Ultra-Wideband Non-Coherent Transceivers
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ABSTRACT
This paper describes two transceiver architectures for a transmission system using Impulse Radio Ultra-wideband, having innovative techniques for receiver synchronization and bit detection. Both systems have low complexity and are suitable for implementation using CMOS technology.

Two distinct techniques for UWB non-coherent transceivers are focused: the Transmitted-Reference (TR) technique with Pulse Position Modulation with a new optimised receiver in order to mitigate the noise and use the energy of the received multipath components; and a new technique that uses Pulse Period Variation (PPV). Both systems are non-coherent and so they adopt autocorrelation receivers as suboptimal receiver technique, which offers a good performance and a simple structure.

Simulations are 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 optimised autocorrelation receivers.

1. Introduction
Ultra-Wideband (UWB) technology is defined as any wireless transmission scheme that occupies a bandwidth of more than 20% of a centre frequency, or more than 500 MHz. 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. UWB systems characteristics are low complexity, low cost, low power consumption and high data rate connectivity [1].

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 [2] which provides a high robustness in dense multipath environments [2]. 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 [3]. These systems demodulate the received signal by correlating it with a well designed template signal. The correlation receiver is optimal over AWGN channels [4]. However, it has to face great design challenges. In order to correlate the received signal with the template, the receiver needs to achieve synchronization with inaccuracy much smaller than one pulse duration. 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. The correlation receiver also requires precise template signal design, which is difficult due to signal linear and nonlinear distortions during transmission [3]. 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 high loss of performance.

Non-coherent receivers [5, 6] overcome those problems and structures as autocorrelation receivers are promising alternatives to coherent receivers.

A technique belonging to this category is the Transmitted-Reference (TR) scheme [3]. 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 allows capturing a large amount of the received energy, despite distortions and multipath propagation.

In this paper we introduce a new data modulation technique, the Pulse Period Variation (PPV) modulation. In PPV 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.

We only address PPM and PPV modulations (studied with Matlab/Simulink). More common OOK, PAM and BPSK modulations are not considered for their poor noise immunity or higher circuit complexity.

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2. Pulse Shape Generator

The pulse-shaping function mostly adopted is the Gaussian pulse, although other functions as the orthogonal Hermite functions can be used to enable multi-user communications. However, the Gaussian pulse, in order to gain more flexibility in the frequency domain, can be multiplied by a phase-shifted sinusoid. The multiplied Gaussian pulses are defined as [2]:

\[ p(t) = u_c(t) \cdot \sqrt{2} \cos(2\pi f_c t + \phi_c) \]

\[ \phi_c = \frac{\pi}{2} \frac{\tau}{\tau_c} \]  

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

In Fig. 1 the Gaussian pulse generator block diagram is presented. In Fig. 2 a sequence of 8 ns duration frame is shown. The generated Gaussian pulses have duration \( T_p = 400 \) ps, and modulate a 5 GHz sinusoid:

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

\[ \tau = \frac{\tau_c}{\sqrt{2}} = 1 \]

and the energy of the pulses is equal to \( E_0 = \frac{\tau^2 \cdot 2\pi}{4} \).

3. 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 [7]:

\[ h(t) = \sum_{l=1}^{L} a_l \cdot \delta(t - \tau_l) \]  

where \( L \) is the number of resolvable paths and \( \tau_l \) is the delay of path \( l \). The multipath delay-spread is defined as \( T_{mds} = \tau_l - \tau_i - T_p \), where \( T_p \) is the width of the transmitted pulses. The received signal \( r(t) \) is given by:

\[ r(t) = s(t) \ast h(t) + w(t) \]  

where \( \ast \) denotes convolution, and \( w(t) \) is the AGWN with power spectral density \( N_0 / 2 \).

4. TR and PPV Modulation Techniques

TR Modulation

Here we briefly recall the Transmitted Reference (TR) modulation; we consider a single-user UWB communication system with binary PPM modulation.

