Miniaturisation of a 2-Bits Reflection Phase Shifter for Phased Array Antenna based on Experimental Realisation

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***Abstract*—In this paper, a controllable reflection type Phase Shifter (PS) is designed, simulated and implemented. The structure of the 2-bits PS consists of branch line coupler, delay lines and six GaAs FET switches controlled in pair. The phase shifting is achieved by turning ON one pair of switches. The circuit design is fabricated using FR4 substrate with dielectric constant equal to 4.7. The size of the realised circuit is 7cm×2.8cm. To reduce this size, two methods are used. First, shortened quarter-wave length transmission line in T model is employed to develop a compact branch-line coupler. Second, a loaded line with capacitor is used to reduce the dimension of delays lines. The two methods are combined to realise a PS with compact size equal to 4.5cm×1.96cm.**

Keywords—Reflection type PS; FET switch; Branch line coupler; Semiconductors technology

I. Introduction

The PS is the key component in phased array antennas used for electronic beam steering. Using digital PS based on semiconductors technology, we can realise an accurate scanning of beam former and a good compatibility with the computer control.

Four classical design topologies are developed to realise digital PS, they are: the switched line, the loaded line, the switched low-pass / high-pass and the reflection theories, each of these methods has its own limitation [1-2]. The topology reflection achieves a low insertion loss and a low phase error, but it presents a poor match over a large bandwidth

Several PSs operating in the L and S band frequency are developed and discussed in the literature. In [3], a reflection type PS characterized by an ultra-band is developed. The structure of the PS is composed of 3 dB hybrid coupler and a pair of novel reflective terminating circuit. The 180° and 90° MMIC PS have demonstrated a phase of 187±7° over 0.5-20 GHz and a phase of 93°±7° over 7-12 GHz.

Another reflection type PSs are presented in [4]. The PSs are implemented at 2.45 GHz in a 0.18 µm CMOS technology. So, an impedance transformed π resonated varactor network is employed to provide 360° phase rang. The measured results of the two PSs show a phase shift range of 120° with insertion losses 5.6 ± 1.6 dB and a phase range larger than 340° with the insertion losses of 10.6±2 dB over the band 2.44-2.55 GHz.

In [5], another reflection type PS is developed, achieving a phase shift over 400° between 1.95 and 2.15 GHz. The circuit is composed of 3 dB hybrid coupler and reflection loads. Measurement results show insertion loss less than 4 dB for 400° phase shift.

Based on the previously mentioned proposals, we suggest in this work a design and the according implementation of the 2-bits reflection type PS operating at the frequency 2.4 GHz for phased array antennas. The proposed structure provides four different phase shifts. Based on the experimental realisation, we show the major drawbacks of our structure which are mainly related to its big size. Therefore, in a second part of this work, we propose a miniaturised version of the PS and address the corresponding simulation results.

This paper is composed of four sections: Section 2 demonstrates our proposed of the 2-bits PS design, simulation, and corresponding implementation results. Section 3 demonstrates our optimisation in size version of the PS and illustrates the corresponding simulation results with ADS. Section 4 describes 2-bits miniaturised PS design and simulation. This paper is enclosed by Section 5 that is the conclusion and the perspectives.

II. 2-Bits Phase Shifter Design and Experimental Realisation

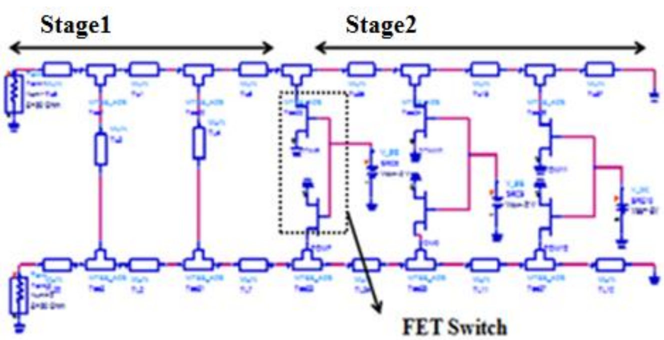
A. Structure design

In Figure 1, a model of a 2-bits reflection type PS is presented. Figure 1 (a) presents the model designed by ADS and Figure 1 (b) illustrates the layout.

