Abstract
In this work we describe indoor multi-ray propagation model for WPAN applications considering reflection and transmission through multilayered structures which are a consequence of the inhomogeneity of the building materials composing the environment. Our model enables the analysis of a communication channel for both the narrow and wide band cases between adjacent rooms and inside single room. We use our model to predict the behavior of UWB signals while multipath propagating within dynamic surroundings.

Key-Words: - Wireless Communications, RF, Indoor Propagation Models, Multipath

1 Introduction
During recent years, wireless communications have become an issue of extensive research. The Wireless Local Area Network (WLAN) is used to communicate between nodes up to few tens of mega bits per second. The conventional designated frequencies for WLAN systems are within the Industrial Scientific and Medical (ISM) 2.4-2.4835 band and the Unlicensed National Information Infrastructure (UNII) 5.15-5.35GHz and 5.725-5.825GHz bands.

The Wireless Personal Area Network (WPAN) system is designed to provide short-range, high-speed multi-media data services to terminals located in rooms or office spaces. Two continuous blocks of frequencies were allocated for indoor short-range networks, the band of 3.1-10.6 GHz and 59-64 GHz one [1]. The unique capabilities of the ultra wideband (UWB) technique may promise reliable links between nodes located in different rooms. It was suggested to integrate the 5GHz and the 60 GHz bands to enable operation in indoor and outdoor environments [2].

One of the principal challenges in realizing high data rate, wireless communication links in indoor or indoor-outdoor scenarios are the phenomena arising during the propagation of electromagnetic waves inside a construction [3]-[4]. Development of a channel model for continuous RF frequencies in indoor environments is required for the analysis of indoor communications in an ultra wide band wireless network which includes a single transmitting and a single receiving antenna. The building blocks of such a model include single frequency propagation coefficients which are later integrated to
obtain the wideband response. The model developed here is in the frequency domain and thus allows analyzing dispersive effects in transmission and reflection of ultra short pulses in UWB communications from building materials which the room is made of in accordance with their complex dielectric coefficients. Transmission of UWB signals is dominated by the frequency dependent dielectric properties of the materials and structures composing the propagation medium. It is well recognized that walls introduce attenuation. However dispersive effects are also expected due to their multi-slab structure, creating a dielectric resonator.

In this paper we present a space-frequency model for both narrowband and wideband links, considering the complex dielectric properties of the materials of the building walls. Based on the approach, a numerical model will be developed enabling examination of the absorptive and dispersive effects on propagation of radio waves through the walls at the super- and extremely high frequencies. For this purpose a library of material characteristics of various materials (concrete, reinforced concrete, plaster, wood, blocks, glass, stone and more) in the standard frequency domain for wireless networks was previously assembled. In this work we will describe a model taking into account transmission through multiple layers which are an outcome of the inhomogeneity of the building materials our indoor environment is made of. Examples include hardened concrete wall and wooden door. Our model enables the analysis of a communication channel for both the narrow and wide band cases between adjacent rooms and inside single room.

In a typical indoor ambience, a radio signal transmitted from a fixed source will encounter multiple objects that produce reflected, diffracted or scattered copies of the signal called multipath components. They can be attenuated in power, delayed in time and shifted in phase and/or frequency from the LOS signal path at the receiver. The multipath and the transmitted signal are summed together at the receiver, which often generates distortion in the received signal relative to the origin. At the discussed frequencies, the deterministic propagation modeling can be realized using ray tracing technique based on geometrical optics (GO). In ray tracing we assume a finite number of reflectors with known location and dielectric properties whether the transmitter and receiver are mobile or fixed. If one of the stations is moving, time variations of the multipath characteristics are deterministic when the parameters of the reflectors are specified exactly over time. Ray tracing method approximates the propagation of electromagnetic waves by representing wavefronts as simple particles. The error of ray tracing approximation is smallest when the receiver is many wavelengths from the nearest scatterer and all the scatterers are large relative to a wavelength and fairly smooth.

The most general ray tracing model includes all attenuated, diffracted and scattered multipath components. This model uses all of the geometrical and dielectric properties of the objects surrounding the transmitter and receiver. Usually, the LOS and reflected paths provide the dominant components of the received signal, since diffraction and scattering losses are high. If the ray tracer is fed with the location of walls and doors only, the measured power in a very specific point in the room may vary from the predicted value,
but the average power measured in that room will sufficiently coincide with the value produced by the ray-tracer \([\ldots]\). In our model we refer to an indoor space free of obstacles (furniture), thus the behavior of the propagation channel can be described by reflections only while diffraction and scattering are neglected. In order to simplify the ray tracing algorithm and to improve computation time, first order specular reflections alone will be regarded.

