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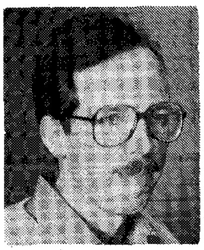

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A Waveguide-Cavity Multiple-Device FET Oscillator

ANDRZEJ MATERKA AND SHIZUO MIZUSHINA, MEMBER, IEEE

Abstract —**A waveguide-cavity oscillator, applicable to power-combining circuits, has been developed using probe for coupling between active device and cavity. No lossy stabilizing element is required. The control of output power, oscillation frequency, and injection locking bandwidth are performed easily. Output power of 44 mW and dc-RF conversion efficiency of 33.2 percent were obtained at 9.2 GHz for a single-device low-power FET oscillator. A simple technique of cascading the pretuned oscillator modules was used to construct multiple-device oscillators incorporating up to four FET's with combining efficiency of about 100 percent.**

I. Introduction

Gallium arsenide FET's offer attractive performance as microwave power sources, especially because of their efficiency and output power capabilities which have been steadily improving for the last few years [1]. For further increase of the output power from FET amplifiers several combining techniques have been developed [2]; however, no work has been reported on adding the power from FET oscillators. In the present paper, a waveguide-cavity oscillator is described which can be used to combine power from individual FET devices with high efficiency.

Manuscript received December 4, 1981; revised March 8, 1982.
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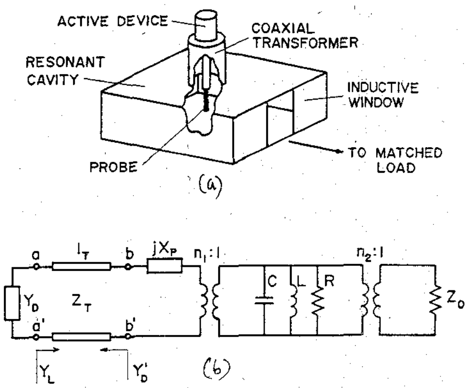

Fig. 1. (a) A waveguide-cavity single-device oscillator and (b) its equivalent circuit.

II. Description and Analysis

Fig. 1(a) shows a single-device oscillator of the proposed type in which an active device is placed at one end of a coaxial transformer. The other end of the transformer is coupled to the cavity by a probe. The cavity in turn is coupled to the matched waveguide through an inductive window. An equivalent circuit of the oscillator, for frequencies close to the cavity dominant resonant frequency, is shown in Fig. 1(b). The active device is represented by a

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one-port whose admittance $Y_D(A, f)$ has its real part negative in a certain frequency range, where A is the amplitude of RF voltage across the device terminals at the frequency f. The LCR parallel circuit represents the cavity. X_P is the probe reactance and the ideal input transformer $(n_1:1)$ represents the probe coupling. Z_T denotes the characteristic impedance of the coaxial transformer whose length is l_T. Z_0 stands for the waveguide characteristic impedance while the ideal output transformer $(n_2:1)$ represents the inductive window coupling.

The active device is loaded by an admittance $Y_L(f)$ which is seen from the port $a-a'$ looking toward the cavity. For sustained steady-state oscillation the following well-known condition must be satisfied:

$$-Y_D(A_O, f_O) = Y_L(f_O) \tag{1}$$

where A_O and f_O are, respectively, an amplitude and frequency of oscillation.

Let us assume that the device admittance is weakly dependent on frequency and monotonously dependent on A. It can then be represented by the so called device line [3], shown in Fig. 2. On the other hand, the probe input admittance plots a circle on the Smith chart, illustrated by the same figure. To obtain the oscillation in the vicinity of the cavity resonant frequency (f_r), a coaxial transformer length must be selected, resulting in a new device line (Fig. 2) which represents the admittance Y_D' seen at the terminals $b-b'$ of the equivalent circuit.

