

Coordinated FMCW and OFDM for Integrated Sensing and Communication

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Abstract—Modern vehicles commonly use Frequency Modulated Continuous Wave (FMCW) radar for environmental sensing and Orthogonal Frequency Division Multiplexing (OFDM) for wireless communication. However, operating these systems independently leads to increased hardware complexity, cost, and inefficient spectrum usage. To address this, we propose a coordinated FMCW-OFDM (Co-FMCW-OFDM) system that enables integrated sensing and communication (ISAC) by allowing sensing and communication to share the same RF front end, antennas, and spectral resources. In the proposed ISAC system, the FMCW signal is superimposed on the OFDM signal and serves dual purposes: facilitating bistatic sensing and enabling channel estimation at the receiver end. Based on proposed Co-FMCW-OFDM waveform, we propose two efficient sensing algorithms—fast cyclic correlation radar (FCCR) and digital mixing and down-sampling (DMD)—which significantly reduce system complexity while accurately estimating target range and velocity. We consider a realistic channel model where delays can take any value, not just integer multiples of the sampling period. This leads to a significantly larger number of effective paths compared to the actual number of targets, which makes the sensing, channel estimation, and interference cancellation more challenging. Leveraging the sensing results, we develop a sensing-aided effective channel estimation method which effectively reconstructs the channel under arbitrary delay condition based on successive interference cancellation and propose an interference cancellation scheme that removes the FMCW signal before the OFDM demodulation. Simulation results demonstrate that the proposed system achieves superior sensing accuracy, improved channel estimation, and lower bit error rate (BER) compared to conventional OFDM systems with embedded pilots. The proposed scheme demonstrates superior BER performance in comparison to the conventional OFDM-plus-FMCW approach.

Index Terms—ISAC, ICAS, FMCW, OFDM, waveform, sensing, bistatic, channel estimation, interference cancellation

I. INTRODUCTION

Modern vehicles commonly use Frequency Modulated Continuous Wave (FMCW) radar for environmental sensing and Orthogonal Frequency Division Multiplexing (OFDM) for wireless communication. However, operating these systems independently leads to increased hardware complexity, cost, and inefficient spectrum usage. By sharing spectrum, hardware and

software, integrated sensing and communication (ISAC) can enhance spectrum efficiency, reduce device size, lower power consumption, and decrease costs for both communication and sensing functions. As a result, ISAC has been envisioned as a key enabler for a number of emerging applications, such as intelligent transportation systems, smart drones, health care and home security [1]–[3].

A. Background and motivation

One of the main challenges in ISAC lies in finding suitable waveforms that can be simultaneously used for information transmission and radar sensing. In recent years, there have been a lot of researches for ISAC waveform design. The design of ISAC waveforms encompasses a wide range of strategies that balance or prioritize sensing and communication objectives depending on the application context. In general, these strategies can be categorized into three paradigms: (1) Communication-Centric Design: Prioritizes communication performance while embedding sensing functionality with minimal impact. Typical approaches reuse existing communication signals (e.g. OFDM) for sensing with slight modifications [4]. (2) Sensing-Centric Design: Focuses on maintaining high-precision sensing capabilities while enabling basic communication features. These designs often originate from radar systems and embed data within radar waveforms [5]. (3) Joint or Balanced Design: Aims to co-optimize both sensing and communication performances, often formulated as multi-objective optimization problems. This includes, but is not limited to, weighted-sum utility functions that combine metrics such as communication rate and sensing accuracy. Optimization can be done in the time-frequency domain through power allocation [6]–[8] or in the spatial domain through precoding [9]–[16]. One of the major disadvantages of such approaches is that the designed waveform is only optimized for the specified environment/channel. Furthermore, such designs in general have very high complexity in solving the optimization problems.

Orthogonal time frequency space (OTFS) and its extensions are promising candidates for ISAC [17]–[23]. Research has shown that OTFS can be used as a dual purpose waveform to enhance communication performance in high mobility environment [17]–[21] and enable reliable sensing for ISAC [22], [23]. However, one of the major obstacles for OTFS is the high complexity receiver.

Orthogonal frequency division multiplexing (OFDM) is the waveform standardized for the fifth generation (5G) and the fourth generation (4G) communications, which enjoys a

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low complexity receiver to combat the channel frequency selectivity and enables a flexible multiple access scheme (OFDMA). Adopting OFDM as the ISAC waveform for both communication and sensing purposes has been extensively studied in recent years [24]–[27]. Ref. [24] introduces a channel estimation-based sensing algorithm using OFDM signals. Ref. [25] explores the use of 5G reference signals (RS) for bistatic sensing. Ref. [26] investigates a bistatic radar sensing system that utilizes 5G signals by combining different reference signals and interpolating between nonuniform grid points to reconstruct the channel for the entire grid. It shows that leveraging decision-feedback from decoded payload data as “pseudo-RS” can improve delay-Doppler detectability and robustness to noise. Additionally, ref. [27] implements a bistatic radar based on standardized 5G waveforms, using decision feedback from decoded payload data as “pseudo-RS” combined with demodulation reference signals (DMRS) to estimate the range, speed, and angle of radar targets.

FMCW radar and OFDM communication devices are widely used in various applications. For instance, in advanced automobiles, FMCW radar is typically employed for environmental sensing, while OFDM communication devices are used to connect with other vehicles or infrastructure. Integrated designs for both systems have the potential to reduce complexity and cost, share spectrum, and enhance performance. Thus, we propose a coordinated FMCW and OFDM (Co-FMCW-OFDM), in which FMCW signal and OFDM signal are designed to be coordinated and synchronized and share the same RF front end, antenna and the spectral resources.

Our motivation is to explore a practical and low-complexity ISAC solution that can be rapidly deployed using existing vehicular hardware without major architectural changes. This addresses a critical need in real-world automotive applications where cost, legacy compatibility, and spectrum efficiency are key constraints. Our approach maximizes the utilization of existing vehicular hardware, thereby facilitating the rapid and practical deployment of ISAC systems.

In contrast to most existing ISAC solutions—which are typically based on either communication-centric waveforms, sensing-centric waveforms, or jointly optimized designs—our approach emphasizes practicality and low complexity. Existing solutions often suffer from trade-offs such as degraded radar accuracy, reduced communication robustness, excessive implementation complexity, limited adaptability to varying environments, or reliance on custom RF front-end architectures.

In the proposed framework, FMCW signals serve dual purposes: facilitating sensing and enabling channel estimation at the receiver end. Different from most papers, which only consider integer delays (path delays are integer multiples of sampling period), we consider practical situations where the delay can have any value. Based on the characteristics of the FMCW signal, we propose two methods to estimate the delays and Doppler frequencies for all paths: (1) digital mixing and down-sampling (DMD), (2) fast cyclic correlation radar (FCCR) [28]. With the estimated delays and Doppler frequencies, we propose a sensing-aided channel estimation algorithm with successive interference cancellation (SIC), which achieves super performance. Since the FMCW signal

is superimposed with the OFDM signal, it is necessary to remove the FMCW signal interference before demodulating the OFDM signal to ensure optimal demodulation performance. By using the information obtained from sensing and channel estimation, the interference caused by FMCW signals is reconstructed and eliminated from the received signal prior to OFDM demodulation.

There have been substantial researches on the combination of FMCW and OFDM signals for ISAC, as seen in [29]–[34]. For example, Ref. [29] proposed a power-domain non-orthogonal ISAC waveform using FMCW and OFDM, and analyzed the impact of sensing signal discretization on overall system performance. Our proposed scheme differs from [29] in several key aspects: (1) Ref [29] employs a non-zero cyclic prefix (CP) for the FMCW signal. In contrast, we adopt an all-zero CP for the FMCW signal, which is inherited from the FMCW radar waveform design with guard interval, and reduces interference to the overlapping OFDM symbols. (2) Ref [29] does not leverage the FMCW signal for channel estimation and does not propose any method for estimating the channel. Our scheme utilizes the FMCW signal for both sensing and channel estimation. Specifically, we propose a sensing-aided channel estimation method based on SIC, which effectively improves performance in dynamic channel conditions. (3) Ref [29] performs interference cancellation after OFDM demodulation and equalization in the frequency domain, which uses the averaged received signal as the estimation of the FMCW signal. It then subtracts the estimated FMCW signal from the received signal. In our approach, interference cancellation is performed in the time domain, prior to OFDM demodulation. We use the estimated channel and the known FMCW signal to cancel the FMCW contribution, which is more accurate and yields better performance. Simulation results demonstrate that this method achieves superior interference suppression and significantly enhances overall system performance.

Ref [30] and [31] proposed a non-orthogonal superposition of FMCW and OFDM signals. In their scheme, the transmitter is coordinated such that the first chirp of each transmission frame is reserved solely for sensing—no OFDM data is transmitted during this chirp. While this avoids mutual interference, it also reduces communication data rate. In contrast, our proposed scheme transmits FMCW and OFDM signals simultaneously, maintaining full data throughput. Ref [32] introduced a waveform design where FMCW and OFDM coexist orthogonally on the same time-frequency resources. This orthogonality is achieved by constraining the FMCW chirp duration to align with pilot subcarriers in the frequency domain, enabling channel estimation. However, this constraint limits the flexibility of the FMCW waveform, potentially compromising radar sensing performance. Ref [33] and [34] explored various interference mitigation strategies, such as time and frequency resource coordination, to enable the co-existence of FMCW and OFDM waveforms.

B. Our Contributions

The following are the main contributions of our work.

