APPLICATION OF WIRELESS SENSOR NETWORKS IN POWER SYSTEM MONITORING AND CONTROL: A Study on Wireless Channel, Interference and physical layer Issues

Amadou Louh
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Amadou Louh

Pr. J. smit
Dr. H. Nikookar
Dr. D. Djairam
G. Bajracharya
H. Lu

TU Delft
Electrical Engineering, Mathematics and Computer Sciences
Departments of Power Engineering and Telecommunications
Preface

This report has been written to present the results of my study on wireless channel and interference in open-air substation environments. A study performed to obtain the degree of Master of Science at the Delft University of Technology (TU Delft). The report also presents two proposals of wireless technologies, for data transmission in these environments. It has been written as a stand-alone document to allow any reader, with a background in power engineering as well as telecommunication, to go through the presented topics without the need of additional materials.

References to scientific papers and books, journal articles and web pages are used throughout this text to enable the interested reader to delve into detailed reading about the discussed topics. Likewise, appendices are mentioned in the text and provided at the end, for those interested in additional concepts and materials such as orthogonality and cyclic prefixing and, Matlab programming files used for the various simulations.

I consider myself fortunate to have been one of the many students who had the opportunity to study at the famous Delft University of technology (TU Delft) and in particular at the excellent faculty of electrical engineering. After obtaining my bachelor degree in electrical engineering, I followed a master program in power engineering and a master program in telecommunication as gently reflected by the multidisciplinary character of my thesis subject.

I am pleased to acknowledge the help, support and guidance of my supervisors: Dr. Homayoun Nikookar, Dr. ir. Dhiradj Djairam, G. Bajracharya and Hao Lu. Furthermore, their many suggestions and remarks have contributed enormously in the writing of this report. Special thanks to dr. ir. S. Meijer for helping me start this thesis work.

Finally, I’ m also grateful to Pr. J. Smit who has given me the opportunity to graduate inside the warm-hearted and enthusiastic research group of High Voltage Engineering.

Delft, november 2009

Amadou Louh
### Abbreviations

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<th>Description</th>
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<tbody>
<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>AIS</td>
<td>Air Insulated Substation</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>bps</td>
<td>bit(s) per second</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
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<tr>
<td>CHP</td>
<td>Combined Heat and Power</td>
</tr>
<tr>
<td>CP</td>
<td>Coupling Path</td>
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<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DS</td>
<td>Distribution Substation</td>
</tr>
<tr>
<td>DSSS</td>
<td>Direct Sequence Spread Spectrum</td>
</tr>
<tr>
<td>EMI</td>
<td>ElectroMagnetic Interference</td>
</tr>
<tr>
<td>EMC</td>
<td>ElectroMagnetic Compatibility</td>
</tr>
<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
</tr>
<tr>
<td>GIS</td>
<td>Gas Insulated Substation/Switchgears</td>
</tr>
<tr>
<td>GS</td>
<td>Generation Substation</td>
</tr>
<tr>
<td>GSM</td>
<td>Global System for Mobile communications or Groupe Spécial Mobile</td>
</tr>
<tr>
<td>GP</td>
<td>Generation Plant</td>
</tr>
<tr>
<td>HV</td>
<td>High Voltage</td>
</tr>
<tr>
<td>HVDC</td>
<td>High Voltage Direct Current</td>
</tr>
<tr>
<td>IS</td>
<td>Interference Source</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter Symbol Interference</td>
</tr>
<tr>
<td>IV</td>
<td>Interference Victim</td>
</tr>
<tr>
<td>Kbps</td>
<td>Kilo bits per second</td>
</tr>
<tr>
<td>kV</td>
<td>Kilo Volt</td>
</tr>
<tr>
<td>LAN</td>
<td>Local Area Network</td>
</tr>
<tr>
<td>LISI</td>
<td>Light Inter Symbol Interference</td>
</tr>
<tr>
<td>LOS</td>
<td>Line Of Sight</td>
</tr>
<tr>
<td>lx</td>
<td>lux (lumen per square meter): a unit for light incident on a surface.</td>
</tr>
<tr>
<td>MAC</td>
<td>Medium Access Control</td>
</tr>
<tr>
<td>MEMS</td>
<td>Micro Electromechanical System</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>O-QPSK</td>
<td>Quadrature Phase Shift Keying with phase offset</td>
</tr>
<tr>
<td>PD</td>
<td>Partial Discharge</td>
</tr>
<tr>
<td>PHY</td>
<td>Physical Layer</td>
</tr>
<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
</tr>
<tr>
<td>PV</td>
<td>Photovoltaic</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>RX</td>
<td>Receiver</td>
</tr>
<tr>
<td>SISI</td>
<td>Strong Inter Symbol Interference</td>
</tr>
<tr>
<td>SS</td>
<td>Sub-transmission System</td>
</tr>
<tr>
<td>TS</td>
<td>Transmission Substation/System</td>
</tr>
<tr>
<td>TX</td>
<td>Transmitter</td>
</tr>
<tr>
<td>XLPE</td>
<td>Cross-Linked Polyethylene</td>
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<tr>
<td>$B_{coh}$</td>
<td>The coherence bandwidth of the wireless channel [Hz]</td>
</tr>
<tr>
<td>$c$</td>
<td>Speed of light [m.s$^{-1}$]</td>
</tr>
<tr>
<td>$d$</td>
<td>The average distance between transmitter and receiver [m]</td>
</tr>
<tr>
<td>$E_b$</td>
<td>Signal energy per bit [dB]</td>
</tr>
<tr>
<td>$f_c$</td>
<td>Carrier frequency of transmitted signals [Hz]</td>
</tr>
<tr>
<td>$f_s$</td>
<td>The sampling frequency [Hz]</td>
</tr>
<tr>
<td>$N$</td>
<td>Number of multipath components in wireless channel [.]$^1$</td>
</tr>
<tr>
<td>$N_0$</td>
<td>Noise power spectral density [dBm/Hz]</td>
</tr>
<tr>
<td>$N_{Symb}$</td>
<td>The number of transmitted BPSK/QPSK/QAM symbols [.]</td>
</tr>
<tr>
<td>$R_{Symb}$</td>
<td>The symbol rate [symb.s$^{-1}$]</td>
</tr>
<tr>
<td>$T_{Signal}$</td>
<td>The duration of the transmitted signal [s]</td>
</tr>
<tr>
<td>$T_{Symb}$</td>
<td>The symbol duration [s]</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Modulus of the wireless channel complex amplitude [.]</td>
</tr>
<tr>
<td>$\alpha_n$</td>
<td>Attenuation of the $n^{th}$ multipath components [.]</td>
</tr>
<tr>
<td>$\gamma$</td>
<td>Wireless channel Path-loss exponent [.]</td>
</tr>
<tr>
<td>$\Phi$</td>
<td>Argument of the wireless channel complex amplitude [radians]</td>
</tr>
<tr>
<td>$\varphi_n$</td>
<td>Path-phase of the $n^{th}$ multipath components [radians]</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>Wavelength of transmitted signals [m]</td>
</tr>
<tr>
<td>$\sigma_n$</td>
<td>Additive White Gaussian Noise standard deviation [dB]</td>
</tr>
<tr>
<td>$\sigma_m$</td>
<td>Standard deviation of the random variable accounting for multipath fading [.]</td>
</tr>
<tr>
<td>$\sigma_s$</td>
<td>Standard deviation of the random variable accounting for shadow fading [dB]</td>
</tr>
<tr>
<td>$\tau_{max}$</td>
<td>Maximal delay spread of the wireless channel [s]</td>
</tr>
<tr>
<td>$\tau_{mean}$</td>
<td>Mean excess delay of the wireless channel [s]</td>
</tr>
<tr>
<td>$\tau_n$</td>
<td>Path-delay of $n^{th}$ multipath components [s]</td>
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<tr>
<td>$\tau_{rms}$</td>
<td>RMS-delay spread of the wireless channel [s]</td>
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$^1$ [.] Meaning that the considered quantity is unit less.
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Chapter 1

Introduction

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1.1 Background of the thesis
1.2 Motivation of the thesis
1.3 Research framework and focus of the thesis
1.4 Outline of the thesis

This chapter is intended to give the reader a background on the motivation of this thesis work. It sketches the larger picture of which the thesis is part of and reveals the essence and structure of the thesis at the same time. The problems that give rise to the subject of the thesis are mentioned in the first section and discussed in more detail in section 1.2. The proposed solution to these problems, which also indicates the research framework of the thesis, together with the focus of the thesis is then presented in section 1.3. The last section describes the outline of the thesis.

1.1 Background of the thesis

The Dutch electricity grid is one of the most reliable in the world [1]. However, in the near and far future, significant changes in the grid will occur. First of all, more renewable energy sources such as wind and solar power are to be integrated into the grid. Secondly, an increase in the application of power electronics is foreseen. Thirdly, a progressive liberalization of the electricity market is taking place. All these changes, as discussed in the following, will affect the grid in terms of reliability of power transfer, stability of the overall system and safety for the grid components and, more importantly, for the people nearby grid facilities. On top of that, most of today’s grid components, which are now practically at the end of their life time, were designed with an old-fashion vertical type of power distribution philosophy in mind. As a result of this, 40-years-old power networks will have to deal with high fluctuations and huge power surges [2].

1.2 Motivation of the thesis

1.2.1 Renewable energy sources

A classical electrical power system, designed to supply electrical power to the consumers, can be subdivided in 3 main systems, namely, the generation, transmission and distribution system, as depicted in Figure 1.1. The generation consists of a small number of large power plants. According to [8], actual power system infrastructure is even 50 years old and definitely shows signs of ageing. The term small here is relative to the huge number of small (compared to conventional power plants) decentralized or distributed generating units as discussed in the following of section 1.1.
conventional power plants, which convert mechanical power from conventional primary energy sources into electrical power [3]. One of the most important aspects in the design and operation of the electrical power system is that the supply and demand of power should be equal at any time, due to the unresolved problems associated with electrical energy storage [4]. As the demand still increases nowadays [5], there is a need for more generation. However, economical, social, political and mostly environmental issues limit the construction or expansion of large conventional power plants [3]. One of the proposals of the European Commission to fight climate change is the increase of the share of renewable energy sources in the total energy production, from six percent in 2005 to twenty and even thirty percent by 2020 [2]. Therefore, more renewable energy sources are to be applied at the generation side.

**Figure 1.1:** A conventional classical electrical power system showing the old fashion vertical type of operation, as the power flow is one-directional, from top (generation system) to bottom (consumer). Generating units are large conventional power plants [picture taken from [3]].

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3 Conventional primary energy sources are energy sources such as, coal, gas, water (hydro power) and uranium (nuclear). These sources of energy are used in large conventional power plants which can deliver hundreds of MW (Mega Watts) of electrical power.
Generation systems relying on renewable energy sources convert primary energy sources of a random nature such as wind, solar irradiance, tidal movement and wave motion, into electrical energy. Generating units based on that type of energy sources, therefore, have an intermittent electrical power output. For example, the electrical power output of a wind turbine is mainly characterized by the downstream wind velocity at the exit of the rotor blades, which can change at any time. Photovoltaic (PV) modules also, have an electrical power output strongly dependent on the diffuse- and the direct-horizontal irradiance, which are influenced by weather changes and clouds movements [3].

Therefore, the problem with renewable energy sources lies in the stochastic character of the electrical power output which flows into the electrical power grid. This has an adverse effect on maintaining the balance between supply and demand. As the load (consumers of electrical energy) varies according to the customer demand in time, balance should be maintained by adjusting the generation. This can be done relatively easily, with a deterministic supply as offered by conventional generation, because the power output can be accurately predicted. The increase in the use of renewable energy sources, with the idea of eventually get rid of conventional energy sources thus means that random load fluctuations will have to be compensated with a random varying generation. This will be a big challenge, if we recall the difficulties face today in electric energy storage [4]. A communication network, relaying real time information on power inputs and outputs at all location where it is needed, (e.g.: control centers) can significantly help maintain the balance.

1.2.2 Increase in application of power electronics

The electrical power system is designed to operate as an AC (Alternating Current) system with a constant frequency of 50 Hz or 60 Hz. However, generating units relying on renewable energy sources, generate electricity at different frequencies. For example, certain types of wind turbines generate electricity as AC with variable frequency. PV modules generate electricity as DC (Direct Current). In order to connect a generating unit to the power grid, the output power of the unit should have a frequency equal to the power system frequency. Therefore, to connect renewable energy sources to the grid, a power electronic converter is needed. Also, the interconnection of energy storage elements (e.g.: fuel cells, batteries and superconductive inductors) in order to level load peaks, requires power electronic components [6]. The problem with power electronics is the introduction of higher harmonics on top of the fundamental system frequency. These higher harmonics are harmful for all equipment designed to function at the power system frequency of 50 Hz or 60 Hz, especially rotating machinery. Harmonics can cause excessive overheating in transformers and generators windings, false operation of circuit breakers and dielectric failure in capacitors, to name a few [7]. A sensor network recording currents and voltages, up to higher harmonics, at critical points in the grid and distributing the information will help in preventing this to happen.

1.2.3 Liberalization of the electricity market

The progressive liberalization of the electricity market will give rise to a noticeable increase in energy trading which will lead to an increase in unpredictable load flows [2]. The problem brought by this electricity market liberalization is the complexity of power flow control, which will change from one-directional to bi-directional. As almost everyone will be able to inject electrical power into the grid, customers can now be considered as part of the generation system. As the number of energy providers will increase, and the customers still free to choose where to buy their energy, a profitable energy market for both parties is
unthinkable without a reliable communication network. Likewise, fair trade is only possible if information on productions and demands is known to everyone involved. This requires again a sensor network, collecting real-time information on these productions and demands in the entire electricity grid.

1.2.4 Vertical to horizontal transformation of electrical power systems

Generating units relying on renewable energy sources are relatively small, as compared to conventional power plants, and spread out over the power system, hence the name distributed power generation systems. Distributed generation transforms the vertical operated power system into a horizontal operated power system, in which the power flow is bi-directional [3]. Figure 1.2 gives an impression of the transformation of the power system from vertical to horizontal, due to the introduction of distributed generation. Figure 1.2 (a) is a simple representation of the vertically operated system of Figure 1.1 and Figure 1.2 (b) the migration to horizontal operation.

![Diagram](image)

**Figure 1.2:** From vertical to horizontal mode of operation. (a) A simple representation of the old fashion vertical operated power system of Figure 1.1. (b) A horizontal operated power system. The number of generating stations (G) increases and is spread out over the system. Therefore, the power flow becomes bi-directional as indicated by the grey arrows. (GS: Generating Substation, TS: Transmission Substation and DS: Distribution Substations).

Power flow control is essential for the stability of the power system. In a horizontally operated power system, this will become more difficult. Again, a sensor network recording real-time information on power flow in the entire system will significantly help stability control-systems.

---

4 Distributed generation includes not only renewable energy sources but also small scale decentralized power generators based on conventional primary energy sources. Combined Heat and Power (CHP) generators and microturbine generators supplied by gas belongs to this family.
All above mentioned changes will increase the probability of occurrence of power system faults, which can spread into cascading failures. Therefore, they endanger the reliability of the grid in terms of continuity of power supply, failure of the infrastructure and safety of the maintenance staff. Actual grid protection tools, to prevent cascading failures to end up in black outs, rely mainly on extensive offline studies. Therefore, by the time instabilities are detected, it is already too late to take proper action. Fast real-time information acquisition and analysis for immediate local ‘healing’ is required. In order to be able to properly control, monitor and assess the condition of the high voltage assets, “intelligence” has to be included into the assets. Reliable and secure communication is the key and primary challenge in deploying self-healing, automated, wide-area grid control systems [8]. The possibility for information exchange between the components and central control systems would therefore be one of the major features of this intelligent system.

1.3 Research framework and focus of the thesis

As stated above, there is a need for real-time measurements at all critical points of the power system and for distribution of this information in the entire grid. The critical points are points where the information is collected. It can be a substation or more precisely a High Voltage component inside a substation like a power transformer. To realize the information exchange, we propose the integration of a wireless sensor network embedded in or on top of the electrical power system. This is a solution still absent in today’s Dutch electricity grid. Figure 1.3 gives an impression of this wireless sensor network.

![Figure 1.3: An impression of a wireless sensor network on top of, or embedded in a future horizontally operated electrical power system. The sensor network nodes (grey bullets) represent the critical points in the grid (e.g.: substations), where real-time measurements are to be performed. These nodes consist of interconnected sub-nodes (e.g.: power transformers). The lines between the nodes symbolize network wireless links used to share or distribute the information obtained from the real-time measurements.](image)

The case for a wireless type of communication is further supported by the ease of installation and maintenance of wireless systems as compared to wired systems. Furthermore, the rapid progress in microelectronics results in manufacture of sensing units of decreasing sizes and

---

5 Examples of power system faults are: generation-load imbalances, sudden changes in grid topology, lack of voltage support and short-circuits [8].
increasing capabilities. Micro Electromechanical systems (MEMS) can have the size of a coin and are equipped with multiple sensors, microprocessors, transceivers and energy supply units. This, of course, reduces the cost significantly and enables the use of many of them [9].

**Focus of the thesis**

The wireless sensor network for monitoring and controlling the electrical power flow inside the electrical power system, and protection of the equipment, can only be implemented if the wireless link between two sensor network nodes of the communication system is reliable. In other words, the acquired information at the sensor nodes must be transferred via the wireless link without errors from one node to the others. This wireless link between two sensor nodes will be the focus of this thesis. The objective of this thesis is to determine whether or not, it is possible to accurately transfer data, wirelessly, in an environment prone to electromagnetic interference, such as substations. For this purpose, a study of the major interference sources or electromagnetic emissions polluting a substation environment is required. We will then focus on the simple system as depicted in Figure 1.4. The maximum allowable and required data rate, for which the probability of bit errors is acceptable, will determine the quality on the wireless link.

![Image](image-url)

**Figure 1.4:** A simplified scheme consisting of 2 sensor network nodes and a wireless link in between. The sensor node can represent 2 power transformers inside a substation. The wireless channel between the nodes contains the major interference sources in a substation environment.

In order to be able to determine the quality of this wireless link and therefore, make qualitative and quantitative statements about the possibilities and limitations of wireless data transfer inside substations, the work has been divided into 3 parts. The first part concentrates on the wireless channel. In the second part, we will look at the sensor nodes and in particular, the type of data they process and the transmission technology they use to exchange their data. In the third part, the overall system is considered, where the accent will be on power system application.

**1.4 Outline of the thesis**

It should be noted that many used concepts in this thesis work have been explained in details, due to the multidisciplinary character of the subject, for the report to be as stand-alone as possible. Any reader with a power engineering or telecommunication background should be able to read through the report without needing additional notes.

The thesis is subdivided in three parts. The first part entitled: *Description and Modeling of the Wireless Channel*, deals with the properties and associated parameters of the wireless
channel, with assumptions made for the special case of substation environments. It is
organized as follows:

- Chapter two reveals the general properties of a wireless channel together with the
  parameters that will be used to build the channel model.

- Chapter three classifies and models the main interference sources present in a
  substation environment. A frequency analysis of interfering electromagnetic waves is
  also provided for a better understanding of interference and modeling of this
  interference.

- Chapter four combines the general channel parameters found in chapter two with the
  channel interference of chapter three, to build the final wireless channel model
  specific for substation’s environments.

- Chapter five implements the channel model derived in chapter four, and tests its
  validity. A simple transmitter and receiver model have been implemented for this
  purpose. A signal is generated at the transmitter, passed through the channel and the
  received signal is analyzed. The obtained simulation results are discussed and
  compared with existing theory on wireless channels.

In the second part entitled: The Sensor Network: Topology and Physical Layer, the major
concern is the data acquired by the sensor nodes and in particular the corresponding data rates.
In this regard, two different types of transmission technologies are discussed. The Zigbee
standard for low data rates, and orthogonal frequency division multiplexing (OFDM) for
higher data rates, are discussed. Bluetooth is not considered because of the small range it
requires (2-10 m). This part is organized as follows:

- Chapter six gives a brief proposition for the network topology and the location of the
  sensor network nodes. The types of nodes are explained and some of the required data
  and corresponding data rates are presented.

- Chapter seven takes into consideration the ZigBee standard as communication
  networking candidates, for low data rate applications. A system using ZigBee (the
  ZNet platform), which has been tested in laboratory and in a real 380 KV open-air
  substation, is also presented.

- Chapter eight considers the use of OFDM at the physical layer to increase the data
  rates. The results of this chapter are based on a Matlab program implementing an
  OFDM transmitter and receiver.

The third and last part summarizes the findings, presents the overall conclusions and
recommendations and includes the list of abbreviations, the references and the appendix. It is
entitled: Wireless Sensor networks for Power Systems Monitoring and Control and organized
as follows:

- Chapter nine provides a summary of all chapters in terms of conclusions. Furthermore,
  some recommendations are given for future work.
PART I: DESCRIPTION AND MODELING OF THE WIRELESS CHANNEL

This first part will be devoted to the wireless channel and in particular the wireless channel inside a substation. The final goal is to derive a suitable model for this channel. In order to achieve that goal we will answer in subsequent chapters, the following questions. What type of channel can we expect? What are the channel properties and how do these affect the transmitted signals? What are the most important channel parameters that we can retrieve from those properties? What types of interference sources can we expect in such a channel and how do they affect the transmitted signal? How can we model this interference and add it to the channel model?

Chapter two will reveal the most important characteristics (properties) of the channel and the related channel parameters needed to derive the channel model. Also, noise added by the channel will be discussed within this chapter because it is considered as being a channel property. Interference in the channel and its effects on the transmitted signals will be the subjects of chapter three. A model for this interference to be added to the final wireless channel model is also derived in that chapter. Based on the information obtained in the two previous chapters, a model that is applicable to this particular case is given in chapter four. Chapter five implements and tests the derived channel model.
Chapter 2

Channel Characteristics and Properties

Contents
2.1 Causality and linearity
2.2 Time invariance
2.3 Multipath property
2.4 Attenuation property
2.5 Dispersion property
2.6 Noise as channel property
2.7 Others properties of the channel
2.8 Conclusions on channel characteristics and properties

This chapter presents the main characteristics and properties of a wireless channel in general, with the emphasis on substations environments. The purpose is to derive from each property, a channel parameter to build a reliable wireless channel model. This model will be used in the analysis of the wireless link between two nodes of the communication system.

2.1 Causality and linearity

We view the wireless channel as a system that responds to excitations caused by transmitted signals \( s(t) \), as depicted in Figure 2.1. The transmitted signals or input signals to the system are time varying electric currents or voltages that will generate electromagnetic waves. The response of the system to these input signals is also a time varying electric current or voltage, \( r(t) \). The causality property is given to a system when it is not able to anticipate on the future. This means that the response of the system is always determined by the past and present course of the excitation and never by its future course. We will consider a channel for which the causality property holds.

