Buletinul Ştiinţific al Universităţii "Politehnica" din Timişoara

Seria ELECTRONICĂ şi TELECOMUNICAŢII
TRANSACTIONS on ELECTRONICS and COMMUNICATIONS

Tom 49(63), Fascicola 2, 2004

UWB communications systems based on orthogonal waveforms set

Florin Crăciun[1], Cosmin Mateescu[1], Octavian Fratu[1], Simona Halunga [1]

Abstract : Ultra wideband (UWB) signals are a new form of time hopping spread spectrum (TH-SS) signals with pulse position modulation (PPM). Typically an UWB is defined as any signal in which the 3 dB bandwidth of the signal is at least 25 percent of its center frequency. UWB signals are using previously allocated RF bands, by hiding signals under the noise floor.
Keywords: spread spectrum, time hopping, ultra-wideband modulation,

I.INTRODUCTION

Recent development in the area of wireless communications systems indicates that ultra wideband (UWB) technology is an attractive solution for short-range multiple-access communications due to a number of attractive characteristics.

The UWB signal is obtained using the impulse radio technique, obtained by combining the Pulse Position Modulation (PPM) with Time Hopping Spread Spectrum (TH-SS) technology. Impulse radio communicates using baseband impulses of very short duration, typically on the order of a nanosecond, thereby spreading the energy of the radio signal very thinly from near d.c to a few gigahertz. Even when those pulses are applied to appropriate designed antennas, they propagates with distortion. The antennas behave as filters, and, even when propagation occurs over free space, a differentiation of the impulse is produces when the wave is radiated. In this paper, the combined effects of the channel and antenna are modeled as a differentiation operation. Hence, the received pulse is the derivate of the transmitted pulse.

Considering that transmitted pulse is $\omega_{tx}(t) = \int_{-t}^{\infty}\omega(\zeta)d\zeta$, the received pulse will be $A\omega(t-\tau)+n(t)$, where the constant A and τ are the attenuation and, respectively, propagation delay experienced by the signal. The noise $n(t)$ is modeled as $AWGN$ with two-sided power density $N_0/2$ Watts/Hz.

In this paper we consider that a UWB pulse is modulated by the second derivate of a Gaussian function $\exp\left(-2\pi\left[\frac{t}{t_n}\right]^2\right)$ properly scaled. In this case the transmitted pulse is:

$$\omega_{tx}(t) = t\exp\left(-2\pi\left[\frac{t}{t_n}\right]^2\right) \quad (1)$$

and the received pulse is :

$$\omega = \left[1-4\pi\left[\frac{t}{t_n}\right]^2\right]\exp\left(-2\pi\left[\frac{t}{t_n}\right]^2\right) \quad (2)$$

where $t_n = 0.4472$ was a optimum value for a measured waveform. Using those values, the pulse duration is $T_\omega = 2$ ns. The normalized correlation function of the impulse $\omega(t)$ is determined by :

$$\gamma_\omega(\tau) = \frac{1}{E_\omega}\int_{-\infty}^{\infty}\omega(t)\omega(t-\tau)d\tau > -1 \ \forall\tau \quad (3)$$

where:

[1] University POLITEHNICA of Bucharest,
Electronics, Telecommunications and Information Theory Faculty,
Telecommunications Department,
1-3 Iuliu Maniu Blvd, 061071,Bucharest 6, Romania
Phone: (4021) 402 4996; Fax: (4021) 410 2379
ofratu@elcom.pub.ro, shalunga@elcom.pub.ro

$$E_\omega = \int_{-\infty}^{+\infty} |\omega(\xi)|^2 d\xi \qquad (4)$$

is the energy of the signal.

The final relation is given by:

$$\gamma_\omega(\tau) = \left[1 - 4\pi\left[\frac{\tau}{t_n}\right]^2 + \frac{4\pi^2}{3}\left[\frac{\tau}{t_n}\right]^4\right]\exp\left(-\pi\left[\frac{\tau}{t_n}\right]^2\right) \qquad (5)$$

The bandwidth for $\omega(t)$ is determined by :

$$F_\omega(f) = \sqrt{2}t_n\left[\frac{\pi(ft_n)^2}{2}\right]\exp\left(-\pi\frac{(ft_n)^2}{2}\right) \qquad (6)$$

which has a maximum at $f = \frac{1}{t_n}\sqrt{\frac{2}{\pi}} = 1.7842$ GHz .

