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Joint Subcarrier and Phase Shifts Optimization for RIS-aided Localization-Communication System

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***Abstract*—Joint localization and communication systems have drawn significant attention due to their high resource utilization. In this paper, we consider a reconfigurable intelligent surface (RIS)-aided simultaneously localization and communication system. We first determine the sum squared position error bound (SPEB) as the localization accuracy metric for the presented localization-communication system. Then, a joint RIS discrete phase shifts design and subcarrier assignment problem is formulated to minimize the SPEB while guaranteeing each user's achievable data rate requirement. For the presented non-convex mixed-integer problem, we propose an iterative algorithm to obtain a suboptimal solution by utilizing the Lagrange duality as well as penalty-based optimization methods. Simulation results are provided to validate the performance of the proposed algorithm.**

***Index Terms*—Reconfigurable intelligent surface, joint localization and communication, subcarrier assignment, RIS phase shifts design.**

I. Introduction

In future communications, localization technology will be integrated into the communication systems as an essential function. Compared to independent localization and communication systems, joint localization and communication systems can improve spectrum and energy efficiency, and reduce hardware overhead by sharing physical platform and spectrum resources [1].

There are growing interests on investing different types of joint localization and communication problems. For example, a MIMO system was considered to maximize the sum rate while guaranteeing the localization performance [2]. The authors in [3] studied a radar and communication integration system to degrade the positioning error and improve the capacity of channels. Although the above-mentioned works can provide both localization and communication services and reduce resource overheads, some key issues remain to be further discussed: how to guarantee localization and communication services when the line-of-sight (LOS) link is completely blocked; how to simultaneously improve localization and communication performance with limited resources.

Recently, reconfigurable intelligent surface (RIS) as new technology has drawn significant attention, which can create a desirable path and reflect the incident signals towards the receivers by tuning the phase shifts of passive reflecting elements without any extra power or spectrum consumption [4]. Most of the existed works have been proposed to explore the capability of RIS techniques in communication and localization systems. Specifically, by joint designing base station (BS) beamforming and RIS configuration with continuous phase shifts, the RIS-based communication systems can effectively improve the system throughput and reduce the energy consumption [5]. Moreover, [6] pointed out that the communication performance and energy efficiency can also be ensured by the RIS with finite discrete phase shifts. Meanwhile, several RIS-aided localization schemes were presented to support far-field, near-field, and indoor localization. For example, [7] investigated an RIS-aided near-field regional localization problem. RIS was also applied to improve the discernibility of the received signal strength (RSS) at adjacent positions in indoor localization scenarios [8].

Based on the above observations, one can find that the LOS blocking and resource-limiting issues can be overcome by introducing RIS into the existing communication and localization systems. However, most prior works only focused on localization or communication independently in RIS-assisted networks. In this paper, we present a downlink RIS-assisted orthogonal frequency division multiple access (OFDMA) system to achieve both localization and communication services. In our considered system, the direct links between the BS and users are blocked due to the obstacle. Specifically, we first introduce an RIS-assisted OFDMA joint localization and communication system model with discrete RIS phase shifts. Then, a sum squared position error bound (SPEB) minimization problem is formulated under individual data rate constraints. For the formulated non-convex mixed-integer problem, we develop a suboptimal algorithm to iterate solve subcarrier assignment and RIS phase shifts by resorting to the Lagrange duality and penalty-based optimization methods. Finally, simulation results are provided to illustrate the performance of the presented algorithm.

The remainder of this paper is organized as follows. Section II describes the RIS-assisted localization-communication system model and provides the performance metrics for communication as well as localization services. In Section III, we first formulate a joint subcarrier and RIS phase shifts optimization problem and develop an iterative algorithm to solve such a non-convex mixed-integer problem. Simulations in Section IV validate the performance of the proposed algorithm. Finally, the conclusions are presented in Section V.

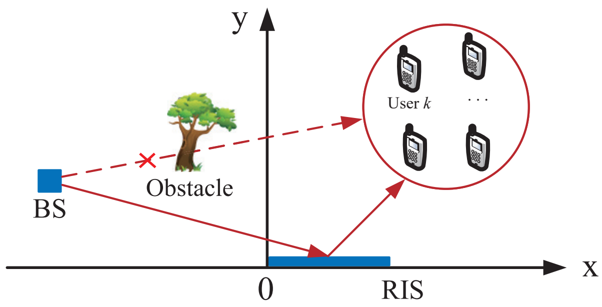


Fig. 1. The RIS-assisted localization and communication system.

