Ka-Band and MMIC pHEMT-Based VCO's with Low Phase-Noise Properties

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***Abstract*— Two pseudomorphic high electron-mobility transistor (pHEMT)-based *Ka*-Band voltage-controlled oscillators (VCO's), which have exhibited novel close-to-carrier phase-noise properties in conjunction with output powers greater than previously reported heterojunction bipolar transistor (HBT)-based oscillators, are presented in this paper. Good low phase noises of at least −70 and −75 dBc/Hz at an offset of 100 kHz around 38 GHz have been measured for the two different VCO designs over reasonable frequency tuning ranges with flat or linear output-power tuning in these ranges. Both designs show a strong dependence between phase noise and tuning-element bias conditions.**

I. Introduction

DIELECTRIC-RESONATOR oscillators offer excellent performance in terms of close-to-carrier phase noise [1]–[3]. However, they are not fully monolithic and the circuits still require careful post-fabrication attention. This is to position the dielectric puck onto the main substrate or onto a second adjacent substrate [4]. The demanding factors of cost, size, reliability, and repeatability made by the developing collision-avoidance radar market still point toward a *fully* monolithic solution to the problem.

Monolithic-microwave integrated-circuit (MMIC) designs for millimeter-wave applications may be based on either MESFET, heterojuction bipolar transitor (HBT), or high electron-mobility transistor (HEMT) technology. Guttich and Klassen *et al.* [5] reported work on a 38-GHz MESFET-based voltage-controlled oscillator (VCO) with an integrated buffer amplifier. A tuning range of 700 MHz was achieved in conjunction with an output power in excess of 18 dBm across the tuning band. However, there is no indication as to the close-to-carrier phase-noise performance of the system. The use of HBT's as the basis of millimeter-wave oscillators has been given much attention recently. HBT-based oscillators have exhibited better close-to-carrier phase-noise performance than HEMT-based oscillators [6]–[9]. This is mainly due to the superior low-frequency noise characteristics of HBT's [10]. However, the output power available from HBT-based oscillators cannot compete with the HEMT-based oscillators [6]–[9]. Investigations into HBT-based VCO's by Guttich and Dieudonne *et al.* [11] based upon InGaP/GaAs HBT's have shown how the VCO close-to-carrier phase-noise performance degrades as the base current in the common emitter VCO is increased to deliver the maximum output power of 2.9 dBm at around 35 GHz. The close-to-carrier phase noise is shown to be around −75 dBc/Hz at an offset of 100 kHz for this maximum output-power bias condition. Additionally, this paper reports that the close-to-carrier phase noise is observed to be dependent upon the tuning-element bias. This dependence is noted to be especially strong for applied voltages of less than 1 V. As a result, it is thought that the HBT technology is still not mature enough for full monolithic integration of collision-avoidance-orientated transceiver front-ends. This paper covers two pseudomorphic high electron-mobility transistor (pHEMT)-based *Ka*-band VCO's, which have exhibited good phase-noise properties in conjunction with output powers greater than current HBT-based oscillators. Both oscillator designs show a strong dependence between phase noise and tuning-element bias conditions.

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II. Tuning-Diode Configuration

The configuration of the tuning diode in the circuit has to be understood for the graphical observations to be correctly interpreted. The bias voltage for the tuning diode is applied to the cathode terminal while the anode terminal has been dc grounded through a via hole. Therefore, when the applied voltage is negative in the graphs, the diode is operating in forward bias, and when the applied voltage is positive, the diode is operating in reverse bias.

III. MMIC Design and Fabrication

Both oscillators are based upon a common source: series feedback topology. The active device is a 4 × 15 μm GaAs pHEMT with a gate length of 0.2 μm and an f_T of 62 GHz. All fabrication was carried out by Phillips Microwave Limeil, France, on 100-μm substrates. To introduce instability, the single-diode-tuned VCO has identical 50-Ω transmission lines stubs shorted to ground using via holes. The gate section of the circuit incorporates a 50-Ω transmission line to establish the required negative conductance at the drain terminal of the active device, a dc blocking capacitor to separate the varactor, and active-device power supplies and an interdigitated varactor diode for tuning purposes. The diode is biased through a radial stub and high-impedance line, and the active device is dc grounded via a high-impedance line. The diode has a low series resistance due to its interdigital design and it provides a larger tuning range than using a drain–source shorted pHEMT as the

