# RECOMMENDATION OF ANTENNAS AND COMMUNICATION TECHNIQUES QUALIFIED FOR IMPLEMENTATION

<table>
<thead>
<tr>
<th>Deliverable Identifier:</th>
<th>D3.1</th>
</tr>
</thead>
<tbody>
<tr>
<td>Delivery Date:</td>
<td>30 April 2017</td>
</tr>
<tr>
<td>Classification:</td>
<td>Public</td>
</tr>
<tr>
<td>Editors:</td>
<td>Athanasios Kanatas, Petros Bithas, UPRC</td>
</tr>
<tr>
<td>Document version:</td>
<td>V2.0</td>
</tr>
</tbody>
</table>

| Contract Start Date:    | 1 May 2015            |
| Duration:               | 36 months             |
| Project coordinator:    | IMST (Germany)        |
| Partners:               | MAN (Germany), TNO (Netherlands), UPRC (Greece) |

**Horizon 2020**

Call: H2020-MG-2014_TwoStages  
Topic: MG-3.5a-2014  
Type of action: RIA  
Proposal number: SEP-210181921  
Proposal acronym: ROADART
# Revision History

<table>
<thead>
<tr>
<th>Version</th>
<th>Date</th>
<th>Description and comments</th>
<th>Edited by</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>07/03/2017</td>
<td>TOC, Sections 3.1-3.2, 4.1-4.2</td>
<td>Petros Bithas</td>
</tr>
<tr>
<td>0.2</td>
<td>10/03/2017</td>
<td>Section 2.1</td>
<td>Antonis Aspreas</td>
</tr>
<tr>
<td>0.3</td>
<td>15/03/2017</td>
<td>Section 2.2</td>
<td>Konstantinos Maliatsos</td>
</tr>
<tr>
<td>0.4</td>
<td>3/04/2017</td>
<td>Section 5.1</td>
<td>Leonidas Marantis</td>
</tr>
<tr>
<td>0.5</td>
<td>11/04/2017</td>
<td>Sections 1, 3.3-3.5, 4.6, 6, 7</td>
<td>Petros Bithas</td>
</tr>
<tr>
<td>0.6a</td>
<td>19/04/2017</td>
<td>Sections 5.3, 5.4</td>
<td>Dimitrios Rongas</td>
</tr>
<tr>
<td>0.6b</td>
<td>19/04/2017</td>
<td>Sections 5.2, 5.5, 5.6, 5.7</td>
<td>T. Paraskevopoulos</td>
</tr>
<tr>
<td>0.6_final</td>
<td>19/04/2017</td>
<td>Sections 4.4, 5.1, 5.2, 5.8, 6.3, 7</td>
<td>Leonidas Marantis</td>
</tr>
<tr>
<td>0.7a</td>
<td>20/04/2017</td>
<td>Section 4.3</td>
<td>Emmanouel Michailidis</td>
</tr>
<tr>
<td>0.7b</td>
<td>21/04/2017</td>
<td>Section 4.5</td>
<td>Kostas Peppas</td>
</tr>
<tr>
<td>0.8</td>
<td>24/04/2017</td>
<td>Executive Summary, Section 7, Review</td>
<td>Athanasios Kanatas</td>
</tr>
<tr>
<td>1.0</td>
<td>27/04/2017</td>
<td>Review</td>
<td>Petros Bithas</td>
</tr>
<tr>
<td>2.0</td>
<td>19/05/2017</td>
<td>Deliverable classification change from Restricted to Public after unanimous agreement of the partners and the PO</td>
<td>C. Oikonomopoulos</td>
</tr>
</tbody>
</table>
Executive Summary

The main drivers for the work that has been carried out in the framework of work-package (WP) 3 were the objectives of the project as they have been provided in the description of work. In particular, the main scope of this WP was to propose specific robust and adaptive communication techniques, algorithms, strategies, and architectures that significantly improve the system performance over the specific vehicular radio channels. From the very beginning, the most promising techniques identified were based on the diversity concept, utilizing multiple antennas. In this context, a major initial target in WP3 was to implement a new simulator for the IEEE 802.11p standard, which is considered as the underlying protocol for inter-vehicular communications (IVCs). In particular, both the physical (PHY) as well as the medium access control (MAC) layers of the IEEE802.11p standard have been implemented in MATLAB/OCTAVE. All major receive diversity reception techniques for truck-to-truck (T2T) and truck-to-infrastructure (T2I) communications have been also implemented and their performance has been evaluated using the simulation platform. A critical part of the research was to identify the consequences of the unique characteristics of the T2T communications. In this context, various effects such as the spatial correlation among the multiple antennas, the existence of interference, and the outdated channel state information (CSI) have been identified and their impact on the diversity gain has been analytically investigated. It was shown that all these parameters seriously deteriorate system’s performance by at least 3dB in terms of the outage probability (OP). Moreover, in cooperative communication scenarios it is critical to identify the required hops between the source and the final destination. Therefore, the issue of the connectivity probability among multiple vehicle-nodes was studied in order to discover the probability of correct packet detection versus the neighbour node distance.

Since the main objective of this WP was to propose low complexity algorithms and architecture designs for T2T communication, several techniques satisfying these requirements have been proposed. More specifically, a new reconfigurable antenna pattern selection scheme is proposed that aims to reduce the complexity of the traditional antenna selection techniques. The new technique offers a clear diversity gain, which is related to the number of the different antenna patterns that can be supported, and a performance improvement of up to 10 dB as compared to the single reception scenario. A new relay selection scheme has been also investigated under different T2T communication scenarios that also aims to reduce the signal processing complexity that the traditional relay selection schemes induce to the system. The new scheme offers an excellent compromise between performance and complexity as compared to existing schemes. In addition, a novel extended open loop beamforming (eOLB) technique that is realized through pattern selection is designed and demonstrated via a proof-of-concept indoor experiment. This technique provides up to 10 dB performance gain per case or 6.5 dB average gain with respect to the use of omnidirectional antennas. Furthermore, the benefits of adopting both space shift keying as well as spatial modulation in IVC scenarios have also been studied. It was shown that both these schemes outperform single transmit/reception case as well as other well-established multiple-input multiple-output schemes, in terms of bit error rate, offering simultaneously significant lower hardware and signal processing complexity.

Another important part of the research carried out in this WP was the investigation of the antenna that will be employed at the development of the ROADART prototype. The research on antennas for ROADART was based on electronically steerable passive array radiator (ESPAR) designs and includes the selection of the antenna position on the truck. The proposed antenna-truck configuration (ESPARs on the truck side mirrors) achieves considerable performance improvement. Two types of 3-element printed ESPAR antenna were designed and fabricated providing simulation and measurement results in good agreement and exhibiting a valuable pattern reconfigurability (3-4 different radiation patterns). An impedance matching network mechanism has been proposed in order to provide satisfying reflection coefficients for all antenna operating states. Moreover, the
number of the antenna operating states was increased to 9 radiation patterns with the design of 5-element ESPAR antennas (a 5-element printed mono/dipole ESPAR and a 5-wire monopole ESPAR). A further expansion of the ESPAR pattern reconfigurability was studied theoretically by exploiting the tuning effect of the varactor diodes that can offer numerous antenna patterns with less array elements.

Finally, based on the proposed techniques and taking also into consideration the requirements/constraints raised by external factors from this WP, for example architecture design and integration constraints, the recommended techniques qualified for implementation were presented.
Contents

1 Introduction ................................................................................................................................. 12
  1.1 Objectives ......................................................................................................................... 12
  1.2 Structure of the Document ............................................................................................... 13

2 IEEE 802.11p Simulator ............................................................................................................. 14
  2.1 PHY Layer ......................................................................................................................... 14
  2.2 MAC Layer ......................................................................................................................... 17
    2.2.1 Initialization of Users ................................................................................................. 20
    2.2.2 Data Generation Procedure .................................................................................... 21
    2.2.3 Simulator Actions per Status .................................................................................. 21
  2.3 Summary ............................................................................................................................. 27

3 Investigations of Impairment Parameters and Performance Analysis in IVC Systems ........... 28
  3.1 Spatial Correlation ............................................................................................................. 28
    3.1.1 Correlated Double Rayleigh Statistics ..................................................................... 28
    3.1.2 SD Receivers Performance ....................................................................................... 30
    3.1.3 Numerical Results .................................................................................................... 32
  3.2 Interfering Effects .............................................................................................................. 35
    3.2.1 System and Channel Model ....................................................................................... 35
    3.2.2 Single Reception Scenario ....................................................................................... 37
    3.2.3 Multichannel Communication Investigation .......................................................... 38
  3.3 Outdated Channel State Information (CSI) ....................................................................... 40
    3.3.1 Asymptotic Analysis .................................................................................................. 40
    3.3.2 Varying Correlation for Realistic Modeling .............................................................. 41
    3.3.3 Numerical Results .................................................................................................... 42
  3.4 Nearest Neighbor Node Discovery .................................................................................... 43
    3.4.1 System Model ........................................................................................................... 44
    3.4.2 Wireless Applications ............................................................................................... 47
  3.5 Summary ............................................................................................................................. 50

4 Low Complexity Algorithms and Architecture Designs for T2T Communications ............... 52
  4.1 A New Reconfigurable Antenna Scheme and its Application to Vehicle-to-Vehicle
      Communications .................................................................................................................. 52
    4.1.1 System and Channel Model ....................................................................................... 52
    4.1.2 Statistical Characteristics .......................................................................................... 53
    4.1.3 Numerical Results .................................................................................................... 56
  4.2 A Cooperative Relay Selection Scheme in V2V Communications under Interference and
      Outdated CSI ....................................................................................................................... 59
    4.2.1 System and Channel Model ....................................................................................... 60
    4.2.2 SIR Statistics .............................................................................................................. 62
    4.2.3 Performance Evaluation ............................................................................................ 64
    4.2.4 Numerical Results .................................................................................................... 65
  4.3 MIMO V2V Communications via Multiple Relays: Relay Selection Over Space-Time
      Correlated Channels ........................................................................................................... 67
    4.3.1 System Model ........................................................................................................... 68
    4.3.2 Application of the Relaying Scheme on ITS ............................................................. 69
    4.3.3 Simulation Results .................................................................................................... 69
List of Figures

Figure 1: 802.11p simulator model. ................................................................. 14
Figure 2: Transmitter software diagram. ............................................................ 15
Figure 3: Receiver software diagram. ................................................................. 16
Figure 4: CSMA/CA implementation in IEEE 802.11p. ................................. 18
Figure 5: EDCA implementation in IEEE 802.11p ............................................ 19
Figure 6: Simulator time controller. ................................................................. 21
Figure 7: Basic flow of the Simulator loop. ....................................................... 22
Figure 8: Data generation procedure. ............................................................... 23
Figure 9: State transition for current State = 0. ............................................... 24
Figure 10: State transition for current State = 1. ............................................. 24
Figure 11: State transition for current State = 2. ............................................. 25
Figure 12: State transition for current State = 3. ............................................. 25
Figure 13: State transition for current State = 4. ............................................. 26
Figure 14: State transition for current State = 5. ............................................. 27
Figure 15: Keyhole channel as simulated using WINNER2. ............................ 33
Figure 16: Histogram using simulation vs. theoretical PDF for Z1 and Z2. ......... 33
Figure 17: OP vs $\rho$, for different values of $\gamma_{th}$. ........................................ 34
Figure 18: ABEP vs $\gamma$, for different values of $\rho$. ....................................... 34
Figure 19: Example of a LOS in SOS conditions. ............................................ 36
Figure 20: OP of single antenna scenario vs M. ............................................... 38
Figure 21: OP of different diversity schemes vs the average input SNR. ............. 40
Figure 22: OP vs the two correlation coefficients. .......................................... 42
Figure 23: OP vs the average input SNR. ......................................................... 43
Figure 24: OP vs the average input SNR for varying correlation. ....................... 43
Figure 25: CCDF of received SNR versus SNR threshold $\gamma_{th}$. ...................... 44
Figure 26: Probability of connectivity versus the neighbor index $n$. .................. 48
Figure 27: Probability of connectivity versus the neighbor index $n$ for $a=3$. ....... 49
Figure 28: Probability of abstention versus the neighbor index $n$ for $a=3.5$. .... 49
Figure 29: Mode of operation of the proposed scheme. ................................... 50
Figure 30: OP vs the average received SNR for different values of $\rho$. ............... 53
Figure 31: BEP vs correlation coefficient $\rho$ and for different values of $\gamma_T$. ... 53
Figure 32: Average BEP vs average SNR for the proposed scheme, MRC, SD, and SISO. .......................................................... 58
Figure 33: Average BEP vs average SNR for the proposed scheme, MRC, SD, and SISO. .......................................................... 59
Figure 34: OP vs $\rho$, for different values of $\gamma_T$. ......................................... 63
Figure 35: OP vs the average SNR of the 2nd hop for different values of $\rho_L$. .... 66
Figure 36: SEP vs number of relays for different values of $\Omega_{th}$. ....................... 67
Figure 37: Simple representation of a V2V multi-relay communication system, where one FD DF relay is selected. ...................... 69
Figure 38: Network throughput versus SNR between nodes 1 and 2 for different degree of local scattering at the vehicle nodes and different maximum multipath elevation angles. ........................................ 70
Figure 39: PER versus SNR for the traffic between nodes 1 and 2 for different degree of local scattering at the vehicle nodes and different maximum multipath elevation angles. ........................................ 71
Figure 40: Network throughput versus SNR for a) nodes 1 and 2 with relaying and MRC reception and b) nodes 3 and 4 with single-antenna point-to-point links for different degree of local scattering and different maximum multipath elevation angles. ........................................ 72
Figure 41: PER versus SNR for the traffic between a) nodes 1 and 2 with relaying and MRC reception and b) nodes 3 and 4 with single-antenna point-to-point links for different degree of local scattering and different maximum multipath elevation angles. ........................................ 73
Figure 42: Representative examples of a cooperative T2T communication link: (a) the “takeover” scenario, (b) the “platooning” scenario. .................. 74
Figure 43: The “top-view” layout of the ESPAR simulation model in CST. The ground plane stripe (printed at the back side of the panel) is also visible since the substrate is left out of the diagram. .................. 75
Figure 44: The two equivalent circuits for the (a) ON ($R_s = 0.9$ $\Omega$, $L = 0.45$ nH) and the (b) OFF ($C_T = 0.3$ pF, $R_p = 1$ K$\Omega$) state of the diode switch. .......... 75
Figure 45: The layout of the 3-printed monopole ESPAR antenna design. Three radiation patterns that correspond to the three ON-OFF combinations of the PIN diodes are roughly depicted. .......... 76
Figure 46: Simulated reflection coefficients ($S_{11}$) versus frequency for the OFF-OFF and ON-OFF antenna states. ........ 77
Figure 74: The effect of parameter “diff” to the return loss ($S_{11}$) versus frequency for the OFF-OFF state.

Figure 75: The layout of the indoor test site. Rx remains stable, Tx moves in four different locations.

Figure 56: A blind bend propagation scenario for V2I communications.

Figure 60: ABEP of uncoded SSK in a V2V environment.

Figure 61: ABEP of multiple antenna schemes in a V2I environment with curved road.

Figure 62: ABEP of SM-QPSK in a V2I environment with curved road.

Figure 63: ABEP bounds of coded SSK in a V2I environment with curved road.

Figure 64: The truck – trailer design along with the highway part underneath it.

Figure 65: Layout of (a) the single monopole and (b) the 3-monopole ESPAR antennas, demonstrating their simple geometry.

Figure 66: Return loss of the single monopole and the 3-element ESPAR.

Figure 67: The far-field radiation patterns of the single monopole antenna (red) and the 3-monopole ESPAR (blue) at the $\theta = 90^\circ$ plane (top-view).

Figure 68: The far-field directivity patterns for the first antenna – truck configuration (monopoles on the roof).

Figure 69: The far-field directivity patterns for the remaining two antenna – truck configurations (monopoles at the side mirrors – top, ESPARs at the side mirrors – bottom).

Figure 70: (a) Average SNR vs. distance for T2T C2 WINNER radio channels. (b) ABEP vs. distance for T2T C2 WINNER radio channels.

Figure 71: The detailed design of the proposed 3-element printed monopole ESPAR antenna in CST.

Figure 72: The “top-view” design of the ESPAR antenna along with the basic design parameters.

Figure 73: The effect of parameter “dis” to the return loss ($S_{11}$) of the printed monopole ESPAR antenna.

Figure 74: The effect of parameter “diff” to the return loss ($S_{11}$) of the printed monopole ESPAR antenna.

Figure 75: The effect of parameter “diff” to the radiation pattern of the printed monopole ESPAR antenna.

Figure 76: The effect of parameter “$W_p$” to the return loss ($S_{11}$) of the printed monopole ESPAR antenna.

Figure 77: The simulated return loss ($S_{11}$) versus frequency for the two states (ON-OFF and both OFF) of the printed monopole ESPAR antenna.

Figure 78: The simulated gain radiation patterns of the printed monopole ESPAR antenna on the (a) $H$ plane and (b) $E$ plane at 5.9 GHz for the OFF-OFF (blue) and the ON-OFF (red) states.

Figure 79: (a) Front-side view and (b) back-side view of the 3-element printed monopole ESPAR after its fabrication and assembly.

Figure 80: Comparison of the measured (straight lines) and simulated (dotted lines) reflection coefficient $S_{11}$ for different switching states.

Figure 81: Measured radiation characteristics (absolute gain) at 5.9 GHz: (a) $H$-plane radiation, (b) $E$-plane radiation for the OFF-OFF (blue) and the ON-OFF (red) states. Simulated results are plotted for comparison with dotted lines.

Figure 82: Perspective view of the wideband dipole with two parasitic elements.

Figure 83: Basic dimensions of the wideband dipole with two parasitic elements.

Figure 84: Reflection coefficient of the wideband mono/dipole with two parasitic elements on the omni-directional state (OFF-OFF).

Figure 85: Radiation pattern of the wideband mono/dipole ESPAR antenna at 5.9 GHz with parasitic elements on the omni-directional (OFF-OFF) state.

Figure 86: Reflection coefficient of the wideband mono/dipole ESPAR antenna with one parasitic element on the directive state (ON-OFF).
Figure 87: Radiation pattern of the wideband mono/dipole ESPAR antenna at 5.9 GHz with one parasitic element on the directive state (ON-OFF).

Figure 88: Reflection coefficient of the wideband mono/dipole ESPAR antenna with both parasitic elements short-circuited (ON-ON state).

Figure 89: Radiation pattern of the wideband mono/dipole ESPAR antenna with both parasitic elements short-circuited (ON-ON state).

Figure 90: Respective view of the wideband mono/dipole ESPAR antenna with the designed biasing network.

Figure 91: The final designs of the wideband mono/dipole ESPAR antenna (a) with DC pins and (b) without DC pins.

Figure 92: Simulated far-field directivity patterns for three different operating states. The OFF-ON state is excluded since it is identical to the ON-OFF (due to the perfect design symmetry).

Figure 93: The fabricated prototype of the wideband mono/dipole ESPAR antenna: (a) front-side view and (b) rear-side view.

Figure 94: Measured reflection coefficients of the mono-dipole ESPAR antenna without balun.

Figure 95: The mono-dipole ESPAR antenna with the bazooka balun attached during measurement.

Figure 96: (a) (a) Simulated reflection coefficients (S11) versus frequency for three different states. The OFF-ON state is excluded since it is identical to the ON-OFF (due to the perfect design symmetry). (b) Measured reflection coefficients (S11) versus frequency for four different states.

Figure 97: Diagonal-view and top-view layouts of the 5-element mono/dipole ESPAR simulation model in CST. The substrate panels were made transparent in order to make visible the ground plane of the mono/dipole and the “back” parasitic dipoles.

Figure 98: The simulated return loss (S11) results for 6 six different operating states.

Figure 99: Simulated radiation patterns of the printed mono/dipole 5-element ESPAR antenna with one element switched ON.

Figure 100: Simulated radiation patterns of the printed mono/dipole 5-element ESPAR antenna with two elements switched ON.

Figure 101: Investigation of a theoretical 3-element dipole ESPAR antenna at 5.9 GHz, loaded with varactor diodes in the parasitic elements.

Figure 102: S11 response of a 3-element dipole ESPAR antenna at 5.9 GHz with varactor capacitance ranging between 0.1-1 pF.

Figure 103: Radiation pattern response of a 3-element dipole ESPAR antenna loaded with varactors at 5.9 GHz.

Figure 104: S11 response of a 3-element dipole ESPAR antenna at 5.9 GHz with varactor capacitance ranging between 0.1-100 pF.

Figure 105: Schematic of the double-stub matching network configuration.

Figure 106: Simplified varactor circuit model.

Figure 107: C-V curve of the SMV 2019 varactor diode, employed in the proposed matching network.

Figure 108: Picture of the double stub matching network in the current 3-element ESPAR antenna design at 5.9 GHz in CST.

Figure 109: Simulated return loss (S11) of the 3-printed monopole ESPAR with and without the double-stub impedance matching network.

Figure 110: Picture of the single stub matching network in the current 3-element ESPAR antenna design at 5.9 GHz in CST.

Figure 111: Simulated return loss (S11) of the 3-printed monopole ESPAR with and without the single-stub impedance matching network.

Figure 112: ESPAR 5-element wire monopole antenna designed to operate at 5.9 GHz ITS band.

Figure 113: Reflection coefficient of the wire 5-elements ESPAR antenna for the two switching states (OFF- and ON-).

Figure 114: Radiation pattern of absolute (IEEE) gain in the XY (azimuth) plane of the 5-element wire ESPAR antenna.

Figure 115: Radiation pattern of absolute (IEEE) gain in the XZ (elevation) plane of the 5-element wire ESPAR antenna.

Figure 116: System concepts for cases 1 and 2.

Figure 117: System concepts for cases 3 and 4.

Figure 118: System concept for case 5.

Figure 119: System concept for case 6.

Figure 120: Performance comparison of the different diversity systems.

Figure 121: System model for proposed relay selection.

Figure 122: Open loop beamforming system model.
List of Tables

Table 1: Minimum Number of Terms of (9) For Accuracy Better Than \(\pm 10^3\) ................................................................. 33
Table 2: Simulator Parameters ................................................................................................................................. 58
Table 3: Diversity Receivers Parameters ............................................................................................................. 58
Table 4: ESPAR Antenna Dimensions .................................................................................................................. 76
Table 5: Truck – Trailer Materials ....................................................................................................................... 95
Table 6: Basic Truck Dimensions ............................................................................................................................ 95
Table 7: ESPAR Dimensions ................................................................................................................................... 97
Table 8: The Equivalent RLC Circuit Values for the Two Different States of the PIN Diode. ..................................... 102
Table 9: Final Dimensions of the 3-Printed Monopole ESPAR Antenna. ................................................................. 105
Table 10: Varactor Circuit Model Parameters ........................................................................................................ 125
Table 11: Dimensions of the 5-Element Wire Monopole ESPAR Antenna. ............................................................. 129
1 Introduction

The main objective of ROADART project is to investigate and optimise the integration of communication units into trucks in the framework of intelligent transportation systems (ITSs). Due to the size of a truck-trailer combination, the architecture approaches investigated for passenger cars cannot be applied. Moreover, it is totally clear that single antenna communication systems will not be sufficient to fulfil the requirements for improved performance. Therefore, novel architecture concepts, communication techniques, and antenna designs should be developed and their performance should be evaluated in order to assure a sufficient quality of service (QoS) for trucks and heavy duty vehicles. The requirements for large vehicles like trucks as well as the peculiarities of the time varying wireless medium are taken into account, initiating the need for new system architectures, e.g., in terms of system partitioning, diversity, and antenna concepts.

1.1 Objectives of WP3

In the framework of WP3, ROADART investigates, evaluates, and measures multi-antenna transmission/reception techniques that can be tailored to the specificities of the truck-to-truck (T2T) and truck-to-infrastructure (T2I) communication links. The scope was to propose specific robust and adaptive techniques, algorithms, strategies, and architectures that significantly improve the performance over the specific vehicular radio channels. More specifically, with the use of the WP2 results and channel models, ROADART:

- Evaluated alternative transmit/receive diversity techniques for T2T and T2I mobile radio channels taking into account the tradeoff between performance and receiver complexity.
- Investigated conventional and alternative beamforming techniques for T2T and T2I channels.
- Introduced the use of parasitic antenna in order to significantly improve beamforming capabilities with the use of compact antennas and a small number of vehicle-mounted radio-frequency (RF) chains.
- Investigated the adoption of spatial modulation (SM) diversity techniques and design robust, low-complexity SM receiver architectures.
- Designed novel, efficient, and functional cooperative communications schemes in order to increase robustness, reliability and throughput.
- Determined an optimal but realistic antenna positioning subjected to specific constraints imposed by the vehicle.
- Alternative configurations of switched parasitic arrays (SPAs) and electronically steerable passive array radiator (ESPAR)s were designed and specific performance measures investigated.

The purpose of this deliverable is to propose the final qualified communication techniques and antenna arrays to be implemented in the ROADART platform. Therefore, in addition to all the above-mentioned objectives, several constraints, coming from other WPs, were taken into consideration for providing the final recommendations. More specifically, several constraints exist due to the system architecture that is proposed in WP4. Moreover, constraints also exist when trying to integrate the proposed solutions in the trucks, due to space limitation and/or power supplies issues. Therefore, the final recommended solutions should satisfy the objectives of this WP subject to the constraints coming from other WPs.
1.2 Structure of the Document

The structure of this document is as follows. In Section 2, the simulation platform that was developed and is based on the IEEE 802.11p standard is described, which is considered as the underlying protocol for intervehicular communications (IVCs). In Section 3, the investigation of the negative consequences of the various wireless medium peculiar characteristics for the T2T and T2I wireless communication systems is presented. In this context, the impact of spatial correlation, interference, and outdated channel state information (CSI) on the performance of IVC systems has been investigated in terms of analytical and simulation studies. Moreover, in the same section, the study of the connectivity probability of vehicular ad-hoc nodes is also provided. In Section 4, various novel low complexity algorithms and architecture designs for T2T communications are presented, including reconfigurable antenna schemes, extended open loop beamforming, and spatial modulation. Moreover, in this section, novel cooperative relaying schemes are also presented and discussed. In Section 5, the research regarding the antenna designs and prototyping is presented. Several antenna designs were simulated, fabricated, and experimentally verified. Furthermore, an investigation was carried out concerning the optimal antenna location on the truck. In Section 6, based on the solutions presented in the previous sections as well as several constraints, the final ROADART recommended techniques are presented along with the qualified ESPAR antenna designs. Finally, the conclusions of the deliverable are provided in Section 7.
2 IEEE 802.11p Simulator

A major objective in this WP was to prepare a simulation platform that is based on the IEEE 802.11p standard, which is considered as the underlying protocol for IVC. This platform may be used to test, evaluate, and examine the PHY and medium access control (MAC) layers of T2T/T2I links as well as the performance of various diversity reception techniques.

2.1 PHY Layer

The following block-diagram represents the standard’s proposed system structure. The system can be divided into 3 main parts, transmitter, receiver, and channel. As a result, an object-oriented approach was selected and an object for each part was created. Besides the object specific functions, a common-use set of functions was created, like fast Fourier transform (FFT) and Inverse Short Training Long Training Frame Preamble. 

Figure 1: 802.11p simulator model.
FFT (IFFT) implementation. Transmitter’s model includes adaptive modulation and coding mechanism, supporting 4 types of modulations and three coding rates. Also, scrambling and interleaving functions are available. Data symbols are modulated through orthogonal frequency division multiplexing (OFDM), which is easy to achieve via proper symbol mapping and use of
FFT. Furthermore, short training preamble, long training preamble and signaling data are created, as spe-

![Diagram](image.png)

Figure 3: Receiver software diagram.

cified at IEEE802.11p standard. Receiver’s model, on the other hand, consists of the transmitter’s “mirror” functions, such as de-mapper, de-modulator, etc. Essential receiver’s functions are
considered to be channel estimation, where CSI is given by long training preamble, and detector, where signal sensing is achieved from reaped correlations of short training preamble. The detection algorithm that has been used is based primarily on Shi-Serpedin proposed algorithm [1].

The software diagrams shown in Figure 2 and Figure 3 offer a more vivid image of the transmitter and receiver functions. Note that, when a vector is followed by a [1 x n] notation, is one-dimensional, where [m x n] denotes an m x n matrix.

2.2 MAC Layer

In the following paragraph, a description of the developed MAC simulator is presented. The basic operation of the MAC sublayer in IEEE 802.11p is presented in Figure 4 and Figure 5. The IEEE 802.11p is a random access protocol utilizing a carrier sense multiple access with collision avoidance (CSMA/CA) technique. Distributed radio access is implemented using the enhanced distributed channel access (EDCA) function.
Figure 4: CSMA/CA implementation in IEEE 802.11p.
The simulator is implemented in MATLAB with an object-oriented approach. Five main classes are defined:

1. The ITSG5_MAC class that initialize global properties for the MAC layer of all network nodes
2. The ITSG5_Simulator class that implements the functionality of an ITSG5 network with multiple network nodes. The ITSG5_simulator_loop method implements the main simulator actions. The ITSG5_Simulator contains and manages the simulator clock, i.e., the simulated time line for the network operation
3. The ITSG5_Transceiver class implements the PHY and MAC procedures per network node. Each network node in the simulator uses an instance of the ITSG5_Transceiver class. The ITSG5_Transceiver inherits properties from the ITSG5_MAC class
4. The ITSG5_Transmitter is a class-property for the ITSG5_Transceiver. ITSG5_Transmitter implements all the PHY functions and operations as described in Subsection 2.1 for transmitter operation. ITSG5_Transceiver controls MAC operation and assigns transmitting operation to its ITSG5_Transmitter property
5. The ITSG5_Receiver is a class-property for the ITSG5_Transceiver. ITSG5_Receiver implements all the PHY functions and operations as described in Subsection 2.1 for receiving operation. ITSG5_Transceiver controls MAC operation and assigns receiving operation to its ITSG5_Receiver property

During the simulator initialization stage, one ITSG5_Simulator instance is produced that performs the main network/simulator tasks. Moreover, based on the selected user generation procedure (implemented as a method in the simulator class), new network nodes are generated either in the initialization stage or continuously during the simulator loop. New network nodes are generated with new ITSG5_Transceiver instances. Each ITSG5_Transceiver instance retains as properties one ITSG5_Transmitter instance and one ITSG5_Receiver instance. At all times, each ITSG5_Transmitter uses either the receiver or transmitter operation.

At this stage, and based on the description of the protocol, the following five transmission types are supported:

- Broadcast – i.e., a transceiver gains access to the medium and broadcasts a QoS data frame. No ACK is expected
- Multicast – i.e., a transceiver gains access to the medium and sends a QoS data frame to a group of users. No ACK is expected
- Unicast without ACK – i.e., the transceiver sends directly a QoS data frame to a specified destination but it does not require an ACK
- Unicast with ACK – i.e. the transceiver sends directly a QoS data frame to a specified destination and an ACK is expected as a response.
- RTS-CTS Unicast with ACK – i.e. the transceiver sends an RTS (ready to send) frame towards a destination. A CTS (clear to send) response is expected. When the CTS is received, then a QoS data frame is send with an expected ACK as a response. RTS-CTS type of transmission is expected for frames with MPDU size greater than 1Kbyte.

The simulator supports the following types of Frames:

- Management frames:
  - Action frames
  - Time advertisement frames
- Control Frames:
  - RTS
  - CTS
  - ACK
- Data Frames:
  - QoS data (since EDCA is used)
  - Null (without practical use for the simulator)

The following status are defined per network node:

0. Idle – Sensing:
1. Waiting to Tx (transmitter) – Sensing:
2. Transmitting (data or ACK)
3. Waiting to transmit ACK
4. Receiving
5. Waiting to receive ACK

In order to implement the slotted operation of CSMA/CA, the simulator implements a time line in nano-seconds. The time line is updated with the use of a “while” loop (until the end of the simulation). The time line is increased using the following rationale:

- Simulator global time increases in slot duration steps, where slot duration is the MAC slot time duration in nanoseconds. The exception in this procedure is the existence of an event at a time instance less than the current slot duration. The existence of an event is specified by a number of counters retained by each network node that participates in the simulator.
- Each network node (user) retains the following counters:
  - Timers that count short interframe spacing (SIFS), arbitration interframe space (AIFS), or extended interframe space (EIFS) duration. (AIFS, SIFS, and EIFS counters – AIFS counter is a 4-vector, since four QoS queues are defined by the standard)
  - Timers that implement the contention procedure for each node and each priority group of data (contention window (CW) timers)
  - Timers that count the duration of the currently transmitted packet from other sources (information acquired with demodulation of the NAV field)
  - Timer that counts the remaining time for transmission for a packet originating by the transceiver (Tx Timer)

All counters are initialized (based on an event) and continuously reduced until reaching zeros. Zeroing of a timer constitutes an event. The simulator time controller is depicted in Figure 6. The general flow of the simulator is described in Figure 7.

2.2.1 Initialization of Users:
Each generated user initially has:
- No data to send
- No information about adjacent network nodes

Therefore, no a-priori knowledge is available at each transmitter.
2.2.2 Data Generation Procedure:
Initially, each user has no data. Based on a predefined method, new data are produced stochastically with a certain rate during each time progression step. Data are produced with a different rate for each QoS data queue of each transceiver. Moreover, the size of the currently produced data frame is stochastically determined. Therefore, the current data frame size is determined randomly between 200 bytes up to 4Kbytes. In Figure 8, the data generation procedure is depicted.

![Data Generation Procedure Diagram](image)

Figure 6: Simulator time controller.

2.2.3 Simulator Actions per Status:
The analysis in the following paragraphs is implemented per transceiver Status.

2.2.3.1 Status 0:
When a node is in status 0, then:
- There are no available data into the QoS data queues to compose a full frame
- The node operates as a receiver, sensing the medium
- The node is operating as a receiver performing carrier sense.
- New data are created during each time step. When the data in one or multiple queues are enough to compose a full frame, then the node moves to Status 1.