II. System Model and Performance Metric

A. System Model

We consider a multiuser downlink transmission system as shown in Fig. 1, in which one BS via RIS serves K single-antenna users for both localization and communication. The system operates at carrier frequency f_c which corresponds to wavelength λ_c. The RIS consists of N_R reflecting elements with interelement spacing $d_R = \frac{\lambda_c}{2}$. Set the first reflecting element on the left as the reference point of RIS, which is located at $\boldsymbol{q}_R = [q_{R,x}, q_{R,y}]^{\mathrm{T}}$. The user k with unknown position is located at position $\boldsymbol{q}_k = [q_{k,x}, q_{k,y}]^{\mathrm{T}}$, while the BS is located at $\boldsymbol{q}_B = [q_{B,x}, q_{B,y}]^{\mathrm{T}}$. The distance d_{rk} and angle φ_{rk} between the RIS element r and user k are given by $d_{rk} = \|\boldsymbol{q}_k - \boldsymbol{q}_r\|$ and $\varphi_{rk} = \tan^{-1}\left(\frac{q_{k,x}-q_{r,x}}{q_{k,y}-q_{r,y}}\right)$, respectively, where $\boldsymbol{q}_r = [q_{r,x}, q_{r,y}]^{\mathrm{T}}$ is the position of the r-th RIS element. The distance d_{Br} and angle φ_{Br} between the BS and the r-th RIS element are constructed similar as d_{rk} and φ_{rk}.

Without loss of generality, we assume that each transmission block is equally divided into the downlink localization and communication phases [2]. Moreover, the OFDMA technology is adopted, and the frequency band with bandwidth W is divided into S subcarriers with the subcarrier spacing $\Delta f = \frac{W}{S}$. For the subcarrier association indicator $\beta_{k,s}$, we have

$$\beta_{k,s} = \begin{cases} 1, & \text{if subcarrier } s \text{ is allocated to user } k, \\ 0, & \text{otherwise.} \end{cases} \tag{1}$$

It is assumed that a subcarrier is exclusively scheduled to one user for both the downlink localization and communication together. Thus, we have the constraint $\sum_{k\in\mathcal{K}} \beta_{k,s} \leq 1, \forall s$. As for the RIS, we denote the reflection-coefficient vector as $\boldsymbol{\vartheta} = [\vartheta_1, ..., \vartheta_r, ..., \vartheta_{N_{\mathrm{R}}}]^{\mathrm{H}}$, where $\vartheta_r = e^{j\theta_r}$. We consider the discrete phase-shift value, i.e.,

$$\theta_r = \frac{2\pi l}{Q}, \;\; l \in \mathcal{Q} \triangleq \{0, ..., Q-1\}, \;\; \forall r \in \mathcal{N}_R. \tag{2}$$

where $Q = 2^b$ represents the number of phase-shift levels, b is the number of quantization bits.

The received signal of user k in the localization or communication phase via subcarrier s can be expressed as

$$y_{k,s}^{l(c)} = \sqrt{p}\sum_{r=1}^{N_{\mathrm{R}}} \vartheta_r^* g_{k,s}^r e^{-j2\pi s\Delta f(\tau_{Br}+\tau_{rk})} s_{k,s}^{l(c)} + n_{k,s}^{l(c)}, \tag{3}$$

where superscript l and c denote localization and communication phases, respectively; $s_{k,s}^l(t)$ with $\mathbb{E}[|s_{k,s}^l|^2] = 1$ and $s_{k,s}^c$ with $\mathbb{E}[|s_{k,s}^c|^2] = 1$ are the transmitted pilot and data signals, respectively; $n_{k,s}$ with zero-mean and variance σ^2 is the observation noise; $p = P_{\mathrm{MAX}}/S$ denotes the subcarrier power, while P_{MAX} is the maximum transmit power of BS; $\tau_{Br} = d_{Br}/c$ and $\tau_{rk} = d_{rk}/c$ are the time of arrivals (TOAs), c is the speed of light. $g_{k,s}^r = \rho_k h_{k,s}^r$, in which $h_{k,s}^r$ stands for the complex channel gain of BS-r-user k link. Note that $d_R \ll d_{Br} + d_{rk}$, we thus assume that the path losses are the same for all elements [6]. Thus, we have $\rho_k = (d_{BR} \cdot d_{Rk})^{-\alpha}$ with path loss exponent α.

B. Performance Metric

For communication service, we apply the individual achievable rate as the performance metric, i.e.,

$$R_k = \sum_{s\in\mathcal{S}} \beta_{k,s} \log_2\left(1 + \frac{p|\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{B}_{k,s}\boldsymbol{\hbar}_{k,s}|^2}{\sigma^2}\right), \tag{4}$$

where $\boldsymbol{B}_{k,s} = \mathrm{diag}\{[g_{k,s}^1, ..., g_{k,s}^{N_R}]^{\mathrm{T}}\}$, $\boldsymbol{\hbar}_{k,s} = [\hbar_{k,s}^1, ..., \hbar_{k,s}^{N_R}]^{\mathrm{T}}$, and $\hbar_{k,s}^r = e^{-j2\pi s\Delta f(\tau_{Br}+\tau_{rk})}$.

