Dynamic-Subarray With Fixed Phase Shifters for Energy-Efficient Terahertz Hybrid Beamforming Under Partial CSI

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Abstract—Terahertz (THz) communications are regarded as a pillar technology for the 6G systems, by offering multi-ten-GHz bandwidth. To overcome the huge propagation loss, THz ultra-massive MIMO systems with hybrid beamforming are proposed to offer high array gain. Notably, the adjustable phase shifters considered in most existing hybrid beamforming studies are power-hungry and difficult to realize in the THz band. Moreover, due to the ultra-massive antennas, full channel-state-information (CSI) is challenging to obtain. To address these practical concerns, in this paper, an energy-efficient dynamic-subarray with fixed phase shifters (DS-FPS) architecture is proposed for THz hybrid beamforming. To compensate for the spectral efficiency loss caused by the fixed phase of FPS, a switch network is inserted to enable dynamic connections. In addition, by considering the partial CSI, we propose a row-successive-decomposition (RSD) algorithm to design the hybrid beamforming matrices for DS-FPS. A row-by-row (RBR) algorithm is further proposed to reduce the computational complexity. Extensive simulation results show that, the proposed DS-FPS architecture with the RSD and RBR algorithms achieves much higher energy efficiency than the existing architectures. Moreover, the spectral efficiency of the DS-FPS architecture with the proposed algorithms is robust to the CSI error.

Index Terms—Terahertz communications, hybrid beamforming, dynamic-subarray, fixed phase shifters, partial CSI.

I. INTRODUCTION

In order to meet the rapid growth of wireless data rates, the Terahertz (THz) band with ultra-wide bandwidth has gained increasing attention [2]. In the THz band, the ultra-wide bandwidth comes with a cost of huge propagation loss, which drastically limits the communication distance [3]. Fortunately, the sub-millimeter wavelength allows the design of array consisting of a large number of antennas at transceivers, e.g., 1024, to enable THz ultra-massive MIMO (UM-MIMO) systems [4]. By utilizing the ultra-massive antennas, the beamforming technology can provide a high array gain to compensate for the path loss and combat the distance problem. Meanwhile, multiple data streams can be supported to offer a multiplexing gain and further improve the spectral efficiency. In the THz UM-MIMO systems, many hardware constraints preclude from using conventional digital beamforming, which, instead, motivates the appealing hybrid beamforming [5], [6], [7]. The hybrid beamforming divides signal processing into the digital baseband domain and analog radio-frequency (RF) domain, which can achieve high spectral efficiency while maintaining low hardware complexity [8], [9].

The fully-connected (FC) and array-of-subarrays (AoSA) architectures are two widely-studied hybrid beamforming architectures [10], [11], [12], [13]. The FC architecture achieves high spectral efficiency while consuming high power [10]. On the contrary, as illustrated in Fig. 1(a), the complexity and power consumption of AoSA are noticeably reduced, while the spectral efficiency is sacrificed [13]. As a result, the energy efficiency, which is defined as the ratio between the spectral efficiency and power consumption, of both these two architectures is unsatisfactory and needs to be enhanced. Moreover, in THz UM-MIMO systems, due to the large number of antennas and the high carrier frequency, full channel state information (CSI) is difficult to acquire. To address these two practical concerns, we develop a novel energy-efficient hybrid beamforming architecture for THz UM-MIMO systems based on partial CSI in this work.

A. Related Work

1) Energy-Efficient Hybrid Beamforming Architecture: There have been many studies aiming at improving the energy efficiency of the hybrid beamforming systems. The authors in [14] proposed to jointly optimize the hybrid beamforming matrices and the resolution of DAC/ADC to enhance the energy efficiency. Another promising direction is utilizing the low-cost switches to provide dynamic connections.
In [15] and [16], the RF chains of the hybrid beamforming architecture are dynamically deactivated to reduce the power consumption and enhance the energy efficiency. Except from the RF chain selection, there have also been many efforts on the dynamic connections between phase shifters and antennas [17], [18], [19], [20], [21], [22]. By inserting switches in the FC architecture, a fully-adaptive-connected (FAC) architecture [17] was proposed, where partial phase shifters are inactive to reduce power consumption. However, the number of remaining active phase shifters is usually larger than the number of antennas, which causes high power consumption. A dynamic hybrid beamforming (DHB) architecture was proposed [18], [19], and [20], as shown in Fig. 1(b). Through the switch network, each antenna dynamically selects one RF chain to connect with to enhance the spectral efficiency. Moreover, the number of phase shifters in the DHB architecture equals the number of antennas, which is less than the FAC architecture and leads to higher energy efficiency.

One remaining problem of the DHB architecture is that the phase shifters are assumed to own high resolution and even infinite resolution, which are impractical and power-hungry. To address this problem, the authors of [21] and [22] proposed to use low-resolution phase shifters in the DHB architecture. However, both the above high-resolution and low-resolution phase shifters in [17], [18], [19], [20], [21], and [22] are still adjustable phase shifters, i.e., phase selection is adjustable, which has high power consumption in the THz band. Fortunately, a fixed phase shifter (FPS) can be adopted, whose phase remains fixed and non-adjustable. Compared to adjustable phase shifters, the THz FPS has substantially lower power consumption, which is more practical. The authors of [23] and [24] proposed to use FPSs, instead of adjustable phase shifters, in the hybrid beamforming architecture, where each antenna is connected with multiple FPSs through switches. As a result, the number of closed switches is usually several times of the number of antennas, which is unbearably large in THz UM-MIMO systems and causes huge power consumption. Therefore, to utilize low-cost FPSs while keeping low power consumption, we propose a novel DS-FPS architecture, as shown in Fig. 1(c). In the proposed DS-FPS architecture, each RF chain connects with multiple FPSs. Each antenna can dynamically select one FPS through one switch such that the number of closed switches equals the number of antennas, which is much smaller than the number of closed switches in [23], [24]. As a result, the energy efficiency of the proposed DS-FPS architecture is improved.

2) Partial CSI: Most of the existing hybrid beamforming studies assume that full CSI is known at both transmitter and receiver. However, the full CSI is hard to obtain in THz UM-MIMO systems due to the large-dimensional channel matrix caused by ultra-massive antennas. To tackle this problem, the authors of [25] considered to design the hybrid beamforming based on partial elements of the channel matrix. Since the overall dimension of the channel matrix in THz UM-MIMO systems is prohibitively large, acquiring partial elements of the channel matrix is still difficult. The authors of [19], [26], and [27] proposed the hybrid beamforming solutions considering the statistical information of the channel. However, due to the lack of well-known general statistical MIMO channel model in the THz band yet, it is also challenging to know the statistical channel information. A more practical partial CSI was considered in [28] and [29], where only the directions and the amplitude of the path gains of the multipath are known. However, the hybrid beamforming algorithms in [28] and [29] were proposed for the FC architecture with adjustable phase shifters and did not consider the use of low-cost FPSs, thus cannot be applied to the proposed DS-FPS architecture in this work. Consequently, by considering the practical partial CSI, novel hybrid beamforming algorithms need to be proposed for the DS-FPS architecture.

B. Our Contributions

In this work, we propose an energy-efficient DS-FPS architecture, by utilizing the low-cost FPS and dynamic switch network, as shown in Fig. 1(c). Moreover, we consider the practical partial CSI, i.e., only the directions...
and the amplitude of the path gains of the multipath are known. Since the number of multipath of THz channel is usually small [30], the number of required parameters in the considered partial CSI is very limited. Furthermore, by considering partial CSI, we propose two CSI-robust hybrid beamforming algorithms for the THz DS-FPS architecture. In the prior and shorter version of this work [1], we concisely investigated the DS-FPS architecture, while the consideration of partial CSI, the corresponding hybrid beamforming algorithms, and the performance comparisons with existing work in terms of spectral and energy efficiencies were not thoroughly studied. The distinctive features of this work are summarized as follows.

- We propose an energy-efficient DS-FPS architecture, by using the low-cost FPS and switch network. With a fixed and nonadjustable phase, the FPS is more practical and consumes less power than the adjustable phase shifter in the THz band, which however, brings spectral efficiency loss. To address this problem, we further design a switch network to enable dynamic connections between the antennas and FPSs. Each antenna can intelligently select one FPS with the proper phase from all FPSs to adapt the THz UM-MIMO channel, which enhances the spectral efficiency.

