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Joint Space-Frequency Block Codes and Signal Alignment for Heterogeneous Networks

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ABSTRACT In this paper, we propose a new diversity-oriented space-frequency block codes (SFBC) and signal alignment (SA) enabled physical network coding (PNC) method for the uplink of heterogeneous networks. The proposed joint Dual-SFBC with SA-PNC design substantially reduces interference and enables connecting a larger number of users when compared with methods adopting interference alignment (IA) or PNC. The main motivation behind the dual SFBC and SA-PNC design is that it allows the efficient coexistence of macro and small cells without any inter-system channel information requirements. Numerical results also verify that the proposed method outperforms the existing SA-PNC static method without any additional information exchange requirement between the two systems while achieving the main benefits of IA and SA-PNC coordinated methods recently proposed.

INDEX TERMS Signal alignment (SA), space frequency block codes (SFBC), physical-layer network coding (PNC), heterogeneous networks (HetNets), small-cell system, macro-cell system.

I. INTRODUCTION

Supporting bandwidth-intensive applications, combined with an enormous increase in data rates and connection density, increasingly burdens present 4G LTE cellular networks. In order to support 1000 times mobile data traffic loading by 2020 [1], [2], emerging 5G technologies are under development to achieve these targets in terms of higher data rates and increased network capacity such as Device to Device (D2D) communications, Internet of Things (IoT), Millimeter Waves (mmW), Massive-MIMO and Machine to Machine (M2M) communications [3]–[6]. Another interesting approach to handle the wireless traffic explosion in future wireless networks is the deployment of a large number of small-cells giving rise to the concept of Heterogeneous Networks (HetNets) [7], [8]. HetNets typically comprise small-cells overlaid within the coverage area of macro-cells, where small-cells are low-power access points (APs) capable of providing improved network capacity without compromising network coverage. Moreover, small-cells can also offload traffic from macro cellular system to improve spectrum usage efficiency [9]. On the other hand, the deployment of small-cells in conjunction with legacy macro-cells substantially complicates interference management in HetNets and calls for coordinated operation between macro and small cells for interference mitigation [10]. Several interference management challenges for 5G HetNets are addressed in [11], where the key challenges for interference management in multi-tier 5G networks are due to the heterogeneity and dense deployment of wireless devices. Moreover, cooperation among different tiers and direct communication among users (i.e. D2D communications) may further complicate the interference mitigation process.

One effective approach to tackle interference in HetNets characterized by high signal to noise ratio (SNR) is interference alignment (IA) [12]–[15]. IA is a degree of freedom (DoF) optimal approach that maximizes system capacity by superimposing, i.e. aligning, interference signals, which means that it can reach the capacity of interference networks at very high SNR. Different schemes based on IA have been proposed to address the interference management problem of HetNets [13], [16]. IA can be achieved either in time, frequency or space dimensions [17]. Due to the wide applications of MIMO technique, the most popular is the IA in space dimensions. Due to its promising performance in interference management, different IA schemes have been designed for several network topologies [12], [17]. The concept of IA has already been used for cellular networks, cognitive radio (CR) networks, relay networks, HetNets, D2D communications and massive MIMO in 5G systems [12].

Another effective approach to deal with interference in HetNets is the physical network coding (PNC) [18]. PNC was first proposed in [19] as a way to exploit interference as a useful signal and PNC has the potential to achieve 100% throughput gain when compared to traditional transmission schemes [20]. For multi-node systems, PNC enables the joint detection and decoding of signals received at relay nodes and simultaneously forwards a linear combination over multiple dimensions to improve system performance. The authors in [21] proposed a linear MIMO transceiver design for HetNets using PNC. A coordinated joint transmitter and receiver design is proposed in order to mitigate the interference between two-way relaying channel for small-cell and uplink channel in macro-cell.

A. RELATED WORK

As compare to IA, in signal alignment (SA) [19], [22] signals are superimposed rather than interference and receivers decode the linear sum of transmitted symbols. Nevertheless, the major constraint in the practical application of IA and SA is the requirement of perfect channel state information (CSI), where accurate CSI must be available at all the nodes to enable applying IA/SA, entailing high transmission overhead [17]. Subsequently, several research efforts have been made using IA/SA to reduce the feedback requirements without degrading system performance [14], [15]. In [14], several IA based cognitive communication techniques for the uplink of HetNets were proposed. Namely, coordinated, static, and uncoordinated IA techniques were proposed to accommodate varying levels of feedback requirements, with the coordinated method providing optimal performance at the expense of extensive feedback requirements and the static method completely eliminating the need for information exchange by sacrificing the system performance. To overcome the shortcomings of the coordinated and static methods proposed in [14], a one-bit method is proposed in [15] to provide near-optimal performance with limited overheads.

Another interesting approach to further enhance interference management in HetNets jointly applies SA and PNC [23], [24]. Joint signal and interference alignment schemes, in which uplink and downlink transmissions are completed over two time slots, has been studied in [23] and a practical SA enabled PNC for the uplink of MIMO networks, where SA based precoding is performed at the transmitters followed by PNC at the relay node, is proposed in [24]. In [25] and [26], we proposed a joint SA and PNC design characterized by limited information exchange for the uplink of HetNets. The proposed methods were shown to be capable of efficiently removing cross-tier interference while increasing the overall system throughput by enabling the connection of a larger number of users when compared to the methods proposed in [14] and [15].

B. CONTRIBUTIONS

In this paper, we propose a new diversity-oriented SA and PNC scheme, which combines dual SFBC [29] with SA-PNC, for the uplink of HetNets. To the best of our knowledge, joint dual-SFBC and SA-PNC, in the context of HetNets has not been addressed in the literature. In [25] and [26] joint SA and PNC based methods for single carrier systems were proposed. These schemes require knowledge at the small-cell user equipment(SUEs) of the macro base station (MBS) and macro user equipment (MUE) channels and this information needs to be exchanged between the two systems. This requirement adds to the complexity of practical systems and creates further signaling overhead. In [27], we proposed a specific case of dual Alamouti scheme to avoid the requirement of inter-system channel information exchange and design a solution that combines it with SA enabled PNC for the multi-carrier systems. In this work, we extended the methods proposed in [27] for a generalized dual-SFBC and SA-PNC scheme considering different group sizes for the normalization constant in order to achieve a balance between performance and digital link requirements.

This new scheme eliminates the main drawback of the previous ones, because the small-cells just need to sense if macro-cell system is using a scheme based on dual-SFBC. This requirement can also be fulfilled using the static scheme of [25] and [26] but diversity is lost and if one wants to preserve it, the inter-system channel information must be exchanged. The dual-SFBC imposes some structure in the data transmission which allows to align the interfering signals without the need of inter-system channel information while preserving the diversity. By combining it with SA-PNC allows to achieve an efficient inter-tier interference management.

The rest of the paper is organized as follows: In section II, we introduce the system model for the macro and small-cell systems. Section III describes the design of the SA-based precoder, subspace and filter matrix based on dual-SFBC scheme. In Section IV, the simulation results and performance comparison of the proposed method with existing methods are presented. Finally, conclusions are provided in Section V.

