A W-Band Self-Oscillating Subharmonic MMIC Mixer

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***Abstract*— A novel 77-GHz subharmonically pumped self-oscillating mixer has been implemented using monolithic-microwave integrated-circuit (MMIC) technology. The microstrip mixer achieves mixing and doubling simultaneously using a single 4 μm $\times$ 15 μm $\times$ 0.15 μm gate-length pseudomorphic high electron-mobility transistor, eliminating the need for an external local oscillator. The total circuit size is 1 mm $\times$ 2 mm, including coplanar probing pads. The mixer exhibits a measured double-sideband (DSB) conversion loss of 11.0 dB at 77.6 GHz and an average measured DSB conversion loss of 15 dB from 70 to 85 GHz, which compares well with simulated results. To the authors' knowledge, this is the first demonstration of a subharmonically pumped MMIC self-oscillating mixer operating with state-of-the-art performance at W-band frequencies.**

***Index Terms*—MMIC, self-oscillating, subharmonic, W-band, mixer.**

I. Introduction

Subharmonic mixers, which downconvert an RF signal using a harmonic of the local oscillator (LO), typically use an antiparallel arrangement of two devices, either diodes [1] or heterojunction FET's (HFET's) [2]. This allows efficient generation of even harmonics of the externally applied LO, which is then used to mix with the RF. These systems have been demonstrated at millimeter and submillimeter frequencies in waveguide and planar environments and show performance approaching those achieved with fundamentally pumped mixers [3]. Conversion losses of 7.0 dB have been achieved at W-band using uniplanar technology with Schottky diodes [1].

In this paper, a different approach has been implemented. Instead of using two devices to generate the second harmonic of the LO, a single active pseudomorphic high electron-mobility transistor (pHEMT) has been used. Also, the device is designed in such a way that it self-oscillates at the LO frequency. The second harmonic of this self-oscillation is used to mix with the W-band RF. This creates a two-port network, which receives an RF and downconverts it to an IF without the need for any external signals. Since the LO frequency is approximately half the RF, the complexity of the device and feedback network is reduced. This makes the design compatible with lower cost monolithic-microwave integrated-circuit (MMIC) gate fabrication techniques. Since fewer devices are used, there is an overall improvement in reliablilty of the circuit, costs are reduced, and dc power requirements are also reduced. Although a pHEMT is used in this paper, the design approach allows other devices such as MESFET's [4] and heterojunction bipolar transistors (HBT's) [5] to be used. Such devices cannot operate in their fundamental mode at W-band, but still could be used in this design.

This paper differs from other solutions by employing a single device to achieve self-oscillation, mixing, and doubling simultaneously, thus leading to reduced MMIC area usage, improved reliability, improved fabrication yield, and reduced power consumption. To the authors' knowledge, this is the first demonstration of a subharmonically pumped MMIC self-oscillating mixer at W-band, and this design represents a substantial improvement over current W-band mixers.

II. Self-Oscillating-Mixer Design

The self-oscillating mixer consists of a common-source feedback oscillator and a gate mixer (see Fig. 1). The oscillator design follows the theory outlined in [6]. This consists of embedding the simulated S-parameters of the device at the frequency of oscillation in an appropriate network so as to synthesize a one-port negative resistance looking into the drain. This generates contours of negative resistance and drain reactance with respect to gate and source reactances. From these contours, the desired negative resistance and reactance is selected and the required gate and source reactances are determined. These reactances are physically realized using open- or short-circuit stubs. This, and all other simulations, was implemented using a large signal model of the 4 μm $\times$ 15 μm 0.15 μm gate-length pHEMT on a Hewlett-Packard (HP) Microwave Design Simulator (MDS). Since the RF (77 GHz) is to be applied at the gate of the device, it is necessary to ensure that the impedance looking into the gate is within the Smith chart boundary. This is achieved by ensuring that negative resistance is not generated by the feedback network at the RF. Since a negative resistance is required at the LO frequency (39 GHz) to enable self-oscillation, a frequency-dependent feedback network must be used to satisfy these conditions. If a short circuit is presented to the source of the device at the RF, then a negative resistance cannot be generated. If a negative resistance was generated, then it would not be possible to match to the RF input impedance

