A MULTIACCESS PROTOCOL FOR HIGH-SPEED WLAN

Pierre A. Humblet[1] Serge Héthuin[2] Louis Ramel[2]

[1]Institut Eurecom
B.P. 193
06904 Sophia-Antipolis, France
and
[2]Thomson CSF - CNI
B.P. 82
92704 Colombes, France

Abstract: There is growing interest in very high speed local area networks capable of transporting ATM and Multimedia traffic, and very wide bands (100's MHz) are becoming available to that effect. This paper introduces and analyzes some characteristics of a protocol to access and utilize all the available bandwidth without putting undue strain on the capabilities of individual transceivers.

I. INTRODUCTION

The following protocol has been suggested [1, 2, 3] for use in high-speed Wireless Local Area Networks. Whenever they have a packet to transmit, nodes send a packet header at a given fixed frequency, followed by the body while the carrier frequency is ramping up, as illustrated in Figure 1.

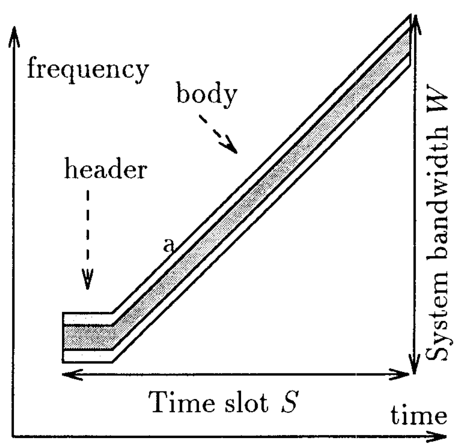

Figure 1: Time-Frequency transmission pattern

Under mild conditions discussed below, if bursts sent by two different transmitters do not overlap during their header, such as bursts b) and c) in Figure 2, then they can be transmitted simultaneously. On the other hand, bursts transmitted too close together, such as d) and e), collide. Thus many nodes can transmit simultaneously, and all nodes together can make full use of the system bandwidth, while their individual Baud rate is only a fraction of that bandwidth. This feature greatly simplifies the design of some proposed wideband systems, e.g. those designed for Asynchronous Transfer Mode traffic.

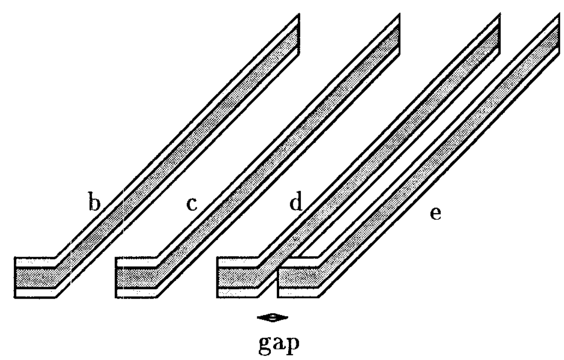

Figure 2: Cases of simultaneous transmission

To fix the reader's idea, we mention the values of the parameters originally suggested for use in the 17 GHz band. Baud rate $1/T$: 2.5 to 20 Mb/s, System bandwidth W: 200 MHz, Time slot length S: $= 600\ \mu s$. We will use these values in our numerical examples throughout the paper.

This proposed system can be compared to Frequency Division Multiplexing (FDMA), using many frequency subbands, with one subband serving as an access control channel, and the other subbands as traffic channels. A packet body can then either be entirely transmitted at one frequency, or alternatively the carrier frequency can be hopped among the various subbands. Compared to FDMA, the proposed approach has the advantage of easily allowing collision avoidance (only required on the signaling frequency) and of not requiring traffic channel assignments. The approach also provides frequency diversity, as does FDMA with frequency hopping, but without suffering from the dead times associated with synthesizer tuning. It transforms a static frequency selective fading into time-varying fading.

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Different aspects of this original approach merit to be studied, relating both to protocol performance issues, and to radio issues. This paper considers only the second aspect. In the next section we will consider some of the basic parameters of the system and model the baseband equivalent channel. This will be followed by a section on optimal baseband processing. Finally we will consider the effect of the frequency sweep on the error probability.

II. BASIC MODELLING ISSUES

This section consider two key issues. First, we study the required time separation to avoid collisions, and the resulting maximum number of simultaneous users. Next, we model the effect of using a swept carrier on a multipath channel.

