EMPIRICAL EIGENANALYSIS OF INDOOR UWB

PROPAGATION CHANNELS

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Abstract

This work aims at characterizing the second order statistics of indoor Ultra-Wideband (UWB) channels using channel sounding techniques. We present measurement results for different scenarios conducted in a laboratory setting at Institut Eurécom. These are based on a eigen-decomposition of the channel autocovariance matrix, which allows for the analysis of the growth in the number of significant degrees of freedom of the channel process as a function of the signaling bandwidth as well as the statistical correlation between different propagation paths. We show empirical eigenvalue distributions as a function of the signal bandwidth for both line-of-sight and non line-of-sight situations. Furthermore, we give examples where paths from different propagation clusters (possibly arising from reflection or diffraction) show strong statistical dependence.

I. INTRODUCTION

The use of Utra-Wide Band (UWB) signaling techniques are being considered for short-range indoor communications, primarily for next generation high bit-rate *Wireless Personal Area Networks (WPAN)*. Initial work in this direction were carried out by Sholtz [1, 2], using the most common form of signaling based on short-term impulses, where information is carried in their position. Such techniques, as well

as others are being considered in the standardization process of the IEEE 802.15.3a WPAN proposal (see http://grouper.ieee.org). At the same time, regulatory aspects are quickly being defined by the FCC.

The expected bandwidths of these systems are one the order of gigahertz, which has significant implications both for systems design and implementation. The goal of this work is to determine the result of these extremely large system bandwidths on the second order statistics of the propagation channel as it is seen by the underlying system. We are primarily interested in assessing the growth in degrees of freedom needed to characterize the channel as a function of the system bandwidth.

Other measurements studies on the UWB propagation channel are appearing, we cite for instance [3-6]. This work is complementary, in the sense that we use state-of-the-art wideband measurement equipment to determine the growth of the number of significant *free degrees of freedom* (*DoF*) of the propagation channel as a function of the signaling bandwidth, based on sub-space techniques. The number of significant DoF is related both to the number of resolvable multipath components and to the diversity order or richness of the indoor channel. Measurements are carried out for both line-of-sight and non-line-of-sight short-range indoor communications.

Section II describes the measurement equipment used in this study as well as the propagation environment. In section III we outline the sub-space methods used for analyzing the second-order statistics of the propagation channel. Section IV describes the numerical results and section V presents the conclusions of this study.

II. UWB CHANNEL MEASUREMENT SCENARIO

A. Equipment and Measurement Setup

The measurement device used in this study is a wideband vector network analyzer (VNA)which allows complex transfer function (e.g. S_{21}) parameter measurements in the frequency domain extending from 10 MHz to 20 GHz. This instrument has low inherent noise (<-110 dBm for a measurement bandwidth of 10 Hz) and high measurement speed (<0.5 ms/point). The maximum number of equally-spaced frequency

samples (amplitude and phase) per measurement is 2001. The measurement data are acquired and controlled remotely using the RSIB protocol over an Ethernet network permitting off-line signal processing and instrument control in MATLAB. The measurement system is shown in figure 2.

In order to perform truly wideband measurements with sufficient resolution several bands can be concatenated using consecutive measurements. In this study we performed measurements from 3 to 9 GHz by concatenating 3 groups of 2001 frequency samples per 2 GHz sub-band (3-5 GHz, 5-7 GHz, 7-9 GHz). This yields a 1MHz spacing between frequency samples. As discussed in the following section, this resolution was found to be sufficient for the analysis of the second order statistics of the channel impulse response. The corresponding maximum time domain resolution is T = 167 ps in a 1µs time-interval.

The use of passive elements in the measurement apparatus (cables, SMA connectors, etc.) imposed a systematic and frequent calibration measurement (which was controlled remotely), in order to compensate undesirable frequency-dependent attenuation factors that could affect the collected data. Following the VNAs manual recommendations, the calibration "through response" type was selected, and the cables and the connectors were included in this calibration.



Fig. 1. S21 Response for the SkyCross UWB Antenna

The wideband antennas employed in this study are omni-directional in the vertical plane and have an approximate bandwidth of 7.5 GHz (varying from 3.1 to 10 GHz). They are not perfectly matched across the entire band, with a VSWR (Voltage Standing Wave Ratio, $VSWR = \frac{1+|\rho|}{1-|\rho|}$, ρ is the wave's coefficient of



Fig. 2. Channel Measurement Setup

reflection) varying from 2 to 5 (as an example the antenna has an efficiency of about 82% at 5.2 GHz [7]). The frequency response of the antenna is shown in figure 1 where we see that it is not completely flat over the band of interest. In what follows, the antennas are considered to be part of the channel response. This is of practical interest, since the true response to the antenna depends heavily on its immediate vicinity, especially in the type of applications envisaged for UWB systems, namely low-height and peer-to-peer communications with the presence of nearby obstacles (people, furniture, walls, etc.). As a result, the true response of the antenna is of secondary importance from a system point-of-view.

