SIGNAL PROCESSING PERSPECTIVES TO RADIO CHANNEL MODELLING

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Abstract— **In this paper, we show that a proper data-model for a measured radio channel consists of deterministic and stochastic components. The deterministic part maps channel parameters to the radio channel using an algebraic function, while the stochastic part is parameterized using second order statistics. We focus our discussion on the stochastic part describing the Dense Multipath Component (DMC) of the channel. We propose to use the von Mises-Fisher distribution to describe the power angular profile at the transmitter and the receiver in azimuth and elevation. This distribution is appropriate for angular data. Furthermore, the uniform angular distribution, used so far to model the DMC, is included as an extrem case. We identify the basic components necessary to implement this channel model for analysis and synthesis purposes. Finally, using channel sounding measurements, we demonstrate that uniform as well as concentrated distributions occur in real world radio propagation scenarios.**

I. Introduction

Future wireless communication systems will exploit the rich spatial and temporal diversity of the radio propagation environment. Already today some WLAN products (e.g. Draft 802.11n) utilizing the MIMO technology are available. At the same time, the usage and the density of different wireless networks such as WLAN and WiMAX has increased. This calls for new advanced channel models, which need to be verified by real-world channel sounding measurements. In this context the reliable estimation and tracking of the model parameters from measurement data is of particular interest.

In this paper, we discuss the modeling of radio propagation channels for wireless Multiple-Input-Multiple-Output (MIMO) communication systems based on channel sounding measurements. We apply an extended model for the spatio-temporally distributed measurement covariance. Such an extension for the Dense Multipath Component has been discussed in [1]. In that work the authors derived a maximum likelihood estimator for a covariance model having von Mises distribution in azimuth at one link end, as well as an exponential delay profile. The analysis in [1] was also limited to simulations with linear antenna arrays.

We extend the angular model used in [1] to both azimuth and elevation domains, as well as to cases where antenna arrays may have arbitrary geometry. Our covariance model comprises of additive white Gaussian measurement noise plus the DMC component [2], which takes the spatial correlation at Tx and Rx as well as the exponential delay profile into account. In our second contribution, we show that uniform as well as concentrated power distributions are observed for DMC in the angular domain in MIMO radio channel measurements.

This paper is structured as follows. In Section II, we derive our radio channel model. In Section III, we outline a state of the art channel parameter estimation algorithm used in this work. Section IV describes the channel sounding measurements used in Section V, where we present MIMO channel measurement results for DMC in a Micro-Cell scenario.

II. Measurement-Based Radio Channel Modeling

Suppose one is given a sequence of channel observations with a description of the measurement system (channel sounder), such as used antenna arrays at the transmitter and the receiver (calibration data), measurement bandwidth and -interval, excitation signal(s), et cetera. Now the task is to estimate parameters of a parametric channel-model, i.e. solving the inverse problem.

Commonly used deterministic models employs a superposition of a finite number of plane waves modeling the propagation. However, it has been show that a deterministic model alone is not sufficient for describing the observation completely [2], [3]. The complexity of the radio channel becomes an issue. It turnes out that a channel observation contains a superposition of a very large number of propagation paths (interactions of the propagating wave with one or multiple scattering points). Due to the limited aperture of a channel sounder there is not enough information in a channel observation to separate/resolve all observed propagation paths.

The data model for a channel observation was extended to include a stochastic component, which describes the contribution of a large amount of (individually weak) spatial distributed scatterers, so called dense multipath [2], [3]. Consequently, the model for a channel observation contains both, a deterministic mapping of parameters (describing strong propagation paths) and a stochastic part (mapping) being determined by second order statistics, i.e. a parametric covariance matrix.

We use only second order moments to describe the stochastic part, but it should be noted that the actual phenomena may have arbitrary order. Since our data model for a channel observation contains a deterministic as well as a stochastic part, it would be classified as a stochastic model.

As mentioned above, the design of this channel-model is solely motivated by the needs of channel parameter estimation.

