

# OFDM-ISAC Beyond CP Limit: Performance Analysis and Mitigation Algorithms

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**Abstract**—Orthogonal frequency division multiplexing (OFDM) is well-suited for integrated sensing and communications (ISAC), yet its cyclic prefix (CP) is dimensioned for communications-grade multipath and is generally insufficient for sensing. When echoes exceed the CP duration, inter-symbol and inter-carrier interference (ISI/ICI) break subcarrier orthogonality and degrade sensing. This paper presents a unified analytical and algorithmic framework for OFDM-ISAC beyond the CP limit. We first develop a general echo model that explicitly captures the structured coupling of ISI and ICI caused by CP insufficiency. Building on this model, we derive closed-form expressions for the sensing signal-to-interference-plus-noise ratio (SINR) and the range-Doppler peak sidelobe level ratio (PSLR), both of which are shown to deteriorate approximately linearly with the normalized excess delay beyond the CP. To mitigate these effects, we propose two standard-compatible successive interference cancellation (SIC) methods: SIC-DFT, a low-complexity DFT-based scheme, and SIC-ESPRIT, a super-resolution subspace approach. Simulations corroborate the analysis and demonstrate consistent gains over representative benchmarks. Both algorithms provide more than 4dB SINR improvement under CP-insufficient conditions, while SIC-ESPRIT reduces range/velocity root-mean-square-errors (RMSE) by about one order of magnitude, approaching the performance achievable with a sufficiently long CP. These results offer both theoretical insight and practical solutions for reliable long-range OFDM-ISAC sensing beyond the CP limit.

**Index Terms**—Integrated sensing and communication (ISAC), orthogonal frequency division multiplexing (OFDM), cyclic prefix (CP), inter-symbol interference (ISI), and inter-carrier interference (ICI).

## I. INTRODUCTION

Integrated sensing and communication (ISAC) has emerged as a pivotal enabling technology for next-generation wireless systems, transforming traditionally isolated radar and communication networks into unified, multifunctional platforms [1]-[3]. By jointly leveraging spectral resources, hardware architectures, and advanced signal processing techniques, ISAC achieves substantial improvements in spectrum efficiency and system integration, enabling emerging applications such as autonomous driving, smart manufacturing, and environment-aware wireless connectivity [4], [5]. Among various waveform candidates, orthogonal frequency division multiplex-

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ing (OFDM) has attracted particular attention owing to its widespread adoption in contemporary wireless standards (e.g., 5G NR, Wi-Fi 6/7) and its inherent suitability for ISAC applications, such as efficient decoupled estimation of delay and Doppler [6], [7], flexible time-frequency resource allocation [8], [9], and low range sidelobes [10], [11].

A fundamental constraint in OFDM-based ISAC systems stems from the cyclic prefix (CP), originally introduced in communication systems to combat inter-symbol interference (ISI) caused by multipath propagation. For communication systems, CP durations are designed based on microsecond-level delay spreads, typically adequate for wireless channels. However, radar sensing applications must cope with echoes arriving from kilometer-scale distances, implying significantly longer round-trip delays. As a result, echoes exceeding the CP duration inevitably induce both ISI and inter-carrier interference (ICI). For instance, considering the 5G NR normal CP configuration with a subcarrier spacing of 120kHz, the standard CP duration is only 0.59μs, theoretically restricting the interference-free sensing range to roughly 90 meters [12]. Thus, practical ISAC systems must reliably operate *beyond the CP limit*, where sensing requirements fundamentally diverge from communication specifications.

Several studies have recently investigated the impact of insufficient CP on OFDM sensing performance [13]-[17]. These analyses reveal that when target echoes extend beyond the CP duration, significant ISI and ICI arise, causing elevated sidelobe levels in the range-Doppler map (RDM) and substantially degraded parameter estimation accuracy [13], [14], [15]. Unlike communication systems, where moderate ISI can typically be mitigated through equalization, radar sensing relies heavily on coherent signal accumulation across subcarriers and symbols, making it inherently more sensitive to structured interference. Existing theoretical analyses, however, often simplify ISI/ICI modeling by assuming statistical independence from interference-free components or by adopting Gaussian approximations. These simplifications limit analytical precision and fail to capture the intrinsic structured coupling of ISI/ICI components in OFDM sensing scenarios. Consequently, a rigorous and accurate analytical performance characterization for OFDM-ISAC systems beyond the CP limit remains largely unexplored.

Aside from performance characterization, recent works have also examined methods for mitigating the detrimental effects of insufficient CP. Current mitigation approaches fall broadly into two categories: transmitter-side waveform design and receiver-side compensation. On the transmitter side, methods

such as CP structure modification [15], [18], [19], pilot pattern redesign [20], and symbol-level waveform optimization [21] have been proposed. Although these designs effectively suppress ISI and ICI, they often compromise spectral efficiency and deviate from standardized OFDM formats. On the receiver side, methods such as time-domain coherent compensation (TDCC) [13], [14], virtual-CP reconstruction [22], frequency-domain coherent compensation (FDCC) [23], and multi-target coherent compensation (MTCC) [23] have been introduced. These receiver-side solutions strengthen the interference-free signal component through coherent compensation, but they simultaneously amplify noise and residual interference, failing to entirely eliminate ISI and ICI. Therefore, effective mitigation algorithms compatible with standard OFDM signaling formats and capable of complete interference cancellation remain absent in the existing literature.

Motivated by these critical gaps, this paper develops a unified analytical and algorithmic framework for OFDM-ISAC sensing beyond the CP limit, encompassing signal modeling, theoretical analysis, and interference mitigation algorithms. The main contributions are summarized as follows:

- We establish a general OFDM-ISAC echo model that explicitly captures the structured coupling between ISI and ICI caused by an insufficient CP. This model reveals how CP insufficiency breaks the separable range-Doppler structure and introduces coherent interference across subcarriers and OFDM symbols, motivating our theoretical performance analysis and algorithm design.
- We derive closed-form analytical expressions for the sensing signal-to-interference-plus-noise ratio (SINR) and the range-Doppler peak sidelobe level ratio (PSLR) without invoking independence or Gaussianity assumptions on the interference terms. The analysis shows that both SINR degradation and sidelobe elevation increase approximately linearly with the normalized excess delay beyond the CP, providing quantitative insight into the trade-off between CP length and sensing performance.
- We propose two frequency-domain interference cancellation algorithms, SIC-DFT and SIC-ESPRIT, to mitigate ISI and ICI due to insufficient CP. SIC-DFT leverages low-complexity DFT-based estimation, while SIC-ESPRIT exploits subspace-based parameter recovery to achieve super-resolution sensing accuracy. Both methods are fully compatible with standard OFDM signaling and require no waveform modification.
- Extensive simulations validate the theoretical analysis and confirm the effectiveness of the proposed algorithms. The results demonstrate that both SIC-DFT and SIC-ESPRIT consistently outperform existing benchmark methods, achieving more than a 4dB SINR improvement under the same CP-insufficient conditions. In particular, SIC-ESPRIT exhibits superior estimation accuracy, reducing both range and velocity root-mean-square-errors (RMSEs) by approximately one order of magnitude compared with competing schemes, approaching the performance of systems employing a sufficiently long CP. These findings verify that the proposed framework enables reliable and

high-fidelity OFDM-ISAC sensing even when operating beyond the CP limit.

*Notation:* Boldface lowercase and uppercase letters denote vectors and matrices, respectively. The operators  $(\cdot)^*$ ,  $(\cdot)^T$ , and  $(\cdot)^H$  denote conjugate, transpose, and Hermitian transpose. The symbols  $\otimes$ ,  $\odot$ , and  $\|\cdot\|_F$  represent Kronecker product, Hadamard product, and Frobenius norm. The notation  $\mathbb{E}\{\cdot\}$  denotes statistical expectation,  $\mathcal{CN}(\mu, \sigma^2)$  represents complex Gaussian distribution,  $\text{diag}\{\cdot\}$  constructs a diagonal matrix from its vector argument,  $\text{vec}(\cdot)$  denotes vectorization, and  $\mathbf{I}_N$  is the  $N \times N$  identity matrix. We use  $j = \sqrt{-1}$  for the imaginary unit and  $\triangleq$  for definitions.

## II. SIGNAL MODEL AND ECHO SIGNAL PROCESSING

### A. Transmit Signal Model

Consider an OFDM frame with  $M$  symbols and  $N$  subcarriers, for which the baseband transmit signal including the CP can be expressed as

$$x(t) = \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} \frac{s_{n,m}}{\sqrt{N}} e^{j2\pi n \Delta_f (t - mT_s - T_{\text{cp}})} g\left(\frac{t - mT_s}{T_s}\right), \quad (1)$$

where  $s_{n,m}$  denotes the data symbol on the  $n$ -th subcarrier of the  $m$ -th OFDM symbol,  $\Delta_f$  denotes the subcarrier spacing, the total OFDM symbol duration including the CP is  $T_s = T + T_{\text{cp}}$  with CP duration  $T_{\text{cp}}$  and “useful” OFDM symbol duration (i.e., IFFT duration)  $T = 1/\Delta_f$ , and  $g(\xi)$  is a rectangular pulse shaping function, i.e.,  $g(\xi) = 1$  for  $0 \leq \xi < 1$  and 0 otherwise. The symbols are normalized such that  $\mathbb{E}\{|s_{n,m}|^2\} = 1$  and  $\mathbb{E}\{s_{n,m}\} = 0$ . For most typical constellations such as PSK and QAM, except BPSK and 8-QAM,  $\mathbb{E}\{s_{n,m}^2\} = 0$  also holds.

### B. Radar Echo Signal Model

The sensing receiver activates its detection window immediately after transmission to ensure that echoes from nearby targets are not missed. Suppose there exist  $Q$  point targets at ranges  $R_q$  with relative radial velocities  $v_q$  and radar cross section (RCS)  $\sigma_{\text{rcs},q}$  for  $q = 1, \dots, Q$ . The echo signal at the sensing receiver is expressed as

$$y(t) = \sum_{q=1}^Q \alpha_q x(t - \tau_q) e^{j2\pi f_{d,q} t} + z(t), \quad (2)$$

where  $\alpha_q$  denotes the complex reflection coefficient of the  $q$ -th target and  $z(t)$  is additive white Gaussian noise (AWGN). The reflection coefficient  $\alpha_q$ , round-trip delay  $\tau_q$  and Doppler shift  $f_{d,q}$  of the  $q$ -th target are respectively defined by

$$\alpha_q = \sqrt{\frac{\sigma_{\text{rcs},q} c_0^2}{(4\pi)^3 R_q^4 f_c^2}}, \quad \tau_q = \frac{2R_q}{c_0}, \quad f_{d,q} = \frac{2v_q f_c}{c_0}, \quad (3)$$

where  $c_0$  is the speed of light and  $f_c$  is the carrier frequency. Following [24]-[26], the Doppler-induced phase rotation over the duration of one OFDM symbol is assumed to be negligible, which is well-justified under standard 5G NR parameter configurations [12]. For instance, even at the highest FR3 frequency  $f_c = 24.5\text{GHz}$ , a target with velocity  $v = 30\text{m/s}$

$$\tilde{y}_{m,q}[i] = \frac{\alpha_q}{\sqrt{N}} \sum_{n=0}^{N-1} s_{n,m-1} e^{j\frac{2\pi}{N}ni} e^{-j2\pi n \Delta_f (\tau_q - T_{cp})} e^{j2\pi(m-1)f_{d,q}T_s} g_1(i) + \frac{\alpha_q}{\sqrt{N}} \sum_{n=0}^{N-1} s_{n,m} e^{j\frac{2\pi}{N}ni} e^{-j2\pi n \Delta_f \tau_q} e^{j2\pi m f_{d,q} T_s} g_2(i). \quad (4)$$

