Adaptive Self-interference Cancellation for Full-duplex Underwater Acoustic Communications using Coded OFDM

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Abstract: *Presently, most underwater acoustic communications (UAC) systems operate using the half-duplex mode. Due to the limited available acoustic bandwidth and transceiver front-end saturation induced by overlapping transducer and hydrophone frequency responses, the transmission and reception need to be separated in time. This in turn leads to inefficient use of time-frequency resources and increased latency. This paper considers an UAC system with in-band full-duplex operation using adaptive cancellation techniques that consider power amplifier non-linearity to remove self-interference (SI). Furthermore, the proposed system utilizes quasi-cyclic low-density parity check codes (QC-LDPC), quadrature phase shift keying (QPSK) modulation and orthogonal frequency division multiplexing (OFDM) to overcome noise and multipath impairments. The performance is evaluated by simulation and using experimental data acquired during sea trials in the North Sea, UK, and demonstrates successful detection of remote transmissions in the presence of SI, which is attenuated by over 90 dB after cancellation.*

Keywords: *Full-duplex, Self-interference Cancellation, Coded OFDM, QC-LDPC.*

1. INTRODUCTION

There are various approaches to establish a communication link between two nodes, e.g. using a cable, or wirelessly using electromagnetic (EM), optical or acoustic waveforms. The latter approach is the only feasible and flexible way to efficiently convey information over short and long distances underwater. Commercial underwater acoustic communication (UAC) systems available today have to combat two key channel impairments to achieve successful communication between two users/nodes, i.e. multipath and Doppler. To combat multipath induced inter-symbol interference (ISI), a decision feedback equalizer (DFE) with timing and phase correction has been used successfully in the past for single-carrier multichannel systems [1]. The concept has been extended to use forward error correction (FEC) to exploit correct decision feedback and iterative decoding demonstrating improved performance at lower signal-to-noise ratio (SNR) levels [2]. Alternatively, ISI can be combated using multicarrier systems, such as orthogonal frequency division multiplexing (OFDM) that converts the frequency-selective multipath into narrowband frequency non-selective channels, which in turn enables the use of an efficient one-tap equalizer per subcarrier [3]. OFDM is typically used in conjunction with FEC to provide frequency diversity and improve performance in the presence of movement induced Doppler [4].

Over the last two decades, UAC has been the subject of intensive research, however, due to the limited bandwidth of the UAC channel, the focus was concentrated mostly on the half-duplex (HD) mode of operation using time division duplexing (TDD). Thus, commercial UAC systems available today are inherently HD-TDD based. This implies that transmission and reception can not overlap in time or frequency. Recently, researchers focused their attention on full-duplex (FD) systems [7]. However, due to the severe bandwidth limitation that is intrinsic to the UAC channel, it is not feasible to achieve high data rate links using full-duplex (FD) through frequency division multiplexing. Therefore, the only approach to implement FD is by allowing transmissions to overlap in both frequency and time and utilize digital or analog cancellation (C) techniques to combat self-interference (SI) [6], [8].

In this paper, we present an FD-UAC system that utilizes quasi-cyclic low-density parity check codes (QC-LDPC), quadrature phase shift keying (QPSK) modulation and orthogonal frequency division multiplexing (OFDM) to overcome noise and multipath impairments. To overcome SI, a SIC method is proposed that taps the attenuated signal from the output of the power amplifier to train the weights of the SIC filter. The performance is demonstrated using experimental data from sea trials in the North Sea of the coast of UK.

2. TRANSCEIVER MODEL

The generic block diagram of the transceiver structure for the proposed coded QC-LDPC OFDM FD system is shown in Fig. 1. The information to be transmitted is first encoded using a Quasi-cyclic LDPC (QC-LDPC) code and interleaved to protect against potential burst channel errors. Subsequently, the interleaved bits are group into pairs of 2 bits and mapped to a Gray-encoded quadrature phase shift keying (QPSK) modulation to improve spectral efficiency in the transmission. Following modulation, the QPSK symbols are grouped in blocks of dimension N and an inverse fast Fourier transform (IFFT) is applied to the block to transfer the signal into to time domain, where a cyclic prefix is appended to protect the OFDM symbol samples from inter-block interference (IBI). The CP length is typically selected to be longer than the expected multipath delay spread of the channel. Finally, the OFDM symbol samples are pulse-shaped and frequency up-converted to a carrier frequency of f_c to exploit the transducer resonance and improve its transmission efficiency. The OFDM waveform is then amplified by an audio power amplifier that drives the transducer. The transmit power is adjusted so that the required signal-to-noise ratio (SNR) is satisfied to achieve the expected bit error rate (BER) of the system.

