Exploiting interactions among signals to decode interfering transmissions with fewer receiving antennas

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A R T I C L E   I N F O

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A B S T R A C T

Due to the hierarchical structure and heterogeneity of most deployed wireless networks, multiple data transmissions using the same communication resource, such as same frequency band and time slot, are likely to occur in the same area. Hence, interference becomes critical to the decoding of interfered signals. There have already been numerous techniques dealing with interferences that can be implemented either at the transmitter (Tx) or the receiver (Rx) or both, to support simultaneous multi-user transmissions. We focus on the reception design in multiple access channels (MACs) where multiple Txs send data to a common Rx simultaneously. By exploiting the constructive/destructive interactions among interfering signals, we propose a Rx structure based on interference combination (ICom) that incorporates zero-forcing (ZF) and successive interference cancellation (SIC). In addition, by exploiting the relationship between the desired signal and interference, reception mode adaptation is employed to further improve the performance of the proposed reception mechanism. The proposed Rx structure does not require coordination at the Tx side and can significantly reduce the number of receiving antennas at a moderate processing cost. The ICom-based reception is shown to be able to make not only a significant improvement of system spectral efficiency (SE) under stringent latency constraints, but also a flexible tradeoff between the requirements of receive antennas and signal processing complexity, thus facilitating its implementation and deployment.

1. Introduction

With the rapid development of wireless communication technologies, most deployed networks are hierarchically organized [1] and have heterogeneous features, making the interferences among them more complex and difficult to manage. On one hand, when new technologies are deployed to accommodate multiple subscribers, more interferences will be likely to happen. On the other hand, as the degree of frequency reuse increases, multiple co-located transmissions may share the same frequency band, thus introducing various types of cochannel interference (CCI). These problems must be addressed adequately, else the improvement of network performance will be limited or even impossible.

User scheduling [2] can be used to select a set of subscribers so as to simplify and handle interferences while blocking unscheduled subscribers. However, bursty traffic – especially in the networks where the number and the transmission demand of users vary widely and dynamically – cannot be supported. As a result, how to exploit the system capacity to accommodate as many subscribers as possible becomes a key issue. The interference caused by a user’s transmission in the network is known to be a structured signal [3]. In contrast to structureless noises, the interference’s signature can then be exploited in the design of a transmission mechanism. Thanks to the array signal processing capability brought by multi-antenna technology, multiple interfering signals can be distinguished in the spatial domain. By generating appropriate directional beams and/or employing receive filters, interferences can be managed effectively. There have been numerous promising interference management (IM) schemes, including interference cancellation (IC) [3], interference alignment (IA) [4], interference neutralization (IN) [5], adaptive filtering, etc. With these schemes, interference can be manipulated separately or jointly at Tx and/or Rx side to recover multiple simultaneous interfering signals.

Although IA and IN emerge as promising IM schemes, their applicability is still limited for the following reasons.

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First, both schemes require coordination at the Tx side. Especially for IN, multiple interfering signals should carry the same information to neutralize them. So, simultaneously-transmitted signals should be shaped interdependently, thereby sacrificing some users’ quality-of-service (QoS) for preservation of the others’. Moreover, in wireless multiple access channels, due to the capability limit and the overhead of mobile terminals, non-interoperability between different operators, etc., coordination among mobile users is expensive and not always possible.

Second, the DoF (Degree of freedom) requirement is critical to the feasibility of IA and IN. Tx should be equipped with multiple antennas so as to form directional beams with differentiated spatial signatures. The number of receiving antennas should also be greater than or equal to the total number of independent data streams from all Txs, so that multiple transmissions can share time and frequency resources via spatial division multiplexing [6,7].

Given specific Tx-side parameter settings, increasing the number of receiving antennas seems an effective way to decode more interfering signals. However, due to space constraint and hardware cost, it is impractical to increase receiving antennas beyond a certain limit, especially for mobile terminals. In the design of wireless networks, multiple concurrent transmissions to a common destination have become popular. It is, therefore, of practical importance to develop a reception mechanism that can decode multiple concurrent signals with fewer receiving antennas.

Note that the overlapping of multiple radio frequency (RF) signals is essentially the superposition of electromagnetic waves [8–10]. Fig. 1 shows the constructive or destructive interactions between two interfering signals of the same strength, and perceived by a common Rx. It illustrates three cases in which the waves in the left part of the figure indicate two signals whereas dark lines in the right part represent their superposition. In Fig. 1(a), when electromagnetic waves interfere constructively, the superposition signal gets stronger than each individual waveform. Fig. 1(b) shows that the superposition has no electrical energy since the two destructively interfering electromagnetic waves weaken each other. Fig. 1(c) shows a general case where the superposition of the two waves becomes arbitrary.

Based on the above observations, we exploit the interactions among multiple interfering signals to reduce the number of receiving antennas while preserving the capability of recovering the interfering signals. The proposed mechanism does not require any Tx-side cooperation — this is especially significant when the Txs are mobile terminals, i.e., all Txs can send signals to a common Rx without consuming any resource for the others. Thanks to the advanced signal processing capability of hardware, the reduction of receiving antennas can be compensated for via more sophisticated signal processing inside the equipments.

The main contributions of this paper are three-fold:

- Development of a reception mechanism that exploits the constructive/destructive interactions among multiple interfering signals, thus reducing the requirement of receiving antennas significantly while preserving the capability of recovering multiple concurrently-transmitted signals.
- Making a flexible tradeoff between the number of receiving antennas and the computational complexity. The complexity can be reduced further by using SIC, and an extension of the proposed reception mechanism is elaborated in more general system settings.
- Derivation of an adaptive reception mode selection criterion based on the relationship between the desired signal and interference, according to which a good tradeoff between interference suppression and desired signal distortion is achieved.

We will use the following notations throughout the paper. The set of complex numbers is denoted by $\mathbb{C}$, while vectors and matrices are represented by bold lower-case and upper-case letters, respectively. The Hermitian (or conjugate transpose), pseudo-inverse and inverse of a vector or a matrix are denoted by $(\cdot)^H$, $(\cdot)^+$( and $(\cdot)^{-1}$, respectively, while $E(\cdot)$ and $\|\cdot\|$ represent statistical expectation and the Euclidean norm.

The rest of this paper is organized as follows. Section 2 discusses the work related to interference management for multi-user communications. Section 3 describes the system model and assumptions, and Section 4 presents how to exploit the combination of interferences. Section 5 discusses two ways to improve the performance of the proposed scheme. Section 6 evaluates the proposed scheme and finally, Section 7 concludes the paper.

2. Related work

Interference alignment (IA) has been under development in recent years [11–16]. Its principle is to confine all interferences to a subspace of minimal dimensions at Rx so as to maximize the available dimensions for the intended signals. Via preprocessing at the Tx side, multiple interfering signals are mapped into a finite subspace, so that desired signal(s) may be sent through a subspace without attenuation. The feasibility of IA is shown in [11] to be highly dependent on system parameters, such as the numbers of Txs and Rxs, configuration of transmitting and receiving antennas, etc. Opportunistic IA was proposed in [12] for a large number of users to harvest the multisizer diversity so as to facilitate the implementation of IA. IA-based coordinated beamforming was proposed in [13] to improve the downlink performance of multiple cell-edge users in multi-user multiple-input multiple-output (MU-MIMO) systems. IA-based uplink IM for two-tier cellular systems was devised in [14]. In order to manage the uplink interference caused by macrocell users at femtocell base stations (FBSs), cooperation between macrocell users with the closest FBSs was proposed with the goal of aligning the received signals of macrocell users in the same subspace at multiple FBSs. The author of [15] designed an IA-based uplink transmission scheme in a simple CR-MIMO systems consisting of one primary user, one secondary user and a common destination. They assumed that the primary channel is completely known at both the primary and the secondary networks (two Txs and one common Rx). In [16], IA is exploited to eliminate and zero-force interference at each secondary user (SU) and primary user (PU), so that interference-free transmission can be achieved by the legitimate cognitive radio network, and the eavesdropping towards PU can be disrupted effectively by the signal from SUs. However, the requirement of coordination among multiple Txs [13–16] is viable for the downlink scenario, but usually impractical for the uplink due to its high overhead.

With IC, the previously-decoded information is subtracted from the mixed signal, and then fewer components are left in the signal for further processing. One typical application of IC is SIC [3,17–19], with which interference from already-detected components is subtracted from the received signal vector, yielding a modified received vector in which, effectively, fewer interferers are present. This is somewhat analogous to decision feedback equalization. SIC is a multi-user detection technique.
that uses the structure of an interference signal to decode multiple concurrent transmissions. It can be adopted for uplink [17] and downlink [18]. Another way of applying IC is to combine it with IA, called IAC [20–22]. It can be applied to the scenarios where neither IA nor IC alone could be used and is expected to provide more DOFs than IA. In [21], IAC was proposed as an effective technique to overcome the limitation of the number of antennas for improving the throughput of MIMO wireless LANs. In [22], an IAC-based uplink transmission scheme was proposed for infrastructured CR-MIMO systems. By exploiting the Rx side collaboration, one Rx decodes its data and feeds to the other. Then, the latter cancels this interference from its received signals. With the proposed strategy, more spatial communication opportunities could be provided for secondary users than traditional IA based schemes.

