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Joint Frequency Ambiguity Resolution and Accurate Timing Estimation in OFDM Systems with Multipath Fading

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A serious disadvantage of orthogonal frequency-division multiplexing (OFDM) is its sensitivity to carrier frequency offset (CFO) and timing offset (TO). For many low-complexity algorithms, the estimation ambiguity exists when the CFO is greater than one or two subcarrier spacing, and the estimated TO is also prone to exceeding the ISI-free interval within the cyclic prefix (CP). This paper presents a method for joint CFO ambiguity resolution and accurate TO estimation in multipath fading. Maximum-likelihood (ML) principle is employed and only one pilot symbol is needed. Frequency ambiguity is resolved and accurate TO can be obtained simultaneously by using the fast Fourier transform (FFT) and one-dimensional (1D) search. Both known and unknown channel order cases are considered. Computer simulations show that the proposed algorithm outperforms some others in the multipath fading channels.

1. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) is an effective technique to deal with the multipath fading channel in high-rate wireless communications [1]. It has been chosen for the European digital audio and video broadcasting standards, as well as for the wireless local-area networking standards IEEE802.11a and HIPERLAN/2. It is also a promising candidate for the fourth-generation (4G) mobile communication standard.

Despite many advantages, OFDM systems are very sensitive to symbol timing offset (TO) and carrier frequency offset (CFO) [2, 3]. A lot of schemes for CFO and TO estimation for OFDM systems have been proposed in the literature [4–12]. However, most low-complexity estimation approaches can only estimate the CFO within one or two subcarrier spacing [4–6]. When the CFO is larger than one subcarrier spacing, the frequency ambiguity would appear. The frequency ambiguity is called integer frequency offset (IFO) because it is the integer multiple of one subcarrier spacing. The part of CFO within one subcarrier spacing is called fractional frequency offset (FFO). Schmidl and Cox [7] presented an efficient algorithm (called SCA for simplicity) for estimating the FFO, IFO, and TO. For the IFO estimation, however, their algorithm requires the observation of two consecutive symbols and supposes that the symbol timing is perfect. Moreover, the broad timing metric plateau inherent in [7] results in a large TO estimation variance. Morelli et al. [8] and Chen and Li [9] enhanced the performance of SCA [7] for the IFO estimation by employing maximum-likelihood (ML) technique (note that if there is no virtual subcarrier, Morelli's method is equivalent to Chen's method). However, their methods require perfect timing still. Park et al. [10] proposed an IFO estimator robust to the timing error, but its performance is unsatisfactory (see Figure 3).

In this paper, an efficient method for joint estimation of the IFO and TO in multipath fading channels is derived. Maximum-likelihood principle is employed and only one pilot symbol is needed. Both of them can be obtained by using the fast Fourier transform (FFT) and one-dimensional (1D) search. The estimation in the cases of known channel order (KCO) and unknown channel order (UCO) are also discussed. Our method for IFO estimation outperforms the methods in [7–10], even if those methods use two pilot symbols. The performance of the proposed method for TO estimation is also better than that of the conventional methods [7, 11] in multipath fading channel. In effect, our approach can be viewed as an extension of the Morelli and Mengali algorithm [13].

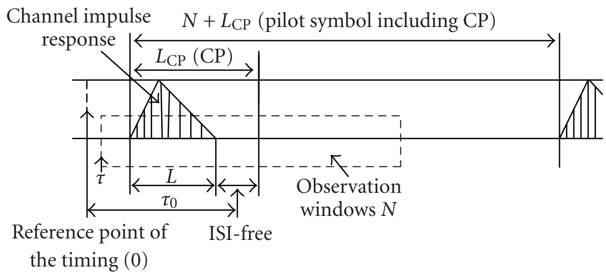

Figure 1: Accurate timing position under multipath fading.

The organization of this paper is as follows. The signal model of OFDM is introduced in Section 2. In Section 3, the algorithm for joint timing and IFO estimation using FFT is developed and the estimation in the cases of UCO and KCO are discussed. Computer simulations are presented in Section 4 to demonstrate the performance of the proposed algorithm with comparisons to the available methods [7, 9–11]. Section 5 concludes the paper.

