ISSN : 2231-0711

Novel Designs of Reflective Type Phase shifters for Millimeter wave application

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***Abstract*—** Design and Implementation of highly efficient Reflective type phase shifters (RTPS) at 60GHz has been proposed in this work. Low loss cost effective design which utilizes less chip area is proposed. At high frequencies, to improve the efficiency of the communication system highly directional high gain antennas have to be used. Phase shifters are one of the most important elements in phased array antenna system. Phase shifters are used to shift the phase of the input signal. To reduce the losses, instead of passive elements active elements such as active inductors and varactors are used in this work. By using active elements the utilization of chip area can be reduced. Three designs of Reflective type phase shifters with spiral inductor, active inductor and MOS varactor as reflective load are proposed in this work. RTPS with spiral inductor as load has produced a constant phase shift of -83^0. A variable range of phase shift has achieved for RTPS with active inductor load and MOS varactor load. At 60 GHz path loss is very high due to the small wavelength thus it is necessary to use high gain antenna. Phased array antennas are capable of producing narrow, high gain beams[1] .

***Keywords*— Active inductor, Hybrid coupler, Millimeter wave, Reflective type phase shifter, Varactor, 60GHz.**

I. INTRODUCTION

Researchers and industries are now more and more interested in millimeter wave spectrum due to the large availability of unlicensed bandwidth in this spectrum. Millimeter waves can be used for short range wireless applications such as wireless local area network (WLAN), wireless personal area network (WPAN)[1]. Millimeter wave spectrum ranges from 30-300GHz.The wavelength of this band is from 10mm-1mm, hence it is known as millimeter band. Unlicensed bandwidth is available in the V band from 57GHz to 64GHz; it is very useful for short range wireless communication due to the strong absorption characteristic by oxygen in the atmosphere. For millimeter wave communication the antennas should have high gain and directivity, to reduce interference and free space loss.

Phase shifters are essential components in a phased array to adjust the phase of each antenna path and steer the beam. Individual antenna in an array can be turned on or off or driven through different phase shifter. The radiation pattern of array can be pointed over a wide range of different directions without physically moving the antennas with the help of these phase shifters. A compact low power and linear phase shifter lowers the cost and complexity of the phased array system to a large extend. A continuous phase shifter reduces the beam pointing error of the phased array antenna.

The millimeter wave systems are usually fabricated using HIC (hybrid integrated circuit) technology. This will consume more area and cost. So MMIC (monolithic millimeter wave integrated circuit) technology is considered as an alternative technology for HIC technology due to its ability to integrate passive and active elements on a single semiconductor substrate. The MMIC has advantages such as high reliability, high productivity, small size and low cost compared to conventional HIC technology [21]. Complementary metal oxide semiconductor (CMOS) technology is the dominating technology for most wireless products below 10GHz, Nowadays with the aggressive scaling of gate length, CMOS technology is pushing further into millimeter wave region. Main advantages of CMOS technology compared to other semiconductor technologies such as GaAs and SiGe are its high reliability, low cost and high device count. GaAs and SiGe have low yield and limited integration[21],[17]. CMOS technology offers high level of integration with RF analog and digital circuits. The Speed of CMOS transistors can be increased by scaling down the gate length.

This work presents three novel designs of Reflective type phase shifters (RTPS) at 60 GHz. Reflective type phase shifters are used to provide both large and small phase shifts. In this work the focus is placed on RTPS to provide a continuous, low-power phase shift over the desired frequency band (at 60 GHz). RTPS with three different reflective loads are presented in this work. The reflective loads used in this work are

1. Spiral inductor
2. varactor
3. Active inductor

Spiral inductor consumes large chip area, and the losses are also high [14,15]. MOS varactor will consume small chip area and the complexity of the design is also less. Active inductor will have high Q factor at high frequencies hence losses will be less. Less chip area is needed to implement active inductors.

II. IMPORTANCE OF 60 GHZ

The 60GHz spectrum is ideally useful for high data rate applications. It is suited for such application because of its unlicensed worldwide availability, high allowable transmit power, and large contiguous bandwidth. 5GHz of contiguous bandwidth at mm wave frequencies around 60GHz was opened by FCC for unlicensed use. the license which is highly expensive. Significance of the frequency around 60GHz is that ,the atmospheric oxygen molecules resonate at this frequency. So the Oxygen molecule will absorb this EM Waves. This will increase the path loss exponentially and it causes attenuation of EM energy at distances beyond a few kilometers. So long distance transmission is not possible and long range interference is also not an issue. This enables frequency reuse.

