Efficient Baseband/RF Feedforward Linearizer through a Mirror Power Amplifier Using Software-Defined Radio and Quadrature Digital Up-Conversion

E. G. Jeckeln, F. M. Ghannouchi, M. Sawan and F. Beauregard*

Department of Electrical and Computer Engineering
Ecole Polytechnique de Montréal
P.O. Box 6079, Station. Centre-Ville, Montreal, QC, Canada, H3C 3A7

*AmpliX Wireless & Satcom
3333 Queen Mary Rd., R-320
Montreal, Quebec, Canada

Abstract — **This paper describes an efficient feedforward linearizer suitable for base station 3G power amplifiers based on and using both software-defined radio and quadrature digital up-conversion technologies. The linearizer accomplishes the extraction of the PA complex nonlinear behavior in real time and performs a numerical quasi-perfect carrier's cancellation that allows the reduction of the power rating of the error amplifier (EA) and eliminates the requirement for any output delay line. This enhances the overall power efficiency of the feedforward amplifier. The entire system is validated using DSP/RF co-simulation for the LDMOS model of a typical 44 dBm class AB power amplifier (PA_{AB}) as the main amplifier.**

I. INTRODUCTION

While the third generation (3G) of mobile radio standards is being defined, the demand for developing ultra-linear microwave transmitter, supporting high crest factor signals is greater than ever. The demanding adjacent channel power ratio (ACPR) requirements of these new systems, i.e., W-CDMA or cdma2000, present a critical issue for transmitter designers if both ultra-linearity and high power efficiency must be met. In fact, the degradation of linearity becomes significant as the PA operates close to saturation where both high power efficiency and high output power emission are achieved. Therefore, for different stimulus levels driving the amplifiers and for a given ACPR specification, the trade-offs between power efficiency and linearity impose an operating condition with poor power efficiency. In this case, linearization techniques become the only possible way to recuperate the non-linearity and to allow optimal trade-offs.

Various linearization methods have been reported and are derived, by any measure, from three main types named: i) Feed-forward, ii) Feedback and iii) Predistortion. Referring to the first one, the feed-forward technique is characterized for its broadband performance [1-2]. It has a high level intermodulation reduction that is achieved with an unconditionally stable condition. A drawback of this technique is its low power efficiency due to the high average power requirement needed in the error amplifier and the insertion loss from couplers and the delay line at the output of PA. The error signal needs to be amplified up to the necessary power level with high linearity and therefore, it constrains the error amplifier to work in Class A (EA_A) with high output back off. One way to minimize the power requirement of the error amplifier is to reduce the residual carrier power from the carrier cancellation loop, but it is limited to the typical value of –30 dBc depending upon the performance of the phase and gain tracking of the cancellation loop.

Today, the high clock speeds in DSPs, which are entering the microwave frequency bands, offer the possibilities to migrate most of the analog signal processing to the digital domain [3-6]. Hence, we propose in this paper an efficient baseband/RF feedforward linearizer that incorporates advanced software defined radio and quadrature digital up converter [7]. By taking advantage of these technologies to digitally translate signal from IF to baseband and vice versa, the linearizer performs the instantaneous characterization of the memoryless transfer function of the PA, which is used to develop the first cancellation loop in digital domain. The added benefits of this technique are that the power requirement of the error amplifier is reduced and the output delay line is removed enhancing the overall power efficiency of the feedforward amplifier.

II. DESCRIPTION OF THE PREDISTORTER

Fig. 1 shows the block diagram of the linearizer. Following the RF signal path, the RF signal is picked up from the input and the output of the PA and translated down to within an alias-free sampling range from DC up to 35 MHz.

