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for High Capacity and Sustainable
5G Ultra-Dense Cellular Networks

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WP2 – mmWave Communications

D2.2: INNOVATIVE MMWAVE TRANSMISSION SCHEMES, CHANNEL MODELS AND ANTENNAS

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Executive Summary

The next generation (5G) of wireless communication will improve upon existing mobile communication standards significantly while taking into account the increasing amount of stress on the limited amount of available spectrum today. The millimetre wave (mmWave) spectrum provides abundant and contiguous resources to solve this issue. The work package 2 (WP2) of the 5G-wireless project is devoted to the modelling of mmWave propagation channels and development of appropriate mmWave beamforming and signal processing techniques.

The deliverable D2.1 paved the way to better understand the state-of-the-art existing in mmWave channel measurements, characterization, modelling, antenna technologies, channel estimation, precoding algorithms and review of the evaluation and analysis tools. In this deliverable D2.2, advances to the state-of-the-art are made by proposing new methodologies in propagation channel characterization modelling, new antenna designs and innovative energy efficient architectures.

In chapter 2, we propose new mmWave channel modelling approaches which are based on channel measurements and simulations. Measurement campaigns have been conducted by Heriot-Watt university and characterization of the mmWave channel from the perspective of non-stationarities occurring in the channel have been modelled. Simulations have been conducted by Siradel by including the impact of blockages in outdoor dense small-cell networks.

In chapter 3, we propose new approaches towards the design of mmWave systems. These include new antenna prototypes which have been designed and prepared for fabrication by TTI Norte. An energy efficient architecture has been proposed by Universitat Politècnica de Catalunya (UPC) to better design the mmWave systems.

In chapter 4, an assessment of mmWave dense small-cells network is performed with the proposed hybrid stochastic – deterministic channel model considering various in-street blockages. The assessment provides an insight into the performance of the mmWave channel in different realistic outdoor scenarios.

In chapter 5, the scenario and approach for the channel data repository to be made available to the 5G wireless project partners is given. The aim of creating the repository is to enable the accessibility of realistic channel data obtained from the Siradel simulators for the project partners. This data can then be used for further analysis and evaluation by the partners in their own studies.
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<td>3G</td>
<td>3rd Generation of mobile communication systems</td>
</tr>
<tr>
<td>3GPP</td>
<td>3rd Generation Partnership Project</td>
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<tr>
<td>4G</td>
<td>4th Generation of mobile communication systems</td>
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<tr>
<td>5G</td>
<td>5th Generation of mobile communication systems</td>
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<tr>
<td>ABEP</td>
<td>Average bit error probability</td>
</tr>
<tr>
<td>AD</td>
<td>Amplitude detector</td>
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<tr>
<td>ADC</td>
<td>Analog-to-digital-converter</td>
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<tr>
<td>AoA</td>
<td>Angle of Arrival</td>
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<tr>
<td>AoD</td>
<td>Angle of Departure</td>
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<tr>
<td>APDP</td>
<td>Averaged Power Delay Profile</td>
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<td>APSK</td>
<td>Amplitude phase shift keying</td>
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<td>ARA</td>
<td>Active receive antennas</td>
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<td>ASR</td>
<td>Angular Stationary Region</td>
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<td>AUT</td>
<td>Antenna Under Test</td>
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<td>BER</td>
<td>Bit error rate</td>
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<tr>
<td>BPCU</td>
<td>Bits per channel use</td>
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<td>BS</td>
<td>Base station</td>
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<td>BW</td>
<td>Bandwidth</td>
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<td>CDF</td>
<td>Cumulative Distributive Function</td>
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<td>CIR</td>
<td>Channel Impulse Response</td>
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<td>CSI</td>
<td>Channel state information</td>
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<td>DFT</td>
<td>Discrete Fourier Transform</td>
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<td>DKED</td>
<td>Double Knife Edge Diffraction</td>
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<tr>
<td>DL</td>
<td>Downlink</td>
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<tr>
<td>DS</td>
<td>Delay Spread</td>
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<tr>
<td>EE</td>
<td>Energy efficiency</td>
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<tr>
<td>EIRP</td>
<td>Effective Isotropic Radiated Power</td>
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<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
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<tr>
<td>FCF</td>
<td>Frequency Correlation Function</td>
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<tr>
<td>FDTD</td>
<td>Finite Difference Time Domain</td>
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<tr>
<td>FnS</td>
<td>Frequency non-stationary</td>
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<td>FSR</td>
<td>Frequency Stationarity Region</td>
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<tr>
<td>GHz</td>
<td>GigaHertz</td>
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<tr>
<td>HPBW</td>
<td>Half power beam width</td>
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<td>HSA</td>
<td>High SNR approximation</td>
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<td>ICI</td>
<td>Inter-cell Interference</td>
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<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
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<tr>
<td>ISD</td>
<td>Inter-site distance</td>
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<td>ITN</td>
<td>International Training Networks</td>
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<td>KED</td>
<td>Knife Edge Diffraction</td>
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<td>Acronym</td>
<td>Description</td>
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<tr>
<td>LMDS</td>
<td>Local Multiple Distribution System</td>
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<td>LOS</td>
<td>Line Of Sight</td>
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<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
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<td>ML</td>
<td>Maximum likelihood</td>
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<td>MmWave</td>
<td>Millimetre wave</td>
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<td>MPC</td>
<td>Multipath Components</td>
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<td>MS</td>
<td>Mobile Station</td>
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<td>MSA</td>
<td>Moderate SNR approximation</td>
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<td>MU-MIMO</td>
<td>Multi-user MIMO</td>
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<td>NLOS</td>
<td>Non Line Of Sight</td>
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<td>O2I</td>
<td>Outdoor to Indoor</td>
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<tr>
<td>O2O</td>
<td>Outdoor to Outdoor</td>
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<td>PDF</td>
<td>Probability density function</td>
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<td>PS</td>
<td>Phase shifter</td>
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<td>Phase shift keying</td>
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<td>PSM</td>
<td>Pilot Symbols Matrix</td>
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<td>QAM</td>
<td>Quadrature amplitude modulation</td>
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<td>RF</td>
<td>Radio frequency</td>
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<td>RMS</td>
<td>Root Mean Square</td>
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<td>RSM</td>
<td>Receive Spatial Modulation</td>
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<tr>
<td>Rx</td>
<td>Receiver</td>
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<td>SAGE</td>
<td>Space Alternating Generalized Expectation</td>
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<tr>
<td>SC</td>
<td>Small-cell</td>
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<tr>
<td>SE</td>
<td>Spectral Efficiency</td>
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<tr>
<td>SINR</td>
<td>Signal to Interference plus Noise Ratio</td>
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<td>Spatial modulation</td>
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<td>SubMiniature version A</td>
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<td>SNR</td>
<td>Signal-to-noise-ratio</td>
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<td>Space Shift Keying</td>
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<tr>
<td>SVD</td>
<td>Singular Value Decomposition</td>
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<td>TDD</td>
<td>Time Division Duplex</td>
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<tr>
<td>TE</td>
<td>Transverse Electric</td>
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<tr>
<td>Tx</td>
<td>Transmitter</td>
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<td>UCA</td>
<td>Uniform Circular Array</td>
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<td>UL</td>
<td>Uplink</td>
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<td>US</td>
<td>Uncorrelated Scattering</td>
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<tr>
<td>UT</td>
<td>User terminal</td>
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<tr>
<td>UTD</td>
<td>Uniform Theory of Diffraction</td>
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<td>UWB</td>
<td>Ultra WideBand</td>
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<tr>
<td>VNA</td>
<td>Virtual Network Analyzer</td>
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<tr>
<td>XFDTD</td>
<td>X (Windows System) Finite Difference Time Domain</td>
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1. Introduction

A key candidate to be a part of the 5G technology is mmWave communication systems. With a frequency range from 30 – 300 GHz, the mmWave band has a lot of available spectrum that could ease the spectrum scarcity issues in existing communication models. This band is sometimes also inaccurately referred to begin from 10 GHz onwards however since similar properties can be observed at frequencies like 28 GHz and 30 GHz, this definition is widely accepted and used. Due to the availability of large spectrum, the bandwidth can also be relatively large which can support very high data rates. As the frequency becomes larger, the wavelength reduces and the interactions with the environment can change significantly. Insignificant in-street objects at lower frequencies (sub 6GHz) like vehicles, humans, vegetation, signage and urban furniture become significant due to the stronger attenuations and even blockage caused by the relative electrical size between the smaller wavelength and the objects. Some assumptions made at lower frequencies like stationarity in space, frequency and time may not hold at such larger frequencies and bandwidths. The expected high data rates will require a larger amount of hardware in the form of wide antenna arrays and thus many RF chains which in turn will consume large amounts of energy. And the process of design and fabrication of antennas at such high frequencies is also a complicated process since the physical size of the antennas reduces considerably. All these factors require new algorithms, channel models, antennas and experimental trials to better understand, characterize and fabricate the mmWave based systems. This deliverable D2.2, addresses the various enhancements and techniques required for the development of mmWave technology by making proposals towards various new channel modelling approaches, experimental trials, antenna designs and energy efficient architectures.

The new channel models, antenna designs and architectures for mmWave technology must work along with existing and upcoming sub-6GHz technologies like massive MIMO. This can be achieved by creating modular designs at mmWave such that the individual blocks can be combined to form any configuration based on the requirements. The channel models should also take this into consideration so that the combined systems work efficiently.

MmWave dense small-cell networks have their own unique scenarios and challenges which can be quite different from those of the traditional sub-6GHz frequencies. At mmWave frequencies, due to the small wavelengths, the relative electrical size of the interacting objects of the environment becomes important. This leads to mmWave specify scenarios in which blockage due to various obstructions have to be considered. For example, a user at a bus stop may temporarily be disconnected from the network when blocked by a bus in close proximity. Such access related mmWave-specific scenarios have to be considered in the channel models for accurate representation of the channel.

For the dense small-cells network to work efficiently, a backhaul network should be easily accessible to the small cells. This could involve challenges of its own as the data rates involved are much higher than existing models. Line-of-sight operation is often a requirement for fixed wireless access. Obstructions caused due to the topography of the scenario like hills and vegetation can cause significant limitations to mmWave based fixed wireless access solutions. For a backhaul solution based on mmWave technology, blockages by interacting objects like poles, metallic boards, urban furniture and overall vegetation have to be considered. These considerations have to be included in the new mmWave channel models to better characterize the channel.
To accurately model a radio (mmWave) channel, channel measurements are required. These measurements provide a reference to the actual channel information, so that an appropriate model can be correspondingly derived. Such channel measurements have been performed by Heriot-Watt University at various mmWave frequencies and derivations to mmWave channel modelling from an angular (spatial) domain perspective have been made in section 2.2.

The complexity and computational effort of a channel model is also a major concern. The channel models should be able to compute accurately the performance of a given channel quickly. Further, different channel modelling approaches like deterministic or stochastic have different pros and cons. A hybrid model however is able to exploit the pros of both deterministic and stochastic models. Such a hybrid stochastic-deterministic approach for channel modelling of mmWave dense small-cell networks including the effects of various obstructions has been proposed by SIRADEL in sections 2.3 and 4.

New antenna designs corresponding to the mmWave frequencies are required which have their own hardware related challenges. The mmWave antennas have been designed, validated by simulations and are being prepared for fabrication by TTI and the details can be found in section 3.2.

Due to the expected high data rates at mmWave frequencies and the higher power consumption in each RF chain, new energy efficient architectures are required. Such an energy efficient mmWave based MIMO transceiver architecture has been proposed by Universitat Politècnica de Catalunya (UPC), details of which can be found in section 3.3.

At mmWave frequencies, due to the small wavelengths the channel can vary significantly as the signal interacts with different objects in the environment. In order for the project partners to access channel data from industrial simulators, that has been calibrated and tested widely, Siradel has made available a channel data repository. This data has been extracted based on the requirements given by the partners. The scenario and further details of the channel data extracted can be found in section 5.
2. MmWave propagation channel

2.1. Introduction

Most PHY-layer techniques designed for 3G and 4G systems have been evaluated and refined through link-level or even multi-link simulations, using stochastic propagation channel models, from the simple tapped delay line models to the MIMO-compliant geometry-based bidirectional models like the ones proposed by the European COST 273 action [1] or the WINNER projects [2][3]. The deterministic approaches, which typically rely on ray-tracing or ray-launching, were primarily devoted to network coverage predictions and network radio-planning in complex environments. Using deterministic tools in test and validation simulation frameworks was more an exception. Today there is a growing interest for the introduction of deterministic or hybrid solutions at early research and final validation stages, in particular by equipment vendors working on 5G systems with large antenna arrays or investigating the above-6GHz spectrum. There are a few reasons for this. The extension of the stochastic channel models to fulfil 5G requirements is a challenge. The fact that channel properties strongly differ between the well-known sub-6GHz frequencies and the millimeter-wave (mmWave) spectrum is one difficulty. The need for accurate spatial consistency and 3D directions in ultra-dense network simulations are two other ones. New channel measurements are required; research activity is intensive today, but such measurements and subsequent model derivation necessarily take time. In addition, higher models complexity is expected.

Besides, the ray-based models are a mature technology, and comparisons to mmWave measurements generally show a good agreement. Deterministic propagation models usually yield a better correlation with measurements than stochastic models, especially in non line-of-sight conditions. Optimized software implementations and ever-growing computational performance (GPU, cloud) make this kind of ray-tracing solutions able to provide numerous and rapid channel realizations. A major constraint remains the use of a geographical map data to represent the main interacting objects in the environment.

The recent METIS channel model [4] proposes a map-based approach as an alternative to stochastic models which may not be applicable in all cases. The map-based model is composed of a fully-specified geographical data (so-called Madrid grid) on which a ray-based prediction is applied. Such model can generate all large-scale and small-scale channel properties or may be used to feed a stochastic modeling with some deterministic large-scale parameters. Besides, the MiWEBA European project has recently released a channel model with hybridization of deterministic reflected paths and stochastic components [5].

Here we have exploited the ray-based model described in [6] and adjusted to 60 GHz [7] to assess the performance of an outdoor mmWave small-cell network and predict the mmWave multi-path channel properties. Stochastic components have been added to the deterministic rays to reproduce the high sensitivity of the multi-link propagation scenario to non-static objects, i.e. the body of the user itself and large vehicles (e.g. buses, trucks or vans) driving in the streets. The blockage of the 60-GHz wave by the human body or large vehicles strongly impacts the link quality and dominant propagation directions, leading to changes in the cell selection, beam selection, interference levels and even the connectivity rate. Blockage by a human body or group of bodies has been intensively studied by the scientific community, through measurements and simulations, into different frequency bands.
including mmWave [8]. The authors already proposed a hybridization between indoor ray-tracing and stochastic human-body blockage in [9] at 2 GHz; the focus was then on the impact of persons moving around the user. The reported work, deals with a significantly different situation where the blockage comes from the user-body itself in the near field and later from a group of users in the far field.