III. REFLECTIVE TYPE PHASE SHIFTER

Reflective type phase shifter consists of a 90^0 hybrid coupler and terminated with two identical reflective loads with equal impedance. The reflective loads should be matched [2][3]. 90^0 hybrid coupler divides the input signal into two signals 90^0 out of phase. These signals reflect from a pair of reflective load and combine in phase at the phase shifter output. The input signal is divided into two parts and each part is reflected at one of the reflective loads. The phase of RTPS output can be controlled by varying the impedance of reflective load. Reflective loads are purely reactive, no power is lost in the reflective loads and all power is coupled into the output. The block diagram of RTPS is shown in fig. 1.

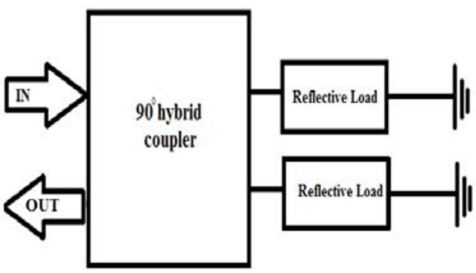

Fig.1 Block diagram of a reflective type phase shifter

The main advantage of this configuration is that input and output impedance matching is preserved, although the coupler is terminated with reflective loads. Let Z_0 and Z_L be the characteristic impedance of the coupler and input impedance of the load then the expression for reflection coefficient is

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} \qquad (1)$$

If Z_L and/or Z_0 are complex quantities then Γ also will be complex quantity and that can be expressed as

$$\Gamma = |\Gamma| e^{j\theta} \qquad (2)$$

where $|\Gamma|$ is the magnitude of reflection co-efficient and θ is the phase angle of reflection co-efficient. The phase of reflection co-efficient can be changed by varying Z_L. We can use variable active inductances and capacitances for this purpose. Due to the advantage in terms of loss, size and power consumption varactors are more preferable. The phase variation of Γ determines the transmission phase of circuit and can be calculated using the expression given below.

$$\Delta\theta = 2\left[\tan^{-1}\left(\frac{Z_{max}}{Z_0}\right) - \tan^{-1}\left(\frac{Z_{min}}{Z_0}\right)\right] \qquad (3)$$

The values for Z_{max} and Z_{min} can be found out by using the following equation:

$$Z_{max} = \frac{1}{\omega_0 C_{min}} \qquad (4)$$

$$Z_{min} = \frac{1}{\omega_0 C_{max}} \qquad (5)$$

where C_{max} and C_{min} be the maximum and minimum varactor capacitances. The value of Cmax and Cmin depends on the centre value C0 of varactor and the tuning ratio of varactor.

$$C_0 = \frac{1}{j\omega z_0} \qquad (6)$$

Tuning ratio of a varactor is also known as capacitance ratio. It is ratio of varactor capacitance at a minimum reverse voltage to maximum reverse voltage

$$C_{max} = C_0\sqrt{t_v} \qquad (7)$$

$$C_{min} = \frac{C_0}{\sqrt{t_v}} \qquad (8)$$

Any load which can provide an impedance mismatch can produce reflections. So we can use an inductor as reflective loads, but only one problem is its reactance should not match with the characteristic impedance of the coupler.

IV. 90^0 HYBRID COUPLER

90^0 Hybrid coupler is a 4-port directional coupler. If the coupler is designed for 3dB coupling then it splits the input power in port 1 into equal powers in port2 and 3, Thus it can be used as power divider. Directional couplers with 3dB coupling are also called hybrid junctions. It has four arms into which power can be fed and from which power can be taken. Fig.2. illustrates signal splitting in a 3dB hybrid coupler. 90^0 hybrid coupler will split the incoming signal into two signal having equal power, but produces some phase shift. If we are giving input to port 1 then port2 is known as through port, port 3 is known as coupled port and port 4 is called as isolated port.

