Nonlinearity and scaling trends of quasiballistic graphene field-effect transistors targeting RF applications

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Abstract
Graphene field-effect transistors (GFETs) based on ballistic transport represent an emerging nanoelectronics device technology with promise to add a new dimension to electronics and replace conventional, silicon technology, especially for radiofrequency applications. The radiofrequency (GHz) static linearity and nonlinearity performance potential of GFETs is analyzed herein in the ballistic transport regime by exploring their static linearity mathematically in the quasiballistic transport regime along with their scaling potential at four different channel lengths. The proposed model explores linked mathematical expressions for the harmonic distortion, intermodulation distortion, and intercept points, which are depicted in graphical form. The second- and third-order harmonics and intermodulation distortions are analyzed with the help of a mathematical analysis of the drain current equation formulated using McKelvey’s flux theory. The presented expressions are validated based on the nonlinear output characteristic curves of the drain current versus the drain voltage for channel lengths of 140, 240, 300, and 1000 nm. The nonlinearity effect and its impact on the use of quasiballistic and ballistic GFETs for radiofrequency electronic applications is one of the important prospects and is tabulated in Table 1 for greater clarity using the particular models and their respective frequencies.

Keywords Graphene field-effect transistor · Quasiballistic transport · Nonlinearity · Harmonic and intermodulation distortion

1 Introduction
The modeling and simulation of GFETs plays an important role in the research and development of technology computer-aided design (TCAD) tools dedicated to two-dimensional (2D) electronic device design. Electronic device design tools are useful to simulate and develop advanced device structures and circuit applications. Conventional modeling of FETs is dedicated to silicon technology only. However, silicon technology-based electronics has reached its scaling limit. Further scaling of the channel length of Si-based metal–oxide–semiconductor field-effect transistors (MOSFETs) results in many short-channel effects and heat generation in integrated circuits (ICs). Thus, based on the continual advance of technology over time and the intense research interest in two-dimensional MOSFET technology recently, graphene has been identified as a novel electronic material. Graphene is an atomically thick 2D material with exceptional electronic properties, being the most promising 2D material for research into future electronic devices because of its very high electron mobility (2 × 10^5 cm² V⁻¹ s⁻¹) [1–3]. Although placing graphene on a silicon oxide substrate can degrade its most fascinating property (with an electron mobility of ~10⁴ cm² V⁻¹ s⁻¹), the scope of graphene for high-speed and RF electronics applications has also been reported [4–6]. Hexagonal boron nitride (hBN) is an appealing 2D dielectric substrate material that has been reported [7] to improve the electron mobility by three- to fourfold (to 4 × 10⁴ cm² V⁻¹ s⁻¹) in comparison with SiO₂. The hBN used as a 2D dielectric material also helps to improve RF performance metrics such as the high intrinsic transit frequency and unity-power-gain frequency of GFETs. Modeling and simulation of graphene FETs on an SiO₂/hBN substrate has been reported since the discovery of graphene. Modeling of graphene based FETs was reported for the first time using a graphene field-effect device in

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The current manuscript is organized into five sections. Following this introduction, the device structure and simulation setup, static nonlinearity model, results and discussion, and conclusion are presented. An appendix is also included to define the explicit drain current expression that is used to find the harmonic distortion (HD) and Intermodulation distortion (IMs). This introduction explains the various modeling approaches applied to GFETs and highlights the scope and research interest in finding a unique and novel way to study the nonlinearity of quasiballistic GFETs, which represents a further novelty of this work.

### 1.1 The device structure and simulation setup

TCAD simulations and the device structure are configured for ballistic and quasiballistic GFETs with four channel lengths (140, 240, 300, and 1000 nm) to describe the quasiballistic and ballistic transport of charge carriers in a graphene channel with lengths from 1000 nm, which is relatively quasiballistic in nature, to 140 nm, which is ballistic in nature. The entire simulation process is thus carried out in four independent steps for the device structures with channel lengths of 1000 nm, 300 nm 240 nm and 140 nm. Meanwhile, Fig. 1 illustrates the 2D view of the device and basic electrostatics characteristics of the device structure which remains the same, namely a four-terminal device with a brown 500-nm-thick silicon substrate on a blue/black 22-nm/10-nm back-gate contact, a 285-nm-thick (300 – 15 nm) back-gate oxide hBN, atomically thick red single-layer graphene, and 15-nm-thick top-gate oxide hBN, plus source, drain, and gate contacts with similar thickness to the back gate. The graphene is modified using the established properties of red polysilicon, such as the electron mobility $\mu_e$, hole mobility $\mu_h$, energy bandgap $E_G$ at room temperature, and work function.