The mmWave indoor channel is also characterized with the support of various measurement campaigns. The measurements have helped to better understand and characterize the non-stationarity regions related to the mmWave systems. This non-stationarity has to be considered in the upcoming channel models, without which the estimated channel properties may be distorted or inaccurate. Since the broad bandwidth resource is the key advantage of mmWave communication, the frequency non-stationarity of mmWave channels is addressed in the following content. Another key characterization of the mmWave communication channel is directional propagation. We found that the directional mmWave channels show the features of Markov states. By modelling the channel by including the Markov state information instead of directional information, the complexity of the channel model can be greatly reduced.

### 2.2. Measurement-based indoor mmWave channel characterization

**2.2.1. Measurement setup and scenario (LoS, NLoS)**

The channel measurements were performed in a basement environment, a big and almost empty room scenario, same as in [10]. The channel sounder consisted of a Virtual Network Analyser VNA) and a large virtual uniform circular array (UCA) with the radius 0.5 m. The channel was measured in

- A sub-6 GHz frequency band: 2-4 GHz,
- and two mmWave frequency bands: 14-16 GHz and 28-30 GHz

*Channel measurement performed in Aalborg University (Denmark)*
Figure 2 Approximation of linear antenna subarrays [10].

There are 750 frequency points within each of the frequency bands (frequency interval is 2.67 MHz). The volume of the measured channel environment is 7.85 m × 7.71 m, see Fig. 1. A Bi-conical antenna with frequency range from 2-30 GHz was used for both transmitter (Tx) and receiver (Rx), and the radiation pattern was omnidirectional in horizontal plane. The Tx bi-conical antenna was fixed at about 1 m height, and for the Rx side, the bi-conical antenna was moved along the trajectory of the virtual UCA at 1 m height. The distance between each two adjacent positions of the virtual UCA was 0.0044 m, which was less than \( \lambda /2 \) at 30 GHz. Both line-of-sight (LOS) and non-LOS (NLOS) measurements were performed, with metal board placed between Tx and Rx for creation of the NLOS situation.

Space-Alternating Generalized Expectation maximization (SAGE) [11] is used in the estimation of the channel parameters. Since the Rayleigh distance of the whole UCA is beyond 400 meters, 16 consecutive virtual antennas are chosen as virtual linear subarray in each sliding window along half of the virtual UCA in the data analysis, see Fig. 2. The size of virtual linear subarray is approximately 0.07 m; the corresponding Rayleigh distance (or Near field range) is shorter than the distance between Tx and Rx.

Channel measurement performed in Shandong University (China)

Other indoor mmWave channel measurements were conducted in Shandong University, China, in a static office environment as depicted in Fig. 3. In this measurement, Keysight N5227A vector network analyzer (VNA) and Keysight E8257D signal generator were used [12]. The intermediate frequency (IF) filter bandwidth of VNA is 1 kHz and the output power of the signal generator is 13 dBm. The measured frequency range is from 59 to 61 GHz with 401 frequency points (0.5 ns delay resolution). The standard horn antennas with 25 dBi gain and 3-dB beam width of 10° at 60 GHz were used in both Tx and Rx. The Tx antenna was placed on the antenna positioner at Tx1 and Tx2 positions. It rotated from 0° to 355° in azimuth angle, and rotated from 30° to 150°, in elevation angle. We assume the LOS path between Tx and Rx is the azimuth angle = 0° and elevation angle = 90°. The Tx rotation step is 5°, in both azimuth and elevation angle for each D-CIR measurement (72 × 25 D-CIRs in total), while the Rx horn antenna was hold still pointing to Tx all the time. The heights of Tx and Rx antennas were both 1.6 m. In the analysis below, we mainly focus on the data measured while Tx is located at TX1 and TX3 positions. We consider the measurement for the Tx positioner at TX1 position as LOS scenario, and that for Tx positioner at TX3 position as NLOS scenario.
2.2.2. Channel characterization with focus on the analysis of stationarity regions

A general averaged power delay profile (APDP) method was proposed in [13], it is used as a metric to determine the stationarity regions of channels in time, frequency, and spatial domains. This method is an extension of the former stationarity studies of channels in time and spatial domain [14]. Moreover, the general APDP method could be further extended to angular domain, i.e. angular APDP (A-APDP) method. In this section, we focus on the studies of frequency stationarity region (FSR) and angular stationary region (ASR) of the mmWave channels, since those researches related to mmWave channels are rarely seen in the literature.

**Frequency stationarity region**

For a static MIMO channel environment, if we measure the channel at a high frequency and at a low frequency respectively, the estimated cluster-maps from those two channel measurements could be very different, so as the statistical properties of the channels. We assume that the channels in each of the frequencies are frequency stationary, but they are in different FSRs. In case we measure the channel with a sounding signal that its bandwidth covers both high and low frequencies, then we consider the measured channel is frequency non-stationary (FnS).
The details of how to calculate the stationarity region of channel in frequency domain can be found in [13]. The idea is illustrated as in Fig. 5. First, we calculate the correlation coefficient of the APDPs with a distance of $\Delta f$ in frequency domain. Then, we decide a threshold to determine the frequency stationarity bandwidth of the channel (or channel sounding signal). We define that if the correlation coefficients above the threshed bandwidth of channel within such stationary bandwidth, then the channel is considered frequency stationary, otherwise, frequency non-stationary.

**Angular stationarity region**

In mmWave communication, the signals from Tx to Rx are normally transmitted in a certain direction, since beamforming technology is widely used to concentrate the power of the signal to one direction, compensating the high path loss. In this case, we should consider the directional propagation of channel, and due to the different channel situations in different directions of the environment, the channel impulse responses (CIRs) measured in different directions are dramatically different.
The estimation of the stationarity region of channel in angular domain is based on the APDP of directional channel impulse responses (D-CIRs). The idea is similar to those in time, frequency, and spatial domain in general APDP method, the procedure of calculating the ASR of channel is roughly the same as that of calculating FSR. The Fig. 6 illustrates the A-APDP method, and the details of calculation can be found in [15].

2.2.3. Channel modelling for indoor mmWave scenario focussed on modelling of frequency and angular non-stationarity

Frequency non-stationary channel modelling
In usual studies, the mmWave channels are normally considered frequency stationary, such as the meeting room, in a sub-6 GHz 4G wireless communication scenario, measured at 60 GHz (center frequency), by the sounding signal with 2GHz bandwidth [16]. However, the frequency stationarity assumption might not hold for some of the mmWave channel scenarios. There are two practical reasons which could cause the mmWave channel FnS. First, similar to UWB (Ultra Wide Band) channels, if the bandwidth of the channel is very broad, the statistical property of channel varies within the bandwidth of the channel [17]. Second, according to the diffuse scattering described in [18], if the scattering area (volume) of objects (scatterers) in the channel is comparable to the wavelength of sounding signals, then the interaction of the sounding signal with the scatterers could change significantly as the frequency of sounding signal changes. In this case, the stationary bandwidth could be much narrower than the bandwidth of the sounding signals, and the uncorrelated scattering (US) assumption in frequency domain is violated. This may happen for mmWave communications in certain channel environments, such as factory, warehouse, mine of coal/metal and chemistry laboratory, etc.

The idea of frequency non-stationary model is that, we split the frequency non-stationary channel into a few frequency stationary sub-band channels, and then we model each sub-band channels and add
them up as one. However, it is not realistic to model each sub-band channel as WINNER II model individually and add them up, the work load is very heavy in this case. Therefore, a novel frequency domain cluster evolution is proposed to simplify the model. It describes the relationship among those sub-band channels, and maintains a reasonable level of similarity for each two consecutive sub-band channels.

In the equation, sum of sub-band channels are denoted by sum of “C”, and in each sub-band channel, there are N clusters and M rays within each cluster. The modelling of first sub-band channel can refer to WINNER–II model, and the descriptions of the rest of the sub-band channels are using cluster evolution in frequency domain.

\[ h(\tau) = \sum_{c=1}^{C} \sum_{n=1}^{N} \sum_{m=1}^{M} F^{-1}[\delta(f - f_c)] a_{n,m} \delta(\tau - \tau_n - \tau_m) \delta(\theta - \theta_n - \theta_m) \]  

\[ 1 \]

Cluster evolutions in time and spatial domain are normally interpreted as moving clusters/Tx/Rx situations in standard channel models and other proposed channel models [19] – [22]. However, cluster evolution in frequency domain is quite different. In our proposed FnS channel model, we use tidal variation to metaphor the cluster evolution in frequency domain as in Fig. 7.

Assume the high tide and low tide situations representing two estimated cluster-maps of two sub-band channels in the data analysis, then how to describe the size of the clusters and the number of clusters are the focus for the cluster evolving from one sub-band channel/FSR to another. The idea is based on that a cluster is a bunch of estimated multipath components (MPCs), which reflect the outcomes of incoming waves interacting with a bunch of correlated scatterers (objects). It is possible to split one cluster into a few clusters, or vice versa, which depends on how we define the correlated scatterers in the estimation of channel parameters, and how we model the inter-cluster and intra-cluster parameters.

The proposed solution describing cluster evolution in frequency domain is as follows:

- Track the survival probability of each single cluster when it evolves from one sub-band channel into another one.
- Pre-define the number of clusters for all sub-band channels. When evolving to another sub-band channel, only generate new cluster if one cluster is dead or if the number of clusters of the sub-band channel increases.
- Update both inter-cluster and intra-cluster parameters by the trends when evolving to another sub-band channel.

The survival probability follows the exponential distribution, i.e. \( \exp(-\lambda \ast \text{number of sub-band channels}) \). In the simulation, in order to approximate the measurement data (from the channel measurement in Fig. 1), the rate parameters \( \lambda \) should be between 0.05 to 0.1 in the sub-band channels within 2–4 GHz band, and be between 0.025 to 0.05 in the sub-band channels within 28–30GHz band. We have observed that the clusters can survive 1–15 sub-band channels (25 in total) in the cluster evolution within 2–4 GHz band, the clusters survive for all 1–3 sub-band channels within 28–30 GHz band (only have 3 sub-band channels in this case). The approximation of the simulated frequency correlation function (FCF) to the measured FCF in both 2–4 GHz band and 28–30 GHz band can be found in Fig. 8.

![Image]

Figure 8 FCF approximations of the simulated FnS channels to the measured FnS channels. The left plot illustrates the FCF approximation of the Fns channel in 2–4 GHz band, and the right plot is related to 28–30 GHz band.

**Directional mmWave channel modeling (angular non-stationarity)**

As former description of the mmWave communications, the signal transmissions between Tx and Rx in mmWave communication are normally directional, therefore the study of directional knowledge of the channel is very important.

In the channel measurement in Fig. 3, we have found that the D-CIRs can be sorted into three types LOS scenarios, which are related to the different propagation situations of mmWave channel in different directions. The three types of D-CIRs are illustrated as in Fig. 9(a), Fig. 9(b), and Fig. 9(c). In the NLOS scenarios, we can sort the D-CIRs into two types, which have the similar feature with the Fig. 9(b) and Fig. 9(c), and therefore we consider them as a subset of three types of D-CIRs LOS scenario.
In the data analysis of mmWave channel measurements, we have estimated the ASRs of the channel, as well as the root mean square delay spread (RMS DSS), K-factors, and the peak power gains based on the D-CIRs in both LOS scenario and NLOS scenario. According to the measured channel environment in Fig. 3, it is obvious that due to the strong LOS path between the Tx and Rx, the D-CIRs measured in those directions are with high similarity, and the ASRs of channel in those directions are very high. While in other directions where the sounding signals are fully scattered, the NLOS components are dominant in the measured D-CIRs, and the ASRs of channel become dramatically smaller. It also make sense in the findings that, in the directions where the ASRs are very large, the measured D-CIRs in corresponding directions are with small RMS DSSs, high K-factors, and high peak power gains. See Fig. 10 and Fig.11 (LOS scenarios, as an example), in those directions where the ASRs of channel are very small, the measured D-CIRs in corresponding directions are with large RMS DSSs, low K-factors, and low peak power gains.
Note that the horn antenna radiation patterns are embedded in the measured data, i.e. measured D-CIRs, because there is no suitable method existed to remove it for our cases at the moment. Therefore, the channel parameters estimated are effected by the channel measurement set up, such as antenna pattern, the number of scanning steps, and the size of scanning step, etc.

\[
h(\tau) = \sum_{l=1}^{L} a_l \delta(\tau - \tau_l) \delta(\phi - \phi_l) \delta(\theta - \theta_l)
\]

(2)

Based on the three Markov states of D-CIRs, herein we propose to use Markov states to simplify the conventional description of the directional channels as above equation. Then, by the knowledge of the Markov state, such as the probability of each Markov state (limiting distribution), or the angular locations of the Markov state, we can model the directional mmWave channel by a group of joint mmWave channel models.

By the peak power of each D-CIR, if we choose -62 dB and -80 dB peak power gain as the limits of those D-CIRs sorted into three Markov states. We let the type 1 D-CIRs as in Fig. 9(a) belong to LOS
state, and the type 2 D-CIRs as in Fig. 9(b) belong to Outage state, and the type 3 D-CIRs as in Fig. 9(c) belong to NLOS state. The one step transition matrix and limiting distribution of Markov states can be estimated, and they can be found in the Fig. 12. We also found that, those three Markov states are in different angular locations. An example of the angularly separated Markov states on the Azimuth plane can be found in Fig. 13.

![Figure 12 Transition Matrix and limiting distribution, based on LOS scenario measurement in Fig. 3.](image)

![Figure 13 Angular separated three Markov states on horizontal plane, based on LOS scenario of measurement in Fig. 3.](image)

From the estimated ASRs of channel in Fig. 10, we found that the D-CIRs in LOS state belong to one ASR, and the D-CIRs in NLOS state and Outage state are over a few ASRs. With the respect of that, we model the D-CIRs in LOS state by stationary channel model, and model the D-CIRs in NLOS state by non-stationary channel model. Since the powers of D-CIRs in Outage state are too weak to maintain
the communications, we consider them as invalid channels, the powers of those directional channels are “zero”.

The system level simulation of the directional mmWave channels is illustrated as the procedure in Fig. 14. We assume the directional channels in LOS scenario are with three Markov states, and we model them by joint channel models that consist of one stationary model and one non-stationary model. Similarly, for the directional channels in NLOS scenario, they are with two Markov states, and we model them by the joint channel models that consist of only one non-stationary model.

