

A Non-Orthogonal ISAC Architecture based on OCDM-IM and Superimposed Training with Iterative SIC Receiver

George Lucas Lima Takemiya, José Miracy de Souza F., Herman Lucas dos Santos and José Carlos Marinello F.

Abstract—This paper proposes a non-orthogonal approach for integrated sensing and communication by superimposing a Zadoff-Chu (ZC) sequence onto an orthogonal chirp division multiplexing with index modulation signal. The ZC sequence acts simultaneously as a radar signal and a pilot for communication channel estimation (CHEST). To mitigate mutual interference, we design an iterative receiver that uses soft-decision successive interference cancellation and a linear minimum mean square error (MMSE) estimator. Numerical results show that this significantly reduces the CHEST MSE while efficiently maintaining data rates. The system achieves a peak-to-sidelobe ratio of 24.3 dB at a 20 dB signal-to-noise ratio, with centimeter-level range accuracy, confirming its effectiveness for future sixth-generation applications.

Keywords—ISAC, OCDM-IM, Superimposed Training, Channel Estimation, Radar Sensing.

I. INTRODUCTION

The sixth generation (6G) of wireless networks is expected to support intelligent and immersive applications, such as autonomous driving and industrial automation, which require extreme reliability and massive connectivity [1]. Given the strain on finite frequency resources caused by growing wireless traffic [2], network architectures must evolve to utilize bandwidth more efficiently while interpreting the physical environment.

In this context, integrated sensing and communication (ISAC) has been widely studied as one of the future technologies for 6G wireless networks [3]. By supporting sensing and communication on a single hardware platform and sharing a common waveform, ISAC achieves advantages such as reduced hardware implementation costs and efficient utilization of spectrum resources. This approach addresses spectrum congestion and supports new applications like autonomous vehicles and smart cities [4].

However, deploying ISAC often requires dedicating specific sensing resources, such as positioning reference signals, which reduces the achievable data rate. To eliminate this overhead, superimposed training (ST) has emerged as a promising technique, where the sensing signal, such as a Zadoff-Chu (ZC) sequence, is arithmetically added to the

communication waveform, fully sharing all available resources [5]. While ST enhances spectral efficiency, it introduces severe mutual interference between the sensing and communication subsystems. To address this issue, successive interference cancellation (SIC) schemes have been explored to estimate and progressively remove the radar interference from the received superimposed signal [6].

To improve the effectiveness of ISAC systems, orthogonal chirp division multiplexing with index modulation (OCDM-IM) has emerged as a promising waveform. By multiplexing linear chirps that are inherently resistant to Doppler shifts, OCDM effectively exploits multipath diversity in fading channels [7], [8]. Furthermore, by conveying information through both the activated subchirp indices and the transmitted data symbols, the OCDM-IM scheme significantly improves spectral efficiency, bit error rate (BER), and reduces peak-to-average power ratio (PAPR). Hence, OCDM-IM has been recently proven as a highly efficient dual-function waveform, since it enhances communication reliability while preserving accurate sensing performance for target estimation [9].

Contributions: shifting away from orthogonal resource allocation, this work proposes a novel non-orthogonal ISAC architecture that superimposes a low-power ZC sensing sequence directly onto an OCDM-IM communication signal in the Fresnel domain. The main contributions of this paper are: (i) we propose an ST-based ISAC system where the superimposed ZC sequence acts both as a signal for radar target detection and as a pilot sequence for channel estimation (CHEST) in the communication system, ensuring high spectral efficiency; (ii) to tackle the severe self-interference at the communication receiver, we design an iterative receiver for the OCDM domain, employing a soft-decision SIC loop assisted with an adaptive linear MMSE (LMMSE) channel estimator to progressively filter the channel distorted information from the pilot sequence; and (iii) to overcome the data noise from the radar signal, we employ coherent pulse integration (CPI), incorporating an adaptive MMSE pulse compression strategy [10] and Chebyshev windowing [11] to control residual interference and achieve deep sidelobe suppression.

