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A Physical Layer Network Design for 5G with Massive MIMO

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Abstract

The future physical layer network coding (PNC) attitude is applied for uplink scenario in binary systems, and the design recommendation serves multiple mobile terminals (MTs) and guarantees unambiguous recovery of the message from each MT. This paper presents a PNC approach for network MIMO (N-MIMO) systems to release the heavy problem of backhaul load. We present a PNC design standard on first based on binary matrix theories, followed by an adaptive optimal mapping selection algorithm based on the proposed design criterion. In order to reduce the real-time computational complexity, a two stage search algorithm for the optimal binary PNC mapping matrix is advanced. Numerical results show that the recommended scheme achieves lower outage probability with reduced backhaul load compared to practical CoMP schemes which quantize the estimated symbols from a log likelihood ratio (LLR) based multiuser detector into binary bits at each access point (AP).

Keywords: Massive MIMO systems; Effective capacity; RF chains; Equivalent precoding.

Introduction

As various wireless multimedia applications are getting more and more popular, the demand for wireless traffic is increasing rapidly, and the massive multi-input-multi output (MIMO) technology has been proposed as a key technology for the next generation (5G) wireless communication systems [1], facilitating to guarantee the increasing demand of user Quality of Experience [2]. Recently, a number of excellent studies have validated that massive MIMO systems are specialized in improving the wireless communication capacity vastly in cellular networks [3]. Apparently, the huge antenna arrays have to be deployed compactly because enough spaces are not available at not only base stations (BSs) but also mobile terminals; therefore, the interaction of mutual coupling among antennas gets so strong that it cannot be ignored in massive MEMO systems [4]. Also, the realistic channel capacity, which is subject to the quality of service (QoS) in multimedia wireless communication systems and the Shannon capacity, are not the same thing [5]. So, exploring a new precoding solution for the 5G massive MIMO multimedia communication systems is necessary.

The concept of network multiple input, multiple output (NMIMO) [6] has been known for some time as a means to overcome the inter-cell interference in fifth generation (5G) dense cellular networks, by allowing multiple access points (APs) to cooperate to serve multiple mobile terminals (MTs). This was implemented in the coordinated multipoint (CoMP) approach standardized in LTE-A [7]. More recently the Cloud Radio Access Network (C-RAN) concept has been proposed, which has similar goals [8]. However these approaches result in large loads on the backhaul network (also referred to as fronthaul in C-RAN) between APs and the central processing unit (CPU), many times the total user data rate.

While there has been previous work addressing backhaul load reduction in CoMP and C-RAN, using, for example, Wyner-Ziv compression [9] or iterative interference cancellation[10],the resulting total backhaul load remains typically several times the total user data rate. A novel approach was introduced in [11], based on physical layer network coding (PNC), which reduces the total backhaul load to be equal to the total user data rate. However most previous research on PNC focused on two-way relay channel (TWRC) or lattice code-based design.

In [11], a PNC scheme based on BPSK is designed for the TWRC. Compute-and-forward approach which generalizes PNC of TWRC to multiuser relay networks by utilizing lattice network coding [12] is presented in [13], and heterogeneous modulation schemes which lead to precoded PNC design are studied in [14]. Work in [15] provides a PNC design in N-MIMO system but the design focused on lower order modulation schemes. In this paper, we present a PNC design with unambiguous detection of messages from all MTs to reduce the backhaul load in binary N-MIMO systems, followed by an adaptive optimal mapping matrix selection algorithm.

The main contributions are listed as follows In Section 2, a system model in which there is a 2D antenna array is described for massive MIMO wireless communications. In Section 3, the effect of mutual coupling on the massive MIMO wireless systems is evaluated by the receive diversity gain. Moreover, an optimal equivalent precoding matrix is proposed to reduce the cost of RF chains and satisfy the multimedia data requirements for 5G massive MEMO multimedia communication systems. Furthermore, the upper bound of effective capacity is derived for 5G massive MIMO multimedia communication systems. Numerical simulations and analysis are presented in Section 4. Finally, Section 5 summarizes the paper.