Notations: Bold upper case, bold lower case letters denote matrices and vectors, respectively. The operators $(\cdot)^T$, $(\cdot)^H$ and $(\cdot)^\dagger$ represent the transpose, Hermitian transpose and pseudo-inverse of a matrix. null($\mathbf{A}$) denotes a matrix which span the null-space of matrix $\mathbf{A}$ and $\mathbf{A} = \text{bdiag}(\mathbf{A}_1, \mathbf{A}_2)$ a block diagonal matrix with entries $\mathbf{A}_1$ and $\mathbf{A}_2$. $\mathbf{I}$ denotes the identity. Let $\mathbf{A}$ be a matrix then $\mathbf{A}_{(i)}$ denotes the row i of $\mathbf{A}$. For a complex number a, $\Re(a)$ and $\Im(a)$ denote its real and imaginary parts, respectively. For a complex vector $\mathbf{x}$ the $(\cdot)^R$

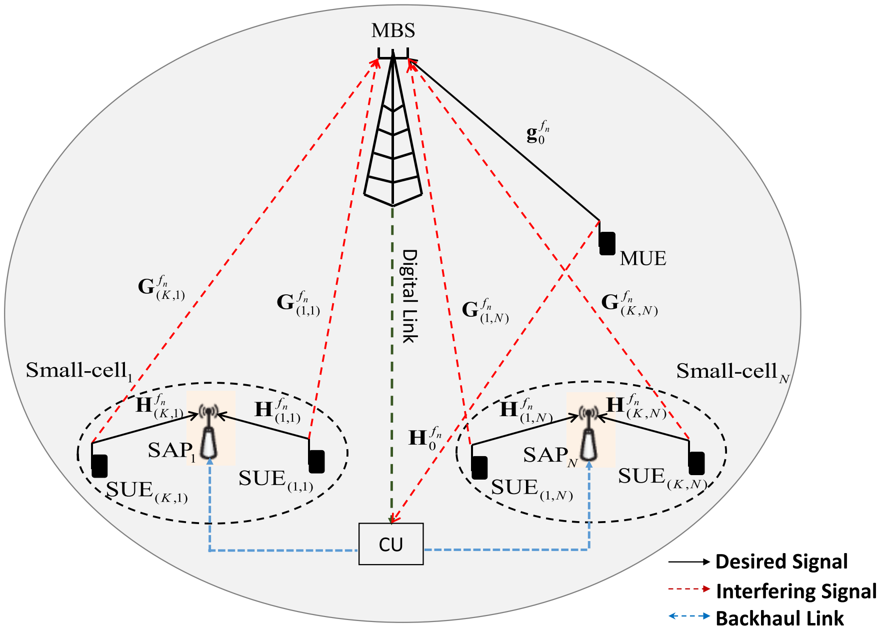

FIGURE 1. System Model: Small-cells within coverage area of macro-cell.

operator is defined by $\mathbf{A}^R = [\Re(\mathbf{A})^T, \Im(\mathbf{A})^T]^T$ and for a complex matrix $\mathbf{A}$ by $\mathbf{A}^R = \begin{bmatrix} \Re(\mathbf{A}) & -\Im(\mathbf{A}) \\ \Im(\mathbf{A}) & \Re(\mathbf{A}) \end{bmatrix}$.

II. SYSTEM MODEL

We consider the uplink of a heterogeneous network with N Small-cells overlaid within the boundaries of a macro-cell utilizing the same set of frequency bands, as shown in Figure 1. Each small-cell access point (SAP) connects K small-cell user equipment (SUEs), with SAPs connected to a central unit (CU) through a backhaul network that transport PHY-layer signals and enables joint processing of received signals. The digital link is a logical channel that connects the two systems (macro and small) and it is used to exchange the information required for the efficient inter-working. This can be considered as the part of the backhaul and our aim is to reduce the signaling information to minimize the traffic overhead in the backhaul. OFDM transmission with N_c subcarriers is assumed and the transmission power per subcarrier for the macro base station (MBS) and SAPs is constrained to P_m and P_s, respectively.

We consider that the MBS serves the macro user equipment denoted by MUE. For the sake of simplicity and without loss of generality, we consider a single user per MBS and adding more macro-cell users will not affect the overall performance, since the interference from the small-cells can be mitigated at the MBS. For the small-cell system, SAP p serves K SUEs denominated by $SUE_{(u,p)}$, $u = \{1, 2 \ldots, K\}$.

A. MACRO-CELL SIGNAL MODEL

For the macro-cell system, we assume that the MUE has a single antenna and the MBS has M_m antennas. The received signal ($\mathbf{y}_m^{f_n}$) at the MBS on subcarrier f_n is expressed as

$$\mathbf{y}_m^{f_n} = \underbrace{\mathbf{g}_0^{f_n} x_0^{f_n}}_{\text{Desired signal}} + \underbrace{\sum_{p=1}^{N}\sum_{u=1}^{K} \mathbf{G}_{(u,p)}^{f_n} \mathbf{x}_{(u,p)}^{f_n}}_{\text{Interference}} + \mathbf{n}_m^{f_n}, \tag{1}$$

where $\mathbf{g}_0^{f_n} \in \mathbb{C}^{M_m}$, $x_0^{f_n} \in \mathbb{C}$, $\mathbf{G}_{(u,p)}^{f_n} \in \mathbb{C}^{M_m \times M_s}$, $\mathbf{x}_{(u,p)}^{f_n} \in \mathbb{C}^{M_s}$ and $\mathbf{n}_m^{f_n} \in \mathbb{C}^{M_m}$, denote the channel between MUE and MBS, the transmitted signal at MUE, the channel between SUE$_{(u,p)}$ and MBS, the transmitted signal at SUEu of small-cell p(SUE$_{(u,p)}$), and the zero mean white Gaussian noise with variance σ^2, respectively. As shown in Figure 1, the received signals of all N small-cells is jointly processed by the CU and considered as a single small-cell with KN SUEs. In this context, equation (1) can be rewritten as

$$\mathbf{y}_m^{f_n} = \underbrace{\mathbf{g}_0^{f_n} x_0^{f_n}}_{\text{Desired signal}} + \underbrace{\sum_{k=1}^{NK} \mathbf{G}_k^{f_n} \mathbf{x}_k^{f_n}}_{\text{Interference}} + \mathbf{n}_m^{f_n}. \tag{2}$$

In the following, we assume that the coherence bandwidth of the macro channel is such that it can be considered as frequency flat over a block of adjacent F subcarriers. Let $f_1, \ldots f_F$, be a block of F adjacent subcarriers then we assume that $\mathbf{g}_0^{f_n} = \mathbf{g}_0$ for $f_n = f_1, \ldots f_F$. However, the channels of different blocks are considered to be independent. Therefore, (2) over a single block is given by [28]

$$\mathbf{y}_m = \underbrace{\mathbf{G}_0 \mathbf{x}_0}_{\text{Desired signal}} + \underbrace{\sum_{k=1}^{NK} \mathbf{G}_k \mathbf{x}_k}_{\text{Interference}} + \mathbf{n}_m, \tag{3}$$

where $\mathbf{y}_m = [(\mathbf{y}_m^{f_1})^T, \ldots, (\mathbf{y}_m^{f_F})^T]^T$ is the received signal at the MBS, $\mathbf{x}_k = [(\mathbf{x}_k^{f_1})^T, \ldots, (\mathbf{x}_k^{f_F})^T]^T$ denotes the transmitted signals at the SUE_k, $\mathbf{G}_0 = \mathrm{bdiag}[\mathbf{g}_0, \ldots, \mathbf{g}_0]$ denotes the macro channel, $\mathbf{G}_k = \mathrm{bdiag}[\mathbf{G}_k^{f_1}, \ldots, \mathbf{G}_k^{f_F}]$ for $k \in \{1, 2, \ldots, NK\}$ denotes the channel k in block format and $\mathbf{n}_m = [(\mathbf{n}_m^{f_1})^T, \ldots, (\mathbf{n}_m^{f_F})^T]^T$ is the zero-mean white Gaussian noise with variance σ^2.