However, considering that the model is tailored for channel measurements (system analysis) having usually a large observation aperture (bandwidth, aperture of the antenna arrays, et cetera) this model may also be suited as a channel generation model (channel synthesis) for e.g. Link-level simulations. The capabilities (apertures) of the communication system of interest should not exceed the resolution capabilities of the employed measurement system.

The radio channel can be described by a channel matrix. In the frequency domain the channel matrix has a simple structure, provided the channel can be treated as time-invariant. The broadband channel matrix $\boldsymbol{\mathcal{H}}$ is block diagonal

$$\boldsymbol{\mathcal{H}} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \ddots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{H}_{M_f} \end{bmatrix} \in \mathbb{C}^{M_f M_R \times M_f M_T}, \qquad (1)$$

where $\mathbf{H}_{m_f} \in \mathbb{C}^{M_R \times M_T}$ denotes the channel matrix for the frequency m_f, M_f is the number of frequency samples, and M_T and M_R are the number of transmit and receive antennas, respectively. From a parameter estimation point of view it may be convenient to express the channel with a vector $\mathbf{h} \in \mathbb{C}^{M_f M_R M_T \times 1}$, which is related to $\boldsymbol{\mathcal{H}}$ according to

$$\mathbf{h} = \mathrm{vec}\left\{ \left[\begin{array}{ccc} \mathrm{vec}\left\{\mathbf{H}_1\right\} & \cdots & \mathrm{vec}\left\{\mathbf{H}_{\mathrm{M_f}}\right\} \end{array} \right]^{\mathrm{T}} \right\}. \qquad (2)$$

Using (1) the input-output relation of the broadband MIMO radio channel in the frequency domain can be expressed by

$$\mathbf{x_r} = \boldsymbol{\mathcal{H}} \mathbf{x_t},$$

where $\mathbf{x}_t$ and $\mathbf{x}_r$ are the transmitted and received symbols, respectively.

According to the discussion above, we split the radio channel into two components $\mathbf{h}_s$, the contribution of the dominant (concentrated) propagation paths having parameters $\mathbf{\Theta}_s$, and $\mathbf{h}_d$ the dense multipath component having parameters $\mathbf{\Theta}_d$. The vector $\mathbf{h} = \mathbf{h}_s + \mathbf{h}_d$ can be understood as a realization of the process $\mathbf{h} \sim \mathcal{CN}\left(\mathbf{h}_s(\mathbf{\Theta}_s), \mathbf{R}_d(\mathbf{\Theta}_d)\right)$. That means, we separate the channel into its first order statistics, the parametric mean $\mathbf{h}_s(\mathbf{\Theta}_s)$ and its second order statistics, describing the dense multipath components with the covariance matrix $\mathbf{R}_d(\mathbf{\Theta}_d)$.

A. Concentrated (Specular) Propagation Paths

Each concentrated or dominant propagation path has a significant contribution to the received signal, and their parameters can be identified using the information available in a sequence of channel observations. That means, they can be distinguished from the distributed diffuse scattering since their magnitude is large, i.e. the likelihood that they belong to the process $\mathbf{h}_d \sim \mathcal{CN}\left(\mathbf{0}, \mathbf{R}_d(\mathbf{\Theta}_d)\right)$ is very small.

The concentrated propagation paths are parameterized by a time-delay, a transmit angle (azimuth and elevation), a receive angle (azimuth and elevation), and the polarimetric path weights. Let P_s be the number of dominant propagation paths in an observation, than $\mathbf{h}_s$ can be expressed by the superposition of the contribution of the individual paths according to

$$\mathbf{h}_s\left(\mathbf{\Theta}_s\right) = \sum_{p=1}^{P_s} \mathbf{s}\left(\boldsymbol{\theta}_p\right),$$

where $\mathbf{s}\left(\boldsymbol{\theta}_p\right)$ is a function mapping the propagation path parameters $\boldsymbol{\theta}_p$ using measurement system information to the observation data domain. Alltogether the parameters $\boldsymbol{\theta}_p$ form the parameter set $\mathbf{\Theta}_s = [\boldsymbol{\theta}_1, \cdots, \boldsymbol{\theta}_{P_s}]$. For a discussion of the parameterization of the concentrated propagation paths and the mapping of the parameters $\boldsymbol{\theta}_p$ to $\mathbf{h}_s$ see [2].

