Antenna Effects and Modelling in UWB Impulse Radio

Short title: UWB Antenna Effects and Modelling

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Abstract— The first part of this communication addresses the question of the representation of UWB antennas, particularly in the Time Domain, in terms of effectiveness, compactness and versatility. The second part of this work deals with the role of the antenna(s) in the effective radio link performance, taking into account antenna dispersion with respect to radio channel dispersion.

Keywords: UWB, antenna, antenna model, antenna impulse response, radio channel, antenna dispersion

1. Introduction

This communication intends to tackle several questions raised by the behaviour of the antennas and the propagation channel in the Time Domain (TD) of UWB (Ultra Wide Band) impulse radio :

1. How to model the UWB antennas, and mainly their TD (or FD) response, since unlike in the narrow band case, they operate as "multidimensional" linear filters (they are represented by a transfer function H for each direction of space (θ,φ)). A subsidiary question arises here : how to characterize the antennas efficiently ? Although it is not a fundamental theoretical question, it is very important in practice : such a transfer function holds a huge amount of data, is difficult to manage in practice and, moreover, is probably not necessary if we address the overall system from a statistical point of view.
2. What is the relative impact of the antennas behaviour (gain and dispersion/distortion) and the channel characteristics (mainly regarding multipath density) on the overall system performances ?

2. Antenna Numerical Models

The starting point of the adopted process is to define an intrinsic and characteristic quantity from which almost all the antenna parameters could be derived. Aiming to consider not only the directional properties but also the *phase* and *polarization* information, this quantity should be a **complex vector transfer function**, i.e., an amplitude transfer function and not a power one : this should consequently rely a quantity which characterize for example the radiated far field (but independent of the distance – so called the *vector* *amplitude* of the field) to a quantity which characterize the source, namely the amplitude of the incoming signal, chosen here as the incident partial wave a_1.

This vector transfer function $\mathcal{H}$, for the transmitting mode (Tx) is defined as [1]:

$$\mathcal{H}(\hat{\mathbf{r}}, f) = \frac{\mathbf{A}(\hat{\mathbf{r}}, f)}{a_1(f)} \quad \text{with} \quad \mathbf{E}(\mathbf{r}, f) = \frac{e^{-jkr}}{r}\sqrt{\frac{\eta_0}{4\pi}}\mathbf{A}(\hat{\mathbf{r}}, f) \tag{1}$$

where $\mathbf{A}(\hat{\mathbf{r}}, f)$ is the vector amplitude of the field in a given state.

Of course, the *vector impulse response*, (or simply the *impulse response*) of the Tx antenna is the inverse Fourier transform of $\mathcal{H}$.

For clarity, only the simple case of linear polarization and omni-directionality is presented here, considering that the extension to the general case is straightforward.

2.1. Time Domain Modelling with IIR Filters

The first proposed model is a representation of the antenna response (in the Time Domain) with an IIR "filter bank", computed from the impulse responses (measured or simulated with an e. m. tool) with the classical and simple Steiglitz-McBride identification method. For the simplified case of linear polarization and omni-directionality (or for a specific azimuth direction, e.g. φ = 0 – Fig. 1), the transfer function is represented by :

$$\mathcal{H}(\theta, z) = \left(\sum_{n=0}^{N_b} b_{n+1}(\theta) z^{-n} \bigg/ \sum_{m=0}^{N_a} a_{m+1}(\theta) z^{-m} \right) \hat{\theta} \tag{2}$$

the coefficients a_n and b_m being identified from the sampled impulse response $\mathcal{H}(\hat{\mathbf{r}}, t_k)$.

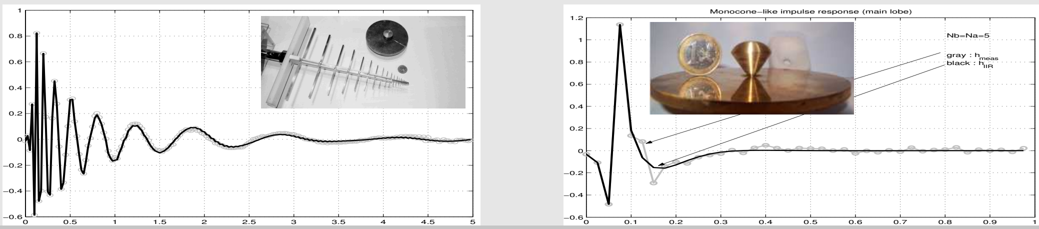

Figure 1 : comparison of the measured (black) and IIR filter impulse responses (grey circles) for a LPDA (left) and a shaped monocone (right).