The macro-cell system uses a dual-SFBC, denoted by $\mathcal{C}$ in the following. As the real and imaginary parts of each symbol may each modulate a separate codeword of code $\mathcal{C}$, in the following, we will consider the real representation of (3) given by

$$\mathbf{y}_m^R = \underbrace{\mathbf{G}_0^R \mathbf{x}_0^R}_{\text{Desired signal}} + \underbrace{\sum_{k=1}^{NK} \mathbf{G}_k^R \mathbf{x}_k^R}_{\text{Interference}} + \mathbf{n}_m^R. \tag{4}$$

The transmitted signal at the MUE over one block is given by

$$\mathbf{x}_0^R = \gamma_0 \mathbf{V}_0 \left[\mathcal{C}, \mathbf{G}_0^R\right] \mathbf{d}_0^R, \tag{5}$$

where $\mathbf{d}_0 \in \mathbb{C}^F$ denotes the data vector drawn from an M-QAM constellation and $\mathbf{d}_0^R \in \mathbb{R}^{S_m}$ with $S_m = 2F$ denotes its real representation. The decoded signal, after the filter matrix $\mathbf{Q}_m[\mathcal{C}] \in \mathbb{R}^{S_m \times 2FM_m}$ is

$$\tilde{\mathbf{d}}_0^R = \mathbf{Q}_m[\mathcal{C}] \mathbf{y}_m^R. \tag{6}$$

The precoder is a function of the dual-SFBC $\mathcal{C}$ and macro channel $\mathbf{G}_0^R$. In the following, we consider the zero-forcing precoder, which is given by

$$\mathbf{V}_0 \left[\mathcal{C}, \mathbf{G}_0^R\right] = \left(\mathbf{Q}_m[\mathcal{C}] \mathbf{G}_0^R\right)^{\dagger} \tag{7}$$

but other precoders may be considered. γ_0 is a normalization constant used to enforce the transmission power at the MUE.

The matrices $\mathbf{V}_0[\mathcal{C}, \mathbf{G}_0^R] \in \mathbb{R}^{S_m \times S_m}$ and $\mathbf{Q}_m[\mathcal{C}] \in \mathbb{R}^{S_m \times 2FM_m}$ are used to encode and decode the data vector $\mathbf{d}_0^R \in \mathbb{R}^{S_m}$. Please note that for a dual-SFBC the encoder and decoder exchange roles. Namely, $\mathbf{Q}_m[\mathcal{C}]$ is the generator of the dual of code $\mathcal{C}$. For example, for the dual-Alamouti code the matrix $\mathbf{Q}_m[\mathcal{C}]$ is the generator matrix of the Alamouti code. Furthermore, the filter matrix $\mathbf{Q}_m[\mathcal{C}]$ is only a function of code $\mathcal{C}$ and independent of the channel. More details about concrete examples for matrix $\mathbf{Q}_m[\mathcal{C}]$ will be given in section III.

B. SMALL-CELL SIGNAL MODEL

For the small-cell system, we assume that the SUEs have M_s and SAPs have N_s antennas, which imply that the CU is effectively utilizing N_sN antennas. In the following, we will consider only the case where the number of antennas at the SUEs is lower or equal than the number of antennas at the MBS ($M_s \leq M_m$), since this is the typical case in practical systems. However, the proposed schemes may be easily extended to the case where $M_s > M_m$.

The received signal at the CU is given by

$$\mathbf{y}_s = \underbrace{\mathbf{h}_0 \mathbf{x}_0}_{\textit{Interference}} + \underbrace{\sum_{k=1}^{NK} \mathbf{H}_k \mathbf{x}_k}_{\text{Desired signal}} + \mathbf{n}_s, \tag{8}$$

where $\mathbf{h}_0^{f_n} \in \mathbb{C}^{N_sN}$ denotes the channel between MUE and the CU at subcarrier f_n, $\mathbf{H}_k^{f_n} \in \mathbb{C}^{N_sN \times M_s}$ represents the channel between SUE_k and the CU at subcarrier f_n, $\mathbf{H}_0 = \mathrm{bdiag}[\mathbf{h}_0^{f_1}, \ldots, \mathbf{h}_0^{f_F}]$, $\mathbf{H}_k = \mathrm{bdiag}[\mathbf{H}_k^{f_1}, \ldots, \mathbf{H}_k^{f_F}]$ and $\mathbf{n}_s$ is the zero mean white Gaussian noise with variance σ^2.

The real representation of (8) is given by

$$\mathbf{y}_s^R = \underbrace{\mathbf{H}_0^R \mathbf{x}_0^R}_{\textit{Interference}} + \underbrace{\sum_{k=1}^{NK} \mathbf{H}_k^R \mathbf{x}_k^R}_{\text{Desired signal}} + \mathbf{n}_s^R, \tag{9}$$

The SUE_k can transmit $M_s - 1$ data streams, per subcarrier, since it is equipped with M_s antennas and the MUE is equipped with a single antenna. This arises because the SUEs need to construct a signal that will allow the MUE signals to be detected without interference at the MBS. By doing that, SUE_k is left with $M_s - 1$ degree of freedom [12].

The transmitted signal at the SUE_k over one block is given by

$$\mathbf{x}_k^R = \gamma_k \mathbf{V}_k \mathbf{d}_k^R, \tag{10}$$

where γ_k is a normalization factor to enforce the transmission power at the SUE_k, $\mathbf{V}_k \in \mathbb{R}^{2FM_s \times S_s}$ is the SUE_k SA-based precoder and $\mathbf{d}_k^R \in \mathbb{R}^{S_s}$, with $S_s = 2F(M_s - 1)$, denotes the real representation of the complex data vector $\mathbf{d}_k \in \mathbb{C}^{F(M_s-1)}$ whose elements are drawn from a M-QAM constellation.

