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Power Scaling of Full-Duplex Two-Way Millimeter-Wave Relay with Massive MIMO

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Abstract—In this paper, the asymptotic performance of a full-duplex (FD) millimeter wave (mmWave) two-way relay (TWR) system with massive multiple-input multiple-output (MIMO) is analyzed. After applying conjugate array response vector-based precoding to a hybrid analog and digital MIMO structure in the mmWave band, the asymptotic spectral efficiency of the system is derived when the number of antennas approaches infinity. Based on the asymptotic performance and different system configurations, the power scaling schemes are investigated. Considering different power scaling strategies at each node, eight power scaling cases are discussed, and the corresponding asymptotic spectral efficiencies are derived. The results show that interference and noise can be fully eliminated by increasing the number of antennas. A high spectral efficiency can be achieved in a system wherein all the nodes are equipped with large antenna arrays. If the number of antennas of node A and node B is limited, then system can achieve considerable spectral efficiency and energy efficiency. For both configurations, the power of node A and node B can be scaled down without spectral efficiency loss. However, if only one node is equipped with limited antennas, the node with massive arrays can achieve a higher spectral efficiency. Moreover, we analyze the effect of the FD SI levels. We show that the spectral efficiency decreases rapidly when the intensity of SI increases, which is related to the ability of SI cancellation. Under ideal conditions, the FD system can achieve twice the spectral efficiency of its conventional half-duplex (HD) counterpart. However, the FD SI and finite antennas degrade the spectral efficiency of the FD system.

Index Terms—Millimeter wave; full-duplex; two-way relay; power scaling; massive MIMO.

I. INTRODUCTION

In two-way relay (TWR) systems, the two nodes can transmit signals to the relay simultaneously, and the relay forwards the received signals during the next transmission period. Based on physical layer network coding, the self-interference (SI) of TWR transmission can be fully cancelled [1]. Compared with one-way relay systems, TWR transmission can potentially double the spectral efficiency. Full-duplex (FD) communications, which transmit and receive in the same frequency band simultaneously, can further improve the spectral efficiency [2], [3]. Combined with FD transmission, an FD TWR relay can transmit and receive signals simultaneously,

and the information-containing messages are exchanged in a single period [4]. However, FD leads to additional SI when the node receives the signal from its own transmitting antennas. In addition, FD TWR means that the nodes and relay have to work in FD mode, and thus, SI exist at every node. The throughput of FD TWR is analyzed in [5]. The FD scheme can achieve almost double the throughput with high signal-to-self-interference plus noise ratios and large packet sizes, which means that strong SI cancellation is necessary for FD systems. In [6], [7], the SI suppression of the FD TWR system is considered. The relay and receiving beamforming is jointly designed to minimize the mean square error (MSE) and SI cancellation in [6]. For imperfect channel state information (CSI), a transceiver design is proposed in [7]. Instead of eliminating the SI completely, the end-to-end performance is maximized by jointly optimizing the beamforming matrix at the relay and the sources [8]. The FD scheme can improve the achievable data rates over the conventional half-duplex (HD) scheme [9].

To further improve the performance, massive multiple-input multiple-output (MIMO) is introduced into TWR systems in [10]–[14]. Massive MIMO helps overcome the SI at the relay. A system of multiple user pairs exchanging information through an FD relay is analyzed in [10]. With a large number of antennas, the power of the users and relay can be reduced without significant performance losses. The users' ergodic rates are derived with a finite number of antennas in [11]. The rates increase with the number of antennas at the relay and decrease with the number of user pairs. In [12], the asymptotic spectral efficiencies of the proposed power-scaling schemes are derived when the number of relay antennas approaches infinity, and both maximum-ratio combining/maximum-ratio transmission (MRC/MRT) and zero-forcing reception/zero-forcing transmission (ZFR/ZFT) schemes are considered. With massive relay antennas, the SI can be reduced, and inter-pair interference and inter-user interference can be eliminated. Imperfect CSI is analyzed in [13]. If the power scaling scheme is properly selected, then the effects of channel estimation errors, SI, other interference and noise can be eliminated when the number of antennas approaches infinity. Moreover, FD TWR outperforms the HD system when the number of antennas is large. The impact of antenna correlation is investigated in [14]. The antenna correlation at both the user transceiver sides and the relay receiver side has detrimental effects on the sum rate. Thus, a larger number of user antennas may not always lead to a higher sum rate.

Millimeter wave (mmWave) and massive MIMO are both

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key promising technologies [15]. When working in the mmWave band (approximately 30 to 300 GHz), mmWave communications can utilize the large available bandwidth [16], [17]. The short wavelength of mmWaves enables tightly packed large antenna arrays, resulting in a high-gain beam. The large available bandwidth and high-gain beam of mmWave communications can provide very high rates in TWR systems, which can be applied in cellular systems [18]. However, different from microwave communications, the high free-space pathloss and tightly packed large antenna arrays of mmWave systems lead to a sparse spatial scattering in the mmWave channel. In addition, conventional fully digital MIMO architecture is no more fit for mmWave transmission due to its high cost and power consumption [19]. In mmWave systems, a hybrid analog-digital MIMO processing structure, including radio frequency (RF) chains and baseband (BB) filters, is often applied to achieve a good balance between cost and performance. An orthogonal matching pursuit (OMP)-based spatially sparse precoding scheme is proposed in [19] and achieves an accurate approximation of the conventional full digital structure.

Thanks to state-of-the-art work on FD mmWave communications [20]–[23], significant FD SI cancellation can be realized. A 60 GHz FD transceiver and 25 GHz magnetic-free non-reciprocal passive circulator implemented in 45 nm SOI CMOS are discussed in [20]. With polarization-based antennas and RF cancellation, 80 dB SI suppression can be achieved. In the antenna domain, polarization isolation can suppress the SI by over 10 dB [21]. The SI intensity is inversely proportional to the transmitting and receiving antenna spacing. A fully integrated 60 GHz direct-conversion transceiver in 45 nm SOI CMOS for FD transmission is proposed in [22]. Over 70 dB SI suppression is achieved by utilizing antenna and RF cancellation. Considering both near-end cross talk and far-end cross talk SI, [23] puts forward an adaptive multi-segment frequency-domain cancellation scheme in an E-band full-duplex backhaul scenario. Combined with antenna isolation, the proposed scheme can cancel the SI to a degree below the noise floor. Compared to microwave communications, FD SI presents different characteristics in the mmWave band [23], [24]. It also contains both line-of-sight (LOS) SI and non-line-of-sight (NLOS) SI. In [25]–[27], FD SI precoding schemes in mmWave systems are proposed. Through proper hybrid precoding design, SI can be significantly suppressed. The joint beamformer design can preserve the dimensions of the transmitting signal and suppress SI [25]. OMP-based SI cancellation precoding design is proposed in [26], in which SI is eliminated through zero space projection. Combined with TWR systems, a mmWave TWR beamforming design is studied in [28]–[30]. In [28], the RF precoder is designed based on the Gram-Schmidt process, and the BB precoder is based on the weighted sum minimum mean square error. A mmWave TWR secrecy beamforming design is analyzed in [30].

To the best of our knowledge, very few results have been obtained for FD mmWave TWR systems. The integration of FD communications, mmWave transmission and the TWR system can achieve a high spectral efficiency if the SI can

be well addressed. Thus, we consider an FD mmWave TWR system with residual FD SI, which contains LOS and NLOS SI. When the number of antennas approaches infinity, the asymptotic performance of the system is analyzed. Then, eight different power scaling schemes are derived. Different antenna configurations are also considered, and the corresponding power scaling schemes are presented. The performances of FD SI and TWR SI are also analyzed. Moreover, the FD system is compared with a conventional HD system. The contributions of this paper are summarized as follows:

- To the best of our knowledge, the asymptotic analysis of large-scale antennas has only been studied in the microwave band. According to the special sparse scattering channel matrix of mmWave communications and the conjugate array response vector-based RF precoding matrix, we derived the asymptotic performance considering an infinite number of antennas. We show that as the number of antennas approaches infinity, the interference and noise in the system can be suppressed.
- We analyzed eight different power scaling schemes and derived the corresponding asymptotic signal-to-interference-plus-noise ratios (SINRs). We show that under ideal conditions, the TWR SI, forwarded noise at the relay, FD SI and noise at the nodes can be fully eliminated. For different power scaling schemes, the end-to-end spectral efficiency performance varies. The power of the nodes can be scaled without significant spectral efficiency loss.
- With a limited number of antennas used for a node, the asymptotic SINRs are derived based on the power scaling schemes. When the power of the nodes is scaled, the system always achieves a comparatively high spectral efficiency in all antenna configurations. When the number of antennas of one node is limited, the asymptotic SINRs approach zero if the power of the node is scaled.
- We study the impacts of FD SI and TWR SI. We show that different power scaling schemes lead to different SI cancellation abilities for different nodes. Overall, the FD SI suppression of the relay is better than that of the nodes. When the power of the node is scaled, a better FD SI suppression performance can be achieved. In addition, the TWR SI can also be eliminated through a large number of antennas.
- We also compare the FD system with the HD counterpart. Under ideal conditions, the FD system can achieve twice the spectral efficiency of the HD system. However, as the intensity of the FD SI increases, the FD system performance decreases. The performance decreases rapidly if the power scaling scheme is not effectively able to suppress FD SI. To achieve a better performance than that of the HD system, an FD SI cancellation scheme is needed.

The remainder of this paper is organized as follows. In Section II, the system model and channel model of the FD mmWave TWR system are presented. Then, the asymptotic analysis and power scaling schemes are derived in Section III. The power scaling and asymptotic analysis are extended to

Fig. 1: System model of the FD two-way mmWave relay.

other antenna configurations in Section IV. The HD counterpart is presented in Section V. In Section VI, the numerical results of the power scaling schemes are shown. Conclusions are drawn in Section VII.

II. SYSTEM MODEL

In this section, we will introduce the system model of the two-way relay. Then, the transmission scheme of the system is illustrated and the spectral efficiency of the system is obtained. The sparse spatial scattering and FD SI channel models are discussed separately.

First, an FD two-way mmWave relay system is considered, as shown in Fig. 1. The relay system is composed of one relay and two nodes: A and B. The two nodes are connected by the relay, and the direct link between the nodes is neglected due to high pathloss. All three nodes operate under the FD transmission mode. Specifically, they transmit and receive simultaneously. During TWR transmission, both nodes transmit information to the relay node, while the relay forwards the received signal to the nodes. Because of its high cost and power consumption, conventional fully digital MIMO processing performed in the microwave band is inappropriate for mmWave systems. Therefore, the hybrid analog-digital MIMO processing structure is applied in this system.

