Compact InP-Based HBT VCOs with a Wide Tuning Range at W- and D-Band

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***Abstract*—Compact monolithic integrated differential voltage-controlled oscillators (VCOs) operating in W-band were realized using InP-based heterojunction bipolar transistors (HBTs). The oscillators, with a total chip size of 0.6 by 0.35 mm^2, are based on a balanced Colpitts-type topology with a coplanar transmission-line resonator. By varying the voltage across the base–collector junction of the HBT in the current mirror and by changing the current in the VCO, the oscillation frequency can be tuned between 84 and 106 GHz. At 100 GHz, a differential voltage swing of 400 mV is obtained, which should be sufficient to drive 100-Gb/s digital logic. By combining the balanced outputs of a similar differential VCO in a push–push configuration, a compact source with close to −10 dBm output power and a tuning range between 138 and 150 GHz is obtained.**

***Index Terms*—Heterojunction bipolar transistors, indium compounds, millimeter-wave bipolar transistor oscillators, monolithic microwave integrated circuits (MMICs), voltage-controlled oscillators (VCOs).**

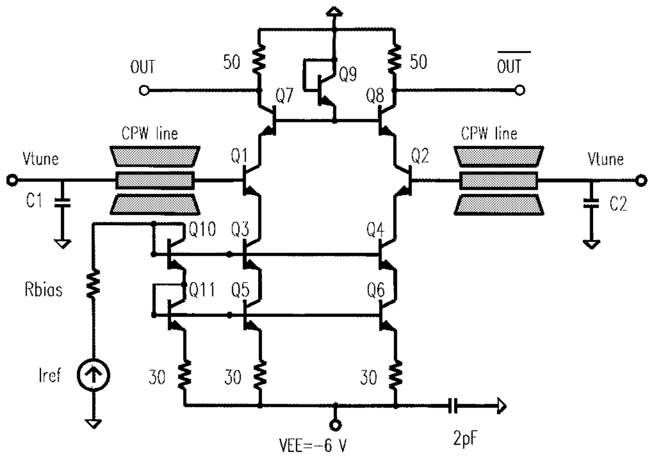

Fig. 1. Circuit diagram of the differential W-band oscillator.

I. Introduction

DUE TO the ever increasing demand for bandwidth, next-generation lightwave systems will operate at data rates of 40, 80, 100 Gb/s or higher. For this type of application and at these high speeds, it is advantageous to integrate both the digital functions such as demultiplexers and frequency dividers and the analog circuits such as clock sources and amplifiers on a single chip. High-speed digital circuits normally use a differential topology; therefore, such integrated systems require as a clock source a millimeter-wave oscillator with balanced output.

Most integrated circuits operating at 60 GHz and above are based on high electron mobility transistors (HEMTs), which allow a high frequency of operation due to their high maximum oscillation frequency $f_{\max}$ [1], [2]. While heterojunction bipolar transistors (HBTs) usually have a lower $f_{\max}$ and provide lower output power at millimeter-wave frequencies, their reduced low-frequency noise makes them attractive for low-phase-noise oscillators. This was demonstrated by Wang *et al.* [3], who reported a fully integrated 62-GHz fundamental oscillator with a phase noise of −104 dBc/Hz at 1-MHz offset. The highest fundamental frequency HBT oscillators reported in the literature so far are operating at 104 and 134 GHz [4] and are based on InGaP/InGaAs HBTs with an $f_{\max}$ of 170 GHz.

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All the reported HEMT and HBT oscillators operating at higher millimeter-wave frequencies are based on circuit topologies using reactive transmission-line matching, resulting in a relatively large chip area. Furthermore, the tuning range of reported oscillators is limited, making it difficult to account for process and temperature variations. In this paper, we present an oscillator design that combines analog and microwave design techniques. As a result, we obtain a fully integrated W-band source with a very compact and uniplanar layout and with the balanced output needed to drive digital logic blocks.

Additionally, when an even higher frequency is needed, the balanced oscillator can be used in a push–push configuration providing an output at twice the frequency [5]. A 108-GHz InP-based HBT push–push source with 2.7-GHz tuning range was reported by Kobayashi *et al.* [6]. In this paper, we demonstrate that our differential voltage-controlled oscillators (VCOs) can be used to realize a monolithic bipolar source with close to −10 dBm output power and with a tuning range between 138 and 150 GHz.

