A Broadband Power Amplifier for 3GHz to 6GHz Range in 130nm CMOS Technology

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***Abstract*—This paper presents the design and analysis of a broadband power amplifier operating in the frequency range from 3 GHz to 6 GHz using 130nm CMOS technology. The amplifier is coupled with a bandpass filter, employing Chebyshev and Bode-Fano design methodologies. The impact of transistor oxide thickness on amplifier performance is investigated, and simulations are carried out to optimize the design for maximum power output and linearity. The proposed amplifier demonstrates promising results, making it suitable for various wireless communication applications.**

***Index Terms*—Broadband, power amplifier, bandpass filter, Chebyshev, Bode-Fano, CMOS technology.**

I. Introduction

The demand for broadband wireless communication systems has fueled the need for high-performance power amplifiers operating in a wide frequency range. In this context, the design of a broadband power amplifier becomes crucial for achieving efficient and reliable communication. This paper focuses on the development of a broadband power amplifier targeting the frequency band from 3 GHz to 6 GHz.

The design of the proposed amplifier involves the integration of a bandpass filter to mitigate unwanted harmonics and improve overall linearity. Two different filter designs, based on Chebyshev and Bode-Fano methodologies, are analyzed to compare their impact on amplifier performance. Additionally, the influence of transistor oxide thickness on the amplifier's characteristics is investigated.

The remainder of this paper is organized as follows: Section II presents the design of the matching network, including the use of Chebyshev and Bode-Fano filters. Section III discusses the design of the broadband power amplifier, considering the impact of transistor oxide thickness and provides simulation results and analysis, and Section IV concludes the paper with insights into potential future work.

II. The Design of Matching Network

Bode and Fano discovered that regardless of the matching conditions used, the matching network and RC load system must satisfy the inequality

$$\int_0^{+\infty} \ln\left(\frac{1}{|\Gamma|}\right) dw \leq \frac{\pi}{RC}, \tag{1}$$

where Γ is the reflection coefficient, and RC depends on technology. Equation (1) shows that an increase in bandwidth leads to an increase in the maximum reflection coefficient in the bandpass.

The Chebyshev filter provides the most efficient system in terms of Equation 1. Therefore, Chebyshev filter design will be used in the design of matching networks.

Two matching networks are analyzed, as shown in Fig. 1. The first network has Chebyshev Type I filter characteristics and is driven by a generator with finite internal impedance. The second network is designed using the Fano method and is driven by a current generator with infinite internal impedance.

A. Prototype Chebyshev Filter

The Chebyshev attenuation function $A(w) = |H(w)|^{-2}$ is given by

$$A(w) = 1 + \epsilon^2 T_n^2(w), \tag{2}$$

where

$$T_n(w) = \begin{cases} \cosh\left(n \cosh^{-1} w\right) & |w| \geq 1, \\ \cos\left(n \cos^{-1} w\right) & |w| < 1, \end{cases} \tag{3}$$

and n is the order of filter.

If we define the ripple coefficient γ as

$$\gamma = \sinh\left(\frac{1}{n}\sinh^{-1}\epsilon^{-1}\right), \tag{4}$$

filter elements g_i, for $r = 1, 2, ..., n-1$, can be expressed as

$$g_r g_{r+1} = \frac{4 \sin\frac{(2r+1)\pi}{2n} \sin\frac{(2e-1)\pi}{2n}}{\gamma^2 + \sin^2\frac{r\pi}{n}} \tag{5a}$$

$$g_n = \frac{2\sin\frac{\pi}{2n}}{\gamma} \tag{5b}$$

By substituting $n = 3$ into equations (5a) and (5b) we get

$$g_1 = \frac{1}{\gamma} \tag{6a}$$

$$g_2 = \frac{\gamma}{\gamma^2 + \frac{3}{4}} \tag{6b}$$

$$g_3 = \frac{1}{\gamma} \tag{6c}$$

Since the minimum and maximum attenuations in the passband are $A_{min} = 0$ and $A_{max} = 1 + \epsilon^2$, ripple in passband is equal to

$$\delta = 10\log_{10}\left(1 + \epsilon^2\right), \tag{7}$$

where ϵ can be expressed by finding the inverse function of (4) as

$$\epsilon = \sinh^{-1}\left(n \sinh^{-1}\gamma\right). \tag{8}$$

For a fixed $\gamma = 1$ and $n = 3$, ripple in the passband is $\delta = 0.0877\,\text{dB}$.

