IMPACT OF THE NUMBER OF FINGERS OF A SELECTIVE RAKE RECEIVER FOR UWB SYSTEMS IN MODIFIED SALEH-VALENZUELA CHANNEL

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ABSTRACT

The study for this paper focuses on searching the optimal number of fingers to be used by a Selective-Rake receiver using different ultra wideband systems. The presence of multipath fading has been taken into account using a modified Saleh-Valenzuela model. Coherent and non-coherent detection has been performed within different combining approaches. Non-coherent techniques have been implemented with and without power estimation. Results show that in non-coherent systems the optimal number of fingers D_{opt} required for the detection is lower than for coherent ones. On the other side, an increase of signal-to-noise ratio causes D_{opt} to get higher, which is the same trend noticed while passing from a line-of-sight (LOS) channel to a non-LOS one.

I. INTRODUCTION

According to the regulations, a radio system is classified as ultra wideband (UWB) if its bandwidth or fractional bandwidth is greater than 500 MHz or 20%, respectively [1]. Therefore, the signal detection in a multipath fading channel is a critical issue, due to the fact that the pulse waveforms characterising an UWB signal must have a really short time duration, on the order of nanoseconds. In this paper we investigate the best choice for the number of fingers to be used for a selective Rake (S-Rake) receiver, based on the best results in terms of bit error rate (BER) given by link lever simulations of several UWB systems at fixed value of received average signal-to-noise ratio E_b/N_0. Both coherent and non-coherent detection approaches have been taken into account, in order to cover the most typical implementations of a wireless system receiver.

II. UWB SYSTEM MODEL

A. Transmitted signal

UWB pulse waveforms: The pulse waveforms implemented in our study are the 4th and 5th derivatives of the Gaussian pulse [2]. This choice ensures to fulfil the UWB emission limits defined by FCC [1] and ETSI [3].

Spreading techniques: Time hopping (TH-UWB) and direct sequence (DS-UWB) spreading approaches have been implemented in our study. In both cases, one transmitted data bit is spread over multiple pulses, in order to achieve a processing gain in reception due to repetition codes.

In TH-UWB systems, the pseudo random (PR) time hopping code defines the transmission instants inside time hopping frames, for each of the users individually. The pulse integration gain (K) defines the number of frames composing a databit. The frame length T_f (pulse repetition interval) in TH-UWB is much longer than the pulse width T_p. From the pulse integration gain, the time length of each databit can be calculated as $T_d = KT_f$.

In DS-UWB approach a PR code is used to spread the data bit into multiple chips like in conventional DS spread spectrum systems still having a chip waveform with a UWB spectrum. In this case the processing gain (PG) defines the number of repetitions of UWB waveforms within a single databit. The databit length is therefore defined as $T_d = PG \cdot T_p$. Refer to [2] for a better insight on UWB spreading techniques

Modulation schemes: Since the goal of this study is to compare coherent and non-coherent combining techniques, a major stress has been given to orthogonal modulation schemes. Pulse position modulation (PPM) is a typical time-shifted modulation [4]. In order to soften the disruptive intra-symbol interference introduced by the multipath channel, a time shift 5 times the length of the pulse (2.5 ns) was chosen. Pulse shape modulation (PSM) is based on the transmission of two different orthogonal waveforms which are, in our case, the 4^{th} and the 5^{th} derivatives of the Gaussian pulse. The antipodal scheme implemented is based on a simple binary pulse amplitude modulation (BPAM).

B. Modified Saleh-Valenzuela Channel

The real-valued channel model used in our study is the modified Saleh-Valenzuela (SV) model proposed by the UWB channel model sub-committee within the IEEE P802.15.3a [5].

The model is based on empirical measurements in indoor environments. Due to clustering phenomena observed in the measurements, the proposed model is derived from the Saleh-Valenzuela one [6] using a log-normal distribution rather than a Rayleigh distribution for the multipath gain magnitude. Independent fading is assumed for each cluster as well as each ray within the cluster. The phase of the channel impulse response can be either 0 or π. Therefore the model contains no imaginary component.

Four different channel implementations are suggested, which are based on the average distance between transmitter and receiver, and on the presence or not of a LOS component.

In our study the following two channel modes have been taken into account:

- SV-1: Presence of LOS component (0-4m), RMS delay spread of 5.28 ns,
- SV-3: NLOS (4-10m) model, RMS delay spread of 14.28 ns,

characterising a channel with a dominant path and one whose power is distributed over many different paths, respectively.

C. Receiver structures

The receiver algorithm is based on the use of a selective Rake (S-RAKE). This kind of Rake requires an estimation of the power levels of the channel taps in order to choose the most powerful ones to feed the detector.