A. Parameters to avoid interferences

We assume that two signals do not interfere if their carrier frequency are at least β/T apart. This ratio corresponds to the frequency separation between bands in a classical FDMA system, and is somewhat larger than the bandwidth of the receiver filter. The number of bands in a classical system is then given by $W/(\beta/T) = WT/\beta$. Taking into account that one band is used for control, the number of traffic channels in an FDMA system is

$$N_f = \frac{WT-\beta}{\beta}.$$

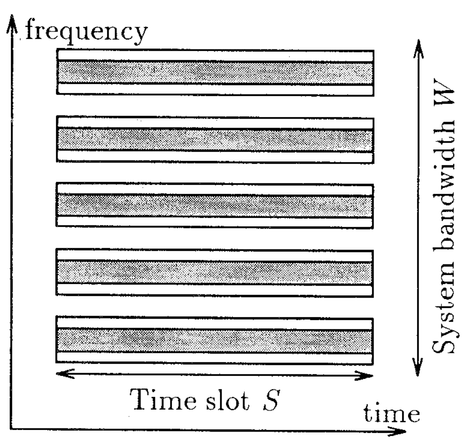

Figure 3: Classical FDMA system

For a system bandwidth $W = 200$ MHz, and for a conservative $\beta = 2$, $N_f = 39$, 9 or 4 for Baud rates of 2.5, 10 and 20 MBaud/s, respectively.

We now turn to the proposed system. We denote by α the fraction of a slot that is used by the fixed frequency header. The duration of a sweep is $(1-\alpha)S$ and during that time the carrier frequency varies by $W - \beta/T$. The speed of variation of the carrier frequency during the sweep is thus given by

$$f_s = \frac{W - \beta/T}{(1-\alpha)S} \qquad (1)$$

Assuming that bursts last about .5 ms and that the sweep is over 200 MHz, $f_s \approx 400$ GHz/s if the terms α and β/T are neglected.

As can be seen by considering the bursts d) and e) in figure 2, two transmissions do not collide if they are separated by at least a header length αS, plus a gap to allow the climbing ramp to "clear" the following header. The length of that gap is given by $(\beta/T)/f_s$, so that by using (1), the time separation t_s is

$$t_s = \alpha S + \frac{\beta}{Tf_s} = \frac{S(\beta - 2\alpha\beta + \alpha WT)}{WT-\beta}.$$

A new transmission of length S can occur every t_s, so the maximum number of simultaneous transmissions with a frequency sweep is given by

$$N_s = \frac{S}{t_s} = \frac{WT-\beta}{\beta - 2\alpha\beta + \alpha WT},$$

independently of S. We see that $N_s \leq N_f$, with equality if the header can be eliminated.

For example, if $\alpha = 1\%$, $N_s = 28$, 8 or 4 for Baud rates of 2.5, 10 and 20 MBaud/s, respectively. Comparing with the numbers given for FDMA, we see that the reduction is largest at low Baud rates. Note that non-interfering slots could be spaced more densely if the signal bandwidth was reduced at the beginning of the headers.

B. Baseband channel model

We now turn to deriving the baseband equivalent channel response of our system operating on a time invariant multipath channel.

We denote by $u(k)$ the information sequence. Each of the $u(k)$ can take one of M values $a, b, \ldots, M$ (usually $M = 2$). The modulated complex baseband signal is

$$x(t) = \sum_k g^{u(k)}(t - kT).$$

The $g^u(t)$'s are the pulse shapes associated with the modulation. For example, for antipodal modulation we would have $g^a(t) = -g^b(t)$ and these waveforms might be rectangular or raised-cosine pulses. More sophisticated modulations, such as those using many subcarriers, can also be considered and analyzed using our notation.

Assuming that the frequency sweep begins at time 0 and neglecting the transmission of the constant frequency header, the signal at the output of the transmitter is given by

$$x'(t) = \sqrt{2}\Re\left(\sum_k g^{u(k)}(t-kT)e^{j2\pi(f_c t + .5 f_s t^2)}\right)$$

where f_c is the carrier frequency at the beginning of the sweep, and f_s has been defined in (1).

The modulated signal is transmitted over a multipath channel with response

$$c'(\tau) = \sum_i c_i \delta(\tau - d_i).$$

In this model the channel is time invariant, a reasonable approximation for millisecond bursts in an indoor environment. The path coefficients c_i and the path delays d_i are independent of frequency. We do not know how accurate this assumption is, but we expect it to be good if enough paths are included in the model. For indoor operation, the delay spread extends to 10 or 100 ns.

