On The Performance of WH-STC-OFDM and WH-SFC-OFDM in Non-Linear Time Variant Channels

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***Abstract*— Orthogonal Frequency Division Multiplexing is being widely used on today's digital communication standards, mainly because its robustness against multipath channels. Although, the high PAPR of the OFDM signal becomes a problem when a mobile communication takes place, because of the power limitation of the mobile unit. One technique that can be used to reduce the PAPR of OFDM signals is to apply the Walsh-Hadamard Transform on the data prior the IFFT. The aim of this paper is to analyze the peak-to-average power ratio PAPR performance of Orthogonal Frequency Division Multiplexing systems, using space-time and space-frequency diversity techniques on a non-linear time variant multipath channel.**

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) [1] is a wireless transmission technique for high data rate communication. The OFDM is robust against frequency selective fading, which makes this technique suitable for some wireless applications, such as: wireless computer networks [2] [3], digital television broadcasting [4] [5] and mobile communications [6]. However, OFDM has some drawbacks that limit its application and/or reduce its performance. Among these drawbacks, the high peak-to-average-power ratio (PAPR) [7] deserves special attention. The Gaussian distribution of the OFDM signal may produce high amplitudes that saturate the output of the power amplifier. This saturation clips the OFDM signal introducing in-band and out-band non-linear interferences [8]. The specialized literature presents several approaches to reduce the PAPR of OFDM signals on linear channel. [9] [10]. A simple and low complexity solution to PAPR is to apply the Walsh-Hadamard Transform (WHT) in the data to be transmitted, before the Inverse Fast Fourier Transform (IFFT) [11]. Obviously, the Inverse Walsh-Hadamard Transform (IWHT) must be applied to the symbols obtained at the output of the Fast Fourier Transform (FFT) in the receiver. This paper analysis the performance of a WH-OFDM system on a non-linear time variant channel. The theoretical and simulation results are compared [12]. In addition, the transmission diversity proposed by Alamouti [13] are integrated to the WH-OFDM, resulting on a WH-STC-OFDM (Walsh-Hadamard Space Time Coding Orthogonal Frequency Division Multiplexing) and a WH-SFC-OFDM (Walsh-Hadamard Space Frequency Coding Orthogonal Frequency Division Multiplexing) [12]. The performance of both approaches on a non-linear time variant channel are also obtained by computational simulations and compared with theoretical results.

This paper is organized as follow: Section II indroduces the basics on OFDM systems, while section III combines this technique with the Space Time Coding introduced in [13]. Section IV presents the integration of the OFDM, STC-OFDM and SFC-OFDM with the Walsh-Hadamard Transform. Section V presents the performance analysis of these systems on a time variant channel and, finally, Section VI presents final comments and conclusions.

II. BASICS ON OFDM

The main advantage of OFDM is the robustness against the impairments introduced by frequency selective channels. In a Non Line of Sight (NLOS) scenario, there are multiple paths between the transmitter and the receiver, as shown in Figure 1. The coherence bandwidth of a channel with J different paths is given by [12]

$$BW_C = \frac{1}{50\sigma_\tau}, \tag{1}$$

where

$$\sigma_\tau = \sqrt{T_2 - T_1^2} \tag{2}$$

and

$$T_1 = \frac{\sum_{j=0}^{J-1} \mathrm{a}_j^2 \mathrm{t}_j}{\sum_{j=0}^{J-1} \mathrm{a}_j^2} \qquad T_2 = \frac{\sum_{j=0}^{J-1} \mathrm{a}_j^2 \mathrm{t}_j^2}{\sum_{j=0}^{J-1} \mathrm{a}_j^2}. \tag{3}$$

The serial data stream of a OFDM system is split in N sub-streams. Each sub-stream modulates a subcarrier and the frequency spacing between two adjacent subcarriers is given by

$$\Delta f = \frac{R_s}{N} \tag{4}$$

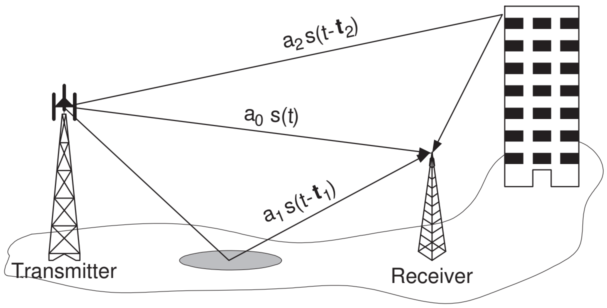

Fig. 1. Multipath channel scenario.

