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1. INTRODUCTION

This paper provides implementation details and performance results for the proposed MIMO system channel model introduced in [1]. The model provides channel transfer matrices derived from the superposition of plane waves. The advantages of the proposed model can be summarized as follows:

1. It preserves joint statistics in temporal, spatial, and frequency domains, as verified through power delay profiles, temporal correlations, and cumulative distributions of capacity.
2. It is easy to implement, requiring reasonable simulation time.
3. Channel matrices can be derived from Node B and UE parameters (e.g., PAS, AOA, AOD) and does not require additional parameters to be specified (e.g., for scatterers).

2. COMPARISON WITH OTHER MODELING TECHNIQUES

There are three popular techniques in MIMO channel modeling: ray-tracing, correlation models, and scatterer models.

The **ray-tracing model** tries to model the exact locations of the scatterers. Free-space propagation, reflection, diffraction and scattering are modeled to follow each of the propagation path through the channel. Reasonably good prediction can be achieved for indoor and short-range outdoor channels, however, it is too complex for the majority of outdoor environments [2] and impractical for system simulations.

In the **correlation model**, the spatial correlation is captured by multiplying a complex Gaussian IID channel matrix by the correlation matrices at the receiver and the transmitter [3],[4]. This is the approach chosen for the MIMO link level channel model [5]. The disadvantage of this approach is that the temporal variations of the channel are either neglected or separately modeled from the spatial correlation. Consequently, the joint statistics in temporal and spatial domains may not be preserved.

The **scattering model**, adopted by the COST259 model [6] and also proposed in [7], assumes a particular distribution of scatterers and generates channel realizations based on the interaction of scatterers and planar wavefronts. The resulting statistical properties of the channels are dependent on the distribution of the scatterers, and these distributions can be chosen to accurately represent a variety of MIMO channel environments. The disadvantages of this approach are 1) the complexity in parameterizing and generating the scattering distributions for a variety of channel environments, and 2) the resulting simulation time. A large number of parameters needs to be specified

Our proposed channel model adopts the classical plane wave assumption, where the electromagnetic field in space is generated by a superposition of plane waves. However, unlike in ray-tracing or scattering models where the wireless propagation medium is explicitly modeled to affect the plane waves, we will treat the propagation medium as a “black box” so that the received signal’s characteristics are derived entirely from statistical properties of the channel. In other words, given the statistical properties such as shadow fading, delay statistics and angular distribution, the model constructs the channel transfer matrices according to these statistics and independent of the “black box”.

Because the wireless medium is not explicitly modeled, the complexity of the proposed model is significantly less than that of either the ray-based or scattering models. As an example, the proposed model requires only 8 minutes of run time to generate the channel matrices for 10 seconds of simulation time for a Pedestrian A model with 4 transmitters and 4 receivers. The simulation was performed using Matlab code, and the run time can be reduced further using C code.

Despite the simplicity of the proposed model, it preserves the joint statistical characteristics in the temporal, spatial, and frequency domains. The ray tracing and scattering models do so as well, but with much greater complexity. Using the simplified black box approach, the model sufficiently captures the statistical properties of the channel in the resulting channel transfer functions. In section 4, we verify the consistency of the statistics and show that the correlations and capacities are consistent with those achieved using the correlation model.

3. DESCRIPTION OF PROPOSED MODEL

A prior version of this channel model was presented in [1]. Consider a UE with N_R receive antennas receiving signals from B Node B’s, each with N_T transmit antennas. The signal from the first Node B is the desired signal, and the others are assumed to be interference. Assume that the power delay profile (PDP) consists of J paths and that it is the same for each of the B bases. We further assume that the PDP is randomly chosen for each UE from some set of PDPs and is fixed for the duration of one drop cycle. The received signal at the UE at time t can be written as the sum of signals received from B Node B’s:

$$\mathbf{r}(t) = \sum_{b=1}^B \sqrt{\mathbf{g}_b} \sum_{j=1}^J \mathbf{H}_{b,j}(t) \mathbf{x}_b(t - \mathbf{t}_{b,j}) + \mathbf{n}(t) \quad (1)$$

