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ICI Cancellation Techniques Based on Data Repetition for OFDM Systems

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1.1 OFDM History

Orthogonal frequency division multiplexing (OFDM), as a special case of multicarrier transmission, is an elegant solution to the problem of high data rate transmission. OFDM was first developed to transmit data streams without inter-symbol interference (ISI) and intercarrier interference (ICI) in the 1960s [1]. A big breakthrough for efficiently implementing the OFDM system was made in the 1970s, when discrete Fourier transform (DFT) was applied to perform baseband modulation and demodulation in OFDM [2]. Another breakthrough in OFDM is the emergence of cyclic prefix (CP) in the 1980s [3], which maintains orthogonality of the transmitted signals over multipath fading channels. In the 1990s, OFDM was exploited for wideband data communications. The first application in the commercial use of OFDM was digital audio broadcasting (DAB) in the 1980s and 1990s, where OFDM guarantees highly reliable data transmission over a high-velocity and complex environment. At the beginning of the 2000s, wireless local area network (WLAN) applied the OFDM technique to the physical layers [4]. In recent years, OFDM has been widely used for fourth generation (4G) and fifth generation (5G) wireless systems to increase the utilization of spectrum resources as well as to combat frequency-selective fading.

The advantages of OFDM are summarized as follows:

- Resistance to frequency-selective fading
- Elimination of ISI and ICI
- Efficient use of the available spectrum
- Recovery of symbol lost by adequate channel coding and interleaving
- Enabling one-tap channel equalization

However, every coin has two sides. The disadvantages of OFDM lie in the following:

- Large dynamic range of transmitted signal, or peak to average power ratio (PAPR)
- Sensitivity to carrier frequency offset (CFO) and Doppler

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1.2 OFDM Principle

Consider an OFDM system of bandwidth B . The entire frequency band is divided into N subbands, each of bandwidth $B_N = B/N$. Given the channel coherence bandwidth B_c satisfying $B_N \ll B_c < B$, which usually holds for a sufficiently large value of N , the frequency-selective fading channel is converted into multiple-frequency flat fading subchannels (or say subcarriers). This can be also explained in the time domain as follows. The symbol duration of the modulated signal on each subcarrier is given by $T_N \approx 1/B_N$. When $B_N \ll B_c$, the symbol duration T_N is much larger than the channel delay spread $T_{\max} \approx 1/B_c$, which indicates that each subcarrier experiences flat fading. Therefore, as long as all subcarriers are orthogonal to each other, high data-rate interference-free parallel transmission can be achieved. How to ensure subcarrier orthogonality and perform the discrete implementation of OFDM will be discussed in detail in this section.

1.2.1 Subcarrier Orthogonality

In an OFDM system, the substream on the k -th subcarrier X_k is linearly modulated relative to the subcarrier frequency f_k , and the modulated signals associated with all the subcarriers are summed together to form the transmitted signal as

$$x(t) = \sum_{k=0}^{N-1} X_k e^{j2\pi f_k t}, \quad (1.1)$$

where $0 \leq t \leq T_N$, $f_k = f_c + k\Delta f$ with Δf representing the subcarrier interval, and f_c denoting the basic carrier frequency.

To ensure orthogonality among all subcarriers, the frequency tones $\{f_0, \dots, f_{N-1}\}$ must satisfy the following condition: given any two subcarriers k and p ,

$$\frac{1}{T_N} \int_0^{T_N} e^{j2\pi f_k t} \cdot e^{-j2\pi f_p t} dt = \begin{cases} 0, & k \neq p, \\ 1, & k = p, \end{cases} \quad (1.2)$$

where $0 \leq k, p \leq N-1$. The solution to Eq. (1.2) turns out to be $\Delta f = n/T_N$, where n is a nonzero integer. The choice of $n = 1$ is preferred as it leads to a most compact spectrum. A snapshot of the OFDM spectrum with $N = 8$ is given in Figure 1.1. Figure 1.1 shows that the spectrum of one subcarrier signal overlaps that of the others. It is expected that as N increases to a large value, the overall spectrum tends to be confined by a rectangle with a range of $N/T_N = B$.

At the receiver, without taking into account the effects of the channel and noise, that is, the received signal $y(t)$ is equal to the transmitted signal $x(t)$, the output signal on the k -th subcarrier Y_k is given by

$$\begin{aligned} Y_k &= \frac{1}{T_N} \int_0^{T_N} y(t) e^{-j2\pi f_k t} dt \\ &= \frac{1}{T_N} \sum_{p=0}^{N-1} X_p \int_0^{T_N} e^{j2\pi(p-k)\Delta f t} dt. \end{aligned} \quad (1.3)$$

Figure 1.1 Spectrum of OFDM signals with $N = 8$, depicted by the dashed line.

Because of the orthogonality in Eq. (1.2), Eq. (1.3) can be simplified as

$$Y_k = X_k. \quad (1.4)$$

From the above, it is clear that the subcarrier orthogonality ensures interference-free parallel data transmission.

Subcarrier orthogonality can be also interpreted in the frequency domain. In the frequency domain, all subcarriers should be orthogonal so as to separate out the corresponding input signal without interference. According to the Nyquist criterion, no ISI occurs when the condition that the coherence bandwidth is no less than $1/T_N$. This implies that the minimum frequency separation required for subcarriers to remain maintain orthogonality over the symbol interval $[0, T_N]$ is $1/T_N$. Obviously, the minimum frequency interval should be set to be $1/T_N$ to save the frequency resource in OFDM.

1.2.2 Discrete Implementation

According to the Nyquist criterion, to recover a signal of bandwidth B without causing any distortion, the sampling frequency must be no less than B . Consider the sampling time interval to be $\Delta t = 1/B$. The time-domain discrete transmitted signal can be obtained from Eq. (1.1) as

$$x_n \triangleq x(n\Delta t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j \frac{2\pi n k}{N}}, \quad (1.5)$$

where $n = 0, 1, \dots, N-1$. Obviously, the output of Eq. (1.5) is the inverse discrete Fourier transform (IDFT) of $\{X_0, X_1, \dots, X_{N-1}\}$.