where R_s is the serial data rate. The bandwidth of each subcarrier is N times smaller than the bandwidth of the serial data stream. If N is large enough, the bandwidth occupied by each subcarrier can be smaller than the channel coherence bandwidth, which means that the fading channel can be considered flat for each subband, as shown in Figure 2.

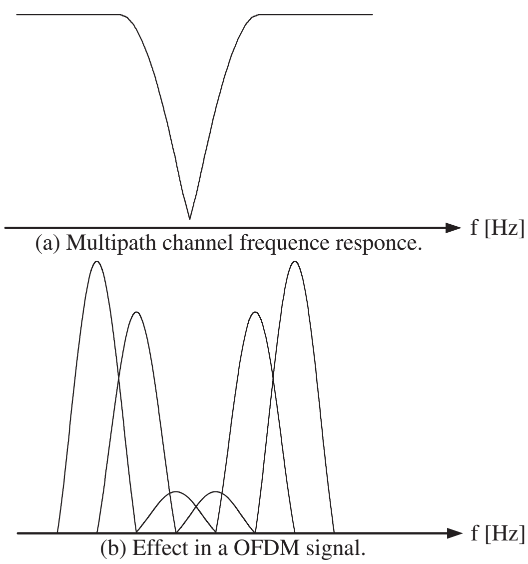

Fig. 2. Effect of a multipath channel in an OFDM signal.

Notice that the OFDM signal can be seen as a N terms Fourier series because it can be represented by the summation of N complex subcarriers weighted by N data symbols. Therefore Inverse Fast Fourier Transform (IFFT) can be used to generate the OFDM symbols. An N-point IFFT is applied to the N symbols, resulting in a complex baseband OFDM signal stated as [7]

$$s[m] = \frac{1}{2N} \sum_{n=0}^{N-1} c[n] e^{j\frac{2\pi n}{N}m}, \tag{5}$$

where $s[m]$ is m^{th} time-domain sample of the OFDM symbol, $0 \leq m \leq N-1$ and $c[n]$ is the in-phase and quadrature symbol transmitted on the n^{th} subcarrier, $0 \leq n \leq N-1$. As can be noticed from (5), the OFDM symbol is the IFFT of the data symbols, $c[n]$. The complex vector, $s[m]$ is transmitted using a in-phase and quadrature modulator. In order to recover the transmitted symbols, the OFDM signal is sampled and the FFT is applied, as presented in Figure 3

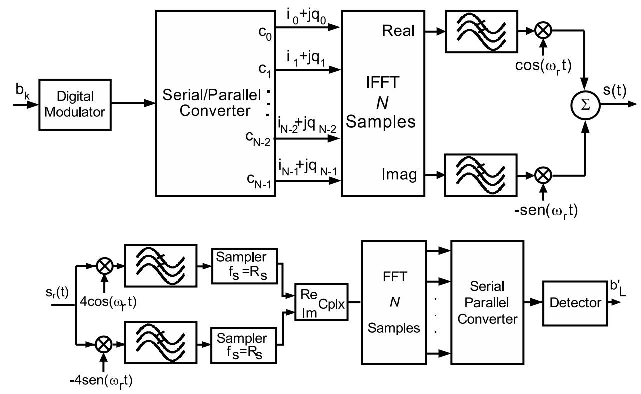

Fig. 3. Block diagram of an OFDM system.

III. Transmission Diversity

The scheme proposed by Alamouti [13] can be associated with OFDM in order to obtain transmission diversity in frequency selective channels. Basically, there are two different approaches to combine the transmission diversity with OFDM that will be shortly described.

A. STC-OFDM

A Space Time Coding OFDM scheme [14] can be achieved by applying the transmission matrix presented in Table I, where the n^{th} subcarrier of two adjacent OFDM symbols are used to construct a code-word.

