

Active bias network-based measurement set-up for the direct characterization of low-frequency noise currents in electron devices

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Abstract- A technology-independent set-up for the full and direct characterization of short-circuit low-frequency noise currents in two-port electron devices is presented. It is based on the use of low-noise transimpedance amplifiers and a specially-designed active bias network, adopted at the collector/drain port of the device. The features of the set-up proposed allow for a fast and accurate estimation of LF noise current power spectra, within a wide range of DC current bias levels, without the need for any re-configurations and by carrying out a single-step procedure to achieve each bias point of interest.

I. Introduction

The experimental characterization of low-frequency (LF) noise at the extrinsic ports of electron devices is a task commonly performed in the fields of technology performance evaluation and development [1,2], device reliability investigation [3,4] and empirical noise modelling [5-7]. All these applications require the accurate estimation of auto- and cross-spectral density functions (i.e., power spectra) of the short-circuit noise currents which describe from a Norton-equivalent (Fig. 1a), circuit-oriented standpoint, the effects of propagation up to the external nodes of several microscopic stochastic phenomena distributed within the active device physical structure (basically carrier population fluctuations, which lead to generation-recombination (G-R) noise, and fundamental scattering, which justifies flicker noise [8]).

The LF noise is usually measured as different flicker and G-R contributions to the current spectral densities within the bandwidth [100 Hz – 100 kHz], the active device being biased at different levels under quiescent operation. As a result, the empirical data obtained at the end of the measurement process are frequency- and bias-dependent functions describing the second-order statistics of the equivalent noise (EN) current generators ($\Delta i_1, \Delta i_2$) of Fig. 1a. However, classical approaches usually lead to an *indirect* estimation of these noise currents. In fact these methods, which are based on the exploitation of low-noise voltage amplifiers, actually measure noise *voltages* at the (suitably loaded) device ports and transform these data into the noise current spectra, by means of numerical algorithms requiring additional information such as the device bias-dependent impedance matrix and the set of different loads used during the tests [9,10].

More recently, alternative solutions to the LF noise device characterization issue have been proposed, which make use of *transimpedance* amplifiers (TIA) in order to directly estimate the noise current spectra [11,12]. The low input impedance shown by these components allows to absorb (nominally) all the noise current supplied by the EN generators of Fig. 1a. The currents are then converted to corresponding voltages by means of the high voltage/current gain of the TIAs, and a two-channel low-noise FFT analyzer can be exploited at the outputs of the TIAs for the power spectra estimation (Fig. 1b). In principle, the advantages introduced by these *direct* techniques strongly simplify the measurement procedure. In general they do not require the use of multi-loads and, under more specific conditions, either the characterization of the device bias-dependent small-signal (SS) response within the LF bandwidth. In addition, since such set-ups do not need any reconfigurations during the measurement process, the overall speed is highly increased and a larger set of different bias conditions can be explored in a shorter time. However, the hypothesis of considering the input of the TIAs as ideal short-circuits is acceptable only when the impedance shown by the device corresponding port is several magnitude orders higher. This can not be the case, for example, of both base and collector ports of a bipolar transistor biased according to a common-emitter configuration, in the presence of a high level of DC collector current. The input impedance of TIAs in ultra low-noise configuration is neither negligible nor constant but tends to rapidly increase with frequency in the bandwidth of interest as well. In addition, at some point of the procedure decreasing values of the collector bias resistor R_2 are needed as the DC bias current I_2 is increased, if reasonable and safe low-level voltage DC batteries are exploited for the power supplying (similar observations are still valid in the case of a FET technology, when the drain port is considered). Thus, also the impedance shown by the resistive bias system can not be

considered as an ideal open-circuit if an exhaustive, wide multi-bias LF noise investigation is needed, the issue of set-up reconfiguration requirements being also re-introduced. Finally, even when a single, fixed value of the bias resistor at port #2 can be considered acceptable, an iterative procedure is needed to step-by-step converge to a given bias point through the control of the DC batteries if a passive bias system is exploited.