Denote $y_n \triangleq y(n\Delta t)$ as the sampled time-domain received signal. Similarly, without taking into account the channel and noise, the discrete frequency-domain signal can be obtained from Eq. (1.3) by

$$Y_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y_n e^{-j\frac{2\pi nk}{N}} = \frac{1}{N} \sum_{k=0}^{N-1} X_k \sum_{n=0}^N e^{j\frac{2\pi}{N}pn} e^{-j\frac{2\pi}{N}kn}, \quad (1.6)$$

which is in fact the DFT of $\{y_0, y_1, \dots, y_{N-1}\}$. The subcarrier orthogonality in Eq. (1.2) also applies to the discrete-time representation:

$$\frac{1}{N} \sum_{n=0}^N e^{j\frac{2\pi}{N}pn} e^{-j\frac{2\pi}{N}kn} = \begin{cases} 1, & k = p, \\ 0, & k \neq p, \end{cases} \quad (1.7)$$

such that $Y_k = X_k$.

In summary, the OFDM modulation and demodulation can be realized by IDFT and DFT, respectively, which is very simple by virtue of the computationally efficient inverse fast Fourier transform (IFFT) and fast Fourier transform (FFT) algorithms.

1.2.3 OFDM in Multipath Channel

When taking into account the multipath channel $\mathbf{h} = [h_0, h_1, \dots, h_{L-1}]$, where L is the number of channel taps, the orthogonality of subcarriers will be unfortunately broken, as shown in Figure 1.2. A CP can be used to avoid ISI when an OFDM signal is transmitted over the multipath channel, where the CP is essentially an identical copy of the last portion of the OFDM symbol. This CP preserves the orthogonality of the subcarriers and prevents ISI between successive OFDM symbols. Specifically, the CP of length $\mu \geq L$, which is composed of the last μ samples of the OFDM symbol, namely $\{x_{N-\mu}, \dots, x_{N-1}\}$, is appended to the beginning of the OFDM symbol. After adding the CP, the new time-domain OFDM symbol yields

$$\tilde{\mathbf{x}} = [\tilde{x}_{-\mu}, \dots, \tilde{x}_{-1}, \tilde{x}_0, \dots, \tilde{x}_{N-1}]^T = [x_{N-\mu}, \dots, x_{N-1}, x_0, \dots, x_{N-1}]^T. \quad (1.8)$$

The effect of CP is illustrated in Figure 1.2, which shows that subcarrier orthogonality is maintained provided that the length of CP is no less than the channel delay spread.

Figure 1.2 The multipath and CP effects.

After passing through the channel, the received signal without noise is

$$y_n = \tilde{x}_n * h_n = \sum_{k=0}^{\mu-1} h_k \tilde{x}_{n-k} = \sum_{k=0}^{\mu-1} h_k x_{\langle n-k \rangle_N} = x_n \otimes h_n, \quad (1.9)$$

where $\langle n - k \rangle_N$ denotes $n - k$ modulo N . This means that by appending a CP to the channel input, the linear convolution associated with the channel impulse response becomes a cyclic convolution.

Let us take another look at the relation in Eq. (1.9) via matrix representation. The channel output including the CP can be expressed in matrix form:

$$\begin{bmatrix} y_{-\mu} \\ \vdots \\ y_{-1} \\ y_0 \\ \vdots \\ y_{N-2} \\ y_{N-1} \end{bmatrix} = \begin{bmatrix} h_0 & 0 & \cdots & \cdots & 0 & \cdots & 0 \\ \vdots & h_0 & 0 & \cdots & 0 & \cdots & 0 \\ h_{L-1} & \vdots & \ddots & \ddots & \vdots & \ddots & \vdots \\ 0 & h_{L-1} & \ddots & h_0 & 0 & \ddots & 0 \\ \vdots & 0 & \ddots & \vdots & h_0 & \ddots & \vdots \\ \vdots & 0 & \ddots & h_{L-1} & \vdots & h_0 & 0 \\ 0 & \cdots & 0 & 0 & h_{L-1} & \cdots & h_0 \end{bmatrix} \begin{bmatrix} x_{N-\mu} \\ \vdots \\ x_{N-1} \\ x_0 \\ \vdots \\ x_{N-2} \\ N-1 \end{bmatrix}. \quad (1.10)$$

The channel matrix of dimensions $(N + \mu) \times (N + \mu)$ in Eq. (1.10) is a Toeplitz matrix. Removing the CP leads to

$$\begin{bmatrix} y_0 \\ y_1 \\ \vdots \\ y_{L-1} \\ y_L \\ \vdots \\ y_{N-1} \end{bmatrix} = \begin{bmatrix} h_0 & 0 & \cdots & 0 & h_{L-1} & \cdots & h_1 \\ h_1 & h_0 & 0 & \cdots & 0 & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & h_{L-1} \\ h_{L-1} & \cdots & h_1 & h_0 & 0 & \cdots & 0 \\ 0 & h_{L-1} & \cdots & h_1 & h_0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \vdots & \vdots & \ddots & \vdots \\ 0 & \cdots & 0 & h_{L-1} & h_{L-2} & \cdots & h_0 \end{bmatrix} \begin{bmatrix} x_0 \\ x_1 \\ \vdots \\ x_{L-1} \\ x_L \\ \vdots \\ x_{N-1} \end{bmatrix}. \quad (1.11)$$

The channel matrix of dimensions $N \times N$ in Eq. (1.11) now becomes a circular matrix, which can be decomposed into the multiplication of an IFFT matrix, a diagonal matrix, and an FFT matrix.

Benefiting from the cyclic convolution, the subcarrier signal Y_k can be directly obtained by X_k and the corresponding frequency response H_k from

$$Y_k = \text{DFT}\{x_n \otimes h_n\} = X_k H_k, \quad (1.12)$$

where $k = 0, 1, \dots, N - 1$ and

$$H_k = \sum_{l=0}^{L-1} h_l e^{-j \frac{2\pi k l}{N}}. \quad (1.13)$$

The overall block diagram of an OFDM transceiver is shown in Figure 1.3. At the transmitter, the incoming bit stream is divided into N substreams via a serial-to-parallel (S/P) converter. Then, all bit substreams are independently mapped to the corresponding constellation points by amplitude phase modulation, such as phase shift keying (PSK) and quadrature amplitude modulation (QAM), yielding $X_0, X_1, \dots, X_{N-1}$. An N -point IFFT operation follows, converting the signal from frequency domain to time domain. After that, a CP

Figure 1.3 Block diagram of OFDM systems.

of length μ is appended to the time-domain signal, which is further passed through a low-pass filter to generate the baseband transmitted signal. Finally, the baseband transmitted signal is converted up to the passband by a carrier frequency f_c , and sent over the channel. At the receiver, the received signal is first converted down to the baseband by using a local carrier signal of carrier frequency f'_c . Then, the receiver follows the reverse baseband processing procedure at the transmitter to finally obtain the estimated bit stream.

