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Millimetre-Wave Based Mobile Radio Access Network for Fifth Generation Integrated Communications (mmMAGIC)

Deliverable D5.2

Final multi-node and multi-antenna transmitter and receiver architectures and schemes
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Abstract

This final deliverable from the WP5 of mmMAGIC project details the multi-antenna, multi-node and the transceiver hardware modelling related analyses and solutions developed over the 2 year project duration. The multi-antenna studies have identified hybrid beamforming as a key solution to meet the complexity and performance needs of mm-wave transceivers. The multi-node support, be it through other access points, rays or other technologies like FSO (Free Space Optics) is identified as an essential element in mm-wave communications, to increase the link reliability. The antenna design and modelling to meet the unique challenges and opportunities in Access point (both radio access and backhaul) and handset domains have been achieved. The hardware impairments (including Phase Noise and Power Amplifier non-linearities) have been studied and effective modelling solutions presented in this work.

Keywords

Antenna array, Analog Beamforming, Digital Beamforming, Hybrid Beamforming, Impairments, mm-wave, Multi-node co-ordination, Non-linearities, Power consumption, Radiation patterns

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Executive summary

This deliverable discusses the mature multi-node/ multi-antenna transceiver architectural and schematic solutions and hardware component models developed by mmMAGIC WP5.

At the start of the project and in D5.1, WP5 considered all transceiver architectural options for beamforming, i.e. Digital, Analog and Hybrid. As the research matured, WP5 focussed more on Hybrid beamforming due to the best trade-off in complexity, cost and performance it offers. This is reflected in this deliverable, where most of the architectural and schematic solutions are based on Hybrid beamforming. We show, through detailed simulations, that Hybrid beamforming is more resilient to challenging radio channel conditions in mm-wave and also to hardware impairments, than Digital or Analog beamforming. From an architectural point of view, we also propose a sub-array based solution, which can reduce the combiner losses and further simplify the Hybrid beamforming operations. Given the technical evidence we show on the flexibility and resilience of Hybrid beamforming, we would recommend this option for most of the mm-wave mobile applications, particularly in radio access.

We have analysed multi-antenna schematic solutions for access, backhaul and relay applications in this deliverable. For access, we have looked at beam synthesis solutions and synthesis of wider beams for broadcasting signals, in addition to the above noted analyses on Hybrid beamforming performance. For backhaul, we introduced in D5.1 a novel, simplified decoding scheme for massive, line of sight MIMO communication channels, which was pitched as a short range backhaul/ fronthaul solution. In this D5.2, we report an extensive validation study based on realistic ray traced channel data (indoor and outdoor) with real antenna patterns, to estimate what proportion of the theoretical performance can be achieved in real environments. The included experiment details and the resulting analyses are a highlight from the year 2 work of WP5. We also provide an analysis into the challenges of providing mm-wave backhaul to a moving hotspot in an urban environment. In terms of relay solutions, we propose a user selection scheme for ad-hoc relay link provision when the direct link (user to the access point) is blocked.

Our analyses into multi-node schemes have shown that the provision of multi-node support is an essential feature for reliable mm-wave communications. The ‘second link’ can be of mm-wave, lower frequency RAT or even of another technology like Free Space Optics (FSO). Having such a second link significantly increases the coverage reliability, ensuring that cell sizes that are economically viable can be achieved. We propose a hybrid mm-wave/FSO multi-hop solution, which can significantly improve the performance where the use of HARQ can mitigate the effects of an imperfect mm-wave power amplifier. We also demonstrate that with the use of a few high rise mm-wave access points to complement the low rise AP network, the LOS probability can be significantly enhanced for greater cell ranges. Further, we investigate the efficacy of relay nodes in improving the Outdoor to Indoor (O2I) coverage in mm-wave networks. It is shown that the contribution of the relay nodes increases with longer cell ranges and we also demonstrate an optimum bandwidth split point (between the relay link and backhaul link) which maximizes the overall spectral efficiency.

The modelling of the performance and impairments of the key hardware components of the transceiver chain is a major contribution from the WP5 work. We have looked at different antenna designs and their performance for handset and access point (for both radio access and backhaul/ fronthaul) applications. We provide optimized antenna array designs in D5.2, to meet the specific challenges and also utilize opportunities in these applications. The impacts of Phase Noise (PN) are captured with a very detailed model and a simpler model and we provide PN impact analysis with both these models. These analyses look at the PN effects in MIMO OFDM systems and also at the possible densities of Phase Tracking Reference Signals (PTRS) in the radio frame, which is currently an active topic in 3GPP NR standardisation. The power amplifier performance in mm-wave frequencies is modelled with both statistical and behavioural models and we show that there is good agreement in both these approaches.
Overall, D5.2 provides a comprehensive picture of the innovative work done in WP5 over the past 2 years, in multi-antenna transceiver design, multi-node co-ordination and hardware performance assessment and impairment modelling.
Contents
1 Introduction ........................................................................................................................................ 1
2 Multi-Antenna Transceiver Schemes and Design ........................................................................ 4
  2.1 Introduction .................................................................................................................................... 4
  2.2 Hybrid Beamforming Transceiver Architecture ............................................................................ 4
  2.3 Hybrid Beamforming with a Subarray Architecture ....................................................................... 5
  2.4 Transceiver Schemes for Access .................................................................................................... 7
    2.4.1 Flexible Multi-User Hybrid Beamforming Design ................................................................. 7
    2.4.2 System Level Comparison of Different Transceiver Architectures for Access ....................... 11
    2.4.3 Digital Beamforming with 1 bit DACs .................................................................................... 14
  2.5 Transceiver Schemes for Backhaul ............................................................................................... 16
    2.5.1 Massive Multiple Input Massive Multiple Output (MMIMMO) for Short Range LOS Links .... 16
    2.5.2 Impact of Beam Misalignment for Moving Hotspot Scenario ................................................. 23
  2.6 Transceiver Design for Relaying ................................................................................................... 27
    2.6.1 Relay Selection in mm-wave Multiuser Systems ...................................................................... 27
  2.7 Mm-wave CSI Aspects: Beam Design ............................................................................................ 31
    2.7.1 Arbitrary Beam Synthesis for Hybrid Beamforming Systems ................................................. 31
    2.7.2 Wide Beams and Broadcasting Signalling ............................................................................... 35
  2.8 Summary ....................................................................................................................................... 36
3 Coordinated Multi-node Scheme Design ......................................................................................... 38
  3.1 Introduction .................................................................................................................................... 38
  3.2 Beam sweeping for multi-node networks ...................................................................................... 40
  3.3 Sequential Hybrid Beamforming Design for Multi-Link mm-wave Communication ................. 43
  3.4 Beam management for mobility ..................................................................................................... 46
    3.4.1 Beam management procedures ............................................................................................... 46
  3.5 Performance of mm-wave based RF-FSO Multi-hop Networks .................................................... 48
    3.5.1 On the Performance of mm-wave based RF-FSO Multi-hop Networks ................................. 48
    3.5.2 System model ........................................................................................................................... 48
    3.5.3 Data Transmission Model ......................................................................................................... 49
    3.5.4 Analytical results ...................................................................................................................... 49
    3.5.5 Simulation results ...................................................................................................................... 50
  3.6 Mm-wave LOS Coverage Enhancements with Coordinated High-Rise APs ............................... 51
    3.6.1 LOS probability ......................................................................................................................... 51
    3.6.2 Evaluation Results .................................................................................................................... 53
  3.7 Direct vs Relay-assisted Access: A System Level Evaluation at mm-waves ......................... 54
  3.8 Joint Hybrid Precoding for Energy-efficient mm-wave Networks ............................................. 57
3.8.1 Introduction ................................................................. 57
3.8.2 System Model .............................................................. 57
3.8.3 Joint Hybrid Precoding .................................................... 58
3.8.4 Conclusion ................................................................. 60
3.9 Summary and key results and observations for System Concept Design .... 60
4 Hardware impairments modelling and performance assessment ............... 62
  4.1 Introduction ................................................................. 62
  4.2 Mm-wave antennas design and models .................................. 62
     4.2.1 Antenna specifications ............................................. 62
     4.2.2 Patch antenna arrays for mm-wave user terminal and access point .... 64
     4.2.3 Massive dipole arrays ............................................. 67
     4.2.4 Transmitarray antennas for access point .......................... 72
     4.2.5 Transmitarray antennas for backhauling/fronthauling ............... 75
  4.3 Power amplifier modelling .................................................. 77
     4.3.1 Behavioural modelling ............................................. 77
     4.3.2 Statistical modelling ............................................... 77
     4.3.3 Scaling laws for energy efficiency analysis ......................... 78
  4.4 A simple and effective Phase Noise model and design of PTRS ............... 80
     4.4.1 The multi-pole/zero PN model ..................................... 80
     4.4.2 Analysis of Phase Tracking Reference Signal (PTRS) design .......... 81
  4.5 Effect of Phase Noise on Uplink Multi-User MIMO-OFDM .................... 82
  4.6 Impact of hardware impairments on system performance achieved by hybrid beamforming for mm-wave access .................................................. 84
     4.6.1 Phase shifter errors .................................................. 84
     4.6.2 Combiner stage losses .............................................. 85
  4.7 Impact of Major RF Impairments on mm-wave Communications using OFDM Waveforms .................................................. 86
  4.8 Summary ................................................................. 89
  5 Conclusions ..................................................................... 90
List of Figures

Figure 2-1: Hybrid Beamforming with Analog RF beamforming ................................................................. 5
Figure 2-2: Fully Connected and Partially Connected Architecture (Subarrays) ................................................. 6
Figure 2-3: Performance Simulation Comparison with Subarrays (NRF = 4) ...................................................... 7
Figure 2-4: Conceptual level descriptions of the proposed hybrid beamforming design ................................. 8
Figure 2-5: BER simulation comparisons for the single user case ................................................................. 10
Figure 2-6: BER comparisons for the single user case after blockage of main path ...................................... 10
Figure 2-7: Performance comparisons for the multi-user case ................................................................. 11
Figure 2-8: Performance comparisons for the multi-user case after blockage of the strongest phat of UE#1 ................................................................. 11
Figure 2-9: CDF of the UE throughput under perfect CSI assumption .......................................................... 13
Figure 2-10: CDF of the UE throughput for different values of $T$ ................................................................. 13
Figure 2-11: Quantized massive MU-MIMO-OFDM downlink system .......................................................... 14
Figure 2-12: Uncoded BER with QPSK .............................................................................................................. 15
Figure 2-13: Achievable sum-rate throughput with Gaussian inputs and mismatch nearest-neighbour decoding ................................................................................................................................. 15
Figure 2-14: MMIMO system parameters (where $\rho_c$ is a normalising factor and $\dagger$ is the transpose conjugate operation, for more details please refer to [MMMAGICD51]) ......................................... 17
Figure 2-15: Bristol Simulated Deployment Scenarios ......................................................................................... 17
Figure 2-16: Helsinki Airport Simulated Deployment Scenarios ......................................................................... 18
Figure 2-17: Antenna models ............................................................................................................................ 18
Figure 2-18: Beam misalignment due to $\Delta x$ .................................................................................................. 24
Figure 2-19: The beam misalignment probability ($P_{\text{mis}}$) vs. the speed variance ............................................ 26
Figure 2-20: The beam misalignment probability ($P_{\text{mis}}$) vs. fixed mmSC beam width ...................................... 26
Figure 2-21: 95% available rate ......................................................................................................................... 27
Figure 2-22: An example of the relay selection process: (a) initial transmission scheduling, (b) blockage detection, (c) broadcasting help message, (d) relay candidate identification (e) relay selection metric and (f) transmission via the relay path ........................................ 28
Figure 2-23: CDF of spectral efficiency loss experienced at (a) the blocked user and (b) the unblocked users ............................................................................................................................................ 30
Figure 2-24: Sub-array HBF with individual power constraint ................................................................................. 34
Figure 2-25: Fully-connected HBF with 2-bit phase quantization and sum power constraint ............................. 35
Figure 2-26: Sub-array (a) and fully-connected (b) HBF MU-MIMO 2 simultaneous beams ................................ 35
Figure 2-27: Example of beam shapes designed using the Widener method described in [Int16] and the technique using amplitude tapering in [QQL16]. Plotted is also the DFT beam and the subelement pattern ........................................................................................................ 36
Figure 3-1: Multi-Node Beam Sweeping (figure modified from original figure in [HKC16]) .............................. 39
Figure 3-2: Example of Beam Sweeping Result in a Multi-Node Setup ............................................................. 40
Figure 3-3: Average Sum Rate for Different Methods to Determine the Best beam pair ................................. 42
Figure 3-4: Parallel beam search frame structure for test beam pair f_k11f_k12, w_k2. .............43

Figure 3-5: Performance comparisons of digital beamforming (DB), reference HBF in [ARA+14], the proposed HBF using perfect CSI of the effective channel (PHBF), PHBF #1, PHBF #2 and PHBF #3 in [ZCS+17]. (a) One-node scenario and (b) 2-node scenario. SNR refers to received SNR at each UE antenna input from BS #1. ........................................45

Figure 3-6: Performance comparisons of the proposed HBF using PHBF #3 in [ZCS+17] with perfect CSI knowledge (PHBF) and different pilot transmission powers (β = 1,2,4). SNR=10 dB. (a) One-node scenario and (b) 2-node scenario. Two-node scenario. SNR refers to received SNR at each UE antenna input from BS #1. ........................................45

Figure 3-7: CDF of SINR on the baseline data transmission when parallel beam training is carried out with different pilot transmission powers. β = 1,2,4. SNR=10 dB. (a) One-node scenario and (b) 2-node scenario. SNR refers to received SNR at each UE antenna input from BS #1. ........................................45

Figure 3-8: Illustration of beam management procedures. (Top) P1 to enable UE measurement on different TRP Tx beams to support selection of TRP Tx beams/UE Rx beam(s) (Middle) P2 to enable UE measurements on different TRP Tx beams to possibly change/select inter/intra-TRP Tx beam(s), (Bottom) P3, to enable UE measurement on the same TRP Tx beam to change UE Rx beam in the case UE uses beamforming .................46

Figure 3-9: Beam based mobility measurements using Level 1 SS. The SS’s may be wide/sector covering (left) or somewhat beamformed (right, coarse beams) depending on the user scenario. The network can use the SS report as a starting point for the Level 2 beam management using P1 or go directly to Level 3 beam management using P2/P3 in the case SS reports already gives some beam direction information (right figure). ......................47

Figure 3-10: In Level 2 based beam reporting and when the load is low (left) the SS reports from the Level 1 reporting can be used to select a restricted beam sweep for the UE which may be periodic. TRPs with no users (as observed from SS measurement reports) need not transmit CSI-RS at all. If the load is high (right), the TRP nodes makes a periodic beam sweep for CSI-RS reporting, when Level 1 doesn’t use beamforming or uses sector covering wide beams (Figure 3-9, left). .............................................................47

Figure 3-11: Level 3 beam management with aperiodically triggered UE specific beamformed CSI-RS (P-2) and UE RX beam refinement (P-3). ........................................................................................................47

Figure 3-12: (a): On the tightness of the analytical results. (b): Outage probability for different PA models and number of HARQ-based retransmissions. ......................................................51

Figure 3-13: Blockage by high building........................................................................52

Figure 3-14: LOS association probability for H_b = 3 and 30 m, respectively (H_{max} = 15 m)......53

Figure 3-15: Illustration of the considered two considered access methods. (a) Direct access vs (b) Relay-assisted access ..........................................................................................54

Figure 3-16: System performance achieved by DA and RA: macro sector spectral efficiency 55

Figure 3-17: System performance achieved by DA and RA: 5th percentile of the UE rate ......55

Figure 3-18: Impact of the bandwidth allocated to the BS-RS link with RA on the macro sector spectral efficiency.............................................................................................................56

Figure 3-20: Average total transmit power v Δv_k = 1K ||Rvbk, v||2 vs. target spectral efficiency per user for a given user drop over 500 channel realisations. “Optimal” - the case of fully-
digital beamforming with the optimal BS mode combination. “Hybrid” - the case of hybrid beamforming.................................................................58

Figure 3-21: CDF of the total power consumption for 3 coordinated BSs including the hardware power consumption $SvPa$ and $SvPs$. The target spectral efficiency is 4 bit/s/Hz......59

Figure 3-22: CDF of the total power consumption for 2 coordinated BSs including the hardware power consumption $SvPa$ and $SvPs$. The target spectral efficiency is 4 bit/s/Hz......59

Figure 4-1: Single rectangular patch element. (a) Geometry, (b) simulated reflection coefficient, and simulated (c) co- and (d) cross-polar components of the realized gain patterns computed at the central frequency.................................................................65

Figure 4-2: 1×4-element linear array based on rectangular patch elements. (a) Geometry, (b) simulated scattering matrix parameters, and simulated co-polar components of the realized gain active patterns computed at the central frequency on the two principal planes (phi = 0° (c) and phi = 90° (d)).................................................................65

Figure 4-3: Simulated co-polar components of the realized gain patterns computed at the central frequency on the plane phi = 0° at 25 GHz. (a) 1×4-, (b) 4×4-, (c) 1×8-, and (d) 8×8-element linear arrays based on rectangular patch elements..............................66

Figure 4-4: Single and crossed dipole: geometry, return loss and gain pattern......................68

Figure 4-5: Input impedance for HV polarization (up) and ±45° polarization (down)........68

Figure 4-6: Radiation patterns for several array configurations: geometry and gain pattern. 69

Figure 4-7: 6×6 array: periodic (up) and non-periodic in the horizontal plane (down)..........70

Figure 4-8: Single polarization element with connector, UE and AP configurations.........70

Figure 4-9: Input impedance measurement of the dipole -45° (left) and radiation pattern measurement of the dipole+45° (right) .....................................................71

Figure 4-10: input impedance (dB) for dipole +45° (left) and -45° (right)........................71

Figure 4-11: 2D radiation pattern (directivity) of the dipole +45° at 26 GHz.....................72

Figure 4-12: 2D radiation pattern (directivity) of the dipole -45° at 26 GHz.....................72

Figure 4-13: (a) Schematic view of the electronically reconfigurable Transmitarray. (b) 1-bit linearly-polarized electronically reconfigurable unit-cell architecture [DCD+16]........73

Figure 4-14: Simulated realized gain computed at the central frequency ($f_0$) as a function of the scanning angle. (a) 1-bit electronically reconfigurable Transmitarray and (b) 2-bit electronically reconfigurable Transmitarray..............................74

Figure 4-15: Simulated realized gain of the 1- and the 2-bit electronically reconfigurable Transmitarrays.................................................................74

Figure 4-16: Optimized phase distribution on the array aperture. 1-bit electronically reconfigurable Transmitarray for scan angle (a) 0° and (b) 20°. 2-bit electronically reconfigurable Transmitarray for scan angle (c) 0° and (d) 20°.....................................................75

Figure 4-17: Schematic view of the 3-bit passive unit-cells in K-band.................................75

Figure 4-18: Frequency response (gain) of the fixed beam Transmitarray operating in K-band as a function of the phase quantization.................................................................76

Figure 4-19: Simulated radiation patterns of fixed beam Transmitarray for backhaul/fronthaul computed at the central frequency as a function of the phase quantization.................................76

Figure 4-20 EIRP of the adjacent channel distortion of an 8x8 element phased array, sweeping one beam in both azimuth and elevation.......................................................77
Figure 4-21(a) Transmit signal PSD vs. AoD, (b) Error signal PSD vs. AoD using a behavioural model, (c) Error signal PSD vs. AoD using the statistical model .............................................. 78

Figure 4-22 Phase noise power spectral density ............................................................... 81

Figure 4-23 BLER results for PTRS densities for a 100 PRB allocation ............................ 82

Figure 4-24 BLER results for PTRS densities for different PRB allocations ........................ 82

Figure 4-25 EVM performance of uplink multi-user MIMO-OFDM (K=2, M=4, Ts = 10 ns) under different PNs. .................................................................................................................................................. 84

Figure 4-26: CDF of the UE throughput with HBF for different values of L_{db} and for P=4 .... 86

Figure 4-27: CDF of the UE throughput with HBF for different values of L_{db} and for P=8, 16 .. 86

Figure 4-28 PDF of difference on the equivalent channel between two consecutive OFDM symbols in terms of EVM as defined in (4.19) for both case I and case II at 28 GHz ............... 88

Figure 4-29: Performance comparisons without and with RF impairments. Two oscillator implementations case I and case II are considered. Operating frequency are (a) 28 GHz and (b) 82 GHz respectively ........................................................................................................... 88
List of Tables

Table 1-1: Mapping of WP5 models, solutions and studies to deliverable sections .................. 3
Table 2-1: Simulation parameters ........................................................................................... 9
Table 2-2: Simulation parameters and parameters for 26 GHz ............................................ 20
Table 2-3: Simulation results with the basic antenna model for 26 GHz ................................. 21
Table 2-4: Comparison between basic and directional antenna model for 26 GHz ............... 23
Table 2-5: Simulation parameters .......................................................................................... 25
Table 2-6: Simulation parameters ......................................................................................... 29
Table 2-7: Performance of the designed beams. .................................................................... 34
Table 3-1: Simulation Parameters .......................................................................................... 42
Table 4-1: Antenna specifications for the mm-wave user terminal aligned with practical deployments ................................................................................................................. 63
Table 4-2: Antenna specifications for the mm-wave access point aligned with practical deployments ................................................................................................................. 63
Table 4-3: Antenna specifications for the mm-wave backhauling/fronthauling aligned with practical deployments ............................................................................................ 63
Table 4-4: Realized gain of the 1×4-element array based on rectangular patch at the central frequency ..................................................................................................................... 66
Table 4-5: Realized gain of the 2×4-element array based on rectangular patch at the central frequency ..................................................................................................................... 66
Table 4-6: Realized gain of the 4×4-element array based on rectangular patch at the central frequency ..................................................................................................................... 67
Table 4-7: Realized gain of the 1×8-element array based on rectangular patch at the central frequency ..................................................................................................................... 67
Table 4-8: Realized gain of the 8×8-element array based on rectangular patch at the central frequency ..................................................................................................................... 67
Table 4-9: Array antenna gains ............................................................................................. 70
Table 4-10: Characteristics of the electronically reconfigurable Transmitarray for mm-wave access point .................................................................................................................. 73
Table 4-11: Realized gain of the 20×20 1-bit Transmitarray computed at the central frequency ($f_0$) ...................................................................................................................... 74
Table 4-12: Realized gain of the 14×14 2-bit Transmitarray computed at the central frequency ($f_0$) ...................................................................................................................... 74
Table 4-13: Characteristics of the fixed-beam Transmitarray for mm-wave backhauling/fronthauling. .................................................................................................................... 76
Table 4-14 Example of parameter sets for the multi-pole/zero phase noise modelling.......... 80
Table 4-15: Performance results for HBF with P=4 RF chains and phase shifter errors ....... 85
## List of Abbreviations

<table>
<thead>
<tr>
<th>ABF</th>
<th>Analog Beamforming</th>
<th>GA</th>
<th>General Assembly</th>
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<tbody>
<tr>
<td>ADC</td>
<td>Analog-to-Digital Converter</td>
<td>Gbps</td>
<td>Gigabit per second</td>
</tr>
<tr>
<td>AP</td>
<td>Access Point</td>
<td>GPRS</td>
<td>General Packet Radio Service</td>
</tr>
<tr>
<td>BS</td>
<td>Base Station</td>
<td>GSM</td>
<td>Global System for Mobile communications</td>
</tr>
<tr>
<td>CAPEX</td>
<td>Capital Expenditure</td>
<td>GSM-R</td>
<td>Global System for Mobile communications – Railway</td>
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<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
<td>HARQ</td>
<td>Hybrid Automatic Repeat reQuest</td>
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<td>CDMA</td>
<td>Code Division Multiple Access</td>
<td>HetNet</td>
<td>Heterogeneous Network</td>
</tr>
<tr>
<td>CO</td>
<td>Confidential</td>
<td>HBF</td>
<td>Hybrid Beamforming</td>
</tr>
<tr>
<td>CoMP</td>
<td>Coordinated MultiPoint</td>
<td>HTD</td>
<td>Horizontal Topic Driver</td>
</tr>
<tr>
<td>CPM</td>
<td>Conference Preparatory Meeting</td>
<td>ICI</td>
<td>Inter-Cell Interference</td>
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<tr>
<td>CPRI</td>
<td>Common Public Radio Interface</td>
<td>ICT</td>
<td>Information and Communications Technology</td>
</tr>
<tr>
<td>CPE</td>
<td>Common Phase Error</td>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
</tr>
<tr>
<td>CRS</td>
<td>Cognitive Radio Systems</td>
<td>IF</td>
<td>Intermediate Frequency</td>
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<td>C-RAN</td>
<td>Cloud Radio Access Network</td>
<td>IMT</td>
<td>International Mobile Communications</td>
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<td>CSI</td>
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<td>INR</td>
<td>Incremental Redundancy</td>
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<td>CSI-RS</td>
<td>Channel State Information-Reference Signal</td>
<td>IPR</td>
<td>Intellectual Property Rights</td>
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<td>C2X</td>
<td>Car-to-Anything</td>
<td>IR</td>
<td>Internal report</td>
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<td>DBF</td>
<td>Digital Beamforming</td>
<td>ISD</td>
<td>Inter-Site Distance</td>
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<tr>
<td>DA</td>
<td>Direct Access</td>
<td>ITU</td>
<td>International Telecommunication Union</td>
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<tr>
<td>DAC</td>
<td>Digital-to-Analog Converter</td>
<td>ITU-R</td>
<td>International Telecommunication Union-Radio</td>
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<td>Downlink Control Information</td>
<td>KPI</td>
<td>Key Performance Indicator</td>
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<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
<td>LAN</td>
<td>Local Area Network</td>
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<td>DL</td>
<td>Downlink</td>
<td>LED</td>
<td>Light Emitting Diode</td>
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<td>DoF</td>
<td>Degree of Freedom</td>
<td>LO</td>
<td>Local Oscillator</td>
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<td>D2D</td>
<td>Device-to-Device</td>
<td>LoS</td>
<td>Line of Sight</td>
</tr>
<tr>
<td>ECC</td>
<td>Electronic Communications Committee</td>
<td>LTE</td>
<td>Long Term Evolution</td>
</tr>
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<td>EHF</td>
<td>Extremely High Frequency</td>
<td>LTE-A</td>
<td>Long Term Evolution-Advanced</td>
</tr>
<tr>
<td>EIRP</td>
<td>Equivalent Isotropically Radiated Power</td>
<td>LTE-NR</td>
<td>Long Term Evolution-New Radio</td>
</tr>
<tr>
<td>EIT</td>
<td>European Institute for Innovation and Technology</td>
<td>M</td>
<td>Milestones</td>
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<td>EMF</td>
<td>Electro Magnetic Field</td>
<td>Mo</td>
<td>Month</td>
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<td>E2E</td>
<td>End-to-End</td>
<td>MA</td>
<td>Multiple Access</td>
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<td>FDD</td>
<td>Frequency Division Duplex</td>
<td>MAC</td>
<td>Medium-access Control</td>
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<td>FSO</td>
<td>Free Space Optics</td>
<td>xiv</td>
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<tr>
<td>VCO</td>
<td>Voltage Controlled Oscillator</td>
<td></td>
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<tr>
<td>WCDMA</td>
<td>Wide Code Division Multiple Access</td>
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<td>Wireless Local Area Network</td>
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<td>Work Package</td>
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<td>WPL</td>
<td>Work Package Leader</td>
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<td>WRC</td>
<td>World Radio-communication Conference</td>
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<tr>
<td>ZF</td>
<td>Zero Forcing</td>
<td></td>
<td></td>
</tr>
<tr>
<td>3GPP</td>
<td>3rd Generation Partnership Project</td>
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</tr>
</tbody>
</table>
1 Introduction

There is now a general understanding of the varied challenges in mm-wave transceiver design and the need for accurate modelling of the transceiver components in the wider research community. Through the extensive studies done in WP5, this deliverable aims to address some of the most significant challenges in multi-node and multi-antenna architectural and schematic solution design as well as in modelling the performance and impairments in the transceiver hardware chain.

The previous deliverable D5.1 of WP5 [MMMAGICD51] (published in M9) should be seen as a precursor to this deliverable D5.2. In D5.1, many of the solutions discussed here were presented in their initial stage with the theoretical background. We do not repeat that information here, but simply refer to D5.1 wherever necessary. Also in D5.1, a complete description of the use cases and KPIs that motivate the WP5 work has been presented. We give only a very brief summary of the use cases and KPIs below, to initiate the D5.2 discussion.

The baseline use case adapted for WP5 work is the 5G immersive experience in targeted hotspots. This use case looks at the early 5G deployments, where the unique 5G immersive experience is supported at targeted, densely crowded hotspots like city centres and transport hubs. It is assumed that an underlay Macro cell network (likely to be 4G LTE) will support these hotspots to provide continuity in connectivity. The use case supports 4 typical data applications with the peak data demand expected to be up to 20 Gbps. A comprehensive analysis of the probabilistic usage of these 4 applications in a typical hotspot was presented in [QHM+16], where the 95% and 99% cumulative user data demand (per cell) comes up to 29 Gbps and 38 Gbps respectively. So the technical work in WP5 is aimed mainly at supporting these peak data rates and cell capacity demands, i.e. in-line with the eMBB mode as seen by 3GPP [3GPP-16]. Full details of the additional requirements placed by this use case can be found in chapter 2 of D5.1 [MMMAGICD51]. In addition, the work presented here pays a special attention on power consumption of mm-wave transceivers, which is expected to be a significant challenge in designing mm-wave mobile compatible systems.

WP5 considers 3 other extended use cases in addition to the base-line use case. These extensions are for extended mobility, coverage and connection density. In mobility, the extension comes from the ‘Moving Hotspots’ use case. This extends the pedestrian mobility in the baseline use case to vehicular speeds in urban areas, around 50 kmph. The coverage extension comes from the ‘50 Mbps+ everywhere’ use case. The limited hotspot coverage is extended in this use case, to provide the edgeless coverage in a wider urban area. The limited connection density of the baseline use case (40 active users per small cell) is extended in this ‘Dense urban society with distributed crowds’ use case, where connection densities up to 150,000 per km² can be considered.

