Abstract:

This document presents the outcomes of the modelling activities of multiple-input multiple-output (MIMO) radio channels performed within the METRA project. A stochastic model of MIMO radio channels is proposed, and its implementation in COSSAP® is described. Guidelines for choosing the input parameters of the model are also given. On the other hand, an experimental set-up for sounding MIMO channels is presented, as well as the environments in which it has been deployed during measurement campaigns. Finally, the results from the measurements are compared to those produced by the COSSAP® model as a mean of validating the proposed stochastic model.

Keyword list: MIMO channel, space-domain modelling, azimuth dispersion, channel sounder
0 EXECUTIVE SUMMARY

This document presents the outcome of the modelling activities of multiple-input multiple-output (MIMO) radio channels performed in the framework of the METRA project. The scope of this project is to study MIMO antenna concepts for the 3rd generation mobile communication systems, namely Universal Mobile Telecommunication Services (UMTS).

These modelling activities have combined two approaches, simulations and experiments. To enable METRA partners and other research teams outside of the METRA consortium to investigate the potentialities of MIMO radio channels, a stochastic model has been proposed. Its major strength lies in the fact that it collapses all the correlation information about the environment under study, usually extracted from the geometrical characteristics of the set-up, in two correlation matrices, which are measured at the terminations of the link. The time-domain characterisation of the set-up is conventionally defined with Power Delay Spectrum\(^1\) (PDS) and Doppler spectrum. Additionally, a review of the open literature has been made, to guide the choice of input parameters and spectra of the proposed model according to the considered environment. Moreover, this stochastic model has been implemented into a COSSAP\(^\circledR\) primitive block. COSSAP\(^\circledR\) is a widely used simulation tool.

Regarding the experimental approach, a MIMO channel sounder has been built on the basis of the Technology in Smart antennas for Universal Advanced Mobile Infrastructure (TSUNAMI) II testbed. The most innovative contribution in that respect has been the design, the building and the operation of a trolley carrying a slide allowing to perform measurement runs of more than 10 wavelengths with smart antenna set-ups. This enables to perform a thorough characterisation of MIMO channels in both space- and time-domain.

A total of 99 positions related to 6 different environments have been investigated with the channel sounder. For each of these positions, the typical parameters to be used as inputs of the proposed stochastic model have been derived, and simulations have been performed. Measurement and simulation results have then been compared through the eigenanalysis of the recorded and synthesised MIMO channel impulse responses (IR). It is shown that the agreement between them is fairly good, which enables to conclude the validity of the model.

\(^1\) The terminology “Power Delay Spectrum” is adopted from [Proakis95, p. 762].
DISCLAIMER

The work associated with this report has been carried out in accordance with the highest technical standards and the METRA partners have endeavoured to achieve the degree of accuracy and reliability appropriate to the work in question. However since the partners have no control over the use to which the information contained within the report is to be put by any other party, any other such party shall be deemed to have satisfied itself as to the suitability and reliability of the information in relation to any particular use, purpose or application.

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# 1 ABBREVIATIONS

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<tr>
<td>AAU</td>
<td>Aalborg University</td>
</tr>
<tr>
<td>ACTS</td>
<td>Advanced Communications Technologies and Services</td>
</tr>
<tr>
<td>AS</td>
<td>Azimuth Spread</td>
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<tr>
<td>BS</td>
<td>Base Station</td>
</tr>
<tr>
<td>BU</td>
<td>Bad Urban</td>
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<tr>
<td>cdf</td>
<td>Cumulative Distribution Function</td>
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<tr>
<td>CW</td>
<td>Continuous Wave</td>
</tr>
<tr>
<td>DoA</td>
<td>Direction of Arrival</td>
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<tr>
<td>DS</td>
<td>Delay Spread</td>
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<tr>
<td>EVD</td>
<td>EigenValue Decomposition</td>
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<tr>
<td>FCCH</td>
<td>Frequency Control CHannel</td>
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<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
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<tr>
<td>IR</td>
<td>Impulse Response</td>
</tr>
<tr>
<td>LOS</td>
<td>Line-Of-Sight</td>
</tr>
<tr>
<td>METRA</td>
<td>Multiple Element Transmit Receive Antennas</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple-Input Multiple-Output</td>
</tr>
<tr>
<td>ML</td>
<td>Maximum Length</td>
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<tr>
<td>MS</td>
<td>Mobile Station</td>
</tr>
<tr>
<td>NLOS</td>
<td>Non-Line-Of-Sight</td>
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<td>PAS</td>
<td>Power Azimuth Spectrum</td>
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<td>PADS</td>
<td>Power Azimuth-Delay Spectrum</td>
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<td>PDS</td>
<td>Power Delay Spectrum</td>
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<tr>
<td>PN</td>
<td>Pseudo-Noise</td>
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<tr>
<td>P/S</td>
<td>Parallel-to-Serial</td>
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<tr>
<td>Rx</td>
<td>Receiver</td>
</tr>
<tr>
<td>SCH</td>
<td>Synchronisation Control Channel</td>
</tr>
<tr>
<td>SDMA</td>
<td>Space Division Multiple Access</td>
</tr>
<tr>
<td>SISO</td>
<td>Single-Input Single-Output</td>
</tr>
<tr>
<td>S/P</td>
<td>Serial-to-Parallel</td>
</tr>
<tr>
<td>std</td>
<td>Standard deviation</td>
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<tr>
<td>SUNBEAM</td>
<td>Smart Universal BEAMforming</td>
</tr>
<tr>
<td>TCH</td>
<td>Traffic Control Channel</td>
</tr>
<tr>
<td>TSUNAMI</td>
<td>Technology in Smart antennas for Universal Advanced Mobile Infrastructure</td>
</tr>
<tr>
<td>TU</td>
<td>Typical Urban</td>
</tr>
<tr>
<td>Tx</td>
<td>Transmitter</td>
</tr>
<tr>
<td>UE</td>
<td>User Equipment</td>
</tr>
<tr>
<td>UMTS</td>
<td>Universal Mobile Telecommunication Services</td>
</tr>
<tr>
<td>WSS</td>
<td>Wide Sense Stationary</td>
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2 INTRODUCTION

Although the characterisation of wireless channels started some decades ago, and has since been the subject of intense research activities, it still attracts much interest. One of the main reasons for this continuing interest is the fact that, until some years ago, most of the modelling activities have focused on the time-domain aspects. This has lead to a bunch of models, which can be sorted according to the outdoor vs. indoor dichotomy. Usually, in outdoor scenarios, the Base Station (BS) is regarded to be placed much higher than the Mobile Station (MS), such that the scatterers which account for the diffuse transmission of the signals are mostly lying close to the MS. On the contrary, the surrounding environment is usually much more similar for MS and BS in indoor scenarios, introducing thus some symmetry in the phenomena. This dichotomy lead to the development of two sets of models, the first one accounting for outdoor, mobile scenarios, while the second one describes indoor, portable ones. The models proposed by [COST89, ITU97] are among the most widely accepted for the outdoor environments. They account for the time dispersion and the time variation of mobile channels. On the other hand, the model proposed in [Saleh87] appropriately describes indoor phenomena.

These time-domain models have been applied successfully until quite recently, when the growing demand for ubiquitous high speed connections pushed researchers to investigate new means of increasing the capacity of wireless channels. As part of these efforts, the use of so-called “smart antennas” for antenna/space diversity, beamforming or even Space Division Multiple Access (SDMA), has been regarded as a powerful improvement [Martin99]. However, the classical models of radio channels were of no immediate help, as they are non-directional. In as such, they do not appropriately model the propagation phenomena in the space domain. There have been many proposals of models solving this lack. Some proposed an upgraded version of time-domain models, such as [Klein96] for outdoor and [Spencer00] for indoor environments. Others proposed new models, based either on a geometric description of the scattering process used to compute PDS and Power Azimuth Spectrum (PAS) according to propagation laws or on empirical models fitting measurement results. [Ertel98, Martin99] propose a comprehensive survey of these efforts.

The target of the METRA project is to study the feasibility of introducing multi-element adaptive antennas into user equipment and the BS for the 3rd generation mobile communication systems. Thus one main objective is to gain a better understanding of the characteristics of the MIMO radio channels in a wide variety of environments. This has been the purpose of Workpackage 2. The following document describes the main results achieved in that respect.

The presented material is split into four chapters accounting for both simulation- and measurement-oriented approaches. Chapter 2 introduces the proposed stochastic model and its implementation in COSSAP®, a widely used simulation tool. A major characteristic of the stochastic model is that, contrary to other directional models, it does not rely on a geometrical description of the environment under study. The spatial correlation information collapses into two matrices defined at the connection terminations. Guidelines for choosing the element values of these two correlation matrices according to the environment under consideration, as well as the PDS and Doppler spectrum characterising its time-domain behaviour, are given in Chapter3. Leaving the simulation world to move to the experimental one, Chapter 4 fully describes the trolley designed, built and used in order to perform the measurement
campaigns. A total of 99 positions in 6 different environments have been considered. Their measurement results are presented in Chapter 5 through the eigenanalysis of the measured IRs. These eigenvalues are compared to those derived from the analysis of the synthesised IRs generated by the COSSAP® implementation of the proposed stochastic model, as a mean to validate it through the matching of the eigenanalysis results.
3 STOCHASTIC MODELLING

The following chapter is dedicated to the modelling of MIMO radio channels. The stochastic model of MIMO radio channels initially proposed in [Pedersen00b] is described in details in section 3.1. This model accounts for both time- and space-domain characteristics. Section 3.2 is dedicated to its COSSAP implementation. Parameters, input and output files of the implementation are documented in section 3.2.5.

3.1 Description of the proposed stochastic model

Let us consider the set-up pictured in Figure 1 with $M$ antennas at the BS and $N$ antennas at the MS. The signals at the BS antenna array are denoted $y(t) = [y_1(t), y_2(t), ..., y_M(t)]^T$, where $y_m(t)$ is the signal at the $m^{th}$ antenna port and $[.]^T$ denotes transposition. Similarly, the signals at the MS are the components of the vector $s(t) = [s_1(t), s_2(t), ..., s_N(t)]^T$.

