The Linearization of Doherty Amplifier

Aleksandar Atanasković, Nataša Maleš-Ilić, Bratislav Milovanović

Abstract — **The linearization of the conventional two-way Doherty amplifier and two configurations of three-way Doherty amplifier has been performed in this paper. The fundamental signals' second harmonics (IM2) and fourth-order nonlinear signals (IM4) are exploited for linearization. In the case of two-way Doherty amplifier IM2 and IM4 signals extracted at the output of peaking cell are injected at the input and output of the carrier cell in Doherty amplifier. The linearization approach achieves very good results in the reduction of the third-order intermodulation products retaining the high efficiency of Doherty amplifier. When three-way Doherty amplifier is considered IM2 and IM4 signals are extracted at the output of peaking cells that operate at different bias conditions. The signals for linearization (IM2+IM4) generated at one peaking cell are led to the input of the carrier amplifier whereas ones produced at the output of another peaking cell are directed to the carrier amplifier output. As a result of linearization the third- an fifth order intermodulation products of Doherty amplifier are reduced, which is accompanied by the high efficiency of Doherty configuration.**

Keywords – **Doherty amplifier, linearization, power added efficiency, second harmonics, fourth-order nonlinear signals.**

I. INTRODUCTION

The current wireless communication systems are characterized by high data rate transmission and transmit power that carries high-peak-to-average ratio signals. Therefore, the high linearity and efficiency of power amplifiers of base stations for the systems are required. The Doherty amplifier is capable of achieving the requirements of the power amplifiers in base station concerning high efficiency [1].

The linearity of high power Doherty amplifier was improved using ″post-distortion-compensation″ [2], the feedforward linearization technique [3], the predistortion linearization technique [4] and combination of those two linearization techniques [5].

The influences of fundamental signals' second harmonics on the third-order intermodulation products (IM3) in microwave amplifiers have been investigated and applied in [6-7]. The linearization approach in [8] suggests the injection of second harmonics (IM2) at the amplifier input, whereas the IM2 signals and fourth-order nonlinear signals (IM4) at frequencies that are close to the second harmonics are fed into the amplifier output. The reduction of the third- and fifth-order intermodulation products (IM5) has been achieved by applying a proposed approach for a wide range of the fundamental signals' power going close to 1-dB compression point.

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However, the linearization effects of IM2 and IM4 signals to Doherty amplifier have not been investigated yet according to the author's best knowledge. Therefore, in this paper Doherty amplifier is linearized by applying the approach where IM2 and IM4 signals are injected together with the fundamental signals into the carrier amplifier input and put at its output. When classical (two-way) Doherty amplifier is concerned the signals for linearization are generated at the output of the peaking amplifier in Doherty configuration. Additionally, the linearization of two configurations of three-way Doherty amplifier is considered. The signals for linearization (IM2 and IM4) are extracted at the output of the peaking cells in Doherty amplifier that are biased at various points to provide the appropriate power levels and phase relations of IM2 and IM4 signals. After been adjusted in amplitude and phase the signals from the output of one peaking amplifier are injected at the input of carrier amplifier while ones appeared at the output of another peaking cell are put to the carrier amplifier output.

A theory relating to the proposed linearization approach is given in the section II for digitally modulated fundamental signals. The design of two-way Doherty amplifier with an additional circuit for linearization is described in section III. Also, the simulated results of linearization in case of digitally modulated signal are represented in the same section. The influence of the proposed linearization approach on the power added efficiency of Doherty amplifier for the range of output power is shown as well.

In section IV the design of two configurations of three-way Doherty amplifier is considered. The section also includes all results referring to the intermodulation products and efficiency obtained in simulation by applying the linearization approach for two sinusoidal as well as digitally modulated signals. The results of linearization in the range of carrier output power for two frequency offsets from the carrier, which relate to the first configuration of Doherty amplifier, are shown in the same section.

II ANALYSIS

Theoretical analysis of the proposed linearization approach is based on the nonlinearity of drain-source current expressed by a polynomial model up to the third-order [9].

The expression for the nonlinearity of LDMOSFET in amplifier circuit, under the assumption of neglecting the memory effect, is represented by eleven terms as given by Eq. (1). The drain-source current is dependent upon two control voltages: v_{gs}–voltage between gate and source and v_{ds}-voltage between drain and source of the transistor. The Eq. (1) introduces the nonlinearity of the drain-source current, i_{ds}, in reference to the voltage v_{gs}, which is represented by the coefficients K_{10} to K_{50}. The higher order nonlinear terms K_{40}

and K_{50} are included into the analysis according to the theory given in [10] that favours the terms of output current as function of v_{gs} up to the fifth-order.

The nonlinearity of drain-source current in terms of v_{ds} is included in Eq. (1) over the coefficients K_{01} to K_{03}. Also, the equation encompasses "mixing" terms K_{11}, K_{12} and K_{21}.

$$\begin{aligned} i_{ds}(v_{gs}, v_{ds}) &= K_{10}v_{gs}(t) + K_{20}v_{gs}^2(t) + K_{30}v_{gs}^3(t) + \\ &+ K_{40}v_{gs}^4(t) + K_{50}v_{gs}^5(t) + \\ &+ K_{01}v_{ds}(t) + K_{02}v_{ds}^2(t) + K_{03}v_{ds}^3(t) + \\ &+ K_{11}v_{gs}(t)v_{ds}(t) + K_{21}v_{gs}^2(t)v_{ds}(t) + K_{12}v_{gs}(t)v_{ds}^2(t) + \ldots \end{aligned} \tag{1}$$

A carrier supplemented with a baseband spectrum $V_B(j\omega)$ represents the spectrum of a digitally modulated fundamental signal: $V_B(j\omega) \otimes \frac{1}{2}\delta(\omega \pm \omega_0)$.

The drain-source current at IM3 and IM5 frequencies can be written by the Eqs. (2) and (3), where ρ_2, φ_2, ρ_4 and φ_4 are the amplitudes and phases of IM2 and IM4 signals driven at the amplifier input, whereas $\rho_2^{(F)}$, $\varphi_2^{(F)}$, $\rho_4^{(F)}$ and $\varphi_4^{(F)}$ represent amplitudes and phases of the IM2 and IM4 signals put at the amplifier output (feedforwarded).

