An 80 GHz SiGe Quadrature Receiver Frontend

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***Abstract*— An In-phase / Quadrature (I/Q) receiver frontend for communication and sensing (e.g. radar) applications in the frequency range between 75 and 86 GHz is presented. The frontend includes a low-noise amplifier (LNA), two down–conversion Gilbert mixers, a branchline coupler for the generation of the in-phase and quadrature LO signals, and LO buffers. The circuit was designed and fabricated in a 200 GHz f_T SiGe:C bipolar technology. The chip performance was characterized by on-wafer measurements. From 75 GHz to 86 GHz, the frontend exhibits a single sideband noise figure of 11 dB, and a conversion gain that is higher than 28 dB. The input-referred compression point is -16 dBm. The chip occupies an area of $1000\mu\mathrm{m} \times 1100\mu\mathrm{m}$. The circuit works from a positive supply voltage of 5.5 V and draws 195 mA.**

I. Introduction

Sensing and data transmission applications initiate a high interest for research and product development in the frequency range around 80 GHz. The particular fields of work up to now include long range and short range automotive radar from 76 GHz to 77 GHz and 77 GHz to 81 GHz [1], respectively, and data communication in the frequency bands 71–76 GHz and 81–86 GHz. These frequency ranges have been dominated by compound III–V semiconductors. Over the past years, improvements in silicon-germanium led to transit frequencies f_T and maximum oscillation frequencies f_{max} beyond 200 GHz, making this technology an alternative choice [2],[3].

A quadrature receiver frontend is an essential component for data transmission and precise sensing systems. At the ISM band around 60 GHz, good results have already been obtained: Floyd et. al. [4] presented various building blocks, like a low noise amplifier with a noise figure of 4.5 dB, or an I/Q mixer with a conversion gain of 18.6 dB and a noise figure of 13.3 dB. In that work, the phase difference of the I/Q outputs varies between 87° and 105° from 57 to 64 GHz. At 77 GHz, Kaleja et al.[5] realized a passive I/Q Schottky diode mixer, exhibiting a conversion loss of 10 dB. In a previous work, a 77 GHz low noise amplifier exhibiting a noise figure of 5 dB and a gain of 9 dB for automotive applications was presented by the authors [6].

This paper presents a direct conversion quadrature receiver frontend for applications around 80 GHz. The circuit was designed for use in both sensing and communication systems. The frontend was implemented in two steps: First, the low noise amplifier and the quadrature receiver were fabricated and tested individually. Then, both circuits were merged to the presented frontend.

II. Circuit Design

The receiver frontend consists of a single–ended low-noise amplifier (LNA), balanced-to-unbalanced transformers to convert the single–ended signal to differential signals, two fully differential direct conversion mixers, LO buffer amplifiers, and a branchline coupler. An overview of the implemented circuit is depicted in Figure 1.

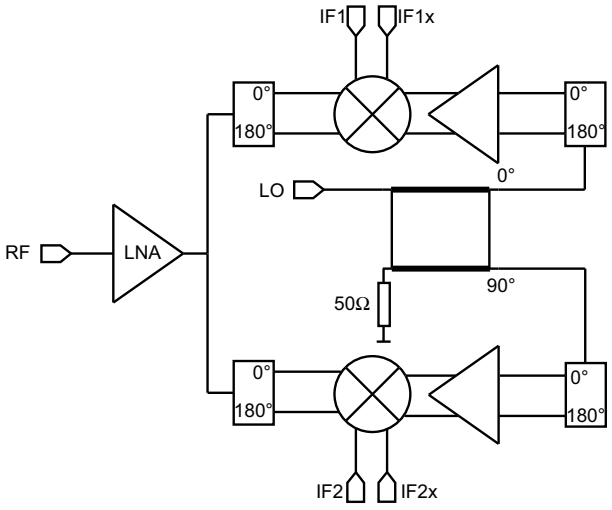

Fig. 1. Block diagram of the IQ frontend.

