MEMORYLESS POLYNOMIAL ADAPTIVE PREDISTORTION

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ABSTRACT

In this paper we investigate algorithms to adaptively adjust the coefficients of memoryless polynomial structures used to precompensate for the nonlinear amplitude and phase distortion of the high-power amplifier in a terrestrial digital television transmitter. The results of the investigation are twofold. First the phase error is a non-Euclidean measure of the absolute symbol error. For small inputs, noise causing a small Euclidean change can create a large phase error. We compensate for this heuristically by not updating the predistorter coefficients for small inputs. This thresholding is shown to decrease the residual error of the phase predistorter. Second,the pre-compensation nature of the amplitude correction requires a modification to the traditional LMS algorithm. This modification will be seen to produce a smaller residual error than traditional LMS. We demonstrate the superior performance of our algorithms via simulations based on the measured characteristics of production high-power amplifiers.

1. INTRODUCTION

In terrestrial transmission the most expensive component in the transmitter is the high power amplifier (HPA). Since the HPA's high cost largely comes from its RF power transistors, the goal is to use as few RF power transistors as possible to create a given output power. However, using too few transistors to meet a given design specification causes the transistors to saturate for larger inputs, creating significant nonlinear amplitude and phase distortion. The tradeoff is between the system's cost and its maximum output power rating (linear amplification range).

Predistortion, the mapping of the input signal through an approximation to the inverse of the HPA's nonlinearity, has been used for years in analog terrestrial television transmitters to help reduce the overall system cost. Low-cost, manually-tunable circuits in the low power, intermediate frequency (IF) stage are the preferred technology. This approach is well-suited to analog television, which experiences a graceful degradation in picture quality with respect to variations in the HPA's nonlinear characteristic due to temperature change and aging. However, digital television, as a result of its increased data rates, suffers from the cliff effect[1]. This causes the received signal quality to be sensitive to variations in the HPA characteristic. Fortunately, the advent of digital television brings with it the ability to use digital filtering, especially adaptive digital filtering.

Using traditional adaptive algorithms to compensate for the time-varying nature of the HPA's nonlinearity in a digital transmitter has already been proposed[1-5]. However, there are problems with directly applying LMS and RLS. In this paper, we modify the LMS and RLS algorithms for updating the coefficients of memoryless polynomial predistorters. The modification to the RLS phase predistortion algorithm is strictly heuristic, we do not update the filter coefficients for small symbol amplitudes because in this situation the phase information has an extremely poor SNR. The modification to the LMS algorithm arises in re-deriving the equations for amplitude predistortion. The update term [7] gains an extra factor proportional to the gradient of the nonlinearity. These new predistortion algorithms will be shown to produce a lower residual error power than direct application of unmodified LMS and RLS.

In this paper we present modified adaptive predistorter algorithms which are to be incorporated into ITS corporation's upcoming line of digital television terrestrial transmission products. In section two we formally identify the adaptive predistortion problem for digital television terrestrial transmission. In section three we derive the general form for using LMS to adjust the weights of the amplitude predistorter and present a heuristic improvement to RLS for phase distortion. Simulation results using the measured characteristic of a production HPA as the nonlinearity, presented in section four, confirm the expected performance gains of the new algorithm. Conclusions and future work will be discussed in the final section.

2. PROBLEM STATEMENT

A typical transmitter is shown in figure 1. In this figure, the data is separated into two channels representing either QAM data, where each channel carries independent information, or VSB data, where one channel is data and the other is its Hilbert transform.

In either case the data is passed through pulse-shaping filters to create the input to the adaptive predistorter. The

[1] Cliff effect describes the phenomenon of error correction providing nearly perfect reception up to the point that the error correction is overloaded when total picture loss is experienced

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pulse-shaping filters provide Nyquist shaping for zero ISI and oversample. The oversampling allows the predistorter to "see" outside the data channel and correct out-of-band emissions created by the nonlinearity of the HPA. The predistorted data is then modulated, amplified and transmitted. We can now formulate the adaptive predistortion problem.

