A High-Performance *W*-Band Uniplanar Subharmonic Mixer

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***Abstract*— A uniplanar subharmonic mixer has been implemented in coplanar waveguide (CPW) technology. The circuit is designed to operate at RF frequencies of 92–96 GHz, IF frequencies of 2–4 GHz, and LO frequencies of 45–46 GHz. Total circuit size excluding probe pads and transitions is less than 0.8 mm × 1.5 mm. The measured minimum single-sideband (SSB) conversion loss is 7.0 dB at an RF of 94 GHz, and represents state-of-the-art performance for a planar W-band subharmonic mixer. The mixer is broad-band with a SSB conversion loss of less than 10 dB over the 83–97-GHz measurement band. The measured LO–RF isolation is better than -40 dB for LO frequencies of 45–46 GHz. The double-sideband (DSB) noise temperature measured using the Y-factor method is 725 K at an LO frequency of 45.5 GHz and an IF frequency of 1.4 GHz. The measured data agrees well with the predicted performance using harmonic-balance analysis (HBA). Potential applications are millimeter-wave receivers for smart munition seekers and automotive-collision-avoidance radars.**

***Index Terms*— Coplanar waveguide, millimeter-wave mixer, MMIC's, subharmonic mixers.**

I. Introduction

SUBHARMONIC mixers (SHM's) downconvert an RF signal which is approximately the nth harmonic of the LO frequency to an IF. For mixing at even harmonics of the LO frequency, SHM's typically use a nonlinear device with an antisymmetric current–voltage characteristic, such as an antiparallel Schottky barrier diode pair or a planar doped barrier (PDB) diode. These mixers have been used primarily at millimeter- and submillimeter-wave frequencies in both waveguide [1] and quasi-optical [2] configurations, and have demonstrated noise and conversion loss performance competitive with fundamental mixers. There is considerable interest in the development of planar circuit alternatives to waveguide components at millimeter-wave frequencies. Microstrip SHM's have been realized at W-band with both antiparallel Schottky barrier diodes [3] and heterostructure field-effect transistors (HFET's) [4], and minimum conversion losses on the order of 10–11 dB have been achieved.

The primary motivation for using subharmonic mixing at millimeter/submillimeter-wave frequencies has been the lack of sufficient LO power at the fundamental frequency. However, advances in solid-state oscillator and multiplier technology have made fundamental mixing a realistic option up through W-band. Still, subharmonic mixing remains attractive due to the ease of LO distribution and the inherent RF/LO isolation. In receiver applications, this isolation is important in order to minimize the radiation of LO power from the RF port.

The purpose of this paper is to implement a 94-GHz subharmonic mixer in uniplanar coplanar waveguide (CPW) technology. The advantages of the CPW over a microstrip are well known [5]. The CPW is a uniplanar structure allowing both shunt and series connection of circuit elements, and simple ground-plane connections without the need for backside metallization or the associated via-holes. Quasi-static operation is possible up to very high frequencies (low dispersion), and the CPW line exhibits lower radiation losses in the odd mode. There has been considerable recent work in the development of millimeter-wave CPW fundamental mixer circuits, including a 64–77-GHz mixer for European automotive radar applications [6], and FET V- to C-band and C- to V-band frequency converters [7], [8].

II. Coplanar Waveguide Design Issues

A test wafer was fabricated with individual circuit elements such as short- and open-circuited stubs, and through-line-reflect (TRL) on-wafer calibration standards. The circuits were fabricated on 535-μm-thick high-resistivity ($>$ 2000 Ω·cm) silicon with a 3000-Å Si_xN_y ($\epsilon_r \approx$4.0) layer grown using plasma-enhanced chemical vapor deposition (PECVD). The CPW line dimensions ($w = 24$ μm, $g = 15$ μm) were chosen to result in a 50-Ω characteristic impedance and maintain quasi-TEM operation up to 120 GHz [9]. A further advantage of this narrow ground–to–ground spacing is that the parasitics at discontinuities (T-junctions, bends, etc.) remain relatively small.

The Si_xN_y layer is necessary to prevent formation of a rectifying contact between the CPW center conductor and the silicon substrate. The Si_xN_y layer must be removed from the gaps in the CPW metallization to prevent excessive line losses, in this case by reactive ion etching (RIE). These losses are due to a parasitic conduction layer at the interface between the high-ρ Si and the Si_xN_y, which arises from the accumulation of surface charges in the Si_xN_y layer. This effect has also been observed with SiN passivations in SiGe monolithic microwave integrated circuits (MMIC's) [10]. The Si_xN_y layer also results in a lowering of the effective dielectric constant of the line; this lowering is significant due to the high