![Figure 1: Arrays in a scattering environment](image)

The wideband MIMO radio channel which describes the connection between the MS and the BS can be expressed as $H(\tau) = \sum_{l=1}^{L} A_l \delta(\tau - \tau_l)$ where $H(\tau) \in \mathbb{C}^{MN}$, $A_l = [\alpha_{mn}^{(l)}]_{M \times N}$ is a complex matrix which describes the linear transformation between the two considered antenna arrays at delay $\tau_l$ and $\alpha_{mn}^{(l)}$ is the complex transmission coefficient from antenna $n$ at the MS to antenna $m$ at the BS.

Notice that this is a simple tapped delay line model, where the channel coefficients at the $L$ delays are represented by matrices. The relation between the vectors $y(t)$ and $s(t)$ can thus be expressed as

$$ y(t) = \int H(\tau) s(t - \tau) d\tau $$

$$ s(t) = \int H^T(\tau) y(t - \tau) d\tau $$

depending on whether the transmission is from MS to BS, or vice versa. The potential gain from applying diversity concepts is strongly dependent on the correlation coefficient between the elements of $H(\tau)$ and thus of $A_l$. 
The spatial correlation function observed at the BS has been studied extensively in the literature for scenarios where the MS is surrounded by scatterers, while there are no local scatterers in the vicinity of the BS antenna array, i.e. typical urban environment. This basically means that the PAS observed at the BS is confined to a relatively narrow beamwidth, as further explained in the next chapter. Consequently, the correlation coefficient between antennas \( m_1 \) and \( m_2 \) at the BS,

\[
\rho_{m_1m_2}^{BS} = \left\langle \left| e_{m_1}^{(l)} \right|^2, \left| e_{m_2}^{(l)} \right|^2 \right\rangle \tag{3}
\]

is easily obtained from the literature assuming that the BS antenna array is elevated above the local scatterers. Notice from (3) that it is assumed that the spatial correlation function at the BS is independent of \( n \). This is a reasonable assumption provided that all antennas at the MS are closely co-located and have the same radiation pattern, so they illuminate the same surrounding scatterers and therefore also generate the same PAS at the BS, i.e. the same spatial correlation function.

The spatial power correlation function observed at the MS has also been extensively studied in the literature. Assuming an MS surrounded by local scatterers, antennas separated by more than \( \frac{\lambda}{2} \), where \( \lambda \) represents the wavelength, can be regarded as practically uncorrelated [Clarke68], so

\[
\rho_{n_1n_2}^{MS} = \left\langle \left| e_{m_1}^{(l)} \right|^2, \left| e_{m_2}^{(l)} \right|^2 \right\rangle \tag{4}
\]

nearly equals zero for \( n_1 \neq n_2 \). However, experimental results reported in [Eggers95] show that in some situations antennas separated with \( \frac{\lambda}{2} \) might be highly correlated, even in indoor environments. Under such conditions, an approximate expression of the spatial correlation function averaged over all possible azimuth orientations of the MS array is derived in [Durgin99]. The latter expression is a function of the azimuth dispersion \( \Lambda \in [0;1] \), where \( \Lambda = 0 \) corresponds to a scenario where the power is coming from one distinct direction only, while \( \Lambda = 1 \) when the PAS is uniformly distributed over the azimuth range \([0^\circ; 360^\circ]\) [Durgin98]. As the MS is typically non-stationary, the results presented in [Durgin99] are very useful since they are averaged over all orientations of the MS array.

Given (3) and (4), let us define the symmetrical correlation matrices \( R_{BS} = \left[ \rho_{pq}^{BS} \right]_{MxM} \) and \( R_{MS} = \left[ \rho_{pq}^{MS} \right]_{NxN} \) for later use. The spatial correlation function at the BS and at the MS does not provide sufficient information to generate the matrices \( A_j \). The correlation of two transmission coefficients connecting two different sets of antennas also needs to be determined, i.e.

\[
\rho_{n_1n_2} = \frac{\left\langle \left| e_{m_1}^{(l)} \right|^2, \left| e_{m_2}^{(l)} \right|^2 \right\rangle}{\rho_{n_1n_2}^{MS} \rho_{m_1m_2}^{BS}} \tag{5}
\]
Neither a theoretical expression for (5) nor experimental results have been published according to the authors' knowledge. An approximation of (5) is therefore proposed in (6). This approximation is motivated by [Eggers93], where it was found that the correlation between two spatially separated antennas with different polarisations is given by the product of the spatial and polarisation correlation coefficients. Relation (6) can be shown to be exact using definitions (3) and (4) and assuming that the average power of the transmission coefficients is identical for a given delay, so \( P_l = E\left|x_{mn}^{(l)}\right|^2 \) for all \( n \in \{1,2,\ldots,N\} \) and \( m \in \{1,2,\ldots,M\} \).

3.2 Description of the COSSAP® implementation

3.2.1 General description

The proposed stochastic model is implemented in a COSSAP® primitive model whose functional sketch is shown in Figure 2. It is a complex single-input single-output (SISO) Finite Impulse Response (FIR) filter whose taps are computed so as to simulate time dispersion, fading and spatial correlation. To simulate MIMO radio channels, it has to be preceded by a parallel-to-serial (P/S) converter with turns the \( M \) signals transmitted from the MS into a single complex signal. Similarly, at the output side, the complex signal is serial-to-parallel (S/P) converted into \( N \) signals impinging the BS. This enables to maintain only one block suitable for a wide range of scenarios. Moreover, the FIR structure enables the user willing to shape the envelope of the IR either to define a synthetic PDS, or to use profiles recorded during measurement campaigns. In the former case, the attenuation and the delay with respect to the first tap are given for each tap in external files read at the initialisation step of the block, see section 3.2.5.2. In the latter case, sampled profiles would be fed directly to the FIR filter. However, this functionality has not been implemented yet. A steering matrix is also applied to take into account Direction of Arrival (DoA).

![COSSAP PRIMITIVE MODEL](image)

**Figure 2: Functional sketch of COSSAP® primitive model MIMO_CHANNEL**
The FIR filter is characterised by three parameters, \( L \), \( \text{MAX}_L \) and \( \text{Sampling\_Frequency\_Hz} \). \( L \) sets the number of taps. It is also the number of elements in the external files read at initialisation. The minimal spacing between two taps is given by the inverse of \( \text{Sampling\_Frequency\_Hz} \). Finally, the maximal Delay Spread (DS)\(^2\) supported by the FIR filter is defined as

\[
\frac{\text{MAX}_L}{\text{Sampling\_Frequency\_Hz}} \tag{7}
\]

This figure defines the span of the internal memory, that is to say the time span over which inputs have to be stored to ensure the FIR filter follows up. In terms of samples, this span is given by \( \text{MAX}_L \).

### 3.2.2 Fast fading

Following the approach in [Klingenbrunn99], the correlated transmission coefficients can be obtained according to

\[
\widetilde{A}_i = \sqrt{P_i} \mathbf{c}_i \tag{8}
\]

where

\[
\widetilde{A}_i = \left[ \alpha_{i1} \alpha_{i2} \cdots \alpha_{iM1} \alpha_{i21} \cdots \alpha_{iMN} \right]_{\text{MNx1}} \tag{9}
\]

\( \mathbf{C} \in \mathbb{R}^{\text{MNxMN}} \) is a symmetrical mapping matrix defining the spatial correlation and

\[
\mathbf{a}_i = \left[ \alpha_{i1}^{(l)} \alpha_{i2}^{(l)} \cdots \alpha_{iMN}^{(l)} \right]_{\text{MNx1}} \tag{10}
\]

with \( \alpha_{x}^{(l)} \) defined as random processes. The fading characteristics of the taps \( \alpha_{x}^{(l)} \) are defined by shaping an oversampled Doppler spectrum in the spatial frequency domain. The inverse Fourier transform of this Doppler spectrum defines the complex random fading coefficients \( \alpha_{x}^{(l)} \) in the spatial domain. Then, it is a simple operation to convert them into the time domain, by taking into account the speed of the mobile.

The fading characteristics of the taps \( \alpha_{x}^{(l)} \) are defined by shaping an oversampled Doppler spectrum in the spatial frequency domain. Depending on the value of parameter \( \text{Doppler\_Type} \), a single predefined Doppler spectrum is used to generate the \( LMN \) taps or \( MN \) user-defined Doppler spectrum are uploaded to create the fading processes.

#### 3.2.2.1 Predefined Doppler spectrum

Two predefined shapes are currently available, Rayleigh or flat. They are defined equivalent in terms of energy. \( \text{Doppler\_Oversampling} \), the oversampling factor, and \( \text{IFFT\_Length} \), the width of the frequency window over which the Doppler spectrum is defined, are two other

\(^2\) The DS is the root second central moment of the PDS.
parameters of the primitive model. Finally, the Doppler effect is completely characterised thanks to parameters Carrier_Frequency_Hz and Velocity_kmh. Indeed, having shaped the Doppler spectrum in the spatial frequency domain\(^3\), its inverse Fourier transform defines the complex random fading coefficients \( a^{(t)}_x \) of (10) in the spatial domain. Then, it is a simple operation to convert them into the time domain, by taking into account the speed of the mobile.

\[
\begin{align*}
    f_m &= \frac{\lambda}{\pi} \\
    DOF f_m &= DOF \frac{\lambda}{\pi}
\end{align*}
\]

\[
\begin{align*}
    \frac{1}{\lambda} \quad DOF \frac{1}{\lambda} \\
    [m^{-1}] \quad [m^{-1}]
\end{align*}
\]

\[
\begin{align*}
    f &\xrightarrow{\mathcal{F}} f'[Hz] \\
    t &\xrightarrow{\mathcal{F}} t[	ext{s}]
\end{align*}
\]

\[
\begin{align*}
    \frac{\lambda}{100 \text{ns}} \\
    \frac{v}{100 \text{ns}}
\end{align*}
\]

\[
\begin{align*}
    \frac{v \Delta t}{100 \text{ns}} \\
    \frac{\lambda}{100 \text{ns}}
\end{align*}
\]

\[
\begin{align*}
    DOF &= \text{Doppler Oversampling Factor}
\end{align*}
\]

**Figure 3: Computation of the fading from an oversampled Doppler spectrum**

The conversion from space to time domain may require interpolation between spatial samples of the fading. Consider a carrier having a frequency of 2.05 GHz. It has \( \lambda = 0.146 \text{m} \). Oversampled by a factor 20, this gives a spatial distance between samples of 7.317 mm. On the other hand, a mobile moving at a speed of 3 km/h covers only 83 nm every 100 ns (Sampling time with a sampling frequency of 10 MHz). Hence, fading properties would need to be updated only every 88 000 activations! However, one should consider this fading process from the symbol point of view. In UMTS uplink, FDD mode, a symbol can be spread by 4 to 256 chips, which means that it can last from 1.04 to 66.66 µs. At 3 km/h speed, the mobile covers 5.56 mm in 66.66 µs, which means that the update of the fading properties consumes at most 2 fading samples/symbol, and 9 samples/slot of 10 symbols.

