High-frequency compact analytical noise model for double-gate metal-oxide-semiconductor field-effect transistor

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Silicon-on-insulator (SOI) metal-oxide-semiconductor field-effect transistors (MOSFETs) are excellent candidates to become an alternative to conventional bulk technologies. The most promising SOI devices for the nanoscale range are based on multiple gate structures such as double-gate (DG) MOSFETs. These devices could be used for high-frequency applications due to the significant increase in the transition frequency \( f_T \) for these devices. For low noise radiofrequency and microwave applications, high-frequency noise models are required. In this work, we present compact expressions to model the drain and gate current noise spectrum densities and their correlation for DG MOSFETs. These expressions depend on the mobile charge densities that are obtained using analytical expressions obtained from modeling the surface potential and the difference of potentials at the surface and at the center of the Si doped layer without the need to solve any transcendental equations. Using this model, the DG MOSFET noise performances are studied. The current and noise models can be easily introduced in circuit simulators. © 2009 American Institute of Physics. [DOI: 10.1063/1.3077279]

I. INTRODUCTION

The use of low power, low noise devices for future electronic applications is becoming more and more important. Especially, silicon-on-insulator (SOI) metal-oxide-semiconductor field-effect transistors (MOSFETs) are excellent candidates to become an alternative to conventional bulk devices. The most promising SOI devices for the nanoscale range are based on multiple gate structures such as the double-gate (DG), triple gate, or fin-shaped field-effect transistors (FinFETs), and surrounding gate or gate-all-around (GAA). DG MOSFET is a very attractive option due to the improvement of the performance in the sub-50 nm gate length,1 favored by the stronger control of the channel by the gate compared to the single-gate transistor. The drain-induced barrier lowering (DIBL), the threshold voltage rolloff, and the off-state leakage can also be significantly reduced.2 With the evolution in MOSFET scaling to shorter channel lengths, the high-frequency (HF) capabilities of the transistor have reached into the gigahertz regime so that rf circuit applications have been steadily growing.3 In addition, new application markets in the microwave and millimeter wave regions have emerged.

For these features, the development of accurate and Computer Aided Design (CAD) compatible compact models are needed.4,4 Although several theoretical, process, and physics-based analyses have been reported related to the dc properties of the undoped devices,5–8 the limitation or degradation of the HF and noise characteristics along the down-scaling of the channel length has only been described by a few authors.9–12 At HF, the MOS transistor noise is mainly dominated by thermal noise (also called Johnson noise or Nyquist noise) coming from the channel.12 The term thermal is due to the origin of this noise, which can be traced to the random thermal motion of carriers in the channel. Accurate noise modeling is a prerequisite for the application of advanced MOSFET technologies to low noise rf design.3,4

In previous papers,9–11 we presented an analytical method for rf and noise model applied to DG SOI Metal Semiconductor Field Effect Transistor (MOSFET) devices. These analyses are based on the segmentation method. In the segmentation method or active transmission line analysis, the channel is split into channel sections or slides. The small-signal parameters and noise source expressions for each channel section can be derived from basic semiconductor equations. This method was used for noise modeling in other devices such as MESFETs,13 High Electron Mobility Transistors (HEMTs),14 MOSFET, Single Gate (SG) SOI,9,15 DG MOSFETs,9–11 GAA,11,16 and FinFETs.17 The application of this method will be used to describe the small-signal and noise equivalent circuit. Although this analysis could be done using circuit analysis techniques9 that are implemented in standard microwave circuit simulators, it would be time consuming due to the large number of circuit elements to be analyzed. Hence, compact analytical expressions for the spectral noise densities are preferred for implementation in commercial circuit simulators.

In this paper, we present compact analytical expressions to model the gate and drain current noise spectrum noise densities and their correlation in short-channel symmetric DG MOSFETs. These expressions depend on the mobile charge density and the drain current. We use here the com-
II. NOISE MODEL