- By considering partial CSI, we formulate the hybrid beamforming problem for the DS-FPS architecture and propose two hybrid beamforming algorithms. Specifically, we first propose a row-successive-decomposition (RSD) algorithm. The key idea is deriving an approximated form of the spectral efficiency, which only relies on partial CSI, and then optimizing each row of the switch network matrix successively. Furthermore, to reduce the computational complexity brought by the successive design, we propose a row-by-row (RBR) algorithm, which decomposes the optimization of each row of the switch network matrix as multiple uncorrelated sub-problems and solves them in parallel.

- We evaluate the performance of the proposed DS-FPS architecture with the RSD and RBR algorithms and analyze the computational complexity. Specifically, we show that the DS-FPS architecture achieves significantly higher energy efficiency than the existing architectures. Moreover, with partial CSI, the spectral efficiency of the RSD and RBR algorithms is similar to the case of full CSI and is robust to the CSI error. Furthermore, we analyze that the computational complexity of both the RSD and RBR algorithms is lower than the existing counterpart algorithms.

The remainder of this paper is organized as follows. In Sec. II, we present the channel model and system model, and formulate the hybrid beamforming problem for the THz DS-FPS architecture. Then, an RSD algorithm and a low-complexity RBR algorithm are proposed to solve the DS-FPS hybrid beamforming problem in Sec. III and Sec. IV, respectively. Furthermore, simulation results are provided in Sec. V. Finally, the conclusion is drawn in Sec. VI.

**Notations:** A is a matrix, a is a vector, α is a scalar. \( I_N \) denotes an \( N \)-dimensional identity matrix. \((\cdot)^T\), \((\cdot)^*\), and \((\cdot)^H\) represent transpose, conjugate, and conjugate transpose. \( \| \cdot \|_p \) is the \( p \)-norm of the vector. \( \| \cdot \|_F \) is the Frobenius norm of the matrix. \( \text{Tr}(\cdot) \) and \( \text{Re}(\cdot) \) denote the trace and real part of the matrix. \( \text{blkdiag}(\cdot) \) denotes the block diagonal matrix. \( \odot \) is the element-wise product. \( \otimes \) represents the Kronecker product.
A \theta_{1i} \text{ denote the azimuth and elevation direction of departure (DoD). Moreover, } d \text{ is the antenna spacing, which is set as half of the wavelength of the central frequency in this work.}

1) Partial CSI: The channel matrix \( \mathbf{H}[k] \) is composed by the DoA, the DoD, and the complex path gain of each path. There have been many studies that jointly estimate the DoA, DoD, and the path gain at either transmitter or receiver [31]. After the feedback, both the transmitter and receiver know the DoA, DoD, and path gain. In this work, we consider the partial CSI scenario, where the transmitter knows the DoD and amplitude of path gain, i.e., \( \mathbf{A}_t[k] \) and \( \bar{\Lambda}[k] \) in (1c), while the receiver knows the DoA and amplitude of path gain, i.e., \( \mathbf{A}_r[k] \) and \( \bar{\Lambda}[k] \), respectively. There have been multiple low-complexity methods [32], [33] which can estimate the DoA at the receiver and estimate the DoD at the transmitter, respectively. During the estimation of DoA and DoD, the amplitude of path gain can also be acquired. Consequently, the considered partial CSI in this work is practical. For sparse channels, e.g., the THz channel in this work, the DoD and DoA are demanding information for hybrid beamforming design. While for rich scattering channels, it has been studied in [34] that the DoD and DoA can be mapped to the correlation matrix which is a realistic requirement for the hybrid beamforming design to achieve high spectral efficiency.

B. System Model of DS-FPS Hybrid Beamforming

Most of the existing hybrid beamforming studies [10], [11], [12], [13], [17], [18], [35] use adjustable phase shifters. For THz communications, due to the higher frequency, the power consumption of adjustable phase shifters becomes prohibitively large and thus, impractical to use. To address this problem, we use low-cost FPSs to construct a DS-FPS architecture, as shown in Fig. 1(c).

We set the number of RF chains as \( L_t \) and each RF chain connects with \( Q \) FPSs. The provided phases of the FPSs are fixed as \( \Phi_1, \Phi_2, \ldots, \Phi_Q \), respectively. One drawback of FPSs is that the provided phase is fixed while the required phase to steer beams varies with different channels. To overcome this drawback, we propose a switch network to enable dynamic connection, where each antenna can select one FPS from all \( L_tQ \) FPSs to connect with, i.e., dynamically selects one proper phase from \( \Phi_1, \Phi_2, \ldots, \Phi_Q \) to perform analog beamforming. Since the required phase of analog beamforming at each antenna may be an arbitrary value in \([0,2\pi)\) when channel varies, we set the phases of FPSs uniformly located between 0 and \( 2\pi \), i.e., \( \Phi_i = \frac{2\pi(i-1)}{Q} \) for \( i = 1,2,\ldots,Q \), to make the beamforming weight error at each antenna no larger than \( \frac{\pi}{Q} \). The system model of the DS-FPS hybrid beamforming architecture at the \( k \)-th subcarrier can be expressed as

\[
y[k] = \mathbf{W}[k]^H \mathbf{H}[k] \mathbf{SFD}[k] \mathbf{s}[k] + \mathbf{W}[k]^H \mathbf{n}[k],
\]

where \( \mathbf{s}[k] \in \mathbb{C}^{N_r \times 1} \) and \( y[k] \in \mathbb{C}^{N_r \times 1} \) denote the transmitted and received signals. \( N_r \) is the number of data streams. \( \mathbf{n}[k] \in \mathbb{C}^{N_r \times 1} \) is the noise vector. \( \mathbf{N}_t \times L_tQ \times L_t \)-dimensional binary matrix \( \mathbf{S} \) and \( L_tQ \times L_t \)-dimensional matrix \( \mathbf{F} \) represent the switch network matrix and the phase matrix of the FPS network at the transmitter, respectively. The phase of one FPS is the same for each subcarrier and the state of switch is identical for each subcarrier. Hence, the frequency index \( [k] \) is omitted in \( \mathbf{F} \) and \( \mathbf{S} \). As a result, \( \mathbf{F} \) can be written as

\[
\mathbf{F} = \text{blkdiag}(\mathbf{f}_1, \ldots, \mathbf{f}_{L_t}),
\]

where \( \mathbf{f} = [e^{j\Phi_1}, e^{j\Phi_2}, \ldots, e^{j\Phi_Q}]^T \) represents the phase vector generated by \( Q \) FPSs of each RF chain. Since each antenna only selects one FPS from \( L_tQ \) FPSs through one closed switch, each row of \( \mathbf{S} \) has only one ‘1’ and other elements are ‘0’s, i.e., \( [\mathbf{S}]_{i0} = 1, i = 1,2,\ldots,N_r \), where \( \mathbf{S} \) denotes the \( i \)-th row of \( \mathbf{S} \). \( \mathbf{D}[k] \in \mathbb{C}^{L_t \times N_r} \) represents the digital beamforming matrix. The transmit power constraint is enforced on \( \mathbf{D}[k] \) as

\[
\| \mathbf{SFD}[k] \|_{\mathbf{F}}^2 = \rho_k \quad \text{and} \quad \sum_{k=1}^{K} \rho_k = \rho, \quad \text{where } \rho_k \text{ is the transmit power at the } k\text{-th subcarrier and } \rho \text{ is the total transmit power of all subcarriers.}
\]

In this work, we consider that \( \rho_k = \rho/K \) and the power allocation across different subcarriers can be a future research direction. \( \mathbf{SFD}[k] \) is the preceding matrix of DS-FPS architecture at the transmitter. \( \mathbf{W}[k] \in \mathbb{C}^{N_r \times N_r} \) is the combining matrix at receiver.