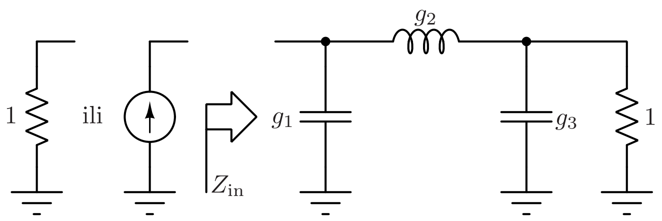

Fig. 1. Prototype matching networks. One is driven by a generator with finite, and the other with infinite internal impedance.

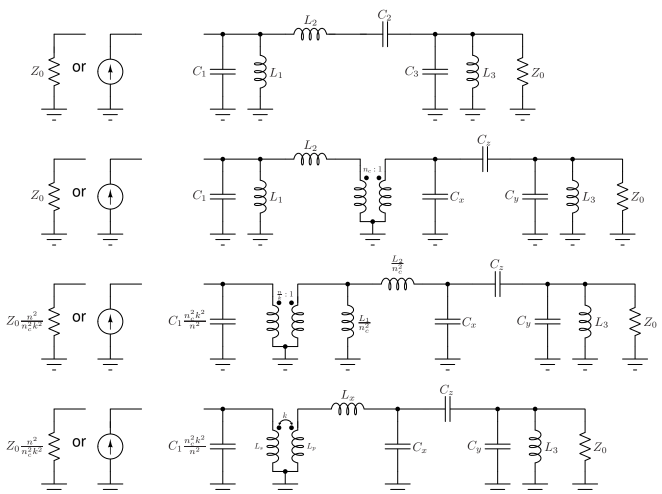

Fig. 2. a) Denormalized bandpass filter. b) Norton transformation of capacitors C2 and C3 . c) Moving an ideal transformer in front of inductor L1 . d) Changing an ideal with a real transformer.

B. Prototype Fano Matching Network

In the case of the Fano design method of a network driven by a generator with an infinite internal impedance, equations (5a) and (5b) become

$$g_r g_{r+1} = \frac{4\sin\frac{(2r+1)\pi}{2n}\sin\frac{(2e-1)\pi}{2n}}{2x^2(1-\cos\frac{r\pi}{n})+\sin^2\frac{r\pi}{n}} \tag{9a}$$

$$g_n = \frac{\sin\frac{\pi}{2n}}{x}, \tag{9b}$$

for $r = 1, 2, ..., n-1$. By substituting $n = 3$ into equations (9a) and (9b) we get

$$g_1 = \frac{12x^2+3}{8x^2+6x} \tag{10a}$$

$$g_2 = \frac{16x}{3\cdot(4x^2+1)} \tag{10b}$$

$$g_3 = \frac{1}{2x} \tag{10c}$$

Ripple in the passband can be found to be

$$\delta = 10\log_{10}\left(\coth^2\left(n\sinh^{-1}x\right)\right). \tag{11}$$

For a fixed $x = 1$ and $n = 3$, ripple in the passband is $\delta = 0.877\,\mathrm{dB}$.

C. Denormalized Filter

In order to get a bandpass filter out of the prototype, which is a lowpass filter normalized to $Z_0 = 1$ and cutoff frequency $w_0 = 1\,\mathrm{rad}$, we need to introduce a substitution

$$s \rightarrow \frac{s^2+w_0^2}{s\Delta w_0}, \tag{12}$$

where w_0 is the central frequency, and Δ is the relative bandwidth.

The substitution (12) suggests that capacitors and inductors from Fig. 1 transform into parallel and series oscillators, respectively, as shown in Fig. 2.a.