The proposed method requires the knowledge of channel $\mathbf{G}_k \in \mathbb{C}^{FM_m \times FM_s}$ at SUE_k, see Section 3. This channel may be acquired by listening to the pilot signals broadcasted by the MBS. Since the macro-cell mode of operation is time division duplex (TDD) and the users associated to small-cells are mainly indoor, $\mathbf{G}_k$ is a quasi-static channel and the estimation processing overhead is small. Furthermore, it is also assumed that the knowledge of dual-SFBC code $\mathcal{C}$ is available at the SUEs.

III. DESCRIPTION OF THE PROPOSED PRECODER, SUBSPACE AND FILTER MATRIX

The design of the precoder and subspace matrix at the SUEs for the previously proposed SA-PNC based methods is dependent on the macro channel $\mathbf{G}_0$ [25], [26]. This assumption requires cooperation between the two systems, which is challenging to realize in practical systems. Therefore, we show that the use of dual SFBC [29] allows the macro and small cell systems to coexist over the same spectrum by eliminating the need of inter-system channel information exchange. Namely, a new diversity-oriented dual SFBC and SA-PNC scheme is proposed. As mentioned, the motivation behind the joint dual SFBC and SA-PNC is that it allows the system to achieve the benefits of SA-PNC coordinated and static methods without suffering from their major drawbacks.

A. SA BASED PRECODING AT SUEs AND DECODING AT THE MBS

We utilize SA based precoding to align all small cells signals at the MBS and instead of considering these signals as interference to be removed, a linear combination of the aggregated signal is decoded. Namely, the MBS must decode its desired data $\mathbf{d}_0^R$ and a linear combination $\gamma_1\mathbf{d}_1^R+\gamma_2\mathbf{d}_2^R+\ldots+\gamma_{NK}\mathbf{d}_{NK}^R$ of the small-cell data symbols.

From (4) and (6) we verify that to enforce zero interference at the MBS and decode $\mathbf{d}_0^R$ the SA precoder of SUE$_k$ must satisfy

$$\mathbf{Q}_m\left[\mathcal{C}\right]\mathbf{G}_k^R\mathbf{V}_k=\mathbf{0}. \tag{11}$$

Furthermore, to perform SA the following constraint must be also satisfied

$$\mathbf{W}_m\left(\mathbf{G}_0^R\mathbf{x}_0^R+\sum_{k=1}^{NK}\gamma_k\mathbf{G}_k^R\mathbf{V}_k\mathbf{d}_k^R\right)=\sum_{k=1}^{NK}\gamma_k\mathbf{d}_k^R, \tag{12}$$

$\forall\mathbf{d}_k^R\in\mathbb{R}^{S_s}$, where $\mathbf{W}_m\in\mathbb{R}^{S_s\times 2MF}$ is the filter matrix used to recover the linear combination of the small-cell symbols at the MBS. Let $\mathbf{R}_1\in\mathbb{R}^{S_s\times 2FM_m}$ and $\mathbf{R}_2\in\mathbb{R}^{S_s\times S_s}$ be two full row rank matrices, then a feasible solution to constraints (11) and (12) is

$$\mathbf{V}_k=\left(\mathbf{W}_{m,1}\mathbf{G}_k^R\right)^{-1}\mathbf{S}_{\mathcal{C}},$$
$$\mathbf{W}_m=\left(\mathbf{W}_{m,0}\mathbf{S}_{\mathcal{C}}\right)^{-1}\mathbf{W}_{m,0}\mathbf{W}_{m,1}, \tag{13}$$

where the matrices $\mathbf{W}_{m,1}\in\mathbb{R}^{(S_m+S_s)\times 2FM_m}$, $\mathbf{S}_{\mathcal{C}}\in\mathbb{R}^{(S_m+S_s)\times S_s}$ and $\mathbf{W}_{m,0}\in\mathbb{R}^{S_s\times(S_m+S_s)}$ are defined by $\mathbf{W}_{m,1}=\left[\mathbf{Q}_m\left[\mathcal{C}\right]^T,\mathbf{R}_1^T\right]^T$, $\mathbf{S}_{\mathcal{C}}=\left[\mathbf{0},\mathbf{R}_2\right]^T$ and $\mathbf{W}_{m,0}=\text{null}\left(\mathbf{W}_{m,1}\mathbf{G}_0^R\right)$. As $\mathbf{R}_1$ and $\mathbf{R}_2$ may be any full row rank matrix we assume that these two matrices are known at both the macro and small-cell terminals, i.e. they are fixed at the beginning of the interaction between the two systems and do not change further.

Note that the SUE$_k$ precoder $\mathbf{V}_k$ is a function of $\mathbf{G}_k$, $\mathbf{R}_1$, $\mathbf{R}_2$ and $\mathbf{Q}_m[\mathcal{C}]$, and the filter matrix $\mathbf{W}_m$ is a function of $\mathbf{G}_0$, $\mathbf{R}_1$, $\mathbf{R}_2$ and $\mathbf{Q}_m[\mathcal{C}]$. Therefore, all parameters are fixed except the channels $\mathbf{G}_k,\ k\in\{0,1,\ldots,KN\}$. As these channels may be estimated at the corresponding terminals, as stated in the system model section, the terminals have all the information required for the computation of the precoder and filter matrix and no inter-system information exchange is required.

B. DECODING PROCESS AT THE CU

After the equalization process, the MBS performs a hard decision procedure to do the mapping of analog signal bits $\mathbf{d}_s^R=\gamma_1\mathbf{d}_1^R+\gamma_2\mathbf{d}_2^R+\ldots+\gamma_{NK}\mathbf{d}_{NK}^R$ to the corresponding constellation point. Afterwards, the MBS maps this constellation point to a binary representation and finally relays it to the CU using the digital link between MBS and CU. The number of bits needed to represent the constellation point will define the data rate requirements. Let us define a set $\mathcal{V}=\{v_1,\ldots,v_G\}$ with cardinality G, where $v_n\neq v_m,\ \forall n\neq m\in\{1,2,\ldots,G\}$ and assume that the value of the normalization constants $\gamma_k,\ k\in\{1,2,\ldots,NK\}$ are selected from $\mathcal{V}$ such that for every element v in $\mathcal{V}$ there is at least one $k\in\{1,\ldots,NK\}$ such that $\gamma_k=v$, then the dimensionality of the constellation of signal $\mathbf{d}_s^R$ is M^G and the data rate requirements are $G\log_2(M)$ bits per channel use. In the next subsection, we describe how to select the cardinality G and the values of elements of set $\mathcal{V}$.

