Design and Measurements of a Novel Subharmonically Pumped Millimeter-Wave Mixer Using Two Single Planar Schottky–Barrier Diodes

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***Abstract*— A novel design of a subharmonically pumped millimeter-wave mixer operating at room temperature was developed and realized. The double sideband conversion loss and mixer noise temperature were measured to be 6.2 dB and 930 K, respectively, at a local oscillator frequency of 73 GHz and an IF of 1.5 GHz. These results are comparable to the best published results measured for subharmonically pumped mixers at similar frequencies. The mixer shows good performance even at IF's up to 9.5 GHz resulting in a useful RF range from 136 GHz to 156 GHz. For the first time a subharmonically pumped millimeter-wave mixer was designed without the use of any scale model measurements or other high-frequency measurements. The whole design process occurred on the basis of computer simulations. Two single low-capacitance planar air-bridge type Schottky–Barrier diodes are used as the mixing elements.**

I. Introduction

Subharmonically pumped millimeter-wave mixers have already been demonstrated very successfully [1], [2]. The antiparallel diode pair used in that type of mixer has the advantage of subharmonic pump and inherent local oscillator (LO) noise suppression [3], [4], both reducing the demands on the LO source. For fundamentally pumped single ended mixers special means are required to superimpose the LO and the RF signal before they enter the mixer input. This is not the case for a subharmonically pumped mixer resulting in a simpler receiving system set-up and a reduction of potential causes for performance degradations by additional loss or standing waves [5].

Since the first published results [1], [2] only little progress has been made in improving the performance of subharmonically pumped millimeter-wave mixers, although several modifications of the design were tested, e.g., [6]–[9]. Just recently a new design was proposed [10] that showed slightly improved performance in comparison to that of the original Bell Laboratories group.

For most of the subharmonically pumped millimeter-wave mixers developed so far, whisker-contacted dot-matrix diodes were used. The mechanical contacting procedure for this type of diode is most difficult and the reproducibility of the characteristic of the contact is poor. The work of [10] was stimulated by recent developments in the fabrication of high-quality, low-capacitance planar integrated Schottky–Barrier diodes [8], [11]–[18]. The planar Schottky–Barrier diodes show good reproducibility of their characteristic parameters and they can easily be mounted on a printed circuit.

Independent of the developments in the field of millimeter-wave mixers the capabilities of various new and further developed commercial high-frequency computer-aided design (CAD) products have increased rapidly in the past decade. The following software products were used to aid the design procedure presented in this work: The high frequency structure simulator (HFSS), the microwave design system (MDS), MAFIA and *em*.

HFSS is a software package of Hewlett-Packard for the analysis of the electrodynamic behavior of passive structures; MDS, also a product of Hewlett-Packard, is an integrated computer-aided engineering software package for microwave devices; MAFIA, an acronym for "the solution of MAxwell's equations using the Finite Integration Algorithm," is a software package for the simulation of various electromagnetic problems, distributed by CST GmbH, Darmstadt, F.R.G.; ***em*** is a software product by Sonnet Software, Inc., Liverpool, NY, USA, for the electromagnetic analysis of predominatly planar structures.

The recent developments in the field of high-quality planar Schottky–Barrier diodes and in the field of high-frequency CAD motivated the investigations performed for this work: This paper describes a new design of a low-noise subharmonically pumped planar diode mixer at 146 GHz. The design is developed by the use of commercially available high-frequency CAD products. No scaled model measurements or other high-frequency measurements are used and no tuning elements (except waveguide backshorts to tune transitions to microstrip line) are provided for the mixer circuit to adjust its performance.

