A UWB Power Delay Profile Data Acquisition for Indoor Environments Using a Spectrum Analyser

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Abstract— **This paper presents a measurement system for indoor characterization of UWB channels. The presented method uses a scalar network analyzer (SNA) or a spectrum analyzer (SA) with tracking generator, connected to two antennas: a receiving and a transmitting one. This setup, which obtains only the magnitude data of the frequency response, makes the method less expensive when compared with other setups presented in the literature. From the magnitude data of the frequency response, a processing algorithm obtains the phase data of the frequency response making it possible to calculate the channel power delay profile. The tests for validation of the method are detailed herein.**

measurements; indoor radio propagation; spectrum analyzer

I. INTRODUCTION

In a typical indoor radio communication system the signal transmitted from TX reaches the RX through different paths with or without direct line of sight. The signals from different paths combine themselves to produce a distorted version of the transmitted signal. Therefore a detailed analysis of such channels is very important to the development, optimization and simulation of new systems.

The literature already presents some methods for this proposes. In one of them, [1], narrow pulses are transmitted, and the received signal, composed of delayed versions of the original pulses, is analyzed with a memory oscilloscope. However, this method is extremely limited for indoor environments because in this kind of channel the reflections occur very close to each another, requiring very narrow pulses for proper time resolution.

Another method, [2], solves this problem using a vector network analyzer that obtains the frequency response with magnitude and phase information, and post-processing through Fast Fourier Transform in order to obtain the power delay profile. Still a VNA is a quite expensive piece of equipment.

The method presented here, based on [3], makes the system less expensive, and more viable, using a scalar network analyzer or a spectrum analyzer with a tracking generator. The measurement setup is presented in figure 1.

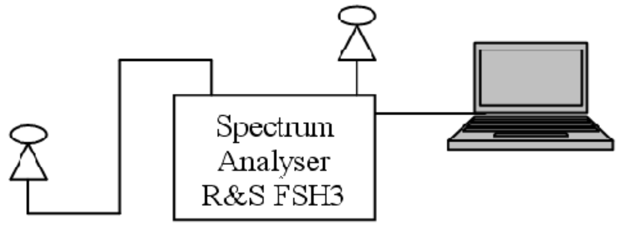


Figure 1. The measurement setup

II. THE MULTIPATH CHANNEL

Due to the randomness of the obstacle position and movement, the multipath communication channel is highly subject to random changes at random moments. Therefore, the channel presents different impulsive response at different instants.

As a consequence, these responses are dependent on two time variables: one related to the channel movement randomness, like mobile scattering objects, and another one related to the multipath component's time of arrival.

Since the speed of scattering objects in an indoor multipath environment is very low (like a walking person), this characteristics may be considered time invariant, and a stationary version of this model can be used, [4], reducing the impulse response to

$$h(\tau) = \sum_k a_k \delta(\tau - \tau_k) e^{i\theta_k} \tag{1}$$

where δ is the Dirac function and a_k, τ_k and θ_k are, respectively, the random variables relating to the amplitude, time of arrival and phase for each k multipath component.

From this impulsive response one can determine the PDP of the received signal by

$$P(\tau) = \frac{|h(\tau)|^2}{\int |h(\tau)|^2 d\tau}. \tag{2}$$

And, from this profile one can extract the most common multipath time scattering parameters used to describe the channel, namely: the mean excess delay, $\bar{\tau}$, and the RMS delay spread, σ_τ, given, respectively, by [3]

$$\bar{\tau} = \int (\tau - \tau_0) P(\tau) d\tau \tag{3}$$

$$\sigma_\tau = \sqrt{\int (\tau - \bar{\tau} - \tau_0)^2 P(\tau) d\tau} \tag{4}$$

where τ_0 is the time of arrival of the shortest multipath.

III. THE IMPULSE RESPONSE ACQUISITION

If the frequency response, $H(j\omega)$, represents a minimum phase system, i.e., if its poles and zeros lay on the left semi-plane, it is possible to obtain the phase data from the magnitude data of the frequency response with the application of the Hilbert Transform, as will be shown.

