COMBINED EQUALIZATION AND NONLINEAR DISTORITON CANCELLATION FOR TRANSMISSION OF QAM SIGNAL IN FIXED WIRELESS CHANNEL

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ABSTRACT

In this paper, we consider combined equalization and nonlinear distortion cancellation for transmission of QAM signals in fixed wireless channel with nonlinear distortion. To reduce nonlinear inter-symbol interference (ISI) due to the use of high power amplifiers in the transmitter, we design a nonlinear equalization scheme by approximating the Volterra series model for the nonlinear distortion. The proposed equalization scheme employs two memoryless nonlinear modules, one at the input of the feedback filter and the other at the input of the decision device of the conventional minimum mean-squared-error decision-feedback equalizer (MMSE-DFE). The performance of the proposed scheme is analyzed in terms of the symbol error rate and verified by computer simulation. The SNR gain of the proposed scheme over the conventional MMSE-DFE increases as the QAM constellation size and/or the roll-off factor of the transmit and receive filters increase, enabling the transmission of high-level QAM signals in fixed wireless channel.

1. INTRODUCTION

Increasing demand for high-speed multimedia services over the limited radio spectrum requires the employment of spectrum-efficient modulation schemes such as multi-level quadrature amplitude modulation (QAM). The use of QAM transceivers requires highly linear power amplifiers with good power efficiency and low signal distortion. Nonlinear distortion (NLD) due to the use of a high-power amplifier (HPA) can be reduced by increasing the back-off of input signal level. However, this may result in inefficient use of power amplifiers.

Techniques for mitigating the NLD due to the use of HPAs have long been investigated in wireless systems such as the satellite communications and digital microwave radio systems [1-3]. Conventional techniques include the use of predistortion in the transmitter [2,3] and nonlinear equalization in the receiver [1,4-6]. The predistortion technique has mostly been considered in the base station that can afford an additional demodulation circuitry in the transmitter to provide the reference signal.

When the predistortion method is not affordable, the effect of NLD can be alleviated by employing nonlinear signal processing techniques, such as the sequence detection [4] and nonlinear equalization methods [1,5,6] in the receiver. The sequence detector can provide a maximum-likelihood estimate of the transmitted data sequence, but its complexity increases exponentially as the constellation size and/or the channel memory increase. The use of nonlinear equalizers has also been considered as a practical alternative. Nonlinear equalizers can be designed by modeling the nonlinear channel using the Volterra series [1,5,6]. Although the nonlinear DFE in [5] can significantly reduce the NLD effect, it may require the use of an excessive number of equalizer taps, making its application impractical. An attempt was proposed to reduce the equalizer complexity by simplifying the Volterra series representation of the nonlinear satellite channel [1]. However, it may not be applicable to the QAM signal because this approximation is valid only for phase shift keying (PSK) signals.

In this paper, we propose a combined equalization and NLD cancellation scheme, called nonlinear distortion canceller decision-feedback equalizer (NDC-DFE), for transmission of QAM signals over fixed wireless channel. Assuming the use of a conventional transceiver scheme with perfect synchronization, the discrete-time channel model can be approximated by considering only dominant nonlinear inter-symbol interference (ISI) terms in the useful signal, while treating the residual nonlinear ISI as an additional noise term. Based on this approximate channel model, the proposed receiver scheme is designed so as to eliminate the dominant nonlinear ISI, significantly reducing the implementation complexity.

Following Introduction, Section 2 describes the system model and derives a baseband equivalent nonlinear ISI model using the Volterra series. The proposed nonlinear receiver is described in Section 3. The performance of the proposed scheme is analyzed and verified by computer simulation in Section 4. Finally, concluding remarks are summarized in Section 5.

