A Compact 140-GHz Radar MMIC with I/Q Downconverter in SiGe BiCMOS Technology

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Abstract—A monostatic D-band Doppler radar transceiver with a newly developed diode-based mixer is proposed which provides an I/Q signal at the baseband output and thus allows to determine the relative radial direction of a moving target. The quadrature downconverter shows a simple and compact structure since no phase shifters or couplers are needed. The transmitted signal is generated by a low-noise differential voltage-controlled oscillator (VCO) with a frequency band of 136 to 146 GHz. The whole transceiver consumes 74 mA from a 3.3 V supply, with the downconverter and baseband amplifiers contributing about 2 mA. The transceiver shows reliable functionality over a wide range of silicon bulk temperature up to 125°C.

Keywords—radar transceiver, quadrature downconverter, BCMOS, diode mixer.

I. INTRODUCTION

In recent years, D-band-frequencies have become increasingly relevant for radar systems. E.g., for automotive applications, frequency bands around 140 GHz are of interest [1]. Especially for Doppler radar applications [2] it can be advantageous to use higher frequencies than the established 77 GHz due to the Doppler frequency’s linear reliance on the transmitted frequency. This paper presents a compact monostatic Doppler sensor which allows the evaluation of I/Q signals at the baseband output. Unlike common I/Q receivers, no dedicated 90-degree coupler or phase shifter is needed which enables a very compact layout realization.

In contrast to bistatic radar systems, monostatic receivers use a single antenna for both transmitting and receiving the signal. This leads to smaller required chip area in case of a chip-integrated antenna or a more compact radar module when an external antenna is used, respectively, and therefore results in better cost efficiency. However, conventional monostatic transceiver concepts typically require coupling structures for RX/TX separation which come along with high losses for both directions [3][4].

To overcome this drawback, a very compact concept of a diode-based monostatic transceiver topology has been introduced in [5] where each differential VCO output is connected to a diode in series as shown in Fig. 1a.

In comparison to other mixer realizations, such diode mixer shows advantages like the simple passive structure and applicability to a wide range of output power. Furthermore, a better signal-to-noise ratio (SNR) can be achieved due to the absence of high-loss coupling structures for separating the RX and TX paths. In a conventional monostatic transceiver, the coupler induces ≥3dB loss for both TX and RX signal. This results in a SNR degradation of ≥ 6 dB. In the proposed approach, essentially only the diode’s insertion loss is relevant for the SNR degradation [5].

Fig. 1. Principle structure of VCO and diode-based mixer with parallel diodes (a) and proposed diode-based I/Q mixer using shifted diodes (b).

In this circuit shown in Fig. 1a, the differential nature of the emitted signal provided by the VCO is maintained while both paths provide identical and therefore redundant baseband information of the incoming signal. However, this basic structure offers several opportunities to gain more information of the signal. The presented circuit extends the above principle to an I/Q transceiver that easily enables to determine the phase difference and thus the relative radial direction of movement of a target. For this purpose, the principal structure of Fig. 1a is adapted by adding shifted λ/8 transmission lines in the signal paths between VCO output and diode or diode and output pads, as shown in Fig. 1b. Thus, both differential paths are not identical any more but still exhibit the same total length from the VCO output to the output pads. These λ/8 lines before and after the diodes cause a 45-degree phase shift each or a total relative phase shift of 90 degrees at the baseband output, respectively, which leads to an I/Q baseband signal. Since the total length between the differential transformer output and the respective output pad is not affected by this diode shifting, there is no effective phase shift for the transmitted signal which still is a fully differential signal with typical advantages like higher robustness against external disturbances. In addition, the used transmission lines are necessary in any case to connect the VCO output to the diodes or the diodes to the output pad and no separate 90-degree coupler is needed. For the intended Doppler...
radar application, an exact 90-degree phase shift is not required since only the sign of the phase difference is relevant. Therefore, the frequency dependence of the phase shift introduced by the λ/8 lines does not limit the operating frequency bandwidth of the circuit. For the frequency range of 136 GHz to 146 GHz the theoretical phase deviation is within ±3.5 degrees. Table 1 provides a summary of the proposed concept in comparison to other recently published D-band radar transceivers.

In the following sections, the proposed transceiver circuit is presented in more detail and its functionality is demonstrated by means of the results of different measurement scenarios.

Table 1: Concept comparison to other D-band radar transceivers.

