Interference Cancellation in In-Band Full-Duplex Underwater Acoustic Systems

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***Abstract*—In-band full-duplex (IBFD) radios can double the spectral efficiency in theory, and their feasibility has been demonstrated by using advanced cancellation methods for terrestrial communications. We investigate the feasibility of acoustic IBFD in the underwater acoustic (UWA) channel, to enhance the spectral efficiency in the severely bandwidth-limited aquatic environment. We discuss the challenges of implementing IBFD UWA systems. Further, we propose an acoustic-specific cancellation scheme to deal with the strong multipath self-interference in IBFD UWA systems. We show, through simulations, that the proposed scheme can effectively suppress different types of interference.**

I. Introduction

Underwater acoustic communications has been widely discussed to support monitoring of our marine environment, exploration of the ocean resources, and minimization of man-made disasters and their impacts. However, the current subsea acoustic communications and networking technologies are still primitive, especially with respect to providing adequate data rates with acceptable reliability. The fundamental difficulty for underwater acoustic communication networks lies in the narrow bandwidth available, with a maximum of only tens of kilohertz [1]–[3]. Recently, radio-frequency (RF) testbed implementations [4]–[6] have proven that in-band full-duplex (IBFD) is feasible and can outperform half-duplex (HD) systems. This approach enables two transceivers to communicate over a bidirectional channel using the same temporal and spectral resources. Therefore, IBFD ideally renders up to double the spectral efficiency with respect to conventional HD systems. Considering that the bandwidth available for underwater acoustics is extremely limited [3], it is reasonable and desirable to investigate the employment of IBFD in underwater acoustic (UWA) systems.

As described in [4]–[6], the suppression or cancellation of the strong self-loop interference caused by its own transmission plays the most critical role in the performance of an IBFD RF transceiver. In [4], an extra transmit chain is used to generate an RF reference signal for cancellation. Implementations in [5] employ multiple antennas or fixed phase shifters to create a null point at the receive antennas. In [6], samples of the RF self-loop interference are interpolated to generate the reference signal for cancellation.

Among the limited work in full-duplex acoustics, underwater acoustic modems in [7] perform full-duplex communications through a frequency division scheme. A full-duplex underwater network, through separate frequency bands, is proposed in [8] to solve the hidden/exposed terminal problems. In [9], full-duplex communications through code division multiple access (CDMA) is used to support simultaneous data transmission and reception in the acoustic channel. Full-duplex communications via frequency division schemes or CDMA does not increase spectral efficiency; however, IBFD communications does.

In this paper, we propose to develop IBFD modems for the underwater acoustic (UWA) channel. First, we discuss three challenges in implementing IBFD UWA systems: 1) strong interference with significant delay due to reflections from the sea surface and seafloor; 2) difficulties of analog cancellation in acoustics; and 3) effectiveness of some existing cancellation methods adopted from IBFD radios for UWA systems. Then, we propose an acoustic-specific cancellation scheme for IBFD UWA systems with the analysis of different types of residual interference. Simulation results show that the proposed scheme performs well when the reflected acoustic returns do not have fast fluctuations. With our proposed scheme, the residual interference after digital cancellation can be suppressed down close to the noise level.

II. Challenges in IBFD UWA system

In this section, we explain the acoustic IBFD system for the marine environment and the associated challenges for interference cancellation. In shallow water, as shown in Fig. 1, there are two types of interference for an IBFD transceiver: 1) self-loop interference from its own transmission; and 2) interference due to reflections from the sea surface and seafloor, called self-multipath interference. In the presence of self-multipath interference, the received signal is

$$r(t) = \sum_{i=0}^{\infty} x(t-\tau_i) * h_i(t) + s(t) + z(t), \qquad (1)$$

where $x(t)$ is the transmitted signal from the self transducer, τ_i denotes the propagation delay of the i-th self-multipath

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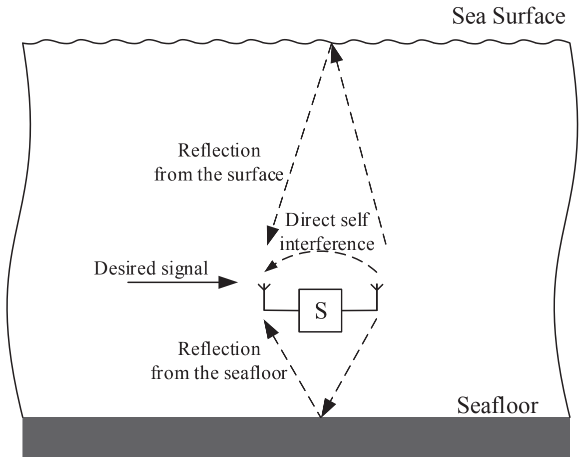

Fig. 1. Self-multipath interference of acoustic IBFD systems in shallow water.