Interference can be not only aligned but also canceled or partially canceled through multiple paths, which are referred to as interference neutralization (IN) [23–26]. IA is a new IM mechanism found from and inherent in interference networks with relays [23,24]. IN strives to properly combine signals arriving from various paths in such a way that the interfering signals are canceled while preserving the desired signals [25]. It can be regarded as a distributed zero-forcing of interference before the interfering signal arrives at the undesired destination [26]. With IN, if there are more than one propagation paths from a source to its interfering destination – which is common in relay systems – multiple copies of an interference signal arriving at each user can add up to zero. In other words, the interference can be eliminated in the air, i.e., neutralized [24]. The authors of [25] constructed a linear distributed IN scheme that encodes in both space and time for separating multiuser uplink-downlink two-way communications. In [26], an aligned IN was proposed in a multi-hop interference network formed by concatenation of two two-user interference channels. The strategy provides a way to align interference terms over each hop in a manner that allows them to be canceled over the air at the last hop.

As for adaptive filtering, at least two reception modes can be adopted. One is the matched filter (MF) which is designed to align with the intended signal so as to maximize the received signal-to-noise ratio (SNR). However, due to the randomness of the desired signal and the interference, they are always not orthogonal to each other, and hence the interference cannot be eliminated by using MF. The other well-known filter type is zero-forcing (ZF), in which the filter is derived by projecting an initial vector onto the orthogonal subspace determined by the interference’s signature to nullify the interference, but at the expense of effective power loss w.r.t. the desired signal.

3. System model and assumptions

As depicted in Fig. 2, the system under consideration consists of K Txs, each with $N_T > 1$ antennas, and one Rx with $N_R > 1$ antennas. Txs have the same transmit power $P_T$ and are mutually independent, i.e., no Tx-side coordination. All Txs transmit to Rx simultaneously, and the system is perfectly synchronized, i.e., signals from Txs are assumed to arrive at the destination at the same time (symbol-synchrony) [27,28].

This assumption was also used in some existing IN studies [23,26]. The channel matrix between an arbitrary Tx, say Tx_l, and Rx is denoted by $h_l \in \mathbb{C}^{N_R \times N_T}$, $l \in \{0, \ldots, K - 1\}$, whose elements are modeled as independent and identically distributed zero-mean unit-variance complex Gaussian random variables. Channel reciprocity is assumed to hold. Channels are characterized by block fading, i.e., channel parameters in a block consisting of several successive transmission cycles remain unchanged and vary randomly between blocks. Rx can accurately obtain all channel state information (CSI) via Txs’ feedback.

The communication scenario shown in Fig. 2 is typical in an uplink (multiple-access) channel in a centralized network where an access point, e.g., a base station in cellular systems, needs to decode signals from multiple subscribers. Another example can be found in cooperative communications where a relay node acts as the common Rx. On one hand, the relay receives the information from its client(s) which will be forwarded to the destination(s). On the other hand, the relay node may have its own data transmission demand, and thus receives a signal from its intended Tx.

For an arbitrary $T_{x_j}$, the bit stream is first mapped into transmit symbols based on a predefined constellation map. Let $S_k = \{s_1, \ldots, s_{L_k}\}$ denote the symbol set adopted by $T_{x_k}$. For simplicity, we assume all the Txs employ an identical modulation scheme, i.e., $S_k = S$, $k \in \{0, \ldots, K - 1\}$. The size of $S$ is $|S| = L$, where $|\cdot|$ represents the cardinality of a finite set. Then, the modulated symbol $x_k \in S$ is mapped onto and sent by $N_T$ antennas of $T_{x_k}$. If $N_T > 1$, precoding may be applied, i.e., $x_k$ is multiplied by a selected weight vector before it is sent out. So, the direction of a transmitted signal can be controlled to enhance communication performance.

Since our focus is on the design of a reception mechanism, for simplicity we assume each transmitter is equipped with a single antenna in the following discussion. This is in some sense a worst-case assumption, since there is no coordination among the transmitting antennas. However, the proposed scheme can be easily extended to the case of $N_T > 1$. Under such condition, one or more directional beams corresponding to beamforming (BF) and spatial multiplexing (SM), respectively, can be sent by a Tx. The transmit power of each Tx, $P_T$, is equally allocated to the beams. If BF is employed at $T_{x_k}$, we have $E(\|x_k\|^2) = P_T$, else $E(\|x_k\|^2) = P_f$, where $x_k$ denotes the transmitted symbol vector. However, in this paper we assume only one data stream is sent by a Tx irrespective of $N_T$, which is in accordance with our main concern.

According to the system model, the common Rx needs to decode all the signal components sent from $K$ Txs. Due to the randomness of spatial channel $h_k$, the received signal at Rx consists of $K$ components, each of which is characterized by a unique spatial feature. If $N_R$ is large enough, i.e., no less than the total number of independent data streams sent by all users, Rx can decode all the signal components by appropriately designing a receive filter. As shown in Fig. 2, we take $x_k$’s recovery as an example, and treat the other $K - 1$ signals carrying $x_1, \ldots, x_{K-1}$, as interferences.

4. Design of a MAC receiver with fewer receiving antennas

By exploiting the constructive/destructive interactions among interfering signals, we can reduce the required number of receiving antennas significantly. We first propose ICom-based signal processing for decoding multiple interfering signals, and then present a MAC Rx structure based on ICom. For the clariety of exposition, these are discussed based on some simplified parameter settings which are followed by a generalized design. Next, the associated computational complexity is analyzed and finally, a brief comparison of the proposed scheme and classical BLAST (Bell Labs Layered Space-Time) architecture is given.
4.1. ICom-based signal processing

We first describe some basic signal processing algorithms that could be employed in MAC Rx design, including matched filter (MF) and zero-forcing (ZF),\(^1\) and then propose the ICom-based ZF. Let us consider \(x_i\)'s decoding in Fig. 2 as an example. Then, the received signal at Rx can be expressed as:

\[
y = h_x x_0 + \sum_{k=1}^{K-1} h_k x_k + n
\]

where the first term on the right-hand side (RHS) of Eq. (1) is the intended signal, the second term represents for the sum of interferences from the other \((K-1)\) Tsxs. \(n\) is an additive white Gaussian noise (AWGN) vector, elements of which have mean 0 and variance \(\sigma_n^2\). In order to decode \(x_0\), Rx applies the filter vector \(w_0\) to obtain the estimated signal:

\[
yhat = w_0^T h_x x_0 + w_0^T \sum_{k=1}^{K-1} h_k x_k + w_0^T n
\]

\(w_0\) can be designed as \(w_0 = t_0^T = h_x / \| h_x \| \) — called the matched filter (MF). So, the receiving power of the desired signal can be maximized, but the interference term remains. The other way to design \(w_0\) is to project \(h_x\) — the spatial feature of the desired signal — onto the orthogonal space spanned by the orthonormal desired basis \(h_x\), i.e., the spatial signature of signals from the other Tsxs. Then, the interference could be nullified at Rx. This type of filter, known as zero-forcing (ZF), is characterized by the orthogonal subspace projection, hence represented as \(t_0^T\). With \(t_0^T\), effective power loss related to the desired signal results since \(\| t_0^T h_x \| = \| h_x \| \) cannot be guaranteed. In order to obtain \(t_0^T\), we first apply Gram–Schmidt [29] to \(h_1, \ldots, h_{K-1}\) to obtain a set of orthonormal basis \(h_1, \ldots, h_{K-1}\). Then, \(t_0^T\) is computed as:

\[
t_0^T = h_0 - \sum_{k=1}^{K-1} h_0^T h_k h_k / \| h_0 - \sum_{k=1}^{K-1} h_0^T h_k h_k \|
\]

Note that \(N_R \geq K\) should be satisfied so as to calculate \(t_0^T\). However, in practice, the number of Tsxs could be greater than \(N_R\). In such a case, without users’ scheduling, the solution of \(t_0^T\) will not be available due to insufficient DoFs of the Rx.

By exploiting the fact that multiple interfering signals from \(K\) single-antenna Tsxs interact, i.e., construct or destruct with each other upon their arrival at Rx, we can design a filter as follows. First, the \(K-1\) interferences are aggregated to produce an effective interference, whose spatial signature is defined by \(h_i = \sum_{k=1}^{K-1} h_{k+1} x_k / \| \sum_{k=1}^{K-1} h_{k+1} x_k \|\), i.e., \(h_i\) is determined by both the channel status and the transmitted symbols w.r.t. the \(K-1\) interferers. Then, the ICom-based filter, denoted by \(g_0^T\), can be readily obtained by projecting \(h_i\) onto the perpendicular direction w.r.t. \(h_0\), and then applying normalization to the result as:

\[
g_0^T = h_0 - h_0^T h_i h_i / \| h_0 - h_0^T h_i h_i \|
\]

\(\)\(^1\) One may adopt other types of detectors such as minimum mean square error (MMSE) instead of MF and ZF in our work. In such a case, our approach can be directly applied.

In Fig. 3, the acquisition of both \(t_0^T\) and \(g_0^T\) is depicted using a three-dimensional graph. For simplicity, we set \(K = 3\), i.e., two interferences. The plane \(\beta\) is determined exclusively by the orthonormal basis consisting of \(h_1\) and \(h_2\), derived from \(h_1\) and \(h_2\). \(N_R\) should be greater than or equal to \(K\) for the existence of the solution of \(t_0^T\). For the clarity of presentation, we first take \(S = \{s_1, s_2\}\) (i.e., BPSK (Binary phase-shift keying)) as an example. If \(s_1 = s_2 = S\) during the current symbol transmission period, we have \(h_1 = (h_1 x_1 + h_2 x_2) / \| h_1 x_1 + h_2 x_2 \|\) = \((h_1 + h_2) / \| h_1 + h_2 \|\). Otherwise, \(s_1\) and \(s_2\) are different, i.e., under the above assumptions we have \(s_1 = - s_2\). Then, \(h_i = (h_1 - h_2) / \| h_1 - h_2 \|\) is obtained. We take \(s_1 = s_2\) as an example (the other situation is similar). As shown in Fig. 3, by normalizing \(h_1 - \sum_{k=1}^{K-1} h_0^T h_k h_k\), \(h_1\) and \(h_2\) are achieved, respectively. That is, an effective interference can be obtained by aggregating the interfering signals. Then, the filter only needs to be orthogonal to the signature of such an interference. In other words, the interference dimension can be significantly reduced, especially when there are many interferers.

