

A Multimodal Dielectric Waveguide-Based Monopulse Radar at 160 GHz

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Abstract—For highly integrated imaging systems above 100 GHz, the complexity and chip area increase significantly with an increasing number of channels. In addition, bulky dielectric lenses prevent applications in spatially restricted surroundings. The presented concept of an imaging monopulse radar with a mechanically flexible front end reduces the required chip area and allows the antenna to be placed in any desired position apart from the sensitive electronics. The radar system is based on a two-channel 160-GHz microwave monolithic integrated circuit (MMIC) feeding a flexible dielectric waveguide. Depending on the angle of the incidence signal, a sum mode (HE₁₁ mode) and a difference mode (HE₂₁ mode) are excited in the dielectric waveguide. MMIC and waveguide are connected by a self-aligning transition reducing the requirements for packaging accuracy. The required chip area of the transition is only $0.022\lambda^2$ with a spacing between the on-chip antennas of $\lambda/4$. The measured ambiguity-free region between -18° and 15° is defined by the modified elliptical lens antenna focusing in the *E*-plane only. A mechanical bending of the flexible waveguide is possible down to a radius of at least 2 cm without affecting the angle estimation capability.

Index Terms—Dielectric waveguide, flexible radar sensor, microwave monolithic integrated circuit (MMIC), millimeter wave, monopulse, multimodal, transition.

I. INTRODUCTION

THE robustness of radar systems allows their usage for distance and velocity measurements in harsh environments, such as level measurements in tanks [1] or in the automotive sector [2], [3]. Especially, frequencies above 100 GHz are attractive for high-resolution applications due to the large available absolute bandwidth. In addition, complete systems can be integrated on monolithic microwave integrated circuits (MMIC), allowing sensors to be mass-produced at low cost. These systems become even more compact by integrating the antenna on-chip, thus avoiding lossy transitions to printed circuit boards (PCBs). This results in many new applications, accompanied by the additional demand for angular information.

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The 3-D scanning of the scenario or synthetic aperture radar (SAR) imaging with single-channel MMICs results in a high-resolution image [4], [5]. These techniques require a long measurement interval and a precise mechanical scanning. The measurement time can be reduced by the combination of several single-channel MMICs to one imaging system [6]. However, a high-precision positioning of the MMICs and the antennas must be ensured to avoid a high sidelobe level. In addition, a complex feeding network is required to perform a coherent angle estimation. A second approach is to increase the number of channels on one MMIC [7]–[10], allowing multiple-input–multiple-output (MIMO) systems, phased arrays, and monopulse systems to be realized. The disadvantage of such a system is the large MMIC area needed due to the large number of antennas with typically used spacings of $\lambda/2$. Both system approaches usually employ additional lenses to focus the beam and to achieve an increased detection range. However, the sensor range remains limited since not all antennas can be placed in the focal point, causing different tilts of the beams [11]. Furthermore, the sensor platform becomes more bulky, and flexible usage is no longer possible.

Especially, in the considered frequency range above 100 GHz, a mechanically flexible front end allows new applications [12]–[14] and to measure directly at the desired position in spatially restricted surroundings. A flexible dielectric waveguide can be used, which can easily bypass longer distances of up to 1 m due to its low attenuation and mechanical bendability [15], [16]. Depending on the application requirements, different radiation patterns can be emitted with different antenna types [17]–[19]. However, an angle estimation based on the evaluation of the phase difference of the signals between several channels is not possible, as only one waveguide front end, and thus only one channel, is available.

The mentioned radar systems are based on single-mode dielectric waveguides. Since the radiation pattern depends on the mode fed into the waveguide, a sum and a difference beam can be radiated when employing higher order modes. Comparable to a multimodal excited metallic waveguide horn [20], an imaging system based on the monopulse principle [21] can be realized with the dielectric waveguide. The angle information is contained in the amplitude ratio of the difference pattern to the sum pattern. This results in an imaging system with flexible front end, offering the advantage to separate the sensitive electronic and the

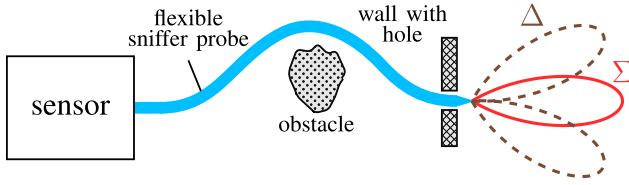


Fig. 1. Schematic of a flexible sniffer probe measuring around an obstacle with sum beam (Σ) and difference beam (Δ).

more robust antenna spatially, for example, in dirty, humid, or flammable surroundings. Furthermore, such sniffer probes enable to measure around obstacles, through holes or in areas, which are difficult to access, as shown in Fig. 1.

A further advantage is the minimization of the area consumption on the MMIC. The excitation structure area of the transition can be reduced to some extent without degrading the performance of the angle estimation. Instead, a distance reduction between the monopulse antennas on chip, typically $\lambda/2$ [22], directly affects beam pattern and, thus, the ambiguity-free region and the steepness of the amplitude ratio.

This article presents a novel imaging radar system at 160 GHz incorporating a mechanically flexible antenna front end with the capability to estimate the direction of arrival using the monopulse principle. The flexible front end is based on a multimodal dielectric waveguide that receives sum and difference pattern for the direction-of-arrival estimation. The connection between front end and MMIC is realized with a self-aligning and compact transition, providing equal transmission of both modes. At first, the overall system concept and the mode propagation in the monopulse radar are explained. In Section III, the individual components of the system, the MMIC, the transition to dielectric waveguide, the dielectric waveguide itself, and the antenna are described and characterized. The measurement setup of the monopulse radar is presented in Section IV. Measurement results and an investigation of the influence of the mechanical flexibility on the angle estimation are stated in Section V.

II. SYSTEM CONCEPT

The proposed system concept is shown in Fig. 2(a). The frequency-modulated continuous-wave (FMCW) radar signal is generated on a two-channel 160-GHz MMIC [23] and couples through a superstrate antenna and a self-aligning waveguide plug into the flexible dielectric waveguide. A Rohacell shell, fitting positively on the pin-shaped profile, is used to fix the flexible waveguide in the desired position. A dielectric lens antenna, focusing the field in the E -plane, is plugged on the dielectric waveguide. Lens and dielectric waveguide are made of the same material to avoid mismatch.

