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RF PA LINEARIZATION BY SIGNALS MODIFIED IN BASEBAND DIGITAL DOMAIN

Aleksandra Đorić[1], Nataša Maleš-Ilić[2], Aleksandar Atanasković[2]

[1]Innovation Center of Advanced Technologies, Niš, Serbia
[2] University of Niš, Faculty of Electronic Engineering, Niš, Serbia

Abstract. *This paper represents the linearization of the RF power amplifier performed by a new approach that combines two different methods exploiting the modified baseband signals. The signals for linearization in both methods are formed and processed in digital domain. The required modified baseband signals for linearization are products of the second order nonlinearity of a nonlinear system fed by the useful baseband signal. In the first method, adequate part of the modified baseband signal is adjusted in amplitude and polarity and injected at the input and output of the amplifier transistor across the series LC resonant circuit. In the second method, the appropriate modified baseband signal set on the appropriate amplitude and phase modulates the fundamental carrier second harmonic, which is then inserted at the input and output of the amplifier transistor. The effects of the combined linearization method are considered on a single stage power amplifier for quadrature amplitude modulated signals characterized with frequency spacing between spectral components up to 60 MHz for different input power levels, as well as for WCDMA digitally modulated signal.*

Key words: *linearization, power amplifier, baseband signal, second harmonic, intermodulation products.*

1. INTRODUCTION

The new generation of the communication technologies and standards impose demanding requirements to new systems in order to increase bit rate, linearity and spectrum efficiency. These requirements present a serious task for the transmitter designers regarding the power amplifier (PA) topology that should support wideband operation, various modulation formats, a diversity of signal bandwidths and frequency ranges, high efficiency, as well as linear operation [1]. For achieving high power efficiency, the power amplifier should operate closer to its compression region distorting the linearity of the output signals. Consequently, significant efforts have been devoted to development of linearization techniques for nonlinear RF and microwave power amplifiers. Various linearization methods

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Corresponding author: Aleksandra Đorić
Innovation Center of Advanced Technologies, Bulevar NikoleTesle 61/5, 18 000 Niš, Serbia
(E-mail: alexdjoric@yahoo.com)

for minimizing nonlinear distortions of power amplifiers have been reported in the literature [2-5]: feedback, feed-forward, predistortion, etc.

In the previously deployed linearization technique that uses the fundamental signals' second-order (IM2) and fourth-order nonlinear signals (IM4) at frequencies around the second harmonics, [6-12], the signals for linearization were generated and prepared for injection at the input and output of the transistor amplifier in the RF analogue signal domain. Linearization effects were validated on the single stage RF power amplifiers throughout the simulation process [6]-[8] and experiments [9], as well as on the Doherty amplifiers [10-12].

In this paper, we combine two linearization methods that exploit modified baseband signals formed and processed in the digital domain [13], [14] in order to linearize the RF amplifier. In the first method, an adequately prepared signal in the baseband is adjusted in amplitude and polarity and injected at the input and output of the amplifier transistor across the LC resonant circuit**.** In the second method, specified baseband signals modulate the carrier second harmonic after appropriate setting on the amplitude and phase, and the formed signal is run at the gate and drain of the amplifier transistor.

The injected signals for the linearization and the fundamental signal are mixed due to the second order nonlinearity of the transistor generating additional third-order nonlinear products that may suppress the original intermodulation products caused by the transistor nonlinear characteristic. The impact of the proposed linearization methods is considered on a single stage power amplifier for QAM signals wherein I and Q components are single tones with maximum spectrum bandwidth 60 MHz, as well as for the WCDMA digitally modulated signal.

The results obtained by the combined method are validated by comparison with the results achieved by the first and second linearization approaches, which are described in [13] and [14]. Since the linearization results for the first method are represented in [13] for only one power level of the QAM signal, in this paper we have analysed its linearization effect for a range of the input signal powers.

This paper is organized as follows: Section II explains the theoretical background of the combined linearization approach; Section III represents the results of linearization of the designed single-stage RF power amplifier obtained by the combined method proposed in this paper, which are also compared to the results obtained by the first linearization approach; In Section IV conclusions are reported.

