Finite Large Antenna Arrays for Massive MIMO: Characterization and System Impact

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Abstract—Massive MIMO is considered a key technology for 5G. Various studies analyze the impact of the number of antennas, relying on channel properties only and assuming uniform antenna gains in very large arrays. In this paper, we investigate the impact of mutual coupling and edge effects on the gain pattern variation in the array. Our analysis focuses on the comparison of patch antennas versus dipoles, representative for the antennas typically used in massive MIMO experiments today. Through simulations and measurements, we show that the finite patch array has a lower gain pattern variation compared with a dipole array. The impact of a large gain pattern variation on the massive MIMO system is that not all antennas contribute equally for all users, and the effective number of antennas seen for a single user is reduced. We show that the effect of this at system level is a decreased rate for all users for the zero-forcing MIMO detector, up to 20% for the patch array and 35% for the dipole array. The maximum ratio combining on the other hand, introduces user unfairness.

Index Terms—Antenna array mutual coupling, antenna measurements, antenna radiation patterns.

I. INTRODUCTION

MASSIVE MIMO proposes a new wireless communication concept relying on an excess number of base-station (BS) antennas, relative to the number of active user terminals. The technique allows for very efficient spatial multiplexing, attainable using linear processing in a time-division duplex mode [1]–[3]. It has been demonstrated to achieve a record spectral efficiency (SE) [4]. Moreover, the technology has the potential to drastically improve energy efficiency [5]. Consequently, massive MIMO addresses several key 5G requirements [6]: it offers a great capacity increase, efficiency [5]. Consequently, massive MIMO addresses several key 5G requirements [6]: it offers a great capacity increase, efficiency 

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In the existing literature [2], [3], there are clear no guidelines of how to select a basic element for a massive MIMO antenna array, although this is really a crucial aspect of a massive MIMO array and system. One thing that is known from the basic MIMO theory is that it is always better if an antenna element in such an array receives as much multipath from all directions as possible. Hence, it has often been assumed that using a quasi-omnidirectional dipole is always better than the more directive patch element.

In this paper, for the first time, the effect of mutual coupling in larger arrays on embedded gains, and specific the consequent impact on the system performance in a massive MIMO system is investigated, both for the more omnidirectional dipole element, and the well-known and widely used patch element. This is done by, including the gain variation into the small-scale fading channel model. The study of how these realized gain variations (a problem more understood in the antenna and propagation community) impact system-level performance (a problem formulation approach typically used in the massive MIMO signal processing community) is novel and of great interest to both communities.

We first study the active gain pattern variation of individual antenna elements in a large massive MIMO array, caused by the mutual coupling between the closely located elements and the edge effects in finite arrays. Both dipoles and patch antennas are considered in the simulation-based assessment, and for the latter results of real-life experiments are also presented. Our antenna measurements rely on measuring 32 active elements in an array, which is enabled by relying on a massive MIMO test bed placed in an anechoic chamber. Consequently, the impact of the gain pattern variation on the achievable SE is highlighted. While a dipole individually features a better omni-directionality, when composed in an array their severe mutual coupling causes drastic directionality on the elements and gain variations over the array. The patch array is shown to be the better choice from the system capacity point of view.

This paper is further organized as follows. First, we introduce a massive MIMO system model with an extended channel model that takes into account the 3-D antenna gain in Section II. Next, the simulation-based assessment of antenna gain variation and directivity of a representative finite large array composed of either dipoles or patch antennas is provided in Section III. The experimental validation is presented in Section IV. The impact of the gain variation on SE at system level is illustrated in Section V. Finally, we conclude this paper by reviewing the main findings, and provide recommendations for the design of large antenna arrays to be used in massive MIMO systems.

The notation used in this paper is as follows: We denote bold face upper (lower) letters as matrices (vectors). Superscripts $H$, $T$, and $−1$ stand for Hermitian transpose, transpose, and inverse, respectively. The matrix $I_K$ denotes an $K \times K$ identity matrix. Moreover, $\otimes$ denotes as Kronecker product, $\text{vec}([])$ represents vectorization of a matrix, det(·) is the determinant of a matrix, and cofactor(·) means the cofactor operation of a matrix. The element in the $k\times h$ row and $m\times h$ column of matrix $A$ is denoted by $[A]_{k,m}$.

