A 10 GHz high-efficiency and low phase-noise negative-resistance oscillator optimized with a virtual loop model

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Abstract: A virtual loop model was built by the transmission analysis with virtual ground method to assist the negative-resistance oscillator design, providing more perspectives on output power and phase-noise optimization. In this work, the virtual loop described the original circuit successfully and the optimizations were effective. A 10 GHz high-efficiency low phase-noise oscillator utilizing an InGaP/GaAs HBT was achieved. The 10.028 GHz oscillator delivered an output power of over 15 dBm with a phase-noise of lower than –107 dBc/Hz at 100 kHz offset. The efficiency of DC to RF transformation was 35%. The results led to a good oscillator figure of merit of –188 dBc/Hz. The measurement results agreed well with those of the simulations.

Key words: oscillator; InGaP/GaAs HBT; low phase-noise; high efficiency; virtual loop; transmission analysis with virtual ground

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1. Introduction

Microwave systems such as wireless communication equipments need frequency sources to generate signals and define the system operating band. Microwave oscillators are essential parts of frequency sources. Low phase-noise oscillators make systems utilize channels more effectively. A lower phase-noise of the oscillator means less interference between channels and that a smaller frequency band is needed for the same data rate. For high frequency systems, power output level and efficiency are always meaningful for reducing amplifier demands and extending the battery life of mobile systems. This is why high-efficiency and low phase-noise oscillators are always preferred. A great deal of work has been reported on the topic[7–9].

Microwave oscillators are mostly designed by the negative-resistance (N-R) method. Focusing on the port parameters of the N-R and the resonator blocks, the N-R method can start up an oscillation easily without dealing with the complex parasitics inside the blocks. But concerning the loaded Q and power, the N-R method is deficient compared to the positive feedback loop transmission method. This is not explicit in the N-R model while calculating the loaded Q while a resonator is connected with a negative resistor, making the total resistance zero or negative, and it is hard for the N-R method to calculate the output power and gain of the active device that was artificially made unstable for oscillation. It is also hard for the loop method to build unilateral signal loops directly through the distributed parasitics at microwave frequencies. If they can be combined together by the method of transmission analysis with virtual ground (TAVG)[1,2], an N-R oscillator can be firstly built, taking advantage of the N-R method, and then the circuit can be transformed into a virtual loop topology and analyzed from the loop transmission viewpoint. By checking the phase and gain of the virtual loop, the phase-noise and power performance of the oscillator can be optimized just as in the low-frequency feedback oscillator design. In this work, N-R oscillation theory assisted with a virtual loop model built by TAVG is used in oscillator design. The oscillator is designed and optimized for the goal of low phase-noise and high-efficiency at 10 GHz, the typical microwave frequency at which many microwave device manufacturers characterize their products. The combined method is shown to be quite effective. The oscillator is designed and optimized quickly, and the measurement results match the design goal well.

2. Theories of the design

When designing microwave oscillators, N-R oscillation theory is widely used for its convenience in starting up an oscillation, and the positive feedback loop theory is well known for its explicit analysis on signal flows. Their models are shown in Fig. 1. Equation (1) shows the requirement for an N-R oscillator in Fig. 1(a), and Equation (2) for a feedback oscillator in Fig. 1(b).

$$\Gamma a \Gamma res = 1 \text{ or } Z_{\mathrm{a}} + Z_{\mathrm{res}} = 0, \tag{1}$$

$$G(j\omega)F(j\omega) = 1. \tag{2}$$

After an oscillation starts up, many parameters such as the phase-noise need to be optimized. Leeson gave the famous equation for oscillator single side band phase-noise as[3,4]:

$$L(\Delta\omega) = 10\lg\left\{\frac{1}{2}\frac{KTF}{P_{\mathrm{s}}}\left[1+\left(\frac{\omega_0}{2QL\Delta\omega}\right)^2\right]\left(1+\frac{\omega_{\mathrm{fc}}}{\Delta\omega}\right)\right\} \text{ (dBc/Hz)}. \tag{3}$$