The CU has access to $2FNN_s+S_s$ equations ($2FNN_s$ from the direct signal and S_s from the digital relayed signal by the MBS). To recover the NKS_s+S_m data symbols, the constraint $2FNN_s+S_s\geq NKS_s+S_m$ must be satisfied. For IA-only methods proposed in [14] and [15], the condition $2FNN_s>NKS_s+S_m$ must be fulfilled. To solve these equations and to decode the desired symbols at the CU, adapted ML decoding or decoding via remodulation such as ZF or MMSE can be used [24]. In this work, we have used a ZF based equalizer given by

$$\mathbf{W}_s=\left(\mathbf{H}_{eq}\Gamma\right)^{\dagger}=\Gamma^{-1}\mathbf{W}_{s,1}, \tag{14}$$

where $\Gamma=\text{bdiag}[\gamma_1\mathbf{I}_{S_c},\ldots,\gamma_{NK}\mathbf{I}_{S_c}]$, $\mathbf{W}_{s,1}=\mathbf{H}_{eq}^{\dagger}$ and

$$\mathbf{H}_{eq}=\begin{bmatrix}\mathbf{H}_0^R & \mathbf{H}_1^R\mathbf{V}_1 & \ldots & \mathbf{H}_{NK}^R\mathbf{V}_{NK}\\ \mathbf{0} & \mathbf{I}_{S_s} & \ldots & \mathbf{I}_{S_s}\end{bmatrix}. \tag{15}$$

Note that from $\mathbf{W}_s\mathbf{y}_s^R$ the CU obtains an estimate of $\mathbf{d}_k^R$, $k\in\{1,\ldots,NK\}$ and from $\mathbf{W}_{s,1}\mathbf{y}_s^R$ an estimate of $\gamma_k\mathbf{d}_k^R$, $k\in\{0,1,\ldots,NK\}$.

It can be verified from the above mentioned conditions that the proposed "joint dual SFBC and SA-PNC" and "SA-PNC only" based methods can accommodate one additional SUE as compared to IA-based methods of [14] and [15]. This gain is due to the coordination between the MBS and CU through the digital link to jointly exploit SA-PNC.

C. SELECTION OF THE NORMALIZATION CONSTANTS

After applying the SA based precoder and filter matrix mentioned in constraints (11) and (12), the MBS received signal can be rewritten as

$$\mathbf{W}_m\mathbf{y}_m^R=\sum_{k=1}^{NK}\gamma_k\mathbf{d}_k^R+\mathbf{W}_m\mathbf{n}_m^R. \tag{16}$$

The main objective is to decode the linear combination of small-cell symbols $\mathbf{d}_s^R=\gamma_1\mathbf{d}_1^R+\gamma_2\mathbf{d}_2^R+\ldots+\gamma_{NK}\mathbf{d}_{NK}^R$ at the MBS. $\mathbf{d}_s^R$ is the superposition of NK M-QAM signals. Without loss of generality, in the following we consider that $\bar{\gamma}_1\geq\bar{\gamma}_2\ldots\geq\bar{\gamma}_{NK}$, where

$$\bar{\gamma}_k=\left(\frac{P_s}{\text{tr}\left(\mathbf{V}_k^H\mathbf{V}_k\right)}\right)^{1/2}. \tag{17}$$

To enforce the power constraint $\mathbf{x}_k^H\mathbf{x}_k\leq P_s, forall k=1,\ldots,K$, we must have

$$\gamma_k\leq\bar{\gamma}_k. \tag{18}$$

The dimensionality of the corresponding superposed constellation depends on the parameters $\gamma_k,\ k \in \{1,\ldots,NK\}$, as discussed in the previous subsection. These parameters are selected by considering two objectives 1) To minimize the bit error rate (BER) at the CU and 2) to reduce the data rate requirements for the digital link between MBS and CU. Therefore, by considering the equivalent channels $\mathbf{H}_k^R\mathbf{V}_k,\ k \in \{0,1,\ldots,NK\}$, the normalization constants $\gamma_k, k \in \{1,\ldots,K\}$ must be selected to solve the following optimization problem

$$(\hat{\gamma}_{1,\ldots}\hat{\gamma}_k) = \underset{\gamma_1,\ldots,\gamma_k\in\mathcal{V}}{\arg} \min \mathrm{BER}[\mathbf{W}_{s,1},\gamma_1,\ldots,\gamma_k] \quad \text{s.t. } G\log_2(M) \le R_L \tag{19}$$

where R_L denotes the maximum digital link capacity between MBS and CU and $\mathrm{BER}(\cdot)$ represents the average bit-error rate at the CU given by

$$\mathrm{BER}[\mathbf{W}_{s,1},\gamma_1,\ldots,\gamma_{NK}] = \frac{1}{NK}\sum_{k=1}^{NK}\mathrm{BER}_k[\gamma_k], \tag{20}$$

$$\mathrm{BER}_k[\gamma_k] \cong \frac{\alpha}{S_s}\sum_{s=1}^{S_s}\mathcal{Q}\left[\sqrt{\beta\mathrm{SNR}_{k,s}[\gamma_k]}\right]. \tag{21}$$

where $\alpha = 4(1-1/\sqrt{M})/\log_2[M]$, $\beta = 3/(M-1)$, $\mathcal{Q}[\cdot]$ denotes the Q-function and $\mathrm{BER}_k[\cdot]$ the BER of SUE_k. $\mathrm{SNR}_{k,s}$ denotes the signal-to-noise ratio of stream s of user k and is given by

$$\mathrm{SNR}_{k,s}[\gamma_k] = \frac{\gamma_k}{(\mathbf{W}_{s,1})_{(k,s)}(\mathbf{W}_{s,1})_{(k,s)}^H}, \tag{22}$$

where $(\mathbf{W}_{s,1})_{(k,s)}$ denotes the row of matrix $\mathbf{W}_{s,1}$ used to estimate the data symbol of stream s of user k.

The $\mathrm{SNR}_{k,s}$ is only a function of γ_k, since $\mathbf{W}_{s,1}$ only depends on the equivalent channel $\mathbf{H}_{eq}$, which is fixed. Therefore, to minimize the BER_k we should maximize the normalization constant γ_k.

Let us say that we have a group of users, denoted by $\mathcal{U}$, to which we assign the same normalization constant, then we can always get a better BER if we select instead $\gamma_k = \max_{u\in\mathcal{U}}\bar{\gamma}_u$. As such the normalization constants must be selected from the set $\{\bar{\gamma}_1,\ldots,\bar{\gamma}_{NK}\}$. Furthermore, as the cardinality of set $\mathcal{V}$ must be G then only the best G elements from $\{\bar{\gamma}_1,\ldots,\bar{\gamma}_{NK}\}$ must be selected.

As the parameter G denotes the cardinality of the set $\mathcal{V}$ problem (19) is a combinatorial optimization problem and therefore hard to solve. To obtain a solution for problem (19) we propose first to fix the cardinality G and the value of the elements of set $\mathcal{V}$ using a suboptimal approach and then select the optimum value for the normalization constants.

The selection of the G elements of set $\mathcal{V}$ is done by dividing the users in G groups with $\overline{G} = NK/G$ users, i.e. users 1 to $\overline{G}$ form one group, users $\overline{G}+1$ to $2\overline{G}$ other group and so on. Therefore, according to the previous description the optimum set $\mathcal{V}$ is $\mathcal{V} = \{\bar{\gamma}_{\overline{G}}, \bar{\gamma}_{2\overline{G}},\ldots,\bar{\gamma}_{NK}\}$ and the optimum value for each normalization constant is

$$\begin{cases} \gamma_1 = \gamma_2 = \ldots = \gamma_{\overline{G}} = \bar{\gamma}_{\overline{G}}, \\ \gamma_{\overline{G}+1} = \ldots = \gamma_{2\overline{G}} = \bar{\gamma}_{2\overline{G}}, \\ \vdots \\ \gamma_{NK-(\overline{G}-1)} = \ldots = \gamma_{NK} = \bar{\gamma}_{NK}. \end{cases} \tag{23}$$

To select the value of parameter G, i.e. the group size, we consider three specific cases: $G = 1$, $G = NK$ and $G = N$ which are denoted by "one group", "NK group" and "N group" in the following. These three cases are described in more detail in the subsequent subsections.

