Integrated Sensing and Communications for Statistical-QoS Provisioning Over 6G M-MIMO Mobile Networks Using FBC

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***Abstract*—Since the 6G mobile wireless networks require the high-performances on both channel state estimations and information transmissions, the technique of *integrated sensing and communication (ISAC)* has attracted considerable research attention due to its ability to sense and communicate by sharing the same frequency band and hardware. However, how to jointly optimize the sensing and communication functions of the ISAC to support the 6G traffic transmissions over a time-varying wireless fading channel has imposed many new challenges not encountered before. To conquer these difficulties, in this paper we propose the ISAC scheme to jointly sense the channel state and transmit the wireless-streaming data using massive multiple-input and multiple-output (massive MIMO) communications over the Rician fading channel. First, we establish the system models for the ISAC scheme under the Rician fading wireless channel and the channel state estimation scheme using the radar sensing feedback. Second, we define the channel state estimation distortion and the capacity-distortion function of a massive MIMO channel to jointly measure the performances of sensing and communication in our ISAC scheme. Third, we integrate the capacity-distortion function with the finite blocklength coding (FBC) scheme by developing the concept of the *ISAC-based ϵ-effective capacity* to implement the statistical delay and error-rate bounded provisioning for supporting the 6G traffic under our ISAC scheme. Finally, we use numerical analyses to validate and evaluate our proposed ISAC scheme with massive MIMO in the non-asymptotic regime.**

***Index Terms*—6G, ISAC, massive MIMO, ϵ-effective capacity, statistical delay and error-rate bounded QoS, FBC.**

I. Introduction

THE 6G wireless networks are anticipated to support various applications with stringent quality-of-service (QoS) requirements, including the *massive ultra-reliable and low-latency communications (mURLLC)*, such as the digital twin (DT), metaverse streaming, dynamic multiple mobile targets tracking, smart-home, and vehicle localizations. These emerging civilian and military applications demand the short delay and low error probability for serving massive mobile users (MUs). Specifically, mURLLC techniques can support military operations by delivering critical information rapidly, accurately, and reliably to the correct individual or organization, and thus, can significantly improve the efficiency of combat operations. To support mURLLC applications, the integration of the *statistical QoS theory* [1–3] with the *finite blocklength coding (FBC)* [4] has been developed as an efficient technique to reduce the transmission latency while mitigating the decoding error rate. Another core feature envisioned for the 6G wireless networks is the extensive mobility of the MUs, which requires the accurate and timely detection and localization capabilities to provide the highly accurate sensing and the high data-rate communication. Towards this end, the integrated sensing and communication (ISAC) has been developed to significantly improve the channel sensing accuracy by using communication signals to increase the spectrum efficiency and reduce the hardware cost [5].

Recent works have studied and proposed various ISAC schemes. The authors of [6] proposed an ISAC processing framework relying on millimeter-wave (mmWave) massive MIMO systems using a compressed sampling perspective. The authors of [7] proposed an IEEE 802.11ad-based radar applications to enable a joint waveform for automotive radar and a potential mmWave vehicular communication system based on the mmWave consumer wireless local area network standard, allowing hardware reuse. The work of [8] designed approaches of intelligent waveforms, which are suitable for simultaneously performing both data transmission and radar sensing, based on classical phase-coded waveforms of wireless communications.

However, these above works mainly focused on the targets' position sensing using ISAC schemes. In this paper, we propose a channel state estimation method using the radar echo received by the massive MIMO base station (BS) in ISAC. First, we define the capacity-distortion function to obtain the channel capacity under a given channel estimation distortion constraint. Then, applying the FBC scheme into the capacity-distortion function, we define the ϵ-effective capacity to measure the communication performance in ISAC, which characterizes the maximum constant arrival rate for a given service process considering the delay decaying rate and the non-vanishing

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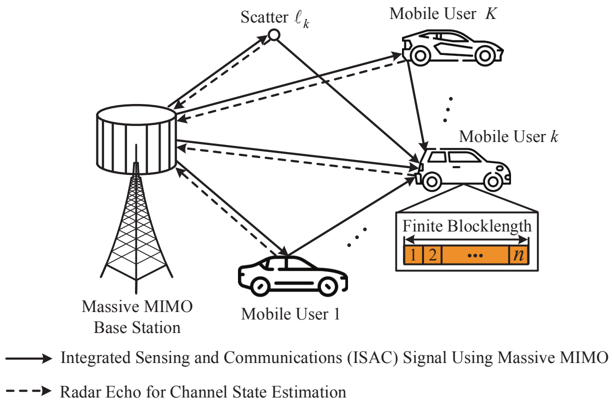

Fig. 1. System models of the mono-static ISAC using massive MIMO channel.

decoding error rate, subject to statistical delay and error-rate bounded QoS constraints, respectively, of the 6G streaming. Finally, we formulate an optimization problem to maximize the ϵ-effective capacity under constraints of the channel state estimation distortion and the ISAC signal's transmit power.

The rest of this paper is organized as follows. Section II establishes the system models for our proposed ISAC scheme. Section III estimates the channel state under the channel estimation distortion using the ISAC scheme. Section IV formulates the optimization problem to maximize the ϵ-effective-capacity. Section V validates and evaluates our developed ISAC schemes. This paper concludes with Section VI.