Since we are dealing with complex modulation we use the following notation to represent the baseband data.

$$I_n + jQ_n = R_n \cos(\theta_n) + jR_n \sin(\theta_n)$$

$$R_n = \sqrt{I_n^2 + Q_n^2} \text{ and } \theta_n = -\arctan\left(Q_n/I_n\right) \quad (1)$$

The input to the HPA is given by,

$$\hat{R}_n \cos(\omega t + \theta_n) \quad (2)$$

The output of the HPA(3) is known[1-5] to contain amplitude distortion(AM/AM) and phase distortion (AM/PM).

$$A_n\left(\hat{R}_n\right) \cos\left(\omega t + \hat{\theta}_n + \phi_n(\hat{R}_n)\right) \quad (3)$$

It is important to note that both distortions, A_n and ϕ_n, depend only on the amplitude of the input to the HPA and are independent of the input phase. This allows us to correct for each of the distortions independently by first compensating for the amplitude distortion and then adding in a phase correction.

The demodulated baseband symbols are

$$\tilde{I}_n = A_n\left(\hat{R}_n\right) \cos\left(\hat{\theta}_n + \phi_n(\hat{R}_n)\right)$$

$$\tilde{Q}_n = A_n\left(\hat{R}_n\right) \sin\left(\hat{\theta}_n + \phi_n(\hat{R}_n)\right) \quad (4)$$

The goal of predistortion is to generate warped values, $\hat{I}_n$ and $\hat{Q}_n$, so that the recovered baseband symbols $\tilde{I}_n$ and $\tilde{Q}_n$ are equal to I_n and Q_n. This requires solving two independent problems: First, the magnitude of the data must be warped such that,

$$A_n\left(\hat{R}_n\right) = R_n \quad (5)$$

Second, the input phase must be warped such that,

$$\hat{\theta}_n + \phi_n\left(\hat{R}_n\right) = \theta_n \quad (6)$$

The optimal solution to (6) problem is obviously,

$$\hat{\theta}_n = \theta_n - \phi_n\left(\hat{R}_n\right) \quad (7)$$

This is a system identification problem. We identify and then subtract out the AM/PM nonlinearity.

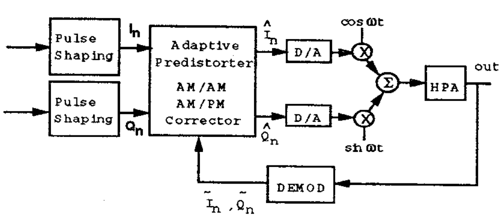

Figure 1: Digital Transmitter

The magnitude problem is more difficult since we must estimate the inverse of the AM/AM. It is this problem that requires the development of the new algorithm in the next section.

To proceed, we must choose structures for the amplitude predistorter and the phase predistorter. It is well known in the literature[6], and through our laboratory measurements, that each of these distortions is well-modeled by a polynomial, the AM/AM by an odd power polynomial and the AM/PM by an even power polynomial. Figure 2 shows the measured and modeled AM/AM distortion while figure 3 shows the measured and modeled AM/PM distortion of a typical HPA. To choose structures for the predistorters we assume that the predistorter/HPA interaction is memoryless. This assumption is valid because the SAW filters used in the IF and RF sections have group delays that are far less than a symbol time and thus introduce no ISI into the system.

Thus, the AM/PM predistortion structure is an even power memoryless polynomial used to identify and then subtract off the additive phase nonlinearity as indicated in (6). A 16th order, even power (9 term) polynomial is sufficient for the worst case.

For the AM/AM predistorter we, as have previous researchers[1], use a higher order odd power memoryless polynomial. The algorithm that we use to update this structure is developed in the following section. A 9th order, odd power (5 term) polynomial is sufficient for the worst case. Note that in figure 2 that the amplification has been normalized to one and that the corresponding maximum output amplitude of the HPA is approximately .62. Thus, no matter how large the input, the output can never exceed .62. It is the goal of the AM/AM predistorter to make the HPA appear linear over the entire input range of (-.62,.62).

3. ALGORITHM DEVELOPMENT

In this section we describe the algorithms used to update the coefficients of the amplitude and phase predistorters.

Phase predistortion is a fairly straightforward system identification problem except for the following caveats. First,

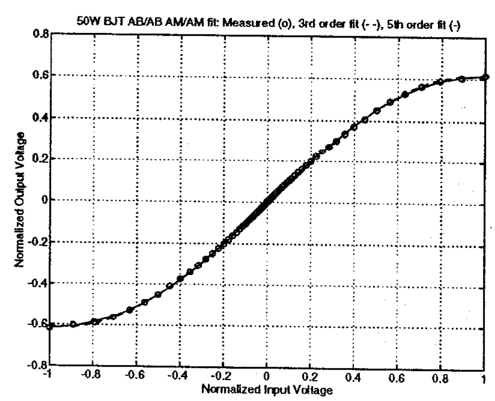

Figure 2: BJT class AB AM Char.

the high order of the phase predistortion polynomial makes LMS converge slowly. Thus, we use RLS to adjust its coefficients. Second, the phase error is not linearly related to the symbol errors: If the symbol amplitude is small, a tiny amount of noise can cause a large phase error that is not representative of the actual symbol error (recall the goal is to make $I_n = \tilde{I}_n$ and $Q_n = \tilde{Q}_n$). To combat this, we do not update the phase predistorter unless the input amplitude is above a chosen level. We call this thresholding and feel it is justifiable since the HPA is linear for small inputs and we get to knowingly discard data with a very low SNR.