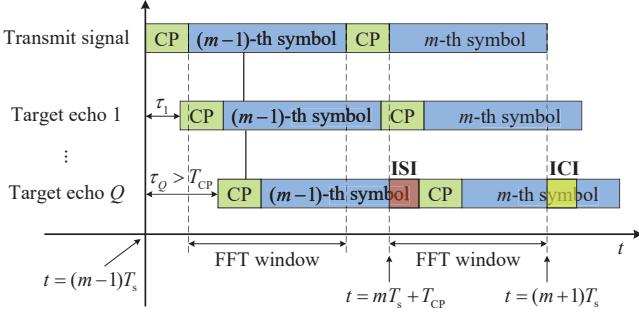


Fig. 1. Illustration of transmit and echo signals. ISI and ICI occur when the CP duration  $T_{cp}$  is shorter than the maximum target delay  $\tau_Q$ .

induces a Doppler shift of only  $f_d = 4.85\text{kHz}$ , which is typically less than  $0.1\Delta_f$ , a commonly used threshold. Therefore, except for very high-speed scenarios, the Doppler-induced ICI is limited, and thus we focus on analysis and mitigation of the CP-induced ICI. Under this assumption, the echo signal in (2) can be rewritten as

$$y(t) \approx \sum_{q=1}^Q \frac{\alpha_q}{\sqrt{N}} \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} s_{n,m} e^{j2\pi n \Delta_f (t - mT_s - \tau_q - T_{cp})} \times e^{j2\pi m f_{d,q} T_s} g\left(\frac{t - mT_s - \tau_q}{T_s}\right) + z(t), \quad (5)$$

At the sensing receiver, the radar echo signal  $y(t)$  is sampled every  $1/B$  sec., where  $B = N\Delta_f$  is the signal bandwidth. The resulting discrete-time echo signal is thus represented as

$$y[i] = \sum_{q=1}^Q \frac{\alpha_q}{\sqrt{N}} \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} s_{n,m} e^{j\frac{2\pi}{N}n(i - mN_s - N_{cp})} \quad (6a)$$

$$\times e^{-j2\pi n \Delta_f \tau_q} e^{j2\pi m f_{d,q} T_s} g\left(\frac{i - mN_s - l_q}{N_s}\right) + z[i], \quad (6b)$$

where  $l_q = [\tau_q B]$ ,  $N_{cp} = T_{cp}B$  denotes the CP length,  $N_s = N + N_{cp}$ , and  $[.]$  is the rounding operation. For the echo signal corresponding to the  $q$ -th target, the  $m$ -th symbol spans the sample indices from  $i = mN_s + l_q$  to  $i = (m+1)N_s + l_q - 1$ . After CP removal, OFDM demodulation is applied to the  $N$  samples from  $i = mN_s + N_{cp}$  to  $i = (m+1)N_s - 1$ .

1) *Sufficient CP Length Case:* When the delay of the  $q$ -th target does not exceed the CP length, i.e.,  $l_q \leq N_{cp}$ , neither ISI nor ICI occurs. In this case, the detection window for the  $m$ -th symbol fully covers the entire duration of the current transmitted symbol (as exemplified by the echo of the first target in Fig. 1), ensuring subcarrier orthogonality and interference-free reception. The discrete-time echo signal for the  $q$ -th target can be expressed as

$$y_{m,q}[i] = \sum_{n=0}^{N-1} \frac{\alpha_q s_{n,m}}{\sqrt{N}} e^{j\frac{2\pi}{N}ni} e^{-j2\pi n \Delta_f \tau_q} e^{j2\pi m f_{d,q} T_s} g\left(\frac{i}{N}\right). \quad (7)$$

After OFDM demodulation, the frequency-domain echo on the  $n$ -th subcarrier of the  $m$ -th symbol from the  $q$ -th target is

$$y_{n,m,q} = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} y_{m,q}[i] e^{-j\frac{2\pi}{N}ni} \quad (8a)$$

$$= \frac{\alpha_q}{N} e^{j2\pi m f_{d,q} T_s} \sum_{n'=0}^{N-1} s_{n',m} e^{-j2\pi n' \Delta_f \tau_q} \sum_{i=0}^{N-1} e^{j\frac{2\pi}{N}(n' - n)i} \quad (8b)$$

$$= \alpha_q s_{n,m} e^{-j2\pi n \Delta_f \tau_q} e^{j2\pi m f_{d,q} T_s}. \quad (8c)$$

2) *Insufficient CP Length Case:* When the delay of the  $q$ -th target exceeds the CP length, i.e.,  $l_q > N_{cp}$ , both ISI and ICI arise and degrade sensing performance. As illustrated in Fig. 1, for the echo of the  $Q$ -th target, the FFT window of the  $m$ -th symbol overlaps the tail of symbol  $m-1$  (yielding ISI) and truncates the body of symbol  $m$  (breaking subcarrier orthogonality and inducing ICI). In this case, the discrete-time echo of the  $q$ -th target within the  $m$ -th symbol is given by (4) at the top of this page, where  $g_1(i) = g\left(\frac{i}{l_q - N_{cp}}\right)$  selects the first  $l_q - N_{cp}$  samples (residual from symbol  $m-1$ ), and  $g_2(i) = g\left(\frac{i - l_q + N_{cp}}{N - l_q + N_{cp}}\right)$  selects the remaining  $N - (l_q - N_{cp})$  samples of symbol  $m$ . After OFDM demodulation, the corresponding frequency-domain echo  $\tilde{y}_{n,m,q}$  can be expressed as<sup>1</sup>

$$\tilde{y}_{n,m,q} = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} \tilde{y}_{m,q}[i] e^{-j\frac{2\pi}{N}ni} \quad (9a)$$

$$= \alpha_q s_{n,m} e^{-j2\pi n \Delta_f \tau_q} e^{j2\pi m f_{d,q} T_s} + y_{n,m,q}^{\text{ISI}} - y_{n,m,q}^{\text{ICI}}, \quad (9b)$$

where  $y_{n,m,q}^{\text{ISI}}$  and  $y_{n,m,q}^{\text{ICI}}$  denote the ISI and ICI components, respectively:

$$y_{n,m,q}^{\text{ISI}} = \alpha_q e^{j2\pi(m-1)f_{d,q} T_s} \sum_{n'=0}^{N-1} s_{n',m-1} e^{j2\pi n' \Delta_f (T_{cp} - \tau_q)} \phi_{n,n'}^q, \quad (10a)$$

$$y_{n,m,q}^{\text{ICI}} = \alpha_q e^{j2\pi m f_{d,q} T_s} \sum_{n'=0}^{N-1} s_{n',m} e^{-j2\pi n' \Delta_f \tau_q} \phi_{n,n'}^q, \quad (10b)$$

$$\text{where } \phi_{n,n'}^q = \frac{1}{N} \sum_{i=0}^{l_q - N_{cp} - 1} e^{j\frac{2\pi}{N}(n' - n)i}.$$

To distinguish between the interference-free and ISI/ICI-contaminated echoes, the targets are partitioned according to their round-trip delays. Without loss of generality, we order the targets such that  $\tau_1 \leq \dots \leq \tau_Q$  and assume that the first  $\tilde{Q}$  targets have round-trip delays satisfying  $l_q \leq N_{cp}$ , for  $q = 1, \dots, \tilde{Q}$ . Thus, the frequency-domain echo on the  $n$ -th

<sup>1</sup>The negative sign before the ICI term originates from the time-domain signal construction: The actual received signal of the  $m$ -th symbol can be viewed as the interference-free signal minus the portion of samples that fall outside the CP.

subcarrier of the  $m$ -th OFDM symbol is given by

$$y_{n,m} = \sum_{q=1}^{\tilde{Q}} y_{n,m,q} + \sum_{q=\tilde{Q}+1}^Q \tilde{y}_{n,m,q} \quad (11a)$$

$$\begin{aligned} &= \sum_{q=1}^Q \underbrace{\alpha_q s_{n,m} e^{-j2\pi n \Delta_f \tau_q} e^{j2\pi m f_{d,q} T_s}}_{\text{ISI/ICI-free component, } y_{n,m}^{\text{free}}} \\ &+ \underbrace{\sum_{q=\tilde{Q}+1}^Q y_{n,m,q}^{\text{ISI}}}_{\text{ISI component, } y_{n,m}^{\text{ISI}}} - \underbrace{\sum_{q=\tilde{Q}+1}^Q y_{n,m,q}^{\text{ICI}}}_{\text{ICI component, } y_{n,m}^{\text{ICI}}} + z_{n,m}, \quad (11b) \end{aligned}$$

where  $z_{n,m} \sim \mathcal{CN}(0, \sigma^2)$  denotes AWGN,  $\sigma^2 = F k_b \Delta_f T_{\text{temp}}$  is the noise power,  $F$  is the receiver's noise figure,  $k_b$  Boltzmann's constant, and  $T_{\text{temp}}$  the equivalent noise temperature.

For notational compactness, we define the frequency-domain steering vector, temporal steering vector, and ISI/ICI-induced phase matrix as

$$\mathbf{b}(\tau_q) \triangleq [1, e^{-j2\pi \Delta_f \tau_q}, \dots, e^{-j2\pi (N-1) \Delta_f \tau_q}]^T, \quad (12a)$$

$$\mathbf{c}(f_{d,q}) \triangleq [1, e^{-j2\pi f_{d,q} T_s}, \dots, e^{-j2\pi (M-1) f_{d,q} T_s}]^T, \quad (12b)$$

$$[\Phi_q]_{n,n'} \triangleq \phi_{n,n'}^q. \quad (12c)$$

The ISI/ICI-free, ISI, and ICI components for the  $m$ -th symbol can thus be written compactly as

$$\mathbf{y}_m^{\text{free}} = \sum_{q=1}^Q \alpha_q \mathbf{b}(\tau_q) [\mathbf{c}^*(f_{d,q})]_m \odot \mathbf{s}_m, \quad (13a)$$

$$\mathbf{y}_m^{\text{ISI}} = \sum_{q=\tilde{Q}+1}^Q \alpha_q \Phi_q (\mathbf{b}(\tau_q - T_{\text{cp}}) [\mathbf{c}^*(f_{d,q})]_{m-1} \odot \mathbf{s}_{m-1}), \quad (13b)$$

$$\mathbf{y}_m^{\text{ICI}} = \sum_{q=\tilde{Q}+1}^Q \alpha_q \Phi_q (\mathbf{b}(\tau_q) [\mathbf{c}^*(f_{d,q})]_m \odot \mathbf{s}_m), \quad (13c)$$

where

$$\mathbf{y}_m^{\text{free}} \triangleq [y_{0,m}^{\text{free}}, \dots, y_{N-1,m}^{\text{free}}]^T, \quad (14a)$$

$$\mathbf{y}_m^{\text{ISI}} \triangleq [y_{0,m}^{\text{ISI}}, \dots, y_{N-1,m}^{\text{ISI}}]^T, \quad (14b)$$

$$\mathbf{y}_m^{\text{ICI}} \triangleq [y_{0,m}^{\text{ICI}}, \dots, y_{N-1,m}^{\text{ICI}}]^T, \quad (14c)$$

$$\mathbf{s}_m \triangleq [s_{0,m}, \dots, s_{N-1,m}]^T. \quad (14d)$$

Aggregating (13) over  $M$  symbols, the ISI/ICI-free, ISI, and ICI components over a frame can be further written in matrix form as

$$\mathbf{Y}_{\text{free}} = \sum_{q=1}^Q \alpha_q (\mathbf{b}(\tau_q) \mathbf{c}^H(f_{d,q}) \odot \mathbf{S}), \quad (15a)$$

$$\mathbf{Y}_{\text{ISI}} = \sum_{q=\tilde{Q}+1}^Q \alpha_q \Phi_q (\mathbf{b}(\tau_q - T_{\text{cp}}) \mathbf{c}^H(f_{d,q}) \odot \mathbf{S}) \mathbf{J}_1, \quad (15b)$$

$$\mathbf{Y}_{\text{ICI}} = \sum_{q=\tilde{Q}+1}^Q \alpha_q \Phi_q (\mathbf{b}(\tau_q) \mathbf{c}^H(f_{d,q}) \odot \mathbf{S}), \quad (15c)$$

