A 77 GHz SiGe Sub-Harmonic Balanced Mixer

Juo-Jung Hung, *Member, IEEE*, Timothy M. Hancock, *Member, IEEE*, and Gabriel M. Rebeiz, *Fellow, IEEE*

***Abstract*—A novel SiGe 77 GHz sub-harmonic balanced mixer is presented with a goal to push the technology to its limit [SiGe2-RF transistor ($f_T = 80$ GHz)]. This new topology uses a compact input network not only to achieve high isolation between the LO and RF ports, but also to result in excellent 2LO-RF isolation. The measured results demonstrate a conversion gain of 0.7 dB at 77 GHz with an LO power of 10 dBm at 38 GHz, LO-RF isolation better than 30 dB, 2LO-RF isolation of 25 dB, and a $P_{1\text{dB}}$ of -8 dBm. The mixer core consumes 4.4 mA at 5 V. The circuit demonstrates that SiGe sub-harmonic mixers have comparable performance with GaAs designs, at a fraction of the cost.**

***Index Terms*—Coplanar waveguides (CPWs), millimeter-wave mixer, monolithic microwave integrated circuits (MMICs), SiGe, sub-harmonic balanced mixer.**

I. Introduction

SUB-HARMONIC mixers (SHMs) offer an alternative solution to fundamental mixers in the millimeter-wave (mm-wave) range since this allows the use of a local oscillator at a relatively low frequency with better phase noise performance and higher output power. For single-ended mixers, the LO AM noise is downconverted to the IF port and as is well known, this AM noise can be rejected by using a balanced mixer.

The most widely used SHM topology is the anti-parallel diode pair (APDP) as implemented in [1]–[3]. Typically, the cut-off frequency of the diode needs to be 3–5 times higher than the operating frequency to result in low conversion loss. However, the Schottky diode available in our process (Atmel SiGe2-RF) has an $f_c \approx 50$ GHz, and is not suitable for 77 GHz SHM applications.

Another commonly used technique is the FET resistive mixer [1], [4]. The transistors are biased in the triode region and the differential LO signals are applied to the gates to modulate the FET linear channel resistance. The RF signal is applied to the drain and the frequency mixing occurs due to the time-variable channel conductance.

Another sub-harmonic mixer topology is presented in [5], [6]. It uses a “multi-level” LO switching core in place of the LO switching quad in the standard Gilbert cell mixer. This design was implemented at 1–2 GHz where a precise quadrature LO signal can be easily obtained using lumped *RC* networks. The transistors in the [5], [6] topologies do not operate well as switches at mm-wave frequencies, and this leads to a poor conversion loss.

Because of the process limitation, the current topologies are not suitable for a 77 GHz SHM design. Therefore, a novel topology using SiGe technology is presented and shows comparable performance to GaAs designs.

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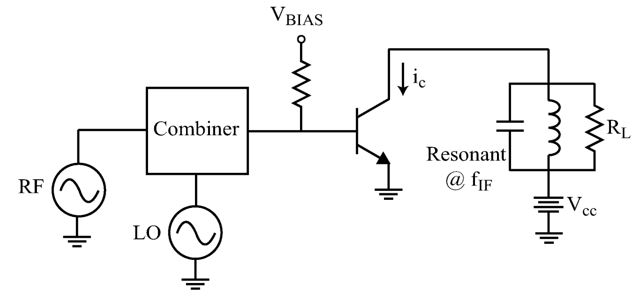

Fig. 1. Simplified schematic of a single-ended transconductance mixer.

II. Design Theory

The mixing principle used in the SHM design is called transconductance mixing in which the time-varying transconductance is the dominant contributor to frequency conversion [1], and a simplified single-ended schematic is shown in Fig. 1, [7]. In this topology, both RF and LO signals are applied to the base together through a combiner which is usually a filter or a directional coupler. The resonator at the output port works as bandpass filter and rejects all frequencies except IF. The LO is a large signal and its voltage across the base and emitter junctions creates a time-varying transconductance $G_m(t)$. In general $G_m(t)$ is periodic but not a sinusoidal signal and can be represented as a Fourier series

$$G_m(t) = G_{m0} + G_{m1}\cos(\omega_{\text{LO}}t) + G_{m2}\cos(2\omega_{\text{LO}}t) + G_{m3}\cos(3\omega_{\text{LO}}t) + \cdots \quad (1)$$

