77 GHz Transceiver Module Using A Low Dielectric Constant Multilayer Structure

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***Abstract* - This paper presents a novel 77 GHz transceiver design based on a simple and compact planar multilayer architecture. The module comprises a specific type of feed-through ratrace mixer and an aperture-coupled patch antenna both using low dielectric constant microstrip material. The broadband design makes the circuit suitable to be incorporated into a MMIC-based automotive radar front-end.**

I. INTRODUCTION

There is a growing demand for millimeter wave radar sensors for industrial and automotive applications. Several mm-wave products are under commercial development both for communication and sensor markets [1-3]. The 77 GHz front-end being developed at Siemens uses a MMIC-based solution [4] with all chips being assembled in flip-chip configuration.

The transceiver section (Fig. 1) of the front-end has a major influence on the overall system performance. Whereas the active part of the front-end (signal source, amplifiers) can be realized with excellent performance on high dielectric constant substrates like alumina, the antenna efficiency is rather poor when radiating structures are built on Al_2O_3 substrates. This was the motivation to investigate alternatives for the antenna/receiver part.

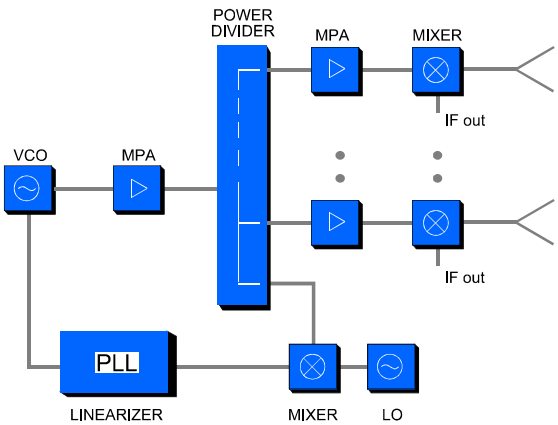

Fig. 1: Schematic of Siemens‘77 GHz ACC radar front-end.

Configurations have already been reported, that provide improvements of the antenna efficiency. Actually, using an „air substrate“ is proposed, whereas antenna patches are being mounted either in flip-chip configuration on alumina [4] or the patches are realized on synthesized low dielectric-constant silicon substrates using MEMS technology [5]. In this paper a third alternative is investigated, namely an aperture coupled patch antenna using a multiple layer structure built up from Teflon, polyimide and low dielectric foam.

Since the ACC front-end configuration uses five or even more narrow antenna beams to discriminate complex traffic situations, it is required to minimize the transceiver complexity per beam. This is realized with the monostatic antenna approach shown in Fig. 2., while the same antenna patch is used for transmit and receive.

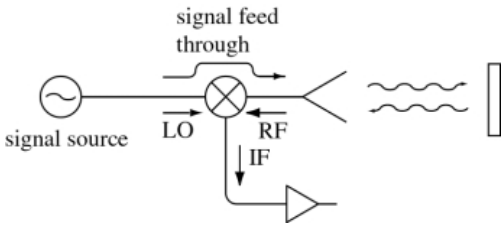

Fig. 2: Monostatic radar setup.

II. ANTENNA

The well-known structure of an aperture-coupled patch antenna is shown in Fig. 3a. Although the assembly is more complicated due to its multiple layer structure, it provides several advantages over a similar direct-coupled patch antenna [6,7]. With aperture coupling the usable bandwidth of the patch radiator is significantly increased. Although the bandwidth is not required for the application, it makes the design more robust with regard to etching tolerances of the microstrip structures and misalignment of the antenna layers. Furthermore, less cross-polarization is achieved due to the symmetrical structure and an elimination of spurious radiation from the feed line. It is also advantageous that the patch is located at the opposite side of the sub-

strate as is the feed line. Hence, it is possible to cover the active circuits of the radar module (VCO, mixer, amplifiers, MMICs) to protect it to environmental influences while not having any disturbance of the antenna radiation pattern.

