Coverage Increase at THz Frequencies: A Cooperative Rate-Splitting Approach

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Abstract—Numerous studies claim that terahertz (THz) communication will be an essential piece of sixth-generation wireless communication systems. Its promising potential also comes with major challenges, in particular the reduced coverage due to harsh propagation loss, hardware constraints, and blockage vulnerability. To increase the coverage of THz communication, we revisit cooperative communication. We propose a new type of cooperative rate-splitting (CRS) called extraction-based CRS (eCRS). Furthermore, we explore two extreme cases of eCRS, namely, identical eCRS and distinct eCRS. To enable the proposed eCRS framework, we design a novel THz cooperative channel model by considering unique characteristics of THz communication. Through mathematical derivations and convex optimization techniques considering the THz cooperative channel model, we derive local optimal solutions for the two cases of eCRS and a global optimal closed form solution for a specific scenario. Finally, we propose a novel channel estimation technique that not only specifies the channel value, but also the time delay of the channel from each cooperating user equipment to fully utilize the THz cooperative channel. In simulation results, we verify the validity of the two cases of our proposed framework and channel estimation technique.

Index Terms—THz communication, cooperative communication, rate-splitting multiple access, coverage increase.

I. INTRODUCTION

Along with the constant evolution of modern technology, wireless communication has steadily developed to satisfy increasing demands of high data traffic in fifth-generation wireless communication systems. To satisfy the demands, wireless communication systems have implemented novel technologies such as massive multiple-input multiple-output (MIMO) systems and millimeter-wave technology [1], [2], [3].

While these technologies are sufficient for current requirements, innovative approaches are in need to meet the requirements of sixth-generation (6G) systems [4].

One promising technology for 6G systems is terahertz (THz) communication, which exploits broad and vacant frequency bands ranging from 0.1 THz to 10 THz [5], [6], [7]. While a primitive THz system has already succeeded in 100 Gbps communication [8]. THz communication is still in its infancy due to limitations in both poor channel conditions and low signal power. In THz communication, the propagation loss degrades the signal power in orders of magnitude larger than the conventional frequency bands, and molecular absorption reduces the signal power even further, where the degradation is exponentially proportional with respect to distance [9], [10]. THz communication systems are also known to be line-of-sight (LoS) link dominant and vulnerable to blockage due to minimal scattering and decreased multipaths [11]. Furthermore, limited hardware specifications are additional factors to surmount. Although various approaches to generate the THz frequency signals were studied, the signal generators suffer from low transmit power [8], [12], [13]. The THz frequency signals were generated using photonic devices and photomixing in [8], but the transmit power was extremely low due to the limited gain of the uni-travelling-carrier photodiode. While the non-photonic device was used to generate signals of 300 GHz carrier frequency in [12], it was inadequate for the high power regime since the power conversion had linearity only in the low transmit power regime.

There have been several studies considering unique characteristics of THz communication. Considering that the far-field assumption does not apply to short-range wireless communication in THz frequencies, it was shown in [14] that multiplexing in the direct LoS environment is possible, where reconfigurable array architectures were proposed to exploit this effect. Other works considered relaying systems, where the studies focused on increasing the performance from the detrimental channels due to the molecular absorption [15], [16].

In this paper, we revisit the concept of cooperative communication to overcome the harsh channel conditions and low signal power devices in THz communication, where cooperative communication has already been considered as a candidate technology for THz systems [17]. Cooperative communication is an idea that gained attention in the early 2000s, and the basic concept is to cooperate among user equipments (UEs) to gain better performance for the overall system, e.g., the strong UE can additionally use its power to increase the coverage of the system by assisting weak
UEs [18], [19]. To be clear, we distinguish the cooperative communication system from the relay system. In the relay system, the relaying devices only transmit messages for other UEs, while in the cooperative communication system, the cooperating UEs not only transmit but also receive messages for themselves. While the concept of cooperative communication was promising, one weakness was the practicality, since no UE would want to use its resource to help other UEs so far. However, nowadays, people possess multiple devices, and the future is envisaged as an Internet-of-Things society, where we can strongly anticipate the increase of devices [20]. By using multiple devices spread in different places, possibly owned by the same user, cooperative communication can be successfully utilized. In addition, cooperative communication can also be achieved between multiple users. In distributed multiple antenna systems, such as military communication, a squad of devices located in different places can cooperate for the common purpose of communication [21]. By incentivizing cooperative communication, e.g., offering lower fees to cooperating users, users can also willingly cooperate to increase the performance of the overall system.

To effectively support cooperative communication, we need to adopt a proper multiple access technique, e.g., non-orthogonal multiple access (NOMA), spatial-division multiple access (SDMA), or rate-splitting multiple access (RSMA) [22], [23], [24]. In NOMA, a successive interference cancellation (SIC) technique is adopted to fully decode the interference, while in SDMA, the interference is fully considered as noise in general. RSMA is a superset of NOMA and SDMA and bridges those two schemes to further increase the performance. This is achieved by enabling RSMA to partially decode interference, thanks to the creation of a common stream and the presence of SIC, and partially treating the remaining interference as noise [25]. To that end, RSMA divides the messages of the UEs into private and common parts, where the private parts are separately encoded into private streams, and the common parts are jointly encoded into a common stream. From the UE standpoint, the common stream is decoded first while the private streams are considered as noise. Then, by removing the decoded common stream, the desired private stream is decoded with reduced interference.

It has been shown that RSMA includes both SDMA and NOMA as extreme cases and has superior performance in both perfect channel state information (CSI) and imperfect CSI cases [24], [25]. Due to its superior characteristics, RSMA gained attention in many studies and was implemented in various applications to increase performance [24]. For example, the concept of cooperative rate-splitting (CRS) was proposed in [26], [27], and [28], where the cooperating UEs relay the common message to increase the minimum rate of the UEs. The sum rate and energy efficiency was improved in a non-orthogonal unicast multicast (NOM) problem with RSMA [29].

In this paper, we consider a THz multiple-user (MU) multiple-input single-output (MISO) downlink system. To model the characteristics of THz communication systems, we consider the channel and hardware factors, e.g., LoS dominant links, blockage vulnerability, limited transmit power, and high sampling rates. In our system model, an access point (AP) is not able to communicate with a UE, denoted as a destination UE (dUE), due to a blocked channel. The other UEs, denoted as medium UEs (mUEs), will then cooperate by sharing their resources to achieve reliable communication for the dUE while the mUEs also decode their own messages as well. By introducing a two-phase cooperative communication system, the AP transmits messages to the mUEs in the first phase, and the mUEs transmit messages to the dUE in the second phase. The contributions of this paper are summarized as follows:

- For the two-phase cooperative communication, we propose a new type of CRS, named extraction-based CRS (eCRS). The proposed eCRS has two main differences compared to the conventional CRS. The first difference appears when the mUEs relay the messages to the dUE in the second phase. In the conventional CRS, the common message includes not only the message for the dUE, but also the messages for the mUEs is directly relayed to the dUE as in [26], [27], and [28]. This may cause an inefficient use of resources since the mUEs transmit messages unrequired for the dUE. In contrast, in eCRS, each mUE extracts the message for the dUE from the common message, and then relays only the message for the dUE in the second phase. Through this approach, the mUEs can use their power solely for the message of the dUE. The second difference appears in the two-phase transmission framework, which explains how the messages are split and combined. While the conventional CRS is only based on a single common message [26], [27], [28], we establish a transmission framework for eCRS by employing multiple common messages, similar to the general RSMA [25].

- We also investigate two extreme cases of eCRS, namely, identical eCRS (ieCRS) and distinct eCRS (DeCRS), where both cases only require one SIC layer. ieCRS, which is the same as NOM in the first phase, transmits an identical common message to the mUEs, and then the mUEs decode-and-forward the identical message to the dUE. In contrast, DeCRS transmits distinct common messages to each mUE, and then each mUE decode-and-forwards the distinct message to the dUE. Optimization problems are formulated for both ieCRS and DeCRS. Through mathematical derivations and convex optimization techniques, we show that ieCRS and DeCRS are both effective in different scenarios.

