A Low Noise, High Gain, Highly Linear Mixer for 77 GHz Automotive Radar Applications in SiGe:C Bipolar Technology

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***Abstract*—This paper presents a modified Gilbert type mixer which is fabricated in a 200 GHz f_T SiGe:C bipolar technology and well suited for 77 GHz bi-static automotive radar applications. The measured single sideband noise figure (*NF_{SSB}*), conversion gain (*CG*), and input-referred 1 dB compression point (*ICP*) of this mixer are 10.8 dB, 21.5 dB, and -5 dBm, respectively. The current consumption is 21 mA under 3.3 V power supply. This mixer shows state-of-the-art noise figure, conversion gain traded-off with 1 dB compression point, and low power consumption.**

I. INTRODUCTION

In the 77 GHz automotive radar application the mixer is the most critical block in the receiver. Though the specification for the linearity in a bi-static radar is smaller than that in a mono-static one, the input-referred 1 dB compression point is still up to around -10 dBm in the real implementation. Due to this high linearity requirement, it is nearly impossible to use the structure of a low noise and high gain LNA preceding a low noise and highly linear mixer for the receiver RF front-end in SiGe bipolar technology. For this case, the mixer has to be at the first stage of the whole receiver, so its noise figure should be as low as possible and its conversion gain as high as possible. To detect the two- or three-dimensional information of a target, there are normally more than one receiver channels needed. Therefore, the power consumption of the mixer should be kept to a low level. Although there are lots of mixers for 77 GHz automotive radar applications in published papers, there is still a big challenge to further explore the circuit art of this kind of mixer with lower noise figure, higher conversion gain, higher linearity and lower power consumption.

In this paper, a modified Gilbert type mixer in a 200 GHz SiGe:C bipolar technology is presented especially for 77 GHz bi-static automotive radar applications. Compared to the other recently published mixers [1, 2, 3, 4, 5, 6], this mixer shows state-of-the-art noise figure, combined with a balanced trade-off between high conversion gain and high 1 dB compression point, and very low power consumption.

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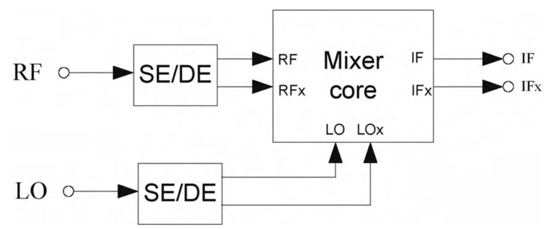

Figure 1. Block diagram of the presented mixer.

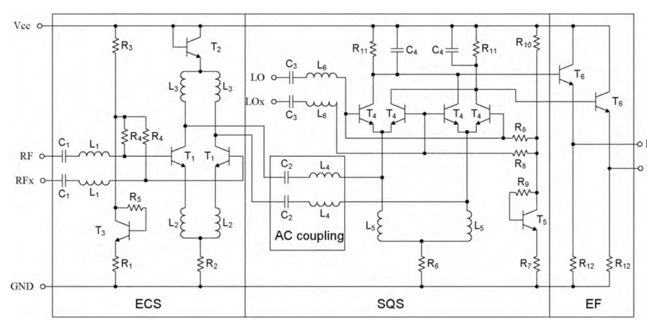

Figure 2. Modified Gilbert type mixer core.

II. THE PRESENTED MIXER

The mixer presented in this work mainly consists of a modified Gilbert type mixer core and on-chip single-ended to differential conversions (SE/DE) for RF and LO input ports, respectively (see Fig. 1). Single-ended RF and LO input ports are very convenient for measurement and application.

