Parameterized Interference Cancellation for Single-Carrier Underwater Acoustic Communications

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***Abstract*—Underwater acoustic communications provide promising solutions for remote and real-time aquatic exploration and monitoring. However, the underwater environment is rich in various kinds of interferences. Those interferences could severely degrade the acoustic communication performance. This work tackles interference cancellation in a single-carrier modulated communication system. Based on the Nyqusit sampling theorem, the interference is parameterized by a finite number of unknown parameters. The Page test is applied to detect the presence of an interfering waveform in the received signal. An iterative receiver is developed, which iteratively performs the interference estimation/cancellation and traditional receiver processing. The proposed receiver is evaluated when the communication waveform is interfered by the ice-cracking impulsive noise and the sonar signal collected from the Arctic. The data processing results reveal that the proposed receiver achieves considerable decoding performance improvement through the iterative interference estimation and cancellation.**

***Index Terms*—Interference cancellation, impulsive noise, narrowband interference, single carrier, underwater acoustic communications**

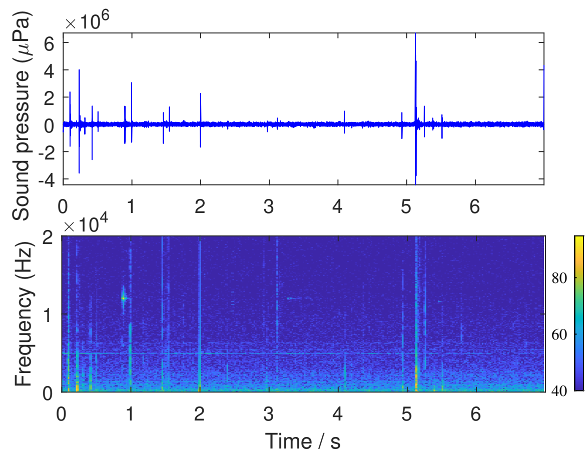

Fig. 1. An example of the impulsive noise collected during the breaking of an ice sheet in Arctic. Top: the time-domain waveform; bottom: the time-frequency spectrum.

I. Introduction

Underwater environments are rich in ambient sound sources. Those sound sources create interferences of various kinds to underwater acoustic (UWA) communications. The external interferences can be divided into two categories according to their time-frequency characteristics: (1) impulsive interference with short time duration and large bandwidth; and (2) narrowband interference with small bandwidth and long time duration. The interferences from, e.g., sonar operations, marine mammals, malicious jamming and natural environment, occur quite often during UWA communications, and lead to significant performance degradation [1]–[3]. Fig. 1 shows an example of the ambient noise collected from the Arctic, which consists of many impulses generated by ice cracking.

Various methods for interference cancellation have been proposed for UWA communications and wireless radio communications. One common method is the thresholding method either in the time domain for the impulsive noise mitigation or in the frequency domain for the narrowband interference mitigation [4]–[6]. Iterative approaches in various forms were developed for interference estimation/cancellation and typical receiver processing (e.g., channel estimation and symbol detection) [5], [6]. Compressive sensing techniques were also applied by exploiting the sparsity of UWA channels and the sparsity of impulsive noise [7]. Statistical methods, such as the sparse Bayesian learning, have also been developed for interference mitigation [8], and with consideration of UWA channel dynamics [9].

Existing interference cancellation algorithms are primarily for the orthogonal frequency-division multiplexing (OFDM) system. Based on an earlier work of interference cancellation for the underwater OFDM [10], this work focuses on the cancellation of interference in a general form for the single-carrier communication system. Different from the thresholding method, the Page test is introduced not only for the interference detection but also for determining the starting time and the ending time of an interfering waveform. Based on the Nyquist sampling theorem, the interference waveform in a general form will be parameterized by a finite number of unknowns. An iterative receiver is then developed for iterative interference estimation/cancellation and symbol detection.

The proposed receiver is validated via data sets collected from the Bohai Gulf, Dalian, China and from the Arctic. The processing results show that the proposed receiver can effectively mitigate the impulsive noise and the narrowband inter-

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ference. Considerable performance improvement is achieved through the iterative interference estimation and cancellation.

II. System Model

A. Single-carrier Transmission

In the single-carrier transmission, the baseband signal in one transmission frame can be expressed as

$$s(t)=\sum_{n=0}^{N_{\mathrm{s}}-1} s[n] g\left(t-n T_{\mathrm{s}}\right), \qquad t \in[0, T], \tag{1}$$

where $\{s[n]\}$ is the information symbol sequence of length N_{s}, $g(t)$ is the pulse shaping waveform, T_{s} is the symbol duration, and $T:=N_{\mathrm{s}} T_{\mathrm{s}}$ is the total waveform duration.