The signal at the receiver is then

$$\begin{aligned} r'(t) &= c'(t) * x'(t) + n'(t) \\ &= \sqrt{2}\Re\Big(\sum_i \sum_k c_i g^{u(k)}(t - d_i - kT) \\ &\qquad e^{j2\pi\left(f_c(t-d_i)+.5f_s(t-d_i)^2\right)}\Big) + n'(t), \end{aligned}$$

where the noise $n'(t)$ is modeled as a Gaussian process with a two-sided spectral density $N_0/2$.

We assume that the front end of the receiver consists of a wideband filter and of a quadrature demodulator with an instantaneous frequency that mirrors that of the modulator. The output of the quadrature demodulator, after low pass filtering to eliminate double frequency terms, can be written in complex form as

$$\begin{aligned} r(t) &= \sum_{i,k} c_i g^{u(k)}(t - d_i - kT) e^{j2\pi\left(-(f_c+f_s t)d_i + .5 f_s d_i^2\right)} + n(t) \\ &= \int d\tau\, c(t,\tau) x(t-\tau) + n(t) \end{aligned}$$

where $n(t)$ is a complex symmetric Gaussian process with spectral density N_0 and

$$c(t,\tau) = \sum_i c_i e^{j2\pi\left(.5 f_s d_i^2 - f_c d_i - f_s d_i t\right)} \delta(\tau - d_i)\,.$$

We can interpret $r(t)$ as the complex baseband signal $x(t)$ at the transmitter passed through an equivalent baseband channel with a **time-varying** impulse response $c(t,\tau)$, and with noise added.

Although we have assumed that the underlying channel is time invariant, the combination of a sweeping carrier frequency and a delay spread yields a classical baseband equivalent multipath response with "Doppler" spread, where the word "Doppler" is used in a non-classical sense. The "Doppler frequency" associated with a path of delay d_i is given by $f_s d_i$.

If f_s = 400 GHz/s and d_i = 100 ns, the Doppler frequency is 40 kHz. For a modulation rate $1/T$ = 10 Mb/s, the product $f_s d_i T = .004$, which is typical for mobile systems, being intermediate between those of GSM and of IS-54 with vehicle speed of 250 km/h.

This Doppler will cause fluctuations in the channel. We will now examine the effect on the structure of the baseband receiver and on transmission performances.

III. BASEBAND PROCESSING

We first examine optimal baseband processing. In presence of additive white Gaussian noise, a Maximum Likelihood sequence detector will measure the channel response and determine the sequence $\hat{u}(k)$ that minimizes the Euclidean distance

$$\left\| r(t) - \sum_{i,k} c_i g^{\hat{u}(k)}(t - d_i - kT) e^{j2\pi\left(-(f_c+f_s t)d_i + .5 f_s d_i^2\right)} \right\|^2 .$$

Proceeding as in [4], this quantity can be expanded as a term that depends only on $r(t)$ minus

$$\begin{aligned} &\sum_k \Re\Big(\int dt \sum_i r(t+d_i) c_i^* e^{j2\pi\left(f_c d_i + .5 f_s d_i^2 + f_s d_i t\right)} \\ &\qquad g^{\hat{u}(k)^*}(t-kT) \\ &- \sum_{i,i'} \int dt\, g^{\hat{u}(k)}(t) g^{\hat{u}(k)^*}(t + d_i - d_{i'}) \\ &\qquad c_i c_{i'}^* e^{-j2\pi\left(d_i - d_{i'}\right)\left(f_c + .5 f_s\left(d_i - d_{i'}\right) + f_s kT + f_s t\right)} \\ &- 2 \sum_{k'<k} \sum_{i,i'} \int dt\, g^{\hat{u}(k)}(t) g^{\hat{u}(k')^*}(t + d_i - d_{i'} + (k-k')T) \\ &\qquad c_i c_{i'}^* e^{-j2\pi\left(d_i - d_{i'}\right)\left(f_c + .5 f_s\left(d_i - d_{i'}\right) + f_s kT + f_s t\right)}\Big) \end{aligned}$$