A medium is considered linear if at any particular point in this medium, different waves can be superposed. The wireless channel in our case is the propagation medium for electromagnetic waves in a substation environment. It consists mostly of air, insulators and metallic structures like transformers, circuit switchers/breakers and arresters, capacitors, wires and cables, insulators supports, etc. Figure 2.2 gives an impression of this propagation medium. For such a medium, the superposition principle holds for the response to any
transmitted (input) signals as described above. We therefore consider the channel as being linear. For the transmitted signals and their responses the following then holds:

Let \( r_i(t) = S[s_i(t)] \) denotes the response of system \( S \) to input \( s_i(t) \). For an input signal \( s(t) \) such that:

\[
s(t) = \sum_{i=1}^{n} A_i s_i(t) ,
\]

with \( n \in \mathbb{N} \) and \( A_i \) any complex constant, linearity means that the following relationship holds:

\[
r(t) = S\{s(t)\} = \sum_{i=1}^{n} A_i r_i(t)
\]

This means that the system or channel, responds to a linear combination of more than two input signals with a correspondingly linear combination of output signals [10].

![An open air substation environment.](image)

**Figure 2.2:** An open air substation environment. The picture shows that the propagation medium for electromagnetic waves in such an environment mostly consists of air, insulator materials and metallic structures.

### 2.2 Time invariance

In general, the properties of a wireless channel vary over time due to motion of transmitters/receivers and/or objects in the channel. These moving objects are reflectors in the transmission path giving rise to a varying number of multipath components. However, we assumed a quasi-static environment, featuring a fixed number \( N \) of multipath components, each with fixed path loss and shadowing. As a result of this, the wireless channel will be **time-invariant**. For the transmitted signals and their responses, the following holds:

Let \( r(t) = S\{s(t)\} \) denotes the response of system \( S \) to input \( s(t) \). Time-Invariance means that the following relationship holds:

\[
r(t - \tau) = S\{s(t - \tau)\}, \forall \tau \in \mathbb{R}
\]

That is the system or channel, responds to a delayed input signal with a correspondingly delayed output signal [10].
2.3 Multipath property

As stated in the previous section and illustrated by Figure 2.2, the wireless channel is almost always full of obstacles which act as reflectors in the transmission path between transmitter and receiver. The environment in a substation is thus favorable for reflections and thus generation of multipath components. As a result of this, a transmitted signal will not follow only one path to reach the receiver but a multitude of distinct paths. In other words, a transmitted single pulse results in a train pulse at the receiver with the number of received pulses depending on the time of transmission. Such a channel is called a multipath channel, and is illustrated in Figure 2.3. In the figure, \( s(t) \) and \( r(t) \) denote the transmitted and received signals, respectively. The number of multipath components is given by \( N \) and is a fixed number, for each link between two communication network nodes, as explained in section 2.2.

![Figure 2.3: Effects of multipath on a transmitted pulse. A single pulse at the transmitter results in a train pulse at the receiver. The train pulse consists of \( N \) multipath components, attenuated and delayed in time.](image)

Applying the superposition principle due to the linearity of the channel, the multipath components (multipath waves) will combine at the receiver to give a resultant signal which can vary significantly in amplitude and phase depending on the bandwidth of the original transmitted signal and the intensity and propagation time of the multipath waves [11].

### 2.3.1 Delay spread

The difference between the propagation time of the first arrived multipath component and the last one is called the maximum delay spread \( \tau_{\text{max}} \) of the channel. It is the time between the arrival of the first multipath component and the last one. In Figure 2.3 it is given by \( \tau_n - \tau_1 \). If the delay spread is small compared to the symbol duration \( T_{\text{symb}} \) of the transmitted signal, there will be little time spreading in the received signal and thus almost no signal distortion (see Figure 2.4 (a) and (b)). In the top figures, the transmitted signal (solid line) and its replicas (multipath components) are shown. The received signal is the sum of the multipath components and is plotted in the bottom figures. The received signal in Figure 2.4 (a) has the same shape as the transmitted signal. In Figure 2.4 (b) this is not the case. It is clear that the sum of the replicas in (b) results in a signal of different shape than the transmitted signal. As a consequence, the received signal is distorted. In terms of signal processing needed at the receiver, the maximum delay spread is a measure for the maximum symbol rate for which no equalization is needed.
The maximum delay spread $\tau_{\text{max}}$ is a good measure for the spreading caused by the wireless channel on the transmitted signals and, therefore, will be used, together with $\tau_{\text{mean}}$ and $\tau_{\text{rms}}$, as channel parameters. $\tau_{\text{mean}}$ and $\tau_{\text{rms}}$ are respectively the mean excess delay and RMS-delay spread. They are computed as follows:

$$\tau_{\text{mean}} = E[\tau] = \frac{\sum_{n=1}^{N} \tau_n P_c(\tau_n)}{\sum_{n=1}^{N} P_c(\tau_n)} \quad (2.4)$$

$$\tau_{\text{rms}} = \sqrt{E[\tau^2] - (E[\tau])^2} \quad (2.5)$$

where, $N$ is the number of multipath components and $P_c(\tau_n)$ the power delay profile\(^6\). It reflects the time distribution of the received signal power. $E[\tau^2]$ is the second moment\(^7\) of the random variable $\tau$, which represents the delays of the multipath components. Common values of RMS-delay spread found in literature [12] for indoor buildings are between 1470 ns and 3500 ns. Values around 30 $\mu$s are found for outdoor environments [11].

### 2.3.2 Coherence bandwidth

We have seen in the time domain that to avoid signal distortion, the maximum delay spread of the channel $\tau_{\text{max}}$ should be significantly less than the symbol duration $T_{\text{symb}}$ of the transmitted signals $\tau_{\text{max}} \ll T_{\text{symb}}$. This statement has an equivalent in the frequency domain, namely:

In order to avoid signal distortion, the transmitted signal bandwidth $B$ should be less than or equal to the coherence bandwidth $B_{\text{coh}}$ of the wireless channel. This is the bandwidth over

---

\(^6\) The power delay profile is often modeled as having a one-sided exponential distribution: $P_c(\tau_n) = \left(\frac{1}{\tau_{\text{mean}}}\right) e^{-\frac{\tau_n}{\tau_{\text{mean}}}}$.

\(^7\) The second moment of random variable $\tau$ is given by: $E[\tau^2] = \left(\sum_{n=1}^{N} \tau_n^2 P_c(\tau_n)\right) / \left(\sum_{n=1}^{N} P_c(\tau_n)\right)$.
which the frequency components fade in the same way (flat fading). In other words, if we denote the channel frequency response by \(|H(f)|\), the coherence bandwidth is the distance \(|f_1 - f_2|\) in frequency domain where \(|H(f_1)|\) is significantly correlated to \(|H(f_2)|\). Equation (2.6) gives the relationship between the coherence bandwidth and the RMS-delay spread.

\[
B_{coh} \approx \frac{1}{2\pi \tau_{rms}} \quad (2.6)
\]

Taking the value of \(\tau_{rms}\) obtained from literature [11] and applying equation (2.6), we find values between 0.04 - 0.1 MHz for the coherence bandwidth for indoor environments. For outdoor environments we can expect a value around 5 kHz.

### 2.4 Attenuation property

An electromagnetic wave is subject to *attenuation* when propagating through the wireless channel. There are many phenomena causing attenuation. In the first place, we can think of path loss. Secondly, shadow-fading due to obstacles in the environment can be mentioned. Lastly, reflection, transmission and absorption by objects in the channel can also contribute in reducing the received signal power, which is called multipath fading.

#### 2.4.1 Path loss

Path loss \(PL(d)\) is the expected attenuation as a function of distance. It is a constant value for a specific distance and is given by the following formula:

\[
PL(d) = PL(d_0, f) \left( \frac{d}{d_0} \right)^\gamma \quad (2.7)
\]

where \(PL(d_0, f)\) is the attenuation at reference distance \(d_0\). The reference distance \(d_0\) should be in the far-field and is usually taken to be equal to one meter in indoor environments. The exponent \(\gamma\) is the path loss exponent and is strongly dependent on the environment. In free-space, \(\gamma\) is 2 and, in factory environments, values between 1.6 and 3.3 are found. The path loss exponent can also be obtained via a minimum mean-square error (MMSE) fit to empirical data [11]. In dB, equation (2.7) becomes:

\[
PL(d)|_{dB} = PL(d_0, f)|_{dB} + 10\gamma\log\left( \frac{d}{d_0} \right) \quad (2.8)
\]

The attenuation at reference distance \(PL(d_0, f)\) depends on the antenna characteristics as given by equation (2.9):

\[
PL(d_0, f) = \left( \frac{4\pi d_0}{\lambda} \right)^2 \cdot \left( \frac{1}{G_{TX}} \right) \cdot \left( \frac{1}{G_{RX}} \right) \quad (2.9)
\]

where \(G_{TX}\) and \(G_{RX}\) are respectively the gains of the transmitting and receiving antennas and \(\lambda = \frac{c}{f_c}\) is the wavelength, with \(c = 3.10^8\) m.s\(^{-1}\). For the sake of simplicity, the antenna gains
are both taken equal to one as an approximation for dipole antennas. \( f_c \) is the operating frequency in Hz \[13\]. In dB, equation (2.9) becomes:

\[
PL(d_0, f)\big|_{dB} = 10\log_{10}\left(\frac{PL(d_0, f)}{1}\right)
\]  

(2.10)

### 2.4.2 Shadow-fading

Shadow-fading accounts for the fluctuations of signal strength around the path loss constant value. Thus, if we measure the received power at all locations at a distance \( d \) from the transmitter – that means on a sphere of radius \( d \) with the transmitter at its center – we will obtain different values which are close to the path loss value. An illustration of this is given in Figure 2.5. The received power is plotted for locations around a circle of radius \( d \) with the transmitter at the center \[12\].

![Figure 2.5: Effects of shadow-fading on received signal power. For a transmitter located at the center of the grey area, the received power at receivers around the area all at distance \( d \) from the transmitter is not always the same.](image)

Those fluctuations are due to obstacles in the channel, causing scattering, diffraction or complete shadowing of the transmitted signal. The variations on received signal power, at a specific distance from the transmitter, are random for all possible directions of signal propagation. They are modeled as a Gaussian random variable \( X_{\text{shadow}} \) with a mean of 0 dB and a standard deviation \( \sigma_s \) in dB.

\[
X_{\text{shadow}} = N(0, \sigma_s) \quad [\text{dB}]
\]  

(2.11)

The standard deviation of the shadow fading is environment dependent and has to be determined from measurements. In indoor environments, a value between 2 and 4 dB is almost always found (\( 2\text{dB} \leq \sigma_s \leq 4\text{dB} \)). In outdoor environments values up to 12 dB can be encountered \[12\].
2.4.3 Multipath fading

Multipath fading is the short distance fluctuation of the signal attenuation due to multipath components of the transmitted signal reaching the receiver at the same time. These multipath components differ in amplitude, phase and arrival time. As a result of this, they will add destructively or constructively at the receiver. The fluctuations are random and modeled as a random variable $X_{\text{multipath}}$ with a mean 0 and a standard deviation $\sigma_m$ (see equation (2.12)). The variations are around the loss consisting of path-loss with addition of shadow fading in linear value as given by equation (2.13).

$$X_{\text{multipath}} = N(0, \sigma_m)$$  \hfill (2.12)

The total loss (attenuation) $L(d)$ at distance $d$ can now be given by the following formula:

$$L(d) = PL(d_0, f) \left( \frac{d}{d_0} \right)^\gamma \cdot 10^{\frac{X_{\text{shadow}}}{10}} + X_{\text{multipath}}$$  \hfill (2.13)

2.5 Dispersion property

As explained in section 2.3, the effect of multipath generates copies of the transmitted signal. These time-delayed signal replicas arrive at the receiver at distinct times. These distinct arrival times or delays will result in distortion of the transmitted signal, due to time or frequency dispersion.

2.5.1 Time dispersion

A channel is considered time dispersive or frequency selective if the multipath delays are distinguishable relative to the symbol period or inverse signal bandwidth. Thus, whether or not our channel will be considered frequency selective depends on the transmitted data as depicted in Figure 2.6.

![Figure 2.6](image)

**Figure 2.6:** The frequency selectivity property of the channel appears when the signal bandwidth is much greater than the channel coherence bandwidth. The signal with bandwidth $B_{\text{narrowband}}$ undergoes flat fading, while the one with bandwidth $B_{\text{wideband}}$ will have some frequencies much more attenuated than others.
The frequency response of the channel has been plotted and the frequency selectivity property of the channel can be seen by considering a wideband and a narrowband signal. While the wideband signal frequency components are not equally amplified, the narrowband signal undergoes flat fading, meaning that its frequency components are all attenuated in the same way. The wireless channel acts like a filter, attenuating certain frequencies while amplifying others. Recalling the definition of the coherence bandwidth in section 2.3.2 and again denoting the channel frequency response by $|H(f)|$, we can infer that a narrowband signal of bandwidth $B_n = [f_{\text{begin}}, f_{\text{end}}]$, does not require equalization at the receiver as $|H(f_1)| \approx |H(f_2)|$ for any $f_1, f_2 \in B_n$.

2.5.2 Frequency dispersion

Doppler effects, due to movements of the transmitter/receiver or the environment, cause short-term fluctuations of the received signal in the time domain. A channel is termed frequency dispersive or time selective when the Doppler shift is comparable to the signal bandwidth [14]. In our case, the channel is considered quasi-static. The transmitters/receivers and the environment are not moving. The few possibilities that motion can be introduced into the channel are people moving in the substation for maintenance or to visit. But this is not often the case and in addition the velocities involved are negligible compare to the speed of the signals. As a result of this, the channel will be considered time invariant and therefore, not frequency dispersive.

2.6 Noise as channel property

In practice, noise will always corrupt the transmitted signal passing through the channel. Therefore, the channel is considered to produce noise. The purpose of this section is to derive a suitable basic model for this noise, which represents thermal and man-made noise at the receiver. The model is based on three assumptions. The first is that the noise is additive, meaning that the received signal $r(t)$ is obtained by adding the noise $n(t)$ to the transmitted signal $s(t)$, where the noise is considered statistically independent of the transmitted signal.

$$r(t) = s(t) + n(t)$$  \hspace{1cm} (2.15)

The second assumption is that the autocorrelation of the noise in the time domain $R_n(\tau)$ is zero for all non-zero time differences $\tau$. In other words, the Power Spectral Density (PSD) of the noise $P_n(f)$ is flat and the noise is termed white. We then have:

$$R_n(\tau) = P_n(f) \delta(\tau)$$  \hspace{1cm} (2.16)

And,

$$P_n(f) = \frac{N_0}{2} \quad [W/Hz]$$  \hspace{1cm} (2.17)

where, $\delta(\tau)$ is the Dirac delta function or unit impulse. $N_0$ is a positive constant. For thermal resistor noise, $N_0$ equals $kT_0$, with $k$ being Boltzmann’s constant and $T_0$ is the absolute temperature. Typically, $N_0$ will be a value slightly above -170 dBm/Hz. Finally, we also
assume that the noise samples have a Gaussian distribution. The noise added by the channel will thus be modeled as a Gaussian random process $n(t)$ with mean 0 and a constant PSD over all frequencies [15]. This random process is called a White Gaussian Noise process and can be represented using the complex lowpass representation as follows:

$$n(t) = \text{Re}\{n_l(t)e^{j2\pi f_0 t}\},$$

where, $n_l(t)$, defines as $n_l(t) + jn_q(t)$, is a complex lowpass Gaussian process with $n_l(t)$ and $n_q(t)$ being two independent, zero-mean, real lowpass Gaussian processes with PSD of $N_0$. Finding the appropriate value of $N_0$ for a substation environment will completely characterize the noise. Therefore, $N_0$ is the only channel parameter accounting for the noise.

2.7. Others properties of the channel

There are of course many other properties to be considered when describing a wireless channel as propagation medium. We can think of uniformity and homogeneity which are related to changes of channel properties as function of space. A channel can also be termed isotropic when direction is to be taken into consideration. Furthermore, the term relaxation can be used to describe a medium when it shows some buffering or “inertia” property. However, to avoid dealing with a very complex model, and likewise to reduce the number of channel parameters, we will limit our self to the most important parameters as mentioned in previous sections.

2.8. Conclusions on channel characteristics and properties

The channel parameters are the parameters which we will use to characterize, model and simulate the wireless channel. From the preceding sections, we are dealing with a causal, linear, time-invariant, attenuating, frequency selective and noisy multipath channel. In the following, we list the above mentioned channel characteristics discussed in previous sections and briefly summarize the parameters they will bring in our channel model.

- The causality and linearity properties will reveal themselves, when we will derive an expression for the received signal in section 4.2. We can think of the parameters associated with these two properties as being implicitly integrated into the final model.

- From the time-invariant property, we will use the parameter $N$ which denotes the fixed number of multipath components between a transmitter and a receiver.

- The distance $d$, the path loss exponent $\gamma$ and the standard deviation of shadow fading $\sigma_s$ all account for the attenuating property.

- From the frequency selective or time dispersive property, we retain the coherence bandwidth of the channel $B_{coh}$.

- The noisy property is fully captured by determining the value of the noise power spectral density $N_0$ as explained in section 2.6.
The multipath property seems to bring a lot of parameters but they are actually linked together and give the same information. We will retain the maximum delay spread $\tau_{\text{max}}$, the RMS-delay spread $\tau_{\text{rms}}$ and the excess mean delay $\tau_{\text{mean}}$.

All the channel parameters with their values are listed again in Table 2.1 together with the channel property for which they account. The values used for the simulations are between brackets.

**Table 2.1:** The wireless channel parameters used for the derivation and implementation of the wireless channel model. The values not specified are computed from the other specified values. Values between brackets are used for the simulations.

<table>
<thead>
<tr>
<th>PROPERTY</th>
<th>PARAMETER</th>
<th>EXPLANATION</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>Causality</td>
<td>IMPLICIT</td>
<td>Introduction of delays in received signal.</td>
<td>No value</td>
</tr>
<tr>
<td>Linearity</td>
<td>IMPLICIT</td>
<td>Superposition or addition of multipath components.</td>
<td>No value</td>
</tr>
<tr>
<td>Time-Invariance</td>
<td>$N$</td>
<td>Number of multipath components.</td>
<td>$&gt;20$ (20)</td>
</tr>
<tr>
<td>Attenuation</td>
<td>$d$</td>
<td>Distance between transmitter and receiver.</td>
<td>20 - 400 m (20)</td>
</tr>
<tr>
<td></td>
<td>$\gamma$</td>
<td>Path loss exponent.</td>
<td>1.6 – 3 (3)</td>
</tr>
<tr>
<td></td>
<td>$\sigma_s$</td>
<td>Standard deviation of shadow fading.</td>
<td>4 - 8 dB (8)</td>
</tr>
<tr>
<td>Multipath</td>
<td>$\tau_{\text{max}}$</td>
<td>Maximum delay spread.</td>
<td>50 ns - 30 $\mu$s (1 $\mu$s)</td>
</tr>
<tr>
<td></td>
<td>$\tau_{\text{rms}}$</td>
<td>RMS-delay spread.</td>
<td>-</td>
</tr>
<tr>
<td></td>
<td>$\tau_{\text{mean}}$</td>
<td>Mean excess delay.</td>
<td>-</td>
</tr>
<tr>
<td>Frequency selectivity</td>
<td>$B_{\text{coh}}$</td>
<td>Coherence bandwidth of the channel.</td>
<td>$1/2\pi\tau_{\text{rms}}$</td>
</tr>
<tr>
<td>Noise</td>
<td>$N_0$</td>
<td>Noise power spectral density.</td>
<td>-170 dbm/Hz</td>
</tr>
</tbody>
</table>
Chapter 3

Interference Analysis and Modeling

Interferences of various kinds can deteriorate the performance of a communication system and corrupt the desired transmitted signal. Therefore, they are a major concern in the design and operation of communication systems. In order to prevent them causing severe signal distortion and dramatically increasing the bit error rate (BER), an understanding of their origins, properties and effects on transmitted electromagnetic waves is of the utmost importance. This chapter will be completely devoted to the main sources of interference that we can encounter in a substation and the types of interference that these sources generate. We will thus take a look at the electromagnetic environment inside a substation as shown in Figure 3.1.

The final objective is to find a mathematical expression which captures the effects of each type of interference and therefore allow their insertion in the channel model derived in next chapter. This will complete the final model of the channel and the final expression of the received signal.

Figure 3.1: An impression of a substation electromagnetic environment. Electromagnetic radiation from interference sources as power lines, transformers, switching devices and lightning strokes contribute to the pollution of the wireless channel.
3.1 EMC and EMI concepts

The electromagnetic environment accounts for all electromagnetic phenomena or events present at a particular location. It thus consists of all electrical signals or electromagnetic emissions that unintentionally leaks out of conductors. Electromagnetic emission can be classified as continuous or transient [16] and conducted or radiated [17], [18].

Electromagnetic compatibility (EMC) is defined as the ability of an electrical device, circuit or system to function properly in its electromagnetic environment without being an interference source for other electrical devices, circuits or systems. It should be noted here that our device, circuit or system, is an information carrying electromagnetic wave propagating through a substation. This is the transmitted signal \( s(t) \) between two sensor nodes inside the substation, and will be referred to as the interference “victim” in the following. The influence of the electromagnetic environment on the interference victim is what we call electromagnetic interference (EMI). The EMI-problem is depicted in Figure 3.2 and shows that a coupling path is required to make the interference victim suffer from its electromagnetic environment [19].

![Figure 3.2: Basic form of the EMI-problem. The interference victim (IV) is the transmitted signal between two sensor nodes. This wave will interfere, via radiated coupling, with all other electromagnetic emissions emanating from the interference sources (IS).](image)

There are two types of coupling paths or coupling mechanisms: conducted coupling or radiated coupling [20]. Keeping in mind the type of interference victim we have, namely the transmitted signal between two sensor nodes, we will only be concerned with radiated coupling. Looking again at Figure 3.2 we can infer that in order to prevent EMI from taking place, there are three options or actions to be considered, as stated below [19]:

1. Action at the source: Stop the interference source from emitting interference.

2. Action at the link between source and victim: Eliminate the coupling paths to prevent the interference from coupling into the interference victim.

3. Action at the victim: Increase the interference immunity level of the interference victim.

From the above mentioned options, only the third one is relevant in our case. Shutting down transformers or de-energizing power transmission lines to prevent them from contributing to
the pollution of the electromagnetic environment will interrupt power delivery\(^8\). Likewise, eliminating the coupling paths, which translates into preventing electromagnetic waves to interfere is not a feasible solution either. Interference of two or more waves resulting in a new wave pattern is a physical law based on the fact that the electric field cannot have two different values and directions at a same point in space. The same holds for the magnetic field. As a result of this, when two or more EM-waves meet in space they superpose (interfere). The third option will be the remaining option. The immunity level of the transmitted signal will be increased by an appropriate choice of transmission technology. This choice will depend on the types of interferences present in the channel.

