

Secure Transmission in Integrated Communication-Radar Systems

Lanxin Yang , Mingjing Wang , Yechao Bai , *Member, IEEE*, Mengting Lou , and Lan Tang , *Member, IEEE*

Abstract—When the base station (BS) transmits information to users, there may be suspicious targets nearby eavesdropping. In order to ensure the communication security while detecting the suspicious targets, we study the secure transmission in integrated communication-radar systems (ICRSs), where the BS not only communicates with users but also detects the suspicious targets with orthogonal frequency division multiplexing (OFDM) signals. To improve the secrecy capacity without degrading the detection performance, we add artificial noise (AN) into OFDM signals. We first calculate the Cramér–Rao lower bound (CRLB) of the target parameters to be estimated when OFDM signals are used as the detection signals. Then, we investigate the transmission optimization scheme, which maximizes the secrecy capacity under the CRLB constraints. To solve the optimization problem, an effective algorithm based on the Lagrange duality method is proposed to obtain the optimal power and subcarrier allocation. The simulation results reveal the tradeoff between secrecy capacity and estimation performance, and the advantage of our scheme over existing schemes in terms of the estimation performance.

Index Terms—Artificial noise (AN), Cramér–Rao lower bound (CRLB), orthogonal frequency division multiplexing (OFDM), resource allocation, secrecy capacity.

I. INTRODUCTION

THE demand for simultaneously realizing communication and perception on the shared spectrum and hardware platform has spawned the development of integrated communication-radar systems (ICRSs). The main approaches for realizing ICRS with shared resources include signal multiplexing and signal sharing [1]. In time division multiplexing mode, communication and sensing are implemented by transmitting communication and radar signals separately in different time slots with shared radio frequency (RF) link [2]. In space division multiplexing mode, the main lobe and side lobe of transmit waveform (or different beams) direct to communication terminals and detection targets via beamforming technology, respectively [3], [4]. Another way to realize ICRS is to share

signals between communication and radar. Some traditional communication signals, such as spread spectrum signals and orthogonal frequency division multiplexing (OFDM) signals, have been studied as alternative waveforms of integrated signals [5], [6].

As a popular candidate for integrated waveform, OFDM signals have been widely applied in wireless communication and radar probing due to their high spectrum efficiency, good delay, and Doppler resolution. In passive radar, OFDM communication signals transmitted by television stations are also used to detect and locate airborne targets. Some algorithms have been proposed to estimate target parameters with passively received OFDM signals [7], [8], [9]. When OFDM signals are employed as dedicated radar probing signals, the symbols on each subcarrier are not random communication signals, but deterministic signals conducive to detection. In [10], complex weights transmitted over the subcarriers were optimized to improve the detection and estimation performance in multipath scenarios. In the coexisting radar and communication system, the radar waveform design needs to consider the impact on the communication system. Shi et al. [11] proposed power minimization based optimal OFDM radar waveform design to improve the estimation performance and reduce the interference of radar signals on communication signals. Temiz et al. [12] designed an optimal OFDM radar waveform by minimizing the distance between communication symbols and the radar interference received at CRs.

In the abovementioned work, the transmit waveforms for communication and radar were designed separately. While in the ICRS with shared signals, the transmitted OFDM waveform needs to be optimized to meet the requirements of communication and detection performance at the same time. In [13], transmit powers of subcarriers were optimized to satisfy the requirements of conditional mutual information for radar and data information rate for communications simultaneously. To ensure parameters identifiability, Dokhanchi et al. [6] designed a frequency-division multiplexing strategy between radar and joint radar-communication frames for OFDM-ICRS. Cheng et al. [14] designed hybrid beamforming for OFDM-ICRS by jointly optimizing the digital and RF beamformers to minimize the spectral efficiency of communication and spatial spectrum matching error of radar. Liu et al. [15] proposed a low peak-to-average power ratio (PAPR) OFDM waveform design method suitable for a large constellation size in integrated passive radar and communication systems.

In the ICRS with shared OFDM waveform, OFDM symbols carrying communication information were used to detect the unknown targets, so that the communication information may be eavesdropped by the target. From the perspective of physical layer security, when the transmitter adopts the coding

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method that can achieve secrecy capacity, it can avoid being eavesdropped. According to the definition of the secrecy channel capacity, we can increase the secrecy capacity by reducing the channel capacity between the transmitter and the target (potential eavesdropper). Adding artificial noise (AN) into the transmitted signal is a common method to degrade the channel of the eavesdropper. In multiuser OFDM systems, the subcarrier/power allocation and AN design were jointly optimized to improve the secrecy rate with a low-complexity Lagrange dual method [16]. Injecting AN into the integrated signals is also regarded as an efficient method to improve the secrecy of the ICRS. Zhou et al. [17] investigated the secure precoding and AN techniques to improve the secrecy rate of ICRS with multiple eavesdroppers by using an alternating optimization algorithm. Su et al. [18] added the AN in the beamforming design of the ICRS to guarantee the communication secrecy while detecting targets. To interfere eavesdroppers, AN was employed to minimize the signal-to-noise ratio (SNR) of targets under the condition of ensuring the signal-to-interference-plus-noise ratio (SINR) of legitimate users [19].

When an unknown target appears in the vicinity of the communication user, the target may be a potential eavesdropper. When OFDM signals transmitted by base station (BS) are used for detection, increasing the signal power at the unknown target helps to improve the estimation performance, but also increases the risk of information leakage. Therefore, in this article, we investigate the secure transmission scheme in ICRSs, in which both secrecy capacity and estimation performance are taken into account. Different from existing work on secure transmission in the ICRS system, we adopt Cramér–Rao Lower bound (CRLB) as performance metric of radar performance instead of desired beampattern in [17], [18], and [19]. As the theoretical lower bound of estimation performance, CRLB can more intuitively reflect the estimation performance of targets at different positions than beampattern. In this article, our goal is to find potential eavesdroppers in the vicinity of communication receivers (CRs). Therefore, we assume that when beamforming vectors in different subcarriers direct to CRs at different locations, the targets (potential eavesdroppers) can be illuminated by the main lobe or side lobe of the transmit waveform. To improve the estimation performance, CRs near targets will be allocated more subcarriers or power. Therefore, instead of designing the beamforming vector as in [17]–[19], the best tradeoff between communication and detection performance is achieved by optimizing subcarrier and power allocation. At the same time, by injecting AN into the transmit signals, the security of information transmission is guaranteed without degrading the detection performance. The main contributions of this article are as follows.

- 1) To enhance the perceptual ability of existing communication systems, we utilize OFDM signals transmitted by BS to communicate and detect unknown targets near CRs. In the considered dual-function system, multiple receivers cooperate to detect and estimate the parameters of unknown targets. Taking OFDM signals embedded with AN as integrated waveform, we calculate the CRLB of target velocity and location estimation and lower bound of CRLB utilizing echo signals received by multiple radar receivers (RRs). The analysis results reveal that CRLBs of estimated parameters are related to the subcarrier and power allocation, but independent of power splitting ratio

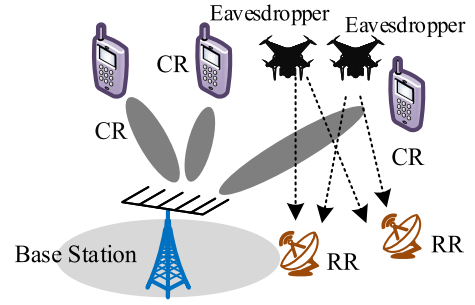


Fig. 1. System model of integrated OFDM communication-radar system with potential eavesdroppers.

of AN. Hence, splitting partial power to AN will not reduce the detection performance.

- 2) Based on the analysis result of CRLB, we can see that the resource allocation that maximizes the security capacity may not meet the target detection requirements. Therefore, resource allocation needs to take into account the performance of both communication and radar. To maximize the secrecy capacity between BS and CRs under CRLB constraints, subcarrier allocation, power allocation, and power splitting ratio are optimized with the Lagrange duality method.
- 3) The numerical results demonstrate the performance of opportunistic detection utilizing communication signals under different system configurations and tradeoff between secrecy capacity and estimation performance in active detection. Some comparison results with existing methods are presented to reflect the advantages of the proposed method.

Notation: This article uses lower-case, bold lower-case, and bold upper-case for the scalars, vectors, and matrices, respectively. $\mathbb{C}$ denotes the set of complex numbers. $|\cdot|$, $(\cdot)^*$, $(\cdot)^T$, $(\cdot)^H$, and $\arg(\cdot)$ denote the absolute value, complex conjugate, transpose, conjugate transpose, and argument, respectively. $[\cdot]^+ \triangleq \max(0, \cdot)$ and $\|\cdot\|$ denotes the 2-norm.

II. SYSTEM MODEL

As shown in Fig. 1, we consider a distributed ICRS, consisting of a dual-function BS and R RRs, which aims to communicate with M CRs and detect $\tilde{M}$ unknown targets at the same time. We assume that the BS is equipped with uniform linear array of N_t antennas and each RR has single antenna. The BS transmits OFDM signals to M CRs, and at the same time, R RRs detect $\tilde{M}$ unknown targets cooperatively. We assume that targets are in the vicinity of CRs. Therefore, when beams in different subcarriers direct to CRs, the targets can be illuminated so that they can be perceived by RRs. Then, echo signals received by R RRs are forwarded to the central processor, which completes the parameters estimation. In case of the targets being potential eavesdroppers, AN is added to the OFDM signals to improve the secrecy capacity between BS and CRs.

The transmitted signal block consists of N_b OFDM symbols with symbol duration $T_s = (N_d + N_p)T$ where T is the sampling period, N_d is the number of subcarriers and N_p is the number of cyclic prefix. Therefore, the transmitted signal is

given by

$$s(t) = \sum_{n=0}^{N_b-1} \sum_{k=0}^{N_d-1} \mathbf{b}_n(k) e^{j2\pi k \frac{t}{N_d T}} u(t - nT_s) \quad (1)$$

where $\mathbf{b}_n(k)$ is the transmit vector on the k th subcarrier in the n th OFDM symbol and $u(t)$ is the rectangular pulse with duration T_s and amplitude A . $\mathbf{b}_n(k)$ includes the transmitted information symbol and AN and can be written as $\mathbf{b}_n(k) = \sum_{m=1}^M \mathbf{w}_m(k) \rho_m(k) (\sqrt{\alpha_m(k)} p_m(k) x_{n,m}(k) + \sqrt{(1 - \alpha_m(k))} p_m(k) z_{n,m}(k))$, where $\mathbf{w}_m(k) \in \mathbb{C}^{N_t \times 1}$ is the transmit beamforming vector of the k th subcarrier with $\|\mathbf{w}_m(k)\| = 1$, $x_{n,m}(k) \sim \mathcal{CN}(0, 1)$ and $z_{n,m}(k) \sim \mathcal{CN}(0, 1)$ are the information symbol and AN in the n th OFDM symbol on the k th subcarrier, respectively, $p_m(k)$, $\alpha_m(k) \in [0, 1]$ and $\rho_m(k) \in \{0, 1\}$ are the transmit power, power splitting ratio and subcarrier allocation factor on the k th subcarrier, respectively. If the subcarrier k is allocated to m th CR, $\rho_m(k) = 1$, else $\rho_m(k) = 0$. To eliminate multiuser interference, each subcarrier is assigned to only one CR, i.e., $\sum_{m=1}^M \rho_m(k) = 1$.