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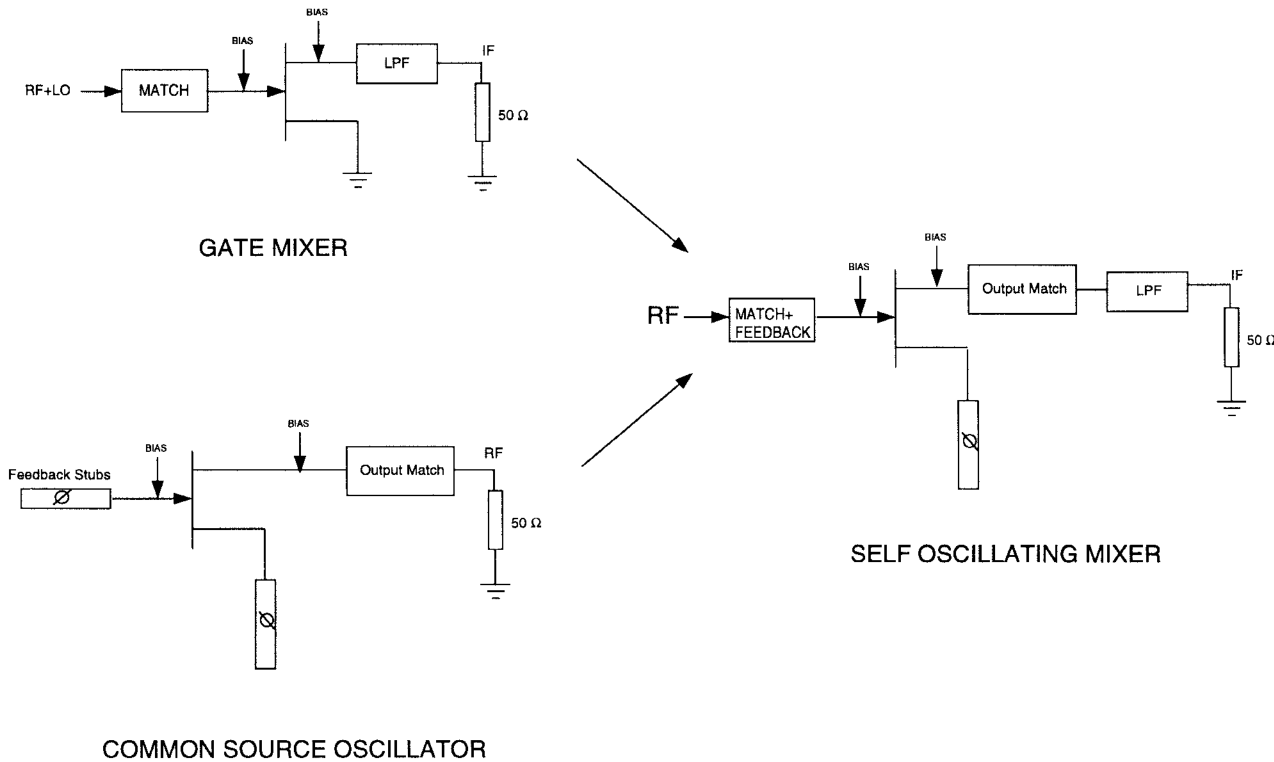

Fig. 1. Schematic of gate mixer and common-source oscillator.

with a passive matching network. The contour plot in Fig. 2 shows the areas of negative resistance with respect to gate and source feedback reactance. It can be seen that, with the source impedance at zero (i.e., short circuit), whatever value of gate reactance is used, a negative resistance cannot be generated. To short circuit the source at 77 GHz, a half-wavelength short-circuit stub is used. This stub presents an open circuit at 39 GHz and, therefore, has minimal effect on the LO feedback circuit. To generate source feedback at 39 GHz, an open-circuit stub is connected to a second-source terminal of the device. This stub generates capacitive feedback at 39 GHz, and its length can be varied in the design stage so as to achieve the optimum negative resistance. Variations in the length of this stub have minimal effect on the RF impedance at the gate at 77 GHz. This is because the half-wavelength stub always creates a short circuit and, thus, the impedance remains approximately constant. This is shown in Fig. 3, where a large change in the length of the stub has little effect on the RF input impedance. The impedance is insensitive to variations in this length and, thus, this length can be used as a design parameter. The length of this stub is selected to maximize negative resistance at the drain of the device. The source feedback network is shown in Fig. 4.