Compared to the existing dynamic architectures in [17], [18], [19], [20], [21], and [22], the major novelty of the DS-FPS architecture is that we use the low-cost FPSs rather than the adjustable phase shifters, which causes that the power consumption of the DS-FPS architecture is lower than the existing dynamic architectures in [17], [18], [19], [20], [21], and [22]. The DS-FPS indeed tackles the power consumption challenge of the THz band, although its spectral efficiency is lower than some existing architectures, e.g., the FC architecture. However, at mmWave systems, since the power consumption of adjustable phase shifters is still acceptable, as considered by most existing studies, it may be unnecessary to use the DS-FPS for enhancing energy efficiency at the cost of spectral efficiency degradation. Hence, the DS-FPS architecture is THz-specific. Note that the existing study [23] also uses FPSs and switch network to realize the hybrid beamforming. The analysis in [23] focuses on the case that \( L_t \geq KN_r \) and the authors prove that the hybrid beamforming architecture in [23] can approach the optimal digital beamforming with sufficient number of FPSs. In this work, we mainly consider a different case that \( L_t \) is similar with \( N_r \), which is a common practice for hybrid beamforming studies [11], [12], [13]. Moreover, the partial CSI has not been analyzed in [23].

C. Design Problem: Maximize the Spectral Efficiency With Partial CSI

The spectral efficiency of the DS-FPS architecture can be expressed as [36]

\[
\begin{align*}
SE &= \frac{1}{K} \sum_{k=1}^{K} \log_2 \left( \left| \mathbf{W}[k] \right|^H \left( \mathbf{W}[k]^H \mathbf{W}[k] \right)^{-1} \mathbf{SFD}[k] \mathbf{D}[k] \mathbf{F}^H \mathbf{H}^H \mathbf{H}[k]^H \mathbf{SFD}[k] \right) \\
&= \frac{1}{K} \sum_{k=1}^{K} \log_2 \left( \left| \mathbf{W}[k] \right|^H \mathbf{W}[k] \right) - \log_2 \left( \left| \mathbf{W}[k] \right|^H \mathbf{F}^H \mathbf{H}^H \mathbf{H}[k]^H \mathbf{SFD}[k] \right),
\end{align*}
\]

where \( \sigma_k^2 \) denotes the noise power of the \( k \)-th subcarrier. \( \mathbf{W}[k] = \mathbf{S}_r \mathbf{F}_r \mathbf{D}_r \) is the combining matrix, with \( \mathbf{S}_r, \mathbf{F}_r, \) and \( \mathbf{D}_r \) being the switch network matrix, the phase matrix, and
the digital beamforming matrix of the DS-FPS architecture at the receiver, which have the similar expressions with those of at the transmitter. The transmitter design and receiver design are usually coupled, which brings an obstacle to solve the design problem. To make the design more tractable, it is a common practice to decouple the design of these two terms in the existing hybrid beamforming studies. Specifically, when designing the precoding matrix at transmitter, the optimal combining is assumed at the receiver and the influence of the combining matrix to (5) is omitted [10], [13], [36]. As a result, the design problem of DS-FPS architecture at the transmitter can be formulated as [10], [13], and [36]

$$\max_{\mathbf{S}, \mathbf{D}} \frac{1}{K} \sum_{k=1}^{K} \log_2 \left( |\mathbf{I}_N + \frac{1}{\sigma_k^2} \mathbf{H}[k]|\mathbf{SFD}[k] \mathbf{D}[k] \mathbf{H}^H \mathbf{S}^H \mathbf{H}[k] |^2 \right)$$

s.t. $\mathbf{S}_{i,l} \in \{0, 1\}, \|\mathbf{S}_{i,l}\|_0 = 1, \forall i, l$

$$\|\mathbf{SFD}[k]\|_F^2 = \rho_k,$$

where $\mathbf{S}_{i,l}$ denotes the element at the $i$th row and the $l$th column of $\mathbf{S}$.

In this section, we propose an RSD algorithm to solve the problem (6). One main challenge is that the objective function is related to the full CSI $\mathbf{H}[k]$, while the transmitter only knows $\mathbf{A}_r[k]$ and $\bar{\mathbf{A}}[k]$. To tackle this problem, we first derive an approximated form of the objective function (6a), which only relies on $\mathbf{A}_r[k]$ and $\bar{\mathbf{A}}[k]$. Then, we decompose the rows of switch network matrix $\mathbf{S}$ successively to transform the intractable design problem as multiple tractable sub-problems, to overcome the non-convex binary constraint of $\mathbf{S}$.

### III. R OW-SUCCESSIVE-DECOMPOSITION (RSD) ALGORITHM

In this section, we propose an RSD algorithm to solve the problem (6). One main challenge is that the objective function is related to the full CSI $\mathbf{H}[k]$, while the transmitter only knows $\mathbf{A}_r[k]$ and $\bar{\mathbf{A}}[k]$. To tackle this problem, we first derive an approximated form of the objective function (6a),
Therefore, in THz UM-MIMO systems with ultra-massive antenna arrays, the approximation in (11b) follows the mixed-product property of the Kronecker product. \( \Psi_1 = \sin(\phi_r)\sin(\theta_r) - \sin(\phi_i)\sin(\theta_i) \), \( \Psi_2 = \cos(\phi_r) - \cos(\phi_i) \). (11d) comes from the fact that \( \sin(\frac{2\pi}{\lambda}dL\Psi_1) \leq 1 \) and \( \sin(\frac{2\pi}{\lambda}dW\Psi_2) \leq 1 \). Since \( i \neq l \), we usually have \( \Psi_1 \neq 0 \), \( \Psi_2 \neq 0 \), and \( \sin(\frac{2\pi}{\lambda}dL\Psi_1)\sin(\frac{2\pi}{\lambda}dW\Psi_2) \neq 0 \). Therefore, in THz UM-MIMO systems with ultra-massive antennas, \( N_r \geq 1024 \), \( \frac{1}{N_r} \sin(\frac{2\pi}{\lambda}dL\Psi_1)\sin(\frac{2\pi}{\lambda}dW\Psi_2) \) is very close to 0 and the approximation \( \frac{1}{N_r}[a_{ri}[k]H a_{rl}[k]] \approx 0 \) holds. The special case is that when the directions of the \( i \)th path and the \( l \)th path are similar, \( \Psi_1 \) and \( \Psi_2 \) are small such that \( \sin(\frac{2\pi}{\lambda}dL\Psi_1)\sin(\frac{2\pi}{\lambda}dW\Psi_2) \) is close to 0 and the approximation error of \( \frac{1}{N_r}[a_{ri}[k]H a_{rl}[k]] \approx 0 \) is large.

We further assess the approximation error in Fig. 2. As shown in Fig. 2(a) and Fig. 2(b), we plot \( \frac{1}{N_r}[a_{ri}[k]H a_{rl}[k]] \) versus the direction of the \( l \)th path, i.e., \( \phi_{rl} \) and \( \theta_{rl} \), where \( \phi_{rl} = 30^\circ \) and \( \theta_{rl} = 60^\circ \). Due to the symmetrical property of the sinusoidal function, we only consider the cases that \( 0^\circ \leq \phi_{rl} \leq 90^\circ \) and \( 0^\circ \leq \theta_{rl} \leq 90^\circ \), while the remaining cases are similar and extensible. For most angles of \( \phi_{rl} \) and \( \theta_{rl} \) which locate in the blue region, \( \frac{1}{N_r}[a_{ri}[k]H a_{rl}[k]] \approx 0 \). When \( \phi_{rl} \) and \( \theta_{rl} \) are very close to \( \phi_{ri} \) and \( \theta_{rl} \), i.e., the \( l \)th path locates in the highlighted region \( \{(\phi_{rl}, \theta_{rl}) | \phi_{rl} - 4^\circ < \phi_{rl} < 4^\circ \land 0^\circ < \theta_{rl} < 90^\circ \}$, \( \frac{1}{N_r}[a_{ri}[k]H a_{rl}[k]] \) is not close to 0. Note that the blue region accounts for a very large proportion of the whole space and the THz channel is usually sparse. Hence, the probability that the \( l \)th path with \( l \neq i \) locates in the highlighted region is very small. Through the above illustrations, we can state that the approximation \( \frac{1}{N_r}[a_{ri}[k]H a_{rl}[k]] \approx 0 \), i.e., \( \frac{1}{N_r}a_{ri}[k]H a_{rl}[k] \approx 0 \), holds reasonably well for THz UM-MIMO systems. For mmWave frequencies, with richer multipath and higher spatial degree of freedom, the approximation error usually increases. For sub 6-GHz band with rich-scattering, the number of paths is very large and the approximation usually not holds. Hence, for mmWave and sub 6-GHz bands, the performance of the proposed algorithms in this work may become worse.