Notation

Capital (small) bold face letters denote matrices (column vectors). Frequency domain components are indicated by a tilde. $(\cdot)^*$, $(\cdot)^T$, and $(\cdot)^H$ represent conjugate, transpose, and conjugate transpose, respectively. $\|\cdot\|$ denotes the Frobenius norm, and $\mathbf{I}_{N\times N}$ denotes the $N \times N$ identity matrix. $\mathrm{Re}(\cdot)$ denotes the real part of a complex number $(\cdot)$. $\mathrm{diag}(\cdot)$ denotes a diagonal matrix constructed by a vector. $*$ denotes the convolution and $\mathrm{fft}(\cdot)$ denotes the FFT of the columns of a matrix.

2. PROBLEM FORMULATION

The OFDM signal is generated by taking the N-point inverse fast Fourier transform (IFFT) of a block of symbols with a linear modulation such as PSK and QAM. The OFDM samples at the output of IFFT are given by

$$x(i) = \frac{\sum_{n=0}^{N-1} \tilde{a}_n[\exp(j2\pi ni/N)]}{\sqrt{N}}, \quad 0 \le i \le N-1, \tag{1}$$

where $\tilde{a}_n$ is modulated data sequence with unit energy. The useful part of each block has the duration of T seconds and is preceded by a cyclic prefix (CP) with the size of L_{CP}, longer than the channel impulse response, so as to eliminate the interference between adjacent blocks. Each OFDM block is serialized for the transmission through the possible unknown time-invariant composition multipath channel. The channel can be denoted by a discrete-time filter $h(l)$ with order L ($L \le L_{\mathrm{CP}}$):

$$h(l) = g_{\mathrm{tr}}(t) * h_p(t) * g_{\mathrm{rec}}(t)|_{t=lT_s-t_0}, \tag{2}$$

where $g_{\mathrm{tr}}(t)$ and $g_{\mathrm{rec}}(t)$ are, respectively, the response of transmitting and receiving filters. $h_p(t)$ is the impulse response of the dispersive channel. $T_s = T/N$ is sampling period, and t_0 is propagation delay. In the presence of a frequency offset f, the samples at the receiving filter output are

$$r(k) = \exp\left[\frac{j2\pi k(\nu_I + \nu_F)}{N}\right]\sum_{l=0}^{L-1} h(l)x(k-l) + w(k), \tag{3}$$

where ν_I and ν_F are, respectively, the IFO and the FFO normalized by the subcarrier space $1/T$, $x(m(N + L_{\mathrm{CP}}) + n)$ is the serialized version of the mth OFDM block with the nth entry, and $w(k)$ denotes zero-mean additive white Gaussian noise (AWGN).

Assuming that a length-N observation window slides through the received data stream (Figure 1), we can obtain observation vectors represented by the following matrix form:

$$\mathbf{r}(\tau) = \mathbf{C}(\nu_F)\mathbf{C}(\nu_I)\mathbf{X}(\tau)\mathbf{h}\xi + \mathbf{w}(\tau), \tag{4}$$

where τ is the start point of observation window, $\xi = \exp[j2\pi\tau(\nu_F + \nu_I)/N]$,

$$\begin{gathered}\mathbf{r}(\tau) = [r(\tau), r(\tau+1), \ldots, r(\tau+N-1)]^T,\\ \mathbf{C}(\nu) = \mathrm{diag}\left(1, \exp\left(\frac{j2\pi\nu}{N}\right), \ldots, \exp\left(\frac{j2\pi\nu(N-1)}{N}\right)\right),\\ [\mathbf{X}(\tau)]_{i,j} = x(i-j), \quad \tau \le i \le N+\tau-1,\ 0 \le j \le L-1,\\ \mathbf{h} = [h(0), h(1), \ldots, h(L-1)]^T,\end{gathered} \tag{5}$$

and $\mathbf{w}(\tau) = [w(\tau), \ldots, w(\tau+N-1)]^T$ is a zero-mean Gaussian vector with covariance matrix

$$\mathbf{C}_{\mathbf{w}} = E\{\mathbf{w}\mathbf{w}^H\} = \sigma^2\mathbf{I}_{N\times N}. \tag{6}$$

As illustrated in Figure 1, as long as the timing estimate is within the ISI-free guard interval, the timing offset, regardless of its values, will not degrade the system performance.

