Phase Shifters with Tunable Reflective Method Using Inductive Coupled Lines

Prof. Nitin Sherje
Professor, Department of Mechanical Engineering, S.K.N.C.O.E Pune India
npsherje@sinhgad.edu

Article History

Article Submission
19 December 2016
Revised Submission
30 January 2017
Article Accepted
10 March 2017
Article Published
31st March 2017

Abstract

A coupler forms essential aspect of a phase shifters with tunable reflector, which is a 3 decibel quadrature coupler (λ/4). In this paper, a model which shows that by reducing the length of the coupler a wide phase range is achieved for a reflective type phase shifter. The approach used in this method is by having a variable instead of constant Even and Odd impedances. The ultimate aim is to design a Reflective Type Phase Shifter which has a very low area, low return and insertion losses and a large phase range. The proposed model is done using Advanced Design System (ADS) and the results are verified for the frequency of 2.4GHz.

Keywords: *3-dB coupler, Reflective Type Phase Shifters, Even and Odd impedances.*

I. Introduction

Tunable phase shifters find its applications in many key areas such as frequency converters, MIMO, satellite systems, microwave instruments. Size of a coupler, insertion and return loss plays an optimum role in the phase shifter design especially reflective type [1][2]. The reason behind this is in MIMO a mobile handset is used which requires a small sized coupler instead of a larger one. Plus the cost also occupies an important role in this. But this reduction in size may affect the return loss of the coupler and hence the size should be reduced with a reasonable reduction in the return loss. Similarly insertion loss also, if a phase shifter is used in the transmitter part then the phase shifter may allow an undesirable power to the transmitting antenna and if the PS is used in the receiver part, then the SNR value of the signal may get degraded if the insertion loss is very high [3][4]. The 3-dB coupler is most commonly used in a RTPS which divides the entire signal into two orthogonal signals of equal energy. Then phase shifter is realized by placing impedance at two ports of the coupler except input and output port. The characteristic impedance is achieved using odd and even impedance by (Zoo* Zoe)=(Zo)2 A steady value is chosen for even and odd impedance and by using the above formula 50 ohms is obtained. In this paper, a model which uses coupler length which is less than (λ/4) to reduce the area and by adjusting the impedance of coupler with wide phase range is obtained [5][6].

II. Proposed Reflective Type Phase Shifters

The RTPS has impedances connected at Coupled and Isolated ports. By variation of the impedances the phase of the signal at the output is changed. This is shown in figure 1.

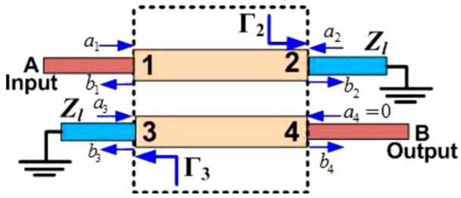

Fig 1: Proposed Reflective Type Phase Shifters

Here the length l is variable, and to obtain a phase change the Load impedance Zl is changed. Here instead of using 3-dB coupler, a coupler of length less that 3-dB is used. Also the mode impedances such as Even and Odd are assumed to be independent variables, which mean they can take any value but square root of their product must be equal to the characteristic impedance [7][8].

The signals which are outgoing can be calculated for the incoming signal a using the formula

$$\mathbf{b=[I - S\Gamma]^{-1} * Sa} \qquad (1)$$

where S = Scattering Parameters
Γ = Reflection coefficient
b = vector of outgoing signals
a = vector of incoming signals

Γ = Diagonal matrix of the reflection coefficient

$$\Gamma 1=\Gamma 4=0 \quad \Gamma 2=\Gamma 3=\frac{(Zl-Zin)}{(Zl+Zin)} \qquad (2)$$

where Zin= Input impedance

$$Zin = Zo + \frac{2 * (Zine * Zino - Zo * Zo)}{Zino + Zine + 2Zo}$$

$$Zino = Zoo * \frac{Zo + jZoo * \tan(\beta l)}{Zoo + jZo * \tan(\beta l)}$$

$$Zine = Zoe * \frac{Zo + jZoe * \tan(\beta l)}{Zoe + jZo * \tan(\beta l)}$$

The S parameter calculation using odd and even impedances are as follows

$$S11 = \frac{(S11e + S11o)}{2}$$

$$S21 = \frac{(S21e + S21o)}{2}$$

$$S31 = \frac{(S11e - S11o)}{2}$$

$$S41 = \frac{(S21e - S21o)}{2}$$

$$S11e = \frac{j\left(\left({Zoe}/{Zo}\right) - \left({Zo}/{Zoe}\right)\right) * \sin(\beta l)}{2\cos(\beta l) + j\left({Zoe}/{Zo} + {Zo}/{Zoe}\right) * \sin(\beta l)}$$

$$S11o = \frac{j\left({Zoo}/{Zo} - {Zo}/{Zoo}\right) * \sin(\beta l)}{2\cos(\beta l) + j\left({Zoo}/{Zo} + {Zo}/{Zoo}\right) * \sin(\beta l)}$$

$$S21e = \frac{2}{2\cos(\beta l) + j\left({Zoe}/{Zo} + {Zo}/{Zoe}\right) * \sin(\beta l)}$$

$$S21o = \frac{2}{2\cos(\beta l) + j\left({Zoo}/{Zo} + {Zo}/{Zoo}\right) * \sin(\beta l)}$$

(3)