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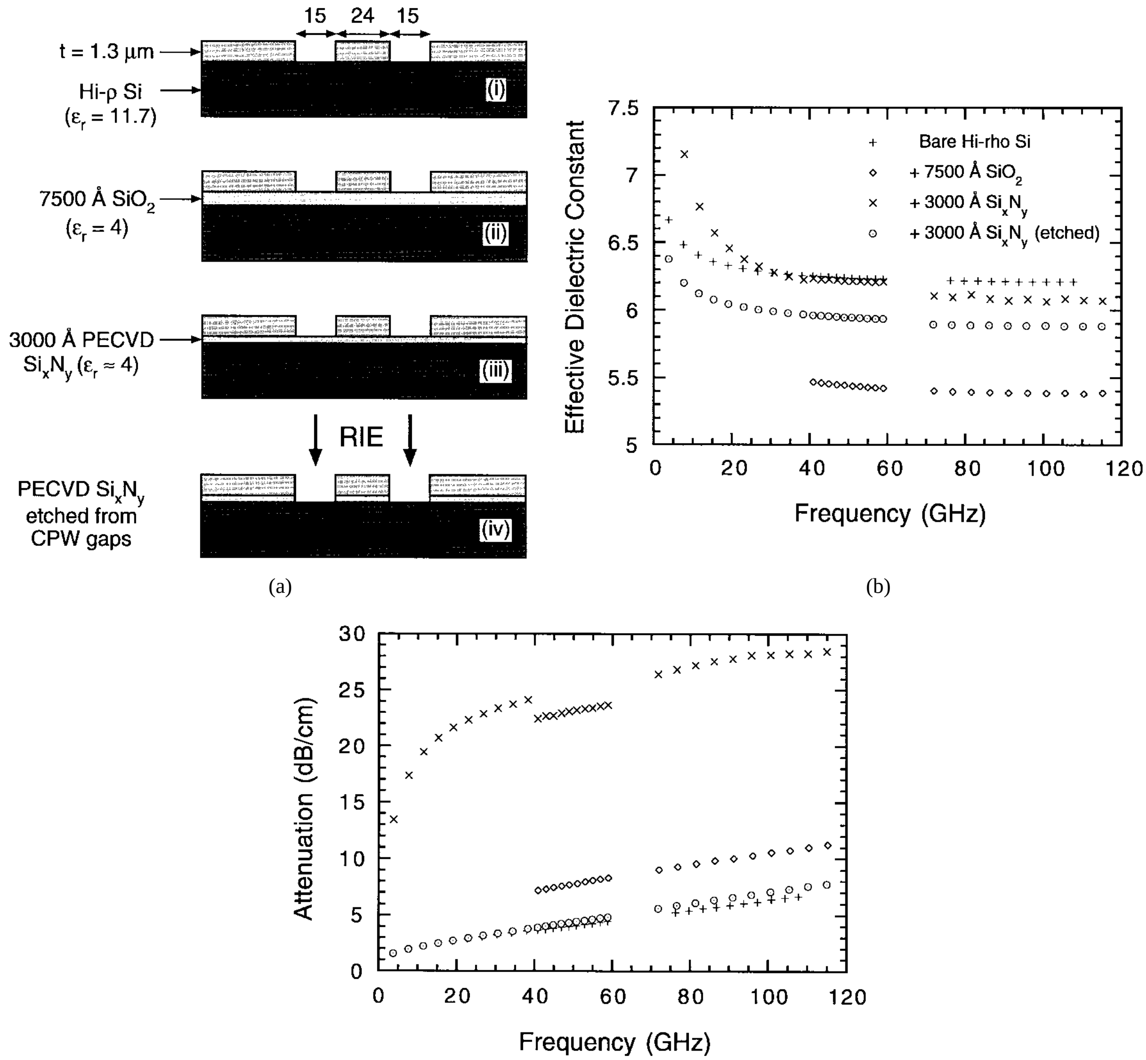


Fig. 1. (a) CPW line geometries with $w = 24$ μm, $g = 15$ μm on various dielectric layers. (i) Bare high-ρ Si. (ii) High-ρ Si with 7500 Å of SiO_2. (iii) High-ρ Si with 3000 Å of PECVD Si_xN_y. (iv) High-ρ Si with 3000 Å of Si_xN_y RIE etched from the CPW gaps. (b) Effective dielectric constant. (c) Attenuation.

concentration of field lines in the CPW gaps with a narrow ground–to–ground spacing ($w = 24$ μm, $g = 15$ μm). Fig. 1 shows the measured values of effective dielectric constant (ϵ_{eff}) and attenuation for various dielectric layer geometries: (i) bare high-ρ Si; (ii) high-ρ Si with 7500 Å of SiO_2; (iii) high-ρ Si with 3000 Å of PECVD Si_xN_y; and (iv) high-ρ Si with 3000 Å of Si_xN_y RIE etched from the CPW gaps. The ϵ_{eff} and attenuation values are determined from on-wafer TRL calibrations using NIST MultiCal software [11]. The sample with unetched Si_xN_y has the closest ϵ_{eff} value to bare Si, but suffers from excessively high losses (28 dB/cm at 94 GHz) due to the aforementioned surface-charge effect. However, when the Si_xN_y is removed from the CPW gaps, the attenuation drops to 6.9 dB/cm at 94 GHz, which is nearly that of the identical CPW line on a bare high-ρ Si substrate (6.1 dB/cm). The mixer circuits were subsequently fabricated on high-ρ Si with 3000 Å of PECVD Si_xN_y RIE etched from the CPW gaps. Distributed elements were redesigned taking into account the measured ϵ_{eff} of 5.9 for this CPW configuration.

III. Mixer Design and Simulation

The subharmonic-mixer circuit schematic is shown in Fig. 2. The circuit is designed to operate at RF frequencies of 92–96 GHz, IF frequencies of 2–4 GHz and LO frequencies of 45–46 GHz. The mixer design is based on University of Virginia SC1T7-D20 GaAs antiparallel Schottky diodes. The 38-μm-thick diode chip is 75 μm $\times$ 195 μm; the small size of the diode chip approximates a monolithic geometry while maintaining the flexibility of a hybrid design. The diode parameters are given in Table I. The Schottky diodes have an etched surface channel design in order to reduce parasitic capacitances. The junction capacitance per anode is 2.5 fF and the total capacitance from anode pad to ohmic contact pad is 16 fF (5-fF junction capacitance in parallel with 11-fF parasitic capacitance). Diode parameters extracted from the dc I–V characteristic by curve fitting and indicate that the individual junctions are exceptionally well matched. The figure-of-merit cutoff frequency ($f_c = 1/2\pi R_s C_T$) is approximately 1.5 THz.