B. Design of Digital Beamforming Matrix \( \mathbf{D}[k] \)

Till now, we obtain an approximated form (9e) of the original objective function (6a), which relies on the partial CSI \( \bar{\mathbf{A}}[k] \) and \( \mathbf{A}_i[k]^H \). Next, we present how to use (9e) to design \( \mathbf{D}[k] \) by assuming that \( \mathbf{S} \) has been determined. The maximization of the original objective (6a) can be transformed as the maximization of (9e) as

\[
\max_{\mathbf{S}, \mathbf{D}[k]} \frac{1}{K} \sum_{k=1}^{K} \log_2 \left( \left| |\mathbf{I}_{N_r} + \frac{N_r}{\sigma_k^2} \bar{\mathbf{A}}[k]\mathbf{A}_i[k]^H \mathbf{S}\mathbf{F}^H \mathbf{H}^H \mathbf{A}_i[k]\bar{\mathbf{A}}[k]^H \right| \right),
\]

(12)

where the constraints of \( \mathbf{S} \) and \( \mathbf{D}[k] \) are the same as in (6). By omitting the transmit power constraint temporarily, the solution of \( \mathbf{D}[k] \) to maximize (12) is

\[
\mathbf{D}[k] = \bar{\mathbf{V}}_{N_r}[k] \hat{\Gamma}[k],
\]

(13)

where \( \bar{\mathbf{V}}_{N_r}[k] \) is the first \( N_s \) columns of \( \bar{\mathbf{V}}[k] \), which comes from the singular value decomposition (SVD) of \( \mathbf{H}_i[k] = \bar{\mathbf{A}}[k]\mathbf{A}_i[k]^H \mathbf{S}\mathbf{F}^H \mathbf{H}^H \mathbf{A}_i[k] \bar{\mathbf{A}}[k]^H \). Moreover, \( \hat{\Gamma}[k] \) is the power allocation matrix, for which the water-filling allocation is the optimal. Despite so, to reduce the computational complexity, we consider the more practical equal-power allocation such that \( \hat{\Gamma}[k] = \sqrt{\frac{P}{N}} \mathbf{I}_{N_r} \). The consideration of water-filling power allocation and the corresponding analog and digital beamforming solution can be considered in the future work.
C. Design of Switch Network Matrix $S$

Next, we design $S$ to maximize (12). The successive decomposition is an effective method to transform the complicated matrix design problem into multiple vector design problems [13], [39]. Inspired by this, we aim to first remove the impact of $D[k]$ and then use the successive decomposition idea to design $S$. By substituting the solution of $D[k]$ in (13) into (12), the $k^{th}$ term of (12) is derived as

$$\log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD [D[k]^H S^H A_k \tilde{\Lambda}_k [A_k]^H] \right)$$

(14a)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} H_{k} [\nu_N, H_{k} [A_k]^H] \right)$$

(14b)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Sigma}_k [\tilde{V}_N, \tilde{V}_e]^H \right)$$

(14c)

$$\leq \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Sigma}_k [\tilde{V}_N, \tilde{V}_e]^H \right)$$

(14d)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD [D[k]^H S^H A_k \tilde{\Lambda}_k [A_k]^H] \right).$$

(14g)

From (14b) to (14g), for $\tilde{U}_k, \tilde{V}_k, [\tilde{V}_N, [\tilde{V}_N, [\tilde{V}_e]$, we omit the index $[k]$ for simplicity. (14c) is the result of applying the SVD of $H_{k} [\nu_N, H_{k} [A_k]^H]$ and $\tilde{\rho}_k = \frac{N_p}{\sigma_k^2}$. (14d) comes from the property that $\log_2 (I + XY) = \log_2 (I + YX)$, where $X = \tilde{\Sigma}_k [\tilde{V}_N, \tilde{V}_e]^H$. Moreover, $\tilde{V}_N$, and $\tilde{V}_e$ denote the first $N_s$ columns and the remaining columns of $\tilde{V}$, respectively. (14e) follows the property of SVD that $\tilde{U}^H \tilde{U} = I$, $\tilde{V}^H \tilde{V}_N = I$, and $\tilde{V}^H \tilde{V}_e = 0$, where $\tilde{V}_N$ represents the first $N_r$ rows and columns of $\tilde{V}$. In the hybrid beamforming system, the number of RF chains $L_r$ is usually larger than or equal to the number of data streams $N_s$ and smaller than the number of multipath $N_p$. Therefore, we use (14f) as an upper bound of (14g), where $\hat{\Sigma}_{L_r}$ represents the first $L_r$ rows and columns of $\hat{\Sigma}$ and the equality holds when $N_s = L_r$. The dimension of $H_{r} [\nu_N]$ is $N_p \times L_r$ such that $H_{r} [\nu_N]$ has at most $L_r$ non-zero singular values, which suggests that $\hat{\Sigma}_{L_r}$ contains all the non-zero singular values of $H_{r} [\nu_N]$. According to the property of SVD, we obtain $\log_2(\frac{N_p}{\sigma_k^2}) = \log_2(\frac{N_p}{\sigma_k^2} A_k [A_k]^H)$, which equals to (14g).

So far, we have derived an upper bound (14g) for (14a), which is uncorrelated with $D[k]$ and $S$ is mitigated. Next, we propose to design $S$ to maximize (14g) rather than directly maximizing (14a). One main difficulty to solve $S$ is the binary constraint (6b) on each row. The authors in [13] proposed to successively decompose the hybrid beamforming matrix in the column-manner to enable the design. Inspired by this, to tackle the row-manner constraint of $S$, we decompose each row of $S$ successively. Although the idea of successive decomposition is similar, the detailed constraints and derivations of the row-manner decomposition in this work are quite different from the column-manner decomposition in [13]. Moreover, some important procedures of the RSD algorithm, including the derived approximated spectral efficiency (9e), the solution of $D[k]$ as well as the derived upper bound (14g) on the approximated spectral efficiency, and the solution of each row of $S$ have not been studied in [13]. To start with, (14g) can be rewritten as

$$\log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD [D[k]^H S^H A_k \tilde{\Lambda}_k [A_k]^H] \right)$$

(15a)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD [D[k]^H S^H A_k] \right)$$

(15b)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD \left( \left[ \tilde{C}_1 N_s - 1 [k], \tilde{C}_N, [k] \right] \right) \right)$$

(15c)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD \left( \left[ \tilde{C}_1 N_s - 1 [k], \tilde{C}_N, [k] \right] \right) \right)$$

(15d)

$$= \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD \left( \left[ \tilde{C}_1 N_s - 1 [k], \tilde{C}_N, [k] \right] \right) \right)$$

(15e)

In (15c), we use $C[k]$ to represent $A_i[k]^H$, where $C_1[k]$ and $C_q[k]$ denote the first $p$ columns and the $q^{th}$ column of $C[k]$, respectively. $S_1$ and $S_q$ denote the first $p$ rows and the $q^{th}$ row of $S$, respectively. In (15d), $A[k] = C_1 N_s - 1 [k] S_1$ and $B[k] = C_{N_s} [k] S_{N_s} F$. (15e) comes from the property of determinant that $|I + X + Y| = |I + X + (I + Y)|$, where $X = \rho_k [\tilde{A}_k]^2 [A_k]^H$, $Y = \tilde{\rho}_k \tilde{A}_k^2 [B[k]^H + A[k]^H] + A[k]^H$, and $T[k] = I_{N_s} + \rho_k \tilde{A}_k [A_k]^H$. We observe that, by substituting $A[k] = C_1 N_s - 1 [k] S_1$, the first term in (15e) can be further expressed as $\log_2(\frac{N_p}{\sigma_k^2}) = \log_2(\frac{N_p}{\sigma_k^2} C_1 N_s - 1 [k] S_1)$, which has the similar structure with (15b). Therefore, we can continue to decompose the first term of (15e) with the similar procedures from (15b) to (15e). After $N_t$ times, (15e) can be represented as

$$\sum_{i=1}^{N_t} \log_2 \left( \frac{N_p}{\sigma_k^2} \tilde{\Lambda}_k [A_k]^H SFD \left( \left[ \tilde{C}_1 N_s - 1 [k], \tilde{C}_N, [k] \right] \right) \right),$$

(16)

where $A_i[k] = C_{1_i} N_s - 1 [k] S_{1_i}$ and $B_i[k] = C_i[k] S_{1_i} F$. $T_i[k] = I_{N_s} + \rho_k \tilde{A}_k [A_k]^H$. When $i = 1$, $A_1[k] = 0$ and $T_1[k] = I_{N_s}$. We have now transformed (14a), which is the $k^{th}$ term of (12), to (16). Recall that we aim to design $S$ to maximize (12). Hence, designing $S$ to maximize (12) is transformed as designing $S$ to maximize the summation of (16) about $k$, which is given at the bottom of the previous page.

