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RESEARCH PAPER

A 240-GHz circularly polarized FMCW radar based on a SiGe transceiver with a lens-coupled on-chip antenna

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A 240-GHz monostatic circular polarized SiGe frequency-modulated continuous wave radar system based on a transceiver chip with a single on-chip antenna is presented. The radar transceiver front-end is implemented in a low-cost 0.13 μm SiGe HBT technology version with cut-off frequencies f_T/f_{max} of 300/450 GHz. The transmit block comprises a wideband ×16 frequency multiplier chain, a three-stage PA, while the receive block consists of a low-noise amplifier, a fundamental quadrature down-conversion mixer, and a three-stage PA to drive the mixer. A differential branch-line coupler and a differential dual-polarized on-chip antenna are added on-chip to realize a fully integrated radar transceiver. All building blocks are implemented fully differential. The use of a single antenna in the circular polarized radar transceiver leads to compact size and high sensitivity. The measured peak-radiated power from the Si-lens equipped radar module is +11 dBm (equivalent isotropically radiated power) at 246 GHz and noise figure is 21 dB. The characterization bandwidth of the radar transceiver is 60 GHz around the center frequency of 240 GHz, and the simulated Tx-to-Rx leakage is below −20 dB from 230 to 260 GHz. After system calibration the resolution of the system to distinguish between two targets at different distance of 3.65 mm is achieved, which is only 21% above the theoretical limit.

Keywords: Radar architecture and systems, Si-based devices and IC technologies

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I. INTRODUCTION

Recently, the consumer and industrial market shows a demand for short-range radar systems because of the enormous applications such as automotive collision avoidance radars [1], healthcare sensors [2], high precision distance and motion detectors [3], non-destructive testing and quality control imaging systems [4], and security screening [5]. The noticeable trend toward higher operating frequencies is mainly motivated by increased lateral and range resolution, and finally, because of the smaller form factors.

Latest publications of radar systems operating in the 220 GHz transmission window and beyond include experimental radar systems based on III–V monolithic microwave integrated circuits (MMICs) [6, 7] as well as CMOS- [8] and SiGe-based [9–11] designs. Although classical rectangular metal waveguide technologies combined with III–V MMICs offer excellent performance in J-band (220–325 GHz) transceivers, their implementation remains quite costly and space consuming. Newest advances in SiGe bipolar technology make the mass production of fully integrated radar sensors at low-cost feasible.

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Most radar systems use linear polarization for its simplicity in implementation. However, linearly polarized radar systems are inefficient in signal-to-noise ratio (SNR) aspect. One reason for this inefficiency is the probability of wave depolarization at the reflecting target. In this case, the reflected signal is rotated by an angle θ and the received signal power is reduced by $\cos\theta$, thus leading to a decreased sensitivity of the radar. Another reason for this inefficiency is that radar transceivers with a single antenna do need a separation between transmit (Tx) and receive (Rx) paths. Because a circulator, which is usually fabricated with ferrite material, cannot be integrated monolithically, an equivalent quasi-circulator relying on directional hybrid couplers is generally used in the integrated circuit due to process compatibility.

However, for such implementation, half of the transmitted signal is lost across the terminating resistor of the coupler. For the reflected signal, again half of the received signal power is transferred to the transmitter where it is dissipated [12]. This kind of architecture with a 120 GHz SiGe radar MMIC and a single antenna with linear polarization, and the Tx and Rx paths being separated by a rat-race coupler was presented in [9]. To mitigate this power loss, a circular-polarization radar can be exploited, wherein a quadrature hybrid coupler in conjunction with a single dual-polarization antenna would be implemented.

Thus, the circularly polarized radar would lead to an improved SNR of at least 6 dB, compared with its linearly polarized counterpart [12]. In this paper, a 240-GHz circularly polarized frequency-modulated continuous wave (FMCW) radar system based on a SiGe transceiver with a single lens-coupled on-chip antenna is presented. The chip is packaged with a 4.5 mm diameter hyper-hemispherical silicon lens.

In the proposed configuration, the silicon lens acts as the chip substrate carrier and the secondary antenna at the same time. Therefore, no external optical components are required to increase its effective gain for terahertz applications. The antenna gain is adjustable, because the diameter of the lens and the length extension can be arbitrarily chosen. Additionally, the chip-on-lens packaging solution provides high radiation pattern quality due to the minimization of the influence of surface waves.

In general, high-gain antennas are needed for radar imaging applications at terahertz frequencies to compensate for the high path loss and to reduce the radar background clutter, thus increasing the SNR of the system. The implementation of two separate lens-coupled antennas on a single chip, one for Tx and one for Rx path, would lead to the off-center position of these two antennas in relation to the main optical axis. This configuration would in general cause a difference in pointing directions of the Tx and Rx antennas; depending on the lens extension. For low lens extension values the effective antenna gain is low, and therefore does not impose any major problem on the radar performance. However, due to the final antenna size, performance deterioration is expected for high gain values (space separation of Tx and Rx beams). Therefore, a two-antenna system, such as that presented in [10], was not targeted here. Additionally, compared with the solution proposed in [10], the here presented system has the advantage of high suppression of the unwanted radio frequency (RF) harmonics. In [10], a combination of a 120 GHz-VCO and a frequency doubler was implemented, where the fundamental harmonic from the VCO is not suppressed enough. Contrary to that, the transceiver implementation presented here offers a PA and a low-noise amplifier (LNA) serving as active 240 GHz bandpass filters.