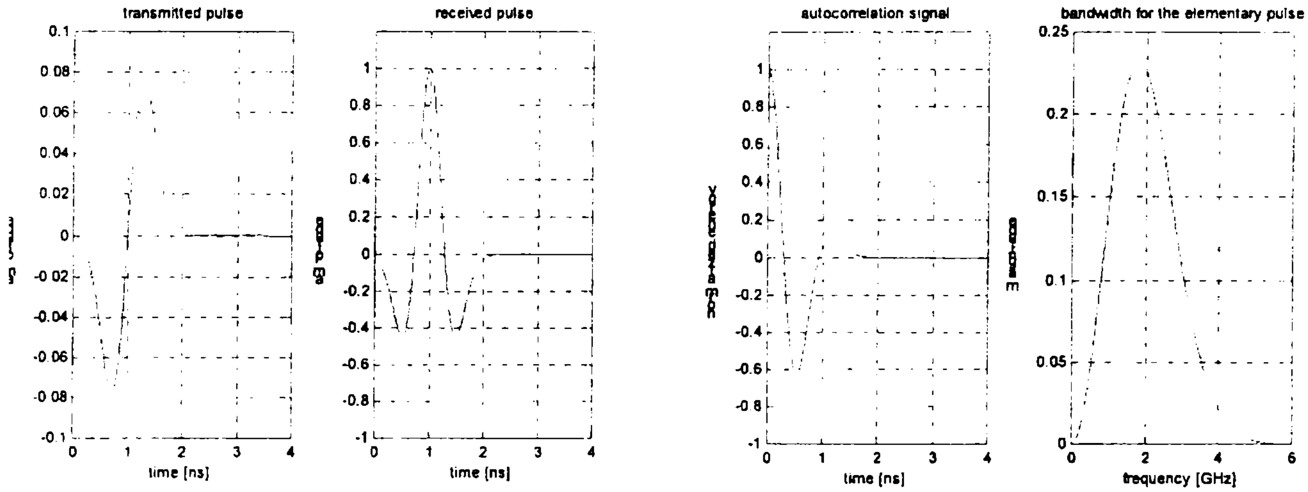

Figure 1. The waveforms for transmitted pulse, received pulse, autocorrelation signal and the bandwidth for the elementary pulse

II. TIME HOPPING SIGNALS

When operates is a densely populated radio environment, the impulse radio interferes with other narrowband radio signals that co-exists in the same frequency range. The design of the UWB signal has to be properly done, such that those narrowband signals will be affected as little as possible. This requirement impose the use of spread spectrum technique.

The simplest method to spread the spectrum of those ultra wide bandwidth low-duty-cycle pulse trains is time hopping, with data modulation at the rate of many pulses per data symbol.

A typical hopping format with pulse position data modulation (PPM) is given by:

$$x^{(\upsilon)}(t) = \sum_{k=0}^{\infty} \omega(t - kT_f - c_k^{(\upsilon)}T_c - \delta^k_{d^{(\upsilon)}_{[k/N_s]}}) \qquad (7)$$

where:

- the superscript υ represent a particular user;
- T_f is the frame interval;
- $c_k^{(\upsilon)}$ is the spreading code used by the υ th user;
- T_c is the spreading code chip interval
- $\delta^k_{d^{(\upsilon)}_{[k/N_s]}}$ is the υ th user data sequence.

The signal transmitted by the υ-th user consists of a large number of monocycle waveforms shifted to different time instants; the transmitted pulse $\omega_{tx}(t)$ is referred to as a monocycle.

A pulse train of the form $\sum_{k=0}^{\infty} \omega(t - kT_f)$ consists of monocycle pulses spaced T_f seconds apart in time. The frame time or pulse repetition time T_f typically may be a hundred to a thousand time the monocycle wide. The result is a signal with a very low duty cycle.

In figure 2 is illustrated an uniform spaced monocycle pulse train and its power spectrum. It can be easity seen that the frequency response of this equally spaced pulse trains include both continuous and discrete spectral lines at regular intervals, so multiple-access signals composed of uniform spaced pulses are vulnerable to occasional catastrophic collisions in which a large numbers of pulses from two signals are received at the same time instant.

