

A Novel Dual-Frequency RF MEMS Phase Shifter

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Abstract—The operating principle of the RF MEMS phase shifter is described. A novel design concept for dual-frequency RF MEMS phase shifter is proposed. By using network theory and CST microwave studio simulation tool, the proposed RF MEMS phase shifter is modelled. Simulation results show that the proposed model has the return loss less than -10dB and the insertion loss more than -2dB in two operating frequencies, both of which are more than 1GHz bandwidth. For one RF MEMS phase shifter, both 180-degree and 90-degree phase shift are realized.

I. INTRODUCTION

With the development of the modern communication technology, the antennas and devices used in satellite communication or mobile communication are requested to be miniaturized and integrated. At the same time, the phase shifter, as a key part of the communication system, is supposed to be operating in different frequencies to reduce the EM interference and achieve miniaturization. Since its operating band and insertion loss will affect the anti-interference ability and the sensitivity of the devices directly. And the cost capacity and ponderance are also influenced ^[1, 2].

Meanwhile Radio Frequency Micro Electro Mechanical Systems (RF MEMS) phase shifters have been shown to provide very low loss compared to solid-state phase shifters for Ka band application. And Distributed MEMS Transmission Line (DMTL) phase shifters have been realized by loading a transmission line using MEMS bridges. When the analogue voltage is used between the centre conductor and the ground wire, the MEMS bridges will be pulled down for forming “up” or “down” states. As a result, the phase velocity is increased to change the phase shift ^[3, 4].

For this reason, a novel dual-frequency RF MEMS phase shifter used in Ka band is proposed in this paper. The equivalent circuit for the distributed MEMS transmission line phase shifter of Ka band is modelled. After the simulation and the analysis of the network of characteristics of the MEMS phase shifter, the phase shifter structure based on this model is

designed and the return loss and the insertion loss of each range is optimized, respectively.

II. THE PRINCIPLE OF THE RF MEMS PHASE SHIFTER

The phase shift of the DMTL phase shifter can be expressed as:

$$\Delta\Phi = \omega\sqrt{L_t C_t} \left(\sqrt{1 + \frac{C_{bd}}{sC_t}} - \sqrt{1 + \frac{C_{bu}}{sC_t}} \right) \quad (1)$$

In the above equation, C_{bu} is the capacitance of the MEMS bridge in the “up” state, while C_{bd} is in the “down” state. And s is the distance between neighboring bridges, L_t and C_t are the per unit length equivalent inductance and capacitance of the unloaded transmission line, respectively, ω is the angular frequency.

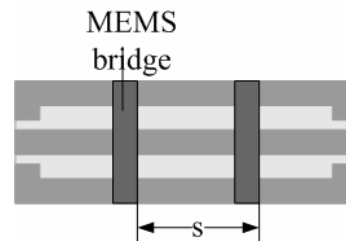


Fig. 1 Structure of the DMTL phase shifter from the top view

According to the transmission line theory, each line can be equivalent to be a circuit which is composed of capacitances and inductances, and the MEMS bridges loaded on the coplanar waveguide (CPW) can be seen as shunt capacitances in the circuit. As a result, the structure and the unit equivalent circuit model of the DMTL phase shifter with MEMS bridges are given in Figure 1 and 2, respectively.

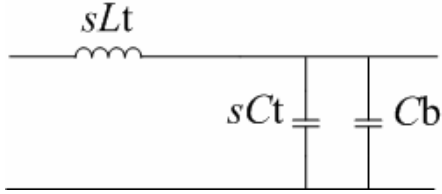


Fig. 2 Unit equivalent circuit model with the MEMS bridges

Here, C_b is the equivalent capacitance of the MEMS bridge. And L_t , C_t , C_b can be calculated through the following equations [4]:

$$C_t = \frac{\sqrt{\epsilon_{r,eff}}}{cZ_0} \quad (2)$$

$$L_t = C_t Z_0^2 \quad (3)$$

$$\epsilon_{r,eff} = \frac{\epsilon_r + 1}{2} \quad (4)$$

$$Z_0 = \frac{\eta_0 K(k')}{4\sqrt{\epsilon_{r,eff}} K(k)} \quad (5)$$

$$k = w/(w + 2g) \quad (6)$$

$$k' = \sqrt{1 - k^2} \quad (7)$$

In above equations, Z_0 is the equivalent resistance of the unloaded transmission line, $K(k)$ is the first elliptic integral, η_0 is the wave impedance in free space; w is the width of the centerguide, and g is the distance between the centerguide and the ground line.