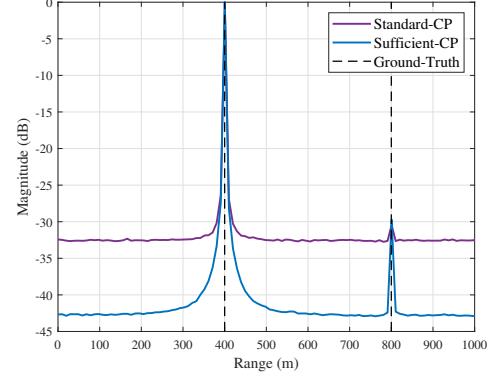


Fig. 2. Range profiles generated using the 2D-DFT sensing algorithm under standard-CP and sufficient-CP configurations.

where  $\mathbf{Y}_{\text{free}} \triangleq [y_0^{\text{free}}, \dots, y_{M-1}^{\text{free}}]$ ,  $\mathbf{Y}_{\text{ISI}} \triangleq [y_0^{\text{ISI}}, \dots, y_{M-1}^{\text{ISI}}]$ ,  $\mathbf{Y}_{\text{ICI}} \triangleq [y_0^{\text{ICI}}, \dots, y_{M-1}^{\text{ICI}}]$ ,  $\mathbf{S} \triangleq [\mathbf{s}_0, \dots, \mathbf{s}_{M-1}]$ , and  $\mathbf{J}_1 \in \mathbb{C}^{M \times M}$  is a shift matrix defined as

$$\mathbf{J}_1 \triangleq \begin{bmatrix} \mathbf{0}_{M-1} & \mathbf{I}_{M-1} \\ 0 & \mathbf{0}_{M-1}^T \end{bmatrix}. \quad (16)$$

Finally, the sensing observation matrix can be expressed as

$$\mathbf{Y} = \mathbf{Y}_{\text{free}} + \mathbf{Y}_{\text{ISI}} - \mathbf{Y}_{\text{ICI}} + \mathbf{Z}, \quad (17)$$

where the noise matrix  $\mathbf{Z} \in \mathbb{C}^{N \times M}$  consists of independent and identically distributed (i.i.d.) complex Gaussian entries, i.e.,  $[\mathbf{Z}]_{n,m} = z_{n,m}$  with  $\text{vec}(\mathbf{Z}) \sim \mathcal{CN}(\mathbf{0}, \sigma^2 \mathbf{I}_{MN})$ . Given the known data matrix  $\mathbf{S}$ , the objective of OFDM-based radar sensing is to jointly estimate the target parameters, i.e. the reflection coefficients  $\{\alpha_q\}_{q=1}^Q$ , delays  $\{\tau_q\}_{q=1}^Q$ , and Doppler shifts  $\{f_{d,q}\}_{q=1}^Q$ , from the observation matrix  $\mathbf{Y}$ .

### III. SENSING PERFORMANCE ANALYSIS

The signal model developed in Section II reveals that an insufficient CP fundamentally distorts the structure of the received echo signal, introducing ISI and ICI. In this section, we analyze how CP insufficiency affects the sensing performance of OFDM-ISAC systems. Specifically, we characterize the ISI/ICI-induced degradation in terms of the SINR and the sidelobe level of the RDM. This analysis provides insight into the sensing impairment caused by an insufficient CP and establishes the theoretical foundation for the interference cancellation algorithms developed in Section IV.

#### A. Impact of Insufficient CP

The interference-free component  $\mathbf{Y}_{\text{free}}$  of the signal model in (15) exhibits a separable structure, represented by a rank- $Q$  sum of outer products  $\mathbf{b}(\tau_q) \mathbf{c}^H(f_{d,q})$  modulated by the data matrix  $\mathbf{S}$ . After matched filtering and 2D-DFT processing, this component produces a well-localized RDM, allowing clear resolution of the target delays and Doppler shifts.

In contrast, the interference components  $\mathbf{Y}_{\text{ISI}}$  and  $\mathbf{Y}_{\text{ICI}}$  depart from this desirable separable structure. Both involve the phase-rotation matrix  $\Phi_q$ , which induces off-diagonal spectral leakage across subcarriers and destroys the subcarrier orthogonality. Moreover, the temporal-shift matrix  $\mathbf{J}_1$  in  $\mathbf{Y}_{\text{ISI}}$  couples

adjacent OFDM symbols and introduces ISI. These distortions prevent representing  $\mathbf{Y}_{\text{ISI}}$  and  $\mathbf{Y}_{\text{ICI}}$  as simple separable rank- $Q$  matrices. As a result, when standard matched filtering and 2D-DFT processing are applied, the interference components manifest as degraded SINR and elevated RDM sidelobes, impairing both delay and Doppler estimation accuracy.

This phenomenon is illustrated in Fig. 2. In this example<sup>2</sup>, two targets are positioned at ranges of 400m and 800m. Under the *standard-CP* configuration defined by 3GPP [12], the CP duration is only  $0.59\mu\text{s}$ , corresponding to an interference-free range of approximately 90m. Consequently, both target echoes arrive well beyond the CP guard interval, inevitably causing ISI and ICI. For reference, we introduce a customized *sufficient-CP* configuration, whose CP duration is extended to  $8.33\mu\text{s}$ , equal to one OFDM symbol period, resulting in an interference-free range of approximately 1250m. Under this sufficient-CP scenario, all target echoes arrive within the CP duration, entirely eliminating ISI/ICI and thus providing an ideal performance baseline. Fig. 2 clearly demonstrates that the standard-CP scenario produces significantly elevated sidelobe levels compared to the sufficient-CP baseline, highlighting the substantial performance degradation induced by CP insufficiency. To rigorously quantify these detrimental impacts, the following sections derive analytical expressions for the resulting SINR and sidelobe characteristics, laying the foundation for our proposed interference-mitigation algorithms.

### B. SINR Analysis

The SINR of the echo signal  $\mathbf{Y}$  is defined as the ratio between the interference-free signal power and the combined interference and noise powers:

$$\text{SINR} \triangleq \frac{\mathbb{E}\{\|\mathbf{Y}_{\text{free}}\|_F^2\}}{\mathbb{E}\{\|\mathbf{Y}_{\text{ISI}} - \mathbf{Y}_{\text{ICI}}\|_F^2\} + \mathbb{E}\{\|\mathbf{Z}\|_F^2\}}. \quad (18)$$

This metric characterizes the relative strength of the desired echo component  $\mathbf{Y}_{\text{free}}$  compared with the interference caused by  $\mathbf{Y}_{\text{ISI}}$ ,  $\mathbf{Y}_{\text{ICI}}$ , and  $\mathbf{Z}$ . It serves as a key indicator of the degradation in delay and Doppler estimation accuracy when the CP is insufficient. Based on the signal model in (15), the SINR can be derived in closed form as follows.

**Proposition 1.** *The SINR of the echo signal  $\mathbf{Y}$  is given by*

$$\text{SINR} = \frac{MN \sum_{q=1}^Q |\alpha_q|^2}{(2M-1)N \sum_{q=\tilde{Q}+1}^Q \rho_q |\alpha_q|^2 + MN \sigma^2}, \quad (19)$$

where  $\rho_q = (l_q - N_{\text{cp}})/N$  is the normalized excess delay beyond the CP for the  $q$ -th target.

*Proof:* See Appendix A. ■

The above expression is derived without assuming the statistical independence of ISI/ICI or approximating them as Gaussian noise, and thus provides a general characterization of SINR performance under CP insufficiency. To gain further

<sup>2</sup>The simulation settings for this example are consistent with those in Table I in the simulation section.

insight, note that for a large number of OFDM symbols ( $M \gg 1$ ), the SINR in (19) is approximately

$$\text{SINR} \approx \frac{\sum_{q=1}^Q |\alpha_q|^2}{2 \sum_{q=\tilde{Q}+1}^Q \rho_q |\alpha_q|^2 + \sigma^2}. \quad (20)$$

This expression highlights the pivotal role of the normalized excess delay beyond the CP,  $\rho_q = (l_q - N_{\text{cp}})/N$ , in determining SINR performance. If all echoes lie within the CP, the interference term vanishes and the SINR reduces to the noise-limited SNR. Once any echo exceeds the CP ( $\rho_q > 0$ ), the interference grows approximately linearly with  $\rho_q$ , and hence the SINR monotonically decreases with the normalized excess delay beyond the CP.

### C. Sidelobe Level Analysis

While SINR quantifies the overall performance degradation, it does not capture the spatial distribution of interference in the delay-Doppler domain. We now analyze how CP insufficiency elevates sidelobe levels in the RDM, which directly impacts the ability to detect weak targets. The RDM obtained using the 2D-DFT can be expressed as [27], [28]

$$\chi = \mathbf{F}_N^H (\mathbf{Y} \odot \mathbf{S}^*) \mathbf{F}_M, \quad (21)$$

where  $\mathbf{F}_M$  and  $\mathbf{F}_N^H$  denote the normalized DFT and inverse-DFT (IDFT) matrices, respectively. Since  $\chi$  is random due to data modulation, target reflection coefficients, and noise, the sensing performance is evaluated using its second-order moment  $\mathbb{E}\{|\chi(l, \nu)|^2\}$ .

**Proposition 2.** *The second-order moment of the RDM is*

$$\mathbb{E}\{|\chi(l, \nu)|^2\} = \sum_{q=1}^Q \frac{|\tilde{\alpha}_q|^2}{MN} |D_N(l - \tilde{l}_q)|^2 |D_M(\nu - \tilde{\nu}_q)|^2 + \sigma_{\text{SL}}^2, \quad (22)$$

where  $D_N(x) = \frac{\sin(\pi x)}{\sin(\pi x/N)} e^{j\pi(N-1)x/N}$  denotes the Dirichlet kernel,  $\tilde{l}_q = \tau_q B$  and  $\tilde{\nu}_q = f_{d,q} M T_s$  represent the normalized delay and Doppler,  $\mu_4 = \mathbb{E}\{|s_{n,m}|^4\}$ , and the effective reflection coefficient  $\tilde{\alpha}_q$  is defined as

$$\tilde{\alpha}_q \triangleq \begin{cases} \alpha_q, & q = 1, \dots, \tilde{Q}; \\ (1 - \rho_q)\alpha_q, & q = \tilde{Q} + 1, \dots, Q. \end{cases} \quad (23)$$

In addition, the sidelobe level is given by

$$\sigma_{\text{SL}}^2 = (\mu_4 - 1) \sum_{q=1}^Q |\tilde{\alpha}_q|^2 + \sum_{q=\tilde{Q}+1}^Q \rho_q (2 - \rho_q) |\alpha_q|^2 + \sigma^2, \quad (24)$$

where the first term originates from the randomness of the communication symbols, and the second term accounts for the sidelobe leakage caused by ISI/ICI.

*Proof:* See Appendix B. ■

When the normalized delay and Doppler are integers,  $\mathbb{E}\{|\chi(l, \nu)|^2\}$  simplifies to

$$\mathbb{E}\{|\chi(l, \nu)|^2\} = \begin{cases} MN |\tilde{\alpha}_q|^2 + \sigma_{\text{SL}}^2, & (l, \nu) = (\tilde{l}_q, \tilde{\nu}_q); \\ \sigma_{\text{SL}}^2, & \text{otherwise.} \end{cases} \quad (25)$$

To further evaluate the sidelobe level, we define the PSLR for the  $q$ -th target as [29]

$$\gamma_q \triangleq \frac{\mathbb{E}\left\{\max_{(l,\nu) \in \mathcal{R}_s} |\chi(l,\nu)|^2\right\}}{\mathbb{E}\{|\chi(\tilde{l}_q, \tilde{\nu}_q)|^2\}}, \quad (26)$$

where  $\mathcal{R}_s = \{(l,\nu) \mid (l,\nu) \neq (\tilde{l}_q, \tilde{\nu}_q), \forall q\}$  denotes the sidelobe region excluding the true target bins. Assuming  $(l,\nu)$  take integer values,  $\chi(l,\nu) \sim \mathcal{CN}(0, \sigma_{\text{SL}}^2)$  in  $\mathcal{R}_s$ . According to [30], [31], the expected peak sidelobe level is

$$\mathbb{E}\left\{\max_{(l,\nu) \in \mathcal{R}_s} |\chi(l,\nu)|^2\right\} = H_Q \sigma_{\text{SL}}^2, \quad (27)$$

where  $H_Q = \sum_{q=1}^{MN-Q} 1/q$  denotes the harmonic number. Then, the PSLR can be expressed as

$$\gamma_q = \frac{H_Q \sigma_{\text{SL}}^2}{MN|\tilde{\alpha}_q|^2 + \sigma_{\text{SL}}^2}. \quad (28)$$

From (28), it is evident that an insufficient CP, characterized by the normalized excess delay  $\rho_q$ , concurrently raises the sidelobe floor and weakens the mainlobe return. As  $\rho_q$  increases (i.e., a larger fraction of the echo falls outside the CP guard interval), the leakage of echo energy into the interference increases the noise floor  $\sigma_{\text{SL}}^2$  (boosting the sidelobe baseline), while the effective target amplitude is reduced to  $\tilde{\alpha}_q = (1 - \rho_q)\alpha_q$ , diminishing the captured mainlobe power. Thus the PSLR  $\gamma_q$  in (28) increases monotonically with CP insufficiency.

