Performance Enhancement of Blind Adaptive Equalizers Using Environmental Knowledge

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***Abstract* – In this paper, a multichannel equalizer that utilizes information of the surrounding environment is proposed. The structure of this receiver consists of a bank of prefilters matched to the deterministic part of the channel impulse response, whose outputs are fed to a conventional blind multichannel equalizer. If information on the source-receiver geometry as well as physical properties of the channel are available, the unknown impulse responses can be estimated using a numerical propagation model. Assuming a Pekeris model for the physical channel and utilizing a ray propagation model, the receiver structure is tested with measured shallow water data gathered during the ROBLINKS 1999 sea trial. The obtained results demonstrate that incorporating environmental knowledge into the signal processing scheme leads to an improved performance compared to a conventional blind receiver.**

1. INTRODUCTION

Since acoustic waves represent the most effective means of directing energy transfer over long distances in seawater, digital acoustic communication has attracted considerable research work to enable high bit-rate underwater transmission. In the last few decades, the focus of using digital acoustic telemetry moved from almost exclusive military applications to more and more commercial ones. Examples for such applications are remote control of autonomous underwater vehicles (AUVs), pollution-monitoring and collection of scientific data recorded at submerged stations without the need for retrieving the instruments. This development has been accompanied by an ever growing need for higher data rates to cope with the huge amount of data to be transmitted. Typical data rates range from a few kilobits per second for simple command and control tasks up to hundreds of kilobits per second for video image transmission [1]. However, these desired high data rates are in contrast to the transmission conditions induced by the underwater telemetry channel. For example the inhomogeneity of a real oceanic environment causes acoustic wavefronts to be dispersed and refracted during their propagation through an underwater medium. In a shallow water scenario the refracted waves can also be reflected at the surface and/or the bottom. In general, more than one propagation path will exist between transmitter and receiver, which will introduce different delays and attenuations to the originally transmitted signal. This leads to intersymbol interferences (ISI) at the receiver, where the sequentially transmitted pulses overlap in time. In order to achieve a high data throughput, combating the possibly time-varying multipath induced by a shallow water acoustic channel is considered to be the most challenging task [2]. Due to the limitation of available bandwidth, phase-coherent receivers must be employed to increase efficiency. One possibility to perform these tasks is to use joint synchronization and adaptive channel equalization. This was thought to be impossible due to the time variability and dispersive nature of the medium, and was first successfully demonstrated in the early 1990's by Stojanovic, Catipovic and Proakis [3]. However, in a time-varying environment the training of the equalizer implies periodic transmission of known data sequences, which reduces the effective data throughput. Therefore, blind equalization methods represent an important alternative and are topic of current research [4, 5, 6].

One example for blind equalization methods is the well known and successfully in numerous practical applications implemented constant modulus algorithm (CMA), which only uses information about the modulus of the signal constellation of the underlying modulation scheme. However, the CMA is a comparable simple method and suffers from slow convergence. From an intuitive point of view it can be expected that the performance of equalizers based on such an algorithm can be enhanced by incorporating knowledge of environmental properties and the source-receiver geometry into the data processing. In [7] it was shown that an improved convergence behavior of blind equalizers can be achieved by utilizing such knowledge for the initialization step. However, this approach uses the available informations only at the startup of the equalization. In order to benefit from the knowledge of the surrounding environment during the whole equalization process this information should be incorporated into the receiver structure in a suitable way.

In 1999 Gomes and Barroso [8] proposed a matched field processing approach derived from the principle of a time-reversal mirror (TRM). The resulting receiver consists of a cascade of

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a bank of filters matched to the impulse responses of the corresponding physical channels, whose outputs are summed and fed into a single channel equalizer to combat residual intersymbol interference. When information of the environment is available, the coefficients of the matched filters can be estimated using a numerical propagation model. However, these models only yield the deterministic part of the impulse responses, which is determined by the multipath structure of the transmission channel. To compensate the mismatch, each model-based matched filter has to be followed by an adaptive filter. Since the output equalizer seems to be redundant in such a receiver structure, we propose to incorporate this equalizer into each receiving channel and to merge it with the adaptive filters needed for fitting the matched filters to the channel impulse response. Jointly updating the coefficients of all these equalizers with a blind algorithm leads to a conventional blind adaptive multichannel equalizer, where, as a preprocessing step, a filtering matched to the deterministic part of the impulse response of each physical channel is performed.