2. THEORETICAL ANALYSIS OF THE COMBINED LINEARIZATION TECHNIQUE

The operational principle of the linearization method proposed herein can be comprehended by the theoretical analysis of the current nonlinearity at the transistor output in the amplifier circuit. The dominant nonlinearity of FETs can be represented by a Taylor-series polynomial model [15-17] in case when the memory effects are neglected

$$\begin{aligned} i_{\mathrm{ds}} = {} & g_{\mathrm{m1}}v_{\mathrm{gs}} + g_{\mathrm{m2}}v_{\mathrm{gs}}^2 + g_{\mathrm{m3}}v_{\mathrm{gs}}^3 + g_{\mathrm{d1}}v_{\mathrm{ds}} + g_{\mathrm{d2}}v_{\mathrm{ds}}^2 + g_{\mathrm{d3}}v_{\mathrm{ds}}^3 + \\ & + g_{\mathrm{m1d1}}v_{\mathrm{gs}}v_{\mathrm{ds}} + g_{\mathrm{m2d1}}v_{\mathrm{gs}}^2v_{\mathrm{ds}} + g_{\mathrm{m1d1}}v_{\mathrm{gs}}v_{\mathrm{ds}}^2 + \ldots \end{aligned} \tag{1}$$

The transistor's drain current (i_{ds}) depends on the voltage between the gate and source (v_{gs}), which is expressed by the transconductance terms labeled by g_{mx}. The dependence of the drain current on the voltage between the drain and source (v_{ds}) is included by the drain conductance terms g_{dy}. In addition, the drain current is a function of voltage between the gate and source and voltage between the drain and source, which are represented by the coefficients g_{mxdy}. The order of each coefficient can be calculated as x + y.

The digitally modulated signal is characterized by the magnitude c(t), phase $\phi(t)$, and carrier frequency ω_0, as

$$\begin{aligned} v_{\text{in}}(t) &= c(t)\cos(\omega_0 t + \phi(t)) = c(t)[\cos(\phi(t))\cos(\omega_0 t) - \sin(\phi(t))\sin(\omega_0 t)] \\ &= v_{\text{s}}(I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \tag{2}$$

where $I = (c(t)/v_s)\cos(\phi(t))$ and $Q = (c(t)/v_s)\sin(\phi(t))$ are the in-phase and quadrature-phase components of the baseband signal.

The second-order nonlinear system generates the output signal given by

$$v_{\text{out\_2\_order}}(t) = v_{\text{in}}^2 = \frac{1}{2}v_{\text{s}}^2[I^2 + Q^2] + \frac{1}{2}v_{\text{s}}^2[(I^2 - Q^2)\cos(2\omega_0 t) - 2IQ\sin(2\omega_0 t)] \tag{3}$$

if the digital signal expressed by Eq.2 is fed at its input.

The linearization approach suggested in this paper utilizes the complete modified baseband signal from the Eq.3: the baseband signal in appropriate form (the first term) together with the fundamental carrier second harmonic modulated by the adequately shaped baseband signal whose modified (new) in-phase and quadrature-phase components are in the forms $I_{new} = I^2 - Q^2$ and $Q_{new} = 2IQ$ (the second and third terms).

Fig. 1 shows the schematic diagram of the amplifier with the linearization circuit that forms, processes, and injects the linearization signals at the input and output of the amplifier transistor in case of the combined linearization method. The baseband part of the signal for linearization in the form $BB_{mod} = I^2 + Q^2$ is multiplied by the coefficients $a_{\{ib/ob\}}$ for amplitude and polarity tuning. The baseband signals modified in this manner are then inserted over the series LC circuit at the gate and drain of the amplifier transistor. Another modified components of the linearization signal, I_{new} and Q_{new}, which modulate the carrier second harmonic, are adjusted in amplitude by $a_{\{i2h/o2h\}}$ and phase by $\theta_{\{i2h/o2h\}}$ across two branches. The prepared signals are then inserted at the input and at output of the amplifier transistor through the bandpass filters. The aforementioned indices consisting of the letters i and o in subscript are related to the signals injected at the input and output of the amplifier transistor, respectively.

The voltage at the gate, given as,

$$\begin{aligned} v_{\text{gs}}(t) &= v_{\text{s}}[I\cos(\omega_0 t) - Q\sin(\omega_0 t)] + a_{\text{ib}}(I^2 + Q^2) + \\ &+ a_{\text{i2h}}e^{-j\theta_{\text{i2h}}}[(I^2 - Q^2)\cos(2\omega_0 t) - 2IQ\sin(2\omega_0 t)] + V_{\text{G}} \end{aligned} \tag{4}$$

is comprised of all signals injected: the fundamental useful signal, the gate bias signal, the modified baseband signal and the fundamental carrier second harmonic modulated by the adequately shaped baseband signal.

