Integration Concepts on Silicon up to 140 GHz

Jürgen Hasch, Hans Irion, Andreas Müller
Robert Bosch GmbH, Central Research and Development, FV/FLO, Postfach 10 60 50, 70049 Stuttgart, Germany

Abstract — **Bandwith and antenna aperture size considerations drive radar applications into ever higher frequency ranges. Yet increasing the operating frequency has its advantages, allowing the integration of distributed circuit elements on silicon, utilizing highly sophisticated microelectronic structuring processes. Using high resistivity silicon and an advanced micromaching technology, millimeter wave circuit components with low loss and small tolerances are presented.**

I. Introduction

Market demands for future radar sensors drive millimeter circuit development to cut down size and greatly reduce costs while still improving sensor performance. One approach to meet these conflicting goals, is to increase the operating frequency of the radar sensor, thereby reducing the circuit size and required antenna aperture. Cost improvements can be realized by moving to integrated circuit processing on high resistivity silicon [1].

II. System Integration

Monolithically integrating all millimeter wave components, active and passive, on a single chip avoids difficult an expensive mounting and contacting.

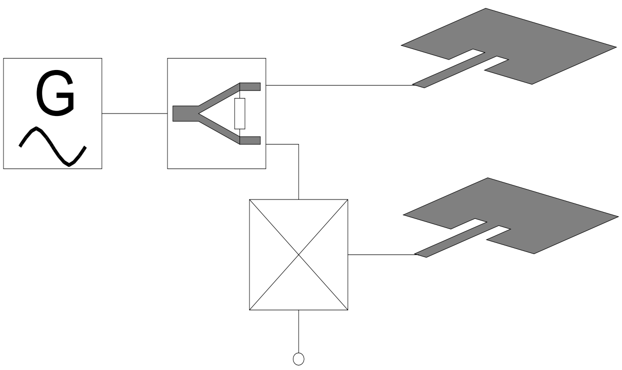

Figure 1 Basic Radar System

Figure 1 shows the high frequency part of a most basic radar system, consisting of a millimeter wave source, a mixer, transmit and receive antennas and some means of signal distribution. For a truly integrated solution, all of these components have to be monolithically integrated on the silicon substrate.

To study the feasibility of such an approach, millimeter wave circuits on high resitivity silicon have been simulated and manufactured.

At first, coplanar waveguide circuits in the lower GHz region have been examined [2]. As the next step, millimeter wave components on micromachined silicon were investigated.

The following sections present some of our results obtained for passive micromachined components.

II. Micromachining Technology

Applicating microstrip circuits in the millimeter wave region requires thin membranes in the region of one tenth of the wavelength to avoid substrate modes. Also a decent conductor width for transmission lines is desirable, to keep the losses small.

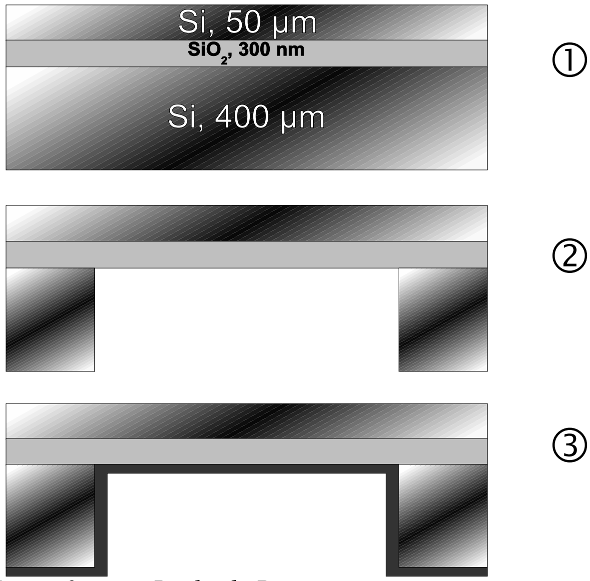

Figure 2 Backside Processing:
1 – Base SOI Wafer
2 – Trench Etching
3 – Backside Metallization

Therefore a membrane thickness of less than 100 µm for applications in the F band (90 to 140 Ghz) is required.

