Per block pre-distortion of a multi-carrier non-linear satellite communication channel

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***Abstract*—In this paper, we propose a pre-distortion algorithm for a two carriers non-linear satellite communication channel. In the considered scenario, two linearly modulated carriers are amplified by the same non-linear power amplifier aboard the satellite. Due to the presence of the shaping filter at the transmitter side and the receiver filter (and possibly interference with filters aboard the satellite), the received signal is affected by non-linear intersymbol interference (ISI) but also non-linear adjacent channel interference (ACI). We consider the case where the two signals are transmitted from the same gateway, so that joint pre-distortion of the two signals is possible. Simulations for the 8PSK modulation show that the algorithm is able to greatly reduce the impact of the non-linear interference.**

I. Introduction

Compensation of the non-linear satellite channel with memory has been already well studied for the single carrier case. The compensation can be applied at the transmitter side, which is called pre-distortion or at the receiver side, which is called equalization. One main advantage of equalization is that it does not require the channel knowledge. Adaptive equalization of the non-linear channel has for instance been presented in [1]. Moreover, the error correcting code can be used to better mitigate the non-linear interference in so-called turbo-equalization structures, as for instance done in [2]. In communication systems with a star topology (such as broadcast or broadband systems), the forward link involves a transmitter gateway and a number of user receivers. In such context, considering that the channel can be assumed to be reasonably well-known, pre-distortion has the advantage that it does not increase the complexity of all satellite terminals. Non-linear pre-distortion after the shaping filter would create out of band interference, which has to be avoided in practical systems. The pre-distortion is thus applied here to the transmitted symbols prior to the shaping filter. The pre-distortion algorithm can consist in feeding a Volterra system with the initial symbols as in [3]. The Volterra model, described in [4], is an analytical representation of a non-linear system with memory. In other solutions, the pre-distortion algorithm can also rely on look-up tables with pre-computed values, as in [5] and in [6] for high order modulations.

The literature on the compensation of a multi-carrier satellite channel is much narrower. The term multi-carrier refers to the case where more than one modulated signal is amplified by the same power amplifier, in opposition to the single carrier case. Equalization (and turbo-equalization) for the multi-carrier case has already shown good result, as demonstrated in [7]. It is then necessary for the receiver to demodulate all carriers to apply the proposed equalization algorithm. Similarly, multi-carriers pre-distortion will be only possible for scenarios where all carriers sharing the channel are available at the same location (e.g. at the transmit gateway). In [8], pre-distortion of the multi-carrier non-linear communication satellite channel has been proposed based on reduced channel model using memory polynomials. Memory polynomials can be seen as a simplified Volterra model, as explained in [9]. The pre-distortion algorithm in [8] is an extension of the order p compensation, described in [10] for the single carrier case. In this pre-distortion method, all interference terms up to order p are cancelled. However, it creates higher order terms interference, which in case of high non-linearities, may be more significant than the cancelled terms, as shown in [11]. Another multi-carrier pre-distortion algorithm could also be the extension to multiple carriers of the pre-distortion method based on the use of look-up tables presented in [6]. However, due to the ACI, more entries have to be taken into account in the look-up table so that the algorithm would quickly become impractical.

In this paper, we propose a new pre-distortion algorithm for a multi-carriers non-linear satellite communication channel. We focus here on a scenario with two carriers. The proposed algorithm is an iterative algorithm which determines the sequence of pre-distorted symbols so that the square error between the received and the initial symbols is jointly minimized for the two carriers. Simulations show that this algorithm is capable of greatly reducing the impact of the ISI and the ACI for the two carriers case.

II. System model

The system model is represented in Fig. 1. At the transmitter side, two independent data sources produce binary outputs which are encoded with a block code with message length K, code length M and code rate $\frac{K}{M}$. The coded bits are interleaved and mapped onto a linear modulation at the rate T_s^{-1}. The