The voltage at the drain, given as,

$$\begin{aligned} v_{\text{ds}}(t) &= v_{\text{o}}[I\cos(\omega_0 t) - Q\sin(\omega_0 t)] - a_{\text{ob}}(I^2 + Q^2) + \\ &- a_{\text{o2h}}e^{-j\theta_{\text{o2h}}}[(I^2 - Q^2)\cos(2\omega_0 t) - 2IQ\sin(2\omega_0 t)] + V_{\text{D}} \end{aligned} \tag{5}$$

consists of the fundamental signal amplified linearly, the drain bias voltage and the signals for linearization including the baseband signal and modulated second harmonic, which are appropriately modified, tuned, and injected together at the amplifier transistor drain. In Eq. 5 $v_{\mathrm{o}}(I\cos(\omega_0 t)-Q\sin(\omega_0 t))$ is the output signal at the fundamental frequency, V_{G} and V_{D} are DC bias voltages supplied at the gate and drain of the amplifier transistor, respectively.

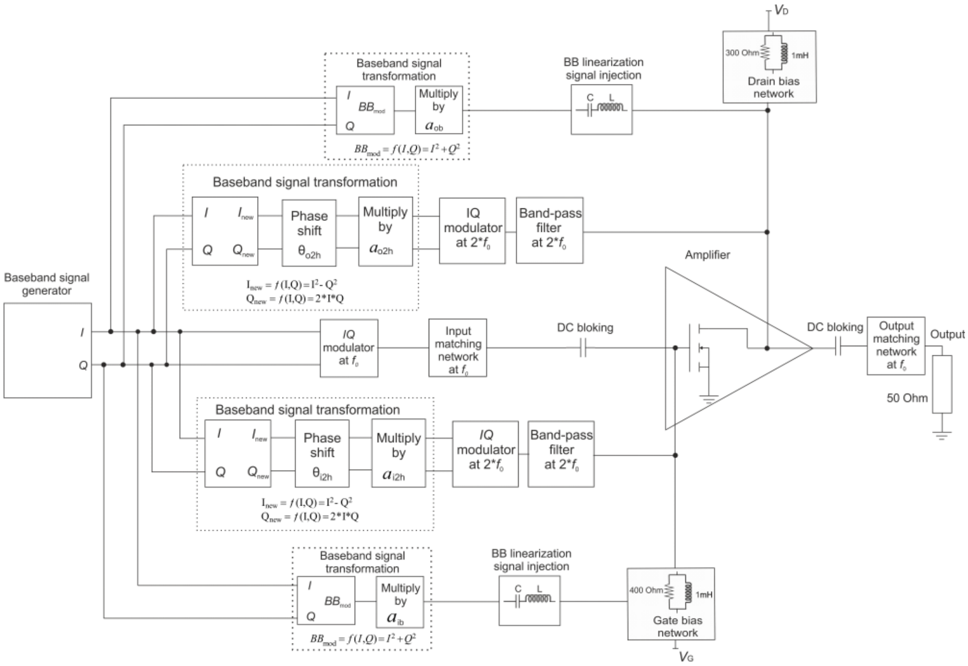


Fig. 1 Schematic diagram of the amplifier linearized by the injection of the modified baseband signals processed in digital domain

The distorted output current is obtained by substituting the Eq. 4 and Eq. 5 in Eq. 1. yielding

$$
\begin{aligned}
i_{\mathrm{ds}}(t)\big|_{IM3} = \Big(& \frac{3}{4} v_{\mathrm{s}}^3 g_{\mathrm{m3}} + 2 a_{\mathrm{ib}} v_{\mathrm{s}} g_{\mathrm{m2}} + a_{\mathrm{i2h}} e^{-j\theta_{\mathrm{i2h}}} v_{\mathrm{s}} g_{\mathrm{m2}} - a_{\mathrm{ob}} v_{\mathrm{s}} g_{\mathrm{m1d1}} + \\
& - \frac{1}{2} a_{\mathrm{o2h}} e^{-j\theta_{\mathrm{o2h}}} v_{\mathrm{s}} g_{\mathrm{m1d1}} + a_{\mathrm{ib}} v_{\mathrm{o}} g_{\mathrm{m1d1}} + \frac{1}{2} a_{\mathrm{i2h}} e^{-j\theta_{\mathrm{i2h}}} v_{\mathrm{o}} g_{\mathrm{m1d1}} + \\
& + \frac{3}{2} v_{\mathrm{s}} v_{\mathrm{o}}^2 g_{\mathrm{m1d2}} + \frac{3}{2} v_{\mathrm{s}}^2 v_{\mathrm{o}} g_{\mathrm{m2d1}} \Big) (I^2+Q^2)[I\cos(\omega_0 t) - Q\sin(\omega_0 t)] \qquad (6)
\end{aligned}
$$