Instead of using lossy and area-consuming couplers on the MMIC, the presented monopulse system feeds both signals $v_{1,Tx}$ and $v_{2,Tx}$ [see Fig. 2(b)] into the dielectric waveguide in-phase and excite the sum mode Σ , which is radiated by the lens antenna. Depending on the angular position of the target, the reflected and received signal excites the sum mode Σ as well as the difference mode Δ in the waveguide, resulting in an angular-dependent amplitude ratio.

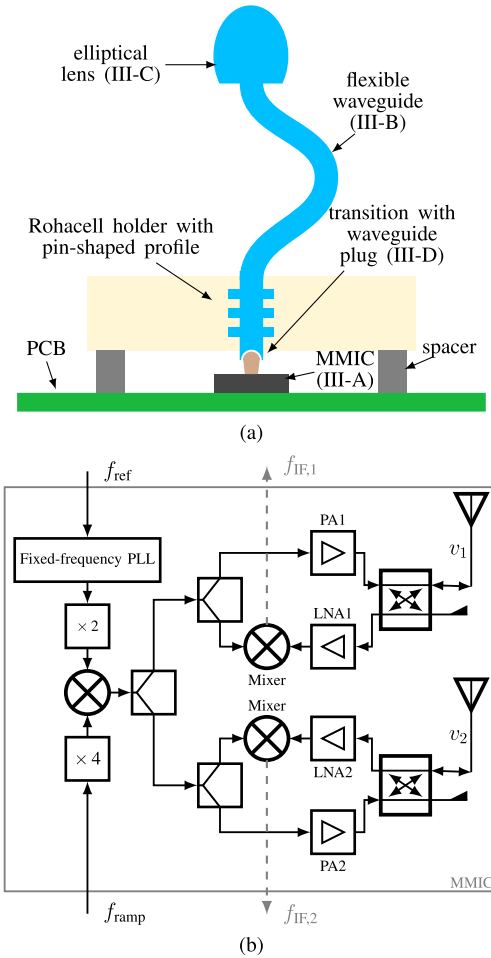


Fig. 2. (a) Cross section of the system concept in the E -plane. (b) Block diagram of the MMIC used for the monopulse radar.

These modes in turn excite the two signals $v_{1,Rx}$ and $v_{2,Rx}$ on the MMIC, with the relationship between the signals on the MMIC and the dielectric waveguide modes being [21]

$$\Sigma = \frac{v_1 + v_2}{\sqrt{2}} \quad (1)$$

$$\Delta = \frac{v_1 - v_2}{\sqrt{2}}. \quad (2)$$

The determination of the sum and difference modes is carried out only in the digital domain after downconverting and sampling the intermediate frequency (IF) signals in order to avoid additional on-chip components.

The ratio determines the quality of the angle estimation as well as the ambiguity-free region and is determined by the front end, in particular by the beam pattern of the antenna. The quality of the angle estimation can be adjusted via the steepness of the ratio curve and it improves with increasing slope. This is determined by the sum and difference beam patterns and their gain. As an example, Fig. 3 shows a sum beam, a difference beam, and the corresponding ratio. The ambiguity-free region is between -67° and 67° and its limits are defined by the first minima in the sum beam.

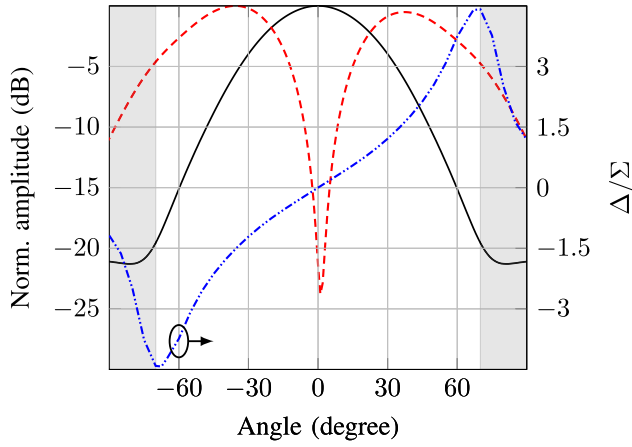


Fig. 3. Exemplary pattern of a monopulse antenna with sum beam Σ (—), difference beam Δ (---), and the corresponding ratio Δ/Σ (-·-·-).

III. SYSTEM COMPONENTS

In the following sections the individual components required for the proposed system are explained in detail.

A. 160-GHz Radar MMIC

The proposed radar system is based on a modified version of the two-channel silicon–germanium (SiGe) MMIC as presented in [23] [see Fig. 2(b)]. Due to its architecture with a frequency offset synthesizer, the MMIC is optimized for low phase noise [24]. The center frequency of the MMIC is at 152 GHz and enables bandwidths of more than 8 GHz. Thus, a high range resolution of 1.88 cm is achievable.

In order to keep the chip area as small as possible and consequently the costs low, several steps were taken. Since the antennas or the transition structures are typically the largest components on an MMIC, a monostatic chip architecture was used to minimize the number of antennas. Furthermore, the excitation structure area for the transition can be minimized since the quality of the angle estimation in this concept is almost independent of the distance of the feeding structures. The only requirement on the structure is to excite the corresponding modes in the waveguide with low losses. In contrast, in conventional antenna arrays, the antenna spacing directly affects the beam pattern and the ambiguity-free region.

These design criteria result in an MMIC area of only $2.24 \text{ mm} \times 1.50 \text{ mm}$.

B. Dielectric Waveguide

The dielectric waveguide does not only constitute the mechanically flexible part of the radar system but also guides sum and difference modes and enables to realize a monopulse radar. Therefore, its electrical characteristics are examined in more detail.