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. A drawback of this system is the performance degradation due to the employment of noisy received signals as reference or template signals in demodulation.

In TR modulation two pulses separated in time from each frame are transmitted. The first pulse is the unmodulated reference pulse, which provides the multipath channel’s impulse response to the demodulator. In TR-PPM the time gap between these two pulses in a frame modulates the binary data, as shown in Fig. 3.

\[ s_{PPM}(t) = \sum_{i=0}^{N_f-1} \left[ E_p p(t - iT_f) + \sqrt{E_p} p(t - iT_f - \Delta t_{i}) \right] \]

where \( N_f \) is the number of frames in one symbol, \( T_f \) is the frame duration, \( E_p \) 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 in each frame, with time delay \( \Delta t_1 \) or \( \Delta t_0 \) if the transmitted bit is “1” or “0”, respectively, \( (\Delta t_1 > \Delta t_0) \). Due to the very low pulse duty-cycle of the signal, \( T_f \), \( \Delta t_1 \) and \( \Delta t_0 \) can be made large enough.
\[ \Delta t_i > \Delta t_0 > T_{md}, \] \[ T_j > 2 \Delta t_i + T_{md} \]

... to prevent multipath induced inter-frame and inter-symbol interferences. TR modulation has the advantage of high performance in dense multipath environments [3].

**PPV Modulation**

In the PPV modulation, defining a basis pulse with arbitrary shape \( p(t) \), the binary data is modulated by sending the pulses with a different pulse repetition period depending on the bit is “1” or “0”, as represented in (6) and shown in Fig. 4.

\[
s(t) = \sum_{m=0}^{N_i} p(t - mT_{\text{rep}}) \tag{6}
\]

where \( N_i \) is the number of pulses sent for each bit and \( T_{\text{rep}} \in \{ T_0, T_1 \} \) represents the pulse repetition period.

![Fig. 4 - Pulse Period Variation Modulation.](image)

The PPV system uses a non-coherent, autocorrelation receiver for data demodulation with no need for channel estimation algorithms. An advantage of the 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: \( T_0 > T_{md}, \text{ and } T_1 > T_{md} \).

5. **Transceivers Implementation**

In the following section the implementation of both the improved TR and PPV transceivers are presented.

5.1– Improved TR-PPM Transceiver

**Emitter**

The TR-PPM emitter block diagram is presented in Fig. 5. In order to modulate the signal with binary information, a square wave 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 \( s_0(t) \) can be represented by:

\[
s_i(t) = \sum_{n=0}^{N_i} [p(t - nT_f) + p(t - nT_f - T_{\text{rep}})] \tag{7}
\]

The transmitted signal \( s(t) \) is represented by equation (8) and is shown in Fig. 6.

\[
s(t) = s_0(t) \times \dot{Q}(t) + s_1(t) \times \dot{Q}(t) \tag{8}
\]

![Fig. 5 – Block Diagram of the TR-PPM Emitter.](image)

![Fig. 6 - TR-PPM signal \( s(t) \). \( T_f = 8 \text{ ns} \). \( N_f = 1 \). \( T_e = 1 \text{ ns} \) and \( T_{\text{e0}} = 600 \text{ ps} \). The binary sequence transmitted is “10”.](image)

**Improved Receiver**

The improved receiver is shown in Fig. 7. The signal is initially passed through an ideal band-pass filter (BPF) with bandwidth 3.1-10.6 GHz, equal to the received signal bandwidth. In the classic receiver this signal passes directly to the parallel branches with 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: \( T_{r1} = T_e; T_{r0} = T_{e0} \).

In this improved version, three blocks are added to the receiver after the input BPF: a multiplier, a low-pass filter (LPF) and a comparator, as shown in Fig. 7.