Fig. 2 is the circuit schematic of the modified Gilbert type mixer core which mainly includes an emitter-coupled trans-admittance stage (ECS), AC coupling, a switch quad stage (SQS), and the output buffer stage – an emitter follower (EF). Compared to the traditional Gilbert type mixer [2, 6], here AC coupling between ECS and SQS is used to have the following advantages: the base-band noise in ECS is decoupled to the IF output port; the DC current of SQS can be far smaller than that of ECS to lower its noise contribution; and the swing of IF or

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IFx is possible to be 2 V (the maximum V_{CE} is around twice of the collector–emitter breakdown voltage of the transistor at open base - BV_{CEO}) under 3.3 V power supply to increase the output compression point.

In Fig. 2, capacitors C_1 and C_3 are for the AC coupling purpose. In ECS, the current mirror with resistor degeneration (T_1, T_3, and R_1~R_5) is the bias circuit. For the low noise application, the DC current density is normally biased at small value to get less noise, but the total current is still high enough to meet high linearity requirement [7]. With such kind of bias circuit, the RF input signal is larger and the "transient" DC current as well, so that the linearity can also be further improved somewhat. With the method described in [8], the compromise between the input matching, the linearity, and the noise performance is made by choosing inductors L_1 and L_2. Diode connected T_2 is for the DC level shift in case that transistor T_1 works in the breakdown region.

In SQS, the DC bias circuit is similar to that in ECS. Moreover, the mixer's conversion gain and noise figure are more dependant on the LO input power than the transistor's DC current, which is the reason why the DC current of SQS can be far below that of ECS. The LO port matching is optimized with L_6. C_2, L_3~L_5, the capacitive output impedance of ECS and the capacitive input impedance of switch quad transistors (T_4) compose a matching network to peak mixer's conversion gain and filter out out-of-band noise. The load of the switch quad (R_{11} and C_4) is a low pass RC filter with 1.5 GHz corner frequency to filter out higher than IF frequency components (mixed or LO leakage). The emitter followers (EF) function as the output buffer to increase the driving ability and reduce the influence of the following base-band process stage on the mixer core's performance.

The single-ended to differential conversions (SE/DE) in Fig. 1 are realized with the circuit shown in Fig. 3 for both RF and LO input ports. Here an LC-Balun is the core circuit. When $R_{unbalanced}$ and $R_{balanced}$ are properly defined, L_3 and C_2 can be chosen. The L-type matching network incorporating the pad (Pad is modeled by C_{pad}) is used to convert $R_{unbalanced}$ to external 50 Ohm. At 77 GHz, the loss of this kind of single-ended to differential conversion is around 1 dB in this mixer.

III. FABRICATION

The presented mixer was implemented in an advanced 200 GHz f_T SiGe:C bipolar technology [9]. The maximum oscillation frequency f_{max} is 275 GHz. Shallow and deep trench isolation is 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 collector–emitter breakdown voltage of the transistors at open base (BV_{CEO}) is 1.5 V, which limits the collector's voltage swing. In the presented mixer, the maximum transient collector-emitter voltage is 3 V which benefits from the bias circuit described in II (the transistor's base impedance is smaller; its collector-emitter breakdown voltage is bigger).

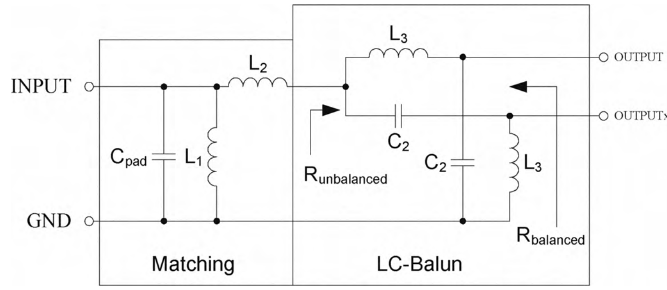

Figure 3. Single-ended to differential conversion in RF or LO input port.

