RF PA Linearization Using Modified Baseband Signal that Modulates Carrier Second Harmonic

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Abstract – **A novel linearization technique for RF power amplifiers based on the modified baseband signal that modulates the fundamental carrier second harmonic is presented in this paper. Signals prepared for linearization are inserted at the input of the amplifier transistor (together with the fundamental useful signal) or injected simultaneously at the input and at the output of the amplifier transistor. In-phase and quadrature-phase components of the modified baseband signal are formed as the products of the second order nonlinearity of a nonlinear system fed by the useful baseband signal. The effects of the proposed linearization method are considered on a single stage power amplifier for quadrature amplitude modulated signals at different input power levels going up to 1dB below saturation region, as well as for WCDMA digitally modulated signal.**

Keywords – **linearization, amplifier, baseband signal, second harmonics.**

I. INTRODUCTION

With the increasing importance of spectral efficiency in mobile communications, the new generation of mobile communication technologies employs linear modulation (e.g. QPSK, QAM) for increasing bite rate and spectrum efficiency. Therefore, the power amplifiers have to process high rate non-constant envelope signals. For achieving high power efficiency, the power amplifier should operate around its compression region which distorts the linearity of the output signals. Different linearization methods of power amplifiers for minimizing the nonlinear distortions have been deployed [1-3]: feedback, feed-forward, predistortion, etc.

In this paper, we propose a new linearization method based on an appropriate digital processing of the baseband signal that modulates the carrier second harmonic. The modulated signal at the second harmonic is then injected at the input of the amplifier transistor together with the fundamental signal and also fed at the transistor output in order to reduce the intermodulation products. The modulated second harmonic and the fundamental signal are mixed due to the second order nonlinearity of the transistor generating the additional third-order nonlinear products that may suppress the original intermodulation products distorted by the transistor nonlinear characteristic. The effects of proposed linearization method are considered on a single stage power amplifier for the case when linearization signal is fed only at the amplifier transistor input and for the case when linearization signals are injected simultaneously at the input and output of the amplifier transistor. The linearization method is considered for QAM signals wherein I and Q components are single tones characterized with a range of frequencies up to 10MHz and different input power levels going up to 1dB below saturation region. Additionally, linearization is carried out for WCDMA digitally modulated signal.

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II. ANALYSIS

Theoretical analysis of the proposed linearization approach is based on the nonlinearity of transistor output current in amplifier circuit. The dominant nonlinearity of FETs can be represented by a Taylor-series polynomial model [4-6] as given by Eq. 1 in case when the memory effects are neglected.

$$\begin{aligned} i_{ds} &= g_{m1}v_{gs} + g_{m2}v_{gs}^2 + g_{m3}v_{gs}^3 + \\ &+ g_{d1}v_{ds} + g_{d2}v_{ds}^2 + g_{d3}v_{ds}^3 + \\ &+ g_{m1d1}v_{gs}v_{ds} + g_{m2d1}v_{gs}^2 v_{ds} + g_{m1d2}v_{gs}v_{ds}^2 + \ldots \end{aligned} \quad (1)$$

Transconductance terms, which represent the drain-source current (i_{ds}) dependence on the gate-source voltage (v_{gs}), are denoted by g_{mx}. The drain-source current dependence on drain-source voltage (v_{ds}) is expressed by the drain-conductance terms g_{dy}. The cross terms that relate drain-source current with the gate and drain voltages are expressed by g_{mxdy}. The order of each coefficient can be calculated as $x + y$.

The input gate voltage of digitally modulated signal is characterized by the magnitude $c(t)$, phase $\varphi(t)$, and carrier frequency ω_0, as given below:

$$\begin{aligned} v_{gs}(t) &= c(t)\cos(\omega_0 t + \varphi(t)) \\ &= c(t)\cos(\varphi(t))\cos(\omega_0 t) - c(t)\sin(\varphi(t))\sin(\omega_0 t) \\ &= v_s\,(I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \quad (2)$$

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

In order to express the in-phase and quadrature-phase components of a digitally modulated signal at second harmonic carrier frequency, the digital signal (Eq. 2) is fed at the input of the second degree nonlinear system with transfer function $v_{out} = v_{in}^2$. The output signal is obtained in the following form:

$$v_{out}(t) = \frac{1}{2}v_s^2(I^2+Q^2) + \frac{1}{2}v_s^2\left[(I^2-Q^2)\cos(2\omega_0) - 2IQ\sin(2\omega_0)\right] \quad (3)$$

In-phase component at second harmonic carrier has the form (I^2-Q^2) while the quadrature-phase component is presented by $2IQ$.