II. SYSTEM MODEL AND CHANNEL FORMULATION

We consider a monostatic ISAC system operating in a highly dispersive environment modeled by a frequency-selective Rician fading channel. The channel impulse response (CIR) consists of a Line-of-Sight (LoS) path, followed by multiple

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Non-LoS (NLoS) echoes. The channel is characterized by a Rician factor K , delay vector τ , and a given power delay profile (PDP).

The OCDM-IM system consists of N orthogonal subcarriers, which are partitioned into g groups. Each group contains $n_c = N/g$ subcarriers, out of which only k are activated for data transmission.

For a generic transmitted signal block $\mathbf{x}_{Fn}$ in the Fresnel domain, the corresponding time-domain signal is generated via inverse discrete Fresnel transform (IDFnT) and appended with a cyclic prefix (CP). At the receiver, after CP removal and fast Fourier Transform (FFT) processing, the discrete frequency-domain signal is modeled as:

$$\mathbf{Y}_{\text{freq}} = \mathbf{H}_{\text{freq}} \mathbf{X}_{\text{freq}} + \boldsymbol{\eta}_{\text{freq}}, \quad (1)$$

where $\mathbf{H}_{\text{freq}}$ is the diagonal channel matrix in the frequency domain, $\mathbf{X}_{\text{freq}}$ is the frequency domain representation of the transmitted block, and $\boldsymbol{\eta}_{\text{freq}}$ is the additive white Gaussian noise (AWGN) in the frequency domain.

For the monostatic radar receiver, the echo is reflected by a target located at a distance d , with velocity v and radar cross-section (RCS) σ_{RCS} . The round-trip delay is $\tau_0 = 2d/c$, where c is the speed of light. The received radar echo is affected by the same multipath environment and thermal noise, maintaining the structural form of the communication channel but scaled by the two-way path loss L_{path} , defined as [12]:

$$L_{\text{path}} = \frac{\lambda^2 \sigma_{\text{RCS}}}{(4\pi)^3 d^4}, \quad (2)$$

where λ represents the signal wavelength.

III. PROPOSED ISAC OCDM-IM ARCHITECTURE

To achieve highly efficient joint radar sensing and communication, we propose a novel non-orthogonal ST scheme within the OCDM-IM framework. This section details the proposed transmit signal architecture, followed by the tailored iterative receiver designs that decouple the superimposed streams.

A. Proposed Transmit Signal Formulation

The incoming bit stream is split to simultaneously select the active subcarrier indices and map M -QAM symbols onto these active subcarriers. The resulting communication signal vector in the Fresnel domain is denoted as $\mathbf{x}_{IM} \in \mathbb{C}^{N \times 1}$.

For sensing functionality and CHEST, a low-power ZC sequence $\mathbf{x}_{ZC} \in \mathbb{C}^{N \times 1}$ is superimposed onto the communication signal. The n -th element of the ZC sequence is defined as:

$$x_{ZC}[n] = \exp\left(-j \frac{\pi \cdot u \cdot n(n + N \bmod 2)}{N}\right), \quad (3)$$

$$n = 0, 1, \dots, N-1,$$

where u represents the root index of the ZC sequence. The composite transmit signal in the Fresnel domain is formulated as:

$$\mathbf{x}_{Fn} = \sqrt{P_c} \mathbf{x}_{IM} + \sqrt{P_r} \mathbf{x}_{ZC}, \quad (4)$$

where P_c and P_r represent the power allocated to the communication and radar signals, respectively, as illustrated in Fig. 1.

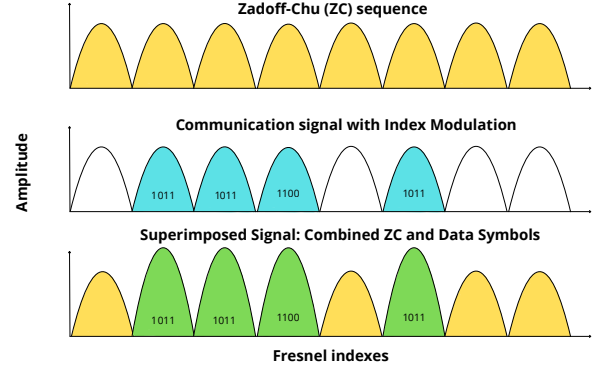


Fig. 1. Superposition of the ZC sequence $\mathbf{x}_{ZC}$ and the OCDM-IM communication signal $\mathbf{x}_{IM}$ ($N = 8$, $g = 1$, $k = 4$).