### 3.2 Interference sources in substation’s wireless channel

As mentioned in the previous section, we are only interested in continuous/transient radiated coupling or electromagnetic radiation from the main interference sources. When the interference victim is located in the far-field of an interference source, the radiated coupling will only be via electromagnetic waves with a fixed ratio of electric and magnetic field strength. In close proximity of the source (near-field) the coupling may either be a predominantly electric or magnetic field [20]. Thus, according to the location of source and victim, we will have to deal with static or quasi-static fields and dynamic fields. Furthermore, this radiated emission can be categorized into narrowband and broadband, in-band and out-of-band, when we compare the frequency contents of source and victim. Based on the types of electromagnetic emissions stated above, we can now classify our interference sources as depicted in Figure 3.3.

![Figure 3.3](image)

\textit{Figure 3.3:} A classification of electromagnetic interference sources, as derived from different scientific papers: [16], [17], [19] and [21]. Narrowband and broadband are with respect to the transmitted signal bandwidth. The same holds for in/out-of-band.

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\(^8\) Shielding the components is not considered here as a solution because not all components in nowadays substations are shielded. Even though some components as cables are shielded, EMI may still results from faulty screens.
The term dynamic in the figure means space and time-varying. These space and time-varying electromagnetic fields (EM-fields) are electromagnetic waves obeying the wave equation as obtained from Maxwell’s equations [22]. We will now present the main interference sources in a substation and classify the type of interference they generate according to Figure 3.3.

### 3.2.1 EM-fields around substation’s equipment

The power system can be subdivided into two main systems, namely the primary system for generation and transport of electrical energy and the secondary system which consists of all the equipments and circuits to measure, control and supervise the electrical energy flow into the primary system [19]. This equipment is found in substations. We can think, for example, of transformers, breakers/arresters and high voltage cables. As soon as there will be a potential difference between parts of those components or a current flow through them, they will produce electric and magnetic fields. Because of the slowly varying nature of these fields, usually varying with extremely low frequencies around the power frequency of 50/60 Hz, they are classified as static or quasi-static, man-made, radiated emission.

### 3.2.2 Switching actions

In high voltage substations, switching actions are needed to protect the equipment in case of faults and to control the energy flow. Think of the closing or opening of circuit breakers to switch electric circuits and equipment in and out of the system or the operation of a surge/lightning arrester to protect the electrical system and substation components against lightning strokes and switching surges. Not to forget the interruption of fault current (fault clearing) by circuit breakers [23]. The making and braking of current by above mentioned high voltage switchgears is an important cause of very high frequency electromagnetic interference. This interference results from transient bus currents and associated EM-fields of significant amplitudes and frequency content. Bus currents of 5 to 20 $\mu$s duration with peak amplitudes of 400 A and frequencies up to 100 MHz have been measured. The associated electric and magnetic fields have peak strengths of several tens of kV/m and A/m [24].

![Figure 3.4: Macroburst in 115 kV AIS with associated micropulses. The radiated emission can have significant frequency components up to 100 MHz [25].](image)

According to [25], measured electric and magnetic fields resulting from switching transients are characterized by their macroburst and micropulse properties. An example of macroburst and associated micropulses is shown in Figure 3.4. A macroburst is defined as the totality of all arcs (transients) produced during a single switching operation. The number of arcs or transients present in the macroburst, their amplitudes with respect to one another, their frequency of occurrence and the total duration of the macroburst allow for a complete characterization of the switching action.
A single arc or individual transient in a macroburst is called a micropulse. Micropulses are characterized by their duration, amplitude and wave shape. Values of those parameters in air insulated substations (AIS) and gas insulated substations (GIS) for different operation voltages and types of breakers are also reported in [25].

To summarize, EMI caused by switching actions is strongly dependent on the type of substation, the operating voltage, the type of breakers and their speed, the electrical characteristics of the bus being excited and the location of the receiver. This will result in a wide range of values for the parameters characterizing those transients as mentioned above. Nevertheless as we are trying to guarantee an error free robust data transfer between two nodes we can only consider the most severe type of switching action interference. The number of micropulses can be as high as 10000. The faster the switch, the lower this number will be. The total duration of the macroburst can last for two seconds\(^9\) and the micro pulse repetition frequency (PRF) can reach 40 kHz. Furthermore, the dominant frequencies can be quite high: 30 MHz – 80 MHz, and even 120 MHz [25].

By choosing a transmission technology operating in the 2.4 GHz range, this type of interference can be classified as transient, broadband and out-of-band man-made radiated emission.

### 3.2.3 Partial discharges and corona

Partial discharge (PD) is a generic name used to refer to any electrical discharges taking place in any electrical insulation under voltage pressure. The discharges are partial because they do not completely bridge the distance between the electrodes [21]. These discharges result from accelerated ionization due to the electric field in the insulation generated by the applied voltage. The resulting PD pulse currents generate unwanted electromagnetic radiation, whose spectra depend on the physical conditions\(^{10}\) at the place where the PD occurs. Measured spectra, for different types of defects, as reported in [26], show maximum frequencies between 1.5 and 2 GHz. However, the radiating structure at the defect place in the HV component can significantly attenuate the PD signals [26].

Corona is a special type of partial discharge, occurring around electrodes in gas or liquid, and can be a troublesome source of interference. It is considered as being a broadband type of interference [16]. In [27], the reference frequency for corona is 0.5 MHz.

In order to determine the spectrum of corona in our region of interest, measurements were performed by using two types of broadband antennas and a narrowband antenna connected to a spectrum analyzer. The corona was generated by mounting a piece of copper wire on a high-voltage electrode. The voltage on the high-voltage electrode could be regulated and was increased (up to 45 kV) until the electric field enhancement caused by the copper wire caused the air to ionize. The radiated emission caused by corona was then picked up by one of the antenna and its spectrum was made visible on the spectrum analyzer. As the used antennas are sensitive for radiation in a certain frequency band, the spectrum was measured only in that frequency band, namely from 500 MHz to 2.9 GHz. The antennas and therefore the frequency band of measurements were chosen to look at the spectrum around the 2.4 GHz

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\(^9\) For hand-operated disconnect, the macroburst duration is in the range of 0.04 – 2 seconds. For motor-operated disconnects, which are mostly applied today, a macroburst can last for approximately 0.6 seconds [C. Wiggins, 1991].

\(^{10}\) The physical conditions includes physical properties of the surfaces at the defect place, temperature, pressure, electric field strength, and spatial distribution of charges [26].
operating frequency of our communication system. The separation distance was varied from 0.5 to 2.5 m. Figure 3.5 shows the results. According to the figure, corona will appear as noise in the GHz range.

![Figure 3.5: Noise, Transmitted signal and Corona spectra in frequency domain of interest as measured by three different types of antennas. Top plot: dielectric wedge antenna. Middle plot: Bow-tie antenna. Bottom plot: monopole antenna. In all three measurements only the transmitted signal was picked up by the antennas. Voltage: 45 kV and measured electric field strength: 10 kV.m$^{-1}$ at 0.75 m from the corona source.](image)

Corona does not appear in the frequency band of measurements. This type of interference will thus be classified as man-made, dynamic, continuous, broadband and out-of-band type of interference. However, the use of a transformer with faulted windings shows more interesting results in the frequency domain. The radiated electromagnetic emission can be due to partial discharges occurring between the transformer windings, due to defective insulation. The generated electromagnetic emission has frequency components up to the GHz range. During the measurements, the voltage is increased step by step to investigate the effects of voltage level on the electromagnetic emission. The measurements results show that an increase in voltage results in an increase in the magnitude of the spectrum of the electromagnetic
emission. Figure 3.6 shows the results. In addition, the spectrum is becoming wider, as higher frequency components of lower amplitudes appear in the spectrum (see Figure 3.6, bottom plot).

![Transmitted signal and Discharge in frequency domain at 30kV](image)

![Transmitted signal and Discharge in frequency domain at 40kV](image)

**Figure 3.6:** Transmitted signal and Discharge spectra with voltage increased from 30 kV (top plot) to 40 kV (bottom plot). The increase in voltage results in appearance of frequency components of non negligible amplitude in the frequency band of operation of the transmitted signal.

### 3.2.4 Lightning

Lightning is a form of electrical breakdown of air over extreme distances. It is a serious type of interference. Lightning will generate currents of varying shapes. However, standardization defines the so-called 1/50 \( \mu \)s wave. It is a wave with a front of 1 \( \mu \)s and a time to half-value of 50 \( \mu \)s [28]. If we rely on the fact that the source current determines the spectrum of the wave and vice versa, we can infer that lightning will generate electromagnetic emission with significant frequency content up to the MHz range. It is a natural type of radiated emission, classified as broad-band and out-of-band interference.

### 3.2.5 Radio frequency emissions and other communication signals

In this section, we will consider all signal generated by communication equipment. Interference is strongly determined by transmission frequencies. Radio frequency emissions are not considered harmful because the frequencies used, are much lower than 2.4 GHz. Therefore, all out-of-band communication signals can be disregarded. However there are communication signals in the same frequency range as the transmitted signal. These are coming from reflected surfaces present in the channel. The substation environment is full of metal surfaces as most of the components are mounted on metallic frames. The transmitted signal will arrived in multiple replicas at the receiver at different time instances due to reflections on those metal surfaces. This was explained in section 2.3 and referred to as multipath. The frequency content of those interfering signals is the same as the frequency content of the source signal. It is thus not possible to filter those signals out at the receiver.
This is the reason why this type of interference is considered as the most dangerous and therefore also the deterministic type (deciding factor) for the choice of transmission technology. This interference is thus classified as man-made, dynamic, continuous, broadband/narrowband and in-band.

3.3 Modeling the interference

3.3.1 The simplified general wave expression

After having classified the different types of interference present in the channel, a mathematical expression for each type of interference will be derive. This mathematical expression will be used in our final model of the wireless channel.

Any wave, thus also EM-waves can be expressed in complex notation, as a linear combination of sinusoidal waves as given by the following equation [22]:

\[ \hat{f}(z,t) = \int_{k=-\infty}^{\infty} \hat{A}(k)e^{j(kz-\omega t)} dk \]  

(3.1)

where, \( k \), the wave number, is equal to \( \frac{2\pi}{\lambda} \) with \( \lambda \) the wavelength. The negative values of \( k \) account for waves travelling in opposite direction as compared to those with a positive \( k \) value. \( \omega \) the angular frequency is equal to \( kv \) with \( v \) the velocity. The coefficients \( \hat{A}(k) \) can be obtained from the theory of Fourier transforms in terms of the initial conditions \( f(z,0) \) and \( f'(z,0) = df(z,0)/dz \) as follows:

\[ \hat{A}(k) = \frac{1}{2\pi} \int_{-\infty}^{\infty} [f(z,0) + \frac{j}{\omega} f'(z,0)]e^{-jkz}dz \]  

(3.2)

By expressing \( k \) in terms of \( \omega \), we find:

\[ k = \frac{\omega}{v} \]  

(3.3)

The velocity \( v \) in equation (3.3) is a positive constant. Combining the two equations, (3.1) and (3.3) we can rewrite (3.1) in terms of \( \omega \) and express any wave as a sum of sinusoidal waves of different frequencies and corresponding complex amplitudes. Those amplitudes are the coefficients in front of the exponential term in equation (3.1) and are frequency dependent. This brings us to the conclusion that two waves interfering with one another will result in a new wave with a frequency spectrum containing all the frequencies present in the two waves, no more, no less. If the spectra of the two waves do not overlap the resulting spectrum contains the two spectra with no change in amplitudes. However, in the case that they do overlap, the coefficients corresponding to the frequencies belonging to the overlap region will change. This is visualized in Figure 3.7 where we are considering two waves \( f_1(t) \) and \( f_2(t) \) with frequency spectrum \( B_1 \) and \( B_2 \) respectively and a fixed location \( z_0 \) in space. Their time-domain expressions are given by equations (3.4) and (3.5) respectively, assuming a discrete\(^{11}\) frequency set.

\(^{11}\) By using a discrete set of frequencies, we replace the integral by a sum.
In the top of Figure 3.7 (a), $B_1$ and $B_2$ do not overlap. By using the superposition principle to write the expression of the resulting wave in the time domain and choosing for simplicity reasons a value of zero for $z_0$, the resulting wave can then be written as follows:

$$\tilde{f}_1(t) = \sum_{\omega \in B_1} \tilde{A}_1(\omega)e^{j(kz_0-\omega t)}$$

(3.4)

$$\tilde{f}_2(t) = \sum_{\omega \in B_2} \tilde{A}_2(\omega)e^{j(kz_0-\omega t)}$$

(3.5)

The spectrum of the resultant wave is shown in the bottom of Figure 3.7 (a). The interference of the two waves is not harmful in this case as we can still distinguish their spectra and therefore differentiate the two waves at the receiver.

In Figure 3.7 (b), there is an overlap region ($B_1 \cap B_2$). To give an expression for the resultant wave, we need to rewrite equations (3.4) and (3.5) as follows:

$$\tilde{f}_1(t) = \tilde{f}_1(t) + \tilde{f}_2(t) = \sum_{\omega \in B_1} \tilde{A}_1(\omega)e^{j(kz_0-\omega t)} + \sum_{\omega \in B_2} \tilde{A}_2(\omega)e^{j(kz_0-\omega t)}$$

(3.6)

and

$$\tilde{f}_2(t) = \sum_{\omega \in B_1 \cap B_2} \tilde{A}_2(\omega)e^{j(kz_0-\omega t)} = \sum_{\omega \in B_2} \tilde{A}_2(\omega)e^{j(kz_0-\omega t)}$$

(3.7)

The sets $B'_1$ and $B'_2$ are respectively the relative complements of $B_2$ and $B_1$ in $B_1$ and $B_2$ respectively [45]. Again, we took $z_0 = 0$, for simplicity. The resultant wave then becomes:

$$\tilde{f}_r(t) = \sum_{\omega \in B'_1} \tilde{A}_1(\omega)e^{j(\omega t)} + \sum_{\omega \in B'_2} \tilde{A}_2(\omega)e^{j(\omega t)} + \sum_{\omega \in B_2} \tilde{A}_2(\omega)e^{j(\omega t)}$$

(3.10)
Notice that the coefficients for the frequencies belonging to $B_1 \cap B_2$ are not known anymore.

Equations (3.4), (3.5), (3.7) and (3.10) can be simplified by looking at the spectra as being centered at a specific center or carrier frequency $\omega_c$, as indicated by the plots of Figure 3.7. An EM-wave can then be mathematically model in complex notation as follows:

$$\tilde{f}(t) = \left( \sum_{\omega \in B} \tilde{A}_1(\omega)e^{j\omega t} \right) e^{j\omega_c t} \quad (3.11)$$

This is the expression for the complex wave function at a fixed location in space. The term between brackets represent the baseband equivalent of the actual wave. It’s a wave with the same amplitude spectrum but concentrated near $\omega = 0$. The real wave function at that specific location is then given by equation (3.12):

$$f(t) = \text{Re}\left\{ \left( \sum_{\omega \in B} \tilde{A}_1(\omega)e^{j\omega t} \right) e^{j\omega_c t} \right\} \quad (3.12)$$

Simulation results obtained by the use of simulation software Matlab, and reported in section 3.4, confirm the derived expressions and the explicative plots of Figure 3.7. In the following, we used the derived expression of equation (3.12) to model the main interference sources present in a substation.

3.3.2 Wave expression for EM-fields around substation’s equipment

The first type of interference is of the quasi-static type. The carrier frequency $\omega_{c1}$ is zero and the frequency range $B_1$ is located at extremely low frequencies ($\omega \in B_1 \approx 0$). This type of interference results in the addition of a DC-component or extremely low varying component\(^{12}\) to the transmitted signal at the receiver. The mathematical expression is given by equation (3.13). It can be considered as being a constant.

$$I_1(t) = \text{Re}\left\{ \left( \sum_{\omega \in B_1} \tilde{A}_1(\omega)e^{j\omega t} \right) e^{j\omega_{c1} t} \right\} = I_1 \quad (3.13)$$

3.3.3 Wave expression for switching actions

The second type of interference has significant frequencies up to the MHz range of the frequency spectrum.

$$I_2(t) = \text{Re}\left\{ \left( \sum_{\omega \in B_2} \tilde{A}_2(\omega)e^{j\omega t} \right) e^{j\omega_{c2} t} \right\} = \text{Re}\left\{ \tilde{I}_2 e^{j\omega_{c2} t} \right\} \quad (3.14)$$

3.3.4 Wave expression for corona

The third type of interference will consist of a part in the MHz range and a non overlapping part in the GHz range.

---
\(^{12}\) This is compared to the transmitted signal operating around 2.4 GHz.
\[ I_3(t) = \text{Re} \left\{ \sum_{\omega \in B_3} \tilde{A}_3(\omega) e^{j\omega t} \right\} + \text{Re} \left\{ \sum_{\omega \in B_3} \tilde{A}_3(\omega) e^{j\omega t} \right\} = \text{Re} \left\{ I_3 e^{j\omega t} + I_3 e^{j\omega t} \right\} \]  (3.15)

The first term will be filtered out while the second will just be noise in the channel.

### 3.3.5 Wave expression for lightning

Interference due to lightning has not been thoroughly investigated. If we rely on the standard model we can however write a similar expression as for switching phenomena:

\[ I_4(t) = \text{Re} \left\{ \sum_{\omega \in B_4} \tilde{A}_4(\omega) e^{j\omega t} \right\} \]  (3.16)

The significant frequencies being located in the MHz range of the spectrum.

### 3.3.6 Wave expression for communication signals

The last type of interference can be subdivided into out-of-band and in-band interference signals as mentioned in section 3.2.5. Only the signal replicas are considered here as they have the same carrier frequency as the transmitted signal. Their baseband equivalents are time-delayed and attenuated versions of the baseband equivalent of the transmitted signal. We then can write:

\[ I_5(t) = \text{Re} \left\{ \sum_{\omega \in B_5} \tilde{A}_5(\omega) e^{j\omega t} \right\} = \text{Re} \left\{ (\alpha g(t - \tau)) e^{j\omega t} \right\} \]  (3.17)

The total interference is a summation of all the different types present in the channel. In the following chapter where we model the channel and derive an expression for the received signal, the interference will thus be given by the following equation:

\[ I(t) = I_1(t) + I_2(t) + I_3(t) + I_4(t) + I_5(t) \]  (3.18)

However, when building a model for this interference to include in the wireless channel model for simulations with Matlab, we will assume that a band-pass filter is used at the receiver and only take into considerations the in-band interference.

### 3.4 Matlab simulations on interference

In this section, we describe the simulation results confirming the equations of section 3.3 of this chapter, on modeling of the different types of interference. We first define a 3-dimensional (3-D) computational space box and we compute the electric field strength at all points of interest in our space as function of time. The Matlab code used for the simulations in this section can be found in Appendix B.
3.4.1 The implemented wave equation

To generate an electromagnetic wave for the simulations we use equation (3.19) [29]. This equation has been derived from Maxwell’s equations and gives the magnitude of the 3 components $E_x$, $E_y$ and $E_z$ of the electric field in space and time. The radiating structure is assumed to be a dipole antenna.

$$
E(\vec{x},t) = \int_{-\infty}^{t_\text{ret}} I(\tau) d\tau \frac{1}{4\pi \varepsilon_0 |\vec{x} - \vec{x}_\text{r}|^2} (3Q - I)P + I(t'\text{ret}) \frac{1}{4\pi c \varepsilon_0 |\vec{x} - \vec{x}_\text{r}|^2} (3Q - I)P
$$

$$
+ \partial_\tau I(t'\text{ret}) \frac{1}{4\pi c^2 \varepsilon_0 |\vec{x} - \vec{x}_\text{r}|} (Q - I)P
$$

(3.19)

where, $E(\vec{x},t) = [E_x(\vec{x},t) \ E_y(\vec{x},t) \ E_z(\vec{x},t)]^T$ represents the 3 components of the E-field at all specified locations in the 3D-computational space box for all time instants of interest. The matrix $Q = \Theta \Theta^T$ with $\Theta = (\vec{x} - \vec{x}_\text{r})/|\vec{x} - \vec{x}_\text{r}|$, is a unit vector in the direction of observation. $I$ and $P$ are the identity matrix and polarization vector respectively.

For the simulations, the polarization vector is chosen to ensure that the electric field is directed along the z-axis. The source current $I(t)$, can be the transmitted signal $s(t)$ representing the information, a delayed version of it $s(t-\tau)$ or an interference current generating the unwanted electromagnetic emission. This source current should be specified as well as its time derivative evaluated at $t'\text{ret}$ (the retarded time) and its integral from minus infinity to $t'\text{ret}$ [29]. An impression of the computed electric field strength as function of place and time is given in Figure 3.8. In the figure, two snapshots of the propagating wave at two different time instances are shown. This scenario represents an interference free space, where the only signal present is the wanted source signal, emanating from the bottom-right corner of the computational box where the transmitter is located and propagating to the left.

![Figure 3.8: Wave propagation in 3-D computational space at two different time instants. The wave emanates from the right corner (transmitter location) and propagates to the left.](image-url)
In Figure 3.9 the source current $i_s(t)$, interference current $i_i(t)$ and electric field at the receiver $E_r(t)$ are shown as well as their spectra. As there is no interference signal, the resultant wave and therefore the electric field at the receiver is only due to the source current. The frequency spectrum of this electric field is located in the same frequency range as the one characterizing the source signal. The bottom-most plot in this figure shows the electric field as function of time at two different locations from the source. The dotted line is at a larger distance than the continuous curve and shows a decrease in amplitude as expected.

To confirm the story of section 3.3 we will now generate two waves. One will be seen as the signal to be transmitted while the other will be the interference signal. We will then take a look at their original spectra and the spectrum of the resultant wave.

### 3.4.2 Resulting wave spectra for different interference signals

**A. Interference and source signals have overlapping spectra**

In this scenario, the interference signal is a time-delayed replica of the transmitted signal. We then assume that a reflected surface is present in the channel and created a second propagation path between the transmitter and the receiver. The two signals which overlap
the frequency domain will interfere in space and therefore at the receiver. Figure 3.10 gives an impression of the signals propagation in the 3D-computational space box.

Figure 3.10: Wave propagation in 3-D computational space at two different time instants, for two waves of overlapping spectra propagating in opposite direction. The wave emanating from the right corner represents the transmitted signal and the other a delayed version of it.

Figure 3.11: Time-domain signals of interest (left) and their frequency spectra (right). The fact that the spectra of the two interfering waves overlap, results in a spectrum of different shape for the received signal. However, the frequency content of the spectrum is still determined by the contents of the sources spectra.
The signals at the receiver as function of time and frequency are shown in Figure 3.11. The resultant wave spectrum differs significantly from the two original spectra in the overlap region.

