Ultra-Wideband Transceiver for Data Transmission in Visual Prosthesis

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Abstract - This study aims to develop a transmission system by Impulse Radio Ultra-wideband with very low complexity to be implemented using CMOS technology. Several reception and modulation techniques have been studied in order to determine which technique presents the best performance, lowest complexity and higher robustness towards synchronism errors. In order to study the system, several models are developed and simulated using Matlab/Simulink software.

In the system design an innovative technique for receiver synchronization and bit detection is developed. A new modulation technique is also introduced.

In this work we study two distinct techniques for UWB non-coherent transceivers, the Transmitted-Reference (TR) technique with Pulse Position Modulation and we introduce a new technique that uses Pulse Period Modulation. The non-coherent systems adopt autocorrelation receivers as a suboptimal receiver technique, which offers a good performance and a simple structure.

In order to mitigate the noise, and use the energy of the received multipath components, some improvements are developed in the receiver structure without increasing complexity.

Simulations were also carried out in the presence of Additive White Gaussian Noise (AWGN) and multipath fading channels, and comparisons are made to demonstrate the performance improvement of the autocorrelation receivers.

I - Introduction

Ultra-Wideband (UWB) technology is defined as any wireless transmission scheme that occupies a bandwidth of more than 25% of a centre frequency, or more than 1.5 GHz [1].

UWB has gained much interest during the last few years as a potential candidate for further wireless short-range data communication. The Federal Communications Commission (FCC) has allocated the spectrum from 3.1 GHz to 10.6 GHz for UWB applications [2].

Ultra-Wideband systems are based on the transmission of pulses with very short duration and low duty-cycles [3]. UWB characteristics are low complexity, low cost, low power consumption and high data rate connectivity [4].

A particular type of UWB communication is Impulse Radio (IR-UWB), where baseband pulses of very short duration are transmitted, typically on the order of a nanosecond, thereby spreading the energy of the radio signal from near dc to a few gigahertz [5] (Fig. 1).

Because of this large signal bandwidth occupancy, UWB systems must follow strict regulations [1] which limit the achievable data rates, transmission range and implementation of power control [6]. On the other hand, the large signal bandwidth provides a high robustness in dense multipath environments [4].

In general, current embodiments of UWB receivers sacrifice performance for low-complexity operation and a large discrepancy in performance exists between these implementations and the theoretically optimal receivers. The most common receiver implementations cited in UWB literature include threshold detectors, coherent correlation or rake receivers and autocorrelation receivers [6].

Many of the approaches in the UWB literature adopt coherent modulation receivers [7]. This systems demodulate the received signal by correlating it with a well designed template signal. The correlation receiver is optimal over AWGN channels [8]. However, it has to face great design challenges. In order to correlate the received signal with the template, the receiver needs to achieve a synchronization with inaccuracy much smaller than one pulse duration [5].

In addition, the radiation and propagation process can act on the transmitted pulse as a filter, whose characteristics vary with time. Therefore, the received signal can be seen as a train of distorted waveforms [9].

The correlation receiver also requires precise template signal design, which is difficult due to linear and nonlinear distortions of the signal during its transmission [7].

A conventional coherent receiver must be able to capture and track the energy associated with a number of multipath replicas. Due to complexity constraint, only a small subset of the received multipath replicas is expected
to be selected and combined, which implicates a high degradation of performance.

A different approach to overcome all the above mentioned disadvantages is based on the use of non-coherent reception techniques [10]. Non-coherent UWB receivers such as the autocorrelation receiver are promising alternatives to coherent receivers.

A technique belonging to this category is the Transmitted-Reference (TR) scheme. TR communication systems operate by transmitting a pair of unmodulated and modulated signals, employing the former to demodulate the latter. This technique does not require channel estimation and allow to capture a large amount of the received energy, despite distortions and multipath propagation. In this work we introduce a new data modulation technique, the Pulse Period Variation modulation. In this technique the binary data is modulated by varying the pulse repetition period. We also introduce a new non-coherent reception technique that uses an autocorrelation receiver for data demodulation.

II – Pulse Shape

One major parameter in UWB radio technology is the choice of a pulsed waveform. A short UWB pulse with less than 1 ns width is used to obtain a large bandwidth.