II. Design Concept

The mixer presented here uses two single planar air-bridge type Schottky–Barrier diodes from the University of Virginia (UVa type SC2T3) mounted in flip-chip geometry on a shielded quartz microstrip circuit, Fig. 1 and Fig. 2. The RF signal is coupled into the microstrip circuit via an E-probe reaching into the RF waveguide (standard WR6 rectangular waveguide). The LO waveguide (standard WR12 rectangu-

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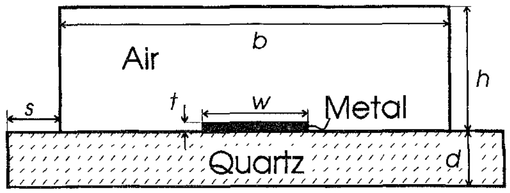


Fig. 1. Cross section of the shielded microstrip transmission line. $d = 50\mu$m is the thickness of the quartz substrate ($\varepsilon_r = 3.75$, $\tan\delta = 0.0015$), $t \approx 2\mu$m is the thickness of the metal strip, $w = 50 \ldots 450\mu$m is the width of the metal strip, $h = 150\mu$m and $b = 800\mu$m are the height and width of the top cover of the line. $s = 50\mu$m is the length the quartz juts out over the top cover to have sufficient mechanical support.

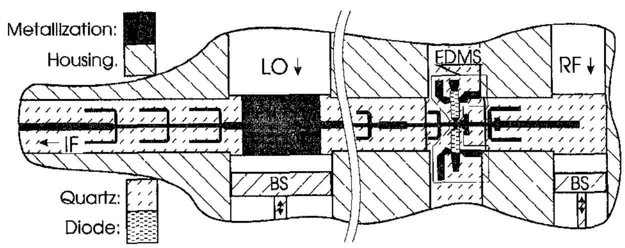


Fig. 2 Schematic overview of the mixer circuit. The cut plane of this sketch is along the E-plane of the RF and LO waveguide, just above the plane of the quartz metallization. The LO and RF waveguides are standard rectangular waveguides (WR12 and WR6, respectively). EDMS: Extended diode mounting structure. BS: Movable backshorts.

lar waveguide) crosses the microstrip transmission line and, together with a LO band-stop filter, provides LO/IF diplexing.

The circuit in the vicinity of the diodes is set up in a second microstrip line channel perpendicularly crossing the main channel of the microstrip circuit. This part of the circuit is called "extended diode mounting structure" (EDMS, see Fig. 2) in the following part of the text. The two diodes are placed antiparallely to each other and in shunt to the main microstrip line. Three microstrip line elements, called "multifrequency short circuits" (MFSC) in the following part of the text, provide the necessary short circuit at the outer pad of each diode for dc, the LO frequency, and the RF. Two open stubs are responsible for the short at the RF and the LO frequency, and a short transmission line to the shielding block provides the dc short to ground.

The performance of a subharmonically pumped mixer sensitively depends on the electrical length between the diodes in the mixer circuit. This indicates the analysis presented in [19] and was confirmed by nonlinear analyzes of simple sample mixer circuits performed with the aid of MDS and an external postprocessing computer program called MIXSIM. MIXSIM, written in IDL, was developed by the author. It uses the results of a harmonic balance analysis (performed by MDS) of a two-diode mixer circuit to calculate noise and conversion loss based on the method presented in [20].

To be able to control the electrical length between the diodes by appropriately designing the external microstrip circuit, two single planar diodes are used in the design presented here instead of a two-diode chip commonly used in other subharmonically pumped planar diode mixers (e.g., [10]).

The design procedure of the mixer circuit can be divided into several parts that are described in more detail in the next section:

First, a transmission line structure is chosen that prevents higher modes from propagating for frequencies below 160 GHz and that allows to realize a broad range of characteristic impedances to ease the design of impedance matching networks.

For the second part of the design procedure, among the few commercially available planar Schottky diodes an appropriate one is chosen. Its parasitic elements in the chosen transmission line environment are analyzed. An equivalent circuit description of the planar diode in this environment is elaborated that is used in the subsequent analysis.

In the third part, the performance of the EDMS (extended diode mounting structure) is optimized for minimum conversion loss using the nonlinear optimization tools of MDS. The nonlinear analysis also gains RF, LO, and IF impedances of the EDMS that have to be matched to the external circuit.

In the forth part, appropriate impedance matching and filtering circuits are designed.

Independently of the previous steps, the transitions from rectangular waveguide (RF and LO input) to 50 Ω-microstrip line are designed by the aid of HFSS to show low reflection over a broad frequency band.