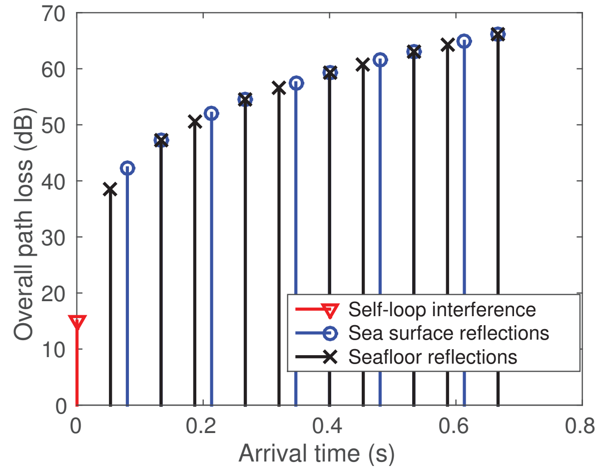

Fig. 2. Path loss of the self-multipath signals in acoustic IBFD systems.

link, $h_i(t)$ is the channel responses (including the amplifier and transducer responses, and the medium effects of the i-th interference link), $s(t)$ is the desired signal from a remote transmitter, and $z(t)$ is the ambient noise in the ocean. Therefore, the first term in (1) is the summation of all interference from the self-loop $x(t-\tau_0)*h_0(t)$ and the later multipath $x(t-\tau_i)*h_i(t)$, $i \geq 1$. Note that $s(t)$ may contain distortion from multipath effects between the transceiver and the remote transmitter.

Different from the self-loop interference in IBFD radios, the reflected interference in the underwater acoustic channel may still be strong, and the interference arrives at the transceiver after a significant delay. These characteristics make the interference suppression much more difficult. Here, we provide an example to calculate the path loss associated with reflected interference for the scenario shown in Fig. 1. We use the acoustic model from [3]:

$$A(l,f) = (l/l_r)^k a(f)^{l-l_r}, \tag{2}$$

where f is the signal frequency, l is the transmission distance, l_r is the reference distance, k models the spreading loss (2 is adopted here), and the absorption coefficient $a(f)$ can be expressed empirically (for frequencies above a few hundred Hz) using Thorp's formula [3]. We assume the speed of sound is 1500 m/s, the acoustic carrier frequency be 25 kHz, and the distances from the transceiver to sea surface and to seafloor are $d_{\text{ts}} = 60$ m and $d_{\text{tf}} = 40$ m, respectively. The calculated path loss and propagation delay of the self-multipath interference are shown in Fig. 2 [1]. Twenty reflected returns from the surface and seafloor are shown. We can see that some of these reflections with long delays (on the order of a tenth of a second) are still strong, creating significant interference to the desired signal in this example.

Even in the deep ocean where the multipath interference can be neglected, additional challenges exist. First, the self-loop interference in the underwater acoustic channel still has a delay several orders of magnitude larger than that in the RF environment, due to the slow speed of sound. In analog cancellation schemes in IBFD radios, the channel response of the self-loop link is estimated, and then an artificial signal is generated to approximate the self-loop interference by properly delaying and attenuating the original "known" transmitted signal. Then, the artificial signal is subtracted from the received signal. In IBFD radios, the self-loop delay is small enough so that it can be generated by printed circuit board lines [6]. In the acoustic channel, the self-loop delay can easily reach tens of ms, which is not realistic to be generated by circuit delay lines.

Second, the acoustic path loss offered by antenna separation is limited for short ranges (within meters). Thus, higher suppression gain is needed. In addition, the ambient noise may be non-white, which means that effective cancellation methods should provide enough suppression for the worst case in the communication band.

Third, some existing cancellation schemes from the RF channel, for example the antenna cancellation scheme in [4], cannot be adopted due to the strong frequency-dependency of acoustic propagation. Following the scheme from [4], we can cancel the interference by adding two acoustic signals destructively at the hydrophone. For example, two transducers are placed at distances d and $d+\lambda/2$ away from the hydrophone (as in [4]). Equivalently, two transducers are placed at the same distance from the hydrophone but a fixed π phase shifter exists between the two transmitted signals (as in [5]). However, these methods can only create a null at the center frequency. The cancellation performance decreases quickly with an increase in the bandwidth.