The design of the LNA is similar to the one presented in [6]. It consists of three common emitter (CE) stages, the first one is shown in Figure 2. Stages two and three basically have the same topology. The base–emitter biasing of Q_1 is done by a current mirror that is decoupled from the signal path by a shorted quarter–wavelength ($\lambda/4$) transformer (T_1). This transformer is replaced by $500\,\Omega$ resistors in the second and third stage. The collector–emitter voltage of Q_1 is held at a constant level of 1.4 V by a voltage divider (R_2, R_3), followed by a transistor in common–collector configuration to deliver the collector current of Q_1. The input of the LNA is matched with transmission lines for a return loss larger than 10 dB (VSWR $< 2:1$). The resulting noise figure is only 0.5 dB above the minimum noise figure of the circuit.

The simulated noise figure of the LNA is 6.5 dB in the frequency range from 75 GHz to 86 GHz. Interstage matching is accomplished by an LC matching network (T_2, C_3). Here, the inductance is replaced by a transmission line that is shorted for high frequencies. The transformed input impedance of the following stage determines the load impedance (the gain) of the CE transistor Q_1. The output of the LNA is matched with transmission lines to a $50\,\Omega$ interface, including the parasitic pad capacitance. The simulated gain of the LNA at 79 GHz is 16 dB. The LNA consumes 32mA.

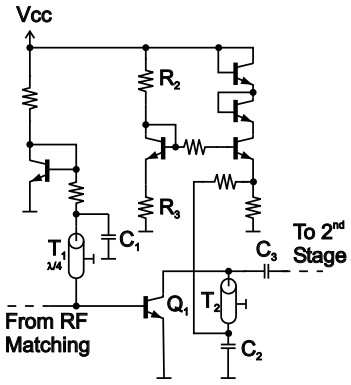

Fig. 2. One stage of the low-noise amplifier.

The power splitting of the RF signal is done as follows: Two transmission lines of 220 μm length transform the input impedance of both RF baluns to a low impedance. This value is transformed to the $50\,\Omega$ interface by a serial transmission line and a shunt capacitor.

The RF and the LO baluns are LC–baluns where the inductors are replaced with transmission lines [7]. With this type of balun, a single-ended to differential conversion and impedance matching is achieved simultaneously. The insertion loss of one balun is 1.6 dB.

The direct conversion mixers are based on the high level mixer presented in [8], but without any degeneration at the lower differential pair. The switching behavior of the four upper transistors in the Gilbert cell has a large influence on the noise figure of the mixer. Thus, the current density in the switching quad is chosen to result in maximum f_T, i.e., 5 mA/μm^2. The amplified thermal noise of the base resistances in the lower differential pair degrades the overall noise figure. Therefore transistors with bases consisting of two fingers and large emitter sizes are chosen. Current mirrors provide the bias current for the mixers. The bias networks at the LO port as well as at the RF port consist of level shifting diodes and resistors. $500\,\Omega$ resistors decouple the biasing from the signal path. The mixer is designed for high impedance external loads, so the internal load of $2 \times 400\,\Omega$ and the bias current of 6 mA determine the conversion gain. The simulated noise figure of one mixer is 12 dB, and the conversion gain is 24 dB. The current consumption of one mixer, including biasing, is 11 mA.

To maintain a high voltage swing at the LO port of the mixer, which is essential for good noise performance, the LO signal is applied to the four switching transistors via an LO buffer. This buffer is depicted in Figure 3. The load of the differential pair $Q_{1,2}$ is formed by an LC circuit in a parallel resonator configuration. The shorted transmission lines $T_{1,2}$ replace the inductors, and the capacitors are the transformed ($T_{3,4}$) input impedances of the emitter followers Q_{3-6} and the parasitic capacitances at the collectors of the differential pair $Q_{1,2}$. The simulated gain of the buffer as shown in Figure 3 is 10 dB in the frequency range around 80 GHz. The buffers consume 60 mA each.

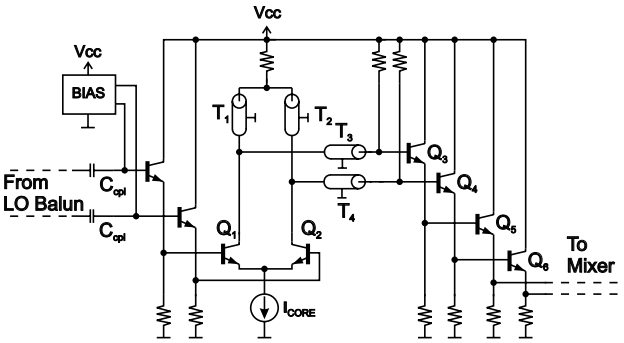

Fig. 3. LO buffer amplifier.