Quasi-cyclic LDPC (QC-LDPC) codes are a special category of low-density parity-check (LDPC) codes. LDPC codes are linear block error-correcting codes constructed using parity-check matrices that are sparse in nature. This implies that the number of '1' entries per row is low compared to the dimension of the parity-check matrix. QC-LDPC codes have been shown to perform very well in terms of error correction capabilities, and are used in modern communications systems, including wireless RF (5G, IEEE 802.11n, 802.16e) and optical (ITU-T G.975.1, ITU-T G.709) communication systems, and hard-disk and solid-state drive (HDD, SSD) storage systems. QC-LDPC codes exhibit a specific structure that enables more efficient decoding than generic LDPC codes. The parity-check matrix of a QC-LDPC code can be expressed as a circulant permutation of a smaller sub-matrix, which makes it quasi-cyclic. This structure allows the decoding algorithm to be implemented using fast Fourier transform (FFT) techniques, which can significantly reduce the computational complexity of the decoding process [5]. This is particularly useful for OFDM systems that have already an FFT implementation available.

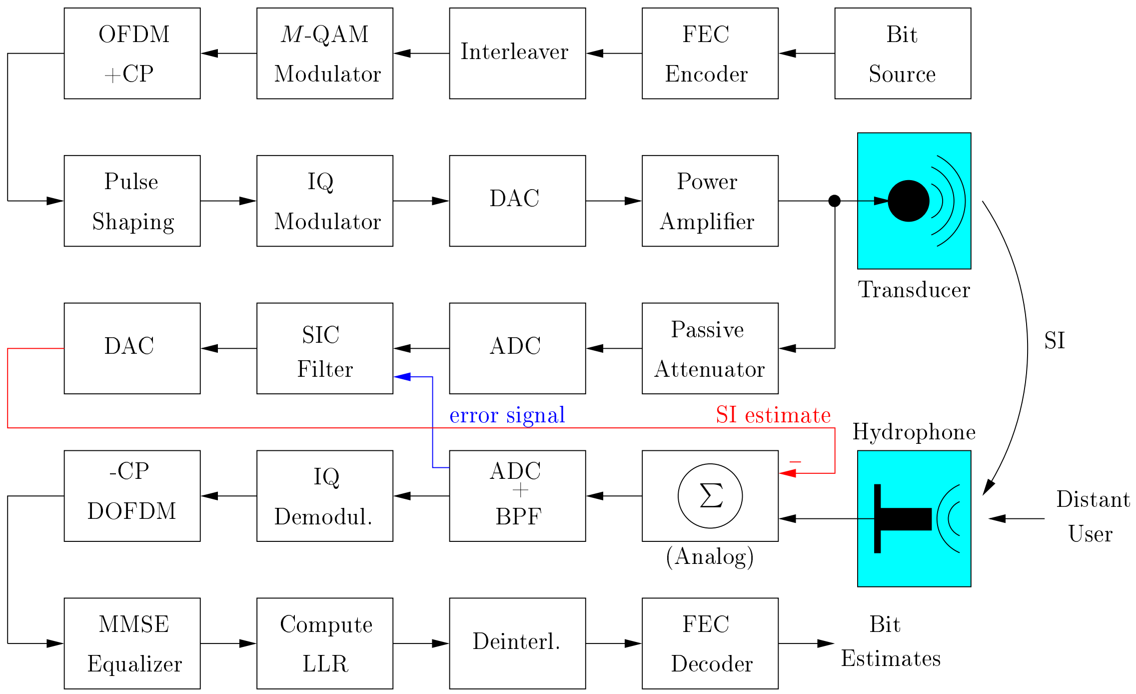

Figure 1: Generic block structure of the proposed FD transceiver.