TABLE I
STC-OFDM TRANSMISSION MATRIX.

	Antenna 0	Antenna 1
n^{th} subcarrier, i^{th} symbol	$c[n]$	$-c[n+1]^*$
n^{th} subcarrier, $(i+1)^{th}$ symbol	$c[n+1]$	$c[n]^*$

Figure 4 presents the block diagram of an STC-OFDM transmitter and Figure 5 presents the block diagram of an STC-OFDM receiver with a single antenna.

The data received at the n^{th} subcarrier of the i^{th} and $(i+1)^{\text{th}}$ OFDM symbol are respectively given by

$$\begin{aligned} S_{r_i}[n] &= c[n]H_0[n] - c^*[n+1]H_1[n] + W_i[n] \\ S_{r_{i+1}}[n] &= c[n+1]H_0[n] + c^*[n]H_1[n] + W_{i+1}[n] \end{aligned} \tag{6}$$

where $c[n]$ is the original data vector, $H_0[n]$ and $H_1[n]$ are the frequency response of the channels at frequency n and $W_i[n]$ is the amplitude spectrum of the noise at the n^{th} subcarrier and time instant iT. Notice that the channel frequency response is considered time invariant during the transmission of two adjacent OFDM symbols [14]. The STC decoder combines these received signals to obtain the diversity gain. Thus, the

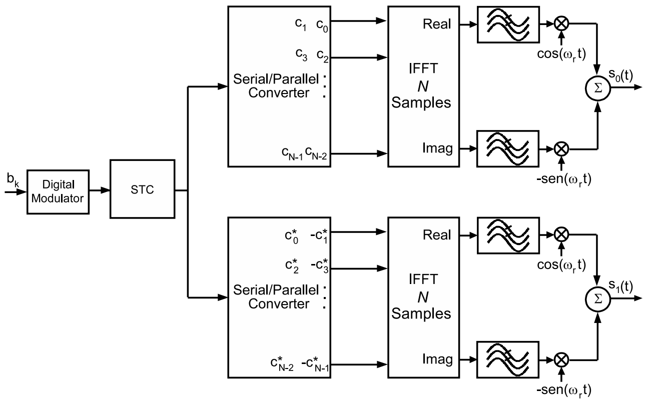

Fig. 4. Block diagram of an STC-OFDM transmitter.

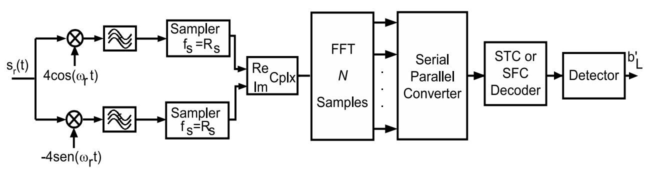

Fig. 5. Block diagram of an STC- or SFC-OFDM receiver.

TABLE II

SFC-OFDM TRANSMISSION MATRIX.

	Antenna 0	Antenna 1
n^{th} subcarrier, i^{th} symbol	$c[n]$	$-c[n+1]^*$
$(n+1)^{th}$ subcarrier, i^{th} symbol	$c[n+1]$	$c[n]^*$

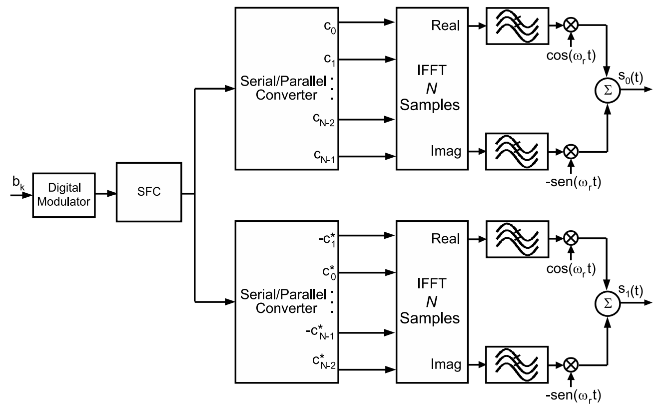

Fig. 6. Block diagram of an SFC-OFDM transmitter.