To obtain the time-domain signal, an IDFnT is applied to the superimposed vector. The time-domain transmitted block is given by:

$$\mathbf{s} = \boldsymbol{\Phi}_{\text{DFnT}}^H \mathbf{x}_{Fn}, \quad (5)$$

where $\boldsymbol{\Phi}_{\text{DFnT}} \in \mathbb{C}^{N \times N}$ is the DFnT matrix described in [13]. Before transmission, a CP of length L_{CP} is included in $\mathbf{s}$ to mitigate the inter-symbol interference (ISI) caused by the dispersive multipath channel.

B. Iterative SIC-Based Communication Receiver

At the communication receiver, the superimposed ZC sequence causes severe self-interference to the OCDM-IM symbols. To decouple the signals, we employ an iterative SIC architecture combined with an LMMSE estimator.

Prior to the first SIC iteration ($i = 1$), the receiver exploits the diagonal channel matrix in the frequency domain. By treating the unknown communication symbols as additive noise with variance P_c , an LMMSE filter, derived by scaling a coarse least squares estimate by the SINR, yields the initial channel estimate $\hat{H}_{\text{freq}}^{(0)}[k]$ via the known ZC sequence $X_{ZC, \text{freq}}[k]$:

$$\hat{H}_{\text{freq}}^{(0)}[k] = \frac{Y_{\text{freq}}[k] / (\sqrt{P_r} X_{ZC, \text{freq}}[k])}{1 + \frac{\sigma_\eta^2 + P_c}{P_r |X_{ZC, \text{freq}}[k]|^2}}, \quad \text{for } k = 0, \dots, N-1, \quad (6)$$

where $|X_{ZC, \text{freq}}[k]|^2$ denotes the squared magnitude of the k -th scalar element, which is constant across all subcarriers due to the sequence's constant envelope property, and σ_η^2 is the AWGN power. This initial estimate is immediately transformed to the time domain, denoised using the thresholding technique described later in this section, and converted back to the frequency domain to provide the refined starting channel $\hat{H}_{\text{freq}}^{(0)}$.

For the i -th SIC iteration, the process begins by removing the estimated radar interference from the frequency-domain received signal $\mathbf{Y}_{\text{freq}}$ to isolate the communication data. This cancellation relies on the channel estimate obtained from the previous iteration, $\hat{H}_{\text{freq}}^{(i-1)}$:

$$\mathbf{Y}_{\text{comm, freq}}^{(i)} = \mathbf{Y}_{\text{freq}} - \sqrt{P_r} (\mathbf{X}_{ZC, \text{freq}} \odot \hat{H}_{\text{freq}}^{(i-1)}), \quad (7)$$

where $\odot$ denotes element-wise multiplication. The "clean" frequency-domain signal $\mathbf{Y}_{\text{comm, freq}}^{(i)}$ then undergoes MMSE

equalization. The k -th element of the equalization weight vector $\mathbf{W}_{\text{freq}}^{(i)}$ also utilizes the previous channel estimate:

$$W_{\text{freq}}^{(i)}[k] = \frac{(\hat{H}_{\text{freq}}^{(i-1)}[k])^*}{|\hat{H}_{\text{freq}}^{(i-1)}[k]|^2 + \sigma_{\text{reg}}^2}, \quad \text{for } k = 0, \dots, N-1, \quad (8)$$

where $(\cdot)^*$ denotes the complex conjugate and σ_{reg}^2 is a regularization factor representing the effective noise variance normalized by the communication signal power. The equalized signal is subsequently converted back to the time domain via IFFT before being transformed to the Fresnel domain via the DFnT to detect the communication symbols:

$$\mathbf{y}_{\text{comm}, \text{Fn}}^{(i)} = \Phi_{\text{DFnT}} \mathcal{F}^{-1} \left\{ \mathbf{W}_{\text{freq}}^{(i)} \odot \mathbf{Y}_{\text{comm}, \text{freq}}^{(i)} \right\}. \quad (9)$$

