n}$ of the cth pre-filter and a time delayed version of the transmitted symbol sequence a_n. Assuming a linear filter, the coefficients of a minimum MSE (MMSE) pre-filter are given by [8]

$$\tilde{\mathbf{f}}_c = \left[\left(\hat{\mathbf{H}}_c\hat{\mathbf{H}}_c^H + \left(\frac{\sigma_a^2}{\sigma_{w,c}^2}\right)^{-1}\mathbf{I}\right)^{-1}\hat{\mathbf{H}}_c\right]^*\mathbf{i}_{n_c} \,, \tag{13}$$

where $\hat{\mathbf{H}}_{\det,c}$ denotes a $L_{\text{pre}} \times (L_{\text{pre}} + M - 1)$ convolution matrix of the estimated deterministic channel according to (6) , and σ^2_a and $\sigma^2_{w,c}$ are the mean power of the transmitted signal a_n and the additive noise $w_{c,n}$ respectively. Vector $\mathbf{i}_{n_c}$ denotes a unit vector with the 1 element at the n_cth position. It should be noted that (13) is explicitly designed to minimize the disturbances of the transmitted signal. In the noise free case it becomes the so called zero-forcing solution [13], while transforming into the channel matched approach (12) if the influence of noise strongly increases. However, despite its advantageous properties, the determination of the MMSE solution demands a higher computational complexity, and requires additional knowledge of the SNR in the receiving channels. A more detailed description of this approach is given in [17].

5.0 EXPERIMENTAL RESULTS

5.1 Experimental Setup

Within the EC-funded project ROBLINKS, a sea trial for underwater communication was conducted in spring 1999 about 10 km off the Dutch coast in the North Sea. For transmission an almost omnidirectional acoustic source was deployed from the stern of the research vessel Tydeman and lowered at a depth of approximately 9 m. The area of the sea trial is a shallow water zone with typical depths of about 20 m. The transmitted signals were received by two different vertical arrays of hydrophones that have been attached to the research platform "Meetpost Noordwijk" (MPN) to sample the sound field vertically at different depths. The distance between the ship and the arrays varied between 1 km and 10 km. More details of the ROBLINKS experiment can be found in [22].

For performance evaluation of the proposed blind receiver structure and its model based extensions, an OQPSK sequence is considered that had been recorded at MPN while the ship moored at a nominal distance of 1 km from the platform[2]. Utilizing a carrier frequency of 3079 Hz about 100000 data bits were continuously transmitted at a rate of 3771.62 bit/s. The signal was received by a vertical uniform linear array consisting of 7 hydrophones that have been placed between 4.4 m depth and 15.2 m depth (i.e. 1.8 m sensor spacing). The outputs of these sensors have then been jointly processed using the multichannel receiver explained in section 3. Equalization as well as pre-filtering has been performed on data that have been oversampled by a factor of two with respect to the underlying symbol rate (i.e. P=2).

5.2 Oceanographic Parameters & Channel Properties

During the sea trial, basic oceanographic parameters of the sea area, like bottom properties and the sound speed profile within the water layer, have been determined, and the transmitter-receiver geometry has been logged continuously. In case of the selected 1 km experiment, the bottom was almost flat, where the range independent water depth was measured as 17.5 m. Samples showed that the sea bottom consist of silty sand with a density of ρ_b = 1900 kg/m³ and a porosity of 41.25%, suggesting a speed of sound within the bottom layer to be c_b = 1640 m/s [7]. Due to spring time, the water layer was well mixed and exhibted an almost isovelocity SSP with a speed of sound of c_w=1489 m/s. Prior to the data transmission, the noise level in each channel has been measured, yielding an estimated SNR of approximately 35 dB per channel.

5.3 Channel Impulse Response

The transfer behaviour of the underwater acoustic channel was experimentally determined by transmitting a linear frequency modulated (FM) sweep signal that covered the frequency band of the succeeding data signals. The top plot in Figure 4 shows the impulse response of the physical transmission channel between the transmitter and a hydrophone in approximately 8 m depth that has been derived from the received FM sweep. The channel clearly exhibits a time spread of about 25 ms, showing the difficult ISI situation that can be expected for the considered transmission distance.

