New Developments on Polymer-based MEMS Phase Shifters

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ABSTRACT

Phase shifter is one of the core components for phased-array antennas which find wide applications in satellite systems, telecommunications, wireless systems, radar systems and tracking systems. The current high-frequency phase shifters are mainly built upon active semiconductor technologies which suffer from high fabrication and assembly costs. In this work, we propose a cost-effective approach to fabricate distributed MEMS transmission line phase shifters on a polymer-glass substrate. The fabrication procedure is compatible with CMOS and post-CMOS processes. The polymer-glass phase shifter was characterized from DC to 26 GHz. The measurements show a phase shift of 120° and a 2.5-dB insertion loss at 26 GHz.

INTRODUCTION

Phase shifters are essential components for many modern communication and radar systems. Conventionally, phase shifters are based on semiconductor or ferrite technologies [1]. Although these technologies proven to be effective, they suffer from some major drawbacks. The high material and fabrication costs, high RF losses and large dimensions increase difficulties for large-array applications. The RF MEMS approach of building phase shifters places a series of distributed MEMS bridges over a coplanar waveguide transmission line (DMTL) [2]. The bridge height is effectively varied by electrostatic forces through biasing MEMS parallel plates. The varied MEMS bridge capacitances slow down the phase velocity and create phase shifts. The DMTL phase shifter has demonstrated higher performance, lower insertion losses, lower power consumption and simpler fabrication processes compared to its semiconductor counterparts [2–3]. However, DMTL phase shifter and its extended approaches are mainly built on quartz [2], intrinsic silicon [1,4] and III-V compound [5] substrates taking advantages of their high resistivity. The material and assembly costs are high.

Comparing to traditional semiconductor materials, polymers are much more cost-effective. Polymers have shown low insertion losses at high frequencies [6–8]. The low dielectric constants, high resistivities and low dissipation factors [9,10] are very attractive for high quality-factor RF applications. Processing polymer does not involve toxic materials nor high thermal budget for annealing. The high thermal stability, high degree of planarization and chemical resistances are compatible with existing IC processing technologies [9,10]. The light weights of polymers are ideally suited for applications where a large number of phase shifters are required. Polymers can also be conformal reducing processing complexity. Many polymers have versatile micromachining abilities which are very attractive for MEMS applications. 3-D micro structures can be realized by photolithography, deep reactive ion etching (DRIE), plasma etching [11] or hot embossing processes.

Due to the superb performance of the RF MEMS approach and advantages of using polymer materials, we propose to build distributed MEMS transmission line phase shifters on a polymer substrate. In this paper, design method, fabrication sequences and measurement results will be discussed in details.

DESIGN CONSIDERATIONS

Figure 1. An overall sketch of a RF MEMS phase shifter.

Substrate and Polymer Dielectric Layer

Unlike intrinsic silicon wafers and III-V compound materials, glass substrates are more cost-effective. Hence, in this work, glass slide is chosen as the substrate material. The polymer will form a layer of insulator to prevent field leakage into the substrate so the resistivity of the glass substrate becomes less important.

Among many polymers, benzocyclobutene (BCB) has a dielectric constant of 2.65, dissipation factor of 0.08% and resistivity of $10^9 \Omega\text{-cm}$. The annealing temperature is 250°C which is below most existing IC thermal budgets [9]. Recent years, researches showed that the BCB has very low insertion losses in the applications of coplanar waveguide transmission lines. For example, an insertion loss of 1 dB/cm at 40 GHz has been demonstrated [6–8]. Due to these outstanding RF properties, BCB is chosen as the dielectric material. Fig. 1 shows an overall sketch of a DMTL phase shifter where the BCB polymer is applied on top of the glass substrate as the dielectric layer.

Coplanar Waveguide Transmission Line and MEMS Bridges

The coplanar waveguide (CPW) transmission line is the foundation block of many RF MEMS devices such as switches, varactors and phase shifters. The uniplanar attribute of CPW lines provides convenience for MEMS structure construction and on-wafer characterization [12]. Fig. 2 shows the basic structure of a CPW line and its unit-length equivalent circuit [2], where L_t and C_t are unit-length inductance and capacitance, respectively.

Figure 2. (a) The CPW line structure. (b) The unit-length equivalent circuit of an unloaded CPW line.

For most RF applications, minimizing the transmission loss is one of the primary goals. Two major factors are the reflection and dissipation losses. To minimize the reflection loss, the characteristic impedance Z_o needs to match with that of the rest of the circuitry. In our case, the circuitry network has a characteristic impedance of 50Ω . Eq. 1 shows the relationship between the characteristic impedance Z_o , unit-length inductance L_t , capacitance C_t and the effective dielectric constant ϵ_{eff} for a CPW line without loading any MEMS bridges. Eq. 2 indicates the phase velocity traveling in a CPW line.

$$Z_o = \sqrt{\frac{L_t}{C_t}} \text{ and } C_t = \frac{\sqrt{\epsilon_{eff}}}{cZ_o} \quad (1)$$

$$V_u = \frac{1}{\sqrt{L_t C_t}} \quad (2)$$

The dissipation loss consists of three mechanisms: 1) the skin effect, 2) metal conductive losses, and 3) dielectric losses. By properly choosing a good conductor, a micro-scaled MEM device can have very low skin effect and conductive losses. The dielectric loss is inherently low due to the usage of glass substrate and BCB polymer.