2 Presentation of Electromagnetic Waves in the Frequency Domain

The electromagnetic field in the time domain is described by the space-time electric \(E(r,t)\) and magnetic \(H(r,t)\) signal vectors. \(r\) stands for the \((x,y,z)\) coordinates, where \((x,y)\) are the transverse coordinates and \(z\) is the axis of propagation. The Fourier transform of the electric field is defined by:

\[
E(r,f) = \int_{-\infty}^{\infty} E(r,t) \cdot e^{-j2\pi ft} dt
\]

where \(f\) denotes the frequency. Similar expression is defined for the Fourier transform \(H(r,f)\) of the magnetic field. Since the electromagnetic signal is real (i.e. \(E^*(r,t) = E(r,t)\)), its Fourier transform satisfies \(E^*(r,f) = E(r,-f)\). A complex representation of the signal is given by the expression [5]:

\[
\tilde{E}(r,t) = E(r,t) + j\hat{E}(r,t)
\]

where

\[
\hat{E}(r,t) = \frac{1}{\pi} * E(r,t) = \int_{-\infty}^{\infty} \frac{E(r,t')}{\pi (t-t')} dt'
\]

is the Hilbert transform of \(E(r,t)\). Fourier transformation of the complex representation (2) results in a ‘phasor-like’ function \(\tilde{E}(r,f)\) defined in the positive frequency domain and related to the Fourier transform by:

\[
\tilde{E}(r,f) = \begin{cases} 
2E(r,f) & f > 0 \\
0 & f < 0
\end{cases}
\]

The Fourier transform can be decomposed in terms of the phasor like functions according to:

\[
E(r,f) = \frac{1}{2} \tilde{E}(r,f) + \frac{1}{2} \tilde{E}'(r,-f)
\]

and the inverse Fourier transform is then:

\[
E(r,t) = \int_{-\infty}^{\infty} E(r,f) \cdot e^{j2\pi ft} df = \text{Re} \int_{0}^{\infty} \tilde{E}(r,f) \cdot e^{j2\pi ft} df
\]
3 Electromagnetic Waves in Dielectric Media

Propagation of electromagnetic waves in a medium can be viewed as a transformation through a linear system. A plane wave propagating in a (homogeneous) medium is given in the frequency domain by:

$$\tilde{E}_{out}(f) = \tilde{E}_{in}(f) \cdot e^{-jkd}$$  \hspace{1cm} (7)

Here, $k(f) = 2\pi f \sqrt{\varepsilon_0 \mu}$ is a frequency dependent propagation factor, where $\varepsilon$ and $\mu$ are the permittivity and the permeability of the material composing the medium, respectively. In a dielectric medium the permeability is equal to that of the vacuum $\mu = \mu_0$ and the permittivity is given by $\varepsilon(f) = \varepsilon_r(f) \cdot \varepsilon_0$. The polarization in dielectric materials with absorption is accounted via a complex relative permittivity:

$$\varepsilon_r(f) = \varepsilon'(f) - je''(f)$$  \hspace{1cm} (8)

The relative permittivity is a frequency dependent quantity with a real part $\varepsilon'(f) \geq 1$ and an imaginary part $\varepsilon''(f) = \frac{\sigma(f)}{2\pi \varepsilon_0}$ where $\sigma$ is conductivity. The ratio between the imaginary and real parts is known as the ‘loss tangent’ $\tan\delta(f) = \frac{\varepsilon''(f)}{\varepsilon'(f)}$. The propagation factor can be presented by:

$$k(f) = \frac{2\pi f}{c} \sqrt{\varepsilon_r(f)} = \frac{2\pi f}{\sqrt{\varepsilon'(f)}} \cdot \sqrt{1 - j \cdot \tan\delta(f)}$$  \hspace{1cm} (9)

where $c = \frac{1}{\sqrt{\mu_0 \varepsilon_0}}$ is the speed of light. The propagating wave can be written as:

$$E_{out}(f) = E_{in}(f) \cdot e^{[-\alpha(f) + j\beta(f)]d}$$  \hspace{1cm} (10)

where $\alpha(f) = -\text{Im}[k(f)]$ is the attenuation coefficient and $\beta(f) = \text{Re}[k(f)]$ is the wave number of the propagating wave. The real and imaginary parts are determined from:

$$\alpha^2(f) - \beta^2(f) = \left(\frac{2\pi f}{c}\right)^2 \varepsilon'(f)$$

$$2\alpha(f)\beta(f) = \left(\frac{2\pi f}{c}\right)^2 \varepsilon'(f) \tan\delta(f)$$  \hspace{1cm} (11)

Solution of the set (11) results in [6]:

$$\alpha(f) = \frac{2\pi f}{c} \sqrt{\frac{\varepsilon'(f)}{2}} \left| \frac{\varepsilon'(f)}{2} \sqrt{1 + \tan^2 \delta(f)} - 1 \right|$$

$$\beta(f) = \frac{2\pi f}{c} \sqrt{\frac{\varepsilon'(f)}{2}} \left| \frac{\varepsilon'(f)}{2} \sqrt{1 + \tan^2 \delta(f)} + 1 \right|$$  \hspace{1cm} (12)

The dielectric properties of several building materials are summarized in Table 1 for frequency of 5GHz.
Consider a dielectric slab of thickness \( d \), as illustrated in figure 1. The problem of transmission and reflection of an incident wave can be treated as done for a Fabry-Perot etalon [7]-[8].