#### IV. ITERATIVE ALGORITHMS FOR BEYOND-CP SENSING

Building on the analytical results in Section III, here we develop practical sensing algorithms that mitigate the ISI and ICI induced by an insufficient CP. The proposed framework adopts an SIC strategy that iteratively estimates target parameters, reconstructs the interference, and refines the interference-free echo signal. Two implementations are presented: a low-complexity SIC-DFT algorithm based on conventional 2D-DFT processing, and a high-resolution SIC-ESPRIT algorithm that exploits subspace-based parameter estimation.

##### A. SIC-DFT: Efficient Iterative Cancellation

For SIC-DFT, the interference-free observation is initialized as  $\mathbf{Y}_{\text{free}}^0 = \mathbf{Y}$ . At the  $k$ -th iteration, the RDM is obtained using

$$\mathbf{x}^k = \mathbf{F}_N^H (\mathbf{Y}_{\text{free}}^k \odot \mathbf{S}^*) \mathbf{F}_M. \quad (29)$$

A constant false alarm rate (CFAR) detector is then applied to  $\mathbf{x}^k$  to identify the targets and obtain the corresponding delay and Doppler estimates  $\hat{\tau}_q^k$  and  $\hat{f}_{d,q}^k$ . Given these estimates, the reflection coefficient of the  $q$ -th target is estimated via least squares (LS) [32]:

$$\hat{\alpha}_q^k = \frac{\mathbf{b}^H(\hat{\tau}_q^k)(\mathbf{Y}_{\text{free}}^k \odot \mathbf{S}^*) \mathbf{c}(\hat{f}_{d,q}^k)}{\|\mathbf{b}(\hat{\tau}_q^k)\|^2 \|\mathbf{c}(\hat{f}_{d,q}^k)\|^2}. \quad (30)$$

Using the current parameter estimates, the ISI and ICI components are reconstructed as

$$\mathbf{Y}_{\text{ISI}}^k = \sum_{q=\tilde{Q}+1}^Q \hat{\alpha}_q^k \Phi_q (\mathbf{b}(\hat{\tau}_q^k - T_{\text{cp}}) \mathbf{c}^H(\hat{f}_{d,q}^k) \odot \mathbf{S}) \mathbf{J}_1, \quad (31a)$$

---

#### Algorithm 1 SIC-DFT Algorithm

---

**Input:**  $\mathbf{Y}, \mathbf{S}, \delta_{\text{th}}, K_{\text{max}}$ .

**Output:**  $\hat{\tau}_q^*, \hat{f}_{d,q}^*, \forall q$

- 1: Initialize the interference-free signal  $\mathbf{Y}_{\text{free}}^0 = \mathbf{Y}, k = 0$ .
  - 2: **repeat**
  - 3:   Compute the RDM from  $\mathbf{Y}_{\text{free}}^k$  using (29).
  - 4:   Estimate delay  $\hat{\tau}_q^k$  and Doppler  $\hat{f}_{d,q}^k$  with a CFAR detector.
  - 5:   Estimate target reflection coefficient  $\hat{\alpha}_q^k$  using (30).
  - 6:   Reconstruct ISI and ICI components  $\mathbf{Y}_{\text{ISI}}^k, \mathbf{Y}_{\text{ICI}}^k$  using (31).
  - 7:   Compute the interference-free signal  $\mathbf{Y}_{\text{free}}^{k+1}$  using (32).
  - 8:    $k := k + 1$ .
  - 9: **until** (33) is satisfied.
  - 10: Return  $\hat{\tau}_q^* = \hat{\tau}_q^k, \hat{f}_{d,q}^* = \hat{f}_{d,q}^k, \forall q$ .
- 

#### Algorithm 2 SIC-ESPRIT Algorithm

---

**Input:**  $\mathbf{Y}, \mathbf{S}, \delta_{\text{th}}, K_{\text{max}}$ .

**Output:**  $\hat{\tau}_q^*, \hat{f}_{d,q}^*, \forall q$

- 1: Initialize the interference-free signal  $\mathbf{Y}_{\text{free}}^0 = \mathbf{Y}, k = 0$ .
  - 2: **repeat**
  - 3:   Estimate the sensing channel by  $\hat{\mathbf{h}} = \text{vec}(\mathbf{Y}_{\text{free}}^k \odot \mathbf{S}^*)$ .
  - 4:   Apply spatial smoothing to generate multiple snapshots and construct the sample covariance matrix  $\hat{\mathbf{R}}$ .
  - 5:   Perform EVD of  $\hat{\mathbf{R}}$  and obtain the signal subspace  $\mathbf{U}_s$ .
  - 6:   Compute  $\mathbf{P}_\tau$  and  $\mathbf{P}_f$  using (37a) and (37b).
  - 7:   Estimate  $\hat{\tau}_q^k$  and  $\hat{f}_{d,q}^k$  using (38a) and (38b).
  - 8:   Estimate target reflection coefficient  $\hat{\alpha}_q^k$  using (30).
  - 9:   Reconstruct ISI and ICI components  $\mathbf{Y}_{\text{ISI}}^k, \mathbf{Y}_{\text{ICI}}^k$  using (31).
  - 10:   Compute the interference-free signal  $\mathbf{Y}_{\text{free}}^{k+1}$  using (32).
  - 11:    $k := k + 1$ .
  - 12: **until** (33) is satisfied.
  - 13: Return  $\hat{\tau}_q^* = \hat{\tau}_q^k, \hat{f}_{d,q}^* = \hat{f}_{d,q}^k, \forall q$ .
- 

$$\mathbf{Y}_{\text{ICI}}^k = \sum_{q=\tilde{Q}+1}^Q \hat{\alpha}_q^k \Phi_q (\mathbf{b}(\hat{\tau}_q^k) \mathbf{c}^H(\hat{f}_{d,q}^k) \odot \mathbf{S}), \quad (31b)$$

and the interference-free observation is then updated as

$$\mathbf{Y}_{\text{free}}^{k+1} = \mathbf{Y} - \mathbf{Y}_{\text{ISI}}^k + \mathbf{Y}_{\text{ICI}}^k. \quad (32)$$

This process repeats until convergence, determined by

$$\frac{\|\mathbf{Y}_{\text{free}}^{k+1}\|_F^2 - \|\mathbf{Y}_{\text{free}}^k\|_F^2}{\|\mathbf{Y}_{\text{free}}^k\|_F^2} < \delta_{\text{th}} \text{ or } k \geq K_{\text{max}}, \quad (33)$$

where  $\delta_{\text{th}}$  and  $K_{\text{max}}$  denote the convergence tolerance and maximum iteration count, respectively.

The SIC-DFT procedure is summarized in Algorithm 1. Each iteration requires a 2D-DFT computation and ISI/ICI reconstruction. With an FFT implementation, the overall complexity is  $\mathcal{O}(N_{\text{iter}} MN(\log N + \log M + Q))$ , where  $N_{\text{iter}}$  is the number of iterations. This cost remains moderate for typical ISAC configurations, making SIC-DFT suitable for real-time implementation.

##### B. SIC-ESPRIT

Although SIC-DFT effectively suppresses interference, its performance is constrained by the coarse resolution and spectral leakage of DFT-based processing. To overcome these limitations, the SIC-ESPRIT algorithm replaces the DFT stage with a super-resolution parameter estimation method based on ESPRIT. A key insight that enables ESPRIT to be applied

despite the ISI/ICI corruption is that, although the interference terms alter the covariance structure, they preserve the fundamental subspace spanned by the target steering vectors. This preservation is formalized as follows:

**Proposition 3.** *The covariance matrix of the sensing channel estimate  $\hat{\mathbf{h}} = \text{vec}(\mathbf{Y} \odot \mathbf{S}^*)$ , accounting for the presence of ISI and ICI, is*

$$\mathbf{R} = \mathbb{E}\{\hat{\mathbf{h}}\hat{\mathbf{h}}^H\} = \mathbf{A}_Q \boldsymbol{\Sigma}_\alpha \mathbf{A}_Q^H + \sigma_{\text{SL}}^2 \mathbf{I}_{MN}, \quad (34)$$

where  $\mathbf{A}_Q = [\mathbf{a}_1, \dots, \mathbf{a}_Q] \in \mathbb{C}^{MN \times Q}$  with  $\mathbf{a}_q = \mathbf{c}^*(f_{d,q}) \otimes \mathbf{b}(\tau_q)$ ,  $\boldsymbol{\Sigma}_\alpha = \text{diag}\{|\tilde{\alpha}_1|^2, \dots, |\tilde{\alpha}_Q|^2\}$ .

*Proof:* See Appendix C.  $\blacksquare$

Proposition 3 reveals that the ISI/ICI manifests itself as a power scaling (through  $\tilde{\alpha}_q$ ) and noise enhancement (through  $\sigma_{\text{SL}}^2$ ), but preserves the column space of  $\mathbf{A}_Q$ . Therefore, subspace methods remain applicable for parameter extraction. Given the preserved subspace structure, we apply ESPRIT to extract high-resolution delay-Doppler estimates. The procedure leverages the shift-invariance properties of the steering vectors, which are briefly outlined below.

1) *Subspace Extraction:* Perform eigenvalue decomposition of the covariance matrix:

$$\mathbf{R} = \mathbf{U}_s \boldsymbol{\Lambda}_s \mathbf{U}_s^H + \sigma_{\text{SL}}^2 \mathbf{U}_n \mathbf{U}_n^H, \quad (35)$$

where  $\mathbf{U}_s \in \mathbb{C}^{MN \times Q}$  spans the signal subspace corresponding to the  $Q$  largest eigenvalues.

2) *Shift-Invariance Exploitation:* Define selection matrices that extract overlapping subarrays:

$$\mathbf{J}_0 = [\mathbf{I}_{M-1}, \mathbf{0}] \otimes [\mathbf{I}_{N-1}, \mathbf{0}], \quad (36a)$$

$$\mathbf{J}_\tau = [\mathbf{I}_{M-1}, \mathbf{0}] \otimes [\mathbf{0}, \mathbf{I}_{N-1}], \quad (36b)$$

$$\mathbf{J}_f = [\mathbf{0}, \mathbf{I}_{M-1}] \otimes [\mathbf{I}_{N-1}, \mathbf{0}]. \quad (36c)$$

These matrices create subspace pairs related by diagonal phase shifts that encode the target parameters.

3) *Parameter Recovery:* Form the rotation matrices:

$$\mathbf{P}_\tau = (\tilde{\mathbf{U}}_s^H \tilde{\mathbf{U}}_s)^{-1} \tilde{\mathbf{U}}_s^H \mathbf{U}_s^\tau, \quad (37a)$$

$$\mathbf{P}_f = (\tilde{\mathbf{U}}_s^H \tilde{\mathbf{U}}_s)^{-1} \tilde{\mathbf{U}}_s^H \mathbf{U}_s^f, \quad (37b)$$

where  $\tilde{\mathbf{U}}_s = \mathbf{J}_0 \mathbf{U}_s$ ,  $\mathbf{U}_s^\tau = \mathbf{J}_\tau \mathbf{U}_s$ , and  $\mathbf{U}_s^f = \mathbf{J}_f \mathbf{U}_s$ . The eigenvalues of  $\mathbf{P}_\tau$  and  $\mathbf{P}_f$  directly yield the parameter estimates:

$$\hat{\tau}_q = -\frac{\angle \lambda_q^\tau}{2\pi \Delta_f}, \quad (38a)$$

$$\hat{f}_{d,q} = \frac{\angle \lambda_q^f}{2\pi T_s}, \quad (38b)$$

where  $\lambda_q^\tau$  and  $\lambda_q^f$  are the  $q$ -th eigenvalues of  $\mathbf{P}_\tau$  and  $\mathbf{P}_f$ , respectively.