1.3 Carrier Frequency Offset Effect

Ideally, f_c is identical to f'_c . However, due to hardware imperfection and/or Doppler, this may not hold, resulting in a CFO $\Delta f_c = f_c - f'_c$. Denote ϵ as the normalized CFO, which is given by

$$\epsilon = \frac{\Delta f_c}{\Delta f}. \quad (1.14)$$

To investigate the effect of CFO, we assume a fixed ϵ . From Eq. (1.9), the time-domain received signal can be written as

$$\begin{aligned} y_n &= \frac{1}{\sqrt{N}} \left(\sum_{l=0}^{L-1} h_l \tilde{x}_{n-l} \right) e^{j \frac{2\pi \epsilon n}{N}} + w_n \\ &= \frac{1}{\sqrt{N}} \left[\sum_{l=0}^{L-1} h_l \sum_{k=0}^{N-1} X_k e^{j \frac{2\pi k(n-l)}{N}} \right] e^{j \frac{2\pi \epsilon n}{N}} + w_n \\ &= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k \left(\sum_{l=0}^{L-1} h_l e^{-j \frac{2\pi kl}{N}} \right) e^{j \frac{2\pi (k+\epsilon)n}{N}} + w_n \\ &= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k H_k e^{j \frac{2\pi (k+\epsilon)n}{N}} + w_n, \end{aligned} \quad (1.15)$$

where w_n is the additive white Gaussian noise (AWGN) and $n = 0, 1, \dots, N-1$.

After DFT, the received signals in the frequency domain can be expressed as

$$Y_m = S_0 H_m X_m + \sum_{k=0, k \neq m}^{N-1} S_{k-m} H_k X_k + W_m, \quad m = 0, 1, \dots, N-1, \quad (1.16)$$

where W_m is the noise in the frequency domain, and

$$S_k = \frac{\sin(\pi\epsilon)}{N \sin(\pi(k + \epsilon)/N)} e^{j\left(\pi\epsilon\left(1 - \frac{1}{N}\right) - \frac{\pi k}{N}\right)} \quad (1.17)$$

is the ICI coefficient [5, 6].

1.3.1 Properties of ICI Coefficients

The ICI coefficient S_k is a periodic function with period N . The amplitude of S_k is plotted in Figure 1.4 with $\epsilon = 0.1$ and $N = 1000$. It can be observed that for large values of $|k|$, $|S_k|$ goes to 0. This means that ICI mainly comes from neighboring subcarriers. The phase of S_k is $\angle S_k = \pi \left[\epsilon \left(1 - \frac{1}{N} \right) - \frac{k\%N}{N} \right]$ for $0 < \epsilon < 1$, which is plotted in Figure 1.5 with $N = 1000$. From its expression or Figure 1.5, it is obvious that for small values of $|k|$, $\angle S_k \approx \pi\epsilon$ and $\angle S_{-k} \approx \pi\epsilon - \pi$. This implies $S_k + S_{-k} \approx 0$. In addition, for small values of ϵ , $\angle S_k \approx 0$ and $\angle S_{-k} \approx -\pi$. This means $S_k + S_{-k}^* \approx 0$. It is worth noting that $S_k + S_{-k} \approx 0$ is a more accurate approximation than $S_k + S_{-k}^* \approx 0$. As we will see in Section 1.4, the mirror-mapping-based schemes are designed on the basis of the aforementioned properties.

1.3.2 Carrier-to-Interference Power Ratio

The carrier-to-interference power ratio (CIR) is a widely used metric for evaluating the system ICI power level without considering the noise power. Suppose that the transmitted data symbols are mutually independent. According to Eq. (1.16), the instantaneous CIR of the OFDM system can be derived as

Figure 1.4 The amplitude of S_k .

Figure 1.5 The phase of S_k .

$$\text{CIR}_{\text{OFDM, Inst}} = \frac{|S_0 H_m|^2}{\sum_{k=0, k \neq m}^{N-1} |S_{k-m} H_k|^2}. \quad (1.18)$$

The average CIR can be calculated by averaging the above instantaneous CIR expression over the distribution of the channel gains. However, the calculation of the average CIR using multiple integral is overly complicated. Therefore, an approximate average CIR expression can be derived by taking the average of the numerator and the denominator of Eq. (1.18) separately. It is a good predictor of the average CIR and is simple enough to compare different schemes [7]. Thus, the approximate average CIR is utilized in this chapter. Then, the CIR of the OFDM system is

$$\text{CIR}_{\text{OFDM}} = \frac{|S_0|^2 \mathbb{E}\{|H_m|^2\}}{\sum_{k=0, k \neq m}^{N-1} |S_{k-m}|^2 \mathbb{E}\{|H_k|^2\}}. \quad (1.19)$$

Owing to the periodic property of S_k and the identities $\mathbb{E}\{|H_k|^2\} = \mathbb{E}\{|H_m|^2\} = \eta$ and $\sum_{k=0}^{N-1} |S_k|^2 = 1$, Eq. (1.19) can be simplified as

$$\text{CIR}_{\text{OFDM}} = \frac{|S_0|^2}{\sum_{k=1}^{N-1} |S_k|^2} = \frac{|S_0|^2}{1 - |S_0|^2}. \quad (1.20)$$

From Eqs. (1.17) and (1.20), an accurate approximation of the CIR can be obtained for a sufficiently large value of N as $\text{CIR}_{\text{OFDM}} \approx \text{sinc}^2(\epsilon)/(1 - \text{sinc}^2(\epsilon))$. It is clear that the CIR of the OFDM system is very sensitive to the CFO.

1.4 ICI Cancellation Techniques

ICI can seriously degrade the performance of OFDM systems. To tackle this problem, several ICI countermeasures have been proposed over the past few years. These techniques can be generally classified into three categories. In the first category, the basis expansion model (BEM) is considered for ICI mitigation [8–10]. In the second category, the ICI is treated explicitly by first estimating the ICI coefficients followed by various ICI cancellation algorithms (see e.g. [11–14]). In the third category, a low-complexity strategy, is taken to treat ICI implicitly. The adjacent-mapping-based ICI cancellation method proposed by Zhao and Haggman [6] is a representative where each data symbol is transmitted over two adjacent subcarriers. Depending on the conversion or the conjugate relation between adjacent subcarrier pairs carrying the same information, there are the adjacent symbol repetition (ASR) scheme and the adjacent conjugate symbol repetition (ACSR) scheme. These schemes can implicitly cancel ICI by combining received signals on adjacent subcarrier pairs.