The generation of stationary model is based on the original Saleh-Valenzuela (SV) channel model [23], [24] in Fig. 15 [25], there is no angular information in the model, it may be considered as the modelling of a SISO channel. All the parameters used to generate the channel coefficients are constants based on the estimated statistical parameters from the data of channel measurement. In the generation of
non-stationary model, we let one or a few statistical channel parameters, such as K-factors, in the stationary model following normal/log-normal distributions, and those parameters are randomly generated in each run of the simulations. The rest of procedure of generating the channel coefficient of non-stationary channel model is exactly the same as those of generating the channel coefficient of the stationary model.

2.2.4. Considerations and improvements required from previous models

The broad bandwidth of the mmWave channel is the key feature of the mmWave communication. However, the very broad bandwidth may violate the stationary assumption, and we need to consider the mmWave channel is frequency non-stationary in some circumstances. We need to be alert to that in the data processing of the channel measurements and the channel modelling, otherwise, the channel properties estimated or the modelled mmWave channels may be inaccurate or distorted. The frequency non-stationary channel modelling could be considered as a complementary part of previous channel models. The cluster evolution in frequency domain can be considered as an extension of cluster evolution in time and spatial domains.

The directional propagation of mmWave signals is another important feature of the mmWave communication, and the directional mmWave channel shows the feature of Markov states. In this case, we can use the Markov states to substitute the directional information of the channel models, and each directional mmWave channel model can be considered as modeling of a single-input-single-output (SISO) channel. By such approach, the overall complexity of the mmWave channel models could be greatly reduced. This channel modelling approach could be considered as an alternative in the future standard mmWave channel models.

2.3. Hybrid Stochastic-Deterministic channel modelling

Dense small-cell mmWave networks are a promising 5G solution, however many technical challenges like those related but not limited to antennas, signal processing and networking are still to be solved. Strong propagation attenuation in the mmWave band is obviously a major constraint for the small-cell coverage, which is consequently restricted to few tens of meters. But it also provides high isolation between cells, and thus leads to low interference levels. These are the two sides of a same coin that must be properly taken into account in the network design problem. At higher frequencies like mmWaves, the relative size of the objects and the signal leads to a much more complex channel requirement, which is affected by various kinds of in-street obstructions like human bodies, large vehicles, urban furniture and other interacting objects in the environment.

Blockage by a human body or group of bodies has been intensively studied by the scientific community, through measurements and simulations, into different frequency bands including mmWave [8]. Siradel already proposed a hybridation between indoor ray-tracing and stochastic human-body blockage in [9] at 2 GHz; the focus was then on the impact of persons moving around the user. Here we deal with a significantly different situation where the blockage comes from the user-body itself. The implemented model is presented and illustrated in section 3. The obstruction by a bus was analyzed in [7], using the same ray-tracing tool.

The studied ray-based model predicts the propagation of multi-paths, including the angles of departures (AoD), angles of arrival (AoA) and field strengths, into suburban or urban environments.
Specular contributions, i.e. multi-paths that result from reflections or diffractions, are computed. Diffuse scattering is another allowed contribution, but it was not enabled here. Indeed its characterization at 60 GHz is still an on-going work in the propagation community. As its strength is generally smaller than first-order reflections, we do not expect any major error.

2.3.1. Blockage due to vegetation

The importance of the environmental details was pointed out in [2] making the use of high-resolution building maps mandatory even in theoretical studies (in order to assess realistic shadowing and interference situations) along with the presence of vegetation as a key feature. Propagation through foliage suffers from a loss that basically increases proportionally with the log of frequency [3]. Its impact is severe at mmWave frequency bands: 11 dB/m can be extrapolated from the ITU curve [3]. The diffraction above and around the vegetation blocks is a major propagation mechanism (for large vegetation) along with transmission through the foliage. The woods can be viewed as almost opaque obstacles like buildings, while the isolated trees and hedges need to be carefully considered. The ray-based propagation model predicts two contributions in presence of vegetation: the first one goes through the vegetation; the second one is the diffraction by the vegetation obstacle (using the knife-diffraction method). Only the strongest contribution is preserved. The implemented technique supports the presence of multiple obstacles along each ray path.

Evaluating the impact of vegetation was part of the study in [2], based on a single small-cell simulation, and visual comparison of coverage maps. The study is extended here; a larger number of small-cell locations are considered, and path-loss statistics versus distance are extracted. In addition, an issue in the implementation of the propagation loss through or above multiple vegetation obstacles has been solved.

The study was conducted in a real European environment with a mix of urban and suburban areas. Ten small-cells were created with an antenna height of 6 meters above the ground (typically installed on lampposts). Locations were selected so they are representative of various possible situations: some are placed close to a crossroads; others in the middle of a street; and others in a large square. Predictions run with different vegetation kinds of representation are compared: the first considered geographical map data does not include any information on vegetation; the second one integrates the largest vegetation blocks (surface greater than 300 m²) as sometimes available in high-resolution map data used today for radio-planning; and the last one represents all vegetation details, including the individual trees. The small-cell locations were chosen from the first geographical map data, i.e. without any knowledge on the surrounding vegetation, in order to not bias the evaluation study. The small-cells were moved few meters away from the original position when they appeared to be close to a tree.

The power maps predicted from the small-cell being the most impacted by vegetation are given in Fig. 16. The displayed power is relative to the free-space received power at 1 m; this facilitates the comparison of shadowing effects at different frequencies, i.e. 2.4 GHz and 60 GHz in this particular scenario. The predicted maps clearly illustrate how critical is the presence of vegetation in mmWave band, even isolated trees.
The results from all ten small-cells are processed to extract the statistics given in Fig. 17. The median received power and two percentile values (at 10% and 90%) are computed on successive distance intervals of 5 meters (the considered distance is the one measured in the horizontal plane). The graphs show how the prediction of the received power versus distance is changing when introducing vegetation details in the geographical map data. The maximum observed difference on the median received power due to the vegetation modeling is 4 dB at 2.4 GHz (distance = 32.5 m), while it is 23 dB (distance = 37.5 m).
2.3.2. Blockage due to large vehicles

At mmWave frequencies, due to the small wavelength, the electrical size of most in-street objects becomes significant. Channel models considering buildings and vegetation [7] can be predicted using purely deterministic approaches, using ray-tracing or ray-launching tools as the positions of these objects are fixed. However, to model in-street objects which are more dynamic in nature (not fixed) like vehicles and humans (self and crowd obstructions) some stochastic components are required for the cost-complexity tradeoff. A hybrid model is proposed, as shown in Fig. 18, where the obstruction by large vehicles and the self-obstruction caused by user body are stochastically introduced into a ray-based deterministic model. To evaluate the proposed model, traditional scenarios based on real city-environment map data are included along with mmWave specific scenarios like fixed users e.g. user at a bus stop.
Most in-street objects like vehicles, pedestrians, lampposts, urban furniture etc. can obstruct the dominant path at mmWave frequencies. The impact of the obstructions may sometimes not be dramatic due to multipaths created by surrounding buildings or large objects. A scenario in which obstruction by a large vehicle (for e.g. bus) of a user is shown in Fig. 19. Here the different propagation paths impacted by the large vehicle obstruction are shown. In Fig. 19, it can be seen that both the direct path and the multipath components can be obstructed by the large vehicle with relatively lower power levels as compared to the path that has not been obstructed. The impact of the obstruction caused by large vehicles and user bodies is illustrated through simulations in section 4. In this section, the focus is on the models that have been used to calculate the various blockage impacts caused by the in-street objects.
2.3.3. Blockage due to user-body self-obstruction

The near-field self-obstruction caused by the user body itself is also of great interest as this kind of obstruction is omnipresent. Various diffraction models have been used in literature [4][8][11]. The Knife Edge Diffraction (KED) models have been widely used to calculate the loss due to human body obstruction [4][8]. And the double knife-edge model (DKED) is found very representative of the reality. It has been proposed in the METIS model [4] to consider the human body as an infinitely long screen with two edges. However, the knife edge model underestimates the shadowing when the antenna is close to the obstructing body. A more appropriate approach is to consider the human body as a cylinder with uniform dielectric properties.

At 60 GHz frequency, the electrical size of the human body is relatively large and the body must be considered as a rounded obstacle instead of a sharp edge. Then the Uniform Theory of Diffraction (UTD) for rounded objects or the rounded edge model by ITU-526 [10] can apply. In particular, the UTD-based creeping wave model was shown to perform well as compared to the exact solution [11], and is proposed for on-body channel predictions.

For implementation simplicity, we have chosen the [10] formulation for our model. An extra attenuation due to a curved obstacle is added to the KED loss. The relevance of this model is well-known for large range propagation, but we decided to evaluate its accuracy for the short-range obstruction that is considered here. It is compared to the creeping wave model results published in [11]. In this deliverable, the human body is modelled as a cylinder of radius $r=0.2$m. Omni-directional antenna is considered for the receiver which is at a distance $d=0.205$m from the centre of the cylinder. The receiver is oriented at different angles $\phi$ as shown in Fig. 20 (a). Frequency is 60 GHz. The obstruction loss from the single knife-edge, creeping wave, and the ITU-526 rounded edge models are plotted together in Fig. 20 (b), showing good agreement between the two latter ones, and strong underestimation from the first one. Note the creeping wave model predictions on this scenario have been validated against measurements [11].

In [9], the impact of the human body self-obstruction was studied for mmWave networks. Self-obstruction is caused due to the user’s body that could obstruct the multipath propagation degrading not just the received power but also the interference, which is desirable. Since the human body self-obstruction is omnipresent, it is used in all simulations conducted for the human body crowd obstruction studies as well. In the proposed implementation here, the body orientation is random and the distance between the body centre and the antenna is $d=0.4$m. In this situation, and if we assume the base station is far away from the user, the shadow region is 60° wide, thus covering 16.7% of the plane.

We decided to consider only one diffracted edge instead of two. The loss in the shadowed area where both diffracted paths are of same order is very strong and the improved accuracy offered by the double-edge approach will not have any impact on the simulations. The red plot in Fig. 20 (b) illustrates the loss that is computed in the remaining of our study, i.e. with $d=0.4$m.
2.3.4. Blockage due to human crowd-obstruction

The effect of human obstruction due to a crowd of people surrounding a user is considered next. To model the human bodies as a crowd surrounding the user, groups of human bodies (clusters) of different sizes are created. Each cluster of a given size is assumed to contain a certain number of closely located human bodies and the cluster itself is considered to be an opaque obstruction. Diffraction occurs at the side and the top of the cluster. The KED model is used in case of the clusters as the receiver is located further away from the cluster as compared to the user body self-obstruction case. The average height of the crowd clusters is considered to be 1.7m and the user equipment is assumed to be located at 2/3 of the average height for e.g. user holding a smartphone. Reflection and scattering on the bodies is not considered.

The human crowd obstruction is applied on the direct path and all multi-paths initially generated by the ray-based simulation, as illustrated in Fig. 21. The clusters surrounding the user are generated in

Figure 20 (a) Shadow region caused by the user-body; Antenna distribution considered in the analysis is represented by blue dots. (b) Obstruction loss from the user-body.

Figure 21 Example of obstruction of an initial reflected path by a body cluster, leading to three diffracted paths (from top and sides of the cluster).
a random way and (in current version of the implementation) independently from one user to the other. The shadow region created by each cluster depends on the distance between the user antenna and the cluster, size of the cluster and the difference between the heights of the antenna and cluster.

Fig. 22 shows the relation between the horizontal shadow region and the distance to the antenna for various cluster sizes (radii). As the distance between the cluster and the antenna increases the shadow region significantly becomes smaller. Therefore, for a cluster size of radius 1.0m the shadow region is less than 40°, which is just 11.1% of the horizontal plane. Further, when the impact of the vertical direction is also added this distance reduces further. In our simulations, the impact due to the clusters beyond 4m is very small especially in case of the smaller sized clusters.

![Figure 22 Shadow region generated by the body cluster, depending on its size and distance to the user antenna.](image)

A random number of clusters are stochastically positioned around the user using a Poisson process based random variable generator. Fig. 23 illustrates how the clusters impact the Power Delay Profile of the two signals S1 and S2 captured by the same user equipment but coming from two different base stations. The multi-path propagation channels are composed of several contributions arriving with specific delays and angles. The power of the paths arriving in the shadow region (or more precisely: in the close vicinity of the shadow region, as significant diffraction loss occurs even without optical path obstruction) are reduced in a similar way for both signals.
The Poisson parameter (\(\lambda\)) which is the expected number of occurrences is varied to achieve different crowd densities. The following scenarios are considered.

**Very crowded scenario:** the crowd is relatively dense in close vicinity of the user. It is representative of open-air events, concerts, crowds created during arrival and departure of flights, trains etc. Larger sized clusters located around 2m from the user are considered.

**Moderately crowded scenario:** the crowd is moderately dense and located at a larger distance from the user. This scenario is representative of crowded shopping streets, public beaches, main road pedestrian intersections etc. The simulations are conducted such that medium to large sized clusters are considered in a range of around 3m from the user.

**Crowded scenarios:** the crowd is sparse and can be located at a distance away from the receiver. This could be in the case of users waiting at a bus stop, tourist spots, waiting areas of airports or train stations, universities, hospitals etc. The simulations are conducted with small to medium sized clusters located up to 4m away from the receiver.

The blockage channel modelling approaches presented in this section are used for the performance evaluation of a small-cell network to analyse the impact of the various blockages like large vehicles, user-body self-obstruction and human-crowd obstruction. In section 4, a 60 GHz small-cell network consisting of 18 small-cells is designed in an urban European city environment. System level analysis
of the designed network is conducted using Monte Carlo simulations. A part of the same network is further used in section 5, to provide MIMO channel samples to the 5Gwireless project partners based on their requirements. The SIRADEL ray-tracing implements the spherical-wave extrapolation for MIMO system predictions, making the model suitable for large antenna arrays even at mmWave frequencies.

Both stochastic and deterministic (or hybrid) based channel modelling approaches require validation through rigorous measurements. The SIRADEL ray-tracing tools are regularly compared and calibrated to mmWave measurements performed by various academic and industrial partners or customers. The various parameters involved in the mmWave channel modelling like the different types of impact due to the vegetation are constantly refined in the tools. The process of conducting measurements is both time consuming and expensive; which makes this work ongoing, as and when more measurements are available the channel models will be further refined. Validation of the human body blockage impact would be good, however such measurements are not planned in the perspective of this project. At SIRADEL, the mmWave activity will now be mainly focused on the exploitation of the modelling approaches presented in this deliverable with multi-user beamforming based network simulations and generation of channel samples.
3. MmWave Systems

3.1. Introduction

Over the last few years there has been an increased interest in the millimeter wave frequency band owing to the vast bandwidth that it offers. MmWave propagation suffers from severe path-loss compared to the sub-6 GHz. However, the small wavelengths of mmWave signals can permit packing large number of antennas elements in small size that enables using the massive antennas arrays at the base station (BS) and also at the user terminal (UT) to maintain high beamforming gain.