The pulse-shaping function that is mostly adopted is the Gaussian monocycle, although other functions as the orthogonal Hermite functions can be used to enable multi-user communications.

The first order Gaussian monocycle can be expressed as [12]:

\[ u_0(t) = K_0 e^{-\frac{t^2}{\tau^2}} \] (1)

where the constant \( K_0 \) define the maximum amplitude and the time constant \( \tau \) define the width of Gaussian pulse \( T_p \).

The energy of \( u_0(t) \) is defined as [12]:

\[ E_0 = \int_{-\infty}^{\infty} u_0(t)^2 dt = \frac{K_0^2 \tau \sqrt{2\pi}}{2} \] (2)

For gaining more flexibility in the frequency domain, the time function can be multiplied and modified by a phase-shifted sinusoid. The multiplied Gaussian pulses are defined as follows [12]:

\[ p(t) = u_0(t) \cdot \sqrt{2} \cos(2\pi f_c + \phi) \] (3)

where \( f_c \) is the shifting frequency and \( \phi \) is an arbitrary phase that can be zero without loss of generality.

III - Data Modulation

In this section we study two modulation techniques, the Pulse Position Modulation (PPM) and introduce the Pulse Period Variation Modulation (PPV). PPM is a common method of modulation employed in IR-UWB [12]. In PPM each pulse is delayed or sent in advance of a regular time scale. By defining a basis pulse with arbitrary shape \( p(t) \), we can modulate the data by the delay parameter \( \tau \) to create a signal \( s(t) \), as shown in equation (4).

\[ s(t) = \sum_{m=1}^{M} p(t - mT - b_m \tau) \] (4)

where \( T \) is the pulse repetition period and \( b_m \in \{0,1\} \) is the binary data.

In the PPV modulation, the binary data is modulated by sending the pulses with a different pulse repetition period whether the bit is “1” or “0”, as represented in (5) and shown in Fig. 2.

\[ s(t) = \sum_{m=0}^{M} p(t - mT \left[ \frac{m}{N_r} \right]) \] (5)

where \( N_r \) is the number of pulses sent for each bit and \( T \left[ \frac{m}{N_r} \right] \in \left\{ T_0, T_1 \right\} \) represents the pulse repetition period.

IV - Channel Model

In order to implement an efficient UWB system it is critical to understand the characteristics of the propagation channel. According to the model proposed by the IEEE 802.15.3a working group, the channel model is based on a modification of the Saleh-Valenzuela’s one. Ignoring the distortion on the transmitted signal due to reflections, diffractions, scattering and the clustering phenomena, UWB multipath channel can be modelled as a linear filter with impulse response as [5]:

\[ Fig.2 - Pulse Period Variation Modulation. \]
where $L$ is the number of resolvable paths and $\tau_l$ is the delay of path $l$.

The multipath delay-spread is defined as $T_{mds} = \tau_l - \tau_s - T_p$ where $T_p$ is the width of the transmitted pulses.

The received signal $r(t)$ can be expressed as [5]:

$$ r(t) = s(t) * h(t) + w(t) $$

where $*$ denotes convolution, and $w(t)$ is the Additive White Gaussian Noise (AGWN) with power spectral density $N_0/2$.

### V – The Transmitted Reference System

We consider a single-user UWB-TR communication system which employs binary PPM modulation.

The primary drawback of this system is the performance degradation associated with the employment of noisy received signals as reference or template signals in demodulation [16]. TR modulation avoids the stringent synchronization requirements that exist in coherent detection techniques by eliminating the need for individual pulse synchronization with locally generated templates [3].

TR modulation is defined as the transmission of a pair of pulses separated in time from each frame. The first pulse is the un-modulated reference pulse, which provides the multipath channel’s impulse response to the demodulator at the receiver end.

In TR-PPM the time gap between these two pulses in a frame modulates the binary data, as shown in Fig. 3.

In TR receivers there is no need for channel estimation algorithms.