where $\mathbf{r}(t)$ is an N_R dimensional vector; \mathbf{g}_b is the relative power from base b which accounts for large scale signal variations due to the path loss and lognormal fading; $\mathbf{H}_{b,j}(t)$ is an N_R -by- N_T matrix which gives the spatial channel coefficients between the UE and base b for the j th path ($j = 1 \dots J$) at time t ; $\mathbf{x}_b(t - \mathbf{t}_j)$ is an N_T dimensional vector which gives the data signal sent by the b th base with delay \mathbf{t}_j induced by the j th path; and $\mathbf{n}(t)$ is the N_R dimensional additive noise vector which accounts for residual noise and interference not accounted for by bases $b = 2 \dots B$. The pulse shaping is accounted for in $\mathbf{x}_b(t - \mathbf{t}_j)$. For the duration of each drop (on the order of 10 seconds), the values \mathbf{g}_b and $\mathbf{t}_{b,j}$ are assumed to be static.

Let $h_{q,s,j}(t)$ be the (q, s) component of the matrix $\mathbf{H}_{b,j}(t)$ ($q=1\dots N_R, s=1\dots N_T$) which represents the complex phase and amplitude of the j th path delay from the s th antenna of base b to the q th antenna of the UE. We drop the dependence on b for simplicity. The value of $h_{q,s,j}(t)$ can be determined by adding a large number of sinusoidal waves whose phases and amplitudes are determined by the geometric parameters of the system:

$$h_{q,s,j}(t) = \sqrt{P_j} \sum_{l=1}^L A_l \sqrt{P_{NodeB,j}(\mathbf{j}_{AOdl})} e^{ikd_s \cos(\mathbf{j}_{AOdl} - \mathbf{j}_s)} \sqrt{P_{UE,j}(\mathbf{j}_{AOAl})} e^{ikd_q \cos(\mathbf{j}_{AOAl} - \mathbf{j}_q)}, \quad (2)$$

where

- P_j is the relative power of the j th path,
- L is the number of sinusoids for the j th path,
- A_l is a complex Gaussian random variable (zero mean, unit variance) which is the complex amplitude of the l th sinusoid,
- \mathbf{j}_{AOdl} is a uniformly distributed random variable between 0 and $2\mathbf{p}$ which is the angle of departure of the l th sinusoid of the j th path from the Node B antenna s (we have not written the dependency on s and j for simplicity),
- \mathbf{j}_{AOAl} is a uniformly distributed random variable between 0 and $2\mathbf{p}$ which is the angle of arrival of the l th sinusoid of the j th path at the UE antenna q (we have not written the dependency on q and j for simplicity),
- $P_{NodeB,j}(\mathbf{j}_{AOdl})$ is the combined antenna response and angular spectra of the j th path at the NodeB as a function of the angle of departure of the l th sinusoid (for example, if the j th path has a Laplacian distribution, $P_{NodeB,j}(\mathbf{q})$ is the product of the antenna response centered about 0 degrees and the Laplacian distribution centered about the mean angle of departure of the j th path),
- $P_{UE,j}(\mathbf{j}_{AOAl})$ is the combined antenna response and angular spectra of the j th path at the UE as a function of the angle of arrival of the l th sinusoid,
- d_s and \mathbf{j}_s are the polar coordinates of Node B antenna s ,
- d_q and \mathbf{j}_q are the polar coordinates of UE antenna q , accounting for the motion of the UE which is function of the direction of travel and velocity (we have not written the dependency on time t for simplicity),
- $k = \frac{2\mathbf{p}}{\lambda}$ is the wave number.

Figure 1 illustrates the effects on the l th plane wave induced by the propagation medium:

- 1) It changes its direction from \mathbf{j}_{AOdl} to \mathbf{j}_{AOAl} .
- 2) It attenuates the signal by a factor $\sqrt{A_l}$.
- 3) It creates an aggregate phase shift which is a function of \mathbf{j}_{AOdl} , \mathbf{j}_{AOAl} , and the coordinates of the Node B and UE antennas.