This recovers the orthogonality of the OCDM chirps, allowing the symbol detector to accurately identify the active indices and demodulate the QAM symbols. Soft-decision probabilities are utilized to reconstruct the estimated communication signal $\hat{\mathbf{x}}_{\text{IM}}^{(i)}$. Finally, the reconstructed data is transformed into the frequency domain ($\hat{\mathbf{X}}_{\text{IM}, \text{freq}}^{(i)}$) and subtracted from the original frequency-domain signal to refine the channel estimate for the next iteration:

$$\mathbf{Y}_{\text{radar}, \text{freq}}^{(i)} = \mathbf{Y}_{\text{freq}} - \sqrt{P_c} (\hat{\mathbf{X}}_{\text{IM}, \text{freq}}^{(i)} \odot \hat{\mathbf{H}}_{\text{freq}}^{(i-1)}). \quad (10)$$

To further suppress residual QAM interference, this updated radar signal $\mathbf{Y}_{\text{radar}, \text{freq}}^{(i)}$ is re-estimated via LMMSE and transformed into the time-domain impulse response $\hat{h}^{(i)}[n]$ via IFFT. A soft-denoising thresholding technique is then applied to completely zero-out noise components beyond the maximum expected delay spread L_{ch} . Following [14], the Rayleigh noise standard deviation σ is robustly estimated from the median of the interference-only tail ($n > L_{\text{ch}}$). Since $\sigma \approx 0.85 \cdot \text{median}$ [15], the three-sigma rule [16] yields an adaptive threshold $\gamma^{(i)} \approx 2.55 \cdot \text{median}$, generating the denoised taps $\hat{h}^{(i)}[n]$:

$$\hat{h}^{(i)}[n] = \begin{cases} \tilde{h}^{(i)}[n] \left(\frac{|\tilde{h}^{(i)}[n]|^2}{|\tilde{h}^{(i)}[n]|^2 + (\gamma^{(i)})^2} \right), & 0 \leq n \leq L_{\text{ch}}, \\ 0, & n > L_{\text{ch}}. \end{cases} \quad (11)$$

After denoising, the refined channel estimate is converted back to the frequency-domain via FFT to obtain $\hat{\mathbf{H}}_{\text{freq}}^{(i)} = \mathcal{F} \left\{ \hat{\mathbf{h}}^{(i)} \right\}$. This updated channel is stored and will be utilized in the interference cancellation step of the subsequent $(i+1)$ -th iteration. This process is repeated for a predefined number of iterations, progressively cleaning both the data symbols and the channel state information (CSI) before feeding the final Log-Likelihood Ratios (LLRs) to a Viterbi decoder.

C. Monostatic Radar Receiver with CPI

Because the superimposed ZC sequence operates at a lower power level relative to the communication signal to minimize self-interference, extracting radar target parameters from a single observation is challenging. To overcome this, the monostatic radar receiver jointly processes the echoed signals over a train of N_p consecutive OCDM-IM blocks.

For each p -th received block, the radar receiver reconstructs the full transmitted reference $\mathbf{S}_{\text{ref}, \text{freq}, p}$ in the frequency domain, accounting for both the known ZC sequence and the known QAM symbols, including the cyclic prefix to preserve the physical structure of the transmitted pulse.

The estimation of target reflections, while suppressing communication interference, is performed in two stages. First, a matched filter (MF) [12] is applied by correlating the received echo $\mathbf{Y}_{\text{freq}, p}$ with the reconstructed reference:

$$\mathbf{R}_{\text{MF}, \text{freq}, p} = \mathbf{Y}_{p, \text{freq}} \odot \mathbf{S}_{\text{ref}, \text{freq}, p}^*. \quad (12)$$

A frequency-domain Wiener filter is then employed to obtain the final instantaneous range response $\mathbf{R}_{p, \text{freq}}$:

$$\mathbf{R}_{p, \text{freq}} = \frac{\mathbf{R}_{\text{MF}, \text{freq}, p}}{|\mathbf{S}_{\text{ref}, \text{freq}, p}|^2 + \alpha}, \quad (13)$$

where α is a regularization factor proportional to the inverse of the signal-to-noise ratio (SNR), ensuring numerical stability during the sensing channel inversion.