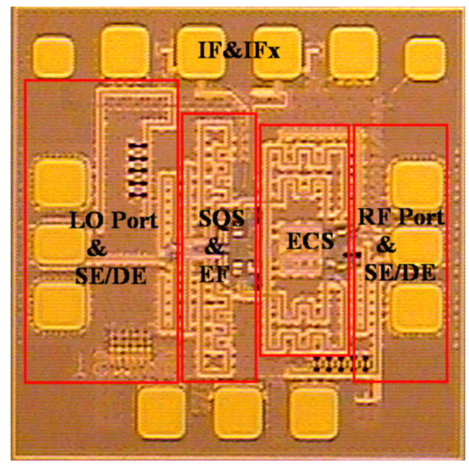

Figure 4. Chip photograph.

This technology provides four metal layers, MIM–capacitors, and different types of resistors. The inductors in the presented mixer are realized by micro-strip lines with top Metal layer M4 as signal path over bottom metal layer M2 as the ground plane. For such kind of configuration, the characteristic impedance of a 6 µm wide M4 line is around 50 Ω and its loss is around 1 dB/mm.

The size of the implemented mixer is 728×728 μm^2 and its photograph is depicted in Fig. 4. It can be seen from this picture that most of the space is occupied by the inductors realized with micro-strip lines. The left side is the mixer's LO port with ground-signal-ground structure (GSG); the right side is the mixer's RF port with the same PAD structure like the LO port; the mixer's differential IF output port is at the top of this chip with GSSG structure.

IV. MEASUREMENT RESULTS

The measurements were performed on wafer at room temperature. The current consumption is 21 mA under 3.3 V power supply and correspondingly the power consumption (P_{DC}) is around 70 mW.

The measurement result of the small signal RF port matching is given in Fig. 5. For comparison, the simulated curve is also attached here. From these curves, the return loss of the mixer's RF port is -16 dB at 77 GHz and the measurement result fits well to the simulated over the whole frequency range (from 10 GHz to 110 GHz), especially above 30 GHz.

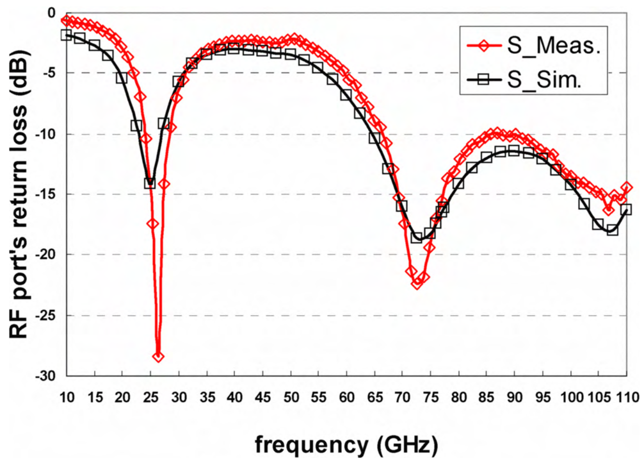

Figure 5. RF port matching.

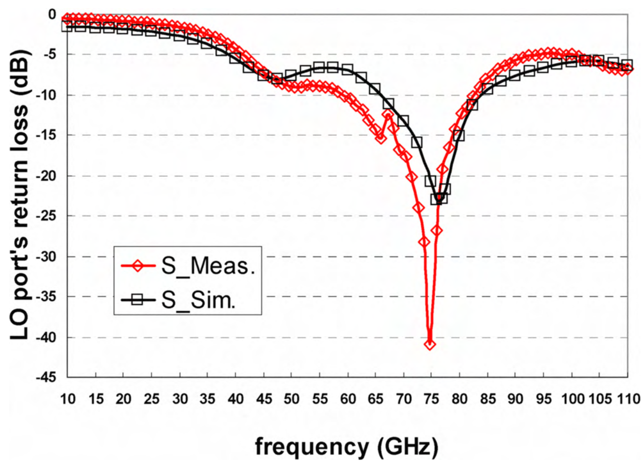

Figure 6. LO port matching.