A. Local noise sources and equivalent noise sources

The DG transistor structure under analysis is shown in Fig. 1, where \( N_a \) is the uniform acceptor concentration in the silicon layer with thickness equal to \( t_{Si} \); \( t_{ox} \) is the equivalent gate dielectric thickness; and \( L \) is the channel length. The transistor is symmetrical, with both gates connected together at \( V_G \). Several noise sources must be considered. The first source is the thermal noise due to parasitic access resistances \( R_G \), \( R_S \), and \( R_D \). Although the noise introduced by these resistances, especially \( R_G \), has a great impact on the final noise factor, their noise contribution is easily modeled using the Nyquist theorem. The other noise sources arise from the intrinsic transistor. The intrinsic contribution is modeled using two correlated current noise sources, \( i_s \) and \( i_d \) (admittance representation). They include the effect of diffusion noise and flicker noise in the channel, induced-gate noise, and shot noise. In principle, flicker noise is a low-frequency noise and it mainly affects the low-frequency performance of the device, so it can be ignored at very HF. However, the contribution of the flicker noise should be considered in designing some radiofrequency circuits such as mixers, oscillators, or frequency dividers that upconvert the low-frequency noise to higher frequency and deteriorate the phase noise or the signal-to-noise ratio. Channel resistance and all terminal resistances contribute to the thermal noise at HF, but typically the channel resistance dominates in the contributions of the thermal noise from the resistances in the device. Induced-gate noise is generated by the capacitive coupling of local noise sources within the channel to the gate, and usually it plays a more important role as the operation frequency goes much higher than the frequency at which channel thermal noise dominates. With the decrease in the oxide thickness due to gate downscaling, the dc tunneling gate current flowing through the oxide in the gate is increasing. This gate current produces an increase in noise due to the shot noise current. In this paper, we will focus on the modeling of the thermal channel noise, the induced-gate noise, and the correlation coefficient.

Let us consider a nonuniform channel as shown in Fig. 2. The channel is split into channel sections or slides, and the small-signal and noise source for each channel section can be derived from semiconductor equations. The local equivalent circuit (Fig. 2) for each channel slide is composed of the gate-to-channel capacitance, the transconductance, and the channel resistance (or conductance). Diffusion noise and gate shot noise can be incorporated into the model by adding two current noise sources for each slide. The local equivalent circuit elements can be obtained from linearizing the current at any channel position \( x \), resulting in the current continuity equation

\[
I(x) = g \left( V_G \frac{dV}{dx} \right) \frac{dV}{dx},
\]

where \( g = W \mu Q \) and \( W, \mu, \) and \( Q \) are the channel width, mobility, and inversion charge density, respectively. The mobility depends on the electric field \( E = -dV/dx \) and \( g \) depends on the channel potential \( V(x) \) and the \( dV/dx \).

The local equivalent circuit elements for a channel slide between \( x \) and \( x + \Delta x \) are given as

\[
C_{gs}(x) = W\Delta x \frac{dQ}{dV_{gc}},
\]

\[
g_m(x) = W\Delta x \frac{dQ}{dV_{gc}} v(x),
\]

\[
g_{ad}(x) = \frac{g(x) + g'(x)E(x)}{\Delta x},
\]

where \( V_{gc}(x) = V_G - V(x) \) is the gate-to-channel potential and \( g' = dg/dE = WQ \mu \mu / dE \).
For HF noise analysis, the channel diffusion noise source and the gate shot noise for each slide are introduced.

Modeling the channel thermal noise of a MOSFET has been an active area of research in recent years. The method to calculate the local noise source due to the carrier fluctuation caused by the diffusion noise in nonequilibrium is to introduce an Einstein-like relationship between the differential mobility and diffusion coefficient:

\[ D = \frac{kT_n}{q} \mu_d, \quad (5) \]

where \( D \) is the nonequilibrium noise diffusivity, \( T_n \) is the noise temperature, and \( \mu_d \) is the differential mobility (\( \mu_d = dv/dE \)). For the typical values of the inversion carrier density in the inversion layer, we can consider that the channel is not degenerated, and the velocity distribution is heated Maxwellian. Under this assumption, following Ref.24 the noise temperature \( T_n \) becomes equal to the carrier temperature \( T_e \), and the spectral current density can be written as:

\[ \overline{I_n^2} = 4qQ(x)D(E) \frac{W}{\Delta x} = 4kT_ng_{cb}(x). \quad (6) \]

Under high channel electric fields, the temperature of electrons in the channel can rise above the lattice temperature. This effect can increase the thermal noise of the device. However, in the drift-diffusion models, an analytical relationship between \( T_e \) and the lateral field \( E(x) \) must be used. One of the more challenging aspects of including the effect of \( T_e \) in the noise model is finding a good model for the electron temperature. A prevalent model in literature expresses the electron mobility as a function of the temperature as follows:

\[ \mu = \mu_e \sqrt{\frac{T_e}{T_s}} \quad (7) \]