B. Measurement Environment

Measurements are performed at spatially different locations for both Line-Of Sight (LOS) and Non Line Of Sight (NLOS). The experiment area is set by fixing the transmitting antenna on a mast at 1 m above the ground on a horizontal linear grid (20 cm) close to VNA and moving the receiver antenna to different locations on a horizontal linear grid (50 cm) in 1 cm steps. In this configuration we assure that the attenuation due to short-term fading effects is virtually constant for all combinations of transmitter and receiver positions. This is explained in more detail in the following subsection. The transmitter/receiver positioning is depicted in figure 3. The height of receiver antenna was also 1m above the ground. This type of propagation scenario clearly targets peer-to-peer applications.



Fig. 3. Transmitter/Receiver positioning

A measurement scenario is described by the transmitter/receiver separation and the presense or lack of a line-of-sight component. The latter was achieved by inserting a large obstacle between the transmitter and receiver in order to block the line-of-sight path. Because of the proximity of the transmitter and receiver, this also has the effect of reducing the amount of scattered paths which reach the receiver and thus the richness of the propagation channel. For the two measurement scenarios with transmitter/receiver separation of 6m representing non line-of-sight and line-of-sight communications, we acquired 1000 different complex frequency responses each. All numerical results reported in this paper refer to the two 6m scenarios. Similar experiments with fewer measurements were also made for transmitter/receiver separations varying from 1-12m in both LOS and NLOS configurations. The measurements were carried out in Institut Eurcom's Mobile Communication Laboratory, which is a typical laboratory environment (radio frequency equipment, computers, tables, chairs, metallic cupboard, glass windows,...), as shown in figure 4, rich in reflective and diffractive objects.



Fig. 4. Channel Measurement Environment

III. UWB EIGENANALYSIS

The classical approach [9] [10] [11] used for the characterization of the selectivity of multipath fading channels, is based on root mean square (rms) delay spread T_d . This parameter or equivalently the *coherence* bandwidth B_c defined as

$$B_c = \frac{k}{T_d} \tag{1}$$

where 0 < k < 1 is a constant, is widely used as a measure of channel frequency selectivity. This approach is generally correct when the signal bandwidth comprises a small number of coherence bandwidths. But in wide band or UWB channels, this study will show that the coherence bandwidth can only be seen as a local measure and does not give an accurate description of channel selectivity. Similar ideas for channel bandwidths on the order of tens of megahertz were also suggested in [12].

A. Eigen-decomposition of covariance matrix

The radio-propagation channel is randomly time-varying due to variations in the environment and mobility of transmitters and receivers. It is classically represented, following the work of Bello [13, 14] by its input delay-spread function $h(t, \tau)$ called also, by abuse of language, the time-varying Channel Impulse Response (CIR). The variable t in the CIR notation represents the time-varying behavior of the channel caused by the mobility of either the transmitter, the receiver or the scatterers. The second variable τ represents the delay domain in which we characterize the channel regarding the most important arriving paths. We consider for each measurement a fixed position at the transmitter and the receiver sides, and a static environment (at least during the time-frame of one measurement). We are thus considering a static channel and we can then simplify the notation of the CIR by dropping its dependence on t. In what follows, $h(\tau)$ is simply replaced by h(t).

Let $\mathbf{h}(\mathbf{t}) = [h_{W,1}(t), h_{W,2}(t), ..., h_{W,N}(t)]$ be the channel process obtained from measurements for N different antenna configurations, where $h_{W,i}(t)$ is expressed as

$$h_i(t) = g_i(t) + n_i(t), i = 1..N,$$
(2)

where $n_i(t)$ is zero-mean additive white Gaussian noise with power spectral density equal to σ_n^2 at all frequencies in the bandwidth of interest. We neglect any non-linear perturbation caused by measurement elements (e.g. VNA amplifiers), which were treated in a more general setting in [19]. We include the frequency response of the antenna as part of the channel response as argued in the previous section, and moreover, the linear response of the equipment is assumed to be perfectly accounted for in the calibration of the measurement apparatus. The noise process, resulting from thermal noise in the receive chain of the VNA and the noise generated by device itself, is assumed to be white in the band of interest. We therefore have that $\mathbf{g}(t) = [g_1(t), g_2(t), ..., g_N(t)]$ are the observations of the noise-free channel process corresponding to N observation positions. Due to the rapid variation of the wave's phase (from 0 to 2π over one wavelength), we can assume that the received electric field at each position represents a zero-mean process, and thus g(t)is taken to be zero-mean. We remark that this does not rule out the possibility of line-of-sight propagation as will be treated shortly.