II. Circuit Design

The circuit diagram of the differential W-band oscillator is shown in Fig. 1. It is based on a balanced Colpitts-type topology, which uses a coplanar transmission line as the inductive resonator element. The base–collector junction of the transistors in the current mirrors acts both as the destabilizing element and as the varactor. By changing the tuning voltage $V_{\rm tune}$ on the bases of HBTs Q1 and Q2, the voltage across the base–collector junction of HBTs Q3 and Q4 in the current mirror is changed. This results in a variation of the base–collector capacitances of Q3 and Q4 ($C_{BC_{Q3}}$ and $C_{BC_{Q4}}$) with a factor of about two (from 18 to 40 fF). These variable differential capacitances provide a way to tune the oscillator frequency without having to use an additional varactor circuit. Taking C1 and C2 in Fig. 1 sufficiently large (C1 and C2 $\gg C_{BE_{Q1}}$ and $C_{BC_{Q3}}$) and with the electrical length of the coplanar resonator well below a quarter wavelength, the oscillation frequency of the VCO can be approximated by the expression for an LC-type Colpitts oscillator

$$f_{\rm VCO} \cong \frac{1}{2 \cdot \Pi \cdot \sqrt{L_{\rm CPW} \cdot \left(\frac{C_{BE_{Q1}} \cdot C_{BC_{Q3}}}{C_{BE_{Q1}} + C_{BC_{Q3}}} \right)}} \quad (1)$$

with $L_{\rm CPW}$ as the effective inductance of the coplanar line and with the second term under the square root reflecting the series connection of the base–emitter capacitance of Q1 ($C_{BE_{Q1}}$) and the base–collector capacitance of Q3 ($C_{BE_{Q3}}$).

By making the connection of the bases of Q3 and Q4 a higher impedance node using a cascode current mirror (Q3–Q5 and Q4–Q6) and a relatively large bias resistor $R_{\rm bias}$, the common mode capacitance at the emitter of Q1 and Q2 is made smaller than the differential capacitance. In this way, only a differential oscillation and no common-mode oscillation can start up.

The cascode transistors Q7 and Q8 provide extra isolation between the oscillator core and the load. While millimeter-wave oscillators are normally matched to the 50-Ω output load by a reactive matching network, in this design we chose to have an on-chip 50-Ω resistive match. This resulted in a more compact layout and a process-tolerant matching at the expense of a lower output power since half of the available radio-frequency (RF) current is dissipated in the on-chip 50-Ω termination resistors. When integrating the VCO together with digital circuitry, this resistor can be removed, increasing the differential voltage swing with a factor of almost two.

By using a differential topology, the area occupied by biasing networks is significantly reduced, as bias can be provided at the virtual ground node. Only three metal–insulator–metal capacitors are needed to ac ground the tuning voltage ($V_{\rm tune}$) and the single -6 V negative bias supply $V_{\rm EE}$. The current flowing through the devices is controlled by an external current source $I_{\rm ref}$ through the bias resistor and the cascode current mirror Q10 and Q11.

A chip microphotograph of the oscillator is shown in Fig. 2. The layout is done in a fully symmetric way using coplanar transmission lines. This coplanar topology was also chosen to improve the shielding of the oscillator from its digital surroundings. The total chip size of the oscillator is 0.35 by 0.55 mm^2, including dc and RF probe pads. The size of the oscillator core is only 0.15 by 0.5 mm^2, making it easy to integrate this chip with digital circuitry. Providing a reactive matching for the cascode output would more than double the size of the oscillator core.

The circuit diagram of the D-band push–push oscillator is shown in Fig. 3. It consists of a differential oscillator similar to the W-band oscillator and a Wilkinson coupler output combiner. Due to the lower maximum oscillation frequency $f_{\rm max}$ of the HBT technology used to realize this design and due to the need for high harmonic content and hence high oscillator loop gain, the differential oscillator was designed for a slightly lower frequency of about 70–75 GHz. Furthermore, the oscillator core was designed for maximum second-harmonic generation. This