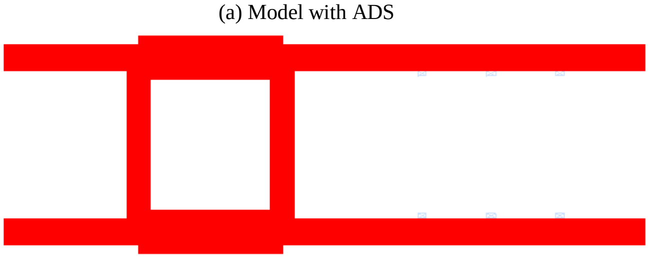
The structure is composed of 3 dB hybrid coupler (stage 1 of Figure 1 (a)), six FET switch and delay lines (stage 2 of Figure 1 (a)). The electric length, denoted θ, is determined by the following:

$$\theta = \beta l \qquad (1)$$

Indeed, to realise a 90° difference phase and to validate the reflection propriety, the length of delay lines is determined for $\theta = 45°$. Also, three delay lines are cascaded in series and connected to direct and coupled ports of the branch line coupler. FET switches is controlled in pair in the gate port, the bias is ±1.6V.



(a) Model with ADS



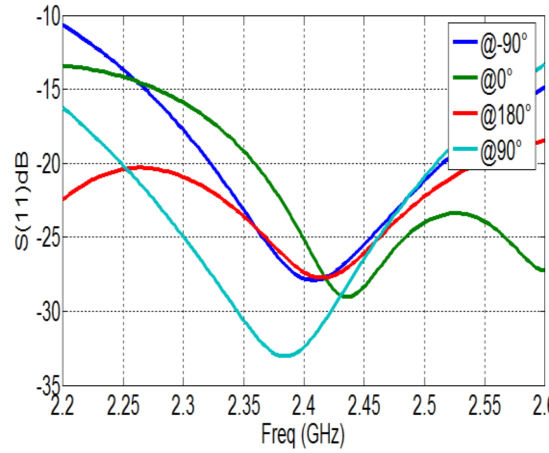
(b) Layout of the PS

Fig. 1. Design of the 2-bit PS

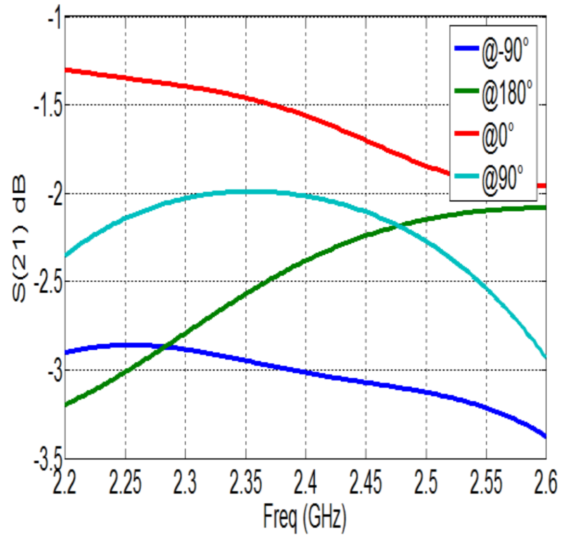
The input signal is first split into two parts having the same amplitude but they are 90 ° phase shifted. According to the states of the switches, the signals are propagated along a delay lines and are reflected and recombined in phase at the output port. By turning ON the different states of the switches, we obtain four output phases. So, using this PS, four pointed direction beam is achieved.