System Model

A massive MIMO wireless transmission system is illustrated in Figure 1. The wireless down-link between user equipment (UE) with multi-antenna and a BS with a 2D rectangular antenna array is studied in this paper. First of all, we define some basic parameters for this model. We define as the wavelength of the carrier, *d* as the antenna spacing of this antenna array, *a* .*a* 1/ as the length of this antenna array, and *b* .*b* 1/ as the width of this antenna array. If we would like to deploy *m* antennas in each row and *n* antennas in each column for this antenna array, then we will have the relationship as listed in (1),

$$d = \frac{a\boldsymbol{\lambda}}{M-1} = \frac{b\boldsymbol{\lambda}}{n-1} \tag{1}$$

And the total number of antennas in this antenna array *M* can be derived easily as follows:

$$M = mn \tag{2}$$

If we define *SNRBS* as the signal-to-noise ratio (SNR) at the BS, H and stand for the small-scale fading matrix and large-scale fading coefficient of the channel in this model respectively, the signal the BS transmits is defined as x, w means the additive white Gaussian noise (AWGN) over wireless channels, and the mutual coupling matrix is configured as K, equivalent precoding matrix is configured as F*eq*, A is defined as steering matrix, then the down-link signal vector received at a UE equipped with *N* antennas can be expressed as follows:

$$y = \sqrt{SNR_{BS}}\,\mathbf{HAKF}_{eq}\beta^{1/2}\,x+w \tag{3}$$

In which **x** is a *Ns x*1 vector and **w** is a *N* x 1 vector. **H**-*CN* (0, *P***I**) is governed by a complex Gaussian distribution and is expressed as follows:

$$H= [\ h_1,\ldots,h_p,\ldots.h_p]^T \in C^{NxP} \tag{4}$$

In which C *n x p* denotes a *NxP* matrix, *P* stands for the number of the independent incident directions,and **h***p*-*CN*(0, **I**) stands for the complex coefficient vector of small-scale fading received from the *p*th incident direction, which is expressed as follows:

$$hp=hp^{(r)} +_j h_p^{(i)} \tag{5}$$

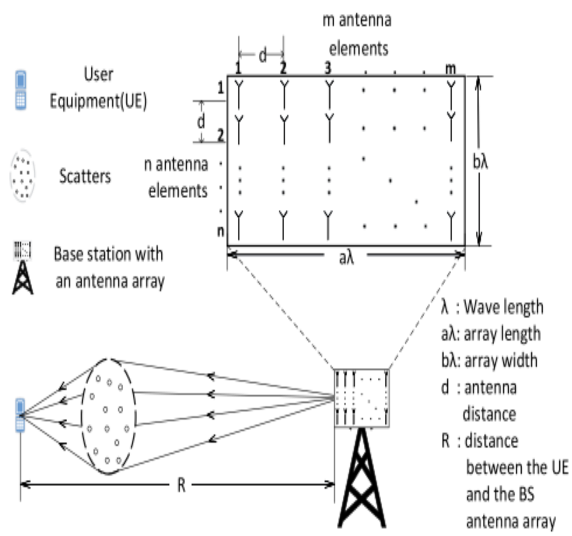

Figure 1. System model.

In which h*p^r* is defined as the real part of h*p* and h*p^i* is defined as the imaginary part of h*p*. Furthermore, both of them are Gaussian random variables distributed independently and identically, whose expectation and variance are 0 and 0.5, respectively.

Definitely, *P* will be very large if considerable scatterers exist in the propagation environment. According to [16], we divide the angular domain into *P* independent incident directions with *P* being large but finite.Here, we assume both of the azimuth angle *q*(*q* D1, ..., *P*), and elevation angle is within the scope of Œ=2, =2. Each independent incident direction corresponds to one steering vector **a** *q*, 2 C*M*1, so all the *P* steering vectors can constitute the steering matrix

A of the rectangular antenna array, which is expressed as follows:

$$A=[a\ (\phi_1,\Theta),\ldots,a(\phi_q,\Theta),\ldots,a(\phi_p,\ \Theta)] \tag{6}$$

If we define $A^q \in C^{nxm}$as the steering matrix of the *q*th incident direction of the rectangular antenna array, we will obtain the following relationship,

$$Vec(A^q)=a((\phi_1,\Theta) \tag{7}$$

in which Vec(A^q) is defined as the matrix vectorization operation. Without loss of generality, we assume the antenna that locates at the first place for both the row and the column of the rectangular antenna array as the reference point ofwhich the phase response is zero. And we normalize amplitude responses of all the antennas of the antenna array as