For each $i$, we observe that $T_i[k], A_i[k],$ and $B_i[k]$ are only related with the first $i^{th}$ rows of $S$ but not related with the remaining rows. According to this property, we propose the following $N_t$-stage method to design $S$ to maximize (17), shown at the bottom of the next page. At the first stage, we design $S_1$ to maximize $\ast$ with $i = 1$. At the second stage, with the determined $S_1$ at the first stage, we design $S_2$ to maximize $\ast$ with $i = 2$. Following this trend, at the $N_t^{th}$ stage, the first $N_t - 1$ rows of $S$ have been determined and
we design $S_{N_t}$ to maximize $\langle \star \rangle$ with $i = N_t$. After that, all rows of $S$, i.e., the whole $S$, can be determined.

Next, we present how to design $S_i$ to maximize $\langle \star \rangle$ for each stage. Without loss of generality, we use the $i^{th}$ stage as an illustration as below.

$$\max_{S_{i_*}} \sum_{k=1}^{K} \log_2 \left( \left| N_p + \rho_k T_i[k] \right|^{-1} \bar{\Lambda}[k] \right)^2 (B_i[k] A_i[k])^H$$

$$+ A_i[k] (B_i[k] B_i[k]^H + B_i[k] B_i[k]^H))$$

$$\text{s.t.} \ S_{i_*,i} \in \{0,1\}, \|S_{i_*}\|_0 = 1, \forall l.$$  (18a)

Due to the structure of $F$ in (4), all the diagonal elements of $FF^H$ are 1. There is only one ‘1’ in $S_{i_*}$ while the other elements are ‘0’s. As a result, regardless how we design $S_{i_*,i}$, $S_i FF^H S_i^H$ equals 1, i.e., $B_i[k] B_i[k]^H = C_{i_*} [k] S_i F F_i S_i^H C_i[k] = C_{i_*} [k] C_i[k]$. Consequently, by substituting $B_i[k] = C_{i_*} [k] S_i F$ and $B_i[k] B_i[k]^H = C_{i_*} [k] C_i[k]$ into (18a), (18b) can be expressed as

$$\sum_{k=1}^{K} \log_2 \left( \left| N_p + \rho_k T_i[k] \right|^{-1} \bar{\Lambda}[k] \right)^2 (C_i[k] S_i F A_i[k])^H$$

$$+ A_i[k] (P^H S_i^H C_i[k]^H + C_i[k] C_i[k]^H))$$

(19)

Since there is only one ‘1’ in $S_{i_*}$, the possible $S_{i_*}$ has only $L_t Q$ choices, where $L_t Q$ is the number of FFSs in the DS-FPS architecture. The number of FFSs is usually limited. Hence, it is reasonable and efficient to use the exhaustive search method to find the optimal $S_{i_*}$ to maximize (19). The flow and pseudocodes of the RSD algorithm are presented in Algorithm 1. After the design of $S$ and $D[k]$, we normalize $D[k]$ to satisfy the transmit power constraint. Note that the value of $\rho_k$ influences the performance of the RSD algorithm. $\rho_k = \frac{\rho_k N_t}{P_{\text{max}}}$ is a proper value in the simulation environment of this work while for other environments $\rho_k$ may need to be changed to realize a satisfied performance.

IV. LOW COMPLEXITY ROW-BY-ROW (RBR) ALGORITHM

In the previous section, we have proposed an RSD algorithm, where the rows of switch network matrix need to be optimized successively, which incurs high complexity. In this section, we propose a low-complexity RBR algorithm, by transforming the design problem into multiple parallel sub-problems and optimizes each row of $S$ in parallel, which effectively reduces the computational complexity.

As analyzed in Sec. III-B, the original design problem can be transformed into problem (12). By treating $SFD[k]$ as a whole beamforming matrix and considering the equal-power allocation, the solution of the optimal unconstrained beamforming matrix to maximize (12) is $P[k] = \hat{V}_{N_*} [k]$, where $\hat{V}_{N_*} [k]$ is the first $N_* \times N_*$ columns of $\hat{V} [k]$, and $\hat{V} [k]$ is derived from the SVD of $\Lambda[k] A_i[k]^H$ such that $\Lambda[k] A_i[k]^H = \hat{U} [k] \hat{\Sigma} [k] \hat{V} [k]^H$. To reduce the computational complexity, rather than directly solving $S$ and $D[k]$ to maximize (12), we aim to design $S$ and $D[k]$ to make $SFD[k]$ close to $P[k]$, as

$$\min_{S,D[k]} \sum_{k=1}^{K} \|P[k] - SFD[k]\|_F^2$$

s.t. $S_{i_*,i} \in \{0,1\}, \|S_{i_*}\|_0 = 1, \forall i, l$  (20a)

$$\|SFD[k]\|_F^2 = \rho_k.$$  (20b)

It is still uneasy to solve the problem (20) due to the non-convex binary constraint and the coupling of $S$ and $D[k]$. Hence, we propose the RBR algorithm to alternatively design $D[k]$ and $S$, i.e., alternatively fix one to optimize another one as follows.

A. Design of Digital Beamforming Matrix $D[k]$

To begin with, we design $D[k]$ when $S$ is fixed. A semi-unitary digital beamforming matrix can mitigate the interference among the data streams and enhance the spectral efficiency [11], [40]. Inspired by this, we enforce a semi-unitary constraint to the digital beamforming matrix, given by $D[k]^H D[k] = I_{N_*}$. By fixing the switch network matrix $S$ and omitting the transmit power constraint temporarily, the problem (20) can be reformulated as

$$\min_{D[k]} \sum_{k=1}^{K} \|P[k] - SFD[k]\|_F^2$$

s.t. $D[k]^H D[k] = I_{N_*}.$  (21)

The solution to (21), which is the orthogonal procrustes problem, is given as [11] and [40]

$$D[k] = \hat{V}_{N_*} [k] \hat{U}[k]^H,$$  (22)

where $L_t \times L_t$ and $N_* \times N_*$-dimensional $\hat{V} [k]$ and $\hat{U}[k]$ are obtained from the SVD of $P[k]^H SFD$, yielding that $P[k]^H SFD = \hat{U}[k] \hat{\Sigma}[k] \hat{V}[k]^H$, and $\hat{V}_{N_*} [k]$ is the first $N_* \times N_*$ columns of $\hat{V} [k]$.

B. Design of Switch Network Matrix $S$

Then, we design $S$ to solve the problem (20), with fixed $D$. By omitting the transmit power constraint temporarily, solving $S$ to minimize (20a) is rearranged as

$$\min_S \sum_{k=1}^{K} \|P[k] - SFD[k]\|_F^2$$  (23a)

$$\frac{1}{K} \sum_{i=1}^{N_t} \sum_{k=1}^{K} \log_2 \left( \left| N_p + \rho_k T_i[k] \right|^{-1} \bar{\Lambda}[k] \right)^2 (B_i[k] A_i[k]^H + A_i[k] B_i[k]^H + B_i[k] B_i[k]^H)),$$  (17)