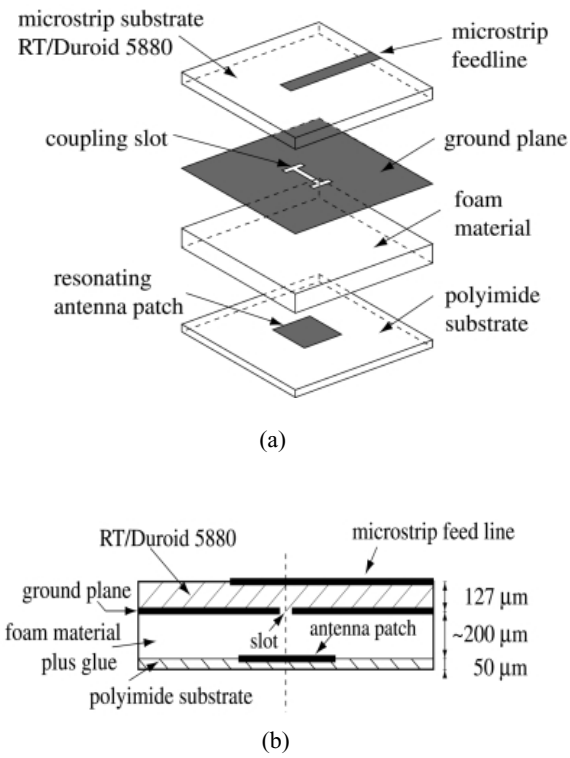

Fig. 3: Perspective view (a) and cross-section (b) of the aperture-coupled microstrip antenna.

A. Structure

The cross-section of the aperture-coupled antenna is illustrated in Fig. 3b. An open-ended microstrip line is the feed. The electromagnetic field couples through an H-shaped slot in the ground plane to the resonating patch, which is built on a polyimide substrate. The patch is physically separated from the slot in the ground plane by an intermediate layer. For this layer foam material is used, which provides sufficient mechanical stability as well as a dielectric constant close to air ($\varepsilon_r \approx 1$). Its thickness is about 200 µm. Both substrates are of low dielectric constant: the feedline is built on Rogers RT/Duroid 5880 with $\varepsilon_r = 2.2$, the polyimide substrate has an ε_r of 3.4. The thicknesses of the substrates are 127 µm and 50 µm, respectively.

The calculated dimensions of the antenna structure (Fig. 4) are: L = W = 1200 µm, D = H = 150 µm, $L_{slot} = 600$ µm, L_{stub} = 700 µm and W_0 = 400 µm resulting in a characteristic impedance of the feed line of 50 Ω.

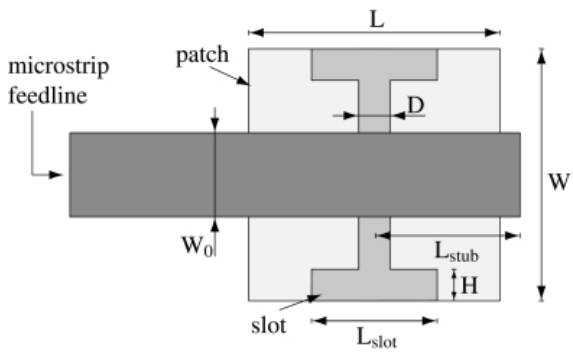

Fig. 4: Top view of the aperture-coupled antenna.

B. Measurements

For the simulations of the input impedance the IE3D field simulation tool from Zealand Software was used [8]. Simulation and measurement did not agree very well. We suppose, that the simulation tool uses a simple model of the electrical field in the ground plane at the coupling slot, which assumes magnetic currents. Experiments have shown that the distance between resonating patch and coupling slot needs to be smaller than predicted in order to get a good coupling. At the same time the resonant frequency is shifted down by about 3.5%. Fig. 5 shows the simulated and measured input impedance of two slightly different antennas: the thickness of the intermediate layer of the simulated antenna was 300 µm instead of 200 µm. The measured –10 dB bandwidth of the aperture-coupled 77 GHz patch is 6.4 GHz (8.3%) (simulation: 7.6 GHz / 9.9%), which is about three times as much as the bandwidth of a comparable direct-coupled patch antenna. A radiation efficiency of about 68% percent was measured. The applied method for the radiation efficiency measurement has been described in [5,9].

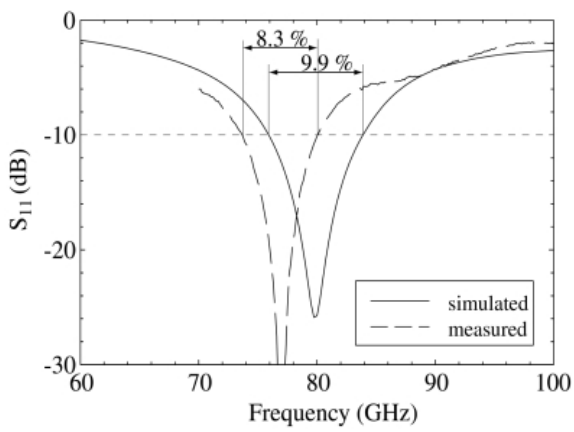

Fig. 5: Simulated and measured input impedance of the aperture-coupled microstrip antenna.

III. MIXER

The developed single-balanced mixer uses a rat-race coupler with two silicon Schottky diodes (Fig. 6). Port 1 is the LO port, port 2 the RF port, the IF is taken out at port 5. A lowpass filter prevents LO and RF leakage. The dc parameters of the diodes are: $R_S = 7.8\ \Omega$, n = 1.45, $\Phi_b = 0.7$ V, $\gamma = 0.5$ and $I_0 = 9*10^{-10}$ A. These values have been obtained by curve fitting from measurement of the I-V dc curve. The junction capacitance is $C_{j0} = 30$ fF according to the manufacturer. Both the balanced structure and the silicon diode (instead of a GaAs diode) were used to achieve a good low-frequency noise performance of the receiver.