1.(1<=c<=m,1<=e<=n) as the element in the steering matrix **A***q*, which locates at the ***c*th** row and ***e*th** column, and it is expressed as follows:

$$A^q_{ce}=\exp\{j\tfrac{2\pi}{\lambda}[(c-1)d\cos\phi_q\sin\theta+(e-1)d\sin\phi q\sin\theta]\} \quad (8)$$

For a rectangular antenna array with **M** elements, we define K ε C^mxm as the corresponding mutual coupling matrix, which is expressed as follows [17]:

$$K=Z_L(Z_LI+Z_M)^{-1} \quad (9)$$

In which *ZL* denotes the antenna load impedance that is constant for each antenna, Z_M denotes the *MxM* mutual impedance matrix, and **I** denotes an *MxM* unit matrix. From Figure 1, $\mathbf{Z}_M$ can be constructed by *nxn* sub-matrices, that is, $Z_M=(Z^{st})\ nxn$, where $\mathbf{Z}^{st}$, as an *mxm* mutual impedance sub-matrix, denotes the mutual impedances between the *m* antennas located at the *s*th (s=1,... ,*n*) row and the *m* antennas located at the t^{th}. (*t*=1,...,*n*) row in the rectangular antenna array. For ease of exposition, we define *Antsm* as the antenna located at the *s*th row and *u*th. (*s* =1,..., *m*; *u* =1, ..., *m*) column of the rectangular antenna array and define *Anttv* as the antenna located at the *t*th row and the *v*th (*t* =1, ..., *m*; *v*=1, ..., *m*)column of the rectangular antenna array, the corresponding distance between which is give

$$d^{st}_{uv} = {}^{d}\sqrt{(t-s)^2} + {}^{d}\sqrt{(v-u)^2} \quad \text{Thus, } Z^{st}$$

As follows:

$$\begin{matrix} Z^{st}_{11} & Z^{st}_{12},\ldots, & Z^{st}_{1m} \\ Z^{st}_{21} & Z^{st22},\ldots, & Z^{st}_{2m} \\ \ldots, & \ldots, & ,\ldots, \\ ,\ldots, & ,\ldots, & ,\ldots, \\ Z^{st}_{m1} & Z^{st}_{m2},\ldots, & Z^{st}_{mm} \end{matrix} \quad (10)$$

Consider a special case where all *M* antenna elements in the rectangular antenna array are dipole antennas with the same parameters. Then the mutual impedance $z^{st}\ uv$ only depends on the antenna spacing and can be obtained with the Electromotive force (EMF) method in [18]. With a fixed antenna spacing *d*, we have the following properties:

$$\begin{aligned} Z^{st}_{uv} &= Z^{st}_{(u+1)\,(v+1)} \\ Z^{st}_{uv} &= Z^{st}_{vu} \end{aligned} \quad (11)$$

Similar properties can be derived for Z^{st} as follows:

$$\begin{aligned} Zst &= Z(s+1)\,(t+1) \\ Z_{st} &= Z_{ts} \end{aligned} \quad (12)$$

Together with (10)-(14), the mutual impedance matrix **Z***M* can be readily obtained. It bears noting that with (10)- (14), the computational complexity can be significantly reduced compared with the direct calculation of the *MxM* entries of **Z***M*, especially with a large *M*.

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The equivalent precoding matrix $\mathbf{F}_{eq}=\mathbf{F}_{RF}\mathbf{F}_{BB}$ consists of baseband precoding matrix $\mathbf{F}BB$ and the RF precoding matrix $\mathbf{F}_{RF}$. *Ns* data streams are transmitted by $N_{RF\ t}$ radio frequency (RF) chains and *M* antennas at the BS. All wireless data are received by $N_{RF}\ r$ RF chains and *N* antennas at the UE. In this case, the detected wireless signals at the UE areexpressed by the following:

$$y = W^{\dagger}_{eq}y = W^{\dagger}_{BB}W^{\dagger}_{Rf}y \tag{13}$$

In which is a conjugate transpose operation, $\mathbf{W}_{eq}$ is a ***NxNs*** equivalent signal detection matrix, which consists of baseband detection matrix $\mathbf{W}_{BB}$, and the **RF** detection matrix $\mathbf{W}_{RF}$, **y** is the received signal vector at antennas of the UE. Essentially, $\mathbf{F}_{RF}$ and $\mathbf{W}_{RF}$ are phase shift matrices used for the signal precoding and detection at the RF chains. Hence, the absolute value of the RF detection matrix $\mathbf{F}RF$ and the RF precoding matrix $\mathbf{F}_{RF}$ is equal to 1.

