LOW LOSS MICROMACHINED FILTERS FOR MILLIMETER-WAVE TELECOMMUNICATION SYSTEMS

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Abstract— **This paper describes the design and the realization of two membrane supported microstrip millimeter-wave planar bandpass filters. Both filters exhibit transmission zeros and a 2.3 dB port-to-port insertion loss for the 37 GHz 3.5% bandwidth 2-pole filter and a 1.5 dB insertion loss for a 60 GHz 8% bandwidth 4-pole filter. The use of membrane technology allows a significant reduction of insertion loss, combined with a reproducible, low cost fabrication process.**

I. INTRODUCTION

MILLIMETER -wave communication systems are expanding rapidly as they offer many advantages over conventional wireless links. They allow the use of very wideband radio links suitable for inter satellite communications and personal communications.

We have focused on 38 GHz radio links for base station communications in PCS networks and upcoming 60 GHz multipurpose telecommunication systems [1] [2]. Waveguide technology results in the best performance, but waveguides suffer from high production costs and result in bulky systems. For conventional passive components (microstrip, CPW), the dielectric and radiation loss in the substrate reduce the filter resonator unloaded Q resulting in high insertion loss and poor rejection performance. It is therefore difficult to build filters with narrow bandwidth, sharp skirts and reasonable insertion loss using microstrip lines. Silicon or GaAs micromachining techniques suppress most of these drawbacks by replacing the substrate with a thin dielectric membrane. In these circuits, the propagation media is air, and the whole circuit or subsystem, including packaging, is fabricated in a collective process. In the past, micromachining techniques have been successfully applied to K and W Band microstrip membrane supported filters [8][9], and have lead to very good results. In this paper, we report conception, fabrication, and measurements of planar self-packaged bandpass filters. Transmission zeros are highly desirable in many applications to achieve a sharp out of band rejection with a reduced number of poles compared to conventional Chebyschev functions. Self packaging using micromachined cavities and vias-grooves allows to reduce ohmic loss in the circuit since it is possible to use large dimensions without increasing radiation loss. Moreover, it is important to note that the membrane filters can be easily combined with active devices (low noise amplifiers, power amplifiers) and high efficiency planar antennas. This will result in very low-loss integrated front-ends for millimeter-wave applications.

II. 37 GHz 2 POLE FILTER

A. Fabrication

A stress compensated 1.4 μm membrane layer consisting of $SiO_2/Si_3N_4/SiO_2$ is deposited on a high resistivity 525 μm thick silicon substrate using thermal oxidation and low pressure chemical vapor deposition. After the membrane is deposited, the circuit is patterned on the top side of the wafer using standard 2.5 μm gold electroplating technique. Next, the silicon is completely etched under

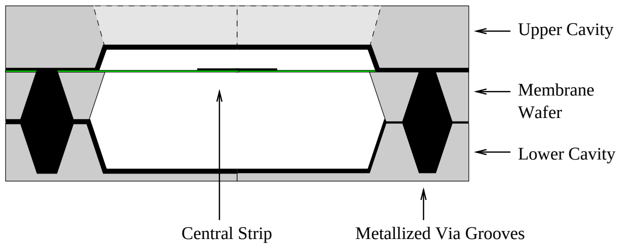

Fig. 1. Transverse section of the microstrip structure

the filter circuit until the filter is left standing on the thin dielectric membrane. At the same time, via grooves are opened all around the filter circuit in order to ensure a complete shielding of the structure. The lower cavity is etched in a 525 μm low resistivity wafer and metalized with gold. The upper cavity is formed by double side etching. First, a selective etch is performed on the upper side of the wafer to begin probe window openings. Next, the lower side is patterned and the wafer is etched 200 μm on both sides to open the probe windows and to form the upper cavity. The upper cavity is also metalized with gold. The three wafers are stacked together and silver epoxied to form a completely shielded, self packaged circuit, as shown in fig. 1.

The operating mode is microstrip with wide dimensions. The strip width was chosen to be 700 μm with a ground plane spacing of 200 μm.