Based on the above discussion, one desired signal (e.g., \(x_0\) among \(K\) transmitted symbols) can be decoded with either traditional ZF \(t_0^T\) or ICom-ZF-based reception \(g_0^T\). With the classical ZF, \(N_R \geq K\) should be satisfied to calculate \(t_0^T\). With ICom-ZF, since the dimension of interference is reduced to 1 irrespective of the number of interferers, two Rx-antennas are enough to decode \(x_0\). Both ZF and ICom-ZF are characterized by interference nulling and can be used in decoding multiple interfering signals. In order to recover \(K\) signals, each signal component is regarded, in turn, as the desired, and the remainders as interferences. By repeating the above process, a set of \(t_0^T\) and \(g_0^T\) can be obtained.

Note that in the above analysis, we assume \(N_F = 1\). When \(N_F > 1\) and preceding is employed, \(h_i\) should be replaced by the signal’s actual spatial feature, i.e., preceding information incorporated with \(h_i\) is used. By comparing the above two ZF schemes, we can see that with classical ZF, \(N_R \geq K\) should be satisfied to calculate \(t_0^T\). Otherwise, DoFs at the Rx is not sufficient to find a solution filter vector being orthogonal to the subspace spanned by the orthonormal basis derived from the spatial features of \(K-1\) interferers. With ICom-ZF, since the dimension of interference is reduced to one regardless of the number of interferers, two receiving antennas are enough to decode \(x_0\).

So far, the advantages of ICom-based reception have been shown under some specific parameter settings, i.e., \(K = 3\), \(|S| = L = 2\), and \(x_1 = x_2 \in S\). However, there are two important remaining issues:

(1) With the ICom-based filter, how to design the Rx structure to distinguish different combinations of transmitted symbols and decode all the other signals, i.e., \(x_1, x_2, k \neq 0\)?
(2) For general \( L \) and \( K \), what is the relationship of \( L \), \( K \) and \( N_p \) with the ICom-based mechanism and how to achieve a flexible tradeoff between the above parameters and computational complexity?

These two issues will be addressed in the following subsections.

4.2. ICom-based receiver structure

In this subsection, we first design the ICom-ZF based Rx structure and then use SIC to reduce computational complexity. We omit the superscript \( O \) in the following discussion for clarity of exposition.

We begin with \( L = 2 \) and assume that the elements of \( S \) are symmetric about the origin. Since the sum of \( h_1 x_1 \) determines the spatial feature of an effective interference, with \( L = 2 \), there will be 4 symbol combination types, i.e., effective interferences. However, due to the symmetric feature of the elements of \( S \), only 2 cases -- i.e., identical and different phases -- need to be considered. Taking \( x_1 \)'s decoding as an example, with \( K = 3 \) and \( N_p = 2 \), Fig. 4 shows the structure of an ICom-ZF receiving branch. The subscript \( i \) and \( i' \) indicate identical and different transmitted symbols, respectively.

As Fig. 4 shows, the mixed input signal \( y \) is split into two parts, each of them goes through a chain, the receive filters of the two chains, denoted by \( g_{0}^{ll} \) and \( g_{0}^{ll'} \) are calculated by substituting \( h_i = (h_{1} + h_{2}) || h_{1} - h_{2} || \) and \( h_{i} = (h_{1} - h_{2}) || h_{1} - h_{2} || \) into Eq. (4), respectively. If the actual symbol combination is in accordance with the first chain where \( x_1 = x_2 \) is assumed, we have \( g_{0}^{ll} (h_{1} + h_{2}) = 0 \) and the interference is thus mitigated. The estimated signal \( \hat{y}_{i} = g_{0}^{ll} h_{0} x_{0} + g_{0}^{ll} n \). Then, the achievable signal-to-noise ratio (SNR) is computed as:

\[
\hat{y}_{i} = \frac{g_{0}^{ll} h_{0} x_{0} + g_{0}^{ll} n}{\sigma_{n}^{2}}.
\]

However, in the second chain \( g_{0}^{ll'} (h_{1} + h_{2}) \neq 0 \). The estimation of \( y \) is:

\[
\hat{y}_{i} = g_{0}^{ll'} h_{0} x_{0} + g_{0}^{ll'} (h_{1} x_{1} + h_{2} x_{2}) + g_{0}^{ll'} n
\]

(6)

where \( x_1 \) can always be replaced by \( x_2 \) since they are identical. The output signal-to-interference-plus-noise ratio (SINR) of the second chain is then:

\[
\hat{y}_{i} = \frac{g_{0}^{ll'} h_{0} x_{0} + g_{0}^{ll'} n}{\sigma_{n}^{2} + 4 g_{0}^{ll'} h_{0} x_{0} + g_{0}^{ll'} n}.
\]

One can see that the interference part still remains in Eq. (7). Both \( \hat{y}_{i} \) and \( \hat{y}_{i} \) are sent to the decision-and-reconstruction module, denoted by \( D_R \) whose output is an \( N_p \times 1 \) signal vector, which is then subtracted from the original received signal \( y \). In the DR modular, the desired data \( x_{i} \) is firstly estimated, denoted by \( \hat{x}_{i} \) and \( \hat{x}_{i} \) respectively, then incorporated with channel information \( h_{i} \), the desired signal is reconstructed. If a chain can perfectly eliminate the interference, its DR module outputs only the desired signal, and thus there is interference plus noise from the first \( D_R \). Therefore, by using the filtering vector \( g_{0}^{ll} \) or \( g_{0}^{ll'} \) again, the interference is mitigated once again and only noise is left. Here, in order to distinguish the filter used twice in each chain, we call the filter before DR the first filter; while the one prior to the comparison modular the second filter. They are identical in a chain. As for a chain that cannot eliminate the interference, there will be signal component(s) from the second filter. Therefore, by comparing the outputs of the two chains, the one with signal will be discarded. The selection result is fed to a selection logic so that the symbol \( x_{i} \) is estimated.

It should be noted that in most cases, by comparing the power of the signals outputted from the first \( D_R \), the decision can be made about which chain produces the right estimate. However, due to the randomness of channel status and transmitted symbols, as well as possible errors incurred by certain modes, some special situations may occur such that direct power comparison of the outputted signals from the first \( D_R \) becomes inapplicable. Therefore, we employ CB and second filter before the comparison module in the design of the ICom-ZF receiving branch. In what follows, we will discuss the special cases and the necessity of CB module and the second filter.

- If a DR module reconstructs a signal which is equal to the sum of all signal components in \( y \), i.e., \( h_{1} x_{1} + h_{2} x_{2} \) in Fig. 4, the first \( D_R \) in the chain will output only noise. Therefore, the output of this chain will be incorrectly selected as the estimated symbol. In such a case, we employ a comparison-and-biasing module, denoted by CB. If the input of CB is below a preset threshold, \( T_{h} \), the input signal is regarded as noise, then a non-zero biasing signal vector \( \varepsilon \) is added before feeding to the second filter so that the output of this chain is discarded. By adopting CB, the influence of DR’s outputting \( y \) in the chain adopting wrong filter can be effectively avoided. However, one should note that if the above situation happens in the chain employing right filter due to deep channel fading or strong noise, the desired data cannot be recovered neither with our method nor other known receivers.

- If a chain employing a wrong filter decodes the right symbol accidentally, and as a consequence, reconstructs the desired signal, the input of the second filter becomes interference plus noise. However, since the filter in this chain cannot mitigate the interference, the second filter will output a non-zero interference plus noise. Then, by adopting the comparison module, the correct chain outputting only noise can be found. That is, such a particular situation does not affect the correctness of our scheme.

- If the output of a DR module happens to be the interference, i.e., \( h_{1} x_{1} + h_{2} x_{2} \), the first \( D_R \) will output the desired signal. In such a case, the input of the last comparison circuit contains signal component(s), so that this chain will be treated as incorrect. That is, the structure given in Fig. 4 is robust to the above exceptional situation.

- If the input of the second filter contains signal component(s), i.e., \( h_{i} x_{i} \neq 0 \), which is coincidentally orthogonal to the filter vector, i.e., \( g_{0}^{ll'} h_{i} x_{i} \neq 0 \), the output of this chain will be mistakenly regarded as correct. For generality, we omit the superscript/subscript \( i \) and \( i' \) here. The above situation occurs only when the spatial feature of desired signal is a linear combination of those of multiple interferences, i.e., \( h_{i} = (x_{0} - x_{j}) || h_{1} x_{1} + h_{2} x_{2} \) where \( x_{i} \neq x_{0} \). Under such a condition, even the chain employing correct filter cannot obtain the desired data. One solution to this problem is to rearrange the decoding order of multiple concurrent signals. However, the independence of various channels can always be guaranteed in practice. Hence, it is reasonable to treat the above situation as a fairly low-probability scenario, and hence is negligible.