Building on the ESPRIT-based delay-Doppler estimation, the proposed SIC-ESPRIT algorithm adopts the same iterative reconstruction-cancellation framework as SIC-DFT. Its main advantage lies in the super-resolution capability of ESPRIT, which enables accurate estimation of closely spaced targets and effectively mitigates spectral leakage, resulting in more precise ISI/ICI reconstruction and improved sensing perfor-

TABLE I: Simulation Parameters

Parameter	Value	Parameter	Value
Carrier frequency $f_c$	28GHz	No. of subcarriers $N$	128
Subcarrier spacing $\Delta_f$	120kHz	No. of symbols $M$	64
Symbol duration $T$	8.33μs	Noise figure $F$	3dB
Standard-CP duration	0.59μs	Ref. temp. $T_{\text{temp}}$	290K
Sufficient-CP duration	8.33μs	Constellation	1024-QAM

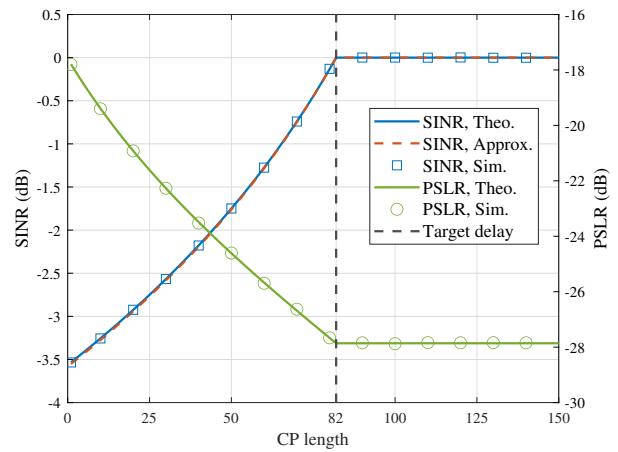


Fig. 3. SINR and PSLR versus CP length, where the target is located at 800m, corresponding to the 82nd delay tap.

mance. In practice, the theoretical covariance matrix  $\mathbf{R}$  is estimated from multiple snapshots obtained via spatial smoothing, which enhances robustness by exploiting the shift-invariance property of the steering vectors [24]. The complete procedure is summarized in Algorithm 2.

From a computational perspective, the dominant cost of SIC-ESPRIT arises from subspace extraction, efficiently implemented using a singular value decomposition of the spatially smoothed data matrix. This step has a complexity of  $\mathcal{O}(MNL^2)$ , where  $L$  is the number of snapshots, leading to an overall complexity of  $\mathcal{O}(N_{\text{iter}} MN(L^2 + Q))$ .

## V. SIMULATION RESULTS

This section presents simulation results to validate the analytical findings from Section III and evaluate the performance of the proposed SIC-DFT and SIC-ESPRIT algorithms developed in Section IV. Unless otherwise noted, all simulations use the common parameter configuration summarized in Table I.

### A. Impact of CP Length on Sensing Performance

We first verify the accuracy of the derived SINR/PSLR expressions and quantify the impact of the CP length on sensing performance. Fig. 3 plots SINR and PSLR versus the CP length for a target at 800m (82nd delay tap) at SNR = 0dB. The theoretical, approximate, and simulated curves closely agree, confirming the validity of our analysis. When the CP is insufficient ( $N_{\text{cp}} < 82$ ), ISI/ICI arise, reducing SINR and increasing the sidelobe level; both degradations grow approximately linearly with the normalized excess delay, consistent with theory. Once  $N_{\text{cp}} \geq 82$ , ISI/ICI vanish and performance converges to the noise-limited bound.

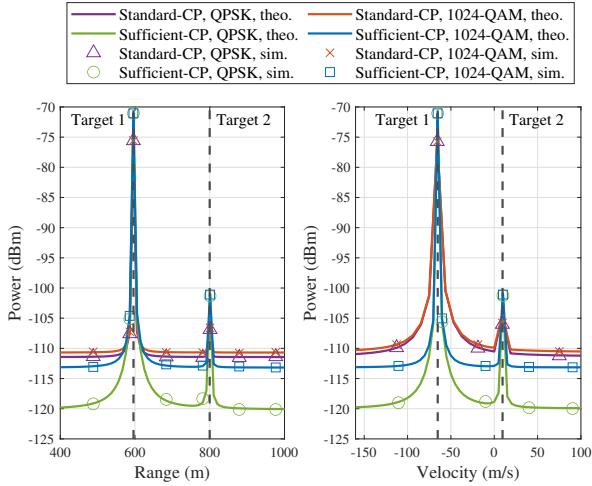


Fig. 4. Range and velocity profiles under different constellations and CP configurations.

Fig. 4 shows representative range/velocity profiles for different CP settings and constellations. The *standard-CP* setting (i.e., normal 3GPP NR CP with  $0.59\mu\text{s}$ ) provides an interference-free range of  $\approx 90\text{m}$ , whereas the *sufficient-CP* (customized long CP with  $8.33\mu\text{s}$ ) baseline extends it to  $\approx 1250\text{m}$ . For targets at  $600\text{m}$  and  $800\text{m}$ , the sufficient-CP incurs no ISI/ICI and thus represents the ideal bound. Theoretical predictions closely match the simulations, and the baseline yields lower sidelobe floors in both range and velocity. Increasing the modulation order from QPSK to 1024-QAM exacerbates sidelobe leakage, resulting in a higher sidelobe floor.

#### B. Performance of SIC-DFT and SIC-ESPRIT Algorithms

We next examine the proposed SIC-DFT and SIC-ESPRIT algorithms when the CP is insufficient (i.e., targets located beyond the standard-CP interference-free range). Fig. 5 illustrates SINR and PSLR performance against the number of algorithm iterations for two representative targets located at ranges of  $600\text{m}$  and  $800\text{m}$ . Both targets are located beyond the standard-CP interference-free range, thus posing significant challenges due to ISI/ICI. As benchmarks, we compare against existing coherent compensation-based methods including TDCC [13], [14], FDCC [23], and MTCC [23], and include the sufficient-CP scenario as an ideal performance bound for reference. Results demonstrate that both SIC-DFT and SIC-ESPRIT rapidly converge within only a few iterations, ultimately achieving SINR gains of more than  $4\text{dB}$  over the benchmark methods. SIC-ESPRIT achieves even better SINR due to its super-resolution delay-Doppler estimates, closely approaching the sufficient-CP bound. The SIC-DFT algorithm substantially reduces PSLR, also reaching the sufficient-CP bound after about 4 iterations. SIC-ESPRIT algorithm is not included in the PSLR comparison since it does not directly generate an RDM.

Fig. 6 compares representative RDMs generated by SIC-DFT under standard-CP conditions compared with conven-

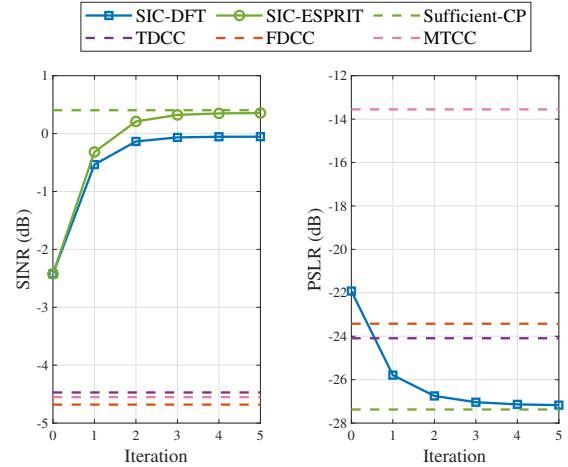


Fig. 5. SINR and PSLR versus number of iterations.

tional DFT-based processing under both standard-CP and sufficient-CP configurations. Here, three targets are present at ranges  $300\text{m}$ ,  $600\text{m}$ , and  $800\text{m}$ . Clearly, even under significant CP insufficiency, SIC-DFT dramatically reduces the sidelobe floor, nearly matching the performance achieved by the sufficient-CP baseline. This indicates that SIC-DFT can effectively reconstruct and cancel ISI/ICI, achieving high-quality target parameter estimation even with a limited CP length.

Finally, Fig. 7 quantifies the range and velocity estimation accuracy by plotting RMSE against sensing SNR. Two targets are uniformly distributed within the range  $(100, 800)\text{m}$  and velocity  $(-150, 150)\text{m/s}$  intervals, which are beyond the standard-CP interference-free zone. SIC-DFT, SIC-ESPRIT, and benchmark schemes (TDCC, FDCC, MTCC, DFT, ESPRIT) are evaluated under standard-CP conditions to assess their robustness. For comparison, we also show conventional DFT and ESPRIT results under the sufficient-CP configuration as ideal RMSE lower bounds. Results highlight that both SIC-DFT and SIC-ESPRIT consistently outperform all benchmarks across the entire SNR range. At moderate SNR (e.g., around  $-10\text{dB}$ ), SIC-ESPRIT achieves approximately one order-of-magnitude lower RMSE than the benchmarks. For range estimation, both SIC-DFT and SIC-ESPRIT nearly achieve the sufficient-CP ideal bound, demonstrating their ability to substantially recover lost sensing performance due to CP insufficiency. In velocity estimation, a visible gap with the bound remains since the sufficient-CP configuration benefits from a much longer observation window, enabling finer Doppler resolution. Overall, these results demonstrate that when the CP is insufficient, the proposed algorithms recover most of the sensing loss induced by ISI/ICI while remaining fully standard-compliant.

## VI. CONCLUSION

This paper has presented a unified analytical and algorithmic framework for OFDM-based ISAC systems operating beyond the CP limit. A general echo model was developed to explicitly characterize the structured ISI/ICI coupling caused

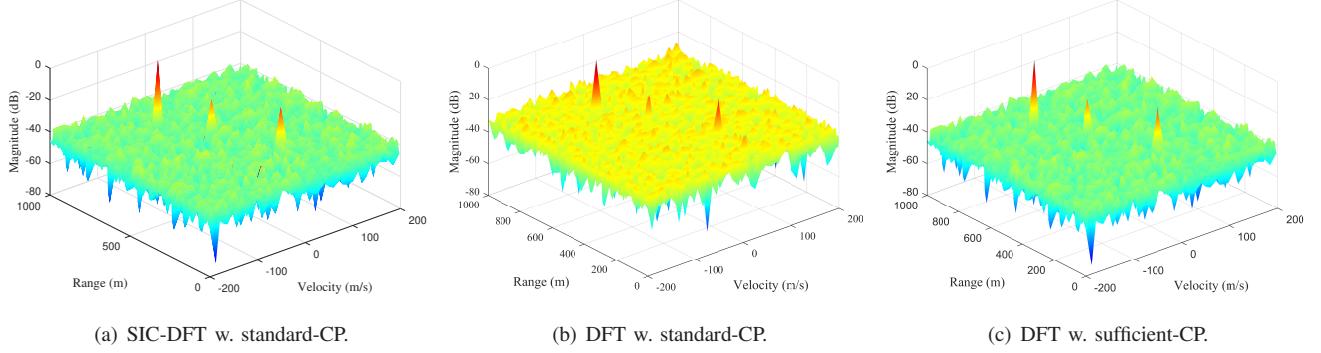


Fig. 6. RDM comparisons.

