Influence of time and processing mismatches between phase and envelope signals in linearization systems using Envelope Elimination and Restoration, application to hiperlan2

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Abstract — **This paper is a theoretical and experimental study of the influence of time and processing mismatches between the envelope and the phase of an OFDM in an Envelope Elimination and Restoration (EER) linearization system. We give the theoretical expression of the power spectral density of the distorted signal in the case without intersymbol interference at the emitter. In the general case, we give the average power of the distortion as a function of the time mismatch. We also propose an approximation of the distorted signal as a delayed version of the original signal, that explains the observed rotations of the constellations at the receiver on the different OFDM carrier. Others effects of EER principle are also studied, such as limitation of envelope and phase bandwidth and non-linearity of the envelope restoration command.**

I. INTRODUCTION

The Envelope Elimination and Restoration (EER) technique was introduces by Kahn [1] for linear amplification of non-constant envelope signals. We study, in this paper, the specifications of a new EER architecture [2] (Fig. 1) for OFDM signal. The principle consists in splitting the modulated signal $x(t)$ in a product of 2 signals $x(t)=a(t)p(t)$, where $a(t)$ is the envelope of $x(t)$ and $p(t)$ is a constant-envelope phase modulated signal. The high efficiency power amplifier (PA) is fed by the RF phase signal via a modulation loop. The envelope (low frequency signal) is restored using the supply voltage of the last stage of the PA.

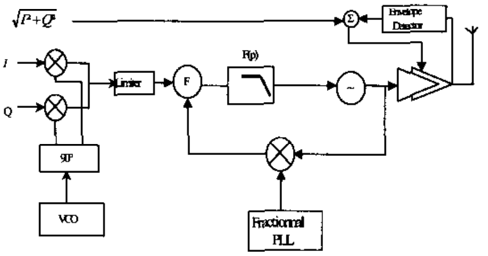

Fig. 1. Envelope Elimination and Restoration principle with translation loop

The first analyzed imperfection concerns the non synchronization of envelope and phase signals. Then we will expose the influence of envelope and phase signal filtering in the transceiver. The last studied point is the influence of the envelope restoration non-linearity.

II. THEORETICAL ANALYSIS

In an OFDM modulator, each block of N symbols is transformed by an inverse FFT (IFFT). The samples A_k at the output of the IFFT, are send serially. they represent the OFDM signal sampled at the symbol period T_S. This output is oversampled at the sampling period T_e and filtered by a low-pass filter with impulse response $s(t)$ before digital to analog conversion, thus generating a complex signal $z(t)$. The oversampling can also be realized by zero-padding the IFFT and in that case s(t) has a rectangular transfer function in frequency. Let $\rho(t)$ be the amplitude and $\Phi(t)$ the phase of $z(t)$: $z(t)=\rho(t)exp(j\Phi(t))$.

$$z(t)=\sum_k A_k s(t-kT_S)=\sum_k |A_k| e^{j\Phi_k} s(t-kT_S).$$

When there is a time mismatch θ between the envelope and the phase, the resulting distorted signal $z_\theta(t)$ can be written: $z_\theta(t)=\rho(t-\theta)e^{j\Phi(t)}$.

A. Case without intersymbol interference at the emitter

We first study the case where the shaping filter $s(t)$ does not introduce intersymbol interference (because the duration of its impulse response is shorter than T_S). In that case, we were able to do the complete theoretical analysis of the influence of the time mismatch θ. We give the results for the power spectral density (psd) of the distorted signal $z_\theta(t)$.

The principle of the demonstration consists, when $\theta<T_S$, in splitting $z_\theta(t)$ in 2 parts $z_1(t)$ and $z_2(t)$ corresponding to the product of $z_\theta(t)$ with 2 windowing periodical signals $w_1(t)$ et $w_2(t)$. The signals $w_1(t)$ and $w_2(t)$ are made of rectangular pulses with respective durations T_S-θ and θ repeated with the period T_S. The signals $z_1(t)$ and $z_2(t)$ are equal to: $z_1(t)=w_1(t)\, z_\theta(t)$ and $z_2(t)=w_2(t)\, z_\theta(t)$, with $w_1(t)+w_2(t)=1$ and $z_1(t)+z_2(t)=z_\theta(t)$.

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The signals $z_1(t)$ and $z_2(t)$ are Pulse Amplitude Modulation (PAM) signals, the power spectral densities of which can be calculated by the Bennett formula. For $z_1(t)$, the shaping pulse $s_1(t)$ is the product of $s(t\text{-}\theta)$ with a rectangular pulse of duration $T_S\text{-}\theta$ noted $R_{TS\text{-}\theta}\,(t\text{-}\theta)$. For $z_2(t)$, the shaping pulse $s_2(t)$ is the product of $s(+T_S\text{-}\theta)$ with a rectangular pulse of duration θ noted $R_\theta(t)$. The signals z_1 and z_2 can be written:

$$z_1(t)=\sum_k A_k s_1(t-kT_S),\; z_2(t)=\sum_k |A_{k-1}|e^{j\Phi_k}s_2(t-kT_S).$$

When $\theta > T_S$, the same type of decomposition can be applied but θ must be replaced by θ_m equal to θ modulo T_S.

