A Si/SiGe HBT Dielectric Resonator Push–Push Oscillator at 58 GHz

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***Abstract*—In this letter, we present a dielectric resonator push–push oscillator at 58 GHz. The microstrip circuit is fabricated in hybrid thin-film technology on a 10 mil Alumina substrate. Flip–chip bonded Si/SiGe HBT's are used as active devices. A maximum output power of −8 dBm and a phase noise of −105 dBc/Hz at an offset frequency of 1 MHz have been measured. At a lower output power of −14 dBm an optimum phase noise of −112 dBc/Hz has been achieved. The mechanical tuning range of the oscillator is approximately 500 MHz.**

***Index Terms*—Dielectric resonator, microstrip circuits, push–push oscillator, SiGe HBT.**

I. Introduction

LOW PHASE noise signal sources are an important precondition for future system applications at millimeter-wave frequencies [1]. It has been shown by using the push–push principle [2], [3] that the usable frequency range of active devices for oscillator applications can be extended even far beyond the maximum frequency of oscillation of the transistor itself [4], [5]. The reason for this result is that in a push–push oscillator the internal operation frequency is only half the output frequency. This fact also allows to use commercially available dielectric resonators (DR's) with resonant frequencies of up to 30 GHz for millimeter-wave oscillators up to 60 GHz. In this work, a *TRANS-TECH* 29-GHz resonator has been used to fabricate a 58 GHz DRO in hybrid thin-film technology on a 10 mil Alumina substrate. The active devices are flip–chip bonded onto the substrate to minimize bonding inductances. Thin-film resistors and beam-lead capacitances are used as additional passive elements.

II. Circuit Principle

In a push–push oscillator, two symmetrical oscillator parts are operating at half the output frequency $f_0 = 1/2 f_{out}$ with a phase difference of 180°. In an appropriate output network, the two signals

$$s_1 = a_1 e^{j\omega_0 t} + a_2 e^{j2\omega_0 t} + a_3 e^{j3\omega_0 t} + \cdots \tag{1}$$

and

$$s_2 = a_1 e^{j\omega_0 (t-\Delta t)} + a_2 e^{j2\omega_0 (t-\Delta t)} + a_3 e^{j3\omega_0 (t-\Delta t)} + \cdots \tag{2}$$

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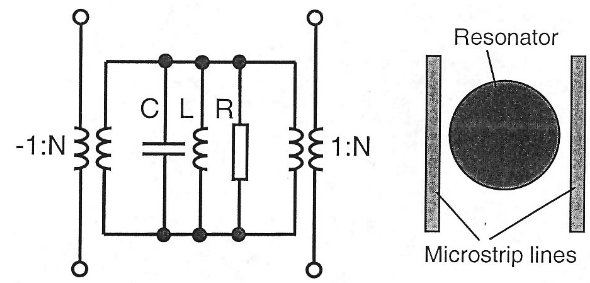

Fig. 1. Equivalent circuit of the dielectric resonator with two coupling lines.

with the phase difference

$$\omega_0 \cdot \Delta t = \pi \tag{3}$$

are added. This results in the output signal

$$s_{\text{out}} = a_1 e^{j\omega_0 t} \cdot (1 + e^{-j\pi}) + a_2 e^{j2\omega_0 t} \cdot (1 + e^{-j2\pi}) + \cdots. \tag{4}$$

In (4), the odd multiples of the frequency f_0 cancel out and the even multiples are added constructively. It is obvious that (3) is crucial for the proper operation of the circuitry. There are, however, several possibilities to enforce this odd mode of operation [3], [4], [6], [7]. Furthermore, equal amplitude values a_1 and, therefore, absolute symmetry of the two network parts is required to achieve a good suppression of the fundamental signal.

In the case of a DRO, the phase difference is assured by coupling the resonator to two parallel microstrip lines [3], [8]. The magnetic field at the fundamental resonance shows a radial symmetry. Therefore, two transmission lines at two opposite sides of the resonator are coupled with equal magnitude and opposite sign, i.e., with the required phase difference.