To prevent destructive phase cancellation caused by high-mobility targets over the integration period, the receiver organizes the sequence of instantaneous responses into a two-dimensional fast-time and slow-time matrix $\mathbf{D} \in \mathbb{C}^{N+L_{\text{cp}} \times N_p}$. The fast-time dimension refers to the internal sampling of a single pulse, capturing the echo delay to determine the targets' range. In contrast, the slow-time dimension tracks the signal changes across consecutive pulses (indexed by p). By observing the phase evolution over the pulse repetition interval (PRI), the receiver can accurately estimate the targets' Doppler shift.

By stacking the $\mathbf{R}_{p, \text{freq}}$ vectors as columns in this matrix, a Chebyshev window $\mathbf{w}$ with 80 dB of attenuation is applied along the fast-time dimension to control range sidelobes [11]. The matrix is then transformed into the time domain via an IFFT across the subcarriers, followed by a FFT across the N_p slow-time pulses to generate a coherent 2D Range-Doppler map $\mathbf{M}$:

$$\mathbf{M}(n, m) = \frac{1}{N_p} \sum_{p=0}^{N_p-1} (\mathcal{F}^{-1} \{ \mathbf{R}_{p, \text{freq}} \odot \mathbf{w} \})_n e^{-j2\pi \frac{mp}{N_p}}, \quad (14)$$

where n represents the range bin and m indicates the discrete Doppler bin.

To estimate the target range, the 2D Range-Doppler map is projected onto the range axis by extracting the maximum absolute magnitude across the Doppler dimension. This process results in a phase-compensated CIR, where the signal energy from all coherent pulses is concentrated. To overcome the resolution limit imposed by the systems' bandwidth, spline interpolation is applied to the CIR profile, enabling sub-sample range estimation with centimeter-level precision.

Finally, to achieve an accurate continuous velocity estimation, the algorithm extracts the complex phasor history of the detected target bin across the slow-time pulses. By performing linear regression on its continuous phase evolution, the system surpasses the resolution limits of discrete Doppler bins, providing precise and continuous velocity measurements.

TABLE I
PARAMETERS OF THE COMMUNICATION SYSTEM (OCDM-IM)

Parameter	Value
Number of subcarriers (N)	256
Number of groups (g)	16
Subcarriers per group (n_c)	16
Active subcarriers per group (k)	8/6
Modulation order	4-QAM
Cyclic prefix length (L_{CP})	64 samples
Power Ratio PR (Radar/Comm)	-3 dB
Channel Rician factor (K)	0 dB
Number of SIC iterations	5
Convolutional encoder	$r = 1/2, (133, 171)_8$
Constraint length (L_c)	7
Traceback depth	35
Interleaver	Block (16 rows)
Monte Carlo simulations	10,000
SNR range (dB)	0 to 20 (step 2)
Maximum expected delay spread (L_{ch})	64
Pulse repetition interval (PRI)	128 μs

 TABLE II
PDP OF THE SIMULATED CHANNELS

Tap	Delay (ns)	Sev. (0 dB)	Mild (10 dB)	Tap	Delay (ns)	Sev. (0 dB)	Mild (10 dB)
1	0	0.0	0.0	8	109	-8.0	-23.0
2	12	-0.2	-8.0	9	135	-10.0	-25.0
3	25	-0.8	-10.0	10	173	-12.0	-28.0
4	31	-1.0	-12.0	11	195	-14.0	-30.0
5	48	-3.0	-15.0	12	220	-16.0	-33.0
6	71	-5.0	-18.0	13	251	-18.0	-36.0
7	92	-6.0	-21.0	14	295	-20.0	-40.0

IV. RESULTS AND DISCUSSIONS

A. Performance Metrics

To evaluate the proposed ISAC system, we analyze¹ both radar and communication performance across varying SNR levels. Radar accuracy is quantified by the root mean square error (RMSE) of range and velocity estimations, while the peak to sidelobe ratio (PSLR) measures the targets' peak prominence against the strongest interference in the 2D Range-Doppler map. For the communication link, we evaluate the CHEST MSE alongside the BER, distinguishing between QAM and index modulation (IM) bits, both before and after Viterbi forward error correction (FEC).