Because the large signal behavior is more important for the LO port, the large signal return loss is measured by setting the LO port power of the instrument to -4 dBm (the maximum power that can be generated). Fig. 6 gives both the measurement and simulation results. It shows that both measured and simulated curves fit well to each other above 76 GHz and the LO port's return loss is -23 dB at 77 GHz. There is an abrupt change at 67 GHz in the measured curve, which is due to the fact that the power of the network analyzer changes at this point (S-parameter measurement above 67 GHz needs an additional mixer).

To get both the conversion gain and the single sideband noise figure of the presented mixer versus its LO port power (P_{LO}) and then find out the minimum LO port power under which the mixer can work, the measurement with the gain method is used. In this measurement, an additional IF base-band amplifier with 10 dB gain is needed to extend the minimum detected noise level of the spectrum analyzer. Fig. 7 shows the de-embedded measurement results when LO frequency (f_{LO}) is 76.5 GHz; IF frequency (f_{IF}) is 440 kHz; and RF port power (P_{RF}) is -20 dBm (which is only needed for the conversion gain measurement). The simulation results under the same conditions are also shown in Fig. 7. From these results, it can be seen that the minimum required LO port power is around -4 dBm in both measurement and simulation. At this point, the conversion gain and the single sideband noise figure in measurement are 21.8 dB and 10.8 dB, respectively; and both of them are better than the simulated by 1 dB.

The measurement results of the conversion gain and the single sideband noise figure versus IF frequency are given in Fig. 8 when LO frequency is 76.5 GHz; LO port power is -2 dBm; and RF port power is -20 dBm (which is needed only for the conversion gain measurement). In this picture, even though there are some ripples in the conversion gain curve and some spurs in the noise figure curve, the main trends of both curves show that the mixer's conversion gain is 21.5 dB in average; and the mixer's single sideband noise figure is 10.5 dB and worsens a little bit when IF frequency is close to 100 kHz, which shows that the flicker noise starts to be noticeable below 200 kHz.

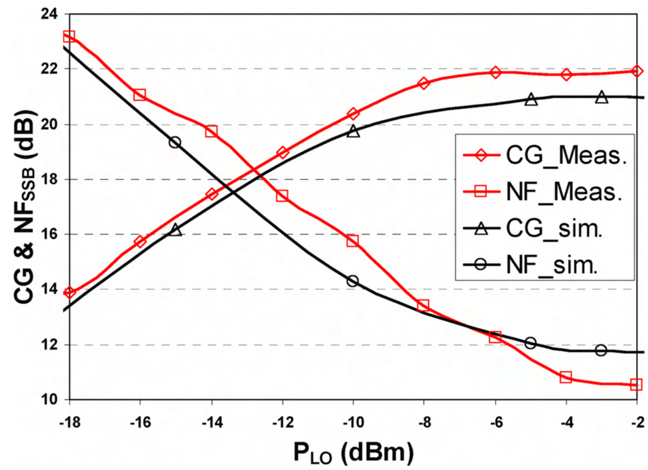

Figure 7. *CG* and NF_{SSB} versus P_{LO} when f_{LO}=76.5GHz, f_{IF}=440 kHz, and P_{RF}=-20 dBm (needed only for the conversion gain measurement).

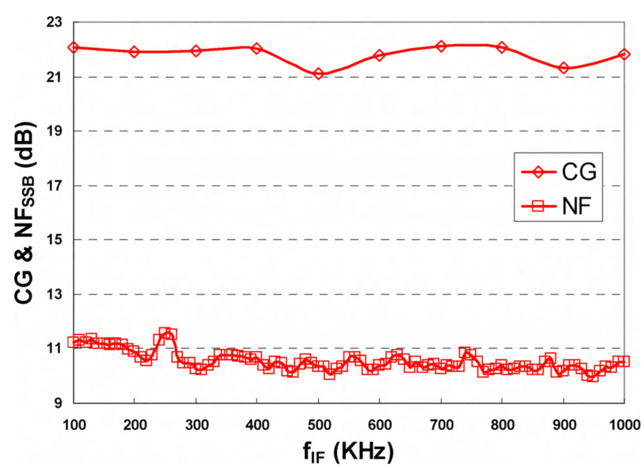

Figure 8. *CG* and NF_{SSB} versus f_{IF} when f_{LO}=76.5GHz, P_{LO}=-2 dBm, and P_{RF}=-20 dBm (needed only for the conversion gain measurement).