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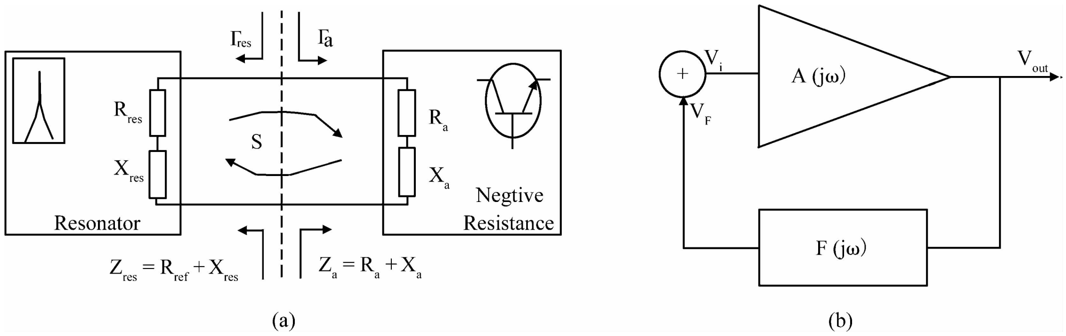

Fig. 1. (a) Negative-resistance oscillator; (b) Feedback oscillator.

Equation (3) shows the keys to low phase-noise, including high oscillator output power P_s, high loaded quality factor Q_L and low active device noise characteristics $F(1+\omega_{fc}/\Delta\omega)$. Using devices with low flicker-noise-corner such as HBTs can meet the third requirement, but the other two are not easy to obtain for the N-R model. Leeson's equation was based on a signal transmission and amplification model in the feedback loop[3,4], as was the maximum oscillator output power evaluating equation[5].

$$P_{osc} = P_{SAT}\left(1-\frac{1}{G}-\frac{\ln G}{G}\right). \tag{4}$$

Alechno introduced the TAVG concept, which can be of help in microwave oscillator analysis[1,2]. The basis of TAVG is that the physical ground points of a circuit can be treated as one normal node, and a normal node can be treated as a new virtual ground. Then the circuit can change its form with the new virtual ground, forming new topologies and view angles. The idea makes it possible to transform the N–R oscillator into a feedback loop model for analysis. Its inspiring advantages are that the loop gain and phase shift can be checked and tuned by adjusting the feedback network, including the original matching network and resonator. This means that the loop power gain and loaded Q are entirely controllable. If a virtual loop standing for the original N-R oscillator can be built, straightforward performance optimizations can be made.

3. Design and measurement

For the requirement of low phase-noise, an InGaP/GaAs HBT was used as the active device for its low flicker-noise-corner and high performance at microwave frequency. The common base topology and the open-short transmission stub resonator using the same type of HBT were used in our earlier work on a 6 GHz oscillator in Ref. [7]. The design was simulated with the Agilent advanced design system (ADS). With the 2.5D electric and magnetic field (EM) simulation tools, called momentum in ADS, full simulated EM models of the design layout were built and co-simulated with the active devices to get accurate results.

The oscillator circuit model based on the N–R oscillation theory is shown in Fig. 2.

As shown in Fig. 2, the HBT forming the N-R block was configured in common base mode. Transmission line TLb

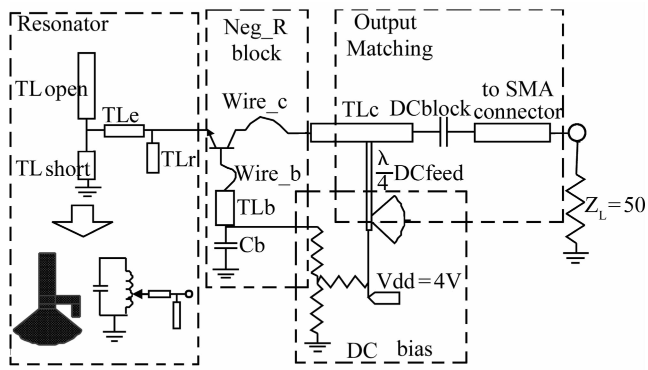

Fig. 2. N–R circuit model of the oscillator.