When the wave is incident at an angle \( \theta_0 \) relative to the normal of the interface between dissimilar media, the field reflection coefficient is given by [9]:

\[
 r = \rho_0 \cdot \frac{1 - e^{-j2k_z(f)d}}{1 - \rho_0^2 e^{-j2k_z(f)d}}
\]

(13)

where \( k_z(f) = k(f) \cos(\theta) = (\alpha + j\beta) \cos(\theta) \) and refractive angle \( \theta \) can be found by generalized Snell’s law:

\[
 \tan(\theta) = \frac{\sin(\theta_0)}{\Re\{\sqrt{\varepsilon_r - \sin^2(\theta_0)}\}}
\]

(14)

The coefficient \( \rho_0 \) is resulted from the Fresnel’s equations [9]-[10]. When the electric field component of the wave is perpendicular to the plane of incidence (TE-wave):

\[
 \rho_{0-TE} = \frac{\eta \cos(\theta_0) - \eta_0 \cos(\theta)}{\eta \cos(\theta_0) + \eta_0 \cos(\theta)}
\]

(15)

In which \( \eta_0 = \sqrt{\mu_0 / \varepsilon_0} \) and \( \eta \) are an intrinsic impedances of the free space and the medium respectively. The latter is given by:

\[
 \eta(f) = \sqrt{\frac{\mu_r \cdot \mu_0}{\varepsilon_r \cdot \varepsilon_0}} = \frac{\eta_0}{\sqrt{\varepsilon_r(f)}}
\]

(16)

4 Wave Propagation in a Dielectric Layer

TABLE 1
Dielectric Properties of Building Materials
Measured at 5 GHz.

<table>
<thead>
<tr>
<th>MATERIALS</th>
<th>5 GHz</th>
<th>5 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>( \varepsilon' )</td>
<td>( \sigma ) [S/m]</td>
</tr>
<tr>
<td>Brick (made of chalk, with holes)</td>
<td>4.12</td>
<td>4.45E-02</td>
</tr>
<tr>
<td>Brick (without holes)</td>
<td>3.3</td>
<td>2.78E-03</td>
</tr>
<tr>
<td>Brick wall</td>
<td>3.56</td>
<td>9.46E-02</td>
</tr>
<tr>
<td>Concrete (one year)</td>
<td>5.5</td>
<td>5.01E-02</td>
</tr>
<tr>
<td>Concrete (40 years)</td>
<td>4.6</td>
<td>6.68E-02</td>
</tr>
<tr>
<td>Wood (a)</td>
<td>2.05</td>
<td>8.23E-02</td>
</tr>
<tr>
<td>Wood (b)</td>
<td>1.65</td>
<td>6.54E-02</td>
</tr>
<tr>
<td>Plasterboard</td>
<td>2.02</td>
<td>1.48E-02</td>
</tr>
<tr>
<td>Chipboard</td>
<td>2.88</td>
<td>1.36E-01</td>
</tr>
<tr>
<td>Glass</td>
<td>5.98</td>
<td>2.99E-01</td>
</tr>
</tbody>
</table>

TABLE 1
Dielectric Properties of Building Materials
Measured at 5 GHz.
For the case of parallel polarization, where the incident electric field vector is in the plane of incidence (TM-wave):

\[ \rho_{0-TM} = \frac{\eta \cos(\theta) - \eta_0 \cos(\theta_0)}{\eta \cos(\theta) + \eta_0 \cos(\theta_0)} \]  

(17)

The field transmission of the electromagnetic wave through the slab is described by the expression:

\[ t = \frac{(1 - \rho_0^2) e^{-jk_d(f)d}}{1 - \rho_0^2 e^{-jk_d(f)d}} \]  

(18)

Transmission, reflection and loss \((L = 1 - |r|^2 - |t|^2)\) of a single layer are depicted in figure 2.

The free spectral range (FSR) between transmission peaks is given for low loss materials (small imaginary dielectric constant) by:

\[ FSR \approx \frac{c}{2d \cdot \sqrt{\varepsilon'(f) \cdot \cos(\theta)}} \left[ 1 + \frac{f}{2\varepsilon'(f)} \cdot \frac{d\varepsilon'(f)}{df} \right]^{-1} \]  

(19)
The full width at half maximum (FWHM) of the transmission peaks is given by:

\[
FWHM = \frac{FSR}{Finesse}
\]

In which the Finesse is defined by:

\[
Finesse = \frac{\pi \sqrt{|\rho_0| e^{-\alpha \cos(\theta) d}}}{1 - |\rho_0| e^{-\alpha \cos(\theta) d}}
\]

A realistic calculation for a concrete layer transmission and reflection coefficients is given in figures 3 and 4. The dielectric constants are taken from Table 2 assuming linear frequency dependence of the material properties.