<table>
<thead>
<tr>
<th>Frequency range</th>
<th>[8]</th>
<th>[9]</th>
<th>[10]</th>
<th>[5]</th>
</tr>
</thead>
<tbody>
<tr>
<td>110-124 GHz</td>
<td>122 GHz*</td>
<td>144.6-150 GHz</td>
<td>136-146 GHz</td>
<td></td>
</tr>
<tr>
<td>Process</td>
<td>130nm SiGe</td>
<td>130nm SiGe</td>
<td>130nm SiGe</td>
<td></td>
</tr>
<tr>
<td>RF channels</td>
<td>RX, 1 TX</td>
<td>RX, 1 TX</td>
<td>RX, 1 TX</td>
<td>RX, 1 TX</td>
</tr>
<tr>
<td>RX mixer</td>
<td>not specified</td>
<td>Gilbert-cell (differential)</td>
<td>Gilbert-cell (differential)</td>
<td>Gilbert-cell (differential)</td>
</tr>
<tr>
<td>IQ realization</td>
<td>n/a</td>
<td>90-degree coupler (branchline)</td>
<td>90-degree coupler (branchline)</td>
<td>90-degree coupler (branchline)</td>
</tr>
<tr>
<td>RX signal generation</td>
<td>on-chip VCO</td>
<td>external LO with on-chip PA</td>
<td>external LO with on-chip PA</td>
<td>on-chip VCO</td>
</tr>
<tr>
<td>on-chip antenna</td>
<td>no</td>
<td>yes (2x1)</td>
<td>no</td>
<td>yes</td>
</tr>
</tbody>
</table>

* center frequency (external LO input)

II. CIRCUIT DESIGN

The transceiver is fed by a differential-output Colpitts push-pull VCO that was designed similar to the concept shown in [6] with the focus on low phase noise, robust behavior regarding varying chip temperature or supply voltage and sufficient output power level without additional consequent power amplifier. The VCO provides an output power of about -2 dBm within a tunable frequency range of 10 GHz around the D-band center frequency of 140 GHz. As shown in Fig. 2, the differential VCO outputs are connected to one transistor (T1, T2) with shorted collector-to-base in each case so that it serves as a mixing diode in both respective differential signal paths. A small bias current through the diodes allows a trade-off between insertion loss for the transmitted signal and conversion loss for the received signal. Since the capacitors C1, C2 form a short circuit for the RF signal, λ/4 lines TL3, TL4 are used to transform this short into an RF open so that the VCO power is entirely led to the antenna output.

As per the monostatic structure, the incoming signal is detected by the same antenna and is affected by a phase shift of both differential paths when reaching the respective diodes where it is mixed with the VCO signal. At the output pads, stub lines to ground represented by TL9, TL10 are added in order to compensate parasitic pad capacitance and provide a return path for the bias current of the mixer diodes.

Each of the λ/8 lines TL5, TL6 contributes a phase shift of π/4 which results in the desired downconversion: The oscillator signal provides a frequency of f0 and shows an arbitrary phase of φ0 after TL1 where it acts as the LO signal

\[ s_{LO1}(t) = \cos(2\pi f_0 t + \varphi_0) \]  

for T1. Due to the differential VCO output and the 45-degree phase shift induced by TL5, the corresponding input signal for T2 can be written as

\[ s_{LO2}(t) = \cos(2\pi f_0 t + \varphi_0) \]  

The antenna receives an echo signal shifted by ±fD relatively to f0 with an unknown phase of φRX. The signal

\[ s_{RX}(t) = \cos(2\pi (f_0 \pm f_D) t + \varphi_{RX} - \frac{\pi}{4}) \]  

reaches T1 with an additional 45-degree phase shift caused by TL6. For TL2, the received signal

\[ s_{RX}(t) = \cos(2\pi (f_0 \pm f_D) t + \varphi_{RX} - \pi) \]  

maintains the relative differential phase shift of the transmitted signal. By neglecting the RF mixing product at 2f0 ± fD, the consequent baseband signals can be written as follows

\[ s_{IF1}(t) = \cos(\pm2\pi f_D t - (\varphi_0 - \varphi_{RX} + \frac{\pi}{4})) \]  

\[ s_{IF2}(t) = \cos(\pm2\pi f_D t - (\varphi_0 - \varphi_{RX} - \frac{\pi}{4})) \]  

By comparing the phase shifts, the intended phase difference of ±π/2 is obtained, depending on the sign of fD.

The frequency fD of the downconverted signal at the baseband output is expected to have a frequency well below 1 MHz. E.g., according to the Doppler equation for monostatic radar systems

\[ f_D = \frac{2\nu_r}{\lambda} \]  

a radial velocity of \( \nu_r = 1 \) m/s leads to a frequency shift of about 930 Hz. The baseband signals are led to the output pads where a coupling capacitor is connected externally and are then fed to a respective buffer amplifier on the same chip. The amplifier output signal can be taken for external measurement, data processing, and evaluation in order to verify the proper functionality of the transceiver system.