- We develop a THz cooperative channel model by considering unique characteristics of THz communication. In conventional cooperative communication, it is assumed that the transmit signals from the multiple mUEs arrive at the dUE in a single tap. However, for cooperative communication in THz communication systems, the signals from the mUEs arrive in different taps due to the large bandwidth and high sampling rate. By judiciously taking the multipaths through the mUEs into account, the THz cooperative channel can obtain additional gain, which has not been considered in existing literature.
We propose a novel channel estimation technique to fully utilize the developed THz cooperative channel model. While the channel gain may be obtainable with conventional channel estimation techniques for frequency selective channels, the THz cooperative channel model requires additional information of both the channel gain and time delay for each specific mUE. By utilizing the characteristics of sequences such as pseudo-noise or Zadoff-chu sequences, we derive an estimation technique to not only estimate the channel gains, but also identify the delays of all the mUEs. In result, we verify that the channel estimation technique works sufficiently well even in the low transmit power regime and also show the performances of the two cases of our proposed framework with imperfect CSI.

The rest of paper is organized as follows. Section II describes the transmission framework of eCRS and details of IeCRS and DeCRS. In Section III, we formulate and solve the minimum rate maximization problem for IeCRS through convex optimization techniques. We also derive an effective closed form solution with minimal complexity and optimal performance for IeCRS. In Section IV, we formulate and solve the optimization problem for DeCRS. Section V describes the specifics of the channel estimation technique adequate for the THz cooperative channel model. Section VI shows the simulation results of the channel estimation technique and cooperative communication techniques with perfect and imperfect CSI. Finally, Section VII concludes our paper.

Notation: Lower and upper boldface letters represent column vectors and matrices. $A^*$ and $A^H$ denote the conjugate, and conjugate transpose of the matrix $A$. $\mathbb{C}^{m \times n}$ and $\mathbb{R}^{m \times n}$ represent the set of all $m \times n$ complex and real matrices. $|\cdot|$ denotes the amplitude of the scalar, and $\| \cdot \|$ represents the $\ell_2$-norm of the vector. $\mathbb{Z}$ denotes the set of integers, $O$ denotes the Big-O notation. The Kronecker product is denoted by $\otimes$. $0_m$ is used for the $m \times 1$ all zero vector, and $I_m$ denotes the $m \times m$ identity matrix. $CN(m, \Sigma)$ denotes the circularly symmetric Gaussian distribution with mean $m$ and variance $\Sigma$.

## II. SYSTEM MODEL

In this paper, we consider the MU-MISO downlink system operating in THz frequency bands. Because of the reduced multipath effect in the THz frequencies, we consider the system in which only LoS links exist. In the system, the AP equipped with $N_t$ antennas serves $(K + 1)$ single antenna UEs. Among all UEs, the $K$ UEs designated as the mUEs have the LoS links from the AP, whereas the UE designated as the dUE experiences blockage of the LoS link from the AP. Since there is no other link to the dUE, it is impossible for the AP to serve every UE with the conventional non-cooperative multiple access techniques such as NOMA, SDMA, and RSMA.

To resolve this issue, we adopt CRS, which enables the AP to communicate with every UE. As illustrated in Fig. 1, CRS operates in two phases. The AP transmits messages to the $K$ mUEs in the first phase, and then the $K$ mUEs transmit messages to the dUE in the second phase. Through this two-phase system, the AP can successfully communicate with the dUE even when the AP-dUE link is blocked. We focus on the single dUE scenario in this paper, while the generalization to the multiple dUE scenario will be studied as future works.

To avoid inefficient use of resource while transmitting the message for the dUE, we propose eCRS that extracts the message for the dUE from the common message. Furthermore, we investigate two extreme cases of eCRS, namely, IeCRS and DeCRS. The details of the framework are explained in the following subsections. Throughout the paper, the $K$ mUEs are indexed by a set $\mathcal{K} = \{1, \ldots, K\}$, and the messages for the $k$-th mUE and the dUE are denoted as $W_k$ and $W_d$ respectively.

### A. Transmission Framework of eCRS

In this subsection, we describe the transmission framework of eCRS. In the first phase, the AP splits the message $W_d$ with respect to the subsets of the set $\mathcal{K}$, where each subset $\mathcal{S}$ corresponds to the group of mUEs that decodes the message. The resulting messages are denoted as $\{W_d^\mathcal{S}\}_{\mathcal{S} \subset \mathcal{K}}$, where $\mathcal{S} \neq \emptyset$. Similarly, the AP splits the message $W_k$ into private part $W_{p,k}$ and multiple common parts $\{W_{c,k}^\mathcal{S}\}_{\mathcal{S} \subset \mathcal{K}}$, where every $\mathcal{S}_i$ includes the element $k$. Next, the AP jointly combines the messages with the same subset $\mathcal{S}$ to generate the common message $W_{c,\mathcal{S}}$, e.g., for $\mathcal{S} = \{1, 2\}$, the messages $W_{c,1}^{\{1,2\}}, W_{c,2}^{\{1,2\}}$, and $W_d^{\{1,2\}}$ are combined into $W_{c}^{\{1,2\}}$. Then, the private part $W_{p,k}$ is encoded into the private stream $s_k$, and the common message $W_c^{\mathcal{S}}$ is encoded into the common stream $s_{c,\mathcal{S}}$. Since every mUE in the subset $\mathcal{S}$ decodes the common stream $s_{c,\mathcal{S}}$, every mUE must decode multiple common streams. While this is possible with multiple SIC layers, our scenario of interest considers the practical case of one SIC layer. In result, we select the subsets so that each mUE only decodes a single common stream. The selection of subsets for eCRS with one
SIC layer should satisfy the conditions given as
\[ \bigcap_{i=1} \{ S_i \} = \emptyset, \quad \bigcup_{i=1} \{ S_i \} = \mathcal{K}, \quad S_i \subset \mathcal{K}, \]
where \( S_i \) is the subset, which contains the group of mUEs that decodes the common stream \( s_c \).

While there can be various approaches to select the best subsets for eCRS with one SIC layer, we focus on two extreme cases, which are IeCRS and DeCRS. The subsets of IeCRS and DeCRS are denoted as \( S_1 = \mathcal{K} \) and \( \{ S_i \}_{i=1}^K = \{ \{1\}, \ldots, \{K\} \} \), respectively. Motivated by the 1-layer RSMA in [30] and [31], IeCRS explores the case where the common message is for all the mUEs. In contrast to IeCRS, DeCRS is the other extreme case with \( K \) common messages, where every mUE has its own common message.

**B. IeCRS**

In IeCRS, the AP selects a single subset as \( S_1 = \mathcal{K} \), where the corresponding common message is intended for all mUEs. Hence, the message \( W_{d,k}^C \) is equal to the message \( W_d \), and the message \( W_k \) is split into the private part \( W_{p,k} \) and common part \( W_{c,k} \). For simplicity, we neglect the index \( K \) in IeCRS. The overall message transmission and reception of IeCRS in the first phase is described in Fig. 2 (a). With the common stream \( s_c \) and private stream \( s_k \), the AP linearly precodes the \((K+1)\) streams and transmits the signal given as
\[ x = f_c s_c + \sum_{k=1}^{K} f_k s_k, \]
where \( f_c \in \mathbb{C}^{N_t \times 1} \) and \( f_k \in \mathbb{C}^{N_t \times 1} \) are the linear precoders for the common stream \( s_c \) and private stream \( s_k \), respectively.

The received signal at the \( k \)-th mUE during the coherence time block is given as
\[ y_k = h_k^H F s + z_k, \]
where \( h_k \in \mathbb{C}^{N_t \times 1} \) is the downlink channel for the AP-(\( k \)-th mUE) link, \( F \in \mathbb{C}^{N_t \times (K+1)} \) is the precoder matrix defined as \( F = [f_1, f_1, \ldots, f_K] \) with the AP power constraint as \( \text{tr}(F F^H) \leq P_{AP} \), \( s \) is the stream vector defined as \( s = [s_c, s_1, \ldots, s_K]^T \) satisfying \( \text{E} \{ s s^H \} = I_{(K+1)} \), and \( z_k \sim \mathcal{CN}(0, N_0) \) is the additive white Gaussian noise (AWGN). Without loss of generality, we set the noise variance \( N_0 \) equal to 1 throughout the paper.

