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Measurement of brightness temperature of two-dimensional electron gas in channel of a high electron mobility transistor at ultralow dissipation power

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ABSTRACT

A technically simple and physically clear method is suggested for direct measurement of the brightness temperature of two-dimensional electron gas (2DEG) in the channel of a high electron mobility transistor (HEMT). The usage of the method was demonstrated with the pseudomorphic HEMT as a specimen. The optimal HEMT dc regime, from the point of view of the "back action" problem, was found to belong to the unsaturated area of the static characteristics possibly corresponding to the ballistic electron transport mode. The proposed method is believed to be a convenient tool to explore the ballistic transport, electron diffusion, 2DEG properties and other electrophysical processes in heterostructures.

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1. Introduction

It is a common knowledge that the impact of a measuring device onto the object-under-test should be minimized if working with signal sources of essentially quantum nature. This is a general problem of non-disturbing quantum measurements. For electronic detecting facilities, especially amplifiers, it implies minimizing the energy, of noise or other origin, irradiated backwards to the object-under-test. The effect is a so called "back action", the phenomenon causing uncontrolled destruction of a quantum state of the object, e.g., qubit decoherencing, etc. [1–3]. The back action is detected and requires a quantitative description in a wide frequency band, orders of magnitude wider than the amplifier operation frequency band. Therefore, the "equivalent noise temperature" (T_n) determined for a relatively narrow operation frequency band of the amplifier (receiver) is an inadequate term here. Instead, the wide-spectrum brightness temperature (T_b) of the amplifier and its active elements should be considered as a proper quantifier, since it characterizes the total power of noise irradiation in a wide frequency band, from infra-low frequencies to optics.

One should clearly distinguish between the brightness (T_b) and noise (T_n) temperatures as referred to Friis formula [4]. The T_n obeys it while T_b does not. Let us illustrate this statement by an example. An ideal (not irradiation-contributing) attenuator with attenuation L placed after a source with temperature T_b will weaken the source irradiation power, in temperature units, down to T_b/L. When attenuation L tends to infinity, the attenuated power falls down to zero. In contrary to T_b, the T_n following Friis formula will rise as T_nL tending to infinity along with the attenuation. Thus, the noise temperature T_n is a conditional value although having a clear physical meaning. In contrary, the brightness temperature T_b, being in a sense an effective temperature since the noise irradiation is not always equilibrium one, characterizes nevertheless a real power flow which is the back action in the current context.

The amplifiers intended for ultra-low temperature applications (to amplify signals from quantum detectors, single electron transistors and variety of other quantum structures) are typically based on field-effect transistors (FETs). Among them, a class of HEMTs is distinguished, the high electron mobility transistors. HEMTs feature a very wide operational frequency band while field-induced (as opposed to thermally-generated) current-carrier electrons, which form two-dimensional electron gas (2DEG) in the channel, principally enable the transistor to operate at temperatures down

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to the absolute zero. Owing these advantages, the HEMTs are widely used in ultra-sensitive readout amplifiers [5,6] for measuring quantum device signals. Consequently, a quantitative description of the back action as applied to HEMTs is a hot issue.

Thermal noise is generated in HEMT input (gate-source) terminals due to power dissipation in the transistor input circuit. The dissipative losses come mainly from the gate metallization resistance, under-gate channel resistance and source resistance. Corresponding irradiation (for perfect matching, or zero input reflection coefficient) is characterized by the gate temperature T_g which is close to the physical temperature of the transistor crystal lattice T_{latt}. Cooling down to cryogenic temperatures is an effective method to suppress the thermal irradiation. If an ultra-deep cooling is supposed, it should be accompanied by a considerable decrease (down to a few microwatts and less) in the transistor consumed/dissipated power to avoid excessive Joule overheating of its active area. The situation with overheating becomes even more severe because of low thermal conductivity of the heterostructures [7]. Small cooling capacity of ultra-low-temperature cryorefrigerators, especially below 100 mK, also strongly limits the HEMT dissipated power. Provided if heat sink is effective, the input-circuit-generated thermal noise is reduced sufficiently and can be neglected regarding back action.