- We propose a coordinated FMCW and OFDM waveform for ISAC. In this scheme, the FMCW radar signal is superimposed on the OFDM communication signal and acts as a pilot for both bistatic sensing and channel estimation at the receiver. FMCW signal and OFDM signal have the same symbol duration and are synchronized in symbol level. A special feature of our design is the adoption of an all-zero cyclic prefix (CP) for the FMCW signal. This has two primary benefits: Firstly, it enables the use of efficient and low-complexity algorithms FCCR and DMD for sensing; Secondly, it reduces the interference from the FMCW signal to the overlapping OFDM symbols, thereby enhancing communication performance.
- We propose two bistatic sensing algorithms, FCCR and DMD, for estimating the range and velocity of targets using FMCW signals. In contrast to traditional FMCW radar systems, where the received signal undergoes analogue matched filtering (mixing with the transmitted analogue FMCW signal), low-pass filtering, and sampling, the proposed DMD algorithm performs digital mixing and down-sampling based on the digitized FMCW signal. The FCCR algorithm, on the other hand, has been shown to be near-optimal for radar sensing of signals with block structures, such as cyclic-prefixed single-carrier (CP-SC) and CP-OFDM waveforms. Thanks to the specific structure of our FMCW signal design, the FCCR algorithm can be directly applied.
- Unlike many prior works that assume delays are integer multiples of the sampling period, we consider a realistic channel model where delays can take arbitrary value, not just integer multiples of the sampling period. This leads to a significantly larger number of effective paths compared to the actual number of targets, which makes the sensing and channel estimation more challenging. We derive a signal model which captures the spreading of energy across multiple taps, increasing the complexity of sensing and estimation. This modelling is more aligned with real-world propagation environments, especially for high-resolution sensing applications, yet has received limited attention in existing ISAC literature. Our contribution lies in both deriving a mathematically tractable signal model under arbitrary delays and highlighting the implications for sensing-channel estimation interaction and interference cancellation.
- Leveraging delay and Doppler frequency estimation from the sensing module, we propose a sensing-aided effective channel estimation method which effectively reconstructs the channel under arbitrary delay condition based on SIC. This method accommodates the arbitrary delays of the effective channel and iteratively refines channel estimates. While SIC is a known technique, its tailored application for sensing-informed estimation in the presence of arbitrary delays—within the Co-FMCW-OFDM framework—is a novel adaptation. We also provide a performance analysis of the channel estimator, which is typically lacking in related works.
- We develop and analyze interference cancellation methods to mitigate the impact of FMCW components on

OFDM demodulation. We present a method that reconstructs the FMCW signal component using the joint sensing-channel estimates and then removes it in the time domain prior to OFDM demodulation. Unlike conventional interference cancellation methods which cancel FMCW interference in frequency domain [29], our method explicitly models the FMCW signal structure and integrates it with real-time sensing results to achieve time domain FMCW signal interference cancellation. This is critical for practical ISAC systems where residual interference from co-existing waveforms can significantly degrade communication performance.

- We demonstrate the feasibility and performance of the proposed scheme through both analysis and simulations. We conduct an extensive simulation study comparing our proposed scheme with (1) a traditional OFDM system with embedded pilot subcarriers, and (2) the method in reference [29]. Previous works often evaluated only sensing or communication separately. Our comparative analysis offers joint evaluation of both sensing and communication performance, including Bistatic Range Doppler accuracy, channel estimation accuracy, and BER. The results demonstrate that our method achieves robust gains across all three metrics, even under realistic channel conditions and imperfect sensing.

Compared to conventional OFDM systems with embedded pilot subcarriers, the advantage of using a superposed FMCW signal as a pilot signal is summarized as follows.

1) In conventional OFDM systems, the channel estimation is based on the embedded pilot subcarriers, which needs interpolation to generate the channels at non-pilot OFDM symbols and non-pilot subcarriers, which results in inaccurate channel estimation in fast time-variant channel. Using superposed FMCW signal to do channel estimation can track the fast time-variant channel and estimate the time-variant channel more accurately. Simulation results indicate that in a rapidly time-varying channel (with high Doppler frequency), the proposed scheme delivers superior sensing performance, achieves more accurate channel estimation, and offers better BER performance compared to the conventional OFDM systems.

2) Due to the periodic nature of OFDM symbols with embedded pilot subcarriers within the OFDM frame, ghost targets can appear in the radar range-Doppler map (RDM). The presence of these ghost targets compromises the ability to detect weak targets, while our proposed method has advantage in detecting weak targets. Simulation results show that, in terms of detecting weak targets, using the FMCW signal for sensing surpass decision feedback-based sensing and channel estimation-based sensing in conventional OFDM systems. Moreover, the Co-FMCW-OFDM yields superior sensing performance without incurring the associated delays inherent in decision feedback mechanisms.

3) The additional advantage of Co-FMCW-OFDM is that: the superposed FMCW signal can be always there, like a broadcasting signal. Receivers can use it for communication and/or sensing. Sometimes, the receiver does not need communication, or when the SNR is too low for communication, the

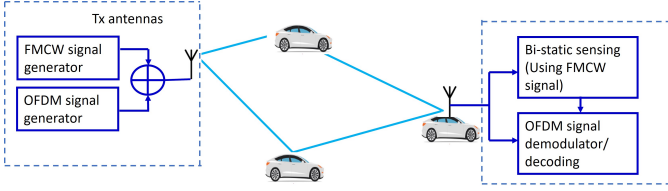


Fig. 1: Coordinated FMCW and OFDM scheme for ISAC

receiver can still use the signal to sense the environment, which can provide real-time information to enable some critical applications.

The remainder of the paper is structured as follows. Section II presents the proposed system model. Section III details FCCR sensing and DMD sensing with a superposed FMCW signal. Section IV presents the sensing-aided channel estimation and FMCW interference cancellation algorithms. Section V provides performance evaluations for the Co-FMCW-OFDM. Section VI first compares the sensing and data demodulation performance of the Co-FMCW-OFDM scheme with that of conventional OFDM systems employing embedded pilot subcarriers. It then compares the Co-FMCW-OFDM scheme with a conventional OFDM-plus-FMCW system as described in [29]. Our work is concluded in Section VII.

II. THE COORDINATED FMCW AND OFDM SYSTEM

In this section, we present the system model, the transmitted signal model, and the received signal model.

A. System Model

The proposed Co-FMCW-OFDM design is shown in Fig. 1. In the coordinated system, a communication transceiver shares the same transmitter antenna as the FMCW radar. The FMCW radar and communications transceiver are coordinated and synchronized.

The same model can be adapted as an ISAC system with superposed FMCW and OFDM waveform, where the FMCW signal is designed as a pilot signal to enable the channel estimation and bistaic sensing at the receiver.

B. Transmitted Signal Model

Fig. 2 illustrates one symbol of the proposed Co-FMCW-OFDM signal for ISAC. The radar transmits the FMCW signal known to the communication receiver. The OFDM communication signal can be written as

$$s_c(t) = \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} a_{m,n} e^{j2\pi n \Delta f t} \Pi(t - mT_{sym}), \quad (1)$$

where, T_{sym} is the duration of one OFDM symbol, $T_{sym} = T + T_{cp}$, $\Delta f = \frac{1}{T}$ is subcarrier spacing, T_{cp} is duration of Cyclic Prefix (CP), $a_{m,n}$ is the multiplexed communication data symbol on n th subcarrier of m th OFDM symbol. $\Pi(t)$ is a rectangular pulse of width T_{sym} , N is number of subcarriers, M is number of OFDM symbols.

Sampling the $s_c(t)$ with sampling period $T_s = \frac{1}{N\Delta f}$, we will get $N_a = N + N_{cp}$ samples for each OFDM symbol, where N_{cp} is the length of CP in number of samples.

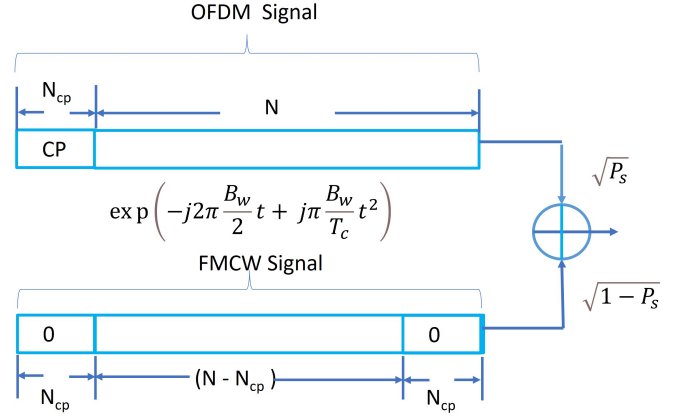


Fig. 2: One symbol of Co-FMCW-OFDM waveform

The FMCW signal is used for sensing and channel estimation at the communication receiver. The baseband FMCW signal is defined as follows,

$$s_r(t) = \begin{cases} 0 & mT_{sym} \leq t \leq mT_{sym} + T_{cp} \\ \exp(\varphi(t)) & mT_{sym} + T_{cp} < t \leq mT_{sym} + T \\ 0 & mT_{sym} + T < t \leq mT_{sym} + T_{cp} + T \end{cases} \quad (2)$$

where $\varphi(t) = j\pi \frac{B_w}{T_c} t^2 - j2\pi \frac{B_w}{2} t$, $m = 0, 1, \dots, M-1$, B_w is the bandwidth, and $T + T_{cp} = T_{sym}$ is the symbol duration of one FMCW symbol (which is same as the one OFDM symbol duration). $T_c = T - T_{cp}$ is the actual non-zero duration of one FMCW symbol. According to equation (2), there is an all-zero cyclic prefix in $s_r(t)$. This enables us to use the FCCR algorithm [28] for sensing, which will be discussed in the next sections.