The general signal model for a TR-PPM modulated signal is [16]:

$$ s_{req}(t) = \sum_{i=0}^{N_f-1} E_0 \sqrt{E_p} p(t - iT_f) + \sqrt{E_p} p(t - iT_f - \Delta t_{i}) $$

where $N_f$ is the number of frames in one symbol, $T_f$ is the frame duration, $E_0$ is the energy of a single pulse and $p(t)$ represents the modulated Gaussian pulse with duration $T_p$.

This $N_f$ frames represent one bit and constitute one symbol with duration $T_s = N_f T_f$. Two pulses are transmitted during each frame, with a time delay $\Delta t_0$ or $\Delta t_0$ if the transmitted bit is “1” or “0” respectively, with $\Delta t_1 > \Delta t_0$.

Due to the very low pulse duty-cycle of the transmitted signal, $T_f$, $\Delta t_1$ and $\Delta t_0$ can be selected to be sufficiently large ($\Delta t_1 > \Delta t_0 > T_{mds}$ and $T_f > 2\Delta t_1 + T_{mds}$) to prevent multipath induced inter-frame and inter-symbol interferences.

At the receiver end, the received signal is initially passed through an ideal bandpass filter (BPF) with centre frequency $f_0$ and one sided bandwidth $W$, chosen to be the same with the signal bandwidth. The optimal choice of $W$ exists to balance the energy capturing capability and the noise reduction effect.

The filtered noise term $n(t)$ is Gaussian with an auto-correlation function:

$$ R_{n(t)}(\tau) = WN_0 \frac{\sin(\pi W \tau)}{\pi W \tau} \cos(2\pi f_0 \tau) $$

The filtered received signal $\hat{r}(t)$ is then delayed by $\Delta t$ and correlated with itself. The receiver structure is shown in Fig. 4.

The performance of TR receivers largely depends on the appropriate use of their integration window in the presence of channel noise. The integration window introduces two important design parameters: length of integration interval and accurate position of the integration interval.

An advantage of TR modulation is its high performance in dense multipath environments [3] [17].

### VI – The PPV system

The PPV system uses a non-coherent, autocorrelation receiver for data demodulation.

The structure of the receiver is analogous to TR receiver and there’s also no need for channel estimation.
algorithms. The basic structure of a PPV receiver is reproduced in Fig. 5.

If the received bit is “0” the pulses have a repetition period equal to $T_0$, thus:

\[
\begin{align*}
& r_{m0}(t) = r(t), r(t-T_0) = r(t)^2 \\
& r_{m1}(t) = r(t), r(t-T_0) = 0
\end{align*}
\]

\[
\Rightarrow \begin{cases} 
 r_{m0}(t) = \int_{-\infty}^{\infty} r(t)^2 = E_p \\
 r_{m1}(t) = 0 
\end{cases}
\]

(10)

In the same way, if the received bit is “1”, the pulse repetition period is $T_1$, resulting:

\[
\begin{align*}
& r_{m0}(t) = r(t), r(t-T_1) = 0 \\
& r_{m1}(t) = r(t), r(t-T_1) = r(t)^2 \Rightarrow \\
& r_{m1}(t) = \int_{-\infty}^{\infty} r(t)^2 = E_p 
\end{align*}
\]

(11)

where $E_p$ is the energy of a single pulse.

The delay lines in the receiver must have a value equal to $T_0$ or $T_1$. An advantage of PPV modulation is its high performance in dense multipath channel. To avoid inter-symbol interferences $T_0$ and $T_1$ must be selected to be greater than the maximum multipath delay spread of the channel:

$T_0 > T_{mds}$ and $T_1 > T_{mds}$

VII – Transceivers Implementation

For the following subsections the Matlab/Simulink software was used in order to implement and simulate the TR-PPM and the PPV systems.

1– TR-PPM Emitter Structure

Fig. 6 presents the block diagram of the Gaussian pulse generator for the emitter of a TR-PPM system.

This circuit allows generating Gaussian pulses with duration $T_p = 400\text{ps}$, modulated by a sinusoid signal with $5\text{GHz}$

\[
p(t) = k_0 \cdot e^{-\frac{t^2}{4\tau^2}} \cdot \sqrt{2} \cos(2\pi f_s t) \quad (12)
\]

where $K_0 = \sqrt{2} = \frac{2E_p}{\tau \sqrt{2\pi}} = 1$ and the pulses energy are equal to $E_0 = \frac{\tau \sqrt{2\pi}}{4}$.