This paper is organized as follows. In Section 2 the signal model is described and the notation is introduced. The underlying basic blind receiver structure is presented in Section 3, and an approach of enhancing the receivers performance by incorporating environmental knowledge is given in Section 4. In Section 5 the employed channel model is introduced, while in Section 6 the obtained results with the proposed model-based receiver structure applied to experimental data from a sea trial in the North Sea are presented.

2. SIGNAL MODEL

In this section the notation for the discrete-time description of a digital communication channel is introduced. Consider a sequence of symbols $s(n)$ that is transmitted over a single channel with impulse response $h(n)$. In fact, $h(n)$ represents the concatenation of the transmit filter, the communication channel and the receive filter, i.e.

$$\begin{aligned} h(n) &= h_{\text{tx}}(n) * h_{\text{ch}}(n) * h_{\text{rx}}(n) \\ &= h_{\text{filt}}(n) * h_{\text{ch}}(n) \,, \end{aligned} \tag{1}$$

where the influence of the transmit and receive filters is described by the impulse response

$$h_{\text{filt}}(n) = h_{\text{tx}}(n) * h_{\text{rx}}(n) \,. \tag{2}$$

Following [8], the impulse response h_{ch} of the underwater channel can be decomposed into a deterministic component $h_{\text{det}}(n)$ representing the macro-multipath structure, and a zero-mean stochastic component $h_{\text{stoch}}(n)$ reflecting the stochastic fluctuations of the channel

$$h_{\text{ch}}(n) = h_{\text{det}}(n) * h_{\text{stoch}}(n) \,. \tag{3}$$

The channel output is disturbed by additive noise $w(n)$, which is assumed to be white and uncorrelated from the information-bearing signal. The resulting received signal

$$x(n) = s(n) * h(n) + w(n) \tag{4}$$

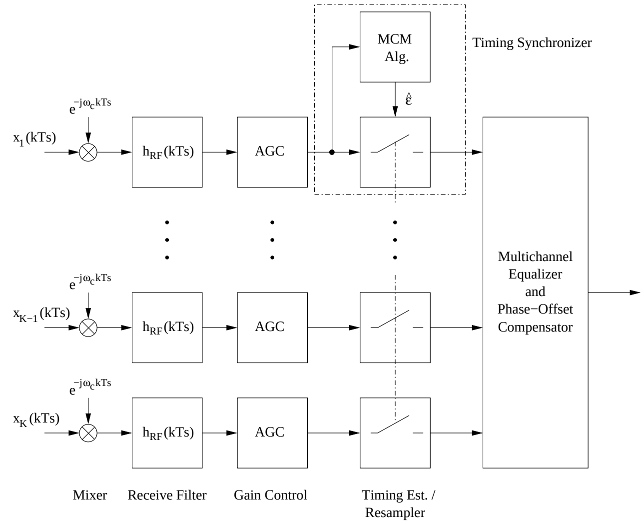


Fig. 1. Multichannel Receiver Architecture

is then fed into an adaptive filter with its coefficients given by the elements of the vector $\boldsymbol{\theta}$. From the equalized signal $y(n)$ the received symbols $\hat{s}(n)$ are estimated by a decision device.

3. BASIC BLIND RECEIVER ARCHITECTURE

Although the simple single channel model presented in the previous section is sufficient for the description of the communication problem, in a real digital receiver several processing steps have to be applied to the received signal prior to equalization. Therefore, this section gives an overview of the underlying blind receiver architecture shown in Figure 1, which also takes the more general case of space-time processing into account.

The signals received at an array of K hydrophones are sampled at a fixed rate $1/T_s$ and subsequently shifted to the baseband using the nominal carrier frequency ω_c. This leads to an error as the actual carrier frequency deviates from the nominal one, e.g. due to Doppler, which has to be compensated for by further processing steps. To reject additive noise outside the frequency band occupied by the transmitted signal, the baseband signals are low-pass filtered utilizing $h_{\text{RF}}(k)$. Subsequently, the mean signal power of each sensor output is normalized to one using an adaptive gain control (AGC). Then, non-data-aided timing recovery is performed from the output of one sensor only yielding an estimate $\hat{\epsilon}$ of the residual timing error ϵ, representing the misalignment between the transmitter and receiver timing grids, where $-0.5 \leq \epsilon \leq 0.5$ is assumed. This estimate is used to dynamically resample the signals in all channels simultaneously, leaving only the task of timing adjustment associated with inter-sensor timing fluctuations to the equalizer. The input of the multichannel equalizer and phase-offset compensator are K signals sampled at a fraction l of the symbol interval T. The multichannel equalizer adaptively reduces ISI induced by the transmission channel. The residual time-varying carrier-phase offset is compensated using a digital phase-locked loop (PLL).