When operating in the mmWave band, all nodes adopt a hybrid analog and digital MIMO structure, as shown in Fig. 2. The precoding structure consists of the RF portion and BB portion with phase shifters and BB processors, respectively. Each RF chain is connected to all the antennas; i.e., it is a fully connected structure. At node A, the numbers of transmitting antennas and receiving antennas are N_{AT} and N_{AR} , respectively. The numbers of transmitting RF chains and receiving RF chains are N_{ARFT} and N_{ARFR} , respectively. The number of data streams is N_{AS} . The corresponding parameters of node A are shown in Fig. 2, where $\mathbf{W}_{BBT}$, $\mathbf{W}_{RFT}$, $\mathbf{W}_{BBR}$ and $\mathbf{W}_{RFR}$ are the precoding matrices. Similarly, N_{BT} , N_{BR} , N_{BRFT} , N_{BRFR} and N_{BS} are the corresponding parameters of node B. N_{RT} , N_{RR} , N_{RRFT} , N_{RRFR} and N_{RS} are the corresponding parameters of the relay.

A. Transmission Scheme

In the first step, after precoding by an $N_{ARFT} \times N_{AS}$ BB precoding matrix $\mathbf{W}_{BBT}$ and $N_{AT} \times N_{ARFT}$ RF precoding matrix $\mathbf{W}_{RFT}$, the n th $N_{AT} \times 1$, the transmitting signal of

node A can be expressed as follows:

$$\mathbf{x}_A(n) = \sqrt{P_A} \mathbf{W}_{RFT} \mathbf{W}_{BBT} \mathbf{s}_A(n), \quad (1)$$

where $\mathbf{s}_A(n)$ denotes $N_{AS} \times 1$ transmitting data with $E[\mathbf{s}_A(n)\mathbf{s}_A(n)^H] = N_{AS}^{-1} \mathbf{I}_{N_{AS}}$ and P_A denotes the transmitting power of node A. The n th $N_{BT} \times 1$ transmitting signal of node B is expressed as

$$\mathbf{x}_B(n) = \sqrt{P_B} \mathbf{G}_{RFT} \mathbf{G}_{BBT} \mathbf{s}_B(n), \quad (2)$$

where $\mathbf{G}_{BBT}$ denotes the $N_{BRFT} \times N_{BS}$ BB precoding matrix, $\mathbf{G}_{RFT}$ denotes the $N_{BT} \times N_{BRFT}$ RF precoding matrix, $\mathbf{s}_B(n)$ denotes the $N_{BS} \times 1$ transmitting data and P_B denotes the transmitting power of node B. Then, the received signal at the relay is

$$\mathbf{y}_R(n) = \mathbf{H}_{AR} \mathbf{x}_A(n) + \mathbf{H}_{BR} \mathbf{x}_B(n) + \mathbf{H}_{RR} \mathbf{x}_R(n) + \mathbf{n}_R(n), \quad (3)$$

where $\mathbf{H}_{AR}$ denotes the $N_{RR} \times N_{AT}$ channel matrix from node A to the relay, $\mathbf{H}_{BR}$ denotes the $N_{RR} \times N_{BT}$ channel matrix from node B to the relay, $\mathbf{H}_{RR}$ denotes the SI channel of the relay, and $\mathbf{x}_R(n)$ denotes the $N_{RT} \times 1$ transmitting signal of the relay. $\mathbf{n}_R(n)$ denotes the $N_{RR} \times 1$ noise vector with independent and identically distributed (i.i.d.) $\mathcal{CN}(0, \sigma_R^2)$, where σ_R^2 denotes the noise variance. At the relay node, the received signal is first processed by the receiving combiner and then precoded by the transmitting precoder. Considering the processing delay, the $N_{RT} \times 1$ transmitting signal of the relay is denoted

$$\mathbf{x}_R(n) = \beta \mathbf{F}_{RFT} \mathbf{F}_{BBT} \mathbf{F}_{BBR} \mathbf{F}_{RFR} \mathbf{y}_R(n-1), \quad (4)$$

where $\mathbf{F}_{RFR}$ denotes the $N_{RRFR} \times N_{RR}$ RF combiner of the receiving node of the relay, $\mathbf{F}_{BBR}$ denotes the $N_{RS} \times N_{RRFR}$ BB combiner of the receiving node of the relay, $\mathbf{F}_{BBT}$ denotes the $N_{RRFT} \times N_{RS}$ BB precoder of the transmitting node of the relay, $\mathbf{F}_{RFT}$ denotes the $N_{RT} \times N_{RRFT}$ RF precoder of the transmitting node of the relay, β denotes the amplification factor designed to satisfy the power constraint of the relay and $\mathbf{y}_R(n-1)$ denotes one time slot delay. Based on the transmitting power constraint of the relay $P_R = \text{tr}\{E[\mathbf{x}_R(n)\mathbf{x}_R(n)^H]\}$, (3) and (4), the amplification factor is expressed as

$$\beta = \sqrt{\frac{P_R}{P_A \Delta_1 + P_B \Delta_2 + P_R \Delta_3 + \sigma_R^2 \Delta_4}}, \quad (5)$$

in which $\Delta_1 = \text{tr}\{(\mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T\}$, $\Delta_2 = \text{tr}\{(\mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T\}$, Δ_3

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$= \text{tr}\{(\mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR})^H \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR}\}$ and $\Delta_4 = \text{tr}\{(\mathbf{F}_T \mathbf{F}_R)^H \mathbf{F}_T \mathbf{F}_R\}$. In (5), we use $\mathbf{F}_T = \mathbf{F}_{\text{RFT}} \mathbf{F}_{\text{BBT}}$, $\mathbf{F}_R = \mathbf{F}_{\text{BBR}} \mathbf{F}_{\text{RFR}}$, $\mathbf{W}_T = \mathbf{W}_{\text{RFT}} \mathbf{W}_{\text{BBT}}$ and $\mathbf{G}_T = \mathbf{G}_{\text{RFT}} \mathbf{G}_{\text{BBT}}$ for simplicity, respectively.

In the second step, the relay broadcasts the received signal to the nodes. The received signal at node A is written as

$$\mathbf{y}_A(n) = \mathbf{H}_{RA} \mathbf{x}_R(n) + \mathbf{H}_{AA} \mathbf{x}_A(n) + \mathbf{n}_A(n), \quad (6)$$

where $\mathbf{H}_{RA}$ denotes the $N_{AR} \times N_{RT}$ channel matrix from the relay to node A, $\mathbf{H}_{AA}$ denotes the $N_{AR} \times N_{AT}$ SI channel of node A and $\mathbf{n}_A(n)$ denotes the $N_{AR} \times 1$ noise vector with i.i.d. $\mathcal{CN}(0, \sigma_A^2)$. Finally, the received data of node A are written as

$$\begin{aligned} \hat{\mathbf{y}}_A(n) &= \mathbf{W}_{\text{BBR}} \mathbf{W}_{\text{RFR}} \mathbf{y}_A(n) \\ &= \mathbf{W}_{\text{BBR}} \mathbf{W}_{\text{RFR}} [\mathbf{H}_{RA} \mathbf{x}_R(n) + \mathbf{H}_{AA} \mathbf{x}_A(n) + \mathbf{n}_A(n)] \end{aligned} \quad (7)$$

where $\mathbf{W}_{\text{BBR}}$ and $\mathbf{W}_{\text{RFR}}$ denote the $N_{AS} \times N_{ARFR}$ BB precoder and the $N_{ARFR} \times N_{AR}$ RF precoder of node A, respectively. By substituting (1), (2), (3) and (4) into (7), we can obtain (8), where $\mathbf{W}_R = \mathbf{W}_{\text{BBR}} \mathbf{W}_{\text{RFR}}$. In (8), the first term is the SI of the TWR transmission, which is the desired signal of node B, and it can be cancelled by utilizing the CSI of the system. The second term is the desired signal of node A; the third term is the propagated FD SI at the relay; the fourth term is the propagated noise of the relay; the fifth term is the SI of node A; and the last term is the noise at node A. After TWR SI cancellation [1], the received signal is expressed as follows:

$$\begin{aligned} \hat{\mathbf{y}}_A(n) &= \beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \sqrt{P_B} \mathbf{G}_T \mathbf{s}_B(n-1) + \\ &\quad \beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR} \mathbf{x}_R(n-1) + \mathbf{W}_R \mathbf{n}_A(n) + \\ &\quad \beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{n}_R(n-1) + \mathbf{W}_R \mathbf{H}_{AA} \sqrt{P_A} \mathbf{W}_T \mathbf{s}_A(n). \end{aligned} \quad (9)$$

Therefore, the SINR at node A is expressed as

$$\gamma_{B \rightarrow A} = \frac{\beta^2 P_B \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{\beta^2 P_R \Delta_{RR} + \beta^2 \sigma_{nR}^2 \Delta_{nR} + P_A \Delta_{AA} + \sigma_{nA}^2 \mathbf{W}_R (\mathbf{W}_R)^H}, \quad (10)$$

where $\Delta_{RR} = \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR} (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR})^H$, $\Delta_{nR} = \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R)^H$ and $\Delta_{AA} = \mathbf{W}_R \mathbf{H}_{AA} \mathbf{W}_T (\mathbf{W}_R \mathbf{H}_{AA} \mathbf{W}_T)^H$. The received signal at node B is expressed as

$$\mathbf{y}_B(n) = \mathbf{H}_{RB} \mathbf{x}_R(n) + \mathbf{H}_{BB} \mathbf{x}_B(n) + \mathbf{n}_B(n), \quad (11)$$

where $\mathbf{H}_{RB}$ denotes the $N_{BR} \times N_{RT}$ channel matrix from the relay to node B, $\mathbf{H}_{BB}$ denotes the $N_{BR} \times N_{BT}$ SI channel matrix of node B and $\mathbf{n}_B(n)$ denotes the $N_{BR} \times 1$ noise vector with i.i.d. $\mathcal{CN}(0, \sigma_B^2)$. The received data of node B are written

as

$$\hat{\mathbf{y}}_B(n) = \mathbf{G}_{\text{BBR}} \mathbf{G}_{\text{RFR}} [\mathbf{H}_{RB} \mathbf{x}_R(n) + \mathbf{H}_{BB} \mathbf{x}_B(n) + \mathbf{n}_B(n)], \quad (12)$$

where $\mathbf{G}_{\text{BBR}}$ denotes the $N_{BS} \times N_{BRFR}$ BB precoder of node B, and $\mathbf{G}_{\text{RFR}}$ denotes the $N_{BRFR} \times N_{BR}$ RF precoder of node B. Similarly, after TWR SI cancellation, the received data of node B are expressed as

$$\begin{aligned} \hat{\mathbf{y}}_B(n) &= \beta \mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \sqrt{P_A} \mathbf{W}_T \mathbf{s}_A(n-1) + \\ &\quad \beta \mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR} \mathbf{x}_R(n-1) + \mathbf{G}_R \mathbf{n}_B(n) + \\ &\quad \beta \mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{n}_R(n-1) + \mathbf{G}_R \mathbf{H}_{BB} \sqrt{P_B} \mathbf{G}_T \mathbf{s}_B(n), \end{aligned} \quad (13)$$

where $\mathbf{G}_R = \mathbf{G}_{\text{BBR}} \mathbf{G}_{\text{RFR}}$. The SINR of node B is written as

$$\gamma_{A \rightarrow B} = \frac{\beta^2 P_A \mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T (\mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H}{\beta^2 P_R \Omega_{RR} + \beta^2 \sigma_{nR}^2 \Omega_{nR} + P_B \Omega_{BB} + \sigma_{nB}^2 \mathbf{G}_R (\mathbf{G}_R)^H}, \quad (14)$$

where $\Omega_{RR} = \mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR} (\mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR})^H$, $\Omega_{nR} = \mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R (\mathbf{G}_R \mathbf{H}_{RB} \mathbf{F}_T \mathbf{F}_R)^H$, $\mathbf{F} = \mathbf{F}_T \mathbf{F}_R$ and $\Omega_{BB} = \mathbf{G}_R \mathbf{H}_{BB} \mathbf{G}_T (\mathbf{G}_R \mathbf{H}_{BB} \mathbf{G}_T)^H$.