III. Design Procedure

A. Microstrip Transmission Line

The microstrip transmission line cross section (Fig. 1) is designed to prevent all higher (nonquasi-TEM) modes below 160 GHz from propagating. The cut-off frequency calculations are performed with the aid of MAFIA. The microstrip transmission line on 50 μm quartz allows a broad range of characteristic impedances: 20 Ω to 80 Ω (in accordance with the I-V definition of characteristic impedance) for metal strip widths from 450 μm to 50 μm, Fig. 3. The range of characteristic impedances ease the realization of impedance matching networks. The propagation characteristics of the microstrip transmission line, i.e. characteristic impedance, effective dielectric constant and attenuation coefficient, are calculated with the aid of HFSS and are used later for the design and optimization of the impedance matching and filtering structures.

B. Planar Diode Model

The active part of the planar Schottky–Barrier diode is described in the usual way by an exponential current-voltage (IV) characteristic

$$I(V) = I_s(\exp(\alpha V) - 1) \tag{1}$$

and a reverse square root capacitance law

$$C(V) = \frac{C_{J0}}{\sqrt{1 - \frac{V}{V_J}}} \tag{2}$$

with

$$\alpha = q/\eta k_B T \tag{3}$$

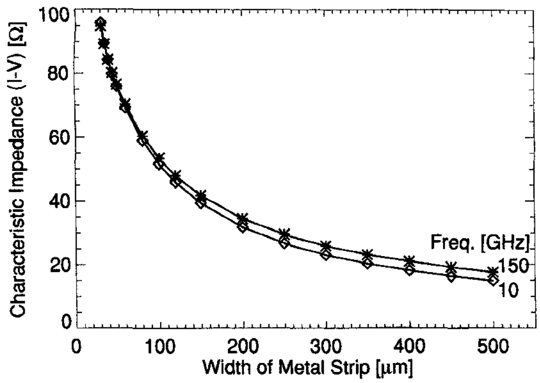


Fig. 3. Characteristic impedance according to the voltage-current definition of the quasi-TEM mode of the shielded microstrip line for various strip widths w and frequencies f. The calculations are performed with the aid of HFSS including dielectric and metal conduction losses (electric loss tangent $\tan\delta = 0.0015$, resistivity $\rho = 4.1 \cdot 10^7 \Omega/\mathrm{m}$).

where V is the junction voltage, i.e. the voltage drop at the diode V_d minus the voltage drop at its series resistance R_s, I_s is the reverse saturation current of the diode, η the ideality factor, T the physical temperature, C_{j0} the zero bias junction capacitance, V_j the built-in potential, q is the electron charge and k_B is Boltzmann's constant.

The respective parameters for the diode used here are shown in Table I as given by the manufacturer. The nominal overall chip dimensions of the SC2T3 diode are 380μm × 130μm × 40 μm (length × width × thickness). In the millimeter-wave frequency range the chip gives rise to significant parasitic elements. An equivalent circuit description is suggested by the manufacturer including especially the pad-to-pad capacitance and the series inductance of the anode finger.

As the equivalent circuit parameters of the chip are expected to be dependent on the specific mounting configuration of the chip in the mixer circuit, the parameters given by the manufacturer can only be used as a rough guideline. For the flip-chip geometry used here a detailed analysis of the parasitics similar to the analysis presented in [21] for a two-diode mounting structure was carried out. The analysis of two whisker-contacted dot-matrix diodes mounted on a suspended-substrate stripline described in [21] revealed a significant coupling between the parasitics of the two diodes. The analysis for the present work yields a detailed equivalent circuit description of the parasitics of a flip-chip mounted single planar diode that is suitable to be included in subsequent nonlinear circuit analyzes.

C. Nonlinear Optimization

The calculation effort to simulate the performance of a nonlinear circuit is comparatively high. The effort rises quickly with the number of harmonic frequencies that are considered and with the complexity of the circuit under consideration. For the computer-aided optimization of the performance of a circuit, several simulations of the circuit are necessary: The optimization software varies the parameters that are to be optimized to minimize a certain error function defining the goal of the optimization. The more parameters that are to be optimized the more circuit simulations have to be performed and the longer the time needed for the software to find the optimum combination of optimization parameters.