Adjusting the weights of the AM/AM predistorter requires developing the predistortion LMS (PLMS) algorithm. Since the AM/AM problem is independent of the AM/PM problem we can use figure 4 to describe the problem. A memoryless polynomial filter, $H_n(R_n)$ operates on the data, R_n, to compensate for the nonlinearity, $A_n(\widehat{R}_n)$. The general form for the polynomial predistorter is given by the following.

$$H_n(R_n) = \widehat{R}_n = \sum_{i=1}^{n} h_k(i) R_k^i \tag{8}$$

The error from this compensation, denoted by e_n, is given by the following.

$$e_n = R_n - \tilde{R}_n = R_n - A\left[H(R_n)\right] \tag{9}$$

In LMS the procedure is to use steepest descent to find the filter coefficients that minimize the mean square output error, $E\left(e_n^2\right)$, by estimating the gradient of the mean square error with its instantaneous value. Thus, the well-known update equation is given by,

$$h_{n+1}(i) = h_n(i) - \alpha \nabla_h |e_n|^2 \tag{10}$$

The difference from traditional LMS created by the prefiltering configuration arises when we evaluate the gradient in (10). The first step is to apply the chain rule.

$$\nabla_h e_n^2 = -2 e_n \nabla_{\widehat{R}_n} A\left(\widehat{R}_n\right) \cdot \nabla_{h_n} \widehat{R}_n \tag{11}$$

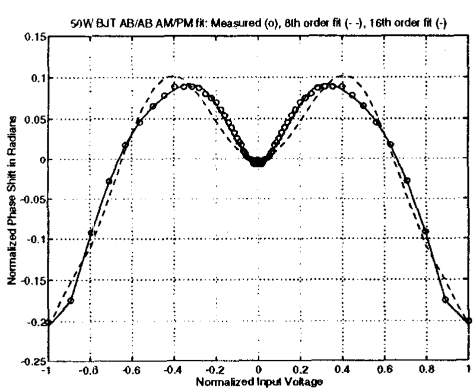

Figure 3: BJT class AB PM Char.

$\nabla_h \widehat{R}_n$ is obtained by considering a single element of the gradient vector along with (8).

$$\frac{\partial}{\partial h_n(i)} \widehat{R}_n = R_n^i. \tag{12}$$

Thus, the PLMS filter update equation becomes

$$h_{n+1}(i) = h_n(i) - \alpha e_n R_n^i \nabla_{\widehat{R}_n} A_n\left(\widehat{R}_n\right) \tag{13}$$

The difference between (13) and traditional polynomial LMS [7] is the factor $\nabla_{\widehat{d}_n} A_n\left(\widehat{d}_n\right)$. In looking at figure 2, this new factor reduces the magnitude of the update where the nonlinearity is strongest and prevents the filter from overcompensating. This will be evidenced by the superior performance of the new algorithm in the next section.

In general, $\nabla_{\widehat{d}_n} A_n\left(\widehat{d}_n\right)$ is not known; however, we can well approximate the saturation characteristic, A_n, with an odd power polynomial and differentiate it explicitly. In practice, a stairstep could be input to an amplifier and the differences of the outputs used as the gradient. Since the saturation characteristic varies slowly, these differences could be calculated at startup each day and stored in a lookup table.

4. SIMULATION RESULTS

In the last section we developed and discussed the algorithms to adjust the coefficients of the AM/AM and AM/PM predistorters. In this section we demonstrate the performance of the algorithms via simulation. This section is broken down into the following three parts: First, we examine the AM/AM corrector by itself. Second, we examine the AM/PM corrector by itself. Finally, we examine the overall performance. All of the simulations use the laboratory measured characteristics of figures 2 and 3 as the distortions.