The signal is multiplied by itself and then filtered by a LPF with 1.8 GHz bandwidth. The resultant pulse has a \( T_e \) duration, superior to the \( T_p \) pulse duration. The delay and the signal attenuation, as well as \( T_e \), are inversely proportional to the LPF bandwidth. The comparator output shown in Fig. 8 has two frames with two pulses per frame and a delay between them of \( T_{r1} \) or \( T_{r0} \), for a transmitted Bit “1” or “0”, respectively.

The signal at the comparator output can have two values: \( V_{\text{OFF}} = 0 \), when its entrance is inferior to the noise level and \( V_{\text{ON}} = A_f \), when its entrance is higher than the noise level. This allows eliminating effectively the noise. This signal is constituted by square pulses of \( A_f \) amplitude and duration approximately equal to \( T_e \) with two pulses per frame and a delay between them of \( T_{r1} \) or \( T_{r0} \), for a transmitted Bit “1” or “0”, respectively.
Multiplier

Fig. 7 – Block Diagram of the improved TR-PPM receiver.

Fig. 8 – Results of the improved TR-PPM receiver.

The signal is then multiplied by a delayed version of itself and later integrated. The input signals of the integrators 1 and 2 are represented by:

\[
R_{m1}(t) = \sum_{n=0}^{N-1} A_i^2 \cdot \text{rect} \left( \frac{t - iT_i - T_i/2}{T_e} \right) \quad \text{if bit} \, "1" \tag{9}
\]

\[
R_{m0}(t) = \sum_{n=0}^{N-1} A_i^2 \cdot \text{rect} \left( \frac{t - iT_i - T_i/2}{T_e} \right) \quad \text{if bit} \, "0" \tag{10}
\]

where \( A_i \) is the square pulse amplitude and \( T_e \) its duration. Each square pulse energy is \( E_i = A_i^2 T_e \) and so the integrated energy per Bit is \( N_i E_i = N_i A_i^2 T_e \).

Simulation results for the improved receiver are presented in Fig. 9, which reflect an improvement regarding those of the classic architecture shown in Fig. 10, as a consequence of the new energy recovering scheme at the receiver input. In this example, \( N_f = 2 \) and “1 0” is the transmitted binary sequence. The square pulses have amplitude \( A_f = 1 \) and the LPF has a 1.8 GHz bandwidth, resulting in \( T_e = 0.8 \) ns.

5.2 – PPV Transceiver

Emitter

The PPV emitter block diagram, represented in Fig. 11, uses a square wave generator to modulate the signal with binary information. The square waves have period \( T_q \) and 50% duty-cycle. Multiple pulses are transmitted for each Bit. \( N_{i1} \) and \( N_{i0} \) are the number of pulses for bit “1” and “0” respectively. As \( T_i > T_0 \) and the bit duration is always \( T_i/2 \), we have \( N_{i1} < N_{i0} \). The result is presented in Fig. 12.

Fig. 9 – Results of the improved TR-PPM receiver.

Fig. 10 – Results of the TR-PPM receiver. Number of frames for bit: \( N_f = 2 \). The received binary sequence is “10”.

Fig. 11 – Block Diagram of the PPV emitter.

Fig. 12 – Time domain representation of signals \( s_i(t) \), \( s_f(t) \)

Receiver

The receiver is represented in Fig. 13. The received signal is first passed through a BPF with 3.1-10.6 GHz bandwidth, followed by a multiplier, a low-pass filter (LPF) with 1.8 GHz bandwidth and a comparator. This receiver structure is similar to that of TR-PPM, but the delay lines have different values: they must be equal to the pulse repetition period \( T_0 \) and \( T_i \) of the modulation.
Fig. 13 – Block Diagram of the improved PPV receiver.

Depending on the received bit, integrator 1 input and integrator 2 input are given by:

\[
R_{i1}(t) = \sum_{i=0}^{N-1} A_i^2 \cdot \text{rect}\left(1 - iT_b - T_i / 2 \right) \quad \text{bit} = "1" \quad (11)
\]

\[
R_{i0}(t) = \sum_{i=0}^{N-1} A_i^2 \cdot \text{rect}\left(1 - iT_b - T_i / 2 \right) \quad \text{if bit} = "0" \quad (12)
\]

respectively. The energy per pulse is \( E_i = A_i^2 T_i \), for each bit “1” is \( N_i E_i \) and for each bit “0” is \( N_0 E_i \).