Mobility can also be represented as a function of the electric field:

\[ \mu = \frac{\mu_e}{\left(1 + \left(\frac{E}{E_c}\right)^p\right)^{1/p}.} \quad (8) \]

Equating Eqs. (7) and (8) and solving for \( T_e/T_s \) give

\[ T_e = T_s \left(1 + \left(\frac{E}{E_c}\right)^p\right)^{2/p}. \quad (9) \]

For \( p=1 \) and \( p=2 \), Eq. (9) shows that electronic temperature has a quadratic dependence for high electric fields.

The analysis of the active transmission line (Fig. 2) using the admittance matrix gives the small-signal admittance matrix and its noise correlation matrix. Alternatively, the analysis could be performed using a standard microwave circuit simulator; however, the analysis is time consuming due to the large number of circuit elements. For the compact modeling of the noise, three methods are usually applied: (1) an equivalent circuit approach, (2) the impedance field method, or (3) the Langevin or Klaasen–Prins (KP) method. Reference 28 shows that the three methods are equivalent and the same final expression for the current noise densities and correlation coefficient was obtained. Thus, the compact modeling methods could be considered as an analytic analysis technique of the active transmission line.

Assuming that the local noise source is spatially uncorrelated and \( \omega \) is the angular frequency, the drain and gate spectral densities and the correlation matrix are obtained using the following expressions:

\[ S_{ig} = \overline{I_g^2} = \frac{1}{Lg_c^2} \int_{V_d} g_c(V)^2 \frac{dV}{g(V)} S_{n} dV, \quad (10) \]

\[ S_{g} = \overline{I_g^2} = \frac{1}{g_c^2} \int_{V_d} \left( \int_{V_s} g_c(V') [Q(V')] - Q(V) \right) \frac{g_c(V)^2}{g(V)} S_{n} dV, \quad (11) \]

\[ S_{g} = \overline{I_g^2} = \frac{1}{g_c^2} \int_{V_d} \left( \int_{V_s} g_c(V') [Q(V')] - Q(V) \right) \frac{g_c(V)^2}{g(V)} S_{n} dV, \quad (12) \]

where

\[ g_c(V,E) = \frac{g(V,E)}{g(V,E) + \frac{\partial g(V,E)}{\partial E} E}. \quad (13) \]

and the corrected length \( L_c \) is given by:

\[ L_c = \int_0^L \frac{g_c(V) dV}{g(V)} = \int_0^L \frac{g_c(V) dV}{g(V)}. \quad (14) \]

The power spectral density for the local noise source is given by

\[ S_{n}(x) = 4kT_L \frac{g(x)}{g_c(x)} \frac{T_n(x)}{T_L}. \quad (15) \]

B. Charge and dc models

In order to calculate the integrals with respect to the potential in Eqs. (10)–(12), we will obtain a relation between the mobile channel charge \( Q \) and the channel potential \( V \) at each channel point \( x \). The electric field at the surface of the silicon layer \( E_s \) is calculated using Poisson’s equation. The following expression is obtained for \( E_s \) as a function of the potential at the surface, \( \phi_0 \), and at the center of the layer, \( \phi_c \),
\[ E_s = \sqrt{\frac{2qN_d \phi_t}{\varepsilon_s}} \times \sqrt{\left( \frac{\phi_s - \phi_t}{\phi_t} \right) + (1 - e^{-\left[(\phi_s-\phi_t)/\phi_t\right]}) e^{(\phi_s-2\phi_f-V)/\phi_t}}. \]

(16)

where \( \phi_t=kT/q \) is the thermal potential, \( k \) is the Boltzmann constant, \( q \) is the electron charge, \( T \) is the temperature in kelvin, \( \varepsilon_s \) is the silicon dielectric permittivity, and \( \phi_F \) is the Fermi potential.