This can be achieved by thinning the wafers after processing, however this approach leads to very fragile wafers, where handling and mounting of the devices gets very difficult. The proposed alternative is thinning of only small specified area by micromachining, where the microstrip circuitry is to be placed.

Figure 2 shows the basic micromachining steps performed to create a 50 µm thin membrane by structuring the backside of a high resistivity SOI wafer. The SOI wafer consists of two layers of silicon with the top layer having a tightly specified thickness and a thin silicon oxide layer to separate the two silicon regions.

Backside micromaching is performed utilizing the so called „Bosch Process“, a high performance anisotropic plasma etch process [3]. The thin silicon oxide layer acts as an etch stop, so after the etching step only the the bottom silicon layer has been removed. After the echh step, the wafer backside is metallized by sputtering an aluminum layer with a thickness of a few microns.

Figure 3 shows the backside of such a micromachined wafer. Several rectangular areas which have been micromachined can be seen. On the top side of the wafer microstrip circuits have been placed at these areas.

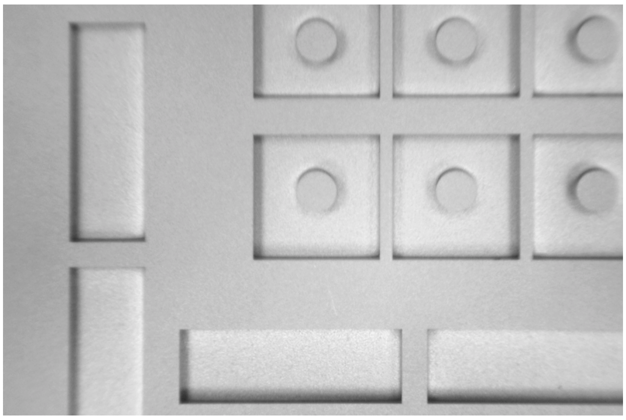

Figure 3 Backside view of the micromachined and metallized silicon wafer

III. Transmission Lines

As the most basic element, microstrip transmission lines with different lengths have been manufactured and their attenuation has been measured.

To be able to measure microstrip lines, coplanar probing heads need to be contacted. Therefore a coplanar waveguide to microstrip transition is needed.

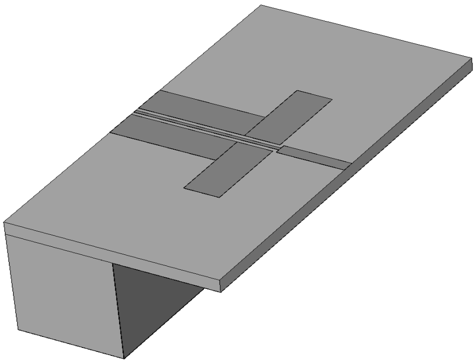

Figure 4 Coplanar waveguide to microstrip line transition

The simulated performance shows an insertion loss of less than 0,5 dB and a reflection coefficient better than 20 dB over the complete F band .

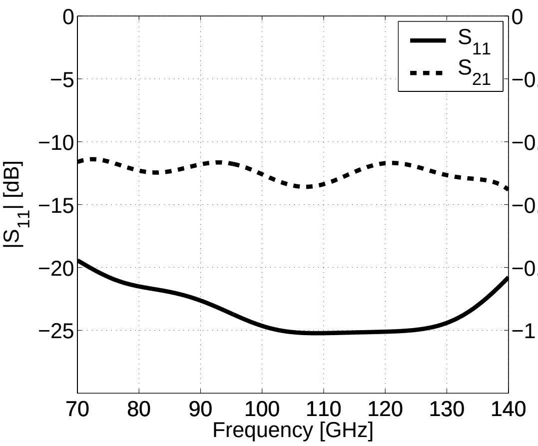

Figure 5 Input reflection and insertion loss of the coplanar waveguide to microstrip transition

The simulated and measured transmission line losses of a microstrip line are shown in Figure 5, with the theoretical values plotted as a dotted line.