1) ONE GROUP: ALL EQUAL CASE

For this first case, we consider the case where $\gamma_k, k \in \{1,\ldots,K\}$ are all equal. For this case the normalization constant is set to

$$\gamma_1 = \gamma_2 = \ldots = \gamma_{NK} = \bar{\gamma}_{NK}. \tag{24}$$

2) NK GROUP: ALL DIFFERENT CASE

For the second case, we consider that $\gamma_k, k \in \{1,\ldots,NK\}$ are all different and the normalization constant is set to

$$\gamma_k = \bar{\gamma}_k, \quad k \in \{1,\ldots,NK\}. \tag{25}$$

3) N GROUP CASE

For the third case, we set the size of the group equal to the number of small-cells. for this case the normalization constants are selected as follows

$$\begin{cases} \gamma_1 = \gamma_2 = \ldots = \gamma_K = \bar{\gamma}_K, \\ \gamma_{K+1} = \ldots = \gamma_{2K} = \bar{\gamma}_{2K}, \\ \vdots \\ \gamma_{NK-K+1} = \ldots = \gamma_{NK} = \bar{\gamma}_{NK}. \end{cases} \tag{26}$$

D. COMPUTATIONAL COMPLEXITY AND FEEDBACK ANALYSIS

The computational complexity and data rate requirements for the digital link are presented in Table 1. Both the computational complexity and data rate requirements depend on the dimensionality of the constellation of vector $\mathbf{d}_s^R$. A description about the dimensionality of the constellation of vector $\mathbf{d}_s^R$ is provided in subsection III.B. The computational complexity is proportional to the number of points of the constellation and the feedback requirement, in bits per channel use, is equal to the logarithm base two of the dimensionality of $\mathbf{d}_s^R$.

For the first case considered in the previous section, i.e. the one Group case, where all the parameters $\gamma_k, k \in \{1,\ldots,K\}$ are equal, i.e. $G = 1$, the dimensions of $\mathbf{d}_s^R$ is on the order of M and the data rate requirements for the digital link is $\log_2(M)$ bits per channel use. Whereas, for the second case (NK Group), $\gamma_k, k \in \{1,\ldots,NK\}$ are all different ($G = NK$) and the dimensionality of $\mathbf{d}_s^R$ is M^{NK} and $NK\log_2(M)$ bits are required to represent the constellation point. Finally, for

TABLE 1. Data rate requirement for three cases.

Cases	Dimensionality of $\mathbf{d}_s^R$	Number of Bits	Data Rate Requirement
One Group	M	$\log_2(M)$	Lowest
N Group	M^N	$N\log_2(M)$	Intermediate
NK Group	M^{NK}	$NK\log_2(M)$	Highest

the third case (N Group) the dimensionality of $\mathbf{d}_s^R$ is M^N and the data rate requirements for the digital link is $N\log_2(M)$ bits per channel use. As may be verified from Table 1, the higher the dimensionality of $\mathbf{d}_s^R$, the larger is the number of bits required to represent the constellation point and therefore the larger the computational complexity and data rate requirements of the digital link. Therefore, the one group case has the lowest, the N group case the intermediate and NK group case the highest digital link data rate requirements and complexity.

E. DUAL-SFBC BASED FILTER MATRIX

In the following, we consider two specific examples of dual-SFBC and present the corresponding filter matrix $\mathbf{Q}_m[\mathcal{C}]$. We consider the dual-Alamouti and dual-Quasi orthogonal codes [29], [30] with data symbols coded in space and frequency. In the following, we consider that $F \geq M_m$ in order to achieve full diversity. From the context of SFBC literature, the channel between adjacent carriers is assumed to be approximately constant (OFDM based systems are usually designed so that channels between a group of adjacent subcarriers are approximately flat).

1) DUAL-ALAMOUTI CODE

For the first example, we consider for simplicity of exposition that the receiver (MBS) utilizes two antennas ($M_m = 2$) and the transmitter (MUE) utilizes a single antenna, as shown in Figure. 2. The filter matrix based on the dual Alamouti code over 2 subcarriers, $F = 2$ is given by

$$\mathbf{Q}_m[\mathcal{C}] = \mathrm{bdiag}\left[\mathbf{Q}_R, \mathbf{Q}_I\right], \tag{27}$$

$$\mathbf{Q}_R = \begin{bmatrix} 1 & 0 & 0 & 1 \\ 0 & 1 & -1 & 0 \end{bmatrix}, \tag{28}$$

$$\mathbf{Q}_I = \begin{bmatrix} 1 & 0 & 0 & -1 \\ 0 & 1 & 1 & 0 \end{bmatrix}. \tag{29}$$

where $\mathbf{Q}_R$ and $\mathbf{Q}_I$ filter the real and imaginary parts of $\mathbf{y}_m$, respectively. Please note that as stated in the system model section, the roles of the encoder and decoder of a dual-SFBC are swapped in relation to the corresponding SFBC. Therefore, if the transpose of matrix $\mathbf{Q}_m[\mathcal{C}]$ is used to encode the data vector $\mathbf{d}_0^R$ the output would be the encoding of a standard Alamouti code.

2) DUAL-QUASI ORTHOGONAL CODE

For the second example, we consider the dual quasi orthogonal code at the macro-cell system. For this case, we consider that the MBS has four antennas ($M_m = 4$) and the MUE a single antenna. The filter matrix based on dual quasi-orthogonal code over 4 subcarriers, $F = 4$ is given by

$$\mathbf{Q}_m[\mathcal{C}] = \mathrm{bdiag}\left[\mathbf{Q}_R, \mathbf{Q}_I\right], \tag{30}$$

$$\mathbf{Q}_R = \begin{bmatrix} 1&0&0&0&0&-1&0&0&0&0&1&0&0&0&0&-1 \\ 0&1&0&0&1&0&0&0&0&0&0&1&0&0&1&0 \\ 0&0&1&0&0&0&0&-1&1&0&0&0&0&-1&0&0 \\ 0&0&0&1&0&0&1&0&0&1&0&0&1&0&0&0 \end{bmatrix}, \tag{31}$$

$$\mathbf{Q}_I = \begin{bmatrix} 1&0&0&0&0&1&0&0&0&0&1&0&0&0&0&1 \\ 0&1&0&0&-1&0&0&0&0&0&0&1&0&0&-1&0 \\ 0&0&1&0&0&0&0&1&1&0&0&0&0&1&0&0 \\ 0&0&0&1&0&0&-1&0&0&1&0&0&-1&0&0&0 \end{bmatrix}. \tag{32}$$

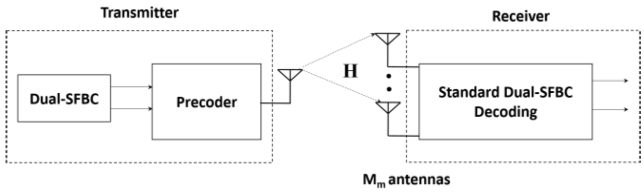

FIGURE 2. Dual-SFBC at the macro-cell system.