The input coupling coefficient (β_1) for the probe, placed at the point of maximum electric field in the rectangular cavity, is given as [4]

$$\beta_1 = \frac{n_1^2 R}{Z_T} = \frac{Q\eta}{2abcZ_T}\left(\frac{\lambda}{\pi}\right)^3 \tan^2\left(\frac{kd_P}{2}\right) \tag{2}$$

where a, b, c are the cavity dimensions, d_P is the probe length, and the other symbols have their customary meaning. On the other hand, the output coupling coefficient (β_2), for inductive window coupling, can be approximated as [5]

$$\beta_2 = \frac{R}{n_2^2 Z_0} \simeq \frac{Q\lambda^2 a^2}{8\pi c^4}\tan^4\left(\frac{\pi d_W}{2a}\right) \tag{3}$$

where d_W is the window width. From (2) and (3), the turns ratios n_1 and n_2 dependences on, respectively, d_P and d_W can be evaluated and used in a design procedure. As in the case of other cavity oscillators, there is a freedom in selecting the values of β_1 and β_2 for a given probe input admittance (corresponding, e.g., to maximum output power from the oscillator). This permits the independent adjustment of the external quality factor of the cavity, related, e.g., to the injection locking bandwidth. The control of the main oscillator parameters can be performed easily in the proposed circuit (by adjusting the probe length d_P, window width d_W, and cavity resonant frequency f_r), without significant constructional changes (cf. [6]).

An N-device oscillator of the present type consists of N active devices, loaded by probes, coupled to the electric field in a $TE_{10\,N}$ rectangular cavity and spaced half a wavelength apart. The condition for mode-free operation

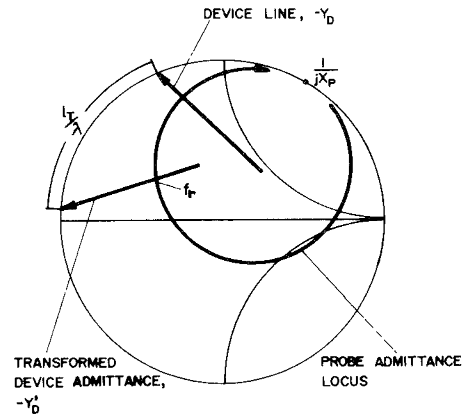

Fig 2. Graphic representation of the oscillation condition.

of this oscillator, with the same resonant mode of the cavity, is given as [7], [8]

$$-Y_D' \neq \frac{1}{jX_P}. \tag{4}$$

It is seen (Fig. 2) that condition (4) is satisfied if the coaxial transformer length is chosen as described earlier.

The approach utilized in this paper is similar to that of Kurokawa [7]. The difference between the properties of these two oscillators originates from the way in which the input coaxial line is connected with the cavity. In the circuit of [7], the extra length of the input line, extended beyond the physical area of coupling, is responsible for an unstable oscillator work, if shortcircuited. To compensate this effect, a lossy element is necessary. On the contrary, in the circuit of Fig. 1(a), the probe constitutes a natural extension of the inner conductor of the coaxial line, and, therefore, no stabilizing load is required.

III. Realization

To obtain oscillation in X-band an arrangement schematically shown in Fig. 3 was used. A microstrip FET circuit was substituted as the active device. A Hewlett–Packard HFET-1101 unit was mounted in this circuit which was fabricated on 0.031 in Duroid substrate. The "active device" was tuned to exhibit a maximum negative conductance at a desired frequency ($f_0 = 9.2$ GHz in the present example). The measured device admittance of a typical FET circuit is shown in Fig. 4. The transistor was biased at $V_{GS} = -0.5$ V ($I_D \simeq I_{DSS}/2$) and $V_{DS} = 4.0$ V.

The measured input admittance of the loading circuit, for several values of window width and fixed position of the cavity tuner, is also plotted in Fig. 4. It is seen that oscillation condition (1) is satisfied in the oscillator circuit discussed for $d_W < 10$ mm. Also, the following should be noted.

a) The output power is dependent on frequency, and, therefore, to obtain the maximum power, oscillation frequency must be kept close to 9.2 GHz.
b) The difference between $G_D(0, f)$ and $G_D(P_0, f)$ at 9.2 GHz is rather small.[1] Then, if the active device is

[1]This property seems to come from the features of the FET device. Further related works are planned on other FET circuit configurations.