The capacitance of the MEMS bridge can be computed through equation 8. ϵ_0 is the dielectric constant in free space, ϵ_{r1} is the dielectric constant of the insulation layer, h is the height of the bridge without any external force, t is the thickness of the insulation layer under the bridge, x is the position of the bridge when analogue voltage is used, A is the area of the bridge on the centerguide [5].

$$C_b = \frac{\epsilon_0 A}{h + \frac{t}{\epsilon_{r1}} - x} (1.25 - 0.20 \frac{x}{h}) \quad (8)$$

III. THE ANALYSIS OF THE DUAL-FREQUENCY PHASE SHIFTER

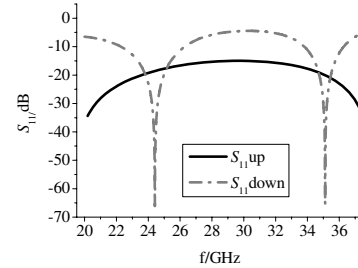
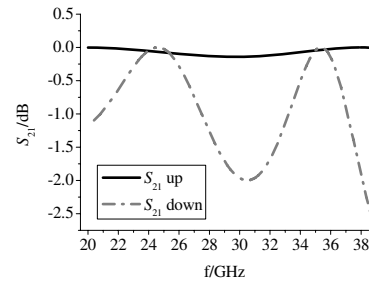
A. Design Considerations of the Dual-frequency Phase Shifter

Based on the former work [6], the insulation layer and the saw-shaped CPW is used in this design, which is used to increase the phase shift for each bridge and decrease the return loss of the device, respectively.

And the matrix A can be obtained through the unit equivalent circuit model discussed above. Then the total

matrix A can be computed through cascade matrix, and the total matrix A is transformed to matrix S . As a result, the S_{11} and S_{21} parameters can be optimized. After the optimization, silicon nitride with the relative dielectric constant 7.5 is chosen as the insulation layer below the bridges, its thickness is $0.5\mu\text{m}$, the width of the bridge is $25\mu\text{m}$, the height of the bridge in two state is $3\mu\text{m}$ and $0.5\mu\text{m}$, respectively, the distance between the bridges is $207\mu\text{m}$, the width of the centre conductor is $100\mu\text{m}$ and that of the saw-shaped CPW is $60\mu\text{m}$, the gap of the CPW and the saw-shaped CPW are $100\mu\text{m}$ and $120\mu\text{m}$, respectively. Therefore, a seven-bridge phase shifter is designed.

Figure 3 and 4 show the results of S_{11} and S_{21} based on the above analysis, respectively.


 Fig. 3 S_{11} of the circuit model analysis

 Fig. 4 S_{21} of the circuit model analysis

It can be seen that S_{11} is less than -10dB within the two ranges: between 22GHz and 28GHz and between 34GHz and 36GHz , while S_{21} is less than -1dB . And S_{11} is less than -20dB at the both centre frequencies: 24GHz and 35GHz .

B. Circuit Analysis of the Phase Shifter

However the analysis above is based on the theory of conformal transformation, and there may be some error since the CPW is equivalent to be inductances and capacitances for Ka band application.

As a result, further circuit analysis using the software ADS is needed.

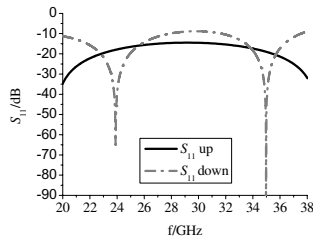


Fig. 5 S_{11} of the ADS analysis

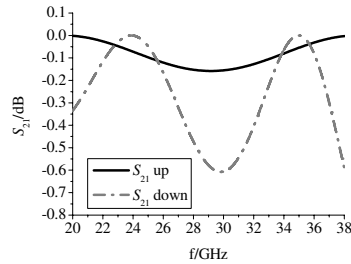


Fig. 6 S_{21} of the ADS analysis

The results of S_{11} and S_{21} of ADS analysis are given in Figure 5 and 6, respectively. It shows that S_{11} is less than -10dB within the two ranges: between 22GHz and 28GHz and between 34GHz and 36GHz, while S_{21} is less than -1dB. And S_{11} is less than -20dB at the both centre frequencies: 24GHz and 35GHz.

Comparing figure 3, 4 and 5, 6, the results of the circuit model analysis and those of the ADS analysis agree very well, so the correctness of the circuit model is verified.

Nevertheless, the analysis in Part II and III is based on the network theory. For Ka band applications, the transmission line loss, surface wave loss and radiation loss need to be considered in practice. For accurate analysis, the full wave EM analysis is necessary.

