1 HIGH-SPEED UNDERWATER ACOUSTIC COMMUNICATIONS

Milica Stojanovic
Massachusetts Institute of Technology

Underwater acoustic communications are a rapidly growing field of research and engineering, driven by the expansion of applications which require underwater data transmission without wired connections. In this chapter, we explore the problems of underwater acoustic communications in three parts. The first part presents an overview of modern applications in underwater data transmission and today's achievements in this area. System requirements are reviewed, and propagation characteristics of underwater acoustic channels are given. It is shown that the majority of underwater acoustic channels are severely band-limited, with signal distortions depending on the link configuration, and ranging from benign to extreme ones caused by time-varying multipath propagation and signal phase variations. Examples of existing systems are given, with emphasis on the methods used for intersymbol interference mitigation. Most of these systems use noncoherent or a differentially coherent signal modulation and detection methods. Phase-coherent detection, which offers better efficiency in bandwidth utilization, is the subject of the second part of this chapter. In this part, the design of high-speed digital communication systems, which rely on powerful equalization and multiple sensor signal processing methods is treated. Theoretical aspects of adaptive multichannel equalization are given, followed by a discussion on adaptive algorithm selection and methods for reducing the receiver complexity. An example of experimental performance analysis is presented, and a DSP implementation of the receiver is described. The concluding part is devoted to future research in the area, which is expected to lead towards the development of high-speed mobile acoustic communication systems and underwater communication networks.

1 CHANNEL CHARACTERISTICS AND SYSTEM DESIGN PRINCIPLES

The need for underwater wireless communications exists in applications such as remote control in off-shore oil industry, pollution monitoring in environmental systems, collection of scientific data recorded at ocean-bottom stations and by unmanned underwater vehicles, speech transmission between divers, and mapping of the ocean floor for objects detection and recovery. Wireless underwater communications can be established by transmission of acoustic waves. Radio waves are of little use because they are severely attenuated, while optical waves suffer from scattering and need high precision in pointing the laser beams. Underwater acoustic communication channels are far from ideal. They have very limited bandwidth, and often cause severe signal dispersion in time and frequency [1]-[5].

Among the first modern underwater communication systems was an underwater telephone, which was developed in the forties in the United States for communication with submarines [2]. This device used a single-sideband (SSB) suppressed carrier amplitude modulation in the frequency range of 8-11 kHz, and it was capable of sending acoustic signals over several kilometers. Today, a new generation of systems is made possible by implementing powerful signal processing and data compression algorithms on digital signal processors (DSPs).

During the past few years, significant advencements have been made in the development of underwater acoustic communication systems, in terms of their operational range and data throughput. Acoustically controlled robots have been used to replace divers in performing maintenance of submerged platforms [9]; high-quality video transmission from the bottom of deepest ocean trenches (6500 km) to a surface ship was established [10]; and data telemetry over horizontal distances in excess of 200 kilometers was demonstrated [19].

The development of efficient communication methods makes new applications possible, which, in turn, impose new requirements on the system performance. Many of the developing applications, both commercial and military, require real-time communication with submarines and autonomous, or unmanned underwater vehicles (AUVs, UUVs). Vehicles, robots and stationary sources on underwater moorings, equipped with oceanographic instruments and cameras, are foreseen to operate together in the future underwater data networks.

System Requirements

In the existing systems, there are usually four kinds of signals that are transmitted: control, telemetry, speech and video signals. The achievable data throughputs, and the reliability of an underwater acoustic communication system, as measured by the bit-error rate, must be determined to suit the bandwidth limitations and distortions of underwater acoustic channels.

Control signals include navigation, status information, and commands for underwater robots, vehicles and submerged instrumentation such as pipeline valves or deep ocean moorings. The data rates up to about 1 kilobit per second (kbps) are sufficient for these operations, but very low bit-error rates may be required [4].

Telemetry data is collected by submerged acoustic instruments such as hydrophones, seismometers, sonars, current-meters, chemical sensors, and it also may include low rate image data. Data rates on the order of one to several tens of kbps are required for these applications. The reliability requirements are not so stringent as for the command signals, and a probability of bit error of $10^{-3} - 10^{-4}$ is acceptable for many of the applications.

Speech signals are transmitted between divers and to a surface station. While the existing, commercially available diver communication systems mostly use analog communications, based on single-sideband modulation of the 3 kHz audio signal, research is advancing in the area of synthetic speech transmission for divers, as digital transmission is expected to provide better reliability. Transmission of digitized speech by linear predictive coding (LPC) methods requires rates on the order of several kbps to achieve close-to-toll quality. The bit error rate tolerance of about 10^{-2} makes it a viable technology for poor quality band-limited underwater channels [12, 13].

Video transmission over underwater acoustic channels requires extremely high compression ratios if an acceptable frame transmission rate is to be achieved. Fortunately, underwater images exhibit low contrast, and preserve satisfactory quality if compressed to few bits per pixel. Compression methods, such as the JPEG (Joint Photographic Experts Group) standard, discrete cosine transform, have been used to transmit 256×256 pixel still images with 2 bits per pixel, at transmission rates of about one frame per 10 seconds [10]. Further reduction of the required transmission rate seems to be possible by using dedicated compression algorithms, e.g., the discrete wavelet transform [14]. Current achievements report on the development of algorithms capable of achieving compression ratios on the order of 100:1. On the other hand, underwater acoustic transmission of television-quality monochrome video would require compression ratios higher than 1000:1. Hence, the required bit rates for video transmission range from higher than

10 kbps, possibly up to several hundreds of kbps. Performance requirements are moderate, as images will have satisfactory quality at bit error rates on the order of $10^{-3} - 10^{-4}$.

Channel Characteristics

Unlike in the majority of other communication channels, the use of underwater acoustic resources has not been regulated yet by standards. The available bandwidth and transmission range in an underwater acoustic channels depend on the signal-to-noise ratio which is primarily determined by transmission loss and noise level. System performance and its information throughput depend on the signal distortions caused by reverberation, or multipath propagation. Channel characteristics are time-varying and depend on the system location.

Range and Bandwidth

Transmission loss is caused by energy spreading and sound absorption. While the energy spreading loss depends only on the propagation distance, the absorption loss increases not only with range but also with frequency, thus setting the limit on the available bandwidth [1]. In addition to this nominal transmission loss, the received signal level is influenced by the spatial variability of the underwater acoustic channel, such as the formation of shadow zones. Transmission loss at a particular location can be predicted by many of the propagation modeling techniques [1] with various degrees of accuracy. Spatial dependence of transmission loss imposes particularly severe problems for communication with moving sources or receivers.

Noise observed in the ocean consists of man-made noise and ambient noise. In deep ocean, ambient noise dominates, while near shores and in the presence of shipping activity, man-made noise significantly increases the noise level. Most of the ambient noise sources can be described as having a continuous spectrum and Gaussian statistics [1]. As a first approximation, the ambient noise power spectral density is assumed to decay at 20 dB/decade, both in shallow and deep water, over frequencies of interest to communication systems design.

Frequency-dependent transmission loss and noise determine the relationship between the available range, bandwidth and SNR at the receiver input. This dependence is illustrated in Fig.1 which shows the frequency dependent term of SNR for several transmission ranges. (The SNR is evaluated assuming spherical spreading, absorption according to Thorp and a 20 dB/dec decay of the noise power spectral density[1].) Evidently, this dependence influences the choice of a carrier frequency for the desired transmission range. In addition, it determines the relationship between the available range and frequency band. As a result, underwater acoustic communication links can be classified according to range. For a long-range system, operating over 10-100 km, the bandwidth is limited to few kHz (for a very long distance on the order of 1000 km, the available bandwidth falls below a kHz). A medium-range system operating over 1-10 km has a bandwidth on the order of 10 kHz. A short-range system operates over distances less than a km with bandwidth in excess of 10 kHz, while only at very short distances below about 100 m, more than a hundred kHz of bandwidth may be available.

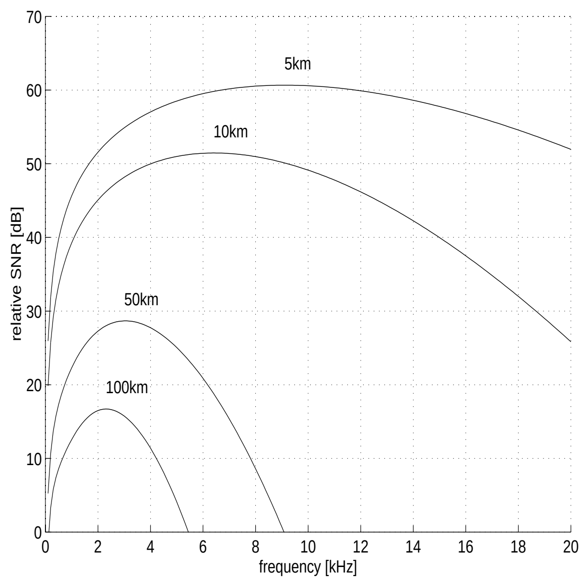

Figure 1: Frequency-dependent portion of SNR.