II. System Models for Our Proposed ISAC-Based Architectures Using Massive MIMO and FBC

As shown in Fig. 1, we consider a cellular network consisting of a massive MIMO BS and totally K moving targeted MUs, where the massive MIMO BS sends the radar sensing and the downlink communication signals simultaneously to these K targeted MUs for tracking the movements of and communicating to these MUs. Assume that there are M_T antennas on the BS and there are M_R antennas for each targeted MU, where $M_T \gg M_R$. The channel state information is available at both the sender (i.e., the BS) and the receiver (i.e., MUs), where the BS obtains the channel state information through the radar echo and MUs obtain the channel state information by using the de-spreading scheme for pilot signals sent by the BS [9, Section III-A].

A. Integrated Sensing and Communications (ISAC) Scheme

Assume that the radar sensing model uses the mono-static model, where the massive MIMO BS transmits omnidirectional radar signals, and receives the reflected radar echo from K targeted MUs. In our proposed radar echo sensing, each MU is considered as a scatterer to reflect the radar signal, and thus, the angle-of-arrival (AoA) and angle-of-departure (AoD) of radar sensing signals for each targeted MU are the same. The steering/response signal vector, denoted by $\mathbf{a}(M,\alpha)$ for M antennas and AoA/AoD equal to α, is defined as follows:

$$\mathbf{a}(M,\alpha) \triangleq \left[1, e^{j\pi\sin(\alpha)}, \dots, e^{j\pi(M-1)\sin(\alpha)}\right]^{\mathsf{T}}, \tag{1}$$

where $(\cdot)^{\mathsf{T}}$ is the transpose operation. For the kth MU's radar-echo and radar-signal, let ϕ_k be the AoA and AoD (where AoA is equal to AoD for radar sensing). We use the Rician fading wireless channel model to characterize the radar sensing channel by assuming that there are L_k scatters (including $(K-1)$ MUs other than the kth MU) between the MU k and BS with $L_k \geq (K-1)$. There is one explicit line-of-sight (LoS) path between each MU and the BS and there are L_k non-line-of-sight (NLoS) paths. Let ℓ_k be the index of the NLoS path set $\{1, 2, \dots, L_k\}$ between the kth targeted MU and the BS. Define $\mathbf{S}_k^{\text{sen}} \in \mathbb{C}^{M_T \times M_T}$ as the radar sensing channel matrix for the kth MU, where $\mathbb{C}^{M_T \times M_T}$ denotes a set of elements each consisting of a complex-valued matrix with M_T rows and M_T columns, which is given by [10, Eq. (30)]

$$\mathbf{S}_k^{\text{sen}} \triangleq \underbrace{\mu_k \mathbf{a}(M_T,\phi_k)\mathbf{a}(M_T,\phi_k)^{\mathsf{T}}}_{\text{Targeted MU } k \text{ in LoS path}} + \underbrace{\sum_{\ell_k=1}^{L_k} \mu_{\ell_k} \mathbf{a}(M_T,\phi_{\ell_k})\mathbf{a}(M_T,\phi_{\ell_k})^{\mathsf{T}}}_{\text{Other MUs and scatters in NLoS paths}} \tag{2}$$

where μ_k is the amplitude of a complex value, characterizing both the signal attenuation and initial phase difference for the radar signal between the massive MIMO BS and the kth MU. In the radar sensing, $\{\mu_k, \phi_k\}, \forall k$, denote the sensing parameters to be estimated and maintain unchanged during a sensing period.

For the massive MIMO downlink communications, we denote by $\mathbf{S}_k^{\text{com}} \in \mathbb{C}^{M_R \times M_T}$ the communication channel state matrix, representing the channel state between all antennas on the kth targeted MU and all antennas on the massive MIMO BS. We also use the Rician fading wireless channel model to characterize the downlink communications. Also, let $\widetilde{\mu}_{k,\ell_k}, \forall \ell_k \in \{1, \dots, L_k\}$, be the complex scattering coefficient of the ℓ_kth multipath, and let ξ_{k,ℓ_k} and ψ_{k,ℓ_k} denote the AoA and AoD, respectively, over the ℓ_kth downlink communication multipath for the kth targeted MU. Further, let $\widetilde{\mu}_{k,0}$, $\xi_{k,0}$, and $\psi_{k,0}$ be the coefficient, AoA, and AoD, respectively, of the LoS path. Then, the communication channel state matrix $\mathbf{S}_k^{\text{com}}$ is given by [10, Eq. (29)]

$$\mathbf{S}_k^{\text{com}} \triangleq \underbrace{\widetilde{\mu}_{k,0} \mathbf{a}(M_R,\xi_{k,0})\mathbf{a}(M_T,\psi_{k,0})^{\mathsf{T}}}_{\text{LoS path to MU } k} + \underbrace{\sum_{\ell_k=1}^{L_k} \widetilde{\mu}_{k,\ell_k} \mathbf{a}(M_R,\xi_{k,\ell_k})\mathbf{a}(M_T,\psi_{k,\ell_k})^{\mathsf{T}}}_{\text{NLoS paths to MU } k}. \tag{3}$$