In this paper a technology independent, fast measurement set-up for the direct and full (i.e., correlation is included) on-wafer characterization of short-circuit LF noise currents of two-port electron devices will be described. It can be considered as an improvement and generalization of the solution proposed in [12], the latter being originally intended for SiGe heterojunction bipolar transistors (HBTs), a technology usually operated at low levels of collector DC current. The major feature introduced in this work is the use of a specially-designed, low-noise *active* bias network exploited at the port #2 of the bipolar/FET (i.e., collector/emitter terminal) which allows to: *i*) investigate the LF noise behaviour of DUTs either in bipolar (BJT,HBT) or field-effect (MOSFET, MESFET, HEMT) technology, *ii*) explore wide regions of the device possible operation in terms of bias levels, including those associated with high collector/emitter DC currents, without the need for set-up re-configurations and by employing safe levels of DC voltage battery, *iii*) achieve each bias point in a single step, with high benefits in terms of measurement speed. In order to achieve a high measurement accuracy and fully exploit, at the same time, the wide bias range and high speed made possible by the use of the active bias network, some additional issues are taken into account in this work with respect to the basic measurement procedure based on the TIAs and related simplified assumptions, such as the characterization in terms of input impedance of all the elements of the proposed set-up, including the DUT.

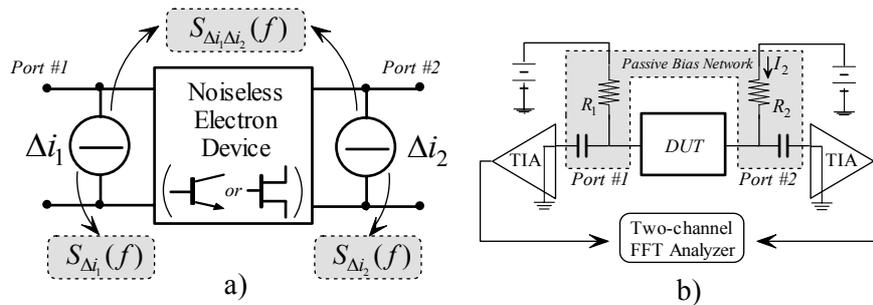


Figure 1. a): Norton-equivalent description of LF noise in terms of short-circuit noise currents and measurands of interest. b): Direct measurement of noise currents by means of ideal transimpedance amplifiers and resistive bias networks.

II. The architecture of the proposed set-up

The set-up is schematically depicted in Fig. 2. Each port of the DUT is accessed through a RF bias-tee [45 MHz – 50 GHz] closed at the RF terminals on a 50-ohm load, which ensures the stability of the device quiescent operation by providing a well-defined, resistive termination at high frequencies.

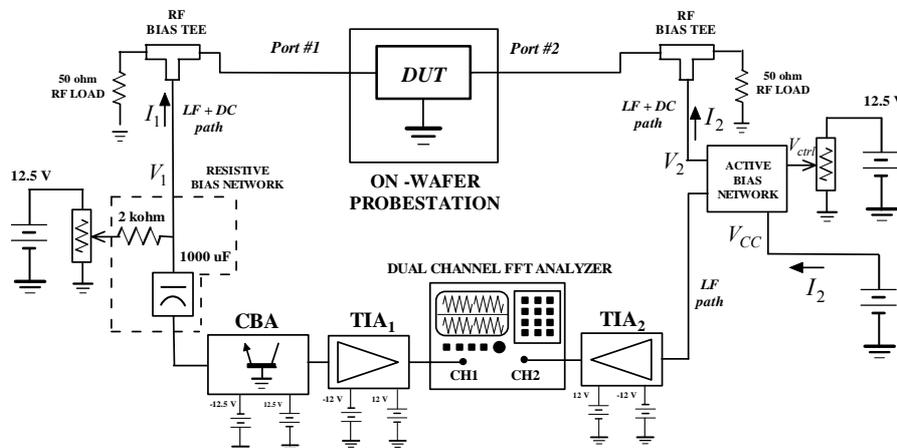


Figure 2. The set-up for the full characterization of LF noise currents in two-port electron devices.

The DC ports of the RF bias-tees are exploited to both supply the bias to the DUT and allow a path for

the LF noise currents from the latter to the corresponding measurement channel. The RF ports of the bias-tees are clearly ideal open-circuits to the LF noise. Port #1 of the DUT is supplied by means of a conventional, resistive bias network, which can feed the base DC current I_1 or apply the gate-source DC voltage V_1 for bipolar or field-effect technology, respectively. A custom SS amplifier (CBA) based on common-base configuration, low-noise transistors can be inserted at the front-end of the measurement channel in order to provide: *i*) a very low, frequency-constant input impedance throughout the LF bandwidth and *ii*) the separation between the DUT and the input-referred LF noise of TIA₁. This feature of the set-up strongly contributes to improve the accuracy achievable on the estimates of $S_{\Delta i_2}(f)$ and correlation $S_{\Delta i_1 \Delta i_2}(f)$, since in most of the DC quiescent operating conditions in which the DUT is measured the latter works as a SS amplifier for LF noise from port #1 to port #2.