In Ref. 19, using a detailed numerical calculation, an empirical expression for the difference of potentials \( \phi_S - \phi_0 \) as a function of the potential difference \( V_D - V_0 \) was found. The surface electric field is analytically calculated using the Lambert function. The charge carrier concentration \( q_n \) along the channel \( Q \) normalized to \( C_{ox} \phi_t \) is determined through the following relation with the surface electric field at each interface:

\[ q_n = \frac{g_S E_s}{C_{ox} \phi_t} - \frac{q_b}{2}, \]

(17)

with

\[ q_b = \frac{g_S N_d \phi_t}{C_{ox} \phi_t}, \]

(18)

where \( t_s \) is the silicon thickness and \( C_{ox} \) is the oxide capacitance, \( C_{ox} = \varepsilon_{ox} / t_{ox} \).

For the typical operating voltage range in this type of transistors, the following relation can be applied:

\[ dV = -\phi_t \left[ \frac{1}{q_n + q_b} + \frac{1}{q_n + q_b} \right] dq_n. \]

(19)

Using Eq. (19), the channel charge can be integrated. Then, the drain current taking into account the velocity saturation and short-channel effects is given by Eq. (37) in Ref. 18 as

\[ I_D = \int_0^{V_D_{sat}} g(V) dV = \frac{2W \mu C_{ox} \phi_t^2}{L_c} \left( \frac{q_n^2 - q_d^2}{2} + 2(q_s - q_d) \right) - q_b \ln \left( \frac{q_n + q_b}{q_b + q_d} \right), \]

(20)

where \( q_n \) and \( q_d \) represent the normalized charge \( q_n \) evaluated at the source \( q_n(V=0) \) and at the effective drain voltage \( q_d(V=V_D) \), respectively, and \( L_c = L - \Delta L \), where \( \Delta L \) is the channel length modulation in the saturation region given by Eq. (36) in Ref. 18.

C. Compact expressions

Conventional compact models use the mobility-longitudinal field relation (8) with \( p=1 \); however, for an accurate description of the velocity saturation, we will use \( p = 2 \), as used in the SDDG model. Using the mobility model (8) with \( p=2 \), we obtain the following expression for the channel per unit length conductance (1):

\[ g(V) = \frac{WQ \mu_s}{\sqrt{1 + (E/E_c)^2}} = \frac{g_0}{\sqrt{1 + (E/E_c)^2}}, \]

(21)

with

\[ g_0 = WQ(V) \mu_s = 2W C_{ox} \phi_t \mu_s q_n(V). \]

(22)

From Eqs. (13) and (15), we obtain

\[ g_s(V) = g(V) [1 + (E/E_c)^2] = g_0 \sqrt{1 + (E/E_c)^2}. \]

(23)

\[ \frac{g_s(V) T_c}{g(V) T_L} = 1 + (E/E_c)^2. \]

(24)

\[ S_d = \frac{4kT_L}{I_D L_c^2} \int_0^{V_D} g_s(V) T_n dV = \frac{4kT_L}{I_D L_c^2} \int_0^{V_D} g_0 dV = \frac{4kT_L}{I_D L_c^2} (2W \mu_s C_{ox} \phi_t)^2 \int_0^{V_D} q_n^2 dV. \]

(25)

According to Ref. 21, the contribution of the velocity saturation region to the output noise current is negligible as the carriers in that region travel at their saturation velocity and they do not respond to the fluctuations of the electric field caused by the voltage noise in that region.21 Thus, integrals (10)–(12) and (14) must be integrated up to the saturation channel voltage or the effective voltage \( V_D \) and \( L \) in Eq. (14) must be replaced with \( L_c \) in short-channel devices to take into account the channel length modulation effect.

Substituting the expression of \( dV \) (19) in Eq. (25), the following compact expression for the drain noise spectral density is obtained:

\[ S_d = \frac{4kT_L}{I_D L_c^2} (2W \mu_s C_{ox} \phi_t)^2 F, \]

(26)

where

\[ F = -\phi_t \left[ \frac{q_n^3}{3} - q_n^2 - q_n q_b + q_b \ln(q_n + q_b) \right] \left. \right|_{q_n}. \]

(27)

Substituting Eq. (8) in Eq. (1), the following relations for the conductance and corrected conductance are found:

\[ g = \left[ \frac{g_0^2 - \left( \frac{I_D}{E_c} \right)^2}{2} \right]^{1/2} = (2W \mu_s C_{ox} \phi_t)^2 \left[ q_n^2 - q_d^2 \right]^{1/2}, \]