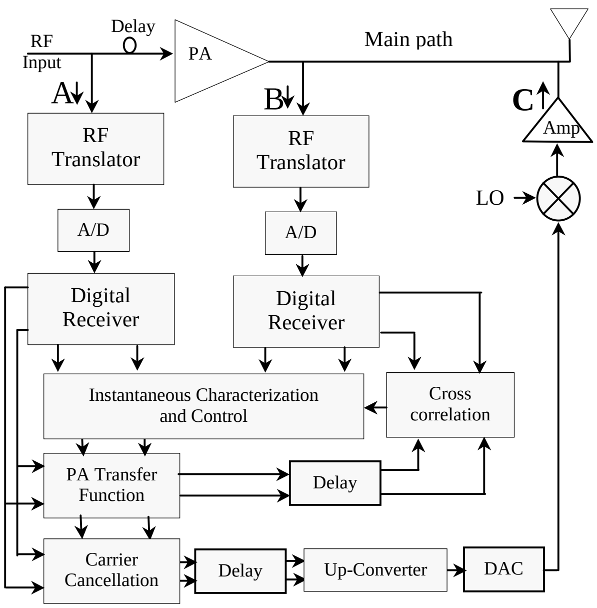

Fig. 1 General block diagram of the linearizer

Right after the translator stages, one for each branch, the signals are conditioned by 12-bit A/D converters into digital samples at the high rate of 70 MHz. The entire subsequence complex down converting, filtering, and decimating is performed digitally by two digital receivers [7]. It is understood that, the complex envelop from the output of the PA in branch "B" bring the information of non-linearity when it is driven further into nonlinear operation mode. This non-linearity information can be discriminated when this complex envelop is referenced to the complex envelop from the input of the PA in branch "A". It means that the instantaneous characterization (i.e., AM-AM and AM-PM curves) can be performed following both complex envelopes variation during a real work condition [8]. It allows correlating the PA transfer function in real time and to control the adaptation update step, through the cross correlation block, whenever it is necessary to be update into the PA transfer function block. Notice that, the cross correlation block compares the distorted signal from the main PA with the distorted signal from the PA transfer function allowing to truck any change due to temperature variation, aging, etc. In order to perform the carrier cancellation loop, the complex envelop from the branch "A" is split into the two paths at the input of the PA transfer function and then, the carrier cancellation block process the suppression given a free carrier error signal. The resulting error signal containing only distortion components is processed through the branch "C" and then added to the main path for distortion cancellation purpose. Notice that, the delay line is placed at the input of the PA, and it is compensated by software.

In terms of algorithms, the translation and filtering process are the two majors signals processing operations performed by the digital receiver blocks. First, a single-sideband complex translation is accomplished by mixing the real signal with the complex output of a digital quadrature local oscillator. Then, a decimation filter conditions the complex baseband signal by fixing an appropriate value of the decimation parameter M [7]. It controls the reduction of the cutoff frequency f_{cutoff} and the sampling rate f_s as follows:

$$f_{cutoff} = f_s/2M \,, \tag{1}$$

$$f' = f_s/M \,. \tag{2}$$

It means that, any signal can be selected digitally from the IF domain and put it into the baseband domain for further processing. In this way, the two receiver output data, that represent the stimulus and response of the PA, are routed through the instantaneous characterization block to carry out the nonlinearity behavioral. Assuming that the input bandpass signal is given by

$$v_i(t) = \mathrm{Re}\{\rho(t) e^{j[\omega_c t \; + \theta(t)]}\} \,, \tag{3}$$

where ω_c is the midband angular frequency, $\rho(t)$ is the amplitude variation and $\theta(t)$ is the phase variation. Then, the bandpass output signal can be represented by

$$v_o(t) = \mathrm{Re}\{g[\rho(t)] e^{j\{\omega_c t \; + \phi[\rho(t)] + \theta(t)\}}\} \,, \tag{4}$$

where $g[\rho(t)]$ and $\phi[\rho(t)]$ are two memoryless nonlinear functions that represent the instantaneous AM-AM and AM-PM curves. Notice that these functions are characterized in terms of the input and output bandpass complex envelopes, without including all harmonics effect.

III. RESULTS AND DISCUSSIONS

The entire system is validated using DSP/RF co-simulation for the LDMOS model of a typical 44 dBm class AB power amplifier. Temperature noise, quantization noise, and impairment from other components have been taken into consideration to realistically model the system. The digital receivers are considered as both narrowband and wideband where the decimation values are ranged from 2 to 2^{17}. The performance of the feedforward system is evaluated by applying CDMA standard signal which stress level is characterized by its complementary cumulative distribution function (CCDF) as shown in Fig 2.