The power spectral density $S_{z_\theta}(f)$ of $z_\theta(t)$ can be calculated from this decomposition. It can be written:

$$S_{z_\theta}(f)=S_{z_1}(f)+S_{z_2}(f)+S_{z_{12}}(f),$$

where $S_{z_1}(f)$ and $S_{z_2}(f)$ are the psd of $z_1(t)$ and $z_2(t)$ and $S_{z_{12}}(f)$ is the Fourier transform of the intercorrelation of $z_1(t)$ and $z_2(t)$. With the hypothesis that the samples A_k at the output of the IFFT are uncorrelated, centered with variance σ_A^2, the different terms are easily calculated:

$$S_{z_1}(f)=\frac{\sigma_A^2}{T_S}|S_1(f)|^2,\; S_{z_2}(f)=\frac{\sigma_A^2}{T_S}|S_2(f)|^2$$
$$S_{z_{12}}(f)=\frac{2E(|A_K|)^2}{T_S}\Re\left(S_1(f)S_2^*(f)\right). \tag{1}$$

expressions where $S_1(f)$ and $S_2(f)$ are the Fourier transform of $s_1(t)$ and $s_2(t)$.

The Fig. 2 shows the theoretical psd obtained by (1) and the experimental estimation of this psd using a technique of averaging of periodograms, in the case of a ratio θ/T_S equal to 10% and a filter with impulse response s(t) equal to a hamming window of duration T_S. .

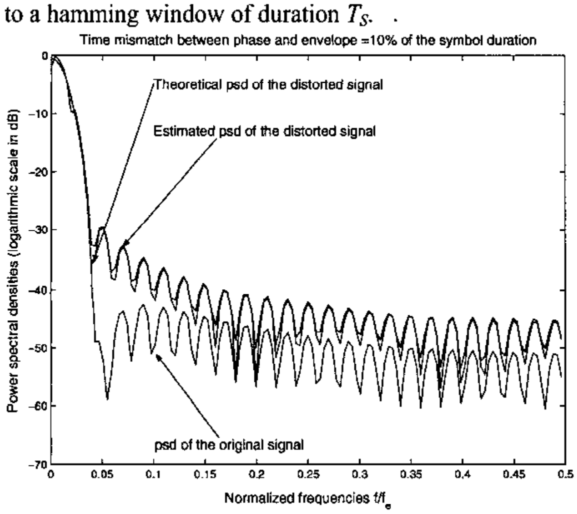

Fig. 2. Theoretical and experimental power spectral densities

B. General case

We now consider the general case, where the impulse response of the shaping filter $s(t)$ may have a duration longer than the duration period T_S.

Calculation of the average power of the distortion introduced by the time mismatch

The distorsion $e(t)$ introduced by the time mismatch θ is defined by $e(t)=z(t)\text{-}z_\theta(t)$. Its average power $B(\theta)$ is equal to:

$$B(\theta)=E\left(|e(t)|^2\right)=2\left(R_\rho(0)-R_\rho(\theta)\right), \tag{2}$$

where $R_\rho(t)$ is the autocorrelation of the amplitude $\rho(t)$.

For OFDM modulations, when the number of subcarriers is large enough, one can consider that the modulated signal is nearly gaussian (central limit theorem). When θ is large enough for $\rho(t)$ and $\rho(t\text{-}\theta)$ to be uncorrelated, the average power of the distortion tends towards $B=\sigma^2(4-\pi)$ and the signal to noise ratio tends towards $SNR=\dfrac{E\left(|z(t)|^2\right)}{E\left(|e(t)|^2\right)}=\dfrac{2}{(4-\pi)}$.

Fig. 3 shows $1/SNR(\theta)$ given by (2) and the EVM at the receiver obtained by simulation.

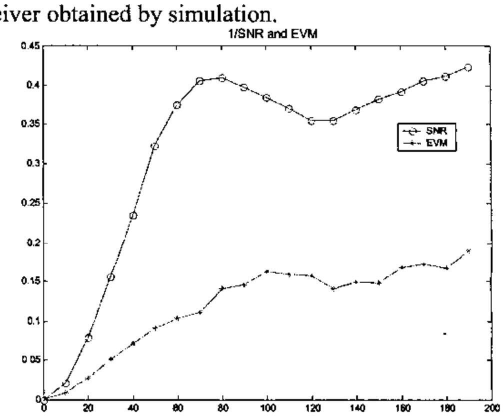

Fig. 3. Theoretical 1/SNR and experimental EVM.