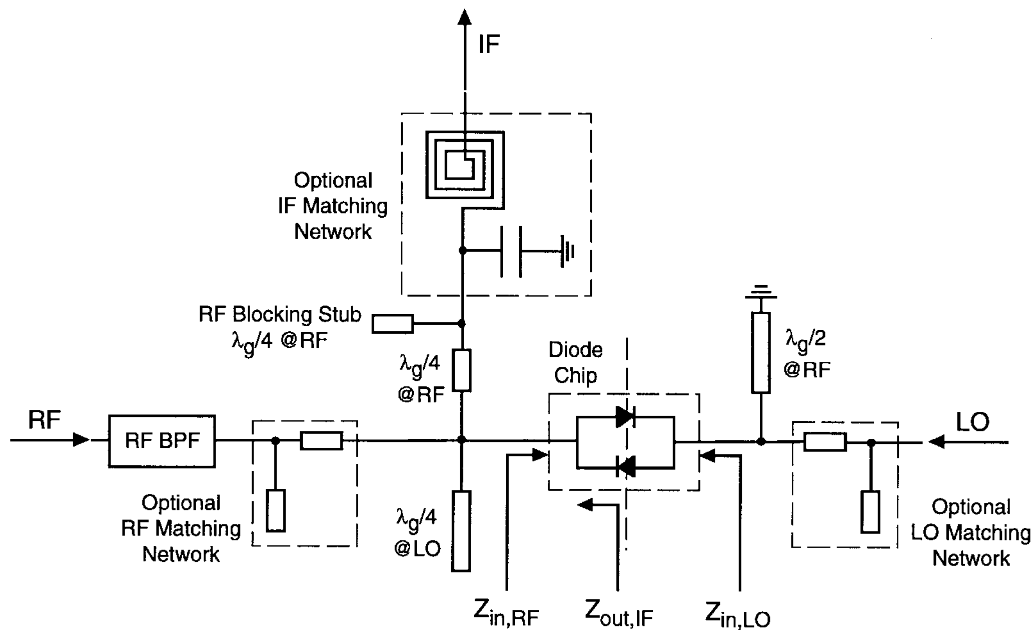


Fig. 2. Schematic of the subharmonic mixer circuit.

TABLE I
UNIVERSITY OF VIRGINIA SC1T7-D20 DIODE PARAMETERS

C_{jo} (fF)	C_p (fF)	I_s (A)	R_s (Ω)	η	ϕ_{bi} (V)	f_c (THz)
2.5	11	$4\text{x}10^{-17}$	6.0	1.16	0.84	1.5

The basic circuit has a $\lambda_{g,\mathrm{RF}}/2$ ($\lambda_{g,\mathrm{LO}}/4$) shorted-circuited stub on the LO side of the diode pair such that the diodes are terminated with a short circuit at the RF frequency, but the LO signal is not affected. The use of a shorted stub also provides a dc/IF return path to ground, and allows for straightforward dc testing of the mounted diodes. Similarly, a $\lambda_{g,\mathrm{LO}}/4$ ($\lambda_{g,\mathrm{RF}}/2$) open-circuited stub is located on the RF side of the diode pair such that the diodes are terminated with an open-circuit at the LO frequency but the RF signal is not affected. The IF signal is extracted from the RF side of the diode pair. The 75–110-GHz bandpass filter provides good rejection at 2–4 GHz, preventing IF leakage to the RF port. An open-circuited $\lambda_{g,\mathrm{RF}}/4$ stub located $\lambda_{g,\mathrm{RF}}/4$ away from the diodes in the IF output circuit prevents RF leakage to the IF port (RF choke).

The RF bandpass filter [Fig. 3(a)] was designed using CPW series open-circuited stubs [12]. A single-series stub section was analyzed using the method-of-moments [13], and the calculated S-parameters were matched to a lumped LC-equivalent circuit for use in the mixer simulations. The two-section bandpass filter can be accurately modeled by cascading the S-parameters of the single-stub section [14]. Fig. 3(b) shows the calculated response of the two-section RF bandpass filter on bare high-resistivity silicon, the corresponding response of the equivalent LC network, and the measured results on high-ρ Si with 3000 Å of $\mathrm{Si}_x\mathrm{N}_y$ RIE etched from the CPW gaps. The measured filter rejection is better than 30 dB below 4 GHz, providing the necessary IF block. Based on the S-parameter measurements the filter will contribute 1.2–1.3 dB of loss to the RF signal from 85 to 100 GHz.