Adaptive Binary PNC Design

In this section, we present the novel PNC design criterion followed by a two-stage optimal mapping selection algorithm for N-MIMO systems with multiple MTs and APs. We define and study the so-called singular fading problem first, followed by proposing a PNC design criterion to resolve this problem, and finally develop a two-stage binary mapping selection algorithm with low computational complexity. The detailed mathematical proof is also provided in this section to support our design criterion

A. Singular Fading in MA Stage

Singular fading in the MA stage is a serious problem which degrades network performance due to the indistinguishable superimposed symbols at an AP under some conditions. We give a simple example here with 2 MTs to define the singular fading problem and summarize the PNC design criterion after. It is worth to mention the design criterion can be extended to a general N-MIMO system with more than 2 MTs. A discussion is given at the end of this sub-section. The singular fading is defined as a situation in which different pairs of transmitted symbols cannot be distinguished at the receiver, mathematically given by:

$$h_j,1^{s}1 + h_j,2^{s}2 = h_j,1^{s'}1 + h_j,2s2, \tag{14}$$

where *si* and *s′ i* stand for the QAM modulated signals at the *itch* MT for $i = 1;\ 2$, and $s_i = s'\ i$. The special channel coefficient vector $\mathbf{h}_{sf}$, $[h_j;1;\ h_j;2]$ is defined as a singular fade state(SFS). In this case, the failure in detection of $[s_1\ s_2]$ and $[s'_1\ s'_2]$ due to the superimposed symbols coincide in the constellation will degrade the network performance. Please note that the solution of (5) is not unique, hence there is more than one SFSs for each QAM scheme.

We can calculate the values of all SFSs by substituting all possible modulated symbol combinations to (5), which is given by :

$$U^{(q)}_{SFS} = \frac{h_{1,2}}{h_{1,1}} = \frac{s_1^{(\gamma)} - s_1^{(\gamma')}}{s_2(^{\gamma'}) - s_2^{(\gamma)}}\ ;\ \forall s1(_{\gamma}'),s1(_{\gamma}')\ \varepsilon\ \Omega, \tag{15}$$

Where U^{q}_{SFS} is an element in the SFS set $\mathbf{V}_{SFS}$, $[r_{SFS}(1);\cdots;r_{SFS}(L)]$ for a QAM modulation scheme and the value of *L* is different for different modulation schemes. According to (5), an SFS raises an ambiguous detection problem when different symbol combinations are transmitted from MTs. We define a *clash* which contains the superimposed symbols with the same value. In order to achieve unambiguous detection, the superimposed symbols in a clash should be mapped to the same NCV. We extend the definition of clash to *cluster* in which the superimposed symbols are mapped to the same NCV to reduce the decoding complexity. In this case, a cluster may contain a set of clashed symbols and individual non-clashed symbols. Then by mapping the superimposed symbols in a cluster to the same NCV, the problem of SFS is solved. Additionally the optimal PNC mapping matrix should keep different NCVs as far apart as possible in order to achieve the unambiguous detection.

By defining *d*min as the minimum Euclidean distance between different clusters, the optimal PNC mapping matrix design criterion is given by:

$$X_j^{(opt)} = G_j^{(opt)} * W_{cluster}^{(I,k)}, \forall W_{cluster}^{(i,k)} \ \varepsilon \ w_{cluster}(k)$$
$$G_j \max {}_{d_{min}}, \text{ for } i,k=1,2,...,mu,$$

Where $d_{min} = \min \Theta(s^{(r)}_{j,sc} \neq \Theta(s^{(r')}_{j,sc}) \ |s^{(r)}_{j,sc} - s^{(r')}_{j,sc}|^2$

$$X^{(r)}_j = \Theta(s^{(r)}_{j,sc)} = G_j * w^{(r)}_{cluster},$$
$$S^{(r)}_{j,sc} = s^{(r)}_1 + u^{(q)}_{SFS} \, s^{(r)}_2,$$
$$\forall \, s^{(r)}_{1,} s^{(r)}_2 \ \varepsilon \ \Omega, \ \forall \, s^{(r)}_{j,sc} \ \varepsilon \ c^{(r)},$$
$$\forall s^{(r')}_{j,\,sc} \ \varepsilon \ c^{(r')}, \ q=1,2,\ldots,L, \tag{16}$$