Mm-wave spectrum consideration is another key area that has significantly evolved since the completion of D5.1 [MMMAGICD51]. In D5.1, we focussed on three wide frequency ranges, i.e. low (6-30 GHz), mid (31-50 GHz) and high (70-100 GHz) for developing the initial schemes. Since then, the European Commission has identified the 24.25-27.5 GHz band (commonly referred to as the 26 GHz band) as a pioneer band for mobile applications in Europe [RSPG16] and the FCC (Federal Communications Commission) has announced their intention to license the 28 GHz band for mobile services in the US [FCC16]. Accordingly, the work developed for D5.2 by WP5 has mostly focussed on the 28 GHz band, with some of the antenna modelling work in 24.25 – 27.5 GHz band and ray tracing based MMIMMO (Massive Multiple Input – Massive Multiple Output) experimental validation work in 26 GHz band as well. Most of the qualitative assessments in these bands will hold for other bands in the mm-wave frequency range as well.

Chapter 2 of this deliverable details the various multi-antenna solutions developed in the project, for access, backhaul and relay applications. After careful considerations (analysis
presented in D5.1 ([MMMAGICD51]) the hybrid beamforming architecture has been selected as the most versatile architecture to satisfy the diverse demands of radio access in mm-wave systems. Many of the solutions presented under radio access detail novel applications of hybrid beamforming. A simulation based comparison of Analog, Digital and Hybrid beamforming schemes is also presented for dense urban deployments. One of the highlights of the WP5 work agenda for the final year of the project was the experimental validation of the achievable spectral efficiencies for the MMIMMO scheme, theoretically developed in D5.1. This experimental work and the evaluated results are presented under the backhaul sub-section, as these short range, LOS communications with massive number of antenna elements mainly suits backhaul deployments. A mm-wave backhaul analysis for moving hotspots completes this backhaul section, where some performance and complexity trade-offs are derived. Under the investigations for mm-wave relay based systems, a user selection scheme for relay operations is presented. The general issues with the multi-antenna beam design and beam management/beam tracking are also covered in this chapter 2.

Multi-node co-ordination will be an essential feature for mm-wave radio access systems, as the single link reliability can be severely impacted by the challenging propagation conditions. Chapter 3 is devoted to the study of different multi-node schemes, where different configurations and technologies are evaluated. In an architectural point of view both stand-alone mm-wave systems and non-stand-alone mm-wave systems (with the likely support of LTE Macro cells) are evaluated. Some of the hybrid beamforming concepts presented earlier in Chapter 2 are extended towards multi-node co-ordinated schemes in this chapter. Also a novel concept of utilizing FSO (Free Space Optical) links in combination with mm-wave links for multi-hop networks is presented with the associated theoretical analysis. Another novel concept presented in Chapter 3 is the opportunistic usage of a few high rise mm-wave access points to complement the coverage provided by a network of low rise mm-wave access points. The effectiveness of this scheme is evaluated theoretically in this work. A comparison of the direct mm-wave access is compared with a relay assisted network of mm-wave APs, in another dimension of multi-node networks. The chapter concludes with an analysis of the Radio Resource Management (RRM) aspects for wider mm-wave networks.

The modelling of the performance and impairments of the essential hardware components in the mm-wave transceiver chain is the focus of chapter 4. This work is pushing the current levels of knowledge in this relatively new area (of application of mm-wave hardware for mobile communications). The performance and limitations of antenna arrays, suited for handset and access point (for both radio access and backhaul) applications have been modelled and in some cases proto-typed in WP5 work. A major part of chapter 4 is devoted detailing the final results from this work. The first analysis on PA (Power Amplifier) linearity performance modelling in D5.1 has been extended to a system level, to include the impact of large antenna arrays on the PA. A second PN (Phase Noise) model is also introduced in chapter 4, which has been used to analyse the effectiveness of different PTRS (Phase Tracking Reference Signal) densities in the frame structures that are now being studied for the 3GPP New Radio (NR) standardisation. A MIMO-OFDM based study on the impact of PN is also presented, using the PN model proposed in D5.1, which is now publicly available as an open source code. The chapter concludes with some analyses on the overall system level impact of these hardware impairments.

Chapter 5 contains the main takeaways, or conclusions that can be derived from the extensive work done in WP5. These conclusions will be presented as segmented per each of the technical chapters and finally as some overall conclusions. The implications of this work on an overall mm-wave system concept design will also be addressed in Chapter 5.

As noted above, D5.2 provides a continuing set of work initially reported in D5.1 and there is frequent referencing to D5.1 for the initial analyses. To help the reader to navigate the 2 deliverables more efficiently the following Table 1-1 is provided, which maps the developed models, solutions and studies to the broad themes in mm-wave transceiver design and the reported sections in the two deliverables.
### Table 1-1: Mapping of WP5 models, solutions and studies to deliverable sections

<table>
<thead>
<tr>
<th>Technology enablers</th>
<th>Access</th>
<th>Backhaul</th>
<th>Relay/ Self-Backhaul</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Antenna Array design</strong></td>
<td>Dual polar dipole array (D5.2 - section 4.2.3)</td>
<td>Switched Beam Transmitarray (D5.2 - section 4.2.5)</td>
<td>Switched Beam Transmitarray (D5.2 - section 4.2.5)</td>
</tr>
<tr>
<td></td>
<td>Patch antenna array (D5.2 - section 4.2.2)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>Electronically reconfigurable Transmitarray (D5.2 - section 4.2.4)</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Multi-antenna schemes</strong></td>
<td>Flexible SU and MU Hybrid BF design (D5.2 - section 2.4.1)</td>
<td>MMIMO B-DFT-SM-MRT decoding scheme (D5.1 - section 3.4.1)</td>
<td>Ad-hoc relay user selection scheme (D5.1 - section 3.5.1 and D5.2 - section 2.6.1)</td>
</tr>
<tr>
<td></td>
<td>Arbitrary beam synthesis for Hybrid BF (D5.2 - section 2.7.1)</td>
<td>MMIMO experimental validation (D5.2 - section 2.5.1)</td>
<td></td>
</tr>
<tr>
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<td>Wide beams for broadcast signalling (D5.2 - section 2.7.2)</td>
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<td>Broadband Hybrid BF (D5.1 - section 3.3.5)</td>
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<tr>
<td></td>
<td>Max SINR Digital BF (D5.1 - section 3.3.3)</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Multi-node schemes</strong></td>
<td>Beam sweeping for multi-node networks (D5.2 - section 3.2)</td>
<td>Hybrid mm-wave/FSO Multi-hop links (D5.2 - section 3.5)</td>
<td>Relay Node enhancements for Q2I coverage (D5.2 - section 3.7)</td>
</tr>
<tr>
<td></td>
<td>Sequential Hybrid BF Design for Multi-links (D5.2 - section 3.3)</td>
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</tr>
<tr>
<td></td>
<td>Beam Management for Mobility (D5.2 - section 3.4)</td>
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<tr>
<td></td>
<td>Complementary coverage with High rise AP (D5.2 - section 3.6)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>Energy efficient joint Hybrid pre-coding (D5.2 - section 3.8)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Model enablers for performance analysis**

| | Progress Beyond SOTA | Power Amplifier | Quantization noise | I/Q Imbalance |
| | Detailed Model (D5.1 - section 4.2.1) | Statistical and Behavioural Models (D5.2 - section 4.3) | Analysis (D5.1 - section 4.3.4 and D5.2 - section 2.4.3) | Analytical Model (D5.1 - section 4.2.3) |
| | Simpler Model (D5.2 - section 4.4) | | | |
2 Multi-Antenna Transceiver Schemes and Design

2.1 Introduction

Contrary to lower frequencies, propagation at mm-wave faces significantly distinct challenges including higher path loss, penetration losses and shadowing. To address these adverse aspects, highly directional transmission and reception has been proposed, by deploying large antenna arrays at the transmitter and receiver [RRE14]. To this end, the small wavelength at higher frequencies can be exploited, allowing to pack a large number of antennas in a small area. The large antenna array sizes as well as the directional transmission coupled with hardware constraints at mm-wave, result in specific challenges on the mm-wave transceiver architecture and design [RRE14].

In this chapter we present several multi-antenna transceiver designs and schemes which have been developed for access, backhaul and relaying for mm-wave communication. Most of these schemes are based on a hybrid beamforming architecture, for which a brief overview is provided in the following, including a discussion about fully connected and subarray-based architectures. We conclude the chapter by discussing some channel state information (CSI) aspects and providing some concluding remarks.

2.2 Hybrid Beamforming Transceiver Architecture

Due to the large number of antennas in the transmit and receive arrays required to enable mm-wave communication, equipping each antenna with a separate radio frequency (RF) transceiver chain along with a high resolution converter, as done with smaller arrays at lower frequencies, would result in a high complexity, cost and power consumption. This is mainly due to the implementation of RF components at mm-wave frequencies, as well as the expected large bandwidths, which impose the requirement on the converters (digital to analog converters (DACs) at the transmitter and analog to digital converters (ADCs) at the receiver) to operate at a high sampling rate. Thus, equipping one converter per antenna in a large antenna array translates inevitably into a high power consumption.

To address these aspects, several transceiver architectures have been considered including analog beamforming with a single RF chain, such as in IEEE 802.11ad. Since this architecture offers limited signal processing capability, other alternatives have also been proposed including digital beamforming with low resolution converters [MH15]. Reducing the precision of the converters enables to reduce the power consumption, which scales roughly exponentially with the number of resolution bits [Mur97]. This enables to have a large antenna array with many active elements at a reduced power consumption and cost. Despite the increased signal processing capabilities compared to analog beamforming, the non-linearity introduced by the quantization leads to limited capacity at high SNR as well as imposes certain challenges on the channel estimation and data detection [MH15].

Besides the previous transceiver architectures, another approach to tackle the power consumption and complexity bottleneck facing mm-wave transceivers is to reduce the number of RF chains (including converters) with the hybrid beamforming architecture [AAL+14] as depicted in Figure 2-1. With hybrid beamforming the number of RF chains \( N_{RF} \) is less than the number of antennas in the array, e.g. \( N_{tx} \) at the transmitter. By splitting the beamforming operation between the analog RF domain and the digital baseband, this architecture provides a tradeoff between performance and complexity / power consumption [AAL+14], at the expense, however, of reduced degrees of freedom for the baseband digital processing (number of streams \( N_s \)). Compared to a fully digital system, the hybrid beamforming architecture poses different challenges for the CSI acquisition and beamforming design due to the constraints on the analog processing and the need of directional transmission for mm-wave.
Figure 2-1: Hybrid Beamforming with Analog RF beamforming

For example, the analog processing is frequency flat, which implies that the analog beamforming matrix is fixed for all subcarriers in a multicarrier system, whereas the digital beamformer can be adapted for each subcarrier. For the wideband hybrid beamforming design, however, the fact that the spatial characteristics of the channel are frequency invariant can be exploited [IVC+16]. Similarly for hybrid beamforming in a multiuser scenario, the design of the analog beamforming needs to consider that it is common for all users. Furthermore, the analog processing can be implemented via a network of phase shifters, RF switches or with a lens antenna array, and can be implemented at different stages including RF, intermediate frequency and baseband [KPS+13]. In case of phase shifters, the entries of the analog beamformer are constrained to have unit modulus. For further discussions about the hybrid beamforming architecture, refer to [MMMAGICD51].

2.3 Hybrid Beamforming with a Subarray Architecture

Assuming that the analog RF beamforming for the hybrid beamforming transceiver architecture is implemented with a network of phase shifters, we would then require one phase shifter per antenna and per RF chain, as depicted for the fully connected architecture in Figure 2-2(a) (for the transmitter). Denoting the number of antennas at the transmitter as \( N \) and the number of available RF chains as \( N_{RF} \), the analog beamformer is implemented with a network of \( N \times N_{RF} \) phase shifters, since the output of each RF chain is connected to all the transmit antennas in a fully connected architecture. Furthermore, a 1 to \( N \) (1:N) divider is required for each RF chain, whereas one adder over \( N_{RF} \) inputs (\( N_{RF}:1 \)) is needed for each antenna as shown in Figure 2-2(a). Such a fully connected architecture, however, might be difficult to implement in practice at higher frequencies. In addition, the fully connected architecture can result in substantial losses when considering that losses in the \( N_{RF} \) dividers scale linearly with \( N \), while the losses in the \( N \) adders scale linearly with \( N_{RF} \) [GVR+16].

To this end, the partially connected architecture has recently been proposed, where the output of each RF chain is connected only to a subset of the transmit antennas [ZHD+15], [HIX+15]. This approach reduces the required number of phase shifters as well as the losses, thereby facilitating the implementation of hybrid beamforming at the expense of reduced design flexibility. If each RF chain (or set of RF chains) is connected to a distinct set of antennas, the architecture is based on subarrays, where each subarray is basically connected to its own transceiver. The partially connected architecture is applicable at both the transmitter and receiver side.
In Figure 2-2(b), a partially connected architecture with subarrays is depicted assuming subarrays of equal size with \( N_{\text{sub}} = N / N_{\text{RF}} \) antennas, i.e. each RF chain is connected to \( N_{\text{sub}} \) antennas. In this case, the analog beamforming is implemented with a network of \( N_{\text{sub}} \times N_{\text{RF}} \) phase shifters, resulting in a factor of \( N_{\text{RF}} \) fewer phase shifters compared to the fully connected case. In addition, the required dividers are smaller, i.e. a \( 1:N_{\text{RF}} \) divider is required for each RF chain, such that the losses in the dividers scale linearly with \( N_{\text{sub}} \) \cite{GVR+16}. Furthermore, note that the subarray architecture does not require adders. Based on the fact that \( N_{\text{RF}} \) fewer phase shifters are required, that the losses in the dividers are \( N_{\text{RF}} \) times smaller and in addition, no adders are needed, the subarray architecture has at least a factor of \( N_{\text{RF}} \) less losses compared to the fully connected case.

The reduced losses for implementing hybrid beamforming with the subarray architecture come at the expense of reduced flexibility which translates into reduced performance. In order to make a comparison with a fully connected architecture, the losses need to be characterized \cite{GVR+16}. By assuming certain values for the losses, however, the performance of each scheme is then coupled to the specific values for the assumed losses. To avoid this, we propose for comparison purposes to employ a simple approach independent of specific values for the losses. Since the subarray architecture has roughly \( N_{\text{RF}} \) less losses compared to the fully connected architecture, we penalize the transmit power for the fully connected case by a factor of \( N_{\text{RF}} \).

To observe the tradeoff between reduced design flexibility and reduced losses introduced by the subarray architecture, we consider the following point-to-point scenario with 64 antennas at the transmitter and 16 antennas at the receiver. For the channel, we assume a narrowband channel model as in \cite{ARA+14} consisting of 8 clusters, with 10 paths per cluster. We assume a hybrid beamforming architecture at the transmitter and receiver with \( N_{\text{RF}} = 4 \) RF chains, i.e. 4 streams are transmitted. For the receiver, we consider a fully connected architecture, whereas for the transmitter, we compare a fully connected architecture with a subarray architecture composed of 4 subarrays each consisting of 16 antennas. We assume perfect CSI and compute the rate with 4 streams based on the Shannon formula. The rate achieved with the two architectures without considering losses is depicted in Figure 2-3, where the hybrid beamforming design for the fully connected case is based on \cite{ARA+14} while the hybrid beamforming design for the subarray architecture is based on \cite{IVU+17}. As expected, the fully connected case outperforms the subarray architecture due to the increased flexibility for the
beamforming design. However, when considering the losses by introducing a power penalty for the fully connected case, the subarray architecture can outperform the fully connected architecture as shown in Figure 2-3. As detailed before, for the given scenario with $N_{RF} = 4$, the power penalty on the fully connected case corresponds to a 6 dB shift to the right, since the subarray architecture has $N_{RF}$ less losses compared to the fully connected case.

![Figure 2-3: Performance Simulation Comparison with Subarrays ($N_{RF} = 4$)](image)

2.4 Transceiver Schemes for Access

After the brief overview of the transceiver architectures as well as discussing hybrid beamforming with subarrays, we now present multi-antenna transceiver designs and schemes for access. Although most of the developed schemes are based on hybrid beamforming, we also discuss analog beamforming and digital beamforming. The presented schemes in general assume a TDD system.

2.4.1 Flexible Multi-User Hybrid Beamforming Design

Consider a base station (BS) serving $J$ users (UEs) in the downlink as shown in Figure 2-4. Assume $N_{BS}$ transmit antennas and $N_{UE,j}$ receive antennas are implemented at the BS and $j$-th UE, respectively. The numbers of RF chains are $M_{BS}$ and $M_{UE,j}$ at the BS and $j$-th UE, respectively. We propose to design the hybrid transmission scheme in three sequential phases based on given scenarios, available system resources and desired performance targets:

- Phase 1: Uncoupled analog beamforming at the BS and UE(s)
- Phase 2: Digital beamforming at the UE
- Phase 3: Precoding/digital beamforming at the BS
As a starting point, we denote the mathematical input-output relations of the hybrid beamforming design in Figure 2-4 as

\[
\hat{s}_j = W_{D,j} W_{A,j} H_j F_A F_D F_s + W_{D,j} W_{A,j} n_j
\]

(2-1)

where \( s_j \) is the \( L_j' \times 1 \) transmitted signal vector for the \( j \)-th UE, \( F_D \) and \( F_A \) are the digital precoder and analog beamformer at the BS, \( H_j \) refers to propagation DL channel from the BS and to the \( j \)-th UE, \( W_{D,j} \) and \( W_{A,j} \) are digital combiner and analog beamformer at the \( j \)-th UE, \( s = [s_1^T, ..., s_J^T] \), \( j = 1, ..., J \). Here, for simplicity, we assume \( L_j' = M_{BS,j} = M_{UE,j} \), \( \sum_{j=1}^{J} L_j' = L \) and \( \sum_{j=1}^{J} M_{BS,j} = M_{BS} \). Furthermore, we point out that the proposed scheme is also applicable for the single user case, where multiple streams are sent to one user.

**Phase 1: Uncoupled analog beamforming at the BS and UE(s)**

In this phase, the BS sets the digital precoder to be \( F_D = I \) and only applies analog beamforming \( F_A \) on the transmitted signal. At the UE side, the RX carries out analog beamforming \( W_{A,j} \) and for equalization considers only the effective channel of each transmission path, i.e., \( W_{D,j} \) is set to be a diagonal matrix. The analog beamformers at the UE and BS are designed to use antenna array response matrices of those channel paths with the largest gains as described in [ZRG16], [MIMGIC51]. For phase 1, the BS needs to know the angles of departure(s) (AoD(s)) of all the used transmission paths, whereas each UE needs to know the angles of arrival (AoAs) and the complex gain(s) of the used transmission path(s) from the BS to this UE.

With the use of the uncoupled analog beamforming in phase 1, the system doesn’t cope with inter-stream interferences in the single-user case and inter-stream/inter-user interferences in the multi-user case. This should suffice if different paths/users are separated well in angular domain at the BS and UE(s) sides, and/or targeted link performance is not very demanding, and/or low computational complexity and overhead is of dominant interest.

**Phase 2: Digital beamforming at the UE**

In this phase, after conducting analog beamforming as in phase 1, each UE estimates the effective channel from the BS to this UE including propagation channel and the used analog beamformers. With reasonable good effective DL channel estimation, the multi-stream interferences can be removed by constructing digital combiner \( W_{D,j} \) using any MIMO.
equalization algorithm available in the literatures at the UE. Compared to digital combiner used in phase 1, \( W_{p,j} \) is most likely a non-diagonal matrix.

After phase 1 and phase 2 processing and without precoding at the BS side, the link performance for single-user case is already close to the achievable performance of reference hybrid beamforming scheme in [ARA+14]. In multi-user scenario, if multiple users that have good angular separations at the BS are scheduled to be served at the same time, it is still possible to achieve reasonably good system performance.

**Phase 3: Precoding/digital beamforming at the BS**

In the last phase, if the system requires multi-user interference mitigation to achieve the targeted performance metrics, the BS can carry out multi-user interference nulling or reduction by applying precoding, e.g., using zero-forcing and MMSE criterions. To this end, however, the BS needs to acquire knowledge of DL effective channel (transmit CSI), which leads to extra system overhead. Meanwhile, the use of zero-forcing or MMSE precoders could potentially increase PAPR of the transmitted signal dramatically. Thus more intelligent and power efficient precoding schemes need to be developed. For the numerical analysis, we focus only on the first two phases which are the most relevant for mm-wave transmission.

First, we consider a single-user case using numerical computer simulations with parameters given in Table 2-1. In Figure 2-5, the performance of the proposed analog (phase 1) / hybrid beamforming (phase 1 + phase 2) are compared with SVD digital beamforming and hybrid beamforming approach in [ARA+14] using 3 RF chains at BS and UE sides with different codebook sizes \( \gamma_{N_{BS}} \) and \( \gamma_{N_{UE}} \), where the vectors in the codebook are equally distributed over the phase domain. For digital beamforming full CSI is assumed, and we compare the results with CSI from all the 5 available paths and from only the 3 used paths. Naturally the former case can provide slightly better performance than the latter case as more CSI is available. For examining the proposed algorithm and reference hybrid beamforming approach in [ARA+14], we assume with knowledge of the 3 used paths and finite size codebook as well. As shown in Figure 2-5, the proposed algorithm can achieve the same performance as the algorithm in [ARA+14] in all cases, but with reduced channel knowledge. Whereas the proposed hybrid beamforming only needs to know AoDs at the BS, the algorithm in [ARA+14] requires full channel knowledge at the BS. In addition, if the targeted uncoded BER is \( 10^{-1} \), analog beamforming in phase 1 can already form a simple solution.

### Table 2-1: Simulation parameters

<table>
<thead>
<tr>
<th>Simulation Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating Frequency</td>
<td>28 GHz</td>
</tr>
<tr>
<td>Path Loss Exponent</td>
<td>2</td>
</tr>
<tr>
<td>BS-UE Distance</td>
<td>50 m</td>
</tr>
<tr>
<td>TX power</td>
<td>27 dBm</td>
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<tr>
<td>Number of Antennas at the BS</td>
<td>64</td>
</tr>
<tr>
<td>AoA (AoD)</td>
<td>Uniform Distribution</td>
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<td>Modulation</td>
<td>Single-carrier 16QAM</td>
</tr>
<tr>
<td>Antenna Array Type</td>
<td>Uniform Linear Array</td>
</tr>
<tr>
<td><strong>Single user case</strong></td>
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</tr>
<tr>
<td>Number of Antennas at the UE</td>
<td>16</td>
</tr>
<tr>
<td>Number of Paths in Channel</td>
<td>5</td>
</tr>
<tr>
<td>Number of Streams(RFs)/Used Paths</td>
<td>3</td>
</tr>
<tr>
<td><strong>Multiuser case</strong></td>
<td></td>
</tr>
<tr>
<td>Number of Antennas at the UE</td>
<td>8</td>
</tr>
<tr>
<td>Number of Paths for each UE</td>
<td>3</td>
</tr>
<tr>
<td>Number of Streams for each UE</td>
<td>2</td>
</tr>
</tbody>
</table>
In the following we assume $\gamma = 1$ for determining the codebook size. Next, the impact of signal blockage is examined, assuming the strongest path is blocked. In this case, for SVD-based digital beamforming, a re-calculation of the digital beamformer at the both BS and UE sides is required. On the other hand, without any re-estimation and re-calculation and by employing a reduced number of RF chains at the receiver, the proposed hybrid beamforming design can immediately remove the blocked signal path and continue transmission other two available paths as shown in Figure 2-6. In the figure, “digital beamforming a. blocking” refers to the case where the digital beamformers are calculated for the 3 paths and this beamformer is continued to be used when the strongest path is blocked, i.e. with only 2 paths. “Proposed hybrid beamforming a. blocking” is defined similarly as before, but for hybrid beamforming. The case “digital beamforming ideal a. blocking” refers to the case where the digital beamformers are calculated for the remaining 2 paths after the strongest path has been blocked.

![Figure 2-5: BER simulation comparisons for the single user case](image)

![Figure 2-6: BER comparisons for the single user case after blockage of main path.](image)

We now study the achievable performance of the proposed hybrid beamforming design framework for a 2-user case with parameters given in Table 2-1. No advanced schedule is applied here. AoDs of different UEs are assumed to be uniformly distributed over $[-0.5\pi, 0.5\pi]$. For reference, we consider eigenbeamforming using perfect CSI on all the $2 \times 3$ paths and the used $2 \times 2$ paths in the channel respectively. As shown in Figure 2-7, without knowledge on all the $2 \times 3$ paths, the eigenbeamforming also suffers from multi-user interferences (blue curves). Without deploying any digital beamformer at the BS, the proposed
hybrid beamforming (phase 1 + phase 2) can achieve performance close to digital beamforming based on the available paths up to SNR at around 0 dB. More sophisticated digital beamformers (phase 3) can be designed later to cope with multi-user interference if more channel information is available at the BS.

![Figure 2-7: Performance comparisons for the multi-user case](image)

Finally, the blocking effect is studied in the multi-user context in Figure 2-8. Assume the strongest path to UE #1 is blocked. Then by modifying the digital beamformer for UE #1 corresponding to the remaining path, UE #1 can continue its reception over the second path as usual. In the figure, “digital beamforming a. blocking” and “digital beamforming ideal a. blocking” refer to the cases that the eigenbeamforming design based on the channel before blocking (2+2 paths) and after blocking (1+2 paths) respectively. It is interesting to observe that the transmission for UE #2 is not influenced using either digital beamforming or the proposed hybrid beamforming. In general, this shows the robustness of the proposed hybrid beamforming design to signal blockage in the multi-user case as well.

![Figure 2-8: Performance comparisons for the multi-user case after blockage of the strongest path of UE#1.](image)

2.4.2 System Level Comparison of Different Transceiver Architectures for Access

As discussed in Section 2.2, by having a number of RF chains which is less than the number of antennas, hybrid beamforming (HBF) represents a good trade-off between performance and complexity for mm-wave access. Several works have already shown that HBF can strongly outperform a pure analog beamforming (ABF) and approach the performance
achieved by fully digital precoding (DP), with as many RF transceiver chains as antennas, at least in scenarios with only one BS serving one [AHA+12], [ARA+14] or multiple UEs [ALH15]. However, more studies are needed in particular on a system level to properly assess these different hardware architectures and corresponding beamforming schemes. Here, we consider a system with multiple BSs, each serving many UEs. Therefore, the UEs served by a BS can be characterized by very different channel conditions: for instance, a BS may be simultaneously serving UEs in line-of-sight (LOS), i.e., with typically good signal to noise ratio (SNR), and UEs in non-line-of-sight (NLOS), i.e., characterized by very strong attenuation and by a frequency selective channel. Therefore, besides the fact that a proper user selection algorithm is needed, the few RF chains of the HBF architecture should be employed in different ways: either to transmit/receive along the different multipath components when serving a specific NLOS UE, or to (spatially) multiplex multiple LOS UEs, if they are sufficiently separated in the angular domain.

In detail, we consider a hexagonal deployment with seven sites, three BSs per site, an inter-site distance (ISD) of 200 m, and wraparound. Each BS is equipped with a horizontal uniform linear array with 64 antennas, each with 8 dBi gain. Moreover, we consider a dense area with 50 full-buffer UEs per site, each assumed to have 1 RF transceiver chain and a beamforming gain of 12 dBi, with the beamformer pointing toward the direction that maximizes the long-term SNR. The carrier frequency is 28 GHz and the system bandwidth is 1 GHz. We consider the 3D channel model proposed in [TNM+14], which is based on measurements done in New York City and characterizes the channel in azimuth/elevation as well as in time/frequency domain. In this scenario we compare the following beamforming schemes:

- ABF: Only one UE is scheduled per subframe on the whole band and the RF beams are designed by using a codebook based on the discrete Fourier transform (DFT) matrix [CYC+11];
- HBF: The BS is equipped with $P$ RF chains and serves up to $P$ UEs per sub-frame, all on the whole band; also for this scheme the analog RF beams are designed by using the DFT codebook;
- DP: The BS employs maximum ratio transmission (MRT) to serve the UEs, different set of UEs can be served on each sub-band and there is no limit in the number of UEs that can be served in each subframe.

A proportional fair scheduler with a greedy user selection algorithm is considered with all the three schemes. More details about the beamformers and the scheduler as well as further numerical results can be found in [GRM+16, GRB+16]. In Figure 2-9, we assume perfect CSI at the BS and depict the cumulative distribution function of the UE throughput. First of all, we notice that very high rates can be achieved at mm-wave, mainly because of the large available bandwidth: even with ABF, which neither exploits spatial multiplexing nor properly deals with multipath fading, a median value of about 500 Mbit/s is achieved. Then, we also observe that HBF, even with only $P = 4$ RF chains, can strongly outperform ABF, by providing a gain of about 100% in the cell border throughput. Moreover, we also observe that the performance of HBF increases when the number of RF chains increases: with $P = 16$, HBF reaches DP in the upper part of the CDF curve (mainly related to UEs in LOS), however there is still some gap in the lower part of the curve (mainly related to NLOS UEs).
In Figure 2-10 we remove the assumption of perfect CSI knowledge and assess the effect of using an outdated CSI at the BS. In detail, we assume that the BS gets an update of the CSI only every $T$ ms. We however consider values of $T$ that cause changes only in the small scale fading components of the channel, whereas the large scale fading parameters like path-loss, shadowing and arrival/departure angles remain constant. Note that in this setup the UE speed is 3 km/h, therefore, for example, $T = 50$ ms corresponds to $f_D T = 3.8$, with $f_D$ being the Doppler frequency. In Figure 2-10, we observe that the throughput achieved with HBF hardly decreases when $T$ increases. On the other hand, the DP significantly decreases when the CSI is outdated. This can be easily explained by observing that DP strongly exploits the precise knowledge of the small scale fading in the precoder design, whereas HBF, where RF beams are wideband, relies less on the small scale fading. While at mm-wave, it is expected that the channel will be tracked at the BS only every few ms (or few tens of ms), i.e., much less than for instance $T = 500$ ms (which is reported here only as a case where there is basically no knowledge of the small scale fading), these results show that achievable HBF performance is much more robust to CSI imperfection when compared to DP performance. For HBF, beamformers can be designed by relying mainly on the large scale fading components of the channel, while DP requires timely CSI knowledge in order to exploit all the degrees of freedom offered by the fully digital design.
2.4.3 Digital Beamforming with 1 bit DACs

As discussed in Section 2.2, to limit cost, power consumption, and hardware complexity, digital beamforming with low precision converters can be considered. In this subsection, we analyse the performance (both in terms of uncoded BER and achievable rates) achievable over the multiuser (MU) MIMO downlink for the case when the base station is equipped with a large number of antennas, each one connected to an individual RF port. In addition, we shall consider the extreme case of a base station equipped with 1-bit resolution DACs and investigate the downlink performance achievable with linear precoders and OFDM over a frequency selective channel. Apart from a reduction in system costs and power consumption, this architectural solution may also significantly lower the load of the fronthaul link connecting the baseband processing unit to the RF ports for mm-wave access.