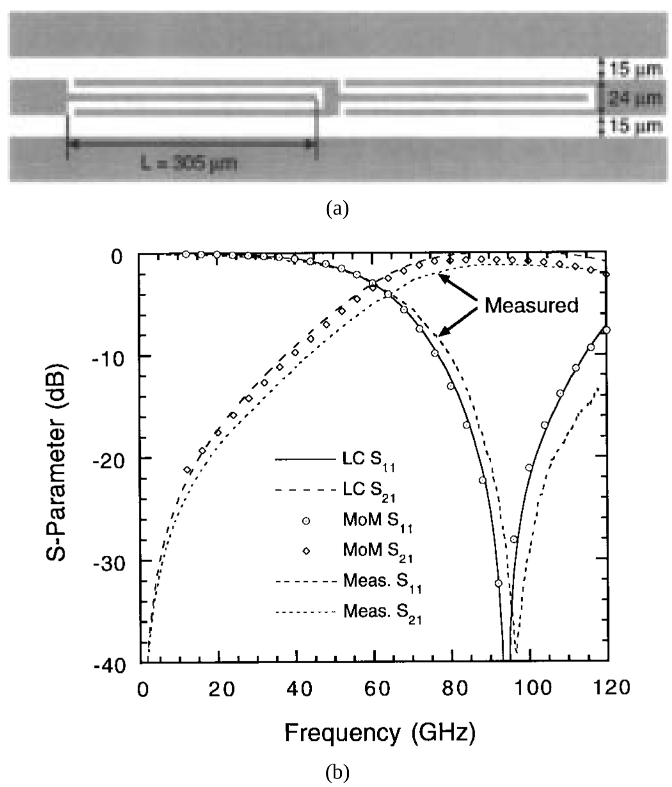


Fig. 3. (a) Two-section CPW open-circuited series stub bandpass filter. (b) S-parameters of bandpass filter from moment-method calculation, simulation of lumped element equivalent circuit, and measurement.

In order to determine the effect of the diode chip, a section of CPW line representing the diode pads was fabricated. This section consists of a CPW line with $w = 34$ μm, $g = 21$ μm with step transitions from the $w = 24$ μm, $g = 15$ μm feed lines. A practice diode chip (an actual SC1T7-D20 chip with bad *I–V* curves) was flip-chip mounted

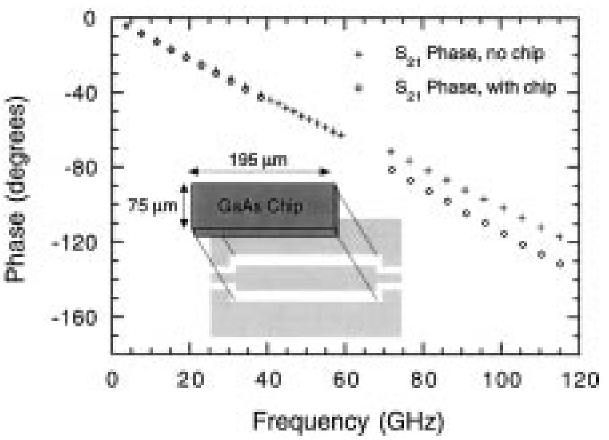


Fig. 4. Phase of S_{21} for diode mount with and without diode chip present.

TABLE II
RESULTS OF THE LIBRA HARMONIC BALANCE ANALYSIS SIMULATION

f_{LO} (GHz)	f_{IF} (GHz)	f_{RF} (GHz)	P_{LO} (dBm)	L_c (dB)	$Z_{in,LO}$ (Ω)	$Z_{in,RF}$ (Ω)
			5.0	6.16	14 - j56	85 + j21
45.5	3 (USB)	94	7.0	5.41	29 - j44	146 - j15
			9.0	6.28	41 - j28	185 -j26

on the line with EPO-TEK H20E silver epoxy.[1] The phase of S_{21} indicates that the diode chip and mounting structure contribute ∼100° of electrical delay at 94 GHz (Fig. 4). This is accounted for in the mixer simulation by including 50° transmission-line sections on either side of the diode junction model. These sections result in reactive terminations at the RF and LO frequencies rather than pure short circuits, and these reactances are included in the diode embedding impedances. This additional line length would not be significant in a fully monolithic implementation, but must be considered here in order to accurately simulate the mixer.

The mixer circuit was simulated using a commercial microwave computer-aided design (CAD) software package (HP EESof Libra v.6.0, [15]) harmonic-balance analysis (HBA) routine. In this paper, the Libra HBA was used to simulate the mixer conversion loss, and RF and LO input impedances. However, since the Libra HBA does not yield the IF output impedance, a reflection-type analysis program adapted from the work in [16] by [2] was used to calculate this quantity. Table II shows the results of the Libra HBA simulations for an LO frequency of 45.5 GHz and an IF of 3 GHz with a 50-Ω IF load. The RF and LO input impedances given in Table II are defined at the edge of the diode chip (see Fig. 2). The mixer IF output impedance calculated using the reflection algorithm was found to be relatively insensitive to the termination (short, open, or 50 Ω) of the higher RF and LO harmonics, and was around 110 Ω for IF frequencies from 2 to 4 GHz.

The Libra HBA mixer simulations are based on ideal transmission-line components and do not account for losses such as conductor loss, junction effects, loss in the RF bandpass filter, etc. Loss in the RF filter was determined from the measured S-parameters to be 1.2–1.3 dB from 85 to 100 GHz. The measured insertion loss of the fabricated open-circuited stub (RF side) is approximately 0.5 dB over 80–96 GHz, while the insertion loss of the short-circuited stub (LO side) is approximately 0.4 dB from 40 to 48 GHz. The combined RF loss in the passive network from the RF-port reference plane to the diode chip will result in a conversion loss ∼2.0 dB higher than predicted by the simulation.

The loss due to the RF mismatch is 1.2 dB, and this can be reduced by the incorporation of an appropriate matching network in the RF path. In this case, it is not clear if an RF-matching network would result in a dramatic improvement in mixer performance, since the additional circuit losses due to the network would mitigate some of the gains from matching, and may not result in accurately predictable impedance levels. On the other hand, an IF-matching network can clearly improve the mixer performance by approximately 0.5 dB. A lumped element network was designed to present a 80–100-Ω load to the mixer over the 2–4-GHz IF bandwidth. However, the fabricated matching network did not provide the desired impedance levels and actually resulted in 0.7–1.0 dB of loss from 2 to 4 GHz. The following section will present results for a subharmonic-mixer circuit without matching networks.