For the localization service, we take the SPEB as the performance metric. Specifically, we first define two position information parameters vectors $\boldsymbol{\eta}_{k,s} = [\boldsymbol{q}_k^{\mathrm{T}}, \Re\{g_{k,s}^1\}, ..., \Re\{g_{k,s}^{N_R}\}, \Im\{g_{k,s}^1\}, ..., \Im\{g_{k,s}^{N_R}\}]^{\mathrm{T}}$ and $\boldsymbol{\xi}_{k,s} = [\boldsymbol{\tau}_k^{\mathrm{T}}, \Re\{g_{k,s}^1\}, ..., \Re\{g_{k,s}^{N_R}\}, \Im\{g_{k,s}^1\}, ..., \Im\{g_{k,s}^{N_R}\}]^{\mathrm{T}}$. The transformation matrix $\boldsymbol{T}_{k,s} = \partial\boldsymbol{\eta}_{k,s}^{\mathrm{T}}/\partial\boldsymbol{\xi}_{k,s} = [\boldsymbol{Q}_k \;\boldsymbol{O}; \boldsymbol{O}\; \boldsymbol{I}_{2N_R}]$, where $\boldsymbol{Q}_k \triangleq \partial\boldsymbol{\tau}_k^{\mathrm{T}}/\partial\boldsymbol{q}_k = [\boldsymbol{x}_{1,k}, ..., \boldsymbol{x}_{N_R,k}]$ with $\boldsymbol{x}_{r,k} \triangleq 1/c[\cos\varphi_{rk}, \sin\varphi_{rk}]^{\mathrm{T}}$. The fisher information matrix (FIM) $\boldsymbol{J}_{\boldsymbol{\xi}_{k,s}}(\boldsymbol{\vartheta})$ is defined as

$$\boldsymbol{J}_{\boldsymbol{\xi}_{k,s}}(\boldsymbol{\vartheta}) \triangleq -\mathbb{E}_{y_{k,s}^l|\boldsymbol{\xi}_{k,s}}\left[\frac{\partial^2 \log f(y_{k,s}^l|\boldsymbol{\xi}_{k,s})}{\partial\boldsymbol{\xi}_{k,s}\partial\boldsymbol{\xi}_{k,s}^{\mathrm{T}}}\right] = \begin{bmatrix} \boldsymbol{G}_{k,s} & \bar{\boldsymbol{G}}_{k,s} \\ \bar{\boldsymbol{G}}_{k,s}^{\mathrm{T}} & \boldsymbol{\Sigma}_{k,s} \end{bmatrix}, \tag{5}$$

where $\log f(y_{k,s}^l|\boldsymbol{\xi}_{k,s})$ is the log-likelihood function, and

$$\boldsymbol{G}_{k,s} = \frac{8\pi^2}{\sigma^2} pW_s^2 \Re\{\boldsymbol{\Gamma}_{k,s}^{\mathrm{H}}\boldsymbol{B}_{k,s}^{\mathrm{H}}\boldsymbol{\vartheta}\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{B}_{k,s}\boldsymbol{\Gamma}_{k,s}\}, \tag{6}$$

$$\bar{\boldsymbol{G}}_{k,s} = \begin{bmatrix} \bar{\boldsymbol{G}}_{k,s}^1 & \bar{\boldsymbol{G}}_{k,s}^2 \end{bmatrix}, \; \boldsymbol{\Sigma}_{k,s} = \begin{bmatrix} \boldsymbol{\Sigma}_{k,s}^1 & \boldsymbol{\Sigma}_{k,s}^2 \\ -\boldsymbol{\Sigma}_{k,s}^2 & \boldsymbol{\Sigma}_{k,s}^1 \end{bmatrix}, \tag{7}$$

$$\begin{aligned} \bar{\boldsymbol{G}}_{k,s}^1 &= \frac{4\pi}{\sigma^2} pW_s \Re\{j\boldsymbol{\Gamma}_{k,s}^{\mathrm{H}}\boldsymbol{B}_{k,s}^{\mathrm{H}}\boldsymbol{\vartheta}\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{\Gamma}_{k,s}\}, \\ \bar{\boldsymbol{G}}_{k,s}^1 &= \frac{4\pi}{\sigma^2} pW_s \Re\{-\boldsymbol{\Gamma}_{k,s}^{\mathrm{H}}\boldsymbol{B}_{k,s}^{\mathrm{H}}\boldsymbol{\vartheta}\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{\Gamma}_{k,s}\}, \\ \boldsymbol{\Sigma}_{k,s}^1 = \frac{2}{\sigma^2} p\Re\{\boldsymbol{\Gamma}_{k,s}^{\mathrm{H}}\boldsymbol{\vartheta}\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{\Gamma}_{k,s}\}&, \boldsymbol{\Sigma}_{k,s}^2 = \frac{2}{\sigma^2} p\Re\{j\boldsymbol{\Gamma}_{k,s}^{\mathrm{H}}\boldsymbol{\vartheta}\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{\Gamma}_{k,s}\}, \end{aligned} \tag{8}$$