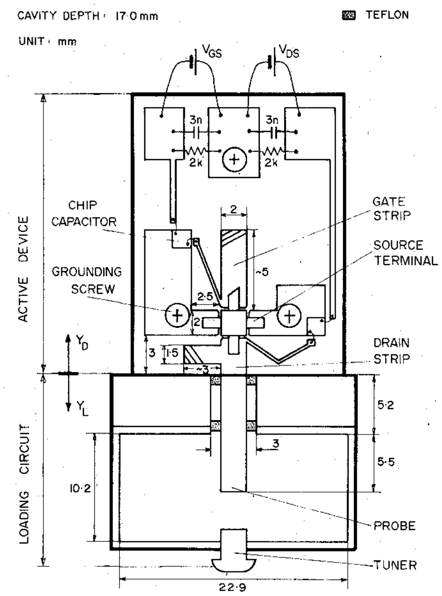

Fig. 3. An experimental single-device FET oscillator.

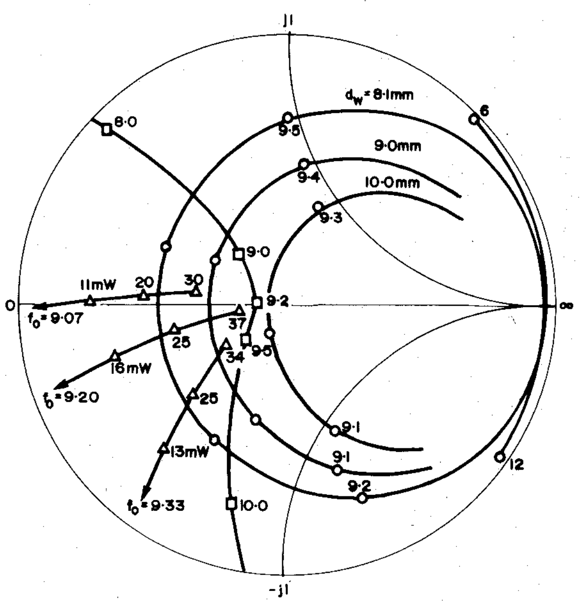

Fig. 4. Measured output admittance Y_D of the FET active circuit and input admittance Y_L of the loading circuit. 0— —0 load admittance locus, $Y_L(f)/Y_0$. □— · —□ small-signal device line, $-Y_D(0,f)/Y_0$. △— —△ large-signal device line, $-Y_D(P_0,f_0)/Y_0$. Frequencies in gigahertz, $Y_0 = 20$ mS, P_0 stands for the RF power generated at f_0.

expected to operate over a very wide temperature range, a compromising effort should be made to ensure stable operation at power near to its maximum. (G_D, the device conductance, depends on temperature.)

IV. Experiments

A. Single-Device Oscillator

Fig. 5 shows the measured characteristics of a single-device oscillator as previously described. No spurious oscillation, tuning difficulties, or hysteresis effects were observed during the experiment. As the window width was varied from 7.0 mm to 9.8 mm, the output power changed from 11 mW to 40 mW and the dc-RF efficiency from 8.5 percent to 29.2 percent. The frequency of oscillation decreased as the window width increased, as shown in Fig. 5. As expected, no oscillation was sustained for $d_W > 9.8$ mm. Also, no peak value in the P_0 dependence on window width was observed (Fig. 5). In spite of this, no related instabilities took place in the experiments conducted in ordinary temperature conditions.

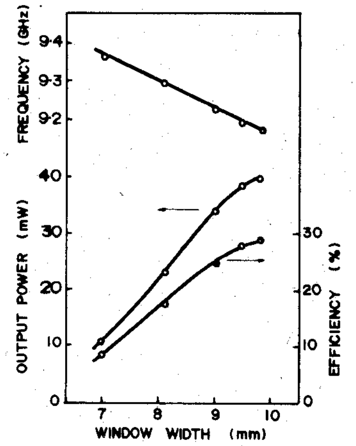

Fig. 5. Measured characteristics of a single-device FET oscillator.