2. SYSTEM MODEL

Consider a fixed wireless communication system whose baseband equivalent scheme is depicted in Fig. 1. The QAM data symbol $a[n]$ is pulse-shaped by a shaping filter $g(t)$ and then transmitted after passing through the HPA. The HPA introduces memoryless nonlinearity whose baseband equivalent output $s(t)$ is described by [7]

$$s(t) = \sum_{m=0}^{\infty} \lambda_m \, | p(t) |^{2m} \, p(t) \tag{1}$$

where λ_m is the m-th order Taylor series coefficient of the memoryless nonlinearity of the HPA and $p(t)$ is the input of the HPA represented as

$$p(t) = \sum_n a[n] g(t - nT) \tag{2}$$

We assume that the impulse response of the channel can be represented by

$$h(\tau; t) = \sum_l h_l(t) \delta(\tau - lT) \tag{3}$$

where $h_l(t)$ is the gain of the l-th path, $\delta(\cdot)$ is the Kronecker delta function and T is the symbol time interval. The received signal $r(t)$ is matched-filtered by $g^*(-t)$ and then sampled at a symbol rate for the receiver operation, where the superscript * denotes the complex conjugate.

Assuming that the combined impulse response of the transmit and receiver filter satisfies the Nyquist criterion, i.e.,

$$g_T(kT) = \int_{-\infty}^{\infty} g(\tau) g^*(\tau - kT) d\tau = \delta(kT) \quad (4)$$

and that the sampling clock is perfectly synchronized to the incoming signal, the received signal sampled at time $t = nT$ can be expressed as [7]

$$r[n] = \sum_{m=1}^{M} \sum_{k_1} \cdots \sum_{k_{2m-1}} \psi^{(2m-1)}(k_1, k_2, ..., k_{2m-1}) \prod_{i=1}^{m} \prod_{j=m+1}^{2m-1} a[n-k_i] a^*[n-k_j] + w[n] \quad (5)$$

where $(2M-1)$ is the order of the nonlinearity, $w[n]$ is the filtered noise sample and $\psi^{(2m-1)}(k_1, k_2, ..., k_{2m-1})$ is the Volterra series kernel represented as

$$\psi^{(2m-1)}(k_1, k_2, ..., k_{2m-1}) = \sum_l h_l f^{(2m-1)}(k_1 - l, k_2 - l, ..., k_{2m-1} - l) \quad (6)$$

Here,

$$f^{(2m-1)}(k_1 - l, k_2 - l, ..., k_{2m-1} - l) = \lambda_{2m-1} \int_{-\infty}^{\infty} g^*(lT - \tau) \prod_{i=1}^{m} \prod_{j=m+1}^{2m-1} g(\tau - k_i T) g^*(\tau - k_j T) d\tau \quad (7)$$

Since the channel gain can be assumed nearly unchanged during a short time interval in a fixed wireless channel, the time-dependency of $h_l(t)$ can be omitted for simplicity of description.

Unless the HPA is operating in severely nonlinear region, the nonlinear distortion of the HPA can be approximated by a third-order Volterra series representation [7]. The received signal (5) can be expressed as

$$\begin{aligned} r[n] = & \sum_l h_l \{ f^{(1)}(0) a[n-l] + f^{(3)}(0,0,0) \mid a[n-l] \mid^2 a[n-l] \} \\ & + \sum_l h_l \sum_{(i,j,k)'} f^{(3)}(i,j,k) a[n-i-l] a[n-j-l] a^*[n-k-l] \\ & + w[n] \end{aligned} \quad (8)$$

where the index $(i, j, k)'$ in the summation denotes all the indices except $i = j = k = 0$. Here, the first term in (8) represents the ISI term making the signal constellation warped and the second term corresponds to the nonlinear ISI yielding the signal constellation clustered [3].

Unlike linear ISI terms, it may not be easy to compensate nonlinear ISI terms by considering the effect of individual terms from the received signal. It may be feasible to compensate dominant warping distortion from the received signal, while treating the clustering nonlinear ISI as an additional noise. This approach can be justified by the fact that the amount of warping distortion is significantly larger than that of clustering ISI when measured before the channel filtering. The power ratio of the warping distortion to the clustering noise can be represented as

$$\Lambda = G \frac{[f^{(3)}(0,0,0)]^2}{\sum_{(i,j,k)'} [f^{(3)}(i,j,k)]^2} \quad (9)$$

where

$$G = \frac{E\{ | a[n] |^6 \}}{E\{ | a[n-i] |^2 | a[n-j] |^2 | a[n-k] |^2 \}} \quad (10)$$

is a constant related to the signal constellation. For example, G is 1.9600, 2.2258 and 2.2922 for 16-, 64- and 256-QAM, respectively. When a square root-raised-cosine (RRC) filter with roll-off factor $\alpha > 0.2$ is used as the transmit and receive filter, the ratio Λ is larger than 10. Approximating the clustering ISI as an additional noise component, the received signal $r[n]$ can further be simplified to