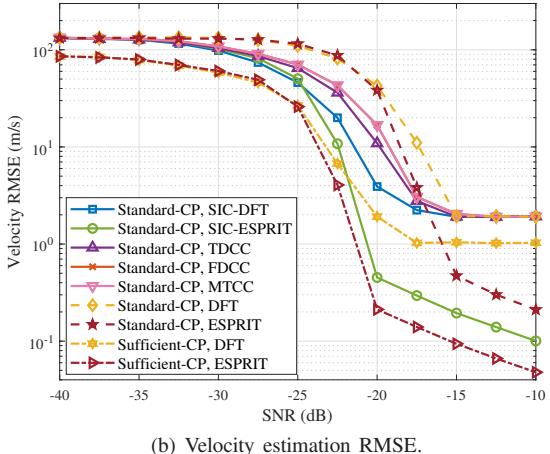
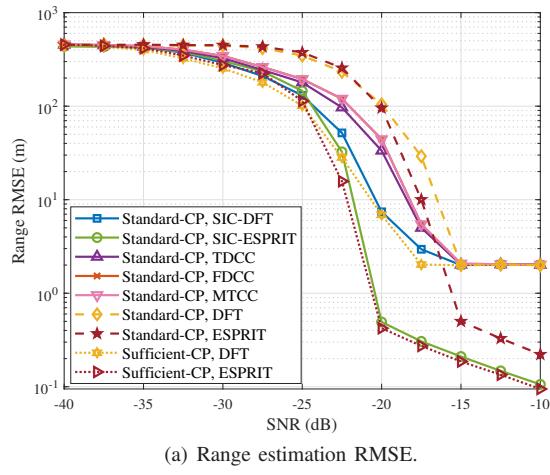


Fig. 7. RMSE for range and velocity estimation versus the sensing SNR.

by insufficient CP, from which closed-form expressions for the sensing SINR and range-Doppler PSLR were derived. The analysis reveals that both SINR degradation and sidelobe elevation increase approximately linearly with the normalized excess delay beyond the CP. To mitigate these effects, two iterative interference cancellation algorithms, SIC-DFT and SIC-ESPRIT, were proposed. Simulation results validate the analytical expressions and show that both proposed algorithms can significantly improve estimation performance compared

with existing benchmarks. These results offer both theoretical insight and practical techniques for achieving reliable, long-range OFDM-ISAC sensing beyond the CP limit.

#### APPENDIX A PROOF OF PROPOSITION 1

Based on the definition of SINR in (18), the proof proceeds by separately evaluating the expected powers of the interference-free component  $\mathbb{E}\{\|\mathbf{Y}_{\text{free}}\|_F^2\}$ , the interference term  $\mathbb{E}\{\|\mathbf{Y}_{\text{ISI}} - \mathbf{Y}_{\text{ICI}}\|_F^2\}$ , and the noise term  $\mathbb{E}\{\|\mathbf{Z}\|_F^2\}$ .

First, the expected power of the interference-free echo signal can be written as

$$\mathbb{E}\{\|\mathbf{Y}_{\text{free}}\|_F^2\} = \mathbb{E}\{\text{Tr}\{\mathbf{Y}_{\text{free}}^H \mathbf{Y}_{\text{free}}\}\} \quad (39a)$$

$$= \mathbb{E}\left\{\sum_{q=1}^Q \sum_{q'=1}^Q \alpha_q \alpha_{q'}^* \text{Tr}\{\mathbf{G}_q \mathbf{G}_{q'}^H\}\right\} \quad (39b)$$

$$= \sum_{q=1}^Q |\alpha_q|^2 \text{Tr}\{\mathbb{E}\{\mathbf{G}_q \mathbf{G}_q^H\}\}, \quad (39c)$$

where we define  $\mathbf{G}_q \triangleq \mathbf{b}(\tau_q) \mathbf{c}^H(f_{d,q}) \odot \mathbf{S}$  for brevity. The cross terms in (39b) vanish due to the statistical independence of the target reflection coefficients. Since the transmitted data symbols  $s_{n,m}$  are i.i.d. with unit power,  $\mathbb{E}\{\mathbf{G}_q \mathbf{G}_q^H\}$  can be derived as

$$\sum_{m=0}^{M-1} |[\mathbf{c}(f_{d,q})]_m|^2 \mathbb{E}\{(\mathbf{b}(\tau_q) \odot \mathbf{s}_m)(\mathbf{b}(\tau_q) \odot \mathbf{s}_m)^H\} \quad (40a)$$

$$= \sum_{m=0}^{M-1} \text{diag}\{\mathbf{b}(\tau_q)\} \mathbb{E}\{\mathbf{s}_m \mathbf{s}_m^H\} \text{diag}\{\mathbf{b}^H(\tau_q)\} \quad (40b)$$

$$= M \mathbf{I}_N. \quad (40c)$$

Thus, the expected power of the interference-free component is given by

$$\mathbb{E}\{\|\mathbf{Y}_{\text{free}}\|_F^2\} = MN \sum_{q=1}^Q |\alpha_q|^2. \quad (41)$$

Next, we evaluate the interference power, which consists of the ISI, ICI, and their cross term:

$$\begin{aligned} & \mathbb{E}\{\|\mathbf{Y}_{\text{ISI}} - \mathbf{Y}_{\text{ICI}}\|_F^2\} \\ &= \mathbb{E}\{\text{Tr}\{(\mathbf{Y}_{\text{ISI}} - \mathbf{Y}_{\text{ICI}})(\mathbf{Y}_{\text{ISI}} - \mathbf{Y}_{\text{ICI}})^H\}\} \end{aligned} \quad (42a)$$

$$= \mathbb{E}\{\|\mathbf{Y}_{\text{ISI}}\|_F^2\} + \mathbb{E}\{\|\mathbf{Y}_{\text{ICI}}\|_F^2\} - 2\Re\{\mathbb{E}\{\text{Tr}\{\mathbf{Y}_{\text{ISI}}^H \mathbf{Y}_{\text{ICI}}\}\}\}. \quad (42\text{b})$$

For the ICI component, the expected power is expressed as

$$\mathbb{E}\{\|\mathbf{Y}_{\text{ICI}}\|_F^2\} = \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \mathbb{E}\{\text{Tr}\{\mathbf{G}_q^H \Phi_q^H \Phi_q \mathbf{G}_q\}\} \quad (43\text{a})$$

$$= \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \text{Tr}\{\Phi_q^H \Phi_q \mathbb{E}\{\mathbf{G}_q \mathbf{G}_q^H\}\}. \quad (43\text{b})$$

Using the derivations in (40c), the key step reduces to evaluating  $\text{Tr}\{\Phi_q \Phi_q^H\}$ , which can be expanded as

$$\text{Tr}\{\Phi_q \Phi_q^H\} = \sum_{n,n'=0}^{N-1} \left| \frac{1}{N} \sum_{i=0}^{l_q - N_{\text{cp}} - 1} e^{j\frac{2\pi}{N}(n'-n)i} \right|^2 \quad (44\text{a})$$

$$= \frac{1}{N^2} \sum_{n,n'=0}^{N-1} \sum_{i,i'=0}^{l_q - N_{\text{cp}} - 1} e^{j\frac{2\pi}{N}(n'-n)(i-i')} \quad (44\text{b})$$

$$= \frac{1}{N} \sum_{i,i'=0}^{l_q - N_{\text{cp}} - 1} \sum_{\Delta=0}^{N-1} e^{j\frac{2\pi}{N}\Delta(i-i')} \quad (44\text{c})$$

$$= \frac{1}{N} \sum_{i,i'=0}^{l_q - N_{\text{cp}} - 1} N\delta_{i,i'} \quad (44\text{d})$$

$$= l_q - N_{\text{cp}}, \quad (44\text{e})$$

where  $\delta_{n,n'}$  denotes the Kronecker delta function. Substituting (40c) and (44e) into (43b) yields

$$\mathbb{E}\{\|\mathbf{Y}_{\text{ICI}}\|_F^2\} = M \sum_{q=\tilde{Q}+1}^Q (l_q - N_{\text{cp}}) |\alpha_q|^2. \quad (45)$$

Similarly, the expectation of the ISI power is written as

$$\mathbb{E}\{\|\mathbf{Y}_{\text{ISI}}\|_F^2\} = \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \mathbb{E}\{\text{Tr}\{\mathbf{J}_1^H \tilde{\mathbf{G}}_q^H \Phi_q^H \Phi_q \tilde{\mathbf{G}}_q \mathbf{J}_1\}\} \quad (46\text{a})$$

$$= \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \text{Tr}\{\Phi_q^H \Phi_q \mathbb{E}\{\tilde{\mathbf{G}}_q \mathbf{J}_1 \mathbf{J}_1^H \tilde{\mathbf{G}}_q^H\}\}, \quad (46\text{b})$$

where  $\tilde{\mathbf{G}}_q \triangleq \mathbf{b}(\tau_q - T_{\text{cp}}) \mathbf{c}^H(f_{d,q}) \odot \mathbf{S}$ . Since  $\mathbf{J}_1 \mathbf{J}_1^H = \text{diag}\{1, \dots, 1, 0\} \in \mathbb{C}^{M \times M}$ , the expectation of  $\tilde{\mathbf{G}}_q \mathbf{J}_1 \mathbf{J}_1^H \tilde{\mathbf{G}}_q^H$  can be expanded as

$$\begin{aligned} & \sum_{m=0}^{M-2} |[\mathbf{c}(f_{d,q})]_m|^2 \mathbb{E}\{(\mathbf{b}(\tau_q - T_{\text{cp}}) \odot \mathbf{s}_m)(\mathbf{b}(\tau_q - T_{\text{cp}}) \odot \mathbf{s}_m)^H\} \\ &= (M-1)\mathbf{I}_N, \end{aligned} \quad (47)$$

which implies

$$\mathbb{E}\{\|\mathbf{Y}_{\text{ISI}}\|_F^2\} = (M-1) \sum_{q=\tilde{Q}+1}^Q (l_q - N_{\text{cp}}) |\alpha_q|^2. \quad (48)$$

Then, we derive the cross-correlation term between the ISI

and ICI components:

$$\mathbb{E}\{\text{Tr}\{\mathbf{Y}_{\text{ISI}}^H \mathbf{Y}_{\text{ICI}}\}\} = \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \text{Tr}\{\Phi_q^H \Phi_q \mathbb{E}\{\mathbf{G}_q \mathbf{J}_1^H \tilde{\mathbf{G}}_q^H\}\}. \quad (49)$$

Note that  $\mathbb{E}\{\mathbf{G}_q \mathbf{J}_1^H \tilde{\mathbf{G}}_q^H\}$  can be derived as

$$\begin{aligned} & \sum_{m,m'} \mathbb{E}\{(\mathbf{b}(\tau_q) \odot \mathbf{s}_m)(\mathbf{b}(\tau_q - T_{\text{cp}}) \odot \mathbf{s}_{m'})^H\} \\ & \times e^{j2\pi(m-m')f_{d,q}T_s} [\mathbf{J}_1^H]_{m,m'} \end{aligned} \quad (50\text{a})$$

$$= \sum_{m,m'} \delta_{m,m'} \text{diag}\{\mathbf{b}(T_{\text{cp}})\} e^{j2\pi(m-m')f_{d,q}T_s} [\mathbf{J}_1^H]_{m,m'} \quad (50\text{b})$$

$$= \sum_{m=0}^{M-1} [\mathbf{J}_1^H]_{m,m} \text{diag}\{\mathbf{b}(T_{\text{cp}})\}. \quad (50\text{c})$$

Since all diagonal entries of  $\mathbf{J}_1^H$  are zero, we obtain

$$\mathbb{E}\{\text{Tr}\{\mathbf{Y}_{\text{ISI}}^H \mathbf{Y}_{\text{ICI}}\}\} = 0, \quad (51)$$

which indicates that the ISI and ICI components are statistically uncorrelated, allowing their powers to be added directly to form the total interference power.