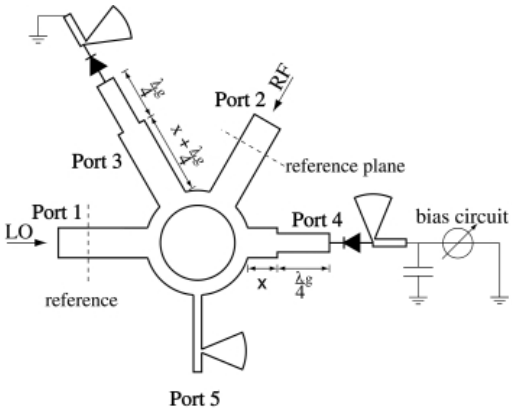

Fig. 6: Layout of the modified ratrace mixer.

In contrast to a conventional ratrace mixer, in this special type of mixer, a part of the local oscillator signal is fed through the mixer to the antenna. This is achieved by using a slight mismatch at the Schottky diodes as well as by adding a λ/4 section in one of the two diode branches. The extra λ/4 section transforms the ratrace coupler into a 0/90° coupler.

A. LO port –to –RF port transmission

The reflection at the diodes is given by the reflection coefficient:

$$|G| = \left|\frac{Z_2 - Z_L}{Z_2 + Z_L}\right|$$

A perfect matching resulting in zero reflection would occur, if the circuit is matched to the conjugate impedance of the diode. The impedance of the diode was simulated to be $Z_L = 41 - j27\ \Omega$ using the program DIODEMX [10], not taking into account the unknown diode inductance L_S. This simulation was done for a frequency of 76.5 GHz, for an LO power of 0 dBm per diode and a dc voltage bias of 0.35 Volt per diode. With no matching network, the calculated reflection coefficient is 0.30 or –5.2 dB. However, measurements showed a transmission of –7 dB, which can be explained by the unknown series inductance. To increase the reflection, an additional mismatching network in form of a $\lambda_g/4$ transformation line was added to both diode branches in order to reach the targeted transmission of –3 dB. The width is 220 µm resulting in a characteristic impedance of 70.7 Ω and a matching to $Z_L = 100\ \Omega$. The resulting reflection coefficient is 0.45 or –3.5 dB.

In a standard ratrace design the reflections would return to the port of the incoming signal. In order to achieve transmission to the isolated port a piece of microstrip line with length of $\lambda_g/4$ is added to port 3. The S-matrix of the modified ratrace is:

$$S = \frac{1}{\sqrt{2}}\begin{bmatrix} 0 & 0 & j & -1 \\ 0 & 0 & j & 1 \\ j & j & 0 & 0 \\ -1 & 1 & 0 & 0 \end{bmatrix}$$

Now an incoming signal a_1 at port 1 is scattered in the following way:

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \frac{1}{\sqrt{2}}\begin{bmatrix} 0 & 0 & j & -1 \\ 0 & 0 & j & 1 \\ j & j & 0 & 0 \\ -1 & 1 & 0 & 0 \end{bmatrix}\begin{bmatrix} a_1 \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ j\frac{1}{\sqrt{2}}a_1 \\ -\frac{1}{\sqrt{2}}a_1 \end{bmatrix}$$

Assuming that port 3 and 4 are open-ended microstrip lines, the reflection coefficient is Γ = 1. The scattering of these reflections is calculated by a second S-matrix multiplication:

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \frac{1}{\sqrt{2}}\begin{bmatrix} 0 & 0 & j & -1 \\ 0 & 0 & j & 1 \\ j & j & 0 & 0 \\ -1 & 1 & 0 & 0 \end{bmatrix}\begin{bmatrix} 0 \\ 0 \\ j\frac{1}{\sqrt{2}}a_1 \\ -\frac{1}{\sqrt{2}}a_1 \end{bmatrix} = \begin{bmatrix} 0 \\ -a_1 \\ 0 \\ 0 \end{bmatrix}$$

The resulting feed-through effect is illustrated in Fig. 7, which shows the measured S-parameters of the modified ratrace structure. This measurement was done with no diodes on the circuit.