$$r[n] = \sum_l h_l \left(a[n-l] + \beta \mid a[n-l] \mid^2 a[n-l] \right) + \upsilon[n] \quad (11)$$

where $f^{(1)}(0)$ can be assumed to be one after appropriate scaling without the loss of generality, $\beta = f^{(3)}(0,0,0)$ is the warping distortion coefficient and $\upsilon[n]$ is the total noise term including the AWGN and clustering nonlinear ISI. Since the clustering nonlinear ISI is a sum of large ISI terms with zero mean, it can be assumed a zero mean Gaussian random variable from the central limit theorem. Thus, the noise $\upsilon[n]$ can be approximated as zero-mean Gaussian noise.

3. NDC-DFE

We consider the design of a receiver scheme that can simultaneously compensate both linear and nonlinear distortion. The received signal model (11) implies that the transmit data symbol $a[n]$ suffers from the warping distortion before being transmitted. It may be practical for the receiver to compensate the warping distortion due to the transmitter after compensating the channel effect. This intuitive idea leads to a receiver structure as depicted in Fig. 2, where the linear ISI is handled by a conventional MMSE-DFE and the warping distortion is compensated by two nonlinear modules named cubic block and memoryless NDC, respectively.

If the warping distortion coefficient is known exactly in advance, we can construct a new signal constellation based on the distorted data symbol $a[n-i] + \beta | a[n-i] |^2 a[n-i]$. Then, the transmitted data can be recovered using a conventional MMSE-DFE, where the decision is made based on the new signal constellation. Note that the coefficient of the optimum equalizer is the same as that of the MMSE-DFE when the transceiver is operating in a linear channel with distortion-free signal constellation. If β is unknown, however, this structure may not be applicable since the decision region based on the distorted signal constellation can not be specified exactly. This problem can be resolved by placing the two memoryless nonlinear modules around the slicer, a memoryless NDC before the slicer and cubic block after the slicer, as shown in Fig. 2. We can obtain a new signal constellation by adjusting the coefficient of the two nonlinear modules using a conventional adaptation algorithm. Coupled with the two memoryless nonlinear modules, we can form a modified slicer for the warped signal constellation.

Assuming that the modified slicer makes a proper decision based on the warped signal constellation, the feed-forward filter (FFF) in the DFE can sufficiently eliminate the pre-cursor ISI since the distortion of the channel on the warped signal constellaiton is linear. Then the FFF can be represented as

$$y[n] = \sum_{i=0}^{I_b} q[i] \left(a[n-i] + \beta \mid a[n-i] \mid^2 a[n-i] \right) + \nu[n] \quad (12)$$

where I_b is the number of the post-cursor taps, $q[i]$ is the overall impulse response of the channel and FFF, and $\nu[n]$ is additive noise term. The output of the cubic block is given by

$$b[n] = \hat{a}[n] + \hat{\beta}_c \mid \hat{a}[n] \mid^2 \hat{a}[n] \quad (13)$$

where $\hat{a}[n]$ is the output of the slicer and $\hat{\beta}_c$ is the coefficient of the cubic block. If the data symbols are correctly decoded and the coefficients of the FBF and cubic block are properly initialized, the post-cursor ISI can be subtracted from (12), yielding the input of the memoryless NDC

$$x[n] = a[n] + \beta\,|a[n]|^2\, a[n] + v[n] \quad (14)$$

after being scaled by $1/q[0]$. The second term in (14) can be cancelled out using a memoryless NDC whose input-output relationship is given by

$$z[n] = x[n] - \hat{\beta}_N\,|x[n]|^2\, x[n] \quad (15)$$

where $\hat{\beta}_N$ is the coefficient of the memoryless NDC. The optimum $\hat{\beta}_c$ should be equal to the warping distortion coefficient in order to completely eliminate the post-cursor nonlinear ISI. The optimization of $\hat{\beta}_N$ is not straightforward due to the noise enhancement problem and will be investigated in depth in the next section.