### 1.2 The static nonlinearity model

The static nonlinearity of GFETs in the quasiballistic regime can be calculated by expanding the drain current using a Taylor series thus:

$$I_{DS} = x_1 V_{GS} + x_2 V_{GS}^2 + x_3 V_{GS}^3 + \ldots + x_n V_{GS}^n,$$  

(1)

where $x_1, x_2, \ldots, x_n$ are the coefficients of the Taylor series and $V_{GS}$ is the gate–source voltage. Meanwhile, the drain current [20] is formulated based on the MFT as

| Ref. | Operating frequency | IIP3 (dBm) | Conversion loss (dB) | L (μm) |
|------|---------------------|------------|----------------------|--------|
| [32] | 10 MHz              | 13.8       | ~30–40               | 2      |
| [33] | 30 GHz              | 12.8       | 19                   | 0.5    |
| [34] | NA                  | 4.9        | 20–22                | 1      |
| [35] | NA                  | 22         | ~15                  | 0.25   |
| [35] | NA                  | 27         | 10                   | 2      |
| [36] | 2 GHz               | 19         | 5                    | 0.75   |
| [37] | 4.3 GHz             | 30         | 10                   | 0.24   |
| [38] | 300 MHz             | 20         | 15                   | 0.5    |
| [39] | NA                  | 17         | 17                   | 2.4    |
| [24] | NA                  | 13.8       | 22                   | 0.44   |
| Proposed | 30 GHz           | 12.6       | 18.4                 | 0.14   |
| Proposed | 30 GHz            | 12.8       | 18.8                 | 0.30   |
where $W$ is the width of the GFET, $n(x)$ is the charge carrier concentration in the graphene channel as an ensemble of the field-dependent charge carrier density, $n_{E}$ and the field-independent charge carrier density, which is also called the residual carrier density. The total carrier concentration contributing to the calculation of the drain current can thus be represented mathematically as [22] as $n(x) = n_{EF} + n_{Res}$, where the field-dependent carrier density becomes more important at large values of $V_{CH}$ such that $qV_{CH} \gg K_{B}T$ and can be expressed mathematically as follows [21]:

$$n_{EF} = \frac{q \pi (K_{B}T)^{2}}{(h \nu f)^{2}} + \frac{(q)^{3}V_{CH}V_{CH}}{\pi (h \nu f)^{2}},$$  (3)

where the symbols have their usual meanings. The field-independent charge carrier density is also an ensemble of thermally excited ($n_{i}$) and electron–hole “puddles” adding a carrier density of [25].

$$n_{pu} = \frac{2}{\pi (h \nu f)^{2}} \left( \Delta^{2}/2 + (K_{B}T)^{2}/6 \right).$$  (4)

Besides, the 2D degenerate thermal velocity $V_{Th}$ of a graphene layer is found to be independent of $V_{GS}$ (the gate to source voltage), which can be expressed as [26] $V_{Th} = V_{Th} \sqrt{2/3}$ when using the value 2 for a 2D graphene sheet. The next element of the drain current is the back-scattering coefficient $r_{bs}$, which is another parameter that is independent of $V_{GS}$ directly and that can be solved mathematically as [22, 27].

Fig. 1 a A cross-sectional view of the proposed GFET device with its basic electrostatic parameters. b The equivalent capacitance of the GFET model. c A view of the proposed model of the GFET with its equivalent capacitance.
\[ r_{bs} = \frac{1.5V_{th}\mu_{eff}}{2K_BT_0} \left( 1 + \coth \left( \frac{Lg[1]}{2K_BT_0} \right) \right) + \lambda^{-1}, \]  
(5)

where \( \lambda \) is the mean free path in the graphene channel, which can be formulated as \( \lambda = \hbar \mu \left( \sqrt{\pi n} / 2 \right) \), where the symbols have their usual meanings. \( \mu_{eff} \) is the quasiballistic phenomenological effective mobility given by Mathiessen’s rule [28], which could be described as

\[ \frac{1}{\mu_{eff}} = \frac{1}{\mu_0} + \frac{1}{\mu_B}, \]  
(6)