In [8], a circularly polarized 60 GHz Doppler radar system is presented where the antenna array is implemented on a low-temperature co-fired ceramic substrate and attached to a CMOS radar transceiver. However, there are two antennas for the Tx and Rx paths. To the best of the authors' knowledge, the currently presented radar transceiver is the first that offers circular polarization with a single on-chip antenna implemented in silicon technology.

In comparison with the conference paper [13], the here presented radar module achieves higher SNR performance due to active cooling. Additionally, simulation results, and a more detailed analysis are added here.

II RADAR TRANSCEIVER CHIP

A) Monolithic implementation

Figure 1(a) shows a block diagram of the implemented circular polarized radar transceiver with a single antenna. It is composed of a differential dual polarization on-chip slot antenna connected to a quadrature hybrid coupler on one side, and the Tx and Rx signal paths on the other side. The local oscillator (LO) signal needs to be applied externally and is frequency multiplied by an integrated ×16 multiplication chain. Its output signal is split in two: one is fed into a power amplifier (Tx-PA) in the Tx path, and the other one is amplified by a LO–PA and serves as LO signal for the quadrature down conversion mixer. The Rx path is accomplished by a LNA. The quadrature hybrid coupler is a differential branch-line coupler and serves as a duplexer to separate the Tx and Rx signal paths. The Tx-PA is connected to one input port of the branch-line coupler, while the LNA is connected to the other one.

In general, the implemented antenna geometry allows obtaining any kind of linear or circular polarization of the transmitted signal by driving of the two differential input ports. A quadrature excitation of the feeds, exploiting a branch-line coupler, leads the transmitted signal to be circularly polarized. The implemented configuration forms a left-hand circular polarized (LHCP) antenna for the transmitted signal. The polarization of the signal being single reflected from the target is therefore changed to right-hand circular polarized (RHCP). The two received portions of signal power from the orthogonal polarization antenna ports are combined at the LNA input through the quadrature coupler.

Figure 1(b) shows a micrograph of the proposed circular polarized radar transceiver with the highlighted building blocks. The circuit is fabricated in IHP's 0.13 μm SiGe BiCMOS technology SG13G2 [14]. The f_T/f_{max} for this technology is 300/450 GHz. The fabricated chip measures 2.4 × 1.2 mm^2.

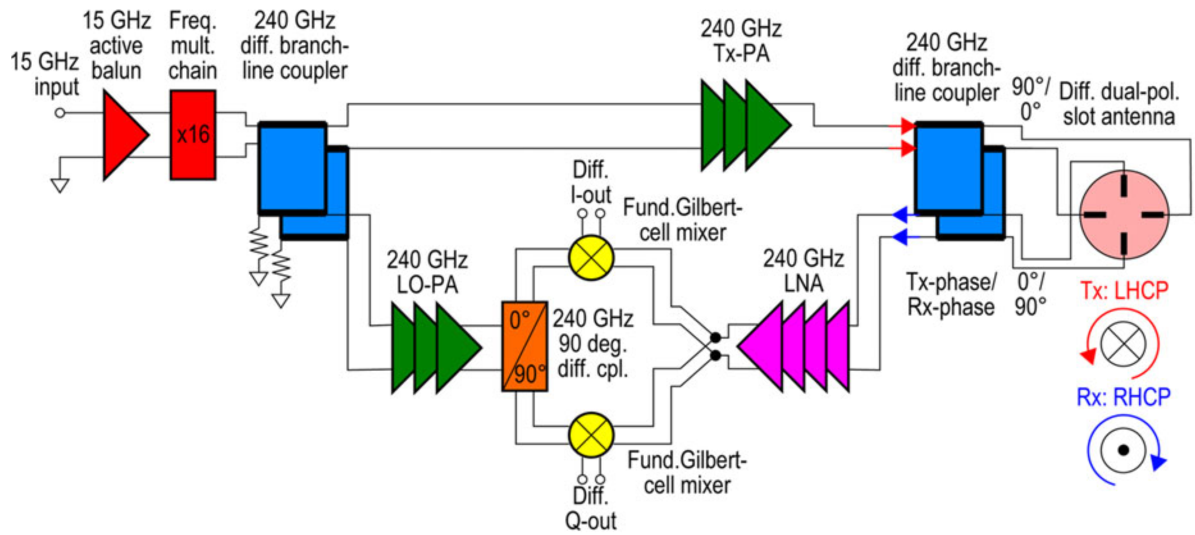
The core of the chip is the ×16 frequency multiplication chain, which consists of four cascaded Gilbert cells. The frequency doubling is realized by feeding the same input signal to the RF and the LO inputs of the Gilbert cell for mixing. No hybrids and intermediate drive amplifiers are used to reduce the chip area and power consumption. At the input of the first frequency doubler an active balun is used. The low-frequency input signal (13.1–16.9 GHz) with an input power of 0–5 dBm has to be externally applied for the radar operation.

The output signal from the frequency multiplication chain is equally split in two using a differential branch-line coupler, where the isolated port is differentially terminated with a 100 Ω resistor. One-half serves as the Tx signal: it is amplified by the three-stage Tx-PA (see Fig. 1(a)) and fed through the branch-line coupler to the antenna. The second half is used as the LO signal for the Rx block. It is amplified by the three-stage LO-PA, split, and phase-shifted by a 90° hybrid coupler, and used to drive the two Gilbert-cell down-conversion mixers fundamentally. The received RF signal is amplified by a four-stage LNA and fed to the RF-input of the quadrature mixer. For robust operation all on-chip building blocks and interfaces are realized fully differential.