The "hot" electron irradiation from the gate-drain channel region is another cause of the back action. A high effective electron temperature in the drain region T_d exceeding the lattice temperature by two orders of magnitude for commonly used saturated HEMT regime is inherent for this mechanism. The irradiation of 2DEG in the gate-drain part of the channel goes backward to input via intrinsic drain-gate capacitance of the transistor. The excitation of waveguide modes by output circuit in a conductive cavity (where the amplifier, often along with a signal source, is placed) is an additional way.

An estimation for T_b of the amplifier input which governs the back action can be easily derived using reverse transmission gain (S_{12}) from the transistor S-matrix and two-temperature Pospieszalski model [8] treating the HEMT amplifier input as a black body. Thus, for an ideal matching of complex impedances of a source and the amplifier:

$$T_b = T_g + |S_{12}|T_d, \qquad (1)$$

The effect of T_d can be roughly estimated assuming (see above) $T_g \approx T_{latt} \approx T_{amb},\ T_d \approx 100T_{latt} \approx 100T_{amb}$ [8,9] for commonly used saturated HEMT regime. For the ultra-low-consumption (unsaturated) HEMT regime [10], as we will see below, T_d can be much smaller, down to $T_d \approx T_{latt} \approx T_{amb}$.

Typically, $|S_{12}|$ is about $-20\ldots-30$ dB at 1 GHz frequency. The $|S_{12}|$ rises almost linearly with frequency so the effect of T_d can prevail over that of T_g. If the amplifier has a high input impedance (S_{12} is defined for 50-Ω network), then the reverse transmission increases stimulating one to search for the ways of T_d reduction.

It is worth noting that the problem of optimal matching of the signal source and amplifier impedances is of great importance when designing an ultra-low noise amplifier. Optimal matching circuit synthesis based on adequate equivalent circuits of the signal source and amplifier first-stage transistor is a classic radio-engineering task. In the case of readout amplifier with low back action, the matching optimization is a special multifactor problem which is out of scope of the paper. We will use the simplified expression (1) as an "upper estimate".

The T_d and T_g are commonly used to calculate basic noise characteristics of a transistor, namely, the minimal noise temperature, optimal source impedance and noise conductivity [8]. Both T_d and T_g figures are extracted from a series of noise measurements by solving inverse problem [9] on the basis of the electrophysical transistor model which is inevitably limited to certain frequency band and temperature range. Integrally, the extraction procedure is sophisticated and ambiguous. The same can be said about other noise models and other similar noise invariants [11]. In the context of the back-action problem, it would be desirable to elaborate a simple method to measure mainly T_d since $T_g \approx T_{latt}$.

First, to measure T_d instrumentally, the contribution of the amplified noise of input circuit to the integral output noise irradiation of the transistor should be excluded. Referring to modern transistors with cut-off frequencies of tens and more gigahertz, such an elimination is hard enough because of stability problem. Moreover, it becomes much more complicated under deep-cooling conditions. However, the ultra-low power consumption of the transistor associated with deep-cooled amplifiers results in decrease in the cut-off frequency by two orders of magnitude while the stability factor exceeds the unity. Consequently, the stability is not further an issue, and direct instrumental measurements of T_d become possible.

Wide-spectrum noise irradiation of the HEMT channel is governed by the electron temperature T_{el} of the 2DEG which by both physical sense and value is close to T_d, as it will be seen from the following. Thus, both T_{el} and T_d are, in fact, brightness temperatures that determine a wide-band noise irradiation from the HEMT amplifier output (bearing in mind matching considerations). In this work we propose a simple method to measure directly the brightness temperature of 2DEG in a HEMT channel. The experimental results are discussed and recommendations on choice of the HEMT dc regime are formulated concerning the back action phenomenon.