B. Dense Multipath Components

A discussion of the model for the power delay profile (PDP) of the DMC can be found in [2], [3]. The model is based on the observation that the PDP has an exponential decay in the delay domain and a base delay, which is related to the distance between the transmitter and receiver. The power delay profile of the dense multipath components for infinite bandwidth has been proposed in [4]–[6]

$$\psi(\tau) = E[|x(\tau)|^2] = \begin{cases} 0, & \tau < \tau'_d \\ \alpha_1/2, & \tau = \tau'_d \\ \alpha_1 e^{-B_d(\tau - \tau'_d)}, & \tau > \tau'_d \end{cases}, \qquad (3)$$

where B_d is the coherence bandwidth, α_1 denotes the maximum power, and τ' is the base delay. The related power spectrum density is given by the Fourier transform of (3) as

$$\psi(\Delta f) = \frac{\alpha_1}{\beta + j2\pi\Delta f} e^{-j2\pi\Delta f \tau'}, \qquad (4)$$

where $\beta = B_d/(M_f f_0)$ is the normalized coherence bandwidth, and f_0 is the sampling interval in the frequency domain. Let $\boldsymbol{\kappa}(\boldsymbol{\theta}_{f,d})$, $\boldsymbol{\theta}_{f,d} = [\alpha_1, \beta, \tau]^{\mathrm{T}}$ denote a sampled version of the correlation function (4). In frequency-domain it may be written as

$$\boldsymbol{\kappa}(\boldsymbol{\theta}_{f,d}) = \frac{\alpha_1}{M_f} \left[\frac{1}{\beta} \quad \cdots \quad \frac{e^{-j2\pi(M_f - 1)\tau}}{\beta + j2\pi\frac{M_f - 1}{M_f}} \right], \qquad (5)$$

where $\tau = \tau' f_0, \tau \in [0, 1)$ is the normalized base delay.

Since the process is assumed to be wide sense stationary in frequency domain, the correlation between components at different frequencies is given by the frequency distance Δf

$$\Psi_f(f_1, f_2) = \psi(f_1 - f_2) = \psi(\Delta f). \qquad (6)$$

The covariance matrix of the diffuse scattering is then a Toeplitz matrix

$$\mathbf{R}_f\left(\boldsymbol{\theta}_{f,d}\right) = \mathrm{toep}\big(\boldsymbol{\kappa}(\boldsymbol{\theta}_{\mathrm{f,d}}), \boldsymbol{\kappa}^{\mathrm{H}}(\boldsymbol{\theta}_{\mathrm{f,d}})\big). \qquad (7)$$

The covariance matrix $\mathbf{R}_f\left(\boldsymbol{\theta}_{f,d}\right) \in \mathbb{C}^{M_f \times M_f}$ describes the distribution of the DMC in the frequency domain. Consequently, the parameter vector $\boldsymbol{\theta}_{f,d}$ describes the frequency correlation of the DMC.

In general, we have to represent the second order statistics of the dense multipath components with a 10-dimensional correlation function between any two points in the parameter domain, where one point i is characterized by frequency, transmit-azimuth and -elevation, and receive-azimuth and

-elevation $(f_i, \varphi_{T,i}, \vartheta_{T,i}, \varphi_{R,i}, \vartheta_{R,i})$. Provided the WSSUS (wide sense stationary uncorrelated scattering) assumption [7] holds in the spatial domains as well, the correlation function can be expressed by distances in the respective domains. This leads to a 5-dimensional correlation function

$$\psi\left(\Delta f, \Delta\varphi_T, \Delta\vartheta_T, \Delta\varphi_R, \Delta\vartheta_R\right).$$

The full covariance matrix $\mathbf{R}_d\left(\mathbf{\Theta}_\mathrm{d}\right) = \mathrm{E}\left\{\mathbf{h}_\mathrm{d}\mathbf{h}_\mathrm{d}^\mathrm{H}\right\}$ of the dense multipath components is of size $M_R M_T M_f \times M_R M_T M_f$.