The receiving operation produces a decision regarding the medium status. If medium status is busy, then the receiver demodulates the headers in order to:
- Update NAV counters and determine the end of the transmission
- Decide if the node is the destination for the specific frame. In this case, the node is moving to State 4 and demodulates.

All the above are shown in Figure 9.

Figure 7: Basic flow of the Simulator loop.
Initialization of User Data \[ i = [ ] \]

\[ i = 1 \]

\[ i <= \text{Num Queues} \]

If Data \[ [i] = [ ] \]

Determine the size of the current Frame \[ \text{curFrameSize}[i] \]

Generate new data for the current time step

Update Data \[ [i] \]

If Data \[ [i] \] >= curFrameSize\[i\]

Frame queued for Tx

While not End of Simulator

---

**Figure 8:** Data generation procedure.

### 2.2.3.2 Status 1:

When a node is in status 1, then:

- There are available data into the QoS data queues to compose a full frame
- The node has initialized and it continuously updates
  - AIFS counters
  - CW counters (if a collision has been already sensed in previous instances)
- The node operates as a receiver, sensing the medium
- If during the sensing procedure, a signal is sensed
- The node reinitializes all AIFS counters
- The node pauses all CW counters
- It remains in State 1, and it tries to extract Destination and NAV information.

If the identified destination is the ID of the node, then the node moves to State 4 and demodulates the signal. If no signal is sensed, and AIFS and CW counters are zeroed, then the node will transmit data and it moves in State 2. If more than on AIFS/CW counter are zeroed simultaneously, then internal collision is detected. The queue with the highest priority is qualified, while Backoff procedure and AIFS counters are reinitialized for the rest of the queues.

All the above are shown in Figure 10.
2.2.3.3 Status 2:
When a node is in status 2, then:
- The node is in transmitter node
- The frame with the highest order from the queue that won contention is transmitted
- If during the current time period, transmission is not completed (indicated by the Tx timer), then the node remains at State 2 until completion
- If Tx timer is zeroed (i.e., transmission is completed), the basic transmission scheme is used (i.e. no ACK) and the node has more data to send then it moves to State 1. If no other data are available, then the node moves to State 0
- If Tx timer is zeroed (i.e., transmission is completed) and ACK or CTS is needed, then the node moves to State 3.

All the above are shown in Figure 11.
2.2.3.4 Status 3:
When a node is in status 3, then:
- The node is waiting to receive an ACK for a frame send during its previous state
- The node will wait for duration EIFS for ACK

Figure 11: State transition for current State = 2.

Figure 12: State transition for current State = 3.
During the EIFS waiting period, the medium should be determined as busy. If EIFS expires with no reception of an ACK, then the node determines that a collision occurred since no response from the destination was received.

If the medium is sensed as busy, then the node moves to Receiver node. After demodulation of the received signal, the node will determine if the desired ACK was received (successful transmission) or a different signal was received (collision detected). All the above are shown in Figure 12.

2.2.3.5 Status 4:

When a node is in status 4, then:
- The node receives and demodulates the signal
- It is assumed that the node has identified itself as a destination of the signal

If the NAV timer for the received frame has not yet expired, then reception continues and the node remains at State 4.

If data reception is completed then:
- If no ACK is needed, then it moves in State 0 or State 1 depending on the availability of data.
- If no ACK is needed, however CRC does not check and collision is detected, CW timers are properly updated.
- If ACK is needed and collision is detected, then the node moves in State 0 or State 1 depending on the availability of data with proper adjustment of CW timers.
- If ACK is needed and no collision is detected, then the node moves in State 5 (waiting to transmit an ACK).

All the above are shown in Figure 13.
2.2.3.6 Status 5:
When a node is in status 5, then it waits SIFS duration and then transmits an ACK for a frame received during its previous state that needs acknowledgement.
If SIFS expires and the medium is considered free, then the node moves to Transmitting Mode State 2 and it sends the ACK.
If during SIFS, the medium status changes to busy, then collision is detected and the node moves either to State 0 (no data available) or State 1 (data available – with necessary CW timer adjustment). All the above are shown in Figure 14.

2.3 Summary
In this section, the new MATLAB/OCTAVE simulator for both PHY and MAC layers of IEEE 802.11p standard that was developed in the framework of this WP is presented. In particular, all the functionalities of the PHY and MAC layers have been developed, based on the latest releases of this standard. Moreover, this simulation platform was utilized to evaluate the performance of all major receive diversity reception techniques for T2T and T2I communications, which have been also implemented in the platform.
3 Investigations of Impairment Parameters and Performance Analysis in IVC Systems

An important objective of this WP was to identify the negative consequences of the various peculiar parameters for the T2T and T2I wireless communication systems. In particular, the characteristics of the T2T communications differ from those of the traditional cellular ones, due to various factors. More specifically,

- Dimension/size constraints exist for both the transmission and reception equipment. Thus, small distances among the antennas are unavoidable, which will result to spatial correlation among the received signals.
- Simultaneous transmissions cannot be easily avoided, due to the random ad-hoc nature of the MAC protocol that is employed in the current ITS-G5 standard. Thus, increased levels of interference are expected.
- The T2T wireless medium is characterized by high mobility, since both the Tx and the receiver (Rx) as well as many of the important scatterers, continuously move. Thus, it is not surprising that the estimation of the channel conditions will not always be accurate, since due to the high mobility its condition will be rapidly changing.

Therefore, motivated by all the above-mentioned observations as well as by the absence of relevant studies on the open technical literature, the negative consequences of spatial correlation, interference, and outdated CSI on the performance of IVC systems have been investigated in terms of analytical and simulation studies. It is noted that the research was mainly based on a widely-employed channel model for V2V communications, which is the double-scattering model. The main idea for this generic model is that depending upon the distance between the Tx and Rx; one or more groups of scatterers are participating in the propagation, by employing transmission paths from each scatterer in one group, to scatterers in other groups. Thus, this forwarding mechanism between the groups of scatterers, defines a new transfer function, which in general consists of a mix of single, double, triple etc. Rayleigh distributed components [2]. Therefore, even though under certain propagation conditions, multi-Rayleigh distributions will become dominant, in general, it is more likely that a mix of a single, a double Rayleigh, and line of sight (LoS) component will exist. In addition to its relatively simple analytical form, the double-scattering model has also been verified in various experimental studies that are based on measurements in different mobile-to-mobile communication environments.

3.1 Spatial Correlation

Regardless of the channel model that has been adopted, an important factor that seriously affects the system’s performance is the existence of correlation among the diversity branches. In general, signal correlation exists in cases where the distance among the diversity antennas is small. In this context, since the potential antenna mounting positions have small dimensions, e.g., mirrors, it will not be surprising that correlation effects will also be present in T2T communication scenarios. Motivated by this observation, in the framework of this WP, the influence of the correlation effects in a T2T communication scenario has been studied for the first time. In order to statistically characterize the correlation between the received signals from the different antennas, the bivariate double-Rayleigh distribution was introduced. Based on this new distribution, it is possible to describe in analytical form, important statistical metrics of the received instantaneous SNR of the output of a selection diversity (SD) receiver. Using the analytical results, the performance deterioration due to the existence of spatial correlation of a dual-branch SD that operates in T2T communication environment was investigated.

3.1.1 Correlated Double Rayleigh Statistics

Let \( X_i \) (\( i = 1, 2, 3 \)) denoting the envelopes of zero mean complex Gaussian random variables (RVs) with marginal probability density function (PDF) given by [3]
where $\Omega_i$ is the mean square value. The double-Rayleigh distribution is a product of two independent Rayleigh RVs [4]. In this context, let $Z_j$, with $j = 1, 2$, be defined as

$$ Z_1 = X_1 \times X_2 $$

$$ Z_2 = X_1 \times X_3 $$

This investigation was based on the fact that $X_2$ and $X_3$ are correlated Rayleigh RVs with joint PDF given by [4]

$$ f_{X_2,X_3}(x,y) = \frac{4xy}{\Omega_1\Omega_2(1-\rho)} \exp\left[-\frac{1}{1-\rho}\left(\frac{x^2}{\Omega_1} + \frac{y^2}{\Omega_2}\right)\right] I_0\left[\frac{2\sqrt{\rho xy}}{(1-\rho)\sqrt{\Omega_1\Omega_2}}\right] $$

(3)

where $I_v(.)$ is the modified Bessel function of the first kind and order $v$ [5] and $\rho = \text{cov}(X_2^2, X_3^2)/\sqrt{\text{var}(X_2^2)\text{var}(X_3^2)}$, with $\text{cov}$ and $\text{var}$ denoting the covariance and the variance, respectively, is the power correlation coefficient ($0 \leq \rho < 1$) [11]. Since $Z_1$ and $Z_2$ constitute products of two RVs, their joint PDF is given by

$$ f_{Z_1,Z_2}(z,w) = \int_0^\infty \frac{1}{x_i} f_{X_1}(x_i) f_{X_2,X_3}\left(\frac{z}{x_i}, \frac{w}{x_i}\right) dx_i. $$

(4)

Substituting (1) and (3) in (4), employing the infinite series representation of the Bessel function, [5] eq. (8.445), making a change of variables, and using [5] eq. (3.471/9), yields the following expressions for the joint PDF of $Z_1$ and $Z_2$

$$ f_{Z_1,Z_2}(z,w) = \frac{8}{\Omega_1\Omega_2(1-\rho)} \sum_{k=0}^\infty \frac{1}{(k!)^2} \left(\frac{\rho}{\Omega_1}\right)^{k+1/2} \left(\frac{\rho}{\Omega_2(1-\rho)^2}\right)^k \left(\frac{z^2}{\Omega_1} + \frac{w^2}{\Omega_2}\right)^{k+1/2} K_{2k+1}\left[\frac{2}{\sqrt{(1-\rho)\Omega_1\Omega_2}}\sqrt{\frac{z^2}{\Omega_1} + \frac{w^2}{\Omega_2}}\right] $$

(5)

where $K_v(.)$ is the modified Bessel function of the second kind and order $v$ [5] eq. (8.432/1). Assuming identical mean square values, i.e., $\Omega_i = \Omega$, the joint cumulative distribution function (CDF) of $Z_1$ and $Z_2$ is given in

$$ F_{Z_1,Z_2}(z,w) = \frac{2}{\Omega^2} \sum_{k=0}^\infty \frac{1}{(k!)^2} \left(\frac{\rho}{\Omega^2}\right)^k \left(\frac{1}{1-\rho}\right)^{k+1/2} \left\{\frac{1}{2} \sum_{p=0}^{2k} \binom{2k}{p} \left(\frac{w^2}{z^2}\right)^p \zeta^{2(p+1)} \xi w^{2(1+k)} + (-1)^k \zeta^{2(k+1)} w^{2(k+1)}\right\} $$

$$ \times \left\{\frac{(-1)^k}{k+1} \sum_{t=2}^{k+1} \frac{1}{k+2-t} F_{k+1,k+2-t;k+2;\frac{w^2}{z^2}} \ln(z^2) + \left(1 - \frac{z^2}{w^2}\right)^{k+1} \ln\left(1 + \frac{w^2}{z^2}\right)\right\} $$

WP3 – T2X Communication Techniques © ROADART consortium
3.1.2 SD Receivers Performance

Let us consider a dual-branch SD receiver operating over an additive white Gaussian noise (AWGN) channel, subject also to flat fading, modeled by the previously discussed distribution. The instantaneous signal to noise (SNR) at the ith branch, with \( i = 1, 2 \), is defined as

\[
\gamma_i = \frac{Z_i^2 E_s}{N_0}
\]

with \( E_s \) denoting the transmitted symbol energy and \( N_0 \) is the power spectral density of the AWGN channel. The corresponding average SNR is given by \( \bar{\gamma}_i = E(Z_i^2)E_s/N_0 \). Additionally, since the instantaneous SNR at the output of the SD receiver is \( \gamma_{sd} = \max\{\gamma_1, \gamma_2\} \), its CDF can be expressed as \( F_{\gamma_{sd}}(\gamma) = F_{\gamma_1}(\gamma) \). Using (6), assuming \( \gamma_i = \bar{\gamma} \) and based on the information provided above, it is not difficult to recognize that the CDF of \( \gamma_{sd} \) is given by

\[
\text{CDF of } \gamma_{sd} = \int_0^\infty \frac{1}{\Gamma(m+n)} \Gamma\left(1+k, m-1\right) \Gamma\left(1+k, n-1\right) F_1\left[m, n ; 1, 2 \right] \frac{e^{-x}}{x^{m+n+k}} dx
\]
Next, based on the analytical expression for the CDF of the output SNR, the performance was investigated using the outage probability (OP) and the average bit error probability (ABEP).

3.1.2.1 Outage Probability
OP is defined as the probability that the SNR falls below a predetermined threshold $\gamma_{th}$ and is given by.

$$P_{out} = F_{\gamma_{sd}}(\gamma_{th})$$

Asymptotic Analysis: In order to clearly understand important system-design parameters, we studied the asymptotic OP. This approach helped us to quantify the amount of performance variations, which are due to the fading effects as well as to the receiver’s architecture. For higher mean square values of $\Omega_i$, the PDF of $X_i$ can be closely approximated by

$$f_{X_i}(x) \approx 2x/D_i.$$ Based on this approximated expression, and by following a similar procedure as the one for deriving (6), the CDF of $\gamma_{sd}$ simplifies to the following expression

$$F_{\gamma_{sd}}(\gamma) \approx \sum_{k=0}^{\infty} \frac{1}{(k!)^2} \left( \frac{\rho}{\gamma} \right)^k \frac{1}{1-\rho} \left[ 1 + \frac{1}{2} \frac{F_1(1)}{\Gamma(2k+1-p)} \gamma^{p+1} + (-1)^2k+2 \right]$$

$$\times \left[ \sum_{p=0}^{\infty} \frac{(1/\xi)^{2p+2k+1}}{p!(2k+p+1)!} \sum_{m=0}^{p} \frac{p^m}{m} \left[ \ln \frac{\sqrt{2\gamma/\xi}}{p+k-m+1} - F_2(1) - F_2(1) - F_2(1) \right] \right]$$

where

$$F_2 = \frac{(-1)^{k+1} \xi}{k+1} \sum_{i=2}^{k+1} \frac{k+1}{k+2-i} \left( 1 - (-1)^{k+1} \right) \ln(2) + \sum_{i=1}^{k+1} \frac{(-1)^i}{k-i+2} + \frac{1}{k+1}$$

$$F_5 = \frac{1}{2} \sum_{m=0}^{p} \frac{p^m}{k+p-m+1} \psi(p+1) - \psi(2k+p+2) \frac{1}{k+m+1}.$$}

From (11), it is obvious that $\rho$ will not affect the diversity order, which is always 2, but only the coding gain.

3.1.2.2 Average Bit Error Probability (ABEP)
Using the CDF-based approach the ABEP is given by [9]
\[
P_b = \int_{0}^{\infty} -P_e(\gamma) F_{\gamma,d}(\gamma)d\gamma.
\]  
(12)

where \(-P'_e(\gamma)\) denotes the negative derivative of the conditional error probability. For example, assuming differential binary phase shift keying (DBPSK) modulation \(-P'_e(\gamma) = \alpha \beta \exp(-\beta \gamma)\), where \(\alpha = 1/2, \beta = 1\) [9]. Substituting (9) in (12), employing [7] eq. (2.6.21/2) and after some mathematical manipulations yields

\[
P_b = \frac{2}{\beta^2} \sum_{k=0}^{\infty} \frac{1}{(k!)^2} \left( \frac{\rho}{\beta^2} \right)^k \left( \frac{1}{1 - \rho} \right)^{k+1} \left\{ \frac{1}{2} \sum_{p=0}^{k} \frac{F_1(p)}{F_2(1)} \right\} + F_4 \left( \frac{1}{2} \right) \left( -1 \right)^{2k+2} \left\{ \sum_{p=0}^{\infty} \frac{1}{p!} \right\}
\]

\[
\frac{1}{(2k+p+1)!} \frac{\Gamma(2k+p+3)}{\gamma^{2p+2k+1}} \left[ 2 \ln \left( \frac{2}{\xi} \right) + \psi(2k+p+3) - \ln(2) \right] - F_2(1) - F_3(1) - F_5 \left\{ \right\}.
\]

(13)

Here, it is important to note that the derivation of (13), and thus the investigation of the ABEP performance, is only possible by using the analytical approach presented in the previous subsection and the CDF expression given in (9).

Asymptotic Analysis: For the higher mean square values of \(\Omega\), substituting (11) in (12), and after some mathematics, the following convenient closed-form expression for the ABEP can be derived

\[
P_b \approx \frac{1}{\beta^2} \left[ \ln(2) + \ln \left( \frac{1+\sqrt{1-\rho}}{2} \right) - \frac{\ln(1-\rho)}{2} \right].
\]

(14)

3.1.3 Numerical Results

In this section, various numerical performance evaluation results will be presented in terms of the OP and the ABEP. Firstly, the rate of convergence of the infinite series given in (9) has been investigated. More specifically, the minimum number of terms, which guarantees accuracy better than \(\pm 10^{-5}\) is presented in Table 1 versus \(\gamma\), for different values of \(\rho\) and \(\bar{\gamma}\). It is clear from these results that only a relatively small number of terms is necessary to achieve an excellent accuracy, while this number is significantly smaller than the corresponding ones for similar studies. Moreover, the number of terms increases as \(\gamma, \rho\) increase as well as with the decrease of \(\bar{\gamma}\). It is worthwhile to mention that very similar results with Table 1 were also obtained by using the other infinite series expressions derived in this work, e.g., (13).

In order to evaluate the applicability of the analytical results in real-world conditions, the WINNER2 channel models were used [10]. The objective was to verify that the extracted distributions are representative for single input multiple output (SIMO) propagation through keyhole channels using measurement-based models. In keyhole channels, the radio environment in the proximity of both Tx and Rx contains multiple scatterers, while the propagation beyond Tx-Rx proximity and between them is clear with absence of complex propagation phenomena. In order to compose a keyhole with WINNER2, bad urban radio channels (non LoS (NLoS) B2 WINNER) were considered in Tx and Rx proximity, while in-between them rural D1 LoS radio channel was assumed.

Moreover, the following assumptions were also made: propagation at 5.9 GHz (ITS), a single-antenna dipole Tx, two closely spaced dipole antennas for the Rx with interelement distance \(\lambda/6\) (\(\lambda\) denoting the wavelength). Due to strong mutual coupling of the array elements, correlation between
the two Rx signals is expected. It is noted that transition between radio environments is not directly supported by WINNER2. Therefore, two virtual relays were considered in the borders of the

Table 1: Minimum Number of Terms of (9) For Accuracy Better Than ±10^5.

<table>
<thead>
<tr>
<th>γ</th>
<th>ρ=0.2</th>
<th>ρ=0.7</th>
<th>ρ=0.2</th>
<th>ρ=0.7</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3</td>
<td>7</td>
<td>1</td>
<td>5</td>
</tr>
<tr>
<td>5</td>
<td>6</td>
<td>8</td>
<td>1</td>
<td>7</td>
</tr>
<tr>
<td>10</td>
<td>9</td>
<td>11</td>
<td>2</td>
<td>8</td>
</tr>
</tbody>
</table>

Figure 15: Keyhole channel as simulated using WINNER2.

Figure 16: Histogram using simulation vs. theoretical PDF for Z1 and Z2.

propagation environments as seen in Figure 15. The virtual relays are using directional antennas (10° aperture), since only specific propagation paths in the direction of the Tx-Rx link pass through the keyhole achieving Tx-Rx connectivity. During simulations, 40,000 WINNER2 narrowband keyhole channels were produced. Due to the closely spaced elements of the array, the correlation coefficient in the Rx B2 channel was set to 0.85. This corresponds to ρ = (0.85)^2 as defined below (3). In Figure 16, histograms for the normalized Rx signal amplitude for the two antennas as produced by the WINNER2 vs. the marginal PDF for the keyhole channels as calculated by (5) are presented. It is evident that the modeled channels can be accurately fitted by the theoretical PDF. Moreover, the correlation between the two WINNER2 simulated Tx-Rx channels was calculated and the result was compared with the theoretical one. Simulations show that the presented
theoretical analysis can be used to successfully characterize keyhole radio channels for SIMO systems.

In Figure 17, using (9) and assuming $\gamma = 9$dB, the OP of dual-branch SD is plotted as a function of the correlation coefficient $\rho$, for different values of the outage threshold, $\gamma_{th}$. It is depicted that as $\rho$

![Figure 17: OP vs $\rho$, for different values of $\gamma_{th}$.](image)

increases, the performance decreases. This decrease is higher for increased values of $\rho$, i.e., $\rho \geq 0.8$, where the diversity gain rapidly minimizes. Moreover, the performance improves with a
D3.1 – Recommendation of Antennas and Communication Techniques Qualified for Implementation  H2020 - 636565

decrease on $\gamma_{th}$. In the same figure, the OP which is based on the asymptotic analysis, obtained using (11), is also included. It is shown that the difference between the exact and the approximated values decreases with the decrease of the switching threshold.

In Figure 18, using (13), the ABEP is plotted as a function of the average input SNR $\bar{\gamma}$, for different values of the correlation coefficient $\rho$. Moreover, based on (14), the asymptotic ABEP is also included. It is depicted that for lower values of $\rho$ the performance is better. It is important to note that the asymptotic curves approximate quite well the exact ones even for moderate values of the average SNR, i.e., $\bar{\gamma} \geq 10$dB, while this approximation improves for lower values of $\rho$. Finally, it is noted that the computer simulations performance results, which are also included in all figures, verify in all cases the validity of the proposed theoretical approach.

3.2 Interfering Effects

As it is well known diversity techniques can improve the performance of T2T communication systems. However, in many practical situations, e.g., due to hidden terminal effect in CSMA based schemes, the performance of these systems can be significantly affected by co-channel interference (CCI). In this context, the influence of interfering effects in an IVC scenario was studied. In particular, based on the multiple scattering channel model, statistical behavior of multiple interfering signals has been investigated for the first time and used to analyze the performance of the scenario under consideration. Specifically, the QoS has been evaluated using the signal-to-interference ratio (SIR) statistics. Moreover, the performance improvement induced by the adoption of diversity techniques has been also analytically investigated.

3.2.1 System and Channel Model

In general, we considered a communication system with 1 transmitting and L receiving antennas, operating in a vehicular communication environment. In our study, LoS conditions exist between the Tx and the Rx, as it is shown in Figure 19. Moreover, Rx is also subject to interfering signals coming from various mobile sources, which however do not have LoS components, as it is also shown in Figure 19. Moreover, we also assume that (in general) the level of interference at the receiver is such that the effect of thermal noise on system performance can be ignored (interference limited scenario). Let us denote the instantaneous SNR of the desired received signal as

$$\gamma_{d_j} = \frac{|h_{d_j}|^2 E_s}{N_0}$$

with $h_{d_j}$ denoting the complex channel gain received at the jth branch, $E_s$ is the average transmitted signal energy and $N_0$ the AWGN power spectral density.

It is assumed that $|h_{d_j}|$ follows the Rice distribution. Thus, the PDF of $\gamma_{d_j}$ under independent and identically distributed (i.i.d.) fading conditions, is given by [3] eq. (2.16)

$$f_{\gamma_{d_j}}(\gamma) = \frac{(1+K)\exp(-K)}{\bar{\gamma}_d} \exp\left[\frac{(1+K)\gamma}{\bar{\gamma}_d}\right] I_0\left[2\sqrt{\frac{K(K+1)\gamma}{\bar{\gamma}_d}}\right]$$  \hspace{1cm} (15)

with

$$\bar{\gamma}_d = E\left(|h_{d_j}|^2\right) E_s/N_0$$

denoting average input SNR per branch, $K$ corresponds to the ratio of the power of the LoS component to the average power of the scattered component. Moreover, the corresponding expression for the CDF of $\gamma_{d_j}$ is given by

$$F_{\gamma_{d_j}}(\gamma) = 1 - Q_1\left(\sqrt{2K} \cdot \sqrt{\frac{2(1+K)\gamma}{\bar{\gamma}_d}}\right)$$  \hspace{1cm} (16)
where \( Q_1(., .) \) denotes the first order Marcum Q-function [3] eq. (4.33). Furthermore, let us denote the instantaneous received interference-to-noise ratio (INR) of the i-th interfering received signal, with \( i \in [1, \ldots, M] \) as \( \gamma_{i,t} = \frac{|h_i|^2 \rho_s}{N_0} = |h_i|^2 \rho_s \) with corresponding average INR equal to \( \bar{\gamma}_{i,t} = E(|h_i|^2) \rho_s \), where \( |h_i| \) denotes the channel gain of the i-th interfering signal with energy \( E_i \).

Furthermore, let us denote the instantaneous received interference-to-noise ratio (INR) of the i-th interfering received signal, with \( i \in [1, \ldots, M] \) as \( \gamma_{i,t} = \frac{|h_i|^2 \rho_s}{N_0} = |h_i|^2 \rho_s \) with corresponding average INR equal to \( \bar{\gamma}_{i,t} = E(|h_i|^2) \rho_s \), where \( |h_i| \) denotes the channel gain of the i-th interfering signal with energy \( E_i \).

Figure 19: Example of a LOS in SOS conditions.

Here, we have adopted the multiple-scattering radio channel for modeling the envelopes of the interfering signals. In this context, we focus our attention to the important case of second order scattering (SOS), which has been found to provide a good explanation for the signal envelope for V2V communication conditions. The SOS is characterized by the following impulse response [2]

\[
C_{2,i}^c = w_0 e^{j \theta} + w_1 H_{1,i} + w_2 H_{2,i} H_{3,i}.
\]

In (17), \( C_0 = w_0 e^{j \theta} \) is the LoS component with constant magnitude and uniformly distributed phase over \([0; 2\pi]\), \( w_0, w_1, \) and \( w_2 \) are non-negative real-valued constants that determine the mixture weights of the LoS, single as well as double scattering components. More specifically, by varying the parameters \( w_i \), different propagation conditions can be modeled, while (17) includes well-known fading distributions as special cases. For example, assuming \( w_2 = 0 \), it coincides to Rice distribution, for \( w_0 = w_1 = 0 \), it coincides to double-Rayleigh distribution and for \( w_0 = w_2 = 0 \), it coincides to Rayleigh distribution.

Here, for the interfering signals, we focus on a scenario of practical interest where only a combination of single and double scattering components exists, i.e., \( w_0 = 0 \). A potential communication scenario satisfying the assumptions made in this work is given in Figure 19. In that case the magnitude of second order scattering, \( |h_i| = |C_{2,i}^c| \), is given by [2] eq. (30)

\[
f_{|h_i|}(r) = 2 \exp\left(\frac{w_1^2}{w_2^2}\right) \sum_{m=0}^{\infty} \frac{(-1)^m \Gamma(-m, w_1^2 / w_2^2)}{m!} \left(\frac{w_1^2}{w_2^2}\right)^{m+1} r^{2m+1}.
\]
D3.1– Recommendation of Antennas and Communication Techniques Qualified for Implementation

H2020 - 636565

where \( \Gamma(\ldots) \) is the upper incomplete gamma function \([5]\) eq. (8.350/2). Based on the definition of \( \gamma_H \), it has been proved that the PDF of \( \gamma_H \) is given by \([11]\) eq. (38).

In the system under consideration, the instantaneous output SIR is given by

\[
\gamma_{\text{out}} = \frac{\gamma_X}{\gamma_I}
\]  

(19)

where \( X \equiv \{s,mrc,\text{sd}\} \) when referring to the instantaneous SNR of the single, maximal ratio combiner (MRC) and SD receivers, respectively, while \( \gamma_I \) denotes the total INR, i.e., \( \gamma_I = \sum_{i=1}^{M} \gamma_{I_i} \).

In \([10]\), it has been proved that the PDF of \( \gamma_I \) can be expressed as

\[
f_{\gamma_I}(\gamma) = \frac{1}{2} \exp \left( \frac{w_1^2 M}{w_2^2} \sum_{n=0}^{\infty} c_{n} \gamma^{M+n} \right) 
\]  

(20)

where

\[
a_0 = \frac{1}{w_2^2} \Gamma \left( 0, \frac{w_1^2}{w_2^2} \rho_s \right), c_0 = a_0^M
\]

\[
a_m = (-1)^m \frac{m!}{m!(m+1)} \Gamma \left( -m, \frac{w_1^2}{w_2^2} \rho_s \right), c_m = \frac{1}{m} \sum_{t=0}^{m} (m-t) a_t c_{m-t}
\]

It is noted that the PDF in (20) converges fast, since in most cases a relatively small number of terms is sufficient, i.e., \(<20\), to achieve a high accuracy. Based on (19), the CDF of \( \gamma_{\text{out}} \) is given by

\[
F_{\gamma_{\text{out}}}(\gamma) = \int_{0}^{\infty} F_{\gamma_X}(\gamma x) f_{\gamma_I}(x) dx.
\]  

(21)

Next, we examined different system and channel model communication scenarios. More specifically, we analyzed i) the influence of multiple interfering signals to the system performance of a single antenna system, and ii) the performance improvement induced by employing diversity in an interference limited scenario. It is obvious that in order to better understand the impact of all these contradictory parameters to the system performance, the best approach is to examine them separately.

3.2.2 Single Reception Scenario

Here, we consider a communication scenario where the Tx communicates with the Rx (with \(L=1\)) via a LoS. Moreover, Rx is also subject to interfering effects coming from \(M\) sources. In this context, substituting (16) and (20) in (21), where we have assumed \(X=s\), using \([13]\) eq. (10) and after some mathematics, yields the following expression for the CDF of \( \gamma_{\text{out}} \)

\[
F_{\gamma_{\text{out}}}(\gamma) = 1 - \exp \left( \frac{w_1^2 M}{w_2^2} \sum_{n=0}^{\infty} c_{n} \frac{\gamma_d^{M+n}}{(1+K)^{M+n}} \right) I_{F_1}(\gamma_{-M-n,1,-K})
\]  

(22)

where \( I_{F_1}(\ldots) \) denotes the confluent hypergeometric function \([5]\) eq. (9.210/1). Using (22), the OP can be studied. The OP is an important performance indicator for identifying the system’s QoS and is defined as the probability that the SIR falls below a predetermined outage threshold \( \gamma_{th} \) and is
given by $P_{\text{out}} = F_{\text{out}}(r_{th})$. In Figure 20, the OP of a single antenna communication scenario is plotted as a function of the number of interfering sources $M$. To obtain this figure, we have assumed strong and weak LoS conditions (for the desired signal), that is $K = 11.4$ and $K = 2$, respectively. Moreover, two scenarios regarding the interfering signals propagation conditions have been studied. In the first one, we have assumed $w_1 = 0.2$ (with $w_2 = \sqrt{1 - w_1^2}$), which results to dominant double-scattering components for the interfering signals, while in the second one $w_1 = 0.8$, which results to dominant single-scattering components. In Figure 20, it is shown that the performance improves as $K$ increases, i.e., strong LoS conditions exist for the desired signal. An interesting observation that comes out of this figure is that for weak LoS conditions ($K = 2$), the OP for both scenarios of scattering are very close. However, when strong LoS conditions exist, the best performance is when the single scattering components of the interfering signals are dominant. Thus, the single scattering propagation for the interfering signals results in lower INR and thus higher SIR.

3.2.3 Multichannel Communication Investigation

Here, we consider the impact of diversity techniques on the system performance. Two well-known schemes will be analytically investigated, namely MRC and the SD.

![Figure 20: OP of single antenna scenario vs M.](image)

3.2.3.1 Maximal Ratio Combining

We consider a communication scenario where the Tx communicates with a Rx (with $L$ antenna branches) supporting MRC reception via a LoS. Moreover, Rx is also subject to interfering signals coming from 1 source. We have assumed that the same interfering signals are present on each diversity branch [12]. In this context, the CDF of the MRC is given by [14] eq. (20)

$$F_{\text{MRC}}(\gamma) = 1 - Q_L\left(\sqrt{2KL}, \frac{2(1 + K)}{\gamma_d} \gamma\right)$$

(23)
where $Q_m(\ldots)$ denotes the generalized Marcum Q-function [3] eq. (4.59). Thus, substituting (20) and (23) in (21), where we have assumed $X \equiv \text{mrc}$, using again [13] eq. (10) and after some mathematics, yields the following expression for the CDF of $\gamma_{\text{out}}$

$$F_{\gamma_{\text{out}}} (\gamma) = 1 - \exp \left( \frac{w_1^2}{w_2^2} \sum_{n=0}^{\infty} \frac{(-1)^n}{(n+1)!} \Gamma \left( n, \frac{w_2^2}{w_1^2} \right) \frac{\Gamma(L+n+1)}{\Gamma(L)} (A_n)^{n+1} F_1(-n-1,L,-KL) \right)$$

(24)

where $A_n = \frac{\gamma_d}{(\rho_s \gamma_2^2 (1+K) \gamma)}$.

3.2.3.2 Selection Diversity

We consider a communication scenario where the Tx communicates with the Rx supporting SD (with $L = 2$ antenna branches) via LoS paths. We have also assumed that the same interfering signals are present on each diversity branch. In this context, the CDF of the SD is given by

$$F_{\gamma_{\text{sd}}} (\gamma) = \left[ 1 - Q_i \left( \sqrt{2K}, \sqrt{\frac{2(1+K)}{\gamma_d}} \gamma \right) \right]^2$$

(25)

In [15], it has been proved that the CDF of the output SIR for the scheme under consideration is given by

$$F_{\gamma_{\text{sir}}} (\gamma) = 1 - \frac{\sum_{n=0}^{\infty} B_n (A_n)^{n+1} \Gamma(n+1)}{K} F_1(-n-1,1,-K) + \sum_{n=0}^{\infty} B_n (A_n)^{n+1} \exp(-K) \frac{\Gamma(n+2)}{K} \frac{M_{-\frac{3}{2},0}(K)}{M_{-\frac{1}{2},0}(K)}$$

$$\times M_{-\frac{1}{2},0}(K) - \exp(-K) \sum_{\ell=1}^{\infty} \frac{(n+1)!}{2(\ell-1)!} \left( \frac{\pi}{2} \right)^{\ell-1} \left( \frac{\pi}{2} \right)^{n+1-\ell} \frac{\Gamma(n+1)}{\Gamma(n+1-\ell)}$$

$$\times \sum_{j=0}^{n+1-\ell+1} \left( \frac{\pi}{2} \right)^{\ell-1} \frac{\Gamma(n+1)}{\Gamma(n+1-\ell-j)}$$

$$\left( \frac{\pi}{2} \right)^{\ell+1} \left( \frac{\pi}{2} \right)^{n+1-j} \frac{\Gamma(n+1)}{\Gamma(n+1-j)}$$

$$\times \sum_{j=0}^{n+1-\ell+1} \left( \frac{\pi}{2} \right)^{\ell+1} \frac{\Gamma(n+1)}{\Gamma(n+1-j)}$$

$$\times \exp(-K) \frac{\Gamma(n+1)}{\Gamma(n+1-\ell)} \frac{M_{-\frac{3}{2},0}(K)}{M_{-\frac{1}{2},0}(K)} \frac{M_{-\frac{1}{2},0}(K)}{M_{-\frac{3}{2},0}(K)}$$

$$\times M_{-\frac{1}{2},0}(K)$$

(26)

where $B_n = (-1)^n \exp \left( \frac{w_1^2}{w_2^2} \right) \frac{\Gamma(n+1)}{\Gamma(n+1)}$ and $M_{\nu,\lambda}(.)$ is the Whittaker function [5] eq. (9.301).