2. Measurement technique

To measure the brightness temperature of a HEMT channel, we used a technique which was principally based on the routine procedure for calibrating the noise of an active four-terminal device with the thermal noise of a reference resistor [12], but modified it with some novelties to take into account inherent features of the object-under-test. While measuring the noise of a cooled HEMT channel, one should bear in mind the following. Firstly, the informational signal has a noise nature and very low power. Secondly, the channel differential resistance in the regime of interest may range, roughly, from 100 to 1000 Ω [13]. If one use a standard noise-measuring equipment at room temperature, the signal would be highly attenuated and distorted by shunting action of the transmission line at frequencies of about tens megahertz and above. At lower frequencies, the $1/f$ type noise expectedly emerges. To eliminate effect of the transmission line, we used a separate amplifier placed near the HEMT in the cryostat. The absence of a mechanical switch alternating connection of the amplifier input to either object-under-test or noise reference is a yet another important feature of our measuring procedure which enhances the result reproducibility.

Although the proposed procedure exploits neither two-temperature noise source (hot/cold loads method), nor fixed cooled attenuator with an external noise source (cold attenuator method), our method could be referenced to as a modification of the well-known Y-factor method.

Simplified diagram of the experimental setup is shown in Fig. 1. The essence of the technique is by-turn measurement and further comparison of the powers of two signals. The first one is produced by output circuit of the transistor-under-test (Q1), the second one is a reference, taken from a variable resistor R having ambient temperature. During the calibration procedure, the resistance R is set equal to the channel differential resistance in a specified point of the transistor static characteristics. The measurement cycle is described below in more detail.

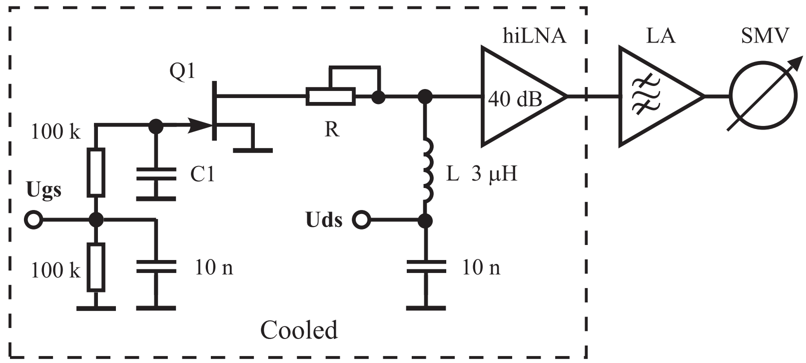

Fig. 1. Simplified diagram of the testing setup.

The transistor gate is ac-shorted to source by C1 to exclude the amplified input circuit noise from the net output signal.

The self inductance of capacitor C1 (SMD 0603, 330 pF) along with the inductance of Q1 gate terminal does not exceed 3 nH. The associated reactance at a mean operation frequency (50 MHz) is inductive and not greater than 4 Ω while the capacitive reactance of the gate-source is more than 3 *k*Ω. Under these conditions, only the noise component representing thermal noise of source terminal resistance is amplified contributing to T_{el}. The source resistance does not exceed 3 Ω for practically all low-power HEMTs. Taking into account that the transistor voltage gain does not exceed 5 under microcurrent dc regime, it is easy to show that contribution of the source noise in the output signal is less than 1%.

The transistor Q1 and the variable resistor R are placed close to an instrumental amplifier to minimize shunting capacitance. A three-stage HEMT (AVAGO ATF35143) high input impedance (100 kΩ), low-noise cooled amplifier (hiLNA in Fig. 1) is used as the instrumental amplifier. The hiLNA gain is about 40 dB, the operational frequency band is 20–100 MHz. The integral noise temperatures of the hiLNA are 2.3 ± 0.5 K and 1.2 ± 0.5 K at the source resistance of 10 Ω and 1000 Ω, respectively, and the ambient temperature $T_{amb} = 4.2$ K. The amplifier circuitry is similar to earlier reported devices [14,15].