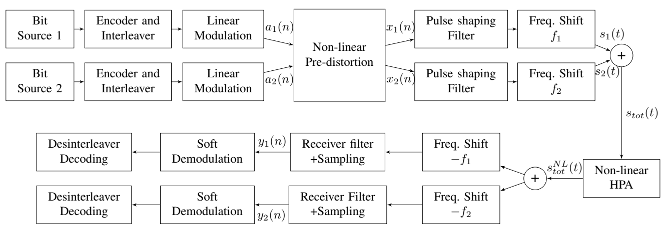

Fig. 1. System model

obtained symbols are denoted as $a_m(n); m = 1, 2$. The index m refers thus to the number of the considered carrier. We denote by N the number of symbols in each block. The pre-distortion block transforms each block of symbols into a block of pre-distorted symbols of equivalent size. The pre-distortion algorithm will be detailed in next section. The pre-distorted symbols, which we denote as $x_m(n); m = 1, 2$, are then converted to an analog signal $x_m(t)$:

$$x_m^{analog}(t) = \sum_{k=-\infty}^{\infty} x_m(k)\delta(t - kT_s) \tag{1}$$

and filtered by a pulse shaping filter $p_m(t); m = 1, 2$:

$$s_m(t) = \int_{\tau=-\infty}^{\infty} x_m^{analog}(\tau) * p_m(t - \tau) \tag{2}$$

Each signal is then translated to a central frequency f_m and the obtained signals are mixed together to produce:

$$s_{tot}(t) = s_1(t)e^{j2\pi f_1 t} + s_2(t)e^{j2\pi f_2 t} \tag{3}$$

Note that the notations are intentionally very similar to [12], except that we consider that the signals are mixed inside the same gateway, and do not come from two physically different transmitters. Therefore, different propagation times or different carrier phases do not have to be taken into account.

The power amplifier aboard the satellite can be seen as a memoryless non-linear device. It can be shown that its input-output relationship is given by:

$$s_{tot}^{NL}(t) = \sum_{i=0}^{\infty} \gamma^{2i+1} s_{tot}(t)|s_{tot}(t)^i|^{2j} \tag{4}$$

where γ^{2i+1} are complex coefficients. At the receiver side, each carrier m is moved to baseband and filtered in order to remove the out of band component:

$$s_m^{rec}(t) = \int_{-\infty}^{\infty} (s_{tot}^{NL}(t - \tau) + n(t - \tau))e^{-j2\pi f_m(t-\tau)} r_m(\tau)d\tau \tag{5}$$

$n(t)$ is the standard additive white Gaussian noise. Finally, the received signals are down-sampled to produce the received symbols $y_m(n); m = 1, 2$. Ideal synchronization is assumed here.

Note that if the shaping and receive filters have a large bandwidth, interference with the IMUX and OMUX filters may occur aboard the satellite. The IMUX filter is a bandpass filter which separate the different single- or multi-carriers channels. The OMUX filter, which is also a bandpass filter, removes the out-of-band components produced by the power amplifier. The filter $p_m(t)$ is in fact the convolution of the pulse shaping filter and the IMUX filter in the considered band. Similarly, the filter $r_m(t)$ is the convolution of the receiver filter with the OMUX filter in the considered frequency band.

The Volterra model is a common tool to describe the input-output relationship of a digital or analog non-linear system with memory [4]. The case of a single carrier satellite communication channel has been described in [10]. [7] expresses the generalization to the multi-carrier case of the first and third order interference terms in the general case of one physical transmitter for each carrier. In the single carrier case, we have:

$$y_1(n) = \sum_{n_1=-\infty}^{\infty} H_1(n_1)x_1(n - n_1) + \sum_{n_i=-\infty}^{\infty} H_3(n_1, n_2, n_3)x_1(n - n_1)x_1(n - n_2)x_1(n - n_3)^* \tag{6}$$

Note that the second term in Equation (6) is in fact a triple sum on the indexes n_1, n_2 and n_3. We consider here third order models, but it can be easily extended to any order. The coefficients $H_m(n_1 ... n_m)$ are called the Volterra coefficients, and can be seen as the extension of the impulse response of a linear system to that of a non-linear system. In the case of a two carriers system, Equation (6) becomes:

$$\begin{aligned} y_m(n) = & \sum_{n_1=-\infty}^{\infty} H_1^{m,1}(n_1)x_1(n - n_1) \\ & + \sum_{n_1=-\infty}^{\infty} H_1^{m,2}(n_1)x_2(n - n_1) \\ + & \sum_{n_i=-\infty}^{\infty} H_3^{m,1,1,1}(n_1, n_2, n_3)x_1(n - n_1)x_1(n - n_2)x_1(n - n_3)^* \\ + & \sum_{n_i=-\infty}^{\infty} H_3^{m,2,1,1}(n_1, n_2, n_3)x_2(n - n_1)x_1(n - n_2)x_1(n - n_3)^* \\ + & ... \\ + & \sum_{n_i=-\infty}^{\infty} H_3^{m,2,2,2}(n_1, n_2, n_3)x_2(n - n_1)x_2(n - n_2)x_2(n - n_3)^* \end{aligned} \tag{7}$$