4. PERFORMANCE ANALYSIS

To evaluate the proposed scheme, the performance of the conventional MMSE-DFE in Fig. 3 and proposed NDC-DFE receiver is analytically compared in terms of the symbol error rate (SER) in the presence of NLD. The FFF of the conventional MMSE-DFE consists of a linear equalizer $C(z)$ that minimizes the power spectrum density (PSD) of the error signal defined by

$$\varepsilon[n] = \sum_k c[k]r[n-k] - a[n] \quad (16)$$

and a causal monic filter $E(z)$ that whitens $\varepsilon[n]$. The feedback filter eliminates the ISI introduced by $E(z)$. It can be shown that the error signal can be decomposed into three terms: the residual linear ISI term $A(z)[H(z)C(z)-1]$, nonlinear ISI term $B(z)H(z)C(z)$ and noise term $V(z)C(z)$, where $A(z)$, $B(z)$ and $V(z)$ represent the z-transform of $a[n]$, $\beta|a[n]|^2 a[n]$ and $\upsilon[n]$, respectively, and $H(z)$ is the transfer function of the channel.

It can be shown that the PSD of the error signal is

$$\begin{aligned} S_\varepsilon(e^{j\omega}) &= S_A(e^{j\omega})\,|H(e^{j\omega})C(e^{j\omega})-1|^2 \\ &+ S_B(e^{j\omega})\,|H(e^{j\omega})|^2|C(e^{j\omega})|^2 + S_V(e^{j\omega})\,|C(e^{j\omega})|^2 \\ &+ 2S_{AB}(e^{j\omega})\,\mathrm{Re}\{[H(e^{j\omega})C(e^{j\omega})-1]H^*(e^{j\omega})C^*(e^{j\omega})\} \end{aligned} \quad (17)$$

where $S_A(e^{j\omega})$, $S_B(e^{j\omega})$ and $S_V(e^{j\omega})$ are the PSD of $a[n]$, $\beta|a[n]|^2 a[n]$ and $\upsilon[n]$, respectively, and $S_{AB}(e^{j\omega})$ is the cross power spectrum density (CPSD) of $a[n]$ and $\beta|a[n]|^2 a[n]$. Since $S_\varepsilon(e^{j\omega})$ is a second-order polynomial of $C(e^{j\omega})$, the optimum filter can uniquely be determined as

$$C_{opt}(e^{j\omega}) = \frac{[S_A(e^{j\omega}) + S_{AB}(e^{j\omega})]H^*(e^{j\omega})}{[S_A(e^{j\omega}) + S_B(e^{j\omega}) + 2S_{AB}(e^{j\omega})]\,|H(e^{j\omega})|^2 + S_V(e^{j\omega})} \quad (18)$$

and the corresponing PSD of the error signal is

$$S_{\varepsilon,opt}(e^{j\omega}) = \frac{[S_A(e^{j\omega})S_B(e^{j\omega}) - S_{AB}^2(e^{j\omega})]\,|H(e^{j\omega})|^2 + S_A(e^{j\omega})S_V(e^{j\omega})}{[S_A(e^{j\omega}) + S_B(e^{j\omega}) + 2S_{AB}(e^{j\omega})]\,|H(e^{j\omega})|^2 + S_V(e^{j\omega})} \quad (19)$$

The mean-squared-error (MSE) of the input of the slicer can be calculated by [8]

$$\eta_{MMSE-DFE} = \langle S_{\varepsilon,opt}(e^{j\omega}) \rangle_G \quad (20)$$

where $\langle f(x) \rangle_G$ denotes the geometric mean of $f(x)$. Then, the corresponding SER can be calculated by

$$P_{MMSE-DFE} \approx KQ\left(\frac{d_{\min}/2}{\sqrt{\eta_{MMSE-DFE}}}\right), \quad (21)$$

where K is the average number of neighboring signal points for a given QAM constellation, $d_{\min}$ is the minimum distance of the signal constellation and $Q(\cdot)$ is Gaussain tail function.