The obtained compression factors range from about 4 to 10 or more (i.e. 75 to 90 % of compression rate), depending on the requested precision and on the impulse response time duration (the longer the spreading the more efficient is the compression).

2.2. Spherical Modes Expansion Modelling

The second model is based on a representation by projection of the antenna impulse response (or transfer function) onto the spherical harmonics basis. It is well known that the field can be developed, at any distance r from the antenna, on the spherical harmonics basis (see e.g. Schelkunoff [2]). The writing of this *projection* is particularly simple for the case of the far field and is given here for the specific case of linear (vertical) polarization :

$$E^{\infty}(\mathbf{r}, f) = -j\eta \frac{e^{-jkr}}{r} \sum_{m,n} \alpha_{mn} \left[m \cot\theta P_n^m(\cos\theta) + P_n^{m+1}(\cos\theta) \right] \cdot e^{jm\varphi} \cdot \hat{\theta} \quad (3)$$

where the functions P_n^m are the associated Legendre polynomials. For the particular case of omni-directionality, and introducing Eq. (1), this expression leads to :

$$\mathcal{H}(\theta, f) = \sum_{n=1}^{\infty} h_n(f) P_n^1(\cos\theta) \cdot \hat{\theta} \quad (4)$$

where the functions $h_n(f)$ are nothing else than the θ-spherical modes of the antenna, which write, taking into account the orthogonality properties of the associated Legendre functions :

$$h_n(f) = \frac{2n+1}{2n(n+1)} \int_0^{\pi} [\mathcal{H}(\theta, f) \cdot \hat{\theta}] P_n^1(\cos\theta) \sin\theta d\theta \quad (5)$$

Of course, the Fourier transform being a linear operator, the equations concerning the impulse response are formally the same. Knowing the impulse response $\mathcal{H}(\theta, t_k)$, or the frequency response $\mathcal{H}(\theta, f_k)$, it is very simple to compute the radiation modes.

The preceding computed modes (in the TD) may, in turn, be represented by IIR filters as in sub-section 2.1., driving to the third model, i.e. :

$$\mathcal{H}(\theta, z) = \sum_{n=1}^{\infty} h_n(z) P_n^1(\cos\theta) \cdot \hat{\theta} \quad \text{with :} \quad h_n(z) = \sum_{p=0}^{B_n} b_{np}(\theta) z^{-p} \Bigg/ \sum_{q=0}^{A_n} a_{nq}(\theta) z^{-q} \quad (6)$$

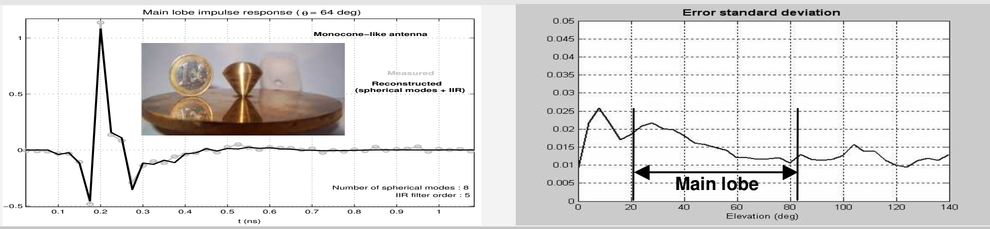

Figure 2 : comparison of the reconstructed and measured impulse response in the main lobe (left) and standard deviation of error versus elevation (right); 8 spherical modes used and 5th order filters

One of the practical interests of this modal expansion, notably considering the data compression objective, lies in the small number of modes required to achieve a good precision. This is particularly true for "small" antennas, but also true for "good" UWB antennas, i.e. antennas for which the gain and the radiation pattern is weakly frequency dependent. This is due to the theoretical fact that the number N_{max} of radiating higher order modes is limited by the "size" of the antenna, $N_{max} \sim kR_{min}$ (R_{min} being the radius of the smallest sphere enclosing the antenna). At the time of writing, it has been observed for several cases, that using for the wave number k in the above equation an average value – instead of the maximal value – (over the BW of interest) gives rise to satisfactory results.

Eventually, the best results (e.g., significantly improved data compression rate for a given error) are obtained when both "compression methods" are applied to the antenna response. Results for the same shaped monocone [3], [4] with 8 spherical modes and 5th order filters are presented (Fig. 2). Considering that a reasonable size of the input matrix $\mathcal{H}(\theta_n, t_k)$ is 36 x 100, the obtained compression factor is about **45** or **97.8 %**.