Assume the FFO is corrected in advance, then the term $\mathbf{C}(\nu_F)$ in (4) can be removed. We construct the matrix $\mathbf{X}$ by pilot symbol $[x_{N-L+1}, \ldots, x_N, x_0, \ldots, x_{N-1}]$ and replace the matrix $\mathbf{X}(\tau)$ in (4) by the matrix $\mathbf{X}$. The term ξ in (4) can be incorporated into the channel parameters $\mathbf{h}$. Then the observed data can be expressed as

$$\mathbf{r}(\tau) = \mathbf{C}(\nu_I)\mathbf{X}\mathbf{h} + \mathbf{w}(\tau). \tag{7}$$

Now, we can find from the first term in the right-hand side of (7) that there are three kinds of unknown parameters in (7), namely TO τ, IFO ν_I, and channel parameters. Assume τ_0 is the offset from a given reference to the ISI-free interval. Our task is to find τ_0 and estimate the IFO ν_I simultaneously based on the observation $\mathbf{r}(\tau)$ for given $\mathbf{X}$.

3. MAXIMUM-LIKELIHOOD ESTIMATION USING FAST FOURIER TRANSFORM

In this section, the ML principle is applied to derive an algorithm for jointly estimating the timing and IFO. The joint estimation problem in the case of unknown channel order is also discussed.

3.1. Derivation of the algorithm

Since all the parameters except for noise in (7) are deterministic, the log-likelihood function of received data can be represented as

$$\ln(L) = \text{const} - 2N\ln(\sigma^2) - \frac{\|\mathbf{r}(\tau) - C(\nu_I)\mathbf{X}\mathbf{h}\|^2}{\sigma^2}. \tag{8}$$

The estimation of τ, ν_I, and $\mathbf{h}$ is the solution of the following joint optimization problem:

$$[\mathbf{h}, \tau, \nu_I] = \min_{\hat{h},\hat{\tau},\hat{\nu}_I} \|\mathbf{r}(\tau) - \mathbf{C}(\nu_I)\mathbf{X}\mathbf{h}\|^2. \tag{9}$$

For given τ and ν_I, the minimum for (9) is

$$\hat{\mathbf{h}} = (\mathbf{X}^H\mathbf{X})^{-1}\mathbf{X}^H\mathbf{C}^H(\nu_I)\mathbf{r}(\tau). \tag{10}$$

Substituting (10) into (9), τ and ν_I can be obtained by maximizing the following cost function:

$$J(\nu_I, \tau) = [\mathbf{C}^H(\nu_I)\mathbf{r}(\tau)]^H\mathbf{P}[\mathbf{C}^H(\nu_I)\mathbf{r}(\tau)] \tag{11}$$

$$= -b(0,\tau) + 2\,\text{Re}\left[\sum_{m=0}^{N-1} b(m,\tau)\exp\left(-\frac{j2\pi m\nu_I}{N}\right)\right], \tag{12}$$

$$b(m,\tau) = \sum_{k=m}^{N-1} [\mathbf{P}]_{k-m,k}\, r^*(k-m+\tau)\, r(k+\tau), \tag{13}$$

where $\mathbf{P} = \mathbf{X}(\mathbf{X}^H\mathbf{X})^{-1}\mathbf{X}^H$ and $[\mathbf{P}]_{i,j}$ is the (i,j)th entry of $\mathbf{P}$.

The main steps in obtaining (12) are outlined in the appendix.

As ν_I and τ are integers, the estimation range of the normalized IFO ν_I is in $[0, N-1]$ and the search range of timing τ is in $[0, L_\tau - 1]$ (assume τ_0 is in $[0, L_\tau - 1]$), where 0 is the reference point of TO and L_τ is the length of TO search.

