**2D Materials-Based Radio-Frequency Wireless Communication and Sensing
Systems for
Internet-of-Things Applications**

Liang Zhu¹, Mohamed Farhat², Khaled Salama², and Pai-Yen Chen¹

*¹Department of Electrical and Computer Engineering, University of Illinois at
Chicago, IL, 60607, USA*

*²Computer, Electrical, and Mathematical Science and Engineering (CEMSE)
Division, King Abdullah University of Science and Technology (KAUST), Thuwal 23955-
6900, Saudi Arabia*

Abstract: Radio-frequency (RF) electronics involving 2D material-based devices and integrated circuits have been widely advocated as a promising platform for future generations of wireless communication and sensing systems. Monolayer and few-layer 2D materials display exceptional electrical, mechanical, and thermal properties, which can be leveraged for building superior key elements in transceiver/receiver RF front ends, including but not limited to amplifiers, mixers, switches, oscillators and modulators that enable various forms of signal modulations. In particular, some 2D materials with high intrinsic mobilities can be used as channels of RF transistors that possess high gain, high cutoff frequency (up to hundreds of gigahertz), and excellent abilities in improving the performance of complex analog and RF circuits. Moreover, the (bio-)chemical sensing capabilities of 2D material-based devices may be integrated into RF circuits and modular building blocks for making monolithically-integrated wireless nanosensors. The robust, flexible, scalable and small-footprint features of these 2D devices further allows for development of flexible electronics (e.g., wearable devices, smart skins and contact lens sensors) that facilitate the practice of ubiquitous sensors and internet-of-things (IoT). In this chapter, we will review recent advances in RF and microwave transistors based on 2D materials such as graphene and beyond, as well as their potential applications as constituent parts of the IoT-hardware.

Keywords: 2D materials, RF and microwave transistors, RF building blocks, wireless sensors, internet-of things

2.1 Introduction

Nowadays, the internet of things (IoT) has had an increasing impact on healthcare, automobiles, agriculture, retail, security, specific environmental controls and so forth [1], [2]. Thanks to rapid advances in nanotechnology and microsystem engineering, as well as ever-expanding wireless technology, nanomaterial-based sensors are envisioned to drive the market of wireless IoT sensors that could be compact, low-cost, wearable, textile-based, and light-weight. When compared to bulky semiconductors, low-dimensional nanomaterials may be significantly more sensitive to the exposed environments and chemical substances. Among various types of nanomaterials, 2D materials including graphene and beyond were the most studied ones in the past decade [3]-[5]. In general, they exist in a bulk form as stacks of strongly bonded monolayers with weak interlayer attraction, and can be successfully isolated as single atomic layers by means of exfoliation or chemical vapor deposition (CVD) methods. Graphene formed by a single layer of graphite was the first two-dimensional (2D) material ever discovered. In 2004, Geim's group, which has been widely credited for the explosive growth of interest in 2D materials, reported the first graphene field-effect transistor (GFET), opening up a new field of 2D electronics [6]. GFETs with ultrahigh carrier mobilities and cut-off frequency (up to 840 GHz [7]) are particularly attractive for RF applications, such as high-frequency power amplifiers and oscillators in modern communication systems. It was even more exciting that graphene transistors or electrodes can be flexible, highly-durable, and optically transparent. Such properties make graphene become the preferred candidate material for wearable and textile electronics. Moreover, it has been demonstrated that the electrical properties of GFETs can be sensitively tuned by chemical gating effects, which enables selective detection of gas, chemical, or biomolecular agents, with sensitivity at the single-molecule level [8], [9].

Noticeably, pristine graphene does not have a bandgap - a property that is essential to maintain a good on/off current ratio for logic circuits. Due to the ambipolar transport property of the gapless graphene, GFETs typically exhibit a "V- shape" drain current-gate voltage characteristic that enables effective conversion of an applied RF signal into its second harmonic [10], [11]. This exotic characteristic was not found in traditional silicon or III-V devices, and can be engineered to make the simplest possible frequency modulator. At terahertz (THz) and infrared frequencies, low-loss Dirac plasmons excited in graphene,

together with tunability of Fermi energy through gating effects, make graphene plasmonics an excellent platform for building reconfigurable infrared optoelectronic components. In this regard, several new applications have been devised, such as reconfigurable optoelectronic devices [12], metasurfaces and metamaterials [13], [14], phase shifters [15] and oscillators (with proper optical pumping) [16].

A patterned graphene, e.g., graphene nanoribbon (GNR), can have a bandgap that depends on the width of the GNR (w) as: $E_g [\text{eV}] = 0.8/w [\text{nm}]$ [17]. Although a GNR can be used to improve on/off ratios of GFETs, it also causes drops in the intrinsic and extrinsic mobilities of graphene. In this context, many new types of 2D materials with sizable intrinsic bandgap (at the level of 1-2 eV) have been investigated [18], [19], aiming to fabricate high-performance switching devices that meet requirements of chip-scale logic circuits. Transition metal dichalcogenide (TMDs) are a class of 2D materials with the chemical formula MX_2 , where M is a transition metal element from group IV (e.g., Ti, Zr, or Hf), group V (e.g., Nb or Ta), or group VI (e.g., Mo, W and so on), and X is a chalcogen (e.g., S, Se or Te). These materials are made-up of layered composite of the form X-M-X , with the chalcogen atoms in two hexagonal planes separated by a plane of metal atoms. To date, 2D transistors with MoS_2 and WeS_2 channels have been demonstrated to have a greater than 10^8 on/off ratio [5]; Such a value makes them suitable for the development of low-power digital circuits for IoT applications. So far, many groups have successfully fabricated wafer-scale TMD-based transistors for analog and digital integrated circuits. [20], [21].

Wireless communication and sensing systems present a continuing evolution in terms of the number of users and of the volume of exchange information, as well as the basis for advancing the IoT systems. However, it is always met with price constraints, an incessant miniaturization, and new specificities that make the design conditions increasingly difficult. For instance, textile and flexible technologies are not possible with traditional solid-state devices that are bulky and rigid, but may be realized with 2D electronics. This chapter will briefly review RF and microwave transistors based on different 2D materials, and their practical uses in RF circuit building blocks illustrated in Fig. 2.1. Fig. 2.1 shows, as an example, the RF architectures of wireless IoT sensors and integrated system, of which 2D electronics can be applied to several highlighted key building blocks. These basic blocks

include passive RF components (e.g., filter and miniature antenna) and active RF components (e.g., RF mixer with a local oscillator, signal modulator/demodulator, variable-gain amplifier, low-noise amplifier (LNA)), memories, digital chip, and analog-to-digital converter (ADC). Replacing current RF front-end components with 2D-based devices and circuits offers several advantages in terms of cost, performance, power consumption, device size, and flexibility. Researchers have envisioned that 2D electronics will create a new paradigm in the design of next-generation wearable electronics, bioimplants, soft robots, and ubiquitous sensors for IoT applications. In the following, the state-of-the-art GFETs and different 2D transistors working in the RF and microwave frequency ranges will be reviewed, followed by introduction of modeling and parameter extraction methods for these 2D transistors. Afterward, an overview of active RF components based on 2D transistors will be presented, and their benefits and tradeoffs will be discussed. Finally, potential applications of 2D electronics in ubiquitous wireless sensors and IoT systems will be mentioned.