II. 3-D SYSTEM MODEL

In this section, we introduce the system model bringing into account a 3-D gain pattern for the antenna elements in the array. The actual 3-D gain pattern at each antenna element depends both on the embedded gain pattern, as well as the various multipath reflections. This requires the establishment of a fairly detailed channel model, including propagation and array gain patterns. To access the impact from the gain variation to the system performance, we later plug the results of arrays consisting of dipoles or patch antennas in Section IV into this channel model and simulate the impact of gain variation to the user achievable rate in Section V-B.

A massive MIMO BS equipped with $M$ antennas communicates with $K$ single-antenna user terminals in the same time–frequency unit. The symbols transmitted from the $K$ users are represented as a vector $x = [s_1, \ldots, s_K]^{T}$ with $s_k$ denoting the average transmit power of user $k$, while $w \sim \mathcal{C}N(\mathbf{0}, I_M)$ is the i.i.d. complex Gaussian distributed noise. $D = [d_1, \ldots, d_K]$ represents the channel, with the channel vector between the $M$-antenna BS and the $k_{th}$ user $d_k \in \mathbb{C}^M$. Originating from the correlation channel model in [14], we decompose the channel vector $d_k$ into three terms, namely, large-scale fading, antenna gain variation, and small-scale fading

$$d_k = \sqrt{\alpha_k} \sum_{c=1}^{C_k} G(\theta_{c,k}, \phi_{c,k}) \Delta_c a(\theta_{c,k}, \phi_{c,k}) \nu_{c,k}$$

where $\alpha_k$ represents the large-scale fading and shadowing effect of user $k$ seen by the whole antenna array and $C_k$ stands for the number of multipath components. The array gain pattern is a diagonal matrix $G(\theta_{c,k}, \phi_{c,k}) = \text{diag}([g_1(\theta_{c,k}, \phi_{c,k}))^{1/2}, \ldots, (g_M(\theta_{c,k}, \phi_{c,k}))^{1/2})$, which represents the different active antenna patterns from different angle of arrivals for each antenna $m$ due to mutual coupling and the edge effect. To represent the rich multipath environment, $\Delta_c$ is an $M \times M$ matrix with binary diagonal elements

$$[\Delta_c]_{m,m} = \begin{cases} 1, & \text{belongs to cluster } c \smallskip \text{0, otherwise} \end{cases}$$

specifying whether the reflection belongs to the multipath cluster $c$. This matches the fact that for a large antenna array, reflections from one cluster do not contribute to all antennas. The steering vector $a(\theta_k, \phi_k)$ of a rectangular matrix is modeled as

$$a(\theta_k, \phi_k) = \text{vec} \left[ \left[ 1, e^{j2\pi \frac{\lambda}{4} \sin \theta_k}, \ldots, e^{j2\pi (\sqrt{M-1})\frac{\lambda}{4} \sin \theta_k} \right]^{T} \right.$$

$$\otimes \left[ 1, e^{j2\pi \frac{\lambda}{4} \sin \phi_k}, \ldots, e^{j2\pi (\sqrt{M-1})\frac{\lambda}{4} \sin \phi_k} \right] \right]$$

where $\gamma$ is the antenna spacing, $\lambda$ is the carrier wavelength, and $\phi_k$ denotes an azimuth of arrival angle. Moreover, $\nu_{c,k} \sim \mathcal{C}N(\theta, I)$ represents a standard complex
Gaussian vector. When there is only a single line-of-sight (LoS) cluster, the model simplifies to

$$d_k = \sqrt{\frac{k}{\Delta_1}} G(\theta_k, \phi_k) a(\theta_k, \phi_k).$$

For the simulation results in Section V, we use the simplified channel model in (5) to consider the effect of pure antenna patterns. However, we develop a more general channel model in (2) illustrating that the assessment of system-level impact of gain variations is not trivial.

