A NON-LINEAR ADAPTIVE EQUALIZATION TECHNIQUE IN DIGITAL SATELLITE COMMUNICATION

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Abstract

An approach to rectify the distortions caused by the Travelling Wave Tube (TWT) or solid-state High Power Amplifiers (HPA) used in digital satellite communication systems and at the same time to correct the effects of intersymbol interference is presented. Linear adaptive signal processing techniques fail to eliminate these nonlinear distortions introduced by nonlinear components in the satellite channel. Here we present a nonlinear adaptive signal processing technique based on Volterra series. This part includes mathematical description of proposed scheme; simulation results will be presented in the second part of the paper.

Key Words: QPSK, QAM, TWT, HPA.

Introduction

Adaptive signal processing techniques have been used in many diverse fields such as echo cancellation, system identification, financial market data analysis etc. They have worked quite well and sometimes have produced remarkable results. Most of these techniques have been developed for linear systems and are linear in nature. Various problems of that type and their solutions are discussed in [1], [13], [25], and [28]. However, there are many problems that are inherently nonlinear, where approaches based on linear adaptive signal processing techniques fail to produce satisfactory results. Digital satellite communication channel is one such example. High power amplifiers are an integral part of any digital satellite communication channel. These amplifiers are either based on traveling wave tube (TWT) or solid-state components and have nonlinear input/output characteristics. The maximum amount of information transfer, with acceptable error rate, over digital

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satellite communication channels is constrained by the available transmission power and the available bandwidth. Availability of power is very limited onboard communication satellites. To use this limited power efficiently, these high power amplifiers usually operate in the nonlinear region of their characteristics. Quadrature Phase-Shift Keying (QPSK) is typically the modulation scheme of choice. It is preferred over other bandwidth and power efficient modulation schemes with acceptable error rate, because of its constant envelope characteristic, which makes the process of demodulation easy, even after passing through a nonlinear power amplifier.

In order to use power and bandwidth efficiently digital satellite communication systems need modulation schemes such as Quadrature Amplitude Modulation (QAM). QAM has by design variable amplitude. When this type of modulation scheme is used in digital satellite communication system, where power amplifiers operate in a nonlinear regime, the output signal is nonlinearly distorted. These distortions are very difficult to rectify using the linear adaptive filtering techniques. There are a number of schemes proposed to counter the adverse effects of digital satellite communication channel's nonlinearity. There exist basically three different approaches to manage this problem. First approach is known as preprocessing. In this scheme, we modify (deliberately distort) the signal before it passes through the nonlinear element present in the system to have an overall linear response. The whole idea of pre-distortion is the transformation of input signal using a nonlinear device before transmission in order to improve the linearity of the overall link. Saleh did a pioneer work on this approach [25]. Many other variants of this scheme are discussed in [4], [8], [17], [24], and [25].

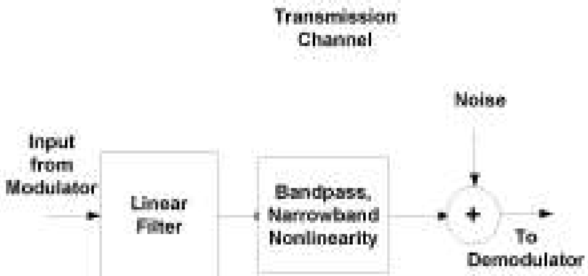
In the second approach, exactly opposite of preprocessing is done. It is known as post-processing. The signal is transmitted in its original form while all processing is done at the receiving end. Received signal is processed to counter the effects of nonlinearity and to recover the signal with an acceptable error rate. The optimal solution for this kind of processing has been found by Mesiya et al. [20]. The solution employs the maximum likelihood sequence estimation (MLSE) technique, which is rather demanding computationally. Other sub-optimal solutions are nonlinear equalization methods proposed by Falconer [9] and Benedetto [2-3]. Some other methods that fall into this category are discussed in [10], [11], [12], [18], [19], [21], and [22]. Our method also belongs to this category. We use a third-order; finite memory Volterra series based equalizer to demodulate the nonlinearly distorted signal. We find it more efficient than a non-adaptive solution because it can track the changes both in the linear part of the channel and in the nonlinearity and compensate for them. This is more suitable for systems where the complexity of the receiver is more tolerable, such as an earth-located receiving station. The increase in complexity becomes an issue if the receiver is a handheld device where size and cost are very important.

The third approach to counter the effects of nonlinearity on the signal is to combine both pre- and post-processing techniques. Here signal is treated on both ends of digital communication system. Signal processing starts on the transmitter side where trellis coding is used to encode and modulate the signal while on the receiver side the Vitterbi algorithm is used to decode the received signal. More information about this scheme can be found in [6], [7], [26], and [27].