The end-to-end sum spectral efficiency of the FD mmWave TWR system is evaluated as

$$R = R_A + R_B, \quad (15)$$

where $R_A = \log_2 |\mathbf{I} + \gamma_{B \rightarrow A}|$ and $R_B = \log_2 |\mathbf{I} + \gamma_{A \rightarrow B}|$.

B. Channel Model for the Links and SI

In contrast to microwave channels, the high-frequency transmission and special antenna structure of mmWave communications lead to sparse scattering environments. Traditional MIMO statistical fading distributions are inaccurate for mmWave channel modeling. Therefore, based on the Saleh-Valenzuela model [19], [31], [32], the sparse spatial scattering in the mmWave channel can be expressed as

$$\mathbf{H} = \sqrt{\frac{N_t N_r}{KL}} \sum_{k=1}^K \sum_{l=1}^L \alpha_{kl} \Psi_r(\phi_{kl}^r, \theta_{kl}^r) \Psi_t(\phi_{kl}^t, \theta_{kl}^t)^H, \quad (16)$$

where N_t denotes the number of transmitting antennas, N_r denotes the number of receiving antennas, and α_{kl} is the complex gain of the l -th path in the k -th cluster. K denotes the number of clusters, and $K \sim \max\{\text{Poisson}(\kappa), 1\}$, where κ is the Poisson distribution parameter and each of the K clusters contains L subpaths. The functions $\Psi_r(\phi_{kl}^r, \theta_{kl}^r)$ and $\Psi_t(\phi_{kl}^t, \theta_{kl}^t)$ represent the normalized receiving and transmitting array response vectors, respectively. The angles ϕ_{kl}^r (θ_{kl}^r) and ϕ_{kl}^t (θ_{kl}^t) are the azimuth (elevation) arrival and departure

$$\begin{aligned} \hat{\mathbf{y}}_A(n) &= \underbrace{\beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \sqrt{P_A} \mathbf{W}_T \mathbf{s}_A(n-1)}_{\text{TWR self-interference}} + \underbrace{\beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \sqrt{P_B} \mathbf{G}_T \mathbf{s}_B(n-1)}_{\text{desired signal}} \\ &\quad + \underbrace{\beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR} \mathbf{x}_R(n-1)}_{\text{FD self-interference at relay}} + \underbrace{\beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{n}_R(n-1)}_{\text{noise at relay}} \\ &\quad + \underbrace{\mathbf{W}_R \mathbf{H}_{AA} \sqrt{P_A} \mathbf{W}_T \mathbf{s}_A(n)}_{\text{FD self-interference at node A}} + \underbrace{\mathbf{W}_R \mathbf{n}_A(n)}_{\text{noise at node A}}, \end{aligned} \quad (8)$$

Fig. 2: The hybrid analog and digital MIMO structure of node A.

angles, respectively. For a uniform planar array (UPA), the response function can be expressed as follows:

$$\Psi_{\text{UPA}}(\phi, \theta) = \frac{1}{\sqrt{N}} \left[1, \dots, e^{j \frac{2\pi}{\lambda} d(m \sin(\phi) \sin(\theta) + n \cos(\theta))}, \dots, e^{j \frac{2\pi}{\lambda} d((W-1) \sin(\phi) \sin(\theta) + (H-1) \cos(\theta))} \right]^T, \quad (17)$$

where d is the inter-element spacing, λ is the carrier wavelength, $0 \leq m \leq W-1$, $0 \leq n \leq H-1$, $N = WH$ and N denotes the number of antennas. For a uniform linear array (ULA), the response function can be expressed as

$$\Psi_{\text{ULA}}(\phi) = \frac{1}{\sqrt{N}} \left[1, e^{j \frac{2\pi}{\lambda} d \sin(\phi)}, \dots, e^{j \frac{2\pi}{\lambda} (N-1) d \sin(\phi)} \right]^T. \quad (18)$$

The array response function is determined by the structure of the antenna array configuration. We use the ULA for analysis. The sequential summation in (16) can be rewritten in matrix form as

$$\mathbf{H} = \sum_{k=1}^K \mathbf{A}_k^r \mathbf{\Lambda}_k (\mathbf{A}_k^t)^H = \mathbf{A}^r \mathbf{\Lambda} (\mathbf{A}^t)^H, \quad (19)$$

where $\mathbf{A}_k^r = [\Psi_{k1}^r, \Psi_{k2}^r, \dots, \Psi_{kL}^r]$, $\mathbf{A}_k^t = [\Psi_{k1}^t, \Psi_{k2}^t, \dots, \Psi_{kL}^t]$, the coefficient matrix $\mathbf{\Lambda}_k = \text{diag}[\gamma \alpha_{k1}, \gamma \alpha_{k2}, \dots, \gamma \alpha_{kL}]$ and $\gamma = \sqrt{\frac{N_t N_r}{KL}}$. The matrices $\mathbf{A}^r$, $\mathbf{\Lambda}$ and $\mathbf{A}^t$ are composed of $\mathbf{A}_k^r$, $\mathbf{\Lambda}_k$ and $\mathbf{A}_k^t$, respectively.

In FD communications, transmitting and receiving simultaneously leads to SI from the transmitting node to the receiving node. During transmission in the mmWave band, the high gain leads to additional SI, namely, NLOS SI, which results from the reflections due to various nearby clusters. As illustrated in [20]–[23], FD SI can be suppressed up to 80 dB through antenna isolation, RF cancellation and digital cancellation schemes. We mainly focus on the residual SI after SI cancellation schemes and the influence of different levels of SI on the system. Therefore, the residual SI consists of

two parts: LOS SI and NLOS SI [23], [24], [26]. The NLOS SI is assumed to be a decayed sparse-scattering mmWave channel. The NLOS SI is written as $\mathbf{H}_{\text{NLOS}}$, which denotes the channel model in (19). LOS SI is the direct link from the transmitting antennas to the receiving antennas, which can be modeled by a near-field model. The channel coefficient between the m th receiving antenna and the n th transmitting antenna is denoted $h_{mn}^{\text{SI}} = \mu_{mn} e^{-j \frac{2\pi}{\lambda} d_{mn}}$, where μ_{mn} is the normalization coefficient and d_{mn} denotes the distance between the m th receiving antenna and the n th transmitting antenna. The SI channel matrix is written as

$$\mathbf{H}_{\text{LOS}} = [\Phi^{\text{SI}r} (\Phi^{\text{SI}t})^H] \circ \mathbf{H}^{\text{SI}}, \quad (20)$$

where the antenna array response vectors are $\Phi^{\text{SI}r} = [1, e^{j \frac{2\pi}{\lambda} d \sin \varphi_2^r}, \dots, e^{j \frac{2\pi}{\lambda} (n_R-1) \sin \varphi_{n_R}^r}]^H$, $\Phi^{\text{SI}t} = [1, e^{j \frac{2\pi}{\lambda} d \sin \varphi_2^t}, \dots, e^{j \frac{2\pi}{\lambda} (n_T-1) \sin \varphi_{n_T}^t}]^H$, φ_n^r and φ_n^t are the included angles of the corresponding antennas, and $\mathbf{H}^{\text{SI}} = \{h_{mn}^{\text{SI}} | n = 1, \dots, n_T, m = 1, \dots, n_R\}$. The operator $\circ$ denotes the Hadamard product, which takes two matrices of the same dimensions and multiplies the corresponding elements of the two matrices. To evaluate different levels of residual SI, the intensity coefficient of the SI channel η_{SI} is introduced. The FD mmWave SI channel can be expressed as

$$\mathbf{H}_{\text{SI}} = \eta_{\text{SI}} (\mathbf{H}_{\text{LOS}} + \mathbf{H}_{\text{NLOS}}). \quad (21)$$

III. ASYMPTOTIC ANALYSIS AND POWER SCALING

In this section, we first introduce the RF precoding design for the asymptotic analysis. Then, we derive the asymptotic property of the channel and the RF precoder, which illustrates the interference suppression capability of large antenna arrays. Based on the asymptotic property, the asymptotic performance and the theoretical upper bound of the relay system are studied as the number of antennas approaches infinity.

According to the different antenna configurations of the three nodes, we analyze Configurations ARB, aRb and ARb below. Large antenna arrays and limited antenna arrays are

respectively indicated by uppercase and lowercase letters. R, A and B denote the relay and two nodes, respectively. Thus, Configuration ARB denotes that all the three nodes are equipped with a large number of antennas, and the configuration aRb denotes that the numbers of antennas of nodes A and B are limited, but that the relay is equipped with massive MIMO. Configuration ARb denotes that the number of antennas is limited only in node B. The power scaling results of Configuration ARB for different power scaling cases are presented in this section. The power scaling schemes for Configurations aRb and ARb are shown in the next section.

A. Precoding Design

In conventional relay systems, maximum ratio combining (MRC) and maximum ratio transmission (MRT) schemes are normally used [1]. However, this approach cannot be employed in the hybrid analog and digital MIMO structure of the mmWave system due to the special structural limitation of the phase shifters. In [33] and [34], an equal gain transmission method is utilized to harvest the large array gain by choosing elements of the conjugate transpose of the channel matrix, which is similar to the MRC and MRT schemes.