The dielectric waveguide is used for frequencies above 100 GHz in radar [12] as well as in communication applications [25] due to the low losses and the low production costs compared with metallic waveguides or PCB transmission lines. The fundamental mode in a rectangular dielectric waveguide

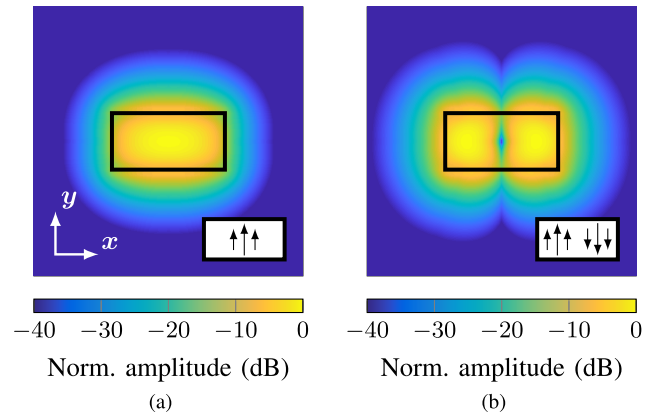


Fig. 4. E -field distribution of (a) HE_{11} mode and (b) HE_{21} mode with polarization in the y -direction. The black square represents the outline of the dielectric waveguide.

is the HE_{11} mode [26]. It has no lower cutoff frequency and guides field components inside and outside of the dielectric medium as shown in the E -field distribution in Fig. 4(a). The field is polarized in the y -direction and has its maximum in the center of the dielectric waveguide.

In comparison, the first higher order mode in the dielectric waveguide, the HE_{21} mode, is shown with its E -field distribution in Fig. 4(b). The field intensity has two maxima in the x -direction with opposite signs and a minimum in the waveguide center. If the mode is radiated, this minimum results in a notch in the radiation pattern at the boresight axis and two split beams. Therefore, the HE_{21} mode corresponds to the differential mode of the monopulse radar since a differential beam is obtained. Together with the fundamental HE_{11} mode radiating a sum beam, a dielectric waveguide-based monopulse front end can be realized.

Since the HE_{21} mode has a cutoff frequency, the field distribution and mode propagation of both modes have to be considered in the design process of the radar front end. The field distribution and cutoff frequency of the dielectric waveguide depend on the waveguide dimensions and the material permittivity. By increasing the dimensions or the permittivity, the field becomes more concentrated in the dielectric, and the propagation of higher order modes becomes possible. Due to the higher field density in the dielectric, the dielectric losses increase. Simultaneously, the losses due to parasitic radiation in bends are reduced. Therefore, a compromise between the propagation of the required modes and a minimum of losses is necessary for the design.

The dielectric waveguide in the presented system has a rectangular cross section of $1050 \mu\text{m} \times 2100 \mu\text{m}$ and is made of high-density polyethylene (HDPE; $\epsilon_r = 2.25$, $\tan \delta = 3.1 \times 10^{-4}$ at 160 GHz [27]). In order to determine the attenuation, several waveguides of different lengths were fed via a WR5 metallic waveguide mode converter [17], and the insertion losses were compared. Since standing waves in the mode converter caused ripples, the measured values were smoothed. The fundamental mode of the dielectric waveguide has a measured attenuation of 10 dB/m at 160 GHz in a straight waveguide, as shown in Fig. 5. The attenuation

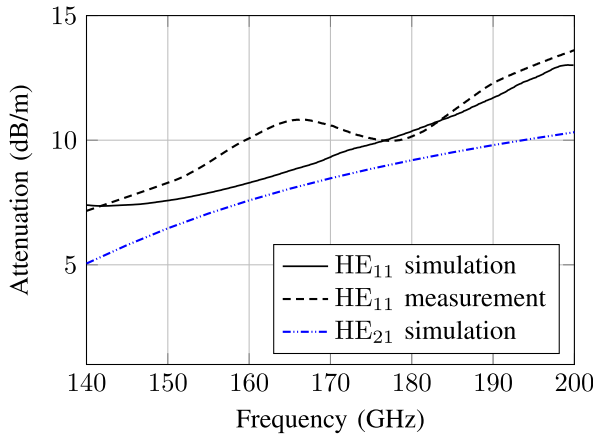


Fig. 5. Simulated and measured attenuation of a straight dielectric waveguide made of HDPE with a rectangular cross section of $1050 \mu\text{m} \times 2100 \mu\text{m}$.

increases at higher frequencies since the field distribution changes and the field is more concentrated in the dielectric waveguide. The measurements agree well with the full-wave simulations although a change of 10^{-4} in the loss tangent already corresponds to a change of approximately 1 dB/m.

The dielectric losses of the HE_{21} mode are lower since the field is less concentrated in the waveguide. The simulated attenuation is 7.5 dB/m at 160 GHz and is less dependent on frequency. A precise measurement of the attenuation is not possible since a mode converter in split block design causes highly varying losses. However, an excitation of the HE_{21} mode is possible with a combination of two-mode converters [17], [28]. The radiation losses of the HE_{21} mode are higher than those of the HE_{11} mode. Thus, the higher order mode determines the minimum bending radius. Since no measurements are possible, the minimum bending radius is derived from [12] because field densities are comparable. The minimum measured bending radius with negligible losses in [12] is 1.5 cm, which also corresponds approximately to the maximum possible mechanical bending due to the larger dimensions. The susceptibility of the modes to distortion in the surroundings of the waveguide is also similar to [12].

Using a monopulse radar with two hybrid modes in the dielectric waveguide, dispersion must also be considered. Due to the different field distributions, both modes propagate with different group velocities, as shown in Fig. 6. The difference of 1.3×10^7 m/s results in a range offset between sum and difference signals in the radar response depending on the waveguide length. The offset can be calibrated using a defined signal path, such as the reflection of the lens.

C. Elliptical Lens Antenna

An important component in the monopulse front end is the antenna, as it significantly determines the ratio of the difference mode to the sum mode by the radiation characteristics of the individual modes. Furthermore, it determines the field of view (FoV), and the maximum range of the system can be set by the gain of the antenna.