<table>
<thead>
<tr>
<th>TABLE 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>DIELECTRIC PROPERTIES OF HARDENED CONCRETE MEASURED THROUGHOUT WIDE SPECTRUM.</td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td>( 1 )</td>
</tr>
<tr>
<td>( \vdots )</td>
</tr>
<tr>
<td>95.9</td>
</tr>
</tbody>
</table>

Fig. 3. Transmission coefficient of a hardened concrete dielectric slab \( d=5 \) cm at normal incidence.
Fig. 4. Reflection coefficient of a hardened concrete dielectric slab d=5 cm at normal incidence.

TABLE 3
DIELECTRIC PROPERTIES OF WOODEN DOOR MEASURED THROUGHOUT WIDE SPECTRUM.

<table>
<thead>
<tr>
<th>( f ) [GHz]</th>
<th>( \varepsilon' )</th>
<th>( \sigma ) [S/m]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.01</td>
<td>2.08</td>
<td>1.35E-02</td>
</tr>
<tr>
<td>( \cdot )</td>
<td>( \cdot )</td>
<td>( \cdot )</td>
</tr>
<tr>
<td>14.31</td>
<td>1.98</td>
<td>1.02E-01</td>
</tr>
</tbody>
</table>

Visibly fading fluctuations of the coefficients magnitude arise from increasing material conductivity as function of frequency. The same calculation for transmission and reflection coefficients of a wooden door layer embedded between two concrete media is depicted in figures 5 and 6. The dielectric constants are taken from Table 3 assuming linear frequency dependence of the material properties.

5 WAVE PROPAGATION IN MULTIPLE LAYERS

Consider the case in which several layers are involved, the situation is depicted in figure 7. In this case in addition to the internal reflections which occur in each single layer and are taken into account in equations (13) and (18) there are multiple reflections from one layer to another.
Fig. 5. Transmission coefficient of a wooden door dielectric slab $d=4.45$ cm at normal incidence.

Fig. 6. Reflection coefficient of a wooden door dielectric slab $d=4.45$ cm at normal incidence.

In this case the total transmission and reflection coefficients $T$ and $R$ can be retrieved by using ABCD method []. Each layer $n$ is represented by a transfer matrix:

$$
\begin{bmatrix}
A_n & B_n \\
C_n & D_n
\end{bmatrix} = \begin{bmatrix}
\cos(k_{zn}d_n) & jZ_n \sin(k_{zn}d_n) \\
jZ_n^{-1} \sin(k_{zn}d_n) & \cos(k_{zn}d_n)
\end{bmatrix}
$$ (22)

Multiplying the matrices of single layers we arrive at total transfer matrix for the entire structure:

$$
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
A_1 & B_1 \\
C_1 & D_1
\end{bmatrix} \begin{bmatrix}
A_2 & B_2 \\
C_2 & D_2
\end{bmatrix} \cdots \begin{bmatrix}
A_N & B_N \\
C_N & D_N
\end{bmatrix}
$$ (23)
Thus general transmission and reflection coefficients are expressed in terms of the total matrix elements:

\[
R = \frac{A + \frac{B}{Z_{\text{out}}} - CZ_{\text{in}} - D \frac{Z_{\text{in}}}{Z_{\text{out}}}}{A + \frac{B}{Z_{\text{out}}} + CZ_{\text{in}} - D \frac{Z_{\text{in}}}{Z_{\text{out}}}} \tag{24}
\]

\[
T = \frac{2}{A + \frac{B}{Z_{\text{out}}} + CZ_{\text{in}} + D \frac{Z_{\text{in}}}{Z_{\text{out}}}} \tag{25}
\]

The impedances of the all structure’s layers and input and output medium are calculated as follows \( Z = \eta / \cos(\theta) \) and \( Z = \eta \cdot \cos(\theta) \) for TE and TM polarization respectively. In practical indoor ambiance input and output medium are identical (free space), what means \( Z_{\text{in}} = Z_{\text{out}} = Z_0 \) and \( \theta_{\text{in}} = \theta_{\text{out}} = \theta_0 \). Therefore, substituting the former equality in (24) and (25) leads to the equations with simpler form:

\[
R = \frac{A + \frac{B}{Z_0} - CZ_0 - D}{A + \frac{B}{Z_0} + CZ_0 + D} \tag{26}
\]

\[
T = \frac{2}{A + \frac{B}{Z_0} + CZ_0 + D} \tag{27}
\]

In particular case we consider a structure which is modeled by a 5 cm layer of hardened concrete followed by a 4.45 cm layer of wooden door which is followed by a 5 cm layer of hardened concrete again (Fig 8).
The power transmission and reflection coefficients are illustrated for a triple layer. Furthermore, we assume that a plane electromagnetic wave approaches the layers at normal incidence. One can appreciate the complexity of the transmission and reflection coefficients by studying figures 9 and 10.