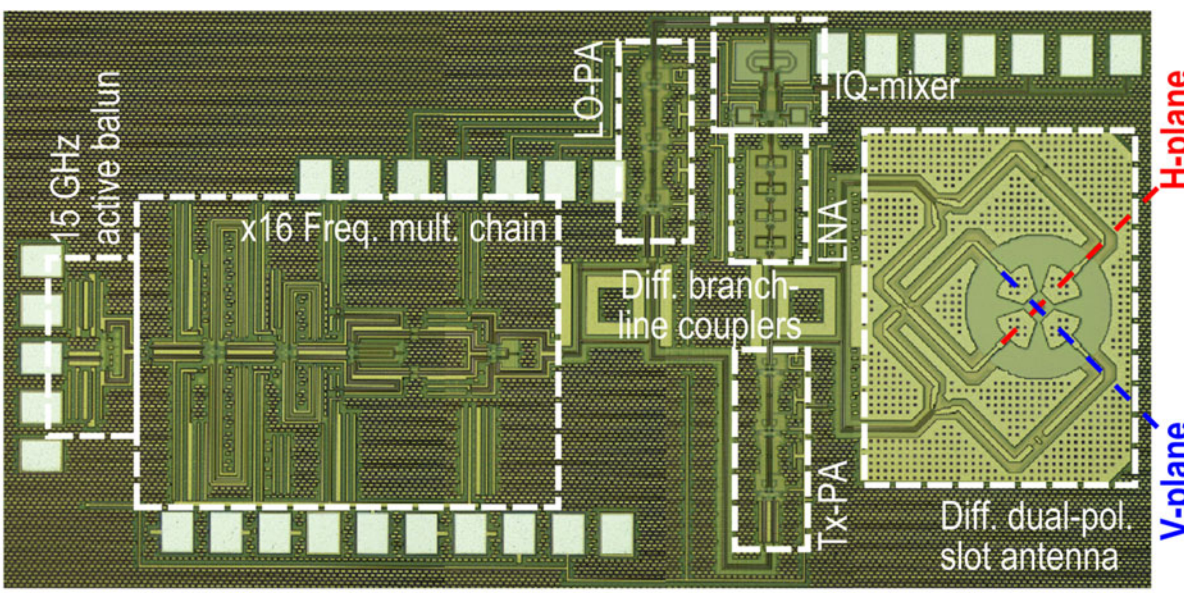
B) Packaged module

Figure 2 shows a cross-section of the Si-lens, the SiGe radar transceiver chip, the Printed Circuit Board (PCB), and the lens surrounding copper heat sink. The heat produced by the chip is transferred through the Si-lens to the copper heat sink. Additionally, the lens is cooled by air convection.

Figure 3 shows the completely assembled 240 GHz circular polarized FMCW radar module equipped with a 4.5-mm



(a)



(b)

Fig. 1. (a) Block diagram, and (b) micrograph of the monolithically integrated $2.4 \times 1.2\ mm^2$ 240 GHz radar transceiver chip implemented in SiGe technology.

silicon lens. The radar transceiver chip is mounted on the back side of the silicon lens, with its on-chip antenna aligned with the lens center. The size of the radar module PCB is 6 cm × 6 cm. The heat produced by the transceiver chip is transferred through the silicon lens to the copper heat sink, which is in direct contact with the lens. For proper heat dissipation, a Peltier element, and a fan are added for active cooling. Under normal operating conditions with the chip power dissipation of 1.4 W, the lens was measured with an IR camera to be at 39 °C.

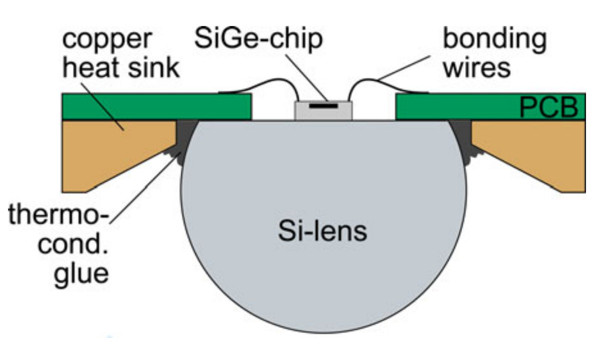


Fig. 2. A cross-section of the Si-lens, the SiGe radar transceiver chip, the PCB, and the lens surrounding copper heat sink.

III. SIMULATION RESULTS

A) Differential branch-line coupler

In the proposed implementation, a quadrature hybrid coupler is used to isolate the Tx and Rx channels and to provide quadrature excitation of the dual-polarization antenna for circular polarization operation. Therefore, amplitude and phase balance of the quadrature hybrid coupler are the crucial parameters for the quality of the radiated circular polarized beam

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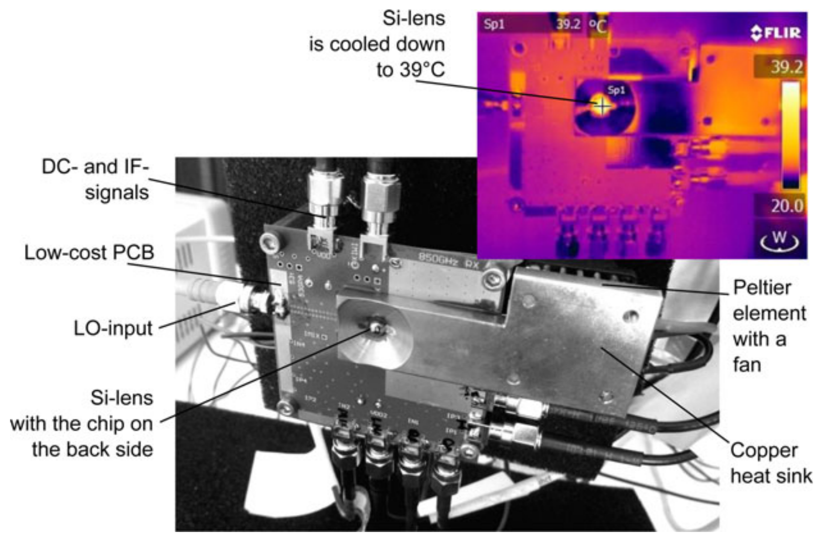

