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THE DEFORMABLE-CHANNEL MODEL-A NEW APPROACH TO HIGH-FREQUENCY MESFET MODELLING

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SUMMARY

High-frequency small-signal circuit parameters are evaluated for a saturated-channel MESFET by including transit-time effects in a rigorous way through a study of induced shape changes in the saturated-channel region. Results agree well with empirical equivalent-circuit parameters for a physical MESFET.

Most non-numerical modelling of the small-signal AC characteristics of MESFETs assumes that the small-signal AC parameters are the same as the small-signal DC ones. In this paper, we present calculations of the high-frequency small-signal AC parameters which go beyond this assumption to include transit-time effects in a rigorous way; the transistor Y-parameters become frequency-dependent in exactly the way that is observed experimentally.

The necessity for including transit-time effects is clear when we examine the behavior of a change packet in the saturated channel. Because the differential mobility along the channel is very small, a charge inhomogeneity must propagate, as happens in the drift region of an IMPATT. However, due to the mobility anisotropy, charge inhomogeneities that appear in the saturated channel relax quickly, changing the channel shape locally. These shape variations propagate down the channel just as compressional waves do in the drift region of an IMPATT and (as we will show) can significantly affect the AC impedances of the device.

Let us assume that the device geometry is the same as that shown in Fig. 1: an epitaxial layer of length L and width a, with a donor doping density of N_D. The width of the saturated channel under DC conditions is denoted by b_o and is determined within the two-region model of Grebene and Ghandhi[1] by the gate and source-drain voltages. Let $\delta I_D(x,t)$ be the spatially-dependent total AC current flowing in the channel and passing through a plane located at the point x, and let $x = L_1$ be the end of the gradual channel. Then if $\tilde{\delta I}_D$ is the source terminal AC current, the IMPATT equation implies that

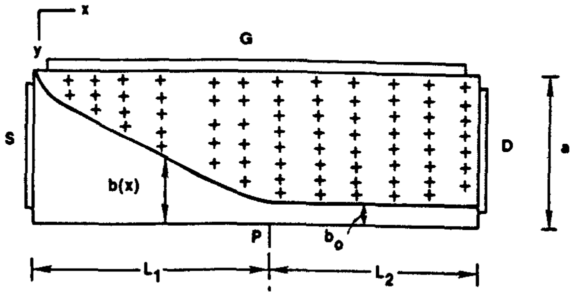

Figure 1. Schematic of a MESFET for the two-region model.

$$\delta I_D(x,t) = \tilde{\delta I}_D e^{j\omega(t - \frac{1}{v_s}[x - L_1])}$$

for the current at each point in the channel. Assume that the only way δI_D can vary is through a change in the channel width δb. Then $\delta I_D(x,t) = N_D e v_s z \delta b(x,t)$, where z is the gate length. This implies that there is also a high-frequency voltage perturbation in the saturated channel region given by

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2}\right) \hat{V}_{HF}(x,y,t)$$

$$= + \frac{N_D e}{\varepsilon_r \varepsilon_o} \left[\delta(y + b_o) + \delta(y - b_o)\right] \hat{\delta b}_{HF}(x,t)$$

where

$$\hat{\delta b}_{HF}(x,t) = \frac{\delta I_D}{N_D e v_s z}\left(e^{-j\frac{\omega}{v_s}[x - L_1]} - 1\right) e^{j\omega t}$$

We have solved this equation using a Fourier expansion

$$V_{HF}(x,y) = e^{j\omega t} \sum_{n=0} V_n(x) \cos\left(n + \frac{1}{2}\right) \frac{\pi y}{a};$$

assuming

$$V_n(x = L_1) = 0 \quad , \quad \left.\frac{\partial V_n}{\partial x}\right|_{x=L} = 0 \; ,$$