The used antenna is an elliptical lens antenna, which is directly connected to the end of the dielectric waveguide

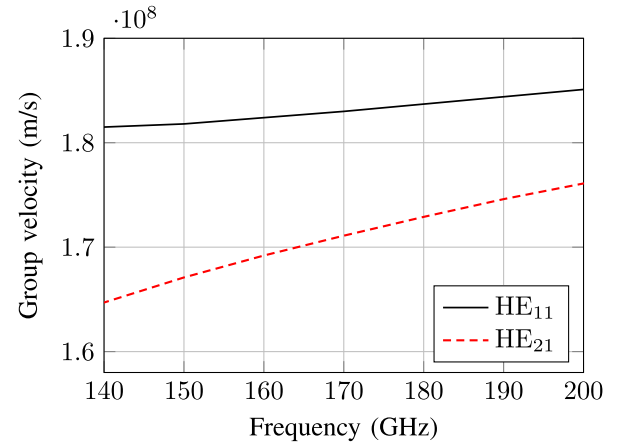


Fig. 6. Simulated group velocity for the HE_{11} and HE_{21} mode over frequency in a dielectric waveguide with a cross section of $1050 \mu\text{m} \times 2100 \mu\text{m}$.

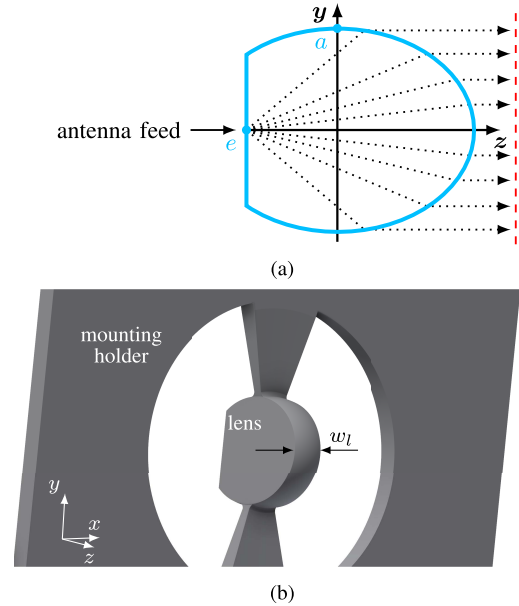


Fig. 7. (a) Schematic of the elliptical lens. (b) 3-D model of the fabricated lens antenna with mount.

as proposed in [17]. The lens is fed at its focal point e by the dielectric waveguide, as shown in Fig. 7(a). Due to the propagation delay within the dielectric material and the angle-dependent refraction, a plane wavefront propagates in free space. A rotation of the ellipse around the z -axis focuses on the wave in the E - and H -planes of the dielectric waveguide. Both modes in the dielectric waveguide are focused equally; however, the radiation patterns differ. Due to the minimum in the H -plane of the HE_{21} mode, a notch is obtained in the radiation pattern at 0° and the pattern corresponds to the difference beam. The sum beam results from the radiation of the HE_{11} mode.

Since the angle is estimated in the H -plane (x - z -plane), the beamwidth in this plane determines the FoV of the sensor. In order to increase the FoV, focusing on the H -plane must be reduced. Therefore, the ellipsoid is cut trimmed to the width of w_l , as shown in Fig. 7(b). With the rectangular base of the

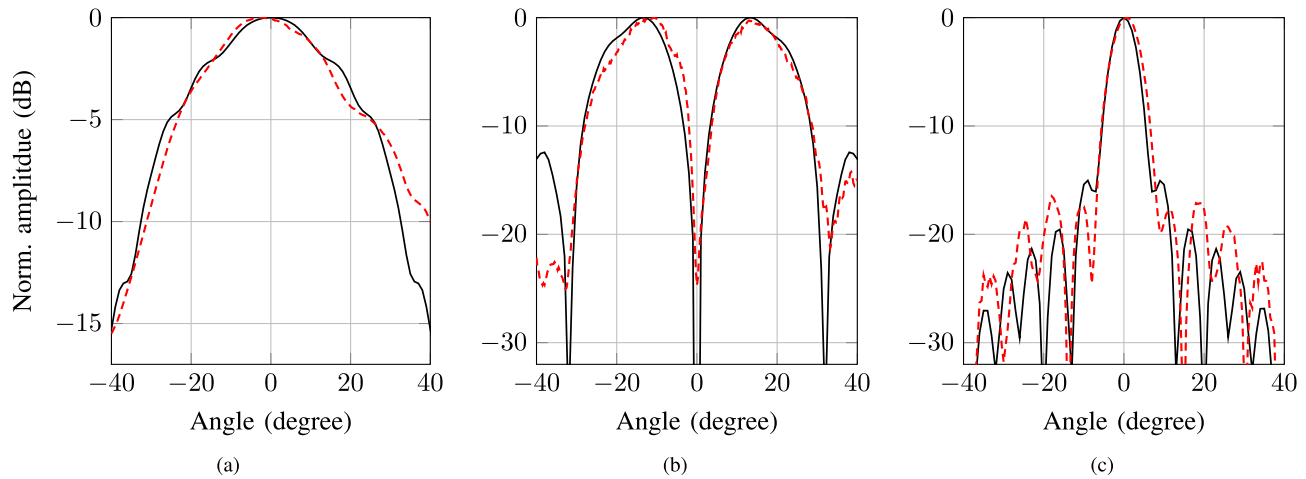


Fig. 8. Simulated (—) and measured (---) radiation pattern of the lens antenna at 160 GHz for (a) sum mode in the E -plane, (b) sum mode in the H -plane, and (c) difference mode in the H -plane.

lens, only focusing on the E -plane is achieved, whereas the H -plane radiates more broadly. The beamwidth as well as the gain of the antenna can be adjusted by the width w_l of the rectangular base.

A further design aspect of a monopulse antenna is the resulting ratio of the difference beam to the sum beam required for the angle estimation. The width w_l of the lens determines the ambiguity-free region and the steepness of the ratio. Increasing the lens width increases the gain and steepness but decreases the ambiguity-free region. Furthermore, the gain of the lens increases as well but to different extents in both modes.