The aspect of antenna design for millimeter wave requires a different approach from the design principles usually followed in the sub 6 GHz range. This stems mainly from the smaller wavelengths that inherently result at the millimeter wave frequencies. On-chip antennas, in-package antenna, parabolic antennas, waveguide array antennas, lens antennas are some of the antennas that have recently gained interest owing to their special characteristics pertaining to millimeter wave antenna design.

3.1.1. Antenna challenges

On-chip and in-package antennas that use dielectric substrates in their design often resort to multilayer technologies [26] to minimize the losses seen by the substrate at higher frequencies. The design of antennas in the sub 6 GHz range uses the characterized properties of the substrate at those frequencies. However often at higher frequencies the substrate characteristics need custom characterization and the substrate can often be lossy at these high frequencies. A significant portion of the RF power is lost to the simple interconnect and dielectric profiles. Therefore, design of planar antennas on substrate technology at millimeter wave can be challenging.

Parabolic reflector based antennas are a good promise at the millimeter wave frequency [27]. They have been extensively used in satellite communication and fixed wireless communication. Their adoption to the mobile millimeter wave communication scenario may be hindered where spatial footprint is of concern.

Waveguide based phased array antennas have been explored in application areas requiring azimuthal/elevation beams-scanning [28]. They are also attractive owing to their higher power handling capabilities. The design of high precision waveguide components at millimeter wave frequencies is necessary owing to the smaller wavelengths.

Lens antennas based on waveguides are an attractive option in millimeter wave frequencies [29]. They can excite multiple beams using a simple waveguide port excitation and can result in high gain beams owing to the focusing aided by the lens. The beamformer for this purpose can also be simpler. The design of the lens based waveguide at the high frequencies can be challenging with concerns such as the net bandwidth obtained, the beamwidth and the scanning range.

3.1.2. Signal processing challenges

Several signal processing challenges arises because of communicating at mmWave frequency band.

- Hardware and power consumption constraints
At sub-6 GHz, the MIMO signal processing occurs in the baseband and these schemes require one radio-frequency (RF) chain and analog-to-digital-converter (ADC) per antenna [32]. Unfortunately, these devices are expensive and power consuming at mmWave band. Moreover, a practical implementation to complete RF chain per antenna for large MIMO systems at mmWave is limited by sizing constraints. Therefore, the traditional baseband signal processing methods become challenging with massive MIMO systems at mmWave frequencies.

- **Channel estimation**

The number of needed pilot symbols for the conventional channel estimation techniques increase with number of antennas [33] and thus these techniques become challenging with large antennas arrays. Also, the spatially sparse nature of the mmWave channel can be exploited for the development of low complexity channel estimation algorithms.

- **Relay-based blockage minimization**

MmWave propagation in urban cellular systems is very sensitive to the blockage caused by the human-body. At 28GHz and 50m TX-RX distance, a full blockage can occur with probability 60% [34]. Therefore, blockage mitigation techniques such as cooperative transmission and relay assisted networks can be applied for mmWave systems. blockage mitigation technique based macro diversity with multiple BSs is studied in [35] where a user chooses one of the BSs within a communication range to mitigate the blockage effect.

- **Channel variations**

The channel fluctuation is a limiting factor for wireless communications at high velocities because it reduces the channel coherence time. For example, at 60 Km/h and 60GHz, the channel coherence time in order of hundreds of microseconds while for today's cellular systems it is in order of milliseconds [36]. Thus, the channel changes faster in mmWave communication, and consequently, the channel needs to be estimated more frequently.

In the following, we discuss the massive MIMO transceiver design challenges in mmWave bands. Next, we summarize the pros and cons of the state of the art. After that, we present novel energy-efficient architecture for receive spatial modulation in large MIMO systems that works with high performance at mmWave bands.

### 3.1.3. Need for new architectures in mmWave bands

In contrast to Deliverable 2.1 [37] where the state of the art for the precoding and channel estimation was proposed, we detail several massive MIMO transceiver architectures as a motivation for the proposed architecture. The conventional MIMO architecture, developed for centimetre-wave based communications, depends on a fully digital structure where each antenna connected to a RF chain and ADC, as it is shown in Fig. 24.
However, the cost and power consumption of a RF is highly dependent on the spectrum band of interest. For example, the power consumption of the different transceiver components at different frequencies [38]-[39]-[40] are illustrated in . At $N_r = 16$, the receiver power consumption of the architecture in Fig. 24 equals 1W at 2.5GHz and 6.2W at 73GHz in addition to the power consumption by the baseband processing. The previous example shows that the receiver power consumption at 73GHz increases six times more than at 2GHz so the conventional MIMO becomes very challenging with large arrays and high frequencies due to the substantial increase in complexity and energy consumption. In the following, we present many state of the art MIMO transceiver architectures that have been developed to reduce the cost and the power consumption.

**Analog MIMO**

Analog MIMO is the simplest transceiver design that can be applied for mmWave systems. In this system, each antenna element is connected to analog phase shifter and all the phase shifters are connected to single RF chain as shown in Fig. 25. The phase shifters weights are digitally controlled according to the beamformers design. Analog precoding can support single-user and single-stream transmission because using only one RF chain at the BS but it is not possible to apply it for multi-user MIMO systems and it is usually used for beam steering [32]-[41]-[42].

<table>
<thead>
<tr>
<th>Device</th>
<th>2.5 GHz</th>
<th>28 GHz</th>
<th>73 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>LNA</td>
<td>5</td>
<td>10</td>
<td>30</td>
</tr>
<tr>
<td>RF Chain</td>
<td>10</td>
<td>40</td>
<td>60</td>
</tr>
<tr>
<td>ADC</td>
<td>50</td>
<td>200</td>
<td>300</td>
</tr>
</tbody>
</table>
**Hybrid analog and digital MIMO**

In hybrid MIMO systems [32], analog devices (phase shifters) are used to reduce the number of RF chains, but still seize the use of many antennas to maintain high gains. As shown in Fig. 26, the beamforming process of the hybrid architecture is divided between analog and digital domains with number of RF chains smaller than number of antennas where it allows multi-user transmission [32]. In [43], the hybrid architecture is introduced for centimetre-wave frequencies. The design algorithms described in this paper can be used for mmWave frequencies. However, they did not consider the properties of mmWave propagation in designing the precoders and combiners.

In [44], the spatially sparse nature of mmWave propagation is leveraged to simplify the hybrid transceiver design. The spectral efficiency of the fully connected hybrid architecture in Fig. 27 approaches the fully digital MIMO with small number of RF chains [38]. However, each RF chain is connected to number of phase shifters equals to the number of receive antennas. Thus, the number of phase shifters increases rapidly by increasing the number of RF chains. Consequently, large number of phase shifters may consume greater power than RF chains and ADCs. In [38], many low complexity hybrid systems have been studied based on phase shifters and switches but their spectral efficiency gap with respect to conventional MIMO is large.
**MIMO with low precision ADCs**
The high-resolution ADC at mmWave frequencies is the most power consuming device (200mW at 28GHz) [32]. The ADC power consumption increases exponentially with its quantization bits [32]. Thus, the MIMO receiver design based on low precision ADCs shown in Fig. 28 is an alternative technique to reduce the power consumption. For example, the 1-bit ADC at sampling rate 240GS/s consumes 10mW [45]. Nonetheless, the use of low bit ADCs hampers the availability of perfect channel knowledge and entails a reduction of the spectral efficiency.

**MIMO based spatial modulation**
Spatial modulation (SM) schemes have been reported to reduce the number of RF chains and to achieve high throughput. In SM systems [46], the BS transmits spatial symbol by activating subset of the transmit antennas and as a result the number of RF elements and ADCs are less than the number of transmit antennas. The basic structure of SM is called space shift keying (SSK) [47]. In SSK, only one transmit antenna is active at given time slot. The input bit stream is converted to index of active transmit antenna, and the SSK receiver detects this index. For the sake of increasing the number of transmitted bits per channel use (BPCU), the active antenna transmits data symbol [46]. For further increasing of the transmitted BPCU, more than one antenna is active simultaneously [48]. However, SM techniques suffer from small antennas gain because most of the transmit antennas are silent.
Further, the UT cannot detect the set of active transmit antennas precisely with low rank channel matrices.

On the contrary, in receive spatial modulation (RSM) systems, the BS transmits spatial symbol by activating subset of the receive antennas [49]. The SM/RSM systems can simplify the MIMO transceiver and achieve high data rates. However, conventional RSM methods use fully digital receiver architecture and zero forcing precoding so it suffers from high performance degradation under low rank channel matrices. Thus, traditional RSM methods are not convenient for mmWave systems. Recently, a novel RSM architecture have been reported in [50] for indoor 60 GHz communication. Generalizing the concept of SM/RSM for high frequency communication is an interesting open research problem.

### 3.2. Design and fabrication of mmWave antennas

Millimeter wave beamforming antennas necessitate a completely different approach to their design and fabrication which introduces many challenges. An attempt is made to detail several of these challenges here though not exhaustive:

1. **Simulation**: In the design of antennas full wave EM solvers based on the Finite Difference Time Domain (FDTD), X (windows system) Finite Difference Time Domain (XFDTD) methods and also the Method of Moments, Integral Equation method are used. These methods and especially the FDTD solver are mesh based solvers which work on discretization of Maxwell’s Equations. The amount of mesh cells generated and the type of mesh cells depend on the nature of the geometry of the design and more strongly on the frequency of operation. For higher frequencies owing to the smaller wavelength the mesh cells size is generally high and this results in consumption of significant simulation time and in some problems the solution may take several days even with multi-threading. Therefore, to arrive at a solution through simulation can also be significantly challenging.

2. **Design**: The design of antennas at the millimeter wave poses challenges pertaining to antenna pattern stability over a wide band, tapping of sufficiently large bandwidths, stable gain over frequency behaviour, phase center localization among a few to mention. The design of millimeter wave interconnects and components based on simulations would require accurate phase and amplitude assessments to aid in their design.

3. **Characterization**: Characterization of millimeter wave antenna can be challenging owing to their extremely small size and unique implementation methods.

4. **Fabrication**: The fabrication challenges mainly stem from the miniature size of the antenna and millimeter wave components. They require the use of high precision fabrication technologies like laser lithography for planar substrate antennas, precision cutting, drilling, milling tools for waveguide components. The use of SMA connectors at this frequency also requires precise positioning and isolated environments.

5. **Measurement**: The measurement of millimeter wave antennas requires precise alignment and positioning of the AUT and the illuminating antenna. In case of On-chip antenna it may be necessary to use a metal probing station to excite the antenna in the laboratory. Devices such as metal probing station being made of metal introduce multipaths and would therefore introduce possible errors in measurements if the MPC are not isolated.
3.2.1. Prototypes of mmWave beamforming antennas

Two prototypes are designed and presented here, namely the 28 GHz endfire antenna element and the 28 GHz and 31 GHz waveguide based dual band lens antenna element. The next sections explain the design and simulation results in detail.

3.2.2. Initial Development Cycle Stages

The prototype for the WP2 identified as ‘Prototype 3’ namely the high directional mmWave antenna element was designed for operation at the 28 GHz using substrate based planar end fire antenna design initially. Since the band of 28 GHz at millimeter wave has been researched well under different propagation scenarios [30], it would be more attractive to design the antenna at 28 GHz. During the later phase, it was decided that, the substrate based antenna element when used at high frequency of 28 GHz introduces high losses and therefore a decrease in gain, it was not very promising to continue the same design for 28 GHz. With all these due considerations, it was decided to take up the design of antenna using a different lens based waveguide technology than the one based on planar substrate.

3.2.3. Prototype 1: 28 GHz endfire antenna element

The design of the 28 GHz antenna prototype is shown in Fig. 29. The antenna element is of an endfire configuration with printed dipole acting as driver whose gain is enhanced by the printed reflector and parasitic director elements. The antenna operates from 27.5 GHz to 34 GHz and has a bandwidth of 21.1%. The antenna is shown in the Fig. 29.

The antenna return loss behaviour is shown in the Fig. 30 and is well below -10 dB for the entire band of operation.
The antenna element in simulation has a gain of 8.1 dBi and the gain pattern characteristic is shown in Fig. 31.

### Figure 30 Return loss characteristic of the antenna

### Figure 31 Gain radiation pattern of the antenna

#### 3.2.4. Prototype2: 28 GHz waveguide based lens antenna element

Millimeter wave band requires the use of high gain multibeam antennas to cast multiple simultaneous beams in desired directions. In course of attaining this objective the bandwidth over which this performance is obtained is a crucial factor. Beamformer networks such as the Butler, Nolan and Blass matrices have proposed schemes to project multiple beams. These and the variants of these beamforming networks often operate over a narrow frequency bandwidth. They operate by aperture sharing and the derived beamsteering capabilities. With the current trends there is a need for antennas with multiple beams, wide angular coverage and operating over a broader bandwidth especially at the 28 GHz band. Antennas based on quasi-optical beamformers [31] using parallel plate waveguide and lens based technology aided by a flared aperture horn for radiation are an attractive option in this regard.
The antenna introduced here consists of multiple horn fed into a parallel plate waveguide connected to a lens that converts cylindrical wave front to planar wave front, before finally radiating out through a flared aperture horn. The design of a dual band multibeam antenna with operation centered around 28 GHz and 31 GHz is discussed next. Fig. 32 shows the design of the antenna in a perspective view. The 3D antenna occupies a volume of 190 mm x 130 mm x 35 mm. All simulations were carried out in full wave solver of Ansoft HFSS.

3.2.4.1. Principle of operation

The antenna design consists of 3 sectoral H plane horns feeding into a common parallel plate waveguide with TE mode excitation for each. The rare horns are placed at 60° inclination w.r.t to the central one to focus beams in the directions of 0°, +60° and -60°. The antenna uses a ridge based custom designed lens as the transformer of the wavefronts. The cylindrical wavefront emitted by the sectoral H plane horn is corrected for the phases using the ridge delay line placed at the corresponding distance to account for the extra path length needed by the cylindrical wavefront to convert to planar wavefront. This results in the amplification or focusing of the gain due to the individual horn. The ridge waveguide transitions into a flared horn structure to further add to the gain.

3.2.4.2. Frequency and directional behaviour of the antenna

The S parameter behaviour of the three ports of the antenna are shown in Fig. 33. The antenna is designed for a dual band behaviour over 28 GHz and 31 GHz band with 1 GHz of bandwidth at -15 dB of S11 for each band. The port to port isolation between the three ports at both the frequencies of interest is below -25 dB as shown in Fig. 34. The two frequencies are currently used in the LMDS (Local Multi-Service Digital System).
multipoint distribution system) and the antenna element can find potential use at this band. The directional beams cast at 0°, +60° and -60° are to form a directional coverage area spanning 120° sector of a typical base station cell. The design could further be extended to other angles as well with marginal variation in the gain.

