Investigations on Millimeter-Wave Massive MIMO Scenarios by Bidirectional Ray-Tracing Using Stationary-Phase Point Approximation

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Abstract

A bidirectional ray-tracing algorithm based on the identification of stationary-phase points is utilized to investigate the average downlink user data rate and the signalinterference characteristics in a millimeter-wave massive multiple-input multiple-output (MIMO) urban scenario. The ray-tracing simulations rely on the conventional bidirectional ray-tracing technique for collecting the information about the incident rays and corresponding wavefronts, then a stationary-phase point approximation is employed for evaluating the antenna transfer function by the reciprocity integral. Maximum-ratio combining and regularized zero-forcing precoding schemes are considered in a single cell massive MIMO scenario where a frequency of 28 GHz is assumed. Relevant comparisons are made with a carrier frequency of 7 GHz as well, where the lower frequency transmission is found beneficial in noise-limited scenarios, due to the smaller transfer function attenuation. In interference-limited scenarios, the millimeter-wave transmission performs similar to the 7 GHz carrier for a comparable base station antenna arrangement, which can, however, be considerably smaller in absolute size. Another important advantage of the millimeter-wave band is of course the larger available absolute frequency bandwidth.

1 Introduction

Massive multiple-input multiple-output (MIMO) and millimeter-wave communication (mmWave) are considered to be two major technologies which may yield significant performance gains in next-generation cellular networks [1]. The massive MIMO concept relies on the use of a relatively large number of base station antennas compared to the number of user antennas, i.e., dozens of users are served by a base station comprising several hundreds of antennas. These large numbers of base station antennas are generally utilized for linear precoding, which considerably improves the network performance by suppressing intra-cell interference, as all the users (within the cell) simultaneously utilize the same frequency resources [2, 3].

Millimeter-wave communications relies on the use of

(carrier) frequencies, which are usually higher than 20 GHz and traditionally have not been utilized often in wireless networks until recently. The primary benefit of utilizing a large carrier frequency is the availability of frequency resources, i.e., a larger bandwidth can be allocated for data transmission, hence, data rate improvements of several orders of magnitude are theoretically achievable [4]. A notable issue concerning the transmission at mmWave frequencies is the so-called path loss problem, where the power delivered from one antenna to the other typically drops as the carrier frequency grows. In order to compensate for this power loss, utilizing a large number of antennas is a common solution in mmWave communications, which is also complementary with the notion of massive MIMO.

Although the massive MIMO and mmWave communication concepts have individually been studied in a comprehensive manner during the last decade, a unified mmWave massive MIMO concept and the peculiarities concerning the modeling and simulation of such systems have not been thoroughly addressed in the literature so far. In this study, ray-tracing simulations are utilized to characterize the propagation characteristics and achievable data rates for an outdoor mmWave massive MIMO scenario. In particular, the average downlink data rate among 16 users and the signal-interference characteristics under two different precoding schemes, maximum-ratio combining (MRC) and regularized zero-forcing (RZF), are investigated. A carrier frequency of 28 GHz is considered and relevant performance comparisons are presented with respect to a 7 GHz carrier as well.

2 Ray-Tracing Simulations

As the complexity of wireless networks has grown in order to manage the ever-increasing demand for mobile connectivity in the last decade, fast and accurate numerical modeling techniques, such as ray-tracing, have become essential tools. However, ray-tracing simulations involve certain challenges, especially in mmWave scenarios. In particular, the conventional ray-detection algorithms, which involve so-called reception spheres with a unidirectional ray launching scheme, usually require a large number of rays to be traced in order to achieve a good simulation accuracy. Furthermore, the number of ray launches should usually be increased at higher frequencies in order to avoid phase errors, thus, the computation time may become prohibitively large at mmWave frequencies [5]. Therefore, a bidirectional ray-tracing algorithm based on the calculation of stationary points (through which the exact rays having the minimum path length pass) is utilized in this study. Here, an exact ray is assumed to represent a path connecting two points in accordance with the Fermat principle of least time, which can be written as [6]

$$s = \int n(\mathbf{r}) dl, \qquad (1)$$

where *s* is the ray path length, *n* is the refraction coefficient and $\mathbf{r} \in \mathbb{R}^3$ is the coordinate vector. The considered method is a superior alternative to the conventional unidirectional ray-tracing for mmWave problems since a good accuracy can be obtained with a relatively small number of ray launches without any particular dependency on the frequency. There are three main steps involved with this approach:

- 1. Both receiver and transmitter antennas are utilized for ray launching, and the rays are collected on an interaction surface, which encloses the receiver.
- 2. The properties of the incident wavefronts (from both receiver and transmitter sites) and field-related quantities on the surface are computed.
- 3. The calculation of the antenna transfer function, which involves the evaluation of a reciprocity integral, is carried out by an asymptotic expansion, rather than full integration. First, stationary-phase points on the interaction surface are identified, and then the integral result is expressed by a single algebraic expression based on an asymptotic expansion.