Approximation of the distorted signal and synchronization

When the time mismatch θ is small enough, one can approximate $z_\theta(t)$ by $z(t)$ delayed by τ, ie. by $z(t\text{-}\tau)$ using a limited development of order 1 around the time instant $t\text{-}\tau$.

The approximation error is $\varepsilon(t)=z_\theta(t)\text{-}\,z(t\text{-}\tau)$. Its average power is minimized for the optimal choice of the delay:

$$\tau=E\left(\frac{\rho'^2(t)}{|z'(t)^2|}\right).$$

And for this choice, it is equal to:

$$E\left(|\varepsilon(t)|^2\right) = \theta^2 \frac{E\left(\rho'^2(t)\right)E\left(\rho^2(t)\phi'^2(t)\right)}{E\left(\rho'^2(t)+\rho^2(t)\phi'^2(t)\right)}.$$

This delay τ, the value of which depends on the shaping filter *s(t)*, corresponds at the receiver, if the synchronization is not modified, to a rotation of the constellation on the carrier f_k by an angle $\exp(j2\pi f_k\tau)$. The Fig. 4 illustrates this result for the constellations obtained on 2 different subcarriers (N=64).

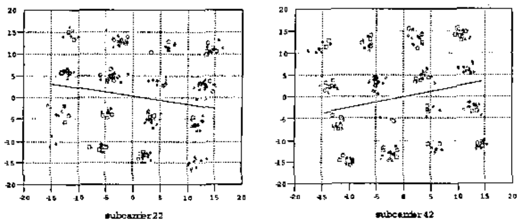

Fig. 4. Rotation of the constellation on 2 subcarriers for θ=10ns.

III. IMPERFECTION SOURCES AND ANALYSIS BASED ON HIPERLAN2 STANDARD

Hiperlan2 transmitter performances are specified in the standard with different parameters from which we will retain here EVM, specified for a 16QAM to 11.22%, and the output spectrum mask. This mask is defined relatively to the carrier power taken in ±9MHz bandwidth. Important points are –20dBc at 11MHz frequency offset from the nominal carrier frequency, -28dBC at 20MHz frequency offset and -40dBc above 30MHz frequency offset. All these points are taken in 1MHz resolution bandwidth.

All results presented are related to these specifications. The OFDM modulation used in the simulations uses N=64 subcarriers, and a 16-QAM alphabet.

A. Time mismatch between envelope and phase signals

The time mistmatch parameter is linked to the synchronization lost between the envelope and phase signals, as they are both following to different paths. This time mismatch generates a rotation of the emitted constellation proportional to the subcarrier frequency offset from the nominal carrier frequency and also proportional to the delay itself. Fig 5 presents the values of the phase rotations of the 64 subcarriers for two different delays, 10ns and 2ns. For 10ns, the phase shift can reach 8.5 degrees and EVM is about 30%. For 2ns, the phase shift is lower than 2 degrees and EVM is about 6.5%. Fig. 6 is the output spectrum obtain for different time delays and displays prominently that it has to be kept under 2ns to fulfill the requirements of the standard.

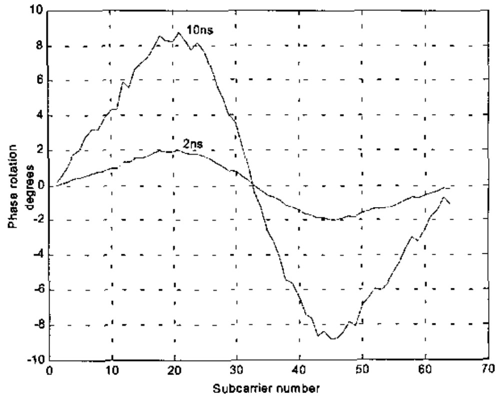

Fig. 5. Subcarrier rotation for 10ns and 2ns delay mismatch

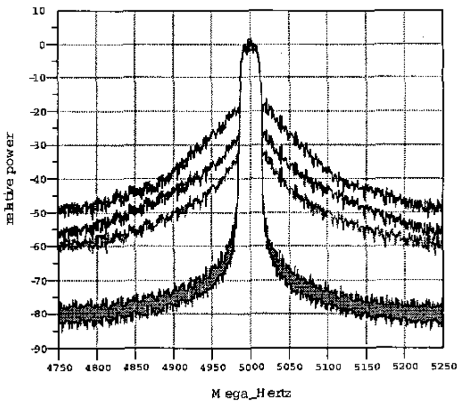

Fig. 6. Output spectrum for various delay mismatch : 2ns, 4ns and 10ns

B. Influence of envelope filtering

The envelope signal, in EER principle, is usually efficiency amplified with a class D power amplifier, and then low pass filtered. In order to define precisely the useful envelope bandwidth, we made simulations with a theoretical raised cosine filter, with zero roll-off. The filter cut off frequency was taken from 30MHz to 100 MHz. The most important effect of this filtering is on output spectrum with important spectral raising above 20MHz offset from the carrier as presented on Fig.7.

