A Novel DSP Architecture of Adaptive Feedforward Linearizer for RF Amplifiers

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Abstract **— A novel adaptive design for feedforward amplifier linearizer with DSP control is proposed in this paper. Compared to existing adaptive architectures, this is a "blind" design which does not require pilot signal and intentional signal perturbation and phase calibration. A polar gradient adaptive algorithm is also developed to support the hardware architecture to provide the unconditional convergence during the full working range of the phase control components. The stability criterion is analyzed. The linearizer performance for multi-tones and for CDMA signal is simulated and demonstrated using EDA design tools.**

I. INTRODUCTION

Among numerous amplifier linearization techniques, feedfoward linearization has been extensively utilized in base-station amplifiers for wireless communication because of its intrinsic advantages on providing high linearity over a wide frequency band. However, it requires very accurate balances of the amplitude, phase and delay over the frequency band. To compensate the parameter changes dynamically due to varying operating conditions such as temperature, input power level and supply voltages, an adaptive control circuitry is essential.

There are a number of adaptive approaches have been proposed. Of the available control algorithms, the gradient method [1]-[2] is considered to be the best choice because no pilot signal [3]-[4] or intentional signal perturbation [5] is needed. In this type of algorithm, gradient signals are used to constantly adjust the circuit parameters in a direction toward the global minimum of the error surface. The gradient signals are generated by performing correlations between the error signal and the reference signal, using either analog circuits [1] or DSP technology [6]. The latter is more advantageous because the DC offset caused by analog mixing can be avoided.

Nevertheless, we found that the conventional gradient architectures have conditional stability problems [7], when the phase control component, a vector modulator, is placed before the main or auxiliary amplifier. This type of configuration is used as an option to minimize the modulator distortion for the main amplifier loop. However, for the error amplifier loop, the phase shifter can only be placed in front of the error amplifier since the other branch is a path of high power where no phase shifter should be put in. Therefore, the loop convergence may be lost when amplifier devices have certain phase shifts. The underlying reason is found to be that the amplifier phase diverts the gradient vector from pointing the global minimum of the error surface. To overcome this problem, a polar gradient algorithm is proposed here. It generates gradients with regarding to the signal amplitude and phase angle, whose pointing directions are irrelevant to the amplifier phase shift. This guarantees the stability over any phase shift of the amplifier. It also gives much better tolerance to the non-ideality of circuit components compared to the conventional gradient approach.

In this paper, the stability of conventional gradient approaches is first analyzed in Section II, followed by the proposal of the new architecture and polar algorithm. In Section III, with the RF/DSP co-simulation capability of Agilent eesof CAD software Advanced Design System (ADS), the amplifier linearizer system is simulated for two-tone signal and for IS95 CDMA signals respectively. The suppression of intermodulation and spectral regrowth is thus demonstrated

II. ARCHITECTURE AND OPERATING PRINCIPLES

The fundamental structure of feedforward amplifier linearizer consists of two signal cancellation loops. Based on the assumption that amplifier output signal is the sum of amplified reference signal and error signal, the reference signal is cancelled from the attenuated amplifier output in the first loop, which leaves only the error signal. In the second loop, the error signal is amplified and cancelled from the amplifier output, which leaves only the amplified reference signal as the final output. Though theoretically feedforward structure can provide completely distortion free output, the actually achievable distortion suppression is dependent on how well the signal cancellation is performed, when the signal phase and delay variation has to been taken into account. The amplifier delay is not so sensitive to the environment and can be compensated by a fixed delay line. Hence the delay

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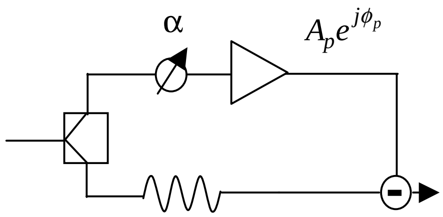


Fig.1 Sketch of conventional signal cancellation loop.

mismatch can normally be ignored when the frequency range is not too wide, while the phase control is of more importance for most of the cases.

As depicted in Fig.1, the conventional implementation of signal cancellation loop is using vector modulator in front of the amplifier to adjust the signal phase and amplitude to match the reference signal [6]. We assume that the complex coefficient for the vector modulator is α and the complex transfer function of the amplifier is $A_p e^{j\phi_p}$ without considering the amplifier delay. Therefore, α should be controlled converging toward $A_p{}^{-1} e^{-j\phi_p}$ for perfect signal cancellation. This is achieved by generating the gradient signal $\Delta\alpha$ from the correlation between the error signal output and the reference signal, which has the following form,

$$\Delta\alpha = G \cdot (1 - A_p e^{j\phi_p} \cdot \alpha) = G \cdot A_p e^{j\phi_p} (A_p^{-1} e^{-j\phi_p} - \alpha) \quad (1)$$

where G is the gain constant of the control loop. In the above formula, $\Delta\alpha$ is zero when α converged to the right value. However, it should be noticed that $\Delta\alpha$ is actually different with the true gradient in a factor $A_p e^{j\phi_p}$. This means the resulted gradient vector points to a direction different with pointing to the global minimum of the error surface in an angle ϕ_p. In the worst case when the