B. Simulation results

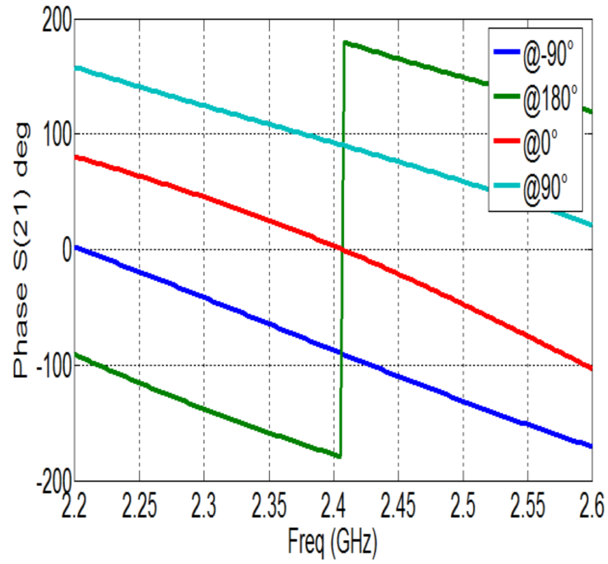
We simulated the proposed structure by using ADS, for the four phases, Figures 2(a), (b) and (c) illustrate the return loss, the insertion loss and the phase shift, respectively.



(a) Return loss



(b) Insertion loss



(c) Phase shift of 2-bits PS

Fig. 2. Simulation results of the 2-bits PS

Simulation results of the four phases show a return loss better than -15 dB, an insertion loss less than -3dB over 2- 2.6 GHz. The phase shift is equal to 90° with error of 0.5° in the centre frequency. The objective of the next paragraph is to validate our simulation through a realised experimentation setup.

C. Validation through PS experimental realisation

Our PS is fabricated using FR4 substrate with dielectric constant ε_r =4.7 and thickness of 1.6 mm. The size of the circuit is 7cm×2.8cm. Six transistors NE3508M04 are used to switch the different states. Bias lines provide the desired polarisation in the gates of the different transistor. The operating frequency is 2.4 GHZ. Figure3 shows the photograph of the fabricated 2-bits reflection type PS based on branch line coupler.

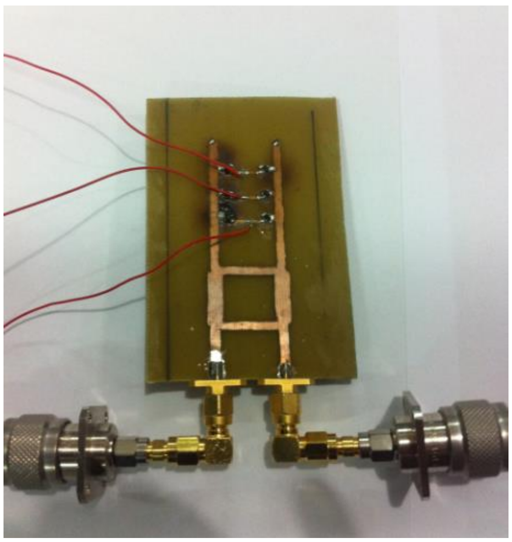
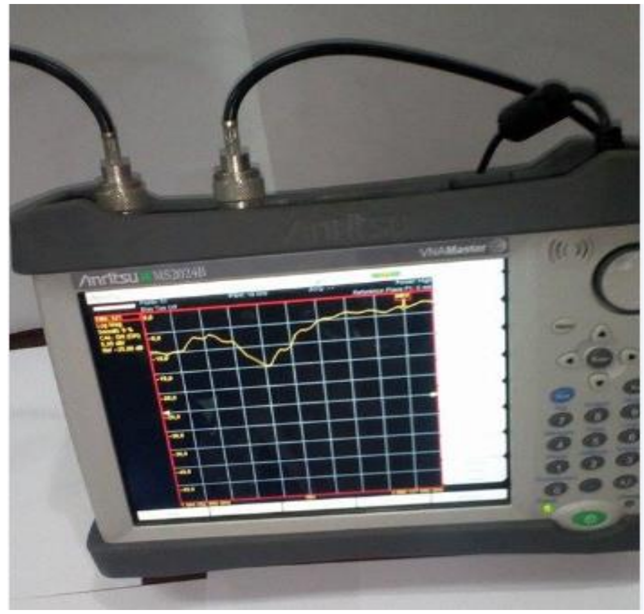


Fig. 3. Fabricated 2-bits PS

According to the three Bias Lines, four possible combinations are provided, depending on the states of the transistors. These combinations are 100, 010, 001 and 000, corresponding respectively to the phases 0°, 270°, 180° and 90°. Using the frequency analyser, we measured the return losses and insertion losses as depicted by Figure 4.