Now we need to denormalize for input/output impedance Z_0. This is done by multiplying all impedances by Z_0 , which means multiplying inductances by Z_0 and capacitances by $\frac{1}{Z_0}$.

Values of capacitors and inductors are

$$C_1 = \frac{g_1}{\Delta w_0 Z_0}, \qquad L_1 = \frac{\Delta Z_0}{w_0 g_1} \tag{13a}$$

$$C_2 = \frac{\Delta}{g_2 w_0 Z_0}, \qquad L_2 = \frac{g_2 Z_0}{w_0 \Delta} \tag{13b}$$

$$C_3 = \frac{g_3}{\Delta w_0 Z_0}, \qquad L_2 = \frac{\Delta Z_0}{w_0 g_3} \tag{13c}$$

In order to scale input impedance by a factor of n_c^{-1} we do Norton transformation under capacitors C_2 and C_3 , as shown in Fig. 2.a. Transformation leads to Fig. 2.b, where transformed capacitances are

$$C_x = n_c(n_c-1)C_2 \tag{14a}$$

$$C_y = (1-n_c)C_2 + C_3 \tag{14b}$$

$$C_z = n_c C_2 \tag{14c}$$

For the transformation to be valid, or equivalently, for circuits from Fig. 2.a and 2.b to be considered equivalent, the following condition must hold true

$$n_c \leq 1 + \frac{C_3}{C_2}. \tag{15}$$

Next step is removal of ideal transformator towards the input. By doing that impidances and inductancces must be corrected by a factor of n_c^{-2} and capacitances by n_c^2, as showrn in Fig. 2.c.

Having a transformator is a requierment. To achive this, another ideal transformator is placed between first capacitor and first inductor, Fig. 2.c. By doing so, impedances to the left of an ideal transforamtor must be corrected by factor of n_c^2.

The mathematical model of a real transformer comprises an ideal transformer, leakage and magnetic inductances. It is evident that $\frac{L_1}{n_c^2}$ and $\frac{L_2}{n_c^2}$ conform to this structure. Consequently, we obtain a real transformer, Fig. 2.d, with

$$L_p = \frac{L_1}{k^2 n_c^2} \tag{16a}$$

$$L_s = L_p n^2 \tag{16b}$$

$$L_x = \frac{k^2(L_1+L_2)-L_1}{k^2 n_c^2} \tag{16c}$$

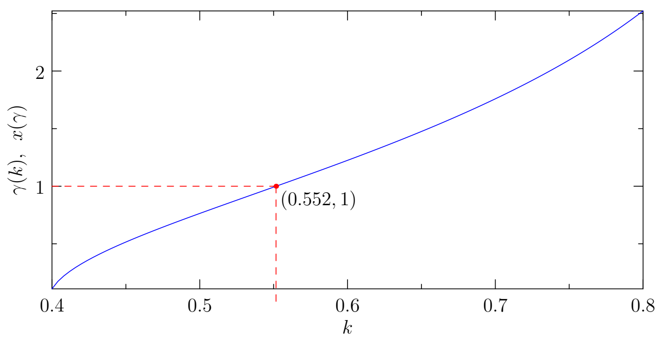

Fig. 3. Functions $\gamma(k)$ and $x(k)$ for $\Delta = \frac{\sqrt{2}}{2}$.

where L_p is inductance at primar, L_s at secundar and L_x is leakage inductance.

If we substitute equations (13) into (14) and (16) we will get the complit set of equations for elements of a filter from Fig. 2.d

$$Z_in = \frac{Z_0 n^2}{k^2 n_c^2}, \tag{17a}$$

$$C_1 = \frac{g_1 k^2 n_c^2}{Z_0 \Delta w_0 n^2}, \tag{17b}$$

$$L_p = \frac{Z_0 \Delta}{w_0 g_1 k^2 n_c^2}, \tag{17c}$$

$$L_s = \frac{Z_0 \Delta n^2}{w_0 g_1 k^2 n_c^2}, \tag{17d}$$

$$L_x = \frac{Z_0(\Delta^2(k^2-1) + g_1 g_2 k^2)}{\Delta w_0 g_1 k^2 n_c^2}, \tag{17e}$$