Thus, by using (24) and (26), the OP of the diversity schemes under consideration can be directly evaluated.

In Figure 21, the OPs of MRC and SD receivers are plotted as a function of the average input SNR. To obtain this figure we have assumed $w_1 = 0.7$, which results to intermediate propagation conditions regarding the interfering signals, while different values of $\rho_s$ have been assumed. In this figure, it is shown that the QoS clearly improves, since the OP decreases, when diversity reception is employed, with MRC having always the best performance, as compared to single channel reception. Interesting observations that come out of this figure is that for lower values of $\rho_s$, the OP improves. This is a reasonable result since an increase on the fading severity in the interfering signals result to a lower INR and thus to a higher SIR. Finally, for the MRC case another plot with $L = 3$ branches is also included. In this plot, it is also verified that using MRC, important diversity gain is achieved, despite the negative consequences of the interfering signals.
### 3.3 Outdated Channel State Information (CSI)

In this part, the research focuses on a multichannel system operating in a T2T communication environment (modeled by the double-Rayleigh distribution) in the presence of AWGN. An antenna selection mechanism has been adopted, which can be implemented at either the Tx or the Rx side. In both cases, the system selects the antenna that provides the highest instantaneous channel gain, a decision that is based on the estimation of the CSI. In this context, the channel gain $Z_j$ that is available at the selection instance, due to the fast time varying nature of the medium, is different from the actual channel gain, $\hat{Z}_j$, at the transmission/reception instance [16]. The discrepancy between the channel gains is measured by the correlation coefficient $\rho$, defined in [17] eq. (9), and thus it depends on the maximum Doppler frequency $f_{Dj}$ and the time delay due to the CSI feedback $T_{Dj}$. In this case, the instantaneous SNR at the jth diversity branch, with $j = 1, 2$, is defined as $\gamma_j = |Z_j|^2\frac{E_s}{N_0}$ with $E_s$ denoting the transmitted symbol energy and $N_0$ noise variance.

![Outage Probability vs Average Input SNR](image)

Figure 21: OP of different diversity schemes vs the average input SNR.

The corresponding average SNR is given by $\bar{\gamma}_j = E(|Z_j|^2)\frac{E_s}{N_0}$. Assuming i.i.d. fading conditions, i.e., $\bar{\gamma}_j = \bar{\gamma}$, the CDF of the actual SNR of the selected branch at the data transmission instance is expressed as

$$F_{\gamma_{\text{out}}} (\gamma) = \frac{2}{\bar{\gamma}} \sum_{i,j=0}^{\infty} \frac{\rho_{ij}^2}{(i!)^2(i!)^2} \left( \frac{1}{\rho_{ij}^2} \right)^{\frac{j+i+2}{2}} \frac{\gamma^{i+j+2}}{\rho_{ij}^{i+j+2} \bar{\gamma}^{i+j+2}} \left[ G_{2,2}^{1,1} \left( \frac{-1-\frac{1}{2}}{1-\frac{1}{2}}, \frac{1}{1-\frac{1}{2}} \right) G_{1,3}^{2,1} \left( \frac{-1}{\rho_{ij}^2}, \frac{-1}{\rho_{ij}^2}, \frac{-1-\alpha_{ij}}{1-\alpha_{ij}} \right) \right]$$

The proof for (27) is given in [17]. Based on the analytical expression for the CDF of the output SNR in (27), the performance of the system under consideration can be evaluated using the criterion of OP that is given by $P_{\text{out}} = F_{\gamma_{\text{out}}} (\gamma_{\text{th}})$.

#### 3.3.1 Asymptotic Analysis

The exact results presented previously do not provide a clear physical insight of the system’s performance. As such, the main concern is to derive an asymptotic closed-form expression...
for $F_{\gamma_{\text{out}}}(\gamma)$. Therefore, here, we focus on the high SNR regime to quantify the amount of performance variations, which are due to the correlation as well as to the receiver’s architecture. In this context, assuming i.i.d. conditions, higher values of $\bar{\rho}$ and using the approach presented in [17], the following closed-form asymptotic expression for the CDF of $\gamma_{\text{out}}$ is obtained

$$F_{\gamma_{\text{out}}}(\gamma) \approx \frac{4\sqrt{\bar{\rho}^{1/2}}}{\sqrt{\bar{\rho}}} \left[ \sqrt{\bar{\rho}} K_i \left( \frac{2\sqrt{\bar{\rho}}}{\sqrt{\bar{\rho}}} \right) \right] \left[ 1 - \frac{1}{\bar{\rho}} \Phi \left( 1 - \frac{1}{\bar{\rho}}, 1, 2 \right) \right]$$  \tag{28}$$

where $\Phi(.,.,.,.)$ denotes the Lerch function [5] eq. (9.55). Moreover, assuming DBPSK, the average bit error probability (ABEP) can be evaluated as $P_{b} = (1/2) \int_{0}^{\infty} \exp(-\gamma) F_{\gamma_{\text{out}}}(\gamma) d\gamma$. Substituting, (28) in this integral and employing [5] eq. (6.631/3), the following approximating expression for the ABEP is obtained

$$P_{b} \approx \frac{1}{\bar{\rho}^{3/4}} \left[ 1 - \exp \left( \frac{1}{\bar{\rho}} \right) W_{-1,1/2} \left( \frac{1}{\bar{\rho}} \right) \right] \left[ 1 - \frac{1}{\bar{\rho}} \Phi \left( 1 - \frac{1}{\bar{\rho}}, 1, 2 \right) \right]$$  \tag{29}$$


3.3.2 Varying Correlation for Realistic Modeling
The previous analysis was based on deterministic values of $\rho_j$. However, in real-world situations, the vehicles (relative) velocity, i.e., $v_j = f_{D_j} \lambda$, and/or the time of arrival between data transmissions, i.e., $T_{D_j}$, may continuously change in a random manner. Under these circumstances, $\rho_j$ will also randomly vary. Therefore, it is reasonable to assume that correlation coefficients $\rho_j$ are given by $\rho_j = g(f_{D_j}, T_{D_j})$, e.g., assuming the classic Jakes spectrum $\rho_j = J_0 \left( 2\pi f_{D_j} T_{D_j} \right)$, where $J_0(.)$ is the Bessel function of the first kind [5] eq. (8.402). More specifically, we consider the case where $T_{D_j}$ are deterministic and thus the variations of $\rho_j$ depend only on $f_{D_j}$. The following analysis holds also in case where $T_{D_j}$ are stochastic and $f_{D_j}$ deterministic. Under these assumptions, (27) expresses the conditional CDF $F_{\gamma_{\text{out}}}(\gamma | \rho_1 = r_1, \rho_2 = r_2)$. In this context, the total OP involves the solution of the following integral:

$$F_{\gamma_{\text{out}}}(\gamma) = \int_{D} F_{\gamma_{\text{out}}}(\gamma | \rho_1 = r_1, \rho_2 = r_2) f_{\rho_1}(r_1) f_{\rho_2}(r_2) dr_1 dr_2. \tag{30}$$

A generic analytical calculation of (30) is unfeasible. An alternative approach is to approximate (30), with proper change of variables from $\rho_j$ to $f_{D_j}$, and discretization of the Doppler domain into $M$ equally spaced intervals resulting to the following formula

$$F_{\gamma_{\text{out}}}(\gamma) \approx \frac{1}{A} \sum_{m=1}^{M} \sum_{n=1}^{M} f_{\alpha} \left( f_{D_1}(m) \right) f_{\gamma_{\text{out}}} \left( \gamma | g\left( f_{D_1}(m) \right) \right) \tag{31}$$

where $f_{D_j}(m)$ is an $M$-point uniform sampling of the Doppler spread field, $f_{D_j}(x)$ is the PDF of $f_{D_j}$, and $A = \prod_{j=1}^{2} \sum_{m=1}^{M} f_{D_j}(f_{D_j}(m))$. 
3.3.3 Numerical Results

In Figure 22, assuming $\tilde{\gamma} = 15$dB and $\gamma_{th} = 3$dB, a contour plot of the OP is depicted as a function of the two temporal correlation coefficients $\rho_1$ and $\rho_2$. It is noted that the antennas are spatially uncorrelated. In this plot, it is shown that as the correlation coefficients increase, e.g., feedback delay diminishes in a transmit antenna selection scenario, the OP improves with an increased rate, while the impact of $\rho_1$ and $\rho_2$ to the system’s performance is identical. In the same figure, it is also depicted the probability of correct decision, which is plotted as a function of $\rho_j$s and obtained via simulations. From this subfigure, it is evident that for lower values of $\rho_j$, the antenna selection becomes random, while clear diversity gain is expected only for $\rho_j > 0.8$.

In Figure 23, assuming $\gamma_{th} = 5$dB, the OP is plotted as a function of the average input SNR for different values of $\rho_1$; $\rho_2$. It is also shown that the performance improves as $\rho_j$ increase. Moreover, in the same figure, the close agreement between the exact and the asymptotic (high SNR) OP is also depicted. It is noted that the approximation improves for lower values of $\rho_j$s. In addition, for comparison purposes, the corresponding performance of a single (without diversity) receiver is also depicted. It is shown that for lower values of $\rho_j$ s, i.e., fast time varying channel, the diversity gain is lost. In this figure, computer simulations performance results are also included, verifying in all cases the validity of the proposed theoretical approach. In Figure 24, assuming that the velocity of both ends of the link is a zero- mean Gaussian RV with $\sigma = 20$ m/sec (72 km/h), the total OP is calculated through simulation. It is noted that the distribution of $\rho_j$s for the specific case is presented as a histogram in the subfigure. In addition, the approximation of (31) as well as the OP from (27), where the mean values of $\rho_j$s are used, are also presented. It is interesting to note that (27) represents an upper bound of the total OP shown in (30), while the approximation given in (31) is quite close to (30). It is also noted that similar observations were also reported in [18], where it was depicted that outdated CSI affects seriously the system’s performance, even when good channel conditions exist [18].

Figure 22: OP vs the two correlation coefficients.
3.4 Nearest Neighbor Node Discovery

Ad-hoc networks are becoming increasingly important in future wireless networks. Transmission based on a variant of IEEE 802.11p wireless standard is the main method for enabling communication between mobile devices, including vehicular ad-hoc networks (VANETs) and wireless sensor networks. A common aspect of all mobile ad-hoc networks (MANETs) is the random position of transmitting and receiving nodes within the communication area. A model that is usually used to characterize the spatial distribution of these nodes is the Poisson point process (PPP). For example, the work in [19] studied the connectivity probability for V2V and V2I
communication scenarios under the assumption that vehicles are distributed on the road following a Poisson distribution. This spatial model assumes an infinite area with uniform distribution of the said nodes. Based on this stochastic model, the distance between adjacent nodes is a random variable.

Many works studied distance distributions in random infinite and finite networks [20]. For a random network with node distribution based on the PPP model and the node of interest located at the origin, the distance between this node and its n-th closest neighbor is a random variable that follows the generalized Gamma distribution [21]. Assuming constant transmit power, the received average SNR is determined by the path loss attenuation, which depends on the transmitter-receiver distance and the path loss exponent of the propagation environment. Therefore, the received SNR is also a random variable. However, most works in this area do not consider small-scale fading, e.g. [22], or consider the simple case of Rayleigh fading [23].

In WP3, we considered the combined effect of path loss and Nakagami-m fading on the outage probability of the n-th closest neighbor node [24]. Closed-form expressions for the PDF and CDF of the received SNR at the n-th closest neighbor were derived. Two examples of wireless applications, where the analytical results are utilized, were considered. In the first one, assuming the node of interest to be a transmitter, the derived expressions determine the probability of correct detection at the n-th nearest neighbor node in a PPP network with different node densities. This information can be used to determine the connectivity probability of broadcast transmissions, that is, to determine the farthest neighbor node that can successfully decode a packet. Secondly, if the node of interest is a potential receiver, the derived expressions provide insight on the probability of a neighbor node to interfere with the node of interest. Assuming that the potential receiving node transmits a beacon signal, e.g., it sends a CTS packet before reception, we can determine the received SNR of the beacon signal at the n-th nearest neighbor node. The distribution of the received SNR can then be used to determine the probability of the n-th nearest neighbor node to abstain from transmitting, so that it does not interfere with the node of interest.

3.4.1 System Model

3.4.1.1 Signal-to-Noise Ratio Analysis

The Poisson process is suitable for modeling uniformly random networks. For a homogeneous m-dimensional Poisson process in a bounded Borel set \( A \subset \mathbb{R}^m \) with intensity \( \lambda \), the probability that \( k \) points exist in \( A \) is

\[
\Pr\{k \text{ points exist in } A\} = e^{-\lambda \Omega(A)} \left( \frac{\lambda \Omega(A)}{k!} \right)^k, k = 0,1,...
\]  

(32)

where \( \Omega(A) \) is the Lebesque measure of \( A \). The Euclidean distance between a point and its n-th nearest neighbor, \( R_n \), is distributed according to the generalized Gamma distribution [22]

\[
f_{R_n}(r) = \frac{m}{r \Gamma(n)} \left( \frac{\lambda c_m r^m}{r} \right)^n \exp\left( -\frac{\lambda c_m r^m}{r} \right)
\]  

(33)

where \( c_m r^m \) is the volume of the m-dimensional ball of radius \( r \) and \( c_m \) is given by
D3.1 – Recommendation of Antennas and Communication Techniques Qualified for Implementation  H2020 - 636565

\[ c_m = \begin{cases} \frac{\pi^{m/2}}{\left(\frac{m}{2}\right)!}, & \text{even } m \\ \frac{\pi^{-m/2}2^m}{m!}\left(\frac{m-1}{2}\right), & \text{odd } m \end{cases} \] (34)

and \( \Gamma(n) \) is the gamma function evaluated at \( n \). In addition, if we want to consider only neighboring nodes that lie within a sector with opening angle \( \phi \), this simply corresponds to a change of the volume from an \( m \)-ball to an \( m \)-sector (with opening angle \( \phi \)) whose volume is \( c_{\phi,m} \). Therefore, the PDF of the distance to the \( n \)-th neighbor in a sector \( \phi \) is given by replacing \( c_m \) by \( c_{\phi,m} \) in (34). For \( m = 1, 2, 3 \), we have \( c_{\phi,1} = 1 \), \( c_{\phi,2} = \phi \), and \( c_{\phi,3} = (2\pi/3)(1 - \cos(\phi)) \), respectively.

Moreover, it is assumed that the transmitted signal undergoes small scale fading modeled by the Nakagami-\( m \) distribution. In that case, the corresponding instantaneous received SNR, \( X \), follows the gamma distribution with PDF

\[ f_X(x) = \frac{m_i^{m_i} x^{m_i-1} \exp\left(-\frac{m_i}{\Omega_s} x\right)}{\Gamma(m_i)\Gamma(\Omega_s)} \left(\lambda c_m\right)^n \] (35)

where \( m_i \) and \( \Omega_s \) are distribution’s shaping and scaling parameters. When the path-loss exponent \( \alpha \) is expressed as \( \alpha = \ell = k \), where \( \ell \) and \( k \) are integers, the PDF of the received SNR at the \( n \)-th closest neighbor is then given by

\[ f_{X_n}(x) = \int_0^\infty f_X(x|r)\,f_{R_n}(r)\,dr = \frac{x^{m_i-1}}{\Gamma(m_i)\Gamma(\Omega_s)} \left(\lambda c_m\right)^n \] (37)

The previous integral can be solved using the following result

\[ \int_0^x \exp\left(-B_1 x^h\right)\exp\left(-B_2 x^h\right) \, dx = \frac{1}{\sqrt{b_1 b_2} B_1}\frac{1}{\left(2\pi\right)^{1/2}}C_{b_1 b_2}^{-1} \Delta\left(b_1,1-\frac{v+1}{b_1}\right) \] (38)

where \( b_1 \) and \( b_2 \) are integers and \( \Delta(m,n) = m, \ldots, m+n-1 \). It is noted that Meijer G functions are built-in function in many mathematical software packages, e.g., Mathematica, Maple, and thus can be directly evaluated. For path-loss exponent expressed as \( \alpha = \ell = k \), where \( \ell \) and \( k \) are integers,
based on the solution given in (38) with $b_1 = \ell$ and $b_2 = km$, the PDF of the SNR at the n-th nearest neighbor node is given in closed-form as

$$f_{X_n}(x) = \frac{\left(\frac{m_s}{P_t}\right)^{m_s} (\lambda c_m)^n}{\Gamma(m_s) \Gamma(n)(2\pi)^{\frac{m_s}{2}}} \left(\frac{m_s x}{kmP_t}\right)^{m_s - 1} \left(\frac{m_s x}{kmP_t}\right)^{m_s - 1} \left(\frac{\lambda c_m}{kmP_t}\right)^n \sum_{k,m} \binom{am+mn}{km,1-kam+mn} \Delta(km,1-kam+mn) \Delta(l,0).$$

Moreover, the corresponding CDF of the received SNR is given by

$$F_{X_n}(x) = \int_0^x f_{X_n}(u) du = \frac{\left(\frac{m_s}{P_t}\right)^{m_s} (\lambda c_m)^n}{\Gamma(m_s) \Gamma(n)(2\pi)^{\frac{m_s}{2}}} \left(\frac{m_s x}{kmP_t}\right)^{m_s - 1} \left(\frac{m_s x}{kmP_t}\right)^{m_s - 1} \left(\frac{\lambda c_m}{kmP_t}\right)^n \sum_{k,m} \binom{am+mn}{km,1-kam+mn} \Delta(km,1-kam+mn) \Delta(l,0).$$

The CDF of $X_n$ is given in closed-form as

$$F_{X_n}(x) = 1 - \frac{\left(\frac{m_s}{P_t}\right)^{m_s} (\lambda c_m)^n}{\Gamma(m_s) \Gamma(n)(2\pi)^{\frac{m_s}{2}}} \left(\frac{m_s x}{kmP_t}\right)^{m_s - 1} \left(\frac{m_s x}{kmP_t}\right)^{m_s - 1} \left(\frac{\lambda c_m}{kmP_t}\right)^n \sum_{k,m} \binom{am+mn}{km,1-kam+mn} \Delta(km,1-kam+mn) \Delta(l,0).$$

where $b_j$ represents the jth term of $\Delta(l; 0)$ and $a_p$ represents the pth term of $\Delta(km; 1 - (2am_s + 2mn)/\ell)$. Moreover, based on the Taylor series approximation for the exponential function in (33), assuming higher values of $P_t$, i.e., $P_t > 70$ dB, and using the previously described framework, the following asymptotic expression for $F_{X_n}(x)$ is obtained

$$F_{X_n}(x) \equiv 1 - \frac{\left(\frac{\lambda c_m}{P_t}\right)^{am_s/m}}{\Gamma(m_s+1) \Gamma(n) \left(\frac{m_s x}{P_t}\right)^{am_s/m}}.$$

### 3.4.1.2 Signal-to-Interference Analysis

Assuming that received signal is also subject to interfering effect, the received SIR is given by
D3.1 Recommendation of Antennas and Communication Techniques Qualified for Implementation

\[ \gamma_{out} = \frac{X}{Y} \]  

(43)

where \( Y \) denotes the instantaneous received interference to noise ratio (INR). In that case, the PDF of \( \gamma_{out} \) is given by

\[ f_{\gamma_{out}}(x) = \left( \frac{m}{\Omega_s} \right)^{m_1} \left( \frac{m}{\Omega_i} \right)^{m_2} \frac{\Gamma(m_i + m_s)}{\Gamma(m_i)\Gamma(m_s)} \left( \frac{m_i + m_s}{\Omega_i + \Omega_s} \right)^{m_i + m_s - 1} x^{m_i - 1}. \]  

(44)

In (44), \( m_i \) is related with the severity of the interference and \( \Omega_i \) denotes the average INR. Based on (44), using [25] eqs. (10, 11, and 21) the PDF of the SIR at the n-th closest neighbor is given by

\[ f_{X_n}(x) = \frac{k^{m_i + m_s + m_{I, n} - 1}}{\Gamma(m_i)\Gamma(m_s)\Gamma(n)} \left( \frac{m_i \Omega_i}{P_i m_i} \right)^{m_i} \frac{m_i + m_{I, n} - 1}{2\pi} \frac{1}{(2\pi)^{3/2}} \frac{m_i + m_{I, n} + n/2}{(\lambda c_m)^{am_i + mn}} x^{m_i} \Delta(km,1-m_i-m_s), \Delta(1-\frac{am_i + mn}{m}), \Delta(km,0) \]  

\[ G_{km, km+n} \]  

(45)

The corresponding CDF expression is given by

\[ F_{X_n}(x) = \frac{k^{m_i + m_s + m_{I, n} - 1}}{\Gamma(m_i)\Gamma(m_s)\Gamma(n)} \left( \frac{m_i \Omega_i}{P_i m_i} \right)^{m_i} \frac{m_i + m_{I, n} - 1}{2\pi} \frac{1}{(2\pi)^{3/2}} \frac{m_i + m_{I, n} + n/2}{(\lambda c_m)^{am_i + mn}} x^{m_i} \Delta(km,1-m_i-m_s), \Delta(1-\frac{am_i + mn}{m}), \Delta(km,0), -\frac{m_i}{m} \]  

\[ G_{km, km+n+1} \]  

(46)

Assuming higher values for \( P_i \), an approximated expression for \( F_{X_n}(x) \) can be derived as

\[ F_{X_n}(x) \approx 1 - \frac{\Gamma(m_i + m_s)\Gamma(am_i / m + n)/m_i}{(\lambda c_m)^{am_i / m} \Gamma(m_i)\Gamma(m_s)\Gamma(n)} \left( \frac{m_i \Omega_i}{m_i P_i} \right)^{m_i} x^{m_i} \]  

(47)

3.4.2 Wireless Applications

3.4.2.1 Probability of Connectivity

In many wireless systems, it is important to determine the coverage area of a transmission. The probability of coverage or probability of connectivity for the n-th nearest neighbor can be defined as the complementary cumulative distribution function (CCDF) of received SNR as

\[ \Pr \{ X_n > \gamma_{th} \} = 1 - F_{X_n}(\gamma_{th}) \]  

(48)

which is the probability that the SNR at the n-th nearest neighbor is greater than the target SNR \( \gamma_{th} \).
For all the numerical results, we consider a 2-dimensional case and a sector of 90 degrees angle, by setting $m = 2$ and $\phi = \pi/4$ in (33) and $c_{\phi,2} = \phi$ in (41). Figure 25 plots the CCDF of received SNR for different values of threshold $\gamma_{th}$ and neighbor index $n$ assuming $\bar{P}_e = 70$ dB, $m_s = 2$, $\alpha = 3$, and $\lambda = 0.001$. It is shown that the connectivity performance improves with the decrease of $\gamma_{th}$ and $n$. Moreover, it is interesting to note that for higher values of $n$, the increase of the threshold $\gamma_{th}$ has a higher impact on the connectivity probability, since the curve for the CCDF becomes steeper as $n$ increases. In order to determine the coverage area of the transmitter, the average distance to the $n$-th neighbor, can be evaluated as

$$E[d_n] = \frac{\Gamma\left(\frac{n+1}{2}\right)}{\Gamma(n)\sqrt{\lambda\phi}}.$$ (49)

This distance determines how far a node can transmit given a minimum required SNR at the receiver, i.e., the length of the longest possible hop for a given transmit power. For example, for PPP density $\lambda = 0.001$ the average distance to the neighbors $n = \{1, 10, 20\}$ obtained from (49) with $m = 2$ and $\phi = \pi/4$ are $E[d_n] = \{32, 111, 158\}$ m. We note that assuming $N = -105$ dBm and $K = -35$ dB, $\bar{P}_e = 70$ dB corresponds to $P_t = 70 + 35 - 105 = 0$ dBm.

Figure 26 plots the probability of connectivity versus the neighbor index $n$ for PPP density $\lambda = \{1, 5, 10\} \times 10^{-3}$, assuming $\bar{P}_e = 65$ dB, $m_s = 2$, path loss exponents $\alpha = 3$ ($\ell = 3$; $k = 1$) and $\alpha = 3.5$ ($\ell = 7$, $k = 2$) in (41), and $\gamma_{th} = 5$ dB. The figure shows the impact of path loss exponent and PPP density $\lambda$ on the probability of connectivity to the $n$-th closest neighbor.

![Figure 25: CCDF of received SNR versus SNR threshold $\gamma_{th}$](image)
In Figure 27 we plot the probability of connectivity versus neighbor index \( n \) for an interference-limited system, that is, for a system where connectivity is based on SIR criteria. We consider the case of \( \alpha = 3.5, \lambda = 0.01, m_s = m_i = 2, \bar{P}_t = 65 \) dB and two values of INR \( \Omega_t = \{5, 15\} \) dB. The plot shows the effect of interference on the probability that the \( n \)th nearest neighbor will successfully detect the transmission. This plot can be used to determine the interference that can be tolerated at the receiver side and estimate the minimum distance that other transmissions could be allowed in order to reduce interference to a potential receiver.
3.4.2.2 Interference Avoidance
The neighbors which can detect the beacon signal abstain from transmission. The n-th nearest neighbor abstains from transmission, with a probability that depends on the received SNR of the packet that notifies the surrounding nodes for a pending reception. For example, we can define the probability that the n-th closest neighbor abstains from transmission to be equal to the probability of the received SNR to be higher than a threshold, that is,
\[ \rho_n = 1 - F_{X_n}(\gamma_{th}). \]  
(50)

This is equivalent to the scenario that neighbor nodes adjust their probability to transmit according to the CCDF of the received SNR. Figure 28 plots the probability that the nth closest neighbor abstains from transmission for threshold values \( \gamma_{th} = \{0, 5, 10\} \) dB and PPP density \( \lambda = 0.005 \), assuming \( P_c = 80 \) dB, \( m_s = 2 \), and path loss exponent \( \alpha = 3.5 (l = 7, k = 2) \) in (41). The asymptotic result of (42) is also plotted. The results show that decreasing the threshold value increases the area around the receiver where a neighbor node is allowed to transmit, as depicted by the top curve of Figure 28, while the asymptotic results are quite useful for lower values of \( \gamma_{th} \) and/or \( n \).

3.5 Summary
An important objective in this WP was to investigate the negative consequences of several channel and network parameters to the system model. In this context, the influence of the spatial correlation effects on the QoS of a V2V communication scenario has been analytically evaluated. Spatial correlation may exist in real world situations, due to the space limitations constraints that exist in vehicles. More specifically, based on the double scattering distribution, the impact of correlation to the performance of a SD system has been evaluated. In addition, in an interference limited scenario, which was also studied, it was shown that interfering effects degrade seriously the system performance, diversity reception can improve this poor situation, provided that channel estimates are close to the exact ones. Moreover, the influence of outdated CSI on the performance of a transmit antenna selection (TAS) scenario has been analytically investigated. Also in this case, it was analytically proved that the performance of diversity systems under the assumption of outdated CSI is considerably reduced, in terms of the diversity gain. Finally, in a VANET communication scenario.
scenario the connectivity probability was also analytically investigated, in interference free and in interference limited scenarios.
4 Low Complexity Algorithms and Architecture Designs for T2T Communications

An important objective of this WP was to propose low complexity algorithms and architecture designs for T2T communication taken into consideration the peculiar parameters of the T2T and T2I wireless channel. Based on the outcomes of the research studies presented previously as well as the requirements for reduced complexity solutions, which was also one of the main objectives of this project, novel diversity reception, relay selection, and multiple input multiple output (MIMO) techniques have been proposed.

4.1 A New Reconfigurable Antenna Scheme and its Application to Vehicle-to-Vehicle Communications

A novel reconfigurable antenna scheme is proposed that is based on a new state selection approach. The new scheme employs two RF chains that support reconfigurable antenna arrays with two states each. This approach allows us to achieve an important diversity gain, satisfying also the strict constraints that exist in the vehicle dimensions as well as cost. In order to further reduce the complexity of the proposed scheme, in each reception period, the antenna reconfigurability is performed in only one of the two RF chains. For this new approach, a novel analytical framework is developed for evaluating its statistical characteristics. The new framework is general enough to take into account the correlation between the antenna patterns that is present in real-world situations. Moreover, as an application example this scheme has been employed in a V2V communication scenario that is based on IEEE 802.11p standard and its performance is analyzed under different channel conditions.

4.1.1 System and Channel Model

We consider a communication system with one transmitter (with one RF chain) and one receiver with two RF chains. In each receiving antenna, reconfigurable antenna arrays exist with two modes (or states) of operation based on the radiation pattern characteristics, i.e., shape, direction, polarization. In the proposed scheme, during a given period of reception, the system is continuously using a specific radiation pattern from one of the two receivers’ RF chain, while from the other one, it selects the state that provides the best performance in terms of the received SNR. Such an approach, where one of the two RF chains is always based on a specific pattern, without any kind of examinations, provides clear practical benefits in terms of synchronization, channel estimation, while also offers reduced number of switching operations. It should be noted that switching among branches (or states) not only consumes power, but also reduces the data throughput and leads to inaccurate phase estimates [26]. The reduced amount of switching operations that the proposed scheme offers, further decreases in cases where time correlated wireless medium exists. In particular, under time correlated fading conditions, assuming that a tagged antenna pattern is selected in time instance \( t \), based on the maximum SNR criterion, it is very likely that in the next instance, \( t+\tau \), the same pattern selection will be also made. However, in our theoretical analysis, in order to provide convenient closed-form expression time-correlated conditions have been omitted.

Assuming that \( \gamma_i \), with \( i \in \{1, 2, 3, 4\} \), denotes the instantaneous received SNR based on the \( i \)-th state, the mode of operation of the proposed scheme is depicted in Figure 29. In this figure, we have assumed that \( \gamma_{\text{max}} = \max\{\gamma_i, \gamma_{i+1}\} \), \( j=3 \) if \( i \in \{1,2\} \), else \( j=3 \). Thus, assuming independence between \( \gamma_i \) and \( \gamma_{\text{max}} \), it is not difficult to recognize that the total instantaneous output SNR is given by

\[
\gamma_{\text{out}} \triangleq \gamma_i + \gamma_{\text{max}}. \tag{51}
\]
In that case, the moment generating function (MGF) of the output SNR for the new scheme is given by

\[ M_{\gamma_{out}} = M_{\gamma_i}(s)M_{\gamma_{max}}(s). \]  \hspace{1cm} (52)

The previous expression may apply to any fading scenario. Here, we will focus on the Rayleigh multipath fading model, which typically agrees very well with experimental data for mobile systems where NLoS path exists between the transmitter’s and receiver’s antennas. Assuming identically distributed fading conditions, the instantaneous received SNR of the i-th state follows the exponential distribution with PDF given by

\[ f_{\gamma_i}(\gamma) = \frac{1}{\overline{\gamma}} \exp\left(-\frac{\gamma}{\overline{\gamma}}\right). \]  \hspace{1cm} (53)

where \( \overline{\gamma} \) denotes the average received SNR. The corresponding expressions for the CDF and the MGF are given as

\[ F_{\gamma_i}(\gamma) = 1 - \exp\left(-\frac{\gamma}{\overline{\gamma}}\right) \]  \hspace{1cm} (54)

\[ M_{\gamma_i}(s) = \frac{1}{1 + s\overline{\gamma}}. \]  \hspace{1cm} (55)

In the following section, we will provide analytical expressions for important statistical metrics of the output SNR for the system under consideration.