This section introduces four ICI cancellation schemes belonging to the third category, called mirror symbol repetition (MSR), mirror conjugate symbol repetition (MCSR), mirror conversion transmission (MCVT), and mirror conjugate transmission (MCJT). Based on the interesting relationship between the interference from a mirror subcarrier pair, they apply data repetition within OFDM symbols according to some carefully designed subcarrier mapping rules and operations such that after combining the mirror subcarrier pairs carrying the same information, the CIR can be significantly improved. The MSR and MCSR schemes repeat the data within one OFDM symbol, while the MCVT and MCJT schemes repeat the data across two consecutive OFDM symbols. All these four ICI cancellation schemes feature the simplicity of implementation and the effectiveness of ICI mitigation.

1.4.1 One-Path Cancellation with Mirror Mapping

Figure 1.6 depicts the system architecture of the ICI one-path cancellation schemes with mirror mapping. Compared with the plain OFDM, the ICI one-path cancellation schemes have two additional modules, i.e. the ICI self-canceling modulation before the IFFT operation at the transmitter and the ICI self-canceling demodulation after the FFT operation at the receiver.

For the ICI self-canceling modulation, the input modulated data symbols are first grouped into transmit blocks. Each block consists of $(N/2 - 1)$ modulated data symbols $\{X\}_{k=1}^{N/2-1}$, which are then mapped onto N subcarriers using the one-to-two mirror-mapping rule as follows:

$$\tilde{X}_k = \begin{cases} 0, & k = 0, N/2 \\ X_k, & k = 1, 2, \dots, N/2 - 1 \\ \mathcal{O}(X_{N-k}), & k = N/2 + 1, \dots, N - 1, \end{cases} \quad (1.21)$$

where $\{\tilde{X}_k\}_{k=0}^{N-1}$ are the actual transmitted data symbols on the OFDM subcarriers. $\mathcal{O}(x)$ is defined as the mapping operation that reflects the relationship between the two modulated data symbols with the same information. Note that $\tilde{X}_{N-k} = \mathcal{O}(\tilde{X}_k)$, $k = 1, 2, \dots, N/2 - 1$ represents mirror mapping.

Figure 1.6 Block diagram of an OFDM transceiver with the ICI self-cancellation modules using mirror mapping in the baseband.

After the ICI self-cancellation demodulation, the received signals on subcarrier m , $m \in \{1, 2, \dots, N/2 - 1\}$ and its corresponding mapped subcarrier pair ($m' = N - m$) will carry the same data information. This signal redundancy renders it possible to improve the ICI mitigation performance through a coherent combining technique:

$$\hat{X}_m = \frac{H_m^* Y_m + \mathcal{O}(H_{N-m}^* Y_{N-m})}{|H_m|^2 + |H_{N-m}|^2}, \quad (1.22)$$

where H_m , H_{N-m} , Y_m , and Y_{N-m} are the channel frequency response (CFR) and the received signals at subcarrier m and its corresponding subcarrier pair ($m' = N - m$), respectively. Note that Eq. (1.22) is essentially maximum ratio combining (MRC).

1.4.1.1 MSR Scheme

For MSR scheme, the conversion operation is applied, which can be represented as $\mathcal{O}(x) = -x$. The 0-th and $N/2$ -th subcarriers are vacant in order to meet the opposite polarity condition. The transmitted data symbols in the frequency-domain (FD) after the ICI self-canceling modulation are $\tilde{X}_1 = -\tilde{X}_{N-1} = X_1$, $\tilde{X}_2 = -\tilde{X}_{N-2} = X_2$, ..., $\tilde{X}_{N/2-1} = -\tilde{X}_{N/2+1} = X_{N/2-1}$, and $\tilde{X}_0 = \tilde{X}_{N/2} = 0$. According to Eq. (1.22), the decision variable at the m -th subcarrier ($m \in \{1, \dots, N/2 - 1\}$) becomes

$$\begin{aligned} \hat{X}_m &= H_m^* Y_m - H_{N-m}^* Y_{N-m} \\ &= (S_0 |H_m|^2 + S_0 |H_{N-m}|^2 - S_{-2m} H_m^* H_{N-m} - S_{2m} H_{N-m}^* H_m) X_m \\ &\quad + \sum_{k=1, k \neq m}^{N/2-1} (S_{k-m} H_m^* H_k + S_{m-k} H_{N-m}^* H_{N-k} - S_{-k-m} H_m^* H_{N-k} - S_{m+k} H_{N-m}^* H_k) X_k \\ &\quad + H_m^* W_m - H_{N-m}^* W_{N-m}. \end{aligned}$$

The factor $1/(|H_m|^2 + |H_{N-m}|^2)$ is removed from the decision variable expression, as it does not affect the CIR value. The CIR at the m -th subcarrier can be expressed as

$$\text{CIR}_{\text{MSR}}(m) = \frac{\mathbb{E}\{|S_0(|H_m|^2 + |H_{N-m}|^2) - S_{-2m}H_m^*H_{N-m} - S_{2m}H_{N-m}^*H_m|^2\}}{\sum_{k=1, k \neq m}^{N/2-1} \mathbb{E}\{|S_{k-m}H_m^*H_k + S_{m-k}H_{N-m}^*H_{N-k} - S_{-k-m}H_m^*H_{N-k} - S_{m+k}H_{N-m}^*H_k|^2\}}.$$

For the flat fading channel, i.e. $L = 1$, the CIR becomes

$$\text{CIR}_{\text{MSR}}(m) = \frac{|2S_0 - S_{2m} - S_{-2m}|^2}{\sum_{k=1, k \neq m}^{N/2-1} |S_{k-m} + S_{-(k-m)} - S_{k+m} - S_{-(k+m)}|^2}. \quad (1.23)$$

See Ref. [15] for proof. As $S_k + S_{-k} \approx 0$, the CIR of the MSR scheme can be markedly improved. The CIR under the multipath Rayleigh fading channel and the detailed derivation are available in [15].