The designed lens has a width $w_l = 4.1$ mm and a semi-minor axis a of 7.5 mm. The dimensions are a compromise between a sufficient gain of 20.6 dBi for both modes and the best possible angle estimation in the FoV from -17° to 17° . The ambiguity-free region and the beamwidth can be kept narrow since the direction of the main beam can be adjusted arbitrarily due to the mechanically flexible front end. The lens is made of HDPE to avoid mismatch at the feeding point. In addition, the cross section of the dielectric waveguide is tapered to a square cross section of $2100\ \mu\text{m} \times 2100\ \mu\text{m}$ at the interface of waveguide and lens to focus the field in the center of the waveguide. Thus, both modes are matched very well with a measured and simulated reflection coefficient below -15 dB from 140 to 180 GHz. Due to the low-loss material, the simulated efficiency of the lens antenna is above 90%.

The simulated and measured radiation patterns for both modes are shown in Figs. 8 and 9. For the measurements, the antenna with the dielectric waveguide was fed by a network analyzer with a rectangular waveguide interface. A mode transformer was used to feed the dielectric waveguide with the respective mode. The lens itself was fixed with a mount connected to the lens in the E -plane. The mount does not affect the radiation pattern since the field of both modes is focused on this plane. The sum beam in the H -plane has a measured angular width of 38° with its maximum at -2° and agrees well with the simulation, as shown in Fig. 8(a). The slight

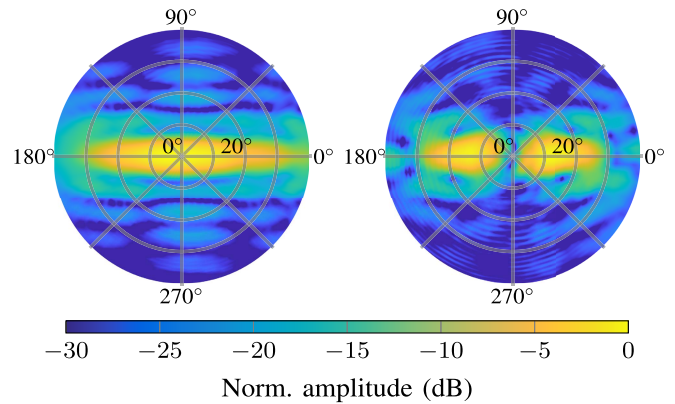


Fig. 9. Measured 3-D radiation pattern at 160 GHz for the sum mode (left) and difference mode (right) with the H -plane orientated horizontally.

shift of the main beam is caused by alignment inaccuracies of the flexible front end. For the difference mode, the two beam directions are $\pm 13^\circ$ with an angular width of approximately 13.5° [see Fig. 8(b)]. The simulated and measured values agree very well.

In the E -plane, the main beam is more focused. The measured 3 dB angular width for the sum beam is 6.2° and coincides very well with the simulated values, as shown in Fig. 8(c). The maximum measured sidelobe level is -15 dB. Due to the two beams for the difference mode, the complete 3-D pattern is shown in Fig. 9 with a 3-dB angular width in the E -plane of 7° .

D. Transition From MMIC to Dielectric Waveguide

With the transition from the MMIC to the dielectric waveguide, both the sum mode and the difference mode of the waveguide can be coupled to the MMIC. The insertion loss should be the same for both modes in order to preserve the expected ratio. In addition, the required chip area should be minimized.

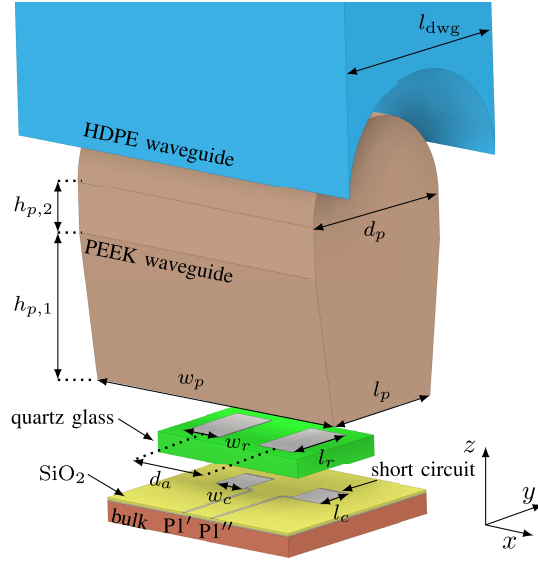


Fig. 10. Perspective view of a 3-D model of the transition from MMIC to dielectric waveguide.

Therefore, the transition presented in [29] is extended and optimized for a transition of both modes. It is composed of several components as shown in Fig. 10 and is explained starting from the MMIC side, although only the sum mode is excited here. The output signals of the two channels (P1' and P1'') feed two shortened $\lambda/4$ -patches on the topmost layer of the SiO₂ back end ($\epsilon_r = 4$ and $\tan \delta = 0.02$ at 160 GHz) on the MMIC. A quartz glass carrier ($\epsilon_r = 3.8$ and $\tan \delta = 0.001$ at 160 GHz) with two $\lambda/2$ -patches is positioned above to efficiently decouple the field out of the back end. The patches excite the respective modes in a rectangular polyether ether ketone (PEEK) waveguide. The last step is the feeding of the flexible HDPE front end using a mechanically decoupled and self-aligning transition.

The center frequency and the matching of the transition are determined by the lengths of the patches ($l_c \approx \lambda/4$ and $l_r \approx \lambda/2$). However, the wavelength of both modes is different. The different behavior also applies to the distance d_a between the patches, which are positioned centrally on top of each other. The distance d_a should be adjusted to obtain a maximum of the field intensity at the patch positions, minimize the required area, and minimize the crosstalk of the channels. Thus, a compromise must be made to excite both modes equally well.

In the PEEK waveguide positioned above the patches on the quartz glass, the respective modes should be excited and the parasitic radiation losses minimized. Therefore, a PEEK waveguide with a permittivity of 3.2 ($\tan \delta = 0.006$ at 160 GHz) is used in the first step instead of a direct coupling into the HDPE waveguide ($\epsilon_r = 2.25$). Further advantages are the mechanical decoupling of the front end and the MMIC as well as a significantly reduced waveguide cross section that simplifies the assembly.