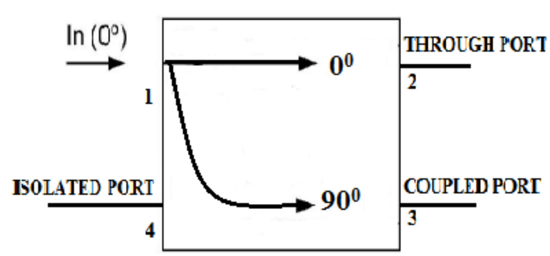

Fig. 2 Block Diagram of a 90^0 hybrid coupler

90^0 Hybrid coupler occupies a large area when they are realized using microstrip and coplanar wave guides. This prevents their integration with other RF circuits. So realization of couplers using lumped elements can be done. This lumped element L-C realization helps to reduce the chip area. In recent years, lumped elements are very attractive in the sense of realizing very small sized circuits since fabrication techniques for inductors and capacitors have improved so as to be able to neglect the parasitic inductance and capacitance even in the ultra high frequency (UHF) band. 90^0 Hybrid coupler realized using the Lumped elements are shown in fig 3. The component values can be calculated using the given equations.

$$C_1 = \frac{1}{\omega_0 Z_0} \qquad (1)$$

$$L_1 = \frac{Z_0}{\omega_0 \sqrt{2}} \qquad (2)$$

$$C_2 = \frac{1}{\omega_0^2 L_1} - C_1 \qquad (3)$$

Equation 1 simply sets the magnitude of the branch impedence $1/\omega_0 C_1$ at frequency ω_0 to Z0. Equation 2 equate sthe series line impedence $\omega_0 L$ to the characteristic impedence 1/√2 Z0. Inductor resonates with capacitance C1+C2 at ω_0.

Z_0 is the characteristic impedance of the Hybrid coupler. ω_0 is the centre frequency. Here we have designed for a center frequency of $f_0 = 60$GHz. The values for C_1, C_2, L_1 are 53fF, 22.65fF, 93pH respectively.

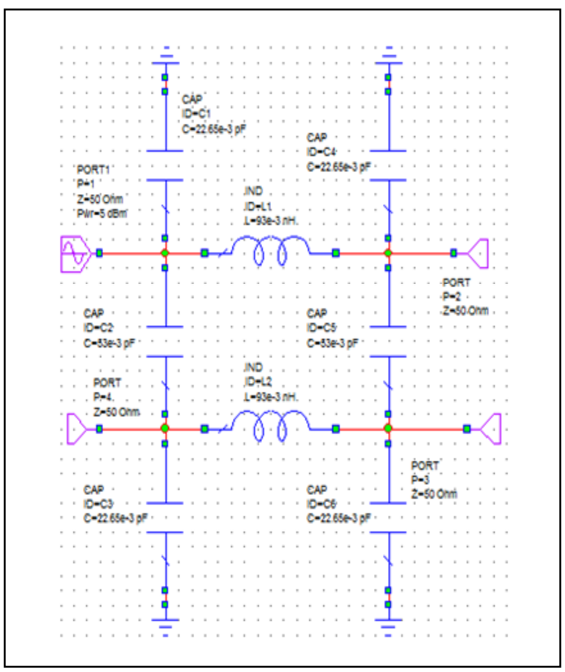

Fig. 3 Lumped element 90^0 Hybrid coupler

The Smatrix obtained for the designed 900 hybrid coupler is given below.

$$\begin{bmatrix} .009\angle 179.5 & .69\angle .043 & .69\angle -89 & .009\angle -89 \\ .71\angle .008 & .009\angle 178.5 & .009\angle -88 & .69\angle -88.97 \\ .69\angle -89.11 & .009\angle -88.97 & .009\angle 179.5 & .71\angle 14 \\ .009\angle -89.11 & .7\angle -89.56 & .71\angle -.1367 & .009\angle 179.3 \end{bmatrix}$$

The input port of coupler is port1.Port 2 is the through port and port 3 is the coupled port. Port 4 is the isolated port. All the four ports of the coupler are terminated with equal impedance. This means all the four ports are matched, Hence no reflections from any of the ports. For an ideal 90^0 hybrid coupler S11=S22=S33=S44=0. This signifies the absence of reflections. For the designed hybrid coupler the values of S11=S22=S33=S44=.009. The values of S14, S41, S23 and S32 for the designed coupler is .009 which is approximately equal to zero. This also satisfies the ideal condition. The 3 dB coupler divides the input power equally between the through port and coupled port. But the signal reach at port 3 ie the coupled port will have some phase shift which is equal to 90.