3. Combined antenna-channel effects

3.1.1. Problem statement and method of analysis

Impulse radio generally make use of a correlator or matched filtering in the receiver (Figure 3). Such a scheme is optimal in the presence of Gaussian noise and for an ideal channel. Unfortunately the latter is not realized in practice, both because of antenna and of channel dispersion/distortion of radiated and received signals. From the antenna designer point of view, it is interesting to investigate the role of antenna characteristics on the radio link performance. However it is not clear in what respect antenna distortion affects this performance. Another trivial way to express this, is the naïve question : "since the channel is so bad, does it make any difference if the antenna differentiates it to third, second or fourth order ?". For this purpose we investigate the performance of a pulse based radio link, evaluated through the signal to noise ratio (SNR) at the correlator output, under white input noise assumption. Cumulative distribution functions (CDF) of the SNR for various antenna/channel combinations are plotted, referenced to the ideal antenna case. Here an ideal antenna is defined as having a unity isotropic transmission function $\mathcal{H}^t$ with no dispersion.

In order to better identify the antenna role, we consider several normalizations of the correlator output signal in the frequency domain :

- no normalization, i.e. the SNR involves the antenna contribution as a whole
- a normalization is applied in order to ensure that in the case of an ideal (free space) channel, the energy of the output signal after correlation is unchanged with respect to an ideal antenna. This feature allows to remove antenna gain from the output signal, and to keep link performance affected only by the antenna dispersion in combination with the channel.
- a normalization is applied in order to ensure that at the Tx level, the EIRP is maintained identical to that of an ideal antenna. This feature is intended to evaluate the influence of the real Tx antenna while respecting the EIRP limitation imposed by regulation.

In addition, two kinds of reference templates in the correlator are used :

- The (ideal) template is the received signal for an ideal (free space) channel and ideal antennas at each both Tx and Rx levels.
- The (antenna filtered) template is the received signal for an ideal (free space) channel and the antennas under consideration at Tx and Rx levels. In other words this template takes into account the antenna characteristics, neglecting the angular dependence.

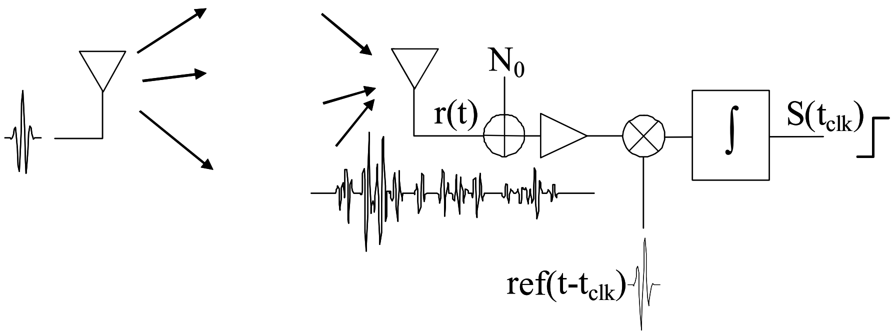

Figure 3: UWB communication system model

3.1.2. Signals, antennas and channels

We use a signal waveform at the transmitter level that is compliant with the FCC mask issued on 14 february 2002. The FCC requires the EIRP to be at most –41.3 dBm between 3.1 and 10.6 GHz, and much less outside this range. The transmitted and received waveforms (with ideal antennas) are shown in Figure 4, they were constructed from an initial rectangular spectrum (sinc function in the time domain), properly windowed. These waveforms will differ in the case of real antennas, due to antenna filtering/distortion. Several non ideal antennas are considered, with various magnitudes of the dispersion :

- the log-periodic dipole array modelled above, operating in the 1-18 GHz band.
- the monocone modelled above, operating in the 3-10 GHz band.
- a poorly matched simulated rectangular horn, with an operation bandwidth 2-5 GHz.

We have used two kinds of UWB channels :

- IEEE-802.15.3a channel model 4), which is a discrete multipath model based on a clustered approach. 4 kinds of channels are categorized, with varying Tx-Rx distance and varying degree of obstructions (LOS/NLOS).
- Experimental channel measurements carried out at ENSTA with a vector network analyser in the 2-10 GHz band. These measurements were made with a 3D automated positioner at one side of the radio link, enabling to obtain a small-scale space-variant statistical set of channels for each antenna position. Monocones on a ground plane were used for these measurements at both ends of the link. In the SNR simulations below for these channels, extra antenna transfer functions were added just like if the original ones were ideal, which obviously they could not be. This procedure is normally incorrect, since the true channel data is filtered twice per radio terminal, instead of once. This has the effect of reducing the channel+antennas bandwidth and increase the dispersion, with respect to the true situation. However the channel “richness” and multipath density are apparently not strongly affected by this procedure. Therefore we will provisionally consider it pertinent.