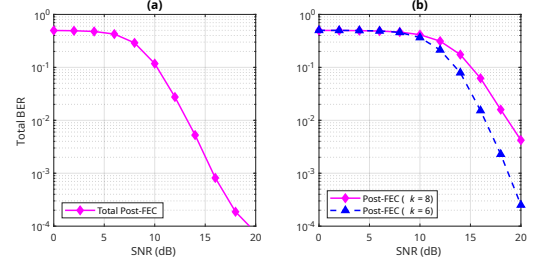
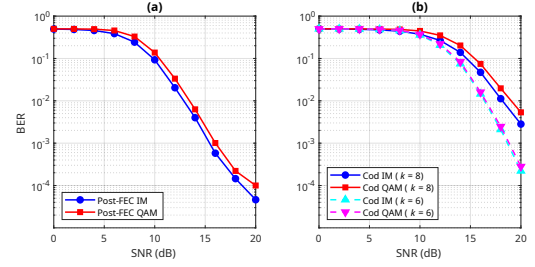
Table I presents the key simulation parameters adopted to obtain the numerical results presented in this section. Additionally, Table II details the PDP for the simulated multipath channels, presenting the delays and average power for both the severe ($K = 0$ dB) and mild ($K = 10$ dB) scenarios. To assess ISAC target detection capabilities, the scenario includes two dynamic targets at 65.42 m ($v = 15$ m/s) and 110.30 m ($v = -12$ m/s).

B. Communication Performance

As can be observed in Fig. 2, the SIC process yields significant gains in CHEST accuracy.

Regarding BER, Fig. 3 and Fig. 4 illustrate the performance for both IM and QAM symbols. The IM bits exhibit higher

¹The MATLAB source code for the proposed ISAC architecture is publicly available at <https://github.com/GeorgeTakemiya/isac-ocdm-im-sic>.

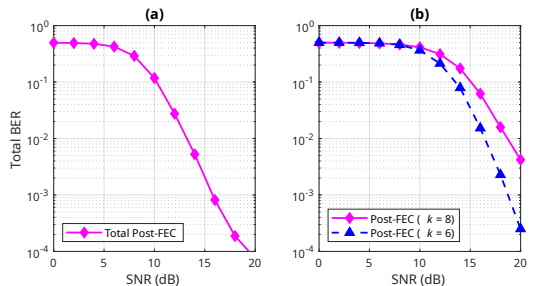

 Fig. 2. Channel estimation MSE versus SNR for (a) mild ($K = 10$) and (b) severe ($K = 0$) fading scenarios.

 Fig. 3. Post-FEC BER performance of IM and QAM symbols vs SNR for (a) mild ($K=10$) and (b) severe ($K=0$) fading scenarios.

robustness than QAM symbols due to the inherent diversity of IM. Furthermore, Fig. 4(b) demonstrates that increasing the sparsity ($k = 6$) in severe scenarios results in a performance gain compared to a denser configuration ($k = 8$). This highlights the fundamental trade-off between spectral efficiency and reliability: the $k = 6$ setup improves robustness at the cost of reducing the throughput to 384 bits per symbol.

C. Radar Performance

Fig. 5 shows the PDP for Targets 1 and 2 at an SNR of 20 dB. While the conventional MF [12] presents a noise floor around -30 dB, the proposed Wiener-based approach successfully suppresses communication-induced interference. Moreover, increasing the CPI length to 32 pulses further refines the estimation precision, as detailed in Table III.

As shown in Fig. 6 and Fig. 7, the proposed method provides a highly linear phase response. The Wiener filter enhances target detection without compromising this phase linearity, preserving the coherence required for accurate Doppler processing even for the more distant Target 2. The quantitative estimation results are summarized in Table III.


 Fig. 4. Total post-FEC BER performance vs SNR for (a) mild ($K=10$) and (b) severe ($K=0$) fading scenarios.