(28)

\[ g_c = \frac{g_0^2}{\left[ \frac{g_0^2 - \left( \frac{I_D}{E_c} \right)^2}{2} \right]^{1/2}} = (2W \mu_s C_{ox} \phi_t)^2 \left[ q_n^2 - q_d^2 \right]^{1/2}, \]

(29)

where \( q_a \) is defined as

\[ q_a = \frac{I_D}{2W \mu_s C_{ox} \phi_t E_c}. \]

(30)

Before evaluating the gate spectral density (11) and the cross spectral density (12), we need to calculate the integral

\[ \int_0^{V_D} \left[ \frac{g_0^2 - \left( \frac{I_D}{E_c} \right)^2}{2} \right]^{1/2} dV. \]
\[
\int_0^{V_D} g_c(V') \left[ Q(V') - Q(V) \right] dV' \\
= \int_0^{V_D} \left( 2W\mu_S C_{ox} \phi q_n^2(V) \right) \left[ q_n^2(V') \right] dV' \\
- \frac{q_n^2}{2\alpha} 2C_{ox} \phi \left[ q_n(V') - q_n(V) \right] dV' \\
= 2C_{ox} \phi 2W\mu_S C_{ox} \phi \left( \int_0^{V_D} q_n^2(V') \left[ q_n^2(V) \right] \right) dV' \\
- \frac{q_n^2}{2\alpha} 2C_{ox} \phi \left[ q_n(V') - q_n(V) \right] dV' \\
= 2C_{ox} \phi 2W\mu_S C_{ox} \phi \left[ A + B q_n(V) \right],
\]

where we used expression (19) to perform a change in variables to integrate with respect to the normalized charge \( q_n \).

\( A \) and \( B \) are defined using the functions \( q_S \) and \( q_D \) as

\[
A = - \phi_L \int \left[ \frac{q_n^2}{q_n^2 - q_S^2} \right] dV' \\
B = \phi_L \int \left[ \frac{q_n^2}{q_n^2 - q_D^2} \right] dV'
\]

Using Eq. (31) the following compact analytical expressions for Eqs. (12) and (13) are found:

\[
S_{11} = \frac{\omega^2 W^2}{P_L^2} \left( \int_0^{V_D} g_c(V') \left[ Q(V') - Q(V) \right] dV' \right)^2 \\
- \left( \int_0^{V_D} g_c(V') \left[ Q(V') \right] dV' \right)^2 \\
= \frac{\omega^2 W^2}{P_L^2 \Delta V} 4kT_L (2W\mu_S C_{ox} \phi_0^2 (2C_{ox} \phi_0)^2
\times \left( \int_0^{V_D} \left[ A + B q_n(V) \right] g_0^2(V) dV' \right)
\]

where

\[
\int_0^{V_D} \left[ A + B q_n(V) \right] g_0^2(V) dV' = A^2+C+2ABD+B^2E
\]

and the functions \( C, D, E \) are defined as

\[
C = \int_0^{V_D} q_n^2(V) dV' \\
D = \int_0^{V_D} q_n^3(V) dV' \\
E = \int_0^{V_D} q_n^4(V) dV'
\]

The cross noise spectral density is calculated by

\[
S_{11,c} = \frac{\omega^2 W^2}{P_L^2 \Delta V} \left( \int_0^{V_D} g_c(V') \left[ Q(V') - Q(V) \right] dV' \right)^2 \\
- \left( \int_0^{V_D} g_c(V') \left[ Q(V') \right] dV' \right)^2 \\
= \frac{\omega^2 W^2}{P_L^2 \Delta V} 4kT_L (2W\mu_S C_{ox} \phi_0^2 (2C_{ox} \phi_0)^2
\times \left( \int_0^{V_D} \left[ A + B q_n(V) \right] g_0^2(V) dV' \right)
\]

where
\[ \int_{0}^{V_D} (A + Bq_n(V))q_n^2(V)dV = A \int_{0}^{V_D} q_n^2(V)dV + B \int_{0}^{V_D} q_n^3(V)dV = AC + BD. \] (40)

Finally, the gate and cross spectral densities are given by

\[ S_{\nu G} = \frac{\omega^2 W^2}{2}\frac{4kT_L(2W\mu_0C_{ox}\phi_0)^3}{L_c^2}2(2C_{ox}\phi_0)^2(A^2C + 2ABD + B^2E), \] (41)

\[ S_{\nu GD} = \frac{j\omega W}{g_{d0}^2}4kT_L(2W\mu_0C_{ox}\phi_0)^3(2C_{ox}\phi_0)(AC + BD). \] (42)