For $m = 1$ (resp. $m = 2$), the first (resp. second) line of Equation (7) is similar to the first line of Equation (6). It represents the linear intersymbol interference (ISI). The third (resp. last) line of Equation (7) is similar to the second line of Equation (6) and represents the non-linear ISI. The second (resp. first) line expresses the linear adjacent channel interference (ACI) and the remaining lines the non-linear ACI. Equation (7) can easily be extended to any order or to a different number of carriers.

III. Pre-distortion Algorithm

The pre-distortion algorithm proposed here is an extension of the algorithm developed in [11] for the single carrier case. The algorithm determines the sequences of pre-distorted symbols $\{x_m(1)...x_m(N)\}; m = 1, 2$, to minimize:

$$\sum_{k=1}^{N} ||y_1(k) - a_1(k)||^2 + \sum_{k=1}^{N} ||y_2(k) - a_2(k)||^2 \quad (8)$$

The proposed pre-distortion algorithm relies on three main characteristics. The first characteristic is that we define the pre-distorted value of $x_1(n)$ only when the values of $x_m(n-1); m = 1, 2$, have been determined. This is clearly a sub-optimum approach since joint optimization is not performed. After $x_1(n)$ has been determined, $x_2(n)$ is then calculated. The value of $x_1(n)$ is chosen to minimize the cost function (8) supposing that the pre-distorted sequences are:

$$\begin{array}{lll} \{x_1(1)...x_1(n-1) & ? & a_1(n+1)...a_1(N)\} \\ \{x_2(1)...x_2(n-1) & a_2(n) & a_2(n+1)...a_2(N)\} \end{array} \quad (9)$$

Similarly, the value of $x_2(n)$ is chosen to minimize (8) supposing that the pre-distorted sequences are:

$$\begin{array}{lll} \{x_1(1)...x_1(n-1) & x_1(n) & a_1(n+1)...a_1(N)\} \\ \{x_2(1)...x_2(n-1) & ? & a_2(n+1)...a_2(N)\} \end{array} \quad (10)$$

The second characteristic of the algorithm is that the considered N-symbols sequences are iteratively processed. After the last pre-distorted value $x_2(N)$ of a block has been determined, the algorithm looks for a new optimum value for $x_1(1)$. The reason is that $x_1(1)$ has been determined assuming that all other symbols were undistorted and thus equal to:

$$\begin{array}{l} \{? \;\; a_1(2)...a_1(N)\} \\ \{a_2(1) \;\; ... \;\; a_2(N)\} \end{array} \quad (11)$$

However, after one iteration of the algorithm, the symbols to transmit do not belong anymore to the constellation alphabet:

$$\begin{array}{l} \{? \;\; x_1(2)...x_1(N)\} \\ \{x_2(1) \;\; ... \;\; x_2(N)\} \end{array} \quad (12)$$

so that the optimization needs to be done again. After $x_1(1)$ has been recalculated, it is then $x_2(1)$ which is recalculated and so on. By iterating several times, the algorithm converges at least towards a local minimum for the cost function (8). In the following, we define $x_m^k(n)$ the pre-distorted value relative to the symbol n of the carrier m after iteration k, and the initial value $x_m^1(n) = a_m(n)$.

Until now, we have considered $x_m(n)$ as the value which minimizes Equation (8) when all other symbols are kept constant as represented in Equation (9). Using Equation (7), The cost function (8) can then be reduced to a non-linear function in the complex variable $x_m(n)$. Two problems occur:

- The coefficients of this non-linear function depend on all Volterra coefficients of the system, so that the complexity to derive these coefficients is prohibitive.
- The minimum of this non-linear function may be hard to find.