This pulse generator allows generating a sequence of frames, each one with duration $T_f$ and a pulse per frame. The simulation results are presented in Fig. 7, in this case $T_f = 8\text{ns}$

The emitter block diagram is presented in Fig. 8. In order to modulate the signal with binary information, a square wave’s generator is added to the emitter structure. The generated square waves have a $N_f \times T_f$ period and 50% duty-cycle.

The signals $s_i(t)$ and $s_o(t)$ can be represented by the equations:
The transmitted signal $s(t)$ is represented by Equation 11. Fig. 9 shows the time domain representation of the TR-PPM signal $s(t)$.

$$s(t) = s_0(t) \times \bar{Q}(t) + s_i(t) \times Q(t) \quad (15)$$

The square wave’s generator is used to modulate the signal with binary information. The square waves have period $T_q$ and 50% duty-cycle. Multiple pulses are transmitted to represent one Bit. $N_{i1}$ and $N_{i0}$ are the number of pulses transmitted to represent bit “1” and “0” respectively. Because $T_i > T_q$ and the duration of one bit is always $T_q / 2$, we have $N_{i1} < N_{i0}$.

The simulations results are presented in Fig. 11.

The simulations results are presented in Fig. 11.

3 — TR-PPM Receiver Structure

The received signal is initially passed through an ideal BPF with bandwidth 3.1-10.6GHz, equal to the received signal bandwidth (Fig. 12).

The receiver have two delay lines $T_{r1}$ and $T_{r0}$. Each line have a delay equal to the one used in the emitter to modulate the signal information: $T_{r1} = T_{i1}$; $T_{r0} = T_{i0}$.

The signals $r_{i1}(t)$ and $r_{i0}(t)$ are defined by the following expressions:

$$r_{i1}(t) = \int r(t) \cdot r(t-T_{r1}) \, dt \quad (16)$$
$$r_{i0}(t) = \int r(t) \cdot r(t-T_{r0}) \, dt \quad (17)$$

When a bit “0” is received, the signals become:

$$r_{i0}(t) = \int r(t) \cdot r(t-T_{r0}) \, dt = \int r(t)^2 = E_0$$
$$r_{i1}(t) = \int r(t) \cdot r(t-T_{r1}) \, dt = 0$$

If instead a bit “1” is received, it’s the opposite situation:

$$r_{i0}(t) = \int r(t) \cdot r(t-T_{r0}) \, dt = 0$$
$$r_{i1}(t) = \int r(t) \cdot r(t-T_{r1}) \, dt = \int r(t)^2 = E_0$$

The simulation results are presented in Fig. 13. The received Bit energy is equal to $N_f \times E_0 \approx 1.25 \times 10^{-10}$.

In the end of the bit reception, we have:

$$r_{i0}(t) = 0 \quad \text{if bit “1” is received}$$
$$r_{i1}(t) = N_f \times E_0 \quad \text{if bit “0” is received}$$
4 – PPV system Receiver Structure.

The received signal is initially passed through an ideal BPF with bandwidth 3.1–10.6GHz, equal to the received signal bandwidth (Fig. 14).

We can see that the receiver structure is very similar to the TR-PPM receiver structure, only the delay lines have different values. This values must be equal to the pulse repetition period used to modulate the signal: $T_0$ and $T_1$.

When a bit “0” is received:

$$\begin{cases} r_{00}(t) = \int r(t) \cdot (t - T_0) \, dt = \int r(t)^2 = E_p \\ r_{10}(t) = \int r(t) \cdot (t - T_1) \, dt = 0 \end{cases}$$  \hspace{1cm} (22)$$

If instead a bit “1” is received:

$$\begin{cases} r_{01}(t) = \int r(t) \cdot (t - T_0) \, dt = 0 \\ r_{11}(t) = \int r(t) \cdot (t - T_1) \, dt = \int r(t)^2 = E_p \end{cases}$$  \hspace{1cm} (23)$$

The total received energy per bit is equal to:

$$\begin{align*} r_{00}(t) &= N_{00} \times E_p \\ r_{10}(t) &= 0 \quad \text{if bit “0” is received} \\ r_{01}(t) &= 0 \\ r_{11}(t) &= N_{11} \times E_p \quad \text{if bit “1” is received} \end{align*}$$  \hspace{1cm} (24)$$

The simulation results are presented in Fig.15.