If \( x_{1} = -x_{2}, g_{0}^{ll'} (h_{1} - h_{2}) = 0 \), whereas \( g_{0}^{ll'} (h_{1} - h_{2}) \neq 0 \). The output is \( \hat{y}_{i} = g_{0}^{ll'} h_{0} x_{0} + g_{0}^{ll'} n \). Similarly to the derivation of Eq. (6), \( \hat{y}_{i} \) is given by:

\[
\hat{y}_{i} = g_{0}^{ll'} h_{0} x_{0} - 2g_{0}^{ll'} h_{1} x_{1} + g_{0}^{ll'} n.
\]

As for the decoding of \( x_{1} \) and \( x_{2} \), the same structure illustrated in Fig. 4 can be directly applied. What we need to do are replacing the desired data \( x_{0} \) by \( x_{1} \) and \( x_{2} \), respectively, redesigning the filter as described in the last subsection, and implementing the ICom-ZF branch repeatedly. Although the above design is with \( K = 3 \), by duplicating the reusable processing module, it can be easily extended to a general \( K \).

Given \( L = 2 \), general \( N_p \) and \( K > N_p \), since the number of decodable signals using the above ICom-ZF receiving branch is \( N_p K - 1 \), the number of required chains becomes \( 2^{K-(N_p K-1)} = 2^{K-N_p} \). Here, we exploit the symmetric property of the elements of \( S \) so that the number of symbol combinations can be reduced by a factor of \( 1/2 \). With ICom-ZF, the filter vector in the \( i \)-th chain of a branch for decoding \( x_{j} \), denoted by \( g_{0}^{ll'} \), can be obtained by projecting \( h_{i} \) onto the orthogonal subspace spanned by an orthonormal basis derived
from \([h_0 \ldots h_{k-1} h_{k+1} \ldots h_{N_R-2}] \Sigma_{k'=N_R-1}^{N_R-1} h_{k'} x_{k'}\), or equivalently computing the inverse of \(H = [h_0 \ldots h_{N_R-2}] \Sigma_{k'=N_R-1}^{N_R-1} h_{k'} x_{k'}\) and taking the transpose of the 4th row vector, and then applying normalization to the result. The subscript 4 could be either 4 or \(d_x\) \((i = i)\) or \(2N_R - 1\) different symbol combinations, respectively.

To make further generalization, when \(L > 2\), a new dimension related to modulation is introduced in addition to \(K\) and \(N_R\) which will cooperatively affect the computational complexity (the number of chains in a branch). In general, the increase of \(L\) and \(K\) will incur more processing load, whereas increasing \(N_R\) will reduce computational complexity at the expense of larger Rx size and more hardware cost. Given \(K\) and \(N_R\), there will be at most \(K - (N_R - 1)\) un-decodable signal components in the above ICom-based decoding branch. Suppose the elements in \(S\) are symmetric about the origin, e.g., QAM (Quadrature amplitude modulation), then the number of chains required in a branch is \(\frac{1}{2} K^{N_R - 1}\).

Fig. 5 illustrates the structure of ICom-ZF receiving branch for general \(K\), \(N_R\) and \(L\). \(\pi = \frac{1}{2} K^{N_R - 1} - 1\) is used to denote the number of chains corresponding to different symbol combinations. There are \(M\) chains in this generalized structure. Matrix \(G_k\) consists of \(N_R - 1\) ZF-based filter vectors \(G_{kA}\), where \(A\) represents an array and \(k \in A\).

With large \(K\) and \(L\) but a small \(N_R\), repeating the above module would impose a huge computational load on the Rx. We employ SIC to reduce this computational load. With SIC, the interference from already-detected components of a transmitted symbol vector is subtracted from the received signal vector, yielding a modified received vector in which fewer interferers are present. So, we need not repeat the module given in Figs. 4 or 5 multiple times in order to decode all the signal components. Instead, we adopt the structure shown in Fig. 6. For simplicity, we use the parameter settings \(K = 3\), \(N_R = 2\) and \(L = 2\) to illustrate the application of SIC. This can be easily extended to more generalized cases. From the figure, by employing the decision feedback information \(x_{k_0}\), the signal component carrying \(x_{k_0}\) can be subtracted from the received signal \(y\). We assume \(x_{k_0} = x_{k_0}\), i.e., decoding is error-free. Note that after interference cancellation, the number of signals is reduced to 2, which is not greater than \(N_R\) any longer. Then, a traditional ZF Rx employing \(W_{ZF} = (H^H H)^{-1} H^H\) where \(H = [h_0 h_1]\) can be directly applied to the residual mixed signal \(y'\) to recover \(x_1\) and \(x_2\).

Compared to the Rx structure which duplicates the modular given in Fig. 4 for three times, the above ICom/SIC-ZF-based Rx can significantly reduce the computational complexity. It should be noted that due to processing and feedback of \(x_{k_0}\), an additional delay will be introduced in decoding \(x_1\) and \(x_2\). Moreover, the decoding accuracy and order will affect the performance as well. Since these problems have been extensively studied before [30–33] and our focus is on exploiting the interaction among interfering signals in the Rx design, some simplifications are made such as neither error propagation nor decision feedback delay is considered, and the decoding order is arbitrarily chosen.

Table 1 illustrates the tradeoff between the required \(N_R\) and the processing complexity for the case of \(L = 2\). For simplicity, we use a virtual chain (VC) to represent the computational complexity. An \(\alpha\)-type VC, denoted by \(V C_{\alpha}\), is defined as the structure shown in Fig. 4, whereas one \(\beta\)-type VC (\(V C_{\beta}\)) is equivalent to a traditional ZF. The table shows that with SIC, the complexity of the Rx structure is reduced significantly, compared to the one without SIC. Thus, in practice we can make a flexible tradeoff between \(N_R\) and processing complexity.

Provided \(N_T = 1\), the number of TxS, \(K\), is equal to the number of signals to Rx. When \(N_T > 1\), multiple data streams may be sent by one Tx, and the title of the first column in Table 1 should be the number of transmitted signals. Since antennas of one Tx can cooperate to generate proper signal patterns, \(N_T = 1\) is in some sense the worst-case situation, i.e., no coordination across all transmit antennas. When \(K = N_R\), for both schemes, without or with SIC, the classical ZF reception can be directly applied. When \(K > N_R\), the DoFs at Rx are insufficient for decoding all the signal components at once, and hence the proposed ICom-based reception scheme can be employed. Note also that the complexities of \(V C_{\alpha}\) and \(V C_{\beta}\) depend on \(N_R\) which determines the dimension of the reception filter. Moreover, \(K - N_R\) affects the number of required \(V C_{\beta}\) in an ICom-ZF receiving branch.

The following discussion about Table 1 is under the condition of \(K > N_R\). When \(K < N_R\), for both schemes, a \(V C_{\beta}\) can be used directly to decode the signals in one stage.

1. As for the reception mechanism without SIC, with one receiving branch, the number of decodable signals is \(N_R - 1\), so there will be \(2^{K - (N_R - 1)}\) combinations of the other \(K - (N_R - 1)\) un-decodable signal components. By exploiting the symmetric property of the elements in \(S\), the number of combinations needs to be considered becomes \(2^{K - N_R}\). Let \(\mu\) be the number of branches employed for decoding signals without SIC. Given \(K > N_R\), by taking into account the situation in which \(K\) is not the integer multiples of \(N_R - 1\), the variable \(\mu\) can be selected from the set \([1, 2, \ldots, \left\lfloor \frac{K}{N_R - 1} \right\rfloor]\), where \(\left\lfloor x \right\rfloor\) denotes rounding of the element to the nearest integer that is greater than or equal to it. Then, the number of decodable signals with \(\mu\) branches is \(\min \left\{ \mu(N_R - 1), K \right\}\).
2. With SIC, the output(s) of each branch is(are) fed to the next stage for interference cancelation so as to reduce the dimension of the mixed signal. As a result, the number of VC_{C_{p}} in a set of sequentially used iCOM-based branches decreases one after the other to simplify the Rx structure. If the number of interferring signal components is larger than N_{r}, we employ one iCOM-ZF receiving branch to recover N_{r}-1 signals. Then, the outputs are fed to the next processing stage. This process is repeated until the number of remaining signals is less than or equal to N_{r}. At this point, a traditional ZF reception module, i.e., one VC_{p} is directly applied. We define the number of iCOM-ZF branches employed for decoding signals incorporating with SIC as η. When the inequality K−η(N_{r}-1) > N_{r} holds, VCs are used. The number of required VCs is computed as $\sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}} \left(1 - \frac{K-\eta}{N_{r}-\eta} \right)$, with which η(N_{r}-1) signal components can be decoded. When K−η(N_{r}-1) ≤ N_{r}, one VC_{p} is employed to recover the remaining signals.

The general expression for the number of required VCs with SIC is:

\[ \sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}} \left(1 - \frac{K-\eta}{N_{r}-\eta} \right) \times VCs_{\eta} \text{ where } \eta \in \{1, 2, \ldots, \left\lfloor \frac{K-N_{r}}{N_{r}-1} \right\rfloor + 1\}, \]

\[ \psi_{\eta} = \begin{cases} 1, & \text{if } \eta \leq \left\lfloor \frac{K-N_{r}}{N_{r}-1} \right\rfloor, \\ 0, & \text{if } \eta > \left\lfloor \frac{K-N_{r}}{N_{r}-1} \right\rfloor, \end{cases} \]

total of min(η(N_{r}-1)+\psi_{N_{r}},K) signal components can be decoded.