(a) Return loss



(b) Insertion loss

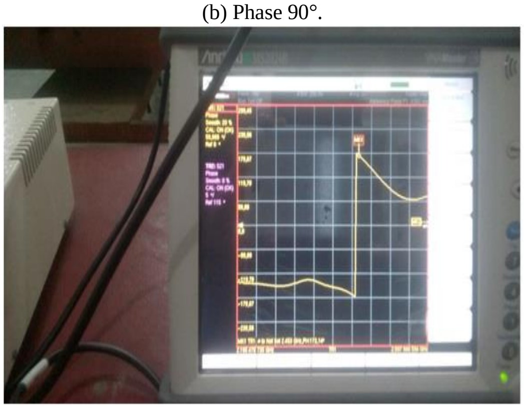
Fig. 4. Measurd results of the 2-bits reflection PS

The Figure 4 (a) illustrates the measured results for only the phase 270°. The measured return loss is less than -10dB over 2.7-2.9 GHz and is equal to -25 dB in 2.8 GHz. The measured insertion loss varied around -3 dB over 2.7-2.9 GHz. A difference of about 300 MHz was provided between the measured and simulated results. This difference is mainly due to the characteristic of the substrate, the error of fabrication and the influence of via. Moreover, the FETs used could insert a certain length of transmission line as well.

In a second hand, since the switching process is done through electronic module, we show in the Figures 5 (a), (b) and (c), the rest of the different phases 0°, 90° and 180°, respectively, obtained in the band 2.1-2.9 GHz.

(a) Phase 0°

(b) Phase 90°.



(c) Phase180°.

Fig. 5. Measured results of the 2-bits reflection PS

According to the measured results, we notice the presence of the same error range of about 300 MHz around the centre frequency of 2.4 GHz used in the simulation. Since the origin of the errors is considered to be the same for all phases, we believe that a better experimental setup could offer better results.

D. Discussion

We notice that the results provided by the experimental setup are close to theoretical simulation. The error occurred was at the order of 300 MHz, this difference is mainly due to the substrate features which are slowly different from those used at the simulation level. So, to conclude with the experimental realisation, we consider that the four phases are provided but not exactly at the desired frequency.

Furthermore, the drawback of such PS is related to its size considered to be as voluminous; therefore, we suggest a compact version of the 2-bits PS. Furthermore, the branch line coupler occupied an important area of the circuit. So, the development of a compact branch-line coupler is very necessary to reduce the size of the PS. Several compact branch-line couplers have been developed in [7], [8] and [9]. Based on these methods, we aim in the next section to propose our own miniaturised model of the PS.

III. 2-Bits Miniaturised PS Design and Simulation

The process of miniaturisation consists of miniaturising both the branch line coupler and the delay lines, presented by stage 1 and stage 2 of Figure1(a), respectively. The following work discusses this issue.

A. Miniaturised branch-line coupler

To reduce the size of the 3-dB branch-line coupler, S. Jung et al [8] employed the technique of open stub with low impedance. The proposed method consists first of replacing a quarter-wave length transmission line by shortened one and making equivalence between them. Then, a low or a high impedance open stub in the shortened quarter –wavelength (in T model, π model and a combination between them) is employed [8].

Based on this proposed method, we use T model shortened quarter-wavelength transmission line with low impedance open stub to realise a compact branch line coupler.

Equivalent quarter-wavelength transmission line of the T-model is presented in Figure 6.

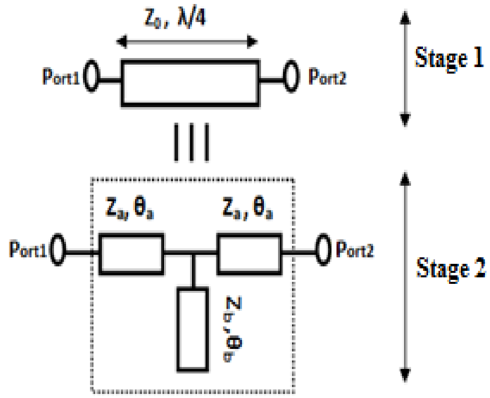


Fig. 6. Equivalent quarter-wavelength transmission line of the T-model

Let us first recall the ABCD matrix of a squared wavelengths transmission line (stage 1 of Figure 6) to be as follows:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\lambda/4} = \begin{bmatrix} 0 & jZ_0 \\ jY_0 & 0 \end{bmatrix} \quad (2)$$

where, Z_0 is the characteristic impedance and $Y_0 = {}^{1}\!/_{Z_0}$.