Multipath

Within a limited bandwidth, the signal is subject to multipath propagation through a channel whose characteristics vary with time and are highly dependent on the location of the transmitter and receiver. In the first place, the multipath spread depends on the link configuration, which is primarily designated as vertical or horizontal. While vertical channels exhibit little time-dispersion, horizontal channels may have extremely long multipath spreads. In a digital communication system which uses a single carrier, multipath propagation causes intersymbol interference (ISI), and an important figure of merit is multipath spread in terms of symbol intervals.

While typical multipath spreads in the commonly used radio channels are on the order of several symbol intervals, in the horizontal underwater acoustic channels they increase to several tens, or a hundred of symbol intervals for moderate to high data rates. For example, a commonly encountered multipath spread of 10 ms in a medium-range shallow water channel causes the ISI to extend over 100 symbols if the system is operating at a rate of 10 kilosymbols per second (ksps).

The mechanisms of multipath formation in the ocean are different in deep and shallow water, and also depend on the frequency and range of transmission. Depending on the system location, there are several typical ways of multipath propagation, determined mostly by the water depth. The definition of shallow and deep water is not a strict one, but usually implies the region of continental shelves, with depth less than about 100 m, and the region past the continental shelves, respectively. One mechanism of multipath formation is by reflections off the bottom, surface and any objects in the water, and this mechanism prevails in shallow water in addition to a possible direct path. Another mechanism, prevalent in deep water, is by ray bending which occurs because the rays of sound tend to reach regions of lower propagation speed. In this way, the sound channel may form by repeated bending of the rays toward the location where sound speed reaches its minimum, called the axis of the deep sound channel. Since there is no loss due to reflections, sound can travel in this way over several thousands of kilometers. Alternatively, the rays bending upwards in deep water may reach the surface focusing in one point where they are reflected, and the process is repeated periodically. The region between two focusing points on the surface is called a convergence zone, and its typical length is 60 -100 km.

The geometry of multipath propagation and its spatial dependence are important for communication systems which use array processing to suppress multipath (e.g., [16], [17]). The design of such systems is often accompanied by the use of a propagation model for predicting the angular distribution of multipath arrivals. Ray theory and the theory of normal modes provide basis for such propagation modeling.

Time-Variation

Associated with each of the deterministic propagation paths (macro-multipaths), which can be modeled accurately, are random signal fluctuations (micro-multipath), which account for the time-variability of the channel response. Some of the random fluctuations can be modeled statistically [1]. These fluctuations include surface scattering due to waves, which is the most important contributor to the overall time variability of the shallow water channel. In deep water, internal waves additionally contribute to the time-variation of the signal propagating along each of deterministic paths.

Surface height displacement can be well modeled as a zero-mean Gaussian random variable, whose power spectrum is completely characterized by the wind speed [1]. Motion of the reflection point results in the Doppler spreading of the surface-reflected signals. Highest Doppler spreads, with values on the order of 10 Hz, are most likely to be found in short and medium range links, which use relatively high frequencies. Note that this effect is present in the channel regardless of the system's mobility.

Statistical channel modeling has significance for communication system design and analysis by simulation. While experimental model-fitting results are limited, short and medium-range channels are often modeled as Rayleigh fading channels. The deep water channel has also been modeled as a Rayleigh fading channel; however, the available measurements are scarce, often making channel modeling a controversial issue [5].

To illustrate the time-varying multipath effects, Figs.2-4 each show an ensemble of channel impulse responses, observed as functions of delay over an interval of time. These responses are estimated from experimental measurements obtained in three typical underwater environments: long-range deep and shallow water, and medium-range shallow water. Relevant system parameters are indicated in the figures.

For a digital communication system which uses adaptive signal processing to track the time-variations of the channel, a relevant parameter is the Doppler spread normalized by the signal bandwidth. This parameter needs to be much less than 1 to enable efficient tracking. Consequently, the implications time-varying multipath bears on the high-speed communication system design are twofold. On one hand, signaling at a high rate causes many adjacent symbols to interfere at the receiver, and requires sophisticated processing to compensate for the ISI. On the other hand, as pulse duration becomes shorter, channel variation over a single symbol interval becomes slower. This allows an adaptive receiver to efficiently track the channel on a symbol-to-symbol basis, or even less frequently, provided, of course, a method for dealing with the resulting time-dispersion.

Examples of Existing Systems

To overcome the difficulties of time-varying multipath dispersion, the design of many underwater acoustic communication systems has so far relied mostly on the use of noncoherent modulation techniques and signaling methods which sacrifice throughput to achieve robustness to channel distortions. Recently, phase-coherent modulation techniques, together with array processing for exploitation of spatial multipath diversity, have been

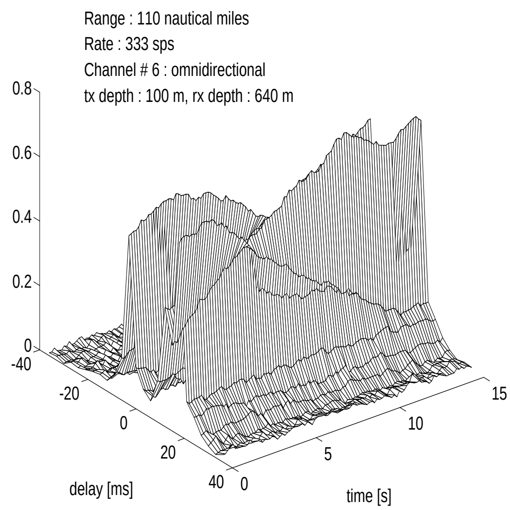

Figure 2: Ensemble of long-range channel responses in deep water (approx. 2000 m) off the coast of California, during the month of January. Carrier frequency is 1 kHz. Range corresponds to three convergence zones. Channel estimates are obtained by recursive least squares estimation using pseudo-random QPSK signals.

shown to provide a feasible means for a more efficient use of the underwater acoustic channel bandwidth. These advancements are expected to result in a new generation of underwater communication systems, with at least an order of magnitude increase in data throughput.

Approaches to system design vary according to the technique used for overcoming the effects of intersymbol interference and signal phase variations. Specifically, these techniques involve the choice of modulation/detection method which provides robustness to the channel impairments, and the choice of transmitter/receiver structure which may include array processing and/or equalization methods. While most of the existing systems operate on the vertical, or the very short-range channels, the systems under development often focus on the severely spread horizontal shallow water channels.

Noncoherent detection of FSK (frequency shift keying) signals has been used to overcome rapid phase variation present in many underwater channels. To deal with the ISI, these systems employ guard times, which are

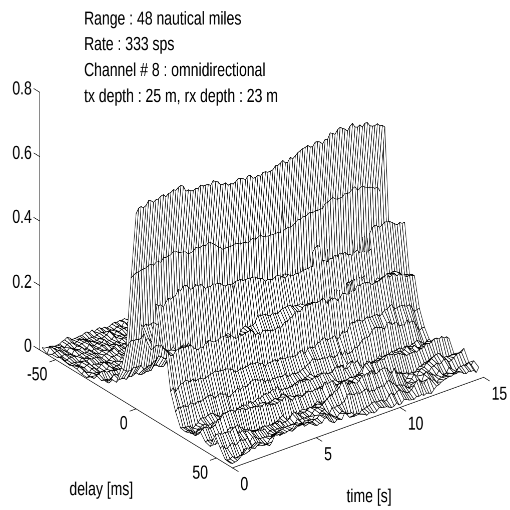

Figure 3: Ensemble of long-range channel responses in shallow water (approx. 50 m) off the coast of New England, during the month of May. Carrier frequency is 1 kHz.

