W-Band Microstrip RF-MEMS Switches and Phase Shifters

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Abstract— **This paper presents W-band microstrip shunt switches on quartz substrates. The −20 dB isolation bandwidth of the switches is limited to around 10% due to the $\lambda_g/4$ radial stubs used, with an insertion loss of −0.2 to −0.5 dB. The microstrip switches are used in a switched-line phase shifter and a reflect-line phase shifter at W-band frequencies, and resulted in good performance over a 5-10% bandwidth.**

Keywords— **MEMS, microwave, millimeter-wave, phase shifters, switches.**

I. Introduction

Shunt and series RF MEMS switches have been demonstrated from 1-120 GHz [1], [2]. While series switches are mainly used from DC-40 GHz, capacitive shunt switches are better suited for 10-120 GHz applications. This paper presents W-band microstrip capacitive shunt switches on quartz substrates with $\lambda_g/4$ radial stubs instead of via-holes to ground. This is similar to the Raytheon work on Ka-band phase shifters [3]. T-match and π-match circuits were designed with very low insertion loss (−0.2 to −0.5 dB) and good isolation response (−20 dB) but they suffer from a relatively narrow bandwidth (5-10 %) due to the $\lambda_g/4$ radial stub. The switches are used to build, for the first time, a one-bit 0/180° switched-line phase shifter and a 2-bit 0/90/180/270° reflect-line phase shifter at W-band frequencies with good performance over a 10% bandwidth.

II. W-band Microstrip MEMS Switches

A. T-match Designs

Microstrip designs offer easier biasing of the individual switches, especially if a large number is used. However, they require via-hole technology and very thin wafers to obtain a low-inductance via-hole to ground. One way to get around the via-hole is to use a $\lambda_g/4$ open stub in order to simulate an RF-short (Fig. 1). The substrate used is 150 μm-thick quartz with $h \leq \lambda_d/10$ to eliminate higher-order modes. The SiCr bias line (1000 Ω/sq) is connected to the low-impedance point of the stub so as to have a minimal effect on the RF signal. The lines are 1.3 μm-thick and result in an attenuation of 0.3 dB/cm for a 50 Ω-line at 90 GHz. The switch is 40 μm wide, 280 μm long, 8,000 Å-thick and is suspended 1.6 μm above the t-line. For a 40 × 70 μm^2 pull-down electrode, the parallel plate up-state capacitance is $C_{upp} = \epsilon_o A/g = 15$ fF and the total up-state capacitance is around 19 fF (30% fringing capacitance). The Si_3N_4 dielectric layer underneath the MEMS bridge is 1,500 Å-thick, resulting in a maximum down-state capacitance of 1.2 pF but, due to dielectric roughness, the down-state capacitance is 300-500 fF. The pull-down voltage is around 40 V.

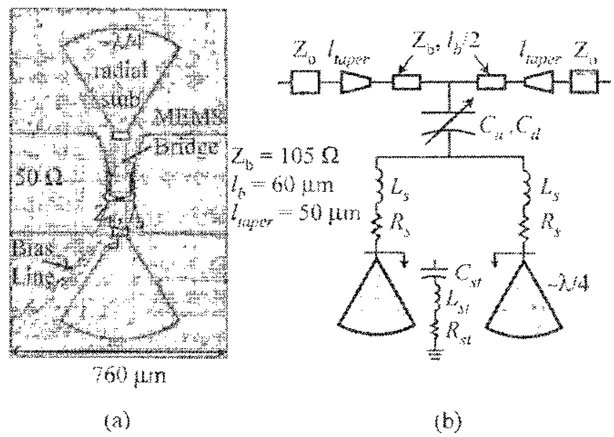

Fig. 1. Microstrip shunt capacitive MEMS switch with two stubs (a) and its equivalent circuit circuit model (b).

In the up-state position, the switch has a low up-state capacitance and results in very low insertion loss. The radial stub has a little effect on the performance of the switch in the up-state position due to the low capacitance value of the bridge. However in the down-state position, the LC resonant frequency is given by the series combination of the radial stub and the MEMS switch. In fact, the radial stub, around its resonant frequency, is equivalent to a series LC tank, and the down-state resonant frequency will depend on CLR series combination of the MEMS bridge and the radial stub. For a 320 μm-long radial stub, the fitted LC model has $L_{st} = 90$ pH and $C_{st} = 37$ fF.