Fig. 1 shows the block diagram of the amplifier with the linearization circuit that processes the second harmonic modulated by the modified baseband signal and injects it into the amplifier transistor input (Fig. 1a) and simultaneously at the input and at the output of the amplifier transistor (Fig. 1b). The baseband signal transformation circuit generates the desired (I^2-Q^2) and $2IQ$ signals, which are separately adjusted in phase by $\theta_{\{i|o\}}$ and multiplied by $a_{\{i|o\}}$ for amplitude tuning. Indexes i and o in subscript are related to the signals prepared for the injection at the input and output of the amplifier transistor, respectively. In the case when linearization is performed by injection of the signal for linearization only at the input of amplifier transistor the coefficient a_o is equal to zero. The baseband signals modified in this way are injected at IQ modulators with central frequency $2\omega_0$ -fundamental carrier second harmonic. The created signal is then inserted at the input (together with the fundamental signal), (Eq. 4) and at the output of the amplifier transistor, Eq. 5:

$$v_{gs}(t) = v_s\left[I\cos(\omega_0) - Q\sin(\omega_0)\right] + a_i e^{-j\theta_i}\frac{1}{2}\left[(I^2-Q^2)\cos(2\omega_0) - 2IQ\sin(2\omega_0)\right] \quad (4)$$

$$v_{ds}(t) = v_o\left[I\cos(\omega_0) - Q\sin(\omega_0)\right] - a_o e^{-j\theta_o}\frac{1}{2}\left[(I^2-Q^2)\cos(2\omega_0) - 2IQ\sin(2\omega_0)\right] \quad (5)$$

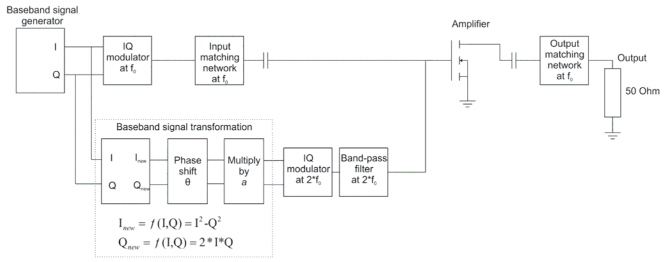
where v_o is the output signal at fundamental frequency.

By substitution of Eq. 4 and 5 in Eq. 1 the distorted output current at the fundamental frequency can be expressed by Eq. 6 which is truncated beyond the third degree nonlinearity.

$$\begin{aligned} i_{ds}(t) = [&v_s g_{m1} + \frac{3}{4}v_s^3 g_{m3}(I^2+Q^2) + \frac{1}{2}a_i e^{-j\theta_i} v_s g_{m2}(I^2+Q^2) \\ &- \frac{1}{4}a_o e^{-j\theta_o} v_s g_{m1d1}(I^2+Q^2) + \frac{1}{4}a_i e^{-j\theta_i} v_o g_{m1d1}(I^2+Q^2) \\ &+ \frac{3}{4}v_s v_o^2 g_{m1d2}(I^2+Q^2) + \frac{3}{4}v_s^2 v_o g_{m2d1}(I^2+Q^2)] \\ &(I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \quad (6)$$

The nonlinearity of drain–source current in terms of the voltage between drain and source, v_{ds}, which is expressed by the coefficients $g_{d1} - g_{d3}$, is assumed to have a negligible contribution to the intermodulation products according to [5-6], so that they are omitted from the equations.