acted as a positive feedback inductor between the base and ground to make the device unstable. The main resonator was made up by the open and short stubs, TLopen and TLshort. The total length of them was about $\lambda/4$, forming a high Q parallel resonator. The position of TLe formed a coupling tap point, controlling the coupling and reflection loss between the resonator and N–R block. TLe also consisted of the model of the pad where the HBT was mounted and the stub TLr was used to resonate out the redundant parameters of the pad. The layout EM model and equivalent LC network model of the resonator network are also shown in Fig. 2. The HBT was mounted on the pad at its emitter, and the other two electrodes were connected with the circuit by wire-bonding. The length of wire was tested for accurate modeling. The load is connected to the collector via a segment of matching line TLc with a DC block capacitor. Practically all capacitors in Fig. 2 except the DC block were replaced by well EM designed radial stubs to eliminate the loss and the errors that might be caused by surface mounted elements and soldering.

After the N–R model was designed to satisfy Eq. (1) so that the oscillation could start up, TAVG was employed to transform the circuit in Fig. 2 into the virtual loop model shown in Fig. 3. The essential work in the transformation was dealing with the relationship between the microwaves' distributed characteristics and the original oscillation conditions. For example, adding a 50 Ω transmission line to the 50 Ω load means nothing but a little more loss for the original common base oscillator in Fig. 2, and adding the line to the resonator like TLe might destroy the N-R oscillation conditions (Eq. (1)) due to the added impedance transformation and phase shift. But if these changes were done to the circuit in Fig. 3, the

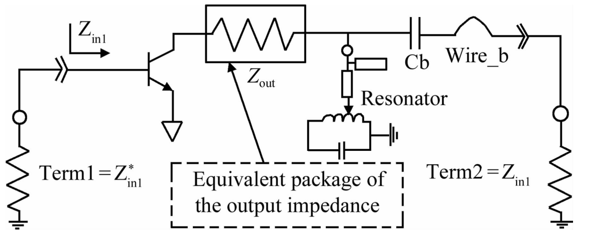

Fig. 3. Virtual loop model of the original N-R oscillator circuit.

results would be completely different. This example indicates that a parameter that seemed to be important in the original circuit may mean nothing in the transformed circuit, and vice versa. How to deal with a parameter in the transformation depends on how the parameter takes part in making up the oscillation conditions. In this work, there were three key tips to build the virtual loop model in Fig. 3.

First, the load with output matching network was calculated and replaced by the equivalent impedance Z_{out}, which can be tuned according to the need of optimization.

Second, modeled as the parallel LC network in Fig. 2, the resonator was found to have the same topology after the transformation. So the layout EM model of the resonator can be reused to co-simulate in the virtual loop model. The base feedback line TLb was completely removed due to the sufficient inductance offered by bonding wire_b.

Third, taking advantage of the good unilateral gain of the HBT, the reference terminators for the open loop transmission analysis were set as a pair of conjugate values according to the input impedance Z_{in1}. In this way, the loop gain could be checked more effectively than with common 50 Ω terms, because the signal along the closed loop was loaded not by the 50 Ω but by the circuit itself.

With the virtual loop, optimizations were made as if dealing with a real positive feedback oscillator. The total output impedance was tuned to increase the device gain for a higher output power, according to Eq. (4). The power and gain consumed by the load was tuned; meanwhile, the loop gain was kept over unity for Eq. (2). The resonator block was tuned to increase the transmission phase slope, which stands for the circuit loaded Q and controls the phase-noise, according to Eq. (3)

After tuning, as in the simulated curves shown in Fig. 4, the deepest phase slope happened at the zero phase point near 10 GHz. The slope result in a loop group-delay defined[5, 7] Q_L of 58, while the reflection group-delay defined Q_L of the main resonator was tuned to 110 with a 50 Ω terminator. The loop gain reserved for the HBT gain-compression and unexpected circuit loss was 2.1 dB, a not very high but sufficient value that set the gain-compression to a low level. The low gain-compression meant less saturation and less noise degradation. The difference between the device gain and the loop gain was caused by Z_{out}, indicating that the power was delivered to the output load in Z_{out}.