A third characteristic of the algorithm is thus to assume a linear relation between the output variations and the variation from $x_m^k(n)$ to $x_m^{k+1}(n)$. To keep the notations simple, we denote $y_m^{old}(n+l); l = 1-n...N-n$, the received symbols when the sequences

$$\begin{array}{lll} \{x_1^{k+1}(1)...x_1^{k+1}(n-1) & x_1^k(n) & x_1^k(n+1)...x_1^k(N)\} \\ \{x_2^{k+1}(1)...x_2^{k+1}(n-1) & x_2^k(n) & x_2^k(n+1)...x_2^k(N)\} \end{array} \quad (13)$$

are transmitted. Similarly, we denote $y_m^{new}(n+l); l = 1-n...N-n$, the received symbols when the sequences

$$\begin{array}{lll} \{x_1^{k+1}(1)...x_1^{k+1}(n-1) & \mathbf{x_1^{k+1}(n)} & x_1^k(n+1)...x_1^k(N)\} \\ \{x_2^{k+1}(1)...x_2^{k+1}(n-1) & x_2^k(n) & x_2^k(n+1)...x_2^k(N)\} \end{array} \quad (14)$$

are transmitted. We consider here that $x_1^k(n)$ has been updated but the reasoning is of course similar for $x_2^k(n)$. We define $\epsilon = x_1^{k+1}(n) - x_1^k(n)$. The linear assumption states that:

$$y_m^{new}(n+l) = y_m^{old}(n+l) + A_l\epsilon + B_l\epsilon^* \quad (15)$$

where A_l and B_l are complex coefficients. Substituting Equation (15) in (8), we see that the calculation of $x_1^{k+1}(n)$ requires only to minimize a second order complex equation with the complex unknown ϵ, which can be easily done using partial derivative.

Note that this linear assumption is not equivalent to neglect the non-linear Volterra terms. As an example, if we consider a single carrier system where the channel consists only in one Volterra coefficient $H_3(0,0,0)$. The linear assumption induces the following approximation:

$$\begin{aligned} y_m^{new}(n) &= H_3(0,0,0)x^{k+1}(n)|x^{k+1}(n)|^2 \\ &= H_3(0,0,0)(x^k(n)+\epsilon)|x^k(n)+\epsilon|^2 \\ &\approx y_m^{old}(n) + H_3(0,0,0)(2|x^k(n)|^2\epsilon + x^k(n)^2\epsilon^*) \end{aligned} \quad (16)$$

In this simple example of a memoryless channel, we have that $A_l = 0$ and $B_l = 0$ except for $l = 0$, where their value are respectively $H_3(0,0,0)2|x^k(n)|^2$ and $H_3(0,0,0)x^k(n)^2$. In practical cases, the coefficients A_l and B_l will depend on all Volterra coefficients and on the pre-distorted sequence so that they remain difficult to derive. In Annex I, we show how they can be calculated in practice without resorting to the Volterra coefficients.

Finally, we introduce a damping factor in the updating process to ensure that the linear assumption is met. The here above calculated value for $x_m^{k+1}(n)$ is only optimum if the linear assumption is met. Since this may not be the case, we take instead the following value for $x_m^{k+1}(n)$:

$$x_m^{k+1}(n)' = x_m^k(n) + \gamma(x_m^{k+1}(n) - x_m^k(n)). \quad (17)$$

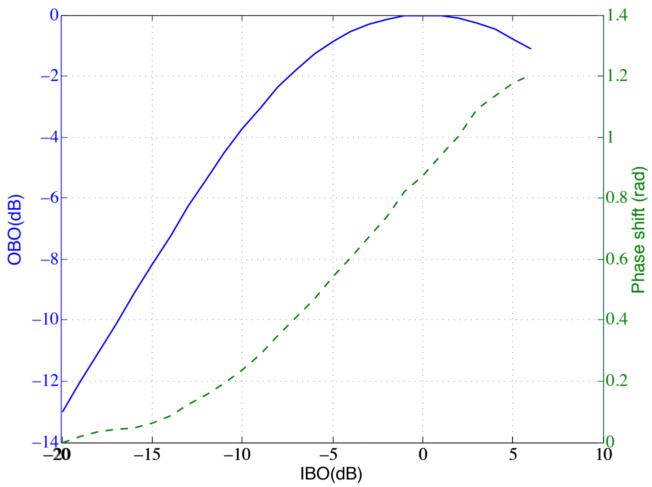

Fig. 2. AM-AM and AM-PM characteristics of the HPA.