Note also that the \( LMN \) uncorrelated complex coefficients \( a^{(t)}_x \) are taken from a single Doppler vector. The vector is divided into \( LMN \) sub-vectors, and coefficients are read within these sub-vectors, as illustrated in Fig. 4. If the length of the simulation requests it, the reading goes on in the neighbouring sub-vector after the initial sub-vector has been exhausted.

---

\(^3\) Shaping only concerns the amplitude. The phase is assumed to be a uniformly distributed random variable.
Figure 4: Reading of fading samples in the Doppler vector

Particular care should be taken when performing this reading process to guarantee that the uncorrelated hypothesis is fulfilled. First of all, the length of these sub-vectors should be a significant multiple of $\lambda$. Moreover, the reading of the taps $a^{(l)}_x$ in the global vector should avoid to be performed according to a regular pattern. In this case, the generation process would be vulnerable to $\sin(x)/x$ shapes, where zeros are regularly distributed along the axis. Regularly distributed taps would then be all null once in a while, which would induce a strong correlation between them. In order to avoid this situation, a deterministic offset is added to the regular sampling pattern to make it irregular, as shown in Figure 5.

Figure 5: Use of irregular sampling to guarantee that the fading samples are uncorrelated

3.2.2.2 User-defined Doppler spectrum

The main difference in the case of user-defined Doppler spectrum with respect to the previous one lies in the fact that instead of using a single pattern, $MN$ patterns are provided, each one of these accounting for the fading process of the channel connection one of the $N$ elements of the MS to one of the $M$ elements of the BS. As it will become clear in Chapter 6, this enables the user to embed the correlation information he might extract from measurements to mimic the behaviour of a MIMO channel he would have previously investigated (so called “phase 1” simulations in Chapter 6).

3.2.3 Spatial correlation

The symmetrical mapping matrix $C$ results in a correlation matrix $\Gamma = CC^T$ where the $(x,y)^{th}$ element of $\Gamma$ is the root power correlation coefficient $\sqrt{\rho_{n_1m_1}}$ between the $x^{th}$ and $y^{th}$ element of $\tilde{A}_i$. These coefficients are computed according to (6) from the symmetrical
correlation matrices $\mathbf{R}_{BS}$ and $\mathbf{R}_{MS}$ fed through external files. The symmetrical mapping matrix $\mathbf{C}$ is easily obtained by applying square root matrix decomposition [Golub96], provided that $\mathbf{\Gamma}$ is non-singular.

Section 4.1 will discuss means to generate the correlation matrices $\mathbf{R}_{BS}$ and $\mathbf{R}_{MS}$. However, it is worth mentioning without delay that the values of the elements of these matrices strongly depend on the way the waves are spread and also on the values on angles of incidence. Should the values of the angles of incidence change, the correlation matrices provided as inputs to the model should be updated accordingly.

### 3.2.4 Steering matrix

The proposed stochastic model only reproduces the correlation metrics and fast fading characteristics of the radio channel, while the phase derivative across the antenna arrays is not necessarily reflected correctly in the model. The current model gives rise to a mean phase variation of $0^\circ$ across the antenna array. This basically means that the mean direction-of-arrival (DoA) of the impinging field correspond to broadside. However, the stochastic model is easily modified to comply with scenarios like the one pictured in Figure 6, where the mean DoA at the BS $\phi_{BS} \neq 0^\circ$.

![Figure 6: Sketch of a scenario where all scatterers are located near the MS so the impinging field at the BS is confined to a narrow azimuth region with a well defined mean DoA](image)

Relation (1) is then modified to

$$y(t) = \mathbf{W}(\phi_{BS}) \int H(\tau) s(t-\tau) d\tau$$

where the steering diagonal matrix is expressed as

$$\mathbf{W}(\phi) = \begin{bmatrix} w_1(\phi) & 0 & \cdots & 0 \\ 0 & w_2(\phi) & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & w_M(\phi) \end{bmatrix}_{M \times M}$$

with $w_m(\phi)$ describing the average phase shift relative to antenna number one assuming that the mean azimuth DoA of the impinging field equals $\phi$. Thus, for an uniform linear antenna array with element spacing $d$, 

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\[ w_m(\varphi) = f_m(\varphi) \exp[-j(m-1)d\lambda^{-1}2\pi \sin(\varphi)] \]  \hspace{1cm} (13)

where \( f_m(\varphi) \) is the complex radiation pattern of antenna \( m \) and \( j \) is the imaginary unit.

From the point of view of the primitive model, most of the numerical values in relation (13) come from block parameters. \( \lambda \) is derived from \text{Carrier\_Frequency\_Hz} whereas \text{Mean\_DoA\_BS\_deg} and \text{Element\_Spacing\_BS\_m} give the mean azimuth DoA and the BS antenna element spacing. Another parameter, \text{Theta\_BS\_deg} defines the mean elevation angle \( \vartheta \). In most of the cases, it will be equal to \( \frac{\pi}{2} \). Both azimuth \( \varphi \) and elevation \( \vartheta \) are requested in order to compute the complex gain of each element. Indeed, for each \((\vartheta, \varphi)\), the gain writes \( |e_\vartheta|^2 + |e_\varphi|^2 \) where \( e_\vartheta \) and \( e_\varphi \) are the complex orthogonal components of the electrical field impinging a BS antenna element. These field values have been measured at discrete values of both azimuth and elevation angles, using angular sampling distance \text{Step\_gai\_deg} and are fed through an external file. To derive the requested gain, a two-dimensional linear interpolation is thus performed between sampled values of the gain to obtain the desired value.

In situations where the antenna signals at the array are assumed statistically independent (uncorrelated), it does not make sense to define a mean DoA, so (1) is applicable without the modification proposed in (11).

3.2.5 Interface documentation

As illustrated in Figure 2, the COSSAP implementation of the stochastic model is characterised by a set of parameters and some input files, called “datasets” in COSSAP terminology. They are listed in the following sections, with some information about their format. On the other hand, the COSSAP primitive model not only produces the result of the convolution of the input signal by the MIMO channel IRs, but it also generates some ASCII .TXT files enabling to control a posteriori the generation of the fading processes. The writing of these control files is activated only in debugging mode (Compilation directive DEBUG set on).

3.2.5.1 Parameters

Table 1 lists the parameters used to configure an instance of the COSSAP implementation of the proposed stochastic model. The type of these parameters, integer (I) or real (R) is also mentioned.
### Table 1: Parameters of MIMO_CHANNEL

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Type</th>
<th>Role</th>
</tr>
</thead>
<tbody>
<tr>
<td>N</td>
<td>I</td>
<td>Number of elements at MS</td>
</tr>
<tr>
<td>M</td>
<td>I</td>
<td>Number of elements at BS</td>
</tr>
<tr>
<td>L</td>
<td>I</td>
<td>Number of taps</td>
</tr>
<tr>
<td>MAX_L</td>
<td>I</td>
<td>Maximum number of supported taps in case of regular spacing</td>
</tr>
<tr>
<td>Sampling_Frequency_Hz</td>
<td>I</td>
<td>Sampling frequency of the taps, its invert defines their minimal spacing</td>
</tr>
<tr>
<td>Velocity_kmh</td>
<td>R</td>
<td>MS speed, in km/h</td>
</tr>
<tr>
<td>Carrier_Frequency_Hz</td>
<td>I</td>
<td>Carrier frequency, in Hz</td>
</tr>
<tr>
<td>IFFT_Length</td>
<td>I</td>
<td>Width of the sampled spatial frequency window used to define the Doppler spectrum</td>
</tr>
<tr>
<td>Doppler_Oversampling</td>
<td>R</td>
<td>Oversampling factor of the Doppler spectrum</td>
</tr>
<tr>
<td>Doppler_Spectrum_Type</td>
<td>I</td>
<td>Shape of the Doppler spectrum: Rayleigh (1), flat (2) or user-defined (3)</td>
</tr>
<tr>
<td>Mean_DoA_BS_deg</td>
<td>R</td>
<td>Mean DoA angle at the BS, in degrees</td>
</tr>
<tr>
<td>Theta_BS_deg</td>
<td>R</td>
<td>Elevation angle of the impinging waves at the BS, in degrees</td>
</tr>
<tr>
<td>ElementSpacing_BS_m</td>
<td>R</td>
<td>Distance between antenna elements at the BS, in m</td>
</tr>
<tr>
<td>Step_gai_deg</td>
<td>I</td>
<td>Sampling angular distance of the BS radiation pattern input dataset</td>
</tr>
<tr>
<td>Random_Seed</td>
<td>I</td>
<td>Seed for the random phase generator</td>
</tr>
</tbody>
</table>

### 3.2.5.2 Input datasets

Besides the parameters listed in the previous section, some other information has to be provided to MIMO_CHANNEL using input datasets: the correlation matrices \( \mathbf{R}_{\text{BS}} \) and \( \mathbf{R}_{\text{MS}} \), the radiation pattern of the receiving antenna element (to compute the steering matrix), the simulated PDS (powers and delays are defined separately), and the Doppler spectra characterising the fading processes. The input datasets are ASCII files containing values organised in matrices. The dimension of these matrices is given in Table 2, where \( N_{\text{Samp}} \) stands for the number of Doppler spectrum samples. The factor 6 in \( \text{antenna}_\text{BS}.\text{gai} \) comes from the fact that the radiation pattern is given in complex values for both horizontal and vertical polarisations, as a function of the azimuth \( \phi \) and the elevation \( \vartheta \). Similarly, \( \text{Doppler}.\text{am} \) contains complex values, thus the factor 2.
### 3.2.5.3 Output files

The elements of the channel matrix $H(\tau)$ are output at every activation in a file called `channel_SIMNO_IT.txt`, where SIMNO stands for the identification number of the COSSAP simulation and IT for the iteration within a simulation (See COSSAP® manuals for more details on this notation). Moreover, the Doppler spectrum read by MIMO_CHANNEL, the fading vector and the computed outputs can also be stored in separate files provided the compilation directive DEBUG is set on. Table 3 summarises the output files.

