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Time-Domain Diversity in Ultra-Wideband MIMO Communications

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The development of ultra-wideband (UWB) communications is impeded by the drastic transmitted power limitations imposed by regulation authorities due to the "polluting" character of these radio emissions with respect to existing services. Technical solutions must be researched in order either to limit the level of spectral pollution by UWB devices or to increase their reception sensitivity. In the present work, we consider pulse-based modulations and investigate time-domain multiple-input multiple-output (MIMO) diversity as one such possible solution. The basic principles of time-domain diversity in the extreme (low multipath density) or intermediate (dense multipath) UWB regimes are addressed, which predict the possibility of a MIMO gain equal to the product $N_t \times N_r$ of the numbers of transmit/receive antenna elements when the channel is not too severe. This analysis is confirmed by simulations using a parametric empirical stochastic double-directional channel model. They confirm the potential interest of MIMO approaches solutions in order to bring a valuable performance gain in UWB communications.

Keywords and phrases: MIMO, diversity, ultra-wideband, radio channel.

1. INTRODUCTION

Ultra-wideband (UWB) technologies are among the "hot topics" in the present days as their specificities are promising for future communications or positioning applications. Extremely cautious regulations are expected however due to the wide emitted radiation spectra which ignore the numerous protected bands. The latter exist for a great variety of scientific, public, or commercial services, and particularly sensitive to electromagnetic pollution are those requiring very low noise levels (spatial scientific services, fixed wireless access, GPS, etc.). The Federal Communications Commission (FCC), for instance, imposes to indoor communications a maximum emitted isotropic radiated power (EIRP) of -41.3 dBm/MHz between 3.1 and 10.6 GHz, and much less outside this band. Although this is still a subject of debate, European authorities will probably adopt conditions at least as stringent as the FCC. In spite of the numerous advantages of UWB, the transmitted power, at most -2.6 dBm but likely several dB less, will thus tend to limit applications to relatively short ranges or to moderate data rates. It is therefore crucial to develop solutions that make the best possible use of the radiated and received power, for the feasibility and the future commercial success of UWB communications systems.

In the present work, we address multiple-input multiple-output (MIMO) techniques as one possible solution to improve the UWB link robustness or its range. MIMO techniques for UWB have recently been investigated in the context of space-time coding for pulse-position modulations under the condition of flat fading [1], and in the case of full-channel equalization in order to improve performance through diversity [2]. Here we consider switched beam or time-domain combining employing multielement antennas at one or at both sides of the radio link. The pertinence of these approaches in a practical radio link is here tested by simulations. Two important issues are to be highlighted.

(i) The signal waveform, which will determine the bandwidth and spectral content, and will strongly affect the fading behaviour of the radio link. We used a waveform specially determined to comply with FCC regulations for indoor communications.

(ii) The channel model, which will obviously affect the link quality to a great extent. We used a stochastic Monte Carlo channel model based on the definition of multipath amplitudes, delays, directions of departures (DODs), and directions of arrival (DOAs), according to given statistical distributions.

In the following, we discuss general features of time-domain combining in MIMO arrays in Section 2. This is followed in Section 3 by a description of the channel model, of antennas and of the pulsed modulation and detection which are used in the simulations presented in Section 4. A final discussion and conclusion is given in Section 5.

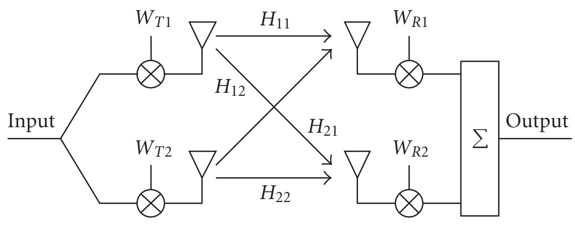

Figure 1: Schematic of a 2×2 linear combining MIMO system. The cross symbol is a product in the frequency domain and a convolution in the time domain.

2. DIVERSITY COMBINING IN THE TIME DOMAIN

2.1. Conventional frequency-domain combining

Taking as an example a 2×2 MIMO system (Figure 1), narrowband optimal combining in the partial 1×2 subsystem is achieved by maximal ratio combining (MRC) complex weights W_{R1} and W_{R2}, which maximize the signal-to-noise ratio (SNR) under the assumption of Rayleigh fading. The second set of weights W_{T1} and W_{T2} on the transmitter side provide an independent set of optimization variables, which can be used to maximize the 2×2 combined signal, provided channel knowledge is available at the transmitter. Applying phase-only weights (equal gain combining, (EGC)) results in a slight suboptimality [3]. The wideband frequency-domain case is basically not different from the narrowband case, except that the weights are frequency dependent, due to rotation of phasors with frequency and channel dispersion. Optimal combining can be achieved through digital processing of transmitted and received signals at baseband.

2.2. Time-domain combining

We consider a discrete channel model described by the space-variant impulse response

$$h(\vec{r}_1,\vec{r}_2,\tau) = \sum_i A_i\delta(\tau - \tau_i(\vec{r}_1,\vec{r}_2)) \tag{1}$$

with

$$\tau_i(\vec{r}_1,\vec{r}_2) = \tau_{i0} - \frac{(\vec{r}_1\cdot\vec{u}_1 - \vec{r}_2\cdot\vec{u}_2)}{c}, \tag{2}$$

where A_i are real amplitudes (positive or negative), and $\tau_i(\vec{r}_1,\vec{r}_2)$ the associated propagation delays for transmitter and receiver positions $\vec{r}_1$ and $\vec{r}_2$, respectively. $\vec{u}_1$ and $\vec{u}_2$ are unit vectors along the propagation direction at the transmitter and receiver locations. Dropping the spatial dependence for notational simplicity, the received signal writes $y(t) = \sum_i A_i f(t-\tau_i)$, where $f(t)$ is the received signal for an ideal nonattenuating nondelaying Dirac channel, yielding for the correlator output, $s(t) = \sum_i A_i g(t-\tau_i)$, where $g(t) = \int f(t')f(t-t')\cdot dt'$ is the ideal received signal autocorrelation. For a pair (m,n) of radiator (transmit side) and sensor (receive side), this can be written $s_{mn}(t) = \sum_i A_i g(t-\tau_{i,mn})$, where the array topologies are contained in the discrete form $\tau_{i,mn}$ of (2). Assuming the technological possibility of applying branch combining (real) coefficients α_m and β_n and branch delays $\Delta\tau_{Tm}$ and $\Delta\tau_{Rn}$, the time-domain combining problem can be stated as follows:

$$\underset{t,\Delta\tau_{Tm},\Delta\tau_{Rn},\alpha_m,\beta_n}{\mathrm{Max}} \left\{ \sum_{m,n,i} \alpha_m\cdot\beta_n\cdot A_i \cdot g(t-\tau_{i0}-\tau_{i,mn}-\Delta\tau_{Tm}-\Delta\tau_{Rn}) \right\} \tag{3}$$

constrained by the relations $\sum_m \alpha_m^2 = 1$ (constant transmitted energy) and $\sum_n \beta_n^2 = 1$ (constant noise energy). Currently the maximization on t is carried out by a sliding correlation, and α_m, β_n, $\Delta\tau_{Tm}$, and $\Delta\tau_{Rn}$ must be found.