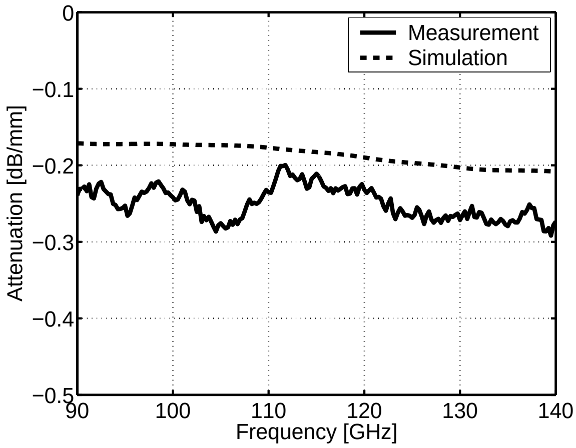

Figure 6: Microstrip Line Attenuation

IV. Filter Circuits

A side-coupled microstrip-bandpass-filter will be discussed. It is designed as a single-pole Chebycheff filter with a center frequency of 120 GHz, a bandwidth of 3 GHz with 0.1 dB ripple. The design was performed by using a standard design-technique as implemented in [4].

The computed circuit geometry is tranferred into a 2D EM simulator using Method of Moments (Agilent Momentum). Optimization was necessary as first results deviated from the aspired behaviour. This is due to the neglect of fringing effects in the design process. Figure 7 shows the single-pole filter manufactured on the SOI wafer.

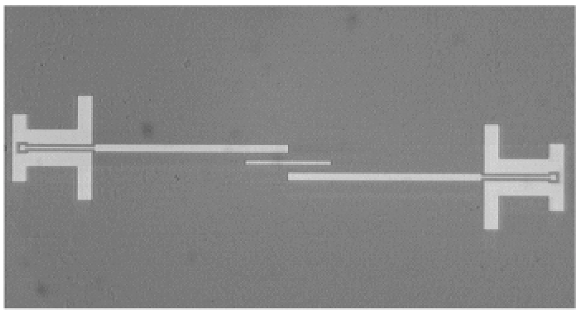

Figure 7 Microstrip Filter on SOI

The measured performance of the bandpass-filter between 90 GHz and 140 GHz agrees well with the simulated results. A slight shift concerning the resonance frequency can be observed. The measurement depicts an absolute value of the input reflection coefficient of -16 dB at a frequency of 122 GHz compared to the simulated result of -11 dB at 120 GHz.

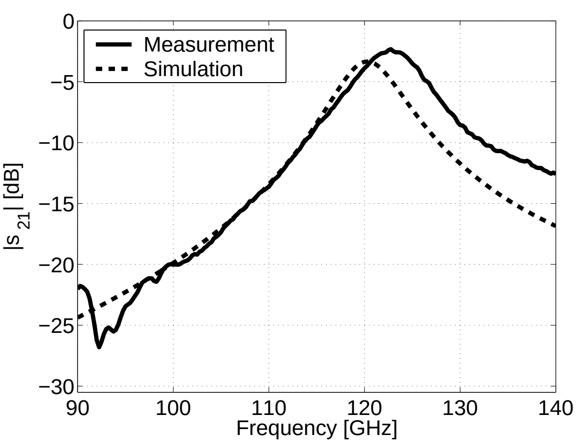

Figure 8 Absolute value of the transmission coefficient of the single-pole microstrip filter.

Figure 8 shows the absolute value of the transmission coefficient for both measurement (solid) and simulation (dashed). An insertion loss of 2.2 dB at a frequency of 122 GHz is measured. The simulation predicted a value of 3.2 dB at a frequency of 120 GHz. The discrepancy is due to a worst-case estimation for the substrate losses.

V. Micromachined Cavity Resonator

A waveguide type cavity resonator filled with silicon can be manufactured utilizing the presented backside micromachining process.

