low-resistivity substrates with $\rho \approx 50\ \Omega \cdot$ cm and higher may be used in the microwave and low millimeter-wave frequency range.

- High-resistivity (FZ) material is more expensive than the low-ρ counterpart and it requires modifications of the standard Si process steps. On the other hand, one yields a substrate material with low-loss high-frequency features comparable to GaAs. Our results demonstrate that high-resistivity silicon is a substrate suitable for microwave and millimeter-wave applications up to W-band. A coplanar element library was developed based on field-oriented simulation. After implementation in commercial CAD software, the library was successfully employed with MMIC design. In conjunction with SiGe HBT's, a technology exists which meets the requirements for low-cost microwave and millimeter-wave mass products.

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Investigation of Microstrip and Coplanar Transmission Lines on Lossy Silicon Substrates Without Backside Metallization

Wolfgang Dürr, Uwe Erben, Andreas Schüppen, Harry Dietrich, and Hermann Schumacher

***Abstract*—Silicon circuits, now penetrating well into the microwave frequency range, use lossy silicon substrates. Consequently, the microwave performance of transmission lines on this substrate becomes increasingly important and has been investigated here up to 20 GHz. It is shown that transmission lines on 20-$\Omega \cdot$ cm substrates have no need for backside metallization and backside via holes. Two models for different line types are derived from measurements and verified against them. A coplanar waveguide (CPW) with an overall width of less than 30 μm was fabricated with an attenuation of 0.5 dB/mm at 20 GHz, which is acceptable for monolithic microwave integrated circuit (MMIC) design.**

***Index Terms*— Coplanar waveguides, microstrip, MMIC's, silicon, transmission lines.**

I. Introduction

Traditionally, silicon technology uses a medium conducting substrate to balance between isolation characteristics and substrate depletion capacitance [1]. The substrate must be connected to the lowest potential, mostly ground, to put the substrate diodes of the active components in blocking direction. The conductivity of the Si substrate and the need of an oxide on top of the substrate are the major differences to conventional microwave substrates.

Traditional monolithic microwave integrated circuit (MMIC) design on GaAs uses two different line types: microstrip lines (MSL's) with a need for backside metallization and via holes, and coplanar waveguides (CPW's) with problems to suppress odd modes and substrate waves. MSL's on standard Si substrates are very lossy, and the backside metallization is shielded by the substrate for lower frequencies. CPW's are strongly influenced by the substrate, even if the oxide is very thick, because of the need to connect the substrate to ground when active components shall be isolated.

Most of the past work on Si substrate is based on a 1971 paper by Hasegawa *et al.* [2]. The authors distinguish between three types

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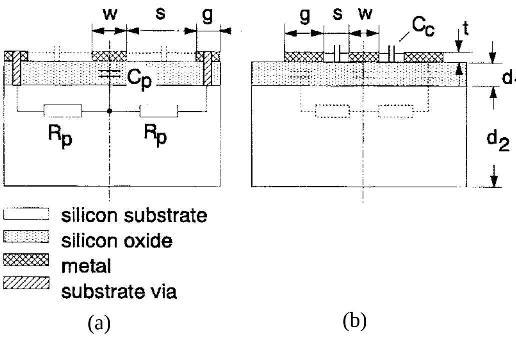

Fig. 1. Cross section of (a) MSL and (b) CPW.

TABLE I
LAYER SEQUENCE OF THE TEST STRUCTURES

Layer	Thickness / μm
metal (AL)	1.5
CVD grown silicon oxide	1.7
thermal field oxide	1.0
$300\Omega/\Box$ p^+ channel stop layer	≈ 1
20 Ωcm silicon substrate	690

of wave propagation: a skin-effect mode, a quasi-TEM mode, and a slow-wave mode, which shows higher effective relative permitivities than that of the Si substrate. The slowing factor is defined as $\sqrt{\varepsilon_{r,\text{eff}}}$.

Up to now, all papers investigating lossy silicon substrates look either at MSL's with backside metallization [3], or at CPW's without connecting the substrate to ground and without consideration of a channel stop layer [4], [5]. This paper reports on two different line types that connect the substrate to ground from the surface with small substrate vias, which corresponds to common procedures in Si integrated circuits (IC's). The first line type is a CPW with an imperfect conducting ground plane. The second type is a transmission line where the lateral distance from the signal line to the ground line is large, but undefined. This is further called the MSL because the conducting substrate that is connected to the ground line represents an imperfect ground plane, which needs to be considered if the width of the line is larger than the oxide thickness. A realized thin MSL shows a measured attenuation of 2 dB/mm at 20 GHz, which is very close to the predictions of conductor-backed MSL [3].