III. Resonator Modeling

For the CAD of the oscillator, a model of the resonator is required that reflects the odd coupling of the two microstrip lines. Therefore, the RLC-equivalent circuit of the resonator is extended with two ideal transformers with response ratios of $1 : N$ and $-1 : N$, respectively (Fig. 1). Since the transformers are only used to reflect the odd coupling of the two lines, a value of $N = 1$ is assumed.

The validity of this equivalent circuit has been verified. The coupling of the resonator both to one and to two microstrip lines has been calculated using the electromagnetic field simulator *HP HFSS*. Although we found that the resonant frequency in the two line case increases about 100 MHz with respect to the one

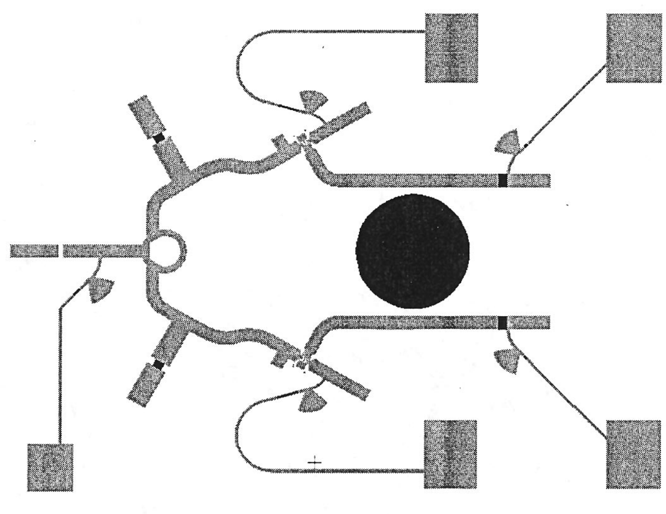

Fig. 2. Layout of the 58 GHz Oscillator.

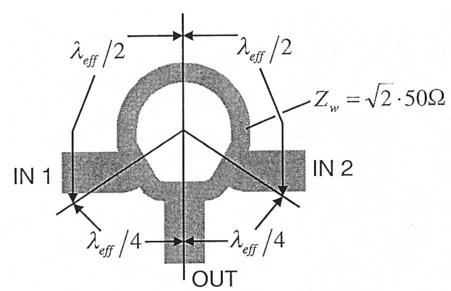

Fig. 3. Layout of the symmetrically modified rat race coupler.

line case, the standard element values for the equivalent circuit R, L, C, (provided by the manufacturer) could be used.

IV. Active Device Modeling

Si/SiGe HBT's, fabricated at the Daimler Chrysler Research Center, Ulm, Germany, have been used as active devices. The double mesa HBT's have two emitter fingers with an active area of $2 \times 8\ \mu\text{m}^2$. They show maximum values of approximately $f_T = 40$ GHz and $f_{\max} = 70$ GHz. A complete large-signal modeling has been done using a VBIC model, where additional parasitic elements have been added [9], [10]. The model parameters are extracted from dc and S-Parameter measurements. For the modeling of parasitic elements, additional open and short dummy structures have been used.

V. Circuit Description

Fig. 2 shows the layout of the entire oscillator circuit with the geometrical dimensions of approximately $10 \times 6\ \text{mm}^2$. On the left side, a beam-lead capacitor decouples the output line from the active circuitry. Then, a modified rat-race coupler is used as a power combiner (Fig. 3). This element is designed to have high transmission s_{12} and s_{13} from the inputs to the output at 58 GHz. Additionally, it must be symmetric with respect to the two inputs in order to maintain symmetry of the two oscillator half circuits. As a result, the fourth input of a typical rat-race coupler is omitted. This type of combiner shows slightly higher transmission compared to a Wilkinson combiner and, therefore, is preferred for our application. Additional stubs at the internal output lines are used to improve isolation at 29 GHz. The resonator is coupled to the two transmission lines on the right side of Fig. 2. The lines are terminated with 50 Ω loads to avoid parasitic reflections. The ground of these resistors is provided by $\lambda/4$ stubs at 29 GHz. Then, at 58 GHz the lines are effectively terminated with open circuits, and the power at 58 GHz is reflected. The third terminal of the transistors, in this configuration the base terminal, is connected to open transmission lines to add positive feedback, i.e., to increase instability of the devices. Finally, the bias network uses mainly 58 GHz stubs and proper positioning of the bias lines to decouple dc and ac signals.