To control the out of band emission (OOBE), we need to apply a pulse shaping filter to $s_c(t)$ and $s_r(t)$. After filtering, the transmitted Co-FMCW-OFDM signal can be written as,

$$x(t) = \sqrt{1 - P_s} \int_{-\infty}^{\infty} g_0(t - \tau) s_r(\tau) d\tau + \sqrt{P_s} \int_{-\infty}^{\infty} g_0(t - \tau) s_c(\tau) d\tau, \quad (3)$$

where $g_0(t)$ is the impulse response of the pulse shaping filter. Assume that the total power budget is 1, and P_s and $1 - P_s$ are the power allocated to OFDM signal and FMCW signal, respectively. P_s is a design parameter, which can be flexibly adjusted to balance the sensing and communication performance.

C. Received Signal Model

At the communication receiver, the signal is filtered by the same filter $g_0(t)$ (matched filtering). After the filtering,

the received signal can be written as

$$\begin{aligned}
 y(t) = & \sqrt{1 - P_s} \sum_{p=1}^P h_p e^{j2\pi f_{d,p} t} \\
 & \cdot \int_{-\infty}^{\infty} g(t - \tau_p - \tau) s_r(\tau) d\tau \\
 & + \sqrt{P_s} \sum_{p=1}^P h_p e^{j2\pi f_{d,p} t} \\
 & \cdot \int_{-\infty}^{\infty} g(t - \tau_p - \tau) s_c(\tau) d\tau + \eta(t), \quad (4)
 \end{aligned}$$

where $g = g_0 * g_0$ (convolution), $\eta(t)$ is the additive white Gaussian Noise (AWGN), P is the number of targets, h_p is the target reflection coefficient, τ_p is the delay associated with the path reflected by target p , $f_{d,p} = \frac{\vartheta_p}{c} f_c$ is the Doppler frequency shift induced by the moving of target p , f_c is carrier frequency, ϑ_p is the relative speed of the target p and c is the speed of light. Radar sensing is to find the delay τ_p and Doppler frequency shift $f_{d,p}$.

We consider a practical channel model, where the delay τ_p may not be an integer multiple of sampling period T_s . Let $\tau_p = l_p T_s + \alpha_p$, with $0 \leq \alpha_p < T_s$ and l_p is an integer number. The raised cosine filter can be used as the pulse shaping filter, whose time domain impulse response is given as

$$g(t) = \frac{\sin(\pi t/T_s)}{\pi t/T_s} \cdot \frac{\cos(\pi \beta t/T_s)}{1 - (2\beta t/T_s)^2}, \quad (5)$$

where β is the roll-off factor. In practical implementations, the filter is time-limited by truncating $g(t)$ to a finite duration. Specially, it is assumed that $g(t)$ is time-limited to the interval $[-\Delta T_s, \Delta T_s]$; that is $g(t) = 0$ for $t > \Delta T_s$ or $t < -\Delta T_s$. Then the sampled received signal is

$$\begin{aligned}
 y(nT_s) = & \sqrt{1 - P_s} \sum_{p=1}^P \sum_{k=-\Delta}^{\Delta} h_{p,k} s_r((n - l_p - k)T_s) \\
 & \cdot e^{j2\pi f_{d,p} n T_s} \\
 & + \sqrt{P_s} \sum_{p=1}^P \sum_{k=-\Delta}^{\Delta} h_{p,k} s_c((n - l_p - k)T_s) \\
 & \cdot e^{j2\pi f_{d,p} n T_s} + \eta(nT_s), \quad (6)
 \end{aligned}$$

where $h_{p,k} = h_p g(kT_s - \alpha_p)$, $p = 1, \dots, P$; $k = -\Delta, \dots, \Delta$. Note that since the pulse shaping filter is a raised cosine filter, it satisfies the property that $g(nT_s) = 1$ when $n = 0$ and $g(nT_s) = 0$ when $n \neq 0$. Therefore, when the delays τ_p of all paths are integer multiples of the sampling period T_s , i.e., $\alpha_p = 0$, the effective channel will contain P paths. However, if τ_p is not an integer multiple of T_s , each path will expand to $(2\Delta + 1)$ effective paths, with delay of each effective path being $(l_p + k)T_s$, $k = -\Delta, \dots, \Delta$. When delays τ_p of all paths are not the integer multiples of T_s , totally there will be $(2\Delta + 1)P$ effective paths.

Fig. 3 illustrates an example of the original channel profile and the corresponding effective channel profile. The original channel is a two-path channel with delay $\tau_1 = 0.5075 \mu s$ and $\tau_2 = 0.8395 \mu s$. Given a sampling period $T_s = 0.016276 \mu s$, we have $\tau_1 = 31.18T_s$ and $\tau_2 = 51.58T_s$. Assume $\Delta = 2$,

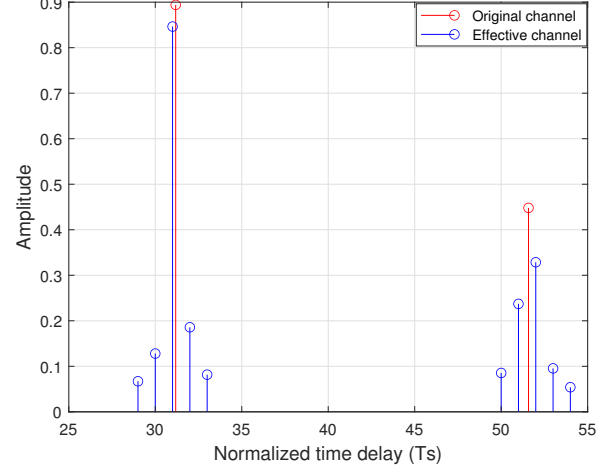


Fig. 3: An example of original and effective channel profile

the effective channel will consist of 10 paths, with path delay being $[29, 30, 31, 32, 33, 50, 51, 52, 53, 54]T_s$.

In general, given an original channel with P paths, the effective channel will consist of $(2\Delta + 1)P$ paths, whose channel coefficient, delay and Doppler frequency can be written as follows:

$$(\tilde{h}_l, \tilde{\tau}_l, \tilde{f}_{d,l}), \quad l = 1, \dots, \tilde{P}, \quad (7)$$

where $\tilde{P} = (2\Delta + 1)P$ is the number of effective paths, $\tilde{\tau}_l = (l_p + k)T_s$ is the time delay, $\tilde{f}_{d,l} = f_{d,p}$ is the Doppler frequency, $\tilde{h}_l = h_p g(kT_s - \alpha_p)$ is the channel coefficient, $l = (2\Delta + 1)(p - 1) + \Delta + k + 1$, $p = 1, \dots, P$; $k = -\Delta, \dots, \Delta$.

In the following we will use the effective channel model consisting of $\tilde{P}$ paths, with the channel coefficient, delay and Doppler frequency defined by equation (7). For notation simplicity, we will drop the sampling period T_s in all the equations, that is, we will use $y(n)$ for $y(nT_s)$, and similarly for others.

The signal $y(n)$ is composed of M symbols with each symbol having a length of N_a samples. In each symbol, the first N_{cp} samples are discarded. After discarding CP, the symbol m is denoted as $\bar{y}_m(n)$ ($n = 0, 1, \dots, N - 1$, $m = 0, 1, \dots, M - 1$). Assume that the normalized delay is always smaller than the CP length, i.e., $(l_p + \Delta) < N_{cp}$, then the received signal for symbol m can be expressed as,

$$\bar{y}_m(n) = \sum_{l=1}^{\tilde{P}} \tilde{h}_l s_m(\langle n - \tilde{\tau}_l \rangle_N) e^{j2\pi n v_l} e^{j2\pi m N_a v_l} + \eta_m(n), \quad (8)$$

where $v_l = \tilde{f}_{d,l}T_s$ is normalized Doppler frequency, $\tilde{\tau}_l = \tilde{\tau}_l/T_s$ is normalized delay, $\tilde{h}_l = \tilde{h}_l e^{j2\pi N_{cp} v_l}$, $\eta_m(n)$ is the additive white Gaussian noise (AWGN). $s_m(n) = \sqrt{1 - P_s} s_{r,m}(n) + \sqrt{P_s} s_{c,m}(n)$, $s_{r,m}(n)$ and $s_{c,m}(n)$ are samples of the FMCW signal and the OFDM signal for symbol m , respectively. $\langle n \rangle_K \triangleq n \bmod K$ denote the remainder of n modulo K .

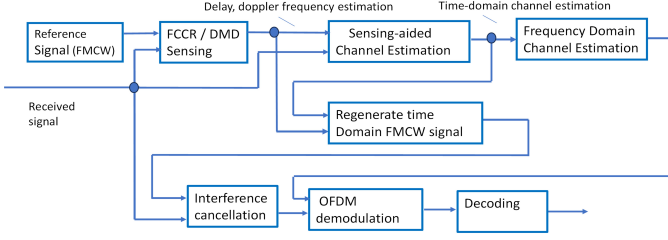


Fig. 4: Receiver Algorithm Diagram

III. BISTATIC SENSING WITH THE SUPERPOSED FMCW SIGNAL

In this section, we present bistatic sensing algorithms with the superposed FMCW signal. Fig. 4 shows the diagram of the proposed receiver algorithm. Both FCCR [28] and the DMD sensing algorithms can be used to estimate the delay and Doppler frequency. By utilizing the information obtained from sensing and channel estimation, interference from FMCW signals is reconstructed and eliminated from the received signal in the time domain before OFDM demodulation.

A. FCCR Sensing Algorithm

The FCCR sensing algorithm [28] was derived for monostatic radar sensing based on the maximum likelihood (ML) principle and can be used for any waveform with a CP structure. Theoretically, it is nearly optimal given that the delay is within the CP range. Here we consider using the FCCR for bistatic sensing with the superposed FMCW signal.