The key elements in the above ceiterion is the values of SFS. In the 2-MT case, (13) can be utilised to determine all possible SFSs for a QAM modulation scheme. However, in multiple-MT (u > 2) case, (13) is no longer suitable for mathematical representation of SFSs because the increased number of MTs in MA stage results in non-linear relationship between modulated symbols and SFS values. This is still an open question for PNC design. In this paper, we utilise 2- MT case as the basis to cope with the multiple-MT case. For example, when 3 MTs are served by an AP, we could use (6) to determine the SFS between MT1 and MT2 first and the received symbol from MT3 is treated as additional noise. Then we may pair MT1 (or MT2) and MT3 for PNC encoding and treat signal from MT2 (or MT1) as additional noise. We utilise received power as threshold for pairing different MTs and this pairing method has been discussed in [19].

B. Algebraic Work for Design Criterion

In this subsection, we illustrate mathematical work to support the proposed PNC design criterion. As mentioned in the previous subsection, we need to carefully design each **G***j* for $j = 1; 2; \cdots; n$, so that **G** = [**G**1*; · · · ;* **G***n*]*T* includes a number of row coefficients which forms

Theorem 1: Assuming **G** = $G_{n\times n}(R)$, where the coefficients are from a commutative ring *R*. Source messages are drawn from a subset of *R* and all source messages can be unambiguously decoded at the destination if and only if the determinant of the transfer matrix is a unit in *R*,

$$\text{Det(G)} = \text{u(R)} \tag{17}$$

Proof: We first prove that (11) gives the sufficient and necessary conditions that make a matrix **B** invertible in NMIMO networks. Suppose **B** is invertible, then there exists A matrix C ε G nxn® such that BC = CB = In.

This implies 1 = det(In) = det(BC) =det(B)det(C).

According to the definition of a unit,we say det(B) ε U(R).

We know $B.adj(B) = adj(B).B = det(B)I_n$. if det(B) ε U(R), we have

$$B.(det(B)^{-1}adj(B))=(det(B)^{-1}adj(B))B = det(B)^{-1}det(B) = I_n \quad (18)$$

Hence $C=(det(B)^{-1}adj(B))$ is the inverse of B since $BC = CB = I_n$.

If B is Invertible, then its Inverse B^{-1} is uniquely deteremined.asssuming B has two inverse ,say C, and C^{-1}

$$B.C = C.B = I_n$$
$$B.C^{-1} = C^{-1}.B = I_n \quad (19)$$

Hence we have

$$C = C.I_n = C.B.C^{-1} = In.C^{-1} = C^{-1}$$

It proves the uniqueness of the invertible matrix B over *R*.
Assume $a \neq a'$, $B \cdot a = F$, $B \cdot a' = F'$, and $F = F'$. This Means $a = B^{-1}. F = B^{-1}. F' = a'$

This contradicts $a \neq a'$. Hence it ensures unambiguously decode ability:

$$B .a \neq B. a', \forall a \neq a'. \quad (20)$$

Let Iν (Gm×n(R)) denotes the ideal in R generated by all ν × ν minors of Gm×n(R), where *ν* = 1*;* 2*;* · · ·*; r* = min (m x n). A ν × ν minor of Gm×n(R) is the determinant of a ν × ν matrix obtained by deleting m - ν rows and n - ν columns. Hence there are (mν*)* (*nν*) minors of size ν × ν. Iν (Gm×n(R)) is the ideal of *R* generated by all these minors.

Design criterion: The destination is able to unambiguously decode *u* source messages if:

1) $u \geq \max\{v | Ann_R(I_v(G_j)) = <0>\}, \forall_j = 1,2,\ldots,n,$
2) $G_j = \arg\max_{M_j} \{I(\tilde{Y}, \dot{F}_j)\};$

Where *⟨x⟩* denotes the ideal generated by *x*. Condition 1 can be proved as follows. According to Laplace's theorem, every (*ν*+1) × (*ν*+1) minor of *Gm*×*n*(*R*) must lie in *Iν* (*Gm*×*n*(*R*)).