An important consideration regarding the feasibility of this method for mmWave scenarios is the (correct) identification of stationary-phase points, since an inaccurate solution may yield significant phase errors and impair the final result, especially in scenarios with many ray paths.

The asymptotic expansion, which represents the result of the reciprocity integral under the assumption $k \to \infty$, can be given as [7]

$$I \approx \frac{2\pi}{k} \frac{f(\mathbf{r}_0)}{\sqrt{\det(\operatorname{Hess}(g(\mathbf{r}_0)))}} e^{jkg(\mathbf{r}_0) + j\frac{\pi}{4}} + \mathscr{O}(k^{-1}), \quad (2)$$

with

$$\oint _{\Psi} \left[(\mathbf{H}_{R} \times \mathbf{E}_{T}) - (\mathbf{H}_{T} \times \mathbf{E}_{R}) \right] \cdot d\mathbf{S} = \iint _{\Psi} f(\mathbf{r}) e^{jkg(\mathbf{r})} dS,$$
(3)

and

$$\Theta(\mathbf{r}) = \left[\left(\mathbf{H}_{R}(\mathbf{r}) \times \mathbf{E}_{T}(\mathbf{r}) \right) - \left(\mathbf{H}_{T}(\mathbf{r}) \times \mathbf{E}_{R}(\mathbf{r}) \right) \right] \cdot \hat{\mathbf{n}},$$

$$f(\mathbf{r}) = |\Theta(\mathbf{r})|, \ g(\mathbf{r}) = \frac{\arg(\Theta(\mathbf{r}))}{k}, \tag{4}$$

where \mathbf{r}_0 is the stationary-phase point, k is the wavenumber, Ψ denotes the interaction surface, $\hat{\mathbf{n}}$ is the unit surface normal vector for Ψ , $\mathbf{E}_{\{R,T\}}$, $\mathbf{H}_{\{R,T\}}$ are the electric and magnetic fields from receiver and transmitter antennas, and f and g are magnitude and phase functions, respectively. Note that the equations (2)-(4) involve only a single stationary-phase point with a single incident wavefront from each site (transmitter and receiver) towards the interaction surface. In case multipaths exist in the problem, the asymptotic expansion is applied to each individual wavefront pair separately. The stationary-phase point \mathbf{r}_0 satisfies $\nabla g(\mathbf{r}_0) = 0$, where g can generally be expressed in terms of the ray path lengths of the incident wavefronts. Hence the computation of stationary-phase points can be articulated as a minimization problem over g which is defined as the sum of two path lengths. The antenna transfer function hbetween a receiver-transmitter pair can then be written as [5]

$$h = -\frac{I}{i_R V_{gen}},\tag{5}$$

where *I* denotes the asymptotic expansion given in Eq. (2), i_R is the port current of the receiving antenna when it is on transmit, and V_{gen} is the generator voltage of the transmitting antenna. It should be noted that many wavefront pairs may not yield a stationary-phase point, though, it is beneficial to consider these contributions as well, since they implicitly represent diffraction effects while expensive UTD-based computations are avoided.

3 Massive MIMO

A downlink transmission case is considered with MRC and RZF precoders, which are expressed as precoding matrices

$$\mathbf{F}^{\mathrm{MRC}} = \mathbf{H}^{H},\tag{6}$$

and

$$\mathbf{F}^{\text{RZF}} = \mathbf{H}^{H} \left(\mathbf{H} \mathbf{H}^{H} + \alpha \mathbf{I} \right)^{-1}, \tag{7}$$

respectively, where **H** is the channel matrix, whose individual elements represent the antenna transfer functions (which can be calculated according to Eq. (5)), i.e., $\mathbf{H}_{(i,j)}$ indicates the channel between the *i*th user (each having a single antenna) and the *j*th base station antenna. It is assumed that the base station comprises 16-256 antennas (which is varied in the simulations) and 16 outdoor users are present in the scenario. The data rate for the *i*th user in the network can be then given by [8]

$$D_i = W \log_2\left(1 + \mathrm{SINR}_i\right),\tag{8}$$

with

$$\operatorname{SINR}_{i} = \frac{P_{\operatorname{TX}} \left(\mathbf{H}_{(i,:)} \mathbf{F}_{(:,i)} \right)^{2}}{P_{\operatorname{TX}} \left[\sum_{k=1, k \neq i}^{16} \left(\mathbf{H}_{(i,:)} \mathbf{F}_{(:,k)} \right)^{2} \right] + \sigma_{n}^{2}}, \qquad (9)$$
$$i = 1, \dots, 16.$$

where P_{TX} is the total transmission power, σ_n^2 is the noise power and *W* is the channel bandwidth. The precoding matrix **F** is either **F**^{MRC} or **F**^{RZF}, which are given in Eq. (6) and (7), respectively.

An urban scenario, which contains 16 buildings arranged as a 4×4 grid and covers a $120 \times 120 \text{ m}^2$ area, is utilized for the simulations. The dielectric properties are assumed to be constant throughout the geometry, i.e., $\varepsilon_r = 4$ for all the buildings as well as the ground.