The measuring circuit is a low-Q parallel tank consisted of the resistor R and HEMT channel resistance, capacitance of the wires, pads and hiLNA input (totally 5 pF) and inductance L (3 μH).

A linear amplifier (LA) with a band-pass filter (40...80 MHz) is placed next to the hiLNA. The specified frequency band (or rather the lower edge) is chosen in order to exclude the $1/f$ noise of Q1.

The output signal level is measured by a square-meter voltmeter (SMV) within an accuracy of better than 1%. Multisection filters are put into supply circuits (not shown in Fig. 1) and the test unit is electromagnetically shielded. The measurement errors are 0.1% and 1.5% for the dc voltages and currents, correspondingly.

The measuring cycle is a three-stage one. First, the resistor R is set to zero resistance, and static characteristics of the transistor are measured. The drain-source resistance r_d is calculated from the data obtained as a function of U_{ds} and U_{gs}. The SMV readouts are taken for each calculated r_d. According to the measurement procedure and Nyquist theorem, the SMV voltage, which is proportional to the output signal power, can be written as:

$$\langle U_m^2 \rangle = \langle U_0^2 \rangle + 4kr_d T_{el} G_m^2 \Delta F_m, \tag{2}$$

where $\langle U_m^2 \rangle$ is the mean square of the measured output voltage at a specified drain-source resistance r_d, $\langle U_0^2 \rangle$ is the mean square of the measured output voltage at $r_c = 0$ (see below), k is the Boltzmann constant, T_{el} is the brightness temperature of noise irradiation of the transistor channel (2DEG temperature in the case of HEMT), G_m is the band-averaged voltage gain and ΔF_m is the effective pass band of the measuring system, correspondingly, at specified r_d.

The second stage is a calibration. The transistor is zero-biased and maximally opened ($U_{ds} = U_{gs} = 0$). The channel resistance r_d is minimal ($r_{dmin} = 8 \pm 1$ Ω for ATF36077) and fully linear. The noise generated by r_d is purely thermal (white). It adds to the noise of the calibration resistor R with resistance r_c. The calibration includes tuning the calibration resistor R in the range of 0...670 Ω while the SMV readout is synchronously taken. In a manner similar to (2), we write the equation for the SMV output signal:

$$\langle U_c^2 \rangle = \langle U_0^2 \rangle + 4kr_c T_{amb} G_c^2 \Delta F_c, \tag{3}$$

where $\langle U_c^2 \rangle$ is the mean square of the measured output voltage at a specified resistance r_c of the calibration resistor R (exactly, $r_{dmin} + r_c$), T_{amb} is the physical temperature of the calibration resistor R, G_c is the band-averaged voltage gain and ΔF_c is the effective pass band of the measuring system, correspondingly, at specified r_c.

Finally, an expression for T_{el} can be derived from (2) and (3) taking $G_c = G_m$, $\Delta F_c = \Delta F_m$. These equalities are valid if the experimental values are taken from the data array with the selection rule $r_d = r_c$. Then we obtain:

$$T_{el} = T_{amb} \frac{\langle U_m^2 \rangle - \langle U_0^2 \rangle}{\langle U_c^2 \rangle - \langle U_0^2 \rangle} \tag{4}$$

To simplify the measurement and calculation procedures, a few assumptions were made when writing (2)–(4).

(i) $\langle U_0^2 \rangle$ = const is assumed, i.e. the noise temperature of the instrumental amplifier (hiLNA) does not depend on the source resistance. Actually, it varies (see above) although always staying below the temperature to measure, T_{el}.

(ii) $\langle U_0^2 \rangle$ is supposed to be measured at $r_c = 0$. In fact, the minimal value $r_{cmin} = r_{dmin} \approx 8$ Ω (for ATF36077).