The transmitted ISAC signal from the massive MIMO BS to all K targeted MUs, denoted by $\mathbf{x} \in \mathbb{C}^{M_T \times 1}$, is given by

$$\mathbf{x} = \mathbf{W}^{\text{com}}\mathbf{q}^{\text{com}} + \mathbf{W}^{\text{sen}}\mathbf{q}^{\text{sen}} \tag{4}$$

where $\mathbf{W}^{\text{com}} \in \mathbb{C}^{M_T \times KM_R}$ and $\mathbf{W}^{\text{sen}} \in \mathbb{C}^{M_T \times KM_R}$ are the communication and sensing precoders for all K targeted MUs, respectively, $\mathbf{q}^{\text{com}} \in \mathbb{C}^{KM_R \times 1}$ is the communication symbols and $\mathbf{q}^{\text{sen}} \in \mathbb{C}^{KM_R \times 1}$ is the probing sensing symbols. The reflected radar sensing signal received by the massive MIMO BS, i.e., radar echo, denoted by $\mathbf{y}^{\text{sen}} \in \mathbb{C}^{M_T \times 1}$, is given by

$$\mathbf{y}^{\text{sen}} = \mathbf{S}^{\text{sen}}\mathbf{x} + \mathbf{n}^{\text{sen}} = \mathbf{S}^{\text{sen}}\left(\mathbf{W}^{\text{com}}\mathbf{q}^{\text{com}} + \mathbf{W}^{\text{sen}}\mathbf{q}^{\text{sen}}\right) + \mathbf{n}^{\text{sen}} \quad (5)$$

where $\mathbf{n}^{\text{sen}} \in \mathbb{C}^{M_T \times 1}$ is the additive white Gaussian noise (AWGN) at all M_T antennas on the massive MIMO BS. The massive MIMO BS uses the received radar echo $\mathbf{y}^{\text{sen}}$ given by Eq. (5) to obtain the AoA of each MU by analyzing ϕ_k and the Doppler parameter by analyzing the reflected $\mathbf{q}^{\text{sen}}$.

B. Integrating Statistical QoS Provisioning Theory with FBC to Support mURLLC Streaming

The 6G wireless networks real-time streaming requires the stringent QoS on *both* transmission *delay* and *decoding error probability*. The FBC technique has been developed to enable small packet communications for adaptive error-control and real-time transmissions, where senders encode the message with short packets (i.e., packets with small numbers of bits) to reduce the transmission latency while constraining and controlling the decoding error probability. We define an FBC scheme in the following definition.

Definition 1: We consider a wireless fading channel, which uses input blockcode set $\mathcal{X}$ and output blockcode set $\mathcal{Y}$. We define that an (n, W_k, ϵ_k)-code, $\forall k \in \{1, 2, \ldots, K\}$, for a state-dependent memoryless channel consists of [4, 11]:

- A message set $\mathcal{W}_k = \{c_{1,k}, \cdots, c_{W_k,k}\}$, $\forall k$, with the cardinality W_k and the message length equal to $\log_2(W_k)$ under FBC, see Figs. 1 and 2.
- A delayed feedback output set $\mathcal{Z}^{i-1}, \forall i \in \{1, \ldots, n\}$. An encoder is a function f: $\mathcal{W}_1 \times \cdots \times \mathcal{W}_K \times \mathcal{Z}^{i-1} \mapsto \mathcal{X}^n$, where $\mathcal{X}^n$ is the set of codewords with length n and $\mathcal{Z}^{i-1}$ is the set of $(i-1)$ feedback outputs.
- The channel state set $\mathcal{S}_k^{\text{com}}, \forall k$. At the receiver end, a decoder produces an estimate of the original message by observing the channel output, according to a function: $\mathcal{Y}^n \times \mathcal{S}_k^{\text{com}} \mapsto \widehat{\mathcal{W}}_k$, where $\mathcal{Y}^n$ is the set of received codewords with length n and $\widehat{\mathcal{W}}_k$ is an estimation of $\mathcal{W}_k$.
- The decoding-error probability, denoted by ϵ_k, is defined as $\epsilon_k \triangleq \frac{1}{W_k}\sum_{w=1}^{W_k} \Pr\{c_{w,k} \neq \widehat{c}_{w,k}\}$, where the message index $w \in \{1, \cdots, W_k\}$, a message $c_{w,k} \in \mathcal{W}_k$ and its estimated message $\widehat{c}_{w,k}$ received at MU k, $\widehat{c}_{w,k} \in \widehat{\mathcal{W}}_k$.
- The channel state estimation function h: $\mathcal{X}^n \times \mathcal{Z}^n \mapsto \widehat{\mathcal{S}}_k^{\text{com}}$, where $\widehat{\mathcal{S}}_k^{\text{com}}$ is the estimation of $\mathcal{S}_k^{\text{com}}$.

where usually $\epsilon_k > 0$ if $n < \infty$. ■

Thus, the triple-tuple (n, W_k, ϵ_k) defines the data source with the cardinality W_k which can successfully transmit the messages with a probability of success $(1 - \epsilon_k)$ over n channel uses. In Fig. 2, $\mathbf{M}^n$ denotes the sequence of $M_i, \forall i \in \{1, \ldots, n\}$, over n channel uses, for any given M_i; and $\mathbf{S}_k^{\text{com}} \in \mathcal{S}_k^{\text{com}}$, and $\widehat{\mathbf{S}}_k^{\text{com}} \in \widehat{\mathcal{S}}_k^{\text{com}}$. The channel input at the ith channel use is $X_i = f(W_1, \ldots, W_K, \mathbf{Z}^{i-1})$, $i \in \{1, 2, \cdots, n\}$, with the given encoding scheme $f(\cdot)$, where $\mathbf{Z}^{i-1} \in \mathcal{Z}^{i-1}$ is the sequence of channel feedback by the $(i-1)$th channel use. In Fig. 2, we define the channel state as $\mathbf{S}^{\text{com}} \triangleq [\mathbf{S}_1^{\text{com}}, \cdots, \mathbf{S}_K^{\text{com}}]^\intercal$, and the channel output sequence as $\mathbf{Y} \triangleq [Y_{1,i}, \cdots, Y_{K,i}]$, where the channel output at the ith channel use $Y_{k,i}$ is obtained from the state $\mathbf{S}_k^{\text{com}}$ and the input X_i according to the transition law $P_{\mathbf{Y}Z|X\mathbf{S}^{\text{com}}}$.