Modulating the baseband signal onto a carrier frequency f_{c} yields the passband waveform

$$\tilde{s}(t)=\operatorname{Re}\left\{s(t) e^{j 2 \pi f_{\mathrm{c}} t}\right\}, \qquad t \in[0, T]. \tag{2}$$

We adopt a path-based model for the underwater channel,

$$\tilde{h}(\tau, t)=\sum_{p=1}^{N_{\mathrm{pa}}} A_{p}(t) \delta\left(\tau-\tau_{p}(t)\right), \tag{3}$$

where N_{pa} is the number of paths, and $A_{p}(t)$ and $\tau_{p}(t)$ are the amplitude and delay of the pth path, respectively. Here we assume that the amplitude does not change during the transmission duration, i.e., $A_{p}(t) \approx A_{p}$, and the path delay can be approximated as $\tau_{p}(t) \approx \tau_{p}$. The channel model can be simplified as

$$\tilde{h}(\tau)=\sum_{p=1}^{N_{\mathrm{pa}}} A_{p} \delta\left(\tau-\tau_{p}\right). \tag{4}$$

In an environment with additive interference, the received passband signal can be written as

$$\tilde{y}(t)=\sum_{p=1}^{N_{\mathrm{pa}}} A_{p} \tilde{s}\left(t-\tau_{p}\right)+\tilde{I}(t)+\tilde{w}(t), \tag{5}$$

where $\tilde{I}(t)$ and $\tilde{w}(t)$ denote the interference and the ambient noise, respectively.

After the bandpass-to-baseband conversion, one can obtain the baseband signal,

$$y(t)=2 \operatorname{LPF}\left[\tilde{y}(t) e^{-j 2 \pi f_{\mathrm{c}} t}\right], \tag{6}$$

where LPF$[\cdot]$ stands for the low-pass filtering. Substituting (2) into (6) yields

$$y(t)=\sum_{n=1}^{N_{\mathrm{s}}-1} \sum_{p=1}^{N_{\mathrm{pa}}} A_{p} s[n] g\left(t-\tau_{p}-n T_{\mathrm{s}}\right)+I(t)+w(t), \tag{7}$$

where $I(t)$ and $w(t)$ denote the interference in the baseband and the baseband ambient noise, respectively.

Denote f_{sB} as the baseband sampling rate. The baseband input-output relationship can be represented in the discrete time as

$$y[m]=\sum_{n=1}^{N_{\mathrm{s}}-1} \sum_{p=1}^{N_{\mathrm{pa}}} A_{p} s[n] g\left(\frac{m}{f_{\mathrm{sB}}}-\tau_{p}-n T_{\mathrm{s}}\right)+I[m]+w[m], \tag{8}$$

$$=\sum_{l=0}^{L-1} s[m-l] h_{l}+I[m]+w[m], \tag{9}$$

where the first term on the right side of (8) can be written as a convolution, with $\{h_{l}\}$ being the discrete channel taps of length L, and $I[m]$ and $w[m]$ are the discrete samples of the interference and the ambient noise, respectively. The ambient noise samples are assumed white Gaussian with variance σ_{w}^{2}.

B. Interference Parameterization

Consider an individual interfering waveform. Denote f_{cI}, B_{I} and T_{I} as its center frequency, bandwidth and time duration, respectively. The interfering waveform can be represented as

$$\tilde{I}_{0}(t)=\operatorname{Re}\left\{I_{0}(t) e^{j 2 \pi f_{\mathrm{cI}} t}\right\}, \qquad t \in\left[0, T_{\mathrm{I}}\right] \tag{10}$$

where $I_{0}(t)$ is the complex baseband waveform. Following the Nyquist sampling theorem [11], it can be parameterizd by $N_{\mathrm{I}}:=\left\lceil B_{\mathrm{I}} T_{\mathrm{I}}\right\rceil$ parameters, namely,

$$I_{0}(t)=\sum_{n=0}^{N_{\mathrm{I}}-1} I_{0}\left(\frac{n}{f_{\mathrm{sBI}}}\right) \operatorname{sinc}\left(\pi f_{\mathrm{sBI}}\left(t-\frac{n}{f_{\mathrm{sBI}}}\right)\right), \tag{11}$$

where $f_{\mathrm{sBI}} \approx B_{\mathrm{I}}$ is the baseband sampling rate according to the sampling theorem.