For the AM/AM corrector figures 5 shows the ensemble average of 60 runs. Uniformly distributed 8 level PAM signals were input to a pulse-shaping filter, a predistorter and then the HPA saturation characteristic. The input range is (-.62,.62), the maximum possible correctable range. The output of this system was then subtracted from the input to compute the error. The same seeds were used for both LMS and PLMS. Each algorithm used a 9th order structure to correct for the saturation characteristic of figure 2. The step sizes used for each algorithm were the best discovered after many experimental runs. Notice in figure 5 that PLMS exhibits a smaller residual error than LMS and a smaller deviation from the minimum error.

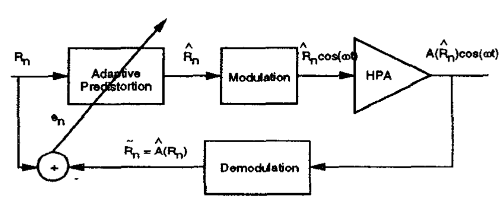

Figure 4: Transmitter

For the AM/PM corrector we demonstrate the improvement available by thresholding. We ran simulations (same noise and seeds) with and without thresholding and compared the magnitudes of their respective phase errors. The simulations were of a 64-QAM transmitter with -60dB of measurement noise. Figure 6 plots the difference in phase error with and without thresholding. Negative error indicates that thresholding was detrimental, positive error indicates that thresholding was beneficial. The proliferation of positive error, even with an SNR of 60 dB, demonstrates the benefit of thresholding. The lower the SNR the more improvement thresholding offers.

Finally, we put it all together in figure 7 and plot the magnitude of $I_n + jQ_n$ minus $\tilde{I}_n + j\hat{Q}_n$ for a 64-QAM transmitter with -60dB of measurement noise. The signal levels for each channel are -7,-5,...5,7. with an average magnitude of 6.4. To correct to the noise floor (-60dB) requires the average residual error magnitude to be .0064. The measured average residual error was 0.006376. This final simulation demonstrates the benefit available with predistortion. We were able to create a perfectly linear transmitter, up through the maximum possible output power.

5. CONCLUSION

In this paper we altered and improved adaptive algorithms for use in memoryless adaptive polynomial predistortion. Phase predistortion was improved with a heuristic modification of the RLS algorithm. Amplitude predistortion was improved by deriving and implementing a predistortion LMS algorithm. Simulations, based on laboratory measured nonlinearities, demonstrated the benefits of these modifications. Issues such as implementation, phase jitter of the oscillators and distortion of analog filters will be addressed in a future full length paper.

6. REFERENCES

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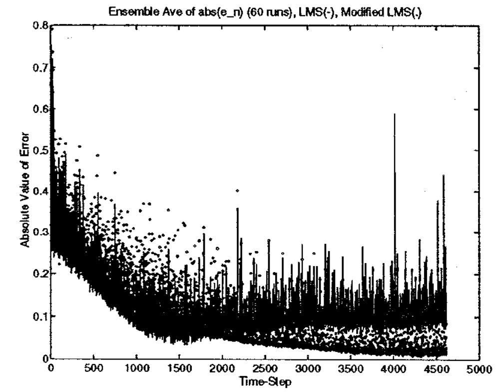

Figure 5: Absolute Error Comparison

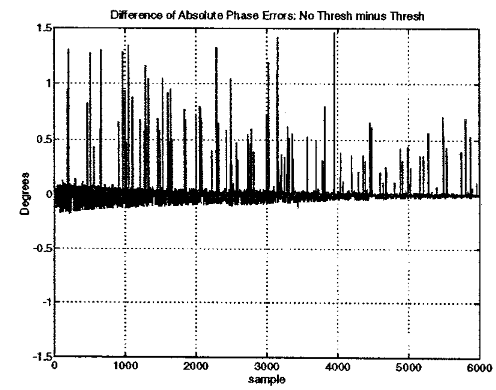

Figure 6: Phase Err w/o Thresh - Phase Err with Thresh

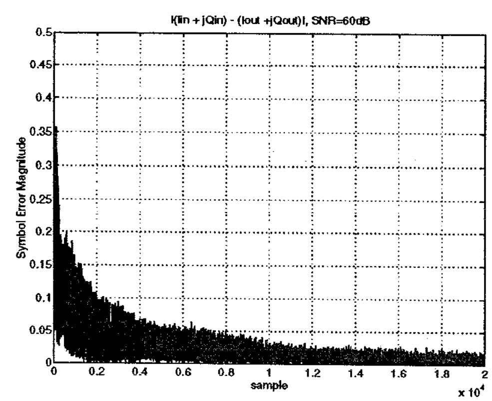

Figure 7: Compensation Comparison