$$C_x = \frac{\Delta n_c (n_c - 1)}{Z_0 w_0 g_2}, \tag{17f}$$

$$C_y = \frac{g_2 g_3 - \Delta^2(n_c^2 - 1)}{Z_0 \Delta w_0 g_2}, \tag{17g}$$

$$C_z = \frac{\Delta n_c}{Z_0 w_0 g_2}, \tag{17h}$$

$$L_3 = \frac{Z_0 \Delta}{w_0 g_3}, \tag{17i}$$

(17j)

where

$$n = \sqrt{\frac{L_s}{L_p}}. \tag{18}$$

From (17e) it can be seen that there is a condition for k that satisfy $L_x = 0$. That condition is

$$k = \sqrt{\frac{\Delta^2}{\Delta^2 + g_1 g_2}}, \quad L_x = 0 \tag{19}$$

If we enforce $\gamma \geq 1$ for a filter exhibiting Chebyshev characteristics and $x \geq 1$ for a filter with current drive, and choose $\Delta = \frac{\sqrt{2}}{2}$, the corresponding minimum values for k are determined as follows

$$k_{min,\gamma} = k_{min,x} = 0.5517. \tag{20}$$

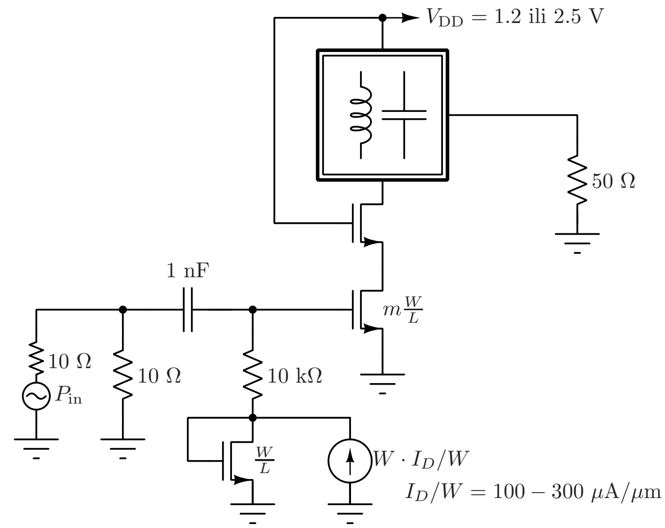

Fig. 4. Block diagram of power amplifier.

Inverse functions of (19) in termas of γ and x when condition $L_x = 0$ is met are

$$\gamma(k, \Delta) = x(k, \Delta) = \frac{1}{2\Delta}\sqrt{\frac{8k^2 - 3\Delta^2(1-k^2)}{1-k^2}}. \tag{21}$$

Fig. 3 shows $\gamma(k)$ and $x(k)$ for $\Delta = \frac{sqrt2}{2}$. Minimum k for which γ, or x is greater then 0, for same Δ is

$$k_{min}|_{\gamma=0,x=0} = 0.3973. \tag{22}$$

III. The Design of Broadband Power Amplifier

Power amplifier and bandpass filter are needed in order to to design a boradband power amplifier. Bandpass filters were designed in previous chapter. Their goal is to compensate parasitic capacitors of output stage of amplifier in frequency range of interest, in this case from 3 GHz to 6 GHz.

In order to get maximum power amplification (for unit transistor amplifier) we have to match output impedance of 50 Ω to optimal load impedance. We could also increase width of transistors to get higher power, but it has it's limits.

A. Simulations and Results

Simulations were conducted for two configurations of transistors in a cascode arrangement, Fig. 4. In the first configuration, both transistors have a similar oxide thickness, approximately $t_{ox} = 2.6$ nm. In this scenario, the width of the unit transistor is 20 μm, and the current density is 192.23 μA/μm.

