Approach for the Design of a Broadband Microwave Power Amplifier in Microstrip Technology for Mobile Communications Systems

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Article Info	ABSTRACT
	This work presents a broadband power amplifier in S-band in microstrip technology. The proposed power amplifier is modeled with a single-stage architecture based on a field effect transistor ATF13786 of Agilent Technologies (hp)®. The used transistor has been biased with transmission lines and it has been matched with networks in the form of λ/4 transformers filters at the input and at the output. This amplifier has been studied and optimized using the Advanced Design System (ADS®) software. The simulation results of the output power and S parameters show excellent characteristics with a satisfactory gain greater than 10.9 dB, low reflections, a saturated output power of 16.4 dBm with a 1 dB compression point at an input power level of about 5 dBm, a maximum PAE of 25.3% and unconditional stability in the desired frequency band. The modeled amplifier can be integrated into mobile communications systems namely LTE mobile networks (2500 to 2690 MHz) and wireless networks using Wi-Fi protocol (2400 to 2485 MHz).

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1.   INTRODUCTION

The design of power amplifiers is today the subject of particular attention in the world of microwaves. On the one hand, the current development of communications systems is such that it requires occupying ever-higher frequency bands. On the other hand, within the sub-systems of transmission-reception, amplifiers occupy a preponderant place because they condition, in general, the quality of transmission of information. The design of a broadband power amplifier requires accurate modeling of the RF active device, effective impedance matching, stability throughout operation and ease of practical implementation [1-3].

The characteristics of a wideband power amplifier are evaluated by its ability to achieve maximum power gain over the required frequency bandwidth, under stable operating conditions with a minimum of amplification stages and the need for high efficiency or linearity can be taken into account if necessary. We note that, for stable operation, it is necessary to evaluate the operating frequency domains in which the active device may be potentially unstable [4, 5].

A single-stage power amplifier can be modeled by the circuit of figure 1, where the matching networks are used on both sides of the transistor to transform the input/output impedance Z_0 in source impedance Z_s and load impedance Z_L [6-8].
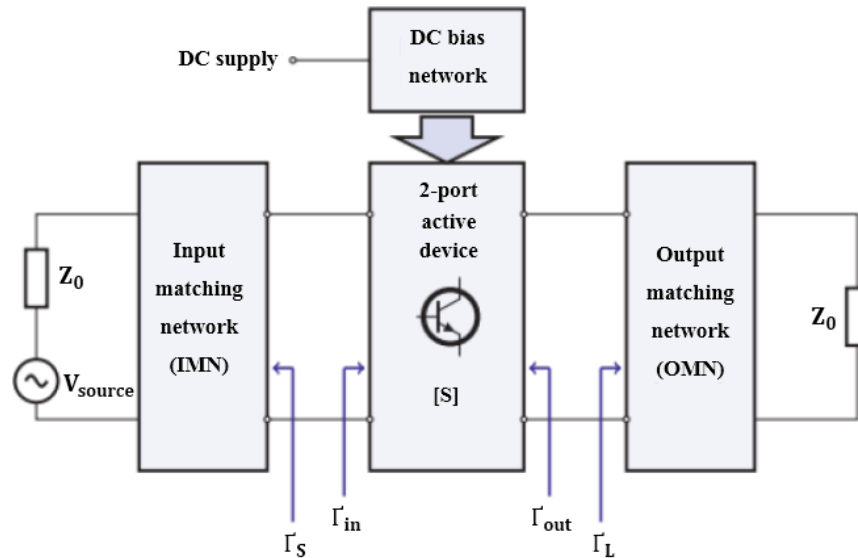
Figure 1. Block diagram of a microwave amplifier [5]

Our work is a contribution to the design of microwave circuits in microstrip technology in order to make amplifiers operating in linear and non-linear modes. We present, in this article, a design methodology, its mathematical formalism as well as the optimization procedure associated with it. Then the study, modeling and characterization of an amplifier for S-band applications are detailed.

The proposed amplifier has high and significant output power, broadband impedance matching, and minimal reflection losses with satisfactory power gain.