In our miniaturisation, we considered the T model with parameters N=1, θc=0 and θd=0 [8]. According to our simplification, the matrix ABCD of a squared wavelengths transmission line (stage 2 of figure 6) could be reduced to the following expression:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{shortned} = G.H.G \quad (3)$$

G and H of equation 3 are the matrices expressed as follows:

$$G = \begin{bmatrix} \cos\theta_a & jZ_a \sin\theta_a \\ jY_a \sin\theta_a & \cos\theta_a \end{bmatrix} \quad (4)$$

$$H = \begin{bmatrix} 1 & 0 \\ jY_b \tan\theta_b & 1 \end{bmatrix} \quad (5)$$

where,

Z_a: characteristic impedance of the shortened quarter-wavelength

θ_a : electric length of the shortened quarter-wavelength

Z_b : open stub characteristic impedance

Y_b: open stub admittance

θ_b : open stub electric length

After resolving equations 2, 4 and 5, we obtain:

$$Z_a = \frac{Z_0}{\tan\theta_a} \quad (6)$$

$$Y_b \tan\theta_b = \frac{2}{Z_a \tan 2\theta_a} \quad (7)$$

We choose a two low impedance open stubs (Zs1 and Zs2) and two electric lengths (θ1 and θ2) to determine the electric lengths (θs1 and θs2) of the two stubs and the characteristic impedances Z1 and Z2 of the direct and the coupled branches coupler. Then, the dimension of the compact coupler is determined using FR4 substrate having relative permittivity 4.4 and height h=1.6 mm, at a centre frequency equal to 2.4 GHz. The layout of the coupler is presented in Figure 7.

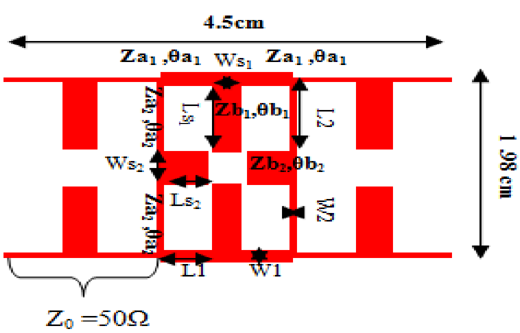
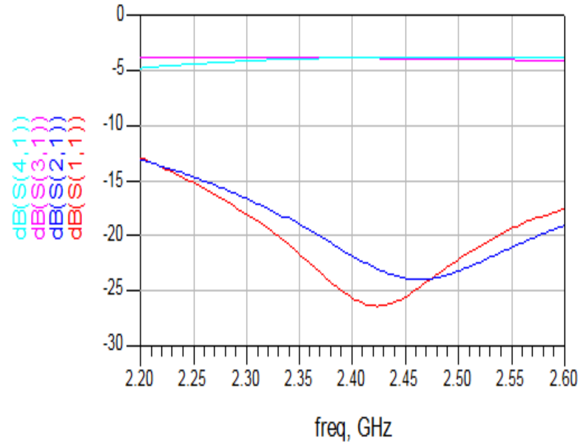


Fig. 7. Design of the T model branche line coupler with a low impedance

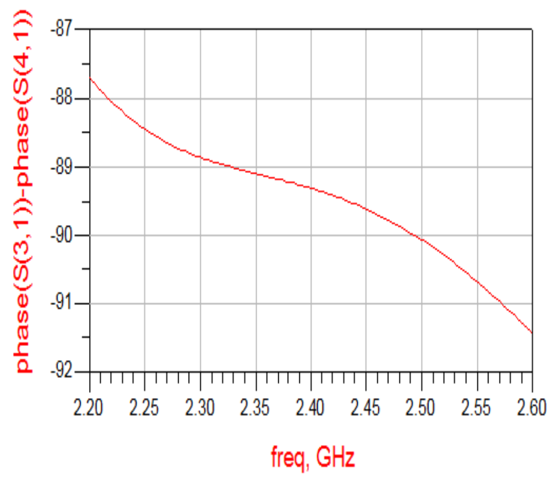
The reduction of the dimensions of the direct and the coupled branches coupler offers a compact size but the adaptation lines present important lengths. Indeed, we employed the same method to reduce the length of the adaptation line.