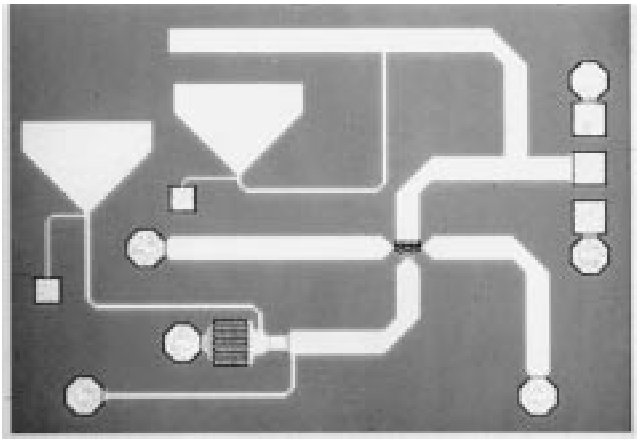

Fig. 1. Photograph of the single-diode-tuned VCO.

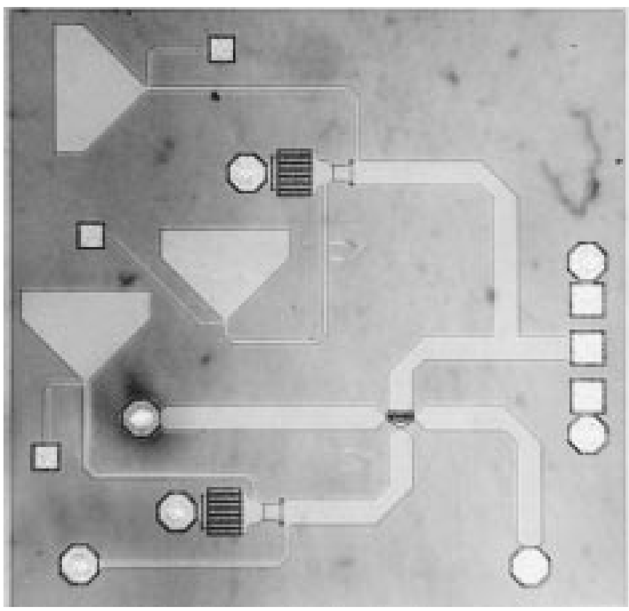

Fig. 2. Photograph of the double-diode-tuned VCO.

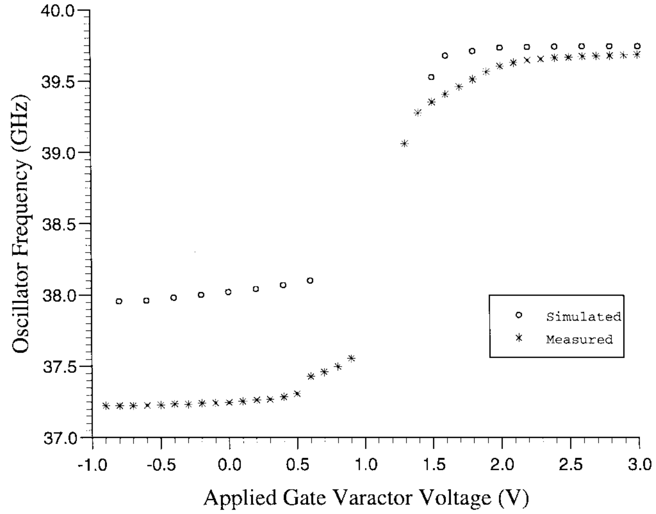

Fig. 3. Measured and nonlinear simulated frequency tuning with applied varactor bias of the single-diode-tuned oscillator.