We assume that channels between BS and CRs are frequency-selective block fading channels. That is, communication channels keep constant during N_b OFDM symbol periods. We suppose that CRs can cancel the added AN embedded in transmit symbols. Therefore, If $\rho_m(k) = 1$, in the n th OFDM symbol, the received signal by m th CR over k th subcarrier is

$$y_{n,m}(k) = h_m(k) \sqrt{\alpha_m(k)} p_m(k) x_{n,m}(k) + v_{n,m}(k) \quad (2)$$

where $h_m(k) = \mathbf{h}_m^H(k) \mathbf{w}_m(k)$, and $\mathbf{h}_m(k) = \sqrt{\frac{\lambda^2}{(4\pi)^2 d_m^2}} (\mathbf{a}(\theta_m) + \sum_{l=1}^L \boldsymbol{\varepsilon}_{l,m} e^{-j2\pi k l / N_d})$ is the channel frequency response of CR m on the k th subcarrier (d_m is the distance between BS and CR m , λ is the wavelength, L is the number of multipath delays, $\boldsymbol{\varepsilon}_{l,m} \in \mathbb{C}^{N_t \times 1}$ is the small-scale complex vector of the l th delay path of the m th CR, $\mathbf{a}(\theta_m) = [1 \ e^{-j\pi \sin(\theta_m)} \dots e^{-j\pi(N_t-1)\sin(\theta_m)}]^T$ denotes the transmit steering vector of the m th CR corresponding to direct path and θ_m is the direction of departure of the m th CR) and $v_{n,m}(k)$ is the additive white Gaussian noise (AWGN) with zero mean and variance σ^2 . Due to the noncooperative relationship between BS and unknown targets, we assume that AN is unable to be eliminated at the target end. Therefore, if $\rho_m(k) = 1$, the received signal by the $\tilde{m}$ th target over subcarrier k is given by

$$\tilde{y}_{n,\tilde{m}}(k) = g_{n,\tilde{m}}^{\tilde{m}}(k) \left(\sqrt{\alpha_m(k)} p_m(k) x_{n,m}(k) + \sqrt{(1 - \alpha_m(k))} p_m(k) z_{n,m}(k) \right) + \tilde{v}_{n,m}(k) \quad (3)$$

where $g_{n,\tilde{m}}^{\tilde{m}}(k) = \mathbf{g}_{n,\tilde{m}}^H(k) \mathbf{w}_m(k)$, and $\mathbf{g}_{n,\tilde{m}}(k) = \sqrt{\frac{\lambda^2}{(4\pi)^2 d_{\tilde{m}}^2}} (\mathbf{a}(\theta_{\tilde{m}}) + \sum_{l=1}^L \tilde{\boldsymbol{\varepsilon}}_{l,\tilde{m}} e^{-j2\pi k l / N_d})$ ($d_{\tilde{m}}$ is the distance between BS and the $\tilde{m}$ th target, $\tilde{\boldsymbol{\varepsilon}}_{l,\tilde{m}} \in \mathbb{C}^{N_t \times 1}$ is the small-scale complex vector of the l th delay path of the $\tilde{m}$ th target, $\mathbf{a}(\theta_{\tilde{m}}) = [1 \ e^{-j\pi \sin(\theta_{\tilde{m}})} \dots e^{-j\pi(N_t-1)\sin(\theta_{\tilde{m}})}]^T$ denotes the transmit steering vector of the $\tilde{m}$ th target corresponding to direct path and $\theta_{\tilde{m}}$ is the direction of departure of the $\tilde{m}$ th target) and $\tilde{v}_{n,m}(k)$ is the AWGN with zero mean and variance σ^2 . To maximize the received power of CRs, we

let $\mathbf{w}_m(k) = \mathbf{h}_m(k) / \|\mathbf{h}_m(k)\|$. That means that when the subcarrier k is allocated to CR m , main lobe of beam on subcarrier k is directed toward CR m and the sidelobe is utilized to illuminate the targets.

We assume that R RRs can receive the echo signals from the targets. Let $\tau_{\tilde{m},r}$ denotes the propagation time of transmitted OFDM symbols from BS, reflected by the $\tilde{m}$ th target to the r th RR, i.e., $\tau_{\tilde{m},r} = \frac{1}{c} (\|\mathbf{p}_0 - \mathbf{x}_{\tilde{m}}\| + \|\mathbf{p}_r - \mathbf{x}_{\tilde{m}}\|)$, where $\mathbf{p}_0 = [p_0^x, p_0^y]^T$, $\mathbf{p}_r = [p_r^x, p_r^y]^T$, and $\mathbf{x}_{\tilde{m}} = [x_{\tilde{m}}^x, x_{\tilde{m}}^y]^T$ are locations of BS, r th RR and the $\tilde{m}$ th target, respectively, and c is the speed of light. The corresponding Doppler shift is $f_{\tilde{m},r} = \frac{\mathbf{v}_{\tilde{m}}^T (\mathbf{p}_0 - \mathbf{x}_{\tilde{m}})}{\lambda \|\mathbf{p}_0 - \mathbf{x}_{\tilde{m}}\|} + \frac{\mathbf{v}_{\tilde{m}}^T (\mathbf{p}_r - \mathbf{x}_{\tilde{m}})}{\lambda \|\mathbf{p}_r - \mathbf{x}_{\tilde{m}}\|}$, where $\mathbf{v}_{\tilde{m}} = [v_{\tilde{m}}^x, v_{\tilde{m}}^y]^T$ is the velocity of the target $\tilde{m}$.

Practically, the received echo signal at the RR r is a sum of echo signals from CRs, targets as well as surrounding scatterers. To simplify the analysis of target estimation performance, we assume that CRs and the targets are spatially separable so that echo signals from CRs have a marginal impact on target detection. Besides, the influence of clutter is incorporated into the noise term. Therefore, the estimation performance of target parameters mainly depends on following echo signals

$$\begin{aligned} y_r(t) &= \sum_{\tilde{m}=1}^{\tilde{M}} \tilde{c}_{\tilde{m}} e^{j2\pi f_{\tilde{m},r} t} \mathbf{a}^H(\theta_{\tilde{m}}) \mathbf{s}(t - \tau_{\tilde{m},r}) + w_r(t) \\ &= \sum_{\tilde{m}=1}^{\tilde{M}} \sum_{n=0}^{N_b-1} \sum_{k=0}^{N_d-1} \tilde{c}_{\tilde{m}} \mathbf{a}^H(\theta_{\tilde{m}}) \mathbf{b}_n(k) e^{j2\pi f_{\tilde{m},r} t} e^{\frac{j2\pi k(t - \tau_{\tilde{m},r})}{N_d T}} \\ &\quad \times u(t - \tau_{\tilde{m},r} - nT_s) + w_r(t) \end{aligned} \quad (4)$$

where $\tilde{c}_{\tilde{m}} = \sqrt{\frac{\lambda^2 \text{RCS}}{(4\pi)^3 d_{\tilde{m}}^2 d_{\tilde{m},r}^2}}$ is the attenuation coefficient including two-propagation loss and reflection coefficient, $d_{\tilde{m},r}$ represent the distance between the target $\tilde{m}$ and the RR r , $w(t)$ is the complex white Gaussian noise with power spectral density n_0 . For point targets, radar cross section (RCS) corresponding to different RRs can be assumed equal. Assume that $\tau_{\tilde{m},r} < N_p T$ for all RRs. The output of the matched filter corresponding to the subcarrier k in the n th OFDM symbol is

$$\begin{aligned} y_{n,r}(k) &= \frac{1}{N_d T} \int_{nT_s}^{nT_s + N_d T} y_r(t) e^{-\frac{j2\pi k t}{N_d T}} dt \\ &\approx \sum_{\tilde{m}=1}^{\tilde{M}} \tilde{c}_{\tilde{m}} \tilde{b}_{\tilde{m}}^{\tilde{m}}(k) e^{j2\pi n \tilde{f}_{\tilde{m},r}} e^{-j2\pi k \tilde{\tau}_{\tilde{m},r}} + \tilde{w}_{n,r}(k) \end{aligned} \quad (5)$$

where $\tilde{b}_{\tilde{m}}^{\tilde{m}}(k) = \mathbf{a}^H(\theta_{\tilde{m}}) \mathbf{b}_n(k)$, $\tilde{f}_{\tilde{m},r} = T_s f_{\tilde{m},r}$ and $\tilde{\tau}_{\tilde{m},r} = \frac{\tau_{\tilde{m},r}}{N_d T}$, $\tilde{w}_{n,r}(k) = \frac{1}{N_d T} \int_{nT_s}^{nT_s + N_d T} w_r(t) e^{-\frac{j2\pi k t}{N_d T}} dt$ is the AWGN with zero mean and variance $\sigma_w^2 = \frac{n_0}{N_d T}$. The approximation in (5) is because the phase rotation within one OFDM block due to the Doppler frequency can be approximated as a constant when $f_{\tilde{m},r} T_s \ll 1$.

III. PERFORMANCE ANALYSIS AND PROBLEM FORMULATION

In this section, we will analyze the communication performance and estimation performance of ICRS. Then, based on the analysis results, an optimization problem in terms of tradeoff between secrecy capacity and CRLB is formulated.

A. Secrecy Capacity of CRs

Based on (2), the channel capacity between BS and CR m on subcarrier k is

$$R_m(k) = \log_2 \left(1 + \frac{|h_m(k)|^2 \alpha_m(k) p_m(k)}{\sigma^2} \right). \quad (6)$$

Based on (3), the channel capacity between BS and the $\tilde{m}$ th target on subcarrier k is given by

$$R_m^{\tilde{m}}(k) = \log_2 \left(1 + \frac{|g_m^{\tilde{m}}(k)|^2 \alpha_m(k) p_m(k)}{\sigma^2 + |g_m^{\tilde{m}}(k)|^2 (1 - \alpha_m(k)) p_m(k)} \right). \quad (7)$$

On the k th subcarrier, to prevent information from being eavesdropped by any target, the secrecy capacity is $R_m^s(k) = [R_m(k) - R_m^e(k)]^+$ [20], $R_m^e(k) = \max_{\tilde{m}} \{R_m^{\tilde{m}}(k)\} = \log_2 \left(1 + \frac{|g_m(k)|^2 \alpha_m(k) p_m(k)}{\sigma^2 + |g_m(k)|^2 (1 - \alpha_m(k)) p_m(k)} \right)$ with $g_m(k) = \max_{\tilde{m}} \{g_m^{\tilde{m}}(k)\}$.

When BS implements OFDM transmission, the system can be regarded as the parallel of Gaussian wiretap channels [21]. Therefore, in the multiuser wiretap OFDM channels, to achieve reliable communication between BS and CRs with targets obtaining no information on all subcarriers, the instantaneous sum secrecy rate is $R_{\text{sum}}^s = \sum_{m=1}^M \kappa_m \sum_{k=0}^{N_d-1} \rho_m(k) R_m^s(k)$ [16], where κ_m is the weight for CR m . If the secrecy capacities of some communication users are small, we can increase the weights of these users to improve their secrecy capacities.

B. CRLB of OFDM Signals

CRLB provides an analytical lower bound for mean square error of parameter estimation. We suppose that the targets move slowly so that their locations hardly change during N_b symbol periods. We will calculate CRLB of location and velocity estimation of unknown targets based on $\{y_{n,r}(k), n = 1, \dots, N_b, r = 1, \dots, R, k = 1, \dots, N_d\}$.