After receiving the signal, following the same procedure with existing RSMA techniques [26], [27], [29], the \( k \)-th mUE first decodes the common stream \( s_c \), which contains the common message \( W_c \). The achievable rate for the common message \( W_c \) at the \( k \)-th mUE is given as
\[ R_{c,k} = \log_2 \left( 1 + \frac{|h_k^H f_c|^2}{|h_k^H f_c|^2 + I_k + 1} \right), \]
where \( I_k = \sum_{i \neq k} |h_k^H f_i|^2 \) is the interference from other private streams. Since every mUE needs to decode the common message \( W_c \), the corresponding rate of \( W_c \) should not exceed the achievable rate \( R_{c,k} \) respect to every \( k \in \mathcal{K} \). The inequality that guarantees the mUEs to successfully decode the common message \( W_c \) is given as
\[ C_d + \sum_{i=1}^{K} C_i \leq R_{c,k}, \quad \forall k \in \mathcal{K}, \]
where \( C_k \) is the rate for the common part \( W_{c,k} \), and \( C_d \) is the rate for the message \( W_d \) in the first phase. Next, the common stream \( s_c \) is cancelled out from the received signal \( y_k \) using the SIC technique. Then, the \( k \)-th mUE decodes the private stream \( s_k \), which contains the private part \( W_{p,k} \). The achievable rate for the private part \( W_{p,k} \) is given as
\[ R_{p,k} = \log_2 \left( 1 + \frac{|h_k^H f_c|^2}{I_k + 1} \right). \]
Finally, the \( k \)-th mUE extracts the common part \( W_{c,k} \) and message \( W_d \) from the common message \( W_c \) and then recombines the private part \( W_{p,k} \) and common part \( W_{c,k} \) into the message \( W_k \). The achievable rate for the message \( W_k \) is given as \( R_k = R_{p,k} + C_k \).

In the second phase of IeCRS, every mUE transmits the message \( W_k \) to the dUE. The overall process of the message transmission and reception in the second phase is described in Fig. 2 (b). The \( k \)-th mUE encodes the message \( W_k \) into the stream \( s_d \) and transmits the precoded signal given as
\[ x_{d,k} = f_{d,k} s_d, \]
where \( f_{d,k} \in \mathbb{C} \) is a linear precoder at the \( k \)-th mUE. The received signal at the dUE in the \( m \)-th time slot is given as
\[ y_d[m] = \sum_{k=1}^{K} g_k x_{d,k}[m - \tau_k] + z_d[m], \]
\[ = \sum_{k=1}^{K} g k s_d[m - \tau_k] + z_d[m], \]
where \( g_k \in \mathbb{C} \) and \( \tau_k \in \mathbb{Z} \) are the channel gain and time delay for the \((k\text{-th mUE})\)-dUE link, respectively. The term \( z_{ik} \) is the AWGN, and \( g_{ik} \) is the effective channel gain of the \((k\text{-th mUE})\)-dUE link defined as \( g_{ik} = g_k f_{d,ik} \).

Due to the large bandwidth and high sampling rate of THz communication, the signals from multiple mUEs arrive in different taps, making the THz cooperative channel in (9) have the form of a frequency selective channel or single frequency network (SFN) system. However, different from the frequency selective channel and conventional SFN [33], the effective channel gain \( g_{ik} \) can be controlled independently at the \(k\text{-th mUE}\) with the precoder \( f_{d,ik} \) to obtain a higher achievable rate. We employ an orthogonal-frequency-division-multiplexing (OFDM) approach to make the channel into \( N_c \) parallel flat-fading channels. After OFDM processing, the received signal at the \(n\text{-th sub-carrier}\) is given as

\[
\tilde{y}_n = g_n \tilde{s}_n + z_n, \tag{10}
\]

where \( \tilde{s}_n \) is the frequency domain stream that satisfies \( \mathbb{E} \{|\tilde{s}_n|^2\} = 1 \), and \( z_n \) is the AWGN. The frequency domain channel \( \tilde{g}_n \) is given as

\[
\tilde{g}_n = \sum_{k=1}^{K} g_k \exp \left( -j2\pi \tau_k n \frac{n}{N_c} \right). \tag{11}
\]

We observe that the \( N_c \) sub-carrier channels depend on the variable \( g_k \). By controlling \( g_k \), i.e., \( f_{d,ik} \), the quality of the sub-carrier channels can improve. By decoding and recombinig the messages from \( N_c \) received signals, the achievable rate for the message \( W_d \) in the second phase is given as

\[
R^{(2)}_d = \frac{1}{N_c + L} \sum_{n=1}^{N_c} \log_2 \left( 1 + |\tilde{g}_n|^2 \right) \approx \frac{1}{N_c} \sum_{n=1}^{N_c} \log_2 \left( 1 + |\bar{g}_n|^2 \right), \tag{12}
\]

where \( L \) is the length of the cyclic prefix. The approximation in (12) is valid since we can set the number of sub-carriers \( N_c \) much larger than \( L \). Finally, the achievable rate for the message \( W_d \) at the dUE is given as

\[
R_d = \min \left\{ C_d, R^{(2)}_d \right\}, \tag{13}
\]

which guarantees the message \( W_d \) to be successfully decoded in both two phases. We design the precoders and message split for IeCRS in Section III.

C. DeCRS

For DeCRS, the AP selects the subsets as \( \{S_i\} = \{1, \cdots, K\} \). The overall message transmission and reception of DeCRS in the first phase is described in Fig. 3 (a). With

\[3\] Since OFDM may suffer from the high peak-to-average power ratio (PAPR) issue, it would be an interesting future work to investigate whether other waveforms, e.g., single-carrier frequency division multiple access (SC-FDMA) [34], with low PAPR can also be adopted in our problem of interest.
private stream $s_k$, which contains the private part $W_{p,k}$. The achievable rate for the private part $W_{p,k}$ is given as

$$R_{p,k} = \log_2 \left( 1 + \frac{|h_k^T e_k|^2}{E_k + I_{c,k} + 1} \right).$$

Finally, the $k$-th mUE extracts the common part $W_{c,k}$ and message $W_d^k$ from the common message $W_c^k$ and then recombines the private part $W_{p,k}$ and common part $W_{c,k}$ into the message $W_k^k$. The achievable rate for the message $W_k^k$ is given as $R_k = R_{p,k} + C_k$ as in IeCRS.

In the second phase of DeCRS, the $k$-th mUE transmits the distinct message $W_d^k$ to the dUE. The overall process of the message transmission and reception in the second phase is described in Fig. 3 (b). The $k$-th mUE encodes the message $W_d^k$ into the stream $s_{d,k}$ and transmits the precoded stream given as

$$x_{d,k} = f_{d,k} s_{d,k},$$

where $f_{d,k} \in \mathbb{C}$ is a linear precoder, and $s_{d,k}$ satisfies $\mathbb{E} \{|s_{d,k}|^2\} = 1$. The received signal at the dUE in the $m$-th time slot is given as

$$y_d[m] = \sum_{k=1}^{K} g_k x_{d,k}[m - \tau_k] + z_d[m]$$

where $g_k$, $\tau_k$, $z_d$, and $g_k$ are defined in IeCRS. While the THz cooperative channel in IeCRS has the form of the frequency selective channel due to the identical message $W_d^k$, the THz cooperative channel in DeCRS has the form of the uplink multiple access channel, which treats the streams from other mUEs as interference. Assuming a conservative scenario that the interference from other mUEs always exists in every time slot, the achievable rate for the message $W_d^k$ in the second phase is given as

$$R_{d,k}^{(2)} = \log_2 \left( 1 + \frac{|g_k|^2}{J_k + 1} \right),$$

where $J_k = \sum_{i \neq k} |g_i|^2$ is the interference from other mUEs. Finally, the achievable rate for the message $W_d^k$ at the dUE is given as

$$R_d = \sum_{k=1}^{K} \min \left\{ C_{d,k}, R_d^{(2)} \right\},$$

which is the summation of the achievable rate for the message $W_d^k$ of every $k \in \mathcal{K}$. The minimum function is used to guarantee the successful decoding of the message $W_d^k$ in both the AP-(k-th mUE) and (k-th mUE)-dUE links. We design the precoders and message split for DeCRS in Section IV.