The collector current i_c is the product of the $G_m(t)$ and input voltage, $V_{\text{BIAS}} + V_{\text{RF}}\cos(\omega_{\text{RF}}t)$

$$\begin{aligned} i_c &= G_m(t)[V_{\text{BIAS}} + V_{\text{RF}}\cos(\omega_{\text{RF}}t)] \\ &= V_{\text{BIAS}}[G_{m0} + G_{m1}\cos(\omega_{\text{LO}}t) + G_{m2}\cos(2\omega_{\text{LO}}t) + \cdots] \\ &\quad + G_{m0}V_{\text{RF}}\cos(\omega_{\text{RF}}t) \\ &\quad + \frac{G_{m1}V_{\text{RF}}}{2}\{\cos[(\omega_{\text{RF}} - \omega_{\text{LO}})t] + \cos[(\omega_{\text{RF}} + \omega_{\text{LO}})t]\} \\ &\quad + \frac{G_{m2}V_{\text{RF}}}{2}\{\cos[(\omega_{\text{RF}} - 2\omega_{\text{LO}})t] + \cos[(\omega_{\text{RF}} + 2\omega_{\text{LO}})t]\} \\ &\quad + \cdots \end{aligned} \quad (2)$$

The output current consists of many mixing products. For example, $(1/2)G_{m1}V_{\text{RF}}\cos[(\omega_{\text{RF}} - \omega_{\text{LO}})t]$ is the fundamental

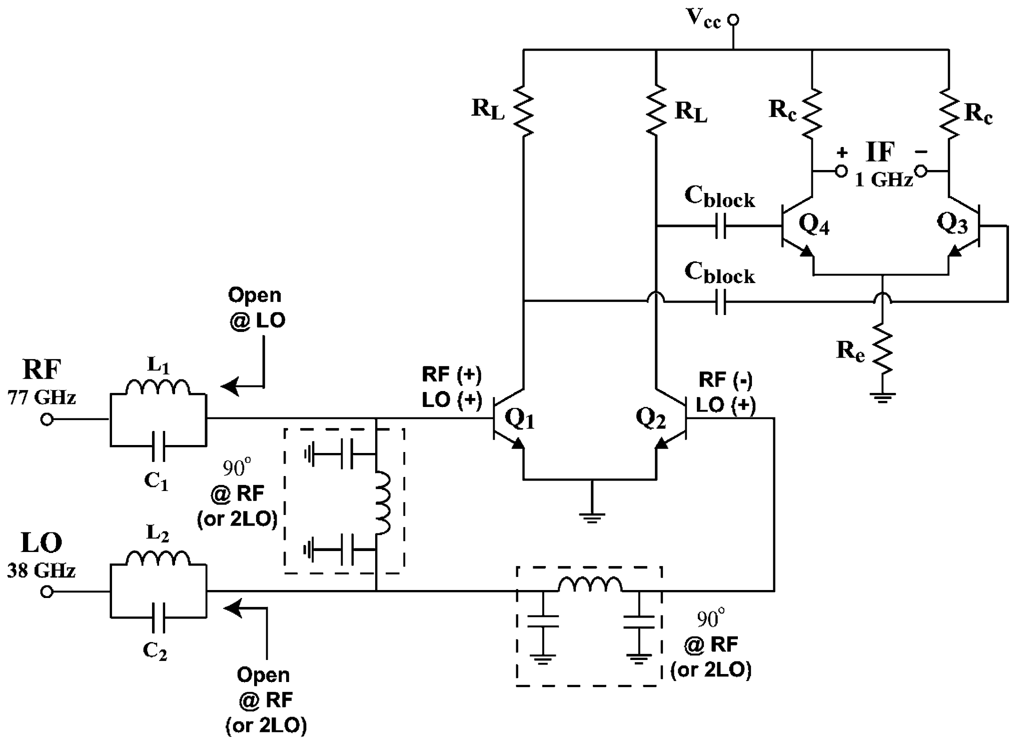

Fig. 2. Simplified schematic of the lumped-element sub-harmonic balanced mixer.

mixing term and some of the transconductance mixers utilize this product as frequency translation [8]. In this paper, it is the $(1/2)G_{m2}V_{\mathrm{RF}}\cos[(\omega_{\mathrm{RF}}-2\omega_{\mathrm{LO}})t]$ term that generates the desired subharmonic mixing product and G_{m2} should be maximized for higher conversion gain. It is well know that the second harmonic signal is a function of conduction duty cycle and its optimum value can be obtained when the transistor is biased close to its turn-on point [7].

