New Non-Quasi-Static Theory for Extracting Small-Signal Parameters Applied to LDMOSFETs

Patrick Roblin, Siraj Akhtar, and Jeffrey Strahler

Abstract—We present analytic formulas for simultaneously extracting the parasitic resistances, inductances, and the intrinsic parameters of a small-signal FET equivalent circuit model including the non-quasi-static (NQS) charging time-constants associated with the gate and drain charges, respectively. For the NQS equivalent circuit topology considered, there exists a continuum of solutions for the circuit parameters, as a function of the source resistance, giving exactly the same frequency response fit. A multi-bias analysis is used to determine the final source resistance. Realistic results are obtained for power LDMOSFETs despite the very small value of the parasitics in these power RF devices.

Index Terms—FET equivalent circuits, LDMOSFET, microwave small-signal modeling, non-quasi-static, parameter extraction.

I. INTRODUCTION

PACKAGED Silicon power LDMOSFET transistors are finding important applications in RF power amplifier design, creating a demand for accurate models for use in circuit simulation [1], [2]. Efficient techniques have been developed for the extraction of the parasitics in quasi-static (QS) models [3], [4]. However, the extraction of R_{gsi} and drain time-delay τ in non-quasi-static (NQS) models are known to be "notoriously unstable," and "show enormous sensitivities" as stated by King *et al.* [5]. An iterative method has been recently proposed by Raskin *et al.* [6]. In this letter, we present analytic formulas for a charge-based model topology relying on two NQS time-constants [7], [8].

II. NEW EXTRACTION THEORY

At microwave frequencies, the iso-thermal microwave response of the LDMOSFET can be well represented by the NQS small-signal model shown in Fig. 1, where the intrinsic Y-parameters are: $Y_{intr,ij} = g_{ij} + j\omega C_{ij}/(1 + j\omega\tau_{ij})$ with $g_{11} = g_{12} = 0$, $g_{22} = g_d$ and $g_{21} = g_m$. In agreement with its associated charge-based NQS large-signal model topology [7] we limit our analysis to a model with two NQS time-constants: $\tau_G = \tau_{12} = \tau_{11}$ for the gate charge and $\tau_D = \tau_{22} = \tau_{21}$ for the drain charge.

In the presence of resistive and inductive parasitics at the terminals, one verifies that the extrinsic Z_{ij} for the NQS model

selected assumes a frequency dependence similar to that of the QS model [3], [4]

$$\mathrm{Re}[Z_{extr,ij}] = R_{ij} + \frac{A_{ij}}{\omega^2 + B_r^2} \tag{1}$$

$$\frac{1}{\omega}\,\mathrm{Im}[Z_{extr,ij}] = L_{ij} + \frac{B_{ij}}{\omega^2 + B_r^2} + \frac{G_{ij}}{\omega^2(\omega^2 + B_r^2)} \tag{2}$$

where the various coefficients R_{ij} are now given by

$$R_{12} = R_s + \delta R_{12} \quad \text{with } \delta R_{12} = \frac{C_{12}\tau_D}{A}$$

$$R_{22} = R_s + R_d + \delta R_{22} \quad \text{with } \delta R_{22} = -\frac{C_{11}\tau_D}{A}$$

$$R_{11} = R_s + R_g + \delta R_{11} \quad \text{with } \delta R_{11} = -\frac{\tau_G\tau_D g_d + C_{22}\tau_G}{A}$$

$$R_{21} = R_s + \delta R_{21} \quad \text{with } \delta R_{21} = \frac{\tau_G\tau_D g_m + C_{21}\tau_G}{A}$$

with $A = C_{12}C_{21} - C_{11}C_{22} + \tau_D(g_m C_{12} - g_d C_{11})$ and δR_{ij} the NQS corrections.

