A 94 GHz UNIPLANAR SUBHARMONIC MIXER

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Abstract - **A uniplanar subharmonic mixer has been implemented in coplanar waveguide 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 is less than 2 mm x 1.5 mm. The measured minimum SSB conversion loss is 9.4 dB at a 94 GHz RF, 2 GHz IF, and an LO power of 9 dBm, and represents state-of-the-art performance for a planar W-band subharmonic mixer. Potential applications for this mixer are millimeter-wave receivers for smart munition seekers and automotive collision avoidance radars.**

I. INTRODUCTION

Subharmonic mixers (SHMs) downconvert an RF signal which is approximately the *n*th harmonic of the local oscillator (LO) frequency to an intermediate frequency (IF). For mixing at even harmonics of the LO frequency, SHMs typically use a non-linear device with an anti-symmetric current-voltage characteristic, such as an anti-parallel 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 subharmonic mixers have been realized at W-Band with both anti-parallel Schottky barrier diodes [3] and HFETs [4]. Minimum conversion losses on the order of 10 dB have been achieved with these circuits. The purpose of this work is to implement a 94 GHz subharmonic mixer in *uniplanar* coplanar waveguide (CPW) technology. The advantages of CPW over microstrip are well known. CPW is a uniplanar structure allowing both shunt *and* series connection of circuit elements, and simple ground-plane connection 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. Also, well-developed on-wafer measurements techniques using coplanar probes simplify circuit testing, particularly at millimeter-wave frequencies. There has been considerable recent work in the development of millimeter-wave CPW circuits [5], including a 64-77 GHz fundamental mixer, for European automotive radar applications.

II. DESIGN AND FABRICATION

The subharmonic mixer circuit schematic is shown in Figure 1. 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 anti-parallel Schottky diodes. The 38 μm-thick diode chip is 75 μm x 195 μm. 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). The diode parameters extracted from the DC I-V characteristic by curve fitting indicate that the individual junctions are extremely well-matched with $I_s = 4 \times 10^{-17}$ A, $R_s = 6.5\ \Omega$, $\eta = 1.163$, and $\phi_{bi} = 0.842$ V. The figure-of-merit cutoff frequency ($f_c = 1/2\pi R_s C_T$) is approximately 1.5 THz.

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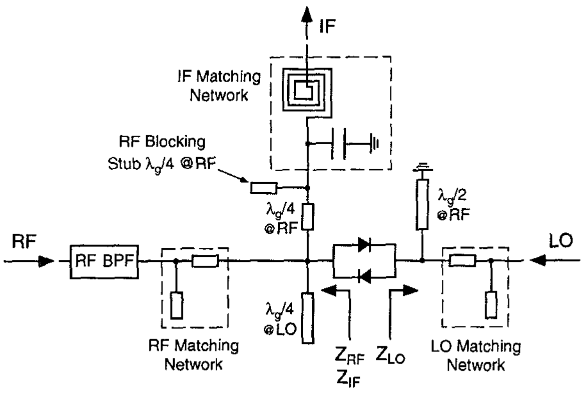

Fig. 1: Schematic of the subharmonic mixer circuit.

The basic circuit has a $\lambda_{g,RF}/2$ ($\lambda_{g,LO}/4$) shorted stub on the opposite side of the diode pair from the RF input, such that the diodes are terminated with a short circuit at the RF frequency. 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 $\lambda_{g,LO}/4$ ($\lambda_{g,RF}/2$) open-circuited stub is located on the opposite side of the diode pair from the LO input. The IF signal is extracted from the RF side of the diode pair. A 75-110 GHz bandpass filter provides good rejection at 2-4 GHz, preventing IF leakage to the RF port. An open-circuited $\lambda_{g,RF}/4$ stub located $\lambda_{g,RF}/4$ away from the diodes in the IF output circuit prevents RF leakage to the IF port. The RF bandpass filter (Fig. 2) was designed using CPW series open-circuited stubs [6]. The overall line dimensions (w = 24 μm, g = 15 μm) were chosen to result in a 50 Ω characteristic impedance and maintain quasi-TEM operation in the frequency range of interest [7]. A single series stub section was analyzed using the method-of-moments, and the S-parameters matched to a lumped LC equivalent circuit. Figure 2 also shows the calculated response of the two-section RF bandpass filter and the corresponding response of the equivalent LC network. The filter rejection is better than 30 dB below 4 GHz, providing the necessary IF block.