Construct two $N \times L_\tau$ matrices $\mathbf{B}$ and $\mathbf{J}$ whose entries are denoted by $b(m,\tau)$ and $J(\nu_I, \tau)$, respectively. The cost function (12) can be expressed in the following matrix form:

$$\mathbf{J} = 2\,\text{Re}\,[\,\text{fft}(\mathbf{B})] - \mathbf{B}_0, \tag{14}$$

where $\mathbf{B}_0$ is an $N \times N$ matrix with the same columns from the first column of $\mathbf{B}$.

The maximum entry of the matrix $\mathbf{J}$ can be obtained by 1D search. It is clear that the indexes of the row and column corresponding to the maximum entry of $\mathbf{J}$ represent the IFO ν_I and the TO τ_0, respectively.

3.2. Unknown channel order case

In fact, there is still a hidden parameter unknown in the data model (7). In order to construct the matrix $\mathbf{X}$, the channel order L should be known in advance. Thus the additional algorithm for the channel order estimation is needed. Furthermore, since the channel order is varying in practice, the matrices $\mathbf{X}$ and $\mathbf{P}$ have to be reconstructed according to different L. However, we find that the estimator is robust to the overestimated channel order. Hence the channel order L can be simply replaced by L_{CP} under the condition of $L_{\text{CP}} \geq L$ which is generally satisfied in OFDM systems. Therefore, we do not need to estimate L and to reconstruct $\mathbf{X}$ and $\mathbf{P}$. Comparisons of the KCO with the UCO will be given in detail next.

3.3. Effects of unknown channel order

Assume the IFO $\nu_I = 13$ and the search range of TO is from 0–18. The cost function $J(\nu_I, \tau)$ in the cases of the KCO and UCO are plotted in Figure 2. It can be seen that the cost function has a narrow timing metric plateau when $\nu_I = 13$ in the case of KCO, whereas it gives a wide timing metric plateau within the ISI-free guard interval in the case of UCO. It should be noted that the wide plateau is likely to be beyond the ISI-free interval to degrade the performance (see Simulation 2 in Section 4). For both the KCO and UCO, the cost functions have the unique tall peak at the IFO metric. However, the IFO metric of the UCO case has higher sidelobes relative to the mainlobe than that of the KCO case. It implies that there is still loss in terms of the performance of the IFO estimation when channel order is unknown (see Simulation 1 in Section 4).

Remarks

(1) Matrix $\mathbf{P}$ can be calculated in advance, which reduces largely the burden of online computations.

(2) The multipath fading channel parameters can be obtained by (10) after both the IFO and TO, are corrected. The phase offset of estimated channel parameters can be compensated by itself in the process of channel equalization.

(3) Only one pilot symbol is needed in the algorithm to estimate the IFO, TO, and channel parameters, and the pilot symbol can be selected as a random sequence.

(4) The proposed algorithm can also be extended to MIMO-OFDM systems directly, if there are a set of pilot symbols, each corresponding to a transmitting antenna.

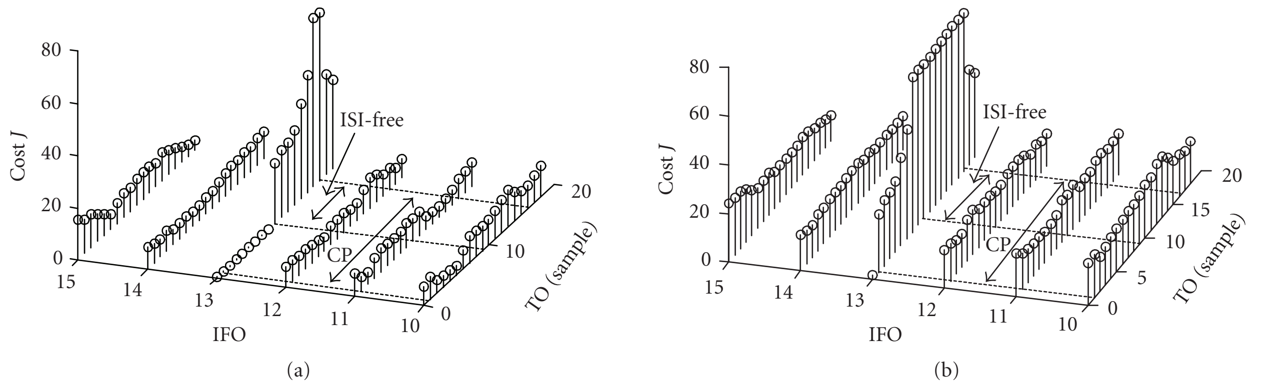

Figure 2: Cost function for joint IFO and TO estimations ($N = 64$, $L_{CP} = 16$, $L = 8$, SNR $= 20$ dB, $\nu_I = 13$): (a) the case of KCO and (b) the case of UCO.