inserted between successive pulses to ensure that all the reverberation vanishes before each subsequent pulse is to be received. The insertion of idle periods of time obviously results in a reduction of the available data throughput. In addition, because fading is correlated among frequencies separated by less than the coherence bandwidth (the inverse of the multipath spread), only those frequency channels with sufficient separation should be used at the same time. This requirement further reduces the system efficiency unless coding is employed so that the adjacent, simultaneously used frequency channels belong to different codewords. A representative noncoherent system [6] uses a multiple FSK modulation technique in the 20-30 kHz band, with maximum bandwidth efficiency of 0.5 bps/Hz. The band is divided into 16 subbands, in each of which a 4-FSK signal is transmitted. This system has successfully been used for telemetry over a 4 km shallow water horizontal path, and a 3 km deep ocean vertical path. It was also used on a less than 1 km long shallow water path, where probabilities of bit error on the order of $10^{-2} - 10^{-3}$ were achieved without coding. The system incorporates the possibility of error correction coding, which improves performance at the expense of reducing the information throughput. The

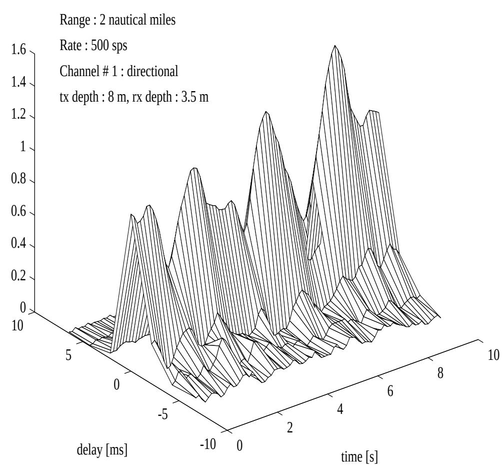

Figure 4: Ensemble of medium-range channel responses in shallow water (approx. 20 m) near the coast of New England, during the month of February. Carrier frequency is 15 kHz.

multiple FSK system is commercially available with a maximum data rate of 1200 bps. This modem has been used in missions of experimental AUVs for vehicle-to-vehicle communication [7].

Despite the low bandwidth efficiency, noncoherent FSK is a good solution for applications where moderate data rates and robust performance are required. Multicarrier modulation techniques in general offer the possibility to eliminate the need for equalization, and as such present an attractive solution for some systems. A system has recently been implemented [8] which uses orthogonal frequency division multiplexing (OFDM) realized with DFT-based filter banks. This system was used on a medium-range channel; however, due to the high frequency separation among the channels (only every fourth channel is used) and relatively long guard times (10 ms guard following a 30 ms pulse), the effective data rate is only 250 bps.

With the goal of increasing the bandwidth efficiency of an underwater acoustic communication system, research focus over the past years has shifted towards phase-coherent modulation techniques, such as PSK (phase shift keying) and QAM (quadrature amplitude modulation).

Depending on the method for carrier synchronization, phase-coherent systems fall into two categories: differentially coherent and purely phase-coherent. The advantage of using differentially encoded PSK (DPSK) with differentially coherent detection is the simple carrier recovery it allows; however, it has a performance loss as compared to coherent detection. Most of the existing systems employ DPSK methods to overcome the problem of carrier phase extraction and tracking. Real-time systems have been implemented mostly for application in vertical and very short range channels, where little multipath is observed and the phase stability is good.

Deep ocean, vertical path channel is used by an image transmission system [10]. This is a 4-DPSK system with carrier frequency of 20 kHz, capable of achieving 16 kbps bottom to surface transmission over 6500 m. The field tests indicate the achievable bit error rates on the order of 10^{-4} with linear equalizer operating under a a least mean squares (LMS) algorithm. Another example of a successfully implemented system for vertical path transmission is that of an underwater image and data transmission system [11]. This system uses a binary DPSK modulation at a rate of 19.2 kbps and a carrier of 53 kHz. The system was used for transmission over 2000 m.

In the very short range channel, where bandwidth in excess of 100 kHz is available, and signal stability is good, phase-coherent system based on 16-QAM has been used [9]. This system operates over 60 m at a carrier frequency of 1 MHz and a data rate of 500 kbps. It is used for communication with an undersea robot which performs maintenance of a submerged platform. A linear equalizer, which uses an LMS algorithm, suffices to reduce the bit error rate from 10^{-4} to 10^{-7} on this channel.

For high-speed communication over longer distances, bandwidth-efficient systems based on phase-coherent signaling methods must allow for considerable ISI in the received signal. These systems employ either some form of array processing, or equalization methods, or a combination thereof, to compensate for the distortions. As an alternative, direct-sequence spread-spectrum has also been used to provide immunity to multipath propagation [15]; however, this technique requires excessive bandwidth.

Array processing has been used both at the transmitter and at the receiver end to isolate a single propagation path and thus eliminate, or at least alleviate the problem of ISI. To excite only a single path of propagation, very large transmitter arrays are required. Instead, the use of parametric sources has been extensively studied [16]. These sources achieve high directivity by relying on the nonlinearity of the medium in the vicinity of a transducer where two or more very high frequencies from the primary projector are mixed highly directive sources rely are mixed, and the resulting signal at difference frequency is transmitted by a virtual array formed in the water column in front of the projector. A major limitation of such a source is in its high power requirements. High directivity implies the problem of pointing errors, and careful positioning is required to ensure absence of multipath. These systems have been employed in short-range shallow water channels where equalization is not deemed feasible due to rapid time-variation of the signal. Binary and quaternary DPSK signals were used achieving data rates of 10 kbps and 20 kbps, respectively, with a carrier frequency of 50 kHz. The estimated bit error rate was on the order $10^{-2} - 10^{-3}$, depending on the range. It was found that this technique is more effective at shorter ranges.

Adaptive beamsteering at the receiver end provides another alternative for reducing signal dispersion in time. The beamformer [17] uses an LMS algorithm to adaptively steer nulls in the direction of a surface reflected wave. The system was tested in shallow water, with DPSK signals transmitted at 10 kbps, and a carrier frequency of 50 kHz, showing a bit error rate of 10^{-2}. Similarly as in the case of a parametric source, it was found that the beamformer encounters difficulties as the range increases relative to depth.

Current state-of-the art in high-speed communications using purely phase-coherent detection of PSK and QAM signals over severely time-spread horizontal channels is based on simultaneous multichannel processing and equalization aided by explicit phase synchronization [18]-[20]. The adaptive multichannel decision-feedback equalizer (DFE) which uses a recursive least squares (RLS) algorithm was tested in a variety of underwater channels, showing satisfactory performance regardless of the link geometry. The achieved data rates of up to 2 kbps over long range channels, and up to 40 kbps over shallow water medium-range channels, are the highest reported to date. In the following section, these methods are discussed in detail.

2 SIGNAL PROCESSING FOR HIGH-SPEED COMMUNICATIONS

In many of the underwater acoustic channels multipath structure may exhibit one or more components which carry the energy similar to that of the principal arrival. As the time progresses, it is not unusual for these components to exceed in energy the principal arrival (e.g., see Fig.2). The fact that the strongest multipath component may not be well defined makes the extraction of carrier reference a difficult task in such a channel.

Receiver Structure

The optimal receiver for multichannel detection in a time-dispersive Gaussian noise channel consists of a bank of filters, one per receiver sensor, each of which is matched to the overall channel response between the transmitter and that sensor [19]. The filter outputs are coherently combined and sampled at the signaling rate. The resulting discrete-time signal represents a set of sufficient statistics for detecting the sequence of transmitted data symbols. The optimal, maximum likelihood sequence estimation (MLSE), even when efficiently implemented using the Viterbi algorithm, has prohibitively high computational complexity to accommodate the ISI span of many of the underwater channels. Linear equalizer on the other hand leads to a computationally simple solution, but lacks the capability to deal with high spectral distortions which are often found in underwater channels. Between these two extreme solutions is the decision-feedback equalizer, which we choose because it offers a good trade-off between performance and complexity [24]. Experimental results have justified this choice.

The optimal receiver gives rise to the structure of an adaptive receiver shown in Fig.5. The input signals to the baseband processor are the A/D converted received signals, translated to baseband using nominal carrier frequency and lowpass filtered. The signals are frame-synchronized prior to any processing. Frame synchronization is accomplished by matched filtering to a known channel probe (a short sequence with good autocorrelation prop erties). The operations of demodulation and frame synchronization can be performed digitally. A signaling frame is shown in Fig.6. It consists of the channel probe, a pause, and a data block which starts with a training sequence. Frame synchronization is performed at the start of each frame, and it provides coarse alignment in time for the duration of a frame. In a mobile scenario, where bit timing may change significantly over the duration of a frame, the channel probe may be used to obtain a coarse estimate of the Doppler shift based on which the signal is resampled. Resampling is efficiently implemented using polyphase filters [22].