Fig. 3. Lens-packaged radar transceiver module with a copper heat sink, a Peltier element, and a fan (located on the back side of the heat sink) for active cooling purposes. The incorporated IR-image shows, that the 4.5 mm Si-lens with the SiGe chip on its back-side is cooled down to 39 °C.

as well as for the achieved SNR performance of the radar system. Additionally, high isolation between the Tx and Rx ports is important to prevent LNA saturation. Here, a branch-line hybrid coupler is favored, because it can be easily connected to the two feeds of a dual-polarization antenna with no need for crossover connections, whereas a rat-race hybrid has the disadvantage that its output arms are not adjacent and a crossover connection would be needed.

In Fig. 4(a), a three-dimensional (3D) simulation model of the designed differential branch-line coupler is shown. The structure consists of two interlocked wire rings; each wire ring is made up of four branch lines having an electrical length of one-quarter wavelength at the center design frequency of 240 GHz. The differential characteristic port impedance at 240 GHz is ca. 100 Ω, and the differential characteristic impedance of the cross branch lines is about 70 Ω. The main coupler structure is realized on the 3 μm-thick top level of metal 2 (TM2), and the underpasses connecting ports of the rings are placed on 2 μm-thick TM1. The bottom of TM2 is approximately 8.8 μm above the ground plane, and the effective dielectric constant of the insulating layers is approximately 4.1. The implemented differential lines are implemented as grounded coplanar strip lines. Each strip line has 2 μm width, with 4.8 μm-spacing for 100 Ω characteristic impedance, and 2.3 μm-spacing for 70 Ω characteristic impedance. Shielding ground metal blocks surround the coupler structure. The differential branch-line coupler occupies an area of $159 \times 156\ \mu m^2$, and its dimensions could be further reduced by folding the arms, like shown in [15].

Using full-wave EM simulation software Ansoft HFSS, the differential branch-line coupler was designed for maximum amplitude and phase balance over the operation bandwidth of 210 to 270 GHz, and for the highest possible isolation at the design center frequency of 240 GHz. The length of the branch lines and the width of underpasses were tuned through the well-known quasi-Newton optimization procedure in Ansoft HFSS.

Figure 4(b) shows the magnitude of simulated S-parameters for the branch-line coupler: return loss (S_{11}), isolation (S_{21}), and coupling (S_{31} and S_{41}). At the design center frequency, the return loss is below −20 dB, the port isolation reaches 30 dB, and the excess coupling loss is about 1.0–1.5 dB. Figure 4(c) shows that both the equal amplitude and quadrature phase conditions were closely matched (within 0.75 dB and 4°) over a wide frequency range from 210 to 280 GHz. In the simulation the isolated port was terminated with 100 Ω. Therefore, the designed differential branch-line coupler is highly suitable for its use as Tx-to-Rx isolator and 90°-phase shifter for circular polarization radar operation.

B) Lens-packaged module

To analyze the radiation efficiency and the Tx-to-Rx leakage of the radar transceiver, a chip-on-lens-packaged assembly was simulated in HFSS. A 3D EM simulation model of the packaged radar transceiver with silicon chip placed on the backside of the 4.5 mm-silicon lens with an extension close to the elliptical position [16] was created. The Si-lens is attached to the PCB, whereas the radar chip is placed inside a rectangular hole of the PCB. The PCB is modeled as a copper metal plate.

A differentially driven dual-polarization slot on-chip antenna in combination with a branch-line coupler was designed for broadband circular polarization operation. The antenna spans over the multi-layer BEOL stack and includes all dummy fillers to be compliant with the process design rules. A very detailed description of the antenna topology is out of scope of this paper. The coupled ports of the branch-line coupler (ports 3 and 4) are connected to the two feeds of the antenna, denoted as H- and V-plane antenna feeds (see Fig. 4(a)). The transmit signal is applied at port 1, and port 2 of the branch-line coupler is the input port of the receiver chain.

The chip thickness is 150 μm and the bulk silicon substrate resistivity is 50 Ω·cm. The simulated radiation efficiency of the complete packaged circular-polarized radar module is 0.31–0.37 over the frequency range of 210 to 270 GHz. In Fig. 5 the simulated return loss at the Tx-port and the Tx-to-Rx leakage of the packaged radar module are shown. The return loss at the Tx-port of the branch-line coupler is −19 dB at the design center frequency, and the Tx-to-Rx isolation is ca. 20 dB for the frequency range of 230 to 260 GHz.