In FCCR, the cyclic correlation of the received signal with the transmitted signal is used for range and speed detection. For bistatic sensing in Co-FMCW-OFDM, we only know the transmitted FMCW signal at the receiver. Thus, we can only use the cyclic correlation of the received signal with the FMCW signal. After discarding the CP, we denote the received signal for m -th symbol duration as $\bar{y}_m(n)$, and the transmitted FMCW signal for m -th symbol duration as $s_{r,m}(n)$, $n = 0, 1, \dots, N-1$. The cyclic correlation between the received signal $\bar{y}_m(n)$ and the transmitted FMCW signal $s_{r,m}(n)$ is

$$r(n, m) = \sum_{l=0}^{N-1} \bar{y}_m(l) s_{r,m}^*(\langle l - n \rangle_N), \quad (9)$$

$$m = 0, 1, \dots, M-1.$$

With similar derivations as in [28], we can get the range and speed detection with the cyclic correlation $r(n, m)$. To save space, we omit the details here. The cyclic correlation $r(n, m)$ can be computed by the fast Fourier transform (FFT) as shown in [28]. After getting $r(n, m)$, we compute the FFT of $r(n, m)$ to get $R(n, k)$ as

$$R(n, k) = \sum_{m=0}^{M-1} r(n, m) e^{j \frac{-2\pi m k}{M_1}}, \quad (10)$$

where $M_1 \geq M$ is a positive integer for the FFT size. Selecting $M_1 > M$ can improve the accuracy of Doppler frequency estimation. Based on the range-Doppler map (RDM)

$|R(n, k)|$, we can find the delay and Doppler frequency of the P targets [28].

B. DMD Sensing Algorithm

In a conventional FMCW radar receiver, the received signal is mixed with the reference signal in the analog domain [35], [36]. This mixed signal is then passed through a low-pass filter to remove high-frequency components and isolate the beat frequency, to reveal the target's range and speed. However, this FMCW radar processing cannot be used in a common communication receiver for bistatic sensing, as a common communication receiver does not have the transmitted analogue FMCW signal and the required hardware. Thus, we propose the DMD sensing algorithm to estimate the delay and Doppler frequency. The key differences between the conventional FMCW sensing algorithm and the DMD sensing algorithm are as follows: 1) In the DMD sensing algorithm, mixing is performed in the digital domain; 2) As the beat frequency is relatively low, a downsampler is used in the DMD algorithm to reduce the sample size, thereby decreasing the implementation complexity.

Adopting the effective channel model described in (6), we can write $y(n)$ as,

$$y(n) = \sqrt{1 - P_s} \sum_{l=1}^{\tilde{P}} \tilde{h}_l s_r(n - \tilde{\varepsilon}_l) e^{j2\pi \tilde{f}_{d,l} n T_s} + \sqrt{P_s} \sum_{l=1}^{\tilde{P}} \tilde{h}_l s_c(n - \tilde{\varepsilon}_l) e^{j2\pi \tilde{f}_{d,l} n T_s} + \eta(n). \quad (11)$$

The known sampled FMCW signal can be written as,

$$s_{ref}(n) = s_r(n) = \exp \left(j\pi \frac{B_w}{T_c} (n T_s)^2 - j2\pi \frac{B_w}{2} n T_s \right). \quad (12)$$

Multiplying the $s_{ref}(n)$ with the conjugate of received signal $y(n)$, we get,

$$\begin{aligned} \Upsilon(n) &= s_{ref}(n) y^*(n) \\ &= \sqrt{(1 - P_s)} \sum_{l=1}^{\tilde{P}} \tilde{h}_l^* s_r^*(n) s_r^*(n - \tilde{\varepsilon}_l) e^{-j2\pi \tilde{f}_{d,l} n T_s} \\ &\quad + \sqrt{P_s} \sum_{l=1}^{\tilde{P}} \tilde{h}_l^* s_r^*(n) s_c^*(n - \tilde{\varepsilon}_l) e^{-j2\pi \tilde{f}_{d,l} n T_s} + s_r(n) \eta^*(n) \\ &= \sqrt{(1 - P_s)} \sum_{l=1}^{\tilde{P}} \tilde{h}_l^* e^{j2\pi f_{bt} n T_s} e^{-j\pi \frac{B_w}{T_c} (\tilde{\tau}_l^2 + T_c \tilde{\tau}_l)} e^{-j2\pi \tilde{f}_{d,l} n T_s} \\ &\quad + \sqrt{P_s} \sum_{l=1}^{\tilde{P}} \tilde{h}_l^* s_r^*(n) s_c^*(n - \tilde{\varepsilon}_l) e^{-j2\pi \tilde{f}_{d,l} n T_s} + s_r(n) \eta^*(n), \end{aligned} \quad (13)$$

where $f_{bt} = \frac{B_w}{T_c} \tilde{\tau}_l$ is the beat frequency. In (13), the first term is the desired signal, the second term is the interference from the OFDM signal, and the third term is the noise. Note that the first term is a narrowband signal; we can use a low-pass filter to reduce the impact of interference and noise. Denote

the output of the low-pass filter as $\Lambda(n)$. The relationship between $\Lambda(n)$ and $\Upsilon(n)$ can be written as,

$$\Lambda(n) = \sum_{l=1}^{N_F} \lambda_l \Upsilon((n-l)), \quad (14)$$

where N_F is the number of taps of the low-pass filter, and $\{\lambda_l, l = 1, \dots, N_F\}$ is the coefficient of the low-pass filter.

$\Lambda(n)$ can be down sampled by a factor of D to reduce the sample rate and implementation complexity. The choice of D shall fulfill 1) $1/(DT_s) \geq 2f_{bt}$, 2) N_a and N_{cp} are divisible by D . After down sampling, an FMCW symbol contains $N_T = N_a/D$ samples including the gap, and $N_D = (N - 2N_{cp})/D$ samples excluding the gap. Denote $z(n, m)$ ($n = 0, 1, \dots, N_D - 1, m = 0, 1, \dots, M - 1$) as the n -th sample of the m -th FMCW symbol. The relationship between $z(n, m)$ and $\Lambda(n)$ is

$$z(n, m) = \Lambda((n + N_b + mN_T)D), \quad (15)$$

$$n = 0, 1, \dots, N_D - 1, m = 0, 1, \dots, M - 1$$

where, $N_b = N_{cp}/D$ is the number of samples to skip for each FMCW symbol during the down sample process.

To get the RDM, we first calculate the column wise FFT of $z(n, m)$, which can be written as,

$$U(l, m) = \sum_{n=0}^{N_D-1} z(n, m) e^{-j \frac{2\pi n l}{N_D}}, \quad (16)$$

$$l = 0, 1, \dots, N_D - 1, m = 0, 1, \dots, M - 1.$$

Then we calculate the row-wise IFFT of $U(l, m)$, which can be written as,

$$\bar{R}(l, k) = \sum_{m=0}^{M-1} U(l, m) e^{j \frac{2\pi m k}{M-1}}, \quad (17)$$

$$l = 0, 1, \dots, N_D - 1, k = 0, 1, \dots, M_1 - 1.$$

$\bar{R}(k, l)$ represents the sensing RDM. Based on $|\bar{R}(k, l)|$, we can determine which points are targets.

IV. SENSING-AIDED CHANNEL ESTIMATION AND INTERFERENCE CANCELLATION

In this section, we present a sensing-aided channel estimation, which estimates the path delays, the Doppler frequencies, and the channel coefficients of the effective channel based on the delay and Doppler frequency estimation obtained from the sensing procedure. By using the information obtained from sensing and channel estimation, interference from FMCW signals is reconstructed and eliminated from the received signal in the time domain before OFDM demodulation.

A. Path delay and Doppler frequency of the effective channel

At the sensing stage, we have found the normalized delay and Doppler frequency of the P targets: $(\hat{l}_p, \hat{k}_p)$ ($p = 1, \dots, P$). Due to the pulse shaping filtering, the effective channel has $\tilde{P} = (2\Delta + 1)P$ paths (see equation (6)). The pulse shaping creates $2\Delta + 1$ pseudo-targets (include the real target) around each real target. For channel estimation purposes, at each of the detected targets, we need to estimate all the $(2\Delta + 1)$ paths. Fortunately, based on the sensing results, we have the estimations of the normalized delay and Doppler frequency of the pseudo-targets as: $\{(\hat{l}_p - \Delta, \hat{k}_p), (\hat{l}_p - \Delta + 1, \hat{k}_p), \dots, (\hat{l}_p + \Delta, \hat{k}_p)\}$, ($p = 1, \dots, P, j = -\Delta, \dots, \Delta$). The strength of each pseudo-target can also be computed directly from the RDM (see equation (10)) as $A_{p,j} = |R(D_{p,j}, K_{p,j})|$, where $D_{p,j} = \hat{l}_p + j$ and $K_{p,j} = \hat{k}_p$ are normalized delay and Doppler frequency of the $(2\Delta + 1)$ pseudo-targets within the target cluster p . Thus, after the sensing, we have the path delays and Dopplers: $\hat{\tau}_p$ and $\hat{f}_{d,p}$, for all the $\tilde{P}$ paths of the effective channel.