$$
\begin{aligned}
i_{\mathrm{ds}}(t)\big|_{IM5} = \Big(& \frac{5}{8} v_{\mathrm{s}}^5 g_{\mathrm{m5}} + 3 a_{\mathrm{ib}}^2 v_{\mathrm{s}} g_{\mathrm{m3}} + \frac{3}{2} a_{\mathrm{i2h}}^2 e^{-j2\theta_{\mathrm{i2h}}} v_{\mathrm{s}} g_{\mathrm{m3}} + a_{\mathrm{ob}}^2 v_{\mathrm{s}} g_{\mathrm{m1d2}} - 2 a_{\mathrm{ib}} a_{\mathrm{ob}} v_{\mathrm{o}} g_{\mathrm{m1d2}} + \\
& + \frac{1}{2} a_{\mathrm{o2h}}^2 e^{-j2\theta_{\mathrm{o2h}}} v_{\mathrm{s}} g_{\mathrm{m1d2}} - a_{\mathrm{i2h}} a_{\mathrm{o2h}} e^{-j(\theta_{\mathrm{i2h}}+\theta_{\mathrm{o2h}})} v_{\mathrm{o}} g_{\mathrm{m1d2}} - 2 a_{\mathrm{ib}} a_{\mathrm{ob}} v_{\mathrm{s}} g_{\mathrm{m2d1}} + \\
& + a_{\mathrm{ib}}^2 v_{\mathrm{o}} g_{\mathrm{m2d1}} + \frac{1}{2} a_{\mathrm{i2h}}^2 e^{-j2\theta_{\mathrm{i2h}}} v_{\mathrm{o}} g_{\mathrm{m2d1}} - a_{\mathrm{i2h}} a_{\mathrm{o2h}} e^{-j(\theta_{\mathrm{i2h}}+\theta_{\mathrm{o2h}})} v_{\mathrm{s}} g_{\mathrm{m2d1}} \Big) \\
& (I^2+Q^2)^2 [I\cos(\omega_0 t) - Q\sin(\omega_0 t)] \qquad (7)
\end{aligned}
$$

where Eq. 6 and Eq. 7 refer to the third- and fifth-order intermodulation products of the drain current at the fundamental frequency, respectively.

The nonlinearity of the drain current in terms of the voltage between the drain and source, v_{ds}, is expressed by the coefficients $g_{d1} – g_{d3}$ and according to [16] and [17] it is assumed to have an inessential impact on the intermodulation products and have been omitted from the equations.

The first term in Eq. 6 represents the signal distorted by the cubic term of the amplifier (g_{m3}), which is considered as a dominant in arousing the third-order intermodulation products, IM3, and spectral regrowth [16], [17]. The g_{m2} second-order transconductance nonlinear products of the fundamental signal and the linearization signals injected at the amplifier transistor gate are expressed as the second and third terms in the Eq.6. The fourth and fifth g_{m1d1}terms are the mixing products between the gate-source voltage of the fundamental signal and the voltage of the linearization signals fed at the amplifier transistor drain. Additionally, the fundamental signal at the output of the transistor mixes with the linearization signals driven at the amplifier transistor input generating the sixth and seventh terms. The drain current at IM3 frequencies includes the mixing products of the third-order, g_{m1d2} and g_{m2d1}, between the drain and gate voltages of the fundamental signal (the eighth and ninth terms in Eq. 6). Since the output signal at the fundamental frequency is considered to be 180 degree out of phase in reference to the input signal, these products reduce each other [17] due to their opposite phases.

According to previous analysis, it is possible to reduce spectral regrowth caused by the third-order distortion of the fundamental signal by selecting the appropriate amplitude and polarity of the modified baseband signal injected at the input (a_{ib}) and output (a_{ob}) of the amplifier transistor, as well as by choosing the adequate amplitude and phase of the modified baseband signal that modulates the second harmonic injected at the input (a_{i2h}, θ_{i2h}) and output (a_{o2h}, θ_{o2h}) of the amplifier transistor.

The first term in Eq.7 is formed due to the amplifier nonlinearity of the fifth-order, (g_{m5}) and expresses the fifth-order intermodulation products of the drain current of the amplifier transistor. The mixed terms between the drain and gate, g_{m1d2} and g_{m2d1} are the products between the fundamental signal and the baseband linearization signal as well as the fundamental signal and the modulated second harmonic, which exist at the amplifier transistor input or output. It is supposed that these terms of the drain current at the IM5 frequencies neutralize each other to a certain extent, which depends on the phase relations between the linearization signals driven at the gate and drain as well as the intensity of the mixing products. However, the second and third g_{m3} terms in Eq. 7 may increase or decrease the IM5 products owing to the signs of the third- and fifth-order nonlinear coefficients g_{m3} and g_{m5} and also to the linearization signal phase θ_{i2h}.