<table>
<thead>
<tr>
<th>File</th>
<th>Content</th>
<th>Remark</th>
</tr>
</thead>
<tbody>
<tr>
<td>Channel_SIMNO_IT.txt</td>
<td>Elements of $H(\tau)$</td>
<td></td>
</tr>
<tr>
<td>Test_SIMNO_IT.txt</td>
<td>Doppler spectrum read from the input dataset Doppler.am</td>
<td>Only produced if compilation directive DEBUG set on</td>
</tr>
<tr>
<td>Fading_SIMNO_IT.txt</td>
<td>Fading vector computed from the input Doppler spectra</td>
<td></td>
</tr>
<tr>
<td>Output_SIMNO_IT.txt</td>
<td>Result of the convolution of the input signal by the channel</td>
<td></td>
</tr>
</tbody>
</table>

**Table 3: Output files of MIMO_CHANNEL**

### 3.2.6 Distribution

The COSSAP® implementation of the stochastic model has been distributed among partners of the METRA project. Their feedback helps to improve the software and its documentation. The source code is also available to third parties, provided they agree on the distribution terms listed in Appendix 2.

### 3.2.7 Web site

Ongoing developments of the model and its implementation, related publications and updates are available at

http://kom.auc.dk/~schum/MIMO/index.html
4 PROPAGATION ISSUES

The purpose of this chapter is to list references in the open literature relevant for the choice of the input parameters of the stochastic model introduced in Chapter 3. Although PAS, Doppler spectrum and PDS are requested, this chapter is mainly focused on the two first spectra, as PDS has already received much attention and researchers are quite accustomed with typical models. The PAS and the Doppler spectrum are thus thoroughly investigated in sections 4.1 and 4.2 respectively, while the PDS is shortly dealt with in section 4.3.

Notice that the Power Azimuth-Delay Spectrum (PADS) embeds the information separately provided by the PAS and the PDS. In the following, the PADS is just split into the product of the PAS and the PDS, although the physical mechanisms leading to these dispersions are correlated. This separateness is discussed in [Pedersen00a] for typical urban environments and in [Spencer00] for indoor ones.

4.1 Power-Azimuth Spectrum (PAS) and correlation matrices

The elements of the symmetrical correlation matrices $R_{BS}$ and $R_{MS}$ introduced in the previous chapter are determined by the correlation properties of the fading characteristics of the wireless radio channel. Following the approach of [Lee73], several authors have presented closed-form expressions of this correlation. These relations enable to compute the value of the correlation of the power (or the envelope) of the signal as a function of the normalised antenna element separation $\frac{d}{\lambda}$, indexed on the angle of incidence and on the Azimuth Spread (AS)\(^4\). These correlation results mostly differ by the underlying assumption related to the PAS.

While the seminal work of [Lee73] modelled the PAS as the $n^{th}$ power of a cosine function, two other distributions are considered in [Fuhl98], namely a truncated Gaussian PAS originally proposed by [Adachi86] and a uniform one introduced by [Salz94]. Correlation results are presented for these two distributions indexed on three different angles of incidence ($0^\circ$, $60^\circ$ and $90^\circ$). Similar results are presented in [Stege00] based on the same assumption of uniformly distributed impinging power, for two different angles of incidence ($0^\circ$ and $60^\circ$). Finally, the uniform distribution is also considered by [Durgin99]. A method is proposed to compute the cross-correlation of elements separated by a distance $d$ as a function of the normalised antenna separation $\frac{d}{\lambda}$ and of a parameter $\Lambda$ which is derived from the PAS, as described in [Durgin98]. To illustrate this method, the correlation is computed assuming a uniform distribution of the impinging power within some azimuth sector. The width of this sector serves as indexing parameter. On the other hand, [Pedersen98] investigated cross-correlation properties in the case of a Laplacian distribution. Looking at all these results, one can notice that the correlation coefficient decreases with increasing AS and with decreasing angle of incidence of the signals, from broadside to end-fire [Fuhl98].

These correlation functions have been derived on the basis of assumptions related to the shape of the PAS and are indexed on its standard deviation and on the mean angle of

\(^4\) The AS is defined as the root second central moment of the PAS.
incidence. Since the latter is usually derived from the Line-Of-Sight (LOS) of the system under study, one comes to the conclusion that the correlation properties of the fading, and consequently the values of the symmetrical correlation matrices $R_{BS}$ and $R_{MS}$, are completely defined by the PAS and its standard deviation. In the remaining of this section, the shape and the spreading of the PAS will be investigated.

As far as the shape of the PAS is concerned, one should distinguish the case of the BS in outdoor environments from all other cases.

In outdoor scenarios, the impinging signals at the BS come within a narrow azimuth sector, leading to definitions of the PAS which are mainly related to the spatial distribution of the scatterers around the MS. It has already been mentioned that a uniform PAS is proposed in [Salz94]. Such a PAS is derived from a uniform spatial distribution of the scatterers around the MS. However, this uniform distribution of the scatterers has been regarded as hardly justified from a physical point of view [Fuhl98]. The PAS had earlier been modelled as the $n^{th}$ power of a cosine function [Lee73], but some authors argue that it does not enable to reach closed-form solutions. It has then been proposed to model the PAS as truncated Gaussian functions [Adachi86]. This corresponds to a Rayleigh distribution of the local scatterers [Laurila98]. More recently, [Pedersen97] introduced the Laplacian function, for it reproduces more accurately measurement results than the truncated Gaussian function: the Laplacian function exhibits a sharp peak in the LOS direction and has long tails.

When considering the PAS at the BS in micro- or pico-cells, or at either end of a wireless communication link in indoor environments, it is no longer possible to constrain scatterers within a well-defined area surrounding the sole MS. Both MS and BS are surrounded by scatterers and the PAS tends to become uniform. A method is proposed in [Döhler00] to convert a PAS defined outdoor in its corresponding indoor version, taking into account transmission through the wall. Due to diffraction, the indoor PAS exhibits more spreading than its outdoor counterpart. The indoor PAS appears to be a periodical repetition of the outdoor one, with the latter standing as a pattern repeated along the propagating non-specular components. The AS is subsequently increased. A similar repetition is proposed by [Spencer00], which extends to the space domain the clustering pattern proposed in the time domain by [Saleh87] for indoor environments. These spatial clusters have an angle of incidence uniformly distributed in $[0, 2\pi]$ but within each cluster, the impinging power is distributed according to a Laplacian function.

Table 4 summarises the different options regarding the shape of the PAS.

<table>
<thead>
<tr>
<th>BS</th>
<th>Outdoor</th>
<th>Microcell</th>
</tr>
</thead>
</table>
|    | Macrocell | • Laplacian [Pedersen97]  
|    | Microcell | • $n^{th}$ power of a cosine function [Lee73]  
|    | Picocell  | • Truncated Gaussian [Adachi86]  
| Indoor | Uniform | • Uniform [Salz94]  
| Indoor | Uniform | Almost uniform [Fuhl98]  

Table 4: PAS shapes
The AS has been measured in a wide variety of environments. Table 5 gathers significant results for each of them. One can notice that the AS is very low in rural environments [Pajusco98] but tends to increase as scattering effects become more and more significant. For instance, it is shown in [Nilsson99] that the AS measured in urban environments is twice as high as in suburban ones. Furthermore, when considering micro- and picocells, for which the BS often stands below rooftop, it is no longer possible to consider scatterers around the sole BS [Fuhl98]. As a result, the PAS tends to become uniform. However, applying the clustering approach, ASs between 20° and 60° are measured within clusters [Wang96, Spencer00].

<table>
<thead>
<tr>
<th>Reference</th>
<th>Carrier frequency [MHz]</th>
<th>Outdoor</th>
<th>Indoor</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Macrocell</td>
<td>Microcell</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Urban</td>
<td>Suburban</td>
</tr>
<tr>
<td>Wang96</td>
<td>1000</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Pedersen98</td>
<td>1800</td>
<td>5°-10°</td>
<td></td>
</tr>
<tr>
<td>Nilsson99</td>
<td>1800</td>
<td>8°</td>
<td>5°</td>
</tr>
<tr>
<td>Eggers95</td>
<td>1845</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Martin98</td>
<td>1873</td>
<td>3°-15°</td>
<td></td>
</tr>
<tr>
<td>Petsersen99</td>
<td>2100</td>
<td>7°-12°</td>
<td>13°-18°</td>
</tr>
<tr>
<td>Kalliola98</td>
<td>2154</td>
<td>10,3°</td>
<td></td>
</tr>
<tr>
<td>Pajusco98</td>
<td>2200</td>
<td></td>
<td>3°</td>
</tr>
<tr>
<td>Spencer00</td>
<td>7000</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 5: Median AS (In indoor environments, the mentioned AS is the one measured within a cluster)

The authors of [Kalliola98] claim that the AS they measured is not typical of suburban environments because of a too short distance between MS and BS. Indeed, this distance impacts on the AS. References [Pedersen98, Martin98] show that the AS increases with decreasing distance between MS and BS, provided that this distance is still much greater than the radius of the circle within the scatterers surrounding the MS are placed, and that the assumption of scatterers distributed around MS holds. Consequently, for a same element separation distance \(d\), MS elements are less correlated than BS ones as they experience a greater AS [Martin98]. However, the conditions mentioned here above should be fulfilled, unless one faces completely opposite behaviours [Pedersen98, Petsersen99]. On the other hand, it is shown in [Pedersen98, Petsersen99] that the height of the BS has also an influence on the AS, as the spreading increases with decreasing antenna height.

4.2 Doppler Spectrum

The expression of the Doppler spectrum has been derived by Clarke [Clarke68]. It is well known that the Doppler spectrum lies within a \([-f_D, f_D]\) bandwidth, when \(f_D\) is the maximal Doppler shift, defined as

\[
f_D = \frac{\nu}{\lambda}
\]  

(14)

where \(\nu\) stands for the velocity of the movement. In as such, this section cannot bring any income as far as the selection of this value is concerned, since the main parameter of the Doppler spectrum is heavily depending on simulation assumptions.
On the other hand, the shape of the Doppler spectrum can be derived knowing the PAS and the radiation pattern of the receiving antenna [Petrus97]. Assuming a flat PAS (scatterers uniformly distributed around the receiving antenna) and an isotropic/omnidirectional antenna leads to the classical U-shaped Clarke’s Doppler spectrum expression as

\[ P(f) \propto \left[ 1 - \left( \frac{f}{f_D} \right) \right]^{-\frac{1}{2}} \quad (15) \]

This assumption is reasonably fulfilled in the case of MS moving in urban environments [TSUNAMI97b, pp. 5-6 – 5-7], for both Typical Urban (TU) and Bad Urban (BU) cases [TSUNAMI97a, pp. A-11 – A-12]. It also applies to both MS and BS in the case of indoor transmissions, as both communication ends are then surrounded by scatterers.