To avoid introducing numerical degeneracy, the total number of plane waves L for each delay should be larger by an order of magnitude than the number of elements in the \mathbf{H} matrix, say $L = 400$ for a system with 4 transmitters and 4 receivers. For simplicity, the number of plane

waves is equal to L for different delays. Limitations on effective rank of \mathbf{H} would then arise from the angular constraints implicit in the angular spectra, as well as from a LOS component, added below. The random variables $A_l, \mathbf{j}_{AOdl}, \mathbf{j}_{AOAl}$ are generated once at the beginning of the drop and kept constant for the duration of N frames. Many parameters derived for the link level channel model such as power delay profiles and angular spectra can be used here in our proposed model. Other parameters such as the mean angles of departure and arrival can be randomly generated at the beginning of each drop for each user. Similarly, the polar coordinates of the array elements can be determined at the beginning of each drop as a function of the array configurations. The coordinates of the UE elements can be tracked in time as a function of the direction of travel and velocity. Hence the time variations in $h_{q,s,j}(t)$ will result from the time-dependent Doppler term implicit in d_q and \mathbf{j}_q .

If it is desired to include a line of sight (LOS) path, defined as a plane wave signal from the Node B to the UE, equation (1) can be rewritten as

$$\mathbf{r}(t) = \sum_{b=1}^B \sqrt{\mathbf{g}_b} \left[\sum_{j=1}^J \mathbf{H}_{b,j}(t) \mathbf{x}_b(t - \mathbf{t}_{b,j}) + \mathbf{H}_b^{LOS}(t) \mathbf{x}_b(t) \right] + \mathbf{n}(t) \quad (3)$$

where $\mathbf{H}_b^{LOS}(t)$ is the N_R -by- N_T LOS matrix whose (q, s) component ($q = 1 \dots N_R, s = 1 \dots N_T$) is given by

$$h_{q,s}^{LOS}(t) = \sqrt{K} e^{ikd_s \cos(\mathbf{j}_{AOD,LOS} - \mathbf{j}_s)} e^{ikd_q \cos(\mathbf{j}_{AOALOS} - \mathbf{j}_q)}, \quad (4)$$

where

K is the Rician K factor

$\mathbf{j}_{AOD,LOS}$ is the angle of departure of the LOS component from the Node B antenna s (we have not written the dependency on s for simplicity),

\mathbf{j}_{AOALOS} is the angle of arrival of the LOS component at the UE antenna q (we have not written the dependency on q for simplicity),

and the other terms are defined above.

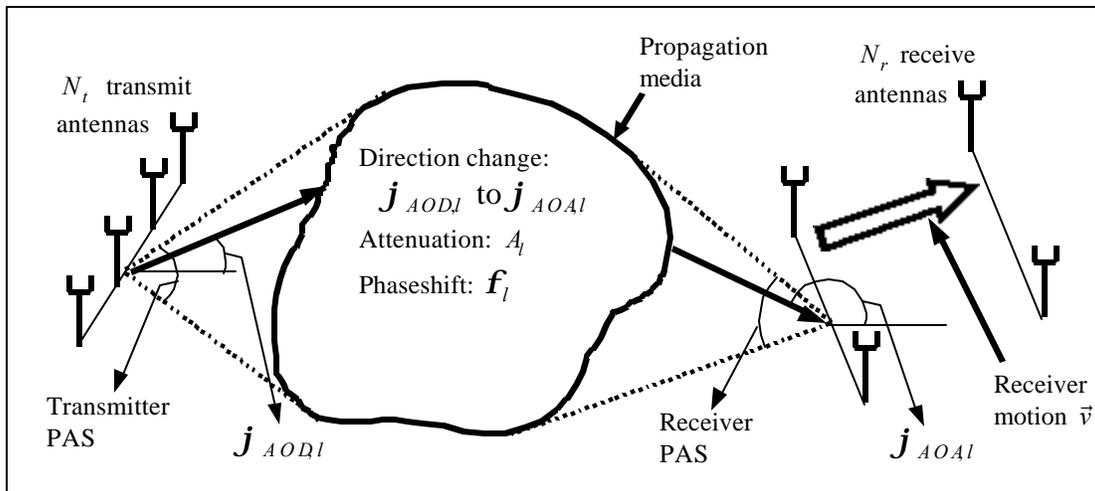


Figure 1. MIMO channel model

4. STATISTICAL PROPERTIES OF THE SIMULATED CHANNEL

This section verifies the statistical consistency of the model with respect to the PDP, temporal correlations, and CDF of the capacity. Without loss of generality, the channel modeling approach is demonstrated for a typical system setup of a 4 by 4 MIMO system (i.e., 4 Node B transmit antennas and 4 receive antennas per UE) with a uniform linear array at both ends. The transmitter array elements have a separation of 4 wavelengths, and the receiver array has a separation of 0.5 wavelengths. We have chosen the number of planar waves to be $L = 400$.