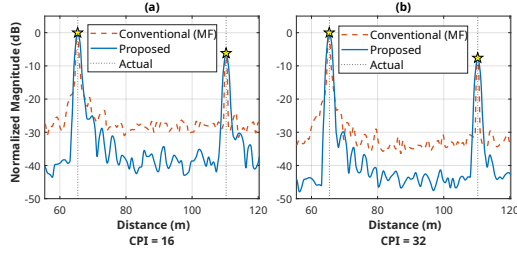


Fig. 5. PDP comparison for distance estimation (a) CPI = 16 and (b) CPI = 32 pulses.

TABLE III

RADAR ESTIMATION PERFORMANCE COMPARISON AT SNR = 20 dB

Target	Method	Actual Values		Est. (CPI = 16)		Est. (CPI = 32)	
		Dist. (m)	Vel. (m/s)	Dist. (m)	Vel. (m/s)	Dist. (m)	Vel. (m/s)
Target 1	Proposed	65.42	15.00	65.36	14.92	65.36	15.06
	Conventional (MF)			65.25	14.93	65.25	15.06
Target 2	Proposed	110.29	-12.00	110.29	-12.11	110.29	-11.98
	Conventional (MF)	110.30	-12.00	110.25	-12.11	110.25	-11.97

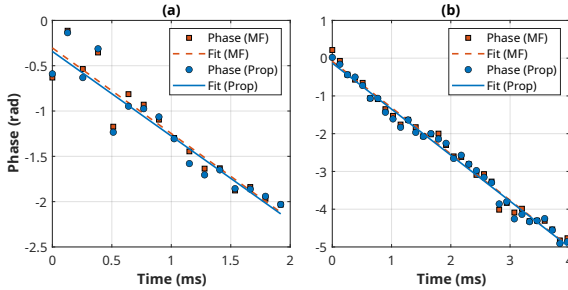


Fig. 7. Velocity estimation of target 2 comparing the performance of the MF and the proposed method (a) CPI = 15 and (b) CPI = 32 pulses.

D. Peak-to-Sidelobe Ratio (PSLR)

As illustrated in Fig. 8, the MF performance is limited by an interference floor around 17-18 dB. In contrast, the proposed method exhibits a consistent PSLR improvement as the SNR increases, reaching 24.3 dB for a CPI of 32 pulses. This confirms that the mitigation strategy effectively expands the dynamic range of the ISAC receiver by suppressing the deterministic data interference.

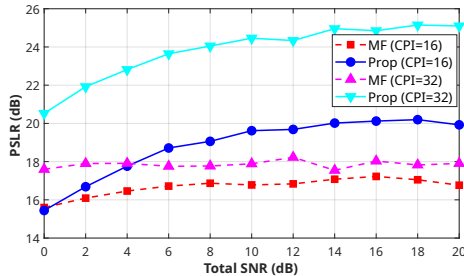


Fig. 8. PSLR versus SNR.

V. CONCLUSION

In this work, a novel non-orthogonal ISAC architecture utilizing OCDM-IM with superimposed ZC sequences with SIC receiver is presented. Because OCDM allows all carriers to share the entire spectrum rather than dividing it into frequencies, this superimposed framework increases spectral

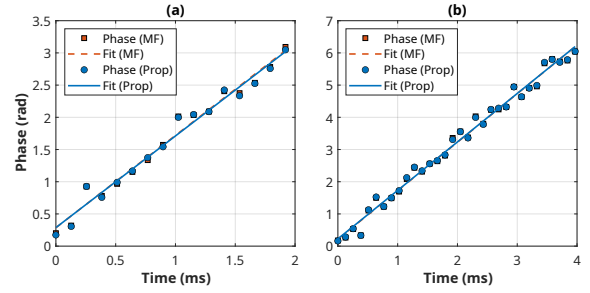


Fig. 6. Velocity estimation of target 1 comparing the performance of the MF and the proposed method (a) CPI = 15 and (b) CPI = 32 pulses.

efficiency without sacrificing high-precision sensing. Simulations demonstrated that the iterative SIC receiver effectively mitigates mutual interference, significantly reducing both BER and CHEST errors. The proposed system provides a robust, high-performance solution for the coexistence of communication and sensing in upcoming 6G networks.