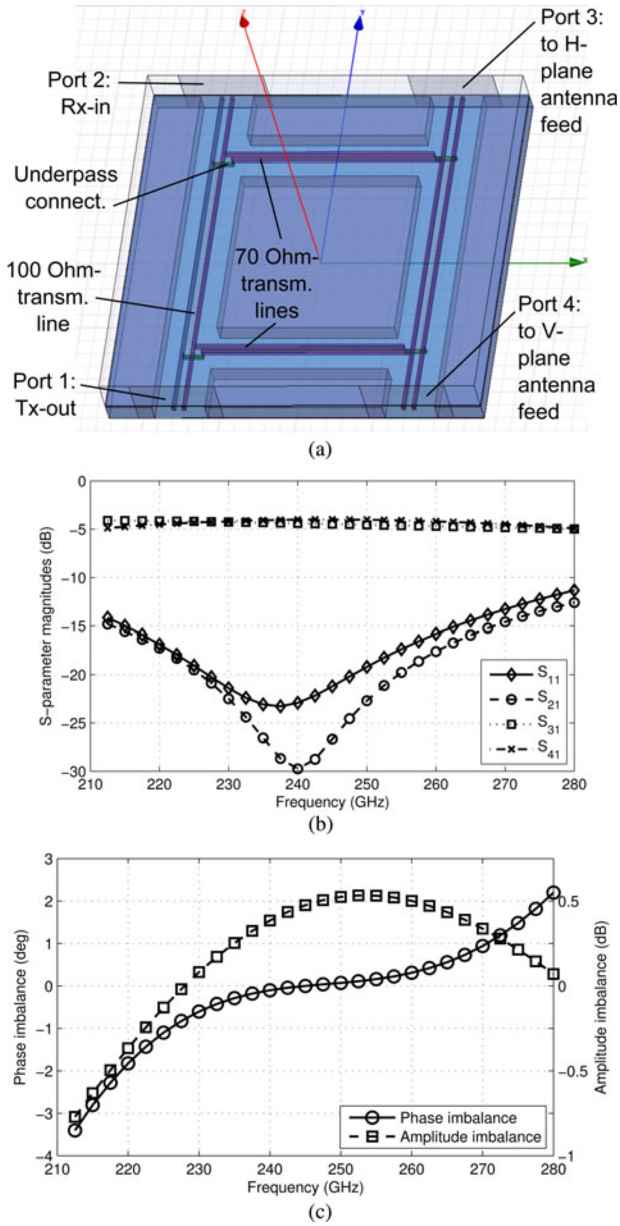

Fig. 4. 3D simulation model of the implemented differential branch-line hybrid coupler (a), simulated *S*-parameter magnitudes of return loss (S_{11}), isolation (S_{21}), and coupling (S_{31} and S_{41}) (b), and simulated phase and amplitude imbalance of the quadrature coupled ports.

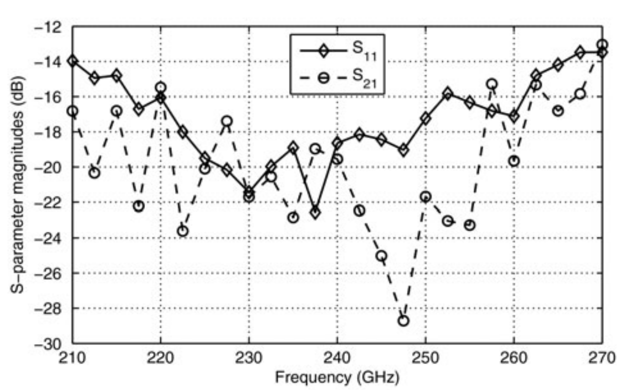

Fig. 5. Simulated return loss at the Tx-port is denoted as S_{11}, and the Tx-to-Rx leakage is referred as S_{21}.

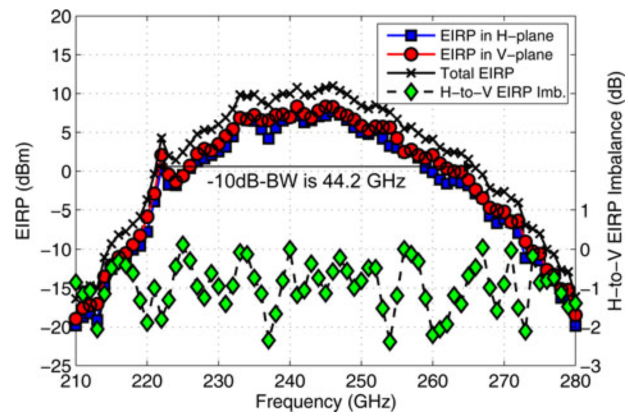

Fig. 6. The EIRP measured with a commercially available pre-calibrated Rx module with a WR-3.4 horn antenna at a 30-cm distance to the lens-integrated radar transceiver. The two portions of the circular polarized EIRP are measured in the both excitation planes of the dual polarization antenna (denoted as *H*- and *V*-planes). Additionally, the imbalance between the EIRP measured in the both polarization planes is plotted. The mean EIRP imbalance is −1.0 dB, and the total peak EIRP is +11 dBm at 246 GHz. The −10 dB operation bandwidth is 44.2 GHz (221.3–265.5 GHz).

IV. TRANSCEIVER MODULE CHARACTERIZATION

A) Transmit functionality

In Fig. 6, the equivalent isotropically radiated power (EIRP) measured in free space is presented. A pre-calibrated Rx reference module with a WR-3.4 20 dBi standard horn antenna is placed at a 30 cm distance to the radar transceiver. The EIRP is measured in the *H*- and *V*-planes of the dual polarization antenna (see Fig. 1(b)). The peak EIRP from the Si-lens equipped radar module measured at 246 GHz is +6.7 dBm in *H*-polarization plane and +8.3 dBm in *V*-polarization plane. Hence, the combined circular polarized peak EIRP is +11 dBm. The −3 dB-operational bandwidth of the combined Tx power is 21.1 GHz, while the −10 dB-bandwidth is 44.2 GHz (221.3–265.5 GHz).