2.3. Time-domain combining in the extreme UWB regime

We now concentrate on the "extreme" UWB regime, that is, for which a perfect time-resolved separation of all multipaths is achieved (Figure 2). In such a situation, it may be approximated that each received pulse is characterized by a given DOA and a given DOD. This approximation expresses the discreteness of the underlying channel model, both in the time and in the angular domains. It may turn out wrong, especially when the scattering objects are spatially extended and thus when the point source model is inadequate. Provisionally, we will accept its validity, implying that the time-domain transfer functions between radiators and sensors obey the general relation $\tau_{i,mn} = \tau_{i,Tm} + \tau_{i,Rn}$. For the 2×2 linear combining system, this yields $\tau_{i,11} - \tau_{i,12} = \tau_{i,21} - \tau_{i,22}$, which expresses that the two arrays are in the far-field region of each other. In this case, the solution to the maximization problem is trivial and writes for a single received pulse $\Delta\tau_{Tm} = -\tau_{Tm}$; $\Delta\tau_{Rn} = -\tau_{Rn}$; $\alpha_m = 1/\sqrt{N_t}$; $\beta_n = \mathrm{sign}(A_i)/\sqrt{N_r}$.

Under a Gaussian noise assumption, EGC is the optimal combining scheme, and is equivalent to MRC since the amplitudes of combined signals are equal. The combiner uses pure delay operators both at the transmitter and receiver levels in the time domain. For the 2×2 case,

$$\begin{aligned} W_{T1} &= \delta(t), & W_{T2} &= \delta(t+\Delta\tau_T), \\ W_{R1} &= \delta(t), & W_{R2} &= \delta(t+\Delta\tau_R). \end{aligned} \tag{4}$$

The final consequence is that we can expect the combined received pulse to be $\sqrt{N_t\cdot N_r}$ higher in magnitude than the SISO pulse, that is, a received energy $N_t\cdot N_r$ higher for the same noise energy, yielding an SNR gain also equal to $N_t\cdot N_r$.

It should be very clearly stated that in this regime no diversity gain in the usual sense can be obtained, since the channel does not fade. In other words, the average SNR antenna gain mentioned above will not translate into an increase of the slope of the bit error rate versus the SNR plot, which is the characteristic of true diversity.

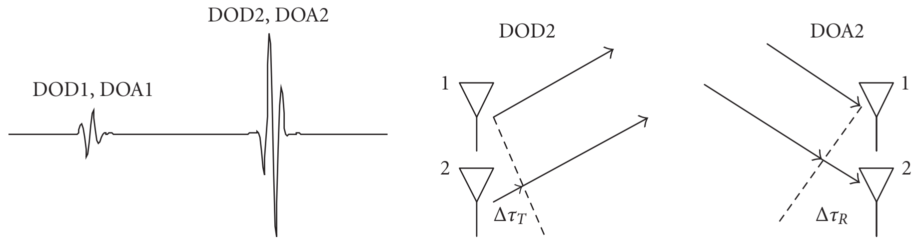

Figure 2: Delayed propagation in the extreme UWB case.

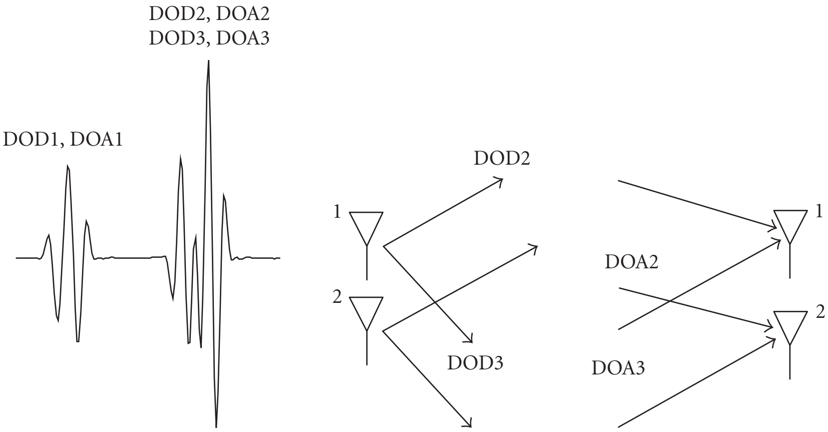

Figure 3: Delayed propagation in the intermediate UWB case.

2.4. Time-domain combining in the intermediate UWB regime (dense multipaths)

The extreme UWB regime will not be realized in most practical situations. This is due to the complex scattering environment of the radio equipments (buildings, furniture, people, etc.), which is responsible for a dense time-domain impulse response in the UWB case. It can be envisioned that in the intermediate UWB regime, several scattered pulses with nearly identical delays interact with each other at the transmitter and receiver levels (Figure 3). Due to the overlapping of pulses with differing DOAs and DODs, optimality can no more be achieved with a single-delay operator per combining branch. Capturing the maximum physically achievable energy would require the temporal alignment of all received pulses whatever the radiator/sensor pair. This can be optimally achieved in SISO, MISO, or SIMO schemes (S = single, M = multiple) through MRC RAKE combining *in addition to* multiple antenna combining. MIMO systems are intrinsically suboptimal, as there is no benefit in RAKE combining *both* at the transmitter and at the receiver level. It can easily be demonstrated for, for example, two "interacting pulses" of amplitudes A_0 and A_1 that the following equality holds

$$\begin{aligned}\operatorname*{Max}_{\alpha_T,\alpha_R}&\left\{\alpha_T\cdot\alpha_R\cdot A_0+\sqrt{1-\alpha_T^2}\cdot\sqrt{1-\alpha_R^2}\cdot A_1\right\}\\&=\operatorname{Max}\left\{A_0,A_1\right\},\end{aligned}\tag{5}$$

where α_T and α_R are RAKE combining coefficients at the transmitter and receiver, respectively. In other words, the gain is nil. RAKE combining on each antenna element of *either* the transmitter *or* the receiver is possible, but in a MIMO scheme, it will be impossible to temporally align both pulses for all combinations of delays, and the relation

$$\Delta\tau_{Tmk}+\Delta\tau_{Rn}=\tau_{i0}-\tau_{i,Tm}+\tau_{i,Rn}\tag{6}$$

cannot hold for all i for a single RAKE index k on the transmitter. Quantitatively, the suboptimality in the intermediate UWB regime depends on two main physical characteristics of the radio system in its environment.

(a) *The angular extension of the scattering objects* seen from transmitter and seen from the receiver. A large angular extension will favour SNR loss in the combining process at either transmitter or receiver. Angular spreads are meaningful parameters to express this effect.

(b) *The radio system bandwidth*, in relation with the time-domain size of the scattering objects. When the former is large, the scattering objects are time resolved, implying it is possible to separate them in the combining process. A 3 GHz bandwidth, for instance, typically yields a 10 cm resolution of the scattering objects, and therefore typically this same size for the sources of scattered radiation if we assume that the lateral and longitudinal sizes of the object are equal.