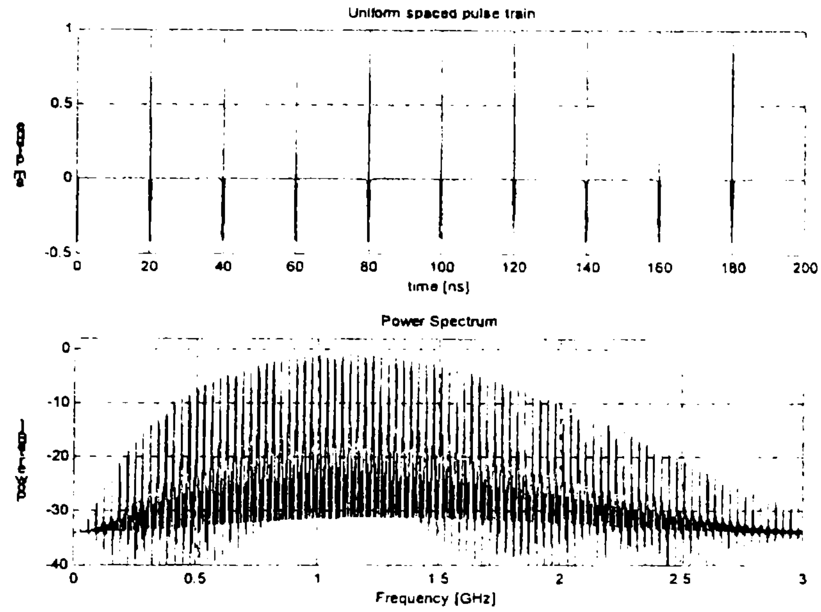

Figure 2. Uniform spaced pulse trains and its power spectrum

To eliminate the catastrophic collisions that may occur in multiple access systems, each link, (indexed by υ) is assigned to a distinct pulse shift pastern $\{c_k^{(\upsilon)}\}$ which is referred refer to as a time hopping code. These pseudorandom codes are periodic with the period N_p, i.e., $c_{k+iN_p}^{(\upsilon)} = c_k^{(\upsilon)}$, $(\forall)i, k$. Each code element is an integer in the range $0 \le c_k^{(\upsilon)} < N_h$. The time hopping code provides therefore an additional time shift to each pulse, in the pulse trains with the *k-th* monocycle undergoing an added shift of $c_k^{(\upsilon)}T_c$. The added time shifts caused by code are discrete times between 0 and $N_h T_c$ seconds.

We further assume that $N_h T_c \le T_f$ and hence the ratio $N_h T_c / T_f$ indicates the fraction of the frame time T_f over which time-hopping is allowed. If $N_h T_c$ is to small, then catastrophic collision remain a significant possibility. Conversely, with a large enough value of $N_h T_c$ and well designed codes, the multiple-access interference in many situation can be modeled as a Gaussian random process.

Because the hopping code is periodic with period N_p, the waveform $\sum_{k=0}^{\infty} \omega(t - kT_f - c_k^{(\upsilon)}T_c)$ is periodic also, with the period $T_p = N_p T_f$.

One effect of the hopping code is that it reduces the power spectral density from the line spectral density ($1/T_f$ apart) of uniformly spaced pulse train with finer line spacing $1/T_p$ apart, as we can see in figure 3.

Comparing figures 2 and 3, one can observe that, when the bearer impulses are not randomized the power spectrum is dominated by spectral lines whereas, when the randomization codes are used, it reduces the spectrum lines, and the power spectrum is predominately continuous.