The imbalance between the EIRP measured in the *H*- and *V*-polarization planes is plotted in Fig. 6, too. In the 210–270 GHz band, the mean EIRP imbalance between the two polarizations is −1.0 dB with standard deviation of 0.6 dB.

B) Receiver sensitivity

Both conversion gain (CG) and noise figure (NF) of the radar Rx chain were characterized with a pre-calibrated Tx reference source equipped with a WR-3.4 20 dBi standard horn antenna. Due to the fact that the reference antenna is linearly polarized, and the radar receiver needs a right-hand circularly polarized signal, the obtained CG and NF values have to be corrected by 3 dB. That is because half of the linearly polarized power impinging on the radar module is lost in the Tx path of the transceiver chip.

The distance between the reference source and the radar module was 30 cm. The output power of the source was carefully calibrated with the absolute power meter PM4 and the received power is measured from IF-outputs of the radar transceiver using two external baluns and an additional 90°-hybrid to combine the quadrature differential IF outputs

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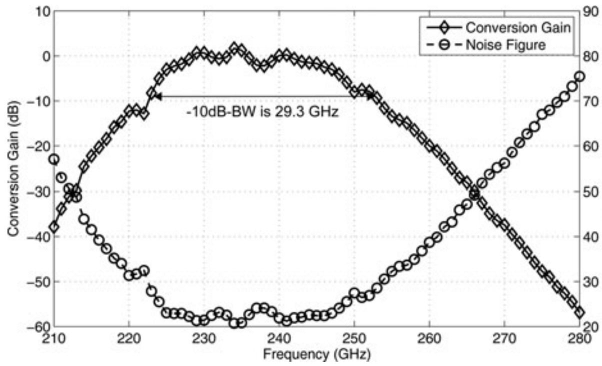

Fig. 7. Conversion gain and DSB noise figure of the radar Rx chain were measured from *I*/*Q*-output ports at IF-frequency of 33 MHz, which were combined by 90°. The distance between the Tx reference source and the radar module was 30 cm. The peak conversion gain of the receiver chain is 1.7 dB and the minimum noise figure is 20.7 dB at 234 GHz. The simulated directivity of the lens-integrated radar transceiver is 19.2 dBi at 240 GHz. The −10 dB operation bandwidth is 29.3 GHz (223.0–252.3 GHz).

to a single-ended signal. The 33 MHz-IF output signal and the output noise spectral density were captured with the spectrum analyzer.

The antenna directivity values for the reference horn antenna and the implemented lens-coupled on-chip dual-polarization antenna (operating in Rx mode) were extrapolated from the manufacturer data sheet (20 dBi at 270 GHz) and from the simulation results (19.2 dBi at 240 GHz), respectively. The frequency behavior of the measured Rx conversion gain and the noise figure can be found in Fig. 7. The maximum conversion gain of the radar receiver chain is 1.7 dB and the minimum DSB noise figure is 20.7 dB at 234 GHz. The −10 dB operation bandwidth measured from the frequency dependence of the conversion gain is 29.3 GHz (223.0–252.3 GHz).

As mentioned above, the temperature of the Si-lens could be cooled down to not less than around 39 °C due to its relatively small size. However, the chip is expected to be at a slightly higher temperature. Therefore, the actual noise figure of the LNA is assumed to be higher than the simulated one of about 14 dB at 25 °C. Also the real gain of the LNA should be lower than the simulated one due to an increased chip temperature. As stated in section III.B, the simulated radiation efficiency of the completely packaged circular-polarized radar module is 0.31–0.37 (over the frequency range of 210 to 270 GHz) that corresponds to about 5 dB loss in Tx mode. Because no anti-reflection coating for the Si-lens is used, the lens reflections in Rx mode do cause some excessive loss (compared with the Tx mode). Therefore, a 6 dB loss in Rx mode is estimated. The facts described above correlate quite well with the measured NF of around 21 dB.

V. FMCW TEST SET-UP

Figure 8 shows a block diagram of the FMCW radar system in the test configuration. The system comprises the 240 GHz circular polarized FMCW radar module, an off-chip linear frequency sweeper, a differential intermediate frequency (IF)-amplifier, a data acquisition unit, and a MATLAB digital signal-processing program. The linear frequency sweeper is used to generate sawtooth up-chirps from 13.1 to 16.9 GHz. It is based on a hybrid combination of a direct digital synthesizer (DDS), a phase-locked loop (PLL), and a voltage controlled oscillator (VCO). The DDS provides a chirp signal, which is used as the reference input for the PLL.

The duration of a sawtooth up-chirp is configured to be 2 ms. The resulting IF signal is sampled at 2 MHz. To reduce the receiver noise influence the obtained 4000 samples per chirp period are coherently averaged over 4000 consecutive IF signals in the time-domain.

In the current FMCW radar setup, the achieved system performance is analyzed on the beat signal obtained while an aluminum trihedral corner reflector shown in Fig. 9(a) is used as a single radar target. To extract an isolated single-target response, the beat signal is band-pass filtered in software. Here, only the acquired in-phase output signal of the receiver quadrature mixer is used for further signal processing steps.

Figure 9(b) presents the bandpass-filtered time-domain beat signal obtained with a trihedral corner reflector placed at 40 cm distance. The resulting IF signal is supposed to be an ideal single tone. However, a noticeable amplitude modulation of the obtained beat signal can be observed. Extraction of this parasitic amplitude modulation in the time domain is needed for further signal equalization. This step is crucial for achieving a near-theoretical system's performance, e.g. the range resolution of the FMCW radar system.