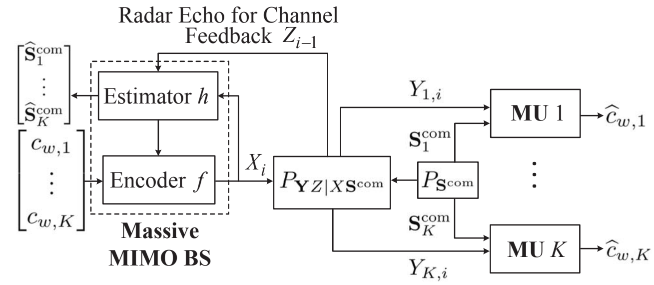

Fig. 2. The ISAC's encoding, decoding, and channel estimation model at the ith channel use with $i = 1, 2, \ldots, n$ to transmit every message for each of K MUs, where n is FBC's codeword blocklength.

III. ISAC-Based Channel Estimation Under Distortion

A. The Channel State Estimation Distortion

To simplify the notations, we use S_k to represent any element in $\mathbf{S}_k^{\text{com}}$, use s_k to represent the realization of S_k, and we use $\hat{s}_k$ and $\hat{\mathbf{s}}_k^n$ to represent the estimation for s_k and $\mathbf{s}_k^n$, respectively. Define $d : \mathcal{S}_k^{\text{com}} \times \widehat{\mathcal{S}}_k^{\text{com}} \mapsto [0, \infty)$ as the channel state estimation distortion function. Using the mean squared error (MSE), we define the channel state estimation distortion of the ISAC scheme in the following definition.

Definition 2: The *channel state estimation distortion* of the ISAC scheme due to the *radar estimation distortion*, denoted by D, is defined as follows [10, Eqs. (3) and (5)] [11]:

$$D \triangleq \mathbb{E}\left[d\left(\mathbf{s}_k^n, \hat{\mathbf{s}}_k^n\left(\mathbf{x}^n, \mathbf{z}^{n-1}\right)\right)\right] \triangleq \frac{1}{n}\sum_{i=1}^{n} \mathbb{E}[d(s_k, \hat{s}_k(x_i, z_{i-1}))]$$

$$\triangleq \frac{1}{n}\sum_{i=1}^{n} \mathbb{E}\left[\|s_k - \hat{s}_k(x_i, z_{i-1})\|_F^2\right] \leq D_{\text{a}} \quad (6)$$

where $\|\cdot\|_F$ is the Frobenius norm; $\mathbf{x}^n \in \mathcal{X}^n$ and $\mathbf{z}^{n-1} \in \mathcal{Z}^{n-1}$ based on Definition 1; $\hat{\mathbf{s}}_k^n\left(\mathbf{x}^n, \mathbf{z}^{n-1}\right)$ and $\hat{s}_k(x_i, z_{i-1})$ are the channel state estimation functions defined in Definition 1; and D_{a} is the *achievable channel estimation distortion* if there exists a sequence of (n, W_k, ϵ_k)-codes defined in Definition 1 such that distortion D defined by Eq. (6) satisfies:

$$\limsup_{n\to\infty}\{D\} \leq D_{\text{a}} \quad (7)$$

and the decoding-error probability ϵ_k defined in Definition 1 satisfies $\epsilon_k \triangleq \frac{1}{W_k}\sum_{w=1}^{W_k} \Pr\{c_{w,k} \neq \widehat{c}_{w,k}\} \to 0$ as $n \to \infty$. ■

B. Channel Estimation Distortion of Rician Fading Channel

We denote by $\mathbf{S}^{\text{com}} = [\mathbf{S}_1^{\text{com}}, \ldots, \mathbf{S}_K^{\text{com}}]^\intercal \in \mathbb{C}^{KM_R \times M_T}$ the channel state between all antennas on all K targeted MUs and all antennas on the massive MIMO BS, and denote by $\widehat{\mathbf{S}}^{\text{com}}$ the estimation of $\mathbf{S}^{\text{com}}$. Define $\mathbf{S}^{\text{up}} \in \mathbb{C}^{M_T \times KM_R}$ as the uplink channel state between all antennas of MUs and all antennas on massive MIMO BS, and define $\widehat{\mathbf{S}}^{\text{up}}$ as the estimation of $\mathbf{S}^{\text{up}}$. Since downlink and uplink channels share the same spectrum, we assume that they are reciprocal within each coherence interval, and hence we can treat the estimation of uplink channel as the true downlink channel. Thus, we have:

$$\mathbf{S}^{\text{up}} = (\mathbf{S}^{\text{com}})^{\mathsf{T}} \text{ and } \widehat{\mathbf{S}}^{\text{up}} = \left(\widehat{\mathbf{S}}^{\text{com}}\right)^{\mathsf{T}}. \tag{8}$$