3. THE SIC FILTER

The SIC filter is a finite impulse response (FIR) filter of length L, which is typically selected to cover the significant part of SI. The filter coefficients, $\mathbf{w}_k$, are optimized by minimizing a cost function. If a least mean squares (LMS) type of algorithm is used, the cost function corresponds to the average mean squared error (MSE), i.e. $J_n = E\{|e_n|^2\}$ [9]. If the normalized

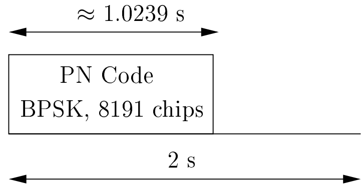

Figure 2: FD training mode.

LMS (NLMS) is used to optimize the SIC filter weights, the cost function $J_n = ||\mathbf{w}_{n+1} - \mathbf{w}_n||^2$ is minimized subject to $\mathbf{w}_{n+1}^H \mathbf{u}_n = d_n$, where d_n is the desired signal, and $\mathbf{u}_n$ is the memory of the FIR filter [10]. NLMS is a constrained optimization type algorithm that tries to minimize the norm of the difference between subsequent filter weights. The computational complexity of these algorithms is low for practical implementation, however, they may suffer by slow convergence. If higher computational complexity can be afforded, a fast implementation of a recursive least squares (RLS) algorithm can be utilized, where the cost function is a weighted sum of the past N errors, i.e. $J_n = \sum_{i=0}^{n} \lambda^{n-i} |d_i - \mathbf{w}_n^H \mathbf{u}_i|^2$ [10]. For SIC filter sizes of $L > 100$, the complexity of the RLS becomes impractical, hence, the fast RLS version can be alternatively used [11], which exploits the fact the only 1 sample in the filter memory changes per iteration. This is leads to a computational complexity of a total of $7L + 14$ operations per iteration and 2 divides [11], which is computationally very attractive for hardware implementations. The reader is referred to the original paper by Cioffi and Kailath [11] for the equations required in the implementation.

3.1. INITIAL TRAINING

The adaptive SIC filter is initially trained using a pseudo-noise (PN) code that is modulated using a binary phase shift keying (BPSK) scheme. The PN code is typically selected to have a good autocorrelation sequence. In practice, an m-sequence is used whose length is selected to be at least as long as the expected local multipath delay spread that causes the SI. Figure 2 shows the signal structure of the initial training required to optimize the coefficients of the adaptive SIC filter. Since the PN code is known, the transceiver can adapt iteratively the coefficients of the adaptive SIC until the mean squared error (MSE) between the known BPSK modulated PN code and the received SI signal is minimized. Prior to being digitised by an analog to digital converter (ADC), the MSE is computed using an analog difference amplifier, i.e. a voltage subtractor, in order to avoid saturation of the ADC from the SI signal. A low-noise, low-distortion, high-bandwidth, rail-to-rail operational amplifier is required (such as the LT1630) with an operating supply voltage range (>12 V) that is much larger than the dynamic range of the ADC (<5 V). It is worth noting that the error signal is tapped before the digital bandpass filter as shown in Figure 1. The adaptive SIC filter is trained for the duration of the PN code and the coefficients are stored at the end of the training for utilization and further adaptation during the FD mode. The normalized least mean square (NLMS) algorithm is used for the optimization of the coefficients whose size is selected to cover up to 1 s of local SI.

4. EXPERIMENTAL RESULTS

Experimental results from seatrials in the North Sea, UK, are presented in this section. Figure 3, 4, and 5 show the location off the coast of Tynemouth, the FDUAC scenario, and the measurement set-up, respectively. The scenario is typical for shallow water channels with the water column depth in the area being $d_w = 48$ m. The Tx transducer and Rx hydrophone were lowered in the water using the same line as shown in Figure 4 at $d_{Tx} = 10$ m and $d_{Rx} = 5$ m, with the Tx at higher depth to avoid interaction with the dynamic surface to the greatest extent. In these arrangement, the toroidal (doughnut like) characteristics in the z-axis are exploited to minimize SI. The signals were recorded using a Zoom F4 audio recorder with a 24-bit ADC precision at a sampling frequency of $f_s = 48$ kHz. The signal after the power amplifier is recorded using a passive attenuator to bring the signal strength to a line level range (1V rms).