The average CIR of the MSR scheme is then given by

$$\text{CIR}_{\text{MSR}} = \frac{2}{N-2} \sum_{m=1}^{N/2-1} \text{CIR}_{\text{MSR}}(m). \quad (1.24)$$

1.4.1.2 MCSR Scheme

For the MCSR scheme, conjugate operation is adopted, which can be represented as $\mathcal{O}(x) = x^*$. Accordingly, we have $\tilde{X}_1 = \tilde{X}_{N-1}^* = X_1$, $\tilde{X}_2 = \tilde{X}_{N-2}^* = X_2$, ..., $\tilde{X}_{N/2-1} = \tilde{X}_{N/2+1}^* = X_{N/2-1}$, and $\tilde{X}_0 = \tilde{X}_{N/2} = 0$. From Eq. (1.22), the decision variable at the m -th subcarrier is given by

$$\begin{aligned} \hat{X}_m &= H_m^* Y_m + H_{N-m} Y_{N-m}^* \\ &= (S_0 |H_m|^2 + S_0^* |H_{N-m}|^2) X_m + \sum_{k=1, k \neq m}^{N/2-1} (S_{k-m} H_m^* H_k + S_{m-k}^* H_{N-m} H_{N-k}^*) X_k \\ &\quad + \sum_{k=N/2+1}^{N-1} (S_{k-m} H_m^* H_k + S_{m-k}^* H_{N-m} H_{N-k}^*) X_{N-k}^* + H_m^* W_m + H_{N-m} W_{N-m}^*. \end{aligned}$$

Then, the CIR of the MCSR scheme can be expressed as

$$\text{CIR}_{\text{MCSR}}(m) = \frac{\mathbb{E}\{|S_0 |H_m|^2 + S_0^* |H_{N-m}|^2|^2\}}{\sum_{k=1, k \notin \{m, N/2\}}^{N-1} \mathbb{E}\{|S_{k-m} H_m^* H_k + S_{m-k}^* H_{N-m} H_{N-k}^*|^2\}}. \quad (1.25)$$

For the flat fading channels, it can be readily verified that the CIR of the MCSR scheme is

$$\text{CIR}_{\text{MCSR}}(m) = \frac{4\Re\{S_0\}^2}{\sum_{k=1, k \notin \{N/2-m, N-m\}}^{N-1} |S_k + S_{-k}^*|^2}. \quad (1.26)$$

As $S_k + S_{-k}^* \approx 0$, the CIR of the MCSR scheme is improved compared with that of the plain OFDM. The CIR under the multipath Rayleigh fading channel is the same as that under flat

Figure 1.7 Block diagram of an OFDM transceiver with the ICI two-path cancellation modules using mirror mapping in the baseband.

fading channels, indicating that the CIR of the MCSR scheme is not affected by the channel length L . See Ref. [15] for proof.

Accordingly, the average CIR of the MCSR scheme is given by

$$\text{CIR}_{\text{MCSR}} = \frac{2}{N-2} \sum_{m=1}^{N/2-1} \text{CIR}_{\text{MCSR}}(m). \quad (1.27)$$

1.4.2 Two-Path Cancellation with Mirror Mapping

Figure 1.7 depicts the system architecture of the ICI two-path cancellation schemes with mirror mapping. The two-path cancellation schemes transmit data symbols in two consecutive OFDM symbols, which are usually referred to as two independent paths separated by time division multiplexing (TDM). Evidently, the main additional operations due to the introduction of two-path cancellation schemes are integrated inside the precoding and decoding modules.

In general, for the two-path cancellation schemes with mirror mapping, at the precoding module, one OFDM symbol input $\{X_k\}_{k=0}^{N-1}$ will become two OFDM symbol outputs $\{\bar{X}_k^{(1)}\}_{k=0}^{N-1}$ and $\{\bar{X}_k^{(2)}\}_{k=0}^{N-1}$, where the first OFDM symbol $\{\bar{X}_k^{(1)}\}_{k=0}^{N-1}$ is identical to the input OFDM symbol, i.e. $\bar{X}_k^{(1)} = X_k$, and the second OFDM symbol $\{\bar{X}_k^{(2)}\}_{k=0}^{N-1}$ obeys the subcarrier mirror-mapping rule and can be obtained as $\bar{X}_k^{(2)} = \mathcal{O}(X_{N-k})$, $k = \{0, \dots, N-1\}$. For ease of exposition, if the conversion operation is utilized for the mapping operation, i.e. $\mathcal{O}(x) = -x$, we refer to it as the MCVT scheme, and if the conjugate operation is used, i.e. $\mathcal{O}(x) = x^*$, we call it the MCJT scheme.

After the deliberate design of the transmitted signal in the precoding module, the received signals at the m -th subcarrier and the $(N-m)$ -th subcarrier of the first OFDM symbol and

the second OFDM symbol, respectively, will carry the same data information. Therefore, at the decoding module, it is reasonable to use MRC for decoding, yielding

$$\hat{X}_m = \frac{H_m^* \bar{Y}_m^{(1)} + \mathcal{O}\{H_{N-m}^* \bar{Y}_{N-m}^{(2)}\}}{|H_m|^2 + |H_{N-m}|^2}, \quad (1.28)$$

where H_m is the CFR at subcarrier m , and $\bar{Y}_m^{(i)}$ is the received signal in the FD corresponding to the i -th transmitted OFDM symbol ($i \in \{1, 2\}$).

From the above, it can be found that compared with the one-path cancellation approach, the two-path cancellation approach is conveniently compatible with the traditional OFDM transceiver design without any modifications.

1.4.2.1 MCVT Scheme

For the MCVT scheme, conversion operation is adopted. The two consecutive transmitted OFDM symbols are of the form $\bar{X}^{(1)} = [X_0, X_1, \dots, X_{N-1}]$ and $\bar{X}^{(2)} = [-X_0, -X_{N-1}, \dots, -X_1]$. From Eq. (1.28), the decision variable at the m -th subcarrier ($m = 0, \dots, N-1$) is given by

$$\begin{aligned} \hat{X}_m &= H_m^* \bar{Y}_m^{(1)} - H_{N-m}^* \bar{Y}_{N-m}^{(2)} \\ &= [S_0(\epsilon)|H_m|^2 + S_0(\epsilon + \Delta\epsilon)|H_{N-m}|^2]X_m \\ &\quad + \sum_{k=0, k \neq m}^{N-1} [S_{k-m}(\epsilon)H_m^* H_k + S_{m-k}(\epsilon + \Delta\epsilon)H_{N-m}^* H_{N-k}]X_k \\ &\quad + H_m^* W_m^{(1)} - H_{N-m}^* W_{N-m}^{(2)}, \end{aligned} \quad (1.29)$$

where the CFO of the first symbol is ϵ , the CFO of the second symbol is $\epsilon + \Delta\epsilon$, and $W_m^{(p)}$ ($p = 1, 2$) is the noise at the m -th subcarrier of the p -th path.