The VNA provides samples of the observed channel process in the frequency domain, $H(k\Delta f)$, where Δf is the frequency separation, in our case 1 MHz. Furthermore, it is a filtered version of the channel response, where the filter corresponds to an ideal bandpass filter of bandwidth W centered at $f_c = (f_{\text{max}} - f_{\text{min}})/2$, as shown in figure 5. After removing the carrier frequency f_c , we denote the complex baseband equivalent filtered channel by $h_W(t)$. By sampling the frequency response in the VNA we obtain an aliased version of $h_W(t)$ denoted by $\tilde{h}_W(t) = \sum_k h_W(t - k/\Delta f)$. The time-domain samples obtained by performing an inverse discrete Fourier transform (IDFT) on the vector $\mathbf{H} = \left(H(0) \quad H(\Delta f) \quad \cdots \quad H((N-1)\Delta f)\right)^T$ are samples of $\tilde{h}_W(t)$ at sampling frequency W Hz. We note that the choice of frequency separation Δf has an impact on how closely $\tilde{h}_W(t)$ approximates $h_W(t)$ in the interval $[0, 1/\Delta f)$. In our case the choice of $\Delta f = 1$ MHz guarantees that the approximation will be very accurate since the typical delay spread of the considered channels is less than 100 ns and therefore time-domain aliasing will not distort the channel measurement. For this reason we assume in what follows that the measurement equipment provides perfect samples of $h_W(t)$.





Our approach to characterize the UWB propagation channel is based on the analysis of the channel sub-

space and the eigen-decomposition of the covariance matrix, $\mathbf{K}_{\mathbf{h}}$, of the samples of $h_W(t)$, denoted by the vector $\mathbf{h} = \left(h_W(0) \quad h_W\left(\frac{1}{W}\right) \quad \cdots \quad h_W\left(\frac{p-1}{W}\right)\right)^T$, where p is the length of the channel used for statistical analysis with $0 \le p \le \frac{1}{\Delta F} - \frac{1}{W}$. This allows us to estimate the number of DoF characterizing $h_W(t)$ [15]. A similar approach for estimating the (finite) unknown number of Gaussian signals using a finite set of noisy observations is described in [16] [17]. These techniques amount to determining the finite dimension of the signal subspace.

In order to estimate the true covariance matrix $\mathbf{K}_{\mathbf{h}}$, we use statistical averages based on observations from (20×50) positions. The sample covariance matrix is a maximum-likelihood estimate, under the assumption of a large number of independent channel observations [18] which arise from the different transmitting and receiving antenna positions as explained earlier. The separation between positions on the grids must be large enough to obtain sufficient variation of the wave's phase, $\Delta \phi^{-1}$, in order to extract all the DoF of the channel. On the other hand, the separation must be small enough to ensure that the distance between transmitter and the receiver (6m in our primary measurement scenario) remains virtually constant. If both constraints are satisfied we can assume that $\mathbf{K}_{\mathbf{h}}$ will depend solely on the slowly-varying parameters (distance, arrival angles, scatterers, geometrical settings,...) and thus will not vary significantly across the set of the total transmitter and receiver positions. The covariance matrix of measured channel samples, \mathbf{h} , is written as

$$\mathbf{K}_{\mathbf{h}} = E[\mathbf{h}\mathbf{h}^{H}] = E[\mathbf{g}\mathbf{g}^{H}] + \sigma_{n}^{2}\mathbf{I}$$
(3)

where g is a vector of samples of the noise-free channel process, and I is the identity matrix. The maximumlikelihood covariance matrix estimate computed from N statistically independent channel observation with length p and p < N is given by [18]

$$\mathbf{K}_{\mathbf{g}}^{N} = \mathbf{K}_{\mathbf{h}}^{N} = \frac{1}{N} \sum_{i=1}^{N} \mathbf{h}_{i} \mathbf{h}_{i}^{H}, \qquad (4)$$