Although this receiver structure has proved its applicability for underwater communications using different kinds of blind multichannel equalizers [6], one can think of enhancing its performance by incorporating additional information on the channel into the signal processing scheme. This approach seems not to be devious since in many communication applications information about the source-receiver geometry as well as physical properties of the medium are available or are at least easy to determine.
In the following section a model-based structure is presented, where prior knowledge of the environment and the source-receiver geometry is utilized to counter multipath effects.

4. INCORPORATING ENVIRONMENTAL KNOWLEDGE

An approach for incorporating the channel physics into the receiver structure was presented by Gomes and Barroso in [8], where the idea of a time-reversal mirror was utilized and which will be explained in Section 4.1. Based on this approach our newly proposed structure is presented in Section 4.2.

4.1. Existing Matched Field Processing Approach

4.1.1. Principle of a Time-Reversal Mirror

A recent field of interest in underwater acoustic communication is the concept of a so called time-reversal mirror (TRM) [9], which utilizes the idea of phase conjugation [10]. In a TRM experiment, a signal is emitted by a point source and received by a transducer array. In the ideal case, retransmitting the time reversed and complex conjugated version of the received signals will yield the overall energy to be focused in the location of the point source, while the resulting signal can be shown to be proportional to the time reversed and complex conjugated version of the originally transmitted signal. These effects are due to the fact that during the retransmitting step the ocean itself acts as a matched filter to its own impulse response.

4.1.2. Utilizing the Time-Reversal Mirror for a model-based Receiver

In [8] the working principle described in 4.1.1 has been used to derive a model-based receiver structure from the concept of a simulated time-reversal mirror[1]. In a simulated TRM the effect of a real TRM is achieved by applying filters matched to the corresponding channel impulse response to the received signals before subsequent combination, instead of actually retransmitting the received signals. This structure is equivalent to the optimum diversity combiner (ODC) proposed by Balaban and Salz in a multichannel receiver context [11], since it was shown in [8] that this combiner can be seen as a cascade of a simulated TRM-like combiner bank of filters followed by a single channel equalizer. The ODC was also the base for the development of the multichannel receiver presented in [3].

[1] For the sake of completeness it should be noted here that a simulated TRM is an alternative description for the concept of passive phase conjugation presented in [13].

However, the drawback of this approach is that the impulse responses of the physical channels needed for the implementation of the matched filters are usually a priori unknown. Therefore, for estimating the impulse responses one can think of transmitting a probe signal prior to data sequences [12, 13]. Since the channel might vary with time this training procedure has to be repeated periodically, leading to a less efficient data throughput. This method is especially not applicable for blind systems that should operate totally unsupervised from the transmitter. In these cases a more suited approach is to estimate the impulse responses from available knowledge of the channel. However, these estimates only represent the deterministic part of the real impulse responses, which in general will stochastically vary. To compensate for these variations as well as for errors due to model mismatch each model-based matched filter has to be followed by an adaptive filter as it is done in [8], leading to the structure shown in Figure 2(a). The adaptation of the coefficients of these filters can either be done jointly or independently from the adaptation of the single channel equalizer.

4.2. Model-Based Prefiltering

Since the single channel equalizer in Figure 2(a) seems to be redundant we propose to incorporate this equalizer into each receiving channel and to merge it with the adaptive filters needed for fitting the matched filters to the channel impulse response. Updating the coefficients of all these equalizers jointly with a blind algorithm leads to a conventional blind adaptive multichannel equalizer [6], where, as a preprocessing step, a filtering matched to the deterministic part of the impulse response of each physical channel is performed (Figure 2(b)). This idea is strongly related to the approach presented in [14], where a model-based matched filter was introduced as an improvement in underwater acoustics compared to a conventional matched filter. The basic working principle will be outlined in the following.
Let $s(n)$ be the signal sequence to be transmitted. Using the notation introduced in Section 2 the signal received by the kth sensor is given by