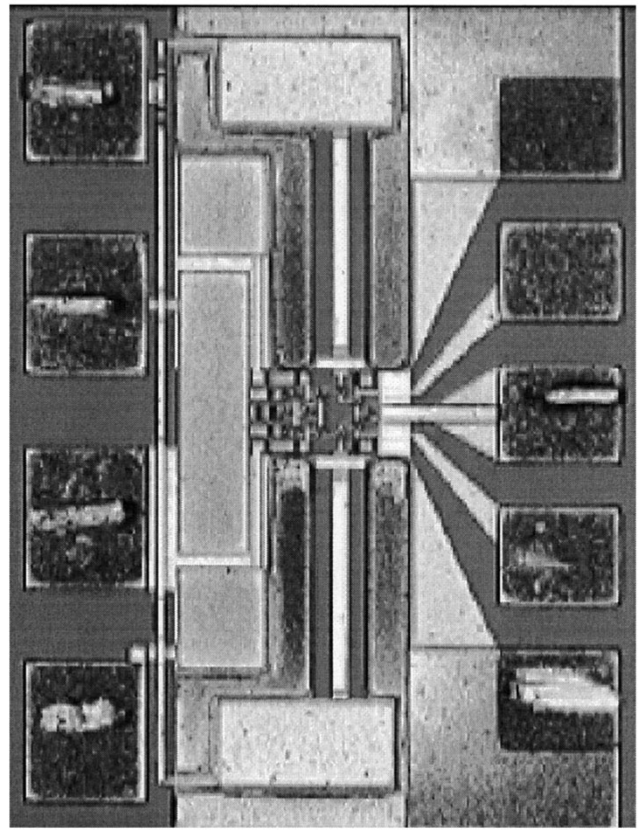

Fig. 2. Microphotograph of the differential W-band oscillator.

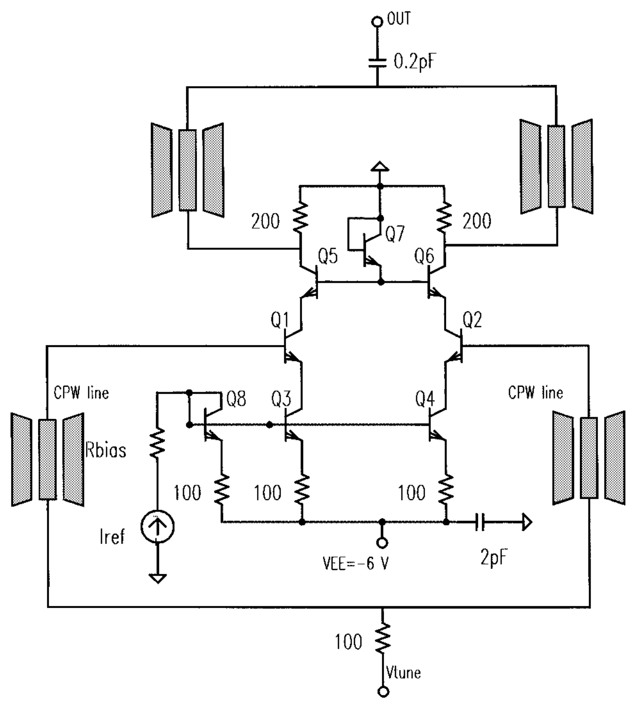

Fig. 3. Circuit diagram of the D-band push–push VCO.