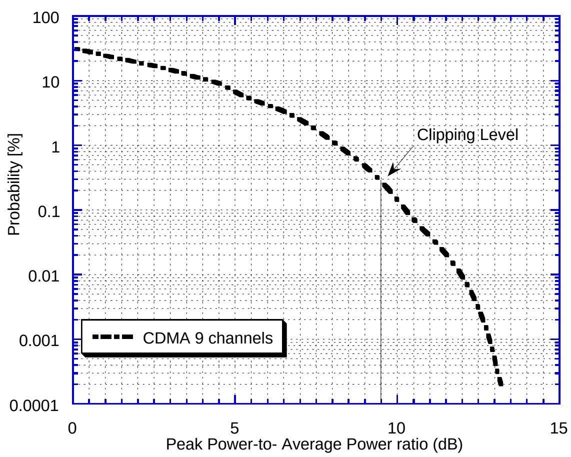

Fig.2 The CCDF plots of CDMA standard signals

The CCDF curves illustrate the statistical property of the nine channel standard signal built in simulation; the simulator generates around 800,000 signal samples that allow attaining the 0.0001% probability value with high stability.

For analysis purpose, the ACPR is evaluated by computing the ratio between the in-band and the out-of-band power spectral densities for three pairs of offset channels. The channels are normalized to the same bandwidths of *30 kHz* at the offset frequencies of *±885 kHz*, *±1.256 MHz*, and *±2.75 MHz*. The normalization factor (NF) is calculated by logging the ratio between the normalization bandwidth NBW, i.e., 30 kHz, and the specified bandwidth BW, as follow:

$$NF(dB) = 10\log\left(\frac{NBW}{BW}\right) \qquad (5)$$

By applying this equation to the J-STD-008 standard requirements, the relative limit values of the power spectral densities for the three offsets are given by:

$$R_1(dB) = 28.8\ dB \qquad (6)$$

$$R_2(dB) = 28.8\,dB + \Delta\ dB \qquad (7)$$

$$R_3(dB) = 47.8\,dB + \Delta\ dB \qquad (8)$$

where Δ is given by:

$$\Delta = \begin{cases} P_{PA}\,(dBm) - 35.8dBm & \text{for } P_{PA} \geq 35.8\,dBm \\ 0 & \text{elsewhere} \end{cases} \qquad (9)$$

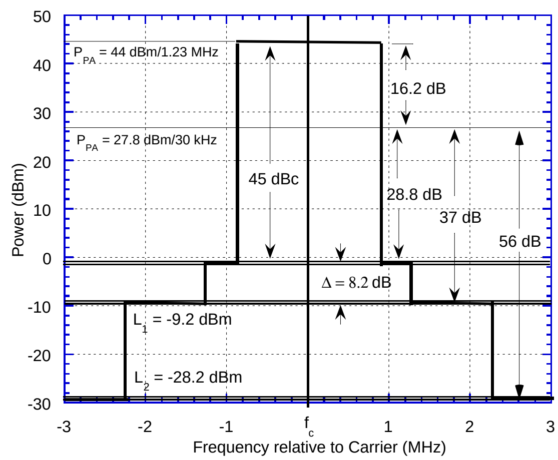

Fig.3 Graphical representation of 30 kHz normalized CDMA standard requirement for *44 dBm* PA

In Fig 3 we can see a graphical representation of requirement for a typical *44 dBm* PA; after normalization the mean power become *27.8 dBm/30 kHz*, Δ take the value *8.2 dB* and the relative limit values are R_1= *28.8 dB*, R_2=*37 dB* and R_3=*56 dB*.

The power calculation for each channel is accomplished by converting the time domain signal to frequency domain using Hanning window and with FFT length of 8192 points. The integration bandwidth (IBW) method is used to calculate both mean channel power and offset channel powers. In addition, the simulator computes an average power for each specified integration channel bandwidth and over a specified number of data acquisitions, *avg.*= 16.