IV. SIMULATION RESULTS AND DISCUSSION

The performance of the proposed joint dual SFBC and SA-PNC method is compared with IA based (coordinated, static and 2-bit) methods proposed in [14] and [15], and SA-PNC (coordinated, static and 2-bit) methods proposed in [25] and [26]. For the dual SFBC, we consider the example of dual Alamouti code ($F = 2$). Since interference can be mitigated for any number of SAPs, adding more SAPs would not impact the performance of the MBS and a scenario with 2 SAPs is considered for the sake of simplicity. Additionally, for the $2n$-bit method, $n = 1$ is considered as using $n > 1$ provides marginal performance improvement. For IA based methods, a total of 3 SUEs are connected (one SAP serves 2 SUEs, while the second SAP serves a single SUE). On the other hand, for the proposed joint dual Alamouti with SA-PNC and SA-PNC based methods (coordinated, static and 2-bit), a total of 4 SUEs are connected (each SAP serves 2 SUEs). It is considered that, for both the proposed joint dual Alamouti with SA-PNC and existing methods of [14], [15], and [26] the CU jointly processes the signals of 2 SAPs. We use ZF decoding at the CU. We consider 4-QAM modulation for all the methods presented in this paper. As a simplest case, we consider 2 antennas at the MBS ($M_m = 2$), while a single antenna is utilized at the MUE. At the small-cell system, we consider that each SUEs and SAPs have 2 antennas ($M_s = N_s = 2$).

SAPs with coverage radii of 600m are assumed to be uniformly placed within the boundaries of an MBS with a coverage radius of 1000m. The power of the MUT is assumed to be 4 times higher than the power of SUEs. We consider the ITU pedestrian B channel model [31] and the path loss exponent is set to 3.5. The SNR at the cell edge is defined as $(P_t R^{-\theta}/\sigma^2)$, where θ is the path loss exponent and R is the cell-radius. The OFDM parameters used for simulating both the macro-cell and small-cell systems are as follows: $FTTsize = 1024$, sampling frequency $f_s = 15.36MHz$, cyclic prefix length $c_p = 5.21\mu s$, 128 subcarriers with 15 kHz separation.

A. MACRO-CELL SYSTEM

The MBS performance for all the three cases are similar and is shown in Figure 3. IA based methods (coordinated, static and 2-bit) and SA-PNC methods (coordinated, static and 2-bit) achieve the same performance (i.e. the BER curves are overlapping), since both methods use the same approach to mitigate interference [25], [26]. The performance of our proposed joint dual Alamouti with SA-PNC approach has a gap of around 3dB when compared to the IA coordinated and SA-PNC coordinated methods but outperforms IA static and SA-PNC static schemes by 8 dB at 10^{-3} BER in spite of having identical information-exchange requirements. We can also see that the coordinated (IA and SA-PNC) methods have the best performance at the expense of highest feedback requirements. The static (IA and SA-PNC) methods provides the worst performance but requiring the lowest feedback constraint. The 2-bit (IA and SA-PNC) schemes can be seen as the intermediate cases since they provide a trade-off between performance and inter-system information exchange requirements. Finally, the joint dual-Alamouti and SA-PNC method overcomes the shortcomings of coordinated and static methods without any inter-system information exchange requirements.

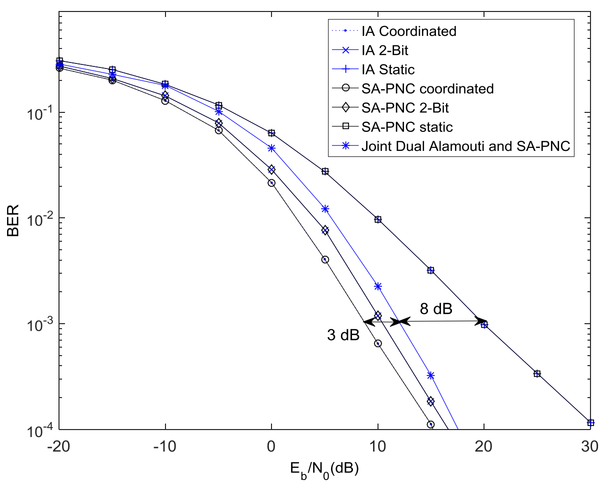

FIGURE 3. **BER for macro-cell system (All cases).**

B. SMALL-CELL SYSTEM

The BER curves of SAPs for the proposed and existing methods (for all three cases) are shown in the following figures. For each case; we compare the results of our proposed joint dual Alamouti with SA-PNC and existing IA and SA-PNC methods using 4-QAM modulation. Hereinafter, for IA based methods, only the curve for the IA coordinated method is presented, as the results for IA static and IA 2-bit methods are similar to the coordinated method [26]. Moreover, the performance of small-cell system depends on the group size of the normalization constant, i.e., increasing the group size will result in improved BER performance. Namely, we present results for "one group", "NK group" and "N group" cases in the following.

1) ONE GROUP CASE

Figure 4 presents the BER performance for the one group case. The SA-PNC (coordinated, 2-bit and static) methods achieves the worst performance when compared to the IA-only based methods. This is due to the fact that the normalization constant has the lowest group size for this case, i.e. results in the worst BER performance. The proposed joint dual Alamouti and SA-PNC scheme provides better performance when compared to the SA-PNC (coordinated, 2-bit and static) methods of [25] and [26] while overcoming the major shortcoming of SA-PNC (coordinated, 2-bit and static) methods, in spite of having no coordination requirements. Moreover, the proposed joint dual Alamouti and SA-PNC scheme and the previously proposed SA-PNC (coordinated, 2-bit and static) methods can serve one additional SAP user as compared to IA-only methods proposed in [14] and [15].