signal delivered to the detector corresponding with the n^{th} subcarrier of the i^{th} OFDM symbol is given by

$$\begin{aligned} d[n] =& H_0^*[n] \cdot S_{r_i}[n] + H_1[n] \cdot S_{r_{i+1}}^*[n] \\ =& \left(|H_0[n]|^2 + |H_1[n]|^2\right) c[n] + H_0^*[n] W_i[n] + \\ &+ H_1[n] W_{i+1}^*[n] \end{aligned} \tag{7}$$

and the signal corresponding to the n^{th} subcarrier of the $(i+1)^{th}$ OFDM symbol is given by

$$\begin{aligned} d[n+1] =& H_0^*[n] \cdot S_{r_{i+1}}[n] - H_1[n] \cdot S_{r_i}^*[n] \\ =& \left(|H_0[n]|^2 + |H_1[n]|^2\right) c[n+1] + \\ &+ H_0[n] W_{i+1}[n] - H_1[n] W_i^*[n] \end{aligned} \tag{8}$$

From (7) and (8) it is possible to state that the STC-OFDM scheme with one antenna at the receiver presents a diversity gain of order 2, which increases the performance of the system on frequency selective time variant channels. Notice that the total diversity gain is obtained if the channel frequency response can be considered time invariant during to at least two adjacent OFDM symbols.This means that the channel coherence time must be larger than two periods of an OFDM symbol [12].

B. SFC-OFDM

Space-Frequency Coding OFDM scheme [15] is similar to STC-OFDM. The main difference is that the code-word in the SFC-OFDM is transmitted in two adjacent subcarriers of the same OFDM symbol, as presented in Table II.

The block diagram of the SFC-OFDM transmitter is presented in Figure 6 and the block diagram of its receiver can be represented by Figure 5.

The signals received at the n^{th} and $(n+1)^{\text{th}}$ sub-carriers of the i^{th} OFDM symbol are respectively given by

$$\begin{aligned} S_{r_i}[n] &= c[n]H_0[n] - c^*[n+1]H_1[n] + W_i[n] \\ S_{r_i}[n+1] &= c[n+1]H_0[n+1] + c^*[n]H_1[n+1] + W_i[n+1]. \end{aligned} \tag{9}$$

The channel frequency response must remain unchanged for two adjacent sub-carriers, which means that $H_x[n] = H_x[n+1]$.

The received signals presented in (9) can be combined to obtain the diversity gain. The signal delivered to the detector corresponding to the n^{th} subcarrier of the i^{th} OFDM symbol is given by

$$\begin{aligned} d[n] =& H_0^*[n] \cdot S_{r_i}[n] + H_1[n] \cdot S_{r_i}^*[n+1] \\ =& \left(|H_0[n]|^2 + |H_1[n]|^2\right) c[n] + H_0^*[n] W_i[n] + \\ &+ H_1[n] W_i^*[n+1] \end{aligned} \tag{10}$$

and the signal corresponding with the $(n+1)^{th}$ subcarrier of the i^{th} OFDM symbol is given by

$$\begin{aligned} d[n+1] =& H_0^*[n] \cdot S_{r_i}[n+1] - H_1[n] \cdot S_{r_i}^*[n] \\ =& \left(|H_0[n]|^2 + |H_1[n]|^2\right) c[n+1] + \\ &+ H_0[n] W_i[n+1] - H_1[n] W_i^*[n] \end{aligned} \tag{11}$$

The channel must be time invariant during one single OFDM symbol. Notice that SFC-OFDM is more robust against Doppler effect than STC-OFDM, but it is more sensible to the frequency selectivity.