In the second configuration, the transistor at the output has a larger oxide thickness, approximately $t_{ox} = 7$ nm. In this case, the widths and current densities of the unit transistors are set to 20 μm and 252 μA/μm for $t_{ox} = 2.6$ nm, and 40 μm and 126 μA/μm for $t_{ox} = 7$ nm.

Simulations for finding parasitic capacitances at the output node, as well as for finding optimal capacitances are done with 20 times the unit transistors. This implies that for the unit transistors, parasitic capacitances should be multiplied by 1/20, optimal impedances and output power by 20.

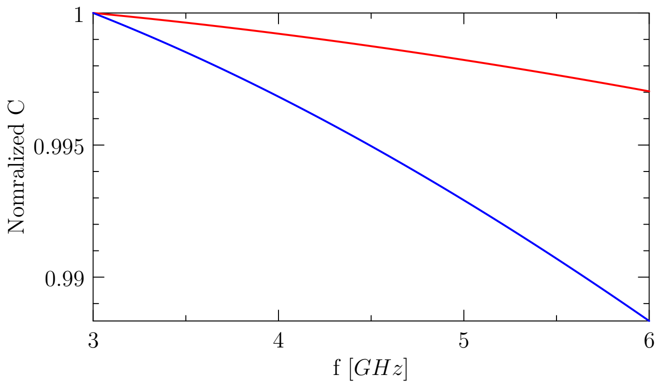

Fig. 5. Relationship between output parasitic capacitance and frequency. The red curve represents the scenario where both transistors possess a similar oxide thickness, while the blue curve illustrates the case where the output transistor has a larger oxide thickness. Simultaion is done for 20 times the unit transistors.

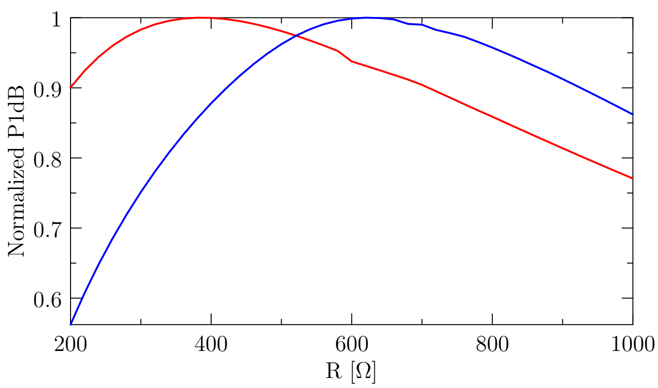

Fig. 6. Relationship between P1dB and load impedance. The red curve represents the scenario where both transistors possess a similar oxide thickness, while the blue curve illustrates the case where the output transistor has a larger oxide thickness. Simultaion is done for 20 times the unit transistors.

Parasitic capacitance at the output of amplifier is found as

$$C_D = \frac{\Im\{Y_{\text{out}}\}}{2\pi f}, \tag{23}$$

where Y_{out} is is admitance seen at the output of the amplifier. Fig. 5 shows normalized parasitic capacitance dpeandens on frequency, for both types of amplifier. We can see that the variation in parasitic capacitance remains quite minimal with respect to changes in frequency. Therefore, the selection of capacitance for the central frequency, $f = 3\sqrt{2}\,\text{GHz}$, is deemed appropriate. Their values are

$$C_D = 17.97\,\text{fF} \tag{24}$$

for the first type of amplifier, featuring transistors with similar oxide thicknesses, and

$$C_D = 23.13\,\text{fF} \tag{25}$$

for the second type of amplifier, characterized by transistors with different oxide thicknesses.

Optimal load impedences for which power gain is maximum can be found by simulating P1dB commpenstaion point for different valuse of resistors. Fig. 6 shows P1dB dpeandens on load impedance for both types of amplifiers. The values of optimal impedances (for unit transistors) are

$$R_{opt} = 380\,\Omega \tag{26}$$

for the first type of amplifier, with transistors of similar oxide thicknesses, and

$$R_{opt} = 620\,\Omega \tag{27}$$

for the second type of amplifier, featuring transistors with different oxide thicknesses.