The active bias network is correspondingly exploited at port #2 of the DUT. Its schematic is shown in Fig. 3. An ultra low-noise instrumentation Op Amp (0.15-pA/ $\sqrt{\text{Hz}}$ @1-kHz input-referred noise current) is employed to propagate and force the externally-supplied voltage V_{ctrl} to the port #2 of the DUT (V_2), while a parallel of eight low-noise PNP transistors (2SA1085) feeds the corresponding DC current I_2 , whose level is set by the chosen bias applied at port #1. Thus, no iterative procedure is needed in order to achieve a given bias condition. The parallel is designed to reduce the residual flicker noise introduced by the single PNP BJT by limiting the maximum current flowing in it, while allowing to feed an overall high bias current (up to 800 mA). Up to moderate levels of DC current (10÷20 mA) the active bias network provides a good high impedance to LF noise. At higher currents, instead, the impedance shown by the parallel of BJTs in the LF bandwidth [100 Hz – 100 kHz] decreases to values to be taken into account in the measurement process.

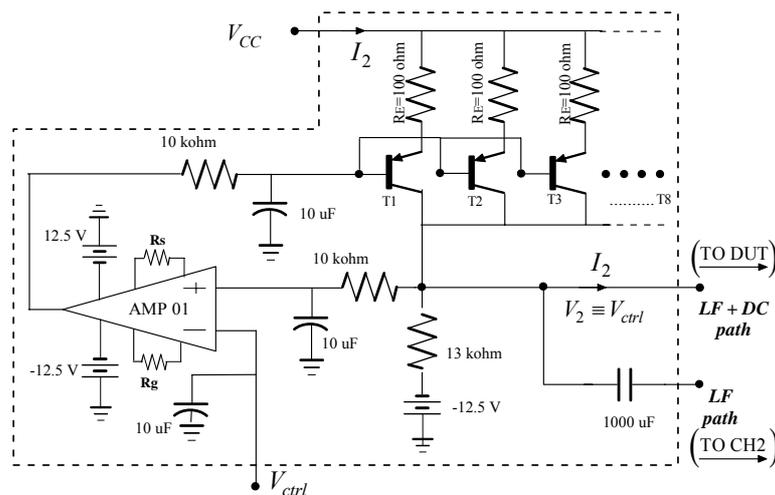


Figure 3. Active bias network topology exploited at port #2 of the DUT.

The drain/collector voltage V_{ctrl} is synthesized within the interval [1,10] V by means of a 12.5-V lead acid DC battery and a precision, wire-wound multi-loop potentiometer. It is worth noting that the DC current I_2 does not flow through such a variable resistor, but instead from the V_{CC} -node. Thus, low-to-high levels of DC current can be supplied to the device port #2 by exploiting a single, safe-level battery, while the measurement process does not depend at all from the variable values set for the potentiometer. Low flicker-noise, thin metal-film technology resistors are used in all the discrete circuits designed, and the set-up is supplied by DC batteries only (apart from the FFT analyzer), including TIAs (SR570).

As introduced in the previous Section, no simplifying hypotheses on the network of impedances seen by the LF noise have been assumed, in order to maximize the overall accuracy achievable and effectively make use of the active bias network also in the presence of high levels of bias current. More precisely, all the possible conductive paths for the noise currents of Fig. 1a and all the non-negligible sources of spurious noise have been taken into account. As far as port #2 is concerned, Fig. 4 shows the equivalent circuit used for the estimation of the actual noise current absorbed by TIA₂. The target noise current Δi_2 is not ideally short-circuited by the input impedance $R_{in2} \neq 0$ of the TIA₂, since a fraction of it flows into the bias-dependent SS output impedance R_{out2} shown by DUT port #2, and another

fraction into the input impedance R_{bias2} of the active bias network. In addition the input-referred, short-circuit LF noise current of both active bias network and TIA₂ loaded on the input (1 MΩ) of the FFT analyzer channel #2 (Δi_{BN2} and Δi_{IN2} , respectively) have to be considered as additive contributions (uncorrelated) to be deperated from the measurement data.