Specifically, we consider the downlink of a single-cell massive MU-MIMO-OFDM as depicted in Figure 2-11. The BS, which is equipped with $B$ antennas, serves $U$ single-antenna users simultaneously and on the same time-frequency resources. At the BS, a linear precoder (e.g., zero-forcing (ZF) precoder or maximal-ratio transmission (MRT) precoder) maps the data symbols into precoded symbols to be sent to the RF port. We assume OFDM transmission over a frequency-selective channel. Hence, the frequency-domain precoded vector is transformed into a time-domain vector by the IDFT at each RF port. Then, the resulting signal is quantized using a pair of 1-bit DACs at each RF port, one for the real part and one for the imaginary part of the time-domain precoded signal. At the UEs, the received time-domain signal is converted to frequency domain through a DFT.

For simplicity, we assume that all RF hardware components other than the DACs (e.g., local oscillators, mixers, power amplifiers, etc.) are ideal and that the ADCs at the UEs have infinite resolution. We also assume that the sampling rate of the DACs at the BS is equal to the sampling rate of the ADCs at the UEs and that the system is perfectly synchronized. In our analysis, we allow a fraction of the OFDM subcarriers to be left unused, which corresponds to oversampling the transmit signal before digital-to-analog conversion, which is advantageous performance wise. Our approach generalizes most analyses available in the literature (see e.g., [SFS16], [LTS+17], [JDC+16], [JMM+16]), which assume symbol-rate sampling and also single-carrier transmission over a frequency-flat channel.

Using Bussgang’s theorem, which allows to characterize the output of a nonlinear operation when the input is Gaussian, we develop a closed-form expression for the signal-to-interference-noise-and-distortion ratio (SINDR) of linear precoders. We then use this expression to obtain an accurate approximation of the uncoded BER achievable with QPSK signalling and a lower bound on the achievable sum-rate downlink throughput. This lower bound corresponds to the rate achievable using Gaussian signalling and mismatched nearest-neighbour decoding at the UEs (see [JDC+17] for more details regarding the closed-form expression and theoretical analysis). We next illustrate our results by focusing on a MU-MIMO-OFDM system in which the BS is equipped with $B = 128$ antennas with $U = 16$ UEs. For
simplicity, we consider a frequency-selective Rayleigh-fading channel with 4 independent channel taps and uniform power-delay profile. The performance over more realistic mm-wave channels, where the channel taps are not necessarily Rayleigh distributed, are expected to be similar. We assume that an OFDM symbol consists of 512 subcarriers, 300 of which are occupied and the rest are left unused to provide oversampling ratio of about 1.7.

In Figure 2-12, we plot the uncoded BER with QPSK for MRT and ZF precoding as a function of the SNR. Both the case of 1-bit DACs and of infinite-resolution DACs are considered. The simulated BER values (given by the markers in the figure) are compared with an analytical approximation (solid lines in the figure) for the BER obtained by assuming that the overall noise at the UEs (which includes MU interference and quantization errors) is Gaussian [JDC+17]. We see from the figure that MRT suffers from a relatively high error floor, which is mainly due to residual MU interference, whereas ZF yields better performance, although the gap at $10^{-4}$ from the performance obtainable with infinite-resolution DACs is about 8 dB. Our closed-form approximation is accurate over the entire range of SNR values considered in the figure. This suggests that the overall noise at the UEs can indeed be modelled accurately as a Gaussian random variable.

![Figure 2-12: Uncoded BER with QPSK](image)

In Figure 2-13, we show the sum-rate throughput achievable with Gaussian inputs and mismatched nearest-neighbour decoding as a function of the SNR. We observe that, similarly to the frequency-flat case [JDC+16], a high sum-rate throughput can be achieved. Specifically, a sum-rate throughput exceeding 64-bits per channel use (corresponding to 4 bits per channel use per UE) can be achieved for SNR values beyond 13 dB.

![Figure 2-13: Achievable sum-rate throughput with Gaussian inputs and mismatch nearest-neighbour decoding](image)
These results hint at that MU-MIMO-OFDM systems equipped with 1-bit DACs and simple linear precoders achieve satisfactory performance despite the severe nonlinearity introduced by the converters. Although our results deal with QPSK modulation, preliminary results reported in [JDC+16b] reveal that higher-order modulation are supported as well. We hasten to add that although this architectural solution appears promising, a thorough characterization of the out-of-band emissions resulting from the use of low-resolution converters, which is not yet available in the literature, is needed to assess its real potential from a deployment perspective.

2.5 Transceiver Schemes for Backhaul

In the previous sections, we have presented multi-antenna transceiver designs and schemes for mm-wave access. Several of the previously presented approaches, however, can also be extended for the backhaul scenario. In the following, we present two further schemes more focused on enabling a mm-wave backhaul, by addressing two relevant aspects for the backhaul, i.e. the large required throughput and the impact of mobility for the backhaul to a moving hotspot.

2.5.1 Massive Multiple Input Massive Multiple Output (MMIMMO) for Short Range LOS Links

Massive multiple input massive multiple output (MMIMMO) communication systems using mm-wave have been identified as one of the most promising technical approaches for 5G in the METIS 2020 project [METISD33]. In [METISD33] a spatial precoding/decoding scheme called Discrete Fourier Transform Spatial Multiplexing with Maximum Ratio Transmission (DFT-SM-MRT) has been introduced [PTR+14]. Applied to very large Uniform Linear Arrays (ULAs) approximatively facing each other, this scheme enables hundreds of bits/s/Hz of spectral efficiency with a very low complexity compared to SVD beamforming. We recall that one can exploit spatial multiplexing even in LOS [MJW05], [SN07], [TMR11]. We also recall that based on [METISD33], when the carrier frequency $f$, the speed of light $c$, the ULA lengths $L$ and the number of antenna elements $N$ verify the following condition:

$$f \sim \frac{cD}{L^2} N$$  \hspace{1cm} (2-2)

with $D >> L$, the MIMO channel has $N$ eigenvalues of same weight.

As part of the mmMAGIC project a new scheme called Block DFT-SM-MRT (B-DFT-SM-MRT) has been presented in [MMMAGICD51] that outperforms DFT-SM-MRT and even with less complexity. It has been assessed for different practical scenarios such as wireless backhaul and self-backhaul solutions for ultra-dense networks, where antennas are embedded in the urban architecture, device-to-device communication solutions, where antennas fit the shape of connected objects (cars, laptops and screens).

However, all these previous studies reported in [METISD33] and [MMMAGICD51] have been performed with very simplified propagation channel and antenna models. In this contribution, we perform the evaluation using more realistic channel models based on ray tracing and measurements in real environments, and using more realistic antenna diagrams.

We recall in Figure 2-14, the main parameters of the MMIMMO communication system. The system consists of two ULAs, separated by a distance $D$, occupying a length $L$, and composed of $N$ antenna elements each. We denote the separation between antenna elements $d = L / N$. The number of data streams being spatially multiplexed is $N_u$. In the case of DFT-SM-MRT, $N_u = N$. In the particular case of B-DFT-SM-MRT, $N_S$ groups of $N_d$ antennas with $N_{cp}$ ‘CP antennas’ are inserted, such that each group consists of $N_d + N_{cp}$ antennas and the number of data streams being spatially multiplexed is $N_u=N_S N_d$. For further details including the method to choose optimum values of $N$, $N_u$, $N_d$, $N_{cp}$ for given $D$ and $L$ values, refer to [MMMAGICD51].
To assess MMIMO techniques with realistic channel models, we have used two ray tracing tools: one for outdoor and one for indoor. Each tool is modelling a real environment and is based on experimental measurements.

**Example of outdoor environment: Bristol City Center**

We chose to model the propagation at 26 GHz on a route inside city centre of Bristol, UK (see Figure 2-15). The employed ray tracer for outdoor identifies the radio wave scatterers using an accurate geometrical database of the physical environment [TN97] [ABA+15]. Note that the same neighbourhood as for the beamforming performance analysis and visualisation (see Section 5.1 in [MMAIC12]) is used. Similar scenario has been adopted in [FRM+17]. As illustrated in Figure 2-15 three examples of links between lamp posts (with antenna arrays along the lamp post) are considered. Point-source 3D ray-tracing is performed from each antenna element of the transmit array to each antenna element of the receive array assuming isotropic elements. The tool provides all the necessary information to compute the channel complex gain for each multipath component (MPC): i.e. the amplitude, phase, time delay, azimuth & elevation AoD and AoA of each multipath component (MPC). The antenna patterns specified in the following are applied per MPC, to yield the MMIMO channel matrix.

**Example of indoor environment: Helsinki Airport**

We chose to model the propagation in the Helsinki airport check-in hall (illustrated in Figure 2-16) at 26 GHz. The employed tool uses an accurate geometrical database of the physical environment [ABA+15] and [FRM+17]. As illustrated in Figure 2-16 three examples of links between lamp posts (with antenna arrays along the lamp post) are considered. Point-source 3D ray-tracing is performed from each antenna element of the transmit array to each antenna element of the receive array assuming isotropic elements. The tool provides all the necessary information to compute the channel complex gain for each multipath component (MPC): i.e. the amplitude, phase, time delay, azimuth & elevation AoD and AoA of each multipath component (MPC). The antenna patterns specified in the following are applied per MPC, to yield the MMIMO channel matrix.
environment, so called a point cloud [Jar16]. The point cloud model is an accurate model of the environment. It includes small objects (e.g. self-check-in machines) which may scatter energy at high carrier frequencies. Our simulator is optimized for the production of realistic channels thanks to comparisons with measurements in the Helsinki airport check-in hall [VJN+16]. Our simulator accurately identifies the locations and the reflection (or scattering) coefficients of scatterers. The information on the scatterers along with the radiation patterns specified in the following allow us to derive MMIMO channels. Different links illustrated in Figure 2-16 are considered: between nearby devices; between signboards; between self-check-in-machines and finally between a signboard and a canopy.

![Antenna Models](image)

**Antenna Models**

Two antenna patterns have been generated by simulation at 26 GHz as illustrated in Figure 2-17: a ‘basic antenna’ (a classical printed dipole on a ground plane described in section 4.2.3) and a “directional antenna” which consists of 5 units of basic antenna. These 5 unit cells are separated by 1.5 wavelength (at 26 GHz).
**Performance Metrics**

The following performance metrics are computed using the same methodology as in [MMMAGICD51] (with updated channel and antenna models instead of the simplified model):

- SE: the spectral efficiency of the considered scheme;
- “Ratio over SVD”: the ratio of the spectral efficiency of the considered scheme over the spectral efficiency of the SVD spatial multiplexing scheme;
- “Ratio over LOS”: the ratio of the spectral efficiency of the considered scheme over the spectral efficiency that would be attained with the same scheme in a pure LOS environment.

Regarding the complexity, we compute the following metrics:

- $\rho_{CO, TX}^{DFT}$ is the ratio of the complexity scaling law of SVD over the complexity scaling law of the DFT-SM-MRT transceiver;
- $\rho_{CO, RX}^{DFT}$ is the ratio of the complexity scaling law of SVD over the complexity scaling law of the DFT-SM-MRT receiver;
- $\rho_{CO, TX}^{BDFT}$ is the ratio of the complexity scaling law of SVD over the complexity scaling law of the B-DFT-SM-MRT transceiver;
- $\rho_{CO, RX}^{BDFT}$ is the ratio of the complexity scaling law of SVD over the complexity scaling law of the B-DFT-SM-MRT receiver.

The larger are these metrics, the better they are. To compute these ratios, we assume a fully digital architecture and we base our complexity evaluation on [GV96]. We use the fact that the complexities of the DFT of size $N$, the SVD of a matrix of size $N \times N$ and the multiplication of two matrices of sizes $N \times M$ and $M \times P$, scale with $O(N \log_2 N)$, $O(N^3)$, and $O(NMP)$. As a consequence, $N_S$ DFTs of complexities scaling with $O((N_U/N_S) \log_2 (N_U/N_S))$ each, result in a total complexity scaling with $O(N_U \log_2 (N_U/N_S))$. With these assumptions and based on the descriptions of the considered schemes given in [MMMAGICD51], we can derive the following expressions of the considered metrics:

\[
\begin{align*}
\rho_{CO, TX}^{DFT} & \sim \frac{N_U^2 + N_U}{\log_2 (N_U/N_S)} \quad (2-3) \\
\rho_{CO, RX}^{DFT} & \sim \frac{N_U^2 + N_U}{\log_2 (N_U/N_S)} \quad (2-4) \\
\rho_{CO, TX}^{BDFT} & \sim \frac{N_U^2 + 2N_U}{\log_2 (N_U/N_S)} \quad (2-5) \\
\rho_{CO, RX}^{BDFT} & \sim \frac{N_U^3 + 2N_U^2}{(N_U + N_S N_{CP})^2 + N_U \log_2 (N_U/N_S)} \quad (2-6)
\end{align*}
\]

**Simulation results**

Table 2-2 lists the simulated scenarios and their corresponding parameters. We compare the performance of the B-DFT-SM-MRT, with the one of DFT-SM-MRT for all the deployment scenarios described before. As already explained, the parameters $N$, $N_d$, $N_s$, $N_{cp}$ and $d$, are optimised according to a method described in [MMMAGICD51]. This method finds $N_d$ and $N$ values being powers of 2 that verify, as much as possible, the equation (2-2).
## Table 2-2: Simulation parameters and parameters for 26 GHz

<table>
<thead>
<tr>
<th>N°</th>
<th>Deployment</th>
<th>$D$ (m)</th>
<th>Scheme</th>
<th>$N$</th>
<th>$L$ (m)</th>
<th>$N_s$</th>
<th>$N_d$</th>
<th>CP</th>
<th>$d$ (mm)</th>
</tr>
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<tbody>
<tr>
<td>0</td>
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<tr>
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<td></td>
<td></td>
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<td></td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Note: * denotes CP parameter.
Table 2-3 provides the result of the evaluation of the performance and complexity metrics listed before. For each deployment, the scenario reaching the best performance is highlighted in bold. In this set of simulations, the basic antenna described previously (see Figure 2-17) has been used.

Except for the deployment scenario 7 (i.e. between the signboard and the canopy), we obtain the same conclusions as in [MMMMAGICD51]:

1. The performance obtained in the realistic environment is close to the one that would be obtained in a pure LOS environment;
2. The performance is quite close to the SVD performance (from 31 to 84% of SVD performance), with a complexity which is considerably much lower (from 24 to 5504 times less complex at the receiver side and from 7 to 123 times less complex at the transmitter side).
3. B-DFT-SM-MRT outperforms DFT-SM-MRT (at high N only) in terms of spectral efficiency, with an even lower complexity (i.e. with a higher value of the complexity ratio metric).

Note that both DFT-SM-MRT and B-DFT-SM-MRT also attain good performance for lower order MIMO (with N= 4 to 16). However for such small N values, the complexity laws do not apply.

Table 2-3: Simulation results with the basic antenna model for 26 GHz (NA=Not Applicable because N is too small)

<table>
<thead>
<tr>
<th>Scenario</th>
<th>ULA Scheme</th>
<th>Performance</th>
<th>Complexity Ratio</th>
</tr>
</thead>
<tbody>
<tr>
<td>N°</td>
<td>Deploy.</td>
<td>D (m)</td>
<td>Scheme</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>25</td>
<td>DFT-SM-MRT</td>
</tr>
<tr>
<td>0*</td>
<td></td>
<td></td>
<td>B-DFT-SM-MRT</td>
</tr>
<tr>
<td>1A</td>
<td>1</td>
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<td>DFT-SM-MRT</td>
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<td>1A*</td>
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<td>B-DFT-SM-MRT</td>
</tr>
<tr>
<td>2B</td>
<td>2</td>
<td>25.97</td>
<td>DFT-SM-MRT</td>
</tr>
<tr>
<td>2B*</td>
<td></td>
<td></td>
<td>B-DFT-SM-MRT</td>
</tr>
<tr>
<td>3A</td>
<td>3</td>
<td>0.5</td>
<td>DFT-SM-MRT</td>
</tr>
<tr>
<td>3B</td>
<td></td>
<td></td>
<td>DFT-SM-MRT</td>
</tr>
</tbody>
</table>
When analyzing the propagation of the deployment scenario 7, we observe that the roof of the canopy creates a strong dominant scatterer. As the parameters of the MIMO system have been optimised assuming a LOS environment, we therefore propose to replace the basic antenna by the directional antenna described earlier (see Figure 2-17). Table 2-4 provides the simulation results for both the basic and the directional antennas. Thanks to the directional antenna, the effective propagation channel becomes closer to a LOS channel, and we therefore observe a strong improvement.
Table 2-4: Comparison between basic and directional antenna model for 26 GHz

<table>
<thead>
<tr>
<th>Scenario</th>
<th>N°</th>
<th>Antenna</th>
<th>N</th>
<th>Performance</th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>SE (bits/s/Hz)</td>
<td>Ratio over LOS (%)</td>
<td>Ratio over SVD (%)</td>
</tr>
<tr>
<td>7</td>
<td>7</td>
<td>Basic</td>
<td>256</td>
<td>113</td>
<td>9</td>
<td>6</td>
</tr>
<tr>
<td>7A</td>
<td>7A</td>
<td>Directional</td>
<td></td>
<td>1174</td>
<td>94</td>
<td>57</td>
</tr>
<tr>
<td>7*</td>
<td>7</td>
<td>Basic</td>
<td>281</td>
<td>281</td>
<td>16</td>
<td>14</td>
</tr>
<tr>
<td>7A*</td>
<td>7A</td>
<td>Directional</td>
<td></td>
<td>1651</td>
<td>93</td>
<td>81</td>
</tr>
<tr>
<td>7**</td>
<td>7</td>
<td>Basic</td>
<td>264</td>
<td>266</td>
<td>15</td>
<td>13</td>
</tr>
<tr>
<td>7A**</td>
<td>7A</td>
<td>Directional</td>
<td></td>
<td>1681</td>
<td>90</td>
<td>82</td>
</tr>
</tbody>
</table>

**Concluding Remarks**

These new simulations of DFT-SM-MRT (from METIS project) and B-DFT-SM-MRT (from mmMAGIC project) confirmed that there is a potential opportunity to provide very high data rate for wireless backhaul in dense urban areas, using short range LOS communications and MMIMMO techniques. These simulations have been performed with advanced channel models of Helsinki Airport check-in hall (as an example of indoor environment) and a route in the city center of Bristol (as an example of outdoor environment), based on ray tracing and more realistic antenna diagrams, at 26 GHz. These results confirm the conclusions from [MMMAGICD51]:

- at very large numbers of antennas, B-DFT-SM-MRT scheme outperforms DFT-SM-MRT and is even less complex;
- both schemes attain hundreds to thousands of bits/s/Hz of spectral efficiency and have a performance which is close to SVD performance (from 31% to 84% of SVD capacity);
- the complexities of both schemes is much lower than the one of SVD (up to around 10 000 and 100 times less complex for the receiver and the transmitter, respectively).

In addition, these new simulations show that even for lower order MIMO (from 4 to 16 antennas) and shorter antenna arrays, DFT-SM-MRT and B-DFT-SM-MRT exhibit a performance which is close to SVD. However, there are still some remaining questions regarding the impact of hardware impairments, imperfect channel knowledge, interference, finite signal to noise ratio and the limitations linked to a practical deployment of long and thin ULAs on objects and architectural elements. Finally, a complete validation of MMIMMO concept would require a full hardware proof-of-concept.

Therefore, we think that MMIMMMO should more be considered for further research of 5G, and that DFT-SM-MRT and B-DFT-SM-MRT applied to lower order MIMO could potentially be interesting for short term 5G.

**2.5.2 Impact of Beam Misalignment for Moving Hotspot Scenario**

In order to provide immersive early 5G experience in targeted hotspots, a small percentage of 5G users may need data rates up to 20 Gbps to support UHD video and 3D virtual reality services [QHN+16], which can be enabled with mm-wave bands. In the moving hotspot use case, we focus on high speed trains and moving vehicles (cars, buses), where passengers expect to have high data rates and high quality of experience all the time, regardless of the vehicle’s speed. While coverage within the small vehicle area (even with higher mm-wave frequencies) does not present many problems, the provision of backhaul connectivity and capacity to the moving vehicles presents major challenges. From the KPI's point of view, the challenge is to achieve the required capacity for large range of vehicle speeds and different
environments where vehicles are moving (highway, rural, suburban, and urban) and the dynamic backhaul setup and fast cells handovers are especially important.

The general layout considered for the wireless backhauling will be the moving cell to be connected to the 5G small cells, which could be positioned in the street furniture (like lamp posts). These mm-wave Small Cells (mmSCs) could dynamically steer the beam direction pointing at the moving hotspots. Mobility introduces additional complications for beam alignment, as the vehicular locations cannot be precisely tracked due to speed variance. In developing analytical models, we look at the achievable data rates as a function of the beam width and investigate how the beam width may impact on the beam misalignment probability under certain speed variance.

If the beam direction of the fixed mmSC cannot be well aligned to the moving hotspot, the backhaul performance can be significantly deteriorated. In this regard, we will derive the beam misalignment probability and show the effects of prediction accuracy on the system performance. As depicted in Figure 2-18, we assume that the average velocity of the vehicle is \( x \) and the variation of the velocity can be modelled as a random variable \( \Delta x \) following Gaussian distribution with zero mean and variance \( \sigma^2 \), denoted as speed variance.

![Figure 2-18: Beam misalignment due to \( \Delta x \)](image)

Because of \( \Delta x \), the prediction of the location for the moving hotspot may not be always accurate since the beam direction of the fixed mmSC is based on average velocity \( x \). Therefore the fixed mmSC beam may not always point at the moving hotspot. As indicated by the simplified antenna model and with \( \varphi_{ml} \) as the main lobe width of the beam, the beamforming gain decreases significantly if \( \varphi > \varphi_{ml}/2 \), i.e., the beam direction of the fixed mmSC is not well aligned to the direction of the moving hotspot. The probability that such thing happens is defined as misalignment probability and derived as follows. If \( \Delta x \geq 0 \), \( \varphi \) can be expressed as

\[
\varphi = \arctan \left( \frac{d_e - xt}{h_f} \right) - \arctan \left( \frac{d_e - (x + \Delta x)t}{h_f} \right) = \arctan \left( \frac{d_e - xt}{h_f} \right) - \arctan \left( \frac{d_e - (x + \Delta x)t}{h_f} \right),
\]

(2-7)

where \( t \) is time, \( d_r \) is the total length of the route and \( h_f \) is the height of the fixed mmSC. We assume for simplicity that as long as \( \varphi \leq \varphi_{ml}/2 \), the fixed mmSC beam is considered to be well aligned. Then the misalignment probability \( P_{mis} \) can be expressed as

\[
P_{mis} = P \left( \arctan \left( \frac{d_e - xt}{h_f} \right) - \arctan \left( \frac{d_e - (x + \Delta x)t}{h_f} \right) > \frac{\varphi_{ml}}{2} \right),
\]

(2-8)

where \( P() \) stands for probability. Then \( P_{mis} \) can be expressed as,
Since the error function is a monotonically increasing function, it can be seen that the misalignment probability increases with speed variance. It also implies that the transmission rate will mostly decrease with a larger speed variance. Another observation is that $P_{\text{mis}}$ decreases with increased beam width. However, the transmission rate may not necessarily increase with a wider beam because wider beam also means smaller beamforming gain.

We also evaluate the moving hotspot use case. For simplicity, we assume that the moving hotspot is able to obtain its own instantaneous location information and the fixed mmSC’s location so that its beam direction can always point at the fixed mmSC. However, the direction of the fixed mmSC beam needs to be steered to track the hotspot movement.

In the following, we evaluate the impact of inaccurate beam alignment due to speed variation. The simulation parameters are listed in Table 2-5. Figure 2-19 illustrates the misalignment probability $P_{\text{mis}}$ with increased speed variance with beam width 11.25 degree. As can be seen, there is a good match between analytical results and the Monte-Carlo simulation results. Basically $P_{\text{mis}}$ increases with speed variance. This is because a constant speed is assumed for the moving hotspot at the fixed mmSC and the location of the moving hotspot is predicted based on this assumption. In this regard, a larger speed variance will apparently introduce more location prediction error and cause worse beam misalignment. However, when the distance between the moving hotspot and the fixed mmSC is large, the impact of speed variance is smaller. It implies that with highly densified fixed mmSCs, prediction error due to speed variance of the moving hotspot might be a critical issue.

### Table 2-5: Simulation parameters

<table>
<thead>
<tr>
<th>Simulation Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Carrier Frequency</td>
<td>28 GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>1 GHz</td>
</tr>
<tr>
<td>$d_r$</td>
<td>50, 100 m</td>
</tr>
<tr>
<td>$h_f$</td>
<td>10 m</td>
</tr>
<tr>
<td>fixed mmSC transmission power</td>
<td>23 dBm</td>
</tr>
<tr>
<td>Noise temperature</td>
<td>300 K</td>
</tr>
<tr>
<td>Moving hotspot speed</td>
<td>20 m/s</td>
</tr>
<tr>
<td>Pathloss Model</td>
<td>LOS path loss models [RHD+15]</td>
</tr>
</tbody>
</table>
We also evaluate the impact of beam width on the misalignment probability in Figure 2-20. A good match is also observed between the analytical and the simulation results. As predicted the misalignment probability decreases with beam width. In an extreme case where the beam width is 360 degree, i.e. omnidirectional, the misalignment probability is 0. However, it should be noted that in such a scenario, a small misalignment probability does not necessarily lead to a higher transmission rate since the beamforming gain of a wider beam also gets smaller, leading to lower transmission rate. Looking at possible implementation values of 20 degree beam width and a practical 50 (m/s)² speed variance, the misalignment probability is around 10%, which does not impact on the data rates (analysed below) significantly.

Now we investigate the effects of beam width and speed variance on the data rates separately. We assume that the fixed mmSC can always track the movement of the hotspot accurately and the beam widths of the fixed mmSC and the moving hotspot are the same. Figure 2-21 shows the 95% available rate with different beam widths for such a case. It decreases with wider beam due to the reduced beamforming gain. When the beam width is 2π, i.e., omnidirectional, the available rate can be very low, e.g., around 1 Gbps. The results re-affirm the previous discussions.
2.6 Transceiver Design for Relaying

The final scenario for which we consider multi-antenna transceiver schemes for mm-wave communication is relaying in a multiuser system. Due to the limited penetration and diffraction capacity of mm-wave signals, the LoS link between the BS and user may easily be blocked by humans in motion or stationary obstacles, causing transmission data rate degradation. Once the blockage happens, a proper relay can be selected to transmit signal from the BS to the blocked user. In the following, we extend the work presented in [MMMAGICD51, Section 3.5] and propose a dual-hop relay selection algorithm for mm-wave multi-user systems.

2.6.1 Relay Selection in mm-wave Multiuser Systems

Consider the mm-wave multi-user system shown in Figure 2-22. The BS equipped with $N_{BS}$ antennas and $N_{RF}$ ($1 < N_{RF} \ll N_{BS}$) RF chains is assumed to communicate with $K$ ($K \geq 1$) users simultaneously via spatial multiplexing. Each user is equipped with $N_{MS}$ antennas and a single RF chain. On the downlink transmission, the BS exploits hybrid beamforming, transmitting $K$ independent data streams, one for every user, in each time slot. The users apply analog only beamforming to receive signals over certain directions. Note that the maximum number of simultaneously served users is limited by the number of BS RF chains, i.e., $K \leq N_{RF}$. For simplicity, we assume that the BS will use all of its $N_{RF}$ RF chains to serve as many users as possible, such that: $K = N_{RF}$.

In the beginning of each time slot, the BS establishes $K$ concurrent transmission links, as shown in Figure 2-22(a). The spatially multiplexed users should be deliberately selected such that the co-channel interference is minimised [YSD+16]. The analog beamformer and digital precoder at the BS, i.e., $P_{RF} = [p_{RF}^1, \ldots, p_{RF}^K]$ and $P_{BB} = [p_{BB}^1, \ldots, p_{BB}^K]$, and the analog combiner at the users, i.e., $\mathbf{c}_k$ for $k = 1, \ldots, K$, are designed to maximise the system sum rate using the two-stage hybrid precoding algorithm, as detailed in [YSD+16, ALH+15]. The achievable spectral efficiency of the $k$-th user is then expressed as:

$$R_k = \log_2(1 + \frac{|c_k H_k p_{RF} p_{BB}|^2}{\sum_{i=1, i \neq k}^K |c_i H_i p_{RF} p_{BB}|^2 + \frac{P}{\sigma^2}})$$

(2-10)

where $H_k$ represents the $N_{MS} \times N_{BS}$ channel matrix of the $k$-th user and $\frac{P}{\sigma^2}$ is the transmit SNR.