After the description of the test structures, the relevant coupling mechanisms will be discussed, which lead to two different line models. The line parameters are deduced from s-parameter measurements. A comparison between the extracted parameters and the measurement up to a frequency of 20 GHz is presented and discussed.

II. Description of the Transmission Lines

The preparation of the microwave-line structures uses a production-ready SiGe HBT technology [6]. The layer topology which is relevant for the transmission lines are summarized in Table I. The cross section of an MSL and a CPW is drawn in Fig. 1(a) and (b), respectively. The length of the lines is chosen to be 5 mm, additional to ground–signal–ground (GSG) pads with 100-μm pitch for on-wafer measurement. For all line types, ground lines are contacted with small vias to the substrate, for the CPW, at the beginning and end of the line, for the MSL, periodically the entire ground line. The lateral distance between all microstrip signal lines and ground lines is choosen greater than 100 μm. The main dimensions of MSL and CPW lines under investigation are summarized in Table II.

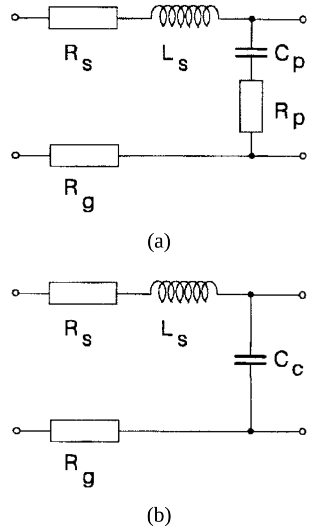

Fig. 2. Models for a small portion of an (a) MSL and (b) CPW.

TABLE II
SUMMARY OF ALL INVESTIGATED
TRANSMISSION-LINE STRUCTURES: MSL AND CPW

Line type	w / μm	s / μm	g / μm	$Z_0'(\Omega)$ @ 10 GHz	$\sqrt{\epsilon_{r,eff}}$ @ 10 GHz	$\alpha(dB/mm)$ @ 20 GHz
MSL	18.5	≥ 100	18.5	53	4	3.8
	9.1		18.5	75	3.5	2.7
	4.2		18.5	100	3	2.0
CPW	18.5	15.9	48.2	42	2.9	1.3
	4.2	4.2	8.4	68	1.8	0.5

III. Model Development

The main difference between the two line types is their field distribution. For the MSL, the main electric field propagates in the oxide, and the magnetic field propagates in both the oxide and the substrate under the conductor, which could provide a slow-wave mode. The maximum slowing factor, if the whole wave would be guided in the Si substrate, would be $\sqrt{\varepsilon_{r,Si}} = \sqrt{11.9} = 3.45$. All of the realized MSL's show slow-wave effects, at least at low frequencies (see Fig. 3). The slow-wave coupling mechanisms can be modeled by a parallel *RC* network [2] [shown in Fig. 1(a)]. CPW's with small gaps guide most of their electric and magnetic fields in the oxide and the air between the conductors and, hence, lower losses and less slow-wave effects are expected. The coupling between the lines can be modeled by a simple capacitance C_c [see Fig. 1(a)]. The dashed elements in the pictures can be neglected. Also, the coupling between an additional backside metallization can be neglected for both line types if the line dimensions are not too large due to the conducting substrate. Both line models should be completed with a series resistance R_s and inductance L_s. The resulting line models for a small portion of an MSL and CPW, which are extended with a series resistance R_g of the lossy ground lines, are shown in Fig. 2.

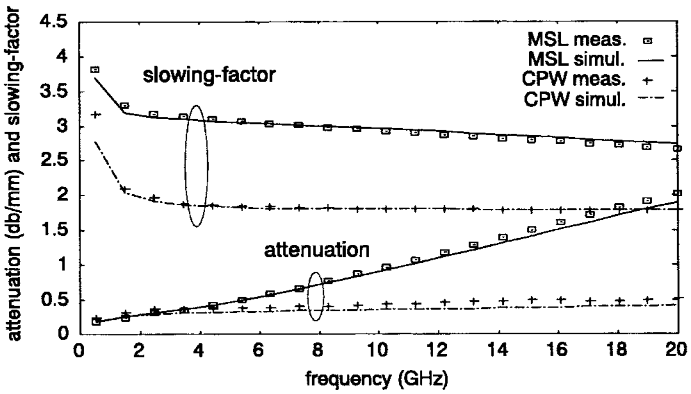

Fig. 3. Comparison of simulated and measured $\sqrt{\varepsilon_{r,\mathrm{eff}}}$ for a 4.2-μm-wide MSL and CPW.

