ACCURATE EXTRACTION METHOD FOR 1/f-NOISE PARAMETERS USED IN GUMMEL-POON TYPE BIPOLAR JUNCTION TRANSISTOR MODELS

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ABSTRACT

In SPICE Gummel-Poon models one 1/f-noise source describes the low frequency noise behaviour of bipolar junction transistors (BJTs). In this paper we present a method to extract the respective model parameters from measured 1/f-noise data without the requirement of exactly determining the corner frequency f_c, i.e. the frequency at which the 1/f-noise equals the device noise floor. This novel method is applicable to low-noise as well as to very noisy devices. We present results of different active devices. Verification of the extracted model parameters is done in an oscillator application. Measured and calculated phase noise results agree within measurement accuracy of ±2dB.

INTRODUCTION

Low frequency noise is an important device characteristic with respect to low noise design of linear and nonlinear circuits. In oscillators, for instance, the 1/f- or flicker-noise strongly influences the phase noise of the complete circuit [1], as the low-frequency noise is upconverted to sidebands of the carrier frequency. Therefore, it is essential for efficient CAD to accurately model the noise behaviour of active devices.

In general, the noise behaviour of bipolar junction transistors is determined by various noise sources [2, 3]. If 1/f-noise is concerned, the dominant noise source is located at the base-emitter transition. In most circuit simulation programs this is addressed by the fact that flicker noise is modelled using only one 1/f-noise source in parallel to the base-emitter diode. This is done, for example, in the standard SPICE Gummel-Poon model [4] and in a similar way in the new industrial standard, the VBIC-model [5].

The model parameters describing low-frequency noise are usually extracted from a consideration of the bias dependent corner frequency f_c [6, 7]. This frequency f_c is defined as the point where the „pink“ 1/f-noise equals the „white“, frequency independent, noise floor. If this noise floor cannot be extracted accurately (this is the case especially for devices with high 1/f noise) the extracted parameters become very inaccurate.

In this paper we present a method to extract low-frequency noise parameters without the need to determine the corner frequency f_c. Therefore, it can be employed to any type of bipolar transistors and the extracted parameters are very accurate. We present results for different commercially available transistors. A verification of the method and the extracted parameters is given by the excellent agreement of measured and simulated phase noise of an oscillator with a carrier frequency of 725 MHz.

THEORETICAL APPROACH

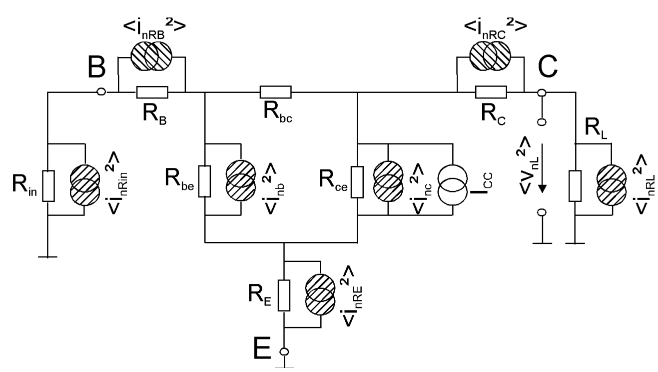

Fig.1: Equivalent circuit of a BJT, including input and output terminations.

Fig. 1 shows the low-frequency equivalent circuit of a BJT including input and output terminations.

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Although several 1/f-noise sources exist within the BJT [2, 3], in the following we consider only the dominant effect in order to be in consistency with the SPICE Gummel-Poon model. Therefore, we assume thermal noise sources in parallel to the resistors R_{in}, R_B, R_E, R_C, and R_L, shot noise sources in parallel to the base-emitter diode and to the current generator I_{CC}, and a flicker-noise source in parallel to the base emitter diode.