Considering the unique channel characteristics in (19) and the orthogonality of the array response vectors [35], the RF precoders are chosen based on the matrices $\mathbf{A}^r$ and $\mathbf{A}^t$ of the corresponding channel. The RF precoding matrices of nodes A and B are given as

$$\begin{aligned} \mathbf{W}_{\text{RFR}} &= (\mathbf{A}_{\text{RA}}^r)^H, & \mathbf{W}_{\text{RFT}} &= \mathbf{A}_{\text{AR}}^t, \\ \mathbf{G}_{\text{RFT}} &= \mathbf{A}_{\text{BR}}^t, & \mathbf{G}_{\text{RFR}} &= (\mathbf{A}_{\text{RB}}^r)^H, \end{aligned} \quad (22)$$

where $\mathbf{A}_{\text{RA}}^r$, $\mathbf{A}_{\text{AR}}^t$, $\mathbf{A}_{\text{BR}}^t$, and $\mathbf{A}_{\text{RB}}^r$ are the array response matrices of $\mathbf{H}_{\text{RA}}$, $\mathbf{H}_{\text{AR}}$, $\mathbf{H}_{\text{BR}}$ and $\mathbf{H}_{\text{RB}}$, respectively. The amplify-and-forward (AF) matrices are designed as

$$\mathbf{F}_{\text{RFT}} = \begin{bmatrix} \mathbf{A}_{\text{RB}}^t & \mathbf{A}_{\text{RA}}^t \end{bmatrix}, \mathbf{F}_{\text{RFR}} = \begin{bmatrix} (\mathbf{A}_{\text{AR}}^r)^H \\ (\mathbf{A}_{\text{BR}}^r)^H \end{bmatrix}, \quad (23)$$

where $\mathbf{A}_{\text{RB}}^t$, $\mathbf{A}_{\text{RA}}^t$, $\mathbf{A}_{\text{AR}}^r$, and $\mathbf{A}_{\text{BR}}^r$ are the array response matrices of $\mathbf{H}_{\text{RB}}$, $\mathbf{H}_{\text{RA}}$, $\mathbf{H}_{\text{AR}}$ and $\mathbf{H}_{\text{BR}}$, respectively. In the next section, the asymptotic property is derived based on the channel characteristics and array response matrices. Therefore, to show the performance of asymptotic analysis, the BB precoding matrices are temporarily omitted.

B. Asymptotic Analysis

In the following sections, we introduce the lemmas to be used when analyzing an asymptotic performance.

Lemma 1: Assuming $\mathbf{A} = [\Psi(\phi_1), \Psi(\phi_2), \dots, \Psi(\phi_L)]$ to be the array response matrix of the mmWave channel and $\Psi(\phi_i) = \frac{1}{\sqrt{N}} [1, e^{j\frac{2\pi}{\lambda} d \sin(\phi_i)}, \dots, e^{j\frac{2\pi}{\lambda} (N-1)d \sin(\phi_i)}]^T$, we can obtain

$$\lim_{N \rightarrow \infty} \Psi^H(\phi_i) \Psi(\phi_j) = \begin{cases} 0 & \text{others} \\ 1 & i = j. \end{cases} \quad (24)$$

Proof: See Appendix A. ■

According to Lemma 1, the array response vector is orthogonal to the other array response vectors in the matrix. Because

ϕ_i and ϕ_j are chosen independently from a continuous distribution, the array response vector is orthogonal to the array response vectors of other channel matrices.

Proposition 1: Assuming that $\mathbf{H} = \mathbf{A}^r \mathbf{\Lambda} (\mathbf{A}^t)^H$, where $\mathbf{\Lambda}$ is the coefficient matrix, $\mathbf{A}^r$ and $\mathbf{A}^t$ are the corresponding array response matrices. When N approaches infinity, we have

$$\lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{H} \mathbf{A}^t}{N^2} = \mathbf{\Lambda}, \quad \lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{H} \tilde{\mathbf{A}}^t}{N^2} = \mathbf{0}, \quad (25)$$

where $\tilde{\mathbf{A}}^t$ is the array response matrix of a different channel matrix $\tilde{\mathbf{H}}$ and $\mathbf{\Lambda}$ is a diagonal matrix.

Proof: See Appendix B. ■

Lemma 2: Denote $\mathbf{H}_{\text{LOS}} = [\Phi^{\text{SIr}} (\Phi^{\text{SI}t})^H] \circ \mathbf{H}^{\text{SI}}$ the SI channel matrix, where Φ^{SIr} and $\Phi^{\text{SI}t}$ are the receiving and transmitting antenna array response vectors of the SI channel. $\mathbf{H}^{\text{SI}} = \{h_{mn} | n = 1, \dots, n_T, m = 1, \dots, n_R\}$, where h_{mn} denotes the channel coefficient. Similarly, according to Lemma 1 and Proposition 1, we can obtain

$$\lim_{N \rightarrow \infty} |\mathbf{A}^H \mathbf{H}_{\text{LOS}}| = \mathbf{0}, \quad (26)$$

where $\mathbf{A}^H$ is the array response matrix of the other channels.

C. Power Scaling Case 1: $P_A = E_A, P_B = E_B, P_R = E_R$

For power scaling case 1, $P_A = E_A, P_B = E_B$ and $P_R = E_R$, where E_A, E_B and E_R are the fixed reference power of each node. $P_A = E_A$ indicates that the transmitting power of node A is not scaled, while $P_A = \frac{E_A}{N}$ indicates that the power is scaled by the number of transmitting antennas. Therefore, case 1 denotes the asymptotic performance of the system, where the power of all the nodes is not scaled. The asymptotic performance can be evaluated by Proposition 2.

Proposition 2: For case 1, when the number of antennas approaches infinity, the asymptotic SINR of node A is given by

$$\gamma_{B \rightarrow A} \rightarrow \infty. \quad (27)$$

Proof: See Appendix C. ■

According to (59), FD self-interference at the relay, noise at the relay, FD self-interference at node A and noise at node A can all be suppressed by a large number of antennas.

D. Power Scaling for Other Cases

Now, we analyze different power scaling cases for the system. Because there are three nodes and the power of each node can be scaled, there are eight different cases in total: Case 2: $P_A = \frac{E_A}{N}, P_B = E_B, P_R = E_R$; Case 3: $P_A = E_A, P_B = \frac{E_B}{N}, P_R = E_R$; Case 4: $P_A = E_A, P_B = E_B, P_R = \frac{E_R}{N}$; Case 5: $P_A = \frac{E_A}{N}, P_B = \frac{E_B}{N}, P_R = E_R$; Case 6: $P_A = \frac{E_A}{N}, P_B = E_B, P_R = \frac{E_R}{N}$; Case 7: $P_A = E_A, P_B = \frac{E_B}{N}, P_R = \frac{E_R}{N}$ and Case 8: $P_A = \frac{E_A}{N}, P_B = \frac{E_B}{N}, P_R = \frac{E_R}{N}$. In cases 2–4, the power of only one node is scaled; in cases 5–7, the power of two nodes is scaled; and the power of all three nodes is scaled in case 8. The power scaling schemes denote that transmitting power of the node can be scaled down as the number of antenna increases. For example, the transmitting power of nodes A and B is scaled down as the number of antenna increases in Case 5, while the transmitting power

$$\gamma_{B \rightarrow A} = \frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^4} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}} \xrightarrow{N \rightarrow \infty} \infty. \quad (29)$$

$$\gamma_{B \rightarrow A} = \frac{N^6 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^6 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^6 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^7} + E_A \frac{\Delta_{AA}}{N} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N}}. \quad (30)$$

of the relay remains constant. Based on the result of power scaling case 1, the performance limits of the other cases are presented. In case 2, $P_A = \frac{E_A}{N}$, $P_B = E_B$, $P_R = E_R$, and the derivation shows a difference in the limit of the amplification factors and the asymptotic SINR. Because $P_A = \frac{E_A}{N}$, we can obtain

$$N^{\frac{5}{2}} \beta = \sqrt{\frac{E_R}{E_A \frac{\Delta_1}{N^6} + E_B \frac{\Delta_2}{N^5} + E_R \frac{\Delta_3}{N^5} + \sigma_{nR}^2 \frac{\Delta_4}{N^5}}} \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}, \quad (28)$$

where $E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\}$ is eliminated due to the power scaling of destination A. The asymptotic SINR is (29).

Similarly, minor differences occur among the other cases that can be derived based on the propositions and lemmas. Specifically, the asymptotic SINR is limited in case 7, which is expressed as (30).

According to (59), the node noise can be written as follows:

$$\frac{\mathbf{W}_R \mathbf{W}_R^H}{N} \xrightarrow{N \rightarrow \infty} \mathbf{I}. \quad (31)$$

Therefore, the noise of node A cannot be eliminated through the power scaling scheme in case 7. The results for all eight cases are summarized in Table I, where the asymptotic SINR of case 7 is

$$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \text{tr}(\mathbf{\Lambda}_{RA}^H \mathbf{\Lambda}_{RA} \mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H)}{\mathbf{I} \sigma_{nA}^2 E_A \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H)}. \quad (32)$$

As shown in Table I, the results show different interference and noise suppression abilities, although the SINRs of the seven cases approach infinity. The index of N in the denominator denotes the capability of the corresponding scheme to eliminate interference. For example, cases 3 and 4 have a lower FD SI index for node A than do the other cases, which means that cases 3 and 4 are more easily influenced by the FD SI of node A. In addition, the index is too low to achieve noise suppression in case 7.

TABLE I: Power scaling cases 1-8 for Configuration ARB

Case	Amplification Factor β	Asymptotic SINR $\gamma_{B \rightarrow A}$
1	$N^{\frac{5}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\} + E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}$	$\frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}} \xrightarrow{N \rightarrow \infty} \infty$
2	$N^{\frac{5}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}$	$\frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^4} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}} \xrightarrow{N \rightarrow \infty} \infty$
3	$N^{\frac{5}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\}}}$	$\frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^7} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^7} + E_A \frac{\Delta_{AA}}{N^2} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^2}} \xrightarrow{N \rightarrow \infty} \infty$
4	$N^3 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\} + E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}$	$\frac{N^6 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^6 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^6 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^2} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^2}} \xrightarrow{N \rightarrow \infty} \infty$
5	$N^2 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\} + E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}$	$\frac{N^4 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^4 \beta^2 E_R \frac{\Delta_{RR}}{N^7} + N^4 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^7} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}} \xrightarrow{N \rightarrow \infty} \infty$
6	$N^3 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}$	$\frac{N^6 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^6 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^6 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^2}} \xrightarrow{N \rightarrow \infty} \infty$
7	$N^3 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\}}}$	$\frac{N^6 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^6 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^6 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^7} + E_A \frac{\Delta_{AA}}{N} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N}} \xrightarrow{N \rightarrow \infty} \infty$
8	$N^{\frac{5}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr}\{\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H\} + E_B \text{tr}\{\mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H\}}}$	$\frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^7} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^2}} \xrightarrow{N \rightarrow \infty} \infty$

Furthermore, the asymptotic SINR of node B can easily be derived by utilizing the above results.