Assuming the stochastic process is separable, we approximate the covariance matrix $\mathbf{R}_d\left(\mathbf{\Theta}_\mathrm{d}\right)$ by a Kronecker-product. Based on this assumption the covariance matrix of the DMC $\mathbf{h}_d$ has the structure

$$\mathbf{R}_d\left(\mathbf{\Theta}_d\right) = \mathbf{R}_R\left(\boldsymbol{\theta}_{R,d}\right) \otimes \mathbf{R}_T\left(\boldsymbol{\theta}_{T,d}\right) \otimes \mathbf{R}_f\left(\boldsymbol{\theta}_{f,d}\right), \quad (8)$$

where $\mathbf{R}_R\left(\boldsymbol{\theta}_d\right) \in \mathbb{C}^{M_R \times M_R}$ and $\mathbf{R}_T\left(\boldsymbol{\theta}_d\right) \in \mathbb{C}^{M_T \times M_T}$ describe the angular distribution / spatial correlation of the dense multipath components at the receiver and the transmitter, respectively. In [2], it was assumed that the DMC are angular white, yielding the full covariance matrix

$$\mathbf{R}_d\left(\theta_d\right) = \mathbf{I}_{M_R} \otimes \mathbf{I}_{M_T} \otimes \mathbf{R}_f\left(\boldsymbol{\theta}_{f,d}\right). \quad (9)$$

Due to our best knowledge, so far no measurement results have been reported, supporting or contradicting this assumption.

However, in [1] it has been shown using simulations that an concentrated angular distribution of the DMC has a significant impact on the estimates of the dominant propagation paths and therefore has to be taken into account. To this end a suitable parametric model for the covariance matrices $\mathbf{R}_T\left(\boldsymbol{\theta}_{T,d}\right)$ and $\mathbf{R}_R\left(\boldsymbol{\theta}_{R,d}\right)$ has to be derived. Assuming the same general behaviour of the DMC at the transmitter as well as at the receiver site, we will derive a general parametric covariance matrix $\mathbf{R}_a$ for directional data.

In [1] the von Mises distribution has been used to model the azimuthal power distribution of the DMC. The von Mises distribution has three properties, which makes it suitable for the modelling problem at hand.

1) It has been designed to describe directional data on full 360°.
2) The uniform distribution is a special case of this distribution.
3) The distribution is parameterized by a location $\boldsymbol{\mu}_a$ and a spread parameter κ_a, very similar to a normal distribution.

However, the von Mises distribution can only be used to describe directional data on a circle. Therefore, we propose to extend this concept and use the von Mises-Fisher distribution [8]. It is a probability distribution on the $(l-1)$-dimensional sphere in $\mathbb{R}^l$. For $l = 2$ it reduces to the von Mises distribution on the circle. The probability density function (pdf) of the von Mises-Fisher distribution for the random, l-dimensional, unit vector $\mathcal{Z}$ is given by

$$p\left(\mathbf{z}; \boldsymbol{\mu}, \kappa\right) = C_l\left(\kappa_a\right) \mathrm{e}^{\kappa_\mathrm{a} \boldsymbol{\mu}_\mathrm{a}^\mathrm{T} \mathbf{z}}, \quad (10)$$

where $\kappa_a \geq 0$, $|\boldsymbol{\mu}_a| = 1$, and $C_l\left(\kappa_a\right)$ is a normalization constant [8]. For a spread of $\kappa_a = 0$ the pdf (10) reduces to a uniform distribution on $\mathbb{S}^{l-1}$.

Let $\mathbf{K}_{\varphi_s,\vartheta_s}$ contain samples drawn from the probability distribution function (10) on the unit sphere, and $\mathbf{G}_a$ be the EADF [9] of an arbitrary antenna array used to observed the radio channel at Tx or Rx. Then

$$\mathbf{R}_a = \mathbf{G}_a \mathbf{A}_{\varphi_s,\vartheta_s} \mathbf{K}_{\varphi_s,\vartheta_s} \mathbf{A}_{\varphi_s,\vartheta_s}^\mathrm{H} \mathbf{G}_a^\mathrm{H} \in \mathbb{C}^{M_a \times M_a} \quad (11)$$

is a model for the covariance matrix of the observed DMC in the antenna element domain.