Since $H(j\omega)$ is composed of two parts, a real $H_R(j\omega)$ and a imaginary one, $H_I(j\omega)$, it can be shown, [6], that the impulsive response $h(\tau)$ is also formed by two parts, an even symmetry part, $h_e(\tau)$, and an odd one, $h_o(\tau)$,

$$h(\tau) = h_e(\tau) + h_o(\tau), \tag{5}$$

with the properties

$$h_e(\tau) = h_e^*(-\tau), \tag{6}$$

and

$$h_o(\tau) = -h_o^*(-\tau), \tag{7}$$

where * represents the conjugated complex.

The even symmetry part and the odd symmetry part can be obtained from the impulsive response by:

$$h_e(\tau) = \frac{1}{2}\left(h(\tau) + h^*(-\tau)\right) \tag{8}$$

$$h_o(\tau) = \frac{1}{2}\left(h(\tau) - h^*(-\tau)\right) \tag{9}$$

Alternatively, if, and only if, $h(\tau)$ is causal, $h_o(\tau)$ can be obtained from $h_e(\tau)$ by:

$$h_o(\tau) = h_e(\tau).sgn(\tau), \tag{10}$$

where

$$sgn(\tau) = 1 \text{ for } \tau > 0 \quad \text{and} \quad sgn(\tau) = -1 \text{ for } \tau < 0. \tag{11}$$

Therefore, if $H_R(j\omega)$ and $H_I(j\omega)$ are the Fourier transforms of $h_e(\tau)$ e $h_o(\tau)$, respectively, we have

$$H(j\omega) = H_R(j\omega) + H_I(j\omega), \tag{12}$$

Additionally, if $h(\tau) = 0$ for $\tau < 0$, the imaginary part $H_I(j\omega)$ can be obtained by:

$$H_I(j\omega) = j\psi\{H_R(j\omega)\}, \tag{13}$$

where $\psi\{.\}$ is the Hilbert Transform operation.

On the other hand, the function $H(j\omega)$ can be written, alternatively, in terms of the magnitude and phase:

$$\widetilde{H}(j\omega) = \ln[H(j\omega)] = \ln|H(j\omega)| + j.\arg[H(j\omega)], \tag{14}$$

where arg [.] represents the phase.

If the system can be considered a minimum phase system, the inverse Fourier transform of $\widetilde{H}(j\omega)$ is causal. It can be shown, [6], that the phase frequency response data and, consecutively the total channel frequency response, can be obtained by

$$\arg[H(j\omega)] = j.\psi\{\ln[|H(j\omega)|]\}, \tag{15}$$

providing, in this way, the phase data from the magnitude data.

IV. ALGORITHM AND SETUP VALIDATION TESTS

This validation was made with the use of the simulation of two causal systems with known impulse response, and by the comparison of two real measurements sets: one obtained with the vector network analyzer and other obtained with a scalar one.

Its objective was to verify if the indoor environments could always be considered a minimum phase response channel and, in the negative case, which are the implications.

A. The First Teste: Mathematical Validation

In the first procedure the impulsive response was determined by three processing methods and their results were compared.

The methods used the continuous Laplace inverse transform, the discrete Fourier inverse transform (with magnitude and phase data) and the set composed of the Hilbert and Fourier inverse transforms (only magnitude data) .

Since this validation's goal was purely mathematical, any known minimum phase frequency response channel could be used, independently of its central frequency and passing band.

So, we chose a 256,4 MHz resonant circuit and a band of 512 MHz to analyze. The resonant circuit is shown in figure 2 and its frequency response appears on figure 3.