in which $\boldsymbol{\Gamma}_{k,s} = \mathrm{diag}\{\boldsymbol{\hbar}_{k,s}\}$ and $W_s = s\Delta f$. Then, the FIM for $\boldsymbol{\eta}_{k,s}$ can be calculated by the chain rule

$$\boldsymbol{J}_{\boldsymbol{\eta}_{k,s}} = \boldsymbol{T}_k \boldsymbol{J}_{\boldsymbol{\xi}_{k,s}} \boldsymbol{T}_k^{\mathrm{T}} = \begin{bmatrix} \boldsymbol{Q}_k\boldsymbol{G}_{k,s}\boldsymbol{Q}_k^{\mathrm{T}} & \boldsymbol{Q}_k\bar{\boldsymbol{G}}_{k,s} \\ \bar{\boldsymbol{G}}_{k,s}^{\mathrm{T}}\boldsymbol{Q}_k^{\mathrm{T}} & \boldsymbol{\Sigma}_{k,s} \end{bmatrix}. \tag{9}$$

According to the definition of equivalent FIM (EFIM) [9], we have

$$\boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\boldsymbol{\vartheta}) = \boldsymbol{Q}_k \boldsymbol{G}_{k,s} \boldsymbol{Q}_k^{\mathrm{T}} - \boldsymbol{Q}_k \bar{\boldsymbol{G}}_{k,s} \boldsymbol{\Sigma}_{k,s}^{-1} \bar{\boldsymbol{G}}_{k,s}^{\mathrm{T}} \boldsymbol{Q}_k^{\mathrm{T}}, \tag{10}$$

Finally, the SPEB for the user k is defined to be

$$\mathcal{P}(\boldsymbol{q}_k;\boldsymbol{\vartheta},\boldsymbol{\beta}) \triangleq \mathrm{Tr}\left\{\Big(\sum_{s\in\mathcal{S}} \beta_{k,s} \boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\boldsymbol{\vartheta})\Big)^{-1}\right\}, \tag{11}$$

where $\mathrm{Tr}\{\cdot\}$ denotes the matrix trace.

III. Problem Formulation and Solution

A. Problem Formulation

The localization performance maximization problem can be formulated as

$$\begin{aligned}
\mathscr{P}_{\mathbf{1}}:\ & \min_{\boldsymbol{\vartheta},\boldsymbol{\beta}} \sum_{k\in\mathcal{K}} \mathcal{P}(\boldsymbol{q}_k;\boldsymbol{\vartheta},\boldsymbol{\beta}) \\
s.t.\quad & \mathbf{C1}:\ \sum_{s\in\mathcal{S}} \beta_{k,s} R_{k,s} \geq R_k^{\min},\ \forall k\in\mathcal{K}, \\
& \mathbf{C2}:\ \beta_{k,s}\in\{0,1\},\ \sum_{k\in\mathcal{K}} \beta_{k,s}(t) \leq 1,\ \forall s\in\mathcal{S}, \\
& \mathbf{C3}:\ \theta_r = \frac{2\pi l}{Q},\ l\in\mathcal{Q},\ \forall r\in\mathcal{N}_R.
\end{aligned} \tag{12}$$

$\mathbf{C1}$ guarantees the communication performance for each user, where $R_k^{\min}$ is the minimum achievable rate requirement of user k; $\mathbf{C2}$ defines the subcarrier scheduling; the phase-shift constraints of RIS elements are given in $\mathbf{C3}$.

B. Subcarrier Assignment and Phase Shifts Design Algorithm

In this subsection, the original problem is decomposed into two subproblems, and then the subcarrier assignment $\boldsymbol{\beta}$ and RIS reflection-coefficient vector $\boldsymbol{\vartheta}$ are optimized sequentially.