Remark 1: In IeCRS, all the mUEs must successfully decode the common message in the first phase. If there are mUEs with weak channel conditions, these mUEs may bound the performance for the first phase. In contrast, since multiple mUEs transmit the same stream in the second phase, the mUEs collaborate to strengthen the signal in the second phase. In result, the common message for multiple mUEs can be detrimental for the first phase, but can be beneficial for the second phase. DeCRS experiences this tradeoff opposite to IeCRS.

Remark 2: For IeCRS, the common message affects all the mUEs, since the common stream contains the messages for all $K$ mUEs. In contrast, the $k$-th common stream for DeCRS only contains the messages for the dUE and $k$-th mUE, thus, it only affects a single mUE. In result, we can expect that the common message will have more impact on IeCRS than DeCRS.

Remark 3: Thanks to its transmission framework, DeCRS can guarantee the privacy of the dUE. Since the message for the dUE is split into multiple messages and the split messages are decoded separately by different mUEs, the whole message for the dUE cannot be perfectly decoded by any of the mUEs.

D. Channel Model

A uniform planar array (UPA) with $N_A$ antennas is implemented at the AP. Hence, the channel for the AP-(k-th mUE) link is given as

$$h_k = \sqrt{\beta_0} \left( d_{AP-mUE}^{k} \right)^{-\alpha} \times \left[ 1, \ldots, \exp \left( j(N_v - 1) \pi \sin \phi_k \right) \right]^T \otimes \left[ 1, \ldots, \exp \left( j(N_h - 1) \cos \phi_k \cos \varphi_k \right) \right]^T,$$

where $\beta_0$ is the path-loss at the unit distance, $d_{AP-mUE}^{k}$ is the distance between the AP and k-th mUE, and $\alpha$ is the path-loss exponent. The array response vector of the channel is expressed with $\phi_k$ and $\varphi_k$, which are the vertical and horizontal angles between the AP and k-th mUE, $N_v$ is the number of vertical antenna elements, and $N_h$ is the number of horizontal antenna elements. The channel for the (k-th mUE)-dUE link is given as

$$g_k = \sqrt{\beta_0} \left( d_{mUE-dUE}^{k} \right)^{-\alpha} \exp (j\theta_k),$$

where $d_{mUE-dUE}^{k}$ is the distance between the k-th mUE and dUE, and $\theta_k$ is the phase of the channel. The corresponding time delay is modeled as

$$\tau_k = \text{round} \left( \frac{f_s d_{mUE-dUE}^{k}}{c} \right),$$

where $\text{round} (\cdot)$ is the round function, $f_s$ is the sampling frequency, and $c$ is the speed of light.

III. IDENTICAL EXTRACTION-BASED CRS

In this section, we formulate the overall problem for IeCRS and solve it through convex optimization techniques. Also, to compensate for the high complexity of the optimization problem, we derive a transmission strategy with minimal complexity. We assume perfect CSI while deriving the precoder and message split. The proposed channel estimation technique is described in Section V.
A. Problem Formulation

To enlarge the coverage of the system, we focus on maximizing the minimum achievable rate of the mUEs and dUE.\(^4\) The overall problem can be formulated as

\[
\text{(P1)}: \max \min \{ R_1, \ldots, R_K, R_d \}
\]

\[
s.t. \quad R_k = R_{p,k} + C_k, \quad \forall k \in \mathcal{K}, \quad \text{(1-a)}
\]

\[
C_d + \sum_{i=1}^{K} C_i \leq R_{c,k}, \quad \forall k \in \mathcal{K}, \quad \text{(1-b)}
\]

\[
R_d = \min \left\{ C_d, R_{d(2)} \right\}, \quad \text{(1-c)}
\]

\[
|\vec{g}_k|^2 \leq |\vec{g}_k|^2 P_k, \quad \forall k \in \mathcal{K}, \quad \text{(1-d)}
\]

where (1-d) is the power constraint of the AP, and (1-c) is the power constraint of the k-th mUE with power \(P_k\).

The optimization variables \(c\) and \(g\) are defined as \(c = [C_1, \ldots, C_K, C_d]^T\) and \(g = [\vec{g}_1, \ldots, \vec{g}_K]^T\), respectively. Since (P1) is non-convex, the problem cannot be directly solved. To resolve this issue, we split the problem into two separate problems, each maximizing \(C_d\) and \(R_{d(2)}\), and then compute \(R_d\) as in (1-c). The problem for the first phase is formulated as

\[
\text{(P2)}: \max \min \{ R_1, \ldots, R_K, C_d \}
\]

\[
s.t. \quad (1-a), (1-b), (1-d),
\]

and the problem for the second phase is formulated as

\[
\text{(P3)}: \max R_{d(2)}
\]

\[
s.t. \quad (1-e).
\]

In the following subsection, we derive the solutions for the problems (P2) and (P3).

B. Proposed Technique

We solve (P2) through a weighted minimum mean square error (WMMSE) approach and (P3) through a successive convex approximation (SCA) approach. Due to the high complexity of the SCA approach, we also derive a closed form solution for (P3) under the low SNR assumption.

From (P2), we observe that the problem is similar to the NOMA transmission scenario with RSMA [29], where we can interpret \([W_1, \ldots, W_K]\) as the unicast messages and \(W_d\) as the multicast message. Thus, we adopt the WMMSE approach to the problem (P2) with some adjustments to match our framework.

The k-th mUE first decodes \(W_c\) and removes the corresponding stream \(s_c\) from the received signal \(y_k\). Next, the k-th mUE decodes the private part \(W_{p,k}\). Through the equalizers \(w_{c,k}\) and \(w_{p,k}\), the k-th mUE estimates the streams \(\hat{s}_{c,k} = w_{c,k} y_k\) and \(\hat{s}_{c,k} = w_{p,k} (y_k - \hat{s}_{c,k})\), respectively. The mean squared errors (MSEs) of the k-th mUE can be expressed as

\[
\varepsilon_{c,k} = \|w_{c,k}\|^2 T_{c,k} - 2 \Re \{w_{c,k} H_k^H f_k\} + 1, \quad \text{(27)}
\]

\[
\varepsilon_{p,k} = \|w_{p,k}\|^2 T_{p,k} - 2 \Re \{w_{p,k} H_k^H f_k\} + 1, \quad \text{(28)}
\]

where \(T_{c,k} = \|h_k^H f_k\|^2 + \sum_{i=1}^{K} |h_k^H f_i|^2 + 1\) and \(T_{p,k} = T_{c,k} - |h_k^H f_k|^2\). To obtain the minimum MSE (MMSE) equalizers, we compute \(w_{c,k}^{\text{MMSE}} = \frac{\bar{H}_k^H T_{c,k}^{-1}}{\varepsilon_{c,k}}\) and \(w_{p,k}^{\text{MMSE}} = \frac{\bar{H}_k^H T_{p,k}^{-1}}{\varepsilon_{p,k}}\).