4.1.2 Statistical Characteristics

4.1.2.1 Generic Scenario

In the following, we consider the practical important scenario where correlation exists between the antenna patterns that are provided by the reconfigurable antenna array and are available to the RF chain. Correlation between the antenna patterns is a reasonable assumption since in many cases due to the proximity of the antennas as well as the surrounding environment it is very likely that the
antenna radiation patterns will be correlated. As a consequence, the instantaneous received SNRs $\gamma_j$ and $\gamma_{j+1}$, based on $j$, $j+1$ patterns, will be also correlated\(^1\), with bivariate PDF given by

$$f_{\gamma_j,\gamma_{j+1}}(x_1, x_2) = \frac{1}{\bar{\gamma}^2(1-\rho)} \exp \left[ -\frac{x_1 + x_2}{(1-\rho)\bar{\gamma}} \right] I_0 \left( \frac{2\sqrt{\rho x_1 x_2}}{(1-\rho)\bar{\gamma}} \right)$$

(56)

where $\rho$ denotes the correlation coefficient between $\gamma_j$ and $\gamma_{j+1}$. As it is proved in [27], the PDF of $\gamma_{\text{out}}$ is given by

$$f_{\gamma_{\text{out}}}(\gamma) = 2 \frac{\gamma}{\bar{\gamma}} \exp \left( -\frac{\gamma}{\bar{\gamma}} \right) \left[ g(\gamma, 0) \sqrt{\frac{2\rho \gamma}{\bar{\gamma}(1-\rho)}} - \frac{\bar{\gamma}}{1-\rho} + \frac{\sqrt{\rho \gamma}}{1-\rho} g(\gamma, 1) \right]$$

(57)

where

$$g(x, i) = \exp \left( -\frac{1+\rho}{\bar{\gamma}(1-\rho)} x \right) I_i \left( \frac{2\sqrt{\rho x}}{\bar{\gamma}(1-\rho)} \right).$$

Moreover, the corresponding CDF expression is given by

$$F_{\gamma_{\text{out}}}(\gamma) = 2 \left( F_{\gamma_{\text{in}}}(\gamma) - \frac{1}{2} \left[ 1 + \exp \left( -\frac{\kappa \gamma}{2} \right) I_0 \left( \frac{2\sqrt{\rho \gamma}}{\bar{\gamma}(1-\rho)} \right) \right] \right)$$

(58)

$$- \exp \left( -\frac{\gamma}{\bar{\gamma}} \right) Q \left[ \sqrt{\frac{2\rho \gamma}{\bar{\gamma}(1-\rho)}}, \sqrt{\frac{2\gamma}{\bar{\gamma}(1-\rho)}} \right] - \bar{\gamma} f_{\gamma_{\text{out}}}(\gamma)$$

where

$$\kappa = \frac{2}{\bar{\gamma}} + \frac{2(1+\rho)}{\bar{\gamma}(1-\rho)}.$$

4.1.2.2 Independent Channel States

Assuming independence between $\gamma_j$ and $\gamma_{j+1}$, the PDF of $\gamma_{\text{max}}$ is given by

$$f_{\gamma_{\text{max}}}(\gamma) = \sum_{i=1}^{\infty} \left( \begin{array}{c} 2 \\ i \end{array} \right) (-1)^{i+1} \frac{i}{\bar{\gamma}} \exp \left( -\frac{i\gamma}{\bar{\gamma}} \right).$$

(59)

Based on it, the corresponding MGF expression is given by

$$M_{\gamma_{\text{max}}}(s) = \sum_{i=1}^{\infty} \left( \begin{array}{c} 2 \\ i \end{array} \right) (-1)^{i+1} \frac{i}{i + \bar{\gamma} s}.$$  

(60)

\(^1\) It is assumed that the reconfigurable antennas between the two RF chains are separated by a distance large enough that guarantees fully decorrelation and thus correlation exists only between states $j$ and $j + 1$.  

WP3 – T2X Communication Techniques © ROADART consortium
Substituting (55) and (60) in (52) and applying the inverse Laplace transform, yields the following simplified expression for the PDF of $\gamma_{\text{out}}$

$$f_{\text{out}}(\gamma) = \frac{2}{\bar{\gamma}} \exp\left(-\frac{\gamma}{\bar{\gamma}}\right) \left[1 + \frac{\gamma}{\bar{\gamma}} - 1\right].$$  \hfill (61)

Based on (61), the corresponding expression for the CDF of $\gamma_{\text{out}}$ can directly be evaluated as

$$F_{\text{out}}(\gamma) = 1 - \exp\left(-\frac{2\gamma}{\bar{\gamma}}\right) - \frac{2\gamma}{\bar{\gamma}} \exp\left(-\frac{\gamma}{\bar{\gamma}}\right).$$  \hfill (62)

4.1.2.3 Asymptotic Analysis

The exact expression for $F_{\gamma_{\text{out}}}(\gamma)$ given in (58) cannot be used to obtain a clear physical insight of the system’s performance. Therefore, here, a simplified asymptotic expression for $F_{\gamma_{\text{out}}}(\gamma)$ will be derived and used to study the diversity and coding gains. For the general case where correlation exists between $\gamma_j$ and $\gamma_{j+1}$, assuming higher values of the average input SNR and using [28] eq. (2), a simplified expression for the PDF of $\gamma_{\text{max}}$ can be extracted as

$$f_{\text{max}}(\gamma) \approx \frac{2}{\bar{\gamma}} \exp\left(-\frac{\gamma}{\bar{\gamma}}\right) \left[1 - \exp\left(-\frac{\gamma}{\bar{\gamma}(1-\rho)}\right)\right].$$  \hfill (63)

Moreover, since for higher values of the average SNR $f_{\gamma}(\gamma) \approx \frac{1}{\bar{\gamma}}$ and based on (63) as well as the analysis presented in [27], the following simple expression for the PDF of $\gamma_{\text{out}}$ is derived

$$f_{\text{out}}(\gamma) \approx \frac{\gamma^2}{1-\rho} \frac{1}{\bar{\gamma}}.$$  \hfill (64)

The corresponding CDF expression is given by

$$F_{\text{out}}(\gamma) \approx \frac{\gamma^3}{3(1-\rho)} \frac{1}{\bar{\gamma}}.$$  \hfill (65)

4.1.2.4 Performance Analysis

Using the previously derived results, important performance metrics of the scheme under consideration were analytically studied. More specifically, the performance was studied using the criteria of OP and the BEP.

**Outage Probability:** The OP is defined as the probability that the instantaneous received SNR falls below a predetermined threshold $\gamma_T$. Therefore, in our case, the OP can be directly evaluated as $P_{\text{out}} = F_{\gamma_{\text{out}}}(\gamma_T)$. It is obvious that (65) is of the form $(G_d\bar{\gamma})^{-0.2}$, where $G_d$ represents the diversity gain and $G_c = \left[\frac{\gamma^4}{3(1-\rho)}\right]^{-1/3}$ is the coding gain. Thus, from the above it is clear that the correlation between the branches does not affect the diversity gain of the system under consideration, which is always equal to 3, but only the coding gain. Nevertheless, as the correlation between the antennas increases, the coding gain decreases and thus the OP performance of the system will also decrease.

**Average Bit Error Probability:** Using the previously derived MGF expressions for $\gamma_{\text{out}}$ and following the MGF-based approach, the BEP can be readily evaluated for a variety of modulation
schemes [3]. More specifically, the BEP can be calculated: i) directly for non-coherent DBPSK, that is 

\[ P_b^{DBPSK} = \frac{1}{\pi \log_2(M)} \int_0^\pi \frac{\log_2(M) \sin^2(\frac{\pi}{M})}{\sin^2(\phi)} d\phi. \]  

(66)

4.1.3 Numerical Results

In this section, based on the previous presented theoretical analysis, several numerically evaluated performance results are provided. These results include an OP investigation, using (58), (62), and (65), and a BEP analysis which can be evaluated with the aid of (52), (55), and (60). Moreover, as an application example, the proposed scheme is also employed in a V2V communication scenario using a IEEE 802.11p-based simulator, which is described in Section 2.

In Figure 30, the OP is plotted as a function of the average received SNR for different values of the correlation coefficient \( \rho \). For comparison purposes the corresponding performances of single input single output (SISO) and MRC diversity (with 2 RF chains) receivers have been also included. It is shown a clear diversity gain of the proposed scheme, as compared to the single receiver. Moreover, the proposed scheme also outperforms MRC even in cases where relative strong correlation exists between the branches. The asymptotic curves, which are also included in the same figure, approximate quite well the exact ones even for moderate values of the average SNR, i.e., 20dB, while the approximation improves for lower values of \( \rho \).

In Figure 31, the BEP of DBPSK is plotted as a function of the correlation coefficient \( \rho \), for different values of the average received SNR. It is shown that the performance decreases with the increase of \( \rho \). Moreover, the performance decreases with the decrease of \( \gamma \), as expected. However, it is interesting to note that the performance deterioration due to the increase of \( \rho \) is more important for higher values of the correlation coefficient. Finally, also in the same figure, assuming \( \rho = 0 \), the diversity order, defined as 

\[ \frac{\log_2(P_b^{DBPSK})}{\log_2(SNR)} \] 

is plotted as a function of the SNR. As it is depicted in this figure, as the SNR tends to infinity, the diversity order approaches 3, verifying the analysis provided in the previous section. Computer simulation performance results are also included in Figure 30 and Figure 31, verifying in all cases the validity of the proposed theoretical approach.

![Figure 30: OP vs the average received SNR for different values of \( \rho \).](image-url)
4.1.3.1 IEEE 802.11p Application Example

The physical layer simulator that is described in Section 2 is used for the evaluation of the proposed diversity scheme. Next, a few details related with the simulation parameters are provided. The binary data are coded and modulated through OFDM. Although the total number of subcarriers is $N = 64$, only $N_{\text{eff}} = 48$ subcarriers are used for data transmission and $N_{\text{pilot}} = 4$ subcarriers are used for pilots transmission. 10 short and 2 long preambles are added as header to the total OFDM symbols for a proper frame construction. The receiver consists of 2 RF chains with 2 states of operation at each RF chain. In addition to the proposed scheme, a SISO and a MRC receiver are also implemented for comparison purposes. In the numerical analysis, the wide sense stationary uncorrelated scattering (WSSUS) assumption was assumed and flat fading Rayleigh channel was used. In order to evaluate the performance of each diversity receiver, a BEP vs SNR diagram is derived, by means of Monte Carlo simulation. To compromise between complexity and fidelity, the noise level is normalized analogous to the desired SNR value and the transmitted signal’s power is normalized to unity. In addition each scheme is simulated with perfect CSI knowledge at the receiver. For each SNR value 104 frames are transmitted. Table 3 summarizes the parameters used for the simulator implementation.

Figure 32 shows the average BEP vs SNR for each diversity scheme with antennas number specified by Table 3. The proposed scheme outperforms SISO, as expected. It also provides better performance from MRC, since it exploits the additional diversity gain offered by the multiple radiation patterns. It is noted that since complexity is translated to hardware processing time, the proposed scheme is a suitable receiver for fast time varying wireless medium, such as those that are typical on intervehicular communication environments, where processing time is not a negligible design parameter.

Further evaluation of the proposed scheme has been performed. Based on [29], a tap delay channel model on the 5 GHz band was developed. A major difference between the model proposed in [29] and the traditional Rayleigh one, is that the non-stationary characteristics of the channel were taken into consideration. Based on the empirical results derived from the measurement campaigns in [29],
the Weibull distributed channel was proposed as a good candidate for modelling such environments. In this model, each channel tap owns Weibull statistical properties, where the shape and scale factors are defined based on the measurement results. Since for more realistic simulation results

Table 2: Simulator Parameters.

<table>
<thead>
<tr>
<th>Parameters</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Center Frequency</td>
<td>5.9GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>10MHz</td>
</tr>
<tr>
<td>Total Subcarriers</td>
<td>64</td>
</tr>
<tr>
<td>Effective Subcarriers</td>
<td>52</td>
</tr>
<tr>
<td>Guard Subcarriers</td>
<td>12</td>
</tr>
<tr>
<td>OFDM symbol duration</td>
<td>8μs</td>
</tr>
<tr>
<td>OFDM Symbols per Frame</td>
<td>100</td>
</tr>
<tr>
<td>Number of Frames</td>
<td>$10^4$</td>
</tr>
<tr>
<td>Coding Rate</td>
<td>$\frac{1}{2}$</td>
</tr>
<tr>
<td>Modulation</td>
<td>BPSK</td>
</tr>
<tr>
<td>Fading Model</td>
<td>Rayleigh</td>
</tr>
</tbody>
</table>

Table 3: Diversity Receivers Parameters.

<table>
<thead>
<tr>
<th>Receiver</th>
<th>RF chains</th>
<th>Total States</th>
<th>Used Antennas</th>
</tr>
</thead>
<tbody>
<tr>
<td>MRC</td>
<td>2</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>Proposed scheme</td>
<td>2</td>
<td>4</td>
<td>2</td>
</tr>
</tbody>
</table>

Figure 32: Average BEP vs average SNR for the proposed scheme, MRC, SD, and SISO.

there is a need to implement time variation, time correlation was inserted between the samples of each tap. Moreover, the probability of correlation existence between two consecutive time instances obeys a simple Markov process, where transition probabilities are empirically derived. This time variation on the channel model implementation, adds the non-stationary characteristics, which are evident on V2V communication scenarios. In addition, tap amplitudes are also correlated, with correlation properties being extracted by the measurements. The measurement campaign took place on four different environments, typically characterized as “Urban with Antenna Outside the
Vehicle”, “Urban-Antenna Inside Car”, “Open-Area High Traffic Density” and “Open-Area Low Traffic Density”. From the four different environments that were implemented, the one that is characterized as “Urban with antenna outside the vehicle” was chosen for the purposes of the simulation. Moreover, no correlation was assumed between the receiving antennas.

Figure 33 shows the average BEP vs SNR for each diversity scheme for SIMO non stationary V2V channel. The performance degradation that is obvious, is caused by the channel’s non stationarity characteristics. However, it is noted that the BEP-SNR curves follow the expected trends. It is no surprising that the proposed scheme performs better than MRC, though the distance between them is considerably shorter in relation to the Rayleigh channel corresponding figure.

Figure 33: Average BEP vs average SNR for the proposed scheme, MRC, SD, and SISO.

4.2 A Cooperative Relay Selection Scheme in V2V Communications under Interference and Outdated CSI

Cooperative communications have been considered as a promising solution for extending coverage and enhancing reliability in contemporary communication networks. The performance of these systems is directly related with the relaying protocol and the selection mechanism that has been considered. As far as the employed protocol is concerned, two major approaches exist, namely decode-and-forward (DF) and amplify-and-forward (AF). With a small cost on the complexity, DF offers better performance as compared to AF, due to the decoding process that is employed in the first phase and guarantees a minimal good source-to-relay link. Moreover, regarding the relay selection mechanism, best relay selection (BRS) has been proposed as an efficient technique for improving performance [30]. In BRS, the best relay, from a set of N available ones, is selected and used for cooperation between the source and the destination. As a result, BRS requires increased processing and feedback load overhead, since for all relays full CSI is necessary, in each packet transmission, while also more switches are expected. This is especially important in ad-hoc networks because, the required fast switching rate impairs system stability and/or leads to inaccurate channel estimations, [31]. VANETs represent an integral part of the ITSs that has gained an important interest by the scientific community and the industry over the past several years. These systems enable the V2V communications and can be applied in a variety of communication scenarios ranging from road-safety and energy-saving improvements to comfort and infotainment. However, in contrast to the traditional cellular mobile radio link, the V2V propagation channel is much more dynamic, since it consists of two non-stationary transceivers, closely located to the ground level. In this context, many works have investigated V2V cooperative scenarios, e.g., [32] [33]. A common observation in previous works is the assumption of noise limited environment.
However, in many practical situations, e.g., in the presence of hidden terminals, the performance of these systems can be significantly affected by CCI. Moreover, previous works in this area, assume that perfect CSI is available at the system for relay selection. However, due to channel estimation errors and feedback delays, such an ideal assumption cannot be established in practice and has only theoretical importance [12]. In WP3 context, an analytical investigation was performed for the influence that interfering effects as well as outdated CSI have to the performance of a V2V cooperative system. In particular, the analysis was based on a well-established channel model for mobile-to-mobile communications such as double-Nakagami (DN) [34]. For this fading model, the PDF was presented for the first time and used to model the correlation between the exact and the outdated versions of the received SNR. Moreover, a new relay selection scheme was adopted, which reduces feedback load processing, while achieves almost similar performance to BRS. Next, the system and channel model under investigation is presented and based on it a stochastic analysis is performed for the received SIR statistics, which is used to analyze the OP and the average symbol error probability (SEP) performances. Finally, capitalizing on this analysis several numerical performance results are presented.

4.2.1 System and Channel Model

We consider a V2V cooperative communication scenario with one source (S), N relays $R_n = \{1, 2, \ldots, N\}$ and one destination (D). We assume that all nodes are equipped with a single antenna and the transmission is realized in two phases. All hops in the scheme under consideration, including the interfering links, experience DN fading with channel gains, $h_{X}$, having PDF given by [34] where $m_{x,k}$ denote the shaping parameter related with the severity of the fading and $\Omega_{x,k} = \{20, 20, 02, 02\}$, when referring to S-$R_n$, $R_n$-D, interfering source to $R_n$, interfering source to D, respectively and $k, \ell \in \{1, 2, 3, 4\}$ being related with the DN double-bounce interaction [2]. The DN distribution is adopted since it provides a realistic description of the V2V channel in situations where both transmitter and receiver are moving. In addition, $R_n$ as well as D are also subject to interfering signals under the assumption that (in general) the level of interference is such that the effect of thermal noise on system’s performance can be ignored, i.e., interference limited scenario. For simplification purposes, and without losing the generality, the analytical results are derived for the case of one interferer. At the first phase, the source transmits a signal only to the set of relays. As a result, the received SIR at the nth relay is given by

$$\gamma^{\text{out}}_{n} = \frac{\gamma_{s_n}}{\gamma_{I_n}}$$

where $\gamma_{s_n} = \frac{E_s |h_{s_n}|^2}{\sigma_{s_n}^2}$ with $E_s$ denoting the transmitter signal energy, $\sigma_{s_n}^2$ the variance of the AWGN and $\gamma_{I_n} = \frac{E_{I_n} |h_{I_n}|^2}{\sigma_{I_n}^2}$ denoting the instantaneous INR received at $R_n$. Moreover, let $C$ denoting the decoding set of the active relays that have correctly decoded the source message [35]. This set will form the basis for the relay that will be selected in the next phase.

4.2.1.1 Relay Selection Scheme

In the following, a new relay-selection scheme is employed that aims at reducing complexity and feedback load processing. The mode of operation of the new scheme is as follows. The relay that
was selected in the previous round of communications, sends request to send packet to the final destination. From this packet, the destination estimates the received instantaneous SNR, which is compared with a predefined threshold, $\gamma_{th}$, and if it exceeds it, then this relay is selected and no further processing is required. Otherwise, a flag packet is send to all relays that successfully received the original message (including the tagged relay) asking for the initiation of a best relay procedure similar to the one given in [36]. From the mathematical point of view and under the assumption of i.i.d. fading, the CDF of the received instantaneous SNR for the link between $R_n$ and $D$, $\tilde{\gamma}_{sel}$, given $C$, can be expressed as [37]

$$F_{\tilde{\gamma}_{sel}|C}(x) = \begin{cases} F_{\gamma_{th}}(x) - F_{\gamma_{th}}(\gamma_{th}) + F_{\gamma_{th}}(\gamma_{th})F_{\gamma_{th}}(x)^{C-1}, & x \geq \gamma_{th} \\ F_{\gamma_{th}}(x)^C, & x < \gamma_{th} \end{cases} (69)$$

where $\gamma_{sel} = \mathbb{E}_h \left[ h_{r_n} \right]^2 / \sigma^2_{h_{r_n}}$, with $f_{\gamma_{sel}}(x)$ obtained by employing a straight forward change of variables in (67) and CDF given by

$$F_{\gamma}(x) = \frac{1}{\Gamma(m_{p,1})\Gamma(m_{p,2})} G^{21} \left( m_{1,2} m_{p,2} x \right) G^{13} \left( m_{x,1} m_{x,2} x \right) \frac{1}{\Omega_{x,1} \Omega_{x,2}}. (70)$$

The corresponding expression for the PDF of $\tilde{\gamma}_{sel}|C$ is

$$f_{\tilde{\gamma}_{sel}|C}(x) = \begin{cases} f_{\gamma_{sel}}(x) + (C-1)F_{\gamma_{th}}(\gamma_{th})f_{\gamma_{sel}}(x)F_{\gamma_{th}}(x)^{C-2}, & x \geq \gamma_{th} \\ Cf_{\gamma_{sel}}(x)F_{\gamma_{th}}(x)^{C-1}, & x < \gamma_{th} \end{cases} (71)$$

After the selection is made, the relay forwards the initial message to the final destination and the received SIR at the destination, given set $C$, can be expressed as

$$\gamma_{out} = \frac{\tilde{\gamma}_{sel}}{\gamma_{sel}} (72)$$

where $\gamma_{dt} = \mathbb{E}_h \left[ h_{d_t} \right]^2 / \sigma^2_{h_{d_t}}$ denotes the received instantaneous INR at destination. The behavior of the proposed scheme depends on the selected switching threshold $\gamma_{th}$, which is related with the instantaneous SNR. More specifically, with an increase on $\gamma_{th}$, its performance improves, approaching that of BRS, with the cost of a higher feedback load processing.

4.2.1.2 CSI Model

In a V2V communication scenario, it is very likely that the fading behavior will change rapidly. In this work, the more practical scenario where the CSI of the $R_n$ - $D$ links is assumed to be outdated, due to the delay existing between the user selection and data transmission phases as well as the fast time varying nature of the medium. Thus, the CSI model employed in [38] will be also adopted. More specifically, at the data transmission instance, the level of imperfection between the actual SNR, $\gamma_{sel}$, and $\tilde{\gamma}_{sel}$, which is available at the selection instance, can be measured based on the correlation coefficient $\rho_i$. For example, assuming the classic Jake’s model $\rho_i = J_0 \left( 2\pi f_d T_d \right)$, with $f_d$ being the maximum Doppler frequency, $T_d$ the time delay due to CSI feedback. Thus, the PDF of the actual received SNR of the selected user at the data transmission instance can be expressed as [39]
\[ f_{\gamma_{sel}}(y) = \int_0^\infty f_{\gamma_{tal},\gamma_{tal}}(y,x) \frac{f_{\gamma_{tal}}(x)}{f_{\gamma_{tal}}(x)} \, dx. \]  

(73)

For evaluating (73), the bivariate DN distribution is required, which has not been presented in the past. Here, the following expression for the joint PDF between \( \gamma_{sel} \) and \( \gamma_{sel} \) has been obtained as

\[
f_{\gamma_{tal},\gamma_{tal}}(x, y) = \sum_{l=0}^{\infty} \sum_{q=0}^{\infty} \frac{4^{l+q}}{l!q!} \left( \prod_{i=1}^{4} \bar{f}_{\gamma_{i,i}} \right) \frac{1}{2} (x y)^{m_{\gamma_{1,1}} + m_{\gamma_{2,2}} + q + l} \frac{1}{2} \left( (x y)^{m_{\gamma_{2,2}} + q} (1 - \rho_{1})^{m_{\gamma_{1,1}} + q + l} (1 - \rho_{2})^{m_{\gamma_{1,1}} + q + l} \right) \times K_{m_{\gamma_{1,1}} - m_{\gamma_{2,2}} + q + l} \left[ \frac{2x^{l/2}}{\sqrt{\bar{f}_{\gamma_{2,2}} \bar{f}_{\gamma_{1,1}} \rho_{12}}} \right] K_{m_{\gamma_{2,2}} - m_{\gamma_{1,1}} + q + l} \left[ \frac{2y^{l/2}}{\sqrt{\bar{f}_{\gamma_{2,2}} \bar{f}_{\gamma_{1,1}} \rho_{12}}} \right]
\]

where \( \rho_{12} = (1 - \rho_{1})(1 - \rho_{2}) \) and \( 0 \leq \rho_{1}, \rho_{2} < 1 \) are the power correlation coefficients of the underlying fading processes of the first and second bounces, respectively. Moreover, in (8), \( \bar{f}_{\gamma_{n,i}} = \frac{a_{\gamma_{n,i}}}{m_{\gamma_{1}}} \) \( \forall i \in \{1,2\} \), \( \bar{f}_{\gamma_{n,i}} = \frac{a_{\gamma_{n,i}}}{m_{\gamma_{2}}} \) \( \forall i \in \{3,4\} \). It is noted that the proof of (74) can be found in [40].

4.2.2 SIR Statistics

In this section, the CDF of the system’s output SIR is derived, while a high SNR analysis is also provided.

4.2.2.1 1st link Statistics

The first phase is characterized by the probability that the nth relay decodes the signal incorrectly, \( P_{off} \), which represents the SEP. In particular, \( P_{off} \) can be evaluated as

\[
P_{off} = a \int_0^\infty \text{erfc} \left( \sqrt{b \gamma} \right) f_{\gamma_{off}}(\gamma) \, d\gamma
\]

(75)

where (a, b) depend on the modulation scheme employed, e.g., for binary phase shift keying (BPSK), \( a = \frac{1}{2}, b = 1 \). Next, in order to simplify the analysis, it is assumed that \( |m_{\gamma_{2,2}} - m_{\gamma_{1,1}}| = 1/2 \) as well as \( |m_{\gamma_{2,2}} - m_{\gamma_{1,1}}| = 1/2 \), i.e., quite similar fading conditions exist for the scattering environments around both S and Rn. Thus, substituting the PDFs of \( \gamma_{sn} \) and \( \gamma_{ln} \) in \( f_{\gamma_{out_{sn}}} (\gamma) = \int_0^\infty \gamma f_{\gamma_{in}} (xy) f_{\gamma_{in}} (x) \, dx \) using [8] eq. (07.34.03.0606.01) as well as [5] eq. (3.326), the PDF of \( \gamma_{out_{2n}} \) is given by

\[
f_{\gamma_{out}} (\gamma) = \frac{\Gamma(2m_{\gamma_{1,1}} + 2m_{\gamma_{1,1}})}{\Gamma(m_{\gamma_{1,1}}) \Gamma(m_{\gamma_{1,1}})} \frac{2^{1-2m_{\gamma_{1,1}}+2m_{\gamma_{1,1}}}}{\pi} \prod_{i=1}^{2} \left( 1 - \frac{1}{\gamma_{i,i}} \right)^{m_{\gamma_{1,1}}} \left( 1 - \frac{1}{\gamma_{i,i}} \right)^{m_{\gamma_{1,1}}}
\]

\[
 \times \gamma^{m_{\gamma_{1,1}}-1} \left( \frac{\gamma}{\sqrt{\gamma_{11}}} + \frac{1}{\gamma_{11}\sqrt{\gamma_{11}}} \right)^{-2m_{\gamma_{1,1}}-2m_{\gamma_{1,1}}}
\]

(76)
Moreover, substituting (76) in (55), using [25] eq. (10), [8] eq. (06.27.26.0006.01) and [25] eq. (21), the following closed-form expression for $P_{off}$ is obtained

$$P_{off} = \frac{a/l \sqrt{\pi}}{\Gamma(m_{t,1})\Gamma(m_{s,2})\Gamma(m_{t,2})} G^{a_{12}} \frac{a_{13}^{1-2m_{t,1}-m_{s,1}}}{\Gamma(m_{t,1})} \frac{1}{\Gamma(m_{s,2})} \left( \frac{2}{\Gamma(m_{t,2})} \right)^{1-2m_{t,1}-m_{s,1}}$$  \hspace{1cm} (77)

### 4.2.2.2 2nd Link Statistics

In the second communication phase, the CDF of the output SIR given $C$, $\gamma_{out|C}$, can be expressed as

$$F_{\gamma_{out}}(\gamma) = \pi^2 \Gamma(2m_{t,1}) \sum_{h,q=0}^{\infty} \left( \sum_{j,h,j,z=0}^{\infty} \prod_{i=1}^{2} \left( \frac{1}{j_i!} \right)^{1-|h+i|/2} \right)$$

$$\times \left( \frac{\rho_{t,1}\rho_{s,2}^2 2^{3-\theta_i-\theta_{s,2}}}{\Gamma(h+m_{r,1})\Gamma(q+m_{r,2})} \right)$$

$$\times A \left( \frac{\gamma}{\tilde{\gamma}_{13,12}^{\theta_{t,1}+\theta_{s,2}}} \right)^{\theta_{t,1}+\theta_{s,2}} \left( \frac{h!}{\gamma_{out|C}^{\theta_{t,1}+\theta_{s,2}}} \right)^{\theta_{t,1}+\theta_{s,2}}$$

where $\tilde{\gamma}_{r,\gamma} = \tilde{\gamma}_{r,\gamma}, \theta_i = h + q + 2m_{r,1} - j_i$ for $i \in \{1,2\}$ and $A$ is given by

$$A = \frac{\Gamma(m_{r,1})\Gamma(m_{r,2})}{\sqrt{\pi} \Gamma(2m_{r,1})} \left( \theta_{t,2} \frac{2\sqrt{\gamma_{th}}}{\sqrt{\gamma_{24,12}}} + \sum_{i=1}^{2} \frac{2^{1-2m_{r,1}}}{\sqrt{\gamma_{24,12}}} \sum_{j=0}^{C-t} \beta_2 \sum_{i=1}^{C-t} \frac{i!}{\gamma_{13,12}^{\theta_{t,1}+\theta_{s,2}}} \right)$$

$$\times \frac{\prod_{j=1}^{2m_{r,1}-1} \frac{1}{\tilde{\gamma}_{13,12}^{\theta_{t,1}+\theta_{s,2}}}}{\beta_1}$$

$$\times \left[ (2-t) \frac{\gamma_{13,12}^{\theta_{t,1}+\theta_{s,2}}}{\gamma_{24,12}^{\theta_{t,1}+\theta_{s,2}}} + (t-1) F_{\gamma_{13,12}}(\gamma_{th}) \right] \left( \theta_{t,2} + \sum_{j=1}^{2m_{r,1}-1} jn_{j,1} \beta_2 \gamma_{th} \right) \Gamma \left( \theta_{t,2} + \sum_{j=1}^{2m_{r,1}-1} jn_{j,1} \beta_2 \gamma_{th} \right).$$

In the above expression $\beta_i = \frac{\ell}{\sqrt{\gamma_{24,12}}} + \ell i \sqrt{\gamma_{13,12}}$.

**Asymptotic Analysis:** The exact expression for $F_{\gamma_{out|C}}(\gamma)$ does not provide a clear physical insight of the system’s performance. In order to provide a simplified expression, the main concern is to derive an asymptotic closed-form expression for $F_{\gamma_{out|C}}(\gamma)$. Therefore, assuming $\rho_1 = \rho_2$, higher values of $\tilde{\gamma}_{r,\gamma}$ and based on the approach presented in [39], a closed-form asymptotic expression for the CDF of $\gamma_{out|C}$ is obtained in as
\[ F_{\text{sel}}(\gamma) = \left[ \left( 2m_{r,1}, 2\sqrt{\gamma_{\text{th}}} / \beta_3 \right) + \sum_{i=1}^{2} (C+1-i) \left( \frac{\sqrt{\pi / m_{r,1}}} {\Gamma(m_{r,1})\Gamma(m_{r,2})} \right)^{C-i} \right]^2 \times \right. \\
\left. \left[ (2-i) \left( \frac{2}{\beta_3} \right)^2 (2m_{r,1}, 2\sqrt{\gamma_{\text{th}}} / \beta_3) + (i-1) F_{\gamma_{\text{th}}} \left( \frac{2}{1-\rho} \right)^2 (2-C)m_{r,1,2} \right] \right] \]

\[ \times \left[ \prod_{i=1}^{2} \left( \frac{1}{\gamma_{l,i}} \right)^{1-2m_{r,2}} \Gamma(m_{r,2}) \right] \frac{\pi(1-\rho)^{1/2} 2m_{r,2}^{-1}} {m_{r,2}^{-1/2} \gamma} \]

where \( \beta_3 = \bar{\nu}_{13}\rho_{12} \).

### 4.2.3 Performance Evaluation

In this section using the previously derived results, important performance metrics of the scheme under consideration were analytically studied. More specifically, the performance was evaluated using the criteria of OP and the SEP.

#### 4.2.3.1 Outage Probability

The OP is defined as the probability that the instantaneous received SIR falls below a predetermined threshold \( \gamma_T \). Therefore, the OP can be evaluated as

\[ P_{\text{out}} = P_{\text{off}}^W + \sum_{k=1}^{N} \left( \left( N + k \right) P_{\text{off}}^{N-k} (1-P_{\text{off}})^k \right) F_{\text{sel}}(\gamma_T). \]  

#### 4.2.3.2 Symbol Error Probability

The SEP can be evaluated as

\[ \bar{P}_{\text{se}} = \bar{P}_{\text{off}}^W + \sum_{k=1}^{N} \left( \left( N + k \right) P_{\text{off}}^{N-k} (1-P_{\text{off}})^k \right) P_{\text{se}} \]  

where

\[ P_{\text{se}} = \sqrt{2\pi \Gamma(2m_{r,1})} \sum_{h,q=0}^{\infty} \sum_{j=0}^{q} \frac{\rho_1^h \rho_2^q \Gamma\left( j + \left\lfloor \frac{1}{2} - q \right\rfloor \right)} {h!q!} \left[ \prod_{i=1}^{2} \left( j_i + \left\lfloor \frac{1}{2} - q \right\rfloor \right) \Gamma\left( m_{r,2} \right)^{-1/2} \right] \]

\[ \times A \frac{\Gamma(2m_{r,1} + \theta_1) \sqrt{h + m_{r,2}} / \theta_1^{2m_{r,1}}} {\Gamma(h + m_{r,2})} \sum_{k=0}^{h+m_{r,2}} \frac{\sqrt{\gamma_{l,2}} \Gamma\left( m_{r,2} \right)^{-1/2}} {k!} \frac{(1-2m_{r,1})_k (\theta_1)_k G^{1/2} \frac{b_{\gamma_{l,2}}}{2\gamma_{l,2}^2} \frac{|-\theta-k|}{2}} {\Gamma(\theta_1 + k) \Gamma(\theta_1 + k + 1) \Gamma(\theta_1 + k + 1/2) \Gamma(\theta_1 + k + 1/2)}. \]

The procedure for (81) is provided in [40].
4.2.4 Numerical Results

In this section, several numerically evaluated performance results complemented by equivalent simulated ones, are provided. These results include the OP investigation, based on (80) and (81). In all cases, the following assumptions have been made for the various parameters of the system and channel models under consideration. For the first link, number of relays \( N = 4 \) (if not otherwise stated), shaping parameter of the desired link \( m_{s,1} = 1.5 \), mean values \( \Omega_{s,i} = 10\text{dB} \), shaping parameter of the interfering links \( m_{I,1} = 1.5 \), average INR \( \bar{\gamma}_{I,i} = 4\text{dB} \). For the second link, shaping parameter of the desired link \( m_{r,1} = 1.5 \), average INR \( \bar{\gamma}_{d,i} = 4\text{dB} \). Moreover, the processing complexity of the proposed scheme is also evaluated based on the approach provided in [41]. More specifically, as a performance indicator for the complexity, the average number of active relays (NAR) is adopted, \( N_{\text{out}} \). It is obvious that as the NAR, which must be examined in the second phase of communications, increases, the processing and feedback load will also increase. In the proposed scheme, NAR can be evaluated using

\[
N_{\text{out}} = P_{o\text{ff}}^N + \sum_{k=1}^{N} \binom{N}{k} P_{o\text{ff}}^{N-k} (1 - P_{o\text{ff}})^k \left[ 1 + (i - 1)F_{\gamma_{r,i}}(\gamma_{\text{th}}) \right].
\]

In Figure 34, assuming \( \Omega_{r,i} = 10\text{dB} \), switching threshold \( \gamma_{\text{th}} = 10\text{dB} \), the OP is plotted as a function of the correlation coefficients \( \rho_1 = \rho_2 = \rho_i \) for different values of the outage threshold \( \gamma_T \). In this figure, it is shown that the OP decreases with the decrease of \( \gamma_T \). It is noted that as \( \rho_i \) increases, i.e., the SNR at the selection instance approaches the one at the reception instance, the performance improves, i.e., OP decreases. This performance improvement is higher for \( \rho_i \to 1 \), while it seems that for lower values of \( \gamma_T \), the performance gain due to the increase of \( \rho_i \) is higher. In Figure 35, assuming \( \gamma_T = 10\text{dB} \), \( \gamma_{\text{th}} = 20\text{dB} \), the OP is plotted as a function of \( \Omega_{r,i} = \bar{\gamma}_{r,i} m_j \), for different values of \( \rho_i \). In this figure it is depicted a gap among performances exists for \( \bar{\gamma}_{r,i} \leq \gamma_{\text{th}} \), which decreases as \( \bar{\gamma}_{r,i} \) increases. This behavior is due to the mode of operation of the new scheme, since for \( \bar{\gamma}_{r,i} \geq \gamma_{\text{th}} \) it is very likely that all second hop links satisfy the switching threshold and thus the number of switches is expected to decrease. Moreover, the asymptotic curves, which are also included in the same figure, approximate quite well the exact ones even for moderate values of the average SNR, i.e., 20dB, while the approximation improves for lower values of \( \rho_i \). In addition, in the same figure, the average NAR is also included for the proposed scheme as well as the corresponding one with BRS. It is shown that for lower values of \( \bar{\gamma}_{r,i} \), the proposed scheme offers an excellent compromise between performance and complexity as compared to BRS.
In Figure 36, assuming $\rho_i = 0.75$, $\gamma_T = 10$dB, the SEP of BPSK is plotted as a function of the number of relays N for different values of $\gamma_{r,i}$ and $\gamma_{th}$. It is depicted in this figure that the performance improves with an increase on $\gamma_{r,i}$, $\gamma_{th}$ and N. More specifically, for lower values of $\gamma_{r,i}$ by setting a larger value for the switching threshold, results to an important improvement on the system performance, which further improves with the increase of N. However, for higher values of $\gamma_{r,i}$ the gain offered due to the increase of $\gamma_{th}$ is smaller. Moreover, it is not necessary to employ more than 2

Figure 35: OP vs the average SNR of the 2nd hop for different values of $\rho_i$. 
D3.1– Recommendation of Antennas and Communication Techniques Qualified for Implementation  H2020 - 636565

relays since for N > 2 the gain achieved is minor. Finally, simulation performance results are also included in all figures, verifying the validity of the proposed theoretical approach.