The optimized parameters were pushed back to the original circuit for validation and nonlinear simulations, such as

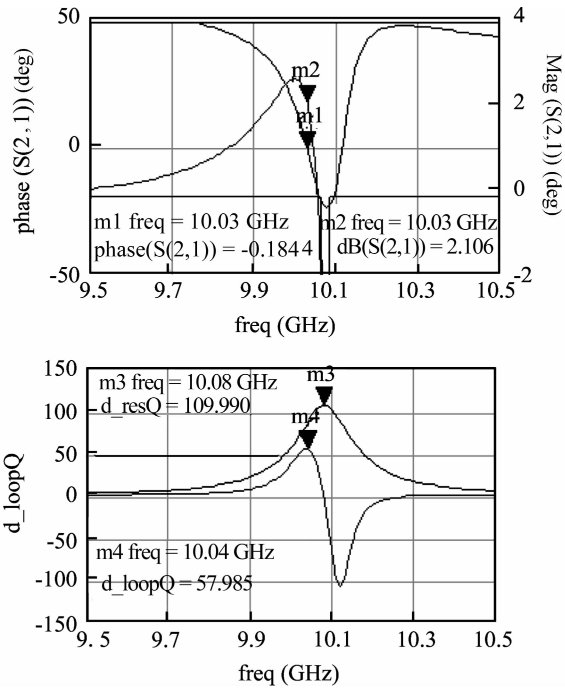

Fig. 4. Optimization results of the virtual loop model.

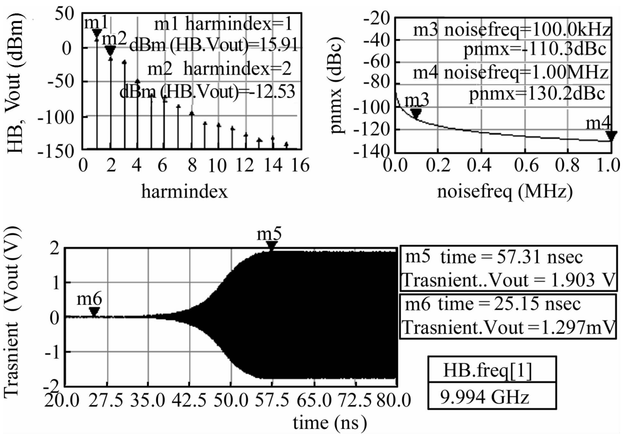

Fig. 5. Nonlinear simulation results of the N-R oscillator.

harmonic balance and transient simulations in ADS. The simulated curves are shown in Fig. 5. Good agreements between the two simulations indicated that the virtual loop we built could exactly describe the core mechanism of the oscillation.

The layout was made on a PTFE board. The oscillator was mounted on a test fixture and measured on an HP 8563E spectrum analyzer (SA). The fixture and the line to the SA had a loss of about 1 dB at 10 GHz. The circuit photographs and measurement results are shown in Fig. 6.