E. TWR SI analysis

According to [1], the TWR SI can be cancelled both through physical layer network coding and by increasing the number of antennas. Based on (8), the SINR at node A without TWR SI cancellation is expressed as (33), where $\Delta_A = \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H$. In case 1, the limit of $\gamma_{B \rightarrow A}$ is written as (34).

According to Proposition 2, we can obtain (35). Therefore, the TWR SI can be eliminated in case 1. However, in cases 3 and 7, the SI cannot be fully suppressed. For example, in case 3, we obtain (36), where index N of Δ_A is insufficient because the power of node B is scaled.

IV. ASYMPTOTIC ANALYSIS FOR OTHER CONFIGURATIONS

In the previous section, the relay and nodes were equipped with large antenna arrays, employing Configuration ARB. In this section, other antenna configurations are considered: Configuration aRb and Configuration ARb. Configuration aRb denotes that the numbers of antennas of node A and node B are limited, but the relay is equipped with massive MIMO.

Configuration ARb denotes that the number of antennas is limited only in node B.

A. Configuration aRb: Power Scaling Case 1

Proposition 3: For case 1, where $P_A = E_A$, $P_B = E_B$ and $P_R = E_R$, when the number of relay antennas approaches infinity, the asymptotic SINR of node A is given by

$$\gamma_{B \rightarrow A} \rightarrow \infty. \quad (37)$$

Proof: See Appendix D. ■

Compared with Configuration ARB in section III, only the relay is equipped with a large number of antennas in Configuration aRb. However, this is sufficient to eliminate the interference and noise in case 1.

B. Configuration aRb: Power Scaling for Other Cases

The results of the other cases can be obtained by following Proposition 3. In case 2, the asymptotic SINR also approaches infinity. However, in cases 3, 4, 5, 6 and 8, the asymptotic SINR is limited. In addition, as the number of relay antennas approaches infinity in case 7, the asymptotic SINR approaches zero. The results are shown in Table II, where $\Gamma_1 = E_A \Delta_{AA} + \sigma_{nA}^2 \mathbf{W}_R (\mathbf{W}_R)^H$ and $\Gamma_2 = E_A \text{tr} \{ \Omega_1 \Omega_1^H \} + E_B \text{tr} \{ \Omega_2 \Omega_2^H \} + \sigma_{nR}^2 N_{RRFR}$. Different

TABLE II: Power scaling cases 1-8 for Configuration aRb

Case	Amplification Factor β	Asymptotic SINR $\gamma_{B \rightarrow A}$
1	$N^{\frac{3}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \Omega_1 \Omega_1^H \} + E_B \text{tr} \{ \Omega_2 \Omega_2^H \}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \infty$
2	$N^{\frac{3}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_B \text{tr} \{ \Omega_2 \Omega_2^H \}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \infty$
3	$N^{\frac{3}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \Omega_1 \Omega_1^H \}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \Omega_3 \Omega_2 \Omega_2^H \Omega_3^H}{E_R \sigma_{nR}^2 \Omega_3 \Omega_3^H + \Gamma_1 E_A \text{tr} \{ \Omega_1 \Omega_1^H \}}$
4	$N^2 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \Omega_1 \Omega_1^H \} + E_B \text{tr} \{ \Omega_2 \Omega_2^H \}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \Omega_3 \Omega_2 \Omega_2^H \Omega_3^H}{[E_A \text{tr} \{ \Omega_1 \Omega_1^H \} + E_B \text{tr} \{ \Omega_2 \Omega_2^H \}] \Gamma_1}$
5	$N \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \Omega_1 \Omega_1^H \} + E_B \text{tr} \{ \Omega_2 \Omega_2^H \} + \sigma_{nR}^2 N_{RRFR}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_B \Omega_3 \Omega_2 \Omega_2^H \Omega_3^H}{\sigma_{nR}^2 \Omega_3 \Omega_3^H}$
6	$N^2 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_B \text{tr} \{ \Omega_2 \Omega_2^H \}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \Omega_3 \Omega_2 \Omega_2^H \Omega_3^H}{\sigma_{nA}^2 \mathbf{W}_R (\mathbf{W}_R)^H E_B \text{tr} \{ \Omega_2 \Omega_2^H \}}$
7	$N^2 \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \Omega_1 \Omega_1^H \}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} 0$
8	$N^{\frac{3}{2}} \beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \Omega_1 \Omega_1^H \} + E_B \text{tr} \{ \Omega_2 \Omega_2^H \} + \sigma_{nR}^2 N_{RRFR}}}$	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \Omega_3 \Omega_2 \Omega_2^H \Omega_3^H}{E_R \sigma_{nR}^2 \Omega_3 \Omega_3^H + \Gamma_2 \sigma_{nA}^2 \mathbf{W}_R (\mathbf{W}_R)^H}$

$$\gamma_{B \rightarrow A} = \frac{\beta^2 P_B \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{\beta^2 P_A \Delta_A + \beta^2 P_R \Delta_{RR} + \beta^2 \sigma_{nR}^2 \Delta_{nR} + P_A \Delta_{AA} + \sigma_{nA}^2 \mathbf{W}_R (\mathbf{W}_R)^H}, \quad (33)$$

$$\gamma_{B \rightarrow A} = \frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_A \frac{\Delta_A}{N^8} + N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}}. \quad (34)$$

$$\begin{aligned}
N^5 \beta^2 E_A \frac{\Delta_A}{N^8} &= N^5 \beta^2 E_A \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H}{N^8} \\
&= N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T}{N^2} \frac{\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T}{N^2} \left(\frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T}{N^2} \frac{\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T}{N^2} \right)^H \\
&\xrightarrow{N \rightarrow \infty} \frac{E_R E_B}{E_A \text{tr} \{ \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \} + E_B \text{tr} \{ \mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H \}} \mathbf{\Lambda}_{RA} \mathbf{0} \mathbf{\Lambda}_{RA}^H = \mathbf{0}.
\end{aligned} \tag{35}$$

$$\begin{aligned}
\gamma_{B \rightarrow A} &= \frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_A \frac{\Delta_A}{N^7} + N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^7} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^7} + E_A \frac{\Delta_{AA}}{N^2} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^2}},
\end{aligned} \tag{36}$$

TABLE III: Power scaling cases 1-8 for Configuration ARb

Case	Asymptotic SINR $\gamma_{B \rightarrow A}$	Asymptotic SINR $\gamma_{A \rightarrow B}$
1	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \mathbf{\Lambda}_{AR}^H)}{E_A \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) \sigma_{nA}^2 \mathbf{I}}$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \infty$
2	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \infty$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \infty$
3	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} 0$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \infty$
4	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} 0$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \frac{E_R E_A \text{tr}(\mathbf{\Omega}_4 \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \mathbf{\Omega}_4^H)}{\text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) [E_B \mathbf{\Omega}_{BB} + \sigma_{nB}^2 \mathbf{G}_R (\mathbf{G}_R)^H]}$
5	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \mathbf{\Lambda}_{AR}^H)}{E_R \sigma_{nR}^2 \text{tr}(\mathbf{\Lambda}_{RA} \mathbf{\Lambda}_{RA}^H) + E_A \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) \sigma_{nA}^2 \mathbf{I}}$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \infty$
6	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \mathbf{\Lambda}_{AR}^H)}{E_A \text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) \sigma_{nA}^2 \mathbf{I}}$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \frac{E_R E_A \text{tr}(\mathbf{\Omega}_4 \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \mathbf{\Omega}_4^H)}{\text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) [E_B \mathbf{\Omega}_{BB} + \sigma_{nB}^2 \mathbf{G}_R (\mathbf{G}_R)^H]}$
7	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} 0$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \frac{E_R E_A \text{tr}(\mathbf{\Omega}_4 \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \mathbf{\Omega}_4^H)}{\text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) \sigma_{nB}^2 \mathbf{G}_R (\mathbf{G}_R)^H}$
8	$\gamma_{B \rightarrow A} \xrightarrow{N \rightarrow \infty} 0$	$\gamma_{A \rightarrow B} \xrightarrow{N \rightarrow \infty} \frac{E_R E_A \text{tr}(\mathbf{\Omega}_4 \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \mathbf{\Omega}_4^H)}{\text{tr}(\mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H) \sigma_{nB}^2 \mathbf{G}_R (\mathbf{G}_R)^H}$

from Configuration ARB, the desired signal in Configuration aRb has a higher index than it does interference. Therefore, the desired signal is suppressed in case 7 and the SINR approaches zero.

C. Power Scaling for Configuration ARb

Configuration ARb denotes the situation in which the number of antennas is limited only in node B. The relay and node A are both equipped with massive MIMO. In contrast to the system configuration above, the transmission signal and SINR are not symmetric for nodes A and B. The data received by nodes A and B are (9) and (13), respectively. For node A in case 2, the power of the receiving node with a massive number of antennas is scaled. However, the same situation cannot be found in $\hat{\mathbf{y}}_B(n)$ for any cases because the receiving node is equipped with a limited number of antennas. Nonetheless, Configuration ARb has characteristics similar to those of the other two configurations. By combining the results in Proposition 2 and Proposition 3, the asymptotic performance can be obtained. The asymptotic performance results are illustrated in Table III, where $\mathbf{\Omega}_4 = (\mathbf{A}_{RB}^r)^H \mathbf{A}_{RB}^r \mathbf{\Lambda}_{RB}$, which shows that the asymptotic SINRs of $\gamma_{B \rightarrow A}$ are zero in four

cases because the SI and noise of node A are not eliminated in these cases. The node with a limited number of antennas always achieves better performance in this configuration.

V. COMPARISON BETWEEN HALF DUPLEX AND FULL DUPLEX

In this section, we introduce the HD scheme and compare it with the FD scheme. In HD mode, the information exchange requires two periods for transmitting and receiving, respectively. Therefore, it can achieve only half of the spectral efficiency compared to FD mode under ideal conditions. In the first period, the nodes transmit signals to the relay. The signal received at the relay is

$$\mathbf{y}_R(n) = \mathbf{H}_{AR} \mathbf{x}_A(n) + \mathbf{H}_{BR} \mathbf{x}_B(n) + \mathbf{n}_R(n). \tag{38}$$

Then, the relay forwards the signals to the nodes in the second period. The transmitting signal of the relay is expressed as follows:

$$\mathbf{x}_R(n) = \beta \mathbf{F}_{RFT} \mathbf{F}_{BBT} \mathbf{F}_{BBR} \mathbf{F}_{RFR} \mathbf{y}_R(n-1). \tag{39}$$

$$\gamma_{B \rightarrow A} = \frac{\beta^2 P_B \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{\beta^2 \sigma_{nR}^2 \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R)^H + \sigma_{nA}^2 \mathbf{W}_R (\mathbf{W}_R)^H}. \quad (42)$$

Because there is no SI, the amplification factor of the HD scheme is written as

$$\beta = \sqrt{\frac{P_R}{P_A \Delta_1 + P_B \Delta_2 + \sigma_{nR}^2 \Delta_3}}, \quad (40)$$

where $\Delta_1 = \text{tr}\{(\mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T\}$, $\Delta_2 = \text{tr}\{(\mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T\}$ and $\Delta_3 = \text{tr}\{(\mathbf{F}_T \mathbf{F}_R)^H \mathbf{F}_T \mathbf{F}_R\}$. The received signal at node A is derived as

$$\hat{\mathbf{y}}_A(n) = \beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \sqrt{P_B} \mathbf{G}_T \mathbf{s}_B(n - \delta) + \beta \mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{n}_R(n - \delta) + \mathbf{W}_R \mathbf{n}_A(n). \quad (41)$$

Therefore, the SINR at node A is expressed as (42).