The overall baseband channel response as a function of delay τ at time t can be written as

$$f_k(\tau,t) = h_k(\tau,t)e^{j\theta_k(t)}, \quad k = 1, \dots K \tag{1}$$

so as to explicitly indicate the carrier phase $\theta_k(t)$ in each of the channels, and the more slowly varying part of the response $h_k(\tau,t)$. The received signal in the k^{th} channel at time t is then modeled as

$$v_k(t) = \sum_n d(n)h_k(t-nT,t)e^{j\theta_k(t)} + \nu_k(t) \tag{2}$$

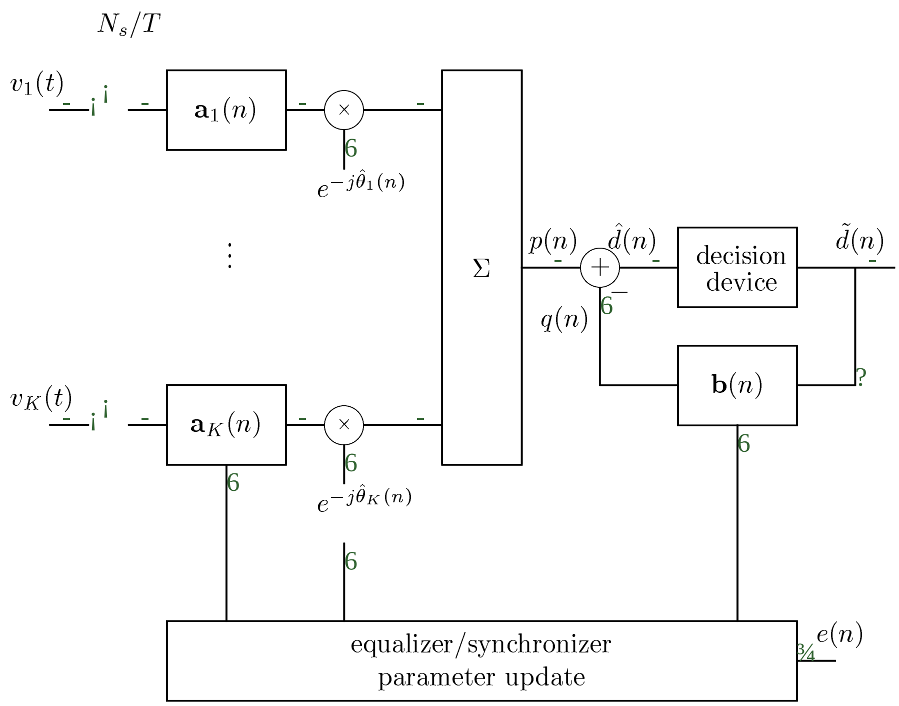

Figure 5: Adaptive multichannel DFE.

where $d(n)$ is the nth transmitted data symbol, and $\nu_k(t)$ is the additive noise.

Knowledge of the channel responses, required by the optimal receiver to implement the matched filters, involves knowledge of the multipath structure, propagation delays and carrier phases. Of these three, carrier phase is the most rapidly changing parameter in the underwater acoustic channel. In classical communication systems, in which carrier and bit synchronization are performed separately from equalization, the presence of strong time-varying multipath affects the performance of a synchronization subsystem, resulting in poor phase tracking capabilities. The residual phase fluctuations on the other hand, impair the equalizer's performance. The equalizer taps, which are complex-valued, may begin to rotate in an attempt to compensate for the residual frequency offset. However, since the

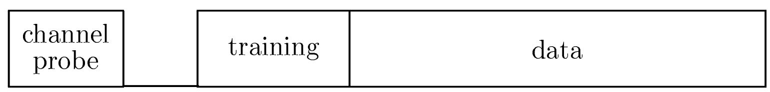

Figure 6: Signaling frame.

rate of convergence of the equalizer tap update algorithm is normally lower than the rate at which the carrier phase changes, the convergence is not achieved. A possible solution to this problem is to jointly perform synchronization and equalization, and it appears to be particularly suitable for the underwater acoustic channels with severe multipath. Basic theoretical aspects of simultaneous multiparameter optimization and data detection can be found in [23].

Baseband processing is performed on signals sampled with as few as 2 samples per symbol interval ($N_s = 2$), since the signals are shaped at the transmitter to have maximal frequency less than $1/T$. Raised-cosine spectrum shaping can be used for this purpose, resulting in a maximal frequency of $(1+\alpha)/2T$, where $\alpha \leq 1$ is the roll-off factor. Since there is no feedback to the analog part of the receiver, the method is suitable for an all-digital implementation.

The sampled signals are processed in the bank of feedforward fractionally spaced equalizers with tap weight vectors denoted by $\mathbf{a}_k$. Sampling is performed starting at an arbitrary time instant, and we assume a sampling rate of $2/T$ without loss of generality. For applications where transmitter and receiver are not moving, but only drifting with water, no explicit adjustment of the sampling clock is needed. It will implicitly be accomplished during the process of adaptive fractionally spaced equalization. The front section of the equalizer will also perform adaptive matched filtering and linear equalization. The output of the feedforward filters is produced once per symbol interval, and since the fractionally spaced equalizers have the capabilities of analog filters they implicitly account for symbol synchronization [24].

Following the feedforward filters is the multichannel carrier phase synchronizer. To correct for the carrier offset, the signals in all channels are phase-shifted by the amount $\hat{\theta}_k$ estimated in the process of joint equalization and synchronization. Depending on the particular channel characteristics, it may not be necessary to have a separate phase-locked loop (PLL) for each of the diversity branches if there is sufficient coherency between the carrier phases in different channels. To accommodate the possibility of large differences in time-varying Doppler frequency shifts observed at different receiver sensors, we consider a general case with as many phase estimates as there are diversity branches.

After coherent combining, the signal is processed by the decision-feedback section of the receiver. A single feedback section is needed because the same information is transmitted in all the channels. The feedback filter with the tap-weight vector $\mathbf{b}$ forms an estimate of the ISI resulting from the previously transmitted symbols (postcursors), which is then cancelled from the linearly processed signal to form an estimate of the transmitted data symbol, based on which the final decision is made.

This receiver structure is applicable to any linear modulation format, such as M-PSK, or M-QAM, the only difference being in the way in which symbol decision is performed. In addition to combining and equalization, signal processing at the receiver may include the operation of decoding if the signal at the transmitter was encoded.

Parameter Optimization

Having established the receiver structure, we can proceed to determine the optimal values of its parameters. The optimization criterion we use is the minimum mean-squared error (MSE) between the estimated data symbol $\hat{d}(n)$ and the transmitted symbol $d(n)$. The receiver parameters to be determined are the tap weights of the multichannel feedforward equalizer, feedback equalizer coefficients, and the carrier phase estimates. In general, there are two ways of computing the equalizer parameters. One is the direct estimation of the equalizer coefficients driven by the output error, and the other is their computation from the estimated channel impulse response. We shall focus on direct estimation.

Assuming the constant channel impulse response and carrier phase in some short interval of time, one arrives at the optimal values of equalization and synchronization parameters. Let the k^{th} channel feedforward equalizer tap weight vector be

$$\mathbf{a}'_k = [a^k_{-N_1} \cdots a^k_{N_2}]^* \tag{3}$$

where the tap weights are taken as conjugate for convenience of notation. At time nT, associated with the detection of the nth data symbol, the input signal samples stored in the k^{th} feedforward equalizer are represented by the vector

$$\mathbf{v}_k(n) = [v_k(nT + N_1T/2) \cdots v_k(nT - N_2T/2)]^T \tag{4}$$

where $N = N_1 + N_2 + 1$ is the number of feedforward taps per channel, and N_1, N_2 are determine to provide the best centering of the signal within the feedforward equalizer, i.e. the highest correlation between the equalizer output and the desired data symbol.