The nonlinearity of the drain-source conductance expressed by coefficients K_{01}, K_{02}, K_{03} is assumed to have a negligible contribution to the intermodulation products according to [9] and [10].

$$\begin{aligned} I_{ds}(j\omega)\big|_{IM3} \approx &\left\{\left[\frac{3}{4}K_{30} + \frac{1}{4}K_{20}\rho_2 e^{-j\varphi_2} + \right.\right. \\ &\left.\left. -\frac{1}{4}K_{11}\rho_2^{(F)}e^{-j\varphi_2^{(F)}} - \frac{1}{4}K_{11}\rho_1\rho_2 e^{-j\varphi_2}\right]\right\} \\ &(V_B(j\omega) \otimes V_B(j\omega) \otimes V_B(j\omega))\} \otimes \frac{1}{2}\delta(\omega \pm \omega_0) \end{aligned} \tag{2}$$

The signal distorted by the cubic term of the amplifier, K_{30}, is included into analysis by Eq. (2) as the first term. The cubic term is considered as a dominant one according to [9] and [10] in causing IM3 products and spectral regrowth. The mixing product of the fundamental signal and second harmonic injected at the amplifier input is expressed as the second term. The mixing term K_{11} (third term) exists due to the reaction between the gate-source voltage of fundamental signal and voltage of second harmonic fed at the amplifier output. Additionally, the fundamental signal at the output of amplifier mingles with the second harmonic that exists at the amplifier input generating K_{11} term. The amplitude of output voltage at fundamental signal frequency that is 180° out of phase in reference to the input signal is denoted as ρ_1. According to this, it is possible to reduce spectral regrowth caused by the third-order distortion of fundamental signal by choosing the appropriate amplitude and phase of both the injected second harmonics (ρ_2 and φ_2) and feedforwarded second harmonics ($\rho_2^{(F)}$ and $\varphi_2^{(F)}$).

The mixing terms between drain and gate, K_{12} and K_{21}, produce drain-source current at IM3 frequencies with the opposite phases so that they reduce each other [10].

$$\begin{aligned} I_{ds}(j\omega)\big|_{IM5} \approx &\left\{\left[\frac{5}{8}K_{50} + \frac{1}{4}K_{20}\rho_4 e^{-j\varphi_4} + \right.\right. \\ &-\frac{1}{4}K_{11}\rho_4^{(F)}e^{-j\varphi_4^{(F)}} - \frac{1}{4}\rho_1 K_{11}\rho_4 e^{-j\varphi_4} + \frac{1}{8}K_{30}\rho_2^2 e^{-j2\varphi_2} + \\ &+\frac{1}{8}K_{12}\rho_2^{(F)2}e^{-j2\varphi_2^{(F)}} + \frac{1}{8}K_{12}\rho_1\rho_2\rho_2^{(F)}e^{-j\left(\varphi_2+\varphi_2^{(F)}\right)} + \\ &\left. -\frac{1}{8}K_{21}\rho_2\rho_2^{(F)}e^{-j\left(\varphi_2+\varphi_2^{(F)}\right)} - \frac{1}{8}K_{21}\rho_1\rho_2^2 e^{-j2\varphi_2}\right] \\ &V_B(j\omega) \otimes V_B(j\omega) \otimes V_B(j\omega) \otimes V_B(j\omega) \otimes V_B(j\omega)\} \\ &\otimes \frac{1}{2}\delta(\omega \pm \omega_0) \end{aligned} \tag{3}$$

In Eq. (3) the first term expresses the drain-source current of the fifth-order intermodulation products (IM5) that is formed between the fundamental signals due to the amplifier nonlinearity of the fifth-order, K_{50}. The second term is made by the reaction between the fundamental and IM4 signal at the amplifier input. The third term is the mixing product between the fundamental signal at amplifier input and IM4 signal fed to its output. Also, the fundamental signal at the amplifier output reacts with the IM4 signal injected at the amplifier input over K_{11} term producing IM5 product (fourth term). Therefore, the original IM5 product (the first term) can be reduced by adjusting the amplitude and phase of IM4 signals that are injected at the input of amplifier and put at its output.

The IM5 products are also expressed in terms of K_{30} coefficient-the fifth term in Eq. (3) made by reaction between two IM2 signals and fundamental one. It is obvious that for the larger amplitude of the fundamental signal the injected second harmonics are supposed to have greater amplitudes as well according to Eq. (2). Since φ_2 should be equal to 180° to reduce IM3 products the phase of the K_{30} term in Eq. (3) is 360°. Accordingly, with the rise in amplitudes of second harmonics, mixing K_{30} term (the fifth term in Eq. (3)) starts increasing the power of IM5 spectrum. Due to the overlapping of IM3 and IM5 spectra the raise of power in the range of IM3 spectrum is unavoidable. Therefore, the power of second harmonics run at the amplifier input should be kept at the reasonable level.

All mixing terms, which stand by K_{12} and K_{21} coefficients in Eq. (3), are generated due to the reaction between two second harmonics and fundamental signal. The signals taken in consideration are observed at the input and output of amplifier. The K_{12} and K_{21} terms produce current at the frequencies of IM5 products with the opposite phases so that they reduce each other. Consequently, their influence to the power of IM3 and IM5 products can be cancelled. As a result,

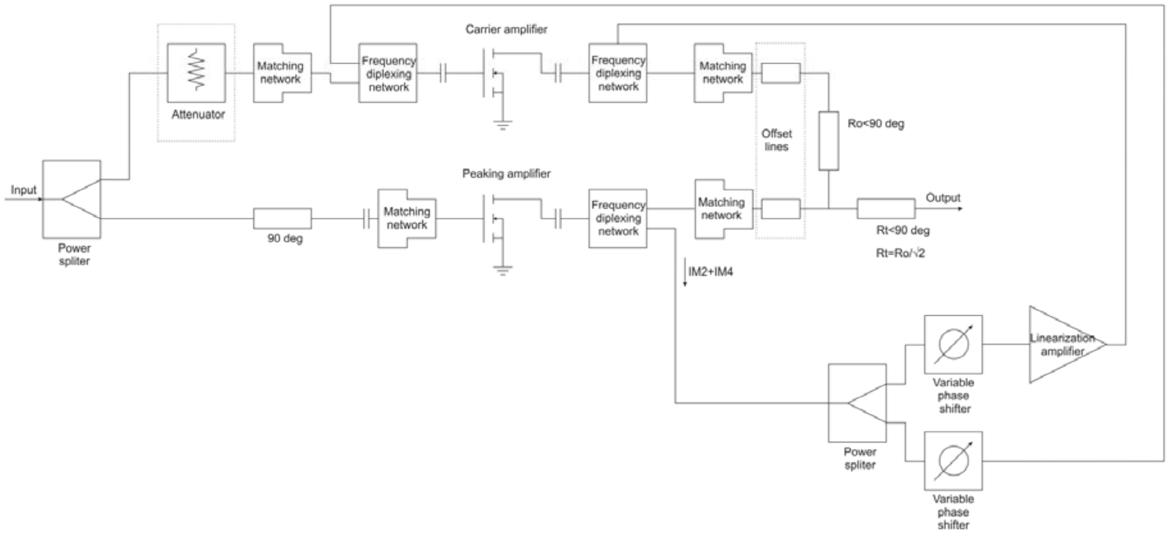

Fig. 1. Doherty amplifier with additional circuit for linearization

the second harmonics fed at the amplifier output are allowed to have power levels that are high enough to reduce IM3 products.

It should be pointed out that if IM2 and IM4 signals are put only at the amplifier output the IM2 and IM4 signals at the amplifier input will have sufficient power (second and fourth terms in Eqs. (2) and (3) will exist) so that can raise the power of IM3 and IM5 products. Therefore, it is necessary to inject IM2 and IM4 signals at the amplifier input that will set the adequate amplitudes and phases of the second and fourth terms in Eqs. (2) and (3) to cancel their undesirable influence to the IM3 and IM5 products.