TABLE I
VALUES OF THE LUMPED COMPONENTS IN THE 77 GHz BALANCED SUB-HARMONIC MIXER

L_1 (pH)	120	R_L (Ω)	800
C_1 (fF)	145	C_{block} (pF)	2
L_2 (pH)	80	R_e (Ω)	400
C_2 (fF)	54	R_c (Ω)	100

III. CIRCUIT DESIGN

The schematic of the 77 GHz sub-harmonic balanced mixer is shown in Fig. 2. Q1 and Q2 are the mixer core and they are biased close to the turn-on voltage in order to maximize the nonlinearity characteristic of the transistors. Two parallel *LC* resonant tanks are used to achieve isolation between the LO and RF ports. L_1 and C_1 present an open circuit at 38 GHz to block the LO signal, while L_2 and C_2 resonate at 77 GHz to prevent RF leakage. The LC π networks are used to provide 90° phase delay at 77 GHz and the whole input network makes the RF signals become out of phase while the LO signals are kept in phase at the input of the transistors. At the output of the transistor, the LO and 2LO signals appear to be common mode and will not be seen at the IF port. Although the unwanted $f_{\mathrm{RF}}-f_{\mathrm{LO}}$ (39 GHz) and desired $f_{\mathrm{RF}}-2f_{\mathrm{LO}}$ (1 GHz) are both differential signals at the output ports, the frequency of $f_{\mathrm{RF}}-f_{\mathrm{LO}}$ is a lot higher than the sub-harmonic mixing term and can be easily filtered out. This balanced topology can be easily integrated with the baseband circuits and it can also reject the AM noise on the LO signal.

After the mixer stage, Q3 and Q4 are used as the output buffer stage to provide the impedance matching to the 50 Ω. Since the IF frequency is lower than 1 GHz, a resistive load R_L is used instead of a choke for small area. A capacitor C_{block} is placed between mixer and buffer stage for DC blocking and the load resistor R_c in the buffer stage is used for output matching. R_e is for DC biasing and it increases the common mode rejection ratio (CMRR) of the output buffer. The values of the lumped components are summarized in Table I.

Except for the isolation between LO and RF ports, the 2LO-RF isolation is also very important in the sub-harmonic mixer design. Since the 2LO and RF frequencies are very close, filtering the unwanted 2LO signal is very difficult if it leaks to the RF port. This leakage signal would cause interference problem with other receivers. The single-ended design was simulated to verify this problem and the simulation result indicates that when RF power is -10 dBm and LO power is 10 dBm, the 2LO signal which appears at the RF port is -4 dBm which is even higher than the RF power.

One of the advantages of this balanced design is that it provides good 2LO-RF isolation compared to other single-ended designs ([9, Fig. 1]). When the LO signals are pumped into the transistors in phase, 2LO signals will be generated at the collectors in phase and will leak through C_μ back to the input of the transistors. As seen in Fig. 3, one of the signals will flow through the 180° delay networks and become out of phase with