According to Eq. (1.29), the CIR of the MCVT scheme can be expressed as

$$\text{CIR}_{\text{MCVT}} = \frac{\mathbb{E}\{|S_0(\epsilon)|H_m|^2 + S_0(\epsilon + \Delta\epsilon)|H_{N-m}|^2\}}{\sum_{k=0, k \neq m}^{N-1} \mathbb{E}\{|S_{k-m}(\epsilon)H_m^* H_k + S_{m-k}(\epsilon + \Delta\epsilon)H_{N-m}^* H_{N-k}|^2\}}. \quad (1.30)$$

For the flat fading channel, the CIR of the MCVT scheme becomes

$$\text{CIR}_{\text{MCVT}} = \frac{|S_0(\epsilon) + S_0(\epsilon + \Delta\epsilon)|^2}{\sum_{k=1}^{N-1} |S_k(\epsilon) + S_{-k}(\epsilon + \Delta\epsilon)|^2}. \quad (1.31)$$

1.4.2.2 MCJT Scheme

For the MCJT scheme, conjugate operation is adopted. In this case, the two consecutive transmitted OFDM symbols are given by $\bar{X}^{(1)} = [X_0, X_1, \dots, X_{N-1}]$ and $\bar{X}^{(2)} = [X_0^*, X_{N-1}^*, \dots, X_1^*]$. According to Eq. (1.28), the decision variable at the m -th ($m \in \{0, 1, \dots, N-1\}$) subcarrier is given by

$$\begin{aligned}
\hat{X}_m &= H_m^* \bar{Y}_m^{(1)} + H_{N-m} \bar{Y}_{N-m}^{(2)*} \\
&= [S_0(\epsilon)|H_m|^2 + S_0^*(\epsilon + \Delta\epsilon)|H_{N-m}|^2]X_m \\
&\quad + \sum_{k=0, k \neq m}^{N-1} [S_{k-m}(\epsilon)H_m^*H_k + S_{m-k}^*(\epsilon + \Delta\epsilon)H_{N-m}H_{N-k}^*]X_k \\
&\quad + H_m^*W_m^{(1)} + H_{N-m}W_{N-m}^{(2)*}.
\end{aligned} \tag{1.32}$$

According to Eq. (1.32), the CIR of the MCVT scheme can be expressed as

$$\text{CIR}_{\text{MCJT}} = \frac{\mathbb{E}\{|S_0(\epsilon)|H_m|^2 + S_0^*(\epsilon + \Delta\epsilon)|H_{N-m}|^2\}}{\sum_{k=0, k \neq m}^{N-1} \mathbb{E}\{|S_{k-m}(\epsilon)H_m^*H_k + S_{m-k}^*(\epsilon + \Delta\epsilon)H_{N-m}H_{N-k}^*|^2\}}. \tag{1.33}$$

For the flat fading channel, the CIR of the MCJT scheme is given by

$$\text{CIR}_{\text{MCJT}} = \frac{|S_0(\epsilon) + S_0^*(\epsilon + \Delta\epsilon)|^2}{\sum_{k=1}^{N-1} |S_k(\epsilon) + S_{-k}^*(\epsilon + \Delta\epsilon)|^2}. \tag{1.34}$$

Similarly to MCSR, we see here that the denominator in the CIR expression is the summation over $S_k(\epsilon) + S_{-k}^*(\epsilon + \Delta\epsilon)$, which is approximately zero for $k \neq 0$ and small $\Delta\epsilon$. In addition, by comparing with Eq. (1.26), it can be found $\text{CIR}_{\text{MCJT}} \approx \text{CIR}_{\text{MCSR}}$ when $\Delta\epsilon = 0$.

Finally, it is worth noting that in flat fading channels, the CFRs at all subcarriers are same, and the CFR coefficients are inherently canceled in the CIR expression. Thus, the CIR of each scheme, namely Eqs. (1.20), (1.23), (1.26), (1.31), and (1.34), has exactly the same expression as that in AWGN channels.

1.4.3 CIR Comparison

An OFDM system with $N = 1024$ subcarriers is considered for comparing the CIR performance among plain OFDM, adjacent-mapping-based schemes, and mirror-mapping-based schemes.

Figure 1.8 presents the CIR results of flat fading channels, i.e. $L = 1$. For the ICI two-path cancellation schemes, CIR results with $\Delta\epsilon = 0$ and $\Delta\epsilon = 0.03$ are presented. From Figure 1.8, the following facts can be observed:

- All mirror-mapping-based schemes express much better CIR performance than plain OFDM.
- MSR and MCSR outperform ASR and ACSR, respectively. This shows that the mirror-mapping rule has better ICI suppression capability than the adjacent-mapping rule. The main reason is that for the mirror-mapping-based schemes, the interference from neighbor subcarriers is sufficiently suppressed, while there is still some residual interference from neighbor subcarriers for the adjacent-mapping-based schemes.
- With $\Delta\epsilon = 0$, the CIR curves of the MCSR scheme and the MCJT scheme coincide, and the CIR of the MSR scheme is close to that of the MCVT scheme. This is because the same operation is adopted for both schemes.
- With $\Delta\epsilon = 0$, the CIR performance of the conversion-based schemes, namely MSR and MCVT, is better than that of the conjugate-based schemes, namely MCSR and MCJT.

Figure 1.8 CIR comparison among the plain OFDM, adjacent-mapping-based schemes, and mirror-mapping-based schemes for different values of ϵ over flat fading channels.

This is because in their CIR expressions $S_k + S_{-k}^* \approx 0$ is a rougher approximation than $S_k + S_{-k} \approx 0$.

- With $\Delta\epsilon = 0.03$, the CIR performance of the MCVT scheme and the MCJT scheme is degraded due to the CFO deviation between the first OFDM symbol and the second OFDM symbol.

Figures 1.9 and 1.10 present the CIR results of multipath Rayleigh fading channels. Assume that each channel tap has the same power, i.e. $\sigma_0^2 = \sigma_1^2 = \dots = \sigma_{L-1}^2$. The effect of the channel length L is investigated. From Figures 1.9 and 1.10, it can be observed that for the multipath Rayleigh fading channels, mirror-mapping-based schemes have better CIR performance compared with the plain OFDM. The CIR performance of the plain OFDM and the conjugate-based schemes, namely MCSR and MCJT, is not affected by the channel length L . For conversion-based schemes, namely MSR and MCVT, the CIR performance is degraded with increase in channel length L . For OFDM-based communications, to maintain the spectral efficiency, N is usually much larger than the channel length L . Thus, conversion-based schemes can effectively reduce ICI for the long OFDM symbol design.