TABLE I
RESULTS OF INJECTION LOCKING EXPERIMENT: $P_0 = 43$ mW, $f_0 = 9.1$ GHz, $P_i/P_0 = 0.1$

d_P (mm)	4.5	5.0	5.4	5.8
d_W (mm)	8.1	8.7	9.5	10.0
Δf_m (MHz)	22	29	38	45

The output power and efficiency dependences on gate and drain voltages were found typical for an FET oscillator [9]. For $d_W = 9.8$ mm, the maximum output power of 44 mW with efficiency of 25.0 percent were measured at $V_{GS} = -0.2$ V and $V_{DS} = 4.5$ V, while the efficiency reached its maximum value of 33.2 percent, with $P_0 = 26$ mW, at $V_{GS} = -0.5$ V, and $V_{DS} = 3.0$ V.

Experiments on injection locking were performed using the single-device oscillator containing a Mitsubishi MGF-1400 device with $I_{DSS} = 65$ mA. The probe length and window width had been changed so that for every given value of d_P the window width was adjusted to obtain $P_0 = 43$ mW. The associated oscillation frequency changes were compensated using the cavity tuner to keep the frequency constant, $f_0 = 9.1$ GHz. Next, the injection locking bandwidth (Δf_m) was measured at $P_i/P_0 = 0.1$, where P_i stands for the injected power. The results are given in Table I. The locking bandwidth was doubled, at power and frequency both constant, without changing the cavity dimensions. In other experiments, the Δf_m values up to 175 MHz were measured for longer probes and reduced cavity depth. A typical output power variation with injection signal frequency (f_i) is shown in Fig. 6.

All the obtained results compare favorably with the predicted oscillator performance.

B. Multiple-Device Oscillators

To combine power from two FET oscillators of the proposed type, the single-device circuit was first built using

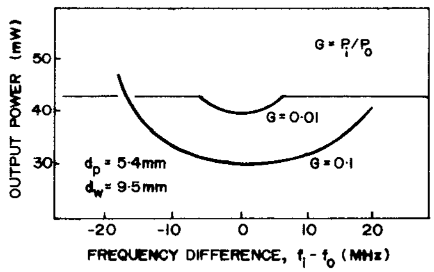

Fig. 6. A typical output power variation with injection signal frequency.

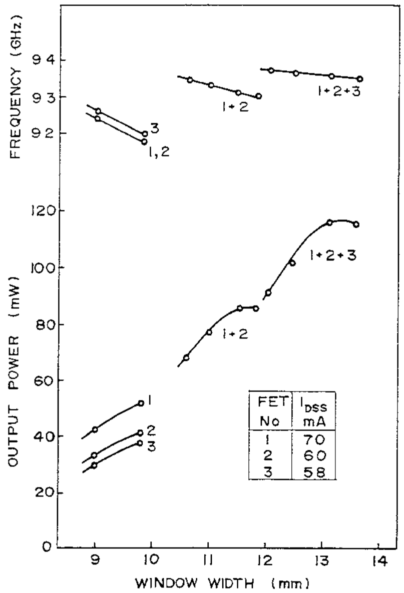

Fig. 7. Measured characteristics of the free-running and cascaded oscillators modules.

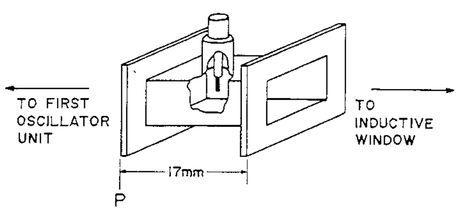

Fig. 8. Single-device module used in the power combining experiments.

an HFET-1101 device. The measured power and frequency characteristics of this oscillator are shown in Fig. 7. Next, the second oscillator was constructed in the form shown in Fig. 8. It was inserted in between the first oscillator and the inductive window to obtain the two-device arrangement. Prior to doing this, the module of Fig. 8 was tuned to have characteristics similar to the first oscillator (Fig. 7). A short circuit was placed at the plane P of the module during tuning.