The performance evaluation has been carried out using two coherent and two non-coherent schemes, whether the phase of the channel taps is recovered or not, respectively. In all cases a perfect time recovery of the multipaths is assumed. For non-coherent schemes, channel amplitude recovery has been considered as a possible choice.

Equal gain combining (EGC) requires a perfect estimation of the phase of each channel taps to correct the offset at the received signal before the detection block. Let us define the received signal for a single databit as

$$r(t) = \sum_{n=1}^{N} a_n s(t-\tau_n) + n(t)\,, \qquad (1)$$

where N is the number of recovered paths, $s(t)$ the transmitted signal, $n(t)$ the Gaussian noise in the channel, $a_n = |a_n| \mathrm{e}^{j\theta_n}$ the gain and τ_n the delay of the n-th multipath, respectively. Being the modified SV model a real model, θ_n will assume only values 0 and π. Thus, the decision variables in EGC will be

$$U_i^{(\mathrm{EGC})} = \sum_{n=1}^{N} \mathrm{e}^{-j\theta_n} \int_0^{T_\mathrm{d}} r(t-\tau_n) w_i(t)dt,\ i = 0,1, \quad (2)$$

where T_d is the databit length and $w_i(t)$ the pulse waveform representing databit "*i*" at the receiver.

For maximal ratio combining (MRC), in addition to phase recovery, the received power level is estimated for each multipath. The decision variable will then assume the form

$$U_i^{(\mathrm{MRC})} = \sum_{n=1}^{N} a_n^* \int_0^{T_\mathrm{d}} r(t-\tau_n) w_i(t)dt. \qquad (3)$$

In non-coherent detection an absolute combiner (AC) can be implemented by summing up the absolute values of the outputs of all the matched filters before feeding the detector. In this case the decision variables will be

$$U_i^{(\mathrm{AC})} = \sum_{n=1}^{N} \left| \int_0^{T_\mathrm{d}} r(t-\tau_n) w_i(t)dt \right|. \qquad (4)$$

An alternative approach is to use a square-law combiner (SLC), i.e.

$$U_i^{(\mathrm{SLC})} = \sum_{n=1}^{N} \left| \int_0^{T_\mathrm{d}} r(t-\tau_n) w_i(t) dt \right|^2 . \qquad (5)$$

The implementation of the S-Rake receiver is based on the non-coherent power estimation (PE) of each single channel path. In order to improve the performance of the system, the knowledge of the channel amplitude can be used to weigh the output of each single correlator, similarly to the MRC approach, but without knowledge of the channel phase. This assumption leads to a more sophisticated implementation of AC, analytically defined as

$$U_i^{(\mathrm{AC+PE})} = \sum_{n=1}^{N} \left(|a_n| \cdot \left| \int_0^{T_\mathrm{d}} r(t-\tau_n) w_i(t) dt \right| \right) . \qquad (6)$$

The expression for $U_i^{(\mathrm{SLC+PE})}$ can be easily derived from (6) by squaring the second absolute value. For more information upon combining schemes, refer to [7].

III. SIMULATION RESULTS

The simulations presented in this study are based on a UWB system having PG = 20 dB. The length of the pulse waveforms is 0.5 ns, leading to a data rate of 20 Mbit/s. In TH case, the pulse integration gain K is 10 dB, that is, each databit is composed by 10 pulse waveforms, whose time location is defined by the PR code.

The modulation schemes are PPM (for TH only), PSM and BPAM (for both TH and DS). The combining approach are MRC and EGC for coherent detection and AC and SLC for non-coherent. PE has been taken into account for both the non-coherent schemes, while only AC results are shown without power estimation. Being BPAM a bipolar modulation, results are shown only using coherent detection. In all cases PE is assumed perfect, that is, the weight used for the Rake fingers are noiseless.

The E_b/N_0, defined as the total average signal-to-noise ratio at the receiver, has been fixed to two values, 8 and 15 dB, respectively. However, the signal-to-noise ratio in the decision variable is less, since not all the paths have been combined.

The detection block is defined by a selective chip-spaced receiver, being its time resolution of 0.5 ns, that is, equal to the length of the pulse waveform.

The channel models were simulated using at least 100 channel realisations for each number of fingers. The channel power has been normalised to 1 over all the channel realisations used in each single simulation.