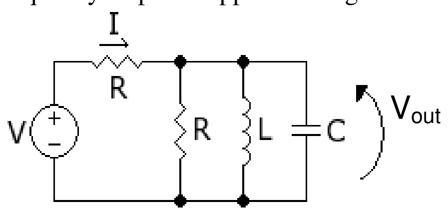


Figure 2. RLC resonant circuit

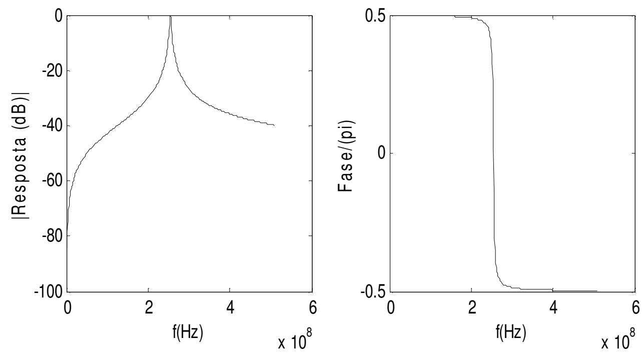


Figure 3.Resonant circuit frequency response

The impulsive response in the Laplace domain is given by

$$H(s) = \frac{s}{RCs^2 + 2s + \frac{R}{L}} \tag{16}$$

where $R = 50\ \Omega$, $L = 239{,}4\ pH$ and $C = 1{,}632\ nF$.

It results in two poles $s_{1,2} = -1{,}2252.10^7 \pm 1{,}5997.10^9 i$, and in one zero $s_z = 0$.

Since the system's zero and poles lay on the left side of the s plane, it can be concluded that the system is a minimum phase one and can be used to test this proposed method.

So, starting with the equation 16, the discrete frequency response $H(\omega)$ was sampled with a $\Delta_f \approx 1MHz$ and, with the inverse fast Fourier transform (IFFT), it was obtained the discrete impulsive response.

In order to finally validate the algorithm, the frequency response's phase data information was discarded and the Hilbert transform was used in the magnitude data information to recover the phase data information, from these values, the IFFT was used to find the discrete impulse response.

These impulse responses are shown in figure 4, where can be observed the existence of a good agreement between them and, from this agreement, the mathematical processing used to find the impulsive response can be considered as validated.

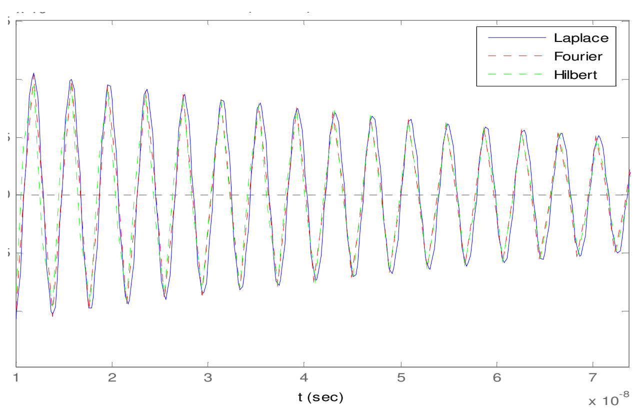


Figure 4. Impulsive response of the RLC resonant circuit

B. The second teste: Channel Computational Simulation

The second part of the validation consisted on the analysis of the frequency data obtained in a simulation of the transmission of a 500 MHz-band signal centered in 2GHz through a channel modeled by two multipath components: a direct and a reflected one.

The simulated channel, shown in figure 5, can be viewed as a free semi space terminated in a conductor wall, with the transmitter and the receiver aligned in the normal direction of the conductor plane.

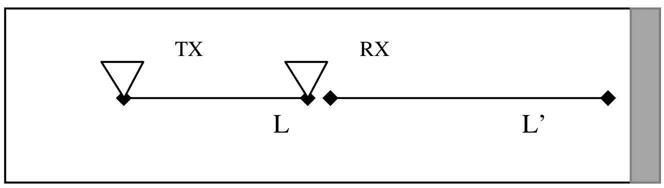
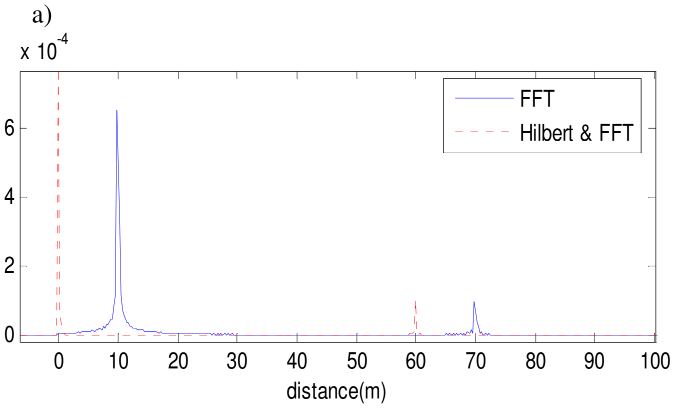


