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Experimental demonstration of phase-coherent underwater acoustic communications using a compact array

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ABSTRACT

High-rate phase coherent acoustic communication systems generally require large arrays to obtain spatial diversity. Large arrays may not be suitable for small underwater platforms such as underwater unmanned vehicles (UUVs). In this paper, acoustic communication utilizing a compact array with closely spaced elements was demonstrated using the data collected during the August 2010 lake experiment. In addition, a bidirectional multichannel decision feedback equalizer (DFE) was developed to enhance the performance of communication systems. The recorded data from three communication ranges of 25 m, 920 m, and 1 300 m were processed to verify the system performance. The results showed an improvement in output symbol signal-to-noise ratio (OSNR) of approximately 3.4 dB, and 4.9 dB for bidirectional two-channel DFE and four-channel DFE over a single-channel DFE, respectively.

1. Introduction

With the increasing demand for ocean exploring, environmental monitoring, and scientific data collection, reliable and high-rate underwater acoustic (UWA) communication techniques for underwater devices will play an important role in the future. The challenge of high-rate UWA communication lies primarily on the unique characteristics of UWA channels, such as severe attenuation, limited bandwidth, time-varying long spread multipath propagation, channel fading, strong background noise, and the Doppler effect (Kilfoyle and Baggeroer, 2000).

Phase coherent single carrier adaptive time-domain decision feedback equalizers (SC-TDE) with an embedded second-order digital phase locked loop (DPLL) have been investigated for decades. The receiver can simultaneously track carrier phases as well as mitigate multipath channel distortions (Stojanovic et al., 1994). To compensate for the effect of channel fading, multichannel receivers are widely used in high-rate phase coherent UWA communication systems (Stojanovic, 2008; Pajovic and Preisig, 2015).

The benefit of spatial diversity in improving the output symbol signal-to-noise ratio (OSNR) was demonstrated in many experiments (Yang, 2007; Zhang and Dong, 2011; Jamshidi and Moezzi, 2015). Theoretical analysis indicates that, to ensure the fading on each element approximately independent, the optimal element spacing is about the signal coherence length (Pajovic and Preisig, 2015). For vertical arrays, the optimal element spacing is on the order of 3–4 wavelengths, while the coherence length is on the order of 30–60 wavelengths for horizontal arrays (Yang and Heaney, 2012). This requirement restricts application of conventional spatial diversity on small underwater platforms. In the past few years, small size receivers, such as vector sensors, have been considered for phase coherent underwater acoustic communication (Song et al., 2011; Han et al., 2015).

In this paper, we proposed phase coherent underwater acoustic communications using a compact array with closely spaced elements to significantly reduce receiver size. The sensor separation of the compact array is about 1/10 wavelength. A compact array with such spaced elements is generally used as a super-gain hydrophone array to improve detection performance of passive sonar. The small sensor separation in compact arrays reduce spatial diversity gain, and a bidirectional multichannel DFE (BiMCE) was proposed for compact arrays to obtains additional diversity. The DFE suffers from error propagation caused by the feedback of incorrect decisions. To mitigate error propagation and improve the performance of a conventional DFE, bidirectional DFE was first introduced in wireless communication (Balakrishnan and Johnson, 2000), and then refined for complementary code keying (CCK) UWA

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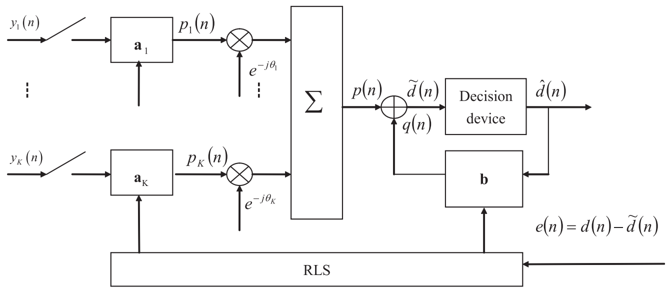

Fig. 1. Block diagram of multichannel decision feedback equalizer with DPLL.

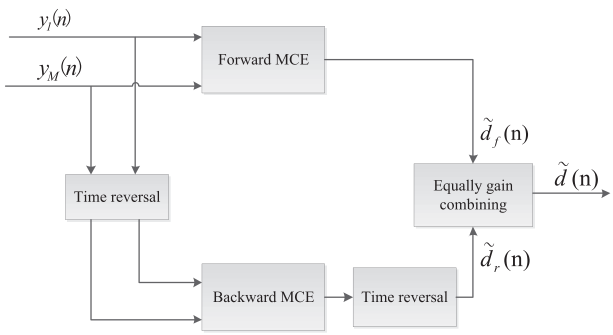

Fig. 2. Structure of bidirectional multichannel DFE.