(iii) It is believed that the dc channel resistance r_d is equal to the ac one, at the measurement frequency (40–60 MHz). In reality, the ac channel resistance is less by a few percent.

All the assumptions made are not too rough, and the systematic absolute error (see section 3) does not exceed 2 K. Nevertheless, it should be taken into account when T_{el} is about several Kelvins. When the measured T_{el} is of some tens Kelvins (the most important and interesting range), the error is determined by accuracy of the instrumental measurement being, by our estimate, of about 10% of the measured value. To the authors' opinion, such an error is acceptable in the context of this work.

Additionally, the following should be clarified here. The drain temperature T_d, as a parameter in the two-temperature Pospieszalski noise model [8], actually represents the effective temperature of 2DEG in the saturated region (roughly, the gate-drain region) of HEMT channel. Meanwhile, the expression (4) that we obtained for T_{el} represents an averaged 2DEG temperature throughout whole channel including the source-gate part with its temperature which is always close to T_{latt}. It is the 2DEG temperature that we designate by T_{el}. It is measured directly and has a clear physical sense. However, T_d and T_{el} do not differ too much since the resistance of the source-gate region is about 1/100 of total channel resistance.

The measurement of T_{el} can, in principle, be done without calibration resistor, using the transistor noise instead at $U_{ds} = 0$, $I_d = 0$. However, in practice such an operation sequence requires an extra (negative) bias voltage which could cause, for some transistors, the gate leakage noise to emerge. Note that for ATF36077 (which is the object-under-test, see below) at $U_{ds} \leqslant 0.2$ V and $U_{gs} = -0.2 - 0.35$ V, according to our measurements, the gate leakage current does not exceed some tenths of nanoampere and gives no noticeable noise contribution.

3. Results and discussion

The pseudomorphic HEMT AVAGO ATF36077 was chosen as a test object. We believe this transistor is a good representative of low-power HEMTs. Additionally, as our practice show, ATF36077 has a very low gate leakage current and a high mechanical stability during multiple thermal cycling. The ATF36077 static characteristics show no hysteresis. All said above makes ATF36077 a suitable object for testing the proposed measurement method.

The bias voltage and, correspondingly, maximal drain current were chosen so that the transistor dissipated power P_c would not exceed 200 μW. The larger P_c values are beyond the subject of this work. Also, at $P_c > 50$ μW the transistor-under-test should be carefully monitored for a parasitic oscillation to emerge which is provoked by the gate grounding. Consumption current of the hiLNA is a suitable indicator of the parasitic oscillations.

The measurements were taken at 4.2 K, 78 K and 290 K when developing the technique. The results of the measurements at 4.2 K are reported below.

Fig. 2 demonstrates the static characteristics of ATF36077. It is seen from the plots that the HEMT operates in unsaturated regime which was studied in detail earlier [13] down to ultra-low (50 mK) temperatures. For each branch of the set with specified U_{gs} the electron temperature was measured in several points.

Let us start with these data to estimate the measurement error. To do this, we note that measured T_{el} is assumed to be equal to T_{amb} when $U_{ds} = 0$, $I_d = 0$. It is quite reasonable from the physical point of view. Then the difference between T_{el} and T_{amb} in the static characteristic branch origin ($U_{ds} = 0$) can serve as a measure of the absolute error. The data show that the systematic error does not exceed 2 K for all four U_{gs}. Fig. 3 shows the electron temperature T_{el} as a function of the drain current I_d at fixed drain-source voltages U_{ds}.

It is clearly seen that $T_{el}(I_d)$ plot tends to a saturation and there is no increase of the electron temperature with current anyway. This fact witnesses for a negligible contribution of the channel diffusion noise induced onto the gate. Therefore, the gate is "well-grounded" while the operation frequency and the circuit design are adequate.