We define the parameters to be estimated as $\mathbf{u} = [\mathbf{u}_1^T, \dots, \mathbf{u}_{\tilde{M}}^T, \mathbf{c}]^T$, where $\mathbf{u}_{\tilde{m}} = [x_{\tilde{m}}^x, x_{\tilde{m}}^y, v_{\tilde{m}}^x, v_{\tilde{m}}^y]^T$ and $\mathbf{c} = [c_1^1, c_2^1, \dots, c_R^1, \dots, c_1^{\tilde{M}}, \dots, c_R^{\tilde{M}}]^T$. According to the chain rule, the Fisher information matrix (FIM) with respect to $\mathbf{u}$ is $\mathbf{J}(\mathbf{u}) = \mathbf{Q} \mathbf{J}(\gamma) \mathbf{Q}^T$ [22], where $\gamma = [\varphi_1, \dots, \varphi_{\tilde{M}}, \mathbf{c}]$, $\varphi_{\tilde{m}} = [\tau_{\tilde{m}}, \mathbf{f}_{\tilde{m}}]$ with $\tau_{\tilde{m}} = [\tau_{\tilde{m},1}, \dots, \tau_{\tilde{m},R}]$ and $\mathbf{f}_{\tilde{m}} = [f_{\tilde{m},1}, \dots, f_{\tilde{m},R}]$. $\mathbf{Q}$ is the following Jacobian matrix:

$$\mathbf{Q} = \frac{\partial \gamma}{\partial \mathbf{u}} = \begin{bmatrix} \mathbf{Q}_1 & \dots & \mathbf{0} & \mathbf{0} \\ \vdots & \ddots & \vdots & \vdots \\ \mathbf{0} & \dots & \mathbf{Q}_{\tilde{M}} & \mathbf{0} \\ \mathbf{0} & \dots & \mathbf{0} & \mathbf{I} \end{bmatrix} \quad (8)$$

where

$$\mathbf{Q}_{\tilde{m}} = \frac{\partial \varphi_{\tilde{m}}}{\partial \mathbf{u}_{\tilde{m}}} = \begin{bmatrix} \mathbf{Q}_{11}^{\tilde{m}} & \mathbf{Q}_{12}^{\tilde{m}} \\ \mathbf{Q}_{21}^{\tilde{m}} & \mathbf{Q}_{22}^{\tilde{m}} \end{bmatrix} = \begin{bmatrix} x_{\tilde{m}}^x & f_{\tilde{m}}^x \\ x_{\tilde{m}}^y & f_{\tilde{m}}^y \\ \mathbf{0} & v_{\tilde{m}}^x \\ \mathbf{0} & v_{\tilde{m}}^y \end{bmatrix} \quad (9)$$

with $x_{\tilde{m}}^x = \frac{\partial \tau_{\tilde{m}}}{\partial x_{\tilde{m}}^x}$, $f_{\tilde{m}}^x = \frac{\partial f_{\tilde{m}}}{\partial x_{\tilde{m}}^x}$, $x_{\tilde{m}}^y = \frac{\partial \tau_{\tilde{m}}}{\partial x_{\tilde{m}}^y}$, $f_{\tilde{m}}^y = \frac{\partial f_{\tilde{m}}}{\partial x_{\tilde{m}}^y}$, $v_{\tilde{m}}^x = \frac{\partial f_{\tilde{m}}}{\partial v_{\tilde{m}}^x}$, $v_{\tilde{m}}^y = \frac{\partial f_{\tilde{m}}}{\partial v_{\tilde{m}}^y}$, and $\mathbf{I}$ is a $\tilde{M}R \times \tilde{M}R$ unit matrix.

Define $\mathbf{y} = [y_{1,1}(1), \dots, y_{n,r}(k), \dots, y_{N_b,R}(N_d)]^T$. Since $\mathbf{y}$ is a Gaussian vector with mean $\boldsymbol{\mu} = [\mu_{1,1}(1), \dots, \mu_{n,r}(k), \dots, \mu_{N_b,R}(N_d)]^T$, where $\mu_{n,r}(k) = \sum_{\tilde{m}}^{\tilde{M}} c_r^{\tilde{m}} b_n^{\tilde{m}}(k) e^{j2\pi n \tilde{f}_{\tilde{m},r}} e^{-j2\pi k \tau_{\tilde{m},r}}$, we have $\mathbf{J}(\gamma) = 2 \text{Re} \left[\frac{1}{\sigma_w^2} \frac{\partial^H \boldsymbol{\mu}}{\partial \gamma} \frac{\partial \boldsymbol{\mu}}{\partial \gamma} \right]$ [22].

By ignoring the coupling of $\{\varphi_{\tilde{m}}\}_{\tilde{m}=1}^{\tilde{M}}$ and $\mathbf{c}$, $\mathbf{J}(\gamma)$ can be written as

$$\mathbf{J}(\gamma) = \text{Re} \begin{bmatrix} \mathbf{F}_1^{(1)} & \mathbf{F}_2^{(1)} & \dots & \mathbf{F}_{\tilde{M}}^{(1)} & \mathbf{0} \\ \mathbf{F}_1^{(2)} & \mathbf{F}_2^{(2)} & \dots & \mathbf{F}_{\tilde{M}}^{(2)} & \mathbf{0} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ \mathbf{F}_1^{(\tilde{M})} & \mathbf{F}_2^{(\tilde{M})} & \dots & \mathbf{F}_{\tilde{M}}^{(\tilde{M})} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{0} & \mathbf{O} \end{bmatrix} \quad (10)$$

where $\mathbf{O}$ is an $\tilde{M} \times \tilde{M}$ matrix and it has no effect on the CRLB of the estimated parameters. In (10), the principal diagonal matrix $\mathbf{F}_{\tilde{m}}^{(\tilde{m})}$ is only related to the $\tilde{m}$ th target and has the major impact on the CRLB of $\mathbf{u}_{\tilde{m}}$. The nonprincipal diagonal matrices denote the coupling of parameters estimation for different targets. $\mathbf{F}_{\tilde{m}}^{(\tilde{m})}$ can be calculated as

$$\mathbf{F}_{\tilde{m}}^{(\tilde{m})} = \frac{2}{\sigma_w^2} \frac{\partial^H \boldsymbol{\mu}}{\partial \varphi_{\tilde{m}}} \frac{\partial \boldsymbol{\mu}}{\partial \varphi_{\tilde{m}}} = \begin{bmatrix} \mathbf{A}_{11}^{(\tilde{m},\tilde{m})} & \mathbf{A}_{12}^{(\tilde{m},\tilde{m})} \\ \mathbf{A}_{12}^{(\tilde{m},\tilde{m})} & \mathbf{A}_{22}^{(\tilde{m},\tilde{m})} \end{bmatrix} \quad (11)$$

where

$$\begin{aligned} [\mathbf{A}_{11}^{(\tilde{m},\tilde{m})}]_{r,r} &= \frac{8\pi^2}{\sigma_w^2} c_r^{\tilde{m}} c_r^{\tilde{m}'} \sum_{n=0}^{N_b-1} \sum_{k=0}^{N_d-1} k^2 b_n^{\tilde{m}}(k) (b_n^{\tilde{m}'}(k))^* \\ &\quad \times e^{j2\pi n (\tilde{f}_{\tilde{m},r} - \tilde{f}_{\tilde{m}',r})} e^{-j2\pi k (\tau_{\tilde{m},r} - \tau_{\tilde{m}',r})} \\ [\mathbf{A}_{12}^{(\tilde{m},\tilde{m})}]_{r,r} &= \frac{-8\pi^2}{\sigma_w^2} c_r^{\tilde{m}} c_r^{\tilde{m}'} \sum_{n=0}^{N_b-1} \sum_{k=0}^{N_d-1} n k b_n^{\tilde{m}}(k) (b_n^{\tilde{m}'}(k))^* \\ &\quad \times e^{j2\pi n (\tilde{f}_{\tilde{m},r} - \tilde{f}_{\tilde{m}',r})} e^{-j2\pi k (\tau_{\tilde{m},r} - \tau_{\tilde{m}',r})} \\ [\mathbf{A}_{22}^{(\tilde{m},\tilde{m})}]_{r,r} &= \frac{8\pi^2}{\sigma_w^2} c_r^{\tilde{m}} c_r^{\tilde{m}'} \sum_{n=0}^{N_b-1} \sum_{k=0}^{N_d-1} n^2 b_n^{\tilde{m}}(k) (b_n^{\tilde{m}'}(k))^* \\ &\quad \times e^{j2\pi n (\tilde{f}_{\tilde{m},r} - \tilde{f}_{\tilde{m}',r})} e^{-j2\pi k (\tau_{\tilde{m},r} - \tau_{\tilde{m}',r})} \end{aligned}$$

and $b_n^{\tilde{m}}(k) (b_n^{\tilde{m}'}(k))^* = \sum_{m=1}^M G_m^{\tilde{m}}(k) (G_m^{\tilde{m}'}(k))^* \rho_m(k) p_m(k)$ with $G_m^{\tilde{m}}(k) = \mathbf{a}^H(\theta_{\tilde{m}}) \mathbf{w}_m(k)$. Based on (8)–(11), we can obtain the following FIM for $\mathbf{u}$

$$\mathbf{J}(\mathbf{u}) = \begin{bmatrix} \mathbf{J}_1^{(1)} & \mathbf{J}_2^{(1)} & \dots & \mathbf{J}_{\tilde{M}}^{(1)} & \mathbf{0} \\ \mathbf{J}_1^{(2)} & \mathbf{J}_2^{(2)} & \dots & \mathbf{J}_{\tilde{M}}^{(2)} & \mathbf{0} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ \mathbf{J}_1^{(\tilde{M})} & \mathbf{J}_2^{(\tilde{M})} & \dots & \mathbf{J}_{\tilde{M}}^{(\tilde{M})} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \dots & \mathbf{0} & \mathbf{O} \end{bmatrix} \quad (12)$$

where $\mathbf{J}_{\tilde{m}}^{(\tilde{m})}$ is a 4×4 matrix with

$$\mathbf{J}_{\tilde{m}}^{(\tilde{m})} = \mathbf{Q}_{\tilde{m}'}^T \mathbf{F}_{\tilde{m}}^{(\tilde{m})} \mathbf{Q}_{\tilde{m}} = \begin{bmatrix} \mathbf{B}_{11}^{(\tilde{m},\tilde{m})} & \mathbf{B}_{12}^{(\tilde{m},\tilde{m})} \\ \mathbf{B}_{12}^{(\tilde{m},\tilde{m})T} & \mathbf{B}_{22}^{(\tilde{m},\tilde{m})} \end{bmatrix} \quad (13)$$

where

$$\begin{aligned}
B_{11}^{(\tilde{m}, \tilde{m}')} &= Q_{11}^{\tilde{m}'} A_{11}^{(\tilde{m}, \tilde{m}')} Q_{11}^{\tilde{m}T} + Q_{12}^{\tilde{m}'} A_{12}^{(\tilde{m}, \tilde{m}')} Q_{11}^{\tilde{m}T} \\
&\quad + Q_{11}^{\tilde{m}'} A_{12}^{(\tilde{m}, \tilde{m}')} Q_{12}^{\tilde{m}T} + Q_{12}^{\tilde{m}'} A_{22}^{(\tilde{m}, \tilde{m}')} Q_{12}^{\tilde{m}T} \\
B_{12}^{(\tilde{m}, \tilde{m}')} &= Q_{11}^{\tilde{m}'} A_{11}^{(\tilde{m}, \tilde{m}')} Q_{21}^{\tilde{m}T} + Q_{12}^{\tilde{m}'} A_{12}^{(\tilde{m}, \tilde{m}')} Q_{21}^{\tilde{m}T} \\
&\quad + Q_{11}^{\tilde{m}'} A_{12}^{(\tilde{m}, \tilde{m}')} Q_{22}^{\tilde{m}T} + Q_{12}^{\tilde{m}'} A_{22}^{(\tilde{m}, \tilde{m}')} Q_{22}^{\tilde{m}T} \\
B_{22}^{(\tilde{m}, \tilde{m}')} &= Q_{21}^{\tilde{m}'} A_{11}^{(\tilde{m}, \tilde{m}')} Q_{21}^{\tilde{m}T} + Q_{22}^{\tilde{m}'} A_{12}^{(\tilde{m}, \tilde{m}')} Q_{21}^{\tilde{m}T} \\
&\quad + Q_{21}^{\tilde{m}'} A_{12}^{(\tilde{m}, \tilde{m}')} Q_{22}^{\tilde{m}T} + Q_{22}^{\tilde{m}'} A_{22}^{(\tilde{m}, \tilde{m}')} Q_{22}^{\tilde{m}T}.
\end{aligned}$$