In Fig. 3, we compare the cancellation performance between an RF and acoustic system. The center frequency and bandwidth for the IBFD radio are 2 GHz and 10 MHz, respectively; and those for the acoustic system are set as 10 kHz and 1 kHz, respectively. As shown, the antenna cancellation for the radio can provide about 35 dB suppression at 1.995 GHz (or 2.005

[1]Note that acoustic propagation at short ranges, within meters, is very complex to model. Often it requires measurements in field tests. This shares some similarities with RF electromagnetic wave propagation, where the near field effects are also hard to delineate.

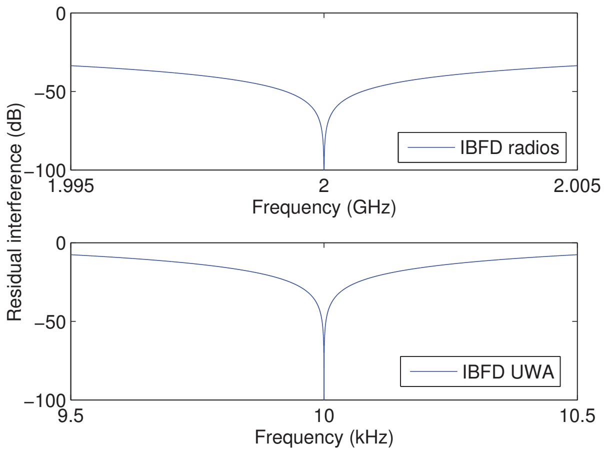

Fig. 3. Residual interference after antenna cancellation for IBFD radios and UWA systems.

GHz). But less than 10 dB suppression can be obtained at the frequency of 9.5 kHz (or 10.5 kHz) for the acoustic system. The result indicates that acoustic antenna cancellation only works with a narrow bandwidth, less than 1 kHz.

In light of these difficulties, we conclude that acoustic-specific methods are necessary to suppress the strong and long-delayed multipath interference.

III. Interference Cancellation for IBFD UWA System

In this section, we describe the proposed IBFD UWA system and analyze its interference performance in the presence of several imperfections. These imperfections include imperfect channel state information (CSI), analog-to-digital converter (ADC) quantization noise, and ambient noise.

To deal with strong self-multipath interference and the large delay, we use a hybrid solution that contains both analog and digital interference schemes, as shown in Fig. 4. The acoustic IBFD transceiver adds one auxiliary transmit chain for interference cancellation, in addition to the common transmit and receive chains in HD systems. We employ multiple buffers in the digital domain to match the large delay of the self-multipath interference. We adopt a directional transmit antenna to provide about 25 dB suppression gain [10].

At the transmitter side, we use orthogonal frequency division multiplexing (OFDM) as the modulation scheme. The information symbol sequence, $x[n]$, is modulated, amplified, and converted into acoustic energy through the use of a directional transducer. At the receiver side, the received signal is converted to an electrical signal by a hydrophone. We assume that there is no hydrophone saturation. Then, after analog and digital cancellation, the desired signal is demodulated via an OFDM demodulator.

A. Residual Self-Multipath Interference from Imperfect CSI

In the auxiliary transmit chain, we treat the self-loop interference and the later multipath interference differently. The reason is that the self-loop interference is much stronger while the self-multipath interference has much longer delays. To deal with the strong self-loop interference, the signal $x[n]$ is first delayed by a buffer and then passed through an artificial channel. The output is

$$\hat{x}_0[n] = x[n - \hat{n}_{\tau_0}] * \hat{h}_0[n], \tag{3}$$

where $*$ is the linear convolution operator, $\hat{n}_{\tau_0}$ denotes the digitized version of the estimated τ_0, and $\hat{h}_0[n]$ is the estimate of $h_0(t)$ in the digital domain. After passing through the DAC and mixer, the output $\hat{x}_0(t)$ is fine-tuned with attenuation and delay. Then it is combined destructively with the received signal. The residual self-loop interference is

$$y_0^{\mathrm{a}}(t) = x(t - \tau_0) * h_0(t) - G \cdot \hat{x}_0(t - \tau_a), \tag{4}$$

where G is the attenuator gain and τ_a is the delay generated in the analog domain. Here, we implement $\hat{h}_0[n]$ in the digital domain rather than the analog domain because this simplifies the design of the analog canceller. Only an attenuator and delay module are required in the analog cancellation.