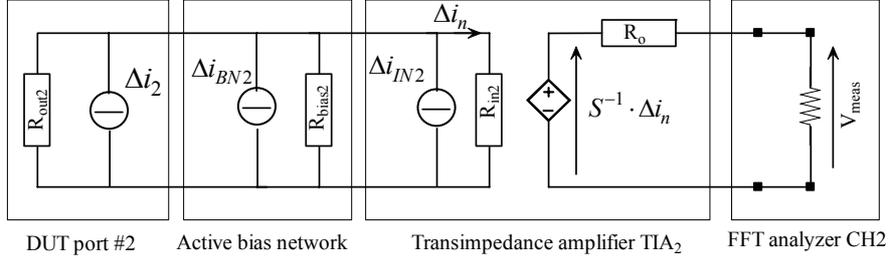


Figure 4. Small-signal equivalent circuit for the characterization of LF noise path at DUT port #2.

The noise current flowing into the TIA₂ is given by:

$$\Delta i_n = k_2 \cdot (\Delta i_2 + \Delta i_{BN2} + \Delta i_{IN2}) \quad ; \quad k_2 = \frac{R_{bias2} // R_{out2}}{R_{in2} + R_{bias2} // R_{out2}} \quad (1)$$

Since the three noise current contributions in Eq. (1) are mutually uncorrelated, it is straightforward to express the relationship between the noise voltage spectral density measured by the FFT analyzer at port #2 and the target auto-spectral density of the short circuit LF noise current Δi_2 at port #2 of the DUT:

$$S_{V_{meas}} = S^{-2} S_{\Delta i_n} = S^{-2} k_2^2 (S_{\Delta i_2} + S_{\Delta i_{BN2}} + S_{\Delta i_{IN2}}) \quad ; \quad S_{\Delta i_2}(f) = S_{V_{meas}} S^2 k_2^{-2} - (S_{\Delta i_{BN2}} + S_{\Delta i_{IN2}}) \quad (2)$$

Thus, for an accurate measurement of DUT LF noise current spectral density $S_{\Delta i_2}(f)$ (and $S_{\Delta i_1 \Delta i_2}(f)$), both the active bias network and the TIA₂ must be characterized in terms of LF noise and input impedance. In particular, two look-up tables for R_{bias2} and Δi_{BN2} are needed by the software associated with the set-up, since these quantities vary with the DC current I_2 supplied to the DUT. As far as the DUT impedance R_{out2} is concerned, the bias-dependent differential response at DC is either usually available, or easily off-line estimated through conventional DC I/V characterization.

As far as port #1 of the device is concerned, the equivalent circuit is shown in Fig. 5.

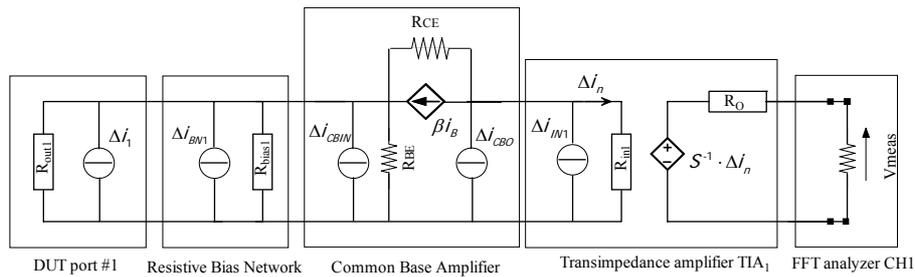


Figure 5. Small-signal equivalent circuit for the characterization of LF noise path at DUT port #1.

To estimate the overall noise current flowing into the TIA₁, it must be considered that the CBA exhibits a unity current gain and a very low input resistance. In the case of a bipolar device, at this port of the DUT R_{out1} is indeed the R_{BE} resistance (DUT in common emitter configuration), which can become quite small when the device is biased at very high current density. For this reason, the CBA with its very low input resistance ($R_{BE} = 8.5$ ohm) is nearly compulsory, since the TIA₁ input resistance is not very small across the bandwidth if compared to R_{out1} . R_{bias1} is a fixed 2-kohm resistance and can be ignored in the computation. Thus, the input current of the CBA is injected nearly unmodified inside the TIA₁, since the CBA current gain is unitary and its output resistance R_{CE} is very high compared to the TIA₁

input impedance R_{in1} (due to the very low CBA collector current: 3.2 mA). The TIA₁ input current Δi_n results

$$\Delta i_n = \Delta i_{IN1} + \Delta i_{CBO} + \Delta i_{CBIN} + \Delta i_{BN1} + \Delta i_1 \quad (3)$$