The CRLBs of the $\tilde{M}$ th target are determined by the diagonal elements of $\mathbf{J}^{-1}(\mathbf{u})$. However, the accurate expressions of CRLBs are too complicated to be used for the subsequent optimization. As a result, we adopt the lower bound of CRLB as the estimation performance metric. The lower bounds of CRLB of the $\tilde{m}$ th target are given by [22]

$$\begin{aligned}
\text{CRLB}(x_{\tilde{m}}^x) &\geq \frac{1}{[B_{11}^{(\tilde{m}, \tilde{m})}]_{1,1}} \\
&= \left(\sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) d_{m,\tilde{m}}^a(k) \right)^{-1} \quad (14)
\end{aligned}$$

$$\begin{aligned}
\text{CRLB}(x_{\tilde{m}}^y) &\geq \frac{1}{[B_{11}^{(\tilde{m}, \tilde{m})}]_{2,2}} \\
&= \left(\sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) d_{m,\tilde{m}}^b(k) \right)^{-1} \quad (15)
\end{aligned}$$

$$\begin{aligned}
\text{CRLB}(v_{\tilde{m}}^x) &\geq \frac{1}{[B_{22}^{(\tilde{m}, \tilde{m})}]_{1,1}} \\
&= \left(\sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) q_{m,\tilde{m}}^a(k) \right)^{-1} \quad (16)
\end{aligned}$$

$$\begin{aligned}
\text{CRLB}(v_{\tilde{m}}^y) &\geq \frac{1}{[B_{22}^{(\tilde{m}, \tilde{m})}]_{2,2}} \\
&= \left(\sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) q_{m,\tilde{m}}^b(k) \right)^{-1} \quad (17)
\end{aligned}$$

where $d_{m,\tilde{m}}^a(k) = \frac{8\pi^2}{\sigma_w^2} |G_{\tilde{m}}^{\tilde{m}}(k)|^2 \sum_{r=1}^R \sum_{n=0}^{N_b-1} (c_r^{\tilde{m}})^2 (k[\mathbf{x}_{\tilde{m}}^x]_r - n[\mathbf{f}_{\tilde{m}}^x]_r)^2$, $d_{m,\tilde{m}}^b(k) = \frac{8\pi^2}{\sigma_w^2} |G_{\tilde{m}}^{\tilde{m}}(k)|^2 \sum_{r=1}^R \sum_{n=0}^{N_b-1} (c_r^{\tilde{m}})^2 (k[\mathbf{x}_{\tilde{m}}^y]_r - n[\mathbf{f}_{\tilde{m}}^y]_r)^2$, $q_{m,\tilde{m}}^a(k) = \frac{8\pi^2}{\sigma_w^2} |G_{\tilde{m}}^{\tilde{m}}(k)|^2 \sum_{r=1}^R \sum_{n=0}^{N_b-1} (c_r^{\tilde{m}})^2 n^2 [\mathbf{v}_{\tilde{m}}^x]_r^2$ and $q_{m,\tilde{m}}^b(k) = \frac{8\pi^2}{\sigma_w^2} |G_{\tilde{m}}^{\tilde{m}}(k)|^2 \sum_{r=1}^R \sum_{n=0}^{N_b-1} (c_r^{\tilde{m}})^2 n^2 [\mathbf{v}_{\tilde{m}}^y]_r^2$.

According to the analysis results of CRLB, we can see that CRLBs of location and velocity estimation only depend on the power and subcarrier allocation, and are independent of power splitting factors.

C. Problem Formulation

Our goal is to maximize the sum secrecy rate of the CRs, under the constraints of the CRLB of the parameter estimation, by optimizing the transmit power, subcarrier allocation, and the power splitting ratio. The problem is formulated as

$$\max_{\rho, \mathbf{p}, \alpha} R_{\text{sum}}^s \quad (18)$$

$$\text{s.t.} \quad [B_{11}^{(\tilde{m}, \tilde{m})}]_{1,1}^{-1} \leq \eta_1, 1 \leq \tilde{m} \leq \tilde{M} \quad (18a)$$

$$[B_{11}^{(\tilde{m}, \tilde{m})}]_{2,2}^{-1} \leq \eta_2, 1 \leq \tilde{m} \leq \tilde{M} \quad (18b)$$

$$[B_{22}^{(\tilde{m}, \tilde{m})}]_{1,1}^{-1} \leq \eta_3, 1 \leq \tilde{m} \leq \tilde{M} \quad (18c)$$

$$[B_{22}^{(\tilde{m}, \tilde{m})}]_{2,2}^{-1} \leq \eta_4, 1 \leq \tilde{m} \leq \tilde{M} \quad (18d)$$

$$\sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) \leq P_t \quad (18e)$$

$$0 \leq p_m(k) \leq P_{\text{peak}}, 0 \leq k \leq N_d - 1, 1 \leq \tilde{m} \leq \tilde{M} \quad (18f)$$

$$0 \leq \alpha_m(k) \leq 1, 0 \leq k \leq N_d - 1, 1 \leq \tilde{m} \leq \tilde{M} \quad (18g)$$

$$\rho_m(k) \in \{0, 1\}, 0 \leq k \leq N_d - 1, 1 \leq \tilde{m} \leq \tilde{M} \quad (18h)$$

$$\sum_{m=1}^M \rho_m(k) \leq 1, 0 \leq k \leq N_d - 1 \quad (18i)$$

where $\rho = [\rho_0(1), \dots, \rho_{N_d-1}(M)]$, $\mathbf{p} = [p_0(1), \dots, p_{N_d-1}(M)]$, $\alpha = [\alpha_0(1), \dots, \alpha_{N_d-1}(M)]$, $\{\eta_i\}_{i=1}^4$ are CRLB constraints. P_t is the total power and P_{peak} denotes the peak power of each subcarrier. The constraints in (18a) and (18b) represent the requirements for location estimation accuracy of each target and ones. The constraints in (18c) and (18d) denote the requirements for velocity estimation accuracy of each target. Equations (18e) and (18f) are total transmit power and peak power constraints, respectively. Equations (18g)–(18i) are the power splitting factor and subcarrier allocation constraints.

IV. OPTIMAL SOLUTION TO PROBLEM (18)

In [23], it has been shown that the duality gap of a similar nonconvex problem to (18) converges to zero as the number of subcarriers goes infinity. Therefore, the Lagrange duality method can be applied to solve (18).

The Lagrangian function associated with the problem (18) is given by

$$\begin{aligned}
\mathcal{L}(\rho, \mathbf{p}, \alpha, \lambda) &= \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m(k) R_m^s(k) \\
&\quad - \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{1,\tilde{m}} \left(1 - \eta_1 \sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) d_{m,\tilde{m}}^a(k) \right) \\
&\quad - \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{2,\tilde{m}} \left(1 - \eta_2 \sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) d_{m,\tilde{m}}^b(k) \right)
\end{aligned}$$

$$\begin{aligned}
& - \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{3,\tilde{m}} \left(1 - \eta_3 \sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) q_{m,\tilde{m}}^a(k) \right) \\
& - \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{4,\tilde{m}} \left(1 - \eta_4 \sum_{m=1}^M \sum_{k=0}^{N_d-1} p_m(k) \rho_m(k) q_{m,\tilde{m}}^b(k) \right) \\
& - \lambda_5 \left(\sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m(k) p_m(k) - P_t \right) \quad (19)
\end{aligned}$$

where $\lambda = [\lambda_{1,\tilde{m}}, \lambda_{2,\tilde{m}}, \lambda_{3,\tilde{m}}, \lambda_{4,\tilde{m}}, \lambda_5], \forall \tilde{m} = 1, \dots, \tilde{M}$ are Lagrange multipliers, and the corresponding dual function is

$$\begin{aligned}
g(\lambda) &= \max_{\rho, \mathbf{p}, \alpha} \mathcal{L}(\rho, \mathbf{p}, \alpha, \lambda) \\
\text{s.t.} \quad & (18\text{f}) - (18\text{i}). \quad (20)
\end{aligned}$$

Based on [24], the optimal λ which makes the duality gap vanish can be obtained by solving the following convex problem:

$$\min_{\lambda \geq 0} g(\lambda). \quad (21)$$

The optimal λ [solution of (21)] and corresponding optimal $(\rho, \mathbf{p}, \alpha)$ [solution of (20)] can be obtained by dual optimization method in [23], which iteratively updating $(\rho, \mathbf{p}, \alpha)$ and λ until convergence. Specifically, for given λ^t (t is the number of iteration), we solve (20) to get $(\rho, \mathbf{p}, \alpha)$, and then λ^t is updated based on following subgradient method:

$$\begin{aligned}
\lambda_{1,\tilde{m}}^{t+1} &= \left[\lambda_{1,\tilde{m}}^t - \xi_{1,\tilde{m}}^t \left(\eta_1 \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m^*(k) p_m^*(k) d_{m,\tilde{m}}^a(k) - 1 \right) \right]^+ \quad (22)
\end{aligned}$$

$$\begin{aligned}
\lambda_{2,\tilde{m}}^{t+1} &= \left[\lambda_{2,\tilde{m}}^t - \xi_{2,\tilde{m}}^t \left(\eta_2 \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m^*(k) p_m^*(k) d_{m,\tilde{m}}^b(k) - 1 \right) \right]^+ \quad (23)
\end{aligned}$$

$$\begin{aligned}
\lambda_{3,\tilde{m}}^{t+1} &= \left[\lambda_{3,\tilde{m}}^t - \xi_{3,\tilde{m}}^t \left(\eta_3 \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m^*(k) p_m^*(k) q_{m,\tilde{m}}^a(k) - 1 \right) \right]^+ \quad (24)
\end{aligned}$$

$$\begin{aligned}
\lambda_{4,\tilde{m}}^{t+1} &= \left[\lambda_{4,\tilde{m}}^t - \xi_{4,\tilde{m}}^t \left(\eta_4 \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m^*(k) p_m^*(k) q_{m,\tilde{m}}^b(k) - 1 \right) \right]^+ \quad (25)
\end{aligned}$$

$$\begin{aligned}
\lambda_5^{t+1} &= \left[\lambda_5^t - \xi_5^t \left(P_{\max} - \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m^*(k) p_m^*(k) \right) \right]^+ \quad (26)
\end{aligned}$$

where $t \geq 0$ is the iteration index, $\rho_m^*(k)$ and $(p_m^*(k), \alpha_m^*(k))$ are determined based on (28) and (38) for given $\lambda^t, \xi = [\xi_{1,\tilde{m}}^t, \xi_{2,\tilde{m}}^t, \xi_{3,\tilde{m}}^t, \xi_{4,\tilde{m}}^t, \xi_5^t], \tilde{m} = 1, \dots, \tilde{M}$ are positive step-sizes.