The 0°/90° phase shift between the two IF output ports is achieved by the use of a branchline coupler at the LO port of the mixer. The coupler is designed using an electrical model, consisting of the damping factor and the effective permittivity of the transmission lines at the operation frequency. The isolated port is terminated with $50\,\Omega$. Transmission lines match the input impedance of the coupler to $50\,\Omega$ parallel to the parasitic pad capacitance. Effects of corners and tee-junctions are neglected. The four branches have a length of 480μm each, which equals $\lambda/4$ at 80 GHz.

III. Technology

The frontend was implemented in an advanced 200 GHz f_T SiGe:C bipolar technology, based on the technology presented in [2]. The maximum oscillation frequency f_{max} of the transistors is 275 GHz. Shallow and deep trench isolation are used. The transistors are fabricated with a double–polysilicon self-aligned emitter base configuration with a SiGe:C base. This base is integrated by selective epitaxial growth. The transistors have a minimum emitter mask width of 0.35 μm, resulting in an effective emitter width of 0.18 μm. The technology additionally features high–voltage transistors with a BV_{CEO} of 5.0 V [9]. Four metal layers, MIM–capacitors, varactors, and different types of resistors are also included.

Transmission lines are essential components in millimeter wave circuits, making their accurate modeling mandatory. A 2D field simulator is used to model the transmission lines, as described in [10]. The transmission lines are implemented with metal layer M4 over metal layer M2. A $5\,\mu$m wide metal M4 signal path over a metal M2 ground path yields a $50\,\Omega$ transmission line. The maximum width of metal M2 is limited to $15\,\mu$m. Therefore, a cheesed structure is used to expand the ground plane. Simulations show that the transmission lines have a loss of 1 dB per millimeter and an effective dielectric

constant of 3.83 at 80 GHz . The $35\,\Omega$ transmission lines of the branchline coupler exhibit a loss of 0.86 dB per millimeter at the same frequency.

IV. Experimental Results

A die photograph of the fabricated I/Q frontend is shown in Figure 4. The left hand side of the chip is occupied by the I/Q mixer, while the LNA covers the right hand side.

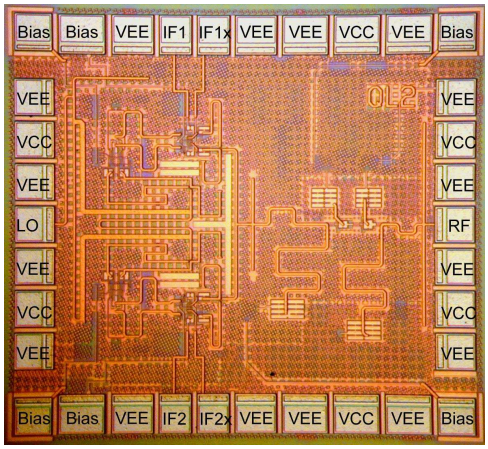

Fig. 4. Die photograph of the frontend. Die size is $1100\,\mu\text{m} \times 1000\,\mu\text{m}$.

All measurements were done on–wafer with probes. The temperature was kept at a constant level of 25°C. The LO power level was set to +1 dBm on-chip. The mixers are designed for high impedance external loads, thus external voltage followers with an input impedance of $100\,\text{k}\Omega$ were attached to the output of the mixers. This provides matching to the $50\,\Omega$ measurement environment. The differential IF signal was combined with a 180° low–frequency hybrid. The voltage followers limit the IF measurement range to an upper frequency of 30 MHz. All off–chip losses from the test setup were de-embedded from the measurement results.

