

Non-Quasi-Static RF Model for SOI FinFET and Its Verification

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Abstract—The radio frequency (RF) model of SOI FinFETs with gate length of 40 nm is verified by using a 3-dimensional (3-D) device simulator. This paper shows the equivalent circuit model which can be used in the circuit analysis simulator. The RMS modeling error of Y-parameter was calculated to be only 0.3 %.

Index Terms—SOI FinFET, radio frequency, model, parameter extraction, model verification

I. INTRODUCTION

Recently, the structures of various multi-gate field-effect transistors have been proposed. The SOI FinFET is a promising candidate for RF applications because of advantages such as no junction capacitance and substrate coupling signal [1-5]. In this device, the channel region is surrounded by the tri-gate. The FinFET structure shows the immunity to the short channel effect and the high transconductance characteristics. The RF modeling of SOI FinFETs is an essential step for the design of high-frequency integrated circuits [5]. The quasi-static (QS) model breaks down when the input signal changes too fast. If the gate signal is varying very fast, the inversion layer charge does not have enough time to respond fully. Therefore, the response of inversion layer charges lags behind the input signal [7]. In order to exactly express the lagging effect for high frequency, the non-quasi-static effects have to be reflected in the RF model of transistors. The RF models of FinFETs have been reported in [3] and [5]. The RF model in [3] is a QS

model. In the previous NQS model [5], the source and drain resistances (R_s and R_d) were located inward the overlap capacitance. However, the outer parts of R_s and R_d are dominant for the thin body FinFETs and therefore a new equivalent circuit with R_s and R_d connected outside of the overlap capacitance is used. Also, new time constants (τ_{m2}) are added to improve the model accuracy. Generally, the second order terms of denominators of Y -parameters are neglected for the easy analysis of Y -parameters [7]. As the operation frequency increases, however, the values of the second order terms become large. Therefore, we include the second order term ($\omega^2 \tau_{m2}^2$) in the small-signal equivalent circuit model. In this paper, we verify the RF model of SOI FinFET with time constants to improve the model accuracy. Accuracy of the model and the extraction method are verified by using the 3-D device simulation data up to 200 GHz.

II. DEVICE STRUCTURE AND RF MODEL

Fig. 1 shows the schematic of the SOI FinFETs with tri-gate structure. The values of physical gate length (L_G), gate height (H_G), and fin width (W_{Fin}) for the simulated structure are 40 nm, 60 nm, and 20 nm, respectively. The cut-off frequency of the device at $V_{GS} = V_{DS} = 1$ V is 248 GHz as shown in Fig. 2. Fig. 3 (a) shows the non-quasi-static (NQS) model of the SOI FinFETs. R_{gs} and R_{gd} are the distributed channel resistances. C_{gs} and C_{gd} are the intrinsic gate-to-source and gate-to-drain capacitances. C_{gd0} and C_{gs0} are the extrinsic gate-to-drain and gate-to-source capacitances, respectively. R_s and R_d are the source and drain resistance, respectively. C_{sdx} is the capacitance due to the DIBL effect in the short-channel MOSFET [6]. In order to accurately describe the NQS

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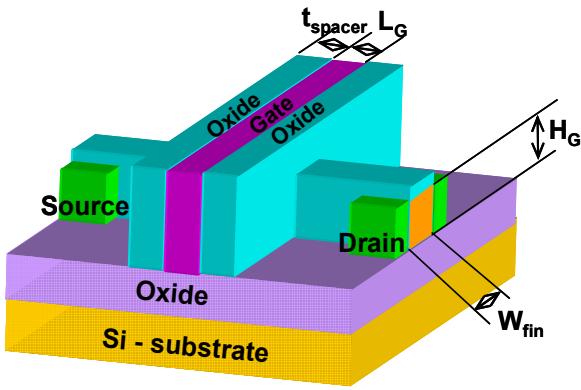


Fig. 1. The 3-D structure for SOI FinFET.

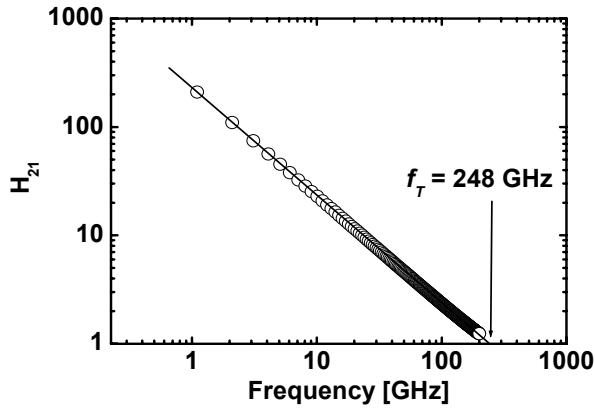


Fig. 2. Frequency-dependent current gain H_{21} of the SOI FinFETs at $VDS = VGS = 1$ V. Cut-off frequency is extracted to be 248 GHz.