The measured oscillating frequency was 10.028 GHz, and the output power was about 15.3 dBm with the 1 dB fixture loss added, as shown in Fig. 6(b). A clear spectrum was achieved without spurs, as Figure 6(c) shows. With the SA HP8563E, the phase noise and FOM can be calculated from Eqs. (5) and (6).

$$\mathrm{PN}(f_{\mathrm{offset}}) = \Delta P(\mathrm{dB}) - 10\,\mathrm{lg}(\mathrm{RBW}), \tag{5}$$

$$\mathrm{FOM} = \mathrm{PN}(f_{\mathrm{offset}}) - 20\,\mathrm{lg}\,\frac{f_0}{f_{\mathrm{offset}}} + 10\,\mathrm{lg}\,\frac{P_{\mathrm{DC}}}{1\mathrm{mW}}, \tag{6}$$

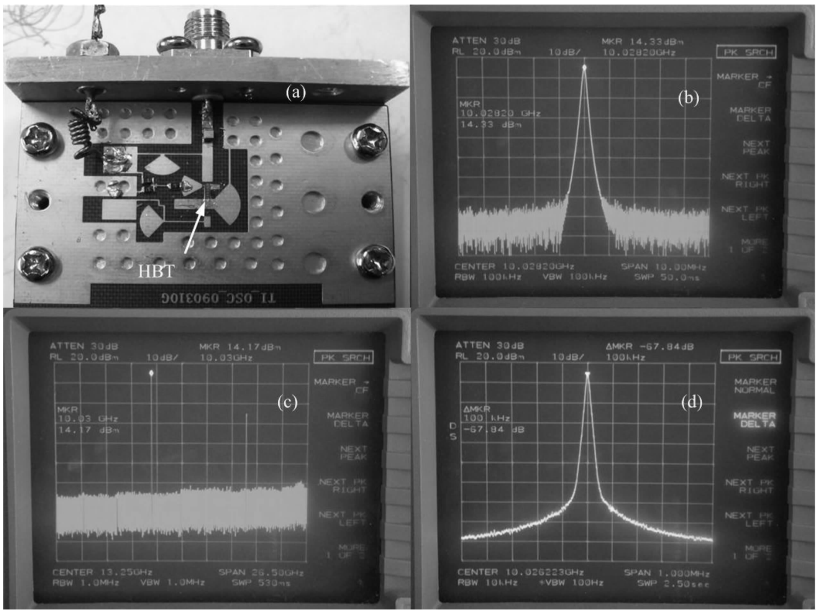

Fig. 6. The realized oscillator and the measurement.

Table 1. Simulated and measured characteristics of the oscillator.

	Frequency (GHz)	Output power (dBm)	Phase-noise at 100 kHz offset (dBc/Hz)
Simulated	9.994 (HB) 10.03 (TAVG)	15.9	–110.3
Measured	10.028	15.3	–107.8

where ΔP stands for the power at a frequency offset proportional to the carrier power in dB. RBW is the operating resolution band of the SA. The data can be read out directly from the SA: as shown in Fig. 6(c), at 100 kHz offset from the carrier, the –67.84 dB ΔP and 10 kHz RBW resulted in a phase-noise of –107.84 dBc/Hz. The total power consumption of the circuit was 4 V × 24 mA = 96 mW. The efficiency of DC to RF transformation and FOM were calculated out to be 36% and –188 dBc/Hz respectively. The FOM was quite good due to the low noise and high efficiency design goal, providing much room for trade-offs with other parameters, such as frequency tuning and circuit downscaling.

Another success of this work was that the measured data was in good agreement with the simulation results, as shown in Table 1.

Additionally, the device was set at class-AB mode in this work. If a higher efficiency mode was used, better efficiency but worse noise and harmonics suppression were expected[8, 9]. The harmonics were not always poor: the –5 dBm 2nd harmonic in this work at 20 GHz was high enough for many applications; it can be taken out by filters and used as a 20 GHz oscillator if needed. If the device saturation was set at a higher level, more harmonics but worse noise was expected[5].

4. Conclusions

The idea of building a virtual loop model for the N-R oscillator by the TAVG method was successfully used in this work. The carefully designed virtual loop and the original N-R circuit model worked in good agreement. The optimized oscillator showed high-efficiency and low phase-noise, and gave quite a good FOM value. The accordance between the simulated and measured results proves that the virtual loop model and careful EM design could be a great deal of help in microwave oscillator design, providing a novel method to design high performance oscillators.

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