By substituting the MMSE equalizers, the MSEs can be expressed as \(\varepsilon_{c,k} = (T_{c,k} - \|h_k^H f_k\|^2) / T_{c,k}\) and \(\varepsilon_{p,k} = (T_{p,k} - \|h_k^H f_k\|^2) / T_{p,k}\). Then, the achievable rates can be expressed in alternate forms as \(R_{c,k} = -\log_2 (\varepsilon_{c,k})\) and \(R_{p,k} = -\log_2 (\varepsilon_{p,k})\). Furthermore, we define the weighted MSEs (WMSEs) as \(\xi_{c,k} = \mu_{c,k} \varepsilon_{c,k} - \ln (\mu_{c,k})\) and \(\xi_{p,k} = \mu_{p,k} \varepsilon_{p,k} - \ln (\mu_{p,k})\), where \(\mu_{c,k}\) and \(\mu_{p,k}\) are the weight variables. In result, the achievable rates can be expressed through the WMSEs as

\[
\xi_{c,k}^{\text{WMSE}} = \min_{\mu_{c,k}} \xi_{c,k} = 1 - R_{c,k} \ln (2), \quad \text{(31)}
\]

\[
\xi_{p,k}^{\text{WMSE}} = \min_{\mu_{p,k}} \xi_{p,k} = 1 - R_{p,k} \ln (2), \quad \text{(32)}
\]

where the optimal weights are derived as \(\mu_{c,k}^{\text{WMSE}} = (\xi_{c,k}^{\text{WMSE}})^{-1}\) and \(\mu_{p,k}^{\text{WMSE}} = (\xi_{p,k}^{\text{WMSE}})^{-1}\).

Through the relationship between the achievable rate and WMSE, the problem (P2) is transformed by interchanging the achievable rates into the WMSE forms as

\[
\text{(P2.1)}: \min_{\nu_{\text{CRS}}} t_0
\]

\[
s.t. \quad X_k + (\xi_{p,k} - 1) / \ln (2) \leq t_0, \quad \forall k \in \mathcal{K}, \quad \text{(21-a)}
\]

\[
(\xi_{c,k} - 1) / \ln (2) \leq X_d + \sum_{i=1}^{K} X_i, \quad \forall k \in \mathcal{K}, \quad \text{(21-b)}
\]

\[
X_d \leq t_0, \quad \text{(21-c)}
\]

where \(x = [X_1, \ldots, X_K, X_d]^T = -c\), and \(t_0\) is a slack variable to express \(\min \{ R_1, \ldots, R_K, C_d \}\). The variables of the optimization problem are defined as the set \(\nu_{\text{CRS}} = \{ F, x, w, \mu, t_0 \}\) with \(w = [w_{p,1}, \ldots, w_{p,K}, w_{c,1}, \ldots, w_{c,K}]^T\) and \(\mu = [\mu_{p,1}, \ldots, \mu_{p,K}, \mu_{c,1}, \ldots, \mu_{c,K}]^T\).

By solving (P2.1), the resulting minimum rate of the system is \(\min \{ R_1, \ldots, R_K, C_d \} = -t_0\). While (P2.1) is non-convex in general, the problem is convex with respect to each variable set \(\{ F, x, t_0 \}\) and \(\{ w, \mu \}\) by fixing the other variable set. Thus, the problem can be efficiently solved iteratively through alternating optimization (AO).

To solve the non-convex problem (P3), we introduce slack variables to adopt the SCA approach. The problem can be
transformed as

$$\text{(P3.1):} \max_{g,a,b,u} \frac{1}{N_c} \sum_{n=1}^{N_c} \log_2(1 + u_n)$$

$$\text{s.t.} \ u_n \leq a_n^2 + b_n^2, \ \forall n \in \mathcal{N},$$

$$a = \Re\{g\}, \ b = \Im\{g\}, \ \forall n \in \mathcal{N}, \text{ (3.1-a), (1-e)},$$

where \( \tilde{g} = [\tilde{g}_1, \cdots, \tilde{g}_{N_c}]^T \). The slack variables are defined as \( a = [a_1, \cdots, a_{N_c}]^T, b = [b_1, \cdots, b_{N_c}]^T \), and \( u = [u_1, \cdots, u_{N_c}]^T \). The sub-carriers are indexed by a set \( \mathcal{N} = \{1, \cdots, N_c\} \). To maximize the objective, \( u_n \) will be maximized until (3.1-a) is met with equality, resulting in \( u_n = |\tilde{g}_n|^2 \). Thus, the problems (P3) and (P3.1) are indeed equivalent. While (P3.1) is non-convex due to (3.1-a), the right-hand side of the constraint is convex, motivating us to use the SCA approach. In result, the surrogate optimization problem for the \( \ell \)-th iteration will be

$$\text{(P3.2):} \max_{g,a,b,u} R_{\ell}^{(2)} = \frac{1}{N_c} \sum_{n=1}^{N_c} \log_2(1 + u_n)$$

$$\text{s.t.} \ u_n \leq \tilde{a}_n^{(\ell)} + \tilde{b}_n^{(\ell)}, \ \forall n \in \mathcal{N}, \text{ (3.2-a), (1-b), (1-e)},$$

where \( \tilde{a}_n^{(\ell)} \) and \( \tilde{b}_n^{(\ell)} \) are the first-order derivatives of the right-hand side of (3.1-a) defined as

$$\tilde{a}_n^{(\ell)} = \left( a_n^{(\ell)} \right)^2 + 2 \left\{ a_n^{(\ell)} \left( a_n - a_n^{(\ell)} \right) \right\}, \text{ (33)}$$

$$\tilde{b}_n^{(\ell)} = \left( b_n^{(\ell)} \right)^2 + 2 \left\{ b_n^{(\ell)} \left( b_n - b_n^{(\ell)} \right) \right\}, \text{ (34)}$$

with the variables \( a_n^{(\ell)} \) and \( b_n^{(\ell)} \) defined as the local points for the \( \ell \)-th iteration. By iteratively solving (P3.2) and updating \( a_n^{(\ell)} \) and \( b_n^{(\ell)} \) with the solutions from the \( (\ell - 1) \)-th iteration, the solution will converge to a local optimum of (P3.1) [35]. By taking the minimum of \( C_d \) and \( R_d^{(2)} \), we obtain the achievable rate \( R_d \). The overall algorithm for IeCRS is shown in Algorithm 1.

**Algorithm 1** Pseudo Code for Minimum Rate Maximization in IeCRS

1. **Initialization**: Set \( \ell_1 = 0, w^{(\ell_1)}, \text{ and } \mu^{(\ell_1)} \).
2. **repeat**
   3. Set \( \ell_1 = \ell_1 + 1 \).
   4. Solve (P2.1) with fixed \( w \) and \( \mu \).
   5. Update \( w^{(\ell_1)} \) and \( \mu^{(\ell_1)} \).
6. **until** \( t_0 \) decreases by a fraction below a predefined threshold.
7. \( C_d = -t_0 \).
8. Set \( \ell_2 = 0, a^{(\ell_2)}, \text{ and } b^{(\ell_2)} \).
9. **repeat**
   10. Set \( \ell_2 = \ell_2 + 1 \).
   11. Solve (P3.2).
   12. Update \( a^{(\ell_2)} \) and \( b^{(\ell_2)} \) as the solutions of (P3.2).
   13. **until** \( R_d^{(2)} \) increases by a fraction below a predefined threshold.
   14. \( R_d = \min \left\{ C_d, R_d^{(2)} \right\} \).

**C. Low Complexity Approach**

While the convex optimization approach is effective in performance, the complexity of iteratively solving (P3.2) is quite high, where the slack variables have the size of the number of OFDM sub-carriers. Due to the excessive use of bandwidth in THz frequencies, the number of sub-carriers are expected to be huge. To compensate for this factor, we propose a low complexity approach to solve (P3).

Due to the low transmit power of the mUEs and severe path-loss of the THz frequency bands, we assume that the DUE operates in the low SNR regime during the second phase. The achievable rate \( R_d^{(2)} \) can then be simplified as

$$\sum_{n=1}^{N_c} \log_2(1 + |\tilde{g}_n|^2) \approx \sum_{n=1}^{N_c} |\tilde{g}_n|^2 / \ln(2). \text{ (35)}$$

We also express \( \tilde{g}_n \) in an alternate form from (11) as \( \tilde{g}_n = \Omega_n^T \tilde{g} \), where \( \Omega_n = [\Omega_{n,1}, \cdots, \Omega_{n,K}]^T \), with \( \Omega_{n,k} = \exp(j2\pi\tau_k n/N_c) \).