To solve (20), the Lagrangian function in (19) is reorganized as

$$\mathcal{L}(\rho, \mathbf{p}, \alpha, \lambda) = \sum_{m=1}^M \sum_{k=0}^{N_d-1} \rho_m(k) \mathcal{L}'_m(k) - \sum_{i=1}^4 \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{i,\tilde{m}} + \lambda_5 P_t \quad (27)$$

where $\mathcal{L}'_m(k) = R_m^s(k) + p_m(k) a(\lambda, k)$ and $a(\lambda, k) = \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{1,\tilde{m}} \eta_1 d_{m,\tilde{m}}^a(k) + \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{2,\tilde{m}} \eta_2 d_{m,\tilde{m}}^b(k) + \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{3,\tilde{m}} \eta_3 q_{m,\tilde{m}}^a(k) + \sum_{\tilde{m}=1}^{\tilde{M}} \lambda_{4,\tilde{m}} \eta_4 q_{m,\tilde{m}}^b(k) - \lambda_5$. Since each subcarrier is only allocated to one CR, in order to maximize $\mathcal{L}(\rho, \mathbf{p}, \alpha, \lambda)$ for given λ , the k th subcarrier should be allocated to the CR with $\max_{m \in \mathcal{M}} \mathcal{L}'_m(k)$. That is

$$\rho_m^*(k) = \begin{cases} 1, & \text{if } m = \arg \max_{m \in \mathcal{M}} \mathcal{L}'_m(k) \\ 0, & \text{otherwise.} \end{cases} \quad (28)$$

To find m which maximizes $\mathcal{L}'_m(k)$ for a given k , we need solve the following M subproblems for each subcarrier

$$\begin{aligned}
& \max_{p_m(k), \alpha_m(k)} \mathcal{L}'_m(k) \\
& \text{s.t.} \quad (18\text{f}) - (18\text{g}). \quad (29)
\end{aligned}$$

By analyzing the expression of $R_m^s(k)$, we can draw that when $(1 - \alpha_m(k))p_m(k) \leq \sigma^2 \left(\frac{1}{|h_m(k)|^2} - \frac{1}{|g_m(k)|^2} \right)$, $R_m^s(k) = 0$, otherwise, $R_m^s(k) > 0$. Considering the influence of the relative relationship between $|h_m(k)|^2$ and $|g_m(k)|^2$ on the solution, we study the following two situations, respectively.

1) Situation I: $|h_m(k)|^2 < |g_m(k)|^2$

Case 1: When $\alpha_m(k) = 1$ or $\alpha_m(k) = 0$ or $(0 < \alpha_m(k) < 1, p_m(k) \leq \mathcal{X}_m^k(\alpha_m(k))$ where $\mathcal{X}_m^k(\alpha_m(k)) = \frac{\sigma^2}{1 - \alpha_m(k)} \left(\frac{1}{|h_m(k)|^2} - \frac{1}{|g_m(k)|^2} \right)$, $R_m(k) \leq R_m^e(k)$ so that $R_m^s(k) = 0$ and $\mathcal{L}'_m(k)$ in (29) can be written as $\mathcal{L}'_m(k) = p_m(k) a(\lambda, k)$.

Because $\mathcal{L}'_m(k)$ is a linear function of $p_m(k)$, when $\alpha_m(k) = 1$ or 0, the optimal value of $p_m(k)$ is 0 when $a(\lambda, k) < 0$, or P_{peak} when $a(\lambda, k) \geq 0$. When $0 < \alpha_m(k) < 1$, the optimal value of $p_m(k)$ is $\min\{P_{\text{peak}}, \mathcal{X}_m^k(\alpha_m(k))\}$. Obviously, when $\alpha_m(k) \rightarrow 1$, $\mathcal{X}_m^k(\alpha_m(k)) > P_{\text{peak}}$ and the optimal value of $p_m(k)$ is P_{peak} . Since whether $\alpha_m(k)$ is equal to 0 or 1, the optimal value of $\mathcal{L}'_m(k)$ is the same, we choose the solution with $\alpha_m(k) = 0$. Because when secure transmission cannot be achieved, a better way is to allocate all power to AN. Therefore, in this case, a set of the optimal value of $(p_m(k), \alpha_m(k))$ is

$$\Psi_1 \triangleq \begin{cases} (0, 0), & \text{if } a(\lambda, k) < 0 \\ (P_{\text{peak}}, 0), & \text{if } a(\lambda, k) \geq 0. \end{cases} \quad (30)$$

Case 2: When $\alpha_m(k) \neq 1$ and $p_m(k) > \mathcal{X}_m^k(\alpha_m(k))$, we have $R_m^s(k) = R_m(k) - R_m^e(k)$. Notice that $\tilde{p}_m(k) > \mathcal{X}_m^k(\tilde{\alpha}_m(k))$ implies $P_{\text{peak}} > \mathcal{X}_m^k(0)$. In this case

$$\mathcal{L}'_m(k) = R_m(k) - R_m^e(k) + a(\lambda, k) p_m(k). \quad (31)$$

By letting $\frac{\partial \mathcal{L}'_m(k)}{\partial \alpha_m(k)} = 0$ and $\frac{\partial \mathcal{L}'_m(k)}{\partial p_m(k)} = 0$, respectively, we have

$$\tilde{\alpha}_m(k) = \frac{1}{2} \left(1 - \frac{\mathcal{X}_m^k(0)}{\tilde{p}_m(k)} \right) \quad (32)$$

and

$$a_1\tilde{p}_m^3(k) + b_1\tilde{p}_m^2(k) + c_1\tilde{p}_m(k) + d_1 = 0 \quad (33)$$

where $a_1 = \ln 2 |h_m(k)|^2 |g_m(k)|^4 \tilde{\alpha}_m(k) (1 - \tilde{\alpha}_m(k)) a(\lambda, k)$, $b_1 = \tilde{\alpha}_m(k) (1 - \tilde{\alpha}_m(k)) |h_m(k)|^2 |g_m(k)|^4 + \ln 2 |g_m(k)|^2 \sigma^2 ((1 - \tilde{\alpha}_m(k)) |g_m(k)|^2 + \tilde{\alpha}_m(k) (2 - \tilde{\alpha}_m(k)) |h_m(k)|^2) a(\lambda, k)$, $c_1 = (\ln 2) \sigma^4 ((2 - \tilde{\alpha}_m(k)) |g_m(k)|^2 + \tilde{\alpha}_m(k) |h_m(k)|^2) a(\lambda, k) + 2\sigma^2 \tilde{\alpha}_m(k) (1 - \tilde{\alpha}_m(k)) |g_m(k)|^2 |h_m(k)|^2$, and $d_1 = \sigma^4 \tilde{\alpha}_m(k) (|h_m(k)|^2 - |g_m(k)|^2) + (\ln 2) \sigma^6 a(\lambda, k)$.

We can get all $(\tilde{p}_m(k), \tilde{\alpha}_m(k))$ by jointly solving (32) and (33). Note that $(\tilde{p}_m(k), \tilde{\alpha}_m(k))$ satisfying (32) and (33) are not necessarily the optimal solution since the optimal solution should satisfy $(1 - \tilde{\alpha}_m(k)) \tilde{p}_m(k) > \mathcal{X}_m^k(0)$, $0 \leq \tilde{\alpha}_m(k) < 1$ and $0 \leq \tilde{p}_m(k) \leq P_{\text{peak}}$.

When $\tilde{p}_m(k) \leq \mathcal{X}_m^k(0)$, we can infer $\tilde{\alpha}_m(k) \leq 0$ and $(1 - \tilde{\alpha}_m(k)) \tilde{p}_m(k) \leq \mathcal{X}_m^k(0)$. Therefore, $R_m^s(k) = 0$ and the optimal solution is the same as (30).

When $\mathcal{X}_m^k(0) < \tilde{p}_m(k) < P_{\text{peak}}$, we have $0 < \tilde{\alpha}_m(k) < 1$ based on (32) ($\tilde{\alpha}_m(k)$ must be smaller than 1 when $|h_m(k)|^2 < |g_m(k)|^2$). Therefore, the optimal solution is $(\tilde{p}_m(k), \tilde{\alpha}_m(k))$. We define Φ_1 as the set of all real roots of (33) that satisfy $\mathcal{X}_m^k(0) < \tilde{p}_m(k) < P_{\text{peak}}$.

When $\tilde{p}_m(k) \geq P_{\text{peak}}$, the optimal solution may be the boundary point $(P_{\text{peak}}, \frac{1}{2}(1 - \frac{\mathcal{X}_m^k(0)}{P_{\text{peak}}}))$. In this case, all possible solutions are summarized as

$$\Psi_2 \triangleq \{(p_m(k), \alpha_m(k)) \mid (p_m(k) \in \Phi_1, \frac{1}{2} \left(1 - \frac{\mathcal{X}_m^k(0)}{p_m(k)}\right)) \cup \left(P_{\text{peak}}, \frac{1}{2} \left(1 - \frac{\mathcal{X}_m^k(0)}{P_{\text{peak}}}\right)\right)\} \quad (34)$$

2) Situation II: $|h_m(k)|^2 \geq |g_m(k)|^2$

In this situation, $R_m^s(k)$ can be expressed as $R_m^s(k) = R_m(k) - R_m^e(k)$, so that the expression of $\mathcal{L}_m^s(k)$ is the same as (31). Therefore, we can obtain all $(\tilde{p}_m(k), \tilde{\alpha}_m(k))$ s by jointly solving (32) and (33). Obviously, $\tilde{\alpha}_m(k) > 0$ when $\tilde{p}_m(k) > 0$. In order to limit $\tilde{\alpha}_m(k) \leq 1$, $\tilde{p}_m(k)$ should meet $\tilde{p}_m(k) \geq -\mathcal{X}_m^k(0)$.

Case 1: Suppose that $-\mathcal{X}_m^k(0) < P_{\text{peak}}$. When $-\mathcal{X}_m^k(0) \leq \tilde{p}_m(k) < P_{\text{peak}}$, we have $0 < \tilde{\alpha}_m(k) \leq 1$. Therefore, the optimal solution is $(\tilde{p}_m(k), \tilde{\alpha}_m(k))$. We define Φ_2 as the set of all real roots of (33), which satisfy $-\mathcal{X}_m^k(0) \leq \tilde{p}_m(k) < P_{\text{peak}}$. When $\tilde{p}_m(k) > P_{\text{peak}}$, the possible optimal solution is the boundary point $(P_{\text{peak}}, \frac{1}{2}(1 - \frac{\mathcal{X}_m^k(0)}{P_{\text{peak}}}))$.