When L is general (other than 2), the base number (=2) in Table 1 should be replaced by L. In practice, Txs may employ different modulation schemes adaptively. Some feedback mechanisms are then required so that Rx can replace the transmission parameters of Txs, based on which the Rx structure is dynamically configured, following the concept of Software-Defined Radio (SDR) [34,35].

It should be noted that the key feature of the proposed mechanism is the utilization of interactions among interfering signals, i.e., we focus on limiting the effect of interference rather than managing individual interferences. Since such interactions are common in wireless communications, their exploitation does not incur additional cost except for an increase of computational complexity at the common receiver which will be analyzed in the following subsection.

Moreover, from Table 1 one may note that the complexity of the proposed structure rapidly grows with an increase of K. To reduce such complexity, we can equip more antennas with the common receiver so as to improve its processing capability, or employ multi-user scheduling to select a set of transmitters to transmit to the receiver so that the complexity is properly controlled. In addition, we employ the proposed receiver structure to determine the combination of symbols carried in the interfering signals, so that the un-decodable components can be mitigated in a processing stage. However, we do not exploit such symbol-combination information for decoding in the following processing stage(s). By utilizing such information, the complexity of the receiver structure can be reduced further. Since it is outside the scope of this paper, we leave it as our future work.

4.3. Analysis of computational complexity

The complexity is quantified in number of real floating point operations (FLOPs) [29]. A real addition, multiplication, or division is counted as one FLOP. A complex addition and multiplication have 2 FLOPs and 6 FLOPs, respectively.

We need \(\sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}}\) stages in total to decode K signals with N_{r} antennas. In each chain of the first \(\sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}}\) stages, the computation includes the calculation of filter matrix G_{\eta} and multiplication of G_{\eta} with the received mixed signal. In order to obtain G_{\eta}, we first use Gaussian-Jordan elimination to compute the inverse of an N_{r} x N_{r} matrix H. This operation takes \(N_{r}^{3}\) complex multiplications/divisions and \(N_{r}^{2}\) complex additions. Then, the first N_{r}-1 row vectors of H are normalized to compose H_{\eta}, Frobenius norm of a 1 x N_{r} vector takes \(4 N_{r}\) FLOPs, and the normalization takes \(N_{r}\) complex divisions, so the FLOP count for normalizing an (N_{r}-1) x N_{r} matrix is \(10 N_{r}^{2}\). Next, the mixed N_{r} x 1 signal is left multiplied by H_{\eta}, which takes \(N_{r}^{2}\) complex multiplications and \(N_{r}-1\) complex additions. Based on the above analysis, the complexity of an arbitrary chain in the first \(\sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}}\) stages is \(8 N_{r}^{3}+14 N_{r}^{2}-18 N_{r}+2\). In each stage \(m \leq \left\lfloor \frac{K-N_{r}}{N_{r}-1} \right\rfloor\), the output $\xi = \frac{1}{2} (H^{H} + H) x$ has \(N_{r}-1\) real additions and $\xi - 1$ real subtractions. As a result, the complexity of the mth stage is \(8 N_{r}^{3}+14 N_{r}^{2}-17 N_{r}+1\).

In the last stage, the number of signal components, say $\psi$, is no greater than N_{r}, and thus traditional ZF can be directly applied and no SNR/SINR comparison is needed. When $\eta = N_{r}$, the operations in the final stage include the inversion of an N_{r} x N_{r} signal feature matrix, normalization of the inverse matrix, and multiplication of the N_{r} x N_{r} filter matrix with an N_{r} x 1 mixed signal, which takes \(8 N_{r}^{3}-4 N_{r}^{2}+2 N_{r}\) FLOPs, \(10 N_{r}^{2}\) and \(8 N_{r}-2 N_{r}\) FLOPs, respectively. The total FLOP count is \(8 N_{r}^{3}+14 N_{r}^{2}\). When $\eta < N_{r}$, the pseudo-inverse of an N_{r} x N_{r} matrix, denoted by $H^{-}\text{null}$ is required. According to the pseudo-inverse calculation $H^{-} = (H^{H} H^{-1} H)^{-1}$, two matrix multiplications and one matrix inversion are needed, which takes \(16\psi^{2} N_{r}^{2} + 8\psi^{2} - 6\psi^{2} - 2\psi N_{r} + 2\psi^{2}\) FLOPs. Additionally, by taking into account the normalization of filter matrix and multiplication of W_{ZF} with mixed signal, of which the FLOP counts are \(16\psi^{2} N_{r} + 8\psi N_{r} - 2\psi\) respectively, the computational complexity of the last stage under $\psi < N_{r}$ is \(16\psi^{2} N_{r}^{2} + 8\psi^{2} - 6\psi^{2} + 16\psi N_{r} - 12\psi^{2}\).

Based on the above discussion, the total computational complexity of the proposed scheme in FLOPs can be approximately computed by $N_{r}^{2} L K\left(\sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}}\right)$, $L\left(\sum_{\eta=1}^{\min\{K-N_{r},(N_{r}-1)\}}\right)$, L(\psi N_{r}^{2} + 2\psi N_{r} - 2\psi) FLOPs. For example, given K = 10 transmitters each employing 4-QAM (i.e., L = 4) to send a single data stream to the common receiver equipped with N_{r} = 4 antennas, the receiver needs approximately 1.06x x 10^{6} FLOPs so as to decode all the received signals. If the receiver can process at the speed of 1GFLOPs per second, the processing delay is only about 1.06 ms.
Table 2
Comparison of different schemes.

<table>
<thead>
<tr>
<th>Feature</th>
<th>IA</th>
<th>IN</th>
<th>MF</th>
<th>ZF</th>
<th>ICom-based</th>
</tr>
</thead>
<tbody>
<tr>
<td>Location of implementation</td>
<td>Tx</td>
<td>Tx</td>
<td>Rx</td>
<td>Rx</td>
<td>Rx</td>
</tr>
<tr>
<td>Rx-side full CSI</td>
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<td></td>
<td>X</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
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<td></td>
<td></td>
<td>X</td>
<td>X</td>
<td></td>
</tr>
<tr>
<td>Rx-side multi-antenna</td>
<td></td>
<td></td>
<td>X</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Symbol-synchrony</td>
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<td>X</td>
<td>X</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Signal attenuation</td>
<td>X</td>
<td>X</td>
<td>X</td>
<td>X</td>
<td>X</td>
</tr>
</tbody>
</table>

4.4. Comparison with other schemes

So far, we can utilize the ICom-based processing to implement a MAC Rx. One may think that the multi-stage structure depicted in Fig. 6 is similar to the BLAST architecture [30,31]. There have been numerous other schemes than our proposed method for handling interferences among concurrent communications, such as IA, IN, ZF, etc. In what follows, we will make a qualitative comparison of our method with the other popular ones. All of them can facilitate the reception of desired signal(s). The results in Table 2 are obtained with K Tx and 1 common Rx. Each Tx sends a single data stream. We consider the reception of one desired signal out of K as an example.

The symbols \(\circ\) and \(\times\) represent having and not having the corresponding feature, respectively. Tx/Rx-side full CSI indicates that the Tx/Rx needs to acquire CSI from all Rxs/Txs, including intended and unintended Rxs/Txs, to itself. The table shows that both IA and IN are implemented at the Tx side whereas the other three schemes are implemented at the Rx side. As a result, multiple antennas are required either at Tx or Rx or both to provide sufficient signal processing capabilities. With IA, both \(N_T\) and \(N_R\) should be at least 2 so that on one hand, the spatial feature of transmitted signals can be adjusted to realize alignment, and on the other hand, the desired signal can be distinguished from the aligned interferences at Rx. To implement IN, a neutralizing signal is generated so as to cancel the interference at Rx, consuming some transmit power. Signals for IN should carry the same data and achieve symbol-synchrony. In the communication scenario shown in Fig. 2, at most \(N_T - 1\) and \(N_R\) signals can be decoded with IA and IN, respectively. However, neither the aligned nor neutralized signals can be recovered. With ZF, in order to acquire the filter vector orthogonal to the subspace determined by the \(K - 1\) interferers, at least \(K\) receiving antennas are needed. MF only needs the CSI of the desired signal, thus saving cost compared to the other schemes. As for the ICom-based processing, since an iterative structure is adopted, \(2\) receiving antennas are sufficient, i.e., one signal is decoded in each processing stage. In addition, both CSI and data carried in the signals are exploited to implement ICom; this is similar to IN. However, in the system model shown in Fig. 2, in order to apply IN, some Tx is required to generate the neutralizing signals instead of sending their own information. This is a non-trivial cost for the system, and a mobile user may not have any incentive to sacrifice its own benefit for others. Our proposed mechanism avoids such a cost. Since IA and ZF are realized by adjusting the transmission beam and the direction of the receive filter’s main lobe, respectively, such an adjustment will cause the loss of effective signal power. As for the ICom-based scheme, the Rx-side full CSI is required to estimate the status of the aggregated interference, whereas the other properties are determined by the reception mode we employed. For example, ICom-ZF requires multiple antennas at the Rx-side and incurs the loss of signal power, whereas ICom-MF does not.