The spectrum mask is just respected with a 40MHZ filter bandwidth. The envelope filter will then have to be larger than 40MHz if we want to fulfill requirement when adding all impairments. EVM is barely degraded since with a 40MHz bandwidth filter, it is about 1%.

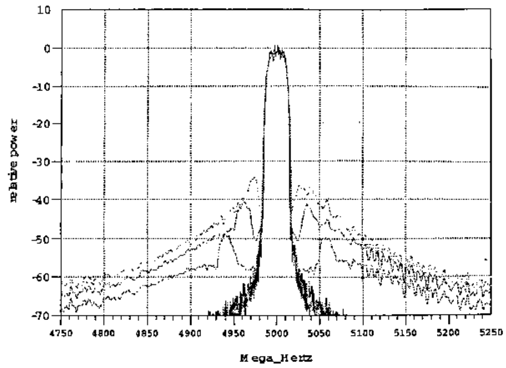

Fig. 7. Output spectrum for different envelope filter bandwidth : 30MHz, 40MHz and 60MHz

C. Influence of phase filtering

Modulation loop principle is efficient for translating phase information with minimum noise. The constant envelope phase modulated signal is presented at the input of the power amplifier with very good noise performances. The signal envelope is filtered by the loop filter and it is, as for envelope signal, important to evaluate the useful bandwidth of this signal. For this, we proceeded the same way as for previous simulation, with a raised cosine filter.

The filter bandwidth is 60MHz, 80MHz, 100MHz and 150MHz. The loop filter as no real effect on EVM (less than 1%) but important spectral raisings appear as shown on Fig.8.

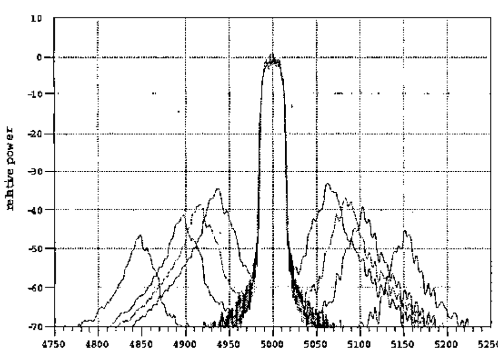

Fig. 8. Output spectrum for different phase signal filter bandwidth : 30MHz, 40MHz and 60MHz

The limitation of the useful phase bandwidth is about 100MHz, but with other impairments, the modulation loop which will have to present a loop bandwidth for the phase signal of more than 150MHz.

D. Influence of the transfer function distortion : output power versus Vcc.

EER principle suffers form non ideal transfer function between the output power and the supply voltage used for restoring the envelope. In order to take into account that phenomenon, we chose to model it with an AM/AM curve. The AM/AM compression was then simulated with a third order approximation of the power amplifier calculated with a 1dB compression point (P1dB). The envelope power has a maximum instantaneous value of –3dBm. The P1dB was then taken at –3dBm, -7dBm and -9dBm. Fig.9 present the spectral effect of this distortion. Unlike previous simulations, the most important impact of this default can be seen on EVM performances with 5% reached when the P1dB is –3dBm.

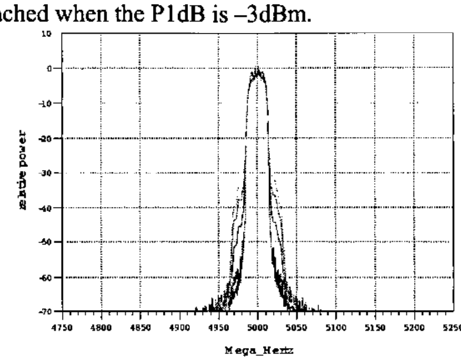

Fig. 9. Output spectrum for compression at -3, -7 and –9dBm

V. CONCLUSION

We have analyzed the influence of different sources of impairments in an EER architecture on OFDM signals. We have studied the power spectral density of the distorted signal, the average power of the distortion and given an approximation of the distorted signal explaining the rotation of constellations on the different subcarriers.

We have applied this analysis to the hiperlan2 standard. We have shown that the time mismatch between envelope and phase signals cannot exceed 2ns, the useful bandwidth cannot be smaller than 40 MHz and 100 MHz respectively for the envelope and phase signals. A small distortion of the transfer function output power versus Vcc can be also tolerated.

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