B. Sensing-Aided Channel Estimation

After finding the path delays and Doppler frequencies, the remaining task is to estimate the channel coefficients of all the paths. We use a SIC algorithm to estimate the time domain channel coefficients. Firstly, we estimate the channel coefficient of the most significant path. Then, we regenerate the signal corresponding to the most significant path and cancel it from the received signal before we estimate the channel coefficient of the second significant path. The procedure continues until we estimate the channel coefficients for all the $\tilde{P}$ paths. For simplicity and without loss of generality, in the following we assume that the paths are ordered according to their strengths. Let $\tilde{h}_p, \tilde{\tau}_p$ and $\tilde{f}_{d,p}$ ($p = 1, \dots, \tilde{P}$) denote the time domain channel coefficient, delay and Doppler frequency of the p th significant path, respectively. Denote $\tilde{\epsilon}_p$ as the normalized delay of the p th significant path, with $\hat{l}_p$ being its estimation, and $\hat{f}_{d,p}$ as the Doppler frequency estimation for the p th significant path. Consequently, the reference signal for the p th significant path can be regenerated as,

$$\tilde{r}_p(n) = \sqrt{1 - P_s s_r} (n - \hat{l}_p) e^{j2\pi \hat{f}_{d,p} n T_s}, \quad (18)$$

$$p = 1, \dots, \tilde{P}.$$

The estimation of the channel coefficient for the most significant path ($p = 1$) can be written as

$$\hat{h}_1 = \frac{\sum_{n=0}^{MN_a-1} y(n) (\tilde{r}_1(n))^*}{\sum_{n=0}^{MN_a-1} |\tilde{r}_1(n)|^2}. \quad (19)$$

After $\hat{h}_1$ is obtained, the signal corresponding to the most significant path can be regenerated as

$$\tilde{y}_1(n) = \hat{h}_1 \sqrt{1 - P_s s_r} (n - \hat{l}_p) e^{j2\pi \hat{f}_{d,p} n T_s}. \quad (20)$$

Assume that we have obtained the estimation of the channel coefficient for the p th significant path ($p < \tilde{P}$). The estimation

of the channel estimation coefficient for the $(p+1)$ th path can be written as

$$\hat{h}_{p+1} = \frac{\sum_{n=0}^{MN_a-1} (y(n) - \tilde{y}_p(n)) (\tilde{r}_{p+1}(n))^*}{\sum_{n=0}^{MN_a-1} |\tilde{r}_{p+1}(n)|^2}, \quad (21)$$

where $\tilde{y}_p(n)$ can be written as

$$\tilde{y}_p(n) = \sum_{i=1}^p \hat{h}_i \sqrt{1 - P_s} s_r(n - \hat{l}_i) e^{j2\pi \hat{f}_{d,i} n T_s}. \quad (22)$$

According to [37], [38], when the Doppler frequency $\hat{f}_{d,p} \ll \Delta f$, the frequency domain channel estimation $\hat{H}(k, m)$ at OFDM symbol m and subcarrier k can be generated from the estimated time domain channel coefficients, delays, and Doppler frequencies as

$$\hat{H}(k, m) = \sum_{p=1}^{\tilde{P}} \hat{h}_p e^{-j2\pi k \Delta f \hat{\tau}_p} e^{j2\pi m T_{sym} \hat{f}_{d,p}}. \quad (23)$$

Similarly, the actual frequency domain channel $H(k, m)$ at OFDM symbol m and subcarrier k can be written as

$$H(k, m) = \sum_{p=1}^{\tilde{P}} \tilde{h}_p e^{-j2\pi k \Delta f \tilde{\tau}_p} e^{j2\pi m T_{sym} \tilde{f}_{d,p}}. \quad (24)$$

The normalized mean square error (NMSE) of the frequency domain channel estimation is defined as

$$\delta_{H, NMSE} = \mathbb{E} \left\{ \frac{\sum_{m=0}^{M-1} \sum_{n=0}^{N_{sc}-1} |\hat{H}(k, m) - H(k, m)|^2}{\sum_{m=0}^{M-1} \sum_{n=0}^{N_{sc}-1} |H(k, m)|^2} \right\}. \quad (25)$$

Generally speaking, $\delta_{H, NMSE}$ consists of two parts: 1) error caused by sensing, i.e., the difference between $\hat{f}_{d,p}$ and $\tilde{f}_{d,p}$, and difference between $\hat{\tau}_p$ and $\tilde{\tau}_p$; 2) error caused by channel estimation, i.e., the difference between $\hat{h}_p$ and $\tilde{h}_p$.

C. Analysis on the Sensing-Aided Channel Estimation

In the following, we analyze the estimation error of the channel coefficient for the most significant path. Note that, as the FMCW signal is superimposed on the OFDM signal, the OFDM signal is treated as interference during the channel estimation process. Assume that the estimations of τ_p and $f_{d,p}$ are ideal, i.e., $\hat{l}_1 = \tilde{\epsilon}_1$ and $\hat{f}_{d,1} = \tilde{f}_{d,1}$. Substituting (18) into (19), we get,

$$\hat{h}_1 = \tilde{h}_1 + \frac{I_{F,1} + I_{O,1} + I_{A,1}}{\sum_{n=0}^{MN_a-1} (1 - P_s) |s_r(n - \tilde{\epsilon}_1)|^2}, \quad (26)$$

where $I_{F,1}$ denotes the interference from other paths of the FMCW signal, and $I_{O,1}$ denotes the interference from the OFDM signal and $I_{A,1}$ denotes the interference from the AWGN.

$$I_{F,1} = \sum_{n=0}^{MN_a-1} \sum_{q=2}^{\tilde{P}} \tilde{h}_q (1 - P_s) s_r(n - \tilde{\epsilon}_q) s_r^*(n - \tilde{\epsilon}_1), \quad (27)$$

$$I_{O,1} = \sum_{n=0}^{MN_a-1} \sum_{q=1}^{\tilde{P}} \tilde{h}_q \sqrt{P_s(1 - P_s)} s_c(n - \tilde{\epsilon}_q) s_r^*(n - \tilde{\epsilon}_1), \quad (28)$$

$$I_{A,1} = \sum_{n=0}^{MN_a-1} \sqrt{1 - P_s} e^{-j2\pi \tilde{f}_{d,p} n T_s} s_r^*(n - \tilde{\epsilon}_1) \eta(n). \quad (29)$$

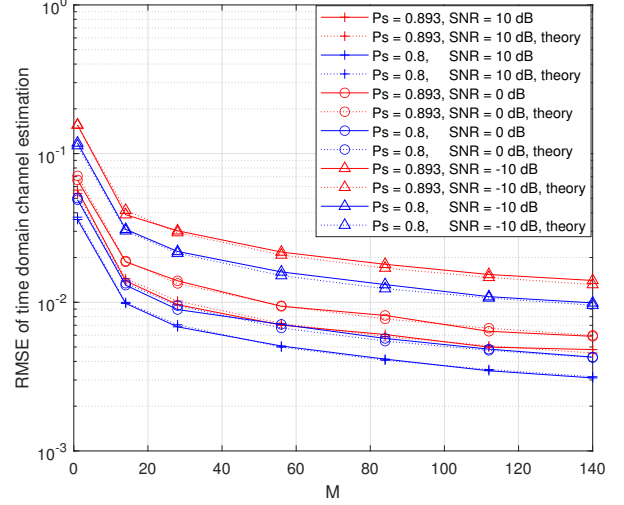


Fig. 5: Simulated and theoretic RMSE of time domain channel estimation

Assume that OFDM signal and FMCW signal are not correlated and $\mathbb{E}\{|s_r(n)|^2\} = \mathbb{E}\{|s_c(n)|^2\} = 1$. The root mean square error (RMSE) of time domain channel estimation for the most significant path can be written as,

$$\begin{aligned} \sigma_{h_1} &= \sqrt{\mathbb{E}\{|\hat{h}_1 - \tilde{h}_1|^2\}} \\ &= \sqrt{\mathbb{E}\left\{\left|\frac{I_{F,1} + I_{O,1} + I_{A,1}}{\sum_{n=0}^{MN_a-1} (1 - P_s) |s_r(n - \tilde{\epsilon}_1)|^2}\right|^2\right\}} \\ &= \sqrt{\frac{\mathbb{E}\{|I_{F,1}|^2\} + \mathbb{E}\{|I_{O,1}|^2\} + \mathbb{E}\{|I_{A,1}|^2\}}{\mathbb{E}\left\{\left|\sum_{n=0}^{MN_a-1} (1 - P_s) |s_r(n - \tilde{\epsilon}_1)|^2\right|^2\right\}}} \quad (30) \end{aligned}$$

When there is only one target and path delay associated with the path reflected by the target is integer multiples of T_s , i.e., $P = \tilde{P} = 1$, σ_{h_1} can be approximated by

$$\sigma_{h_1} \approx \sqrt{\frac{T + T_{cp}}{T - T_{cp}} \frac{N}{N_{sc}} \frac{P_s}{(1 - P_s) MN_a} + \frac{\sigma_n^2}{(1 - P_s) MN_a}}, \quad (31)$$

where $N_{sc} < N$ is the actual number of subcarriers to carry data symbols and σ_n^2 is the noise variance. Equation (31) indicates that σ_{h_1} is inversely proportional to $\sqrt{MN_a}$.

Fig. 5 shows the simulated σ_{h_1} and theoretic RMSE when there is one target and the path delay is an integer multiple of T_s ($P = \tilde{P} = 1$). The Doppler frequency is 900 Hz and subcarrier interval is 15 KHz. From Fig. 5 we can see that σ_{h_1} decreases as M and SNR increase and the simulated σ_{h_1} and theoretic σ_{h_1} are well aligned.