where \( \mu_0 \) is \( 16 \epsilon/(h\nu)^2 \) \( q^3n_{imp} \) formulated in Ref. [29]. This part of the mobility is inversely proportional to the injected impurity density and plays an important scattering role. Now, the ballistic transport contribution to the mobility \( \mu_B \) is \( qL/mV_{th}2D \); i.e., the ballistic phenomenological mobility is inversely proportional to the thermal velocity. It is thus obvious that the total or effective mobility of two-dimensional graphene material depends on both the impurities as well as the thermal velocity but is independent of the gate to source voltage \( V_{GS} \). Another parameter in the drain current is \( I_1(\omega_F - \omega_{DS})/I_1(\omega_F) \), i.e., the Fermi–Dirac integral of order 1 defined by Blakemore, where \( \omega_F \) is equal to \( (E_F - E_C)/K_BT \) and \( \omega_{DS} \) is equal to \( (qV_{DS})/K_BT \) with \( E_F \) being equal to \( qV_C \). So, the first Fermi–Dirac integral \( \omega_F \) also depends on the gate to source voltage \( V_{GS} \). An additional and interesting fact is that \( V_{CH} \) is another parameter that depends on \( V_{GS} \) in the drain current formulation [12], which can be elaborated mathematically as a function of the parametric capacitances in the channel of a 2DGFET

\[ \frac{\partial I_{DS}}{\partial V_{GS}} = \frac{AC(2V_{gs,top} + C_{top} + 2V_{bs,back} + C_{box})P[R]}{C_{top} + C_{box} + 0.5C_q[(1 + P - Q)(R)]}, \]  
(9)

where \( A, B, \) and \( C \) are constants taking values of \( WV_{th}^3, q\pi(K_BT)^2/3(h\nu)^2 \), and \( q^3/\pi(h\nu)^3 \), respectively, where all the elementary constants have their usual meanings. Note in Eq. (7) that the carrier density \( n(x) \) parameter includes contributions that are both dependent on and independent of the electric field, but in the calculations below, only the \( n_{imp} \) part is used with the residual carrier density to minimize the leakage current. By simplifying and modifying the drain current in Eq. (8) as shown in the Appendix and applying Eq. (7), a direct relation between the gate and reference voltages can be obtained as

\[ \frac{\partial I_{DS}}{\partial V_{GS}} = \frac{1.5V_{th}\mu_{eff}}{2K_BT_0} \left( \frac{1 - \lambda}{\lambda} \right), \]  
(7)

\[ I_{DS} = \frac{A(B + CV_{ch}^2(x))\left( 1 - r_{bs} \right) \left( 1 - \lambda \right)}{E_{1}(\omega_F - \omega_{DS})/E_{1}(\omega_F)} \]  
(8)

where \( A, B, \) and \( C \) are constants taking values of \( WV_{th}^3, q\pi(K_BT)^2/3(h\nu)^2 \), and \( q^3/\pi(h\nu)^3 \), respectively, being functions of the drain voltage, effective mobility, mean free path, and channel length as described in detail in the Appendix.

The linearity and nonlinearity of a GFET in the ballistic transport regime, as stated above in Eq. (1), can be calculated by differentiating the drain current \( (I_{DS}) \) with respect to the gate to source voltage \( (V_{GS}) \). In this way, the coefficients of the Taylor series can be found to elaborate the nonlinearity of the GFET as reported in Ref. [23] \( (V_{gs} - V_{gs0} - V_{ds}) \). The unknown coefficients of the Taylor series can be found by simply differentiating the drain current equation for the bilayer GFET. The coefficients of the Taylor series \( x_1, x_2, x_3 \) can be further used to formulate the HD and intermodulation distortions as follows:

\[ \frac{\partial I_{DS}}{\partial V_{GS}} = \frac{AC(2V_{gs,top} + C_{top} + 2C_{box})P[R]}{C_{top} + C_{box} + 0.5C_q[(1 + P - Q)(R)]}, \]  
(9)

\[ x_2 = \frac{\partial^2 I_{DS}}{\partial V_{GS}^2} = \frac{-AC(2C_{top} + 2C_{box} - 2V_{gs,top} + C_{top} + 2V_{bs,back} + C_{box})P[R]}{C_{top} + C_{box} + 0.5C_q[(1 + P - Q)(R)]}, \]  
(10)
\[
\frac{\delta I_{DS}}{\delta V_{3GS}} = \frac{AC(2 \times 2C_{tox} \times C_{box})P(R)}{C_{tox} + C_{box} + 0.5C_q((1 + P - Q)(R))}.
\] (11)