In summary, based on the CIR results of flat fading channels and multipath Rayleigh fading channels, we conclude that mirror-mapping-based schemes can effectively mitigate ICI.

1.5 Experiment on Sea

Fast time varying, phase variation, and Doppler shift constitute the features of the underwater acoustic (UWA) communication channel. As a consequence, the effect of ICI becomes significantly critical for OFDM-based UWA communication systems. The applicability of mirror-mapping-based schemes in UWA communications is verified in this section by sea experiments.

Figure 1.9 CIR comparison among the plain OFDM, the MSR scheme, and the MCSR scheme for different values of ε over multipath Rayleigh fading channels.

Figure 1.10 CIR comparison among the plain OFDM, the MCVT scheme, and the MCJT scheme for different values of ε over multipath Rayleigh fading channels.

1.5.1 Experiment Settings

The experiment was conducted in a sea area about 3 km east of Gushan, Taiwan, on 21–22 May 2013, which is illustrated in Figure 1.11. Three nodes were deployed, i.e. node 4, node 5, and node 9, each of which consisted of one transducer and four hydrophones.

Figure 1.11 Geographical locations of transceiver nodes. The GPS coordinates of nodes 4, 5, and 9 are (N22.66038, E120.21450), (N22.64844, E120.21405), and (N22.64169, E120.23450), respectively. Their relative distances are 1549.58 m (from node 4 to node 5), 2187.60 m (from node 5 to node 9), and 3377.72 m (from node 9 to node 4). Source: Map data: google, © Digital Globe, © Kingway Ltd, © GeoForce Technologies, © TerraMetrics.

The sea depth is around 20 m and the node depth is around 10 m. The transducer and the hydrophones may drift due to waves. During the sea test, the nodes transmitted with each other and the received data packets were recorded.

The basic system parameters are provided in Table 1.1. The bandwidth of the system is 5.36 kHz. The total number of subcarriers is 1600, within which there are 1278 data subcarriers, 214 pilot subcarriers, and 108 null subcarriers. The guard interval has a length of 50 ms, which is much longer than the maximum channel delay spread. The pilot subcarriers are used for channel estimation. For the ICI two-path cancellation schemes, the pilots are uniformly inserted among the data subcarriers. However, for the ICI self-cancellation schemes, to avoid loss of spectral efficiency, the mirror-mapped pilot structure has to be adopted. Thus, the ICI self-cancellation schemes do not facilitate OFDM system standardization. In addition, quadrature phase shift keying (QPSK) modulation is adopted for the mirror-mapping-based schemes.

To demonstrate the performance of the mirror-mapping-based schemes, there are two benchmark candidates, both of which have the same spectral efficiency as the mirror-mapping-based schemes. The first one is the plain OFDM with BPSK modulation (OFDM-B). The second one is the OFDM scheme with half of the total subcarriers occupied by QPSK data symbols and each of the data symbols surrounded by two null subcarriers (OFDM-QH). For OFDM-QH, the ICI from direct neighbors is removed. Figure 1.12

Table 1.1 System parameters.

Sampling rate at the transmitter	48 kHz
Sampling rate at the receiver	48 kHz
Signal bandwidth	5.36 kHz
Carrier frequency	17 kHz
Number of total subcarriers	1600
Number of data subcarriers	1278
Number of pilot subcarriers	214
Number of null subcarriers	108
Subcarrier spacing	3.35 Hz
OFDM symbol duration	299 ms
Guard interval	50 ms
Number of hydrophones	4

Figure 1.12 BER performance comparison between OFDM-QH and OFDM-B for CFOs $\epsilon = 0, 0.1, 0.2$. The total subcarrier number is $N = 1024$, the bandwidth is 6 kHz, and the channel spread length is 9 ms. For OFDM-QH, Gray code is adopted for QPSK modulation.

compares the bit error rate (BER) performance of OFDM-B and OFDM-QH under different CFOs. It is found that OFDM-B has lower BER than OFDM-QH in low-medium signal-to-noise-ratio (SNR) range ($\text{SNR} < 20$, $\text{BER} > 10^{-3}$). This is because of the higher symbol detection error for QPSK modulation. Considering the high signal attenuation of UWA channels and relatively low SNR at the receive hydrophones, OFDM-B, the one with better BER, is chosen as the benchmark in the sea experiment.

Preamble	Plain OFDM (BPSK) 5 symbols	ICI Cancellation (QPSK) 10 symbols	Postamble
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Figure 1.13 Frame structure.

The frame structure for transmission is given in Figure 1.13. The transmitted frame consists of a preamble, transmitted OFDM data symbols, and a postamble. We allocate 5 OFDM symbols with BPSK modulation for the plain OFDM and 10 OFDM symbols with QPSK modulation for either the ICI self-cancellation schemes or the two-path cancellation schemes. Therefore, 17 OFDM symbols are involved in each frame in the experiment. In addition, Gray code is adopted for QPSK modulation. Depending on the ICI cancellation structure at the last 10 OFDM symbols, there are four different frames, implementing the MSR scheme, the MCSR scheme, the MCVT scheme, and the MCJT scheme, respectively. In the experiment, ignoring preamble and CP overhead, the data rate of mirror-mapping-based schemes and OFDM-B is 4.27 kbps.

For time synchronization, the received signal is correlated with the preamble and the postamble to obtain the starting and ending time of the received data frame. Then, the resampling factor is calculated by comparing the received signal length and the transmitted signal length. Finally, to remove the major Doppler effect, the received signal is resampled according to the resampling factor [16].

The CFR of each subcarrier can be estimated via the received signals on the pilot subcarriers. The CFRs on the pilot subcarriers are estimated first. Then the CFRs on the data subcarriers can be obtained through the piecewise cubic spline interpolation. To combat the time-varying feature of UWA channels, channel estimation is done for each OFDM symbol.