The self-aligning feature of the transition is achieved with the cylindrically rounded PEEK waveguide, allowing the counterpart in the mechanically flexible HDPE waveguide to align

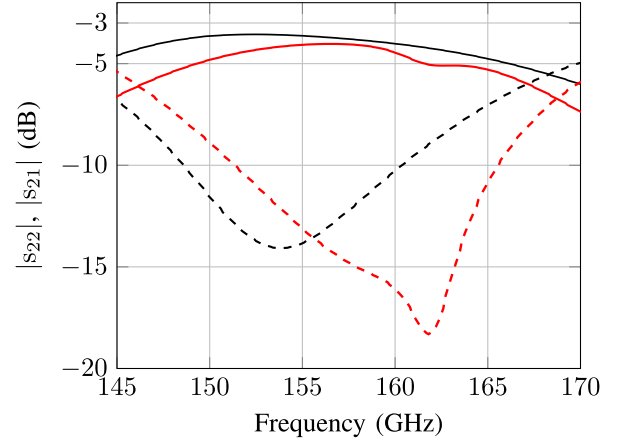


Fig. 11. Simulated transmission (—) and reflection coefficient (---) of sum mode (—/---) and difference mode (—/---) for the transition from dielectric waveguide to MMIC.

TABLE I
DIMENSIONS FOR THE TRANSITION FROM MMIC
TO DIELECTRIC WAVEGUIDE IN μm

patches	w_c	l_c	w_r	l_r	d_a
	180	240	220	390	470
PEEK	w_p	l_p	$h_{p,1}$	$h_{p,2}$	d_p
	1500	700	750	250	900

itself in case of misalignment or mechanical stress. An air gap of 75 μm is provided between PEEK and HDPE to allow for larger tolerances during assembly.

The transition was designed and optimized using a full-wave simulation with ports at the dielectric waveguide (port P2) and at the microstrip lines (ports P1' and P1''). The insertion loss of the two modes is calculated by the superposition of the two signals at ports P1' and P1''. The same is done in the radar system; however, only with the two IF signals in the digital domain. As shown in Fig. 11, the maximum simulated transmission $|S_{21,\Sigma}|$ is -3.5 dB at 152.4 GHz for the dimensions in Table I. The center frequency of the difference mode is shifted to higher frequencies, and the maximum transmission $|S_{21,\Delta}|$ is -4 dB at 156.5 GHz. The deviation of less than 1 dB for both insertion losses in the used frequency range results in slight changes in the amplitude ratio. The losses of the transition are caused by radiation losses, dielectric losses, and conduction losses. For the sum mode, 23% of the total losses result from radiation. Over 58% of the total losses are caused on the MMIC due to the lossy materials. The radiation losses in the difference mode are with over 30% of the total losses higher due to the different field distribution. However, the material losses are slightly lower.

When considering the reflection coefficients, a distinction must be made between transmitting and receiving. For the receive case, the reflection coefficient of the sum mode $|S_{22,\Sigma}|$ at P2 is below -10 dB from 148 to 160 GHz. The reflection coefficient of the difference mode $|S_{22,\Delta}|$ has a -10 dB bandwidth of 14.2 GHz from 151.3 to 165.5 GHz. For transmitting, an exact determination of the reflection factor

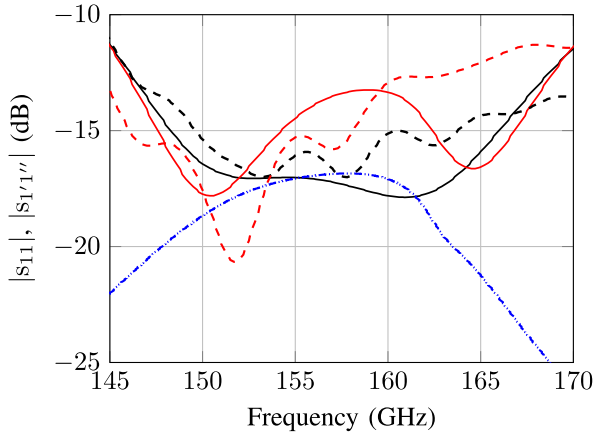


Fig. 12. Simulated (—) and measured (---) reflection coefficient $|s_{11}|$ of sum mode (—/---), difference mode (—/---), and the simulated crosstalk (---) of both channels.

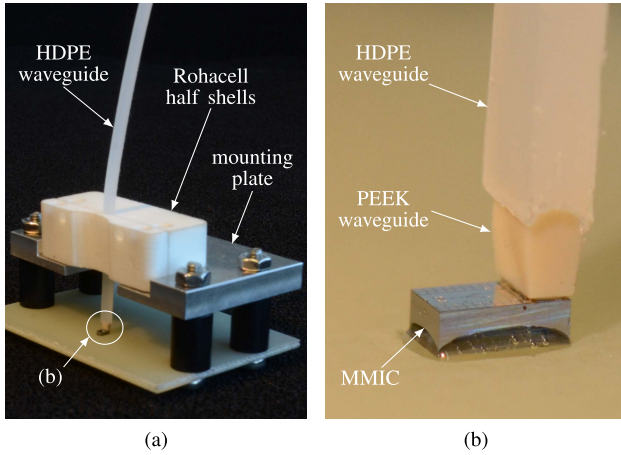


Fig. 13. (a) Realized transition from MMIC to the flexible dielectric waveguide. (b) Detailed view of the MMIC section.

is not possible because both $P1'$ and $P1''$ must be excited simultaneously. The crosstalk $|s_{1'1''}|$ is then added to the reflection coefficient $|s_{11}|$. The simulated reflection coefficient for both sum mode and difference mode is below -10 dB in the frequency range from 145 to 170 GHz since the crosstalk is below -16 dB, as shown in Fig. 12. The measured reflection coefficients of the realized transition (see Fig. 13) agree well with the simulated results. The slight deviations result from the measurement principle using a Y-divider to split the signal first. The nonidealities of the Y-divider are not considered in the simulation.