4. SIMULATION RESULTS AND DISCUSSIONS

The performance of the proposed approach to joint estimation of the IFO and TO is evaluated by computer simulations. Consider an OFDM system with 64 subcarriers and the length of cyclic prefix with 16 samples. The QPSK symbol modulation is employed. The additive channel noise is zero-mean white Gaussian. The delay-power-spectrum function is exponential. The channel order L is varying between 8 and 16. The TX/RX filters in the simulations are raised-cosine rolloff filters with a rolloff factor 0.5. The performance of the estimated IFO is evaluated by means of the probability of failure (POF), $\Pr\{\hat{\nu}_I \neq \nu_I\}$. The performance of the estimated TO is evaluated by mean square error (MSE) and the timing error is counted with reference to the bound of the ISI-free guard interval.

Simulation 1 (performance of integer frequency offset estimation). In Figure 3, the POF of the proposed method for the IFO estimation using one pilot symbol is compared with that of the SCA [7] and Chen's method [9]. Firstly, we use Minn's method [11] to obtain the timing. And then, SCA and Chen's method are used to estimate the IFO. Note that the SCA and Chen's method are based on two pilot symbols. Park's method using one pilot symbol [10] with 32 virtual subcarriers is also plotted in Figure 3. The timing error is assumed within $\tau_0 \pm 3$ for the estimator in [10]. The simulations were performed with 100 000 runs. As shown in Figure 3, our method has smaller POF than other methods even in the case of UCO. Similar to the previous simulation, the estimated performance in the KCO case is better than that in the UCO case.

Simulation 2 (performance of timing offset estimation). Figure 4 shows the MSE of the proposed and conventional methods for the TO estimation. We can observe that our method outperforms both the SCA [7] and Minn's method [11] in both the KCO and UCO cases. It is also noted that in the KCO case, the proposed method has a much smaller MSE than in the UCO case. The reason is that the timing metric plateau of the cost function in the UCO case is beyond the ISI-free interval.

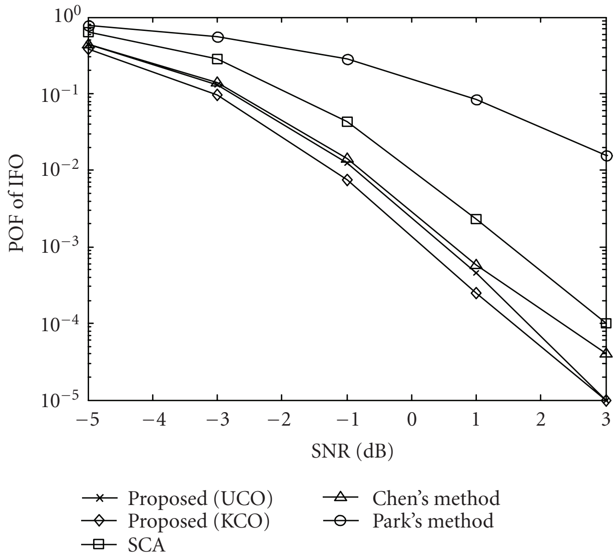

Figure 3: IFO performance comparison for the proposed method, SCA, Chen's method, and Park's method ($N = 64$, $L_{CP} = 16$, $\nu_I = 13$). Note that only the pilot symbol of Park's method has virtual subcarriers.