To illustrate the difference between, the so far dominating approach to model DMC, (9) and the proposed model (8) using (11) (Approach #2), realizations of the two stochastic processes has been generated and compared with a channel sounding measurement. In Figure 1 the three power profiles in the transmit-azimuth and delay domain are shown. Clearly, the reconstructed spatially correlated process resembles the measured process much better. At the same time, all three processes share the same PDP as shown in Figure 2.

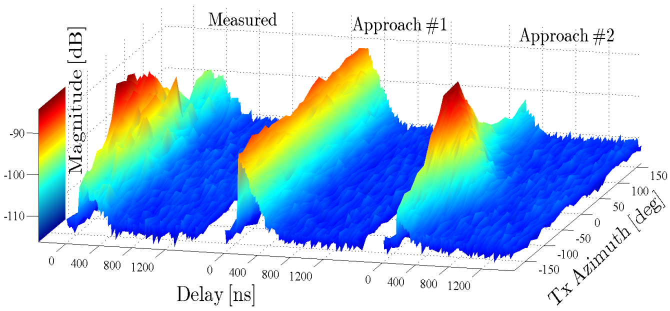

Fig. 1. Illustration of different approaches to model DMC in the angular domain. Clearly Approach #2 matches the measured power-profile.

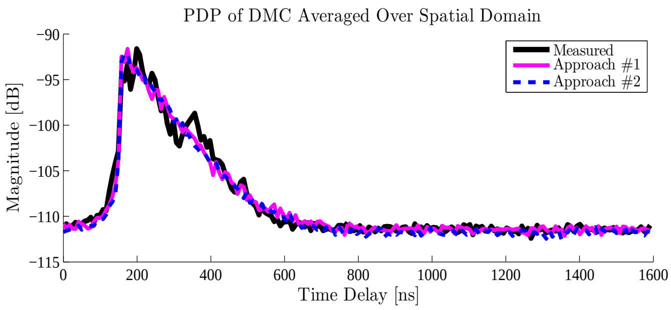

Fig. 2. PDP averaged over all Rx and Tx antennas of DMC in the delay domain, both approaches match to the measured power-profile.

III. Channel Measurements and Parameter Estimation

An observation $\mathbf{x}$ of the radio channel $\mathbf{h}$ acquired with a channelsounder [10] contains also additive measurement noise $\mathbf{w} \in \mathbb{C}^{M_f N_R N_T \times 1}$. The measurement noise can be modelled as a realization of the i.i.d. circular complex Normal distributed process $\mathcal{CN}\left(\mathbf{0}, \sigma_w^2 \mathbf{I}\right)$, where σ_w^2 is the noise variance (power). Since both contributions $\mathbf{w}$ and $\mathbf{h}_d$ are realizations of a circular complex Normal process, a channel observation is distributed according to

$$\mathbf{x} \sim \mathcal{CN}\left(\mathbf{h}_s(\boldsymbol{\Theta}_s), \mathbf{R}_d(\boldsymbol{\Theta}_d) + \sigma_w^2 \mathbf{I}\right).$$

A maximum likelihood estimator (MLE) for the parameters $\boldsymbol{\Theta}_s$, $\boldsymbol{\theta}_{f,d}$, and σ_w^2 RIMAX has been proposed in [2], [11]. It exploits the fact that the parameters of the two components of the channel model are asymptotically independent. Therefore, one can decouple the estimation problem into two estimation problems [2]. The resulting algorithm is iterative and alternates between the maximization of the likelihood function with respect to the parameters $\boldsymbol{\Theta}_s$ and $\boldsymbol{\theta}_{f,d}$.