REFERENCES

- [1] D. C. Nguyen, M. Ding, P. N. Pathirana, A. Seneviratne, J. Li, D. Niyato, O. Dobre, and H. V. Poor, "6G Internet of Things: A Comprehensive Survey," *IEEE Internet of Things Journal*, vol. 9, no. 1, pp. 359–383, 2022.
- [2] I. F. Akyildiz, A. Kak, and S. Nie, "6G and Beyond: The Future of Wireless Communications Systems," *IEEE Access*, vol. 8, pp. 111285–111295, 2020.
- [3] N. González-Prelcic, M. F. Keskin, O. Kaltiokallio, M. Valkama, D. Dardari, X. Shen, Y. Shen, M. Bayraktar, and H. Wymeersch, "The Integrated Sensing and Communication Revolution for 6G: Vision, Techniques, and Applications," *Proceedings of the IEEE*, pp. 1–0, 2024.
- [4] W. Zhou, R. Zhang, G. Chen, and W. Wu, "Integrated Sensing and Communication Waveform Design: A Survey," *IEEE Open Journal of the Communications Society*, vol. 3, pp. 1930–1949, 2022.
- [5] L. Méndez-Monsanto, K. Chen-Hu, M. J. Fernández-Getino García, and A. G. Armada, "Simultaneous Channel Estimation and Sensing with Superimposed Training in OFDM," in *2025 IEEE Global Communications Conference (GLOBECOM)*, IEEE, 2025.
- [6] Z. Wang, X. Liu, L. Xiong, and M. Peng, "Successive Interference Cancellation for Communication and Radar Coexistence," in *2022 IEEE 95th Vehicular Technology Conference: (VTC2022-Spring)*, IEEE, 2022.
- [7] L. Giroto de Oliveira, B. Nuss, M. B. Alabd, A. Diebold, M. Pauli, and T. Zwick, "Joint Radar-Communication Systems: Modulation Schemes and System Design," *IEEE Transactions on Microwave Theory and Techniques*, vol. 70, no. 3, pp. 1521–1551, 2022.
- [8] J. Liu, P. Yang, T. Q. S. Quek, Y. Xiao, and W. Xiang, "Orthogonal Chirp Division Multiplexing With Index Modulation," *IEEE Transactions on Communications*, vol. 72, no. 8, pp. 4577–4590, 2024.
- [9] Y. Zhao, P. Yang, J. Liu, L. Qian, S. Yang, G. Wu, Y. Xiao, and T. Q. S. Quek, "Orthogonal Chirp Division Multiplexing With Index Modulation for ISAC-Based Communication Systems," in *2025 IEEE 102nd Vehicular Technology Conference (VTC2025-Fall)*, IEEE, 2025.
- [10] S. Blunt and K. Gerlach, "Adaptive pulse compression via MMSE estimation," *IEEE Transactions on Aerospace and Electronic Systems*, vol. 42, no. 2, pp. 572–584, 2006.
- [11] A. W. Doerry, "Catalog of Window Taper Functions for Sidelobe Control," Tech. Rep. SAND2017-4042, Sandia National Laboratories (SNL), Albuquerque, NM and Livermore, CA, April 2017.
- [12] B. R. Mahafza, *Radar Signal Analysis and Processing Using MATLAB*. CRC Press, 2008.
- [13] X. Ouyang and J. Zhao, "Orthogonal Chirp Division Multiplexing," *IEEE Transactions on Communications*, vol. 64, no. 9, pp. 3946–3957, 2016.
- [14] D. Donoho, "De-noising by soft-thresholding," *IEEE Transactions on Information Theory*, vol. 41, no. 3, pp. 613–627, 1995.
- [15] A. Papoulis, *Probability, Random Variables, and Stochastic Processes*. McGraw-Hill, 3 ed., 1991.
- [16] F. Pukelsheim, "The three sigma rule," *The American Statistician*, vol. 48, pp. 88–91, May 1994.