These expressions depend on the drain current \( I_{DD} \), and expressions \( A, B, C, D, \) and \( E \), which are functions of \( q_s \) and \( q_D \). The length \( L_c \) is given by expression (14). Using Eqs. (21), (28), and (29), \( L_c \) is given by

\[ L_c = L - \frac{B}{G}, \] (43)

where

\[ G = \phi_i \int_{q_s}^{q_D} (q_n^2 - q_s^2)^{1/2} \left[ 1 + \frac{1}{q_n} + \frac{1}{q_n + q_b} \right] dq_n \]

\[ = \phi_i \left[ \frac{1}{2} (q_s^2 - q_n^2)^{1/2} - \frac{1}{2} q_n \ln[q_n + (q_s^2 - q_n^2)^{1/2}] + \frac{q_n^2}{(q_n^2 - q_s^2)^{1/2}} \ln[-2q_n^2 + 2(-q_n^2)^{1/2}(q_s^2 - q_n^2)^{1/2}] - \ln q_n + q_s^2 - q_n^2 \right] \]

\[ + G_1(q_s - q_n) + G_1^2(2q_s^2 - 2q_n^2 - 2q_n + q_b) + 2q_n(q_s^2 - q_n^2)^{1/2}G_1^2(2q_s^2 - 2q_n^2 - 2q_n + q_b) \] (44)

and

\[ G_1 = (q_n + q_b)^2 - 2(q_n + q_b)q_b + q_n^2 - q_b^2. \] (45)

Although the spectral densities (26), (41), and (42) allow circuit simulators to analyze noise performances, in order to compare different devices, some noise parameters have been proposed in literature for MOSFETs. Following Van der Ziel notation,\(^{20}\) the drain current noise spectrum density is expressed as

\[ S_{\nu D} = 4kT_Lg_{d0}g_s^2, \] (46)

where \( 4kT_L^2g_{d0} \) is the current noise spectrum associated with a conductance \( g_{d0} \) at temperature \( T_L \). In the case of a MOSFET, however, since a channel conductance \( g_{d0} \) depends on a bias voltage, \( g_{d0} \) is fixed to the value obtained when \( V_{DS} = 0 \).

As mentioned above, the physical origin of the induced-gate noise \( S_{\nu G} \) is the capacitive coupling of the channel conductance to the gate capacitance. Based on this assumption, Van der Ziel\(^{20}\) derived a simple expression

\[ S_{\nu G} = \frac{4kT_Lg_{d0}g_s^2}{g_{d0}^2} \left( \frac{\omega C_{gs}}{5g_{d0}} \right)^2, \] (47)

where \( C_{gs} \) is the gate-source capacitance. The expression is valid in the saturation region for frequencies below 1/3 of the cut-off frequency \( f_T \) and this frequency range is sufficient for most real applications of the device. Theoretically, the gate excess noise factor, \( \beta \), tends to 4/3 and the imaginary part of coefficient \( \text{Im}(C) \) tends to 0.4 in saturation for the case of long-channel MOSFETs.

### III. RESULTS AND DISCUSSION

In order to verify the compact expressions presented in Sec. II, we have compared the drain and gate current noise spectral densities, and the imaginary part of correlation coefficient calculated using the segmentation method and the compact model SDDG. Figure 3 shows the results as a function of drain voltage, for a long-channel \( (L=1 \ \mu m) \) case.
using a constant mobility without transversal degradation. A good agreement between the two methods is obtained in all the range of interest.

Compact MOSFET noise models are mostly based on the conventional KP method. The main difference between the improved KP method and the equivalent circuit is that the conventional KP method neglects the lateral nonuniformity of the mobility along the channel. In conventional KP method, the drain spectrum density is calculated as Eq. (25) with \( g_\tau = g \) and \( L_c = L \) (or the effective length \( L_c \) in the short-channel case) as follows:

\[
S_d = \frac{4kT_L}{I_D L^2} \int_0^{V_D} g^2(V) dV
\]

Following a similar procedure, integral (48) could be analytically calculated as function of \( q_S \) and \( q_D \) using Eq. (19).