$|S_{12}| = .69$
$|S_{21}| = .71$
$\angle S_{12} = .04$
$\angle S_{21} = .008$
$|S_{13}| = |S_{31}| = .69$ and $\angle S_{13} = \angle S_{31} = -89$

From these values it is evident that for the particular coupler the input power is equally splitted between the through port and coupled port. The phase shift produced by the coupler is -89 degrees which is approximately equal to 90 degrees.

V. MOS MODEL

MOS modeling has been done for the designing of RTPS with MOS varactor load and MOS Active inductor load. Modeling of MOSFET at 60 GHZ has been done. Modelling a MOSFET at 60 GHz is one of the main challenge in millimeter wave circuit designing. The main challenge of the compact circuit model is to be able to describe all the operation modes of a semiconductor device. Another challenge is the inaccurate behaviour of the model at high frequencies. The conventional approach is modeling a geometry dependent capacitance with a single lumped element. With shrinking line widths, scalable device sizes and the requirement of higher frequency operation has put more and more weight on accurate device characterization [7],[18]. The models have to be very accurate and the extraction of model parameters must be done carefully.

Accurate small-signal equivalent circuit model is an important step in circuit design. Transistors equivalent small signal model is shown in fig 4. Small signal model consists of extrinsic elements and intrinsic elements. Intrinsic parameters include Cgs, Cgd, Cds, Ri, g_m and g_{ds}. Lg, Ls, Ld, Rg, Rs, Rd are the extrinsic parameters.

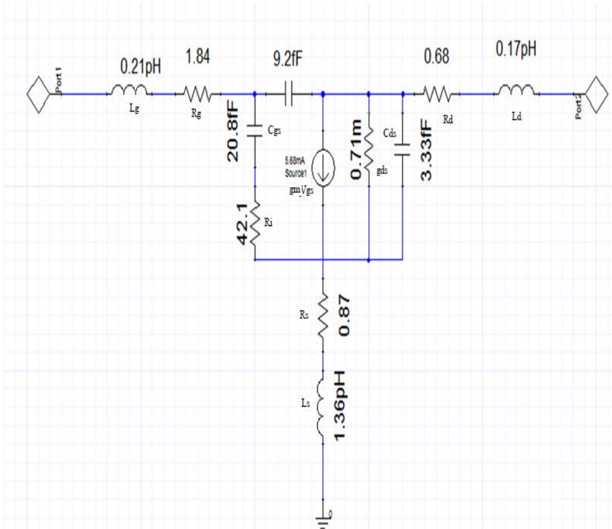

Fig. 4 Designed small signal model at 60 GHz

TABLE I

VALUES FOR INTRINSIC AND EXTRINSIC PARAMETERS

Intrinsic parameter	value	Extrinsic Parameter	Value
Ri	42.1Ω	Rg	1.84 Ω
Cgs	20.8fF	Rs	0.87 Ω
Cgd	9.20 fF	Rd	0.68 Ω
Cds	3.33 fF	Lg	0.21pF
Gm	5.68mS	Ls	1.36pF
Gds	0.71mS	Ld	0.17pF

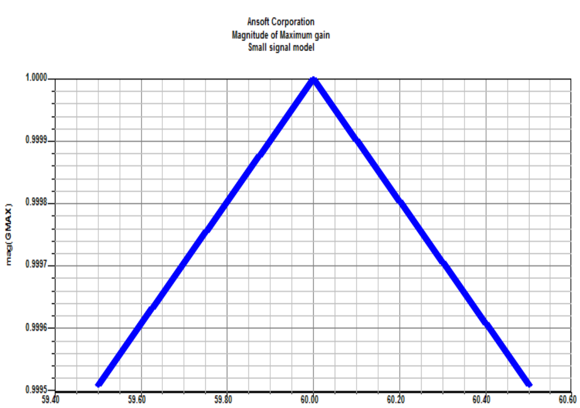

Fig. 5 Magnitude of Gain Vs Frequency

The magnitude of gain is maximum at 60 GHz. So this model can be very well used for the designing of devices or circuits at 60 GHz. Further designing of RTPS is carried out using this modelled MOSFET.