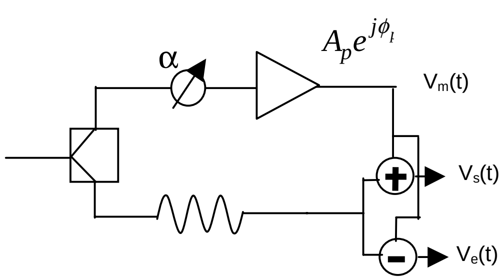
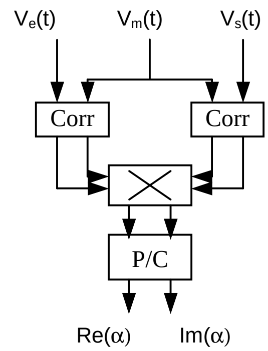


Fig.2 Sketch of proposed signal cancellation loop.



Corr: Complex RF Correlator
X : Complex Baseband Multiplier
P/C : Polar to Cartesian Converter

Fig.3 Block diagram of polar gradient algorithm.

amplifier has 180 degree phase shift, the gradient vector points to the contrary direction of convergence. Since we do not have control in the amplifier phase when we select the devices, this will cause the instability of the circuits for certain amplifier phase shifts. The existing of this problem is also confirmed in the simulations we carried out, which will be described in Section III.

To overcome this problem, a novel implementation of the signal cancellation and detection loop is shown in Fig.2. What is different is that the combiner generates not only the difference but also the sum of two signals. Both of them are used to generate the gradient signals in the polar coordinates. The block diagram of the algorithm is depicted in Fig.3. Basically it performs two correlations and one complex multiplication, which results in the gradients respectively for amplitude and aspect of α

$$\begin{cases} \Delta|\alpha| = G \cdot (1 - A_p^2 |\alpha|^2) \\ \Delta\angle\alpha = G \cdot \sin(-\phi_p - \angle\alpha) \end{cases} \quad (2)$$

As we can see from (2), these gradients are true gradients that ensure the amplitude and aspect of α to converge respectively. They are also independent to the amplifier phase angle. The control coefficients for the vector modulator can thus be generated by using a polar-to-Cartisan converter.

Compared to the conventional scheme, this new configuration has the drawback of needing one more detector. However, if we notice that in the main amplifier

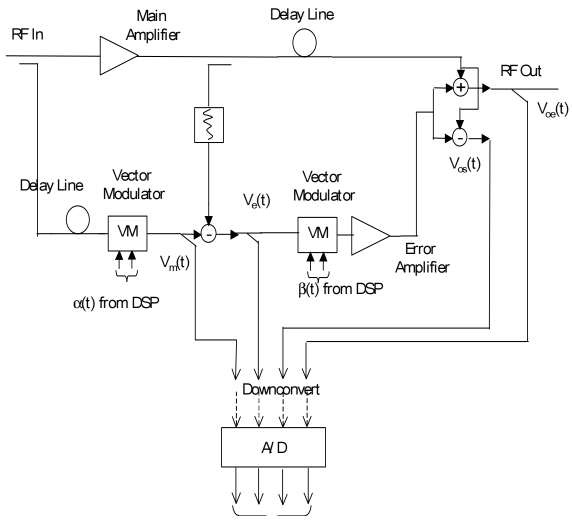


Fig.4 Schematic of proposed adaptive feedforward amplifier linearizer.

loop, the vector modulator can be placed in the reference branch to get unconditional stability [2], the above architecture based on polar gradients can be applied only to the error amplifier loop. Therefore, a unconditional stable feedfoward architecture with the minimum hardware expense is obtained as shown in Fig.4.

III. Simulation and Validation

To illustrate the efficacy of the algorithm, a few simulations are carried out using ADS RF/DSP simulator.

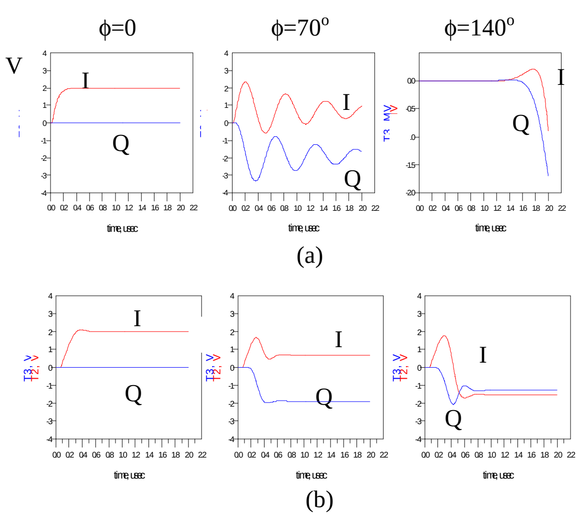


Fig.5 Convergence simulation of a(t) versus different amplifier phase shift (a) using algorithm proposed in [6]. (b) using polar gradient algorithm.