In the received signal $y(t)$, we denote N_{int} as the total number of individual interfering waveforms that are non-overlapping in time. The interfering waveforms may differ in the center frequency, the bandwidth or the time duration. For the ℓth interference, denote $f_{\mathrm{cI},\ell}$, $B_{\mathrm{I},\ell}$ and $T_{\mathrm{I},\ell}$ as its center frequency, bandwidth and time duration, respectively. It can be parameterized as in (10) and (11). Denote $\tau_{\mathrm{I},\ell}$ as its arrival time. The interference in collection can be represented as

$$\begin{aligned}\tilde{I}(t)&=\sum_{\ell=1}^{N_{\mathrm{int}}} \tilde{I}_{\ell}\left(t-\tau_{\mathrm{I}, \ell}\right) p_{T_{\mathrm{I}, \ell}}\left(t-\tau_{\mathrm{I}, \ell}\right), \\ &=\operatorname{Re}\left\{\sum_{\ell=1}^{N_{\mathrm{int}}} I_{\ell}\left(t-\tau_{\mathrm{I}, \ell}\right) p_{T_{\mathrm{I}, \ell}}\left(t-\tau_{\mathrm{I}, \ell}\right) e^{j 2 \pi f_{\mathrm{cI}, \ell}\left(t-\tau_{\mathrm{I}, \ell}\right)}\right\},\end{aligned} \tag{12}$$

where $p_{T}(t):=1$ for $t \in[0, T)$ and zero elsewhere, is introduced to indicate the location and duration of each interference. Note that since all the interference are non-overlapping, at one time $I(t)$ only has maximally one interference.

After the passband-to-baseband conversion, the interference in baseband can be derived as

$$I(t)=\sum_{\ell=1}^{N_{\mathrm{int}}} I_{\ell}\left(t-\tau_{\mathrm{I}, \ell}\right) p_{T_{\mathrm{I}, \ell}}\left(t-\tau_{\mathrm{I}, \ell}\right) e^{j 2 \pi\left[\left(f_{\mathrm{cI}, \ell}-f_{\mathrm{c}}\right) t-f_{\mathrm{cI}, \ell} \tau_{\mathrm{I}, \ell}\right]}. \tag{13}$$

Taking $f_{\rm sB}$ as the baseband sampling rate, the interference can be represented in the discrete time as

$$I[m] = \sum_{\ell=1}^{N_{\rm int}} p_{m,\ell} e^{j2\pi(f_{{\rm cI},\ell}-f_{\rm c})\frac{m}{f_{\rm sB}}} \sum_{n=0}^{N_{{\rm I},\ell}-1} u_{n,\ell}\varrho_{m,n,\ell}, \qquad (14)$$

with

$$p_{m,\ell} := p_{T_{{\rm I},\ell}}\left(\frac{m}{f_{\rm sB}} - \tau_{{\rm I},\ell}\right), \qquad u_{n,\ell} := I_\ell\left(\frac{n}{f_{{\rm sBI},\ell}}\right),$$
$$\varrho_{m,n,\ell} := {\rm sinc}\left[\pi f_{{\rm sBI},\ell}\left(\frac{m}{f_{\rm sB}} - \tau_{{\rm I},\ell} - \frac{n}{f_{{\rm sBI},\ell}}\right)\right] e^{-j2\pi f_{{\rm cI},\ell}\tau_{{\rm I},\ell}}.$$

where $p_{m,\ell} = 0$ or 1 indicates if the ℓth interference contributes to the mth sample. Despite the sum operation in (14), since the interference are non-overlapping, $I[m]$ only consists of the contribution from at most one interference.

Collect the $\{I[m]\}$ which only contain the ℓth interfering waveform (i.e., $\forall m$ with $p_{m,\ell} \neq 0$), and stack them into a vector $\mathbf{i}_\ell$ of length L_ℓ. Stack $\{u_{n,\ell}; n = 0, \cdots, N_{{\rm I},\ell}\}$ into a vector $\mathbf{u}_\ell$ of length $N_{{\rm I},\ell}$. Define a diagonal matrix $\mathbf{\Lambda}_\ell$ of size $L_\ell \times L_\ell$, with the mth diagonal element $[\mathbf{\Lambda}_\ell]_m = e^{j2\pi(f_{{\rm cI},\ell}-f_{\rm c})\frac{m}{f_{\rm sB}}}$. Define a matrix $\mathbf{\Gamma}_\ell$ of size $L_\ell \times N_{{\rm I},\ell}$, with the (m,n)th element $[\mathbf{\Gamma}_\ell]_{m,n} = \varrho_{m,n,\ell}$. Based on (14), we have

$$\mathbf{i}_\ell = \mathbf{\Lambda}_\ell \mathbf{\Gamma}_\ell \mathbf{u}_\ell. \qquad (15)$$