For the alignment, a high accuracy is required for the quartz glass and the PEEK waveguide on the MMIC. They should be positioned with an accuracy better than $40\ \mu\text{m}$, since not only the insertion loss for the difference mode increases but also the insertion loss of the sum mode decreases, resulting in a changed amplitude ratio. The mechanical decoupling and self-alignment of the HDPE waveguide front end and the PEEK waveguide is advantageous. The required assembly accuracy is reduced and a misalignment of $100\ \mu\text{m}$ increases the losses only by 0.2 dB.

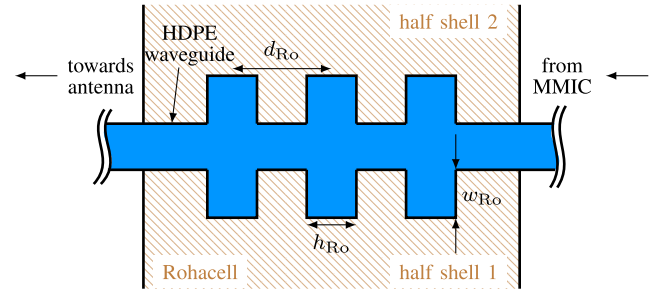


Fig. 14. Schematic cross section of the Rohacell holder.

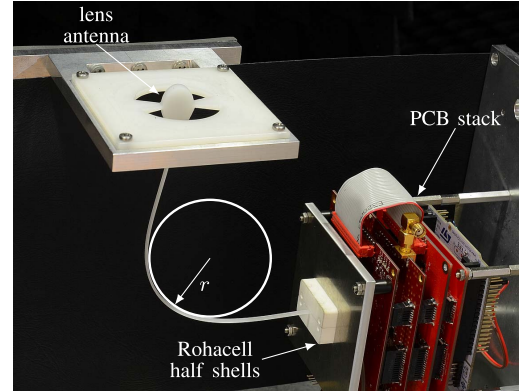


Fig. 15. Measurement setup with bent dielectric waveguide for the determination of the calibration data.

A further advantage of the transition is the compact design of the MMIC with a center-to-center distance of the shortened patches of $d_a = 0.24\lambda$. In addition, these patches only require an area of $0.011\lambda^2$ each, minimizing the required MMIC area and thus the costs.

E. Rohacell Mount

To fix the flexible front end above the MMIC, a mount is required with a minimal influence on the field distribution in the HDPE waveguide. According to Section III-B, the material surrounding a waveguide and the cross section must remain constant. Therefore, a Rohacell ($\epsilon_r = 1.05$, $\tan \delta = 0.005$) half shell as shown in Fig. 14 is used, enclosing the waveguide and fitting positively on the pins. These pins ($w_{Ro} = 1\ \text{mm}$ and $h_{Ro} = 0.75\ \text{mm}$) are placed at a distance of $d_{Ro} = 1.75\ \text{mm}$ to each other and cause radiation losses, which increases the insertion loss by approximately 0.6 dB. The losses of the difference mode are again slightly higher. Both Rohacell half shells are fixed on a mounting plate 1 cm above the PCB, as shown in Fig. 13(a).

IV. MEASUREMENT SETUP

To demonstrate the functionality of the proposed monopulse radar with flexible front end, radar measurements were performed in an anechoic chamber. The radar system consists of individual PCBs, separated in power supply, IF amplification, and signal conditioning. The PCBs are combined in a stack, as shown in Fig. 15. On top of the stack, the PCB with the radar MMIC and the flexible front end is positioned.

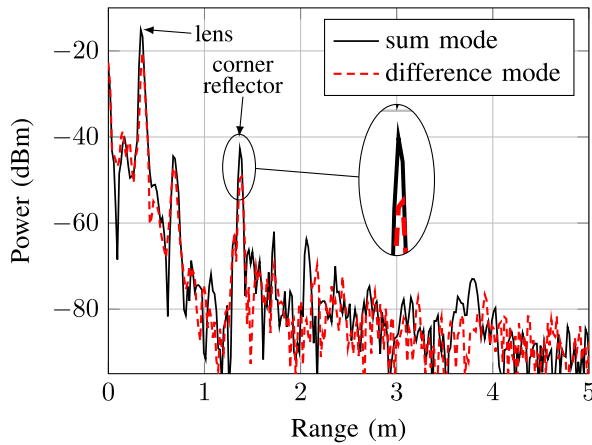


Fig. 16. Range spectrum of the sum mode and the difference mode for a corner reflector in a distance of 1 m to the antenna for $\varphi = 5^\circ$.

For the measurements, the antenna position is fixed, and a corner reflector with a radar cross section (RCS) of 4.9 m^2 is moved around the antenna by means of a robotic arm. The distance of the corner reflector is set to 1.0 m, and only the angular position in the H -plane is changed from -30° to 30° . A radar measurement was performed for each angular position, and the ratio of difference mode to sum mode in the measured range cell was determined from the IF signals of the radar measurement. This results in a calibration data set, which is used for angle estimation.

Since the front end can be moved mechanically, the same measurement procedure was performed with a bent waveguide at different radii r . The resulting data sets were compared with the original calibration data set, and the limits of the system were determined. All measurements were performed with a ramp duration of $100 \mu\text{s}$ and a modulation bandwidth of 8 GHz.

V. MEASUREMENT RESULTS

For the determination of the ratio, the amplitude of the target reflection is required. It is extracted from the range spectrum, as shown in Fig. 16. The distance between the corner reflector and the lens reflection is 1 m and the measured signal-to-noise ratio is approximately 40 dB. Since the corner reflector is positioned at 5° , the amplitude of the sum mode is 6.5 dB above the difference mode level.

The measured ratio of the difference mode Δ and the sum mode Σ for a straight dielectric waveguide is shown in Fig. 17. The ambiguity-free region is between -18° and 15° , and the simulated ratio agrees well with the measurements. The deviations are caused by a misalignment of the dielectric waveguide, which is only plugged into the antenna. A slight twist of the dielectric waveguide already affects the polarization and the mode excitation. Furthermore, in the simulation model, only the waveguide without transition is considered.