**Fig. 9.** Power transmission coefficient of a triple slab. Wood layer d=4.45 cm, Concrete layers d=5 cm each, normal incidence.

**Fig. 10.** Power reflection coefficient of a triple slab. Wood layer d=4.45 cm, Concrete layers d=5 cm each, normal incidence.

### 6 UWB Propagation in Multiple Layers

The challenge of evaluating UWB transmission and reflection coefficients is tantamount to evaluating the transmission and reflection coefficients for each frequency component that the UWB pulse is made of and then integrating the results, this can be done as
follows. Consider the wireless link depicted in figure 11. It was shown in [11] that given the channel modulated input signal to be:

$$E_{in}(t) = \text{Re}\{A_{in}(t) \cdot e^{j2\pi f_o t}\} \quad (28)$$

And the channel modulated output signal to be:

$$E_{out}(t) = \text{Re}\{A_{out}(t) \cdot e^{j2\pi f_o t}\} \quad (29)$$

In which \( f_0 \) is the modulation frequency. The unmodulated output and input signals are related by:

$$A_{out}(t) = I_{out}(t) - jQ_{out}(t) = \int_{-\infty}^{+\infty} A_m(f) \cdot H(f + f_0) \cdot e^{j2\pi f t} \quad (30)$$

![Wireless link schematics](image)

Fig. 11. Wireless link schematics.

The procedure for calculating the output pulse is given in figure 12:
For reflected pulses the transfer function \( H(f) \) is given by the single frequency reflection coefficient while for transmitted pulses the single frequency transmission coefficient serves the same purpose. Consider an input pulse of the form:

\[
A_{in}(t) = e^{-\frac{t^2}{2\sigma_n^2}} \quad (31)
\]

Assume that the pulse has a (standard deviation) width \( \sigma_n = 0.1ns \), the pulse is depicted in figure 13. The Fourier transform of this pulse will also be a Gaussian given by:

\[
A_{in}(f) = \sqrt{2\pi} \cdot \sigma_n \cdot e^{-\frac{1}{2}(2\pi\sigma_n f)^2} \quad (32)
\]

with a (standard deviation) width \( \sigma_f = \frac{1}{2\pi\sigma_n} = 1.6GHz \), the normalized Fourier transform is described in figure 14.

Fig. 13. A Gaussian pulse of standard deviation width of 0.1 ns.

Fig. 14. The normalized Fourier transform of the Gaussian pulse (fig. 13).
In what follows we consider a transmission and reflection of the pulse described above centered at $f_0 = 5GHz$ through a triple layer dielectric slab (fig.8) at normal incidence. The results calculated using equations (26, 27, 30, 32) are given in figures 14 and 15.

![Transmission of a Gaussian pulse through triple slab](image1)

**Fig. 15.** Transmission of a Gaussian pulse through triple slab (fig. 8) at normal incidence.

![Reflection of a Gaussian pulse from triple slab](image2)

**Fig. 16.** Reflection of a Gaussian pulse from triple slab (fig. 8) at normal incidence.

It can be clearly seen the triple layer slab introduces a delay and that the multiple reflections inside that slab create additional pulses called an echoes with decreasing amplitude (due to absorption) of which the first ones are discernible. The amplitudes of both initial pulses ($T$ and $R$) nearly comply with the values of the coefficients at central frequency due to power delay in time domain.

Notice that the first reflection has a zero time delay since the signal is assumed to be injected at zero distance from the concrete slab. Moreover, its amplitude meets the value of Fresnel’s reflection coefficient of the first layer (hardened concrete) and it is thickness independent.

Multiple reflections inside the layers result in a complex structure of reflected and transmitted pulses which in some cases overlap each other generating inter pulse...
interference (IPI). Third transmitted pulse consists of two constructively interfered pulses traveling via distinct paths with same length (delay). This may have a profound effect on UWB communications in which reflections from early signals may appear as "noise" for later signals giving rise to inter symbol interference (ISI). Finally, Gaussian waveforms preserve their envelope and exhibit phase distortion during propagation through a dispersive slab.

The delays can be easily calculated by dividing each layer thickness $d_n$ by the velocity of wave $v_n$ in corresponding layer, approximately:

$$\Delta t_n = \frac{d_n}{v_n} \approx \frac{d_n \sqrt{\varepsilon_r}}{c}$$  \hspace{1cm} (33)$$

and then summing the results for relevant pulse. In the above $c$ is the speed of light in vacuum and $\varepsilon_r$ is the real relative permittivity of the material layer at central frequency.