The assumption of p < N holds in our case since we ensure that the length of sampled CIR is less than 1000 samples which represents the total number of channel observations for one scenario. For small d/λ , the as- ${}^{1}\Delta\phi = \frac{2\pi d}{\lambda}$ is the wave's phase variation between two positions, d is the corresponding distance and λ is the wavelength varying from 3 to 10cm for a UWB channel bandwidth ranging from 3 to 10GHz sumption of independent observations and thus that the channel samples are spatially decorrelated is justified in an indoor MIMO setting in [20]. In the context of our measurements, the multiple transmitter/receiver grid can equivalently be seen as a large(50×20) MIMO system.

The covariance matrix is Hermitian and positive definite. For this reason, a unitary matrix U_h exists such that the Karhunen-Loève (KL) expansion gives

$$\mathbf{K}_{\mathbf{h}}^{N} = \mathbf{U}_{\mathbf{h}} \Lambda_{\mathbf{h}} \mathbf{U}_{\mathbf{h}}^{H} = \sum_{i=1}^{N} \lambda_{i}(\mathbf{h}) \psi_{i}(\mathbf{h}) \psi_{i}^{H}(\mathbf{h}); \ \mathbf{U}_{\mathbf{h}}^{H} \mathbf{U}_{\mathbf{h}} = \mathbf{I}_{N},$$
(5)

where $\lambda_1(\mathbf{h}) \geq \lambda_2(\mathbf{h}) \geq ... \geq \lambda_N(\mathbf{h})$, $\psi_i(\mathbf{h})$ is the i^{th} column of $\mathbf{U}_{\mathbf{h}}$ and \mathbf{I}_N is the $N \times N$ identity matrix with N number of samples. $\lambda_i(\mathbf{h})$ and $\psi_i(\mathbf{h})$ are the i^{th} eigenvalues and eigenvectors of $\mathbf{K}_{\mathbf{h}}^N$, respectively. Decomposing (3) into principal and noise components yields

$$\mathbf{U}_{s,\mathbf{h}} = [\psi_1(\mathbf{h}), \psi_2(\mathbf{h}), ..., \psi_p(\mathbf{h})];$$
$$\lambda_1(\mathbf{h}) \ge \lambda_2(\mathbf{h}) \ge ... \ge \lambda_L(\mathbf{h});$$
$$\mathbf{U}_{n,\mathbf{h}} = [\psi_{L+1}(\mathbf{h}), \psi_{L+2}(\mathbf{h}), ..., \psi_N(\mathbf{h})]$$
$$\lambda_{L+1}(\mathbf{h}) \ge \lambda_{L+2}(\mathbf{h}) \ge ... \ge \lambda_N(\mathbf{h}).$$

where $\mathbf{U}_{s,\mathbf{h}} \perp \mathbf{U}_{n,\mathbf{h}}$. $\mathbf{U}_{s,\mathbf{h}}$ defines the subspace containing both signal and noise components, whereas $\mathbf{U}_{n,\mathbf{h}}$ defines the noise-only subspace. L is the number of significant eigenvalues which also represents the channel degrees of freedom [16], in the sense that any set of observations can be characterized by a set of approximately L independent random variables which excite L modes (their corresponding eigenvectors). In our case, the eigenvectors correspond to the samples of the eigenfunctions characterizing the propagation environment. They convey information regarding the statistical correlation between paths arriving at different time instants, since if in a particular eigenvector, ψ_i , we find several significantly-spaced high-energy samples (paths), the same random variable with variance $\lambda_i(\mathbf{h})$ excites these samples and thus they exhibit statistical correlation.

B. Characterization of the Total Received Energy

Following the eigen-decomposition, we can define the normalized total received energy in an observation as $X = \mathbf{h}^H \mathbf{h} / \overline{\mathbf{h}^H \mathbf{h}}$ where the average channel energy is denoted by $\overline{\mathbf{h}^H \mathbf{h}}$. We must distinguish two important cases, namely the presence or lack of a line-of-sight path in the received wavefront. In general we may express the filtered impulse response of the channel, assuming a discrete set of scatterers, as

$$h_W(t) = A_0 e^{j\phi_0} \operatorname{sinc}(W(t - t_0)) + \sum_{i>0} A_i e^{j\phi_i} \operatorname{sinc}(W(t - t_i))$$
(6)

where A_0 represents the strength of the line-of-sight component, and $A_{i>0}$ the random amplitudes of scattered components, and t_i are the delays of each component. For a particular channel sample we obtain

$$h(n) = A_0 e^{j\phi_0} \operatorname{sinc}(W(n/W - t_0)) + \sum_{i>0} A_i e^{j\phi_i} \operatorname{sinc}(W(n/W - t_i))$$
(7)