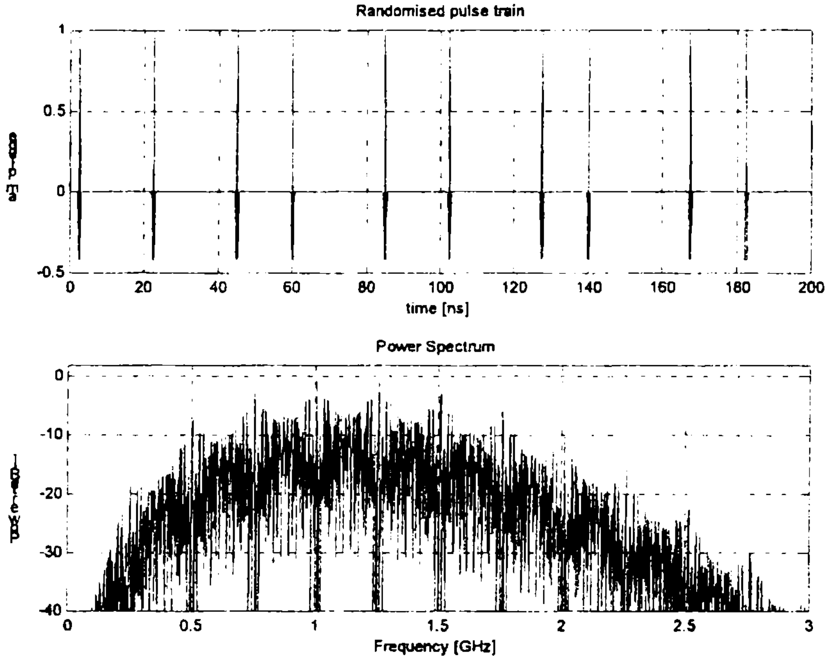

Figure 3. Randomised pulse train and its power spectrum

The data sequence $\{d_l^{(\upsilon)}\}$ of user υ is an M-ary ($1 \le d_m^{(\upsilon)} \le M$) symbol stream that convey the information offered by the in some form to digital data. This information is transported over the radio channel using the above pulse train, by modifying the pulse positions. It should be noted that the pulse time delay introduced due to modulation is relatively small compared with the time delay resulting from pulse spreading with the spreading code (pulse randomization), so the effects of pulse position modulation on the power spectrum are insignificant.

The received signal can be modeled as:

$$r(t) = \sum_{\upsilon=1}^{N_u} A^{(\upsilon)} x^{(\upsilon)}(t - \tau^{(\upsilon)}) + n(t) \quad (8)$$

where

- $A^{(\upsilon)}$ is the attenuation of user υ's signal over the radio channel,
- $\tau^{(\upsilon)}$ represents time asynchronism between the clocks of the user υ's transmitter and receiver, and
- *n(t)* represents non-multiple-access interference modeled as AWGN.

If the receiver wants to demodulate a particular user signal (let's say user 1), representing the m-th data symbol $d_m^{(1)}$, where $d_m^{(1)}$ is one of M equally-likely symbols, then the received signal $r(t)$ is:

$$r(t) = A^{(1)} X^{(1)}_{m,d_m^{(1)}}(t - \tau^{(1)}) + n_{tot}(t) \ , t \in \tau_{\mathrm{m}} \quad (9)$$

where

$$\tau_m = [mN_s T_f + \tau^{(1)}, (m+1)N_s T_f + \tau^{(1)}] \quad (10)$$

and

$$n_{tot}(t) = \sum_{\upsilon=2}^{N_u} A^{(\upsilon)} x^{(\upsilon)}(t - \tau^{(\upsilon)}) + n(t) \quad (11)$$

When the receiver is perfectly synchronized to the first user signal (e.g. having learned the value of $\tau^{(1)}$), the receiver is able to determine the sequence $\{\tau_m\}$ of time intervals, with interval τ_m containing the waveform representing data symbol $d_m^{(1)}$. In this case the detection problem reduces to coherent detection of M equally-energy, equally-likely signals in the presence of multiple–access interference in addition with AWGN, and therefore the optimal receiver is a complicated structure that takes advantage of all of the receiver's knowledge regarding the characteristics of multiple-access interference.

Due to the complexity of the analysis, the multi-user detector will not be considered here. Instead we will assume that $n_{tot}(t)$ is a zero-mean Gaussian random process. Hence, the detection problem becomes coherent detection of M equal-energy, equally-like signals in presence of a mean-zero Gaussian interference in addition to AWGN.

In this paper we consider data been carried by orthogonal signals.