1) Subproblem 1: Subcarrier Assignment: For any given RIS phase shifts $\hat{\boldsymbol{\vartheta}}$, the subcarrier assignment subproblem is reduced to

$$\begin{aligned}
\mathscr{P}_{\mathbf{1-1}}:\ & \min_{\boldsymbol{\beta}} \sum_{k\in\mathcal{K}} \mathrm{Tr}\left\{\Big(\sum_{s\in\mathcal{S}} \beta_{k,s} \boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}})\Big)^{-1}\right\} \\
s.t.\quad & \mathbf{C1}-\mathbf{C2},
\end{aligned} \tag{13}$$

To solve problem $\mathscr{P}_{\mathbf{1-1}}$ efficiently, we transform this problem into a SDP by exploiting the semi-definite property of EFIM. In particular, let $\mathrm{Tr}\{(\sum_{s\in\mathcal{S}} \beta_{k,s} \boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}}))^{-1}\} = \mathrm{Tr}\{\frac{\omega_k}{2}\boldsymbol{I}_2\}$, where ω_k is an auxiliary variable. Based on the Schur's complement lemma, problem $\mathscr{P}_{\mathbf{1-1}}$ then is equivalent to a SDP problem given by

$$\begin{aligned}
\mathscr{P}_{\mathbf{1-1'}}:\ & \min_{\boldsymbol{\beta},\boldsymbol{\omega}} \sum_{k\in\mathcal{K}} \omega_k \\
s.t.\quad & \mathbf{C1}-\mathbf{C2}, \\
& \mathbf{C4}:\ \boldsymbol{X}_k = \begin{bmatrix} \frac{\omega_k}{2}\boldsymbol{I}_2 & \boldsymbol{I}_2 \\ \boldsymbol{I}_2 & \sum_{s\in\mathcal{S}} \beta_{k,s} \boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}}) \end{bmatrix} \succeq 0, \\
& \forall k\in\mathcal{K}.
\end{aligned} \tag{14}$$

where $\boldsymbol{\omega} = [\omega_1,\omega_2,...,\omega_k]^{\mathrm{T}} \in \mathbb{C}^{K\times 1}$.

Due to the binary variable $\beta_{k,s}$, $\mathscr{P}_{\mathbf{1-1'}}$ is a mix-integer SDP, which is a NP-hard problem. Fortunately, as shown in [10], the duality gap between the dual function and the primal problem is assumed to be negligible when the number of subcarriers is greater than 32. This provides us an inspiration to solve $\mathscr{P}_{\mathbf{1-1'}}$ utilizing the Lagrangian dual method. Specifically, we first define $\boldsymbol{B}$ as the set of all possible subcarrier assignment schemes $\boldsymbol{\beta}$ satisfying $\mathbf{C2}$. As such, the Lagrange dual function of $\mathscr{P}_{\mathbf{1-1'}}$ can be written as

$$\mathcal{D}(\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda}) = \min_{\boldsymbol{\beta}\in\boldsymbol{B},\boldsymbol{\omega}} \mathcal{L}(\boldsymbol{\beta},\boldsymbol{\omega},\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda}), \tag{15}$$

where the partial Lagrange function associated with $\mathscr{P}_{\mathbf{1-1'}}$ is given by

$$\begin{aligned}
\mathcal{L}(\boldsymbol{\beta},\boldsymbol{\omega},\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda}) &= \sum_{k\in\mathcal{K}} \omega_k - \sum_{k\in\mathcal{K}} \mathrm{Tr}(\boldsymbol{Z}_k\boldsymbol{X}_k) \\
&+ \sum_{k\in\mathcal{K}} \lambda_k \left(R_k^{\min} - \sum_{s\in\mathcal{S}} \log_2(1+\frac{p|\hat{\boldsymbol{\vartheta}}^H \boldsymbol{B}_{k,s}\boldsymbol{\hbar}_{k,s}|^2}{\sigma^2})\right),
\end{aligned} \tag{16}$$

where $\boldsymbol{\lambda} = [\lambda_1,...,\lambda_K]^{\mathrm{T}} \succeq \mathbf{0}$ and $\boldsymbol{Z}_k \succeq \mathbf{0}$ are Lagrange multipliers with respect to $\mathbf{C1}$ and $\mathbf{C4}$, respectively. As such, the dual problem of problem $\mathscr{P}_{\mathbf{1-1'}}$ is given by

$$\begin{aligned}
& \max_{\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda}} \mathcal{D}(\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda}) \\
& \quad s.t.\quad \boldsymbol{\lambda}\succeq\mathbf{0},\ \boldsymbol{Z}_k\succeq\mathbf{0},\ \forall k\in\mathcal{K}.
\end{aligned} \tag{17}$$

Due to the convex dual problem (17), the primal problem $\mathscr{P}_{\mathbf{1-1'}}$ can be solved by handling the problem (17) iteratively. This thus leads to a two-layer problem, where the inner layer minimizes the Lagrange function $\mathcal{L}$ over $(\boldsymbol{\beta},\boldsymbol{\omega})$ for a given dual point $(\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda})$, while the outer layer maximizes Lagrange dual function $\mathcal{D}$ over $(\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda})$ for given $(\boldsymbol{\beta},\boldsymbol{\omega})$, until the termination requirement is achieved. The details are presented as follows.