With a 90° bend with radius r (see Fig. 15), the influence of the mechanical flexibility on the system performance was investigated. The measured ratios for different radii in the angular range from -30° and 30° are shown in Fig. 18.

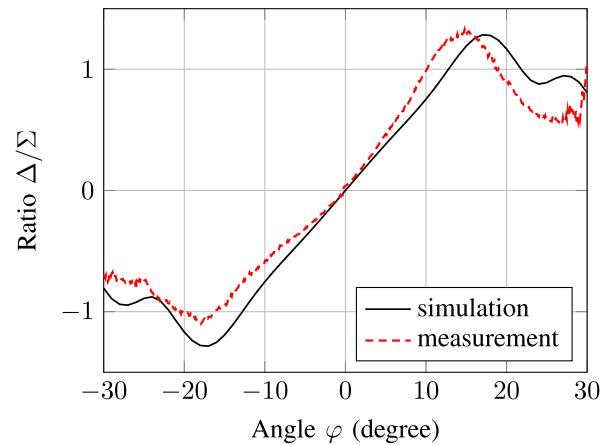


Fig. 17. Simulated and measured ratios of difference mode Δ to sum mode Σ for the angular range from -30° and 30° .

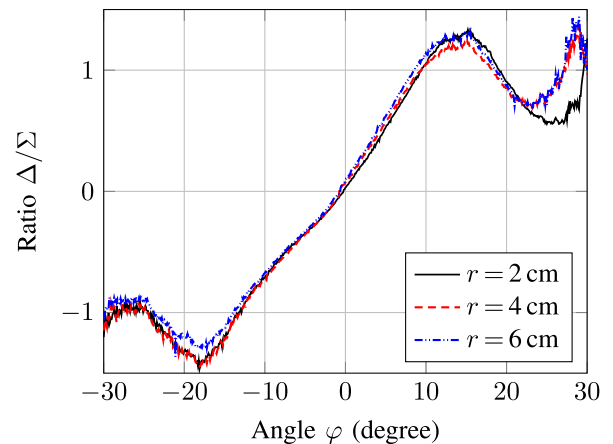


Fig. 18. Measured ratio of difference mode Δ to sum mode Σ for different radii in the angular range from -30° and 30° .

The ambiguity-free region of all measurements is approximately between -18° and 15° . The ratios differ especially at angles larger than $\pm 10^\circ$, resulting in an incorrect estimate. In the angular range between -10° and 10° , however, the estimation error is smaller than 1° and decreases toward lower values of φ . The reason for the deviations is not the bending of the waveguide but the exact alignment in the measurement setup. By changing the radius, the setup has to be moved, and in particular, a twisting of the waveguide in the antenna might occur. If the bending would have an influence on the measured ratio, a reduced slope of the curve would be expected with a decreasing radius since the difference mode is radiated more strongly than the sum mode. Even smaller bending radii were not investigated since this is mechanically only possible with a fixation of the dielectric waveguide. Compared to the measured ratio with a straight waveguide, the deviations are slightly larger due to the modified measurement setup.

The estimated angles for 50 radar measurements using the calibration data of the straight waveguide are shown in Fig. 19. The target was positioned at a distance of 1 m to the antenna at angles φ of 0° , 5° , and 10° . The dielectric waveguide was first straight ($\text{---}\times\text{---}$) and then bent with a radius of 2 cm ($\text{---}\diamond\text{---}$).

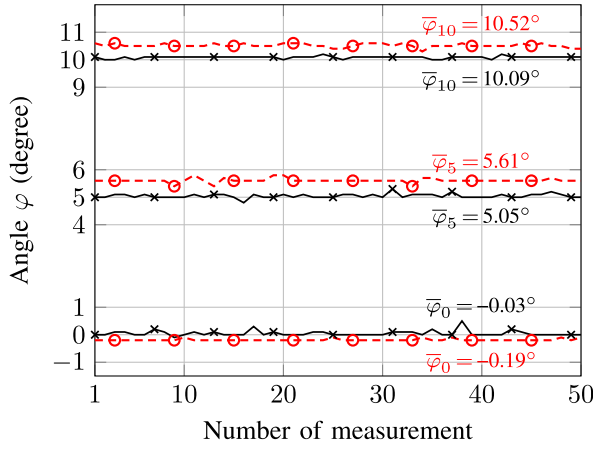


Fig. 19. Estimated angles from 50 radar measurements with a straight (—x—) and a bent dielectric waveguide (—○—) with radius $r = 2$ cm at φ_i ($i = \{0^\circ, 5^\circ, 10^\circ\}$).

The accuracy of the mean estimated angle $\bar{\varphi}$ for the system with the straight dielectric waveguide is better than 0.1° . The mean estimated angle $\bar{\varphi}$ of the system with the bent waveguide differs from the actual angle by less than 0.7° for the abovementioned reasons. However, the standard deviation of both setups is approximately equal and below 0.08° for the measured scenario. This corresponds well with the theoretically calculated value of 0.02° for the standard deviation of the angle estimation in monopulse radars according to [30].

The error sources and the resulting incorrect estimates could be avoided by fabricating both waveguide and antenna of one block.

VI. CONCLUSION

In this article, an imaging system at 160 GHz with a flexible front end based on a multimodal dielectric waveguide is presented. The system enables radar measurements with angle estimation in harsh surroundings and requires a minimum MMIC area.

The system is based on a two-channel MMIC that couples the signal via a self-aligning transition to the dielectric waveguide with an insertion loss for both modes below 4 dB. The feeding of the dielectric waveguide is designed to minimize the required MMIC area ($2 \times 0.011\lambda^2$) with a spacing between the excitation structures of $\lambda/4$. In the dielectric waveguide, the fundamental mode HE_{11} and the mode HE_{21} are propagating, which are received as sum and difference patterns via a modified lens antenna focusing the radiation pattern in the E -plane. The direction of arrival of the signal can be estimated using a digital amplitude monopulse principle in the angular range from -18° to 15° . Due to low losses in the dielectric waveguide (10 dB/m and 7.5 dB/m) and the mechanical flexibility, signals can be distributed over comparably long distances. The influence of bends with a radius of 2 cm on the angle estimation is negligible.

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