Let $r_{\epsilon_h} = (h_0(t) - G \cdot \hat{h}_0(t))/h_0(t)$ and $r_{\epsilon_\tau} = (\tau_0 - \hat{\tau}_0 - \tau_a)/\tau_0$ denote the error ratios of the channel estimation and the mismatch of delay, respectively. We assume $r_{\epsilon_h}, r_{\epsilon_\tau} \sim \mathcal{CN}(0, \sigma_\epsilon^2)$. In a ideal case, $\sigma_\epsilon^2 = 0$, which means that we can obtain ideal CSI and perfect analog circuits and there is no residual interference. Otherwise, due to imperfect channel estimation and implementation, subsequent digital cancellation is necessary. After passing through an ADC, the digitized signal is combined destructively with its replica as digital cancellation. In particular, we view all processing modules prior to digital cancellation collectively as a “channel”. The “channel” response ($h_{\mathrm{res}}[n]$) is estimated to reconstruct the residual self-loop interference. The residual digital interference is

$$y_0^{\mathrm{d}}[n] = x[n - n_{\hat{\tau}_0}] * \hat{h}_{\mathrm{res}}[n] - y_0^{\mathrm{a}}[n], \tag{5}$$

where $\hat{h}_{\mathrm{res}}[n]$ is the estimate of $h_{\mathrm{res}}[n]$.

To cancel the reflected interference, we use different buffers to match the large delays from different reflections. Due to the path loss at longer distances, the reflected interference is weaker than the self-loop interference. In the example of Section II, the strongest reflected interference is about 25 dB below the self-loop interference. We only use digital cancellation for the reflected interference. Let $h_i[n]$ and n_{τ_i} denote the digitized version of $h_i(t)$ and τ_i, respectively. Then, the residual interference for the i-th reflection is

$$y_i^{\mathrm{d}}[n] = x[n - n_{\hat{\tau}_i}] * \hat{h}_i[n] - x[n - n_{\tau_i}][n] * h_i[n], \tag{6}$$

where $n_{\hat{\tau}_i}$ and $\hat{h}_i[n]$ are the estimates of τ_i and $h_i[n]$, respectively. We also assume that the channel estimation error ratios for reflections obey a complex Gaussian distribution. Note that, if some reflected interference is strong when the transceiver is deployed close to the ocean boundares, we still need to adopt analog cancellation schemes, as for the self-loop interference cancellation. Here, we make the assumption that the reflected interference does not vary between one training of the interference cancellation to another.

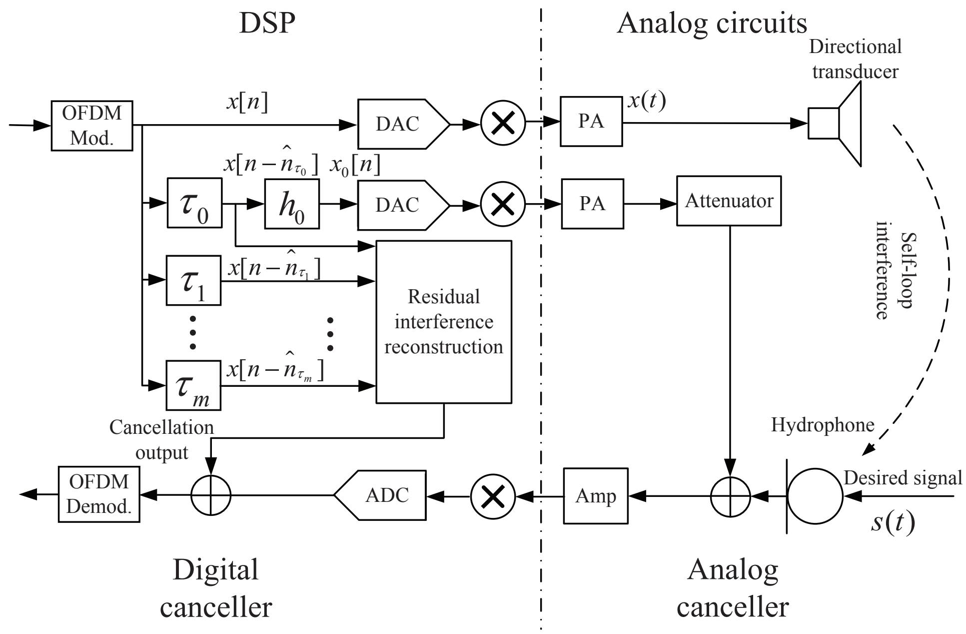

Fig. 4. Block diagram of the proposed IBFD system.