These coefficients can be used in the further calculations to describe the nonlinearity as the GFET approaches the quasiballistic transport regime, which can be represented mathematically using the HD and first-, second-, and third-order intermodulation as described below. The first order already provides basic and useful information regarding the device model, while the harmonic distortion of second order is

\[
HD_2 = \frac{1}{2} \left| \frac{\delta I_{DS}}{\delta V_{GS}} \right| V_{GS} = \frac{(2C_{tox}^2 + 2C_{box}^2 - 2V_{gs top} \times C_{tox} \times 2V_{bs back} \times C_{box})V_m}{(2V_{gs top} \times C_{tox}^2 + 2V_{bs back} \times C_{box}^2 - 2V_{gs top} \times C_{tox} \times 2V_{bs back} \times C_{tox})},
\] (12)

and the harmonic distortion of third order is

\[
HD_3 = \frac{1}{4} \left| \frac{\delta I_{DS}}{\delta V_{GS}} \right|^2 V_{GS} = \frac{(C_{tox} \times C_{box})V_m^2}{(2V_{gs top} \times C_{tox}^2 + 2V_{bs back} \times C_{box}^2 - 2V_{gs top} \times C_{tox} \times 2V_{bs back} \times C_{tox})},
\] (13)

Similarly, the first-order intermodulation distortion can be found very easily, after which the second-order intermodulation distortion becomes

\[
IM_2 = \left| \frac{\delta I_{DS}}{\delta V_{GS}} \right| V_{GS} = \frac{(2C_{tox}^2 + 2C_{box}^2 - 2V_{gs top} \times C_{tox} \times 2V_{bs back} \times C_{box})V_m}{(2V_{gs top} \times C_{tox}^2 + 2V_{bs back} \times C_{box}^2 - 2V_{gs top} \times C_{tox} \times 2V_{bs back} \times C_{tox})},
\] (14)

and the third-order intermodulation distortion is

\[
IM_3 = \frac{3}{4} \left| \frac{\delta I_{DS}}{\delta V_{GS}} \right|^2 V_{GS} = \frac{3(C_{tox} \times C_{box})V_m^2}{4(2V_{gs top} \times C_{tox}^2 + 2V_{bs back} \times C_{box}^2 - 2V_{gs top} \times C_{tox} \times 2V_{bs back} \times C_{tox})}.
\] (15)

### 2 Results and discussion

The nonlinearity of a quasiballistic GFET can be studied statically as well as based on the observation of the potential performance at high frequencies. The high-frequency performance of a GFET is affected by the presence of parasitic capacitances at short channel lengths, which add a nonlinearity effect to its characteristics. The previous section explained the static nonlinearity using mathematical equations, whereas this section presents and validates the nonlinearity of the GFET in the quasiballistic regime in graphical and tabular form in a very novel way. The nonlinearity affects and influences the performance of the GFET through the addition of harmonic and intermodulation distortions, which result in a reduction in the gain of the GFET, a shift in the direct-current (DC) offsets, and

![Fig. 2](image)

*Fig. 2* The nonlinear behavior of a bilayer graphene field-effect transistors (GFET), showing the static nonlinearity on a column, stacked, and area chart for the proposed model as well as CMOS field-effect devices.
cross-modulation of the amplitude modulation (AM)/phase modulation (PM), etc.

Table 1 presents and compares the RF nonlinearity characteristics of GFETs with different channel lengths and frequencies, as well as the third-order input intercept point (IIP3) and total gain compression, for previously available and the proposed model. The bar graph in Fig. 2 shows the static and RF nonlinearity in a column, stacked, and area chart for the proposed model as well as CMOS field-effect devices. The analytical outcomes of the proposed GFET model and the simulated results are compared with those of CMOS technology based on the HD and second- and third-order intermodulation distortions. The HD and IMs validate the nonlinear character of the proposed ballistic GFET model.