The time delays of the first and second transmitted signals are:

$$\Delta t_{\text{first transmission}} \approx 2 \cdot \Delta t_{\text{concrete}} + \Delta t_{\text{wood}} \approx 1.1\text{ns}$$  \hspace{1cm} (34)$$

$$\Delta t_{\text{second transmission}} \approx 2 \cdot \Delta t_{\text{concrete}} + 3 \cdot \Delta t_{\text{wood}} \approx 1.5\text{ns}$$  \hspace{1cm} (35)$$

The time delays of the second and third reflected signals are:

$$\Delta t_{\text{second reflection}} \approx 2 \cdot \Delta t_{\text{concrete}} \approx 0.88\text{ns}$$  \hspace{1cm} (36)$$

$$\Delta t_{\text{third reflection}} \approx 2 \cdot \Delta t_{\text{concrete}} + 2 \cdot \Delta t_{\text{wood}} \approx 1.3\text{ns}$$  \hspace{1cm} (37)$$

7  INDOOR MULTI-RAY MODEL

The multi-ray model is a general case of the well known two-ray model for more than two reflected components. The indoor environment is presented by a spatial coordinates $(x,y,z)$ which correspond to length, width and height respectively as drawn in figure 17. In the given geometry multipath propagation takes place including line of sight (LOS) and six reflections, seven rays in total.
All four walls, floor (wall 3) and ceiling (wall 6) in the depicted scenario are made of homogeneous hardened concrete with 10 cm thickness, when walls 4 and 5 are placed in front of walls 1 and 2 respectively. The total field is a sum of all the fields arrived at the receiver site via various paths and given by:

\[
\begin{align*}
\hat{E}_{\text{Total}} &= \frac{1}{d} \cdot e^{-\frac{2\pi}{c}d} + \\
&+ \frac{1}{d_{11} + d_{12}} \cdot e^{-\frac{2\pi}{c}(d_{11} + d_{12})} \cdot \rho_{\text{wall1}} \\
&+ \frac{1}{d_{21} + d_{22}} \cdot e^{-\frac{2\pi}{c}(d_{21} + d_{22})} \cdot \rho_{\text{wall2}} \\
&+ \frac{1}{d_{31} + d_{32}} \cdot e^{-\frac{2\pi}{c}(d_{31} + d_{32})} \cdot \rho_{\text{floor}} \\
&+ \frac{1}{d_{41} + d_{42}} \cdot e^{-\frac{2\pi}{c}(d_{41} + d_{42})} \cdot \rho_{\text{wall4}} \\
&+ \frac{1}{d_{51} + d_{52}} \cdot e^{-\frac{2\pi}{c}(d_{51} + d_{52})} \cdot \rho_{\text{wall5}} \\
&+ \frac{1}{d_{61} + d_{62}} \cdot e^{-\frac{2\pi}{c}(d_{61} + d_{62})} \cdot \rho_{\text{ceiling}} 
\end{align*}
\]

The received power is calculated according to a modified Friis’s transmission equation which takes into account additional paths:
\[ P_r = G_r \cdot \frac{c}{4\pi fd} \cdot \left( \frac{\vec{E}_{\text{Total}}}{\vec{E}_{\text{LOS}}} \right)^2 \cdot P_t \cdot G_t \]  \hspace{1cm} (39)

where \( \vec{E}_{\text{LOS}} \) is the first item in (38).

In order to obtain comprehensive information about contribution of the surroundings to the received power, isotropic antennas \( (G_t = G_r = 0 \text{dBi}) \) with transmit power of \( P_t = 0 \text{dBm} \) are utilized.

Setting proper dimensions of the geometry allows building any scenario. Reckon a room having (5,4,3)m in size with transmitter and receiver in it located at (1,1,2)m and (4,3,1)m respectively. Frequency responses of the propagation channel in the in-room scenario for both polarizations are depicted in figures 18 and 19 comparatively to transfer function of LOS (single direct ray). Polarizations are emanated from oblique incidences of paths with reflections at the walls. The received power shown is normalized to zero by central frequency \( f_0 = 5 \text{GHz} \) occupying 5 GHz spectrum.

![Fig. 18. Received TE-power (7 rays) compared to LOS power in in-room scenario at \( f_0 = 5 \text{GHz} \).](image-url)
LOS power decays inversely proportional to the frequency squared. Power contributions by reflections diminish path loss up to 9.4 dB for narrow band. Assume a Gaussian pulse with standard deviation of \( \sigma_n = 0.2\text{ns} \) as transmitted UWB signal. The result of its propagation through the in-room channel in time domain is illustrated in figures 20 and 21.
The first pulse is received by LOS path whereas all the rest arose from a reverberation. Owing to symmetry between locations of the transmitter and receiver, each of three major reflections consists of two constructively interfered pulses reflected from two opposite walls in following order (3,6), (2,5) and (1,4). The last discernible pulse is actually a consequence of multiple reflections inside the walls (1,4) whereas echoes (3,6) and (2,5) are concealed under the main reflections (2,5) and (1,4) correspondingly. For the usage of same walls material, the differences between the amplitudes of the reflected pulses are function of path length and incident angle only.

In this geometry reverberation signals may fulfill communication link for obstructed LOS. Thicker walls will produce more delayed but exposed echoes with lower amplitude which are negligible contributors in either event. Concerning asymmetric positions of the stations, six reflected pulses will be received severally. As it was mentioned, the model supposes smooth surfaces which reflect specularly. In reality most of the building materials have rough surfaces which reflect diffusely. Diffuse reflection will be interpreted as a partial specular reflection meaning lower signal power at the receiver site according to criterion that defines surface roughness.