III. ORTHOGONAL SIGNALS

Orthogonal signals (OR) represents a particular case of PPM TH-SS signals, for which the data is given by :

$$d_i^k = [(k+i-1)\bmod M]T_{OR} \tag{12}$$

The construction of orthogonal signals is therefore given by :

$$S_i(t) = \sum_{k=0}^{N_s-1} \omega\left(t - kT_f - [(k+i-1)\bmod M]T_{OR}\right),\ i = 1,2,\ldots,M \tag{13}$$

where $T_{OR} > T_\omega$. In this paper a simulation has been performed for $N_u = 1$, and therefore the time hopping sequence $\{c_k^{(\upsilon)}\}$ and the delay $\tau^{(1)}$ have no effect in the correlation properties of the PPM signals, and they were omitted in this analysis.

For the OR PPM signals the normalized correlation coefficients are given by :

$$\alpha_{i,j} = \begin{cases} 1, & \text{if } i = j \\ 0, & \text{if } i \neq j \end{cases} \tag{14}$$

hence the normalized correlation matrix A_{OR} is a $M \times M$ identity matrix.

For a fixed impulse waveform $\omega(t)$, N_s and T_f, the orthogonal signal depends only on T_{OR}. In the presence if AWGN any $T_{OR} > T_\omega$ will perform identically, so we choose $T_{OR} = 2T_\omega$.

To detect all the M signals we will need to correlate the input signal with all the M reference signals. In order to design the receiver for the OR PPM signals we consider:

$$x(t) = S_j(t-\tau^{(1)} - C_0^1(t-\tau^{(1)})) + n(t) \tag{15}$$

where S_j is one of the signals in (14),

$$C_m^{(\upsilon)}(t) = \sum_{k=mN_s}^{(m+1)N_s-1} T_c c_k^{(\upsilon)} p(t-kT_f) \tag{16}$$

and *n(t)* is AWGN. In this case, each of the M channel correlation output can be written :

$$\begin{aligned} y_i &= \int_0^{N_sT_f} x(t) S_i(t-\tau^{(1)} - C_0^1(t-\tau^{(1)}))dt \\ &= \sum_{k=0}^{Ns-1} \sum_{q=0}^{M-1} \delta_{q,[(k+i-1)\bmod M]} z(k,q) \end{aligned} \tag{17}$$

where:

$$z(k,q) = \int_{kT_f+\tau^{(1)}+c_k^{(1)}T_c+qT_{OR}}^{kT_f+\tau^{(1)}+c_k^{(1)}T_c+(q+1)T_{OR}} x(t)\omega(t-kT_f-\tau^{(1)}-c_k^{(1)}T_c-qT_{OR})dt \tag{18}$$

and $\delta_{q,q'}$ is the Kronecker delta. From the expression for y_i, i = 1,2,...M it is clear that the receiver neerds only one corelator and M store and sum circuits. This is illustrated in figure 4.