$$x_k(n) = s(n) * h_k(n) + w_k(n)\,, \tag{5}$$

where $h_k(n)$ denotes the impulse response of the kth physical channel and $w_k(n)$ represents additive noise in the kth receiving channel ($k = 1, \ldots, K$). Performing a filtering matched to the deterministic part of the channel using a model-based estimate $\hat{h}_{\text{det},k}(n)$ of the channel impulse response yields

$$\begin{aligned} \tilde{x}_k(n) &= x_k(n) * \hat{h}^*_{\text{det},k}(-n) \\ &= s(n) * h_{\text{filt}}(n) * h_{\text{stoch},k}(n) \\ &\quad * h_{\text{det},k}(n) * \hat{h}^*_{\text{det},k}(-n) \\ &\quad + w_k(n) * \hat{h}^*_k(-n) \\ &= s(n) * \tilde{h}_k(n) + \tilde{w}_k(n)\,, \end{aligned} \tag{6}$$

which is then used as the input of the kth channel of a multichannel equalizer.

Neglecting the disturbance $\tilde{w}_k(n) = w_k(n) * \hat{h}^*_{\text{det},k}(n)$ in (6), the overall impulse response between the transmitter and the kth equalizer input is given by

$$\begin{aligned} \tilde{h}_k(n) &= h_{\text{filt}}(n) * h_{\text{stoch},k}(n) \\ &\quad * h_{\text{det},k}(n) * \hat{h}^*_{\text{det},k}(-n)\,, \end{aligned} \tag{7}$$

which consists on the influence of the transmit and receive filters, the unknown stochastic component of the channel, and the cross-correlation

$$r_{h\hat{h}}(n) = h_{\text{det},k}(n) * \hat{h}^*_{\text{det},k}(-n) \tag{8}$$

of the real and the estimated deterministic channel component. If we assume to have an almost perfect estimate of $h_{\text{det},k}(n)$ except for an unknown time delay n_0, $r_{h\hat{h}}(n)$ will be a pulse with its peak centered around n_0 and its shape defined by $h_{\text{det},k}$. Under this assumption, it follows from (6) that in the noise free case

$$\tilde{x}_k(n) \approx s(n - n_0) * (h_{\text{filt}}(n) * h_{\text{stoch},k}(n))\,, \tag{9}$$

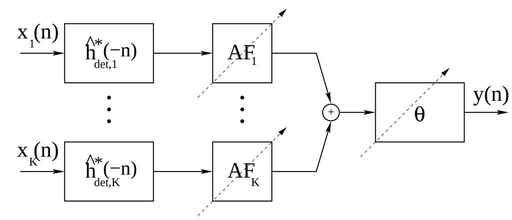
i.e. due to the prefilter step the disturbance originally induced by $h_{\text{det},k}(n)$ is approximately replaced by a mere time delay. This will lead to a reduced ISI, since the deterministic part of the channel impulse response represents the macro multipath structure of the channel. The residual distortion is then only caused by the stochastic component and the system filters. Since the influence of the latter can be reduced by a suitable design[2], the main task of the equalizer is to counter stochastic fluctuations of the real channel from the modeled channel.

It should be noted that the equalization performance in such a model-based receiver is strongly dependent on the channel modeling. In case of severe model mismatch the received signal is disturbed by an arbitrary impulse response, and, therefore, the prefilter will in general decrease the receiver performance. If the deterministic component of the channel impulse response is varying with time, the channel properties have to be tracked in order to assure that the modeled impulse response is well fitted to the real environment. The needed update of the model leads to an increased complexity of both the structure and the used algorithms compared to a non model-based blind receiver. However, if the deterministic component of the channel is stable enough during the transmission a static model should yield some improvement in the equalizer performance. In the following a simple ray model for estimating the channel impulse response is presented, which proved to be sufficient for the investigated data.

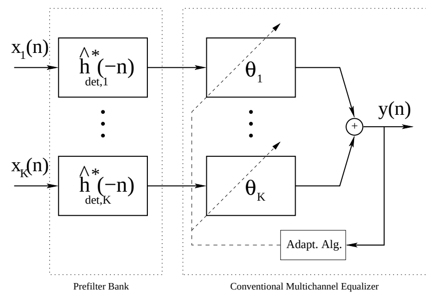
5. CHANNEL MODELING

For modeling the shallow-water acoustic channel we use the Pekeris model [15], which is a simple two-layer propagation model consisting of a water layer of depth D_w and an infinite fluid halfspace representing the bottom. For both layers

[2] Since a filter matched to the transmit pulse has not been applied in our receiver structure presented in Section 3, this filtering task was left to the equalizer during the analysis of the experimental data in Section 6.