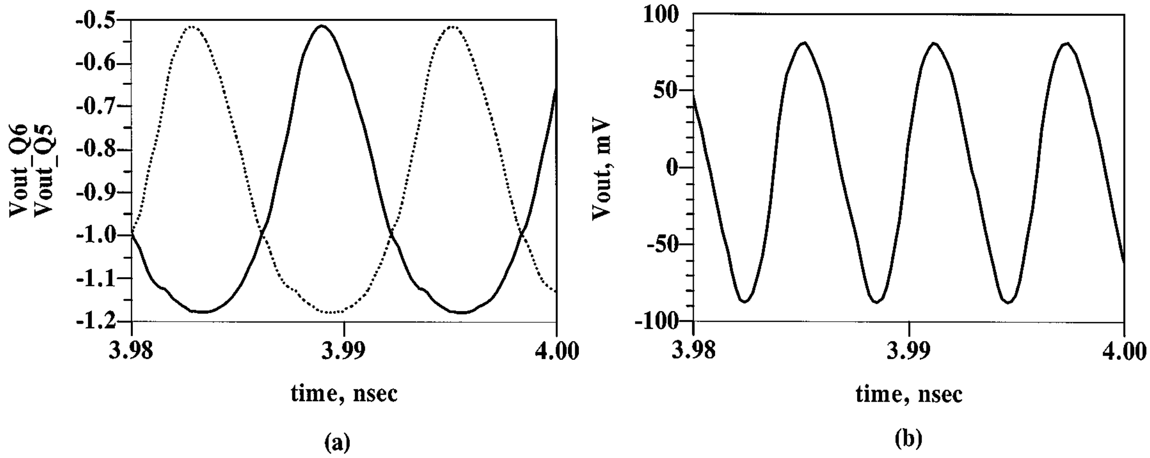

Fig. 4. Waveforms simulated for the D-band push–push VCO: (a) clipped signal at the two internal balanced outputs before the combiner and (b) waveform at double the frequency after the combiner.

was achieved by making the coplanar resonator between the two negative resistances continuous and by using an internal load resistance of 200 Ω instead of 50 Ω. The biasing through a higher resistance together with the large voltage swing causes signal clipping at the two balanced outputs due to forward conduction of the base–collector diodes of Q5 and Q6. As shown in Fig. 4(a), due to this clipping, a strong even harmonic content can be generated.

The two differential signals are summed by a power combiner that is similar to a Wilkinson coupler. This coupler is designed to operate at the fundamental frequency, so the length of the line is a quarter-wavelength length at 70 GHz. While the output node of the combiner is seen as a short circuit for the differential mode of the oscillator core, the quarter-wavelength will transform this short circuit to an open circuit at the internal load resistor, making sure that the differential voltage swing at this point stays high enough to generate enough harmonic content. According to simulations, a voltage swing of about 160 mV at double the frequency is obtained after the combiner, as shown in Fig. 4(b).

A chip microphotograph of the oscillator is shown in Fig. 5. Again, the layout is done in a fully symmetric way using coplanar transmission lines. A good symmetry is crucial for a good suppression of the fundamental and odd harmonics at the output of the push–push VCO. The total chip size of the oscillator is 0.5 by 0.7 mm^2, including dc and RF probe pads.

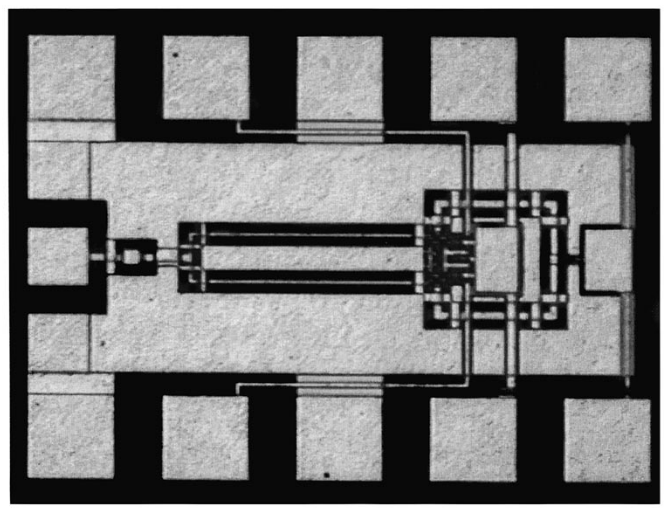

Fig. 5. Microphotograph of the D-band push–push VCO.

III. Device Technology and Performance

The W-band oscillators were realized using an all-optical lithography single-heterojunction InGaAs/InP HBT process, developed at Bell Laboratories, Lucent Technologies [7]. InGaAs/InP bipolar transistors offer the advantages over GaAs/AlGaAs HBTs of a lower turn-on voltage, higher electron mobility, better thermal dissipation, and better microwave performance, while still obtaining a high collector-to-base breakdown voltage. To improve device switching time and reduce power requirements, device dimensions have been continually decreased down to an emitter finger width of 1.2 μm. Obtaining these small device dimensions requires precise control of the fabrication process. We have developed a fabrication procedure for small and reliable InP-based HBTs using a process involving both wet etching and electron cyclotron resonance plasma etching, as described in more detail in [8]. HBTs with nominal 1.2 by 6 μm^2 emitter dimensions, measured on the same wafer as the oscillators, show a maximum cutoff frequency f_T of 150 GHz and a maximum oscillation frequency $f_{\max}$ in excess of 200 GHz. Fig. 6 shows the measured current gain and unilateral transducer power gain for a 1.2 by 6 μm^2 emitter device biased at a collector current of 5 mA, which is similar to the current of the devices in the W-band oscillator and slightly lower then the current for peak f_T. At this current, a transit frequency f_T of 135 GHz and a maximum oscillation frequency $f_{\max}$ of about 230 GHz can be extrapolated from measurements up to 110 GHz. The measured maximum available gain at 100 GHz is 6 dB, which should be sufficient for oscillator application at W-band.