At a 5 m distance, this yields 1.2° angular spread, which is quite small. It thus clearly appears that the wider the band, the better the combining capability of the arrays. However, this statement should be appreciated cautiously since it applies to combining with a single RAKE finger. A large bandwidth also means less energy in individual scattering events, since the scattering cross-section of objects suffering time filtering in the UWB electronics will be smaller. It is well known that maximizing the energy capture in pulse-based UWB receivers requires many RAKE fingers in the dense multipath [4].

The multipath density is very important for the time-domain combining performance in the intermediate UWB regime, expressed by the nonzero probability of overlapping pulses. The fine characteristics of the radio channel are of utmost importance here, as they determine to what extent the multiantenna combining scheme will be able to mitigate angular dispersion, according to angular interpulse correlations. There is unfortunately no recognized model for interpulse correlations, which are intimately related to the physics of propagation and the properties of the scattering objects.

2.5. *Array size limitations*

Time-domain combining of pulse-based UWB signals does not involve the same size limitations as for frequency combining of narrowband signals. In the latter case, it is well known that the main requirement stems from the need to avoid too much correlation between signals received by close sensors. Correlated signals also mean a greater probability of simultaneous fading events, while the major role of spatial diversity is to combat fading. Since UWB transmission is reputed to exhibit little fading, this limitation does not operate in the same manner. In the extreme UWB regime, it can be stated that the correlation is 1, since changing the position of an antenna element only shifts the received signal in time, not its amplitude or shape. However, since this amplitude is unaffected, there is no disadvantage from this point of view in having close elements.

Nevertheless, elementary cleverness tells us that two radiators or sensors cannot be placed too closely. This is a physical rather than a signaling issue, which involves electromagnetic coupling acting as a short circuit between these antenna elements. Coupling ultimately functionally reduces a pair of elements to an equivalent single one, thereby cancelling any advantage brought by the array [5]. It was found by investigation of a realistic monopole UWB array configuration that such coupling had still moderate effects on the array radiation in the FCC spectral band, down to 4 cm intersensor distance.

At the other end, the discussion of Section 2.3 indicates that optimal combining in the extreme UWB regime only requires application of specified delays at both the transmitter and receiver antenna elements. Changing the array size will increase or decrease only the relative delays, but by no means the effectiveness of the combining process. Since these delays are both proportional to angle cosines and to the intersensor distances, we see that the combining electronics should be capable to achieve a sufficient delay excursion, which is an implementation limitation. Another implementation limitation lies in the required precision of the applied delay, which is all the higher as the array is small. Therefore the array size design will be a tradeoff, taking into account electromagnetic and electronic issues. It should also be recalled that UWB arrays generally do not exhibit secondary lobes which are commonplace in narrowband arrays, which is due to the very small temporal overlap of time-shifted pulses.

Another consideration applies to the array size: the larger the array, the higher the probability of overlap between differing echoes, which may reduce the link performance since we expect a suboptimality in the intermediate UWB regime.

3. SIGNAL WAVEFORM AND CHANNEL MODEL

3.1. *Signal waveforms*

FCC issued on 14 February 2002 an authorization for UWB devices intended for a few applications, among which are indoor communications. 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. This imposes a particular care to be exerted by designers of radio transmitters, in order to respect this spectral mask. Regarding the received signal, it should first be recalled that an "ideal" antenna, that is, a nondispersive antenna (phase linear versus frequency) with a frequency-independent gain magnitude, has an effective receiving area scaling with the wavelength squared. This means that for a flat transmission spectrum, the received spectrum has a fundamental downward slope of −6 dB per octave (fundamental receiving antenna filtering). In the plot below, this appears as a shift of the normalized received spectrum peak towards low frequencies (free-space propagation), as compared to the transmitted spectrum. In the present work, we use a transmitted signal obtained by inverse Fourier transform of a bandpass flat spectrum, properly windowed in the time domain to limit its duration (0.8 nanoseconds). According to the parameters chosen for the simulations, the transmitted half-power bandwidth is 2.81 GHz, and the received half-power bandwidth is 2.94 GHz. It is possible to find waveforms that make a better use of the spectral mask; however, the duration of the pulse will generally increase because of the Heisenberg relation.

3.2. *Channel model*

We use a space-variant discrete channel model. For a given position of the receiving and transmitting antennas, the channel is described as a discrete sum of multipaths, each characterized by its delay, its amplitude, its DOD, and its DOA. As compared to the more usual wideband channel models, the present model has the following differences.

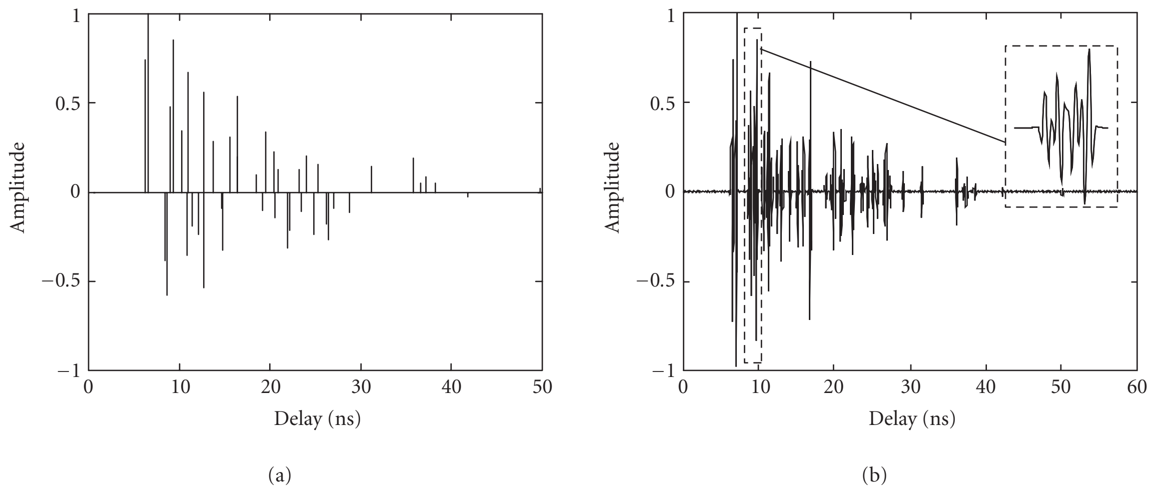

Figure 4: Example of a highly time-dispersive UWB channel (Case 4 channel); (a) infinite bandwidth; (b) received signal resulting from the waveform filtered by the channel.

(i) Since we deal with *real* and not complex signals, the channel impulse response (CIR) is real as well and not complex. This means, in particular, that the path amplitudes are real, positive, or negative. According to physical intuition, the signs and amplitudes of the received paths are related to the elementary events experienced by the pulsed waves, that is, specular metallic reflections, diffraction, transmission, and so forth.

