Short Papers

Equivalent Circuit of the Schottky-Barrier Field-Effect Transistor at Microwave Frequencies

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***Abstract*—Johnson's high-frequency representation theory for MOSFET's [1], experimentally confirmed by Hopkins up to 1 GHz [2], is extended in this short paper for SBFET's and is found to substantially agree with data [3], [7], for 1-μm- and $\frac{1}{2}$-μm-gate GaAs SBFET's up to 12 GHz. Regenerative-feedback conductance [2], [4], [5], not accounted for by conventional models [3], [6], is seen to be present in SBFET's at microwave frequencies.**

INTRODUCTION

It has been noted that the high-frequency reverse-transfer conductance G_{12} of the MOSFET exhibits a positive sign [2],[4],[5] when not entirely negligible, and displays square-law dependence on frequency. A positive sign for G_{12} is contrary to predictions of conventional MOSFET equivalent-circuit representations [4]. That is, direct resistive feedback and degenerative-source resistance both lead to a negative sign for G_{12}, and source-lead inductance usually gives rise to an identifiable resonance as well as strongly varying feedback terms. Johnson [1] suggested a MOSFET representation similar to that shown in Fig. 1 to account for a positive G_{12} and its square-law dependence on frequency. This representation can be distinguished from conventional representations by the incorporation of a drain-to-channel internal-feedback capacitor C_2. The channel resistance r_c, in common with C_2 and the gate-control capacitance C_1, gives rise to a regenerative-conductance term given by $G_{12} = \omega^2 C_1 C_2 r_c$.

The work by Johnson [1] was verified and extended by Hopkins [2], for MOSFET's, by measurements up to 1 GHz that were supplemented by transmission-line analysis. Since Hopkins' work has been little publicized, it appears that equivalent circuits of a similar nature, as shown in Fig. 1, have not been used to represent SBFET's, in spite of the fact that regenerative effects due to internal-feedback capacitance are appreciable in microwave SBFET's as discussed in this short paper. In addition, the representation suggested by Johnson is extended for SBFET's to include effects of parasitic resistances.

DISCUSSION

Even though the previous work [1],[2] was related to MOSFET representation, it is applicable also for representation of SBFET's because the linear circuit elements of the MOSFET and SBFET appear in the same configuration. As with the MOSFET, a drain-to-channel

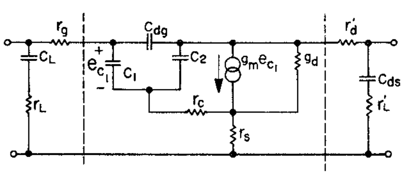

Fig. 1. SBFET equivalent circuit with internal feedback.

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internal-feedback capacitor C_2 is incorporated. A physical interpretation of C_2 can be given as the linear term of a nonlinear capacitance composed of parallel contributions from drain-to-channel feedback via the substrate, and an effective space-charge capacitance at the drain end of the channel. Physical parameters for these effects have not as yet been carefully derived.

Element values C_1, r_c, parasitic elements, etc., appropriate to SBFET's, should be used in the equivalent circuit. Certain parasitic elements are more pronounced in SBFET's than in MOSFET's. These elements are the source-to-channel plus contact resistance denoted by r_s, and the drain-to-channel plus contact resistance denoted by r_d'. In addition, gate-spreading resistance r_g is more important in SBFET's than in MOSFET's because SBFET's are usually used in a much higher frequency range. When these parasitic elements are included, along with Johnson's representation of the internal device, the equivalent circuit of Fig. 1 results where the elements are defined as follows:

- C_1 gate-to-channel capacitance;
- C_2 drain-to-channel internal-feedback capacitance;
- C_{dg} drain-to-gate feedback capacitance;
- C_L input parasitic capacitance;
- C_{ds} output-drain diode and parasitic capacitance;
- g_m low-frequency transconductance;
- g_d output conductance;
- r_c channel resistance;
- r_s source-to-channel plus contact resistance;
- r_g gate-spreading resistance;
- r_d' drain-to-channel plus contact resistance;
- r_L substrate-input spreading resistance;
- r_L' substrate-output spreading resistance.

Normally, the common-source resistor r_s is a degenerative element; hence, the admittance parameters were derived to examine the possibility of competing regenerative effects due to the C_1-C_2-r_c branch with the suspected degenerative effects of the parasitic source-to-gate resistance. It was found that the sign of G_{12} remains positive, and G_{12} still displays ω^2 dependence. The derived admittance parameters internal to r_g and r_d' are shown in Table I. A conclusion that can be drawn from this analysis is that the source resistance appears as though it adds to the channel resistance. Due to the phase shift caused by C_2, the feedback-conductance term remains regenerative even when source resistance is appreciable.