The D-band push–push VCO was realized in an all-optical lithography single-heterojunction AlInAs/InGaAs HBT foundry process, developed and fabricated at HRL Laboratories, Malibu, CA [9]. HBTs with nominal 1 by 3 μm^2 emitter

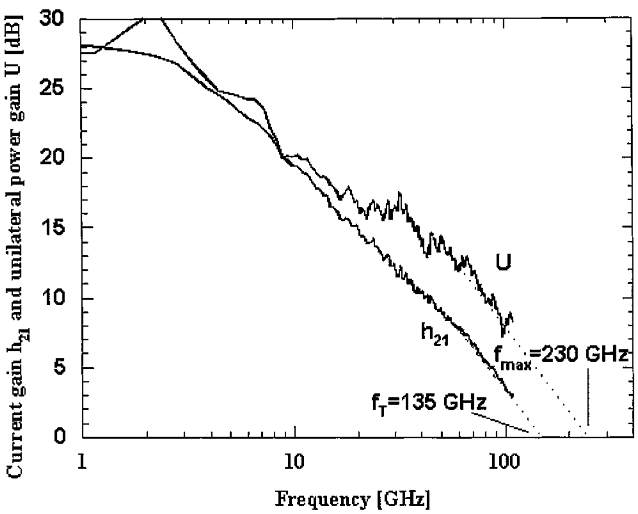

Fig. 6. Measured current gain and unilateral power gain and extrapolated f_T and $f_{\max}$ for a 1.2 by 6 μm^2 emitter single-heterojunction InGaAs/InP HBT biased at a collector current of 5 mA.

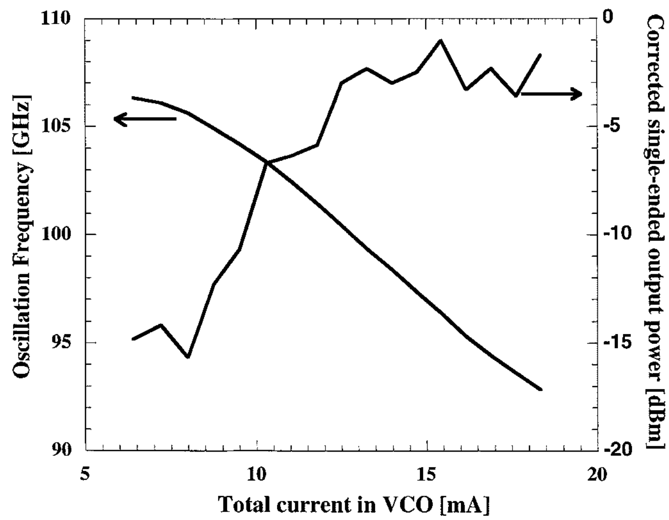

Fig. 8. Measured oscillation frequency and single-ended output power of the differential W-band oscillator as a function of the current in the VCO ($V_{\rm EE} = -6$ V, $V_{\rm tune} = -2.5$).

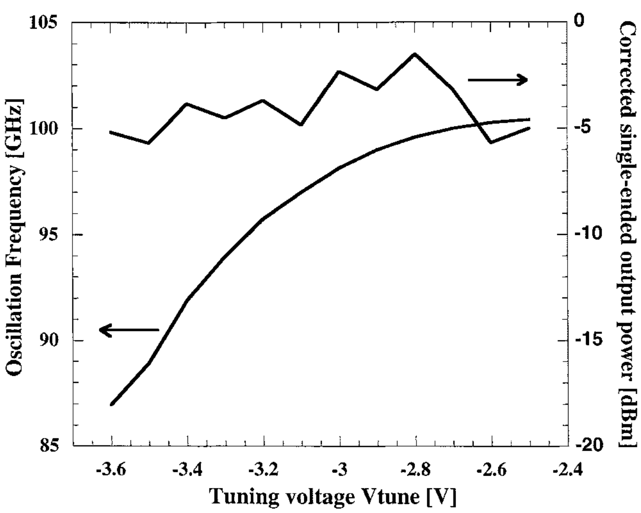

Fig. 7. Measured oscillation frequency and single-ended output power of the differential W-band oscillator as a function of the tuning voltage (total VCO current = 16 mA).