2.2 RF performance of 2D transistors

Low-dimensional nanostructures and nanomaterials, such as semiconductor nanowires and carbon nanotubes (CNT) have been extensively studied for developing field-effect transistors with a ballistic channel, which potentially pushes the cutoff frequency to the terahertz region [22], [23]. In practice, however, their operation frequencies were limited to tens of GHz due to the high contact resistance of nanotubes/nanowires with an infinitesimal diameter [24], [25]. To address the fabrication challenge and drastically reduce the contact resistance, 2D materials have been exploited to scale up the drain current of nanotransistors by significantly increasing the device's channel width. In 2004, the first 2D material, namely graphene, was isolated from graphite and transferred into FETs [6]. Since graphene has a linear energy-momentum ($E-k$) dispersion in the low-energy region, electrons and holes in graphene are considered as massless relativistic particles with an extremely high mobility (Dirac fermions) and an energy-independent velocity that arise in ballistic transports (if scattering with phonons or impurities is excluded). Further research has demonstrated that GFETs can possess an ultrahigh electronic mobility that exceeds $200,000 \text{ cm}^2/\text{Vs}$ [26], a high saturation velocity ($5.5 \times 10^7 \text{ cm/s}$ [27]), and a great carrier

density modulation by electric field ($\sim 2 \times 10^8$ A/cm [28]), which make graphene appealing for high-speed electronic circuits on flexible substrates.

In this section, we will review essential features of high-frequency transistors based on graphene and types of 2D materials, as well as figures of merits (FOMs) used for accessing their performance. Figs. 2.2(a) and (b) respectively show the schematics and the small-signal equivalent circuit model for a 2D material-based FET, of which important elements include the gate-source capacitance C_{gs} , the gate-drain capacitance C_{gd} , the transconductance g_m , the differential drain resistance r_{ds} (the inverse of the drain conductance g_{ds}), and the contact resistances and parasitics due to packages. The transport and small-signal behavior of a transistor are generally described by the drift-diffusion model. As a representative example, we first discuss the modeling of transport properties of GFETs in Fig. 2.2(a). Considering the symmetric low-energy electronic structure in graphene, electron and hole sheet densities (n and p) are derived from the integration of density of states $D(E)$ weighted by the Fermi-Dirac distribution $f(E)$, and the total mobile charge density Q_s in graphene is obtained as [29]:

$$\begin{aligned} Q_s &= q \times (p - n) = q \times \left(\int_{-\infty}^0 D(E) [1 - f(E)] dE - \int_0^{\infty} D(E) f(E) dE \right) \\ &= q \frac{2}{\pi (\hbar v_F)^2} \int_0^{\infty} E \left[\frac{1}{e^{(E+E_F)/K_B T} + 1} - \frac{1}{e^{(E-E_F)/K_B T} + 1} \right] dE, \end{aligned} \quad (1)$$

where $\hbar$ is the reduced Planck constant, E_F is the Fermi level, v_F is the Fermi velocity ($v_F = 10^8$ cm/s), and K_B is the Boltzmann constant. Based on the quantum capacitance definition: $C_q = -\partial(Q_s)/\partial V_{ch}$, we may obtain an explicit expression for quantum capacitance as [30]:

$$C_q = \frac{2q^2 K_B T}{\pi (\hbar v_F)^2} \ln \left[2 \left(1 + \cosh \frac{q V_{ch}}{K_B T} \right) \right], \quad (2)$$

where $V_{ch} = E_F / q$ is the surface potential. Under the condition $q V_{ch} \gg K_B T$, which is generally valid, the expression for quantum capacitance can be simplified to $C_q = K |V_{ch}|$, where $K = 2q^3 / [\pi (\hbar v_F)^2]$. As a result, the net mobile charge density can be derived as

$Q_s = \int C_q dV_{ch} = -\frac{1}{2} C_q V_{ch}$. In a backgate-GFET, since any variation in the Fermi level by an applied gate voltage is equivalent to the voltage drop across the quantum capacitance C_q , by applying Kirchhoff's law, the surface potential is obtained as [31]-[33]:

$$V_{ch}(x) = E_F(x)/q = \frac{C_{ox}}{C_{ox} + C_q/2} (V_{GS} - \Delta V_{cnp} - V(x)), \quad (3)$$

where $C_{ox} = \epsilon_{ox}/t_{ox}$ is the oxide capacitance, and ΔV_{cnp} is the backgate voltage at the Dirac point where the carrier sheet density becomes minimal for zero drain and source voltages. ΔV_{cnp} comprises work-function differences between the gates of the graphene channel, interfacial charges at the graphene-oxide interface, and the charge neutrality point shift ΔV_{cnp} due to intentional or unintentional doping of graphene (e.g. chemical doping). The channel voltage $V(x)$, as a function of position in the channel, can be modeled using the gradual channel approximation: $V(x) = (x/L)V_{DS}$, which is zero at the source end ($x = 0$) and equal to the drain-to-source voltage V_{DS} at the drain end ($x = L$). Since C_q is also a function of V_{ch} , equations (2) and (3) must be solved self-consistently to evaluate the actual V_{ch} , which may have an explicit expression as:

$$V_{ch} = \frac{-C_{ox} + \sqrt{C_{ox}^2 \pm 2K[V_{GS} - \Delta V_{cnp} - V]C_{ox}}}{\pm K}. \quad (4)$$

For a relatively thick gate oxide, i.e. $C_{ox} \ll C_q$, typically valid for backgate transistors,

$V_{ch}(x)$ can be simply expressed as $V_{ch}(x) = E_F(x)/q = \hbar v_F \sqrt{\pi C_{ox} [V_{GS} - (x/L)V_{DS}]/q^3}$.