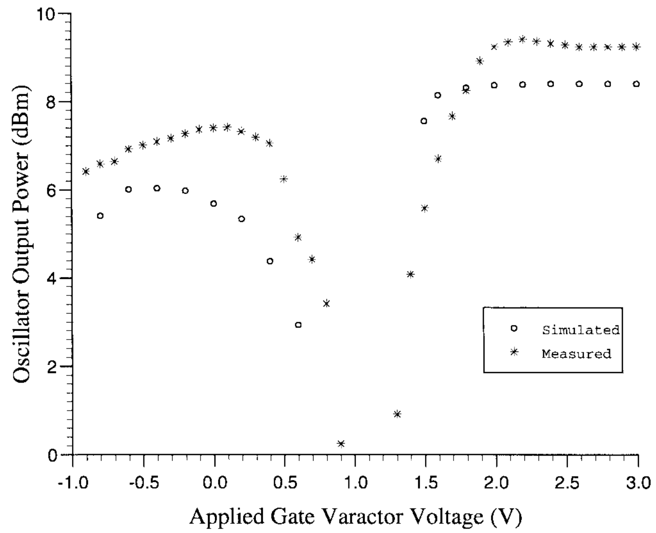

Fig. 4. Measured and nonlinear simulated power tuning with applied varactor bias of the single-diode-tuned oscillator.

tuning element. The output of the oscillator employs a single open-circuit stub-matching network initially determined using small-signal techniques for maximum power transfer and fine tuned using the HP MDS harmonic-balance simulator. The drain bias is applied in a similar fashion as the varactor bias. A full description of the design procedure has been reported elsewhere [12]. A photograph of the single-diode-tuned VCO is shown in Fig. 1. The chip size is 2.0×1.6 mm^2. The pHEMT is situated to the right of center with the drain terminal at the top and the gate terminal at the bottom. The tuning diode is in the bottom left area and the various bias supplies are applied to the pads extending from the radial stubs. The output is via the three pads on the right into an appropriate coplanar-waveguide probing station. The double-diode-tuned VCO has the same source stub and gate resonator sections as the single-diode-tuned VCO. The matching network involves the use of a second tuning diode, bias network, and appropriate dc block instead of a simple open-circuit stub. A photograph of the double-diode-tuned VCO is shown in Fig. 2. The chip size is 2.0×2.0 mm^2. The pHEMT is in the bottom-right quarter with the drain at the top and the gate at the bottom. Again, the gate tuning diode is in the bottom left. The matching-network tuning diode is in the top center. All the bias supplies are again applied to the appropriate pads extending from the radial stubs.

IV. Measurements and Discussion

A. Single-Diode-Tuned VCO

Fig. 3 shows the measured and nonlinear simulated frequency tuning curve for the single-diode-tuned oscillator, and Fig. 4 displays the associated measured and nonlinear simulated power tuning for the same circuit. The power dropout is due to the steady-state oscillation condition $B_D(A_o, \omega_o) + B_L(\omega_o) = 0$ no longer being satisfied, where B_D is the synthesized one-port device susceptance, and is a function of oscillation amplitude A, and oscillation frequency ω and B_L

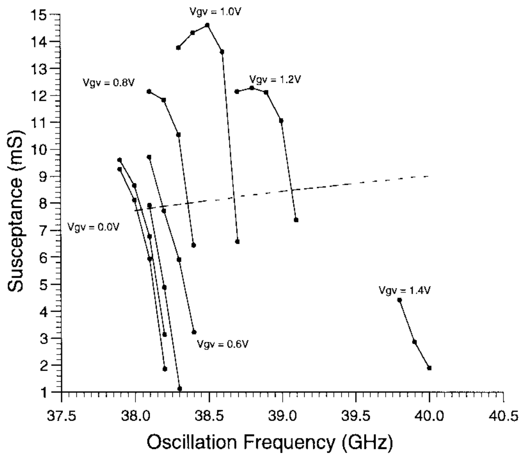

Fig. 5. Synthesized one-port device small-signal susceptance as a function of applied gate varactor voltage and associated load small-signal susceptance versus frequency.