Figure 9(c) shows the corresponding normalized received power extracted from the magnitude of that bandpass-filtered time-domain beat signal after Hilbert-transformation. From this figure, operation RF bandwidth for the entire radar transceiver can be derived, if a target with a $1/\lambda^2$-frequency dependent radar cross-section (like a metallic plate or a corner reflector) is considered. The performance test results in a −6 dB-operation bandwidth for the radar transceiver of 14 GHz, while a −20 dB-bandwidth is 33 GHz with peak power at 235 GHz.

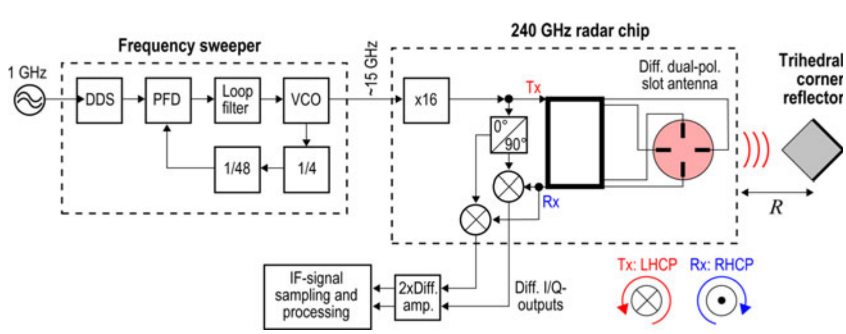

Fig. 8. Simplified block diagram of the presented FMCW radar system. The proof-of-concept for FMCW radar application is carried out using an aluminum trihedral corner reflector placed at distance *R* from the radar transceiver module.

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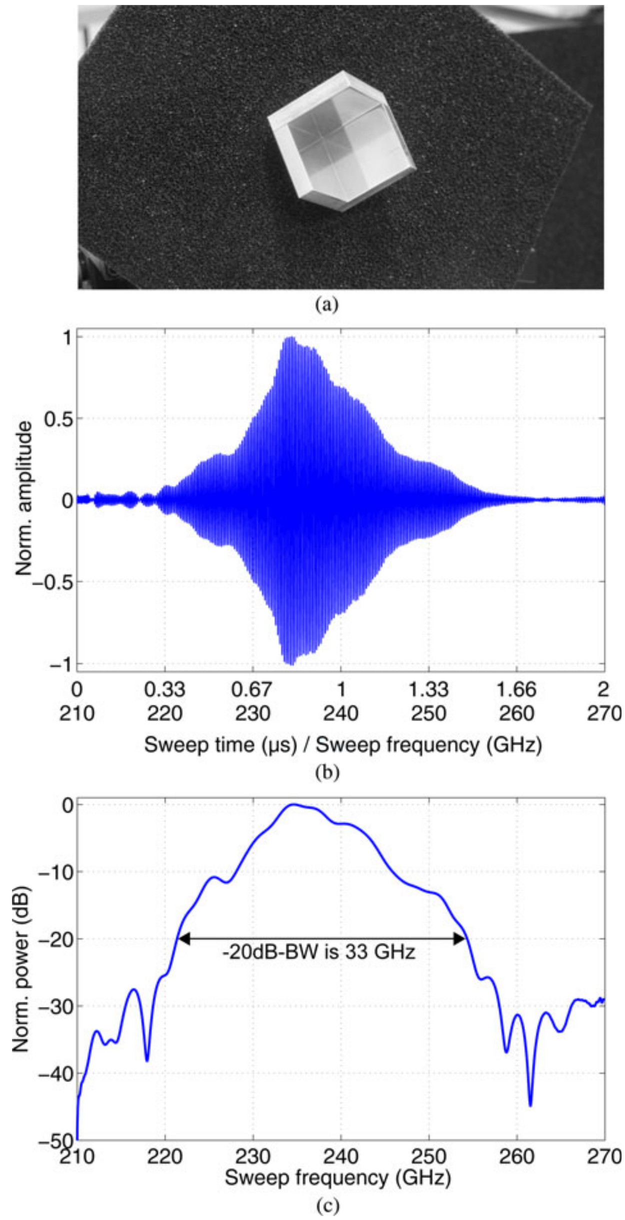

Fig. 9. (a) The aluminum trihedral corner reflector with edge length of 20 mm, which is used to characterize the range resolution achieved in the current FMCW radar setup. (b) Bandpass-filtered time-domain beat signal obtained with a trihedral corner reflector placed at 40 cm distance. The sweep time-scale is completed by the corresponding sweep frequency scale. (c) Normalized received power extracted from the magnitude of the bandpass-filtered time-domain beat signal plotted in (b). The −20 dB operational bandwidth is 33 GHz.

The maximum distance for the used corner cube was not measured; however, it can be roughly estimated by the following calculation. The radar cross-section (RCS) of a trihedral 20 mm-corner reflector at 240 GHz is 3.86 m^2. The simulated antenna directivity is about 19 dB at 240 GHz. Therefore, the estimated maximum radar range for an arbitrary SNR of 20 dB with the current 20 mm-corner cube is 15.9 m, when an equivalent noise bandwidth of 750 Hz (with Hanning-window), and a noise figure of 21 dB are assumed.