3. LINEARIZATION RESULTS

In order to estimate the effects of linearization, the proposed combined approach is applied on the broadband RF amplifier designed in Agilent Advanced Design System-ADS software. The designing process is based on the nonlinear MET model of the Freescale transistor MRF281S LDMOSFET and includes synthesis of the input and output broadband matching circuits with the lumped elements [7]. The source and load impedances of the amplifier transistor were determined by the source-pull and load-pull analysis in ADS entailing high drain efficiency and maximum output power [8]. The amplifier circuit was designed to operate over the frequency range 0.7 GHz-1.1 GHz. We considered the influence of the bandpass filters (ideal elements from the ADS library) connected to the gate and drain of the amplifier transistor to supply the linearization signal, which comprises the modulated fundamental carrier second harmonic, to the amplifier circuit. The series LC circuits that enable injection of the modified baseband signals for linearization into the amplifier were also included into analysis.

The gain, power-added efficiency (PAE) and output power of the amplifier loaded by the LC circuits and bandpass filters, in terms of the input power, is shown in Fig. 2 for the single-tone excitation. It can be noted that the maximum gain observed at 1 GHz is slightly greater than 22 dB, showing a variation of approximately 2 dB with the change of the excitation signal frequency. The power added efficiency at labeled frequencies deviates from the PAE at 1 GHz by maximum 5%, whereas the maximum PAE is 50 % at 1 GHz at maximum output power of around 36 dBm.

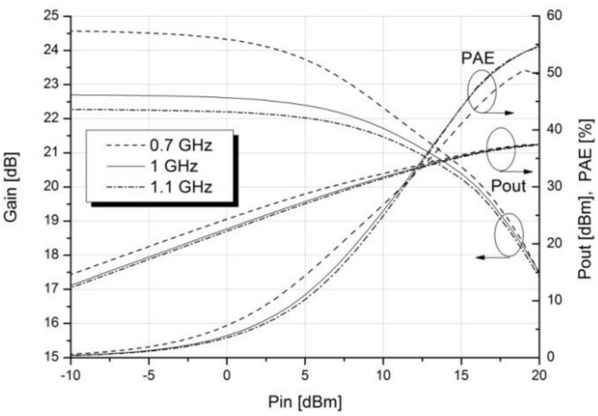


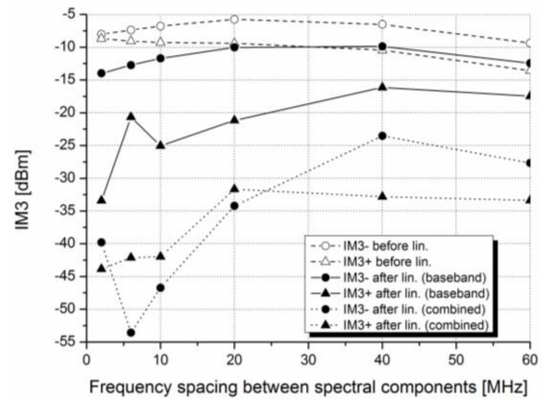
Fig. 2 Gain, PAE and Pout of the design amplifier with the series LC circuits and bandpass filters loading gate and drain of the amplifier transistor

The designed power amplifier was tested for the QAM modulated signals whose spectrum contains two frequency components separated by 2 MHz up to 60 MHz with centre frequency of 1 GHz. Through the ADS simulations, timed source component named QAM was used as a source of the signals. The analysis was carried out for different fundamental signal power levels at the amplifier input: 0 dBm, 3 dBm and 7 dBm. The power levels of the third-order and the fifth-order intermodulation products, before and after the linearization, in terms of the frequency interval between the spectral components of the QAM signal are presented in Fig. 3 and Fig. 4 for different input power levels. It should indicate that the values of the linearization coefficients $a_{\{ib/ob\}}$ for amplitude and polarity tuning of the baseband signals and coefficients $a_{\{i2h/o2h\}}$ and $\theta_{\{i2h/o2h\}}$ for amplitude and phase adjustment of the linearization signals that modulate the carrier second harmonic, were obtained by the optimization process in ADS for each considered input signal power level. The Random optimization of the adjustable coefficients of the linearization signals was carried out with the aim to suppress the third-order intermodulation products and to restrain the fifth-order intermodulation products at the levels below the reduced IM3 products.