IV. Determination of the Model Parameters

The values of the model parameters are deduced from s-parameter measurements. After an line–reflect–match (LRM) calibration with a CASCADE calibration substrate, the influence of the bonding pads were subtracted using y-parameters. Additionaly, the discontinuity at the beginning and end of the CPW was take into account with a parallel capacitance of 20 fF for the small CPW and 10 fF for the wide CPW, which describe the additional coupling. From these extracted s-parameters, the propagation coefficient γ and the characteristic impedance Z_0 can be approximated by the concept of even- and odd-mode excitation [7]. With $z = Z_0/50\ \Omega$, $S_e = S_{11} + S_{21}$, and $S_o = S_{11} - S_{21}$ it follows that

$$z^2 = \left(\frac{1+S_e}{1-S_e}\right)\left(\frac{1+S_o}{1-S_o}\right) \tag{1}$$

$$\tanh^2 \gamma \frac{l}{2} = \frac{\left(\dfrac{1+S_o}{1-S_o}\right)}{\left(\dfrac{1+S_e}{1-S_e}\right)}. \tag{2}$$

This approximation usually is thought to be only valid for low frequencies when the transmission-line length is less than a half-wavelength [8], but if the complex functions are generalized over the complete complex plane and attention is paid to the continuity condition of the phase of all used complex functions, the approximation remains valid up to higher frequencies. For example, the inverse of the tanh could be extended to [9]

$$\operatorname{arctanh}(\underline{w}) = \frac{1}{2}\ \ln\frac{1+\underline{w}}{1-\underline{w}} \tag{3}$$

with

$$\ln\frac{1+\underline{w}}{1-\underline{w}} = \ln(x+jy) = \frac{1}{2}\ \ln(x^2+y^2) + j\left(\arctan\frac{y}{x} + 2k\pi\right), \quad \text{with} \quad k \in \mathrm{N}. \tag{4}$$

The values of L_s, R_s, C_p, and R_p can be derived from the complex characteristic impedance Z_0 and the complex propagation coefficient γ by the following [5]:

$$Z_0 = Z_0' + jZ_0'' = \sqrt{\frac{Z_s}{Y_p}} \tag{5}$$

$$\gamma = \alpha + j\beta = \sqrt{Z_s Y_p} \tag{6}$$

with

$$Z_s = R_s + j\omega L_s \tag{7}$$

and

$$Y_p = \frac{1}{R_p + \dfrac{1}{j\omega C_p}}. \tag{8}$$

The slowing factor can be calculated from β as follows:

$$\sqrt{\varepsilon_{r,\mathrm{eff}}} = \frac{\beta \cdot c_0}{\omega}. \tag{9}$$

The extracted value of the series resistance corresponds to the sum of the resistance of the signal line and the two ground lines, and increases with higher frequencies due to the skin effect. The values of the line resistance R_s and R_g, including the skin effect, can be easily deduced from the conductivity ϱ_{Al} of the Al metallization and the line dimensions

$$R_s = \frac{\varrho_{\mathrm{Al}} \cdot l}{w \cdot t}\sqrt{1 + \frac{f}{f_0}} \tag{10}$$

with f_0 as the skin-effect corner frequency, which can be determined from the extracted series resistance. R_p of the MSL that corresponds to the substrate resistance, which can be calculated from the measurable sheet resistance of the substrate, including the channel stopper, is here, approximately $\varrho_{\mathrm{Sub}}/d \approx 300\ \Omega/\square$ by

$$R_p = \frac{\varrho_{\mathrm{Sub}}}{d} \cdot \frac{s}{l}. \tag{11}$$

The extracted values of L_s and C_c remain nearly constant over the whole frequency range and the value of R_p for the small CPW is, in fact, nearly zero. For the MSL, the extracted values of C_p and R_p decrease with higher frequencies, which is comparable to the results of [8]. This was modeled by an additional factor

$$k = \frac{1}{\left|1 + j\dfrac{f}{f_1}\right|} \tag{12}$$

where f_1 is a second corner frequency. The modified values of C_p and R_p are then $C_p^* = C_p \cdot k$ and $R_p^* = R_p \cdot k$. For exact modeling of the lines with a length of 5 mm up to a frequency of 20 GHz, a chain of 16 equivalent circuits, depicted in Fig. 2, is used.