The T_{el} as a function of U_{ds} at fixed drain currents is plotted in Fig. 4. It should be noted at once that doubled electron temperature ($T_{el} = 2T_{amb}$) is observed at the voltages much higher than expected threshold values (of order of 1 mV) approximately following from the condition $kT = eU_{ds}$. The effect could evidence for the ballistic (collision-free) electron transport at corresponding parts of the static characteristics. Indeed, the main mechanisms of transport electron scattering are suppressed in the considered case: acoustic phonons are frozen out to great extent while optical phonons cannot be excited because of low (< 30 mV) voltage over the channel. Intervalley scattering is not observed in such low fields. 2DEG scattering by donor impurities is absent in HEMT in first approximation. Other mechanisms are minor ones. Hence, the ballistic regime is quite expectable. Surely, the question needs additional studies and should be considered here as an illustration of the proposed method application.

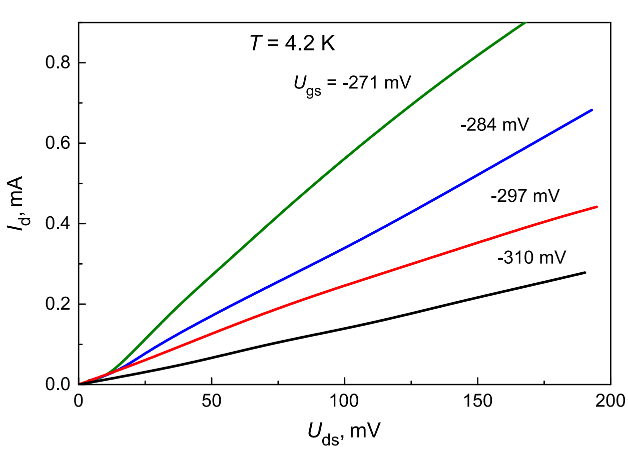

Fig. 2. Static I–V characteristics of ATF36077 at 4.2 K.

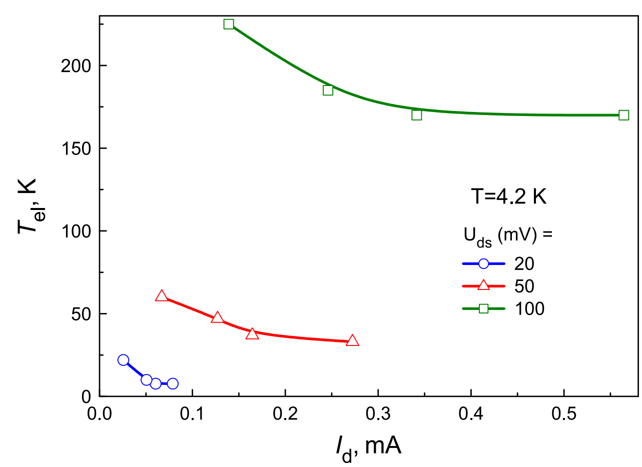

Fig. 3. Brightness temperature of 2DEG T_{el} as a function of the drain current I_d at fixed drain-source voltages U_{ds}.

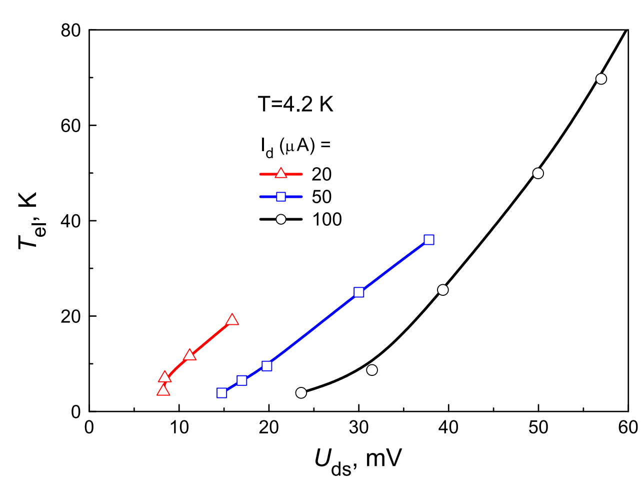

Fig. 4. Brightness temperature of 2DEG T_{el} as a function of the drain-source voltage U_{ds} at fixed drain currents I_d.