In addition, the noise power is given by

$$\mathbb{E}\{\|\mathbf{Z}\|_F^2\} = MN\sigma^2. \quad (52)$$

Combining the above results, the SINR of the echo signal can be written as

$$\text{SINR} = \frac{MN \sum_{q=1}^Q |\alpha_q|^2}{(2M-1)N \sum_{q=\tilde{Q}+1}^Q \rho_q |\alpha_q|^2 + MN\sigma^2}. \quad (53)$$

## APPENDIX B PROOF OF PROPOSITION 2

Since the sensing observation contains the interference-free signal, ISI, ICI, and noise, the RDM can be recast as

$$\chi = \chi_{\text{free}} + \chi_{\text{ISI}} - \chi_{\text{ICI}} + \chi_{\text{awgn}}, \quad (54)$$

detailed expressions for which are given below:

$$\chi_{\text{free}}(l, \nu) = \sum_{q=1}^Q \frac{\alpha_q}{\sqrt{MN}} \sum_{m,n} |s_{n,m}|^2 e^{j\theta_{n,m,q}}, \quad (55\text{a})$$

$$\chi_{\text{ISI}}(l, \nu) = \sum_{q=\tilde{Q}+1}^Q \frac{\alpha_q}{\sqrt{MN}} \sum_{m,n,n'} s_{n',m-1} s_{n,m}^* \psi_{n,n',m}^{\text{ISI}}, \quad (55\text{b})$$

$$\chi_{\text{ICI}}(l, \nu) = \sum_{q=\tilde{Q}+1}^Q \frac{\alpha_q}{\sqrt{MN}} \sum_{m,n,n'} s_{n',m} s_{n,m}^* \psi_{n,n',m}^{\text{ICI}}, \quad (55\text{c})$$

$$\chi_{\text{awgn}}(l, \nu) = \frac{1}{\sqrt{MN}} \sum_{m,n} z_{n,m} s_{n,m}^* e^{j\frac{2\pi}{N}nl} e^{-j\frac{2\pi}{M}m\nu}, \quad (55\text{d})$$

$$\theta_{n,m,q} = \frac{2\pi}{N} n(l - \tilde{l}_q) + \frac{2\pi}{M} m(\tilde{\nu}_q - \nu), \quad (55\text{e})$$

$$\psi_{n,n',m}^{\text{ISI}} = \phi_{n,n'}^q e^{j\frac{2\pi}{N}(nl-n'(\tilde{l}_q - N_{\text{cp}}))} e^{j\frac{2\pi}{M}((m-1)\tilde{\nu}_q - m\nu)}, \quad (55\text{f})$$

$$\psi_{n,n',m}^{\text{ICI}} = \phi_{n,n'}^q e^{j\frac{2\pi}{N}(nl-n'\tilde{l}_q)} e^{j\frac{2\pi}{M}m(\tilde{\nu}_q - \nu)}. \quad (55\text{g})$$

Thus, the second-order moment of the RDM is written as

$$\begin{aligned} \mathbb{E}\{|\chi(l, \nu)|^2\} &= \mathbb{E}\{|\chi_{\text{free}}(l, \nu)|^2\} + \mathbb{E}\{|\chi_{\text{ISI}}(l, \nu)|^2\} \\ &+ \mathbb{E}\{|\chi_{\text{ICI}}(l, \nu)|^2\} + \mathbb{E}\{|\chi_{\text{awgn}}(l, \nu)|^2\} + 2\Re\left\{\mathbb{E}\{\chi_{\text{sig}}(l, \nu) (\chi_{\text{awgn}}(l, \nu))^*\}\right\} + \mathbb{E}\{\chi_{\text{free}}(l, \nu)(\chi_{\text{ISI}}(l, \nu))^*\} - \mathbb{E}\{\chi_{\text{free}}(l, \nu)(\chi_{\text{ICI}}(l, \nu))^*\} - \mathbb{E}\{\chi_{\text{ISI}}(l, \nu)(\chi_{\text{ICI}}(l, \nu))^*\}, \end{aligned} \quad (56)$$

where  $\chi_{\text{sig}} = \chi_{\text{free}} + \chi_{\text{ISI}} - \chi_{\text{ICI}}$ , and the last term collects the cross-correlations between the interference-free signal, ISI, ICI, and noise components.

In the subsequent derivations, we will repeatedly encounter expectation terms of the form  $\mathbb{E}\{s_A s_B^* s_C^* s_D\}$ , where  $A, B, C, D$  denote time-frequency indices (e.g.,  $A = (n', m)$ ,  $B = (n, m)$ , etc.). Under the assumptions used throughout the paper, where the symbols are i.i.d. across time/frequency,  $\mathbb{E}\{s_{n,m}\} = \mathbb{E}\{s_{n,m}^2\} = 0$ , and  $\mathbb{E}\{|s_{n,m}|^2\} = 1$ , we have

$$\mathbb{E}\{s_A s_B^* s_C^* s_D\} = \begin{cases} 1, & A = B \neq C = D; \\ 1, & A = C \neq B = D; \\ \mu_4, & A = B = C = D; \\ 0, & \text{otherwise.} \end{cases} \quad (57)$$

We now analyze each term in (56). First, the second-order moment of the interference-free component is

$$\begin{aligned} \mathbb{E}\{|\chi_{\text{free}}(l, \nu)|^2\} &= \sum_{q=1}^Q \frac{|\alpha_q|^2}{MN} \sum_{m,n} \sum_{m',n'} \mathbb{E}\{|s_{n,m}|^2 |s_{n',m'}|^2\} e^{j(\theta_{n,m,q} - \theta_{n',m',q})} \end{aligned} \quad (58a)$$

$$= \sum_{q=1}^Q \frac{|\alpha_q|^2}{MN} \sum_{\substack{m,n \\ (m',n') \neq (m,n)}} e^{j(\theta_{n,m,q} - \theta_{n',m',q})} + \mu_4 \sum_{q=1}^Q |\alpha_q|^2 \quad (58b)$$

$$= \sum_{q=1}^Q \frac{|\alpha_q|^2}{MN} \left( \left| \sum_{m,n} e^{j\theta_{n,m,q}} \right|^2 - MN \right) + \mu_4 \sum_{q=1}^Q |\alpha_q|^2 \quad (58c)$$

$$= \sum_{q=1}^Q \frac{|\alpha_q|^2}{MN} |D_N(l - \tilde{l}_q)|^2 |D_M(\nu - \tilde{\nu}_q)|^2 + (\mu_4 - 1) \sum_{q=1}^Q |\alpha_q|^2. \quad (58d)$$

Next, the second-order moment of the ISI contribution can be expressed as

$$\begin{aligned} \mathbb{E}\{|\chi_{\text{ISI}}(l, \nu)|^2\} &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{\substack{m_1,n_1,n'_1 \\ m_2,n_2,n'_2}} \mathbb{E}\{s_{n'_1,m_1-1} s_{n_1,m_1}^* s_{n'_2,m_2-1}^* s_{n_2,m_2}\} \\ &\times \psi_{n_1,n'_1,m_1}^{\text{ISI}} (\psi_{n_2,n'_2,m_2}^{\text{ISI}})^* \end{aligned} \quad (59a)$$

$$= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} \sum_{n'=0}^{N-1} |\psi_{n,n',m}^{\text{ISI}}|^2 \quad (59b)$$

$$= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{N} \sum_{n=0}^{N-1} \sum_{n'=0}^{N-1} |\phi_{n,n'}^q|^2 \quad (59c)$$

$$= \sum_{q=\tilde{Q}+1}^Q \rho_q |\alpha_q|^2, \quad (59d)$$

where (59c) follows from (44). Similarly, the second-order moment of the ICI contribution is given by

$$\begin{aligned} \mathbb{E}\{|\chi_{\text{ICI}}(l, \nu)|^2\} &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{\substack{m_1,n_1,n'_1 \\ m_2,n_2,n'_2}} \mathbb{E}\{s_{n'_1,m_1} s_{n_1,m_1}^* s_{n'_2,m_2}^* s_{n_2,m_2}\} \\ &\times \psi_{n_1,n'_1,m_1}^{\text{ICI}} (\psi_{n_2,n'_2,m_2}^{\text{ICI}})^* \end{aligned} \quad (60a)$$

$$\begin{aligned} &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{m_1,n_1} \sum_{(m_2,n_2) \neq (m_1,n_1)} \rho_q^2 e^{j(\theta_{n_1,m_1,q} - \theta_{n_2,m_2,q})} \\ &+ \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{m,n,n' \neq n} |\psi_{n,n',m}^{\text{ICI}}|^2 + \mu_4 \sum_{m,n} |\rho_q e^{j\theta_{n,m,q}}|^2 \end{aligned} \quad (60b)$$

$$= \sum_{q=\tilde{Q}+1}^Q \frac{\rho_q^2 |\alpha_q|^2}{MN} \left| \sum_{m,n} e^{j\theta_{n,m,q}} \right|^2 + (\rho_q + (\mu_4 - 2)\rho_q^2) |\alpha_q|^2 \quad (60c)$$

$$\begin{aligned} &= \sum_{q=\tilde{Q}+1}^Q \frac{\rho_q^2 |\alpha_q|^2}{MN} |D_N(l - \tilde{l}_q)|^2 |D_M(\nu - \tilde{\nu}_q)|^2 \\ &+ \sum_{q=\tilde{Q}+1}^Q ((\mu_4 - 1)\rho_q^2 + \rho_q(1 - \rho_q)) |\alpha_q|^2. \end{aligned} \quad (60d)$$

Finally, the noise contribution is

$$\begin{aligned} \mathbb{E}\{|\chi_{\text{awgn}}(l, \nu)|^2\} &= \frac{1}{MN} \sum_{m,n} \sum_{m',n'} \mathbb{E}\{z_{n,m} z_{n',m'}^* s_{n,m}^* s_{n',m'}\} \\ &\times e^{j\frac{2\pi}{N}(n-n')} l e^{j\frac{2\pi}{M}(m'-m)\nu} \end{aligned} \quad (61a)$$

$$= \frac{\sigma^2}{MN} \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} \mathbb{E}\{|s_{n,m}|^2\} \quad (61b)$$

$$= \sigma^2. \quad (61c)$$

Having derived the auto-correlation terms, we proceed to examine the cross-correlation terms between different components in (56). Specifically, the cross-correlation between  $\chi_{\text{sig}}(l, \nu)$  and  $\chi_{\text{awgn}}(l, \nu)$  can be expressed as

$$\begin{aligned} \mathbb{E}\{\chi_{\text{sig}}(l, \nu)(\chi_{\text{awgn}}(l, \nu))^*\} &= \frac{1}{MN} \mathbb{E}\left\{\chi_{\text{sig}}(l, \nu) \sum_{m,n} z_{n,m}^* s_{n,m} e^{-j\frac{2\pi}{N}nl} e^{j\frac{2\pi}{M}m\nu}\right\} \end{aligned} \quad (62a)$$

$$= \frac{1}{MN} \mathbb{E}_{\alpha,s} \left\{ \chi_{\text{sig}}(l, \nu) \sum_{m,n} \mathbb{E}_z \{z_{n,m}^*\} s_{n,m} e^{-j\frac{2\pi}{N}nl} e^{j\frac{2\pi}{M}m\nu} \right\} \quad (62b)$$

$$= 0. \quad (62c)$$

The cross-correlation between the interference-free and ISI components is

$$\begin{aligned} \mathbb{E}\{\chi_{\text{free}}(l, \nu)(\chi_{\text{ISI}}(l, \nu))^*\} &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{\substack{m_1,n_1 \\ m_2,n_2,n'_2}} \mathbb{E}\{|s_{n_1,m_1}|^2 s_{n'_2,m_2-1}^* s_{n_2,m_2}\} \end{aligned}$$

$$\times e^{j\theta_{n_1, m_1, q}} (\psi_{n_2, n'_2, m_2}^{\text{ISI}})^* \quad (63a)$$

$$= 0. \quad (63b)$$

Similarly, the cross-correlation between  $\chi_{\text{free}}(l, \nu)$  and  $\chi_{\text{ICI}}(l, \nu)$  is

$$\begin{aligned} & \mathbb{E}\{\chi_{\text{free}}(l, \nu)(\chi_{\text{ICI}}(l, \nu))^*\} \\ &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{\substack{m_1, n_1, \\ m_2, n_2, n'_2}} \mathbb{E}\{|s_{n_1, m_1}|^2 s_{n'_2, m_2}^* s_{n_2, m_2}\} \end{aligned} \quad (64a)$$

$$\times e^{j\theta_{n_1, m_1, q}} (\psi_{n_2, n'_2, m_2}^{\text{ICI}})^* \quad (64b)$$

$$\begin{aligned} &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{m_1, n_1} \sum_{(m_2, n_2) \neq (m_1, n_1)} \rho_q e^{j(\theta_{n_1, m_1, q} - \theta_{n_2, m_2, q})} \\ &\quad + \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{m, n} \mu_4 \rho_q \end{aligned} \quad (64c)$$

$$\begin{aligned} &= \sum_{q=\tilde{Q}+1}^Q \frac{\rho_q |\alpha_q|^2}{MN} |D_N(l - \tilde{l}_q)|^2 |D_M(\nu - \tilde{\nu}_q)|^2 + (\mu_4 - 1) \rho_q |\alpha_q|^2. \end{aligned} \quad (64d)$$