We compared two cases: when the linearization was achieved by insertion of the only modified baseband signals at the input and output of the amplifier transistor (the first or baseband method) [13] and when the linearization was performed by the combined approach, i.e. a simultaneous injection of the adequately modified baseband signals together with the second harmonic of the fundamental carrier modulated by another differently modified baseband signal at the input and output of the amplifier transistor. The combined linearization approach encompasses the linearization methods aforementioned above as the first (baseband) and second methods, [13], [14].

Figure 3 represents the third-order intermodulation products, IM3, before and after the linearization for the compared linearization cases. It can be noted that, greater reduction of the IM3 products for all input power levels over the considered power range was achieved by the combined linearization approach proposed in this paper. The effects of the linearization method that exploits only the modified baseband signal was proposed and tested in [13] for a specific input power level of the QAM signal and a range of input power for the WCDMA signal. In this paper, we obtained the linearization results for a power range of the QAM signal. It can be indicated that the suppression of the IM3 products attained for the combined approach suggested in this paper is greater for around 25 dB to 15 dB in comparison with the results of the baseband approach from [13] for the frequency spacing between the QAM spectral components from 2 MHz to 20 MHz. The IM3 products reduction grade is around minimum 25 dB for frequency separation of 20 MHz when the combined approach is run. A general observation is that, as the input power increases and the frequency span becomes wider, the IM3 products drop rate decreases. The IM3products are lessened by 10 dBin the case of 7 dBm input power and 60 MHz frequency span when the combined method is applied, that is still much better result referring to the reduction of only a few decibels in case of the baseband method linearization. Moreover, it should be stressed that the results achieved by the combined method are also notably better in comparison with the IM3 products decrease represented in [14] wherein the fundamental carrier second harmonic modulated by the shaped baseband signal was utilized for the linearization. Better linearization results are obvious in the whole signal power range and frequency spacing between the spectral components, that is especially significant for a larger spacing: e.g. for input power 7 dBm and spacing 20 MHz, the IM3 products are hardly lowered by a few decibels in the linearization approach from [14], whereas in this paper the combined approach decreases the IM3 products by 26 dB.

The influence of the performed linearization approaches on the fifth-order intermodulation products, IM5, is presented in Fig. 4. The simulation shows that the IM5 products are lessened by minimum 10 dB for frequency interval between signals up to 20 MHz, while, by applying the modified baseband linearization signals, the IM5 products stayed unaltered in reference to the state before the linearization for almost all considered input power levels and frequency spacing between the QAM components. An exception is noted at input power of 3 dBm where the reduction of IM5 is 6 dB to 13 dB.



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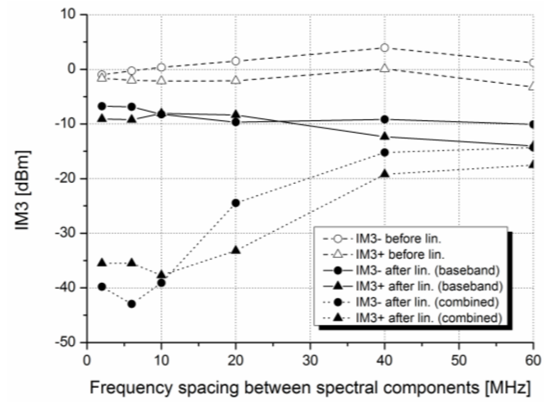

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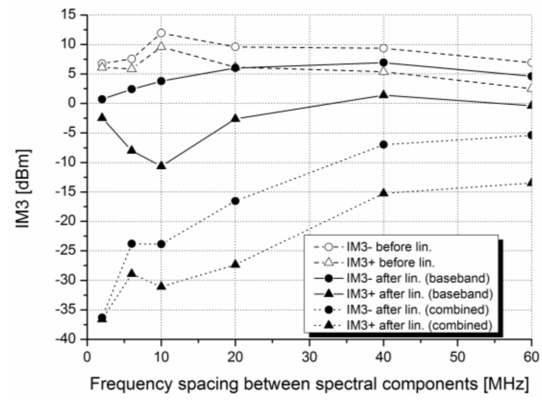
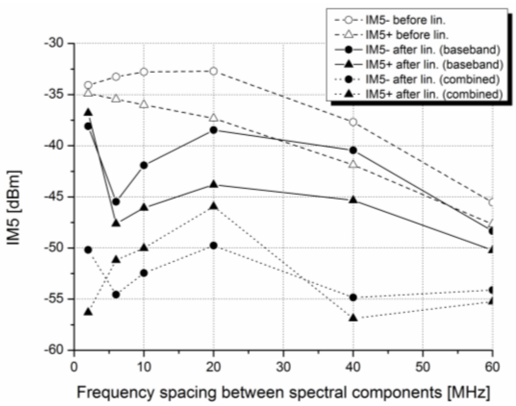


c)