(*)
\[ s.t. S_{i,l} \in \{0, 1\}, \|S_i\|_0 = 1, \forall i, l, \tag{23b} \]

where (23) is an integer programming problem associated with a matrix variable, which is inefficient to solve. To make the problem more tractable, we rewrite the \( k^{th} \) term of (23a) as

\[
\begin{align*}
|P[k] - SFD[k]|_F^2 \\
= \text{Tr}((P[k] - SFD[k])(P[k] - SFD[k])^H) \\
= \text{Tr}(P[k]P[k]^H) + \text{Tr}(SFD[k]D[k]^HP^HS^H) \\
- 2\text{Tr}(\text{Re}(SFD[k]P[k]^H)) \\
= \text{Tr}(P[k]P[k]^H) + \text{Tr}(SFK[k] \begin{bmatrix} I_{N_t} & 0 \\ 0 & K[k] \end{bmatrix} H P^HS^H) \\
- 2\text{Tr}(\text{Re}(SFD[k]P[k]^H)) \\
\leq \text{Tr}(P[k]P[k]^H) + \text{Tr}(K[k]K[k]^H F^HS^H SF) \\
- 2\text{Tr}(\text{Re}(SFD[k]P[k]^H)) \\
= \text{Tr}(P[k]P[k]^H) + \text{Tr}(F^HS^H SF) - 2\text{Tr}(\text{Re}(SFD[k]P[k]^H)), \tag{24f}
\end{align*}
\]

where \( K[k] \begin{bmatrix} I_{N_t} & 0 \\ 0 & K[k] \end{bmatrix} \) in (24d) is the SVD of \( D[k]D[k]^H \)

since we have \( D[k]D[k]^H = I_{N_t} \). (24e) comes from the property of matrix trace. The inequality (24f) follows that the diagonal elements of the Hermitian matrix \( K[k] \begin{bmatrix} I_{N_t} & 0 \\ 0 & K[k] \end{bmatrix} \) are no smaller than zero and the equality holds when \( D[k] \) is a square matrix, i.e., \( L_t = N_s \). Therefore, the \( k^{th} \) term of (23a) can be relaxed as (24g), where \( \text{Tr}(P[k]P[k]^H) \) is known and fixed. According to the structure of \( F \) in (4) and the constraint \( \|S_i\|_0 = 1 \), regardless how we design \( S \), \( \text{Tr}(F^HS^H SF) = \|SF\|_F^2 \) is a constant \( N_t \). Hence, to minimize (24g) is equivalent to maximize \( \text{Tr}(\text{Re}(SFD[k]P[k]^H)) \). Note that (24g) is a relaxed form of the \( k^{th} \) term of (23a). Therefore, minimizing (23a) can be relaxed as maximizing the summation of \( \sum_{k=1}^{K} \text{Tr}(\text{Re}(SFD[k]P[k]^H)) \) about \( k \), given by \( \sum_{k=1}^{K} \text{Tr}(\text{Re}(SFD[k]P[k]^H)) \). According to the property of matrix trace, \( \sum_{k=1}^{K} \text{Tr}(\text{Re}(SFD[k]P[k]^H)) \) is equivalent to

\[
\sum_{i=1}^{N_t} \sum_{k=1}^{K} \text{Re}(S_iFD[k]P_i[k]^H), \tag{25}
\]

where \( S_i \) and \( P_i[k] \) are the \( i^{th} \) row of \( S \) and \( P[k] \), respectively. According to (25), we have decomposed each row of \( S \) as \( N_t \) uncorrelated parts. As a result, designing \( S \) to maximize (25) is equivalent to separately designing \( S_i \) to maximize \( \sum_{k=1}^{K} \text{Re}(S_iFD[k]P_i[k]^H) \), for \( i = 1, 2, ..., N_t \), which can be stated as

\[
\begin{align*}
\max_{S_i} S_i \sum_{k=1}^{K} \text{Re}(FD[k]P_i[k]^H) & \tag{26a} \\
\text{s.t. } S_{i,l} \in \{0, 1\}, \|S_i\|_0 = 1, \forall l. & \tag{26b}
\end{align*}
\]

Following the binary property of the row vector \( S_i \), i.e., \( S_{i,l} \in \{0, 1\}, \forall l \) and \( \|S_i\|_0 = 1 \), maximizing (26a) is equivalent to finding the position of the maximal element of the column vector \( \sum_{k=1}^{K} \text{Re}(FD[k]P_i[k]^H) \), which is a simple ranking problem and can be solved efficiently based on the sorting algorithm in solvers. Consequently, by denoting \( p_{\text{max}} \) as the position of the maximal element, the optimal solution of \( S_i \) to the problem (26) is

\[
S_i = \begin{bmatrix} 0, \ldots, 0 \end{bmatrix} \begin{bmatrix} 1 \end{bmatrix} \begin{bmatrix} 0, \ldots, 0 \end{bmatrix} \quad \text{for} \quad P_{\text{max}} \leq L_t Q - p_{\text{max}}. \tag{27}
\]

Then, by solving problem (26) with \( i = 1, ..., N_t \) in parallel, the solution of \( S \) to problem (23) is obtained. Based on the aforementioned procedures, in the proposed RBR algorithm, we can alternatively solve \( D[k] \) and \( S \) via (22) and (27) until convergence. After that, we normalize \( D[k] \) to satisfy the transmit power constraint. The pseudocodes of the RBR algorithm are described in Algorithm 2.

**Algorithm 2 RBR Algorithm**

**Input:** \( \tilde{A}[k], A_s[k], \) and \( F, k = 1, 2, ..., K \)

**01:** Calculate each \( P[k] = \tilde{V}_N[k] \) and initialize \( S \) randomly

**02:** Repeat

**03:** Solve \( D[k] \) via (22), for \( k = 1, 2, ..., K \)

**04:** for \( i = 1 : N_t \)

**05:** Solve \( S_i \) via (27)

**06:** end for

**07:** Until convergence

**08:** Normalize each \( D[k] \) as \( D[k] \leftarrow \frac{\sqrt{P_{\text{max}}}}{\|SFD[k]\|_F} D[k] \)

**Output:** \( S \) and \( D[k], k = 1, 2, ..., K \)

---

V. SIMULATION RESULTS AND ANALYSIS

In this section, we evaluate the performance of the proposed DS-FPS architecture as well as the RSD and RBR algorithms. The simulation setup is given in Sec. V-A. We first evaluate the spectral efficiency and energy efficiency of the DS-FPS architecture with the RSD and RBR algorithms in Sec. V-B. Then, we analyze the impact of CSI on the spectral efficiency for the DS-FPS architecture in Sec. V-C. Furthermore, we analyze the computational complexity and convergence of the RSD and RBR algorithms in Sec. V-D.

---

**A. Simulation Setup**

The operating frequency is 0.3 THz, with 5 GHz bandwidth. The number of subcarriers is \( K = 5 \) and the noise power is −84 dBm for each subcarrier. The number of antennas at transmitter and receiver is equal, i.e., \( N_t = N_r \), which is set as 128, 256, 512, and 1024, respectively. The number of RF chains at transmitter and receiver is \( L_t = 4 \). The number of data streams is \( N_s = 4 \). The communication distance is set as 40m. The number of multipath is less than 5. The multipath components as well as the path gains and directions are generated via the multi-ray model [30]. We adopt the DS-FPS architecture at both transmitter and receiver with the same system parameters. For fair comparison, the counterpart architectures are also used at two sides.
1) Competitors of Proposed DS-FPS Architecture: We compare our scheme, i.e., the proposed DS-FPS architecture with RSD and RBR algorithms, with the following schemes: i) the FC architecture with the hybrid beamforming algorithm in [12], ii) the DHB architecture with the modified algorithm in [22], iii) the FPS-group-connected (FPS-GC) architecture with the FPS-AltMin algorithm in [24], iv) the AoSA architecture with SIC algorithm in [13], v) the subconnected with phase shifter selection (SPSS) architecture with the algorithm in [41], and vi) the subconnected with reduced number of phase shifters and phase shifter selection (SRPS) architecture with the algorithm in [41] and [42]. For the FPS-GC architecture, the number of groups is set as 2. The SPSS is based on the AoSA architecture, while a dedicated switch is used between each antenna and phase shifter. \( N_t (1 - \frac{1}{\beta}) \) antennas and phase shifters are turned off through controlling the switches. The number of remaining active antennas, phase shifters, and switches is \( N_t / \beta \). In SRPS architecture, the number of phase shifters is \( \frac{N_t}{\beta} \). Each phase shifter connects to \( \beta \) adjacent antennas through a 1-to-\( \beta \) switch and then selects one antenna with largest contribution to the spectral efficiency. In this work, we set \( \beta = 2 \) for both SPSS and SRPS architectures. The SPSS and SRPS use adjustable phase shifters while the proposed DS-FPS uses low-cost FPSs. Moreover, the switches in the DS-FPS are used to allow each antenna selecting one proper FPS such that all antennas are active, while in the SPSS and SRPS architectures partial antennas are turned off through switches. For all the above algorithms which involve iterations, we set the number of iterations as 10. For the algorithms which need the transmit power as one input, we set \( \rho = 20 \) dBm as the input transmit power.