Figure 4 compares the results obtained using the segmentation method, the compact analytical model SDDG, and the analytical model using conventional KP approach (48) for a SDDG MOSFET with \( L = 0.5 \mu m \) and \( V_{DS} = 2 \) V. A good agreement between the three methods is obtained for low drain currents. When \( V_{GS} \) is increased, \( g \) differs more from \( g_0 \), thus Eqs. (48) and (25) produce different results. Moreover, when the gate length \( L_c \) decreases, it differs from the effective length \( L_c \). Figure 5 shows the normalized corrected length \( L_c/L_e \), which tends to 1 in linear region and increase with channel length reduction. As a conclusion the noise model based on the conventional KP method is not a good approach to model channel noise for short-channel devices.

The bias behavior of excess noise factors and correlation coefficient for different channel lengths are shown in the next figures. Now, we consider transversal field mobility degradation and velocity saturation. Figures 6–8 show the drain and gate excess noise factors, and the imaginary part of the correlation coefficient as a function of drain voltage \( t_\tau = 34 \) nm, \( t_{ox} = 2 \) nm, and \( V_{GS} = 1 \) V, respectively [calculated by Eqs. (26), (41), and (42)]. These effects introduce some changes with respect to the classical Van der Ziel long-channel MOSFET model. Due to the velocity saturation effect, \( V_{DSsat} \) is smaller as the channel length scales down at high gate voltages, which results in the slight increase in thermal noise compared to that of the long channel as shown in Fig. 6. This result is in good agreement with the trend determined experimentally in a single-gate MOSFET. The \( \gamma \) factor tends to 1 for \( V_{DS} = 0 \) and its value increases steadily with the drain bias in the saturation regime, which is attributed to the channel length modulation effect through the corrected length \( L_c \), the potential distribution, and the related mobility distribution along the channel. Such increase becomes more critical for devices with smaller channel lengths, because the channel length modulation has larger effects for

\[
C = \frac{1}{\sqrt{(g_\tau - g_0)^2 + \left(\frac{g''_0}{g_0}\right)^2}}
\]
the short-channel devices. This bias behavior was found in experimental results for single-gate MOSFETs.\textsuperscript{12,20} In Fig. 7, the $\beta$ factor tends to 4/3 for the long-channel case and increases with the reduction in $L$. Figure 8 shows our simulation results of the imaginary part of the correlation coefficient. They are slightly smaller than the value 0.4 for the long-channel case, and reduce to zero at $V_{DS}$=0. This coefficient decreases with channel length reduction.

For DG transistors with $t_S$=34 nm and $t_{ox}$=2 nm, Figs. 9–11 give the drain and gate excess noise factors and the imaginary part of the correlation coefficient as a function of gate voltage, at a fixed drain voltage of 2V. The drain excess noise factor $\gamma$ increases with the gate voltage for all channel lengths (Fig. 9). The gate excess noise factor $\beta$ (Fig. 10) is found to depend strongly on the gate voltage for short-channel devices and it can exceed the value of the long-channel case considerably. The behavior of the imaginary part of the correlation coefficient as a function of gate voltage is the same for all channel lengths (Fig. 11). However, it decreases with the reduction in the channel length.

IV. CONCLUSIONS

In this paper we present a compact analytical noise model using the new compact analytical model for short-channel SDDG MOSFETs, which considers doped silicon layer in a wide range of doping concentrations and short-channel transistors. We have developed a compact model for the channel noise, the induced-gate noise, and the cross-noise between drain and gate noises, based on the improved KP approach. This method had been found equivalent to an equivalent circuit approach, and the impedance field method. The expressions obtained are analytical and they depend on the mobile charge at the source and drain ends of the channel.

The noise compact model calculations are compared with the values obtained using the segmentation method, where the active transmission line is analyzed using circuit nodal analysis. The good agreement obtained between both methods validates the analytical expressions presented. When we do not consider transversal field mobility degradation, the values of the excess noise factors and correlation coefficients predicted in the long-channel SG MOSFETs by the Van der Ziel\textsuperscript{20} model are obtained. Moreover, the compact noise model reproduces the measured noise bias behavior for any gate length found for SG MOSFETs in literature, without the need of additional parameters. Therefore, the proposed noise model is very promising for being used in circuit simulators with DG MOSFETs.

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