The output of the k^{th} feedforward equalizer, after phase correction by the amount $\hat{\theta}_k$, is given as

$$p_k(n) = \mathbf{a}_k' \mathbf{v}_k(n) e^{-j\hat{\theta}_k} \tag{5}$$

and the coherent combination of all diversity channels is

$$p(n) = \sum_{i=k}^{K} p_k(n). \tag{6}$$

The feedback filter coefficients are arranged in a vector

$$\mathbf{b}' = [b_1 \cdots b_M]^* \tag{7}$$

and the vector of M previous decisions, currently stored in the feedback filter, is denoted as

$$\tilde{\mathbf{d}}(n) = [\tilde{d}(n-1) \cdots \tilde{d}(n-M)]^T. \tag{8}$$

The output of the feedback filter is now defined as

$$q(n) = \mathbf{b}' \tilde{\mathbf{d}}(n). \tag{9}$$

The estimate of the data symbol at time n is

$$\hat{d}(n) = p(n) - q(n) \tag{10}$$

from which the decision $\tilde{d}(n)$ is obtained as the closest signal point. The resulting estimation error is

$$e(n) = d(n) - \hat{d}(n). \tag{11}$$

The receiver parameters are optimized based on joint minimization of the MSE

$$E = E\{|e^2(n)|\} \tag{12}$$

with respect to $\{\mathbf{a}_k\}$, $\mathbf{b}$, and $\{\hat{\theta}_k\}$.

To find the optimal values of the equalizer coefficients, it is convenient to group all the coefficients into a composite vector $\mathbf{c}$, and to express the estimate $\hat{d}(n)$ as

$$\begin{aligned} \hat{d}(n) &= [\mathbf{a}_1' \cdots \mathbf{a}_K' \ \ -\mathbf{b}'] \begin{bmatrix} \mathbf{v}_1(n) e^{-j\hat{\theta}_1} \\ \vdots \\ \mathbf{v}_K(n) e^{-j\hat{\theta}_K} \\ \tilde{\mathbf{d}}(n) \end{bmatrix} \\ &= \mathbf{c}' \mathbf{u}(n). \end{aligned} \tag{13}$$

The MSE can now be expressed as a function of the composite equalizer vector $\mathbf{c}$, as

$$\begin{aligned} E &= E\{|d(n) - \mathbf{c}'\mathbf{u}(n)|^2\} \\ &= R_{dd} - 2Re\{\mathbf{c}'\mathbf{R}_{ud}\} + \mathbf{c}'\mathbf{R}_{uu}\mathbf{c} \end{aligned} \tag{14}$$

where we have used the notation $\mathbf{R}_{xy} = E\{\mathbf{x}(n)\mathbf{y}'(n)\}$ for the crosscorrelations. The value of $\mathbf{c}$ that minimizes the MSE is given by

$$\mathbf{c} = \mathbf{R}_{uu}^{-1}\mathbf{R}_{ud}. \tag{15}$$

The optimal values of the estimates of the carrier phases, $\hat{\theta}_k$, are most easily found if the estimate $\hat{d}(n)$ is represented as

$$\begin{aligned} \hat{d}(n) &= p_k(n) + \sum_{j\neq k} p_j(n) - q(n) \\ &= \mathbf{a}_k'\mathbf{v}_k(n)e^{-j\hat{\theta}_k} + \pi_k(n). \end{aligned} \tag{16}$$

The second term in the last expression is independent of $\hat{\theta}_k$, which makes it possible to express the MSE as

$$\begin{aligned} E &= E\{|d(n) - \pi_k(n) - \mathbf{a}_k'\mathbf{v}_k(n)e^{-j\hat{\theta}_k}|^2\} \\ &= -2Re\{\mathbf{a}_k'E\{\mathbf{v}_k(n)[d(n) - \pi_k(n)]^*\}e^{-j\hat{\theta}_k}\} \\ &+ \text{ terms independent of } \hat{\theta}_k \end{aligned} \tag{17}$$

The optimal values $\hat{\theta}_k$ satisfy the gradient equations

$$\frac{\partial E}{\partial \hat{\theta}_k} = -2Im\{\mathbf{a}_k'E\{\mathbf{v}_k(n)[d(n) - \pi_k(n)]^*\}e^{-j\hat{\theta}_k}\} = 0, \quad k = 1, \ldots, K. \tag{18}$$

In order to track the time-varying optimal solution for the receiver parameters, equations (15), (18) should be solved recursively, using updated values of possibly time-varying crosscorrelations. Updating can be carried out continuously, i.e. symbol-by-symbol, or in a block-adaptive manner, in which the parameters are updated only during short training blocks.

The simplest form of an adaptive algorithm is the combination of LMS algorithm for the equalizer coefficients update, and the first-order stochastic gradient update for the digital PLL. Such an algorithm, however, often fails on an underwater acoustic channel, primarily due to the poor phase tracking capabilities. Improved phase tracking capabilities result from the use of a second-order stochastic gradient update for the phase estimates. The instantaneous estimate of the MSE gradient (18) is proportional to

$$\Phi_k(n) = Im\{\mathbf{a}_k'\mathbf{v}_k(n)[d(n) - \pi_k(n)]^*e^{-j\hat{\theta}_k}\} \tag{19}$$

This quantity represents an equivalent of the k^{th} phase detector output. Using the fact that

$$d(n) - \pi_k(n) = p_k(n) + e(n) \tag{20}$$

the expression (19) is rewritten as

$$\Phi_k(n) = Im\{p_k(n)[p_k(n)+e(n)]^*\} = Im\{p_k(n)e^*(n)\}, \quad k = 1, \ldots K. \tag{21}$$

The second-order phase update equations are given by

$$\hat{\theta}_k(n+1) = \hat{\theta}_k(n) + K_{f_1}\Phi_k(n) + K_{f_2}\sum_{m=0}^{n}\Phi_k(m), \quad k = 1, \ldots K. \tag{22}$$

It is assumed here that the same proportional and integral tracking constants K_{f_1}, K_{f_2} are used in all diversity channels, and that perfect loop integration is used. Alternative tracking strategies are of course possible. Also, an analogous derivation holds for the case of a single phase estimate for all the channels.

The equalizer coefficients, ideally given by (15), are computed using a suitable adaptive algorithm, driven by the error $e(n)$ and the input $\mathbf{u}(n)$, computed using the current values of the phase estimates. In the training mode, the data symbols $d(n)$ are taken from a known training sequence, while in the decision-directed mode, the values $\tilde{d}(n)$ are used to update the receiver parameters.

Choice of the Adaptive Algorithm

Although the underwater acoustic channels are generally confined to low data rates as compared to many other communication channels, the encountered channel distortions require complex signal processing methods, resulting in high computational load which may exceed the capabilities of the available programmable DSP platforms. Consequently, choosing an efficient adaptive algorithm for receiver implementation, be it for array processing, equalization or both, is an important system design task.

In a majority of recent studies, the LMS-based algorithms are considered due to their low computational complexity, which is linear in the total number of coefficients N [13],[17], [29]. However, the LMS algorithm has a convergence time which may become unacceptably long when large adaptive filters are used (20 N as opposed to 2 N of the RLS algorithm). The total number of coefficients may be very large (more than 100 taps is often needed for spatial and temporal processing in medium and long-range shallow water channels). In addition, the LMS algorithm is very sensitive to the choice of step-size. To overcome this problem, self-optimized LMS algorithms have been used [29], but this results in increased complexity, and increased convergence time.

RLS algorithms, on the other hand, have better convergence properties but higher computational complexity. The quadratic complexity of the standard (form II) RLS algorithm [25] is too high when large adaptive filters need to be implemented. In general, it is desirable that the algorithm be of linear complexity, a property shared by the fast RLS algorithms. A numerically stable fast RLS algorithm [26] has been used for the multichannel equalizer tests in off-line processing, the results of which are shown in Sec.2. A later version of a fast and modular RLS algorithm is given in [27].

Despite its quadratic complexity, a square-root RLS algorithm [28] has been used for real-time implementation because of its excellent numerical stability. The advantage of this algorithm is that it allows the receiver parameters to be updated at arbitrary time instants, rather than every symbol interval, as required by the fast algorithm of [26]. It thus reduces the computational load per each detected symbol. In addition, the updating intervals can be determined adaptively, based on monitoring the mean squared error. Such adaptation methods are especially suitable for use with high transmission rates, where long ISI requires large adaptive filters, but eliminates the need to update the receiver parameters every symbol interval. A different class of adaptive filters, which also have the desired convergence properties and numerical stability, are the lattice filters. RLS lattice algorithms are described in [30].

Reduced-Complexity Receiver Structures

Regardless of the adaptive algorithm used, its computational complexity is proportional to the number of receiver parameters (tap-weights). In addition to focusing on low-complexity algorithms, one may search for a way to reduce the receiver size. Although the use of spatial combining reduces residual ISI and allows shorter equalizers to be used, a broadband combiner may still require a large number of taps to be updated, limiting the practical number of receiving channels to a few. Besides the increase in computational time, very large adaptive filters, which must operate with computationally efficient algorithms, imply increased sensitivity to numerical errors. Unfortunately, some of the fast RLS algorithms [26] preserve numerical stability only at the expense of sacrificing the tracking speed. Another disadvantage of large multichannel equalizers, and perhaps the critical one, lies in their increased noise enhancement, which significantly limits the gain obtained by increasing the number of input channels.