IV. FULL WAVE ANALYSIS OF THE RF CHARACTERISTICS

Figure 7 and 8 are the results of the return loss (S_{11}) and insertion loss (S_{21}) using full wave analysis with the help of CST software based on the MEMS phase shifter designed above, respectively, when the centre frequency is 26GHz, more or less. In the “up” state, the insertion loss of the phase shifter is more than -0.4dB between 24GHz and 28GHz, and the return loss is less than -20dB, while in the “down” state, the insertion loss of the phase shifter is more than -2dB between 25GHz and 27GHz, and the return loss is less than -10dB. And the return loss is less than -20dB at the centre frequency f_0 26GHz.

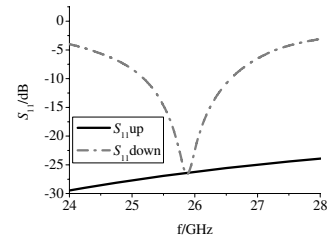


Fig. 7 Results of the return loss of the full wave analysis when $f_0=26$ GHz

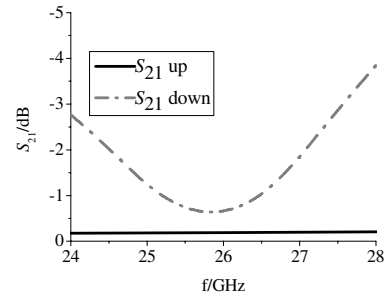


Fig. 8 Results of the return loss of the full wave analysis when $f_0=26$ GHz

Figure 9 and 10 show the results of the return loss (S_{11}) and insertion loss (S_{21}) using full wave analysis, respectively. The return loss of the phase shifter is less than -20dB between 31GHz and 38GHz, and the insertion loss is more than -1dB in the “up” state, while in the “down” state, the return loss of the phase shifter is less than -10dB between 34.5GHz and 35.5GHz, and the insertion loss is less than -2dB. And the return loss is less than -20dB at the centre frequency f_0 35GHz.

It can be seen that results of the full wave analysis and those of ADS circuit analysis agree well, though one of the two centre frequencies changes a little due to the loss of Ka band mentioned above, which cannot be analysed through the circuit analysis.

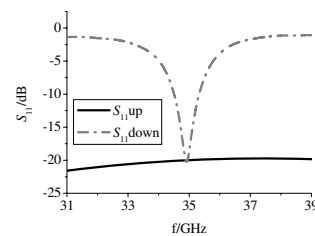


Fig. 9 Results of the return loss of the full wave analysis when $f_0=35$ GHz

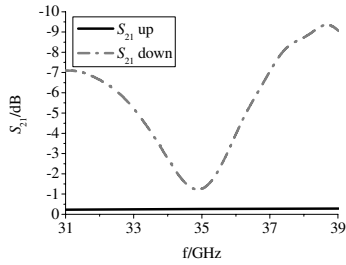


Fig. 10 Results of the insertion loss of the full wave analysis when $f_0=35$ GHz

Figure 11 and 12 are the results of the phase shift of the full wave analysis in two ranges, respectively. Results show that the phase shift is nearly linearity in the two ranges: from 25GHz to 27GHz and from 34.5GHz to 35.5GHz. 180-degree and 90-degree phase shift with only 7 bridges are obtained, respectively. Therefore, the performance improved a lot compared with the conventional phase shifter.

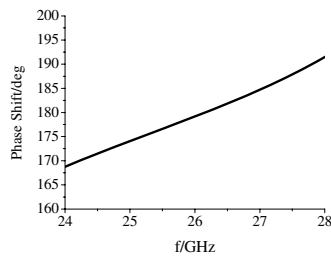


Fig. 11 Results of the phase shift of the full wave analysis when $f_0=26$ GHz

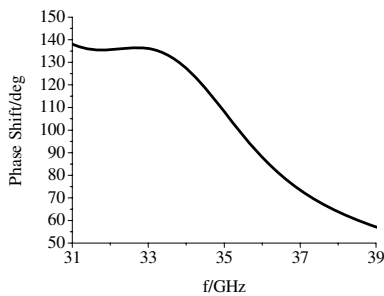


Fig. 12 Results of the phase shift of the full wave analysis when $f_0=35$ GHz

V. CONCLUSIONS

In this paper, a Ka band RF MEMS phase shifter operating at two different frequencies is proposed. Compared with the conventional phase shifter, the return loss at the two center frequencies 26GHz and 35GHz can reach -20dB, and is less than -10dB in more than 1GHz bandwidth, while the insertion loss is more than -2dB. For realizing 90-degree and 180-degree phase shift, only one phase shifter is needed, leading to decrease the total size of the phase shifter and reduce the EM interference. This new structure is significant for the design of dual-frequency RF MEMS phase shifter and reduce the EM interference for Ka band applications.

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