The important aspect that governs the routine of the phase shifter are ϕ, k at the input terminal and output. For Fig1, they can be calculated as

$$S_{AA} = S_{BB} = (b1/a1) \quad S_{BA} = (b2/a2) \quad \phi = \text{angle}(S_{BA}) \qquad (4)$$

Assuming that the loads are connected at ports 2 & 3 then the varactor at these ports are varied such that Cv changes between maximum(Cmax) and minimum(Cmin) values and the phase range of the output signal is calculated using

$$\Delta\phi = \phi\ (\text{when } Cv=Cmin) \text{ -- } \phi(\text{when } Cv=Cmax) \qquad (5)$$

as a function of coupler length and mode impedance. The varactor being a capacitor is assumed to have a capacitor ratio Rc = Cmax / Cmin to be 10. Thus Cmax and Cmin are assumed to have values of 2pF and 0.2pF. The capacitor is not assumed to be greater than 10 because of feasibility and hence instead of increasing the capacitor ratio an inductor is added in series to the varactor to get a higher phase shift. At the same time the inductor value should not exceed a threshold because even a higher inductance will reduce the phase range [9][10].

The odd impedance is assumed to have a low value and the even impedance a higher value to have a higher level of coupling between the coupling lines to get the desired phase. And also care should be taken such that the return loss does not come above -10dB and the reduction of the length of the coupler should be stopped if the return loss goes above -10dB.

III. ANALYSIS AND SIMULATION RESULTS

The presented design is tested using phase shifters at 2.4GHz with 10dB return loss. This frequency band is selected as it is used modern implementations including WLANs with MIMO front ends. This setup is controlled by tunable phase shifters.

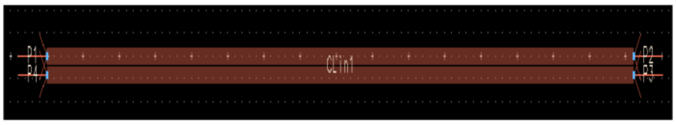

Fig 2: Layout of (λ/4)

The proposed design has a precise setup that has small resistance value, moderate impedance values and small coupled structure with minimum L and C value that reduces stray capacitances. The required values are inductor L=6.5nH , Cmin=0.2pF, Rc=10 , Zoo=10Ω, Zoe=225Ω.

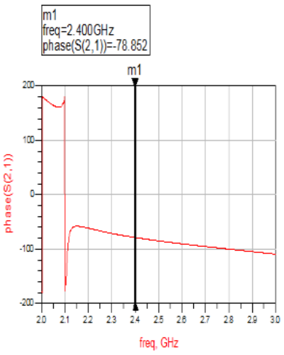

Fig 3: Simulated Results of phase for C=0.2pF and 2pF(λ/4)

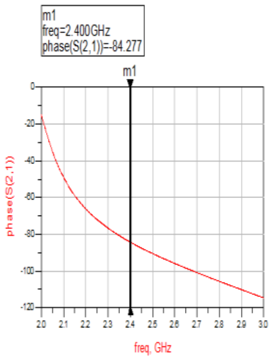

Fig 4: Simulated Results of phase for C=0.5pF and 2pF(λ/8)

The presented design is tested using phase shifters at 2.4GHz with 10dB return loss. This frequency band is selected as it is used modern implementations including WLANs with MIMO front ends. This setup is controlled by tunable phase shifters along with *λ/8 PS* .

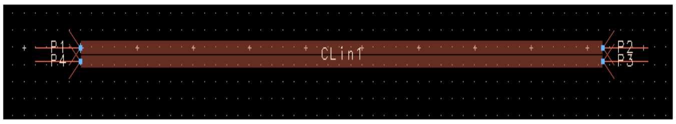

Fig 5: Layout for (λ/7)

The proposed design has a precise setup that has small resistance value, moderate impedance values and small coupled structure with minimum L and C value that reduces stray capacitances. The required values are inductor L=6.5nH , Cmin=0.2pF, Rc=10 , Zoo=10Ω, Zoe=225Ω.

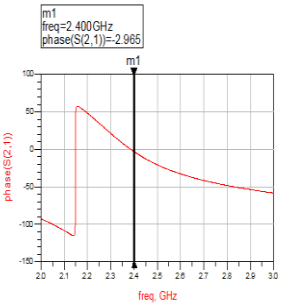

Fig 6: Simulated Phase for C=2pF and 0.2Pf(λ/7)

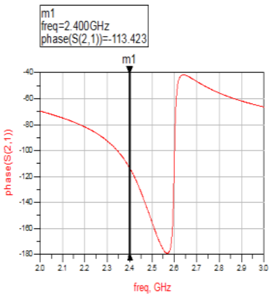

Fig 7: Simulated Phase for C=5pF and 0.8Pf(λ/7)

The simulated results are tabulated in table 1 and table 2 respectively.

Table 1: Coupler length Vs Phase range

S NO	COUPLER LENGTH	PHASE RANGE(DEGREES)
1	λ/4	6
3	λ/7	-111

Table 2: Coupler length Vs insertion loss and return loss

S NO	COUPLER LENTH	INSERTION LOSS(dB)		RETURN LOSS(dB)	
		MIN	MAX	MIN	MAX
1	λ/4	6.56	0.71	5.9	8.4
3	λ/7	0.28	0.3	12.5	19.8

V. Conclusion

Thus from the observation 3-dB coupler which is of length λ/4 gives a very less phase range compared to the lengths which are less than 3-dB coupler. And the return loss value is also considerably around 10dB for less lengths. Hence for the above values of Even and Odd impedance (λ/7) proves to be the better and optimum solution instead of (λ/4).

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