**Algorithm 1** Pseudo Code for Minimum Rate Maximization in IeCRS

1. **Initialization**: Set \( \ell_1 = 0, w^{(\ell_1)}, \text{ and } \mu^{(\ell_1)} \).
2. **repeat**
   3. Set \( \ell_1 = \ell_1 + 1 \).
   4. Solve (P2.1) with fixed \( w \) and \( \mu \).
   5. Update \( w^{(\ell_1)} \) and \( \mu^{(\ell_1)} \).
6. **until** \( t_0 \) decreases by a fraction below a predefined threshold.
7. \( C_d = -t_0 \).
8. Set \( \ell_2 = 0, a^{(\ell_2)}, \text{ and } b^{(\ell_2)} \).
9. **repeat**
   10. Set \( \ell_2 = \ell_2 + 1 \).
   11. Solve (P3.2).
   12. Update \( a^{(\ell_2)} \) and \( b^{(\ell_2)} \) as the solutions of (P3.2).
   13. **until** \( R_d^{(2)} \) increases by a fraction below a predefined threshold.
   14. \( R_d = \min \left\{ C_d, R_d^{(2)} \right\} \).

**Proof**: For \( a \neq b \), \( \{a, b\} \in \mathcal{K} \), the \( (a, b) \)-th element of \( \Omega \) can be derived as

$$\Omega[a, b] = \sum_{n=1}^{N_c} \exp(j2\pi(\tau_a - \tau_b)n/N_c)$$

$$= \sum_{n=1}^{N_c} \omega_{a,b} n^{(k)} = \omega_{a,b} 1 - \omega_{a,b}^{N_c} 1 - \omega_{a,b}$$

$$= 0, \text{ (38)}$$

where \( \omega_{a,b} = \exp(j2\pi(\tau_a - \tau_b)/N_c) \). The \( a \)-th diagonal element of \( \Omega \) can be derived as \( N_c \) in the same way, which finishes the proof.

Using Lemma 1, (P3) can be expressed as

$$\text{(P3.3):} \max_{g} \frac{1}{\ln(2)} \| \tilde{g} \|^2$$

$$\text{s.t.} \ (1-e).$$

From (P3.3), we observe that the optimal communication strategy in the low SNR regime is to simply use all the power of the mUEs, where their phase values are irrelevant. This

5In the rare occasion when the delays of multiple mUEs overlap, the optimal strategy is derived in a similar manner. In result, the mUEs have to use all its power, and the received signals from the mUEs with the same delays must have equal phase values.
strategy results in a drastic reduction of complexity, adequate for scenarios with huge numbers of sub-carriers.

IV. DISTINCT EXTRACTION-BASED CRS
A. Problem Formulation

Similar to IeCRS, we aim to maximize the minimum achievable rate of the system. The overall problem is formulated as

\[
(P4) : \max_{F, \mathbf{c}, \mathbf{g}} \min \{ R_1, \ldots, R_K, R_d \}
\]

\[
s.t. \quad C_{d,k} + C_k \leq R_{c,k}, \quad \forall k \in \mathcal{K},
\]

\[
R_d = \sum_{k=1}^{K} \min \{ C_{d,k}, R_i^{(2)} \},
\]

\[(1-a), (1-d), (1-e),\]

where \( \mathbf{c} = [C_1, \cdots, C_K, C_{d,1}, \cdots, C_{d,K}] \) for DeCRS.

B. Proposed Technique

In this subsection, we solve (P4) through the WMMSE approach. We assume that the dUE uses the equalizer \( w_{d,k} \) to estimate the stream \( \hat{s}_{d,k} = w_{d,k}y_d \). Similar to IeCRS, we define the MSEs of the \( k \)-th mUE as

\[
\varepsilon_{c,k} = |w_{c,k}|^2 T_{c,k} - 2 \Re\{ w_{c,k} \mathbf{h}_k^H \mathbf{f}_c \} + 1,
\]

\[
\varepsilon_{p,k} = |w_{p,k}|^2 T_{p,k} - 2 \Re\{ w_{p,k} \mathbf{h}_k^H \mathbf{f}_p \} + 1,
\]

where \( T_{c,k} = \sum_{l=1}^{K} |\mathbf{h}_k^H \mathbf{f}_l|^2 + \sum_{l=1}^{K} |\mathbf{h}_k^H \mathbf{f}_d|^2 + 1 \) and \( T_{p,k} = T_{c,k} - |\mathbf{h}_k^H \mathbf{f}_d|^2 \). We define the MSE of the stream for the dUE from the \( k \)-th mUE as

\[
\varepsilon_{d,k} = |w_{d,k}|^2 T_{d,k} - 2 \Re\{ w_{d,k} \hat{y}_k \} + 1,
\]

where \( T_{d,k} = \sum_{l=1}^{K} |\hat{y}_l|^2 + 1 \). The definitions of the WMMSEs and weights are neglected due to redundancy.

In result, (P4) can be transformed as

\[
(P4.1) : \min_{\mathcal{V}_{DeCRS}} t_0
\]

\[
s.t. \quad X_k + (\xi_k - 1)/\ln(2) \leq t_0, \quad \forall k \in \mathcal{K},
\]

\[
(\xi_k - 1)/\ln(2) \leq X_{d,k} + X_k, \quad \forall k \in \mathcal{K},
\]

\[
X_{d,k} \leq t_k, \quad \forall k \in \mathcal{K}, \quad (4.1-c)
\]

\[
(\xi_{d,k} - 1)/\ln(2) \leq t_k, \quad \forall k \in \mathcal{K},
\]

\[
\sum_{k=1}^{K} t_k \leq t_0,
\]

\[(1-d), (1-e),\]

where the set of optimization variables is defined as \( \mathcal{V}_{DeCRS} = \{ F, \mathbf{x}, \mathbf{w}, \mathbf{\mu}, t_0, t \} \), with \( \mathbf{x} = -\mathbf{c} \) and \( t = [t_1, \cdots, t_K]^T \).

By introducing the slack variables \( t_k \) for every \( k \in \mathcal{K} \), the constraint (4-b) is expressed by the constraints (4.1-a)-(4.1-e), where (4.1-c) limits the rate of \( C_{d,k} \), (4.1-d) limits the rate of \( R_i^{(2)} \), and (4.1-e) sums up the rates of the mUEs. Through the AO approach, (P4.1) is solved to reach a local optimum. The overall algorithm for DeCRS is shown in Algorithm 2.

Remark 4: Even if there exist some mUEs where the channels between the mUEs and dUE are blocked, the proposed techniques can adapt to this scenario. For IeCRS, while blockage will reduce the overall channel gain, the proposed technique still works without any change. For DeCRS, if the (k-th mUE)-dUE link is blocked, the AP will not allocate the message for the dUE to the k-th mUE, i.e., \( C_{d,k} = 0 \). Thus, DeCRS also can adapt to blockage between the mUEs and dUE.

C. Convergence and Complexity Analyses

For IeCRS, the original problem (P1) is split into two independent problems (P2) and (P3). The problems (P2) and (P3) are solved iteratively through (P2.1) and (P3.2), where the problems are based on the WMMSE and SCA approaches, respectively. By considering the facts that the WMMSE and SCA approaches converge to stationary points [28], [35], and that (P2) and (P3) are independent problems from (P1), we conclude that the solution for IeCRS converges to a local optimum solution.

To analyze the complexity of IeCRS, first note that the problems (P2.1) and (P3.2) are both quadratically constrained quadratic program (QCQP) problems, which have the complexities of \( O(\|KN\|_1^{3.5}) \) and \( O(\|N_c\|_1^{3.5}) \), respectively [36]. With the threshold \( \epsilon \), the number of iterations for convergence are approximated as \( O(\log(\epsilon^{-1})) \), and the worst-case complexity of Algorithm 1 can be expressed as \( O(\log(\epsilon^{-1})(\|KN\|_1^{3.5} + \|N_c\|_1^{3.5})) \). The low complexity approach solves (P3) in closed form, and the worst-case complexity of the low complexity approach can be expressed as \( O(\log(\epsilon^{-1})|KN|_1^{3.5}) \).

For DeCRS, the problem (P4) is solved iteratively through (P4.1) based on the WMMSE approach. Also, (P4.1) is a QCQP problem. Thus, derived in the same manner as IeCRS, the solution of Algorithm 2 converges to a local optimum point and has the worst-case complexity as \( O(\log(\epsilon^{-1})|KN|_1^{3.5}) \).