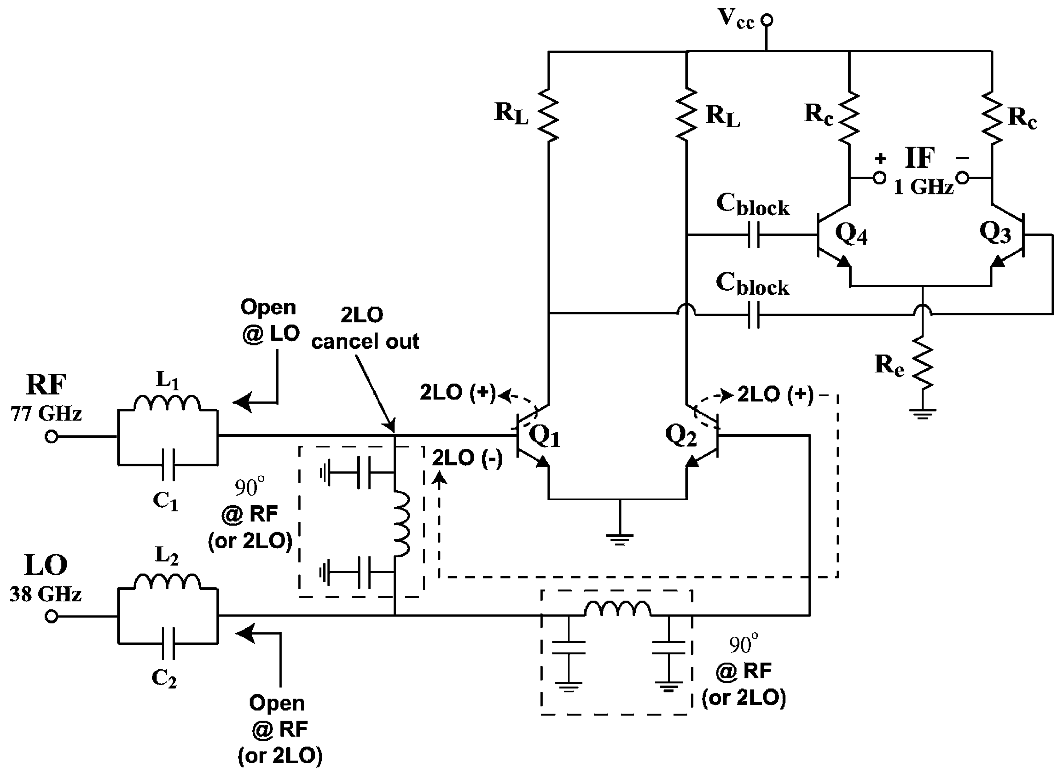

Fig. 3. The 2LO circuit with cancellation at the RF port.

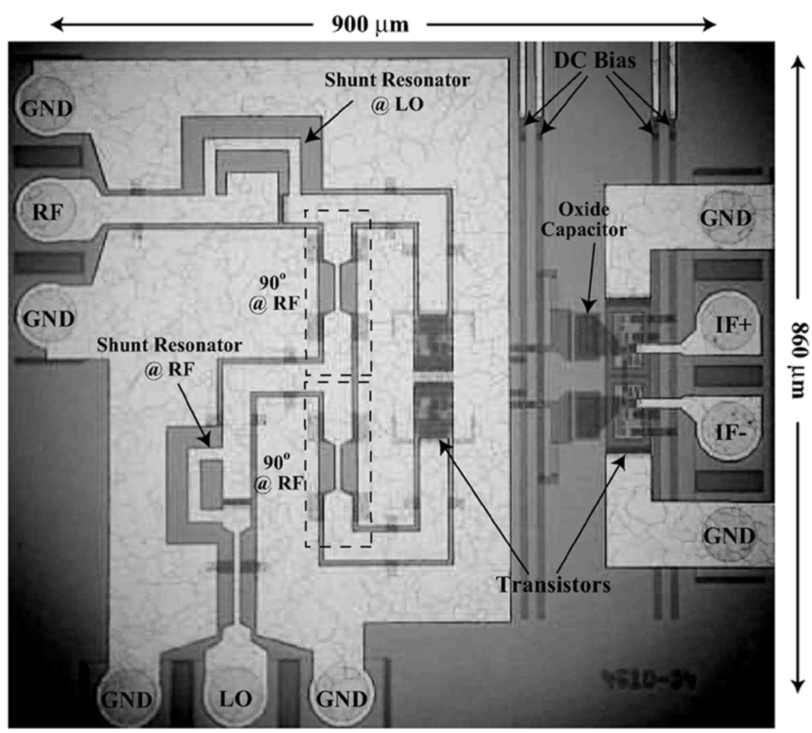

Fig. 4. Microphotograph of the 77 GHz sub-harmonic balanced mixer.

the other one. Therefore, the 2LO signals will be cancelled with each other and this enhances the 2LO-RF isolation.

The microphotograph of the sub-harmonic mixer is shown in Fig. 4 and the area without pads is 900×860 μm^2. The CPW line is built using metal-3 and the equalization bridge is made of metal-2. C_1 and C_2 are made using oxide capacitor between metal-2 and metal-3 (0.03 fF/μm^2). L_1 and L_2 are implemented using a narrow metal line. The phase-delay π networks are built using quasi-lumped LC components. The L and C values are realized using high-impedance and low-impedance t-lines, and their values for a 90° phase delay are calculated to be 100 pH and 40 fF, respectively. These values are used as a starting point