Stack all the $\{I[m]\}$ within one transmission frame into a vector $\mathbf{i}$ of length $N_{\rm fra}$. Stack all the $\{\mathbf{u}_\ell\}$ into a long vector $\mathbf{u}$ of length $N_{\mathbf{u}} := \sum_{\ell=1}^{N_{\rm int}} N_{{\rm I},\ell}$. We have

$$\mathbf{i} = \mathbf{\Lambda}\mathbf{\Gamma}\mathbf{u}, \qquad (16)$$

where $\mathbf{\Lambda}$ is a diagonal matrix of size $N_{\rm fra} \times N_{\rm fra}$, formed by $\{\mathbf{\Lambda}_\ell\}$ according to the positions of the interfering waveforms (i.e., $\{p_{m,\ell}\}$'s); and $\mathbf{\Gamma}$ is a block matrix of size $N_{\rm fra} \times N_{\mathbf{u}}$, formed by $\{\mathbf{\Gamma}_\ell\}$ also according to the positions of the interfering waveforms.

Based on (8), stack the discrete samples $\{y[m]\}$ and $\{w[m]\}$ into vectors of length $N_{\rm fra}$, respectively. Stack the symbols $\{s[n]\}$ into a vector of length $N_{\rm s}$. We have

$$\mathbf{y} = \mathbf{H}\mathbf{s} + \mathbf{\Lambda}\mathbf{\Gamma}\mathbf{u} + \mathbf{w}, \qquad (17)$$

where $\mathbf{H}$ is a Toeplitz channel matrix formed by $\{h_l\}$.

III. Iterative Receiver Design for Parameterized Interference Cancellation

The received signal model in (17) shows a set of parameters to be estimated: the discrete channel taps $\{h_l\}$ (or $\mathbf{H}$) , the information symbols $\mathbf{s}$, and the interference parameters (including the time-of-arrivals $\{\tau_{{\rm I},\ell}\}$, time durations $\{T_{{\rm I},\ell}\}$, center frequencies $\{f_{{\rm cI},\ell}\}$, bandwidths $\{B_{{\rm I},\ell}\}$ and parameterization vectors $\{\mathbf{u}_\ell\}$). In this work, we propose an iterative receiver for interference mitigation. The receiver diagram is depicted in Fig. 2.

The receiver processing modules are briefly described in the following.

- *Preprocessing:* Taking the received passband signal as input, the preprocessing module estimates and compensates the Doppler effect that is induced by the platform mobility or channel dynamics, and performs the passband-to-baseband conversion.
- *Interference detection:* We introduce a sequential detector, the Page test [12], [13], for interference detection. Besides detecting the presence of an interfering waveform, the Page test determines the starting time and the ending time of the interference. Additionally, the Page test can be also applied to the Fourier transform of the time domain interfering waveform to determine its bandwidth and center frequency.
  If any interference is detected, the receiver will perform interference-aware channel estimation and iterative processing for interference estimation and cancellation. Otherwise, traditional receiver processing methods will be applied.
- *Channel estimation:* The discrete channel taps are estimated based on the received training sequence.
- *Interference estimation and cancellation:* In the iterative processing, based on the estimated information symbols in the previous iteration and the estimated channel matrix, one can estimate the interference parameterization vector $\mathbf{u}$ and then reconstruct the interference. The reconstructed interference can then be subtracted from the received signal in baseband.
- *Channel equalization:* In this work, the time-domain decision feedback equalizer combined with the time reversal technique is used for channel equalization.
- *Convolutional decoding:* Channel decoding and deinterlveaing will be applied to generate the original information symbols.

In the iteration processing, the iteration stops when the number of iterations reaches a pre-determined threshold $I_{\max}$ or the information symbols are successfully decoded.

We next provide detailed descriptions of the interference detection, interference-aware channel estimation, interference estimation and cancellation, and channel equalization and decoding.

A. Page Test-based Interference Detection

The Page test is a cumulative sum (CUSUM) detector, which was originally designed for the detection of a change, such as the change from a signal-absent state to a signal-present state [12]. It can also be modified to detect the change from signal presence to signal absence [13]. In this work, the Page test is applied for detection of interfering waveforms, and for estimation of the starting time and the ending time of each interfering waveform. The detailed steps are listed in Algorithm 1.