Therefore, when $\tilde{p}_m(k) \geq -\mathcal{X}_m^k(0)$, possible solutions are summarized as

$$\Psi_3 \triangleq \{(p_m(k), \alpha_m(k)) \mid (p_m(k) \in \Phi_2, \frac{1}{2} \left(1 - \frac{\mathcal{X}_m^k(0)}{p_m(k)}\right)) \cup \left(P_{\text{peak}}, \frac{1}{2} \left(1 - \frac{\mathcal{X}_m^k(0)}{P_{\text{peak}}}\right)\right)\} \quad (35)$$

When $\tilde{p}_m(k) < -\mathcal{X}_m^k(0)$, we have $\tilde{\alpha}_m(k) > 1$. In this case, we consider the boundary point with $\alpha_m(k) = 1$. When $\alpha_m(k) = 1$, corresponding optimal $p_m(k)$ is the solution of (33) with $\tilde{\alpha}_m(k) = 1$. Let Φ_3 denote the set of all real roots of (33) that satisfy $0 \leq \tilde{p}_m(k) \leq P_{\text{peak}}$. All possible optimal solutions

TABLE I
SIMULATION PARAMETERS

Parameter	Quantity	Parameter	Quantity	Parameter	Quantity
M	3, 5, 7	N_d	48, 64, 128	N_t	6
R	4, 5, 6, 8	N_p	6	n_0	10^{-21}
f_c	2 GHz	RCS	0.1	N_b	32
$\frac{1}{N_d T}$	15 KHz	L	6	P_t	3–14 W

are summarized as

$$\Psi_4 \triangleq \{(p_m(k), \alpha_m(k)) \mid (p_m(k) \in \Phi_3, 1) \cup (P_{\text{peak}}, 1) \cup (0, 1)\} \quad (36)$$

Case 2: Suppose that $-\mathcal{X}_m^k(0) > P_{\text{peak}}$. When $\tilde{p}_m(k) < -\mathcal{X}_m^k(0)$, $\tilde{\alpha}_m(k) > 1$. Therefore, $(\tilde{p}_m(k), \tilde{\alpha}_m(k))$ satisfying the constraints does not exist. We just need to consider the boundary point with $\alpha_m(k) = 1$ (the point with $\alpha_m(k) = 0$ corresponds to smaller $\mathcal{L}_m^s(k)$). Thus, the optimal solution is the same as Ψ_4 .

By taking together of situation I and situation II, we define $\mathcal{F}$ as a set of candidates of $(p_m(k), \alpha_m(k))$

$$\mathcal{F} = \begin{cases} \Psi_1 \cup \Psi_2, & \text{if } P_{\text{peak}} > \mathcal{X}_m^k(0) \\ & \text{and } |h_m(k)|^2 < |g_m(k)|^2 \\ \Psi_1, & \text{if } P_{\text{peak}} \leq \mathcal{X}_m^k(0) \\ & \text{and } |h_m(k)|^2 < |g_m(k)|^2 \\ \Psi_3 \cup \Psi_4, & \text{if } P_{\text{peak}} > -\mathcal{X}_m^k(0) \\ & \text{and } |h_m(k)|^2 \geq |g_m(k)|^2 \\ \Psi_4, & \text{if } P_{\text{peak}} \leq -\mathcal{X}_m^k(0) \\ & \text{and } |h_m(k)|^2 \geq |g_m(k)|^2. \end{cases} \quad (37)$$

Then, we can select the optimal $(p_m^*(k), \alpha_m^*(k))$ according to

$$(p_m^*(k), \alpha_m^*(k)) = \arg \max_{(p_m(k), \alpha_m(k)) \in \mathcal{F}} \mathcal{L}_m^s(k). \quad (38)$$

The subgradient updates in (22)–(26) are converged to the optimal λ as long as ξ is chosen to be sufficiently small [25]. A common criterion for choosing the step sizes is that ξ must be square summable [23], but not absolute summable. When the norm of the subgradient is bounded, the following choice: $\xi(i) = \beta/\sqrt{i}$, $i = 1, \dots, 5$ for some constant β is guaranteed to converge to the optimal λ .

The complexity of this iterative algorithm is analyzed as follows. For given Lagrange multipliers λ^t , the problem in (20) is decomposed into MN_d subproblems, as in (29). For each subproblem, $\mathcal{O}(4\tilde{M} + 1)$ computations are needed for solving $a(\lambda, k)$ in (33). Therefore, the complexity is $\mathcal{O}((4\tilde{M} + 1)MN_d)$ for solving (20). Last, the complexity of updating λ based on subgradient method is polynomial in the number of dual variables $4\tilde{M} + 1$. As a result, the overall complexity of the proposed algorithm for solving problem (18) is $\mathcal{O}((4\tilde{M} + 1)^{(q+1)}MN_d)$, where q is a positive constant. Note that the complexity is polynomial. To sum up, the algorithm to solve (18) is summarized as Algorithm 1.

V. SIMULATION RESULTS

In this section, we show some numerical results to evaluate the performance of the proposed algorithm. The parameters for the simulations are provided in Table I. In the system configuration, the relative positions of the BS, CRs, targets, and RRs are

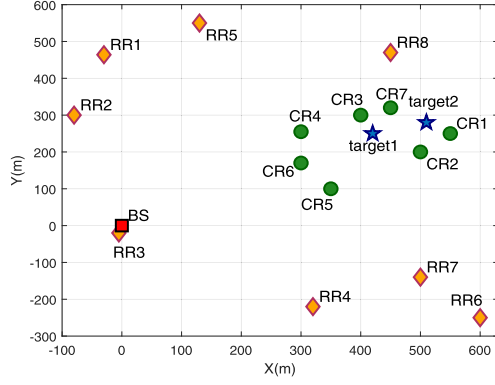


Fig. 2. Relative positions of the BS, CRs, targets, and RRs.

Algorithm 1: Optimal Algorithm for Problem (18).

```

1: initial  $\lambda, \xi, \theta_i$ ;
2: while  $|\lambda_{i,\tilde{m}}^{t+1} - \lambda_{i,\tilde{m}}^t| < \theta_i, i = 1, \dots, 5, \tilde{m} = 1, \dots, \tilde{M}$  do
3:   if  $(|h_m(k)|^2 < |g_m(k)|^2)$ 
4:     if  $(P_{\text{peak}} > \mathcal{X}_m^k(0))$   $\mathcal{F} = \Psi_1 \cup \Psi_2$ 
5:   else  $\mathcal{F} = \Psi_1$ 
6:   end
7: end
8: if  $(|h_m(k)|^2 \geq |g_m(k)|^2)$ 
9:   if  $(P_{\text{peak}} > -\mathcal{X}_m^k(0))$   $\mathcal{F} = \Psi_3 \cup \Psi_4$ 
10:  else  $\mathcal{F} = \Psi_4$ 
11:  end
12: end
13: Compute  $(p_m^*(k), \alpha_m^*(k))$  and  $\rho_m^*(k)$  by (38) and (28)
14: Update  $\lambda$  according to (22)–(26).
15: end while

```

presented in Fig. 2. BS is located at $(0, 0)m$, the location of the target1 and target2 are $(420, 250)m$ and $(510, 280)m$, respectively. The velocity of the two targets are both $(15, 20)m/s$. The location of CRs are $[(550, 250), (500, 200), (400, 300), (300, 255), (300, 100), (300, 170), (450, 320)]m$. The Locations of RRs are $[(-30, 464), (-80, 300), (-5, -20), (320, -220), (130, 550), (600, -250), (500, -140), (450, 470)]m$. In the channel model, elements in $\varepsilon_{l,m}$ and $\tilde{\varepsilon}_{l,\tilde{m}}$ are independent identically distributed Gaussian variables with zero mean and variance $\sigma_{\varepsilon_{l,m}}^2$ and $\sigma_{\tilde{\varepsilon}_{l,\tilde{m}}}^2$. For different multipath components, we set $\sum_{l=1}^L \sigma_{\varepsilon_{l,m}}^2 = 0.1$ and $\sum_{l=1}^L \sigma_{\tilde{\varepsilon}_{l,\tilde{m}}}^2 = 0.1$. The weight for each CR is $\kappa_m = 1, \forall m$. The peak power on each subcarrier is set to $P_{\text{peak}} = P_t$.

A. CRLB Performance With Opportunistic Detection in Single-Target Scenario

In this section, we evaluate CRLB performance of opportunistic detection in multiuser single-target OFDM systems. Opportunistic detection is similar to the concept of passive radar. In this case, transmit parameters are optimized by solving (18) without considering CRLB constraints in (18a)–(18d). Then, CRLBs of target parameters and lower bounds of CRLBs in Fig. 3 are calculated based on (12) and (14)–(17). We will show the detection ability of communication BS when the target is illuminated by communication signals unintentionally. In Fig. 3(a)–(c), the vertical coordinate is the average CRLB of

target location estimation. In Fig. 3(d)–(f), the ordinate is the average CRLB of target velocity estimation. In Fig. 3(a), (b), (d) and (e), we set $M = 3$ and CR1–CR3 in Fig. 2 are chosen.

In Fig. 3(a) and (d), we demonstrate the relationship between CRLB of location and velocity estimation and total transmit power with varying number of RRs. We set $N_d = 64$. When $R = 4$, RR1–RR4 are employed, and when $R = 6$, RR1–RR6 are employed. Obviously, CRLBs of location and velocity will gradually decrease with the increase of total transmit power, and the change of CRLB tends to slow down with the increase of P_t . Multireceiver cooperation can take advantage of receive diversity to improve estimation performance. Therefore, we can see that increasing R can significantly improve the estimation performance. Besides, we compare the lower bounds of CRLB in (14)–(17) with the exact CRLBs. It can be observed that the gap between lower bound and exact CRLB decreases with the increase of R or P_t , especially in the high SNR region, the gap will approach zero. That is because the Fisher matrix tends to diagonally dominant in the high SNR region.

In Fig. 3(b) and (e), we show the effect of number of subcarrier N_d on the CRLB of location and velocity estimation under different power constraints. We set $R = 6$ and $N_d = 48, 64, 128$, respectively. We keep the subcarrier bandwidth unchanged, so the signal bandwidth will increase with the increase of N_d . From Fig. 3(b), we can see that increasing bandwidth helps to improve the performance of location estimation, and the gap between lower bounds and exact CRLB also shrinks with the increase of N_d . Meanwhile, from Fig. 3(e), we can see that increasing bandwidth will result in a slight performance loss in velocity estimation. That is because for fixed subcarrier bandwidth, increasing N_d will decrease the symbol period T_s of OFDM. From (16) and (17), we can see that CRLB of velocity estimation is inversely proportional to T_s . Therefore, unlike location estimation, which benefits from large bandwidth, velocity estimation benefits from large symbol period.

Fig. 3(c) and (f) illustrates the impact of the number of CRs on CRLB of location and velocity estimation, respectively. We set $R = 6, N_d = 64$. When $M = 3$, we choose CR1–CR3 and when $M = 5$, we choose CR1–CR5. In passive detection, the estimation performance of target parameters is related to the positions of CRs. The closer the CRs are to the target, the stronger the signal power irradiated to the target. Therefore, in the studied ICRSs, when the target approaches the CRs for eavesdropping, it will be detected by signals sent by the BS.