(ii) We expect a much greater number of paths due to the extremely large bandwidth. The multipath density has been experimentally ascertained [6, 7, 8].

This model, in particular, assumes an identical number of DODs and DOAs, for a given channel realization. This is an approximation of possible reality, since "path junctions" may exist in certain circumstances, and eventually lead to keyholes [9]. The statistics and related parameters of path amplitudes, delays, DOD, and DOA are obviously a crucial issue for the pertinence of the channel model. Here we have made the following assumption.

(i) Path delays are distributed according to a Poisson law in fixed-delay bin durations, with a parameter of the Poisson function (mean number of paths per bin) decreasing as a function of the bin number. This allows to account for the rarity of significant paths with increasing delays.
(ii) Path amplitudes are governed by a Ricean distribution, whose K factor is randomly generated within certain limits (uniform law); the signs of the path amplitudes are also randomly chosen (uniform law).
(iii) Path DOAs and DODs are governed by a Gaussian distribution in both azimuth and elevation.

For a given channel realization, the knowledge of DOAs and DODs for each path allows to compute the additional path delay when either the transmitting or the receiving antenna is moved (small antenna approximation). This property will be used to compute the channel variation from one sensor to another in an array, or from one antenna position to another intended to generate a large statistical set of channels. In the simulations shown below, a few channel sets considered interesting to test the behaviour of multielement UWB antennas have been generated, according to the following approaches.

Case 1. A channel with little temporal dispersion (well-separated paths) and little azimuth angular spreads (10° standard deviation).

Case 2. A channel with little temporal dispersion but large azimuth angular spreads (60° standard deviation).

Case 3. A channel with strong temporal complexity (many close paths) and small azimuth angular spreads (10° standard deviation).

Case 4. A channel with strong temporal complexity and large azimuth angular spreads (60° standard deviation).

In all cases, the elevation angular spread was kept constant (10° around 0°). An example of a Case 4 channel is shown in Figure 4, both for the (infinite bandwidth) discrete path amplitudes and for the finite bandwidth case when the transmitted waveform is filtered by the channel. It can be seen in the zoomed inset that the signal waveform may strongly depart from the ideal received signal waveform of Figure 5.

3.3. Modulation and detection

Several modulation schemes have been considered:

(i) pulse-amplitude modulation (PAM);
(ii) binary pulse-position modulation (BPPM);

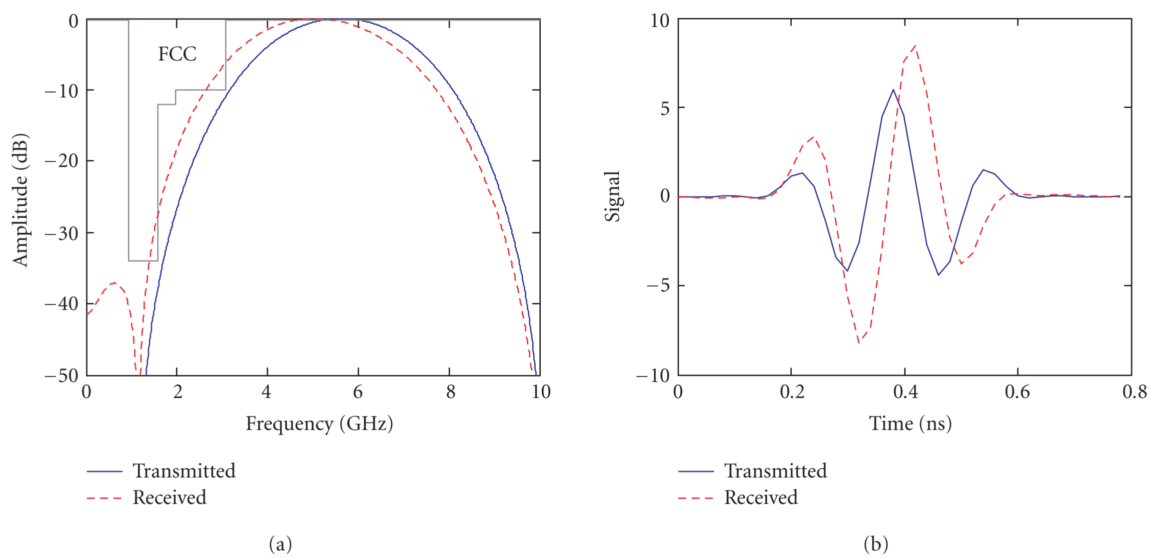

Figure 5: (a) Spectra of the transmitted and received signals, in comparison with the FCC mask. (b) Temporal waveforms.

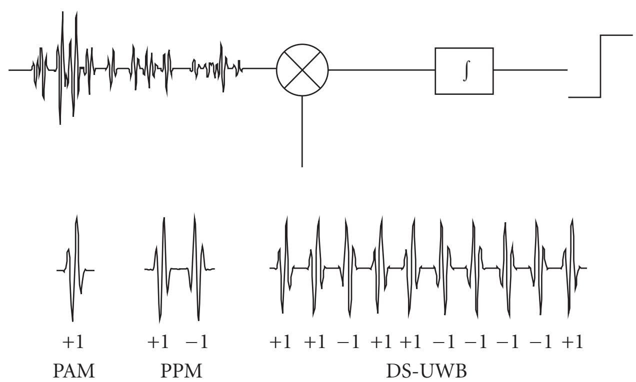

Figure 6: Matched filter correlation and detection.

(iii) direct-sequence code division multiplexing access (DS-UWB), whereby a series of repeated pulses is multiplied by a PN sequence of ± 1.

In the PAM and BPPM cases, it is assumed that the pulse repetition period is greater than the full-channel response duration. This does not apply to DS-UWB, where the total sequence temporal length may or may not exceed this duration. In all cases, detection is supposed to be operated at the output of an ideal correlator, achieving matched filtering between the received signal and a template (Figure 6). In other words, no equalization is performed. In the case of BPPM, the template is built from an ideal waveform together with its delayed replica multiplied by -1. However, a single pulse is transmitted, the decision being made according to the sign of the correlator output. In the other cases, the template is the time-reversed ideal received waveform. As will be seen below, there is hardly any difference regarding MIMO antenna gain for these various modulation schemes.

3.4. Antennas

In the simulations, two types of "generic" elementary radiators/sensors were considered (see Figure 7).

(i) An omnidirectional doublet-like elementary radiator, with a synthetized, real, frequency-independent radiation pattern (gain 1.7 dBi). Such an element was used as a reference radiator for SISO and SIMO schemes, and also within a circular array for SIMO and MIMO schemes. In this case, we assume that the radiating elements within the array are electromagnetically uncoupled [5].