The detector takes the baseband sample sequence $\{y[n]\}$ as input. After squaring the complex samples $r[n] := |y[n]|^2$, a bias b is introduced by the Page test to inhibit the false alarm. Define $g[n] := r[n] - b$. Before the detection of an interfering waveform, the test statistic is a cumulative sum, taking the form as

$$W[n] = \max\{0, W[n-1] + g[n]\}. \qquad (18)$$

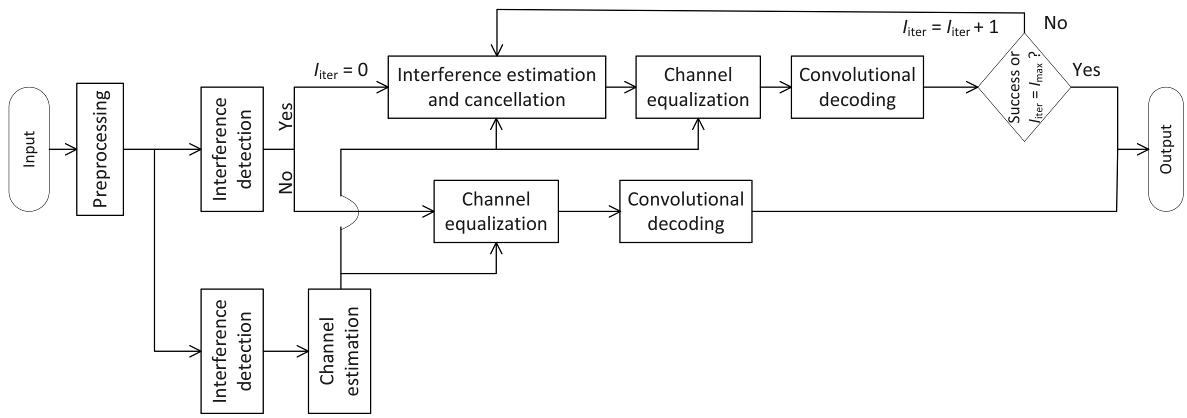

Fig. 2. Illustration of the receiver diagram for interference cancellation. $I_{\max}$: the total number of iterations in the iterative receiver processing.

When $W[n]$ exceeds a pre-determined threshold μ_0, one can declare the detection of an interfering waveform. The moment of zero crossing right before the detection can be taken as the starting time of the interference (see Algorithm 1).

Upon the detection of an intefering waveform, the test statistic will be updated as

$$W[n] = \min\{\mu_0 + \mu_1, W[n-1] + g[n]\}. \tag{19}$$

When $W[n]$ becomes less than a pre-determined threshold μ_1, one can declare the ending of the interfering waveform. The moment of crossing the threshold $(\mu_0 + \mu_1)$ right before the detection of the lagging edge can be taken the ending time of the interference (see Algorithm 1). The time duration between the starting time and the ending time is the time duration of the interfering waveform.

The same operation as above can be applied to the frequency transform of the interfering waveform to determine its frequency band.

B. Interference-aware Channel Estimation

In the real system, a training signal formed by a number of pilot symbols is typically introduced in the beginning of a frame for channel estimation. Denote N_{p} as the number of pilot symbols. Denote $\mathbf{y}_{\text{tr}}$ as a vector formed by the received baseband samples in the training period. The system model in (17) can be reformulated as

$$\mathbf{y}_{\text{tr}} = \mathbf{P}\mathbf{h} + \mathbf{\Lambda}_{\text{tr}}\mathbf{\Gamma}_{\text{tr}}\mathbf{u}_{\text{tr}} + \mathbf{w}_{\text{tr}}, \tag{20}$$

where $(\mathbf{\Lambda}_{\text{tr}}\mathbf{\Gamma}_{\text{tr}}\mathbf{u}_{\text{tr}})$ is the interference, $\mathbf{h} := [h_0, h_1, \cdots, h_{L-1}]^{\text{T}}$ is the unknown channel vector, and $\mathbf{P}$ is a matrix formed by the pilot symbols,

$$\mathbf{P} := \begin{bmatrix} p[L-1] & \cdots & p[1] & p[0] \\ p[L] & \cdots & p[2] & p[1] \\ \vdots & \ddots & \ddots & \vdots \\ p[N_{\text{p}}-1] & \cdots & p[N_{\text{p}}-L+1] & p[N_{\text{p}}-L] \end{bmatrix}. \tag{21}$$