The isolation bandwidth of a single-stub design is very narrow (3-4%) due to the stub frequency response. One can improve the switch isolation by using two stubs at both ends of the bridge. The total series resistance to ground will be cut by half because the currents have two paths to the RF-ground. Also, the total stub inductance will be divided by 2 and the total stub capacitance will be multiplied by 2 giving the same resonance frequency (ω_o). The main effect of the two-stub design is to double the isolation bandwidth.

Fig. 2 presents the measured results for a switch with two 320 μm-long stubs. The reflection loss is better than

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−15 to −20 dB over the entire W-band frequencies and the insertion loss is around −0.2 dB (the reference planes are 760 μm apart as shown in Fig. 1). The 88-93 GHz range is not shown because of poor calibration in this frequency range. The microstrip line is tapered underneath the bridge from 340 μm (50 Ω) to 70 μm (100 Ω) resulting in an inductive T-match for the switch (Fig. 1b). Full-wave Sonnet [4] simulation predicts a reflection loss better than −25 dB from 75-105 GHz for a bridge height of 1.6 μm (C_u = 19 fF), but the reflection loss measurements are only accurate to −20 dB due to TRL calibration accuracy at W-band.

In the down-state position, Sonnet simulations agree well with the measurements and the measured maximum isolation is −25 dB with a −20 dB isolation bandwidth of 4 GHz. The relative permittivity used for silicon nitride in Sonnet simulations is $\epsilon_r = 2.7$ (C_d = 450 fF) to account for the roughness of the dielectric underneath the MEMS bridge. In this case, the resonant frequency is mainly given by the radial stub capacitance (C_{st} = 37 fF) and is not significantly affected by changing C_d from 300 to 500 fF (or ϵ_r from 1.8 to 3). The ADS [5] *fitted* model results in C_d = 450 fF, L_s = 15 pH and a total resistance $R_s + R_{st} = 3$ Ω (Fig. 1). The relatively high resistance is due to radiation from the radial stubs and the bridge resistance (∼ 1 Ω). Sonnet predicts a little bit higher isolation (−28 dB or $R_s + R_{st} = 2$ Ω).

B. π-match designs

The bandwidth of the MEMS microstrip switch can be further improved using a π-matching network (Fig. 3). The stubs are 240 μm-long and the 75 Ω-line (180 μm-wide) matching line is 500 μm-long. The tapered line and the small section underneath the MEMS bridges contribute also to the π-matching network. Fig. 4 shows the measured S-parameters in the up-state and down-state positions. In the up-state position, the reflection loss is better than −15 dB from 75 to 92 GHz and the insertion loss is around −0.5 dB. In the down state position, an isolation of −20 dB is achieved from 82-91 GHz (10%). In this case, Sonnet full-wave simulations and the measured results agree very well. However the *fitted* ADS circuit model does not predict well the behavior of the circuit in the up and down-state positions. This is due to the resonant structures used and the small but significant coupling between the different elements of the π-network.