III. GAIN PATTERN IN LARGE ARRAYS:
Dipoles Versus Patch Antennas

It is favorable for each antenna element in massive MIMO to have equal gain from all directions so as to efficiently exploit the multipath in the wireless environment. Typically, researchers assume an antenna element that preserves its characteristics in an array environment [9], [10]. However, in practice the mutual coupling between closely spaced elements may noticeably affect the embedded element radiation pattern, making it different from the pattern of a single element.

An accurate computational analysis of such influence requires a full wave solver, which is capable of taking into account the mutual coupling between elements and is able to calculate the embedded gain pattern of each element. In this paper, CST microwave studio has been used to compare the gain patterns of a single-antenna element, a finite array, and an infinite phased array. Since it is of interest to compare the qualitative performance of different types of antenna element, a more directional and a more omnidirectional antenna element have been considered. The first type is a microstrip patch antenna and the second type is a half wavelength dipole that generates an omnidirectional pattern in the H-plane.

The microstrip patch prototype consists of a square patch of 31 mm with two merged U-slots with width 1.4 mm. Then, the patch and slot shapes were deformed to polygons using the optimization procedure in CST to cover the frequency bands 2.4–2.62 and 3.4–3.6 GHz. The main comparison in this paper has been performed at 2.6 GHz. A single patch is shown in Fig. 1. The patch is etched on a 1.6 mm FR4 substrate mounted on 5 mm nylon spacers above another 1.6 mm FR4 substrate. The antenna dimensions are 70 × 70 mm. The dimension of dipole is about 51.3 × 2 mm. Both types of finite arrays are illustrated in Fig. 2,1 with an element spacing of 71 mm.

A first estimation of mutual coupling can be obtained from the analysis of the simulated S-parameters as shown in Fig. 3 for the elements in the center and in the corner. All elements in the array are consecutively numbered from the left bottom corner as shown in Fig. 2. The simulated reflection coefficient for a single element are also plotted with curves labeled single in superscript. The simulated mutual coupling between the dipoles in Fig. 3(b) is higher in comparison with the simulated mutual coupling between patch antennas in Fig. 3(a) by around 6 dB. Furthermore, in order to illustrate the accuracy of these simulations, representative measurements were performed in an anechoic chamber using a spectrum analyzer Keysight N9344C with a tracking generator; a typical agreement is illustrated in Fig. 3(a) for s27,28.

Consider the $k$th user and a single element in the BS in a LoS scenario. The power $p_k^{(r)}$ received by the element can be estimated using the well-known Friis transmission formula

$$p_k^{(r)} = p_k^{(r)} g_k^{(r)} r_k g_k^{(r)}$$

where $p_k^{(r)}$ is the transmit power from the user and $g_k^{(r)}$ is its realized gain. $g_k^{(r)}$ is the embedded realized gain or active gain pattern of the element in the BS, and $r_k = (\lambda/(4\pi \Delta r_k))^2$ is the inverse of free-space pathloss with distance $\Delta r_k$ between the $k$th transmitter and the element.

As for an array, the variation in the received power per element is coupled with the embedded gain variation of the elements, so from now on we will focus only on the receive realized embedded gain. For simplicity, the superscript $(r)$ is omitted. For an infinite array, the embedded gain is identical for all elements and can be easily calculated. The calculation reduces to the analysis of a unit cell taking into account a phase shift between neighboring elements. This phase shift depends on the main Floquet harmonic in the direction $(\theta_k, \phi_k)$. The embedded realized gain $g_{m,k}^{(r)}$ in the infinite array of $m$th element is modulated by the reflection coefficient $\Gamma_{m,k}^{(r)}$ [15].
The simulation result is shown in Fig. 4.\textsuperscript{2} When the reflection coefficient goes to 1 for some direction(s), the embedded realized gain goes to zero. These directions are called scan blindness angles (SBAs). Note that in practice, the reflection at SBA can be smaller than 1 due to the losses in dielectric and metal of the antenna elements. The far-field components can be obtained by analyzing the transmission from the antenna port to the main Floquet harmonic. One of the obvious conclusions of this paper is that a strong mutual coupling between elements can completely destroy the omnidirectional pattern of the dipole.