We see from Fig. 15 that the total energy received for each bit “0” is greater than the energy received for each bit “1”, this is because $N_{00} < N_{10}$.

5 – Bit End Detector and Bit Decision Block

The Bit End detection block for the TR-PPM and PPV receivers is presented in Fig. 16.

This detector is made of a low-pass filter, a derivative block and comparator.

In Fig. 17 are presented the simulation results of the TR-PPM decision block during the reception of a Bit “1”, in this case with $N_f = 2$ and $T_f = 16$ ns.

The decision block function is to determine what Bit was received. This decision is made in the moment the end of the Bit reception is detected. The decision block for both TR-PPM and PPV receivers is represented in Fig. 18 and is made of a triggered subsystem with a comparator and a delay line.
For the TR-PPM receiver the delay line value is equal to \( T_f \). For the PPV receiver his value is equal to \( T_0 \).

As an example this block was simulated using the TR-PPM receiver. Ten frames per Bit are transmitted, thus \( N_f = 10 \). The signal in the output of the integration blocks has a value equal to the sum of the received pulses energy (Fig. 19). Therefore, in the end of the 10 frames reception, the Bit decision’s Block input signal has a value equal to \( 10 \times E_0 \) if the received Bit was “1” and \(-10 \times E_0\) in case the received Bit was “0”.

The functioning of the Bit decision block is the follows: When the end of the Bit reception is detected in the comparator input, this signal is, delayed \( T_f \) units in time, and then compared with two predefined values:

\[
V_0 = \frac{(N_f - 1) \times E_0}{2} \tag{26}
\]

\[
V_1 = \frac{(N_f - 1) \times E_0}{2} \tag{27}
\]

If the input signal is greater than \( V_1 \), it means the received Bit was “1” and if is lower than \( V_0 \), the received Bit was “0”.

6 - TR-PPM Receiver Optimization

Three blocks are added to the receiver scheme; a multiplier, a low-pass filter (LPF) and a comparator, as shown in Fig. 20.

The signal on the comparator output can have two values: \( V_{OFF} = 0 \), in case its entrance is inferior to the noise level and \( V_{ON} = A_T \), in case its entrance is higher than the noise level. This allows eliminating effectively the noise. This signal is constituted by square pulses of \( A_T \) amplitude and duration approximately equal to \( T_e \) with two pulses per frame and a delay between them of \( T_{r1} \) or \( T_{r0} \), in case the transmitted Bit in the frame is “1” or “0”, respectively. The signal is then multiplied by a
delayed version of itself and later integrated. The input signals of the integrators 1 and 2 can be represented by:

\[
R_{in}(t) = \sum_{i=0}^{N-1} A^2_r \cdot \text{rect} \left( \frac{t - iT_c - T_{e0} - T_c}{T_c} \right)
\]

if bit "1" (28)

\[
R_{in}(t) = \sum_{i=0}^{N-1} A^2_r \cdot \text{rect} \left( \frac{t - iT_c - T_{e0} - T_c}{T_c} \right)
\]

if bit "0" (29)

where \( A_r \) is the amplitude of the square pulse and \( T_c \) its duration.

The energy of each square pulse is given by \( E_r = A^2_r T_c \). Thus, the integrated energy per Bit is equal to \( N_f E_r = N_f A^2_r T_c \) (24).

In Fig. 23 are presented the simulation results. In this example, \( N_f = 2 \) and "1 0" is the transmitted binary sequence. The square pulses have amplitude \( A_r = 1 \).

The LPF has a bandwidth of 1.8 GHz, resulting \( T_c = 0.8 \) ns.