This suggests an ascending chain of ideals in *R*:

$$<0> = I_r{+}1(G_j) \subseteq I_r(G_j) \subseteq ,\ldots, \subseteq I_1(G_j) \subseteq I_0(G_j) = R.$$

Computing the annihilator of each ideal in (18) produces another ascending chain of ideals,

$$<0> = A_{nnR}(R) \subseteq A_{nnR}(I_1(G_j)) \subseteq ,\ldots,$$
$$\subseteq Ann_R(I_r(Gj)) \subseteq Ann_R(<0>) = R. \quad (21)$$

It is obvious that:

- $AnnR(Ik(Gj)) \neq <0>$
- $AnnR(Ik'(Gj)) \neq <0>, \forall k \leq k$

The maximum value of v which satisfies Ann$R$ (I_v ($\mathbf{G}$j)) = (0) guarantees that Ik ($\mathbf{G}_j$) $2\ R,\ 8k < v$. Hence we define the rank of $\mathbf{G}_j$ as rk($\mathbf{G}_j$) = max $fv\ j$ Ann$_R$(Iv($\mathbf{G}$j)) = (0)g.

Suppose that $\mathbf{G}k\ 2\ Gm\times p(R)$ and $\mathbf{G}k'\ 2\ Gm\times n(R)$, then rk($\mathbf{G}_k$ **x** $G_{k'}$) ≤ min $f_{rk}(\mathbf{G}_k)$; rk($\mathbf{G}k'$)g, and we can easily prove that 0 ≤ rk($Gm\times n(R)$) ≤ min$fm;\ n)$. Thus, in order to guarantee there are at least u unambiguous linear equations available at the CPU, rk($\mathbf{G}_j$) must be at least u, $8j = 1;\ 2;\ \cdots;\ n$.

Condition 1 is a strict definition which ensures unambiguous decodability of the u sources. Condition 2 ensures that the selected coefficient matrix maximises the mutual information of the particular layer, giving finally the maximum overall throughput.

C. Binary Adaptive Mapping Selection Algorithm

Following the design criterion presented in the previous subsection, a binary adaptive mapping selection (BMAS) algorithm is designed in this subsection. Like the work i [20], our proposed algorithm comprises two procedures, the first of which is an Off-line search and the second is an Online search.

Before the proposed Off-line search, the values of SFS for QAM modulation schemes can be determined by using (6). Then we define an $L \times u_j$ matrix $\mathbf{H}_{SFS} = [\mathbf{h}_{sfs1};\ \cdots;\ \mathbf{h}_{sfsL}]^T$ contains all SFS for a QAM modulation scheme, where $\mathbf{h}^{(q)}_{sfs} = [h^{(v)}_{SFS}\ (q)\ ]$ for $q = 1;\ 2;\ \cdots;\ L$. By substituting all $vSF\ S$s into (10), the superimposed symbols can be obtained. According to the design criterion, the mapping matrices used at each AP should encode the superimposed constellation points within one cluster to the same NCV and additionally, the Euclidean distance between different NCVs should b maximised. In that case, an exhaustive search among all $m \times mu$ binary matrices is implemented in the proposed Offline search to calculate all possible NCVs using (15) and hence the minimum Euclidean distance between different clusters can be obtained by using (16). The optimal mapping matrix candidate for each SFS are selected according to the design criterion (17) and stored at the APs and CPU. During the real-time transmission, a search among the stored matrix candidates is applied and this procedure is called On-line search algorithm. Please note according to the unambiguous decoding criterion, an invertible global mapping matrix with maximum value of $d$min should be selected in the proposed On-line search.

A potential issue of the proposed BMAS algorithm is that the large number of SFSs in QAM schemes with higher modulation orders increases the computational complexity and searching latency in On-line search. For example, a search among 389 4 × 8 binary matrices for 16QAM to resolve all SFSs puts the proposed On-line search algorithm in realtime application to a serious trouble. However, according to our research we found that many different SFSs generate the same clashes so that they coul be resolved by the same binary matrices. In this case, the number of mapping matrices searched in the On-line search can be reduced because differ ent SFSs with the same clashes will be resolved by the same mapping matrix. Furthermore, we found that the appearanc probability of each SFS is different and we can ignore those "nonactive" SFSs with low appearance probabilities to reduce the number of mapping matrices utilised in the proposed Online search with eliminated performance degradation.