Channel Model

A typical satellite communication link consists of both uplink and downlink channels. A simplified channel model is shown in Figure 1. We have lumped together all parts of the channel before the high power amplifier into one block and we name it linear filter block. We have also ignored the additive white noise in the uplink part because the signal is very strong, there is a very small noise component and its effects are minimal [5]. These simplifications allow us to make the mathematical description tractable. Downlink noise is added to the system because now the signal-to-noise ration (SNR) is not that high, and the additive noise affects considerably the performance of the system. In this channel model, we have assumed that all components of the channel (demodualtor, decision device, adaptive system, etc) except the additive noise are parts of the receiver. The downlink channel is linear in nature and its effects on the signal can be compensated easily in the receiver with well-established techniques.

Figure 1: Simplified Transmission Channel Model



We use the model shown in Figure 2 to represent the nonlinearity present in communication channel, as proposed by Saleh [25] as well as other researchers [14-16], and [24]. The nonlinearity present in the communication system introduces two kinds of distortion, Amplitude-to-Amplitude (AM-AM) and Amplitude-to-Phase (AM-PM). Both distortions are not easy to counter.

Let the input signal be

$$x(t) = r(t)cos(\omega t + \theta(t)) \quad (1)$$

where is the carrier frequency, and and $\theta(t)$ are modulated envelope and phase, respectively. The top branch of the model is simply in-phase nonlinearity and the bottom branch is the quadrature nonlinearity. The output of these branches can be represented as:

and

(2)

whereas the total output of this quadrature model is the sum of these two expressions. The expressions (2), in order to be consistent, require that

and

(3)

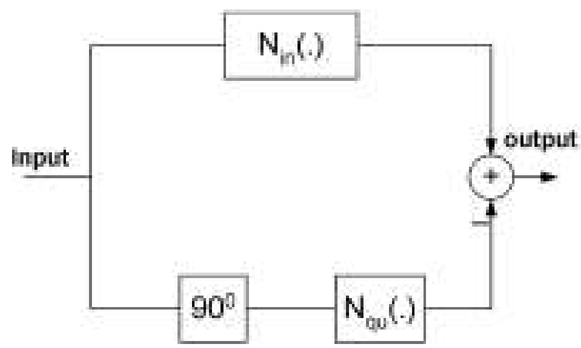
where and $\Phi(r)$ are amplitude distortion and phase shift caused by the non-linearity, respectively. Saleh [25] has proposed two-parameter approximation to represent and $N_{qu}(r)$ in terms of the $r(t)$.

(4)

$$N_{qu}(r) = \alpha_{qu} r^3 / (1 + \beta_{qu} r^2)^2 \quad (5)$$

He further determined the coefficients α, β for different TWT amplifiers and found the values that give a very good match to the empirical input/output characteristics of TWT amplifiers. The nonlinearity in our analysis is memoryless and time-invariant, with flat frequency response over the bandwidth of the input signal [16]. Therefore, we assume that the nonlinearity is in effect a narrowband device. As a direct result of this assumption, the 90-degree phase shift does not affect the statistics of the input signal that passes through the nonlinearity.

Fig. 2 Analytic Model for TWT Nonlinearity



The nonlinearity in the system creates two types of distortions: AM-AM and AM-PM. The effects of phase distortion, due to AM-PM, are more worrisome for communication systems.

Signal Propagation in the Channel

We start with the QAM output of the transmitter. A QAM signal pulse is defined as

(6)

where , denotes the set of M possible amplitudes corresponding to $M = 2^x$ possible x-bit blocks for both in-phase and quadrature terms, is the carrier frequency, and *g(t)* is the shaping waveform (envelop) of the signal

The signal amplitudes and B_i are equal to $(2i-1-M)d$ for different discrete values of *i*, where *2d* is the minimum difference between two adjacent signal amplitudes. The waveform *g(t)* is a real-valued signal pulse whose shape influences the spectrum of the transmitted signal. We have selected the usual form of *g(t)* equal to

(7)

The output of the QAM modulator is then defined as:

$$f_i(t) = [A_i \frac{1}{2}(1-\cos(2\pi t/T)\cos\omega t -$$

(8)

This signal is the input to the transmission channel in our model, where it is first processed through the linear filter block.

The linear passband filter used in our model is defined as

(9)

The filter introduces the intersymbol interference (ISI) in the communication channel. The value of t_0/T controls the amount of ISI in the channel.