Using LineCalc ADS, the dimensions of the compact coupler are obtained as follows: Ls1= 6.9 mm, Ws1 =2.9 mm, Ls2=4.4 mm, Ws2=3.3 mm, L1= 4.9 mm, W1=1.2 mm, L2= 6.5 mm and W2=0.8 mm. Therefore, we obtain a size reduction of about 50% compared to the size of the conventional coupler.

The simulation results of the compact branch line coupler are presented in figure 8.



(a) Sij parameter



(b) Phase shift

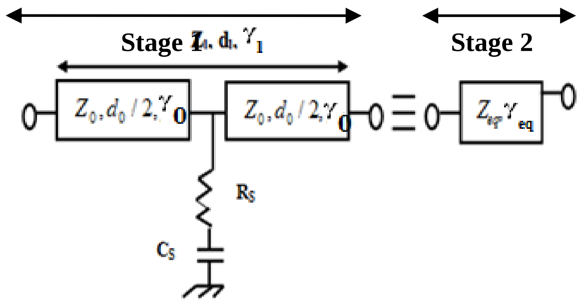
Fig. 8. Simulation results of the miniaturised line

The return losses (S11) and the isolation (S21) are better than -15 dB over 400 MHz, while the coupling (S31 and S41) are varying between 3 dB and 4 dB. The phase shift between port 3 and 4 is 90°± 2° over the frequency range 2.2 GHz-2.6 GHz.

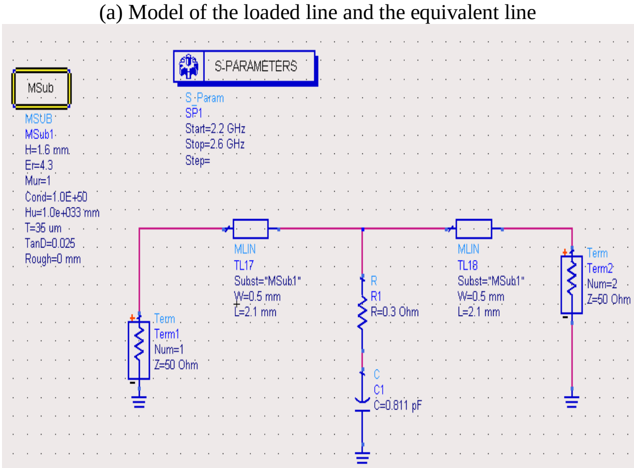
B. Miniaturisation of the transmission line using distributed elements

Miniaturisation technique using loaded line consists of adding distributed elements in parallel or in serial of the propagation line. This technique is used to reduce the filter [10-11-13], the resonator [12] and the coupler [14].

In this section, the miniaturised technique used consists of replacing the transmission line by sections of loaded line with parallel capacitor. The loaded line section (stage 1 of Figure 9 (a)) is equivalent to a transmission line characterised by a low propagation velocity (stage2 of Figure 9 (a)). This method can offer a size reduction lower than 50% [15-16].