4.3 MIMO V2V Communications via Multiple Relays: Relay Selection Over Space-Time Correlated Channels

The performance of SISO and MIMO V2V systems in DF multi-relay wireless networks has been investigated, where multiple vehicle relay nodes, are assigned to assist a source in forwarding its information to a destination. A full-duplex (FD) operation mode has been assumed, which facilitates frequency reuse and enables data reception and transmission at the same time in a single frequency band. Nevertheless, it also allows the reception of a certain amount of loop-interference (LI) due to signal leakage from the relays’ transmission to their own receiver. The residual LI can be minimized by adaptively control the power of the system [42]. The performance of wireless relay networks is significantly improved by selection of relays [43], [44] with respect to specific metrics. Hence, a relay selection policy has been implemented, which is directly related to the radio channel [45] and the space-time correlation properties regarding the MIMO system. Note that the degree of correlation is a complicated function of the degree of scattering and the antenna inter-element spacing at the source, relay, and destination [46]. Hence, three-dimensional (3-D) regular-shaped geometry-based stochastic models (RS-GBSMs) for SISO and MIMO V2V relay-based channels have been utilized. This is the first time that relay selection is performed through the use of 3-D channels in inter-vehicular communications.

To improve the resource utilization and reduce the hardware complexity, opportunistic relay selection has been performed, where only one relay is activated [47]. The relay selection is based on the minimization of the transmission power, i.e., the selected FD DF relay requires the minimum sum of powers in the source-relay (S-R) and relay-destination (R-D) links. Aiming at reducing the amount of CSI overhead, a reactive relay selection policy has been examined, instead of a proactive relay selection policy. Since the power of the selected relay is adapted, CSI is required only in the R-D link, whereas in the S-R link, the source always transmits with a fixed power level. The

Figure 36: SEP vs number of relays for different values of $\Omega_{r,i}$. 

WP3 – T2X Communication Techniques © ROADART consortium  page 67
The proposed relay selection policy has been examined in terms of the throughput and the packet error probability (PER) performance for various fading conditions.

### 4.3.1 System Model

An IVC system has been considered, where the wireless radio channels are frequency-flat. All the links exhibit block fading, which is considered constant during one time-slot and changes independently in the next time-slot. AWGN channels have been assumed, where the noise N has zero mean and variance n, i.e., \( N \sim CN(0,n) \). The source communicates with a destination through a set of W FD DF relays. To aid our analysis, the subscripts S, D, and R are affiliated with the source, the destination, and the selected relay, respectively. Then, the S-R-D system can be separated into the S-R and R-D subsystems. As shown in Figure 37, the direct link between source and destination is obstructed due to high attenuation in the propagation medium. The source and destination are equipped with single antennas, whereas the relay nodes are equipped with two antennas; one for reception of the source’s signal and one for transmission towards the destination. As far as the MIMO V2V system is concerned, the source and destination are equipped with one uniform linear array (ULA), whereas the relay is equipped with two ULAs; one for reception of the source’s signal and one for transmission towards the destination. It is considered that the multiple-antenna receivers use an MRC diversity scheme.

Due to DF relaying, the relays decode the received signal and then re-encode it for transmission to the destination. Hence, simultaneous reception and transmission take place resulting in LI from the relay’s output antenna to the relay’s input antenna. However, depending on the vehicle size, the antennas at the relays can be isolated and perfect LI cancellation can be achieved. Additional LI cancellation and suppression techniques can be also applied to further mitigate LI effect. The source is considered saturated and has always data to transmit. Besides, the data rate is equal to bits-per-channel-use (BPCU). In each time-slot, one relay is selected to establish S-R-D communication through FD transmissions. Since instantaneous CSI knowledge is required and M relays must be examined each time, the implementation complexity is increased in proactive relay selection scenarios leading to CSI overhead, especially in propagation scenarios with special mobility properties, i.e., high vehicle velocity and rapid changing road topologies. To reduce the amount of CSI overhead, the reactive policy is proposed, where the source power level is fixed and the signal is broadcasted towards all the relays. The relay’s transmission power levels are dynamically changed and power adaptation is performed. Hence, only the CSI of the R-D link is required for the distributed relay selection. In particular, this procedure intends to achieve a specific QoS level mapped to an SNR value based on the CSI of the R-D sub-channel, which is fed back to the source in order to perform the relay selection procedure. Then, power expenditure can be significantly reduced and network service provision is assured for extended time periods.

To denote whether or not the destination has successfully received the source’s packet, Acknowledgement/Negative-Acknowledgement (ACK/NACK) messages are sent by the destination, which in turn triggers a contention procedure, towards the source through the selected relay. Providing that a node fulfills the selection requirements, it can always deliver the packet to the destination. Otherwise, no candidates are available for relaying, and hence the source retransmits the data itself. At the start of each time-slot the relays should exchange signaling with the destination by sending pilot sequences. The latter estimates the CSI and notifies the relays on this CSI. The selection takes place among the relays which decoded the source’s signal. The set of these relays is denoted as \( \mathcal{w} \) where \( \mathcal{w} \) is the SNR in the R-D link and \( \mathcal{w} \) is an instantaneous SNR value at the receiver. The best relay is denoted by \( b_R \) and is chosen as [45]

\[
b_R = \arg\min_{w \in \mathcal{w}} (P_{w}) = \arg\min_{w \in \mathcal{w}} \left( \frac{n\text{SNR}}{g_{R,D}} \right), \tag{82}
\]
where $P_{re}$ is the relay’s transmission power and $g_{R_e,D}$ is the channel power in the R-D link.

4.3.2 Application of the Relaying Scheme on ITS

The application of the proposed system model and the relaying scheme on ITS-G5 standard [48] has been investigated in [49] [50]. The PHY and MAC layer of ITS-G5 is based on the most popular relevant standard, the IEEE 802.11p [51], with 10 MHz signal bandwidth, and has been presented in Section 2. It has been assumed that a safety critical ITS application is used, e.g., cooperative adaptive cruise control, which means that the ITS-G5 control channel is used. In ITS applications, the transmit data uses random access, i.e., the IEEE 802.11p. For safety critical applications, all vehicles have the obligation to state their presence and inform other vehicles for status changes periodically. For typical applications, the minimum transmission frequency of ITS messages is 10 to 25 Hz.

In the investigated system, 25 Hz minimum transmission frequency is assumed, which means that all vehicles participating in the network should broadcast at least one data packet every 40 msec. For the considered system, this means that every 40 msec a given vehicle will receive data from all the vehicles in the network. The size of each ITS message is typically less than 1kByte for reduced latency and they are transmitted using OFDM. These signals can be used to extract CSI for all the links with the adjacent nodes. However, in most cases, the nodes will exchange signals more frequently with much higher rate and therefore CSI is updated much more often. Relays are assumed to have FD capabilities. This means that each relay is able to directly retransmit the received message from the source without the need to content for access in the medium. Moreover, possible throughput performance degradation due to the retransmission procedure is significantly reduced.

![Diagram](image)

Figure 37: Simple representation of a V2V multi-relay communication system, where one FD DF relay is selected.

4.3.3 Simulation Results

This section presents results regarding the performance evaluation of the proposed reactive relay selection technique. In particular, the throughput and the PER are investigated assuming an ITS-G5 transmission scheme. In addition, $W = 2$ FD DF relays have been assumed and a complete ITS-G5 PHY and MAC simulator has been developed. The evaluated network consists of 6 nodes. Nodes 1 and 2 have no direct connectivity and therefore, a relay is necessary for communication. Nodes 4 and 5 are FD DF relays, while the two remaining nodes are conventional transceivers. The address field of the MAC header, which is not used in ITS-G5, is properly filled to indicate the need for relaying for every packet transmitted by Nodes 1 and 2. All messages are sent through the ITS-G5 control channel at 5.9 GHz with 10 MHz bandwidth. The basic rate (BPSK $\frac{1}{2}$ Coding) was used.
All Nodes are mobile with average speed 24 km/h. The values of the model parameters used are $R_S = R_R = R_D = 50$ m, $d_{SR} = d_{RD} = 1$ km, $\omega_S = \omega_R = \omega_D = \pi/4$, $\mu_S = \mu_R = \mu_D = 0^\circ$, $\gamma_S = \gamma_R = \gamma_D = 0^\circ$, $f_{S,max} = f_{R,max} = f_{D,max} = 100$ Hz. The simulation was performed for $P_A = K_A = L_A = Q_A = 20$, $P_E = K_E = L_E = Q_E = 3$, and 50 simulation trials.

The simulation results examine the effect of the use of outdated CSI in the relay selection scheme for different distribution of the effective scattering objects, i.e., different degree of local scattering in the azimuth domain (parameter $k = k_S = k_{SR} = k_{RD} = k_D$) at the vehicle nodes and different maximum value of the multipath elevation angles (parameter $\beta_{max} = \beta_{S,max} = \beta_{SR,max} = \beta_{RD,max} = \beta_{D,max}$). Since the multipath elevation angles are expected to be high in dense-urban areas, where the scatterers are usually dense and tall, values for the maximum multipath elevation angles up to $80^\circ$ are considered. During simulation, each node transmitted information with an average rate of two packets per msec. Thus, relay selection has been performed according to (82) based on the estimated SNR from the previous measurement. Performance of relay selection assuming perfect CSI is also provided as reference. Results for the achieved network throughput between nodes 1 and 2 are presented in Figure 38. In Figure 39, the results of PER for the ITS-G5 network and particularly for the link between nodes 1 and 2 are demonstrated. Packet losses may occur either due to poor channel conditions or due to collisions since a random access scheme is used.

The results show that the effects of outdated CSI in the degradation of performance increases as a) the degree of local scattering decreases and as b) the maximum multipath elevation angles increase. The aforementioned behavior is expected, since the wireless radio channel changes more rapidly, when the scatterers are distributed in a much larger range of azimuth and elevation angles. As the source, the relay, and the destination move among the effective scatterers in their vicinity, the Doppler effects become more significant and the regions of stationarity becomes smaller. These results also justify the inclusion of the third dimension of the proposed channel model.

For the MIMO V2V system, 6 nodes have been also assumed. Nodes 1 and 2 are unable to communicate directly with each other. The relaying mechanism has been introduced in the ITS-G5

![Figure 38](image_url): Network throughput versus SNR between nodes 1 and 2 for different degree of local scattering at the vehicle nodes and different maximum multipath elevation angles.
standard in order to establish connectivity between nodes 1 and 2. Nodes 5 and 6 have FD DF relaying capability and therefore can be used as relays, while nodes 3 and 4 are conventional single-antenna transceivers. Moreover, nodes 1, 2, 5, and 6 use two antennas. In transmitting mode, nodes 1 and 2 equally share the available power at both antennas. In receiving mode, an MRC algorithm is applied in order to fully exploit the diversity gain. It is noted that no standard modification is necessary in order to embed the MRC operation in an ITS-G5 system. The relaying mechanism was implemented in the ITS-G5 standard with the following modifications: Nodes 1 and 2 use the Address 4 field of the MAC header in order to request relaying and indicate the preferred relaying node. Therefore, if the Address 4 is empty then conventional transmission occurs. Otherwise, the node indicated in the specific address will act as a relay. Transmission is performed in the ITS-G5 Control Channel (10 MHz bandwidth), where 48 of the 64 ITS-G5 OFDM subcarriers were used. The transmitted packet was generated randomly with an average size of 1kbyte. The average SNR at the receiver varied from 10 to 20 dB. The carrier frequency was 5.9 GHz. Communication is performed with the ITS-G5 basic rate (BPSK ½ Coding). The values of the model parameters used to generate the MIMO channel matrix are $\delta_S = \delta_{RR} = \delta_{RT} = \delta_D = \lambda$, $\theta_S = \theta_{RR} = \theta_{RT} = \theta_D = 90^\circ$, $\psi_S = \psi_{RR} = \psi_{RT} = \psi_D = 45^\circ$, $\mu_S = \mu_D = 0^\circ$, $\psi_{SR} = \psi_{RD} = \pi/6$, $\gamma_S = \gamma_D = 0^\circ$, $\gamma_R = \pi/4$, and $f_{S,max} = f_{R,max} = f_{D,max} = 100$ Hz. The selected maximum Doppler frequency corresponds to an average speed of 24 km/h.

The simulation model has evaluated the gains offered by the use of an MRC diversity receiver. Moreover, the outdated CSI effects have been considered. Since the preferred relay station has been defined in the Address 4 field of the previously transmitted packet, the outdated CSI may have significant impact on the system performance, when no significant temporal correlation exists. Performance has been evaluated for different distributions of the effective scattering objects, i.e., different degree of local scattering in the azimuth domain (parameter $k = k_S = k_{SR} = k_{RD} = k_D$) and different maximum value of the multipath elevation angles (parameter $\beta_{max} = \beta_{Smax} = \beta_{SRmax} = \beta_{RDmax} = \beta_{Dmax}$). Note that large $k$ and small $\beta_{max}$ values indicate that the local scatterers are concentrated in specific angles at azimuth and elevation planes correspondingly. On the contrary, small $k$ and large $\beta_{max}$ values indicate that the local scatterers are distributed in a much larger range of azimuth and elevation angles. Then, the wireless radio channel rapidly changes and the Doppler effects become more significant. Figure 40 presents the achieved net throughput between nodes 1 and 2 with DF relaying and MRC reception. To evaluate the diversity and relaying gain, the
performance of conventional single-antenna point-to-point transmission between nodes 3 and 4 is also presented. In Figure 41, the results of PER for the ITS-G5 network are also demonstrated.

Packet losses may occur either due to poor channel conditions or due to collisions in the random access scheme. More specifically, in high SNR packet loss occurs mostly due to packet collisions and thus the gain of the diversity technique is decreased. The results show that a) there are significant gains from the use of the MRC diversity receiver in the link performance and b) the performance degrades, when the scatterers are distributed in a large range of azimuth and/or elevation angles. These results also justify the inclusion of the third dimension of the geometrical channel model.

Figure 40: Network throughput versus SNR for a) nodes 1 and 2 with relaying and MRC reception and b) nodes 3 and 4 with single-antenna point-to-point links for different degree of local scattering and different maximum multipath elevation angles.

Figure 41: Packet Error Rate for a) nodes 1 and 2 with relaying and MRC reception and b) nodes 3 and 4 with single-antenna point-to-point links for different degree of local scattering and different maximum multipath elevation angles.
4.4 Beamforming and Limited Feedback Techniques

An important objective in the framework of this WP was to design and evaluate beamforming transmission by implementing adaptive control algorithm schemes suitable for T2T and T2I channels. Towards satisfying this objective, electronically switched parasitic array radiators (ESPARs) have been employed [52], [53], [54], which provide the required pattern reconfigurability. The research work presented in this section involves a proof-of-concept indoor experiment where an extended open loop beamforming (eOLB) is realized through pattern selection using pattern reconfigurable antennas [55]. Specifically, the Tx-Rx system performs a beam switching and selection technique. This technique provides adaptive operation to a cooperative system selecting the best beam for each case and maximizing SNR. Furthermore, this strategy can be expanded in both ends of the T2T and T2I link. In that case, ESPAR antennas can be used in every truck (operating in both transmitting and receiving mode) and no feedback link will be required in order to communicate the pattern selection decision. Figure 42 presents two typical examples of a cooperative T2T communication system. The beam combination for the “takeover” scenario employs the omnidirectional patterns of the antenna, whereas for a “platooning” scenario, the best combination employs the directive antenna beams that point each other.

The eOLB technique is accomplished by using two ESPAR antennas, which can be regarded as excellent candidates to satisfy the necessity of pattern reconfigurability. The ESPAR antennas exhibit significantly reduced complexity and low cost compared to other MIMO systems with multiple RF-chains and considerable inter-element distances. ESPARs are able to operate with only one active element, and as a result with a single RF-chain, while preserving the majority of the features of a MIMO system, such as beamforming and spatial multiplexing capability [56]. The rest of the radiators that comprise the ESPAR antenna are passive parasitic elements that do not involve complex feeding and control circuits and can be distributed around the active element in several arrangements and in extremely small distances (λ/5 – λ/16). This leads to the second main advantage of the ESPAR antennas, which is the compact size that they can maintain. The comparably small dimensions can provide fitness and applicability in cases where severely limited space is available, e.g., inside the side mirror of a truck.

![Diagram of ESPARs in a cooperative T2T communication system](a)
Section 4.4.1 presents a new printed ESPAR antenna design along with the simulation results. Moreover, it includes the fabrication details of the two prototypes and the $S_{11}$ and radiation pattern measurement results. Measured return loss and far-field patterns demonstrate a good agreement with the design. In Section 4.4.2, the proof-of-concept indoor experiment is demonstrated along with the SNR comparison results. A series of indoor measurement tests were held in the University of Piraeus premises in order to demonstrate the beamforming advantage of the antenna and evaluate the enhancement of an IEEE 802.11p system. The obtained results prove that the proposed ESPAR equipped system achieves a significant SNR increase.

4.4.1 Configuration of the Printed ESPAR Antenna

4.4.1.1 ESPAR antenna design

The ESPAR antenna model is designed in the CST 3D electro-magnetic simulator [57]. The detailed layout of the antenna is presented in Figure 43, along with the basic parameterized dimensions of the structure. The proposed ESPAR consists of three elements (one active printed monopole and two parasitic printed monopoles). A RO-4725 JXR (57 x 40 mm) dielectric substrate ($h = 0.78$ mm, $\varepsilon_r = 2.55$), purchased from Rogers Corporation, is employed in order to “accommodate” the antenna elements along with the controlling circuit. The active printed antenna element is composed by an end-launch (edge-fed) SMA connector, a microstrip quarter wavelength transformer (to convert the 50 Ω impedance to the theoretical 37 Ω input impedance) and the printed $\lambda/4$ monopole.
Figure 43: The “top-view” layout of the ESPAR simulation model in CST. The ground plane stripe (printed at the back side of the panel) is also visible since the substrate is left out of the diagram.

The dielectric panel is excluded from the illustration in order to make the ground plane at the back side visible. Beamforming is implemented by allocating two parasitic printed monopoles next to the active element (one at each side), in a relatively short distance ($\lambda/6$). Strong mutual coupling is induced and radiation pattern reconfigurability is achieved by employing two PIN diode electronic switches (with two operation states – ON or OFF) in order to control which parasitic element is grounded. The monopole is connected to the ground plane stripe by a plated via, forming an L-shaped reflector that focuses the antenna beam to the opposite direction. Figure 44 demonstrates the two equivalent circuits that are employed in order to simulate the two PIN diode states.

The number of possible pattern combinations is four. However, the ON-ON operating mode is excluded since it does not provide a satisfactory return loss response for the desired frequency of 3.55 GHz, ending up with three feasible pattern states (OFF-OFF, ON-OFF, OFF-ON). The antenna design also includes a two-branch DC bias network that is necessary in order to apply voltage to the two PIN diodes and change their operating mode. Two RF chokes are also realized by two chip inductors that are located at the start of the DC bias lines. Figure 45 illustrates a diagram of the printed ESPAR antenna along with the three different radiation patterns that correspond to the three PIN diode combinations (0,0 – OFF-OFF / 1,0 – ON-OFF / 0,1 – OFF-ON). The basic dimensions of the antenna design are given in Table 4 in mm.
Figure 45: The layout of the 3-printed monopole ESPAR antenna design. Three radiation patterns that correspond to the three ON-OFF combinations of the PIN diodes are roughly depicted.

Table 4: ESPAR Antenna Dimensions

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>t</td>
<td>thickness of the copper and the ground plane layers</td>
<td>0.035</td>
</tr>
<tr>
<td>L_m</td>
<td>length of the active printed monopole</td>
<td>13.8</td>
</tr>
<tr>
<td>W_m</td>
<td>width of the active printed monopole</td>
<td>1.8</td>
</tr>
<tr>
<td>L_p</td>
<td>length of the parasitic monopole</td>
<td>16.2</td>
</tr>
<tr>
<td>W_p</td>
<td>width of the parasitic monopole</td>
<td>3.0</td>
</tr>
<tr>
<td>L_qw</td>
<td>length of the $\lambda/4$ transformer</td>
<td>12.0</td>
</tr>
<tr>
<td>W_qw</td>
<td>width of the $\lambda/4$ transformer</td>
<td>2.3</td>
</tr>
<tr>
<td>dis</td>
<td>distance between the active and the parasitic elements</td>
<td>$\lambda/6$</td>
</tr>
<tr>
<td>gap</td>
<td>gap between the parasitic monopole and the grounding metal pad</td>
<td>0.4</td>
</tr>
<tr>
<td>via_r</td>
<td>radius of the plated via (through hole)</td>
<td>0.3</td>
</tr>
</tbody>
</table>

4.4.1.2 Simulation Results

Figure 46 depicts the simulated reflection coefficient ($S_{11}$) of the printed ESPAR antenna for the two basic operating states of OFF-OFF and ON-OFF. The OFF-ON mode is not included in the simulation plot since it is identical to the ON-OFF mode (due to the design symmetry). As it is apparent from Figure 46, both $S_{11}$ responses exhibit a satisfying return loss (below -15 dB at 3.55 GHz) and a remarkably wide -10 dB bandwidth (both over 22%). The two corresponding radiation patterns (blue/omni – red/directive) of the antenna are plotted in Figure 47 for the E plane (left) and the H plane (right). A significant 2-6 dB directivity increase can be noticed between the two operation modes, depending on which $\theta$ angle we cut the 3D pattern.
Figure 46: Simulated reflection coefficients ($S_{11}$) versus frequency for the OFF-OFF and ON-OFF antenna states.

![Simulated reflection coefficients](image)

Figure 47: The corresponding far-field 3.55 GHz directivity radiation patterns of the printed ESPAR antenna at the (a) E plane and (b) the H plane.

4.4.1.3 Measurement Results of the ESPAR Antenna

Figure 48 shows a prototype of the 3-element printed ESPAR antenna after its fabrication and assembly. All the necessary components (connector, PIN diodes, inductors, DC pins) were soldered by hand. The RF choke inductors were purchased from Coilcraft (0302CS-34NXJLU, 34 nH). The control circuit is based on the SMP1320-040LF PIN diode electronic switches manufactured by Skyworks. The simulated and the measured reflection coefficients ($S_{11}$) are presented in Figure 49. A slight discrepancy is observed between the simulation and the measurement results, probably due to fabrication and soldering inaccuracies. However, the measured $S_{11}$ parameter for both states remains
Figure 48: The prototype printed ESPAR antenna. The close distance between the active and the parasitic elements makes the array relatively compact.

Figure 49: The simulated and the measured reflection coefficients ($S_{11}$) versus frequency for the OFF-OFF and ON-OFF states of the antenna.

below -15 dB and the operating bandwidth is preserved in sufficient levels (above 14%). Figure 50(a) contains the two corresponding measured far-field radiation patterns at the E plane for the 3.55 GHz. A considerable 5-6 dB difference between the directivities of the two states can be noticed. A small deviation at the expected upward tilt angle of the radiation is also observed. This is presumably caused by the big plastic structure that is employed in order to mount the antenna in the anechoic chamber. In Figure 50(b), the two far-field directivity patterns are plotted at the H plane. The OFF-OFF mode demonstrates a non-perfectly (quasi) omnidirectional behavior (“squeezed shape”), which is mainly induced by the planar structure of the antenna along this axis.
4.4.2 System Level Experiment and Performance Evaluation

In order to validate the beam-shaping advantage of the introduced ESPAR design, a real-time over-the-air test is performed using the two prototype ESPARs with a software defined radio (SDR) implementation of an IEEE 802.11p [58] Transmitter (Tx) and Receiver (Rx). SDR is a radio communication system where most of the traditional hardware components, are implemented on software level. The main idea of SDR is that the digitalization of the signal should be as close as possible to the antenna, thus most, if not all, of the physical layer functions can be modified through firmware, operating on general-purpose processors. As a result, there is an advanced flexibility on supporting multiple waveform standards and adapt to any transmission feature based on a selected technique. SDR systems are capable to operate in an ultra-wide range of frequencies and compatible with multiple radio interfaces. A good introduction to the topic can be found in [59] and [60].

Since, the initial intention was to evaluate the antennas in vehicular systems, an in-house implementation of the IEEE 802.11p was developed. The SDR hardware that was used for digitization of the waveforms was the universal software radio peripheral (USRP) N210 [61] with SBX daughter-boards. The USRP N210 is an open source hardware platform that uses UHD open source middleware and connects to the host computer via Gigabit ethernet interface. Both receiver and transmitter operations can be implemented on the board on a wide range of frequencies, up to 4.4 GHz. The main subsystems provided by the USRP motherboard are ADC/DAC, clock synchronization, automatic gain control (AGC), and field programmable gate array (FPGA).

Design (for Tx) and analysis (for Rx) of the wireless standard waveform was performed by two laptops using in-house developed software, implemented in C++ (in order to achieve real time processing). The host laptops are also connected with two single board computers. The single-board computers provide through their general purpose input/output (GPIO) interface the necessary voltage in order to drive the antenna PIN diodes. The Tx transmits periodically a single signal frame every 50 msec. The IEEE 802.11p frame starts with a sequence of 10 Short Preamble Signals, where two of them are π-shifted, followed by two long preambles. The initial Rx operations include signal detection, preamble detection and synchronization. The access medium is monitored until an incoming signal is identified by the detector. A second detector, detects IEEE 802.11p preambles and achieves synchronization based on [1]. Moreover, after synchronization, the Rx uses the
preamble overhead to estimate Received Signal Strength and SNR. Consequently, it automatically switches to a different pattern. After measuring all patterns, Rx decides on the pattern of preference. A simple scenario with two portable units (Tx and Rx) is considered. Rx performs beam diversity, Tx performs beamforming after relevant feedback. Each unit is mounted on an office trolley and uses the proposed printed ESPAR reconfigurable antenna (see Figure 51). During the evaluation test, the Tx sends a signal frame every 50msec. After Rx synchronization, signal reception, and SNR estimation, the host Rx automatically sends a command to the single-board computer, in order to change the Rx pattern. For experimentation purposes, the host Rx laptop is also connected through Ethernet with the Tx single board computer. Thus, the Rx is also able to control the Tx pattern. This is an indirect way to emulate the CSI feedback from Rx to Tx, that can be used for beamforming. The carrier frequency for the experiments was 3.55 GHz (5.9 not supported by the specific USRP). Figure 52 presents an illustration of the eOLB experiment set up (Tx and Rx) including all the equipment that is used in each unit. In Figure 53 and Figure 54, the set up diagram of the Tx and the Rx are illustrated respectively. Tx and Rx are able to select among three patterns each: 0-0 (OFF-OFF), 1-0 (ON-OFF) and 0-1 (OFF-ON). Thus, 9 possible pattern combinations can be derived. During the experiment, the Rx spans all 9 patterns periodically and selects the combination that provides the best SNR. Figure 55 illustrates the four different cases that are tested for the aforementioned scenario. In all four cases, the Rx remains in a fixed position. It should be noted that the planar ESPAR is always positioned in parallel with the trolley’s long side orientation (see Figure 51 for clarification). In the first case, the Tx is placed aligned at the left side of the Rx (with a 4m separation). In the second case, the Tx is moved to the opposite side of the Rx (to the right) at the same distance. At the third case, the Tx returns back to its first position but it is dis-oriented (rotated by 90°). Finally, the fourth case involves a NLOS (Non-Line of Sight) condition measurement, where the Tx is positioned out of the room.

(a) (b)

Figure 51: (a) The two portable units (Tx and Rx) as they were mounted on the trolleys. (b) Each unit consists of a USRP, a single-board, a voltage divider and a prototype printed ESPAR antenna.
Figure 52: The set up of the eOLB indoor experiment. Both Tx and Rx units consist of an ESPAR antenna, a voltage divider, a Beaglebone, a USRP N210 unit and a laptop.

Figure 53: The set up diagram of the Tx.
The measurements in Positions 1 and 2 produced similar results. In the first position, 99.6% of the measurements selected pattern Tx(1,0)-Rx(0,1) as the best selection, while in position 2, 99.2% of the cases indicated pattern Tx(0,1)-Rx(1,0). In both cases, the dominant patterns were the expected ones, i.e. the combination that maximizes radiated power through the Line Of Sight (LOS) direction. In Figure 56, the results are presented as a set of empirical CDFs of the SNR Gain of the pattern combos relative to the OMNI combination Tx(0,0)-Rx(0,0), which is denoted by the vertical red line. The superiority of the aforementioned combinations is undoubtable, providing in real channel conditions more than 6.5dB average gain for each case. The performance of combinations Tx(0,0)-Rx(0,1) and Tx(0,0)-Rx(1,0) (i.e. OMNI transmitter) is, as expected, 2-3 dB inferior. Moreover, the performance of the “opposite” pattern combos (the patterns focusing on the opposite of the LOS direction) is presented. These patterns are outperformed by the OMNI combo by more than 2.5 dB on average, while more than 10% of measurements provide gain lower than -4 dB.
In Position 3 (see Figure 57(a)), the dominant pattern combination is Tx(0,0)-Rx(0,1), i.e. Tx is OMNI while Rx is selecting the pattern that targets Tx. However, the superiority of the pattern is not as clear as in Positions 1 and 2. Despite the fact that, it provides average gain of 4.2 dB, pattern combinations Tx(1,0)-Rx(0,1) and Tx(0,1)-Rx(0,1) were measured more suitable with higher SNR in 32% of the cases (average SNR gain 3.38, 3.9 dB respectively). Since no Tx pattern maximizes radiated power towards Tx-Rx direction, the SNR gain reduces. However, SNR gain remains relatively high for every possible Tx pattern selection. Finally, the results of Point 4 are presented in Figure 57(b). It is noted that in the NLOS scenario, there was intense mobility of scatterers in the measurement room. In this case, no dominant pattern can be identified and only 3 combinations actually outperform the OMNI combo. However, by assuming that through the periodic measurement, the best SNR pattern combination is selected, the SNR gain (relative to the OMNI combo) is quite significant, exceeding 7.32 dB.
4.5 Spatial Modulation

Throughout this task, the introduction of spatial modulation (SM) in T2T/T2I wireless links has been investigated. SM can be considered as an ideal candidate for IVC systems, since it can improve system performance while minimizing transceiver complexity and cost. More specifically, SM entirely avoids inter-channel interference (ICI) and inter-antenna synchronization, and only requires a single RF chain at the transmitter, which is of critical importance for the small-sized vehicle mounted transmission systems. Moreover, the receiver design is inherently simpler compared to other multiple antenna systems, e.g., V-BLAST, since complicated ICI cancelation algorithms are not required [62]. Furthermore, with the advent of multiple antennas technologies incorporated on vehicles, it is envisaged that transmission technologies such as SM can be efficiently used in ITS systems for safety devices communications, ad hoc peer-to-peer networks and V2V communications. The specific outcomes of this task are as follows [63] [64]:

- A complete mathematical framework has been developed for the performance analysis of SM operating over T2T/T2I channels. New analytical expressions for the BEP of space shift keying (SSK) operating over multiple scattering conditions have been derived. When the transmitter is equipped with two antennas the resulting analytical expressions are exact, whereas for an arbitrary number of transmit apertures tight upper bounds have been obtained. Tight lower bounds for the ABEP of SM systems operating over multiple scattering conditions have been derived. Finally, an asymptotic analysis for SM has been carried out that provides useful insight as to the factors affecting system performance.
- In order to evaluate the performance of SM in a realistic manner, a physically motivated model for fading channels, namely the called second order scattering (SOS) radio propagation channel has been considered.
- New methods for further improving the error performance of SM in T2T/T2I wireless links have been considered, namely the incorporation of channel coding and transmit diversity schemes that increase the achievable diversity gain.
- The performance of SM systems has been compared to the one of other conventional MIMO schemes, such as VBLAST with zero forcing (ZF) and successive interference cancellation (SIC) receivers, using simulation.