Denote by n_{p} the length of uplink training sequence. Let $\mathbf{X}^{(\text{p})} \in \mathbb{C}^{KM_R \times n_{\text{p}}}$ be the uplink pilot training sequences sent by K MUs. The matrix $\mathbf{X}^{(\text{p})}$ fulfills the total power constraint $\text{tr}\left\{\left(\mathbf{X}^{(\text{p})}\right)^H \mathbf{X}^{(\text{p})}\right\} = \rho^{\text{ul}}$, where $(\cdot)^H$ is the conjugate transpose and $\text{tr}\{\cdot\}$ is the trace of a matrix. The received singal at massive MIMO BS, denoted by $\mathbf{Y}^{(\text{p})} \in \mathbb{C}^{M_T \times n_{\text{p}}}$, is given by

$$\mathbf{Y}^{(\text{p})} = \mathbf{S}^{\text{up}}\mathbf{X}^{(\text{p})} + \mathbf{N} \tag{9}$$

where $\mathbf{N} \in \mathbb{C}^{M_T \times n_{\text{p}}}$ is the AWGN matrix uncorrelated with the channel $\mathbf{S}^{\text{up}}$. Define $\widetilde{\mathbf{X}}^{(\text{p})} \in \mathbb{C}^{n_{\text{p}}M_T \times KM_RM_T}$ as follows:

$$\widetilde{\mathbf{X}}^{(\text{p})} \triangleq \left(\mathbf{X}^{(\text{p})}\right)^{\mathsf{T}} \otimes \mathbf{I}_{M_T} \tag{10}$$

where $\otimes$ is the Kronecker product and $\mathbf{I}_{M_T}$ is the $M_T \times M_T$ identity matrix. Define $\text{vec}(\mathbf{M})$ as the column vector obtained by stacking the columns of the matrix $\mathbf{M}$ and define $\mathbf{C}_{\mathbf{M}}$ as the covariance matrix of $\mathbf{M}$. Define $\boldsymbol{\varrho} \in \mathbb{C}^{n_{\text{p}}M_T \times 1}$ as follows:

$$\boldsymbol{\varrho} \triangleq \text{vec}(\mathbf{Y}^{(\text{p})}) - \widetilde{\mathbf{X}}^{(\text{p})}\text{vec}\left(\overline{\mathbf{S}}^{\text{up}}\right) - \text{vec}\left(\overline{\mathbf{N}}\right) \tag{11}$$

and $\overline{\mathbf{N}} \in \mathbb{C}^{M_T \times n_{\text{p}}}$ is the mean of $\mathbf{N}$. Let $\widehat{\mathbf{S}}^{\text{up}}$ be the minimum mean square error (MMSE) estimator of $\mathbf{S}^{\text{up}}$, and according to [12, Eqs. (5)-(6)], $\text{vec}\left(\widehat{\mathbf{S}}^{\text{up}}\right)$ is given by

$$\text{vec}\left(\widehat{\mathbf{S}}^{\text{up}}\right) = \text{vec}\left(\overline{\mathbf{S}}^{\text{up}}\right) + \mathbf{C}_{\text{vec}(\mathbf{S}^{\text{up}})}\left(\widetilde{\mathbf{X}}^{(\text{p})}\right)^H \times \left[\widetilde{\mathbf{X}}^{(\text{p})}\mathbf{C}_{\text{vec}(\mathbf{S}^{\text{up}})}\left(\widetilde{\mathbf{X}}^{(\text{p})}\right)^H + \mathbf{C}_{\text{vec}(\mathbf{N})}\right]^{-1} \boldsymbol{\varrho}. \tag{12}$$

where $\overline{\mathbf{S}}^{\text{up}}$ is the expectation operation for each element of the matrix $\mathbf{S}^{\text{up}}$ and $\mathbf{C}_{\text{vec}(\mathbf{N})} \in \mathbb{C}^{n_{\text{p}}M_T \times n_{\text{p}}M_T}$. The channel estimation $\widehat{\mathbf{S}}^{\text{up}}$ for the Rician fading channel should satisfy the following condition [12, Eq. (8)]:

$$D \triangleq \mathbb{E}\left[\left\|\text{vec}\left(\mathbf{S}^{\text{up}}\right) - \text{vec}\left(\widehat{\mathbf{S}}^{\text{up}}\right)\right\|_F^2\right] \leq D_{\text{a}}. \tag{13}$$

Thus, the *downlink channel estimation*, denoted by $\widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}}) \in \mathbb{C}^{KM_R \times M_T}$, subject to the achievable channel estimation distortion D_{a} as specified in Eq. (13), is given by:

$$\widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}}) = \left(\widehat{\mathbf{S}}^{\text{up}}\right)^{\mathsf{T}} = \arg \min_{\left(\widehat{\mathbf{S}}^{\text{up}}\right)^{\mathsf{T}} \in \mathcal{S}_k^{\text{com}}} \{D\} \quad \text{s.t.: } D \leq D_{\text{a}}. \tag{14}$$

IV. Maximization of the ISAC-Based ϵ-Effective Capacity

A. The Capacity-Distortion Function

Let $\mathbf{q}_k^{\text{com}} \in \mathbb{C}^{M_R \times 1}$ be the communication symbols for the kth targeted MU and let $\mathbf{q}^{\text{com}} = [\mathbf{q}_1^{\text{com}}, \ldots, \mathbf{q}_K^{\text{com}}]^{\mathsf{T}} \in \mathbb{C}^{KM_R \times 1}$ be the communication symbols for all K MUs. Define B_k as the maximum transmit power allocation for the kth targeted MU. Let $\mathbf{W}_k^{\text{com}} \in \mathbb{C}^{M_T \times M_R}$ be the communication beamforming matrix (i.e., precoder) for the kth targeted MU and let $\mathbf{W}^{\text{com}} = [\sqrt{B_1/M_T}\mathbf{W}_1^{\text{com}}, \ldots, \sqrt{B_K/M_T}\mathbf{W}_K^{\text{com}}] \in \mathbb{C}^{M_T \times KM_R}$ be the communication precoder for all K MUs. We re-write $\mathbf{S}^{\text{com}} = \widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}}) - \widetilde{\mathbf{S}}^{\text{com}}$, where $\widetilde{\mathbf{S}}^{\text{com}} \in \mathbb{C}^{KM_R \times M_T}$ is the error of the estimation. Then, the received ISAC signal through communication channels at all antennas on K MUs, denoted by $\mathbf{y}^{\text{com}} \in \mathbb{C}^{KM_R \times 1}$, is given by