In the characterization of the energy of the channel, $\mathbf{h}^H \mathbf{h}$, the phase offset of the direct path can clearly be neglected, and thus we can assume that ϕ_0 is zero, even if it is truly a random quantity. Moreover, for a dense multipath environemtn, each h(n) can be approximated by a Gaussian random variable with mean $A_0 \operatorname{sinc}(W(n/W - t_0))$ as is classically assumed [21]. We can thus express the channel-energy moment generating function as

$$G_X(s) = E[e^{sx}] = \int_{-\infty}^{+\infty} e^{sx} f_X(x) dx = L(f_X(x))|_{s=-s}$$
(8)

L is the Laplace transform operator and $f_X(x)$ is the probability density function (pdf) of variable *X*. We can see from last equation that $G_X(s)$ and $f_X(x)$ are Laplace transform pairs with s = -s. $G_X(s)$ can also be expressed as follows using the covariance matrix \mathbf{K}_h

$$G_X(s) = \frac{e^{\mathbf{m}^H (\mathbf{I} - s\mathbf{K}_h)^{-1} \mathbf{m}}}{\det(I - s\mathbf{K}_h)}$$
(9)

where the vector **m** has components $m(n) = A_0 \operatorname{sinc}(W(n/W - t_0))$ [22]. We will assume that \mathbf{K}_h is characterized by L significant eigenvalues, $\lambda_i(\mathbf{h}) : i = 1..L$, which are positive. Furthermore, we constrain our treatment of the channel energy to the non-line-of-sight case since no attempt was made to estimate the line-of-sight component strength, A_0 , from our collected data. It is easily shown using the initial value theorem [23] that the probability density function of the normalized channel energy can be approximated around the origin as

$$f_X(x) \approx \frac{x^{(L-1)}}{(L-1)!} \frac{1}{\det(K_{\mathbf{h}})} = \frac{x^{(L-1)}}{(L-1)!} \frac{1}{\prod_i \lambda_i(\mathbf{h})}, \text{for} 0 \le x <<\min(\lambda_i(\mathbf{h}))$$
(10)

This indicates that around the origin X behaves as an Erlang- L^2 variable with parameter $2\sigma^2 = \sqrt[L]{\prod_i \lambda_i(\mathbf{h})}$. The approximate cumulative distribution function can be expressed in terms of the incomplete Gamma function, or in terms of the Marcum-Q function, $Q_L(0, x)$, as

$$F_X(x) = \gamma \left(\frac{x}{2\sigma^2}, L\right) = 1 - Q_L \left(0, \frac{\sqrt{x}}{\sigma}\right)$$
$$\approx \frac{1}{\Gamma(L+1)} \left(\frac{x}{2\sigma^2}\right)^L, \quad 0 \le x \le \prod_i \lambda_i(\mathbf{h})$$
$$= \frac{x^L}{\Gamma(L+1) \prod_i \lambda_i(\mathbf{h})}.$$
(11)

The above approximations can be found in [24]. Since the approximation is valid for small X, we are characterizing the region of the CDF in its lefthand tail. From the above expression we see that we can use the slope of log(cdf) to gain insight regarding the the number of significant DoF for a particular average channel strength. This can also be seen equivalently as determining the inherent diversity-order of the channel [8].

IV. EMPIRICAL MEASUREMENT RESULTS

A. Empirical Distribution Of Eigenvalues For Indoor UWB Channels

The empirical results presented in this section are obtained from the LOS/NLOS 6m scenarios described earlier. The sampled CIR is calculated from 6003 samples of the frequency response using IFFT, limited to 1000 samples for the computation of the sample covariance matrix, $\mathbf{K}_{\mathbf{h}}^{N}$. In figures 6 and 7 we show the sample mean power delay profile (diag($\mathbf{K}_{\mathbf{h}}^{N}$)) for the LOS and NLOS cases. We see that the LOS



Fig. 6. Power Intensity Profile in LOS situation



Fig. 7. Power Intensity Profile in NLOS situation

is characterized by two dominant clusters, one of which is around the LOS path, which is clearly visible. Secondary scatterers comprise significant energy nontheless. In the NLOS case there is dense scattering for short delays and the same secondary cluster as the LOS scenario, resulting surely from multiple reflections in the room.

