DISTRIBUTED SCATTERING IN RADIO CHANNELS AND ITS CONTRIBUTION TO MIMO CHANNEL CAPACITY

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ABSTRACT

A well accepted radio channel model approximates the radio channel impulse response by a superposition of a finite number of propagation paths. Wideband radio channel measurements can be approximated using concentrated propagation paths with an accuracy of close to 100% down to only 30% or less, depending on the scenario. For this reason it has been proposed to extend the data model by an additional component describing the distributed scattering (dense multipath component, DMC) of the radio channel. The model for the DMC is parameterized by three parameters, a base delay, the coherence bandwidth or delay-spread, and the attenuation factor of the dense multipath component. It is shown in this paper that the distributed scattering contributes significantly to the capacity of the MIMO-wideband radio channel. In fact it can be oberved in channel sounding measurements that the distributed scattering can contribute more to the channel capacity than the concentrated propagation paths.

Key words: radio channel modeling, MIMO, channel measurements, channel sounding, channel capacity.

1. INTRODUCTION

The interest in the multidimensional structure of the mobile radio channel is still growing. This is mainly due to the fact that future beyond 3G wireless systems will employ multiantenna transceivers in order to improve spectral efficiency and radio link quality. Consequently, realistic channel models that are verified by real-world measurement campaigns are needed especially for transceiver design and network planning purposes. Channel sounding and related propagation parameter estimation are key tasks in creating such channel models. In particular, the double-directional modeling of the radio channel (Steinbauer et al., 2001) has attracted a lot of interest because it gives a better physical insight into the wave propagation mechanism in real radio environments and it has, to some extend, the ability to remove the measurement antenna influence from the channel observation. Moreover, studying and comparing the performance of various MIMO (multiple-input-multiple-output) transceiver structures requires such advanced channel models as well.

To summarize, a model for the wireless radio propagation channel is required for the synthesis of radio channels. And it is needed as well for the analysis of radio channels in the field of radio channel measurements. The requirements on the model are different for the synthesis and the analysis task. One of the differences is the complexity of the model in terms of its parameters. In principle, for channel synthesis the channel model can have arbitrary many parameters. In channel analysis this is not true. Here the number of independent free parameters of the model can not be chosen arbitrary, the model complexity is rather determined by the available amount of information in the channel observations used to determine (estimate) the parameters. A well accepted (synthesis) radio channel model approximates the radio channel impulse response by a superposition of a finite number of propagation paths. This approach is valid for generating a realization of the radio channel, as long as the observed apertures in time, frequency and space are small. The necessary information for this channel model is a statistical model of the path parameters. The modelling accuracy can be controlled by the number of propagation paths used to generate the channel. These parameters must be derived from measurements of the radio channel. The parameter estimates are used to derive sufficient statistics for the radio channel propagation path parameters.

For the parameter estimation algorithm, we need a channel model to describe the observations. It has to be a model, whose parameters can be estimated from the measurement data. Due to the observations uncertainty, usually determined by the measurement noise, calibration and modelling errors, the parameter estimation resolution and accuracy is limited. For a given model, the minimum achievable parameter variance can be determined using the Cramér-Rao-Lower bound (Scharf, 1990). The

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chosen model is said to be too complex for the available amount of information if it turns out that the lower bound on the variance of one or several model parameters renders some of the parameter estimates meaningless. For the problem at hand this means, in contrast to the synthesis task one cannot increase the number of (concentrated) propagation paths beyond a certain limit in order to enhance the accuracy of the radio channel model. It should be noted that several researchers have discovered this fact while analysing channel sounding measurements. Typically, wideband radio channel measurements can be approximated using concentrated propagation paths with an accuracy of close to 100% down to only 30% or less, depending on the scenario, see e.g. (Richter, 2004). For this reason it has been proposed to extend the data model by an additional component describing the distributed scattering (dense multipath component) of the radio channel (Richter, 2004, 2005). The dense multipath components describe the contribution of the vast number of weak propagation paths, which cannot be estimated individually. The model is parameterized by three parameters, a base delay, the coherence bandwidth or delay-spread, and the attenuation factor of the dense multipath component. In this paper it is shown that the distributed scattering contributes significantly to the capacity of the MIMO-wideband radio channel. In fact it can be oberved in channel sounding measurements that the distributed scattering can contribute more to the channel capacity than the concentrated propagation paths.

The paper is structured as follows. In Section 2, the channel model is outlined. In Section 3, channel estimation algorithm used to separate the contribution of the DMC and the specular propagation paths to the radio channel are discussed. In Section 4, the estimation of the channel capacity is described. In the same section the influence of the measurement noise to the estimated channel capacity is discussed. In Section 5, the MIMO channel capacity estimated from a channel sounding measurement in a micro-cell scenario is presented. And finally, Section 6 concludes the paper.

2. RADIO CHANNEL MODEL

The radio channel is usually 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

$$\boldsymbol{\mathcal{H}} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \ddots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{H}_{N_f} \end{bmatrix} \in \mathbb{C}^{M_f N_R \times M_f N_T}, \quad (1)$$

where $\mathbf{H}_k \in \mathbb{C}^{N_R \times N_T}$ denotes the channel matrix for the frequency k. From a parameter estimation point of view it may be convenient to express the channel with a vector $\mathbf{h} \in \mathbb{C}^{M_f N_R N_T \times 1}$, which is related to $\boldsymbol{\mathcal{H}}$ according to

$$\mathbf{h} = \text{vec}\left\{ [\ \text{vec}\{\mathbf{H}_1\} \quad \cdots \quad \text{vec}\{\mathbf{H}_{\mathrm{M_f}}\}\ ]^{\mathrm{T}} \right\}. \quad (2)$$

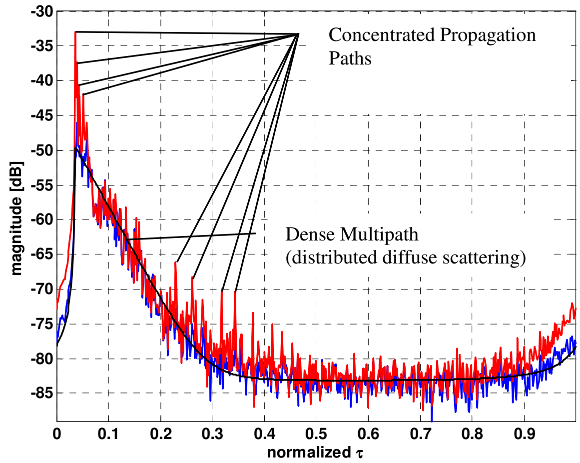

Figure 1. Components of the data model used to describe a radio channel observation. The red graph is a power delay profile of a measured channel impulse response (IR). The blue line represents the power delay profile of the same IR after removing the contribution of specular propagation paths. The black line shows a estimated PDP of the dense multipath components.

The vector $\mathbf{h} = \mathbf{h}_s + \mathbf{h}_d$ can be understood as a realization of the process $\mathbf{h} \sim \mathcal{N}_c(\mathbf{h}_s(\boldsymbol{\theta}_{sp}), \mathbf{R}_d(\boldsymbol{\theta}_{dmc}))$. That means, we separate the channel into its first order statistics, the parametric mean $\mathbf{h}_s(\boldsymbol{\theta}_{sp})$ and its second order statistics, describing the dense multipath components with the covariance matrix $\mathbf{R}_d(\boldsymbol{\theta}_{dmc})$.

2.1. Concentrated Propagation Paths

The concentrated or specular propagation paths are propagation paths which contribute individually significant to the received power. That means they can be distinguished from the distributed diffuse scattering, see also Figure 1. I.e. the likelihood that they belong to the process $\mathbf{h}_d$ is very small, since their magnitude is large. The concentrated propagation paths are parameterized by a time-delay, a transmit angle (azimuth and elevation), a receive angle (azimuth and elevation), and a polarimetric path weight matrix. For a discussion of the parameterization of the concentrated propagation paths and the mapping of the parameters $\boldsymbol{\theta}_{sp}$ to $\mathbf{h}_s$ see Richter (2005).

2.2. Dense Multipath Components

A discussion of the model for the DMC can be found in (Richter, 2005). The model is based on the observation that the power delay profile has an exponential decay over time-delay 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 (Erceg et al., 1999; Cassioli et al., 2002; Pedersen et al. , 2000) (see also

Fig. 1)

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

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

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

where $\beta = B_d/(Mf_0)$ is the normalized coherence bandwidth, and f_0 is the sampling interval in the frequency domain. Let $\boldsymbol{\kappa}(\boldsymbol{\theta}_{c_k})$, $\boldsymbol{\theta}_{dmc} = [\alpha_1, \beta, \tau]^{\mathrm{T}}$ denote a sampled version of the correlation function (4). In frequency-domain it may be written as

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

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

$$\Psi(f_1, f_2) = \psi(f_1 - f_2). \quad (6)$$

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

$$\mathbf{R}_f(\boldsymbol{\theta}_{dmc}) = \mathrm{toep}\big(\boldsymbol{\kappa}(\boldsymbol{\theta}_{\mathrm{dmc}}), \boldsymbol{\kappa}^{\mathrm{H}}(\boldsymbol{\theta}_{\mathrm{dmc}})\big). \quad (7)$$

The covariance matrix $\mathbf{R}_f(\boldsymbol{\theta}_{dmc})$ derived so far describes the distribution of the DMC only in the frequency domain. In general, we have to represent the second order statistics of the dense multipath components with a 10-dimensional function, i.e.

$$\psi(f_1, f_2, \varphi_{T,1}, \varphi_{T_2}, \vartheta_{T,1}, \vartheta_{T,2}, \varphi_{R,1}, \varphi_{R,2}, \vartheta_{R,1}, \vartheta_{R,2}).$$

Provided the WSSUS assumption (Bello , 1963) can be applied 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

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

The full covariance matrix $\mathbf{R}_d(\theta_{dmc}) = \mathrm{E}\left\{\mathbf{h}\mathbf{h}^{\mathrm{H}}\right\}$ of the dense multipath components is of size $N_R N_T M_f \times N_R N_T M_f$.

Since the available measurement apertures in the the spatial (angular) domain available are small, no satisfactory parametric models for the complete covariance matrix could be developed so far. Hence, it is assumed that the DMC are i.i.d. in the remaining domains. Consequently, the covariance matrix of the DMC $\mathbf{h}_d$ is assumed to have structure

$$\mathbf{R}_d(\theta_{dmc}) = \mathbf{R}_R(\theta_{dmc}) \otimes \mathbf{R}_T(\theta_{dmc}) \otimes \mathbf{R}_f(\theta_{dmc}),$$

where $\mathbf{R}_R(\theta_{dmc}) \in C^{N_R \times N_R}$ and $\mathbf{R}_T(\theta_{dmc}) \in C^{N_T \times N_T}$ describe the spatial distribution of the dense multipath components at the receiver and the transmitter position, respectively. Furthermore, since the process is assumed to be spatially i.i.d., the full covariance matrix has structure

$$\mathbf{R}_d(\theta_{dmc}) = \mathbf{I}_{M_R} \otimes \mathbf{I}_{M_T} \otimes \mathbf{R}_f(\theta_{dmc}).$$

One should note, that a model, that takes also the angular distribution of the dense multipath components into account, has been proposed in (Ribeiro et al., 2005). It can be used to describe the spatial correlation of the DMC at the transmitter as well as at the receiver site. The model is based on a mixture of Van Mises distributions. However, the model has not been verified so far by measurements.

3. CHANNEL MEASUREMENTS AND PARAMETER ESTIMATION

An observation $\mathbf{x}$ of the radio channel $\mathbf{h}$ acquired with a channelsounder (www.channelsounder.com , 2006; www.propsim.com , 2006) contains also additive measurement noise $\mathbf{w} \in \mathbb{C}^{M_f N_R N_T \times 1}$. The measurement noise is a realization of the i.i.d. circular complex Normal distributed process $\mathcal{N}_c(\mathbf{0}, \sigma_w^2 \mathbf{I})$, 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

$$\mathbf{x} \sim \mathcal{N}_c\left(\mathbf{h}_s(\boldsymbol{\theta}_{sp}), \mathbf{R}_d(\boldsymbol{\theta}_{dmc}) + \sigma_w^2 \mathbf{I}\right).$$

A maximum likelihood estimator (MLE) for the parameters $\boldsymbol{\theta}_{sp}$, $\boldsymbol{\theta}_{dmc}$, and σ_w^2 RIMAX has been proposed in (Thomä et al. , 2004; Richter, 2005). It exploits the fact that the parameters of the two components of the channel model are asymptotically independent. Therfore, one can decouple the estimation problem into two estimation problems (Richter, 2005). The resulting algorithm is iterative and alternates between the maximization of the likelihood function with respect to the parameters $\boldsymbol{\theta}_{sp}$ and $\boldsymbol{\theta}_{dmc}$.

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 (Richter et al., 2005) 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 $\boldsymbol{\theta}_{sp}$ using an extended Kalman Filter (EKF) in (Richter et al., 2005). The state-space model, and in turn the estimator have been refined in (Salmi et al., 2006). 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 (Richter et al., 2006) a state-space model for the parameters $\boldsymbol{\theta}_{dmc}$ 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 $\boldsymbol{\theta}_{sp}$ and $\boldsymbol{\theta}_{dmc}$ is the best estimator available nowadays. Here best means in terms of estimation variance and computational complexity.

4. RADIO CHANNEL CAPACITY ESTIMATION

For a time invariant channel, the mutual information [bits] or channel capacity [bps/Hz] is given by (Telatar , 1999)

$$c\left(\boldsymbol{\mathcal{H}}\right) = \log_2\left(\det\left(\mathbf{I} + \frac{\rho}{N_T}\boldsymbol{\mathcal{H}}\boldsymbol{\mathcal{H}}^{\mathrm{H}}\right)\right). \quad (8)$$

Since the broadband channel matrix $\boldsymbol{\mathcal{H}}$ is assumed to be a block diagonal matrix (1), the channel capacity (8) can also be expressed as

$$c\left(\boldsymbol{\mathcal{H}}\right) = \sum_{k=1}^{M_f}\log_2\left(\det\left(\mathbf{I} + \frac{\rho}{N_T}\mathbf{H}_k\mathbf{H}_k^{\mathrm{H}}\right)\right). \quad (9)$$

Let $\lambda_{i,k}$ be the eigenvalues of $\mathbf{H}_k\mathbf{H}_k^{\mathrm{H}}$, than (9) can also be computed by

$$c\left(\boldsymbol{\mathcal{H}}\right) = \sum_{k=1}^{M_f}\sum_{i=1}^{M_r}\log_2\left(1 + \frac{\rho}{N_T}\lambda_{i,k}\right). \quad (10)$$

The channel matrices $\mathbf{H}_k$ are not accessible from radio channel measurements. The measured channel matrices are disturbed by i.i.d. cirular Normal distributed noise (11).

$$\tilde{\mathbf{H}}_k = \mathbf{H}_k + \sigma_{\mathrm{w}}^2\mathbf{W} \quad (11)$$

The expected value of the eigenvalues of $\tilde{\mathbf{H}}_k\tilde{\mathbf{H}}_k^{\mathrm{H}}$ are related to the eigenvalues $\lambda_{i,k}$ according to

$$\mathrm{E_w}\left\{\tilde{\lambda}_{\mathrm{i,k}}\right\} = \lambda_{\mathrm{i,k}} + \sigma_{\mathrm{w}}^2, \quad (12)$$

if $\mathrm{E_w}\left\{\tilde{\lambda}_{\mathrm{i,k}}\right\} >> \sigma_{\mathrm{w}}^2$ holds. Having an estimate of σ_w^2 one can reduce the influence of the measurement noise on the estimated channel capacity, by correcting the estimated eigenvalues.

To show the influence of the measurement noise on the estimated channel capacity a Monte-Carlo simulation with 1000 realizations has been carried out. The channel matrices had i.i.d. Rayleigh-fading elements with unit variance. The number of transmit antennas and receive antennas has been chosen as $N_T = 16$ and $N_R = 16$, respectively. Three simulations with different ranks of the channel matrix (4, 8, and 16) have been conducted. The results are shown in Figure 2-7. The influence of the measurement noise on the estimated channel capacity is significant. The MIMO-capacity of channels having very low rank cannot be estimated reliably for low measurement SNRs. Even if the variance of the measurement noise is taken into account the measurement SNR has to be at least 20dB to get reasonable accurate estimates (less than 10% error) of the channel capacity for a tranceiver-SNR of 10dB. With increasing rank of the channel matrix the accuracy of the estimated channel capacity is improving. In the simulated example, the channel capacity for a tranceiver SNR $\rho = 10$dB can be estimated with less than 10% error, if the rank is larger than 8 and the measurement SNR is larger than 7dB. Altogether, this example shows that the influence of observation noise on the estimated MIMO channel capacity is significant, and can not be neglected.

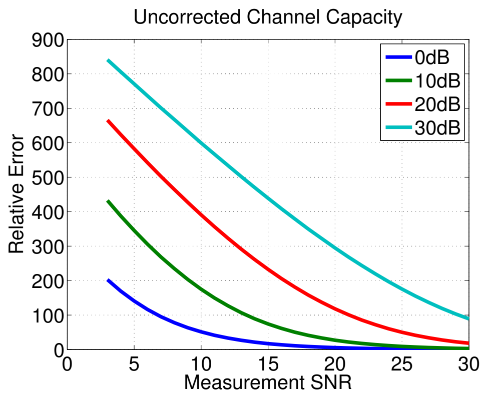

Figure 2. Estimated channel capacity of a 16×16 *Rayleigh fading MIMO channel having* $\mathrm{rank} = 1$ *for a tranceiver SNR of 0dB, 10dB, 20dB and 30dB, without reduction of the measurement noise influence.*

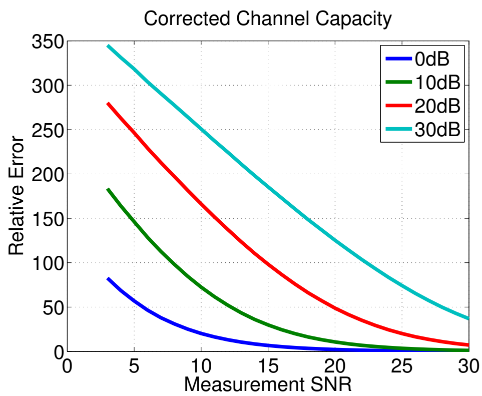

Figure 3. Estimated channel capacity of a 16×16 *Rayleigh fading MIMO channel having* $\mathrm{rank} = 1$ *for a tranceiver SNR of 0dB, 10dB, 20dB and 30dB, with reduction of the measurement noise influence.*

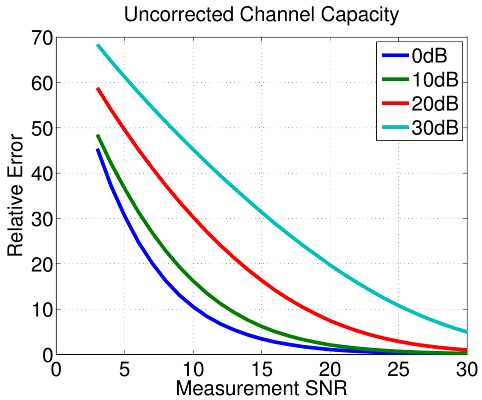

Figure 4. Estimated channel capacity of a 16×16 *Rayleigh fading MIMO channel having* $\text{rank} = 8$ *for a tranceiver SNR of 0dB, 10dB, 20dB and 30dB, without reduction of the measurement noise influence.*

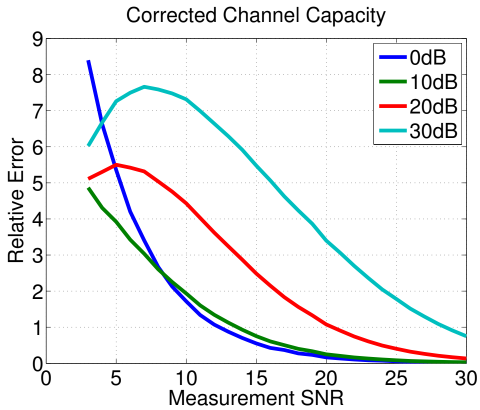

Figure 5. Estimated channel capacity of a 16×16 *Rayleigh fading MIMO channel having* $\text{rank} = 8$ *for a tranceiver SNR of 0dB, 10dB, 20dB and 30dB, with reduction of the measurement noise influence.*