VI. RTPS WITH SPIRAL INDUCTOR LOAD

Design of RTPS with spiral inductor load is shown in fig 6. The characteristic impedance of the designed coupler is 50Ω. Hence whenever there is an impedance mismatch between the load and coupler, reflections will takes place. Here Z_0 is 50 Ω and Z_L is 2.3nH.There is impedance mismatch between the load and coupler hence the above circuit will acts as a reflective type phase shifter. The phase shift produced by the spiral inductor can e find out by using the equation 4.

$$\theta = -\left(\tan^{-1}\left(\frac{X_L}{Z_0}\right)\right) \qquad (4)$$

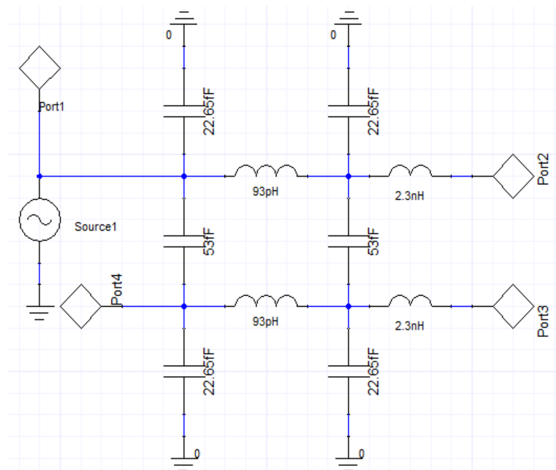

Fig. 6 RTPS with spiral inductor load

$$\begin{bmatrix} .001\angle-174 & .07\angle-83.51 & .08\angle-173.2 & .9\angle-83.51 \\ .07\angle-82.73 & .9\angle6.584 & 0.0001\angle104.1 & .08\angle-83.51 \\ .08\angle-174 & 0.0001\angle103.7 & .9\angle6.58 & .07\angle-173.3 \\ .9\angle-83.43 & .08\angle-83.52 & .07\angle-173.5 & .001\angle-173.5 \end{bmatrix}$$

The first type of RTPS in this work is RTPS with spiral inductor load.. Analysis of the S matrix obtained for the design has been done to verify the working of the phase shifter as RTPS. The S matrix for RTPS with spiral inductor load is given. For an ideal RTPS $|S_{41}| = |S_{14}| = 1$. For the designed RTPS $|S_{41}| = |S_{14}| = .9$ which is approximately equal to one. S11 and S44 represent the signals which are returned back to the input port. Ideally it should be zero. For the particular design $|S_{11}| = |S_{44}| = .03$. The magnitude of Reflection coefficient is given by |S22| and |S33|. Ideally S22 and S33 of RTPS is 1. For the above designed RTPS |S22| =.9; |S33| =.9. Hence the values of |S22| and |S33| approximately equal to that ideal RTPS. Since the reflection coefficient is maximum reflection takes place. It is shown in fig 7.

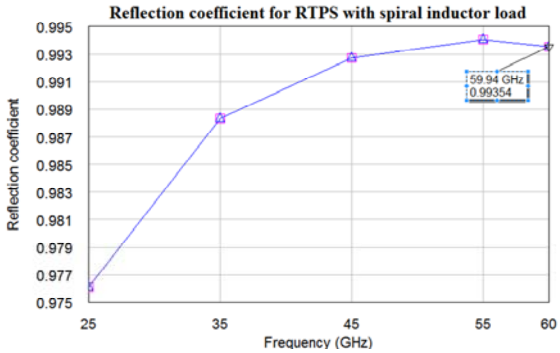

Fig. 7 Reflection coefficient for RTPS with spiral inductor load

VII. RTPS With MOS Varactor Load

The circuit shown in fig 8 is designed in such a way that it has to operate as reflective type phase shifter. By connecting the source and drain terminal together and giving a control voltage across the S/D and gate a variable capacitance can be produced[2]. This will act as a capacitor with gate terminal as one plate and S/D terminal as the other plate of a capacitor. Positive terminal of the control voltage is connected to Source/Drain terminal and the negative terminal of the control voltage is connected to the Gate terminal of the MOSFET. So the device is reverse biased voltage increases width of depletion region also increases. As the width of depletion region increases the capacitance value decreases. The control voltage is connected to the source /Drain terminal. As the control voltage changes the capacitance formed between the S/D and gate also changes. The tuning ratio of the designed varactor is two. RTPS with varactor load is shown in fig 9.