For the model based estimation of the channel transfer behaviour the underwater channel was assumed to be sufficiently described by a Pekeris wave guide [12]. In this approach the water layer and the bottom are modelled as homogeneous fluid layers with density and speed of sound in each layer being spatially constant. Utilizing this model, the environmental information summarized in section 5.2 together with the logged transmitter-receiver geometry have been evaluated using the propagation channel simulator (PROSIM) software package [14] to obtain model based estimates of the channel impulse responses. The bottom plot in Figure 4 shows the PROSIM prediction of the channel corresponding to the considered setup. Despite the utilized rough modelling of the channel, the obtained estimate is in good agreement with the measured impulse response.

[2] The GPS-logged horizontal distance between the source and the hydrophone array was 1103 m during the transmission.

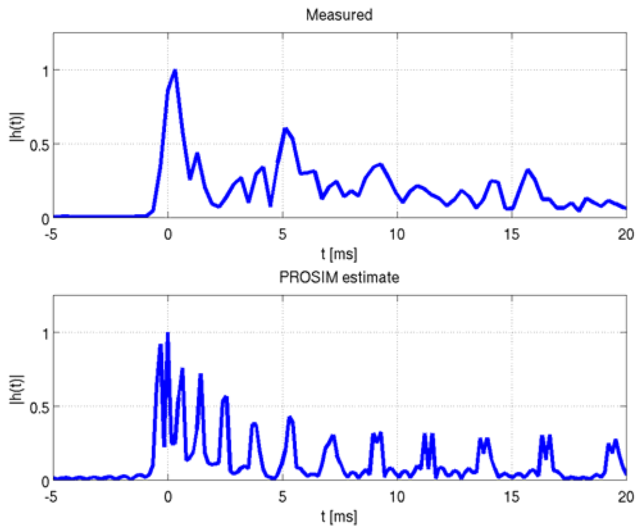

Figure 4: Impulse response of physical transmission channel between transmitter in 9 m depth and receiving hydrophone in 8 m depth; measured (top) and model based estimate (bottom)

5.4 Results

For processing the array outputs, the equalizer filter in each receiving channel have been implemented with a baud-spaced length of L=16. The top plot in Figure 5 shows the evolution of the bit error rate (BER) in dependence on time for pure blind equalization using the CMA, while the middle and bottom plot present the results obtained by extending the receiver structure utilizing the proposed model-based pre-filter approaches. The step size parameter μ_{CMA} has been found by optimization for each case. The pre-filters have been designed using the PROSIM estimated channel impulse responses in equation (12) and (13) respectively. In case of the MMSE pre-equalization the baud-spaced length of the pre-filter was restricted to L_{pre}=30.

The obtained results show that pure blind multichannel equalization (i.e. without pre-filtering) is able to counter the influence of the severe 1 km channel, as can be seen from the obtained BER that is decreasing with increasing time. However, despite processing information from 7 different spatial channels, it suffers from slow tracking behaviour, allowing for momentary occurring BERs of more than 10% after initial convergence. These error events correspond to short-time changes in the transmission conditions, which have to be countered by the adaptive equalizer. However, convergence rate of the proposed linear CMA equalizer is too slow to let **f** converge to the desired solution that compensate the severe time-spread of the transmission channel. In contrast to this, the model-based pre-filter steps are able to catch the *deterministic* signal distortions in the receiving channels. As a result, the transmission system between the transmitter and the cth equalization filter input exhibits a considerably reduced time-spread. The task remaining to the adaptive equalizer is then to compensate for residual ISI that might arise from a mismatch of the deterministically designed pre-filter to the actual channel conditions. Therefore, the equalizer has to cope with deviations from the deterministic transmission conditions only rather than equalizing the complete channel, which leads to an improved tracking behaviour of the blind receiver. This can be clearly seen from the significantly lower BERs in case of pre-filtering, where the MMSE pre-equalization approach outperforms the channel matched pre-filtering results at the expense of additional necessary information, i.e. the SNR in each receiving channel. The model-based pre-filter steps reduce the overall number of bit errors from 8511 in case of pure CMA equalization to 2553 when applying the channel matched pre-filter, and to only 849 wrongly decided bits when the receiver is extended by the MMSE pre-equalizer.