Simulations results are shown in Fig. 14. To note that the total energy received for each bit “0” is greater than that received for each bit “1” because \( N_i < N_0 \).

5.3 – Bit End Detector and Bit Decision Block

The Bit End Detector Block is equal for the TR-PPM and PPV receivers and is presented in Fig. 15. Its function is to establish the end of bit time.

This detector is made of a LPF, a derivative block and a comparator. In Fig. 16 simulation results of the TR-PPM Bit End Detector Block during the reception of a Bit “1”, with \( N_f = 2 \) and \( T_f = 16 \text{ ns} \), are presented.

Fig. 16 – Results of the Bit End Detector Block.

The Bit Decision Block determines what Bit is received when the end of the Bit reception is detected. This decision block for both TR-PPM and PPV receivers is shown in Fig. 17 and is made of a triggered subsystem with a comparator and a delay line: The delay line value for TR-PPM is \( T_f \) and for PPV is \( T_0 \). An example of this block performance with TR-PPM for ten frames per Bit (\( N_f = 10 \)) is shown in Fig. 18.

Fig. 17 – Block Diagram of the Bit Decision Block.

Fig. 18 – Results of the Bit Decision Block.

The signal at the output of the integration block has a value equal to the sum of the energies of the received pulses, in this case \( 10 \times E_0 \) if the received Bit was “1” and \( -10 \times E_0 \) if the received Bit was “0”. The Bit Decision Block works as 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:

\[
V_0 = -(N_f - 1) \times E_0 / 2 \quad \text{and} \quad V_1 = (N_f - 1) \times E_0 / 2 \quad (13)
\]

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

6. Synchronism and Noise Results

Any relevant difference between the 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 this delay in emission and reception. We simulate the TR-PPM and PPV receiver for $\varepsilon \in [-500\text{ps}, 500\text{ps}]$ and $T_p = 400\text{ps}$ of pulse duration. The results are presented in Fig. 19.

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 depends on both the error $\varepsilon$ and the square pulses width $T_s$, namely: $r_{h,i} = A_f^2(T_s - |\varepsilon|)$. So, these receivers have high robustness towards synchronism errors, as by increasing $T_s$ the signal amplitude $r_{h,i}$ also increases. It is then possible to increase the receptor robustness towards $|\varepsilon|$, by controlling $T_s$. In Fig. 20 the simulation results of the classic TR-PPM receiver and the improved TR-PPM receiver in the presence of a AGWN channel without multipath reflections are presented. For this example: $N_f = 2$; $T_f = 16\text{ ns}$; $T_s = 4,5\text{ ns}$. In the classic TR-PPM receiver case, the integrator output signal is affected by AWGN noise, so $R_{n}(T, i) \neq 0$. However, in the improved TR-PPM case, the signal is not affected by AWGN noise, resulting an increase in the output SNR.

To note that both PPV and TR-PPM receivers may not need channel estimation and can capture the energy associated with a high number of multipath replicas, so the overall received energy per pulse is increased.

Finally, Fig. 21 shows the results of the improved TR-PPM transceiver with a multipath channel with impulse response as $h(t) = 0$.

7. Conclusions

New implementations of a non-coherent TR-PPM transceiver system and a new non-coherent PPV system are presented in this paper. The PPV scheme and the improvement implemented in the TR-PPM transceiver result in good performance and robustness towards noisy channels with multipath characteristics, making the system less sensitive to synchronism errors. In the improved TR-PPM case, the significant increase of the system performance is achieved without a relevant increase of its complexity. Simulation results confirm the expected results for both the transceivers. Also, both systems have low complexity and are suitable for implementation using CMOS technology.

References