TABLE I[1]
PARAMETERS OF THE SC2T3 DIODE CHIP

	Parameter	Value
IV parameters*	$1/\alpha$	30.4 mV
	I_s	$1.63 \cdot 10^{-16}$ A
	η, (T=300 K)	1.175
CV parameters	C_{j0}	5.5 fF
	V_j	950 mV

Even on modern high-performance workstations the calculation time needed for the nonlinear optimization of circuits with moderate complexity considering about ten harmonics of the fundamental frequency may easily exceed an hour if more than five parameters are to be optimized.

To reduce calculation time the part of the circuit considered in the nonlinear optimization should be as simple as possible and the number of optimization parameters should be low. Therefore, only the EDMS is taken into consideration for the nonlinear optimization here. To reduce the calculation effort further, the microstrip line elements of this part of the circuit are all modeled as ideal transmission lines for the nonlinear optimization.

The LO and RF source impedances are matched to the respective input impedances of the EDMS. The IF load impedance is set to 50 Ω.

In the optimization procedure the length and characteristic impedances of the transmission lines are allowed to vary within a reasonable range to minimize the conversion loss from the matched RF source to the 50 Ω IF load. The following parameters (see Table II) result from the optimization: Single sideband (SSB) conversion loss L_C, RF and LO input impedance of the EDMS Z_{RF} and Z_{LO}, neccessary LO power P_{LO}, lengths and impedances of the open stubs of the MFSC's and the transmission lines connecting the diodes to the main microstrip line.

D. Linear Optimization

The results of the nonlinear optimization form the starting point for the linear optimization part, i.e. the design of the microstrip circuit of the mixer. The ideal transmission lines used for the nonlinear optimization are realized by suitable microstrip transmission lines, matching networks are designed to match the RF and LO input impedance to 50 Ω, and suitable networks for the filtering of IF, LO frequency, and RF are designed.

[1]Parameters as specified by the manufacturer or derived from manufacturer's parameters (*), respectively.

TABLE II[2]
NONLINEAR OPTIMIZATION RESULTS

Parameter	Explanation	Value
L_C	SSB conversion loss	6.25 dB
T_M	SSB noise temperature	460. K
$\Re(Z_{\mathrm{RF}})$	Real part of RF source impedance	9.5 Ω
$\Im(Z_{\mathrm{RF}})$	Imaginary part of RF input impedance	−1.8 Ω
$\Re(Z_{\mathrm{LO}})$	Real part of LO input impedance	6.2 Ω
$\Im(Z_{\mathrm{LO}})$	Imaginary part of LO input impedance	−15.2 Ω
P_{LO}	LO power available from source	7.4 mW

The following method was applied to determine the optimal layout of the different parts of the microstrip line circuit: The microstrip transmission line parameters calculated with the aid of HFSS were used to define a transmission line model for MDS. Models for the microstrip line discontinuities like bends, steps in width, and T-junctions were taken from the system model library of MDS. Using these models, standard procedures of MDS were used to help design the parts of the microstrip circuit.

The layout obtained in this way was analyzed by ***em*** to check the performance predictions of MDS. Critical parts of the circuit like open stub lengths were fine tuned interactively with the help of the ***em*** results.

This two-step procedure (first step with the help of MDS, second with the help of ***em***) was chosen because of the following reasons: The models used in the MDS analyzes allow computer-aided optimization of relatively complex circuits within a moderate calculation time. But the models neglect effects like coupling and higher mode generation which may be significant in the millimeter-wave range.

Like every electromagnetic field simulator, ***em*** on the other hand accounts for all electrodynamic effects that are probably neglected in models for circuit simulators, but typical simulation times are an order of magnitude higher in comparison to a circuit simulator. Thus, the ***em*** results are expected to be more accurate, but the calculation time is typically too high to perform computer-aided optimizations.

Using the procedure just described, the circuit shown in Fig. 2 was designed part by part with the exception of the waveguide to microstrip line transitions. The resulting circuit is predicted to have the following performance: The RF insertion loss from the RF waveguide to microstrip line transition to the EDMS is 0.4 dB; the RF input voltage standing-wave ratio (VSWR) of the same part is less than 1.2 for 8 GHz bandwidth; the LO mismatch loss from the LO waveguide to microstrip line transition to the EDMS is 1.0 dB; the IF mismatch loss for the same part is less than 0.2 dB for IF less than 6 GHz; the RF rejection in the direction of the IF path is better than 50 dB for 20 GHz bandwith.