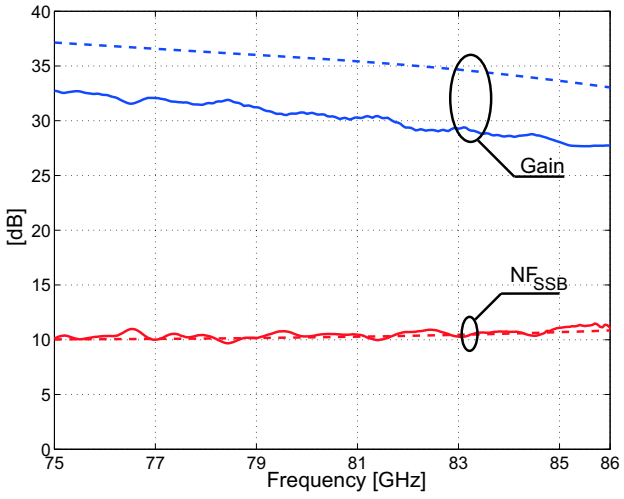

Fig. 5. Measured (solid lines) and simulated (dashed lines) singles sideband noise figure and conversion gain of the frontend (IF = 10 MHz).

The noise figure was measured using HP's 8970B noise figure meter, which limits the lower IF range to 10 MHz. The single sideband (SSB) noise figure and the conversion gain are depicted in Figure 5. This plot also shows a comparison with the simulated results. While the noise figure matches simulations (11 dB), the gain deviates by 5 dB over the whole frequency range. Additional measurements show that the gain of the LNA differs from the simulation by 4 dB.

The 90° phase shift of the in phase and the quadrature output is a crucial parameter for the performance of communication as well as precision sensor systems. This parameter was measured using a frequency counter with phase measurement capability (HP 53132A). The result is shown in Figure 6, along with the simulated phase difference. The measured results are in good agreement with the simulations.

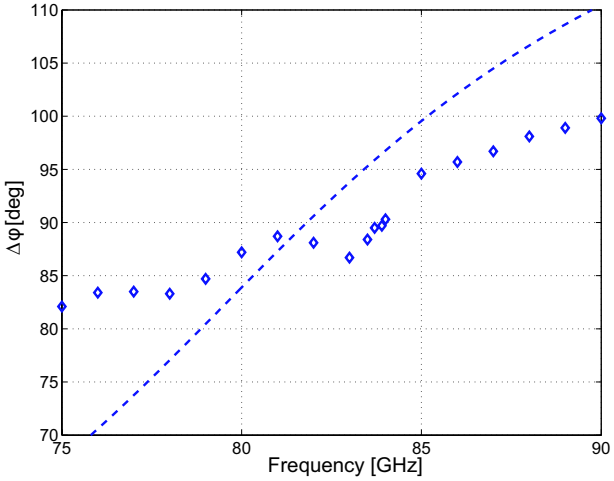

Fig. 6. Measured mean phase difference and simulated (dashed lines) phase difference of the IF outputs (IF = 10 MHz, 500 averages).

The relative phase difference from 90° is within ±8° from 75 GHz to 88 GHz. Figure 7 shows the conversion gain and noise figure imbalance of the two mixer outputs. The gain differs by less than 0.7 dB. The difference in noise figure is negligible.

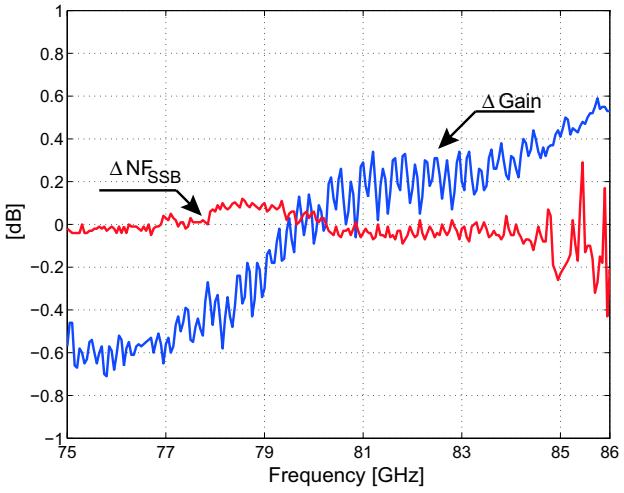

Fig. 7. Measured gain and noise figure imbalance of the two output ports (IF = 10 MHz).