Simulation 3 (word error rate (WER) performance). Suppose a CFO including both FFO and IFO has an arbitrary subcarrier spacing inside [0, 64]. Figure 5 compares the WER performance of the system (by the use of SCA [7] to joint FFO and coarse TO estimation along with the proposed method) with that of the system with ideal timing and frequency synchronization. The channel parameters can be obtained by (10) and the phase offset is compensated by itself in the process of channel equalization. 128 000 words were used to obtain the results. It can be seen that for high SNRs, the proposed method, after the SCA [7], has essentially the same WER performance as the ideal system even in the case of UCO. The result indicates that although the replacement of L by L_{CP} impacts the performance of the TO and IFO estimates considerably, the impact of the replacement on the system WER is negligible in high SNR.

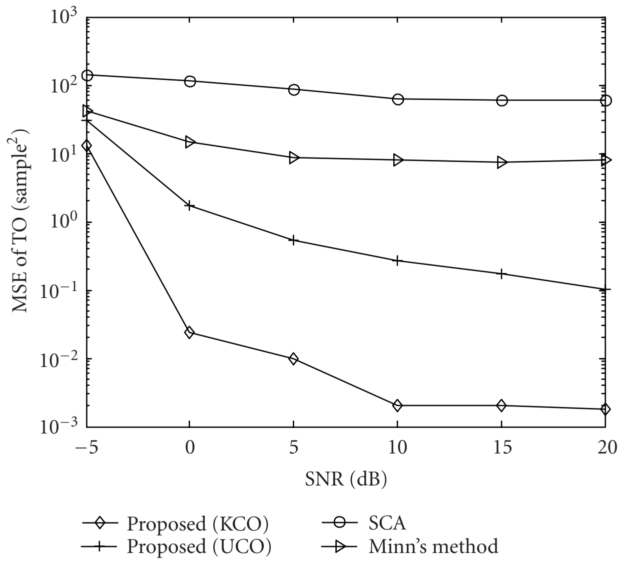

Figure 4: TO performance comparison for the proposed method, SCA, and Minn's method ($N = 64$, $L_{\text{CP}} = 16$, $\nu_I = 13$).

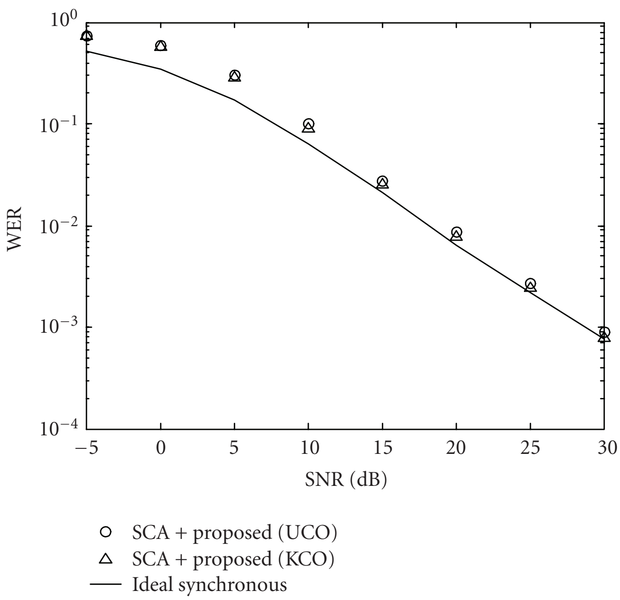

Figure 5: WER performance comparison for the system using proposed method along with SCA and the ideal synchronized system. SCA is used to estimate the FFO and coarse TO.

5. CONCLUSIONS

A method for joint frequency ambiguity resolution (or IFO estimation) and TO estimation using one pilot symbol for OFDM system is proposed. The FFT and the 1D search are employed to obtain the accurate estimation of the TO and IFO. Especially, when channel order is known, the performance of both the IFO and TO can be improved considerably. The replacement of channel order by the length of CP leads to the negligible loss in terms of the WER of systems.