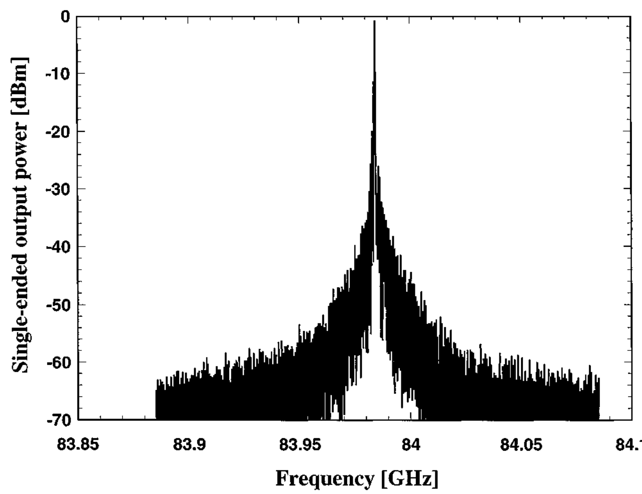

Fig. 9. Detail of the measured spectrum of the W-band VCO tuned for the lowest frequency.

dimensions, measured on the same wafer as the oscillators, show a maximum cutoff frequency f_T of 150 GHz and a maximum oscillation frequency $f_{\max}$ of 170 GHz.

IV. Results and Discussion

Spectral measurements from 75 to 110 GHz were performed using a 50-GHz spectrum analyzer extended in frequency with a Millitech CDA-10 75–100 GHz waveguide block downconverter. On-wafer measurements were performed using a dc to 110 GHz 1-mm coaxial probe connected to the WR-10 waveguide downconverter using a coaxial-to-waveguide transition.

The measured frequency of oscillation and output power versus the VCO tuning voltage is plotted in Fig. 7. A wide tuning range from 87 to 100 GHz is observed. The output power, which has been corrected for the mixer conversion loss and for loss in the probe and cables, is -3 dBm over the frequency range, constant to within our measurement error. For the measurement, a bias voltage $V_{\rm EE}$ of -6 V and a total VCO current of 13 mA was applied. This bias current corresponds to an emitter current density of 80 kA/cm^2 and a total power dissipation of 78 mW.

By changing the current in the VCO, the tuning range can be further increased. As shown in Fig. 8, for a fixed tuning voltage of -2.5 V, the oscillation frequency can be decreased from 106 to 92 GHz by increasing the total current in the VCO from 6 to 18 mA. For the lowest currents, however, a decrease of the output power can be observed. By increasing the VCO current together with decreasing the voltage across the current mirror, a minimum oscillation frequency of 84 GHz can be obtained.

A detail of the measured spectrum of the VCO at this bias condition is shown in Fig. 9. The extremely high frequency and sensitivity to noise introduced by the bias supplies complicates an accurate determination of the phase noise. Depending on tuning frequency, a phase noise between -80 and -85 dBc/Hz at 1-MHz offset from the carrier can be estimated from the measured downconverted spectrum using a spectrum analyzer.

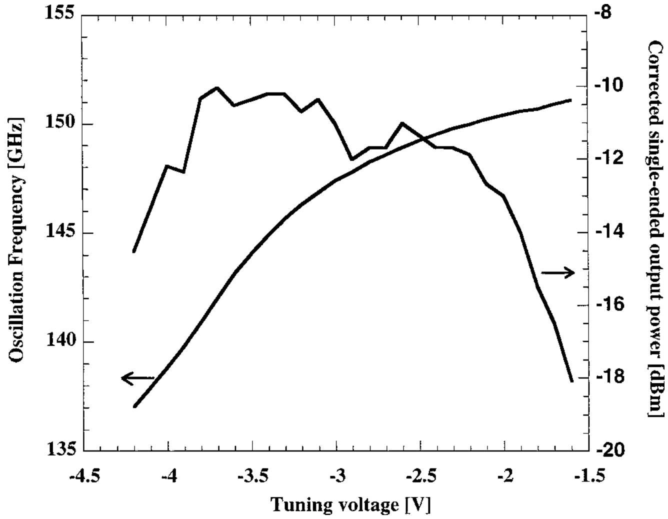

Fig. 10. Measured frequency tuning characteristics of the 150-GHz push–push VCO.