The gate feedback circuit consists of a coupled line bandpass filter, which is open circuited at the LO frequency and acts as a transmission line at the RF. At the LO, the filter has approximately the same characteristics as a 0.1λ open-circuit stub. This was determined by observing the stopband input impedance of the filter. To present the correct reactance, a length of transmission line whose length is 0.1λ less than the determined length is placed between the gate and filter. The gate feedback arrangement is shown in Fig. 5.

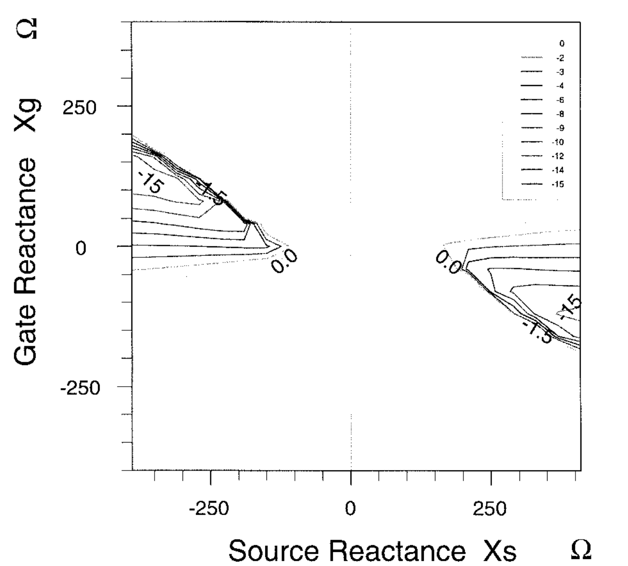

Fig. 2. Contour plot showing areas of negative resistance with respect to gate and source reactance at 77 GHz.

The free-running oscillation was produced by using a single-stub matching network to present the load impedance

$$z_l = \frac{-r_d}{3} - jx_d \qquad (1)$$

where r_d is the negative resistance at the drain, which was determined by simulations, and x_d is the drain reactance.

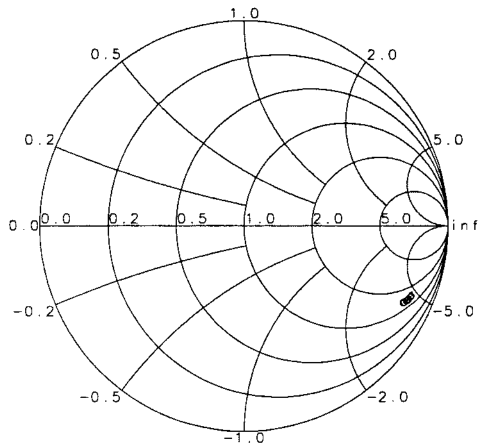

Fig. 3. Variation of gate impedance with respect to variations in the length of the open-circuit stub length x. Length x varies from 0 to 10 mm, step 0.1 mm. Frequency $=$ 77 GHz.

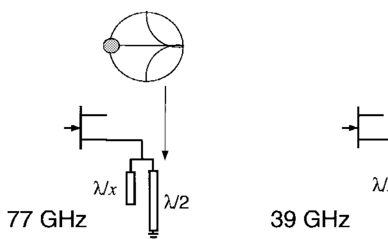

Fig. 4. Schematic of the source feedback arrangement. X is a optimization design parameter used to maximize $-R$ at the drain.

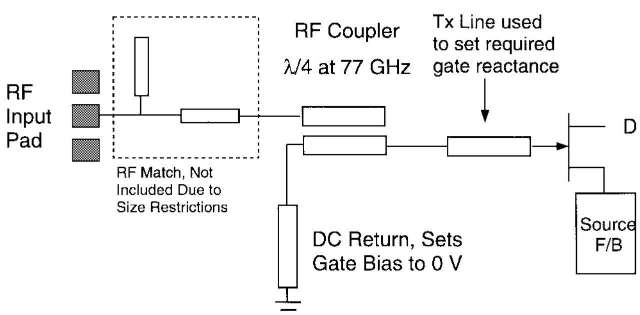

Fig. 5. Schematic of the gate feedback and RF input arrangement.