B. Interference and source signals have non overlapping spectra

To generate two signals of non-overlapping spectra, we use the same signals as above. The only difference is that the transmitted signal is now up-converted to a carrier frequency of 15 MHz. The results are shown in Figures 3.12 and 3.13. The interference source (IS) is located in the middle of the computational box. The transmitter and receiver locations are unchanged.

Figure 3.12: Wave propagation in 3-D computational space. The interference source is located at the center of the 3-D computational space and generates a wave of different spectrum than the transmitted signal wave emanating from the right.

Figure 3.13: Time-domain signals of interest (left) and their frequency spectra (right). As the spectra of the two interfering signals do not overlap, the spectrum of the received signal consists of both unchanged spectra. The frequency content of the received electric field is again determined by the contents of the two sources spectra.
3.5. Conclusions on interference analysis and simulations

In this chapter the focus was on the main types of interference sources present in a substation environment namely, high EM-fields around substation’s equipment, switching actions, corona, lightning, radio frequency emissions and signal reflections. All the sources generate unwanted radiated emission, which interfere with the information carrying transmitted signal between two nodes of the communication network. To better understand the effects of this interference on the transmitted signal, a frequency domain analysis of all the signals was performed. The results are summarized here below.

First of all, the types of interference have been classified. Figure 3.3 gives an overview of the classification tree used for this purpose. From the classification, only signal replicas due to reflections in the substation environment share the frequency spectrum with the transmitted signal. They are therefore classified as an in-band type of interference. The other types of above mentioned interference sources are all located at a lower side of the spectrum, with significant frequencies up to the MHz range of the spectrum. They are classified as an out-band type of interference.

Secondly, the interference analysis was based on some assumptions, statements or physical laws known from literature. A Matlab program based on Maxwell’s equations has been written and simulations have been ran to verify those assumptions, statements or laws. They are listed in the following.

- The frequency content of the current (interference current or source current) determines the frequency content of the waves they produce and vice versa.
- Interference of EM-waves with non-overlapping spectra in the frequency domain results in a resultant wave with a spectrum consisting of all the individual spectra. Frequency-band selection by filtering can select a spectrum specific to one of the interfering wave and therefore differentiate the waves at the receiver.
- In the case of overlapping spectra, the waves can not be differentiated at the receiver if the individual spectra are not known before the waves interfere. The resultant spectrum is namely totally different than the original spectra. We should then make sure that the transmitted signal spectrum does not coincide (overlap) with the interference spectrum.

Based on the above mentioned conclusions, we can further infer that:

- Choosing an operating frequency of 2.4 GHz or higher, locates the communication signals at a safe place in the frequency spectrum for the most types of interference sources.
- The most harmful type of interference for the wireless link are the signal replicas. They are the key in determining the suitable transmission technology. Therefore, their effects on the transmitted signal deserve more attention and will be studied further in chapter 5 section 5.3.
The communication system taken into consideration is shown in Figure 4.1. The objective of this chapter is to derive a suitable model for the channel to run some simulations in order to predict the behavior of the channel as well as its limitations.

The information input \( m \) in Figure 4.1 represents the data collected by a sensor or a sensor node of the embedded wireless sensor network. More information on the sensors and their signals can be found in chapter 6, section 6.3.

### 4.1 The transmitted signal

The transmitted signal is a physical bandpass waveform with non-zero spectrum concentrated near the carrier frequency \( f_c \). Equation (4.1) gives an expression for the transmitted signal [13]:

\[
s(t) = \text{Re}\left\{g(t)e^{j2\pi f_c t}\right\} = S_I(t)\cos(2\pi f_c t) - S_Q(t)\sin(2\pi f_c t)
\]

where, \( g(t) \) is the low-pass complex envelope of \( s(t) \). It is a baseband waveform with non-zero spectrum concentrated near \( f \) is zero. \( S_I(t) \) and \( S_Q(t) \) are the in-phase and quadrature components of \( s(t) \) respectively.

This representation can allow us to remove the carrier terms whenever we want, when analyzing our communication system and just use the equivalent low-pass representations for the transmitted signal, the received signal and the channel impulse response.
4.2. The received signal

The received signal $r(t)$ is the sum of the line-of-sight (LOS) path and all resolvable multipath components. This is due to the causality and linearity properties of the wireless channel. In other words the received signal consists of multiple time-delayed and attenuated signal replicas. In the following we derive a mathematical expression for the received signal.

**Derivation of received signal**

Starting from the transmitted signal, we can write down the following association:

\[ s(t) \leftrightarrow g(t) \quad \text{(i)} \]

meaning that the complex envelope $g(t)$, belongs to the signal $s(t)$ according to equation (4.1). The received signal will first be a time-delayed version of the transmitted signal. Based on equation (4.1) and considering a delay $\tau(t)$, we can then write:

\[ s(t - \tau(t)) = \text{Re}\{ g(t - \tau(t))e^{j2\pi f_d(t)(t - \tau(t))} \} = \text{Re}\{ g(t - \tau(t))e^{-j2\pi f_d(t)\tau(t)}e^{j2\pi f_d(t)} \}, \quad (4.2) \]

Leading to the following association:

\[ s(t - \tau(t)) \leftrightarrow g(t - \tau(t))e^{-j2\pi f_d(t)\tau(t)} \quad \text{(ii)} \]

This time-delayed signal will be attenuated. Let’s call the attenuation, which in the general case depends on distance and time, $\alpha(d,t)$. We then obtain the third association relating the attenuated, time-delayed signal to its complex envelope:

\[ \alpha(d,t)s(t - \tau(t)) \leftrightarrow \alpha(d,t)g(t - \tau(t))e^{-j2\pi f_d(t)\tau(t)} \quad \text{(iii)} \]

Using the expression for the complex envelope obtained in (iii), the received signal can then be written as:

\[ r(t) = \text{Re}\{ \alpha(d,t)g(t - \tau(t))e^{-j2\pi f_d(t)\tau(t)}e^{j2\pi f_d(t)} \} \quad (4.3) \]

Because of motion, the received signal will not only be attenuated and time-delayed but will also experience a phase change. The Doppler phase shift $\varphi_d(t)$ is given by the following equation:

\[ \varphi_d(t) = \int \left[ 2\pi f_d(t) \right] d\tau \quad (4.4) \]

with $f_d = (v/\lambda)\cos(\theta(t))$ the Doppler frequency shift. $v$ is the velocity related to the motion of the transceivers and $\lambda$ is the wavelength of the wave. $\theta(t)$ is the angle of arrival of the wave relative to the direction of motion of the transmitter or receiver. Adding the Doppler phase shift to equation (4.3), the expression for the received signal becomes:

\[ r(t) = \text{Re}\{ \alpha(d,t)g(t - \tau(t))e^{j2\pi f_d(t)(t - \tau(t)) + \varphi_d(t)} \} \quad (4.5) \]

Taking now multipath into consideration, the received signal will be the sum of all $N(t)$ signal replicas.

\[ r(t) = \text{Re}\left\{ \sum_{n=1}^{N(t)} \alpha_n(d,t)g(t - \tau_n(t))e^{j2\pi f_d(t)(t - \tau_n(t)) + \varphi_n(t)} \right\} \quad (4.6) \]

Let’s now introduce the total phase $\varphi_n(t)$ such that:

\[ \varphi_n(t) = 2\pi f_d(t)\tau_n(t) - \varphi_n(t) \quad (4.7) \]
Equation (4.6) simplifies into:

\[ r(t) = \text{Re} \left\{ \sum_{n=1}^{N} \alpha_n(t) e^{-j\phi_n(t)} g(t - \tau_n(t)) e^{j2\pi f_c t} \right\} \]  

(4.8)

We can now make some simplifications to adapt this general expression to our particular case. First of all we are dealing with a quasi-static environment. As a result of this, time variations in the number of multipath components, the attenuation and the delays will be removed. Furthermore there will be no Doppler phase shift (If \( v = 0 \), then \( f_D = 0 \) and \( \phi_D(t) = 0 \)). The received signal becomes:

\[ r(t) = \text{Re} \left\{ \sum_{n=1}^{N} \alpha_n d e^{-j\phi_n} g(t - \tau_n) e^{j2\pi f_c t} \right\} \]  

(4.9)

with \( \phi_n = 2\pi f_c \tau_n \). The expression within the squared bracket in equation (4.9) is the complex envelope \( \tilde{g}(t) \) of the received signal. It shows us how the channel affects the complex envelope \( g(t) \) of the transmitted signal. The fourth association can now be made:

\[ r(t) \leftrightarrow \tilde{g}(t) = \sum_{n=1}^{N} \alpha_n d e^{-j\phi_n} g(t - \tau_n) \]  

(iv)

If we assume that the transmitted signal is narrowband, meaning that the maximum delay spread of the channel is much less than the symbol duration of the transmitted signal, then the following assumption also holds:

\[ g(t - \tau_n) = g(t) \]  

(4.10)

The received signal will then be:

\[ r(t) = \text{Re} \left\{ \sum_{n=1}^{N} \alpha_n d e^{-j\phi_n} g(t) e^{j2\pi f_c t} \right\} \]  

(4.11)

And the last association follows as:

\[ r(t) \leftrightarrow \tilde{g}(t) = \sum_{n=1}^{N} \alpha_n d e^{-j\phi_n} g(t) \]  

(v)

It should be noted that in the case the narrowband assumption does not hold, the received signal consist of a superposition of different symbols arriving at the same time and not only the same symbol shifted in time.

The received signal \( r(t) \) can also be written in terms of its in-phase and quadrature components \( r_I(t) \) and \( r_Q(t) \) as follows:

\[ r(t) = r_I(t) \cos(2\pi f_c t) - r_Q(t) \sin(2\pi f_c t) \]  

(4.12)

where for a narrowband signal,

\[
\begin{align*}
    r_I(t) &= \sum_{n=1}^{N} \alpha_n d \cos(\phi_n) S_I(t) + \sum_{n=1}^{N} \alpha_n d \sin(\phi_n) S_Q(t) \\
    r_Q(t) &= \sum_{n=1}^{N} \alpha_n d \cos(\phi_n) S_Q(t) - \sum_{n=1}^{N} \alpha_n d \sin(\phi_n) S_I(t)
\end{align*}
\]  

(4.13)

In the next sections we explain how the attenuation coefficients \( \alpha_n \), the path delays \( \tau_n \) and the path phases \( \phi_n \) are obtained to complete the channel model and perform some simulations.
4.2.1 Generation of attenuation coefficients

The attenuation coefficients are represented by the variable $\alpha_n(d_n)$, with $n \in \{1, \ldots, N\}$ where $N$ is the number of multipath components. The distance travelled by the $n^{th}$ multipath or the length of the $n^{th}$ path is denoted by $d_n$. In section 2.4, we found an expression for the total attenuation (loss) as function of distance as given by equation (2.12). This total loss $L(d)$ at distance $d$ can also be written in terms of signal power into the transmitting antenna $P_{TX}$ and signal power from the receiving antenna $P_{RX}$ as follow:

$$L(d) = \frac{P_{TX}}{P_{RX}} \text{ or in } \text{dB, } L(d)_{\text{dB}} = P_{TX}_{\text{dB}} - P_{RX}_{\text{dB}} \quad (4.14)$$

The transmitter and receiver power are computed from the expressions for the transmitted and received signal, equation (4.1) and (4.11) respectively. We then have:

$$s(t) = \text{Re}\{g(t)e^{j2\pi f t}\} \Rightarrow P_{TX} = \frac{1}{2}\langle |g(t)|^2 \rangle$$
$$r(t) = \text{Re}\{\tilde{g}(t)e^{j2\pi f t}\} \Rightarrow P_{RX} = \frac{1}{2}\langle |\tilde{g}(t)|^2 \rangle \quad (4.15)$$

The total loss then becomes:

$$L(d) = \frac{\langle |g(t)|^2 \rangle}{\langle |\tilde{g}(t)|^2 \rangle} = \frac{\langle |g(t)|^2 \rangle}{\langle \sum_{n=1}^{N} \alpha_n(d)e^{-j\phi_n}g(t) |^2 \rangle} \quad (4.16)$$

We will now focus on the denominator of equation (4.16) to derive the expression for the total loss at distance $d$.

Writing down the summation in the expression of the denominator it is obvious that the term $g(t)$ is a common factor and can thus be placed in front of the sum:

$$\langle \sum_{n=1}^{N} \alpha_n(d)e^{-j\phi_n}g(t) |^2 \rangle = \langle |g(t)|^2 \sum_{n=1}^{N} \alpha_n(d)e^{-j\phi_n} |^2 \rangle \quad (I)$$

The complex envelope can also be written as:

$$g(t) = |g(t)|e^{j\arg\{g(t)\}} \quad (II)$$

Furthermore let’s introduce the complex amplitude $A_{\alpha, \Phi}$, which is the result of the sum term as depicted in Figure 4.2 below.

![Figure 4.2: The complex amplitude in the complex plane.](image)
We thus can write: 

\[ \sum_{n=1}^{N} \alpha_n(d) e^{-j\phi_n} = \sum_{n=1}^{N} \alpha_n(d) \cos(\varphi_n) - j \sum_{n=1}^{N} \alpha_n(d) \sin(\varphi_n) = \alpha(d) e^{j\Phi(d)} = A_{\alpha, \Phi} \] 

(III)

with, [30]:

\[ \alpha(d) = \sqrt{\left( \sum_{n=1}^{N} \alpha_n(d) \cos(\varphi_n) \right)^2 + \left( \sum_{n=1}^{N} \alpha_n(d) \sin(\varphi_n) \right)^2} \quad \text{and} \quad \Phi(d) = \tan^{-1} \left( \frac{-\sum_{n=1}^{N} \alpha_n(d) \sin(\varphi_n)}{\sum_{n=1}^{N} \alpha_n(d) \cos(\varphi_n)} \right) \] 

(IV)

Adding equation (III) to (I) and recalling that the modulus of the product of two complex number is equal to the product of the moduli, the denominator of equation (4.16) becomes:

\[ \left\langle \left| g(t) \right| A_{\alpha, d} \right\rangle^2 = \left\langle \left| A_{\alpha, d} \right| \right\rangle^2 \left\langle \left| g(t) \right| \right\rangle^2 = \left\langle \left( \alpha(d) \right)^2 \left| g(t) \right|^2 \right\rangle \] 

(V)

The time average operator \( \left\langle \left[ . \right] \right\rangle \) is defined as follow [13]:

\[ \left\langle \left[ . \right] \right\rangle = \lim_{T \to \infty} \frac{1}{T} \int_{-\frac{T}{2}}^{\frac{T}{2}} \left[ . \right] dt \] 

(VII)

Applying this definition to equation (V) and realizing that \( \alpha^2(d) \) is a real constant, the denominator of equation (4.16) becomes:

\[ \left\langle \alpha^2(d) \left| g(t) \right|^2 \right\rangle = \lim_{T \to \infty} \frac{1}{T} \int_{-\frac{T}{2}}^{\frac{T}{2}} \alpha^2(d) \left| g(t) \right|^2 dt = \alpha^2(d) \lim_{T \to \infty} \frac{1}{T} \int_{-\frac{T}{2}}^{\frac{T}{2}} \left| g(t) \right|^2 dt = \alpha^2(d) \left\langle \left| g(t) \right|^2 \right\rangle \] 

(VIII)

Having found an expression for the denominator the total loss is rewritten as:

\[ L(d) = \frac{1}{\alpha^2(d)} \] 

(4.17)

where, 

\[ \alpha^2(d) = \left( \sum_{n=1}^{N} \alpha_n(d) \cos(\varphi_n) \right)^2 + \left( \sum_{n=1}^{N} \alpha_n(d) \sin(\varphi_n) \right)^2 \] 

(4.18)

The attenuation coefficients can now be generated according to equation (4.17) as follows:

\[ \alpha_n(d_n) = \frac{1}{\sqrt{L(d_n)}} \] 

(4.19)

### 4.2.2 Generation of path-delays

The path-delays are represented by the variable \( \tau_n \), with \( n \in \{1, \ldots, N\} \) where \( N \) is the number of multipath components. In other words \( \tau_n \) is the arrival time of the \( n^{th} \) multipath component as depicted in Figure 4.3. Those arrival times are all comprise in the time interval between 0 and the maximum delay spread \( \tau_{\text{max}} \). For \( t > 0 \), \( \tau_n \in (0, \tau_{\text{max}}] \), and the number of arrival is an integer that cannot decrease with time. This is a counting process, which counts the number of arrivals as time goes on.
Figure 4.3: Arrival times or path-delays $\tau_n$ and inter-arrival times $X_n$, for $N$ multipath components. The last path-delay $\tau_N$ determines the maximal delay spread.

Furthermore, the time of each arrival is completely random, but there is a known average number of arrivals per unit time $\lambda$. This average rate is given by the following equation:

$$\lambda = \frac{N}{\tau_{\text{max}}} = \frac{N}{\tau_N}$$  \hspace{1cm} (4.20)

We can thus model the number of arrivals in any time interval $(t_0, t_1] \subset (0, \tau_{\text{max}}]$ as a Poisson random variable with expected value $\lambda(t_1 - t_0)$. Furthermore, for any pair of non-overlapping intervals $(t_0, t_1]$ and $(t_2, t_3]$ the numbers of arrivals are independent random variables. The two preceding remarks make our counting process a Poisson process of rate $\lambda$. For such a process the inter-arrival times $X_1, X_2, \ldots, X_N$ (see Figure 4.3) are independent, identically distributed random sequence with the exponential probability function as given below [31]:

$$f_X(x) = \begin{cases} \lambda e^{-\lambda x} & x \geq 0 \\ 0 & \text{otherwise} \end{cases}$$  \hspace{1cm} (4.21)

From Figure 4.3, we observe that the $i^{th}$ path-delay is the cumulative sum of the previous inter-arrival times $X_n$ ($n = 1, \ldots, i$) up to the $i^{th}$ inter-arrival time $X_i$. Thus after generating the inter-arrival times $X_n$ and replacing the first arrival time $X_1$ by the propagation time of the Line Of Sight (LOS) component (see equation (4.22)), we compute the path-delays according to equation (4.23):

$$X_1 = \frac{d}{c}$$  \hspace{1cm} (4.22)

$$\tau_i = X_1 + X_2 + \ldots + X_i$$  \hspace{1cm} (4.23)

The variable $d$ is the distance between the transmitter and the receiver and $c$ accounts for the speed of the signals, which is taken to be the speed of light: $c = 3 \times 10^8 \text{ m/s}$. 
4.2.3 Generation of path-phases

The path-phases are represented by the variable $\varphi_n$, with $n \in \{1, \ldots, N\}$, where $N$ is the number of multipath components. Removing the time variance in equation (4.7) and assuming a Doppler phase shift of zero, we obtain for the variable $\varphi_n$:

$$\varphi_n = 2\pi f_c \tau_n$$

(4.24)

where, $f_c$ is the carrier frequency and $\tau_n$ the path-delay of the $n^{th}$ multipath component. The path-phases have unit of radians and are distributed over $[0, 2\pi]$. If $f_c$ is large – think of values in the GHz range – the path-phases can go through a 360° rotation for a small change in path-delays. Therefore we scale the phases to remain in the interval $[0, 2\pi]$ as follows:

$$\varphi_n = (n + r)2\pi \rightarrow \varphi_n = r2\pi$$

(4.25)

In equation (4.25), $n \in \mathbb{N}$ and $r = f_c\tau_n - n \in \mathbb{R}$ . This also means that the path-phases can be viewed as a random variable uniformly distributed over $[0, 2\pi]$.

4.3 The Finite Impulse Response (FIR) model

From the previous sections, it can be concluded that the wireless channel is a causal, linear and time-invariant system, with a fixed number of multipath-components. Therefore, it can be modeled as a Finite Impulse Response (FIR) filter, with impulse response $h(t)$, such that the received signal $r(t)$ is obtained by taking the convolution of the transmitted signal $s(t)$ with the channel impulse response $h(t)$, as given by equation (4.26).

$$r(t) = h(t) * s(t)$$

(4.26)

In terms of low-pass equivalents, the received signal is found by taking the real part of the convolution of the equivalent low-pass input signal or complex envelope $g(t)$, with the equivalent low-pass time-invariant channel impulse response $h_l(t)$, up-converted to the carrier frequency as follows:

$$r(t) = \text{Re}\{(h_l(t) * g(t))e^{j2\pi f_c t}\} \triangleq \text{Re}\left\{\int_{-\infty}^{\infty} h_l(t)g(t-\tau)d\tau e^{j2\pi f_c t}\right\}$$

(4.27)

Comparing equation (4.9) with (4.27), we can derive an expression for $h_l(t)$:

$$h_l(t) = \sum_{n=1}^{N} \alpha_n(d)e^{-j\varphi_n}\delta(t-\tau_n)$$

(4.28)

The variable $t$ represents the delay spread and not time, as we are dealing with a time-invariant or stationary channel [11]. The channel impulse response $h(t)$ is then given by equation (4.29):

---

13 The operator $\lfloor x \rfloor$ denotes the floor of $x$, the greatest integer less than or equal to $x$ [R. Jonhsonbaugh, 2001].
\( h(t) = 2 \text{Re}\left\{ h_i(t) e^{j2\pi f_i t} \right\} \) (4.29)

The factor of 2 is to avoid constant factors in the expression of \( H(f) \), the transfer function or frequency response of the channel in equation (4.31) [11]. Taking the Fourier transform of the equivalent low-pass channel impulse response \( h_i(t) \), we obtain \( H_i(f) \):

\[
H_i(f) = \sum_{n=0}^{N} \alpha_n(d) e^{j(2\pi f_i \tau_n + \Phi_n)}
\] (4.30)

The frequency response of the channel \( H(f) \) then becomes:

\[
H(f) = H_i(f - f_c) + H_i^*(-f - f_c)
\] (4.31)

### 4.4 The Final model

In the final model, the effects of noise and interference are added to the FIR model. The received signal, as given by equation (4.9) then becomes:

\[
r(t) = \text{Re}\left\{ \left[ \sum_{n=1}^{N} \alpha_n(d) e^{-j\phi_n} g(t - \tau_n) \right] e^{j2\pi f_i t} + \text{Re}\left\{ n_i(t) e^{j2\pi f_i t} \right\} + I(t) \right\}
\] (4.32)

The summation term contains the interference due to signal replicas, while the term \( I(t) \) represents all out-of-band types of interference as discussed in previous chapter. This interference is located around another carrier frequency \( f_{ci} \) and can be written according to equation (3.12) as follows:

\[
I(t) = \text{Re}\left\{ \sum_{\omega \in B_i} \tilde{A}_i(\omega) e^{j\omega t} \right\}
\] (4.33)

where, \( B_i \) represents the total bandwidth of the resultant wave which comprises all out-of-band types of interferences, and \( \omega_{ci} = 2\pi f_{ci} \). The three terms of equation (4.32) can then be joined together by using the fact that the real part of the sum of two or more complex numbers is equal to the sum of the real parts. We then obtain:

\[
r(t) = \text{Re}\left\{ \left[ \sum_{n=1}^{N} \alpha_n(d) e^{-j\phi_n} g(t - \tau_n) \right] + n_i(t) \right\} e^{j2\pi f_i t} + \left[ \sum_{\omega \in B_i} \tilde{A}_i(\omega) e^{j\omega t} \right] e^{j\omega_t t}
\] (4.34)

This final expression for the received signal contains the transmitted signal, the channel noise and the channel interference.
4.5 Conclusions on channel model

In this chapter, a model for the wireless channel in a substation environment, based on a derivation of the received signal at a fixed location, is proposed. The communication system considered consists of a transmitter, a receiver and a wireless channel in between. The wireless channel is modeled as a finite impulse response filter. Convolution of the transmitted signal with this finite impulse response filter results in a received signal as given by equation (4.5). By adding the noise and the interference present in the channel to this received signal, a new expression for the received signal as given by equation (4.34) and repeated here below is found:

$$r(t) = \text{Re} \left\{ \left( \sum_{n=1}^{N} \alpha_n(d) e^{-j\phi_n} g(t - \tau_n) \right) + n_i(t) e^{j2\pi f_i t} + \left( \sum_{\omega \in B_i} \tilde{A}_i(\omega) e^{j\omega t} \right) e^{j2\pi f_i t} \right\}$$

The variables in this expression are briefly summarized in the following:

- $N$ denotes the number of multipath components. It is a fixed number for two specific nodes in the communication system.
- The coefficient $\alpha_n(d)$ represents the attenuation of the $n^{th}$ multipath component, with $d$ the fixed distance between the two nodes. These coefficients are generated by considering the total loss in signal strength due to distance (path loss) and the effects of shadowing, according to equation (4.19).
- The time of arrival of the $n^{th}$ multipath component $\tau_n$ is defined as a cumulative sum of independent, identically distributed inter-arrival times with an exponential probability density function.
- The exponent $\phi_n$ accounts for the path phase of the $n^{th}$ multipath component due to their relative delays $\tau_n$. These path phases are considered uniformly distributed over $[0, 2\pi]$.
- $n_i(t)$ is a complex low-pass Gaussian process representing the additive white Gaussian noise (AWGN) in the channel.