All sources are uncorrelated and the rms values of the resulting noise currents are given by

$$\sqrt{\overline{i_{nR_i}^2}} = \left(\frac{4 \cdot k \cdot T}{R_i} \cdot \Delta f\right)^{\frac{1}{2}} \text{ for } R_i = R_{in}, R_B, R_E, R_C, R_L, \quad (1)$$

$$\sqrt{\overline{i_{nc}^2}} = \left(2 \cdot e \cdot I \cdot \Delta f\right)^{\frac{1}{2}}, \quad (2)$$

$$\sqrt{\overline{i_{nb}^2}} = \left(2 \cdot e \cdot I_b \cdot \Delta f + KF \cdot \frac{I_b^{AF}}{f^{\alpha}} \cdot \Delta f\right)^{\frac{1}{2}}. \quad (3)$$

In consistency to the SPICE model α is set to unity. If we concentrate on frequencies below the corner frequency f_c, we can neglect all noise sources besides the 1/f noise source. $\sqrt{\overline{i_{1/f}^2}} = \sqrt{\overline{i_{nb}^2}}\Big|_{f<f_c}$ can be related to the measured noise voltage $\sqrt{\overline{v_{nL}^2}}$ at the load by means of linear circuit theory. In general, this gives a quite complicated relation, which can be found for example in reference [6].

In almost any practical case it will be possible to make the following simplifications. If R_{in} is chosen much larger than R_{be}, which can be estimated by $R_{be} \approx \frac{n_{be} \cdot k \cdot T}{e \cdot I_b}$, we can neglect both R_{in} and R_B. In the forward mode of operation R_{bc} can be set to infinity [4]. R_{ce} models the Early-effect, i. e. base width modulation, and can therefore be estimated to $R_{ce} \approx \frac{VAF}{I_c}$ [4]. The load impedance is given by the input of the noise measurement equipment as R_L = 50Ω. The resistor R_{ce} is usually much larger than R_C+R_E+R_L and can be neglected. As a result we get the equation

$$\sqrt{\overline{i_{1/f}^2}} = \frac{1}{\beta(i_b) \cdot R_L} \sqrt{\overline{v_{nL}^2}}, \quad (4)$$

in which the bias dependency of the low-frequency current gain b is included. Using equations (3), (4) and

$$S_V(f, i_b) = 10 \cdot \log_{10}(\overline{v_{nL}^2}), \quad (5)$$

we get for a fixed frequency f_0

$$AF \cdot \log_{10}(i_b) + \log_{10}(KF) = \frac{S_v(f_0, i_b)}{10} - 2 \cdot \log_{10}\left(R_L \cdot \beta(i_b)\right) + \log_{10}(f_0). \quad (6)$$

A least-squares fit of equation (6) to the linear equation $y = a \cdot x + b$ yields

$$AF = a \quad (7)$$

and

$$KF = 10^b. \quad (8)$$

To determine $S_V(f_0, i_b)$ from the measured noise data, we consider that

$$\overline{v_n^2} \cdot f = \left(R_L \cdot \beta(i_n)\right)^2 \cdot KF \cdot i_b^{AF} = \overline{v_{nL}^2(f = 1Hz)}. \quad (9)$$

As a result we can calculate $S_V(1Hz, i_b)$ as the mean value of $S_V(f, i_b) + 10 \cdot \log_{10}(f)$ within the frequency range from f_{low} to f_{high}. The lower frequency limit f_{low} is given by the measurement equipment as the lowest frequency providing reliable measurement values. The upper limit f_{high} has to be chosen below the corner frequency f_c.

MEASUREMENT SETUP

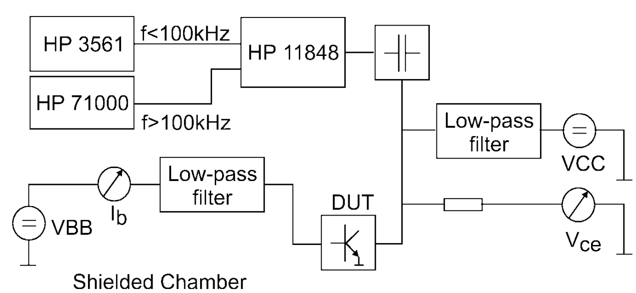
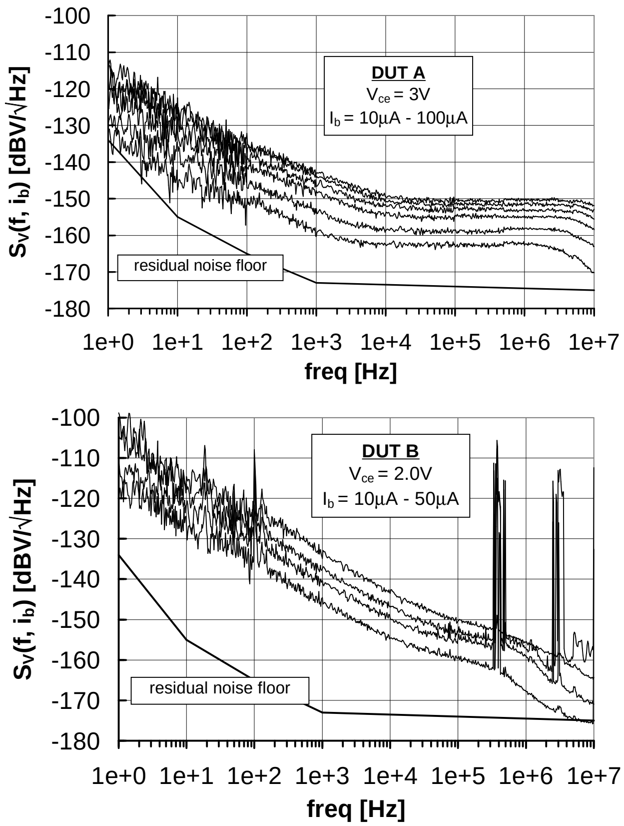


