Physical Communication 18 (2016) 40–48

Contents lists available at ScienceDirect

Physical Communication

journal homepage: www.elsevier.com/locate/phycom

Full length article

DSP based real-time single carrier underwater acoustic communications using frequency domain turbo equalization

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ARTICLE INFO

Article history:
Received 25 March 2015
Received in revised form 9 October 2015
Accepted 10 November 2015
Available online 2 December 2015

Keywords:
Turbo equalization
DSP
Underwater acoustic communication

ABSTRACT

In this paper, a DSP based real-time single carrier underwater acoustic communication system is developed. The designed receiver system operates in frequency domain for low complexity equalization. The turbo principle is incorporated into channel equalization and channel estimation processes to improve the system reliability. The channel is treated as stable within each small block and iteratively estimated based on LMS method and soft information provided by channel decoder. The soft LMS channel estimation is able to achieve better performance than conventional phase compensation scheme in single carrier frequency domain system. The receiver system is implemented on a floating point DSP running at 350 MHZ. Through careful optimization, the real-time processing is achieved with sufficient margin for both single-hydrophone receiver and multi-hydrophone systems. The designed system has been tested by both tank experiment and sea trials and achieves satisfied performance.

1. Introduction

In recent years, there has been an increasing interest in the development of high performance underwater acoustic communication networks for various underwater applications. One of the key issues for the success of the underwater acoustic networks is to build up a reliable and high performance physical underwater acoustic communication link.

To develop reliable and high performance underwater acoustic communication link, various techniques have been proposed and investigated [1]. Recently, orthogonal frequency-division multiplexing (OFDM) and single carrier frequency domain equalization (SC-FDE) have been extensively studied for underwater communication due to their low computational complexity. OFDM transmission with two-stage Doppler compensation has been proposed for underwater communication for single-input multiple-output (SIMO) [2] and multiple-input multiple-output (MIMO) [3] situations. Compared with OFDM transmission, SC-FDE exhibits the property of low peak to average power ratio due to its single carrier transmission nature while the low complexity frequency domain equalization could reduce the complexity of the system. The investigation of SC-FDE for underwater acoustic communication has been carried out in [4] for SISO situation and in [5] for MIMO situation.

To further improve the reliability of underwater communication, iterative turbo detection technique where the extrinsic information from channel decoder is feedback to the equalizer has been widely investigated in underwater acoustic communication applications [6–9]. The adaptive time domain turbo equalization for SISO and MIMO underwater communication was proposed in [6]. In [7], a MIMO time domain turbo equalizer with hybrid soft interference cancellation and reliability ordering was proposed for single carrier communication. A frequency domain turbo equalizer with successive interference cancellation and phase correction was investigated in [8] for

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http://dx.doi.org/10.1016/j.phycom.2015.11.001
1874-4907/

single carrier MIMO transmission. Iterative receivers for distributed MIMO OFDM were investigated in [9].

The above mentioned works are carried out in the view of underwater acoustic communication research and real-time implementation is not considered, which makes those works still far away from practical underwater acoustic communication system. Some work has been done on the development of the underwater acoustic modem [10–16]. The designs in commercial products [10–12] are either based on non-coherent FSK detection [10,12] or direct spread spectrum technique [11]. All of them operate in a non iterative manner and inherently have low band efficiency. [16] considered the design of low cost modem and the system was also based on FSK. The Micro-Modem in [13] can operate in two modes: (1) low data rate mode based on non-coherent FSK detection and (2) high data rate mode based on PSK. But the time domain decision feedback equalizer is used and operates in non iterative way which may suffer performance degradation due to error propagation. [14] investigates the implementation of DSP based SISO and MIMO OFDM underwater modem and the receiver also operates non-iteratively. The DSP implementation of turbo equalizer has been studied in [17]. The equalizer operates in time domain while the taps of equalizer are calculated in frequency domain. However, the channel state information is assumed to be known and not updated during iteration process. The underwater acoustic modem with turbo equalizer is also investigated in [15]. However, the time domain equalizer inherently exhibits higher computational complexity than its frequency domain counterpart.

Our work focus on the issue of real-time implementation of frequency domain turbo equalizer for underwater acoustic communication. We design a real-time underwater acoustic communication receiver based on a digital signal processor (DSP). In this paper, we present our receiver structure and DSP implementation issues. To verify the system performance, both tank and sea experiments are carried out. Experiments show that we can achieve real-time processing. The BERs achieved in the experiments are below 5×10^{-3} for single-hydrophone system and less than 10^{-3} for two-hydrophone system.