In case when both amplitudes and phases of IM2 and IM4 signals are not related so that can suppress the IM3 and IM5 products simultaneously one kind of intermodulation products will not be lowered sufficiently or, unfortunately, they can increase in power. This situation is more probable when only one source of IM2 and IM4 signals is used that will be shown through the example in further text.

III. Two-Way Doherty Amplifier

Advanced Design System-ADS software has been used for the design of conventional two-way Doherty amplifier which schematic diagram is shown in Fig. 1.

The carrier and peaking amplifiers have input and output matching circuits which transform from the input impedance of the device to 50Ω and from the optimum load impedance Z_{opt} to 50Ω, [11]. In the low-power region, the peaking amplifier should be an open circuit and load impedance of the carrier amplifier should be doubled to $2Z_{opt}$ by a quarter-wave impedance transformer with the characteristic impedance R_0=50Ω. Also, quarter-wave transmission line with the characteristic impedance $R_t = R_0\sqrt{2}$ transforms 50Ω to 25Ω that is load impedance of output combining circuit when the peaking amplifier is turned on in higher power region. Phase difference of 90° is required at the inputs of the carrier and peaking amplifier to compensate for the same phase difference between those two amplifiers causing by the quarter-wave impedance transformer at the output.

The carrier and peaking cells were designed using Freescale's MRF281S LDMOSFET which non-linear MET model is incorporated in ADS library. The transistor shows a 4-W peak envelope power. The matching impedances for source and load at 2.14GHz are $Z_S = 3.1 - j3.5\,\Omega$ and $Z_L = 11.36 + j7.94\,\Omega$, respectively.

The output impedance of the LDMOSFET is strongly reactive with low resistance so in low-power region considerably power leaks from the carrier amplifier to the peaking amplifier. The output impedance seen at the output of the peaking transistor is transformed to the open by the output matching circuit and the proper offset line [11]. The length of the offset line in the output circuit of the carrier amplifier is chosen to be 22.5°, the same as that for the peaking amplifier.

The carrier amplifier is biased in class-AB ($V_D = 26\text{V}$, $V_G = 5.1\text{V}$ (13.5%I_{DSS})), whereas the peaking amplifier operates in class-C ($V_D = 26\text{V}$, $V_G = 3.2\text{V}$). According to the analysis performed in [12] Doherty amplifier with uneven power drive of the carrier and peaking cells operates more linear and generates full power from both amplifiers producing in that way more power than configuration with identical power level. Therefore, for the two-way Doherty amplifier considered in this paper the carrier amplifier is driven with 2.4dB lower power than peaking amplifier.

Power level of intermodulation products depends on nonlinearity of both amplifying cells but especially on the carrier amplifier which goes to saturation before peaking amplifier. The second harmonics (IM2) generated at the output of the peaking amplifier are extracted through a diplexer circuit [13] shown in Fig. 2 that was designed to separate the fundamental signals and their second harmonics. However, there are IM4 signals at the output of the peaking amplifier that cannot be neglected. They are also led together with the second harmonics through two paths at the input and output of the carrier amplifier.

Having sufficient power, the signals for linearization that should be injected at the input are tuned only in phase, whereas the ones directed at the output of carrier amplifier are adjusted in amplitude and phase through the amplifier and phase shifter as depicted in Fig. 1. They are inserted at the carrier amplifier input and output by the diplexers.

Since IM2 and IM4 signals influence not only IM3 but also IM5 products referring to Eqs. (2) and (3) it is possible to reduce both kinds of intermodulation products, which depends on the relations between amplitudes as well as phases of the IM2 and IM4 signals generated at the peaking amplifier output. However, when required relations are not fulfilled, only one kind of intermodulation products can be reduced sufficiently, while the other will be raised.

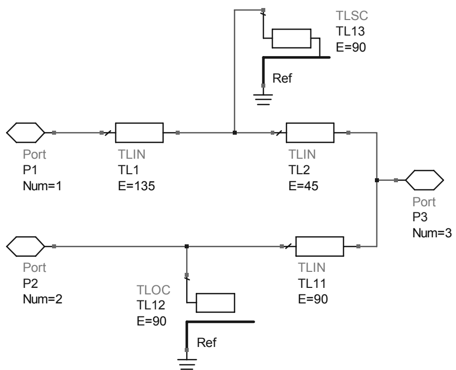

Fig. 2. Diplexer

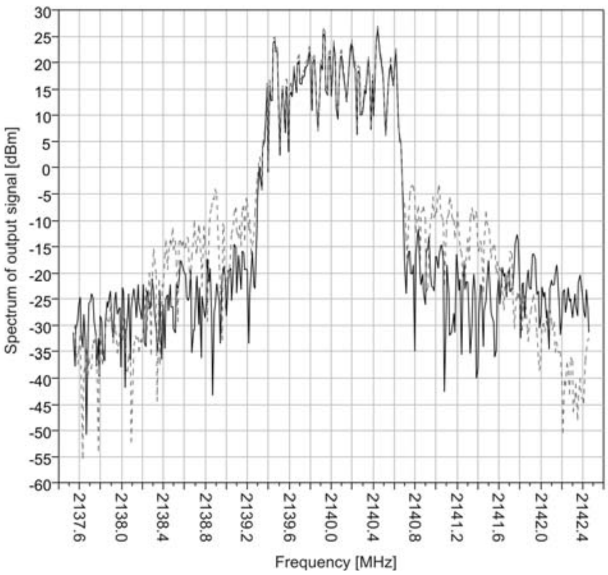

Fig. 3. Simulated spectrum of the output signal of two-way Doherty amplifier for OQPSK digitally modulated signal before (dashed line) and after linearization (solid line) for 36dBm output power

The Doherty amplifier was simulated for OQPSK digitally modulated signals with 1.25MHz spectrum width, carrier at frequency 2.14GHz with output power 36dBm. It should be noticed that peak-to-average power ratio in this case is 6dB. The improvement in adjacent channel power ratio (ACPR) for ±900kHz offset from carrier frequency over 30kHz bandwidth gained by applying the linearization approach proposed herein is 9dB. However, ACPR for the offsets that belongs to the region of IM5 products gets worse by approximately 8dB. It can be seen from Fig. 3 which compares the output spectra before and after linearization.

The power added efficiency (PAE) for a single carrier is shown in Fig. 4 that compares PAE of the Doherty amplifier linearized in a proposed manner with the case of Doherty amplifier when DC supply of the additional circuit for linearization is not counted. Moreover, the figure includes PAE of class-AB amplifier achieved with AB-biasing of both cells of Doherty amplifier. Power added efficiency of designed Doherty without linearization at the maximum output power is 47% while at the 6dB back-off point it drops to 26.9%. At 36dBm power point, which represents 3dB back-off from a maximum output power, PAE of linearized Doherty amplifier is only 6.1% lower than that in case when linearization is not applied (36.8%). It can be seen that PAE of class-AB amplifier is lower than that for Doherty amplifier even in case when the additional linearization circuit is considered.