These issues motivate the search for a different multichannel processing strategy in which the size of the adaptive filter will be reduced, but multipath diversity gain preserved. One way to reduce the complexity of the multichannel equalizer of Fig.5 is by the method of pre-combining. Suppose that a large number of input channels K, say more than 10, is available. If each of the K channels is followed by a 50 tap feedforward adaptive filter, the total number of coefficients in the feedforward combining section is more than 500. A reduced-complexity multichannel DFE is shown in Fig.7. It uses a a pre-combiner to reduce a large number of input channels K to a smaller number P for subsequent multichannel equalization. More than one channel at the output of the combiner is usually required if the full diversity gain is to be preserved, but this number, which depends on the type of the transmission channel, is often small.

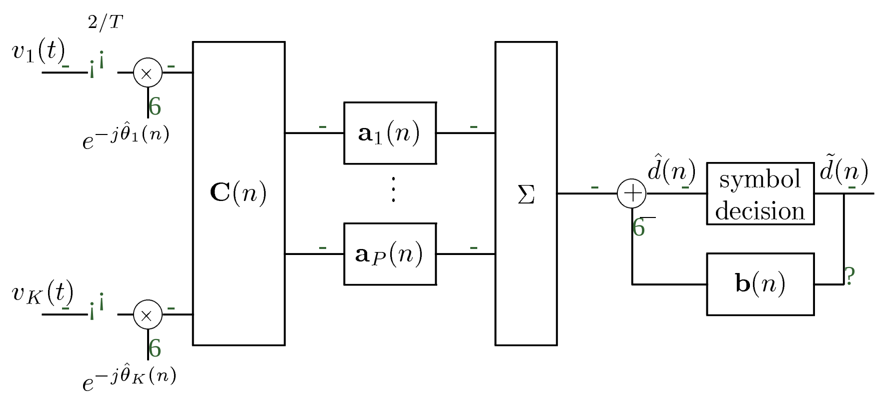

Figure 7: Adaptive multichannel DFE with pre-combining.

The fact that diversity gain may be preserved is explained by multipath correlation across the receiver array. To preserve this processing gain, the $K \times P$ coefficients of the transformation $\mathbf{C}$ are determined jointly with the rest of receiver parameters in a manner analogous to that of Sec.2. The details of receiver optimization can be found in [20]. The adaptive receiver version is based on parallel use of two adaptive algorithms, one for the pre-combiner and one for the P-channel equalizer, where both algorithms are driven by the common error signal $e(n)$. When there is a large number of input channels, and long feedforward equalizers are needed, reduction in the number of adaptively adjusted feedforward taps ($KP + PN$ instead of

KN) can be significant. The choice of P can be made to achieve the least degradation in performance with respect to the full complexity K-channel equalizer, while still keeping the receiver complexity at an acceptable level. We shall illustrate this method shortly by an example.

Another approach in the design of efficient receiver structures is to focus on reducing the number of equalizer taps. A conventional equalizer is designed to span all of the channel response. However, if the channel is characterized by several distinct multipath arrivals separated in time by intervals of negligible reverberation, an equalizer may be designed to have fewer taps. Such method, termed sparse equalization, was applied to detection of QPSK signals recorded in the Arctic waters, for which results show an order of magnitude reduction in computational load [31]. By reducing the number of adaptively adjusted parameters, this approach also makes it possible to use simple updating algorithms, such as standard RLS algorithms which have good numerical stability. Finally, in channels which are naturally sparse, discarding the low-magnitude equalizer taps in fact results in improved performance since no unnecessary noise is processed.

Experimental Performance Analysis

The method of adaptive multichannel combining and equalization was demonstrated to be effective in a variety underwater channels with fundamentally different mechanisms of sound propagation. These channels include the long and medium-range, deep and shallow water channels of Figs.2-4. We shall examine the receiver performance on a long-range shallow water channel.

The experiment we shall describe was conducted by the Woods Hole Oceanographic Institution in the region of New England Continental Shelf, in May of 1992. The transmission ranges were between 15 and 65 nautical miles, and the receiver was positioned in about 50 m deep water with a vertical array of 20 sensors spanning depths from 15 to 35 m. The transmitter power was 193 dB re μPa, and a carrier frequency of 1 kHz was used, with a transducer bandwidth of 1 kHz. At 48 nautical miles (approximately 90 km) an ensemble of estimated channel responses is shown in Fig.3. The channel response consists of a fairly stable main arrival followed by extremely long, unstable multipath. The modulation formats were QPSK, 8-QAM and 8-PSK, and the symbol rates were varied from 1 to 1000 symbols per second.

The signaling frame, as shown in Fig.6, was used with the channel probe consisting of a 13 element Barker code transmitted in phase and in quadrature at the data rate. The signals were shaped using a cosine roll-off filter with roll-off factor α=0.5, digitally implemented by truncating the impulse response to ± 2 symbol intervals.

Fig.8 shows the results of signal processing for the transmission rate of 500 sps. The modulation format is 8-PSK, resulting in the bit rate of 1500 bps. Three channels are combined in this example, and they are numbered according to depth, starting from the one closest to the surface. The upper left plot shows the snapshots of the three channel responses, as estimated from the preamble. There is substantial coherence between the channel response magnitudes due to the low separation between the array elements. The ISI is such that the scatter plot of unequalized, symbol rate sampled received signal is completely smeared, even if the signal is phase-synchronized.

The receiver parameters are indicated in the figure: N and M denote the number of feedforward and feedback coefficients, respectively, f.f. denotes the forgetting factor of the RLS algorithm, which accounts for the exponential weighting of the past data, and K_{f_1}, K_{f_2} denote the tracking constants of the PLLs. P_e is the fraction of erroneous decisions in the processed data block. When a single-channel equalizer is used, the scatter plot of estimated data symbols $\hat{d}(n)$ is shown in the upper right corner. Although the single-channel algorithm converges, the performance is not very good, with an estimated probability of error on the order of 10^{-2}. Performance of the multichannel equalizer with $K = 3$ channels is illustrated in the remaining three plots. The estimated MSE indicates steady convergence in decision-directed mode of operation. The number of training symbols required by the RLS algorithm is about twice the total number of adaptively adjusted coefficients. Phase estimates for the three channels are shown together, after each has been scaled to remove the constant frequency-offset term (the average of the Doppler shifts is indicated in the plot). Combining the three channels eliminates all the decision errors and results in about 3 dB better output SNR than in the case of a single-channel equalizer. The multichannel gain normally increases with the number of input channels, but is limited by residual ISI, the noise enhancement in the feedforward equalizers, and the amount of diversity available in the given channel.

As the data rate is increased, these limitations begin to dominate the receiver performance. At 1000 sps, which was the maximum data rate available in the given frequency range, the extent of ISI becomes high enough to prevent successful operation of a single-channel equalizer. A five channel equalizer opens the eye in the output scatter plot, but each feedforward filter needs close to a hundred taps. The reduced-complexity multichannel equalizer offers a solution in this case. Fig.9 shows the result of QPSK signal processing using a $K = 7$ to $P = 3$ pre-combiner followed by a three-channel DFE. Excellent performance, with a clearly open eye and no detected errors is achieved. A very interesting conclusion is also drawn from this study, regarding the receiver performance as a function of the reduced number of channels P. With the total number of input channels fixed to $K = 7$, the reduced number of channels P was varied, and in each case the steady state MSE was observed. With $P = 1$, the performance is poor. The output SNR (the inverse of the MSE) initially increases rapidly with an increase in the number of channels P. However, after a certain value, in this case $P = 3$, the performance saturates, and the SNR retains the same value with further increase in P. In other words, the receiver achieves the full ($K = 7$ in this case) multichannel processing gain at a complexity determined by the number of equalizer channels $P = 3$. The optimal value of P is a parameter inherent to the transmission channel and the system configuration.

Theoretical performance analysis of the adaptive multichannel equalizer requires knowledge of the channel and noise statistics. Under the assumption of a Rayleigh fading channel, an approximate analysis is possible, and it is discussed in [21]. This analysis, as well as the experimental results, confirms the fact that low signaling rates at which ISI may be negligible are not a good choice for rapidly varying channels. On such channels, choosing a higher signaling rate results in better performance through improved tracking capability of the adaptive algorithm.