One can then derive the formulas given in Table I for calculating the extrinsic and intrinsic model parameters in terms of the known fit-parameters A_{ij}, R_{ij}, L_{ij}, B_r^2, and G_{ij} [3] and the unknown source resistance R_s. For this two-τ NQS model, there exists therefore a continuum of solutions as a function of R_s which yield the exact same fit to the Z_{ij} or S_{ij} parameters. However, note that the capacitances C_{11} and C_{12}, the extrinsic g_D and g_M, and the parasitic inductances L_s, L_g, and L_d are independent of R_s and hence uniquely extracted.

III. EXPERIMENTAL RESULTS

For this work we have used Motorola's MRF 181 RF power, n-channel enhancement mode, lateral MOSFET [9]. Isothermal microwave data were acquired for various bias points and the text fixture contribution deembedded using a TRL calibration procedure.

First the regression procedure of Raskin *et al.* [4] was used to establish the linear relationships giving the R_{ij} and L_{ij} coefficients in terms of R_{11} and L_{12}, respectively. Next the independent coefficients R_{11}, A_{ij} and B_r^2 were obtained from a simultaneous least-square-fit of all the $\mathrm{Re}[Z_{ij}]$. Finally, G_{21} and L_{12} were obtained from a simultaneous fit of all the $\mathrm{Im}[Z_{ij}]$ using in addition the theoretical model constraints: $A_{ij} = \pm g_{ij}/A + B_r B_{ij}$. A typical fit obtained is shown in Fig. 2.

Once the extrinsic parameters $Z_{extr,ij}$ are fitted, the extrinsic and intrinsic model parameters are calculated as a function of R_s using the expressions given in Table I. Fig. 3 shows
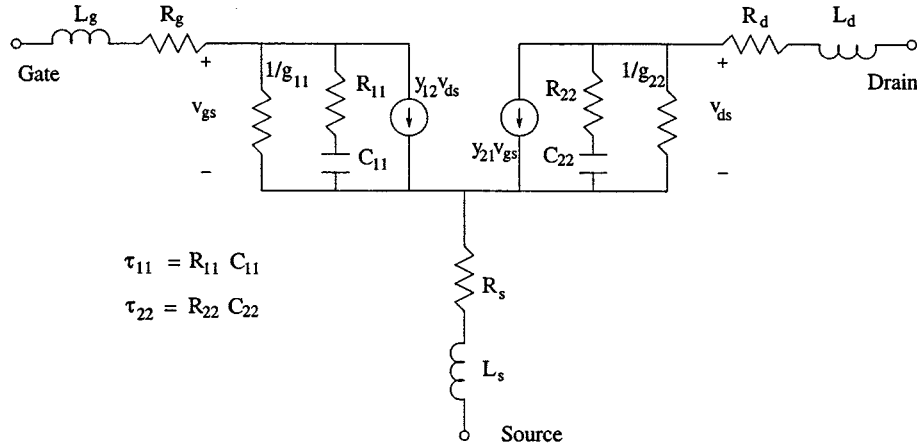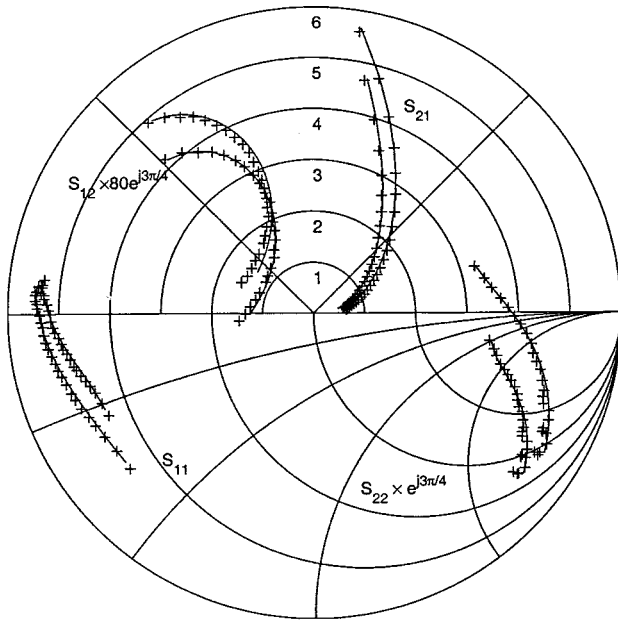
Fig. 1.   NQS small-signal equivalent-circuit for the high-frequency response

Fig. 2.   Extrinsic S-parameters measured from 0.4–2 GHz for 2 different bias points in saturation ($V_{GS} = 4$ and 4.75 V).