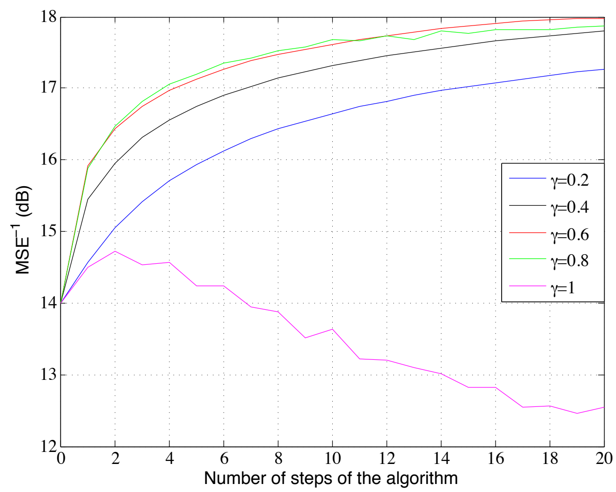

Fig. 3. Mean square error after each step of the pre-distortion algorithm for different values of γ

γ is a damping factor, with $\gamma \in]0,1]$. It can be easily shown that even if $x_m^{k+1}(n)'$ is not the optimum value for the minimization of (8), it still decreases the value of (8) if the linear assumption is met. It is then always possible to define γ small enough to validate the linear assumption, so that $x_m^{k+1}(n)'$ at least decreases the square error. After a sufficient number of iterations, the algorithm converges then to an (at least local) optimum.

IV. NUMERICAL RESULTS

The chosen modulation for the simulations is the 8-PSK modulation, which offers a good spectral efficiency, and is representative of state of the art satellite broadband systems in Ka band, with multiple beam coverage and frequency reuse. The code is the LDPC code from the DVB-S2 standard (see [13]) with code rate 3/4 and the interleaver is the associated interleaver described in the DVB-S2 standard. The shaping and receiver filters are a square-root raised cosine filter with a roll-off factor β of 0.2. The power amplifier model is derived using the AM-AM and AM-PM characteristics of a typical traveling wave tube amplifier (TWTA), as illustrated in Fig. 2, where the input back-off (IBO) and output back-off (OBO) are defined relatively to the amplifier saturation We consider that the two carriers have the same data rate T_s^{-1}. A null guard board is considered, which means that if f_1 is the central frequency of the first carrier and f_2 the central frequency of the second carrier, we have:

$$f_2 = f_1 + (1+\beta)T_s^{-1} \tag{18}$$

Fig. 3 represents the mean square error (MSE) after each iteration of the algorithm for different damping factors γ when the amplifier is operated at saturation (IBO=0dB). As expected, the converging time of the algorithm increases for smaller values of γ. However, we see that for large values of γ, the algorithm does not converge anymore. This means that the linear assumption is not valid anymore.

Fig. 4 represents the MSE after 50 iterations of the algorithm for different operating points of the amplifier. In case of no pre-distortion, the single carrier case exhibits a smaller MSE for each IBO compared to the two carriers case. This is expected since no adjacent channel interference occurs in the single carrier case. The pre-distortion algorithm decreases the MSE in both cases, but the gain of the algorithm on the MSE is higher for the single carrier case. The initial interference level is however higher in the two carriers scenario. Therefore, similar improvements of the interference level will be more profitable on the link budget in the two carriers case, as the interference term will be more significant in the signal-to-noise-plus-interference ratio (SNIR).

Fig 5 represents the bit error rate (BER) as a function of the SNR for the two carriers channel, for an IBO of 0dB. The pre-distortion algorithm allows to gain about 2dB compared to the case of no pre-distortion. It is only separated from few tenths of dB from the AWGN channel without interference.