(ii) A directional element for beam-switched SIMO or MIMO schemes, with the same elevation dependence as the doublet, and an azimuth dependence constructed by cubic interpolation over 5 points: the beam steering angle, the two half-power angles of the beam, and the beam boundaries where the gain cancels. The total beam width was chosen to be twice the half-power beam width in order to represent a realistic switched beam antenna, neglecting secondary lobes. We find a relative antenna gain comparing well with the ideal

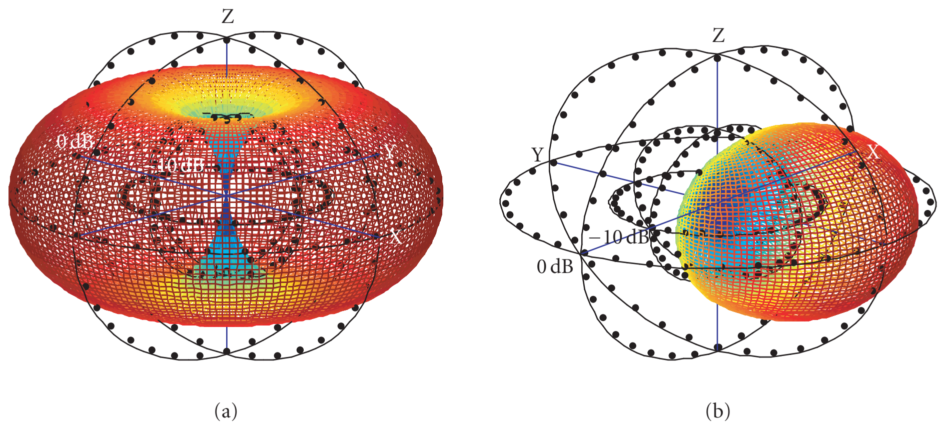

Figure 7: "Generic" sensors radiation patterns: (a) omnidirectional and (b) directional (90° half-power beam width for use in a 4-sensor circular array).

case of a perfect beam-switched antenna, that is, for which the extra gain offered by beam formation is equal to the number of beams (e.g., 6 dB higher than the omnidirectional radiator for a 4-beam antenna).

4. RESULTS

4.1. Simulation methodology

In all simulations below, the performance evaluation criterion is the SNR gain of the MIMO system with respect to the SISO antenna configuration for the same channel. The SNR gain is computed on the basis of the correlation magnitude squared, divided by the integrated noise power over the duration of the correlation signal acquisition. Each set of channels contains Monte Carlo random realizations, and for each of them, 4 positions of the Rx antenna array on corners of a 10 cm square have been defined in order to enlarge the size of the set ("small-scale" statistical averaging).

4.1.1. Simulation procedure for switched diversity

It is assumed that the transmitting (receiving) array is composed of $N_t(N_r)$ directional antenna elements, each covering a sector of width $360°/N_t$ $(360°/N_r)$. This yields a total number $N_t \times N_r$ of MIMO transmit/receive beam pattern combinations. For each channel realization, it is thus assumed that the best sliding correlation between the received signal and the ideal template over all combinations is selected in the switched diversity mode.

4.1.2. Simulation procedure for the time-domain combining

In the case of time-domain combining, the delay applied on each antenna branch has to be determined in order to maximize the SNR gain. Synchronization is carried out hypothetically as follows.

(i) The best sliding correlation is obtained on a reference radiator/sensor ($n_{t0} = 1$, $n_{r0} = 1$) pair (nominal receiver sending a training pulse sequence, nominal transmitter acquiring the synchronization).

(ii) Subsequently, the best correlations for the n_t/n_{r0} channels are acquired, that is, the nominal receiver sends an acquisition signal on n_{r0}, and each n_t antenna determines the best synchronization delay $\Delta\tau_{Tn_t}$ (phase 1)

(iii) Subsequently, the best correlation for the n_{t0}/n_r channels is acquired, that is, the nominal transmitter sends an acquisition signal on n_{t0}, and each n_r antenna determines the best synchronization delay $\Delta\tau_{Rn_r}$ (phase 2).

(iv) Finally, both the transmitter and the receiver synchronize the signals of their radiators/sensors by applying on each (n_r, n_t) radiator/sensor pair the optimal delays $(\Delta\tau_{Rn_r}, \Delta\tau_{Tn_t})$ given in Section 2.3; the signals are combined in the receiver, and the communication link is considered established.

We consider the following combining schemes.

(i) Simple resynchronization; that is, the transmitted and received signals in the transmitter/receiver are simply added (quoted below as "simple combining").

(ii) Resynchronization with polarity correction; in this case, the synchronization is acquired by searching for correlation maxima in absolute value, and before combining, the signal is corrected for polarity if necessary. This scheme corresponds in narrowband systems to "equal gain combining" (quoted below as "coherent combining").

(iii) The same scheme as above, with an MRC weighting factor multiplying each antenna signal before combining in the reception.

In all cases, a single RAKE finger will be used throughout the simulations.