4.1 Power Delay Profile -- Temporal Dispersion

The temporal dispersion of the channel is modeled by the power shaping of $\sqrt{P_j}$ according to a given PDP. From (2), for any transmitter-receiver pair, by averaging the $h_{q,s,j}(t)$ entries in time, the average received power for a given path will collapse to the desired power given by $\sqrt{P_j}$. Figure 2 shows the actual PDP for the Pedestrian A channel and the measured PDP derived from the averaged H matrices based on (2). The simulation produces almost exact shape of the average PDPs. Since the rays for different multipath delays are different, it is generally assumed that the multipath components are not correlated.

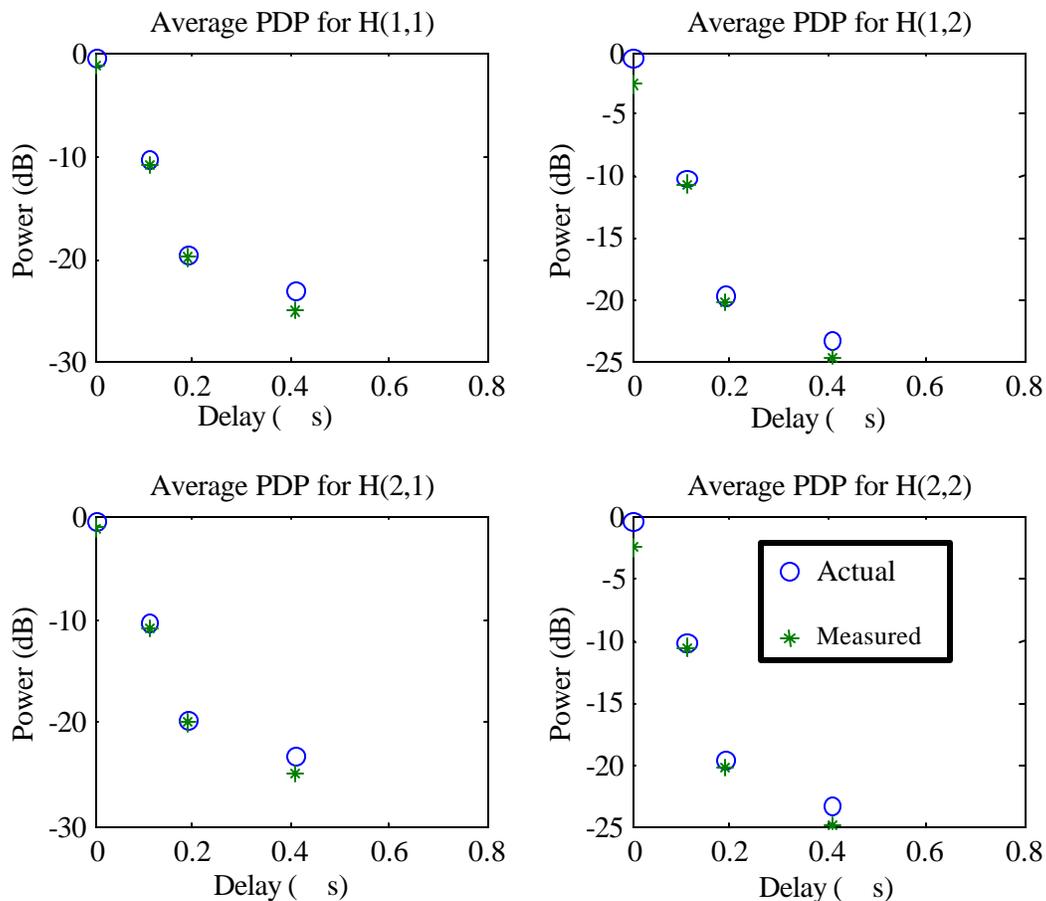


Figure 2. Power delay profile for Pedestrian A channel

4.2 Power Azimuth Spectrum -- Spatial Dispersion

The spatial dispersion of the MIMO channel is modeled by PAS of the AOD at the transmitter and the PAS of the AOA at the receiver. Typical PAS from the measurements are Laplacian distributed AOD with 2 degree rms angular spread at the base and 35 degree rms angular spread at the mobile [8]. In general, the PAS at the mobile and PAS at the base can be dependent. In such a case, a joint PAS needs to be used in (2). If the scattering medium is rich enough, then it is reasonable to assume that the PASs at the mobile and base are independent. One way to generate such independent PASs is to have L_1 AOAs and L_2 AODs, and let the total of $L = L_1 L_2$ rays have all the possible combination of AOA and AOD. By doing this, power shaping of PAS at the base will have no effects on the PAS at the mobile. As a result of the angular spread at the transmitter and receiver, the entries of H are spatially correlated. For the results in this section, we use $L_1 = L_2 = 20$.