IV. FABRICATION AND MEASUREMENT

The 94-GHz uniplanar-mixer circuit (Fig. 5) was fabricated with probe pads at the RF, LO, and IF ports to accommodate on-wafer testing with 150-μm-pitch probes. The probe-pad–to–CPW-line transition is identical to that of the TRL calibration standards fabricated along with the test circuits. The circuit size excluding probe pads and transitions is 0.8 mm × 1.5 mm. Airbridges are included at various points in the circuit, particularly junctions, to suppress excitation of the undesired slotline (even) mode in the CPW line. The circuit was fabricated on 535-μm-thick high-resistivity (> 2000 Ω·cm) silicon with a 3000-Å PECVD-grown Si_xN_y layer. As discussed above, the Si_xN_y in the CPW gaps was removed using RIE. The CPW center conductors and ground planes are 1.3-μm-thick evaporated Ti/Al/Ti/Au, which corresponds to five skin-depths at 94 GHz, and three-and-one-half skin-depths at 45.5 GHz. The 24-μm-wide airbridges are 4-μm thick electroplated gold at a height of 3.5 μm above the CPW line. The antiparallel diode chip is mounted using flip-chip technology and bonded to the circuit using EPO-TEK H20E silver epoxy.

A. Downconversion Loss

The single-sideband (SSB) downconversion-loss-measurement setup is shown in Fig. 6. The RF source is an HP W85104A W-band source module controlled by an HP 85106C millimeter-wave network analyzer operating in CW mode; the RF signal is delivered to the mixer via a W-band Picoprobe.[2] The LO source is a 42–46 GHz mechanically tuned Gunn diode oscillator; the LO power is varied using a WR-19 level-set attenuator and delivered to the mixer via a 67-GHz Picoprobe. The 2–4-GHz IF signal is extracted by a third probe and the IF power level is measured using a spectrum analyzer. The waveguide/cable, probe and transition losses at the RF, LO, and IF frequencies are determined using

[1] EPO-TEK H20E is a product of Epoxy Technology, Inc., Billerica, MA.

[2] Picoprobe is a product of GGB Industries, Inc., Naples, FL.

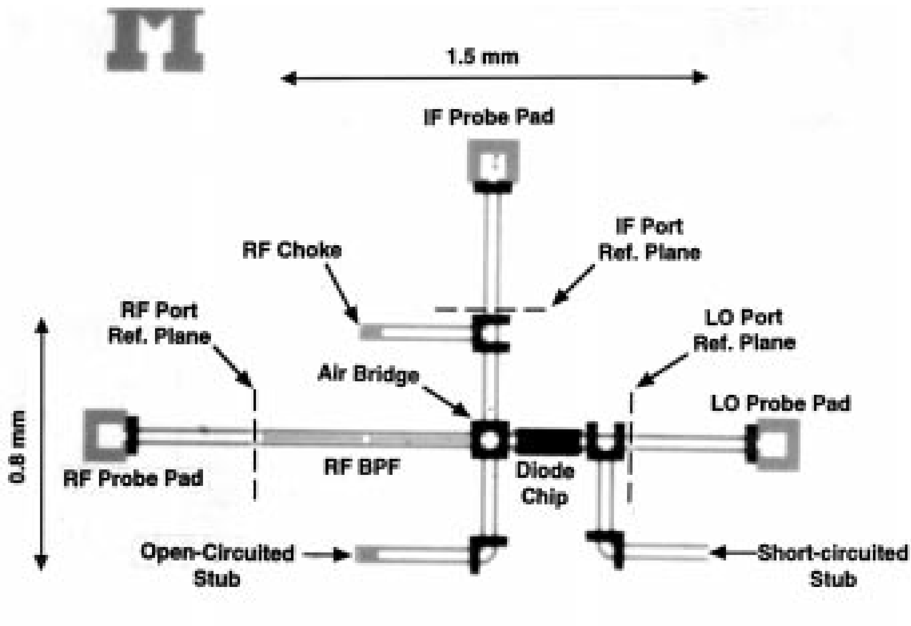


Fig. 5. Photograph of the fabricated CPW subharmonic mixer circuit. Total circuit size excluding probe pads and transitions is 0.8 mm × 1.5 mm. The size of the 38-μm-thick diode chip is 75 μm × 195 μm.

a two-tier calibration approach and are deembedded from the measured data. The measured conversion loss represents the loss from the RF-port reference plane to the IF-port reference plane, and the LO power is defined at the LO-port reference plane (see Fig. 5).

Fig. 7 shows the measured SSB downconversion loss versus RF frequency at the LO power corresponding to the minimum conversion loss, and versus LO Power at an RF frequency of 94 GHz. The measured minimum SSB conversion loss is 7.0 dB at an RF of 94 GHz, an IF of 2–4 GHz, and an LO power of 8.5 dBm, and represents state-of-the-art performance for a planar W-band subharmonic mixer. Furthermore, the mixer is broad-band with a SSB conversion loss of 7–8 dB from 90 to 96 GHz, and less than 10 dB over the 83–97 GHz measurement band. When the ~2.0-dB loss in the RF bandpass filter and open-circuited stub T-junction is added to the simulated conversion loss, there is excellent agreement between measurement and simulation.