Usually a pulse $g^u(t)$ is only significant on an interval $[0, LT]$ where L is a small integer, and the integral in the second and third terms on the previous expression can be limited to that interval. Consequently in the sum above we only need to consider k' in the interval $[k-L, k-1]$ (assuming the differential delays $|d_i - d_{i'}| \le T$). Moreover, for the typical values given above, the terms $f_s\,(d_i - d_{i'})^2$ and $f_s t\,(d_i - d_{i'})$ can also be neglected. The receiver must thus find the sequence $\hat{u}(k)$ that maximizes

$$\begin{aligned} &\textstyle\sum_k \Re\Big(\int dt \sum_i r(t+d_i) c_i^* e^{j2\pi(f_c d_i + f_s d_i t)} g^{\hat{u}(k)^*}(t-kT) \\ &\quad - \textstyle\sum_{i,i'} c_i c_{i'}^* e^{-j2\pi(d_i - d_{i'})(f_c + f_s kT)} \Big(\rho_{\hat{u}(k),\hat{u}(k)}(d_i - d_{i'}) \\ &\qquad + 2 \textstyle\sum_{l=1}^{L} \rho_{\hat{u}(k),\hat{u}(k-l)}\,(d_i - d_{i'} + lT)\Big)\Big) \end{aligned}$$

where

$$\rho_{u,u'}\,(\tau) = \int dt\, g^u(t) g^{u'^*}\,(t+\tau)\,.$$

Only the first term depends on the received signal. As shown in Figure 4, such a receiver can be implemented as the concatenation of a slowly time-varying channel matched filter, which would be implemented digitally and would track the channel response, a bank of filters matched to the $g^u(t)$ and sampled every kT, and a time-varying (because of the presence of $f_s kT$) Viterbi equalizer with M^L states, the state at time k consisting of the $\hat{u}(k-l), l = 1, \ldots, L$. In a typical application $M = 2$, modulation is antipodal so that only one filter is required, and the memory L is a small integer that depends on the pulse shape, being 1 for rectangular pulses. Lack of space prevents us from interpreting this expression in special cases and to consider the role of the frequency sweep. We turn directly to the error probability analysis.

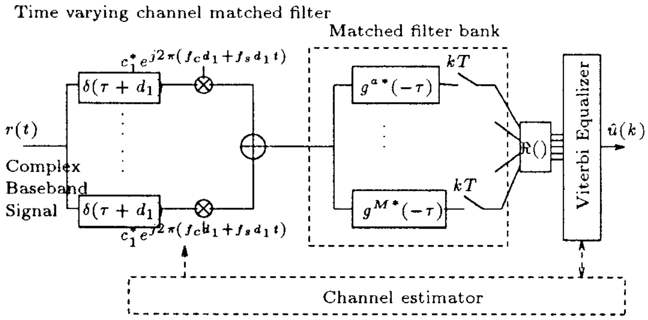

Figure 4: Diagram of an optimal receiver

IV. ERROR PROBABILITY ANALYSIS

Given that the sequence $u(k)$ is transmitted, an error occurs if the received signal $r(t)$ is closer to the signal produced by some $\hat{u}(k)$ than to the signal produced by $u(k)$. Conditioned on the channel characteristics, the probability of an error between $u(k)$ and $\hat{u}(k)$ is given by

$$Q\left(\sqrt{\frac{1}{2N_0}}\left\|\sum_k \int d\tau\, c(t,\tau)\left(g^{u(k)}(\tau)-g^{\hat{u}(k)}(\tau)\right)\right\|\right)$$

$$= Q\Big(\sqrt{\frac{1}{2N_0}}\Big(\sum_k \sum_{i,i'} c_i c_{i'}^* e^{-j2\pi(d_i-d_{i'})(f_c+f_s kT)}$$

$$\Big(R_{u(k),\hat{u}(k),u(k),\hat{u}(k)}(d_i-d_{i'})$$

$$+2\sum_{l=1}^{L} R_{u(k),\hat{u}(k),u(k-l),\hat{u}(k-l)}\left(d_i-d_{i'}+lT\right)\Big)\Big)^{.5}\Big)$$

where $Q(x)$ denotes the Gaussian complementary distribution function $Q(x)=\int_x^\infty du\, \exp(-u^2/2)/\sqrt{2\pi}$, and where

$$R_{u_1,u_2,u_1',u_2'}(\tau)=\int dt\,(g^{u_1}(t)-g^{u_2}(t))(g^{u_1'*}(t+\tau)-g^{u_2'*}(t+\tau)).$$

We interpret this expression first in the case of uncoded systems, and then with interleaving and coding.