If a sequence of observations is available, RIMAX does not exploit the fact that the channel-parameters are correlated in time to reduce the variance of their estimates. Their correlation is only exploited to reduce the computational complexity. In [12] it has been proposed to use a state-space model to describe the evolution of channel parameters in time. This state-space model has been applied to estimate $\boldsymbol{\Theta}_s$ using an extended Kalman Filter (EKF) in [12]. The state-space model, and in turn the estimator have been refined in [13]. The state-space based approach yields estimates with lower variance compared with the RIMAX estimates, if a sequence of channel observations is available, what is usually the case. Furthermore, the EKF based estimator provides a significant reduction in computational complexity compared with RIMAX. In [14] a state-space model for the parameters $\boldsymbol{\theta}_{f,d}$ has been proposed, and it has been shown that this parameters can be estimated by an EKF based estimator, as well. Again, the proposed estimator provides estimates with lower variance than RIMAX and reduces the computational complexity further. The EKF based estimator for $\boldsymbol{\Theta}_s$ and $\boldsymbol{\Theta}_d$ is one among the best estimators available nowadays. Here best means in terms of estimation variance and computational complexity.

IV. Measurement Setup

The channel sounding data used in this work has been measured in the city center of Ilmenau, Germany [15]. The channel sounder used was a RUSK ATM (MIMO) [10]. The measurement setup applied in the measurement campaign is outlined in Table I. A map of the measurement scenario is shown in Figure 3. The map shows the start- and end-points of the individual measurement routes, and the orientation of the fixed access-point (AP) antenna array. Out of the eight available measurements the data taken along route #1 (point (3) to point (16)) and along route #2 (from (16) going to (26)) have been used in this work. The trolley with the MT (mobile terminal) antenna array has been driven on the right side of the street shown in Figure 4. The trolley with the transmit-antenna-array and the transmitter are in the foreground of Figure 5. In the background of the same picture the position of the linear receive-array is visible (middle-left of the picture).

Every $20.48\mathrm{ms}$ a channel observation has been measured. The radio-channel along route #1 has been measured for $\approx 51\mathrm{s}$ and for $\approx 59\mathrm{s}$ along route #2, yielding $K_1 = 2472$ and $K_2 = 2865$ channel observations, respectively.

TABLE I

Measurement Setup

Channel sounder:	RUSK ATM (MIMO)
Carrier frequency:	5.2 GHz
Measurement bandwidth:	120 MHz
Maximum multipath delay:	$1.6\mu s$
Transmit power at the antenna:	approx. $200\mathrm{mW}$
AP antenna:	An 8-element uniform lin. patch array (PULA8) with $\approx$ 0.49λ element spacing, polarimetric $\approx$ 120 deg directivity, about $4\mathrm{m}$ above ground
MT antenna:	A 16-element uniform circular array with radius $\approx 0.38\lambda$, vertically polarized, on the top of a trolley, and about $1.3\mathrm{m}$ above ground

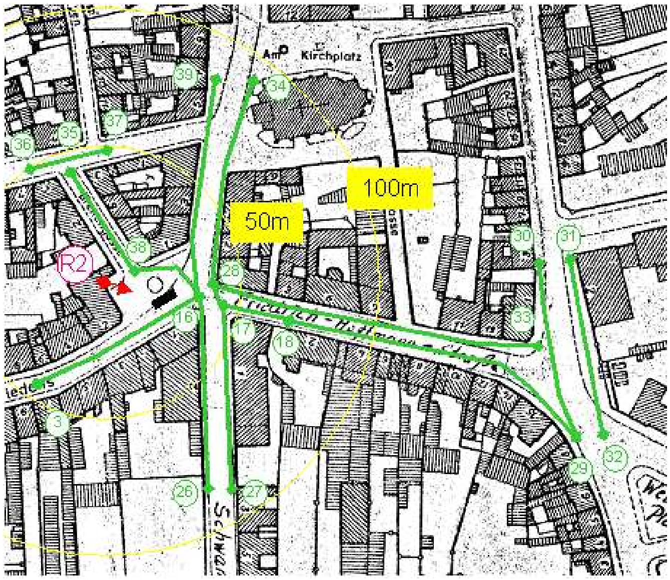

Fig. 3. Map of measurement scenario. The measurements have been carried out in August 2004, downtown Ilmenau, Germany.