This is however usually not the case for the BS in outdoor scenarios. In such scenarios, the PAS measured at the BS is usually constrained within a narrow bandwidth, which leads to Doppler spectrum shapes differing from the classical U-shaped Clarke’s one. Conversely, MS illuminated by BS exhibiting such narrow beams that they do not illuminate a significant fraction of the surrounding scatterers lead to unconventional Doppler spectrum [Petrus97]. It is also shown in [TSUNAMI97b, pp. 5-7 – 5-8] that a truncated Gaussian shape fits measurements in rural environments, where a few scattering directions dominate. [Clarke97] shows on the other hand that the Doppler spectrum appears to be uniform when one considers 3-D isotropically scattered field instead of the usual 2-D representation. Such 3-D models apply in indoor scenarios. Based on a similar 3-D model, a rectangular Doppler spectrum (with an additional spectral line in LOS cases) is proposed in [Vatalaro97] for low-gain handheld terminals in microcellular urban environments.

### 4.3 Power-Delay Spectrum (PDS)

The PDS has been widely studied as part of the time-domain characterisation of wireless radio channels. In accordance with [COST89], the PDS is accurately modelled by a one-sided exponential decaying function

\[ P(t) = \begin{cases} \exp\left(-\frac{t}{\sigma_D}\right) & t > 0 \\ 0 & \text{otherwise} \end{cases} \quad (16) \]

where \( \sigma_D \) represents the DS. Values of the DS have been proposed for the different environments. Table 6 summarises some of these results. However, in some environment classes, the power is not monotonically decaying with time, as waves arrive at the termination gathered in clusters. This clustering effect has been modelled in the so-called “hilly” outdoor environments of [COST89]. A similar clustering process has been identified in [Saleh87] for indoor environments.
Based on this exponential decay modelling of the power, tap-based channel models have been proposed [COST89, ITU97]. They differ in the number, the spacing and the relative power of the taps, according to the environment under study. Such taps settings can be used to define the PDS of the proposed stochastic model.
5 EXPERIMENTAL HARDWARE AND GEOMETRY

As part of the WP2, a vast measurement campaign was planned to investigate the MIMO concept in real environments. The experience and excellence gained in the two European Advanced Communications Technologies and Services (ACTS) TSUNAMI II [WWW TSUNAMI II] and Smart Universal BEAMforming (SUNBEAM) [WWW SUNBEAM] projects render AAU the ideal partner to perform field measurement results. This expertise gave AAU the opportunity to be one of the first research groups to provide measurement field tests of the MIMO radio propagation channel to the research community [Kermoal00a, Kermoal00b].

One goal of these measurements was the evaluation of the spatial correlation which exists between several elements of an antenna array. Another aim of the measurements was to derive realistic parameters of the MIMO radio channel and to fed them into the COSSAP® implementation of the stochastic channel model described in section Description of the COSSAP® implementation in order to validate it.

The objective of this chapter is therefore to describe the measurement campaign set-up used for the WP2 of METRA. This chapter is organised as follows. The measurement set-up is presented in section 5.1. The environments where the measurement campaign was conducted are described in section 5.2. The antenna topologies employed are outlined in section 5.3.

5.1 The measurement set-up

5.1.1 General description

MIMO measurements with $M \times N$ set-up, where $M$ and $N$ are the number of elements at the BS and MS respectively, were performed. A simplified sketch of the MIMO set-up is presented in Figure 7, where the transmitter (Tx) at the MS is on the left and the stationary receiver (Rx) located at the BS is on the right. In the working assumptions of the METRA project presented in the deliverable D3.1 [Ylitalo00], it was decided that each environment would provide its own $M \times N$ set-up as briefly summarised in Table 7. In Table 7, the term “port” was used instead of “antenna” in order to avoid any potential confusion when dealing with polarisation diversity (1 antenna = 2 ports).

With respect to these assumptions, a scenario where $M=N=4$ was considered for the measurement campaign. As it will be explained in section 5.1.2, the decision of implementing a 4x4 set-up and the performance of the Rx end enabled to perform simultaneous measurements with two arrays. A second artefact, more thoroughly explained in section 5.1.3, enabled similarly the implementation of two arrays at the Tx end. The set-up configuration is summarised in Table 8. This, eventually, increases the statistical figure of the propagation channel measurement campaign.

In Table 8, horizontal #1 and horizontal #2 represent two horizontal polarised sleeve dipoles which are perpendicularly positioned with respect to each other. After post-processing, the combination of the two antennas would be almost equivalent to the radiation pattern of a horizontally polarised loop antenna. As indicated in Table 8, three different antennas where
used at the MS. The measurement procedure was such that a trolley was physically locked to a specific position and three measurements were consecutively performed with the different polarisations as described in Figure 8. This was feasible since the propagation channel was ensured to be stationary.

![Propagation diagram](image)

**Figure 7: Configuration set-up for parallel channel sounding MIMO measurements**

<table>
<thead>
<tr>
<th>Propagation environment</th>
<th>Environment</th>
<th>MS</th>
<th>BS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Indoor to Indoor Picocell</td>
<td>2(-4) ports</td>
<td>4 ports</td>
<td></td>
</tr>
<tr>
<td>Outdoor to Indoor Microcell</td>
<td>2 ports</td>
<td>4 ports</td>
<td></td>
</tr>
<tr>
<td>Outdoor to Outdoor Macrocell</td>
<td>2 ports</td>
<td>8 ports</td>
<td></td>
</tr>
</tbody>
</table>

**Table 7: D3.1 summary working assumptions with respect to the number of elements to introduce at each end of the MIMO channel**

<table>
<thead>
<tr>
<th>Number of arrays</th>
<th>Type</th>
<th>Number of ports per array</th>
<th>Polarisation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>meas. #1</td>
</tr>
<tr>
<td>MS</td>
<td>2</td>
<td>dipole</td>
<td>4</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BS</td>
<td>1</td>
<td>dipole</td>
<td>4</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>patch</td>
<td>2</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2</td>
</tr>
</tbody>
</table>

**Table 8: METRA experimental antenna set-up**
5.1.2 The stand alone testbed

The involvement of AAU in the TSUNAMI II and SUNBEAM project provided a stand-alone testbed which was upgraded for MIMO measurements. A more thorough description of this testbed is given in [Frederiksen98].

At the Tx, the common RF signal was sent to a 1-to-4 RF switch which toggled the different antenna ports on and off, with a switch time of 50 $\mu$s as shown in Figure 9. In this way, only one transmit antenna is active at a time, thus providing isolation between the transmit antennas. Furthermore, since the switching is relatively fast, it approximates a parallel transmission for low mobile speeds. The TCH (Traffic Channel) /FCCH (Frequency Control Channel) /SCH (Synchronisation Channel) frame in Figure 9 is transmitted on the last used antenna (i.e. ant #4 in the figure), and is used at the receiver for synchronisation purposes. Channel sounding measurements were performed every 20 ms at a carrier frequency of 2.05 GHz (UMTS band) and a chip rate of 4.096 Mchip/s. The transmitted power was 25 dBm after the switch. The complex narrowband information has been extracted from the wideband channel data.

At the receiver (BS side), the system is capable of measuring 8 parallel channels at the same time due to duplication of hardware. During the initial synchronisation of the system, the receiver software searches for the FCCH, which is characterised as a CW (Continuous Wave) and is easy to detect. When this FCCH is found, the receiver program searches for the SCH, and adjusts some local timers to obtain a lock to the transmitter. After this, the receiver goes into a locked state, where the estimation of the radio channel is performed. At first, the program starts the sampling at the time where the first transmitted sequence is expected (i.e. ant #1 in Figure 9). The sampling module stores all the transmitted data in a temporary buffer. For each block of input data, corresponding to 50 $\mu$s or one Pseudo-Noise (PN) segment, the following actions are performed on the signals from the eight Rx antennas:

Figure 8: Measurement procedure
• Calibrate for receiver imperfections (DC offsets, gain differences, phase imbalances and phase differences.

• Estimate the channel IR by correlating with the known PN sequence.

• Transfer the estimated IRs for later processing and storage.

When all 4 PN segments have been processed, the receiving branch with the highest input power is read and processed to update the local timers and to verify that the system is in the locked mode. During the processing of the PN segments each segment gets a number associated such that it is possible to distinguish these during the post-processing. For later processing it is therefore very important to note/measure which Tx antenna is active for the different PN segments.

Each PN segment holds a repeated 127 bit ML (Maximum Length) sequence sampled such that 40 extra bits are used in each end resulting in a sequence length of 127+2*40= 207 bits as shown in Figure 10. The reason for this repetition is for the sequence to appear as a repeated ML sequence and reconstruct the nice auto-correlation properties during the estimation of the radio channel as presented in Figure 11.

Since the BS hold eight parallel Rx channels, it is feasible to have two simultaneous independent arrays with 4 sensors as previously mentioned in section 5.1.1.

---

**Figure 9:** Sounding burst every 20 ms with a switching time of 50 µs between each antenna

**Figure 10:** Illustration of the transmitted PN sequence

**Figure 11:** Auto-correlation of the PN sequence used
5.1.3 Mechanical hardware and measurement procedure

The purposes of these measurement campaigns were the evaluation of the spatial correlation which exists between several elements of an antenna array and the extraction of realistic parameters of the MIMO radio channel in order to feed them into the COSSAP® implementation of the stochastic channel model. Previous work at AAU on MIMO channel characterisation, reported in [Kermoal00a], provided measurement results where a dipole described a circular motion. The spatial correlation could easily be extracted from the results of this measurement campaign but this method failed to provide any information in terms of Doppler spectrum since the angle of arrival changes as the dipole rotates. Therefore, a linear motion of the antenna array was considered instead of a circular one.

5.1.3.1 Mechanical hardware

The MS uses two trolleys as shown in Figure 12. One trolley was carrying all the electronic hardware of the transmitter. The other one, later referred as the “satellite”, was equipped with a linear slide used to move the antenna array at the MS. The two trolleys were connected by 10 m coaxial and signal cables. The purpose of using two trolleys was to avoid any interference from metallic surfaces. Preliminary measurements had shown that the set-up could be very sensitive to any metallic surface located on the trolley itself. Consequently, the satellite trolley was made of wood and the metallic part of the linear slide was shielded by a layer of microwave absorbers.