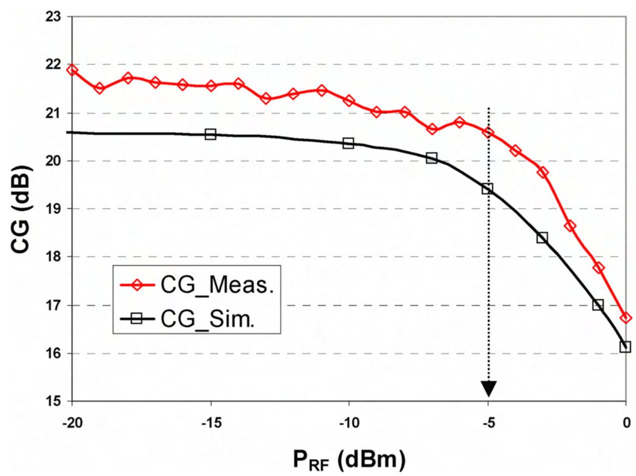

Figure 9. Input-referred 1 dB Compression point.

TABLE I. STATE-OF-THE-ART MIXERS FOR 77GHZ RADAR.

	NF_{SSB} [dB]	*CG* [dB]	*ICP* [dBm]	V_{CC} [V]	P_{DC} [mW]	*FoM*
[1]*	12.8	20	-14.7	3	360	0.032
[2]**	14	24	-30	-5	300	0.001
[3]	16.5	11	0	5.5	413	0.120
[4]**	18.4	13.4	-12	4.5	176	0.015
[5]	16	15.5	-3	5.5	187	0.250
[6]	11.2	15	2.5	5.5	335	1.489
This work	10.8	21.5	-5	3.3	70	2.960

* superheterodyne downconverter ** includes an IF buffer

For the compression measurement, in case of the compression effect of the external IF amplifier on the presented high gain mixer, the external IF amplifier is switched to low gain mode. With this solution, the reliable compression measurement is done and Fig. 9 gives the result when LO frequency is 76.5 GHz; LO port power is -2 dBm; and IF frequency is 100 kHz. The simulation result under the same conditions is also given in this figure for comparison. According to these curves, the input-referred 1dB compression points in both measurement and simulation are identical at -5 dBm; and the measured conversion gain is larger than the simulated by 1dB, which is consistent to the previous results which are shown in Fig. 7.

V. CONCLUSION

In this paper, a low noise, high gain, highly linear mixer with a modified Gilbert structure is presented. The measurement results fit well to the simulated ones. The power consumption of this mixer is around 70 mW. The measured conversion gain, single sideband noise figure, and input-referred 1 dB compression point of this mixer at 76.5 GHz are 21.5 dB, 10.8 dB, and -5 dBm, respectively.

A comparison with state-of-the-art mixers [1, 2, 3, 4, 5, 6] is given in Table I (based on the latest results and the table shown in [6]). The figures of merit (*FoM*) for these mixers are also calculated by (1), which is similar to the *FoM* definition for an LNA [10].

$$FoM = \frac{(ICP+9)\times G\times f}{(F-1)\times P_{DC}} \tag{1}$$

Here, both *ICP* and P_{DC} are in mW; *G* and *F* are the linear value of the conversion gain and the single sideband noise figure, respectively; and *f* is the interesting frequency in GHz and 76.5 for all mixers in table I. From this table, it can be seen that the presented mixer has the highest *FoM* and shows state-of-the-art noise figure, conversion gain traded-off with 1 dB compression point, and low power consumption. This mixer is well suited for 77 GHz bi-static mode automotive radar applications.

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