The plot of the differential IF output voltage versus RF input power is shown in Figure 8. Due to the external high impedance load of the mixer, the output voltage is depicted instead of the output power. The input-referred 1 dB compression point is -16 dBm. The differential saturated output voltage is 13 dBV_{pp}, which is equal to a linear differential output voltage of 4.47 V_{pp}. The 1 dB compression point instead of the third-order intercept point is taken as a figure of merit for the usability of this frontend in data communication systems due to the lack of a third RF source for the two tone measurement setup.

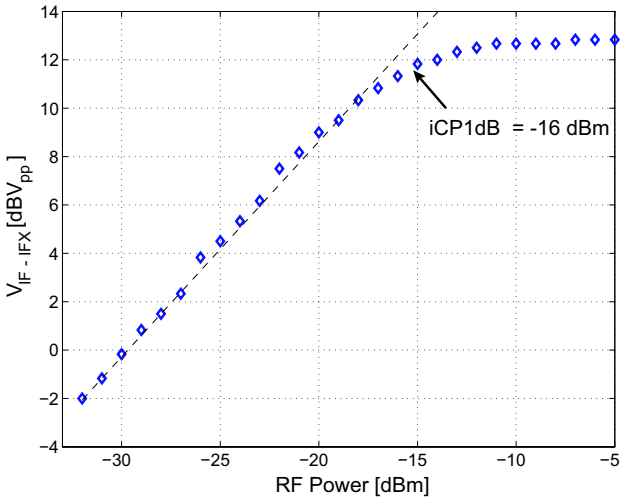

Fig. 8. Differential IF output voltage versus input power at 79 GHz (IF = 10 MHz).

The RF and LO port matching were measured using millimeter-wave probes and a 110 GHz network analyzer from Agilent. The results are shown in Figure 9. The RF port has a return loss larger than 10 dB in the frequency range from 75 GHz to 89 GHz. The return loss of the LO port is larger than 10 dB in the frequency range from 50 GHz to 90 GHz. These results show very good agreement with the simulations, as indicated in Figure 9. The measured LO to RF isolation is better than 35 dB from DC to 110 GHz.

V. Conclusion

An I/Q receiver frontend for sensing and communication purposes was designed, fabricated in a SiGe:C bipolar technology, and characterized. The frontend features a conversion gain of more than 28 dB and a noise figure of 11 dB over a frequency range from 75 GHz to 86 GHz. The two IF outputs are in quadrature at a frequency of 84 GHz, and they are within 8° out of quadrature from 75 GHz to 88 GHz. The gain imbalance between these outputs is smaller than 0.7 dB, and the imbalance in noise figure is smaller than 0.4 dB. The input-referred 1 dB compression point is -16 dBm.

The performance of this frontend indicates its suitability for communication systems in the frequency range from 81 GHz to 86 GHz, or for precision sensing applications around 79 GHz. From the technology point of view, the obtained

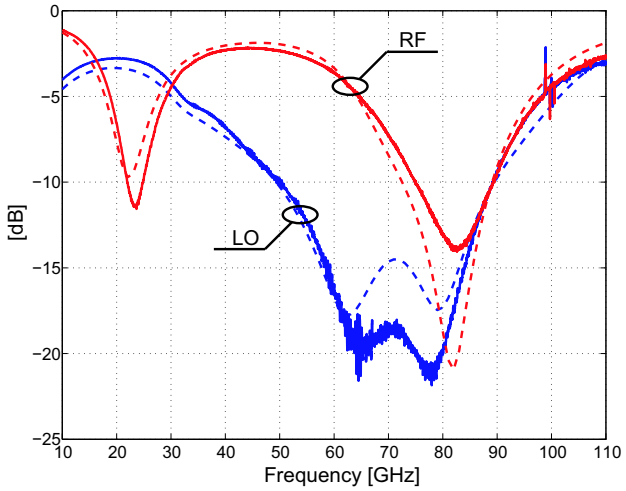

Fig. 9. Measured (solid lines) and simulated (dashed lines) LO and RF port matching of the frontend.

results show that very high integration levels in millimeter–wave frequency ranges are feasible in SiGe.

VI. Acknowledgement

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