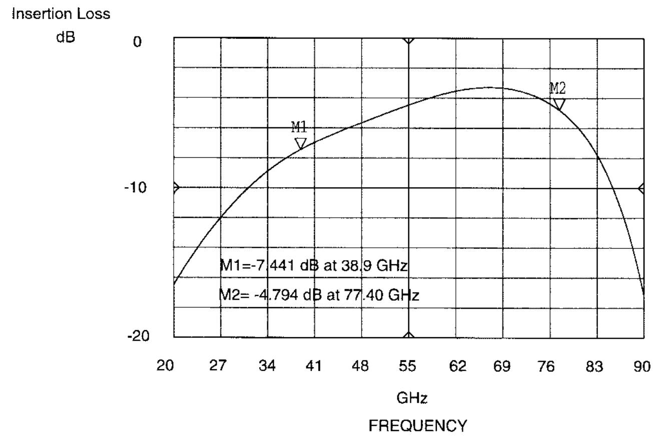

Fig. 6. Simulated performance of the gate feedback circuit.

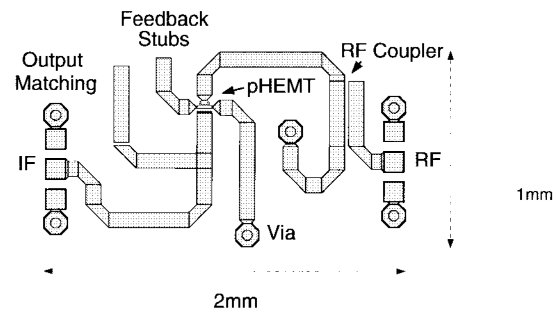

Fig. 7. Schematic of MMIC layout.

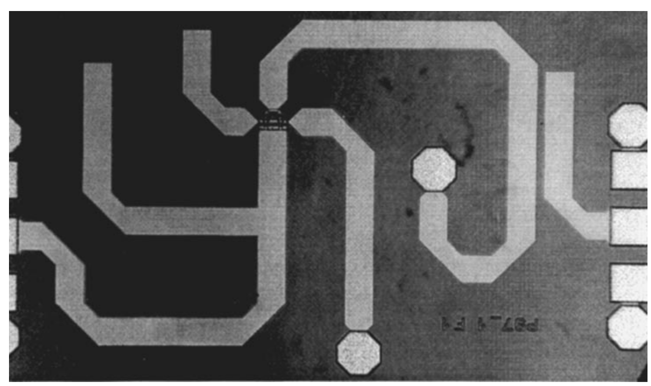

Fig. 8. Photograph of MMIC layout. Dimensions: 1 mm $\times$ 2 mm.

This match produces the maximum power at the fundamental free-running oscillation [7].

Due to size restrictions, the RF filter was reduced to a single quarter-wave coupled line. This section has a higher loss than a multiple section coupled-line filter, but occupies much less space on the MMIC. The simulated loss of this section is 5.3 dB at 77 GHz. For the same reason, it was not possible to include RF matching on the MMIC, resulting in an RF mismatch, which contributes approximately 2–3-dB loss to the mixer.

III. Simulation and Measurements

The mixer was simulated using the HP MDS two-tone oscillator harmonic-balance analysis. The models used for the passive and active elements were provided by the fabrication foundry (PML, France). These models have proven to be very accurate up to 60 GHz; beyond this, the frequency accuracy reduces.

