CHANNEL SHORTENING FOR ENHANCING PASSIVE UHF RFID PERFORMANCE

Taoufik Ben Jabeur & Abdullah Kadri

Qatar Mobility Innovations Center (QMIC), Qatar Science and Technology Park, Doha, Qatar
Email: taoufikb,abdullahk@qmic.com

ABSTRACT

This paper proposes using a channel shortening equalizer (CSE) to improve the performance of passive ultrahigh frequency radio frequency identification (UHF RFID) systems. In UHF RFID systems, the reader interrogates RF tags by transmitting continuous wave (CW) signals that power up the internal integrated circuitry of the tags that in turn, backscatter these signals to the reader after embedding their unique information. The overall performance of passive UHF RFID systems depends heavily on the power level of the signal impinged on the tag which is a function of the multipath channel environment in which the reader and the tags are deployed. In this paper, a channel shortening equalizer with a new constraint that exploits the knowledge of the propagation channel and the nature of the CW signal is proposed to boost the power of the impinged signal on the tag. The results show that using the proposed equalizer enhances the power level significantly which results in better performance.

Index Terms— Passive UHF RFID system, Channel shortening equalizer, SIMO system.

1. INTRODUCTION

Passive ultrahigh frequency radio frequency identification (UHF RFID) technology represents an effective way for identification of objects that are distant few meters from the reader. Due to their low cost equipments, the price of a RF tag doesn't exceed 0.20$, passive UHF RFID systems are used for many applications including access control, industrial automation, and indoor identification (clothes, IT equipments, etc.) [1]. Contrary to the conventional wireless systems, where both transmitter and receiver have their own power sources for their internal circuitries, passive UHF RFID system depends on the reader's transmitted CW signals to power up the internal integrated circuitry (IC) of the interrogated passive tags. Each tag backscatters a certain percentage of the impinged signal to the reader after embedding its own information. For successful tag detection, the power of the received signal at the tag should be higher than its threshold, and the power level of the backscattered signal received by the reader should be higher than its sensitivity [2]. In normal deployments, the presence of obstacles cause signals diffraction and reflection and thus lead to generate multipath fading channel (indoor environment). The performance of passive UHF RFID systems in a such environment has been studied in [3, 4].

Channel shortening equalizers (CSEs) are widely used in digital communication, mostly in Orthogonal Frequency Division Multiplexing (OFDM) systems, to reduce the length of the transmission channel to the length of the guard interval (GI). This reduction ensures a perfect channel equalization with a low complexity computation [5]. Furthermore, channel shortening leads to a reduction in bit error probability (BER) by controlling the combined channel equalizer [6]. The condition of perfect channel shortening can not be reached in systems that use one transmission antenna and one receiving antenna. Therefore, several CSEs are discussed for systems that use either single input multiple output (SIMO) or multiple input multiple output (MIMO) systems [7, 8].

In this paper, we propose using a channel shortening equalizer (CSE) to improve the performance of passive UHF RFID system deployed in multipath and fading environment. Usually, an RF tag has a simple IC that modulates and backscatters the received CW signal after embedding its unique information. Adding channel shortening function to RF tags is impractical due to the computation complexity and the need for power source. Therefore, to avoid this barrier, we propose using a CSE at the reader side. In order to achieve a perfect channel shortening, we consider a UHF RFID system with a reader that is equipped with two antennas and a tag that is equipped with one antenna. The contribution of this paper is in proposing a new algorithm for channel shortening for passive UHF RFID systems based on *apriori* knowledge of the propagation channels and the transmitted CW signal. Also, this paper shows that the power level of the impinged signal on the tag is improved and therefore, the overall performance of the system can be enhanced.

The rest of this paper is organized as follows. The system model is shown in Section 2. In Section 3, effects of multipath channel fading on the CW signal is illustrated. In Section 4, a review of SIMO CSE is given. Section 5, a new method to enhance the passive RFID power on the tag is proposed. Sections 6 and 7 are dedicated to the simulation results and conclusions, respectively.

This work was made possible by NPRP grant #NPRP 4–726–2–272 from the Qatar National Research Fund (a member of Qatar Foundation). The statements made herein are solely the responsibility of the authors.