APPENDIX

This appendix outlines the main steps in obtaining (12):

$$\begin{aligned}
J(\nu_I,\tau) &= \left[\mathbf{C}^H(\nu_I)\mathbf{r}(\tau)\right]^H \mathbf{P}\left[\mathbf{C}^H(\nu_I)\mathbf{r}(\tau)\right] \\
&= \sum_{i=0}^{N-1}\sum_{k=0}^{N-1}[\mathbf{P}]_{i,k} r^*(\tau+i) r(\tau+k) \\
&\quad \times \exp\left\{-\frac{j2\pi\nu_I(k-i)}{N}\right\} \\
&\overset{m=k-i}{=} \sum_{m=-N+1}^{N-1}\sum_{k=m}^{N-1+m}[\mathbf{P}]_{k-m,k} r^*(\tau+k-m) r(\tau+k) \\
&\quad \times \exp\left(-\frac{j2\pi\nu_I m}{N}\right) \\
&= -\sum_{k=0}^{N}[P]_{k,k} r^*(k+\tau) r(k+\tau) \\
&\quad + \sum_{m=0}^{N-1}\sum_{k=m}^{N-1+m}[\mathbf{P}]_{k-m,k} r^*(k-m+\tau) r(k+\tau) \\
&\quad \times \exp\left(-\frac{j2\pi\nu_I m}{N}\right) \\
&\quad + \sum_{m=-N+1}^{0}\sum_{k=m}^{N-1+m}[\mathbf{P}]_{k-m,k} r^*(k-m+\tau) r(k+\tau) \\
&\quad \times \exp\left(-\frac{j2\pi\nu_I m}{N}\right).
\end{aligned} \tag{A.1}$$

The third term in the right-hand side of (A.1) can be transformed as follows:

$$\begin{aligned}
&\sum_{m=-N+1}^{0}\sum_{k=m}^{N-1+m}[\mathbf{P}]_{k-m,k} r^*(k-m+\tau) r(\tau+k) \\
&\quad \times \exp\left(-\frac{j2\pi\nu_I m}{N}\right) \\
&\overset{m'=-m}{=} \sum_{m'=0}^{N-1}\sum_{k=-m'}^{N-1-m'}[\mathbf{P}]_{k+m',k} r^*(k+m'+\tau) r(k+\tau) \\
&\quad \times \exp\left(\frac{j2\pi\nu_I m'}{N}\right) \\
&\overset{k'=k+m'}{=} \sum_{m'=0}^{N-1}\sum_{k'=0}^{N-1}[\mathbf{P}]_{k',k'-m'} r^*(k'+\tau) r(k'-m'+\tau) \\
&\quad \times \exp\left(\frac{j2\pi\nu_I m'}{N}\right) \\
&= \sum_{m=0}^{N-1}\sum_{k=0}^{N-1}[\mathbf{P}]_{k,k-m} r^*(k+\tau) r(k-m+\tau) \\
&\quad \times \exp\left(\frac{j2\pi\nu_I m}{N}\right).
\end{aligned} \tag{A.2}$$

Note

(1) Because $\mathbf{P}$ is an $N \times N$ matrix, the range of k in (A.1) and (A.2) is from m to $N-1$.

(2) Because $\mathbf{P}$ is a projection matrix, $[\mathbf{P}]_{k-m,k} = ([\mathbf{P}]_{k,k-m})^*$.

Substituting (A.2) into (A.1) results in

$$\begin{aligned} J(v_I,\tau) &= [\mathbf{C}^H(v_I)\mathbf{r}(\tau)]^H\mathbf{P}[\mathbf{C}^H(v_I)\mathbf{r}(\tau)] \\ &= -\sum_{k=0}^{N}[P]_{k,k}r^*(k+\tau)r(k+\tau) \\ &\quad + 2\,\mathrm{Re}\left\{\sum_{m=0}^{N-1}\sum_{k=m}^{N-1}[\mathbf{P}]_{k-m,k}r^*(k-m+\tau)\right. \\ &\quad \left. \times r(k+\tau)\exp\left(-\frac{j2\pi v_I m}{N}\right)\right\} \end{aligned} \tag{A.3}$$

$$= -b(0,\tau) + 2\,\mathrm{Re}\left[\sum_{m=0}^{N-1}b(m,\tau)\exp\left(-\frac{j2\pi m v_I}{N}\right)\right] \tag{A.4}$$

$$b(m,\tau) = \sum_{k=m}^{N-1}[\mathbf{P}]_{k-m,k}r^*(k-m+\tau)r(k+\tau). \tag{A.5}$$

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