When blockage is detected on any of the transmission links, e.g., an active user experiences a sudden decrease in spectral efficiency, the following procedures should be performed to select the optimal relay out of the idle users which maximises the effective spectral efficiency on the relay path while minimising the interference to other unblocked transmissions:
- Once the blockage is detected, the blocked user $b$, e.g., user 2 in Figure 2-22(b), broadcasts 'help' messages to users in its neighbourhood on the control channel, e.g., a channel operating in lower frequency;
- The users that stay in idle, e.g., user 3 to 7 in Figure 2-22(c), upon receiving the 'help' message, measure the link strength from themselves to the blocked user, and then send feedback to the BS via the control channel to inform their availability and the quantized link strength $G_{rb}$;
- The BS collects the feedback from the feasible relay candidates, e.g. user 3, 4 and 7 in Figure 2-22(d), which have LoS links to both the BS and the blocked user, and then evaluates the link strength from itself to these candidates, i.e., $G_{BSr}$, to choose the optimal relay that maximises the selection metric, e.g., user 7 in Figure 2-22(d).
- In the first hop, the BS transmits data streams to the selected relay, i.e., user $r$, and other spatially multiplexed users, i.e., user $k$ for all $k \neq b$, in one time slot, e.g., user 7 and user 1 in Figure 2-22(f);
- In the second hop, when the relay finishes reception, it transmits the received data to the blocked user $b$ in another time slot, as shown in Figure 2-22(f).

Figure 2-22: An example of the relay selection process: (a) initial transmission scheduling, (b) blockage detection, (c) broadcasting help message, (d) relay candidate identification (e) relay selection metric and (f) transmission via the relay path.
For the mm-wave multiuser system shown in Figure 2-22, the relay selection metric should be well designed to balance between the spectral efficiency and co-channel interference. It is important to note that assuming two equal timeslots are allocated, the effective spectral efficiency of the relay path $R_E$ is restricted by the minimum between those of the two hops, i.e., $R_E = \min(R^1_r, R^2_r)$. Given that in the first hop the BS replaces the blocked user with the chosen relay for data transmission, the achievable rate at the active user $k$ $R_k$ and the relay $R_r$ are therefore defined using Eq. (2.10). It is assumed that the transmission in the second hop is deliberately scheduled such that little interference is caused. The spectral efficiency is then, $R^2_r = \log_2(1 + |c_r^H H b c_b|^2 / (\sigma^2 P))$, where $H_b$ denotes the $N_{MS} \times N_{MS}$ channel matrix, and $c_r$ and $c_b$ represent the analog beamforming vectors applied at the relay and the blocked user respectively. In addition, it has been shown in [YAD+16] that the interference level among the concurrent transmissions can be effectively measured with the angular distance between the relay and the unblocked links, i.e., $\xi(r,k) = \sin(\alpha_{rk})$. We therefore define the selection metric as: $G_{BSr} G_{rb} \prod_{k \neq b} \xi(r,k)$, as illustrated in Figure 2-22(f). The user $r$ (in idle) that maximises the product of the joint link strength on the relay path $g_{BSr} g_{rb}$ and the angular distances $\xi(r,k)$ to the unblocked users will be selected as the optimal relay.

The performance of the proposed relay selection algorithm is evaluated via numerical simulations in this section. We consider the mm-wave multiuser system shown in Figure 2-22. User channels are assumed to have LoS paths to the BS and generated using the QuaDriGa channel model [MMMAGICD21]. The detailed simulation parameters are listed in Table 2-6.

### Table 2-6: Simulation parameters

<table>
<thead>
<tr>
<th>Simulation parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency [GHz]</td>
<td>28</td>
</tr>
<tr>
<td>Cell radius [m]</td>
<td>10</td>
</tr>
<tr>
<td>Number of total users</td>
<td>20</td>
</tr>
<tr>
<td>Number of BS antennas $N_{BS}$</td>
<td>32</td>
</tr>
<tr>
<td>Number of BS RF chains $N_{RF}$</td>
<td>2</td>
</tr>
<tr>
<td>Number of user antennas $N_{MS}$</td>
<td>8</td>
</tr>
<tr>
<td>Number of static blockers</td>
<td>10</td>
</tr>
<tr>
<td>BS/Blocker height [m]</td>
<td>5</td>
</tr>
<tr>
<td>User height [m]</td>
<td>1.5</td>
</tr>
<tr>
<td>User/Blocker radius [m]</td>
<td>5</td>
</tr>
<tr>
<td>Number of cellular realizations</td>
<td>100</td>
</tr>
<tr>
<td>Number of blockage realizations</td>
<td>100</td>
</tr>
<tr>
<td>Transmit SNR [dB]</td>
<td>35</td>
</tr>
</tbody>
</table>

Performance comparisons are carried out between random relay selection and the proposed algorithm. Figure 2-23(a) shows the cumulative distribution function (CDF) of the loss on spectral efficiency experienced at the blocked user $b$, as the BS switches data transmission from the direct link to the relay path, i.e., $L = ((R_b - R_E) / R_b) \times 100\%$, where $R_b$ represents the achievable rate without blockage, as defined in Eq. (2.10). Using the proposed relay selection algorithm, the blocked user will suffer a loss less than 30\% at a rate of 70\%, which then decreases to 40.7\% for a loss less than 10\%. As to its random counterpart, significant performance degradations can be observed. The corresponding rates are 27.9\% for 30\% loss.
and only 10.5% for 10% loss. This is thanks to the properly designed relay selection metric that accounts for link strength of both the two hops forming the relay path, based on which only the candidate that optimizes the trade-off between co-channel interference and effective spectral efficiency will be selected as the relay. In addition, it is noted in Figure 2-23(a) that transmitting data via the relay path can even improve the spectral efficiency, providing a loss with minus value at a rate of 25.3%. This is especially the case for users far from the BS. Given that mm-wave signals attenuate significantly over distance, deliberately breaking a long-hop into multiple short ones can improve the network throughput, as reported in [LSW+09, QCS+11] and Section 3.7.

Moreover, for the considered multiuser systems, the BS always establishes concurrent transmission links to serve multiple users at the same time, which causes co-channel interference. Once the direct link between the BS to the blocked user is replaced by the relay path, the performance of other spatial multiplexed users, i.e., user, \( k \neq b \), may be affected accordingly, given that the first hop of the relay path may change the interference level. Figure 2-23(b) shows the CDF of the spectral efficiency loss at the unblocked users. Owing to the highly directional links exploited in mm-wave systems, the unblocked user is far less sensitive
to relay selection schemes as compared with its blocked counterpart. It suffers a loss less than 3.1% using the proposed relay selection algorithm at the rate of 98.0%, while a loss of 52.1% using random relay selection. However, a poorly designed relay path can still largely interfere with other concurrent transmissions, causing a spectral efficiency loss as high as 97.8% at the unblocked users.

2.7 Mm-wave CSI Aspects: Beam Design

As discussed before, the large size of the antenna arrays envisioned for mm-wave systems imposes challenges for CSI acquisition. This is mainly due to the architectural constraints and the need of directional transmission to overcome the propagation effects at higher frequencies. To this end, closed-loop beam training protocols have been proposed (e.g. [IEEE802153c] [IEEESIT]) involving either an exhaustive search over all possible codebook beams or a hierarchical search, where the beams are narrowed down at each stage. In any case, beam synthesis is a relevant aspect to consider for the CSI acquisition, which will be discussed in the following.

The required directional transmission also impacts the design of broadcast channels, which are performed with an omnidirectional transmission in conventional systems, e.g. at lower frequencies. Hence, we also discuss the design of wide beams for control and broadcast signalling.

2.7.1 Arbitrary Beam Synthesis for Hybrid Beamforming Systems

For future mm-wave mobile communication systems, the use of analog/hybrid beamforming is envisioned to be an important aspect. The synthesis of beams is a key technology to enable the best possible operation during beam search, data transmission and MU-MIMO operation. In this subsection, we present a method for synthesizing beams based on previous work in radar technology considering phased array antennas. With the proposed technique, it is possible to generate a desired beam of any shape with the constraints of the desired target transceiver antenna frontend. It is not constrained to a certain antenna array geometry, and can handle 1D, 2D and even 3D antenna array geometries like cylindrical arrays. The numerical examples show that the method can synthesize beams by considering a user defined tradeoff between gain, transition width and passband ripples. Based on requirements on the beam shape, this work formulates an optimization problem similar to [S16] and [MMR+12]. Afterwards the optimization problem is solved numerically. This work includes the specific constraints of hybrid beamforming and low resolution phase shifters. The superscripts s and f are used to distinguish between subarray and fully-connected hybrid beamforming, which were discussed in Section 2.3.

In the following we develop a strategy to synthesize arbitrary beams based on the formulation of an optimization problem. Furthermore, we show how different constraints can be used to model the restrictions of different systems.

The array factor $A(u, a)$ of an antenna array is defined as

$$A(u, a) = a^T p(u), [p(u)]_n = e^{j\frac{2\pi}{\lambda} x_n(u)},$$

where $a$ is the beamforming vector, $u$ is the spatial direction combining the azimuth and elevation angle. The scalar $x_n(u)$ is the distance to the plane defined by the normal vector $u$ and a reference point. A common choice for the reference point is the position of the first antenna, in this case $x_1(u) = 0$.

The objective of synthesizing an arbitrary beam pattern can be formulated as a weighted $L^p$ norm between the desired pattern $D(u)$ and the absolute value of the actual array factor $|A(u, a)|$

$$f(a) = \left( \int W^p(u) |A(u, a)| - D(u) |^p du \right)^{\frac{1}{p}},$$

(2-12)
where \( W(u) \) is the weighting. This objective function itself is convex over its domain, but the constraints on \( a \) shown in the following subsections lead to a non-convex optimization problem. This problem formulation ignores the phase of the array factor, since we require only the magnitude to be of a specific shape. By only optimizing over the array factor we do not take the pattern of the antennas into account. As described in [S16], to account for an antenna pattern it is only necessary to divide \( D(u) \) and \( W(u) \) by the pattern of the antenna elements.

The sub-array and the fully connected hybrid beamforming system have different constraints on the beamforming vector \( a \), since as indicated in Section 2.2, with subarrays the RF chains are not connected to all antennas. For the case without phase quantization a digital phase shift would be redundant with the ones in the RF domain and are therefore not considered in this case. The number of antennas is denoted by \( M \), the number of RF-chains is \( M_{RFE} \) and the number of antennas per sub-array is \( M_{c} \). The beamforming vectors for sub-array hybrid beamforming combine the beamforming vectors \( w_{i}^{s} \) of each sub-array in the following form

\[
a = W^{s} \alpha^{s} = \begin{bmatrix}
w_{1}^{s} & 0 & \cdots & 0 \\
0 & w_{2}^{s} & \ddots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & w_{M_{RFE}}^{s}
\end{bmatrix}, \tag{2-13}
\]

where \( w_{i}^{s} \) is defined as

\[
w_{i}^{s} = [e^{j \theta_{i,1}^{f}} \ e^{j \theta_{i,2}^{f}} \ \cdots \ e^{j \theta_{i,M_{RFE}}^{f}}]^{T}. \tag{2-14}\]

For the fully connected case the beamforming vector is constraint to be of the form

\[
a = W^{f} \alpha^{f} = [w_{1}^{f} \ w_{2}^{f} \ \cdots \ w_{M_{RFE}}^{f}] = \begin{bmatrix}
e^{j \theta_{1,1}^{f}} & e^{j \theta_{1,2}^{f}} & \cdots & e^{j \theta_{1,M_{RFE}}^{f}} \\
e^{j \theta_{2,1}^{f}} & e^{j \theta_{2,2}^{f}} & \ddots & \vdots \\
\vdots & \vdots & \ddots & \vdots \\
e^{j \theta_{M_{RFE},1}^{f}} & e^{j \theta_{M_{RFE},2}^{f}} & \cdots & e^{j \theta_{M_{RFE},M_{RFE}}^{f}}
\end{bmatrix} \begin{bmatrix}
\alpha_{1}^{f} \\
\alpha_{2}^{f} \\
\vdots \\
\alpha_{M_{RFE}}^{f}
\end{bmatrix}. \tag{2-15}\]

In addition to the constraints stemming from hybrid beamforming it is always necessary to introduce power constraints. Here it is possible consider an individual power constraint

\[
|\alpha|_{m} \leq 1 \ \forall m = \{1, 2, \cdots, M\}, \tag{2-16}\]

or a sum power constraint

\[
\|\alpha\| \leq 1. \tag{2-17}\]

It is also possible that the resolution of the phase shifters is limited. This means that the values of \( \theta_{i,j}^{f} \) are from a finite set of possibilities

\[
\theta_{i,j}^{f} = -\pi + k_{i,j} \frac{2\pi}{K} \ \forall i, j \text{ and } k \in \{0, 1, \cdots, K - 1\}, \tag{2-18}\]

where \( K \) is the number of possible phases. Due to the finite resolution of the RF phase shifters, a digital baseband phase shift needs to be taken into account. Therefore, in addition to the scaling \( \alpha^{s} \) and \( \alpha^{f} \), we need to take a phase shift \( \xi^{s} \) and \( \xi^{f} \) into account. For the case of sub-array hybrid beamforming with limited resolution RF phase shifters the beamforming vector \( a \) takes the form

\[
a = W^{s} (\alpha^{s} \circ \xi^{s}), \tag{2-19}\]

where \( \circ \) is the Hadamard product (element-wise multiplication) and \( \xi^{s} \) are the digital phase shifts defined as

\[
\xi^{s} = [e^{j \xi_{1}^{s}} \ e^{j \xi_{2}^{s}} \ \cdots \ e^{j \xi_{K_{RFE}}^{s}}]^{T}. \tag{2-20}\]
The formulation for the fully-connected case does also contain additional phase shifts of the baseband signals in a similar fashion. It is important to keep in mind that with the resolution constraints the problem is converted to a mixed integer non-linear programming (MINLP) problem.

Combining the objective function with the constraints associated with the hardware capabilities lead to the following optimization problem

$$\min_a f(a) \ s.t. \ g(a) \leq 0, h(a) = 0$$ \hspace{1cm} (2-21)

where $g(a)$ and $h(a)$ are the constraints modelling the desired hardware capabilities. The weighting $W(u)$, the desired pattern $D(u)$ and the choice of $p$ in $f(a)$, determine which point in the trade-off gain, passband ripple and transition width is going to be targeted.

To compare the designed beams we need to first define some measure of the difference of the designed beams. Some of the measures are similar to the ones defined in [DBG+16]. The first metric is the average gain in the desired directions. Directly connected to the average gain is the maximum ripple of the array factor in the desired directions. To not distort the result, the transition region is excluded from the search of the maximum ripple. A very important criterion to evaluate the performance of a beam for initial access is the overlap of adjacent beams of the same width. Here we evaluate the area at which the distance between two beams is less than 5 dB relative to the total area of one beam. The last measure is the maximum side-lobe relative to the average gain in the desired directions.

In the following section, beams synthesized by the described method are shown. For all systems, the transmitter is equipped with 4 RF-chains ($M_{\text{RFE}} = 4$), connected to 64 Antenna elements, forming an ULA with half-wavelength inter-element spacing. Since the antenna array is one dimensional, it is sufficient to look at only one spatial direction. The following plots are referring to angle $\psi = \lambda/2 \sin \varphi$, where $\varphi$ is the geometric angle between a line connecting all antennas and the direction of a planar wavefront.

For an ULA, the spatial direction $u$ is fully represented by $\psi$, therefore $W(u)$, $D(u)$ and $A(u, a)$ depend only on $\psi$. Since the magnitude of each element of is less or equal to one, if a perfect flat beam without side lobes could be constructed, it would have the array factor $D_{\text{max}} = \sqrt{N \beta D_{\text{max}}}$. As also described in [S16] such a beam cannot be realized, therefore $D(\psi)$ is equal to $\beta D_{\text{max}}$ at the desired directions and equal to zero, elsewhere.

For all systems we set $p = 4$ in the objective function. This ensures equal gain and side lobe ripples. The integral of the objective function over all spatial directions in the objective function is approximated by a finite sum. To ensure a sufficient approximation, the interval is split into 512 elements. As described in [S16] the computational complexity can be significantly reduced by reformulating the problem to use FFT/IFFTs to calculate $A(\psi, a)$ and the derivatives of the objective function.

For each system, the optimization process was started by considering several initializations. Since the used non-linear programming (NLP) and MINLP solvers only guarantee to find a local minimum for a non-convex problem, the results were compared and the implementation leading towards the minimum objective function was selected.

Results for synthesizing different beams are shown in Figure 2-24 and Figure 2-25. Given the previously defined criteria the performance is quantified in Table 2-7. The beams in Figure 2-26 are designed to show that it is also possible to support MU-MIMO. The beams can simultaneously be used to serve two users in different locations.
Figure 2-24: Sub-array HBF with individual power constraint.

Figure 2-25: Fully-connected HBF with 2-bit phase quantization and sum power constraint.

Table 2-7: Performance of the designed beams.

<table>
<thead>
<tr>
<th>Beam</th>
<th>avg. gain dB</th>
<th>max ripple dB</th>
<th>overlap in %</th>
<th>max sidelobe in dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Figure 2-24(a)</td>
<td>18.2</td>
<td>4.00</td>
<td>2.44</td>
<td>-17.4</td>
</tr>
<tr>
<td>Figure 2-24(b)</td>
<td>21.7</td>
<td>2.89</td>
<td>3.22</td>
<td>-16.2</td>
</tr>
<tr>
<td>Figure 2-24(c)</td>
<td>26.3</td>
<td>2.76</td>
<td>7.21</td>
<td>-16.3</td>
</tr>
<tr>
<td>Figure 2-25(a)</td>
<td>2.52</td>
<td>3.90</td>
<td>7.66</td>
<td>-10.3</td>
</tr>
<tr>
<td>Figure 2-25(b)</td>
<td>5.50</td>
<td>3.01</td>
<td>6.54</td>
<td>-10.1</td>
</tr>
<tr>
<td>Figure 2-25(c)</td>
<td>8.23</td>
<td>1.47</td>
<td>6.63</td>
<td>-12.7</td>
</tr>
</tbody>
</table>
2.7.2 Wide Beams and Broadcasting Signalling

It is commonly assumed that large arrays are only capable of transmitting/receiving in terms of narrow beams, which poses an issue for random access and transmission of cell-specific reference signalling. A wide-beam approach which maximizes the power utilization of the array, which leads to less loss of array gain compared to conventional methods, is defined as follows. The main idea lies in applying a widener function \( f_n(p, c) \) on a base DFT weight vector. For the \( N \)-antenna uniform linear array (ULA), we get the precoding vector

\[
\mathbf{w}_{\text{wide}} = \mathbf{w}_{\text{DFT}} \odot [e^{j f_0(p, c)}, \ldots, e^{j f_{N-1}(p, c)}]^T,
\]

where for \( n = 0, \ldots, N - 1 \). The widener function, [Int16], is given by

\[
f_n(p, c) = 4\pi c \left( \frac{1}{2(N-1)} + \frac{(n-N)^p}{N-1} \right)
\]

in which \( \odot \) denotes the Hadamard product. Parameters \((p, c)\) allows for further optimization of the beam shape given an immediate objective. Simulation of the proposed scheme for a 128x1 ULA, compared with state of the art found in [Int16, QQL16] is shown in Figure 2-27.
Figure 2-27: Example of beam shapes designed using the Widener method described in [Int16] and the technique using amplitude tapering in [QQL16]. Plotted is also the DFT beam and the sub-element pattern.

One advantage of this approach is that it requires phase-only pre-coding, which provides a simpler implementation which may be required at mm-wave frequencies.

2.8 Summary

In this chapter, we have discussed the subarray architecture as an enabler to implement hybrid beamforming with reduced complexity and losses. Despite the reduced design flexibility with subarrays, the performance with such an architecture might actually outperform the fully connected case when accounting for the losses in the analog network of phase shifters.

We have also discussed multi-antenna transceiver schemes for mm-wave communication in the following scenarios: access, backhaul and relaying. We have presented a flexibly hybrid beamforming design for the access in multiuser systems which is able to achieve the same performance as state of the art approaches but with a reduced channel knowledge. The robustness of the proposed scheme against signal blockage has been discussed, since after a blockage the beamformer with hybrid beamforming does not need to be recomputed as with digital beamforming. For the access in dense urban areas, a system level comparison between analog beamforming, hybrid beamforming and digital precoding has been provided, showing that hybrid beamforming matches the performance of digital beamforming under LOS, but suffers a performance gap in case of NLOS. However, when considering outdated CSI we have shown that hybrid beamforming is more robust than digital precoding, since the hybrid beamforming design can rely on large scale parameters. As an alternative for addressing the power consumption bottleneck in mm-wave systems, digital beamforming with 1 bit DACs has also been discussed. Despite the severe non-linearity introduced by the converters, it has been shown that such approach is able to achieve a satisfactory performance.

Multi-antenna transceiver schemes for mm-wave backhaul have been considered, by considering two important aspects related to this scenario: the large required throughput and the impact of mobility for a moving hotspot. For addressing the high spectral efficiency requirement, we discussed the massive multiple input massive multiple output communication system for short LOS links based on a scheme which is able to achieve a significant portion of the capacity with a large reduction in complexity. The proposed technique has been validated with realistic channels for an indoor and an outdoor environment. For the backhaul to a moving hotspot, the impact of inaccurate beam alignment due to speed variation has been
analysed. It has been shown that beam misalignment increases with a denser deployment of small cells, as well as with decreased beam width. Although broader beams reduce the beam misalignment, this comes at the expense of reduced antenna gain and performance.

Regarding transceiver schemes for implementing mm-wave relaying, a dual-hop relay selection algorithm has been discussed to aid a user which has suffered a signal blockage. For this purpose, a relay selection metric has been proposed to provide a balance between the spectral efficiency and the co-channel interference resulting from the transmission to the relay. The results indicate a reduced performance degradation with the proposed relay selection metric.

In the last part of the chapter, we have discussed the beam synthesis as a relevant aspect for CSI acquisition for broadcasting. To this end, beam synthesis has been investigated by taking into account a user-defined trade-off between gain, transition width and passband ripples. In addition, constraints from hybrid beamforming and low resolution phase shifters have been considered. In contrast to the common notion that large antenna arrays are capable of generating only narrow beams, we have also shown how a wide beam can be generated with a large antenna array. The proposed approach is based on a phase-only precoding scheme, and hence eases the implementation at higher frequencies of wide beams for broadcast and control signals.
3 Coordinated Multi-node Scheme Design

3.1 Introduction
In the D5.1 [MMAGICD51] multi-antenna chapter we discussed the need for highly directional transmission in mm-wave systems. For this reason, it is often argued that the interference for mm-wave cellular systems might not be so detrimental as compared to current deployments at lower frequencies. However, whether interference plays a significant role or not, actually depends on the deployment, i.e. on the BS density as well as the capabilities of the antenna array at the users. In fact, due to the limited range of mm-wave communication, a high BS density might be required to achieve an acceptable coverage. For ultra-dense deployments, users might have a line-of-sight to several BSs, eventually leading to higher interference, in particular if we need to accept a less precise hardware-constrained (e.g. due to phase noise or limited DAC/ADC resolution) beamforming approach.

Nevertheless, in contrast to the traditional multi-node cooperation at lower frequencies, due to the sparse multipath and high penetration loss the main goal of multi-node coordination at mm-waves might be to reduce signal outage due to sudden blockages and avoiding intermittent interference, instead of simply providing higher data rates. With peak-power limited mm-wave nodes, there will also be coverage gains due to increased aggregated power using multi-node joint transmission.

Thus, coordinated multi-node schemes might play a crucial role in dense mm-wave network deployment scenarios, in which several cooperating nodes can help to obtain

–Macro-diversity gains towards shadowing/blocking,
–Power gains in cases when the system is peak power limited per node,
–Artificially increase multipath and thus (distributed) MIMO rank to better support (distributed) spatial multiplexing and massive MIMO gains at sparse mm-wave channels,
–Support integrated multi-node/multi-beam soft handover,
–Efficient load balancing for energy efficient operations.

In the following sections, we report the work performed within some of those areas beyond the results reported in D5.1 [MMAGICD51], but we start with a short review of the results in D5.1 [MMAGICD51].

In D5.1 [MMAGICD51] we started by reviewing the state of the art in multi-node cooperation, and we discussed the key challenges and opportunities such as dynamic node and user clustering and the need to co-design with the backhaul network and network architecture. To that end, D3.2 [MMAGICD32] has devoted several sections discussing concepts for integration of 5G mm-wave systems such as Multi-Connectivity, LTE-NR tight interworking, RRC diversity, Multi-band system integration, Cell clustering, Mobility state transition, Self-Backhauling, Access-integrated backhaul in fixed wireless access, Joint optimization of access and backhaul, and Interference coordination. The interested reader are advised to read further in [MMAGICD32].

The initial multi-node results in D5.1 [MMAGICD51] showed that there are substantial gains by combining mm-wave links with FSO links, in particular in combination with joint HARQ. Then, three studies were devoted to coverage analysis. The conclusions of the first analytical study showed that the multiple LOS BSs available in a small cell range allows exploiting the coordination among multiple BSs to improve the coverage for a typical UE. In the second study, based on ray tracing, it was shown that increasing the number of BSs efficiently reduces the outage probability. However, the coverage for narrow streets and for users blocked by buildings or foliage is still a challenge, to which the use of multi-node schemes might be a solution. The third study focused on the uplink coverage with base stations employing partial-zero-forcing. It was shown that the coverage increases with the total number
of antennas, but also that there exists an optimal fraction of the antennas that should be devoted to cancel the nearest few strongest interferers. Then, we shortly introduced the study on Joint hybrid precoding for energy-efficient mm-wave networks that was completed towards this deliverable and is included in Section 3.8. We concluded the contributions in D5.1 with initial system level results on analog beamforming for mm-wave access. In the considered outdoor urban environment and using the pure analog beamforming scheme, very good SINR can be achieved for the UEs very close to the BS (that are mainly in LOS) and acceptable SINR can be also obtained for the UEs at the cell edge. These results are very different from the ones observed at lower frequencies with LTE, in which the system would be mainly interference limited.

In the following sections, we present the final results on technical studies towards cooperative and coordinated multi-node scheme design. We start by a section 3.2 proposing a multi-cell beam sweeping method, then with section 3.3 on sequential hybrid beamforming design for multi-link mm-wave communication, followed by section 3.4 on beam management for mobility, section 3.5 performance of mm-wave-based RF-FSO multi-hop networks, section 3.6 on mm-wave LoS coverage enhancements with coordinated high-rise access points, section 3.7 on direct vs relay-assisted access, and ends with section 3.8 on joint hybrid precoding for energy-efficient mm-wave networks, and then we conclude the chapter.

3.2 Beam Sweeping for Multi-Node Networks

As mentioned before, highly directional transmission and reception is required to alleviate the adverse propagation effects at higher frequencies, to enable mm-wave communication [RRE14]. To this end, beam sweeping at the BS and the UE is needed, allowing to find the best beam pair to setup a reliable link, i.e. the BS finds the best transmit (Tx) beam, whereas the user finds the best receive (Rx) beam in the downlink (DL). With such a procedure, the best beam pair is usually determined at the UE based solely on the signal strength to its own BS, i.e. the best beam pair is that with the best SNR. In the case of a network with multiple BSs, however, the transmission from the BSs in other cells needs to be considered for obtaining the best beam pair. For the UE to properly distinguish the signal in its serving cell, the serving BS and the interfering BSs need to transmit orthogonal downlink pilot sequences as traditionally done in low dimensional multi-antenna systems at lower frequencies. The downlink pilot sequence assigned to each BS is transmitted with each trained transmit beam from the given BS, i.e. it is repeated throughout the beam sweeping procedure (Figure 3-1).

![Figure 3-1: Multi-Node Beam Sweeping (figure modified from original figure in [HKC16])](image-url)
Despite the previous argument, [ANR+17] has recently proposed to reuse DL pilots and avoid orthogonal pilots among neighboring cells for mm-wave cellular networks. This result is basically motivated from the assumption that the effect of intercell interference significantly decreases as the width of the Tx beams is reduced in mm-wave communication. In case of a full pilot reuse among neighboring cells as proposed in [ANR+17], this implies that the decision for finding the best beam pair is based on the total receive power, consisting of the signal to its own BS plus interference from neighboring BSs. Hence, such an approach does not really consider the effect of the interference, which cannot be distinguished anyway with a full pilot reuse. In case the intercell interference can be neglected, the approach proposed in [ANR+17] with full pilot reuse and finding the best beam pair based on the total receive power would be similar to employing orthogonal DL pilots and finding the best beam pair based solely on the SNR. Although the interference might be reduced on average for mm-wave, in some scenarios a UE could have a line of sight (LOS) or a strong multipath to interfering BSs. In fact, as pointed out in [BH15], the probability of having a LOS to interfering BSs increases for denser deployment of BSs, i.e. as cells become smaller. Therefore, even though the average interference is reduced due to the directional transmission in neighboring cells, a UE could still experience strong interference. Thus, in contrast to the previous two approaches discussed before, we stress the need to properly consider the interference in the beam pair selection.

To this end, consider a simple setup with a UE connected to BS 1 and having two interfering BSs as depicted in Figure 3-1, where the BSs employ $N_{BS}=10$ Tx beams and the UE uses $N_{MS}=8$ Rx beams for the beam sweeping. After the multi-cell beam sweeping, the UE communicates its preferred Tx beam to BS 1. To this end, denote the strength of the receive signal by the shades of brown for the possible beam pairs between BS 1 and the UE. Hence, we can see from Figure 3-2 that the SNR-based best beam pair would consist of the Tx beam #3 and the Rx beam #4. Furthermore, in Figure 3-2 we represent the strength of the interfering signals with the different beam pairs with shades of gray. Selecting the best beam pair based solely on the SNR, ignores that the UE could also receive a strong interference with Rx beam #4 if the BS 2 employs Tx beam #9 for the data transmission to its users.