We also present here results for two channels, measured in the microwave laboratory of ENSTA : for the first the antennas are in unobstructed line of sight (LOS) at a distance of about one meter of each other. For the second (NLOS), the positioner location is unchanged, but the other antenna is placed in a neighbouring room, only separated from the laboratory by a very thin windowed wall and a closet. The LOS distance is typically 3m. In both cases it is clear that the environment favours a rich multipath structure (Figure 6).

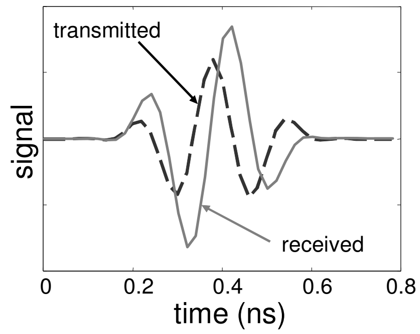

Figure 4: temporal (ideal) waveforms

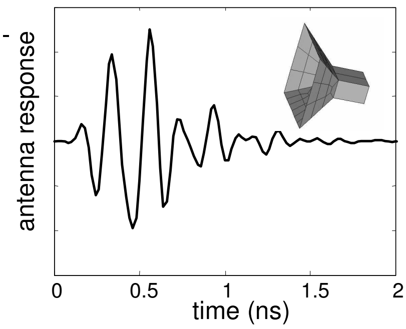

Figure 5: antenna impulse response for the simulated horn (inset)

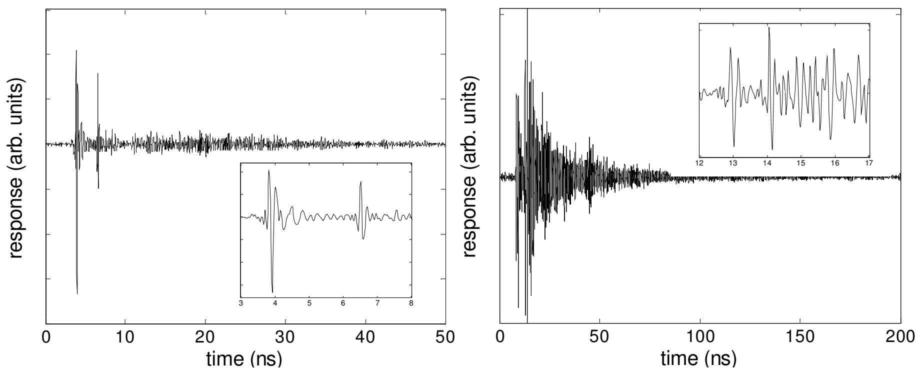

Figure 6: left : LOS channel in the microwave laboratory of ENSTA ; right NLOS channel

3.1.3. Results

Below are shown the statistical distributions of the modification of SNR by the replacement of ideal antennas by real antennas. First the effect of the various normalization conditions can be appreciated on Figure 7 : the antenna gain is directly visible on the SNR (full black curves). When a constant EIRP is imposed, this gain disappears whenever a real antenna is placed at the transmitter side. When a constant received power is imposed for an ideal (free space) channel, the antenna gain of both antennas is removed (grey curves). Still the SNR is different from its value for ideal antennas, which stems from the influence of antenna dispersion. In spite of the fact that the same antennas are used on both sides (right part of Figure 7), the SNR loss by imposing constant received power on an ideal channel is not twice the SNR loss by imposing constant EIRP. Again this results from the complicated interplay between antenna dispersion and channel dispersion. Such a difference is not seen in channels with less dense multipath.

In Figure 8 are shown the SNR distributions in the case of normalization by constant received power (all antenna gains removed), highlighting the role of antenna dispersion vs. channel dispersion, for the three antennas and IEEE channels CM1 and CM4. Several remarks can be drawn :

- The SNR is almost always worse than for ideal antennas.
- In general, the SNR is degraded when the ideal reference template is used, and less or not degraded when the antenna filtered template is used. The exception is essentially for CM4 channel, which is also the densest in terms of multipaths, and for the horn which also exhibits the highest dispersion.
- There is a greater dispersion of the SNR for the log-periodic antenna and the horn than for the monocone. This result is expected on the basis of the relative magnitudes of antenna dispersion.
- Although not shown in the plots, the SNR degradation is generally larger for two real antennas than for a single one, not necessarily by a factor two though.

The results in the case of ENSTA LOS and NLOS channels are basically on the same trends, with some similarities with IEEE-CM1 and CM4 respectively.