Finally, we briefly compare the proposed multi-stage ICom/SIC-based Rx with classical BLAST architectures. Forchini [30] proposed a diagonally-layered architecture, now widely known as D-BLAST, which is also called ‘Horizontal BLAST’ (H-BLAST) [32]. There have been several variations of D-BLAST, and V-BLAST [31] is a simplified version, which tries to reduce computational complexity at the loss of transmit diversity. Our Rx structure and BLAST are similar in the following aspects. First, in each stage (layer), one or more data streams are considered as the desired while treating the remains as interferences. Second, the detected symbol(s) is(are) used to regenerate its(their) interference contribution which is then subtracted from all the other layers. Third, these operations are repeated for the lower layers, finishing at the top layer. Finally, both mechanisms assume that all symbols at previous layers have been detected correctly, and hence their interferences to the other layers can be completely mitigated. However, there are two main differences between the two. First, either D-BLAST or V-BLAST decodes one data stream in one layer, whereas ICom/SIC could recover multiple data streams in one layer as long as \(N_R > 2\). Second, both BLAST and the proposed scheme reduce the signal dimension by employing SIC, but since we treat all the interferences as one equivalent disturbance, the number of layers required is much less than that with BLAST. However, ICom/SIC incurs additional computational complexity.

5. Detection-order optimization and reception-mode adaptation

In the above discussion, we focus on the ZF-based Rx design while the decoding order of multiple concurrent transmissions is arbitrarily chosen. However, by optimizing the detection order or/and adaptive selection of reception mode, the system’s SE can be improved further. We will give a brief discussion about the first measure and elaborate on the reception mode adaptation between matched filter (MF) and ICom-ZF by exploiting the interrelationship of multiple signals’ spatial features.

5.1. Detection order optimization

V-BLAST (Vertical BLAST), a wireless communication architecture, has been implemented in real time in a laboratory without considering the error propagation [31]. However, error propagation is inevitable and should thus be minimized. This is because V-BLAST uses SIC, and hence the Rx detects the user’s data, then immediately cancels the interference to enable the next detection. As a result, the accuracy of signal detection should be improved as much as possible, so as to enhance the system’s overall performance. Based on the above observation, OSIC (Ordered Successive Interference Cancellation) can be to substitute SIC in V-BLAST [31]. OSIC employs “sequential nulling and cancellation” detection. At each time instant, instead of jointly decoding signals from all transmit antennas, OSIC first decodes the strongest signal using a nulling vector, then cancels the effect of this strongest signal from each of the received signals, and finally proceeds to decode the other “strongest” among the remaining signals, and so on. [36, 37] showed the BER (Bit Error Rate) performance of MMSE (Minimum Mean Square Error) detector with and without ordering schemes in Rayleigh and Rician Channels, proving better performance of a given detector with optimal ordering. Thus, by introducing OSIC, the performance of V-BLAST can be improved by combating error propagation. One should notice that even without accounting for error propagation, detection-order optimization still affects system performance, e.g., [31] compared performance using nulling only and combining nulling and optimally-ordered cancellation, and showed that BER is improved by optimal ordering irrespective of SNR. Based on the above discussion, by employing detection order optimization, we can also improve the performance of the proposed ICom/SIC/ZF mechanism. As for the criterion for determining the detection order, we adopt the “best first” cancellation approach [31], i.e., in our system model the signal passing through the channel with the highest gain will be detected first.
5.2. Adaptation of reception mode

We elaborate on adaptation between ICom-ZF and MF at Rx. Let us consider the reception of $x_0$ as an example. As shown in Fig. 7, spatial feature of the transmission of $x_k$ is determined by $h_0$. With MF, the filter vector $f^M$ matches $h_0$, thus no loss of the desired signal power. However, the interference is not mitigated, i.e., the projection of the signature of the interference, $h_i$, onto the direction determined by $h_0$ is non-zero. As for ICom-ZF, the filter vector $g^D$ is designed to be orthogonal to the direction of the effective interference, $h_i = \sum_{k=1}^{K-1} h_k x_k$, such that the interference can be eliminated. However, this incurs loss of a desired signal power. In practice, to improve system performance, the above two reception modes can be selected adaptively by exploiting both the relative strength of spatial correlation between mutually interfering signals.

According to the discussion in Section 4, given $N_R = 2$ and $K = 3$, with the assumption of $x_1 = x_2$, we have $g^D(h_0 + h_1) = 0$. Without ambiguity, in the following discussion we omit the subscript $i$ for simplicity. By noting that $\|g_0\| = 1$, $g^D(h_0)g^D(h_0)^H = \|h_0\|^2 \cos^2 \theta_0$, where $\cos \theta_0 = \|g_0\| = 1$, we can obtain the achievable SNR in Eq. (5) can be rewritten as:

$$\gamma_{(ICom-ZF)}^M = \frac{P_T}{\sigma^2} \|h_0\|^2 \cos^2 \theta_0.$$  \hspace{3cm} (9)

Similarly, we can define the spatial correlation between $f_0$ and $h_0$ as $\cos \phi_0 = \frac{\langle f_0, h_0 \rangle}{\|f_0\| \|h_0\|}$. Under $N_R = 2$, $\cos^2 \phi_0 + \cos^2 \theta_0 = 1$ holds.

The achievable SNR of $x_0$ with MF is computed as:

$$\gamma_{(MF)} = \frac{P_T \|f_0\|^2 \|h_0\|^2}{\sigma^2 + P_T \sum_{k=1}^{K-1} \|f_0\| \|h_k\|^2 \|f_0\| \|h_k\|^2 \cos^2 \phi_k}.$$  \hspace{3cm} (10)

Note that $\phi_k = \frac{\|h_k\|^2}{\langle f_0, h_k \rangle}$ denotes the desired signal to noise ratio (SNR) at the Rx. By defining the left hand side (LHS) and the right hand side (RHS) of the above inequality as $\xi_{(\gamma)}$ and $\zeta_{(\gamma)}$ respectively, the criterion of reception mode adaptation can be expressed as follows:

$$\begin{cases} \text{If } \xi_{(\gamma)} < \zeta_{(\gamma)}, \text{ ICom-ZF is adopted} \\
\text{If } \xi_{(\gamma)} \geq \zeta_{(\gamma)}, \text{ MF is adopted} \end{cases}$$  \hspace{3cm} (14)

where $\xi_{(\gamma)} = \frac{\|f_0\|^2}{\|h_0\|^2}$ represents the interference to the signal ratio (ISR). $\zeta_{(\gamma)}$ indicates a dividing line derived from $\gamma_{(ICom-ZF)}^M = \gamma_{(MF)}$. In practice, both $\xi_{(\gamma)}$ and $\zeta_{(\gamma)}$ can be obtained by using some locally measurable information such as SNR, ISR, correlation indices $\cos^2 \phi_k$ and $\cos^2 \theta_k$, making it easy for implementation. Although we take the reception of $x_0$ as an example, the reception of other signals is similar. We only need to replace the subscript $i$ by the index of intended signal and modify the expression of the sum of interferences.

Eq. (13) shows that provided with specific $h_0$ and $h_0$, $\zeta_{(\gamma)}$ is a function of $\cos^2 \phi_0$ and $\cos^2 \theta_0$. Since there are more than 3 independent variables, we cannot graphically show the validity of the rule in Eq. (14). However, by noting that all $h_k$s are of the same statistical characteristic, so are $\cos \phi_0$, $\cos \theta_0$, and hence we can use the expectation of $\cos^2 \phi_0$ w.r.t. $\cos^2 \theta_0$, denoted by $E(\cos^2 \phi_0)$, to approximate the accurate $\zeta_{(\gamma)}^D$ in Eq. (14), so that the number of independent variables can be reduced to 1. $E(\xi_{(\gamma)}^D)$ can be expressed as:

$$E(\xi_{(\gamma)}^D) = \frac{\sum_{k=1}^{K-1} \|h_k\|^2}{\sigma^2} = \frac{\sum_{k=1}^{K-1} \|h_k\|^2}{\sigma^2} = \frac{2}{\sigma^2}.$$  \hspace{3cm} (15)

Fig. 8 shows the adaptive selection of reception mode for $x_0$ versus $\cos^2 \theta_0$ and $\zeta_{(\gamma)}$ under $\|h_0\| = 2.3679$ and 10log$_{10}(P_T/\sigma^2) = 5$ dB. The dots are acquired by calculating $\gamma_{(ICom-ZF)}^M$ and $\gamma_{(MF)}^M$ according to Eqs. (9) and (10), and then choosing the larger one by using Eq. (14). Note that the red asterisk and dark diamond are divided by $E(\xi_{(\gamma)}^D)$. Since $E(\xi_{(\gamma)}^D)$ is only an approximation, it can be seen from Fig. 8 that when $\cos^2 \theta_0$ ranges from 0.1 to 0.6, the use of $E(\xi_{(\gamma)}^D)$ incurs a portion of wrong decisions, i.e., selecting the reception mode with inferior SNR/SNR. However, when $\cos^2 \theta_0$ is close to 0 (or 1), $\gamma_{(MF)}^M$ (or $\gamma_{(ICom-ZF)}^M$) outperforms $\gamma_{(ICom-ZF)}^M$ (or $\gamma_{(MF)}^M$) with a probability approximately equal to 1, so the use of $E(\xi_{(\gamma)}^D)$ provides considerably good accuracy.