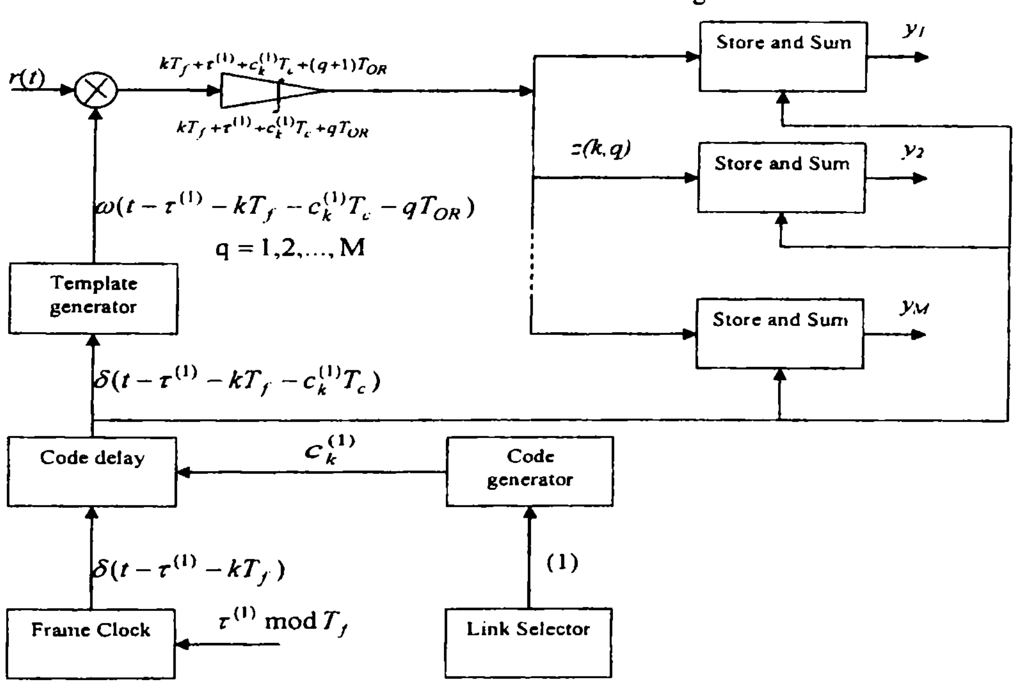

Figure 4. Receiver block diagram for the reception of the first user's signals

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is the energy of the signal.

The final relation is given by:

$$\gamma_\omega(\tau) = \left[1 - 4\pi\left[\frac{\tau}{t_n}\right]^2 + \frac{4\pi^2}{3}\left[\frac{\tau}{t_n}\right]^4\right]\exp\left(-\pi\left[\frac{\tau}{t_n}\right]^2\right) \qquad (5)$$

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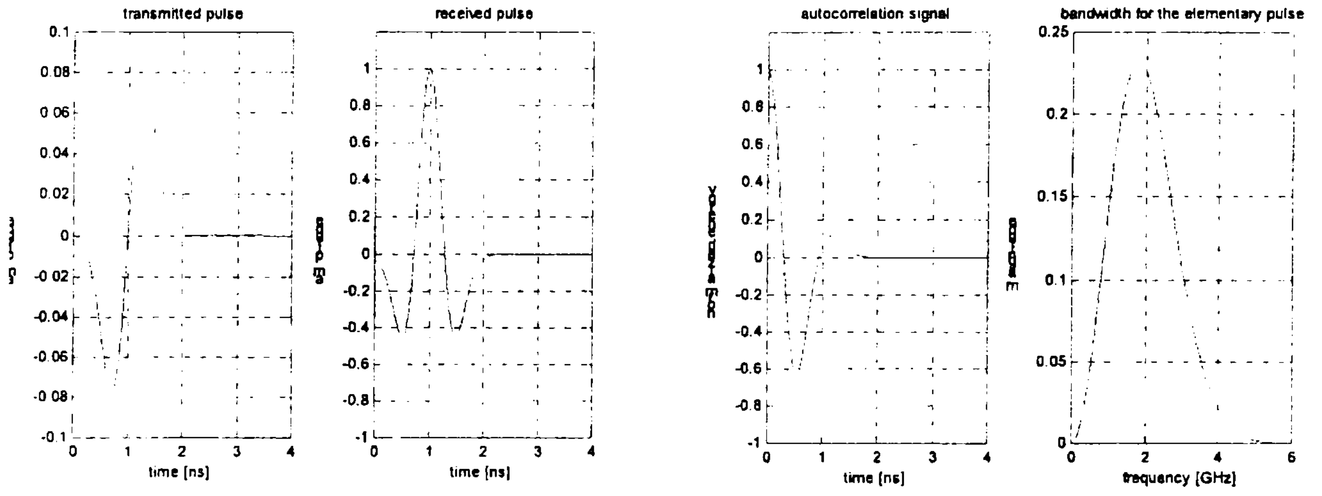

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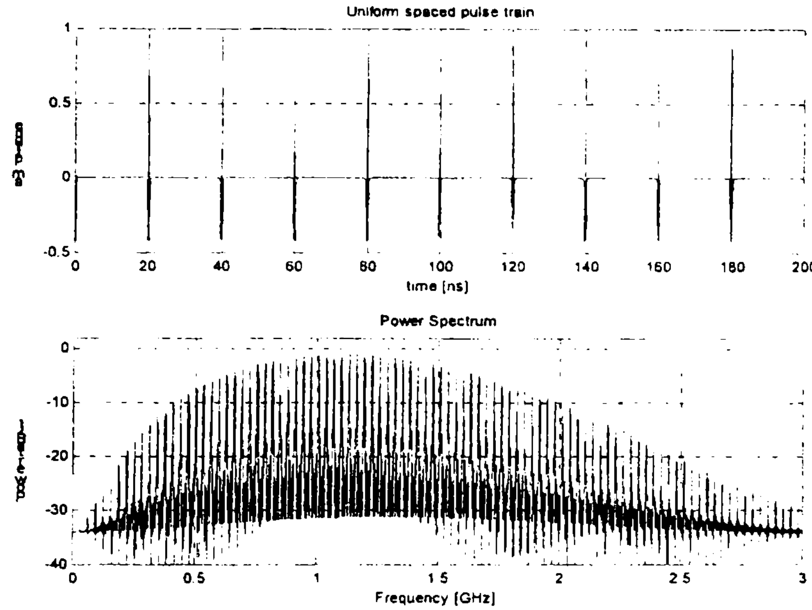