In figures 8 and 9 we plot, for the LOS and NLOS measurements, the fraction of the captured energy for M considered eigenvalues defined by $E_M = \frac{\sum_{i=1}^{M} \lambda_i(\mathbf{h})}{\sum_{i=1}^{N} \lambda_i(\mathbf{h})}$ where N is the total number of eigenvalues. We ²The Erlang-L probability density function of variable X is given by $f_X(x) = \frac{x^{L-1}}{(2\sigma^2)^L \Gamma(L)} e^{-\frac{x}{2\sigma^2}}$, x > 0 and L an integer

remark in both plots, for the narrowband cases, that the majority of the channel energy (more than 90%) is confined in small number of significant eigenvalues whereas in the wide bandwidth case, the channel energy is spread over a large number of eigenvalues (i.e. DoF). The results are obtained for a sampling rate, in the frequency domain, of 1 sample per megahertz. This sampling resolution was verified to be sufficient to capture all channel degrees of freedom and at the same time minimize the effect of time-domain aliasing as mentioned previously. Indeed, figure 10 shows that increasing the sampling rate from 1 sample per 2 megahertz to 1 sample per megahertz does not modify channel eigenvalue distribution.



Fig. 8. Fraction of the captured energy versus number of significant eigenvalues in LOS situation, resolution 1 sample per megahertz

Figure 11, plotted for 98% captured energy, shows that the number of significant eigenvalues increases with the channel bandwidth, although not linearly. We see that in both cases, the increase is linear until approximately 200MHz, where a saturation effect begins to occur. This means that until this critical bandwidth the signal bandwidth does not provide sufficient resolution to resolve all eigenvalues, or equivalently the complete number of multipath components. Beyond this point, however, the channel is degenerate in the sense that all paths can be resolved. We note that the number of significant eigenvalues in the saturation region is markedly higher in the LOS case. This can be explained by the fact that the object placed between the transmitter and receiver also shadows a significant portion of non-line-of-sight components and the richness of the channel is therefore reduced.



Fig. 9. Fraction of captured energy versus number of significant eigenvalues in NLOS situation



Fig. 10. Fraction of captured energy versus number of significant eigenvalues in LOS situation, resolution 1 sample per 2 megahertz

The analysis of the empirical cdf of the normalized energy for the LOS case in figure 12 shows that the cdf for a 6 GHz bandwidth exhibits a very sharp slope around the mean signal strength (it is compared to a diversity 1 Rayleigh channel to highlight the extreme behavior of these channels). We also compare the empirical cdf curve to that one obtained using the approximation in 11. We can see that the number of significant eigenvalues is directly related to the steepness of cdf curve. The high number of degrees of freedom shows that UWB channels can be considered as deterministic (non-fading) in practice, provided receivers exploiting the full channel energy are employed.



Fig. 11. Evolution number of eigenvalue for LOS and NLOS cases UWB measurement

B. Eigenfunction Analysis In LOS And NLOS Situations

Figures 13, 14 and 15 show some sampled eigenfunctions in LOS settings corresponding to the most significant eigenvalues. These figures show that the channel exhibits a clustered behavior, as can be seen when we plot the CIR shown in figures 13 and 14. It is remarkable that from our analysis we see that paths from different clusters have comparable strengths in the same eigenfunction. As a result, these clusters are strongly statistically dependent as was described in the previous section.



Fig. 12. The empirical Cumulative Distribution Function



Fig. 13. Eigenfunction corresponding to $\lambda_1 = 0.4445$ in LOS situation



Fig. 14. Eigenfunction corresponding to $\lambda_5 = 0.0314$ in the LOS situation

V. CONCLUSION

In this work we present results from UWB channel measurement campaigns conducted at Institut Eurcom laboratories. These measurements confirm previous results performed in other laboratories (e.g. [3]) regarding the clustered behavior of the UWB channel as well as the multipath richness exposed by the extreme bandwidth. The new results published in this work are twofold; first we show the evolution of the number of channel eigenvalues as a function of the system bandwidth for both LOS and NLOS scenarios, where we see that the number of eigenvalues tends to saturate for the extreme bandwidth of UWB systems. This seems to



Fig. 15. Eigenfunction corresponding to $\lambda_{10} = 0.0133$ in LOS situation

suggest that all significant multipath components can be resolved. Secondly we show that there is a strong statistical dependence between paths coming from different clusters, which is assumed not to be the case in most propagation models.

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