978-0-9928626-3-3/15/$31.00 ©2015 IEEE

2. SYSTEM MODEL

The configuration of the passive UHF RFID system is shown in Fig. 1 where the system is composed of one RFID reader and one RF tag. The RFID reader has two antennas and the RF tag has one antenna. The assumption is that the same CW signal $s_{cw}(t)$ is transmitted via both antennas to the tag at the same instant. A Time EQualizer (TEQ) $w_i(t)$ $(i = 1, 2)$ is added at each antenna of the RFID reader. Taking into account the presence of the two TEQs, the instantaneous transmitted signals $x_i(t)$, $(i = 1, 2)$ at the two antennas are given by:

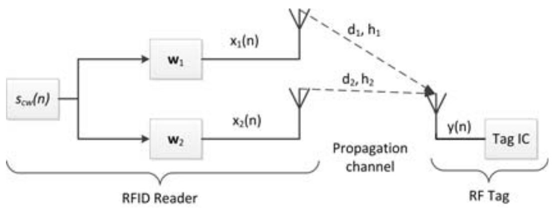

Fig. 1*: Passive UHF RFID system configuration.*

$$x_i(t) = s_{\text{cw}}(t) \star w_i(t) \tag{1}$$

Thus, the instantaneous received signal $y(t, d_1, d_2)$ at a tag located at distance d_1 m from the first antenna and d_2 m from the second antenna is expressed by:

$$y(t, d_1, d_2) = h_1(t, d_1) \star x_1(t) + h_2(t, d_2) \star x_2(t) + b(t) \tag{2}$$

where $h_1(t, d_1)$ and $h_2(t, d_2)$ are the channel impulse responses associated with the first and second antennas, respectively and $b(t)$ is the additive white Gaussian noise (AWGN). In this model, the function $h_i(t, d_i)$ represents the propagation channel that includes the effect of the forward link fading, which its envelope can follow either Rician or Rayleigh distribution, and the channel pathloss $L(d_i), i = (1, 2)$. Without loss of generality, we assume that the propagation channel is linear time-invariant and therefore, $h_i(t, d_i)$ can be modeled as [9]:

$$h_i(t, d_i) = \sum_m \alpha_{i,m} \delta(t - \tau_{i,m}) \tag{3}$$

where i refers to the antenna index ($i = 1, 2$) and m is the path index. $\alpha_{i,m}$ is the path coefficient and $\tau_{i,m}$ corresponds to the path delay. For the rest of this paper, we use a discrete form instead the continuous form. Thus, the discrete time n and frequency k can be expressed as: $n = tf_s$, $k = 2f\frac{M}{f_s}$, where f_s is the sampling frequency, M is total number of samples in frequency domain (FD). Therefore, the ith channel $\mathbf{h}_i$ with size L can be given by the following vector:

$$\mathbf{h}_i = [h_i(0), \cdots, h_i(L-1)]^T \tag{4}$$

We assume that both TEQs $\mathbf{w}_i$, have the same size with length q and they are expressed as:

$$\mathbf{w}_i = [w_i(0), \cdots, w_i(q-1)]^T$$

The CW signal $s_{cw}(n)$ is chosen as a tone and it can be extended to multi-tone signal or linear frequency modulation (LFM) signal:

$$s_{cw}(n) = A\cos(2\pi\frac{k_0 n}{N}) \tag{5}$$

where A is the amplitude of CW signal and it is assumed to equal 1, N is the size of signal, and k_0 is the carrier frequency index. Using the commutative property of the convolution operator, the received signal given by Eq. 2 can be rewritten as:

$$\begin{aligned} y(n) =& s_{cw}(n) \star w_1(n) \star h_1(n) + s_{cw}(n) \star w_2(n) \star h_2(n) + b(n) \\ =& s_{cw}(n) \star (h_1(n) \star w_1(n) + h_2(n) \star w_2(n)) + b(n) \end{aligned} \tag{6}$$