Figure 2. Uniform spaced pulse trains and its power spectrum

To eliminate the catastrophic collisions that may occur in multiple access systems, each link, (indexed by υ) is assigned to a distinct pulse shift pastern $\left\{c_k^{(\upsilon)}\right\}$ which is referred refer to as a time hopping code. These pseudorandom codes are periodic with the period N_p, i.e., $c_{k+iN_p}^{(\upsilon)} = c_k^{(\upsilon)}$, $(\forall) i, k$. Each code element is an integer in the range $0 \le c_k^{(\upsilon)} < N_h$. The time hopping code provides therefore an additional time shift to each pulse, in the pulse trains with the *k-th* monocycle undergoing an added shift of $c_k^{(\upsilon)} T_c$. The added time shifts caused by code are discrete times between 0 and $N_h T_c$ seconds.

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Comparing figures 2 and 3, one can observe that, when the bearer impulses are not randomized the power spectrum is dominated by spectral lines whereas, when the randomization codes are used, it reduces the spectrum lines, and the power spectrum is predominately continuous.

In figures 5 and 6 can be observed the benefits of using block waveform modulation. By using values for M higher than 2, it is possible to increase the number of users for a fixed probability of error as it is shown in figure 5.

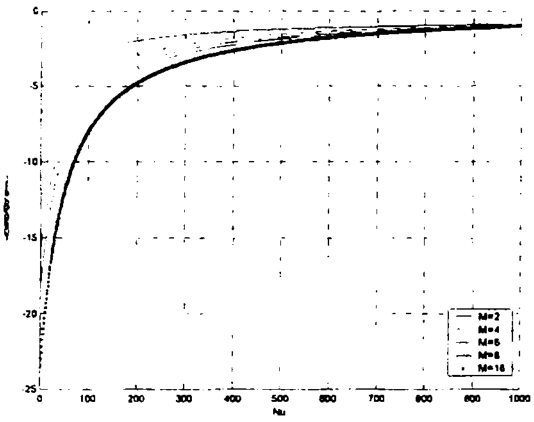

Figure 5.The base 10 logarithm of probability of bit error for orthogonal PPM as function of *Nu* user for different values of M,using *Rb* =9.6 *Kbps*.

If we use a high bit data rate and lower bit error probability we obtain he curves shown in figure 6.

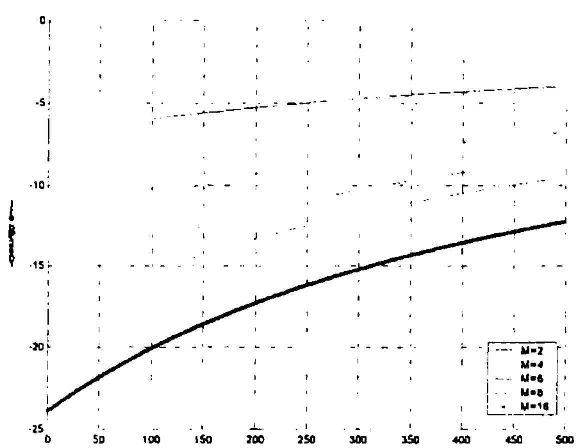

Figure 6. The base 10 logarithm of the probability of bit error for orthogonal PPM signals as function of *Nu* for different values of M, using *Rb* = *1048 Kbps*.

V. CONCLUSION

In this paper we have shown that for applications requiring high data rate (1024 Kbps) combined with low probability of bit error (10^{-8}), impulse radio modulation using orthogonal signals is potentially able to support hundred of users.

Similarly, for applications requiring low data rate (9.6 Kbps) and moderate probability of error (10^{-4}) impulse radio modulation using orthogonal signals is potentially able to support thousands of users.

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