For performance analysis of the NDC-DFE, we assume that the FFF and FBF are initialized as the optimum MMSE-DFE and $\hat{\beta}_c = \beta$. It can easily be shown that the variance of $v[n]$ in (14) is equal to that of an MMSE-DFE without the NLD [8]

$$E\{|v[n]|^2\} = \left\langle \frac{S_A(e^{j\omega})S_W(e^{j\omega})}{S_A(e^{j\omega})\,|H(e^{j\omega})|^2 + S_W(e^{j\omega})} \right\rangle_G \quad (22)$$

Substituting (14) into (15), it can be observed that the output of the memoryless NDC has residual noise term comprising weighted high-order products of the data symbol and $v[n]$. The variance of the residual noise is dependent on the magnitude of the data symbol. We need to calculate the conditional SER at first. Assuming that the transmitted data symbol is A, the slicer error can be represented as

$$e(A,\hat{\beta}_N) = -\beta|A|^2 A + v[n] + \hat{\beta}_N\left|UA + v[n]\right|^2\left(UA + v[n]\right) \quad (23)$$

where $U = 1 + \beta|A|^2$. Letting $N_i = E\{|v[n]|^i\}$ and $S_i = |A|^i$, it can be shown that the mean and covariance of $e(A,\hat{\beta}_N)$ are respectively represented as

$$\mu(A,\hat{\beta}_N) = -\beta S_2 A + \hat{\beta}_N(U^3 S_2 + 2UN_2)A \quad (24)$$

$$\sigma^2(A,\hat{\beta}_N) = \{\varsigma_2\hat{\beta}_N + \varsigma_1\hat{\beta}_N + \varsigma_0\} - |\mu(A,\hat{\beta}_N)|^2 \quad (25)$$

where

$$\begin{aligned} \varsigma_0 &= \beta^2 S_6 + N_2 \\ \varsigma_1 &= 2\beta(U^3 S_4 + 2US_2N_2) + 2(2U^2S_2N_2 + N_4) \\ \varsigma_2 &= U^6 S_6 + N_6 + 9U^4 S_4 N_2 + 9U^2 S_2 N_4 \ . \end{aligned} \quad (26)$$

Thus, the conditional SER can be calculated by

$$\begin{aligned} P(A,\hat{\beta}_N) \approx\ & K_{left}(A)\,Q\left(\sqrt{\frac{2[1-\mathrm{Re}\{\mu(A,\hat{\beta}_N)\}]^2}{\sigma^2(A,\hat{\beta}_N)}}\right) + K_{right}(A)\,Q\left(\sqrt{\frac{2[1+\mathrm{Re}\{\mu(A,\hat{\beta}_N)\}]^2}{\sigma^2(A,\hat{\beta}_N)}}\right) \\ &+ K_{up}(A)\,Q\left(\sqrt{\frac{2[1-\mathrm{Im}\{\mu(A,\hat{\beta}_N)\}]^2}{\sigma^2(A,\hat{\beta}_N)}}\right) + K_{down}(A)\,Q\left(\sqrt{\frac{2[1+\mathrm{Im}\{\mu(A,\hat{\beta}_N)\}]^2}{\sigma^2(A,\hat{\beta}_N)}}\right) \end{aligned} \quad (27)$$

where $K_{dir}(A)$ is the number of the nearst neighboring signal points in direction *dir* with respect to the data symbol A which is equal to either 0 or 1. The SER for a given $\hat{\beta}_N$ can be calculated by averaging the conditional SER over the data symbols

$$P_S(\hat{\beta}_N) = \sum_A P(A,\hat{\beta}_N)\,p_A(A) \quad (28)$$

where is is the probability of the data symbol A. The optimum SER can be obtained by

$$\hat{P}_S = \min_{\hat{\beta}_N} P(\hat{\beta}_N) \quad (29)$$

To evaluate the performance of the proposed NDC-DFE receiver, we consider the QAM signal transmission over a frequency selective fixed wireless channel whose impulse response is $\{-0.2+j0.3,\ 1.0,\ 0.5-j0.2,\ -0.1+j0.1\}$. The transmit data is pulse shaped using an RRC filter with roll-off factor 0.3. The third order nonlinear channel in Fig. 1 is generated using the baseband equivalent model of (8), where the NLD is defined as

$$\gamma_N = 10\log_{10}\frac{E\{|p(t)|^2\}}{E\{|s(t)-p(t)|^2\}}\ (dB) \quad (30)$$

The tap size of the FFF and FBF is 15 and 5, respectively.