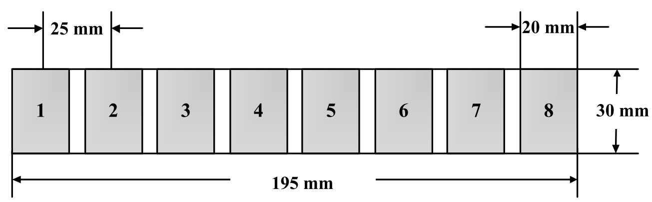

Fig. 3. Structure of the horizontal compact array with closely spaced elements.

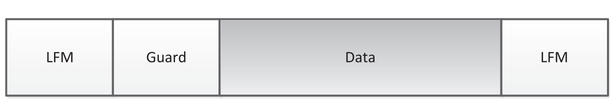

Fig. 4. Transmitted signal structure.

communication in our previous work (He et al., 2009). It was also adopted for time-reversal underwater acoustic communication (Song, 2012) and single carriers with frequency domain equalizers (He et al., 2012). The proposed BiMCE consists of two parallel MCE structures, one to equalize the received multichannel signal in a causal fashion and the other equalize the time-reversed version of the received multichannel signal in a non-causal fashion.

The reminder of this paper is organized as follows. Multichannel DFE with the DPLL algorithm is reviewed in Section 2. Section 3 discusses the proposed method. Section 4 illustrates the experimental configuration, analyzes the recorded data, and presents the experiment results. Conclusions are given in Section 5.

2. Adaptive multichannel DFE

In conventional SC-TDE communication, transmitted signals are corrupted by multipath and noise interference. To combat channel fading, spatial diversity is achieved by multiple spatially separated sensors. The structure of the multichannel DFE (MCE) and the adaptive

Table 1
Receiver parameters.

Parameters	Description	Values
f_s	Sampling frequency	20kHz
B	Bandwidth	4 kHz
T_c	Symbol duration	0.5 ms
M	Total sensor numbers	8
K	Oversampling rate	2
N_p	The training symbol length	1000
N_f	Feedforward filter order	20/100
N_b	Backward filter order	20/80
K_{f_1}	Proportional tracking constants in PLL	0.001
K_{f_2}	Integral tracking constants in PLL	0.001
λ	RLS forgetting factor in DFE	0.999
Range	Communication range	20, 920, 1 300 m
Rate	Communication rate	2 kbps
d	sensor separation	λ/10

algorithm widely used in UWA communication are reviewed in this section.

The transmitted signal is represented in baseband form as

$$u(t)=\sum_n d(n)g(t-nT), \tag{1}$$

where $d(n)$ are the M-ary phase shift keying (MPSK) modulated data symbols transmitted every T seconds, and $g(t)$ is the transmitter pulse shape filter.

The received signal at the input to the k^{th} equalizer branch is modeled as

$$y_k(t)=\sum_n d(n)h_k(t-nT,t)e^{j\theta_k(t)}+w_k(t), k=1, \quad \ldots \quad K, \tag{2}$$

where h_k (τ,t) is the overall channel response of the k-th channel (including physical channels and transceiver filters), $w_k(t)$ is the additive white Gaussian noise, and θ_k is the phase rotation caused by symbol timing offset and Doppler shift. Without loss of generality, we assume a sampling rate of 2/T for the received signal.

The MCE is effective for removing the ISI induced by multipath propagation. When the channel is unknown, the equalizer tap-weights are determined by minimizing the mean square error (MSE) of the input received data symbols and the recovered data symbols. Then, channel tracking is accomplished by the adaptive algorithm combined recursive least squares (RLS) and a second-order DPLL (Stojanovic et al., 1994).

The structure of the MCE is shown in Fig. 1. It consists of a bank of adaptive feedforward filters, one per receiving sensor, followed by a decision feedback filter. The feedforward filter coefficients a_k arranges in vector $\mathbf{a}_k$, while the feedback filter coefficients b arranges in vector $\mathbf{b}$. The coefficients of the forward and feedback filters are iteratively updated and phase compensation is required separately for each sensor. The order of feedforward and feedback filters are M and N, respectively. The input received signal for the feedforward filter is

$$\mathbf{y}_k(n)=[y_k(n) \quad y_k(n-1) \quad ,\ldots, \quad y_k(n-M+1)]^T, \tag{3}$$

where$[\cdot]^T$ denotes the matrix transpose. The MCE embedded with the DPLL algorithm is described as follows:

Step 1: The feedforward filtering on the sequence of M input symbols, and phase compensation are performed once per symbol, yielding:

$$p(n)=\sum p_k(n), \tag{4}$$

where

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