4.5.1 System and Channel model

4.5.1.1 System model

A generic IVC system has been considered, where both vehicles, equipped with multiple antennas, are in motion and operating under local scattering environments as depicted in Figure 58. We assume that the source vehicle employs \( N_t \) transmit antennas and the destination vehicle has \( N_r \) receive antennas. The transmitter encodes blocks of \( \log_2(N_t) \) data bits into the index of a single transmit antenna, \( t_i \), for \( i = 1, \ldots, N_t \). Such a block of bits is hereafter referred to as “message.” It is assumed that the \( N_t \) messages are transmitted by the encoder with equal probability. During each time slot, the \( t_i \)-th antenna is switched on for data transmission while all other antennas are kept silent, i.e., \( N_t - 1 \) antennas do not transmit. At the receiver, an \( N_t \) hypothesis detection problem is solved to retrieve the active transmit antenna index, which results in the estimation of the unique sequence of bits emitted by the transmitter.

4.5.1.2 Channel model

Depending on the propagation characteristics around the communicating vehicles, a variety of vehicular environments are often distinguished in the literature, such as highways, rural roads, suburban and urban streets. This variety of environments will result in different propagation conditions, which can be modeled by a weighted combination of LoS, single scattering and double
scattering fading. Throughout this task, the SOS distribution has been adopted to model signal fluctuations, since it has been found to accurately approximate small-scale fading for a wide range of IVC scenarios.

In Figure 58, an example of a communication scenario is depicted, where SOS conditions exist under the assumption of a LoS path. For this scenario, a vehicle denoted as $V_1$ is moving in a straight road and communicates with another vehicle $V_2$. It is assumed that due to surrounding buildings and objects, three scattering clusters exist. More specifically, a part of the electromagnetic energy transmitted by $V_1$ is scattered by the cluster surrounding Building 1. Each scatterer in this cluster also scatters some energy to the cluster surrounding Building 3 and the remaining energy is received by vehicle $V_2$. Scattering also occurs from cluster surrounding Building 2 and the resulting energy is received by $V_2$. The resulting channel transfer function is therefore affected by two components, i.e., it follows the so-called double-Rayleigh distribution, while the remainder is modeled as a single Rayleigh. In addition to the scattering, there is a strongly dominant signal along a LoS between $V_1$ and $V_2$. Hereafter, it is also assumed that the underlying channel is quasi-static, which is well justified for IVC scenarios in rush-hour traffic. The transmitted signal can be expressed as a linear combination of signal components with constant, Rayleigh, and double-Rayleigh distributed amplitudes.

This multiple scattering model yields as special cases the Rician, Rayleigh, and double-Rayleigh fading channels. The complex channel gains $\alpha_{i,\ell}$, $i = 1, \ldots, N_t$ and $\ell = 1, \ldots, N_r$ are defined as

$$\alpha_{i,\ell} = a_{i,\ell} \exp\left(\imath \phi_{i,\ell}\right)$$

where $a_{i,\ell}$ and $\phi_{i,\ell}$ are the envelopes and phases of the link from the $i$-th transmit antenna to the $\ell$-th receive antenna, respectively. Under this considered multiple scattered scenario, $\alpha_{i,\ell}$ can be expressed as [2]

$$\alpha_{i,\ell} = w_{0,i,\ell} \exp(\imath \phi_{i,\ell}) + w_{1,i,\ell} H_{1,i,\ell} + w_{2,i,\ell} H_{2,i,\ell} + H_{3,i,\ell} = \sum_{n=0}^{3} C_{n,i,\ell}$$

(83)

where $C_{0,i,\ell} = w_{0,i,\ell} \exp(\imath \phi_{i,\ell})$, is the LoS component with constant magnitude and uniformly distributed phase over $[0, 2\pi)$, $C_{1,i,\ell} = w_{1,i,\ell} H_{1,i,\ell}$, $C_{2,i,\ell} = w_{2,i,\ell} H_{2,i,\ell}$, and $C_{3,i,\ell} = H_{3,i,\ell}$, where $H_{n,i,\ell}$ are i.i.d. isotropic Gaussian processes having zero mean and unit variance, and are nonnegative real-valued constants that determine the mixture weights of the multiple scattering components.

When a LoS component is not present, the scenario corresponds to a blind bend propagation environment for V2I communications (see Figure 59).
For this scenario, a vehicle moving along a curved street communicates with a fixed road-side unit (RSU) deployed on the other side of the curved road. The RSU, which acts as a fixed receiver, provides the driver with information related to traffic and road conditions. Due to the curvature of the street, the driver is not able to see the road ahead and thus the LoS component can be ignored. It is noted that this model does not consider higher-order scattering terms since first- and second-order components bear more energy than higher order components. Thus, by omitting higher order components, the resulting channel model is considerably simplified without significant accuracy loss. The IVC scenarios that were used to obtain the performance evaluation results include a V2I scenario and a V2V scenario. In the V2I scenario, the source vehicle is equipped with two transmit antennas and moves in a curved road (no LoS component exists), whereas the RSU is equipped with $N_r \geq 1$ antennas. In the V2V scenario, the source vehicle is equipped with two transmit antennas whereas the destination vehicle has $N_r \geq 1$ receive antennas operating in a multiple scattering environment with a LoS component.

4.5.2 Average Bit Error Probability of Uncoded SSK

A tight upper bound for the ABEP of SSK can be obtained from [65]

$$
\overline{P} \leq \frac{N_t^{-1}}{\log_2(N_t)} \sum_{t_1=1}^{N_t} \sum_{t_2=t_1+1}^{N_t} N_b(t_1, t_2) \text{PEP}_{SSK}(t_1 \rightarrow t_2)
$$

where $N_b(t_1 \rightarrow t_2)$ is the number of bit errors having occurred when the receiver decides that antenna $t_2$ instead of antenna $t_1$ has been active and $\text{PEP}_{SSK}(t_1 \rightarrow t_2)$ denotes the average pairwise error probability (PEP) related to the pair of transmit antennas $t_1$ and $t_2$, for all $t_1, t_2 = 1, 2...N_t$, defined as [65]

$$
\text{PEP}_{SSK}(t_1 \rightarrow t_2) = E \left\{ Q \left( \sqrt{\frac{N_t}{2}} \sum_{\ell=1}^{N_t} Z^{t_1, t_2}_\ell \right) \right\} = \frac{1}{\pi} \int_0^{\pi/2} \prod_{j=1}^{N_{t_1}} \prod_{\ell=1}^{N_t} M Z_{\ell, t_1, t_2} \left( \frac{\overline{\gamma}}{2 \sin^2 \theta} \right) d\theta
$$

where $\overline{\gamma} = E_m/4N_0$ is the symbol-energy-to-noise-spectral density ratio with $E_m$ being the average energy transmitted by each antenna that emits a nonzero signal, $N_0$ the noise double-sided power spectral density and $Z_{\ell, t_1, t_2} = |a_{t_2, \ell} \exp(j\phi_{t_2, \ell}) - a_{t_1, \ell} \exp(j\phi_{t_1, \ell})|$. When two transmit antennas are used, i.e. when $N_t = 2$, the ABEP of SSK is equal to $\text{PEP}_{SSK}(t_1 \rightarrow t_2)$. 

---

Figure 59: A blind bend propagation scenario for V2I communications.
Using the tight approximation for the Gaussian Q-function, i.e., $Q(x) = (1/12)\exp(-x^2/2) + (1/4)\exp(-2x^2/3)$ as well as the definition of the MGF of a random variable $X$ an expression accurately approximating (8) can be deduced as [66]

$$
\bar{P} \approx \frac{1}{12} \prod_{i=1}^{N_r} M_{Z_{n,i}}(\sqrt{\gamma}) + \frac{1}{4} \prod_{i=1}^{N_r} M_{Z_{n,i}}\left(\frac{2\sqrt{\gamma}}{3}\right).
$$

(86)

The MGFs in (8) can be expressed as [67]

$$
M_{Z_{n,i}}(s) = \frac{1}{2\pi} \int_{0}^{\infty} \text{Re} \{ R \Phi_{a_{n,i}}(R)\Phi_{a_{n,i}}(R)\} dR
$$

(87)

where $\Phi_{a_{n,i}}$ are the Hankel transforms of the envelopes $a_{n,i}$. Due to the independence of $C_{n,i}$ the Hankel transforms in (8) can be expressed as [2]

$$
\Phi_{a_{n,i}}(R) = \prod_{n=0}^{2} \Phi_{e_{n,i}}(R)
$$

(88)

where

$$
\Phi_{e_{n,i}}(R) = J_0(w_{0,i}, R)
$$

$$
\Phi_{e_{n,i}}(R) = \exp\left(-w_{1,i}, R^2 / 4\right)
$$

$$
\Phi_{e_{2,i}}(R) = \frac{4}{4 + w_{2,i}, R^2}.
$$

(89)

For a V2I scenario where a LOS component is absent, the MGF in (8) can be deduced in closed form whereas for the V2V scenario it can be evaluated numerically in an efficient manner by means of numerical integration.

In order to provide further insight on the impact of system parameters to the error rate performance, an asymptotic analysis at high values of the average SNR has been carried out. In particular, a generic analytical expression, which becomes asymptotically tight at high SNR values, can be derived for the average PEP appearing in (8), as follows [68]

$$
\text{PEP}_{SSK}(t_1 \rightarrow t_2) \approx \frac{2^{N_r-1} \Gamma(N_r + 1)}{\sqrt{\pi} \Gamma(N_r + 1) \prod_{i=1}^{N_r} c_i} \gamma^{-N_r}
$$

(90)

where

$$
c_i = \frac{1}{4} \int_{0}^{\infty} \prod_{i=1}^{2} \Phi_{a_{n,i}}(\sqrt{y}) dy.
$$

(91)

As it becomes obvious the diversity order of the considered system equals to the number of receiving antennas where the channel parameters affect the coding gain only.

4.5.3 Average Bit Error Probability of Coded SSK

When coded SSK is employed, the input signal is first encoded by a convolutional encoder. The encoded data are interleaved by a random block interleaver and transmitted through the wireless channel using SSK modulation. It is assumed that perfect interleaving at the transmitter and
deinterleaving at the receiver is used. At the receiver, hard or soft decision decoding are considered. For hard decision decoding, a union bound on the ABEP, can be obtained as [63]

$$\bar{P}_{ub} \leq \sum_{k=d_{\text{free}}}^{\infty} c_k Z_k$$

(92)

where $d_{\text{free}}$ is the free distance of the code,

$$Z_k = \left\{ \begin{array}{ll}
\sum_{e=(k+1)/2}^{k} \left( \frac{k}{k/e} \right) p^e (1-p)^{k-e}, & k \text{ odd} \\
\frac{1}{2} \left( \frac{k}{k/2} \right) p^{k/2} (1-p)^{k/2} + \sum_{e=(k/2)+1}^{k} \left( \frac{k}{k/e} \right) p^e (1-p)^{k-e}, & k \text{ even} 
\end{array} \right.$$ 

(93)

with

$$p = \frac{1}{\log_2(N_t)} \sum_{t_1=1}^{N_t} \sum_{t_2=1}^{N_t} N_k(t_1,t_2) \text{PEP}_{\text{SSK}}(t_1 \rightarrow t_2).$$

(94)

The coefficients $c_k$ can be deduced as

$$\left. \frac{\partial T(D,N)}{\partial N} \right|_{N=1} = \sum_{k=d_{\text{free}}}^{\infty} c_k D^k.$$ 

(95)

In (9), $T(D,N)$ is the transfer function of the code, $N$ is an indicator variable taking into account the number of the erroneous bits and $D$ depends on the underlying PEP expression. A union bound on the ABEP of the considered coded communication system can be obtained by employing a Chernoff bound on the Gaussian Q-function as

$$\bar{P}_{ub} \leq \left. \frac{\partial T(D,N)}{\partial N} \right|_{N=1,D=q}$$

(96)

where

$$q = \left\{ \begin{array}{ll}
\prod_{t=1}^{N_t} M z_{t/2} \left( \frac{P}{2} \right), & N_t = 2 \\
\prod_{t_1=1}^{N_{t_1}} \prod_{t_2=1}^{N_{t_2}} M z_{t_1/2} z_{t_2/2} \left( \frac{P}{2} \right), & N_t > 2.
\end{array} \right.$$ 

(97)

### 4.5.4 Average Bit Error Probability of SM

When SM instead of SSK is employed, the considered MIMO system can send information via $M$ complex symbols, $\chi_j = |\chi_j| \exp(i\theta_j)$, $j = 1,2,...,M$. Without loss of generality, it is assumed that the considered system operates under i.i.d fading with constant-modulus modulation, i.e., $|\chi_j| = \kappa_0$. The ABEP of SM can be tightly bounded as [65]

$$\bar{P} \leq \text{ABEP}_{\text{signal}} + \text{ABEP}_{\text{spatial}} + \text{ABEP}_{\text{joint}}$$

(98)

where $\text{ABEP}_{\text{signal}}$, $\text{ABEP}_{\text{spatial}}$ and $\text{ABEP}_{\text{joint}}$ show how the error performance of SM is affected by the signal constellation diagram, the spatial constellation diagram and the interaction of both signal
and space constellation diagrams, respectively. Under generalized fading, the term \( \text{ABEP}_{\text{signal}} \) when either M-ary phase shift keying (M-PSK) or M-ary quadrature amplitude modulation (M-QAM) are employed, can be readily evaluated. Assuming constant modulus modulation \( \text{ABEP}_{\text{spatial}} \) and \( \text{ABEP}_{\text{joint}} \) can be deduced as [65]

\[
\text{ABEP}_{\text{spatial}} = \frac{N_t \log_2(N_i)}{2 \log_2(N_i M)} P_{\text{SM}}(t_1 \rightarrow t_2)
\]

\[
\text{ABEP}_{\text{joint}} = \left[ \frac{M(N_t - 1) \log_2(M)}{2 \log_2(N_i M)} + \frac{N_t(M - 1) \log_2(N_i)}{2 \log_2(N_i M)} \right] P_{\text{SM}}(t_1 \rightarrow t_2)
\]

(99)

where \( P_{\text{SM}}(t_1 \rightarrow t_2) \) can be readily obtained from \( P_{\text{SSK}}(t_1 \rightarrow t_2) \) by multiplying the average SNR by \( \kappa_0 \).

4.5.5 Transmit Diversity Schemes for SSK

Conventional SSK–MIMO systems can offer a diversity gain that depends on only the number of receive–antennas. Thus, it is of interest to develop new SSK concepts with a higher diversity gain, and, so, with a better ABEP. Hereafter we consider the so-called transmit orthogonal signal diversity – SSK (TOSD-SSK) scheme where the coherent summation of signals in (84) is avoided and instead the complex channel gains in (8) sum up incoherently, i.e., in power.

The TOSD–SSK modulation scheme can work when the signals have just either good cross– or good auto–correlation properties. The achieved error probability conditioned upon fading channel statistics is as follows [69]:

\[
\text{ABEP}(\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_{N_t}) \leq \text{ABEP}^B(\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_{N_t}) = \frac{1}{N_t - 1} \sum_{i=1}^{N_t} \sum_{j=1}^{N_t} Q\left( \sqrt{\frac{1}{N_t} \sum_{l=1}^{N_t} \left| \alpha_{i,l} \right|^2 + \left| \alpha_{i,j} \right|^2} \right)
\]

(100)

As it can be observed, the end–to–end SNR does not depend on the differences of complex multiply scattered fading channels, but it is given by their incoherent summation. This introduces an increase in the diversity order or \( 2N_t \), that is two times the diversity order achieved by the conventional SSK. When two transmit antennas are employed, a full diversity gain of \( N_i N_t \) can be achieved.

4.5.6 Performance Evaluation

In Figure 60, the analytical BEP of SSK for the V2V scenario with LoS component is illustrated as a function of the SNR per symbol, using different values of receiving antennas. The exact BEP is also evaluated using Monte-Carlo simulations and it is included as a reference (illustrated by square patterns). It is observed that the derived BEP provides a very tight approximation for the entire SNR region. For all cases considered, it is observed that the asymptotic analysis correctly predicts the diversity gain of the SSK transmission scheme and that the asymptotic ABEP coincides with the exact one in the high SNR regime. In Figure 61, the performance of SSK with two transmit and receive antennas in a V2I scenario is compared with the performance of several well-established MIMO schemes with BPSK, such as i) MIMO 2x2 with ZF equalizer, ii) MIMO 2x2 with minimum mean square error (MMSE) equalizer and SIC, iii) MIMO with ML detection and iv) space time block codes (STBC) based on the Alamouti scheme with \( N_t = 2 \) and \( N_r = 1 \). As a reference, the performance of the SISO scheme is also depicted in the same figure. As it can be observed, SSK outperforms SISO and MIMO with ZF for medium and high SNR values. Moreover, it even outperforms MIMO with MMSE-SIC at high SNR regime.
Figure 60: ABEP of uncoded SSK in a V2V environment.

Figure 61: ABEP of multiple antenna schemes in a V2I environment with curved road.

However, SSK performs worse than STBC and MIMO with ML. Nevertheless, this disadvantage is compensated by its significant lower hardware and signal processing complexity compared to these two schemes. The analytical framework developed in this task offers a simple and reliable method
to estimate BEP performance of SM systems in multiple scattering channel models without resorting to

![Figure 62: ABEP of SM-QPSK in a V2I environment with curved road.](image)

![Figure 63: ABEP bounds of coded SSK in a V2I environment with curved road.](image)

lengthy simulations. Therefore, it is useful for the development, design, and test of future V2V and V2I communication systems. In Figure 62, the ABEP of MIMO SM-QPSK (M = 4), for the V2I
scenario with non LoS component is illustrated as a function of SNR, using different values of $N_r$. It is observed from Figure 62 that the derived ABEP improves as SNR or $N_r$ increases.

Figure 63 considers a coded SSK system employing a rate 1/2 convolutional encoder. The transfer function of the code is given by [70]

$$T(D, N) = \frac{ND^5}{1-2ND} = \sum_{k=5}^{\infty} 2^{k-5} N^{k-4} D^k.$$  (101)

It is shown, that the incorporation of convolutional coding significantly improves the performance of SSK systems, even when a small number of receive antennas is employed. Specifically, when three receiving antennas are employed, i.e. when $N_r = 3$, the proposed system configuration achieves an ABEP of $10^{-14}$ at 14.2 dB and 8 dB for hard and soft decoded systems, respectively.

### 4.6 Summary

The main target of this WP was to propose novel robust and adaptive techniques, algorithms, strategies and architectures that significantly improve the performance over T2T and T2I communication systems. Towards satisfying this objective, in this section, several approaches were presented. More specifically, a new reconfigurable antenna pattern selection scheme is proposed and analyzed. In particular, assuming the generic scenario where correlation exists between the antenna patterns, a novel analytical framework is developed for describing important statistical characteristics of the proposed scheme. The new scheme, is applied in a V2V communication scenario, offering important advantages such as fewer physical antennas as well as fixed hardware complexity in terms of receiving chains. Moreover, a new threshold-based relay selection scheme has been adopted, which reduces the overhead processing that the conventional BRS approach induces to the system. The performance of the proposed scheme has been evaluated, using well-known performance metrics, namely OP and SEP. It is depicted that outdated CSI and interference decrease the system’s performance, while in many cases the new scheme outperforms well-known ones in terms of the performance versus complexity trade-off.

In addition, a reactive relay selection policy has been applied based on instantaneous CSI knowledge, which selects the relay requiring the minimum power expenditure to forward the source’s signal towards the destination. In this context, the network throughput and PER performance for an ITS-G5 network has been evaluated for various scattering conditions considering that all the vehicle nodes are equipped with single antennas or uniform linear multi-element antenna arrays. Moreover, a proof-of-concept indoor experiment is carried out where an eOLB is realized through pattern selection. Two prototype 3-printed monopole ESPAR antennas are used in this experiment and demonstrate their beam shaping capabilities using an SDR test-bed.

Through measurements, it is demonstrated that using an SNR-based pattern selection algorithm, significant SNR gains are achieved, that exceed 6.5 dB on average.

Finally, the SSK transmission is introduced in IVC operating over a multiple scattering channel model, which can be used to model various propagation scenarios. Analytical expressions for uncoded and coded SSK transmissions over this fading model were derived for the first time. It was shown that SSK outperforms other well-known MIMO schemes, offering also considerable reduced complexity.
5 Antenna Designs and Prototyping

The dynamic nature of the vehicular environment is mainly induced by the rich multipath scattering effects and the strong interferences that are caused by the numerous surrounding vehicles and buildings. In order to overcome the above challenges and maintain an acceptable quality of the V2V link, various diversity, MIMO, and beamforming techniques can be employed. However, these techniques require the use of multiple antennas which is opposed to the space and cost constraints in antenna-type selection and placement, dictated by automotive manufacturers. So far, the majority of the proposed antennas for ITS systems correspond to conventional models with only one radiating element (wire monopole, PIFA, printed monopole) [71], [72] that produce fixed beam patterns (mostly omnidirectional) and is quite difficult to support the aforementioned digital techniques. Only a few antenna examples can be found in the literature that offer an increased system capacity [73], [74], [75].

A reconfigurable-pattern antenna that produces multiple radiation patterns, could substitute a multiple antenna system and offer the desired degrees of freedom. A special category of reconfigurable antennas are the parasitic array antennas that can electronically adjust the loads of the parasitic elements, hence offering the capability of pattern reconfigurability. Moreover, they can be considered as excellent candidates for vehicular integration, due to the minimum space that occupy and their low fabrication expenses.

Electronically switched parasitic array radiator (ESPAR) antennas consist of a single active antenna element and several parasitic array elements in a specific geometrical arrangement (linear, circular etc.) [52], [54]. The main advantage of ESPAR antennas over phased array antennas is the absence of a feeding network and the close coexistence of several parasitic elements (in a small fraction of wavelength) that offer reduced complexity, smaller size and low manufacturing costs. Due to the close proximity between the active and the parasitic elements strong mutual coupling effects are formed, which are responsible for the powerful induced currents on the parasitics. Thus, the total radiation pattern of the array is generated from the direct currents owing to the active element and the indirect currents owing to the parasitics. The pattern reconfigurability of ESPAR antennas is then achieved by controlling the mutual coupling by tuning the impedances of the parasitic elements either with RF switches (PIN diodes) or variable capacitances (varactor diodes).

A large number of ESPAR antennas has been proposed in the literature using wire monopole antennas [54], [76], [77] and printed antennas [78], [79], [80], [81], [82]. Particularly interested on printed antenna designs due to the conformal low profile, these are divided into printed dipoles [78], printed monopoles [79], slots [80], printed patches [81], and PIFAs [82]. Dipole, patch and slot antennas have a theoretical electrical length of half wavelength, which is twice the size of monopole antennas. Moreover, in the majority of the literature available, only in [79] it is attempted to test the ESPAR antenna in frequencies above 5 GHz, where significant interaction with diodes' packaging occurs. However, there is still not enough evidence for the performance of ESPAR antennas at higher frequencies.

The investigation that is presented in this section involves the research that has been carried out for the selection of an appropriate antenna array for the ROADART hardware prototype development. Solutions with a significant number of array elements were excluded from our project planning for two basic reasons. Firstly, the level of complication of the final antenna design should be preserved as low as possible to avoid high manufacturing expenses, large antenna dimensions and excess RF cabling (e.g., for an antenna array that is embedded in the side mirror of the truck). Moreover, an antenna array with a dense circuitry consumes a greater amount of power. On the other hand, the antenna array configuration needs to demonstrate high reconfigurability and satisfy various MIMO and adaptive beamforming applications. Therefore, the main investigation is based on ESPAR antennas.
5.1 Antenna Positioning for T2T Communications

The first step of our research on the ROADART antennas concerned the antenna positioning and the selection of the most suitable antenna location on the truck [83]. This is very crucial for the performance of the T2T and T2I communication link due to the significant scattering that is induced by the truck cabin. In combination with the antenna positioning investigation, the concept of the ESPAR antenna is also studied and compared to a conventional monopole antenna. In order to evaluate the antenna performance in various T2T communication links over a realistic propagation environment, three appropriately designed antenna-truck model combinations were developed. The antenna - truck models are implemented in the CST 3D electro-magnetic simulator. The performance of the ITSs depends on the quality of the inter-vehicular link for V2V and V2I communication systems. V2V/V2I communications present specific challenges due to the fact that: i) both transmitter (Tx) and receiver (Rx) are located at the same height, ii) significant scattering is presented at both Tx and Rx since many metallic surfaces are in the close proximity of the antennas, iii) high speed movement is expected for the Tx and/or Rx vehicles, iv) low-power wideband signals are transmitted, v) the distance range between Tx and Rx may vary significantly [84]. The main requirement for ITS applications is the establishment of highly reliable links, since safety messages may be continuously transmitted by the vehicles. In these systems, the position of the antennas on the vehicle plays a crucial role, since the vehicle itself may destructively affect the transmission. Diversity techniques may improve the performance of the link as presented in [85] [86]. The problem of link reliability becomes even more crucial in T2T communications, since the truck trailer may cause even more significant connectivity issues.

This research study uses a simple truck model in order to investigate the effect of the antenna position in T2T links. Two antennas for each vehicle are used for ITS communications located either on the roof of the truck cabin or the truck side mirrors. Furthermore, the use of an ESPAR antenna with beam selection capabilities is proposed in order to improve link performance. The simulated results are very promising since significant improvement is achieved in both Rx SNR as well as the ABEP with the use of the proposed antenna configuration. In Section 5.1.1, the developed Antenna-Truck model is presented. In Section 5.1.2, evaluation of the investigated antenna configurations is performed using the IST-WINNER channel model.

5.1.1 The Antenna – Truck Model

In order to evaluate the antenna performance in various T2T communication links over a realistic propagation environment, three appropriately designed antenna-truck model combinations were developed. The antenna-truck models are implemented in the CST 3D electro-magnetic simulator [57]. Specifically, the three combinations involve two different types of antennas (single monopole, 3-element ESPAR) that are accurately modeled, investigated, and simulated along with a truck and trailer in two different positions (roof and side mirrors), with the intention to produce representative far-field patterns. Consequently, the simulated antenna patterns are imported into the IST-WINNER MATLAB model to evaluate a realistic propagation scenario.

5.1.1.1 Truck – Trailer Model

The truck-trailer design is mainly based on the “TGX-XLX International - National Long-haul Sleeper” model by MAN. Practically, it corresponds to a profoundly simplified version of the actual model that merely satisfies the basic dimensions and the main materials that are employed for the truck manufacture. In addition, two layers of asphalt and dry soil are placed underneath the truck-trailer in order to emulate the highway environment. Figure 64 illustrates the truck-trailer model as it was designed in CST. This high degree of simplification was proved to be necessary in order to preserve the memory usage (meshing) and the simulation duration in fairly low levels. However, special attention was given on the selection of the materials and their correspondence to the correct truck components, especially for the ones that are located in the antenna’s proximity. Table 5 includes a list of the materials used in the model along with the equivalent truck parts and dielectric
constant values. The basic dimensions of the truck-trailer design are given in Table 6 in mm.

![Figure 64: The truck–trailer design along with the highway part underneath it.](image)

### Table 5: Truck–Trailer Materials.

<table>
<thead>
<tr>
<th>Material</th>
<th>Truck components</th>
<th>$\varepsilon_r$</th>
</tr>
</thead>
<tbody>
<tr>
<td>PEC (metal / steel)</td>
<td>cabin body (chassis), roof, pillars, doors, hubcaps, trailer</td>
<td>infinite</td>
</tr>
<tr>
<td>plastic (polyethylene)</td>
<td>Bumpers, top front and side parts of cabin</td>
<td>2.25</td>
</tr>
<tr>
<td>glass (Pyrex)</td>
<td>Front and side windows</td>
<td>4.82</td>
</tr>
<tr>
<td>rubber</td>
<td>tires</td>
<td>3</td>
</tr>
<tr>
<td>asphalt</td>
<td>-</td>
<td>2.6</td>
</tr>
<tr>
<td>dry soil</td>
<td>-</td>
<td>2.27</td>
</tr>
</tbody>
</table>

### Table 6: Basic Truck Dimensions.

<table>
<thead>
<tr>
<th>Description</th>
<th>Dimension</th>
</tr>
</thead>
<tbody>
<tr>
<td>overall cabin height</td>
<td>3,895</td>
</tr>
<tr>
<td>front overhang</td>
<td>1,475</td>
</tr>
<tr>
<td>exterior cabin width</td>
<td>2,440</td>
</tr>
<tr>
<td>exterior cabin length</td>
<td>2,280</td>
</tr>
<tr>
<td>trailer length</td>
<td>12,000</td>
</tr>
<tr>
<td>trailer width</td>
<td>2,440</td>
</tr>
<tr>
<td>tire radius</td>
<td>537</td>
</tr>
<tr>
<td>tire width</td>
<td>318</td>
</tr>
</tbody>
</table>

#### 5.1.1.2 Antenna Designs

This section provides a brief description of the two antenna designs (single monopole, 3-monopole...
ESPAR) that are employed in the IST-WINNER implementation and are designed to operate at 5.9 GHz. Figure 65 demonstrates their layout. The first antenna design is a simple monopole radiator located above a circular ground plane of 50 mm diameter. The radius of the metal pin is equal to 0.7 mm and its length reaches up to 10.9 mm. The pin is fed by a co-axial line (waveguide port) composed by a Teflon inner layer and a PEC outer layer. As it is illustrated in Figure 66, the monopole demonstrates a satisfying return loss (~41 dB at 5.9 GHz) and a remarkably wide operating bandwidth over 28%. The top-view ($\theta = 90$) far-field pattern cut that is presented in Figure 67, confirms the monopole’s omni-directional behavior. The second design corresponds to a 3-element ESPAR antenna.

![Figure 65: Layout of (a) the single monopole and (b) the 3-monopole ESPAR antennas, demonstrating their simple geometry.](image)

![Figure 66: Return loss of the single monopole and the 3-element ESPAR.](image)

In the investigated case, the ESPAR is based on the preceding monopole radiator and is composed by three monopole radiators (one active and two parasitics) in a linear configuration. The parasitic elements are placed in a $\lambda/6$ distance away from the active radiator. The required objective of the antenna structure is to achieve a reconfigurable far-field pattern, i.e. the ability to change the direction of its radiation pattern from one side to the opposite with simple ON-OFF switching of the parasitic monopoles. Switching so far is realized by short/open circuit. Nevertheless, future work will involve the use of PIN diodes and varactors. The dimensions of the active monopole are slightly re-adjusted in order to achieve a satisfactory impedance matching and obtain significantly
low return loss. The dimensions of the parasitics are designed fairly smaller (shorter and thinner metal pin) than the active monopole with the intention to enhance the directivity of the array. Table 7 presents the dimensions of the 3-element ESPAR geometry.

As presented in Figure 66, the operating -10 dB bandwidth of the ESPAR is considerably smaller than the monopole’s, but it still remains in satisfying levels (400 MHz, almost 7%). The ESPARs radiation pattern, illustrated in Figure 67, demonstrates the reconfigurability aspect of the antenna and indicates a considerable 2.9 dB directivity increase in the desired direction.

![Far-field radiation pattern](image)

Figure 67: The far-field radiation patterns of the single monopole antenna (red) and the 3-monopole ESPAR (blue) at the θ = 90° plane (top-view).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>L_m</td>
<td>length of the active monopole</td>
<td>10.85</td>
</tr>
<tr>
<td>radius</td>
<td>radius of the active monopole</td>
<td>1.16</td>
</tr>
<tr>
<td>diff</td>
<td>difference between active and parasitic lengths</td>
<td>1.5</td>
</tr>
<tr>
<td>dis</td>
<td>distance between active and parasitic</td>
<td>λ/6</td>
</tr>
<tr>
<td>gap</td>
<td>gap between parasitic and ground plane</td>
<td>0.5</td>
</tr>
<tr>
<td>rad</td>
<td>radius of the parasitics</td>
<td>0.7</td>
</tr>
</tbody>
</table>

5.1.1.3 Antenna-Truck Configurations

In order to investigate the issue of a suitable antenna position selection, two main location candidates, the roof of the truck cabin and the side mirrors, are selected and tested. The IST-WINNER channel model evaluation of the next section is carried out for the following three antenna-truck configurations:

- Two monopole radiators on the roof of the cabin, located in the middle point of the cabin
length, separated by a 600 mm distance.

- Two monopole radiators at the position of the two side mirrors, 3,000 mm above the asphalt level and 400 mm beyond the edge of the truck cabin.
- Two 3-monopole ESPAR radiators at the position of the two side mirrors.

Figure 68 and Figure 69 include the three 3D directivity patterns for the three cases respectively. It is quite obvious that for the roof configuration, the obtained irregular pattern is mainly caused by the shielding of the metal trailer. Specifically, a significant part of the radiation is reflected on the trailer resulting in a critical restriction of the radiation, especially for the angles away from the azimuth plane. However, when the two monopole radiators are placed at the side mirror positions, the scatterers that are located in the really close proximity of the two radiators are reduced and the effect of the truck trailer system is reasonably decreased. Therefore, the directivity pattern seems to obtain a more isotropic characteristic. For the third case (ESPAR antennas), a significant part of the radiation is directed towards the corresponding desired direction (back direction for the leading truck, front direction for the following truck). This fact, combined with the LoS conditions, results in a substantial performance improvement.