Next in the same scenario consider a transmission channel involving half wavelength (3 cm at 5 GHz) dipole antennas installed vertically for both transmitter and receiver. For the sake of UWB channel characterization, dipole radiation pattern must be viewed not only as function of elevation angles but as frequency dependent as well. Antennas locations specify the elevation angles which are different from a maximal gain direction (90°). The radiation pattern in frequency domain is given by:

$$ V(\theta, \phi, f) = \frac{\cos \left( \frac{\pi f L}{c} \cdot \cos(\theta) \right) - \cos \left( \frac{\pi f L}{c} \right)}{\sin(\theta)} $$

where $L$ is dipole length, here $\theta$ and $\phi$ denote elevation and azimuth angles. It is plotted in figure 22 for existing seven directions within the room when each two of six reflected multipath components from opposite walls are equal due to symmetry.
All the paths except (3,6) have similar elevation angles which are very close to maximal direction (unity gain at central frequency) due to little difference in dipole heights relatively to direct separation distance. Thus rays reflected from walls (1,4 and 2,5) propagate almost in horizontal plane. Rays reflected from floor and ceiling pass straight in vertical plane hence they acquire smaller elevation angle (lower gain).

Figure 23 demonstrates an influence of the frequency dependent radiation pattern upon total received power for TE polarization comparatively to LOS.

![Graph showing radiation pattern and received power comparison](image)

**Fig. 22.** Radiation patterns of half wavelength dipole for available directions in in-room scenario.

**Fig. 23.** Received TE-power (7 rays) compared to LOS power in in-room scenario with dipole antennas at \( f_0 = 5GHz \).

Obviously radiation pattern dominates the frequency response of the channel causing a dramatic change in its general tendency. Maximal narrow band contribution of 8.6 dB is achieved. The result of the same ultra short temporal pulse propagation through this transfer function appears in figure 24.
Fig. 24. Received TE-signal (7 rays) in in-room scenario with dipole antennas at \( f_0 = 5GHz \).

In time domain slightly reduced amplitudes of the pulses LOS, (2,5) and (1,4) are observed whereas the reflections (3,6) are significantly lessened reasoning from a values in which rays intersect the radiation pattern. Again Gaussian waveforms successfully deal with dispersion of the antennas retaining their lineshapes.

Back to isotropic radiators and turn the room into a corridor by elongation of the \( x \) axis unto 50m. In an in-corridor scenario (50,4,3)m the transmitter is stationary (1,1,2)m and the receiver is moving from (2,3,1)m to (49,3,1)m along the \( x \) axis. The received power (TE-wave) as function of direct distance between the antennas for narrow band (central frequency) is shown in figure 25.

Fig. 25. Received TE-power (7 rays) compared to LOS power in in-corridor scenario at 5 GHz (narrowband).

LOS power decays inversely proportional to the distance squared ranging from 2.45 m to 48.05 m. Power contribution attains its maximum of 13.1 dB at the distance of 41.45 m from the transmitter due to existent waveguide effect in the corridor. Regarding wide band, the channel is probed when the receiver has a fixed coordinates at (49,3,1)m. The
propagation of the same UWB pulse for TE-wave in contrast with LOS is resulted in figure 26.

![Graph showing received signal over time](image)

**Fig. 26.** Received TE-signal (7 rays) in in-corridor scenario at \( f_0 = 5\text{GHz} \).

The primary pulse is an integration of the reflections from the walls 2,5,3,6 and LOS due to similar delays. Its envelope is slightly distorted but easily recognizable as Gaussian. Together, combined pulses contribute 5 dB to the single ray. Symmetric reflections from the walls 1 and 4 are constructively interfered into the secondary pulse and always lagged behind the rest (longer path) in the given geometry. As the receiver recedes, incident angles of the walls 1,4 and 2,5,3,6 approach zenith (0°) and grazing (90°) angles respectively. Especially while it reaches the furthest position, the lengths of LOS path and paths with reflections (2,5,3,6) become relatively equal one to another, consequently the delays also. During onward movement of the receiver, the pulses gradually overlap each other until complete merger. The longer the temporal pulse the sooner they merge and the higher power they add in final location of the receiver. In Table 4 the wave guiding contributions of the first received signal are compared under different standard deviations for both polarizations.

<table>
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<tr>
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<tbody>
<tr>
<td>0.2</td>
<td>5</td>
<td>2.8</td>
</tr>
<tr>
<td>0.5</td>
<td>6.7</td>
<td>4.7</td>
</tr>
<tr>
<td>0.8</td>
<td>7</td>
<td>5</td>
</tr>
</tbody>
</table>

The portions of contributed power decrease in relation to evenly growing standard deviation. Once all necessary contributors (2,5,3,6) are perfectly integrated, signal power comes to “saturation”. Thereby, longer signal having \( \sigma_m = 1\text{ns} \) supplies approximately the same extra power as previous one (0.8 ns).
In the continuation of wide band analysis, the peak of the total received TE-pulse $\sigma_{in} = 0.2\, ns$ (five rays) is compared to the signal amplitude of LOS ray as function of distance (mobile receiver) in figure 27. For this purpose the data are collected every 20 cm with numerical procedure execution (fig. 12) in each step when elapsed time is around 145 s.