4.3 Doppler Spectrum -- Frequency Dispersion

The frequency dispersion of the MIMO channel is caused by the motion of the receiver in a spatially dispersive channel, and it is modeled implicitly in the phase term of $e^{ikd_q \cos(j_{AOA} - j_q)}$. The Doppler spread is a function of angular spread, AOA, and mobile velocity. Due to the motion of the mobile and angular dispersion at the receiver, the entries of H are temporally correlated.

4.4 Correlation Results

Figure 3 shows the temporal correlation results for the channel with 35 degree of angular spread at the UE and speed of 10 km/hr. Note that the results of temporal correlation are shown in terms of the wave lengths. These results can be mapped to correlation in time according to the UE velocity. For small separation distances or short time separations, the exact correlation and the simulated correlation results are very close. For large separations, the simulated correlations are slightly larger than the actual values. This difference is mainly caused by the limited discrete samples in the angular domain. By increasing the number of plane waves, the difference decreases. Hence reducing the complexity of the model by reducing L , the model results in higher correlations, leading to more conservative, lower capacity estimates. This characteristic of the channel model is desirable for the system simulations. Even in the range of large separations, the correlation of simulated channels is comparable to the correlation level of Jakes model, which is generally assumed uncorrelated. In summary, our channel model provides accurate correlation results for the purpose of algorithm evaluations.