5.1.2 Radio Channel Modeling

In order to evaluate the three presented configurations in section 5.1.1.3, simulation analysis was performed. Two trucks were assumed, equipped with an ITS-G5 [87] system. ITS-G5 physical layer is based on IEEE 802.11p [88] at 5.9 GHz carrier frequency with 10 MHz bandwidth. The two trucks are assumed to move on a highway crossing a rural area and both trucks move in the right lane of the highway. Moreover, they are moving in a straight line formation (no turns). The speed of the trucks was 70 km/h (approximately 20 m/sec). The radio channel model that was used for evaluation was the IST-WINNER2 channel model [89]. C1 channel models were considered, which correspond to a rural macro-cell scenario. The patterns that were extracted in section 5.1.1.3 were imported in the IST-WINNER2 channel using the modification of [90]:

\[
[H_b(t;\tau)]_{\omega_1} = \sqrt{\frac{p(k)}{(K_{LoS} + 1)N_k}} e^{j\sigma_{z,a} e^{2\pi f (z_{k,a} - z_{v,a}) - \nu_{v,a} \cos(z_{k,a} - z_{v,a})}}
\]

(102)

Figure 68: The far-field directivity patterns for the first antenna – truck configuration (monopoles on the roof).
Figure 69: The far-field directivity patterns for the remaining two antenna – truck configurations (monopoles at the side mirrors – top, ESPARs at the side mirrors – bottom).

More specifically, $H_b$ is a diagonal $(1 + \sum N_k) \times (1 + \sum N_k)$ matrix containing the complex path gains of all scatterers, where $N_k$ is the number of rays for the $k$th cluster of scatterers, $v_{Tx}$, $v_{Rx}$ the Tx and Rx velocities, $\zeta_{k,n}^{AA}$, $\zeta_{k,n}^{AD}$ the angles of arrival and departure for the $n$th ray of the $k$th cluster, $p(k)$ the path gains for each cluster ($p(0) = K_{LOS}$) and $\zeta_{Rx/Tx}$ the orientation of the Rx/Tx arrays. All angles are given with reference to the Global coordinate system of the WINNER model. After the calculation of $H_b$, the channel matrix is given by:

$$H = P_{rx}^H H_b(t, \tau) P_{tx}$$

(103)

where $P$'s are $(1 + \sum N_k) \times 2$ matrices containing the patterns for each of the two Tx and Rx antenna elements towards the location of the scatterers. It is noted that the patterns in $P$ should include phase differences due to the positioning of the antennas in the array (i.e. the steering matrix is applied to each element pattern).

Nevertheless, extra modifications of the channel model were needed in order to support mobility for both trucks (Tx and Rx) and aligned movement in a straight line with the antennas properly placed and rotated, at the correct height (Section 5.1.1.1). Since the objective of this investigation is to evaluate the antenna configurations on the truck, no elaborate Tx or Rx diversity scheme was used.
Thus, it is assumed that the ITS signal is transmitted simultaneously by both antennas of the Tx and it is combined with simple addition at the Rx antennas.

Three basic scenarios were investigated and simulated:

- The two antennas are mounted on the roof of both trucks.
- The two antennas are mounted on the side mirrors of the trucks
- Two 3-element ESPAR antennas are mounted on the side mirrors of the trucks. The preceding truck focuses the antenna beam to the back, while the following truck focuses the antenna beam to the front.

10,000 channels were simulated. The same set of channels was investigated for all three configurations. During each simulated session, 100 ITS-G5 bursts of 1kByte packets were transmitted every 10 msec, in order to take into account the Doppler effects due to the high-speed mobility of the Tx and Rx. The transmitted power was 33 dBm (1W per element), which corresponds to the maximum allowed EIRP for an ITS system that uses the ITS-G5 Control Channel (safety messages). The noise level at the Rx was assumed -118 dBm. In Figure 70(a), the average Rx SNR vs. Tx-Rx separation distance is presented. It is clear that the SNR at the Rx significantly increases with the placement of the antennas at the side-mirrors. Moreover, additional SNR improvement is noticed with the use of a simple ESPAR antenna with ON-OFF elements. The SNR gain due to the use of ESPARs on the side mirrors reaches 10 dB for distances > 250 m compared with the performance of the system with roof mounted antennas.

The ABEP is also extracted for the three configurations. The error rate is calculated assuming the basic ITS-G5 transmission mode, with BPSK modulation and 1/2 convolutional coding. The ABEP is calculated over the wideband system, i.e. over the 48 subcarriers that are used for data transmission in the OFDM IEEE 802.11p. In Figure 70(b), the ABEP vs. Tx-Rx separation distance is plotted. It is noted that due to the frequency selectivity of the channel, the SNR per subcarrier may vary significantly relative to the average SNR.

![Figure 70](image_url)

Figure 70: (a) Average SNR vs. distance for T2T C2 WINNER radio channels. (b) ABEP vs. distance for T2T C2 WINNER radio channels.

The improvement in the operation of the system when the antennas are placed on the side mirrors is impressive. In addition, the use of an ESPAR system will furthermore reduce the achieved ABEP. According to the simulation results, the ESPAR system performs better on average for Tx-Rx distance 120 m than the roof-mounted antenna system for Tx-Rx distance of 50 m. However, it should be noted that the IST-WINNER2 model supports 2D propagation, therefore the effect of the 3D pattern is not fully taken into account. It is expected, due to the effect of the trailer on the pattern...
in the first scenario (Figure 68), that the performance improvement from the placement of the antennas on the side mirrors will be further increased.

5.2 The 3-Printed Monopole ESPAR Antenna

The first type of antenna that was analyzed and investigated was a printed monopole ESPAR antenna with three elements. The design demonstrates the beam-switching capabilities of ESPAR antennas at 5.9 GHz, occupying minimum space while maximizing mutual coupling mechanism [91]. The remainder of this section is organized as follows: in Section 5.2.1, the geometrical characteristics of the planar ESPAR antenna are described in detail, together with the EM simulation results for the two switching states. A parametric investigation is carried out to optimize the final antenna dimensions. In Section 5.2.2, the fabricated prototype is presented. Moreover, an experimental validation is performed to prove the close agreement of the measured return loss and radiation patterns with the EM simulation results.

5.2.1 Antenna Configuration

The ESPAR antenna that is presented in this paper is designed and fabricated in a planar structure. Specifically, it employs three printed monopole antennas as the ESPAR elements (one active and two parasitic). Figure 71 depicts the exact layout of the 3-printed monopole ESPAR antenna model with an overall size of 32 x 30 x 0.8 mm. The substrate that is employed is a RO4725-JXR dielectric panel manufactured by Rogers Corporation. It has a thickness of \( h = 0.78 \) mm and it exhibits an electric permittivity of \( \varepsilon_r = 2.55 \) [92]. The antenna substrate is chosen based on the best feasible “emulation” of free space that provides, at the same time, adequate mechanical support to the antenna. Thus, a relatively thin rigid substrate panel is utilized, with an \( \varepsilon_r \) value as close as possible to 1.

![Image of the antenna setup](image)

Figure 71: The detailed design of the proposed 3-element printed monopole ESPAR antenna in CST.

5.2.1.1 Description of the proposed ESPAR antenna design

The CST 3D electromagnetic solver was employed for the ESPAR antenna design [57]. For the design of the printed monopole element, a microstrip quarter-wavelength (\( \lambda/4 \)) transformer is employed in combination with an end-launch (edge-fed) SMA connector in order to provide a 50 \( \Omega \) impedance and feed appropriately the active element of the array. Specifically, the \( \lambda/4 \) transformer is designed just before (adjacent) to the printed monopole and transforms the theoretical 37 \( \Omega \) monopole's input impedance to the 50 \( \Omega \) impedance of the connector. Figure 72 demonstrates the “top-view” layout of the ESPAR design where the substrate panel is transparent in order to reveal
the metallization at the back. As it is shown in this illustration, there is a strip of copper layer at the back side of the substrate panel that serves as the ground plane of the \( \lambda/4 \) transformation microstrip structure.

![Illustration of the ESPAR antenna design](image)

Figure 72: The “top-view” design of the ESPAR antenna along with the basic design parameters.

The close distance (\( \lambda/5 \)) between each parasitic element and the active printed monopole causes strong mutual coupling effects and provides a reconfigurable radiation pattern to the printed ESPAR antenna according to which parasitic monopole is connected to the ground plane. The grounding of the parasitic printed monopole is achieved using a plated via (through hole) that is allocated inside a metal pad next to each parasitic element. The pattern reconfigurability is attained by employing two PIN diodes (Skyworks SMP-1320 / 040LF package) \[93]\) that are placed between the parasitics and the metal pads and act as two-states (ON-OFF) electronic switches. In principle, PIN diodes achieve faster switching speeds than mechanical switches (typically less than 100 ns) and can be placed in packaging occupying a fraction of the size of mechanical RF switches. When the PIN diode is ON, the equivalent parasitic monopole is grounded (connected to the ground plane stripe at the back), forming a large L-shaped reflector that reflects the array beam towards the opposite side. In this way, three different radiation patterns can be obtained based on the three different PIN diodes' combinations (both OFF, ON-OFF and OFF-ON). The fourth state with both the PIN diodes ON and thus, both parasitic monopoles acting as reflectors, provides a radiation pattern that is extremely similar to the both-OFF state, and therefore, it is not analyzed in depth.

In addition, in Figure 72, the DC bias network connected to each parasitic monopole is clearly illustrated, which is employed in order to supply the appropriate voltage to the PIN diode and alter its state (bias the PIN diode). The DC bias network includes a narrow printed DC bias line, a metal pad, a DC header pin (which is soldered to the metal pad) and an inductor. The 18 nH inductor (Coilcraft 0302CS-18NXJLU) operates as an RF choke and prevents the AC current to pass at the DC bias network and disturb the radiation pattern of the ESPAR antenna. The two different states of the PIN diode (forward/ON-reverse/OFF) were simulated using the two equivalent RLC circuit configurations that are illustrated in Figure 44 (a) and (b) respectively. The values of the circuit elements are provided by the data sheet of the SMP1320 PIN diode and are included in Table 8.

<table>
<thead>
<tr>
<th>Element</th>
<th>( R_s )</th>
<th>( L )</th>
<th>( C_T )</th>
<th>( R_L )</th>
</tr>
</thead>
</table>

Table 8: The Equivalent RLC Circuit Values for the Two Different States of the PIN Diode.
5.2.1.2 Parametric Study of Parasitic Elements

In order to optimize the physical parameters of the proposed ESPAR antenna, a parametric investigation was carried out at the parasitic elements. The parametrized ESPAR antenna model consisted of one parasitic element in the ON state and the other parasitic element at the OFF state in order to optimize the directive antenna performance specifically. Another purpose of this investigation was to reduce the size of the ESPAR antenna without sacrificing its performance. Three main design parameters contribute to the antenna performance, specifically the distance between the active and the parasitic elements \( (\text{dis}) \), the difference in length between the active and parasitic elements \( (\text{diff}) \) and the width of the parasitic elements \( (W_p) \).

At first, the parameter \( \text{dis} \) is varied between \( \lambda/4 \) and \( \lambda/6 \) and its effect to the return loss \( (S_{11}) \) is shown in Figure 73. It is seen that the resonance frequency is shifted upwards from 5.6 GHz to 6.2 GHz and achieves a resonance at 5.9 GHz when \( \text{dis} = \lambda/5 \). In terms of the radiation pattern of the ESPAR antenna, no significant effect is noticed for different values of the \( \text{dis} \) parameter. Furthermore, the effect of parameter \( \text{diff} \) that determines the length of the parasitic elements is shown in Figure 75 and Figure 75. It is varied between \( \pm 2 \text{mm} \) from the length of the active element. The return loss of the antenna at 5.9 GHz (Figure 75) is minimised for values close to 0 mm and reaches its minimum value for \( \text{diff} = -1 \text{ mm} \). As far as the radiation performance is concerned (the radiation pattern at the elevation plane is shown in Figure 75), reaches the maximum gain at 6.05 dB. The final parameter that has been considered in this parametric study is parameter \( W_p \), which shows a decrease in resonance frequency while it is varied from 0.4 to 1.4 mm (Figure 76).

![Graph](image-url)

Figure 73: The effect of parameter “\( \text{dis} \)” to the return loss \( (S_{11}) \) of the printed monopole ESPAR antenna.

While it has been shown that all these three parameters can contribute to the resonance frequency response of the ESPAR antenna, only the \( \text{diff} \) parameter affects the radiation performance. Therefore, the parameters \( \text{dis} \) and \( W_p \) are not considered that crucial for the design procedure since any effect that they induce in the antenna return loss can be easily compensated by adjusting the length of the active monopole \( L_m \). After an extensive parametric investigation, the basic geometrical specifications and final dimensions of the ESPAR antenna are presented in Table 9.
5.2.1.3 Simulation Results

The simulated return loss ($S_{11}$) of the optimized printed monopole ESPAR for two different states (both OFF, ON-OFF) is illustrated in Figure 77. It is observed that the simulated 3-element ESPAR with one parasitic element grounded (ON-OFF state) demonstrates a considerably low return loss (-31.0 dB) relatively close to the desired frequency (5.89 GHz) and a satisfactory -10 dB impedance.
bandwidth of 8.6%. As it is expected, the state alteration of the ESPAR antenna (both OFF) generates a small difference in the return loss response of the antenna that is justified by the effect of the parasitics loading on the antenna's input impedance. In this case, the resonating frequency is slightly higher (6.1 GHz) but still exhibits a decent return loss of -14.7 dB at 5.9 GHz and an 11.9% impedance bandwidth.

Figure 78 demonstrates the equivalent radiation patterns of the ESPAR antenna for the two aforementioned states, plotted on the H plane (left) and on the E plane (right) at 5.9 GHz in polar format. When both PIN diodes are OFF, we can notice that the radiation pattern of the array is completely symmetrical. However, the pattern does not exhibit a perfectly omni-directional behaviour (“squeezed shape”), which can be easily predicted if we take into account the existence of

Table 9: Final Dimensions of the 3-Printed Monopole ESPAR Antenna.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>h</td>
<td>thickness of the RO 4725-JXR substrate layer</td>
<td>0.78</td>
</tr>
<tr>
<td>t</td>
<td>thickness of the copper and the ground plane layers</td>
<td>0.035</td>
</tr>
<tr>
<td>L_m</td>
<td>length of the active printed monopole</td>
<td>7.4</td>
</tr>
<tr>
<td>W_m</td>
<td>width of the active printed monopole</td>
<td>1.65</td>
</tr>
<tr>
<td>L_p</td>
<td>length of the parasitic monopole</td>
<td>8.8</td>
</tr>
<tr>
<td>W_p</td>
<td>width of the parasitic monopole</td>
<td>0.7</td>
</tr>
<tr>
<td>L_qw</td>
<td>length of the λ/4 transformer</td>
<td>9.0</td>
</tr>
<tr>
<td>W_qw</td>
<td>width of the λ/4 transformer</td>
<td>3.1</td>
</tr>
<tr>
<td>dis</td>
<td>distance between the active and the parasitic elements</td>
<td>λ/5</td>
</tr>
<tr>
<td>gap</td>
<td>gap between the parasitic monopole and the grounding metal pad</td>
<td>1.0</td>
</tr>
<tr>
<td>Parameter</td>
<td>Description</td>
<td>Value (mm)</td>
</tr>
<tr>
<td>-----------</td>
<td>-------------</td>
<td>------------</td>
</tr>
<tr>
<td>h</td>
<td>thickness of the RO 4725-JXR substrate layer</td>
<td>0.78</td>
</tr>
<tr>
<td>r_via</td>
<td>radius of the platted via (through hole)</td>
<td>0.3</td>
</tr>
</tbody>
</table>

Figure 77: The simulated return loss ($S_{11}$) versus frequency for the two states (ON-OFF and both OFF) of the printed monopole ESPAR antenna.

Figure 78: The simulated gain radiation patterns of the printed monopole ESPAR antenna on the (a) H plane and (b) E plane at 5.9 GHz, for the OFF-OFF (blue) and the ON-OFF (red) states.

the substrate panel and the parasitic elements (even if they are not connected). When one parasitic monopole is grounded (ON-OFF state), the radiation reflects on the L-shaped reflector and the ESPAR beam points out towards the opposite direction, resulting in a significant increase on the
antenna's gain (almost 3 dB).

5.2.2 Experimental Verification

5.2.2.1 Fabrication

The fabrication of the array was carried out using etching technique. The assembly and the soldering (by hand) of the components (PIN diodes, inductors, SMA connector, DC header pins) was performed in IMST GmbH. Figure 79 includes two pictures of the manufactured printed monopole 3-element ESPAR antenna after its fabrication and assembly. The photographs clearly depict the relatively small dimensions of the ESPAR antenna and the ability to obtain a reconfigurable radiation pattern in such a compact antenna structure. Additionally, a DC bias circuit board was made in order to supply the suitable DC voltage to the PIN diodes. A 1.5 V battery and two 100 Ω potentiometers were employed for adjusting the voltage to the correct value (0.85 V) that corresponds to the equivalent circuit values (Table 8) that were used in the simulation.

![Figure 79: (a) Front-side view and (b) back-side view of the 3-element printed monopole ESPAR after its fabrication and assembly.](image)

5.2.2.2 Return Loss Characteristics

Figure 80 demonstrates the simulated reflection coefficients of the OFF-OFF and ON-OFF/OFF-ON states compared with the corresponding measured results, exhibiting a clear agreement. The small deviations can be attributed to manufacture and soldering inaccuracies and should be considered as acceptable for prototype development. As it is expected, the resonant frequency is slightly shifted when a DC biasing is applied. For the measured OFF-OFF state, the resonant frequency is observed at 6.1 GHz, achieving a satisfying impedance bandwidth of 550 MHz (9.3 %). For the measured ON-OFF/OFF-ON states, where a DC biasing is applied in one of the parasitic elements, the resonant frequency is obtained at 5.9 GHz and 5.92 GHz demonstrating a satisfying design symmetry. The impedance bandwidth slightly decreases to around 440 MHz (7.5 %). The level of $S_{11}$ magnitude in these four cases drops below -25 dB, which justifies an excellent impedance matching. When the DC bias is applied to the parasitic elements, $S_{11}$ improves to -34 dB (at the desirable frequency of 5.9 GHz), since the antenna has been designed to achieve the optimal performance at the ON-OFF / OFF-ON states.

5.2.2.3 Radiation Characteristics

The radiation pattern measurements of the 3-element printed monopole ESPAR antenna have been carried out in the indoor far-field antenna test range of IMST GmbH. This is a fully automated
antenna measurement facility (700 MHz-13 GHz) that is accommodated in a completely shielded anechoic chamber (5 x 3 x 3m) and employs a vector network analyzer (VNA) and a spherical scanning positioner in order to acquire the far-field data around the AUT. In order to derive the absolute ESPAR

![Graph](image_url)

**Figure 80:** Comparison of the measured (straight lines) and simulated (dotted lines) reflection coefficient $S_{11}$ for different switching states.

![Graph](image_url)

**Figure 81:** Measured radiation characteristics (absolute gain) at 5.9 GHz: (a) H-plane radiation, (b) E-plane radiation for the OFF-OFF (blue) and the ON-OFF (red) states. Simulated results are plotted for comparison with dotted lines.

Antenna gain, the gain-transfer method is employed using the calibrated gain of a standard gain
horn (SH800 model manufactured by MVG). Figure 81 shows the measured far-field radiation patterns produced by the ESPAR antenna on the two orthogonal cuts (H-plane and E-plane) for the OFF-OFF (blue) and the ON-OFF (red) operating modes. As it can be noticed in Figure 81(a), when both PIN diodes are OFF, the ESPAR antenna demonstrates a perfectly symmetrical behavior (non-perfectly omni-directional though), achieving equal directivity at the two edges of the antenna structure (where the two parasitic elements are allocated). When a DC bias is applied in the PIN diode connected to the parasitic element (ON-OFF state), the pattern becomes directional, with maximum gain appearing at the opposite side from the biased parasitic element. A maximum gain of 5.3 dBi is achieved at $\theta = -85^\circ$ (5 tilted upwards). A 2.5 – 3 dB gain enhancement from the OFF-OFF state to the ON-OFF state is achieved at the H-plane. Figure 81 compares the measured and the simulated radiation patterns that are obtained at 5.9 GHz for both basic operating modes, demonstrating a good agreement. The insignificant discrepancies between the measured and the simulated patterns can be attributed to the interactions of the ESPAR antenna with the RF cables, DC cables and the plastic mounting structure during the measurement acquisition.

The design of a novel 3-printed monopole ESPAR antenna (operating at 5.9 GHz) with a reconfigurable radiation pattern has been proposed for T2T communications in this section. A good agreement between the simulated and the experimental results is indicated. A satisfactory impedance matching performance for the three main switching antenna states is observed. A considerable 3 dB gain increase is achieved when the ESPAR is switched to its directive mode (ON-OFF). However, the proposed antenna design presents three drawbacks:

- Reflection coefficients ($S_{11}$) for the two antenna states (ON-OFF and OFF-OFF) have a small deviation due to the different parasitics loading and its effect on the antenna’s input impedance.
- Non-perfectly (quasi) omnidirectional radiation pattern on H-plane for the OFF-OFF state.
- Radiation is tilted upwards due to the ground plane presence (E-plane).

5.3 The Printed Mono/Dipole 3-Element ESPAR Design

In order to implement a planar 3-element ESPAR antenna that demonstrates pattern reconfigurability and provides directive patterns with their maximum on the H-plane, a wideband “taper-shaped” monopole is also designed and simulated. The monopole is accompanied by a mirrored ground plane, which causes the design to appear like a dipole (hence, the “mono/dipole” name) and two parasitic dipoles that are ground detached and thus, the directive beams are not tilted upwards. Moreover, the investigation of this wideband mono/dipole, was carried out, and simulated to verify the potential of a planar ESPAR antenna maintaining the reflection coefficient level to acceptable values. The mono/dipole design is based on a mirrored trapezoidal monopole arm.

5.3.1 Primary Designs and Simulations

The first simulated design is depicted in Figure 82 and Figure 83, containing one trapezoidal shaped dipole with a coplanar waveguide (CPW) line as feeding line. The design contains two parasitic
Elements close to the active element with their centers at the same height. As the radiation pattern (that the active element is expected to exhibit) is omnidirectional, the parasitic elements stand vertically to the radiation lobe of the antenna.

The feeding method of the active element involves the two gaps opposing the feeding line of the CPW. Thus, the metallization of the lower trapezoidal arm is sliced into two pieces that is expected to operate as the second arm of the dipole. This way, although the simulation results show that resonance occur in the desired frequency, the impedance of the two dipole arms is not the same. The antenna is expected to be unbalanced due to two basic reasons:

- The phases that enter each arm differ, so the final behavior of the antenna should be predicted via simulations.
- The resistance of the upper arm will differ from the resistance of the metallic sheet of the lower arm.

The first problem can be solved in the simulation process as the results of the phase difference are obvious in the reflection coefficient of the antenna. The second problem can be identified only via the measurement of the manufactured antenna and can be solved with the use of a balun design before
The antenna feeding point. The parasitic elements were simulated as open lines without introducing any circuit element on their junction, while an SMA connector was simulated with the overall structure in order to estimate and eliminate any unwanted effect of its placement to the antenna substrate. The antenna was tuned to resonate in the desired frequency of 5.9 GHz as it is shown in Figure 84. The -10dB bandwidth of the antenna computed as 3.05GHz starting from 4.35GHz. The presence of the SMA connector in the simulation environment causes several dimensional changes, compared to a conventional dipole dimensions, as the electrical length of the overall antenna is altered.

As long as the parasitic elements are not connected (open circuited) the antenna exhibits a quasi-omni-directional radiation pattern, plotted in Figure 85. The presence of the parasitic elements that are not short-circuited does not affect the radiation pattern, but only the impedance of the active element, which is matched through the simulation. The state with one short-circuited parasitic element is also investigated to verify the reconfigurability of the antenna. To simulate the antenna behavior, the upper and lower parts of the left parasitic element are connected with a bond wire (galvanic contact). That way, the electrical length of the left parasitic element is increased, by adding both arms’ lengths, forming a director on the antenna substrate. The reflection coefficient of the antenna is changed (Figure 86) while keeping a usable level, maintaining a low value of -23 dB at the frequency of 5.9 GHz. The bandwidth of the antenna with respect to the -10 dB level is also maintained wide (from 4.35 GHz to 7.3 GHz).

The radiation pattern of the ESPAR antenna with one short-circuited parasitic element is illustrated in Figure 87. The omni-directional pattern is changed to a directive radiation pattern, with its main lobe directed towards the side of the short-circuited parasitic element. Without connecting any parasitics, the ESPAR antenna exhibits a directivity of 1.98 dBi at the sides of the antenna. When one parasitic element in ON, the directivity is increased to 2.7 dBi. On the other side of the antenna (not connected parasitic element), the radiation pattern decreases to values close to 1.1 dBi.
Figure 85: Radiation pattern of the wideband mono/dipole ESPAR antenna at 5.9 GHz with parasitic elements on the omni-directional (OFF-OFF) state.

Figure 86: Reflection coefficient of the wideband mono/dipole ESPAR antenna with one parasitic element on the directive state (ON-OFF).
Consequently, the combination with both parasitic elements connected is tested. However, the results are not that satisfactory. The reflection coefficient (Figure 88) seems to be higher than the previous ESPAR states, as the antenna resonance is detuned from the desired frequency of 5.9 GHz. Moreover, the usable bandwidth of the antenna drops as the impedance of the active element is mismatched with respect to the 50 Ω impedance of the simulated feeding port.

The radiation pattern of the ON-ON state (Figure 89) seems to be similar to the OFF-OFF state of the ESPAR antenna (quasi omni-directional pattern). On this state, no advantage is found in respect to the pattern reconfigurability, as the overall performance of the antenna is obviously lower due to the antenna’s impedance mismatch and the higher reflection coefficient.
5.3.2 Investigation with Circuit Elements and Bias Lines

Since the design with the wideband mono/dipole has given satisfying results regarding the pattern reconfigurability, the switching mechanism along with the network, is introduced in this section. The selection of the switching mechanism that is needed to control the parasitic elements and its integration on the antenna design is considered necessary in order to produce the most reliable results for the final design.

The switching mechanism that is chosen for the transition from the open to the short circuited parasitic element is the use of PIN diodes. Placing the diodes on the middle of the parasitic element and devising the way to bias them, can provide the desired switching scenario for the ESPAR antenna. As the main structure of the antenna is maintained, the dimensions of the antenna are slightly adjusted to preserve the resonance of the antenna at 5.9 GHz. The biasing network is designed around the antenna (Figure 90) so that its effect on the radiation characteristics of the final radiator is negligible. The dominant field on the radiation of the dipole is the E-field, thus, the biasing network that is designed to supply DC voltage to the diodes, is placed extending vertically to the mono/dipole in order to decrease any unwanted resonance on its metallization.

![Gain (dBi)](image)

**Figure 89:** Radiation pattern of the wideband mono/dipole ESPAR antenna with both parasitic elements short-circuited (ON-ON state).
Figure 90: Respective view of the wideband mono/dipole ESPAR antenna with the designed biasing network.

Furthermore, four inductors are placed as RF chokes, not only to prevent the flow of the RF signal towards the DC voltage supply, but also to uncouple the bias network from the radiator. The selected diodes are the SMP1320-SC76 from Skyworks. The diodes are placed on special pads designed on the edges of each parasitic element for manufacturing ease and soldering simplification. The RF chokes are placed on the same pads in order to cut the RF signal propagation to the biasing lines. Through a series of simulations, it is demonstrated that the extension of the biasing lines to the edges of the substrate does not affect the final ESPAR antenna characteristics. Extra pads are placed on the ends of the biasing lines in order to solder DC pins or directly DC cable towards a DC source. 

The main body of all the circuit elements (components: PIN diodes, inductors) is simulated as silicon with the specific dimensions and topology given in the datasheets. The volume of the silicon that was added in the simulation model has impact on the resonance sharpness of the ESPAR antenna. This is due to the high relative permittivity of the material that is positioned close to the active element.

5.3.3 The Final Design of the Mono/Dipole 3-Element ESPAR Antenna

However, the final design of the planar mono/dipole 3-element ESPAR antenna exhibits two main alterations. Firstly, the wideband “taper-shaped” monopole is not fed by CPW but by a microstrip line. Therefore, the mirrored “taper-shaped” ground plane is now accommodated on the back side of the ESPAR panel. Secondly, the parasitic printed dipoles are bended in order to enhance the directivity at the ON states of the ESPAR antenna. Both design alterations are clearly illustrated in Figure 91. The active element is optimized for wideband application, so as the operation of the parasitic elements won’t increase the reflection coefficient of the antenna to undesired levels. Pattern diversity is achieved through switchable PIN diodes. In opposition to the printed monopole ESPAR, the mono/dipole ESPAR operates in directive mode, which means that when a parasitic dipole is ON, it acts as a director and not as a reflector (beam is directed towards the parasitic). Furthermore, the operation of the parasitic elements is not that effective to the reflection coefficient of the antenna (S_{11} is not shifted to undesired frequencies or dramatically increased between different ESPAR states). Thus, the mono/dipole ESPAR antenna design expands the pattern reconfigurability that is achieved by the printed monopole design and offers 4 different operating states (OFF-OFF, ON-OFF, OFF-ON, ON-ON). This advantage is depicted in Figure 92.
of the mono/dipole ESPAR antenna were fabricated by IMST GmbH (see Figure 93) to verify the antenna’s simulated performance, in two identical versions which differ on the biasing network design (with and without DC pins).

Figure 91: The final designs of the wideband mono/dipole ESPAR antenna (a) with DC pins and (b) without DC pins.

Figure 92: Simulated far-field directivity patterns for three different operating states. The OFF-ON state is excluded since it is identical to the ON-OFF (due to the perfect design symmetry).
During the first round of the return loss antenna measurements, a few difficulties emerged, as the reflection coefficient of the antenna was not the expected (see Figure 94). The measured results seemed sensitive to the feeding cable positioning and holding, due to back traveling currents through the cable outer conductor. In order to balance the current magnitudes in the two “dipole” arms, a second approach is followed, utilizing a balun structure for the frequency of 5.9 GHz. The improvised “bazooka” balun is implemented by a λ/4 cylindrical conductor (thick paper and copper foil with conducting glue) that is in galvanic contact with the feeding cable’s outer conductor (Figure 95). As the balun exhibits frequency selective behavior, the back-traveling waves for the frequencies near the 5.9 GHz are canceled, while resonating at the surface of the balun. Therefore, the measured reflection coefficient of the antenna is considerably fostered demonstrating an excellent impedance bandwidth and a satisfactory return loss for three antenna operating modes (see Figure 96).

Figure 93: The fabricated prototype of the wideband mono/dipole ESPAR antenna: (a) front-side view and (b) rear-side view.

Figure 94: Measured reflection coefficients of the mono-dipole ESPAR antenna without balun.
Figure 95: The mono-dipole ESPAR antenna with the bazooka balun attached during measurement.

Figure 96: (a) Simulated reflection coefficients ($S_{11}$) versus frequency for three different states. The OFF-ON state is excluded since it is identical to the ON-OFF (due to the perfect design symmetry). (b) Measured reflection coefficients ($S_{11}$) versus frequency for four different states.

5.4 The Printed Mono/Dipole 5-Element ESPAR Design

The objective for the diversity and possibly the beam-space MIMO scenario, regarding the ESPAR antenna design, is the extension of pattern reconfigurability and the achievement of more different radiation patterns from one ESPAR antenna. Using the same active element (mono/dipole antenna) and maintaining the single RF chain, a five element ESPAR design is investigated. Consisting of one active monopole and four parasitic dipoles displaced from the latter’s substrate panel, the antenna is packed into three stacked boards (see Figure 97). The reconfigurability is obtained by using PIN diodes and switching the parasitic elements ON and OFF. Maintaining the reflection coefficient at a satisfying level (for at least 6 different usable antenna states, see Figure 98) the antenna provides numerous radiation patterns, rotating the main lobe direction to four different directions, switching one parasitic element at a time. At the “all OFF” state, antenna radiates in a quasi omni-directional pattern, while the switching of one parasitic directs the radiation towards the switched element. Figure 99 includes the simulated radiation patterns for the single element switching, keeping as reference the omni-directional pattern of the “all OFF” state.
Figure 97: Diagonal-view and top-view layouts of the 5-element mono/dipole ESPAR simulation model in CST. The substrate panels were made transparent in order to make visible the ground plane of the mono/dipole and the “back” parasitic dipoles.

Figure 98: The simulated return loss ($S_{11}$) results for 6 different operating states.

Figure 99: Simulated radiation patterns of the printed mono/dipole 5-element ESPAR antenna with one element switched ON.
The extension of the antenna operation with two parasitic elements switched ON at the same time, is also studied, providing four extra radiation patterns, when switching elements in pairs. This way the antenna provides the additional patterns of two rear, front, left and right switched ON elements, directing the main lobe towards the rear, front, left and right side of the ESPAR antenna respectively. The radiation patterns of these states are shown in Figure 100. Including the “all OFF” state of the ESPAR design, a reconfigurability of 9 different patterns can be realized. To achieve this scope though, a matching network is needed to improve the ESPAR’s reflection coefficient at certain states, such as the combination of the two ON rear parasitic elements, which offers a reflection coefficient value near the -10 dB level.

Figure 100: Simulated radiation patterns of the printed mono/dipole 5-element ESPAR antenna with two elements switched ON.

5.5 Investigation of ESPAR Antenna with Varactor Diodes

To achieve an increased number of operating states in ESPAR antennas, and therefore, improve their beamforming capability, varactor diodes are employed replacing PIN diodes that have been used so far. Varactor (or variable reactor) diodes are commonly used in ESPAR antennas to provide continuous frequency and/or pattern reconfigurability in comparison to PIN diodes that offer discrete states only [52], [94], [95]. When a reverse DC bias voltage is applied to a varactor, its capacitance is altered. Therefore, by integrating varactor diodes in the parasitic elements of ESPAR antennas, their characteristic (load) impedance ($Z_L$) can be controlled, altering the coupling mechanism between the active and the parasitic elements, and finally obtaining different radiation patterns.