effects for transistor, the time constants τ_{m1} and τ_{m2} by the transport delay of charges in the channel region are included in the RF equivalent circuit model [7]. Y_m with complex coefficient is given by following equation.

$$Y_m = \frac{g_m}{1 + j\omega\tau_{m1} - \omega^2\tau_{m2}^2} \quad (1)$$

C_{gd0} , C_{gs0} , R_s , and R_d are obtained by using the method presented in [5]. In order to simplify the Y-parameter equations, the assumptions that $\omega^2 R_{gs}^2 C_{gs}^2 \ll 1$, $\omega^2 R_{gd}^2 C_{gd}^2 \ll 1$, $\omega^2 \tau_{m1}^2 \ll 1$, and $\omega^2 \tau_{m2}^2 \ll 1$ are used. After the de-embedding of C_{gd0} , C_{gs0} , R_s and R_d , the Y-parameters of equivalent circuit can be approximated as follows.

$$\text{Re}(Y_{11}^{\text{int}}) \approx \omega^2 (R_{gs} C_{gs}^2 + R_{gd} C_{gd}^2) \quad (2)$$

$$\text{Im}(Y_{11}^{\text{int}}) \approx \omega (C_{gs} + C_{gd}) \quad (3)$$

$$\text{Re}(Y_{12}^{\text{int}}) \approx -\omega^2 R_{gd} C_{gd}^2 \quad (4)$$

$$\text{Im}(Y_{12}^{\text{int}}) \approx -\omega C_{gd} \quad (5)$$

$$\text{Re}(Y_{21}^{\text{int}}) \approx g_m - \omega^2 (g_m \tau_{m2}^2 + R_{gd} C_{gd}^2) \quad (6)$$

$$\text{Im}(Y_{21}^{\text{int}}) \approx -\omega (C_{gd} + g_m \tau_{m1}) \quad (7)$$

$$\text{Re}(Y_{22}^{\text{int}}) \approx g_{ds} + \omega^2 R_{gd} C_{gd}^2 \quad (8)$$

$$\text{Im}(Y_{22}^{\text{int}}) \approx \omega (C_{gd} + C_{sdx} - g_{ds} \tau_{m3}) \quad (9)$$

Y^{int} means the Y-parameters of the intrinsic device after de-embedding R_s , R_d , C_{gd0} , and C_{gs0} . The small-signal parameters C_{gd} , C_{gs} , R_{gd} , R_{gs} , g_m , g_{ds} , τ_{m1} , τ_{m2}^2 and C_{sdx} can be expressed by (10)-(18).

$$C_{gd} = -\frac{\text{Im}(Y_{12}^{\text{int}})}{\omega} \quad (10)$$

$$C_{gs} = \frac{\text{Im}(Y_{11}^{\text{int}}) + \text{Im}(Y_{12}^{\text{int}})}{\omega} \quad (11)$$

$$R_{gd} = -\frac{\text{Re}(Y_{12}^{\text{int}})}{\omega^2 C_{gd}^2} \quad (12)$$

$$R_{gs} = \frac{1}{C_{gs}^2} \left[\frac{\text{Re}(Y_{11}^{\text{int}})}{\omega^2} - R_{gd} C_{gd}^2 \right] \quad (13)$$

$$g_m = \text{Re}(Y_{21}^{\text{int}}) \Big|_{\omega^2=0} \quad (14)$$

$$g_{ds} = \text{Re}(Y_{22}^{\text{int}}) \Big|_{\omega^2=0} \quad (15)$$

$$\tau_{m1} = -\frac{1}{g_m} \left[\frac{\text{Im}(Y_{21}^{\text{int}})}{\omega} + C_{gd} \right] \quad (16)$$

$$\tau_{m2}^2 = \frac{-\text{Re}(Y_{21}^{\text{int}}) + g_m - \omega^2 R_{gd} C_{gs}^2}{\omega^2 g_m} \quad (17)$$

$$C_{sdx} = \frac{\text{Im}(Y_{22}^{\text{int}})}{\omega} - C_{gd} + g_{ds} \tau_{m3} \quad (18)$$

Fig. 3(b) shows the equivalent circuit to use the model directly in circuit analysis simulator such as HSPICE. In the proposed equivalent circuit, the coefficient of the voltage-controlled current source for g_m is complex. We could simulate the model by adding several passive elements (C_I , L_I , and R_I) to the equivalent circuit of Fig. 3 (a). The relationship between V_{gs} of Fig. 3(a) and V_I of Fig. 3(b) can be represented by (19).