$$\begin{aligned}\mathbf{y}^{\text{com}} &= \mathbf{S}^{\text{com}}\mathbf{x} + \mathbf{n}^{\text{com}} = \left(\widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}}) - \widetilde{\mathbf{S}}^{\text{com}}\right)\mathbf{x} + \mathbf{n}^{\text{com}} \\ &= \widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}})\mathbf{W}^{\text{com}}\mathbf{q}^{\text{com}} + \widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}})\mathbf{W}^{\text{sen}}\mathbf{q}^{\text{sen}} + \mathbf{n}^{\text{com}} \\ &\quad - \widetilde{\mathbf{S}}^{\text{com}}\mathbf{W}^{\text{com}}\mathbf{q}^{\text{com}} - \widetilde{\mathbf{S}}^{\text{com}}\mathbf{W}^{\text{sen}}\mathbf{q}^{\text{sen}}\end{aligned} \tag{15}$$

where $\mathbf{n}^{\text{com}} \in \mathbb{C}^{KM_R \times 1}$ is the AWGN at all antennas of all targets. Let y_m^{com} be the mth row of the vector $\mathbf{y}^{\text{com}}$, representing the received ISAC signal at the mth antenna of a target ($\forall m \in \{1, \ldots, KM_R\}$ and assuming that m is the index for the antenna belonging to the kth targeted MU), and y_m^{com} is given by [13, Eq. (12)]

$$y_m^{\text{com}} = \sqrt{\frac{B_k}{M_T}}\widehat{\mathbf{s}}_m^{\text{com}}(D_{\text{a}})\mathbf{w}_m^{\text{com}}q_m^{\text{com}} + \Omega_m \tag{16}$$

where Ω_m is the *effective additive noise* at the mth antenna defined by

$$\begin{aligned}\Omega_m \triangleq &\sum_{\substack{i=1,\\ i\neq m}}^{kM_R}\left(\sqrt{\frac{B_i}{M_T}}\widehat{\mathbf{s}}_m^{\text{com}}(D_{\text{a}})\mathbf{w}_i^{\text{com}}q_i^{\text{com}}\right) + \widehat{\mathbf{s}}_m^{\text{com}}(D_{\text{a}})\mathbf{W}^{\text{sen}}\mathbf{q}^{\text{sen}} \\ &+ n_m^{\text{com}} - \widetilde{\mathbf{s}}_m^{\text{com}}\mathbf{W}^{\text{com}}\mathbf{q}^{\text{com}} - \widetilde{\mathbf{s}}_m^{\text{com}}\mathbf{W}^{\text{sen}}\mathbf{q}^{\text{sen}}\end{aligned} \tag{17}$$

where $\widehat{\mathbf{s}}_m^{\text{com}}(D_{\text{a}}) \in \mathbb{C}^{1 \times M_T}$ is the mth row of the matrix $\widehat{\mathbf{S}}^{\text{com}}(D_{\text{a}})$, $\mathbf{w}_m^{\text{com}}$ is the mth column of the matrix $\mathbf{W}^{\text{com}}$, n_m^{com} is the mth row of the vector $\mathbf{n}^{\text{com}}$, and B_i is the transmit power allocation for an MU to which i belongs. Using Eq. (16), the signal to noise ratio (SNR), denoted by $\varphi_m(B_k, D_{\text{a}})$, on the mth antenna of the target is derived as follows:

$$\varphi_m(B_k, D_{\text{a}}) = \frac{\frac{B_k}{M_T}\text{Var}\left[\widehat{\mathbf{s}}_m^{\text{com}}(D_{\text{a}})\mathbf{w}_m^{\text{com}}q_m^{\text{com}}\right]}{\text{Var}[\Omega_m]}. \tag{18}$$

Define the *channel estimation cost function*, denoted by $c_k(x)$, as follows [11, Eq. (8)]:

$$c_k(x) \triangleq \mathbb{E}\left[d\left(s_k, \hat{s}_k^*(x,z)\right) \middle| X = x\right] \tag{19}$$

where $\hat{s}_k^*(x,z)$ is the optimal estimation. To jointly measure the channel capacity and the channel state estimation distortion in ISAC, we employ the capacity-distortion function [5, 11] to represent the tradeoff between the channel-capacity and the sensing distortion. We define the capacity-distortion function [11, Eq. (11)], for the kth targeted MU, denoted by $C_k(B_k, D_{\text{a}})$, as follows:

$$\begin{aligned}C_k(B_k, D_{\text{a}}) &= \sum_{m=(k-1)M_R+1}^{kM_R} \max_{P_X \in \mathcal{P}_{D_{\text{a}}}} \left\{I\left(X; Y \middle| \mathbf{s}_m^{\text{com}}\right)\right\} \\ &\overset{\text{(a)}}{=} \sum_{m=(k-1)M_R+1}^{kM_R} \log_2(1 + \varphi_m(B_k, D_{\text{a}}))\end{aligned} \tag{20}$$

$$\text{s.t.: C1}: \ D = \sum\nolimits_{x \in \mathcal{X}} c_k(x)P_X(x) \leq D_{\text{a}} \tag{21}$$

$$\text{C2}: \ \frac{1}{n}\sum_{i=1}^{n} \mathbb{E}\left[b(X_i)\right] \leq B_k$$

where $\mathcal{P}_{D_{\rm a}} \triangleq \left\{P_X : \sum_{x\in\mathcal{X}} P_X(x)c_k^*(x) \leq D_{\rm a}\right\}$, $c_k^*(x)$ is the minimum $c_k(x)$ that can be achieved for a given signaling $x \in \mathcal{X}$, $I(\cdot;\cdot|\cdot)$ is the conditional mutual information, (a) holds due to [13, Eq. (16)], $\varphi_m(B_k, D_{\rm a})$ is the SNR derived by Eq. (18) as a function of $D_{\rm a}$, and $b(X_i)$ is the transmit power allocation for X_i. When $D_{\rm a} \to \infty$, Eq. (20) is the traditional capacity maximization problem without considering the distortion.