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we can evaluate the coefficients $V_n(x)$, and from them evaluate (a) the voltage drop down the saturated channel due to the transit-time effect and (b) the change induced on the gate by the shape variation. From these relations we get the Y-matrix elements:

$$Y_{11} = i\omega C_{gs} - \frac{r_{ds}G(\omega)}{r_{ds} + Z(\omega)} g_m \quad Y_{12} = \frac{G(\omega)}{r_{ds} + Z(\omega)}$$

$$Y_{21} = \frac{-r_{ds}}{r_{ds} + Z(\omega)} g_m \quad Y_{22} = \frac{1}{r_{ds} + Z(\omega)}$$

where

$$Z(\omega) = -4\alpha\zeta \frac{V_p}{I_s} \sum_{n=0}^{\infty} A_n f_n(\zeta)$$

and

$$G(\omega) = 1 - j\zeta - e^{-j\zeta} + 2j\zeta^2 \sum_{n=0}^{\infty} B_n f_n(\zeta)$$

$$A_n = \frac{\sinh[(1-p)\sigma_n/\alpha]}{\sigma_n^2 \cosh[\sigma_n/\alpha]} \qquad B_n = \frac{\cosh[p\ \sigma_n/\alpha]}{\sigma_n^2 \cosh[\sigma_n/\alpha]}$$

$$f_n(\zeta) = \frac{j\sigma_n e^{-j\zeta} - (-1)^n \zeta}{\zeta^2 - \sigma_n^2}$$

$$\sigma_n = (n + \tfrac{1}{2})\,\pi \quad \alpha = L_2/a \quad \zeta = \frac{\omega}{v_s} L_2 \quad p = b_o/a$$

where V_p is the pinch-off voltage, I_s the saturation current, and r_{ds}, g_m, and C_{gs} are the DC small-signal parameters. Note that the presence of $G(\omega)$ makes the transistor intrinsically bilateral at high frequencies, even when parasitic capacitances are neglected. Note also that $Z(\omega)$, $G(\omega) \to 0$ as $\omega \to 0$, so that we recover the usual quasi-DC Y-matrix.

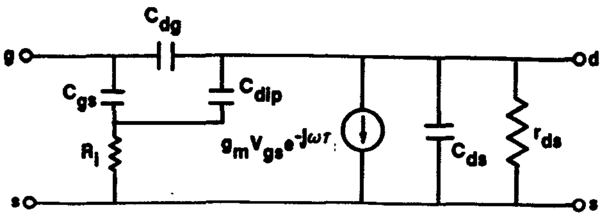

Figure 2. High-frequency MESFET equivalent circuit; the elements C_{dip}, R_i, and τ are all empirical parameters.

The standard equivalent circuit model for the MESFET at high frequencies (Fig. 2) has in addition to the usual quasi-DC capacitances C_{gd}, C_{gs}, and C_{ds}, two other elements — C_{dip} and R_i. Their presence is required because we must account for the experimental fact that a measurement of the admittance Y_{12}, which should be pure imaginary (it relates the feedback gate current to the drain voltage, which can only be mediated by C_{gd} for low frequencies),[2] reveals that ReY_{12} is in fact positive and varies with ω^2. Using the circuit topology shown in Fig. 2 yields the following low-frequency value for ReY_{12}:

$$ReY_{12} \cong \omega^2 C_{dip} R_i C_{gs}.$$

In Ref. 3, the response of a real 1-μm transistor is analyzed and small-signal circuit parameters are given. The numbers given in Ref. 3 that are relevant to our analysis are: C_{dip} = 0.12 pf, C_{gs} = 0.62 pf and R_i = 2.651. Inserting these numbers into Eq. (1), we determine that the coefficient of ω^2 is 0.0322 $(ps)^2$/ohm. If we now invoke our model, and assume $L_2 \sim L$ (the channel is completely saturated) and $p \sim 0$ (which is true in most cases anyway, i.e., the channel is always extremely thin), we predict that at low frequencies:

$$ReY_{12} \cong \frac{1}{2} \left(\frac{L}{v_s}\right)^2 [1 + 4C_o] r_{ds}^{-1}\, \omega^2$$

$$C_o = \sum_{n=0}^{\infty} \frac{1}{\sigma_n^3 \cosh \frac{\sigma_n}{\alpha}}.$$

Using the same parameters as in Ref. 3, we determine that the coefficient of ω^2 in Eq. (2) is 0.0321 $(ps)^2$/ohm, in close agreement with the empirical equivalent circuit result.

Our expression for the Y matrix points up a rather significant feature of the analysis: compared to r_{ds}, the modulus of the impedance $Z(\omega)$ actually is quite small, so there is effectively no transit-time factor $e^{-j\omega\tau}$ in the transconductance. This also is in agreement with experiments: the time τ which is found empirically from fitting data is always much smaller than the transit time. Therefore, we maintain that what experimenters are actually seeing is the small complex part of $Z(\omega)$, which gives g_m a small imaginary part, and not some fictitious "phase delay" down the channel.

IV. REFERENCES

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