(a) Model of the loaded line and the equivalent line



(b) Electric model of the loaded line using ADS

Fig. 9. Design of the miniaturised transmission line

The capacitors CS and the electric length θ0 are determined as follows [17-18]:

$$\cos\theta_l = \cos\theta_0 - 0.5 Z_0 C_s w \sin\theta_0 \qquad (8)$$

$$Zc_l = Z_0 \sqrt{\frac{1-0.5Z_0C_s w tg\dfrac{\theta_0}{2}}{1+0.5Z_0C_s w \cot\dfrac{\theta_0}{2}}} \qquad (9)$$

where,

θ_l : Electric length of loaded line

θ_0 : Electric length of not loaded line

Z_l : characteristic impedances of the loaded line

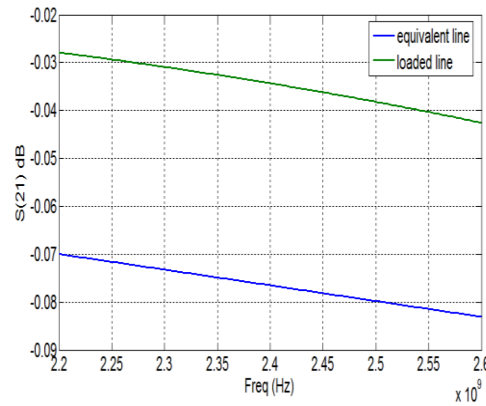
Z_0 : characteristic impedances of the not loaded line

w: resonance frequency

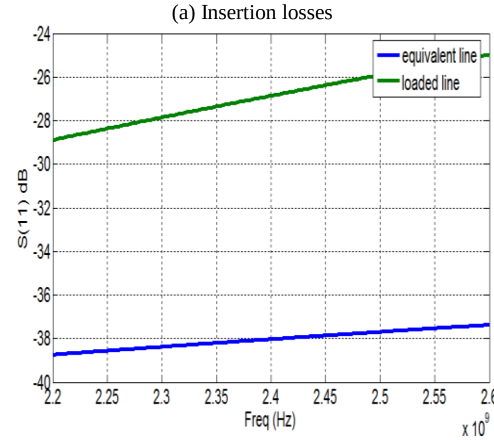
C_s : capacitor CMS

Based on this method, we can replace the delay lines characterised by electric length equal to θ0=45° by a sections of loaded line with capacitor. When we consider the centre frequency at 2.4 GHz, Zl =50Ω, Z0 =110 and θl=45°, we can determine the value of the capacitor and the dimension of the not loaded line.

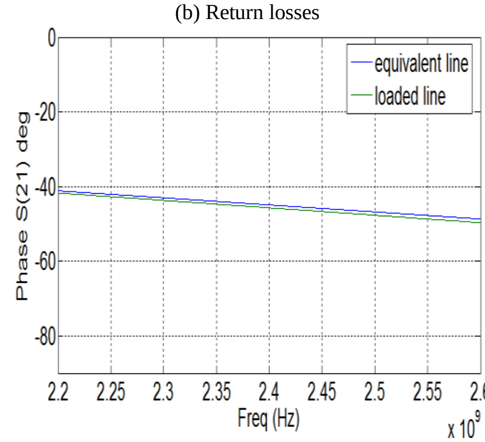
At this level, we thought about validating the dimensions by considering their effect on the performances of the line. To this end, we simulated the line not loaded and the equivalent line as depicted in Figure 10.



(a) Insertion losses



(b) Return losses



(c) Phase shift

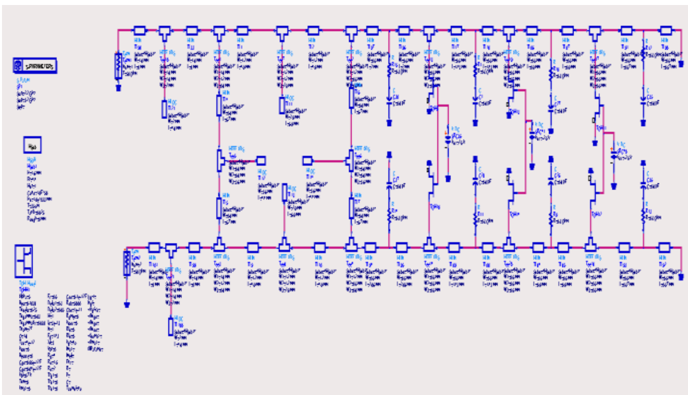
Fig. 10. Simulation results

The length and the width of the equivalent line at the center frequency of 2.4 GHz are respectively equal to leq=8mm and weq=3mm. Using this miniaturised technical, the propagation line is replaced by two sections of not loaded lines separated in the middle by a capacitor. The dimension of the not loaded line is 2.1 mm and the capacitor is 0.8 pf. Indeed, we obtain a reduction size of 50%.