The above mentioned variables describe the in-band type of interference, noise included. The next variables in the above expression for the received signal, are related to the out-band type of interference.

- $B_i$ represents the total bandwidth of the resultant wave which comprises all out-of-band types of interferences.
- $\tilde{A}_i(\omega)$ are Fourier coefficients and represent the amplitude of the frequency spectrum of the interference resultant wave.

In the next chapter, the validity of the channel model is verified by analyzing the simulation results obtained.
Chapter 5

Channel Model Simulations Results

Contents
5.1 Description of the simulation system
5.2 Channel model impulse and frequency response
5.3 Wideband signal versus narrowband signal
5.4 Addition of noise to the channel model
5.5 Bit Error Rate as function of channel parameters
5.6 Conclusions on channel model simulations

In this chapter, the wireless channel model implemented in Matlab, according to the discussions and equations provided in previous chapters, will be tested. The fundamental channel parameters derived in chapter two (see Table 2.1) are the input arguments to the Matlab simulation program. This means that the number of multipath components $N$, the average distance $d$ between transmitter and receiver and the other in Table 2.1 mentioned parameters can be changed to adjust the channel behavior to meet our needs. Interference is added to the model according to the discussions of chapter three. The implementation is mainly based on the equations provided in chapter four. The Matlab code for the following simulations can be found in Appendix C.

5.1 Description of the simulation system

The system that has been used to perform the simulations and test the validity of our channel model is depicted in Figure 5.1. The transmitted signal $s(t)$ is obtained according to equation (4.1) as shown in the figure. A random bit stream, representative for a digitized sensor signal, is first generated. The bits are then mapped or coded into symbols belonging to one of the three possible constellations shown in Figure 5.2, namely binary phase shift keying (BPSK), quadrature phase shift keying (QPSK) or quadrature amplitude modulation (QAM). The obtained coded digital signal is made continuous by convolution with a pulse shaping function $p(t)$ to give the transmitted signal complex envelope $g(t)$. Up-conversion of $g(t)$ to the carrier frequency $f_c$ and real part retrieval generates the transmitted signal $s(t)$.

![Figure 5.1: The simulation system to test and validate the channel model. An analog transmitted signal is generated from a random bit stream and passed through the channel model of chapter four. At the receiver, sampling and QPSK demodulation are performed, and the estimated symbols are analyzed in the complex plane.](image-url)
The transmitted signal $s(t)$ is then passed through the wireless channel to be received as $r(t)$. Equation (5.1) describes the pulse shaping function.

$$p(t) = \begin{cases} 1, & t = 0 \\ \frac{\sin(\pi t)}{\pi t}, & t \neq 0 \end{cases}$$ (5.1)

Figure 5.2: The possible transmitted symbols constellations. Mostly QPSK will be used for the simulations.

5.2 Channel model impulse and frequency responses

Let us now take a look at the wireless channel model. Any physical wireless channel can be described by its impulse and frequency responses. The impulse response is the response of the channel to a single transmitted pulse and gives an indication about the number of multipath present in the channel with their corresponding attenuation. The frequency response is the Fourier transform of the impulse response and reveals the coherence bandwidth of the channel. Knowing the frequency response, we can derive a good estimate for the maximum data rate supported by the channel without the need of equalization at the receiver. Figure 5.3 shows the normalized impulse response of a wireless channel consisting of twenty multipath components ($N = 20$), with and without shadow-fading.

The amplitudes of the impulses in the impulse response were determined as explained in section 4.2.1. The standard deviation of shadow-fading $\sigma_s$ is taken to be zero and four respectively. The times of arrival are computed according to a Poisson process as discussed in section 4.2.2. The maximum delay spread $\tau_{\text{max}}$ equals 1 $\mu$s. From the figure, we observe a
decrease in amplitude of the multipath components with increasing delay, obeying the propagation law of decreasing signal power density as the travel time increases. However, addition of shadow fading can cause certain amplitudes at higher delays to be greater than at lower delays. The corresponding frequency response plots are shown in Figure 5.4, for a set of frequencies around the carrier frequency of 2.4 GHz.

![Figure 5.4: Channel Frequency response around 2.4 GHz, for the channel of Figure 5.3, N = 20. (a) \(\sigma_s = 0 \text{ dB}\) and (b) \(\sigma_s = 4 \text{ dB}\). The responses show how different frequencies of the transmitted signal will be affected by the channel.](image)

For the channels of Figure 5.3 and 5.4, the computed coherence bandwidth (see equation (2.6)) equals 500 kHz. The frequency response plots show that the channels are frequency selective, and thus time dispersive for a signal whose bandwidth will be greater than the above mentioned coherence bandwidth. In that case, some frequencies of the transmitted signal will be more attenuated than other frequencies. This is due to the multipath nature of the channels.

In order to clearly understand the effects of multipath, we will simulate a one-path channel \((N = 1)\). Figure 5.5 shows its frequency response, which is flat for all frequencies. A signal going through such a channel will undergo flat-fading and will, therefore, not be distorted in time.

![Figure 5.5: Normalized channel impulse and frequency response, N = 1. The frequency response is flat for all frequencies. A signal passing through this channel will not be distorted.](image)

In the remaining part of this chapter, we will let a signal pass through the channel and analyze the received signal. Following the discussion above, we will distinguish between two types of signals. A narrowband signal which has a bandwidth smaller than the channel coherence bandwidth and a wideband signal for which the opposite holds.
5.3 Wideband signal versus narrowband signal

5.3.1 Wideband signal through the channel model

In order to generate a wideband signal, we have to make sure that the signal bandwidth is significantly greater than the channel coherence bandwidth. This translates in time domain into a signal with a symbol duration that is smaller than the maximum delay spread. Therefore, we choose to send 1706 QPSK symbols in a time interval of 224 $\mu$s. This is the same rate as is used in the Digital Video Broadcasting (DVB) standard for the European terrestrial digital television (DTV) service. Table 5.1 summarizes the values specified for the simulations. The table also shows the symbol duration and symbol rate computed from the two first specified parameters.

Table 5.1: Values specified in the Matlab code to perform the wideband signal simulations.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Meaning</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_{\text{Signal}}$</td>
<td>The signal duration</td>
<td>224 $\mu$s</td>
</tr>
<tr>
<td>$N_{\text{Symb}}$</td>
<td>The total number of transmitted QPSK symbols</td>
<td>1706</td>
</tr>
<tr>
<td>$f_c$</td>
<td>The carrier frequency</td>
<td>2.4 GHz</td>
</tr>
<tr>
<td>$f_s$</td>
<td>The sampling frequency</td>
<td>10 GHz</td>
</tr>
<tr>
<td>$N$</td>
<td>The number of multipath components</td>
<td>20</td>
</tr>
<tr>
<td>$\tau_{\text{max}}$</td>
<td>The maximum channel delay spread</td>
<td>1 $\mu$s</td>
</tr>
<tr>
<td>$d$</td>
<td>The distance between transmitter and receiver</td>
<td>20 m</td>
</tr>
<tr>
<td>$\gamma$</td>
<td>The path-loss exponent</td>
<td>3</td>
</tr>
<tr>
<td>$\sigma_s$</td>
<td>The standard deviation of shadow fading</td>
<td>4 dB</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Computed parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_{\text{Symb}}$</td>
</tr>
<tr>
<td>$R_{\text{Symb}}$</td>
</tr>
</tbody>
</table>

In the following, we look at the frequency response through which the wideband signal is passed, for a frequency band equal to the signal bandwidth, around the carrier frequency. Figure 5.6 shows the results.

![Normalized channel impulse response](image)

Figure 5.6: Normalized channel impulse response (left plot) and channel frequency response (right plot), $N = 20$, $\sigma_s = 4$ dB. For the considered transmitted signal bandwidth, the channel response is not flat, resulting in ISI.

The channel frequency response is not flat. Inter Symbol Interference (ISI) will occur as depicted in Figure 5.7. Different symbols arrive at the same time at the receiver. If we consider the transmission of a signal consisting of one symbol, the top picture of Figure 5.8
shows all the replicas of the signal as function of time for the channel of Figure 5.6. The bottom plot in Figure 5.8 is the sum of all signal replicas and represents the band-limited channel response of the channel. This band-limited channel response is the result of convolving the channel impulse response, as depicted in Figure 5.6 (left plot), with a pulse as used for pulse shaping. The period of the pulse is the same as the symbol period. It is clear from the figure that if a second symbol is sent directly after the first, it will interfere (superpose) at the receiver with delayed replicas of the first symbol.

Figure 5.7: The transmitted (left) and received (middle) symbols in the complex plane for a wideband signal, \( N = 20 \), \( \sigma_s = 4\text{dB} \). Due to ISI, the received symbols are not only belonging to the QPSK alphabet anymore. In the rightmost plot, the data is shifted back before demodulation.

Notice that the channel also introduces a phase shift on the data as explained in section 4.2.1, due to the random phases added by each multipath component. Back-shifting is required before QPSK demodulation is performed, as shown in Figure 5.7.

![Transmitted Symbols](image1)
![Received Symbols](image2)
![Received shifted Symbols](image3)

Figure 5.8: Pulse replicas and normalized band-limited channel response for a wideband signal, \( N = 20 \), \( \sigma_s = 4\text{dB} \). The band-limited channel response shows that a single transmitted pulse (blue curve in top plot) appears at the receiver as multiple pulses (bottom plot).

Let’s now increase the symbol rate by taking a value two times higher than in the previous case. \( T_{\text{Symb}} = 65.65 \text{ ns} \) and \( R_{\text{Symb}} = 15.2321 \text{ MSymb.s}^{-1} \). The results are shown in Figures 5.9 – 5.11. The channel is again frequency selective (see Figure 5.9) and ISI occurs at the receiver as shown in Figure 5.10. However, the received symbols this time form four distinct clusters in the complex plane. Passing these received symbols through a QPSK demodulator at the receiver results in unambiguous retrieval of the data. This was not the case in the previous
simulations where, after QPSK demodulation, more than 300 received symbols out of the 1706 transmitted symbols were not correctly estimated. We can then distinguish between two types of ISI. “Strong” ISI (SISI) with errors at the receiver due to wrong symbol estimates and “light” ISI (LISI) where the interference is not harmful for the data. The question that will then arise is: how can we increase the symbol rate and have a better channel performance in terms of number of errors at the receiver?

Figure 5.9: Normalized channel impulse response and channel frequency response, $N = 20$, $\sigma_z = 4dB$.

Figure 5.10: The transmitted (left) and received (middle) symbols in the complex plane for a wideband signal, $N = 20$, $\sigma_z = 4dB$. In the right most plot, the symbols are divided by the channel complex amplitude.

Figure 5.11: Pulse replicas and normalized band-limited channel response for a wideband signal, $N = 20$, $\sigma_z = 4dB$. This time, the band-limited channel response shows that a single transmitted pulse (blue curve in top plot) appears at the receiver as one pulse of significant amplitude (bottom plot).

The answer to this question lies in the plots of the band-limited channel response. In Figure 5.11, the response of the channel to a pulse of duration $T_{Symb}$ results in twenty pulses of
different amplitudes whose sum more or less resemble a single pulse. This situation is very suitable for signal transmission. A single pulse at the input of the channel results in a single pulse at the output of the channel. In Figure 5.8, the band-limited response results in two pulses of significant amplitudes separated in time by a time interval in the same order as a symbol duration. If we transmit two symbols of different polarities the one after the other, the second pulse of the band-limited response will cancel the second transmitted symbol. Such a channel will result in strong ISI with errors, wrong symbol estimates, at the receiver.

The last presented wideband signal simulations can be viewed in Figures 5.12 - 5.14. The symbol rate is similar to the rate used in the previous simulations, namely $15.2321 \text{MSymb.s}^{-1}$. This time the ISI is strong as shown by Figure 5.14, resulting in errors at the receiver after QPSK demodulation (see Figure 5.14).

**Figure 5.12:** Normalized channel impulse response, power delay profile in dB and channel frequency response, $N = 20$, $\sigma_s = 4\text{dB}$, for the simulations of Figure 5.13 and 5.14.

**Figure 5.13:** Pulse replicas and normalized band-limited channel response for a wideband signal, $N = 20$, $\sigma_s = 4\text{dB}$. A single transmitted pulse results in a signal with multiple significant pulses.

**Figure 5.14:** The transmitted (left), received (middle) and received back-shifted (right) symbols in the complex plane, $N = 20$, $\sigma_s = 4\text{dB}$. SISI occurs and the QPSK symbols cannot be recovered.
From above presented simulation results, we can infer that not only the maximum delay spread or RMS-delay spread is of importance in determining the maximum symbol rate, but also the strength of the replicas with respect to each other. That is why we also look at the RMS-delay spread and the mean excess delay spread which takes also the path amplitudes into consideration as given by equation (2.4). However the condition imposed by the RMS-delay spread on the maximum symbol rate without the need for equalization is too strong. We can allow for higher data rates \(B_{\text{Signal}} > B_{\text{Coh}}\) if we look at the band-limited channel response.

There are then two criteria to decide whether or not the ISI can be termed as strong or light: The number of significant maxima in the band-limited channel response and their separation in time as compared to the symbol duration.

### 5.3.2 Narrowband signal through the channel model

As stated above, a narrowband signal is a signal with a bandwidth or inverse symbol duration smaller than the coherence bandwidth. We will now send a narrowband signal through the channel and observe the received symbols. The changed values specified for the simulations are shown in Table 5.2 (for the other values see Table 5.1). The table also shows the derived symbol duration and symbol rate.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Meaning</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>(T_{\text{signal}})</td>
<td>The signal duration</td>
<td>22.4 ms</td>
</tr>
<tr>
<td>(N_{\text{symb}})</td>
<td>The total number of transmitted QPSK symbols</td>
<td>896</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Computed parameters</th>
<th>(T_{\text{symb}})</th>
<th>The symbol duration: (T_{\text{signal}}/N_{\text{symb}})</th>
<th>25 (\mu)s</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R_{\text{symb}})</td>
<td>The symbol rate: (1/T_{\text{symb}})</td>
<td>40 kSymb.s(^{-1})</td>
<td></td>
</tr>
</tbody>
</table>

The simulations results are shown in Figures 5.15 - 17. The impulse response and frequency response of the channel are shown in Figure 5.15. The frequency response is plotted for a frequency interval equal to the signal bandwidth around the carrier frequency. It is a “flat” response. It should be noticed that, the scale of the vertical axis in the frequency response plot considerably differs from previous frequency response plots.

![Normalized channel impulse response](image)

![Channel frequency response](image)

**Figure 5.15:** Normalized channel impulse response, power delay profile in dB and channel frequency response, \(N = 20\), \(\sigma_d = 4\)dB. The frequency response of the channel is almost flat for a frequency band equal to the signal bandwidth.
If we now look at the band-limited channel response, there is no time distortion as the sum of all signal replicas results in a signal of the same shape as the transmitted one. This is shown in Figure 5.16. The symbol duration of 25 $\mu$s is significantly greater than the maximum delay spread of 1 $\mu$s.

![Figure 5.16: Pulse replicas and normalized band-limited channel response for a narrowband signal, $N = 20$, $\sigma_s = 4$ dB. A narrowband transmitted pulse appears undistorted at the receiver.](image1)

There is no ISI as confirmed by the middle plot of Figure 5.17. However, the channel does introduce a phase and amplitude change on the transmitted symbols. In this case, back shifting is required at the receiver. This can easily be achieved, if the channel complex amplitude, responsible for the shift of the symbols, is known at the receiver. Because the transmitted data contains a training sequence, it is possible to invert the effects of the channel complex amplitude at the receiver. The right most plot of Figure 5.17 shows the result. The received symbols all equal the transmitted symbols after back shifting.

![Figure 5.17: The transmitted (left), received (middle) and back-shifted (right) received symbols in the complex plane for a narrowband signal, $N = 20$, $\sigma_s = 4$ dB. There is no ISI, but the channel does introduce a phase shift on the transmitted symbols, due to the complex attenuation coefficients of the multipath components.](image2)
5.4 Addition of noise to the channel model

Noise is added to the channel model by considering it as being additive, white and Gaussian (AWGN). We will consider a narrowband signal with symbol duration of 25 µs. The simulation parameters are reported in Tables 5.1. The noise power is specified in terms of signal to noise power ratio (SNR) in dB. We assume a transmitted signal power of one. The noise mean power is then 1/SNR. The simulation results are presented in Figures 5.18 and 5.19, for SNR of 20 dB and 5 dB, respectively. The obtained clusters in Figure 5.18 (c) are not due to ISI but to the noise addition. As long as the clusters do not overlap, perfect data recovery is possible at the receiver. However, if the noise power increases such that the clusters do overlap, errors will be made at the receiver.

![Figure 5.18: (a) The transmitted symbols, (b) the received symbols and (c) the received shifted symbols. The noise addition results in a spreading of the data around the original symbols. SNR = 20 dB.](image)

A decrease of signal to noise ratio to 5 dB as in Figure 5.19 results in wider clusters as expected.

![Figure 5.19: (a) The transmitted symbols, (b) the received symbols and (c) the received shifted symbols. SNR = 5 dB. An increase in SNR increases the spreading around the original transmitted symbols.](image)

In the case of a wideband signal, the spreading of the data around the original symbols due to noise will add to the spreading caused by ISI, and increases the probability of errors at the receiver. In the following sections, we will test the validity of the channel model further, by looking at how the different channel parameters influence the number of errors at the receiver, after passage of the transmitted signal through the wireless channel model.
5.5 Bit Error Rate as function of channel parameters

It is clear from the preceding results that interference and noise will deteriorate the transmitted signal. This will result in wrong symbol estimates at the receiver. A measure for the errors in received symbols is the bit error rate (BER). In the subsequent sections we will investigate the effects of noise, number of multipath components, distance between transceivers and symbol rate on the received symbols respectively.

5.5.1 Bit Error Rate as function of SNR

In order to study the effects of noise on the transmitted data we vary the SNR from 0 to 10 dB in steps of 0.5 dB. The other channel model parameters are kept constant and their values are the same as in Table 5.2. For each value of SNR, 100 simulations are performed. The bit error rate is computed after each of the 100 simulations. From the 100 obtained values of BER, an average is computed and used as final value to estimate the BER at the particular SNR. In each trial, 1000 BPSK or QPSK symbols are sent. The transmitted signal is a narrowband signal in order to prevent ISI to influence the results. The results are shown in Figure 5.20. In the figure, the theoretical value of BER for BPSK according to the formula given by equation 5.2 is also shown [11].

\[ P_b = Q\left( \sqrt{\frac{E_b}{N_0}} \right) = \frac{1}{2} \text{erfc}\left( \sqrt{\frac{E_b}{N_0}} \right) \]  

(5.2)

where, \( P_b \) is the probability of bit error. The ratio \( E_b/N_0 \) is the SNR for binary signaling. \( E_b \) is the signal energy per bit and \( N_0 \) is related to the noise power spectral density as given by equation (2.17). The same formula is also used for QPSK, but with the SNR divided by two.

![Figure 5.20: Bit Error Rate (BER) for BPSK and QPSK transmitted symbols as function of Signal to Noise Ratio (\( E_b/N_0 \)). The theoretical and simulated values are shown for both modulations.](image)

The figure shows a close match between the theoretical values and the simulated values. This confirms the validity of the transceivers and channel model.
5.5.2 Bit Error Rate as function of number of multipath components

To study the effects of the number of multipath components on the transmitted data, the previous simulations are repeated for different values of \( N \) namely: 1, 10 and 50. The results for QPSK are shown in Figure 5.21. Again, the simulations are performed with a narrowband signal in order to prevent ISI to influence the results.

![Figure 5.21: Bit Error Rate (BER) for QPSK modulated transmitted symbols as function of Signal to Noise Ratio (\( E_b/N_0 \)), for 3 different values of number of multipath components \( N \).](image)

Figure 5.21 shows that the higher the number of multipath, the smaller the slope of the probability of bit errors as function of SNR will be. An increase in SNR does improve the BER, but not as much as in a single path channel. The higher the number of multipath components, the worse the channel will be. As stated in [32], increasing the SNR will only improve the BER if the receiver is not suffering from multipath components.

5.5.3 Bit Error Rate as function of distance between transceivers

The distance between the transmitter and receiver also affects the data transmission. Figure 5.22 shows the BER as the distance increases from 1 to 150 m, in steps of 10 meters. The other channel parameters are the same as in Table 5.4, only the SNR is now kept to the constant value of 90 dB while the distance \( d \) is varied.

![Figure 5.22: Bit Error Rate (BER) for QPSK modulated transmitted symbols as function of distance between transmitter and receiver, for 2 different values of number of multipath components \( N \).](image)
As expected, an increase in separation distance between transmitter and receiver results in a degradation of received signal and, thus, more errors at the receiver. This is very important when many nodes are to be linked together to form a network. The BER puts a constraint on the maximum separation distance between the nodes. Again, this figure shows that an increase in the number of multipath components deteriorates the quality of the wireless link.