![Diagram of trolley setup](image)

Figure 12: Graphical representation of the positioning of the two trolley: MS hardware trolley and the “satellite”

A picture of the “satellite” during one measurement run is shown in Figure 13 where the shield can be seen. One constraint of the measurement campaign was to ensure that the investigated environment would be time stationary. Consequently, the measurements were made free from people moving around the environment. This was possible by performing the measurements overnight.
Figure 13: MS “satellite” during a measurement

The MS (i.e.: the actual antenna array here and not to be confused with the trolley) was moved along the slide over a distance of $11.8\lambda$. The choice of the slide length was based on the analysis of Monte-Carlo simulations. The standard deviation (std) of the correlation coefficient of a Rayleigh channel was computed over 1000 realisations. As a result, one can see from Figure 14 that for a distance of $11.8\lambda$, the std of the correlation is less than 0.09. This is considered acceptable regarding the fact that the distance must be much greater than $20\lambda$ to reach a very low std. Such a large distance is not practical if the WSS (Wide Sense Stationary) condition is to be considered when dealing with picocell and microcell environments.

Figure 14: Standard deviation of the correlation coefficient from Monte-Carlo simulations over 1000 realisations of a Rayleigh channel
5.1.3.2 Measurement procedure

The measurement procedure is described as follows and illustrated in Figure 15. For a stationary position of the satellite, the antenna array moves forward during 5 s (i.e., from A to B) while the received signals are being recorded. When the array reaches the end of the slide (B) it moves backward while the slide is moved perpendicularly to the motion of the array over a distance of 0.4λ. Starting from C, a new set of measurements is then collected (C to D). This artefact provides two sets of measurements for each MS location. As earlier mentioned in section 5.1.1, two antenna arrays will thus be generated after postprocessing as shown in Figure 18 of section 5.3.2. The specific distance of 0.4λ for the perpendicular displacement of the slide is later explained in Section 5.3.

The acceleration and deceleration phases of the array moving along the slide were compensated and postprocessed in the measurement analysis.

![Diagram showing measurement procedure](image)

**Figure 15: Measurement procedure at the MS**

5.2 Description of the investigated environments

Two classes of environment were experimentally investigated within the METRA project period: Picocell and Microcell. These measurements where undertaken in northern Denmark in the city of Aalborg. The two main locations were on the campus of Aalborg University and the Aalborg International Airport. Moreover, measured data from previous measurement campaigns at AAU were used to investigate the Macrocell environment.

The measurement campaigns were undertaken within several buildings. Three different environments were selected for their different characteristics. For each environment several MS locations were selected to provide a set of measurements where both LOS and NLOS were present. Moreover, several BS locations were selected in addition to the MS locations in order to increase the statistical information of the environment. This is shown in Figure 16 where the arrows represent the positioning of the MS.

Table 9 presents a summary of the different measured environments with a brief description of each environment. In relation to Table 9, Table 10 summarised the number of measurement positions which have been made during the METRA measurement campaign, ending with a total of 99 investigated positions.
Figure 16: Example of the different positions of the BS and the MS for a Picocell environment (i.e. FB7B2)

<table>
<thead>
<tr>
<th>Environment</th>
<th>Environmental class</th>
<th>Description of the environmental class</th>
</tr>
</thead>
<tbody>
<tr>
<td>Picocell</td>
<td>Novi2</td>
<td>Novi2 provides an example of a building with several small offices on the same floor. Notice that the outside window were so-called “energy windows”, meaning that they wear a very thin metallic “film” giving reflections of RF-signals.</td>
</tr>
<tr>
<td></td>
<td>Novi3</td>
<td>Novi3 is a reception hall. It provides a large open indoor environment with two floors, which could easily illustrate a conference hall or a shopping galleria scenario. Notice that the outside windows of the building are metallically shielded.</td>
</tr>
<tr>
<td></td>
<td>Nokia</td>
<td>Nokia illustrates a typical modern open office environment. Notice that the outside windows of the building are metallically shielded</td>
</tr>
<tr>
<td></td>
<td>FB7B2</td>
<td>FB7B2 provides also an example of a building with several small offices on the same floor, but the outside window of the building are not metallically shielded.</td>
</tr>
<tr>
<td></td>
<td>Aalborg Airport</td>
<td>This is the middle size international airport of Northern Denmark. It provides relatively large open area for an indoor environment.</td>
</tr>
<tr>
<td>Microcell</td>
<td>FrB7</td>
<td>This is an indoor to outdoor microcell. The indoor environment is the same as in FB7B2. In this configuration the MS is located inside while the BS is positioned outside. The actual array antennas are placed on a mast which itself is mounted on the top of a FIAT van.</td>
</tr>
</tbody>
</table>

Table 9: Summary of the different measured environments
<table>
<thead>
<tr>
<th>Environmental class</th>
<th>Number of BS positions</th>
<th>Number of MS positions per BS</th>
<th>Total number of positions</th>
</tr>
</thead>
<tbody>
<tr>
<td>Novi 2</td>
<td>3</td>
<td>7</td>
<td>21</td>
</tr>
<tr>
<td>Novi 3</td>
<td>2</td>
<td>9</td>
<td>18</td>
</tr>
<tr>
<td>Nokia</td>
<td>3</td>
<td>6</td>
<td>18</td>
</tr>
<tr>
<td>FB7B2</td>
<td>1</td>
<td>6</td>
<td>14</td>
</tr>
<tr>
<td>FrB7</td>
<td>2</td>
<td>8</td>
<td>16</td>
</tr>
<tr>
<td>Airport</td>
<td>2</td>
<td>8</td>
<td>12</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>99</td>
</tr>
</tbody>
</table>

Table 10: Summary of the number of investigated environments

5.3 Antenna topologies

Vertical polarised sleeve dipoles, with an average return loss of approximately 14 dB and a cross-polar discrimination of 20 dB were used during the measurement campaign. While these antennas provided omnidirectional radiation pattern when used as single element, an issue got raised when they were to be used as elements of an array due to the mutual coupling effect.

5.3.1 Mutual coupling effect

An issue encountered when measuring MIMO radio channels is the mutual coupling between the elements of the antenna array. Figure 17 presents the measured and simulated co-polarised radiation patterns in azimuth of a sleeve dipole when positioned in a 4 element linear array with a separation of 0.5 $\lambda$. The antenna array was simulated using AWAS® [AWAS95], a tool consisting of a kernel for numerical analysis of wire antennas and scatterers. Our confidence in this simulating tool is comforted by the reasonably good matching between the measured and the simulated results as illustrated in Figure 17.

Clearly, the effect of coupling drastically impacts the shape of the radiation pattern since, from measurements, a reduction in gain of 10 dB can be noticed for rays coming from 0° or 180°. Such variances in the radiation pattern would directly impact the correlation coefficient (if no postprocessing compensation were made) since the waves coming to the array elements would not be weighted identically. Therefore, in order to measure directly the correlation properties of the MIMO radio channel, it is essential to use array elements with an omnidirectional characteristic. Using AWAS®, different scenarios were investigated to provide an antenna array with an optimum omnidirectional radiation pattern.
Figure 17: Coupling effect on measured and simulated co-polarised radiation patterns in azimuth of one of the 4 elements of a linear antenna array with 0.5 $\lambda$ separation. Notice the squeezing effect on both radiation patterns.

5.3.2 Compensation of the mutual coupling

At the MS, an added constraint to the issue of mutual coupling is the separation between the elements of the antenna array to be considered. In order to provide DoA information, it was decided to use $0.4\lambda$ separation between the elements to enable the discrimination of waves coming from the back from waves coming from the front. Figure 18 presents a sketch of the antenna array used and also explains how from an interleaved antenna array it is feasible to generate a conventional linear array.

Figure 18: Top view drawing of the antenna array used during the measurement campaign and the postprocessed artefact to generate a linear antenna array at the MS
The interleaved solution provides an actual separation of $1.1\lambda$ between the elements compared to a traditional linear array. Furthermore the $3\lambda$ separation will both cancel the coupling between the two pairs of elements and create two separate arrays. The x and y axis offsets of the array elements are postprocessed to obtain two uniform linear arrays, each having four elements separated by $0.4\lambda$.

Simulated co-polarised radiation pattern results in azimuth of the four elements of the antenna array used during the measurement campaign, illustrated in Figure 19, showed that the influence of the mutual coupling has been significantly reduced thanks to the interleaved array solution.

![Simulated co-polarised radiation pattern in azimuth of the four elements of the antenna array used during the measurement campaign](image)

Figure 19: Simulated co-polarised radiation patterns in azimuth of the four elements of the antenna array used during the measurement campaign

At the BS, two arrays used different antenna topologies. The first array consisted of a uniform linear array with four elements and a spacing of $1.5\lambda$ was employed since the mutual coupling influence is low with such spacing. Moreover, contrary to the MS, no DoA analysis is planned at the BS. The second array consisted of two dual polarised ($\pm 45^\circ$) patch antennas with $3\lambda$ separation between them. Table 11 summarises the different antenna topologies used with respect to the investigated environments listed in Table 9.

<table>
<thead>
<tr>
<th>Environmental class</th>
<th>MS</th>
<th>BS dipole</th>
<th>BS patch</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>height (m)</td>
<td>antenna set-up (top view)</td>
<td>height (m)</td>
</tr>
<tr>
<td>Novi2</td>
<td>1.69</td>
<td>vertical</td>
<td>2.34</td>
</tr>
<tr>
<td>Novi3</td>
<td>1.69</td>
<td></td>
<td>2.34</td>
</tr>
<tr>
<td>Nokia</td>
<td>1.69</td>
<td>horizontal #1</td>
<td>2.34</td>
</tr>
<tr>
<td>FB7B2</td>
<td>1.69</td>
<td></td>
<td>2.04</td>
</tr>
<tr>
<td>Aalborg Airport</td>
<td>1.69</td>
<td>horizontal #2</td>
<td>2.53</td>
</tr>
<tr>
<td>FrB7</td>
<td>1.69</td>
<td></td>
<td>5.7</td>
</tr>
</tbody>
</table>

Table 11: Summary of the different antenna topologies
5.3.3 Influence of the radiation pattern on the correlation coefficient

The interleaved array (Figure 19) solution provides an improvement in the shape of the radiation pattern when compared to the case of the original linear array (Figure 17). However, a question remains on how important is the influence of the new radiation pattern. Indeed it could be argued that the residual variation (within the region of 2 to 3 dB) of the radiation pattern observed in Figure 19 could still influence the correlation coefficient. This would drastically impact on the conclusion whether the decorrelation observed on the measured data is due to the spatial separation of the array elements only or a combination of both space and radiation pattern diversities.