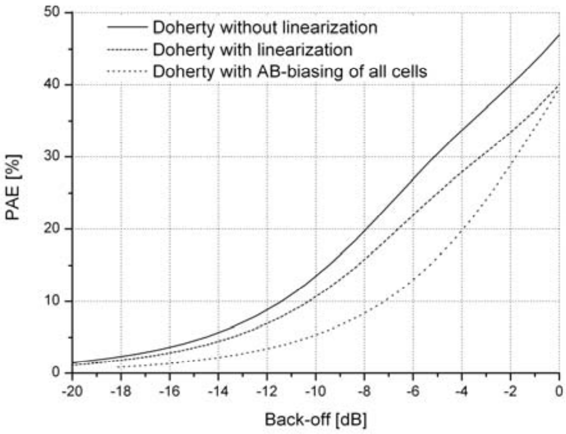

Fig. 4. Power added efficiency of two-way Doherty amplifier

IV. THREE-WAY DOHERTY AMPLIFIER

A. First configuration

In a classical Doherty amplifier operation high efficiency is obtained over a range of 6dB below maximum output power. The concept of multistage Doherty amplifier is used to maintain the efficiency in the backoff region that can be extended beyond the classical design. Three-way Doherty amplifier shown in Fig. 5 was designed on the theoretical principle given in [11] by using ADS.

The carrier and peaking cells were designed using Freescale's MRF281S LDMOSFET. The carrier amplifier is biased in class-AB, $V_G = 5.1\text{V}$ (13.5%I_{DSS}) whereas one peaking amplifier operates in class-B $V_G = 3.8V$ and the other one is biased in class-C, $V_G = 0.6\text{V}$. The drain bias voltage is the same for all amplifiers, $V_D = 26\text{V}$. Uneven power drive of the carrier and peaking cells in the Doherty amplifier is adjusted to be 1:1.4:1.4 in order to produce maximum power from all-cells and operate more linear.

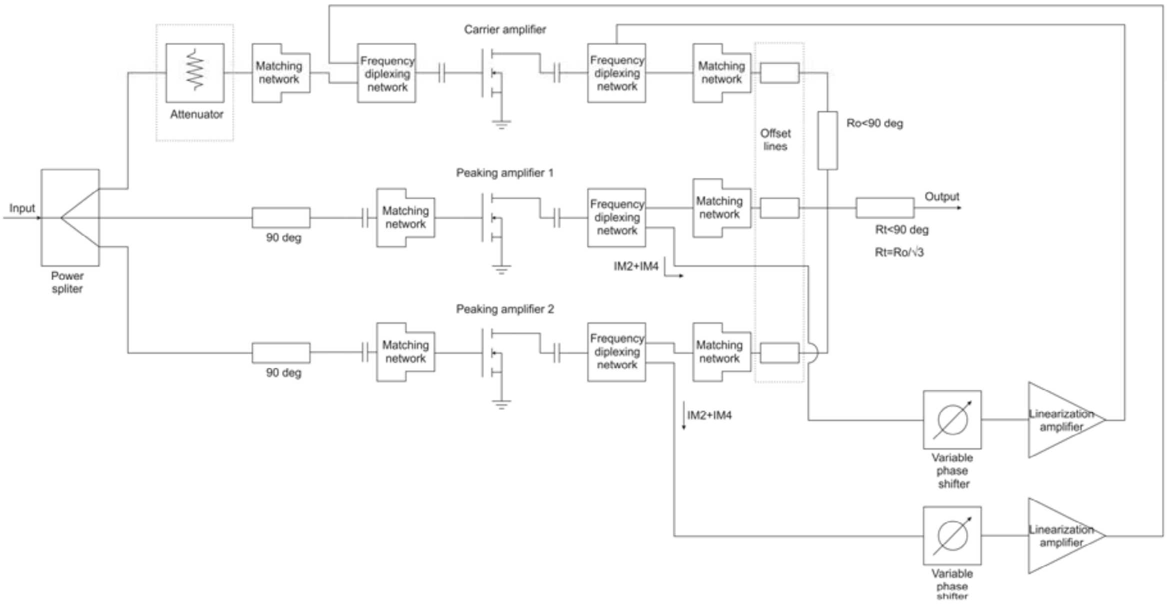

Fig. 5. Three-way Doherty amplifier with additional circuit for linearization (first configuration)

The load impedance seen at the offset line of the carrier amplifier is $3\cdot 50\Omega$ at a low-power region while it is still 50Ω at a high-power region. The input and output matching circuits of amplifier cells in three-way Doherty configuration are designed as described in Section III for the case of classical two-way Doherty amplifier. Since the optimum output impedances of amplifying cells are all matched to 50Ω in the high-power region the characteristic impedance of the output impedance transformer should be $R_t = 50\Omega/\sqrt{3}$. Driving signals of the peaking amplifiers should be shifted in phase by -90° in reference to the input signals of carrier amplifier in order to compensate for the phase shift caused by the quarter-wave transformer with the characteristic impedance $R_0 = 50\Omega$ in output combining circuit of Doherty amplifier. The same length of the offset lines (22.5°) in output circuits of the carrier and peaking amplifiers is implemented.

The IM2 and IM4 signals generated at the output of the peaking amplifiers are extracted through diplexer circuits. Those signals turn up at the output of peaking amplifier 1 are led to the input of the carrier amplifier. The peaking amplifier 2 generates IM2 and IM4 signals that are directed to the output of the carrier amplifier. The IM2 and IM4 signals are tuned in amplitude and phase by the amplifier and phase shifter over two paths as given in Fig. 5. The signals are inserted at the carrier amplifier by the diplexers.

The results of linearization are shown in Fig. 6 for two sinusoidal signals at the input of Doherty amplifier at frequencies 2.139GHz and 2.141GHz. The case before linearization is compared with the results achieved after the application of the linearization technique proposed. It follows from the figure that IM3 products are reduced by approximately 19dB for the output power of each signals 35dBm. Additionally, IM5 products are lowered by 5dB. It is evident from the figure that optimized amplitudes and phases of the signals for linearization (IM2 and IM4) suppress the IM3 products to the lower power levels than the IM5 products. If optimized parameters are set to some other values the IM3 and IM5 products will be reduced to the even power levels.