By applying the lemmas and propositions above, the asymptotic SINR can be easily obtained for all configurations and cases.

VI. NUMERICAL RESULTS

In the following section, the numerical results of the FD mmWave TWR system are presented. First, the power scaling performance of different antenna configurations is illustrated. Then, the influence of the SI level is analyzed based on the expression of the asymptotic SINR. Finally, we compare the FD scheme with conventional HD transmission. Three types of antenna configurations are considered: Configuration ARB, Configuration aRb and Configuration ARb. In Configuration ARB, all nodes are equipped with massive MIMO. Configuration aRb denotes that the numbers of antennas of node A and node B are limited ($N_{AT} = N_{BT} = 16$), and Configuration ARb denotes that only the number of antennas of node B is limited ($N_{BT} = 16$). Assume that $E_A = 10\text{dB}$, $E_R = 20\text{dB}$, $E_B = 10\text{dB}$ and $N_{RRFR} = N_{RRFT} = N_{ARFR} = N_{ARFT} = N_{BRFR} = N_{BRFT} = 4$. Unless otherwise noted, the SI level is $\eta_{SI} = -30\text{dB}$. The numerical results in this section are based on Monte Carlo simulations.

Fig. 3: Sum spectral efficiency of cases 1-8 with Configuration ARB.

A. Power Scaling Performance

Fig. 3 shows the asymptotic performance of the FD mmWave TWR networks under different power scaling cases in Configuration ARB. In Table I, only the asymptotic SINR results of $\gamma_{B \rightarrow A}$ are presented. However, the asymptotic SINR of node B can be easily obtained because of the symmetrical structure of the network. For example, the asymptotic SINR $\gamma_{B \rightarrow A}$ in case 1 is equal to the asymptotic SINR $\gamma_{A \rightarrow B}$ in case 1 because the expressions of the two nodes are the same. The asymptotic SINR $\gamma_{B \rightarrow A}$ in case 2 denotes the case where the transmitting power is scaled at the receiving node of the system, while the asymptotic SINR $\gamma_{A \rightarrow B}$ in case 2 and $\gamma_{B \rightarrow A}$ in case 3 denote cases where the transmitting power is scaled at the transmitting node. Therefore, the sum spectral efficiency in case 2 is equal to that in case 3. Case Xa ($X = 2, 3, 6, 7$) denotes the spectral efficiency of $\gamma_{B \rightarrow A}$. Specifically, the sum of cases 2a and 3a is equal to the sum spectral efficiency in cases 2/3. In Fig. 3, we show the sum spectral efficiency in (15). As shown in Table I, all the SINRs approach infinity in theory, except in case 7; however, the numerical results show significant differences among the cases as the number of antennas varies from 16 to 512. According to (49-51, 55), the interference and noise approach zero as the number of antennas approaches infinity. The index of N , which denotes the interference elimination ability, is different under interference and noise in different cases. Compared with cases 4 and 8, cases 1 and 5 achieve much better performances with higher indexes under the SI of the node and the forwarded noise of the relay. Note that cases 1 and 5 achieve similar performances, while the power of the nodes is scaled in case 5. In addition, power scaling at one of the nodes (case 2/3 and case 6/7) leads to evident performance differences.

Fig. 4 shows the asymptotic performance of the system in Configuration aRb. The numbers of antennas of node A and node B are limited in Configuration aRb, and only the

Fig. 4: Sum spectral efficiency of cases 1-8 with Configuration aRb.

number of antennas of the relay approaches infinity. In contrast to Configuration ARB, most of the asymptotic SINRs are limited, as shown in Table II. Similarly, the asymptotic SINR of node B can be obtained through the symmetrical structure. The numerical results in Fig. 4 show similar features to the results in Configuration ARB: cases 1, 5 and 2/3 substantially outperform cases 4 and 8. As illustrated in Table II, the SI of the relay, the SI of the nodes and the noise of the nodes are eliminated by the large number of antennas, and only the noise of the relay remains in case 5. However, the SI and noise of the nodes remain in case 4, and the noise of the relay and nodes remains in case 8. In addition, the asymptotic SINR of case 7a approaches zero.

Fig. 5: Sum spectral efficiency of cases 1-8 with Configuration ARb.

Fig. 5 shows the asymptotic performance in Configuration ARb. As illustrated in Table III, the asymptotic SINRs of the nodes are no longer symmetric. In case 2, both $\gamma_{B \rightarrow A}$ and $\gamma_{A \rightarrow B}$ approach infinity. In cases 1 and 5, only $\gamma_{A \rightarrow B}$ approaches infinity, and $\gamma_{B \rightarrow A}$ is limited. $\gamma_{A \rightarrow B}$ approaches zero in cases 3, 4, 7 and 8. In addition, both $\gamma_{B \rightarrow A}$ and $\gamma_{A \rightarrow B}$ are limited in case 6. It is obvious that the performance will decrease if the power of the node given limited numbers of antennas (node B in Fig. 5) is scaled. When only the power of node A is scaled, a higher spectral efficiency can be achieved.

Fig. 6: Comparison of three antenna configurations.

Fig. 6 shows a comparison of the three configurations (Configuration ARB, Configuration aRb and Configuration ARb) illustrated above. The three best cases (cases 1, 2 and 5) are chosen from Fig. 3-5. By comparing the best performance of each configuration, we obtain $ARB > ARb > aRb$. In addition, different power scaling schemes are suitable for different configurations to achieve a high performance. In Configuration ARB, case 1 and case 5 achieve similar high spectral efficiencies, which reveals that the power of the two nodes can be scaled without performance loss. Therefore, case 5 can achieve a higher energy efficiency in Configuration ARB. To achieve a higher spectral efficiency, it is better to scale the power of the node with massive antennas (case 2) for Configuration ARb. However, case 2 results in asymmetric spectral efficiency of the two nodes. The power scaling scheme can achieve considerable performance in Configuration aRb, though only the relay is equipped with massive antennas. Most importantly, case 5 shows both a high spectral efficiency and energy efficiency in all cases. In conclusion, Configuration ARB, wherein all the nodes are equipped with large antenna arrays, can achieve a high spectral efficiency with the power of node A and B scaled down. Configuration aRb, wherein only the relay node is equipped with large antenna arrays, can achieve considerable spectral efficiency and energy efficiency with limited antennas. Configuration ARb, wherein node A and the relay are equipped with large antenna arrays, can achieve a higher spectral efficiency at the node with massive arrays.

B. Impacts of SI Levels

Fig. 7 shows the effects of the SI levels in Configuration ARB with $\eta_{SI} = -10dB$ and $\eta_{SI} = -30dB$. Compared with Fig. 3, all the spectral efficiencies decrease due to the higher SI, especially for cases 1 and 2/3. As shown in Table I, the index of N of Δ_{AA} is relatively low in those cases. Note that the SI shows little influence in case 8, although the spectral efficiency is low. In addition, the results show that the SI cancellation ability of massive MIMO is limited.

Fig. 7: Impacts of SI levels.

Fig. 8: HD sum spectral efficiency of cases 1-8 with Configuration ARB.

C. FD vs. HD

Fig. 8 shows the asymptotic performance of the conventional HD counterpart in Configuration ARB. As illustrated in Section V, there is no SI in HD transmission mode, and only noise at the relay and noise at the nodes are considered. Compared with Fig. 3, the HD system achieves half of the spectral efficiency in all cases. Thus, SI is almost eliminated in Fig. 3. Note that the additional information transmission period of the HD system leads to half the spectral efficiency loss. In Fig. 9, a comparison between FD and HD with increasing SI is presented, and the number of antennas is $N = 512$. Cases 1 and 5 show a better performance in the low-SI level region, while case 2/3 has a lower gradient. The results show that the FD system outperforms the HD system only when the SI level is under -5dB . Therefore, an additional SI cancellation design is needed to achieve better performance.

Fig. 9: FD vs HD with increasing SI levels.

VII. CONCLUSIONS

Next, we analyze an FD mmWave TWR system with massive MIMO. By considering the sparse scattering channel, the hybrid analog and digital MIMO structure and the SI of

FD transmission with the conjugate array response vector-based RF precoding matrix, the asymptotic performance of the system is analyzed. Three different antenna configurations are studied. For each configuration, the asymptotic SINRs of eight power scaling schemes are derived. Under ideal conditions, the TWR SI, forwarded noise at the relay, FD SI and noise at the nodes can be fully eliminated in Configuration ARB. The numerical results show that power scaling case 5 achieves considerable spectral efficiency, even when the power of the nodes is scaled down. However, Configuration ARB shows a poor performance because the asymptotic SINR of node A approaches zero in half of the power scaling cases. In Configuration aRB, power scaling cases 1 and 5 achieve comparatively high spectral efficiencies, even if only the relay is equipped with large number of antennas. In Configuration ARB and Configuration aRB, the power of node A and node B can be scaled down without spectral efficiency loss. Moreover, the effects of the SI levels are analyzed. The performance of all cases decreases, especially that of cases 1, 3a and 4. This is mainly due to the SI cancellation capabilities of the corresponding power scaling scheme based on the derived asymptotic SINR. In addition, the FD system is compared with its conventional HD counterpart. The results show that the SI can be fully eliminated with relatively low SI and that the FD system can achieve twice the spectral efficiency with massive MIMO under ideal conditions.