Related Applications

The principles of multichannel equalization are not limited to the problem of ISI suppression, but apply to a general problem of interference suppression. The sources of interference in underwater acoustic communications include external interference, such as noise coming from on-board machinery or the underwater vehicle launch noise. The internal noise has signal-like characteristics, and it arises in the form of echo in full-duplex systems, and as multiple-access interference in underwater communication networks. In configuration as a noise canceler, the reference of external interfering signal is fed to one of the multichannel receiver inputs, while neither the receiver structure, nor its algorithm change. Performance results with band-limited white noise and multiple sinusoidal interference [32] show the receiver's effectiveness in cancelling the interference while simultaneously detecting the desired signal, which is achieved by virtue of having the training sequence.

A multiple-access communication system represents a special case of structured interference environment. In this situation, two or more sources transmit to a common receiver overlapping in both time and frequency. Since the bandwidth is so scarce in underwater acoustic channels, little processing gain is used to separate the different users' signals by code-division.

The fact that transmission loss varies significantly with range additionally contributes to the near-far effect in the underwater acoustic channels. Two categories of multiuser receivers that have been considered are the centralized receiver, at which the signals of all the users have to be detected (e.g., up-link reception at a surface buoy which serves as a central network node), and the decentralized receiver, at which only the desired user's signal needs to be detected (e.g., down-link reception by an ocean-bottom node). The adaptive multichannel receiver of Fig.7 was experimentally shown to have excellent capabilities in the role of a decentralized multiuser detector, operating without the knowledge of the interfering signal. A multiuser multichannel equalizer, based on the principles of joint equalization, synchronization and multiple-access interference cancellation was successfully tested in experimental trials [33]. Array processing plays a crucial role in the detection of multiuser signals, but is associated with the problem of computational complexity.

Receiver Implementation

Digital technology for underwater acoustic communication systems implementation is a natural choice, having in mind the operating frequencies of these systems. Current converter technology can easily digitize signals which are in the frequency range of interest to underwater acoustic communications, with enough dynamic range to support required receiver selectivity. Another advantage of digital processing is that receiver structures can be easily replicated and tolerance problems associated with analog designs are eliminated. This is of particular importance for operations like array processing which may be delay-sensitive. In addition, digital processing of the signal can compensate for non-ideal operation of the analog front-end.

Programmable DSP platforms currently support operations of more than 100 Mips (mega instructions per second) in fixed-point implementations and comparatively high speeds in floating-point implementations. In addition, multiple DSP boards are commercially available having much greater processing power, and so are specialized processors which support computationally extensive operations specific to communications signal processing.

The advantage of DSP implementation is that it offers great flexibility necessary in early stages of modem development. It also allows easy reconfiguration and changes in the system design (modulation format, equalization algorithm, coding scheme). Small series and the lack of standards also make custom-specific digital circuits very expensive to manufacture for underwater acoustic applications. Although the DSP solutions require more power, this is not a major limitation since various power-reduction techniques, such as putting the transmitter/receiver into sleep mode, can be applied on the system level.

The receiver algorithm described in this section has been implemented on a programmable DSP platform at the Woods Hole Oceanographic Institution. The DSP board, completed in January 1997, has since been under experimental evaluation, in both stationary and mobile communication scenarios. An earlier implementation provided QPSK signaling at 5 kbps, and a carrier frequency of 15 kHz, and was successfully tested in the under-ice shallow water environment [34]. In the current configuration, the modem can support up to eight input channels.

The size of the board is 8 in $\times$ 3.5 in $\times$ 1.7 in. The board contains eight AD converters, each with sampling capability programmable to up to 200 kHz and 13 bits of resolution. Two output channels have 12-bit DA converters with sampling rate of 500 kHz each.The DSP is a TI320C44, with processing speed of 60 Mflops. The programmable instrument interface includes serial and parallel capabilities and a high-speed data download/upload capability. The memory consists of 6 Mbytes of static RAM and 1 Mbyte of flash RAM. The modem uses a real-time operating system, Acoustic Modem System (AMS), which was developed at the Woods Hole Oceanographic Institution.

In the transmit mode, the modem requires 30 W when coupled to a 180 dBreμPa source. In the receive mode, the power requirement is 8 W, for active reception. Power is supplied by a lithium battery pack, which fits into a cylindrical pressure housing, approximately the size of the board. The modem is powered up into active reception mode by any of a number of wake-up mechanisms. These mechanisms include external activity on the serial port, real-time alarm clock, battery-low signal, or acoustic wake-up by low-power energy detection of a specially encoded FSK sequence. In the passive mode, modem can either be in hibernation, in which it uses only 1 mW of power from a smaller battery and takes about half a second to wake up, or in sleep mode, in which it uses about 20 mW and can wake up instantly. The battery life-time is approximately 2 years in the passive mode, or 2 days of continuous active reception.

Demodulation is performed digitally. Currently, the modem is programmed to use a single second-order PLL for the active channels, with an option for using a multichannel digital PLL. The adaptive filter sizes may be externally determined or chosen autonomously based on the channel mesurements obtained from the Barker probe signal. The adaptive algorithm is a square root RLS, with on-line determination of the updating intervals. The complexity of the algorithm is 9 N^2 per update.

3 AREAS OF FURTHER DEVELOPMENT

At this stage in the development of underwater acoustic communication techniques, with the feasibility of high-rate communications established, a number of research topics emerged which will influence the design of future systems. Some of these topics we have already discussed, namely the reduced-complexity receiver structures and algorithms to enable **efficient receiver implementations**. Beside theoretical advances, future implementations that may require lower power consumption and cost, possibly targeting ASIC (application-specific integrated circuits) realizations of computationally extensive modem functions (coding, equalization) while other will remain in DSP (speech coding, network related functions and functions that require flexibility). Below, we identify several areas of communication engineering in which further developments are likely to provide improved capabilities of future high-speed underwater acoustic communication systems.

Signal Processing

As we have seen at the beginning of this chapter, underwater acoustic communications find application in a variety of autonomous systems. If the adaptive receiver algorithms are to be used in such systems, external assistance in system operation must be minimized. In the first place, receiver initialization involves adjustment of receiver parameters, such as tracking constants in the adaptive algorithm that we have described. These parameters are adjusted according to the instantaneous channel conditions before the actual signal detection can begin. In addition, an increase the noise level, caused for instance by a passing ship, may temporarily disable the communication. The receiver then must wait for the next available training sequence in order to be re-initialized. Two directions are to be pursued towards the development of self-optimized systems. One is further development of **adaptive algorithms** which have the ability to adjust their tracking parameters to fine channel changes. Another is the development of **blind system recovery** techniques, not only for equalization, but also for array processing and synchronization that can operate without a training sequence or a reference signal.

In addition, **speech coding** techniques and **data compression** algorithms suitable for low-contrast underwater images, as well related speech and **image processing** methods, are expected to enable high-rate information transmission over band-limited underwater acoustic channels.

Modulation and Coding

Achieving high throughputs over band-limited channels is conditioned on the use of bandwidth-efficient modulation and coding techniques [24]. Today, results for underwater acoustic communications are confined to signaling schemes whose bandwidth efficiency is at most 3 to 4 bps/Hz. Consequently, there is a need to investigate the performance of higher-level modulation methods. The major concern in doing so is the sensitivity of high-level signal constellations to both phase jitter and time-varying intersymbol interference, which are abundant in the majority of underwater acoustic channels. A possible solution to this problem is the use of special **signal constellation shaping**, which provides more robustness to channel impairments than the conventional PSK or QAM. One method of generating such signals is by differential encoding of both phase and amplitude, which results in high-level constellations that permit differentially coherent demodulation, thus eliminating the need for explicit phase tracking and gain control.

Trellis-coded modulation is well suited for vertical channels which have minimal dispersion. Their use on the horizontal channels requires further investigation because it is associated with the problem of efficient decoding in the presence of long ISI. Namely, the delay in decoding poses problems for an adaptive equalizer which relies on the feedback of instantaneous decisions. Maximum-likelihood sequence estimation, i.e., joint channel estimation and data detection is likely to require implementation of a reduced-complexity sequence estimation algorithm, a number of which is documented in contemporary literature. The design and application of suitable methods for joint equalization and decoding to underwater acoustic channels merits further investigation.

Error correction coding will provide the basis for increasing the reliability of poor-quality underwater acoustic channels. Adequate coding methods, as well as efficient decoding depend on the particular channel and modulation technique used. For example, concatenated codes may be considered for time-varying multipath fading underwater channels.