5.5.4 Bit Error Rate as function of symbol rate

The symbol rate determines whether or not ISI will take place. For these simulations, all channel parameters are kept constant, except for the symbol duration. By changing the symbol duration and keeping the number of transmitted QPSK symbols constant, the symbol rate is varied. The plot of Figure 5.23 shows how the BER changes as the symbol rate increases. As expected, when the symbol rate increases, the number of errors at the receiver also increases.

This figure shows that depending on the application, the wireless channel limits the maximal allowable symbol rate.

![Figure 5.23: Bit Error Rate (BER) for QPSK modulated transmitted symbols as function of symbol rate, for 10 multipath components.](image)

Simulation results limit the allowable data rate to less than 1 Mbps. It should be noticed here that the power at the receiver has been normalized and the noise power scaled accordingly. For the channel model, the parameters in Table 2.1 have been used.

5.6 Conclusions on channel model simulations

In this chapter the wireless channel model, based on the discussions and equations of previous chapters, has been tested and validated. The main channel parameters derived in chapter 2 (see table 2.1) were used together with the model presented in chapter 4 in which interference as discussed in chapter 3 has been included.
The validity of the channel model has been confirmed by the simulation results which can be sufficiently explained by the existing theory on wireless channels. The main results are listed here below:

- The frequency response of the channel for a frequency band equal to the signal bandwidth is a good indicator for signal distortion and thus ISI. A flat response means no signal distortion and thus no ISI.

- The maximum delay spread of the channel should be significantly less than the symbol duration or inverse signal bandwidth to avoid ISI. The wireless channel therefore, imposes upper bounds for the maximum allowable symbol rate.

- The attenuation coefficients and path phases of the multipath components multiply the transmitted signal complex envelope by a complex amplitude. For static channels, the transmitted signal can then always be reconstructed at the receiver after dividing the received signal by this complex amplitude.

- The BER depends on the channel parameters. An increase of SNR decreases the BER. On the contrary, increasing the distance will results in poorer link quality. Likewise, an increase in symbol rate results in an increase of BER.

- A significant increase of SNR does not result in a significant increase in BER if the receiver is suffering from multipath components. The higher the number of multipath components, the smaller the increase in BER will be.

In addition, we also found that:

- A non-flat frequency response does not necessarily mean that errors will occur in the transmission due to ISI (compare Figure 5.8 and 5.11). We can distinguish between strong ISI (SISI), where errors are introduced after demodulation at the receiver and light ISI (LISI), which do not cause errors at the receiver. The LISI therefore, allows for symbol rates significantly higher than the coherence bandwidth of the channel. A look at the band-limited channel response in the time domain can give a reliable assessment whether or not the ISI is strong or light.

- Multipath components have an adverse effect on the quality of the wireless link even for narrowband signals.

- With normalized power at the receiver and the values specified in Table 2.1 for the channel parameters, simulation results limit the maximum data rate to less than 1Mbps.
PART II: THE SENSOR NETWORK

Having described and modeled the wireless channel which serves as link between the nodes, we can now focus on the nodes and the way they communicate with each other.

A schematic drawing of the electrical power grid, (see Figure 1.1), shows that it forms a network of interconnected nodes. The nodes are the generating stations, the substations and the load centers, while the links between them, are the overhead and underground transmission and distribution circuits. Not surprisingly, those nodes will also be nodes of the sensor network, and in particular the substations as explained in chapter 6. The chapter is devoted to the topology of the network and the network nodes. Here, we explain what we mean by a node, the types of nodes needed and used as well as the type of information they gather (collect), process and distribute. This information will of course require specific data rates, which are deterministic for the type of transmission technology.

For low data rates applications (less than 250 Kbps), the ZigBee standard is proposed and studied in Chapter 7. A ZigBee-compliant system: the ZNet platform is tested in laboratory and in a real 380 KV open-air substation.

The last chapter presents OFDM as an alternative for higher data rates. An OFDM transmitter is implemented in Matlab and the transmitted signals are passed through the channel model of the first part for evaluation of system performance.
The network nodes and the way they are connected to each other (network topology), as well as the measured data, significantly determine the bandwidth of the transmitted signals. This bandwidth is essential, because when compared with the channel coherence bandwidth, tells whether or not a signal will be subjected to ISI. In the first section of this chapter, the sensor network topology is presented. The sensor network nodes are the subjects of the remaining sections. The nodes of the network can be small or big as depicted in Figure 6.1. The big nodes are the substations as explained in section 6.2. The last section presents the small nodes, which are all part of a big node.

**Figure 6.1:** The overall sensor network and its nodes. The big nodes are the substations in the electrical power grid. They are interconnected with each other and all consist of networks of interconnected small nodes.
6.1 The Network topology

When designing a communication network, there are many different types of topologies to be considered. The communication network depicted in Figure 6.1 covers the entire electrical power system and mainly consists of 2 types of networks: a network of interconnected big nodes and many similar networks of interconnected small nodes. For each type of network, a different topology can be used. The big nodes can be connected in a mesh fashion via base stations using GSM (Global system for Mobile communications) technology, via existing wired systems like PSTN (Public Switched Telephone Network) or even via the internet. This is shown in Figure 6.2.

As the focus is on the wireless link inside substations, only the topology of the small nodes networks will be further discussed. In the remaining sections, the term sensor network refers to the interconnection of small nodes.

6.1.1 The concept of locality

In chapter one, it was mentioned that instabilities in power systems are mostly detected too late to take counteracting measures. This is due to the fact that actual grid protection tools to prevent cascading failures rely mainly on extensive offline studies. Fast real-time information acquisition and analysis for immediate local ‘healing’ is inevitable. By local, we refer to the actual fault location or more precisely the nearest point at which the fault is detected.

6.1.2 Distributed versus centralized approach

In a centralized approach, all the small nodes are connected to a central node in a slave-master structure. There is no direct link between the nodes. The network is simple to implement and control. In addition, a hierarchy is introduced. This has a huge advantage when considering the whole network of interconnected big nodes (substations). If a substation is interested in what is happening in another substation, only the master node of that substation has to be addressed. All the information is namely located there. The major drawback with this type of topology is the complete failure of the network in case the central
node fails. This is not the case in a distributed approach, where all the nodes are equal, and the information therefore spread out into the whole network. This makes local healing possible. Failure of an arbitrary node only results in a small loss of information as the rest of the network remains in operation. As opposed to the centralized approach, the network becomes more complex and therefore more difficult to manage. Furthermore, the hierarchical structure offered by the former approach is lost with its benefit. However, as the robustness of the electrical power grid is our objective, the main advantage of the distributed approach can not be disregarded. The advantages of both worlds have to be combined.

6.1.3 The hybrid solution

The hybrid solution takes the benefits of both worlds. While communicating with each other, the nodes inform at the same time a central node. In [8] the electrical power system is compared with the human body to explain the importance of local healing. If the foot gets hurt, it is immediately treated locally to prevent infection of the leg and rest of the body. In the mean time the foot alerts the head about the injury and the local treatment. This requires a distributed and centralized approach simultaneously implemented. The proposed network topology then looks like depicted in Figure 6.3.

![Figure 6.3: The hybrid approach for the communication network topology. The small nodes are connected to all the nodes in their coverage area and to a central node, forming clusters.](image)

The nodes are organized in clusters and one of the cluster members is elected as master node of the cluster. As the node are equal, failure of a master node results in replacement of the node by another member. Another advantage of the cluster structure is that it reduces the transmitting power of the nodes and therefore increases the life time of the energy supply units. As stated earlier the topology will influence the transmitted signal bandwidth. For example, a system addressing 32 nodes each seconds and transferring 2048 bits per second, requires a throughput of 65 kbps. Higher throughput of 4 Mbps are also required for system addressing 32 nodes with a data packet size of 64 bytes and a response time of 4 ms on a Token passing network [33].
6.2 The big nodes: Substations

A big node consists of many small nodes linked together to form a wireless sensor network. Those big nodes are also linked together into a network to share information. Substations represent big nodes in the network, as shown in Figures 6.1 and 6.2. The reason why they are chosen as big nodes has to do with their important role in the electrical power grid. They are designed to accomplish many functions as listed below:

- Connect generating stations to the transmission and distribution systems.
- Change the voltages from one level to another.
- Regulate the voltages to compensate for system voltage changes.
- Switch transmission and distribution circuits in and out of the system.
- Interconnect the electric systems of various companies.
- Measure electric power qualities flowing in the circuits.
- Control reactive kilovolt-amperes supplied to and the flow of reactive kilovolt-amperes in the circuits.
- Eliminate lightning and other electrical surges from the system.
- Convert alternative current (AC) to direct current (DC) or vise versa in High Voltage DC (HVDC) substations.

In short, substations can reveal the state of the power grid at any moment and help switch defective circuits and components out of service in case a fault or short circuit develops in the power grid to protect the rest of the equipment [23]. Therefore they are excellent locations for the deployment of sensor networks. Furthermore, interconnection of substations results in total coverage of the entire power system. Likewise, the network can easily be expanded to neighboring countries, by addition of substations to the existing network. As stated before, the substations consist of many small nodes linked together. The small nodes are discussed in the next section.

6.3 The small nodes: Sensor nodes

Small nodes are sensor nodes and consist of four major units: a sensing unit (one or more sensors), a processing unit if needed, a communication unit (transmitter/receiver) and an energy supply unit. Their essential role is to continuously sense all variables indicating the state of the electrical power grid and send this information where needed. Temperature for example can give an indication about the state of a power transformer. The collected information is partially processed and exchanged between the nodes. There are important system performance requirements, for the implementation of Local Area Networks (LAN) inside substations. One can think for example of time synchronization, timing constraints and delivery time. Phasor measurements and data sharing require a time synchronization accuracy of $1 \mu s \pm 0.5 \mu s$. About timing constraints, 3 categories of data transfer intervals are defined: less than $10 \, ms$ (high speed), between $10 \, ms$ and $1 \, s$ (medium speed) and greater than $1 \, s$ (low speed). The delivery times or times to send routinely updated data must be less than the update interval [33]. The quantities sensed as well as the topology and communication protocols used, determine the speed at which the data should be transferred.
6.3.1 Temperature and Humidity Sensors

The amount of power that can be transmitted via a distribution circuit is limited by the circuit’s rating which in turn depends on the ratings of its components. For a distribution network in the electrical power grid, those components are overhead lines, transformers and other substation components. Their ratings are defined by the voltages and currents at which they are designed to work and strongly depend on temperature. There are manufacturer-specified operating temperatures and care must be taken to ensure that the components operate at those temperatures to avoid their premature aging and failure. Temperature measurements are a good indicator on whether or not a system is operating in normal conditions and therefore, a necessity in monitoring the state of the grid [34]. Today’s temperature measurements are done in recording intervals of 10 minutes [35]. A 16-bit encoding for a temperature sample results in a data rate much lower than 1 bit per second (bps). The same holds for humidity measurements.

6.3.2 Current and Voltage Measurements

To ensure uninterrupted service, the currents and voltages must be kept within the acceptable limits. Therefore, currents and voltages should also be continuously monitored. Actual monitoring systems (e.g.: SASensor\(^{14}\)), record currents and voltages up to the 50\(^{th}\) harmonic. With a fundamental frequency of 50 Hz, the 50\(^{th}\) harmonic has a frequency of 2500 Hz [46]. According to the Nyquist criterion, the minimum sampling rate allowed to reconstruct the current and voltage waveforms without error is 2 times the highest frequency. Therefore a sampling rate of 5000 Hz or more is required. Representing a sample by 16 bits will require a bit rate of 80 Kilo bits per second (Kbps) or higher.

6.3.3 Inductive Sensors for Partial Discharge Measurements

Partial Discharges (PD) are the main cause of breakdown when AC current is used. They are breakdown phenomena occurring in cavities in dielectrics or along dielectric interfaces and cause a slow erosion of the material. As time goes on, a growing pit with branches is formed in the cavity. This process is called treeing and causes a complete breakdown of the dielectric as the tree reaches the electrodes. This phenomenon can take place in many High Voltage (HV) equipments like transformers, Gas Insulated Switchgears (GIS) and cables at operating voltage [28]. For example, in HV cables of the Crossed-Linked Polyethylene (XLPE) type, which are commonly used nowadays for the underground transmission of voltages up to 500kV, PD originate from manufacturing defects, non proper installation and/or accessory assembling [36].

PD manifests itself, when present in cable insulation, by wave pulses consisting of energy frequencies up to hundreds of MHz [37]. Special designed inductive sensors can capture those pulses. A 16 bit encoding for a measured sample will lead to a bit rate in the order of tens of Mbps. The sensor signal may be processed, amplified and finally transmitted through the wireless channel with a smaller data rate. The inductive sensors can be internal or external. In the first case they are built-in, that means installed in the cable, while in the second case they are mounted on grounding cable [36]. Figure 6.4 shows a cable with inductive sensors to measure PD activity.

\(^{14}\) SASensor is a system for accurate data acquisition, power quality monitoring, fault detection, overcurrent protection, etc. It consists of current and voltage interface modules connected to traditional current and voltage transformers, with sampling rates chosen to digitize harmonic currents and voltages up to the 50\(^{th}\) harmonics [SASensor, 2008].
In the figure, S₁, S₂ and S₃ are the three internal inductive sensors. Their locations (cable terminations and joints) are chosen with respect to the types of defects that may cause PD activity such as: surface discharges on termination, missing insulation screen at the cable/joint transition and cable splice improperly prepared. We will retain the terminations as the best locations because of the ease of access and possibility to mount an antenna for transmission and reception of measured data, after installation of the cable. T₁ and T₂ are the first and second termination of the cable.

6.4 Conclusions on Network nodes and Data

The actual topology of the electrical power grid and the objective of making it more robust determine the topology of the embedded sensor network. Substations, because of their important role in the control of power flow inside the grid, are locations by excellence for the deployment of sensor networks. They are referred as big nodes and can be connected to each other directly via base stations using GSM technology, via the PSTN network or even via internet (see Figure 6.2).

The small nodes are the sensing units inside the substations or big nodes. They gather the information needed to capture the state of the power system at any time. Temperatures, humidities, currents and voltages are for example continuously (with some time intervals) measured. The hybrid solution, as depicted in Figure 6.3, is chosen to connect the small nodes with each other. The most important criteria for selection of this topology are locality, hierarchy and robustness. The topology and associated communication protocols influence the bandwidth of the transmitted signals. Throughput of 65 Kbps and 4 Mbps can be expected.

The data rates associated with the measured quantities are to be considered for the choice of transmission technology as discussed in chapter 7 and 8. Most measured quantities have data rates less than 80 Kbps. However, not only the sensed quantities determine the data rates to be supported by the wireless link. The network topology and communication protocols also put constraints on these data rates. Furthermore, nodes can be equipped with more than one sensor. This can also increase the data rate, if all the data have to be sent in a certain time interval.
Chapter 7

ZigBee as Transmission Technology

Contents
7.1 Why ZigBee?
7.2 A ZigBee compliant system: The ZNet Platform
7.3 The limitations of the system
7.4 Conclusions on ZigBee

The discussions of previous chapter on the type of information needed to capture the state of the electrical power grid, revealed different data rates for the sensed quantities. Temperature and humidity are examples of low data rates quantities. This chapter therefore, focuses on a communication standard for low data rates applications: ZigBee. A ZigBee-compliant off the shelf equipment (the ZNet platform), which has been used for real time measurements in a real substation environment, is presented.

7.1 Why ZigBee?

ZigBee is a standard defining a set of communication protocols for short-range low-data-rate wireless networking. ZigBee is therefore designed for mostly battery-powered applications, with as main requirements: low data rate, low cost and long battery life time. In many ZigBee applications, the devices have duty cycles\(^\text{15}\) of less than 1% and the data rates are lower than 250 Kbps. This is sufficient for most of the quantities to be measured, as reported in section 6.3. In Figure 7.1, a comparison is made between 3 types of short range wireless networking classes, in terms of power consumption, complexity and cost against data rate [32].

\[\text{Figure 7.1: Comparison of ZigBee with 2 other types of short-range wireless networking classes [32]. ZigBee appears as a good candidate for low data rates. Bluetooth has a too small range.}\]

The IEEE 802.15.4 standard has been adopted by ZigBee as Physical Layer (PHY) and Medium Access Control (MAC) protocols. A study of OFDM for higher data rates will be

\(^{15}\) The duty cycle is defined as the ratio of the time the device is active to the total time. A device that wakes up every minute and stays active for 600 ms, has a duty cycle of 1%.
done in the next chapter. Table 7.1 gives a summary of the operating frequency band of interest and corresponding data rate, with modulation and spreading method, for the latest version of IEEE 802.15.4 [32]. The optional specifications are omitted.

**Table 7.1: 2.4 GHz frequency band of operation and corresponding data rate for the latest version of IEEE 802.15.4 standard.**

<table>
<thead>
<tr>
<th>Frequency Bands (MHz)</th>
<th>Number of Channels</th>
<th>Modulation</th>
<th>Chip Rate (Kchip.s⁻¹)</th>
<th>Bit Rate (Kbps)</th>
<th>Symbol Rate (Ksymbol.s⁻¹)</th>
<th>Spreading Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>2400-2483.5 (2.4 GHz band)</td>
<td>16</td>
<td>O-QPSK¹⁶</td>
<td>2000</td>
<td>250</td>
<td>62.5</td>
<td>DSSS¹⁷ (16-orthogonal)</td>
</tr>
</tbody>
</table>

As discussed in part I, we are mainly interested in the 2.4 GHz frequency band of operation. The spreading method for this operation band is Direct Sequence Spread Spectrum (DSSS). Each transmitted symbol, which consists of 4 bits, is mapped into a unique 32-bit sequence or chip sequence. Therefore there are 16 unique sequences for each ZigBee symbol. Likewise, the effective bandwidth of the signal passing through the wireless channel is 8 times greater than the real signal bandwidth. If the signal bandwidth is 250 KHz, the effective bandwidth for signal transmission will become 2 MHz. IEEE 802.15.4 signals are thus considered wideband signals [38]. DSSS increases the number of simultaneous data transfers by the use of different quasi-orthogonal¹⁸ chip sequences. Another advantage of ZigBee is that it increases coverage by mesh and tree networking (see Figure 7.2). Hierarchy and distribution are present as discussed in section 6.1.3.

**Figure 7.2:** ZigBee mesh networking increases the coverage range of the network by using coordinators nodes and end devices. A topology that is close to the hybrid concept of section 6.1.3. (picture taken from [32]).

In summary, the allowable data rates, the range or coverage area ability and low-power consumption of ZigBee system are the main reasons why it has been chosen for application in electrical power system monitoring and control in this thesis work.

¹⁶ O-QPSK or offset-QPSK is a modified form of QPSK which supports more efficient amplification by eliminating 180° phase transitions between the In-phase and Quadrature components of the signal.

¹⁷ DSSS: Direct Sequence Spread Spectrum is a spreading method to create robustness against multipath fading and increase the receiver sensitivity level.

¹⁸ Orthogonal sequences are sequences for which the cross-correlation function is equal to zero, meaning that they are totally dissimilar.
7.2 A ZigBee compliant system: The ZNet platform

As stated above, ZigBee appears to be a good candidate for deployment of a sensor network inside substations. Therefore, a ZigBee-compliant system has been tested in laboratory and in a real 380 kV open-air substation. The system is described in the following and some measurement results are presented.

7.2.1 Description of the ZNet platform

The ZNet platform is a wireless sensor network developed in Germany by the JLM innovation firm [39]. The sensor nodes are equipped with a sensing unit, a microcontroller, a wireless transceiver and energy storage unit. Figure 7.3 shows a ZNet sensor node. The width and length are 4.5 cm and 6.5 cm, respectively. The nodes can be configured as end-device or router. An end-device is asleep most of the time and wakes up to check new messages or send new measurement results. A router on the contrary is awake all the time and relays messages of other nodes to extend the network coverage. The coordinator of the ZNet platform is a USB-stick connected to a computer or laptop in which the ZNet software is installed. Figure 7.2 shows end-devices, routers and a coordinator.

![Figure 7.3: A ZNet sensor node with sensing, processing, transceiver and energy storage units.](image)

The sensing unit can gather information on quantities such as temperature, humidity and light intensity. The temperature and humidity are measured with a SHT-11 sensor from Sensirion for a range of -20 to 100 °C and 0 to 100% relative Humidity\(^\text{19}\) (r.H), respectively. The light sensor is a Rhom BH1710VFC sensor and measures light in the range of 1 to 65536 lux (lx). The transceivers are ZigBee modules with built-in antenna, operating in the 2.4 GHz frequency band. The energy storage unit consists of 3 AAA (Alkaline) 1.5 V batteries. There are also possibilities to connect 4 - 6 V and 200 mA power supplies [39].

\(^{19}\) Relative Humidity (r.H) is given in percent and is the ratio of the actual water vapor pressure to the saturated vapor pressure at a prescribed temperature.
7.2.2 Measurements results

The measurements have been done in the High Voltage (HV) laboratory of the Delft University of technology (TU Delft) and at the 380 kV open air substation of Dodewaard, The Netherlands.

A. Measurements at TU Delft HV-Laboratory

In the HV-laboratory, corona (the interference source) is generated by mounting a piece of copper wire on a high-voltage electrode. The voltage on the high-voltage electrode can be regulated and is increased until the electric field enhancement caused by the copper wire causes air to ionize. The E-field strength is measured at the sensor node location and the source voltage is increased until the value of the E-field strength equals the reference level of 10 kV/m, as reported in the ICNIRP\textsuperscript{20} guidelines for limiting occupational exposure to time-varying electromagnetic fields [40]. The radiated emission caused by corona is picked-up by 3 types of antenna connected to a spectrum analyzer: A bow-tie antenna, a dielectric wedge antenna and a monopole antenna (See Figure 7.4). The antenna frequency ranges are: 400 MHz – 4 GHz, 85 MHz – 2.5 GHz and 2 GHz – 2.5 GHz, for the dielectric wedge antenna, bow-tie and monopole antennas, respectively [47], [48].

The interference source (corona source) is then located between the transmitting and receiving sensor nodes as depicted in Figure 7.5. The end device records the ambient temperature, humidity and light intensity. This is first done with the interference source not active. The measurements results are shown in Figure 7.6. Then the voltage on the high-voltage electrode is increased until corona is generated. Many measurements are performed, at different voltage levels ranging from 30 kV to 50 kV. Figure 7.7 shows measurements results at a voltage level of 40 kV. The results are the same for other voltage levels. The temperature for example is always 22.3\textdegree C on average.