Fig. 9 illustrates the structure of a receiving branch employing adaptation between ICom/SIC-ZF and MF for general K, $N_R$ and L. The filter vector $f^M$ ($j = 1, \ldots, N_R - 2$) in matrix $F_{0,0,\ldots,N_R-2}$ is calculated as $f^M = h_j/\|h_j\|$. The ICom-ZF $(0,0,\ldots,N_R-2)$ is then referred to the structure given in Fig. 5 with which we can get an actual combination of interferences, i.e., $h_0 = \sum_{k=1}^{K-1} h_k x_k$, and hence the correct chain can be determined.
Then, with the information of filter matrix \( G_k \) employed by the correct chain, the correlation between the spatial feature of each desired signal and its receive filter vector, i.e., \( \{\cos^2 \theta_0, \ldots, \cos^2 \theta_{N_k-1}\} \), is yielded. In addition, based on the feedback CSI from each Tx, the spatial correlation between each desired signal carrying \( x_i \) and interfering signal carrying \( x_j \), where \( i = 0, \ldots, N_k - 2 \) and \( j = 0, \ldots, N_k - 1 \), can be obtained. To reduce complexity, we do not realize reception mode adaptation w.r.t. each individual signal, but treat the set of signals to be decoded as a whole instead, i.e., based on the comparison of \( \sum_{i=0}^{N_k-2} v_i \) and \( \sum_{i=0}^{N_k-2} v_i \) for \( i \), an appropriate reception mode is selected. The adaptive branch given in Fig. 9 can be adopted as a substitution of the structure illustrated in Fig. 5 so as to realize a multi-stage Rx adapting between ICom-ZF and MF which is similar to the one presented in Fig. 6. For space limitation, we omit the details.

6. Evaluation

We use simulation with MATLAB to demonstrate the performance of the proposed reception mechanism. \( K \) Txs and one common Rx are used for the simulation. Each Tx is equipped with \( N_f \) antennas and sends a single data stream with power \( P_f \). Rx has \( N_R \) antennas. When \( N_f = 1 \), Tx sends omnidirectionally. Otherwise, Tx employs singular value decomposition (SVD) based precoding in terms of its channel status with the destination, and adopts the principle eigen-mode to transmit to Rx. Pre-coders and receive filters should be normalized for power gain fairness. The following two subsections will cover the evaluation of system spectral efficiency (SE) and delay performance.

6.1. Spectral efficiency

We simulate the proposed ICom/SIC-ZF as well as four other methods including MF, traditional ZF, IA, and IN for comparison of their SE. With the MF, a total of \( K \) filter vectors each of which matches a desired signal are generated by Rx and combined to form a filter bank, and \( K \) signals are then decoded. The SINR of the \( k \)-th (\( k \in \{0, \ldots, K - 1\} \)) desired signal can be obtained as \( S_k^M = \frac{P_f X_k^M |B_k^M|^2}{\sum_{i \neq k} P_f X_i^M |B_i^M|^2 + \sigma^2} \), where \( X_k^M \) denotes the MF for the signal from Tx\( k \) and \( B_k \) represents the signal feature that may include precoding information when Tx\( k \) is equipped with multiple antennas. When \( K \leq N_R \), all of the \( K \) signals can be recovered by employing ZF, thus making IA, IN and ICom/SIC-ZF unnecessary. That is, ICom/SIC-ZF has advantages over the classical ZF only when \( K > N_R \). Under such condition, IA, IN and ICom/SIC-ZF are implemented, respectively, as follows.

- With IA, \( K - (N_R - 1) \) of \( K \) signals are randomly selected and adjusted to align in one direction so that the rest of \( N_R - 1 \) signals may be decoded by applying ZF. The performance of IAC is referred to that of IA since it decodes the same number of signals as IA.
- IN can be employed to recover \( N_R \) signals. In the simulation, we randomly pair signals originating from the \( K - N_R \) Txs. Then, in each pair, we adjust the signal with a higher channel gain to neutralize the one with lower gain so that additional power cost can be avoided. Note that when \( K - N_R \) is an odd number, one Tx will be left unpaired — we simply turn off this Tx to eliminate its interference.
- As for ICom/SIC-ZF, \( K-(N_R-1) \) signals are first randomly selected and treated as an effective interference. ZF is then employed to decode the other \( N_R - 1 \) signals. By combining with SIC, all \( K \) signals can be recovered successively.

Besides the above-mentioned methods, we also compare our scheme with non-orthogonal multiple access (NOMA) which has been introduced recently as one of the key technologies for the next generation of wireless networks referred to as 5G [38,39]. The basic principle of NOMA is to serve multiple users by power domain or code domain multiplexing at transmitter and SIC at receiver [39]. In our work, we focus on the power domain NOMA. Two uplink NOMA schemes are studied. The first one is given in [38] where the base station (common receiver) divides the users into pairs and detects every pair using SIC. Moreover, the optimum user pairing set is obtained by employing
the Hungarian algorithm. However, the method in [38] obtains the optimum user pairing set without inter-pair interference management (IM), thus incurring poor system SE. As for the second one, the array processing capabilities brought by multi-antenna configurations are exploited to suppress or mitigate interference and hence improve the system SE. In this scheme, all TxS are first grouped with each group containing at most $N_R$ Txs, such that the system's inter-group correlation (an indicator of inter-group interference) is minimized. We define the spatial feature of the signal sent from $T_x_i$ and perceived by the common receiver as $h_i$ (note that if a precoder $p_i$ is employed by $T_x_i$ for pre-processing data $s_i$, the spatial feature $h_i$ should be replaced by $(h_i, p_i)$). The $K$ Txs are divided into $N_g$ groups denoted by $G_{1, ..., G_{N_g}}$, respectively, then the system's inter-group correlation can be calculated as $\sum_{G_{m} \in G_{1, ..., G_{N_g}}} \sum_{T_x \in G_{m}} \sum_{T_x' \in G_{m'}} \langle h_i, h_{i'} \rangle^2$ where $(h_i, h_{i'})$ represents the inner product of vectors $h_i$ and $h_{i'}$. Next, we decode signals group-by-group following the descending order of the group's sum gain, i.e., $\sum_{T_x \in G_{m}} |h_i|^2$ where $G_{m} \in \{G_{1, ..., G_{N_g}}\}$. In decoding the signals of group $G_{m}$, the inner-group interference, i.e., interference among signals of the same group, is suppressed by employing ZF reception, whereas for the inter-group interference, only that from the signals of later decoded groups should be taken into account since that incurred by the signals of previously decoded groups has been eliminated with SIC. This way, both inter-group and inner-group interference can be managed while realizing the concept of NOMA. Based on the descriptions of the above two uplink NOMA schemes, we call the first one given in [38] user-pairing NOMA without IM whereas the latter is called user-grouping NOMA with IM.

For conciseness, we will use a general form $[N_T \ N_R \ K \ K_g]$ to denote the parameters for different methods, where $K_g$ is the number of decodable signals for given $N_T$, $N_R$, and $K$. Fig. 10 shows the effects of $K$ on the system SE. In order to show the advantage of proposed ICom/SIC-ZF, the simulation is under the condition of $K > N_R$. However, in such a case a traditional ZF is not applicable. For space limitation, we do not show the effects of $N_T$ and $N_R$ on the system SE. The details can be found in [10]. Fig. 11 shows the system SE with different $K$ satisfying the minimum antenna requirement for recovering all $K$ signals.

Fig. 10 shows the system SE of different schemes with fixed $N_T$, $N_R$, and different $K$s. Note that both IA and IN require at least 2 Tx-antennas, so we set $N_T = 2$, $N_R = 3$ and $K \in \{4, 5, 6\}$. Note that with IA, IN and ICom-ZF, the decoded signals are interference-free, whereas with MF, CCI exists. When $K > N_R$, $K_g$ with IA and IN is limited by $N_R$, and hence their SE is independent of $K$. Specifically, IN can recover $N_R$ signals while IA only decodes $N_R - 1$. As a result, IN yields a higher SE than IA. On the other hand, both MF and ICom/SIC-ZF can decode $K$ signals without any constraint on $N_R$, but ICom/SIC-ZF can eliminate CCI, whereas MF cannot. As $K$ increases, $K_g$ of MF and ICom/SIC-ZF grows, improving system SE. But with MF, the aggregate interference $\sum_{k=1}^{K_g} x_k^H h_{1:k} \sum_{k=1}^{K_g} x_k^H h_{1:k}^2$ for user $k$ becomes stronger and hence degrades the system SE. From the expression of $g_k^M$, we can see that when SNR is low, noise dominates the achievable SE, and MF maximizes the effective power of intended signal, thus improving its SE as SNR gets higher. Given an extremely low SNR, IA and IN cannot achieve an obvious gain from interference management. In contrast, MF can gain from supporting multiple Txs. Thus, under a very low SNR, MF slightly outperforms IA and IN. When SNR becomes higher, the interference becomes the dominant factor in limiting SE. MF cannot eliminate CCI, so $g_k^M$ saturates at a high SNR with fixed $K$ and is inversely proportional to $K$ for a given SNR. As can be seen from Fig. 10, in a high SNR region, MF’s SE saturates and does not vary with $K$. ICom/SIC-ZF can mitigate interference, making its SE superior to the other three methods and increase with $K$.

By noting that NOMA always puts emphasis on doubling the simultaneously supported subscribers in the multi-user system, for clarity, we only simulated the NOMA schemes under $K = 2N_R$. As Fig. 10 shows, the system SE of NOMA grows with an increase of SNR. User-grouping NOMA outperforms the user-pairing NOMA due to its effective management of both inner-group and inter-group interferences. Moreover, user-grouping NOMA also excels MF and IA, but is inferior to IN and our proposed scheme. Due to the disturbance from the later decoded signal to the first decoded one within a user-pair and un-managed inter-pair interferences, SE of user-pairing NOMA saturates in a high SNR region, which is similarly to that of MF; however, since user-pairing NOMA employs SIC to partially mitigate interference within a user-pair and obtains the optimum user pairing set maximizing the system SE, its SE performance is superior to that of MF.