A distinctive bend of the curve taken at drain current of 100 μA may be associated with the threshold of optical phonon scattering mechanism activation (approximately 30 meV for the A3B5 semiconductors). The corresponding dc regime ($I_d \approx 0.1$ mA, $U_{ds} \leqslant 35$ mV) can be recommended for the first-stage transistors in amplifiers with minimized back action intended to function at sub-Kelvin temperatures. The HEMT ability of working at ultra-low supply voltages in unsaturated region of the static characteristics was reported earlier in [10,16]. The absence of the "bend" in the plots corresponding to drain currents 20 and 50 μA is most likely due to expansion of the under-gate depletion region towards the source. This effect probably causes the paradoxical rise of T_{el} with decreasing the channel current (Fig. 4) as well. When comparing Figs. 2–4 one can see that different physical mechanisms of the electron diffusion and drift (their specific nature is not important in the case) are clearly distinguished in the plots of T_{el} as a function of dc regime while are indiscernible in the static characteristics. Thus, the electron temperature measurement yields an important information for exploring the current flow processes.

Let us take conventionally $|S_{12}| = -10$ dB as a measure of the back action. Also, we define (using Eq. (1)) the criterion of acceptable irradiation in direction of the signal source: the temperature of such an irradiation must not be higher than $2T_{latt}$, that is $T_{el} \leqslant 20T_{amb}$. Bearing in mind these definitions, a recommendation useful in designing amplifiers with reduced back action can be derived from Figs. 3 and 4. Namely, the maximal acceptable value of U_{ds} is 100 mV for HEMT operating at $T_{amb} = 4.2$ K.

The above estimates and the plots analysis is rather qualitative. They aimed, in the context of this work, to demonstration of reasonability of measurement of the electron temperature in studying electrophysical processes in HEMTs and finding the optimal regimes from the point of view of the back action problem.

To estimate the back action at low frequencies where $1/f$ noise predominates, it is necessary to measure the amplifier noise temperature with standard procedure [5].

4. Summary of used temperature terms

1. T_{amb} is the HEMT environment temperature, here $T_{amb} = 4.2$ K.
2. T_{latt} is the physical temperature of a HEMT crystal lattice in the channel region, Joule-overheated by drain current, $T_{latt} \geqslant T_{amb}$.
3. T_b is the brightness temperature which is the power of a noise irradiation (here: the irradiation of a read-out amplifier backward to a signal source, i.e. the back action) calibrated by a black body irradiation and expressed in Kelvins.
4. T_{el} is the brightness temperature of 2DEG, here: the power of electrical noise of the HEMT channel expressed in Kelvins. As a rule, $T_d > T_{el} \gg T_{latt}$.

5,6. T_g and T_d are the gate and drain temperatures, correspondingly, which are parameters in the phenomenological noise Pospieszalski model [8]. Physically, T_g is close to T_{latt} while T_d can be considered as an effective electron temperature in drain part of a HEMT channel, so $T_d \gg T_g > T_{latt}$.

7. T_n is the equivalent noise temperature which is the characteristic of an amplifier. Unlike T_b, the T_n obeys Friis formula. Here: $T_n \approx 0.3T_{amb}$.

5. Conclusion

A technique is suggested in this paper for direct measurement of the brightness temperature of 2DEG in a HEMT. The technical simplicity and clear physical sense of the measurement results makes the proposed method a convenient tool for further studies of delicate electrophysical processes such as the ballistic electron transport and diffusion, 2DEG properties and so on.

Grounding on the results of the 2DEG brightness temperature measurements, recommendations are put for choosing the HEMT dc regimes which are the best concerning back action. It is found that the optimal HEMT operation points to minimize back action lie in unsaturated area of the static characteristics that possibly corresponds to the electron ballistic transit.

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