V. Measurement Results

The EKF based parameter estimator mentioned in Section III has been applied to the channel sounding observations $\mathbf{x}_k$, yielding channel parameter estimates for the concentrated propagation paths $\hat{\boldsymbol{\Theta}}_{s,k}$. Using the estimates, the contribution of the dominant paths have been reconstructed yielding an

Fig. 4. The measured street scenario, a view from point 17 to point 27 on the map shown in Figure 3.

Fig. 5. Location of the receive antenna array (middle left in the photo). A polarimetric eight element linear array (16 ports) has been used (Location: $50°41'06.60"\mathrm{N}, 10°54'51.64"\mathrm{E}$). The receive array had the role of an access point antenna in the measurement scenario. In the foreground is the trolley with the transmitter and the transmit-array. A 16-element circular array has been used, acting as the mobile station antenna.

estimate $\hat{\mathbf{h}}_{s,k}$ of $\mathbf{h}_{s,k}$. Using $\hat{\mathbf{h}}_{s,k}$ an estimate of the DMC and noise at time k can be computed according to

$$\hat{\mathbf{x}}_{d,k} = \mathbf{x}_k - \hat{\mathbf{h}}_{s,k}. \tag{12}$$

Using a Hann-window an estimate of the PDP of the DMC has been computed from (12). Then, using a beamformer, the angular power distribution in azimuth at the transmitter has been estimated for all 16 receive antennas. The Power-Delay-Transmit-Azimuth profiles (PDTAP) have been averaged yielding an average PDTAP $\hat{\mathbf{R}}_{dta,k}$ at k. Since the parameters of the DMC are slow time variant the PDTAPs have been averaged in time using

$$\hat{\mathbf{P}}_{dta,k} = (1-\psi)\hat{\mathbf{R}}_{dta,k} + \psi\hat{\mathbf{P}}_{dta,k-1},$$

with $\psi = 0.01$. The average PDTAP at 10 positions in both routes are shown in Figure 6 and Figure 7 (darker colors indicate higher powers). From the results the following observations can be made.

1) The distribution of the DMC in azimuth at the transmitter is almost uniform at the beginning of the measurements, i.e. in line-of-sight (LOS) and obstructed line-of-sight (OLOS), i.e. the assumption of a white process in the azimuth angle domain is valid.
2) The power distribution of the DMC in azimuth at the transmitter is changing slowly to a concentrated distribution when the transmitter enters the non-LOS (NLOS) area of the measurement routes, i.e. the assumption of a white process in the azimuth angle domain is wrong.
3) If the power angular profile is concentrated in azimuth, most of the power is propagating along the street canyon in the analyzed scenario.

This result shows clearly that the distribution of the DMC can be uniform as well as concentrated in the angular domain depending on the propagation scenario. Consequently, it is not sufficient to model their distribution using (9), instead the improved model (8) using (11) should be used. Futhermore, this analysis explains the reduction of MIMO channel capacity in the NLOS area of route #1, which has been observed in [16].

VI. Conclusions

Using channel sounding measurements and high resolution channel parameter estimation, we have shown in this paper that the power distribution of the dense multipath component of mobile radio channels has to be modeled using a distribution that can vary between an uniform distribution on the intervall $\langle -\pi, +\pi \rangle$ and an angular concentrated distribution with a small spread. We propose to use a distribution which is especially tailored for the angular domain, e.g. the von Mises-Fisher Distribution [8]. Furthermore, we have shown in this paper that the correlation of DMC can be well approximated using the Kronecker-structured covariance matrix (8), in combination with expression (11), which in turn is based on the distribution function (10).

Acknowledgments

This work was partialy funded by the WILATI project, a joint project between three scandinavian universities. WILATI is part of the NORDITE research programme, funded by Finnish, Swedish and Norwegian national research institutes Tekes, Vinnova, and RCN, respectively

Furthermore, the authors would like to thank Institute of Communications and Measurement Engineering of Ilmenau University of Technology and Medav GmbH Germany for providing the RUSK channel sounder measurement data, as well as for the fruitfull cooperation over the last years.