V. CONCLUSION AND FUTURE WORK

In this paper, we propose a new algorithm for the pre-distortion of a two carriers satellite communication link. This algorithm allows to greatly reduce the impact of the non-linear interference on the demodulation performance and can easily be extended to any number of carriers. However, the complexity of the algorithm is very high, since it requires a lot of channel simulations to find the linear coefficients linking the input variation and the resulting output variations (as explained in Appendix A). The more carriers in the system, the more channel simulations will be required. Future work will therefore include a decrease of the complexity of the algorithm. The algorithm performance will also be assessed for larger signal bandwidths, where interference with the IMUX and OMUX filters has to be taken into account.

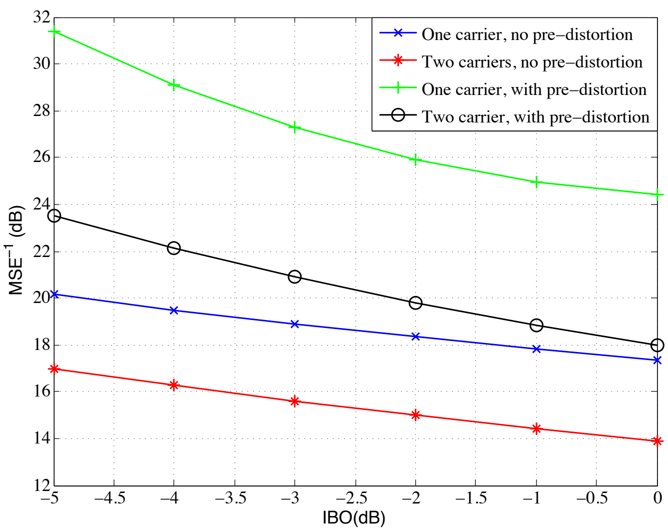

Fig. 4. Mean square error in function of the IBO with and without pre-distortion algorithm.

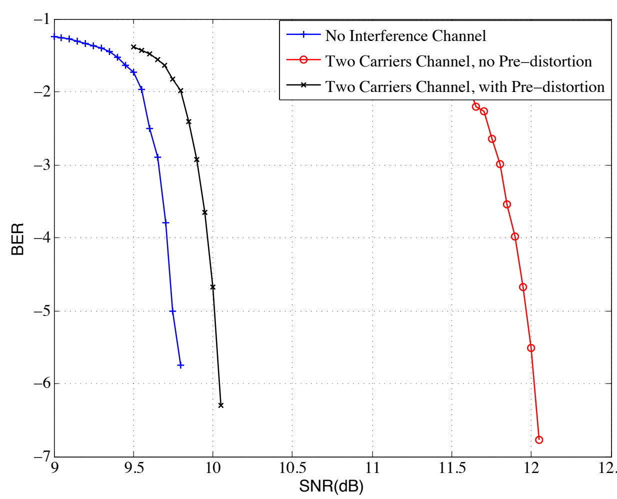

Fig. 5. BER comparison at IBO=0dB .

APPENDIX A
IDENTIFICATION OF THE COEFFICIENTS A_l AND B_l

In this annex, we describe how to identify the coefficients A_l and B_l from Equation (15) assuming that the pre-distorted symbols $x_1^{k+1}(n)$ is being calculated. We will therefore simulate the channel for the following input sequences:

$$\begin{array}{lll} \{x_1^{k+1}(1)...x_1^{k+1}(n-1) & X & x_1^k(n+1)...x_1^k(N)\} \\ \{x_2^{k+1}(1)...x_2^{k+1}(n-1) & x_2^k(n) & x_2^k(n+1)...x_2^k(N)\} \end{array} \tag{19}$$

for 3 different values of X:

- $X = x_1^k(n)$
- $X = x_1^k(n) + \epsilon_r$
- $X = x_1^k(n) + \epsilon_i$

where ϵ_r and ϵ_i are arbitrary small respectively real and pure imaginary numbers. We denote the corresponding channel outputs as respectively $y_m^1(n+l), y_m^2(n+l)$ and $y_m^3(n+l)$. Since ϵ_r and ϵ_i are chosen very small, we can assume that the linear assumption is met. Using the values of $y_m^1(n+l), y_m^2(n+l)$ and $y_m^3(n+l)$ in Equation (15) and the corresponding channel inputs, we can easily determine the values of the coefficients A_l and B_l.

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