2. BIASING NETWORKS

Biasing networks are important elements for the design of amplifiers. They provide the means of efficient excitation for the transistors. The biasing parameters may vary with the technology of the transistors; however, the design principles for the biasing methods are basically the same. The objective of this axis is to include a complete treatment of biasing techniques used for UHF frequencies to millimeter waves [6-10].

2.1 Biasing of transistors

The biasing of transistors comprises two parts: the selection of biasing Q-point for optimum performance of the device in terms of gain, noise figure, output power, PAE and linearity, and the biasing networks. Both are essential to the operation of amplifiers, especially regarding the low levels of noise and high power [5, 6].

Basically, a biasing network is composed of a DC block and RF inductance, The RF inductance aims to have a very high impedance at the operating frequency to prevent RF from leaking through the biasing network. In this case, the effect of the biasing network is negligible when it has a very high impedance for matching networks. Chip capacitors in hybrid circuits and MIM capacitors in MMICs are generally used as DC bias blocks. Ideally, the capacitor should have a short circuit at operating frequencies. The values of the capacities must be selected so that $\omega C = 2\pi f C$ be very large. Therefore, $Z_C = -j(1/\omega C)$ is very small and the signal will pass through the capacitor with little loss or reflection [6].

2.2 Microstrip biasing circuit

The RF choke at microwave frequencies is generally realized using a high-impedance $\lambda/4$ line, also known as a parallel stub terminated by a bypass capacitor C_B, as shown in figure 2a [6]. A DC block may be a capacitor or a back wave coupler 3dB described in reference [6]. For low RF leakage across the biasing network, the impedance ratio of the stub in parallel (Z_S) and the line impedance (Z_0) through, Z_S/Z_0, The low–high impedance lines serve as a lowpass filter, which prevents the microwave signal leaking into the bias port. In this case, the bandwidth increases when the impedance of the stub increases. For VSWR ≤ 1.05, the bandwidth for $Z_S = 100\ \Omega$ is about 12%. In order to further increase the bandwidth, two sections of quarter-wave transmission lines are used. If an open circuit is required across the main line for RF signals, a quarter-wave high-impedance line followed by an open circuited quarter-wave low-impedance line are connected. The configuration is illustrated in figure 2b [6]. Assuming that the characteristic impedance Z_0 of the through line is the same as the impedance of the system, the normalized admittance with the load ($Y_0 = 1/Z_0$) is

$$y = \frac{Y}{Y_0} = 1 + \frac{Z_0}{Z_{in}} \tag{1}$$

With

$$Z_{in} = j Z_1 \frac{Z_1 tan\theta_1 tan\theta_2 - Z_2}{Z_1 tan\theta_2 + Z_2 tan\theta_1} \tag{2}$$

Here, θ_1, Z_1 and θ_2, Z_2 are respectively the electrical length of the line, the characteristic impedance of the first and second line sections. Under matched conditions, $Z_{in} = \infty$ and $y = 1$. The maximum bandwidth is obtained when Z_1 / Z_2 is wide. For example, with $Z_1 = 100\ \Omega$, $Z_2 = 10\ \Omega$, $Z_0 = 50\ \Omega$, and VSWR = 1.2, the maximum bandwidth is about 40%. Generally, the low impedance line section is replaced by a radial line section, as shown in Figure 2c [6, 11]. This arrangement provides better bandwidth than a $\lambda/4$ open circuited line section and is smaller in size. The transmission line medium could be either strip line or microstrip. In this biasing scheme, the biasing circuitry becomes an integral part of the amplifier design [6].
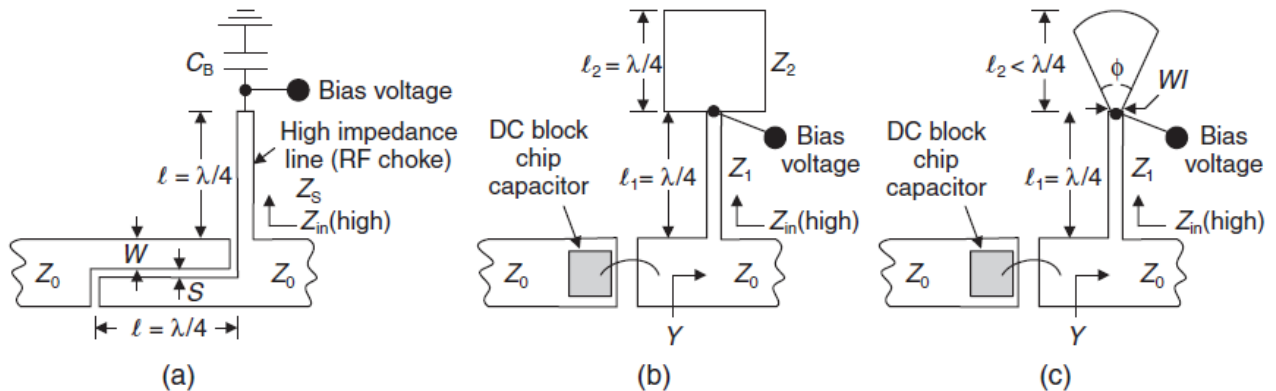