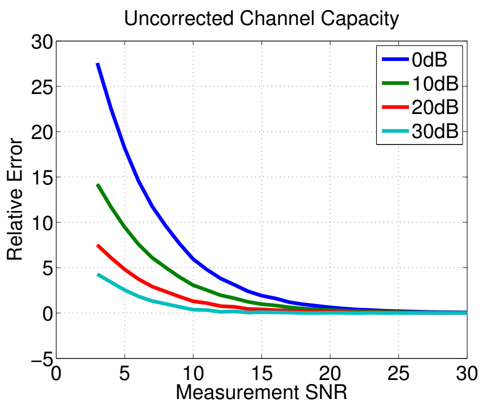

Figure 6. Estimated channel capacity of a 16×16 *Rayleigh fading MIMO channel having* $\text{rank} = 16$ *for a tranceiver SNR of 0dB, 10dB, 20dB and 30dB, without reduction of the measurement noise influence.*

5. ESTIMATION EXAMPLE

The EKF based parameter estimator mentioned in Section 3 has been applied to channel sounding data, measured in the city center of Ilmenau,Germany (Trautwein et al., 2005). The channel sounder used was a RUSK ATM (MIMO) (www.channelsounder.com , 2006). The measurement setup applied in the measurement campaign is outlined in Table 1. A map of the measurement scenario is shown in Figure 11. 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 the route from point (3) to point (16) has been used in this example. The trolley with the MS antenna array has been driven on the right side of the street. Every 20.48ms a channel observation has been measured. The radio channel has been measured for $\approx 60\text{s}$ leading to more than 2900 channel observations.

Figure 8 shows estimates of the powers of the whole radio channel, the power of the specular propagation paths, and the power of the DMC. As a reference also the total noise power is shown in the same figure. At $\approx 37s$ the channel changes from NLOS to LOS. From $\approx 50s$ to $\approx 57s$ the LOS was again obstructed by a van parked close to point (16).

Figure 9 shows the relative contribution of the DMC to the received power. The dense multipath components contribute sometimes $\approx 90\%$ to the radio transmission. Figure 10 shows the MIMO channel capacity with optimum power control at the transmitter, for a tranceiver SNR of $\rho = 10\text{dB}$. If the propagation is mainly supported by the dense multipath components the channel capacity is about 90% of the capacity of an equivalent 16×16 Rayleigh fading channel (straight line at $39.6\text{bps}/\text{Hz}$). For the computation of the equivalent Rayleigh fading channel the influence of the antenna elements of the AP array (120 deg directivity) has been taken into account. Since the received power is changeing slowly over time, TX power control can be considered feasible in the measured scenario.

6. CONCLUSION

The contribution of the distributed diffuse scattering to terrestrical radio propagation is significant. The contribution in terms of received power varies between 10% and more than 90%. In the analyzed channel sounding measurement the contribution to the channel capacity was sometimes close to 100%. In this situation the channel capacity is close to the capacity of a spatially i.i.d. Rayleigh fading channel. This result supports the hypothesis that the distributed scattering has usually a wide angular spread at the transmitter and the receiver. However,

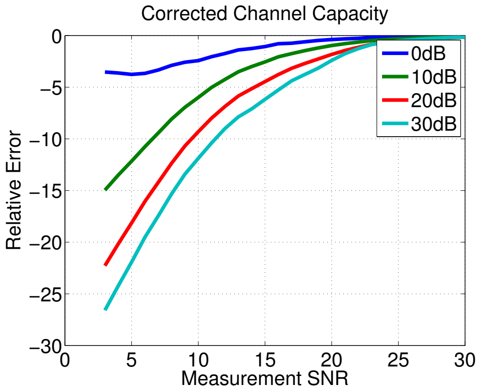

Figure 7. Estimated channel capacity of a 16×16 *Rayleigh fading MIMO channel having* $\mathrm{rank} = 16$ *for a tranceiver SNR of 0dB, 10dB, 20dB and 30dB, with reduction of the measurement noise influence.*