Fig. 4 depicts the SER performance of uncoded 64- and 256-QAM transceivers when the NLD is 22dB and 28dB, respectively. This corresponds to a weak nonlinear distortion condition, i.e., the NLD level is slightly higher than the background noise level. It can be seen that the analysis agrees quite well with the simulation results and that the proposed NDC-DFE outperforms the MMSE-DFE, providing an SNR gain of 3.2dB and 3.8dB in the 64- and 256-QAM transceiver at 10^{-4} SER, respectively.

Fig. 5 depicts the analytic SNR required for transmission of 16-, 64- and 256-QAM signals at 10^{-4} SER for different values of γ_N. It can be seen that the proposed NDC-DFE is getting more effective to the NLD as the QAM level increases. This is because the performance degradation of the NDC-DFE is mainly due to the noise enhancement by the memoryless NDC. In the case of high level QAM signaling such as 64- or 256-QAM, the variance of the enhanced noise is much smaller than that of the background noise level. When the NLD is large, the MMSE-DFE cannot effectively reduce the nonlinear ISI particularly in the outer-most signals because linear filters are not effective for removing the signal-dependent distortion.

When the NLD is the major impairment affecting the receiver performanace, it may be desirable to reduce the transmit power level. This HPA back-off can alleviate the performance degration due to the NLD at the expense of reduced link margin to the background noise. Total performance degradation of the receiver in the presence of NLD can be determined by a sum of the performance degration due to the NLD and the reduction of the noise margin due to the HPA back-off. The performance degradation caused by the NLD is defined as the additional SNR required to obtain a specific SER compared to the required SNR with the use of a linear HPA.

Fig. 6 depicts total degradation of the MMSE-DFE and NDC-DFE for 256-QAM signalling at 10^{-4} SER as a function of the HPA back-off and the roll-off factor α. It can be seen that the proposed NDC-DFE requires less back-off and thus results in total degradation less than the MMSE-DFE. Note that the difference in total degradation at the optimum back-off is reduced as α decreases. It is mainly due to the fact that the disregarded clustering ISI terms become comparable to that of the cancelled nonlinear ISI and begins to dominate the receiver performance as α decreases.

5. CONSLUSIONS

We have proposed a simple nonlinear equalization scheme that can effectively reducce the NLD effect. The structure of the NDC-DFE is derived by approximating the Volterra series representation of a nonlinear channel. The performance of the proposed NDC-DFE is analyzed and verified by computer simulation. Numerical results show that the NDC-DFE is quite robust to a wide range of the NLD when the QAM constellation size is large. The performance gain of the NDC-DFE over the MMSE-DFE becomes larger as the constellation size and/or the NLD increase, quite suitable for transmission of high-level QAM signals in fixed wireless channel.

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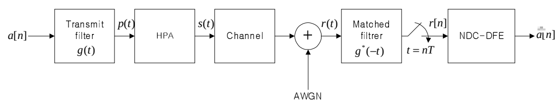

Fig. 1. System model

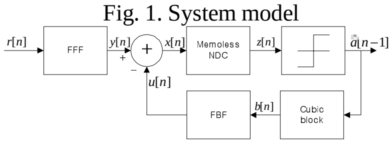

Fig. 2. Proposed NDC-DFE receiver.

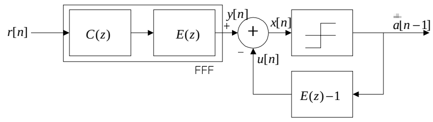

Fig. 3. Conventional MMSE-DFE receiver.

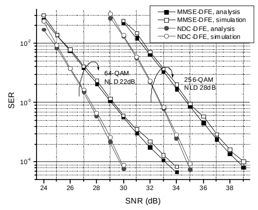

Fig. 4. SER performance of MMSE-DFE and NDC-DFE.

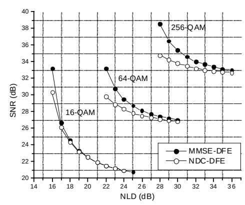

Fig. 5. Required SNR for 10^{-4} SER.

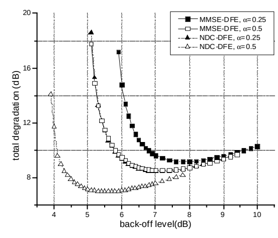

Fig. 6. Total performance degradation for 256-QAM.