In order to investigate this, a Jake’s model simulation (i.e. with scatterers uniformly distributed from $[0, 2\pi]$ over $10\lambda$ (i.e. to remain within the same travelled distance than for the measurements) was generated to highlight the influence of the radiation pattern on the correlation coefficient value. Figure 20 graphically depicts the concept of this analysis.

- **Scenario 1**: Let us assume the auto-correlation function $S_{11}$ of a Rayleigh signal recorded with a dipole having an omnidirectional radiation pattern. $S_{11}$ represents a Bessel function as expected. This function indicates that, for instance, at $0.4\lambda$ separation a correlation coefficient of 0 is present.

- **Scenario 2**: The cross-correlation $S_{12}$ from a second dipole placed at $0.4\lambda$ from the first dipole should give a correlation value of 0 if the radiation pattern was omnidirectional as shown in the dash line curve. Any alteration in the correlation value due to the change in the radiation pattern would clearly be seen as shown with the solid line curve.

Figure 20: Graphical explanation of the potential influence of the radiation pattern on the correlation coefficient
Figure 21 illustrates the simulated results of the cross-correlation function between the signal received at each antenna with respect to the first antenna in the case of an interleaved array. Additionally, similar results when a linear array using $0.4\lambda$ separation is simulated are presented in Figure 22. One can see from Figure 21, that the dashed line representing $S_{11}$ to $S_{14}$ when using an omnidirectional antenna can hardly be differentiated from $S_{11}$ to $S_{14}$ using the interleaved antenna. Comparing Figure 21 and Figure 22 one can see that the differences exhibited in Figure 21 are negligible. Furthermore, Figure 22 highlights the impact of the squeeze in the radiation pattern and consequently the importance of our choice in compensating the mutual coupling effect. In view of these results, one can conclude that the measured correlation behaviour is only due to the spatial separation between the elements. Consequently our constraint in terms of omnidirectionality of the radiation pattern is fulfilled.

**Figure 21:** Cross-correlation of the 4 elements of the interleaved array antenna using Monte-Carlo simulations showing the influence of the new radiation pattern on the correlation coefficient

**Figure 22:** Cross-correlation of the 4 elements of a linear array with $0.4\lambda$ separation using Monte-Carlo simulations showing the influence of the squeeze in the radiation pattern on the correlation coefficient
6 MODEL VALIDATION

6.1 General description

Besides the characterisation of the environments, another aim of the measurements was to
derive realistic parameters of the MIMO radio channel and to feed them into the COSSAP®
implementation of the stochastic channel model described in chapter 3 in order to validate it.
The input parameters given to the COSSAP® MIMO model are the spatial power correlation
matrices $R_{MS}$ and $R_{BS}$ at the MS and BS respectively, in addition to the Doppler spectrum of
the signal. Typical values extracted from the 6 different environmental classes described in
section 5.2 will be presented in Appendix 1.

Notice that the correlation matrices are the result of an average over the reference antennas
with respect to which the measured spatial power correlation matrices are computed. On the
other hand, an averaged measured Doppler spectrum is defined by averaging it over all the
$MN$ paths. The averaged measured Doppler power spectra are normalised in frequency to
their maximum Doppler shift $f_D$ and in power to their maximum value. In order to check the
matching between the measured and simulated results, the eigenanalysis of their respective
channel correlation matrices have been performed.

This chapter is organised as follows. The eigenanalysis method is introduced in section 6.2.
The validation procedure based on the results of the eigenanalysis is presented in section 6.3.
The analysis procedure is then illustrated in section 6.4. Finally, the validation results
presented in Appendix 1 are discussed in section 0.

6.2 The eigenanalysis method

The eigenvalue decomposition (EVD) of the instantaneous correlation matrix $R$ defined as
$R = H H^H$, where $[.]^H$ represents Hermitian transposition, has been chosen to serve as a
benchmark of the validation process. This method enables to estimate the number of
independent channels between two terminals in a rich scattering environment [Andersen00].
A channel matrix $H$ may offer $k$ parallel subchannels with different mean gains, with $k$
defined as $k = \text{Rank}(R) \leq \min(M,N)$ where the function $\text{Rank}(.)$ and $\min(.)$ returns the rank
of the matrix argument and the minimum value of the arguments respectively. The $k^{th}$
eigenvalue is to be visualised as the power gain of the $k^{th}$ channel [Andersen00].

The results of the eigenanalysis of the measured and simulated narrowband information are
presented in Appendix 1. In the following, the eigenvalues are represented by $\lambda_k$, which is not
to be confused with the wavelength. The main objective of this chapter is not the
investigation of the propagation behaviour itself but mainly the investigation of the validity
of the proposed model. However, it is interesting to highlight that the correlation and the
decorrelation at both the BS and the MS is illustrated by the gain obtained in each eigenvalue
($\lambda_1$ to $\lambda_4$). A strong eigenvalue represents a situation where the single antenna elements at
both arrays are so uncorrelated that orthogonal subchannels exist between the ends of the
link. Vice-versa, if the correlation level is high, as it is the case when the eigenvalues are
weak, it is not possible to distinguish a large number of orthogonal subchannels.
6.3 Validation procedure

In order to validate the model, it is necessary to base our conclusions on sufficient statistics, which in the current case means a large number of eigenvalues sets identified for each measurement position. Therefore, a 4×4 scenario has been considered, since at most 4 eigenvalues can be expected.

The simulated data sets have been generated following a two-phase procedure:

- **Phase 1**
  
The measured complex Doppler spectra (i.e. with phase and magnitude information) of the MN paths \( h_{mn} \) are fed into the model. One should be aware that, by providing complex information through the Doppler spectra, the correlation information about the investigated position is simultaneously provided. Consequently, \( R_{MS} \) and \( R_{BS} \) are mapped to the identity matrix. Phase 1 is a first validation of the model with respect to the implementation of the fading.

- **Phase 2**
  
The true average spatial power correlation matrices \( R_{MS} \) and \( R_{BS} \) are applied to the model MN i.i.d. fading variables. Within the model, their amplitudes are shaped by the average Doppler power spectrum and their phase is uniformly distributed over \([0, 2\pi]\).

6.4 Analysis procedure

The analysis procedure applied to the simulated data is illustrated in Figure 23. For each position, 100 iterations have been performed, using different seeds of the random generator defining the phase of the Doppler spectrum. Each iteration counts as many samples as measurement samples collected during a slide run length. The EVD has been performed on the instantaneous correlation matrix \( R \) of each iteration sample in order to identify the corresponding eigenvalues. Averaging over the samples of one iteration, the averaged eigenvalues \( \tilde{\lambda}_k \) are derived, where \( k = 1...4 \) and \( l = 1...100 \). A second averaging operation has then been performed over the 100 iterations to derive the 2nd-order moment \( std(\tilde{\lambda}_k) \) related to the position under study.

In Appendix 1, cdfs of the eigenvalues are presented for each environmental class. Each figure exhibits three different kinds of curves:

(i) The cdfs of the measured eigenvalues \( \tilde{\lambda}_k \) for a full run of the Tx along the linear slide are displayed with a solid line.

(ii) On the same graph the cdfs (over 100 iterations) of the simulated eigenvalues \( \lambda_{kl} \) are illustrated by a dashed line. The eigenvalues are normalised to the mean gain value of a single Tx and single Rx element \( h_{mn} \) of the channel matrix \( H \).

(iii) In addition to these two sets of eigenvalues boundaries represented by two parallel dotted lines will be drawn, as also shown in Figure 23. These boundaries represent
left- and right-translated versions of the cdfs of simulated eigenvalues. The span of the translation is given by $\text{std}(\tilde{\lambda}_s)$.

Figure 23: Analysis procedure

Figure 24 is a typical example of the cdfs of the computed eigenvalues. Using as criterion the fact that the cdf of the measured eigenvalue $\tilde{\lambda}_s$ lies within the boundaries introduced here above, the inspection of the figures presented in Figure 24 as well as in Appendix 1 illustrates that the proposed stochastic model is validated for the correlated propagation mechanisms within a wide set of environments.

Figure 24: Eigenvalue results for measured and simulated data, environmental class Novi 2
6.5 Global validations

The purpose of this section is to propose a global analysis encompassing the 99 positions presented in Table 10 all together, in order to measure the matching rate of the stochastic model with simulations on basis of the eigenanalysis results. This analysis will rely on cdfs and 2nd order statistics of the averaged mean simulated eigenvalues $\bar{\lambda}_k$.

6.5.1 cdfs

As part of the cdfs validation, the error between these of the averaged mean simulated eigenvalues and the mean measured ones has been computed at some outage level for each position and for each eigenvalue. Figure 26 and Figure 27 illustrate these errors for an outage level of 10 % and 50 % respectively. A maximum error threshold is derived from Phase 1 simulation results as the maximum error measured at the same outage levels. The rationale for this choice is that Phase 1 simulation results should in principle be identical to measurements, for the correlation information is embedded in their Doppler spectrum. Any discrepancy between measurements and Phase 1 simulation results, as shown in Figure 25, can thus be attributed to simulation alea, that might plague Phase 2 simulations as well. Hence, the maximum error measured on Phase 1 results is applied as a tolerance for comparison of Phase 2 results with respect to measurements. For the measurement campaign of concern, it has been found that the Phase 1 error was upper-bounded by 2 dB. Hence, a tolerance of +/- 2dB is applied to the comparison of the Phase 2 simulation results to the measured ones. This tolerance threshold appears as horizontal lines in Figure 26 and Figure 27. Vertical lines are also drawn to differentiate the 6 environmental classes. Similar lines appear in Figure 27 and Figure 30.