The output of the linear filter is obtained by convolving the signal with the impulse response of the filter. Let the signal at the output of the linear filter be :

(10)

The output of the linear filter is a cumbersome expression. However, in a real situation , which yields a simplified expression for the in-phase signal:

$$-\frac{\cos(\frac{2\pi t}{T})+\frac{2\pi t_0}{T}\sin(\frac{2\pi t}{T})}{2(1+4\pi^2 t_0^2/T^2)})]; \qquad 0 \le t \le T$$

(11)

Similarly, the quadrature signal expression becomes:

$$-\frac{\cos(\frac{2\pi t}{T})+\frac{2\pi t_0}{T}\sin(\frac{2\pi t}{T})}{2(1+4\pi^2 t_0^2/T^2)})]; \qquad 0 \le t \le T$$

(12)

The sum of these two expressions is the output of the filter for an input of solitary signal pulse.

(13)

The signal after convolution spreads out, and the extent of this spread depends on the ratio . As a result of this spread the output signal contains tails from previous pulses. We consider a range of ratios for which only two previous signals render significant contribution. Denote the output signal of the linear part of the system with the ISI effects taken into account by . Here indices *in* and *qu* stand for the in-phase and quadrature components of the signal, respectively, and indices *i,j,k* are the amplitude indices of the present and two preceding signal pulses, respectively. The following expression demonstrates the effects of ISI on the in-phase signal:

$$w_{ijk,\mathrm{in}}(t) = \cos\omega t \cdot \{A_{1i}[\frac{(1-e^{-t/t_0})}{2} + \frac{e^{-t/t_0}}{2(1+4\pi^2 t_0^2/T^2)} -$$

$$-\frac{\cos(\frac{2\pi t}{T})+\frac{2\pi t_0}{T}\sin(\frac{2\pi t}{T})}{2(1+4\pi^2 t_0^2/T^2)}] + A_{2j}[\frac{e^{-t/t_0}-e^{-(t+T)/t_0}}{2} + \frac{-e^{-t/t_0}+e^{-(t+T)/t_0}}{2(1+4\pi^2 t_0^2/T^2)}] +$$

$$A_{3k}[\frac{e^{-(t+T)/t_0}-e^{-(t+2T)/t_0}}{2}+\frac{-e^{-(t+T)/t_0}+e^{-(t+2T)/t_0}}{2(1+4\pi^2 t_0^2/T^2)}]\}\cdot[u(t)-u(t-T)] \quad (14)$$

A similar expression for the quadrature signal is given below:

$$w_{ijk,\mathrm{qu}}(t)=\sin\omega t\cdot\{B_{1i}[\frac{(1-e^{-t/t_0})}{2}+\frac{e^{-t/t_0}}{2(1+4\pi^2 t_0^2/T^2)}-$$
$$-\frac{\cos(\frac{2\pi t}{T})+\frac{2\pi t_0}{T}\sin(\frac{2\pi t}{T})}{2(1+4\pi^2 t_0^2/T^2)}]+B_{2j}[\frac{e^{-t/t_0}-e^{-(t+T)/t_0}}{2}+\frac{-e^{-t/t_0}+e^{-(t+T)/t_0}}{2(1+4\pi^2 t_0^2/T^2)}]+$$
$$+B_{3k}[\frac{e^{-(t+T)/t_0}-e^{-(t+2T)/t_0}}{2}+\frac{-e^{-(t+T)/t_0}+e^{-(t+2T)/t_0}}{2(1+4\pi^2 t_0^2/T^2)}]\}\cdot[u(t)-u(t-T)] \quad (15)$$

The signal amplitudesare equal to $(2i-1-M)d$ different discrete values for the present pulse that started at time *t*. The signal amplitudesare equal to different discrete values for the first tail i.e. the tail of the previous pulse that started at time *(t-T)* before the present pulse. Similarly, the signal amplitudesare equal to different discrete values for the second tail – that of the signal pulse that started at time *(t- 2T)*.

The sum of (14) and (15) represents the signal produced by the linear filter before the nonlinearity, with the effect of the intersymbol interference taken into account. This effect depends on value of the ratio. In practice, the parameters of satellite communication channel are chosen so as to avoid very strong ISI, which means that the ratio is not large. Table 1 provides data about the contribution of energy from the present pulse and the tails of two previous pulses for the case when =0.25. For instance, suppose that our pulse is , and the values of and *T* are equal to one. The energy of this pulse is then equal to . When this pulse passes through the linear filter, it spreads out. Now the energy contribution from this pulse during the signaling interval is 0.239. The tails of two previous pulses also corrupt this pulse, the first tail contributing 0.044 and the second 0.004 to the total energy. It is readily seen that the contribution of all other tails is of the order of and can be neglected.