B. Upconversion Loss

The mixer circuit was also measured as an upconverter. The probe arrangement is the same as in the downconversion experiment. In this case, an IF signal is generated using an HP 8350B sweep oscillator in CW mode and injected at the IF port. The upconverted RF signal is measured with a W-band waveguide harmonic mixer connected at the RF-waveguide reference plane whose output is measured with the spectrum analyzer. Again, all waveguide/cable, probe, and transition losses are deembedded from the measured data.

The measured port–to–port upconversion loss versus RF frequency at the LO power corresponding to the minimum conversion loss is shown in Fig. 8. The minimum measured upconversion loss is 7.0 dB at 89 GHz, and the upconversion loss at 94 GHz is 8.5 dB. The measured upconversion loss remains below 10 dB from 86 to 96 GHz. The upconverter data has more fluctuations than the downconverter data due to the lower absolute power levels being measured by the spectrum analyzer. In the downconversion experiment, the IF power reaching the spectrum analyzer was on the order of −30 to −40 dBm, while in the upconversion experiment, the power reaching the spectrum analyzer is on the order of −70 to −80 dBm due to the 40-dB conversion loss of the WR-10 harmonic mixer.

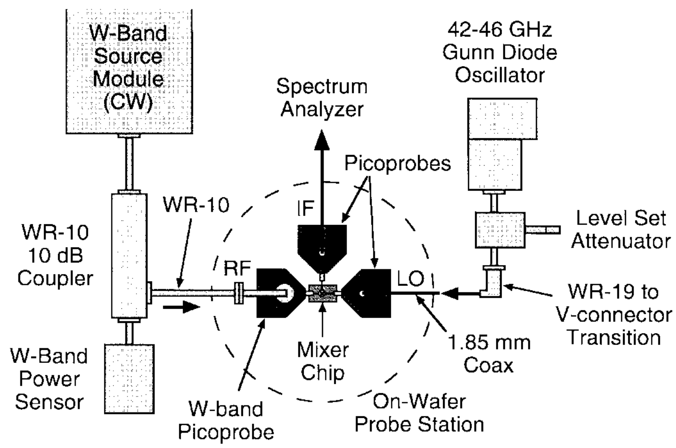


Fig. 6. Mixer downconversion-loss-measurement set-up.

C. Isolation

Fig. 9(a) shows measured and simulated port–to–port LO–RF and LO–IF isolation. The LO–RF isolation was measured with the LO power delivered to the mixer as shown in Fig. 6. The RF port is probed with a second 67-GHz Picoprobe connected to a U-band waveguide harmonic mixer via a 1.85-mm coaxial cable and appropriate transitions, and the IF port is terminated in a matched load. The output of the U-band harmonic mixer is measured on the spectrum analyzer. The measured LO–RF isolation agrees well with the

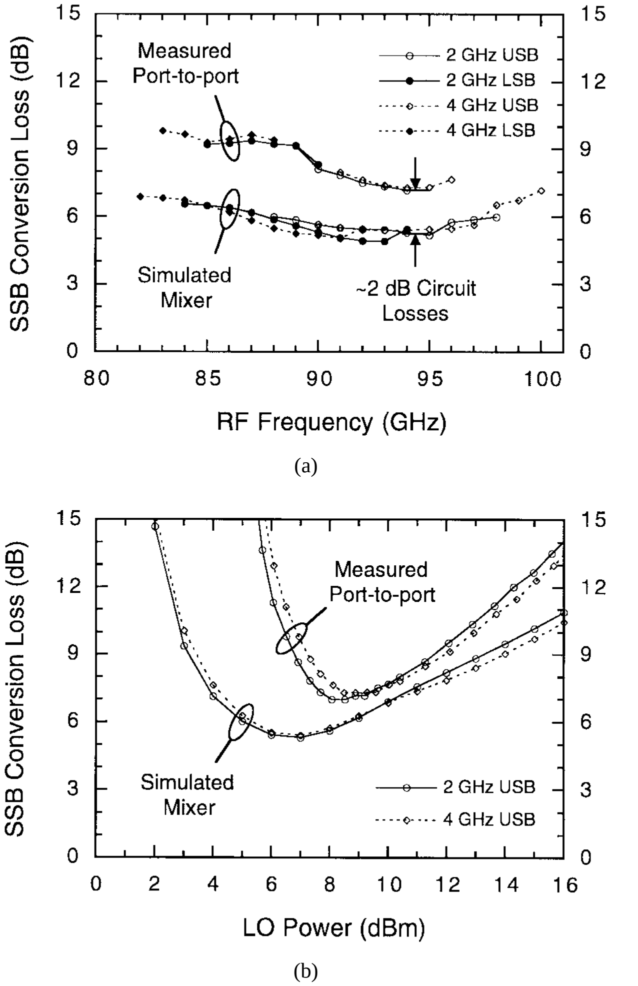


Fig. 7. Measured and simulated port-to-port SSB downconversion loss (a) versus RF frequency at LO power corresponding to the minimum conversion loss (2 and 4 GHz IF) and (b) versus LO Power at an RF frequency of 94 GHz (2 and 4 GHz IF).