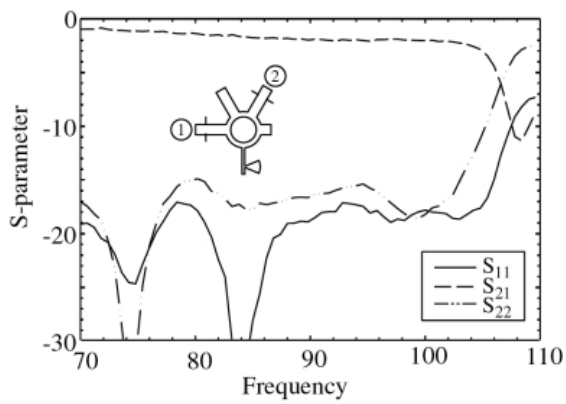

Fig. 7: Measured S-parameters of open-ended (no diodes on the circuit) ratrace coupler.

B. Measurements

Since the developed transceiver section is intended for use in a MMIC-based radar front-end and since available W-band MMICs are not capable to provide high power levels, the mixer had to be optimized for a LO input power of a few mW. For the measurements of conversion loss shown in Fig. 8 and LO port – to – RF port transmission shown in Fig. 9, a typical LO power of +6 dBm was applied. Diode biasing significantly enhances the mixer performance, while without dc bias a very poor conversion occurs since the LO does not pump the diodes sufficiently. Under appropriate bias conditions (about 1.6 mA per diode), the mixer had a conversion loss between 12 and 14 dB from 70 to 77 GHz, while the LO port - to - RF port transmission was around the targeted –3 dB.

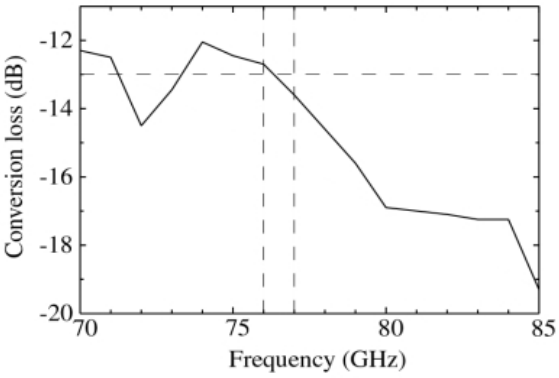

Fig. 8: Conversion loss of the ratrace mixer with P_{LO} = +6 dBm, IF = 220 MHz and I_{bias} = 1.6 mA per diode.

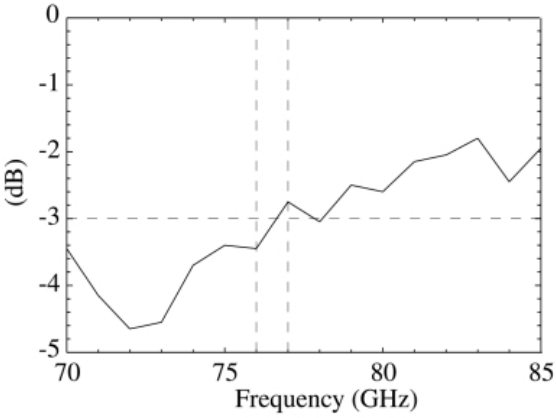

Fig. 9: LO port – to – RF port transmission of the ratrace mixer (P_{LO} = +6 dBm), bias.

Taking into account the 3 dB transmission/feed–through loss, the effective conversion loss compared to a standard ratrace mixer would be around 10 dB, which is acceptable.

The minimum conversion loss for the 3 measurements (P_{LO} = 6 dBm, 3 dBm, 0 dBm) is achieved for different bias currents (1.6 mA, 1.4 mA, 1.0 mA, respectively). Fig. 10 shows the dependence of the conversion loss versus bias current for LO power levels between 0 and + 6dBm.

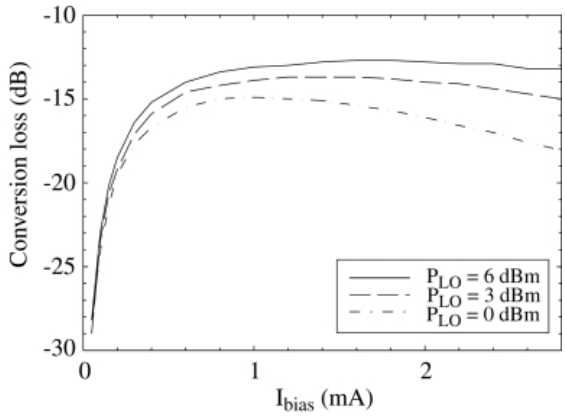

Fig. 10: Conversion loss of the ratrace mixer for different LO power levels (IF = 220 MHz, RF ≈ 77GHz).

IV. CONCLUSIONS

An alternative 77 GHz transceiver approach has been implemented successfully. The achieved performance figures, namely a high design bandwidth (8.3%), a high antenna radiation efficiency (68%) and a reasonable mixer conversion loss (13dB) as well as the tolerance-insensitive design makes the developed circuits attractive for integration into a MMIC-based 77 GHz automotive radar front-end.

V. REFERENCES

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