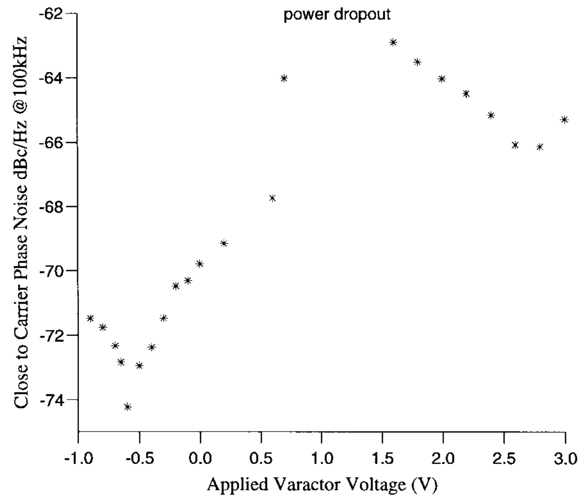

Fig. 6. Measured close-to-carrier phase noise at 100-kHz offset with applied varactor bias of the single-diode-tuned oscillator.

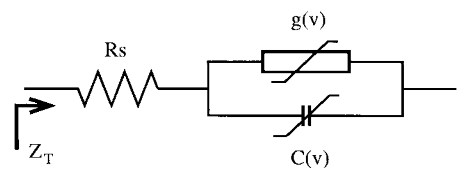

Fig. 7. Simple equivalent circuit of varactor tuning diode.

is the total susceptance looking out from the synthesized one-port device, and is a function of oscillation frequency only. The subscript o denotes the steady-state values for these elements. Fig. 5 shows the synthesized one-port device small-signal susceptance (solid lines) and the magnitude of the total-load small-signal susceptance (dotted lines) of the oscillator against frequency. The device susceptance is represented as a function of the applied gate varactor voltage. The dots on the device susceptance lines indicate the linear simulation points, which are spaced 0.1 GHz apart on each line. The corresponding device conductance at these simulation points indicate that oscillations are possible, as they satisfy the condition for the start up of oscillations, i.e., $-G_D(A,\omega) > G_L(\omega)$. The predicted frequency of oscillation is where the device susceptance line and load susceptance line intersect. As the applied gate varactor voltage is tuned, the linear small-signal simulation shows how the oscillation frequency moves. Initially, when the applied varactor voltage is low, the predicted change in the oscillation frequency is small. As the applied varactor voltage increases, the rate of change of frequency increases until (between the applied varactor bias of 1.2 and 1.4 V) the device susceptance line and load susceptance line no longer intersect. This corresponds with the power dropout predicted by the nonlinear simulations, and is observed in the measured results of Fig. 4.

The good correlation between measured and nonlinear simulated results allows the designer to have good confidence in the device models and in the design procedure employed. Clearly, there are two areas which may be of interest depending on the application. For average power output and a low-frequency tuning range, the diode should be operated in forward and low reverse bias operation (0.6–$-$0.4 V). Giving an output power of $+7.2$ dBm $\pm$ 0.4 dBm with a 50-MHz/V tuning range centered upon 37.250 GHz. For a slightly higher power requirement with a larger tuning range and also without the need to switch the polarity of the tuning voltage, the higher reverse-bias tuning (-1.8–-2.8 V) should be used. This gives $+8.8$ dBm $\pm$ 0.6 dBm, with a 150-MHz/V tuning range centered upon 39.600 GHz.

The measurements shown in Fig. 6 show how the close-to-carrier phase noise at a 100-kHz offset varies with the varactor bias. The measurements were recorded from an HP8562A Spectrum Analyzer and associated *Ka*-band external mixer. This variation exhibits a significant sensitivity around the area where the bias applied to the varactor tuning diode causes the diode to be just turned on in forward conduction. The peak of this sensitivity is at the point where current just begins to flow in the varactor. To the authors' knowledge, this type of behavior has not been reported for any pHEMT-based VCO using an interdigital topology for the tuning element. Apart from possibly a very small increase as the diode is pushed further into forward bias, the series resistance of the tuning diode may be assumed to be a fixed value [13]. This assumption is supported by application notes on the foundry model, which states, for small dc currents, the access resistance (R_s) is constant.[1] Consider the simple equivalent circuit of the diode, as shown in Fig. 7. The real part of the input impedance (R_T) is characterized by

$$\Re e\{Z_T\} = R_T(V) = R_S + \frac{g(V)}{g^2(V) + \omega^2 C^2(V)}. \quad (1)$$

Due to the magnitudes of the elements involved, one may assume that $\omega^2 C^2(V) \approx 0$, leaving

$$R_T(V) = R_S + \frac{1}{g(V)}. \quad (2)$$

As $I = (V/R_T)$ and R_S is essentially a constant with applied voltage, $g(V)$ will increase nonlinearly with an increase

[1] *DO2AH Design Manual*, vol. 1.1, Phillips Microwave Limeil, France, 1996.