D. Time domain interference cancellation

In time domain, the regenerated FMCW signal can be written as,

$$\hat{r}_{fmcw}(n) = \sum_{p=1}^{\tilde{P}} \hat{h}_p \sqrt{1 - P_s} s_r(n - \hat{l}_p) e^{j2\pi \hat{f}_{d,p} n T_s}. \quad (32)$$

The time domain interference cancellation can be written as

$$\begin{aligned} y_I(n) &= y(n) - \hat{r}_{fmcw}(n) \\ &= \sqrt{1 - P_s} \sum_{p=1}^{\tilde{P}} \tilde{h}_p s_r(n - \tilde{\epsilon}_p) e^{j2\pi \tilde{f}_{d,p} n T_s} \\ &\quad + \sqrt{P_s} \sum_{p=1}^{\tilde{P}} \tilde{h}_p s_c(n - \tilde{\epsilon}_p) e^{j2\pi \tilde{f}_{d,p} n T_s} \\ &\quad - \sum_{p=1}^{\tilde{P}} \hat{h}_p \sqrt{1 - P_s} s_r(n - \hat{l}_p) e^{j2\pi \hat{f}_{d,p} n T_s} + \eta(n). \end{aligned} \quad (33)$$

At the ideal case of $\hat{h}_p = \tilde{h}_p$, $\hat{l}_p = \tilde{\epsilon}_p$, $\hat{f}_{d,p} = \tilde{f}_{d,p}$, (33) can be expressed as

$$y_I(n) = \sqrt{P_s} \sum_{p=1}^{\tilde{P}} \tilde{h}_p s_c(n - \tilde{\epsilon}_p) e^{j2\pi \tilde{f}_{d,p} n T_s} + \eta(t). \quad (34)$$

E. The computational complexity

In the following, we provide a complexity analysis of the sensing-aided channel estimation, the interference cancellation, the FCCR sensing, and the DMD algorithm.

In (18), regenerating the reference signal for the p th significant path needs $8MN_a$ real multiplication and $4MN_a$ real addition. In (19), calculating the channel coefficient needs $6MN_a + 4$ real multiplication and $5MN_a$ real addition. In (20), generating the signal corresponding to the path needs $12MN_a$ real multiplication and $6MN_a$ real addition. Canceling interference needs $2MN_a$ real addition. In summary, $26MN_a + 4$ real multiplications and $17MN_a$ real additions are needed to calculate the channel coefficient of one effective path. Since there are $\tilde{P}$ effective paths, sensing-aided channel estimation needs $\tilde{P}(26MN_a + 4)$ real multiplication and $17\tilde{P}MN_a$ real addition.

The FCCR algorithm requires $4MN \log_2 N + 4MN + 2NM_1 \log_2 M_1$ real multiplication and $6MN \log_2 N + 2MN + 4NM_1 \log_2 M_1$ real addition.

In the DMD algorithm, after down-sampling, the number of samples per chirp becomes $N_D = (N - 2N_{cp})/D$. The DMD algorithm needs $2MN_D \log_2 N_D + 2MN_D + 2N_D M_1 \log_2 M_1 + 2N_D M_1 + 2MN_a N_F + 4MN_a$ real multiplication and $4MN_D \log_2 N_D + 4N_D M_1 \log_2 M_1 + 2MN_a N_F + MN_a$ real addition.

The interference cancellation needs $12\tilde{P}MN_a$ real multiplication and $8\tilde{P}MN_a$ real addition. The computational complexity is summarized in Table I.

TABLE I: The computational complexity

Algorithm	Real multiplication	Real addition
Sensing-aided channel estimation	$\tilde{P}(26MN_a + 4)$	$17\tilde{P}MN_a$
FCCR sensing	$4MN \log_2 N + 4MN + 2NM_1 \log_2 M_1$	$6MN \log_2 N + 2MN + 4NM_1 \log_2 M_1$
DMD sensing	$2MN_D \log_2 N_D + 2N_D M_1 \log_2 M_1 + 2N_D M_1 + 2MN_a N_F + 2MN_D + 4MN_a$	$4MN_D \log_2 N_D + 4N_D M_1 \log_2 M_1 + 2MN_a N_F + MN_a$
Interference cancellation	$12\tilde{P}MN_a$	$8\tilde{P}MN_a$

V. PERFORMANCE EVALUATIONS

In this section, we evaluate the performance of the Co-FMCW-OFDM scheme. For simplicity, we only consider two targets, the reflection power of the two targets is 0 dB and -6 dB, respectively. The simulation parameters are shown in Table II.

TABLE II: Simulation Parameters

Parameters	Values
Number of OFDM subcarriers	$N = 4096$
Number of subcarriers to carrier data	$N_{sc} = 3112$
Length of CP	$N_{cp} = 288$
Subcarrier spacing	$\Delta f = 15$ KHz
Channel bandwidth	50 MHz
Sampling rate	61.44 MHz
Carrier frequency	$f_c = 23.6$ GHz
Modulation scheme	QPSK
LDPC code	(1944, 972), coding rate 0.5
Power weight for OFDM signal	$P_s = 0.8930$
Number of OFDM/FMCW symbols	$M = 140$
Relative reflection power of two targets	[0 -6] dB
Range of first target	Uniformly distributed in [48.8 244.14] m
Range of second target	Uniformly distributed in [146.48 341.79] m

A. Sensing Performance Evaluations

We compare the FCCR sensing and the DMD sensing under two scenarios: 1) scenario a, the speed of the first and second target is uniformly distributed in [45.7 91.5] km/hr, and the corresponding Doppler frequency is uniformly distributed in [1.0 2.0] KHz. Note that in this scenario, the ratio of average Doppler frequency to subcarrier spacing is 10%. 2) scenario b, the speed of first and second target is uniformly distributed in [228.8 274.5] km/hr, the corresponding Doppler frequency is uniformly distributed in [5.0 6.0] KHz. Note that, in this scenario, the ratio of average Doppler frequency to subcarrier spacing is 36.67%. According to the setting, the theoretic range and speed estimation resolution are $R_r = \frac{cT_s}{2} = 2.44$ m and $R_v = \frac{c}{2MM_1 T_{sym} f_c} = 0.068$ m/s, respectively [28].

Fig. 6 and Fig. 7 compare the RMSE of range and speed estimations between the FCCR and DMD sensing under scenario a (denoted by SA) and scenario b (denoted by SB). Under both scenarios, FCCR sensing outperforms DMD sensing by 3 dB for the range and speed estimation. We can see that under both scenarios, at high SNR, the RMSE of speed is 0.036 m/s, which is approximately half of the speed resolutions.

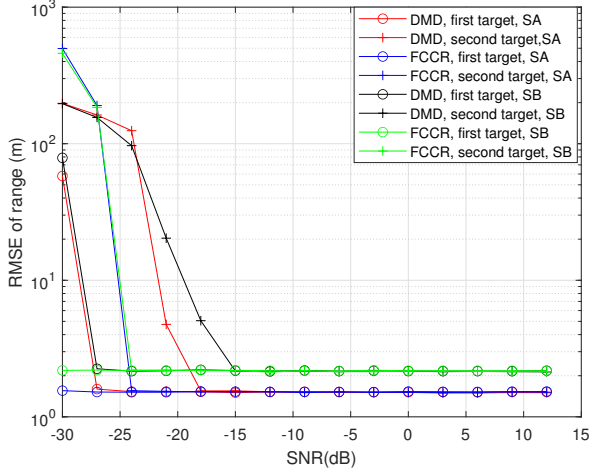


Fig. 6: Comparison of range RMSE between DMD sensing and FCCR sensing

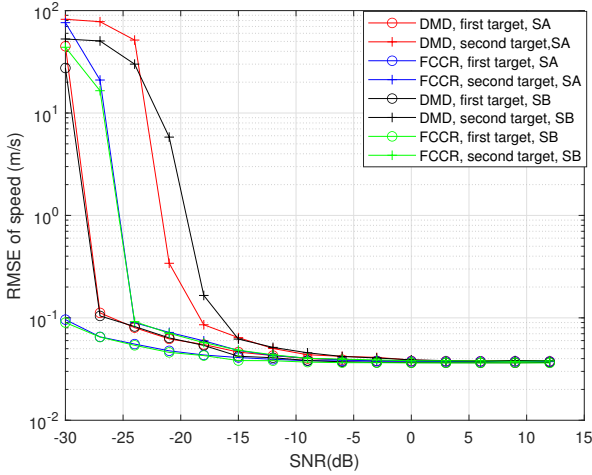


Fig. 7: Comparison of speed RMSE between DMD sensing and FCCR sensing

In scenario a, where the Doppler frequency is relatively low, the RMSE of range at high SNR is 1.46 m. In scenario b, where the Doppler frequency is relatively high, the RMSE of range at high SNR is 2.13 m. Therefore, when Doppler frequency is low, FCCR and DMD sensing achieves a range accuracy close to half of the range resolution at high SNR. However, when Doppler frequency is high, FCCR and DMD sensing achieves a range accuracy larger than half of the range resolution. This is because, at high Doppler frequency, the approximations used in the FCCR sensing are not met exactly [28], leading to degraded sensing performance.

To illustrate the impact of Doppler frequency on the RMSE of range, Fig. 8 shows the RMSE of range at 12 dB SNR as a function of average Doppler frequency. We can see that as the average Doppler frequency increases, the RMSE of range also increases, indicating a sensing accuracy degradation. When the average Doppler frequency is increased from 1.5 KHz to 7.5 KHz, the RMSE of range is degraded from 1.49 m to 2.6

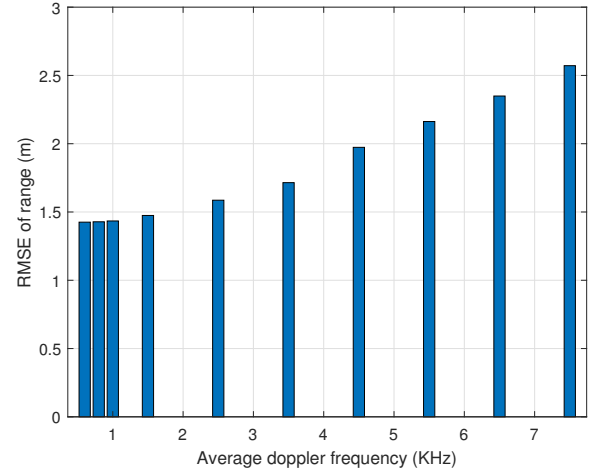


Fig. 8: Range RMSE versus Doppler frequency

m.