The two noise current sources of the CBA are correlated, hence in the analysis of the measured noise power spectral density at the FFT analyzer port #1, their cross-spectrum must also be included:

$$S_{V_{meas}} = S^{-2} \cdot S_{\Delta i_n} = S^{-2} \left(S_{\Delta i_{IN1}} + S_{\Delta i_{CBO}} + S_{\Delta i_{CBIN}} + 2 \cdot \text{Re} \left\{ S_{\Delta i_{CBIN} \Delta i_{CBO}} \right\} + S_{\Delta i_{BN1}} + S_{\Delta i_1} \right) \quad (4)$$

Thus, the DUT noise current auto-spectral density at port #1 is given by the expression:

$$S_{\Delta i_1}(f) = S_{V_{meas}} \cdot S^2 - \left(S_{\Delta i_{IN1}} + S_{\Delta i_{CBO}} + S_{\Delta i_{CBIN}} + 2 \cdot \text{Re} \left\{ S_{\Delta i_{CBIN} \Delta i_{CBO}} \right\} + S_{\Delta i_{BN1}} \right) \quad (5)$$

From equations (2) and (5), it is straightforward to compute the current cross-spectrum as a function of the cross-spectrum measured by the two-channel FFT analyzer as

$$S_{\Delta i_1 \Delta i_2}(f) = S^2 \cdot S_{V_{meas} V_{meas2}} / k_2 \quad (6)$$

Equations (2), (5) and (6) accurately describe the LF noise characteristics of the DUT. However, for bipolar devices at very high bias levels the hypothesis of considering as negligible the fraction of Δi_1 that is partitioned within R_{out1} does not hold anymore. In addition, such a noise current is amplified by the DUT current gain β into an additive collector noise. Nevertheless, in such a case the accuracy of the set-up can be preserved by simply taking into account the current partition $k_1 = R_{out1} / (R_{out1} + R_{BE})$ at port #1 within Eq. (3), adding the corresponding collector noise contribution in Eq. (1) and considering the accordingly modified Eqs. (2)-(4)-(6) as a linear system in the unknowns $S_{\Delta i_1}$, $S_{\Delta i_2}$ and $S_{\Delta i_1 \Delta i_2}$.

III. Experimental examples

As an example of the set of data obtained from the characterization of the proposed set-up and exploited in order to accurately estimate the LF noise current power spectra of the DUT, Fig. 5 provides some information about the active bias network. On the left the behaviour of its input impedance with respect to the DC current supplied is shown. Up to moderate current density technologies, R_{bias2} can be considered as an open-circuit; however, for the full characterization of power devices, the dependence $R_{bias2}(I_2)$ must be taken into account in Eq. (1). On the right-side of Fig. 5 the dependence of the noise current Δi_{BN2} spectral density on the bias current I_2 is given.

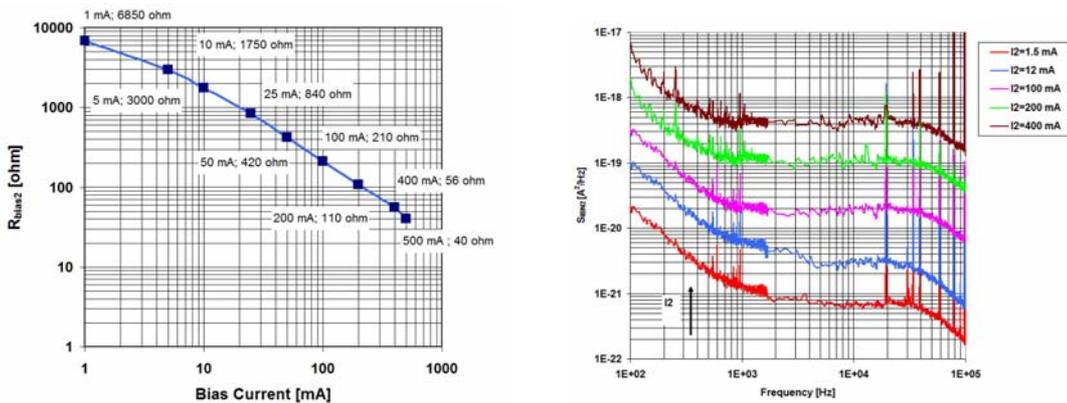


Figure 5. Off-line characterization data for the active bias network: input impedance (left) and input-referred LF noise current (right) vs. the DC current supplied.