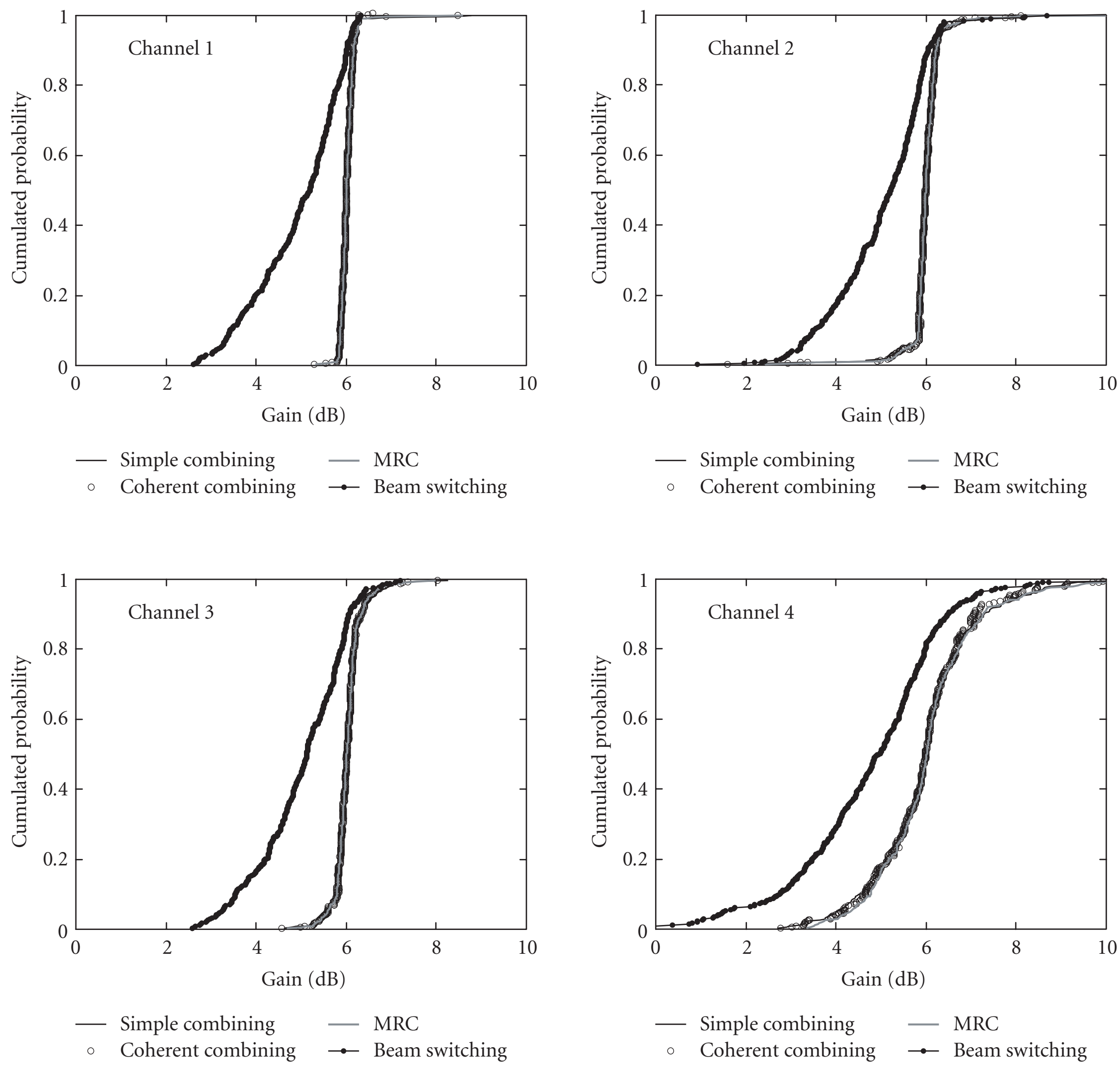

Figure 8: Performance for a 1×4 SIMO architecture. All three time-domain schemes yield superimposed curves, while beam switching consistently yields smaller gains.

4.2. *Ideal synchronization results*

In the present paragraph, ideal synchronization is assumed, that is, the maximum possible correlation between received signals and template for a continuously variable delay is achieved.

It clearly appears from Figure 8 that channels 1–3 behave similarly, with an SNR gain in combining schemes close to 6 dB for a 1×4 SIMO scheme. This implies that we are in the equivalent extreme UWB regime for all these channels. Only channel 4, which is highly dispersive in both the temporal and angular domains, behaves differently. The median SNR gain is also close to 6 dB; however, the cumulative distribution function (CDF) is much more broadly distributed with gains typically ranging from 3 to 10 dB whatever the time-domain combining scheme. This can be explained according to the discussion of Section 2.3. Due to the multipath density and dispersivity, a few interacting pulses can interfere destructively or constructively which can only be imperfectly mitigated by the combiner. This is the signature of fading, which although strongly reduced is not totally absent in dense multipaths channels. All time-domain combining schemes behave identically. This is expected for channels 1–3, less so for channel 4. Beam switching is naturally a less performant scheme due to the suboptimality of the discrete beam steering capability. MIMO results exhibit the same kind of behaviour, in agreement with Section 2.3. It can be seen that the suboptimality of the time-domain combining scheme increases with the channel dispersivity in the temporal and angular domains (Figure 9). In addition, it also increases with the number of transmitter/receiver antenna elements. This is due to the increased array size,

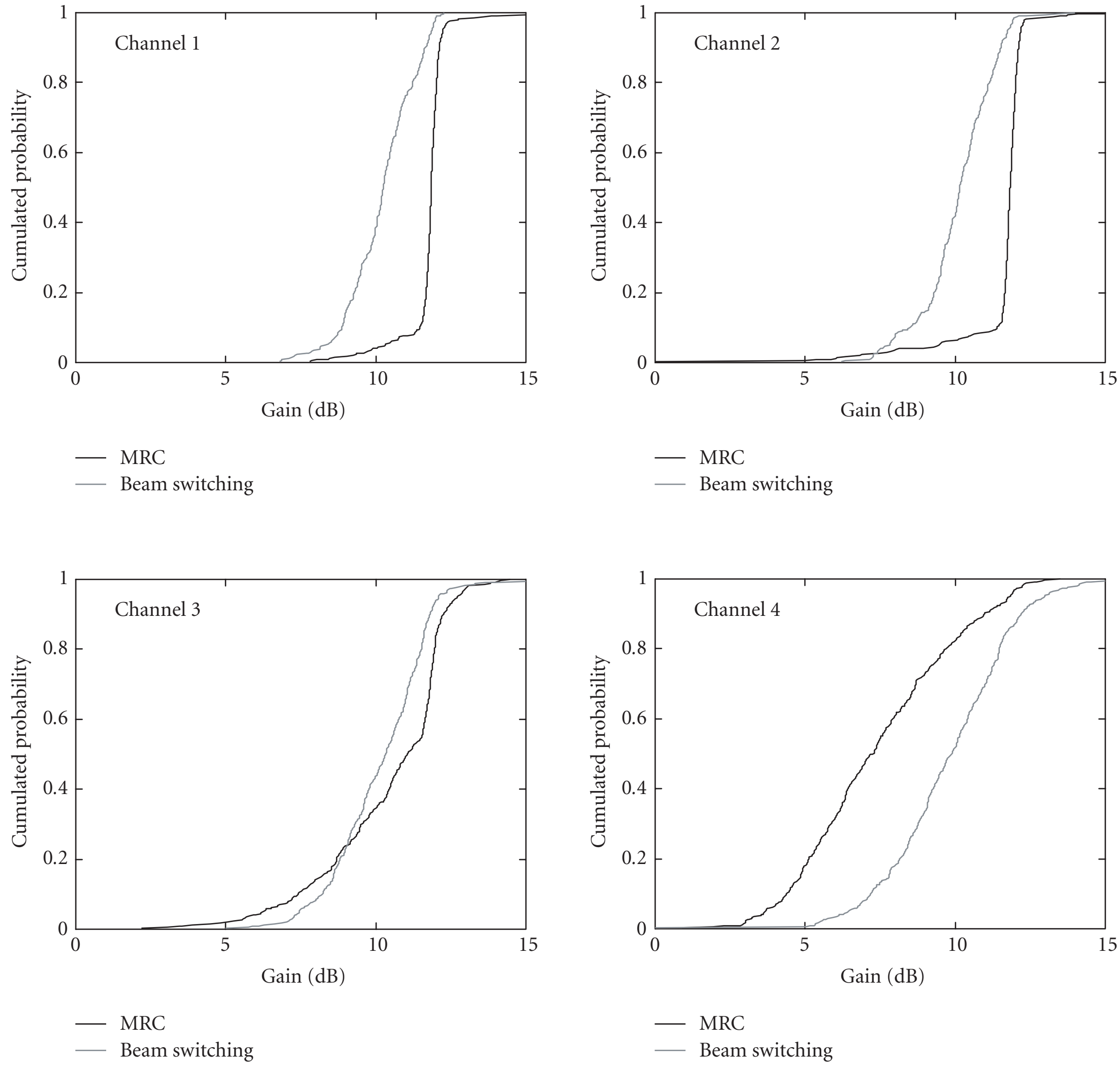

Figure 9: Performance for a 4×4 MIMO architecture.

which amplifies the synchronization problem for the overlapping multiple received pulses in the case of a dense multipath channel. For channel 4 and for 4×4 or 8×8 MIMO schemes, beam switching is thus more performant than MRC time-domain combining. By artificially reducing the array size to a very small value, the full MIMO gain can be recovered (Figure 10). Unfortunately, this is unfeasible in practice due to electromagnetic coupling. This poor performance of the combining scheme actually occurs when the DOA and DOD globally exhibit a large angular spread. However, in reality, it is likely that the angular correlations originated from the physics of propagation will involve *clustering*, whereby the echoes will be grouped either in delay or in angle, both at their departure and at their arrival. Since UWB detection is carried out in a narrow delay range, it is possible that much less angular dispersion will be involved in the detected signals than in the whole CIR. This can be easily simulated by creating clusters in the multipath channel, each of them being angularly narrow although the global channel is highly angularly dispersive. By comparing Figures 11 and 10, it can be seen that indeed there is a much better performance of the combining diversity performance of an 8×8 MIMO architecture for this clustered channel (Figure 12).