Fig. 3 Third-order intermodulation products of the RF power amplifier for QAM signal before and after the linearization for different input power levels: a) 0 dBm, b) 3 dBm, c) 7 dBm



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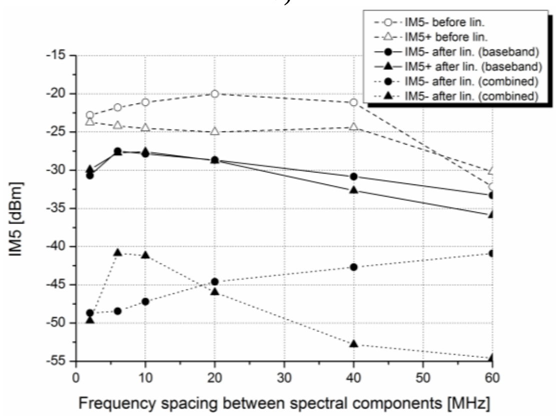

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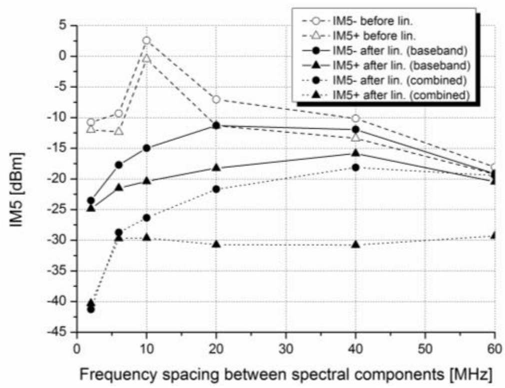

c)

Fig. 4 Fifth-order intermodulation products of the RF power amplifier for the QAM signalbefore and after the linearization for different input power levels: a) 0 dBm, b) 3 dBm, c) 7 dBm

By application of the combined linearization method, theIM5 products reduction grade significantly increases in relation to the previous method and depends on the power and frequency spacing between the QAM signal spectral components in the similar manner as the results for the IM3 products behave: the linearization results are significantly better than when only modified baseband signal is used for linearization. The reduction grade goes from 14 dB at 10 MHz signal spacing for the specified power range until 8 dB at 60 MHz spacing and 3 dBm input power level. At 7 dBm input power, the IM5 products descend by 27 dB at 10 MHz spacing, whereas they are retained at the level before the linearization at 60 MHz spectral component frequency spacing. In comparison with the results achieved in [14], where only modulated second harmonics carried out linearization, the better results concerning the IM5 products reduction are obtained by the combined method proposed herein. Namely, the results from [14] show that the IM5 products stayed unchanged or lowered for a few decibels in almost every analysed case (0 dBm to 7 dBm input signal power and 10 MHz to 20 MHz frequency spacing). It should indicate that the IM5 products are not lessened in reference to the level before the linearization at 20 MHz spectral component spacing, whereas we have accomplished the IM5 suppression even until frequency spacing of 60 MHz for 0 dBm and 3 dBm input power by the combined linearization approach.

Additionally, the influence of the suggested linearization methods was also investigated for the WCDMA signal which has 1 GHz centre frequency, a spectrum width of 3.84 MHz and peak to average power ratio (PAPR) of 6 dB in a range of fundamental signal average output power. The adequate values of the coefficients $a_{\{ib/ob\}}$, $a_{\{i2h/o2h\}}$ and $\theta_{\{i2h/o2h\}}$ for the required linearization results were determined by ADS optimization. It should be noticed that the linearization coefficients obtained for the WCDMA signal differ from the values achieved for the QAM signals. The linearization results obtained by the combined linearization approach are also compared with the results gained by the method that uses only baseband modified signal [13], as indicated in Fig. 5 and Fig. 6. The similar observation relating to the linearization results of the baseband linearization approach and the combined approach is imposed for the WCDMA signal as for the QAM signal previously considered. In the baseband linearization case, the adjacent channel power ratio-ACPR is enhanced around 10 dB at power levels greater than 24 dBm in the range of dominant third-order intermodulation products at ±4MHz offset from the carrier (Fig. 5), while in the range of dominant fifth-order intermodulation products at ±8 MHz offset from the carrier, the ACPR is restrained at the power levels before the linearization with the exception at the higher observed power levels where it is improved by a few decibels. Comparing to the results of the PA linearization gained in this paper by applying the combined linearization method, we may indicate that the improvement of ACPR observed at ±4MHz offset from the carrier is better for maximally 5 dB in relation to the baseband approach.