The rest of the paper is organized as follows. The transmitter scheme and receiver system are introduced in Sections 2 and 3 respectively. The DSP implementation issues are introduced in Section 4 and Section 5 presents the results and analysis. Finally, Section 6 gives some conclusions.

Notation. Upper-case (resp. lower-case) bold letters represent matrices (resp. column vectors) with $[A]_{n,m}$ (resp. a_n) denoting the (n, m)th (resp. nth) entry of $\mathbf{A}$ (resp. $\mathbf{a}$); $(\cdot)^T$, $\mathrm{conj}(\cdot)$ and $(\cdot)^H$ denote transpose, conjugate and Hermitian transpose operators, respectively. $\mathbf{F}_N$ and $\mathbf{I}_N$ represent N by N normalized FFT matrix and N by N unit matrix respectively. e_i is a column vector with ith entrance being 1 and other entrances being 0 and $\mathbf{1}_G$ is a G by 1 column vector with all the entrances being 1. $\tilde{\mathbf{a}}$ denotes the FFT of $\mathbf{a}$, $\hat{a}$ represents the estimation of a, $\bar{a}$ is the expectation of a, tri($\mathbf{A}$) means the trace of matrix $\mathbf{A}$ and diag ($\mathbf{A}$) denotes the diagonal of matrix $\mathbf{A}$. $\otimes$ and $\odot$ denote Kronecker product and Hadamard product.

Table 1
Symbol mapping.

00	01	10	11
$\frac{-1-i}{\sqrt{2}}$	$\frac{-1+i}{\sqrt{2}}$	$\frac{1-i}{\sqrt{2}}$	$\frac{1+i}{\sqrt{2}}$

2. Transmitter

The information bits a_k are divided into blocks with block length N, then the information bits in each block are encoded by a rate $\frac{1}{2}$ convolutional code with the generation polynomial as $[1\,21/37]_8$. The coded bits $c_k[m]$ are interleaved by a random interleaver, and then mapped to QPSK symbols $u_k[m]$ with respect to the given constellation set $\mathbb{S} = \{\alpha_q\}_{q=1}^4$ as in Table 1 for the mth block. A PN sequence v_k is cascaded in the tail of each block as unique words (UW). The length of UW is enough long to avoid inter block interference (IBI). Therefore, the transmitted baseband signals for the mth data block can be denoted as $\mathbf{s}[m] = \left[\mathbf{u}[m]^T \mathbf{v}^T\right]^T$. The generated transmitted baseband signals are pulse shaped by a raised cosine filter and upconverted to passband. The diagram of transmitter is shown in Fig. 1.

3. Receiver

At receiver, the received passband signal is first downconverted to baseband. In practical communication, the received signal usually suffers carrier frequency offset (CFO) due to Doppler effect and clock drifting between transmitter and receiver. To compensate CFO, three identical pilot data blocks with block duration of N_c symbols are transmitted. The choice of N_c should be larger than the channel delay spread. Therefore, CFO f_d is estimated as [18]

$$\hat{f}_d = \frac{1}{\pi N_c} \sum_{i=N_c}^{2N_c-1} \arg\left\{x[i]\mathrm{conj}\left(x[i+N_c]\right)\right\}, \tag{1}$$

where $x[i]$ is the received signal corresponding to the CFO pilot. After CFO compensation and with the quasi-static channel assumption i.e. channel is stable within every receiving block, the CFO compensated time-domain baseband signals within kth receiving block are written as

$$\mathbf{y}[k] = \mathbf{H}[k]\underline{\mathbf{s}}[k] + \mathbf{w}[k] \tag{2}$$

where $\mathbf{y}[k]$ is composed of the CFO compensated baseband signals in the kth block and $\mathbf{H}[k]$ is a circulant matrix with the first column as

$$\left[h[0;\ k] \cdots h[L-1;\ k]\ 0 \cdots 0\right]^T \tag{3}$$

where $h[i;\ k]$ represents the baseband channel impulse response (CIR) in the kth block and $\mathbf{w}[k]$ is the zero mean Gaussian white noise with variance $\sigma_0^2[k]$. $\underline{\mathbf{s}}[k]$ is composed of transmitted baseband signals considering the oversampling factor G. More specifically, $\underline{s}_i[k]$ is defined as

$$\underline{s}_i[k] = \begin{cases} s_{\frac{i}{G}}[k] & \text{if } \dfrac{i}{G} \text{ is integer} \\ 0 & \text{else.} \end{cases} \tag{4}$$