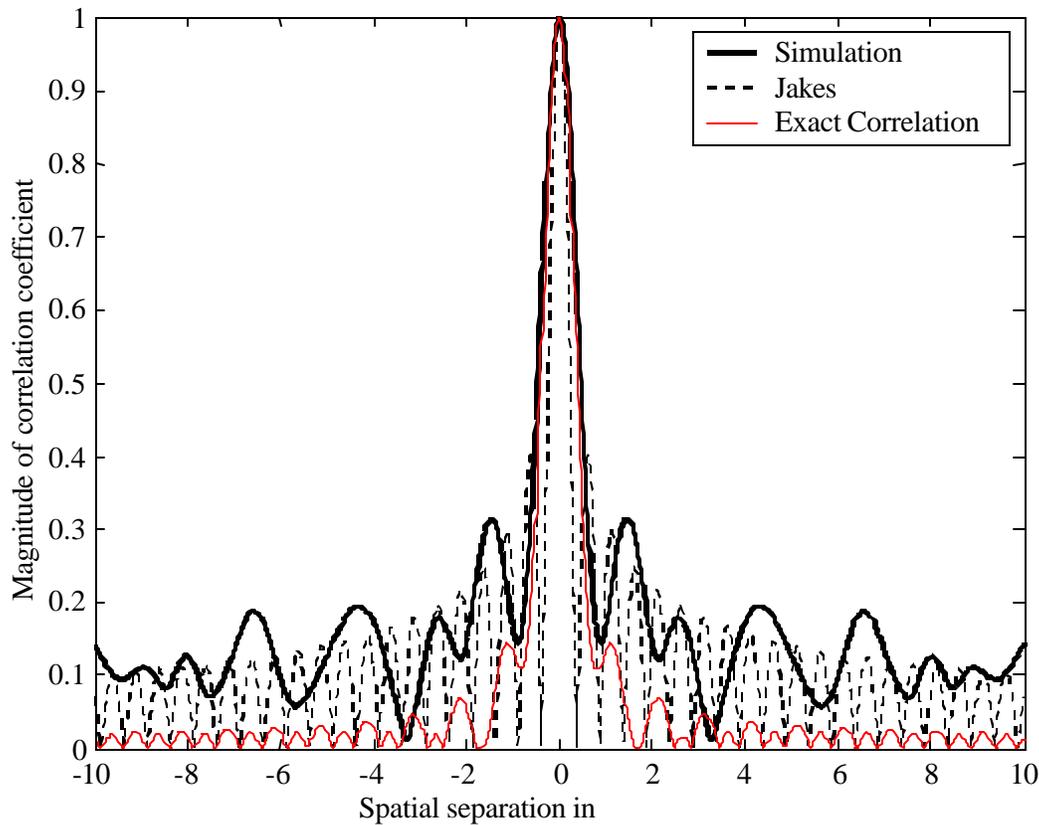
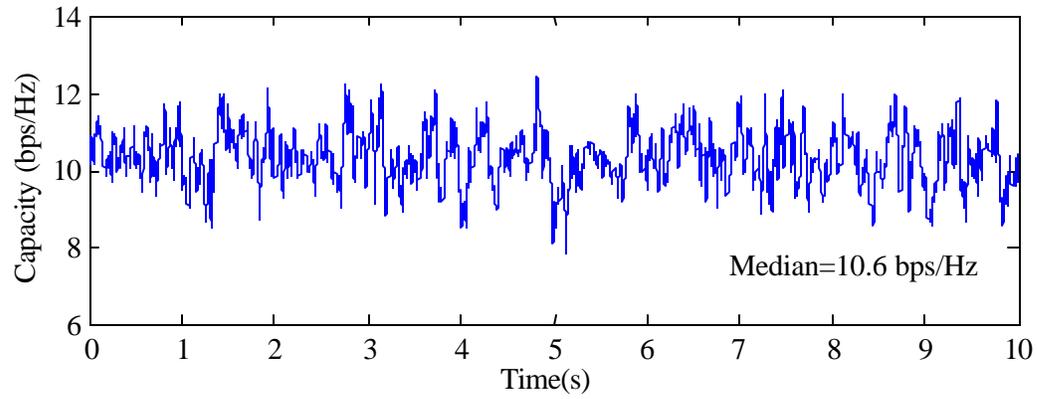


Figure 3. Temporal correlations of $H(1,1)$

4.5 Capacity Results

The attainable MIMO capacity is a function of the correlation of the H matrices. The Shannon capacity of the channel is evaluated for the simulated H matrices and compared to the capacity of a complex Gaussian i.i.d. channel. One example is shown in Figure 4 where the simulated H matrices are for the Pedestrian A PDP, Laplacian PAS with 5 degree angular spread at the Node B and 35 degree angular spread at the UE. The UE speed is 10 km/hr with the direction of travel perpendicular to the mean AOA. The capacity is calculated for the equivalent narrowband channel from the PDPs by summing the multipath component for each of the subchannels. Unlike the complex Gaussian i.i.d. case, where the channels are uncorrelated in time, the simulated H matrices exhibit temporal correlation. The CDFs of these channels are shown in Figure 5. Due to the limited angular spreads at the base station and the mobile, the capacity of the simulated channel is less than the Gaussian i.i.d. channel. Note that the simulated capacity CDF based on the proposed plane wave model is almost exactly as the CDF derived from the correlation based model.

Laplacian PAS, Node B spread 5deg, UE spread 35 deg, V=10Km/hr, AOA=0deg, f = 2GHz