Mixer modeling was performed using a harmonic balance program adapted by Kormanyos *et al.* [2] for subharmonic mixing using antiparallel Schottky diodes. This program allows mixer simulation in the presence of arbitrary embedding impedances and is a useful design tool for designing mixer matching networks. The diode input impedances for optimum conversion loss are 35 - j45 Ω at a 94 GHz RF, 100 - j325 Ω at a 45.5 GHz LO, and 100 Ω at a 3 GHz IF. A single-stub RF matching network, consisting of a 50 Ω, 35° through-line with a 60 Ω, 54° open-circuited stub, was designed to present 35 + j45 Ω to the diodes at 94 GHz. A single-stub LO matching network consisting of a 80 Ω, 59° through-line with a 60 Ω, 62° open-circuited stub, was designed to present 100 + j150 Ω to the diodes at 45.5 GHz. The IF matching network is a lumped element design in order to obtain a more compact structure than possible with either a single-stub tuner or a stepped-impedance filter. The IF matching network is an inductor-capacitor-inductor (1.4 nH-0.5 pF-1.4 nH) "T" network which is designed to provide 80-100 Ω load to the mixer over the 2-4 GHz IF bandwidth. The CPW rectangular spiral inductor design was adapted from [8].

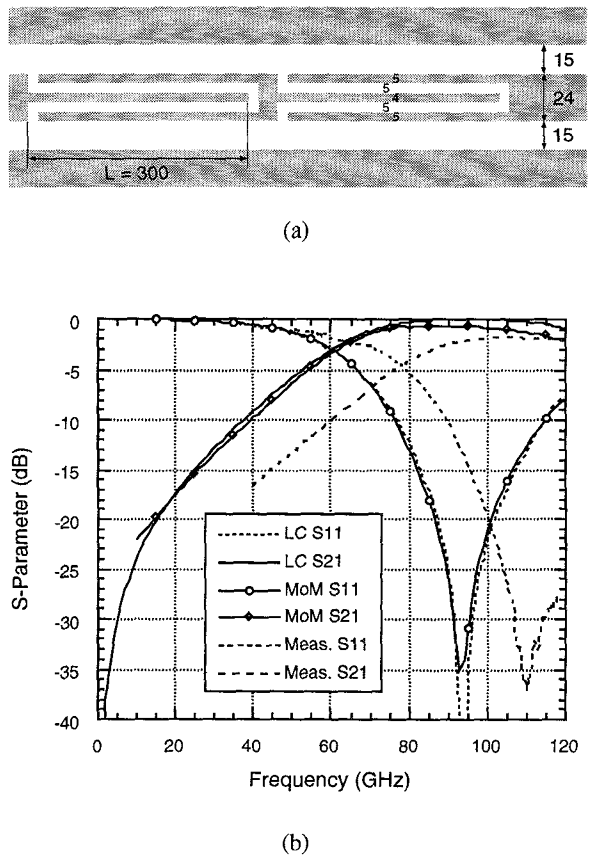

Fig. 2: (a) Two-section CPW open-circuited series stub bandpass filter; (b) S-parameters of bandpass filter from moment method calculations, response of equivalent lumped element LC network, and measured performance.

The mixer circuit design was implemented in HP-EESof Communication Design Suite v5.0 [9] and tested using harmonic balance analysis. A minimum SSB conversion loss of 5.7 dB is predicted for a 94 GHz RF,

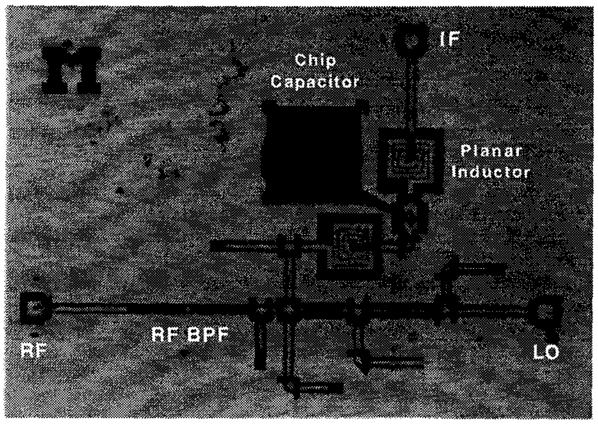

Fig. 3: Photograph of the fabricated CPW subharmonic mixer circuit.

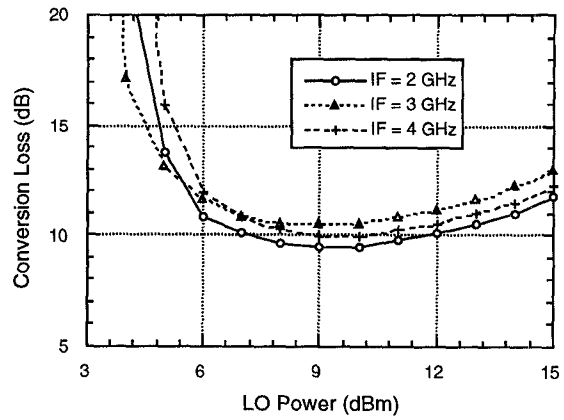

Fig. 4: Measured mixer SSB conversion loss vs. LO Power at RF = 94 GHz for 2, 3, and 4 GHz IF frequencies. Minimum conversion loss is 9.4 dB at 2 GHz IF and 9 dBm LO power.