B. Maximizing the ISAC-Based ϵ-Effective Capacity

The statistical delay-bounded QoS guarantees [1–3] have been shown to be powerful in analyzing queuing behavior for the stochastic arrival and service processes over the time-varying wireless fading channels. The key statistical-QoS performance metric is the *effective capacity* which measures the maximum packet's constant arrival rate such that the given statistical *delay-bounded* QoS can be guaranteed. Based on the large deviation principle (LDP) [2], the queue-length process $Q_k(t)$ for the kth MU converges in distribution to a random variable $Q_k(\infty)$ such that

$$-\lim_{Q_{{\rm th},k}\to\infty} \frac{\log(\Pr\{Q_k(\infty) > Q_{{\rm th},k}\})}{Q_{{\rm th},k}} = \theta_k \tag{22}$$

where $Q_{{\rm th},k}$ is the queue length threshold (bound) and $\theta_k > 0$ is defined as the *QoS exponent* for MU k.

However, the conventional statistical-QoS theory modeled by Eq. (22) focuses only on the statistical delay-bounded QoS without considering the transmission reliability, which is thus not feasible to support mURLLC in our proposed ISAC-based 6G massive-MIMO/FBC wireless networks. To remedy these deficiencies, we propose to integrate the ISAC technique with the FBC-based effective-capacity theory to support *both* the statistical *delay* and *error-rate* bounded QoS provisioning. In particular, applying ISAC channel-state estimation-distortion and capacity-distortion functions to derive the decoding error probability function, we develop the new statistical QoS performance metric called: the *ISAC-based ϵ-effective capacity* to guarantee *both* the statistical *delay* and *error-rate* bounded QoS provisioning for our proposed ISAC-based massive-MIMO/FBC wireless networks through the following definition.

Definition 3: For an (n, W_k, ϵ_k)-code, the *ISAC-based ϵ-effective capacity*, denoted by $EC_k(\theta_k, \epsilon_k(B_k, D_{\rm a}))$, as the function of the delay QoS exponent θ_k, ISAC's *achievable channel estimation distortion* $D_{\rm a}$, and FBC's decoding-error probability $\epsilon_k(B_k, D_{\rm a})$, is defined as the maximum constant arrival rate for a given service process subject to *both* the statistical *delay* and *error-rate* bounded QoS requirements, which is derived as follows:

$$EC_k(\theta_k, \epsilon_k(B_k, D_{\rm a})) = -\frac{1}{n\theta_k}\log\Big\{\mathbb{E}_{r_k}[\epsilon_k(B_k,D_{\rm a})] + \mathbb{E}_{r_k}[(1-\epsilon_k(B_k,D_{\rm a}))]e^{-\theta_k \log_2(W_k)}\Big\} \tag{23}$$

where $\mathbb{E}_{r_k}\{\cdot\}$ denotes the expectation with respect to the random distance r_k between the kth MU and the BS, θ_k is derived in Eq. (22), and the decoding-error probability $\epsilon_k(B_k, D_{\rm a})$ defined in Definition 1, which measures the error-rate bounded QoS requirement, is derived by [4, Eq. (1)]:

$$\epsilon_k(B_k, D_{\rm a}) \approx Q\left(\frac{C_k(B_k, D_{\rm a}) - \frac{\log_2(W_k)}{n}}{\sqrt{V_k(B_k, D_{\rm a})/n}}\right) \tag{24}$$

where $D_{\rm a}$ is ISAC's *achievable channel estimation distortion* specified by Eqs. (6) and (7), $Q(\cdot)$ is the *Q*-function, $C_k(B_k, D_{\rm a})$ is the channel capacity derived in Eq. (20), and $V_k(B_k, D_{\rm a})$ is the *channel dispersion* of the kth targeted MU, which is given as follows [4, Eq. (293)]:

$$V_k(B_k,D_{\rm a}) \approx \sum_{\substack{m=\\(k-1)M_R+1}}^{kM_R} \left[1 - \frac{1}{(1+\varphi_m(B_k, D_{\rm a}))^2}\right] \tag{25}$$

where $\varphi_m(B_k, D_{\rm a})$ is the SNR derived by Eq. (18) as a function of $D_{\rm a}$ specified by Eqs. (6) and (7). ■

Using Definition 3, we can formulate the optimization problem for our proposed ISAC scheme to maximize the aggregate ISAC-based ϵ-effective capacity over all K targeted MUs as follows:

$$\max_{\widehat{\mathbf{s}}_m^{\rm com}(D_{\rm a}),\forall m} \left\{\sum_{k=1}^{K} EC_k(\theta_k, \epsilon_k(B_k, D_{\rm a}))\right\} \text{s.t.: C1 and C2} \tag{26}$$

where the maximization operation is with respect to $\widehat{\mathbf{s}}_m^{\rm com}(D_{\rm a}), \forall m$, specified in Eqs. (16) and (17) and $EC_k(\theta_k, \epsilon_k(B_k, D_{\rm a}))$ is derived in Eq. (23).