The internal admittance parameters should be finally transformed through the resistors r_g and r_d'. These transformed expressions become intractable; nevertheless, the effect of r_g is very small [3]. The transformation through r_d' can have considerable influence since r_d' may be typically 20 Ω or higher. Furthermore, certain devices, particularly high-field GaAs FET's, may exhibit negative resistance in the region between the gate and drain even before spontaneous oscillations occur. These effects have not been completely analyzed, nor has a representation theory been developed. In an attempt to evaluate the extent to which the predicted trends associated with the internal-device structure are displayed by measured terminal parameters, scattering parameters reported by Baechtold [3] were converted to admittance parameters, and the feedback admittance parameters are shown in Fig. 2.

It can be seen by inspection of Fig. 2 that: 1) the sign of G_{12} is positive; 2) G_{12} exhibits dependence on (frequency) [2]; 3) B_{12} exhibits nearly linear dependence on frequency. Apparently, the trends associated with the internal elements of the model are not significantly obscured by effects of terminal resistors r_g and r_d' for this particular SBFET. Careful analysis of parameter data, not presented here, does, in addition, provide a determination of specific element values for use in the equivalent circuit of Fig. 1.

TABLE I
INTERNAL ADMITTANCE PARAMETERS

$g'_{11} \cong \omega^2 C_1^2 (r_c + r_s)$	$b'_{11} \cong \omega \left[C_1 (1+g_m r_s) + C_{dg}\right]$
$g'_{12} \cong \omega^2 C_1 C_2 (r_c + r_s)$	$b'_{12} \cong -\omega (C_{dg} + C_1 g_d r_s)$
$g'_{21} \cong g_m$	
$g'_{22} \cong \frac{1}{r_d} (1 + g_m r_s)$	$b'_{21} \cong -\omega \left[(C_1+C_2) g_m (r_s+r_c) + C_1 g_d r_s + C_{dg}\right]$
	$b'_{22} \cong \omega \left[C_{dg} + C_2 \left[1+g_m (r_s+r_c)\right] + C_1 g_d r_s g_m (r_s+r_c)\right]$

Note: At high frequencies, all terms exhibit the common denominator $[1 + g_m r_s + \omega^2 (C_1 + C_2)^2 (r_c + r_s)^2]$.

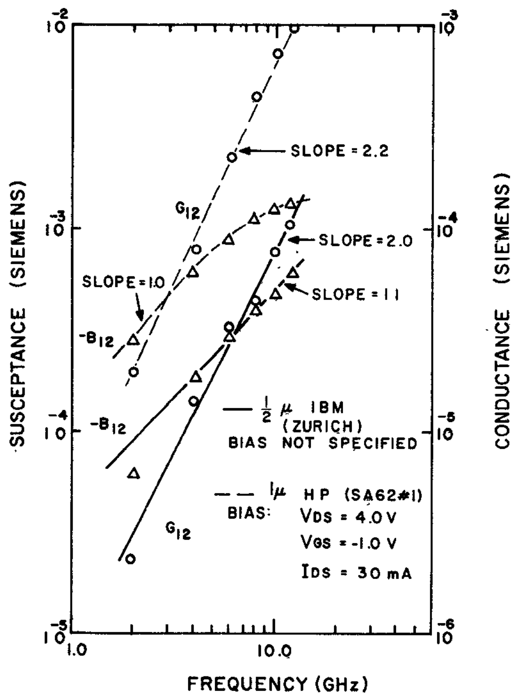

Fig. 2. Feedback admittance parameters for ½-μm- and 1-μm-gate-length GaAs SBFET's.

This representation for SBFET's, based on the original suggestion by Johnson for MOSFET's, is more significant than may be apparent from the parameters of the particular short ½-μm-channel SBFET discussed previously in which G_{12} is considerably less than $|B_{12}|$ up to 10 GHz. Beyond 10 GHz (closer to the gate cutoff frequency [2]), where considerable gain can still be realized, the representation discussed here will be useful in calculations of gain, stability, and broad-band matching networks by accounting for appreciable regeneration due to internal feedback.

Precise data supplied by Liechti for 1-μm-gate-length GaAs SBFET's confirm the trends reported here, as shown also by Fig. 2. In these devices, G_{12} is nearly equal to $|B_{12}|$ at 12 GHz; consequently the regenerative feedback effect is appreciable over the 2-GHz–12-GHz frequency range. In Liechti *et al.* [6], heavy emphasis was placed on "correct" modeling of 1-μm-gate-length SBFET's, but without incorporation of an internal-feedback capacitor C_2. Remarkable agreement between calculated and measured scattering parameters was apparent from the comparison presented; however, only three experimental S_{12} points were shown, whereas a full-set comparison was made for the remaining parameters. In comparing a full set of S_{12} data [7] to the calculated curve [6], a disparity was noted [8]. The representation discussed in this short paper accounts for the disparity, whereas conventional representations [3],[6] do not.