APPENDIX A PROOF OF LEMMA 1

Let $\Psi(\phi_i) = \frac{1}{\sqrt{N}} \left[1, e^{j\frac{2\pi}{\lambda} d \sin(\phi_i)}, \dots, e^{j\frac{2\pi}{\lambda} (N-1)d \sin(\phi_i)} \right]^T$. Then, we obtain [35]

$$\begin{aligned} \lim_{\substack{i \neq j \\ N \rightarrow \infty}} |\Psi^H(\phi_i) \Psi(\phi_j)| &= \lim_{\substack{i \neq j \\ N \rightarrow \infty}} \frac{1}{N} \left| \sum_{n=0}^{N-1} e^{j\frac{2\pi}{\lambda} n d [\sin(\phi_j) - \sin(\phi_i)]} \right| \\ &= \lim_{\substack{i \neq j \\ N \rightarrow \infty}} \frac{1}{N} \left| \frac{1 - e^{j\frac{2\pi}{\lambda} N d [\sin(\phi_j) - \sin(\phi_i)]}}{1 - e^{j\frac{2\pi}{\lambda} d [\sin(\phi_j) - \sin(\phi_i)]}} \right| \\ &\leq \lim_{\substack{i \neq j \\ N \rightarrow \infty}} \frac{1}{N} \frac{2}{\left| 1 - e^{j\frac{2\pi}{\lambda} d [\sin(\phi_j) - \sin(\phi_i)]} \right|} \\ &= 0. \end{aligned} \quad (43)$$

When $i = j$, it is obvious that $\lim_{\substack{i=j \\ N \rightarrow \infty}} |\Psi^H(\phi_i) \Psi(\phi_j)| =$

$$\lim_{\substack{i=j \\ N \rightarrow \infty}} \frac{1}{N} \left| \sum_{n=0}^{N-1} e^{j\frac{2\pi}{\lambda} n d 0} \right| = 1.$$

APPENDIX B PROOF OF PROPOSITION 1

According to (19) and Lemma 1, we can obtain (44), where orthogonality and the special case of $i = j$ in Lemma 1 are utilized. If the array response matrix from another channel matrix is applied, all the vectors are orthogonal, and we can obtain (45).

$$\begin{aligned}
\lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{H} \mathbf{A}^t}{N^2} &= \lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{A}^r \mathbf{\Lambda} (\mathbf{A}^t)^H \mathbf{A}^t}{N^2} = \lim_{N \rightarrow \infty} \frac{[\Psi_1^r \ \Psi_2^r \ \cdots \ \Psi_{KL}^r]^H [\Psi_1^r \ \Psi_2^r \ \cdots \ \Psi_{KL}^r]}{N} \mathbf{\Lambda} \frac{(\mathbf{A}^t)^H \mathbf{A}^t}{N} \\
&= \lim_{N \rightarrow \infty} \frac{\begin{bmatrix} (\Psi_1^r)^H \Psi_1^r & (\Psi_1^r)^H \Psi_2^r & \cdots & (\Psi_1^r)^H \Psi_{KL}^r \\ \vdots & \vdots & \ddots & \vdots \\ (\Psi_{KL}^r)^H \Psi_1^r & (\Psi_{KL}^r)^H \Psi_2^r & \cdots & (\Psi_{KL}^r)^H \Psi_{KL}^r \end{bmatrix}}{N} \mathbf{\Lambda} \frac{(\mathbf{A}^t)^H \mathbf{A}^t}{N} = \lim_{N \rightarrow \infty} \frac{N \mathbf{I}}{N} \mathbf{\Lambda} \frac{(\mathbf{A}^t)^H \mathbf{A}^t}{N} = \mathbf{\Lambda},
\end{aligned} \quad (44)$$

$$\begin{aligned}
\lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{H} \tilde{\mathbf{A}}^t}{N^2} &= \lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{A}^r \mathbf{\Lambda} (\mathbf{A}^t)^H \tilde{\mathbf{A}}^t}{N^2} = \lim_{N \rightarrow \infty} \frac{(\mathbf{A}^r)^H \mathbf{A}^r}{N} \mathbf{\Lambda} \frac{(\mathbf{A}^t)^H \tilde{\mathbf{A}}^t}{N} \\
&= \lim_{N \rightarrow \infty} \mathbf{I} \mathbf{\Lambda} \frac{\mathbf{0}}{N} = \mathbf{0}.
\end{aligned} \quad (45)$$

APPENDIX C PROOF OF PROPOSITION 2

Upon substituting $P_A = E_A$, $P_B = E_B$, and $P_R = E_R$ into (5), the amplification factor can be written as follows:

$$N^{\frac{5}{2}} \beta = \sqrt{\frac{E_R}{E_A \frac{\Delta_1}{N^5} + E_B \frac{\Delta_2}{N^5} + E_R \frac{\Delta_3}{N^5} + \sigma_{nR}^2 \frac{\Delta_4}{N^5}}}, \quad (46)$$

where Δ_1 , Δ_2 , Δ_3 and Δ_4 can be rewritten as $\Delta_1 = \text{tr} \left\{ (\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T) (\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H \mathbf{F}_T^H \mathbf{F}_T \right\}$, $\Delta_2 = \text{tr} \left\{ (\mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T) (\mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H \mathbf{F}_T^H \mathbf{F}_T \right\}$, $\Delta_3 = \text{tr} \left\{ (\mathbf{F}_R \mathbf{H}_{RR}) (\mathbf{F}_R \mathbf{H}_{RR})^H \mathbf{F}_T^H \mathbf{F}_T \right\}$ and $\Delta_4 = \text{tr} \left\{ \mathbf{F}_R \mathbf{F}_R^H \mathbf{F}_T^H \mathbf{F}_T \right\}$, respectively. According to Proposition 1, (19), (22) and (23), we can obtain

$$\frac{\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T}{N^2} = \left[\frac{(\mathbf{A}_{AR}^r)^H \mathbf{A}_{AR}^r}{N} \mathbf{\Lambda}_{AR} \frac{(\mathbf{A}_{AR}^t)^H \mathbf{A}_{AR}^t}{N} \right] \xrightarrow{N \rightarrow \infty} \begin{bmatrix} \mathbf{\Lambda}_{AR} \\ \mathbf{0} \end{bmatrix}, \quad (47)$$

$$\begin{aligned}
\frac{\mathbf{F}_T^H \mathbf{F}_T}{N} &= \frac{\begin{bmatrix} (\mathbf{A}_{RB}^t)^H \\ (\mathbf{A}_{RA}^t)^H \end{bmatrix} \begin{bmatrix} \mathbf{A}_{RB}^t & \mathbf{A}_{RA}^t \end{bmatrix}}{N} \\
&= \begin{bmatrix} \frac{(\mathbf{A}_{RB}^t)^H \mathbf{A}_{RB}^t}{N} & \frac{(\mathbf{A}_{RB}^t)^H \mathbf{A}_{RA}^t}{N} \\ \frac{(\mathbf{A}_{RA}^t)^H \mathbf{A}_{RB}^t}{N} & \frac{(\mathbf{A}_{RA}^t)^H \mathbf{A}_{RA}^t}{N} \end{bmatrix} \xrightarrow{N \rightarrow \infty} \mathbf{I}.
\end{aligned} \quad (48)$$

By substituting (47) and (48) into Δ_1 , we can obtain

$$\begin{aligned}
\frac{\Delta_1}{N^5} &= \text{tr} \left\{ \frac{(\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)}{N^2} \frac{(\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H}{N^2} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N} \right\} \\
&\xrightarrow{N \rightarrow \infty} \text{tr} \left\{ \begin{bmatrix} \mathbf{\Lambda}_{AR} \\ \mathbf{0} \end{bmatrix} \begin{bmatrix} \mathbf{\Lambda}_{AR}^H & \mathbf{0} \end{bmatrix} \mathbf{I} \right\} = \text{tr} \{ \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \}.
\end{aligned} \quad (49)$$

Similarly, $\frac{\Delta_2}{N^5}$ and $\frac{\Delta_4}{N^5}$ can be obtained as

$$\begin{aligned}
\frac{\Delta_2}{N^5} &= \text{tr} \left\{ \frac{(\mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)}{N^2} \frac{(\mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^2} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N} \right\} \\
&\xrightarrow{N \rightarrow \infty} \text{tr} \left\{ \begin{bmatrix} \mathbf{\Lambda}_{BR} \\ \mathbf{0} \end{bmatrix} \begin{bmatrix} \mathbf{\Lambda}_{BR}^H & \mathbf{0} \end{bmatrix} \mathbf{I} \right\} = \text{tr} \{ \mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H \},
\end{aligned} \quad (50)$$

$$\frac{\Delta_4}{N^5} = \text{tr} \left\{ \frac{\mathbf{F}_R \mathbf{F}_R^H}{N} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N} \frac{1}{N^3} \right\} \xrightarrow{N \rightarrow \infty} \text{tr} \{ \mathbf{I} \} = 0, \quad (51)$$

respectively. Because the SI channel $\mathbf{H}_{RR}$, which contains LOS and NLOS SI, is different, the result of $\frac{\Delta_3}{N^5}$ is derived as follows. First, the expression of Δ_3 is expanded as

$$\begin{aligned}
\frac{\Delta_3}{N^5} &= \text{tr} \left\{ \frac{\mathbf{F}_R \mathbf{H}_{NLOS} (\mathbf{F}_R \mathbf{H}_{NLOS})^H}{N^3} + \frac{\mathbf{F}_R \mathbf{H}_{NLOS} (\mathbf{F}_R \mathbf{H}_{LOS})^H}{N^3} \right. \\
&\quad \left. + \frac{\mathbf{F}_R \mathbf{H}_{LOS} (\mathbf{F}_R \mathbf{H}_{NLOS})^H}{N^3} + \frac{\mathbf{F}_R \mathbf{H}_{LOS} (\mathbf{F}_R \mathbf{H}_{LOS})^H}{N^3} \right\} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N^2}
\end{aligned} \quad (52)$$

Similarly, according to Proposition 1, the limit of the first term is written as

$$\begin{aligned}
&\frac{\mathbf{F}_R \mathbf{H}_{NLOS} (\mathbf{F}_R \mathbf{H}_{NLOS})^H}{N^3} \\
&= \frac{\mathbf{F}_R \mathbf{A}_{NLOS}^r}{N} \mathbf{\Lambda}_{NLOS} \frac{(\mathbf{A}_{NLOS}^t)^H \mathbf{A}_{NLOS}^t}{N} \mathbf{\Lambda}_{NLOS}^H \frac{(\mathbf{A}_{NLOS}^r)^H \mathbf{F}_R^H}{N} \\
&\xrightarrow{N \rightarrow \infty} \frac{\mathbf{0}}{N} \mathbf{\Lambda}_{NLOS} \mathbf{I} \mathbf{\Lambda}_{NLOS}^H \frac{\mathbf{0}}{N} \rightarrow \mathbf{0}.
\end{aligned} \quad (53)$$