(a) Matched Field Processing Approach



(b) Model-Based Prefiltering

Fig. 2. Receiver Structures

the densities ρ_w and ρ_b as well as the sound speeds c_w and c_b are assumed constant and known. Furthermore, it is assumed that $c_b > c_w$. Using these assumptions the wave propagation within the acoustic channel can be simulated with a numerical propagation model from which the corresponding transfer function can be estimated.

In the following a ray based numerical propagation model is presented that is used to determine the impulse response of the underwater acoustic channel.

5.0.1. Ray Propagation Model

In the Pekeris model acoustic waves are not refracted due to the assumed constant sound speed. This attracts the use of a ray propagation model, since in this case the eigenrays between the transmitter and the receiver are comparable easy to determine. In fact, the properties of these rays can be directly computed from the source-receiver geometry, as will be shown in the following.

Let d be the horizontal distance between a transmitter at depth D_{TX} and a receiver at depth D_{RX}. Using fundamental laws of trigonometry it can be shown that the length of an eigenray, which is reflected at the surface first and which is reflected i times in a whole during the propagation, is given by

$$r_i = \sqrt{(D_{\text{TX}} + a_i D_w + b_i D_{\text{RX}})^2 + d^2}, \tag{10}$$

for $i = 1, 2, \ldots$, where

$$a_1 = 0 \tag{11}$$

$$a_{i+1} = a_i + (1 + (-1)^{i+1}) \tag{12}$$

$$b_i = (-1)^{i+1}\,. \tag{13}$$

Similarly, the absolute value of the corresponding grazing angle $\theta_{w,i}$ can be shown to be

$$|\theta_{w,i}| = \arctan\left(\frac{|D_{\text{TX}} + a_i D_w + b_i D_{\text{RX}}|}{d}\right), \quad (14)$$

where a_i and b_i are given by (11)-(13). For eigenrays that are reflected at the bottom first, equations (10) and (14) hold true by replacing the depths of transmitter and receiver by their corresponding heights

$$H_{\text{TX}} = D_w - D_{\text{TX}} \quad , \quad H_{\text{RX}} = D_w - D_{\text{RX}} \quad (15)$$

over the bottom.

During the interaction of the acoustic waves with the boundaries only a part of the energy is reflected. The fraction of the reflected component is described by a coefficient $0 \leq R \leq 1$, which is known as the Rayleigh coefficient. In case of an interface between two fluid layers, which equals the bottom in the Pekeris model, the Rayleigh coefficient is given by [16]

$$R_{\text{bottom}} = \frac{\rho_b c_b / \sin\theta_b - \rho_w c_w / \sin\theta_w}{\rho_b c_b / \sin\theta_b + \rho_w c_w / \sin\theta_w}, \quad (16)$$

where the grazing angle θ_w of the incident wave and the angle θ_b of the wave transmitted into the bottom obey Snell's law of refraction

$$c_b \cos\theta_w = c_w \cos\theta_b \,. \quad (17)$$

For the interaction of the acoustic waves with the sea surface we assumed to have an almost perfect reflection [17]

$$R_{\text{surface}} \approx 1 \,. \quad (18)$$

5.0.2. Estimating the Channel Impulse Response

Since a multipath channel acts like a delay line, the baseband impulse response of the deterministic channel component can be modeled in time domain as a superposition of time delayed and attenuated pulses

$$\hat{h}_{\text{det}}(t) = \sum_{p=0}^{P} \alpha_p q(t - \tau_p), \quad (19)$$

where it is assumed to have one direct path ($p = 0$) and P additional multipaths. Considering the restricted bandwidth of the underwater channel, the pulses $q(t)$ have been chosen to be the impulse response of a lowpass filter covering the bandwidth of the underwater channel. The attenuation factors α_p have been assumed to be the product of reflection coefficients given by (16) and (18), corresponding to the occurring bounces along the pth propagation path. The time delay τ_p is determined by

$$\tau_p = \frac{r_p}{c_w}, \quad (20)$$

where r_p is the length of the pth path that can be calculated using (10).