The linear minimum mean square error (LMMSE) estimation of $\mathbf{h}$ is

$$\hat{\mathbf{h}} = \mathbf{P}^{\text{H}}\left(\mathbf{P}\mathbf{P}^{\text{H}} + \sigma_u^2\mathbf{\Lambda}_{\text{tr}}\mathbf{\Gamma}_{\text{tr}}\mathbf{\Gamma}_{\text{tr}}^{\text{H}}\mathbf{\Lambda}_{\text{tr}}^{\text{H}} + \sigma_w^2\mathbf{I}_{N_{\text{p}}-L+1}\right)^{-1}\mathbf{y}_{\text{tr}} \tag{22}$$

Algorithm 1 The Page test for interference detection

Input: Magnitude square of the complex baseband samples, $r[n] = |y[n]|^2$; and the frame length: N_{fra}
Result: Detection decision; if interference present, estimation of the starting time index N_{start} and the ending time index N_{end} of the intefering waveform
Get $g[n] = r[n] - b$; b: a bias required by the Page test
Set $W[0] = 0$, $n = 1$

while $n < N_{\text{fra}}$ **do**
  if $W[n-1] < \mu_0$ **then**
    Section 1:
    Set $W[n] = \max\{0, W[n-1] + g[n]\}$
    if $W[n] = 0$ **then**
      Set $N_{\text{start}} = n$
    if $W[n] < \mu_0$ **then**
      Set $n = n + 1$ and go to **Section 1**
    else
      The leading edge of an interfering waveform is detected;
      Estimation of the starting time index is N_{start};
      Set $W[n] = \mu_0 + \mu_1, n = n + 1$
  else
    Section 2:
    Set $W[n] = \min\{\mu_0 + \mu_1, W[n-1] + g[n]\}$
    if If $W_n = \mu_0 + \mu_1$ **then**
      Set $N_{\text{end}} = n$
    if $W[n] > \mu_0$ **then**
      Set $n = n + 1$ and go to **Section 2**
    else
      The lagging edge of an interfering waveform is detected;
      Estimate of the ending time index is N_{end};
      Set $W[n] = 0, n = n + 1$

where the elements of $\mathbf{u}_{\text{tr}}$ are assumed independently and identically distributed with variance σ_u^2. The result can be directly applied to the interference-free environment by removing the term corresponding to the interference.

C. Interference Estimation and Cancellation

Upon the detection of an interfering waveform, the receiver will perform iterative processing for interference estimation and cancellation. Denote ℓ as the interfering waveform index. Denote $\tau_{\mathrm{I},\ell}$ and $T_{\mathrm{I},\ell}$ as the time of arrival and the time duration of the ℓth interfering waveform, respectively. Denote $\mathbf{y}_\ell$ as the received signal within the time interval $[\tau_{\mathrm{I},\ell}, \tau_{\mathrm{I},\ell}+T_{\mathrm{I},\ell})$. Denote $\mathbf{s}_\ell$ as the information symbols carried by $\mathbf{y}_\ell$.

Consider the ith iteration in the iterative processing of $\mathbf{y}_\ell$. Based on the channel estimation $\mathbf{H}$ and the estimated information symbols in the $(i-1)$th iteration, the desired signal can be reconstructed as $\hat{\mathbf{H}}\hat{\mathbf{s}}_\ell^{(i-1)}$. Based on (17), define

$$\mathbf{i}_\ell := \mathbf{y}_\ell - \hat{\mathbf{H}}\hat{\mathbf{s}}_\ell^{(i-1)} = \hat{\mathbf{B}}_\ell \mathbf{u}_\ell + \bar{\mathbf{w}}_\ell, \tag{23}$$

where $\hat{\mathbf{B}}_\ell := \hat{\mathbf{\Lambda}}_\ell \hat{\mathbf{\Gamma}}_\ell$, $\bar{\mathbf{w}}_\ell := \left(\mathbf{H}\mathbf{s}_\ell - \hat{\mathbf{H}}\hat{\mathbf{s}}_\ell^{(i-1)}\right) + \mathbf{w}_\ell$. Here, $\hat{\mathbf{\Lambda}}_\ell$ and $\hat{\mathbf{\Gamma}}_\ell$ are formed based on the estimations of the interference parameters $(f_{\mathrm{cI},\ell}, B_{\mathrm{I},\ell}, \tau_{\mathrm{I},\ell})$ (c.f. (15)), and $\bar{\mathbf{w}}_\ell$ denotes an equivalent additive noise which consists of the ambient noise and the residual noise due to imperfect channel and information symbol estimation.