In a finite array, the situation is quite different. There, because of the edge effect, i.e., the fact that the elements at the edges see a different environment compared with the elements in the middle, the embedded gains of the elements are not identical. In this paper, the maximum gain variation over the elements was obtained in three steps. First, for each direction of incidence $(\theta_k, \phi_k)$, the embedded gains of all elements $g_{f}^{m,k}(\theta_k, \phi_k)$ are calculated, where the superscript $f$ stands for finite array. Second, for a given $\theta_k$ and $\phi_k$, the maximum difference between two embedded element realized gains is calculated over the whole array $\max_{m,n} (g_{m,k}^{f}(\theta_k, \phi_k) - g_{n,k}^{f}(\theta_k, \phi_k))$. Finally, this maximum difference can be studied as a function of direction as depicted in Fig. 5. Two very important observations can be made. First, the maximum gain variation increases considerably when the angle $\theta_k$ approaches the SBA. Second, the patch array shows a lower gain variation between elements at angles closer to the direction normal to the array. This means that, counter-intuitively, the more directive patch elements are the better choice from the point of view of gain variation.

In order to study the dynamic range of the array, for each angle $\theta_k$, we plotted $\max_{m,n} (g_{m,k}^{f}(\theta_k, \phi_k) - g_{n,k}^{f}(\theta_k, \phi_k))$, $\min_{m,n} (g_{m,k}^{f}(\theta_k, \phi_k))$, and $\text{mean}_{m,n} (g_{m,k}^{f}(\theta_k, \phi_k))$ of the embedded gains in Fig. 6. It is clearly proven that the role of mutual coupling is very destructive: elements that are intrinsically omnidirectional when isolated do not provide an omnidirectional coverage any more in the finite array environment. As long as $\theta_k$ is less than $60^\circ$, the dynamic range of the patch element is around 5 dB, which is 5 dB less than that of the dipole array.

\section{Measured Active Gain Patterns}

In order to validate the active gain variations predicted by the simulations, measurements were performed on the finite 32-element patch array in receive. The operating frequency was 2.6 GHz and, obviously, the element distance was 71 mm. Both the patch array and a wide-band horn (EMCO 3115) transmit antenna were located inside the anechoic chamber at KU Leuven with 7 m of distance in between, as shown in Fig. 7. The patch array was fixed on a cylindrical holder mounted on a positioner capable of rotating in the azimuthal
plane. Each patch was connected via 18 m RF cables to MIMO test bed outputs. 3 The dimension of the patch array is about 44 × 44 cm. Following the horn specification, the 3-D beamwidth in the E-plane is of 53° and 48° in the H-plane. So the array illumination should remain relatively uniform and the incident field variation is considerably smaller in comparison with the variation of measured power levels between elements. So all observed variations in the received power levels can be attributed to mutual coupling between antenna elements. 4 The dependence of the gain variation on the angle was validated by performing measurements in the following zenith angles −75° : 5° : 75° (31 discrete angles in the y–z plane) while fixing the azimuth angle $\phi_k$ to 90°. Note that, while assuming a thermal noise level of −174 dBm/Hz, the SNR of this measurement was above 50 dB. Details of the RF settings are given in Table I.

The synchronized power measurement from 32 antennas was accomplished by a massive MIMO system termed MIMO framework [16] running in the KU Leuven (KUL) MaMi test bed. From which 16 (2 RF ports each) universal software radio peripherals (USRPs) jointed together as a BS as shown in Fig. 8.

For the user side, a single USRP was connected to the horn antenna as a transmitter. The received power strength of the 32-element was calculated from the uplink data symbols synchronized by an LTE-like frame structure. At each $\theta_k$, 30 s of signal strength were recorded and the statistics of maximum, minimum, and mean from 32 antennas were plotted in Fig. 9. We observe that there is a high power gain variation among the antenna array while the zenith angle deviates from 0°. In addition, the measurements agree with the CST simulation in several aspects. First, the received gain is quite flat when $|\theta_k| \leq \pm 20^\circ$ and within this region, there is a low variation of around 3 dB. Second, the maximum

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3The anechoic chamber has an asymmetrical opening for RF cables and the positions of the RF cables are also not ideally symmetrical. Thus the real setup is a little asymmetrical due to several supporting elements leading to a slightly asymmetrical response.