2) Energy Efficiency Model: The energy efficiency \( EE \) is defined as the ratio between the spectral efficiency and the power consumption at transmitter \( P_{TX} \) and receiver \( P_{RX} \), i.e.,
\[
EE = \frac{SE}{P_{TX} + P_{RX}}
\]
We adopt the power consumption model of the hybrid beamforming studies [20], [22]. The power consumption of the DS-FPS architecture at transmitter can be expressed as
\[
P_{TX} = P_{\text{common}} + P_{\text{SW}}N_t + P_{\text{FPS}}N_{\text{FPS}} + P_{\text{DAC}} + P_{\text{RF}}N_t + P_{\text{PA}}N_t + \rho. \quad (28)
\]
where \( P_{\text{common}} = P_{BB} + P_{\text{DAC}}L_t + P_{\text{RF}}L_t + P_{\text{PA}}N_t + \rho \cdot \), \( P_{BB}, P_{\text{DAC}}, P_{\text{RF}}, P_{\text{PA}}, P_{\text{SW}}, P_{\text{FPS}} \) denote the power consumption of the baseband, DAC, RF chain, power amplifier, switch, and FPS, respectively. The corresponding multipliers denote the quantity of these devices used in the architecture. \( \rho \) is the transmit power at transmitter. \( P_{\text{common}} \) is usually the same for different hybrid beamforming architectures. \( P_{\text{analag}} \) is the analog beamforming part of power consumption, which is different for various hybrid beamforming architectures according to the used hardware components. For the DS-FPS architecture, \( P_{\text{analag}} \) is composed by the power consumed by switches and FPSs as expressed in (28).

In the DS-FPS architecture, the total number of FPSs is \( L_tQ \). Each antenna connects to all FPSs through \( L_tQ \) switches. Since each antenna only selects one FPS from all FPSs to connect with. Only one of the \( L_tQ \) switches is closed and the others are disconnected which do not consume power. Hence, the number of switches which consume power equals the number of antennas \( N_t \). In this work, \( N_t, L_t \), and \( \rho \) are not design variables such that \( P_{\text{common}} + P_{\text{SW}}N_t \) can be referred to as the static terms which are not related to the hybrid beamforming design. We denote the FPSs which are selected by at least one switch as the active FPSs and the others as the non-active FPSs which do not consume power. The number of active FPSs \( N_{\text{FPS}} \) equals the number of non-zero columns of \( S \). Hence, \( P_{\text{FPS}}N_{\text{FPS}} \) is related to the design variable \( S \) and is the dynamic term.

For FC, AoSA, DHB, and FPS-GC architectures, the common part is the same with (28), while the analog beamforming part needs to be changed by accounting their own devices. In FPS-GC architecture, \( N_{\text{SW}} \) and \( N_{\text{FPS}} \) should be determined by the FPS-AltMin algorithm [24]. For SPSS and SRPS architectures, only \( N_t/\beta \) antennas are active such that \( P_{\text{common}} = P_{BB} + P_{\text{DAC}}L_t + P_{\text{RF}}L_t + P_{\text{PA}}N_t/\beta + \rho \). For the analog beamforming part, the number of phase shifters and switches is \( N_t/\beta \). Hence, the consumed power of analog beamforming part of both SPSS and SRPS architectures is \( P_{\text{PS}}N_t/\beta + P_{\text{SW}}N_t/\beta \). We summarize the power consumption at transmitter, i.e., \( P_{TX} \), of different architectures in TABLE I. For each architecture, the expression of \( P_{RX} \) is similar with \( P_{TX} \), by substituting the DAC and power amplifier with ADC and low noise power and then deleting \( \rho \).

We use the typical power consumption reported by the existing THz studies, mainly around 0.3 THz, in the unit of mW as follows. The power consumption of DAC, ADC, power amplifier, low noise amplifier, switch, RF chain, and baseband is \( P_{\text{DAC}} = 110 \), \( P_{\text{ADC}} = 158.6 \), \( P_{\text{PA}} = 49 \), \( P_{\text{LNA}} = 53 \), \( P_{\text{SW}} = 9 \), \( P_{\text{RF}} = 43 \), and \( P_{BB} = 200 \), respectively [8], [43]. [44], [45], [46], [47], [48]. It has been reported in [49] that \( P_{BB} = 52, 39, \) and 26 for 3, 2, and 1-bit phase shifters. The FC, AoSA, SPSS and SRPS architectures use the infinite-resolution phase shifter, while the DHB architecture [22] uses the 2-bit or 1-bit resolution phase shifters. Since the infinite-resolution phase shifter is ideal, we use the power consumption of 3-bit phase shifter to represent its power consumption. Hence, for FC, AoSA, SPSS, and SRPS architectures, \( P_{PS} = 52 \), while for DHB architecture, \( P_{PS} = 39 \) for 2-bit phase shifter and \( P_{PS} = 26 \) for 1-bit phase shifter. The FPS is usually realized by a passive delay element which provides a fixed phase adjustment with no power consumption and an amplifier with small gain to keep the output amplitude and power of the FPSs with different phases at the same level. Since the passive delay element with different phases in [50] has a small power difference,
we consider an amplifier with about 4.8 dB gain which consumes 16.8 mW [51] to keep the equal output amplitude and power of FPSs. Hence, the power consumption of FPS is set as $P_{\text{FPS}} = 16.8$.

In this work, we consider the use of full-bit DACs/ADCs. It has been studied in [14] that we can use the power consumption of DAC/ADC (>8 bit) to represent the power consumption of full-bit DAC/ADC, for which we use the power consumption of 9-bit DAC in [43], i.e., $P_{\text{DAC}} = 110$ mW, and 12-bit ADC in [44], i.e., $P_{\text{ADC}} = 158.6$ mW. The use of low-bit DACs/ADCs may be a potential research direction.

The joint optimization of the hybrid beamforming matrices and the bit of the DACs/ADCs has been analyzed in [14] and a channel estimation algorithm for hybrid beamforming with low-bit ADCs has been proposed in [52].

B. Spectral Efficiency and Energy Efficiency of the Proposed DS-FPS Architecture

In the simulations, the DS-FPS architecture with the RSD and RBR algorithms only knows partial CSI. For the other architectures and algorithms, full CSI is assumed to be known. In FC, AoSA, SPSS, and SRPS architectures, the infinite-resolution phase shifters are used. In DHB architecture, 2-bit resolution phase shifters are used, unless stated otherwise. In DS-FPS and FPS-GC architectures, FPSs are used.

As shown in Fig. 3, for the DS-FPS architecture, the RSD algorithm yields higher spectral efficiency than the low-complexity RBR algorithm. Furthermore, both the spectral efficiencies of the DS-FPS architecture with either RSD or RBR algorithm are higher than the DHB, the FPS-GC, and the AoSA architectures. The SPSS and SRPS architectures aim to reduce the power consumption by deactivating partial antennas, which decreases the spectral efficiency at the same time. As a result, the spectral efficiency of the SPSS and SRPS architectures is much lower than the DS-FPS architecture. The SPSS has higher spectral efficiency than the SRPS due to the more flexible switch connections. Furthermore, the spectral efficiencies of the DS-FPS architecture with the RSD and RBR algorithms are lower than the FC architecture. Therefore, compared to the existing architectures, the proposed DS-FPS architecture with the RSD and RBR algorithms can achieve a satisfied spectral efficiency.

