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Abstract:- We present here our initial findings on low frequency noise measurements for shallow emitter poly-silicon contacted transistors. Structures from two particular GEC Plessey Semiconductors technologies were selected for measurement, to assess the change in low frequency noise performance and modelling issues as bipolar technologies move to ever shallower emitter junctions and smaller emitter windows. We also demonstrate an alternative approach to the problem of identifying the intrinsic corner frequency. Details of our measurement technique are given along with the results of some simulations based on SPICE coefficients extracted from these measurements.

INTRODUCTION

‘1/f’ Noise in Bipolar Transistors.

The study of low frequency or ‘1/f’ noise is of importance in many physical systems and particularly in electronics. Bipolar transistors have been used in place of MOS devices where low frequency noise performance is a major concern as they tend to exhibit good low frequency noise spectra. The trend in modern bipolar technologies however is towards shallower emitter junctions and smaller emitter windows. With auto-doping poly-silicon as the material contacting the single crystal. This, coupled with a resultant increase in current densities may have a direct and significant effect on the low frequency noise performance of the latest generation of bipolar transistors. With many designers using bipolar devices in R.F. front ends, particularly for cell phone applications, a study of ‘1/f’ noise measurement and modelling is growing. For the process development engineer too a study of low frequency noise performance can be an aid to understanding other electrical characteristics of a particular bipolar technology.

Sources of ‘1/f’ Noise in Bipolar Transistors.

Unlike other forms of noise within a bipolar transistor which are largely irreversible, ‘1/f’ noise is known largely to be defect density driven and as such reflects the quality of certain interfaces within the device. Generation and recombination events at slow traps within the emitter-base depletion region or changes in diffusion velocity within the bulk emitter will all contribute to the excess noise current at low frequency. Green [1] has specifically studied the low frequency noise spectra of a junction isolated single crystal emitter device, a design which is well suited to minimising ‘1/f’ contributions and one which represents a typical bipolar process of the last generation. The results of Green’s measurements show a strong correlation between the flicker noise and the non ideal component of the base current. It was concluded that the flicker noise could be modelled as a single noise current generator placed between the emitter and base junctions of the device having an intensity

\[
\frac{\overline{I}}{I_f} = K I_g \Delta f / f^\gamma
\]

where the constant K is related to the base-emitter contact area, \(I_g\) represents the non ideal portion of the forward base current, \(\Delta f\) is the measurement bandwidth and \(\gamma\approx 1\). More recent work [2][3][4] has suggested that trapping at residual oxide interfaces between the poly-silicon and single crystal material contributes significantly to the noise current and it is also known from studies of poly-silicon resistors that grain boundary effects in the bulk emitter can also add to the low frequency excess noise. It might be expected then, that a simple model of the form given in Eqn (1) would prove inadequate for modelling more up to date technologies.

NOISE MEASUREMENT SYSTEM

The basic requirements for any low frequency noise measurement system are an amplifier system with suitable gain and a spectrum analyser to cover the frequency range of interest. The problem with looking for ‘1/f’ spectra is that it is all too easy to see the distribution associated with the amplification system and not the device under test. Ideally for a bipolar technology we would want to be able to set the bias point of the device by means of a current source in the base circuit of the DUT, and we would also require that the base-collector voltage be adjustable. The amplifier system chosen was a purpose built low noise unit fabricated by Lancaster University Dept of Physics and based on a design which had been used for their own noise measurements [1][8]. An overall schematic of the amplifier system is shown in Fig(1.0).

Fig(1.0) The low frequency noise measurement and bias system.

The amplifier system consists of a fixed gain (1E+2) low noise pre-amplifier based on a circuit by Stefanovich [8][9] followed by a variable gain stage (1 to 1E+3) [8], the DUT forms an active part of the amplifier. Biasing is either adjusted manually for highest noise resolution or by an automatic system with feedback from the collector of the device, which is connected to a user selected resistor value \(R_c\). The base circuit is decoupled with an R-C network which
can be changed as necessary to alter the input impedance seen by the DUT. The bandwidth of the pre-amplifier was designed at 8 mHz to 25 kHz although in practice this was degraded by parasitic R-C contributions. Test signals could be injected at points of interest within the system to enable an accurate calibration of system gain and noise floor for example. The output from the amplifier system was fed directly to a Stanford Research Systems SR770 FFT Network Analyser. This instrument has a measurement bandwidth of 476 µHz to 100 kHz which is well suited to our area of interest.