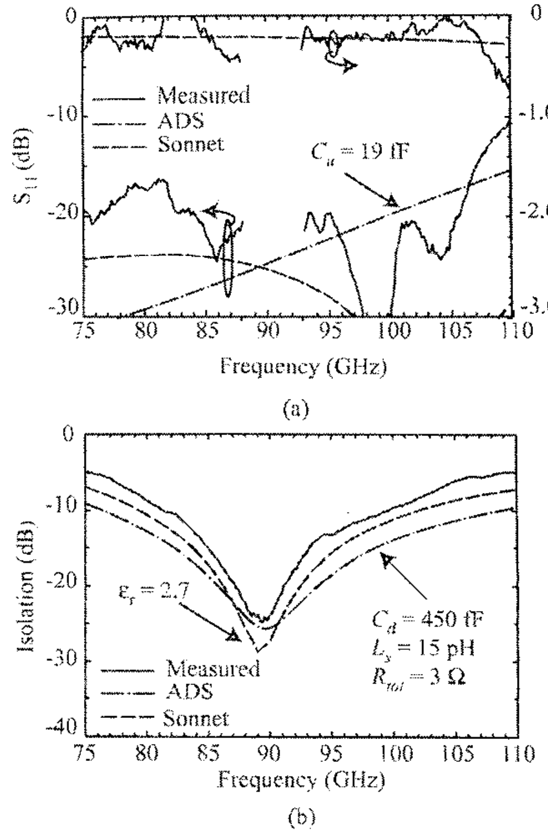
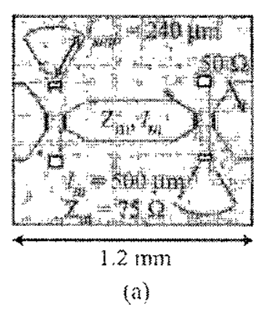


Fig. 2. Measured S-parameters of a T-match microstrip switch with two 320 μm-long stubs: (a) in the up-state position and (b) in the down-state position.



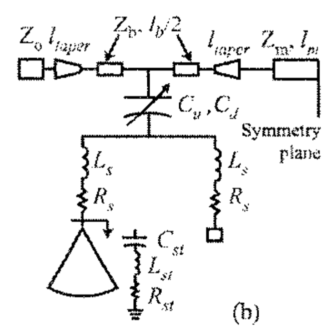


Fig. 3. Microstrip π-match design (a) and its equivalent circuit model (b).

III. W-BAND PHASE SHIFTERS

Many MEMS phase shifters have been developed in the last 3-4 years [6], but little work has been done at W-band frequencies [7]. This section discusses the design and measurements of W-band switched-line and reflect-line microstrip MEMS phase shifters on quartz substrates. The switched-line design is a one bit phase shifter, switching between 0 and 180°. The reflect-line design is a 2-bit phase shifter with 0/90/180/270° states.

A. One-bit switched-line phase shifter

Fig. 5 presents a 0/180° switched-line phase shifter. The operation of the phase shifter is similar to the Ka-band Raytheon design [3]. Instead of placing the MEMS bridges $\lambda_g/4$ away from the T-junction, they are placed $3\lambda_g/4$ away since, for the $\lambda_g/4$ case, the radial stub is very close to the T-junction and results in significant coupling. The difference in line length between the two paths is $\lambda_g/2$. If the

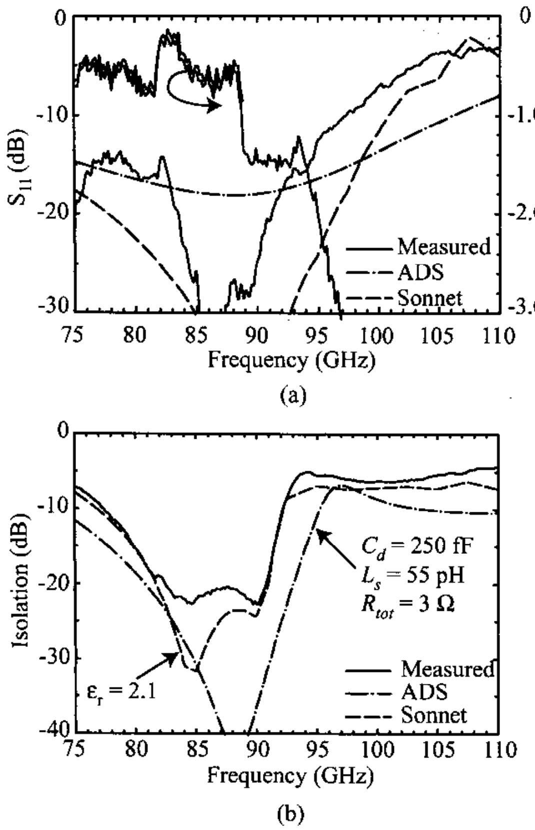


Fig. 4. Measured S-parameters of a π-match microstrip switch with two 240 μm-long stubs: (a) in the up-state position and (b) in the down-state position.