The results of power-combining experiments using these two single-device modules connected in cascade are again shown in Fig. 7. The curve representing the output power is shifted toward the higher values of window width as compared to those obtained for the single-device circuits, in agreement with [7]. At $d_W = 11.8$ mm, output power reaches the maximum value of 86 mW which is less than the sum of the maximum power measured for the free-running single-device modules (51.0 mW + 41.5 mW). This is

TABLE II
RESULTS OF POWER-COMBINING EXPERIMENTS: $f_0 = 9.2$ GHz

NUMBER OF DEVICES	WINDOW WIDTH (mm)	OUTPUT POWER (mW)	DC-RF EFFICIENCY (%)	COMBINING EFFICIENCY (%)
2	11.8	94	29.3	102
3	13.6	133	29.7	103
4	14.0	171	28.8	98

because the frequency of oscillation was, in the case of two-device circuits, higher by about 120 MHz than the frequency at which the single-device modules had been tuned for maximum power (Fig. 7). After tuning the cavity resonant frequency downward by a screw-tuner, output power of 94 mW at 9.2 GHz was obtained from the two-device oscillator with $d_W = 11.8$ mm (see Table II) which corresponds to a combining efficiency of 102 percent.

A three-device oscillator was constructed by adding the third pretuned module to the previously characterized pair. Again, an oscillation frequency increase was observed (about 170 MHz) as compared to the free-running frequency of single-device circuits (Fig. 7). The oscillator performance obtained after cavity tuning is presented in Table II along with the data measured for the four-device circuit which was constructed and tuned in the same way as its predecessors. In all cases, combining efficiency of about 100 percent was obtained, showing once again the usefulness of the cavity-combining arrangements [10].

All the multiple-device oscillators readily generate the combined power after applying dc bias with no power or frequency discontinuities at any bias conditions. The output power and dc-RF conversion efficiency dependences on gate and drain voltages are similar to those of single-device circuits.

V. CONCLUSIONS

A waveguide-cavity oscillator, applicable to power-combining circuits, has been developed using probe for input coupling. Results of experiments at 9.2 GHz using low-power FET's demonstrate dc-RF efficiency of 33.2 percent for a single-device circuit and about 100 percent of combining efficiency for multiple-device oscillators, incorporating up to four transistors. Other solid-state active devices are also applicable to the proposed circuit. Experimental data from a stable four-Gunn device (NEC GD-511 AA) combiner have been obtained in our laboratory.

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Monolithic Integration of a Dielectric Millimeter-Wave Antenna and Mixer Diode: An Embryonic Millimeter-Wave IC

CHINGCHI YAO, STEVEN E. SCHWARZ, SENIOR MEMBER, IEEE, AND B. J. BLUMENSTOCK

***Abstract*—A monolithic silicon integrated circuit consisting of a mixer diode and an all-dielectric receiving antenna has been built and tested at 85 GHz. Radiation is coupled into the device optically with a coupling loss of 2.7 dB. No external metal structure is required for coupling. The design can be used efficiently at considerably higher frequencies, and can be elaborated into more complex integrated circuits. From measurements of video responsivity the losses of various parts of the device are estimated. A simple theory of conversion efficiency is found to agree well with experiments; this theory is then used to predict the performance of improved versions of the device. The conversion efficiency obtained with this demonstration device is low; it is shown, however, that acceptable conversion efficiencies can be obtained with a more advanced diode fabrication technology using epitaxial Si or GaAs. Integrated millimeter-wave receivers of this kind should be suitable for short-path terrestrial communications, in applications where compactness and low cost are required.**

Manuscript received December 30, 1981; revised March 2, 1982. This work was sponsored in part by the U.S. Army Research Office under Grant DAAG29-79-C-0134, in part by the National Science Foundation under Grant ECS-8009033, in part by the Air Force Office of Scientific Research, United States Air Force under Contract F49620-79-C-0178, and in part by the U.S. Army Mobility Equipment Research and Development Command under Contract DAAK70-80-C-0134.

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I. Introduction

A NUMBER OF different waveguide technologies are available for use in the "near-millimeter" regime of 100–300 GHz. These include conventional hollow metal waveguide, fin line, various strip lines, microstrip, dielectric

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