We denote the combined channel-time equalizer by $(\mathbf{c})$ of size (L+q-1) which is defined as:

$$\mathbf{c} = [c(0), \cdots, c(L+q-1)]^T = \mathbf{H}\mathbf{w} \tag{7}$$

where $\mathbf{H}$ is the channel matrix defined as follows:

$$\mathbf{H} = \begin{bmatrix} \mathbf{h}(0) & \mathbf{0} & \cdots & \cdots & \cdots & \cdots & \mathbf{0} \\ \mathbf{h}(1) & \mathbf{h}(0) & \mathbf{0} & \cdots & \cdots & \cdots & \vdots \\ \vdots & \ddots & \ddots & \mathbf{0} & \cdots & \cdots & \vdots \\ \mathbf{h}(L-1) & \cdots & \cdots & \mathbf{h}(0) & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \mathbf{0} \\ \mathbf{0} & \cdots & \mathbf{0} & \mathbf{h}(L-1) & \cdots & \cdots & \mathbf{h}(0) \end{bmatrix}$$

where $\mathbf{h}(i) = [h_1(i), h_2(i)]$ and the TEQ vector is denoted as:

$$\mathbf{w} = [w_1(0), w_2(0), w_1(1), w_2(1), \cdots, w_1(q-1), w_2(q-1)]^T \tag{8}$$

The objective of the CSE is to reduce the combined channel-time equalizer to:

$$\mathbf{c} = \underbrace{\mathbf{h_1} \star \mathbf{w_1}}_{\mathbf{c}_1} + \underbrace{\mathbf{h_2} \star \mathbf{w_2}}_{\mathbf{c}_2} = [\mathbf{0}_{1,D}, \mathbf{v}^T, \mathbf{0}_{1,L+q-1-\nu-D}]^T \tag{9}$$

where $\mathbf{v} = [v(0), \cdots, v(\nu-1)]^T$ is the target impulse response filter (TIR) of size ν, and D is the equalizer delay.

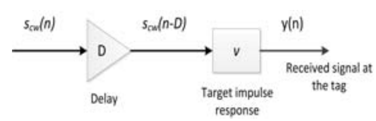

Fig. 2*: Equivalent system using channel shortening equalizer.*

Assuming that the perfect CSE is achieved, the system model illustrated in Fig. 1 can be replaced by the one shown in Fig. 2. We note that the CSE will be equivalent to the channel equalization when $\nu = 1$. Moreover, we observe that the passive UHF RFID with two transmission antennas at the reader and one antenna at the tag can be virtually replaced by a SISO passive UHF RFID system with TIR $\mathbf{v}$ channel, as illustrated in Fig. 2. Before the optimization of the TIR $\mathbf{v}$, we introduce the effect of the multipath channel on the CW signal in the next section.

3. EFFECTS OF MULTIPATH FADING CHANNEL

In this section, the effects of multipath fading channels on the proposed CW signals is analyzed. An example of multipath fading channel is shown in the time domain (TD) in Fig. 3-a. The frequency domain (FD) of the channel is illustrated in Fig. 3-b where we show also the CW signal. We note here that the energy of the CW signal is concentrated only on one tone with amplitude equals to 1. Due to the effect of the multipath channel, the filtered CW signal in the FD is still concentrated on the tone but its energy in this tone is reduced and the level of the noise outside the tone is increased.

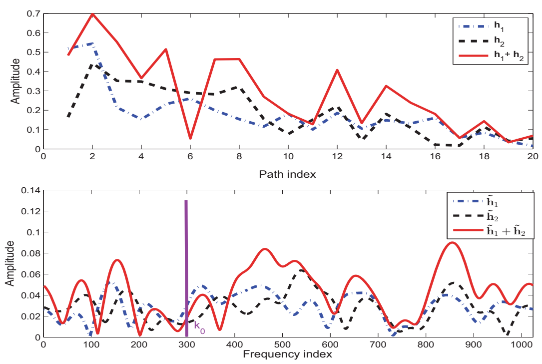

Fig. 3*: Example of a multi-path Rayleigh channel with length $L = 20$ in time domain a) and Frequency domain b) .*

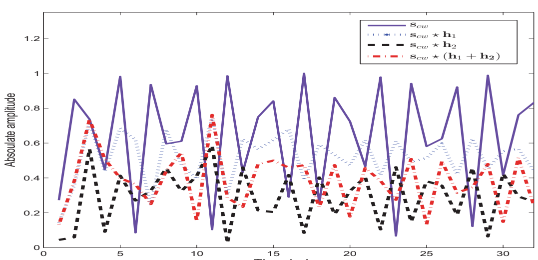

Fig. 4*: CW and the filtered CW signals using channel* 1*, channel* 2 *and the summation of both channels.*

In TD, the absolute amplitude of the original CW signal varies between 0 and 1. Due the effect of the multipath channel, the absolute amplitude of the filtered CW signal is reduced to the range $[0, 0.7]$ as illustrated in Fig. 4 and the energy loss of the CW signal is significant.