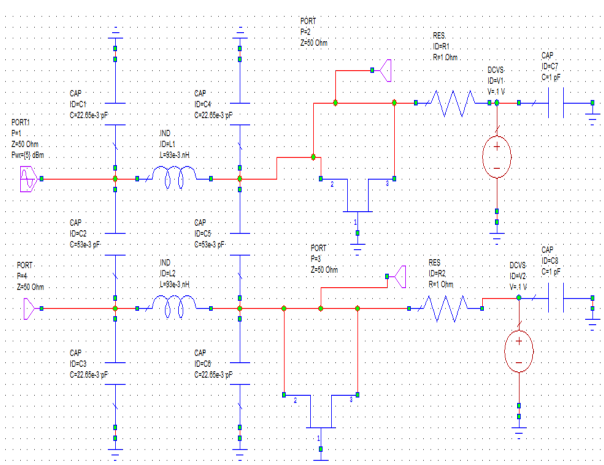

Fig. 8 RTPS with Varactor load

The RTPS with varactor load can produce a continuous variable phase shift, and the range of phase shift can be determined by the equation 5.

$$\Delta\theta = 2\left[\tan^{-1}\left(\frac{Z_{max}}{Z_0}\right) - \tan^{-1}\left(\frac{Z_{min}}{Z_0}\right)\right] \quad (5)$$

The S matrix for the given phase shifter is shown below

$$\begin{bmatrix} .001\angle .37 & .02\angle -90.05 & .02\angle .095 & .96\angle 90.05 \\ .02\angle -90.07 & .96\angle -177.5 & .00002\angle -89.26 & .02\angle .09 \\ .02\angle .37 & .00002\angle -89.53 & .96\angle -179 & .02\angle -89.53 \\ .96\angle 90.4 & .02\angle .4 & .02\angle -89.41 & .001\angle .46 \end{bmatrix}$$

As the magnitude of control voltage changes from .1-1 the output phase also changes from 90.4^0-122.3^0 .ll title and author details must be in single-column format and must be centered. Comparing with the S matrix of ideal reflective type phase shifter |S11|=|S44|=.001 which approximately equal to zero. Hence the signal returned back to the input port is negligible .so return loss is very less. |S14|=|S41|=.96 where port 1 is the input port and port 4 is the output port. Maximum signal is reflected back to the output port. $\Delta\theta = 39.62^0$.

VIII. RTPS With Active Inductor Load

For the designing of a variable RTPS variable loads are required. One RTPS with variable varactor load that we have discussed already. The circuit schematic of active inductor is shown in fig 9. At high frequencies active inductors are more appropriate than varactor. Because losses are less, Less area is required. Hence we have designed an active inductor. The inductive reactance of active inductor will vary with respect to the applied control voltage. As the control voltage is increasing the inductance will decrease. Also high frequency operation is achieved by cancelling the stray capacitance. The change in inductance with respect to the control voltage is given in the table shown in table 2.

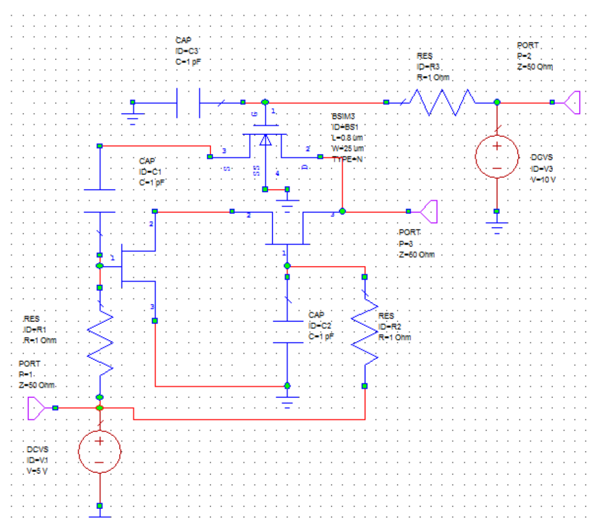

Fig. 9 Active inductor circuit schematic

TABLE II

INDUCTANCE VALUES FOR ACTIVE INDUCTOR

Voltage	Inductance
5	.8
6	.274
7	.253
8	.246
9	.241
10	.239
15	.223
20	.218

The third type of RTPS is with Active as load. The active inductor can vary with respect to its control voltage. As the voltage changes the inductance also changes. Hence the load impedance of the RTPS also changes. So a variable phase shift is obtained. The S matrix of RTPS with Active inductor load is given below