On the other hand, for an ultrathin high-K oxide, $V_{ch}(x)$ reduces to a simple form

$V_{ch}(x) = E_F(x)/q \approx V_{GS} - (x/L)V_{DS}$. The drift-diffusion transport model suggests the

drain-to-source current $I_{DS} = qW|Q_s(x)|v_{drift}(x)$, where W is the gate width, v_{drift} is the

drift velocity. Using a soft saturation model, consistent with Monte Carlo simulations,

$v_{drift}(x)$ that depends on the channel electric field F can be expressed as [31]-[33]:

$$v_{\text{drift}} = \frac{\mu F}{\left[1 + (\mu F / v_{\text{sat}})^\gamma\right]^{1/\gamma}}, \quad (5)$$

where μ is the carrier low-field mobility and v_{sat} is the saturation velocity. Here, we assume that v_{sat} is a material constant that is equal to $v_F / 2$ (which is close to GFET's maximum predicted values), and $\gamma = 1$ for GFETs. Applying $F = -dV / dx$ and integrating the resulting equation over the device length (L), the drain current becomes [31]-[33]:

$$I_{\text{DS}} = \mu \frac{W}{L_{\text{eff}}} \int_0^{V_{\text{DS}}} |Q_s| dV, \quad (6)$$

where the effective channel length accounting for the effect of saturation velocity. In order to have an explicit expression for the drain current, it is more convenient to solve the integral in (6) using V_{ch} as the integration variable and consistently express Q_s as a function of V_{ch}

$$I_{\text{DS}} = \mu \frac{W}{L_{\text{eff}}} \int_{V_{\text{ch,D}}}^{V_{\text{ch,S}}} |Q_s(V_{\text{ch}})| \frac{dV}{dV_{\text{ch}}} dV_{\text{ch}}, \quad (7)$$

where the channel potential at the drain and source $V_{\text{ch,D}}$ and $V_{\text{ch,S}}$ are determined as $V_{\text{ch}}(V=V_{\text{DS}})$ and $V_{\text{ch}}(V=0)$, respectively. Finally, the explicit expression for drain current $I_{\text{DS}}(V_{\text{DS}}, V_{\text{GS}})$, as a function of bias voltages V_{DS} and V_{GS} , can be obtained as

$$\begin{aligned} I_{\text{DS}} &= \frac{\mu K}{2} \frac{W}{L_{\text{eff}}} \left[-\frac{V_{\text{ch}}^3}{3} - \text{sgn}(V_{\text{ch}}) \frac{KV_{\text{ch}}^4}{4C_{\text{ox}}} \right] \bigg|_{V_{\text{ch}}=V_{\text{ch,S}}}^{V_{\text{ch}}=V_{\text{ch,D}}} \\ &= \frac{\mu K}{2} \frac{W}{L_{\text{eff}}} \left\{ \left[-\frac{V_{\text{ch,D}}^3}{3} - \text{sgn}(V_{\text{ch,D}}) \frac{KV_{\text{ch,D}}^4}{4C_{\text{ox}}} \right] - \left[-\frac{V_{\text{ch,S}}^3}{3} - \text{sgn}(V_{\text{ch,S}}) \frac{KV_{\text{ch,S}}^4}{4C_{\text{ox}}} \right] \right\}. \end{aligned} \quad (8)$$

For moderate input RF voltage, the high-frequency behavior of GFETs can be modeled by a small-signal equivalent circuit shown in Fig. 2.2(b). The transconductance g_m and the drain conductance g_{ds} of GFET can be described as:

$$g_m = \frac{dI_{\text{DS}}}{dV_{\text{GS}}} \bigg|_{V_{\text{DS}}=\text{constant}}. \quad (9a)$$

$$g_{ds} = \frac{1}{r_{ds}} = \left. \frac{dI_{DS}}{dV_{DS}} \right|_{V_{GS}=\text{const}}. \quad (9b)$$

The mobile channel charge, depending on the bias voltages V_{GS} and V_{DS} , can be modeled by the gate-to-source capacitance C_{gs} and the gate-to-drain capacitance C_{gd} as

$$C_{gs} = - \left. \frac{dQ_{ch}}{dV_{GS}} \right|_{V_{DS}=\text{const}}, \quad (10a)$$

and

$$C_{ds} = - \left. \frac{dQ_{ch}}{dV_{DS}} \right|_{V_{GS}=\text{const}}, \quad (10b)$$

where the total net charge in the channel Q_{ch} is given by

$$Q_{ch} = W \int_0^L Q_s dx = qW \int_0^L [p(x) - n(x)] dx. \quad (11)$$

After some mathematical manipulation, (11) can be rewritten as

$$Q_{ch} = W \int_0^{V_{DS}} Q_s(V_{ch}) \left(\frac{Q_s(V_{ch}) \mu W}{I_{DS}} + \frac{\mu}{v_{sat}} \right) \frac{dV}{dV_{ch}} dV_{ch}. \quad (12)$$

There are also other parasitic components, which have minor effects provided the operating frequency is sufficiently low.

The performance of a transistor, when used for amplification of an AC signal, is determined by several factors, which include intrinsic gain G_{int} , the current gain h_{21} (see the 2-ports network model in Fig. 2.2(c)) and the unilateral power gain U . The cutoff frequency f_T and the maximum frequency of oscillation f_{max} of a RF transistor is closely related to the current and power gains. The cutoff frequency is the frequency at which the magnitude of h_{21} drops to unity (0 dB) and the maximum frequency of oscillation f_{max} signifies that the unilateral power gain U equals unity (0 dB). For example, Fig. 2.2(d) shows h_{21} and U of a GFET, which are frequency-dependent and roll off with increasing frequency at a slope of -20 dB/decade. f_T and f_{max} mark the upper frequency limits of the RF transistors because beyond these frequencies the transistors fail to amplify the input AC signal. From the equivalent circuit model in Fig. 2.2(b), the intrinsic gain G_{int} , intrinsic cutoff frequency f_{T-int} and intrinsic maximum frequency $f_{max-int}$ of the transistor can be deduced, respectively [3].

$$G_{\text{int}} = \frac{g_m}{g_{ds}} , \quad (13a)$$

$$f_{T-\text{int}} = \frac{g_m}{2\pi(C_{gs} + C_{gd})} , \quad (13b)$$

$$f_{\text{max-int}} = \frac{g_m}{4\pi C_{gs}} \times \frac{1}{\sqrt{g_{ds} R_i}} . \quad (13c)$$

An useful approximation for the extrinsic frequencies are [34]-[36]

$$f_T = \frac{g_m}{2\pi(C_{gs} + C_{gd})} \frac{1}{1 + g_{ds}(R_S + R_D) + \frac{C_{gd}g_m(R_S + R_D)}{C_{gs} + C_{gd}}} , \quad (14a)$$

$$f_{\text{max}} = \frac{g_m}{4\pi C_{gs}} \frac{1}{\sqrt{g_{ds}(R_i + R_S + R_G) + g_m R_G \frac{C_{gd}}{C_{gs}}}} . \quad (14b)$$