4.3. *Iterative and time-discrete synchronization*

The perfect synchronization scheme depicted in the last section allows to evaluate the ultimate time-domain combining performance for a given channel, but is not realistic in a practical communication system. The present section considers a somewhat more realistic case based on the following changes.

(i) Continuous delays are replaced by discrete delays, in order to mimic a clock-based timer.

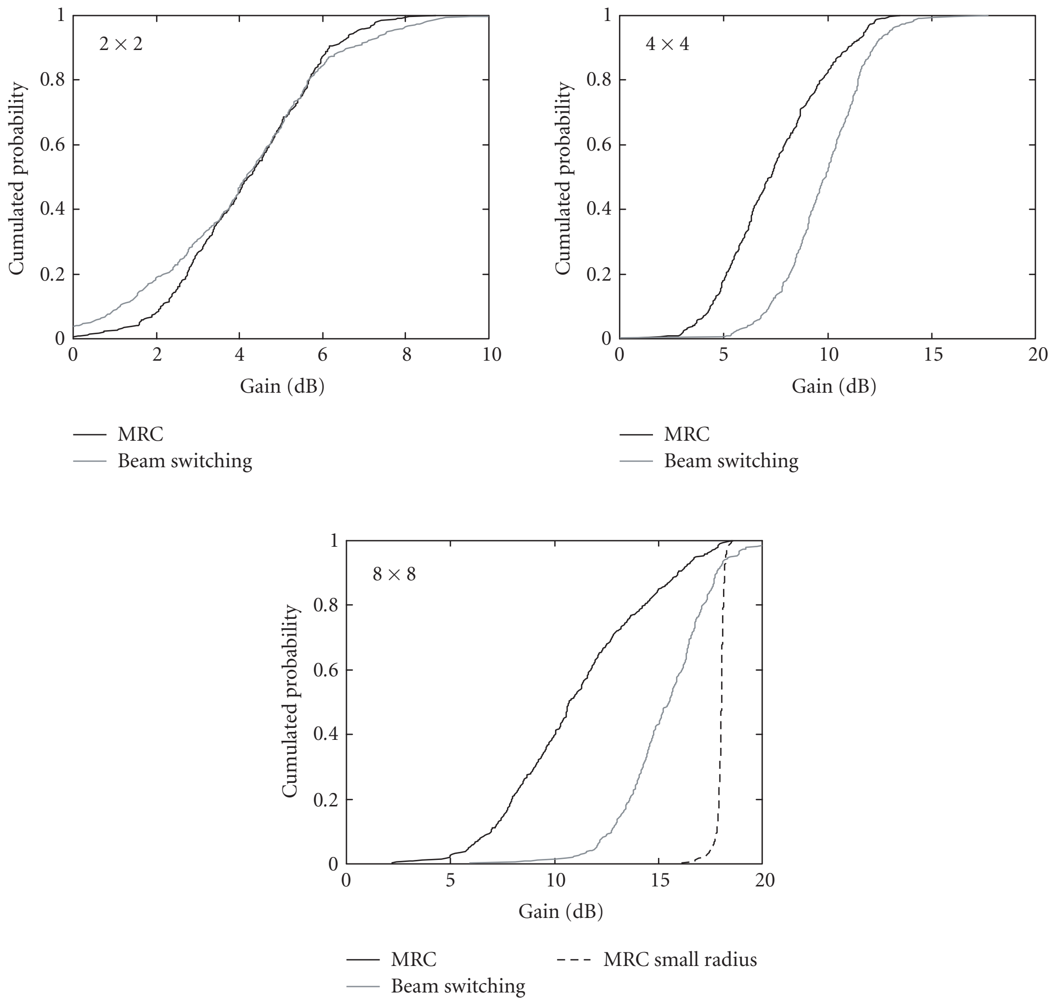

Figure 10: Performance of MIMO architectures of various array dimensions (channel 4).

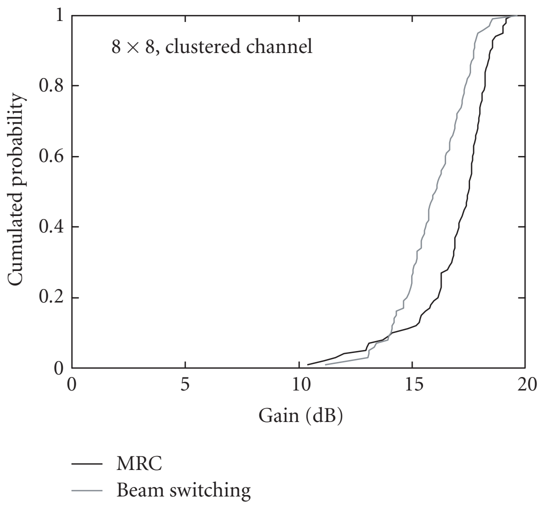

Figure 11: Comparison of a switched beam and a simple combining scheme for a *clustered* highly time and angle dispersive channel (8×8 MIMO architecture).

(ii) The synchronization of received signals on the succession of antenna elements in either phase 1 or phase 2 is done iteratively: the best synchronization is searched on the first element, then on the second, then on the third, and so forth (Figure 13). The algorithm loops in case further improvement is possible after a first round of synchronization on all elements. At each iteration, a single correlation between the template and a received sensor signal is computed.

It can be seen in Figure 14 that the performance degradation is essentially due to the inability of the algorithm to reach long applied delays. Only short delays are needed for time-domain combining in the extreme UWB regime, since they must compensate for propagation delay over the array size. However, in the dense multipath, long delays bring an additional delay diversity advantage which helps improve the SNR gain. The results of Figure 14 were obtained for only 9 discrete delays separated by increments of 0.08 nanoseconds, and required between a total of 20 to 35 iterations of the algorithm for a given channel realization.