Figure 5. Simulated channel

The analysis was made by setting the distance L' at 30m and L at 10m, and thereafter, the distance L was set at 30 m and L' at 10m. The impulse responses obtained are presented in the figure 6.

a)



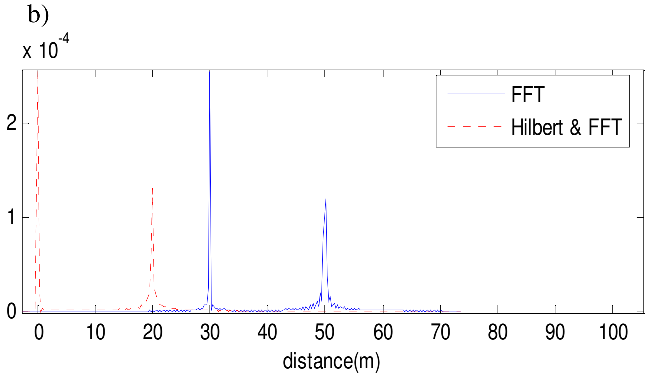


Figure 6. Impulsive responses in space domains for a) $L^{'} = 30m$ and $L = 10m$ and for b) $L^{'} = 10m$ and $L = 30m$

We can see that in both cases, the impulse response obtained from the magnitude and phase data is delayed from that one obtained only from the magnitude data. This delay corresponds exactly to the time of arrival of the direct multipath component, i.e., the Hilbert method considers that the time of arrival of the first ray is $\tau_0 = 0$, when the other method consider the propagation time from TX to RX, i.e. $\tau_0 = \frac{L}{c}$, where c is the light speed.

C. The third test: Comparison Between Two Measured sets: Using a VNA and a SNA.

The last test consisted in the acquisition of a real channel's frequency response with a VNA and in the calculation of the impulsive response in two different ways:

- Using the magnitude and phase data of the frequency response to obtain the impulse response by the inverse Fourier transform, and
- Using only the magnitude data of the frequency response, with the goal to simulate the usage of a scalar network analyzer, to obtain de impulse response by the Hilbert and inverse Fourier transforms.

The analyzed channel was the lab of microwave measurements in the University of São Paulo. It is a room divided by wood-and-glass dry walls into seven parts.

The antennas were, placed in the biggest room, separated by 6.5 meters in line of sight. The lab's floor plan is presented in figure 7.

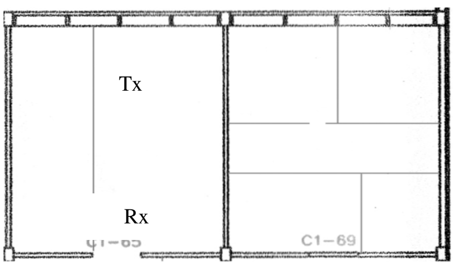


Figure 7. USP's measurements lab

The data acquisition consisted on the average of five scans of the band between 1,75GHz e 2,25GHz. During this process, there was nobody walking in the rooms and the internal doors were open. The external door was kept closed to avoid external influences in the channel. Figure 8 shows the obtained impulsive responses.