In addition, combined with Lemma 2, we can obtain

$$\begin{aligned}
&\frac{\mathbf{F}_R \mathbf{H}_{NLOS} (\mathbf{F}_R \mathbf{H}_{LOS})^H}{N^3} \\
&= \frac{\mathbf{F}_R \mathbf{A}_{NLOS}^r}{N} \mathbf{\Lambda}_{NLOS} \frac{(\mathbf{A}_{NLOS}^t)^H \mathbf{H}_{LOS}^H \mathbf{F}_R^H}{N} \xrightarrow{N \rightarrow \infty} 0, \\
&\frac{\mathbf{F}_R \mathbf{H}_{LOS} (\mathbf{F}_R \mathbf{H}_{NLOS})^H}{N^3} \xrightarrow{N \rightarrow \infty} 0, \\
&\frac{\mathbf{F}_R \mathbf{H}_{LOS} (\mathbf{F}_R \mathbf{H}_{LOS})^H}{N^3} \xrightarrow{N \rightarrow \infty} 0.
\end{aligned} \quad (54)$$

Therefore, the limit of $\frac{\Delta_3}{N^5}$ is

$$\frac{\Delta_3}{N^5} = \text{tr} \left\{ \frac{(\mathbf{F}_R \mathbf{H}_{RR})(\mathbf{F}_R \mathbf{H}_{RR})^H \mathbf{F}_T^H \mathbf{F}_T}{N^3} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N^2} \right\} \xrightarrow{N \rightarrow \infty} 0. \quad (55)$$

The limit of the amplification factor is

$$N^{\frac{5}{2}}\beta = \sqrt{\frac{E_R}{E_A \frac{\Delta_1}{N^5} + E_B \frac{\Delta_2}{N^5} + E_R \frac{\Delta_3}{N^5} + \sigma_{nR}^2 \frac{\Delta_4}{N^5}}} \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \} + E_B \text{tr} \{ \mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H \}}}. \quad (56)$$

Now, we will derive the asymptotic SINR of node A. The limit of $\gamma_{B \rightarrow A}$ is written as (57).

Additionally, according to Proposition 1, the numerator is derived as follows:

$$N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8} \xrightarrow{N \rightarrow \infty} \frac{E_R E_B \mathbf{\Lambda}_{RA} \mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H \mathbf{\Lambda}_{RA}^H}{E_A \text{tr} \{ \mathbf{\Lambda}_{AR} \mathbf{\Lambda}_{AR}^H \} + E_B \text{tr} \{ \mathbf{\Lambda}_{BR} \mathbf{\Lambda}_{BR}^H \}}. \quad (58)$$

Considering Lemma 2, the following limit can be obtained:

$$\begin{aligned} \frac{\Delta_{RR}}{N^8} &= \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T}{N^2} \frac{\mathbf{F}_R \mathbf{H}_{RR}}{N^2} \left(\frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T}{N^2} \frac{\mathbf{F}_R \mathbf{H}_{RR}}{N^2} \right)^H \xrightarrow{N \rightarrow \infty} 0, \\ \frac{\Delta_{nR}}{N^8} &= \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T}{N^2} \frac{\mathbf{F}_R \mathbf{F}_R^H}{N} \left(\frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T}{N^2} \right)^H \frac{\mathbf{I}}{N^3} \xrightarrow{N \rightarrow \infty} 0, \\ \frac{\Delta_{AA}}{N^3} &= \frac{\mathbf{W}_R \mathbf{H}_{AA} \mathbf{W}_T}{N} \left(\frac{\mathbf{W}_R \mathbf{H}_{AA} \mathbf{W}_T}{N} \right)^H \frac{\mathbf{I}}{N} \xrightarrow{N \rightarrow \infty} 0, \\ \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3} &= \frac{\mathbf{W}_R \mathbf{W}_R^H}{N} \frac{\mathbf{I}}{N^2} \xrightarrow{N \rightarrow \infty} 0. \end{aligned} \quad (59)$$

As a result, we can obtain $\gamma_{B \rightarrow A} \rightarrow \infty$.

APPENDIX D PROOF OF PROPOSITION 3

The amplification factor can be expressed as

$$N^{\frac{3}{2}}\beta = \sqrt{\frac{E_R}{E_A \frac{\Delta_1}{N^3} + E_B \frac{\Delta_2}{N^3} + E_R \frac{\Delta_3}{N^3} + \sigma_{nR}^2 \frac{\Delta_4}{N^3}}}. \quad (60)$$

Because only the number of relay antennas approaches infinity, the limit of $\Delta_1 \sim \Delta_4$ can be derived as

$$\begin{aligned} \frac{\Delta_1}{N^3} &= \text{tr} \left\{ \frac{(\mathbf{A}_{AR}^r)^H \mathbf{A}_{AR}^r}{N} \mathbf{\Lambda}_{AR} (\mathbf{A}_{AR}^t)^H \mathbf{A}_{AR}^t \frac{(\mathbf{F}_R \mathbf{H}_{AR} \mathbf{W}_T)^H}{N} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N} \right\} \\ &\xrightarrow{N \rightarrow \infty} \text{tr} \left\{ \mathbf{\Lambda}_{AR} (\mathbf{A}_{AR}^t)^H \mathbf{A}_{AR}^t \left[\mathbf{\Lambda}_{AR} (\mathbf{A}_{AR}^t)^H \mathbf{A}_{AR}^t \right]^H \right\} \\ &= \text{tr} \{ \mathbf{\Omega}_1 \mathbf{\Omega}_1^H \}, \end{aligned} \quad (61)$$

where $\mathbf{\Omega}_1 = \mathbf{\Lambda}_{AR} (\mathbf{A}_{AR}^t)^H \mathbf{A}_{AR}^t$. Similarly, $\frac{\Delta_2}{N^5}$, $\frac{\Delta_3}{N^3}$ and $\frac{\Delta_4}{N^5}$ can be obtained as

$$\begin{aligned} \frac{\Delta_2}{N^3} &= \text{tr} \left\{ \frac{(\mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)}{N} \frac{(\mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N} \right\} \\ &\xrightarrow{N \rightarrow \infty} \text{tr} \left\{ \mathbf{\Lambda}_{BR} (\mathbf{A}_{BR}^t)^H \mathbf{A}_{BR}^t \left[\mathbf{\Lambda}_{BR} (\mathbf{A}_{BR}^t)^H \mathbf{A}_{BR}^t \right]^H \right\} \\ &= \text{tr} \{ \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \}, \end{aligned} \quad (62)$$

$$\frac{\Delta_3}{N^3} = \text{tr} \left\{ \frac{\mathbf{F}_R \mathbf{H}_{RR} (\mathbf{F}_R \mathbf{H}_{RR})^H}{N^2} \frac{\mathbf{F}_T^H \mathbf{F}_T}{N} \right\} \xrightarrow{N \rightarrow \infty} 0, \quad (63)$$

$$\frac{\Delta_4}{N^3} = \text{tr} \left\{ \frac{\mathbf{F}_R \mathbf{F}_R^H}{N} \frac{\mathbf{F}_T \mathbf{F}_T^H}{N} \frac{1}{N} \right\} \xrightarrow{N \rightarrow \infty} 0,$$

where $\mathbf{\Omega}_2 = \mathbf{\Lambda}_{BR} (\mathbf{A}_{BR}^t)^H \mathbf{A}_{BR}^t$. Therefore, the limit of the amplification factor is

$$N^{\frac{3}{2}}\beta \xrightarrow{N \rightarrow \infty} \sqrt{\frac{E_R}{E_A \text{tr} \{ \mathbf{\Omega}_1 \mathbf{\Omega}_1^H \} + E_B \text{tr} \{ \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \}}}. \quad (64)$$

Now, we will derive the asymptotic SINR of node A. The limit of $\gamma_{B \rightarrow A}$ is written as (65).

According to Proposition 1, the numerator is derived as follows:

$$\begin{aligned} N^3 \beta^2 E_B \mathbf{W}_R \frac{\mathbf{H}_{RA} \mathbf{F}_T}{N} \frac{\mathbf{F}_R \mathbf{H}_{BR}}{N} \mathbf{G}_T \left(\mathbf{W}_R \frac{\mathbf{H}_{RA} \mathbf{F}_T}{N} \frac{\mathbf{F}_R \mathbf{H}_{BR}}{N} \mathbf{G}_T \right)^H \\ \xrightarrow{N \rightarrow \infty} \frac{E_R \mathbf{\Omega}_3 \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \mathbf{\Omega}_3^H}{E_A \text{tr} \{ \mathbf{\Omega}_1 \mathbf{\Omega}_1^H \} + E_B \text{tr} \{ \mathbf{\Omega}_2 \mathbf{\Omega}_2^H \}}, \end{aligned} \quad (66)$$

where $\mathbf{\Omega}_3 = (\mathbf{A}_{RA}^r)^H \mathbf{A}_{RA}^r \mathbf{\Lambda}_{RA}$. Similarly, the following limit

$$\gamma_{B \rightarrow A} = \frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}}. \quad (57)$$

$$\gamma_{B \rightarrow A} = \frac{N^5 \beta^2 E_B \frac{\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T (\mathbf{W}_R \mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{BR} \mathbf{G}_T)^H}{N^8}}{N^5 \beta^2 E_R \frac{\Delta_{RR}}{N^8} + N^5 \beta^2 \sigma_{nR}^2 \frac{\Delta_{nR}}{N^8} + E_A \frac{\Delta_{AA}}{N^3} + \sigma_{nA}^2 \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N^3}}. \quad (65)$$

can be obtained:

$$\begin{aligned}\frac{\Delta_{RR}}{N^4} &= \mathbf{W}_R \frac{\mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR}}{N^2} \left(\mathbf{W}_R \frac{\mathbf{H}_{RA} \mathbf{F}_T \mathbf{F}_R \mathbf{H}_{RR}}{N^2} \right)^H \xrightarrow{N \rightarrow \infty} \mathbf{0}, \\ \frac{\Delta_{nR}}{N^4} &= \mathbf{W}_R \frac{\mathbf{H}_{RA} \mathbf{F}_T}{N} \frac{\mathbf{F}_R \mathbf{F}_R^H}{N} \left(\mathbf{W}_R \frac{\mathbf{H}_{RA} \mathbf{F}_T}{N} \right)^H \frac{\mathbf{I}}{N} \xrightarrow{N \rightarrow \infty} \mathbf{0}.\end{aligned}\quad (67)$$

Because node A has a limited number of antennas, the SI and noise at node A approach zero as

$$\begin{aligned}\frac{\Delta_{AA}}{N} &= \frac{\mathbf{W}_R \mathbf{H}_{AA} \mathbf{W}_T (\mathbf{W}_R \mathbf{H}_{AA} \mathbf{W}_T)^H}{N} \xrightarrow{N \rightarrow \infty} \mathbf{0}, \\ &\quad \frac{\mathbf{W}_R (\mathbf{W}_R)^H}{N} \xrightarrow{N \rightarrow \infty} \mathbf{0}.\end{aligned}\quad (68)$$

As a result, we can obtain $\gamma_{B \rightarrow A} \rightarrow \infty$.

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