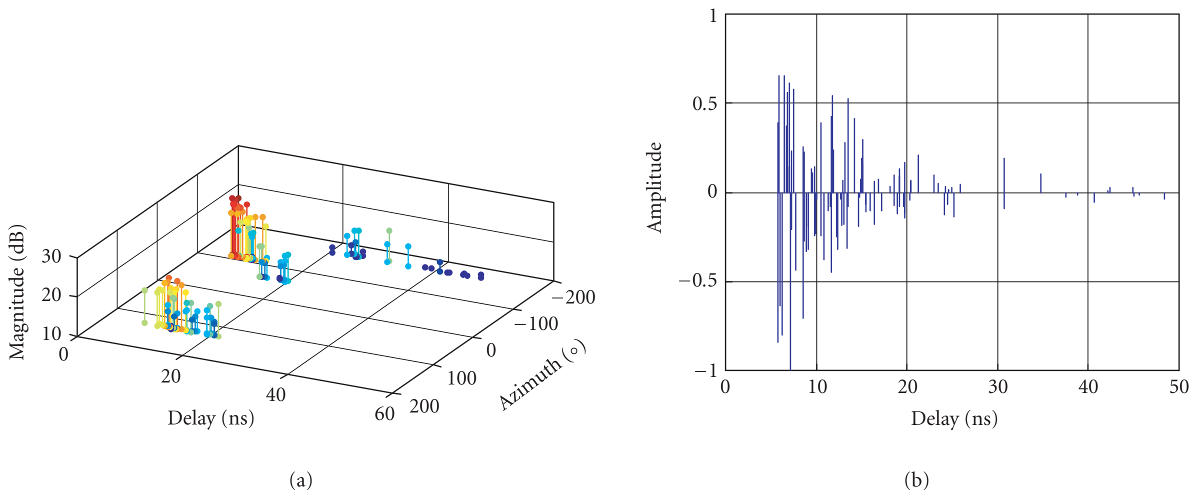

Figure 12: Example of a clustered UWB channel (4 clusters). (a) Delay-azimuth plot-relative power by pseudocolor. (b) Infinite BW amplitude impulse response.

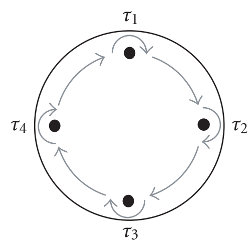

Figure 13: Principle of the iterative synchronization algorithm on a 4-sensor array.

4.4. Comparison of PAM, BPPM, and DS-UWB schemes

The comparison of the various modulation schemes has been performed for channel 4, with 0.5 nanoseconds time lag between pulses for BPPM (very small overlapping), and 20 pulses in the DS-UWB sequence. The plots of Figure 15 (time-discrete synchronization) show that there is no significant difference in the time-domain combining performance for all three schemes.

5. CONCLUSION

We have investigated and evaluated by simulations a few SIMO and MIMO diversity schemes which can improve the performance of a UWB radio link, in the hope to surmount the heavy constraints imposed by stringent regulations (low transmitted power). Using a stochastic channel model based on discrete multipaths and taking into account optimal waveforms intended to respect allowed spectral masks, we find that MIMO techniques may bring an improvement close to the $N_t \times N_r$ SNR gain for N_t transmitters and N_r receivers. This requires synchronizing and combining the signals emitted by the various radiators and received by the various sensors. However, the characteristics of the channel dramatically influence the effective SNR gain, and poor results for MIMO architectures are obtained for a highly time and angular dispersive channel. This is due to the intrinsic deficiency of the synchronization scheme for overlapping received pulses, whose delays are angularly dependent both at the transmitter and at the receiver (equation (6)). Angular clustering of the propagation channel is very much favourable from this point of view. Beam switching works fairly well, and might be an effective technique to obtain SNR gain for a moderate complexity cost. Being aware of the strong limitations imposed on UWB communications by regulation institutions, such a gain may be extremely valuable on the receiver to increase the link margin, or on the transmitter to reduce the total emitted power. Still these results rely on a rather simplistic channel model which should be tested against reality, in particular, as regards the importance of clusters in the angular and delay domains. Also the existence of diffuse scattering, known to contain an important fraction of the radiated energy, has not been accounted for and might complicate the analysis of UWB-MIMO effects. Another key issue is the performance of antennas, since dispersionless antennas are very difficult to design in UWB, especially under other constraints such as size, cost, and so forth.

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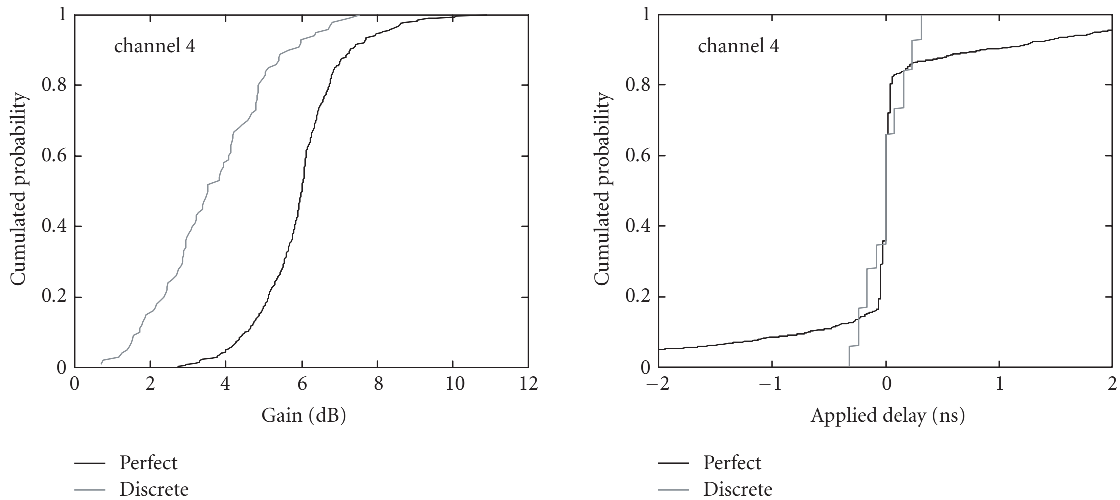

Figure 14: Performance of perfect versus time-discrete synchronization on a 1×4 SIMO array.

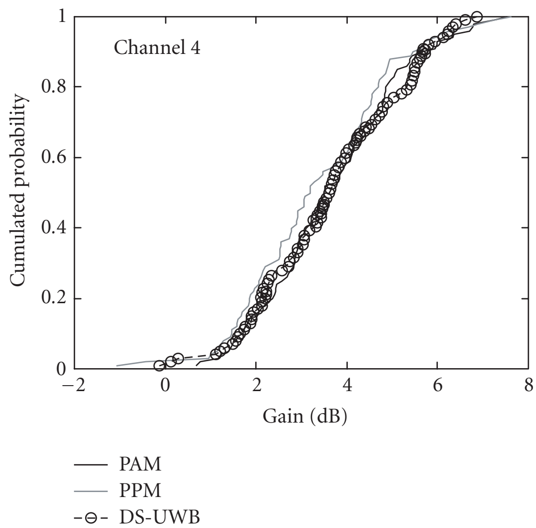

Figure 15: Performance of a 1×4 SIMO array (simple combining, time-discrete synchronization) for the various modulations investigated.

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