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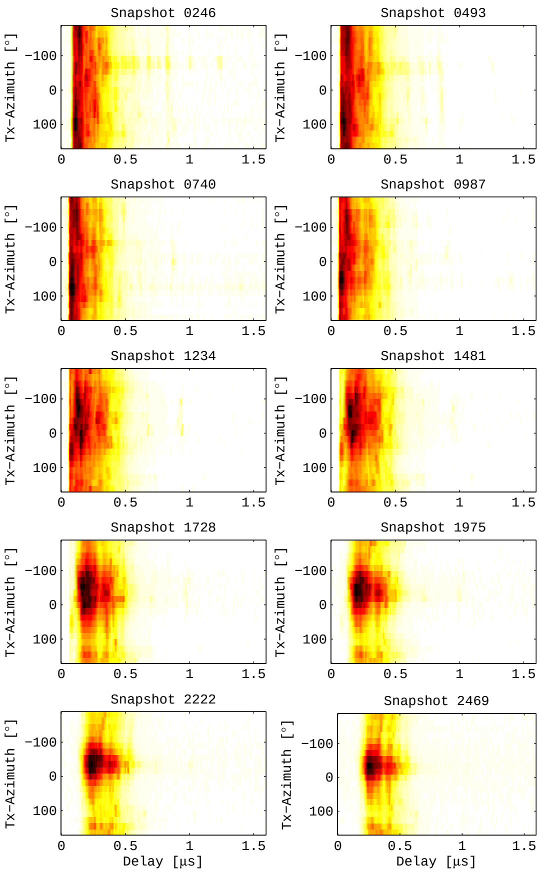

Fig. 6. Power distribution of dense multipath component in delay and transmit-azimuth. The 10 samples are taken from route #1, starting at point 16 and ending at point 3, see Figure 3. At the beginning of the route (line of sight/ obstructed line of sight) the power distribution is almost uniform in azimuth at the Tx (darker colors indicate higher powers). The distribution changes to a concentrated distribution in azimuth when the transmitter enters the entrance of the street (non line of sight scenario).

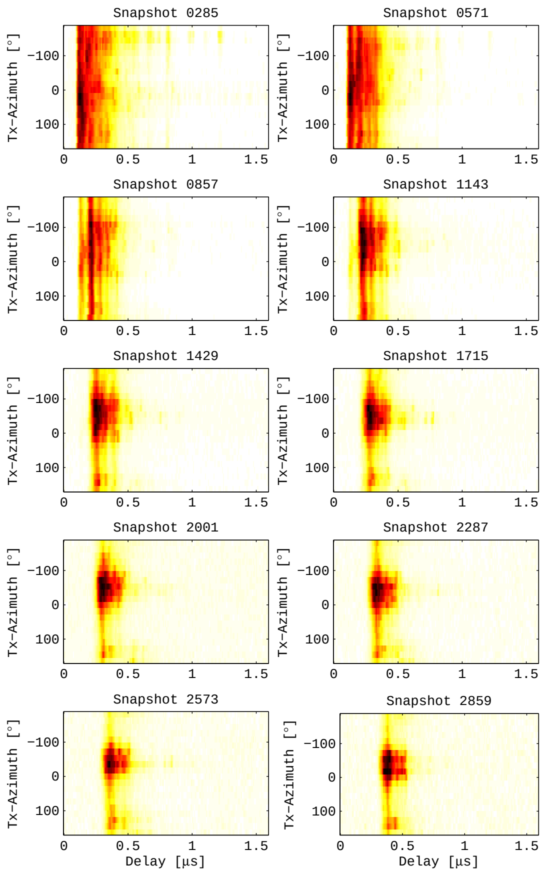

Fig. 7. Power distribution of dense multipath component in delay and transmit-azimuth. The 10 samples are taken from route #3, starting at point 16 and ending at point 26, see Figure 3. At the beginning of the route (line of sight/ obstructed line of sight) the power distribution is almost uniform in azimuth at the Tx (darker colors indicate higher powers). The distribution changes to a concentrated distribution in azimuth when the transmitter enters the entrance of the street (non line of sight scenario).

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