Additionally, the ACPR improvement is better by a few decibels in the range of dominant fifth-order intermodulation products when the combined method is utilized than in the first approach. Moreover, in reference [14], we analysed the ACPR of the WCDMA signal before and after the linearization by applying the modulated fundamental carrier second harmonic for only 11 dBm input signal power (output power of 29 dBm), where the ACPR was enhanced more than 10 dB at ±4MHz offset from the carrier. It is spotted from Fig. 5 that the combined method gives around 15 dB ACPR improvement at that output power level.

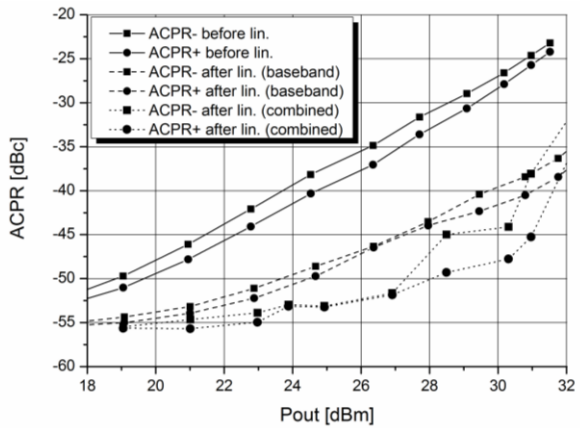

Fig. 5 ACPR before the linearization (solid line) and after the linearization (dashed and dotted lines) at ±4 MHz offset from the carrier (the range of the dominant third-order distortion) for the WCDMA digitally modulated signal in a terms of average output power

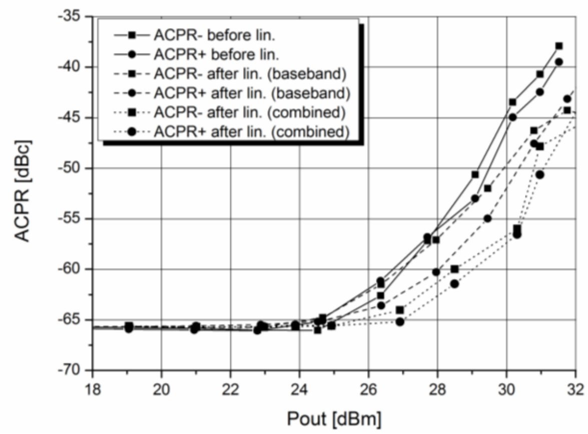

Fig. 6 ACPR before the linearization (solid line) and after the linearization (dashed and dotted lines) at ±8 MHz offset from the carrier (the range of the dominant fifth-order distortion) for the WCDMA digitally modulated signal in a terms of average output power

4. CONCLUSION

This paper presents a new linearization approach that combines variously modified baseband signals where one modulates the second harmonic of the fundamental carrier. The proposed linearization method uses the I and Q signals that are adequately processed in the digital domain with the aim to form the signals for the linearization which are inserted into the gate and drain of the RF power amplifier transistor. The analysis of the impact of the proposed combined linearization techniques on the intermodulation products suppression is assessed in simulation by ADS for the QAM signal whose I and Q components are sinusoidal signals and the spectrum contains two frequency components symmetrical around the carrier frequency. The linearization effects for different input power levels and different frequency spacing between the signal spectral components are examined for the proposed linearization method. Also, the obtained results are compared to the results achieved when the only modified baseband signals are fed at the amplifier circuit. It may be noted that the significantly better results are achieved in the reduction of the third-and fifth-order nonlinearity of the amplifier by the combined linearization method in comparison with the method that uses only the modified baseband linearization signal. The same may be inferred regarding the nonlinearity suppression by the combined method in reference to the results given in the literature that were reached by the method that performs linearization by the second harmonics of the fundamental carrier modulated by adequately modified baseband signals. Additionally, the linearization influence is also demonstrated for the WCDMA digitally modulated signal. The combined linearization method gives also greater improvement of the ACPR for the WCDMA digitally modulated signal in the range of the dominant third-order as well as the fifth-order distortions relative to the two mentioned linerization approaches.

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