The first term in Eq. 6 represents linearly amplified signal. The signal distorted by the cubic term of the amplifier (g_{m3}), which is considered as a dominant [5-6] in causing the third-order intermodulation products-IM3 and spectral regrowth, is included into the analysis by Eq. 6 as the second term. The mixing product of the fundamental signal and second harmonic injected at the transistor input is expressed as the third term.



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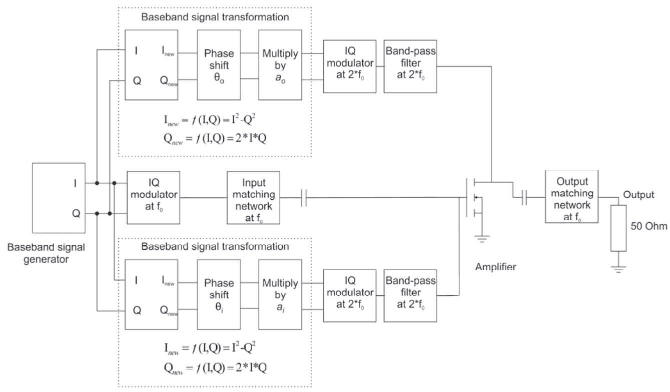


b)

Fig. 1. Schematic diagram of the amplifier linearization with the injection of the second harmonic modulated by the modified baseband signal a) at the input of the amplifier; b) at the input and output of the amplifier

Fig. 2. ADS timed source component - quadrature amplitude modulated carrier with single I and Q modulating tones

The fourth term, g_{m1d1}, exists due to the reaction between the gate-source voltage of the fundamental signal and voltage of the second harmonic fed at the amplifier transistor output. Additionally, the fundamental signal at the output of transistor mixed with the second harmonic driven at the amplifier transistor input generates a fifth term. The output signal at fundamental frequency v_o is considered to be 180 degree out of phase in reference to the input signal v_s. The mixed terms between drain and gate g_{m1d2} and g_{m2d1}, produce drain–source current at IM3 frequencies with the opposite phases (sixth and seventh terms in Eq. 6), so that they reduce each other [6]. According to the 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 phase of the modified baseband signal which modulates second harmonic injected at the input (a_i, θ_i) and output (a_o, θ_o) of the amplifier transistor.

III. PA DESIGN

Agilent Advanced Design System-ADS software was used for designing the broadband RF amplifier [7]. The amplifier was designed to operate over the frequency range 0.7GHz-1.1GHz on the bases of the MET model of the Freescale transistor MRF281S LDMOSFET. The source and load impedances $Z_s = (5.5 + j15)\Omega$ and $Z_l = (12.5 + j27.5)\Omega$, respectively, were obtained by load-pull and source-pull analysis in ADS. Transistor is biased to operate in class-AB, V_D=26V, V_G=5.1V (13.5% I_{DSS}).

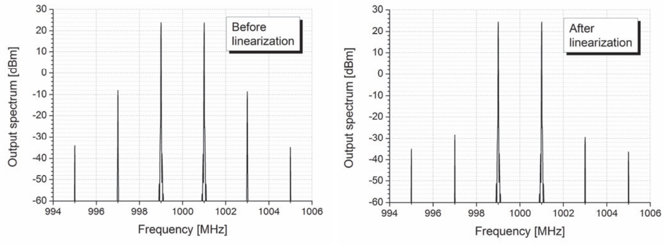
In order to design a broadband amplifier circuit, the input and output matching circuits of the transistor are based on the filter structures with lumped elements. Primarily, a lowpass prototype of filter with the order N=3 was designed.

The method of minimum reflection [8-9] was used for calculating the normalized admittances of the prototype elements. The values of the reactive elements of the lowpass filter were calculated by using the appropriate transformations [10]. Then, lowpass filter was transformed into a bandpass filter in a way that each series element was replaced by series resonant circuit, and each parallel element was replaced by a parallel resonant circuit at $\omega_0 = 1\text{GHz}$ [10].

The signal for linearization is delivered to the input and output of the amplifier transistor throughout the ideal band-pass filters characterized at 2GHz centre frequency and 0.5GHz bandwidth (Fig. 1).