![Figure 3-2: Example of Beam Sweeping Result in a Multi-Node Setup](image)

To address this issue, the UE could communicate to BS 1, which Tx beams should be avoided by neighboring BSs. Assuming the users in the neighboring cells also report to their respective BSs which beams should be avoided by their interfering BSs, this implies the BSs might not be able to serve the users with the preferred Tx beams reported by the users in each cell. In this case, the users would need to communicate another preferred beam taking into account these constraints. Another possibility to address this issue would consist in some coordination between the BSs, allowing the BSs to jointly select the best Tx beams to its users.

Nevertheless, the previous discussed approaches require more feedback from the users as well as results in increased delay in the beam selection procedure. To avoid this, the UE should select the best beam pair by taking into account the possible interference from neighboring BSs. The aim is that the UE should make this decision locally without involving its own BS or for that matter the interfering BSs. Since the UE does not know beforehand which
Tx beams are going to be employed by the interfering BSs, one approach to take the interference into account is by considering the average interference on each Rx beam, i.e. the average interference with the different Tx beams from the interfering BSs when employing a given Rx beam at the UE. To this end, denote the receive signal for the UE as \( S_{ij} \) for the \( j \)-th Tx beam and the \( i \)-th Rx beam between BS 1 and the UE, whereas the interfering signal with the \( k \)-th Tx beam and the \( i \)-th Rx beam between the \( m \)-th interfering BS and the UE is denoted as \( I_{m,ik} \). Assuming the interfering BSs uniformly select each of its \( N_{BS} \) Tx beams for data transmission, i.e. with probability \( 1/N_{BS} \), the UE can select its best beam pair \((i^*, j^*)\), i.e. \( j^* \)-th Tx beam and \( i^* \)-th Rx beam, based on the beams that maximize the following metric:

\[
(i^*, j^*) = \arg\max_{i,j} \frac{S_{ij}}{\sum_{m \in M} \frac{1}{N_{BS}} \sum_{k=1}^{N_{BS}} I_{m,ik} + \sigma_n^2},
\]

where \( \sigma_n^2 \) is the noise variance and \( M \) is the set of interfering BSs. To be able to compute this metric, the UE needs to distinguish the signal coming from different BSs, i.e. different DL pilots are required among neighboring BSs to obtain \( S_{ij} \) and \( I_{m,ik} \) for the different beam pair combinations, that is, to obtain the entries of a table as the one given in Figure 3-2. Hence, an approach based on full pilot reuse would not work. Note that the entries of such a table might also be necessary for considering other multi-node coordination schemes. With the proposed approach, the UE can make the decision, without any extra signaling, and in fact, this decision is totally transparent at the BS, i.e. the BS is unaware of the interference conditions experienced by the UE.

Another possibility for considering the interference, is by considering the worst case interference, such that the beam pair would be selected as follows:

\[
(i^*, j^*) = \arg\max_{i,j} \frac{S_{ij}}{\sum_{m \in M} \max_k I_{m,ik} + \sigma_n^2}.
\]

To evaluate the proposed beam sweeping method for a multi-node scenario, we consider the beam sweeping for a single UE in a cell, which is interfered by 8 neighboring BSs. Further simulation parameters are listed in Table 3-1. The Tx and Rx codebooks consist of steering vectors at uniformly sampled phases. The average rate based on the Shannon formula is depicted in Figure 3-3 for three different methods for determining the best beam pair at the UE: i) based solely on the signal to its connected BS (SNR based, i.e. ignoring the interference), ii) based on the average interference per Rx beam, and iii) based on the maximum interference from each interfering BS. As it can be seen from the figure, there is a performance degradation as the SNR increases, when the interference from neighboring BSs is not taken into account in the beam pair selection. The performance loss increases with increasing SIR as well as with increasing number of interfering BSs. Considering the interference with the proposed methods enables the UE to select a different Rx beam than the one that would be selected based solely on SNR, and in this way avoid potential strong interference from neighboring BSs. Simulation results show that at 10 dB almost 60% of the time the UE selects a different Rx beam than the one that would be selected if the decision would be based solely on the signal strength to its own BS. Although we have assumed a single user, the presented methods can also be extended to multiple users.
Table 3-1: Simulation Parameters

<table>
<thead>
<tr>
<th>Simulation parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of interfering BS</td>
<td>8</td>
</tr>
<tr>
<td>Signal to Interference Ratio (per interfering cell)</td>
<td>-10 dB</td>
</tr>
<tr>
<td>Number of BS antennas (for each BS)</td>
<td>64</td>
</tr>
<tr>
<td>Number of antennas at the UE</td>
<td>16</td>
</tr>
<tr>
<td>Number of Tx beams (for each BS)</td>
<td>$N_{BS} = 64$</td>
</tr>
<tr>
<td>Number of Rx beams at the UE</td>
<td>$N_{MS} = 16$</td>
</tr>
<tr>
<td>Number of independent paths</td>
<td>10</td>
</tr>
<tr>
<td>Channel for path from connected BS to UE</td>
<td>Gaussian with unit variance</td>
</tr>
<tr>
<td>Channel for path from interfering BS to UE</td>
<td>Gaussian with variance 0.1 (SIR = -10dB)</td>
</tr>
</tbody>
</table>

Figure 3-3: Average Sum Rate for Different Methods to Determine the Best beam pair

For the comparison shown in Figure 3-3, the overhead due to the multi-cell beam sweeping has not been taken into account, since the overhead is the same for all the compared schemes, i.e. the number of Tx beams to be trained by each BS is the same. In addition, we also assume that the 9 BSs considered in the setup transmit orthogonal DL pilots, which would enable to obtain the entries of a table as depicted in Figure 3-2. Note that although only the signal strength to its own BS is required by the UE for the SNR-based approach, each BS still needs to send orthogonal pilots to enable the users in each cell to distinguish the desired signal from the interference. In principle, the training overhead could be reduced by reusing the DL pilots as proposed in [ANR+17], e.g. by assuming a full pilot reuse among the neighboring cells. As mentioned before, the decision in this case is based on the total received power, consisting of the desired signal plus interference, and hence, the best beam pair could be very different than when considering the SNR or the interference as well. In fact, with a full pilot reuse and a total received power based approach for determining the best beam pair, the UE could counterintuitively select an Rx beam which potentially could receive the largest interference! Actually, such an Rx beam would be avoided with the proposed schemes which take the interference into account. The approach in [ANR+17] can only be considered in case all the potential inter-cell interference can be neglected.
3.3 Sequential Hybrid Beamforming Design for Multi-Link mm-wave Communication

In this subsection, we discuss a sequential hybrid beamforming design for single-user multi-link transmission over mm-wave frequency bands. By applying the proposed two-step approach, hybrid beamforming becomes an add-on feature that can be easily switched on over an analog beamforming enabled system without interrupting its operation whenever system requires.

As a starting point, a baseline data communication link is established via traditional analog beamforming at both the BS (BS #1) and UE. If an extra RF chain is available at the UE, it can continue to probe the propagation environment at the same frequencies. In case the environment is favorable and system resources allow, a secondary data communication link is established to enable multi-stream transmission. In principle, the secondary link could be served by the same BS (BS #1) or another BS (BS #2). The purpose of establishing the secondary data communication link is to obtain diversity gain or spatial multiplexing gain, depending on digital beamformer design at the UE and BS(s). The detailed formulations are presented in [ZCS+17].

To initialize the secondary data communication link, a parallel beam search scheme is also proposed in [ZCS+17], which helps the UE/BS to find a suitable beam pair with given optimization criteria without interrupting the baseline data communication. As shown in [ZCS+17], with proper pilot signal design and by observing only the relations between signal energy and interference energy without carrying out actual data demodulation, full and accurate synchronization will not be required for this parallel beam search algorithm.

\[
\text{Baseline Data} \quad \begin{bmatrix} S_{1,1,k_1,k_2} & S_{1,2,k_1,k_2} & \cdots & S_{1,p,k_1,k_2} \\ S_{2,1,k_1,k_2} & S_{2,2,k_1,k_2} & \cdots & S_{2,p,k_1,k_2} \end{bmatrix}
\]

\[
\text{Beam Training} \quad \begin{bmatrix} \tilde{s}_{1,k_1,k_2} \\ \tilde{s}_{2,k_1,k_2} \end{bmatrix}
\]

**Figure 3-4: Parallel beam search frame structure for test beam pair \( \{\tilde{f}_k^1, \tilde{f}_k^2, \tilde{w}_{k_2}\} \).**

In detail, an example parallel beam search frame structure is shown in Figure 3-4. \( S_{1,p,k_1,k_2} \) and \( S_{2,p,k_1,k_2} \) are data symbols transmitted over the baseline link with identical transmission power \( \alpha_s^2 \cdot \tilde{f}_k^1 \tilde{f}_k^2 \) and \( \tilde{w}_{k_2} \) refer to the \( k_1 \)-th and \( k_2 \)-th entries of the codebooks at the BS #1 (BS #2) and the UE respectively. \( S_{p,k_1,k_2} \) refers to the \( p \)-th pilot symbol, \( p = 1, \ldots, P \), sent from BS #1 (single-node case) or BS #2 (2-node case) for probing beam pair \( \{\tilde{f}_k^1, \tilde{f}_k^2, \tilde{w}_{k_2}\} \) over the corresponding secondary test link. Notice that the detailed frame structure arrangement and pilot symbol number \( P \) for each test beam pair can be flexibly modified and designed as long as the required observations can be obtained for carrying out the proposed estimation algorithms.

Now, assuming perfect synchronization, the \( p \)-th 2 × 2 data/pilot mixed matrix is transmitted and received using the test beam pair \( \{\tilde{f}_k^1, \tilde{f}_k^2, \tilde{w}_{k_2}\} \) over two symbol durations. The corresponding observation at the UE becomes

\[
y_{p,k_1,k_2} = \tilde{H}_{z,k_1,k_2} \tilde{s}_{p,k_1,k_2} + n_{p,k_1,k_2},
\]

where \( \tilde{H}_{z,k_1,k_2} \) is the 2 × 2 effective testing channel. \( n_{p,k_1,k_2} \) refers to an additive channel noise vector after applying the test analog beamformer at the UE and

\[
\tilde{s}_{p,k_1,k_2} = \begin{bmatrix} S_{1,p,k_1,k_2} \\ \cdots \\ S_{2,p,k_1,k_2} \end{bmatrix}
\]

Based on (3-1), the training signal causes interference on the data transmission. Due to the sparsity of the mm-wave propagation channel and directional nature of mm-wave communication, the interference level is expected to be rather low for most test beam pairs. In
In this context, the power of the pilot transmission $\sigma_{sp}^2 = E \left[ |S_{p,k_1,k_2}|^2 \right]$ is also a design parameter and meanwhile is subject to a sum power constraint at the BS. With higher $\sigma_{sp}^2$, good estimation quality can be achieved with shorter pilot length $P$, yet the interference on the data transmission is larger. With lower $\sigma_{sp}^2$, good estimation quality can be achieved with longer pilot length $P$ and the interference on the data transmission is smaller. In case the interference level is non-negligible, more advanced signal processing algorithms should be deployed, which forms a research topic for future study.

Next, as one concrete example, we consider a DL transmission system that intends to increase the throughput by adding one independent transmission stream over its baseline data transmission. Three estimation approaches are proposed in [ZCS+17]. In the first approach (PHBF #1 in [ZCS+17]), we assume that data symbols are correctly demodulated and inherently the data/pilot mixed matrix $S_{p,k_1,k_2}$ is known. In the second and third approaches (PHBF #2 and PHBF #3 in [ZCS+17]), we assume that no data demodulation is carried out before the beam probe process starts and the data/pilot mixed matrix $S_{p,k_1,k_2}$ is only partially known.

Then we examine the proposed two-step HBF design in [ZCS+17] using extensive computer simulations. The transmitted power is 27 dBm and the distance between the BS #1 and UE is set to be 30 m. In both the single- and 2-node cases, equal transmission power per stream is assumed. The overall channel power between the BS #2 and the UE is minus infinite to 6dB less than that between the BS #1 and UE. The QuaDRiGa geometry-based 3D stochastic channel simulator at 28 GHz is deployed based on measurement results obtained in the mmMAGIC project [JRB+14]. An $8 \times 8$ planar antenna array and a $1 \times 8$ uniform linear array are implemented at the BS and the UE sides respectively. The codebook entries are assumed to be equally distributed over the steering angle domain for elevation angle based on given codebook sizes $K_{BS,L} = K_{BS,L} \times K_{BS,E}$, $t = 1,2$, at the BS side and $K_{UE,E} = K_{UE,A} \times 1$ at the UE side. Here $K_{BS,L}(K_{UE,A})$ denotes the number of codebook entries at the azimuth angle and $K_{BS,E}$ represents the number of codebook entries at the elevation angle, and we select $K_{BS,L} = 32$, $K_{BS,E} = K_{UE,A} = 16$ and $P = 10$. As one numerical example, assuming sampling rate at 122.88 MHz, the exhaustive search requires a 0.66 ms probe time period.

In Figure 3-5 (a-b), achievable sum-rates of the proposed two-step sequential HBF design are evaluated in both the single-node and the 2-node scenarios. The optimal HBF design using perfect CSI for the single-node scenario [ARA+14] is deployed for comparison purpose. In both single-node and multi-node scenarios, with given design parameters, PHBF #1 as well as PHBF #3 can effectively provide performance close to the case using perfect CSI of the effective channel. In general, the proposed hybrid beam search method PHBF #3 in [ZCS+17] appears to be the most efficient estimation approach with limited $P$.

Next the impact of the design parameters is evaluated using PHBF #3. As shown in Figure 3-6 (a-b), we compare the achievable sum rates using three different $\sigma_{sp}^2$ and pilot symbol length $P$. Denoting $\beta = \frac{\sigma_{sp}^2}{\sigma_{sp}^2}$, there is a clear trade-off between the pilot transmission power and pilot symbol length. As shown in Figure 3-7 (a-b), the cumulative density function (CDF) of the SINR of the baseline data transmission under the proposed parallel beam training scheme is compared with different $\beta$ values. It shows that parallel beam training has little impact on the baseline data transmission in the SNR range of interest for mm-wave communication.
Figure 3-5: Performance comparisons of digital beamforming (DB), reference HBF in [ARA+14], the proposed HBF using perfect CSI of the effective channel (PHBF), PHBF #1, PHBF #2 and PHBF #3 in [ZCS+17]. (a) One-node scenario and (b) 2-node scenario. SNR refers to received SNR at each UE antenna input from BS #1.

Figure 3-6: Performance comparisons of the proposed HBF using PHBF #3 in [ZCS+17] with perfect CSI knowledge (PHBF) and different pilot transmission powers ($\beta = 1, 2, 4$). SNR=10 dB. (a) One-node scenario and (b) 2-node scenario. Two-node scenario. SNR refers to received SNR at each UE antenna input from BS #1.

Figure 3-7: CDF of SINR on the baseline data transmission when parallel beam training is carried out with different pilot transmission powers. $\beta = 1, 2, 4$. SNR=10 dB. (a) One-node scenario and (b) 2-node scenario. SNR refers to received SNR at each UE antenna input from BS #1.
3.4 Beam management for mobility

To support random access and mobility at mm-wave frequencies, a toolbox for beam management is necessary, [Eri17-1]. This toolbox consists of a set of procedures that can be configured depending on the deployment, the capability of the hardware, the backhaul quality, the traffic type for a given UE, etc.

3.4.1 Beam management procedures

The three underlying procedures (P1-3) are outlined in [Eri17-2], which assumes a degree of beamforming capability in the UE, and are illustrated in Figure 3-8.

![Beams](image)

**Figure 3-8**: Illustration of beam management procedures. (Top) P1 to enable UE measurement on different TRP Tx beams to support selection of TRP Tx beams/UE Rx beam(s) (Middle) P2 to enable UE measurements on different TRP Tx beams to possibly change/select inter/intra-TRP Tx beam(s), (Bottom) P3, to enable UE measurement on the same TRP Tx beam to change UE Rx beam in the case UE uses beamforming

The beam-management is performed on three levels, depending on scenario and quality of CSI, and backhaul. These are as follows.

- **Level 1**: Event-triggered synchronization signal (SS) based reporting using L3. Can be useful to bootstrap the beam management procedures in level 2 and 3, in case SS is beam-formed in beams that are narrower than the sector.
- **Level 2**: CSI-RS based feedback (P1), typically a cell-specific and periodic beam sweep for a group of UEs for their beam finding.
- **Level 3**: CSI-RS based feedback (P2/P3), typically a UE-specific and aperiodic beam sweep for single UE for its beam refinement.

Each SS block contains PSS (Primary SS), SSS (Secondary SS), and PBCH, and the different blocks within a burst may be transmitted either in a beam sweeping or a wide beam manner depending on the deployment scenario and coverage requirements.

**Level 1: Beam based mobility measurements**

The beam based mobility reporting using SS measurements typically provides coarse/wide beam selection information for a UE which can be used for initial PDCCH transmission, PUCCH reception and low code rate PDSCH, PUSCH transmissions. This should be sufficient to configure the L1/L2 beam management. The UE selects its receive and transmit beamforming weights autonomously.
Figure 3-9: Beam based mobility measurements using Level 1 SS. The SS’s may be wide/sector covering (left) or somewhat beamformed (right, coarse beams) depending on the user scenario. The network can use the SS report as a starting point for the Level 2 beam management using P1 or go directly to Level 3 beam management using P2/P3 in the case SS reports already gives some beam direction information (right figure).

Level 2: Beam management (P1 using CSI-RS)
A sweep of beam-formed CSI-RS is used and the UE reports a single beam and CSI useful for link adaptation, at least for rank 1 or 2 transmissions. The report can be used to schedule small packets, delay sensitive packets or as an initial estimate of useful beams to further refine in Level 3 beam management. Typically, many beams are covered in one sweep burst and multiple UEs are simultaneously measuring and reporting based on the one beam sweep. The UE determines its RX/TX beamforming autonomously and there is no beam indication in the scheduling DCI.

Figure 3-10: In Level 2 based beam reporting and when the load is low (left) the SS reports from the Level 1 reporting can be used to select a restricted beam sweep for the UE which may be periodic. TRPs with no users (as observed from SS measurement reports) need not transmit CSI-RS at all. If the load is high (right), the TRP nodes makes a periodic beam sweep for CSI-RS reporting, when Level 1 doesn’t use beamforming or uses sector covering wide beams (Figure 3-9, left).

Level 3: Beam management (P2 and P3 using CSI-RS)
Aperiodically triggered reporting on a set of UE specifically beamformed CSI-RS and CSI reporting. The process P-3 is used to refine the UE beam and a beam indication is supported in the scheduling DCI. Level 3 reporting is typically triggered when there is a large data packet to transmit which will consume many subframes. Level 3 reporting may also be more frequently used at higher carrier frequencies when P3 may be needed for some terminals that does not have omni-coverage.

Figure 3-11: Level 3 beam management with aperiodically triggered UE specific beamformed CSI-RS (P-2) and UE RX beam refinement (P-3).
3.5 Performance of mm-wave based RF-FSO Multi-hop Networks

3.5.1 On the Performance of mm-wave based RF-FSO Multi-hop Networks

The next generation of wireless networks must provide coverage for everyone everywhere at any time. To address these demands, a combination of different techniques is considered, among which free-space optical (FSO) communication is very promising. Coherent FSO systems provide fiber-like data rates through the atmosphere using lasers. Thus, FSO can be used for a wide range of applications such as last-mile access, fiber back-up, backhauling and multi-hop networks. In the radio frequency (RF) domain, on the other hand, millimeter wave (mm-wave) communication has emerged as a key enabler to obtain sufficiently large bandwidths so that it is possible to achieve data rates comparable to those in the FSO links. In this perspective, the combination of FSO and mm-wave based RF links is considered as a powerful candidate for high-rate reliable communication.

In this work, we study the data transmission efficiency of multi-hop RF-FSO systems from an information theoretic point of view. Considering the mm-wave characteristics of the RF links and heterodyne detection technique in the FSO links, we derive closed-form expressions for the system outage probability and evaluate the effect of imperfect hardware on the system performance. Our results are obtained for the decode-and-forward relaying approach in different cases with and without hybrid automatic repeat request (HARQ). Finally, note that the work presented in this deliverable is based on our extended papers [MSB+17a, MSB+17b] in which we study the outage probability and the ergodic rate of the RF-FSO based multi-hop and mesh networks and determine the required number of antennas in each hop to guarantee different rates. Thus, the interested readers are referred to [MSB+17a, MSB+17b] for more detailed discussions.

3.5.2 System model

Consider a \( T_{\text{total}} \)-hop RF-FSO system, with \( T \) RF-based hops and \( \tilde{T} = T_{\text{total}} - T \) FSO-based hops. As seen in the following, the outage probability is independent of the order of the hops. Thus, we do not need to specify the order of the RF- and FSO-based hops. The \( i \)-th, \( i = 1, \ldots, T \), RF-based hop uses a multiple-input-single-output (MISO) setup with \( N_i \) transmit antennas. We present the analytical results for the quasi-static Rician channel model which is an appropriate model for near line-of-sight conditions and has been well established for different mm-wave based applications [MSB+17a, MSB+17b].

Let us denote the probability density function (PDF) and the cumulative distribution function (CDF) of a random variable \( X \) by \( f_X(\cdot) \) and \( F_X(\cdot) \), respectively. Denoting the sum channel at the receiver of the \( i \)-th hop by \( G_i = \sum_{j=1}^{N_i} g_{ij}^i \), with \( g_{ij}^i \) being the channel gain between the \( j \)-th antenna and the receiver, we have

\[
 f_{G_i}(x) = \frac{(K_{i-1})^x e^{-K_{i-1}x}}{\Omega_i} \frac{(K_i+1)^x}{K_i N_i \Omega_i} \frac{\Gamma(N_i-1)}{\Gamma(N_i-1)} \left( 2 \sqrt{\frac{K_i (K_i+1) N_i \Omega_i}{\Omega_i}} \right),
\]

where \( \Gamma(\cdot) \) denotes the \( n \)-th order modified Bessel function of the first kind and \( \Omega_i \) and \( K_i \) are the fading parameters.

Finally, to take the non-ideal hardware into account, we consider the power amplifier (PA) efficiency model of [MSC+15], [MSE+16] where the output power at each antenna of the \( i \)-th hop is determined according to

\[
 \frac{P_i}{P_{\text{cons},i}} = \varepsilon_i \left( \frac{P_i}{P_{\text{max},i}} \right)^{\nu_i},
\]

where \( P_i, P_{\text{cons},i} \) and \( P_{\text{max},i} \) are the output, the consumed and the maximum output power of the PA, \( \varepsilon_i \) is the PA efficiency and \( \nu_i \) is a parameter depending on the PA class.
The FSO links, on the other hand, are assumed to have single transmit/receive terminals. Here, we present the results for cases with exponential and Gamma-Gamma distributions of the FSO links where the channel gain $\hat{G}_i$ follows

$$f_{\hat{G}_i}(x) = \lambda_i e^{-\lambda_i x} \quad (3-5)$$

and

$$f_{\hat{G}_i}(x) = \frac{a_i^b_i}{\Gamma(a_i) \Gamma(b_i)} x^{a_i+b_i-1} K_{a_i-b_i}(2\sqrt{a_i b_i x}) \quad (3-6)$$

respectively. Here, $\lambda_i, a_i$ and $b_i$ are the fading parameters. Also, $K_n(\cdot)$ denotes the modified Bessel function of the second kind of order $n$ and $\Gamma(x)$ is the Gamma function.

### 3.5.3 Data Transmission Model

We consider the decode-and-forward technique where at each hop the received message is decoded and re-encoded, if it is correctly decoded. Therefore, the message is successfully received by the destination if it is correctly decoded in all hops. Otherwise, outage occurs. As the most promising HARQ approach leading to highest throughput/lowest outage probability [MSB+17a], [MSC+17], [MSE+16], we consider the incremental redundancy (INR) HARQ with a maximum of $M_i$ retransmissions in the $i$-th hop, $i = 1, \ldots, t_{\text{total}}$, hop. Using INR HARQ with a maximum of $M_i$ retransmissions, $q_i$ information nats are encoded into a parent codeword of length $M_i L$ channel uses. The parent codeword is then divided into $M_i$ sub-codewords of length $L$ channel uses which are sent in the successive transmission rounds. Thus, the equivalent data rate, i.e., the code rate, at the end of round $m$ is $\frac{q_i}{m L}$ nats-per-channel-use (npcu) where $R_i = \frac{q_i}{L}$ denotes the initial code rate in the $i$-th hop. In each round, the receiver combines all received sub-code words to decode the message. The retransmission continues until the message is correctly decoded or the maximum permitted transmission round is reached. Note that setting $M_i = 1, \forall i$, represents the cases without HARQ, i.e., open-loop communication.

### 3.5.4 Analytical results

Because independent channel realizations are experienced in different hops, the system outage probability is given by

$$\Pr(\text{Outage}) = 1 - \prod_{i=1}^{t_{\text{total}}} (1 - \phi_i) \prod_{i=1}^{t_{\text{total}}} (1 - \tilde{\phi}_i), \quad (3-7)$$

where $\phi_i$ and $\tilde{\phi}_i$ are the outage probability in the $i$-th RF- and FSO-based hops, respectively. Thus, to analyze the outage probability, we should determine $\phi_i$ and $\tilde{\phi}_i$ which are given by:

$$\phi_i = \Pr \left( \frac{1}{M_i \hat{G}_i} \sum_{\ell=1}^{M_i} \sum_{g=(m-1)\hat{G}_i+1}^{m \hat{G}_i} \log \left( 1 + \frac{\epsilon_i \rho_i^{\text{cons}}}{\rho_i^{\text{max}} \alpha_i} G_i(c) \right) \leq \frac{R_i}{M_i} \right) \quad (3-8)$$

and

$$\tilde{\phi}_i = \Pr \left( \frac{1}{M_i \hat{G}_i} \sum_{\ell=1}^{M_i} \sum_{g=(m-1)\hat{G}_i+1}^{m \hat{G}_i} \log \left( 1 + \tilde{\rho}_i \alpha_i \hat{G}_i(c) \right) \leq \frac{R_i}{M_i} \right). \quad (3-9)$$
Here, $c_i$ and $\tilde{c}_i$ are the number of channel realizations experienced in each codeword transmission of the RF- and FSO-based hops, respectively.

**Lemma 1.** The probability (3-8) is given by

$$\phi_i = \frac{1}{2} \left( 1 + \text{erf}\left( \frac{\sqrt{M_i C_i (x-\mu_i)}}{2\sigma_i^2} \right) \right), \quad (3-10)$$

with $\mu_i$ and $\sigma_i$ given in [MSB+17b], Eq. (12)-(13).

Proof. See [MSB+17b, Lemma 2].

**Lemma 2.** The probability (3-9) is given by

$$\tilde{\phi}_i = \frac{1}{2} \left( 1 + \text{erf}\left( \frac{\sqrt{M_i C_i (x-\mu_i)}}{2\sigma_i^2} \right) \right). \quad (3-11)$$

Here, $\tilde{\mu}_i$ and $\tilde{\sigma}_i$ are constant values given by [MSB+17a, Eq. (17)-(18)] for the exponential distributions of the FSO link and by [MSE+16, Eq. (43)-(44)] in the cases with Gamma-Gamma PDF of the FSO link.

Proof. See [MSB+17b], Lemma 3.

### 3.5.5 Simulation results

We present the simulation figures for a dual-hop RF-FSO setup with one RF- and one FSO-based hop. In Figure 3-12(a-b), the results are presented for the exponential and the Gamma-Gamma PDFs of the FSO link, respectively, while the RF hop follows Rician PDF. We set the fading parameters of (3-3), (3-5) and (3-6) to $a_i = 4.3939, b_i = 2.5636, \lambda_i = 1, K_i = 0.01, \Omega_i = 1$ which lead to the unit mean and variance of the RF and FSO hops. Figure 3-12(a) verifies the accuracy of the theoretical results where we show the outage probability for the cases with $M_i = 1, R_i = 1, 2, C_i = 10, \tilde{C}_i = 30$ and with an ideal PA. With an ideal PA we set $\varepsilon_i = 1, \nu_i = 0, P_i^{\text{max}} = \infty$ in (3-42). Then, Figure 3-12(b) evaluates the effect of HARQ and PAs efficiency on the system outage probability. Here, the results are obtained for $\varepsilon_i = 0.75, \nu_i = 0.5, P_i^{\text{max}} = 25 \text{ dB}, M_i = 1, 2, 3, R_i = 3, C_i = 10, \tilde{C}_i = 20$.

The results indicate that:

1) the analytical results of Lemmas 1-2 mimic the exact results with very high accuracy. As a result, Lemmas 1-2 can be effectively used to analyze the data transmission efficiency of the RF-FSO multi-hop networks, as well as the multi-hop networks with only the RF- or the FSO-based communication.

2) With no HARQ, the efficiency of the RF-based PAs affects the system performance considerably. For instance, with the parameter settings of the Figure 3-12(b) and outage probability $10^{-4}$, the PAs inefficiency increases the required power by 3.5 dB.

3) On the other hand, the HARQ can effectively compensate the effect of imperfect PAs, and the difference between the outage probability of the cases with ideal and non-ideal PAs is negligible for $M_i > 1$. 


4) Finally, the HARQ improves the energy efficiency significantly. As an example, consider the outage probability $10^{-4}$, an ideal PA and the parameter settings of Figure 3-12(b). Then, compared to the open-loop communication, i.e., $M_f = 1$, the implementation of HARQ with a maximum of 2 and 3 retransmissions reduces the required power by 13 and 17 dB, respectively.

![Figure 3-12 (a): On the tightness of the analytical results. (b): Outage probability for different PA models and number of HARQ-based retransmissions.](image)

3.6 mm-wave LOS Coverage Enhancements with Coordinated High-Rise Access Points

Although it has been shown in [IKO+15] that high data rate via mm-wave communication can be supported by surprisingly rich NLOS links through reflected paths, there still could be a considerable performance degradation compared to LOS links. It becomes more crucial to maintain a LOS link for users requiring very high data rate to support the 5G immersive experiences [YHN+16]. In this regard, we focus on LOS coverage in this paper and study the LOS coverage enhancement provided by multi-node coordination. In particular, we consider the coordination between two sets of mm-wave APs: 1) low-rise APs installed on street furniture; and 2) high-rise APs installed on high buildings, and try to answer the following three questions:

1. How many LOS APs can be observed by a typical user equipment (UE) in this specific network scenario?
2. What is the probability that the user is associated with a LOS mm-wave AP (therefore covered by a LOS connection)?
3. What is the improvement in coverage when coordination among two sets of APs is considered?