Reflected rays begin to endow from 4.07 m with top value of 10 dB accomplished at 33.47 m whereas the contributions of TM-pulse commence from 25.5 m and gain superior limit of 5.9 dB at 34.67 m. Once more, owing to symmetry between rays reflected from opposite walls UWB pulses interfered constructively. On the other hand, destructive interference (negative contribution) was spotted in certain segments while using narrower pulses. Additionally, severer and oftener power detriment may occur for asymmetric antennas location, i.e. power delivered by single ray is several dB higher than overall received power.

8 60 GHz

60 GHz radio is expected to provide high throughput up to several Gbps for short range communications. It is recognized that for high walls attenuation (735 dB/m for hardened concrete) and great free space propagation loss (68 dB at 1 m) reliable links at 60 GHz are essentially confined to a single room. The path loss is compensated by higher transmit power while nothing can be done about obstructions. The main propagation mechanisms for 60 GHz indoor wireless are LOS and reflections due to smaller wavelength of 5 mm compared to objects.

In pursuance of the above, transmission through thin homogeneous slab and in-room propagation are reviewed. The transmission and reflection at normal incidence of ultra
short pulse (0.2 ns) through window glass with thickness of 2.84 mm and relative permittivity of \( \varepsilon_r = 2.613 - j0.1686 \) at 60 GHz are submitted in figures 28 and 29.

Fig. 28. Transmission of a Gaussian pulse through window glass at normal incidence.

Fig. 29. Reflection of a Gaussian pulse from window glass at normal incidence.

Despite high attenuation of the material (570 dB/m), the echoes are not absorbed in thin layer but indiscernible because they are obscured beneath the main pulses due to short delay (one way trip 15.3 fs) compared to standard deviation. Inlying echoes cause insignificant time shift in major pulses which ensues in transmission and reflection joint delays of 15 fs and 4 fs. The peaks of the signals precisely agree with the values of the transfer function at central frequency due to absence of power delay in time domain (unified envelope) against thicker slab.

In-room symmetric scenario with isotropic radiators and hardened concrete walls (\( \varepsilon_r = 6.5 - j0.343 \)) is contemplated for 60 GHz indoor applications. Note that reflection coefficient (-7.19 dB) has almost frequency invariant straight line form declining barely by 0.06 dB within 10 GHz band around central frequency of \( f_0 = 60\text{GHz} \). Fluctuations wane due to greater wall thickness (10 cm) contrary to penetration depth of
\( \delta_p = \frac{1}{\alpha} \approx 1.5\text{cm} \). The UWB TE-signal (0.2 ns) received in this scenario is shown in figure 30.

![Figure 30](image)

**Fig. 30.** Received TE-signal (7 rays) in in-room scenario at \( f_0 = 60\text{GHz} \).

In contrast to the results for 5 GHz carrier (fig. 20) some principal differences stand out. The most prominent one is entirely attenuated echoes that can be sighted via their phases. Therewith, the amplitude of LOS signal is lowered by 21.8 dB which is only a free space path loss disparity between two frequencies and strength reduction of reverberant pulses is 22.1 dB in average due to lack of echoes (preferable situation) and alteration in reflection coefficient in addition to free space loss. Bottom line, dielectric properties of the material is not a crucial factor in 60 GHz indoor propagation since the real part of complex permittivity does not vary substantially throughout the spectrum.

9 **Conclusion**

A theory for multi layer transmission and reflection was developed for both narrow band (single frequency) and wide band (UWB = short temporal pulse) cases taking into account frequency dependent dielectric constant. Then the multi-ray model was built for UWB signal propagation prediction in indoor wireless channel. This flexible model allows user to configure diverse geometrical environments comprising adjustable antennas locations and grants both frequency and distance dependent wideband results simultaneously. In order to evaluate the model there are measurements to be made and collated with simulations.

In room, the durability of Gaussian waveforms to both medium and radiation pattern dispersion was verified whilst their phases were distorted. In corridor, UWB incorporated pulses have distorted envelope with fluctuating positive to negative contribution power. The model is applicable for primal situation assessment by preliminary run-ups before employing expensive and computationally complex tools (heavy duty software) that require long simulation time. The indoor environment can be expanded by attaching
adjacent rooms behind each wall of the single room. It is possible to simulate a new scenario with definite receivers locations in the supplementary areas using both the transmission and reflection at once.

Future plans are scanning the room by means of changing receiver location in order to obtain power delay profile (PDP) in each point and calculate mean delay spread?

References: