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A combined turbo code with adaptive predistortion scheme for a non-linear channel

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Abstract. In this paper a distortion compensator suitable for improving the Digital Video Broadcasting Satellite (DVB-S) standard is presented for nonlinear amplification using a Travelling Wave Tube Amplifier (TWTA). The combined predistortion with Turbo Codes (PDTC) scheme is based on a feedforward neural network within an extended Kalman filter algorithm to estimate the coefficients and a conventional turbo code. The proposed scheme mitigates the degradation due to nonlinear amplification and additive white Gaussian noise by an adaptive predistorter and a turbo code, respectively. The results obtained by the PDTC scheme are compared with the theory curve using 16-QAM modulation. Computer simulation confirms that for the case of 16-QAM signals amplified with TWTA a performance improvement of E_b/N_o equal to 4 dB at a BER of 10^{-4} is obtained when compared with the theory curve.

1. Introduction

The European Telecommunication Standards Institute (ETSI) has produced standards for the transmission over satellites using high spectral efficiency multilevel quadrature modulation technique (M-QAM) such as 4QAM and 16QAM [5]. Although the M-QAM signals has a high spectral efficiency it is not suitable for mobile satellite communications because are very sensitive to both non-linear (AM/AM) and (AM/PM) distortions introduced by the high power amplifier kind of TWTA [13]. To compensate for these distortions two techniques have been reported in the literature. One technique is at the transmitter side using data predistortion [9], [12] or signal predistortion [4], [6], [10] and the other technique is at the receiver side using adaptive nonlinear equalization [3], [15].

In this paper a combined turbo code with adaptive baseband predistortion scheme using 16-QAM signals over a nonlinear satellite channel is presented. The turbo code is used to mitigate the additive white Gaussian noise and the predistorter is used to reduce the nonlinearities of the AM/AM (amplitude) conversion and AM/PM (phase) conversion of the amplified signal.

Turbo codes were originally proposed by Berrou et. al. [2] in an effort to achieve near Shannon capacity in the presence of additive white Gaussian

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noise. This technique is obtained by concatenating two or more convolutional codes in parallel with the information sequence divided into blocks of a certain length. The input of the upper component code is the original information block and the input to the lower component code is an interleaved version of the information block. Therefore, the encoded sequence consists of the information sequence, the parity bits of the upper encoder and the parity bits of the lower encoder. Moreover, some puncturing schemes are used to obtain turbo codes with different rates. In order to decode these codes a modified version of the classic maximum a posteriori algorithm (MAP) due to Bahl et al has been used [1].

The predistorter is based on a feedforward neural network (FNN). Recently, the FNN has been applied to modeling nonlinear memoryless channels such as TWTA allowing an efficient approximation of the inverse TWTA transfer function [8]. However, these systems have been trained using a Multilayer Perceptron (MLP) with the classical back-propagation algorithm and slow convergence rate due to the gradient algorithm and the information ignorance of the early data. In this paper a feedforward neural predistorter trained with an Extended Kalman Filter (EKF) algorithm is proposed, which increases the convergence speed [11].

The remainder of this paper is organized as follows: In section 2, a systems description of the propose scheme is presented. In Section 3, a predistortion technique with a feedforward neural network is used. The supervised learning algorithm to adjust the coefficients of the neuronal predistorter is presented in Section 4. In Section 5, the results from simulation are discussed and finally in the last section conclusions are presented.

2. System Model

The baseband-equivalent functional block diagram of the transmission system is shown in Figure 1 for 16-QAM signals. The input data bits are encoded using a turbo code and the constituent codes are recursive systematic convolutional (RSC) codes. The encoder takes as input the data sequence $d_k \in \{0, 1\}$ of length N_d, which is assumed to be statistically independent and produces an output with three streams, the systematic bits $d_k = x^s_k$, the parity bits $x^p_{1,k}$ of the upper component encoder with input d_k and the parity bits $x^p_{2,k}$ of the lower component encoder with the interleaved d_k as input. To achieve an overall coding rate of 1/2 a puncturing scheme is used in such a fashion that only odd parity bits of the upper constituent encoder and even parity bits of the lower constituent encoder are transmitted $x^p_k = \left\{x^p_{1,1}, x^p_{2,2}, x^p_{1,3}, x^p_{2,4}, x^p_{1,5}, x^p_{2,6}, ...\right\}$. The resulting concatenated code can be considered as two constituent codes, one generated from the non-interleaved information sequence and other one from the interleaved sequence. At the Mapper device, the systematic stream x_k^s and coded stream x_k^p are mapped to complex symbols q_k using Gray mapping. The transmitter filter is implemented as a square-root raised cosine (SRRC) filter distributed at the transmitter and receiver with 49 taps and a roll-off parameter of 0.5 and an over-sample factor of 8 samples per symbol. The modulated baseband signal $x(t)$ is first pre-distorted and nonlinearly amplified, then propagated over an (AWGN) channel.

The signal amplified is represented as:

$$z(t) = A(r_y(t)) \exp\left(j \cdot (\theta_y(t) + \Phi(r_y(t)))\right) \tag{1}$$

where r_y and θ_y are the amplitude and phase of the predistorted complex signal $y(t)$. The function $A(.)$ and $\Phi(.)$ denote AM/AM conversion (nonlinear amplitude) and AM/PM conversion (nonlinear phase); respectively.

For a TWTA, the expressions for $A(\cdot)$ and $\Phi(\cdot)$ are given by [13]:

$$A(r_y) = \frac{\alpha_a r_y}{1 + \beta_a r_y^2} \tag{2}$$

$$\Phi(r_y) = \frac{\alpha_\Phi r_y^2}{1 + \beta_\Phi r_y^2} \tag{3}$$

In this paper, we will use a normalized amplifier with coefficients $\alpha_a = 1.9638$, $\beta_a = 0.9945$, $\alpha_\Phi = 2.5293$ y $\beta_\Phi = 2.8168$.

The nonlinear distortion of a high power amplifier depends on the backoff. The input backoff (IBO) power is defined as the ratio of the saturation input power, where the output power begins to saturate, to the average input power:

$$IBO = 10 \log_{10}\left(\frac{P_{i,sat}}{P_{i,avg}}\right) \tag{4}$$

where $P_{i,sat}$ is the saturation input power and $P_{i,avg}$ is the average power at the input of the TWTA.

At time t, the received signal $r(t)$ is defined by:

$$r(t) = z(t) + n(t) \tag{5}$$

where $n(t)$ is complex additive white Gaussian noise with two-sided spectral density $N_0/2$.

The received signal $r(t)$ is passed through the matched filter (SRRC) and then sampled at the symbol rate $1/T$. The sequence at the output of the sampler v_k is fed to the constellation Demapper. The Demapper splits the complex symbols v_k into y_k^s and y_k^p components and puts them into a decision device, where they are demodulated independently against their respective decision boundaries. The Depuncture block simply reverses the process of the Puncture block, converting the received parity bit stream y_k^p to parallel sequences $\left\{y_{1,k}^p, y_{2,k}^p\right\}$ and placing zeros in locations of the

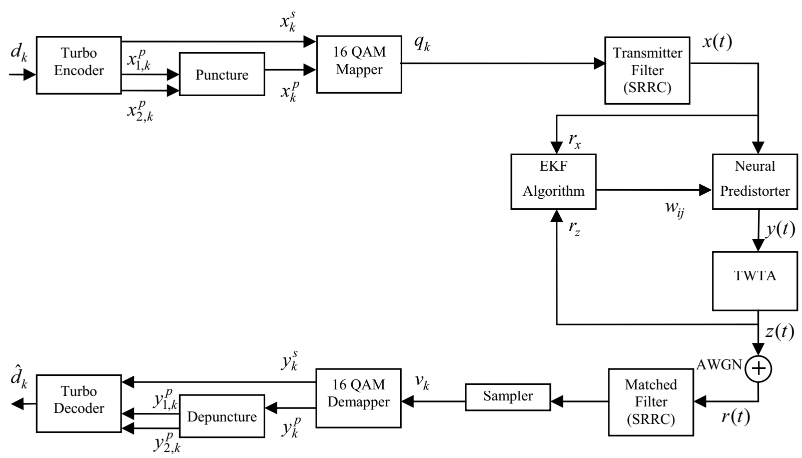

Figure 1. Block diagram of the baseband transmission

punctured bits. The received parallel bit stream $\left\{y_k^s, y_{1,k}^p, y_{2,k}^p\right\}$ is decoded by the turbo decoding process, which generates the estimated output bits stream $\hat{d}_k$. Turbo decoding is an iterative procedure, which makes use of the sub-optimum maximum a posteriori (MAP) algorithm. The derivation of this algorithm has been well documented in previous papers [1], [2].

3. Adaptive Predistortion for TWTA

The adaptive predistortion model used in this paper will be described by both nonlinear amplitude and phase conversions as:

$$y(t) = M(r_x(t)) \exp\left[j\left(\theta_x + N(r_x(t))\right)\right] \tag{6}$$

where r_x and θ_x represent the amplitude and phase of the input signal to predistorter $x(t)$.

Using equation (1) and equation (6) we obtain the complex signal envelope at the TWTA output as:

$$z(t) = A\left[M(r_x)\right]\exp\left[j\left(\theta_x + N(r_x) + \Phi(M(r_x))\right)\right] \tag{7}$$

In order to obtain an ideal predistorter the signal $z(t)$ will be equivalent to the input signal$x(t)$. This is:

$$r_z(t) = r_x(t) \quad \text{and} \quad \theta_z(t) = \theta_x(t) \tag{8}$$

Then, we obtain:

$$\begin{aligned} &r_z(t) = r_x(t) \\ &A\left[M(r_x(t))\right] = r_x(t) \\ &A^{-1}\left(A\left[M(r_x(t))\right]\right) = A^{-1}\left(r_x(t)\right) \\ &M\left(r_x(t)\right) = A^{-1}\left(r_x(t)\right) \end{aligned} \tag{9}$$

where $M(.)$ represent inverse amplitude function of the TWTA and:

$$\begin{aligned} &\theta_z(t) = \theta_x(t) \\ &\theta_x(t) + N\left[r_x(t)\right] + \Phi\left[M\left(r_x(t)\right)\right] = \theta_x(t) \\ &N\left[r_x(t)\right] = -\Phi\left[M\left(r_x(t)\right)\right] \end{aligned} \tag{10}$$

where $N(.)$ represent inverse phase function of the TWTA.

Therefore, predistorted output $y(t)$ is obtained by:

$$y(t) = A^{-1}\left(r_x(t)\right)\exp\left[j\left(\theta_x(t) - \Phi\left(A^{-1}\left(r_x(t)\right)\right)\right)\right] \tag{11}$$

Therefore, in order to obtain the predistortion function $f_{PD}(.) = y(t)$ it is only necessary to find the real-valued function$A^{-1}(\cdot)$. To approximate the $A^{-1}(\cdot)$ function a feedforward neural network is used with a simple hidden layer and the weights are determined from a finite number of samples of the function $A(\cdot)$.

4. Training of the Adaptive Predistorter

During the training process the signal $x(t)$ is equal to$y(t)$, but during direct-decision mode the signal $y(t)$ will be the desired predistorted signal. The training process was done using the trial and error method. In order to implement the training process, it is necessary to obtain a database Γ containing the output amplitude $r_z(n)$ of the TWTA and the corresponding desired output $r_x(n)$, $\Gamma = \{r_z(n), r_x(n); n = 1, ..., N_s\}$, where the $N_s$value represents the sample number of the function $A(\cdot)$ and the input $r_z(n)$ at the FNN is obtained as

$$\begin{aligned} r_z(n) &= \frac{|z(n)|}{\max\{|z(n)|\}} \\ r_x(n) &= \frac{|x(n)|}{\max\{|x(n)|\}} \end{aligned} \tag{12}$$

The output of the FNN is obtained as:

$$\begin{aligned} y_{NN}(n) &= \sum_{j=1}^{N_h} b_j h_j \\ h_j &= \Psi(a_j r_z(n)) \end{aligned} \tag{13}$$

where the N_h value represents the number of nodes in the hidden layer and. The weights b_j and a_j are the interconnections of the output and hidden layer, respectively, and $\Psi(.)$ represents the sigmoid activation function of the hidden nodes.

Training weights of the neural predistorter could be approached by the use of the Extended Kalman Filter algorithm [11]. The goal of the EKF algorithm is to find the weights vector that minimizes the cost function defined by:

$$E(w) = \frac{1}{N_s}\sum_{n=1}^{N_s} e^2(n) = \frac{1}{N_s}\sum_{n=1}^{N_s} \left[r_x(n) - y_{NN}(r_z(n), w)\right]^2 \tag{14}$$

The problem is to find the optimal weights that can be treated as a nonlinear parametrical identification problem. Let us assume that all the weights in the FNN are arranged in a weight vector $w = [a_1, ..., a_{N_h}, b_1, ..., b_{N_h}]$. Then the following state/observation equations can be written [14]:

$$\begin{aligned} w(n+1) &= w(n) \\ r_x(n) &= f_{pd}\left[r_z(n), w(n)\right] + u(n) \end{aligned} \tag{15}$$

The measurement of the non-linear mapping $f_{PD}(.)$ is approximated by a FNN where the weights are the model parameters w. Furthermore, we assume that the initial state $w(0)$ and the sequence $\{u(n)\}$ are independent and Gaussian distributed.

The output of the FNN is $y_{NN}(n) = f_{PD}\left[r_z(n), \hat{w}(n)\right]$, where $\hat{w}(n)$ represents the estimated weights at time n. The estimated updates $\hat{w}(n+1)$ are obtained from the EKF algorithm [14], which is a minimum variance estimator based on a Taylor series expansion of the nonlinear function $f_{PD}(\cdot)$ around the previous estimation.

The EKF equations for the updates $\hat{w}(n+1)$ are given by [14]:

$$K(n) = P(n-1)F(n)\left[R(n) + F^T(n)P(n-1)F(n)\right]^{-1} \tag{16}$$

$$P(n) = P(n-1) - K(n)F^T(n)P(n-1) \tag{17}$$

$$\hat{w}(n+1) = \hat{w}(n) + K(n)\left[r_x(n) - f_{PD}(r_z(n), \hat{w}(n))\right] \tag{18}$$

where $F(.)$represents the Jacobean matrix.

5. Simulation results

The performance of the proposed scheme was evaluated by computer simulation with a maximum length of $N_d = 10^5$data bits. A $(13, 11)_8$ convolutional turbo code was considered with an S-random modulo-K interleaver of size $L = 512$bits, $S = 13$ and $K = 2$ [7].The decision on the data was taken after decoding, where the decoded data $\left\{\hat{d}_k\right\}$ was compared with the transmitted data $\{d_k\}$ to calculate the BER.

The parameters of the predistorter were estimated during the training process using $N_s = 160$ samples of the amplitude $A(\cdot)$ and the TWTA was operated with $IBO = -1$dB. The neural predistorter for 16-QAM signals was configured with one input node, one output node and eight hidden nodes. This structure is denoted as PDTC(1,8,1). In the training process the initial weights, $w(0)$, were initialized by a Gaussian random process with a normal distribution $N(0, 1)$. The initial weight covariance matrix and the measurement noise covariance were set to $P(0) = 10^2$ and$R(0) = 10^{-6}$. The training process was run with 30 trials and the mean square error (MSE) after convergence was approximately10^{-4}.

In Figures 2 we investigate the BER performance versus signal to noise ratio E_b/N_0, for a value of input back off equal to -1dB, where the nonlinearities introduced by the TWTA is more predominant. From Figure 2 it can be seen that a BER of 8×10^{-4} is achieved at a $E_b/N_0$of 10 dB when the number of decoding iteration is low. This is because the efficiency of the compensator PDTC(1,8,1) is poor for a small number of decoding iterations. However, when the number iterations increases, then the efficiency of the PDTC(1,8,1) scheme increases rapidly achievement a gain of approximately 4 dB over theory curve at a $E_b/N_0$of 8 dB.

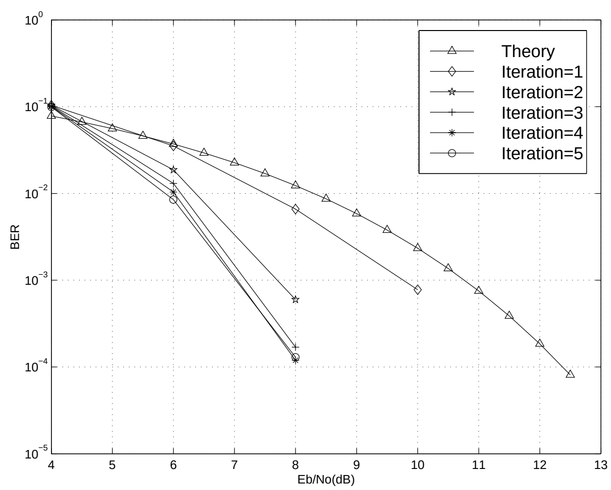

Figure 2. BER performance with TWTA, IBO=-1dB, 16QAM and AWGN.

6. Conclusion

In this paper a distortion compensator suitable for improving the Digital Video Broadcasting Satellite (DVB-S) standard has been presented for non-linear amplification using TWTA, which can be adapted as an alternative for the (DVB-S) standard if the concatenated convolution code specified in the standard is substituted by conventional turbo decoding.

The proposed scheme is based on a forward neural network to mitigate non-linear amplification and the turbo code has been used to mitigate Gaussian noise effect. From simulation results, the proposed scheme for the case of TWTA with 16-QAM signal achieve a gain of E_b/N_0 approximately of 4dB at a BER of approximately 10^{-4} when is compared with the theoretical curve.

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