1.5.2 Experiment Results

To obtain the BER results and enable a fair comparison of different schemes, the packages with indices “M0000043.DAT,” “M0000044.DAT,” “M0000046.DAT,” “M0000047.DAT,” “M0000049.DAT,” “M0000050.DAT,” “M0000052.DAT,” and “M0000053.DAT” are utilized with their information provided in Table 1.2. They are transmitted consecutively with similar receive SNR levels and through the same transmitter–receiver (T-R) pair. Figure 1.14 illustrates the BER performance of OFDM-B, the mirror-mapping-based ICI cancellation schemes, and the plain OFDM with QPSK modulation (OFDM-Q). The BER of OFDM-Q is calculated by directly decoding the mirror-mapping-based schemes without combining data subcarrier pairs, and thus ICI is not suppressed. As expected, all mirror-mapping-based schemes have lower BER than OFDM-B and OFDM-Q for all hydrophones. This confirms that mirror-mapping-based schemes can achieve superior ICI mitigation in OFDM UWA communications. Unlike with the CIR results, it is observed that all four schemes have similar BER performance. This is not surprising since the CIR performance is not directly related to the BER performance [17]. The theoretical analyses of the proposed mirror-mapping-based schemes over frequency-selective UWA channels in terms of BER will be considered as our future work.

Table 1.2 Packet information.

Packet index	Scheme	Effective SNR	Transmit time	T-R pair
M0000043.DAT	MCJT	12.47	18:16:00 05/21/13	N5-N9
M0000044.DAT	MCJT	12.55	18:16:25 05/21/13	N5-N9
M0000046.DAT	MCSR	12.27	18:16:55 05/21/13	N5-N9
M0000047.DAT	MCSR	12.15	18:17:21 05/21/13	N5-N9
M0000049.DAT	MSR	12.21	18:17:51 05/21/13	N5-N9
M0000050.DAT	MSR	12.29	18:18:16 05/21/13	N5-N9
M0000052.DAT	MCVT	12.31	18:18:46 05/21/13	N5-N9
M0000053.DAT	MCVT	12.36	18:19:11 05/21/13	N5-N9

Figure 1.14 BER performance of the plain OFDM and the mirror-mapping-based schemes.

1.6 Summary

In this chapter, four effective low-complexity mirror-mapping-based ICI cancellation schemes without explicitly estimating ICI coefficients or CFO were introduced. ICI self-cancellation schemes, namely MSR and MCSR, are robust against the CFO deviation between the first OFDM symbol and the second OFDM symbol. However, they require the mirror-mapped pilot structure, which does not facilitate OFDM system standardization. In addition, compared with conjugate-based schemes, conversion-based schemes, namely MSR and MCVT, have better CIR performance. However, their CIR performance is degraded for multipath Rayleigh fading channels with large channel length. Thus, scheme selection depends on the actual system requirement and channel conditions. Finally, all mirror-mapping-based schemes were tested in a recent sea experiment conducted in Taiwan in May 2013. Decoding results showed that the proposed

mirror-mapping-based schemes provide much lower BER than plain OFDM. This confirms that mirror-mapping-based schemes are very effective for ICI mitigation.

References

- 1 Chang, R.W. (1966). Synthesis of band-limited orthogonal signals for multichannel data transmission. *Bell System Technical Journal* 45 (10): 1775–1796.
- 2 Weinstein, S. and Ebert, P. (1971). Data transmission by frequency-division multiplexing using the discrete Fourier transform. *IEEE transactions on Communication Technology* 19 (5): 628–634.
- 3 Peled, A. and Ruiz, A. (1980). Frequency domain data transmission using reduced computational complexity algorithms. ICASSP'80. IEEE International Conference on Acoustics, Speech, and Signal Processing, Volume 5. IEEE, pp. 964–967.
- 4 Hwang, T., Yang, C., Wu, G. et al. (2008). OFDM and its wireless applications: a survey. *IEEE transactions on Vehicular Technology* 58 (4): 1673–1694.
- 5 Moose, P.H. (1994). A technique for orthogonal frequency division multiplexing frequency offset correction. *IEEE Transactions on Communications* 42 (10): 2908–2914.
- 6 Zhao, Y. and Haggman, S.G. (2001). Inter-carrier interference self-cancellation scheme for OFDM mobile communication systems. *IEEE Transactions on Communications* 49 (7): 1185–1191.
- 7 Lee, J., Lou, H., Toumpakaris, D., and Cioffi, J.M. (2006). SNR analysis of OFDM systems in the presence of carrier frequency offset for fading channels. *IEEE Transactions on Wireless Communications* 5 (12): 3360–3364.
- 8 Qu, F. and Yang, L. (2008). Basis expansion model for underwater acoustic channels? OCEANS 2008, Quebec City, QC, Canada, September 2008, pp. 1–7.
- 9 Leus, G. and van Walree, P.A. (2008). Multiband OFDM for covert acoustic communications. *IEEE Journal on Selected Areas in Communications* 26 (9): 1662–1673.
- 10 Mason, S.F., Berger, C.R., Zhou, S. et al. (2009). Receiver comparisons on an OFDM design for Doppler spread channels. OCEANS 2009-EUROPE, Bremen, Germany, May 2009, pp. 1–7.
- 11 Tu, K., Fertionani, D., Duman, T.M. et al. (2011). Mitigation of inter-carrier interference for OFDM over time-varying underwater acoustic channels. *IEEE Journal of Oceanic Engineering* 36 (2): 156–171.
- 12 Huang, X. and Wu, H. (2007). Robust and efficient inter-carrier interference mitigation for OFDM systems in time-varying fading channels. *IEEE Transactions on Vehicular Technology* 56 (5): 2517–2528.
- 13 Tang, Z., Cannizzaro, R.C., Leus, G., and Banelli, P. (2007). Pilot-assisted time-varying channel estimation for OFDM systems. *IEEE Transactions on Signal Processing* 55 (5): 2226–2238.
- 14 Huang, J., Zhou, S., Huang, J. et al. (2011). Progressive inter-carrier interference equalization for OFDM transmission over time-varying underwater acoustic channels. *IEEE Journal of Selected Topics in Signal Processing* 5 (8): 1524–1536.

- 15 Cheng, X., Wen, M., Cheng, X., and Yang, L. (2015). Effective mirror-mapping-based intercarrier interference cancellation for OFDM underwater acoustic communications. *Ad Hoc Networks (Elsevier)* 34: 5–16.
- 16 Li, B., Zhou, S., Stojanovic, M. et al. (2008). Multicarrier communication over underwater acoustic channels with nonuniform Doppler shifts. *IEEE Journal of Oceanic Engineering* 33 (2): 198–209.
- 17 Beaulieu, N.C. and Tan, P. (2007). On the use of correlative coding for OFDM. 2007 IEEE International Conference on Communications, Glasgow, UK, June 2007, pp. 756–761.