Fig. 11 plots system SE for different $K$s satisfying the minimum antenna requirement (MAR) for recovering all $K$ signals. We only study MF, ZF and ICom/SIC-ZF because under $K > N_R$, neither IA nor IN can decode all the $K$ signals while for $K \leq N_R$, ZF is adopted, instead of IA, IN and ICom/SIC-ZF. Since MF, ZF and ICom/SIC-ZF do not have any requirement for multiple Tx-antennas, we simply set $N_T = 1, K$ is selected from the set $\{3, 4, 5\}$. From Fig. 11 we can see that with ZF and ICom/SIC-ZF, the system SE grows with $K$. Provided with the same $K$, ZF outputs the best SE since its Rx has $K$ antennas whereas the other two methods only use 2 antennas. ICom/SIC-ZF can decode $K$ interference-free signals, so its SE performance ranks the second. MF improves its achievable SE with SNR in a low SNR region, which saturates at high SNR, and is independent of $K$. The analysis is the same as that in Fig. 10.

Fig. 12 plots the system SE of ICom/SIC-RMA (ICom/SIC with reception mode adaptation) in comparison with other mechanisms under $N_T = 1, N_R = 2$ and different $K$s. Although the MF-based scheme can decode all the data streams in one stage, for a fair comparison, we let SIC-MF decode at most $N_R - 1$ streams (the same as ICom/SIC-ZF) in one intermediate stage, and then employs interference cancellation to subtract the decoded signal components from the received signal before recovering other data streams in the next stage. Without loss of generality, the decoding order is arbitrarily determined. Given $K = 4$, ICom/SIC-ZF (or SIC-MF) needs two intermediate stages, each of the stages can decode $N_R - 1 = 1$ signal, and one ZF (or MF) module for the last 2 signal components. Similarly, provided $K = 6$, a total of five stages are required. As can be seen from the figure, with the same $K$,
SIC-MF outperforms MF in SE due to the fact that the negative impact of interference on the decoding of desired signals is reduced stage by stage by employing SIC. When SNR is low, SIC-MF yields higher SE than ICom/SIC-ZF. As SNR goes up, SE of SIC-MF saturates. The analysis can be referred to that of Fig. 10. Compared to the schemes with a fixed reception mode, SE can be enhanced by employing reception mode adaptation. As Fig. 12 shows, MF is preferred at low SNR, while as SNR increases, ZF outperforms MF.

Fig. 13 shows the system SE of various mechanisms under $N_{RF} = 1$, $K = 6$ and different $N_{RF}$. ICom/SIC-RMA is shown to yield the best SE. The analysis is the same as that given in Fig. 12. As $N_{RF}$ increases, the diversity gain at the Rx grows, hence improving the SE of all mechanisms.

Note that in the above evaluation, error-propagation is not considered, so SE of the proposed scheme increases with $K$. When the decoding is imperfect, the last decoded signal is subject to more interference than those recovered in the previous stage, so that its SE is deteriorated, rendering the saturation of system SE at a higher $K$.

6.2. Latency

We now study the tradeoff between $N_{RF}$ and $K$, as well as the modulation order $L$. under a certain delay constraint. The latency includes processing delay ($D_p$) which is determined by the reception algorithm's complexity and the processor's capability at the Rx, and transmission delay ($D_t$) which depends on the volume of traffic, bandwidth, and received SNR/SINR. For traditional ZF, IA and IN, they have a hard capacity which is subject to the system's DoFs, whereas for MF and the proposed ICom/SIC-ZF, the capacity has a soft feature, i.e., the number of accommodated users is independent of $N_{RF}$.

As for $D_p$, we take 100 ms as the latency bound for users traffic [40]. Based on the achievable processing speed of the base station, 76.8GFLOPs [41], and a mobile station, from several to tens of GFLOPs [42], we use 1GFLOPs as a reference processing speed in the following evaluation. Fig. 14 shows a total $D_p$ of $K = 10$ signals under different $N_{RF}$ and $L$. Once $N_{RF}$ increases with $L$, decreases with $N_{RF}$. Assuming $L > 8$ and $N_{RF} < 4$, $D_p$ is too large to be acceptable. In this case, either increasing $N_{RF}$ or reducing the number of simultaneously supported users is the feasible way to guarantee that each user’s $D_p$ is under the given threshold. Fig. 15 plots the total $D_p$ under different $K$ and $N_{RF}$. All users are assumed to adopt QPSK. $D_p$ increases with $K$ and decreases with $N_{RF}$. Under $K > 16$ and $N_{RF} < 8$, $D_p$ becomes unacceptable. In order to decrease $D_p$, we can either increase $N_{RF}$ or reduce $K$.

Fig. 16 shows the requirement of $N_{RF}$ under the 100 ms $D_p$-constraint, different $L$ and $K$. BPSK, QPSK, 8PSK, 16-QAM and 64-QAM are considered. Given $L \leq 16$ and $K \leq 18$, at least 2 Rx-antennas can be saved with the proposed reception mechanism. When $K = 10$ and $L = 4$, only 2 Rx-antennas are required to support all the users simultaneously. However, when a high order modulation (e.g., $L \geq 64$) is employed, the traditional ZF should be directly employed since no Rx-antenna savings can be achieved (see Table 3).

Since WLANs have been widely deployed, we adopt IEEE 802.11ac, an emerging WLANs protocol, as an example to evaluate the delay ($D_t$) performance of different schemes. By assuming that each user uses 20 MHz bandwidth to send a single spatial stream to the access point, the achievable data rates under different modulation and coding schemes (MCS) are given in Table 3 [43]. With traditional ZF, IA and IN, $K$ users are assumed to fairly share the system’s DoFs via user scheduling such as round-robin [44], etc., so that the effective bandwidth of each user is the total system bandwidth divided by $K$. In contrast, with MF and ICom/SIC-ZF, each user exclusively uses the system bandwidth. Provided that all users have the same transmit power and experience statistically identical channel fading, traditional ZF, IA, IN and ICom/SIC-ZF yield the same user’s spectral efficiency, since they all mitigate CCI, and hence outperform MF. We define the system bandwidth, an arbitrary user's data volume and its SE as $W$, $V$, and $C$, respectively. By omitting other types of latency introduced by channel estimation, user scheduling, etc., $D_t$ of a user with ZF, IA and IN is given as $KV/(N_{RF}W)$. With ICom/SIC-ZF, one user’s $D_t$ is computed as $V/(WC)$. In practice, $K$ is always greater than $N_{RF}$, so $D_t$ of ICom/SIC-ZF is smaller than that of traditional ZF, IA and IN. Without loss of generality, we only evaluate $D_t$ of traditional ZF and ICom/SIC-ZF in the following.

Fig. 17 plots $D_t$ of traditional ZF and the proposed ICom/SIC-ZF under the 100 ms $D_p$-constraint. The traffic volume ($V$) of each user’s session is assumed to be 1000Mb. For a fair comparison, both strategies are simulated with the same $N_{RF}$. According to Table 3, data rate ($WC$) grows as $L$ increases, so delays ($D_t$) of both schemes decrease accordingly. With the $D_p$-constraint, the minimum $N_{RF}$ of ICom/SIC-ZF with different $K$s can be found from Fig. 16. It can be seen from Fig. 17 that the curve of $D_t$ of ICom/SIC-ZF is below that of ZF. Moreover, with ICom/SIC-ZF, when BPSK is adopted ($L = 2$), 2 Rx-antennas are enough to support up to 20 users simultaneously, whereas for $L > 2$, more Rx-antennas are required as $K$ increases so as to meet the 100 ms $D_p$-constraint. Hence, provided with the same $N_{RF}$ for ICom/SIC-ZF, $D_t$ of ZF grows as $K$ increases when $L = 2$, but decreases with an increase of $K$ when $L > 2$. In contrast, with ICom/SIC-ZF each user exclusively occupies $W$, and hence its $D_t$ does not vary with $K$.

To summarize, under a certain $D_p$ threshold, $D_t$ of the proposed mechanism outperforms that of ZF, IA and IN for a large $V$ and approximates them under small $V$ values.

Table 3

<table>
<thead>
<tr>
<th>MCS index</th>
<th>Modulation</th>
<th>Code rate</th>
<th>20 MHz data rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>BPSK</td>
<td>1/2</td>
<td>7.2 Mbps</td>
</tr>
<tr>
<td>1</td>
<td>QPSK</td>
<td>1/2</td>
<td>14.4 Mbps</td>
</tr>
<tr>
<td>3</td>
<td>16-QAM</td>
<td>1/2</td>
<td>28.9 Mbps</td>
</tr>
<tr>
<td>5</td>
<td>64-QAM</td>
<td>2/3</td>
<td>57.8 Mbps</td>
</tr>
</tbody>
</table>

Fig. 17. Comparison of transmission delays under the 100 ms $D_p$-constraint.
7. Conclusion

In this paper, we proposed an iCom-based MAC-Rx structure employing ZF and SIC, namely iCom/SIC-ZF. By exploiting the constructive/destructive interactions among interfering signals, the dimension of interference can be reduced significantly. \( K > N_x \) signals can then be decoded successively. With iCom/SIC-ZF, no Tx-side cooperation is required. All Tx's send signals to the common Rx without any sacrifice for the others. In addition, by exploiting the relative strength of and spatial correlation between mutually interfering signals, we employed reception mode adaptation to further improve the performance of the proposed Rx structure. By comparing it with other methods, the proposed reception mechanism is shown to achieve a remarkable improvement of system spectral efficiency, latency performance and flexible tradeoff between the requirement of Rx-antennas and computational complexity, thus facilitating the implementation of practical communication systems.

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