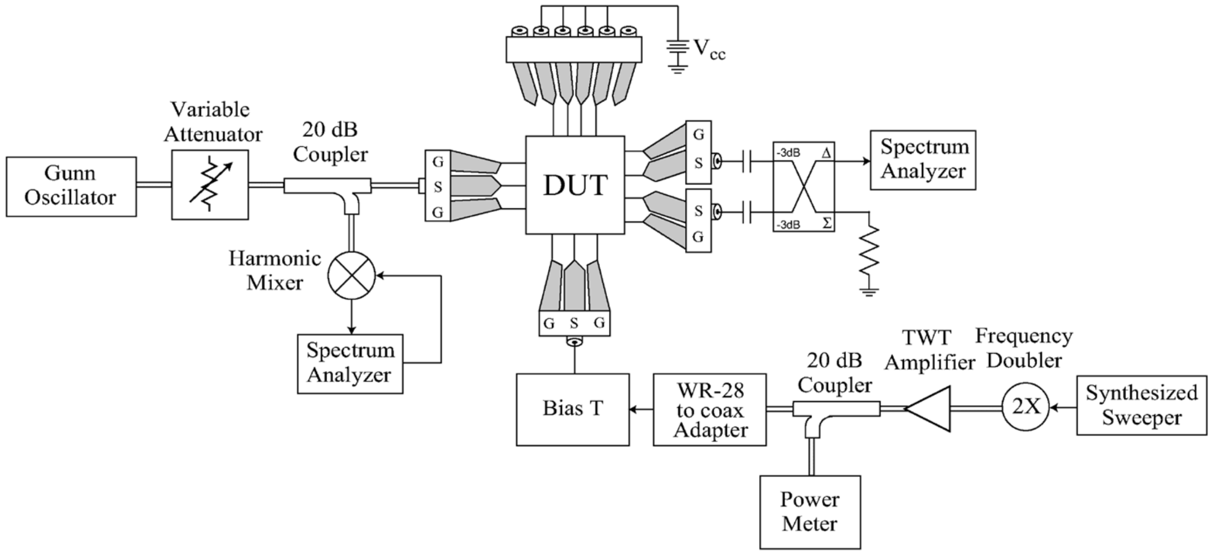

Fig. 5. Block diagram of the 77 GHz sub-harmonic mixer setup.

for simulation, and optimization was done by using a full-wave electromagnetic simulator (Sonnet[1]) to take the effect of the tapered transition between the high and low impedance t-lines. All the passive components were simulated using Sonnet and their S-parameters were exported to ADS[2] for the harmonic balance simulation with the transistor model. Simulation indicates that the Q of the unloaded resonant tank are 30 and 23 at 38 GHz and 77 GHz, respectively.

IV. Measured Results

The test setup is shown in Fig. 5. A mechanically tuned Gunn oscillator is used as the 77 GHz RF source followed by a variable attenuator and a 20 dB directional coupler. A W-band harmonic mixer (HP 11 970W, 75–100 GHz) and a spectrum analyzer (HP 8562A, 9 KHz-26.5 GHz) are connected to the coupled port of the coupler to monitor the RF power and frequency. The through port of the coupler is connected to the 150 μm pitch W-band probe (Picoprobe Model 120, 75–110 GHz) to the DUT.

The 38 GHz LO signal is generated by the synthesized sweeper (HP 83 624A, 2–20 GHz) followed by a frequency doubler (Pacific Ka2, 26.5–40 GHz) and a traveling wave tube (TWT) amplifier (Hughes 800IH). Since the gain of the frequency doubler and TWT amplifier varies a lot with input power and frequency, a power sensor (HP R4886A) followed by an power meter (HP 437B) is connected to the coupled port of the 20 dB directional coupler to monitor the power levels of the through port. Using a WR-28 to coax adapter, the through port of the coupler is connected to the bias-Tee (HP 11 612B, 45 MHz-50 GHz) to a 150 μm pitch Picoprobe (Model 40A, DC-40 GHz). The DC bias are applied to the chip using a MCW-14 multi-contact wedge, and the output IF signal is measured using a differential Picoprobe through the 180° hybrid to the spectrum analyzer (HP 8564E, 9 KHz-40 GHz). The loss of probes were measured separately and used to correct the measured results. To calibrate the effect of the coupler,

[1]Sonnet, ver. 9.52, Sonnet Software Inc., Syracuse, NY 1986–2003 USA.

[2]ADS 2003A, Agilent Technology Inc., Palo Alto, CA 1983–2003 USA.

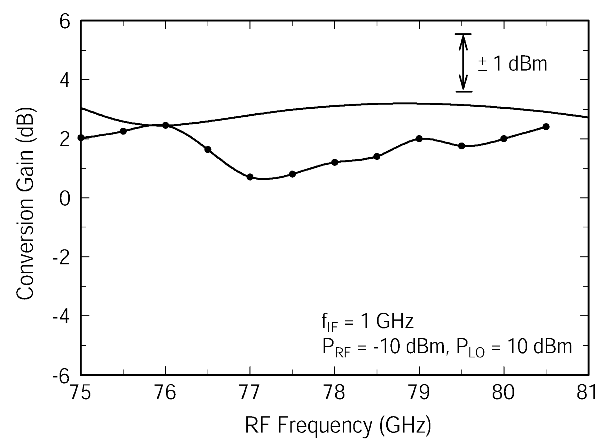

Fig. 6. Measured (dotted line) and simulated (solid line) conversion gain of the 77 GHz sub-harmonic balanced mixer for $f_{\mathrm{IF}} = 1$ GHz.

coax adapter and bias-Tee on the LO signal path, the output of the bias-Tee is disconnected from the probe and measured by another power meter to detect the accurate power pumped into the probe. The offset between the power measured at the output of bias-Tee and at the coupled port of the coupler is used to calibrate the measured results. The same calibration procedure was done to the RF signal path and we believe that the measured results are accurate to within ± 1 dB, which comes from the sensitive waveguide connection and frequency response of the harmonic mixer.