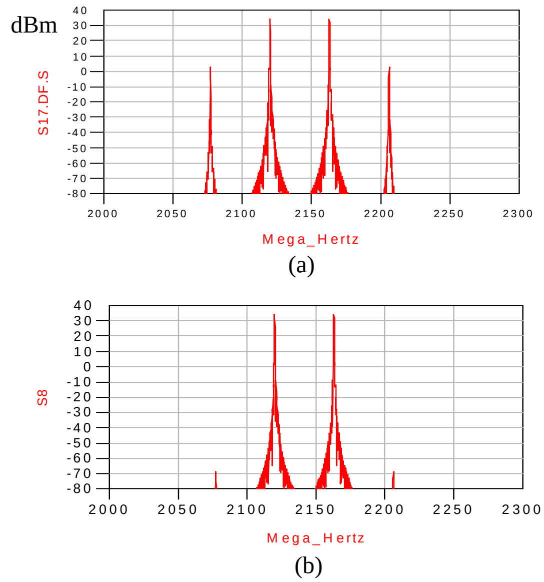


Fig.6 Output spectra with two-tone excitation. (a) before linearization (b) after linearization.

The first simulation is to test the signal cancellation loop using two tones at 2.12 GHz and 2.16 GHz. Fig.5 (a) shows the initial convergence of the control voltage $\alpha(t)$ for amplifier phase shifts at 0, 70 and 140 degrees respectively, using the conventional architecture [6]. As we can see, the algorithm converges within 300 ns when the amplifier has no phase shift. It converges much slower when the phase shift is 70 degrees. It fails to converge when the phase shift is 140 degrees. On the other hand, the simulation result shown in Fig.5 (b) using polar gradient architecture always converges within 800 ns. The second simulation is two-tone test of the whole feedforward linearizer architecture. We use the ADS nonlinear amplifier model with an assumed 140 degree phase shift. Setting the operation point of the main amplifier at the 3-dB back-off from the 1-dB compression, the output frequency spectra at a few micro-seconds after the linearizer acts are plotted in Fig.6 (b), compared against the spectra with linearization in Fig.6 (a). It shows a 74 dB reduction of the IMD level and the IMD3 is 104 dBc.

Unlike the steady state multi-tone signals, CDMA signals has a time-varying wave form and a high peak-to average ratio. Thus the amplifier linearization for CDMA signal is a much tougher problem. Here, a dynamic simulation is also carried out for IS95 CDMA signals using the proposed feedforward architecture. The signal has a center frequency at 2.14 GHz and a 16 MHz bandwidth. The signal peak-to-average ratio is about 8~12 dB. The main amplifier is assumed to have a 1-dB compression at 51dBm and the AM-PM modulation is assumed to be 1.5dB/degree. The operating point is chosen so that the average output power is at 10dB back-off from the 1-dB compression. The error amplifier is assumed to have 1-dB compression point at 41 dBm and a 90 degree phase shift. The simulation shows that the first loop converges within 300 micro-seconds while the second loop converges within 800 miro-seconds. The resulted signal spectra is plotted in Fig.7, from which we can see about 35 dB reduction of spectral regrowth is obtained when compared to the signal spectra without through linearization. In fact, after examining the original signal, we found that the linearizer output spectra has basically no difference with the original signal spectra without any distortion, which means almost perfect suppression of distortion is achieved using the proposed architecture.

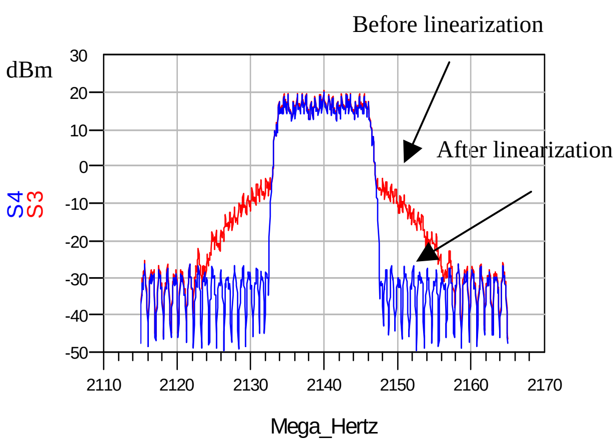


Fig.7 Spectral regrowth reduction for IS95 CDMA.

IV. Conclusion

A novel DSP architecture for adaptive feedforward amplifier linearizer has been proposed by using a new design of signal cancellation and detection circuit. To control the circuit adaptively, a robust algorithm has been developed, based upon the gradient concept in polar coordinates. The proposed architecture has the property of unconditional convergence of the control loop independent to the amplifier phase shift. Various simulations have been carried out to validate the approach. The linearizer has demonstrated more than 70 dB intermodulation suppression for multi-tone signal and 35 dB spectral regrowth reduction for CDMA signal. The hardware implementation of the architecture is currently in process in UCLA.

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