For the given optimal primal variables $(\boldsymbol{\beta}^*,\boldsymbol{\omega}^*)$, the subgradient method mentioned in [10] can be used to find the dual point. In each iteration, the dual variables $\{\boldsymbol{Z}_k\}_{k\in K}$ and $\boldsymbol{\lambda}$ are updated based on the subgradients of $\mathcal{D}(\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda})$, we have

$$\begin{aligned}
\triangle\lambda_k &= R_k^{\min} - \sum_{s\in\mathcal{S}} \beta^*_{k,s} \log_2(1+\frac{p|\hat{\boldsymbol{\vartheta}}^{\mathrm{H}} \boldsymbol{B}_{k,s}\boldsymbol{\hbar}_{k,s}|^2}{\sigma^2}), \\
\triangle\boldsymbol{Z}_k &= \begin{bmatrix} \frac{\omega_k^*}{2}\boldsymbol{I}_2 & \boldsymbol{I}_2 \\ \boldsymbol{I}_2 & \sum_{s\in\mathcal{S}} \beta^*_{k,s} \boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}}) \end{bmatrix}.
\end{aligned} \tag{18}$$

With the subgradient projection method mentioned in [10], the Lagrange multipliers can be updated by

$$\begin{aligned}
\lambda_k(\bar{t}+1) &= [\lambda_k(\bar{t}) - \alpha_\lambda^{\bar{t}} \triangle\lambda_k]^+, \\
\boldsymbol{Z}_k(\bar{t}+1) &= [\boldsymbol{Z}_k(\bar{t}) - \alpha_{\boldsymbol{Z}}^{\bar{t}} \triangle\boldsymbol{Z}_k]^+.
\end{aligned} \tag{19}$$

Here, $\bar{t}$ is iteration index, $\alpha_\lambda^{\bar{t}}$ and $\alpha_{\boldsymbol{Z}}^{\bar{t}}$ are the positive step sizes [11], and $[x]^+ \triangleq \max(0,x)$.

Next, we derive the primal variables $(\boldsymbol{\beta},\boldsymbol{\omega})$ based on the obtained dual variables. Assume that $\boldsymbol{\omega}$ has been done, the dual function can be rewritten as

$$\begin{aligned}\mathcal{D}(\{\boldsymbol{Z}_k\}_{k\in K},\boldsymbol{\lambda}) = \min_{\boldsymbol{\beta}} \sum_{k\in\mathcal{K}}\sum_{s\in\mathcal{S}}\beta_{k,s}M_{k,s} + \sum_{k\in\mathcal{K}}\omega_k \\ + \sum_{k\in\mathcal{K}}\lambda_k R_k^{\min} - \mathrm{Tr}(\frac{\omega_k}{2}\boldsymbol{Z}_k^{(1)} + \boldsymbol{Z}_k^{(2)} + \boldsymbol{Z}_k^{(3)}),\end{aligned} \tag{20}$$

where

$$M_{k,s} \triangleq -\lambda_k \log_2(1+\frac{p|\hat{\boldsymbol{\vartheta}}^{\mathrm{H}}\boldsymbol{B}_{k,s}\boldsymbol{\hbar}_{k,s}|^2}{\sigma^2}) - \mathrm{Tr}(\boldsymbol{Z}_k^{(4)}\boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}})), \tag{21}$$

and $\boldsymbol{Z}_k^{(1)} = [\boldsymbol{Z}_k]_{1:2,1:2}$, $\boldsymbol{Z}_k^{(2)} = [\boldsymbol{Z}_k]_{1:2,2:3}$, $\boldsymbol{Z}_k^{(3)} = [\boldsymbol{Z}_k]_{3:4,1:2}$, $\boldsymbol{Z}_k^{(4)} = [\boldsymbol{Z}_k]_{3:4,3:4}$. The term $M_{k,s}$ can be interpreted as the marginal benefit of $\beta_{k,s}$ [12]. Accordingly, to minimize the Lagrangian in (20), the subcarrier assignment decision can be obtained as follows,

$$\beta^*_{k,s} = \begin{cases} 1, & \arg\min_{k'} M_{k',s}, \\ 0, & \text{otherwise}. \end{cases} \tag{22}$$