CONCLUSIONS

Transformed measured S-parameters for 1-μm and ½-μm GaAs SBFET's [3],[7] to Y-parameters agree with the Y-parameter predictions of the representation presented here that incorporates an internal drain-to-channel feedback capacitor. The effect of the capacitor is regenerative, is appreciable at microwave frequencies, and is not changed in nature by the presence of parasitic-source resistance.

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Broad-Band Varactor-Tuned IMPATT-Diode Oscillator

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***Abstract*—A varactor-tuned IMPATT-diode oscillator with a continuous and monotonic tuning bandwidth of 27 percent and a potential tuning range in excess of 40 percent is described. The results of a computer program which optimizes the tuning bandwidth of the equivalent circuit of the voltage-controlled oscillator (VCO) are presented. The VCO consists of two varactors located symmetrically on each side of an IMPATT diode all mounted in a ridged waveguide with two matched outputs into 50-Ω coaxial. Experimental results on bandwidth, power output, frequency linearity, and FM noise are presented.**

I. INTRODUCTION

Satellite and spacecraft applications have created a need for microwave, solid-state, electronically tuned oscillators which are tunable over a broad frequency range, exhibit high dc-to-RF conversion efficiency, and have low noise properties. Broad tuning range means fewer oscillators are needed for a given receiver bandwidth, and obviates the need for complicated switching schemes. Efficiency is essential for satellite and spacecraft applications, and low noise is a requirement for good receiver sensitivity.

Solid-state devices have been in existence for years which have the capability of meeting these requirements, but have not as yet realized their full potential. The main stumbling block has been that the devices are difficult to impedance match over a broad frequency range. Tunnel diodes are not acceptable for these applications because of their low power levels [1], whereas TRAPATT diode oscillators are inherently narrow band because the fundamental and first several harmonically related frequencies must be matched and tuned simultaneously [2]. Transferred-electron devices (TED's) are ruled out because of system efficiency requirements. The TED's generally operate at less than 5-percent dc-to-RF efficiency, and if YIG tuned, the voltage-controlled oscillator (VCO) efficiency would be less than 1 percent [3].

IMPATT diodes possess all of the characteristics which satisfy the power, efficiency, and broad-band negative resistance requirements. However, because of the circuit problem of broad-band impedance matching, IMPATT diode VCO's have been limited to 10-percent bandwidths [4]–[7]. It is the purpose of this short paper to describe an impedance-matching circuit which theoretically yields a total tuning range in excess of 40 percent and to present experimental results for this circuit configuration which are in excess of 27 percent. This was accomplished without significant sacrifice in power output and efficiency, and with an improvement of 8–10 dB in noise performance over previously reported state-of-the-art results.

This short paper is divided into two parts consisting of the theoretical model of the VCO and the experimental results. A computer program is used to analyze the theoretical model of the VCO which consists of the large-signal admittance of an IMPATT diode and the equivalent circuit of the varactor-loaded microwave circuit. Experimental techniques are described which were used to determine the equivalent-circuit element values. Finally, the overall performance (bandwidth, output power, efficiency, and FM noise) of a VCO is presented.

II. EQUIVALENT CIRCUIT

The equivalent circuit of a voltage-controlled IMPATT-diode oscillator consists of the large-signal circuit model of the diode and the voltage-dependent equivalent circuit of the varactor and microwave circuit. The operation of the oscillator is governed by the solution to the following equation:

$$Z_D(\omega_0,V_{\mathrm{RF}},I_{\mathrm{dc}}) + Z_C(\omega_0,V_B) = 0 \tag{1}$$

where

- Z_D the IMPATT-diode impedance (Ω);
- Z_C the circuit impedance (Ω);
- ω_0 the oscillator frequency (rad/s);
- V_{RF} the diode RF voltage (V);
- I_{dc} the diode bias current (mA);
- V_B the varactor bias voltage (V).

The VCO performance with respect to tuning bandwidth and power output may be analyzed as a function of diode and circuit parameters from the solution of this equation. However, due to the complexity of the equivalent circuits of both Z_D and Z_C, this cannot be solved in closed form but must be solved by an iterative technique at each operating point. This section of the short paper discusses the models for both of the impedances Z_D and Z_C, and presents results of a computer program which solves (1) at each operating point.

For the large-signal characterization of the IMPATT diode, a simple Read-diode model was incorporated [8],[9]. For a given dc bias current, the small-signal impedance of a typical IMPATT diode as calculated by this program is shown in Fig. 1 on the extended Smith chart. The large-signal model was used to generate a matrix of impedances as a function of frequency and RF voltage. This impedance matrix represents the impedance function Z_D in (1).

Since Z_C must be the negative of Z_D from (1), the frequency dependence of Z_C can be deduced from a plot of $-Z_D$, as is also shown in Fig. 1 for the small-signal impedance case. This figure

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