B. Single resonator and 37 GHz filter

A single microstrip resonator has been fabricated in order to measure its unloaded Q. We have obtained an unloaded Q of 420 at 37 GHz. The basic principle of operation for the filter is the same as presented in [3]. The input / output ports are coupled capacitively while the hairpin resonators are coupled magnetically. This principle has been widely used [5] [6] [7] in both planar and non planar devices in the past. The inter resonator coupling value has been computed from a two poles Chebyschev 0.1 dB ripple prototype. Physical dimensions

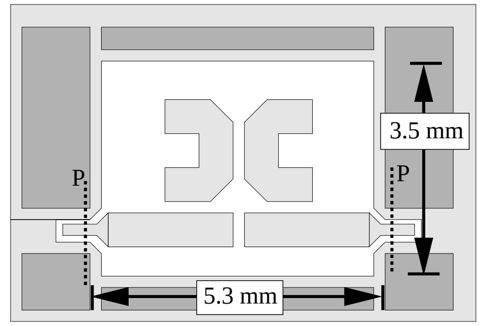

Fig. 2. 2 pole filter layout. The dark areas show the via grooves locations. P denotes the calibration planes.

have been computed using HP Momentum in a similar manner to full wave coupled cavity filter design. First, the single resonator resonance frequency is adjusted using full wave simulation. Next, the inter-resonator coupling is adjusted in a quasi-free oscillation configuration. In that case, input and output couplings are maintained very low and the coupling value can be computed from the odd and even mode frequencies of the two resonators. For the feed lines, the gap has been adjusted with full wave simulator. Resulting filter layout is shown fig. 2.

Measurements were done using an HP 8510C network analyzer Calibration was done using a TRL procedure with NIST MULTICAL software [4]. The calibration planes are taken on silicon CPW ports. The results presented include the CPW to microstrip transition. Full wave computations obtained from Momentum agree very well with measurements. The transition has not been taken into account by these simulations, and the measured results are agree very well with the simulations. Note that S_{11} rose from -15 to -12 dB, and the external Q also increased due to the effect of the dielectric discontinuity. Insertion loss is 2.3 dB, with 3.5% relative bandwidth.

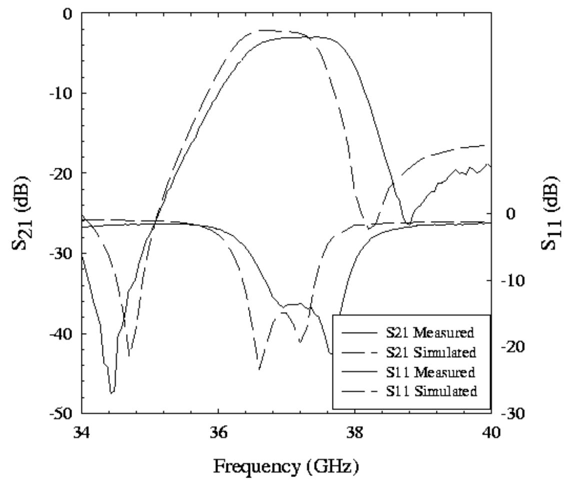

Fig. 3. Measured and simulated response of the 2-pole filter

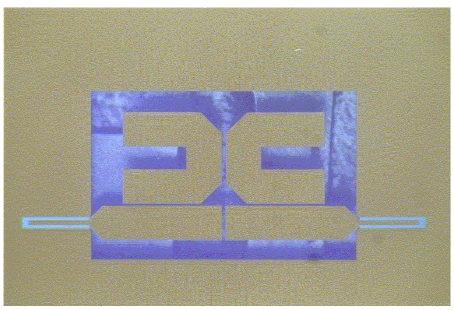

Fig. 4. Photograph of the actual 2-pole filter.