Channel Modeling and Simulation

Performance analysis of candidate system design methods is subject to the availability of an accurate channel model. While there exists a vast knowledge of both deterministic and statistical modeling of sound propagation underwater, the implications this knowledge bears on the communication channel modeling has only recently received more attention. To date, there is no single statistical channel model widely accepted for any of the underwater acoustic channels. The statistical channel measurements which describe the channel behavior on a time scale of interest to high-speed communication systems are scarce, and focus exclusively on stationary communication scenarios. In a mobile underwater acoustic channel, vehicle speed will be the primary factor determining the time-coherence properties of the channel, and consequently the system design.

Mobile Underwater Communications

The problem of channel variability, already present in applications with a stationary transmitter and receiver, becomes a major limitation for a mobile underwater acoustic communication system. The ratio of the vehicle speed (a fast vehicle moves at several tens of knots, or a nautical miles per hour) to the speed of sound (1500 m/s) exceeds its counterpart in the mobile radio channels by several orders of magnitude, making the problem of time-synchronization very difficult in the underwater acoustic channel. Apart from the carrier phase variation, the mobile underwater acoustic systems will have to deal with the motion-induced pulse compression and dilation. In addition to frame-by-frame signal resampling, to compensate for the motion-induced distortion at high vehicle speeds, algorithms for continuous tracking of the time-varying symbol delay in the presence of underwater multipath are being considered.

Communication Networks

The design of underwater communication networks is characterized by the bandwidth limitation and the long propagation delays in these channels. Due to the bandwidth limitation, frequency-division multiple-access is possible only to a limited extent. Time-division multiple-access, on the other hand, is associated with the problem of efficient protocol design, which arises because of the long propagation delays. As we have already mentioned, a possible solution in such a situation is a system based on code-division multiple-access; however, here again the designer is faced with the problem of low processing gains due to the bandwidth limitation. Consequently, two areas of research are being pursued towards the development of underwater acoustic communication networks. One is the design of network and data link protocols, suited for long propagation delays and strict power requirements encountered in the underwater environment [36]. Another area is that of multiuser detection which allows simultaneous detection of multiple users' signals, thus reducing the need for re-transmissions. Finally, if and when mobile users are enabled to access an underwater network, mobile networking will have to be considered in the light of underwater acoustic channels.

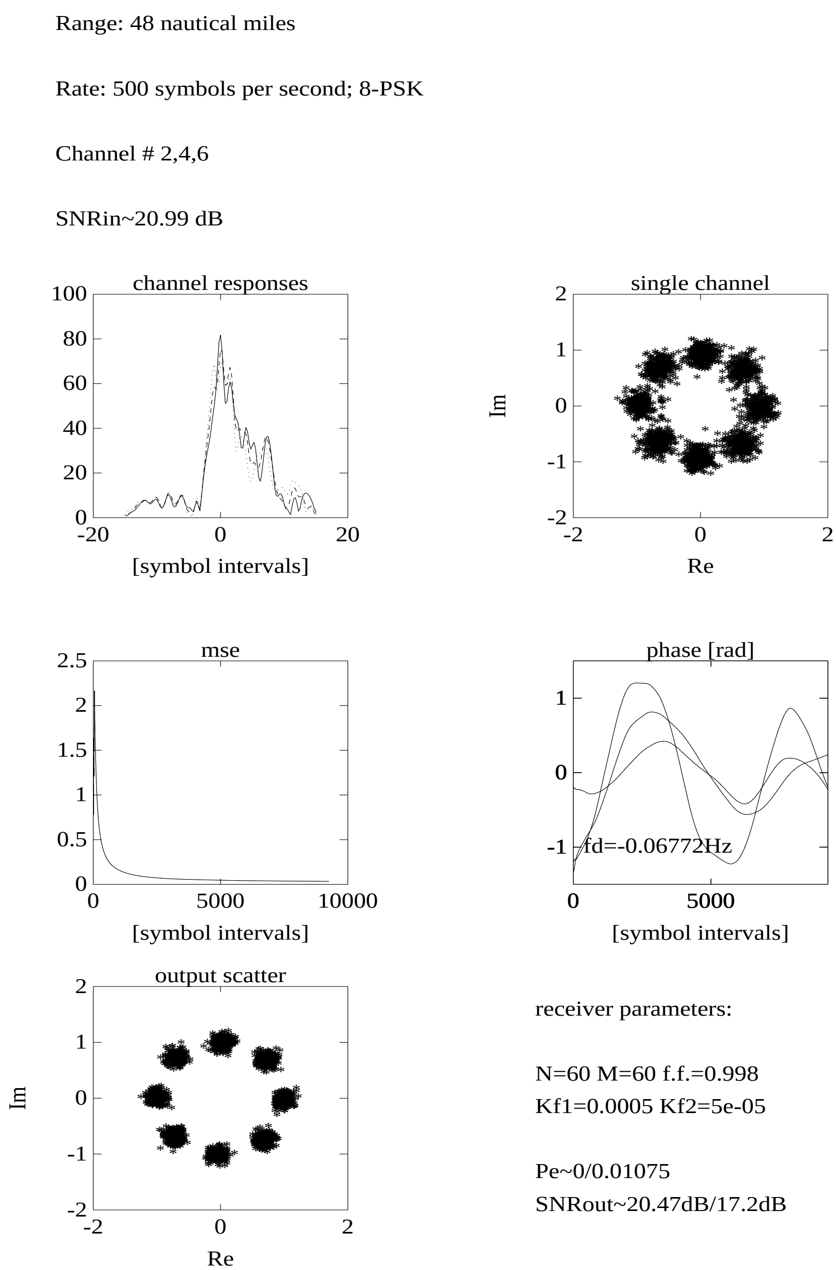

Figure 8: Results of signal processing using adaptive multichannel DFE ($K = 3$). Experimental data from New England Continental Shelf.

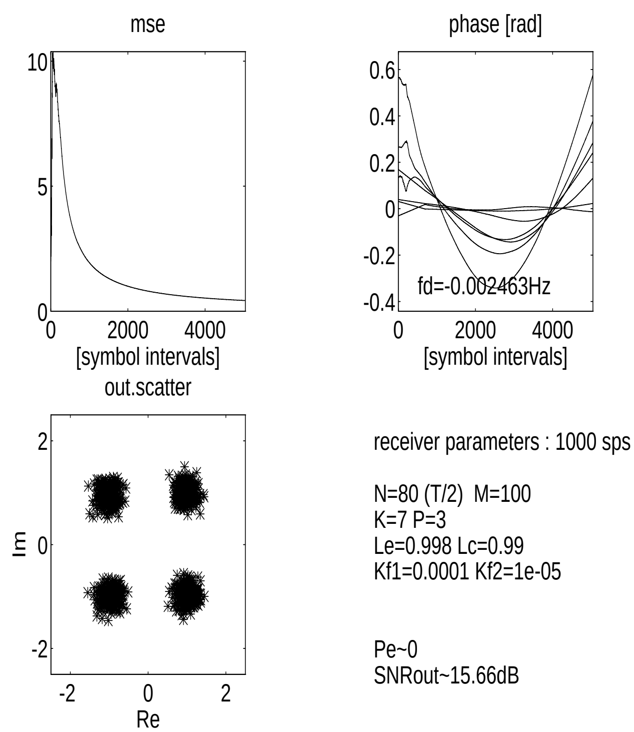

Figure 9: Results of signal processing using reduced-complexity adaptive multichannel DFE ($K = 7, P = 3$). Experimental data from New England Continental Shelf.

BIBLIOGRAPHICAL NOTES

Reference [1] is a classical text which contains a profound treatment of underwater sound propagation and can serve a communications engineer unfamiliar with the propagation characteristics of underwater acoustic channels. Review articles [2]-[5] provide basic concepts of underwater acoustic communications. These articles also contain extensive reference lists for more information on the current state of research and development in the field. Specific examples of communication system design for various applications, such as telemetry, image and speech transmission are given in references [6]-[17], most of which are recent conference publications. Articles [18]-[23] treat system design and performance analysis for high-speed communications based on phase-coherent detection methods. They also contain a number of experimental results. Fundamental principles of digital communications and adaptive filtering are presented in classical texts [24] and [25]. References [26]-[30] contain adaptive algorithms which are of interest to the problems at hand. Finally, references [31]-[36] treat diverse topics in high-speed underwater communications, ranging from efficient modem implementation issues to future underwater data networks.

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