A parametric study is carried out in CST EM simulator to investigate the effect of the variable varactor diode’s capacitance to the performance of a theoretical 3-element dipole ESPAR antenna. The theoretical ESPAR antenna, shown in Figure 101, consists of one active and two parasitic dipole elements spaced $\lambda/4$ apart. In each of the parasitic elements a varactor diode is connected. The diode’s capacitance is varied between 0.1 pF to 1 pF to examine the effect in the reflection coefficient ($S_{11}$) and radiation pattern for antenna operation at the 5.9 GHz ITS band. It is seen in Figure 102 that the $|S_{11}|$ reduces quickly to values below -10 dB for capacitance values of $C > 0.3$ pF. A continuous reduction trend is noticed that shows no saturation point. Respectively, the gain of the radiation pattern (Figure 103) degrades proportionally with the increased load capacitance. Hence, varactor diodes with even a small capacitance range can strongly affect the input impedance and the radiation gain of ESPAR antennas.
Figure 101: Investigation of a theoretical 3-element dipole ESPAR antenna at 5.9 GHz, loaded with varactor diodes in the parasitic elements.

Figure 102: $S_{11}$ response of a 3-element dipole ESPAR antenna at 5.9 GHz with varactor capacitance ranging between 0.1-1 pF.
Figure 103: Radiation pattern response of a 3-element dipole ESPAR antenna loaded with varactors at 5.9 GHz.

Figure 104: $S_{11}$ response of a 3-element dipole ESPAR antenna at 5.9 GHz with varactor capacitance ranging between 0.1-100 pF.

Since no saturation point of the $S_{11}$ has been found for the range of capacitances 0.1-1 pF an additional simulation is undertaken with capacitance ranging from 0.1-100 pF to investigate a larger range of capacitances (Figure 104). It is understood that for capacitances above 1 pF (red curve)
there is a minor change in the behavior of the dipole ESPAR antenna reflection coefficient ($S_{11}$). For 10 pF and 100 pF values (blue and orange curves respectively), a negligible difference is spotted.

The above antenna behavior determines the choice of the suitable varactor model to operate in 5.9 GHz ITS band. The Skyworks SMV 2019 varactor is selected with capacitance range close to 1 pF (specifically 2.22 pF capacitance at 0 V to 0.3 pF at 20 V).

5.6 Implementation of 3-Element Printed ESPAR Antenna with Integrated Impedance Matching Network

The small deviation in the reflection coefficient results ($S_{11}$) between the different switching states (OFF-OFF, ON-OFF) of the ESPAR antenna that are presented in Section 5.2, led to the investigation of an impedance matching network. The objective is then to create a reconfigurable matching network with varactor diodes in order to achieve a similar $S_{11}$ response for all the ON/OFF combinations of the PIN diodes (for all the operating antenna states).

A tunable matching network is designed that can provide a stable impedance matching for all the different biasing states of the ESPAR antenna. Several implementations have been proposed in the literature with L,C lumped elements and open-or-short circuited parallel stubs [96], [97]. The single-stub and double-stub matching technique is favourable for the printed antenna configuration since it can be integrated to the microstrip transmission line. Moreover, this matching technique provides a broad range of load impedances that can be matched to 50 Ω. Hence, in the following section two different matching networks have been realised so far: the double-stub and single-stub matching networks implemented in the presented 3-element printed ESPAR antenna.

5.6.1 The Double-Stub Impedance Matching Network

In the double stub configuration [98], two stubs (segments of transmission line) are inserted at predefined locations in parallel to the transmission line that feeds the antenna. They can be either open or short-circuited stubs. If the load impedance $Z_L$ changes, one should simply replace the stubs with another set of different lengths. To be able to match a wide range of load impedances and therefore realize different stubs length, variable capacitors (varactors) are employed, which they are inserted in series with the stubs.
The procedure that was followed to design the double stub matching network is explained below:

- The distance between the two stubs is selected so that a wide range of load resistances can be matched, according to equation \(\text{Re}\{Z_L\} \geq \frac{1}{\lambda_0} \sin^2 \left(\frac{2\pi d}{\lambda_0}\right)\). A \(\lambda_0/10\) distance is selected so that load resistances above 17.3 \(\Omega\) can be matched to 50 \(\Omega\).

- The length of the first stub is selected so that the admittance at the location of the second stub (before the second stub is inserted) has a real part equal to the characteristic admittance of the line \(\text{Re}\{Y_{IN}\} = 0.02\).

- The length of the second stub is selected to eliminate the imaginary part of the admittance at the location of insertion of the second stub.

The load impedance, which is referred to the input impedance of the printed monopole element, is given below from de-embedding the antenna input impedance (as given in Section 5.2) from the SMA connector and the \(\lambda/4\) transmission line responses:

- \(Z_{\text{load (ON-OFF)}} = 24.3 + j3.3\)
- \(Z_{\text{load (OFF-OFF)}} = 20.5 - j11\)

The proposed matching network, which is integrated in the printed monopole ESPAR antenna design is presented in Figure 108, consists of two shorted stubs loaded with varactor diodes separated at a fixed distance. The two shunt stubs are placed in the opposite sides of the microstrip line that feeds the active element to have minimum effect in the antenna radiation pattern. The varactor diodes are either biased or unbiased with DC voltage to realize different lengths of the shunt stubs.

However, an important restriction in the implementation of the matching network with varactor diodes needs to be taken into account. This has to do with the range of DC bias voltages that are available at the truck cabin. Currently, a 0 – 3.3 V DC bias range is offered that corresponds to a small fraction of the total capacitance range of the varactor. Hence, a careful design process to achieve an adequate impedance matching with the small tuning range for all the different antenna
The final dimensions of the double stub matching network are derived based on microwave theory after an extensive EM simulations campaign.

The varactor diodes model used, is SMV 2019 with capacitance ranging from 2.22 pF at 0 V to 0.95 pF at 3.3 V. The simplified varactor circuit model is inserted in CST as a lumped element to describe the varactor’s behavior in the matching network, as it is schematically shown in Figure 106.

![Figure 106: Simplified varactor circuit model.](image)

The following equation describes the varactor’s total capacitance \( C_J \) tunable behavior in respect to the applied DC voltage. It is the parallel combination of junction capacitance and package capacitance \( C_P \).

\[
C_T(V_R) = \frac{C_{J0}}{1 + \frac{V_R}{V_J}} + C_P
\]  

(104)

The ESPAR antenna design, including the double-stub matching network, is shown in Figure 108 as resulted from the EM simulation in CST Microwave Studio suite. The resonance frequency in the OFF state, which was found at 6.1 GHz, moves now to 5.9 GHz with an impedance matching improvement of 20 dB in the \( S_{11} \). In the ON state, the \( S_{11} \) is maintained at 5.9 GHz with \( S_{11} = -17 \) dB (see Figure 109).

Table 10: Varactor Circuit Model Parameters.

<table>
<thead>
<tr>
<th>Varactor</th>
<th>( C_{J0} ) (pF)</th>
<th>( V_J ) (V)</th>
<th>( M )</th>
<th>( C_P ) (pF)</th>
<th>( R_s ) (Ω)</th>
<th>( L_s ) (nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SMV2019</td>
<td>2.25</td>
<td>3.5</td>
<td>1.4</td>
<td>0.07</td>
<td>4.8</td>
<td>0.7</td>
</tr>
</tbody>
</table>
5.6.2 The Single-Stub Impedance Matching Network

The feasibility of achieving a satisfying impedance matching with a single stub element is also investigated in an attempt to reduce the RF system complexity. However, this requires the presence of a wider range of possible capacitances and effectively possible stub lengths. Hence, the single varactor needs to be biased with a broader DC bias voltage range. This is achieved by inserting an operational amplifier (ISL 28118) to step up the available 3.3 V up to 11.5 V. The ESPAR antenna design, including the single-stub matching network, is shown in Figure 110.
Figure 108: Picture of the double stub matching network in the current 3-element ESPAR antenna design at 5.9 GHz in CST.

Figure 109: Simulated return loss (S11) of the 3-printed monopole ESPAR with and without the double-stub impedance matching network.
Figure 110: Picture of the single stub matching network in the current 3-element ESPAR antenna design at 5.9 GHz in CST.

Figure 111: Simulated return loss (S11) of the 3-printed monopole ESPAR with and without the single-stub impedance matching network.

5.7 Implementation of 5-Wire Monopole ESPAR Antenna

One of the main disadvantages of the two planar ESPAR antenna designs that have been presented in the previous sections is the inferior omnidirectional (all OFF) state (mainly due to the “squized shape” pattern). Moreover, the main beam in the ON state is slightly tilted from the horizontal plane.
and can only be pointed in two opposite directions. A reasonable solution to these issues is to avoid the printed elements and employ a wire monopole as the array element. The wire ESPAR antenna appears in the literature for more than a decade as an ideal candidate to effectively support beam switching capability at the horizontal plane [99], [100], [101]. So far, the proposed ESPAR antennas are designed at 2.4 GHz, where the resulting size is impractical for antenna integration on the truck. In [6], a size reduction is accomplished to the overall volume of the array. In this section, an adaptation of the wire ESPAR antenna design to operate at 5.9 GHz ITS band is presented.

The proposed 5-wire monopole ESPAR antenna consists of one active and four parasitic elements in circular configuration. The wire monopole elements are placed on top of a ground plane which is equivalent to a dipole ESPAR antenna array proposed by Harrington in [52]. By employing four parasitic elements, fixed beams every 45° in azimuth can be formed by switching ON a single or two PIN diodes in the parasitic elements producing a total of 8 different directive radiation patterns. Hence, a 360° coverage in the horizontal plane can be achieved, while a symmetrical omnidirectional pattern is achieved when switching OFF all the parasitic elements. The DC biasing network essential for the control of the PIN diodes, is printed on the bottom side of the Rogers substrate. The proposed 5-element wire monopole ESPAR antenna is shown in Figure 112 and the structural dimensions are given in Table 11.

![Figure 112: ESPAR 5-element wire monopole antenna designed to operate at 5.9 GHz ITS band.](image)

<table>
<thead>
<tr>
<th>Structural Dimensions</th>
<th>In terms of wavelength</th>
<th>Absolute</th>
</tr>
</thead>
<tbody>
<tr>
<td>L_m</td>
<td>0.28·λ</td>
<td>14.24 mm</td>
</tr>
<tr>
<td>L_p</td>
<td>0.25·λ</td>
<td>12.71 mm</td>
</tr>
<tr>
<td>d</td>
<td>0.117·λ</td>
<td>5.95 mm</td>
</tr>
<tr>
<td>r_g</td>
<td>0.4·λ</td>
<td>20.34 mm</td>
</tr>
<tr>
<td>h_g</td>
<td>0.3·λ</td>
<td>15.25 mm</td>
</tr>
<tr>
<td>r_m</td>
<td>0.023·λ</td>
<td>1.15 mm</td>
</tr>
<tr>
<td>r_p</td>
<td>0.011·λ</td>
<td>0.55 mm</td>
</tr>
</tbody>
</table>
With the current configuration, where each PIN diode switches ON and OFF one of the parasitic elements, a fixed beam at predefined angles is obtained, with a gain increase from the omnidirectional (OFF) state of about 4 dB. More specifically, when switching ON a single PIN diode in a parasitic element, the main beam is directed at \( \frac{\pi}{4} \) intervals while when switching ON two parasitic elements the main beam appears at \( \frac{m\pi}{4} \) where \( n \) is even and \( m \) is odd number. It should be noted that each parasitic element behaves as a reflector in the ON state, thus forcing the main beam to the opposite direction. In Figure 113, the simulated reflection coefficient \( S_{11} \) is shown. It is seen that in both ON & OFF states, the \( S_{11} \) remains below the limit of -10 dB, which guarantees a good impedance matching.

**Figure 113:** Reflection coefficient of the wire 5-elements ESPAR antenna for the two switching states (OFF- and ON-).

**Figure 114:** Radiation pattern of absolute (IEEE) gain in the XY (azimuth) plane of the 5-element wire ESPAR antenna.
In Figure 114, four main patterns are produced when switching ON a single (Figure 114a) or two (Figure 114b) parasitic elements. The absolute maximum gain with a single element ON is found to be 4.75 dB, while it is increased to 5.5 dB with two elements ON. Since the gain in the OFF state gain is 0.8 dB, a difference of almost 5 dB is achieved.

Maximum gain is maintained in the horizontal plane (0° elevation) with ground skirt of height $h_g$ (Figure 115). The capability of this wire ESPAR antenna to eliminate the tilt of the pattern in elevation, makes it more efficient than the printed ESPAR antenna of Section 5.2, for V2V communications that take place for low elevation angles in the horizontal plane. In a future version, a continuous 360° pattern coverage will be attempted by replacing PIN diodes with varactor diodes.

5.8 Summary

Section 5 focuses on the antenna investigations for the ROADART platform prototype array. Firstly, a thorough research study was performed in order to select the best suitable position for the antenna on the truck. Specifically, a 3-monopole ESPAR antenna configuration is proposed to be used at the side mirrors of the truck. Significant performance improvement is achieved compared with the use of monopoles. Evaluation was performed with the development of a truck-antenna model and the use of the IST-WINNER channel model. SNR improvement up to 10 dB is noted.

Moreover, a thorough investigation was performed on various designs of ESPAR antenna that utilize PIN diodes as electronic switches in order to generate a reconfigurable radiation pattern. The first design that it is proposed is a novel 3-printed monopole ESPAR antenna operating at 5.9 GHz. A good agreement between the simulated and the experimental results is indicated. A satisfactory impedance matching performance for the three main switching antenna states is observed. A considerable 3 dB gain increase is achieved when the ESPAR is switched to its directive mode (ON-OFF). The second introduced antenna design is the wideband 3-element mono/dipole ESPAR antenna that provides a satisfactory impedance matching for all the antenna states and corrects the radiation tilt that is observed on the E-plane of the previous design. An expanded version of the mono/dipole ESPAR with four parasitic printed dipoles is also proposed in order to provide 9 different radiation patterns. Consequently, a theoretical investigation is carried out by replacing PIN diodes with varactor diodes in order to achieve an increased number of operating states in ESPAR antennas, and therefore, improve their beamforming capability.
Moreover, a 3-printed monopole ESPAR antenna with integrated impedance matching network is presented. The main objective of this model is the correction of the small deviation in the reflection coefficient results $S_{11}$ between the different switching states (OFF-OFF, ON-OFF) of the ESPAR antenna. Finally, a 5-wire monopole ESPAR antenna was investigated. This antenna design offers a reasonable solution to the quasi omni-directional radiation pattern and the slight tilt of the radiation from the horizontal plane.
6 Final Recommendation of Antennas and Communication Techniques

The main objective of this WP was to propose robust and adaptive techniques, algorithms, strategies, and architectures that significantly improve the system performance over the vehicular radio channels. Towards satisfying this objective, various techniques have been investigated, including diversity systems at the receiver and/or transmitter sides, spatial modulation techniques, and reconfigurable antenna arrays. Moreover, as far as cooperative relaying protocols are concerned, several approaches have also been studied, e.g., threshold based relaying selection or reactive policies for low CSI overhead. In addition, a plethora of parameters regarding wireless medium peculiarities, i.e., the time varying nature of the wireless medium, the influence of interference, the spatial correlation, should be considered. Nevertheless, several constraints are imposed by the system architecture, the communication unit, the integration of the system to the track, the need for low latency, and the vehicular communication standards. At the following paragraphs, a few details regarding these constraints will be provided, that will help to understand the rationale behind the decisions for the final recommendations that will follow.

6.1 Constraints for T2X Communication Techniques

All the proposed techniques in this WP considerably improve the system performance, taking also into consideration various generic parameters. However, since the plan is to integrate these systems in real trucks using communication units with limited resources, several additional requirements/constraints should be considered. Next, a few details regarding these requirements are provided.

**System Architecture:** The system architecture is based on two main parts; the RF module and the (baseband) communication unit. In the RF module, simple signal processing procedures are performed including packet detection, frequency correction, FFT, and equalization (please see deliverable 4.1). This approach allows the implementation of the receive diversity algorithm entirely on the communication unit. The communication unit, is capable of simultaneously process signals from at most 2 RF modules in order to support low latency operation. One control port per radio is included in order to support antenna beam switching from 2 reconfigurable antennas per RF module. From this architecture design, the following outcomes can be extracted

- MRC can only be performed in the communication unit
- MRC with at most 4 baseband signals can be supported
- Beam switching among the beams provided by a reconfigurable antenna can be supported.

However, at most 2 reconfigurable antennas per RF module are supported.

**Integration:** The dimensions of the side mirrors, which probably will host the antennas on the trucks, are limited. Therefore, in order to avoid spatial correlation effects, and their negative consequences as it is already shown in Section 3.1, only 2 antennas should be integrated in each mirror.

**Standardization issues:** The communication techniques that have been proposed in the framework of this WP, offer an excellent performance with reduced complexity, as compared to the existing ones. However, some of them, e.g., spatial modulation, require also modifications on the existing ITS-G5 standard, mainly regarding the use of partial channel information at the transmitter. It is almost certain that future ITS standards will support this kind of operation. However, the final techniques that will be recommended by ROADART to be implemented and integrated in the trucks should be compatible with the existing standards.
6.2 Qualification of Communication Techniques to be Implemented

Based on the rationale that has been presented in the previous section, the recommended techniques for the ROADART platform are specified. These techniques provide considerable performance improvements, while also respect all the constraints and requirements imposed by the wireless medium behaviour as well as issues coming from the implementation and integration procedures.

6.2.1 Diversity and Cooperative Techniques

The system concept of all ROADART qualified diversity techniques are depicted in the following figures, complemented by a few details regarding their mode of operation as well as some performance comparison results.

![Figure 116: System concepts for cases 1 and 2.](image)

In Case 1, as shown in Figure 116, one antenna monopole is connected to each RF module and the outputs of these modules feed the communication unit, via a digital link, where the MRC algorithm is implemented. In Case 2, two antenna monopoles are connected to each RF module, where the one offering the best performance, in terms of the received SNR, is selected for further processing. The selection is performed in the communication unit and the decision is sent to the RF module via a ctrl signal.

Case 3, as shown in Figure 117, has a similar design with 2. The only difference is that all the received signals of the four monopoles, following the initial processing from the RF module, are combined under the MRC principle in the communication unit. In Case 4, 1 ESPAR antenna is connected to each RF module. The pattern offering the maximum received SNR is selected at the communication unit and this info returns to the RF module via the control port.
In Case 5, as shown in Figure 118, 1 ESPAR antenna and one monopole are connected to each RF module. The pattern selection for the ESPAR antenna is performed in a similar manner as in case 4. In the communication unit, all the received signals are combined based on the principle of MRC. In Case 6, as shown in Figure 119, 2 ESPAR antennas are connected to each RF module. In the communication unit, all the received signals are combined based on the principle of MRC.

Next, a performance comparison of all the different system designs is provided. In particular, in Figure 120, assuming a simple communication scenario, under Rayleigh fading conditions and BPSK modulation, the BEP performance of all schemes under consideration is presented as a function of the average received SNR. In this figure, it is shown that the best performance is obtained with case 6, while the worst with case 1. However, for all scenarios under investigation, a considerable performance improvement is realized, as compared to the basic scenario, where no diversity reception is assumed. Interesting observations that come along with the comparisons made in this figure are the following. Case 4, with the two ESPAR antennas, provide better performance
as compared to cases 2 and 3, despite the fact that both of them support 4 antennas. Therefore, with this choice, all

![Performance comparison of the different diversity systems.](image)

Figure 120: Performance comparison of the different diversity systems.

objectives of improved performance, reduced complexity and cost are achieved. Moreover, the best option in terms of performance is when two ESPAR antennas are employed per RF module, i.e., case 6. However, it noteworthy here that considering one monopole and one ESPAR per RF module as in case 5, provides some benefits. These benefits include issues related to the synchronization, the induced system complexity, and the plethora of the different vehicular communication applications that can be supported. Finally, it should be noted that all the aforementioned techniques satisfy the constraints described in Section Constraints for T2X Communication Techniques.

As far the cooperative techniques are concerned, the relay selection scheme presented in Section 4.2 is qualified. The system model of this scheme is presented in Figure 121. In particular, one source (S) attempts to communicate with the destination with the aid of N relays (for simplification purposes in Figure 121 only 2 relays have been considered). The relay that was selected in the previous round of communications, e.g., relay 1 (R1) sends request to send packet to the final destination. From this packet, the destination estimates the received instantaneous SNR, which is compared with a predefined threshold, $\gamma_{th}$. If the estimated value of the SNR exceeds that threshold, then this relay is selected and no further processing is required. Otherwise, a flag packet is send to all relays that successfully received the original message (including the tagged relay) asking for the initiation of a best relay procedure. Therefore, the proposed scheme offers an excellent compromise between performance and complexity as compared to existing ones. Moreover, this technique is compatible with the existing standards and thus no modification is required. More details regarding this scheme can be found in section 4.2.
6.2.2 Low Complexity Transceiving Techniques with T2T Communications

As far as low complexity transceiving techniques are concerned, the final recommendation of the ROADART project is the extended open loop beamforming scheme that is already presented in Section 4.4. The eOLB technique can be applied with no feedback from the receiver to the transmitter. This becomes feasible since the communication unit of every truck operates as Tx and Rx and can use the information acquired from the receiving mode when it transmits. Therefore, an amendment to the standards is not necessary. Moreover, this scheme satisfies all the constraints described in Section 6.1 “Constraints for T2X Communication Techniques”.

6.3 Qualification of Antenna Designs to be Implemented

The recommendations that are extracted from the antenna investigations are classified in two parts. The position of the antenna on the truck and the type of the antenna that is going to be implemented in ROADART platform. As far as antenna positioning on the truck is concerned, two different locations were investigated: the roof of the truck and the side mirrors. The final recommendation of the ROADART is to place the antenna on the side mirrors of the truck. According to the results that are presented in Figure 70 of Section 5.1.2, it is clear that the positioning of the antenna on the side mirrors causes a sharp increase to the SNR at the Rx. Furthermore, a considerable reduction of the ABEP (vs Tx-Rx separation) is observed.

As far as the antenna type is concerned, the general recommendation is to use ESPAR antennas. The ability of the ESPAR array to adapt its radiation beam according to the noise environment (and steer it towards a desired direction or suppress it in the direction of an interfering source) can be considered highly preferable in T2T communications. In addition, it would be desirable for an automotive antenna to be compact in order to have no impact in the vehicle’s design, avoid causing additional air drag or be installed in relatively small spaces (e.g. truck side mirror). An ESPAR antenna can meet both aforementioned specifications. Furthermore, the single active array element corresponds to a single RF chain (less cabling), a simple feeding network and uncomplicated control circuitry, and thus, it considerably decreases antenna complexity and cost.
Specifically, two designs among the several presented ESPAR antennas are qualified as the best suitable candidates for the final implementation and usage in ROADART final demonstration. Firstly, the 3-printed ESPAR antenna with the integrated impedance matching network provides a valuable solution to one of the main drawbacks of ESPAR antennas, while preserving the advantage of the 3 dB gain increase between the OFF-OFF and the ON-OFF states. In particular, the proposed design manages to appropriately adjust the antenna impedance and thus, correct the small deviations in the reflection coefficient results between the different switching states of the antenna. This is achieved by a tunable matching network (with varactor diodes) that produces a similar $S_{11}$ response for all the ON/OFF combinations of the PIN diodes (for all the operating antenna states). The double-stub matching technique is qualified as favourable since it can be easily integrated to the microstrip transmission line.

The 5-wire monopole ESPAR antenna design is also recommended for three basic reasons. Initially, the utilization of four parasitic elements and the ability to switch on several ON/OFF combinations extends the pattern reconfigurability of the antenna. It offers 9 antenna operating states that correspond to 9 different radiation patterns (full azimuth coverage with 8 directive beams every 45°). Moreover, the wire monopole selection as the ESPAR’s element along with the circular ground plane and the circular arrangement of the parasitic elements provides a full azimuth symmetry at the antenna design and therefore, provides a perfectly omni-directional radiation pattern at the “all OFF” antenna state. Finally, no radiation upward tilt is observed since a ground skirt is employed in order to preserve the maximum directivity of the beam on the horizontal plane.
7 Conclusions

This report describes the efforts made in WP3 towards the identification and recommendation of advanced antennas and communication techniques for T2T communications. In this context, a set of algorithms and techniques were introduced, while their performance assessment was also provided. In the following text, it is recapitulated how the main project/WP objectives were addressed.

The main objective of WP3 in ROADART was to investigate, evaluate, and measure multi-antenna transmission techniques that can be tailored to the specificities of the T2T and T2I communication links. To achieve this objective, a new simulator for the IEEE 802.11p standard, has been implemented in MATLAB/OCTAVE, while all major receive diversity reception techniques for T2T and T2I communications have been also implemented and their performance was evaluated in the simulation platform. Moreover, the proposed diversity/cooperative techniques should take into consideration all the peculiar parameters of the T2T/T2I wireless links. Towards satisfying this objective, various effects have been identified and their impact on the diversity/cooperative performance has been analytically investigated, such as the spatial correlation among the antennas, the existence of interfering effects, and the existence of outdated CSI.

Another important objective in ROADART was to design transmit/receive diversity techniques as well as relay selection for T2T and T2I mobile radio channels in order to increase reliability, robustness and throughput, taking into account the trade-off between performance and complexity. Towards satisfying this objective, a new reconfigurable antenna pattern selection scheme is proposed that aims to reduce the complexity of the traditional antenna selection techniques. Moreover, a new relay selection scheme has been also investigated under different T2T communication scenarios that also aim to reduce the signal processing complexity that the traditional relay selection schemes induce to the system. Another key objective, was to investigate alternative beamforming techniques for T2T and T2I channels by employing parasitic antenna arrays. Towards satisfying this objective, a novel eOLB technique that is realized through pattern selection is designed and demonstrated via a proof-of-concept indoor experiment. The development of efficient spatial modulation diversity techniques was another important objective in task. In this context, the adoption of SSK as well as SM in IVC scenarios have been studied and the benefits of this approach in terms of complexity versus performance trade-off have been investigated, based on a comparison with several well-established MIMO schemes.

The design of low-cost small sized antennas and the testing of their positions on the vehicles was an important objective of this WP. A sub-objective on this area was to introduce the parasitic antenna arrays in vehicular communication and implement special parasitic antenna designs. It was shown that the proposed antenna-truck configuration (ESPARs on the truck side mirrors) achieved considerable performance improvement. Two types of 3-element printed ESPAR antenna were designed and fabricated providing simulation and measurement results in good agreement and exhibiting a valuable pattern reconfigurability (3-4 different radiation patterns). An impedance matching network mechanism has been proposed to provide satisfying reflection coefficients at all antenna states. Moreover, the number of the antenna operating states increased to 9 radiation patterns in simulation by investigating the design of two 5-element ESPAR antennas (5-element printed mono/dipole ESPAR and 5-wire monopole ESPAR). Finally, a further expansion of the ESPAR pattern reconfigurability is studied theoretically by exploiting the tuning effect of the varactor diodes that can offer numerous antenna patterns with less array elements.

Concluding, the techniques that are recommended for adoption to the final ROADART communication platform were presented. The recommendations considered all the constraints and requirements imposed by the wireless medium behaviour as well as issues coming from the implementation and integration procedures.
8 Bibliography


[87] E. T. S. I. (ETSI), Final draft ETSI ES 202 663 V1.1.0, Intelligent Transport Systems (ITS); European profile standard for the physical and medium access control layer of Intelligent Transport Systems operating in the 5 GHz frequency band, Nov 2009..


[92] RO4725JXR & RO4730JXR Antenna Grade Laminates, Rogers Corporation, Application Note..

[93] SMP1320 Series: Low resistance, low capacitance, plastic packaged PIN diodes, Skyworks Solutions Inc., Application Note..


## 9 Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ABEP</td>
<td>Average Bit Error Probability</td>
</tr>
<tr>
<td>ACK/NACK</td>
<td>Acknowledgement/Negative-Acknowledgement</td>
</tr>
<tr>
<td>AF</td>
<td>Amplify and Forward</td>
</tr>
<tr>
<td>AGC</td>
<td>Automatic Gain Control</td>
</tr>
<tr>
<td>AIFS</td>
<td>Arbitration Interframe Space</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BPCU</td>
<td>Bits-Per-Channel-Use</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>BRS</td>
<td>Best Relay Selection</td>
</tr>
<tr>
<td>CCDF</td>
<td>Complementary Cumulative Distribution Function</td>
</tr>
<tr>
<td>CCI</td>
<td>Co-Channel Interference</td>
</tr>
<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
</tr>
<tr>
<td>CPW</td>
<td>CoPlanar Waveguide</td>
</tr>
<tr>
<td>CSI</td>
<td>Channel State Information</td>
</tr>
<tr>
<td>CRC</td>
<td>Cyclic Redundancy Check</td>
</tr>
<tr>
<td>CSMA/CA</td>
<td>Carrier Sense Multiple access with Collision Avoidance</td>
</tr>
<tr>
<td>CTS</td>
<td>Clear to Send</td>
</tr>
<tr>
<td>CW</td>
<td>Contention Window</td>
</tr>
<tr>
<td>DBPSK</td>
<td>Differentially Binary Phase Shift Keying</td>
</tr>
<tr>
<td>DF</td>
<td>Decode-and-Forward</td>
</tr>
<tr>
<td>DN</td>
<td>Double-Nakagami</td>
</tr>
<tr>
<td>EDCA</td>
<td>Enhanced Distributed Channel Access</td>
</tr>
<tr>
<td>EIFS</td>
<td>Extended Interframe Space</td>
</tr>
<tr>
<td>eOLB</td>
<td>Extended Open Loop Beamforming</td>
</tr>
<tr>
<td>ESPAR</td>
<td>Electronically Switched Parasitic Array Radiator</td>
</tr>
<tr>
<td>FD</td>
<td>Full-Duplex</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>FPGA</td>
<td>Field Programmable Gate Array</td>
</tr>
<tr>
<td>ICI</td>
<td>InterChannel Interference</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse Fast Fourier Transform</td>
</tr>
<tr>
<td>i.i.d.</td>
<td>Independent and identically distributed</td>
</tr>
<tr>
<td>INR</td>
<td>Interference to Noise Ratio</td>
</tr>
<tr>
<td>ITS</td>
<td>Intelligent Transportation Systems</td>
</tr>
<tr>
<td>IVC</td>
<td>Intevehicular Communications</td>
</tr>
<tr>
<td>LI</td>
<td>Loop-Interference</td>
</tr>
<tr>
<td>LoS</td>
<td>Line of Sight</td>
</tr>
<tr>
<td>MAC</td>
<td>Media Access Control</td>
</tr>
<tr>
<td>MANET</td>
<td>Mobile Ad-Hoc Networks</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
</tr>
<tr>
<td>----------</td>
<td>--------------------------------------------------</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
</tr>
<tr>
<td>MGF</td>
<td>Moments Generating Function</td>
</tr>
<tr>
<td>MMSE</td>
<td>Minimum Mean Square Error</td>
</tr>
<tr>
<td>MRC</td>
<td>Maximum Ratio Combining</td>
</tr>
<tr>
<td>NAR</td>
<td>Number of Active Relays</td>
</tr>
<tr>
<td>NAV</td>
<td>Network Allocation Vector</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>OP</td>
<td>Outage Probability</td>
</tr>
<tr>
<td>PDF</td>
<td>Probability Density Function</td>
</tr>
<tr>
<td>PEP</td>
<td>Packet Error Probability</td>
</tr>
<tr>
<td>PHY</td>
<td>Physical Layer</td>
</tr>
<tr>
<td>PPP</td>
<td>Poisson Point Process</td>
</tr>
<tr>
<td>PSK</td>
<td>Phase Shift Keying</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of Service</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>R-D</td>
<td>Relay-Destination</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RS-GBSM</td>
<td>Regular Shaped Geometry-Based Stochastic Models</td>
</tr>
<tr>
<td>RTS</td>
<td>Ready to Send</td>
</tr>
<tr>
<td>RV</td>
<td>Random Variable</td>
</tr>
<tr>
<td>S-R</td>
<td>Source-Relay</td>
</tr>
<tr>
<td>SD</td>
<td>Selection Diversity</td>
</tr>
<tr>
<td>SEP</td>
<td>Symbol Error Probability</td>
</tr>
<tr>
<td>SIC</td>
<td>Successive Interference Cancellation</td>
</tr>
<tr>
<td>SIFS</td>
<td>Short Interframe Spacing</td>
</tr>
<tr>
<td>SIMO</td>
<td>Single Input Multiple Output</td>
</tr>
<tr>
<td>SIR</td>
<td>Signal-to-Interference</td>
</tr>
<tr>
<td>SISO</td>
<td>Single-Input Single-Output</td>
</tr>
<tr>
<td>SM</td>
<td>Spatial Modulation</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>SOS</td>
<td>Second Order Scattering</td>
</tr>
<tr>
<td>SSK</td>
<td>Space Shift Keying</td>
</tr>
<tr>
<td>STBC</td>
<td>Space Time Block Codes</td>
</tr>
<tr>
<td>RSU</td>
<td>Road-Side Unit</td>
</tr>
<tr>
<td>TOSD-SSK</td>
<td>Transmit Orthogonal Signal Diversity</td>
</tr>
<tr>
<td>T2I</td>
<td>Truck-to-Infrastructure</td>
</tr>
<tr>
<td>T2T</td>
<td>Truck-to-Truck</td>
</tr>
<tr>
<td>ULA</td>
<td>Uniform Linear Array</td>
</tr>
<tr>
<td>USRP</td>
<td>Universal Software Radio Peripheral</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
</tr>
<tr>
<td>----------</td>
<td>--------------------------------------------------</td>
</tr>
<tr>
<td>VANET</td>
<td>Vehicular Ad-hoc Networks</td>
</tr>
<tr>
<td>V2V</td>
<td>Vehicle to Vehicle</td>
</tr>
<tr>
<td>VNA</td>
<td>Vector Network Analyzer</td>
</tr>
<tr>
<td>WSSUS</td>
<td>Wide Sense Stationary Uncorrelated Scattering</td>
</tr>
<tr>
<td>ZF</td>
<td>Zero-Forcing</td>
</tr>
<tr>
<td>3-D</td>
<td>Three-Dimensional</td>
</tr>
</tbody>
</table>