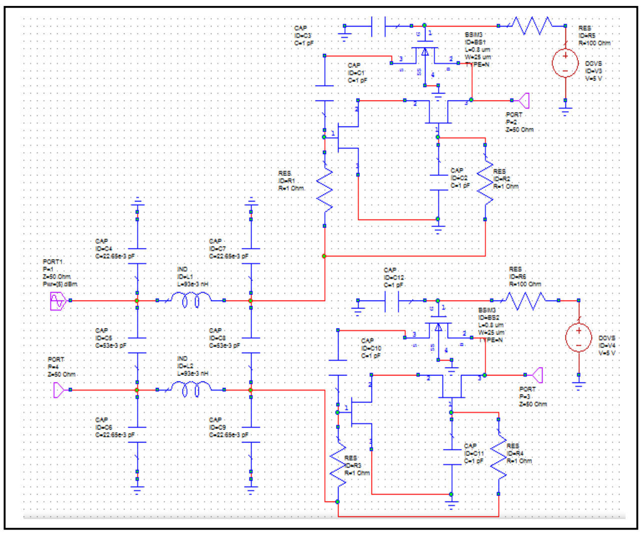

Fig. 10 RTPS with Active inductor load

$$\begin{bmatrix} .001\angle 10.77 & .021\angle -169.1 & .02\angle -79.33 & .953\angle 96.22 \\ .02\angle -169.5 & .8\angle -152.2 & .00002\angle 112.4 & .024\angle 79.27 \\ .02\angle -79.33 & .00002\angle 112.1 & .8\angle -152.2 & .01\angle -169.5 \\ .95\angle 96.22 & .02\angle 78.97 & .01\angle -169.1 & .001\angle 10.79 \end{bmatrix}$$

Comparing with the S matrix of ideal reflective type phase shifter |S11|=|S44|=.001 which approximately equal to zero. |S14|=|S41|=.95 where port 1 is the input port and port 4 is the output port. Maximum signal is reflected back to the output port. The phase shift obtained from the design is the phase of S41 and S14.This design can produce a variable phase shift.

$$\Delta\theta = 2\left(\tan^{-1}\left(\frac{X_m}{Z_0}\right) - \tan^{-1}\left(\frac{X_n}{Z_0}\right)\right) \qquad (6)$$

The theoretical phase shift range is .6 and the obtained phase shift range is .4. So the design is theoretically proved.

The reflection coefficients are S_{22} and S_{33}. The magnitude of Reflection coefficient is given by $|S_{22}|$ and $|S_{33}|$. Ideally S22 and S33 of RTPS is 1. For the above designed RTPS $|S_{22}|$ =.8, $|S_{33}|$ =.8. The transmission co-efficient is represented by the magnitudes of S23 and S32. Ideally transmission coefficient is zero. $|S_{23}|$= .00002, $|S_{32}|$= .00002 . Both $|S_{23}|$ and $|S_{32}|$ are having very small values approximately equal to zero.

Three types of Reflective type phase shifters with three different reflective loads are designed and their performance are compared.

TABLE III

COMPARISON OF DESIGNED RTPS

RTPS load	Phase shift	Insertion loss[dB]	Return loss[dB]
Spiral inductor	83.4	-21.7	-28
Varactor	90.5-122	-42.07	-41.89
Active inductor	96.22-95.84	-29.44	-59.23

IX. CONCLUSION

Three types of reflective type phase shifters at 60GHz were designed. Its performance are verified. Among the three designs two of them are of variable in nature. RTPS with MOS varactor load and RTPS with Active inductor load are the two variable phase shifter. MOS varactor and MOS active inductor are designed using the modelled MOSFET. The operating frequency of the model is 60 GHz. Gain is maximum at 60 GHz. By changing the control voltage the capacitance of varactor and inductance of active inductor changes. Hence the load impedance changes. RTPS produce some phase shift according to the change in load impedance. The losses are high for spiral inductor. Also the area required for fabricating spiral inductor is high. Varactor utilises less chip area. The phase shift range is highest for RTPS with varactor load among the three designs. But the losses are little high for RTPS with varactor load compared to that of RTPS with active inductor load.

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