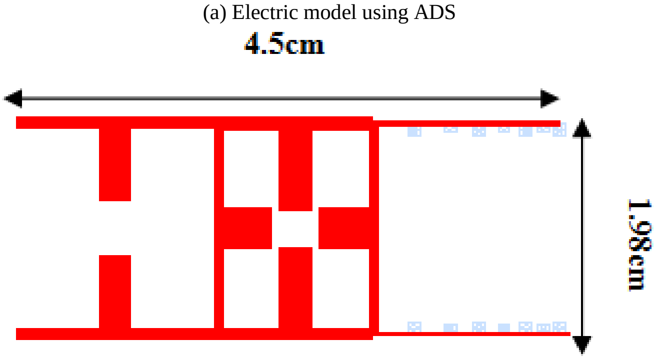
The simulation results of the two lines showed the same performances with a little difference. The Return losses of the line not loaded is better than the return losses of the equivalent line. In the other hand, the insertion losses of the equivalent line are important.

C. 2-bits miniaturised reflection PS

In figure 11, the design and the layout of the whole miniaturised 2-bits reflection PS are illustrated. The value of capacitor is equal to 0.8 pf and the series resistance is 0.3 Ω. The length and the width of delay lines are respectively equal to 4.2 mm and 0.5 mm at the centre frequency 2.4 GHz. Then, the dimensions are optimised with ADS to obtain the adequate results.



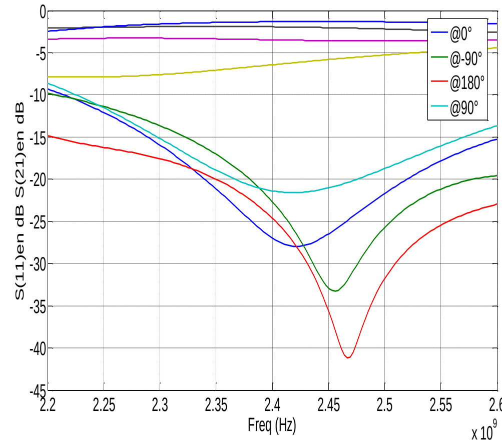
(a) Electric model using ADS



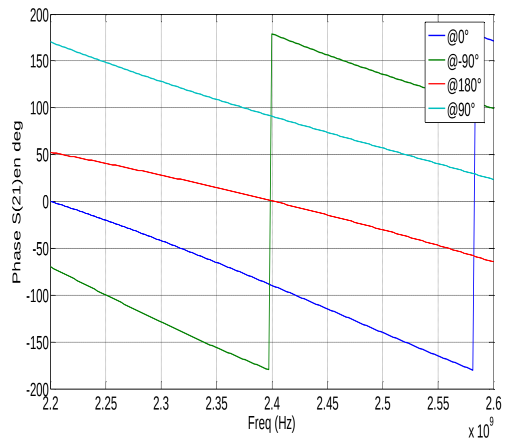
(b) Layout

Fig. 11. Design of miniaturised transmission line

The size of the circuit is reduced to 4, 5cm×1.9cm whereas the size of the conventional PS is 7cm×2.8cm. Simulation results are presented in figure 12.



(a) Sij parameter



(b) Phase shift

Fig. 12. Simulated results

We obtain the same performance of the conventional PS, but we notice an increase of the insertion losses especially for the phase 180°. The phase shift between the four states is equal to 90° in the centre frequency and we obtain a good adaptation.

IV. Conclusion

In this study, we presented the design and the implementation of a 2-bits reflection type PS. We showed that the results obtained by the experimental setup are closed to the simulation at an order of 300 MHz. The PS under study presented a big size and a miniaturisation of the structure was addressed. To this end, different miniaturisation techniques were employed to reduce the circuit size and obtain the same performance. The proposed structure can provide 4 states with 90° of phase shift. Semiconductor devices were used to control the proposed PS, these devices are characterised by a fast switching speed, which permitted providing a fast beam steering and beam shaping. The perspectives of this work address the design and implementation of 4-bits and 6-bits PS.

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