IV. Walsh Hadamard Transform and Transmission Diversity

The Walsh-Hadamard Transform (WHT) consists on multiplying a $2m$-length vector by a $2m \times 2m$ matrix given by

$$\Omega_{2m} = \begin{bmatrix} \Omega_m & \Omega_m \\ \Omega_m & -\Omega_m \end{bmatrix}, \qquad (12)$$

where $\Omega_1 = +1$. For instance the 2×2 Walsh-Hadamard Matrix is given by

$$\Omega_2 = \begin{bmatrix} \Omega_1 & \Omega_1 \\ \Omega_1 & -\Omega_1 \end{bmatrix} = \begin{bmatrix} +1 & +1 \\ +1 & -1 \end{bmatrix} \qquad (13)$$

The Walsh-Hadamard Transform [17] of a data symbol vector, $c[n]$, is given by

$$\vec{c}_\Omega = \vec{c} \times \frac{\Omega_N}{\sqrt{N}} \qquad (14)$$

where N is the length of the data symbol vector. The normalization constant $\sqrt{N}$ is used to keep the average energy of the vector unchanged. Notice from (14) that the WHT of the data symbols results on a vector with same length, where each coefficient is a linear combination of the N data symbols. Since the coefficients of the WHT are always "+1" and "-1", the IFFT of the vector $c_\Omega[n]$ presents low PAPR (Peak to Average Power Ratio) when compared to a conventional OFDM signal [11], as can be seen in Figure 7.

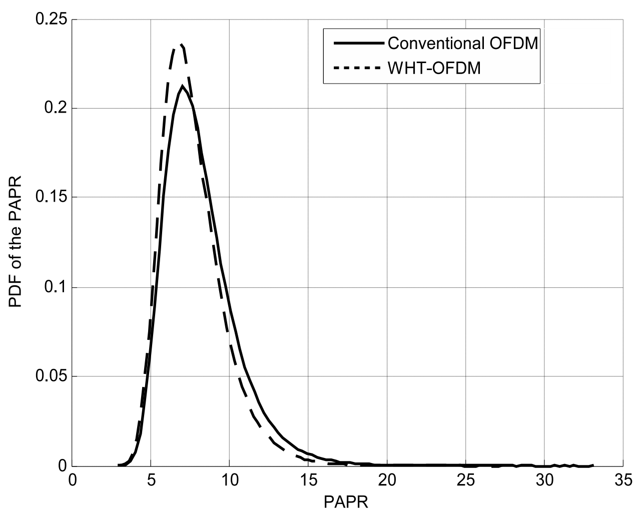

Fig. 7. Probability density function of the PAPR of a WHT-OFDM and conventional OFDM.

The Inverse Walsh-Hadamard Transform (IWHT) is also obtained by multiplying the transformed vector, $c_\Omega[n]$, by the normalized Walsh-Hadamard matrix, Ω_N, that is

$$\vec{c} = \vec{c}_\Omega \times \frac{\Omega_N}{\sqrt{N}}. \qquad (15)$$

The WHT can be easily integrated to the OFDM system, by applying the WHT to the data symbols prior the IFFT and applying its inverse after the FFT in the receiver, as shown in Figure 8.

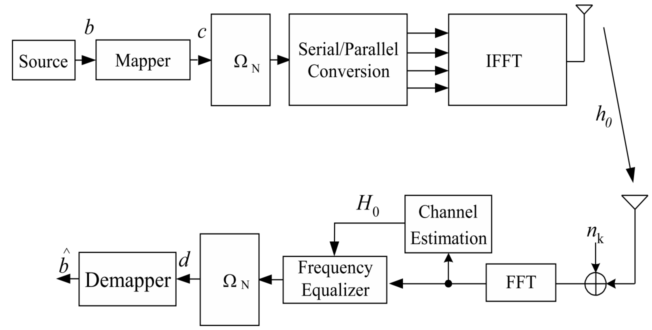

Fig. 8. Block diagram of a WHT-OFDM system.

The transmitted OFDM symbol is given by

$$s = \text{IFFT}\left(\frac{\vec{c}\ \Omega_N}{\sqrt{N}}\right) \qquad (16)$$

and the symbols delivered to the demapper are given by

$$\vec{d} = \frac{\text{FFT}\left(s * h_0 + \vec{w}\right)}{\hat{H}_0} \times \frac{\Omega_N}{\sqrt{N}} = \vec{c} + \frac{\vec{W}\ \Omega_N}{\sqrt{N}\ \hat{H}_0} \qquad (17)$$

where h_0 is the channel impulse response, $\hat{H}_0$ is the channel frequency response estimation and $\vec{w}$ is an AWGN vector of length N.

The STC or SFC OFDM scheme can also be integrated to the WHT, as shown in Figure 9.