![Figure 25: Error at 10% outage from the cdf of the eigenvalues between the measured and the simulated data, Phase 1. The discrepancy between measurements and Phase 1 simulation results lays the tolerance boundary of +/- 2dB](image.png)
Figure 26: Error at 10% outage from the cdf of the eigenvalues between the measured and the simulated data, Phase 2

Figure 27: Error at 50% outage from the cdf of the eigenvalues between the measured and the simulated data, Phase 2

Only 44 out of the 396 (4 eigenvalues × 99 positions) points shown in either Figure 26 or Figure 27 lie outside of the tolerance region. Moreover, the majority of outliers seems to be related to two close positions, which might indicate that some problems occurred during the collect of results.
One of the assumptions of the stochastic model is the WSS condition. However, it can be seen from the measured received power in Figure 28 that, in some cases this assumption is not respected. Indeed, the changes in the fluctuation of the signal from 0 up to 7\( \lambda \) compared to the rest of the run clearly denote that the antenna at the MS moved from a LOS to NLOS situation. In terms of spatial power correlation, the antenna elements are more correlated in the first part of the measurement run than in the other. Consequently when looking at the eigenvalue analysis of the measured samples in Figure 29, the abrupt change in the gain of the first (strongest) eigenvalue clearly indicates a change of correlation properties due to the influence of the fast variation in the received signal. However, when considering the simulated results this phenomenon is not reproduced. This is due to the fact that the stochastic model does not account for transitions between scenarios characterised by different correlation properties, as a result of its WSS assumption.

However, from the results presented in Appendix 1, based on the eigenanalysis of the instantaneous correlation matrix of the channel, one can conclude that the stochastic model appropriately renders the correlated propagation mechanisms.

![Figure 28: Measured received power envelope in the situation where the WSS assumption is not respected](image-url)
Figure 29: Eigenvalue results for simulated and measured data

6.5.2 2nd order validation

Figure 30 represents the $k = 1\ldots4$ standard deviations $\text{std}(\bar{\lambda}_k)$ of the mean eigenvalue $\bar{\lambda}_U, l = 1\ldots100$ for each position. The computation of the standard deviation is performed over the set of 100 iterations, while the mean eigenvalue is computed over a slide run length. The standard deviation of the first, strongest eigenvalue $\text{std}(\bar{\lambda})$ lie within 0 dB and 1 dB while the three others exhibit stronger fluctuations from 2 dB to 7 dB. It is quite natural that the weakest eigenvalues exhibit the widest fluctuations, since their nominal value is low.

Figure 30: Standard deviation $\text{std}(\bar{\lambda})$ of the mean eigenvalue per position
Figure 31 shows the cdf of the standard deviations of the mean eigenvalue $\overline{\lambda}_k$ presented in Figure 30. For the set of available statistics (100 slide run lengths for 99 positions), one can see that the standard deviation of the first eigenvalue is less than 1 dB in more than 95% in the cases. It is therefore fair to conclude at the end of this eigenanalysis that the stochastic channel model is validated.

Figure 31: cdf of the standard deviation $std(\overline{\lambda}_k)$ of the mean eigenvalues over all 99 positions
7 CONCLUSIONS

The present document has presented the main results achieved within Workpackage 2. Its purpose was to model the characteristics of the MIMO radio channels in a wide variety of environments. An innovative stochastic model which combines correlation information gathered at both ends of the communication link has been described in details and its COSSAP® implementation has been documented. One of the characteristics of this model is that it relies on a limited set of input parameters. A literature review has been presented in order to guide the choice of consistent parameters. On the other hand, similar sets of parameters have been extracted from measurement campaigns. During these campaigns, a total of 99 positions has been investigated in 6 different environments. Finally, eigenanalysis has been performed on both measured and simulated data. From the fair matching between the eigenanalysis results, one can conclude that the proposed stochastic model is validated.
8 REFERENCES


[WWW TSUNAMI II] http://www.era.co.uk/tsunami/tsunami2.htm

9 APPENDIX 1 - VALIDATION MODEL PARAMETERS AND RESULTS

In this annex, a typical set of input parameters and the corresponding eigenanalysis results are presented for each of the 6 environmental classes.

9.1 Environmental class: Novi 2

9.1.1 Input parameters

\[
R_n = \begin{bmatrix}
1.0000 & 0.5442 & 0.0968 & 0.1284 \\
0.5442 & 1.0000 & 0.2622 & 0.1631 \\
0.0968 & 0.2622 & 1.0000 & 0.2104 \\
0.1284 & 0.1631 & 0.2104 & 1.0000
\end{bmatrix}
\]

\[
R_m = \begin{bmatrix}
1.0000 & 0.3409 & 0.2893 & 0.0772 \\
0.3409 & 1.0000 & 0.2354 & 0.1814 \\
0.2893 & 0.2354 & 1.0000 & 0.2403 \\
0.0772 & 0.1814 & 0.2403 & 1.0000
\end{bmatrix}
\]

Figure 32: Average empirical Doppler power spectrum

9.1.2 Eigenanalysis results

Figure 33: Eigenvalue results for simulated and measured data
9.2 Environmental class: Novi 3

9.2.1 Input parameters

\[
R_{ai} = \begin{bmatrix}
1.0000 & 0.1538 & 0.1330 & 0.1613 \\
0.1538 & 1.0000 & 0.0906 & 0.0653 \\
0.1330 & 0.0906 & 1.0000 & 0.2082 \\
0.1613 & 0.0653 & 0.2082 & 1.0000
\end{bmatrix}
\]

\[
R_{ar} = \begin{bmatrix}
1.0000 & 0.2623 & 0.0617 & 0.1309 \\
0.2623 & 1.0000 & 0.1605 & 0.1583 \\
0.0617 & 0.1605 & 1.0000 & 0.0521 \\
0.1309 & 0.1583 & 0.0521 & 1.0000
\end{bmatrix}
\]

Figure 34: Average empirical Doppler power spectrum

9.2.2 Eigenanalysis results

Figure 35: Eigenvalue results for simulated and measured data
9.3 Environmental class: Nokia

9.3.1 Input parameters

\[
R_{\text{ai}} = \begin{bmatrix}
1.0000 & 0.4154 & 0.2057 & 0.1997 \\
0.4154 & 1.0000 & 0.3336 & 0.3453 \\
0.2057 & 0.3336 & 1.0000 & 0.5226 \\
0.1997 & 0.3453 & 0.5226 & 1.0000
\end{bmatrix}
\]

\[
R_{\text{si}} = \begin{bmatrix}
1.0000 & 0.3644 & 0.0685 & 0.3566 \\
0.3644 & 1.0000 & 0.3245 & 0.1848 \\
0.0685 & 0.3245 & 1.0000 & 0.3093 \\
0.3566 & 0.1848 & 0.3093 & 1.0000
\end{bmatrix}
\]

Figure 36: Average empirical Doppler power spectrum

9.3.2 Eigenanalysis results

Figure 37: Eigenvalue results for simulated and measured data
9.4 Environmental class: FrB7

9.4.1 Input parameters

\[
R_{ss} = \begin{bmatrix}
1.0000 & 0.2012 & 0.3162 & 0.1429 \\
0.2012 & 1.0000 & 0.4596 & 0.1458 \\
0.3162 & 0.4596 & 1.0000 & 0.2908 \\
0.1429 & 0.1458 & 0.2908 & 1.0000
\end{bmatrix}
\]

\[
R_{ns} = \begin{bmatrix}
1.0000 & 0.0989 & 0.0508 & 0.1089 \\
0.0989 & 1.0000 & 0.1020 & 0.1490 \\
0.0508 & 0.1020 & 1.0000 & 0.0871 \\
0.1089 & 0.1490 & 0.0871 & 1.0000
\end{bmatrix}
\]

Figure 38: Average empirical Doppler power spectrum

9.4.2 Eigenanalysis results

Figure 39: Eigenvalue results for simulated and measured data
9.5 Environmental class: FB7B2

9.5.1 Input parameters

\[
R_{ai} = \begin{bmatrix}
1.0000 & 0.3169 & 0.3863 & 0.0838 \\
0.3169 & 1.0000 & 0.7128 & 0.5626 \\
0.3863 & 0.7128 & 1.0000 & 0.5354 \\
0.0838 & 0.5626 & 0.5354 & 1.0000 \\
\end{bmatrix}
\]

\[
R_{ai} = \begin{bmatrix}
1.0000 & 0.1317 & 0.1992 & 0.2315 \\
0.1317 & 1.0000 & 0.1493 & 0.1907 \\
0.1992 & 0.1493 & 1.0000 & 0.1996 \\
0.2315 & 0.1907 & 0.1996 & 1.0000 \\
\end{bmatrix}
\]

Figure 40: Average empirical Doppler power spectrum

9.5.2 Eigenanalysis results

Figure 41: Eigenvalue results for simulated and measured data
9.6 Environmental class: Aalborg International Airport

9.6.1 Input parameters

\[
R_{ss} = \begin{bmatrix}
1.0000 & 0.1486 & 0.1450 & 0.1085 \\
0.1486 & 1.0000 & 0.3063 & 0.2162 \\
0.1450 & 0.3063 & 1.0000 & 0.2097 \\
0.1085 & 0.2162 & 0.2097 & 1.0000
\end{bmatrix}
\]

\[
R_{sr} = \begin{bmatrix}
1.0000 & 0.2682 & 0.1907 & 0.4118 \\
0.2682 & 1.0000 & 0.2701 & 0.1507 \\
0.1907 & 0.2701 & 1.0000 & 0.2822 \\
0.4118 & 0.1507 & 0.2822 & 1.0000
\end{bmatrix}
\]

![Figure 42: Average empirical Doppler power spectrum](image)

9.6.2 Eigenanalysis results

![Figure 43: Eigenvalue results for simulated and measured data](image)
10 APPENDIX 2 - DISTRIBUTION TERMS

The COSSAP® primitive model developed by AAU\(^5\) is free of use to any party having approved beforehand and on an individual basis the terms of the following agreement:

1. The receiving party agrees to acknowledge AAU’s parenthood on the COSSAP® block by referencing in every publication it may produce in the future based on the use of this block the AAU’s papers presented at IST Mobile Communications Summit 2000 or AAU’s newer related publications;

2. The receiving party agrees to acknowledge co-operation with IST project METRA in every publication it may produce in the future based on the use of this block;

3. The receiving party agrees not to distribute the source code to third parties;

4. In order to ensure that any enhancement might benefit to the whole community using the block, the receiving party agrees to notify AAU of any change and/or improvement of the source code, and to document it.

\(^5\) Contact person at AAU: Prof. Preben E. Mogensen <pm@cpk.auc.dk>, Niels Jernesvej 12, DK-9220 Aalborg East, Denmark.