\textsuperscript{20} ICNIRP: International Commission on Non-Ionizing Radiation Protection.
The transmission is not perturbed by the corona. A picture of the frequency spectrum (see Figure 3.5) of the signal generated by the measurement antenna reveals the absence of corona around 2.4 GHz. However, the case of a transformer with faulted windings, as discussed in section 3.2.3, showed spectra with frequency components up to the GHz range. Even though the interference cause by the discharges has frequency components in the GHz range, the measured temperature, humidity and light intensity are not different from the measurements shown by Figure 7.6. This is due to the fact that ZigBee uses DSSS at the physical layer. The transmitted data is coded (data spreading by the chip sequence) at the transmitter. At the receiver the dispersing of the data will keep all wideband interference spreaded and spread all narrowband interference.

![Figure 7.5: The test set-up in the HV-laboratory. A transformer is used to raise the voltage at a high-voltage electrode on which a sharp point is fixed to generate corona. This interference source is located between a coordinator node and an end device during data transmission.](image)

![Figure 7.6: Ambient temperature, Humidity and Light intensity measurements in HV-Laboratory in absence of corona. The measured values are almost constant.](image)
B. Field test at 380 kV open air substation in Dodewaard, The Netherlands

At the 380 kV open air substation of Dodewaard, a sensor node is arbitrarily positioned in the substation, as end device. The coordinator node remains at a fixed location, with possibility to be connected to a power supply, as shown in Figure 7.8. The figure also shows one of the locations of the end device.

![Figure 7.8: The coordinator node location and one of the end-device locations, for measurements performed at the 380 kV open air substation of Dodewaard (The Netherlands).](image)
The temperature, humidity and light intensity are first measured at a sufficient distance away from the substation (train station Hemmen-Dodewaard at approximately 600m from the substation) where we can assume there is no corona or its amplitude is negligible. The distance between transmitter and receiver is approximately 10 m. The measured ambient temperature, humidity and light intensity are shown in Figure 7.9. The average temperature is 18.2 °C and the humidity 18.2 % r.H. In the following, the main purpose of the measurements has been the determination of the range of an end device inside the substation. Therefore, measurements have been performed at different locations with the distance between the end device and the coordinator node increasing.

The end device is moved away from the coordinator node and at each 10 m, the temperature sensor is approached by a heat source (lighter). The results are shown in Figure 7.10. From the first part of the measurements, before the first peak, the average ambient temperature is also found to be 18.2 °C. The ambient humidity measurements are also consistent with the 18 % r.H. found outside the substation.

**Figure 7.9:** Ambient temperature, Humidity and Light intensity measurements, at train station Hemmen-Dodewaard, located at approximately 600m from the substation.

**Figure 7.10:** Ambient temperature and Humidity measurements. The distance between coordinator and end device is increased by moving the end device. At each 10 m the sensors are approached by a heat source.
The system stops recording the measurements when the link quality is bad and the measurements therefore not reliable. This happens at an average distance of 100 m between coordinator node and end device.

7.3 The limitations of the System

The wireless devices sold in Europe must comply with EMC Directive 89/336/EEC [41]. As stated above, ETSI EN 300 328 document characterizes IEEE 802.15.4 signals as wideband signals. According to the standards, there are limitations with respect to the maximum data rates, the maximal distance or coverage and the bandwidth of the transmitted signals. The allowable BER depends on the application.

7.3.1 Maximum bit rate

The maximal bit rate is set by the IEEE 802.15.4 standard. It should be between 20 and 250 Kbps. The ZNet platform measures temperature, humidity and light at a maximal data rate of 1 sample per second.

7.3.2 Maximal distance between nodes

According to the standard (see table of Figure 7.1) and as derived from the measurements a maximal distance of 100 m can be taken as reference for the separation distance between 2 small nodes of the network. Inside the substation environment, the average distance at which the link was broken is also approximately 100 m, despite the huge number of obstacles. According to the specifications, a separation distance of 150 m can be achieved in open field [39].

7.3.3 Power constraints

The values specified here are imposed by the IEEE 802.15.4, for the 2.4 GHz frequency band of operation. The transmitted signal power needs to be adjustable with respect to the separation distance between the nodes and has to be at least -3 dBm (0.5 mW). The PSD of the transmitted signal is limited to -30 dBm. The lowest received signal power, for which the packet error rate is less than 1% is called the receiver sensitivity. A receiver sensitivity of -85 dBm (3.16 pW) is required. The maximum transmit power, according to European regulations [42], is 10 mW/MHz.

7.4 Conclusions on ZigBee

The ZigBee standard defines a set of communication protocols for short-range low-data-rate wireless networking. Figure 7.1 gives a summary of the supported maximal bit rate (250 Kbps) and range (100 m) as compared to 2 others short-range wireless standards. ZigBee is designed for mostly battery-powered applications, with as main requirements: low data rate, low cost and long battery life time. That makes it suitable for sensor networks applications. Therefore, a ZigBee-compliant device (the ZNet platform) has been tested in laboratory and in a real substation environment: the 380 kV substation of Dodewaard in The Netherlands.

According to the test results obtained from both test locations, corona is not a harmful type of interference for data transmitted with the ZigBee system. The frequency analysis of the radiated emissions shows no corona in the 2.4 GHz frequency band of operation. However, it
has been shown that discharge pulses generated by a transformer with faulted windings, can appear in the GHz frequency range at sufficiently high voltage (see Figure 3.6). Broadband antennas connected to a spectrum analyzer can then be used to detect the occurrence of partial discharges in a transformer during operation without having to connect anything to the transformer. This method can further be investigated for use in diagnostic of high voltage assets.

Furthermore, the many obstacles (metallic structures) in the substation do not reduce the maximal range relative to the maximal distance of 100 m as mentioned in the IEEE 802.15.4 standard.
OFDM as Transmission Technology

If we are to transmit video images for surveillance or diagnostic purposes, or sensor data requiring higher data rates, such as inductive sensors data recording PD pulses, we need another modulation technique at the physical layer. This chapter investigates the possibilities of transmitting with data rates significantly higher than ZigBee allows. The IEEE 802.11.g standard supports data rates up to 54 Mbps using Orthogonal Frequency Division Multiplexing (OFDM). The problem with higher data rates is the need for more bandwidth. Signals with higher bandwidths are more sensitive to the channel frequency selectivity property. In the first section OFDM is proposed as solution for higher data rates. The OFDM technique is then explained in sections 8.2 and 8.3. In section 8.4, simulation results are presented, in which a system with OFDM implemented at the receiver is compared to a system without OFDM. The chapter is closed with some conclusions.

8.1 Why OFDM?

As we have seen in Part I, we are dealing with a time dispersive or frequency selective channel. Furthermore, ZigBee limits the data rates to not higher than 250 Kbps. Orthogonal Frequency Division Multiplexing (OFDM) belongs to the family of multi-carrier modulation. The basic idea in multi-carrier modulation is to divide the data stream (to be transmitted bit stream) into many substreams to be transmitted over distinct orthogonal subchannels or subcarriers centered at different subcarrier frequencies. By this division, a wideband signal is broke up into many parallel narrowband signals. If we now make sure that those narrowband signals have a bandwidth smaller or equal than the channel coherence bandwidth, we can get around the frequency selectivity property of the channel. This is illustrated by Figure 2.6. Furthermore, the addition of a cyclic prefix to data blocks in OFDM completely removes the ISI. This will be explained in the next section.

8.2 OFDM

The objective, as discussed in previous section, is to break up a wideband signal into $N_s$ parallel narrowband signals. The complex envelope of the OFDM signal, over a $T_{ss}$ seconds time interval ($0 < t < T_{ss}$), can be written as [13]: 
\[ g(t) = A_c \sum_{n=0}^{N_s-1} w_n \varphi_n(t) \]  

(8.1)

where, \( A_c \) is the carrier amplitude and \( N_s \) the number of subcarriers or subchannels. \( w_n \) represents the \( N_s \) input symbols of duration \( T_{\text{symb}} \). \( w_n \in \{w_0, w_1, ..., w_{N_s-1}\} \). The orthogonal functions (carriers) \( \varphi_n(t) \) are defined as:  

\[ \varphi_n(t) = e^{j2\pi f_n t} \]  

with \( f_n = \frac{1}{T_{\text{symb}}}(n - \frac{N_s-1}{2}) \). Orthogonality is discussed in Appendix F. The term \((N_s-1)/2\) is called the frequency offset. \( T_{\text{symb}} \) is the symbol duration on each subcarrier (subchannel) in seconds.

### 8.2.1 The OFDM transmitter

The OFDM transmitter as depicted in Figure 8.1 and implemented in Matlab for simulation purposes (see section 8.4), generates the transmitted signal \( s(t) \). A bit stream, representing data from a sensor, is QPSK modulated. The symbols are then distributed on parallel branches for Inverse Fast Fourier Transform (IFFT) computations. After that, a cyclic prefix is added to the data and the parallel stream is converted into a serial stream. This parallel stream represents the discrete complex envelope of the transmitted signal. Digital to analogue conversion (pulse shaping) and up conversion to the carrier frequency are the final steps. The transmitted signal is of course a real signal.

![Figure 8.1: The OFDM transmitter.](image)

The transmitted signal \( s(t) \) is the same as given by equation (4.1). Substitution of above mentioned expression (see equation (8.1)) for the complex envelope in equation (4.1), after replacement of \( \varphi_n(t) \) by its expression and suppression of the frequency offset term, results in the following equation:

\[ s(t) = \text{Re}\left\{ g(t)e^{j2\pi f_c t} \right\} = \text{Re}\left\{ A_c \sum_{n=0}^{N_s-1} w_n e^{j\frac{2\pi}{T_{\text{symb}}} n} e^{j2\pi f_c t} \right\} \]  

(8.2)

The discrete time domain can now be introduced, by replacing the continuous time variable \( t \) by its discrete equivalent \( kT_s \) \( (k \in \mathbb{Z}) \), with \( T_s \) being the sample period. Let’s now choose \( T_s \) to be equal to \( T_{\text{symb}}/N_s \). By doing so, a handy formula for the complex envelope of the OFDM signal as described in the following is found. In addition, notice that by making this choice, the Nyquist criterion is not violated, because \( N_s \) is always equal or greater than 2. The complex envelope then becomes:
\[ g(kT_s) = g[k] = A \sum_{n=0}^{N_s-1} w_ne^{j\frac{2\pi nk}{N}} \quad (8.3) \]

Equation (8.3) represents the \( k^{th} \) sample of the time continuous complex envelope of equation (8.1), over the \( T_{ss} \) time interval. This is what needs to be generated for all values of \( k \), in order to divide the original high rate symbol stream into \( N_s \) parallel lower rate substreams. The variable \( k \) accounts thus for the samples in the time interval of \( T_{ss} \) seconds. The number of samples \( K \) on that time interval is given by equation (8.4). It is equal to the number of subchannels \( N_s \).

\[ K = \frac{T_{ss}}{T_s} = \frac{T_{ss}}{N_s} = N_s \quad (8.4) \]

The variable \( n \) in turn, accounts for the subcarriers frequencies. Having \( N_s \) subchannels or subcarriers, we also have \( N_s \) subcarrier frequencies. We then can write:

\[
\begin{align*}
  k &= 0,1,2,\ldots,N_s-1, \\
  n &= 0,1,2,\ldots,N_s-1.
\end{align*}
\quad (8.5)
\]

Let’s now take a look at the expression for the inverse discrete Fourier transform (IDFT) as given by the following equation [13]:

\[ x[k] = \frac{1}{N} \sum_{n=0}^{N-1} X[n]e^{j\frac{2\pi nk}{N}}, \quad k = 0,1,\ldots,N-1. \quad (8.6) \]

This expression generates \( N \) values representing samples of the waveform \( x(t) \). Comparing this expression to equation (8.3), we see that for \( 1/N = A_c \) and \( N = N_s \) it is exactly what is needed. The OFDM signal can thus be generated by using an IDFT algorithm or via hardware a fast implementation of that called an inverse fast Fourier transform (IFFT). In other words, by passing the symbols \( w_n \) to the IFFT-function we generate the sampled version of the complex envelope of the OFDM signal of equation (8.3). This is illustrated in Figure 8.2. During the time interval of \( T_{ss} \) seconds, one symbol is carried on each of the \( N_s \) branches. Having obtained the \( N_s \) samples of the complex envelope of the signal to be transmitted on \( N_s \) parallel branches, the last step consists of putting them back in a serial stream, to really obtain the discrete representation of the complex envelope. The next step will then be up-conversion to the operational carrier frequency \( f_c \) and transmission of the real part, after pulse shaping. However, before doing that the addition of a cyclic prefix to the samples of the complex envelope is needed as explained in the next section.

![Figure 8.2: Translation of \( N_s \) symbols into \( N_s \) time samples of OFDM signal complex envelope.](image-url)
8.2.2 The cyclic prefix addition

The IDFT has been used to obtain the complex envelope of the transmitted signal. At the receiver, the reverse operations have to be performed. This means that a discrete Fourier transform (DFT) or its fast implementation called fast Fourier transform (FFT) must be computed on the received data. This received data is the channel response to the transmitted signal. Having modeled our channel as LTI, the channel response is a linear convolution between the channel impulse response and the input or transmitted signal. This is given by equation (8.7) for the discrete case, where \( y[k] \), \( h[k] \) and \( x[k] \) denote respectively the received discrete signal, the discrete channel impulse response and the discrete transmitted signal.

\[
y[k] = h[k] * x[k] = \sum_{\kappa} h[\kappa] x[k - \kappa] \tag{8.7}
\]

From the definition of the DFT, circular convolution in the time domain is equivalent to multiplication in the frequency domain [11]:

\[
DFT \{h[k] \otimes x[k]\} = H[i]X[i] = Y[i], \quad 0 \leq i \leq N - 1 \tag{8.8}
\]

Equation (8.8) says that, the transmitted sequence \( x[k] \) can be recovered at the receiver, if the channel impulse response and transmitted signal are circularly convoluted, by taking the IDFT of \( Y[i]/H[i] \). The channel impulse response is assumed known at the receiver. Thus we need to transform our linear convolution into a circular one. This can be done by addition of a cyclic prefix to the transmitted signal. The \( N \)-point circular convolution is defined as follows:

\[
h[k] \otimes x[k] = \sum_{\kappa} h[\kappa] x[k - \kappa]_N \tag{8.9}
\]

The term \( x[k - \kappa]_N \) denotes a periodic version of \( x[k - \kappa] \) with period \( N \). Comparing equation (8.7) and (8.9), some periodicity needs to be add to the input sequence. This is done by addition of the cyclic prefix as explained in Appendix A. Following the notations and derivations of the previous section, we can now order the \( N_s \) samples of the complex envelope of the transmitted signal by a parallel to serial converter and add the cyclic prefix to it as depicted in Figure 8.3.

![Figure 8.3: Cyclic prefix addition to OFDM signal complex envelope.](image)
8.3 The OFDM parameters

8.3.1 Number of subcarriers/subchannels

The number of subcarriers or subchannels $N_s$ is chosen to ensure that each subchannel experiences relatively flat fading. In other words, $N_s$ is chosen such that the subchannel bandwidth is less than the coherence bandwidth $B_{coh}$ of the channel. This also means that $N_s$ is chosen to make the symbol duration time $T_{ss}$ on each subchannel, much greater than the maximal delay spread $\tau_{max}$ of the wireless channel. We then can write:

$$N_s = \frac{T_{ss}}{T_{symb}} \gg \tau_{max}$$  \hspace{1cm} (8.10)

Increasing the number of subcarriers increases the symbol duration $T_{ss}$. For the simulations in section 8.4, $N_s$ is chosen to be a power of two, to speed up the IFFT and FFT computations.

8.3.2 Guard Time

The guard time is the duration of the cyclic prefix. It should be longer than the maximal delay spread of the channel.

8.3.3 Subcarrier spacing

The subcarrier frequencies, as mentioned in section 8.2, are determined by the following expression:

$$f_n = \frac{1}{T_{ss}} \left( n - \frac{N_s - 1}{2} \right) = f_0 + \frac{n}{T_{ss}}$$  \hspace{1cm} (8.11)

Subsequent subchannels are then spaced $1/T_{ss}$ Hz apart.

8.3.4 Modulation type per subcarrier

For the sake of simplicity, the same modulation type is used on all subcarriers. For the simulations we therefore use QPSK modulation on all subcarriers.

8.4 OFDM simulation results

The transmitter depicted in Figure 8.1 is used to generate the OFDM signal for the simulations. We first consider the two simple channels, of two and three paths, as depicted in Figure 8.4. The Matlab code to run the simulations is reported in Appendix D.
Figure 8.4: The two simple channels used to understand the effects of OFDM on the transmitted QPSK symbols. (a) a 2-path channel and (b) a 3-path channel.

Figures 8.5 and 8.6 show the results for these two channels, respectively. The transmitted symbols are shown as well as the received symbols, with and without OFDM modulation at the transmitter.

Figure 8.5: Transmitted QPSK symbols and received symbols without and with OFDM modulation at the transmitter, in a 2-path channel (delays: 0, 0.5 $\mu$s).

Figure 8.6: Transmitted QPSK symbols and received symbols without and with OFDM modulation at the transmitter, in a 2-path channel (delays: 0, 0.5 $\mu$s).

To understand why the received symbols are located in the complex plane on circles or loops around the original symbols, we first need to understand how the symbols are retrieved at the receiver. This is explained in the next section.

8.4.1 The shape of the received signals in the complex plane

For the transmitter of Figure 8.1, the receiver first down-convert the received signal to retrieve the signal complex envelope $g(t)$, which is then sampled. Afterwards, serial to parallel and cyclic prefix removal are performed on the discrete complex envelope. The
remaining samples are then used as input to the FFT function. The result of this operation gives the transmitted QPSK symbols back. The FFT operation in an analog implementation is depicted in Figure 8.7.

Figure 8.7: Received QPSK symbols on each branch, after down-conversion and integration.

On each branch (subcarrier), the complex envelope is down-converted with the subcarrier frequency of that branch. The resulting signal is then integrated over a time interval equal to the symbol duration on a branch \((0 < t < T_{ss})\), to give the transmitted symbol on that branch \((w_n)\). This is given by equation (8.12) for the \(n^{th}\) branch.

\[
w_n = \int_{0}^{T_{ss}} g(t)e^{-j2\pi f_n t} dt
\]

An attenuated and delayed version of \(g(t)\) will result in substitution of \(g(t)\) by \(\alpha g(t-\tau)\) in equation (8.14). In Figure 8.5, the received complex envelope consists of the contribution of two paths. The path-delays are 0 and 0.5 \(\mu s\) and the corresponding path-amplitudes (attenuations) are 1 and 0.5 respectively. The retrieved symbol on the \(n^{th}\) branch is then:

\[
\int_{0}^{\tau_{ss}} \left[ g(t) + \alpha g(t-\tau) \right] e^{-j2\pi f_n t} dt = w_n + \alpha w_n e^{j\psi_n}
\]

It consists of the original symbol and a phase-shifted attenuated second symbol. The phase difference \(\psi_n\) depends on the delay \(\tau\) and the subcarrier frequency \(f_n\). Figure 8.8 gives an illustration in the complex plane, for a symbol transmitted on two different subchannels.

Figure 8.8: Received QPSK symbol on two different branches, after down-conversion and integration. The second path generates a second symbol which adds up to the original symbol. The phase of the second symbol depends on the subcarrier frequency and causes a rotation around the original signal.
As the phases $\psi_n$ are between $[0, 2\pi)$, the received symbols are located on a circle around the original transmitted symbol. For a channel with $N$ multipath components, the expression for the received symbol on the $n^{th}$ branch then becomes:

$$w_n = \sum_{m=1}^{N} \alpha_m e^{-j\phi_m} e^{j\psi_{n,m}}$$

(8.14)

where,

$$\left\{ \begin{array}{ll}
\varphi_m = 2\pi f_c \tau_m \\
\psi_{n,m} = -2\pi f_c \tau_m
\end{array} \right. \quad \text{with } m \in \{1, 2, ..., N\} \text{ and } n \in \{0, ..., N_r-1\}$$

(8.15)

In Figure 8.6, the nice circular form is lost because the delay of the third path causes the phase difference of the third contribution with respect to the second to change more rapidly. Figure 8.9 gives an illustration.

In the right-most plot of Figure 8.9, the tops of the smallest arrows are the locations of the retrieved symbols in the complex plane. They are still going around the original transmitted symbol located at the center of the bigger circle, but not following a circle. Depending on the speed at which the phase of a multipath component makes a complete revolution, the final shape will deviates from a circle, sometimes having knots. Figure 8.10 (a) shows the shape of the received symbols in the complex plane for a “1” symbol sent on all branches. The drawn circles indicate the amplitudes of the second and third paths, which forms the limits in which the symbols will appears.

Figure 8.9: Received QPSK symbol on two different branches, after down-conversion and integration. The third path phase changes more rapidly than the second path phase, as indicated by the small arrows. Therefore, the received constellation deviates from a circle.

Figure 8.10: Received symbols for a “1” sent on all branches (subcarriers). The diameters of the drawn circles represent the gains of the 3 paths (1, 2/3 and 1/3). The path-delays are: 0, 1/3 and 2/3 µs, respectively.
8.4.2 The effect of the number of subcarriers

In this section, the number of subcarriers is varied. Figure 8.11 shows the effect of the number of subcarriers on the received symbols in the complex plane, for the 3-path channel with path-delays 0, 1/3 and 2/3 µs.

![Figure 8.11](image)

Figure 8.11: Received symbols with increasing number of subcarriers. An increase in the number of subcarriers results in a decrease of signal bandwidth on the subcarrier branches. The signals on the branches become more narrowband and are therefore subjected to flat fading.

As the number of subcarriers increases, the bandwidth of the subchannel signals decreases. The subchannel signals are becoming more narrowband (as compared to the channel coherence bandwidth) and therefore less sensitive to ISI. However, more subcarriers means smaller subcarrier spacing and more sensitivity to phase noise and frequency offset [43].

8.4.3 The effect of the cyclic prefix length

The following simulation is performed to study the effects of cyclic prefix length. 64 OFDM symbols are sent, each consisting of 64 QPSK symbols. The number of subcarriers is then 64. The length of the cyclic prefix is increased from 0 to 64. Figure 8.12 shows the results. From the received constellations, an increase in cyclic prefix length or guard time results in less ISI. However, the higher the length of the cyclic prefix, the more we loose in useful data rate. From the figure, we can observe a limit for the value of this guard time at which no improvement is seen in the received constellation. For the simulation, the maximal delay spread of the channel was taken to be 1 µs. Therefore a guard time greater than 1 µs is not needed.

The figure also shows that the cyclic prefix does not remove the circular shape of the received symbols around the original symbols. In order to achieve that, equalization is needed at the receiver to reverse the effects of the channel on the transmitted symbols. Addition of a cyclic prefix is then required to transform the linear convolution of the transmitted signal with the channel impulse response into a circular convolution.
8.4.4 Equalization

Equalization or reversing the channel effects is a widely used technique in a static environment where the channel impulse response can be considered time-invariant. The advantage with OFDM is the ease with which equalization can be performed. It turns out to be a simple scaling [44]. However, exact knowledge of the used subcarrier frequencies is required.