![Figure 33 S parameter behaviour at the three ports of the antenna element](image)

![Figure 34 Port to port isolation between the three ports of the antenna element](image)

### 3.2.4.3. Gain of the dual band multibeam antenna

The multi beam antenna at 28 GHz has a gain of 15 dBi at the central beam and a gain of 19 dBi at the other two symmetric directions. The horn antenna alone has a gain of 11 dBi. So, the gain amplification through the lens is almost double (19 dBi at the side angles) than that provided by a horn antenna.
The same gain behaviour is seen in the other band of 31GHz with a slight variation in the gain of the order of 1 dBi in the central beam and 0.1 dBi in the 60° beams. The gain plots vs azimuthal scan is shown for the two frequencies in the Fig. 35 and Fig. 36.

![Figure 35 Multibeam gain pattern vs azimuth at 28 GHz](image)

![Figure 36 Multibeam gain pattern vs azimuth at 31GHz](image)

### 3.2.5. Conclusion and future work

The presented antenna element offers promise with high directional beams with high gain and narrow beamwidth for millimeter wave applications at two frequencies. The antenna element could find use in small cell or macro cell environments where directional high gain beams are a necessity.

Currently the antenna design is being improved upon. Suitable adaptations will be made in the antenna design to accommodate the fabrication requirements. In the future, the fabrication and assembly of the antenna element would be taken up followed by the measurement and characterization of the antenna patterns in the anechoic chamber.
3.3. Energy efficient architecture

The following work is partially presented in [51] and [52].

3.3.1. Energy efficient massive MIMO transmission

In the following, we propose a novel RSM architecture for large MIMO systems aimed to reduce the power consumption at the user terminal, while attaining a significant throughput. The energy consumption reduction is obtained by exploiting non-linear detector (amplitude detector), which reduces the number of RF chains and ADCs. Next, we propose a RSM transmission scheme where the BS transmits spatial and modulation symbols per channel use. We show that the optimal spatial symbol detector is a threshold detector that can be implemented by using one-bit ADC. We derive closed form expressions for the detection threshold at different signal-to-noise-ratio (SNR) regions showing that a simple threshold can be obtained at high SNR and its performance approaches the exact threshold. Then, we propose a time-division-duplex (TDD) transmission protocol aimed to reduce the training overhead where the channel knowledge is required only at the BS. After that, power consumption analysis is performed to the proposed, hybrid and fully digital MIMO systems. Finally, we derive expressions for the average bit error probability (ABEP) in the presence and absence of the threshold estimation error showing that a small number of pilot symbols is needed. Simulation results show that the power consumption and the energy efficiency of the proposed RSM architecture outperform the hybrid and conventional MIMO systems. Performance comparison based on ABEP is performed between the proposed system and conventional MIMO showing that a suitable constellation selection can reduce the performance gap. We present the conclusion and the future research challenges.

We propose a RSM MIMO system where the BS is fully digital and the UT structure is illustrated at Fig. 37. The proposed receiver relies on energy-efficient devices that can be represented as follows:

- **RF chain and high precision ADC**
  These devices are the most power consuming so we use two RF chains and two high precision ADCs for any number of receiver antennas.

- **Amplitude detector (AD)**
  It is a cheap analog device that detects the amplitude of the RF signal in absolute value and operates at high frequencies with high sensitivity and negligible power consumption [53]. TABLE I in [54] shows that the AD can operate at frequency range from 1GHz to 85GHz, with sensitivity from -5dBm to -50dBm.

- **1-bit ADC**
  It consumes very low power as the power consumption of the ADC grows exponentially with number of quantization bits.

- **Phase shifters (PSs)**
  We use number of PSs equals to $N_r$ for the uplink transmission only while the hybrid system in Fig. 37 uses phase shifters much more than $N_r$. 
3.3.2. System model, precoder design and challenges to be addressed

We consider the downlink of a large MIMO single user that operates in the mmWave outdoor narrowband channel environment. The BS and UT are equipped with $N_f$ and $N_r$ antennas respectively. Based on the properties of mmWave propagation, we consider reciprocal propagation environment. We exploit the channel reciprocity by considering TDD system where the channel state information (CSI) is needed only at the BS. In Fig. 37, displayed in red, we consider a low complexity uplink UT circuit based on analog PSs. During the uplink training, the UT sends pilot symbols so that the BS can acquire the CSI. This can be achieved, since the optimal training pilot symbols matrix (PSM) for the least squares channel estimation can be selected as a discrete fourier-transform (DFT) basis [33] that can be implemented by PSs. Since massive mmWave MIMO systems suffer from high path loss and antenna correlation, antenna selection at the receiver is necessary. The BS selects the most suitable $N_a$ antennas based on the channel knowledge (assumed perfect). After that, the BS informs the UT by the $N_a$ active receive antennas (ARA). The BS transmits data vector to the ARA that comprises spatial and modulation symbols. The transmitted data vector can be written as

$$x_i^j = \sqrt{\alpha p} B s_i x^j$$  \(3\)

where the spatial symbol $s_i \in \mathcal{R}^{N_a \times 1}$ contains $N_a$ bits from the input data bits, $i \in \{1,...,2^{N_a}-1\}$, we assume that the all-zeros spatial symbol is not allowed, the modulation symbol $x_j$ is a symbol from certain constellation with size $M$, $j \in \{1,...,M\}$, the number of transmitted bits per data vector are $(N_a + \log_2 M)$, $B \in \mathcal{C}^{N_r \times N_a}$ is the precoding matrix, $P$ is the average transmit power and we adjust the transmitted power by a normalization factor $\alpha = (\text{Tr}(B^H B))^{-1}$ where $\text{Tr}(\cdot)$ is the trace operator. The received signal vector is given by

$$y = \sqrt{\alpha p} H B s_i x^j + n$$  \(4\)

where $H \in \mathcal{C}^{N_r \times N_t}$ is the channel matrix and $n \in \mathcal{C}^{N_r \times 1}$ is the generated noise vector where its coefficients are independent and identically distributed (i.i.d) zero mean circularly symmetric complex Gaussian random variables and each has variance $\sigma^2$. Let us define the matrix $H_a$ as the channel matrix.
from the BS to the selected ARAs. In order to direct the data vector to the ARA, we design the precoding matrix \( B \) as a zero forcing precoder that can be expressed as

\[
B = H_d^H (H_d H_d^H)^{-1}
\]

The received signal by the \( k^{\text{th}} \) active antenna can be expressed as

\[
y_k = \sqrt{\alpha p} \ s_j \ x_j + n_k
\]

### 3.3.3. Channel model

Millimeter wave channels have limited scattering clusters due to the high path loss. Moreover, using large arrays increases the antennas correlation. Therefore, we choose the narrowband clustered channel model that is widely used for outdoor mmWave channels [44]. In this model, the channel matrix can be expressed as

\[
H = \sqrt{\frac{N_c N_r}{N_c N_y}} \sum_{i=1}^{N_c} \sum_{j=1}^{N_y} g_{ij} \ L_T (\theta_{ij}^r) \ L_T (\phi_{ij}^t) V_r (\theta_{ij}^r) V_t (\phi_{ij}^t) H
\]

where \( N_c \) represents number of scattering clusters, \( N_r \) is the number of rays per cluster, \( g_{ij} \) is the complex gain, \( \theta_{ij}^r, \phi_{ij}^t \) are the elevation and azimuth angles of arrivals and departures, \( L_T (\theta_{ij}^r), L_T (\phi_{ij}^t) \) are receive and transmit directional antennas gains, \( V_r (\theta_{ij}^r), V_t (\phi_{ij}^t) H \) are receive and transmit array response vectors. We consider uniform linear arrays where the \( N \) antennas normalized response vector can be expressed as

\[
V (\Phi) = \frac{1}{\sqrt{N}} \begin{bmatrix} 1, e^{jk d \sin(\Phi)}, ..., e^{j(N-1) k d \sin(\Phi)} \end{bmatrix}^T
\]

where \( k = \frac{2\pi}{\lambda} \) and \( d \) is the inter-antenna spacing. The directional antennas gain can be expressed as

\[
 L (\Phi) = \begin{cases} 1, & \Phi \in [\Phi_{\text{min}}, \Phi_{\text{max}}] \\ 0, & \text{else} \end{cases}
\]

where \( \Phi_{\text{min}} \) and \( \Phi_{\text{max}} \) determine the transmission sector angle.

### 3.3.4. Receive antenna selection

Since massive mmWave MIMO systems suffer from high path loss and antenna correlation, antenna selection at the receiver is necessary. For a given \( N_a \), the BS selects the best \( N_a \) ARA such that the received power is maximized that is, \( \alpha \) is maximized, \( \alpha^{-1} = \text{Tr}(H_H^H) - 1 \). The maximization of \( \alpha \) can be done through the exhaustive search. However, the exhaustive search increases the computational complexity especially with large number of ARAs. A low complexity, fast and efficient receive antennas selection algorithms are considered as a future work topic.

### 3.3.5. Spatial and modulation symbols detection

From eq. (6) each ARA receives either (\( \sqrt{\alpha p} \ x + n \)) or (\( 0 + n \)) according to if the transmitted spatial bit is one or zero respectively. In order to recover the spatial symbol, each ARA is connected to AD and
one-bit ADC as illustrated by example in Fig. 38. The AD measures the amplitude of the received signal and it is compared to a predefined threshold at the one-bit ADC. Thus, the output signals from the ADCs represent the spatial symbol.

In the spatial symbol detection, we will show that both per antenna detection and joint detection lead to the same results.

**Separate spatial symbol detection**

The downlink circuit in Fig. 38 shows that each receive antenna is connected to RF AD. The signal provided by the AD at the kth active antenna is

$$a_k = \sqrt{(\sqrt{ap} s_{ik} x_{jl} + n_{kl})^2 + (\sqrt{ap} s_{ik} x_{jQ} + n_{kQ})^2}$$

(10)

where the indices l and Q represent the in-phase and quadrature components. The probability density function (PDF) of the received amplitude at the kth active antenna follows either a Rice or Rayleigh distributions [55]. The PDF of the received amplitude can be expressed as

$$f(a_k) = \begin{cases} 
\frac{2a_k}{\sigma^2} e^{-\frac{a_k^2 + ap}{\sigma^2}} I_0 \left(\frac{2a_k\sqrt{ap}}{\sigma^2}\right), & s_{ik} = 1 \\
\frac{2a_k}{\sigma^2} e^{-\frac{a_k^2}{\sigma^2}}, & s_{ik} = 0 
\end{cases}$$

(11)

We consider maximum likelihood (ML) detector per ARA to decide if the received spatial bit is one or zero. The detection problem per the kth antenna can be formulated as

$$f(a_k \mid s_{ik} = 1) \frac{s_{ik}^2}{s_{ik}^2} f(a_k \mid s_{ik} = 0)$$

(12)
According to the problem in (12), the estimated spatial bit for the $k^{th}$ antenna can be expressed as

$$s_{ik} = \begin{cases} 1, & a_k > \gamma \\ 0, & a_k < \gamma \end{cases}$$

where $\gamma$ is a threshold that results from solution of the problem in (14). In the following, we present three ways to determine $\gamma$ based on the received SNR.

**Exact threshold:** We can obtain the exact value of the threshold directly by solving (14) numerically. However, this solution will increase the UT circuit complexity.

**Moderate SNR approximation (MSA):** At moderate SNR, we can approximate $I_0(x) \approx \frac{e^x}{\sqrt{2\pi x}}$ in problem (13) and calculate the threshold by solving the following equation

$$\frac{\sigma^2}{4\pi\sqrt{\alpha p}} e^{4\gamma\sqrt{\alpha p} - 2\alpha p} = 1$$

putting it in the form $xe^x = c$ as

$$\frac{-4\gamma\sqrt{\alpha p}}{\sigma^2} e^{-4\gamma\sqrt{\alpha p}} = -\frac{1}{\pi} e^{2\alpha p}$$

The solution of the equation in (15) can be given by

$$\gamma = \frac{-\sigma^2}{4\sqrt{\alpha p}} W_{-1}\left(-\frac{1}{\pi} e^{\frac{2\alpha p}{\sigma^2}}\right)$$

where $W_{-1}(x)$ is one of the main branches of the Lambert W function [56]. Nevertheless, we must calculate the threshold in (16) numerically and this leads to increase in the UT circuit complexity.

**High SNR approximation (HSA):** At high SNR, the left-hand side of equation (14) takes either infinity or zero values based on the sign of the term in the exponential power. Therefore, we can obtain a simple threshold that can be expressed as

$$\gamma = \frac{\sqrt{\alpha p}}{2}$$
Joint spatial symbol detection

In order to jointly detect the spatial symbol bits, we apply the ML detector based on the received amplitudes from all of the active antennas. Since the received amplitudes are i.i.d, their joint PDF can be expressed as

\[ f(a) = \prod_{k=1}^{N_{a}} f(a_k) \]  
(20)

The joint ML detection problem can be formulated as

\[ \hat{s}_i = \arg \max \{ f(a | s_i) \} \]  
(21)

For the sake of simplicity, let us consider the case when the number of the ARA is two. In this case, the joint ML detection problem can be expressed as

\[ \hat{s}_i = \arg \max \{ f_{01}, f_{10}, f_{11} \} \]  
(22)

where \( f_{nm} = f_a(a | s_{i1} = n, s_{i2} = m) \).

As an illustrative example, we decide symbol \([0 1]^T\) if the following conditions are satisfied

\[ f_{01} = f_{00}f_2 > f_{10} = f_{00}f_1 \]  
(23)

\[ f_{01} = f_{00}f_2 > f_{11} = f_{00}f_1f_2 \]  
(24)

where \( f_k \) can be given by

\[ f_k = e^{-\frac{\alpha_p}{\sigma^2} I_0 \left( \frac{2\alpha_p \sqrt{\alpha_p}}{\sigma^2} \right)} \]  
(25)

Inequalities in (23) and (24) imply that \( a_1 < \gamma \) and \( a_2 > \gamma \) respectively. See equations (14) and (15) for more illustration. The analysis of the joint detection can be extended for any number of ARA. Therefore, the joint detection results are same as per antenna detection.

Modulation symbol detection

At first, we detect the spatial symbol and then we detect the modulation symbol. According to the example illustrated in Fig. 39, all the ARA associated to a spatial bit \( s_{ik} = 1 \) receive the same modulation symbol. The \( k^{th} \) ARA is connected to RF switch that passes the signal only if the estimated spatial bit \( \hat{s}_{ik} = 1 \). All the signals that pass through the switches are combined. After that, the combined signal passes through RF chain and a single ADC to detect the modulation symbol.
The combined signal can be given as

$$y_c = \sum_{k=1}^{N_a} s_{ik} \hat{s}_{ik} \sqrt{\alpha_p} x_j + \hat{s}_{ik} n_k$$  \hspace{1cm} (26)$$

This combined signal passes through the RF chain and the high precision ADCs to be detected.