3 GHz IF, and a LO power of 5 dBm (3 mW). This includes 0.7 dB loss (not including conductor loss) through the RF bandpass filter calculated by the moment method analysis. Since the simulations are based on ideal transmission line components, there will be additional losses due to conductor loss, junction effects, etc. In light of these effects, we expect to obtain port-to-port conversion loss of 7-8 dB with this design.

A photograph of the fabricated circuit is shown in Figure 3. The circuit size excluding the probe pads is less than 2 mm x 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 is fabricated on 535 µm-thick high-resistivity (> 2000 Ω·cm) silicon with a 7500 Å furnace-grown SiO_2 layer. The SiO_2 layer is necessary to prevent formation of a rectifying contact between the CPW center conductor and the silicon substrate. The CPW center conductors and ground planes are 1.3 µm thick evaporated Ti/Al/Ti/Au, which corresponds to 5 skin-depths at 94 GHz, and 3.5 skin-depths at 45.5 GHz. The 24 µm wide airbridges are plated gold 4 µm thick 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 with silver epoxy. The chip capacitor is a Metelics MBIC-1002 binary trimming capacitor (500 µm x 500 µm) with 0.25, 0.5, 1.0, and 2.0 pF capacitances available. The capacitor is mounted on the ground plane with silver epoxy and connected to the IF matching circuit with a 18 µm (0.7 mil) diameter bondwire. In a monolithic version of the circuit, the chip capacitor could be easily replaced by a MIM capacitor, resulting in a more compact layout.

III. MEASUREMENTS

The mixer circuit was designed to accommodate testing on a microwave/millimeter-wave probe station. RF and LO signals were delivered to the circuit using appropriate frequency band Picoprobes [10]. The RF source was an HP W85104A W-Band source module controlled by an HP 85106C Millimeter-Wave Network Analyzer operating in CW mode. The LO source was a 45-46 GHz varactor tuned Gunn oscillator with a maximum output power of 240 mW at 46 GHz. The LO power was varied using a level-set attenuator. The 2-4 GHz IF signal is extracted by a third probe and measured using a spectrum analyzer. The measured port-to-port SSB conversion loss of the subharmonic mixer circuit is shown in Figure 4. RF losses from the sampling point (waveguide 10 dB coupler) to the reference plane of the mixer RF port are deembedded from the measurements. IF losses from the mixer IF port to the spectrum analyzer have not been deembedded but are expected to be small. The minimum measured SSB conversion loss is 9.4 dB at a 94 GHz RF, 2 GHz IF, and an LO power of 9 dBm.

In addition to the mixer circuits, various passive components which make up the mixer were fabricated on the same silicon substrate to allow for independent characterization. A 50 Ω CPW line (w = 24 µm, g = 15 µm) fabricated on the > 2000 Ω·cm high-resistivity silicon substrate with a 7500 Å furnace-grown SiO_2 layer has a measured attenuation of 7.5 dB/cm at 45 GHz and 10.2 dB/cm at 94 GHz determined from an on-wafer calibration using the NIST MultiCal software [11]. This is compared to a 50 Ω line on bare high-resistivity silicon which exhibits a loss of 3.9 dB/cm at 45 GHz and 6.1 dB/cm at 94 GHz. The RF signal to the mixer propagates over approximately 1 mm, so the line attenuation adds 1 dB to the mixer conversion loss at 94 GHz.

Furthermore, the effective dielectric constant over the frequency range of interest drops from 6.2 for the bare silicon substrate to 5.4 for the substrate with 7500 Å of SiO_2 because a greater number of field lines are propagating in the lower dielectric constant material. This can be seen from the measured response of the RF bandpass filter shown in Figure 2. The lower ε_{eff} results in a 15% increase in the resonant frequency of the filter. The measured loss through the filter is 2.1 dB at 94 GHz. The lower ε_{eff} also results in higher resonant frequencies for the various stubs in mixer circuit. The next mixer iteration will be fabricated on a 3000 Å-thick layer of PECVD Si_xN_y, which will reduce both the loss and the decrease in ε_{eff}. We anticipate a 1.5-2 dB improvement in the mixer conversion loss, as well as a lower LO power requirement, as a result of reduced millimeter-wave losses, corrected stub resonant frequencies, and further optimization of the IF matching network.

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