Complex Gaussian i.i.d. capacity

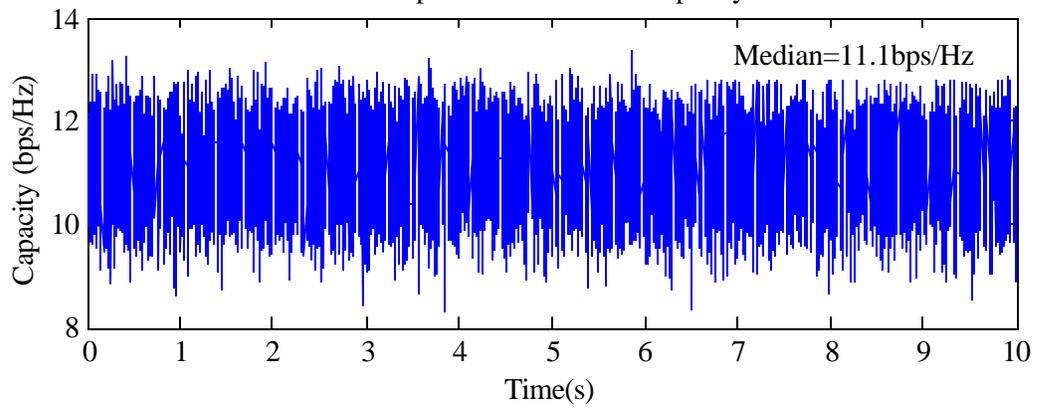


Figure. 4. Capacity variation for simulated channel and Gaussian i.i.d. channel

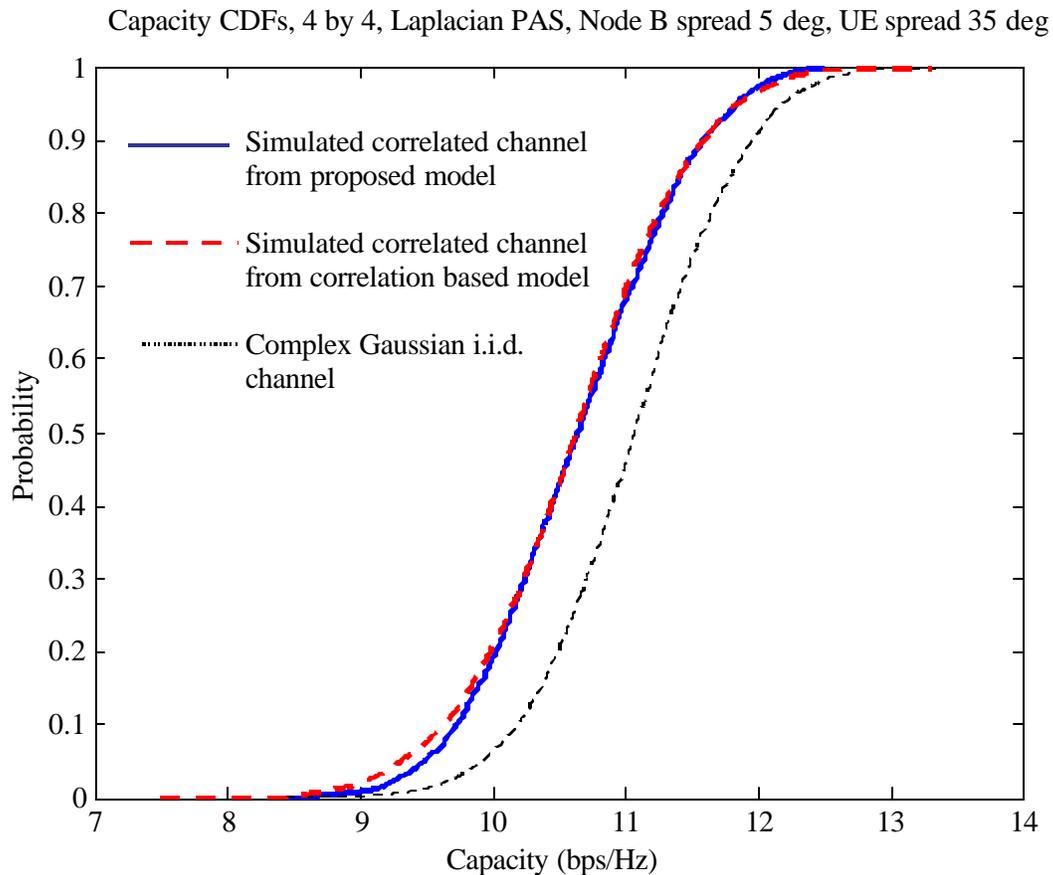


Figure 5. Cumulative distribution functions of capacity

5. SUMMARY

This paper presented a wideband MIMO system-level channel modeling technique to generate channel transfer matrices that are consistent statistically in space, time and frequency for use in the MIMO system level simulations. The simulated channels are verified with respect to power delay profile, correlation, and capacity CDFs. Numerical results showed a good match with the expected statistical properties. The proposed technique is much simpler to implement than other models such as ray-tracing and scattering that give comparably reliable statistical results.

6. REFERENCES

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