This section validates and compares the dynamic nonlinear characteristic curves of the ballistic GFET with various channel lengths (\(L = 140, 240, 300,\) and 1000 nm) as shown in Fig. 3. The nonlinearity effects are clearly seen in Fig. 3a–d in the \(I-V\) characteristic curve at very low gate voltages, revealing a gradual nonlinear increment and turn with kinks at higher values of the drain voltage. The characteristic curve crosses over for different gate voltages, as clearly seen in Fig. 3, because of the kinks in the \(I-V\) characteristic curves. These crossovers of the characteristic curves with increasing gate voltage prove and justify the linearity, scaling, and nonlinearity of the characteristic behavior of the drain current as the GFET approaches the ballistic regime. The scaling of the channel length and the nonlinearity of the drain current are closely related, as seen in all four plots in Fig. 3 for various channel lengths (\(L = 140, 240, 300,\) and 1000 nm) with increasing drain to source voltage (from 0 to 2 V). The output characteristic curve of the drain current versus the drain voltage in Fig. 3a confirms the highly nonlinear behavior for the channel length of 1000 nm. The results shown in Fig. 3b for a channel length of 300 nm also reveal nonlinearity at low drain voltage values of 0–1 V, but a linear profile thereafter. Figure 3c shows an almost linear curve of current versus voltage, while a kink is present at 1.25 V at a low gate voltage value, being absent at higher values of the gate voltage. Figure 3d shows a
linear curve of the output drain current versus the voltage at all drain voltages in the range from 0 to 2 V. This work, inspired by the static nonlinearity in the nonballistic transport regime [23], thus presents the static and dynamic nonlinearity of the quasiballistic GFET model, which is validated and compared in the bar graphs in Fig. 2, the RF frequency nonlinearity characteristics presents in Table 1, and the dynamic nonlinearity drain current characteristics in Fig. 3. The dynamic nonlinear characteristic behavior matches very well when compared with reported works [16, 21]. All the simulation results are shown in Fig. 3 as lines, while all the analytical outcome values are indicated by square symbols.

3 Conclusions

Since the nonlinearity of each device is a strong source of noise in nanoelectronics device applications based on GFETs, the linearity and nonlinearity of GFETs operating in the ballistic transport regime are of critical importance. The nonlinearity characteristic behavior of such devices is thus presented herein, being of significant importance in the current era of advanced nanotechnology. The results of the proposed model and simulations for GFETs with various, scaled channel lengths are presented as line graphs.

Appendix

The drain current for a four-terminal GFET device can be modified to simplify the calculations and enable effective modeling by electronic design automation (EDA) tool developers. From Eqs. (3) and (4), the total charge density can be obtained as

\[ n(x) = \frac{(q)^3V_{CH}|V_{CH}|}{\pi (hv_{F})^2} + C \ldots, \]

where \( C \) is a constant, the symbols used in Eqs. (3) and (4) have their usual meanings, and \( v_{F} \) is the Fermi velocity \( (10^6 \text{ m/s}) \) at room temperature [1]. For a particular value of the extrinsic electric field \( (E) \), fixed channel length \( (140 \text{ nm or 300 nm}) \), and mean free path length, mentioned in detail in the “Introduction” section, the back-scattering coefficient \( (r_{bs}) \) can also be simplified explicitly from Eq. (5) as a function of \( V_{ds} \) and \( \mu_{eff} \), where \( C \) is again a constant with other symbols having their usual meanings.

\[ r_{bs} = \frac{1.5V_{ds}\mu_{eff}}{2LK_{B}T_{c}}, \]

Meanwhile, the first-order Fermi–Dirac integral solved by Blackmore [30, 31] can also be simplified more explicitly as a function of \( V_{ds} \) to become

\[ \mathcal{E}_1(\omega_{s}) = \int_{0}^{\infty} \frac{1}{1+\exp\left(\frac{E-E_{F}}{k_{B}T}\right)} \, dE, \]  

which can be approximated as \( \exp - \frac{(E-E_{F})}{k_{B}T} \), thus \( \mathcal{E}_1(\omega_{s} - \omega_{DS}) / \mathcal{E}_1(\omega_{s}) \) can be expressed as a function of the drain voltage only thus \( \exp - \frac{V_{ds}}{k_{B}T} \) after simplification of the exponential. A proper drain current and gate voltage relation can thus be expressed as

\[ I_{DS} = \frac{A(B + CV_{CH}^2(x)) \left( 1 - \frac{1.5V_{ds}\mu_{eff}}{2LK_{B}T_{c}} \left[ 1 - e^{-V_{ds}/k_{B}T} \right] \right)}{\left[ 1 + \frac{1.5V_{ds}\mu_{eff}}{2LK_{B}T_{c}} \left( \frac{1.5V_{ds}\mu_{eff}}{2LK_{B}T_{c}} \right)^2 e^{-V_{ds}/k_{B}T} \right]} \].

Declarations

Conflict of interest The authors confirm that there are no conflicts of interest.

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