**MEASUREMENT TECHNIQUE**

Samples were selected at random from devices which had been bonded out into ceramic DIL packages. No screening was employed other than rejecting devices which exhibited burst noise in the wide-band region of a full (100kHz) spectrum sweep. Devices were biased at the required quiescent current using the manual bias system and allowed to settle for as long as an hour. The battery discharge characteristics were consulted to ensure that the supply was operating during its most stable period of operation. Battery noise generally manifests itself as a ‘faster than 1/f’ component and can usually be easily identified in the spectra. Spectra were sampled in four linear frequency sweeps covering the range 1 Hz to 60 kHz. Each sweep was recorded as a power spectral density (equivalent 1Hz measurement bandwidth per frequency bin) so that the full spectrum could be reconstructed from its constituent parts. For devices with corner frequencies in the kHz range, provision must be made to correct for the upper 3dB point of the amplifier system. The SR770 is a single channel FFT analyser and so transfer functions must be measured and stored in memory before any spectra are collected. These can then be applied via the waveform math functions to correct for amplifier roll-off.

Fig (2.0) shows the combined amplifier and DUT transfer function, obtained by injecting a ‘chirp’ source (from the FFT analyser) into the base circuit of the transistor. The transistor was biased at the base current of interest and after a settling period the raw spectra were measured and corrected by multiplication with the relevant transfer function. Unfiltered data was processed using the S.A.S. software suite to remove any periodic content from the trace.

**MEASUREMENT INTERPRETATION**

Fig (3.0) shows a typical raw low frequency, four part noise spectrum for an NPN device biased at a base current of 3μA. Fig (3.1) shows the same spectrum after filtering out 50 Hz harmonics.

Features of interest are the flat noise floor to the high frequency end of the plot, the knee region and below this we would wish to identify a region where the noise intensity increases with a gradient of close to 10db per decade of frequency ($\lambda \approx 1$ in Eqn (1)). Below this region the noise intensity rises at a rate progressively faster than this and is believed to be due to battery noise. As we move towards lower frequencies the spectra become increasingly excursive due to the necessity of reduced averaging. A useful figure of merit is the so called ‘corner frequency’ defined either as the frequency at which the noise intensity is 3dB above the thermal noise floor, or the intersection of the ‘1/f’ line with the extended thermal noise floor. Fig (4.0) illustrates the identification of the ‘1/f’ portion of a noise spectra and the extraction of the associated corner frequency. Costa et al [4] distinguish between the measured corner frequency $F^c$ and the true ‘intrinsic’ corner frequency $F^c$. SPICE models the ‘1/f’ component for bipolar transistors using a current generator having an intensity given by:

$$\left(\frac{I_n}{I^F_b}\right)_{1/f} = KF I^F_b \Delta f / f$$

(2)
The ‘intrinsic’ corner frequency is defined as that frequency at which the intensity of the ‘1/f’ noise is equal to that due to the base shot noise alone. It is this value that we should use for extracting the SPICE parameters $K_F$ and $A_F$. Unfortunately other noise sources will contribute to the thermal noise floor and so any extraction of $K_F$ and $A_F$ will be in error. Costa [4] calculates a correction factor for $F_c$ in terms of the internal device impedances and other thermal noise sources as:

$$F_c = F_c' \left[ a I_b + b C_c V_{BB} + d C_s G_0 \over a I_b \right]$$

where the quantities ‘a’ through ‘g’ are complex transimpedances given in terms of the conductances of the transistor equivalent noise circuit shown in Fig (5.0). Analysis of this corrective function is exceedingly complicated and requires that the transistor be restricted to a region of operation that allows considerable simplification of Eqn (3).