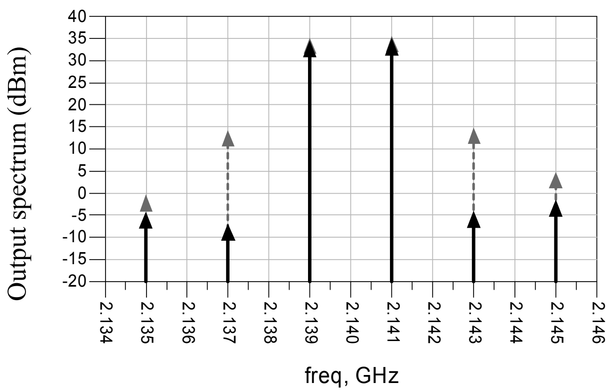

Fig. 6. Output spectrum of three-way Doherty amplifier in the first configuration for 35dBm average output power of fundamental signals; before (dashed line) and after the linearization (solid line)

The power added efficiency for a single carrier is shown in Fig. 7. The PAE of three-way Doherty amplifier is 40.5% in case when linearization is not applied at 38dBm output power (3dB back-off). If the consumption of the additional circuit for linearization is concerned then PAE of Doherty amplifier is 35.6%. On the top of that PAE of class-AB amplifier is significantly lower even in case when additional linearization circuit is considered.

The PAE of linearized three-way Doherty amplifier that is 23.1% at 6dB back-off point (35dBm) is 1.2% higher in reference to the classical Doherty two-way amplifier where PAE is 21.9% for the same back-off.

Additionally, three-way Doherty amplifier was tested for OQPSK digitally modulated signals with 1.25MHz spectrum width, carrier at frequency 2.14GHz with output power 36dBm. Fig. 8 illustrates 13dB improvement in ACPR for ±900kHz offset from carrier frequency over 30kHz bandwidth procured by applying simultaneous injection of IM2 and IM4 signals at the input and output of carrier amplifier. Also, the figure represents ACPR for 2085-2115kHz offset, the range that belongs to the spectrum of IM5 products, where ACPR is improved by approximately 6dB. This is a satisfactory enhancement in comparison with the results acquired for two-way Doherty amplifier when only one peaking cell generate IM2 and IM4 signals for linearization.

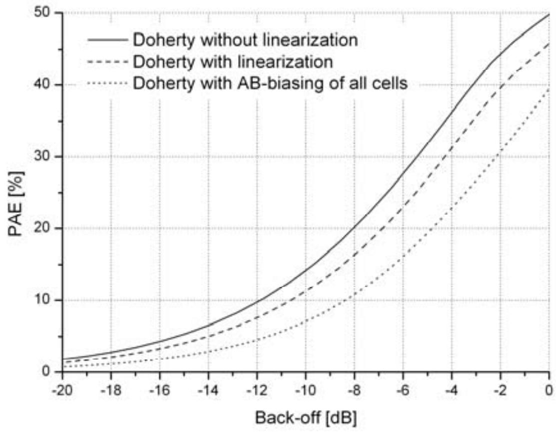

Fig. 7. Power added efficiency of three-way Doherty amplifier in the first configuration

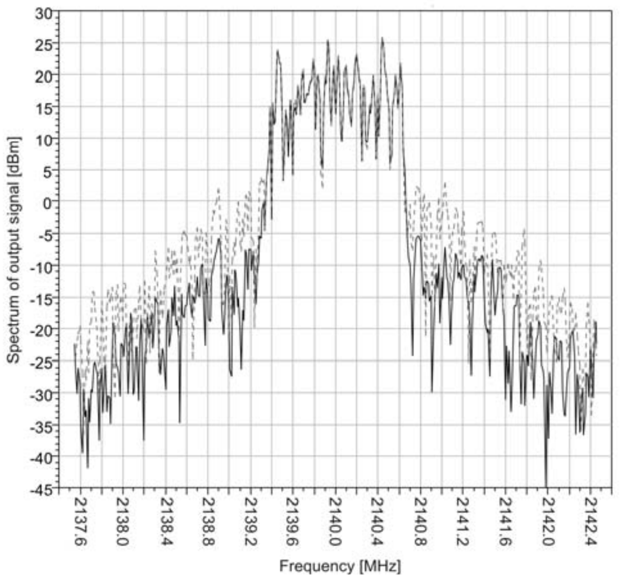

Fig. 8. Simulated spectrum of the output signal of three-way Doherty amplifier in the first configuration for OQPSK digitally modulated signal before (dashed line) and after linearization (solid line) for 36dBm output power

The results from Fig. 9 show the effects of three-way Doherty amplifier linearization accomplished for the output power ranging from 30dBm to 39dBm. These results are compared to the case when linearization is not carried out. Fig. 9a) relates to the power reduction of IM3 and IM5 products at 900kHz and 2100kHz offsets from carrier frequency, whereas the Fig. 9b) shows results connected to the opposite offsets. The presented results relate to the case when the amplitudes and phases of IM2 and IM4 signals are adjusted on the optimal values for 36dBm output power. It is evident from these figures that the linearization with the proposed approach gives satisfactory results in improvement of ACPR for both offsets in the range of output power (approximately 33dBm-38dBm).

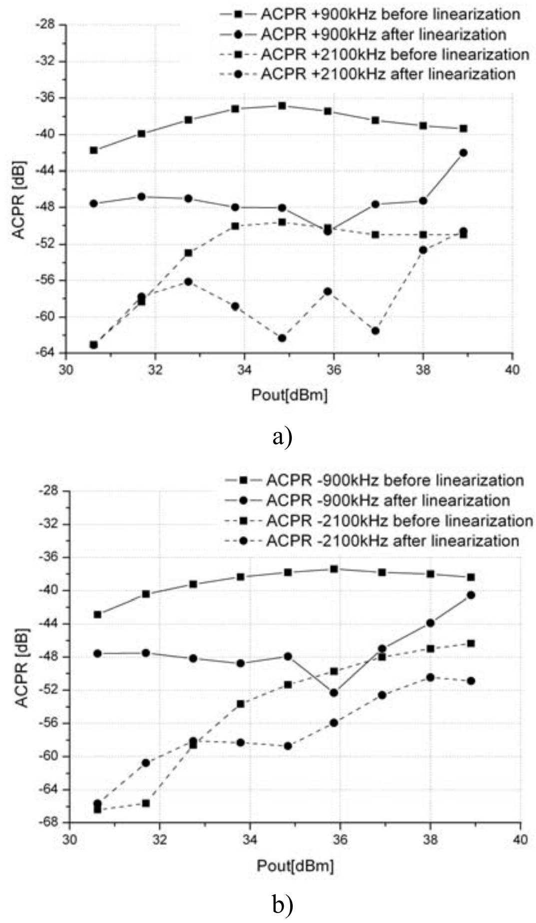

Fig. 9. ACPR before and after linearization in case of three-way Doherty amplifier in the first configuration for a power range at offsets from carrier frequency; a) +900kHz and +2100kHz, b) -900kHz and -2100kHz

B. Second Configuration

In the Doherty configuration proposed in [14] the characteristic impedances of the quarter-wave transmission lines in output combining network were found to be 120Ω and 30Ω as shown in Fig. 10. The values of those impedances were determined to meet the required Doherty amplifier operation of peak efficiency at 6dB and 12dB backoff.

The carrier and peaking cells were designed using Freescale’s MRF281S LDMOSFET. On the bases of the study performed in [14] the carrier and peaking 1 amplifier should provide the same power at first backoff point (6dB). At the maximum output power the carrier amplifier should deliver the same power as at the first backoff point, whereas the peaking amplifiers 1 and 2 should, respectively, produce three and four times more power than carrier amplifier. Also, their phases lag 90° and 180°, respectively, behind that of carrier amplifier.

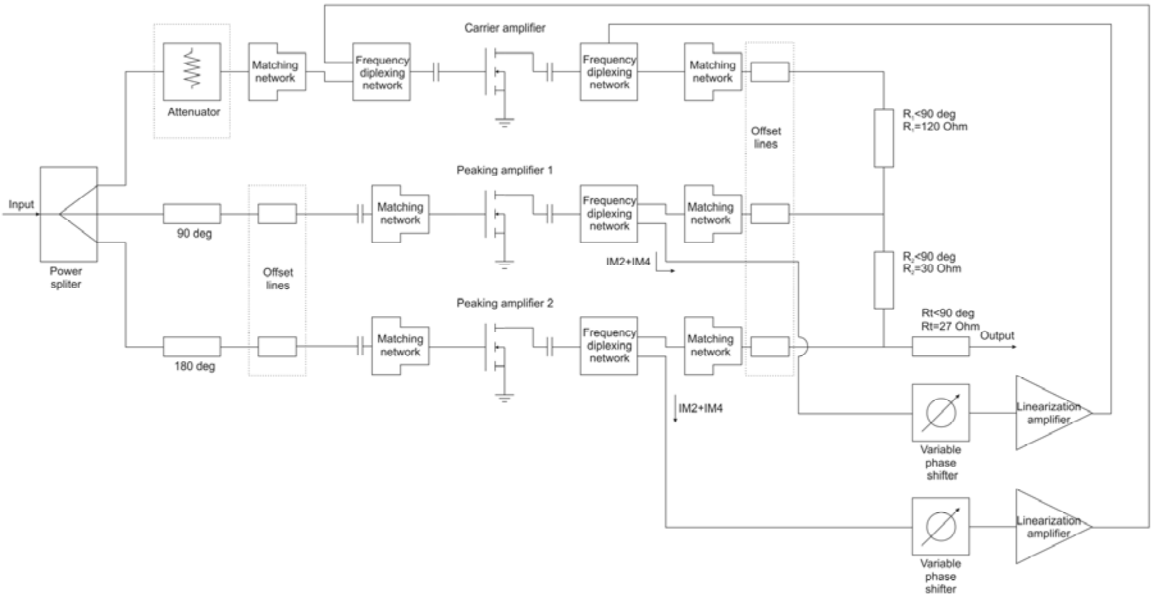

Fig. 10. Three-way Doherty amplifier with additional circuit for linearization (the second configuration)

The carrier amplifier is biased at class-AB, $V_G = 4.6\text{V}$ (6.5%I_{DSS}), whereas the two peaking amplifiers operate in class-C, (peaking 1 amplifier $V_G = 2.6\text{V}$ and peaking 2 $V_G = 0.1\text{V}$).

The drain bias voltage is the same for all amplifiers, $V_D = 26\text{V}$. Driving signal for the carrier amplifier is 1dB lower than that of the peaking amplifiers.

The output matching circuits are designed to transform the output impedances of transistors in amplifier cells to the impedances that provide the required output power of each amplifier in Doherty configuration expressed by Eqs. (4) to (6), [15].

$$P_{carrier} = P_{outDPA}/8 \tag{4}$$

$$P_{peaking1} = P_{outDPA} \cdot (3/8) \tag{5}$$

$$P_{peakin2} = P_{outDPA}/2 \tag{6}$$

where P_{outDPA} is maximum output power of Doherty amplifier.

Therefore, in the high-power range output impedance of the carrier amplifier should be $8 \cdot R_L$, whereas output impedance of the peaking amplifier 1 should be $8 \cdot R_L/3$, and that of the peaking amplifier 2 should be $2 \cdot R_L$, where R_L is $R_{opt\_p2}/2$ if R_{opt_p2} is optimum termination of the peaking amplifier 2. In the three-way Doherty amplifier design the output impedances of carrier and peaking amplifiers are determined to be 120Ω, 40Ω and 30Ω, respectively, while $R_L = 15\Omega$ obtained from 50Ω by the quarter-wave transformer with characteristic impedance $R_t = 27\Omega$ as denoted in Fig. 10.

Offset lines are incorporated at the output of peaking amplifier cells to minimize the effective loading of the peaking amplifiers in state when those amplifiers do not operate (low-power range). The insertion of 40Ω line at the output of the peaking amplifier 1 and 30Ω line at the output of the peaking amplifier 2 will not only rotate output impedances of the peaking cells to as high as possible values but will also change the phase of the peaking amplifiers when they are active, distorting the desired phase relations in Doherty amplifier. If the delay lines are inserted at the input of the peaking amplifiers or an appropriated offset line is adjusted at the output of the carrier amplifier the compensation of the phase discrepancies can be carried out. Length of the offsets lines at output circuits of carrier and peaking amplifiers 1 and 2 are 28°, 45° and 51°, respectively.

As in the case of the first configuration of three-way Doherty amplifier IM2 and IM4 signals appeared at the output of two peaking amplifiers are adjusted over two paths and injected at the input and output of the carrier amplifier. The characteristics of PAE obtained in simulation for the designed Doherty configuration with and without the additional circuit for linearization are compared in Fig. 11 with PAE of the Doherty configuration presented previously.

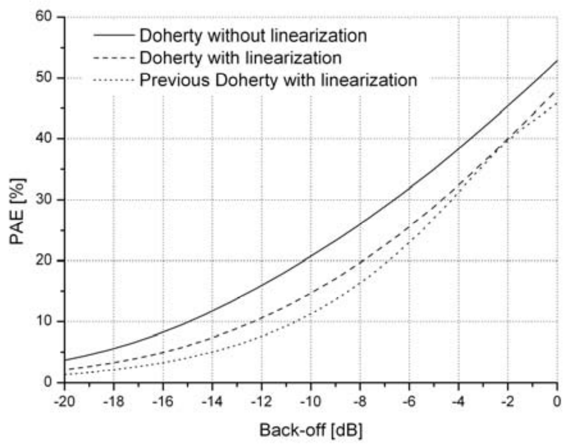

Fig. 11. Power added efficiency of three-way Doherty amplifier in the second configuration

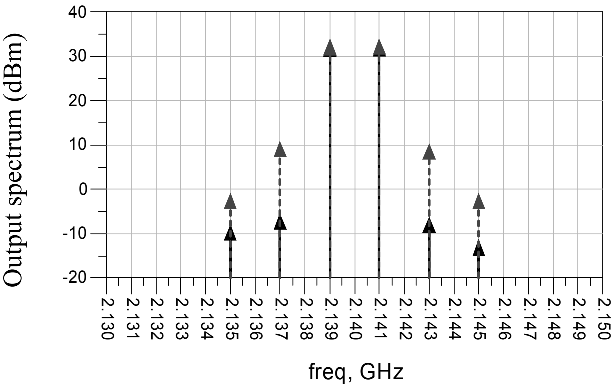

Fig. 12. Output spectrum of three-way Doherty amplifier in the second configuration for 33dBm average output power of fundamental signals; before (dashed line)and after the linearization (solid line) in case of 2MHz signal space

The effects of linearization are clear from Fig. 12. In the case of two sinusoidal fundamental signals at frequencies 2.139GHz and 2.141GHz at Doherty amplifier input when they have 33dBm output power, IM3 and IM5 product are reduced by 18dB and 7dB, respectively. The intermodulation products are descended for lower grade when fundamental signals differ in frequency five times more. Fig. 13 shows the results of lineariarization achieved for two sinusoidal fundamental signals at frequencies 2.135GHz and 2.145GHz and output power 33dBm and 34dBm, respectively. The asymmetry in output spectrum is consequence of the narrow-band matching and bias circuit design; therefore with the broadband matching and biasing the symmetry of output spectrum components will be preserved [7, 8].

It follows from the figure that IM3 products are lessened asymmetrically (18dB and 10dB) while one IM5 product is suppressed only a few decibels and the other increases for 0.5dB. Generally, the intermodulation products are lower than before linearization despite the broader signal space.

In the case of 36dBm total output power of fundamental signals that corresponds to 5dB back-off point Fig. 11 shows 28.9% PAE.

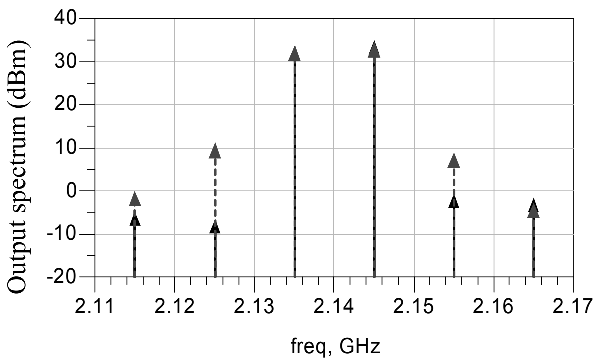

Fig. 13. Output spectrum of three-way Doherty amplifier in the second configuration for 33dBm average output power of fundamental signals; before (dashed line)and after the linearization (solid line) in case of 10MHz signal space

The linearization results for OQPSK digitally modulated signals with 1.25MHz spectrum width, carrier at frequency 2.14GHz with output power 36dBm are shown in Fig. 14. The improvement in ACPR at ±900kHz and ±2100kHz offsets from carrier frequency over 30kHz bandwidth is 18dB and 10dB, respectively.

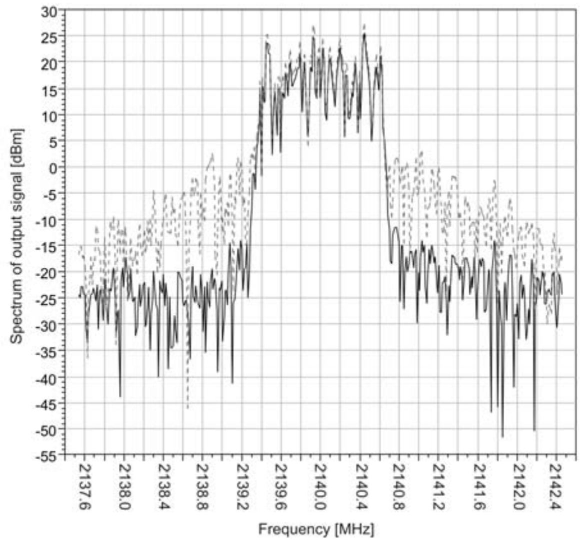

Fig. 14. Simulated spectrum of the output signal of three-way Doherty amplifier in the second configuration for OQPSK digitally modulated signal before (dashed line) and after linearization (solid line) for 36dBm output power

The peaking amplifiers that are forced to operate in class-C have tipically lower gain compared to the class-AB of carrier amplifier. Therefore, the periphery of the peaking amplifiers needs to be enlarged to compensate for the gain reduction in the class-C amplifiers. The device periphery calculated for the determined back-off levels (6dB and 12dB) resultes in a ratio 1:3:4, [14], [15]. However, the choice was limited by the device availability so that 1:2.5:2.5 device size ratio was selected.

The choice of the carrier cell transistor was the same as previous while for the peaking cells in three-way Doherty amplifier Freescale's MRF282S LDMOSFET was utilized. Non-linear MET model included in ADS library exhibits a 10-W peak envelope power level. The drain bias voltage $V_D = 26\text{V}$ is the same for all amplifiers. The carrier amplifier was biased at class-AB, $V_G = 5.1\text{V}$ (13.5%I_{DSS}), peaking amplifiers 1 and 2 were set to operate at class-C, $V_G = 3V$ and $V_G = 1.5\text{V}$, respectively. The matching impedances of MRF282S LDMOSFET for source and load at 2.14GHz are $Z_S = 1.85 + j1.6\,\Omega$ and $Z_L = 3.55 - j0.2\,\Omega$. The inpit matching was performed for 50Ω while output matching circuits were designed to transform the optimum output impdance of the carrier and two peaking cells to 120Ω, 40Ω and 30Ω, respectively. The length of offset lines at the peaking amplifer output circuits determined to lower leackage current at low-power region are 10° and 12°. In order to compensate for the phase variaty at the output combining circuit the signals at the input of the peaking amplifiers were delayed for the appropriate values. The peaking amplifiers were driven by signals with 1dB higher power than that of the carrier amplifier. Maximum output power achieved by this Doherty configuration is 44dBm.

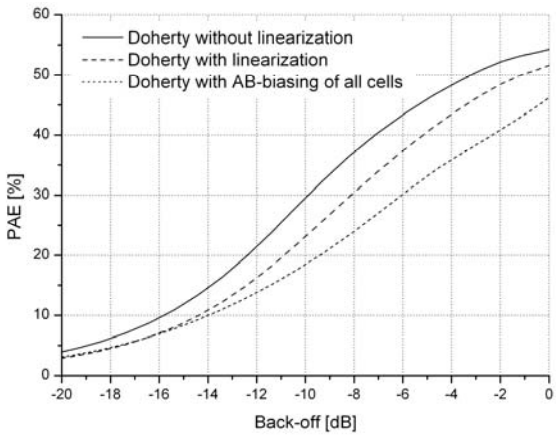

Fig. 15. Power added efficiency of three-way Doherty amplifier in the second configuration for device size ratio 1:2.5:2.5

Power added efficiency for the designed three-way Doherty amplifier is presented in Fig. 15 showing considerably better PAE characterisstic comparing to class AB amplifier. It follows from the figure that PAE aslo shows exelent improvement, even in case of included consumption of the additional linearization circuit, in reference to the case when the same devices (1:1:1 device size) were used for carrier and two peaking amplifiers (Fig. 11). For example, PAE of Doherty with linearization is 3.5% higher at the maximum output power and 5.7% at the 12dB back-off point than in configuration analysed previosly while at 3dB and 6dB back-off that difference is 9.9% and 11.8%, respectively.