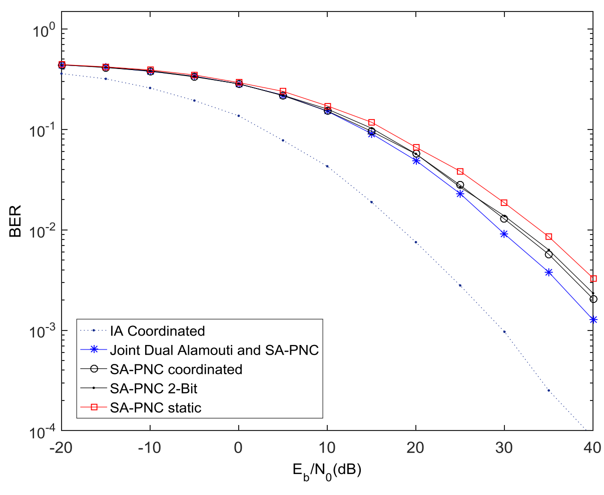

FIGURE 4. **BER for small-cell system (one group case).**

2) NK GROUP CASE

In Figure 5, the BER performance for the NK group case is presented. As we can see, the SA-PNC (coordinated and 2-Bit) methods achieve near to optimal performance (i.e., the curves of SA-PNC method are similar with the IA coordinated method). For the SA-PNC static method, the performance is much better when compared to the one group case (a gap of around 3dB for a target BER of 10^{-3}). Moreover, our proposed joint dual Alamouti

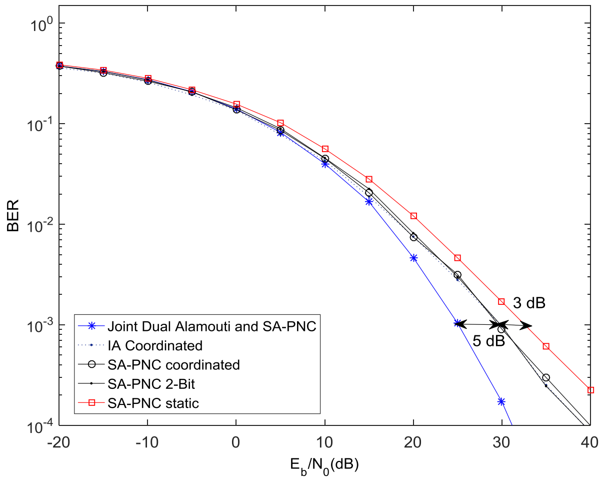

FIGURE 5. BER for small-cell system (NK group case).

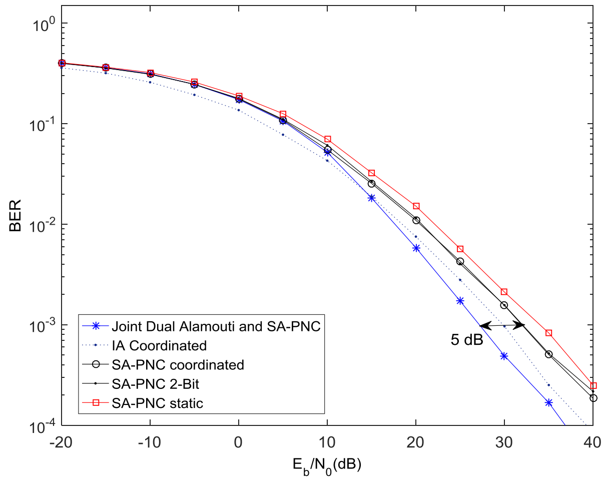

FIGURE 6. BER for small-cell system (*N* group case).

and SA-PNC scheme outperforms the IA-coordinated and SA-PNC (coordinated, 2-bit) methods (a gap of around 5 dB for a target BER of 10^{-3}) and provides around 8dB performance gain over the SA-PNC static method, with similar inter-system information. As for the previous case, our proposed joint dual Alamouti and SA-PNC and previously proposed SA-PNC methods can connect additional SUEs as compare to IA based methods without any performance degradation.

3) N GROUP CASE

The BER performance for the *G* group case (i.e. the intermediate generic case) is presented in Figure 6. As discussed in Section III-B, that case-3 provides an intermediate approach to achieve the commitment between the BER and digital link data rate requirements. As it can be noticed from Figure 6, the SA-PNC (coordinated and 2-bit) methods approach the performance of case-2, and a gap of around 2dB for a BER of 10^{-3} is observed for the SA-PNC coordinated and SA-PNC 2-bit methods and a gap of around 4dB is observed for the SA-PNC static method when compared to the IA-coordinated method. Once again, our proposed joint dual Alamouti and SA-PNC scheme outperforms the IA and SA-PNC methods and provides 3dB, 5dB and 8dB better performances as compared to IA-coordinated, SA-PNC (coordinated and 2-bit), and SA-PNC static methods, respectively while connecting an additional SUE as in the previous cases.

In Figure 7, we compare the BER performances of the proposed joint dual Alamouti and SA-PNC for all three cases. The *NK*group case, achieves the best performance, since the group size is the biggest for this case hence results in the best performance. The one group case provides the worst performance due to the fact that it has the lowest group size. The *N* group case is the intermediate case, where it achieves the performance close to *K* group case, a gap of around 2 dB for a target BER of 10^{-3}.

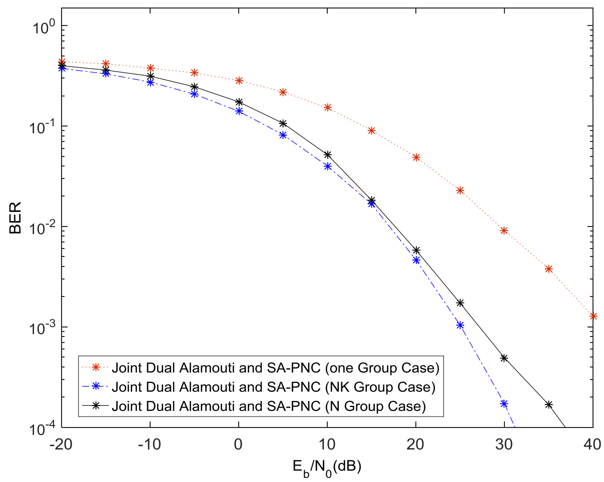

FIGURE 7. BER performance for joint dual SFBC and SA-PNC scheme (All cases).

C. TRADE-OFF BETWEEN PERFORMANCE AND FEEDBACK REQUIREMENTS

As verified from numerical results, for the macro-cell system the performance of our proposed dual Alamouti and SA-PNC and previously proposed SA-PNC methods [25], [26] are identical for all three cases. However, for the small-cell system, the dual Alamouti with SA-PNC and only SA-PNC based methods with *NK* group case achieves the best performance while the one group case provides the worst performance and the *N* group case sits in between. Therefore, we can improve the performance of small-cells by acquiring a higher digital link data rate and without causing any performance degradation at the macro-cell system. For the case with IA-based methods, the IA (coordinated, 2-bit and static) methods provides the similar performance at the small-cells. On the other hand, the IA static, IA 2-bit and IA coordinated schemes achieve the worst, intermediate and best performance, respectively at the macro-cell system.

Therefore, there exists a trade-off between BER and coordination requirements between the two systems.

V. CONCLUSIONS

In this paper, we proposed a new multicarrier diversity-oriented joint dual SFBC and SA-PNC scheme for the uplink of HetNets. The proposed scheme overcomes the shortcomings of IA static and SA-PNC (coordinated, 2-bit and static) previously proposed methods without imposing any additional coordination requirements. The key motivation behind the utilization of diversity-oriented SFBC at the macro-cell system is to facilitate inter-system coexistence without any inter-system channel information requirements, where small-cells only need to sense that the macro-cell system is using a scheme based on dual SFBC. The numerical results showed that, the proposed dual SFBC and SA-PNC method promises to provide close to optimal performance at the macro-cell system and achieves much better performance at the small-cell system, since the diversity order is increased by exploiting the benefits of dual SFBC and utilizing interference as a useful signal by using the joint SA enabled PNC approach.

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