For simplicity, other disturbing effects like volume attenuation, frequency dependent losses and possible impacts on the signal phase due to boundary interactions have been ignored in this model. Effects due to Doppler can be considered in the model by allowing the time delays τ_p to be time-variant [18]. However, since for the investigated experimental data the influence of Doppler was not very significant, the impulse response given in (19) was not refined that way. In general, all information regarding the signal phase will be randomly corrupted by the stochastic component of the channel, and, therefore, the compensation of these deviations of the modeled channel from the real one are subject to the subsequent equalization process in our model-based structure.

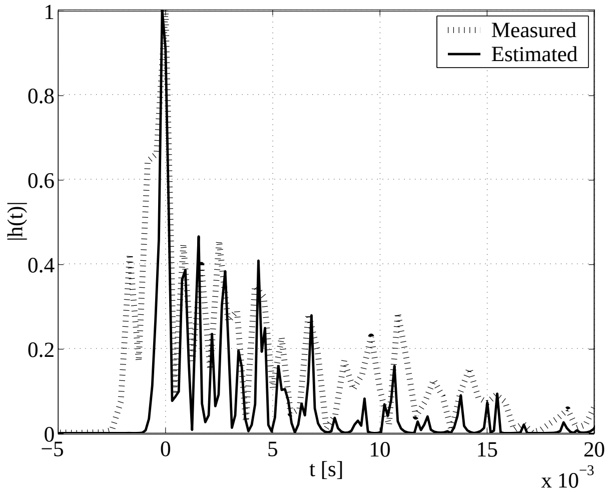
6. EXPERIMENTAL RESULTS

Within the EC-funded project ROBLINKS, in spring 1999, a sea trial for underwater communication was conducted in the North Sea about 10 km off the Dutch coast. At the transmitting side an almost omnidirectional acoustic source was deployed from the stern of a support ship and lowered to a depth of approximately 9 m. The total water depth in this shallow water area is about 20 m. The signals were received by a 6-element vertical array, consisting of three pairs of hydrophones, separated horizontally by 15 cm, that sampled the sound field at depths of approximately 7, 11, and 15 m below the sea surface. The array was fixed to an oceanographic platform, which also hosted all recording facilities. The support ship anchored at various positions with distances of 1, 2, 5, and 10 km from the platform. In addition, the ship was sailing at moderate speed the last two days to obtain measurements with a moving source. More details on the ROBLINKS experiments can be found in [19].

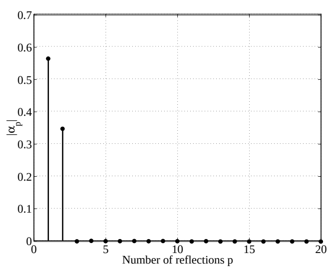
To examine the performance of the proposed model-based receiver structure, we analyzed an MSK data sequence at a rate of 3233.81 bit/s that had been recorded while the ship moored at a distance of 2 km from the platform. The carrier frequency was 3079 Hz and approximately 87000 data bits were continuously transmitted. The signals received at the hydrophones 2, 4, and 6 (corresponding to water depth 7, and 11, and 15 m) of the 6-element array were jointly processed in a multichannel receiver. During time synchronization, an oversampling factor of $l = 2$ was used.

Prior to the information-bearing signal, a linear FM sweep of 200 ms duration spanning the frequency band from 1 to 5 kHz was transmitted to obtain initial information about the channel responses. Figure 3(a) shows the result of a matched-filter analysis of the FM sweep received at hydrophone 6 (dotted line) . It is apparent that the delay spread of the channel is in the order of 5 ms, corresponding to 16 symbol intervals. This gives some guidelines for choosing a proper length of the multichannel equalizer filters.