IV. SIMULATED RESULTS

In order to evaluate the impact of the proposed linearization technique on the designed power amplifier, QAM tests were performed in ADS. Different QAM modulated signals for testing with a carrier frequency of 1GHz comprise the in-phase and quadrature-phase components at frequencies 1MHz, 3MHz, 5MHz and 10MHz.



a) b)

Fig. 3. Output spectra for 0dBm input power and 2MHz frequency spacing between spectral components of the QAM signal a) before and b) after linearization

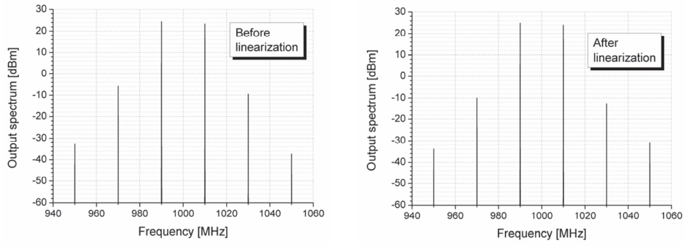


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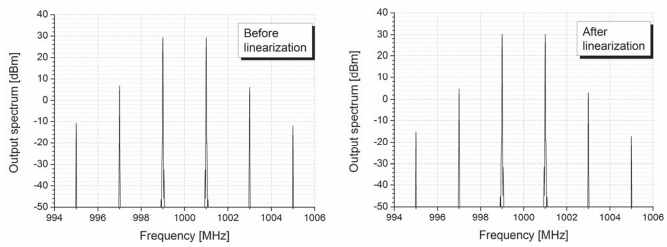
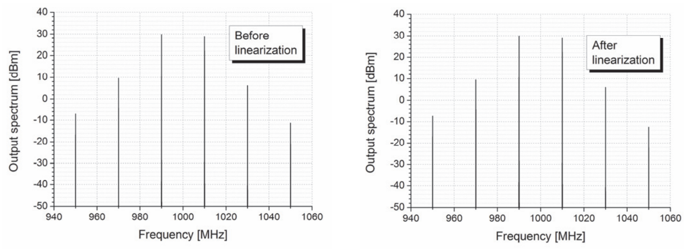
Fig. 4. Output spectra for 0dBm input power and 20MHz frequency spacing between spectral components of the QAM signal a) before and b) after linearization



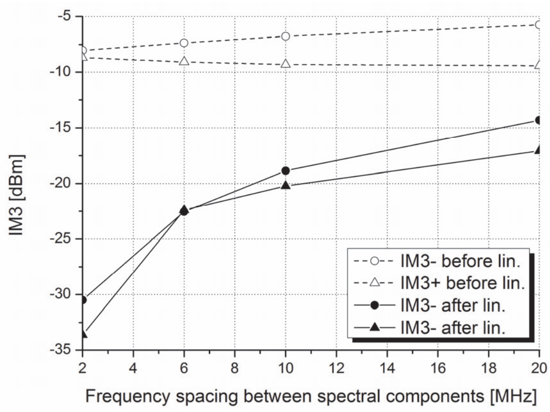
a) b)

Fig. 5. Output spectra for 7dBm input power and 2MHz frequency spacing between spectral components of the QAM signal a) before and b) after linearization



b) c)

Fig. 6. Output spectra for 7dBm input power and 20MHz frequency spacing between spectral components of the QAM signal a) before and b) after linearization



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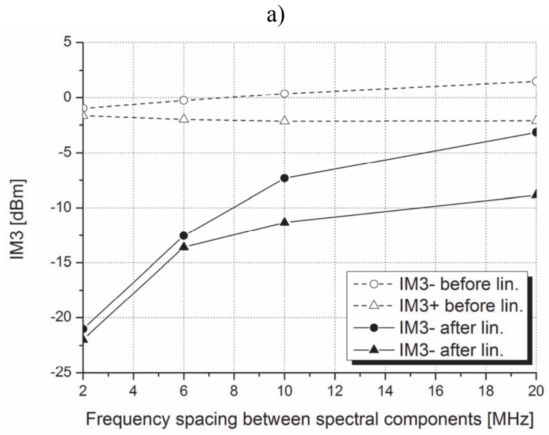