B. Channel Estimation and Demodulation Performance Evaluations

In this subsection, we evaluate the proposed sensing-aided channel estimation performance and BER performance under two scenarios: 1) scenario 1, the speed of the first and second target is uniformly distributed in [22.88 45.76] km/hr, and the corresponding Doppler frequency is uniformly distributed in [0.5 1.0] KHz. 2) scenario 2, the speed of the first and second target is uniformly distributed in [45.76 68.64] km/hr, the corresponding Doppler frequency is uniformly distributed in [1.0 1.5] KHz. Fig. 9 shows the NMSE of the time domain channel estimation under scenario 1 (denoted by S1) and scenario 2 (denoted by S2). The NMSE of the time domain channel estimation is defined as

$$\delta_{h,NMSE} = \mathbb{E} \left\{ \frac{\sum_{p=1}^{\bar{P}} |\hat{h}_p - \tilde{h}_p|^2}{\sum_{p=1}^{\bar{P}} |\tilde{h}_p|^2} \right\}. \quad (35)$$

In Fig. 9, “with SIC” denotes the performance of our proposed channel estimation with the SIC algorithm, as shown in subsection IV-B. To demonstrate the advantage of SIC, we also show the channel estimation performance without SIC, denoted by “without SIC”. “FCCR” denotes the performance using FCCR sensing, while “ideal” denotes the performance when sensing is ideal. Fig. 9 shows that the proposed channel estimation with SIC achieves better performance than channel estimation without interference cancellation.

Fig. 10 illustrates the BER performance comparison among perfect FMCW interference cancellation (denoted by “perfect IC”), actual FMCW interference cancellation (denoted by “actual IC”) and without FMCW interference cancellation (denoted by “without IC”) under scenario 1 and scenario 2. In this figure, FCCR is used for sensing. The gap between “actual IC” and “without IC” illustrates the gain obtained by FMCW interference cancellation. The gap between “perfect IC” and “actual IC” illustrates the performance loss due to

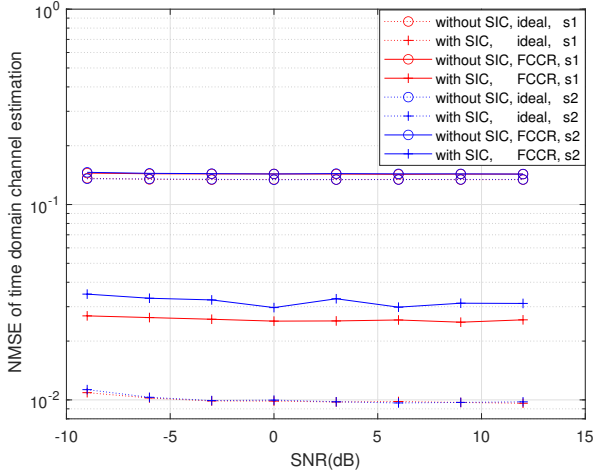


Fig. 9: NMSE of time domain channel estimation

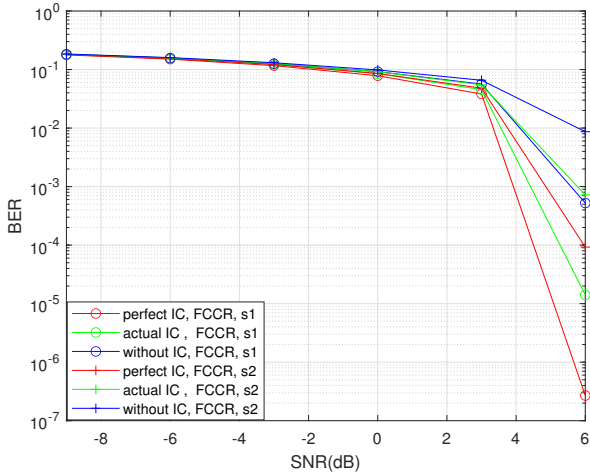


Fig. 10: Coded BER performances of the proposed system

non-perfect FMCW interference cancellation. The non-perfect FMCW interference cancellation is due to inaccuracies in the estimation of $\hat{h}_p$, $\hat{\tau}_p$, and $\hat{f}_{d,p}$ causing the regenerated FMCW signal to differ from the actual FMCW signal. Consequently, the interference from the FMCW signal cannot be entirely eliminated.

As expected, BER performance degrades as the Doppler frequency increases. The reason for this is that the orthogonality among the OFDM subcarriers are destroyed to some extent. Smaller Doppler frequency leads to a larger gap between “perfect IC” and “actual IC”. Smaller Doppler frequency also leads to a larger gap between “actual IC” and “without IC”.

C. Impact of P_s on communication BER performance

In the Co-FMCW-OFDM scheme, P_s , which represents the power allocated to OFDM signal, is a key design parameter. Increasing P_s improves communication SNR but degrades the SNR for sensing and channel estimation. Fig. 11 illustrates the coded BER performance as a function of the power ratio between OFDM and FMCW signals, defined as $R =$

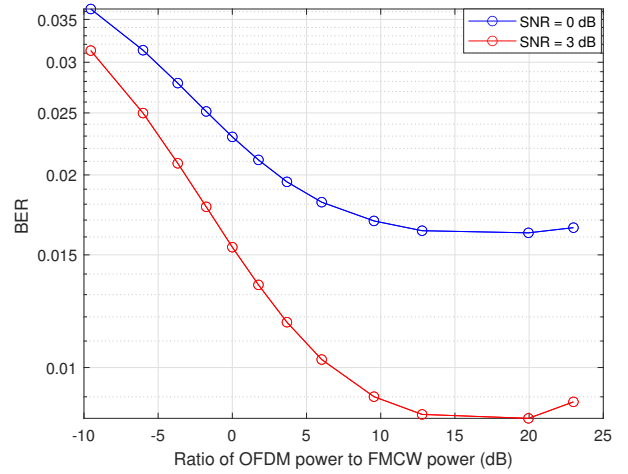


Fig. 11: Coded BER performances as a function of the power ratio between OFDM and FMCW signals

$10\log_{10}(P_s/(1 - P_s))$. In the simulation, two targets are considered with reflection powers of 0 dB and -6 dB, respectively. Their Doppler frequencies are uniformly distributed in the range [0.5 1.0] KHz and the number of symbols used for sensing is $M = 28$. As shown in Fig. 11, increasing the OFDM-to-FMCW power ratio decreases the power available for FMCW signals, thus degrading sensing and channel estimation performance. However, the communication SNR improves with higher P_s , leading to an improvement in BER until a point where further increases in P_s begin to negatively impact BER performance as well. Depending on the application requirements, P_s can be tuned to balance sensing and communication performance. For instances, P_s can be selected such that the sensing SNR remains above a required threshold while minimizing the communication BER.

VI. PERFORMANCE COMPARISON BETWEEN THE PROPOSED SYSTEM AND CONVENTIONAL SYSTEM

In this section, we first compare the performance of the proposed scheme with the conventional OFDM scheme employing embedded pilot subcarrier. Then we compare the proposed scheme with OFDM-plus-FMCW system [29].

A. Comparison with the conventional OFDM system

For the conventional OFDM scheme, we consider a similar frame and slot structure as that of the 5G standard: each slot contains 14 OFDM symbols, and each frame consists of 10 slots. Within each slot of 14 OFDM symbols, there are 3 OFDM symbols with Demodulation Reference Symbols (DMRS), located at the OFDM symbol 3, 8, and 12. In each of the three OFDM symbols, half of the subcarriers are used as DMRS (pilot subcarriers), while the remaining subcarriers carry user data. The DMRS pilot subcarriers in adjacent DMRS OFDM symbols alternate between even and odd subcarriers. In the conventional OFDM scheme, these DMRS pilot symbols are used for the channel estimation.

We evaluate the two schemes by ensuring that the overhead of DMRS in the conventional OFDM scheme is the same as the overhead of the FMCW signal in the Co-FMCW-OFDM scheme. Since the conventional OFDM scheme includes 3 DMRS OFDM symbols, with each symbol having only half of the subcarriers as pilot subcarriers, the DMRS overhead is calculated as $3/14/2 = 0.107$. Consequently, the FMCW signal power is set to $1 - P_s = 0.107$ in the Co-FMCW-OFDM scheme.

For the conventional OFDM scheme, we consider two methods for bistatic sensing: 1) channel estimation-based sensing and 2) decision feedback-based sensing. In the first method, the frequency domain channel is initially estimated with the DMRS pilots [24] as

$$\bar{H}(k_1, m_1) = \frac{\bar{Y}(k_1, m_1)}{b_{m_1, k_1}}, (k_1, m_1) \in \Omega_s. \quad (36)$$

Based on $\bar{H}(k_1, m_1)$ ($(k_1, m_1) \in \Omega_s$), a linear interpolation [25] is used to obtain the channel estimation $\tilde{H}(k, m)$ ($k = 0, 1, \dots, N-1, m = 0, 1, \dots, M-1$) for OFDM symbol m and subcarrier k ,

$$\tilde{H}(k, m) = \bar{H}(k, m_1) + \frac{\bar{H}(k, m_2) - \bar{H}(k, m_1)}{m_2 - m_1} (m - m_1). \quad (37)$$

Based on the estimated frequency domain channel, the well-known OFDM radar is used for the range and Doppler estimation [24], [25], [37].