### 3.3.6. Communication protocol

In this section, we present a DL TDD transmission protocol for the proposed system. The proposed protocol can work efficiently for low and average mobility systems where the channel coherence time can be larger than the TDD frame length [57].

<table>
<thead>
<tr>
<th>Table 2: Proposed TDD frame</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Uplink Training</strong></td>
</tr>
</tbody>
</table>

**Uplink training**

As illustrated in Fig. 40, the UT transmits pilot symbols to allow the BS to acquire the CSI. We propose a simple UL circuit based phase shifters that can achieve the optimal training PSM.
In channel estimation, the optimal scaled least squares PSM can be implemented by phase shifters as it can be designed on DFT basis as described in eq. (15) of section IV in [33]. In this method, $N_r$ pilot symbols are needed to estimate the channel [33]. The channel estimation error $\epsilon$ was given by eq. (22) in [2] and can be expressed as

$$\epsilon = \frac{\sigma^2 N_r^2 N_t \text{Tr}[E[H H^H]]}{\sigma^2 N_r^2 N_t + P_t \text{Tr}[E[H H^H]]}$$

where $P_t$ is the transmit power. When $N_t$ or $N_r$ increases as in the massive MIMO case, the estimation error in (27) can be asymptotically approximated as eq. (23) in [33]

$$\epsilon_{\infty} = \text{Tr}[E[H H^H]]$$

where the error depends on the strength of the channel. By considering the clustered mmWave channel model in (7) with path-loss $P_l$, the asymptotic channel estimation error in (28) becomes $\frac{N_t N_r}{P_l}$.

**Downlink training**

The BS sends pilot symbols to allow the UT estimates the detection threshold. We consider that the entries of the transmitted spatial symbol are all ones.

In order to estimate the detection threshold, the UT estimates the average received signal amplitude and the noise level from the outputs of the ADs connected to the ARA. The joint PDF of the received amplitudes can be expressed as

$$f(a \mid 1_0) = \prod_{k=1}^{N} \frac{a_{k1}}{\sigma^2} e^{-\frac{a_{k1}^2 + \rho^2}{\sigma^2}} I_0 \left( \frac{2a_{k1}\rho}{\sigma^2} \right)$$

Figure 40: Uplink Training
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where $1_a$ is all-ones spatial symbol, $\rho = \sqrt{ap_k}$, $a_{k1}$ measured amplitude at the $k^{th}$ active antenna according to $s_{ik} = 1$, $N = N_p N_a$ and $N_p$ is the number of pilot symbols. We design the amplitude estimator $\rho$ to maximize

$$\log f(a | 1_a) = \sum_{k=1}^{N} \log \left( \frac{2a_{k1}}{\sigma^2} \right) - \frac{a_{k1}^2 + \rho^2}{\sigma^2} + \log I_0 \left( \frac{2a_{k1}\rho}{\sigma^2} \right)$$

(30)

By using the fact that $I_0(x) = \frac{e^x}{\sqrt{2\pi x}}$ for large $x$, we can simplify equation (28) as

$$\log I_0 \left( \frac{2a_{k1}\rho}{\sigma^2} \right) = \frac{2a_{k1}\rho}{\sigma^2} - \frac{1}{2} \log 4\pi a_{k1}\rho$$

(31)

In order to find the ML amplitude estimator $\hat{\rho}_{ML}$ we solve the problem, $\frac{\delta}{\delta \rho} \log f(a | 1_a) = 0$, that can be expressed as

$$\sum_{k=1}^{N} -2\rho \frac{a_{k1}}{\sigma^2} + 2 \frac{a_{k1}}{\sigma^2} = \frac{1}{2}$$

(32)

From (30), the $\hat{\rho}_{ML}$ can be expressed as

$$\hat{\rho}_{ML} = \frac{\Sigma_{k=1}^{N} a_{k1}}{2N} + \frac{1}{2} \sqrt{\left( \frac{\Sigma_{k=1}^{N} a_{k1}}{N} \right)^2 - \sigma^2}$$

(33)

The ML estimator of the noise variance $\sigma_{ML}^2$ can be obtained by solving, $\frac{\delta}{\delta \sigma^2} \log f(a | 1_a) = 0$ as

$$\sigma_{ML}^2 = \frac{2}{N} \sum_{k=1}^{N} (a_{k1} - \rho)^2$$

(34)

By solving the equations in (31,32) simultaneously, a closed form expression for $\hat{\rho}_{ML}$ can be given as

$$\hat{\rho}_{ML} = \frac{2 \Sigma_{k=1}^{N} a_{k1}}{3N} + \frac{1}{3} \sqrt{4 \left( \frac{\Sigma_{k=1}^{N} a_{k1}}{N} \right)^2 - \frac{4N}{3} a_{k1}^2}$$

(35)

3.3.7. Energy consumption analysis

According to [38], the power consumption of the different receiver components can be expressed in.
Table 3: Power consumption of the different receiver components

<table>
<thead>
<tr>
<th>Component</th>
<th>Power Consumption</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{LNA}$</td>
<td>$P_{ref}$</td>
</tr>
<tr>
<td>$P_{ADC}$</td>
<td>10 $P_{ref}$</td>
</tr>
<tr>
<td>$P_{PS}$</td>
<td>1.5 $P_{ref}$</td>
</tr>
<tr>
<td>$P_{SW}$</td>
<td>0.25 $P_{ref}$</td>
</tr>
<tr>
<td>$P_{RF \text{ chain}}$</td>
<td>2 $P_{ref}$</td>
</tr>
<tr>
<td>$P_{BB}$</td>
<td>10 $P_{ref}$</td>
</tr>
</tbody>
</table>

where $P_{PS}$, $P_{SW}$ and $P_{BB}$ are the phase shifter, switch and baseband power consumption respectively.

The DL receiver power consumption of the proposed system $P_P$ (Fig. 37), conventional MIMO $P_C$ (Fig. 24) and hybrid MIMO $P_H$ (Fig. 27) can be expressed as

$$P_P = N_r (P_{LNA} + P_{SW}) + (P_{RF \text{ chain}} + P_{ADC}) + P_{BB}$$  \hspace{1cm} (36)$$

$$P_C = N_r (P_{LNA} + P_{RF \text{ chain}} + P_{ADC}) + P_{BB}$$  \hspace{1cm} (37)$$

$$P_H = N_r (N_{rf} + 1)P_{LNA} + N_r N_{rf} P_{PS} + 2N_{rf} (P_{RF \text{ chain}} + P_{ADC}) + P_{BB}$$  \hspace{1cm} (38)$$

where $N_{rf}$ is the number of RF chains.

We define the energy efficiency $EE$ as the transmission bit rate for a given bit error rate (BER) per the receiver power consumption that can be expressed as

$$EE = \frac{BPCU \times BW}{P_r}$$  \hspace{1cm} (39)$$

where $BW$ is the bandwidth and $P_r$ is the receiver power consumption.

### 3.3.8. Performance analysis

We evaluate the proposed system performance based on the ABEP that can be expressed as

$$\text{ABEP} = \frac{N_a P_{es} + \log_2 (M) P_{em}}{N_a + \log_2 (M)}$$  \hspace{1cm} (40)$$

where $P_{es}$ and $P_{em}$ are the spatial and modulation bit error probabilities respectively that can be given by

$$P_{es} = 0.5(P_1 + P_0)$$  \hspace{1cm} (41)$$

where $P_1 = \Pr(a_{k1} < \gamma)$ and $P_0 = \Pr(a_{k0} > \gamma)$.

$$P_{em} = \sum_{i=1}^{2^{N_a-1}} \sum_{n=1}^{2^{N_a}} \text{BEP}(x_j \in \mathbb{C}_M \mid \mathbb{s}_n, \mathbb{s}_i) \Pr(\mathbb{s}_n \mid \mathbb{s}_i) \Pr(\mathbb{s}_i)$$  \hspace{1cm} (42)$$

where $\text{BEP}(x_j \in \mathbb{C}_M \mid \mathbb{s}_n, \mathbb{s}_i)$ is the bit error probability of M-quadrature-amplitude-modulation (M-QAM), M-phase-shift keying (M-PSK) or M-amplitude-phase-shift-keying (MAPSK) constellations [58]. The probabilities in equation (36) can be expressed as

$$\text{BEP}(x_j \in \mathbb{C}_M \mid \mathbb{s}_n, \mathbb{s}_i) = \text{BEP}(x_j \in \mathbb{C}_M \mid \text{SNR}_c)$$  \hspace{1cm} (43)$$
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$$\text{SNR}_c = \frac{(b_{m_1}^{(m_1 m_2)})^2 \alpha P}{b_{m_1}^{(m_1 m_2)} + b_{m_1}^{(0)}} \sigma^2$$  \hspace{1cm} (44)

$$\Pr (s_n | s_i) = P_1 b_{m_1}^{(m_1)} (1 - P_1) b_{m_1}^{(0)} P_0 b_{m_1}^{(0)} (1 - P_0) b_{m_1}^{(0)}$$  \hspace{1cm} (45)

$$\Pr (s_i) = \frac{1}{2^{N_a - 1}}$$  \hspace{1cm} (46)

where $b_{m_1}^{(m_1 m_2)}$ is the number of entries in $s_i$ that each one of them equals $m_1$ and its corresponding entry in $s_n$ equals $m_2$ and $\sum_{m_1,m_2 \in [0,1]} b_{m_1}^{(m_1 m_2)} = N_a$.

**Perfect threshold**

If the UT knows the threshold perfectly, the probabilities $P_1$ and $P_0$ can be expressed by the cumulative density function of Rice and Rayleigh distributions [55] as

$$P_1 = 1 - Q_1 \left( \frac{\sqrt{2} \alpha P}{\sigma}, \frac{\sqrt{2} \gamma}{\sigma} \right) \hspace{1cm} P_0 = e^{-\gamma^2}$$  \hspace{1cm} (47)

where $Q_1(x)$ is the first order Marcum Q-function.

**Estimated HSA threshold**

The ML estimator is asymptotically gaussian and we consider that the gaussianity is prevailed. The estimated HSA threshold $\hat{\gamma} \sim \mathcal{N}(\mu_\gamma, \sigma_\gamma^2)$ are the mean and variance of $\gamma$. Since the received amplitudes are positive, we can express $P_1$ and $P_0$ as

$$P_1 = \Pr \left( \frac{\hat{\gamma}}{a_{k1}} > 1 \right) \hspace{1cm} P_0 = \Pr \left( \frac{\hat{\gamma}}{a_{k0}} < 1 \right)$$  \hspace{1cm} (48)

In order to find the probabilities in equation (4), we put it in the following form

$$P_1 = \Pr \left( t_{\delta,n,1} > \frac{\sigma}{\sigma_{\gamma}} \right) \hspace{1cm} P_0 = \Pr \left( t_{\delta,n} < \frac{\sigma}{\sigma_{\gamma}} \right)$$  \hspace{1cm} (49)

where $t_{\delta,n}$ is Non-central t distribution, $t_{\delta,n,1}$ is Doubly-noncentral t distribution [59], $n$ is the degrees of freedom, $\delta, l$ are the non-centrality parameters, $\delta = \frac{\mu_\gamma}{\sigma_{\gamma}}$, $n=2$ and $l = \frac{2\alpha P}{\sigma^2}$.

Closed form expressions for $P_1$ and $P_0$ with threshold estimation error can be given as

$$P_1 = 1 - T_{\delta,n,1} \left( \frac{\sigma}{\sigma_{\gamma}} \right) \hspace{1cm} P_0 = T_{\delta,n} \left( \frac{\sigma}{\sigma_{\gamma}} \right)$$  \hspace{1cm} (50)
where $T_{\delta,n}$ and $T_{\delta,n,1}$ are the cumulative density functions of Non-central and Doubly-non-central t distributions \cite{59}. In order to determine the probabilities in equation (48), we need the mean and the variance of $\hat{\gamma}$. The estimated HSA threshold $\hat{\gamma}$ can be expressed as

$$\hat{\gamma} = \frac{1}{2} \hat{\rho}_{ML}$$  \hfill (51)

From the asymptotic properties of the ML estimator \cite{26}, the mean and variance of $\hat{\gamma}$ can be expressed as

$$\mu_{\hat{\gamma}} = \frac{1}{2} \rho \quad \sigma^2_{\hat{\gamma}} = \frac{1}{4} [I_{\hat{\theta}}^{-1}]_{11}$$  \hfill (52)

where $I_{\theta}$ is a $2 \times 2$ fisher information matrix \cite{60} that its elements can be expressed as

$$[I_{\theta}]_{11} = -E\left[\frac{\partial^2 f(\mathbf{a} | 1_\alpha)}{\partial \rho^2}\right] = \frac{2N}{\sigma^2} - \frac{N}{2\rho^2}$$  \hfill (53)

$$[I_{\theta}]_{12} = [I_{\theta}]_{21} = -E\left[\frac{\partial^2 f(\mathbf{a} | 1_\alpha)}{\partial \rho \partial \sigma^2}\right] = \frac{2N}{\sigma^4} (\mu_1 - \rho)$$  \hfill (54)

$$[I_{\theta}]_{22} = -E\left[\frac{\partial^2 f(\mathbf{a} | 1_\alpha)}{\partial \sigma^2 \partial \sigma^2}\right] = \frac{2N}{\sigma^6} (\mu_2 + \rho^2 - 2\rho \mu_1) - \frac{N}{2\sigma^4}$$  \hfill (55)

where $\mu_1 = E[a_{k1}]$ and $\mu_2 = E[a_{k1}^2]$. The inverse of the fisher information matrix can be expressed as

$$I_{\hat{\theta}}^{-1} = \frac{1}{[I_{\theta}]_{11}[I_{\theta}]_{22} - [I_{\theta}]_{12}^2} \begin{bmatrix} [I_{\theta}]_{22} & -[I_{\theta}]_{12} \\ -[I_{\theta}]_{21} & [I_{\theta}]_{11} \end{bmatrix}$$  \hfill (56)

From equations (50, 54), the variance $\sigma^2_{\hat{\gamma}}$ can be given as

$$\sigma^2_{\hat{\gamma}} = \frac{1}{4} \frac{[I_{\theta}]_{12}^2}{[I_{\theta}]_{11}[I_{\theta}]_{22} - [I_{\theta}]_{12}^2}$$  \hfill (57)

**BER**

In simulation environment, we consider that $g_{ll}$ are i.i.d. $CN(0, \sigma^2_0)$ where $\sigma^2_0$ is designed such that $E[Tr\{H^*H\}] = N_c N_r, \quad N_c = 8, \quad N_r = 20$, the elevation and azimuth angles $(\theta_{l1}, \phi_{l1})$ have Laplacian distributions with uniform random means $(\theta_{l1}, \phi_{l1})$, angular spreads $\sigma_{\theta} = \sigma_{\phi} = 1$, SNR = $\frac{\rho}{\sigma^2_0}$, the width of the transmission angle is 50 ° and the user has omnidirectional antenna array. We compare the performance of the proposed system with singular value decomposition (SVD) based precoding and decoding of the FD MIMO system. We allocate the power at SVD such that all the activated modes achieve the same received SNR \cite{61}.
Fig. 41 shows the ABEP of the proposed system with exact threshold compared to FD MIMO at 32×8 MIMO system. The performance of the proposed system without receive antenna selection is inferior to that with antenna selection. Moreover, the proposed system performance with constant amplitude constellation and 4 spatial bits approaches the FD MIMO.