The base station is placed in the middle of the building grid at a height of 5 m above the ground, while the user locations are determined randomly and placed 1 m above the ground. In order to prevent dominant channels to occur for particular users, it was ensured that the distance between the base station and the users is larger than 30 m. All the base station and user antennas are assumed to have vertically polarized isotropic patterns, and the antennas at the base station are arranged as a 16×16 cylindrical array. Two different antenna spacing configurations for the base station array were considered, in particular, a separation of λ for both carriers, and a separation of $\lambda/4$ for only the 7 GHz carrier case, which implies a restriction in terms of physical dimensions. Here, λ denotes the wavelength at the corresponding frequency.

4 Numerical Results

An investigation of the average user data rate with respect to the number of base station antenna elements was performed under two different precoding schemes, namely, maximumratio combining and regularized zero-forcing. Ray-tracing simulations were carried out at two different frequencies, 28 GHz and 7 GHz with up to 7 reflections. Certain diffraction effects were implicitly considered via the bidirectional ray-tracing where the wavefront pairs, which do not yield a stationary-phase point, are also included in the calculation of the transfer function. A cubic interaction surface with a side length of 1 m was utilized around all the receiver antennas. As stated before, two different array configurations were considered with an element spacing of the array at 28 GHz of λ , and at 7 GHz of λ and of $\lambda/4$. A fixed channel bandwidth of 100 MHz was assumed for the 7 GHz carrier, and a varying bandwidth of 100 - 400 MHz for the 28 GHz case. The overall transmission power at the base station is 1 W in all cases. The results are shown in Fig. 1 and in Fig. 2.

The results for the MRC precoder in Fig. 1(a) indicate that the average user data rate at 28 GHz is very close to that



Figure 1. Average user downlink data rate under MRC precoder for (a) fixed channel bandwidth of 100 MHz and (b) fixed array element spacing of λ .

at 7 GHz with λ spacing. On the other hand, the 7 GHz case with $\lambda/4$ spacing is of course considerably worse, in particular also for an increasing number of base station antenna elements. This implies an advantage for the 28 GHz carrier, if the physical dimensions of the base station are constrained (note that a $\lambda/4$ -array at 7 GHz can still be larger than a λ -array at 28 GHz, in general). If no physical constraints apply to the base station (i.e., an element spacing of λ is possible at any carrier frequency), significant improvements in data rate can still be achieved by utilizing a larger channel bandwidth at 28 GHz (shown in Fig. 1(b)), which may not be feasible at 7 GHz due to the scarcity of frequency resources. As the data rate scales almost linearly with the channel bandwidth, an improvement of 300% can be obtained.

The data rate and its variation over the number of TX antenna elements is with the RZF precoder and a fixed channel bandwidth of 100 MHz (as shown in Fig. 2(a)) very different from that with the MRC precoder. With the RZF precoder, the λ -spaced array at 7 GHz clearly outperforms the corresponding array configuration at 28 GHz (i.e., with λ -spacing), where the $\lambda/4$ -array at 7 GHz demonstrates a similar performance as the mmWave array. Since the RZF precoder suppresses the inter-user interference much more effectively than the MRC precoder, the data rate is improved drastically (compared to the MRC case) and is mostly determined by the signal power (constant noise power density per frequency bandwidth is assumed in all





Figure 2. Average user downlink data rate under RZF precoder for (a) fixed channel bandwidth of 100 MHz and (b) fixed array element spacing of λ .

cases). Hence, the transmission channel attenuation plays a crucial role, i.e., as the attenuation is stronger at higher frequencies due to the smaller effective area of the receiving antennas, the signal power as well as the data rate performance at 7 GHz is generally better as compared to the 28 GHz case in the absence of interference. Considerable improvements in data rate could nonetheless be achieved by increasing the channel bandwidth at mmWave frequencies, similar to the MRC precoding case, as shown in Fig. 2(b). It should be noted though that RZF precoding may not be suitable for large scenarios, which consist of many more users (and correspondingly a much larger number of base station antennas), as the computational overhead might grow drastically. In such cases, MRC precoding is usually preferable due to lower complexity.

5 Conclusion

A downlink data rate analysis for a millimeter-wave massive MIMO scenario was performed by means of a bidirectional ray-tracing algorithm, which utilizes stationaryphase point approximation. Two different precoders, namely, maximum-ratio combining and regularized zeroforcing were considered in a scenario with 16 users and up to 256 base station antennas. A 28 GHz carrier frequency was assumed and relevant comparisons with another carrier frequency at 7 GHz were presented. In particular, the larger available frequency resources at mmWave frequencies yield significant improvements in data rate for both MRC and RZF precoding schemes. When an identical channel bandwidth is utilized for both carriers, the major advantage of the mmWave case turns out to be the smaller physical footprint of the base station. The array configuration for the 7 GHz carrier may yield a better performance, in particular in noise-limited scenarios, but typically not without an according growth in the physical dimensions.

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