These predictions cannot be measured directly. Even if adequate measurement equipment would be available for the necessary millimeter wave measurements, the effort for manufacturing the single parts of the circuit would still be high. The predictions are confirmed afterwards by the agreement of the measured and predicted mixer performance.

E. Waveguide Transitions

The LO and RF waveguide to microstrip line transitions were designed with the help of HFSS and MDS disregarding dielectric and conduction losses. The RF transition was optimized to show low signal insertion loss over a broad bandwidth. The LO transition was especially designed to show low IF reflection along the microstrip line crossing the LO waveguide; the LO insertion loss is not too critical, because sufficient LO power is available.

The analysis of the LO transition also allows the determination of the optimal position of the LO band stop filter that prevents the LO signal from propagating in the direction of the IF output. The position of this filter together with the movable waveguide backshort determines the LO reflection of the transition.

For the structure simulations performed by HFSS the moving backshorts of the transitions are removed and a second waveguide port is defined at these planes instead. The HFSS simulation results are then used to define a three-port (in case of the RF transition) or a four-port (in case of the LO transition), respectively, for the circuit simulator MDS. Within MDS the circuits are completed by a rectangular waveguide transmission line of variable length, short-circuited at one end and attached to one of the waveguide ports of the three-port or four-port, respectively. The transition's performance

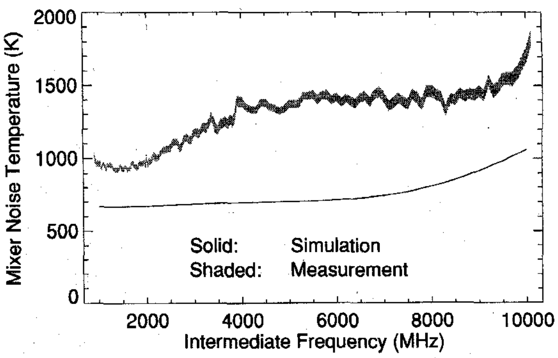
[2]The values given apply to the extended diode mounting structure (EDMS) not to the whole mixer. The optimization was performed for 2 to 5 GHz IF. L_C and $\Im(Z_{RF})$ are slightly dependent on the IF and a mean value is given here. The LO and RF source impedances are conjugate matched to the respective input impedances. The image band termination impedance equals the RF source impedance. The mixer noise temperature was determined afterwards by the postprocessing procedure MIXSIM.

in dependence on the backshort position can then be analyzed by MDS.

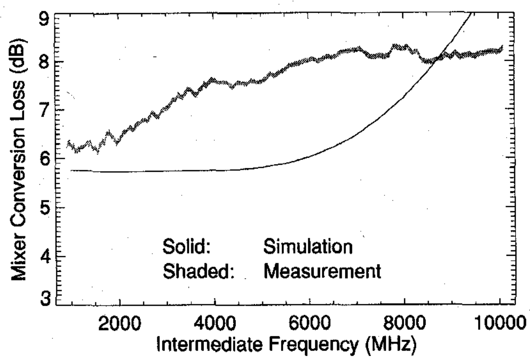
This method has two main advantages: First, the geometries of the transitions with the backshorts removed are symmetrical with respect to the plane perpendicular to the quartz substrate, cutting the width of the metal strip into half. There is a second plane of symmetry of the LO transition cutting the height of the LO waveguide into half. Making use of the symmetry allows a significant reduction of the HFSS simulation times. Second, the transitions' performances can be calculated quasi instantaneously for every backshort setting having analyzed only a single structure by HFSS. (Note that calculation times are only a fraction of a second for the circuit simulation by MDS, but several hours for the structure simulation by HFSS.)

The transitions were analyzed for different E-probe shapes. The optimal shapes are shown in Fig. 2. The E-probe of the RF transition reaches just half the height of the waveguide into it having the width of the 50 Ω microstrip line. Due to the calculations this transition shows less than 0.3 dB insertion loss over the whole WR6-band (110 to 170 GHz) for the optimal backshort position.