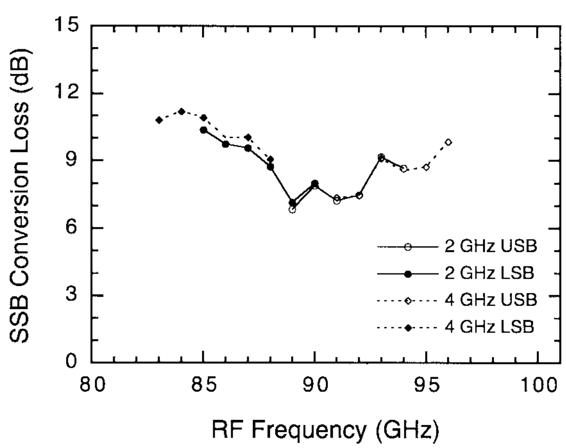


Fig. 8. Measured port–to–port upconversion loss versus RF frequency at LO power corresponding to the minimum upconversion loss (2 and 4 GHz IF).

simulation and is better than -40 dB for LO frequencies of 45–46 GHz. The LO–RF isolation is expected to remain better than -30 dB from 43 to 48 GHz. The LO–IF isolation was also measured using a similar setup, with the 67-GHz probe at the IF port and the RF port terminated in a matched load.

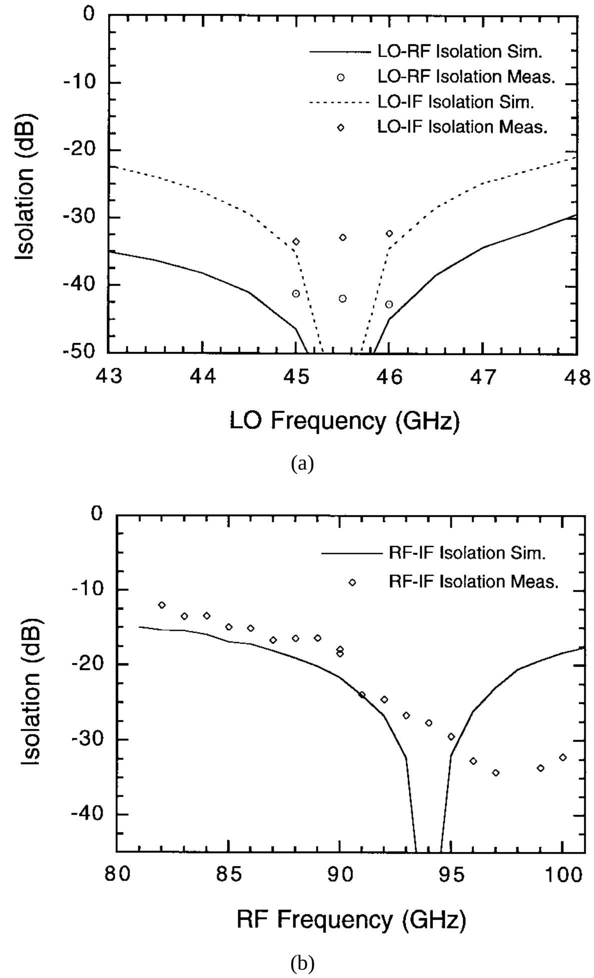


Fig. 9. (a) Measured and simulated port–to–port LO–RF and LO–IF isolations. (b) Measured and simulated port–to–port RF–IF isolation.

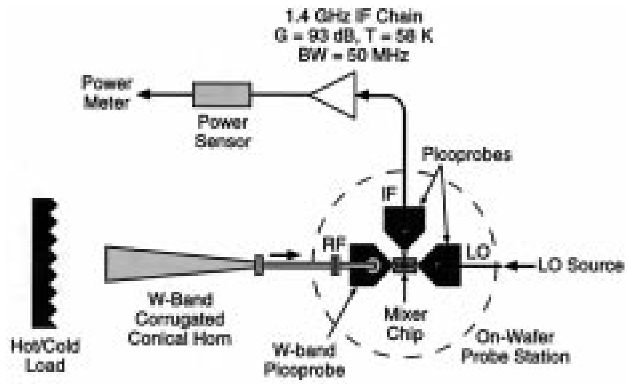
The LO–IF isolation also agrees well with the simulation and is below -30 dB from 45 to 46 GHz.

Fig. 9(b) shows the measured and simulated port–to–port RF–IF isolation. The RF–IF isolation was measured with the RF source configuration shown in Fig. 6. The IF port was probed with a second W-band probe connected to a W-band waveguide harmonic mixer. The measured RF–IF isolation has a minimum around 98 GHz, which is due to the open-circuited $\lambda_{g,\mathrm{RF}}/4$ stub (located $\lambda_{g,\mathrm{RF}}/4$ away from the diodes in the IF output circuit) resonating at 97.4 GHz instead of the design frequency of 94 GHz. Still, the circuit provides better than -25 dB of RF–IF isolation from 90 to 100 GHz.

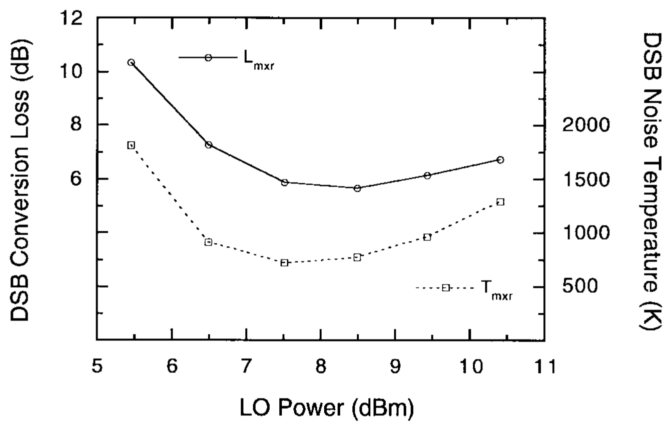
D. Noise Temperature

Noise temperature measurements were performed using the Y-factor method [17] and the measurement setup shown in Fig. 10(a). A corrugated conical horn (scalar feed horn) with a WR-10 feed was connected to the RF input of the mixer. Microwave absorber (ECCOSORB VHP-2-NRL)[3] at room temperature (290 K) placed in the near field of the horn provided the hot load while an identical absorber immersed

[3]ECCOSORB VHP-2-NRL is a product of Emerson and Cuming, Canton, MA USA.