3.6.1 LOS probability

The height of APs has significant impact on the probability of LOS links as shown in Figure 3-13, where both building A and B are in between the AP and the UE but the LOS link is blocked by the higher building A.
We study two different deployment options: 1) irregular deployment where the AP deployment follows a homogeneous Poisson point process (PPP) and 2) regular deployment where the AP deployment is the conventional hexagon pattern.

Considering the model in Figure 3-13, the building A will block the LOS link as long as its height \( h \) is larger than \( h_y \). The blocking probability can be expressed as

\[
P_{\text{blk}} = \int_0^R f(y) P(h > h_y) dy
\]

where \( f(y) \) is the distribution function of the distance between the building and the UE and \( h_y \) can be expressed as

\[
h_y = \frac{yH_B + (R - y)H_U}{R}
\]

where \( H_B \) is the AP height, \( H_U \) is the UE height and \( R \) is the distance between the AP and the UE. Assuming the distance between the buildings and the UE follows uniform distribution in the range of \([0, R]\) and the height of buildings follows uniform distribution in the range of \([0, H_{\text{max}}]\), the blocking probability can be derived as

\[
P_{\text{blk}} = \frac{(H_{\text{max}} - H_U)^2}{2H_{\text{max}}(H_B - H_U)}
\]

When the height is taken into consideration, \( \beta \) should be scaled by \( P_{\text{blk}} \) as \( \beta' = \beta P_{\text{blk}} \) [TRR14], where \( \beta \) is a parameter determined by size and density of blockages. Therefore, the UE LOS association probability is given by

\[
P_{\text{LOS}} = \int_0^\infty e^{-2\pi x} \left[ 1 - \psi_L(x) \right] \psi_L(x) e^{-2\pi x U(x)} dx
\]

where

\[
Y(x) = \frac{x^2}{2} - U(x), U(x) = \int_0^r \rho(p) dp = \frac{1}{\beta P_{\text{blk}}} - \frac{1}{\beta P_{\text{blk}}} e^{-\beta x} \left( \frac{1}{\beta P_{\text{blk}}} + x \right).
\]

Here \( \lambda \) is the density of access points, \( \rho(t) \) is a general LOS probability function and \( \psi_L(x) \) is

\[
\psi_L(x) = \left( \frac{C_N}{C_L} \right)^{\alpha_L / \alpha_N} x^{\alpha_L / \alpha_N},
\]

where \( C_N \) and \( C_L \) are the intercepts of the LOS and NLOS path loss formulas, respectively, and \( \alpha_L \) and \( \alpha_N \) are the path loss exponents for LOS and NLOS, respectively. The above results can be easily extended to a case beyond point to point.
3.6.2 Evaluation Results

We also show some numerical results to demonstrate the impact of height of AP on LOS association probability and the potential improvement of LOS coverage using the proposed multi-node coordination. The system parameters are given in Table 3-2.

<table>
<thead>
<tr>
<th>Simulation Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Carrier Frequency</td>
<td>28 GHz</td>
</tr>
<tr>
<td>Cell radius (m)</td>
<td>up to 1000</td>
</tr>
<tr>
<td>$a_{hi}$</td>
<td>2</td>
</tr>
<tr>
<td>$a_{li}$</td>
<td>4</td>
</tr>
<tr>
<td>$H_0$ (m)</td>
<td>1.5</td>
</tr>
<tr>
<td>$H_{\text{low}}$ (m)</td>
<td>3, 10, 15 and 30</td>
</tr>
<tr>
<td>$\beta$</td>
<td>0.0709</td>
</tr>
</tbody>
</table>

Figure 3-14 shows the LOS association probability for different AP heights with increased average cell radius. It can be easily seen that the height of the AP and cell density play crucial roles in determining the UE LOS association probability. When a lower AP is installed on the street furniture, e.g., lamp post, with height 3 m, the LOS association probability reduces rapidly with increased cell radius, i.e., reduced cell density. However, when an AP is installed in a high building with height 30 m, the LOS association probability decreases much slower. It means that the high-rise APs can always provide significantly high LOS coverage even with low cell density.

![Figure 3-14: LOS association probability for $H_b = 3$ and $30$ m, respectively ($H_{\text{max}} = 15$ m)](image)

From the simulation results, we can see that the LOS probability can be significantly enhanced by increasing the height of APs. However, high-rise APs cannot be deployed as dense as low-rise APs because high buildings are not always as available as street furniture such as lamp poles. Therefore we consider installing only a small number of high-rise APs on top of a large amount of low-rise APs and show the LOS coverage improvement due to the joint deployment. In Table 3-3, we show that for a target overall LOS probability 0.95, how many high-rise APs (30 m) need to be installed together with each 100 low-rise APs (3 m) in a regular deployment scenario. It can be seen that only very few high-rise APs are needed to enhance the LOS coverage when the blocking building height is low. However, with increased blocking building height, more and more high-rise APs are needed to provide target LOS coverage.
Table 3-3: Number of high-rise APs

<table>
<thead>
<tr>
<th>Blocking Building Height (m)</th>
<th>$P_{LOS}$ with Low-Rise AP only</th>
<th>Number of High-rise APs</th>
<th>$P_{LOS}$ with Joint Deployment</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>0.93</td>
<td>1</td>
<td>0.9990</td>
</tr>
<tr>
<td>5</td>
<td>0.44</td>
<td>3</td>
<td>0.9578</td>
</tr>
<tr>
<td>10</td>
<td>0.02</td>
<td>25</td>
<td>0.9636</td>
</tr>
<tr>
<td>15</td>
<td>0.0005</td>
<td>100</td>
<td>0.9513</td>
</tr>
</tbody>
</table>

3.7 Direct vs Relay-assisted Access: A System Level Evaluation at mm-waves

Penetration loss (PL) strongly increases when the carrier frequency increases: first measurement campaigns [ZMS+13, Table II] have shown that the PL can reach values of 40 dB already at 28 GHz, which is 100 times higher than the value of 20 dB, originally proposed for frequencies below 6 GHz. Therefore, indoor user equipment devices (UEs) might be strongly noise limited when served by an outdoor base station (BS) or small cell, experiencing extremely low signal to interference plus noise ratio (SINR). Since most of the UEs are typically indoor, it is important to develop solutions at mm-wave frequencies that are capable of providing good coverage to these UEs.

With this objective, we analyse here the benefits of having relay stations (RSs) to improve network coverage. In particular, we compare two main configurations, as depicted in Figure 3-15:

- Direct access (DA): no RS is deployed and the BSs directly serve the UEs; the whole bandwidth is available, but low SINR are experienced by the UEs because of the high BS-UE distance and PL.
- Relay-assisted access (RA): RSs are deployed, BSs provide the backhaul to decode-and-forward RSs, which, in turn, serve the UEs; bandwidth is split among access and backhaul, but higher SINR is observed because of the reduced RS-UE distance.

These two configurations are compared in a system operating at 73 GHz with a bandwidth of 1 GHz, with 7 sites, 3 macro BSs per site and wraparound. We assume 4 RSs per macro sector and 5 indoor full-buffer UEs dropped within a radius of 40 m around each RS (full hotspot UE distribution), i.e., 20 indoor UEs overall per macro sector. Macro BSs are equipped with 256 antennas and have a maximum transmit power of 36 dBm, whereas RSs are equipped with 32 antennas and have a maximum transmit power of 30 dBm. For the channel model we implement a simplified 2D version of the model developed in WP2 [MmMAGICCD21]. Moreover, we assume that both BSs and RSs can exploit a fully digital beamforming architecture.

Figure 3-15: Illustration of the considered two considered access methods. (a) Direct access vs (b) Relay-assisted access
For the DA scenario, we assume proportional fair scheduling with a greedy user selection at the BSs, equal power allocation among the UEs served in each transmission time interval (TTI), and zero forcing (ZF) beamforming.

With RA we assume that the backhaul and access links use distinct contiguous sub-bands, i.e., the BS-RS links do not interfere with the RS-UE links and vice versa. Moreover, in the backhaul link, i.e., BS-RS, all the 4 RSs are served by their anchor BS in each TTI, with equal power allocation and using multi-user eigenmode transmission (MET) [BH07]. In the access link of RA, i.e., RS-UE, the RSs operate similarly to the BS of the DA configuration: proportional fair scheduling with greedy user selection, equal power allocation and ZF are used.

In Figure 3-16 and Figure 3-17 we compare DA against RA by showing the sector spectral efficiency (SSE), i.e., the spectral efficiency obtained in the area covered by a macro sector, and the cell border throughput. The performance is evaluated based on the 5th percentile of the UE rate, for an inter-site distance (ISD) that varies from 250 m to 1000 m. There, we also assume that the sub-bands allocated to access and backhaul are equally large, i.e., 500 MHz each. As expected, the performance of both the schemes gets worse when the ISD increases, as the path-loss attenuation becomes more dominant and therefore lower spectral efficiencies are achieved. Moreover, as with RA the UEs are served by close-by RSs, we observe in Figure 3-17 that RA outperforms DA in the cell border throughput: indeed, RSs are particularly helpful in serving the UEs that are far away from the macro BSs. On the other hand, using RSs to serve UEs that are close to the macro BSs (even if they are indoor), can be detrimental, because the gain due to the higher SINR is more than offset by the loss due to the smaller bandwidth available for the access link. Indeed, we observe in Figure 3-16 that DA slightly outperforms RA in the SSE, although only for ISD≤500 m.

![Figure 3-16: System performance achieved by DA and RA: macro sector spectral efficiency](image)

![Figure 3-17: System performance achieved by DA and RA: 5th percentile of the UE rate](image)
In the previous results the available bandwidth with RA was equally split among the backhaul and access links. In Figure 3-18 and Figure 3-19 we instead denote with $BW_B$ the bandwidth allocated to the backhaul, i.e., to the BS-RS links, and consider a system bandwidth of 1 GHz and a backhaul bandwidth of $BW_B$ ranging from 300 up to 900 MHz. The corresponding bandwidth allocated to the access link is thus 700 MHz down to 100 MHz. We report the results obtained for ISD=250 m. In Figure 3-19 we observe that the cell border throughput increases when $BW_B$ increases, at least for the values considered here. In fact, the UEs experiencing a lower throughput are the ones far away from the macro BS: with RA, the RS serving these UEs, although close to them because of the hotspot UE distribution, is far from the macro BS as well. As a consequence, the backhaul link tends to be the bottleneck for these UEs and increasing $BW_B$ improves the throughput of these UEs at the cell border. Therefore, when looking only at the cell border throughput, most of the 90% of the bandwidth should be allocated to the backhaul. Clearly, one should also note that the cell border throughput does not monotonically increase with $BW_B$. There is an optimal value, which however lies after 90%. On the contrary, if we look at Figure 3-18, we observe that the sector spectral efficiency first increases when $BW_B$ increases, and then starts decreasing after reaching the optimal value (which is about 40%). This happens because the sector spectral efficiency is affected also by the UEs close to the BSs, for which the backhaul link is not such a bottleneck.

![Figure 3-18](image)

**Figure 3-18:** Impact of the bandwidth allocated to the BS-RS link with RA on the macro sector spectral efficiency

![Figure 3-19](image)

**Figure 3-19:** Impact of the bandwidth allocated to the BS-RS link with RA on the 5th percentile of the UE rate
Finally, looking at the complete results together, we can conclude that RSs at mm-wave can be very beneficial in improving system performance: In fact, for an exemplary ISD=500 m, DA and RA achieve the same SSE, but RA provides a gain of about 120% in the cell border throughput.

### 3.8 Joint Hybrid Precoding for Energy-efficient mm-wave Networks

#### 3.8.1 Introduction

Mm-wave hybrid beamforming design, maximizing spectral efficiency, has been shown to give close performance to the fully digital precoding scheme [RMG+16], [ARA+14]. However, only a few works have studied the energy efficiency of the precoder design. When joint transmission is allowed, a major research problem is to jointly design the analog and the digital precoder such that the total power consumption is minimised and the quality of service of each user is satisfied.

In this work, we adopt the hybrid precoding structure, by allowing joint transmissions from multiple BSs to each user. We study the minimum transmit power required by each BS in order to satisfy a spectral efficiency constraint for each user.

#### 3.8.2 System Model

We consider a mm-wave network consisting of $M$ BSs and $K$ users. BS $v$ is assumed to be equipped with $N_v$ antennas and $S_v$ RF chains, where $S_v < N_v$. Each user is assumed to be equipped with a single antenna and is not pre-associated with any BSs. Due to the fact that strict synchronization among BSs is difficult to achieve, we assume that each user uses successive interference cancellation technique to sequentially detect multiple streams from different BSs. The digital precoder for user $k$ designed by BS $v$ is denoted by $b_{k,v} \in \mathbb{C}^{S_v \times 1}$ and the analog precoder at BS $v$ is denoted by $R_v \in \mathbb{C}^{N_v \times S_v}$, where each element has the same magnitude. In this work, we adopt the clustered channel model so that the channel coefficients from BS $v$ to user $k$ is given by

$$h_{k,v} = \left[ \frac{N_v L_{k,v} \sum_{i=1}^{N_{cl,v}} \sum_{j=1}^{N_{ray,v}} \beta_{i,l} a(\alpha_{i,l})} {\sqrt{N_{cli} N_{ray,v}}} \right],$$

(3.17)

where $L_{k,v}$ denotes the path loss, $N_{cl,v}$ is the number of scattering clusters, $N_{ray,v}$ is the number of multipaths within the same cluster, $\beta_{i,l}$ is the channel gain and $\alpha_{i,l}$ is the antenna steering vector evaluated at the angles of departure from BS $v$ to user $k$. The received signal-to-interference-plus-noise ratio is given by

$$\gamma_k = \frac{\sum_{v=1}^{M} |h_{k,v} R_v b_{k,v}|^2}{I_k + \sigma_k^2},$$

(3.18)

where $I_k = \sum_{v=1}^{M} \sum_{i=1,i \neq k}^{K} |b_{i,v}^* R_v^* h_{k,v}|^2$ is the interference power and $\sigma_k^2$ is the additive noise power. Note that if a user $l$ is not served by BS $v$, we can simply set $b_{k,v} = 0$, thus by jointly designing the analog and digital precoder, we also get a solution for the user association problem. Furthermore, we introduce a sleep mode to each BS and define the expected transmit power from BS $v$ as

$$P_v = \Delta_v \sum_{k=1}^{K} |R_v b_{k,v}|^2 + S_v P_a, \text{ if the BS is active, }$$

$$P_v = S_v P_s, \text{ if the BS is at sleep mode. Here } \Delta_v \text{ models the power amplifier efficiency and } S_v P_a, S_v P_s \text{ denote the hardware power consumption for the active mode and sleep mode, respectively. Our objective is to find an analog precoder for each BS and a digital precoder for each user-BS pair such that total power consumption is minimized, while fulfilling a maximum per-BS power constraint and a rate constraint for each user. The problem can be formally stated as below:}$$

$$\min_{R_v, b_{k,v}} \sum_{v=1}^{M} \alpha_v P_v$$

(3.19)

s.t. $\gamma_k \geq \Gamma_k, \quad \forall k$

(3.20)
where \( a_v \) is the weighting parameter, \( \Gamma_k \) is used to guarantee a minimum spectral efficiency for each user and the last constraint is the unit modulus constraint on the analog precoders.

### 3.8.3 Joint Hybrid Precoding

Due to the unit modulus constraint, the optimization problem for hybrid precoding is usually too complex to find a global optimal solution, thus most research tries to find a suboptimal solution that gives close performance to the fully digital precoding scheme. In a similar manner as [LBS+16], we reformed the above problem as a semidefinite program, conditioned on fixed analog precoders. Our algorithm starts by solving for the optimal fully digital precoder without the analog constraint. Then, we initialise the analog precoder as the element-wise normalisation of the digital precoder and conditioned on the analog precoder, we obtain the digital precoder by solving the optimisation problem (3-19)-(3-22).

![Figure 3-20](image)

**Figure 3-20:** Average total transmit power \( \sum_v \Delta_v \sum_{k=1}^K ||[R_v, b_k]||^2 \) vs. target spectral efficiency per user for a given user drop over 500 channel realisations. “Optimal” - the case of fully-digital beamforming with the optimal BS mode combination. “Hybrid” - the case of hybrid beamforming.

The simulation results are obtained for 3 BSs (1 macro BSs and 2 small BSs) and 4 users. The macro BS (MBs) is assumed to have 32 antennas and 4 RF chains. The small BSs (SBs) are assumed to have 8 antennas and 4 RF chains. The maximum transmit power is set to 46 dBm for the macro BS and 43 dBm for small BSs.
Figure 3-21: CDF of the total power consumption for 3 coordinated BSs including the hardware power consumption $S_V P_a$ and $S_V P_x$. The target spectral efficiency is 4 bit/s/Hz.

Figure 3-22: CDF of the total power consumption for 2 coordinated BSs including the hardware power consumption $S_V P_a$ and $S_V P_x$. The target spectral efficiency is 4 bit/s/Hz.

Figure 3-20 shows the total transmit power for the hybrid precoding scheme and the fully digital-only precoding scheme with the optimal BS mode combination. As expected, the hybrid precoding scheme achieves less spectral efficiency than the fully-digital precoding while transmitting with the same power. Note that, even though the maximum transmit power of MBs and SBs is assumed to be 46 dBm and 43 dBm, respectively, the average total transmit power needed for 3 BSs to achieve a target per-user spectral efficiency of 2.5 bit/s/Hz is approximately 1W (30 dBm), and the average total transmit power increases with the target per-user spectral efficiency.

Figure 3-21 shows that the total power consumption of the hybrid beamforming scheme is much lower than that of the digital beamforming scheme due to the reduced number of RF chains. In Figure 3-22, we show the total power consumption of 1 macro BS and 1 small BS. By comparing the CDF in Figure 3-21 with the CDF in Figure 3-22, we find that, even though the number of coordinated BSs decreases, the total power consumption increases. This is because more coordinated BSs provide a better chance for a user to be jointly served by BSs with good channel conditions, while decreasing the number of BSs would result increased transmit power from the remaining BSs in order to serve users experiencing blockages or...
deep fading, and the increase in the transmit power outweighs the decrease in the hardware power consumption under our parameter settings. Thus, we have showed that our joint precoding scheme achieves better energy focusing beams and association schemes that reduce shadowing effects and propagation losses. It demonstrates the importance of BSs coordination in terms of energy efficiency and load balance and adding more BSs into coordination would help reduce the total power consumption of the network.

3.8.4 Conclusion

Our proposed algorithm provides a low complexity solution to analyse the energy efficiency and load balance in hybrid precoding mm-wave networks. It is shown that, by jointly designing hybrid precoders, the total transmit powers can be reduced while satisfying the spectral efficiency constraint for each user. Compared to a fully digital beamforming the average total transmit power is naturally larger for the more constrained hybrid precoders, but the overall power consumption is lower for a given targeted rate for each user due to the fewer RF chains. We also showed that there are large gains in total energy efficiency with joint multi-node multi-stream transmission, and it increases with the number of cooperative BSs. In addition, the distribution of the power consumption over the cooperative nodes becomes more even, which might be useful for meeting potential EIRP restrictions of mm-wave nodes.

3.9 Summary and key results and observations for System Concept Design

In this chapter we have reported the work within the mmMAGIC project related to multi-node transmission and reception schemes.

We started by a short review of the results in [MMMAGICD51].

Then, a multi-node coordinated beam sweeping approach was proposed and analysed, showing that there are substantial gains compared to neglecting the multi-node interference in dense multi-node networks.

We proceeded with a sequential hybrid beamforming design for multi-link mm-wave communication in which a two-step precoding approach is adopted. With that approach hybrid beamforming becomes an add-on feature that can be easily switched on over an analog beamforming enabled system without interrupting its operation whenever the system requires. It was shown that such a parallel beam training has little impact on the baseline data transmission in the SNR range of interest for mm-wave communication.

We then proceeded with presenting a flexible beam-management approach to support random access and mobility at mm-wave frequencies, assuming a certain degree of beamforming capability at the UE. As shown, the toolbox consists of a set of procedures that can be configured depending on the deployment, the capability of the hardware, the backhaul quality, the traffic type for a given UE, etc.

In D5.1 [MMMAGICD51] we showed the potential of RF-FSO systems in single-hop deployments, and here we generalized the concept to multi-hop networks. In particular, we investigated the performance of a mm-wave-based RF-FSO system for multi-hop networks using HARQ. In the illustrated dual-hop scenario (one RF and one FSO hop), we showed that the HARQ based dual-hop RF-FSO system improves the energy efficiency significantly. With a maximum of 2 and 3 HARQ retransmissions, the required average power is reduced by 13 and 17 dB, respectively.

Such hybrid links have the potential to be deployed in mm-wave LOS coverage enhancements with coordinated high-rise access points, which was the topic of the next section. In particular, LOS coverage is attractive in mm-wave networks, but since it is difficult to find enough positions to deploy high-rise access points in dense networks, we investigated the potential of a joint deployment of high-rise and low-rise APs. It was shown that only very
few high-rise APs are needed to enhance the LOS coverage when the blocking building height is low, whereas with increased blocking building height, more and more high-rise APs are needed to provide a targeted LOS coverage probability. The merits of multi-hopping were further analysed in the following section based on a system level evaluation at mm-waves. A relay-assisted system with decode-and-forward RSs was compared to a direct access system. Naturally, the cell-edge users benefit from the relay-assistance, but it was also shown that RSs at mm-wave can be very beneficial in improving system performance. In particular, for the exemplary ISD=500 m, the relay-assisted scheme provides a gain of about 120% in the cell border throughput, and the same sector spectral efficiency as for the direct scheme.

The chapter concludes with a study on joint hybrid precoding for energy-efficient mm-wave networks. The proposed algorithm provides a low complexity solution to obtain the hybrid precoders and to analyse the energy efficiency and load balance in hybrid precoding mm-wave networks. It was shown that, by jointly designing hybrid precoders, the total transmit powers can be reduced while satisfying the spectral efficiency constraint for each user. Compared to a fully digital beamforming the average total transmit power is naturally larger for the more constrained hybrid precoders, but the overall power consumption is lower for a given targeted rate for each user due to the fewer RF chains. We also showed that there are large gains in total energy efficiency with joint multi-node multi-stream transmission, and it increases with the number of cooperative BSs. In addition, the distribution of the power consumption over the cooperative nodes becomes more even, which might be useful for meeting potential EIRP restrictions of mm-wave nodes.

To summarize the key findings of cooperative and coordinated and multi-node scheme design towards a system concept:

- Multi-node cooperation and coordination is very useful for mm-wave networks, from both beam sweeping, throughput, outage, energy efficiency, load balancing and peak power reduction point of view (EIRP). Thus, it should be an integrated part of efficient mm-wave networks.
- Decode-and-forward relaying can be deployed to substantially improve the cell edge without sacrificing the spectral efficiency.
- Hybrid RF-FSO links have a large potential for backhauling, and multi-hop Hybrid RF-FSO connections in combination with a few high-rise APs have a great potential for efficient LoS coverage.
- Cooperative hybrid beamforming has a great potential to meet key performance targets at a lower cost and energy consumption than fully digital beamforming, and the gains are increasing with the number of nodes (disregarding backhaul overhead for the user data, which potentially could be supported with multi-cast in the backhaul, but there have been not enough resources to study that in the project).
- Sequential hybrid beamforming is a flexible way to support adaptive dual/multi-connectivity. In certain deployment scenarios, even a fully analog beamforming approach is competitive.
- In interference limited scenarios, due to the sparse multi-path and high potential for LoS channels, coverage can be increased in the uplink by increasing the total number of receive antennas, and use an optimal fraction of the antennas to cancel the nearest few strongest interferers.
- To support random access and mobility at mm-wave frequencies, a flexible toolbox for beam management is necessary.
4 Hardware impairments modelling and performance assessment

4.1 Introduction

Hardware impairments are related to imperfections and constraints, which include phase noise, I/Q-imbalance, sampling jitter, sampling frequency offset, carrier frequency offset, Power Amplifier (PA) nonlinearity, quantization noise, antennas impairments etc. Such impairments would cause distortion of the transmitted and received signal and lead to increased Error Vector Magnitude (EVM) and Bit Error Rate (BER) depending on the waveform. Besides, the impairment would also cause distortion on adjacent channels.

The mm-wave communication systems are severely hardware constrained due to physical limitations. This leads to a much larger need for relying on the array-gain in large antenna systems, which in turn adds on to the total system complexity. Further on, components and manufacturing methods so far are also less refined at mm-wave frequencies, compared to traditional carrier frequencies for mobile communications. This further implies that we need to choose hardware with reasonable cost, size and power consumption, which very often may be associated with strong impairments for the actual implementation. Compared to its lower frequency counterparts, the detrimental effects of RF and antenna impairments will be more pronounced in mm-wave frequencies. It is necessary to take them into account already at the system design stage.

In [MM MAGICD51] different models have been proposed for phase noise, power amplifier, I/Q imbalance, ADC and antennas, together with analysis on hardware impairment effect on system performance.

Here in section 4.2 we present an extensive study on antenna models at mm-wave for different frequency bands. Different array topologies are presented for terminal, access and backhaul.

Section 4.3 extends the studies on PA modelling for large arrays and introduces a stochastic model to describe the PA distortion for certain applications such as link- or network-level evaluations.

A phase noise model and its Matlab implementation was proposed for “high” and “low” noise conditions [MM MAGICD51]. Here an additional model is presented in section 4.4 together with an analysis of phase tracking reference signal. The impact of hardware impairments, such as phase noise and phase amplitude errors, was previously studied for hybrid beamforming schemes. The impact of phase noise on local oscillator distribution, according to different strategies, has been investigated in MIMO link. In section 4.5 the study has been extended to Multi-User MIMO OFDM.

Finally section 4.6 analyses the impact of the errors introduced by phase shifters and the losses added by the combiner stage in front of the antenna elements for Hybrid Beam Forming (HBF) architecture.

4.2 Mm-wave antennas design and models

In this section we present investigations on different multi-antenna configurations. Starting from antenna specifications (§ 4.2.1) we present mm-wave array designs based on patch elements (§ 4.2.2 and massive dipole arrays (§ 4.2.3). Finally, realistic transmit arrays are considered for access point (§ 4.2.4) and backhaul (§ 4.2.5) applications.

4.2.1 Antenna specifications

Mm-wave antenna design for handsets carry the usual problems of size and cost constraints and issues of antenna element coupling and blocking effects with the hands. However the
smaller inter-element distances make even planar arrays possible within handsets, to some degree. Typical terminal antenna specifications are presented in Table 4-1.

Table 4-1: Antenna specifications for the mm-wave user terminal aligned with practical deployments.

<table>
<thead>
<tr>
<th></th>
<th>X-band</th>
<th>K-, Ka-band</th>
<th>V- or E-band</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Array topology</strong></td>
<td>Linear array</td>
<td>Linear array</td>
<td>Linear/Square array</td>
</tr>
<tr>
<td><strong>Typ. Frequency bands</strong></td>
<td>14.4 – 15.35 GHz</td>
<td>24.25 – 27.5 GHz</td>
<td>71-76 or 81 – 86 GHz</td>
</tr>
<tr>
<td><strong>Typ. Gain (dBi)</strong></td>
<td>10 – 15</td>
<td>15</td>
<td>15 – 20</td>
</tr>
<tr>
<td><strong>Typ. minimum size¹</strong></td>
<td>1x4 (11 dBi Gain)</td>
<td>1x8 (15 dBi Gain)</td>
<td>4x4 (17 dBi Gain)</td>
</tr>
<tr>
<td><strong>Polarization</strong></td>
<td>Linear/Dual linear</td>
<td>Linear/Dual linear</td>
<td>Linear/Dual linear</td>
</tr>
<tr>
<td><strong>Beam-steering</strong></td>
<td>±40° (on one plane)</td>
<td>±40° (on one plane)</td>
<td>±40° (on one plane)</td>
</tr>
<tr>
<td><strong>Beam-forming</strong></td>
<td>Digital/Hybrid</td>
<td>Digital/Hybrid</td>
<td>Analog/Hybrid</td>
</tr>
</tbody>
</table>

5G base stations includes two functionalities: the radio access link, which guarantee the bidirectional link between users and access point and the backhaul/fronthaul link connecting one access point to the core-network or to a common baseband unit. Typical antenna specifications for both applications are presented in Table 4-2 and Table 4-3. In general, for radio access high gain (> 20 dBi) antennas with analog or hybrid beamforming capability are needed to manage both multi-users and mobility. In the case of the backhaul/fronthaul link, a gain > 30 dBi and fixed beam or a limited scanning capability (±10° on one plane) are required. In fact, for this kind of link, beam-steering could be used to implement a self-alignment function. Other practical constraints for the access point antennas are the limited antenna size, cost and complexity. It is important to note that the typical array size given are for ideal lossless arrays. In practice, as shown in the following sections, this depends on the antenna efficiency, and the number of antennas could be increased to reach the desired gain.

Table 4-2: Antenna specifications for the mm-wave access point aligned with practical deployments.