All connections to the chip are made using coplanar probes. The RF is applied through a W-band air coplanar probe with 1.2-dB loss, and the intermediate frequency (IF) is extracted using a 0–50-GHz coplanar probe, which has 1-dB loss. Bias is applied to the drain of the device using an external bias tee. No bias network is required for the gate since the maximum transconductance occurs at $V_{\rm GS} = 0$ V, i.e., self-biased. A short-circuited stub provides a dc return path and, therefore, the correct bias of 0 V. The simulated insertion loss of the gate circuit is shown in Fig. 6. The schematic of the MMIC is shown in Fig. 7 and a photograph of the chip is shown in Fig. 8. The measurement arrangements are shown in Fig. 9. An external low-pass filter was used to suppress mixing

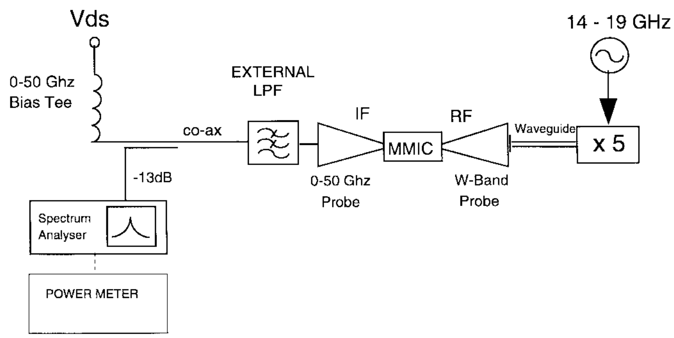

Fig. 9. Schematic of measurement arrangement.

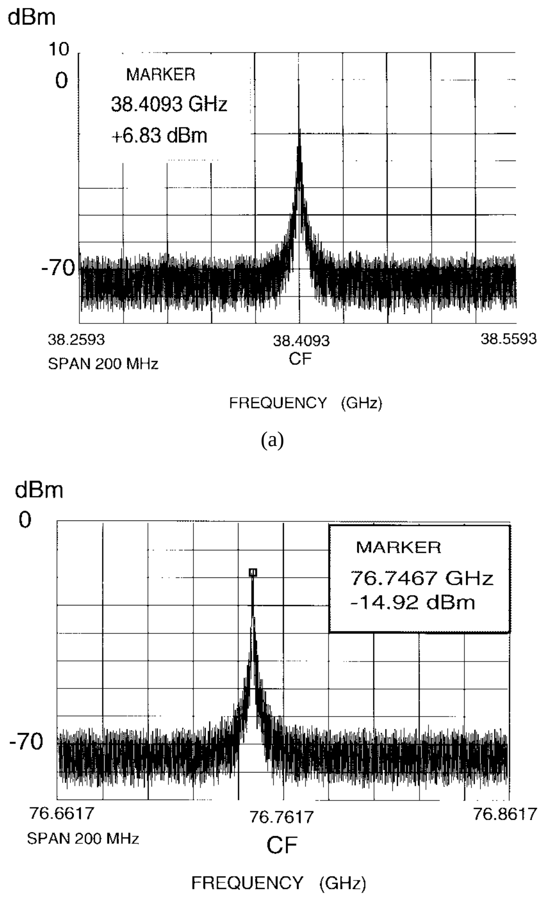

Fig. 10. (a) Self oscillation spectrum. (b) Second harmonic spectrum.

products other than the IF. Removal of this filter enabled LO power and frequency measurements to be made. A wide-band (2–50 GHz) 13-dB coupler was used to extract the IF, which was measured using a spectrum analyzer. The coupler's loss characteristics were measured and were subtracted from IF power measurements.

Fig. 10 shows the fundamental free-running oscillation spectrum and its second harmonic. This measurement was taken from the drain of the device. The oscillator was designed to operate at 39 GHz with a power of 6 dBm, and the measured frequency and power was 38.4 GHz and 6.2 dBm, respectively. The phase noise of the fundamental signal is -76 dBc/Hz at 100 KHz from the carrier.

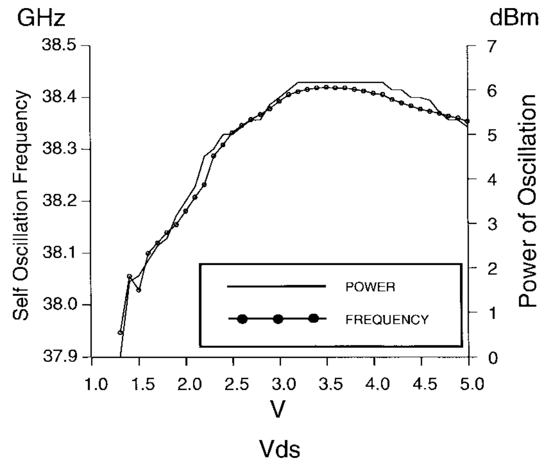

Fig. 11. Frequency tuning and fundamental power variation with respect to drain bias.