In the proposed setup, the system calibration is based on an IF signal, which is obtained when the distance between the radar transceiver and the corner reflector is set to 40 cm. This calibration signal is bandpass-filtered and the resulting time-domain amplitude modulation function is extracted after Hilbert-transformation. Subsequently, the corner reflector is placed at a distance of 60 cm, and the obtained beat signal is recorded again. The inverse of the extracted amplitude modulation function is applied to this beat signal in the time domain. After the amplitude of the beat signal has been corrected, the signal processing is carried out via standard fast Fourier transform (FFT) with zero-padding up to 131 072 points.

The so-called point spread function (PSF) of an imaging radar system indicates its resolution capability. To attain high-range resolution the transmitted bandwidth has to be maximized. The proposed radar system operates in a bandwidth of 60 GHz, ranging from 210 to 270 GHz. The RF bandwidth of 60 GHz results in the theoretical range resolution of the radar system (using rectangular window with correction factor of 1.21 for main-lobe full-width at −6 dB) of: $\Delta R_{\mathrm{rect},-6\mathrm{dB}} = c/2B \cdot 1.21 = 3.03$ mm, where c is the speed of light, and $B = 60$ GHz is the transmitted bandwidth.

Figure 10 presents the IF spectrum for the raw beat signal, and for the high-pass (HP)-filtered and amplitude corrected beat signal (using Hann-window) resulting from the test setup, where the corner reflector is placed at 60 cm distance from the radar module. For the HP-filtering an equiripple FIR HP-filter with a corner frequency of 112.3 kHz (attenuation of 1 dB in the pass-band), and a stop frequency of 102.3 kHz (attenuation of 60 dB in the stop-band) was applied to the beat signal, which contains the response from the target (located at a 60 cm distance) at 122.3 kHz. The IF frequency of the tone corresponding to the calibration target (located at about 40 cm distance) is 82.3 kHz. For the calibration procedure an equiripple FIR BP-filter with a pass-bandwidth of 20 kHz (with 1 dB attenuation), and a stop-bandwidth of 25 kHz (with 60 dB attenuation) around the center frequency of 82.3 kHz was applied.

Here, the resulting range resolution is measured using the main-lobe full-width at −6 dB. With uniform weighting the range resolution reaches 3.65 mm that is 21% higher than

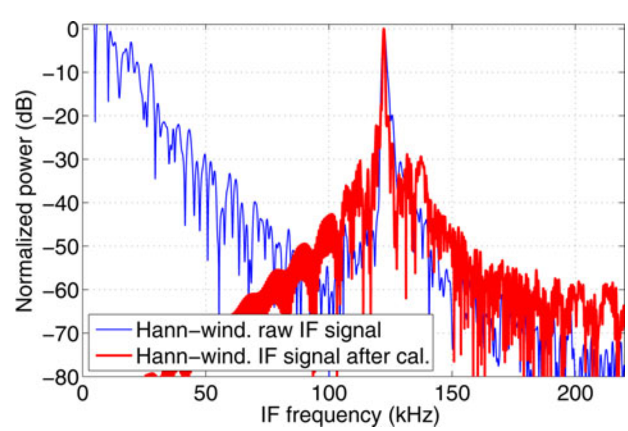

Fig. 10. The resulting IF spectrum of the raw beat signal and of the HP-filtered calibrated beat signal; Hanning window is applied in both cases. The corner reflector is placed at a distance of 60 cm from the radar transceiver module. The calibration is based on an IF signal, which was obtained when the distance between the radar transceiver and the corner reflector was set to 40 cm. A strong peak at the beat frequency of 122.3 kHz is observed for the current frequency chirp configuration. The data on the plot are normalized to the value of that peak power. The equivalent noise bandwidth is 750 Hz.

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the theoretical limit. When the Hanning-windowed results are compared, the measured range resolution descends to 5.09 mm that is only 2% above the corresponding theoretical limit for a Hanning-windowed signal.

As mentioned above, the wanted target response in the current setup occurs at 122.3 kHz. The beat signal of that frequency is created, when the frequency multiplier chain provides the wanted 16th RF harmonic. However, through extensive analysis of the implemented frequency multiplier chain, some unwanted RF spurious tones were found, which correspond to frequency multiplication factors of $\times 14$, $\times 15$ and $\times 18$. When the chip LO signal is swept to cover the RF frequency range of 210 to 270 GHz at the 16th harmonic, these spurious RF harmonics do land in the pass band of the PA and produce spurious IF beat tones. In the current experiment, the unwanted spurious tones corresponding to the $\times 14$ multiplication factor is located at about 107 kHz, the $\times 15$-tone at about 115 kHz, and the $\times 18$-tone at about 138 kHz. For the case of the uncalibrated raw beat signal the spurious IF tones are attenuated by minimum 40 dB. When the magnitude equalization step is applied, the level of spurious tones is still below -30 dB compared with the wanted response.

VI. CONCLUSION

A 240-GHz short-range circular polarized FMCW radar system based on a SiGe transceiver with a single lens-integrated on-chip antenna has been presented. The architecture is based on frequency multiplication, which offers a large operation bandwidth of 60 GHz. After successful calibration procedure, a measured range resolution of 3.65 mm was achieved, which is only 21% above the theoretical limit of 3.03 mm. The packaged radar module achieves a peak radiated power of $+11$ dBm (EIRP) and a noise figure of 21 dB. To the best of the authors knowledge, the currently presented radar transceiver is the first one with circular polarization and single on-chip antenna implemented in silicon technology. The presented lens-packaged radar transceiver allows the system to be used for a wide variety of short-range applications, such as SAR/ISAR imaging, 3D scanning imaging for security, providing a broad range of image modalities.

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