Finally, the cross-correlation between  $\chi_{\text{ISI}}(l, \nu)$  and  $\chi_{\text{ICI}}(l, \nu)$  can be written as

$$\begin{aligned} & \mathbb{E}\{\chi_{\text{ISI}}(l, \nu)(\chi_{\text{ICI}}(l, \nu))^*\} \\ &= \sum_{q=\tilde{Q}+1}^Q \frac{|\alpha_q|^2}{MN} \sum_{\substack{m_1, n_1, n'_1 \\ m_2, n_2, n'_2}} \mathbb{E}\{s_{n'_1, m_1-1} s_{n_1, m_1}^* s_{n'_2, m_2}^* s_{n_2, m_2}\} \\ &\quad \times \psi_{n_1, n'_1, m_1}^{\text{ISI}} \psi_{n_2, n'_2, m_2}^{\text{ICI}} \end{aligned} \quad (65a)$$

$$= 0. \quad (65b)$$

Based on the above derivations, the second-order moment of the RDM is given by

$$\mathbb{E}\{|\chi(l, \nu)|^2\} = \sum_{q=1}^Q \frac{|\tilde{\alpha}_q|^2}{MN} |D_N(l - \tilde{l}_q)|^2 |D_M(\nu - \tilde{\nu}_q)|^2 + \sigma_{\text{SL}}^2. \quad (66)$$

## APPENDIX C PROOF OF PROPOSITION 3

To derive the covariance matrix of  $\hat{\mathbf{h}}$ , we first decompose the  $i$ -th entry of  $\hat{\mathbf{h}}$  as

$$\hat{h}_i = \hat{h}_i^{\text{free}} + \hat{h}_i^{\text{ISI}} - \hat{h}_i^{\text{ICI}} + \hat{h}_i^{\text{awgn}}, \quad (67)$$

where  $\hat{h}_i^{\text{free}} = y_{n, m}^{\text{free}} s_{n, m}^*$ ,  $\hat{h}_i^{\text{ISI}} = y_{n, m}^{\text{ISI}} s_{n, m}^*$ ,  $\hat{h}_i^{\text{ICI}} = y_{n, m}^{\text{ICI}} s_{n, m}^*$ ,  $\hat{h}_i^{\text{awgn}} = z_{n, m} s_{n, m}^*$ , and  $i = n + mN$ . Accordingly, the  $(i, j)$ -th entry of the covariance matrix  $\mathbf{R}$  can be written as

$$\begin{aligned} [\mathbf{R}]_{i,j} &= \mathbb{E}\{\hat{h}_i^{\text{free}}(\hat{h}_j^{\text{free}})^*\} + \mathbb{E}\{\hat{h}_i^{\text{ISI}}(\hat{h}_j^{\text{ISI}})^*\} + \mathbb{E}\{\hat{h}_i^{\text{ICI}}(\hat{h}_j^{\text{ICI}})^*\} \\ &\quad + \mathbb{E}\{\hat{h}_i^{\text{awgn}}(\hat{h}_j^{\text{awgn}})^*\} + 2\Re\{\mathbb{E}\{\hat{h}_i^{\text{free}}(\hat{h}_j^{\text{ISI}})^*\} - \mathbb{E}\{\hat{h}_i^{\text{free}}(\hat{h}_j^{\text{ICI}})^*\}\} \\ &\quad - \mathbb{E}\{\hat{h}_i^{\text{ISI}}(\hat{h}_j^{\text{ICI}})^*\} + \mathbb{E}\{(\hat{h}_i^{\text{free}} + \hat{h}_i^{\text{ISI}} - \hat{h}_i^{\text{ICI}})(\hat{h}_j^{\text{awgn}})^*\}. \end{aligned} \quad (68)$$

We now evaluate each term in (68). Based on the signal model in (11), the covariance of the interference-free compo-

nent  $\hat{h}_i^{\text{free}}$  is

$$\begin{aligned} \mathbb{E}\{\hat{h}_i^{\text{free}}(\hat{h}_j^{\text{free}})^*\} &= \sum_{q=1}^Q |\alpha_q|^2 \mathbb{E}\{|s_{n, m}|^2 |s_{n', m'}|^2\} [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j \\ &= \begin{cases} \mu_4 \sum_{q=1}^Q |\alpha_q|^2, & i = j; \\ \sum_{q=1}^Q |\alpha_q|^2 [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j, & i \neq j. \end{cases} \end{aligned} \quad (69a)$$

$$= \begin{cases} \mu_4 \sum_{q=1}^Q |\alpha_q|^2, & i = j; \\ \sum_{q=1}^Q |\alpha_q|^2 [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j, & i \neq j. \end{cases} \quad (69b)$$

The covariance of the ISI component  $\hat{h}_i^{\text{ISI}}$  can be obtained by

$$\begin{aligned} & \mathbb{E}\{\hat{h}_i^{\text{ISI}}(\hat{h}_j^{\text{ISI}})^*\} \\ &= \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \sum_{n_1, n_2} \mathbb{E}\{s_{n_1, m-1} s_{n, m}^* s_{n_2, m'-1} s_{n', m'}\} \phi_{n, n_1}^q \\ &\quad \times (\phi_{n', n_2}^q)^* e^{j2\pi(n_1 - n_2)\Delta_f(T_{\text{cp}} - \tau_q)} e^{j2\pi(m - m')f_{\text{d}, q}T_s} \end{aligned} \quad (70a)$$

$$= \begin{cases} \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \sum_{n'=0}^{N-1} |\phi_{n, n'}^q|^2, & i = j; \\ 0, & i \neq j; \end{cases} \quad (70b)$$

$$= \begin{cases} \sum_{q=\tilde{Q}+1}^Q \rho_q |\alpha_q|^2, & i = j; \\ 0, & i \neq j. \end{cases} \quad (70c)$$

Similarly, the covariance of the ICI component  $\hat{h}_i^{\text{ICI}}$  is

$$\mathbb{E}\{\hat{h}_i^{\text{ICI}}(\hat{h}_j^{\text{ICI}})^*\} \quad (71a)$$

$$\begin{aligned} &= \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \sum_{n_1, n_2} \mathbb{E}\{s_{n_1, m} s_{n, m}^* s_{n_2, m'}^* s_{n', m'}\} \phi_{n, n_1}^q (\phi_{n', n_2}^q)^* \\ &\quad \times e^{-j2\pi(n_1 - n_2)\Delta_f \tau_q} e^{j2\pi(m - m')f_{\text{d}, q}T_s} \end{aligned} \quad (71b)$$

$$= \begin{cases} \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 (\sum_{n_1 \neq n} |\phi_{n, n_1}^q|^2 + \mu_4 \rho_q^2), & i = j; \\ \sum_{q=\tilde{Q}+1}^Q \rho_q^2 |\alpha_q|^2 [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j, & i \neq j. \end{cases} \quad (71c)$$

$$= \begin{cases} \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 ((\mu_4 - 1)\rho_q^2 + \rho_q), & i = j; \\ \sum_{q=\tilde{Q}+1}^Q \rho_q^2 |\alpha_q|^2 [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j, & i \neq j. \end{cases} \quad (71d)$$

The covariance of the noise component  $\hat{h}_i^{\text{awgn}}$  is given by

$$\mathbb{E}\{\hat{h}_i^{\text{awgn}}(\hat{h}_j^{\text{awgn}})^*\} = \mathbb{E}\{z_{n, m} s_{n, m}^* z_{n', m'}^* s_{n', m'}\} \quad (72a)$$

$$= \delta_{i,j} \sigma^2. \quad (72b)$$

Next, we derive the cross covariances between different components. The cross covariance between  $\hat{h}_i^{\text{free}}$  and  $\hat{h}_j^{\text{ISI}}$  is

$$\begin{aligned} \mathbb{E}\{\hat{h}_i^{\text{free}}(\hat{h}_j^{\text{ISI}})^*\} &= \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \sum_{n_1=0}^{N-1} \mathbb{E}\{|s_{n, m}|^2 s_{n_1, m'-1}^* s_{n', m'}\} \\ &\quad \times (\phi_{n', n_1}^q)^* e^{j2\pi\Delta_f(n_1(\tau_q - T_{\text{cp}}) - n\tau_q)} e^{j2\pi(m - m'+1)f_{\text{d}, q}T_s} \\ &= 0. \end{aligned} \quad (73a)$$

$$(73b)$$

The cross covariance between  $\hat{h}_i^{\text{free}}$  and  $\hat{h}_j^{\text{ICI}}$  is given by

$$\begin{aligned} & \mathbb{E}\{\hat{h}_i^{\text{free}}(\hat{h}_j^{\text{ICI}})^*\} \\ &= \sum_{q=\tilde{Q}+1}^Q |\alpha_q|^2 \sum_{n_1=0}^{N-1} \mathbb{E}\{|s_{n, m}|^2 s_{n_1, m'}^* s_{n', m'}\} (\phi_{n', n_1}^q)^* \\ &\quad \times e^{-j2\pi(n_1 - n)\Delta_f \tau_q} e^{j2\pi(m - m')f_{\text{d}, q}T_s} \end{aligned} \quad (74a)$$

$$= \begin{cases} \sum_{q=\bar{Q}+1}^Q \mu_4 \rho_q |\alpha_q|^2, & i = j; \\ \sum_{q=\bar{Q}+1}^Q \rho_q |\alpha_q|^2 [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j, & i \neq j. \end{cases} \quad (74b)$$

The cross covariance between  $\hat{h}_i^{\text{ISI}}$  and  $\hat{h}_j^{\text{ICI}}$  is

$$\mathbb{E}\{\hat{h}_i^{\text{ISI}}(\hat{h}_j^{\text{ICI}})^*\} = \sum_{q=\bar{Q}+1}^Q |\alpha_q|^2 \sum_{n_1, n_2} \mathbb{E}\{s_{n_1, m-1} s_{n, m}^* s_{n_2, m'}^* s_{n', m'}\} \\ \times \phi_{n, n_1}^q (\phi_{n', n_2}^q)^* e^{j2\pi\Delta_f(n_2\tau_q - n_1(\tau_q - T_{\text{cp}}))} e^{j2\pi(m-m'-1)f_{d,q}T_s} \quad (75a)$$

$$= 0. \quad (75b)$$

Finally, the cross covariance between the signal and noise components can be written as

$$\mathbb{E}\{(\hat{h}_i^{\text{free}} + \hat{h}_i^{\text{ISI}} - \hat{h}_i^{\text{ICI}})(\hat{h}_j^{\text{awgn}})^*\} = \mathbb{E}\{(\hat{h}_i^{\text{free}} + \hat{h}_i^{\text{ISI}} - \hat{h}_i^{\text{ICI}})s_{n', m'}\} \mathbb{E}\{z_{n', m'}^*\} \quad (76a)$$

$$= 0. \quad (76b)$$

Based on the above results, the  $(i, j)$ -th entry of the covariance matrix  $\mathbf{R}$  can be written as

$$[\mathbf{R}]_{i,j} = \begin{cases} \sum_{q=1}^Q |\tilde{\alpha}_q|^2 + \sigma_{\text{SL}}^2, & i = j; \\ \sum_{q=1}^Q |\tilde{\alpha}_q|^2 [\mathbf{a}_q]_i [\mathbf{a}_q^*]_j, & i \neq j. \end{cases} \quad (77a)$$

Therefore, the covariance matrix can be expressed in compact form as

$$\mathbf{R} = \sum_{q=1}^Q |\tilde{\alpha}_q|^2 \mathbf{a}_q \mathbf{a}_q^H + \sigma_{\text{SL}}^2 \mathbf{I}_{NM} \quad (78)$$

$$= \mathbf{A}_Q \boldsymbol{\Sigma}_\alpha \mathbf{A}_Q^H + \sigma_{\text{SL}}^2 \mathbf{I}_{NM}.$$

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