The characteristic impedance is decreased when loading a CPW line with MEMS bridges. It adds a shunt capacitance, C_b , in parallel with the unit-length capacitance, C_t . Fig. 3 illustrates a CPW line with MEMS bridges loaded and its unit-length equivalent circuit. The capacitance of the loaded CPW transmission line is modeled to increase to $C_l = C_b/s + C_t$ [2], where s is the spacing between two adjacent MEMS bridges. The characteristic impedance of the loaded CPW line is decreased by the C_b , as shown in Eq. 3 [2]. Therefore, the dimensions of an unloaded CPW line need to be carefully designed so that the characteristic impedance of the loaded CPW line can be close to 50 Ω . In our design, the CPW line has a signal width (W) of 100 μm , signal-ground spacing (S) of 100 μm , similar to *Barker's* design in [2]. The unloaded characteristic impedance was calculated to be around 90 Ω .

Figure 3. (a) The loaded CPW line structure. (b) The unit-length equivalent circuit of a loaded CPW line.

$$Z_l = \sqrt{\frac{L_t}{C_t + C_b/s}} \quad (3)$$

The MEMS capacitance slows down the wave velocity, as indicated by Eq. 4 [2]. Thus a phase shift is introduced by the slowed wave velocity. Eq. 5 describes the relationship between the phase shifting value $\delta\phi$, unloaded phase velocity V_u and loaded phase velocity V_l , where f is the frequency and l is the device length [3].

$$V_l = \frac{1}{\sqrt{L_t(C_t + C_b/s)}} \quad (4)$$

$$\delta\phi = 2\pi fl \left(\frac{1}{V_u} - \frac{1}{V_l} \right) \quad (5)$$

$$C_b = \epsilon_o \frac{A}{d} \text{ and } A = W \times w \quad (6)$$

The capacitance of each MEMS parallel-plate bridge is decided by its dimensions in Eq. 6, where ϵ_o is the free-space dielectric constant, A is the bridge area overlapped with the CPW signal line, W is the CPW signal line width, w is the bridge width, and d is the height. The parallel-plate capacitance model does not take the edge effects into account. A biasing voltage produces an electrostatic force pulling the MEMS bridges towards to the CPW signal line. Thereby, a larger MEMS capacitance is achieved. In our design, a bridge with a 60- μm width is chosen after trading off the elements of phase shifts, losses and characteristic impedance.

Other Considerations

The relationship between the biasing voltage V_{bias} and bridge height d is shown in Eq. 7 [3], where d_o is the original bridge height, and k is the spring constant of the bridge membrane which is determined by the material strength, bridge size and surface residue level. A larger original bridge height d_o will produce a larger range of MEMS capacitance variation, however a large biasing voltage is needed as well to actuate the parallel plate, as implied by Eq. 7 [3]. A smaller bridge height needs less biasing voltage. Nonetheless, a small bridge height is susceptible to electrical noises and false operations may be introduced. After carefully evaluating these contradicting elements, a bridge height of $0.744 \mu\text{m}$ is designed.

$$V_{bias} = \sqrt{\frac{2k}{\epsilon_o W w}} d^2 (d_o - d) \quad (7)$$

Only one third of the original bridge height (d_o) can be used to achieve variable capacitances. Beyond the one-third limitation, the MEMS bridges will collapse onto the CPW signal line. This phenomenon is called the “pull-in” condition. The voltage at which the pull-in condition happens is called the “pull-in voltage”, as described by Eq. 8 [3].

$$V_p = \sqrt{\frac{8k}{27\epsilon_o W w}} d_o^3 \quad (8)$$

The operational frequency is limited by the *Bragg* frequency, beyond which the CPW line becomes reactive and the return loss increases significantly [2, 13]. The *Bragg* frequency is described in Eq. 9.

$$f_{Bragg} = \frac{1}{\pi s \sqrt{L_t (C_t + C_b / s)}} \quad (9)$$

Eqs. 6 and 9 indicate a larger bridge width (W), larger bridge spacing (s) or lower bridge height (d) results in a lower *Bragg* frequency. In this work, the designed *Bragg* frequency is above 60 GHz, which is more than twice of the upper limit of our measurement ability (26 GHz).

The phase shifter was characterized by two GSG probes which have a $150\text{-}\mu\text{m}$ spacing and a $50\text{-}\mu\text{m}$ tip width. The phase shifter CPW line consists of a $100\text{-}\mu\text{m}$ spacing and a $100\text{-}\mu\text{m}$ signal width. In order to accommodate the dimensional differences and avoid an abrupt impedance change, a taper type matching circuit was designed, as shown in Fig. 4. The insertion loss of such a matching circuit is estimated below 0.15 dB/cm [8, 14].