4. SIMO CHANNEL SHORTENING EQUALIZER

First, we assume that both channels $\mathbf{h}_1$ and $\mathbf{h}_2$ are known or estimated and they are time invariant. We partition the channel filtering matrix $\mathbf{H}$ as follows:

$$\mathbf{H} = \left[\mathbf{H}_{head}^T \, \mathbf{H}_{win}^T \, \mathbf{H}_{tail}^T\right]^T \tag{10}$$

where $\mathbf{H}_{head}, \mathbf{H}_{win}$ and $\mathbf{H}_{tail}$ are made from the first D, intermediate (ν), and last $(L+q-\nu-D-1)$ rows of $\mathbf{H}$, respectively. In particular, we have

$$\mathbf{H}_{win} = \begin{bmatrix} \mathbf{h}(D-1) & \cdots & \mathbf{h}(0) & \mathbf{0} & \cdots & \cdots & \mathbf{0} \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \mathbf{h}(D+\nu) & \cdots & \cdots & \mathbf{h}(0) & \mathbf{0} & \cdots & \mathbf{0} \end{bmatrix}$$

Eq. 9 is equivalent to:

$$\mathbf{H}_{out} = \begin{bmatrix} \mathbf{H}_{head} \\ \mathbf{H}_{tail} \end{bmatrix} \mathbf{w} = \mathbf{0}_{L+q-1-\nu,1}. \tag{11}$$

The set of equalizers that achieve CSE at GI length is a linear subspace given by the right kernel of $[\mathbf{H}_{head} \, \mathbf{H}_{tail}]$. If $\mathbf{H}$ is full column rank, then $[\mathbf{H}_{head} \, \mathbf{H}_{tail}]$ is also full column rank and its right kernel has the following dimension (freedom degree):

$$\eta = N_t \times q - (L+q-\nu-1) = q - (L-\nu-1) \tag{12}$$

where N_t represents the number of transmitted antennas in the proposed system $(N_t = 2)$.
Eq. 12 shows that the presence of only one antenna $N_t = 1$ is insufficient to reach the perfect CSE due the absence of the freedom degree $(\eta < 0)$. Starting from two antennas, we are able to achieve the perfect CSE and the kernel dimension increases when the length of equalizers increases. Consequently, for a large TEQ size, we would have enough freedom degrees η to achieve simultaneously CSE and enhancement of the spectrum of TIR. In other words, we are able to reduce the number of the taps with transforming the channel to a constructive channel rather than destructive.

5. THE PROPOSED METHOD

Let $\mathbf{N}$ be the right kernel of the matrix $\mathbf{H}_{out}$, and thus it will be also the kernel of the square matrix $\mathbf{R}_{out} = \mathbf{H}_{out}^H \mathbf{H}_{out}$. Using eigenvector decomposition, the matrix $\mathbf{R}_{out}$ with size $2q \times 2q$ can be decomposed as:

$$\mathbf{R}_{out} = \mathbf{P}\mathbf{\Delta}\mathbf{P}^H \tag{13}$$

where $\mathbf{P}$ is an eigenvector matrix and $\mathbf{\Delta}$ is a diagonal matrix contains the sorted eigenvalues of the matrix $\mathbf{R}_{out}$:

$$\mathbf{\Delta} = diag[\lambda_0, \lambda_1, \cdots, \lambda_{2q-\eta-1}, \mathbf{0}_{q,1}]$$

and the eigenvalues of $\mathbf{R}_{out}$ are sorted as: $\lambda_0 \geq \lambda_1 \geq \cdots \geq \lambda_{2q-\eta-1}$. Thus, the kernel matrix $\mathbf{N}$ can be easily obtained by the last η rows of the matrix $\mathbf{P}$. We note that each TEQ $\mathbf{w}$ that ensures the perfect CSE, can be written as:

$$\mathbf{w} = \mathbf{N}\mathbf{u} \tag{14}$$

where $\mathbf{u}$ is non null vector with length $2q-\eta$. The perfect CSE does not necessarily improve the received power at the tag as the received power depends on the multipath channel effect of the TIR. Therefore, the control of the TIR is required to ensure a high received power. The CW signal s_{cw} represents a tone and thus, its spectra amplitude is concentrated only on one frequency bin k_0 as illustrated in Fig. 3-b. In order to enhance the power of the continuous wave, the energy of the optimum filter in the FD should be concentrated around the frequency bins k_0 and thus, it has a narrow band. Due to the uncertainty principle, if the filter has a small window in FD then it has a large window in the TD and thus to reach a high improvement of the received power, the required filter should have a large window in the TD. But, the size of the window in the TD is reduced by using CSE. Therefore, we propose a new constraint that is able to achieve a good trade-off between channel shortening and received power enhancement. The TIR filter in FD can be expressed as:

$$\tilde{\mathbf{v}} = \mathbf{F}\mathbf{H}_1\mathbf{N}\mathbf{u} \tag{15}$$

where $\mathbf{F}$ is $N \times N$ Fourier matrix and $\mathbf{H}_1$ is an extension of the channel matrix $\mathbf{H}$ with size $N \times N$ instead of $(L + q - 1) \times 2q$. $\mathbf{N}$ is the kernel matrix and $\mathbf{u}$ is a vector with length κ that should be optimized to enhance the received power. Let $B_f = [k_0 - \frac{\delta_k}{2} + 1, k_0 + \frac{\delta_k}{2}]$ be the band of the desired window in the FD around the frequency bins k_0 . The TIR in FD in the band B_f can be expressed as:

$$\tilde{\mathbf{v}}_{B_f} = \mathbf{F}(B_f, :)\mathbf{H}_1\mathbf{N}\mathbf{u} \tag{16}$$

$$\begin{bmatrix} \tilde{v}(k_0 - \frac{\delta_k}{2} + 1) \\ \vdots \\ \tilde{v}(k_0) \\ \vdots \\ \tilde{v}(k_0 + \frac{\delta_k}{2}) \end{bmatrix} = \begin{bmatrix} \mathbf{F}(k_0 - \frac{\delta_k}{2} + 1, :) \\ \vdots \\ \mathbf{F}(k_0, :) \\ \vdots \\ \mathbf{F}(k_0 - \frac{\delta_k}{2}, :) \end{bmatrix} \mathbf{H}_1\mathbf{N}\mathbf{u}$$

Therefore, the optimized vector $\mathbf{u}_{opt}$ is computed under the constraint of maximization the energy of the TIR $\left|\tilde{\mathbf{v}}_{B_f}\right|^2$ and thus it is the solution of the following equation:

$$\mathbf{u}_{opt} = \underset{\mathbf{u} \in \mathbb{C}^{\kappa}}{\operatorname{argmax}} \ \mathbf{u}^H\mathbf{N}^H\mathbf{H}_1^H\mathbf{F}^H(B_f, :)\mathbf{F}(B_f, :)\mathbf{H}_1\mathbf{N}\mathbf{u} \tag{17}$$

The solution of Eq. 17 is given by the eigenvector associated to the highest eigenvalue of the matrix $\mathbf{T} = \mathbf{N}^H\mathbf{H}_1^H\mathbf{F}^H(B_f, :)\mathbf{F}(B_f, :)\mathbf{H}_1\mathbf{N}$. Finally, the proposed TEQ is obtained by: $\mathbf{w}_{opt} = \mathbf{N}\mathbf{u}_{opt}$

6. SIMULATIONS RESULTS

The 2×1 SIMO passive UHF RFID system in non-ideal environment with two multipaths transmission channels is used in this section. For each transmitted antennas, we use the same CW s_{cw} that is composed of 1024 samples with a carrier frequency index $k_0 = 300$. Each channel is randomly chosen from the mutlipath Rayleigh channel. More precisely, each path $h_{1,2}(i)$ follows a Complex Gaussian distribution with variance $\sigma_i = \frac{e^{-ai}}{\sum_{i=1}^{L} e^{-ai}}$ where $L = 20$ represents the total number of propagation paths and the parameter $a = 0.3$. Each TEQ $\mathbf{w}_{1,2}$ contains 24 taps and the size of the TIR equals to 8 and the length of the suitable band B_f around the carrier frequency index $k_0 = 300$ is fixed to $\delta_k = 32$. Fig. 5-a shows