As mentioned in section 8.4.1, the increase in subcarrier frequencies, from the first branch to the last branch, causes the phases to rotate and therefore the received symbols to form loops. Errors are made when the loops overlap. By limiting the subcarrier frequencies to small values by reducing the subcarrier spacing, we can avoid the formation of loops and reduce the number of errors. The idea consists of applying zero-padding or cyclic prefixing to the front and tail of the OFDM symbol, before computing the IFFT. Figure 8.13 shows the results, for the 3-path channel of Figure 8.4 for which the path-delays are now: 0, 0.2 and 0.7 µs.

In Figure 8.13 (a), 1024 QPSK symbols are sent in a time interval of 224 µs. At the receiver, 265 symbols were wrong. The use of OFDM without equalization results in approximately
the same number of errors (240), due to the overlapping of the symbols, as shown in Figure 8.13 (c). However, the use of 2048 subcarriers for 1024 symbols, results in no errors at the receiver. The received constellation is shown in Figure 8.13 (d). This technique will be use in the following simulations, where the transmitted signal is passed through the channel model derived in chapter 4. The values for the channel parameters are similar to the values reported in Table 2.1 and used in chapter 5. The maximum data rate is increased from less to 1 Mbps to tens of Mbps with the use of 2048 subcarriers for 1024 QPSK symbols.

8.5 Conclusions on OFDM

The OFDM simulation results show that wideband data transmission is possible with OFDM.

- Multipath channels introduce new symbols on the subcarrier branches at the receiver. These new symbols are attenuated and phase-shifted version of the original transmitted symbol on a particular subcarrier branch at the transmitter. The received symbol on a branch is then the sum of all symbols on that branch as given by equation 8.17.
- The attenuations of the new symbols are equal to the path-gains. The phase-shifts depend on the subcarrier frequency and delays of the multipath components as given by equation 8.18. This results in received constellations of circular shapes or in the form of loops.
- The loops formed by the received symbols around the original transmitted symbol are confined to two circles whose diameters are depending on the amplitudes of the multipath components (see Figure 8.10). This will pose a big problem in the presence of strong echoes, which are likely to occur in a substation environment.
- The increase of the number of subcarriers as well as of the cyclic prefix length (guard time), reduces the ISI. However, addition of a cyclic prefix reduces the useful data rate. Therefore it should be as low as possible. The simulation results show that there is a maximum cyclic prefix length beyond which the performance of the receiver does not increase anymore. This occurs when the guard time equals the maximal delay spread of the channel.
- To get rid of the loops, equalization is required. We can also force the phases to be restricted to certain values such that they do not make complete revolutions, by reducing the subcarrier spacing. This technique requires zero-padding or cyclic prefixing to the front and tail of the OFDM symbol before IFFT. An advantage of this technique is that knowledge of the channel is not required.
- The use of above mentioned technique in combination with OFDM increases the data rate in substation environment by a factor of 10, if twice the number of QPSK symbols is used for the number of subcarriers.
PART III: WIRELESS SENSOR NETWORK FOR POWER SYSTEM MONITORING AND CONTROL

This last part offers a summary of the previous parts and presents all the conclusions and recommendations, in one chapter.

Chapter nine summarizes the study made on wireless channel and interference in substation environments, together with the proposed transmission technologies for application of sensor networks to these environments.

All the conclusions are presented and some recommendations for future works are mentioned at the end of the chapter.
Chapter 9

Conclusions and Recommendations

Contents

- 9.1 Introduction
- 9.2 The wireless sensor network requirements
- 9.3 The wireless channel inside substations
- 9.4 ZigBee and OFDM as candidates for the communication network
- 9.5 Recommendations

9.1 Introduction

Current power systems are undergoing several changes, as discussed in section 1.2. Stability of the overall system, security of the working staff and protection of the High Voltage equipment are the major concerns in this respect. Integration of a wireless sensor network on top of, or embedded in these power systems, will help overcome the problems associated with above mentioned changes. However, wireless data transmission via electromagnetic waves is vulnerable to any electromagnetic emission sharing the same medium.

In this chapter, we summarizes the findings of previous chapters, regarding the application of wireless technology, and in particular wireless sensor networking, inside the electrical power system for monitoring and control. Section 9.2 discusses the requirements of the system in terms of topology and data rates. Furthermore, an attempt is made to indicate to which extend wireless data transfer is possible in an interference prone environment like a substation. Therefore, in section 9.3, the wireless channel inside a substation is presented, including the major types of interferences. Section 9.4 proposed the use of wireless networking standard ZigBee and analyses the use of OFDM at the physical layer.

9.2 The wireless sensor network requirements

The wireless sensor network is intended to make the electrical power grid more reliable, robust and safe, as discussed in section 1.3. In order to achieve that goal, it has to fulfill some requirements in terms of topology, node capacity (throughput) and link reliability. Other requirements are imposed by international standards. The International Electrotechnical Commission (IEC), for example, gives common specifications for apparatus operating near High Voltage equipments [50]. All of which will be the subject of a sub-section in this section.

9.2.1 The sensor network topology

One of the most obvious requirements regarding topology is that the sensor network covers the entire power system (global coverage), while allowing local autonomy. As indicated in chapter six, the sensor network consists of interconnected big nodes (e.g.: substations), which all consist of interconnected small nodes (see Figure 6.1). The interconnection of big nodes takes care of global coverage and the interconnection of small nodes allows locality. The
small nodes are the High Voltage components inside substations. Therefore, the overall communication network is subdivided into two networks. The network topology discussed here, only concerns the interconnection of small nodes. The big nodes can be connected in a mesh fashion, by the use of GSM technology, the PSTN system and the internet, as explained in section 6.1 and depicted in Figure 6.2.

The sensor network consisting of interconnected small nodes, should allow local fault detection and local healing. Furthermore, it is essential to have a central point where all information on the state of a substation can be found. These two requirements asked for a distributed and centralized approach respectively. The appropriate topology should then keep both aspects into consideration. This is possible, if a hybrid topology is used as explained in section 6.1.3 and depicted in Figure 6.3, where the small nodes are grouped into clusters, electing a cluster member as coordinator.

9.2.2 The sensor network nodes

There are two types of network nodes: the big and small nodes. The big nodes are the substations, because of their important role in controlling the power flow in the electrical power system. Furthermore, interconnection of all substations (big nodes), results in total coverage of the power system, and allows scalability of the communication network (extension to neighboring facilities).

The small nodes are all the components inside a substation, which are connected to each other, via a wireless link. In most open-air substations, they are separated by distances varying from 10 to hundreds of meters. Their role is to gather as much information as possible on the state of the grid at their location. Therefore, they consist of sensors measuring all relevant quantities for this purpose, such as temperature, humidity, voltages, currents and partial discharges (see section 6.3). These measurements put constraints on the sensor nodes capacity in terms of data rates and data processing speed. For most of the sensed quantities, such as temperature, humidity and solar irradiance, data rates less than tens of bps are sufficient. Voltages and currents can require data rates of 80 kbps or more. If a node measures more quantities at the same time, higher data rates are required to send all the information on time. Transmission of PD pulses or surveillance images requires tens of Mbps.

9.2.3 The sensor network links between the nodes

The big nodes can be connected via wireless or wired links, as mentioned above. For the small nodes, a wireless link has been proposed, because of the ease of implementation and maintenance, as compared to wired systems, especially for a huge number of sensor nodes. But a wireless link means that the transmitted signals are exposed to any unintentional and intentional electromagnetic emission in a substation. Robustness against the main interference sources in substation environments and security are the major constraints of the wireless link.

9.3 The wireless channel inside substations

The wireless channel inside substations depends on the substation. We will consider only open-air substations. The main channel characteristics and associated parameters are presented, as well as, the major types of interference present in such environments. For a thorough understanding, the reader is referred to chapter 2 and 3 respectively.
9.3.1 The main characteristics of the wireless channel

The wireless channel is causal, linear and time-invariant. Therefore, it can be modeled as an LTI filter. Any electromagnetic wave passing through the channel will be attenuated because of path-loss and shadow fading. The metallic structures in the channel and the ground generate multipath components. The number of multipath is fixed for any couple of sensor node in the channel. The path-delays are then also fixed for two specific nodes. For all possible links, an average value can be use for the maximum delay spread and coherence bandwidth of the channel. Depending on the bandwidth of the transmitted signals the channel can then becomes frequency selective. Doppler effects due to movements are neglected and the channel is considered not frequency dispersive. The noise in the channel is considered additive, white and Gaussian, and is completely characterized by the noise power spectral density. Table 9.1 presents all main characteristics and associated channel parameters considered. Their values are obtained from literature, measurements and otherwise estimated with the idea of modeling a substation environment.

Table 9.1: The wireless channel parameters used for the derivation and implementation of the wireless channel model, specific for substation environments. The values not specified are computed from the other specified values.

<table>
<thead>
<tr>
<th>PROPERTY</th>
<th>PARAMETER</th>
<th>EXPLANATION</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>Causality</td>
<td>IMPLICIT</td>
<td>Introduction of delays in received signal.</td>
<td>No value</td>
</tr>
<tr>
<td>Linearity</td>
<td>IMPLICIT</td>
<td>Superposition or addition of multipath components.</td>
<td>No value</td>
</tr>
<tr>
<td>Time-Invariance</td>
<td>( N )</td>
<td>Number of multipath components. Due to the huge number of metallic structures, we assume a very high number of multipath components.</td>
<td>&gt; 20</td>
</tr>
<tr>
<td>Attenuation</td>
<td>( d )</td>
<td>Distance between transmitter and receiver as estimated from distance measurements between components at the 380 kV substation of Dodewaard, The Netherlands.</td>
<td>20 - 400 m</td>
</tr>
<tr>
<td>( \gamma )</td>
<td></td>
<td>Path loss exponent of metal working factory environment [11], [12].</td>
<td>3</td>
</tr>
<tr>
<td>( \sigma_s )</td>
<td></td>
<td>Standard deviation of shadow fading in obstructed metal working factory is 6.8 dB [12].</td>
<td>8dB</td>
</tr>
<tr>
<td>Multipath</td>
<td>( \tau_{\text{max}} )</td>
<td>Maximal delay spread.</td>
<td>1 ( \mu s )</td>
</tr>
<tr>
<td></td>
<td>( \tau_{\text{rms}} )</td>
<td>RMS-delay spread. According to [11]: 50 ns/30 ( \mu s ) for indoor/outdoor.</td>
<td>-</td>
</tr>
<tr>
<td></td>
<td>( \tau_{\text{mean}} )</td>
<td>Mean excess delay.</td>
<td>-</td>
</tr>
<tr>
<td>Frequency selectivity</td>
<td>( B_{\text{coh}} )</td>
<td>Coherence bandwidth of the channel.</td>
<td>-</td>
</tr>
<tr>
<td>Noise</td>
<td>( N_0 )</td>
<td>Noise power spectral density [15].</td>
<td>-170 dbm/Hz</td>
</tr>
</tbody>
</table>

9.3.2 The main interference sources in the wireless channel

The term “interference sources” refers to all electromagnetic emission unintentionally polluting the wireless channel in open-air substations. These emissions can be classified as natural or man-made, static or dynamic and continuous or transient. Furthermore, they can
also be classified as in-band or out-of-band and narrowband or broadband, when compared to
the transmitted signal bandwidth and spectrum location (see section 3.2 and Figure 3.3) [16],
[20]. The spectra of the interference waves are then required. Electromagnetic wave
interference analysis in the frequency domain reveals that these spectra can be estimated from
the interference source currents (see section 3.4). Therefore all electromagnetic emissions, for
which the spectra have not been measured, were estimated from their source currents.

We distinguished five major types of interference sources: High Voltage components,
switching actions, partial discharges, lightning surges and transmitted signal replicas [16],
[19], [21]. High Voltage components, generates extremely low varying electromagnetic fields,
considered static. Switching actions can produce signals with dominant frequencies up to 120
MHz [24], [25]. Likewise corona’s significant frequency components can be found up to 40
MHz [26]. According to [28], lightning is also located in the MHz range in the frequency
domain. For a system operating around 2.4 GHz, all these types of interferences can be
classified as out-of-band. However, the electromagnetic emission generated by partial
discharge pulses, originating from insulation defects, has significant frequency components
up to 2 GHz and higher [26]. Likewise, the remaining type of interference, transmitted signal
replicas (echoes), will always share the same spectrum as the transmitted signal [11]. They
are therefore, classified as in-band and represent the most dangerous type of interference.

From above discussion, a model of the wireless channel can be derived and implemented,
based on the expression for the received signal as given by the following equation:

\[ r(t) = \text{Re} \left\{ \sum_{n=1}^{N} \alpha_n (d) e^{-j\varphi_n} g(t - \tau_n) + n_i(t) \right\} e^{j2\pi f_{ci} t} + \sum_{\omega \in B_2} A_\omega e^{j\omega t} e^{j2\pi f_{ci} t} \right\} \quad (9.1) \]

where, \( \alpha_n, \varphi_n \) and \( \tau_n \) are the path-gains, path-phases and path-delays respectively. Section 4.5
explains how they can be generated. The function \( g(t) \) is the complex envelope of the
transmitted signal and \( n_i(t) \) represents the complex low-pass Gaussian process accounting for
AWGN. The first term of equation (9.1), between square brackets, is then a summation of
transmitted signal, in-band type of interference and in-band noise. The second term, between
square brackets, represents all out-of-band types of interferences. The carrier frequency \( f_{ci} \) of
the out-of-band interference is chosen such that its spectrum does not overlap with the
spectrum of the first term of equation (9.1). In the following, simulations are performed, in
order to determine the maximal achievable data rate for wireless transmission inside
substations, without any type of coding.

### 9.3.3 Wireless channel model simulation results

The channel model considered in this section was derived in chapter 1, 2, 3 and 4. Table 9.1
gives the main channel parameters. The BER in such a channel increases with distance, data
rate, noise PSD and number of multipath components. Simulation results, limit the allowable
data rate, for such a channel, to 1 Mbps.

### 9.4 ZigBee and OFDM as candidates for the communication network

The discussions of previous sections, which were motivated in previous chapters, together
provide guidelines for the choice of a suitable transmission technology for the sensor network.
Frequency domain analysis of the major interference sources in a substation, restrict the
frequency of operation of the sensor network to extremely high frequencies (GHz range). Furthermore, the required topology (hybrid type) and sensed quantities put constraints on the transmitted signal bandwidth. Further aspects as power constraints, which have not been discussed in details, should be beard in mind for application of sensor networks to power systems. In this respect, ZigBee has been chosen for low data rates, and a study has been done for the use of OFDM for higher data rates.

9.4.1 ZigBee as networking technology

The reasons why the ZigBee standard has been proposed for the sensor network are discussed in the following. First of all, ZigBee allows for large coverage by use of coordinator nodes and end-devices, as depicted in Figure 7.2. Secondly, ZigBee can operate in the frequency band around 2.4 GHz. Last but not least, there are many cheap, off-the shelf ZigBee compliant devices available, all designed for low power applications. The only disadvantage of ZigBee can be in the achievable data rates.

The ZNet platform is a system for temperature, humidity and solar irradiance measurements. The sensor nodes are equipped with transceivers, using the ZigBee standard, to wirelessly distribute the information obtained from their measurements. It has been partially tested in the 380 kV substation of Dodewaard, The Netherlands. The system works also in the presence of corona.

9.4.2 OFDM at the physical layer

OFDM is the preferred multicarrier modulation technique for higher data rates because of the ease of equalization at the receiver. This has been discussed in chapter eight. The main advantage of OFDM is the possibility to achieve higher data rates. A technique consisting of zero-padding or cyclic prefixing in front and tail of the OFDM symbol, before IFFT computations, results in an increase in allowable data rate, by a factor of 10 and more. However complexity, cost and power consumption are major drawbacks.

9.5 Recommendations

9.5.1 Improvement of simulation results

- To improve the channel model simulation results, the channel and transceivers models can be improved by using raised-cosine pulses for the pulse shaping, O-QPSK in combination with OFDM instead of QPSK.
- Addition of equalization to OFDM receiver, to determine the improvement in data rate, as compared to single carrier.
- To improve the frequency analysis simulation results, the memory limitations issues should be solved to simulate waves from signals modulated at 2.4 GHz and incorporate PD current pulses with rise times of 1 ns as interference current source.

9.5.2 Further research

- Channel sounding in substation environments should be done to retrieve good estimates for the channel parameters and accordingly refine the channel model.
- Intensive on-site measurements of spurious electromagnetic emissions from the different interference sources should be done to have a clear picture of the polluted part of the frequency spectrum.
- In depth study of required data rates and effects of topology on these data rates is of the utmost importance.
- EMC at transceivers.
- Investigate the possibility to cope with power issues. Sensors with self-generating power supplies are much more attractive for the point of view of maintenance.
- Look further at security issues (intentional jamming, terrorist actions, etc).
References


Appendix A: Orthogonality and Cyclic Prefix

Orthogonality

In mathematics, the functions in the set \{φₙ(𝑡)\} are orthogonal over time interval \(a < t < b\) if and only if [13]:

\[
\int_a^b \varphi_n(t)\varphi_m^*(t)dt = \begin{cases} 
0, & n \neq m \\
K_n, & n = m
\end{cases} = K_n\delta_{nm} \quad (8.2)
\]

The star in equations (8.2) denotes the complex conjugate. \(\delta_{nm}\) is the Kronecker delta function and \(K_n\) are constants [13]. In the appendix we demonstrate how the set of exponential functions \{φₙ(𝑡)\} used in equation (8.1) are orthogonal over a \(T_{ss}\) seconds time interval, \(0 < t < T_{ss}\). The subcarriers are then spaced \(1/T_{ss}\) Hz apart.

Proof:

Let’s define \(φₙ(t)\) and \(φₘ(t)\) with \(n\) and \(m\) \(\in\{0,1,2,\ldots,N_s-1\}\) as follows:

\[
φₙ(t) = e^{j2\pi n t} = e^{j2\pi n \sigma_s t} \quad \text{and} \quad φₘ(t) = e^{j2\pi m t} = e^{-j2\pi m \sigma_s t}
\]

Applying the orthogonality criterion as postulated in equation (8.3), we obtain:

\[
\int_a^b φₙ(t)φₘ^*(t)dt = \int_0^{T_{ss}} e^{j2\pi (n-m) t} dt = \begin{cases} 
0, & n = m \\
T_{ss}, & n \neq m
\end{cases}
\]

For \(n = m\), \(n - m = 0 \Rightarrow e^{j2\pi (n-m) t} = 1 \Rightarrow \int_0^{T_{ss}} e^{j2\pi (n-m) t} dt = \int_0^{T_{ss}} dt = T_{ss}.
\]

For \(n \neq m\), \(n - m = 0 \Rightarrow \int_0^{T_{ss}} e^{j2\pi (n-m) t} dt = \frac{e^{j2\pi (n-m) 0} - 1}{j \frac{2\pi}{T_{ss}} (n-m)} = 0.
\]

This is due to the fact that \(n - m \in \mathbb{Z}\) and \(e^{j2\pi (n-m)} = \cos[2\pi(n-m)] + j\sin[2\pi(n-m)] = 1.

Cyclic prefix

Consider the sequence \(x[k] = x[0], \ldots, x[N-1]\) of length \(N\) and a discrete channel impulse response \(h[k] = h[0], \ldots, h[µ]\) of length \(µ+1\). Let’s now append a part of this sequence, namely the last \(µ\) values of \(x[k]: \{x[N-µ], \ldots, x[N-1]\}\) to it as header (see Figure (a)). This yields a new sequence of length \(N + µ\), \(x_{cp}[k]\), \(µ \leq k \leq N - 1\).

Figure (a): Prefix addition to sequence \(x[k]\).

The new sequence \(x_{cp}[k]\) is a periodic version of \(x[k]\) with period \(N\), for \(-µ \leq k \leq N - 1\):
\[ x_{cp}[k] = x[k]_N \]  

Using this new sequence \( x_{cp}[k] \) as input to a discrete time channel with impulse response \( h[k] \) will result in the output sequence \( y[k] \), \( 0 \leq k \leq N - 1 \) as given by the following equation:

\[
y[k] = h[k] \ast x_{cp}[k] = \sum_{\kappa} h[\kappa]x_{cp}[k-\kappa] = \sum_{\kappa} h[\kappa]x[k-\kappa]_N = h[k] \otimes x[k] \tag{ii}
\]

Equation (ii) shows that the addition of the header has transformed the linear convolution into a circular one. This header is called the cyclic prefix. The equality \( x_{cp}[k-\kappa] = x[k-\kappa]_N \) in equation (ii) is only valid for \( 0 \leq k \leq N - 1 \). Thus, the linear convolution becomes a circular convolution for \( 0 \leq k \leq N - 1 \). We can then discard the first \( \mu \) samples of \( y[k] \), \( -\mu \leq k \leq 0 \) to recover \( x[k] \), \( 0 \leq k \leq N - 1 \). Notice also that the length of the cyclic prefix is the same as the length of the channel impulse response. The ISI caused by the channel and affecting the first \( \mu \) samples of a data block can then be removed.
Appendix B: Matlab Code for Interference Frequency Analysis Simulations
(See the attached CD-ROM)

The file “Interference Analysis” on the Attached CD-ROM contains all the m-files needed to run the Matlab program on interference analysis of electromagnetic waves. The m-files are:

- Space_volume.m
- Source_current.m
- Interference_current.m
- Source_cur_integral.m
- Interf_cur_integral
- Cur_derivative.m
- E_field_at_receiver.m
- E_field_at_receiver_2sources.m
- E_field_strength.m
- E_field_strength_2sources.m
- Plots_file.m

Appendix C: Matlab Code for Channel Model Simulations
(See the attached CD-ROM)

The file “channel model” on the Attached CD-ROM contains all the m-files needed to perform the simulations of chapter 5. The Matlab m-files are:

- data_generation.m
- gen_delays.m
- gen_phases.m
- gen_gains.m
- pulse.m
- imp_response.m
- intToQPSK.m
- wichanmod.m
- BER_SNR.m
- BERvsDIST.m
- BERvsSymbRate.m
- SigThroughChan.m
- Tr_signal_generation.m

Appendix D: Matlab Code for OFDM Simulations
(See the attached CD-ROM)

The file “OFDM simulations” on the Attached CD-ROM contains all the m-files needed to perform the simulations of chapter 8. The Matlab m-files are:

- SigThroughChanOFDM.m
- Wichanmod_ofdm.m
- Wichanmod_ofdm_test.m

Appendix E: Test Set-up Components Data Sheets
(See the attached CD-ROM)

The file “Test set-up” on the Attached CD-ROM contains all the information on the components used for the test set-up. Data sheets or references to data sheets are provided when available.