Fig. 2: Low frequency noise measurement setup.

In order to extract the low-frequency noise-parameters, mesurements of the noise voltage power spectral density across the load resistance have to be performed at several bias conditions. Fig. 2 shows the schematic of the measurement setup. Both base current and collector voltage are fed via low-pass filters in order to achieve a low system noise level. The instruments to measure i_b, V_{ce} and i_c are disconnected during the measurements to avoid any influence. It should be mentioned that V_{ce} must be measured directly at the DUT to be sure that any voltage drop across bias resistors is considered correctly. The noise voltage is fed via a blocking capacitor to the baseband noise input of the HP3048 phase noise measurement system [8]. The amplified noise power is then detected using an FFT analyzer HP3561A in the frequency range from 1Hz to 100kHz and using an analog spectrum analyzer HP70000 in the range from 100kHz to 10MHz. By replacing the DUT with resistors, the noise floor of the entire measurement setup is checked. In the frequency range of interest the system noise floor is found to be at least 10dB below the measured noise data of the transistors (see Fig. 3).

EXPERIMENTAL RESULTS



Figs. 3a, 3b: Power spectral densities of the noise voltage for two different transistors.

Figs. 3a and 3b show the measured noise voltage spectra of two different BJTs. DUT A is a BFR93A at V_{ce} = 3V and six different base currents from 10mA to 100mA. The respective collector currents are between 0.8mA and 10.4mA. DUT B is an integrated RF transistor that shows a very high 1/f-noise. Up to the corner frequency at about 5kHz (DUT A) and about 50kHz (DUT B) the 1/f-noise is dominant. At higher frequencies the internal noise floor of the BJT can be seen. The decrease of the measured noise power at frequencies above 1MHz is found to be due to the internal base-emitter capacitance which is in parallel to the noise source. Additionally, the dc-current gain $\beta(i_b)$ has been measured (or simulated if exact SPICE parameters have been available) at the same bias points. According to equation (9) we multiply the measured noise spectra with frequency (see Fig. 4) and calculate the mean values between some Hz and a few kHz to get $S_V(1Hz, i_b)$.

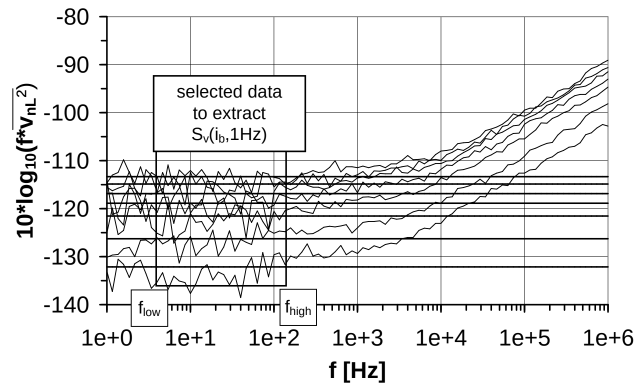


Fig. 4: Extraction of $S_V(1Hz, i_b)$.