Fig. 4 evaluates the energy efficiency as well as the spectral efficiency concurrently. The energy efficiencies of the DS-FPS architecture with the RSD and RBR algorithms are significantly higher than the other architectures, due to the low power consumption of the FPSs and switches. The power consumed by switches is only a small part of the overall power consumption of DS-FPS architecture, as can be observed in TABLE I. Although the spectral efficiency of the DS-FPS architecture is lower than the FC architecture, the energy efficiency of the DS-FPS architecture is remarkably higher. The DHB architecture aims to use 2-bit and 1-bit phase shifter to reduce the power consumption and achieve good trade-off between energy efficiency and spectral efficiency. However, the power consumption of 2-bit and 1-bit phase shifter is still higher than the FPS. Moreover, the quantity of phase shifters in DHB architecture equals $N_t$, i.e., 1024, which is much larger than the number of FPSs in DS-FPS architecture, i.e., 32. Hence, the power consumption of the DS-FPS architecture is much lower than the DHB architecture. Compared to the DHB architecture with 1-bit or 2-bit phase shifters, the energy efficiency enhancement of the proposed DS-FPS architecture exceeds 0.05 bits/s/Hz/W. The spectral efficiency and energy efficiency of the proposed DS-FPS architecture are noticeably higher than the FPS-GC and AoSA architectures. With $\beta = 2$, the SPSS and SRPS architectures deactivate half antennas,
and SRPS architectures and lower than the FC architecture, as shown in Fig. 6. On the other hand, with more antennas, the power consumption is higher such that the energy efficiencies of all architectures decrease, as shown in Fig. 7. Particularly, for various numbers of antennas, the superiority on energy efficiency of the DS-FPS architecture stands out clearly.

In Fig. 8, we evaluate the impact of the number of FPSs on the spectral efficiency and energy efficiency of the proposed DS-FPS architecture. The number of FPSs from left to right is 12, 16, 24, 32, 40, 56, 76, 96, and 128, respectively. The spectral efficiencies of the DS-FPS architecture with RSD and RBR algorithms increase with the number of FPSs, since the number of provided phases of the FPS network increases. However, with more FPSs, the power consumption also increases. As a result, the energy efficiency first increases with the number of FPSs and then decreases with the number of FPSs. Consequently, to achieve high energy efficiency, the number of FPSs should not be too large and a proper value needs to be selected.

C. Impact of Partial and Inaccurate CSI

We evaluate the impact of partial and inaccurate CSI on the spectral efficiency of the DS-FPS architecture. The proposed RSD and RBR algorithms are designed for the case of partial CSI. While with the following adjustment, the RSD and RBR algorithms can also work for the case of full CSI. 

In Fig. 9, we evaluate the spectral efficiency of the RSD and RBR algorithms under inaccurate partial CSI at TX and RX, i.e., the known $\bar{A}[k]$, $A_t[k]$, and $A_r[k]$ are inaccurate. We use $\bar{A}[k]$ to represent the level of accuracy. The inaccurate $\bar{A}[k]$ and inaccurate $A_t[k]$ are represented as

$$\bar{A}[k] = \xi A[k] + \sqrt{1 - \xi^2}(E_1 \odot A[k])$$

$$A_t[k] = e^{j(\xi \angle(A_t[k]) + \sqrt{1 - \xi^2}(E_2 \odot \angle(A_t[k])))}$$

where $E_1$ and $E_2$ denote the error matrix with elements following the distribution of i.i.d $\mathcal{N}(0, 1)$. The inaccurate $A_t[k]$ has the similar expression with (30). Fig. 9 shows the spectral efficiency of the RSD and RBR algorithms by varying
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TABLE II

COMPUTATIONAL COMPLEXITY ANALYSIS OF THE PROPOSED RSD AND RBR ALGORITHMS

| Operation                                                                 | RSD algorithm | Complexity       |
|---------------------------------------------------------------------------|---------------|-----------------|
| Design $S_i$ via maximizing (19) for $i = 1, 2, ..., N_t$                   |               | $O(KN_tN_r^3L_tQ)$ |
| Calculate $D[k]$ through (13), for $k = 1, 2, ..., K$                     |               | $O(KN_tL_t^2Q)$  |
| Normalize each $D[k]$ as $D[k] = \frac{\sqrt{\xi}}{|\text{SFD}[k]|} D[k]$ |               | $O(KN_tL_tN_s)$  |
| Overall                                                                   |               | $O(KN_tN_r^2L_tQ)$ |

| Operation                                                                 | RBR algorithm | Complexity       |
|---------------------------------------------------------------------------|---------------|-----------------|
| Calculate $\bar{A}[k]A_r[k]^H$, for $k = 1, 2, ..., K$                    |               | $O(KN_tN_r^2)$  |
| Calculate each $P[k]$ through $N_s$-truncated SVD of $\bar{A}[k]A_r[k]^H$, for $k = 1, 2, ..., K$ |               | $O(K(N_t + N_p)N_s^2)$ |
| Design $S_i$ via (27) for $i = 1, 2, ..., N_t$                            |               | $O(KN_tL_tN_s)$  |
| Normalize each $D[k]$ as $D[k] = \frac{\sqrt{\xi}}{|\text{SFD}[k]|} D[k]$ |               | $O(KN_tL_tN_s)$  |
| Overall (M iterations)                                                    |               | $O(KM N_tL_tQ)$  |

the value of $\xi$. The dash lines represent the case of full CSI which is accurate and not related to $\xi$. Compared to the case of full CSI, the spectral efficiency loss of the RSD and RBR algorithms under inaccurate partial CSI grows with the smaller $\xi$. When $\xi = 0.7$, the spectral efficiency of the RSD and RBR algorithms is about 70% of the case of full CSI, which reveals that the proposed RSD and RBR algorithms are robust to the CSI error.

D. Convergence and Computational Complexity Analysis

Fig. 10 shows the convergence performance of the RBR algorithm with partial CSI. We evaluate the average objective function of the RBR algorithm over all subcarriers, i.e., $\frac{1}{K} \sum_{k=1}^{K} \| P[k] - \text{SFD}[k] \|^2_F$, versus the number of iterations. The simulation results show that the RBR algorithm converges fast with about 10 iterations.

As listed in Table II, we analyze the computational complexities of the RSD and RBR algorithms. In THz UMMIMO systems, $N_t$ is usually very large, e.g., 1024, $L_t$, $N_s$, $N_p$, and $Q$ are usually small, e.g., $L_t = N_s = 4$, $N_p < 5$, and $Q = 8$ in our simulations. Without loss of generality, we have $O(L_t) \approx O(N_s) \approx O(N_p) \approx O(Q) \ll O(N_t)$. The RSD algorithm does not involve iterations, whose overall computational complexity is $O(KN_tN_r^3L_tQ)$. The RBR algorithm includes iterations and we denote $M$ as the number of iterations. The overall computational complexity of RBR algorithm is $O(KM N_tL_tQ)$. Specifically, for the calculation of $P[k]$, RBR algorithm needs to calculate the $N_s$-truncated SVD of $\bar{A}[k]A_r[k]^H$ to obtain the first $N_s$ columns of the right singular matrix, whose complexity is $O((N_t + N_p)N_s^2)$ [54]. Through the simulation in Fig. 10, we observe that, $M \approx 10$ and is usually smaller than $N_p^3$. Consequently, the computational complexity of RBR algorithm is lower than the RSD algorithm.

The computational complexities of the algorithm in [12] for FC, the algorithm in [22] for DHB, the SIC algorithm [13] for AoSA, and the FPS-AltMin algorithm [24] for FPS-GC are $O(KN_t^3)$, $O(KN_t^3L_t)$, $O(N_t^2L_t)$, and $O(N_tL_tQ \cdot \log_2(N_tL_tQ))$, respectively. The key idea of the algorithms for SPSS and SRPS architectures [41], [42] is utilizing the first $L_t$ columns of the right singular matrix of channel to design the analog beamforming matrix and then use the effective channel to design the digital beamforming matrix. Although with different strategy about the turning off of the antennas, the computational complexity of both these two algorithms is proportional to $O(N_t^2)$. Consequently, the computational complexity of the proposed RBR and RSD algorithms is linearly related with the number of antennas $N_t$, which is lower than the counterpart algorithms.

VI. CONCLUSION

In this paper, we have proposed an energy-efficient DS-FPS architecture for THz hybrid beamforming systems,
using low-cost FPSs. A switch network is used to enable the dynamic connections between the FPSs and antennas. Moreover, to account for practical partial CSI, we have proposed an RSD algorithm to design the hybrid beamforming matrices for the DS-FPS architecture. To further reduce the computational complexity of the RSD algorithm, an RBR algorithm has been developed. Extensive simulation results have shown that the DS-FPS architecture achieves remarkably higher energy efficiency than the existing architectures. Moreover, by using the RSD and RBR algorithms, the spectral efficiency of the RSD and RBR algorithms is linearly related with the number of antennas, which is lower than the existing algorithms.

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