The E-probe of the LO transition should be as large as possible if low IF reflection along the microstrip line is wanted. The transition acts as a series induction for the IF that can partly be compensated by the capacitive load of a large E-probe. However, if the E-probe gets too large the tunability of the transition for good LO transmission is degraded. The optimal E-probe (Fig. 2) is a compromise between LO and IF performance. The IF VSWR is calculated to be 1.36 at 4 GHz and 2.11 at 10 GHz. With a 50 μm misplacement of the open circuit plane of the LO band stop filter the transition can still be tuned to a LO VSWR of better than 1.2 according to the calculations.



(a)



(b)

Fig. 4. Measured and simulated DSB mixer noise temperature and conversion loss versus IF. LO frequency $f_{LO} = 73$ GHz, RF $f_{RF} = 2f_{LO} \pm f_{IF}$. The vertical widths of the shaded areas mark the statistical measurement uncertainties. The systematical measurement uncertainties are about 100 K for the noise temperature and 0.05 dB for the conversion loss. The effects of the horn antenna and the transition from microstrip line to the SMA connector of the IF output are not considered in the simulation.

IV. Mixer Performance

A. Simulation

To predict the performance of the whole mixer two corrections have to be made to the results of the nonlinear optimization of the EDMS presented in Table II. First, the values have to be recalculated using the dc parameters of the diode chips as actually measured after mounting them in the mixer circuit instead of the nominal values of Table I. Second, the values have to be corrected for the loss of the remaining parts of the mixer circuit, namely the RF and IF insertion loss.

The series resistance of the chips in the circuit was measured to be 8.5 Ω instead of the expected 4 Ω (3Ω specified for the diodes plus 1 Ω accounting for additional line resistance). The unexpected high R_s was probably caused by the diode pad to microstrip line connection glued by epo-tek adhesive (epo-tek is a registered trademark of Epoxy Tech. Company). The noise temperature and conversion loss of the EDMS was recalculated to 900 K and 7.7 dB (SSB), respectively.

The RF insertion loss including the RF input waveguide, the transition to microstrip line and the matching and filtering microstrip circuit was estimated to be about 0.76 dB. The IF loss disregarding the microstrip to SMA connector transition was estimated to be about 0.45 dB at 4 GHz.

The results of detailed, frequency-resolved calculations of the double sideband (DSB) mixer noise temperature and conversion loss accounting for all corrections just mentioned, and including the frequency dependent results of the nonlinear simulations and performance of the matching and filtering microstrip circuit are shown as the solid line in Fig. 4.

B. Measurement

A modified hot/cold measurement was carried out to determine the noise and loss of the mixer. At the input to the first amplifier of the IF detection system a step attenuator was placed to allow the variation of the IF system input noise temperature T_{IF}. By measuring the system noise temperature T_{Sys} for various values of T_{IF}, the mixer noise temperature T_M and conversion loss L_C are determined by a linear regression using the well-known equation

$$T_{Sys} = T_M + L_C T_{IF}. \tag{4}$$

A HP8970S noise figure measurement system was used to frequency-resolved measure the noise power at the IF amplifier output. A computer was used to control the step attenuator and the HP8970S and to acquire the measurement data via an IEEE

TABLE III[3]
PUBLISHED PERFORMANCE OF SUBHARMONICALLY PUMPED MIXERS

RF (GHz)	$T_{M,\mathrm{DSB}}$(K)	$L_{M,\mathrm{DSB}}$(dB)	Remark	Reference
95	1050	5.5	Whisker Contacted Diodes	[1]
146	930	6.2	Two Single Planar Diodes	This work
180	?	12.8	Integrated planar Antenna	[22]
180	1375 (System Noise)		Antiparallel Planar Diode Chip	[9]
180	1200	7.5	Whisker Contacted Diodes	[7]
205	795	5.7	Antiparallel Planar Diode Chip	[10]
	860	5.7		
	995	6.7		
205	900	6.6	Whisker Contacted Diodes	[2]

interface bus. The results of the measurements are shown in Fig. 4 as the shaded curve.

The DSB mixer noise temperature and conversion loss measured around 1.5 GHz IF is 930 K and 6.2 dB, respectively. These values are comparable to the best values published for similar mixers (Table III).