To extract the intrinsic transistor parameters and further de-embed the parasitic elements, it is necessary to measure the S-parameters of a RF transistor shown in Fig. 2.2(e). In this experimental setup, the ground-signal-ground (GSG) pads are placed at the gate and drain terminals with a source electrode serving as the common ground [37]. The extrinsic current gain h_{21} can be derived from the measured reflection coefficients (S_{11} and S_{22}) and transmission coefficients (S_{12} and S_{21}):

$$h_{21} = \frac{-2S_{21}}{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}} . \quad (15)$$

The intrinsic device data is obtained by employing a two-step de-embedding process with the device under test (DUT) model shown in Fig. 2.2(c). The admittance parameters (Y parameters) of the DUT are obtained from extrinsic measurement data (transition from S parameters to Y parameters), and afterwards the parasitics (obtained from open and short testing structures) are de-embedded to determine the intrinsic device frequency responses. The open transistor structure with no gate channel involved determines parallel parasitics

(Y_{p1} , Y_{p2} , Y_{p3}) and the short structure with gate, drain and source shorted decides the series parasitics (Z_{L1} , Z_{L2} and Z_{L3}). These series parasitics are very critical for modern small channel FETs

$$Y_{\text{transistor}} = ((Y_{\text{dut}} - Y_{\text{open}})^{-1} - (Y_{\text{short}} - Y_{\text{open}})^{-1})^{-1}, \quad (16)$$

where $Y_{\text{transistor}}$ is the intrinsic Y-parameters of the intrinsic RF transistor, Y_{dut} is the Y-parameter of DUT (Extrinsic), Y_{short} is the Y-parameter of the DUT with gate, drain and source shorted, Y_{open} is the DUT with no channel material (graphene, TMD, etc.) [38]. The comparison of the current gain h_{21} before and after de-embedding of graphene transistor is shown in Fig. 2.2(f), in which the de-embedding h_{21} maintains rolling off with increasing frequency at a slope of -20 dB/dec.

In 2009, the IBM group was the first to comprehensively study the high-frequency performance of graphene RF transistors with variable gate lengths and a predicted cut-off frequency of 26 GHz [37]. In that work, the small-signal current gain h_{21} of the graphene transistors extracted from S -parameter measurement was found to follow the $1/f$ law, which is analogous to conventional FETs. Besides, the cut-off frequency is a strong function of the drain and gate voltages, and is proportional to the dc transconductance g_m . In particular, the cutoff frequency was found to be inversely proportional to the square of the gate length ($\sim 1/L^2$). A peak f_T as high as 26 GHz was predicted for a gate length of 150 nm, as shown in Figs. 2.3(a)-(b). Hereafter, the high frequency performance of GFETs began to draw attention and a growing number of studies have been reported. In early 2010, a graphene MOSFET with a 240 nm gate showing a f_T of 100 GHz has been reported [20], soon followed by a 144 nm gate graphene transistor with f_T approaching 300 GHz later that year [39]. Finally, in 2012, a GFET with a 40-nm-long gate showing 350 GHz f_T [10] and a 67 nm gate GFET with a cutoff frequency f_T 427 GHz have been prepared [21]. The frequency response of the initial devices with only limited cutoff frequency, normally several gigahertz [40], [41], is primarily restricted by two factors. First, the dielectric integration process uses an oxidative functionalization layer, resulting in a serious degradation of the electronic properties of graphene with carrier mobility as low as 400 cm^2/Vs . Second, in these initial devices, a large gap exists between source/gate and/or gate/drain electrodes, in which the graphene is not effectively modulated by the gate but functions as a series access resistance. To overcome the limitation of graphene transistors

and improve their RF performance, various gating configurations and different dielectric/substrate materials have been studied. For the gating configurations, dual gate and self-alignment gate graphene transistor are two common alternatives. Different from the first top gate GFET, dual gate structure consists of both local top-gate and global back-gate, of which the access resistance of the graphene transistor is modulated by the back-gate through electrostatic doping [Figs. 2.3(c)-(d)]. By varying the back-gate voltage, the access resistance can be significantly reduced by more than a half, contributing to a four-fold increase of transconductance g_m . As a result, an intrinsic cutoff frequency of 50 GHz is achieved in this 350 nm gate graphene transistor [42] [Figs. 2.3(e)-(f)]. Another viable gating scenario to mitigate the access resistance is through optimum alignment of the source/drain and the gate electrodes [43], [44]. In this manner, Liao's group developed a self-aligned approach, which utilized a metal/oxide $\text{Co}_2\text{Si}/\text{Al}_2\text{O}_3$ core/shell nanowire as the top gate and integrated it on graphene through a physical assembly process at room temperature, successfully demonstrating a graphene transistor with cutoff frequency f_T of up to 300 GHz, as shown in Figs. 2.4(a)-(c) [39]. Take one step further, in order to eliminate the imperfect contact interface with graphene since the diamond-like cross section of this $\text{Co}_2\text{Si}/\text{Al}_2\text{O}_3$ core/shell nanowire, they proposed 100 nm channel long GaN nanowire [Figs. 2.4(d)-(e)] with a triangular cross-section and a flat side surface as the self-aligned gate, which exhibits a scaled transconductance exceeding 2 mS/ μm and a projected intrinsic f_T at $V_{\text{ds}} = 1$ V is 840 GHz [7], [45]. On the other hand, as for the degradation of graphene mobility, the interfaces existing in graphene/substrate and graphene/top-gate dielectric insulator, are two dominant sources of scattering in graphene devices, resulting in this significant degradation of electrical transport due to the two-dimensional nature of graphene. In 2011, Balandin's group mitigated the scattering effect associated with trapped charges through the usage of DLC film as the GFET substrate, which is advantageous because of the high phonon energy in sp^3 - hybridized carbon (165 meV) and its nonpolar, as well as the non-hydrophilic nature [46], [47]. Another desirable substrate is silicon carbide, which likewise possesses a high intrinsic phonon energy of 116 meV, and additionally, high quality graphene sheets can be directly grown on its surface by epitaxy. In these suitable substrate alternatives, the intrinsic cutoff frequency of graphene transistors may reach 300 GHz, based on both wafer-scale CVD grown graphene and epitaxial

graphene on SiC [48], [49]. In brief, to further alleviate the scattering effect between the 2D graphene and the dielectric surface during the integration process, miscellaneous types of oxidative functionalization layer, such as Al_2O_3 [21], HFO [20], [50], Y_2O_3 [51], and Boron-Nitride [19], have also been studied extensively to receive the cutoff frequency as high as 400 GHz.