B. Tradeoff Between Communication and Radar Performance in Single-Target Scenario

In this section, we present the tradeoff between average secrecy capacity and average CRLBs of location and velocity by optimizing transmit parameters in single-target scenario. The secrecy capacity in terms of bps/Hz is defined as R_{sum}^s/N_d . We set $P_t = 10$ W and $M = 3$, where CR1–CR3 are chosen. To determine the value range of $\{\eta_i\}_{i=1}^4$ for each channel realization, we first optimize (ρ, p, α) by solving (18) without (18a)–(18d), and then calculate the lower bounds of CRLBs using (14)–(17) with optimized parameters. CRLBs calculated by this method represent the estimation performance in opportunistic detection. When we set $\{\eta_i\}_{i=1}^4$ in (18) to be lower bounds of CRLB in opportunistic detection, the problem in (18) must be feasible. To evaluate the average performance, we set the value of $\{\eta_i\}_{i=1}^4$ to the maximum value among all channel realizations, and

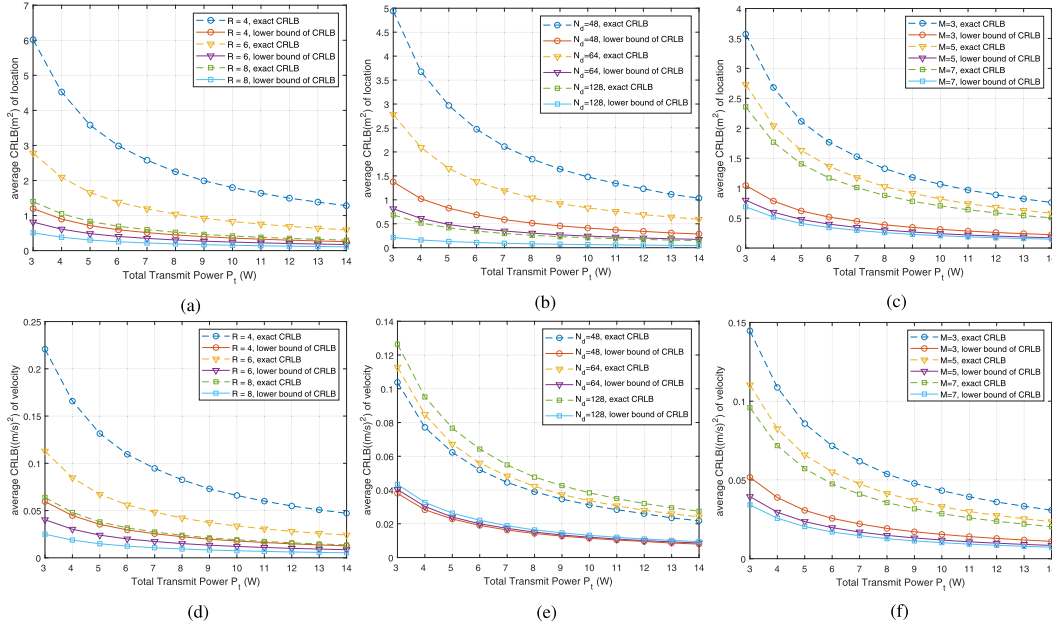


Fig. 3. Average CRLB of (a)–(c) location and (d)–(f) velocity estimation versus total transmit power with different number of RRs, different number of subcarriers and different number of CRs. (a) Different numbers of RRs. (b) Different number of subcarriers. (c) Different number of CRs. (d) Different numbers of RRs. (e) Different numbers of subcarriers. (f) Different number of CRs.

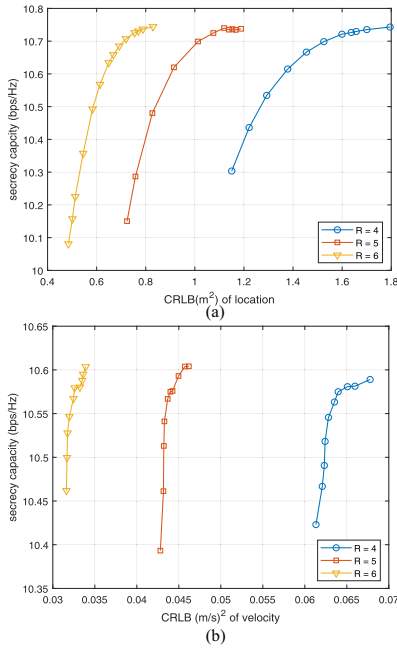


Fig. 4. Tradeoff between secrecy capacity and CRLB with different number of RRs.

then gradually reduce $\{\eta_i\}_{i=1}^4$ until there is no solution to the optimization problem for some channel realization.

In Figs. 4(a) and 5(a), η_3 and η_4 in (18) are fixed to the maximum values of η_3 and η_4 among all channel realizations, and η_1 and η_2 are gradually reduced from the maximum values. For example, in Fig. 4(a), when $R = 6$, $\eta_3 = 0.0068$ and $\eta_4 = 0.0173$. The abscissa is average accurate CRLB of location estimation calculated with the optimized transmit parameters

based on (12). In Figs. 4(b) and 5(b), the abscissa is average CRLB of velocity estimation with η_1 and η_2 fixed. It is analytically obtained that the CRLBs will no longer change with the increase of $\{\eta_i\}_{i=1}^2$ or $\{\eta_i\}_{i=3}^4$ when the CRLBs reach the upper bounds. Take Fig. 4(a) as an example. When $R = 6$, the maximum CRLB is about 0.81. This upper bound is equivalent to CRLB of opportunistic detection when CRLB constraints in (18a) and (18b) are not considered. There also exist lower bounds of the CRLBs, and the optimization problem in (18) is infeasible when $\{\eta_i\}_{i=1}^2$ or $\{\eta_i\}_{i=3}^4$ decreases to a certain value.

In Fig. 4, we show the tradeoff performance between the CRLB and the secrecy capacity with varying number of RRs. The number of subcarriers is set to $N_d = 64$. When $R = 4$, RR1–RR4 are employed. And RR1–RR5 are employed when $R = 5$. It can be seen that secrecy capacity and estimation performance are two conflicting metrics. Improving the secrecy capacity will weaken the radar detection performance to some extent. In this case, increasing the number of RRs is an effective method to improve the tradeoff performance between communication and detection. As shown in Fig. 4, for a given secrecy capacity, increasing the number of RRs can effectively improve the detection performance. By comparing Fig. 4(a) and (b), we can see that changing CRLB constraints of location estimation has a greater impact on the communication performance than varying the CRLB constraints of velocity estimation. Besides, the subcarrier/power allocation and AN power splitting ratio scheme under a certain channel is visualized in Fig. 6 when $M = 3$, $N_d = 64$, $R = 6$. In Fig. 6(a), CRLB of location estimation is at its worst 0.8295. It can be seen that 21.87%, 35.94%, and 42.19% subcarriers are allocated to CR1–CR3, respectively, and nearly half of the power on each subcarrier is allocated to AN to maximize the total security capacity. Fig. 6(b) shows that when CRLB decreases to 0.4849, more subcarriers and power are allocated to CR3. This is because the target is

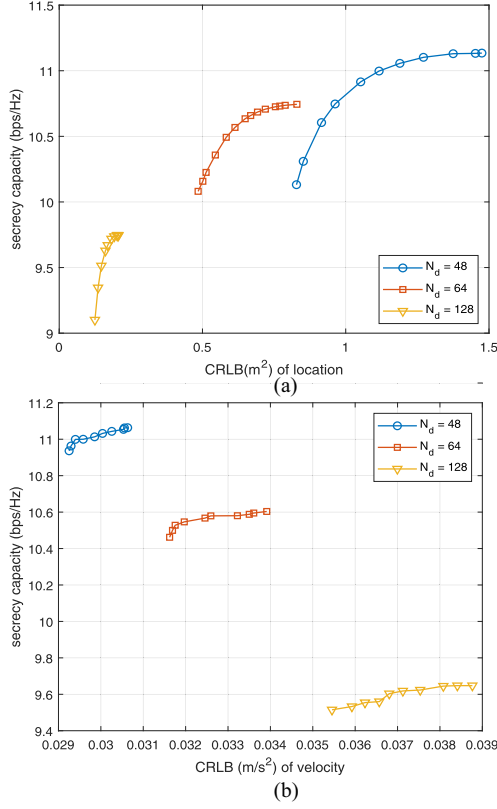


Fig. 5. Tradeoff between secrecy capacity and CRLB with different number of subcarriers.

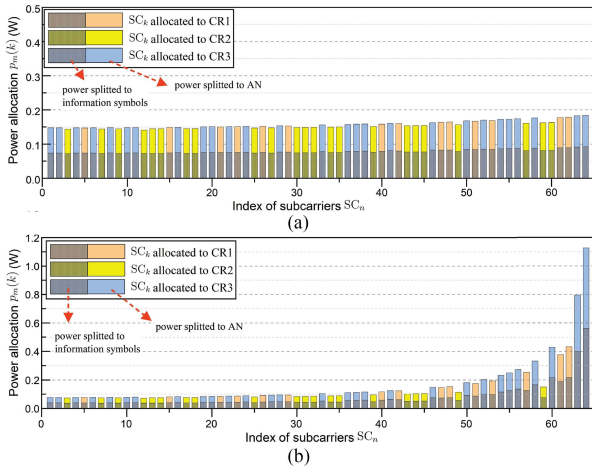


Fig. 6. Subcarrier/power allocation and power splitting ratio scheme when $M = 3$, $N_d = 64$, and $R = 6$.

closest to CR3 and allocating more resources to CR3 will help to enhance the signal energy radiated on the target.

Fig. 5 demonstrates the tradeoff performance under different number of subcarriers N_d . In this case, we set $R = 6$ and $N_d = 48, 64, 128$, respectively. In Fig. 5(a), it can be observed that increasing N_d will decrease CRLB of location estimation, but also reduce the secrecy capacity. This is because the secrecy capacity is equal to R_{sum}^s/N_d . Although increasing bandwidth leads to an increase in R_{sum}^s , the secrecy capacity drops slightly with the increase of N_d . In Fig. 5(b), both the secrecy capacity

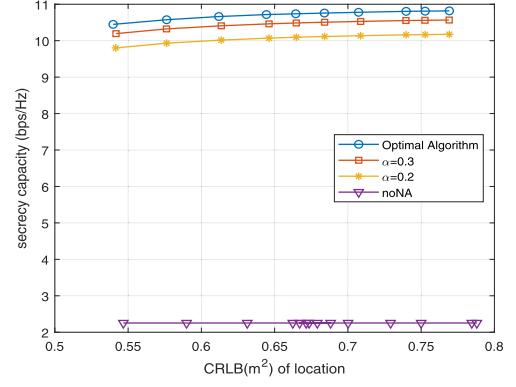


Fig. 7. Tradeoff between secrecy capacity and CRLB of location with different power splitting ratios.

and CRLB of velocity estimation suffer from slight performance loss as N_d increases, which is consistent with the results in the previous section. Besides, the CRLB constraints of location estimation have a more obvious impact on the secrecy capacity compared with the velocity.