Fig. 42 represents the ABEP of the proposed system at different thresholds and different numbers of receive antennas. Applying HSA threshold leads to the lowest complexity and the performance gap is less than 1 dB with respect to the exact threshold. The ABEP with threshold estimation error is very close to that with perfect threshold by using only one downlink pilot symbol. Increasing the number of receive antennas while $N_d$ is fixed boosts the receive antennas gain and as a result the ABEP is improved.
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Throughput

We consider $P_{ref}= 20\text{mW}$, we consider 28GHz carrier frequency with 1GHz bandwidth (BW), noise power = $-84\text{dBm}$, the distance between BS and UT is 50 meters, path-loss = 120dB, $P_t = 40\text{dBm}$, $BER = 10^{-5}$, $N_t = 32$. We determine the number of BPCU by applying the following algorithms

- **Conventional MIMO:**
  We use SVD precoding and decoding and fix the same constellation for all active modes. We allocate the power such that each active mode achieves symbol error probability $10^{-5}$. We repeat this procedure for different constellations and take the maximum number of transmitted bits.

- **Hybrid MIMO:**
  The SVD decoder is designed using hybrid precoding [12] and the same procedure as in conventional MIMO is performed.

- **Proposed system:**
  We fix the constellation symbol and select the spatial symbol to get $BER = 10^{-5}$, repeat for different constellations and take the maximum number of transmitted bits.

![Figure 42: ABEP of the proposed system at $N_t=32$, $N_a=4$, 16-PSK, perfect and estimated thresholds](image)

Security: Public
In Fig. 43, the bit rate is the number of transmitted BPCU multiplied by the BW at BER = 10^{-5}. The bit rate increases with $N_r$ so the conventional MIMO system is superior to the other architectures. The gap between the proposed RSM system and the conventional MIMO arose because of using only one RF chain.

**Energy consumption**

In Fig. 44, the proposed RSM receiver consumes the lowest power because it uses only one RF chain and one ADC. In hybrid MIMO, the number of phase shifters increase rapidly with $N_{rf}$ so it consumes higher power than conventional MIMO when $N_{rf}$ increases.

In Fig. 45, the bit rate of the proposed RSM increases with $N_r$ with small additional receiver power consumption so its EE is the best while the hybrid system is the worst.
Sensitivity to CSI errors
Evaluating the performance of the proposed system taking into consideration the CSI errors at the BS is a topic for future research.

3.3.9. Conclusion and future research
We have considered the downlink of a massive MIMO single user operating in the mmWave outdoor narrowband channel environment. We developed a novel RSM architecture aimed to reduce the power consumption at the UT and achieve high throughput. We proposed zero forcing precoding
scheme and derived the required decoder, providing expressions for the ABEP. We presented a transmission protocol aimed to reduce the training overhead where the channel is required only at the BS. We studied the power consumption and the EE of different MIMO transceivers showing that the proposed system outperforms hybrid and conventional MIMO systems. A performance comparison is performed between the proposed system and FD MIMO in terms of ABEP showing that an appropriate constellation selection can reduce the performance gap.

**Future research**

In this subsection, we present the expected future research topics

- **Receive antenna selection:**
  Fast and efficient receive antenna selection algorithms to maximize the received signal power with performance approaches the exhaustive search method but with much lower computational complexity.

- **Higher order spatial modulation:**
  Extend the proposed architecture for higher order spatial modulation schemes by using multiple-bit ADCs at the receiver side and consider the channel estimation errors.

- **Multi-user:**
  Extend the proposed architecture for the multi-user case.

- **Wideband-channels:**
  Developing RSM precoding scheme for wideband channels.

- **Secondment:**
  Evaluate the performance of the proposed algorithms and study the effect of the blockage and investigate blockage mitigation techniques at SIRADEL company during the secondment interval.

- **Channel estimation:**
  Developing low complexity channel estimation algorithms for RSM MIMO systems at mmWave frequencies.
4. Assessment of mmWave dense small-cell networks

The channel modelling approaches presented in section 2.3 are utilized in this section to evaluate the channel performance in the presence of various in-street obstructions like large vehicles, user-body self-obstruction and human crowd obstruction at mmWave frequencies. A small-cell network is first designed in an urban European city environment in the SIRADEL ray-tracing tools. The blockage impact is then evaluated using system level analysis based on Monte-Carlo simulations using beamforming antennas at the transmitter. Studies are performed for various scenarios like single-cell, two-cell (street level) and network wide impact of the blockages.

4.1.1. Setup of scenario and simulation parameters

The proposed small-cell network is shown in Fig. 46. It has been designed manually, after different inter-site distances have been tested, and provides close to full coverage on the target service area. It is composed of 18 small-cells over a 0.090 km² wide area. The average inter-site distance (ISD) is 78 m, while the maximum ISD is 110 m. Following subsections analyze in more details the network and user performance depending on the beam-switching antenna, data traffic and local obstruction.

The objective of this simulation-based study consists in deploying small-cell antennas so that the main streets and squares in a 300 m × 300 m wide area are covered with broadband mobile access. The small-cells are assumed to be installed on lampposts, i.e. on one side of the street, at height 7 m above the ground. The location of each small-cell must be chosen so it is in line-of-sight from at least one neighbour small-cell, and preferably two. In this way, we make possible the deployment of a wireless mesh network that would likely operate in the mmWave spectrum as well.

Each small-cell has only one sector. It is equipped with an automatically steerable antenna. The half-power beamwidth (HPBW) is 22° with maximum gain 18.5 dBi. The radiation pattern is assumed to be perfectly shaped, i.e. without any significant side-lobe, and following the model given in [4]. The transmit power is adjusted such the EIRP is 40 dBm as allowed in current FCC rules.
Figure 37 SIRADEL S simulator with small-cell deployment (red dots) and user locations (sidewalks: blue lines) in a real European city environment.

Target users are assumed to be all outdoors. It may be objected that some signal penetrates through the windows of buildings in the small-cell close vicinity, however the resulted indoor coverage is very limited and poor; the number of indoor users served in that way must be very small. Relays can obviously be used to combat the strong penetration loss; an antenna installed on the building exterior façade is connected to the outdoor small-cell, while another antenna feeds a local indoor wireless access network. Simulating such a system requires the definition of a more complex scenario and was out of scope for this study. Thus the considered users are all distributed in the streets and open squares. The locations we decided to serve are represented by the blue breaklines in Fig. 47. Remark that courtyards or small streets are not part of the coverage objective, as those areas are difficult to serve and probably with lower traffic demand.

Two active user densities are computed and compared, i.e. 200 users/km² and 1000 users/km², in order to assess the evolution of the inter-cell interference and user performance. Other scenario parameters are listed in Table I.

The simulator is similar to the one run in [6] for analysis of beamforming and beam-switching antennas in a 2-GHz small-cell network. Users are randomly dropped in the prediction area at successive Monte-Carlo iterations. At each iteration, the simulator determines the attachment and selects the antenna beam orientation that maximizes the downlink (DL) signal-to-noise ratio (SNR) of each user. A procedure similar to the beamforming training specified in IEEE 802.11ad standard [7] permits to find the suitable beam orientation with azimuth resolution 11°, i.e. half the antenna HPBW.
The simulator calculates the DL interference levels depending on the antenna beam orientation and bandwidth occupancy by other-cell users. For both the calculation of the useful and interfering powers, the isotropic multi-path field strengths are combined with the antenna gains given in the propagation directions.

Beamforming is realized in the horizontal plane only, meaning that the azimuth of the beam is adjusted for each served user, while the elevation of the maximum gain is always in 0°. A spectrum resource can only be used to serve a single user, i.e. multi-user beamforming is not supported. And there is no inter-cell coordination. Non full-buffer (or finite) traffic is considered. The demand of each active user is limited to 15 Mbps.

| System          | - PHY waveform: Single-Carrier (SC)  
|                 | - Central frequency: 60 GHz         
|                 | - Bandwidth: 200 MHz                
|                 | - Mapping table: from IEEE 802.11ad, with 12 adaptive modulation schemes (MCS)  
|                 | - PHY net data rate: 45.6 to 525.0 Mbps |
| Users           | - Noise figure: 7 dB                
|                 | - Antenna: omni-directional, 5 dBi  
|                 | - Demand: 15 Mbps                   
|                 | - Active user density: 1000 users/km² |
| Power budget    | - Impairment loss: 8 dB             
|                 | - DL Rx sensitivity: -79 to -84 dBm  |

| Human crowd     | - λ(radius=0.2m) = 5              
|                 | - λ(radius=0.4m) = 4              
|                 | - λ(radius=0.8m) = 2              
|                 | - Max range from user antenna = 2m (very crowded), 3m (moderately crowded), 4m (crowded) |

| Other           | - Monte-Carlo: 30 iterations       
|                 | - Traffic: Non full-buffer         |

The performance statistics are obtained from a Monte-Carlo process, where users are dropped randomly along the sidewalks with finite throughput demand. Multi-path propagation, cell selection, beam selection and interference calculations are executed at each iteration, as explained with more details in [7]. The IEEE 802.11ad link performance is considered [10].

An orientation is randomly set to each user, following a given probability density function p(θ) where θ is the angular distance between the sidewalk direction and the body-to-equipment direction. For simplicity, a uniform distribution has been considered in the present study. The user-body obstruction affects the multi-path propagation data based on the model described in section 1.

Large vehicles are randomly dropped in the streets, as shown in Fig. 47 (a) by purple rectangles. The number of vehicles has been arbitrarily fixed: 170/km². Their length is 12 m. They completely block any incident propagation path, but they also produce new reflections and diffractions. Metallic material is considered.
**Inter-cell interference**

Fig. 48 (a) and Fig. 48 (b) show respectively the cumulative distribution function (CDF) of the inter-cell interference levels and signal-to-interference-plus-noise-ratio (SINR) for two different user densities. Except for a very small occurrence percentage, the interference level with 200 users/km² is below and even far below the receiver sensitivity (-79 dBm), meaning that the network performance is almost not affected by the interference. When going up to 1000 users/km², the percentage of allocated resources by each small-cell strongly increase (the average cell load goes from 4.1% to 22.6%, and the load in the most crowded cell goes from 7.6% to 46.4%). Thus the interference levels grow significantly; the percentage of them above the receiver sensitivity goes up to 20%. The peak interference value, measured at the 90-percentile, increases from -85 dBm to -76 dBm. Consequently, the SINR perceived by the users is degraded. Only in a quite limited way however: the mean SINR decreases by 2.6 dB, which is reasonable knowing that the amount of allocated resources in the network has been multiplied by a factor of 5.5. Actually the strong propagation loss and shadowing, as well as the beamsteering-based spatial filtering, permit to efficiently isolate users located in two different cells.
The same analysis was conducted with another antenna, having a smaller beamwidth (15°), a steering azimuth resolution of 6°, and a maximum gain of 21.9 dBi. The 40 dBm EIRP is maintained (maximum radiated power currently allowed by FCC for 60-GHz for point-to-point links), meaning that the transmit power has been reduced from 21.5 dBm to 18.1 dBm. As EIRP is kept constant, a few dB decrease of both the received and interfering signal strengths is observed, which comes from a reduction in the multi-path combination gain; the decrease on interfering levels is significantly greater than the one on the useful signal in average. All this was expected. An improvement in SINR levels was expected as well; the inverse trend is observed. The reason is that most users suffer from the signal strength degradation, while only a small percentage of links were interference-limited, thus benefit from the interference decrease. The conclusion is that: if the antenna gain difference is compensated by a transmit power adjustment, and the global interference strength suffered by the network is low, then a smaller beamwidth does not necessarily lead to better performance.

**User performance**

The user quality of service has been measured from three metrics: user outage, mean DL spectral efficiency (SE), and the mean DL spectral efficiency at cell-edge, which is the average spectral efficiency computed from the 10% worst-served user. Fig. 49 gives the evolution of those metrics depending on the user density. Outage is twice when increasing the user traffic. The mean SE and mean cell-edge SE are respectively degraded by 3% and 13%.

![User performance vs User density](image)
4.1.2. Impact of bus obstruction

Because of the small-cell height and transmission frequency, the coverage may be strongly affected by in-street obstructions, like static objects (e.g. urban furniture), human bodies or high vehicles. Actually the obstruction of a dominant path can be partly mitigated by the system by pointing the antenna towards another path. This mechanism can be properly predicted if multi-path channel data is available. As an example, in Fig. 19, the bus is obstructing the dominant direct-path, but significant signal strength still reaches the user location from a reflection on the building façade.

The effect of bus obstruction on a fixed user located at the blue star shown in Fig. 47 (a) is simulated in this section. The street containing the two adjacent small-cells from the user is the only street considered in this simulation. The active user densities are set to the same values as in previous sections and the users are randomly located based on the Monte-Carlo simulator. The horizontal distance between the fixed user location and the small-cell located to the west, shown in Fig. 47 (a), is 37 m and the small-cell to the east of the fixed user is 33 m. Both the small-cells are located on the opposite side of the street from the user. This represents a scenario in which a bus obstruction impacts a user located at a fixed position for e.g. user at a bus stop.

The mean SINR and the mean inter-cell interference simulated for this fixed user without the bus obstruction with 200 users/km² are 21 dB and -143 dBm respectively, while the connection is provided by the east small-cell. Various simulated bus positions are described by their relative positions to the fixed user as shown in Fig. 50. The front of the bus (of length 12 m) is used as the reference to calculate this relative distance. When the bus is located on the west side of the fixed user, the relative distance is negative. When the bus is positioned at a relative distance of -6 m, it obstructs the dominant path from the east small-cell to the fixed user; the mean signal level decreases by 20 dB; the fixed user is served by the west small-cell. The same change in cell assignment occurs for some other neighbouring users; however all users in the vicinity that are not obstructed by the bus are still connected to the east small-cell, which is closer. This situation creates a higher level of interference as seen in Fig. 50, with a mean interference approaching the receiver sensitivity. When the relative distance is equal or greater than 8 m, the dominant path from the east small-cell to the fixed user is free. At a distance of +14 m, the neighbouring users are no more obstructed and then remain attached to the east small-cell as well, the inter-cell interference reaches the same value as the case without the bus obstruction (-143 dBm and -83 dBm for respectively 200 and 1000 users/km²).