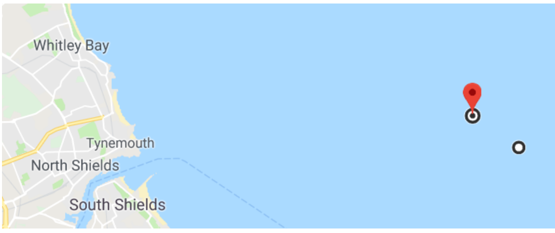

Figure 3: Location: Co-ordinates: (start) 55 00.981,-01 13.185 (stop) 55 01.534, -01 14.627.

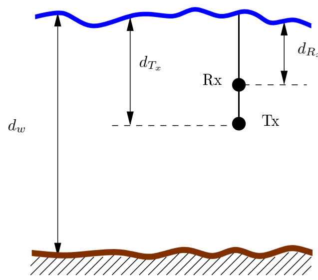

Figure 4: FDUAC scenario: hard bottom, $d_{Tx} = 10$ m, $d_{Rx} = 5$ m, $d_w = 48$ m.

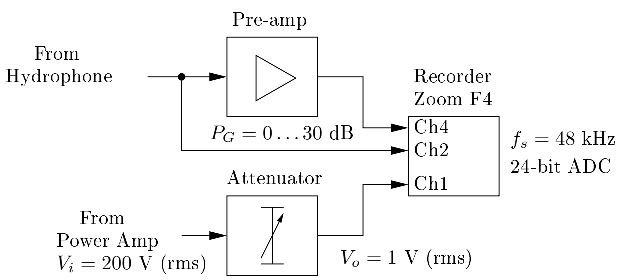

Figure 5: Measurement setup.

Figure 6 shows the magnitude (envelope) a typical SI channel impulse response (CIR) that was obtained using correlation processing between the original clean signal and the received

Parameter	Value
Carrier Frequency f_c (kHz)	12
Bandwidth B (kHz)	8
Number of Subcarriers	2048
Cyclic Prefix	64
Code Rate	1/2

Table 1: System parameters.

signal from pre-amplified channel. In [6] it was demonstrated that the envelope of the SI-CIR, $h_{SI}(\tau)$ can be empirically approximated as

$$|h_{SI}(\tau)| = \frac{0.15}{1 + \frac{\tau}{0.04\ \mathrm{s}}} \tag{1}$$

This is demonstrated in Figure 6, where the envelope of the SI-CIR is shown as red line.

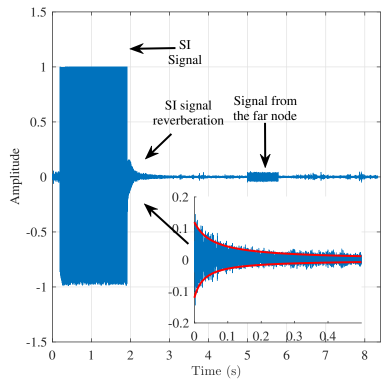

Figure 6: The signal transmitted by local node along with the signal received from the far node.

The noise spectra present in the measured underwater acoustic channel are shown before and after bandpass filtering in Figure 7. The frequency range spans from 0 to half of the sampling frequency, i.e. 24 kHz. A closer look at the figures reveals that the combination of inband SI and ambient noise sets the level of the background at -30 dB. The ambient noise consists of anthropogenic noise (tone at approx. 19 kHz), natural background noise caused by various sources such as rain, wind, waves, and biological activities like fish sounds, marine mammal vocalizations, and snapping shrimp noise. After bandpass filtering the out of band interferences are clearly removed and the signal is prepared for frequency downconversion and as input for the SIC filter.

Figure 8 shows received signal constellations for a user located at 1 km in distance. The constellations are shown before and after zero-forcing (ZF) equalization, and after clipping to prevent numerical instabilities in the log-likelihood ratio (LLR) conversion required by the QC-LDPC decoder. Figure 9 shows the bit error rate over time (packet index) for two FD transmissions at distances of 1 km and 500 m, respectively. It is worth noting that the performance of the 1 km transmission achieves a lower BER due to the shorter multipath spread of the underlying CIRs.