Fig. 7 illustrates the tradeoff performance between secrecy capacity and CRLBs of location with different power splitting ratio $\alpha_m(k)$ s. In this case, we set $R = 6$, $N_d = 64$. We compare four schemes with different $\alpha_m(k)$ s. In the first scheme, $\alpha_m(k)$ s are obtained by Lagrange duality method in Algorithm 1. In the second and third schemes, we fix $\alpha_m(k) = 0.2$ and $\alpha_m(k) = 0.3$ for any m and k , respectively, while the $p_m(k)$ and $\rho_m(k)$ are still optimized as in Algorithm 1. The fourth one is the scheme without AN (noAN), i.e., $\alpha_m(k) = 1$. As suggested in the analytical result in (14)–(17), CRLBs of location estimation only depend on the power and subcarrier allocation and are independent of $\alpha_m(k)$ s. Therefore, when $\alpha_m(k)$ s take different values, the value ranges of CRLB are basically the same. However, as shown in Fig. 7, optimizing $\alpha_m(k)$ s can improve the secrecy capacity significantly for given CRLB constraints.

C. Tradeoff Between Communication and Radar Performance in Multitarget Scenario

In this section, we give some simulation results in multiple targets scenario. The parameters for the simulations are set to $N_d = 64$, $M = 3$, $P_t = 10$ W, and $\tilde{M} = 2$.

In Fig. 8, we evaluate CRLB performance of opportunistic detection in multiuser multitarget OFDM systems with $R = 4, 6$. We can see that in our settings the parameters estimation performance of target 1 is better than that of target 2. As in the single-target scenario, CRLBs of two targets location estimation gradually decrease and the change of CRLBs tends to slow down with the increase of P_t . In Fig. 8, the curves with “target1, lower bound of CRLB” are plotted by calculating diagonal elements of $[\mathbf{J}_1^{(1)}]^{-1}$ in (13) and those with “target1, exact CRLB” are plotted by calculating diagonal elements of $[\mathbf{J}(u)]^{-1}$ in (12). The curves with “target1, lower bound of CRLB” represent the estimation performance of single target when the interference between targets is ignored. When targets are far separated, interference between targets can be ignored. However, when target 2 is close to target 1, the interference will degrade the estimation performance, as shown in Fig. 8. Meanwhile, it can be observed

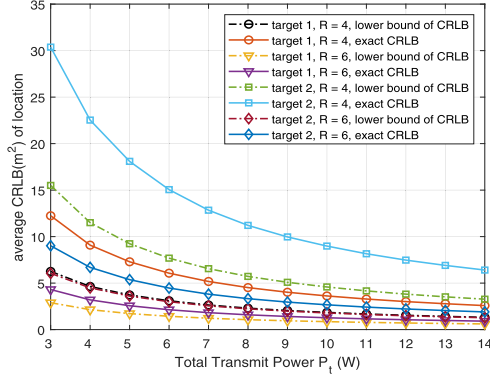


Fig. 8. Average CRLB of location estimation versus total transmit power in multitarget scenario.

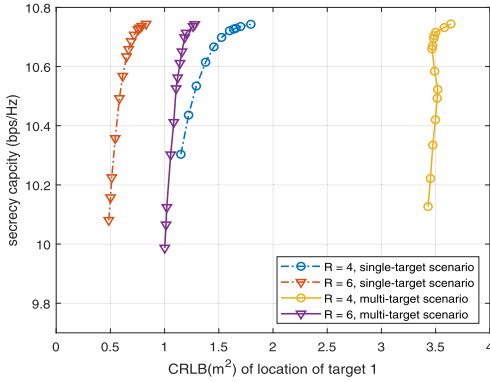


Fig. 9. Tradeoff between secrecy capacity and CRLB of location in single-target and multitarget scenarios.

that the gap between lower bound and exact CRLB decreases with the increase of P_t , which denotes that increasing the total transmit power can reduce the adverse effects of intertarget interference.

In Fig. 9, we present the tradeoff between secrecy capacity and CRLB of target 1's location estimation. Compared with the single-target scenario, the secrecy channel capacity between BS and CRs decreases slightly and the estimation performance of target 1 becomes worse. This is because the more targets, the lower secrecy channel capacity, which prevents information leakage to any target. Moreover, the interference between the two targets parameters estimation leads to an increase in the CRLB.

D. Comparison Schemes

In this section, we will compare the proposed method with those in [19] and [26], which also investigated the tradeoff between secrecy capacity and detection performance by optimizing the transmit parameters. Since the authors in [19] and [26] did not consider OFDM systems, we extend the schemes in [19] and [26] to OFDM systems in order to make a fair comparison with our scheme. In comparison scheme 1, we assign each subcarrier to all communication users and then optimize the beamforming vector and the AN covariance matrix on each carrier according to the method in [19]. The radar performance metric is the desired beam pattern. In comparison scheme 2, we use the same system model as in scheme 1, but replace the radar performance

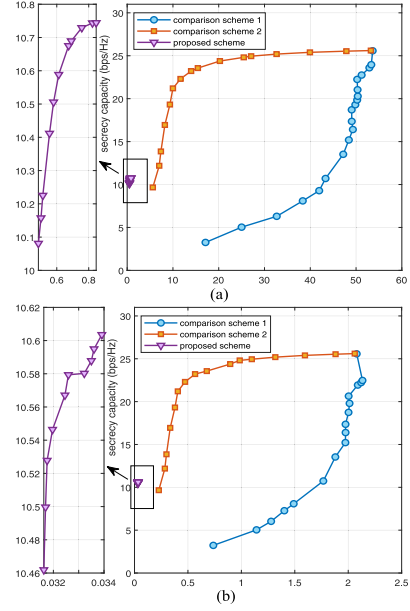


Fig. 10. Tradeoff between the secrecy capacity and CRLBs in different schemes. ($P_t = 10$ W, $N_d = 64$, $M = 3$, $R = 6$).

metric with SNR after matched filtering in [26]. We focus on the single-target case for comparison as in [19] and [26].

In the comparison scheme 1 [19], the transmit signal is $s(t) = \sum_{n=0}^{N_b-1} \sum_{k=0}^{N_d-1} b_n(k) e^{j2\pi k \frac{t}{N_d T}} u(t - nT_s)$. Consider AN-aided transmit beamforming, $b_n(k) = \mathbf{W}(k) \mathbf{x}_n(k) + \mathbf{z}_n(k)$, where $\mathbf{x}_n(k) \in \mathbb{C}^{M \times 1}$ is the information symbol vector with $\mathbb{E}[\mathbf{x}_n(k) \mathbf{x}_n^H(k)] = \mathbf{I}$, $\mathbf{W}(k) = [\mathbf{w}_1(k), \mathbf{w}_2(k), \dots, \mathbf{w}_M(k)] \in \mathbb{C}^{N_t \times M}$ is the beamforming matrix, $\mathbf{w}_m(k)$ represents the beamforming vector of the m th user on the k th subcarrier and $\mathbf{z}_n(k) \sim \mathcal{CN}(0, \mathbf{R}_Z(k))$ is the AN vector in the n th OFDM symbol on the k th subcarrier. The covariance matrix of $b_n(k)$ is $\mathbf{R}_{b_k} = \mathbb{E}[b_n(k) b_n^H(k)] = \sum_{m=1}^M \mathbf{W}_m(k) + \mathbf{R}_Z(k)$, where $\mathbf{W}_m(k) \triangleq \mathbf{w}_m(k) \mathbf{w}_m^H(k)$. For the target 1, the beam pattern is $\mathbf{a}^H(\theta_1) \mathbf{R}_{b_k} \mathbf{a}(\theta_1)$. We assume that the total power is equally allocated on each subcarrier. For each subcarrier, we optimize $\mathbf{W}(k)$ and $\mathbf{R}_Z(k)$ to maximize the secrecy capacity under the beam pattern constraint utilizing the optimization method in [19].

In comparison scheme 2, we use the same system model as comparison scheme 1, but replace the radar performance metric with the matched filtered SNR in [26] (this metric determines the detection performance of the radar). On the k th subcarrier, the maximum SNR after matched filtering is $\text{SNR}_k = \bar{c}_r^2 N_b R |\tilde{b}_n(k)|^2 / \sigma_w^2$, where $\bar{c}_r = \frac{1}{R} \sum_{r=1}^R c_r$, $\tilde{b}_n(k) = \mathbf{a}^H(\theta_1) \mathbf{b}_n(k)$ and $\sigma_w^2 = \frac{n_0}{N_d T}$. For each subcarrier, we optimize $\mathbf{W}(k)$ and $\mathbf{R}_Z(k)$ to maximize the secrecy capacity under SNR_k constraint utilizing the method in [26].

In both comparison scheme 1 and 2, we utilize the optimized transmission parameters to calculate CRLB to evaluate the estimation performance. As stated in Section III-B, we can calculate CRLB of location and velocity estimation of unknown targets based on $\{y_{n,r}(k), n = 1, \dots, N_b, r = 1, \dots, R, k = 1, \dots, N_d\}$. The tradeoff between the secrecy capacity and CRLBs of location and velocity in comparison schemes and proposed scheme are shown in Fig. 10.

In comparison scheme 1, varying the SINR threshold of each CR or the deviation between designed beampattern and ideal beampattern have an impact on the estimation performance. We first design the ideal beampattern aligned to target 1 according to [19], and then use it as the baseline to optimize the beamforming vector on each subcarrier. We find that the CRLB decreases slightly when the gap between the actual and the ideal beampattern is reduced, and the CRLB decreases significantly when the SINR threshold of CR decreases, i.e., the estimation performance is improved at the expense of reducing the secrecy capacity. When the secrecy capacity is reduced to 3.25 b/s/Hz, the CRLB of location and velocity are 17.113 and 0.737, which are still significantly larger than the CRLB in our scheme. In comparison scheme 2, the estimation performance metric is SNR_k after matched filtering. Increasing SNR_k can improve the estimation performance with slight rate loss when CRLBs of location and velocity are larger than 20 and 0.7, respectively. However, when CRLBs of location and velocity are smaller than 20 and 0.7, respectively, secrecy capacity will decrease sharply with the decrease of CRLB. When $\text{SNR}_k > 14$ dB, the problem is infeasible so that the lowest CRLB of location and velocity are 5.6027 and 0.2272, respectively. Therefore, with the same secrecy capacity 10 b/s/Hz, the proposed scheme is obviously superior to comparison schemes in terms of estimation performance. On the other hand, both the comparison schemes 1 and 2 adopted space division multiplexing on each subcarrier. By optimizing beamforming vectors to control multiuser interference, schemes 1 and 2 can obtain higher sum secrecy capacity than our scheme, but cannot obtain a better estimation performance.

VI. CONCLUSION

In this article, we investigated secure transmission in ICRS. To improve the secrecy capacity between BS and CRs, and meanwhile detect the targets accurately, AN is added into OFDM signals. We analyze the CRLB of target parameters and the analysis results show that CRLB only depends on the power and subcarrier allocation, and is independent of power splitting ratio. That means for given power and subcarrier allocation, we can improve the secrecy capacity by adjusting the power splitting factor without affecting the detection performance. To achieve the best tradeoff between communication and detection performance, we maximize the secrecy capacity between BS and CRs under the constraints of CRLB and total power. Lagrange duality method is utilized to solve the problem to obtain the optimal power/subcarrier allocation and power splitting ratio. The numerical results show the tradeoff between secrecy capacity and detection performance and demonstrate the advantage of our work on the aspect of estimation performance. As future research direction, the work in this article can be further improved by designing integrated OFDM signals with low PAPR, considering extended target detection, and extending the proposed scheme to communication network with multiple cooperative transmit BSs.

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