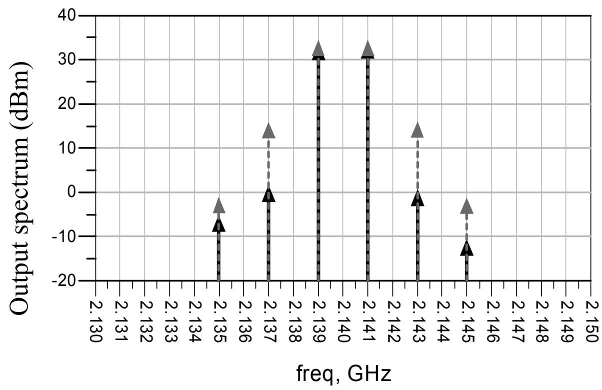

Fig. 16. Output spectra of three-way Doherty amplifier in the second configuration for device size ratio 1:2.5:2.5 for 33dBm average output power of fundamental signals; before (dashed line) and after the linearization (solid line) in case of 2MHz signal space

Two-tone test of three-way Doherty amplifier at frequencies 2.139GHz and 2.141GHz gives results of linearization with the use of IM2 and IM4 signals in Fig. 16. It compares output spectra before and after the linearization in case of 33dBm output power of each fundamental signal. It can be noticed that the levels of IM3 and IM5 products for the Doherty amplifier with the device periphery 1:2.5:2.5 are higher than that of the 1:1:1 design (Fig. 12), though the IM3 and IM5 products are suppressed for 15dB and 5dB, respectively.

The output spectra obtained in simulation before and after linearization for OQPSK digitally modulated signal with the same characteristic as in previous tests are compared in Fig. 17. For 36dBm output power (8dB back-off), ACPR at two offsets (±900kHz and ±2100kHz) is improved for 11dB. It follows from Fig. 15 that PAE at this power level is 30.5%.

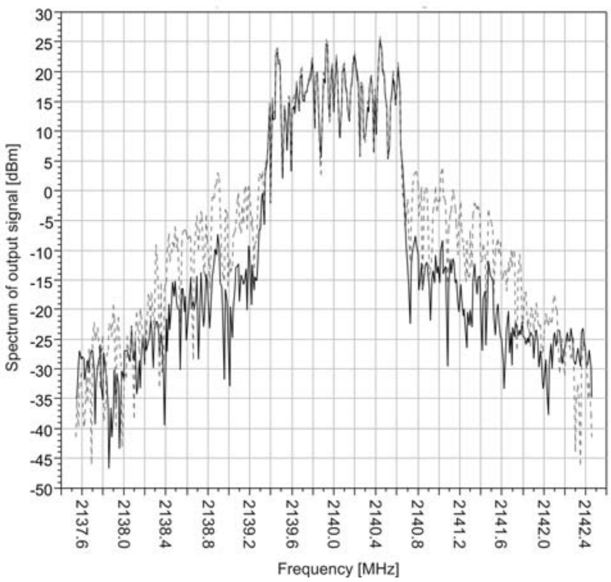

Fig. 17. Simulated spectrum of the output voltage for three-way Doherty amplifier in the second configuration for device size ratio 1:2.5:2.5 for OQPSK digitally modulated signal before (dashed line) and after linearization (solid line) for 36dBm output power

V. CONCLUSION

This paper presents the linearization of Doherty amplifier in classical configuration (two-way) as well as three-way Doherty amplifier by the simultaneous injection of the second harmonics and fourth-order nonlinear signals (IM2 and IM4) at the input and output of the carrier amplifier. Those signals are generated at the output of peaking amplifiers that are biased at different points to produce adequate amplitude and phase relations between IM2 and IM4 signals. Two different configurations of three-way Doherty amplifier are considered. Additionally, the second configuration is analysed with LDMOSFET in carrier and peaking amplifies in periphery relations 1:1:1 and 1:2.5:2.5.

When two-way Doherty amplifier is considered the result is very good reduction of IM3 products accompanied with the high efficiency of Doherty amplifier whereas the IM5 products cannot be suppressed at all. Power added efficiency at the maximum power is 40% while at the 6dB back-off point it drops to 21.9%.

For the three-way Doherty amplifier the linearization approach achieves very good results in the reduction of both IM3 and IM5 products retaining the high efficiency of Doherty amplifier (23.1% at 6dB back-off point).

The three-way Doherty amplifier in the second configuration with device periphery relation 1:1:1 shows 25.6% PAE at 6dB back-off point that is 2.5% higher than that in the latter case. When periphery relation is 1:2.5:2.5 PAE at back-off points 6dB and 12 dB is 37.4% and 16.3%, respectively, that is approximately 11.8% and 5.7% better than in the case of equal device periphery.

The results of linearization for two configurations of three-way Doherty amplifier are summarized in Table I for two-tone test signal and Table II for digitally modulated signal. Tables also content a brief overview of the linearization effects achieved by some other linearization techniques available in the literature.

Table I Results relating to three-way Doherty amplifier for two-tone test signal

Ref.	Output pow. (dBm)/back-off (dB)	PAE (%)	Imp. IM3/IM5	Tone space (MHz)	Lin. Tech.
[4]	45/10.4		9.5/1.5	10MHz	PD
first configuration					
This work	35/3	35.6	19/5	2MHz	*
second configuration (1:1:1) device size ratio					
This work	33/5	28.9	18/7	2MHz	*
This work	33/5	28.9	10/1	10MHz	*
second configuration (1:2.5:2.5) device size ratio					
This work	33/8	30.5	15/5	2MHz	*

Table II Results relating to three-way Doherty amplifier for digitally modulated signals

Ref.	Output pow. (dBm)/back-off (dB)	PAE (DE) (%)	Imp. ACLR (ACPR) (dB)–(offset)	Test signal	Lin. Tech.
[4]	40/15.4	10.4	1-5MHz	WCDMA	PD
[15]	31.5/10	DE 41.6	16-5MHz 19-10MHz)	WCDMA	Dig.PD without memory corr.
[16]	38.5/11.5	53	15-5MHz 5-10MHz	WCDMA	Dig.PD with memory effect comp.
First configuration					
This work	36/5	28	13-0.9MHz 6-2.1MHz	OQPSK	*
second configuration (1:1:1) device size ratio					
This work	36/5	28.9	18-0.9MHz 10-2.1MHz	OQPSK	*
second configuration (1:2.5:2.5) device size ratio					
This work	36/8	30.5	11-0.9MHz 11-2.1MHz	OQPSK	*

*Linearization with the second harmonics and fourth-order nonlinear signals (IM2 and IM4)

It should be pointed out that the crucial matter in the linearization approach used for Doherty amplifier is the possibility to exploit the peaking amplifiers as the sources of signals for linearization and therefore avoid the necessity for additional nonlinear sources that will increase the circuit complexity and total energy consumption**.**

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