III. 60 GHz 4 POLE FILTER WITH TRANSMISSION ZEROS

A. Fabrication

The structure used for the 4-pole filter is the same as previously described, except that the strip width was reduced to 500 μm in order to be able to build the U shaped resonator. To maintain high Q_u, we have increased the cavity height from 200 to 250 μm. Moreover, metalization patterning was done using 1 μm evaporated gold and lift-off procedure in order to take advantage of the smooth metalization surface obtained. Two $\lambda/2$ resonators are added, with their lengths adjusted to keep the same resonant frequency. This layout, presented in fig. 5, allows a small structure in

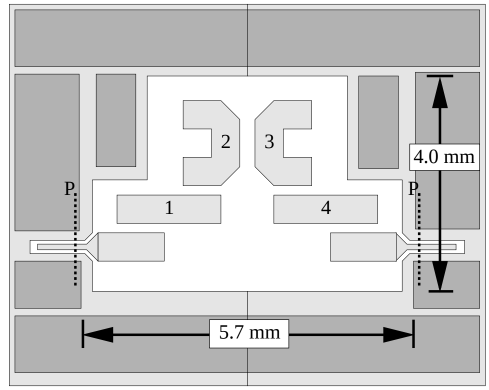

Fig. 5. 4 pole filter layout. The dark areas denote the vias grooves that shield the circuit.

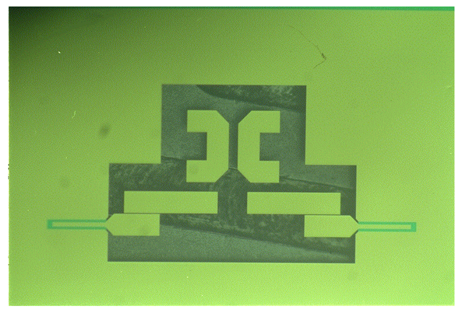

Fig. 6. Photograph of the actual 4-pole filter.

length, which prevents the excitation of parasitic TM cavity modes. Shielding is made using via grooves and one may notice the via grooves on each side of the U shaped resonators. The measured unloaded Q of a single resonator is 450 at 60 GHz.

B. Filter design and results

The filter design was done using HP Momentum according to a 4-pole 8% bandwidth Chebyschev prototype. Cross coupling between resonators 1 and 4 was computed using the formulas given in [5], and Chebyschev coupling coefficients can be modified to keep a correct response using formulas given in the same paper. In our case, cross coupling coefficient is low enough to keep a Chebyschev response in the pass band without modifying coupling coefficients. Let k_{nm} be the coupling coefficient between resonator n and m(see fig. 5). In our

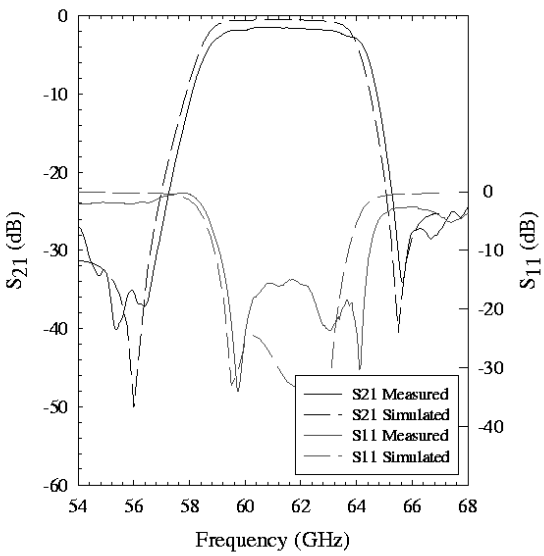

Fig. 7. Measured and simulated response of the 4-pole elliptic filter

design, k_{12} and k_{23} and k_{14} are not independent. In order to be able to determine k_{12}, we first determine the resonator spacing to get the proper k_{23}. Next the gap width is adjusted to get the correct k_{14}. Finally, gap between resonators 1 and 2 is adjusted to match the correct k_{12}.

The measured and simulated responses are presented on fig. 7. The agreement between simulated and measured responses is very good, and the frequency shift between simulated and measured response is less than 1% of the center frequency. The relative passband is 8% and the measured port to port (including transition) insertion loss is 1.5 dB with a return loss below -14 dB. The measured out of band rejection is better than 35 dB, and the whole filter is smaller than 4 x 6 mm^2.

IV. CONCLUSION

In this paper, we have presented measured results obtained after conception, realization and measurement of micromachined filters with transmission zeros. The measured insertion losses are very low compared to other planar filters at these frequencies. These circuits are very low cost considering that they are made with MMIC fabrication techniques. The micromachined filters are easy to integrate with active MMICs using the input / output CPW lines.

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