TABLE I
MODEL PARAMETERS IN TERMS OF THE Z_{ij} FIT PARAMETERS

$L_s = L_{12} = L_{21}$	$L_g = L_{11} - L_s$
$L_d = L_{22} - L_s$	$B_r = \frac{B_{11}B_r^2 - G_{11}}{A_{11}}$
$C_{11} = \left\{ -\frac{G_{11}}{B_r^2} + \frac{A_{12}}{A_{22}} \frac{G_{21}}{B_r^2} \right\}^{-1}$	$C_{12} = \left\{ -\frac{G_{21}}{B_r^2} + \frac{A_{22}}{A_{12}} \frac{G_{11}}{B_r^2} \right\}^{-1}$
$g_M = \frac{-G_{21}}{G_{11}(R_{22}+A_{22}/B_r^2) - G_{21}(R_{12}+A_{12}/B_r^2)}$	$g_D = \frac{G_{11}}{G_{11}(R_{22}+A_{22}/B_r^2) - G_{21}(R_{12}+A_{12}/B_r^2)}$
$R_d = \left(-1 + \frac{A_{22}}{A_{12}}\right) R_s + \left(R_{22} - R_{12}\frac{A_{22}}{A_{12}}\right)$	$\tau_D = \frac{R_s - R_{12}}{B_r(R_{12}-R_s+A_{12}/B_r^2)}$
$c_{11p} = -\frac{A_{22}}{A_{12}} R_s + \left[\frac{A_{22}}{B_r^2} + R_{12}A_{22}\right]$	$c_{12p} = R_s - R_{12} - \frac{A_{12}}{B_r^2}$
$B = -\frac{B_r^2}{c_{11p}G_{11}+c_{12p}G_{21}}$	$R_{p21} = (R_s - R_{21})\frac{B}{B_r}$
$g_m = \frac{BG_{21}}{B_r^2}$	$g_d = -\frac{BG_{11}}{B_r^2}$
$I_{p21} = B_{21}\frac{B}{B_r} + B(R_{21}-R_s)$	$\tau_G = -\frac{I_{p21}}{2g_m} \pm \left(\left(\frac{I_{p21}}{2g_m}\right)^2 + \frac{R_{p21}}{g_m}\right)^{1/2}$
$C_{21} = -[I_{p21} + (\tau_D + \tau_G)g_m]$	$C_{22} = -\frac{B}{C_{11}}\left(\frac{1}{B_r} - \frac{C_{12}C_{21}}{B} + \tau_D\right)$
$R_{p11} = \tau_G\tau_D g_d + \tau_G C_{22}$	$R_g = R_{11} - R_s + \frac{R_{p11}B_r}{B}$

for different biases (V_{GS}, V_{DS}), the extracted R_d and R_g as functions of R_s. **The R_d and R_g loci are limited to values of R_s giving the correct physical sign for all parameters.** Note that the bias points were selected to be in the saturation region. The same line-symbol is used for curves with the same V_{GS}. The loci of R_d versus R_s indicate that a common R_d value is obtained for $R_s = 0.136 \pm 0.002$ Ω for all gate and drain voltages. The drain resistance at the intercept point slightly increases from $R_d(V_{GS} = 4V, V_{DS}) = 0.67\pm0.005$ Ω to $R_d(V_{GS} = 4.75 V, V_{DS}) = 0.785 \pm 0.015$ Ω as the gate voltage varies from 4–4.75 V **and is fairly insensitive to the drain voltage. This indicates that in saturation, the drain resistance is effectively a nonlinear function of the saturation drain-current.** No intercept is found from the R_g loci.