Following the least squares (LS) criterion, the vector $\mathbf{u}_\ell$ can be estimated as

$$\hat{\mathbf{u}}_\ell = [\hat{\mathbf{B}}_\ell^{\mathrm{H}} \hat{\mathbf{B}}_\ell]^{-1} \hat{\mathbf{B}}_\ell \mathbf{i}_\ell. \tag{24}$$

The interference can then be reconstructed and subtracted from $\mathbf{y}_\ell$. Define

$$\mathbf{z}_\ell := \mathbf{y}_\ell - \hat{\mathbf{B}}_\ell \hat{\mathbf{u}}_\ell = \mathbf{H}\mathbf{s}_\ell + \check{\mathbf{w}}_\ell, \tag{25}$$

where $\check{\mathbf{w}}_\ell := (\mathbf{B}_\ell \mathbf{u}_\ell - \hat{\mathbf{B}}_\ell \hat{\mathbf{u}}_\ell) + \mathbf{w}_\ell$ is an equivalent noise which consists of the ambient noise and the residual interference. The channel equalization and decoding can then be performed based on $\mathbf{z}_\ell$.

D. Channel Equalization and Decoding

The time reversal (TR) technique is applied before the decision feedback equalization (DFE). Specifically, the TR computes a convolution between the useful signal and the time-reversed channel, namely,

$$z_{\mathrm{TR}}[n] = \sum_{l=0}^{L-1} z[n-l] h_{L-1-l}. \tag{26}$$

In multi-channel systems, the TR sequences from multiple channels can be directly summed together to form a single data stream for subsequent processing.

The single stream is processed by DFE. For the nth symbol, denote $\mathbf{c}_{\mathrm{ff}}[n]$ of length P_1 as the feedforward coefficient vector, and denote $\mathbf{c}_{\mathrm{fb}}[n]$ of length P_2 as the feedback coefficient vector. Stack the previous symbol estimates into a vector of length P_2, $\hat{\mathbf{s}}[n-1] := [\hat{s}[n-P_2, \cdots, \hat{s}[n-1]]]^{\mathrm{T}}$. Stack the measurements into a vector of length P_1, $\mathbf{z}_{\mathrm{TR}}[n] := [z_{\mathrm{TR}}[n], \cdots, z_{\mathrm{TR}}[n+P_1-1]]^{\mathrm{T}}$. The nth symbol can be estimated as

$$\hat{s}[n] = \mathbf{c}_{\mathrm{ff}}^{\mathrm{H}}[n] \mathbf{z}_{\mathrm{TR}}[n] e^{-j\theta[n]} + \mathbf{c}_{\mathrm{fb}}^{\mathrm{H}}[n] \hat{\mathbf{s}}[n-1], \tag{27}$$

TABLE I

TRANSMISSION PARAMETERS IN THE BOHAI GULF EXPERIMENT

center frequency	f_{c}	6 kHz
bandwidth	B	2 kHz
# of information symbols	N_{s}	7500
# of pilot symbols	N_{p}	500
sampling rate	f_{s}	48 kHz
convolutional coding rate	r_{c}	1/2

where $\theta[n]$ is the output from a phase-locked loop. Specifically,

$$\theta[n+1] = \theta[n] + \lambda_1 \phi[n] + \lambda_2 \sum_{m=0}^{n} \phi[m], \tag{28}$$

with λ_1 and λ_2 being the proportional constant and the integral tracking constant, respectively, and

$$\phi[n] = \mathrm{Im}\left\{ \mathbf{c}_{\mathrm{ff}}^{\mathrm{H}}[n] \mathbf{z}_{\mathrm{TR}}[n] e^{-j\theta[n]} \mathrm{Conj}\left(\hat{s}[n] + \mathbf{c}_{\mathrm{fb}}^{\mathrm{H}}[n] \hat{\mathbf{s}}[n-1]\right) \right\}. \tag{29}$$

The DEF coefficient vectors are updated according to the recursive least squares (RLS) method [14].

Based on the estimated symbol $\hat{s}[n]$, a hard decision can be made according to the symbol constellation. The data symbols are then sent to a decoder for error correction. In this work, a convolutional code is used. In the iterative receiver, the decoded information symbols will then be used for interference estimation and cancellation.

IV. EXPERIMENTAL RESULTS

The proposed receiver is evaluated by field experimental data sets. The communication waveform was collected from a sea experiment in the Bohai Gulf, Dalian, China, in January 2015. The waveform parameters are listed in Table I. We consider two types of interferences: impulsive noise and sonar pulses, both collected from the Artic in August 2018 during the 9th Chinese National Arctic Research Expedition [15]. The impulsive noise (depicted in Fig. 1) was generated by ice cracking. The sonar pulse was a single-tone signal at 6 kHz and has a time duration of 30 ms. Two consecutive sonar pulses were separated by a second.