B. Effects from Ambient Noise and ADC Quantization Noise

We adopt the empirical formulations in [1] to model the ambient noise in the ocean. The model in [1] has four noise components, i.e., turbulence, shipping, waves and thermal noise. Here we generate the ambient noise as a function of the wind speed in m/s.

We assume that the ADC quantization noise q_n is uniformly distributed. For an ADC with m effective quantization bits, the quantization noise power can be calculated as [11]

$$P_q = \frac{1}{\alpha^2} \frac{1}{12 \cdot 2^{2m-2}} = \frac{\sigma_q^2}{\alpha^2}, \tag{7}$$

where σ_q^2 is the quantization noise variance, and α is the gain of the receiver's power amplifier. If the residual interference at the input of the ADC is strong, the gain value α should be set small to avoid saturation. However, smaller gain increases the quantization noise power. For a 14-bit ADC, which has 12 effective bits, the average noise power over a frequency band of $25-35$ kHz is plotted in Fig. 5. We can see that the noise power increases as the residual analog interference increases. Thus, strong analog cancellation should be implemented. Otherwise, the transceiver would experience high noise levels, which are difficult to suppress at the later stages.

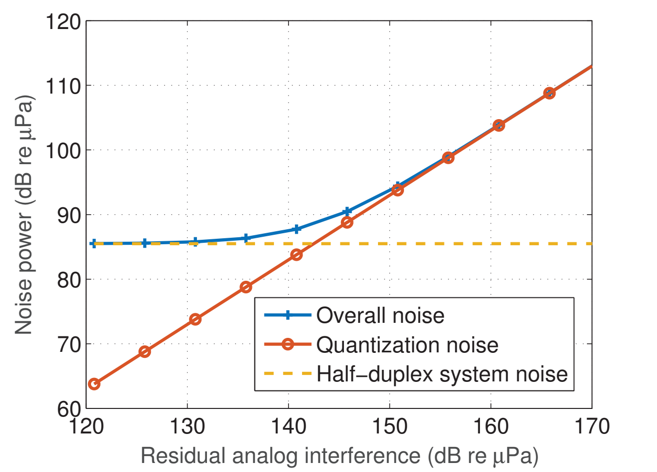

Fig. 5. Receiver noise power as a function of the residual analog interference level.

IV. Simulation Results

In this section, we show a number of computer simulations to demonstrate the effectiveness of the proposed interference cancellation scheme. In the simulations, the acoustic IBFD system has a center frequency of 25 kHz and a bandwidth of 10 kHz. The average transmit power level is 180 dB re 1 μPa. We use the ITC-3001 as the transducer, which has a directional gain of 25 dB [10]. A 14-bit ADC is used in the receiving chain. Other parameters are shown in Table I.

We first consider the case of the deep ocean, where there is only self-loop interference. The cancellation performance of the proposed scheme with different error ratios are shown in Fig. 6. Due to the usage of directional transmissions and the path loss effect, the average received power is about 40 dB lower than the average transmit power. The analog cancellation can offer about 40 and 20 dB suppression gain when the error ratio variance is set as -60 and -40 dB, respectively. Figure 7 shows the average residual analog and digital interference levels as a function of the error ratio variance, with the receiver noise floor plotted. We can see that the performance of analog cancellation highly depends both on the accuracy of the channel estimation and on the exact match of the propagation

TABLE I
SIMULATION PARAMETERS.

Center frequency	$f_c = 25$ kHz
Bandwidth	$B = 10$ kHz
Sound speed in the ocean	$v = 1500$ m/s
Separation of the transmit and receive antennas	$d_{tr} = 0.5$ m
Modulation	BPSK
OFDM subcarrier number	$N = 1024$
Cyclic prefix ratio	0.125

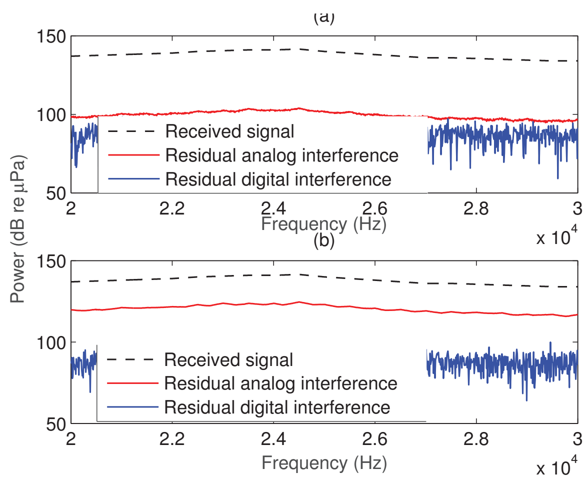

Fig. 6. Analog and digital cancellation performance with different error ratios: (a) -60 dB and (b) -40 dB.