Even though graphene transistor satisfies the demand of RF and microwave characteristic in some applications, the lack of high on/off current ratio and also the relative difficulty to achieve saturation due to the absence of a band-gap lead to inapplicability for logic/digital circuit applications and limited power gain for RF/analog applications, respectively. Compared to graphene, which has no bandgap, TMDs such as MoS_2 , MoSe_2 , WS_2 and WSe_2 have sizable bandgaps changing from indirect to direct in single layers, allowing applications such as transistors, photodetectors and electroluminescent devices. Some of the excellent electronic properties of monolayer TMD include: (1) high electrical conductivity with a bandgap around 1–2 eV, capable of providing sufficient on/off ratio for digital/logic electronics applications, while exhibiting decently high cutoff frequency and power gain for analog/RF electronics; (2) atomic-scale thickness with no dangling bonds at the interface allows excellent control over the channel with low parasitic capacitances; (3) relatively high effective mass of carriers contributes to the reduction of source to drain tunneling current. In the initial stage, the majority of works on TMD-based FETs have focused on the improvement of their digital performance, i.e., the switching capability and the on/off current ratio. In 2011, the first TMD-transistor was reported with the single layer MoS_2 as channel material to fit the logic circuit application with the on/off current ratio up to 10^8 [5]. Since then, over the past few years, many experimental efforts have demonstrated the excellent on/off ratio in between 10^4 and 10^8 [18] with a subthreshold swing approaching 60 mV/dec [52] for MoS_2 field effect transistor (FET), and more recently for WS_2 [53] and WSe_2 transistors [19]. On the other hand, the saturation phenomenon of TMD-transistor under small bias provides potential applications for high power gain electronic devices, and recently, some of the experimental and simulation works have explored on analog/RF performance analysis of MoS_2 -FETs [54], [55]. The systematic investigation of top-gated single-layer, double-layer, and even tri-layer MoS_2 RF transistors with gate lengths scaled down to 200 nm have been reported with the cutoff

frequency of 2 GHz before the de-embedding procedure and as high as 6 GHz after the de-embedding procedure [55], [56] [Figs. 2.5(a)-(b)]. Additionally, the recent numerical works for MoS₂-FETs have been limited to intrinsic analog/RF performance analysis, and show an intrinsic gain higher than 20 dB [57] and the intrinsic cutoff frequency approaching terahertz for sub-5-nm channel of MoS₂ [58].

The high performance TMD-transistor faces the same challenges as graphene transistors, namely: 1) the degradation of TMD material mobility due to the scattering effect; 2) the contact resistance at the metal-MoS₂ interfaces. It should be emphasized that, given the fact carrier mobility of semiconductors tends to reduce with the increasing gap energy, TMD material with 1-2 eV bandgap will inevitably have smaller mobility compared to zero bandgap graphene. For back-gate MoS₂ transistors, the upper mobility limit of defect-free undoped material have been calculated and it ranges from 130 cm²/Vs [59] to 410 cm²/Vs [60], and MoS₂ device mobilities between 1 and 300 cm²/Vs have been extracted from experimental results [61], [62]. In particular, the mobility reported for back-gated MoS₂-FETs (a few to a few tens of cm²/Vs) is commonly lower than the observed mobility in top-gated ones (several tens to hundreds of cm²/Vs), of which the electron mobilities of up to 1000 cm²/Vs [63] have been discovered. As for the other TMD materials, WSe₂ with a hole mobility of 250 cm²/Vs has been reported and in few-layer phosphorene, a remarkable hole mobility exceeding 1000 cm²/Vs has been measured [19]. In terms of the aspect of contact resistance, the first contact test experiments for 2D materials have focused on graphene, and up to now metal-graphene contact resistances as low as 0.01–0.2 Ω mm have been achieved [64], [65]. This already comes close to the resistance of state-of-the-art metal contacts on Si and III–V semiconductors. Synchronously, contact resistance data for metal–TMD material have been reported as well. In 2004, a new strategy based on phase engineering to design low-resistance contacts have been developed, in which it is proposed a methodology to locally pattern the metallic-1T phase on ultrathin semiconducting MoS₂ nanosheets and use them as electrodes. As a result, it was successfully demonstrated low contact resistances of 200–300 Ω μ m at zero gate bias with the properties in air as mobility values of $\sim$ 50 cm²/Vs, subthreshold swing values of 90-100 mV/decade, and on/off ratios higher than 10⁷ [66]. On the other hand, by developing a surface dopant-profiling technique, Feng’s group reported the first high hole mobility WSe₂ ML-FETs with body thickness of $\sim$ 0.7 nm. The use of heavily p-

doped contacts is essential in lowering the metal contact resistances by orders of magnitude and enabling the demonstration of p-FETs with peak effective mobility of $\sim 250 \text{ cm}^2/\text{Vs}$, which is near ideal subthreshold swing (SS) of $\sim 60 \text{ mV/decade}$ and high on/off ratio of $>10^6$ [19].

Along with the development of high mobility 2D TMD-material and low contact resistance techniques, the achievements related to TMD transistors made so far are remarkable and further progress is expected. Due to their relatively low channel mobility, MoS₂-FETs show worse f_T performance over graphene transistors, however, they have big advantage of a semiconducting channel with a sufficiently wide gap, resulting in excellent switch-off as well as good drain current saturation, and thus reasonable power gain and f_{max} . For the state-of-the-art TMD-FETs, the experimental work on few-layer MoS₂-FETs has achieved a cutoff frequency of around 6.7 GHz for 250 nm channel with a maximum oscillation frequency up to 5.3 GHz [54]. Hereafter, large domain bilayer MoS₂ transistor with gate length 90 nm has been studied with extrinsic cutoff frequency $f_T = 7.2 \text{ GHz}$ and extrinsic maximum oscillation frequency $f_{max} = 23 \text{ GHz}$ [67]. The current record RF performance for MoS₂-FETs is defined by a 68 nm gate transistor on a rigid SiO₂/Si substrate presenting an f_T of 42 GHz and an f_{max} of 50 GHz, and its counterpart on a flexible substrate with $f_T = 13.5 \text{ GHz}$ and $f_{max} = 10.5 \text{ GHz}$ [68], thus, rendering MoS₂-FETs a very promising device for flexible digital and RF applications [Figs. 2.5(c)-(e)].

It is worthwhile to point out that the TMD-transistors can be even used for THz applications. TMD-based FETs have shown superior intrinsic RF FOMs in the range of terahertz that can surpass the performance of conventional high-frequency channel materials. In 2D material based devices, a cutoff frequency of about 1.4 THz has been reported for graphene transistor, but their intrinsic gain and maximum oscillation frequency have been found to be about 10 dB and 50 GHz, respectively [69], while for MoS₂-FETs, the previous simulated results have shown an intrinsic gain higher than 20 dB [57] and the intrinsic cutoff frequency approaching terahertz for sub-5-nm channel of MoS₂ [58]. Therefore, 2D TMD-based FETs seem to have a great potential to allow RF performance above 1 THz. The optimization of material geometry and choice, as well as an effective device design strategy can allow such frequency of operation despite the presence of realistic parasitic values.