Figure 2. Simplified microwave biasing circuits: (a) coupled-line DC block and $\lambda/4$ transformer, (b) MIM capacitor DC block and two $\lambda/4$ transformers, and (c) MIM capacitor DC block and a $\lambda/4$ transformer in series with a radial line stub [6].

3. IMPEDANCE MATCHING

Impedance matching plays an important role in the design of microwave circuits. The design of a microwave amplifier with inadequate impedance matching will influence the stability of the circuit and reduce its efficiency. In fact, the use of matched devices is related to several objectives. The most important factor is the maximum power transfer.

One way to match one or more circuits is to insert a quarter-wave line between the impedance of the load and the input impedance line and to choose a suitable characteristic impedance. This is one of the simplest techniques to perform impedance matching in a narrow band [9, 11].

Impedance transformers generally include uniform quarter-wave line sections in cascade. Discontinuities result from different impedance jumps, such as a change in the width of the microstrip line. Upon selection of the matching network, certain factors must be taken into account, such as the complexity, the bandwidth and the necessary adjustment according to the desired application.

The relation which links the characteristic impedance Z_0 and the load impedance Z_L to the impedances characterizing the multi-section transformer is calculated from the following formula [12]:

$$\ln \frac{Z_{n+1}}{Z_n} = 2^{-N} C_n^N \ln \frac{Z_L}{Z_0} \tag{3}$$

To achieve matching, we will study the case of three sections of length ($\lambda/4$), as shown in figure 3:

The equivalent characteristic impedances obtained by applying the equation (3) are given by:

$$Z_1 = \sqrt[8]{Z_L Z_0^7} \tag{4}$$

$$Z_2 = \sqrt{Z_0 Z_L} \tag{5}$$

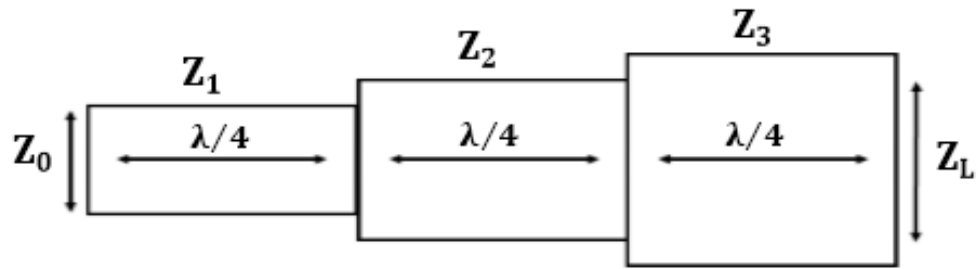$$Z_3 = \sqrt[8]{Z_0 Z_L^7} \tag{6}$$

Figure 3. A three sections quarter-wave transformer [12]

4. STABILITY

A power amplifier must not oscillate whatever the impedance presented to it, both at the input and at the output. The stability criteria make it possible to rule on the risks of instability of the power amplifier.
So the combination of the condition of Rollet [15-21]:

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\triangle|^2}{2|S_{12}S_{21}|} > 1 \tag{7}$$

With one of the following conditions:

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\triangle|^2 > 0 \tag{8}$$

$$B_2 = 1 - |S_{11}|^2 + |S_{22}|^2 - |\triangle|^2 > 0 \tag{9}$$

$$|\triangle| = |S_{11}S_{22} - S_{12}S_{21}| < 1 \tag{10}$$

$$1 - |S_{11}|^2 > |S_{12}S_{21}| \tag{11}$$

$$1 - |S_{22}|^2 > |S_{12}S_{21}| \tag{12}$$

provide unconditional stability over the entire of the considered frequency band [15].
The S_{ij} parameters are the scattering parameters of the transistor or the amplifier.

A new and unique stability criterion has been developed recently [17, 21]. The condition:

$$\mu_1 = \frac{1 - |S_{11}|^2}{|S_{22} - S_{11}^* \triangle| + |S_{12}S_{21}|} > 1 \tag{13}$$

Or :

$$\mu_2 = \frac{1 - |S_{22}|^2}{|S_{11} - S_{22}^* \triangle| + |S_{12}S_{21}|} > 1 \tag{14}$$

alone ensures unconditional stability.

5. PROPOSED BROADBAND POWER AMPLIFIER

Using the concepts mentioned above, the proposed amplifier is shown in Figure 4. For the design of our broadband power amplifier we used the commercialized transistor ATF13786 of Agilent Technologies (hp)®, which is a field effect transistor.

In the design process, the biasing condition (known as DC Q-point) for the active device must be carefully selected because it determines the basic performance in terms of stability, output power, gain and PAE of the entire power amplifier. The biasing network used is shown in figure 4. The proposed power amplifier is biased with two voltage sources V_{DS} of + 2,8 V and V_{GS} of -0,5 V. As necessary for bandwidth, the RF choke is realised by using a high impedance transmission line of length $\lambda/4$ followed by a radial stub [6, 11]. This combination helps prevent RF energy from entering the power supply and increases bandwidth.
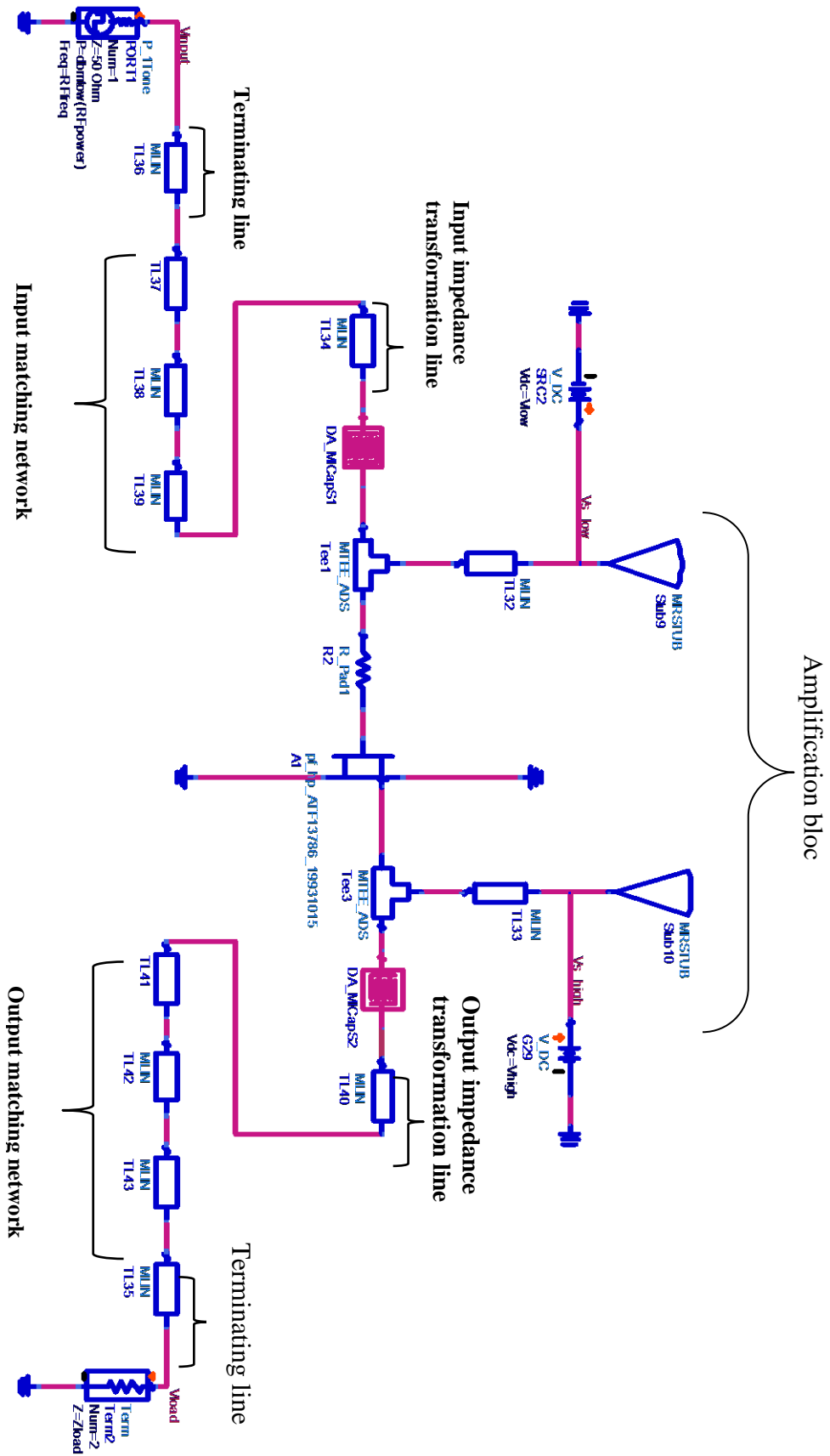
Figure 4. Schematic of the proposed single-stage broadband power amplifier