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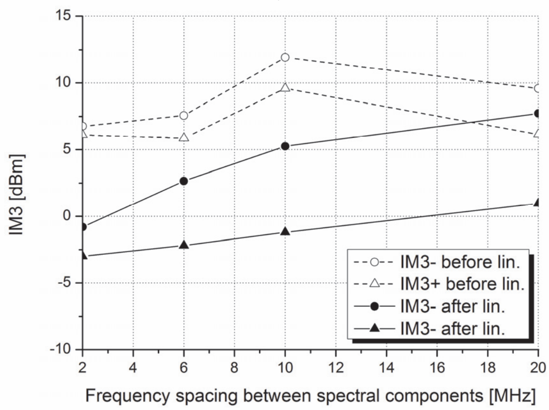
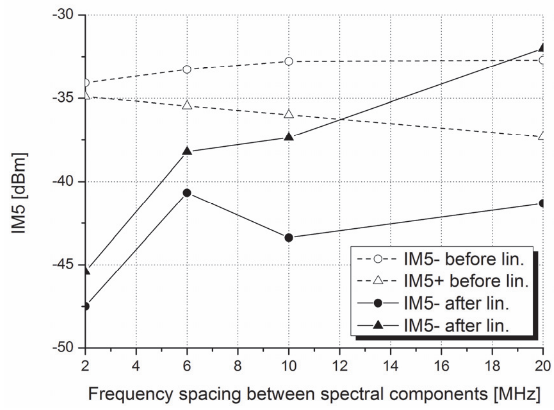


c)

Fig. 7. Third-order intermodulation products
(lower: IM3- & upper: IM3+) for input power levels:
a) P_{in}=0dBm, b) P_{in} =3dBm, c) P_{in} =7dBm



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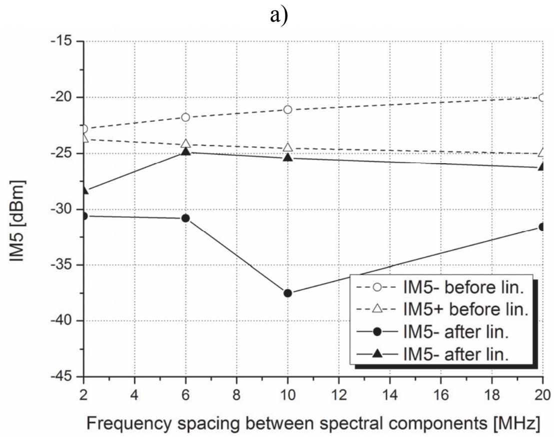

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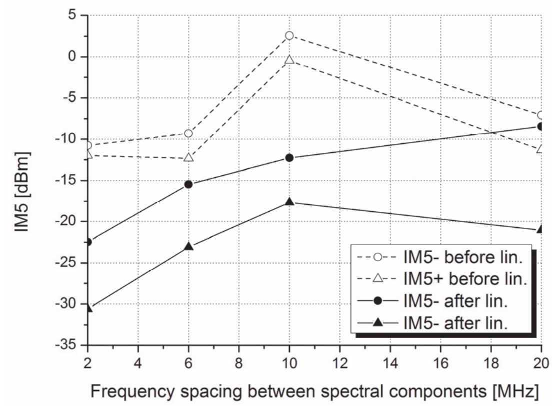


c)

Fig. 8. Fifth-order intermodulation products
(lower:IM5- & upper: IM5+) for input power levels:
a) P_{in}=0dBm, b) P_{in}=3dBm, c) P_{in} =7dBm

The spectrum of these QAM signals contains two frequency components separated by 2MHz, 6MHz, 10MHz and 20MHz at the above mentioned centre frequency. ADS timed source component named QAM was used as a source of signals (Fig. 2). The analysis was carried out for different input signal power levels. The output spectra before and after the linearization when linearization signal are fed only at the amplifier transistor input are represented in Fig. 3 for 0dBm input power and 2MHz span between spectral components. The improvement of IM3 products is 20dB. For the case when spacing between spectral components is 20MHz suppression of the IM3 products is only 4dB (Fig. 4). When the input power increases, the reduction of the IM3 products could not be attained even for narrower spacing between spectral components, so that the IM3 products are suppressed only 2dB for 7dBm input power and 2MHz spacing between spectral components as shown in Fig. 5. Figure 6 represents output spectra before and after linearization for 7dBm input power and 20MHz spacing between signals and indicates that the IM3 products cannot be reduced in case of wider spacing between spectral components and growing power of input signal if the linearization signal is injected at the amplifier transistor input. Even though the theoretical analysis of the linearization technique does not take into account the IM5 products, it follows from the Figs. 3 to 5 that the IM5 products retain at the level before linearization or deteriorate for a few dB with the exception of the case of 7dBm input power and 2MHz frequency spacing where they are lowered for approximately 5dB.

Power level of the third-order intermodulation products of QAM signal, before and after the linearization, when the linearization signals are injected simultaneously at the input and output of the amplifier transistor is presented in Fig. 7. The IM3 product power is drawn in terms of the frequency interval between the spectral components of QAM signal for fundamental signal power levels at the amplifier input 0dBm, 3dBm and 7dBm.

It can be noted that the IM3 products are lessened in a considered power range by applying the linearization. The suppression of the IM3 products is high, more than 20dB, for lower power (0dBm and 3dBm) and narrower frequency spacing between the QAM spectral components (2MHz). In case when the frequency interval between signals is 6MHz, the IM3 products are reduced by 15dB and 12dB for considered power levels of the fundamental signals. As the power increases and the frequency span becomes larger, the IM3 products drop rate falls down, so that it is only a few dB in the case of 7dBm input power and 20MHz frequency interval between signals.

The influence of the performed linearization technique on the fifth-order intermodulation products, IM5, is presented in Fig. 8. The simulation shows that they are stayed unchanged or lowered for a few decibels in almost every analysed case.

The output spectra of QAM signal before and after the linearization are depicted in Fig. 9 for 3dBm input power and 2MHz frequency spacing between spectral components. We may observe that IM3 products descend for 20dB, whereas the IM5 products go down by 5dB after the linearization. Additionally, the fundamental signal power increases for 1.1dB. Accordingly, another advantage of the applied linearization method is the augmentation of the output power of the fundamental signals by decibel or a few tenths of a dB after the linearization, as shown in Fig. 10.

Moreover, the proposed linearization approach was verified for a WCDMA signal at 1 GHz centre frequency, a spectrum width of 3.84MHz in case of 5dBm and 11dBm input power level of fundamental signal. The results of the analysis are shown in Fig. 11.

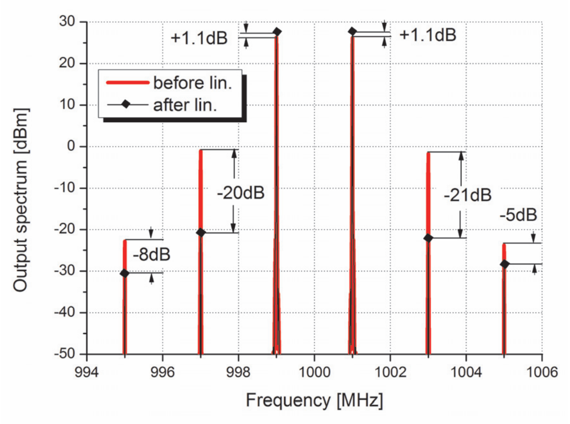


Fig. 9. Output spectra before and after linearization for 3dBm input power and 2MHz frequency spacing between spectral components of the QAM signal

Adjacent channel power ratio-ACPR is enhanced about 7dB for 5dBm input power and more than 10dB for 11dBm input power (saturation point) observed at ±4MHz offset from the carrier.