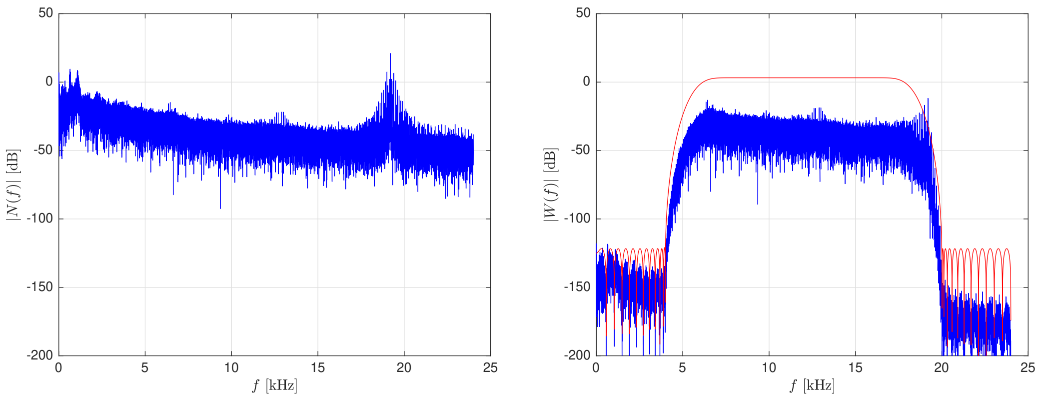

Figure 7: Noise of the raw and filtered received signal during silent period.

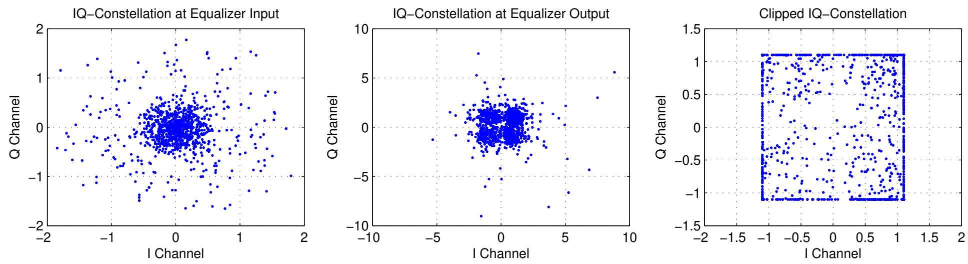

Figure 8: Received signal constellations.

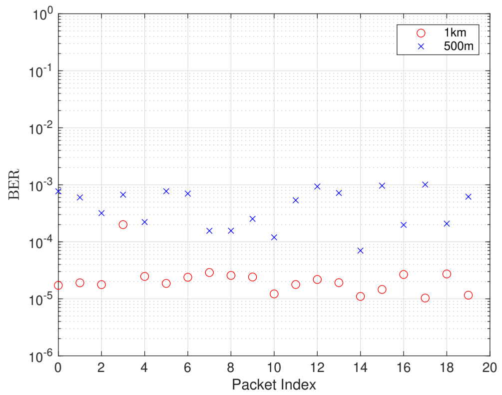

Figure 9: BER over time for 1 km and 500 m FD transmissions.

5. CONCLUSIONS

In this paper, we presented an FD-UAC system and demonstrated its performance using experimental signals from sea trials in the North Sea of the coast of UK. The proposed system utilized QC-LDPC codes, QPSK modulation and OFDM to overcome noise and multipath impairments. To combat SI, a cancellation method was proposed that taps the attenuated signal from the output of the power amplifier to train the weights of the SIC filter. Results demonstrated the performance of the proposed transceiver by decoding successfully distant transmissions from a user located at 1 km in distance. Future work will focus on SIC implementation using infinite impulse response filters (IIR) to reduce the complexity of the long FIR filter implementation and on the development of fast algorithms for the filter weight update of the IIR filters.

6. ACKNOWLEDGEMENTS

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