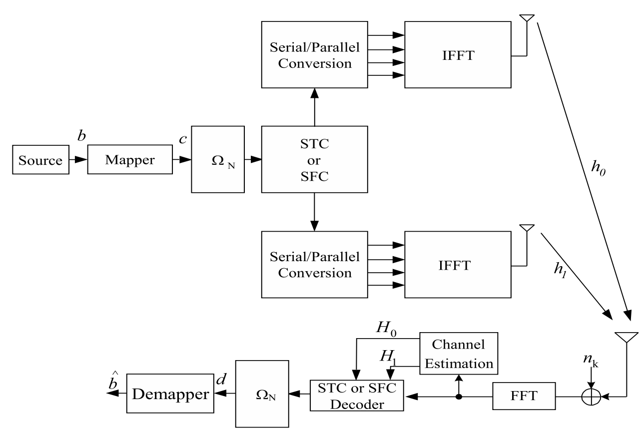

Fig. 9. Block diagram of a WHT-OFDM system with transmission diversity.

Assuming a WH-STC-OFDM system, the transmitted OFDM symbols at antenna 0 and antenna 1 in the time instant iT are respectively given by

$$\begin{aligned} s_{0,i} &= \text{IFFT}\left[\frac{\vec{c_i}\Omega_N}{\sqrt{N}}\right] \\ s_{1,i} &= \text{IFFT}\left[-\frac{\left(\vec{c}_{i+1}\Omega_N\right)^*}{\sqrt{N}}\right] = \text{IFFT}\left[-\frac{\vec{c^*}_{i+1}\Omega_N}{\sqrt{N}}\right], \end{aligned} \qquad (18)$$

where $\vec{c_i}$ is the transmitted symbol vector of length N at instant iT.

The OFDM symbols transmitted by the antennas 0 and 1 at

$(i+1)T$ are respectively given by

$$\begin{aligned} s_{0,i+1} &= \text{IFFT}\left[\frac{\vec{c}_{i+1}\Omega_N}{\sqrt{N}}\right] \\ s_{1,i+1} &= \text{IFFT}\left[\frac{(\vec{c}_i\Omega_N)^*}{\sqrt{N}}\right] = \text{IFFT}\left[\frac{\vec{c}_i^*\Omega_N}{\sqrt{N}}\right]. \end{aligned} \tag{19}$$

The received signals at iT is given by

$$\begin{aligned} r_i &= \text{FFT}\left[s_{0,i} * h_0 + s_{1,i} * h_1 + \vec{w}_i\right] \\ &= \frac{\vec{c}_i\Omega_N}{\sqrt{N}} \times H_0 - \frac{\vec{c^*}_{i+1}\Omega_N}{\sqrt{N}} \times H_1 + \vec{W}_i, \end{aligned} \tag{20}$$

while the received signal at $(i+1)T$ is given by

$$\begin{aligned} r_{i+1} &= \text{FFT}\left(s_{0,i+1} * h_0 + s_{1,i+1} * h_1 + \vec{w}_{i+1}\right) \\ &= \frac{\vec{c}_{i+1}\ \Omega_N}{\sqrt{N}} H_0 + \frac{\vec{c}_i^*\ \Omega_N}{\sqrt{N}} H_1 + \vec{W}_{i+1}. \end{aligned} \tag{21}$$

Eq. (20) and (21) can be combined to obtain the diversity gain. The received vector at the input of the IWHT block at instant iT is given by

$$\begin{aligned} \vec{d}_{\Omega_i} &= H_0^* r_i + H_1 r_{i+1}^* \\ &= \left(|H_0|^2 + |H_1|^2\right)\frac{\vec{c}_i\Omega_N}{\sqrt{N}} + H_0^*\vec{W}_i + H_1\vec{W}_{i+1}^* \end{aligned} \tag{22}$$

and the received vector at the input of the IWHT block at instant $(i+1)T$ is given by

$$\begin{aligned} \vec{d}_{\Omega_{i+1}} &= H_0^* r_{i+1} - H_1 r_i^* \\ &= \left(|H_0|^2 + |H_1|^2\right)\frac{\vec{c}_{i+1}\Omega_N}{\sqrt{N}} + H_0^*\vec{W}_{i+1} + H_1\vec{W}_i^*. \end{aligned} \tag{23}$$