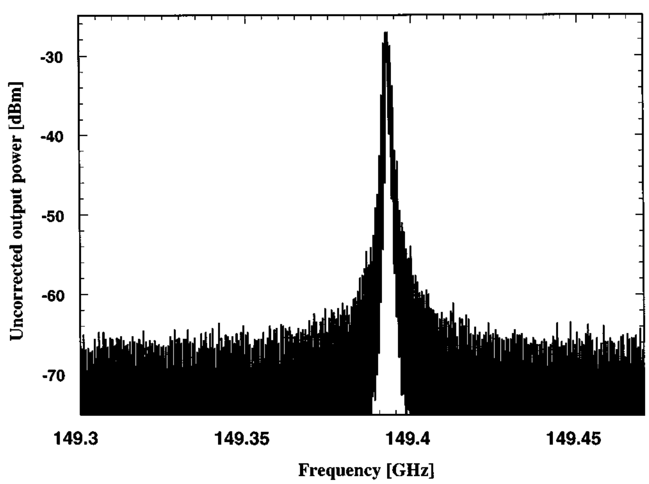

Fig. 11. Detail of the uncorrected measured spectrum of the 150-GHz push–push VCO (spectrum analyzer resolution and video bandwidth = 1 MHz; mixer losses are about 15 dB).

Mainly due to our extended tuning range, this value is about 5 dB higher than that of the lowest phase-noise monolithic W-band VCO (-88 dBc/Hz at 1-MHz offset [6]). The phase noise is better than those reported for HEMT VCOs with similar tuning range. For instance, a W-band HEMT oscillator with a phase noise of -67 dBc/Hz at 1 MHz was reported in [10].

Spectral measurements from 140 to 170 GHz on the D-band push–push VCO were performed on-wafer using a GGB Industries model 220 WR-05 waveguide probe together with a Millitech 140–170 GHz downconverter block, consisting of an active multiplier feeding a second-harmonic mixer. The measured frequency of oscillation versus tuning voltage is plotted in Fig. 10. An oscillation frequency between 137 and 151 GHz is obtained. The measured intermediate-frequency (IF) power from the downconverter is about -25 dBm. Taking a total downconversion loss of the mixer, waveguide probe, and cables of about 15 dB, a maximum output power close to -10 dBm is obtained. Additional spectral measurements in W-band showed that the fundamental frequency component of this second-harmonic oscillator is suppressed below the noise floor of the spectrum analyzer, indicating the good symmetry of a monolithic integration.

A detail of the measured spectrum of the push–push VCO is shown in Fig. 11. Again, the frequency and sensitivity to noise on the bias supplies complicates an accurate determination of the phase noise. From the downconverted IF spectrum, a phase noise of about -75 dBc/Hz at 1-MHz offset from the carrier is measured. This value is higher than that of the fundamental differential VCO due to the frequency doubling and also because the oscillator was designed for nonlinear operation.

V. Conclusions

Differential voltage-controlled oscillators operating up to 106 GHz were realized using InP HBT technology. By using the base–collector junction of the transistors in the current mirror as a varactor, a tuning range of 20% is achieved. This circuit is, to our knowledge, the first fully differential circuit operating in W-band. While using a compact differential topology with no reactive output matching or bias networks, the performance of this VCO in terms of phase noise, output power, and maximum frequency of operation is comparable with the best published monolithic W-band oscillators. This circuit will find application as the clock source in next-generation digital building blocks operating at speeds up to 100 Gb/s.

The differential topology is also used to extend the frequency range of integrated sources by using a push–push topology, as was demonstrated by the realization of a monolithic 136–150 GHz source. This VCO is the highest frequency integrated source based on bipolar technology, to the best of our knowledge. By using an InP-based HBT technology with higher maximum oscillation frequency, a further extension in the sub-millimeter-wave range could be obtained. Such oscillators will find application in ultra-high-speed lightwave and advanced imaging and remote sensing applications.

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