The measured and simulation results are shown in Figs. 6–12. Fig. 6 shows the conversion gain of the mixer when IF is fixed at 1 GHz, and indicates that the conversion gain is 0.7 dB when RF frequency is 77 GHz. The mixer is broad-band with a conversion gain of 0.7–2.5 dB from 75–81 GHz. The simulated gain of the buffer stage is 11 dB, so the conversion gain of the mixer core alone is -10.3 to -8.5 dB at this frequency range. Fig. 7. presents the IF frequency response. The conversion gain increases when the IF frequency decreases due to the frequency response of the output buffer. The gain drops when the IF frequency is below 100 MHz due to the DC blocking cap between

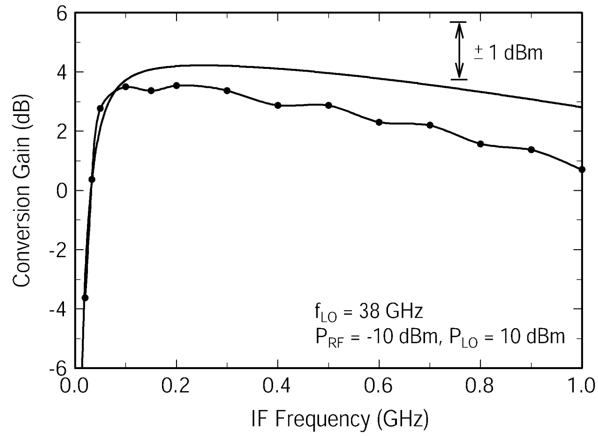

Fig. 7. Measured (dotted line) and simulated (solid line) conversion gain of the 77 GHz sub-harmonic balanced mixer for f_{IF} from 20 MHz to 1 GHz.

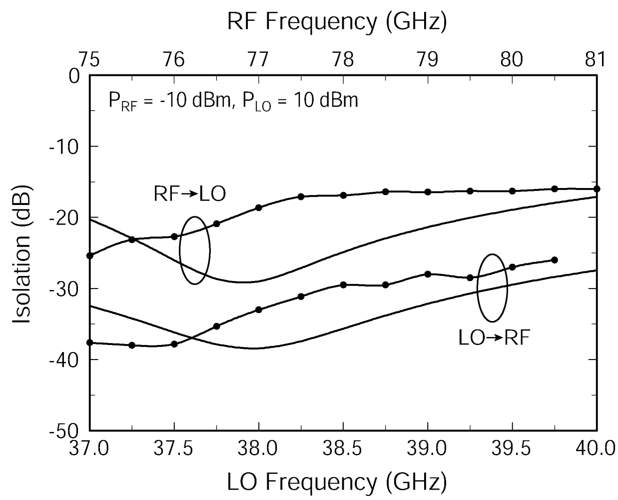

Fig. 8. Measured (dotted line) and simulated (solid line) LO to RF, and RF to LO isolation.

mixer and buffer stage. In Figs. 6 and 7, the difference between the simulated and measured conversion gain is less than 2 dB, and this could be due to the inaccuracy of the transistor model close to the f_T, or to inaccuracies in RF power measurements at 77 GHz.

Fig. 8 shows measured and simulated port-to-port LO-RF isolation. It was measured with the LO power delivered to the mixer as shown in Fig. 5. The RF port is probed using another DC-40 GHz Picoprobe to the spectrum analyzer (HP 8564E), and the IF port is terminated to a matched load. The measured LO-RF isolation is better than 30 dB for the intended operation frequencies (37–38.5 GHz for the automotive radar system), and is close to the simulated result except that the resonant frequency shifts down by 0.5 GHz. The RF-LO isolation is measured using the Gunn oscillator as a source and the harmonic mixer with a spectrum analyzer as a detector through two W-band probes to the RF and LO ports. The measured result is also presented in Fig. 8, and is below -19 dB for the RF frequency from 76–77 GHz. The resonant frequency shifts down from the designed 77 GHz to 75 or 74 GHz which is equivalent to about 2.6–3.9% of deviation. Both LO-RF and RF-LO isolation can be improved by re-tuning the resonant *LC* tank for the correct operating frequencies.