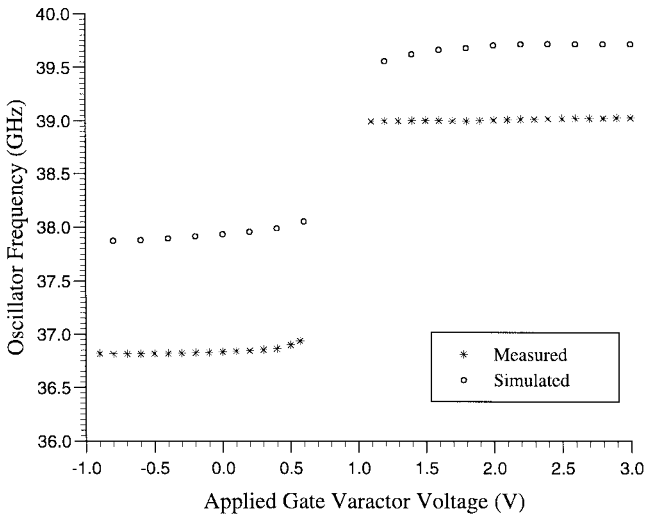

Fig. 8. Measured and nonlinear simulated frequency tuning of the double-diode-tuned VCO. The applied matching-network varactor voltage is constant at 2.1 V for reverse-bias operation.

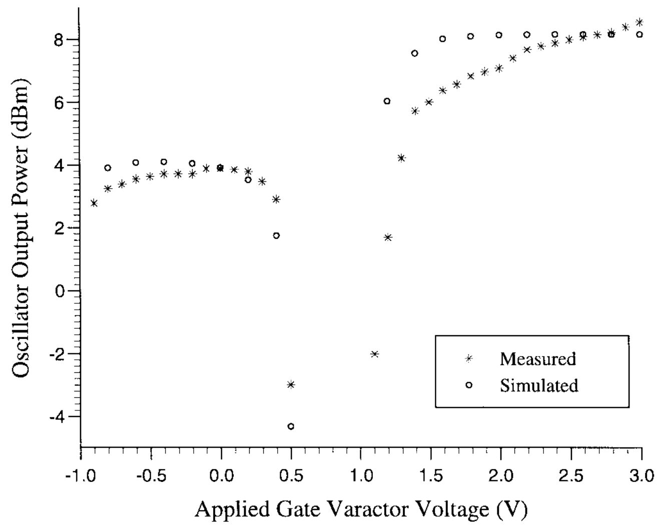

Fig. 9. Measured and nonlinear simulated power tuning of the double-diode-tuned VCO. The applied matching-network varactor voltage is constant at 2.1 V for reverse-bias operation.

in forward bias. This result is expected as where $g(V) = A.I(V)$ is a constant, as shown by Maas [13]. Subsequently, $R_T(V)$ would decrease nonlinearly with increasing applied voltage, but never drop below R_S. If the Q factor of the diode is governed by $Q \propto (1/R_T(V)C(V))$, the improvement of the phase noise in forward bias could be explained by the characteristics of R_T and $C(V)$. By developing a minima in the relationship $R_T(V)C(V)$ at the diode turn-on voltage, a peak in the Q of the diode and the associated resonator section in which it is embedded results. This could lead to the reduction seen in the close-to-carrier phase noise at the diode turn-on point when operated in forward bias. From these measurements, it would seem that there is a tradeoff between the output power, tuning range, and phase noise. Such that when the phase noise is improved, the output power and tuning range decrease.