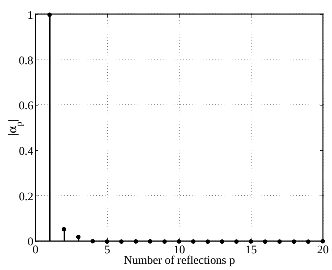
Additionally in Figure 3(a) a least squares fit of the model (19) is plotted (solid line), where besides the direct path the existence of the first 20 eigenrays with a surface reflection first and 20 eigenrays with a bottom reflection first has been assumed, resulting in a 41-path multipath scenario. The modulus of the



(a) Channel Impulse Response

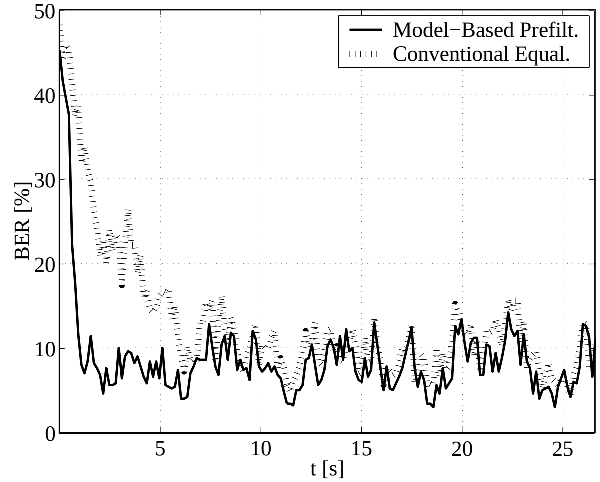


(b) Attenuation parameters for paths reflected at surface first

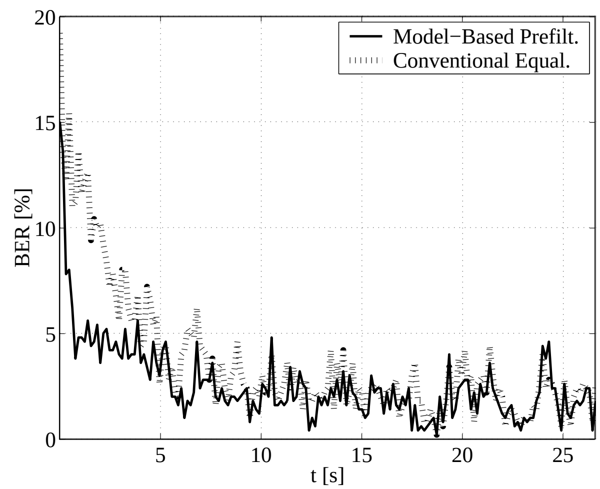


(c) Attenuation parameters for paths reflected at bottom first

Fig. 3. Impulse Response (Hydrophone 6)



(a) Single Channel Equalization



(b) Multichannel Equalization

Fig. 4. Bit Error Rates

corresponding estimates for the attenuation factor $\hat{\alpha}_p$ is given in Figure 3(b) and Figure 3(c) respectively. These plots indicate that in the assumed model only eigenrays with few reflections will contribute significantly to the impulse response of the channel. Similar observations were made for the other two channels. Therefore, all model-based prefilters were designed to model spreading paths with a surface reflection first up to a maximum number of 4 reflections in a whole. For the hydrophones 2 and 4 only the first bottom reflected path was assumed, while for hydrophone 6 bottom reflected first paths with up to 2 reflections have been considered.

The length of the adaptive equalizers has been chosen to $L_f = 31$, and the constant modulus algorithm was used for jointly updating the filter coefficients. The used step-size parameter for this algorithm was found by optimization at each case.

The results obtained from applying the proposed model-based equalization structure to the experimental data are shown in Figure 4(a) for a single channel case, and in Figure 4(b) for a multichannel case, where the bit error rate (BER) over time is plotted and compared to the performance of a conventional equalizer. For the single channel equalization, only the outputs of hydrophone 6 have been used, while in the multichannel case the outputs of hydrophones 2, 4, and 6 have been jointly processed. Although the coarse model presented in Section 5 has been utilized, the obtained results show an improved performance by using the model-based structure: the overall bit error rate is reduced from 13.13% to 8.78% in the single channel case and from 3.67% to 2.56% in the multichannel case. The small improvement in the latter case is due to the already good performance, that can be archived without prefiltering by just combining data from different sensors. However, the model-based prefiltering also results in a faster convergence, which is especially important for blind equalizers that utilize simple but slow converging algorithms like the CMA.

7. CONCLUSION

We proposed a blind model-based receiver structure that incorporates information on the environment by applying filters matched to the deterministic part of the channel impulse response before equalization in order to counter for most of the multipath. The needed but unknown impulse responses have been estimated from prior knowledge of the source-receiver geometry and physical properties of the medium utilizing a ray propagation model. Results with experimental data have demonstrated the improved performance of the model-based structure compared to a conventional blind receiver.

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