4It is also important to remember that the radiation pattern of a patch element in the E-plane is not symmetrical. As a consequence, we do not expect any symmetrical gain measurements in the vertical set of elements for any incident angle.
received gain decreases noticeably for larger zenith angles while the gain variation is increasing. The measured gain range at each incident angle is summarized in Fig. 10, which follows the simulation trend with a higher level of about 1 dB. The higher level can be explained by the presence of various supporting elements located in the array environment that were not taken into account during the simulation. To see how the gain variation distributed along the panel with related to different angle of arrivals, we further map the measured gain of each element with its position on the panel at zenith angles 40° and −40° for both simulation and measurement. The received power were normalized to the mean power and shown in Fig. 11. Again, the simulation results are perfectly symmetric for both angles. In addition, the measurement result at θ_k = 40° matches the simulation quite well over the whole map. For the angle at θ_k = −40°, our measurements show larger deviation from the simulation, which is caused by multipath reflections caused by our openings in the anechoic chamber, as well as induced currents on the RF cables. We observe a larger gain variation in the edge elements compared with the center elements, this is the edge effect. It is very important to note that different elements are sensitive to very different directions, i.e., there is a severe gain variation that varies with incident angle. In any case, when the signal comes from different angles, an antenna element that receives a higher power in one direction does not always receive a higher power from the other direction. We should point out that gain variation also increases the required dynamic range for a fixed-point system implementation, as the automatic gain control in the receiver is not capable of jointly optimizing the received power levels from different directions.

V. Gain Variation and Spectral Efficiency

We have seen that there is a considerable gain variation over the array. Also, there is a different level of gain variation for patch and dipole antenna arrays. In this section, we compare

| Parameters      | Gain    |
|-----------------|---------|
| Horn            | 8.8 dBi |
| Patch           | 6 dBi   |
| TX Power        | 20dBm   |
| RX Gain         | 33.5dBi |
| Cable Loss      | -23.4dB |
| Free Space Path Loss | -57.6 dB |
| Received Level  | -12.7dBm|
the impact on single user achievable rate in a massive MIMO system. First, to theoretically show how the gain variation affects the single user achievable rate, we introduce the SE metric for both linear maximum ratio combining (MRC) and zero-forcing (ZF) detectors. Then, we apply the measured gain variation from the patch array and the CST simulated gain from the dipole array, respectively, to examine the impact of array pattern variation on a massive MIMO system.

A. Spectral Efficiency of MIMO Detectors

Under the assumption that the BS has perfect channel state information and the channel is ergodic, the uplink ergodic achievable rate can be represented as

$$R_{mrc} = \mathbb{E} \left\{ \log_2 \left( 1 + \frac{x_k \|d_k\|^4}{x_k \sum_{i=1, i \neq k}^K \|d_i^H d_i\|^2 + \|d_k\|^2} \right) \right\}$$

(7)

and

$$R_{zf} = \mathbb{E} \left\{ \log_2 \left( 1 + \frac{x_k}{\| (D^H D)^{-1} \|_{k,k}} \right) \right\}.$$  

(8)

The MRC per user rate $R_{mrc}$ in (7) illustrates the two main effects that determine the SE of massive MIMO.

1) Due to the array gain, the SNR without considering interuser-interference (IUI) term increases linearly with the antenna array size. In our system model, we given the noise power $\sigma^2 = 1$, so $\text{SNR} = x_k \|d_k\|^4 / \|d_k\|^2$, meaning that is best to have a maximal number of antennas. Antennas with a low gain, do not contribute and reduce the effective number of antennas seen.

2) The user separation enables to spatially multiplex multiple users based on their unique signature at the antenna array. The interuser correlation term $\|d_k^H d_i\|^2$ in the denominator of (7), when considering only two users for simplicity, the IUI term can be represented as

$$\|d_k^H d_i\|^2 = \|d_k\|^2 \|d_i\|^2 \cos \theta_{ki}^2$$

(9)

where $\cos \theta_{ki}$ is the angle between $d_k$ and $d_i$. Suppose due to gain pattern variation, user $k$ has a higher channel vector two-norm than user $i$. We then obtain the signal to interference ratio (SIR) relationship between user $k$ and $i$ as

$$\text{SIR}_i \leq \text{SIR}_k \iff \frac{\|d_i\|^2}{\|d_k\|^2} \leq \frac{\|d_k\|^2}{\|d_i\|^2}.$$  

(10)

We call this user unfairness caused by antenna gain pattern variation.