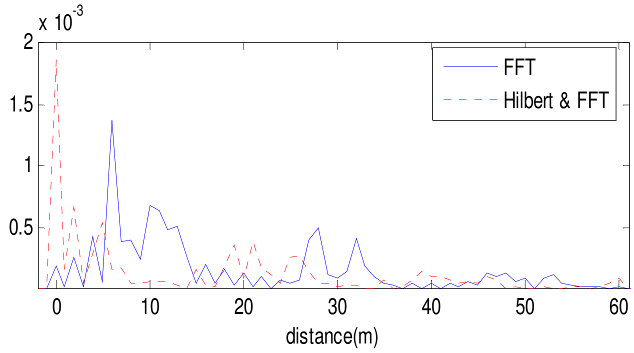


Figure 8. Impulse response obtained from the VNA data.

Analyzing the two responses, it is possible to note that they present the same multipath components, but in different instants. The first one is delayed by 22ns, what is equivalent to travel approximately 6.5m, exactly the distance between the antennas.

The multipath time scattering parameters find for this specific data set are presented in the table 1.

Table 1. Scattering parameters

	FFT	Hilbert & FFT
τ_0	22ns	0s
$\bar{\tau}$	23,02ns	13,37ns
σ_τ	116ns	30,16ns

V. Considerations About The Usage of This Methos in a Non-Minimum Phase Channel

The results above show that both methods present the same multipath components in the obtained impulse response, even the impulse responses are delayed one to another.

This occurs because two identical impulse responses that are out of phase have the same Fourier transform magnitudes, but different phases, and the processing algorithm used here recovers only the one which has the minimum phase.

As shown in [6], starting from a minimum phase transfer function, $H(j\omega)$, is it possible to obtain a new function, maximum phase, $F(j\omega)$, by the multiplication of $H(j\omega)$ by an all pass function $A(j\omega)$, as in the equation:

$$F(j\omega) = H(j\omega)\,A(j\omega) \tag{17}$$

where

$$A(j\omega) = \frac{(z^{-1} - z_0{}^*)}{1 - z^{-1}z_0} \, |z = e^{j\omega}$$ (18)

So, since that $|A(j\omega)| = 1$, the functions $F(j\omega)$, and $H(j\omega)$ will have the same magnitude, but different phases.

In this way, a minimum phase transfer function, $H(j\omega)$, with N zeros, has the same magnitude as the others $2^N - 1$ distinct non minimum phase transfer functions. The 2^N transfer functions that embody this set of functions, in accordance with the Parseval's theorem, have the same total energy ε, given by

$$\varepsilon = \int_0^\infty |h(\tau)|^2 d\tau$$ (19)

The corresponding impulse responses of this set (obtained from the transfer function zeros displacement by the all pass function multiplication), also, have the same total energy but different part-time energy.

In this way, the minimum phase transfer function multiplication by an all-pass filter tends to spread the impulse response in time, without any change in the total energy.

So, the assumption of minimum phase channel results in $h(\tau)$ with a minimum energy spread. And, in the same way, the assumption of maximum phase channel results in a $h(\tau)$ with maximum energy spread.

In this way, it is shown that in the case of non minimum phase, e.g., the third example, the obtained data using a scalar network analyzer or a spectrum analyzer with a tracking generator can be used to give the minimum energy spread in the channel power delay profile, and then, if necessary, the data can be processed with the goal to give the maximum-phase transfer function and so obtain the maximum energy spread.

Herein, starting from the tests results, one may conclude that the set of setup and algorithm processing is validated to be used in an indoor UWB power delay profile measurements campaign

VI. CONCLUSION

A procedure to obtain empirical data about the UWB time dispersion in indoor environments has been presented.

For this propose, simulations of the impulsive responses of known channels were made, and they showed that the system and the algorithm processing are viable.

But, a comparison between two impulsive responses, obtained with a vector network analyzer and with the proposed method, showed that the second one presents a time advanced impulse response.

This occurred because the proposed algorithm, which uses the Hilbert transform, supposes a minimum phase system, meaning that this method gives only the minimum spread time impulsive response.

So, even though the analyzed channel was not a minimum phase, this method can still give the bounds for the spread impulse response in time.

Since the method is validated, the next step consists in a measurement campaign to be done in the same building, already characterized in narrow band by [3]. With this measurement data, it will be possible to have a complete characterization of that building in order to help the development of simulation tools, as done in [7], for wide band.

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