(a)



(b)

Fig. 10. (a) Measurement setup for measuring the DSB conversion loss and noise temperature using the Y-factor method. (b) The measured DSB conversion loss and noise temperature at an IF of 1.4 GHz and an LO frequency of 45.5 GHz. The measured conversion loss represents the loss from the RF-port reference plane to the IF-port reference plane, and the LO power is defined at the LO-port reference plane (see Fig. 5).

in liquid nitrogen (77 K) provided the cold load. Based on past experience with this type of measurement [18], a cold temperature of 85 K is used in the Y-factor calculations since the absorber is not a perfect black body radiator. The IF port of the mixer was connected to a 1.4-GHz IF chain with a gain of 93 dB, noise temperature of 58 K, and a bandwidth of 50 MHz. RF and IF waveguide/cable, probe, and transition losses are deembedded from the measured data. The efficiency of the corrugated conical horn is 94% and is included as an additional 0.3 dB of RF loss prior to the mixer.

The measured double-sideband (DSB) noise temperature and conversion loss are shown in Fig. 10(b). The DSB noise temperature is 725 K (DSB NF $=$ 5.4 dB) at an IF of 1.4 GHz, an LO frequency of 45.5 GHz, and an LO power of 7.5 dBm. The DSB conversion loss is 5.7 dB at an LO power of 8.5 dBm. This roughly corresponds to a SSB conversion loss of 8.7 dB. With an LO frequency of 45.5 GHz and an IF of 1.4 GHz, the upper-sideband (USB) is at 92.4 GHz and the lower-sideband (LSB) is at 89.6 GHz. The measured SSB conversion losses at 92.4 GHz (7.5 dB) and at 89.6 GHz (8.5 dB) result in an average SSB value of 8.0 dB, which is in reasonably close agreement with the Y-factor method results. It is expected that the mixer will result in a DSB noise

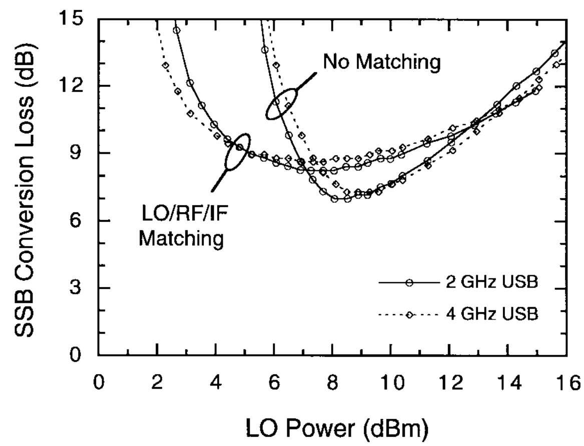


Fig. 11. Measured port–to–port SSB downconversion loss versus LO Power at an RF frequency of 94 GHz (2 and 4 GHz IF) for mixer circuit without matching networks and mixer circuit with LO, RF, and IF matching networks.

temperature of approximately 650 K [5.1-dB DSB noise figure (NF)] at an LO frequency around 46.5 GHz.

V. Conclusions and Future Work

A high-performance W-band subharmonic mixer has been realized in uniplanar CPW technology. The total circuit size excluding probe pads and transitions is less than 0. 8 mm $\times$ 1. 5 mm. The measured minimum SSB conversion loss of 7. 0 dB at an RF of 94 GHz and an IF of 2–4 GHz represents state-of-the-art performance for a planar W-band subharmonic mixer, and is even competitive with fundamental planar Schottky diode mixers. The mixer is broad-band with a SSB conversion loss of less than 10 dB over the 83–97 GHz measurement band. The measured LO–RF isolation is better than -40 dB for LO frequencies of 45–46 GHz, which is important in receiver applications to minimize radiation of LO power from the RF port. The mixer is expected to have a minimum DSB noise temperature of 650 K. These results are in excellent agreement with HBA simulations after including passive circuit losses in the RF path. The excellent performance of this subharmonic-mixer design is due to the high-quality antiparallel diode chip and the well-characterized CPW components. It is important to note that commercial microwave CAD programs accurately predicted the performance of this mixer, showing the maturity of mixer-design techniques up to 100 GHz.

Current work at the University of Michigan is focused on lowering the LO power requirements of the subharmonic mixer. A 94-GHz subharmonic mixer was built with simple single-stub RF and LO matching networks, and the lumped element IF matching network mentioned in Section III. The measured results at an RF of 94 GHz (Fig. 11) indicate an improved LO match, with a SSB conversion loss less than 9.5 dB for LO powers from 4 dBm (2.5 mW) to 12 dBm (15.8 mW). To the authors' knowledge, this represents the lowest LO power requirement of any W-band planar mixer to date. The minimum measured SSB conversion loss of 8.2 dB is 1.2 dB worse than the mixer without matching networks, due primarily to additional losses in the nonoptimal lumped element IF matching network and the RF matching network. Still this conversion loss is approximately 2 dB better than previous planar SHM's at W-band. Future work includes optimizing the matching network designs to realize 7–8-dB conversion loss at a 3-mW power level.

Other future work includes extending the design for operation at an RF of 140–150 GHz, and integrating the mixer with planar antennas to realize high-performance millimeter-wave front-ends. The mixer design can readily be extended to a fully monolithic implementation.

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