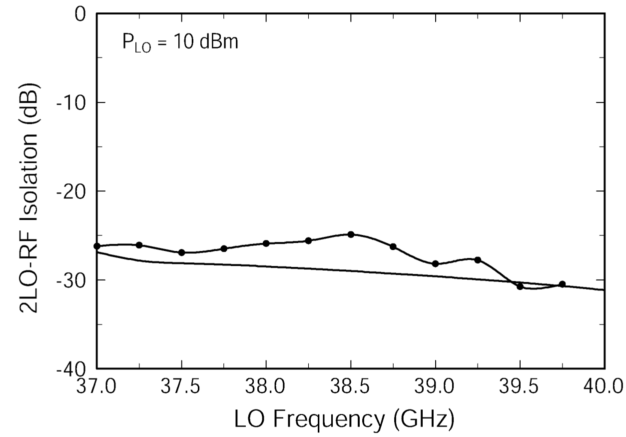

Fig. 9. Measured (dotted line) and simulated (solid line) 2LO to RF isolation.

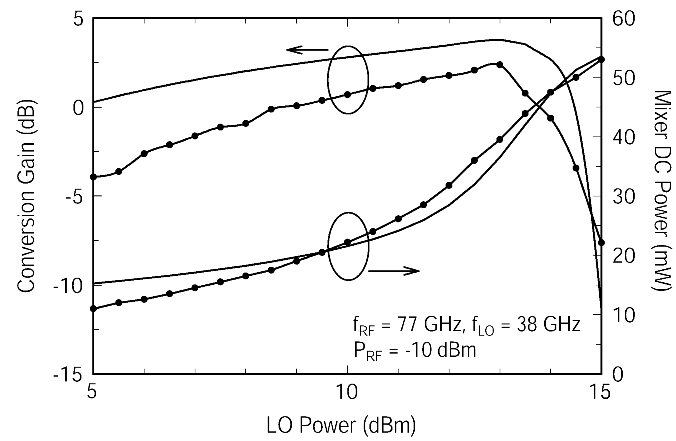

Fig. 10. Measured (dotted line) and simulated (solid line) conversion gain and DC power consumption of the mixer core versus LO power.

In Fig. 9, the measured 2LO-RF isolation also agrees well with the simulation and is better than 25 dB for the LO signal from 37 to 38.5 GHz. Since the RF resonant tank does not present an excellent open-circuit at the middle point of the 180° delay network, its loading effect causes the imbalance of the RF delay path and therefore the cancelation does not work perfectly well. Still, 25–30 dB is a very good isolation from a lumped network at W-band frequencies.

Fig. 10 shows the conversion gain and DC power consumption of mixer core versus LO power. When the LO power is 10 dBm, a conversion gain of 0.7 dB is obtained with 22 mW of DC power consumption. Since the mixer stage is biased close to the class B region, the DC current increases with the LO power. As the LO power is larger than 13 dBm, the mixer starts to saturate and the conversion gain drops sharply. The buffer stage consumes 40 mW DC power and contributes 11 dB of conversion gain. In the future, the IF amp would consume less current since the impedance of the following stage is several kΩ instead of 50 Ω termination. The measured IF output power versus RF power is shown in Fig. 11. The measured 1 dB compression point is -8 dBm which is 3 dB lower than the simulation result.

Two-tone intermodulation measurement was not done due to the lack of a second W-band source. The simulation result is

TABLE II
V- AND W-BAND SUB-HARMONIC MIXER PERFORMANCES

Device / f_T or f_c (GHz)	$f_{\rm RF}$ (GHz)	LO harm.	C.G. (dB)	LO-RF Iso. (dB)	P_{1dB} (dBm)	$\mathrm{P}_{\rm LO}$ (dBm)	Chip size without pads (mm^2)	Features	Ref
SiGe HBT / 80	*77*	2^{nd}	*+0.7 (-10.3)*	$>$ *30*	*-8 (+2.4)*	*10*	*0.9 x 0.86 (0.8)*	balanced, w/ IF buffer (Simu. w/o IF buffer)	*This Work*
GaAs Schottky diode / 670	60	4^{th}	-11.3	$>$ 33	-	7	3 x 2.3 (7)	balanced*, w/o IF buffer	[3]
GaAs PHEMT	60	4^{th}	+3.4	$>$ 40	-9	13	1.8 x 1.7 (3.1)	single, w/ IF buffer	[8]
GaAs Schottky diode / 180 + PHEMT	60	4^{th}	+0.8	$>$ 40	-11	12	1.9 x 2.6 (5)	balanced*, w/ IF buffer	[12]
GaAs PHEMT	77	2^{nd}	-17	-	0	5	1.3 x 0.84 (1.1)	balanced, w/o IF buffer	[4]
GaAs PHEMT / $>$ 85	89	2^{nd}	-4.7	$>$ 15	-	11	0.7 x 1.8 (1.3)	single, w/o IF buffer	[9]
GaAs Schottky diode / 1500	94	2^{nd}	-7	$>$ 40	-	8.5	0.8 x 1.5 (1.2)	balanced*, w/o IF buffer	[2]
GaAs Schottky diode	94	4^{th}	-11.4	-	-9	10	1.2 x 0.8 (1)	balanced*, w/o IF buffer	[13]

(*Back-to-back Schottky diode design. Results in AM noise rejection and some balanced mixer properties.)