III. TECHNOLOGY

The proposed circuit was realized in Infineon’s 130 nm SiGe BiCMOS process B11HFC with fT/fmax of 250 GHz/370 GHz [7]. The metal stack consisting of six copper layers can be seen in Fig. 3. The thickest metal layer M6 is used for microstrip transmission lines (Z0 = 50 Ohms) with M4 as associated ground plane. A chip micrograph of the transceiver is shown in Fig. 3 with the size of 800×1160 µm² (including pads).
When connected to a 3.3 V supply the transceiver consumes a current that amounts to 74 mA. The major part of 72 mA is related to the VCO core including its bias current (40 mA) and divider chain (32 mA) while the receiver including baseband amplifiers consumes about 2 mA.

For output power measurement of the transceiver chip a VDI Erickson PM5B power meter was used in combination with a D-band waveguide probe which was connected to one of the output signal pads while the other one was terminated on chip with 50 Ohms. Since a single-ended RF probe was used, 3 dB are added to consider the differential output path. According to its datasheet, the used D-band probe has an insertion loss of 2 dB in addition to the approximated waveguide and transition losses which were experimentally determined as 1.4 dB. This leads to a total power loss of 6.4 dB which has already been added to the measurement results shown in Fig. 4 which presents both the output power of the transceiver in total as well as of the VCO alone which allows to estimate the insertion loss caused by the mixer structure.

For a chip temperature of 25°C, the measured output power of the transceiver is about 2 to 2.5 dB lower than the output of the VCO itself which can be largely attributed to the diode’s insertion loss of each path and to a lesser degree is caused by losses due to transmission line attenuation. In terms of tuning frequency characteristics or phase noise behavior, the VCO performance is very little affected by the additional transceiver circuit in comparison to its stand-alone measurement results. The performance of the system was characterized within a temperature range of -40°C to +125°C. In Fig. 4 the respective results for +50°C and +125°C are shown. The circuit remains functional over the full measured temperature range and the output power decreases by a maximum of 1.5 dB.

The connection to the power meter was then replaced by a D-band pyramidal horn antenna with 23 dBi gain in order to radiate the signal power and conduct continuous wave (CW) radar measurements. The mixer output pads were connected to the respective on-chip baseband amplifier input pad via an external coupling capacitor. The amplifier outputs were contacted by the probe of a digital oscilloscope in order to measure the transient baseband signal and save the transient signal for further digital processing and frequency analysis.

To verify the principal functionality of the continuous wave radar transceiver, a corner reflector was placed in front of the horn antenna and moved along the main radiation direction. The detected transient signal showed typical behavior of a monostatic CW Doppler radar with clear changes in frequency depending on the axial target velocity. In Fig. 5a, the transient Doppler signals at the outputs are shown when the target reflector was moved towards the antenna from frontal direction and with constant velocity of about 0.11 m/s while the VCO provided a signal at 140 GHz. The associated spectrum presented in Fig. 5b shows one single sharp peak at the equivalent Doppler frequency according to equation (7) which represents the target.

For another measurement scenario, the corner reflector was mounted on a pendulum that oscillated along the radial axis in front of the antenna with low angle of displacement. In this case, a maximum Doppler frequency of about 1000 Hz was observed when the pendulum passes its equilibrium position. When a window function is shifted over the transient signal and the spectrum is calculated for each point and its selected neighborhood within the window separately, the instantaneous frequency over time is obtained as shown on the left axis in Fig. 6a. For reasons of simplicity, the signal of only one output is shown since the second one’s behavior is equivalent. It can clearly be observed how the target reaches maximum velocity at the equilibrium position and has a velocity of zero when it reaches the points of maximum deflection which correspond to maximum or zero instantaneous frequency, respectively.

The corresponding instantaneous radial velocity is shown on the right axis. Eventually, the windowed frequency spectra are used to determine the instantaneous phase over time. Unlike the figures of frequency shown before, the phase signals of both baseband outputs are not congruent but affected by the relative phase shift induced by the shifted diodes. Depending on which signal is delayed, a sign of +1 or -1, respectively, can be...
assigned to the difference angle or relative phase shift of their instantaneous phases to represent the radial direction of the target. In the case of the pendulum, the target is alternating approaching the antenna and moving away in a mostly evenly manner that is only limited by some nonideal influences like slight deviations from the ideal radial axis or deceleration due to noticeable friction at the fulcrum of the pendulum. This behavior can be clearly seen in Fig. 6b where the instantaneous phase difference is alternating according to the turning points of the pendulum with zero frequency.

![Instantaneous Doppler frequency over time and corresponding radial velocity](image1)

**Fig. 6.** Instantaneous Doppler frequency over time and corresponding radial velocity (a) and normalized relative phase shift of the I/Q output signals (b) for a target mounted on an oscillating pendulum.

V. CONCLUSION

A diode-based differential CW radar transceiver with high temperature robustness has been realized that uses a single receive/transmit antenna whilst circumventing the typical drawbacks of conventional monostatic transceiver systems. It shows a concept of providing an I/Q baseband signal without the need of a separate 90-degree phase shifter which allows to easily and reliably detect the relative radial velocity and direction of a moving target. The functionality has been shown by means of different measurement scenarios.

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