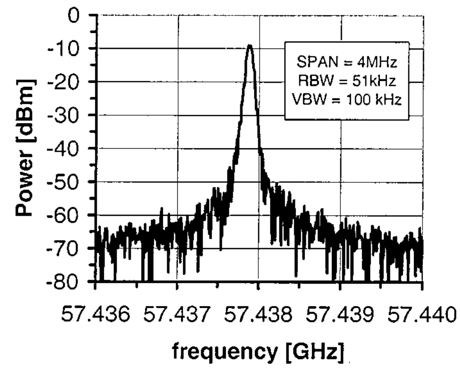

Fig. 4. Spectrum of the oscillator signal at 58 GHz.

VI. Experimental Results

Several microstrip oscillator circuits were fabricated in thin-film technology on 10 mil alumina substrates. Measurements were performed using an *HP 71 000* modular spectrum analyzer and appropriate harmonic mixers. Fig. 4 shows a typical measured output spectrum. The oscillation frequency is approximately 57.4 GHz with a mechanical tuning range of about 500 MHz, i.e., from 57.4 to 57.9 GHz. As expected, we found a dependence of the output power at 58 GHz on the length of the transmission line between the resonator and the terminating 50 Ω resistors. A maximum power of -8 dBm was achieved with a phase noise of -105 dBc/Hz at an offset frequency of 1 MHz. According to circuit simulations, considerably more output power could be expected from future designs using six finger HBT's instead of the two finger devices used in this work. Experimentally, we found a trade off between output power and phase noise. At a moderate output power of about -14 dBm, we measured a single sideband phase noise of -86 dBc/Hz and -112 dBc/Hz at offset frequencies of 100 kHz and 1 MHz, respectively. For these measurements an *HP 3048* phase noise measurement system was used. Suppression of the fundamental signal at 29 GHz simultaneously was measured to be -17 dBc. Table I compares the results of this work to previously published data of similar oscillators. The output power we achieved is relatively low. However, this problem can be overcome using larger devices in future designs. The

TABLE I
COMPARISON OF PUBLISHED DATA OF MM-WAVE OSCILLATORS

	Rheinfelder [11]	Funabashi (DRO) [12]	Kashiwa [13]	this work (DRO)		Inoue (DRO) [14]	Kawasaki [15]	Huei-Wang [16]	Aoki [17]	Wenger (DRO) [18]
aktive element	SiGe-HBT	AlGaAs/ InGaAs-HFET	AlGaAs/ InGaAs-HEMT	SiGe-HBT		AlGaAs/ InGaAs-HFET	InGaP/ InGaAs/ GaAs-HEMT	InP-HBT	AlGaAs-HBT	InGaAs/ GaAs-HFET
f_T/f_{max} [GHz]	70/95	-	-	40/70		-/240	-	60/120	130/180	90/180
f_0 [GHz]	47	55	56	57.5	57.5	60	60	62	60	81
P_0 [dBm]	13	3.7	11	-8	-14	6.9	6.7	4	-9	0
L_{ssb}(100kHz) [dBc/Hz]	-99	-88	-85	-	-86	-80	-63	-78	-	-
L_{ssb}(1MHz) [dBc/Hz]	-	-	-103	-105	-112	-104	-87	-105	-80	-90

measured phase noise values are comparable or better than similar oscillators using GaAs- or InP-based HEMT's and HBT's as active devices.

VII. CONCLUSION

The push–push oscillator principle allows to use commercially available dielectric resonators with resonance frequencies up to 30 GHz for millimeter-wave oscillators up to 60 GHz. An equivalent circuit for a DR coupled to two microstrip lines, verified by electromagnetic field simulations, has been given. A 58 GHz hybrid DRO has been presented. A maximum output power of −8 dBm with a phase noise of −105 dBc/Hz at an offset frequency of 1 MHz has been achieved. An optimum phase noise of −112 dBc/Hz at an offset frequency of 1 MHz has been measured at an output power of −14 dBm. The mechanical tuning range of the oscillator is approximately 57.4–57.9 GHz. The presented results clearly demonstrate the suitability of the push–push technique for the design of millimeter-wave DRO's using SiGe HBT's.