To maximize the output power, it is essential to utilize a transistor width multiplication factor that results in the minimum optimal impedance. This is because, in such a scenario, the multiplication factor attains its maximum value, and the output power is directly proportional to this factor. As depicted in Figure 3, the ripple is inversely proportional to k. Equation (17a) further illustrates that Z_{in} is inversely proportional to k. Consequently, for the minimum optimal impedance, the ripple reaches its maximum. Therefore, to achieve higher power, it is necessary to deliberately increase the ripple.

For $n_c = n_{c,\text{max}}$ and $L_x = 0$ and $n = 1$ and $k = k_{\text{min}} = 0.5512$, we obtain $Z_{\text{in, min}} = 15.21\,\Omega$ with a multiplication factor $M = 25$ (for transistors with the same oxide thickness) and $M = 41$ (for transistors with different oxide thickness) for a filter with Chebyshev characteristics. Additionally, for Bode-Fano filter design, we have $Z_{\text{in, min}} = 38.46\,\Omega$ and $M = 10$ (for transistors with the same oxide thickness) and $M = 16$ (for transistors with different oxide thickness).

Table I presents the values of all elements for filter with Chebyshev characteristics and filter designed with Bode-Fano method.

TABLE II
VALUES OF CAPACITOR C_1

	Same oxide thickness	Different oxide thickness
Chebyshev	3.04 pF	2.54 pF
Bode-Fano	1.30 pF	1.11 pF

TABLE III
MAXIMUM VALUE OF COMPRESSION POINT

	Same oxide thickness	Different oxide thickness
Chebyshev	11.93 dBm	21.19 dBm
Bode-Fano	7.26 dBm	16.55 dBm

TABLE I
VALUES OF FILTER ELEMENTS

	Z_{in} [Ω]	C_1 [pF]	L_s [nF]	L_p [nF]	C_x [pF]	C_z [pF]	L_3 [nF]
Chebyshev	15.21	3.49	0.403	0.403	3.49	1.52	1.33
Bode-Fano	38.46	1.48	0.952	0.952	1.09	1.03	2.65

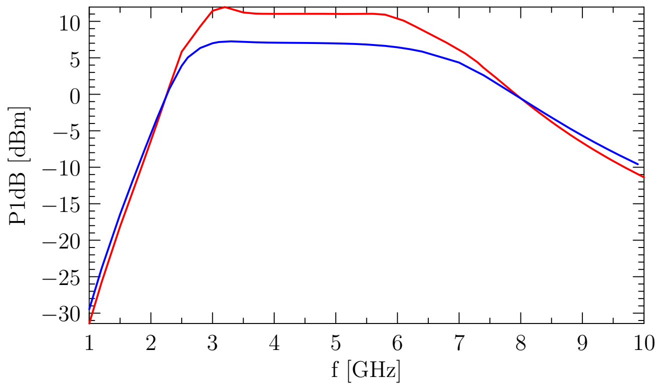

Fig. 7. The P1dB values are observed within the frequency range from 1 GHz to 10 GHz. The blue line corresponds to a filter designed using the Bode-Fano method, while the red line represents a filter exhibiting Chebyshev characteristics.

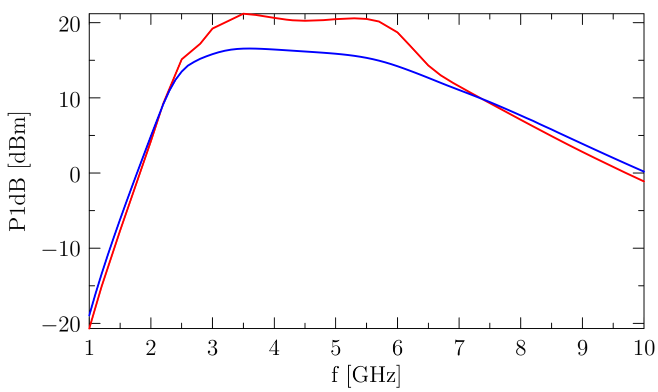

Fig. 8. P1dB values are measured across the frequency range of 1 GHz to 10 GHz for amplifiers with transistors having the different oxide thickness. The blue line represents a filter designed using the Bode-Fano method, while the red line corresponds to a filter exhibiting Chebyshev characteristics.