A. Uncoded systems

The most likely errors are those with a single symbol error in position m, we consider them for simplicity. As $\hat{u}(k)=u(k)$ for all $k\neq m$, all $R_{u_1(k),u_2(k),u_1'(k-l),u_2'(k-l)}$ in the previous expression are 0, except one for the $k=m$ where the error occurs, with $l=0$. To simplify the discussion we consider only binary antipodal modulation with pulse shape $g(t)$. Then if $u(m)\neq\hat{u}(m)$, $R_{u(m),\hat{u}(m),u(m),\hat{u}(m)}(\tau)=4\int dt\, g(t)g^*(t+\tau)=4\rho_g(\tau)$. $\rho_g(0)$ represents the pulse energy denoted $\mathcal{E}$. We will focus on this particularly simple case, and consider two extreme forms of modulations.

If $g(t)$ is a very wideband direct sequence spread-spectrum sequence (which makes little sense in the context considered here), then $\rho_g(d_i-d_{i'})=0$ if $i\neq i'$, and the error probability simplifies to

$$Q\left(\sqrt{\frac{2\mathcal{E}}{N_0}\sum_i |c_i|^2}\right).$$

The contributions of all paths add coherently, minimizing the effects of fading. The frequency sweep has absolutely no effects on the performance.

Consider now the other extreme case case where $g(t)$ is a rectangular pulse of width T, and where the Baud time T is long compared to the delay spread. We can then admit that $\rho_g(d_i-d_{i'})\approx\rho_g(0)$ (this is essentially a "flat fading assumption", but only over the modulation bandwidth of the signal) and the error probability becomes

$$Q\left(\sqrt{\frac{2\mathcal{E}}{N_0}\left|\sum_i c_i \exp\left(-j2\pi d_i\left(f_c+f_s mT\right)\right)\right|^2}\right). \quad (2)$$

The error probability depends on m because the signal to noise ratio is multiplied by a factor containing m. The position of the components of $\sum_i c_i \exp\left(-j2\pi d_i\left(f_c+f_s mT\right)\right)$ are represented in the top part of figure 5.

As the index m varies over the symbols in a slot, the terms $f_s mT$ varies from 0 to the system bandwidth W and the phase $2\pi d_i f_s mT$ will grow to $2\pi d_i W$, which can be large for the typical values given above. Note that it is independent of the Baud rate and of the slot size.

As m varies the vectors in figure 5 rotate with different angular speeds. If $2\pi(d_i-d_j)W$ is large, the vectors i and j will be aligned many times during a sweep, and it is thus likely that at some point during the sweep the phases of the factors multiplying the c_i's in (2) will be such that their contributions combine in the most unfavorable fashion, multiplying the signal to noise ratio by $\left|\sum_i \pm|c_i|\right|^2$, where the $\pm$ are chosen to minimize the sum. This is illustrated in the bottom part of figure 5. We can analyze the situation for some typical cases.

In many wireless LANs, a strong direct path exists. The worst case with the $\pm$ above is when the direct path is $+$, and all other paths $-$. For the system to perform well, it is necessary that the **magnitude** c_1 of the direct path be larger than the sum of the magnitudes of all the other paths. This is to be contrasted with the more favorable situation of Rician fading, which does not have a very negative effect as long as the **power** $|c_1|^2$ of the direct path is larger than the sum of the powers of all the other paths.

To quantify that statement, suppose there is only a direct path with power $|c_1|^2=\alpha^2$ and a secondary path where $|c_2|$ has a Rayleigh distribution (depending on the antenna position) with unit average power.

An outage occurs if the combination of the paths gives rise to a signal with power less than β^2, i.e. to a signal lying in the circle centered at the origin in the top part of figure 5. At an arbitrary frequency, the combination of the two paths give rise to Rician statistics and the outage

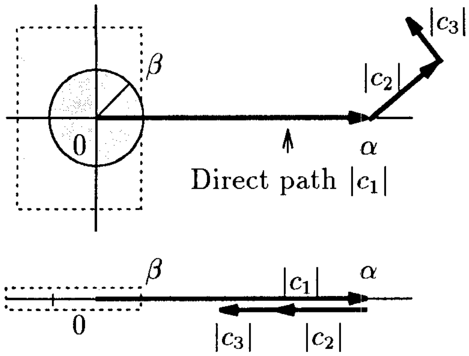

Figure 5: Illustration of fading as a function of frequency

probability P_1 is given by

$$P_1 = 1 - Q_1\left(\sqrt{2}\alpha, \sqrt{2}\beta\right) \quad (3)$$
$$< Q\left(\sqrt{2}(\alpha - \beta)\right) \quad (4)$$
$$< .5\exp\left(-(\alpha-\beta)^2\right),$$

where Q_1 denotes the Marcum Q function [5]. The upper-bound (4) is simply the probability of lying in the dotted half plane.