B. Double-Diode-Tuned VCO

The measured and nonlinear simulated frequency tuning of the double-diode-tuned VCO with variation of the applied gate resonator-varactor voltage and with the matching-network varactor operating in reverse bias at an applied voltage of 2.1 V is shown in Fig. 8. The variation of the output frequency with change in the applied matching-network varactor voltage is small compared to the change with the gate varactor voltage. For clarity, only the measurements with the applied matching-network varactor voltage of 2.1 V are shown. The measured and nonlinear simulated power tuning with applied gate varactor voltage of the double-diode-tuned VCO is shown in Fig. 9. Again, for clarity, only the measurements recorded for the applied matching-network varactor voltage of 2.1 V are shown. As with the single-diode case, the power dropout may be explained by the susceptance condition for steady-state oscillations no longer being satisfied as the gate varactor voltage is tuned. Again, there are clearly two areas which may be of interest depending upon the application. For low-power output and an average-frequency tuning range, the

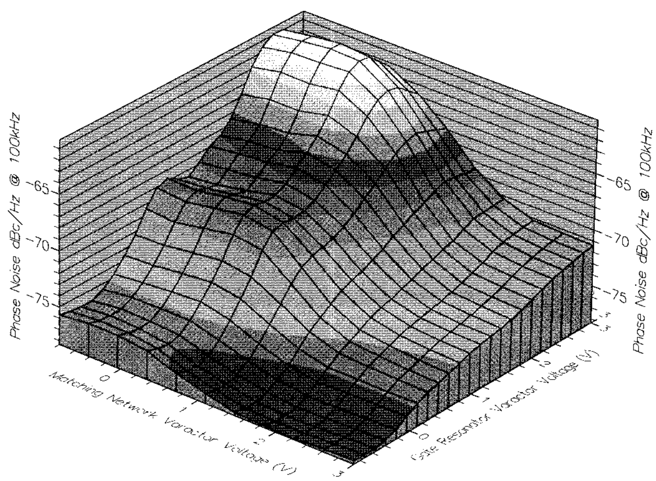

Fig. 10. Contour of measured close-to-carrier phase noise with applied varactor bias of the double-diode-tuned oscillator.

diode should be operated in forward and low reverse-bias conditions (−0.6–0.4 V applied). Giving a flat output power of +3.7 dBm ± 0.2 dBm with a 49-MHz/V tuning range centered upon 36.840 GHz. For a higher power requirement and also without the need to switch the polarity of the tuning voltage, the higher reverse bias tuning (1.8–2.8 V applied) should be employed. This gives a linear power characteristic of +7.5 dBm ± 0.6 dBm, with a 30-MHz/V tuning range centered upon 39.020 GHz. The measured results are again in good agreement with the simulations. Fig. 10 shows the measured variations of the oscillator close-to-carrier phase noise with the change in bias applied to both the gate resonator-varactor diode and the matching-network varactor diode for the double-diode-tuned VCO. It may be seen that the phase noise is strongly dependent upon the bias conditions. The best operating regime (in terms of lowest phase noise) is when the gate resonator-varactor diode is operated in forward bias—as was shown earlier in the case of the single-diode-tuned VCO. Again, the characteristic minima of the close-to-carrier phase noise was observed at the point where the diode is just beginning to conduct in forward-bias operation. The matching-network varactor bias also strongly effects the measured phase noise. This could be due to the circuit locus and device line intersection angle moving closer to the optimum angle for lowest phase-noise operation $(\pi/2)$ as the diode is tuned. The best results are obtained when this diode is operated in reverse bias with an applied voltage in excess of 1.8 V to guarantee a close-to-carrier phase noise of at least -70 dBc/Hz at an offset of 100 kHz for the whole gate resonator-varactor tuning range. Utilizing the combination of a forward-biased gate varactor and a reverse-biased matching-network varactor, the resultant measured close-to-carrier phase noise of at least -75 dBc/Hz at an offset of 100 kHz is comparable with some reported HBT-based MMIC oscillators in this frequency band [6], [11].

V. Conclusions

Two *Ka*-band pHEMT-based MMIC VCO's have been developed using GaAs technology. The first, which incorporates a single varactor diode as the tuning element, exhibited a previously unreported sensitivity of the close-to-carrier phase noise with the diode operating in forward bias. The second design employs a novel second varactor diode to replace the open-circuit stub in the output matching network. The measured results revealed a similar close-to-carrier phase-noise sensitivity with tuning-element bias and exhibited the good low phase-noise properties for a pHEMT-based MMIC VCO of -75 dBc/Hz around 38 GHz at an offset of 100 kHz. Both diodes bias voltages could be altered for tuning purposes.

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