Figs. 1 and 2 show the behaviour of some results of simulations which evaluated BER of the system as a function of the number of Rake fingers D for E_b/N_0 = 8 and 15 dB. Particularly, Fig. 1 represents DS-PSM for SV-1, and Fig. 2 is TH-PPM in SV-3 case. The absence of power estimation in both coherent and non-coherent approach (EGC and AC, respectively) generates a minimum value of BER for a defined D_opt, which can be clearly chosen as the minimum. This behaviour is more evident in SV-1, where the presence of a LOS component makes the minimum BER appear for a lower D, if compared to SV-3. In MRC case, due to the perfect weight used in estimation, the performances of the systems are continuously improving as D increases. Thus, the optimal value has been chosen where the BER performance tends to saturate. The difference in performance between AC+PE and SLC+PE is almost negligible. Then the choice between the two systems is only a real system implementation issue. The improvement given to the system by the use of coherent detection is characterised by a BER around 10 times lower if compared with the equivalent non-coherent implementation for a fixed value of SNR and D.

Table 1 depicts the optimal number of fingers for each of the systems taken into account. As mention above, since performance of AC and SLC are nearly superimposing, only AC and AC+PE results are depicted. The table gives some general trends, such as D_opt generally increases with the E_b/N_0. However, this effect is more remarkable for non-coherent systems. D_opt is higher in SV3, due to absence of LOS component. D_opt is also clearly lower for non-coherent systems. Antipodal modulations show the same results for a fixed E_b/N_0 and channel model, both in terms of number of fingers and BER. Among orthogonal modulations, TH-PPM is in average the one which shows the lowest values of D_{opt}, despite of poorer BER results.

IV. Conclusions

The study done for this paper focused on fixing the optimal number of fingers of a S-Rake receiver for several different UWB systems for coherent and non-coherent detection with or without perfect power estimation. Results show that the required optimal number of fingers generally increases with E_b/N_0, as well as if a SV channel model without LOS component is chosen. Moreover, non-coherent detection algorithms show lower values of optimal number of fingers, despite of poorer BER results.

V. Acknowledgments

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Table 1. Optimal number of Rake fingers for different receiver algorithms.

SV	Eb/N0 [dB]	UWB system concept	Optimal number of fingers (*D*)			
			Coher.		Non coher.	
			MRC	EGC	AC + PE	AC
1	8	DS-PSM	12	8	4	4
		TH-PSM	12	8	4	4
		TH-PPM	10	10	4	2
		DS-BPAM	12	10	-	-
		TH-BPAM	12	10	-	-
	15	DS-PSM	14	8	10	8
		TH-PSM	20	10	10	6
		TH-PPM	14	8	8	6
		DS-BPAM	15	10	-	-
		TH-BPAM	15	10	-	-
3	8	DS-PSM	18	16	8	6
		TH-PSM	20	18	6	6
		TH-PPM	20	14	4	4
		DS-BPAM	15	15	-	-
		TH-BPAM	15	20	-	-
	15	DS-PSM	20	16	16	12
		TH-PSM	16	16	16	12
		TH-PPM	20	18	10	6
		DS-BPAM	18	15	-	-
		TH-BPAM	18	20	-	-

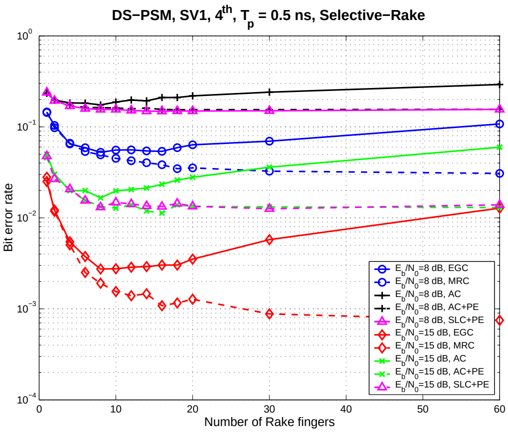

Figure 1. BER as function of the number of fingers for DS-PSM in SV-1.

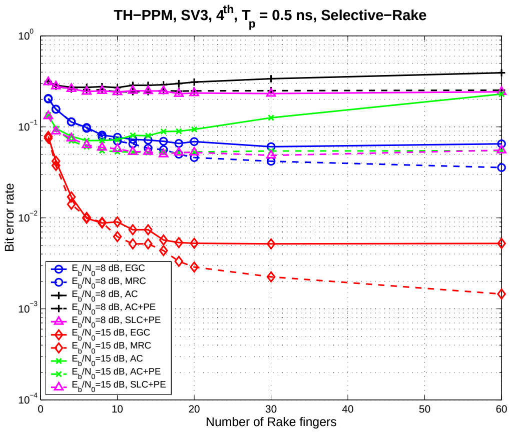

Figure 2. BER as function of the number of fingers for TH-PPM in SV-3