2.3 Frequency mixers and signal modulators based on 2D transistors

Due to the Dirac-cone dispersion of graphene, major carriers of a GFET can be varied from electrons (i.e., n-dominated transport) to holes (i.e., p-dominated transport) by sweeping the dc bias voltage. The gate voltage that gives the minimum conductance of the graphene channel is called the Dirac-point voltage or the charge neutrality point (CNP) voltage V_{cnp} . Such an ambipolar transport property in GFETs, with a symmetric “V-shape” I_{ds} - V_{gs} curve, possesses the potential for building the new generation of RF circuits and components, such as frequency multipliers and signal modulators [70]-[72]. In modern communication devices, the baseband signal that contains information must be embedded into the RF carrier signal through proper modulation. Common types of signal modulation include the amplitude modulation (AM), frequency modulation (FM), and phase-shift keying (PSK). A high-frequency GFET illustrated in Fig. 2.6 may allow a simple implementation for realizing all these modulation schemes [70]. To achieve the digital modulation, the RF carrier signal ($V_{carrier}$) and the data bitstream (V_{data}) are added together and applied to the gate of the transistor, as shown in Figs. 2.6(a)-(b). The amplitude modulation [left panel in Fig. 2.6(b)] is achieved by utilizing the transconductance change over the gate voltage difference in an GFET. The frequency modulation [right panel in Fig. 2.6(b)] is achieved by interchanging the bias point from a region dominated by electrons (or holes) to the charge neutrality point. Similarly, the phase modulation [middle panel in Fig. 2.6(b)] is achieved by changing the bias point from a region dominated by electrons (or holes) to a region dominated by holes (or electrons). Apart from these basic binary modulation schemes, a GFET-based modulator also allows the quaternary digital modulation schemes, such as quaternary amplitude-shift keying (4-ASK) and quadrature phase-shift keying (QPSK) [73], [74], thereby representing a drastic reduction in circuit complexity and a capacity enhancement when compared with conventional modulators. Fig. 2.6(c) shows the example of the quaternary modulation scheme (4-ASK) done by combining the binary amplitude-shift keying (BASK) and the binary phase-shift keying (BPSK). It is obviously seen that the four bias points, corresponding to 00, 01, 10 and 11, are used in this 4-ASK scheme such that both phase and amplitude information can be exploited to distinguish the quaternary signal encoded in the carrier wave. As a result, the data transfer rate can be doubled when compared with binary modulations.

Similarly, a high-frequency GFET can be also be exploited to make frequency mixers and multipliers, with at least one stage of filter removed. These RF components are often used in frequency synthesizers and communication circuits, and are the critical frequency-conversion elements in the building block of modern RF transceivers/receivers. When an GFET shown in the right panel of Fig. 2.6(b) is used as a frequency doubler, it can effectively suppress the odd-order intermodulation and contribute to lower spurious emissions among the circuit because of its symmetrical ambipolarity, namely the symmetric I_{ds} - V_{gs} characteristic centered on V_{cnp} . In 2009, the first GFET-based mixer has been demonstrated, with a third-order intercept point of +13.8 dBm [75]. However, the conversion loss of this mixer is as low as 40 dB and the operating frequency is limited to 10 MHz, which are primarily restricted by the poor quality of graphene, as well as the relatively large gate length and gate capacitance C_{gs} . Later on, the wafer-scale high-performance GFET has been fabricated by IBM [76] for making an RF mixer, which exhibits an improved conversion loss ~ 27 dB and an operating frequency of 10 GHz. In the meanwhile, this GFET-based mixer has exceptional thermal stability with little reduction in performance (up to 400 kelvins). Take one step further, in 2013, an integrated graphene radio frequency circuit was reported [77], which can simultaneously operate as a radio frequency signal amplifier, filter and the down-conversion mixer with the conversion loss as low as -10 dB at $f_{LO} = 4.3$ GHz as shown in Figs. 2.7(a)-(c). We note that when compared to conventional mixers based on III-V semiconductor transistors (which generally require complicated external passive components), the GFET-based RF mixer is simple and may significantly reduce the fabrication cost and area of occupation [76]. Furthermore, in order to increase its operating frequency to the millimeter-wave region, monolithically-integrated subharmonic GFET-mixers have been recently demonstrated, with the operating frequency ranging from 2 GHz to 200 GHz [71]. One example of the subharmonic resistive GFET-mixer implemented with coplanar waveguide technologies is shown in Figs. 2.7(d)-(f), which exhibits a conversion loss of 29 dB across the 185 – 210 GHz band under 12.5 – 11.5 dBm local oscillator (LO) pump power, a >15 dB LO – RF isolation, and a 3 dB IF bandwidth of 15 GHz. Different from conventional RF mixers, this new platform can effectively suppress the LO noise and simplifies the separation between LO and RF signals.

2.4 Integrated wireless IoT sensors

Atomically-thin 2D materials with strong conductivity dependency on chemical/photo/electrostatic gating effects have had a revolutionary impact on electronic sensors [78], [79]. Due to the relatively low density of states of 2D materials, a small change of carrier concentration due to doping effects may dramatically vary the sheet conductance of 2D materials, thus enabling nanosensors with ultrahigh sensitivity. Moreover, carbonous materials, such as carbon graphene and carbon nanotubes, are biocompatible and have good specificity because surface functionalization can readily create selective receptors on carbon atoms. The state-of-the-art graphene and MoS₂ ion-sensitive transistors have demonstrated a single-molecule sensitivity [80]-[82]. Nonetheless, it remains challenging to integrate 2D nanodevices with traditional silicon-based front-end modules.

In wireless sensor and radar systems, nonlinear harmonic tags, which receive a RF monotone and backscatter a second/sub-harmonics, are commonly used to suppress clutters, echoes, cross-talks and other electromagnetic interferences, provided that the sensor/tag has a compact (electrically-small) size [83], [84]. As discussed in the previous section, GFETs can exhibit a cutoff frequency (f_T) up to hundreds of GHz, which is higher than most of existing solid-state transistors. Its unique symmetric and “V-shape” drain current-gate voltage characteristic allows one to make the simplest possible frequency doubler which covers the 5G/6G, Bluetooth, and Wi-Fi bands [42]. Further, since the electrical conductivity of graphene is highly sensitive to gas, chemical or biological dopants, a single back-gate or side-gate GFET may realize a chemically-sensitive frequency doubler, in which a shift in Dirac point (or charge natural point V_{cnp}) caused by chemical gating effects can modulate the second-harmonic output signal and thus vary the frequency conversion efficiency [9], as illustrated in Fig. 2.8. In this case, the information of the concentration of the specific dopant is encoded in the frequency-modulated RF signal, which in some sense, is similar to the frequency modulation (FM) used for wireless communication. As a result, high-performance GFETs having the capabilities of biochemical sensing and frequency modulation may pave a promising way towards a compact and ultrasensitive harmonic-transponder sensor (or harmonic sensor).