The DC Block was realized with a Microstrip Interdigital Capacitor (DA_MICapS1 and DA_MICapS2) in order to avoid the use of the localized elements to facilitate the fabrication of the circuit and to reduce the noise and the energy losses.

As shown in figures 4 and 5, the input and output matching networks were designed as simple as possible to reduce both the consumption of the layout and loss of gain. The design of the two matching networks depends on the frequency band in which this broadband power amplifier is to be designed.
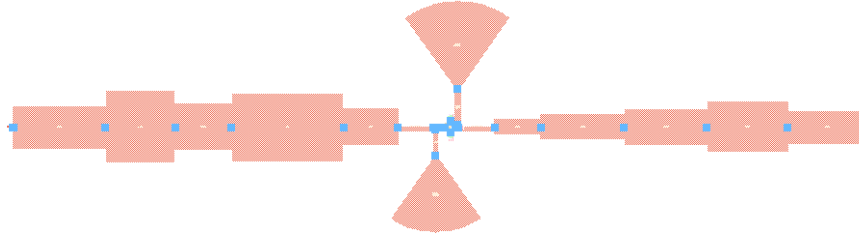

Figure 5. Layout of the proposed single-stage broadband power amplifier

In this work, the proposed amplifier is implemented on a substrate of type FR4 epoxy, with a permittivity (ε_r) equal to 4.32 and a thickness equal to h = 1.6 mm.

6. RESULTS AND DISCUSSION

The complete amplifier circuit was simulated using the ADS software of Keysight®. Figures 6 to 11 present the results of these simulations which will constitute the reference data for the experimental characterization.