As for $\boldsymbol{\omega}$, since $\mathrm{Tr}\{(\sum_{s\in\mathcal{S}}\beta_{k,s}\boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}}))^{-1}\} = \mathrm{Tr}\{\frac{\omega_k}{2}\boldsymbol{I}_2\}$, thus, we have

$$\omega_k^* = \mathrm{Tr}\{(\sum_{s\in\mathcal{S}}\beta^*_{k,s}\boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\hat{\boldsymbol{\vartheta}}))^{-1}\}. \tag{23}$$

2) RIS Phase Design Algorithm: For a subcarrier assignment $\boldsymbol{\beta}^*$, $\mathscr{P}_1$ becomes

$$\begin{aligned}\mathscr{P}_{\mathbf{1-2}}: \ \min_{\boldsymbol{\vartheta}} \ & \sum_{k\in\mathcal{K}}\mathrm{Tr}\left\{\Big(\sum_{s\in\mathcal{S}}\beta^*_{k,s}\boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\boldsymbol{\vartheta})\Big)^{-1}\right\} \\ s.t. \quad & \mathbf{C1}: \ \sum_{s\in\mathcal{S}}\beta^*_{k,s}\log_2(1+\frac{p|\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{B}_{k,s}\boldsymbol{\hbar}_{k,s}|^2}{\sigma^2}) \geq R_k^{\min}, \\ & \mathbf{C3}: \ \theta_r = \frac{2\pi l}{Q}, \ l\in\mathcal{Q}, \ \forall r\in\mathcal{N}_R.\end{aligned} \tag{24}$$

Apparently, $\mathscr{P}_{\mathbf{1-2}}$ is an integer problem because of the constraint $\mathbf{C3}$. In order to make the problem $\mathscr{P}_{\mathbf{1-2}}$ tractable, we next introduce a penalty-based method. In particular, the penalized version of $\mathscr{P}_{\mathbf{1-2}}$ can be formulated as follows,

$$\min_{\boldsymbol{\vartheta}} \ \sum_{k\in\mathcal{K}}\mathrm{Tr}\left\{\Big(\sum_{s\in\mathcal{S}}\beta^*_{k,s}\boldsymbol{J}^{(e)}_{\boldsymbol{\eta}_{k,s}}(\boldsymbol{\vartheta})\Big)^{-1}\right\} + \varrho\sum_{k\in\mathcal{K}}\Psi_k(\boldsymbol{\vartheta}) \tag{25a}$$

$$s.t. \quad \mathbf{C2}, \tag{25b}$$

where

$$\Psi_k(\boldsymbol{\vartheta}) \triangleq \left[\Big[R_k^{\min} - \sum_{s\in\mathcal{S}}\beta^*_{k,s}\log_2(1+\frac{p|\boldsymbol{\vartheta}^{\mathrm{H}}\boldsymbol{B}_{k,s}\boldsymbol{\hbar}_{k,s}|^2}{\sigma^2})\Big]^+\right]^2,$$

and $\varrho > 0$ is the penalty coefficient that can be used to penalize the violation of inequality constraints in $\mathbf{C1}$. A two-layer iterative algorithm is proposed: for the inner layer, $\boldsymbol{\vartheta}$ solved by the penalized problem (25); while for the outer layer, we update the penalty coefficient by $\varrho \leftarrow \varepsilon\varrho$, $\varepsilon > 1$. Since the matrix inverse operation in the objective function and the integer constraint in (25b), the problem (25) is still a complicated non-convex optimization problem for any given ϱ. Nevertheless, we can find that the number of quantization bits b is usually small in practice. According to this, the one-dimensional search method can be resorted. Specifically, we can minimize (25a) by alternately searching the optimal phase-shift of each element with the other phase-shift of elements fixed, and until the convergence is reached.

IV. Simulation Results

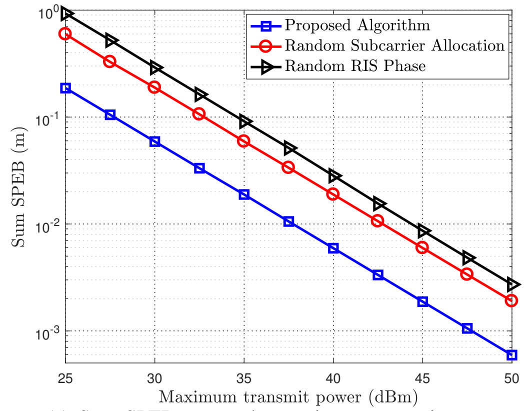
The BS is located at [-100m, 20m], the reference point of RIS is located at [0, 0], and users are randomly distributed in a circle area centered at [15m, 30m] with a radius of 10m. The subcarrier S is chosen as 40, the duality gap thus can be ignored. The phase-shift θ_r, $r\in\mathcal{N}_R$, is initialized by randomly choosing l from $\mathcal{Q}$. The simulation parameters are set as follows unless otherwise stated: $f_c = 3$ GHz, $\alpha = 2$, $\sigma^2 = -169$ dBm, $N_R = 10 \sim 50$, $W = 10$ MHz, $b = 2$, $R_k^{\min} = 100$ bps, $\varrho = 0.1$, and $\varepsilon = 1.3$.