Fig. 2.9 shows the evaluation of RF spectrum for a back-gate GFET in an air exposure-ethanol recovery modulation cycle [8]. In this experiment, the GFET is first exposed to ambient air such that the chemical gating occurs due to adsorption of oxidizers (e.g., NO₂, O₂ in air) that induce charge transfer between the graphene and the adsorbed substance. Fig. 2.9(a) shows the spectra of the modulated output signals for a monotone AC input signal with a 3-V peak-to-peak voltage at the back gate, under a 100-mV DC drain bias. The output vector, which comprises the fundamental tone at 25 kHz and the second-harmonic tone at 50 kHz, has a signal-to-noise ratio (SNR) of <50 dB, 16 dB>. Figs. 2.9(b)–(c) present the detected signals in the air-exposure modulation phase, during which the output vector decreases to <43 dB, 9.5 dB> SNR after 2 minutes, and to <41 dB, 0 dB> SNR after 4 minutes. It is clearly seen that the signal strength of the second-harmonic tone is detuned from 16 to 0 dB due to the shift in V_{cnp} , which serves as a good indicator of the level of chemical exposures. On the other hand, for a doped graphene channel with an up/down-shifted V_{cnp} , the fundamental tone is amplified due to the increased linearity of I_{ds} - V_{gs} curve seen by the time-varying sinusoidal input voltage. After the wet treatment, the output can be reset to its original signal shape, as shown in Fig. 2.9(a).

Although a single GFET can serve as a multi-functional sensing-modulator, it may not be suitable to build a fully-passive harmonic sensor because the device still needs a drain-to-source dc bias (i.e., a power source, such as the piezoelectric transducer, micro-battery, rectification circuit, or solar cell). Some recent seminal works have proposed the practical realization of self-powered harmonic sensors based on the graphene integrated circuits. For instance, a quad-ring mixer or dual-ring mixer [Fig. 2.10(a)] based on connected GFETs can achieve the chemically-sensitive frequency modulation, without any external power source. Fig. 2.10(b) shows the frequency conversion efficiency versus V_{cnp} (simulation) for a self-powered quad-ring multiplier based on four GFETs, showing that the frequency conversion efficiency strongly depends on the concentration of surface absorbates which determine the level of chemical gating. In the balanced condition, the conversion efficiency (the voltage ratio between second harmonic and fundamental tone) is zero if all four graphene receptors are matched to the zero V_{cnp} condition [85]. The strength of the detected second harmonic reflects the value of V_{cnp} , which could be shifted due to the existence of p-/n-type (bio-)chemical dopants. A dual-ring GFET-based sensing-modulator may further

reduce the device footprint and fabrication complexity, while offering improved linearity and reliability compared with the quad-ring design [86]. The all-graphene sensing-modulator can be mounted to the RF front end that include RFID/NFC chips and antennas for building the battery-free wireless IoT sensor that can alarm specific chemical or biological events in real time [85], [86], as schematically shown in Fig. 2.10(a).

2.5 RF energy harvesting using 2D electronic devices

In brief, apart from the aforementioned RF applications associated with 2D transistors, the 2D material can also be utilized to generate other types of high-performance electronic devices, such as 2D Schottky diodes shown in Fig. 2.11 which has been demonstrated recently for energy harvesting platform. Due to mechanical robustness and stability nature of 2D materials, a flexible rectenna based on the MoS₂ Schottky diode has been developed to harvest the ambient RF and Wi-Fi radiations in the global-satellite-positioning band (1.58 GHz and 1.22 GHz), cellular-communications fourth-generation (4G) long-term-evolution band (1.7 GHz and 1.9 GHz), Bluetooth (2.4 GHz), and Wi-Fi channels (2.4 GHz and 5.9 GHz) [87]. Such a lateral diode configuration based on MoS₂ semiconducting – metallic (2H – 1T/1T') phase heterostructure can have a much reduced junction capacitance C_j (which could be lower than 10 fF) and a small effective series resistance R_s , which in turn results in a power conversion efficiency of up to 40 % and a high cutoff frequency of 10 GHz, as shown in Fig. 2.11(b)-(c). This integrated MoS₂-based rectifier provides a universal energy-harvesting building block that can be fused into various flexible electronic devices for IoT applications.

2.6 Conclusion

In this chapter, we provide a comprehensive review of active 2D devices and circuits which can be utilized in the RF front-end building blocks for the next-generation wireless communication and IoT systems. We first discuss characterization and modeling of 2D material-based transistors working at high frequencies. Then, we mention several representative 2D devices used for amplification, mixing, multiplication, modulation/demodulation, and processing of AC signals. There are a rich variety of RF

and analog applications enabled by newly discovered electronic properties of 2D materials. Finally, we have discussed the potential of 2D materials in the development of wireless nanosensors and energy harvesting platforms for the low-power, near-ubiquitous IoT sensors and integrated systems.

Fig. 2.1. Schematics circuit diagram for a wireless sensor, of which high-performance 2D material-based devices can be exploited in several active components (highlighted with gray shadows).

Fig. 2.2. (a) Schematic of a 2D material transistor. (b) Small-signal equivalent circuit of a FET. (c) RF model of conventional FETs under test, where Y_{p1} , Y_{p2} and Y_{p3} are parallel parasitics and Z_{L1} , Z_{L2} and Z_{L3} are series parasitics (They represent FETs contact pads and interconnects). (d) Small-signal current gain h_{21} and unilateral power gain U of an RF FETs versus frequency. Reprinted with permission from (Ref. [3]). Copyright (2013) IEEE. (e) Measured scattering parameters, S_{11} , S_{12} , S_{21} , and S_{22} , of a graphene transistor for de-embedding. (f) The current gain h_{21} calculated from the measured S parameters as a function of frequency. The solid line corresponds to the ideal $1/f$ dependence, or equivalently, -20 dB/decade slope, of the current gain. The cutoff frequency is determined to be 4 GHz. The drain and gate voltages are 1.6 and 0.5 V, respectively. Reprinted with permission from (Ref. [37]). Copyright (2009) American Chemical Society.