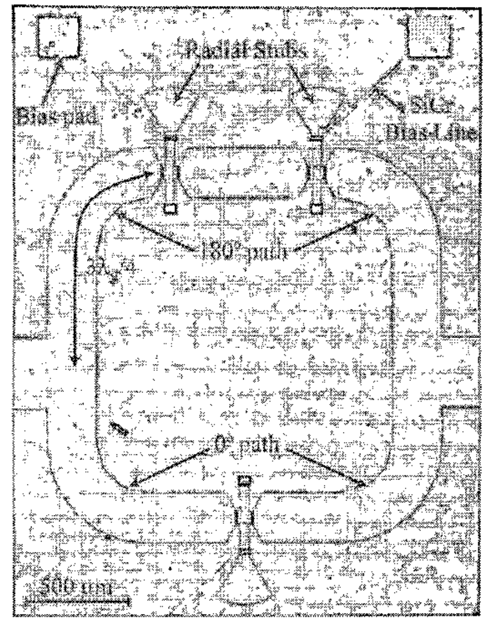


Fig. 5. Microstrip W-band 0/180° switched-line phase shifter (2.3 × 2.5 mm^2).

bridge is pulled down in the lower path, it creates an open-circuit at the T-junction, and the signal passes by the upper path with a phase delay of 180°. It is clear that this design will be narrow-band due to the radial stubs and the $3\lambda_g/4$ approach.

Fig. 6 shows the measured response for the 180° phase shifter. The reflection loss is better than -10 dB from 90-100 GHz and the average insertion loss is -2.5 dB. The loss is a bit high for a microstrip line loss of 0.3 dB/cm and we believe that it is mainly due to radiation in the structure.

B. Two-bit reflect-line phase shifter

A reflect-line phase shifter (Fig. 7) consists of a 3-dB coupler and identical transmission lines at the thru and coupled ports loaded with MEMS switches. When the MEMS switches are actuated in pairs, they create a short circuit to ground and the microwave signal is reflected back to the coupler. Having identical lines will insure that the reflected signals will add coherently at the "isolated port" of the 3-dB coupler and which is now considered as the output port of the reflect-line phase shifter.

Fig. 7 presents a W-band microstrip reflect-line phase shifter. The 3-dB coupler is a 90° branch-line coupler with a characteristic impedance of 97 Ω. The coupler is matched to 50 Ω with a 70 Ω $\lambda/4$ transformer (not shown in the picture). This phase shifter is fabricated on a 150 μm-thick quartz wafer.

Fig. 8 presents the measured S-parameters and the phase-shift of the 2-bit reflect-line phase shifter. The problem in this design is its relatively narrow bandwidth (80-85 GHz) and the errors in the phase shifts. The narrow bandwidth is due to the fact that the 3-dB coupler results in a good match if its "thru" and "coupled" ports are matched to 50 Ω. However, in this design, these ports present a reactive reflection to the coupler. Also, the phase shifts at the output of the reflect-line design are not equal to the phase shifts presented at each of the coupler arms and careful design and optimization should be done in order to achieve an accurate response. The measured phase shifts at 82 GHz are 0, 66° (90° bit), 168° (180° bit) and 282° (270° bit), and are within two-significant bits (±22.5°) of the two-bit design. In any case, this design demonstrates an average insertion loss of -3 dB at 80-85 GHz. This means that a loss of 1.5 dB/bit is achieved with the reflect-line phase shifter and is 1.6 dB/bit better than solid-state state-of-the-art phase shifters presented in literature [8]. More optimization should be done in order to get a wider bandwidth and the correct phase shifts.