Figure 4. The taper impedance transition.

FABRICATION PROCESS

A 10- μm BCB polymer layer was applied on a glass substrate using the spin-on process and inert gas annealing process (Fig. 5(a)). Coplanar waveguide transmission lines were defined on top of the BCB polymer using a sandwiched metal layer of chromium-aluminum-chromium (Fig. 5(b)) with a thickness of 500-8000-500 Å. The two layers of chromium function as an adhesion material. Without such adhesive materials, the aluminum would be easily peeled off in later processes. A silicon nitride layer of 3000 Å was grown on the CPW signal line using the RF sputtering technique (Fig. 5(c)). This silicon nitride layer will prevent MEMS bridges from touching the CPW signal line and resulting in electrical short circuit. A sacrificial layer was applied to define the bridge height using positive photoresist (Fig. 5(d)). A copper layer was deposited using thermal evaporation, followed by photolithography and etching processes to form the MEMS bridge patterns (Fig. 5(e)). In the final step, the devices were soaked into photoresist stripper solution for release (Fig. 5(f)).

Figure 5. Fabrication processes. (a) Polymer deposition. (b) CPW transmission line formation. (c) Silicon nitride growth. (d) Sacrificial layer deposition. (e) Bridge formation. (f) Device release.

Figure 6. (a) A microscopic photo of a phase shifter on the polymer dielectric layer. (b) A SEM photo of a MEMS bridge over a CPW signal line.

EXPERIMENTAL RESULTS

The DMTL phase shifter was successfully fabricated on a BCB-glass substrate. Fig. 6(a) shows a phase shifter photo under the microscope and Fig. 6(b) shows a SEM picture of a MEMS bridge over the CPW signal line. The dimensions of the RF MEMS phase shifter are shown in Table 1, where W is the CPW signal width, S is the CPW signal-ground spacing, s is the spacing between two adjacent bridges, w is the bridge width, n is the bridge number and d_o is the original bridge height.

Substrate	W (μm)	S (μm)	s (μm)	w (μm)	n	d_o (μm)
BCB-Glass	100	100	400	60	16	0.744

Table 1. Phase shifter dimensions.

The device was evaluated by an *Agilent 8510* network analyzer and two *Cascade Microtech Infinity GSG-150* probes from DC to 26 GHz. The scattering parameter measurements were calibrated by a *Cascade Microtech ISS-005-016* impedance standard substrate. The DC actuating voltages were applied by an *HP4142B* DC source through an *HP11612* biasing network.

Figure 7. Experimental results of the phase shifter. (a) Phase shifts versus frequency with different biasing voltages. (b) Insertion and return losses from DC to 26GHz.

Fig. 7(a) shows that at a 10-V biasing voltage, the phase shift reaches 18° at 26 GHz, whereas the phase shift is 95° at the same frequency at a 20-V biasing voltage. The phase shifts demonstrates a linear increase versus frequency under both biasing voltages. Our testing equipment is limited below 26GHz. This linearity implies that the phase shifting values will keep increasing above 26 GHz. Fig. 7(b) shows the scattering coefficients s_{11} and s_{21} of the phase shifter at different actuation voltages. The insertion losses are below 2.5 dB up to 26GHz regardless of biasing voltages.

The maximum achievable phase shift is limited by the pull-in condition. Under the pull-in condition, the calculated phase shift is 140° at 26 GHz using Eqs. 4-8. The experimental result shows, with a biasing voltage of 25 V, the bridge height is reduced to $0.5 \mu\text{m}$ which is measured by a *Veeco* optical profiler. The bridge height reduction is very close to one third of the original bridge height. The phase shift at near the pull-in situation is 120° . The discrepancy may be due to the non-zero surface stress residue of the bridges and the Young's Modulus is higher than the predicted one based on zero-surface stress. The loaded characteristic impedance Z_l can be extracted from the first peak of the return-loss spectral curve where the phase shifter behaves as a quarter-wavelength transformer [2]. Using this Z_l value, the MEMS bridge capacitance is calculated by Eq. 3. The *Bragg* frequency is then extracted from Eq. 9. Under this near pull-in condition, the Z_l is 38Ω , C_b is 120 fF and the *Bragg* frequency is 63 GHz, respectively.

CONCLUSIONS

In this paper, we investigated the design method, fabrication sequences and measurement results of DMTL phase shifters on polymer substrates. Our measurement indicates that the phase shift reaches 120° at 26 GHz. The insertion loss is less than 2.5 dB. The fabrication processes use low-cost materials and are compatible with CMOS processing.

ACKNOWLEDGEMENT

This research is sponsored by the National Science Foundation DMI #0428884. The authors would like to express their gratitude to Dr. N. Scott Barker for his help and valuable suggestions on fabrication, Dr. Ronald Carter for his help on characterization supports, and Dr. Nasir Basit and Mr. Dennis Bueno for their cleanroom assistance.

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