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REFERENCES

[1] W. Menzel, "Future applications," in *Silicon-Based Millimeter-Wave Devices*, J. F. Luy and P. Russer, Eds. Berlin, Germany: Springer-Verlag, 1994.

[2] Z. Nativ and Y. Shur, "push–push VCO design with CAD tools," *Microwave J.*, pp. 127–132, Feb. 1989.

[3] A. M. Pavio and M. A. Smith, "Push-push dielectric resonator oscillator," in *IEEE MTT-S Dig.*, 1985, pp. 266–269.

[4] F. X. Sinnesbichler, H. Geltinger, and G. R. Olbrich, "A 38-GHz push–push oscillator based on 25-GHz f_T BJT's," *IEEE Microwave Guided Wave Lett.*, vol. 9, pp. 151–153, April 1999.

[5] J. R. Bender and C. Wong, "Push-push design extendes bipolar frequency range," *Microwaves RF*, pp. 91–98, Oct. 1983.

[6] D. M. Smith, J. C. Canyon, and D. L. Tait, "25–42 GHz GaAs heterojunction bipolar transistor low phase noise push–push VCOs," in *IEEE MTT-S Dig.*, 1989, pp. 725–728.

[7] A.-S. Hyun *et al.*, "K-band hair-pin resonator oscillators," in *IEEE MTT-S Dig.*, 1999, pp. 725–728.

[8] C.-M. Liu and C. Y. Ho, "On the design of a voltage-tuned push–push dielectric resonator oscillator," *Microwave J.*, pp. 165–174, June 1990.

[9] C. C. Mc Andrew *et al.*, "VBIC95, the vertical bipolar inter-company model," *IEEE J. Solid-State Circuits*, vol. 31, pp. 1476–1483, Oct. 1996.

[10] F. X. Sinnesbichler and G. R. Olbrich, "Extraction of VBIC model parameters for SiGe HBT's including thermal effects, self-heating and 1/f-noise," in *Proc. 5th Europ. IC-CAP Meeting*, Marseille, France, June 1999.

[11] C. N. Rheinfelder, K. M. Strohm, L. Metzger, H. Kibbel, J.-F. Luy, and W. Heinrich, "47 GHz SiGe-MMIC ocillator," in *IEEE MTT-Symp.*, 1999, pp. 5–8.

[12] M. Funabashi *et al.*, "A V-band AlGaAs/InGaAs heterojunction FET MMIC dielectric resonator oscillator," in *16th IEEE GaAs IC Symp.*, 1994, pp. 30–33.

[13] T. Kashiwa *et al.*, "V-band high-power low phase-noise monolithic oscillators and investigation of low phase-noise performance at high drain bias," *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 1559–1565, Oct. 1998.

[14] T. Inoue, "60 GHz dielectrically stabilized monolithic voltage controlled oscillator," in *25th Eur. Microwave Conf.*, vol. 2, 1995, pp. 281–284.

[15] Y. Kawasaki, K. Shirakawa, Y. Ohashi, and T. Saito, "60 GHz monolithoc oscillator using InGaP/InGaAs/GaAs HEMT technology," in *IEEE MTT-Symp.*, 1995, pp. 541–544.

[16] H. Wang *et al.*, "Low phase noise millimeter-wave frequency sources using InP-based HBT MMIC technology," *IEEE J. Solid-State Circuits*, vol. 31, pp. 1419–1425, Oct. 1996.

[17] I. Aoki *et al.*, "GHz AlGaAs HBT oscillator," in *18th Annu. GaAs IC Symp.*, 1996, pp. 281–284.

[18] J. Wenger and U. Güttich, "Ka- and W-band PM-HFET DRO's," *IEEE Microwave Guided Wave Lett.*, vol. 3, pp. 191–193, June 1993.