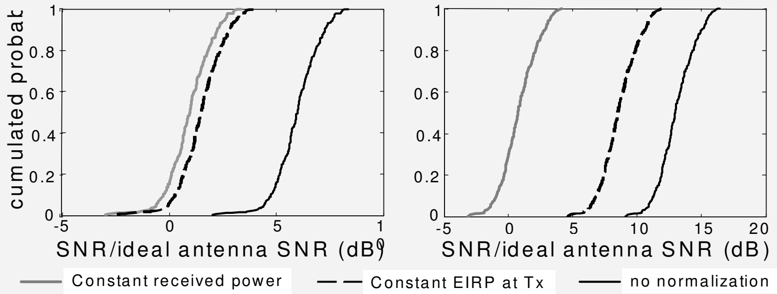

Figure 7: distribution of SNR with respect to ideal antennas and various normalization conditions ; simulated horn and ENSTA NLOS channel ; left : real antenna only at Rx ; right :real antennas both at Rx and Tx

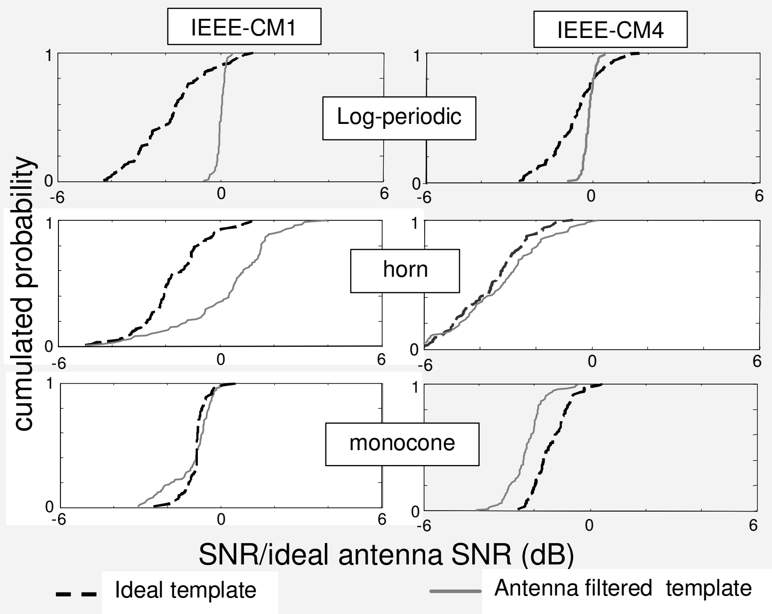

Figure 8: distribution of SNR with respect to ideal antennas IEEE channels) ; real antennas both at Rx and Tx (constant received power).

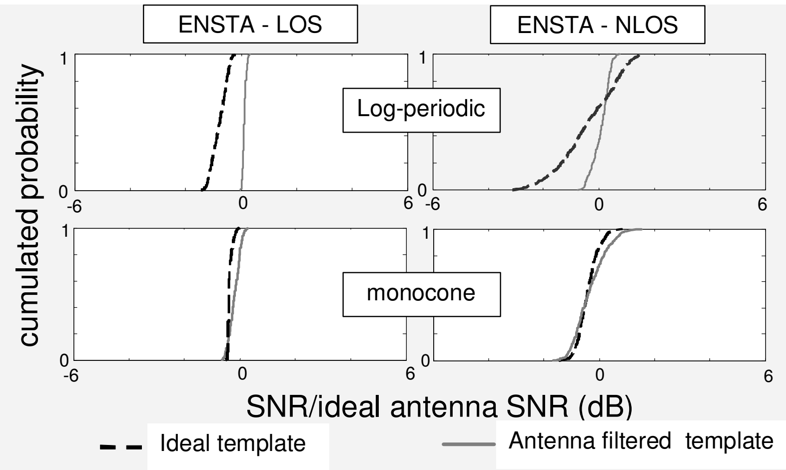

Figure 9: distribution of SNR with respect to ideal antennas (ENSTA channels) ; real antennas both at Rx and Tx (constant received power).

4. Conclusion

Numerical models for UWB antennas have been proposed to improve data handling, and may be useful for other downstream analysis tools such as simulations at the radio link level.

The role of antennas in the performance of a radio link has also been investigated, based on a distinction between antenna gain, and antenna dispersion combined with channel dispersion. It has been shown that a low dispersion antenna also exhibited moderate SNR degradation with respect to an ideal antenna (~ 0-2 dB), while a dispersive antenna (horn) could exhibit up to 6 dB degradation in the indoor channels considered.

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