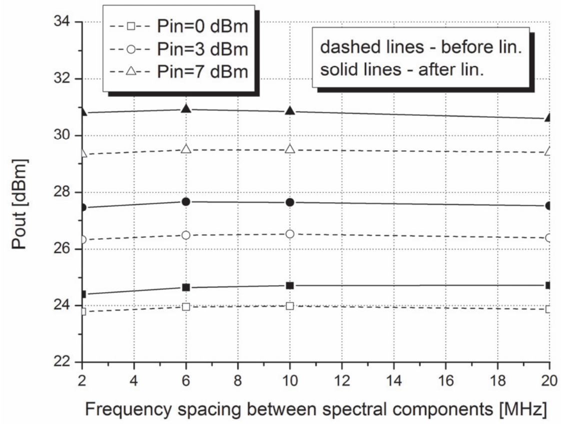
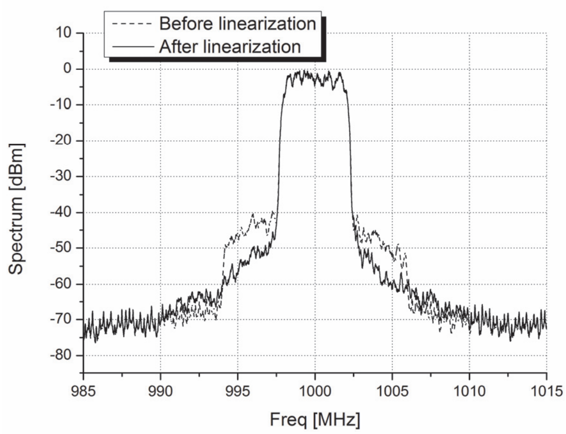
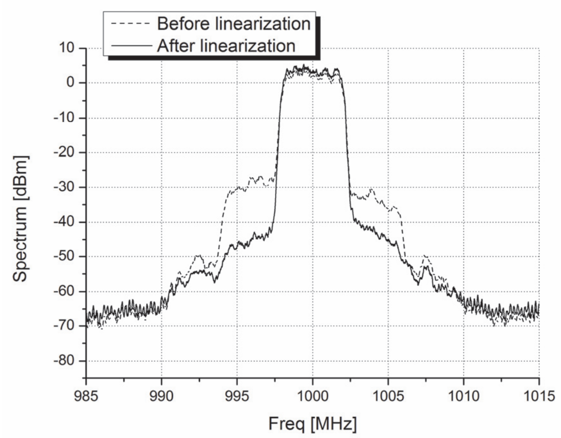


Fig. 10. Output power of fundamental signals before and after the linearization



a)



b)

Fig. 11. Output spectra for WCDMA digitally modulated signal before and after the linearization for a) 5dBm; b) 11dBm input power

V. Conclusion

This paper presents the new linearization technique, which uses the modified baseband signal that modulates the second harmonic of the carrier. The analysis of the impact of the proposed linearization technique on suppression of the intermodulation products is assessed for QAM signal simulation test in ADS. Two spectrum components of QAM signal shifted in frequency by ±1MHz, ±3MHz, ±5MHz and ±10MHz in reference to the carrier frequency 1GHz are simultaneously driven at the amplifier for different power levels of fundamental signal up to 1dB below saturation region. The linearization effect is demonstrated when signal for linearization, i.e. the second harmonic modulated by the modified baseband signal, is inserted only at the amplifier transistor input but for lower power and small frequency interval between spectrum components. However, very satisfactory reduction of the third-order nonlinearity of the amplifier is achieved even for higher power levels in case when signals for linearization are fed at the amplifier transistor input and output simultaneously. Moreover, the technique neither shows a significant downtrend of the IM5 products nor deteriorates them. Additionally, the proposed linearization method shows also an improvement of ACPR for WCDMA digitally modulated signal. However, it can be noticed that the suppression rate of the intermodulation products drops when the power levels and interval between QAM signal spectral components grow up.

ACKNOWLEDGEMENT

This is an extended version of the paper "A Novel Linearization Technique Based on Modified Baseband Signal that Modulates Carrier Second Harmonic" presented at the 11th International Conference on Telecommunications in Modern Satellite, Cable and Broadcasting Services - TELSIKS 2013, held in October 2013 in Niš, Serbia. This work was supported by the Ministry of Education, Science and Technological development of Republic of Serbia, the projects number TR-32052.

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