V. CHANNEL ESTIMATION

In the transmission framework of eCRS, the AP requires the CSI of both the downlink and THz cooperative channels to find the optimal beamformers for the system. In the reception process, the k-th mUE requires the CSI of the AP-(k-th mUE) link, and the dUE requires the CSI of every
link in the THz cooperative channel to decode the intended message. We exploit the channel reciprocity by employing the time division duplexing (TDD). The channel of the AP-(k-th mUE) link, which appears in (3) and (15) can be estimated by the conventional MU-MISO downlink channel estimation techniques [37], [38], [39]. For example, the K mUEs transmit orthogonal pilot signals to the AP, and the AP estimates the reciprocal channel of the AP-(k-th mUE) link from the received pilot signal. We omit the detailed process of the channel estimation for the downlink channel since it is well described in other literatures.

We focus on the channel estimation for the (k-th mUE)-dUE link, which appears in (8) and (20). Although the channel has the form of the frequency selective channel, the conventional estimation technique is not suitable since the AP requires not only the channel gain but also the time delay of the specific mUE. Hence, we propose a novel channel estimation technique for the THz cooperative channel, which first estimates the time delay \( \tau_k \) and then the channel gain \( g_k \).

In prior, the k-th mUE and dUE share the information about the length \( N_p \) pilot given as
\[
\psi_k = [\psi_k[0], \ldots, \psi_k[N_p - 1]]^T,
\]
which satisfies \( \|\psi_k\|^2 = N_p \). Thus, the k-th mUE transmits the signal during \( N_k \) time slots given as
\[
x_{d,k}[m] = \sqrt{P}\psi_k[m], \quad m = 0, \ldots, N_p - 1,
\]
where \( P \) is the transmit power of the pilot signal. The dUE receives the pilot signals through the THz cooperative channel during \( N_r \) time slots. The length \( N_r \) should satisfy the inequality \( \tau_{\text{max}} + N_p \leq N_r \) so that every pilot signal is received at the dUE, where \( \tau_{\text{max}} \) is defined as \( \tau_{\text{max}} = \max_{k} \tau_k \). The received signal at the dUE in the \( m \)-th time slot is given as
\[
y_d[m] = \sum_{k=1}^{K} g_k x_{d,k}[m - \tau_k] + z_d[m] = \sum_{k=1}^{K} g_k \sqrt{P}\psi_k[m - \tau_k] + z_d[m].
\]
The received signals during \( N_r \) time slots can be reformulated as
\[
y = \sqrt{P}\Psi \mathbf{g} + \mathbf{z},
\]
where \( y = [y_d[0], \ldots, y_d[N_r - 1]]^T \), \( \mathbf{g} = [g_1, \ldots, g_K]^T \), and \( z = [z_d[0], \ldots, z_d[N_r - 1]] \). The matrix \( \Psi \) is defined as \( \Psi = [\psi_1(\tau_1), \ldots, \psi_K(\tau_K)] \) where \( \psi_k(\tau) \) is a \( \tau \)-shifted vector of the \( \psi_k \) defined as \( \psi_k(\tau) = [0^T, \psi_k^T, 0^T_{(N_r-N_p-\tau)}]^T \).

A. Time Delay Estimation

The dUE estimates the time delay \( \tau_k \) and then regenerates the \( k \)-th column of the matrix \( \Psi \) by exploiting the estimated time delay \( \hat{\tau}_k \) and pilot \( \psi_k \). To estimate the time delay \( \tau_k \), the dUE projects the \( \tau \)-shifted vector \( \psi_k(\tau) \) onto the received vector \( y \) such as
\[
r_k(\tau) = \sqrt{P} \left( \psi_k(\tau) \right)^H \left( \psi_k(\tau) \right) g_k + \sum_{i \neq k} \left( \psi_i(\tau) \right)^H \psi_i(\tau) g_i \right) \right) + \left( \psi_k(\tau) \right)^H \mathbf{z},
\]
where \( \tau \) ranges from 0 to \( \tau_{\text{max}} \). We propose a maximum projection (MP) estimator given as
\[
\hat{\tau}_k = \arg \max_{\tau} |r_k(\tau)|,
\]
which searches for \( \tau \) such that the magnitude of the projected value is maximized. To fully utilize the MP estimator, we implement the pilot that has a pseudo-noise property, where the auto-correlation is given as
\[
R_k(\tau_1 - \tau_2) = \left( \psi_k(\tau_1) \right)^H \psi_k(\tau_2) \ll \|\psi_k\|^2, \quad \forall \tau_1 \neq \tau_2.
\]
We also assume that the pilot has the cross-correlation given as
\[
R_{k,k'}(\tau_1, \tau_2) = \left| \left( \psi_k(\tau_1) \right)^H \psi_{k'}(\tau_2) \right| \approx 0, \quad \forall k \neq k',
\]
which can suppress the interference term in (46). With the idealistic properties of the pilot, the magnitude of the projected value \( r_k(\tau) \) is given as
\[
|r_k(\tau)| \approx \sqrt{P} \left| \left( \psi_k(\tau) \right)^H \psi_k(\tau) g_k + \left( \psi_k(\tau) \right)^H \mathbf{z} \right| \ll |r_k(\tau_k)|, \quad \forall \tau \neq \tau_k.
\]
Hence, the MP estimator can estimate the exact time delay \( \tau_k \) with the idealistic properties. In this paper, we implement the Zadoff-chu sequence in [40] for the pilot, where the idealistic properties hold when the sequence length is sufficiently long.

B. Channel Gain Estimation

From the estimated time delay \( \hat{\tau}_k \) and pilot \( \psi_k \), the dUE regenerates the matrix \( \tilde{\Psi} \) such as
\[
\tilde{\Psi} = \left[ \psi_1^{(\tau_1)}, \ldots, \psi_K^{(\tau_K)} \right].
\]
We implement the least-square (LS) estimation technique to estimate the channel such as
\[
\mathbf{g} = \left( \tilde{\Psi}^H \tilde{\Psi} \right)^{-1} \tilde{\Psi}^H y.
\]
In our scenario of interest, the dUE cannot feedback the CSI to the AP through the THz frequency bands since the LoS link between the AP and dUE is blocked. In practice, mobile devices may be able to utilize multiple frequency bands. Hence, we assume that the dUE feeds back the CSI of THz cooperative channel through lower frequency bands, where only a limited amount of the message transmission is available.
m and \( m \)UEs is fixed as \( P \)
pilot lengths, which approximates to the idealistic properties
we implement the Zadoff-chu sequence with sufficiently long
cross-correlation decrease as the pilot length increases. Hence,
of \( \tau \)
uniformly spread in a box with diagonal coordinates
at \( [0,1,0] \), respectively, and the noise power spectral density is
verified with the delay error rate (DER) of the time delay
of the AP is assumed as \( \tau \)
throughout the simulations. Unless stated otherwise, the power
power of the AP and mUEs range from \(-10\) dBm to \(20\) dBm
the limited transmit power of current THz signal generators [41], the transmit
power of the AP and mUEs range from \(-10\) dBm to \(20\) dBm
of every mUE to be equal and the pilot length as 100.

In Fig. 4 (a), we measure the auto-correlation and cross-correlation
of the Zadoff-chu sequence. The auto-correlation value is averaged out for the
correlation of the Zadoff-chu sequence with respect to the
pilot length. The auto-correlation and cross-correlation of the Zadoff-chu
sequence in Fig. 4 (a), the DER and NMSE performances.

Fig. 4. Performances of the proposed channel estimation technique with
respect to the pilot length \( N_p \), where \( K = 5 \) and \( N_f = 16 \).