<table>
<thead>
<tr>
<th></th>
<th>X-band</th>
<th>K-, Ka-band</th>
<th>V- or E-band</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Array topology</strong></td>
<td>Square array</td>
<td>Square array</td>
<td>Square array</td>
</tr>
<tr>
<td><strong>Typ. Frequency bands</strong></td>
<td>14.4 – 15.35 GHz</td>
<td>24.25 – 27.5 GHz</td>
<td>71-76 or 81 – 86 GHz</td>
</tr>
<tr>
<td><strong>Typ. Gain (dBi)</strong></td>
<td>15 – 20</td>
<td>20 – 25</td>
<td>32</td>
</tr>
<tr>
<td><strong>Typ. minimum size¹</strong></td>
<td>4x4 (17 dBi Gain)</td>
<td>6x6 (20 dBi Gain)</td>
<td>20x20 (31 dBi Gain)</td>
</tr>
<tr>
<td><strong>Polarization</strong></td>
<td>Linear/Dual linear</td>
<td>Linear/Dual linear</td>
<td>Linear/Dual linear</td>
</tr>
<tr>
<td><strong>Beam-steering</strong></td>
<td>±60° (2D window)</td>
<td>±60° (2D window)</td>
<td>±60° (2D window)</td>
</tr>
<tr>
<td><strong>Beam-forming</strong></td>
<td>Digital/Hybrid</td>
<td>Digital/Hybrid</td>
<td>Analog/Hybrid</td>
</tr>
</tbody>
</table>

Table 4-3: Antenna specifications for the mm-wave backhauling/fronthauling aligned with practical deployments.

¹ Ideal lossless case
### 4.2.2 Patch antenna arrays for mm-wave user terminal and access point

Five linearly-polarized antenna array configurations, respectively with $1\times4$, $2\times4$, $4\times4$, $1\times8$, and $8\times8$ identical elements, based on a rectangular patch element have been designed and simulated using the commercial software CST Microwave Studio to extract the full scattering matrix, the single element radiation patterns including the coupling effects, and the full-array radiation patterns computed as a function of the scanning angle in a spatial window of $\pm 60^\circ$. The proposed design is based on a simple architecture (two metal layers, a probe, and a dielectric substrate) and could be selected as a solution for the mm-wave user terminal or access point antennas. Digital, hybrid, or analog beamforming are supported by opportunely designing the phase-shifts, the feed networks, and the transceiver architecture. Dual-polarization can also be easily implemented by including a second radio-frequency access (probe) to excite the orthogonal mode of the rectangular patch.

The proposed single element (Figure 4-1(a)) has been designed considering standard printed circuit board (PCB) technology and optimized in the band $24.25 - 27.5$ GHz considering a reflection coefficient $< -6$ dBi on the desired frequency band. The commercial high-performance dielectric substrate Rogers RT/Duroid 6002 ($\varepsilon_r = 2.94$, $\tan \delta = 0.0012$, and thickness $762$ µm) has been selected for this design. A rectangular microstrip patch antenna is printed on the top layer of the dielectric substrate, it is fed by using a metallized via. A hole is realized in the ground plane printed on the bottom layer of the substrate to permit the connection to the connectors or feed lines to the probe. All the geometrical characteristics of the single element patch antenna are presented in Figure 4-1(a). The simulated reflection coefficient and realized gain patterns computed at the central frequency and as a function of the cut-plane are presented in Figure 4-1(b), Figure 4-1(c) and Figure 4-1(d), respectively.

The optimized rectangular patch element was used to implement the five array geometries, which have been fully simulated to extract all the antenna parameters. All the simulated parameters have been extracted to create a database and shared with the consortium to implement generic beamforming functions (digital, analog, or hybrid). The array geometry in the case of the $1\times4$-element array and its scattering matrix and active radiation patterns (realized gain including the losses in the dielectric substrate, in the conductor and due to the antenna mismatching) are presented in Figure 4-2.
Figure 4-1: Single rectangular patch element. (a) Geometry, (b) simulated reflection coefficient, and simulated (c) co- and (d) cross-polar components of the realized gain patterns computed at the central frequency.

Figure 4-2: 1×4-element linear array based on rectangular patch elements. (a) Geometry, (b) simulated scattering matrix parameters, and simulated co-polar components of the realized gain active patterns computed at the central frequency on the two principal planes (phi = 0° (c) and phi = 90° (d)).
Figure 4-3: Simulated co-polar components of the realized gain patterns computed at the central frequency on the plane phi = 0° at 25 GHz. (a) 1×4-, (b) 4×4-, (c) 1×8-, and (d) 8×8-element linear arrays based on rectangular patch elements.

The simulated realized gain radiation patterns computed as a function of the scanning angle when the phase coefficient associated to each array element have also been calculated and the data included in the antenna databases. The results at the central frequency are presented in Figure 4-3 and synthetized in the Tables 4-4 to 4-8 below.

Table 4-4: Realized gain of the 1×4-element array based on rectangular patch at the central frequency.

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Max gain scan angle (deg)</th>
<th>Max gain (dBi)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>11.8</td>
<td>11.8</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>11.7</td>
<td>11.7</td>
</tr>
<tr>
<td>20</td>
<td>20</td>
<td>11.6</td>
<td>11.6</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
<td>11.4</td>
<td>11.4</td>
</tr>
<tr>
<td>40</td>
<td>35</td>
<td>11.0</td>
<td>10.6</td>
</tr>
<tr>
<td>50</td>
<td>40</td>
<td>10.1</td>
<td>9.15</td>
</tr>
<tr>
<td>60</td>
<td>45</td>
<td>9.00</td>
<td>6.96</td>
</tr>
</tbody>
</table>

Table 4-5: Realized gain of the 2×4-element array based on rectangular patch at the central frequency.

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Max gain scan angle (deg)</th>
<th>Max gain (dBi)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>13.5</td>
<td>13.5</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>13.5</td>
<td>13.5</td>
</tr>
<tr>
<td>20</td>
<td>20</td>
<td>13.3</td>
<td>13.3</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
<td>12.9</td>
<td>12.9</td>
</tr>
<tr>
<td>40</td>
<td>35</td>
<td>12.4</td>
<td>12.2</td>
</tr>
<tr>
<td>50</td>
<td>45</td>
<td>11.5</td>
<td>11.0</td>
</tr>
<tr>
<td>60</td>
<td>50</td>
<td>10.4</td>
<td>8.98</td>
</tr>
</tbody>
</table>
Table 4-6: Realized gain of the 4×4-element array based on rectangular patch at the central frequency.

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Max gain scan angle (deg)</th>
<th>Max gain (dBi)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>16.3</td>
<td>16.3</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>16.2</td>
<td>16.2</td>
</tr>
<tr>
<td>20</td>
<td>20</td>
<td>16.0</td>
<td>16.0</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
<td>15.6</td>
<td>15.6</td>
</tr>
<tr>
<td>40</td>
<td>35</td>
<td>15.1</td>
<td>15.0</td>
</tr>
<tr>
<td>50</td>
<td>45</td>
<td>14.2</td>
<td>13.7</td>
</tr>
<tr>
<td>60</td>
<td>50</td>
<td>13.1</td>
<td>11.8</td>
</tr>
</tbody>
</table>

Table 4-7: Realized gain of the 1×8-element array based on rectangular patch at the central frequency.

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Max gain scan angle (deg)</th>
<th>Max gain (dBi)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>14.7</td>
<td>14.7</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>14.6</td>
<td>14.6</td>
</tr>
<tr>
<td>20</td>
<td>20</td>
<td>14.5</td>
<td>14.5</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
<td>14.3</td>
<td>14.3</td>
</tr>
<tr>
<td>40</td>
<td>40</td>
<td>13.7</td>
<td>13.7</td>
</tr>
<tr>
<td>50</td>
<td>45</td>
<td>12.5</td>
<td>12.3</td>
</tr>
<tr>
<td>60</td>
<td>55</td>
<td>10.5</td>
<td>9.42</td>
</tr>
</tbody>
</table>

Table 4-8: Realized gain of the 8×8-element array based on rectangular patch at the central frequency.

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Max gain scan angle (deg)</th>
<th>Max gain (dBi)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>22.6</td>
<td>22.6</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>22.5</td>
<td>22.5</td>
</tr>
<tr>
<td>20</td>
<td>20</td>
<td>22.2</td>
<td>22.2</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
<td>21.7</td>
<td>21.7</td>
</tr>
<tr>
<td>40</td>
<td>40</td>
<td>20.9</td>
<td>20.9</td>
</tr>
<tr>
<td>50</td>
<td>50</td>
<td>19.8</td>
<td>19.8</td>
</tr>
<tr>
<td>60</td>
<td>55</td>
<td>18.4</td>
<td>17.9</td>
</tr>
</tbody>
</table>

4.2.3 Massive dipole arrays

Considering the frequency bandwidth (24.25-27.5 GHz), the single element used for this study is a classical printed dipole on a ground plane. This element works in dual polarization thanks to the crossed-dipole configuration in Figure 4-4 with a return loss below -10 dB.
To be able to scan the beam up to 60° without grating lobes, the array step should be around 0.5 \( \lambda \). Taking into account the size of the dipole equivalent to the spacing, The configuration ±45° polarization is more suited thanks to the setting-up of the dipoles which generates less coupling as we can see it for a 1x4 array on the Figure 4-5 where the input impedances of this ±45° configuration remain unchanged with regard to the single element represents Figure 4-4 contrary to those of the HV configuration.

**Figure 4-4**: Single and crossed dipole: geometry, return loss and gain pattern.

**Figure 4-5**: Input impedance for HV polarization (up) and ±45° polarization (down).
Several configurations (1x4, 2x4, 4x4, 1x8, 2x8, 4x8 and 8x8) have been simulated (Figure 4-6).

Figure 4-6: Radiation patterns for several array configurations: geometry and gain pattern.

The gain values for the different configurations are summarized in the Table 4-9 for low scanned angle and for the higher value (60°).
Table 4-9: Array antenna gains.

<table>
<thead>
<tr>
<th>Array configuration</th>
<th>1x4</th>
<th>2x4</th>
<th>4x4</th>
<th>1x8</th>
<th>2x8</th>
<th>4x8</th>
<th>8x8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain (dBi) for scanned angle &lt; 30°</td>
<td>13.0</td>
<td>14.3</td>
<td>16.9</td>
<td>15.7</td>
<td>17.0</td>
<td>19.7</td>
<td>22.6</td>
</tr>
<tr>
<td>Gain (dBi) for scanned angle = 60°</td>
<td>10.0</td>
<td>11.0</td>
<td>12.8</td>
<td>10.6</td>
<td>12.0</td>
<td>14.7</td>
<td>17.7</td>
</tr>
</tbody>
</table>

Equivalent gain can be achieved with a reduced number of elements (6x6 instead of 8x8) with a larger spacing (0.7 λ) at the cost of a reduced scanned angle of 25°. This limit where grating lobes appear could be enlarged by destroying the periodicity of the array. Figure 4-7 shows the radiation patterns for two configurations of a 6x6 array for a 40° scanned angle. This lower number of elements could be interesting to reduce the cost for moderate scanned angle needs.

Figure 4-7: 6x6 array: periodic (up) and non-periodic in the horizontal plane (down).

Taking into account the step size of the array and the need to use connectors to feed each element of the breadboards, it is impossible to place two connectors on the crossed dipole, so only a single polarization element with compact connector is used for the array antennas for the User Equipment (2 arrays of 1x4 dipoles) and for an Access Point antenna (2x8 dipoles). To keep lower level of coupling between elements, the breadboards have a "45° polarization" (Figure 4-8).

Figure 4-8: Single polarization element with connector, UE and AP configurations.
A first iteration has been done to evaluate the antenna performances dispersion due to the manufacturing. The two elements of the ±45° crossed dipoles have been manufactured as two single polarization elements on a 50x50 mm ground plane. For the simulations, only the compact connector (SMP type) has been included in the structure to analyse an additional adapter (SMP to K) is used for the experiments. The two antennas have been assembled and tested (Figure 4-9).

![Antenna assemblies](image)

**Figure 4-9:** input impedance measurement of the dipole -45° (left) and radiation pattern measurement of the dipole+45° (right)

The return losses for the two dipoles are presented Figure 4-10. There is a difference between the simulated and measured input impedances for the two dipoles due to some defects of misalignment during the realization (connectors/microstripline) and to the additional adapter.

![Graphs](image)

**Figure 4-10:** input impedance (dB) for dipole +45° (left) and -45° (right)

The radiation patterns have been presented Figure 4-11 for the dipole +45° and Figure 4-12 for the dipole +45°. The radiation patterns are plotted total filed and in a two dimensional representation where the central point corresponds to the vertical direction normal to the ground plane and the external circle to the θ = 90° direction (horizontal plane). The simulated and measured radiation patterns are comparable for the two dipoles, the measured and simulated directivities for the dipole +45° and -45° are respectively: 8.3 dB, 8.28 dB, 8.47 dB and 8.41 dB.
4.2.4 Transmitarray antennas for access point

In this section, electronically reconfigurable Transmitarrays (Figure 4-13(a)) for mm-wave access points operating in the K-band (24.25 – 27.5 GHz) are presented and analyzed. The proposed Transmitarray performance has been extrapolated starting from the previously developed linearly-polarized electronically reconfigurable unit-cells (Figure 4-13(b)) with 1-bit of phase compensation [DCD+16],[DCD+17]. The unit-cell details and results have been presented in [DCD+16] and the possibility to develop a full electronically reconfigurable Transmitarray with analog beamforming capability has been experimentally demonstrated in [DCD+17]. The unit-cell architecture presented in Figure 4-13(b) is based on four metal layers printed on two dielectric substrates. Two p-i-n diodes have been integrated on the unit-cell to electronically control the transmission phases and implement the 1-bit phase quantization. The p-i-n diodes are biased by using a current in the range of 2 – 10 mA, corresponding to an insertion loss in the range 2.1 – 1.2 dB. 1-bit of phase quantization has been selected in order to limit the full array power consumption, reduce the unit-cell complexity, insertion loss, and cost, and increase the transmission bandwidth. The realized unit-cell operates in the band 27 – 30.2 GHz (3-dB relative transmission bandwidth equal to 14.6%) with insertion loss between 1.09 dB and 1.29 dB when ±10 mA are used to bias the p-i-n diodes [DCD+16]. The full Transmitarray performance has been extracted by using an in-house optimization tool previously developed and demonstrated at CEA-Leti. Considering the schematic view presented in Figure 4-13(a), the full Transmitarray power budget is estimated starting from the input power on the focal source P1 and considering the power terms P2 (radiated power by the focal source), P3 (absorbed power on the receiver side of the Transmitarray considering reflection and spill-over losses), P4 (transmitted power on the transmission layer of the Transmitarray considering insertion, conductor end dielectric losses), and the Transmitarray radiated power P5.
Two different electronically reconfigurable Transmitarrays with analog beamforming capability, a 20×20 unit-cells (UC) with 1-bit of phase quantization and a 14×14 unit-cells with 2-bits of phase quantization, have been analyzed and their performance extracted. The 2-bit unit-cell performance has been extrapolated starting from the experimentally validated 1-bit design. The fundamental parameters and characteristics of the two Transmitarrays, considering the 1-bit and 2-bit UCs, are presented in Table 4-10. The synthesis of the beamforming performances have been presented in Table 4-11 and in Table 4-12 in the case of the electronically reconfigurable 1-bit Transmitarray and of the 2-bit Transmitarray, respectively.

The phase shift on the array aperture is optimized to focus the beam in a desired direction and electronically controlled by tuning the p-i-n diodes integrated on the unit-cells. Also in this case the realized gain patterns have been extracted as a function of the frequency and the scan angle and included in a database, which has been shared with the consortium.

![Figure 4-13](image)

Figure 4-13: (a) Schematic view of the electronically reconfigurable Transmitarray. (b) 1-bit linearly-polarized electronically reconfigurable unit-cell architecture [DCD+16].

### Table 4-10: Characteristics of the electronically reconfigurable Transmitarray for AP

<table>
<thead>
<tr>
<th>Parameter</th>
<th>1-bit</th>
<th>2-bit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of UC</td>
<td>20×20</td>
<td>14×14</td>
</tr>
<tr>
<td>UC size (λ₀²)</td>
<td>0.5×0.5</td>
<td>0.5×0.5</td>
</tr>
<tr>
<td>Array size (λ₀²)</td>
<td>10×10</td>
<td>7×7</td>
</tr>
<tr>
<td>Number of phase states</td>
<td>2</td>
<td>4</td>
</tr>
<tr>
<td>Relative phase-shift</td>
<td>180°</td>
<td>90°</td>
</tr>
<tr>
<td>Number of p-i-n diodes</td>
<td>800</td>
<td>784</td>
</tr>
<tr>
<td>Focal distance (λ₀)</td>
<td>6</td>
<td>4.5</td>
</tr>
<tr>
<td>FS gain at f₀ (dBi)</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>Gain at f₀ (dBi)</td>
<td>23.7</td>
<td>23.4</td>
</tr>
<tr>
<td>Total loss at f₀ (dB)</td>
<td>2.5</td>
<td>2.9</td>
</tr>
</tbody>
</table>
The realized gain radiation patterns computed as a function of the scanning angle and the frequency responses in the case of the 1- and of the 2-bit electronically reconfigurable Transmitarrays have been plotted in Figure 4-14 and Figure 4-15, respectively. In Figure 4-15, the frequency axis has been normalized to the central frequency \( f_0 \) in order to have a general representation independent from the frequency. In the case of the K-band the lower frequency (24.25 GHz) is equal to 0.94 and the higher frequency (27.5 GHz) is equal to 1.06.

Table 4-11: Realized gain of the 20×20 1-bit Transmitarray computed at the central frequency \( f_0 \).

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>23.9</td>
</tr>
<tr>
<td>10</td>
<td>23.6</td>
</tr>
<tr>
<td>20</td>
<td>22.9</td>
</tr>
<tr>
<td>30</td>
<td>22.2</td>
</tr>
<tr>
<td>40</td>
<td>20.8</td>
</tr>
<tr>
<td>50</td>
<td>20.2</td>
</tr>
<tr>
<td>60</td>
<td>18.3</td>
</tr>
</tbody>
</table>

Table 4-12: Realized gain of the 14×14 2-bit Transmitarray computed at the central frequency \( f_0 \).

<table>
<thead>
<tr>
<th>Desired scan angle (deg)</th>
<th>Gain at desired scan angle (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>23.1</td>
</tr>
<tr>
<td>10</td>
<td>22.8</td>
</tr>
<tr>
<td>20</td>
<td>22.5</td>
</tr>
<tr>
<td>30</td>
<td>21.3</td>
</tr>
<tr>
<td>40</td>
<td>20.3</td>
</tr>
<tr>
<td>50</td>
<td>18.7</td>
</tr>
<tr>
<td>60</td>
<td>17.2</td>
</tr>
</tbody>
</table>

Figure 4-14: Simulated realized gain computed at the central frequency \( f_0 \) as a function of the scanning angle. (a) 1-bit electronically reconfigurable Transmitarray and (b) 2-bit electronically reconfigurable Transmitarray.

Figure 4-15: Simulated realized gain of the 1- and the 2-bit electronically reconfigurable Transmitarrays.
The realized optimized phase distributions on the array aperture computed as a function of the scanning angle (0° and 20°) in the case of the 1- and of the 2-bit electronically reconfigurable Transmittarrays have been plotted in Figure 4-16.

![Figure 4-16: Optimized phase distribution on the array aperture. 1-bit electronically reconfigurable Transmittarray for scan angle (a) 0° and (b) 20°. 2-bit electronically reconfigurable Transmittarray for scan angle (c) 0° and (d) 20°.](image)

4.2.5 Transmittarray antennas for backhauling/fronthauling

In the case of backhaul/fronthaul applications at K-band (24.25 – 27.5 GHz), the analyzed Transmittarrays are based on a passive unit-cell with 3-bit phase quantization. In general, for backhaul/fronthaul applications, a gain > 30 dBi and fixed beam or a limited scanning capability (±10° on one plane) are required. Here, beam-steering is used to implement the self-alignment function. The architecture of the unit-cells used in the study is presented in Figure 4-17 and is based on three metal layers (top patch, ground plane, and bottom patch) printed on two dielectric substrates. Four different architectures have been used to implement the required phase states. As presented in [JCH+17],[DCD+17b] for each architecture a 1-bit unit-cell is designed by physically rotating one patch of an angle equal to 180° around the metallized vias used to connect the patch antennas printed on the top and bottom layer of the dielectric structure.

![Figure 4-17: Schematic view of the 3-bit passive unit-cells in K-band.](image)

The unit-cells have been used to implement three 40×40 fixed-beam Transmittarrays with 1-, 2- or 3-bit phase quantization. The frequency response of the arrays is presented in Figure 4-18 and their fundamental characteristics are synthetized in Table 4-13. The simulated
radiation patterns of the three Transmitarrays computed at the central frequency \( f_0 \) are presented in Figure 4-19 and compared to the ETSI requirements for point-to-point applications in K-band [ETSI10-302217].

![Simulated radiation patterns of fixed beam Transmitarray for backhaul/fronthaul computed at the central frequency as a function of the phase quantization.](image)

**Figure 4-19**: Simulated radiation patterns of fixed beam Transmitarray for backhaul/fronthaul computed at the central frequency as a function of the phase quantization.

**Table 4-13**: Characteristics of the fixed-beam Transmitarray for mm-wave backhauling/fronthauling.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>1-bit</th>
<th>2-bit</th>
<th>3-bit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of UC</td>
<td>40×40</td>
<td>40×40</td>
<td>40×40</td>
</tr>
<tr>
<td>UC size ( \lambda_0^2 )</td>
<td>0.5×0.5</td>
<td>0.5×0.5</td>
<td>0.5×0.5</td>
</tr>
<tr>
<td>Array size ( \lambda_0^2 )</td>
<td>20×20</td>
<td>20×20</td>
<td>20×20</td>
</tr>
<tr>
<td>Number of phase states</td>
<td>2</td>
<td>4</td>
<td>8</td>
</tr>
<tr>
<td>Relative phase-shift</td>
<td>180°</td>
<td>90°</td>
<td>45°</td>
</tr>
<tr>
<td>Focal distance ( \lambda_0 )</td>
<td>12.2</td>
<td>12.2</td>
<td>12.2</td>
</tr>
<tr>
<td>FS gain at ( f_0 ) (dBi)</td>
<td>11.5</td>
<td>11.5</td>
<td>11.5</td>
</tr>
<tr>
<td>Gain at ( f_0 ) (dBi)</td>
<td>30.0</td>
<td>32.8</td>
<td>33.6</td>
</tr>
<tr>
<td>Total loss at ( f_0 ) (dB)</td>
<td>1.8</td>
<td>1.9</td>
<td>1.8</td>
</tr>
</tbody>
</table>

Switched-beam capability could be easily implanted by using multi focal source architecture as presented in [MDS+16].
4.3 Power amplifier modelling

We will describe the memory-polynomial based framework for modeling power amplifiers in an active array under a non-constant load impedance. Further on, an improved stochastic power amplifier model suitable for system-level simulations is discussed.

4.3.1 Behavioural modelling

In general, the area of power amplifier modelling is rather mature relying on decades of research in non-linear systems models. The Volterra-series framework, [Sch80], have provided a solid ground for power amplifier modelling from which specializations such as the generalized memory polynomial (GMP) [MMK+16], has arisen.

Large antenna arrays are not only beneficial, but rather necessary to provide sufficient link-budget. When introducing power amplifiers in environments such as large, dense antenna arrays, they become subject to mutual coupling which introduces a new source of distortion not covered in the regular Volterra-based framework. For this reason, development in the field of power amplifier modelling for antenna arrays has flourished the last couple of years with some progress based on a GMP basis, in which the mutual coupling effects are modelled via a secondary variable [FBH+14, HGS+17].

Using the new modelling framework, we may examine the behaviour of power amplifier distortion in active arrays. As shown in Figure 4-20, the mutual coupling introduces a steering-angle dependent distortion behaviour. For massive MU-MIMO, this behaviour takes on a more random behaviour as the number of users increases, which makes the distortion more independent from the transmit signal, [GSE+14].

![Adj channel EIRP of a 8x8 element phased array](image)

**Figure 4-20** EIRP of the adjacent channel distortion of an 8x8 element phased array, sweeping one beam in both azimuth and elevation.

4.3.2 Statistical modelling

For certain applications such as link- or network-level evaluations, it may be practical to limit the modelling to knowledge about the statistical properties to the power amplifier distortion.
One of the simplest approaches is to model the impairments as a static, multiplicative gain and phase error, which allows for simple approximate SINR evaluation through analytical manipulation, [ADG15]. A more involved approach is based upon computing the distortion covariance matrix, [MGE+16]. The entries in the distorted transmit covariance matrix can be developed from the corresponding behavioral model. We will exemplify this using a single-antenna system here. Using a simple, static third order polynomial, we write the output $y$ as a function of the input $x$ as

$$y = \theta_1 x + \theta_2 x|x|^2$$

in which $\theta_n$ are the parameters. Provided that the input signal is Gaussian, which in the case of most multi-user or multi-carrier scenarios due to the central limit theorem, we can use the Bussgang theorem to decompose the model into a first order stochastic approximation which is written as

$$y = ax + w$$

The parameters are defined and computed for the third order polynomial case as

$$\alpha = \frac{E[y'x]}{\sigma_x^2} = \theta_1 + 2\theta_2 \sigma_x^2$$

$$\sigma_w^2 = E[(y - ax)^2] = 2|\theta_2|^2(3\sigma_x^6 + 2\sigma_x^8)$$

It is interesting to notice that the distortion term 1) does not depend on the linear term in the model, and 2) grows cubically with transmit power, $\sigma_x^2$.

In the multi-antenna case, these parameters must be computed across the array for which the corresponding multi-variate model can be formulated as

$$Y = \Lambda X + W$$

where $X \sim \text{CN}(0, C_{xx})$, $W \sim \text{CN}(0, C_{ww})$ and $\Lambda = \text{diag}(\alpha_1, \ldots, \alpha_M)$. To evaluate the spatial consistency of the statistical model, simulations were performed using an 8x1-antenna ULA. An OFDM signal consisting of 600 subcarriers was frequency-selectively pre-coded into two angle-of-departures (AoD). The estimated spatial behaviour of the error signal of the statistical model fits well with the corresponding behavioural model part, as illustrated in Figure 4-21.

\begin{figure}[h]
\centering
\begin{subfigure}{0.3	extwidth}
\includegraphics[width=\textwidth]{transmit_signal_PSD_vs_AoD.png}
\caption{Transmit signal PSD vs. AoD}
\end{subfigure}\hfil
\begin{subfigure}{0.3	extwidth}
\includegraphics[width=\textwidth]{error_signal_PSD_vs_AoD_behavioral_model.png}
\caption{Error signal PSD vs. AoD using a behavioural model}
\end{subfigure}\hfil
\begin{subfigure}{0.3	extwidth}
\includegraphics[width=\textwidth]{error_signal_PSD_vs_AoD_statistical_model.png}
\caption{Error signal PSD vs. AoD using the statistical model}
\end{subfigure}
\caption{Figure 4-21 (a) Transmit signal PSD vs. AoD, (b) Error signal PSD vs. AoD using a behavioural model, (c) Error signal PSD vs. AoD using the statistical model.}
\end{figure}

### 4.3.3 Scaling laws for energy efficiency analysis

The power consumption of key radio hardware components is heavily dependent on performance parameters as well as bandwidth and operating frequency. Here, we present a
brief analysis of published scaling laws which are necessary to identify performance vs. power consumption trade-offs.

**Power amplifiers: Output power and efficiency**

The output power capability for any given semiconductor technology is primarily proportional to the product between the electric breakdown field $E$ and the charge carrier velocity $v_s$.

$$ P \propto \frac{1}{f^2} \left( \frac{E v_s}{2 \pi} \right)^2 $$  \hspace{1cm} (4-6)

It is further inversely proportional to the squared transition frequency, which implies that the effort needed to generate transmit power at mm-wave frequency increases in a quadratic manner. Generally, this also impacts the available transducer gain which, combined with increased impact of the skin effect due to the small geometries, further hampers the potential power efficiency compared to low frequency amplifiers.

**Data-converters**

For data-converters, and mainly A/D converters, a well-established figure of merit is the Schreier FOM. It tells us the relation between key parameters such as dynamic range (DR), bandwidth (BW) and consumed power (P).

$$ \text{FOM} = (\text{DR})_{db} + 10 \log_{10} \left( \frac{\text{BW}}{P} \right) $$  \hspace{1cm} (4-7)

Here, we can make two important observations. For a given converter FOM:

- 3 dB DR has the same power cost as doubling the BW.
- Adding one ENOB (Effective Number of Bits) quadruples the power

For the case of D/A converters, the same scaling applies.

**Local oscillators**

For local oscillators, the well-known Leeson’s figure of merit is commonly used. Ignoring the 1/f noise, the proportionality boils down to

$$ L(df) \propto \frac{P_{DC}}{df^2} $$  \hspace{1cm} (4-8)

in which $P_{DC}$ is the power consumption. We see that for any given oscillator quality, $L(df)$, the power consumption follows the square of the frequency.

**General trends**

As we may observe in the previous sections, keeping the performance up while moving toward mm-wave frequency and very large bandwidth, is quite costly in terms of power consumption. These limitations should be taken into consideration to a large extent compared to when designing mm-wave systems, than what has been done traditionally in sub-6 GHz frequency bands.
4.4 A simple and effective Phase Noise model and design of PTRS

A detailed Phase Noise (PN) model from the work of WP5 was introduced in D5.1. This PN model was shared with the wider research community as an open source code. A simpler, yet effective PN model was also studied in the later stages of WP5 work and it is introduced herewith. The model is easily adaptable for the analysis of CPE (Common Phase Error) in practical systems and we also present some of the related results in this section.

4.4.1 The multi-pole/zero PN model

The presented multi-pole/zero model is an extension to the single pole/zero model adapted for IEEE P802.15 [IEEE-06]. With a few, carefully chosen poles and zeros, it was found that the phase noise spectra of practical oscillators can be effectively matched. The power spectral density (PSD) behaviour of the proposed model is given by the following equation:

$$S(f) = PSD_0 \prod_{n=1}^{N} \left( 1 + \left( \frac{f}{f_{p,n}} \right)^2 \right) \left( 1 + \left( \frac{f}{f_{z,n}} \right)^2 \right)^{-1}$$

(4-9)

The model gives the following advantages:

- Practical phase noise power spectra can be well approximated with a few pole/zeros, as it gives more flexibility than a single pole/zero model. The challenge is to identify the correct poles and zeros to suit to a practical oscillator/ frequency synthesizer.