Applying the IWHT in (22) and (23), respectively, leads to

$$\begin{aligned} \vec{d}_i &= \left(|H_0|^2 + |H_1|^2\right)\vec{c}_i + \left(H_0^*\vec{W}_i + H_1\vec{W}_{i+1}^*\right) \times \frac{\Omega_N}{\sqrt{N}} \\ \vec{d}_{i+1} &= \left(|H_0|^2 + |H_1|^2\right)\vec{c}_{i+1} + \left(H_0^*\vec{W}_{i+1} + H_1\vec{W}_i^*\right) \times \frac{\Omega_N}{\sqrt{N}}. \end{aligned} \tag{24}$$

The analysis of the WH-SFC-OFDM is analogous to that presented above. On a linear channel, the performance of WH-STC-(SFC)-OFDM is equivalent to the performance of STC-(SFC)-OFDM. However, the performance of the WH-STC-(SFC)-OFDM is expected to perform better on a non-linear channel, due the reduction in the PAPR introduced by the use of WHT.

V. Performance Analysis

The PAPR reduction of the OFDM symbol reflects on the performance of the system on non-linear channels, where the high peaks of the signal are clipped by the power amplifier [8]. The performance of a conventional OFDM system in a non-linear frequency selective and time-variant channel can be approximately estimated by (25) [12], where l is the clipping threshold, σ_n^2 is the noise variance, σ_r^2 is the variance of the orthogonal complex Gaussian that generates the Rayleigh distribution [16], 2ν is the minimum distance between two adjacent symbols of the employed constellation and $\bar{\mu}$ is the average number of neighbors in the constellation. Figure 10 shows a performance comparison of a WH-OFDM system on a mobile non-linear channel conventional OFDM system with a conventional OFDM system. A 64-QAM constellation is used and the clipping threshold, l is made equal to three.

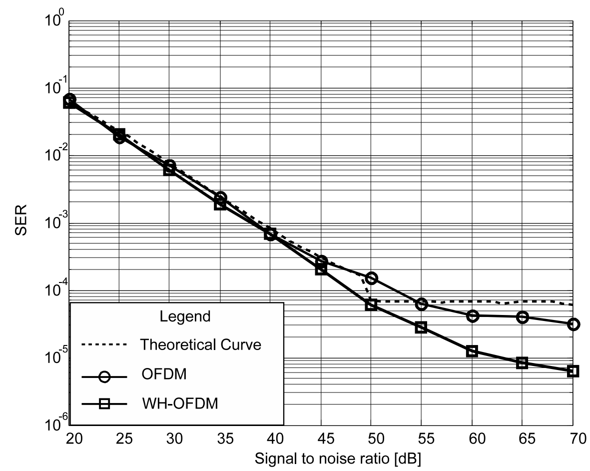

Fig. 10. Performance of a 64-QAM OFDM and 64-QAM WH-OFDM systems in a non-linear mobile channel with $l = 3$.

From Figure 10, it is observed that the WHT reduces the error floor of the system, resulting in a better performance of the WH-OFDM on non-linear channels. However, this performance gain is only verified at high SNR (above 45 dB), which may lead to conclude that the improvement in performance gain does not payoff the complexity introduced by the WHT. For STC-OFDM, the scenario is quit different, as can be seen in Figure 11. Again, a 64-QAM constellation and clipping threshold equal to three are used in the simulation. Figure 11 shows that the WHT sensibly reduces the error floor of the system for SNR above 25dB. The performance gain is also higher than the gain obtained without the STC, which means that the WHT is more effective when combined to STC-OFDM.