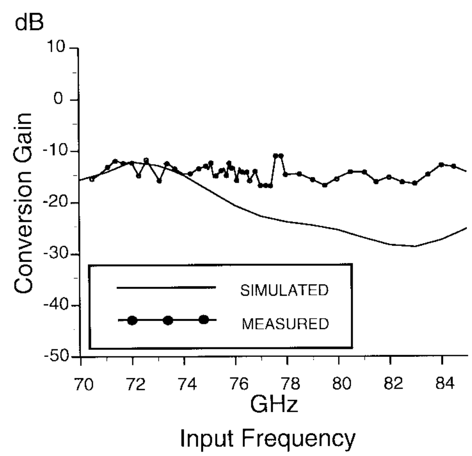

Fig. 12. Measured and simulated conversion loss of mixer.

The free-running oscillation frequency could be tuned over a 0.85-GHz range by varying the drain voltage. The drain–voltage tuning also caused a variation in the power of the self-oscillation, as can be seen in Fig. 11. This power variation is not a problem for this design since the oscillator is designed to operate at a fixed frequency. To achieve a tunable LO with constant power over the tuning range, a varactor diode could be placed on the source or on the output matching network [8]. As expected, the tuning range is increased to 1.7 GHz for the second harmonic.

A signal was then applied to the RF port and conversion-loss measurements were taken. The mixer exhibits very wide-band performance. This is due to the lack of an RF matching circuit. With such a circuit included, the spot frequency performance would be enhanced, but the bandwidth would be reduced. Wide-band performance can also be attributed to the fact that the RF filter has a very wide bandwidth. The simulated performance of this filter is shown in Fig. 6. As can be seen, the insertion loss of the filter is high and the rejection at 38 GHz is poor. This means that the RF–LO isolation is relatively poor at 10 dB. More MMIC space would be required to improve RF–LO isolation and to enable a full RF bandpass filter to be realized. Such a filter could provide adequate isolation because the RF and LO frequencies are distinctly different. The tradeoff for this isolation would be a limitation in bandwidth to the bandwidth of the RF filter used.

The simulated and measured conversion losses are shown in Fig. 12. Very good agreement is obtained from 70 to 74 GHz. The differences at higher frequencies arise from the upper frequency limitations of the passive and active models used. The models show excellent agreement up to 60 GHz, but accuracy reduces beyond this frequency. The mixer has been operated up to 95 GHz, where it exhibits a conversion loss of 28 dB. These measurements do not take into account the loss incurred from the use of a single coupled line for the RF filter. The loss of this coupler could not be measured since it is embedded in the circuit. This coupler could introduce as much as 5.3-dB loss to the circuit. Improvements in this filter could improve conversion performance by 2–3 dB.

IV. Conclusion

A novel MMIC self-oscillating mixer which uses a single pHEMT to achieve mixing and doubling has shown a broadband measured conversion loss of 15 dB from 70 to 85 GHz. The total circuit size is 1 mm $\times$ 2mm, including probing pads. The circuit design was based on a new design approach, which enables mixing and oscillation to occur at two separate frequencies. The design incorporates a novel feedback network, which allows the generation of a subharmonic self-oscillation and retains an RF impedance, which can be matched with a passive network. The measured results exhibit excellent agreement with simulated results. The free-running oscillation displays a tuning range of 1.7 GHz, which compares well with the 1.8-GHz predicted range. At 77.6 GHz, the mixer exhibits a measured conversion loss of 11.0 dB. This mixer has the potential to make a significant contribution to lowering the complexity and cost of 77-GHz systems.

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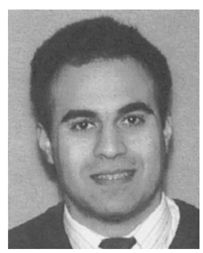

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