C. Comparison Between Measurement and Simulation

The conversion loss obtained from the simulation is in good agreement with the measured conversion loss for 1 to 2.5 GHz IF (5.8 dB instead of 6.2 dB), whereas the simulated noise temperature is significantly lower than measured (690 K instead of 920 K for low IF). Both, measurement and simulation show a degradation of the mixer performance for increasing IF, but the measured mixer performance starts to degrade above 2.5 GHz whereas the simulations predict the onset of the degradation at about 5 GHz IF.

The most likely reason for the difference between measurement and simulation are the idealizations made in the modeling of the EDMS. As mentioned above, ideal transmission lines were used to model the microstrip line structures there. Coupling between close parts of the different lines and higher mode coupling was neglected.

A confirmation of this supposition is given by the measured current imbalance of the pumped diodes: As the geometry of the diode chips is asymmetrical with respect to the exchange of anode and cathode, no perfectly symmetrical conditions for the two diodes can be realized by the EDMS. This results in an imbalance of the pumped diodes, giving rise to a net dc current measurable at the IF connector. The nonlinear analysis of the EDMS predicts a dc current of about 500 μA. The dc current actually measured at the IF connector of the pumped mixer was 2 mA, a considerably high value normally indicating bad mixer performance.

An imbalance between the two diodes in a subharmonically pumped mixer gives rise to degradation in the noise temperature [4] which could explain the observed difference between measurement and simulation.

V. CONCLUSION

For the first time a subharmonically pumped millimeter-wave mixer was designed solely with the use of high-frequency design software. The measured mixer performance is comparable to the best results published for similar mixers and is slightly worse than predicted by the simulations.

The most likely reason for the slight deviation between measurement and simulation was discussed. It indicates that more care has to be taken in modeling the EDMS if the agreement is to be further improved. An improvement of the model for the EDMS is also expected to allow to further improve the mixer's performance because one gains a better knowledge of the EDMS impedances and one is allowed to better balance the two diodes.

Despite the deviation between measurement and simulation the existing high-frequency CAD software products were demonstrated to be helpful for the design of a millimeter-wave mixer. This suggests that these products are also useful for the development of other active millimeter-wave devices. An example of how HFSS and MDS are used to analyze a planar diode doubler from 85 GHz to 170 GHz is given in [23].

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[3]The noise temperature and conversion loss values are all double sideband (DSB) values. If single sideband values were found in the literature they are transformed to DSB values assuming: $T_{M,\mathrm{SSB}} = 2 \cdot T_{M,\mathrm{DSB}}$ and $L_{C,\mathrm{SSB}} = L_{C,\mathrm{DSB}}(\mathrm{dB}) = 3$ dB.

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REFERENCES

[1] T. F. McMaster, M. V. Schneider, and W. W. Snell, "Millimeter-wave recievers with subharmonic pump," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, no. 12, Dec. 1976.

[2] E. R. Carlson and M. V. Schneider, "Subharmonically pumped millimeter-wave reciever," in *Dig. 4th Int. Conf. Infrared Millimeter Waves*, Dec. 10–15, 1979, pp. 82–83.

[3] M. Cohn, J. E. Degenford, and B. A. Newman, "Harmonic mixing with antiparallel diode pair," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, no. 8, Aug. 1975.

[4] P. S. Henry, B. S. Glance, and M. V. Schneider, "Local-oscillator noise cancellation in the subharmonically pumped down-converter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, May 1976.

[5] P. F. Goldsmith and N. Z. Scoville, "Reduction of baseline ripple in millimeter radio spectra by quasioptical phase modulation," *Astron. Astrophys.*, no. 82, pp. 337–339, 1980.

[6] I. Galin, "An improved structure for subharmonically pumped millimeter-wave mixers," in *Dig. 12th Int. Conf. on Infrared and Millimeter Waves*, Dec. 1987, pp. 123–124.

[7] C. M. Mann, D. N. Matheson, and M. R. B. Jones, "183 GHz double diode subharmonically pumped mixer," *Int. J. Infrared Millim. Waves*, vol. 10, no. 9, 1989.

[8] J. W. Archer, R. A. Batchelor, and C. J. Smith, "Low-parasitic, planar Schottky diodes for millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 38, no. 1, Jan. 1990.