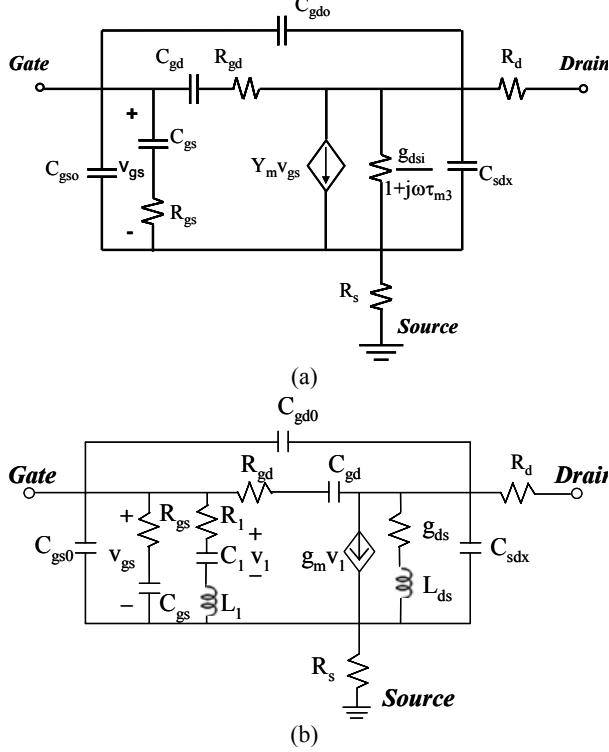


Fig. 3. RF equivalent circuit models. (a) The RF small-signal equivalent circuit of SOI FinFET. Y_m has the complex coefficient with τ_{m1} and τ_{m2} . (b) The model was modified to remove the complex coefficients.

$$\begin{aligned}
 g_m v_1 &= -\frac{\frac{1}{j\omega C_1}}{R_1 + j\omega L_1 + \frac{1}{j\omega C_1}} g_m v_{gs} \\
 &= \frac{1}{1 + j\omega R_1 C_1 - \omega^2 L_1 C_1} g_m v_{gs} \quad (19) \\
 &= \frac{1}{1 + j\omega \tau_{m1} - \omega^2 \tau_{m2}^2} g_m v_{gs}
 \end{aligned}$$

In order not to upset the model, the new elements must draw a negligible current [7]. C_1 , L_1 , and R_1 can be extracted to be $0.001C_{gs}$, τ_{m2}^2/C_1 , and τ_{m1}/C_1 . We use the relationship between C_1 and $0.001C_{gs}$ as presented in [7]. Because C_1 is much smaller than C_{gs} , the impedance due to C_1 is very large. Therefore, the new elements draw a negligible current. C_1 , L_1 , and R_1 are not the physical parameters. These three added parameters are added to simulate the model directly in the circuit analysis simulator such as HSPICE. The time constant τ_{m3} of g_{ds} term in Fig. 3(a) are expressed as $\tau_{m3} = L_{ds}g_{ds}$ [7]. By the Y-parameter model in [7] (as shown in page 481), τ_{m3} is

equal to τ_{m1} . After the de-embedding of C_{gd0} , C_{gs0} , R_s , and R_d , the extraction of the small-signal parameters was performed. All the extracted parameters are summarized in Table 1. Using the extracted parameters, $\omega^2 R_{gs}^2 C_{gs}^2$, $\omega^2 R_{gd}^2 C_{gd}^2$, $\omega^2 \tau_{m1}^2$, and $\omega^2 \tau_{m2}^2$ are calculated to be 4.47×10^{-5} , 1.67×10^{-5} , 1.03×10^{-3} , and 1.46×10^{-4} at 32 GHz , respectively. Fig. 4 compares the results of the proposed model with 3-D simulation. The device was biased at $V_{GS} = V_{DS} = 1\text{ V}$.

It shows that the modeled Y -parameters (solid line) by using the extracted parameters fit the 3-D simulated ones (symbol: \circ) well. The RMS error between the 3-D simulation and the modeled Y -parameters is only 0.3 % up to 200 GHz without any optimization after the parameter extraction step. The RMS error of the previous model ([5], symbol: Δ) is 1.1 %. The error of model is reduced in this work due to the addition of the second order time constant and the change of location of R_s and R_d . The difference of errors between the proposed and previous models in [5] is 0.8 %. Among this quantity, 0.65 % is contributed by the modification of R_s/R_d location and 0.15 % is contributed by the addition of $\omega^2 \tau_{m2}^2$. The effect of $\omega^2 \tau_{m2}^2$ is prominent at Y_{21} -parameters. At 200 GHz , $\omega \tau_{m1}$ and $\omega \tau_{m2}$ are calculated to be 0.207 and 0.006 using the extracted parameters in Table 1, respectively. $\omega^2 \tau_{m2}^2$ is about 3 percent of $\omega \tau_{m1}$. Although $\omega^2 \tau_{m2}^2$ is much smaller than $\omega \tau_{m1}$, the error of Y_{21} -parameters at 200 GHz falls from 2.7 % to 1.2 % by adding $\omega^2 \tau_{m2}^2$. In case that $\omega^2 \tau_{m2}^2$ is added to the previous model in [5], the model error up to 200 GHz falls from 1.1 % to 0.95 %. In the that R_s and R_d are connected outside the overlap capacitances, the error of model falls from 0.95 % to 0.3 %. The change of location of R_s and R_d is a dominant factor for the reduction of model error.