Figure 6 shows us; on the frequency band [2-3] GHz, the losses by reflection in input (S_{11}) and in output (S_{22}) for low levels of excitation. From figure 6, we can observe that the input reflection losses are less than -10.647 dB and can reach the value -36.523 dB at 2.88 GHz. On the other hand, at the output, these losses are less than - 11.455 dB and can reach the value -44.869 dB at 2.73 GHz.
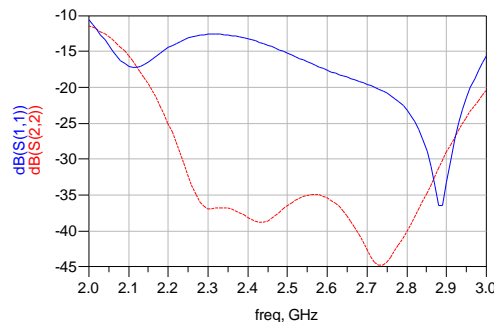

Figure 6. Reflexion coefficients S_{11} and S_{22}

The simulated gain S_{21} for the proposed BPA is given in figure 7. As can be seen, the small signal gain varies between a minimum value of 10.944 dB at 3 GHz and a maximum value of 13.979 dB at 2.07 GHz. Inverse isolation S_{12} represents the internal return of the output to the input of a two-port device. For simulated inverse isolation of the proposed BPA. As can be seen in figure 7, S_{12} varies between a minimum value of -25.767 dB at 2 GHz and a maximum value of -23.889 dB at 2.95 GHz.

In terms of practical operation of a power amplifier, it is necessary that S_{12} be as small as possible because the lower reverse isolation value is the higher degree of isolation between the output and the input, as well as the higher degree of stability of a given stage.
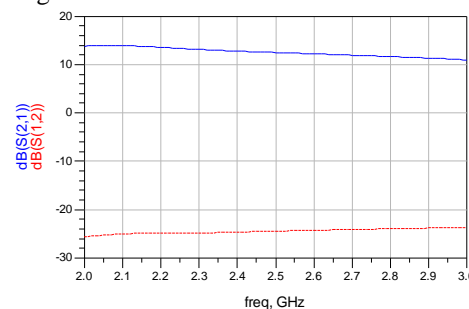

Figure 7. Transmission coefficients S_{21} and S_{12}

Figure 8 shows the values of the stability coefficients k (StabFact1) and μ (Muprime1 and Mu1) simulated on the band [2-3] GHz. The stability factors are greater than 1 over the entire operating band, the unconditional stability conditions of the proposed BPA are met. This means that any source or load can be connected to the input or the output of BPA without risk of becoming unstable or producing oscillations.
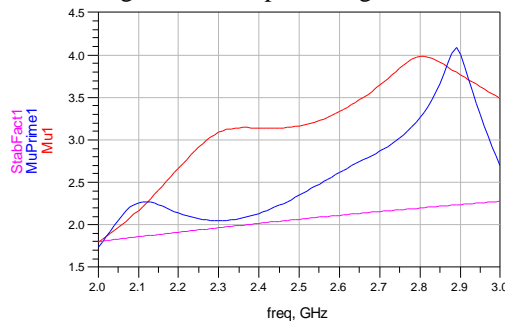

Figure 8. Stability factors

The 1 dB compression point is defined as the point at which the output power level corresponds to a 1 dB difference from the ideal linear behavior. The corresponding input power level is used to mark the boundary between linear and nonlinear regions.

Figures 9 and 10 show the simulated output power and the PAE of the proposed BPA. The curves were plotted at the central operating frequency equal to 2.5 GHz. For each of them, the results are presented according to the input power P_e of the circuit.

Figure 9 shows that the simulated output power reaches a maximum of 16.403 dBm for 20 dBm at the input power, resulting in an output power of 43.682 mW. The proposed BPA achieves a 1 dB compression at an input power level of about 5 dBm. As shown in figure 10, at 2.5 GHz, the simulated PAE reaches a maximum value of 25.317% for an input power of 6 dBm.
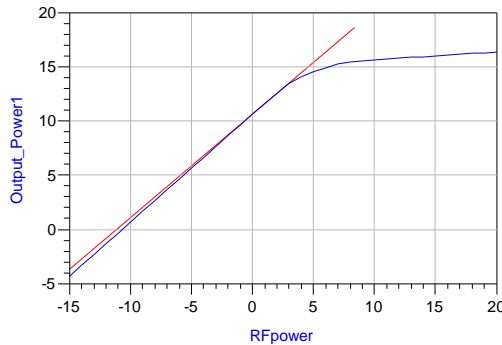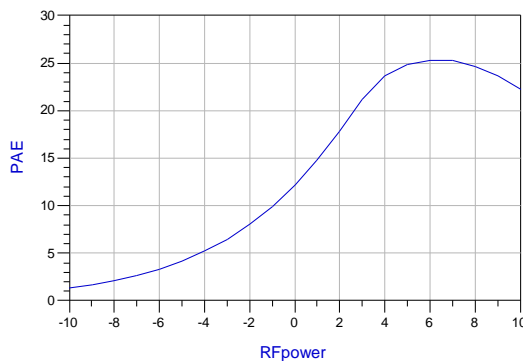