Fig. 2.3. (a) Optical and scanning electron microscopy image of the device layout with ground-signal-ground accesses for the drain and the gate. (b) Measured current gain h_{21} as a function of frequency of a GFET with L_G 150 nm, showing a cutoff frequency at 26 GHz. The dashed line corresponds to the ideal $1/f$ dependence for h_{21} . Inset: the maximum f_T as a function gate length for the four GFETs measured. Reprinted with permission from (Ref. [37]). Copyright (2009) American Chemical Society. (c) Device schematic of the dual-gate graphene transistor. (d) SEM image of a double-channel graphene transistor. The channel width is 27 μm, and the gate length is 350 nm for each channel. (e) Peak transconductance g_m as a function of series resistance R_S . (f) RF performance of a 350-nm-gate GFET, showing a current gain at -20 dB/dec and a cutoff frequency f_T of 50 GHz. Reprinted with permission from (Ref. [42]). Copyright (2010) IEEE.

Fig. 2.4. Graphene transistors with a self-aligned nanowire gate. (a) Schematic illustration of the three-dimensional view of the device layout. (b) Transconductance at $V_{ds} = -1$ V as a function of top-gate voltage V_{TG} before (black) and after (red) the deposition of the self-aligned Pt source/drain electrodes. (c) Small-signal current gain $|h_{21}|$ vs. frequency f at $V_{ds} = -1$ V for a device with gate length = 144 nm. Reprinted with permission from (Ref. [39]). Copyright (2010) Springer Nature. (d) A cross-sectional SEM image of a self-aligned graphene device with a GaN nanowire gate, highlighting the flat interface between graphene and triangular GaN nanowire. Reprinted with permission from (Ref. [7]). Copyright (2010) American Chemical Society. (e) The projected intrinsic cutoff frequency f_T at $V_{ds} = 1$ V calculated using $f_{T-int} = g_m/2\pi C_g$, versus channel length. Reprinted with permission from (Ref. [45]). Copyright (2012) Elsevier.

Fig. 2.5. (a) Schematic illustration of the three-dimensional view of the MoS₂ FETs layout. (b) Trilayer MoS₂ with cutoff frequency $f_T = 6$ GHz at applied $V_{ds} = 2.5$ V and $V_{TG} = -1$ V. The current gain h_{21} decreases with increasing frequency, following the -20 dB/dec slope expected for conventional FETs. Reprinted with permission from (Ref. [55]). Copyright (2014) American Chemical Society. (c) A schematic illustration of a dual-channel self-aligned MoS₂ FETs with transferred gate stacks, and the inset shows the schematic cross-section of the self-aligned device. (d) Small-signal current gain $|h_{21}|$ versus frequency for three devices with a channel length of 216 nm (blue), 116 nm (red) and 68 nm (black). The cutoff frequencies are 13.5, 26 and 42 GHz at a d.c. bias of 5V and gate bias of 2.1, 2.3 and 1.9V, respectively (e) Maximum available gain (MAG) versus frequency for three devices with the same parameters above. Reprinted with permission from (Ref. [68]). Copyright (2014) Springer Nature.

Fig. 2.6. (a) A circuit diagram with a false-colour image of graphene transistors connected in a common-source configuration. The V_{data} signal is the digital data that is encoded onto the carrier signal V_{carrier} . The V_{data} signal is a square wave for all three binary digital modulation schemes and a four level step-like wave for the 4-ASK modulation scheme. (b) Illustrations of amplitude, frequency and phase modulation of a sinusoidal wave achieved by operating a single ambipolar graphene transistor at different gate biases. (c) A time domain plot of the 4-ASK modulation of carrier signal. The inset is an illustration describing the four operating gate bias points used in 4-ASK. V_{DD} is 1 V. Reprinted with permission from (Ref. [70]). Copyright (2012) Springer Nature.

Fig. 2.7. (a) Optical micrograph of the integrated circuit under testing, which comprises 11 active and passive components. (b) Output spectrum of graphene receiver with 900 nm gate length. RF input signal was 20 dBm at 4.3 GHz and LO input signal was 2 dBm at 4.2 GHz. The down converted IF output signal at 100 MHz was 30 dBm. (c) Normalized output power as a function of RF. Reprinted with permission from (Ref. [77]). Copyright (2014) Springer Nature. (d) Equivalent circuit diagram illustrating the selected mixer circuit topology. (e) Micrograph of the fabricated mixer circuit including the dimensions. (f) Conversion loss versus RF frequency. The LO frequency is swept and the IF frequency fixed. Inset: P_{IF} versus P_{RF} at $f_{RF} = 200$ GHz. Reprinted with permission from (Ref. [71]). Copyright (2017) IEEE.

Fig. 2.8. (a) Schematics of a GFET-based sensing-modulator. (b) Mechanism of mixed AM/FM GFET modulation; Red: if V_{cnp} is zero, then the output consists only of a strong harmonic tone; Yellow: when V_{cnp} shifts away from 0 V, a large fundamental tone is present and the harmonic tone decreases; Blue: when V_{cnp} shifts higher, both the fundamental tone and the harmonic tone decrease, but the harmonic tone decreases more rapidly. Reprinted with permission from (Ref. [85]). Copyright (2016) AIP Publishing.

Fig. 2.9. Realization of air exposure-wet treatment AM/FM mixed modulation cycle on a chemical gate GFET with a single tone back-gate input carrier signal at 25 kHz. (a) Initially when V_{cnp} is around 3 V, the output vector is <50dB, 16dB> SNR, with the noise level at -80dBV. (b) V_{cnp} shifts to higher voltage after 2 minutes exposure and the output drops to <43dB, 9.5dB> SNR. (c) Further air exposure for another 2 minutes results in an even higher V_{cnp} and the output is <43dB, 0dB> SNR. The operation condition can be reversed by the ethanol based wet treatment. Reprinted with permission from (Ref. [8]). Copyright (2016) Springer Nature.

Fig. 2.10. (a) The harmonic sensor system based on the compact, self-powered GFET sensing-modulator. (b) Conversion efficiency versus V_{cnp} in each GFET receptor for the qual-ring sensing-modulator in (a).

Fig. 2.11. Flexible rectenna based on a 2D self-aligned MoS₂-heterostructure Schottky diode. (a) Schematic of a lateral Schottky diode based on a MoS₂ semiconducting–metallic (2H–1T/1T') phase junction. The gold layer forms an Ohmic contact with metallic (1T/1T') MoS₂, which also forms an Ohmic contact with semiconducting (2H) MoS₂. The palladium layer forms a Schottky contact with semiconducting (2H) MoS₂. (b) Power efficiency of MoS₂ rectifiers as a function of input power in the Wi-Fi band. (c) Output voltage as a function of frequency of the input signal for RF power P_{RF} = 5 mW, which implies the cutoff frequency of this MoS₂ rectifier to be 10 GHz. Reprinted with permission from (Ref. [87]). Copyright (2019) Springer Nature.

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