Several important results have been obtained, the main of them are summarized in table 1. The OBO referenced to Single Carrier (SC) saturation, the Crest Factor (peak power-to-average power (CF)), the word-length of bits, the ACPR (R_1, R_2, R_3) and the power efficiency $\eta\%$ are chosen as important factors to analyze the performance of the whole system.

The numbers in the Table 1 indicate the operating condition of the feedforward amplifier (PA_{FF}) under the standard requirement. The PA_{FF} achieves 44 dBm with an overall power efficiency of 10%. The result revels about 5% greater than a conventional feedforward. Notice that the PA_{AB} operate at 9.5 dB OBO to achieve 45 dBm with an efficiency of 16%. Under this operating condition, the clipping level becomes 3.5 dB from the CF as shown in Fig. 2. The total insertion losses of the couplers are considerate to be 0.55 dB. To achieve an acceptable distortion cancellation, the P_{1dB} of the EA_A was optimized given an average power capability of 28.8 dBm operating at 18 dB of OBO. This is a critical operation condition showing a poor efficiency of 1.5%.

TABLE 1
OPERATING CONDITION OF THE PA_{FF} UNDER THE STANDARD REQUIREMENT

	OBO [dB]	P_{out} [dBm]	η [%]	R_1 [dB]	R_2 [dB]	R_3 [dB]
PA_{AB}	9.5	45	16	33	36	52.8
EA_A	18	28.8	1.5	-	-	-
PA_{FF}	-	44	10	43	45	56.2

IV. CONCLUSION

An efficient baseband/RF feedforward linearizer has been presented. This new design improves the overall power efficiency by about 5% in comparison with conventional feedforward systems. The improvement is achieved by eliminating the delay line at the output of the power amplifier and by reducing the average power rating of the error amplifier. Progress in this technique should be a control loop for up-converter's LO in branch C to perform good matching frequency between mean path and error path. Finally, the fact that the system can be controlled and tuned by software, this design provides an attractive solution for feedforward amplifiers mass production with almost no need for unit's test bench adjustment.

REFERENCES

[1] R. Meyer, R. Eschenbach and W. Edgerley, Jr." A Wide-Band Feedforward Amplifier", IEEE J of Solid-State Circuits, vol.sc-9, no. 6, pp. 442-428, December 1974.

[2] J. Cavers, "Adaptation Behavior of a Feedforward Amplifier Linearizer", IEEE Transactions on Vehicular Technology, vol. 44, no 1, pp. 31-40, February 1995.

[3] Y. Nagata, "Linear Amplification Technique for Digital Mobil Communications", in Proc. IEEE Veh. Technol. Conf., San Francisco, CA, 1989, pp. 159-164.

[4] J. Cavers, "Amplifier Linearization Using a Digital Predistorter with Fast Adaptation and Low Memory Requirement", IEEE Transactions on Vehicular Technology, vol. 39, no 4, pp 374-382, November 1990.

[5] E.G. Jeckeln, F.M. Ghannouchi and Mohamad Sawan, "Adaptive Digital Predistorter for Power Amplifiers with Real Time Modeling of Memoryless Complex Gains", IEEE MTT-S 1996 International Microwave Symposium, San Francisco, CA, June 1996.

[6] E.G. Jeckeln, F.M. Ghannouchi and Mohamad Sawan, " An L Band Adaptive Digital Predistorter for Power Amplifiers Using Direct I-Q Modem ", IEEE MTT-S 1998 International Microwave Symposium, Baltimore, MA, June 1998.

[7]E. G. Jeckeln, F. M. Ghannouchi and F. Beauregard "Adaptive Baseband/RF Predistorter for Power Amplifiers through Instantaneous AM-AM and AM-PM Characterization Using Digital Receivers" IEEE MTT-S International Microwave Symposium, Boston, June 11-16, 2000, pp. 489-492.

[8] E. G. Jeckeln, F. M. Ghannouchi and M. Sawan, ‘‘ Non Iterative Adaptive Digital Predistortion Technique for Power Amplifiers linearization ‘‘, U.S. Patent 6,072,364, June 6, 2000.