In the performance evaluation, the received signal with interference contamination is emulated as a weighted summation of the communication waveform and the interference. The signal-to-interference ratio (SIR) is defined as $\mathrm{SIR} := P_{\mathrm{s}}/P_{\mathrm{I}}$, where P_{s} is the average power of the desired signal, and P_{I} is the average power of the interference within the interference band. White Guassian noise is added into the received communication waveform to adjust the received signal-to-noise ratio (SNR). In the receiver processing, the bandwidth and time duration of both kinds of interferences are determined by the Page test.

Figures 3 and 4 depict the processing results corresponding to the impulsive noise. Compared to the conventional receiver without interference cancellation, the proposed receiver achieves significant performance improvement by explicit interference estimation and cancellation. Moreover, as the SIR decreases, performance of the proposed receiver gets closer to that of the conventional receiver in the interference-free environment. Fig. 4 shows the performance improvement as the

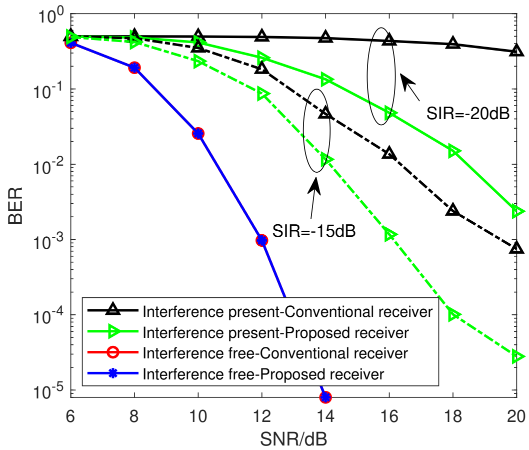

Fig. 3. Performance of several receivers in the presence of the impulsive noise.

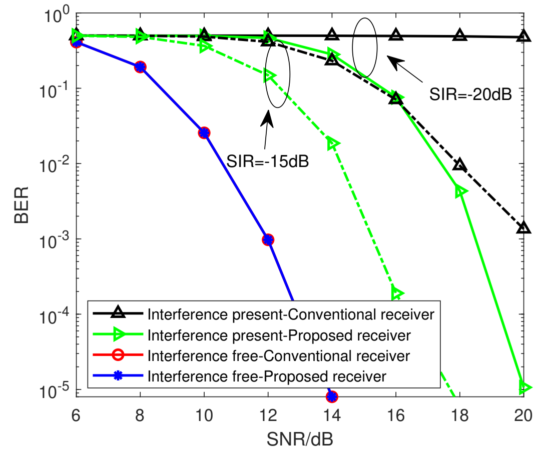

Fig. 5. Performance of several receivers in the presence of the narrowband interference.

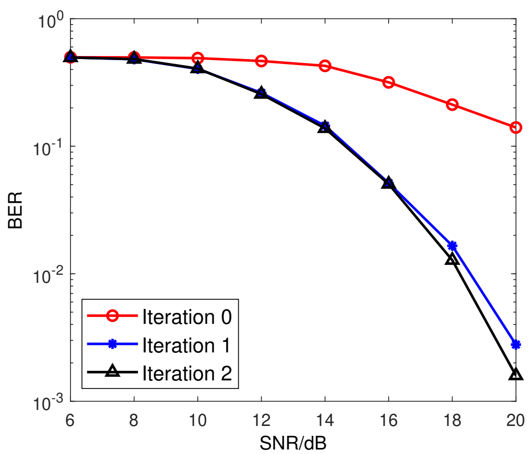

Fig. 4. Performance of the proposed receiver with different number of iterations and in the presence of the impulsive noise, $\text{SIR} = -20$ dB.

number of iterations increases. One can observe a considerable performance improvement from iteration 0 to iteration 1. The performance convergences almost within two iterations.

Figure 5 depicts the processing results corresponding to the sonar interference. One can have similar observations as those for the impulsive noise. Particularly, the performance gap between the proposed receiver and the conventional receiver appears larger than that of the impulsive noise.

V. Conclusions

This paper developed an iterative receiver for interference cancellation in the single-carrier modulated UWA communications. The interfering waveform is detected via the Page test. After parameterzing the interference by a finite number of unknowns, an iterative receiver that performs iterative interference estimation/cancellation and symbol detection was designed. The experimental results showed that the proposed receiver can effectively mitigate the impulsive noise and the narrowband interference.

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