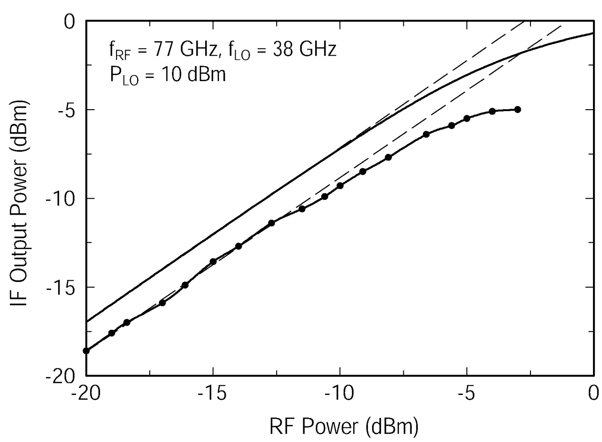

Fig. 11. Measured (dotted line) and simulated (solid line) IF output power versus RF power.

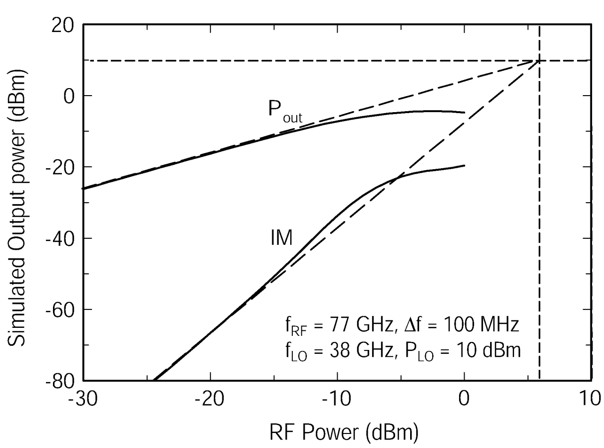

Fig. 12. Simulated fundamental and third harmonic of the IF output power versus RF power.

presented in Fig. 12 which shows that the third-order input intercept point (IIP3) is +6 dBm. The LO AM noise can also be suppressed by this balanced topology, and the simulated LO AM noise rejection is 28 dB. The simulated noise figure (NF) of the SHM mixer is 23 dB which is good enough for the automotive radar system as long as an LNA with moderate noise figure is preceded by the mixer. For the LNA in [10] with a gain of 15 dB and a noise figure of 5.6 dB, the overall front-end NF (LNA + mixer) will become 10 dB. This is much lower than the required NF of 16 dB [11].

V. SUMMARY

A comparison between different V- and W-band sub-harmonic mixers is presented in Table II. Our work is the only one using SiGe technology and the designed frequency is very close to the f_T of the device. In order to make a good comparison, the simulation result of this design without using IF buffer is also presented. Some of the SHM designs use the 2nd harmonic and others use the 4th harmonic of the LO to mix with the RF signal. Typically, a 3-dB conversion degradation is observed when going from 2X to 4X sub-harmonic mixing [3]. Overall, our SiGe mixer achieves comparable conversion gain using similar LO input powers compared to other GaAs designs. It also has a good linearity performance and occupies a small area.

VI. CONCLUSION

A novel SiGe 77 GHz sub-harmonic balanced mixer is presented to push the technology to its limit ($f_T = 80$ GHz). The input network not only achieves the isolation between LO and RF ports, but also results in excellent 2LO-RF isolation. The measured results demonstrate that this SiGe sub-harmonic mixer has comparable performance with the other GaAs designs.

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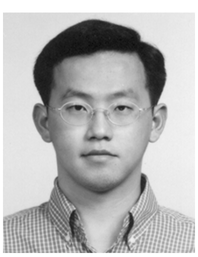

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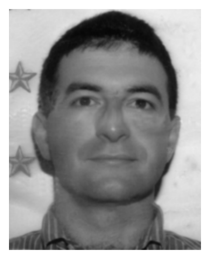

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