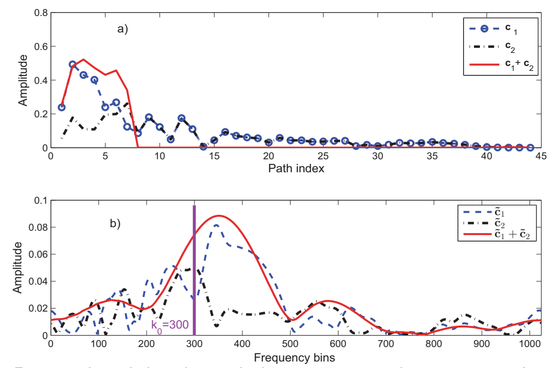

Fig. 5*: Results of the channel shortening equalizer in time domain a):* $\mathbf{c}_1 = \mathbf{w}^i \star \mathbf{h}$ *and in the Frequency domain b)*

that the proposed algorithm is able to shorten both channels form length $L = 20$ to TIR with length $\nu = 8$ with a delay $D = 0$.

In the FD, we observe clearly from Fig. 5-b that the proposed constraint leads to obtain a high energy of the TIR around the carrier frequency $k_0 = 300$ index. In the TD, the effect of high TIR energy around the carrier frequency index of the CW signal is shown in Fig. 6. We can observe that the absolute amplitude of the filtered CW signal using a TIR is increased. Whereas, the destructive effect f the multipath channels is well shown in Fig. 4. This is due to the low channel energy around the frequency carrier illustrated in Fig 3-b. The power of the filtered signal is averaged over $10,000$ realizations of multipath Rayleigh channels. We assume that there is no energy loss coming from the used RFID system. For different configurations of the filtered CW signals, we calculate the probability that the power of the received signal is less than a threshold l, $p(P_{y,s} \leq l)$. We observe from Fig. 7

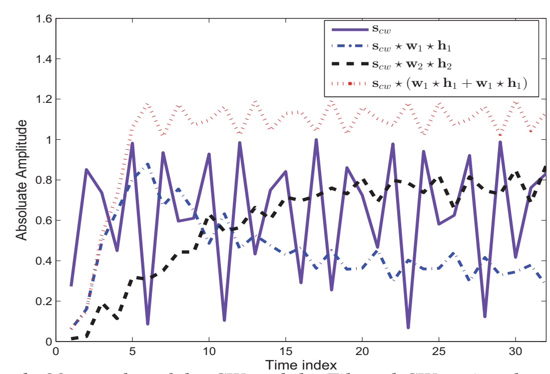

Fig. 6*: 32 samples of the CW and the Filtered CWs using the combined channel-equalizer* $\mathbf{c}_1$*,* $\mathbf{c}_2$ *and* $\mathbf{c}_1 + \mathbf{c}_2$ *in the time domain*

that for 95% of the cases, the proposed technique allowed us to maintain the power of the received signal at the tag superior or equal to the power of the CW signal. Moreover, 75% of the power of the received CW signal is increased at minimum two times of the initial energy. Whereas, without the proposed technique, only 50% of the cases the power of the received signal is superior to the initial energy if the SISO system is considered. Due to the presence of the diversity, the presence of two channels leads to maintain a reasonable energy on the tag. Several parameters affect the performance of

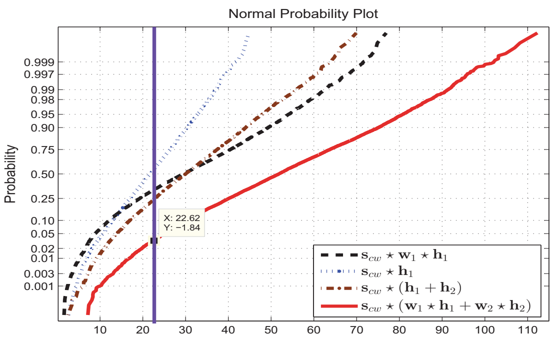

Fig. 7: *Probability that the power of the received signal is less than a threshold l,* $p(P_{y,s} \leq l$

the proposed method among them the size of the TIR. In fact, increasing the size of the TIR induces an augmentation of the freedom degree η and thus the energy control of the TIR in the FD is more efficient. We use the metric G_r to evaluate the performance of the proposed method, defined as follows:

$$G_r = \frac{P_y - P_{s_{\text{cw}}}}{P_{s_{\text{cw}}}} \times 100 \qquad (18)$$

where $P_{s_{\text{cw}}}$ is the power of the CW signal, defined as: $P_{s_{\text{cw}}} = \frac{\left(\sum |s(i)|^2\right)^{\frac{1}{2}}}{N}$ and P_y is the power of the filtered CW signal.
In Table 1, we observe that the presence of two channels (diversity) enhances the received power on the tag. The proposed method strengthen the previous results and a significantly enhancement the power of the received signal on the tag is observed mainly for a high value of ν. For example, when $\nu = 16$, the received power exceeds two times of original power of the CW signal. Moreover, we observe clearly that the power on the tag is increased when ν increases. This results is explained by the fact that a augmentation of freedom degree η leads to improve the control of the TIR in FD.

Table 1: *Effect of the TIR size on the received power at the RF tag*

	$\nu = 6$		$\nu = 8$		$\nu = 12$	
	P_y	G_r	P_y	G_r	P_y	G_r
$\mathbf{s} \star \mathbf{h}_1$	21.55	-4.73%	21.70	-4.6%	21.66	-4.24%
$\mathbf{s} \star (\mathbf{h}_1 + \mathbf{h}_2)$	30.17	33.38%	30.34	34.13%	30.23	33.64%
$\mathbf{s} \star \mathbf{c}_1$	24.96	10.34%	30.25	33.73%	38.98	72.33%
$\mathbf{s} \star (\mathbf{c}_1 + \mathbf{c}_2)$	39.91	76.44%	52.22	130.86%	72.03	218.44%

7. CONCLUSIONS

In this paper, we applied a channel shortening equalizer with a new constraint that exploits the knowledge of the continuous wave signal to a SIMO passive UHF RFID system. The proposed constraint enables the control of the TIR in order to obtain a high energy in FD around the carrier frequency of the CW signal. Due to this constraint, the received power at the tag has been maintained at high energy and its power exceeds two times the power of the CW signal. Moreover, the principle of the proposed technique can be used for any kind of CWs such as multi-tone signals and linear FM signals.

REFERENCES

[1] K. Finkenzeller, *RFID Handbook Fundamentals and Applications in Contactless Smart Cards and Identifications,*, ser. ISBN: 0-470-84402-7, L. John Wiley & Sons, Ed. Second Edition, 2004.

[2] R. Want, “An introduction to RFID technology,” *IEEE Pervasive Computing*, vol. 5, no. 1, pp. 25–33, January 2006.

[3] A. Bekkali, S. Zou, A. Kadri, M. Crisp, and R. Penty, “Performance analysis of passive UHF RFID systems under cascaded fading channels and interference effects,” *Wireless Communications, IEEE Transactions on*, vol. 14, no. 3, pp. 1421 – 1433, 2015.

[4] A. Lazaro, D. Girbau, and D. Salinas, “Radio link budgets for UHF RFID on multipath environments,” *IEEE Transactions on Antennas and Propagation*, vol. 57, no. 4, pp. 1241–1251, April 2009.

[5] P. Melsa, R. Younce, and C. Rohrs, “Impulse response shortening for discrete multitone transceivers,” *Communications, IEEE Transactions on*, vol. 44, no. 12, pp. 1662–1672, Dec 1996.

[6] T. B. Jabeur, K. Abed-Meraim, and H. Boujemaa, “Channel shortening techniques for differential encoded OFDM,” *Physical Communication*, vol. 5, no. 1, pp. 47 – 60, 2012.

[7] H. Gazzah and S. Mclaughlin, “An sos-based blind channel shortening algorithm,” in *Acoustics, Speech and Signal Processing, 2006. ICASSP 2006 Proceedings. 2006 IEEE International Conference on*, vol. 4, May 2006, pp. IV–IV.

[8] J. Walsh, R. Martin, A. Klein, N. Xenias, J. Pagnotta, and C. Johnson, “Necessary and sufficient conditions for perfect channel shortening and implications for interference mitigation,” in *Signal Processing Advances in Wireless Communications, 2004 IEEE 5th Workshop on*, July 2004, pp. 391–395.

[9] D. Tse and P. Viswanath, *Fundamentals of Wireless Communication*. Cambridge University Press, 2005.