From Figure 12 same conclusion can be drawn for the WH-SFC-OFDM. However, the performance gain is even higher than the that obtained with WH-STC-OFDM. Although the performances of the WH-SFC-OFDM and WH-STC-OFDM are almost the same on a non-linear time-variant channel, the performance of SFC-OFDM is poorer than that for STC-OFDM. This happens because SFC-OFDM is high sensible to Intra-carrier Interference (ICI) introduced by the clipping. This behavior was expected because the reduction of the PAPR also reduces the clipping introduced by the power amplifier.

$$p_e \approx \frac{2\mathrm{Q}(l)\bar{\mu}}{\sqrt{2\pi}\,\sigma_n\sigma_r^2}\int_0^\infty\int_{-\infty}^\infty r\exp\left(-\frac{n^2}{2\sigma_n^2}-\frac{r^2}{2\sigma_r^2}\right)\mathrm{Q}\left\{\left[\frac{r\left(\nu-n\right)}{\frac{2}{\sqrt{3N}\,\pi l^2}}\right]^{\frac{1}{3}}\right\}\mathrm{d}n\mathrm{d}r+\frac{\bar{\mu}}{\sigma_r^2}\int_0^\infty r\exp\left(-\frac{r^2}{2\sigma_r^2}\right)\mathrm{Q}\left(\frac{r\nu}{\sigma_n}\right)\mathrm{d}r \tag{25}$$

Since the clipping causes ICI which results on degradation of the space-frequency codeword for the SFC, it is expected that the reduction of this non-linearity improves the performance of the SFC-OFDM system.

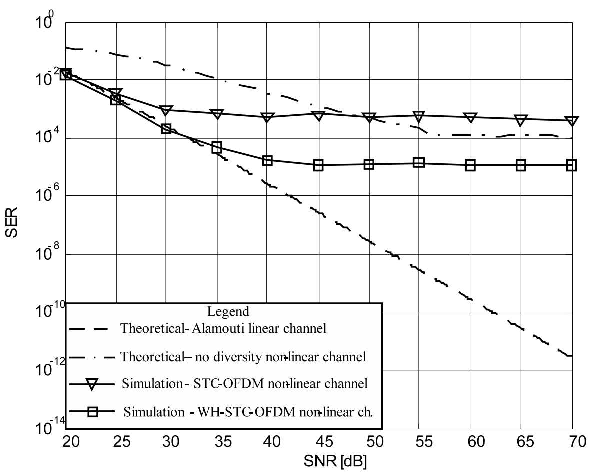

Fig. 11. Performance of a 64-QAM STC-OFDM and 64-QAM WH-STC-OFDM systems in a non-linear mobile channel with $l = 3$.

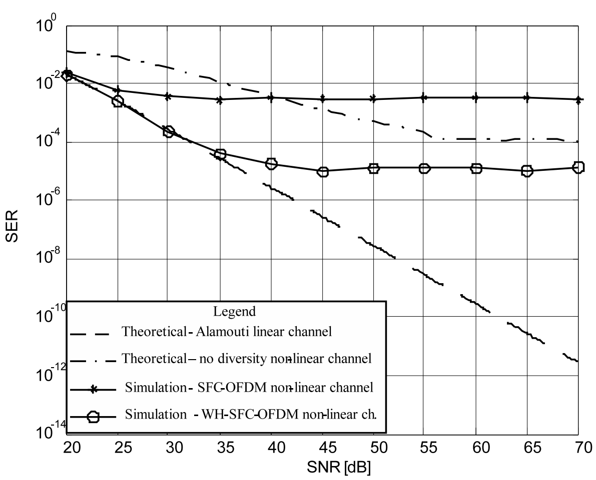

Fig. 12. Performance of a 64-QAM SFC-OFDM and 64-QAM WH-SFC-OFDM systems in a non-linear mobile channel with $l = 3$.

VI. Conclusions

The high PAPR of OFDM symbols is one of the main causes of poor performance on non-linear channels. The reduction of the PAPR can be achieved by several different approaches, for instance, by applying the Wash-Hadamard Transform on the information symbols prior the IFFT.

This paper has compared the performance of conventional OFDM and WH-OFDM, as well STC-(SFC)-OFDM and WH-STC-(SFC)-OFDM, on a mobile non-linear channel. The performance gain obtained by the WHT considering just OFDM is smaller than the performance gain obtained when considering the STC. The gain is even larger when the SFC is considered. These results lead to the conclusion that the Walsh-Hadamard Transform shall be used with some diversity OFDM schemes, since the gain obtained with no diversity OFDM scheme is relevant only for high signal to noise ratio.

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