In the decision feedback-based sensing, the receiver first performs demodulation and decoding based on the estimated channel described above to obtain the decoded payload data. The decoded payload data is then re-encoded, modulated and mapped onto OFDM data subcarriers. This regenerated OFDM signal is used as known signals or “Pseudo-RS” for FCCR sensing. For a fair comparison, we also use FCCR sensing for the Co-FMCW-OFDM scheme.

First, we compare the sensing performances under scenario 1 and scenario 2 defined in Section V-B. The simulation parameters are shown in Table II. Fig. 12 and Fig. 13 compare the RMSE of the range and speed estimations, respectively, at scenario 2. Due to the page limit, we do not show the simulation results at scenario 1. In the figures, the performance of decision feedback-based sensing and channel estimation-based sensing are denoted by “DF” and “CE”, respectively. We see that, for target 1, the Co-FMCW-OFDM scheme outperforms the OFDM scheme by 3 dB. For target 2, the Co-FMCW-OFDM scheme demonstrates a significant advantage over the OFDM scheme. This phenomenon can be explained as follows: since there are only 3 DMRS symbols within an OFDM frame, interpolation is used to estimate the channel across the entire grid. However, when the Doppler frequency is high, the channel changes rapidly. Thus, the interpolation cannot provide accurate channel estimations for the complete grid, leading to degraded sensing performance in channel estimation-based sensing schemes. In scenario 2, the Doppler frequency is high, and the orthogonality among OFDM subcarriers is destroyed to some extent. The combined effect of degraded channel estimation performance and destroyed

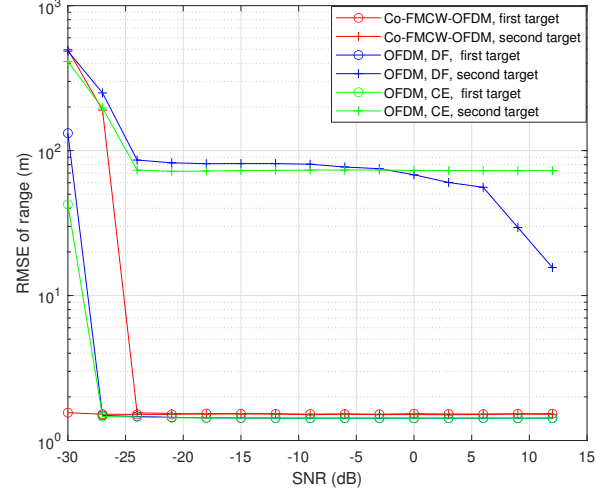


Fig. 12: Comparison of range RMSE between Co-FMCW-OFDM and the conventional OFDM (scenario 2)

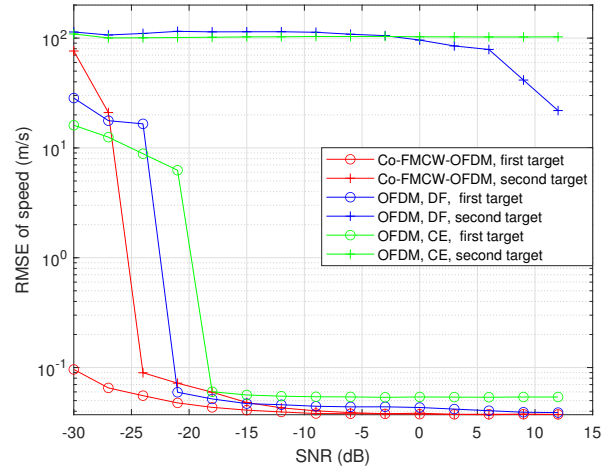


Fig. 13: Comparison of speed RMSE between Co-FMCW-OFDM and the conventional OFDM (scenario 2)

orthogonality results in large errors in the decoded payload data. Using erroneous decoded payload data as “pseudo-RS” will degrade the decision feedback-based sensing performance.

Fig. 14 compares the NMSE of the frequency domain channel estimations under scenario 1 (denoted by S1) and scenario 2 (denoted by S2). We see that the proposed scheme achieves much better performance than the conventional OFDM scheme. The reason is that, in the conventional OFDM scheme, the channel estimation for the entire grid is obtained through interpolation, which results in inaccurate channel estimation for fast time-variant channels. The higher the Doppler frequency, the worse the channel estimation accuracy. While in the proposed scheme, the superimposed pilot FMCW signal is effectively used to accurately estimate the delays, the Doppler frequencies, and the time domain channel coefficients, which are then utilized to generate more accurate frequency domain channel estimations for the entire grid. This is especially

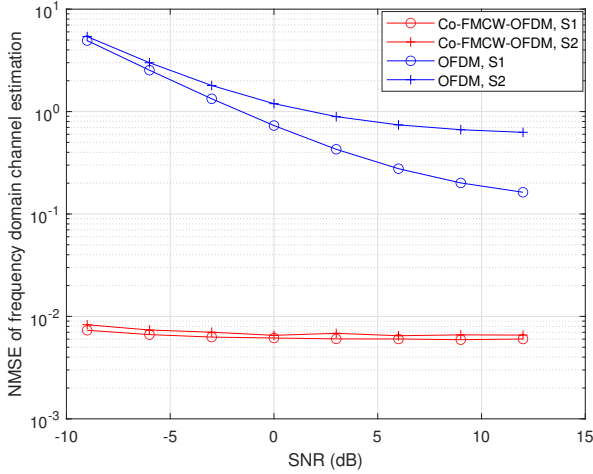


Fig. 14: Comparison of channel estimation performance between Co-FMCW-OFDM and the conventional OFDM

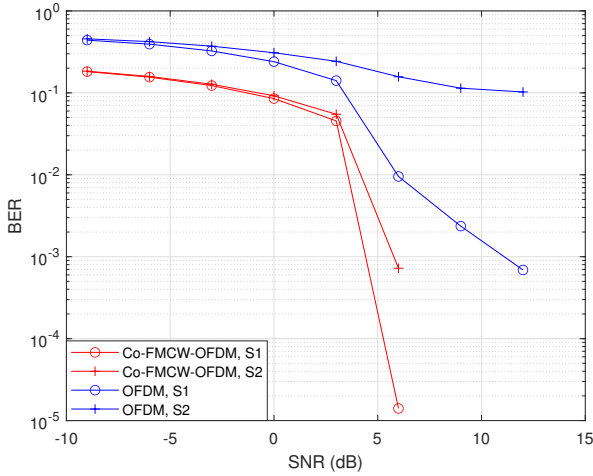


Fig. 15: Comparison of BER between Co-FMCW-OFDM and the conventional OFDM

helpful for fast time-variant channels. We can also see that, as the Doppler frequency becomes larger, the performance gap between the proposed scheme and the conventional OFDM scheme becomes more obvious.

Fig. 15 compares the BER performances. At both scenarios, the proposed scheme achieves significantly better BER performance than the conventional OFDM scheme. The reason is the same as described above.

B. Comparison with the OFDM-plus-FMCW system

In the following we present a comparison of the BER performance between the Co-FMCW-OFDM and the OFDM-plus-FMCW in [29]. The simulation scenario involves a single target with the SNR of 10 dB. The target's Doppler frequency is uniformly distributed in the range [0.5, 1.0] kHz. We evaluate two cases where the number of symbols used for sensing is $M = 56$ and $M = 140$. Since [29] does not provide a channel estimation algorithm, we assume an ideal

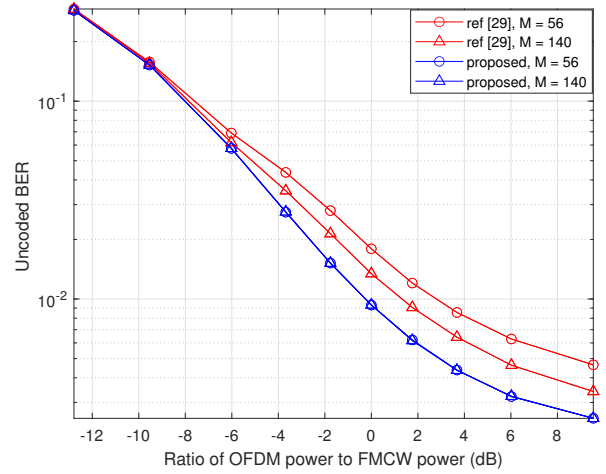


Fig. 16: Comparison of BER between Co-FMCW-OFDM and OFDM-plus-FMCW

frequency-domain channel for both OFDM demodulation and FMCW interference cancellation in their scheme. In contrast, our method employs the proposed sensing-aided channel estimation for FMCW interference regeneration and cancellation.

As shown in Fig. 16, the simulation results show that the Co-FMCW-OFDM consistently achieves lower unencoded BER, even when we use an **estimated channel**, whereas [29] adopts an **ideal channel**, for interference cancellation. As expected, increasing M improves the performance of [29] due to enhanced accuracy in the interference estimation via M symbol averaging. However, in our scheme, the FMCW interference is regenerated and cancelled directly in the time domain, making it less sensitive to change of M .

VII. CONCLUSION

We have proposed the Co-FMCW-OFDM for ISAC. In Co-FMCW-OFDM, the FMCW signal serves dual purposes: facilitating sensing and enabling channel estimation at the receiver end. We have proposed the FCCR and DMD bistatic sensing algorithms to estimate the delays and Doppler frequencies of all targets based on the FMCW pilots. With the estimated delays and Doppler frequencies, we have proposed a sensing-aided channel estimation based on the practical channel model and SIC, which achieves super estimation performance. Furthermore, a method has been presented to cancel the interference from the FMCW signals at the receiver prior to the OFDM demodulation. We have conducted comparative analyses and simulations on the sensing, channel estimation, and data demodulation performances. Simulation results show that the proposed system exhibits superior sensing, channel estimation and BER performances compared to the conventional OFDM system and the OFDM-plus-FMCW in [29].

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