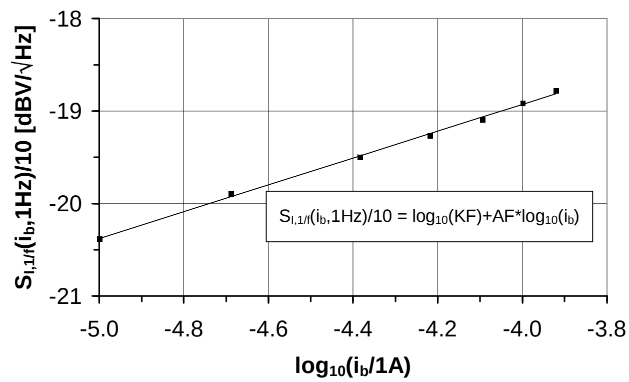


Fig. 5: Extraction of AF and KF.

The results are converted into noise current spectra $S_{I,1/f}(1Hz, i_b)$ by means of equation (6) and plotted against $\log_{10}$(ib). We extract AF and KF performing a least-squares-fit (Fig. 5). Finally, the parameters can be finetuned by simulating the resulting low frequency noise of the transistor. The model of the DUT is embedded into the circuit configuration given during the measurements. The simulated results can then be adjusted to the measured data by optimization of the parameters AF and KF. Our investigations show that only very slight modifications are necessary if good SPICE parameters are available for the DUT.

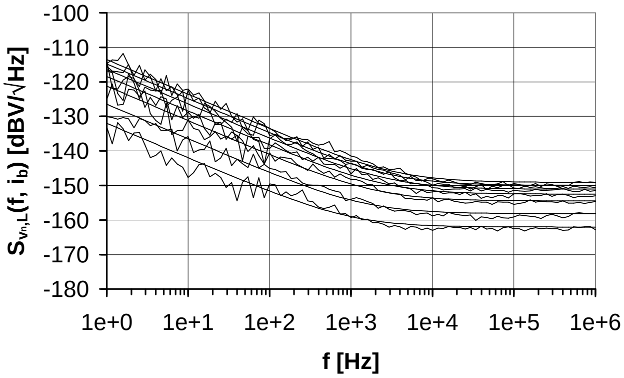


Fig. 6: Measured and simulated power spectral densities of low frequency noise.

Fig. 6 shows a plot of measured and simulated noise voltage spectra of the BFR93A at the bias conditions mentioned before. The results agree very well within the entire frequency range. Table 1 gives parameter values for six different commercially available transistors.

Type	BFR92P	BFR93A	BFR106
AF	1.340	1.478	1.131
KF	64.73e-15	107.4e-15	3.192e-15

Type	BFR183	BFQ81	BFP420
AF	0.9655	1.508	1.665
KF	5.706e-15	490.9e-15	895.7e-15

Table 1: Low-frequency noise parameters for several transistors.

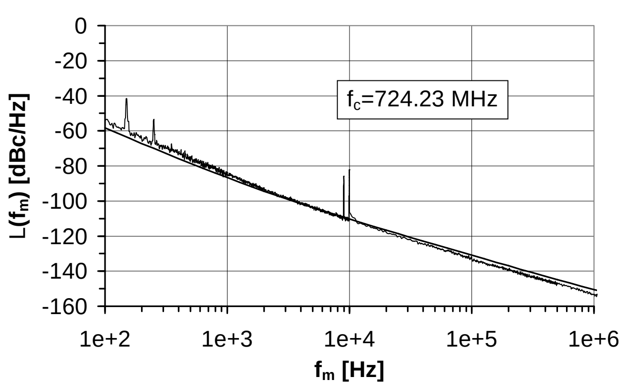


Fig. 7: Measured and simulated phase noise of a 725 MHz Oscillator.

For verification of the noise parameter extraction method, a fixed frequency oscillator at 725 MHz was built with a BFR93A transistor. The output power is -4.1dBm. The oscillation frequency is determined by a a ceramical coxial resonator. Resulting phase noise is measured with a *HP 3048 phase noise measurement system* using the frequency discriminator method. The corresponding simulation is performed with *LIBRA* (Hewlett-Packard). The very good agreement of measured and simulated phase noise data (±2dB) validates the extracted low-frequency noise parameter values (Fig. 7).

CONCLUSION

Modeling low frequency noise of active devices is an important aspect in low-noise CAD of linear and nonlinear circuits. We have presented a generally applicable method to extract the respective parameters AF and KF for Gummel-Poon type BJT models. Our method is based on measurements of the 1/f-noise voltage spectra at the load of a DUT. The extraction uses the 1/f-part of the measured spectra and therefore does not require the determination of the corner frequency f_c. We have presented results for various commercially available BJTs. A verification of the parameters is performed comparing simulated and measured phase noise data of a 725 MHz oscillator.

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