VI. SIMULATION RESULTS

In this section, we verify the performances of the channel
estimation technique and the two cases of eCRS with one SIC
layer. For the simulations, the carrier frequency, bandwidth,
and path-loss exponent are fixed as \( f_c = 0.3 \) THz, \( B = 1 \) GHz,
and \( \alpha = 2 \), respectively, and the noise power spectral density is
fixed as \( N_0 = -174 \) dBm/Hz. Considering the limited transmit
power of current THz signal generators [41], the transmit
power of the AP and mUEs range from \(-10\) dBm to \(20\) dBm
throughout the simulations. Unless stated otherwise, the power
of the AP is assumed as \( P_{AP} = 20 \) dBm, and the power of the
mUEs is fixed as \( P_k = 0 \) dBm. The AP and dUE are located
at \([0,4,1] \) m and \([8,4,0] \) m, respectively, and the mUEs are
uniformly spread in a box with diagonal coordinates \([2,0,0] \)
m and \([6,8,0] \) m.

The performance of the channel estimation technique is
verified with the delay error rate (DER) of the time delay
and the normalized mean square error (NMSE) of the channel
gain, which are given as

\[
\text{DER} = \frac{1}{K} \sum_{k=1}^{K} \text{Pr} (\hat{\tau}_k \neq \tau_k), \quad (53)
\]

\[
\text{NMSE} = \mathbb{E} \left[ \frac{\| \hat{g} - g \|^2}{\| g \|^2} \right], \quad (54)
\]

respectively.

In Fig. 4 (a), we measure the auto-correlation and cross-correlation
of the Zadoff-chu sequence with respect to the pilot length. The auto-correlation value is averaged out for the
cases of \( \tau_1 \neq \tau_2 \). The auto-correlation and cross-correlation
are normalized with \( N_p \) so that the auto-correlation for the case
of \( \tau_1 = \tau_2 \) is one. We observe that the auto-correlation and
cross-correlation decrease as the pilot length increases. Hence,
we implement the Zadoff-chu sequence with sufficiently long
pilot lengths, which approximates to the idealistic properties
given in (48) and (49). In Fig. 4 (b), the DER and NMSE are measured with respect to the pilot length to verify the performance of the proposed channel estimation technique.
The DER decreases as the pilot length increases and eventually
saturates to zero. This is because the DER performance
strongly depends on the idealistic properties in (48) and (49).
The NMSE of the channel gain also decreases as the pilot
length increases since the matrix \( \Psi \) for the LS estimation
is highly related to the DER performance.

In Fig. 5, we investigate the DER and NMSE of the
proposed channel estimation technique with respect to the
transmit power of the mUEs. We set the transmit power of every mUE to be equal and the pilot length as 100.
We also investigate three different cases by changing the number of mUEs as 5, 10, and 15. In Fig. 5 (a), the DER
decreases as the transmit power increases for all three cases
due to the increasing signal-to-interference-plus-noise ratio
(SINR). However, in the high transmit power regime, the DER
tends to saturate since the SINR saturates as the transmit
power increases. As the number of mUEs increases, the DER
increases because the interference term in (46) increases due
to the degradation of the cross-correlation property. For the
NMSE in Fig. 5 (b), all three cases decrease as the transmit
power increases. In the high power regime, the cases with
the number of mUEs of 10 and 15 saturate because the DER
directly affects the channel gain estimation through the
estimated delays \( \{ \hat{\tau}_1, \cdots, \hat{\tau}_K \} \).

The cases of eCRS with one SIC layer are denoted as
IeCRS, DeCRS, and LOW, where IeCRS and DeCRS adopt
the convex optimization approach in IeCRS and DeCRS cases,
respectively, and LOW adopts the low complexity approach for
IeCRS. To compare the results of our proposed framework,
we consider three types of benchmarks, namely, identical
cooperative NOMA (IC-NOMA), distinct cooperative NOMA
(DC-NOMA), and single tap (ST). IC-NOMA and DC-NOMA
are IeCRS and DeCRS without using common messages, similar to NOMA, respectively. ST is the ideal case where the signals from the mUEs arrive simultaneously while using IeCRS. The precoders for the first phase of IeCRS, LOW, IC-NOMA, and ST are initialized by using maximum ratio transmission (MRT) combined with singular value decomposition (SVD) as in [26]. The precoders for the first phase of DeCRS and DC-NOMA are initialized by using MRT for both the private and common streams. The second phase for all cases is initialized by the mUEs using all their power.

In Fig. 6, we plot the achievable rate with respect to the number of mUEs. Unlike conventional transmission techniques, where the rate decreases as the number of mUEs increases, we observe that there exists a certain performance peak for our proposed framework. This peak is due to the two-phase nature of eCRS. For the first phase, the transmission is similar to the conventional MU-MISO downlink channel, where the achievable rate decreases as the number of mUEs increases. For the second phase, as the number of mUEs increases, the achievable rate of the dUE naturally increases. In result, the performance increases for a small number of mUEs, where the performance bottleneck is from the second phase, and the performance decreases for a large number of mUEs due to the bottleneck of the first phase.

We observe that IeCRS and LOW have similar performances. This states that the low SNR assumption for the second phase is valid for our scenario of interest. DeCRS seems to have lower performance compared to IeCRS and LOW. This is due to two factors. First, the achievable rate that we consider is already a lower bound. Similar to IeCRS, different signals from the mUEs arrive in different instances in general. Since we neglected this factor, the performance is degraded. Also, DeCRS adopts more streams for both the first and second phases. Since the mUEs and dUE are all equipped with a single antenna, the interference is crucial. We expect that with multiple antenna mUEs and dUEs, the performance of DeCRS will increase drastically, which is left for future work. Finding the balance of the interference and noise through the common message of the mUEs, we observe that IeCRS outperforms IC-NOMA. However, as expected, DeCRS has the same performance as DC-NOMA since the common messages do not control the interference between the mUEs. We observe that due to the signals arriving in different taps, the performance of IeCRS is upper bounded with ST, showing that there is an inevitable performance loss to consider fast-sampling communication systems. Note that, the signals arriving in different taps is beneficial for DeCRS, since the interference will decrease. However, this will be detrimental for IeCRS, since this would decrease the strength of the overall signal.

In Fig. 7, we plot the achievable rate with respect to the number of mUEs, where the AP and mUEs have low transmit power. Similar to Fig. 6, all cases have performance peaks, where the peaks are shifted to the left due to the low transmit power of the AP. The proposed techniques all have similar performances when there are a small number of mUEs. This is because with low SNR and small interference, the achievable rate of DeCRS becomes effectively the same as that of LOW, which can be easily derived from (23). Another result to note is that DeCRS outperforms IeCRS when there are a large number of mUEs. For DeCRS, the AP can concentrate the power to specific mUEs by selecting a few mUEs for cooperation. We also observe that there is no significant performance improvement of IeCRS compared to IC-NOMA, since the benefit of the common message that finds the balance between the interference and noise is limited due to the dominant AWGN.

In Fig. 8, the ratios between the achievable rates of perfect and imperfect CSI cases of IeCRS, LOW, and DeCRS are simulated. The length of the coherence time block is set to 1200 as in [42] and [43], which corresponds to 1.2 μsec in our system of interest, and we assume that the perfect CSI case has no estimation time to observe the tradeoff between the pilot length and achievable rate. We observe that the ratio of each technique increases, owing to the reduced channel estimation error, until it reaches a peak point. Then, the ratio starts to decrease since the impact of reduced transmission time becomes larger than the performance gain from the accurate channel estimation. Hence, the pilot length should be considered to maximize the minimum achievable rate for practical communication scenarios.

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3While various weights were given for the initialization of the common and private streams of DeCRS and DC-NOMA, we verified that this does not affect the performances of both techniques.
In this paper, we proposed a novel message transmission framework to increase the coverage for a THz MU-MISO downlink system through cooperative communication using RSMA. For message transmissions, we proposed eCRS, and explored two specific cases, which are IeCRS and DeCRS. Based on the novel THz cooperative channel model, we derived local optimal solutions of IeCRS and DeCRS through convex optimization techniques as well as a closed form solution for IeCRS in the low SNR regime. Finally, to successfully use our framework in practice, we proposed a channel estimation technique to detect the channel gains and time delays of the THz cooperative channel model. Through simulation results, we confirmed that our proposed message transmission framework has considerable performance, and that our estimation technique successfully captures the full capabilities of the THz cooperative channel model.

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