Numerical Results

In this section, we evaluate the outage probability (Pout) performance of the proposed BMAS algorithm compared to that of ideal and non-ideal CoMP over different N-MIMO networks. 4QAM and 16QAM modulation schemes are employed in the simulation. Before the QAM modulation, a convolution code with rate of 2=3 is applied and more powerful coding schemes can be employed to enhance the performance. In ideal CoMP, the bandwidth of backhaul network is assumed unlimited so that a joint multiuser ML detector is employed. While in non-ideal CoMP, bandwidth-limited backhaul network is considered and a 2-bit/4-bit quantizer which quantizes the estimated symbols from an LLR based multiuser detection into binary bits is employed. We illustrate the *P*out performance comparisons between different approaches in a 5-node network (2MT-2AP-CPU) and then extend to 6-node (3MT-2AP-CPU) and 7-node (3MT-3AP-CPU) scenarios.

Fig. 2 illustrates *P*out performance comparison among different schemes over a 2MT-2AP-CPU N-MIMO network. Single antenna is employed at each node. We note that the ideal CoMP achieves the optimal *P*out due to the utilisation of joint ML detection with unlimited backhaul. While a performance degradation is observed in non-ideal CoMP when quantizers with limited bits are utilised to release the high backhaul load burden. As illustrated in the figure, the proposed BMAS approach is superior to the non-ideal CoMP with reduced backhaul load, even when an increased quantisation bits is used in the non-ideal CoMP approach. Also the proposed BMAS approach achieves the same diversity order as ideal and non-ideal CoMP. When 16QAM is employed at each MT, degradation of the *P*out performances are observed from the figure. In the proposed BMAS algorithm, 50 most active SFSs are utilised and according to comparison in the figure, the proposed algorithm outperforms the non-ideal CoMP with half backhaul load.

In Fig. 2, the outage probability performances of different schemes in 6-node and 7-node networks are presented, respectively. In the networks with 3 MTs, an increased backhaul load (6 bits in BMAS and 72 bits in non-ideal CoMP with 4-bit quantiser) is required. Similar to the performances in Fig.2, ideal CoMP achieves the optimal *P*out in both networks and the proposed BMAS approach is superior to the non-ideal CoMP with much less backhaul load.

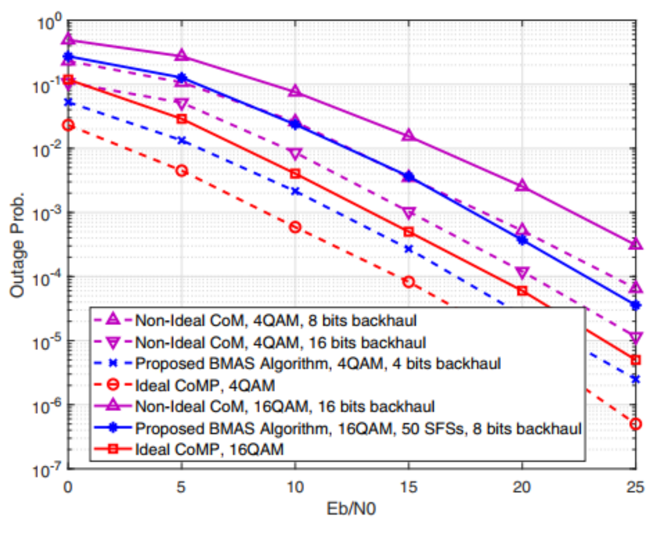

Figure2 The Outage Probability Performance of 2MT-2AP-CPU network.

V. CONCLUSION

In this paper we present, based on the mutual coupling effect, an optimal equivalent precoding matrix has been suggested to maximize the available rate and save the cost of RF chains for 5G massive MIMO multimedia communication systems. We also novel PNC design criterion for binary systems and based on this criterion, a two-stage optimal mapping matrix selection algorithm with low computational complexity is developed for the uplink of N-MIMO networks in order to reduce backhaul load. Considering the requirements of multimedia wireless communications, the upper bound of the effective capacity has been derived for 5G massive MIMO multimedia communication systems with the QoS statistical exponent constraint.

We Compared with the conventional ZF precoding matrix, numerical results show that the proposed optimal equivalent precoding matrix can obviously improve the available rate for 5G massive MIMO multimedia communication systems. The proposed BMAS algorithm is divided into two parts in order to reduce the real-time computational complexity for practical implementation, and a study of applying the proposed design criterion and search algorithm in binary systems with full-duplex Aps has been started. Comparing with the non-ideal CoMP, the proposed algorithm achieves lower outage probability with a reduced traffic in backhaul networks.

In the future work, we propose taking into account the QoS statistical exponent constraints, a more efficient signal detection precoding algorithm is worth exploring towards better performance of the multimedia massive MEMO communication systems, to fulfill the customers demand.

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