V. Performance Evaluations

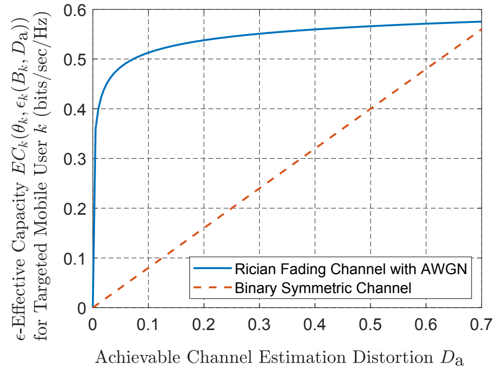

Fig. 3. The ϵ-effective capacity $EC_k(\theta_k, \epsilon_k(B_k, D_{\rm a}))$ for different values of the achievable channel estimation distortion $D_{\rm a}$ under Rician fading channel with AWGN and binary symmetric channel.

In Fig. 3, we compare the ϵ-effective capacity $EC_k(\theta_k, \epsilon_k(B_k, D_{\rm a}))$ for different values of the achievable channel estimation distortion $D_{\rm a}$ under a Rician fading channel with AWGN and a binary symmetric channel. The Rician fading channel with AWGN using the system model given by Section II. The binary symmetric channel supposes that the input symbols are complemented with the probability 0.5. We can observe from Fig. 3 that the Rician fading channel with AWGN can achieve a larger ϵ-effective capacity than the binary symmetric channel under a given $D_{\rm a}$, showing that the Rician fading channel with AWGN can provide a

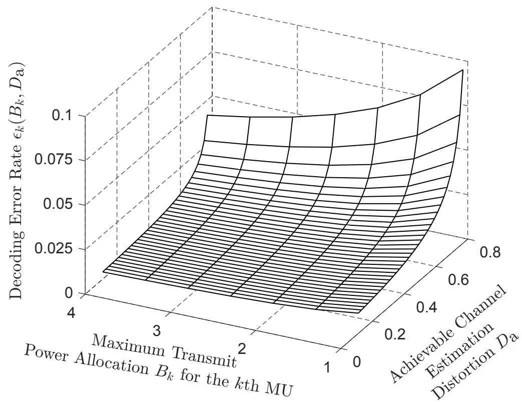

Fig. 4. The decoding error rate $\epsilon_k(B_k, D_\mathrm{a})$ under different values of maximum transmit power allocation B_k and achievable channel estimation distortion D_a.

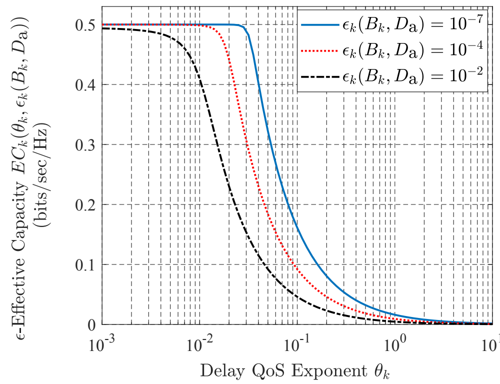

Fig. 5. The function of ϵ-effective capacity under different values of the delay QoS exponent θ_k and the decoding error rate $\epsilon_k(B_k, D_\mathrm{a})$.

more accurate channel state feedback. We can also observe that the ϵ-effective capacity monotonically increases as D_a increases, showing that the capacity-distortion function given by Eq. (20) approaches the traditional capacity maximization problem without considering the distortion.

Figure 4 shows the decoding error rate $\epsilon_k(B_k, D_\mathrm{a})$ under different values of the maximum transmit power allocation B_k and the achievable channel estimation distortion D_a. We observe from Fig. 4 that $\epsilon_k(B_k, D_\mathrm{a})$ monotonically decreases as B_k increases, since a larger transmit power allocation indicates a better channel condition. We also observe from Fig. 4 that $\epsilon_k(B_k, D_\mathrm{a})$ monotonically increases as D_a increases, since a less accurate channel state estimation yields a larger decoding error probability.

Figure 5 shows the ϵ-effective capacity $EC_k(\theta_k, \epsilon_k(B_k, D_\mathrm{a}))$ under different values of the delay QoS exponent θ_k and the decoding error rate $\epsilon_k(B_k, D_\mathrm{a})$. We observe that the ϵ-effective capacity is a monotonically decreasing function of the delay QoS exponent θ_k, because the θ_k indicates the statistical delay QoS stringency and a channel with a looser delay QoS requirement can support a larger data arrival rate. Fig. 5 reveals that the ϵ-effective capacity is also a monotonically decreasing function of the decoding error rate $\epsilon_k(B_k, D_\mathrm{a})$. This is because a smaller decoding error rate indicates a better channel quality and larger achievable rate, yielding a larger ϵ-effective capacity.

VI. Conclusions

To estimate the channel state, we have proposed an ISAC scheme using a joint radar and communication system to sense the channel state and convey messages simultaneously. We have proposed to measure the communication performance of the ISAC scheme using the ϵ-effective capacity, and proposed to measure the sensing performance using the channel estimation distortion. Having modeled the channel estimation distortion for ISAC's sensing performances, we have developed the new statistical-QoS metrics of the *ISAC-based ϵ-effective capacity* to measure the communications QoS performances. Then, we have formulated the optimization problem to maximize the aggregate ISAC-based ϵ-effective capacity. Using numerical analyses, we have validated and evaluated our proposed ISAC scheme over massive MIMO communication networks in the non-asymptotic regime.

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