Finally, all simulated users (incl. the fixed user) are served throughout the various considered bus positions and there is no outage. However the SINR and interference evolutions shown in Fig. 50 are specific to the predicted scenario, and thus cannot be generalized, they properly stress how in-street obstruction can dramatically change the radio conditions, cell assignment and the network performance. The simulation also indicates that the chosen cell density permits the network to offer seamless connection, although the link quality can locally be severely degraded.
4.1.3. Impact of user-body self-obstruction

Single cell simulation
The impact of user-body obstruction is first evaluated in a single-cell scenario. One of the small-cell shown in Fig. 47 (a) has been chosen and users were dropped on the sidewalks 40 m around. Fig. 51 shows the cumulative distribution function (CDF) of the signal-to-noise-ratio (SNR) with and without the impact of user-body obstruction. The users suffer a significant degradation; the average SNR reduces by 1.7 dB and the 10% percentile decreases by 3.5 dB. However the impact is very small for the highest percentiles; the SNR reduces by only 0.3 dB at the 90% percentile.

Two-cells simulation
The previous scenario is extended to the case of a street containing two small-cells separated by 80m. The purpose is to assess the impact of user-body obstruction on inter-cell interference and user performance. Twenty users in average are dropped at each Monte-Carlo iteration into the portion of the street located between the small-cells. Note there is no outage in absence of user-body obstruction.
Fig. 52 and Fig. 53 show the CDF of respectively the inter-cell interference levels and the signal-to-interference-plus-noise-ratio (SINR), with and without user-body consideration. The user-body obstruction leads to lower interference levels. The median interference reduces significantly by 6.7 dB while the most critical interference levels (at highest percentiles) exhibits lower variations; i.e. 0.6 dB at the 90% percentile and lower than 0.1 dB at the 95% percentile.

The SINR CDF exhibits degradation caused by the signal reduction such as observed in the previous single cell scenario; the SINR reduces by 2.9 dB and 0.5 dB for respectively the 10% and 50% percentiles. But there is insignificant degradation for the 90% percentile. The impact of the body obstruction has been partly mitigated via two mechanisms: 1) a strong SNR reduction leads the user to switch to the other cell; 2) when interference is greater or similar to the noise level (which is generally the case here), the reduction of the interference levels may compensate part of the signal decrease. This is particularly true for users for which the body severely obstructs the closest small-cell; those users move to the other cell providing lower signal strength; but then the interference that is coming from the closest cell is negligible. Finally, the simulation results indicate that a major effect of the user-body obstruction is higher inter-cell isolation.

![Figure 43 Two-cell scenario: CDF of Inter-cell Interference levels.](image-url)
The basic studies presented here were chosen to properly validate, illustrate and make understandable the user performance evolutions. The following study deals with a more complex scenario.

**Small-cell network performance**

The effect of the user-body blockage and interactions with large vehicles is now assessed through the whole network of small-cells scenario. The density of active user s is 1000 users/km2, meaning that 90 users in average are dropped at each iteration.

The network performance is computed in three different scenarios: 1) no stochastic component; 2) consideration of the user-body obstruction; 3) consideration of both user-body and vehicles together. The results for respectively the Inter-cell interference and the SINR are given in Fig. 54 and Fig. 55.

![Figure 44 Two-cell scenario: CDF of SINR.](image)

![Figure 45 Whole network scenario: CDF of Inter-cell Interference level.](image)
The reduction of the inter-cell interference due to the user-body blockage is confirmed in Fig. 54, with an average decrease of 1.9 dB. As observed in the two-cell scenario, this positive effect, as well as the optimal cell selection process we consider in the simulation, makes the SINR degradation quite limited. As illustrated in Fig. 55, the 10% SINR percentile decreases by 1.1 dB while the median SINR decreases by 0.2 dB due to body obstruction. Adding the vehicle obstruction has an insignificant impact on link quality. Actually, the impact of vehicle obstruction is local only and has been compensated by two mechanisms (the same as for the body obstruction but at a higher degree): 1) The obstruction of the dominant path can be partly mitigated by the system by pointing the antenna towards another path; 2) The reduction of the SINR leads the user to switch to the other cell and benefit from significant interference reductions due to vehicle obstruction. This latter mechanism is emphasized in the streets where small-cells are distributed alternatively along each side of the street. Remark that as observed in the single bus obstruction scenario investigated in [7], there are some specific cases where interferences may dramatically increase but their occurrence appears to be small. The SINR improvement at highest percentiles that was observed in the two-cell scenario is no more visible when assessing the performance of a larger network.

4.1.4. Impact of human crowd obstruction

The effect of the user-body obstruction is evaluated through system level simulations considering both the user body self-obstruction and the impact due to the different sizes of crowds. The network performance is assessed in five different situations: 1) no body obstruction (not realistic); 2) self-body obstruction; 3-5) self-body and crowd obstruction with different crowd densities.

In the first step, the deterministic predictions of the channel are obtained using a ray-based approach with multipath propagation. Due to the street-canyon like nature of the scenario a rich multipath environment is created. These predictions are then fed into a stochastic system level simulator which uses a Monte Carlo scheme. Users are randomly dropped along the sidewalks (marked in blue lines in Fig. 47 (a) in the selected area. In Fig. 56, the positions for the different crowd densities in a single iteration are shown. In the case of the very crowded scenario, large sized clusters in a dense...
environment are located very close to the mobile station. In case of the moderately crowded scenario, the clusters are located further away and are composed mostly of medium sized clusters. Finally, in case of the crowded scenario, small to medium sized clusters are mostly considered located at a distance which is even further from the mobile station. The user body is randomly oriented at each iteration, to include the self-obstructing impact.

The channel performance is evaluated by considering the three different crowd densities along with the impact in the absence of any crowd but only with the user body self-obstruction and finally without consideration of any human body blockage (purely deterministic). The Cumulative Distributive Function (cdf) results for the SINR and the Inter-cell Interference (ICI) levels is shown in Fig. 57 and 58.
The inter-cell interference (ICI) reduces as the number and size of the obstacles is increased in the scenario. This can be seen in the Fig. 57, in which the probability of interference levels is the least for the case without the human-body obstructions. At the median value, the interference level without the human body obstruction is -86.4 dBm whereas for the very crowded scenario it is -93.5 dBm. This is due to the fact that the clusters, especially large ones in the case of a very crowded scenario are located very close to the mobile station which causes large shadow region that not only block the signal but also the interference. This can be seen as a positive impact due to the blockages. At lower percentiles, the very crowded scenario and the moderately crowded scenario have similar interference levels whereas the other obstruction scenarios tend to the no human body obstruction levels.
For the three different crowd scenarios, it can be seen from Fig. 58 that the SINR values reduce as the density of the crowd increases. The impact of only the user-body self-obstruction as compared to the one without any human blockage is quite limited which is a result of the lower interference levels (positive impact) and the process of optimal cell selection, used in the simulations. The 10th percentile value of the SINR without human body obstruction is 7.9 dB whereas it is -2.6 dB at the densely crowded scenario. Therefore, even with the optimal cell selection process and the reduction of the interference there still exists a reduction in SINR by 10.5 dB.

For a service that requires SINR > 0 dB, the difference in percentiles between the extremes of not considering any human bodies (deterministic) vs. the densely crowded obstruction is 7.8% (from 4.7% to 12.5%). This means that just due to the addition of the human body obstruction (self and dense crowd), an extra 7.8% users may be denied service. This can be seen as a significant impact of blockage due to the human body obstruction, which must be considered in the channel modelling for mmWave dense small-cell networks.

In conclusion, the topology of future outdoor mmWave small-cell networks must adapt to various constraints like the propagation loss, presence of opaque or quasi-opaque static obstacles in the streets (vegetation, urban furniture), and high sensibility to in-street moving obstacles (vehicles, bodies). Accurate geographical map data and an enhanced ray-based model are coupled here with a Monte-Carlo system-level simulator. A network with 18 small-cells over 0.09 km² wide area gives almost full coverage in main streets of a suburban environment. The impact of the steerable antenna beamwidth and user density on the inter-cell interference and user performance are assessed as well as the degradation suffered by a bus obstruction. Good robustness of the network versus interference and in-street obstruction was observed, however further evaluations are required before complete conclusions can be drawn.

Hybriding ray-based models with stochastic components is at an early stage. The propagation framework should later be extended with interactions on other bodies and in-street objects. The user-body is shown to sometimes bring positive effect, with stronger inter-cell isolation. Besides, the degradation on the useful signal strength is not affecting too much the global network performance, as soon as the victim users can switch to another cell with no body masking. The obstruction by the user-body and vehicles finally leads to a quite limited reduction of the SINR. Significant degradation occurs only on the very worse-served users: 1.1 dB on the 10% percentile; and 0.2 dB on the median percentile. Analysis will continue in future studies in order to better identify, understand and model the main non-static factors that should be considered in mmWave network design. Based on the results, it can clearly been seen that the impact of large vehicles in the street can be reduced by physically placing the small-cell base stations on opposite sides of the streets in an asymmetrical manner.

The human body obstruction as compared to no human body obstruction causes significant degradation of 2.8 dB and 1.5 dB in case of densely and moderately crowded scenarios respectively. This impact can be reduced by physically placing the antennas at higher and asymmetric positions. Also, the optimal cell selection process helps to reduce the blockage impact. Hybrid ray-based models with stochastic components are at an early stage and measurements are required for further improvement and calibration of these models. The proposed model should later be extended to scenarios with mobility in which the users are moving in between the small-cells. Adding dynamic dimensions (i.e. time variant channel and network scheduling) remains a major perspective.
5. Channel data repository

A channel data repository was created for the 5Gwireless project partners based on the requirements given in Table 5. The scenario used for the simulated channel data is from a small-cell network of 5 small-cells at 28 GHz. The total area considered is 0.015 km², with an average ISD (Inter-site distance) of 78 m, obtained from a larger network designed previously in section 2.2. All the same cells are considered to be located on top of lampposts at a height of 8 m above the ground. The small-cells consist of 32 MIMO isotropic antennas, randomly oriented in the direction of the streets (towards the small cells). All the users are located outdoors on the deployment lines which represent the footpath as shown in Fig. 59. At the user end, the number of antennas is 16 and a user density of 30 users for the given area is used. The results are presented as matrices corresponding to the individual user antennas (16). Each matrix consists of 40x32 complex coefficients of the channel data, where there are 40 subcarriers for each of the 32 BS antennas. The same set of data is made available for all 30 users in the small-cell network.

![Figure 50 Five small-cells scenario in urban European environment](image)

**Table 5. 5Gwireless partner requirements for channel data**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>28 GHz</td>
</tr>
<tr>
<td>Scenario</td>
<td>Outdoor, urban</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>500 MHz</td>
</tr>
<tr>
<td>User equipment</td>
<td>16 antennas, uniform linear array</td>
</tr>
<tr>
<td>Base station</td>
<td>32 antennas</td>
</tr>
</tbody>
</table>

Security: Public
6. Conclusions and future work

The new algorithms, channel models, antennas and experimental trials have been reported and discussed in this deliverable. With the introduction of mmWave systems, high data rates are expected due to the large availability of the spectrum and bandwidth. The large bandwidth requires a new perspective towards the channel modelling, firstly related to non-stationarity. Indeed non-stationarity could lead to inaccurate or distorted estimations of the mmWave channels. New cluster evolution rules are proposed, which can be added to the existing channel models. Another important feature of mmWave channels, based on measurements observations, is that the directional information can be represented by Markov states. By applying this approach the complexity of the estimation of the channel can be reduced significantly.

The mmWave channels are highly sensitive to the details in the interacting environment due to the small wavelengths. We suggest the channel modelling can be performed through a deterministic approach (precise terrain representation and ray-tracing) to obtain highly accurate estimations for different scenarios. However, the difficulty is more on the modelling of moving or small objects, which have properties that are more stochastic in nature. Therefore hybrid models are required that can utilize the advantages of both deterministic and stochastic approaches. Such a hybrid ray-based mmWave channel model is proposed in this deliverable considering various in-street obstacles in outdoor channels. The obstacles include fixed objects like buildings and vegetation which can be deterministically represented and considered in simulation. Whereas, the non-static objects like human bodies and large vehicles are estimated using stochastic or hybrid approaches. The human body obstruction as compared to no human body obstruction causes significant degradation of 2.8 dB and 1.5 dB in case of densely and moderately crowded scenarios respectively. This impact can be reduced by physically placing the antennas at higher and asymmetric positions along the streets. Also, the optimal cell selection process helps to reduce the blockage impact. Hybrid ray-based models with stochastic components are at an early stage and measurements are required for further improvement and calibration of these models. The proposed model should later be extended to scenarios with mobility in which the users are moving in between the small-cells. Adding dynamic dimensions (i.e. time variant channel and network scheduling) remains a major perspective.

The high data rates expected at mmWaves would also lead to higher power consumption from the perspective of individual RF chains connected to each antenna element in the traditional setup. To reduce this power consumption, a novel energy efficient RSM architecture is proposed for mmWave outdoor channels. By utilizing a non-linear detector (amplitude detector), the number of RF chains has been reduced which reduces the power consumption in the system. The proposed architecture outperforms energy efficiency in both hybrid and conventional MIMO systems which is evaluated through simulations. Further, a suitable constellation selection is made to reduce the difference in performance between the proposed system and conventional MIMO.

New mmWave antennas were designed and validated by simulations considering the antenna pattern stability over a wideband. Simulations of the performance of the designed antenna require significantly larger duration at mmWaves as the antenna sizes are smaller and there are a large number of mesh cells in the solver. Antenna pattern stability, stable gains and phase centre localization over the large bandwidths are some other concerns related to the design of the antennas. From a fabrication perspective, high precision fabrication technologies like laser lithography are
required. Due to the small size of the antennas precise alignment and positioning during the fabrication and measurement phases is also required. With these considerations, a prototype has been designed and is under preparation for fabrication at 28 GHz in this deliverable.

Analysis and evaluation will continue. From a hardware perspective, new techniques to accommodate fabrication challenges for mmWave frequencies are required. After the fabrication and assembly of the antennas, the measurements and characterization of the antennas will be performed in an anechoic chamber. Multi-user channels, antennas and architectures for the mmWave systems will also be considered. In a multi-user channel, multiple users can communicate simultaneously through the same channel. This can be achieved among other approaches by considering beamforming at the antenna design stage such that an antenna can support multiple beams simultaneously to serve multiple users. Energy efficient architecture proposed in this deliverable will also be extended to a multi-user case. This approach can increase the capacity of the system however further studies are required on the entire system including the effects of interference. Antenna selection schemes, higher order spatial modulation, wideband channels and channel estimation are also other important perspectives to be included in the architecture of the mmWave systems.
References

D2.2: Innovative mmWave transmission schemes, channel models and antennas