- Provides an easy framework to convert the PSD of analog phase noise to that of discrete-time phase noise (i.e., baseband version) for simulation by using the bilinear transform with given pole/zeros. This is transforming the s-domain multi-pole/zero function to the z-domain.

Table 4-14 shows two parameter sets which are obtained from practical oscillators operating at 30GHz and 60GHz, respectively. We call them “Set-A” and “Set-B” for simplicity.

<table>
<thead>
<tr>
<th>Parameter Set-A</th>
<th>Parameter Set-B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Carrier frequency ($f_{c,base}$)</td>
<td>30 GHz</td>
</tr>
<tr>
<td>PSD0 (dBc/Hz)</td>
<td>-79.4</td>
</tr>
<tr>
<td>Fp (MHz)</td>
<td>[0.1, 0.2, 8]</td>
</tr>
<tr>
<td>Fz (MHz)</td>
<td>[1.8, 2.2, 40]</td>
</tr>
</tbody>
</table>

Figure 4-22 shows the power spectral densities (PSD) for 3 carrier frequencies with these parameters. If the operating carrier frequency is changed, the PSD is shifted by $20 \log_{10} \left( \frac{f_c}{f_{c,base}} \right)$ dBc/Hz.
4.4.2 Analysis of Phase Tracking Reference Signal (PTRS) design

PTRS design is an active topic in 3GPP NR standardisation work, which commenced in mid-2016 [3GPP-16]. The PTRS are known pilot symbols, inserted into the radio sub-frame at the transmitter, so the receiver can correct the Common Phase Error (CPE). The CPE is a main component of PN and it rotates the symbol constellations across all sub-carriers in an OFDM system by an equal amount. CPE will cause errors in higher order modulation schemes (64QAM and above) as the constellations are closely packed. There is an obvious trade-off in the allowable density of PTRS, w.r.t the acceptable BLER (block error rate) and the pilot overheads.

Using the above PN model, the CPE was synthesized in an OFDM based 5G NR transmission and the performance for different PTRS densities was evaluated. In 3GPP NR terminology the Physical Resource Block (PRB) occupies 12 sub-carriers in the frequency domain and 7 symbols in the time domain (as in LTE). PRB is the basic unit of resource allocation to the user and multiple PRBs can be allocated, depending on the data rate of the supported user. Assuming that the user is allocated 100 PRBs (equivalent to 20MHz BW) the performance of different PTRS densities is illustrated in Figure 4-23. The performance of similar frequency domain PTRS densities but when the user is having different PRB allocations is shown in Figure 4-24. The terms SCS refer to sub-carrier spacing and TDLC is the TDL (Tapped Delay Line) variant of the above 6 GHz channel model developed by 3GPP.

The results in Figure 4-23 show that the BLER performance degrades when both the frequency domain (PTRS in sub-carrier per 4 or 16 PRBs) and time domain (PTRS in each or each 2 or each 4 symbols) densities are reduced. There is a bigger performance gap from 1 to 2 symbol densities than 2 to 4 symbol densities in both plots. In Figure 4-24, the performance is less sensitive to time domain PTRS density when the user is allocated a narrower bandwidth (fewer PRBs). The overall results indicate that the time domain PTRS density needs to be virtually every symbol, but in the frequency domain, PTRS can be in a sub-carrier per every few PRBs. Also for fewer PRB allocations, the PTRS densities can be reduced by increasing the time gap thus reducing the overheads. This work was also submitted to 3GPP RAN1 [3GPP-RAN1-17] and discussed in the meeting.
4.5 Effect of Phase Noise on Uplink Multi-User MIMO-OFDM

For simplicity, we assume $K$ users (each equipped with a single antenna along with a free-running oscillator) and $M (> K)$ antennas at the BS. The BS uses zero-forcing (ZF) decoder for multi-user detection. In order to focus on the phase noise (PN) effect, we assume perfect channel estimation and that the cyclic prefix (CP) of the OFDM symbol is longer than the channel length. Let $N$ be the number of OFDM subcarriers, $F$ be an $N \times N$ discrete Fourier transform (DFT) matrix, whose elements are given by $\exp \left( -j \frac{2\pi kl}{N} \right) \sqrt{N}$ ($k, l = 0, \ldots, N-1$), and $\phi_i$ and $\theta$ be $N \times 1$ vectors containing the time-domain PNs during one OFDM symbol at the $i$th user ($i = 1, \ldots, K$) and at the BS, respectively. The frequency-domain expression of the UL transmission (from $K$ users to the BS) in the presence of PNs is given as

$$ y = (G_R \otimes I_M) H G x + w, \quad (4.10) $$

where $H$ is a $MN \times KN$ block diagonal channel matrix whose $k$th diagonal block entry $H_k$ is the $M \times K$ channel transfer functions (CTFs) at the $k$th subcarrier, $x = [x_0^T \ x_1^T \ \cdots \ \sim^T \sim]_\sim$ is the $KN \times 1$ signal vector with $x_i$ denoting the $K \times 1$ transmitted signal vector (from all the users) at the $k$th subcarrier, $y = [y_0^T \ y_1^T \ \cdots \ \sim^T \sim]$ is the $MN \times 1$ signal vector with $y_i$ denoting the $M \times 1$ received signal vector at the $k$th subcarrier, $w$ is a $MN \times 1$ additive white Gaussian noise (AWGN)
vector, $G_R = F \text{diag}\left(\exp(j\theta)\right) F^H$ is an $N \times N$ matrix of the PN spectral components of the BS oscillator, and $G_T$ is a $KN \times KN$ matrix consisting the $N$ spectral components of the PNs of all the $K$ users. The $(k, l)$th entry of $G_R$ is denoted as $g_{R(k-l)N}^{tx}$, where $(k-l)N$ denotes $(k-l) \mod N$. The $(k, l)$th block of $G_T$ is $g_{T(k-l)N}^{tx}$, where the $K \times 1$ vector $g_{T(k-l)N}^{tx}$ consists of the corresponding spectral components of the PNs from the $K$ users. By separating the CPE and ICI terms $G_T = I_N \otimes \text{diag}(g_{0}^{tx}) + P_I$, and $G_R = g_{0}^{rx} I_N + P_R$, (4.10) can be rewritten as

$$y = g_{0}^{rx} H \left( I_N \otimes \text{diag}(g_{0}^{tx}) \right) x + e + w,$$  \hfill (4.11)

where $g_{0}^{rx}$ is the CPE of the PN at the BS, $g_{0}^{tx}$ consists of the CPEs of the PNs from the $K$ users, and the ICI term $e$ is

$$e = (P_R \otimes I_M) H P_I x + g_{0}^{rx} H P_I x + (P_R \otimes I_M) H \left( I_N \otimes \text{diag}(g_{0}^{tx}) \right) x.$$  \hfill (4.12)

The receive signal at the $k$th subcarrier can be expressed as

$$y_k = H_k \text{diag}(g_0^{tx}) x_k + e_k + w_k,$$  \hfill (4.13)

where $g_0^{tx} = g_0^{rx} g_0^{tx}$, and $e_k$ and $w_k$ denote ICIs and AWGNs at the $k$th subcarrier, respectively.

The CPEs $g_0^{tx}$ can be estimated as

$$\hat{g}_0^{tx} = \frac{1}{N_p} \sum_{k \in S_p} \left( H_k \text{diag}(x_k) \right)^\dagger y_k,$$  \hfill (4.14)

where $S_p$ denotes the set of the $N_p$ scattered pilots. The CPEs can be corrected at the ZF decoder as

$$\hat{x}_k = \text{diag}(\hat{g}_0^{tx})^{-1} H_k^\dagger y_k,$$  \hfill (4.15)

where $\hat{x}_k$ denote the detected signals at the $k$th subcarrier. Note that, for convenience of analysis, we assume perfect channel estimation. In practice, the estimated channel in the preamble will contain an initial CPE. In the payload, the CPE will be different. Thus, it is necessary to estimate the relative CPE with respect to the initial CPE. Mathematically, it is equivalent to assume that the estimated channel contains the preamble CPE and then the relative CPE in the payload can be estimated by (4.15).

Assuming Wiener PN with 3-dB bandwidth of $\beta$, oscillators at different users are of the same quality (i.e., PNs from different user have the same 3-dB bandwidth $\beta$), perfect CPE estimation, and spatially white Rayleigh fading at each subcarrier, the EVM of the uplink multi-user MIMO-OFDM system is readily derived as

$$\xi_k = \frac{\left(M - K\right) \gamma_0 \left( N^2 - \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} \exp(-2\pi\beta|n-m|T_s) \right) + N^2}{\left(M - K\right) \gamma_0 \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} \exp(-2\pi\beta|n-m|T_s)},$$  \hfill (4.16)

where $\gamma_0$ represents the SNR and $T_s$ denotes the sampling duration.
For simulations, we assume there are 512 subcarriers including 32 scattered pilots. The remaining active subcarriers are loaded with QAM symbols. The multipath fading channel is a 4-tap Rayleigh fading channel, where the taps are at the 0, 20, 30, and 60th time samples with equal average tap gain of 0.25, and that the channel stay constant within 40 OFDM symbols after which an independent channel realization is drawn (in total 100 channel realizations are generated). The CP length of the OFDM is set to 64 so that there is no inter-symbol interference due to the delay spread. The PNs from different users are independent yet follow the same (Wiener process) distribution with the same 3-dB PN bandwidth $\beta$. Figure 4-25 shows the EVM performance of the MU-MIMO-OFDM system (in the presence of multiple PNs) with two users and four BS antennas. As a reference, the ideal case (no PN) is also plotted in the same figure. As can be seen, with modest PN ($\beta \leq 500$ Hz) the simulated EVM with CPE correction agrees well with that of the theoretical one. This implies that the CPE correction can eliminate the CPEs of the multiple PNs. Note that, as $\beta$ increases, the CPE estimation becomes less accurate. As a result, it is also shown that the EVM performance with CPE correction is slightly worse than its theoretical counterpart as $\beta$ increases to 1000 Hz.

![Figure 4-25 EVM performance of uplink multi-user MIMO-OFDM (K=2, M=4, Ts = 10 ns) under different PNs.](image)

4.6 Impact of hardware impairments on system performance achieved by hybrid beamforming for mm-wave access

While hybrid beamforming (HBF) represents a good trade-off between complexity and performance when compared to pure analog beamforming (ABF) and fully digital precoding (DP), it is also characterized by some specific hardware impairments (HW) that need to be properly assessed. In particular, in this section we analyse the impact of a) the errors introduced by phase shifters and b) the losses added by the combiner stage in front of the antenna element. Regarding the simulation setup and channel model, we consider the system already described in Section 2.

4.6.1 Phase shifter errors

Radio frequency (RF) beamforming can be implemented by applying phase shifts that steer the transmit power toward a certain direction. In practical systems, there is an error in the shift that is applied by the phase shifter, mainly (but not only) due to the limited number of bits that is used to control the phase shifter itself [SSO09, LCT+13]. Here, we model this error $\Theta$ as a uniform random variable in the interval $[-\Theta_{\text{max}}, \Theta_{\text{max}}]$, we assume it to be independent among different phase shifters and approximate it as a constant on the whole band [BJ09]. In Table 4-15 we report the average and fifth percentile UE throughput achieved by a system where BSs employ HBF with 64 antennas and $P=4$ RF transceiver chains for $\Theta_{\text{max}}=2^0, 6^0, 10^0$. Note that in practical systems an error of $\Theta_{\text{max}}=10^0$ is to be considered as very high [SSO09, LCT+13]; a much lower error is typically expected, and therefore we report these results only.
as a worst case scenario. Table 4-15 shows that the impact of the phase shifter error on system performance is rather minor, causing a maximum loss in the average UE throughput up to 3.5%. More detailed results [GRB+16, Table 4] show also that LOS UEs are a bit more affected by this impairment when compared to NLOS UEs: however, the percentage loss is always in the order of few percentage points, confirming that other impairments (as for example the combiner stage loss detailed in the next sub-section) are typically impacting more the system performance than the phase shifter errors.

<table>
<thead>
<tr>
<th>5th percentile UE throughput</th>
<th>No error</th>
<th>$\theta_{\text{max}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>2°</td>
</tr>
<tr>
<td>[Mbit/s]</td>
<td>421.13</td>
<td>418.97</td>
</tr>
<tr>
<td>[% loss]</td>
<td></td>
<td>-0.51%</td>
</tr>
<tr>
<td>Average UE throughput</td>
<td>[Mbit/s]</td>
<td>1520.33</td>
</tr>
<tr>
<td>[% loss]</td>
<td></td>
<td>-0.12%</td>
</tr>
</tbody>
</table>

### 4.6.2 Combiner stage losses

With HBF the signals coming from different RF transceiver chains are added together to feed an antenna element. Practical combiners always introduce a power loss [GVR+16], which increases with the number of combining stages (i.e., RF chains). This power loss can be (at least partially) compensated by introducing a power amplifier after this combiner stage. This, however, requires the integration of a power amplifier in front of each antenna element, which might not always be cost-effective, in particular with HBF schemes whose main advantage (when compared to DP) is the reduced number of RF transceiver chains. Clearly, taking this loss into account introduces a trade-off in the number of RF transceiver chains that should be used: more transceiver chains allow better multiplexing and beamforming capabilities, but results in a reduction of the effective transmit power. The total power loss $L_{\text{dB}}$ can be written as $L_{\text{dB}} = L_{\text{dB,loss}}$, where $L_{\text{dB}}$ is the loss introduced by a combiner stage with only two input branches [GRB+16, Section 5.2]. In Figure 4-26 and Figure 4-27 we report the CDF of the UE throughput with HBF, P=4, 8, 16 RF chains and $L_{\text{dB}} = 0, 1, 3$ dB, where 0/1/3 dB means full/partial/no compensation of the combiner stage loss. These results show that the combiner stage loss can have a great impact on the system performance at mm-wave. In detail, the percentage loss for P=4 (when compared to the ideal case with $L_{\text{dB}} = 0$ dB) increases from about 5% with $L_{\text{dB}} = 1$ dB to about 20% with $L_{\text{dB}} = 3$ dB. Even more impact is observed with a higher number of RF chains: the percentage loss for P=16 increases from about 10% with $L_{\text{dB}} = 1$ dB to about 50% with $L_{\text{dB}} = 3$ dB. Moreover, from Figure 4-27 we observe that, when $L_{\text{dB}} = 3$ dB, the scheme with 8 RF chains outperforms the scheme with 16 transceiver chains, highlighting the fact that there is an optimal number of RF chains that should be used with HBF, which depends on many system parameters including the combiner stage loss.
4.7 Impact of Major RF Impairments on mm-wave Communications using OFDM Waveforms

Mm-wave communication systems are generally hardware-constrained. As described in [MMMAGICD51], the link and system performance are degraded by several RF impairments, e.g., local oscillator phase noise and PA nonlinearity. All those RF impairments, in practice, co-exist in the transceivers.

As one concrete example, we consider a mm-wave communication link that suffers from joint effects of phase noise, PA nonlinearity and I/Q imbalance at the TX side and phase noise and I/Q imbalance at the RX side [ZZG+16]. For simplicity, we only consider analog beamforming in this mm-wave communication link. Combining the propagation channel and analog beamformers at the TX and RX, we denote the effective channel between TX transceiver and RX transceiver as a SISO channel impulse response \( h(t) \). At the \( m \)-th symbol interval, OFDM modulated symbol \( s_m(t) \) with \( N \) subcarriers \( (s_{k,m} \text{ at the } k \text{-th subcarrier}) \) is transmitted from TX side. Assuming the cyclic prefix is longer than the overall delay spread of \( h(t) \), the signal reception under the considered RF impairments at the \( k \)-th subcarrier at the RX reads...
\[ Z_{k,m} \approx \bar{H}_{k,m} S_{k,m} + \bar{H}_{k,m}^* S_{N-k+1,m}^* + \varepsilon_{ICL,R,k,m} + H_k J_{0,R,m} G_{1,R,k} (\varepsilon_{ICL,T,k,m} + \mu_{NL,k,m}) + n_{k,m} \]  

(4.17)

where \( H_k \) denotes the effective channel frequency response at the \( k \)-th subcarrier, \( \bar{H}_{k,m} = \hat{\varepsilon} \hat{a}_1 J_{0,T,m} \) is the overall effective channel including the linear gain of the propagation channel, PA, I/Q imbalance as well as CPE rotation at TX and RX, and \( n_{k,m} \) is the zero-mean complex Gaussian distributed channel noise.

As shown in (4.17), in addition to channel noise the received signal \( Z_{k,m} \) includes the following terms:

1) The desired signal \( \bar{H}_{k,m} S_{k,m} \). Even if the OFDM symbol duration is much smaller than the coherence time, the overall effective channel \( \bar{H}_{k,m} \) changes from one OFDM symbol to another due to the time varying CPE. In case that the CPE changes dramatically in consecutive OFDM symbol durations, a reference signal is required for CPE estimation in every OFDM symbol duration.

2) Mirror-frequency interference \( \bar{H}_{k,m} S_{N-k+1,m}^* \). This term is caused by the I/Q imbalance effect and its parameters change very slowly. The resulting interference could be mitigated using reasonable system resources, e.g., if a pilot signal is available at the mirror-frequency subcarriers.

3) ICI term due to phase noise \( \varepsilon_{ICL,R,k,m} + H_k J_{0,R,m} G_{1,R,k} \varepsilon_{ICL,T,k,m} \). The level of interference depends on the characteristics of phase noise and also the used subcarrier spacing in OFDM modulation. Combined with the effects of a frequency-selective channel, ICI mitigation can be very challenging. In addition to continuously improved integrated circuit design at mm-wave frequency, devising efficient ICI cancellation schemes, e.g., that only mitigate major interference from neighbouring subcarriers could form a reasonable solution. Such digital compensation approaches, however, require again a dedicated reference signal structure in every OFDM symbol.

4) Nonlinear distortion term \( H_k J_{0,R,m} G_{1,R,k} \mu_{NL,k,m} \). This term is caused by PA nonlinear behavior and also varies in different OFDM symbol durations.

Next, based on (4.17), we examine the impact of the considered RF impairments using extensive computer simulations with the initial air interface design proposed by the mmMAGIC project [VZW+16]. Assume the system operates at a central frequency of 28 GHz or 82 GHz. The channel delay profile in [RHD+15] is implemented to model the multi-path effect of the propagation channel. The used OFDM waveform consists of 2048 subcarriers with 60 kHz subcarrier bandwidth at 28 GHz and 480 kHz subcarrier bandwidth at 82 GHz. The CP length is assumed to be 144 samples. The example subcarrier modulation is set to be 16 QAM. The array gain is assumed to be 30 dB. The SNR in the simulation is defined as received SNR plus array gain.

Regarding the impairment parameter settings, two phase noise models are proposed in the mmMAGIC project [MMMMAGICD51]. The “low” phase noise model corresponds to the use of an oscillator with reasonable performance whereas the “high” phase noise model considers the case that a low-cost and low-power transceiver is used and the oscillator quality is very poor. Here, two cases are considered. Case I: good quality oscillators are used at the both TX and RX. Case II: good quality oscillator is used at the TX and low quality oscillator is used at the RX. The PA nonlinearity is modeled by the sum of 3rd, 5th, 7th, 9th order memoryless polynomials. The frequency-selectivity I/Q imbalances are modeled with three-tap branch filters corresponding to image rejection ratio (IRR) of 25-40 dB. All the parameters used in the simulations are translating from measurement results in lower frequencies [ZRL+16] as the nonlinear behaviors of I/Q imbalance and PA nonlinearity aren’t essentially changed. For reference purpose, perfect I/Q match is also considered in the simulation in order to demonstrate the achievable performance with I/Q imbalance compensation.
First, we examine the level of effective channel fluctuations stemming from varying CPEs. In detail, the total effective channels in two consecutive OFDM symbol durations \( \tilde{H}_{k,m+1} \) and \( \tilde{H}_{k,m} \) are compared in terms of EVM as

\[
EVM = 10 \log_{10} \left( \frac{|\tilde{H}_{k,m+1} - \tilde{H}_{k,m}|^2}{|\tilde{H}_{k,m}|^2} \right)
\]  

(4-18)

The resulting probability density functions (PDF) are illustrated in Figure 4-28. It clearly shows that CPE tracking is needed in each OFDM symbol with operating frequency at 82 GHz.

Then, performance degradation due to RF impairments is demonstrated in terms of average achievable sum rates over \( N \) subcarriers as

\[
R \approx \frac{1}{N} \sum_{k=0}^{N-1} E\left[\log_2 \left( 1 + \gamma_k \right) \right]
\]

(4-19)

where \( \gamma_k \) refers to the instantaneous SINR at the \( k \)-th subcarrier. As shown in Figure 4-29 in considering SNR of 10-20 dB at 28 GHz, the used air interface design is robust to the considered RF impairments even when the high phase noise model is used at the RX side. On the other hand, at 82 GHz, the phase noise can be problematic with the “high” phase noise model. Performance degradation is non-trivial even at the SNR of 15-20 dB.

In general, with reasonably good transceiver implementation, the air interface proposed by the mmMAGIC project [VZV+16] is robust to the considered impairments. In case of highly spectrum efficient transmission and/or low-power and low-cost transceiver implementation, digital/analog compensation schemes need to be deployed.

**Figure 4-28** PDF of difference on the equivalent channel between two consecutive OFDM symbols in terms of EVM as defined in (4.19) for both case I and case II at 28 GHz

**Figure 4-29**: Performance comparisons without and with RF impairments. Two oscillator implementations case I and case II are considered. Operating frequency are (a) 28 GHz and (b) 82 GHz respectively.
4.8 Summary

In the framework of WP5, we first identified the hardware and radio frequency impairments highlighting different modelling approaches for characterizing the impact of each of them at waveform level, link level and system level, respectively.

Different models have been proposed for phase noise, power amplifier, I/Q imbalance, ADC and antennas.

Starting from antenna specifications for terminal, access and backhaul, different antenna arrays have been studied. It was shown how losses and reduced efficiency can reduce the antenna performance (w.r.t. the ideal lossless ones) and yield a higher number of antennas to reach the target antenna performance. Based on realistic implementations, a database of antenna models including the full scattering matrix and the full antenna patterns for different steering directions has been provided. In particular in the 24.25-27.5 GHz band patch and double polarized antenna arrays up to 8x8 elements have been considered. For the base station Transmatria array technology has been investigated. One of the advantages of this technology is to reduce the losses due to the feeding lines and have high efficiency. An electronically reconfigurable Transmatria allowing fine steering for radio access has been proposed. With this solution, the phase control is directly implemented on the unit cell, by employing p-i-n diode switches, instead of using phase shifters as classically done in phased arrays. For backhaul applications, fixed or switched beam solutions can be adopted. The performance of such arrays as a function of phase quantization (i.e. states of the unit cells) has been investigated. In particular, a specific 3-bits unit cell was introduced to realize a 40×40 transmit arrays Transmatria reaching the requirements for backhaul applications.

A behavioural model of Power Amplifier (PA) distortion in active arrays was investigated and a stochastic model was proposed for link or network level simulations. It was shown that the estimated spatial behaviour of the error signal of the statistical model fits well with the corresponding behavioural model part. Also, the scaling laws for energy efficiency analysis were presented, showing that when moving towards mm-wave frequency and very large bandwidth, it is quite costly in terms of power consumption.

A phase noise (PN) model, also employed for WP4 waveform design was introduced in D5.1 [MMMAGIC D51]. The impact of phase noise on the local oscillator, according to different strategies, has been investigated in MIMO links. Here we also evaluated the performance of the MU-MIMO-OFDM system (in the presence of multiple PNs) with two users and four BS antennas. It was shown that Common Phase Error (CPE) correction can eliminate the CPEs of the multiple PNs.

An additional simple PN model and design of Phase Tracking Reference Signal (PTRS) was illustrated. The overall results indicate that the time domain PTRS density needs to be virtually every symbol, but in the frequency domain, PTRS can be in a sub-carrier per every few Physical Resource Blocks (PRBs). This work was also submitted to 3GPP RAN1 [3GPP-RAN1-17].

The impact of phase noise on the local oscillator, according to different strategies, has been investigated in MIMO links. Low precision ADC receiver performance has been addressed [MMMAGICD51]. Also the impact of hardware impairments, such as phase noise, power amplifier, and phase amplitude errors, has been investigated for hybrid beamforming schemes. In practice, all those RF impairments co-exist in the transceivers. In section 4.7 it was shown that, with realistic transceiver architecture, the air interface proposed by the mmMAGIC project [VZV+16], [MMMAGIC-D42] is robust to the considered impairments.
5 Conclusions

This deliverable captures the extensive body of work carried out by WP5 in the research areas of multi-antenna and multi-node communications and hardware performance/impairment modelling for mm-wave mobile systems. As set out in the project proposal, we have analysed and proposed advanced solutions for radio access, backhaul and relay deployments with the aim of enhancing the key KPIs captured in the WP5 adapted use cases [mmMAGIC/D5.1]. In hardware performance and impairment modelling, we have analysed the key components in the mm-wave transceiver chain and provided enhanced models, which have now been used extensively in other WP5s and as open source resources. It is expected that the presented solutions and analyses in this deliverable will significantly contribute to address the challenges in developing mm-wave transceivers and systems for 5G mobile communications.

The analysis of multi-antenna schemes in chapter 2 has emphasized the general choice of hybrid beamforming (HBF) as the preferred beamforming method (over analog and digital beamforming), due to the flexibility it provides to meet the varying system requirements. Under radio access, a sub-array architecture for hybrid beamforming is proposed and it is shown that the performance can be better than the digital BF counterpart, when combiner losses in the fully connected digital BF architecture are considered. A flexible HBF scheme, which overcomes the need for accurate channel knowledge is developed. A system level multi-user analysis of the ABF, HBF and DBF schemes has shown that the simpler HBF schemes can match the performance of highly complex DBF schemes, when there is sufficient user separation in LOS conditions. Under the study of backhaul schemes, an experimental validation of the MMIMO scheme presented in D5.1 [mmMAGIC/D5.1] for realistic channels and antenna patterns is conducted. It is shown that the massive spectral efficiencies theoretically recorded can be achieved to a large extent (up to 84%) in these more realistic channel conditions. An analysis on backhaul provision for moving hotspots highlights the performance trade-offs in having wider beams and the negative impact of higher vehicular speed variances on the achievable data rates. Under the analysis of mm-wave relay systems, a relay selection scheme, based on the candidate relay user location and the SNR that can be supported, is shown to provide greatly reduced spectrum efficiency loss compared to random relay selection. A related study on relays as multi-nodes to provide O2I coverage is reported in chapter 3. This work shows the trade-offs between the bandwidth allocations for relay access and backhaul links and enhanced performance of relay links when the ISDs are larger.

In chapter 3 we have investigated various multi-node configurations to increase the mm-wave radio link reliability. In the sequential hybrid beamforming (HBF) design for multi-link mm-wave communication in which a two-step precoding approach is adopted, it is shown that the proposed pre-coding methods achieve sum rates closer to DBF in the multi (two) node scenarios. In the analysis with hybrid FSO and mm-wave links it was shown that with proper H-ARQ in a multi-hop system the impact of imperfect PA in the mm-wave links can be effectively controlled. A separate analysis with high rise APs complementing the coverage provided by a network of low rise APs was also presented. The results show that the LOS probability can be significantly increased by having a few high rise APs, at a height above the local clutter level.

The hardware impairment and performance modelling studies are reported in chapter 4. We provide modelling results for mm-wave handsets and access points, for radio access and backhaul applications. The antenna array solutions are derived to meet the different challenges and also make use of different opportunities in these scenarios. A relatively novel array architecture – the Transmitarray – is presented and the different trade-offs in terms of phase shifter complexity and the radiation pattern losses are discussed. In terms of impairments the phase noise is studied in detail in this project. A comprehensive PN model presented in D5.1 [mmMAGIC/D5.1] is used to derive the MIMO-OFDM performance with independent oscillators in the uplink, and results indicate that effective CPE compensation can be achieved with the proposed method for multiple PN sources. A simpler second PN model is
also presented in this deliverable and it is used to analyse the effectiveness of different PTRS
densities in channel and sub-frame conditions currently studied by 3GPP. This analysis
indicates that the higher densities in the time domain have much larger impact in reducing PN
when the users are allocated larger amounts of physical resource blocks (PRBs). A
behavioural study of the Power Amplifiers is also presented for analysing the system impact of
the PA with large antenna arrays. The error signal analysis for coupling an antenna array to a
PA transmitting an OFDM signal shows nearly identical performance for this behavioural
analysis to the more conventional statistical analysis. Further studies of the impacts of these
impairments in the overall system performance have indicated that the Hybrid beamforming
mechanism is more resilient to combiner and phase shifter losses in the hardware chains.

The following key take-away points can be stated to aid the development of an overall mm-
wave system concept.

- The HBF architecture provides the best trade-off between complexity and performance
and hence is recommended for the majority of mm-wave radio access configurations. The
system level analyses have shown that the HBF can match the DBF performance in LOS conditions (which are more likely in smaller mm-wave cells). HBF is also more resilient to errors/losses caused by hardware impairments and imperfections.

- The multi-node configurations (be it through AP, relay or some other technology like
FSO) are necessary to increase the link reliability of mm-wave systems. This ensures
coverage can be extended to reasonable cell sizes that enable viable deployments. The 'second' link can provide diversity or spatial multiplexing gains and will be an essential component in mm-wave system design.

- Bespoke antenna design solutions are needed for handset and AP (both Radio Access
and Backhaul) configurations, taking the specific challenges and also utilizing
opportunities. For radio access wider beams with wider scan angles will be needed,
while for backhaul, very narrow beam with much narrower scan angles (for self-
alignment) can be supported. The Transmitarray design can meet both sets of
challenges with varying levels of complexity. For handset arrays, the usual constraints
of size, hand shadowing and cost must be met on top of providing very wide scan
angles.

- The hardware impairments place significant performance boundaries on mm-wave
systems. These must be properly modelled, taking the overall system behaviour into
account. The impact of these impairments should be mitigated with proper waveform,
frame structure and numerology design, which is the domain of WP4. The Phase
Noise and power amplifier models developed by WP5 have been used in these related
analyses.
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