On the other hand, the performance of the ZF detector can be understood by looking into

$$\| (D^H D)^{-1} \|_{k,k} = \frac{\text{det}(D^H D)}{\text{cofactor}(D^H D)_{k,k}} \approx \|d_k\|^2.$$  

(11)

Here, the Hadamard inequality is applied in the approximation. Hence, we can observe that the achievable rate is directly proportional to the two-norm of the channel vector, including the antenna gain pattern.

B. Simulated Gain Variation Impact

To simulate the impact on measured antenna gain variation on system SE, we consider a LoS scenario with $M = 32$ and $K = 2$. The two users are assumed to have equal distance to the BS, so we say they share a common large-scale fading $\alpha_k = 1$. Moreover, good user (user one) locates in a higher power and less gain variation region, i.e., in the zenith angles $|\theta_k| \leq 35^\circ$ (15 discrete locations). While a second bad user locates outside this region, i.e., in zenith angles $35^\circ < |\theta_k| \leq 75^\circ$ (16 discrete locations), as illustrated in Fig. 12. Both of their azimuth angles are distributed at a very limited region $\phi_k = 88^\circ : 1 : 92^\circ$. Furthermore, no power control is considered for simplicity, and the transmitted power $x_k$ is assumed to be equal for both users.

We compare the single user achievable rate of both users for the measured patch array and the simulated dipole array. As a patch antenna has higher embedded gain and can be referenced from Fig. 6, the peak power of patch and dipole arrays are normalized to 0 and $-3 \text{ dB}$, respectively. A reference scenario without gain variation, the peak gain for all angles is set to 0 dB, is also given. Only one user in the no gain variation case is plotted for comparison, as both users have equal performance. First, the per user achievable rate of the MRC detector is plotted as shown in Fig. 13. For each realization, we randomly put one user in the good and one user in the bad region, calculate the rates, and average the two rates over ten thousand realizations. The good user apparently benefits when coexisting with a bad user. A more severe user unfairness is experienced for the dipole array, as the gain pattern variations are more pronounced here. The gain pattern variation increases the rate of good users up to 6% and decreases the rate of bad users up to 24% at an intermediate SNR $= 25$ dB. Second, the ZF achievable rate is shown in Fig. 14. From (11), we see the achievable rate is directly proportional to the received user power and this matches the result that achievable rate of patch is in general higher than that of dipole. If we compare the reference with the bad-power user of dipole array, there is a huge SNR loss by 10 dB and can be improved by 3 dB if instead applying the patch array.

Figs. 13 and 14 are obtained under the assumption that there are always two users actively communicating in the system. The conclusion of the MRC method is that the achievable rates of both users are coupled. The good user causes a larger
CHEN et al.: FINITE LARGE ANTENNA ARRAYS FOR MASSIVE MIMO: CHARACTERIZATION AND SYSTEM IMPACT 6719

The performance of each user should be evaluated by the performance of the bad user, which receives a better channel condition. The performance of a fair communication system, all users should receive similar achievable rate instead of some benefits more if the user induces less IUI, and that is why the achievable rate of the good user is higher than the achievable rate of no gain variation case. We should notice that when there is no gain variation, the two users receive the same peak power from all directions. Moreover, we should highlight that for a gain pattern variation, the two users receive the same peak power from all directions. This gain pattern variation is potentially beneficial for user separation, the main effect is that the received power from each user is decreased because of suboptimal antenna gains. For the MRC detector, the system-level impact leads to user unfairness as this detector exploits the decreased correlation of the users maximally while disadvantaging the user in a suboptimal angle. For ZF, our assessment shows that all users are disadvantaged by the antenna gain variation, and see a lower rate than a system with ideal identical antennas. Our future work is to investigate appropriate topologies and configurations of the antenna array to reduce the impact of such large gain variation effects.

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