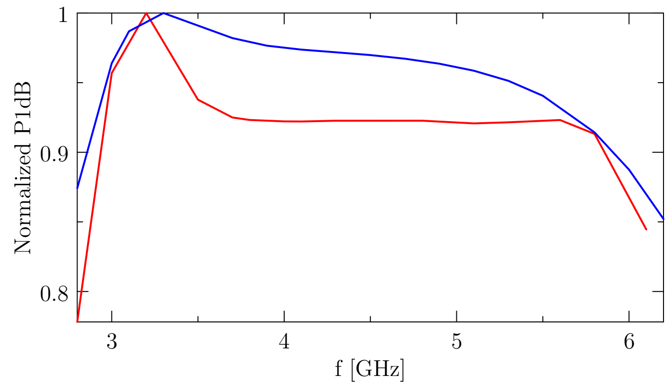

Fig. 9. The P1dB values are observed within the frequency range from 1 GHz to 10 GHz. The blue line corresponds to a filter designed using the Bode-Fano method, while the red line represents a filter exhibiting Chebyshev characteristics.

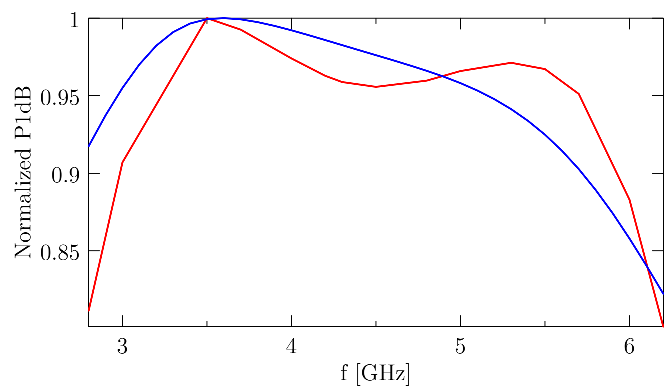

Fig. 10. P1dB values are measured across the frequency range of 1 GHz to 10 GHz for amplifiers with transistors having the different oxide thickness. The blue line represents a filter designed using the Bode-Fano method, while the red line corresponds to a filter exhibiting Chebyshev characteristics.

The only distinguishing factor between amplifiers with transistors having the same and different oxide thickness is the value of the capacitor C_1. Due to the presence of parasitic capacitance at the output node, it is possible to integrate that capacitance into the capacitor C_1 of the filter. Table II presents the values of C_1 corresponding to each combination of filter and amplifier types.

Table III displays the maximum P1dB values within the passband for each combination of filter and amplifier types. These values were measured for $k = k_{\min} = 0.5512$.

Figures 7 and 8 illustrate the P1dB values measured across the frequency range from 1 GHz to 10 GHz. It is evident that filters with Chebyshev characteristics (depicted by red lines) provide higher power in both cases. Furthermore, Figures 9 and 10 showcase the ripple in the passband. It is clear that the filter designed using the Bode-Fano method (depicted by blue lines) exhibits a smoother response in the passband for both cases.

IV. Conclusion

This work presents a comprehensive study on the design of a broadband power amplifier in the 3GHz to 6GHz range, utilizing Chebyshev and Bode-Fano matching networks. Through analytical analysis and optimization, prototype filters were designed, considering ideal and real transformers for practical applications.

Simulations for various transistor configurations show competitive results in terms of P1dB compression point and power gain. The comparison between Chebyshev and Bode-Fano filters reveals the latter's advantages in passband ripple and frequency response smoothness.

In summary, this research contributes valuable insights into the design of broadband power amplifiers. The presented methodology provides a foundation for further exploration, aiding in the development of high-performance communication systems. Future work may involve additional optimizations and practical implementations to enhance applicability in specific scenarios.