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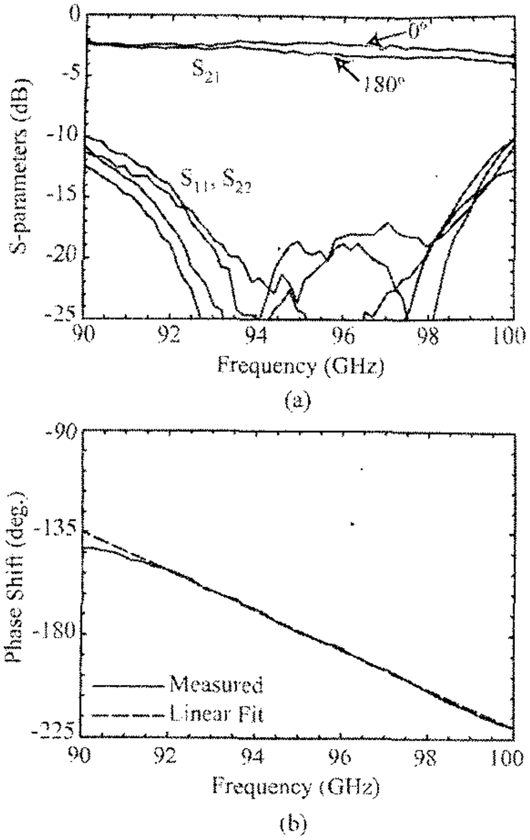

Fig. 6. Microstrip W-band 0/180° switched line phase shifter measured performance: (a) S-parameters and (b) Phase shift in degrees.

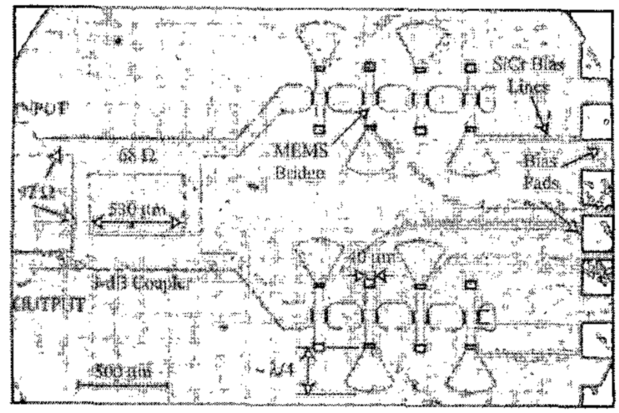

Fig. 7. W-band 2-bit MEMS reflect-line phase shifter in microstrip implementation.

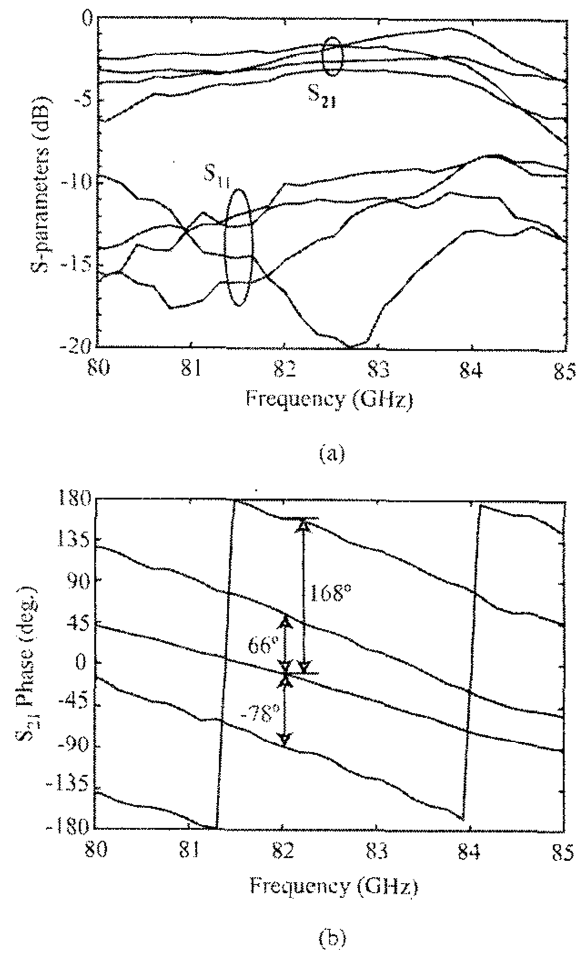

Fig. 8. Measured S-parameters and phase shift of the W-band 2-bit MEMS reflect-line phase shifter in microstrip implementation.

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