Figure 6 illustrates actual LF noise characterization for two different field-effect technologies, typically exploited for microwave power applications: a 1.5-mm total periphery GaAs pHEMT (pseudomorphic High Electron Mobility Transistor) (left) and a 1.2-mm GaN HEMT (right). In both cases the investigation of interest (i.e., $S_{\Delta i_2}(f)$) is successfully carried out from low to high levels of bias drain current. Figure 7, instead, shows measurement data for the LF current spectra (including cross-spectral density) for a GaInP-GaAs 6x30- μ m HBT (Heterojunction Bipolar Transistor).

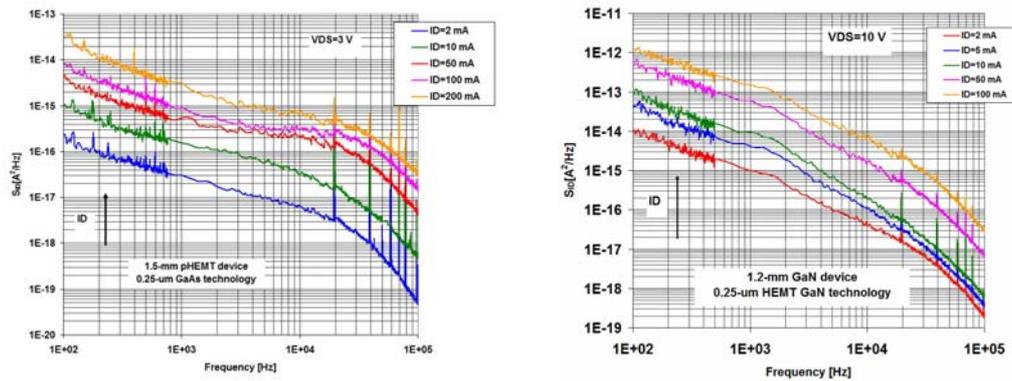


Figure 6. Examples of LF noise characterization for two different field-effect technologies.

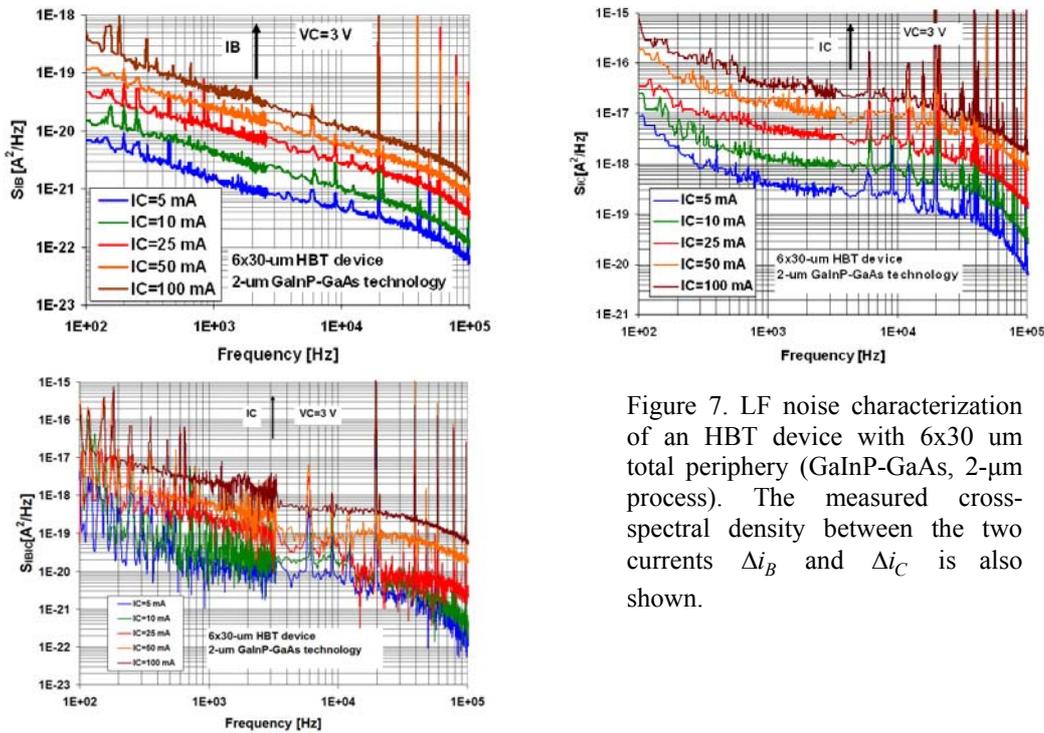


Figure 7. LF noise characterization of an HBT device with 6x30 um total periphery (GaInP-GaAs, 2-um process). The measured cross-spectral density between the two currents Δi_B and Δi_C is also shown.

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