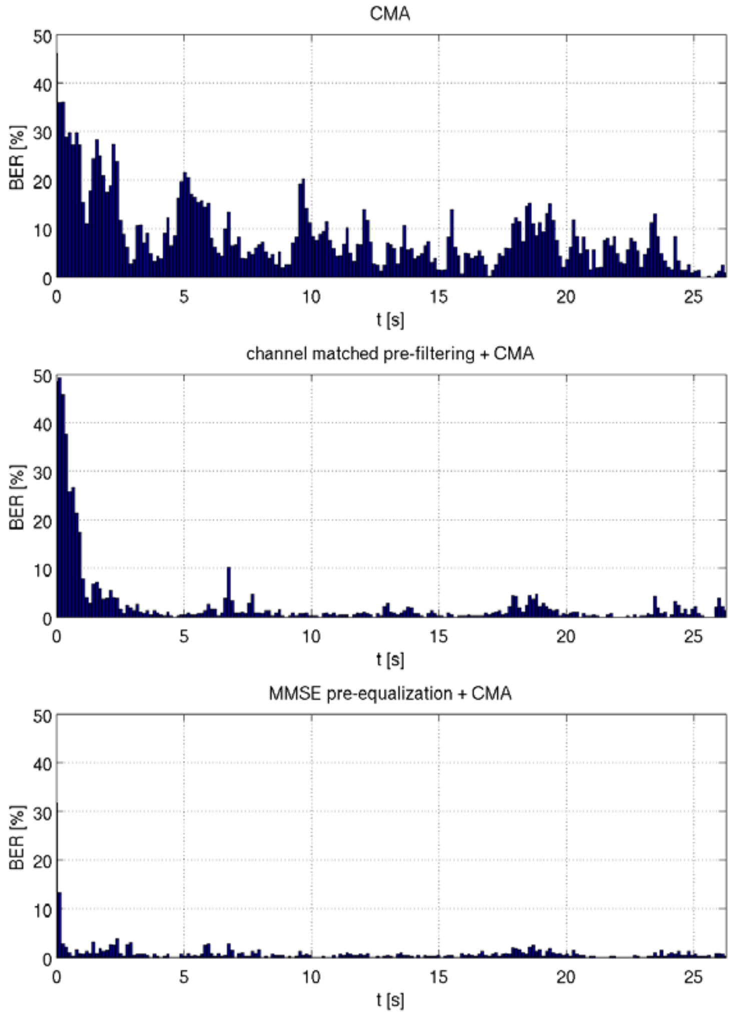

Figure 5: Resulting bit error rate (BER) vs. time for CMA equalization (top), channel matched pre-filtering & CMA equalization (middle), and MMSE pre-equalization & CMA equalization (bottom)

6.0 CONCLUSION

In this paper we presented a multichannel receiver for underwater acoustic communication that does not rely on any co-operation by the transmitter. For adaptation of the utilized linear multichannel equalizer the constant modulus algorithm has been used. If some knowledge of the channel physics and the transmitter-receiver geometry is available, it can be used to predict the deterministic transfer behaviour of the channel. To improve the performance of the blind adaptive equalizer, the additional channel information can be incorporated into the blind receiver structure by introducing a suitable pre-filtering that accounts for deterministic signal distortion induced by the channel. We proposed two model based approaches for the pre-filter design. The results obtained by applying these structures to measured data from a sea trial show the general applicability of blind signal processing methods for underwater acoustic communication as well as the improvement potential of the proposed model based approaches.

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