III. CONCLUSIONS

A compact and accurate RF model of the SOI FinFETs for the strong inversion operation region was presented. The procedure of parameter extraction consisted of 3-D simulation, de-embedding of extrinsic parameters, Y -parameter analysis, simplification of Y -parameters by several assumptions, derivation of extraction equations,

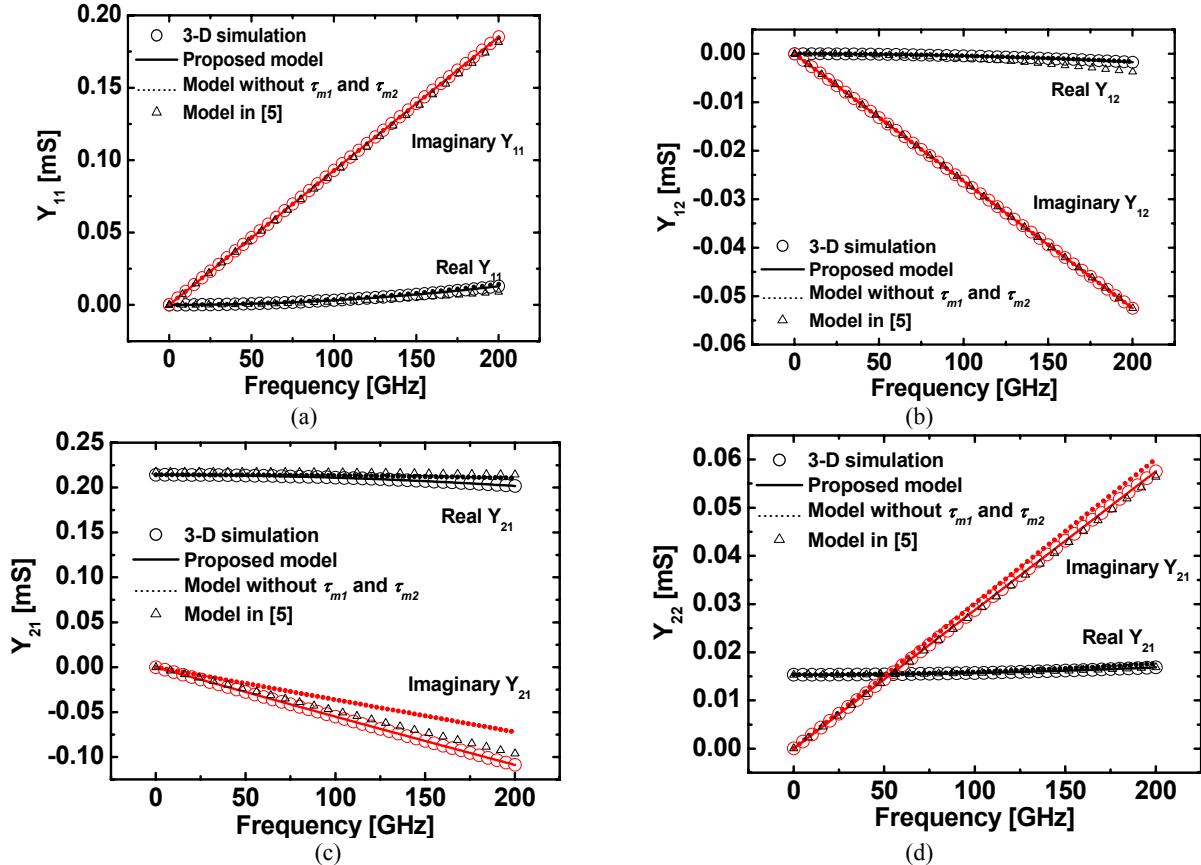


Fig. 4. The comparison of proposed model (solid line), model without time constants (dashed line), previous model (symbol: Δ), and 3-D simulation (symbol: o) at $V_{GS} = V_{DS} = 1$ V. (a) Y_{11} , (b) Y_{12} , (c) Y_{21} , and (d) Y_{22} .

and extractions. Without any complex fitting and optimization steps, the total modeling error of Y -parameter was calculated to be only 0.3 % up to 200 GHz.

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