Conform the received bit was “1” or “0” the integrator 1 input is given by:

\[
R_{i1}(t) = \sum_{i=0}^{N-1} A^2_r \cdot \text{rect} \left( \frac{t - iT_c - T_{e0} - T_c}{T_c} \right)
\]

bit "1" (30)

Or the integrator 2 input:

\[
R_{i2}(t) = \sum_{i=0}^{N-1} A^2_r \cdot \text{rect} \left( \frac{t - iT_c - T_{e0} - T_c}{T_c} \right)
\]

bit "0" (31)

The simulations results are represented in fig.25

\[
\text{Fig.23 – Simulation results of the improved TR-PPM receiver.}
\]

7 – PPV receiver optimization

In Fig. 24 are presented the simulation results. The low-pass filter has a bandwidth of 1.8 GHz.

\[
\text{Fig.24 – Block Diagram of the improved PPV receiver.}
\]

The energy received per bit is equal to \( E_r = A^2_r T_c \). Thus the energy received for each bit “1” is \( N_{i1} E_r = N_{i1} A^2_r T_c \) and for each bit “0” is \( N_{i0} E_r = N_{i0} A^2_r T_c \). As seen before \( N_{i0} > N_{i1} \), so the total energy received for each bit “0” is greater than the energy received for each bit “1”.

IX – Results

It is necessary to have a very precise control of the emitter and receiver delay lines. Any difference between these delays in emission and reception, leads to a significant degradation of the system performance. The synchronism error, \( \varepsilon \), has been defined as the difference between the delay lines in emission and reception. We simulate the TR-PPM and PPV receivers for \( \varepsilon \in [-500\,\text{ps}, 500\,\text{ps}] \) with pulse duration \( T_p = 400\,\text{ps} \).

The simulation results for the TR-PPM and PPV receiver are presented in Fig. 26.
The simulation results of the improved TR-PPM and PPV receiver in the presence of the same synchronism error are presented in Fig. 27.

In the improved TR-PPM or PPV receiver, the signal amplitude in the integrators output linearly decreases with the increase of $\varepsilon$. The output signal amplitude is function not only of the synchronism error $\varepsilon$, but also of the square pulses width $T_e$:

$$r_{0,1} = A_T^2 (T_e - |\varepsilon|)$$  \hspace{1cm} (32)

These receivers have higher robustness towards synchronism errors, as by increasing $T_e$ the signal amplitude $r_{0,1}$ also increases facing the same synchronism error. It is then possible to increase the receptor robustness towards $|\varepsilon|$, by controlling $T_e$.

In Fig. 28 are presented the simulation results of the TR-PPM receiver and the improved TR-PPM receiver in the presence of a AGWN channel without multipath reflections. In this example we consider:

$$N_f = 2; \ T_f = 16 \text{ ns}; \ T_{\text{rs}} = 4.5 \text{ ns}.$$ In the TR-PPM receiver case, the integrator output signal is affected by AWGN noise, which means that $R_s(T_{\text{rs}}) \neq 0$. We also see that the signal in the improved TR-PPM is not affected by AWGN noise which causes an increase in output SNR.

As seen before in section VII.7, in both PPV and improved PPV systems, the total energy received for each transmitted bit “1” is lower than the total energy received for each bit “0”, resulting in a lower SNR when receiving a bit “1”. This can be a major disadvantage when the system is transmitting over a channel with heavy AWGN noise. Thus the improved TR-PPM system is a preferable choice for transmissions over AWGN channels.

Finally, Figs. 29 and 30 shows the results of the improved TR-PPM receiver simulation in the presence of a multipath channel with impulse response as in (5) and $T_{\text{mds}} \approx 60 \text{ ns}$. For simplicity of representation we consider $W(t) \cong 0$.
Although the TR-PPM receiver do not require channel estimation is able to capture the energy associated with a high number of multipath replicas, so the overall energy received per pulse is increased.

X — Conclusion

New implementations of a non-coherent TR-PPM transceiver system and a new non-coherent PPV system are presented in this study. Some improvements were developed in the receivers systems in order to increase its performance and robustness towards noisy channels with multipath characteristics, creating the system less sensitive to synchronism errors. System simulations proved the significant increase of the system performance in result of the optimizations, yet without increasing their complexity.

References