An outage occurs at some point during the sweep if $|c_2|$ exceeds $|c_1| - \beta$ but not $|c_1| + \beta$. $|c_2|^2$ has a unit mean exponential distribution, thus the probability P_a of an outage occurring at some point is given by

$$P_a = \exp\left(-(\alpha-\beta)^2\right) - \exp\left(-(\alpha+\beta)^2\right)$$
$$\approx \exp\left(-(\alpha-\beta)^2\right). \quad (5)$$

where (5) is the probability of lying in the left half space.

The expression (3) appears in dotted line in figure 6 for $\beta = 1$, while expression (5) appears in dashes. The horizontal axis represents the power α^2 of the direct path, normalized to the expected power of the secondary path.

The solid lines in figure 6 represent curves (obtained by numerical convolution) when there are 2, 3, or 4 secondary paths, all with independent Rayleigh statistics, and with a **total** average power of unity (the situation would be much worse if they had individual unit average powers). Expression (3) remains valid for that case.

We see that in presence of a strong direct path, the frequency sweep causes relatively little degradation in error probability, specially if the number of secondary paths is small. This situation can be obtained by using directional antennas.

We should point out that we have used Rayleigh statistics for the secondary paths only for lack of precise knowledge about the actual distribution. We do not expect our conclusions to be strongly affected by that statistics.

On the other hand, it should be clear that in absence of a dominant path the frequency response will exhibit deep fades, and that dismal performances are likely to be met at some point during the sweep.

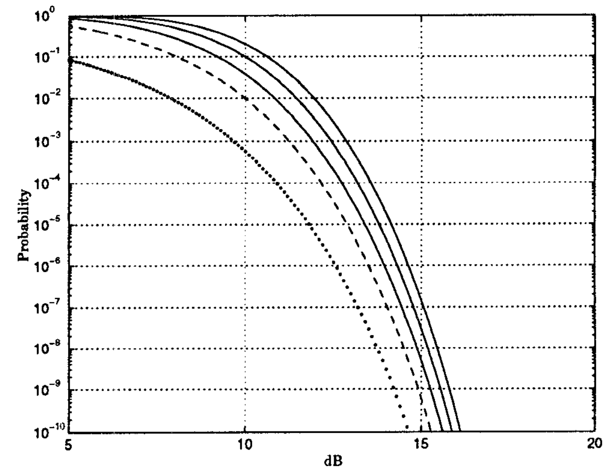

Figure 6: Outage probabilities without and with frequency sweep. $\beta = 1$.

B. Coded systems

One of the strengths of the proposed system is the ease with which diversity can be obtained in a single slot, due to the presence of the frequency sweep. This diversity can be exploited by interleaving and coding. The coherence bandwidth of the channel is about $1/d$, where d is the delay spread [5]. Thus the interleaving depth should be larger than $1/df_s \approx S/dW$, and the degree of diversity is about dW, i.e. 4 for $W = 200$ MHz and $d = 20$ ns. This can already be well exploited by a rate 1/2 convolutional code with 4 states.

This system may permit reliable operation in situations with a large delay spread and without direct path, where classical FDMA with Automatic Repeat reQuest would suffer from retransmission delays. Further studies will consider this possibility and the choice of modulation.

References

[1] L. Ramel, "Procédé de transmission par paquets et émetteur/récepteur mettant en oeuvre ce procédé." French Patent, Dec. 1993.

[2] L. Ramel, "Technical contribution," tech. rep., ETSI Res 10, Sophia-Antipolis, France, May 1995.

[3] L. Ramel and S. Héthuin, "Protocole de communication à étalement de spectre original et réaliste," in *CESP '95*, (Grenoble), Nov. 1995.

[4] A. J. Viterbi and J. K. Omura, *Principles of Digital Communication and Coding.* New York: Mc-Graw-Hill, 1979.

[5] J. G. Proakis, *Digital Communications.* New York: McGraw-Hill, 1983.