Fig. 2 shows the impact of the maximum transmit power of the BS on the localization and communication performance, in which $N_R = 50$ and $K = 4$. We can observe that the system performances are positively correlated with the total transmit power from the BS. This is because with P_{MAX} increases, the received signal-to-noise ratio (SNR) enhances, resulting in performance improvement. More importantly, we can find that our proposed algorithm outperforms the other two schemes. It is due to the fact that the proposed algorithm performs subcarrier assignment and phase shifts design by exploiting channel gains and users' positions, which can effectively reduce the signal attenuation caused by transmission and enhance the RIS's passive beamforming gain, thereby improving the system performance. Therefore, we can conclude that subcarrier assignment and RIS phase shifts design are necessary for the presented system.

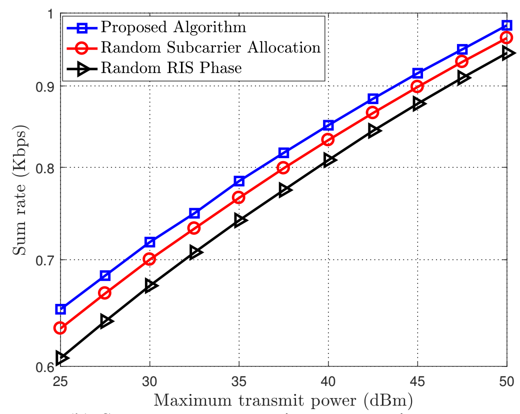
Fig. 3 depicts the impact of RIS reflecting elements number on the performance of individual user, in which $P_{\mathrm{MAX}} = 30$ dBm. It is observed that the individual localization error (achievable rate) declines (grows) and gradually flattens as the number of RIS reflecting elements increases. Such phenomenon is reasonable since a large-scale RIS can better improve the spectrum utilization of wireless communication, and gradually converges to a stable value as the size increases due to finite resources. Besides, for a given N_R, the individual user's performance will decrease as the number of users in the considered region increases. According to this, we can increase the RIS elements number to guarantee the user's performance in limited system resource cases.

V. Conclusion

This paper has proposed a novel RIS-aided OFDMA localization-communication system and investigated joint subcarrier assignment as well as RIS phase shifts optimization to improve system performance. An iterative algorithm based on Lagrange duality and penalty-based methods has been

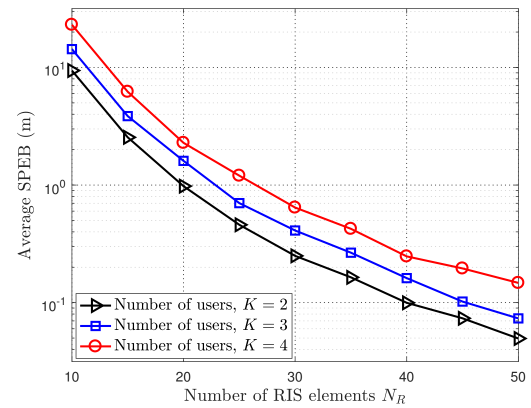


(a) Sum SPEB versus the maximum transmit power

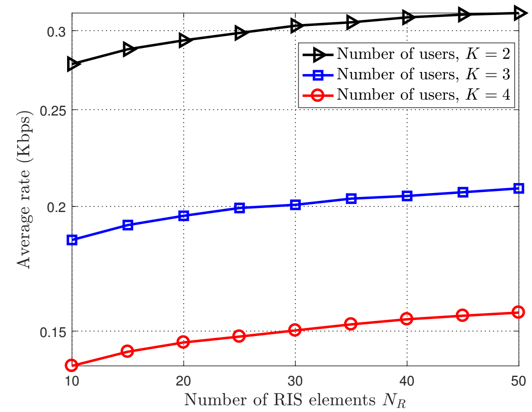


(b) Sum rate versus maximum transmit power

Fig. 2. Performance evaluations of proposed Algorithm.



(a) Average SPEB versus the number of RIS elements



(b) Average rate versus the number of RIS elements

Fig. 3. Impact of the RIS elements number on system performance

developed to implement subcarrier assignment and RIS phase shifts alternately. Simulation results have been presented to demonstrate the effectiveness of the proposed algorithm. In addition, the results unveil that, in resource-limited cases, the performance of the system can be guaranteed by increasing the size of RIS.

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