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Original scientific paper

DOHERTY AMPLIFIER LINEARIZATION BY DIGITAL INJECTION METHODS*

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Abstract. *Verification of two linearization methods, applied on asymmetrical two-way microstrip Doherty amplifier in experiment and on symmetrical two-way Doherty amplifier in simulation, is performed in this paper. The laboratory set-ups are formed to generate the baseband nonlinear linearization signals of the second-order. After being tuned in magnitude and phase in the digital domain the linearization signals modulate the second harmonics of fundamental carrier. In the first method, adequately processed signals are then inserted at the input and output of the main Doherty amplifier transistor, whereas in the second method, they are injected at the outputs of the Doherty main and auxiliary amplifier transistors. The experimental results are obtained for 64QAM digitally modulated signals. As a proof of concept, the linearization methods are also verified in simulation, for Doherty amplifier designed to work in 5G band below 6 GHz, utilizing 20 MHz LTE signal.*

Key words: *Doherty amplifier, baseband signal, second harmonic, linearization, experimental verification.*

1. INTRODUCTION

In modern wireless communications, the efficiency of RF transmitters largely depends on the efficiency of power amplifiers (PA), so the development of 5G/6G systems requires new PA architectures that will ensure that amplifiers hold high efficiency while maintaining good linearity. Therefore, it is necessary to find a compromise between the key parameters of the PA, such as efficiency, power output and linearity. With the classic architecture of the PA, this is not easy to be achieved, and it is very difficult to optimize all the key parameters of PA. Usually the optimal design of the PA for one parameter leads to the degradation of another important parameter; therefore, the solution of this problem is to design an energy efficient PA, which is then to be linearized by one of the appropriate linearization techniques. The PA characterized by high efficiency is Doherty

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topology (DA), which is widely used in the contemporary wireless communication systems. In recent time, different linearization techniques [2] are used for PA nonlinearity compensation, such as feedback linearization [3-4], feedforward linearization [5-6], digital predistortion [7-9] and digital injection methods [10-11].

We deployed in earlier work the digital linearization technique [10], [12-14] which processes the *I* and *Q* signals to generate the adequate 2nd order baseband linearization signals adjusted in the magnitudes and phase angles. These signals are then driven at the gate and drain of the amplifier transistor, after modulating the 2nd harmonic of the fundamental carrier, in order to lower the nonlinearity of the single stage PA [10], [12], and the two-way DA [10], [13]. In [14], DA was linearized by inserting the modulated signals for linearization at the outputs of the main and auxiliary amplifier transistors. The comparison of two digital linearization methods was carried out in simulation on the designed broadband microstrip DA, for different two-tone signal power and maximum tone separation of 30 MHz, as well as for OFDM signal.

In this paper, the various experiments are performed on Doherty amplifier fabricated in microstrip technology [15] for evaluation of two linearization methods. The tests were realized for 64QAM signal with useful spectrum bandwidth of 2 MHz. Measured results show the adjacent channel power ration -ACPR at dominant third-order intermodulation products and fifth-order intermodulation products. To confirm the efficiency of the linearization method, in this paper the verification also was performed in the simulation procedure for DA designed to operate at frequency 3.5 GHz [16]. Simulation was performed for a 20 MHz LTE signal and various output power levels up to 1-dB compression point.

2. ANALYSIS

The applied linearization methods can be explained theoretically by modeling the nonlinearity of the amplifier transistor by using a Taylor-series polynomial model, which does not include memory effect. The FET output current (i_{ds}) in terms of the gate-source voltage (v_{gs}), and drain-source voltage (v_{ds}) is given by Eq. 1, [10], [12], [15].

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

where g_{mx} represents transconductance terms, g_{dy} is the drain-conductance terms and g_{mxdy} is mixed terms (the order of each coefficient can be calculated as $x+y$), and *m*/*a* relates to the main and auxiliary amplifiers in Doherty circuit. The nonlinear terms defined by the coefficients $g_{d1} - g_{d3}$ can be neglected according to the previous performed analysis. Also, the mixing terms of the 3rd order g_{m1d2} and g_{m2d1} produce the 3rd order intermodulation products (IM3) that can be considered to reduce each other to some extent, so that they are omitted from the final equations that relate to the IM3 of DA output current given in text below, based on the results obtained in [10], [12], [15]. However, those mixing terms are included into the equations that describe the 5th order intermodulation products of DA output current (IM5), so that the influence of the injected 2nd order linearization signals to the IM5 can be explained.

The basband signals for linearization are formed by the adeqate processing of the in-phase I and quadrature-phase Q components of the digital signal resulting in the in-phase linearization component – $I_{IM2} = (I_2 - Q_2)$, and quadrature-phase component $-Q_{IM2} = 2IQ$, which are the products of the 2nd order nonlinearity. Those signals are tuned in magnitude by $a^{m/a}_{\{i|o\}}$ and phase by $\theta^{m/a}_{\{i|o\}}$, where adaptation coefficients are denoted by i and o in subscript for the injection of the signals for the linearization at the input and output of the transistor in amplifier. The baseband signals prepared in this manner then modulate fundamental carrier second harmonic. In the first linearization approach applied in this paper for the Doherty amplifier linearization, the signal for linearization are inserted at the input (together with the fundamental signal) as given by Eq. 2 and at the output of the main amplifier transistor in DA, Eq. 3, whereas in the second approach the signals for linearization are led to the transistor output of the main, Eq. 3, and auxiliary stages, Eq. 4, in DA.

$$\begin{aligned} v^m_{gs} &= v^m_s[I\cos(\omega_0 t) - Q\sin(\omega_0 t)] + \\ &+ a^m_i e^{-j\theta^m_i}\frac{1}{2}[(I^2 - Q^2)\cos(2\omega_0) - 2IQ\sin(2\omega_0)] \end{aligned} \tag{2}$$

$$\begin{aligned} v^m_{ds} &= v^m_o[I\cos(\omega_0 t) - Q\sin(\omega_0 t)] + \\ &- a^m_o e^{-j\theta^m_o}\frac{1}{2}[(I^2 - Q^2)\cos(2\omega_0) - 2IQ\sin(2\omega_0)] \end{aligned} \tag{3}$$

$$\begin{aligned} v^a_{ds} &= v^a_o[I\cos(\omega_0 t) - Q\sin(\omega_0 t)] + \\ &- a^a_o e^{-j\theta^a_o}\frac{1}{2}[(I^2 - Q^2)\cos(2\omega_0) - 2IQ\sin(2\omega_0)] \end{aligned} \tag{4}$$

where v^m_s, and v^m_o, are the magnitudes of the input and output signal of the main amplifier transistor at fundamental frequency, and v^a_o is the magnitude of the output signal of the auxiliary amplifier transistor at fundamental frequency.

The distorted output current of the Doherty amplifier analysed for the IM3 products is expressed by Eq. 5 when the first linearization method is applied and by Eq. 6 for the second linearization method. The IM5 products are included into Eqs. 7 and 8 for both linearization methods.

$$\begin{aligned} i_{out}\Big|^{1st}_{IM3} = \Bigg[&\frac{3}{4}(v^m_s)^3 g_{m3} + \frac{3}{4}(v^a_s)^3 g_{m3} + \frac{1}{2}a^m_i e^{-j\theta^m_i} v^m_s g_{m2} \\ &- \frac{1}{4}a^m_o e^{-j\theta^m_o} v^m_s g_{m1d1} + \frac{1}{4}a^m_i e^{-j\theta^m_i} v^m_o g_{m1d1}\Bigg](I^2 + Q^2)(I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \tag{5}$$

$$\begin{aligned} i_{out}\Big|^{2nd}_{IM3} = \Bigg[&\frac{3}{4}(v^m_s)^3 g_{m3} + \frac{3}{4}(v^a_s)^3 g_{m3} + \\ &- \frac{1}{4}a^m_o e^{-j\theta^m_o} v^m_s g_{m1d1} - \frac{1}{4}a^a_o e^{-j\theta^a_o} v^a_s g_{m1d1}\Bigg](I^2 + Q^2)(I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \tag{6}$$

$$\begin{aligned} i_{out}\Big|_{IM5}^{1st} = \Bigg[& \frac{5}{8}(v_s^m)^5 g_{m5} + \frac{5}{8}(v_s^a)^5 g_{m5} + \frac{3}{2}(a_i^m)^2 e^{-j2\theta_i^m} v_s^m g_{m3} + \\ & + \frac{1}{2}(a_o^m)^2 e^{-j2\theta_o^m} v_s^m g_{m1d2} - a_i^m a_o^m e^{-j(\theta_i^m+\theta_o^m)} v_o^m g_{m1d2} + \\ & + \frac{1}{2}(a_i^m)^2 e^{-j2\theta_i^m} v_o^m g_{m2d1} - a_i^m a_o^m e^{-j(\theta_i^m+\theta_o^m)} v_s^m g_{m2d1} \Bigg] (I^2+Q^2)^2 (I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \tag{7}$$

$$\begin{aligned} i_{out}\Big|_{IM5}^{2nd} = \Bigg[& \frac{5}{8}(v_s^m)^5 g_{m5} + \frac{5}{8}(v_s^a)^5 g_{m5} + \\ & + \frac{1}{2}(a_o^m)^2 e^{-j2\theta_o^m} v_s^m g_{m2d1} + \frac{1}{2}(a_o^a)^2 e^{-j2\theta_o^a} v_s^a g_{m2d1} \Bigg] (I^2+Q^2)^2 (I\cos(\omega_0 t) - Q\sin(\omega_0 t)) \end{aligned} \tag{8}$$

The 1st and 2nd terms in Eqs. 5 and 6 represent DA linearity degradation by the 3rd order nonlinearity of the amplifier stages. The 3rd to 5th terms in Eq. 5 are the nonlinear products of the second order between the linearization signal injected at the input and at the output of the main amplifier transistor in DA and fundamental signals. The 3rd and 4th terms in Eq. 6 relate to the mixing terms of the 2nd order between the fundamental signals and the signals for the linearization put at the output of the main and auxiliary amplifier in DA. It can be observed that the nonlinear terms of the 2nd order generated due to the injection of the signals for the linearization can reduce the originally produced IM3 distortion by the adequate adjustment of the magnitude and phase of the signals for the linearization, [10], [12], [15]. Equations 7 and 8 define the IM5 products of the DA output current generated by the 5th order nonlinearity of DA main and auxiliary stages by the 1st and 2nd terms. Additional terms for two linearization methods are the 3rd order products that mix the linearization signal with the fundamental signals, which can reduce original IM5 products if their magnitudes and phases are related appropriately, [10], [12], [15].

3. DA DESIGN

Two Doherty amplifiers were designed to verify proposed linearization method: two-way asymmetrical Doherty amplifier operating at 900 MHz central frequency [15] and symmetrical Doherty amplifier operating at 3.5 GHz central frequency [16].

The linearization effects were examined in experiment on the fabricated two-way asymmetrical Doherty amplifier shown in Figure 1, which consist of: 1. main amplifier and frequency diplexers; 2. auxiliary amplifier and frequency diplexer; 3. offset line and output combining networks; 4. Pi attenuator; 5. power combiner for injection of the signal for linearization at auxiliary amplifier output, 6. port for the injection of the linearization signal at the main transistor input, 7. port for the injection of the linearization signal at the main transistor output. Detailed description of the Doherty amplifier design can be found in [15], (it should point out that Wilkinson power combiner denoted as 5. in Figure 1 was used for another purpose in the linearization method exploited in [15]). In this paper, one port of the combiner was utilized for the linearization. The maximal transducer gain 9 dB was measured for the fabricated two-way asymmetrical Doherty amplifier for the main amplifier biased in class-AB (V_D = 5 V, V_{GM} =−3 V), and the auxiliary amplifier operating in class-C regime (V_D = 5 V, V_{GA} =−5 V) when AP602A-2 GaAs MESFET transistor was used in amplifying cells. Moreover, measured 1-dB compression point of DA is at 15 dBm output power and 18 dBm maximum output power is achieved.

Symmetrical two-way Doherty amplifier that operates at 5G band below 6 GHz was designed according to the instructions given in [16]. The DA PA was designed by using CGH40010F GAN HEMT transistor. The drain voltage is 28 V, whereas the gate voltage of the main and auxiliary amplifier is -2.8 V, and -5.7 V, respectively. Main characteristics that relate to Gain, 1-dB compression point, power added efficiency PAE, DC power consumption etc. were represented in Table 1 for frequency 3.5 GHz. Additionally, the Gain, Gain compression, PAE and Supply current are shown in Figure 2 in the range of DA output power.

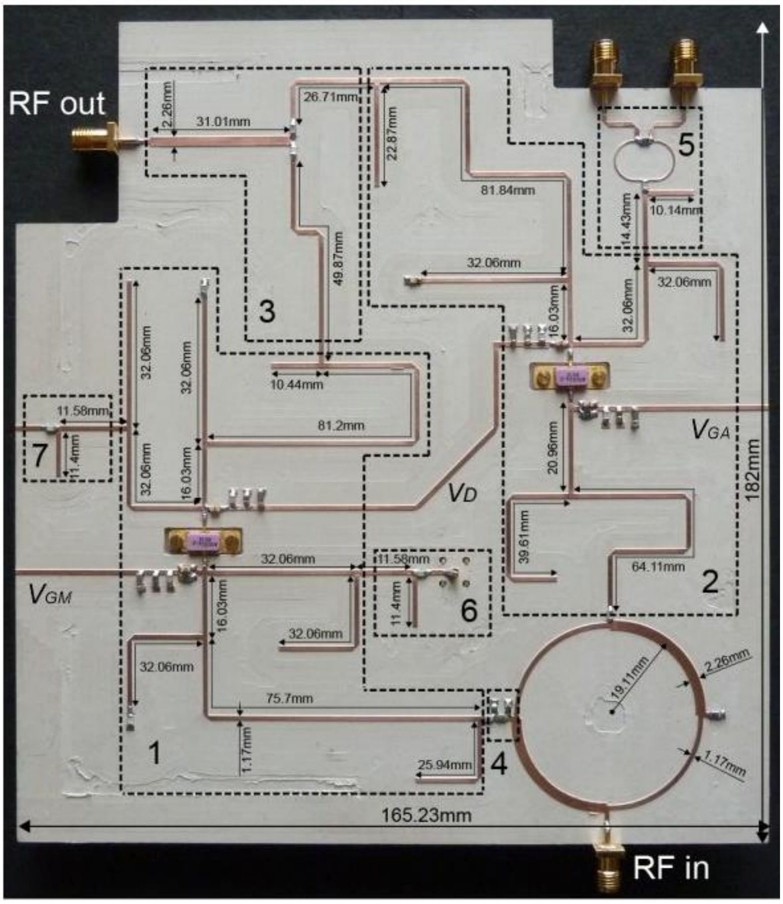
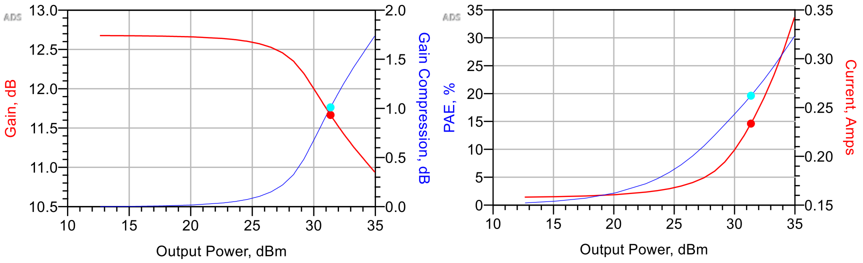


Fig. 1 Asymmetrical two-way Doherty amplifier (all dimensions are in millimeters)

Table 1 Characterization of two-way symmetrical DA 3.5 GHz

Gain Compression (dB)	Output Power (dBm)	Transducer Power Gain (dB)	Power- Added Efficiency, %	DC Power Consumpt. Watts	High Supply Current	Thermal Dissipation Watts
1.010	31.357	11.667	19.622	6.537	0.233	5.253



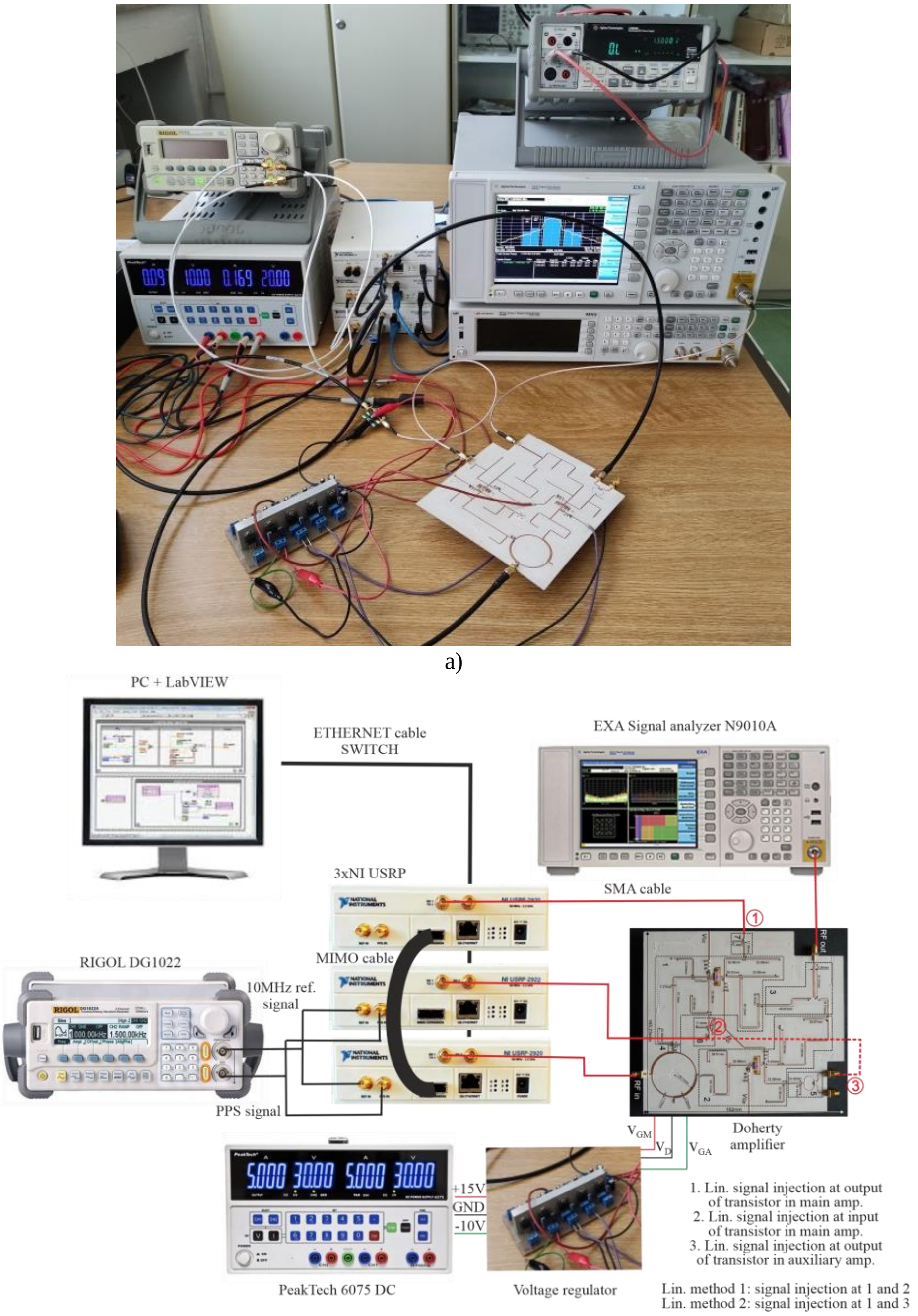
a) b)

Fig. 2 Symmetrical two-way Doherty amplifier characteristics: a) Gain and Gain compression; b) PAE and supply current

4. MEASUREMENT SET-UP

The measurement set-up shown in Figure 3 was established to verify in experiments the linearization methods developed by our researcher group [17]. The linearization methods of power amplifiers are based on the 2nd order baseband nonlinear digital signals, which adequately modified and processed in the baseband, modulate the fundamental carrier second harmonic.

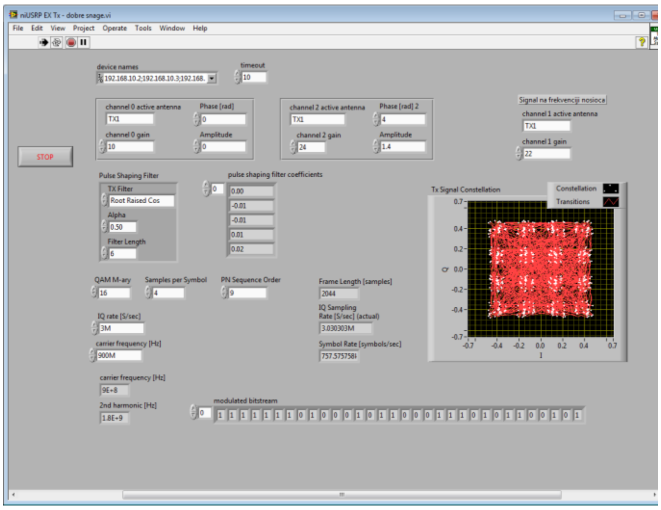
The measurement system was designed to enable verification of the linearization methods, which generally use two linearization signals, which after the digital processing in the baseband modulate the second harmonic of the fundamental signal. The measuring system can generate three independent, synchronized signals – the fundamental signal and two linearization signals at the frequency of fundamental signal second harmonic. Two NI USRP 2920 models and one NI USRP 2922 model were used for the measurement system, which are connected to the computer via an Ethernet switch. The LabVIEW environment was used for implementation of the interface for management and control of the NI USRP devices (Figure 4). A unique challenge during the implementation of the measurement system was the synchronization of the NI USRP devices. The synchronization of the NI USRP devices was performed by a MIMO cable which was used to synchronize two of the three USRP devices (using MIMO EXPANSION input), while the third device was synchronized with the previous two via external referent 10 MHz and 1 PPS signals. RIGOL DG1022 two-channel function/arbitrary waveform generator was used as the generator of these signals. The outputs from the function generator were distributed to the corresponding inputs of the NI USRP device (REF IN and PPS IN).



a)

b)

Fig. 3 Experimental verification of linearization methods 1 and 2: a) measurement set-up; b) schematic diagram



device names: 192.168.10.2;192.168.10.3;192.168.
timeout: 10
channel 0 active antenna: TX1
Phase [rad]: 0
channel 0 gain: 10
Amplitude: 0
channel 2 active antenna: TX1
Phase [rad] 2: 4
channel 2 gain: 24
Amplitude: 1.4
Signal na frekvenciji nosioca
channel 1 active antenna: TX1
channel 1 gain: 22
Pulse Shaping Filter
TX Filter: Root Raised Cos
Alpha: 0.50
Filter Length: 6
pulse shaping filter coefficients: 0 | 0.00, -0.01, -0.01, 0.01, 0.02
QAM M-ary: 16
Samples per Symbol: 4
PN Sequence Order: 9
Frame Length [samples]: 2044
IQ rate [S/sec]: 3M
IQ Sampling Rate [S/sec] (actual): 3.030303M
carrier frequency [Hz]: 900M
Symbol Rate [symbols/sec]: 757.575758k
carrier frequency [Hz]: 9E+8
2nd harmonic [Hz]: 1.8E+9
modulated bitstream: 0 | 1 1 1 1 1 1 1 0 1 0 0 0 1 0 1 1 0 0 0 1 1 1 0 1 0 1 1 0 0 1 0 1

Fig. 4 Interface for management and control of the NI USRP devices

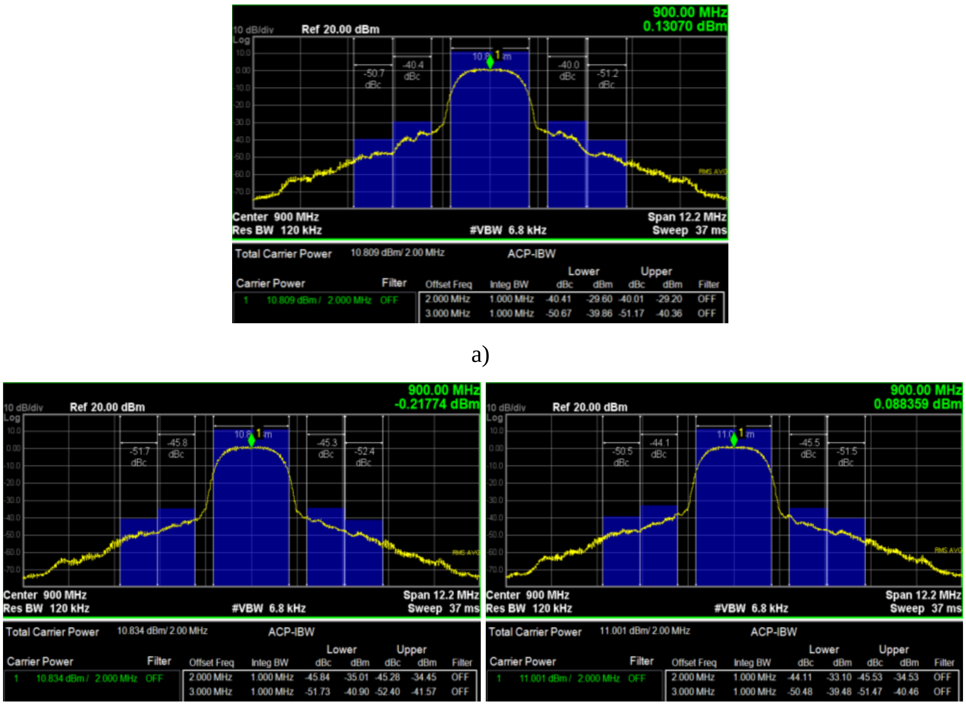
When using the NI USRP with the LabVIEW environment, it is necessary to provide high processing power, large amount and high speed of memory as well as fast Ethernet connection to the computer on which the NI USRPs are connected, which is especially required if multiple USRP devices are used at the same time. The lack of any of these resources can significantly affect the reliable operation of the NI USRP and lead to frequent downtime [17].

To demonstrate the results of linearization, useful 64QAM signal, the signals for linearization and their control in magnitude and phase were performed by the USRP platforms. The linearization effects were examined on the fabricated two-way asymmetrical Doherty amplifier operating at 900 MHz central frequency, shown in Figure 1.

The measurements of output spectra, the adjacent channel power ratios-ACPRs, for the states before and after the linearization carried out for 64QAM modulation format and different signal power levels were spotted in EXA Signal analyzer N9010A.

5. RESULTS OF LINEARIZATION

Asymmetrical two-way Doherty amplifier was tested for 64QAM signal with 2 MHz useful channel bandwidth. Central frequency of operation is 900 MHz. The linearization effects were measured on the fabricated DA for different input signal power levels 1 dBm to 5 dBm. The presented results shown in Figures 5 to 7 compare the ACPRs obtained without and with applying two digital linearization methods: 1) the first-standard method that injects signals for the linearization at the gate and drain of the transistor in the main cell of the DA and 2) the second-modified method, where the linearization signals are put at the drain of the main and auxiliary amplifier transistors in the DA.



a)

Total Carrier Power 10.809 dBm/ 2.00 MHz ACP-IBW

	Carrier Power	Filter
1	10.809 dBm / 2.000 MHz	OFF

		Lower		Upper		
Offset Freq	Integ BW	dBc	dBm	dBc	dBm	Filter
2.000 MHz	1.000 MHz	-40.41	-29.60	-40.01	-29.20	OFF
3.000 MHz	1.000 MHz	-50.67	-39.86	-51.17	-40.36	OFF

b)

Total Carrier Power 10.834 dBm/ 2.00 MHz ACP-IBW

	Carrier Power	Filter
1	10.834 dBm / 2.000 MHz	OFF

		Lower		Upper		
Offset Freq	Integ BW	dBc	dBm	dBc	dBm	Filter
2.000 MHz	1.000 MHz	-45.84	-35.01	-45.28	-34.45	OFF
3.000 MHz	1.000 MHz	-51.73	-40.90	-52.40	-41.57	OFF

c)

Total Carrier Power 11.001 dBm/ 2.00 MHz ACP-IBW

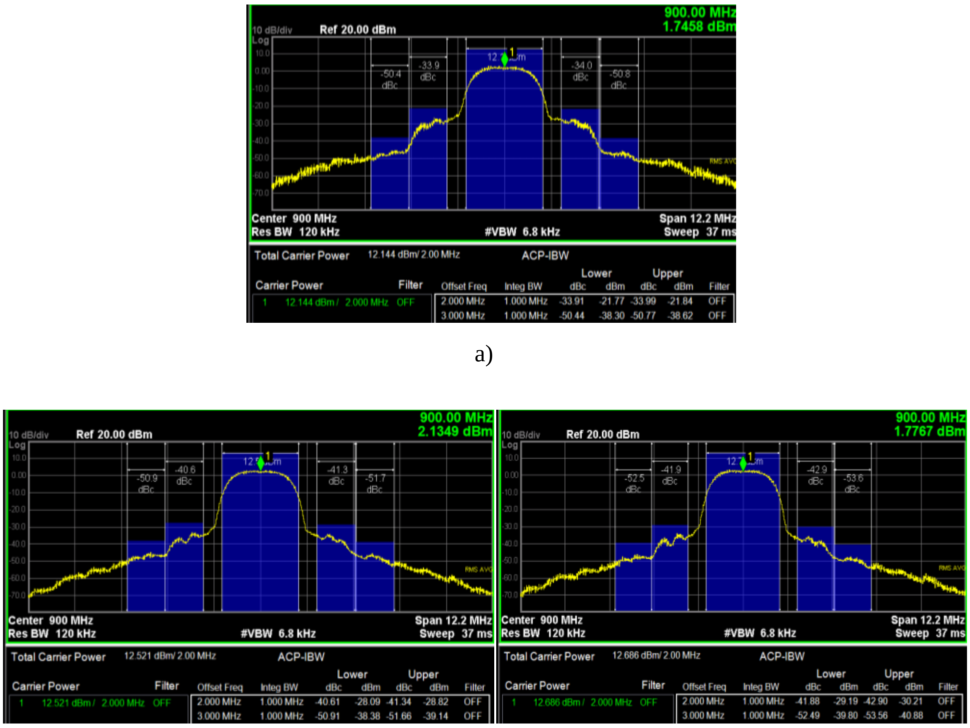
	Carrier Power	Filter
1	11.001 dBm / 2.000 MHz	OFF

		Lower		Upper		
Offset Freq	Integ BW	dBc	dBm	dBc	dBm	Filter
2.000 MHz	1.000 MHz	-44.11	-33.10	-45.53	-34.53	OFF
3.000 MHz	1.000 MHz	-50.48	-39.48	-51.47	-40.46	OFF

Fig. 5 Output spectrum for 64QAM signal of 2 MHz useful signal frequency bandwidth for input signal power 1 dBm: a) before linearization; b) after linearization by 1st method; c) after linearization by 2nd method

The results of ACPRs are illustrated in the lower and upper adjacent channels (at ±2 MHz offset from carrier where IM3 products are dominant) and in the alternate channels (at ±3 MHz offset from carrier where IM5 products are dominant). We can observe for 1 dBm input power, that the ACPR in the adjacent channels is improved by 4 dB for both linearization methods, whereas for 3 dBm input power they become better by 6 dB for the 1st method and 8 dB for the 2nd method. With the power increase to 5 dBm, ACPRs decreases by 3 dB and 5 dB in the 1st and 2nd methods, respectively. No evident improvement in the alternate channels can be noticed for 1 dBm and 3 dBm input power levels, but it is 4 dB in case of 5 dBm power.

Comparing the measured results with the simulated results represented in [14], we can infer that the 2nd linearization method achieves slightly better ACPRs improvement in the adjacent channels, especially for higher power, as it was also deduced in [14] when simulated results were analyzed. Even though the simulated results attained for the two-tone test show more apparent improvement when the 2nd method is used, it should indicate that for the OFDM signal test in simulation, the less divergence between results accomplished with two linearization methods can be observed for higher power, closer to amplifier saturation region.



Center 900 MHz
Res BW 120 kHz
#VBW 6.8 kHz
Span 12.2 MHz
Sweep 37 ms

Total Carrier Power 12.144 dBm/ 2.00 MHz ACP-IBW

Carrier Power	Filter
1 12.144 dBm / 2.000 MHz	OFF

		Lower		Upper		
Offset Freq	Integ BW	dBc	dBm	dBc	dBm	Filter
2.000 MHz	1.000 MHz	-33.91	-21.77	-33.99	-21.84	OFF
3.000 MHz	1.000 MHz	-50.44	-38.30	-50.77	-38.62	OFF

a)

Center 900 MHz
Res BW 120 kHz
#VBW 6.8 kHz
Span 12.2 MHz
Sweep 37 ms

Total Carrier Power 12.521 dBm/ 2.00 MHz ACP-IBW

Carrier Power	Filter
1 12.521 dBm / 2.000 MHz	OFF

		Lower		Upper		
Offset Freq	Integ BW	dBc	dBm	dBc	dBm	Filter
2.000 MHz	1.000 MHz	-40.61	-28.09	-41.34	-28.82	OFF
3.000 MHz	1.000 MHz	-50.91	-38.38	-51.66	-39.14	OFF

b)

Center 900 MHz
Res BW 120 kHz
#VBW 6.8 kHz
Span 12.2 MHz
Sweep 37 ms

Total Carrier Power 12.686 dBm/ 2.00 MHz ACP-IBW

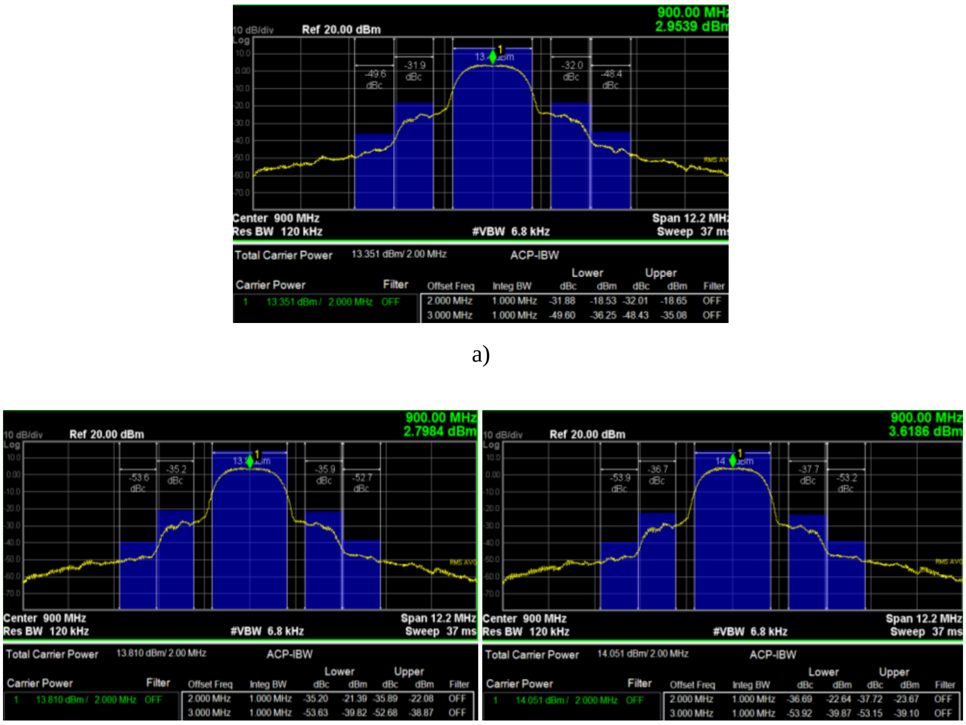
Carrier Power	Filter
1 12.686 dBm / 2.000 MHz	OFF

		Lower		Upper		
Offset Freq	Integ BW	dBc	dBm	dBc	dBm	Filter
2.000 MHz	1.000 MHz	-41.88	-29.19	-42.90	-30.21	OFF
3.000 MHz	1.000 MHz	-52.49	-39.80	-53.56	-40.88	OFF

c)

Fig. 6 Output spectrum for 64QAM signal of 2 MHz useful signal frequency bandwidth for input signal power 3 dBm: a) before linearization; b) after linearization by 1st method; c) after linearization by 2nd method

Symmetrical two-way Doherty amplifier was simulated for 20 MHz LTE signal at 3.5 GHz central frequency of operation. Both linearization methods were considered for different power levels up to 1-dB compression point. Simulation results obtained without and with applying linearization methods are shown in Figures 8 to 10. It can be observed from the Figures that the ACPR at ±20 MHz offsets from the carrier frequency over 2 MHz bandwidth is improved for about 6 dB at 32 dBm output power (near 1-dB compression point) for the 1st method. For lower power levels, ACPR improvement is much better: about 11 dB at 27 dBm output power and nearly 12 dB at 22 dBm output power for the 1st method. For all power levels, the 2nd method shows an improvement in ACPR of 1 dB to 2 dB more comparing to the 1st method. Also, a slight asymmetry in the ACPR reduction can be observed at lower (-20 MHz offset) and upper (+20 MHz offset) adjacent channels for both methods.



a)

Total Carrier Power 13.351 dBm/ 2.00 MHz

Carrier Power			Filter	Offset Freq	Integ BW	Lower dBc	Lower dBm	Upper dBc	Upper dBm	Filter
1	13.351 dBm /	2.000 MHz	OFF	2.000 MHz	1.000 MHz	-31.88	-18.53	-32.01	-18.65	OFF
				3.000 MHz	1.000 MHz	-49.60	-36.25	-48.43	-35.08	OFF

b)

Total Carrier Power 13.810 dBm/ 2.00 MHz

Carrier Power			Filter	Offset Freq	Integ BW	Lower dBc	Lower dBm	Upper dBc	Upper dBm	Filter
1	13.810 dBm /	2.000 MHz	OFF	2.000 MHz	1.000 MHz	-35.20	-21.39	-35.89	-22.08	OFF
				3.000 MHz	1.000 MHz	-53.63	-39.82	-52.68	-38.87	OFF

c)

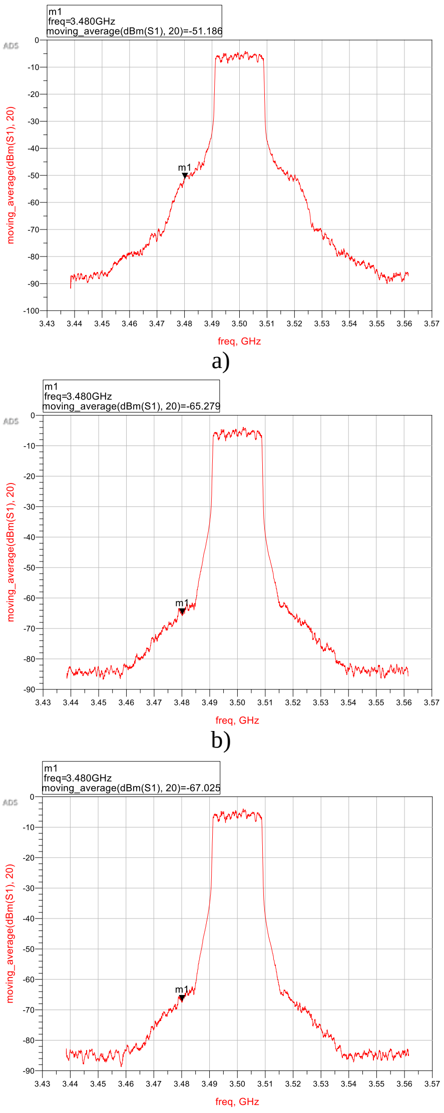
Total Carrier Power 14.051 dBm/ 2.00 MHz

Carrier Power			Filter	Offset Freq	Integ BW	Lower dBc	Lower dBm	Upper dBc	Upper dBm	Filter
1	14.051 dBm /	2.000 MHz	OFF	2.000 MHz	1.000 MHz	-36.69	-22.64	-37.72	-23.67	OFF
				3.000 MHz	1.000 MHz	-53.92	-39.87	-53.15	-39.10	OFF

Fig. 7 Output spectrum for 64QAM signal of 2 MHz useful signal frequency bandwidth for input signal power 5 dBm: a) before linearization; b) after linearization by 1st method; c) after linearization by 2nd method

6. CONCLUSION

Experimental results of the linearization of asymmetrical Doherty amplifier fabricated in microstrip technology obtained by applying two digital linearization methods are presented in this paper, as well as simulation results for symmetrical Doherty amplifier designed to operate in 5G band below 6 GHz. The linearization methods utilize the adequately processed baseband digital signals that modulate the second harmonic of the fundamental carrier. In the 1st linearization method, formed signals for the linearization are injected at the input and output of main transistor in Doherty amplifier, while in the 2nd method these signals are led to the outputs of the main and auxiliary amplifier transistors in the DA circuit. The NI USRP platforms programmed by LabView software were used for generation of the useful 64QAM signals for DA test and measurements of ACPRs in adjacent and alternate channels for various input power levels. Additionally, these platforms form the signals for linearization, and process them in amplitude and phase. Measurements performed by signal analyzer illustrate the results of the linearization for two applied linearization methods and compare them to the states before the linearization.



a)

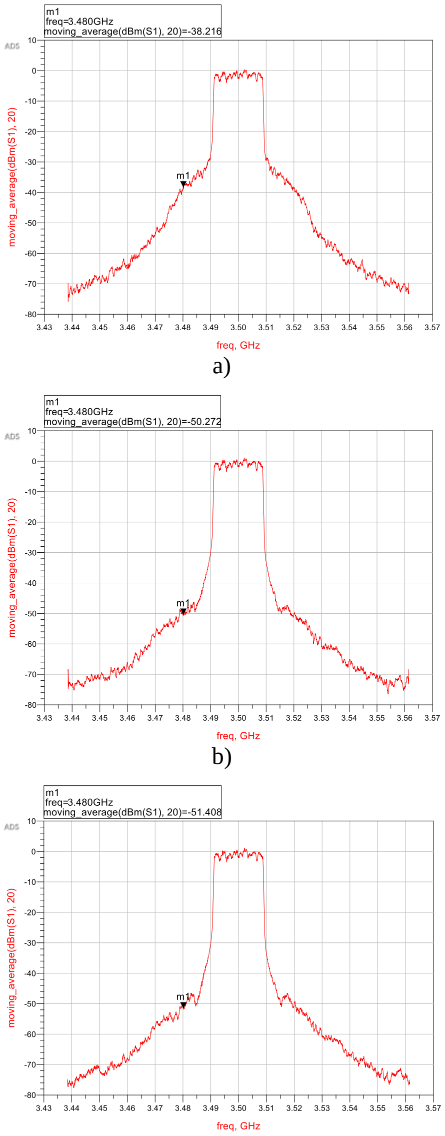
b)

c)

Fig. 8 Output spectrum for LTE signal at 22 dBm output power: a) before linearization; b) after linearization by 1st method; c) after linearization by 2nd method

On the bases of the achieved experimental results, it can be noticed that the 2nd method provides slightly better results for higher power then the application of the 1st method regarding adjacent channels, where the 3rd order IM products are dominant. The same conclusion can be derived for the alternate channels (the band of dominant 5th order IM products) but these results of only 1 dB or 2 dB are inconsiderable, except for the 5 dBm input power where higher improvements of ACPRs were attained in case of both linearization methods.

Based on the obtained linearization results in simulation for symmetrical DA for the 20 MHz LTE signal at 3.5 GHz, it can be assumed that the proposed linearization method can be successfully used for 5G band signals with a bandwidth of 20 MHz. The test for wider 5G modulation formats is a subject of further analysis.

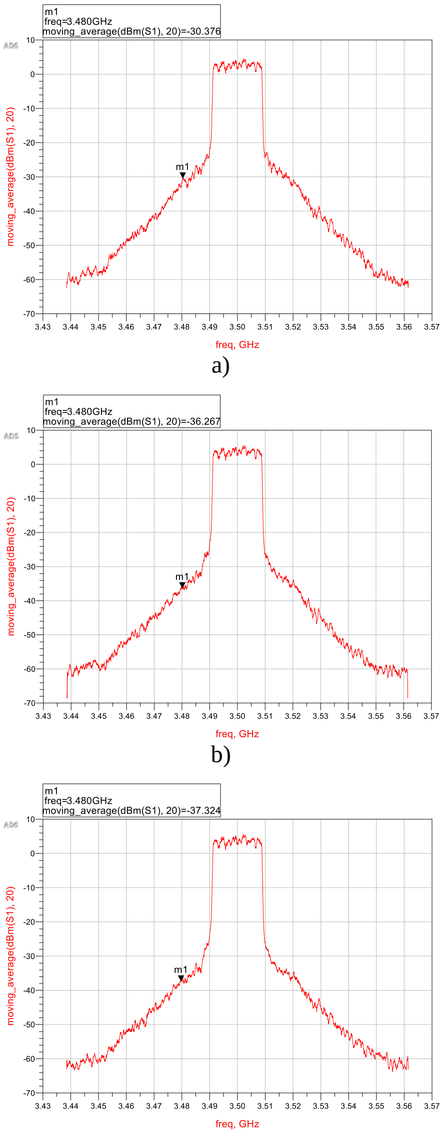


a)

b)

c)

Fig. 9 Output spectrum for LTE signal at 27 dBm output power: a) before linearization; b) after linearization by 1st method; c) after linearization by 2nd method



a)

b)

c)

Fig. 10 Output spectrum for LTE signal at 32 dBm output power: a) before linearization; b) after linearization by 1st method; c) after linearization by 2nd method

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