Figure 9. Output power versus the input power

Figure 10. Added power efficiency versus the input power

Figure 11 shows the curve of the noise figure as a function of frequency. The simulated noise figure F_dB varies between 4.386 and 6.168 dB over the frequency band [2-3] GHz. The lowest NFmin value is 1.847 dB obtained at 2 GHz, as shown in figure 11.
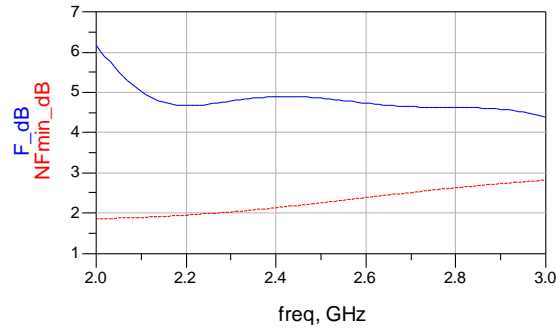
Figure 11. Noise figure

7. COMPARISON TABLE

In the table 1, the performance of the designed broadband amplifier is summarized and compared to the state of art of broadband power amplifiers operating in the S band.

Comparing current work with state-of-art of the BPA [22-24], we can clearly see that the proposed broadband matching techniques are useful, and the designed BPA achieves excellent bandwidth with good matching on all frequency bands of operation. The simulated reflection loss shows good results with a bandwidth of 1 GHz compared to articles [22-24].

In the frequency range from 2 GHz to 3 GHz, the proposed BPA achieves a maximum gain of 13.979 dB, and a P_{sat} of 16.403 dBm with a PAE_{max} of 25.317% and a P_{1dB} of approximately 5 dBm under power supply voltages of + 2.8V for V_{ds} and -0.5 V for V_{gs}.

Given broadband input power and broadband frequency coverage, the proposed BPA offers better broadband performance than other BPAs [22-24], and has high gain, excellent bandwidth, good output power and simple circuits. It is a good candidate for wideband PAs operating in the S band.

Table 1. State of art of broadband power amplifiers

Ref.	Freq. (GHz)	Gain (dB)	S(1,1) (dB)	S(2,2) (dB)	S(1,2) (dB)	Psat (dBm)	PAE (%)	NF (dB)	VDS (V)
[This work]	[2-3]	13,9	−10.6 / −36,5	-11,4 / -44,8	-23,8 / -25,7	16,4	25,3	< 6, 1	2.8
[22]	[1,9-2,7]	11	-	-	-	28.1	13.7	-	2.5
[23]	[1,8-2,8]	25	−12 / −25	−9 / −19	-	28	6.1	-	5
[24]	[1,75-2,15]	10 - 11,7	−11 / −22	−13 / −19	−18 / −19	8	-	-	12

8. CONCLUSION

In conclusion, the design and analysis of a microwave power amplifier in microstrip technology was presented. The simulation results show that the proposed amplifier offers excellent performance for the applications of LTE mobile networks and wireless networks using Wi-Fi protocol. The proposed design has a power gain of 13.979 dB with reverse isolation below -23.8 dB under a power supply of 2.8V for V_{ds} and   -0.5V for V_{gs}. The proposed power amplifier has successfully achieved good performance in terms of stability and impedance matching. This amplifier is capable of delivering an output power of up to 16.403 dBm.

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