The absence of intercept originates from both the sensitivity of the extracted R_g loci to the data noise in the measured S-parameters and the weak dependence of R_g upon R_s (R_g loci are horizontal). From the vertical distribution of R_g loci we estimate that the average value of the gate resistance is $R_g = 0.75\pm0.05$ Ω which still provides an acceptable margin of error.

The values for the parasitics using the QS regression method [4] were verified to correlate well with the NQS analysis, however the present analysis also yields the NQS time-constants: $\tau_G = 1.830$ ps and $\tau_D = 0.668$ ps which are known to be difficult to extract [5]. Note that δR_{21} is in average 42% of R_{21}. On the other hand, δR_{12} is in average only 5.6% of R_{12} so that for the particular device considered here, the assumption $R_s = R_{12}$ instead of the multi-bias analysis, could have alternately been used in the equations of Table I for calculating all the remaining NQS model parameters.
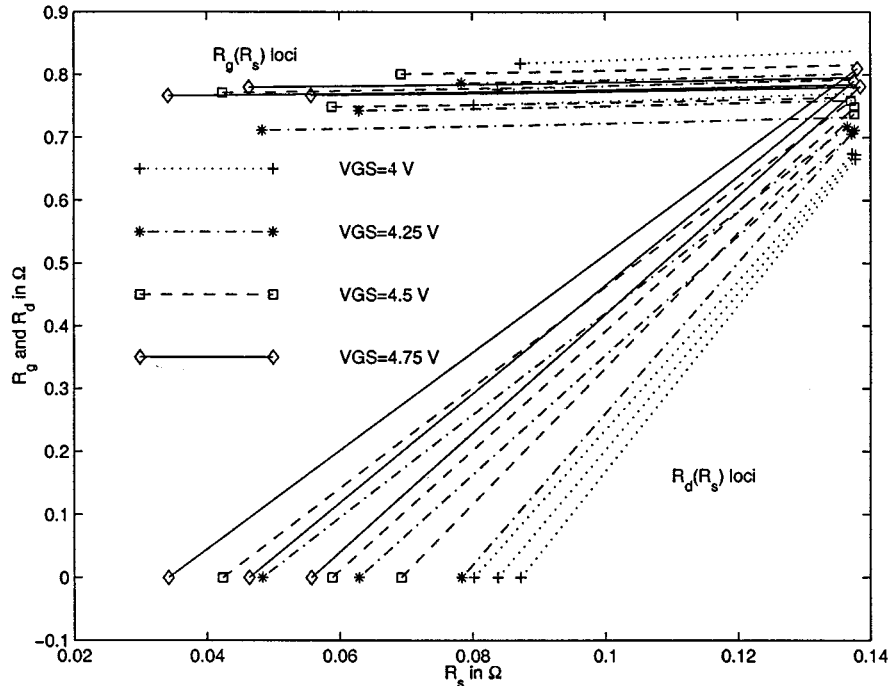
Fig. 3.   Loci of R_d and R_g as a function of R_s for various gate and drain voltages. The same line-symbol is used for the loci of different V_{DS} with the same V_{GS}. Note that the loci are limited to values of R_s giving the correct physical sign for all parameters. The source resistance at all the R_d intercept point is about $R_s = 0.136\ \Omega$. The drain resistance R_d at the R_d intercept point slightly increases from 0.67 to 0.785 Ω as V_{GS} increases from 4 to 4.75 V.

IV. CONCLUSION

We have presented a new extraction theory which provides analytic formulas to extract the two time-constants of a NQS model simultaneously with all the parasitics and other intrinsic model parameters. It is to be noted, however, that the assumption $\tau_{12} = \tau_{11}$ and $\tau_{21} = \tau_{22}$ used by the charge-based NQS model selected might not be necessarily applicable to all FETs. However, realistic results were obtained for various LDMOSFETs in spite of the very small values of the parasitics resistances in these power devices.

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