The expression for the in-phase and quadrature signal can be simplified further, as shown below.

$$w_{ijk,\mathrm{in}}(t)=\cos(\omega t)\{\frac{A_{1,i}}{2}-[\cos(\frac{2\pi t}{T})+\frac{2\pi t_0}{T}\sin(\frac{2\pi t}{T})]A_{1,i}D_1+e^{-t/t_0}A_{ijk}\} \quad (16)$$

where $A_{ijk}=[c_1A_{2j}+c_2A_{3k}-c_3A_{1i}]$, and c_1 ,c_2 and c_3 represent constant terms independent of *t*.

Table 1: Energy Contribution of Present Signal and Two Previous Pulses. The Pulse Shape is

	Main Pulse	Plus the First Tail	Plus Both the Tails
Energy Contribution	0.239	0.283	0.287

For the quadrature part of the signal, we have

(17)

where $B_{ijk}=[c_1B_{2j}+c_2B_{3k}-c_3B_{1i}]$, and c_1 ,c_2 and c_3 represent constant terms independent of *t*. For the sake of simpler notations, we omit the indices *i*, *j*, *k* henceforth. Then the QAM signal at the output of the linear filter can be written as

(18)

where

(19)

and

$$\psi(t)=\{\frac{B_{1i}}{2}-[\cos(\frac{2\pi t}{T})+\frac{2\pi t_0}{T}\sin(\frac{2\pi t}{T})]B_{1,i}D_1+e^{-t/t_0}B_{ijk}\} \quad (20)$$

Expression (18) can be rewritten in the form:

$$w_{ijk}(t)=r(t)\cos(\omega t+\theta(t)) \quad (21)$$

where

and.

Subsequently, this signal is passed through the nonlinearity. As explained in the previous section, we will be using the analytical approximation (2) and (3) (Cf. [25]),

The form of expression (21) is convenient for analysis of the influence of nonlinearity when (21) is the input to the nonlinear block. When this signal passes through the nonlinear block, the output has an expression following from (2), namely,:

(22)

where

After substituting the values for and in expression (22) and further simplification, we get the expression for the output of the nonlinearity shown below:

(23)

The signal received arriving the receiver is usually corrupted by additive Gaussian noise,. Therefore, the signal at the input of the demodulator is:

$$s_{nl}(t)+n(t)$$

The demodulator processes the received signal and generates the in-phase and quadrature outputs to be further processed by the equalizer. The action of the demodulator and the non-linear equalizer is to be presented in the second part of this paper.

Volterra Series for Nonlinear Equalization

Linear equalization (LE) based on the representation of the output signal as a linear combination of the present and several previous input signals, has been found to be quite effective for fighting ISI in linear channels(e.g. [13]). However, this approach fails to correct distortions and ISI in channels with substantial nonlinearity. Nonlinear equalization (NLE) is a natural generalization of the LE technique.

We describe below the use of Volterra series for a nonlinear equalization procedure. Volterra series has been called "power series with memory" since it expresses the output of a nonlinear system in terms of powers (in general, products) of its inputs related to consecutive discrete time intervals . A general expression for Volterra series is shown here:

(24)

It is easy to see that Volterra series is a generalization of a linear moving average filter. We use a finite memory, third-order Volterra filter in our nonlinear adaptive equalization technique. The expression for this filter is given below:

$$y_n = \sum_{i=0}^{M} a_i x_{n-i} + \sum_{i=0}^{M}\sum_{j=0}^{M} b_{ij} x_{n-i} x_{n-j} + \sum_{i=0}^{M}\sum_{j=0}^{M}\sum_{k=0}^{M} c_{ijk} x_{n-i} x_{n-j} x_{n-k} \tag{25}$$

where , and ( ) are the equalizer weights that are updated at each iteration. We also used the symmetry property to keep the number of terms to a minimum in the transversal filter. To justify the use of the third-order Volterra filter in our equalizer, we expand the both in-phase and quadrature nonlinearities into series form. In our channel model (Cf. Sec. 2), in-phase nonlinearity can be expressed as

(26)

Similarly, quadrature nonlinearity can be expressed as

(27)

where are constants and i=1,2,3,... The typical values for and for a normalized input are 1.6397, 0.2038, 0.0618 and 0.1332, respectively [25].

Since the values of and are quite small, contributions from higher order terms are very small and can be neglected. We have designed a finite memory, third-order Volterra series based adaptive equalizer to minimize the effects of nonlinearity on variable amplitude modulations schemes such as QAM. Further details about the design and operation of Volterra series based NLE will be presented in the companion paper along with the simulations and results obtained there.

Conclusion

The analysis of the linear part of the communication channel shows that it is sufficient to take into account only two previous consecutive signal pulses when considering the effects of ISI in the channel. Therefore, we can limit ourselves with the memory depth M=2 in the equalization scheme. Similarly, the analysis of the effects of nonlinearity shows that it is sufficient to employ third-order Volterra filter in NLE. The second paper will consider more details of the NLE implementation.

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