The resonant frequency of an ideal circular waveguide cavity resonator with height *h* and radius *a* for a TM type resonant mode is given in equation (1) [5]:

$$f_r = \frac{1}{2\pi\sqrt{\mu\epsilon}}\sqrt{\left(\frac{\chi_{m,n}}{a}\right)^2 + \left(\frac{p\pi}{h}\right)^2} \quad (1)$$

The fundamental mode of an waveguide resonator is the TM_{010} mode, if the ratio *h*/*a* is smaller than 2.03, which is true for the resonator presented in this paper.

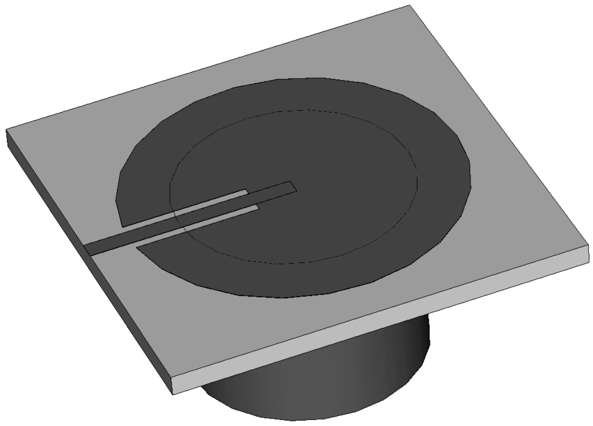

Figure 9 *Micromachined waveguide cavity type resonator with microstrip excitation*

Figure 9 shows the basic geometry of such a resonator realized using SOI substrate. On the back side is the metallized cavity, on top the metal layer forming a cover to reduce radiation from the cavity. The resonator is excited by a microstrip line leading into the cover.

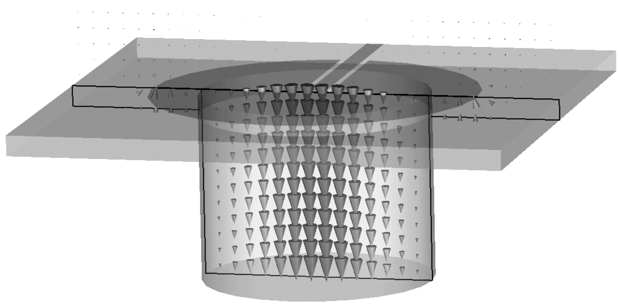

Figure 10 *Electrical field distribution for the TM_{010} mode in the resonator cavity*

The quality factor of the resonator is determined by conductor losses of the aluminum metallization, dielectric losses in the silicon and radiation losses caused by the open structure of the resonator cap.

Figure 11 shows the simulated reflection coefficient (S_{11}) of the cavity resonator with a height of 450 µm and a radius of 280 µm. The TM_{010} resonance frequency is at about 120.5 GHz

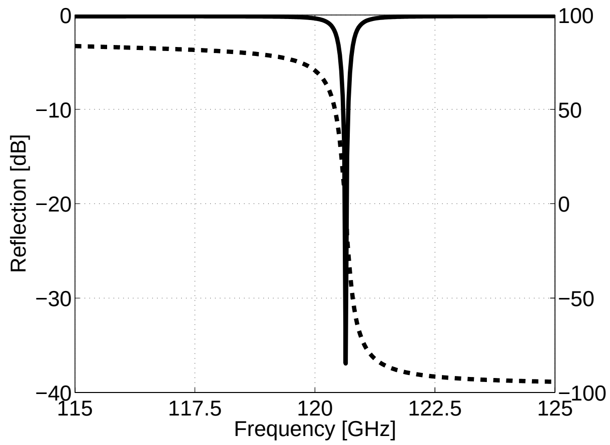

Figure 11 *Simulated reflection coefficient and phase of cavity resonator*

The value of the Q factor has been computed from the calculated S parameters using curve fitting of the Q circle as described in [6]. For the TM_{010} resonance, a value of $Q \approx 300$ has been found.

VI. Conclusion

A number of micromachined millimeter wave components have been investigated. Measurements up to 140 GHz prove the good performance of microstrip circuits and allow their application in future millimeter wave sensors.

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