V. Measurement Results and Discussion

Table II shows the extracted line parameters of all transmission lines. A comparison between the measured and simulated characteristics of an MSL and a CPW with $w = 4.2\ \mu$m is depicted in Fig. 3. For the MSL with decreasing linewidth, the characteristic wave impedance increases, while the attenuation and the slowing factor decreases. This agrees with the above model because a small MSL minimizes the coupling with the substrate and guides more fields in the air and oxide and, therefore, lowers the slow-wave effect. Two different CPW's are investigated. The effective relative permittivity for the smaller line drops to a value of 3.2, which is very close to the optimum of $\varepsilon_{r,\mathrm{eff}} \approx (\varepsilon_{r,\mathrm{SiO_2}} + 1)/2 \approx 2.5$. This indicates that the wave propagation is mainly concentrated in the small gap between signal line and ground, like a conventional CPW. This leads to a reduced attenuation of 0.5 dB/mm at a frequency of 20 GHz, with a characteristic impedance of 65 Ω, which is acceptable for interconnections in MMIC design. This line also has an acceptable overall dimension of less than 30 μm. The wider line shows a remarkable slow-wave effect because of their wide signal line, which couples with the substrate. This line can be seen as a mixture between a CPW and an MSL. Therefore, it should be possible to guide two MSL's in parallel with low cross coupling using only a distance of 20–30 μm. This is, in fact, a big advantage of MMIC design on lossy Si substrates and partially cancels the higher losses of the MSL.

VI. Conclusion

We have investigated the performance of microstrip and coplanar lines up to a frequency of 20 GHz. Model parameters are deduced from measurements and compared with their physical meaning. For both microwave-line types, the attenuation decreases with decreasing linewidth. The smallest attenuation, 0.5 dB/mm, will be achieved for a coplanar waveguide with only 4.2-μm center conductor width. Less than 30 μm in overall width will be necessary for such a CPW, which is acceptable for the application in silicon-based MMIC's. Both line types do not need backside metallization or backside via holes. Due to the small line dimensions, both line types are very attractive for low-cost MMIC designs up to 20 GHz because of minimizing the real estate on wafer.

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Circularly Polarized Millimeter-Wave Rectenna on Silicon Substrate

Ralph H. Rasshofer, Markus O. Thieme, and Erwin M. Biebl

***Abstract*— A circularly polarized (CP) silicon-integrated W-band rectenna (rectifying antenna) for use in six-port polarimetric radar systems was designed, numerically optimized, and fabricated. A rigorous numerical optimization was performed by using the supergrid method (SGM). The rectenna applies a novel low-loss purely planar dual-patch antenna (DPA) layout, which allows the receiver to be manufactured using monolithic integration. The measurement results for the receiver demonstrate an excellent cross-polarization discrimination (XPD) $>$14 dB @ 76 GHz over a wide range of the scan angle (12 dB @ $\pm 20^\circ$).**

***Index Terms*—Active antenna, automotive radar, circular polarization, circularly polarized antenna.**

I. Introduction

Prototypes of millimeter-wave polarimetric radar systems have successfully demonstrated their capability in automotive applications, e.g., measurement of the lateral and longitudinal velocity [1], evaluation of the road condition [2], and collision avoidance [3]. However, the availability of miniaturized low-cost components is a prerequisite for serving the high-volume automotive market. To reduce manufacturing costs, monolithically integrated components can be conveniently employed.

Though a variety of designs for linearly polarized (LP) integrated direct-detection receivers were developed [4], existing designs for circularly polarized (CP) antennas do not comply with monolithic integration since they do not employ a purely planar circuit layout [5], [6]. We, therefore, propose a novel purely planar dual-patch rectenna, which is fully compatible with monolithic integration. The use of two almost-quadratic patches eliminates the need of phase shifters, resulting in a smaller chip size and higher sensitivity of the receiver. We applied a generic two-step approach, comprising a simple transmission-line equivalent-circuit model of the antenna and full-wave electromagnetic (EM) simulation and optimization tools to dimension the layout and optimize the receiver's cross-polarization discrimination (XPD) and sensitivity. We fabricated the rectenna on a high-resistivity silicon substrate with backside metallization using a flip-chip technique to integrate the detector diode. Experiments demonstrate the efficiency of this receiver conception. Although primarily intended for use in six-port polarimetric radar systems [7], [8], the rectenna can find many applications in short-haul communication systems as well. In this paper, we present design and fabrication as well as measurement results of this novel silicon-based receiver chip.

II. Physical Condition for Circular Polarization

CP operation requires phase quadrature and equal magnitude of the antenna's orthogonal current components. This can be accomplished by feeding the radiating elements of the antenna with a phase shifter, e.g., a 90° hybrid. However, for millimeter-wave integrated antennas, this approach has several drawbacks. The use of a phase shifter increases the chip size and causes additional ohmic losses.

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