[9] P. H. Ostdiek, T. W. Crowe, and I. Galin, "Integration of an antiparallel pair of Schottky barrier diodes in millimeter wave mixers," in *Dig. 15th Int. Conf. on Infrared and Millimeter Waves*, Orlando, FL, Dec. 1990, pp. 401–403.

[10] P. H. Siegel, R. J. Dengler, I. Mehdi, J. E. Oswald, W. L. Bishop, T. W. Crowe, and R. J. Mattauch, "Measurements on a 215-GHz subharmonically pumped waveguide mixer using planar back-to-back air-bridge Schottky diodes," *IEEE Trans. Microwave Theory Tech.*, vol. 41, no. 11, Nov. 1993.

[11] D. G. Garfield, R. J. Mattauch, and W. L. Bishop, "Design, fabrication and testing of a novel planar Schottky–Barrier diode for mm and submm wavelength," in *IEEE Southeastcon Proc.*, Knoxville, TN, Apr. 1988, pp. 154–160.

[12] D. G. Garfield, R. J. Mattauch, and S. Weinreb, "RF performance of a novel planar millimeter-wave diode incorporating an etched surface channel," *IEEE Trans. Microwave Theory Tech.*, vol. 39, no. 1, Jan. 1991.

[13] J. L. Bowers, W. L. Bishop, and T. W. Crowe, "The reliability of planar GaAs Schottky diodes," in *Dig. 17th Int. Conf. on Infrared and Millimeter Waves*, Pasadena, CA, Dec. 1992, pp. 216–217.

[14] W. L. Bishop, E. R. Meiburg, R. J. Mattauch, and T. W. Crowe, "A micron-thickness, planar Schottky diode chip for terahertz applications with theoretical minimum parasitic capacitance," in *MTT-S Int. Microwave Symp. Dig.*, Dallas, TX, 1990.

[15] W. L. Bishop, R. J. Mattauch, T. W. Crowe, and L. Poli, "A planar Schottky diode for submillimeter wavelength," in *Dig. 15th Int. Conf. on Infrared and Millimeter Waves*, Orlando, FL, Dec. 1990.

[16] W. L. Bishop, T. W. Crowe, R. J. Mattauch, and P. H. Ostdiek, "Planar Schottky barrier mixer diodes for space applications at submillimeter wavelength," *Microwave Opt. Technol. Lett.*, vol. 4, no. 1, Jan. 1991.

[17] W. L. Bishop, T. W. Crowe, R. J. Mattauch, and H. Dossal, "Planar GaAs diodes for THz frequency mixing applications," in *Dig. 3rd Int. Symp. Space THz Technology*, Univ. of Michigan, Ann Arbor, MI, Mar. 1992.

[18] ——, "Planar GaAs diode fabrication: progress and challenges," in *Dig. 4th Int. Symp. on Space THz Technology*, University of California, Los Angeles, Mar. 30–Apr. 1, 1993.

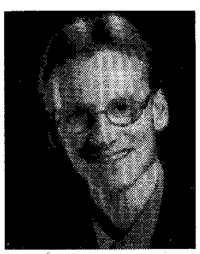
[19] A. R. Kerr, "Noise and loss in balanced and subharmonically pumped mixers: Part II – Application," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, no. 12, pp. 944–950, Dec. 1979.

[20] ——, "Noise and loss in balanced and subharmonically pumped mixers: Part I – Theory," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, no. 12, pp. 938–943, Dec. 1979.

[21] S. D. Vogel, "Analysis of a diode mounting structure of a subharmonically pumped millimeter-wave mixer," in *Dig. 18th Int. Conf. on Infrared and Millimeter Waves*, Colchester, Essex, U.K., Sept. 1993.

[22] B. K. Kormanyos, C. C. Ling, G. M. Rebeiz, P. H. Ostdiek, W. L. Bishop, and T. W. Crowe, "A planar wideband millimeter-wave subharmonic reciever," in *Proc. Int. Microwave Symp.*, Boston, MA, June 9-13, 1991.

[23] J. Tuovinen and N. R. Erickson, "Finite element analysis of a planar diode doubler," in *Dig. 4th Int. Symp. on Space THz Technol.*, University of California, Los Angeles, pp. 326–339, Mar. 30–Apr. 1, 1993.



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