$$\frac{d}{d I_B} \log(F_c) = A_F - 1$$

$K_F$ can then be obtained by taking the intercept value of $\log(F_c)$ and computing:

$$K_F = 2q \cdot 10^{\text{intercept}}$$

**MEASUREMENT RESULTS**

We chose to measure devices from two distinct technologies, a device from a 1µm (as drawn) self aligned bipolar technology (HE) denoted wb10 and a device from a pre-production shrunk version of HE with an (as drawn) emitter width of 0.6µm (process HG device yf4). These two devices represent typical transistor structures from the design kits available for both technologies, and were chosen to cover either end of the low frequency noise performance spectrum. Process HE was known to exhibit good mid band noise performance and the device chosen was known not to be excessively noisy. The device chosen from process HG was one which would maximise any ‘1/f’ contribution and pose the greatest challenge for parametric extraction and modelling. Further, the HG devices measured were from a series of experimental batches which had non-standard final RTA schedules. This was thought to have resulted in residual oxide in the emitter. Fig(6.0) and Fig(7.0) show Gummel plots of typical samples of both devices.
It can clearly be seen that the yf4 device shows a marked non-ideal portion of the base current extending far into the near-ideal region, whereas the HE device shows the more usual low current degradation of the base current with a much smaller recombination component. Fig(8.0) and Fig(9.0) show low frequency noise spectra for each device at different base currents. Some of the traces are composite traces from several devices (of the same type). The reproducibility was generally good enough for this not to degrade the trace significantly.

Fig(10.0) shows an example of parametric extraction of $K_F$ and $A_F$ for a sample wb10 device. Here we are plotting the log of the extracted corner frequency against log$(I_B)$ for five bias points. The fact that we observe a good degree of linearity would indicate that our extracted $F_c$ values are close to the intrinsic $F_c$ values and that any correction would be slight. It can be seen that the corner frequencies for our HG samples are much higher than those for the HE samples. Extraction of the $K_F$ and $A_F$ parameters for both sets of results reveals a range of values for $K_F$ similar to those seen by Quon et al [2]. For the wb10 device we have $A_F=1.4$, $K_F=7.6e^{-15}$ and as expected at the other end of the spectrum we have the yf4 device with $A_F=2.0$, $K_F=5e^{-9}$. It must be stressed however that both these samples were chosen so that they would represent either end of the noise performance spectrum.

Noise simulations were performed using $K_F$ and $A_F$ parameters extracted as detailed. All simulations were conducted within the Cadence DF II environment using the SpectreS simulator. All other device parameters within the model having been derived by a combination of D.C., noise ($rb$ verification) and ‘S’ parameter measurement. The purpose of the simulation is two-fold. Firstly we chose to simulate the whole test harness with the DUT as an integral part of the measurement system. This means that we do not have to worry about the operating point of the transistor or calculation of transimpedances. All we require is that the amplifier system be accurately characterized. It is then possible to compare measurement data directly to simulation results. Any error in our values for $A_F$ and $K_F$ can then be corrected by allowing optimization. This is allowable because the model we are fitting is behavioural and not based on any real measurable entity. Secondly we can come to some decision as to whether the simple model with its fixed ‘$1/f$’ distribution is valid. Fig(11.0) represents our test system with bias resistor network, decoupling and a lumped amplification stage. Fig(12.0) shows our simulated yf4 data plotted on the same axes as measured data for two bias points. It can be seen that there is good agreement between the simulated and measured data in terms of the observed (measurement) corner frequency and the ‘$1/f$’ portion of the spectra. The discrepancy between the flat ‘thermal’ regions is a result of errors in other model parameters, most probably the $rb$ parameter.
CONCLUSIONS

Flicker noise parameters have been extracted for two advanced Bipolar Transistor technologies. The technique does not depend on tight control of the operating point and input/output impedance of the device and the subsequent calculation of many complex terms to effect a correction to the extracted corner frequency. We choose rather to simulate the noise spectrum of the device ‘in harness’ and optimize against measured data. Although we have chosen to model our worst case scenario for a modern bipolar transistor process (poly-silicon emitter, high current density and residual interface oxide) the simple SPICE model with a single noise current source seems sufficient.

REFERENCES


Fig(11.0) Noise circuit for simulation.

Fig(12.0) Simulated and measured spectra.