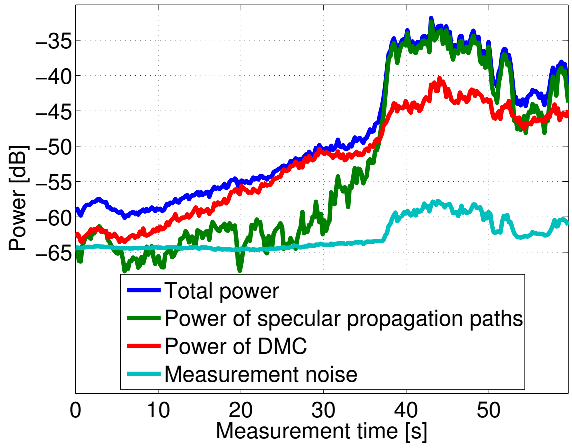

Figure 8. Estimated contribution to the received power of the channel components for route 2.

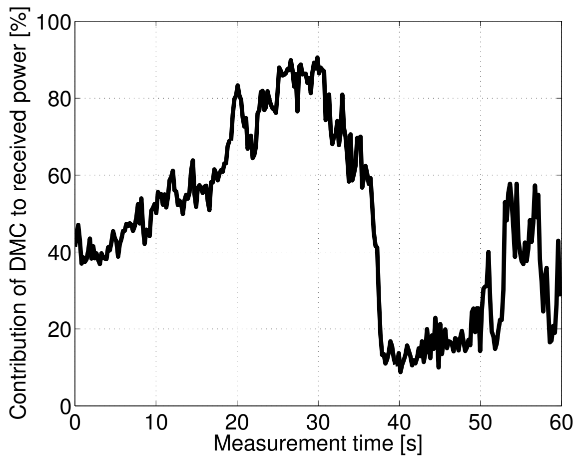

Figure 9. Estimated contribution of the DMC to the received power in measurement route 2.

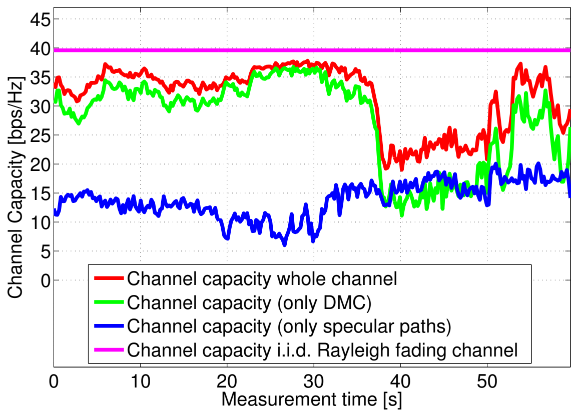

Figure 10. Estimated channel capacities for route 2 with an average tranceiver SNR of $10\mathrm{dB}$

Table 1. 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 $\approx$ $120\,\mathrm{deg}$ directivity, about $4\mathrm{m}$ above ground
MT antenna:	A 16-element uniform circular array with radius $\approx$ 0.38λ, vertically polarized, on the top of a trolley, and about $1.3\mathrm{m}$ above ground

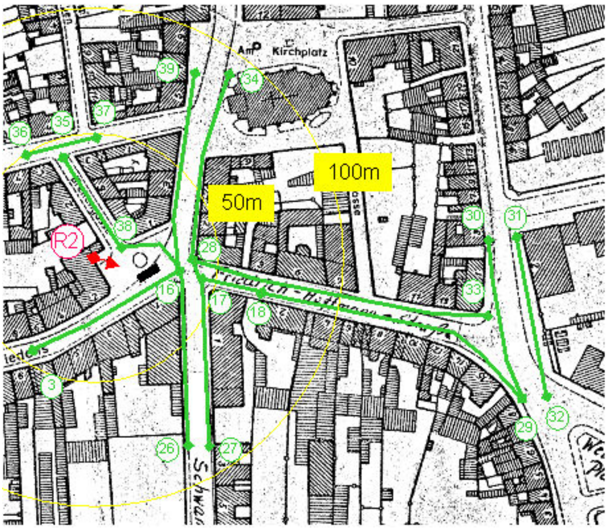

Figure 11. Map of measurement scenario.

for reliable modelling of the DMC the analysis of more MIMO measurements in different scenarios is necessary.

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