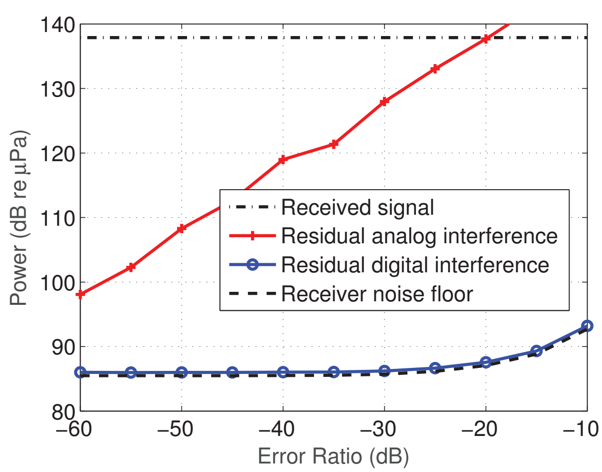

Fig. 7. Residual interference as a function of the error ratio variance.

delay. However, the interference can be suppressed to the noise level via the digital cancellation, regardless of the residual interference level after the analog cancellation. In addition, the receiver noise floor and residual digital interference begin to increase with large error ratios. This is because the analog cancellation cannot provide enough suppression in this case, which leads to large quantization noise.

In shallow water, the strength of the reflected interference

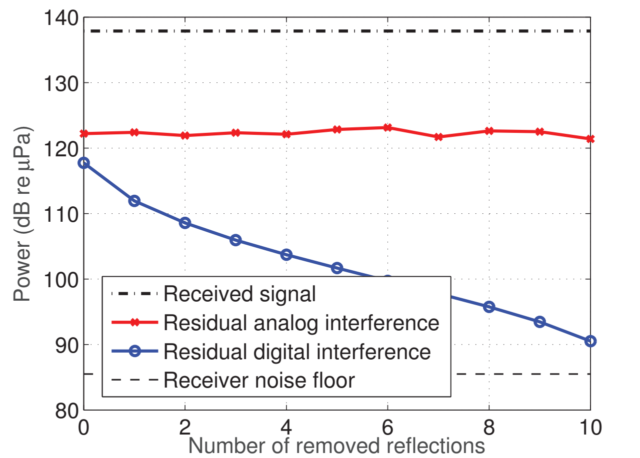

Fig. 8. Residual interference level for different numbers of removed reflections.

depends on the distance between the ocean boundaries and the receive antenna. Here, we assume $d_{st} = 60$ m and $d_{sf} = 40$ m. Figure 8 shows the residual interference level as a function of M_r,the number of removed reflections. Particularly, $M_r = 0$ means that we only reconstruct and remove the self-loop interference in the digital canceller. If $M_r = 5$, we reconstruct and remove the first 5 reflected interference. As shown in Fig. 3, the digital cancellation experiences enhanced performance as the number of the removed interference increases. The residual digital interference is about 90 dB when the first ten reflected reflections are cancelled. It is still higher than the receiver noise floor, due to the accumulation of the residual digital interference for each cancellation. In this scenario, the OFDM demodulator may experience degraded performance.

V. CONCLUSIONS AND FUTURE WORK

In this paper, we proposed to develop acoustic IBFD systems, which may double the spectral efficiency in the underwater environment. By analyzing the challenges in implementing IBFD UWA systems, we presented a hybrid design that includes both analog and digital cancellation for the underwater acoustic environment. Further, we analyzed its performance in the presence of imperfect CSI, ambient noise, and ADC quantization noise. In the deep ocean environment, we showed that the combination of directional transmission, analog cancellation, and digital cancellation can be used to suppress the self-loop interference. In shallow water, the self-multipath interference may pose challenges to the IBFD transceiver, if it is not carefully treated.

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